



IRE Convention Record

Part 7—

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Transmitters, Receivers, and Audio

SESSIONS ON . . .

Broadcast Transmission Systems I — TV Broadcasting

Audio I — General

Broadcast Transmission Systems II — Color Television

Audio II — Symposium: Music, High Fidelity, and
the Listener

Audio III — Seminar: Magnetic Recording for the
Engineer

Symposium on Spurious Radiation

Broadcast and Television Receivers

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Audio

Broadcast and Television Receivers

Broadcast Transmission Systems

Presented at the IRE National Convention, New York, N.Y., March 21-24, 1955

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IRE CONVENTION RECORD
1955 NATIONAL CONVENTION

PART 7 - TRANSMITTERS, RECEIVERS, AND AUDIO

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SYNCHRONIZATION OF MULTIPLEX SYSTEMS FOR RECORDING VIDEO SIGNALS ON MAGNETIC TAPE

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Summary

The over-all and inter-channel timing accuracy requirements for time-division-multiplex, magnetic-tape recording systems are discussed. The effects of tape flutter and skew on timing accuracy are considered, and methods for achieving good synchronization despite flutter and skew are described. The effects of inadequate synchronization on the signal output of the system are shown.

Introduction

Consideration of the many factors involved in the design of magnetic recording systems for operation in the megacycle frequency region led to the choice of a time-division multiplex system. In this type of system, information contained in the wide-band signal to be recorded is distributed by a time sampling process among a number of narrow-band channels, the outputs of which are then recorded on a corresponding number of parallel magnetic tape tracks. A discussion of the basic theory of operation and channel response requirements is given in a companion paper.⁽¹⁾ The present paper is concerned primarily with the proper synchronization of time division multiplex systems.

Basic System

A block diagram of an eight channel multiplex tape recording system illustrative of the techniques to be discussed is shown in Figure 1. The selection of eight channels for discussion is arbitrary, and no particular restriction is imposed on the number of channels except that the complexity of equipment required and the difficulty of maintaining channel uniformity makes a system with a very large number of channels impractical.

The input signal is passed through the Input Low-Pass Filter, having a cut-off frequency, f , corresponding to the desired upper frequency limit to be recorded. The signal output of this filter is then fed simultaneously to eight Sampler channels which sequentially sample the signal at a channel sampling rate of $2f/8$. The channel samples are then passed through the Even-Ringing Filters having a cut-off frequency of $f/8$ and into the corresponding eight channels of the Tape Recorder. On playback each of the recorded channel signals is sequentially sampled in the Resampler, combined in the Adder, and filtered by the Output Low-Pass Filter to restore the original bandwidth limited waveform.

Successful operation of such a system requires exceedingly precise synchronization of

the sampling and resampling process. Basic synchronization is provided by the Sync Generator unit which generates time reference signals for the Sampler and Resampler. The Resampler synchronizing signal from the Sync Generator is recorded on a separate tape track, at some frequency, such as $f/12$, which is sub-harmonically related to the channel sampling frequency $f/4$, since the highest frequency that can be recorded in a channel is $f/8$. The synchronizing signal from the tape is multiplied in the Resampler Synchronizer to the channel sampling frequency, $f/4$, for synchronization of the Resampler. Two relatively-low-frequency reference signals are also provided by the Sync Generator for recording on the two outside tape tracks to provide tape skew compensation as described later.

Channel Timing Relationships

The sampling process which takes place in the Sampler unit is shown in Figure 2. At the top of the figure a square-wave input signal is shown, which for convenience only, is synchronously related to the sampling rate. This signal contains no frequency components higher than f cps. The samples, therefore, occur at time intervals of $1/2 f$ seconds, and some 24 successive samples are shown in Figure 2. Although ideal narrow samples are shown in the figure, the actual sample width in practice may occupy a substantial portion of the period $1/2 f$ seconds without serious degradation of high frequency response. It can be seen in Fig. 2 that the sampling pulses are allocated sequentially to the various channels. Thus, Channel 1 receives samples 1, 9, 17, -----; Channel 2 receives samples 2, 10, 18, -----; etc. The time interval between samples in any channel is accordingly $8/2f$ sec. As is shown in the previously referenced paper⁽¹⁾ the minimum bandwidth required for accurate transmission of the channel sampling information through the tape recorder is $f/8$ cps. To cite an example, in an eight-channel system designed to record 4 mc the high-frequency response limit of each tape channel must be at least 500 kc. Correspondingly, the time between samples of the input signal will be 0.125 microseconds, and the time between individual channel samples will be 1.0 microseconds.

The waveforms of the channel samples is altered in passing through the Even-Ringing Filters prior to recording on the tape. Fig. 3 shows the response of a typical channel to a step function applied to the input of the system. The channel samples, which occur at intervals of $n/2f$ seconds (where n is the number of channels, and f is the

highest frequency component of the input signal), are passed through the Even-Ringing Filters and produce the waveform shown at the bottom of Fig. 3. Note that the output of the channel filter is also a step function, but since it contains no frequencies higher than f/n cps, the rise time is greater than the system input function, and has both a leading and a lagging transient oscillation of period n/f seconds.

Synchronizing Accuracy Requirements

To reconstruct the original signal from the recorded samples requires that each of the tape channels be sampled precisely at the point on the tape corresponding to the correct instantaneous amplitude of the desired samples. This process of sampling the output of the tape channels will frequently be referred to in this paper as "re-sampling." It is not a sufficient condition in itself that each tape channel be resampled at the proper point; it is required in addition that the resulting resamples which are combined in the Adder unit be spaced exactly $1/2f$ sec. from channel to channel. Due to the mechanical vagaries of the tape medium, these two requirements may not be satisfied simultaneously without the corrective measures to be discussed.

Re-sampling mis-timing in any channel results in an error signal at the output of the system, characterized by spurious frequency components at the channel sampling frequency and integral multiples thereof. In some recording applications, such as television, where there is usually a synchronous relationship between the system input signal and the sampling frequencies, re-sampling time errors may result in the formation of "ghost" images of the desired signal. To gain some appreciation of the magnitude of the electrical and mechanical problems associated with accurate re-sampling, consider the case of a system designed to record 4 mc. In this system two successive samples are only 0.125 microseconds apart in time, which at a typical tape speed of 100 inches per second, means that the two successive samples are spaced only 12.5 millionths of an inch (about 0.3 microns) apart in the direction of tape travel. Experimental and analytical evidence indicates that for satisfactory system performance individual samples should have a time-position error (both with respect to their recorded position on the tape, and in actual playback time) less than $n/20f$; or less than 0.05 microsec. in a 4 mc system.

In practical tape recording systems three major sources of timing error have been encountered: first, substantial fixed differences in delay time from channel to channel may exist, particularly where separate recording and playback heads are involved; second, there are time-varying delays which differ from channel to channel due to the effects of tape skew; and third, there are time-varying delays which are the same in each channel due to tape speed variations (flutter). These time-error sources and methods for compensation of them will now be discussed.

In a time division multiplex system of the general type under discussion in this paper there is considerable fixed time delay in each channel. It is requisite to satisfactory operation that the total time delay of all channels be the same within a very small fraction of the channel filter ringing period, n/f sec.

The effect on the output of an eight-channel system due to a single mis-timed channel is shown in Fig. 4. The input signal to the system is shown on the top line of the figure, followed by the correct values of the re-samples of Channels No. 1 and 2. The waveform of the channel playback signal is shown by the dashed-line curves passing through the peaks of the re-samples. Channel No. 3 has a time-delay that is $1/f$ sec. less than the other seven channels. Therefore, the correct amplitudes of the Channel No. 3 re-samples occur $1/f$ sec. earlier than at the points where this channel will actually be re-sampled; the resulting incorrect Channel No. 3 re-samples are shown on line five of the figure. The addition of all channel samples results in the pulse train shown at the bottom line of the figure, and the system output after filtering will have the waveform shown by the dashed-line curve connecting the samples. The effect of the Channel No. 3 mis-timing is observed as an output amplitude error recurrent at the channel sampling rate.

Two major sources of fixed differences in channel delay time are encountered in practical system design. First, the normal tolerances specified for circuit elements, particularly those embodied in the channel filters, account for appreciable time-delay differences between channels. Second, small misalignments of the individual recording and playback heads in multichannel head assemblies normally exist. Of these two sources of delay differences, head misalignment is by far the most significant, especially when separate recording and playback heads are employed.

A typical multichannel recording head assembly is shown in Fig. 5. This particular assembly contains 20 individual ferrite-cored heads with mu-metal shielding between each head. These heads are designed to provide satisfactory recording of wavelengths as short as 0.0005" (0.5 mils), and therefore when used at a tape speed of 100 inches per second render useful channel response to about 200 kc. In the process of assembly of such heads great care is taken to insure that each gap is mechanically as well aligned as possible with all the others in the assembly. Individual head alignment can be held to within ± 0.1 mils of the center line of the gaps. At a tape speed of 100 i.p.s. 0.1 mil represents 1 microsecond of time difference, and it is therefore possible for two adjacent heads in such an assembly to have a time-position difference of as much as 2 microseconds with respect to each other. The possible total time difference between channels is increased to about 4 microsec. when a separate playback head built to the same toler-

ances is employed. In actual operation even greater differences in head delays than can be explained by mechanical misalignment of the gaps are found to exist. In some cases measured channels delay differences could be explained only by adding the delay corresponding to the thickness of the gap spacer to the normal alignment tolerance. It was concluded that there is no certainty as to which edge of the recording head gap will define the remanent field on the tape, this apparently being a matter of the relative sharpness of the two gap edges.

Fixed, unequal channel delays such as described above can be satisfactorily compensated for by installing adjustable delay lines in each tape channel. The amount of delay added to any channel by this method is equal to the difference between the delay of that channel and that of the channel having the greatest amount of delay.

Inequalities in Channel Delay Due To Tape Skew

Multiplex tape recording systems of the type covered by this paper are subject to time-varying inter-channel timing errors arising from small angular changes in the position of the tape with respect to the recording and playback heads. This effect is shown in Fig. 6. Two possible extremes of skew angle of the tape are shown at the top of the figure, while below is shown the desired skew-free alignment of tape and head. A number of tape transports designed for moving tape at speeds from 100 to 200 i.p.s. have exhibited maximum skew angles of the order of ± 0.02 degrees (0.35 milliradians) from the desired perfect alignment. In terms of time variation between the center and outside edges of a $\frac{1}{2}$ "-wide tape running at a speed of 100 i.p.s. the above amount of skew will cause nearly ± 1 microsec. change in effective time delay of a tape track near an outside edge of the tape with respect to a track in the center of the tape. In view of the channel timing accuracy requirements given earlier in this paper, it is evident that the above skew error is completely intolerable.

The effect of tape skew on the time position of the channel samples is shown in Fig. 7 for an eight-channel system. The no-skew condition is given at the top of the figure. The dots indicate the time location on the tape of the correct values that should be re-sampled in each channel. Two extreme conditions of skew are shown in the center and bottom portion of the figure, where in one case skew has approximately doubled the time between the tape channel samples, and in the other case all eight channel samples occur almost simultaneously.

In Fig. 8 is a graphical representation of the effects of skew on a step-function input signal to the system. Actual time positions and amplitudes of the correct samples on the tape are shown for each channel, and the dots represent the times at which re-sampling actually takes place in each channel. The pulse train which results from adding the individual channel re-samples is shown near the bottom of the figure, as is the output

signal from the system after filtering. The waveform of the input signal has been almost completely destroyed by a train of leading and lagging error signals.

Several methods are available for correction of the variation in channel time delays due to skew. One method, which is completely electronic, is to store the correct values of the channel samples in box-car storage circuits, and read them out in correct time sequence and spacing. Another method is to servo control the angular position of the playback head assembly with respect to the tape such that the time positions of the channel samples are correctly maintained. The latter method will be described in this paper.

Playback Head Skew Servo

Servo control of the angular position of the playback head involves movement of the fairly massive playback head assembly at tape skew frequencies, and it is therefore requisite to determine the frequency spectrum of tape skew. Fig. 9 shows a plot of relative skew amplitude of a typical tape transport versus frequency. The larger amplitude components of skew are seen to be concentrated in the region below about 25 cps, with measurable skew components extending as high as 500 cps. Therefore, the angular position of the playback head should be servo-controlled over a corresponding range of frequencies. Successful operation of a head servo as a method of correcting channel delay errors due to skew assumes that the time delay variations in any channel are proportional to the distance of that channel from the center of the tape. Measurements have shown that this assumption is valid, at least to an excellent first approximation, and the tape can be considered as a rigid web which maintains its shape during the skewing process.

A block diagram of a playback head skew servo is shown in Fig. 10. For operation of this system sinusoidal reference signals are recorded on the two outside tape tracks. On playback the two reference signals are compared in phase to provide the necessary error signal for the servo. The error signal is amplified in a dc amplifier and fed to an electro-magnetic driver unit, which later rotates the head assembly about the center of the tape in such a direction that the phase error of the reference signals is reduced. Thus, the playback head assembly is made to follow the skewing of the tape, and hence, to preserve correct channel time delays, at least within its performance limits.

The effectiveness of the head servo in reducing timing errors due to skew is illustrated by Fig. 11. In this figure are shown actual oscillograms of relative skew error in a tape transport with and without the head servo in operation. A reduction in timing error of more than 10 to 1 is indicated, particularly for the lower-frequency skew components. By means of such a servo it has been possible to reduce peak channel timing errors

due to skew to about ± 0.1 microsec., which was adequate for satisfactory operation of the multiplex systems with which it was employed.

Time-Varying Channel Delays Due To Flutter

In time-division multiplex systems variations in tape speed (flutter) during the recording and playback process have the effect of producing time varying channel delays which are the same in all channels, since it can be visualized that the samples recorded on the tape arrive at the playback head either earlier or later than their correct time position, depending upon the instantaneous value of flutter. Considered from a more-familiar point of view, flutter causes the recorded information in each track to be frequency modulated at the flutter frequency. Accurate re-sampling of the tape in the presence of flutter requires that the re-sampler unit be synchronized from a signal which is frequency modulated in an identical manner to the channel information. Great care must be taken to insure that the phase frequency characteristic of the synchronizing signal channel is identical with that of the information channels; otherwise, the re-sampler will not properly track the recorded samples in the presence of flutter.

In a properly synchronized time division multiplex system flutter has the same effect on the output signal as in a single-channel recorder, in that the output signal is frequency modulated. The allowable flutter of a tape transport depends upon the nature of the signal to be recorded. For recording many types of signals a peak-to-peak tape flutter amplitude of 0.1 percent is acceptable. Flutter is defined here as the ratio of the speed variations to the average tape speed. For recording television signals, especially if industry standards for frequency stability are to be met, flutter will have to be several orders of magnitude better than the above figure. It may prove more practical to meet the television flutter requirement by electronic compensation methods rather than by exhaustive machine techniques.

Fig. 12 is an oscillogram showing the flutter characteristics of one 100 i.p.s. tape transport developed for recording video frequencies. Peak-to-peak flutter is seen to be approximately 0.07 percent. However, this oscillogram was made with a pen recorder which does not respond to flutter components above about 100 cps. Total peak-to-peak flutter of this machine as observed on an oscilloscope where all frequency components up to several kc were included was about 0.1 percent.

The synchronization problem is then one of insuring correct re-sampling of the output of each tape channel in the presence of flutter. The frequency at which the signal channels must be re-sampled is higher than can be directly recorded on a single channel (assuming all channels have the same bandwidth limitation); hence, it must be derived on playback from a sub-harmonic

of the channel re-sampling frequency. Although a recorded sub-harmonic which is one-half the channel re-sampling frequency is within the frequency range of the recording channels, it has been more advantageous to use a frequency which is one-third the channel re-sampling frequency, since signals recorded near the upper frequency limit of a channel exhibit greater amplitude modulation than those recorded in the middle range.

Several methods of deriving the channel re-sampling signal from the recorded sub-harmonic frequency have been used, and two of these are shown in the block diagrams of Fig. 13 for an eight-channel multiplex system.

In the multiplier-type of re-sampler synchronizer shown in Fig. 13a, the sync signal from the tape recorder of frequency $f/12$ cps is fed through a continuously variable delay line to an amplifier-clipper stage. The delay line is variable over a range of at least $4/f$ seconds, which corresponds to the period of the channel samples, and permits exact adjustment of the phase of the re-sampling pulses with respect to the recorded channel samples. After amplification and clipping to remove amplitude modulation, the sync signal is passed through a band-pass filter and amplifier tuned to the third harmonic of the sync signal, thus forming the required $f/4$ cps re-sampler synchronizing signal. This type of re-sampler synchronizer can be made to give an output signal which exactly follows the flutter present on its input signal, but very careful filtering techniques are required to remove the sub-harmonic sync signal from the output of the synchronizer, and to achieve satisfactory phase delay characteristics.

The locked-oscillator type re-sampler synchronizer utilizes a reactance-tube-controlled oscillator operating at the channel re-sampling frequency $f/4$ cps. The output of the oscillator is coupled to a buffer amplifier which supplies the synchronizing signal to the re-sampler, and also drives a frequency-divider, which operates at $1/3$ the re-sampler frequency, or the same frequency, $f/12$ cps, as the sync signal from the tape recorder. The sync signal from the tape recorder is fed through a variable delay line, of the same type described above for the multiplier-type synchronizer, and after amplification is compared in a phase detector with the output of the frequency divider. The resulting error signal from the phase detector is amplified and applied to the reactance-tube to control the frequency of the oscillator. One advantage of this type of synchronizer is that its output signal can be kept very free from sub-harmonic content. The locked oscillator type of synchronizer does not directly follow frequency changes caused by flutter, due to low-pass filtering action in the phase detector and reactance-tube control circuit. This is equivalent to a time delay, and if the cut-off frequency of the frequency control circuit is not substantially higher than any flutter frequencies which may be present, serious re-sampling time errors will result.

The effect of an error in re-sampling phase on a step function system input is shown in Figure 14. In this figure the time positions and proper amplitudes of the correct samples on the tape are indicated for the eight channels, and the dots indicate the time at which re-sampling actually occurs due to incorrect re-sampling phase. The envelope of the pulse train resulting from adding the channel re-samples is shown at the bottom of the figure and corresponds to the system signal output. It will be noted that the desired single transition is altered by the addition of alternating error ("ghost") signals which both lead and lag the desired transfer time. Since flutter which is incorrectly tracked by the re-sampling synchronizer results in instantaneous re-sampling phase errors, its effect on the output signal will be to cause ghost-type errors of a complex and continually varying character at each change in signal level.

As an example of how well the requirements for accurate re-sampling synchronization can be met in the presence of flutter, a typical re-sampling synchronizer of the multiplier-type exhibited a re-sampling time error of only ± 0.1 microsecond for $\pm 1\%$ simulated flutter magnitude.

Performance of Practical Systems

Advanced development work has been done on a number of multiplex tape recording systems which utilize the synchronizing techniques discussed in this paper. These systems have been intended primarily for military applications, and bandwidths have ranged from 1 to 3.4 mc.

Figs. 15 and 16 show, respectively, the square-wave and pulse response of a 6-channel, 1.1 mc magnetic recording system, at various square-wave frequencies and signal pulse widths. The non-ideal input pulse shapes shown in the figure are the result of limiting the frequency com-

ponents of the input signal to 1.1 mc. It can be observed that the recording system causes relatively little degradation of rise times, and the resolution of narrow pulses is very close to the ideal limit for a 1 mc system. Such a system is useful for recording many types of pulse and communication signals. The "grass" on the output signals gives indication of a fairly limited signal-to-noise ratio, but much further improvement is achievable within the framework of the basic techniques which have been shown.

The type of multiplex system described in this paper is designed for recording any type of random signal within the useable frequency range of the system. Television signals have been recorded, but picture quality was not up to broadcast standards. As was indicated above, most of the General Electric major development effort to date has been directed toward the military requirements for wide-band tape recording systems rather than toward television, but the same basic techniques are applicable to either field of use.

Acknowledgment

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References

1. B. G. Walker, "Channel Response Requirements of Multiplex Systems for Recording Video Signals on Magnetic Tape," Convention Record of the I.R.E. 1955.

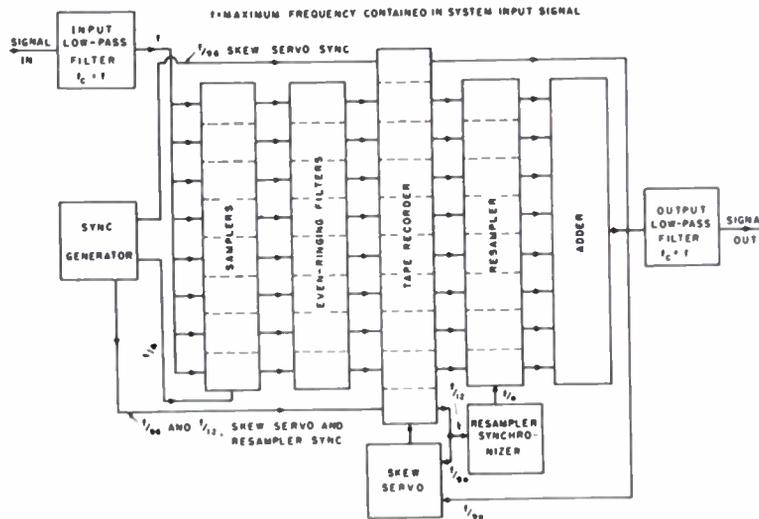


Fig. 1

Block Diagram of Eight-Channel System

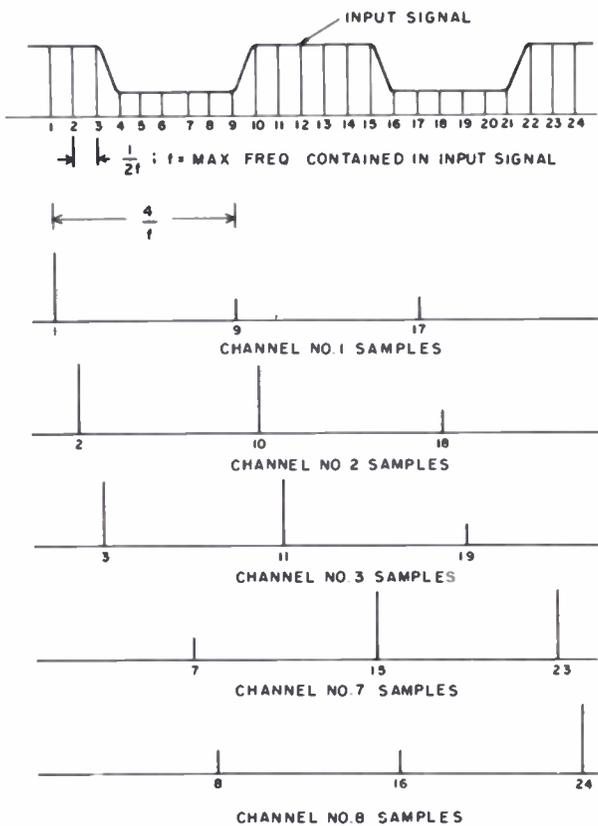


Fig. 2
Time Relation of Channel Samples in
Eight-Channel Multiplex System

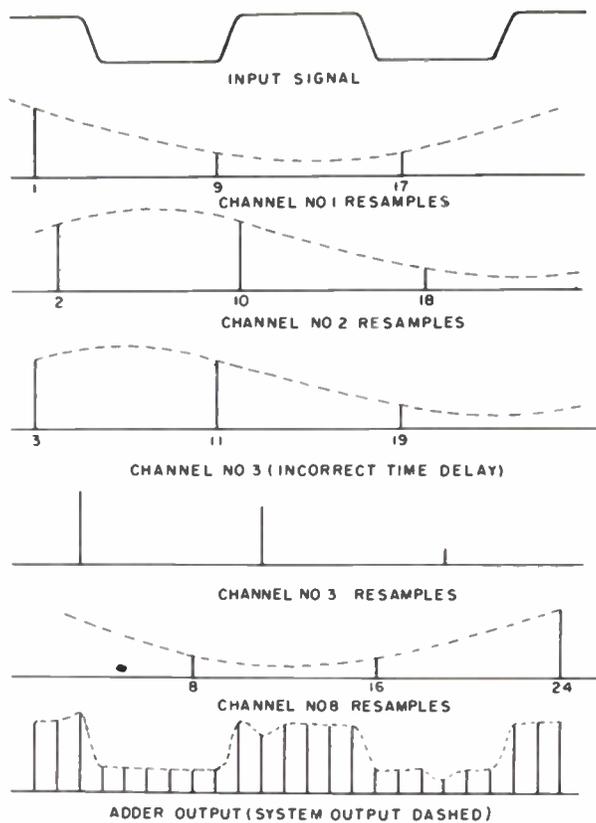


Fig. 4
Effect of Unequalized Channel Time Delays

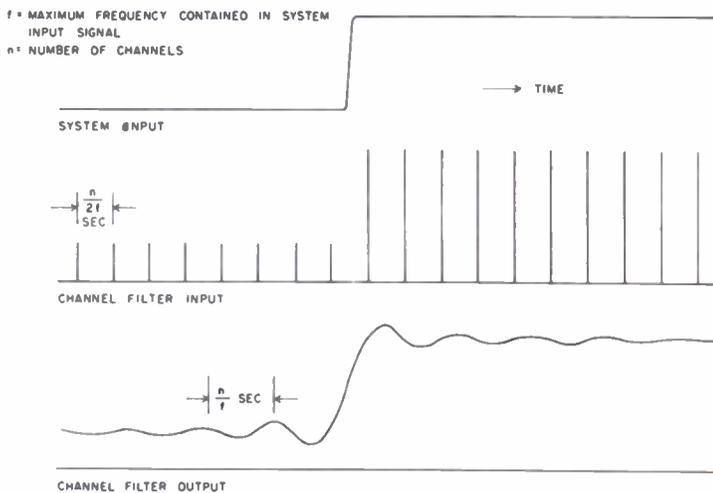


Fig. 3
Response of Channels to Step-Function System
Input Signal

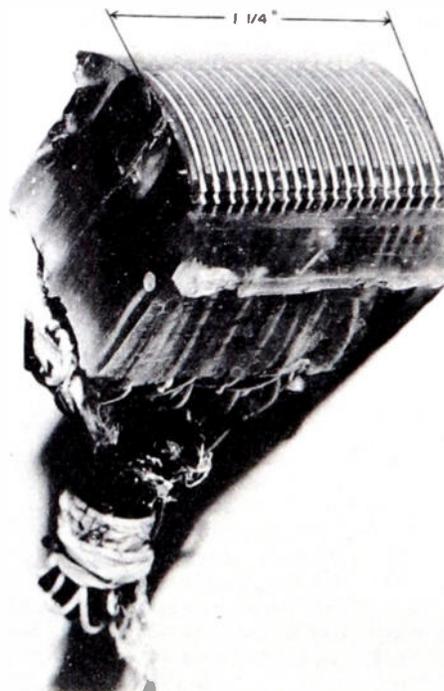
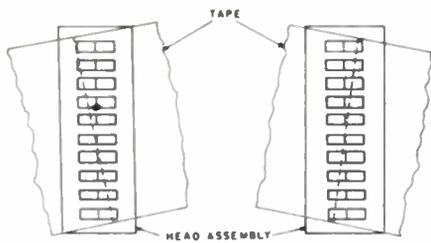
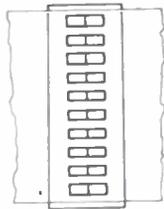


Fig. 5
Typical Wide-Band Multichannel Recording
Head Assembly



EXTREME CONDITIONS OF SKEW
(ANGLES GREATLY EXAGGERATED)



NO-SKEW CONDITION

Fig. 6
Skew of Tape Relative to Heads

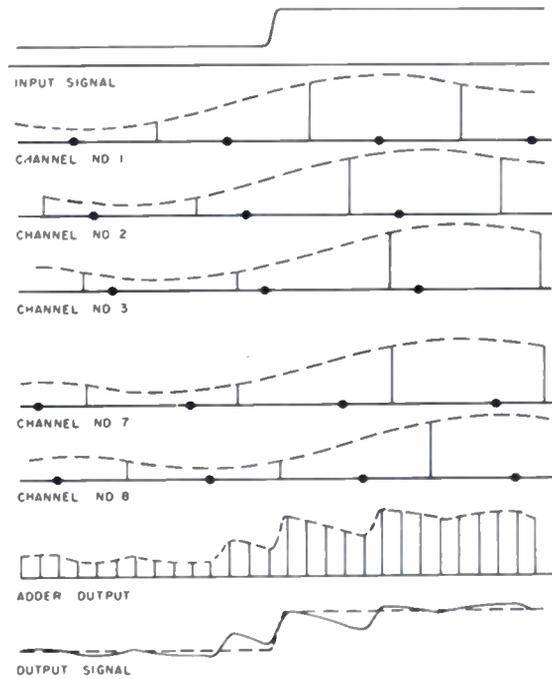


Fig. 8
Effect of Skew on System Output Signal

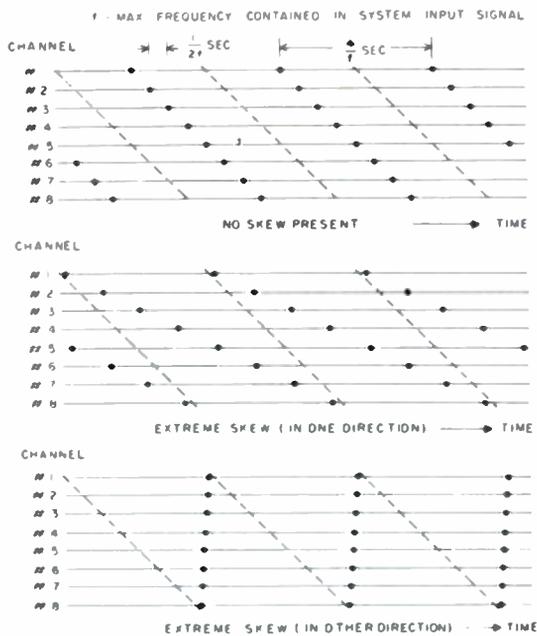


Fig. 7
Time Position of Channel Samples in
the Presence of Skew for
Eight-Channel System

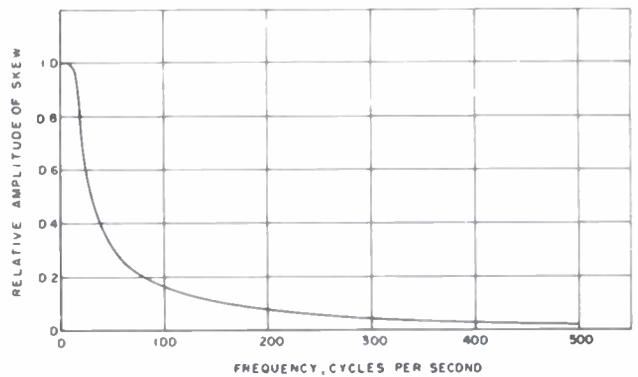


Fig. 9
Frequency Spectrum of Skew

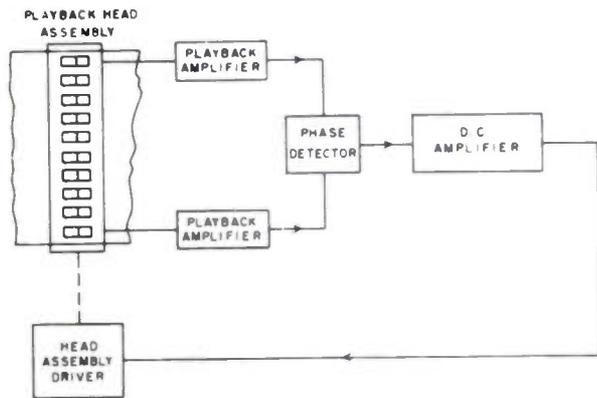


Fig. 10
Playback Head Servo Block Diagram

10

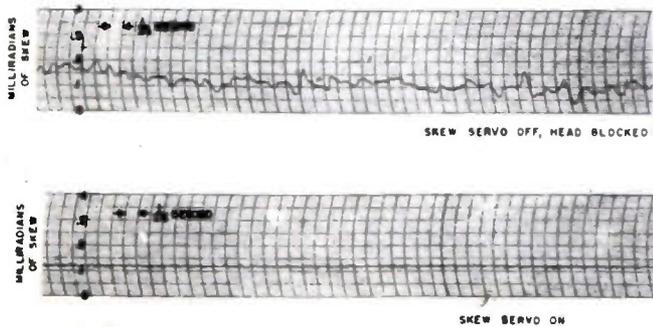


Fig. 11
Oscillograms Showing Reduction in Skew Error
Achieved by Playback Head Servo

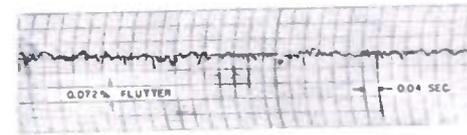


Fig. 12
Oscillograms Showing Flutter Characteristics
of a Tape Transport Developed for
Video Recording

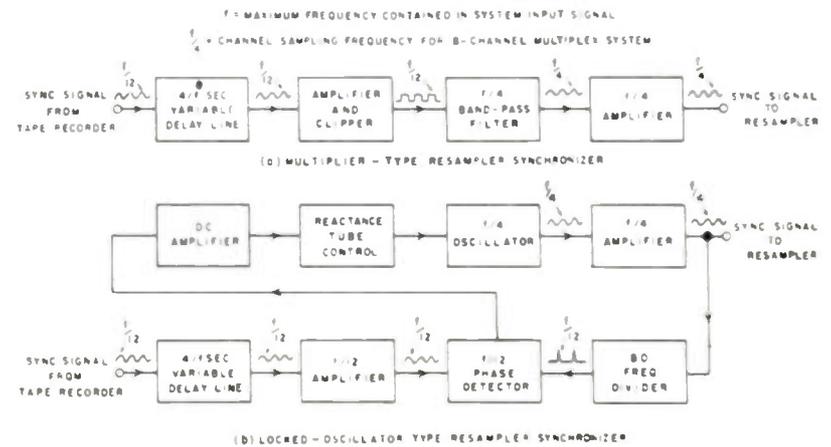


Fig. 13
Block Diagrams of Two Types of Resampling
Synchronizer for Eight Channel System

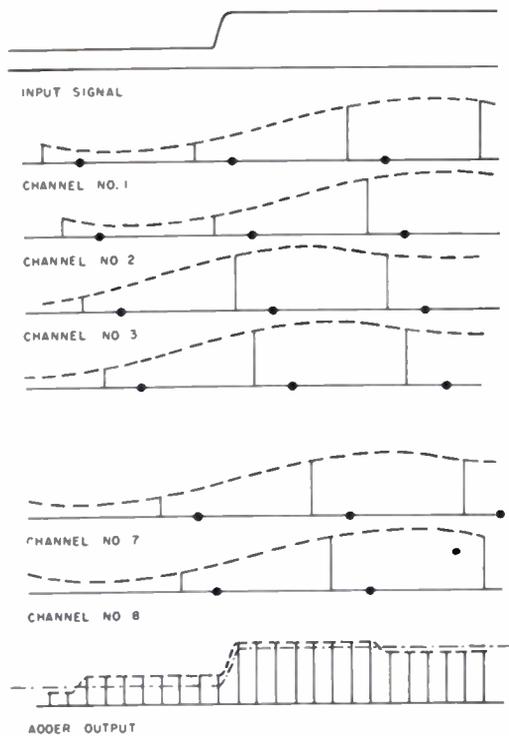


Fig. 14
Effect of Incorrect Resampling Phase on System Output Signal

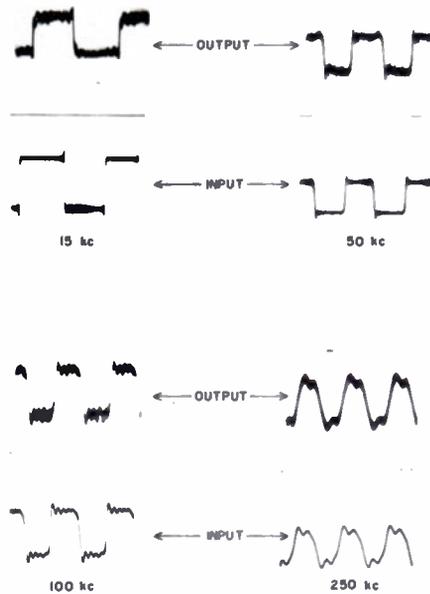


Fig. 15
Square-Wave Response of a 1.1 mc Magnetic Recording System

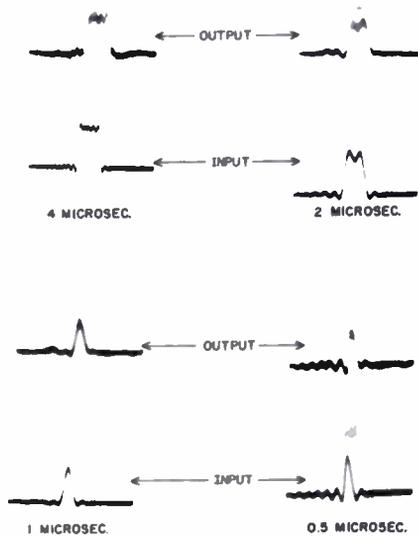


Fig. 16
Pulse Response of a 1.1 mc Magnetic Recording System

CHANNEL RESPONSE REQUIREMENTS OF MULTIPLEX SYSTEMS
FOR RECORDING VIDEO SIGNALS ON MAGNETIC TAPE

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Summary

Time-division multiplexing has been employed as a means of recording a wide-band video signal on multiple-track magnetic tape. The desired frequency response in the individual tape channels is discussed, and a simple method for relating the sampling errors to the deviations in the frequency response is presented. The effects of sampling errors on the output video signals are shown.

Introduction

A system capable of recording video signals on magnetic tape and reproducing them with reasonable fidelity has many potential applications. The recording of television programs, radar data, telemetering signals, and motion pictures are examples. Most of the applications require a system with a bandwidth of about four or five megacycles. It is difficult to achieve this much bandwidth on a single track of magnetic tape; therefore, multiplexing systems have been developed in which the broad-band video signal is broken down into several narrow-band signals that can be recorded on tracks of magnetic tape operating in parallel. The purpose of this paper is to discuss the requirements placed on the frequency response of the narrow-band channels in a time-division multiplex system, and what effects deviations from the ideal response have on the overall system operation.

A block diagram of an eight channel time division multiplexing system is shown in Fig. 1. The theory of operation of the system is based on the sampling theorem which may be stated as follows: If a function $f(t)$ contains no frequencies higher than f_0 cycles per second, it is completely determined by giving its ordinates at a series of points spaced $1/2f_0$ seconds apart.¹

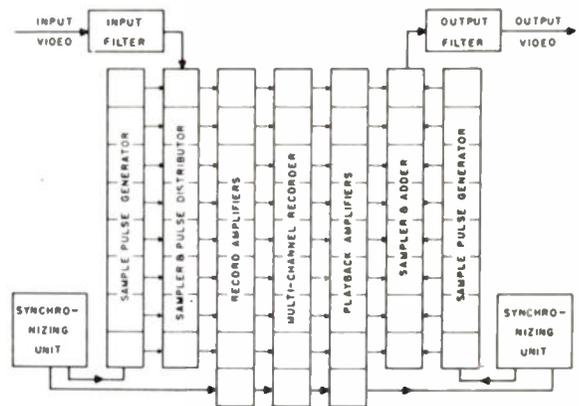


Fig. 1
Block Diagram of 8-Channel Time-Division
Multiplex System

In this system the input signal is passed through a low-pass filter to insure that no frequencies greater than f_0 enter the system. Next, the signal is sampled at a rate of $2 f_0$ times per second, and the resulting samples are distributed sequentially to the narrow-band tape channels. The first sample goes to channel #1, the second sample to channel #2, . . . the eighth sample to channel #8, the ninth sample to channel #1, and so on. During playback, the outputs from the individual channels are sampled and the results combined to form a set of sample pulses identical to that taken from the input video signal. These pulses are passed through a low-pass filter having a cut-off frequency of f_0 , and the output is the reconstructed video signal.

Ideal Channel Response

Each channel receives sample pulses at a rate f_s which is one-eighth the video sampling frequency, $2 f_0$. It follows directly from the sampling theorem that if one desires to put independent samples into a channel at a rate of f_s and recover them with no pulse-to-pulse cross-talk, the channel must have a bandwidth of at least $f_s/2$. However, simply having the required bandwidth does not insure that the desired output can be easily obtained. A channel which has a flat amplitude characteristic and linear phase from zero to $f_s/2$, and transmits no frequencies greater than $f_s/2$ is considered ideal since the input samples can be fed into it at the maximum rate and regained with no cross-talk simply by sampling the output signal at the proper times. This channel has an impulse response of the form:

$$\frac{\sin \pi f_s t}{\pi f_s t}$$

It will give an output waveform of this type when excited with a sampling pulse if it is much

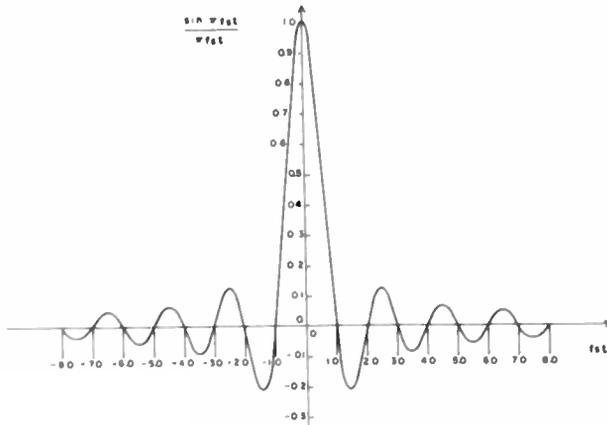


Fig. 2
Ideal Channel Impulse Response

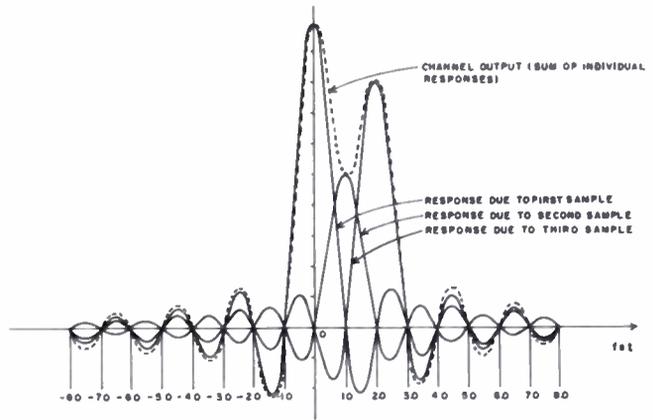


Fig. 3
Response of Ideal Channel to Several Sample Pulses

shorter than $1/f_s$. This ideal response is shown in Figure 2, and the response due to several sample pulses spaced by intervals of $1/f_s$ is shown in Figure 3. Note that if the channel output is sampled at times corresponding to $f_s t = 0, 1, 2, \text{etc.}$, each resulting sample is uniquely related to the amplitude of the corresponding input sample. This condition comes about because the channel impulse response passes through zero at all but one of the sampling points. If, due to an error in the synchronizing circuits, a channel output is sampled at incorrect times cross-talk will be experienced. Since the ringing due to an individual sample pulse persists over a considerable number of samples, the effects of the crosstalk will appear as a whole series of errors in the output. Because of this fact, it may appear desirable to damp the channel impulse response so that it is essentially zero outside of the two sampling intervals in which it is rising to its peak and decreasing again to zero. It is possible to construct channels having this type of impulse response, but the upper frequency limit of the channel must be considerably greater than that required for the even-ringing type of response shown in Figure 2.

Analysis of Practical Channel Response

In a practical system a channel impulse response will not be perfect, but will exhibit some "zero-crossing" errors; i.e., it will not be exactly zero at the sampling points. These zero-crossing errors can be easily measured in a system and their effect on the system operation can be readily determined. In order to

design and align the compensation networks which produce the proper phase and amplitude characteristics in the channels, it is necessary to find a convenient method of relating the zero-crossing errors to the frequency response of the channels. Application of the techniques developed by Wheeler for interpreting phase and amplitude distortion in terms of paired echoes serves this purpose.² Details of this analysis are given in the Appendix

The results that can be obtained by this method are as follows: Consider the frequency response curves shown in Figure 4. (a) is the ideal response; (b) the response achieved in a practical system; and (c) the difference between the ideal and the practical response. $\alpha(\omega)$ and $\beta(\omega)$ are defined as the deviations of the practical amplitude and phase response respectively from the ideal. Figure 5 shows the impulse response of the practical channel. The zero-crossing error, ϵ_n , is defined as the ratio of the impulse response at the n^{th} sampling point ($f_s t = n$) to its maximum value ($f_s t = 0$). If the deviations of the practical response from the ideal are small, then a good approximation for $\beta(\omega)$ is given by:

$$\phi(\omega) = \sum_{-\infty}^{\infty} \phi_n \sin \frac{n\omega}{f_s} \quad \text{for } |\omega| \leq \pi f_s \quad (1)$$

where:

$$\phi_n = \epsilon_n - \epsilon_{-n} \quad (2)$$

An expression for $\alpha(\omega)$ cannot be derived from the zero crossing errors; however, this is not too important since the amplitude response can be easily measured by other methods. It is important to determine what zero-crossing errors a given $\alpha(\omega)$ and $\beta(\omega)$ will cause. These can also be obtained from this analysis, and the expressions are as follows:

$$\epsilon_n = \alpha_{2n} + \frac{\phi_n}{2} \quad \text{for } n > 0 \quad (3)$$

$$\epsilon_n = \alpha_{-2n} - \frac{\phi_n}{2} \quad \text{for } n < 0 \quad (4)$$

where ϕ_n is the Fourier Series coefficient of $\beta(\omega)$ in equation (1) and α_n is the Fourier Series coefficient of $\alpha(\omega)$ as expanded in equation (5),

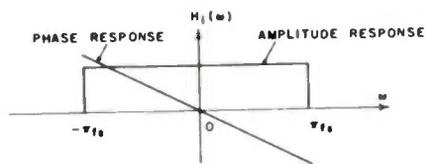
$$\alpha(\omega) = \sum_{-\infty}^{\infty} \alpha_n \cos \frac{n\omega}{2f_s} \quad \text{for } |\omega| \leq 2\pi f_s \quad (5)$$

Effects of Sampling Errors on Output Signal

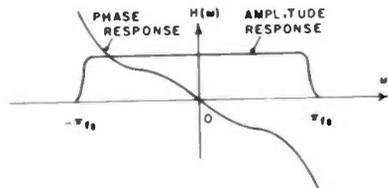
The effects of zero-crossing errors on the output signal can best be understood by considering the results due to a single zero-crossing error in each channel. Once this case is clearly established the generalization to more zero-crossing errors is obvious. The following discussion is based on the assumption that the only zero-crossing error is ϵ_1 , and that all channels are identical. As illustrated in Figure 6(a), the output samples from a channel at time $f_s t = n$ is equal to the desired output, S_n , plus an error $\epsilon_n S_{n-1}$. Figure 6(b) shows the output samples from a channel due to a pulse input signal, and Figure 6(c) shows the composite of the samples from all the channels. The effect of the zero-crossing error, ϵ_1 , is to produce a ghost in the output which is delayed $1/f_s$ seconds with respect to the desired output and is ϵ_1 times the desired signal in amplitude. The effect of any zero-crossing error, ϵ_n , is to produce a ghost delayed by n/f_s seconds and having an amplitude equal to ϵ_n times the main signal. For negative values of n , the resulting ghosts are leading rather than lagging the main signal.

For some applications, such as the recording of digital information, small ghosts do not degrade the output signals appreciably. On the other hand, a five per cent ghost in a television picture is objectionable.

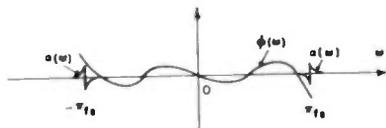
The perceptibility of ghosts in television signals can be reduced by appropriately alternating the polarity of the sample pulses as they are fed into the channels. Up to this point it has been assumed that all sample pulses were the same polarity. Suppose that a circuit is added between the sampler and the record amplifier in each channel which reverses the polarity of every other pair of pulses, and that a decoding circuit is placed in the playback sampler to restore the sample signal to its original form. The samples that are fed into one channel due to a step-function input are shown in Figure 7(a). The error signals, as they appear in the channel output, are shown in Figure 7(b), and in Figure 7(c) as they appear in the output signal (after going through the decoding circuit). The important point to note is that every other error signal is reversed in polarity so that the result of the zero-crossing error is a ghost with its polarity alternating at the sampling frequency. If the sampling frequency



(a) - FREQUENCY RESPONSE OF IDEAL CHANNEL

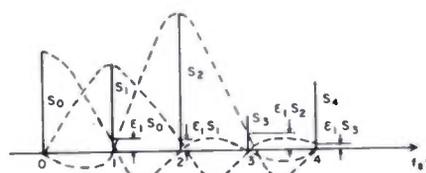


(b) - TYPICAL FREQUENCY RESPONSE OF PRACTICAL CHANNEL

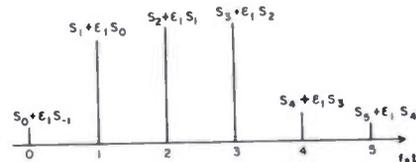


(c) - DEVIATIONS OF PRACTICAL RESPONSE FROM IDEAL CHANNEL

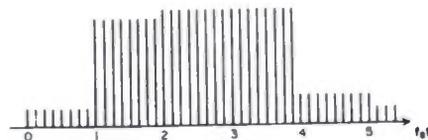
Fig. 4
Channel Frequency Responses



(a) OUTPUT SAMPLES FOR CHANNEL HAVING SINGLE ZERO-CROSSING ERROR, ϵ_1



(b) OUTPUT SAMPLES FROM CHANNEL DUE TO INPUT PULSE



(c) COMPOSITE SAMPLE OUTPUT (SUM OF ALL CHANNEL SAMPLES) FOR INPUT PULSE

Fig. 6

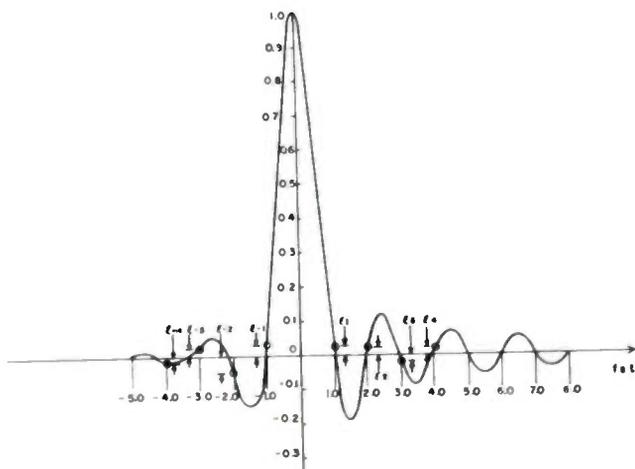
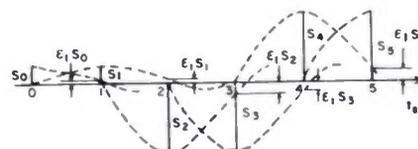
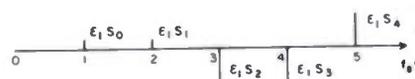


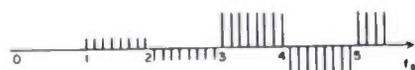
Fig. 5
Practical Channel Impulse Response



(a) ALTERNATING SAMPLES IN CHANNEL DUE TO STEP-FUNCTION INPUT



(b) ERROR SIGNALS AS THEY OCCUR IN A CHANNEL



(c) ERROR SIGNALS AS THEY APPEAR IN THE OUTPUT AFTER THE DECODER

Fig. 7

is an odd multiple of one-half the frame rate, then it can be shown that at any point in the picture a ghost will be positive in one frame and negative on the next. This technique will reduce the effects of ghosts considerably, especially if they are small. Cancellation is never perfect due to non-linearity in the picture tube and failure of the eye to integrate the light output perfectly from one frame to the next.

In general, it is not desirable to use this technique of alternating pulse polarities in systems used in applications other than television. In pulse recording for example, the effects of small ghosts can usually be easily recognized and removed. If the polarity of the ghosts is alternating the problem is much more difficult.

Conclusions

A system utilizing most of the techniques described in this paper has been constructed and evaluated. The results proved satisfactory for some applications and demonstrated the feasibility of the approach; however, further development is required to bring the quality of the output signal up to broadcast television standards.

Acknowledgment

A large portion of the work presented in this paper was developed on a project sponsored by the United States Air Force.

References

- (1) Shannon, C. E. - "Communication in the Presence of Noise" Proc IRE, Vol 33, Pp. 10-21 Jan 1949
- (2) Wheeler, H. A. - "The Interpretation of Amplitude and Phase Distortion in Terms of Paired Echoes" Proc IRE, Vol. 27, Pp 359-385; June 1939

Appendix

Ideally, the channels should have the frequency response shown in Figure 4(a) which may be expressed in equation form as follows:

$$H_I(\omega) = A(\omega) e^{-j\omega \tilde{t}} \quad (6)$$

$$\text{where: } A(\omega) = 1 \text{ for } |\omega| \leq \pi f_s$$

$$A(\omega) = 0 \text{ for } |\omega| > \pi f_s$$

The impulse response of such a filter is given by the Fourier Transform of $H_I(\omega)$,

$$h_I(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H_I(\omega) e^{j\omega t} d\omega \quad (7)$$

Substituting from (6) into (7) and integrating, one obtains:

$$h_I(t) = f_s \frac{\sin \pi f_s (t - \tilde{t})}{\pi f_s (t - \tilde{t})} \quad (8)$$

This is the ideal impulse response. Now the problem is to find the impulse response of a practical channel having the following frequency response.

$$H(\omega) = [A(\omega) + \alpha(\omega)] e^{-j[\omega \tilde{t} + \phi(\omega)]} \quad (9)$$

If $H(\omega)$ approaches the ideal response quite closely, then it may be assumed that $\alpha(\omega) \ll 1$ and $\phi(\omega) \ll 1$. By expanding $e^{-j\phi(\omega)}$ in a Taylor series and neglecting all terms containing powers of $\alpha(\omega)$ and $\phi(\omega)$ greater than unity and products of $\alpha(\omega)$ and $\phi(\omega)$, equation (9) becomes:

$$H(\omega) \cong [A(\omega) + \alpha(\omega) - j A(\omega) \phi(\omega)] e^{j\omega \tilde{t}} \quad (10)$$

Thus the impulse response of the channel may be written as:

$$h(t) = h_I(t) + h_\alpha(t - \tilde{t}) + h_\phi(t - \tilde{t}) \quad (11)$$

where $h_I(t)$ is given by equation (8) and,

$$h_\alpha(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} \alpha(\omega) e^{j\omega t} d\omega \quad (12)$$

$$h_\phi(t) = \frac{-j}{2\pi} \int_{-\infty}^{\infty} \phi(\omega) A(\omega) e^{j\omega t} d\omega \quad (13)$$

It is convenient at this point to expand $\alpha(\omega)$ and $\phi(\omega)$ in Fourier Series and substitute into equations (12) and (13). An equation for $\phi(\omega)$ is required only over the frequency range of $-f_s$ to $+f_s$, therefore it may be expanded in a Fourier Series over this interval. An equation for $\alpha(\omega)$ is required over the whole frequency range. However, if the practical response deviates only a small amount from the ideal, it is reasonable

to assume that $\alpha(\omega) = 0$ for $|\omega| > 2\pi f_s$. Therefore, $\alpha(\omega)$ can be expanded in a Fourier Series over the interval $-2\pi f_s$ to $+2\pi f_s$. The Fourier expansions are as follows:

$$\alpha(\omega) = \sum_{-\infty}^{\infty} \alpha_n \cos \frac{n\omega}{2f_s} \quad \text{for } |\omega| \leq 2\pi f_s \quad (14)$$

$$\alpha(\omega) = 0 \quad \text{for } |\omega| > 2\pi f_s$$

$$\phi(\omega) = \sum_{-\infty}^{\infty} \phi_n \sin \frac{n\omega}{f_s} \quad \text{for } |\omega| \leq \pi f_s \quad (15)$$

Substituting for $\alpha(\omega)$ and $\phi(\omega)$ from (14) and (15) into (12) and (13) yields:

$$h_a(t) = f_s \sum_{n=-\infty}^{\infty} \alpha_n \left\{ \frac{\sin(2\pi f_s t + n\pi)}{2\pi f_s t + n\pi} + \frac{\sin(2\pi f_s t - n\pi)}{2\pi f_s t - n\pi} \right\} \quad (16)$$

$$h_\phi(t) = f_s \sum_{n=-\infty}^{\infty} \phi_n \left\{ \frac{\sin(\pi f_s t - n\pi)}{\pi f_s t - n\pi} - \frac{\sin(\pi f_s t + n\pi)}{\pi f_s t + n\pi} \right\} \quad (17)$$

Substituting into equation (11) from (8), (16) and (17) and evaluating the result at the sampling points ($f_s(t-\tau) = n$), one obtains the following:

$$\begin{aligned} h\left(\frac{n}{f_s} + \tau\right) &= f_s \quad \text{for } n=0 \\ &= f_s \left(\alpha_{2n} + \frac{\phi_n}{2} \right) \quad \text{for } n > 0 \\ &= f_s \left(\alpha_{-2n} + \frac{\phi_n}{2} \right) \quad \text{for } n < 0 \end{aligned} \quad (18)$$

Thus from the definition of the zero-crossing errors, ϵ_n ,

$$\epsilon_n = \alpha_{2n} + \frac{\phi_n}{2} \quad \text{for } n > 0 \quad (19)$$

$$\epsilon_n = \alpha_{-2n} - \frac{\phi_n}{2} \quad \text{for } n < 0$$

From (19) it follows that,

$$\phi_n = \epsilon_n - \epsilon_{-n} \quad (20)$$

FERRITE HEADS FOR RECORDING IN THE MEGACYCLE RANGE

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Summary

The low losses and hardness of ferrites have made them the logical subject of investigation for use in magnetic recording heads, particularly when higher than audio frequencies are to be handled. It was thought that the hardness would produce good wearing characteristics, and that the main disadvantage would be poor resolution due to the granularity of the ferrites.

A program for the development of high density ferrites was initiated. This program has produced a ferrite from which heads with a very short and well defined gap can be produced. These ferrites have a permeability of 1000 or better up to 4 mc and losses sufficiently low to allow operation as playback heads over this frequency range.

Using the improved ferrites, heads have been fabricated with discernable response to .125 mil wavelength and with marginally useful response to .16 mil wavelength. Saturated signal to noise figures greater than 40 db are possible. The maximum saturated signal at the head terminals was around 0.3 millivolt for a head with an inductance of 1 millihenry.

When using these short gap heads as record heads it was not possible to saturate the tape. The ferrite head saturates about 12 db. below tape saturation. This difficulty can be circumvented by using a longer record gap.

The greatest deficiency of the ferrite heads is a short life. Although the gross wearing qualities are better than metallic heads, there is a local erosion at the gap edges; this deterioration occurs primarily on the trailing edge of the gap. Noticeable degradation in the resolution performance occurs in less than 10 hours wear at 100 inches per second in some cases. The use of a glaze material in the gap improves the wear characteristics and improvements in this

fabrication technique along with further improvements in ferrites may solve the ferrite head wear problems.

Text

Because of their relative hardness and low losses, ferrites were early considered a potential core material for magnetic recording heads. As early as 1948 some ferrite heads with an effective gap of around .75 mil were built and tested. Since that time ferrite heads have appeared commercially, mostly for pulse applications where the head was spaced from the medium. In the field of contact heads, ferrites have not fared quite so well. Early thoughts seemed to indicate the following disadvantages:

1. difficulty in fabrication due to hardness,
2. brittleness leading to easy chipping,
3. poor resolution due to granularity.

With the possible exception of the chipping, the above disadvantages have not proven serious. Ferrites can be molded and then ground and lapped, and this process could well prove to be more economical for production than the handling of thin metallic laminations.

The work described in this paper was done as part of a wide band magnetic recording development. It was desired to build heads which could be operated at bias or signal frequencies in the low megacycle range with as high a resolution as possible. Ferrites seemed to satisfy the high frequency requirements.

It was thought at that time that the loss of resolution due to rough gap edges caused by granularity of the ferrite would be the most serious problem, therefore a material development program was initiated to produce a more homogeneous and dense ferrite with satisfactory magnetic and physical properties for use in heads. In terms of the original aims, this program was

quite successful. Ferrites were produced with satisfactory permeabilities and Q to be used as playback heads up to five megacycles; Figure 1 shows a graph of permeability and Q as a function of frequency. That these ferrites had the necessary physical properties to make sharp recording head gaps is shown in Figure 9. It will be noted that the ferrite is free from large voids and blow holes and that the gap edges are quite straight and uniform. Additional evidence of the sharpness of the gap edges is shown in Figure 2. The sharpness of the nulls is a characteristic of relatively sharp and parallel edges.

The head just described (Fig. 2) is not a high resolution head; it could be used out to 0.5 mil wavelength. Photomicrographs of two higher resolution heads are shown in Figure 3. It will be noted that the gap edges appear straight and parallel and free from large irregularities. The sharpness of a gap edge is significant only when related to the recorded wavelength; from this point of view the edges are not sharp and straight but have irregularities which are comparable to the gap length. The minimum gap length for head #15 (Fig. 3) is around .06 mil and for head #18 is around .03 mil therefore the irregularities, although relatively large, are quite small on an absolute scale. The wavelength response of these heads is shown in Figure 4; the problem of gap alignment at very short wavelengths was dodged by recording and playing back on the same head. There were some differences in output between the heads but data on such factors as front and back gap reluctance and the effect of potting strains was not sufficient to attach specific significance to these output variations. It is significant that the curves do not show the sharp null of head #10 this is evidence that the gap edges are less sharp relative to the recorded wavelengths at which the null should occur. In Figure 5 are shown frequency response curves for head #15, taken at 100 and 160 inches per second tape speed. It will be noted that the improvement in frequency response due to the increase in speed is not as great as one would expect. There was evidence that an air film between the head and tape was produced at 100 inches/second. A spacing loss of .02 mils will account for the decreased resolution. These ferrite heads have

a resolution which is at least as good as any metallic heads which have been made available for comparison at this time, a discernable output was noted at .125 mil wavelength. In terms of the equalization required, these heads would be useful to at least .2 mil recorded wavelength. The relatively low output would put a severe requirement on the associated amplifiers.

It has been demonstrated that ferrite heads with good short wavelength resolution can be fabricated, however, they are not, at this stage of development, a satisfactory general purpose head at these resolutions. As a record head they have the following shortcomings: relatively low saturation flux density, low Curie temperature, and erosion of the gap edges. Playback heads suffer from erosion at the gap edges.

The Curie temperature for these particular ferrites was fairly low, around 65°C. The use of these heads as record heads in an ambient temperature of 25°C is marginal; the rise due to combined bias and record current may cause the total temperature to exceed the Curie temperature. They would obviously not satisfy military specifications. This limitation does not appear to be fundamental. A further material development should raise the Curie temperature some, although perhaps at the sacrifice of some of the other properties.

The low saturation flux density is not a serious handicap in conventional playback heads, or in wide gap record heads. In record heads which have gaps as small as those described above, saturation becomes a serious problem. Since the recording process in a gap type head depends upon leakage flux, and since the relative amount of leakage flux with a very fine gap is very small, it follows that the flux density in the core, and especially the gap edges must be high. It has been found that with these ferrites and gap lengths saturation does occur. In Figure 6 are shown Input-Output curves for a ferrite head and a Brush BK1090 head for comparison; the same head was used for playback in each case. It will be noted that the first break in slope occur at 8 db. output for the Brush head but at -1 db for the ferrite head. It is not obvious from the data that the

Brush head is saturating but certainly the ferrite head is saturating well below tape saturation. This difficulty may be reduced by using materials with higher saturation flux density or by increasing the record gap length. Ferrites as a class tend to have a low saturation flux density so that it appears wider gaps are necessary in ferrite record heads to avoid saturation. It should be emphasized that the use of a wider gap does not mean that a decrease in the sharpness of the gap edge is allowable. The use of a wide gap record head would require special attention to the gap edges in order to retain resolution and to reduce record gap anomalies in the frequency response curve.

Originally thought of as wear resistant heads because of their hardness, ferrite heads, at present, have wearing qualities which are poor. Wear shows up as an erosion of the tape contact surface. In many instances this erosion may appear all over the surface, but in most cases it is concentrated at the gap edges, the worst location as far as head performance is concerned. Some wear tests were performed on these ferrite heads by running them at a tape speed of 100 inches per second and then measuring the wavelength response at low tape speeds. A mu-metal head was run at the same time for control purposes.

A ferrite head (Head #5) with a Hysol gap spacer was wear tested for a total of 93 hours, corresponding to 2,400,000 feet of tape. The tape used was 3-M type 111 acetate backed tape, and the normal force between head and tape was around 75 grams. No pressure pads were used. In Figure 7 are shown photomicrographs of the gap edges after 35 and 83 hours of wear. The original gap was similar to that of head #15 but somewhat more irregular. The gap after 35 hours of wear shows a definite wearing pattern. There are long scratches which do not appear to be serious except as a possible site for further erosion. After the scratches, erosion appears; actual erosion can be considered in two categories, surface erosion and gap edge erosion. Surface erosion may occur anywhere on a surface where the conditions are favorable, and in itself is not detrimental to head performance. The type of erosion which is very serious in ferrite heads occurs at the gap edges. In the 35 hour photo-

micrograph, the directional qualities of the gap edge erosion are quite marked. The trailing edge which is directed against the direction of tape travel, is badly eroded while the leading edge which the tape slides off of shows very little evidence of erosion. Intuitively this situation seems reasonable. After 93 hours of wear the trailing edge is much more badly eroded, and some erosion is starting to occur at the leading edge; the surface erosion has also increased appreciably. In Figure 8 are shown wavelength response data showing the deterioration in performance as a playback head as the result of wear. It can be seen that the bulk of the deterioration in performance has occurred in less than 23 hours of wear. The deterioration then progresses slowly and there is evidence that a usefully long life could be realized at 0.5 mil wavelength if the direction of tape travel was not reversed. If the tape direction is reversed the uneroded gap edge will erode rapidly so that no portion of the gap would be sharp or well defined. Evidence based on some experience with ferrite heads designed for 1.0 mil useful resolution indicates that, when both gap edges deteriorate, the shortest useful wavelength will be around 1.0 mil.

Wear data obtained on both sintered ferrites and single crystals without fabricated gaps indicate that the intrinsic wearing properties are appreciably better than those experienced with fabricated heads. For that reason methods of making the gap area physically more like an ungapped ferrite have been devised. One thought is that when the gap is very short the tape surface cannot get down into the gap region and erode the trailing gap edge; if the joint were perfect this certainly appears reasonable. It appears unlikely, at this time, that a head with usefully high output will have a short enough gap to successfully resist wear. For this reason it appears that something must be done in the gap or within the material in order to decrease this gap edge erosion.

A technique which holds some promise, is to fill the gap with a glaze material which is non-magnetic, bonds well to the gap faces, and is hard. Such a glazed gap head (Head #10) was fabricated and subjected to wear tests at 100 inches per second. A photomicrograph of this

head is shown in Figure 9. Most of the gap was clean and straight when new. After 48 hours of wear (1,450,000 feet of tape) the head has a large amount of surface erosion and the gap edges have eroded somewhat. The directional wear qualities are not very obvious on this head. It can be seen that this head has not eroded as much in 48 hours as the previous head had in 35 hours. In Figure 10 are shown wavelength responses taken after 21 and 48 hours of wear. After 21 hours the gap edges were relatively sharp giving a well defined null but there was apparently a loss in resolution. After 48 hours the gap edges were irregular enough to almost completely suppress the second peak, although the resolution at wavelengths longer than 0.5 mil was substantially unchanged. Comparison of the wear on heads #5 and #10 indicated that the glazing technique has apparently increased the resistance to wear. Since there is no reason to consider this glaze optimum, it appears that there is something to be gained by this technique.

In conclusion it can be stated that ferrite heads can be constructed which have resolutions, when new, comparable to metallic heads, and that they compare favorably in performance with metallic heads constructed from thin laminations for high frequency use. They are, however, deficient in wearing qualities; ferrites suffer a reduced performance with wear while metallic heads tend to retain performance although perhaps wearing faster. There are grounds for hopes that the future will produce high resolution ferrite heads which are satisfactory from all view-points.

The contributions of colleagues are gratefully acknowledged; in particular, Aaron P. Greifer and Fredrick G. Keihn who developed the ferrites and fabricated the heads, and Louis A. Budell who participated in the evaluation and testing. This work was accomplished under the sponsorship of the United States Air Force.

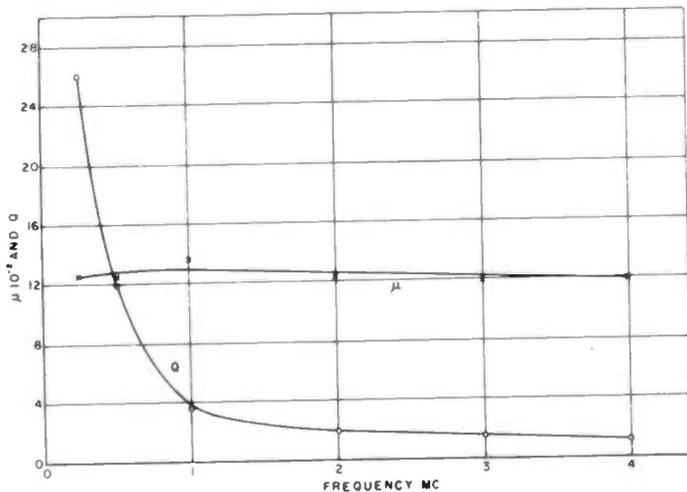
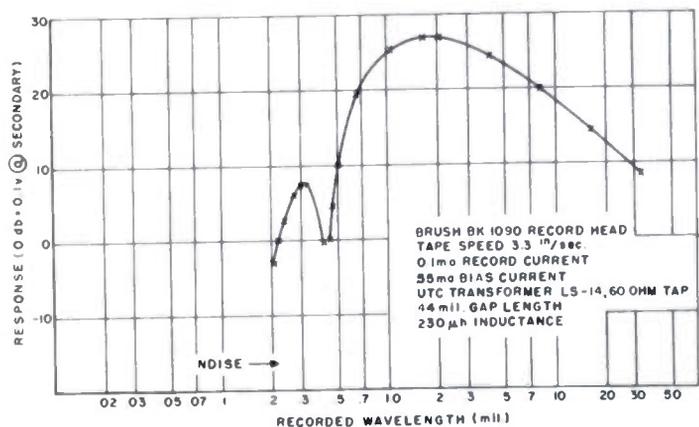
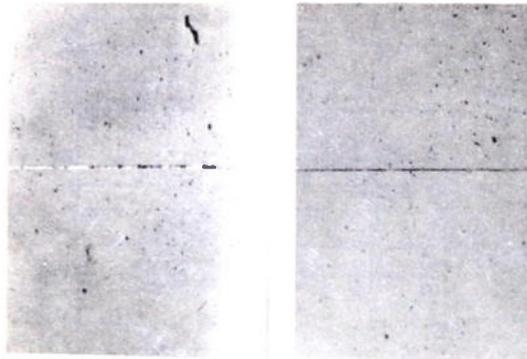


Fig. 1
Permeability & Q vs. Frequency,
Ferrite Sample 13.90

Fig. 2
Wavelength Response, Head No. 10,
Ferrite with Glazed Gap





HEAD NO. 15, 330X
SILVER SPACER
GAP LENGTH .07 MIL

HEAD NO. 18, 500X
HYSOL SPACER
GAP LENGTH .02 MIL

Fig. 3
High Resolution Ferrite Head Gaps

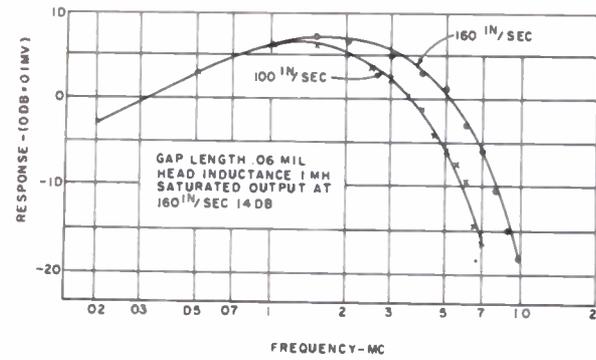


Fig. 5
Frequency Response Head No. 15

22

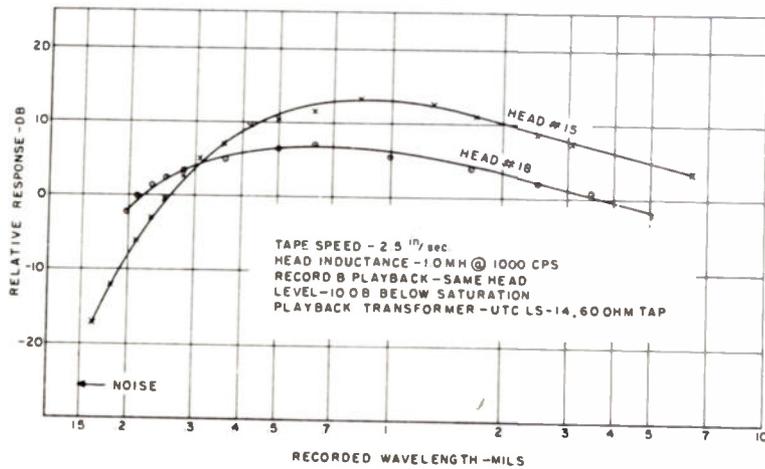


Fig. 4
Wavelength Response Ferrite Heads

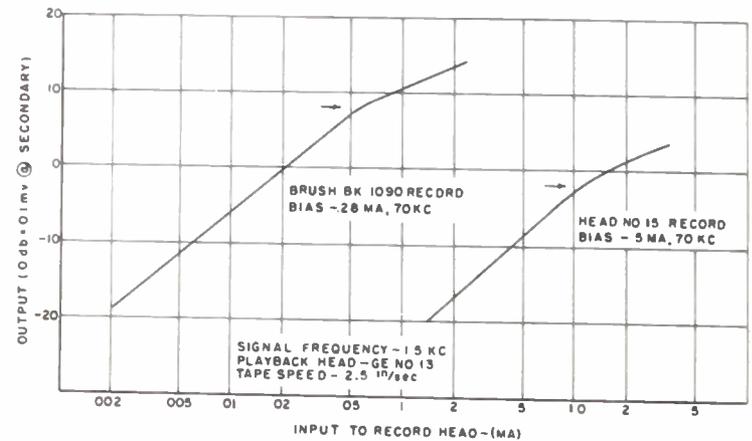
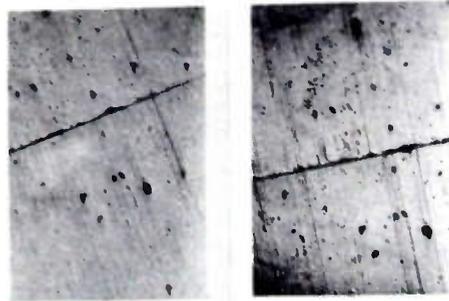


Fig. 6
INPUT-OUTPUT Curves

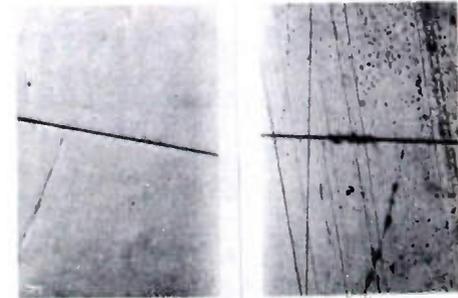


AFTER 35 HOURS
WEAR AT 100 in/sec.

AFTER 83 HOURS
WEAR AT 100 in/sec.

MYSOL GAP SPACER
MINIMUM GAP LENGTH .04 MIL
AVERAGE GAP LENGTH .1 MIL
MAGNIFICATION 135 X

Fig. 7
Ferrite Head No. 5



NEW GAP LENGTH 0.2 MIL

AFTER 48 HOURS
WEAR AT 100 in/sec

MAGNIFICATION 135 X

Fig. 9
Ferrite Head No. 10 With Glazed Gap Spacer

23

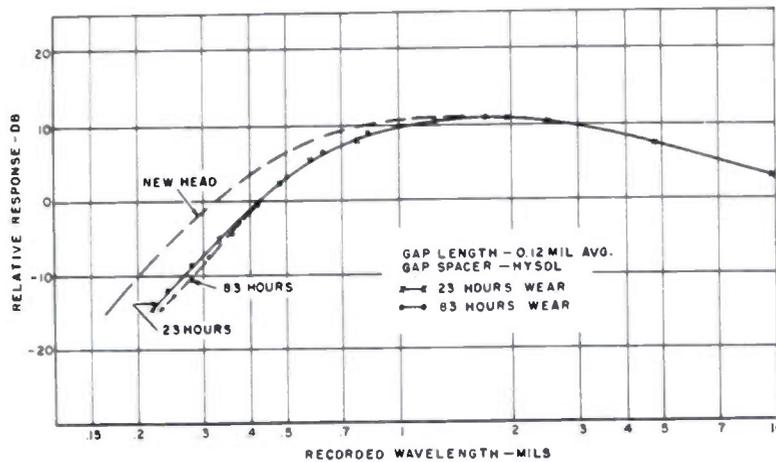


Fig. 8
Effect of Wear on Resolution, Ferrite Head No. 5,
Wear at 100 in/sec

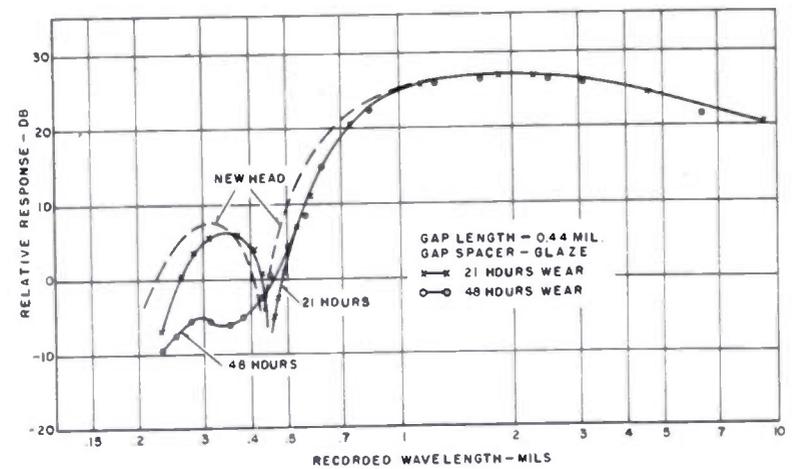


Fig. 10
Effect of Wear Resolution, Ferrite Head No. 10,
Wear at 100 in/sec

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Introduction

The following measuring technique has been developed to compare the attenuations of relatively short samples of r-f transmission lines. These lines are of the type used in FM and UHF-TV broadcast installations, and have low attenuation factors throughout the frequency range up to at least 1,000 mc.

The measuring procedure devised requires no unconventional test equipment or destructive line alterations and has given good results on line samples only 6 to 20 feet long. The general technique employed is to close both ends of the line sample, couple into one end and then measure with an impedance meter and signal source the input impedance of the line as a resonant coaxial (TEM) cavity.

This resonance data for the different line samples with minor restrictions gives a relative comparison of their losses due to series resistance and shunt conductance. This data can also be used to calculate directly the line attenuation in decibels per 100 feet, and an example of this calculation is described herein.

The main restriction to the technique is that the simple equations relating the cavity resonator losses to the matched line attenuation apply only for uniformly distributed loss factors. The series resistance loss is, in general, so distributed, but if the shunt loss is due to spaced insulator beads, it will be a function of the voltage standing-wave pattern for the particular resonant mode. The unloaded, resonant "Q₀" is measured at the frequencies where the line sample is $(1/2n)/4$ wavelengths long. This is not a severe restriction, since even for a typical 20 foot sample of air dielectric line, the resonances occur about 24.6 mc apart above the first resonance at 12.3 MC.

In closing this section it should be explained that a simple technique for closely evaluating the Z₀ of "smooth" lines already exists. This consists of first measuring "C_a", the quasi-static capacity of an air dielectric section of the line, in puf/foot and the added capacity of a typical insulator bead of dielectric ε. The equation $Z_0 = \frac{1015}{C_a}$ ohms is then used with the bead capacity added as a distributed effect by the term $l(\sqrt{\epsilon}-1)$.

Attenuation Measurement Process

A typical equipment set-up is shown in Figure 1. The line sample is terminated at the far end by a shorting plate. The near end is fitted with a coax-fed coupling probe separated less than 0.05 inches. The signal generator must have good frequency stability after warm-up, it must have a good vernier interpolating dial, and it must be free of incidental FM when amplitude modulated or else operated without modulation. The impedance or admittance meter must have "good laboratory accuracy" and the receiver or vswr indicator requirements are conventional with the exception that it must indicate cw level if the signal generator is not modulated.

After the equipment has fully warmed-up, the input Z' or Y' referred to the location of the coupling probe is measured as the carrier frequency is tuned in small steps across one of the resonance modes. This data is plotted on a normalized circle chart, "Smith Chart," to determine the general outline of the "resonance circle." If it has a convenient diameter, (between about 1/5 to 3/5 of the chart diameter), additional points are taken to enable a smooth plot of this "circle." If the measured circle is Z'/Z₀, rotate it 180° on the chart, inverting to Y'/Y₀. If the "circle" diameter is not suitable it can be increased or decreased by increasing or decreasing the capacity of the coupling probe.

The susceptance $1/jB_f$ of the probe ends fringe capacitance "C_f" is then estimated from the Y' data at frequencies well below and above the resonance effect. This term is practically constant across the narrow resonance band and can be numerically subtracted to give $\frac{Y}{Y_0}$, see Figure 2a. The plot of $\frac{Y}{Y_0}$ should be a

perfect circle and is graphically rotated 180° on the circle chart, inverting into Z/Z_0 .

The Z/Z_0 curve is now centered on the real axis at the ∞ end of the chart, and should fit as a member of the constant R circles on the Z chart. Since Z should be a perfect circle, if the plotted data shows experimental scatter, the plot could be refined by graphical smoothing or by plotting R and jX vs frequency on rectangular paper and following "precision" techniques.

The three frequencies f₁, f_r, and f_h, on this resonance circle Z are noted for: -jX = R, jX = 0 and jX = R, respectively.

The unloaded Q_0 of a high Q resonant circuit is

$$Q_0 = \frac{f_r}{f_h - f_l} \quad (1)$$

$$\text{If } B_g = 2\pi f/V_g \text{ radians/length,} \quad (2)$$

$$\text{line attenuation "OC}_g\text{"} = B_g/2Q_0 \text{ nepers/length,} \quad (3)$$

$$= 8.69 f/V_g Q_0 \text{ db/ft if } V_g \text{ in ft/sec.} \quad (4)$$

$$= 2.79 \times 10^{-2} \frac{f_r(\text{mc})}{Q_0} \times \frac{V_c}{V_g} \text{ in db/ft} \quad (5)$$

This procedure for capacitively coupling to the high impedance end of the cable is very convenient for checking the OC_g of flexible cables. The coupling can often be provided by loosening the plug until the center contacts just barely separate.

A Practical Example

The measurements and calculations for a 6 foot sample of RG-9A/U polyethylene insulated line are presented to illustrate the actual technique. The specified characteristics of this line at 400 mc are: $Z_0 = 51$ ohms, $OC_g = 5.2$ db/100' and $V_g = 0.659$ Vc.

A resonance close to 400 mc, assuming $V_g = 0.66$ Vc, would be the 3.75 TEM mode at 402 mc. Figure 2b shows the impedance plot of this resonance at 402.25 MC. The measured points are 250 kc apart. The resonance "circle" is not perfectly circular, due to errors in the impedance measurements, but it closely resembles the circle for $R/Z_0 = 3.3$ and the marks for f_l and f_h are 1.15 mc apart.

$$\text{So } Q_0 = 402.25/1.15 = 350$$

From eq. 5,

$$\begin{aligned} OC_g &= 2.79 \times 10^{-2} \times \frac{402.25}{350} \times \frac{1}{.65} \\ &= 4.95 \times 10^{-2} \text{ db/ft.} \\ &= 4.95 \text{ db/100 ft.} \end{aligned}$$

This value of measured attenuation is within 5% of the nominal value and confirms the technique. This 5% deviation may be due to the following factors: deviation of line sample's characteristics from nominal value, errors in impedance measurement, and errors in frequency interpolation. The bandwidth measurements must be made with considerable precision and for best accuracy it would be desirable to employ some technique more refined than interpolations with the signal generator vernier dial. One such refinement would be to mix a small portion of the signal generator output with the output of a fixed VHF signal generator and measure the frequency of the difference signals for f_l and f_h with an electronic counter-timer or

a LF receiver.

The technique as described requires that one end of the line sample be open circuited. The same technique is applicable when the input end of the line is short circuited. In this case the center conductor bullet is added and a coupling loop with an area of 1 or 2 sq. inches is mounted on the end plate. The center conductor resonates in all the half-wave TEM modes which have a low impedance at both ends and the coupling loop reactance is " X_L ". The resonance circle $Y/Y_0 = Z_0/(Z' - jXL)$ on a locus of constant G and is suitable for determining Q_0 directly.

Theory

The portions of this paper which warrant theoretical discussion are equations 1 and 3. The general theory of coupled resonant cavities and their evaluation by vswr measurements has been rigorously presented,¹ and the vswr technique could be applied to this case but it requires the use of a coefficient for cavity loading. When the resonator Q_0 is so high that the conventional theory for coupled lumped circuits² applies quite well, it leads directly to a simple technique for evaluating Q_0 from input impedance data.

If the coupling coefficient is small, the cavity is lightly loaded and its loaded Q , " Q_L " approaches Q_0 , but when the coupling is increased to make the resonance effect large enough for accurate measurements, the cavity becomes loaded and Q_L decreases. This means that response-bandwidth measurements of Q_L generally require a correction factor to determine Q_0 . If, instead, the coupled impedance admittance at the coupling point is measured as a function of frequency, its variation with frequency is simply related to Q_0 without dependence on the coupling coefficient.

The probe-coupled case can be viewed as a very high- Q parallel resonant circuit which is coupled into by the probe capacitance " C_p ". The probe end fringe capacitance " C_f " presents the shunt susceptance " $Y' = j\omega C_f$ " across the input where Y' is measured. When the term $-j\omega L_f$ is subtracted from Y' , the resonance term Y remains.

At natural resonance the line end shows a pure resistance $4Z_0$. This is too high for direct measurements with an impedance meter having about the same Z_0 , and the probe capacity serves to shift the system's resonant frequency far enough down to reduce the resistive term of the line by a factor on the order of $1/Q$. This frequency is so far away from its natural resonance that the line end now appears as a relatively low resistance, R_s , in series with a high inductive reactance, X_{L_s} , and

1. J. C. Slater, "Microwave Electronics," p. 93-95, Van Nostrand, N. Y.
2. Ramo & Whinnery, "Fields and Waves in Modern Radio," 2nd Ed. p. 464-470, Wiley, N. Y.

when the parallel circuit is converted to its series-equivalent, see Fig. 3. Both R_s and X_s are increased slightly but in the same proportion far off resonance by the line's resonating capacity, So:

$$Z = R_s + j(X_{L_s} - X_{C_p})$$

Therefore the plot of Z on an impedance circle chart is a circle of constant R, and the frequencies f_1 and f_h are those where $\pm jX = R$. These frequencies are used to calculate Q_0 since $Q_0 = f_r / f_h - f_1$.

The loop coupled case can be viewed as a loss-less, low inductance primary loop magnetically coupled by the mutual "M" to the high-Q series resonant secondary, see Fig. 3. The measured input impedance referred to the location of the loop is Z'.

$$Z' = jX_{L_p} + \frac{\omega M^2}{R_s + j(X_{L_s} - X_{C_s})}$$

Since the Q is high, the terms of X_{L_s} , ωM , and R_s are constant through the resonance and jX_{L_p} can be subtracted to give Z:

$$Z = \frac{\omega M^2}{R_s + j(X_{L_s} - X_{C_s})} = 1/Y$$

$$Y = G + jB = KR_s + K(X_{L_s} - X_{C_s})$$

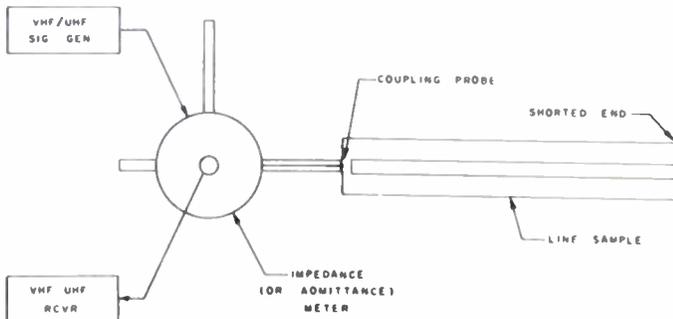
Therefore the plot of Y on an admittance circle chart is a circle of constant G, and the frequencies of f_1 and f_h are those for $\pm jB = G$. The effect of changing M is to change K but not f_1 or f_h .

Equation 3 will be derived from the conventional definition:

$$Q = \frac{\text{Stored Energy}}{\text{Loss per Cycle}} = \frac{U_M}{W_L / 2\pi f}$$

where U_M is the stored magnetic energy in Joules for MKS units.

Assuming a uniform coax guide of:
 Characteristic impedance Z_0 ohms;
 Inductance L_0 henrys per meter;
 Capacitance C_0 farads per meter;
 Resistance loss R_0 ohms / meter, series only



By conventional theory:

$$2\pi f \sqrt{L_0 C_0} = \beta_g \text{ radians per meter;}$$

$$Z_0 = \sqrt{L_0 / C_0} \text{ ohms}$$

$$OC_0 = R_0 / 2Z_0 \text{ in nepers/meter}$$

If line current i_0 flows in length "l"

$$U_M = L_0 l i_0^2 \text{ Joules}$$

$$W_L = R_0 l i_0^2 \text{ Joules/second}$$

$$Q_0 = L_0 2\pi f / R_0 = L_0 2\pi f / 2Z_0 OC_g$$

$$\text{After clearing terms, } OC_g = \frac{R_g}{2Q_0} \text{ nepers/m (3)}$$

By equation 3 it is seen that the length of the line sample does not enter directly into the determination of OC_g . This is logical since in even a short sample the losses in the end plates are negligible. Since the resonant cavity $Q = 2\pi$ stored energy/loss per cycle, it is obvious that a change in line length to add or subtract one or more resonant $\lambda_0/2$ cylindrical sections will not change this ratio for Q. Although the above calculations are for series conductor loss only, a uniformly distributed shunt loss would also be satisfied by (3).

The basic principles of this attenuation-measuring technique apply to any resonant system and consequently to any mode in any guided transmission system. Obviously for system's resonance, equation (3) has to be applied by using "waveguide wavelengths" in the case of waveguides or non-TEM modes in coaxial, twin-wire, strip lines, etc.

Conclusions

The described technique for measuring the attenuation and velocity of propagation of coaxial lines and other transmission systems makes possible convenient and accurate measurements on short sample sections. In addition it does not require any specialized test equipment or destructive modifications to the test samples.

Fig. 1
Equipment Set-up

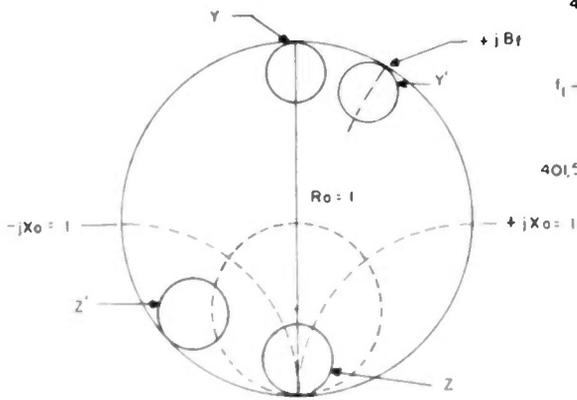


Fig. 2a
Paths of Resonance Circles

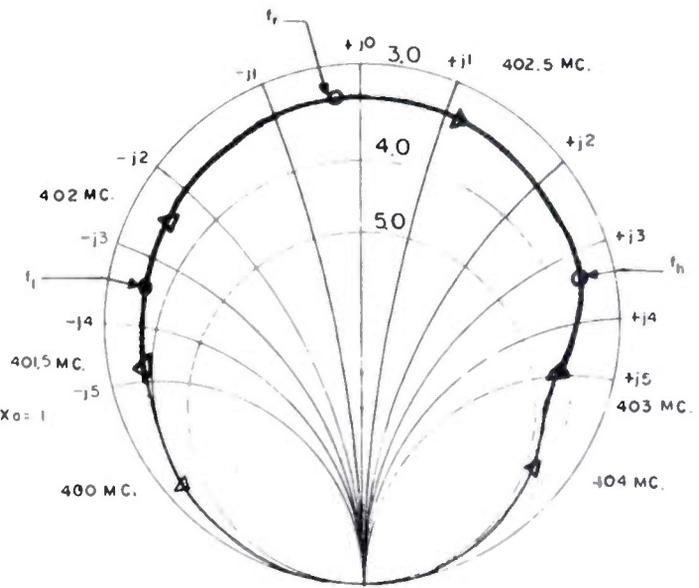
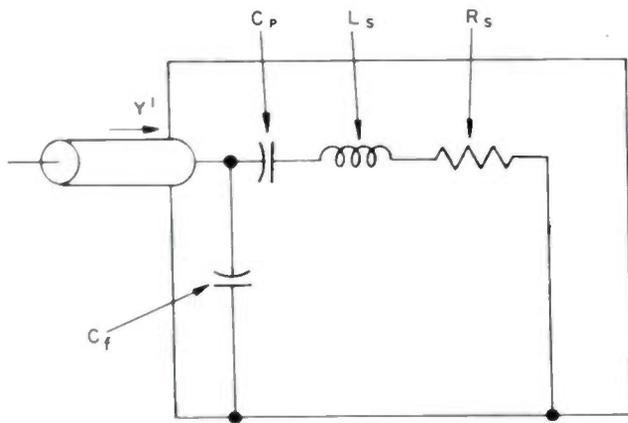
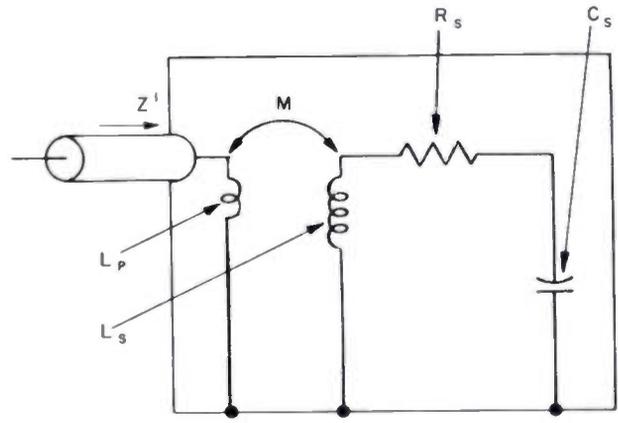


Fig. 2b
Expanded plot Z/Z_0 for RG-9A/U Sample



$$Y' = jB_f + \frac{1}{R_s + j(X_{L_s} - X_{C_p})}$$

PROBE COUPLING



$$Z' = jX_{L_p} + \frac{\omega^2 M^2}{R_s + j(X_{L_s} - X_{C_s})}$$

LOOP COUPLING

Fig. 3
Equivalent, Lumped circuits.

A NEW TELEVISION TRANSMITTING ANTENNA

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Abstract

An omnidirectional traveling wave antenna suitable for high gain television transmitting has been devised. It embodies a novel solution to the problem of overcoming the inherent tendency in wave antennas for the beam direction to change as the frequency varies over the operating range. While the particular construction described lends itself most ideally to applications in Channels 7 to 13, it is conceivable that the basic principle of operation can be adapted to other types of construction for use in the low and uhf channels. The antenna is highly designable.

Brief Theory

Basically, the antenna consists of a long spiralled, linear array of elementary radiators spaced moderately close together and lightly coupled to a uniform, dissipationless transmission line. The coupling is just sufficient for a traveling wave introduced into one end of the structure to be largely, if not entirely, dissipated by radiation upon its arrival at the remote terminal. In order for the antenna to function as intended, it is essential that the radiators have doublet-type patterns in the azimuthal plane. Granted this, it follows that the spiralling of the array advances the effective phase progression along the aperture by the amount of the spiral. The beam direction, δ , with respect to the array normal is therefore given by

$$\sin \delta = \frac{\beta - 2\pi}{\beta_0},$$

where β and β_0 are the internal and free space phase propagation constants respectively in radians per turn length of the spiral. When the spiral rate is precisely one turn per electrical wavelength in the loaded transmission line, the maximum radiation is in the broadside direction, and the highly circular pancake-type beam that results has linear progressive azimuthal phase.

In order to stabilize the beam in the horizontal direction in the neighborhood of a given frequency, it is obviously necessary for β to assume the stationary value, 2π , at that frequency. Similar restrictions apply for other beam directions. This condition can be realized by causing series or shunt coupled radiators to

introduce reactance or susceptance functions respectively which have negative slopes of the proper magnitude at radiator resonance.

A high degree of phase correction over a stated band requires, on the one hand, a fair attenuation rate and a suitable radiator Q , depending upon the bandwidth and the correction tolerance. High directivity, on the other hand, is usually achieved by using long apertures in which the relative excitation level has everywhere an appreciable value. The higher the directivity becomes, the more stringent become the needs for beam stability, and the less becomes the allowable attenuation rate. Since the obtainable phase correction must inevitably decrease with increasing aperture length, one concludes that the directivity which can be achieved using a given distribution is limited by the tolerance placed on the phase function. This is true, but the limit is not a serious one.

Well corrected, end-fed antennas having directivities of the order of 20 relative to a half-wavelength dipole can be built for any channel in the 7 to 13 range. By center feeding apertures longer than this, the emphasis on phase correction is greatly reduced, and antennas having very high directivity become quite feasible.

A very acceptable beam pattern having good directivity and no nulls is produced by an exponential distribution, which is fortunately associated with the simplest of all structures, namely the uniform. The beam shape can be modified to a considerable extent, however, and the beam itself pointed at any desired elevation by specifying the spiral rate and strength of radiator coupling to the line as functions of distance along the aperture. Due to the progressive azimuthal phase character of the resultant fields, the spiral rate may be used as a design parameter to achieve any effective phase distribution whose fine structure is coarse compared to the radiator spacing. The required amplitude distribution (within reasonable limits) is obtained by adjusting the coupling of the radiators to the line.

Of utmost importance is the fact that the controls on the amplitude and phase can be made substantially independent of each other. In a practi-

cal design, one would use four radiating units per turn of the spiral. Successive units, then being orthogonal to each other, would be blind to radiation coupling which could seriously alter their intended relative phase and amplitude. Units two spaces removed from each other would be essentially in phase so that radiation coupling, though present in some degree, would not adversely affect the excitation.

The spacing of the radiators being approximately one-quarter wavelength along the transmission line, and the loading being small compared to the characteristic impedance or admittance of the unloaded line, the input impedance of the antenna when terminated in a matched extension of its own line is practically the same as that of the unloaded line itself. Hence, the usual input impedance bandwidth problems are absent. It is estimated that in practice the net attenuation between the extremities of the antenna would be of the order of 20 to 25 db, which would result in the input reflection coefficient being less than 0.05 under any terminal condition.

With such a high insertion loss through the antenna, it is probable that transmitters not differing greatly in frequency could operate successfully into both ends of the antenna simultaneously without objectionable cross modulation. This suggests an interesting possibility for the diplexerless introduction of the aural and visual transmitters into the one antenna.

Construction

One suggested embodiment which appears to be practical for Channels 7 to 13 inclusive utilizes a coaxial transmission line of sufficiently large dimensions to comprise a free-standing mast. A radiating unit consists of a diametrical pair of short, narrow, probe-coupled, axial slots excited out of phase to achieve the required doublet pattern in the azimuthal plane. The space pattern of such a unit has a symmetrical

shape approximately equal to that of a tangent pair of identical spheres whose line of centers is normal to the array axis. The primary source thus manifests a rather high directivity in the broadside direction, which is just what is desired. This not only tends to minimize radiation coupling between alternate units, but improves the effective continuity of the distribution.

Since transverse-electric slots in cylinders generate only transverse-electric radiation fields, the resultant field of the array is entirely free of cross-polarized radiation at all elevation angles. Because the effective excitation is so nearly continuous, the aperture may be used with a maximum of efficiency toward achieving the required distribution.

Conclusions

The striking physical simplicity, the high power handling capacity, and the relatively inexpensive and durable construction of this antenna are highly commendable features, but probably the most important of all is the fact that it is thoroughly designable to a specified performance by an engineer sitting at his desk. Only the scantest of empirical information is required. A significant contribution to the art appears to have been made in providing a practical physical means for approximating the mathematically prescribed complex aperture distributions required for various given beam shapes.

Author's Note

The work briefly described here is part of a dissertation being submitted to the University of Pennsylvania in fulfillment of a requirement for the Ph.D. It was performed at The Ohio State University Antenna Laboratory under a contract sponsored by the RCA Victor Division of the Radio Corporation of America. A comprehensive paper is in preparation.

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Summary

The basic theory of the coaxial short-line filter is reviewed from a survey of the literature. Specifications for spurious emission filters for television transmission service are arbitrarily assigned. Step by step design of a high power VHF filter is outlined and experimentally verified. Discussion includes 25 KW channels 2-13, 50 KW channels 7-13, and 12.5 KW channels 14-83 low pass filters. Image parameter design with constant K and m-derived (series and shunt) sections is used throughout. Effect of insertion of the filter in the transmission system is studied.

Introduction

The term "Spurious Emission Filter" is, in a sense, a misnomer. Measurements on a large number of television transmitters have shown that only the fundamental harmonics require attenuation if the F.C.C. specifications on spurious radiation¹ are to be satisfied. Consequently, the spurious emission filter, intuitively a band-pass device, is usually in practice a low-pass or a harmonic filter. Since it must be assumed that the magnitude of the harmonic signal appearing at the transmitter output terminals is a function of the final amplifier only, the spurious emission filter has as requisites low fundamental loss and high power handling capacity. This suggests physically large, high-Q elements, which are not usually realized in a lumped circuit. Standard coaxial transmission line elements, however, may be used to construct satisfactory high-power filters for the VHF and UHF television bands at all existing power levels.

Coaxial line filters can be generally classified as: (a) resonant-line filters, (b) short-line filters.

* The development work described here was sponsored by Allen B. Du Mont Labs., Communication Products Division.

Resonant-line filters providing adequate harmonic attenuation and utilizing a minimum of elements are both practical and economical but they are by nature channel selective and do not readily lend themselves to large quantity production. On the other hand, the short-line filter, in essence a broad-band device, has the advantage of being in general non-channel selective and requires no field adjustment, two obvious economies over the resonant-line type. Therefore, it is the short-line filter which will be discussed here.

The purpose of the paper is two-fold; (a) some observations are passed along to the development engineer which it is hoped will help to facilitate future designs; (b) personnel associated with television transmitting equipment, but not necessarily with the development thereof, may derive a better understanding of the transmission line power filter which has only recently appeared on the electronic scene. For information pertaining to the resonant-line type coaxial filter, which takes many interesting forms, the reader is referred to the literature 2,3.

Coaxial Short-Line Filters

It is desirable to base the design of short-line filters on classical filter theory, and this will be realized if short-line sections can be shown to simulate the requisite elements of the classical filter. From the succeeding brief analysis⁴ some necessary relationships are derived.

The input impedance, Z_1 , of a length, l , of transmission line of characteristic impedance, Z_0 , is given by

$$Z_1 = Z_0 \frac{Z_L \cos \beta l + j Z_0 \sin \beta l}{Z_0 \cos \beta l + j Z_L \sin \beta l} \quad (1)$$

where Z_L = terminating impedance

$$Z_0 = [L_0/C_0]^{1/2}$$

$$\beta = \omega(L_0 C_0)^{1/2}$$

Equation (1) may also be written

$$Z_1 = \frac{Z_L + jZ_o \tan \beta l}{1 + j \frac{Z_L}{Z_o} \tan \beta l} \quad (2)$$

If the inequality

$$\left| \frac{Z_L}{Z_o} \tan \beta l \right| \ll 1 \quad \text{holds} \quad (3)$$

and if $\tan \beta l \approx \beta l$
then (2) becomes

$$Z_1 = Z_L + j\omega L_o l \quad (4)$$

a series inductance plus the terminating impedance.

We can also write (2) as

$$Y_1 = \frac{\frac{1}{Z_L} + j \frac{1}{Z_o} \tan \beta l}{1 + j \frac{Z_o}{Z_L} \tan \beta l} \quad (5)$$

and if the inequality

$$\left| \frac{Z_o}{Z_L} \tan \beta l \right| \ll 1 \quad \text{holds} \quad (6)$$

with $\tan \beta l \approx \beta l$
then (5) becomes

$$Y_1 = \frac{1}{Z_L} + j\omega C_o l \quad (7)$$

a shunt capacitance across the terminating impedance.

For a coaxial line

$$L_o = \frac{Z_o}{V} \quad \text{henrys per cm} \quad (8)$$

$$C_o = \frac{1}{Z_o V} \quad \text{farads per cm} \quad (9)$$

So from (4) and (8), the inductance, L ,
of a short length of line

$$L = \frac{Z_o l}{V} \quad (10)$$

and from (7) and (9)

$$C = \frac{l}{Z_o V} \quad (11)$$

The development of (4) and (7) establishes
the design parameters for the short-line

filter as (a) characteristic impedances
(b) line lengths. The limitation on the
parameter (a) is physical, being depend-
ent primarily on the power to be trans-
ferred through the section⁵. The
parameter (b) effects the stop-band
performance with respect to both attenu-
ation and location of spurious pass-
bands². The latter occurs for values of
 $\theta = n\pi$, where n is integral and θ is
defined as the phase shift along a line
length l , and is given by

$$\theta = \frac{\omega l}{V_o} \quad (12)$$

Values of $\theta_L \approx 40^\circ$ for inductive
lengths and $\theta_C \approx 20^\circ$ for capacitive
lengths have been found practical for
low-pass power filters. θ_L being greater
than θ_C , the first spurious pass-band
is generally due to θ_L and has a width
which is approximately equal to $(\theta_C/\theta_L)f_c$.
The peak of the pass-band, though, is
extremely sharp, and the "width", as
defined here, represents points in the
spurious pass-band where considerable
attenuation still exists.

The discontinuity susceptance present
at the impedance transition where the
short-line elements of the low-pass filter
are joined together has been thoroughly
analyzed⁶. The discontinuity is a shunt
capacitance and for low-pass filters can
conveniently be lumped into the capaci-
tance section, the necessary modification
of which is given by²

$$\bar{\theta}_C = 2 \arctan \left[\tan \frac{\theta_C}{2} - B_C Z_{oc} \right] \quad (13)$$

$$B_C Z_{oc} = 5 f_c D C_d' Z_{oc} 10^{-5} \quad (14)$$

where

- $\bar{\theta}_C$ = the corrected phase angle
- D = outer diameter in inches
- f_c = cutoff frequency in megacycles
- C_d' = uuf/cm (as per reference #6)

With equations 10 through 14 as a
point of departure, the design of a
composite low-pass filter can be
attempted.

Specifications for Filters Used on Tele- vision Transmitters

On the basis of measurements made on
a large number of visual and aural trans-
mitters⁷ and in view of possible tighten-

ing of future F.C.C. specifications on spurious radiation, realistic minimum values of harmonic attenuation by the spurious emission filter would seem to be

- 2nd harmonic - 55 db
- 3rd harmonic - 45 db
- 4th harmonic - 40 db
- 5th harmonic - 20 db
- 6th and higher - usually very little.

This would provide a system level of at least -80 db on all harmonics. To minimize pass-band insertion loss and effect on transmitter loading, the filter VSWR throughout the operating channel should be a maximum of 1.1. The heat rise of the filter while passing its maximum rated power should be consistent with transmission line specifications.

VHF 25 KW Filters

The required decreasing attenuation levels for successive harmonics as outlined in the previous paragraph and the 1.1 VSWR specification suggests the use of m-derived sections. If $m = .6$ terminating half sections are employed, the image impedance of a conventional low-pass ladder filter will be within 5% of its characteristic impedance up to $.80 f_c$. The VHF television low-band (54-88 mc) consequently is conveniently handled by two low-pass filters, channels 2, 3, 4 and channels 5, 6 both of which will supply the required harmonic attenuation with a minimum of sections. For the VHF television high-band, a single filter suffices for channels 7-13.

The first filter discussed is the channel 2, 3, 4 unit which is required to pass a peak power of 25 KW. The lumped element values are calculated by the image parameter method. A cutoff frequency of 100 mc is chosen because the stop-band attenuation is met simply and the 80% pass-band includes channel 4. (Figure 1)

As the first step in the transformation of the filter to transmission line sections Z_{OL} is computed from the phase angle θ_L which is arbitrarily chosen to position the first spurious pass-band conveniently, i.e., somewhere above the fourth harmonic region of channel 4. A good starting point is to assume L_2 (the largest series inductance) (Figure 1) responsible for the first pass-band, although intuitively it may seem that $2L_3$ will produce a lower frequency pass-band, particularly if the shunt leg $L_4 + C_3$ presents a high impedance at or about the $\lambda/2$ frequency of $2L_3$. It just so happens that the first pass-band depends

on $L_2, 2L_3$ and $L_4 + C_3$, so it is necessary in the design of the inductive sections to utilize the smallest possible phase angles consistent with the power requirements. A $\theta_{L_2} = .8$ radians or 46° will develop a pass-band at about 390 mcs, so initially

$$Z_{OL} = \frac{6.28 \times 10^8 \times .1312 \times 10^{-6}}{.8} = 103 \Omega$$

A power requirement of 25 KW suggests either $3 \frac{1}{8}$ " or $6 \frac{1}{8}$ " transmission line. $3 \frac{1}{8}$ " line with a $3/8$ " inner conductor produces a $Z_{OL} = 125 \Omega$ which is greater than that arbitrarily calculated, but the higher impedance produces a more desirable phase angle per section. Estimation of the heat rise of the line section under full power is attempted by calculating the stable operating temperature of a $3/8$ " copper rod in free space. If use is made of the empirical thermodynamic relationship¹⁰

$$Q = UA \Delta t \quad (15)$$

where $Q = BTU/HR$

$$U = 2 \frac{BTU}{HR-DEGREES F-FT^2}$$

A = cross-sectional conduction area
 Δt = temperature rise degrees F

the temperature rise of the $3/8$ " copper rod is computed to be $110^\circ F$.

The inverse square law of classical physics suggests that the temperature rise of the outer conductor is proportional to the rise of the inner conductor times the square root of the ratio of the radii, which in this specific case would be $38^\circ F$.

If the internal filter sections are silver brazed and teflon dielectric is used, the computed full power operating temperatures indicate that $3 \frac{1}{8}$ " line can be rated at 25 KW for VHF short-line filter construction.

Referring to the schematic diagram of Figure 1, the short-line section lengths for L_2 and $2L_3$ are calculated using equation (8) with $Z_{OL} = 125 \Omega$. The first spurious pass-band due to $2L_3$ is found at 262 mcs. but this depends, among other things, on the leg L_4 and C_3 shunting a high impedance across the line, which is not the case at 262 mcs. The spurious pass-band due to L_2 is calculated at 500 mcs. The actual first spurious pass-band, therefore, should lie somewhere between 262 and 500 mcs. since it depends on mutual impedance relationships between the various short-line

sections, and is probably a function of $L_2 / (2L_3 + L_2)$ ¹¹. This being the hypothesis, the frequency of the first actual pass-band is estimated as being 346 mcs., which is well above the fourth harmonic frequency of channel 4.

The impedance $Z_{oc} = 7.8 \Omega$ provides a satisfactory phase angle for C_2 of 17° . This impedance is realized with a $3 \frac{1}{8}$ " line of solid teflon dielectric and an inner conductor of $2 \frac{1}{2}$ ". If the dielectric, which has a width of $\frac{1}{4}$ " is extended another $\frac{1}{4}$ " at each end of the short-line low impedance section (Figure 2), a voltage breakdown safety factor of 5 to 1 at a 25 KW level is obtained¹². In addition, the fringing field at the impedance transition is partially in the dielectric, increasing the discontinuity susceptance and lowering the value of Θ_c (equation 13). As has been pointed out previously, a low value of Θ_c is desired, Θ_c being the limiting factor on the width of the first spurious pass-band. For this specific case $\Theta_c = 13^\circ$, making the pass-band at 346 mcs. about 20 mcs. wide. The composite filter construction can now proceed. The legs $L_1 C_1$ and $L_4 C_3$ are of course realized with coaxial tee sections (Figure 3). The calculation of C_1 and C_3 is straightforward, except that in closing off the outer conductor as indicated in Figure 3, an additional parallel plate capacitance is added. The latter suggests a simple method for compensation of the effects of mechanical tolerances on the electrical performance of the filter. If the design values of C_1 and C_3 are purposely constructed somewhat less than required, the outer conductor plate can be made adjustable and the filter can be "tuned" for minimum VSWR in the usable pass-band without appreciably effecting the stop-band performance.

Figure 4 shows the construction of the composite filter. The inner conductor, except at the terminals, is built of solid copper. Use of solid copper in the low impedance sections permits better conductivity of heat away from the $\frac{3}{8}$ " rods than would occur if hollow copper drums were used. The thermal conductivity of the teflon dielectric is enough to afford a good path to the outer conductor. Additional cooling is supplied by drilling four $\frac{1}{4}$ " holes radially spaced through the length of C_2 and locating heavy wire mesh screens on the outer conductor, (Figure 4). Natural convection currents permit the hot air in the filter to escape through the center screen to be replaced by cooler air drawn in through the end screens. Expansion of the inner conductor is allowed along the series bullets located near each flange. The

flanges are swivel type affording maximum physical flexibility when mounting the unit in the transmission system. At 25 KW levels, though, it is necessary that the center screen face up to provide the maximum effective convection cooling.

Experimental verification of the paper design of the channel 2, 3, 4 25 KW filter was almost completely realized. Its transmission characteristics are given in Figure 5 where close correlation with the characteristics of Figure 1 are apparent. The major discrepancy involves the location of the first spurious pass-band which appears 26 mcs. lower than calculated. However, it was found that removal of the two support beads near the center tee shifted the pass-band out to 335 mcs. The support beads could not be sufficiently undercut due to the small diameter of the inner conductor and consequently represent discontinuities (of no effect in the pass-band) which tend to "tune" with the leg $L_4 C_3$ in determining the spurious pass-band center frequency. The beads were replaced and the pass-band left at 320 mcs. since it did not fall on any specific harmonic frequency. The spurious pass-band width (Figure 5) is about 25 mcs. at the 30 db points. The insertion loss at the 5th harmonic frequencies of the channel 3 visual and aural carriers is about 20 db, which is adequate. The target values of 55, 45 and 40 db for 2nd, 3rd and 4th harmonic attenuation, respectively, are met.

The channel 5, 6 25 KW filter (Figure 6) is a replica of the channel 2, 3, 4 unit, differing only in its linear dimensions.

A single filter serves the VHF high band, channel 7-13, at the 25 KW level. It duplicates the low-band design except that two additional constant-K sections are added giving the symmetrical configuration $m_1/2 + 2K + m_2 + 2K + m_1/2$ and a stop-band attenuation which is literally out of sight, (Figure 7). A similar stop-band performance could have been achieved on the low-band filters by the addition of one or two constant-K sections but the filter lengths begin to get excessive and a system attenuation of 80 db does not demand it. It is not inconceivable, however, that some situation may require, say 120 db system attenuation on a specific harmonic. If so, note (Figure 4) that the end section of the filter may be removed from the main body and as many constant-K sections added as desired.

With reference to heat rise, the power tests for the 25 KW filters were

performed on the channel 5, 6 prototype. It may be recalled that the calculated temperature rise was 110°F for the inner conductor. Experimentally, the inner conductor stabilized between 200° and 225°F with an ambient of 95°F under continuous full power operation over a period of four hours. The outer conductor stabilized at 120°F, which was more or less as expected.

Figure 8 is a photograph of the three 25 KW VHF filters. These units lend themselves readily to production in reasonable quantities without any deterioration in performance. The mechanical tolerances on machined parts are a strict $\pm .002$ " and on most line lengths $\pm 1/64$ ", ensuring uniformity of product.

50 KW VHF High-Band Filter, Channels 7-13

If operating temperatures similar to those observed on the 25 KW filters are to be realized at 50 KW, the coaxial line diameters should be increased.

Intuitively, therefore, it is only necessary to scale up the 3 1/8" VHF high-band filter to 6 1/8", but closer investigation reveals that the shunt elements of the series m-derived sections are inconsistent with the larger line diameter. Lowering Z_{oL} of the shunt sections is an apparent solution, but this lowers the frequency of the first spurious pass-band objectionably.

A satisfactory solution can be obtained by utilizing shunt m-derived terminating half sections in the symmetrical ladder structure $m/2 + 3K + m/2$.

Figure 10 illustrates the physical realization of the shunt m-derived half section as used on the 50 KW channel 7-13 spurious emission filter. The effective impedance of the series leg L, C , (Figure 9) must be duplicated by the outer conductor coaxial cavity for proper operation of the filter. The correct cavity impedance Z_{o2} is calculated as 25 ohms¹³, and its length should be $\lambda/4$ at 375 mcs. (the frequency of maximum attenuation for the $m = .6$ section). It is convenient, however, to construct the cavity from standard transmission line, 3 1/8" and 6 1/8"; consequently the Z_{o2} value of 39 ohms is obtained, which, it will be seen, is not fatal. Since Z_{o2} is given as 39 ohms, equations (16) and (17) may be equated and $\tan \theta$ is found to equal .685 at 195 mcs. This places the quarter-wave resonant frequency of the 39 ohm cavity at 447 mcs. If the resonant frequency is raised to 490 mcs. and an adjustable capacitive slug inserted across the cavity mouth (Figure 10), the proper pass-band

impedance is obtained by adjusting the slug for minimum filter VSWR. Furthermore, the slug has much more effect on the cavity resonant frequency than its pass-band impedance, the result being that when minimum pass-band VSWR is achieved, the actual resonant frequency of the capacitive loaded cavity is 385 mcs., close enough to 375 mcs. for proper stop-band performance. Needless to say, the slug also provides a convenient method for "tuning out" mechanical discrepancies. The 1" gap at the cavity mouth is chosen on the basis of arc over requirements.

In the balance of the channel 7-13 50 KW filter (Figure 11) $Z_{oL} = 125$ ohms and $Z_{oC} = 7.8$ ohms are again used, except that overlapping of the dielectric is carried to greater lengths because of the higher voltages involved. Consequently, calculation of the exact value of θ_c is so difficult that the correct lengths of the low impedance sections are finally obtained by cut-and-try. The inner conductor is constructed of silver plated brass, its expansion being allowed along the center bullet.

Furthermore, the inner conductor is physically anchored with the outer conductor at the teflon sleeve of C_z by a teflon pin held in place externally with a hose clamp. Additional inner conductor lateral support is supplied by anchor insulator bullets at the filter terminals. The brass tuning slug has a teflon cup on the end which protrudes into the cavity, providing a long arc over path.

The damping filter, shown coupled into the 5394-A input cavity (Figure 11) is a lumped element high-pass filter with a cutoff frequency of 300 mcs., its stop-band insertion loss in the region 174-216 mcs. being greater than 40 db. The purpose of the damping filter is to provide sufficient loading on the line feeding the 5394-A so that the VSWR on harmonics is restricted to 20 to 1, precluding the possibility of harmonics arcing over the spurious emission filter or the input feed line. The impedance of the input cavity being much greater on harmonics than in the channel pass-band, adequate loading is obtained by light coupling with no noticeable effect on the pass-band VSWR.

As expected, the stable operating temperature of the 50 KW filter under full power is approximately the same as the full power temperature of the 25 KW design.

Each of the three constant-K sections is convection cooled by perforating the

outer conductor (Figure 11). The first spurious pass-band appears 200 mc. lower than calculated, but not low enough to cause trouble. This discrepancy apparently has to do with the height of the capacitive sections and is discussed in the section on UHF filters. The pass-band peak is over 10 db down which may be due to moding inside the filter proper.

UHF Filters

The 50 KW design is applicable to the UHF. Three filters, having a power rating of 12.5 KW and respective cutoff frequencies of 800, 1000 and 1190 cover the seventy channel UHF television band. Due to its lower power rating, the UHF filter (Figure 12) becomes physically quite simple. The only variations from the 50 KW design lie in the closer approximated outer conductor cavity impedances and the 12.5 ohm Z_{oc} . With reference to the latter, the higher impedance was originally chosen to give the capacitive short-line sections additional length since the major part of the calculated capacity at 7.8 ohms was supplied by the impedance discontinuity. However, an additional spurious pass-band relatively close to the cutoff frequency has been observed on these filters which is hypothetically a function of the physical height of the capacitance short-line section. If the section is viewed as a foreshortened quarter-wave stub support on a high impedance line, the attenuation of the composite filter will depend only on its series elements at the resonant frequency of the stub supports, and when impedance relationships are optimum throughout the filter, a spurious pass-band of low attenuation exists. This hypothesis has been experimentally verified by changing the impedance of the capacitance sections with a resultant shift in the spurious pass-band center frequency. Like the pass-band due to the inductive lengths, this additional pass-band is extremely sharp at the peak and narrow at the base. On the 1000 mc. cutoff filter which operates on channels 37-59, the peak of the additional pass-band was observed at 1725 mcs. at an attenuation level of 25 db, disrupting an otherwise 60 db stop-band extending from 1250 to over 3000 mcs. The channel 60-83 filter, $f_c = 1190$ has a 30 db peak at 1825 mcs. Both filters use $Z_{oc} = 12.5$ ohms and, fortunately, the pass-band being discussed lies between the second and third harmonic regions of the respective operating channels of the filters. A reversion to 7.8 ohms was necessary on the channel 14-36 filter ($f_c = 800$ mc.) to shift the 25 db spurious pass-band from 1650 mcs. (12.5 ohms) to 1375 mcs., just below the third

harmonic of channel 14.

The channel 14-36 filter passed 15 KW of peak power continuously for four hours with a resultant outer conductor heat rise of only 25°F.

Insertion of the Filter in the Transmission System

There are some theoretical and empirical considerations concerning the insertion of a spurious emission filter into the television transmission system which it is worth while to discuss. A filter with a VSWR ≤ 1.1 will contribute little to the VSWR of the system in the operating channel of the transmitter (the maximum possible system VSWR being equal to the product of the VSWR's of all the components in the system. Theoretically, therefore, the filter can be inserted anywhere in the transmission system. In practice, location of the filter directly at or close to the transmitter output terminals provides two obvious advantages: (a) the filter need not be pressurized, a distinct mechanical design economy, thus natural convection cooling replaces costly blowers and associated interlocks; (b) adverse effect of high harmonic VSWR on transmitter test equipment and indicators is more conveniently eliminated. With reference to the latter, if a filter is in the system, non-frequency sensitive devices such as envelope detectors, diode demodulators, broad-band directional couplers are likely to produce erroneous information about the channel pass-band. This has been traced to the presence of high harmonic VSWR on the transmission line into which the devices were coupled. Consequently, the coupling probes of such devices should succeed rather than precede the filter, and this is most easily accomplished when the filter is physically close to the transmitter.

The behavior of the filter in the stop-band can be computed or measured. Initially the attenuation is probably calculated utilizing the attenuation function which depends only on the type and number of sections in the composite filter. Secondly, the filter insertion loss is measured in the laboratory by the substitution method using real fixed source and terminating impedances. If the filter is non-dissipative and perfectly designed, the two methods should not be at variance by more than 6 db¹⁴. When the filter is inserted in the television transmission system, it is the system loss rather than the filter loss which must be considered since the nature of the terminating impedance on the filter is nebulous. This situation has been

analyzed³.

The analysis shows that under certain conditions it is possible for the filter to become completely transparent, in which case the total power generated by the transmitter is delivered to the load. This occurs when the ratio of the VSWR's at the filter input and output is unity. On the other hand, it is pointed out that the equal VSWR's may produce a db system loss twice that of the filter⁵. The case of filter transparency seems to be a gloomy one at first glance, but mathematical probability and the basis of the analysis on non-dissipative network shows that nature is with us. The practical television transmission system is of course, not non-dissipative. Consider the hypothetical case of total transparency on a typical installation where the transmission line run to the antenna will usually be about 400 feet of 3 1/8" line.

The filter reflection coefficient is no less than .999 for frequencies at which the attenuation of the filter is 40 db or more². For transparency the load reflection coefficient is also .999. The magnitude of the voltage insertion ratio is given by⁶

$$\left| \frac{E_R'}{E_R} \right| = e^{\alpha x} \left| \frac{1 - K_R K_G e^{-2\tau x}}{1 - K_R K_G} \right| \quad (18)$$

where $\alpha = .384 \times 10^{-4}$ nepers/ft. for a 3 1/8" 50 Ω air dielectric copper coaxial line at 50 mcs.

$$x = 400 \text{ ft.}$$

$$\tau = (1+j)\alpha$$

$$K_R = K_G = .999$$

from which

$$\left| \frac{E_R'}{E_R} \right| \approx \left| \frac{.040}{.001} \right| = 40$$

an insertion loss of 32 db. The key factor in equation (18) is $e^{-2\tau x}$ upon which the insertion loss of the system depends.

As the transmission line separating the filter from the antenna reduces in length, the possibility of low system harmonic attenuation increases. On the other hand, the mathematical probability for total transparency tends toward zero. In fact, the probability that the filter transmission loss will decrease by as much as 6 db is only .25. All of which seems to indicate that large decreases in system harmonic attenuation are not

generally likely but that it is probably worth while to consider in some specific cases.

Of course, the real danger resulting from high VSWR is that a few watts of harmonic power may be sufficient to arc over the line or the filter. This phenomenon has been observed on 50 KW transmitters where 3 1/8" line is commonly used for short runs inside the transmitter room, but it should be noted that the observed arc overs occurred only on full power stopping and starting transients, not under steady operation. As previously mentioned, the damping filter used on the 5394-A is designed to eliminate the possibility of arc over by restricting the VSWR on the input line due to the filter in the harmonic region.

References

1. F.C.C. Docket #10353.
2. Very High-Frequency Techniques; Radio Research Laboratory, Harvard University; McGraw-Hill 1947; Vol. II, Chapters 26 and 27.
3. Ragan, G.; Microwave Transmission Circuits; McGraw-Hill 1948.
4. Cuccia, C. & Hegbar, H.; RCA Review, Dec. 1947, p. 743.
5. We are discussing power filters here. Obvious limitations which are inconsistent with power design occur when the parameter approaches either extreme.
6. Whinnery, Jamieson, Robbins; Coaxial Line Discontinuities; Proc. I.R.E., Nov. 1944.
7. Towlson, H.; Report to RETMA Subcommittee TR4.1 on TV Transmitters Spurious Radiation.
8. Since mechanical discrepancies will exist, a theoretical design VSWR = 1.05 should produce a VSWR \leq 1.1 in practice.
9. The term "convenient" has a double meaning. A more complex structure could have been designed to cover channels 2-6 with VSWR \leq 1.1 and adequate stop-band attenuation. By keeping the filter design simple, development and consumer costs are minimized.
10. Faires, V. M.; Applied Thermodynamics; Macmillan 1947; p. 428.

11. Guesstimate of the center frequency of the first spurious pass-band:

$$f_{PB} = \left[262 + (500 - 262) \left(\frac{L_2}{2L_3 + L_2} \right) \right]$$

12. Pim, J. A.; The Electrical Breakdown Strength of Air at UHF; Proc. I.R.E., 1949.

13. The calculation of the correct cavity impedance proceeds as follows: for the lumped filter, the impedance of the leg $L, C,$

$$Z = \frac{j\omega m R}{1 - \frac{\omega^2}{\omega_c^2} (1 - m^2)} \quad (16)$$

and for a short circuited transmission line of length $< \pi/2$

$$X_L = j Z_0 \tan \theta \quad (17)$$

letting $X_L = Z$, and $f = 195$ mcs. (mid-frequency VHF high-band)

$$\theta = \left(\frac{195}{375} \right) \frac{\pi}{2}$$

and $Z_{02} = 25$ ohms.

14. Guillemin, E.; Communications Networks; Cohn Wiley 1935; Vol. II, p. 290. The interaction loss is being disregarded for reasons mentioned in the text.

15. In essence, the subject treatment in Microwave Transmission Circuits differs from that in Guillemin only in that the latter assumes real terminations.

16. Johnson, W. C.; Transmission Lines and Networks; McGraw-Hill 1950, p. 121.

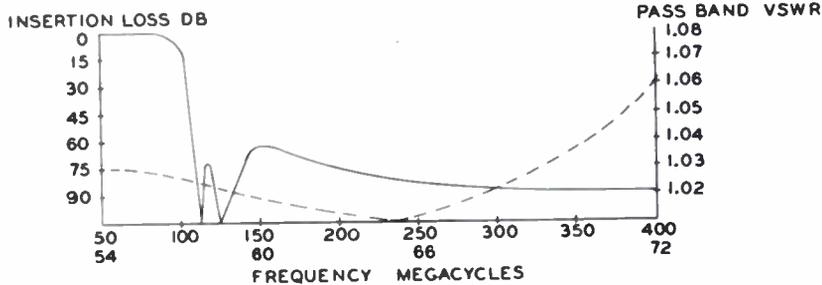
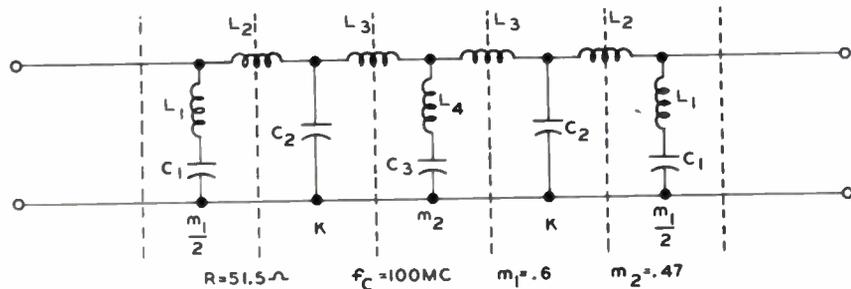


Fig. 1
Theoretical Filter, Channels 2-4

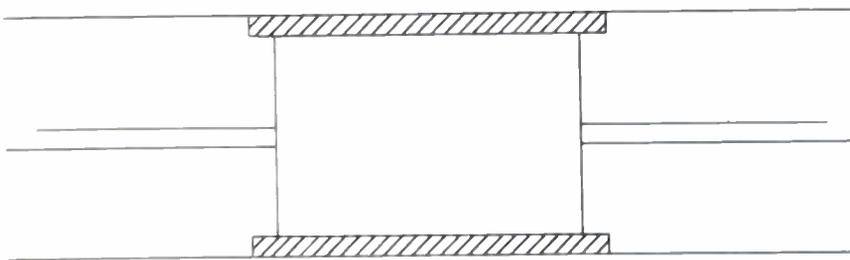


Fig. 2
Overlapping of the Dielectric in the Low Impedance Section to Provide Longer Voltage Breakdown Path

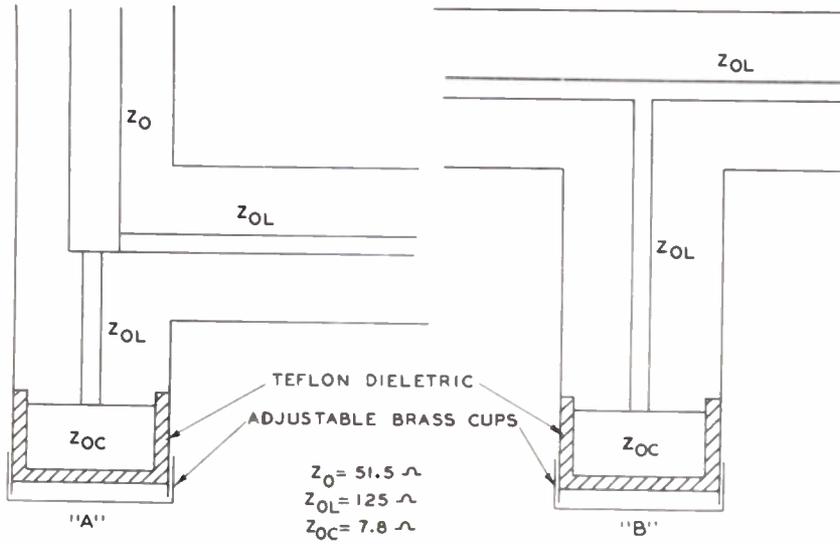


Fig. 3
Coaxial Series m-derived Sections
"A" Terminating Half Section
"B" Full Section

38

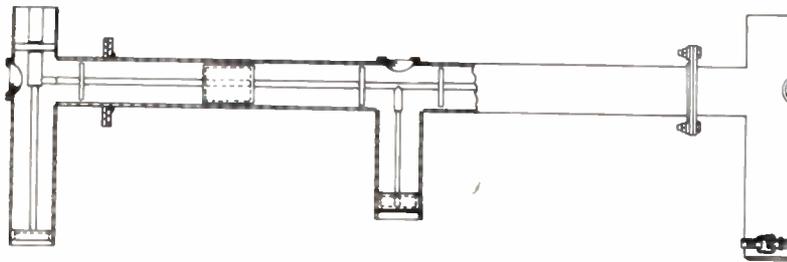


Fig. 4
The Du Mont 5391-A Channels 2, 3 & 4
25 KW Spurious Emission Filter

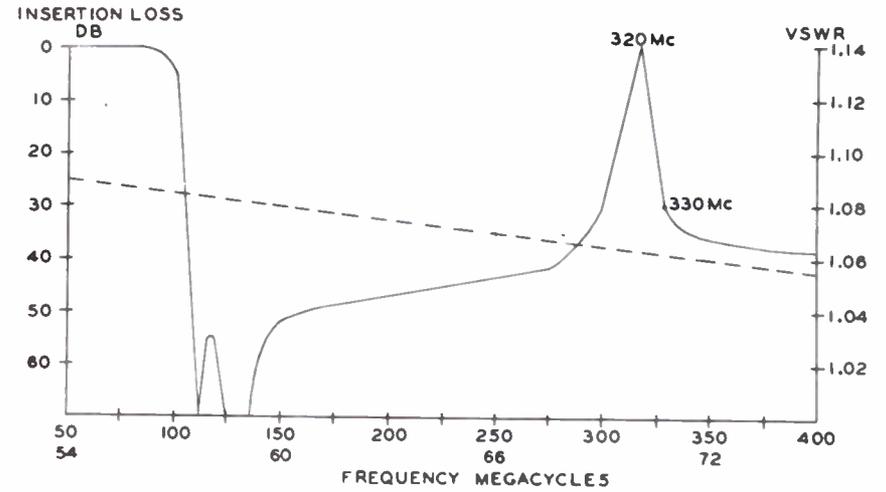


Fig. 5
Measured Transmission Characteristics of the
5391-A

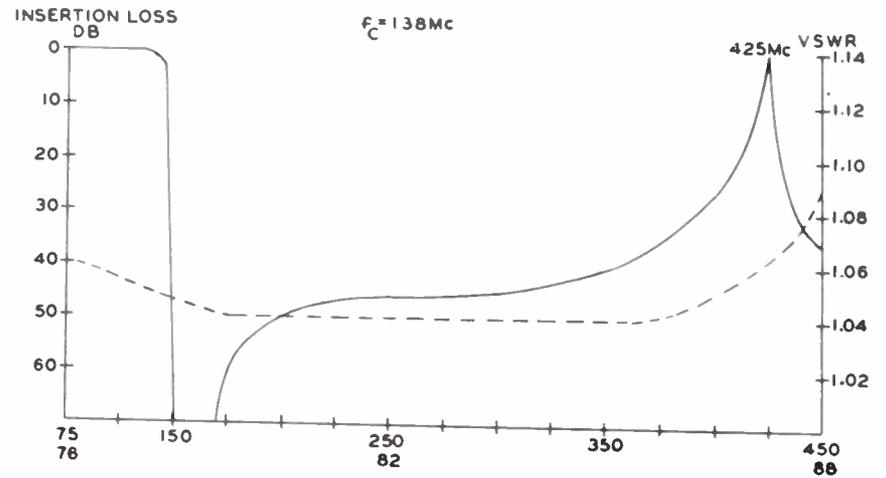


Fig. 6
Measured Transmission Characteristics of the
5392-A

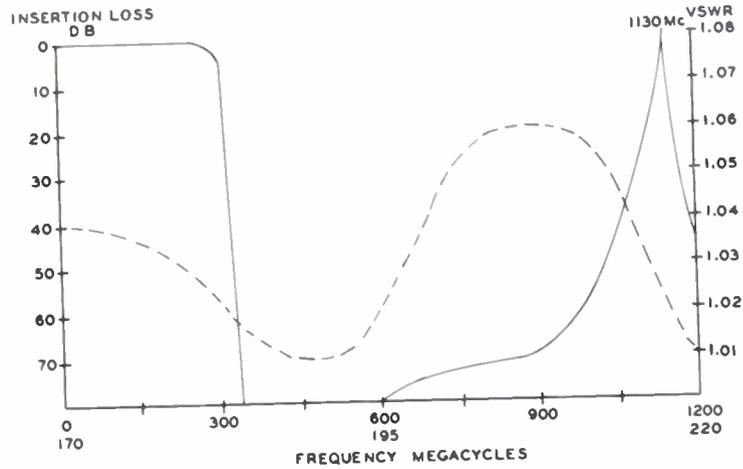


Fig. 7

Measured Transmission Characteristics of the 5393-A

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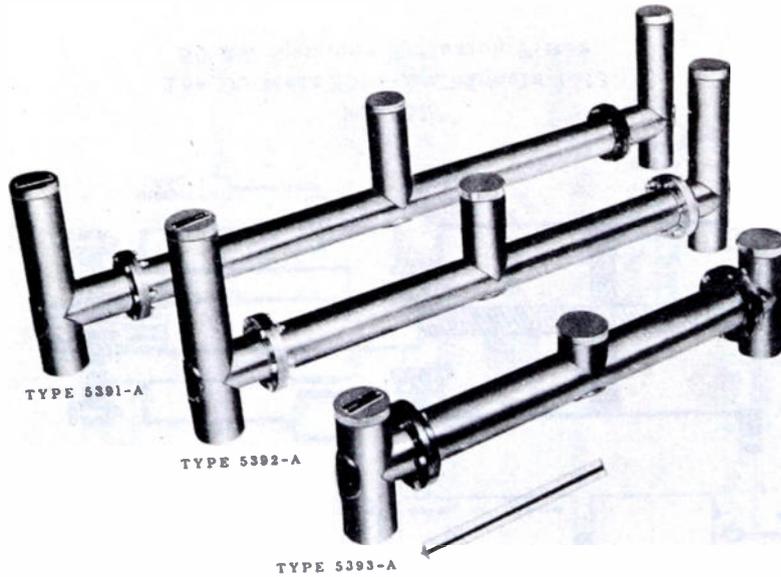


Fig. 8

25 KW VHF Transmission Line Filters

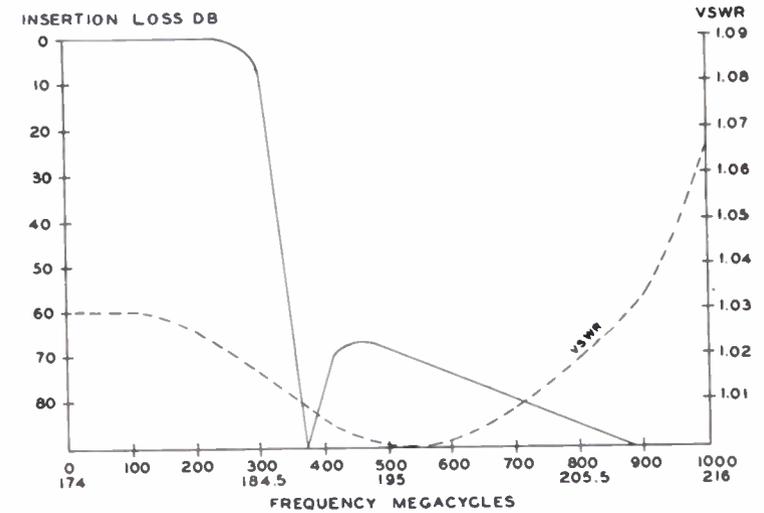
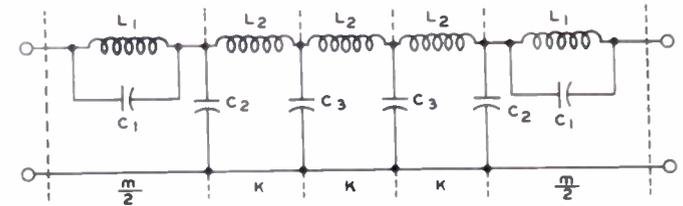


Fig. 9

Theoretical Filter, Channels 7 and 13

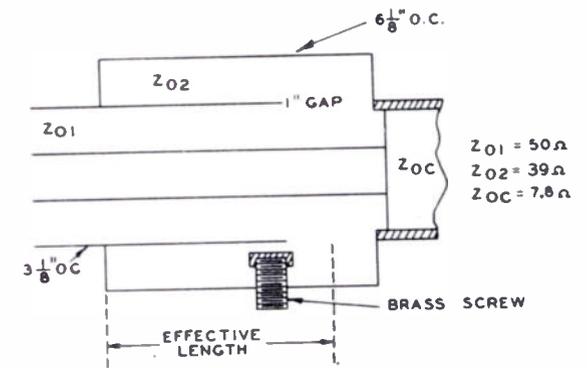


Fig. 10

Coaxial Shunt m-derived Terminating Half-Section

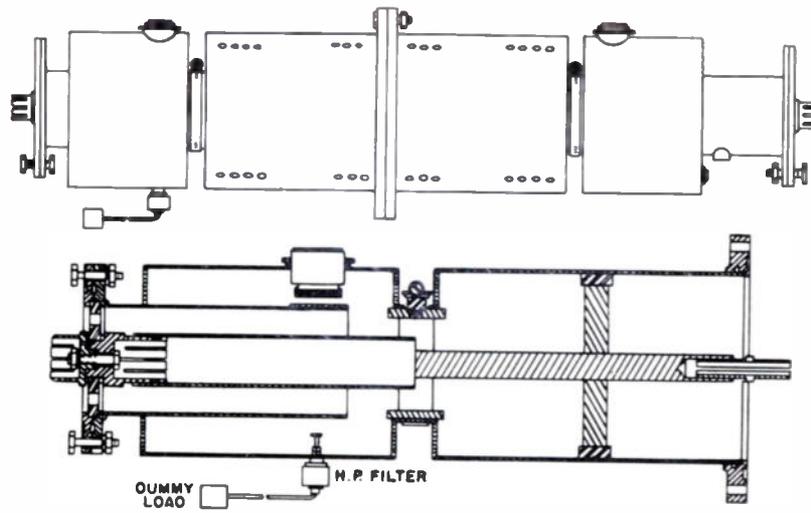


Fig. 11
The Du Mont 5394-A Channels 7-13
50 KW Spurious Emission Filter

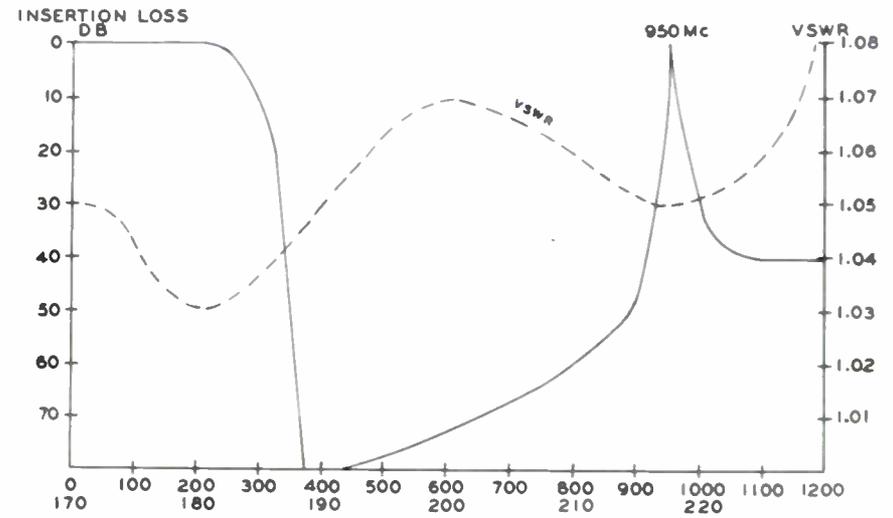
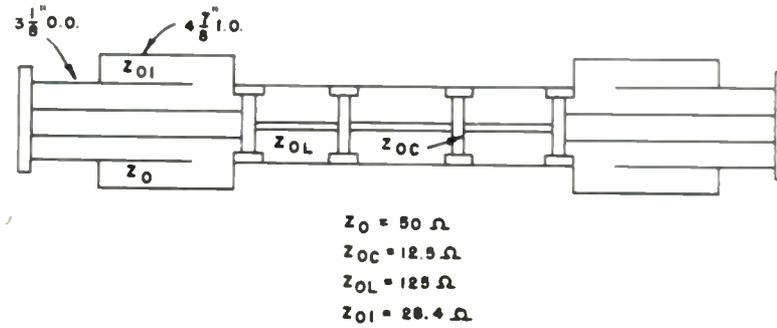


Fig. 12
Measured Transmission Characteristics of the
5394-A



$Z_0 = 50 \Omega$
 $Z_{0c} = 12.5 \Omega$
 $Z_{0L} = 125 \Omega$
 $Z_{0I} = 28.4 \Omega$

Fig. 13
UHF Transmission Line Low-Pass Filter

ELECTRONICALLY CONTROLLED AUDIO FILTERS*

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Summary

The use of filters whose cutoff characteristics are controllable by electronic means is often desirable in problems dealing with audio signals. Based on the recent work on fixed RC active filters by J. G. Linvill, variable active low-pass and high-pass filters have been developed using transistor negative impedance converters.

The design theory of such filters is summarized and measured characteristics and other experimental results are presented. An application, in which the cutoff characteristics are controlled by the incoming audio signal for use in formant tracking, will be described and experimental results given.

Introduction

In certain problems dealing with audio signals** it is desirable to use filters whose cutoff characteristics are continuously variable in accordance with an applied voltage or current signal, preferably with a minimum of delay. The requirement of variable cutoff characteristics implies (a) that some of the filter components can be varied almost instantaneously by means of a suitable control signal, (b) that such a control signal can be obtained. The control signal may either be related to some property of the incoming audio signal, or it may be generated by independent means.

In the application considered a low-pass and high-pass filter were required, both having cutoff frequencies continuously variable between 200 and 1000 cps.

Theoretical Considerations

The filter design adopted was based on the recent work on fixed RC active filters by J. G. Linvill¹. In this design a negative-impedance converter is used in addition to passive elements; however, the magnitude of the reactive filter elements used here can be varied by electronic means so that electronically variable filter cutoff frequencies are obtained.

* This work was performed under Contract AF19 (604)-1039 with the Air Force Cambridge Research Center of the Air Research and Development Command.

** e.g. formant tracking in speech analysis.

The block diagram of either the low-pass or the high-pass filter constructed is shown in Fig. 1. The transfer function of the entire filter in terms of the open circuit driving-point and transfer impedances of the individual networks can be shown to be equal to

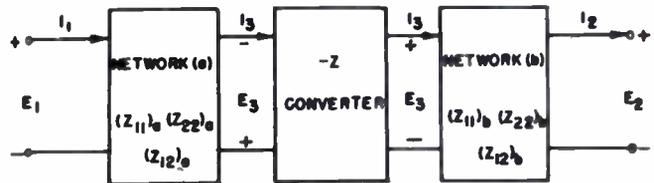


Fig. 1. Block Diagram of HP or LP Filter.

$$H(s) = z_{21} \left| \begin{matrix} E_2 \\ I_2=0 \\ I_1 \end{matrix} \right|_{I_2=0}^{-1}$$

$$= \left[\frac{-1}{(z_{12})_a} \cdot \frac{(z_{11})_b}{(z_{12})_b} + \frac{(z_{22})_a}{(z_{12})_a} \cdot \frac{1}{(z_{12})_b} \right]$$

$$= \frac{(z_{12})_a (z_{12})_b}{(z_{22})_a (z_{11})_b}$$

H(s) will be a ratio of two polynomials and can be written

$$H(s) = \frac{N(s)}{D(s)} \quad (2)$$

In order to synthesize H(s) by means of RC and RL networks only, it is convenient to divide both N(s) and D(s) by

$$M(s) = (s - \sigma_1)(s - \sigma_2) \dots (s - \sigma_n) \quad (3)$$

where $\sigma_1, \sigma_2 \dots \sigma_n$ represent the poles of $(z_{22})_a$ and $(z_{11})_b$.

The denominator of H(s) becomes

$$D'(s) = \frac{D(s)}{\prod_{i=1}^n (s - \sigma_i)} \quad (4)$$

$$= k_0 + \sum_{i=1}^n \frac{k_i}{s - \sigma_i} = (z_{22})_a - (z_{11})_b,$$

while the numerator is equal to

$$N'(s) = \frac{N(s)}{\prod_{i=1}^n (s - \sigma_i)} = (z_{12})_a (z_{12})_b. \quad (5)$$

If in eq. 4 the positive k 's are associated with $(z_{22})_a$ and the negative ones with $(z_{11})_b$, the networks a and b can be synthesized, e.g. by Cauer's continued fraction expansion. The negative sign before $(z_{11})_b$ is supplied by the negative-impedance converter.

Low-Pass Filter

The desired low-pass filter characteristic is of the Butterworth type ($n = 3$), so that, in normalized form,

$$H(s) = \frac{k}{(s + 1)(s^2 + s + 1)} \quad (6)$$

The distribution of the poles of $H(s)$ is shown in Fig. 2.

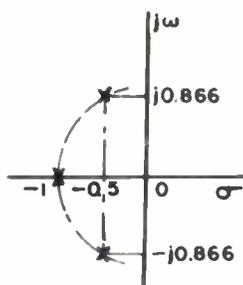


Fig. 2. Pole Pattern for Butterworth LP Filter.

A considerable freedom of choice exists about the location of the σ_i 's. In the present design, $\sigma_1 = -0.3$, $\sigma_2 = -1.05$ and $\sigma_3 = -1.5$ have been chosen. The circuit element values, after scaling to a suitable frequency and impedance level, are given in Fig. 3. As indicated in the diagram, three electronically controllable capacitors are used in the filter. The negative-impedance converter simulates at its input terminals the negative of the physical impedance connected across its output terminals, which in this case is composed of a 1300-ohm resistor and a capacitor whose value varies between 0.08047 and 0.4023 μ f.

Negative Impedance Converter

The requirements which the negative impedance converter must meet exceed the possibilities of a combination of passive elements alone. The converter must therefore contain active elements as well. As indicated by Fig. 4, two transistors

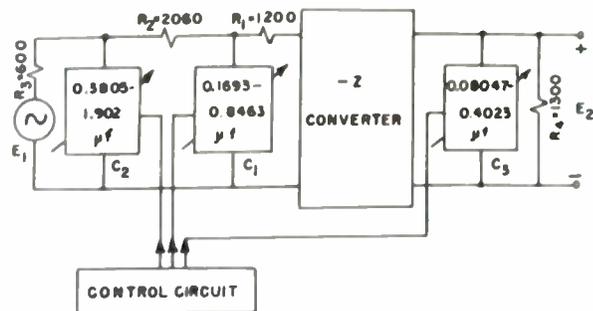


Fig. 3. Simplified Circuit Diagram of Electronically Controllable LP Filter.

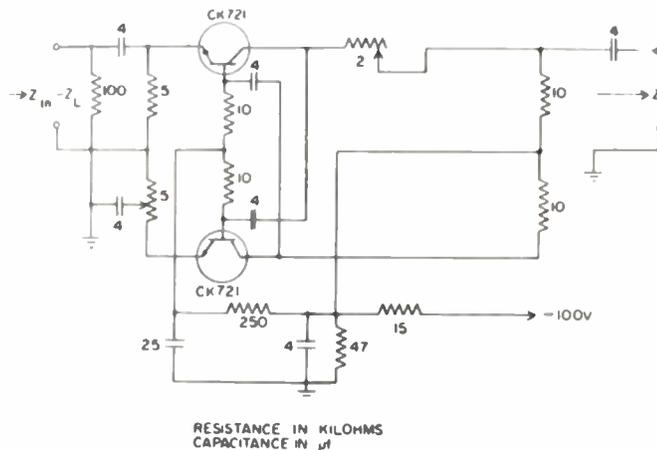


Fig. 4. Transistor Negative-impedance Converter.

are included in the negative-impedance-converter circuit. When Z_L is connected to the output terminals, the apparent input impedance Z_{in} is approximately equal to $-Z_L$. Since this circuit differs only in minor details from the original design described by J. G. Linvill¹, a detailed description of its operation will be omitted here. Figure 5 represents the values of the input impedance magnitude $|Z|$ and phase angle θ for a purely resistive $Z_L = 1.02$ kohm. It is seen that $|Z|$ is substantially constant for 200 cps $< f < 20$ kc, and increases rapidly for frequencies lower than 200 cps. The phase varies from -10.5° to $+12.5^\circ$ for 100 cps $< f < 10$ kc.

Variable C Circuit

In the electronically controllable low-pass filter three electronically controllable capacitors are required. A three-tube circuit, whose apparent input capacitance can be varied by means of a varying dc voltage, has been designed. The operation, which is based on the so-called Miller effect, can be explained by means of the simplified circuit shown in Fig. 6.

Assuming that the loading of the plate

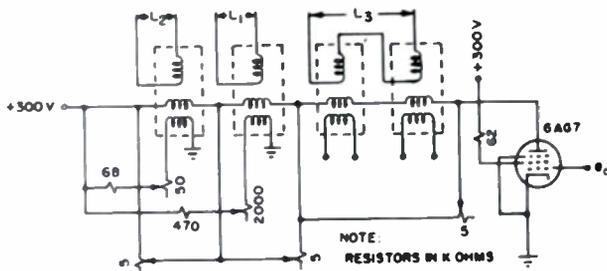


Fig. 10. Control Circuit for Electronically Controlled HP Filter.

Tracking of the Electronically Controllable Inductors

As in the case of the electronically controllable capacitors, the inductors should change by the same ratio for a given change in control signal. This is achieved by connecting all the main control windings of all increductors in series and bypassing each of them by a suitable resistor so that only the desired fraction of the total varying control current passes through the winding. If a constant difference in magnetization between any two Increductors is desired, this can be achieved by means of the second control winding, usually called "bias" winding, which is connected to +300v through a suitable resistor.

The measured variation of the electronically controllable inductors is given in Fig. 11.

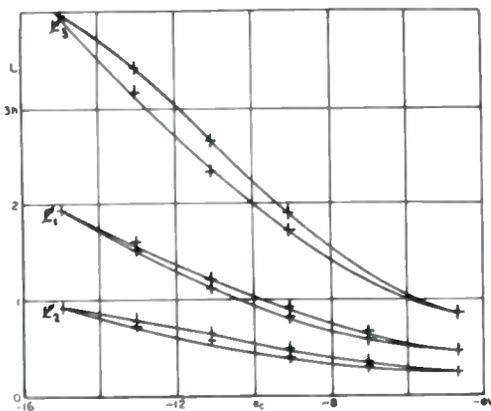


Fig. 11. Variation of L_1 , L_2 and L_3 vs. Control Voltage e_c .

Because of the large control-winding inductance and comparatively small parallel resistance, the time constant of the control circuit is somewhat excessive. It is believed that this drawback can be minimized by using a separate control pentode for each Increductor and by reducing the total control current, rather than by passing a part of

it through the parallel resistor. Another difficulty encountered with the use of Increductors is a certain amount of hysteresis. Both these effects will be demonstrated presently by means of dynamic-behavior oscillograms, representing the output signals of the filters.

Experimental Circuit

Figure 12 represents the electronically controllable high-pass filter, without the control

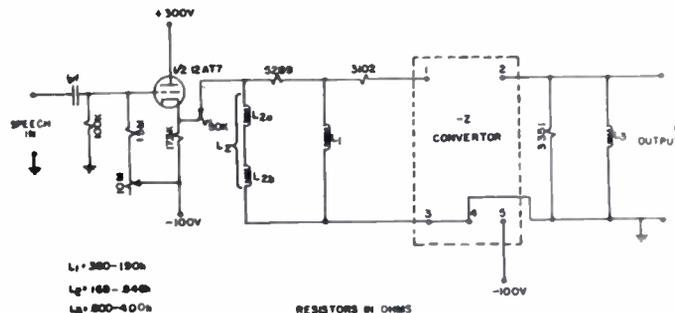


Fig. 12. Electronically Controlled HP Filter.

circuit which has already been shown in Fig. 10. As in the case of the low-pass filter, two of the variable reactances are before, one behind the negative-impedance converter. In order to secure more favorable conditions with respect to hysteresis and Q, L_2 is composed of two Increductors L_{2a} and L_{2b} . In this way it is possible to operate all Increductors nearer to their optimum ranges.

Experimental Results

Low-Pass Filter

A family of curves representing the sinusoidal steady-state response of the low-pass filter for various values of the control voltage e_c is given in Fig. 13.

The attenuation in the stop-band is at a rate of about 17 db/octave. The theoretically expected value is 18 db/octave. Once the attenuation of 35 db is reached, the response remains below this value for all higher frequencies. The residual output at these higher frequencies seems to be mostly due to high-frequency noise generated by the transistor-negative-impedance circuit, with a smaller component due to 60 cps hum. A larger input signal will increase the difference between the pass-band and stop-band response, but at the same time a distortion of the sinusoidal output is observed.

Figure 14A represents the output of the low-pass filter for three values of the control voltage, corresponding to three cutoff frequencies -

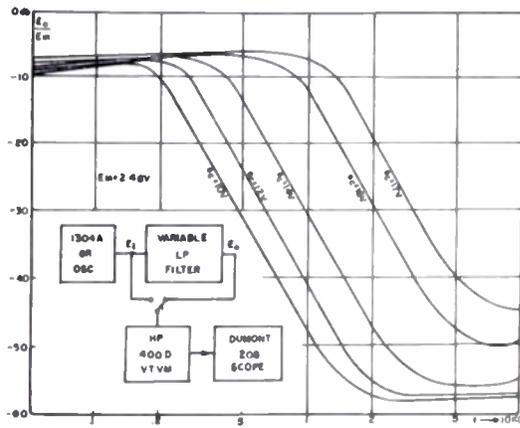


Fig. 13. Response of Electronically Controlled LP Filter for Various Values of Control Voltage e_c

200 cps, 500 cps and 1000 cps - when the frequency of a constant-magnitude signal sweeps between 100 and 1500 cps. In Fig. 14B the input-signal frequency is fixed but the control voltage varies sinusoidally at various rates. In Figs. 15A and B the control signal varies in a saw-tooth and square-wave fashion, respectively, the remaining conditions being the same as in the preceding figure.

High-Pass Filter

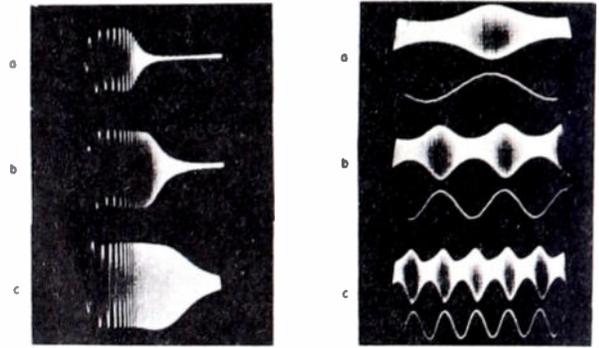
A family of measured response curves of the electronically controllable high-pass filter is given in Fig. 16. It is seen that the desired 5:1 variation in cutoff frequency can be achieved, and the filter attenuates at a rate of about 16 db/octave, with a stop-band attenuation of 45 db or more.

The response for a constant-magnitude sweeping-frequency input signal is demonstrated by Fig. 17A, again for three fixed values of the control signal corresponding to three cutoff frequencies. The response of the filter in the case of a constant sinusoidal signal, with the control voltage varying in a square-wave fashion, is shown in Fig. 17B. Although the control voltage assumes the new value almost instantaneously, the output reaches the new steady-state response only after considerable delay.

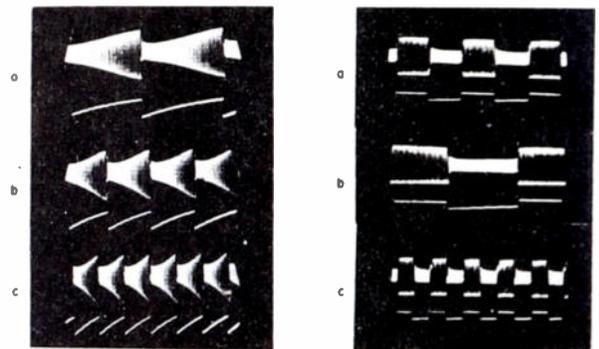
Figure 18 represents the output of the filter when the control voltage is varied very slowly in a triangular fashion, so that the sluggishness of the control circuit is minimized. This oscillogram is indicative of the effect of hysteresis upon the filter response. Variations of up to about 10 percent of the total inductance value have been observed when the Increductor was subjected to the maximum control-current variation, required for the 5:1 change in inductance.

Application to Formant Tracking

In speech analysis, it is often desirable to obtain automatically an approximate indication of



14A. (left) $f_s = 100 - 1500$ cps, e_c adjusted for $f_c = 200$ cps (a), 500 cps (b), 1000 cps (c).
14B. (right) $f_s = 500$ cps, sinusoidal control voltage, $f_{cv} = 5$ cps (a), 10 cps (b), 20 cps (c).



15A. (left) $f_s = 500$ cps, saw-tooth control voltage, $f_{cv} = 5$ cps (a), 10 cps (b), 17 cps (c).
15B. (right) $f_s = 500$ cps, square-wave control voltage, $f_{cv} = 10$ cps (a), 5 cps (b), 20 cps (c).

Figs. 14 & 15. Filter Output for Varying Input Signal Frequency and Control Voltage.

the values of formant* frequency. The block diagram of Fig. 19 represents a scheme for the

* The short-time power spectrum of most speech sounds has several broad peaks along the frequency axis; these concentrations of sound power are sometimes called formants. The formant frequency indicates their approximate location on the frequency axis.

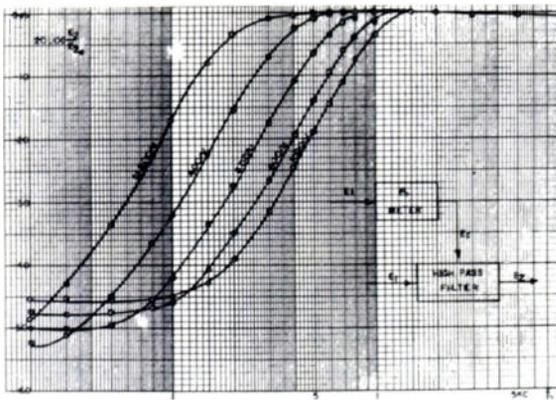
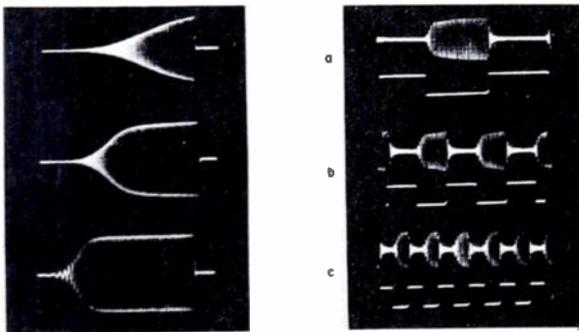


Fig. 16. Response Curves of Electronically Controlled HP Filter for Various Values of Input Frequency.



- A. (left)
 e_s = sinusoidal sweep
 0-1500 cps, 2v rms
 f_i = 1000 cps (a), 500 cps (b)
 200 cps (c)
- B. (right)
 f_s = 300 cps e_s = 2v rms
 e_s = -8.8v d.c. plus square wave 10v peak to peak
 f_c = 5 cps (a) 10 cps (b)
 20 cps (c)

Fig. 17. Dynamic Response Characteristics of Electronically Controllable High-Pass Filter.

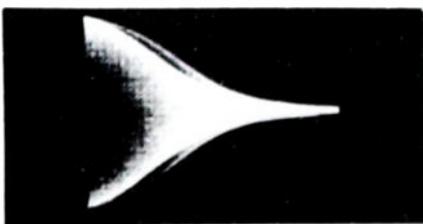


Fig. 18. Hysteresis Display, Sweep Rate 0.33 cps, Triangular Variation of e_c Between 3.6v and 16v.

separation of the formants. An 850 cps LP filter eliminates that part of the power spectrum which, as a rule, does not contain the first formant. The density of zero-crossings of the remaining signal is converted into a voltage signal which controls the cutoff frequencies of a low-pass and a high-pass filter. The output of the low-pass filter should contain essentially only the first formant and a voltage proportional to the first formant frequency should appear at F_1 . Similarly, the output of the high-pass filter should contain the second and possibly higher formants.

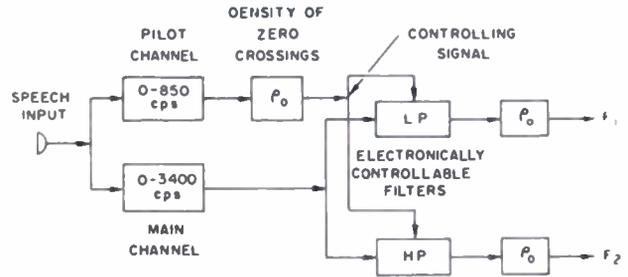


Fig. 19. A Scheme for Formant Tracking.

As an example of the performance of the electronically controllable low-pass filter, sound spectrograms of the sound sequence i-ε-a-o-u are shown in Fig. 20.

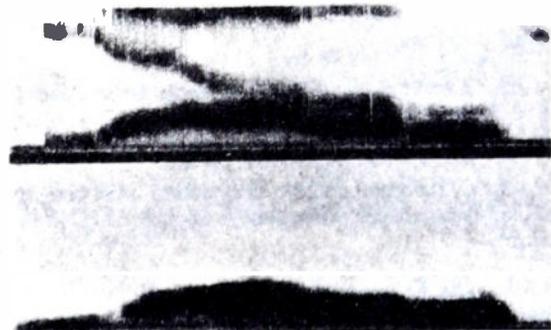


Fig. 20. Sonograms for i-ε-a-o-u. (A) wide band, no filter, (B) wide band, variable LP filter. Note: The Sona-graph was set on HS (high shaping).

In these spectrograms, the formants are represented as dark bars whose vertical position, corresponding to frequency, varies with time. Fig. 20A represents the sound before filtering, while Fig. 20B shows the spectrogram after the sound sequence has been filtered through the electronically controllable low-pass filter. It is seen that the separation of the first formant is accomplished reasonably well even if the second formant is very close to the first formant, as for example for o and u.

In Fig. 21 the behavior of the electronically controllable high-pass filter is indicated by

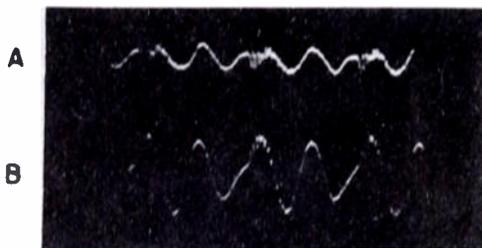


Fig. 21. Filtered (A) and Unfiltered (B) Samples of the Sound "i".

means of the oscillograms of the filtered and unfiltered sound "i". It is seen that the lower frequency components have been attenuated considerably, although for the given purpose more attenuation would be desirable.

Conclusions and Limitations

In view of the investigations which have been reported in this paper, and subject to the limitations listed below, the following conclusions can be made:

1. The performance of the constructed electronically controllable low-pass and high-pass filters have been found to be close to the expected behavior in the pass-band and in the cutoff region.
2. A suitable voltage source is needed for the control of the cutoff frequency of the low-pass filter.
3. A current source is needed for the control of the cutoff frequency of the high-pass filter.
4. The 5:1 range of the cutoff frequency is close to the maximum value which can be achieved with the present low-pass filter.
5. In the case of the high-pass filter, the range of the cutoff frequency is presently limited by the maximum current of the controlling pentode. However, if a larger current source were used, a deterioration of the slope in the

cutoff region is to be expected because of the lower Q of the inductors under such conditions.

6. In the stop-band the response is at least 35 db below the maximum response for the low-pass filter and about 45 db for the high-pass filter. A larger attenuation would be desirable, particularly in the low-pass case. The residual signal seems to be predominantly high-frequency noise, with a smaller hum component.

7. At present, the transistor-negative-impedance converter is the limiting factor for maximum signal size without distortion. However, the limitations imposed by the variable reactance elements are only slightly above the limitations of the transistor circuit.

8. Hysteresis causes the apparent inductance of the saturable inductors to vary by about 10 percent, for the same value of the control current.

Acknowledgement

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Abstract

An experimental investigation has been made of distortion in Class B amplifiers using 50 M watt junction transistors. Qualitative explanations are given for the results which show that distortion depends largely on 1) source impedance; 2) emitter bias; 3) signal current amplitude; and in grounded emitter stages, 4) transistor balance. Correct design will depend on the proper compromise between distortion, gain, and standby power drain.

Introduction

This article will deal only with distortion in low power drain Class B, grounded base and grounded emitter, junction transistor amplifiers. Grounded collector amplifiers will not be mentioned as they do not exhibit the type of distortion studied here, even though they are quite important. Space does not permit discussion of bias stability which may also be an important problem.

Feedback as a method of distortion reduction will not be considered since it too is a topic in itself. It may be said, however, that feedback is not the best cure for distortion here. As will be shown, distortion in the range of 25% can be reduced to less than 5% with proper circuit design. Such a reduction might be difficult to obtain with feedback due to the low power gain of the output stage, and the necessity of feeding back around not only the output transformer but also around the usual transformer at the input to the Class B stage. Both these transformers will normally be designed for small size, and will have correspondingly poor amplitude and phase characteristics. Thus, only small amounts of local feedback are likely to be found in these amplifiers.

Grounded Base, Class B (Fig. 6)

Two facts stand out: 1. All the results to be shown vary remarkably little from transistor to transistor, even for those of different manufacture. The various curves shown invariably exhibit the same form. 2. Distortion is almost entirely due to nonlinearity of the emitter circuit, due to the fact that emitter resistance r_e varies inversely with emitter current.

* This article is based in part on work done at the Microwave Research Institute under Signal Corps Engineering Labs Contract DA-36-039 sc-42525.

Figure 1 shows oscilloscope photos of typical Class B distortion. The output current waveform at the top of Figure 1 results from the emitter current vs emitter voltage characteristic shown below it, for a pair of push-pull transistors with no emitter bias. Each emitter is seen to effectively conduct less than half the time due to the very high emitter input resistance in the low current region.

This may be remedied by a small emitter voltage bias in the forward direction, or by attempting to feed from a high impedance source. With a current source there would be no distortion since the input current would be independent of the nonlinear input impedance.

Source Impedance

Figure 2 shows collector current waveforms with a 25Ω and a 1000Ω source impedance. It can be seen that there is considerably less distortion with the latter, although it still has a noticeable notch, indicating higher harmonic distortion.

Figure 3 shows the maximum 3rd harmonic distortion vs source impedance, with no emitter bias. The "maximum" refers to the fact that distortion varies also with signal level, and that the points of Figure 3 correspond to the peak of Figure 7.

Figure 3 shows that low distortion may be obtained by using a high source impedance. The trouble with using this method of reducing distortion is that a high source impedance gives low power gain. The low distortion that may be obtained this way shows that there is little or no distortion introduced by the fall off of alpha at very low currents.

Emitter Bias

Figure 4 shows scope photos of collector current waveforms with a little emitter bias, and without. Also shown are the two separate collector current waveforms, proving that even with bias each is still conducting essentially 50% of the time. It is surprising that such a large improvement may be made with only a small bias.

Figure 5 shows the dependence of third harmonic distortion on emitter bias. Bias is measured by measuring the DC collector current, I_C , which flows in the collector as a result of emitter bias. I_C is a handier parameter to use than

emitter voltage. It is more easily measured than V_e and more important since it is a measure of standby power drain. Also, for a good transistor I_c is practically independent of collector voltage and is roughly proportional to V_e .

The curve of figure 5 was taken at a constant (and fairly low) source impedance. The fact that this curve is of maximum distortion refers to dependence on signal level. The points correspond to the peaks of figure 9.

The top point on the curve is with no emitter bias, and 16 microamps DC flows in the collector. This curve shows clearly that as the emitter bias is increased the distortion decreases. With 100 μ A flowing in the collector, which gives very little standby power drain, distortion is fairly low. This corresponds roughly to a 0.1-volt emitter to base bias.

Figure 6 shows how the emitter bias is obtained from a divider across a 1.5-volt battery. The bias is developed across a low impedance, here 62 Ω , in an attempt to obtain a constant voltage bias. Since the full-wave rectified signal current flows through the 62 Ω , some bias in the wrong direction is developed which offsets the applied bias. The divider may be bypassed by a capacitor to give better AC gain, but since the DC component still flows through the 62 Ω the resistance level must be kept low.

In these tests no C was used, and the variable resistor R was on the order of 1 K Ω for 100 μ A DC flowing in the collectors. The corresponding emitter bias at room temperature is on the order of 0.1 volt. (The 62 Ω must be counted in the source impedance when unbypassed.)

Signal Current

Another important variable that distortion depends on is the size of the AC emitter current swing. Here collector current swing is measured because it is easier to measure than emitter current swing and also is equal to the emitter current swing for all practical purposes.

Figure 7 shows odd harmonic distortion as a function of the peak collector current swing in a Class B grounded base amplifier, with no emitter bias, and with a constant source impedance. It is seen that the 3rd harmonic predominates here, and may be very large.

The dependence of distortion on output current swing is at first surprising but is easily explained. Note that:

- 1) Distortion is very low for low AC current swings;
- 2) It reaches a large peak for slightly larger current swings; and
- 3) It drops off steadily for large current swings.

The explanation is this:

1) In the region of low AC current swing, the operation is in the region between dotted lines shown on the input characteristic in figure 8a. This portion of the curve is fairly linear, even though the input impedance is very high and the gain very low. The presence of a low gain region for small signals precludes this type of operation in most cases, but it can be seen here that the distortion should drop off to a low value when the input gets into this low current swing region.

2) For slightly larger signals, the current swing is half in one region and half in the other, and distortion is at its worst. This corresponds to the peak in distortion seen at some fairly small value of current swing.

3) As the signal becomes larger the proportion of the cycle that the signal spends in this high input impedance region becomes less and less, so on a per cent basis the distortion falls off. Fourier analysis based on the broken line characteristic of figure 8b predicts that 3rd harmonic distortion will fall off approximately proportionately to the size of the swing. It is seen that this happens.

Eventually at high current swings distortion will begin to rise due to fall off of alpha. However, in good 50 MW transistors, and grounded base operation, this does not begin to occur until about 25 to 30 ma collector swing, which is about the limit of collector dissipation anyway. For large power transistors this may not hold true, but most good 50 MW transistors will show only a 1% or so rise in distortion at these high collector swings. Thus, grounded base amplifiers have been operated at around 30 ma peak swing, 100 microamps or less DC collector current due to emitter bias, and 1 to 2% distortion.

Figure 9 shows a family of curves of 3rd harmonic distortion vs AC collector current swing. The DC collector current is taken as the parameter here.

The top curve, with 13 microamps flowing in the collector is for no emitter bias. As the emitter bias is increased and more DC collector current flows, the distortion peaks become smaller and smaller, and flatten out until with 95 microamps DC flowing the distortion is at a fairly low level.

The remarkable thing here is that such a small amount of bias gives such a good reduction in distortion. Note that even for the lowest distortion shown, the standby power drain will be negligible since less than 100 microamps flow in each collector.

Figure 10 shows the same amplifier with emitter bias such that 95 microamps DC flows in the collectors, and here the scale is blown up to

show how the higher order harmonics come in. It is seen that these also are greatly reduced so the third harmonic still predominates, although at larger current swings the higher order harmonics may become appreciable. In general, it is seen here that distortion is quite small, having been reduced from about 20% to about 3% for an increase of DC collector current drain of only 80 microamps or so.

Load Impedance

As might be expected, since there is almost no distortion in the grounded base collector characteristics, distortion is practically independent of load, as Figure 11 shows. Here the AC collector current swing was kept constant as the load was varied. It is seen that over a wide range of loads, distortion remained constant, both with and without emitter bias. Note that with constant AC current swing, power output here varies directly with load.

This and the preceding figures point up the fact that, for the type of operation described here, distortion is practically independent of power output, and that it depends instead on output current swing.

The even harmonics depend almost entirely on the unbalance in the transistors, and are pretty much independent of current swing. The 2nd harmonic was about 1% and the k th about 0.1% in this case, both with and without emitter bias.

For the grounded base connection, both the even and odd harmonic distortion is almost completely independent of frequency for the audio frequency range.

One point that should be made is that all the grounded base distortion behavior presented here is remarkably constant from transistor to transistor, even for transistors from different manufacturers. The magnitude of the distortion may change somewhat, but the shape of the curves and the dependence on bias and swing does not change. Thus, the general behavior of grounded base Class B circuits is quite predictable as far as distortion goes.

Grounded Emitter, Class B (Fig. 12)

Several facts stand out here: 1. There is an appreciable distortion of the grounded emitter collector characteristics, which varies from transistor to transistor so that distortion is not as predictable as for grounded base. This also means that distortion depends somewhat on load, and usually increases with power output. 2. There may be considerable even order harmonic distortion due to unbalance in both phase and amplitude of the current gain. 3. Odd harmonic distortion acts pretty much like that in grounded base, since it too is largely a result of nonlinearity in the base-emitter circuit. Thus, 3rd harmonic distortion decreases with increased source impedance.

As in grounded base it also decreases with increased emitter bias, and with increased current swing. Somewhat larger biases may be necessary, however, to get as low a 3rd harmonic as with grounded base.

Source Impedance

Besides affecting 3rd harmonic distortion, source impedance affects even harmonic distortion, due to the unbalance in alpha between the two transistors.

Low Impedance Source: If fed from a zero impedance voltage source, the voltage gain for good high alpha transistors is approximately

$$G \approx \frac{\alpha R_L}{r_e}$$

where R_L is the load, r_e the emitter resistance. Fourier series analysis shows that the per cent 2nd harmonic distortion is then approximately

$$\% D \approx 20(\alpha_1 - \alpha_2)$$

where α_1 and α_2 are the two current gains. This distortion will be small, and can be balanced out by putting an external resistor R_E in one emitter, where the size of R_E necessary for balance is

$$R_E \approx (\alpha_1 - \alpha_2)r_e \text{ approximately.}$$

This will usually amount to only a few ohms, and is put in the emitter of the transistor with the high alpha.

The circuit is shown in Figure 12, and the emitter bias is obtained exactly in the same way as for grounded base.

High Impedance Source: If the amplifier is fed from a current source instead, then unbalance in α shows as a much larger unbalance in the current gain,

$$\beta = \frac{\alpha}{1 - \alpha}$$

If $\Delta = \beta_1 - \beta_2$ is defined as the unbalance in the betas, then the per cent 2nd harmonic is

$$\% D \approx \frac{20\Delta}{\beta}$$

where β is the average current gain. This distortion may easily run to 10% or more, since a spread of two to one in beta is common for most manufacturers.

To balance this out an external resistance R_E large enough to change the current gain must be used, so that the current gain A_1 on the high beta side is reduced from $A_1 = \beta_1$ to

$$A_1 = \frac{\alpha_1}{1 - \alpha_1 + \frac{R_E}{r_c}}$$

For balance $A_1 = \beta_2$ which gives

$$R_E = \frac{r_m A}{\beta^2}$$

which easily may run as high as 10K.

Thus, as the source impedance increases, 2nd harmonic distortion increases from perhaps 1 to 15% while 3rd harmonic distortion decreases. Clearly there is some optimum source impedance, depending on the unbalance. Also, if it is desired to balance out the 2nd harmonic with an external emitter resistance, this resistance may have to be increased from a few ohms at low source impedance to 10K or more at high source impedance. This latter value may be too high to be practical, making balance impossible for a constant current source.

Figure 13 shows how both third and fifth harmonic distortion fall off with increase in source impedance, while second and fourth both rise. The total harmonic distortion curve shows a broad optimum from 500 to 700 ohms source impedance. This amplifier has emitter bias, with about 500 microamps DC flowing in each collector. The second harmonic was balanced to a minimum at low source impedance by an $R_E = 7\Omega$. This value was then left in the emitter as the source impedance was increased.

Phase Shift

It is very easy to get distortion in grounded emitter due to unequal phase shifts of the two halves of the waveforms. This in turn is due to the two transistors having unequal alpha cutoff frequencies. This effect is exaggerated in grounded emitter since the current cutoff and its attendant phase shift is lowered in frequency by the magnitude of the current gain.

Fourier series analysis shows that for a small difference in phase shifts, the per cent 2nd harmonic distortion is approximately $\% D = 20\theta$, where θ is the difference in phase shifts in radians or

$$\% D = 20f\beta \left(\frac{1}{f'_{\alpha O}} - \frac{1}{f''_{\alpha O}} \right)$$

where the $f_{\alpha O}$ are the two alpha cutoff frequencies. Distortion thus increases linearly with operating frequency f . For the GE 2N43, with a production spread on cutoffs of 0.5 to 2.5 MC and $\beta = 50$ a second harmonic distortion of 4.8% at 3KC is possible. The linear behavior of this distortion with frequency can easily be observed.

This type of 2nd harmonic distortion may often be excessive in complementary symmetry circuits, where the alpha cutoffs of the NPN and PNP may be quite dissimilar.

Signal Current

The variation of distortion with signal current is similar to grounded base. Figure 14 shows odd harmonic distortion in a grounded emitter Class B amplifier. Here, for no emitter bias, it is seen that the third harmonic is quite enormous. In Figure 15, with emitter bias, it is seen that distortion is greatly reduced. In this case 500 microamps DC flowed in the collectors which is somewhat larger than that necessary for grounded base. Distortion will usually increase again at currents larger than shown in Figure 15, due to distortion in the collector characteristics. Thus, the grounded emitter circuit will almost always exhibit more distortion than grounded base for the same output power and standby drain. On the other hand, the grounded emitter has extra power gain which when used in local feedback may give it the better over-all performance.

In conclusion, Class B offers many advantages to the circuit designer. Both high efficiencies and low distortion can be obtained, and with very little standby power drain. 70 to 75% efficiencies, standby collector currents of only 100 μ A or so, and about a quarter watt out of a pair of 50 milli-watt transistors make these circuits worthwhile.

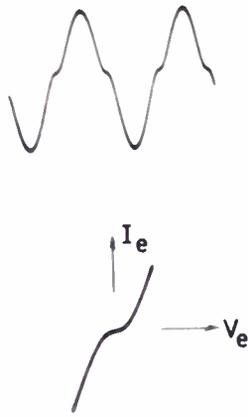


Fig. 1 Top: Output current waveform.
Bottom: Input emitter current vs emitter voltage characteristic.

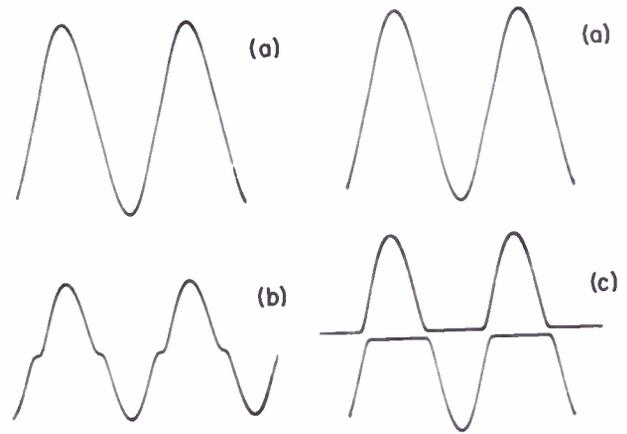


Fig. 4 G. B. Output current waveforms
(a) With a small emitter bias
(b) With no emitter bias
(c) Current in each collector, with bias as in (a).



Fig. 2 Collector current waveforms
Top: 25 Ohm source impedance
Bottom: 1000 Ohm source impedance.

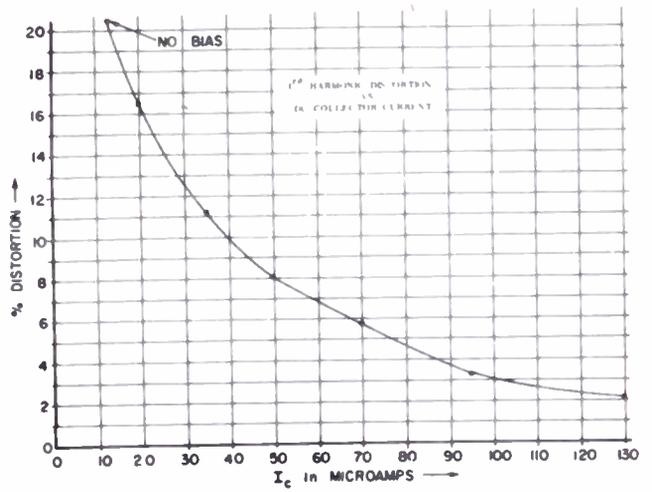


Fig. 5 G. B. Maximum third harmonic distortion vs emitter bias (as indicated by DC collector current). The top point is with no bias.

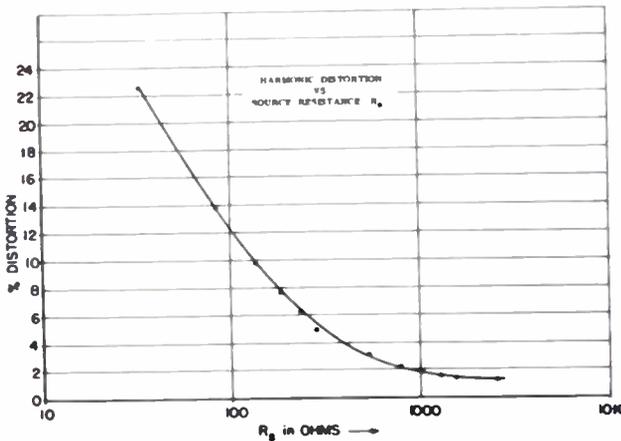


Fig. 3 G. B. Maximum third harmonic distortion vs source impedance. No emitter bias.

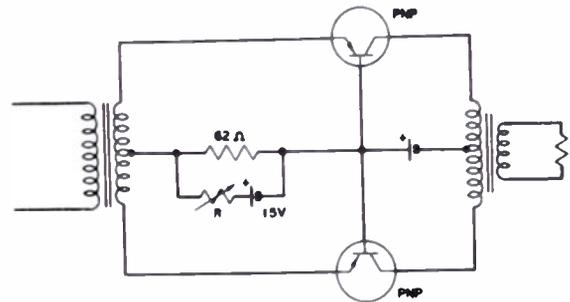


Fig. 6 Class B grounded base circuit.

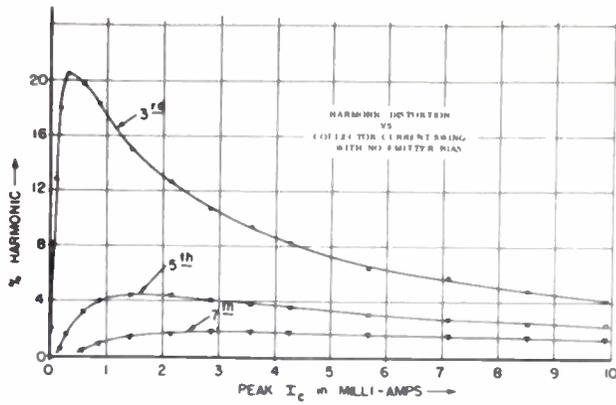


Fig. 7 G. B. 3rd, 5th, and 7th harmonic distortion vs peak collector current swing. No emitter bias.

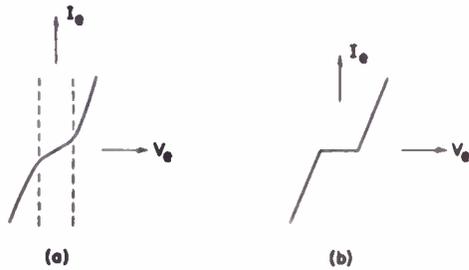


Fig. 8 (a) Input characteristic: emitter current vs emitter voltage, with no emitter bias.
(b) Idealized characteristic used to predict distortion for large signals.

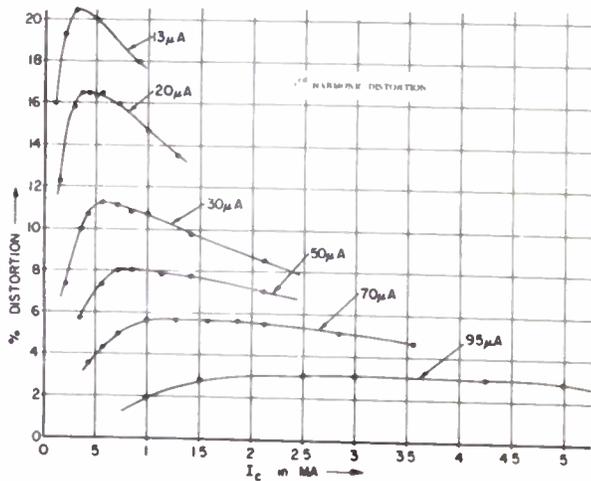


Fig. 9 G. B. Third harmonic distortion vs peak collector current swing, at different values of emitter bias (indicated by DC collector current in μ A).

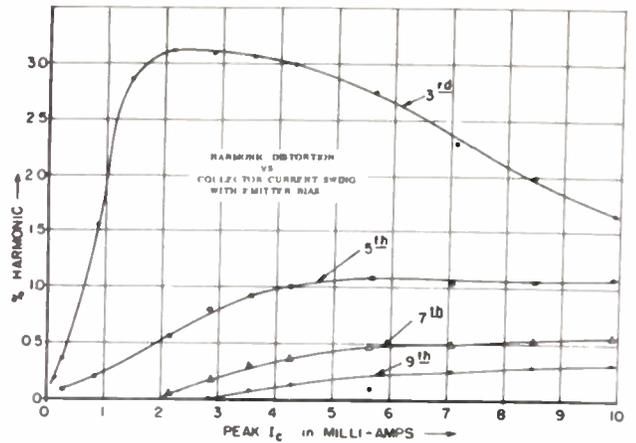


Fig. 10 G. B. Odd harmonic distortion vs peak collector current swing, with 95 μ A DC flowing in each collector as a result of emitter bias.

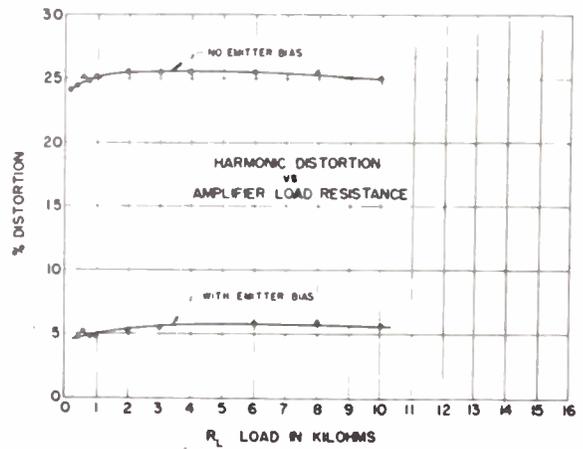


Fig. 11 G. B. Harmonic distortion vs amplifier load resistance.

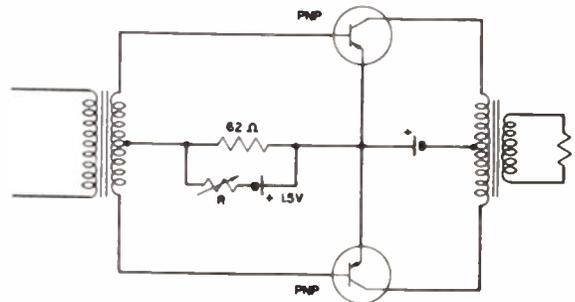


Fig. 12 Class B grounded emitter circuit.

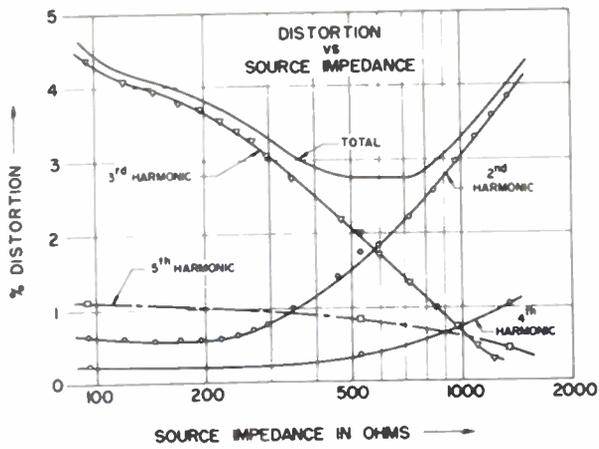


Fig. 13 G. E. Distortion vs source impedance, with emitter bias.

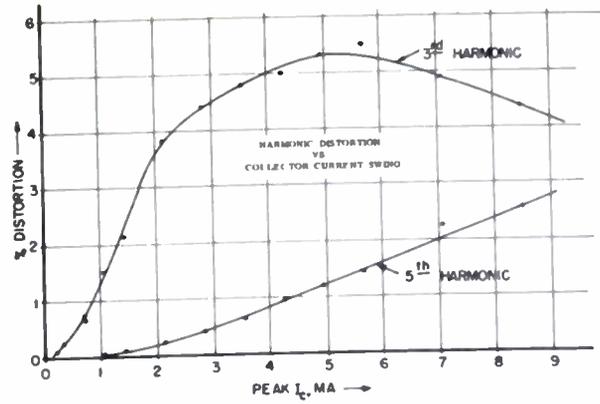


Fig. 15 G. E. Odd harmonic distortion vs peak collector current swing. With emitter bias (500 μ A DC flowing in each collector).

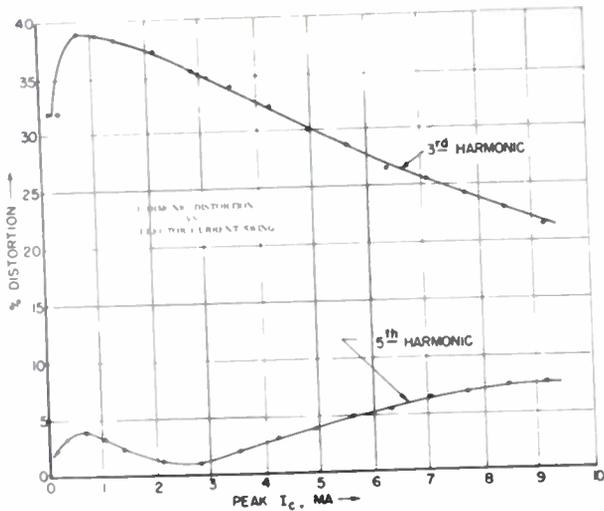


Fig. 14 G. E. Odd harmonic distortion vs peak collector current swing. No emitter bias.

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Abstract

This paper deals with the design and description of a "V" filter which is to be used to distinguish between power spectra having the same central frequency and average power but different dispersions. The power spectra that are discussed have a Gaussian distribution curve when plotted on a logarithmic frequency scale. The filter is adaptable for use in the analysis of speech and its use in this analysis will be discussed. The design of the filter including graphical means of establishing the filter characteristics from the dispersion of the given power spectra is included.

Introduction

The work carried on at Northeastern University on the analysis of speech has as its goal the coding of speech into signals which may be transmitted over a channel of considerably narrower bandwidth than now required for normal voice communication. In order to apply appropriate techniques for the extraction of these parameters it is desirable initially to separate speech into major classes such as turbulent and non-turbulent sounds. A device for performing the initial separation is described on pp. 14, 18-20 of reference 1. The present paper is concerned with the analysis of the output from this device which contains those sounds referred to as being turbulent in nature.

These turbulent speech sounds have similar properties to noise signals passed through filters, as seen by the typical power spectra for *s* (as in *see*) and *ʃ* (as in *ship*) shown in Fig. 1. Experimental results have shown that knowledge of the center frequency of these power spectra is not always sufficient information to distinguish one sound from the other. The additional parameters which synthesis of these turbulent sounds has indicated as being necessary are the dispersion and the envelope of the original sound.

The work described in this paper has been initiated to determine means of obtaining a measure of the dispersion of noise signals of this type. This measure of dispersion may be used to classify a turbulent speech sound into one of two arbitrary groups, which is a discrete type of analysis, where the mean value of the dispersion of each group is known beforehand. Also, a continuous type of analysis has been suggested where the output gives a continuous rather than discrete indication of dispersion.

Theory

The development of the network, as discussed below, assumes that the center frequency is known. This is justified since this value can be obtained by using an average frequency meter. For the purpose of the development which follows it is assumed that the input spectrum is a Gaussian curve when plotted against the logarithm of frequency, as shown in Fig. 2.

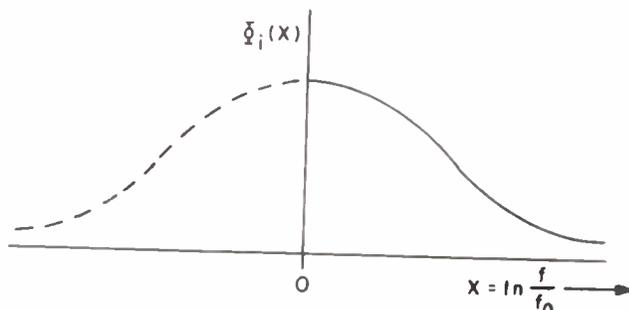


Fig. 2. Assumed Input Power Spectrum.

The equation of this curve is:

$$Q_1(x) = \frac{1}{\sigma\sqrt{2\pi}} \exp\left(-\frac{x^2}{2\sigma^2}\right)$$

where the logarithmic frequency *x* has been normalized with respect to the central frequency *f*₀. The ordinate *Q*₁(*x*) is similar to the normalized power spectral density, in that areas under this curve are assumed proportional to power in corresponding bands and, as indicated by the equation, the total area is normalized to unity. The signals are assumed to have the same *Q*₁(*x*) except for different values, *σ*₁ and *σ*₂, of the parameter *σ*.

The choice of the network characteristic to accomplish the analysis was based upon the availability of methods for synthesizing this type of network as well as the simplifications which are introduced into the development by this choice. The network characteristic of the "V" filter is shown in Fig. 3. Using the notation *ln* for natural logarithm its equation is:

$$\ln |H| = \begin{cases} \ln a + kx & (0 \leq x \leq x_c) \\ 0 & (x_c \leq x < \infty) \end{cases}$$

or

$$|H| = \begin{cases} a e^{kx} & (0 \leq x \leq x_c) \\ 1 & (x_c \leq x < \infty) \end{cases}$$

*This work was performed under Contract No. AF19(604)-1039 with the Air Force Cambridge Research Center of the Air Research and Development Command.

where

$$a = \left(\frac{E_2}{E_1} \right) < 1$$

$$f_0$$

and

$$x_c = -\frac{\ln a}{k}$$

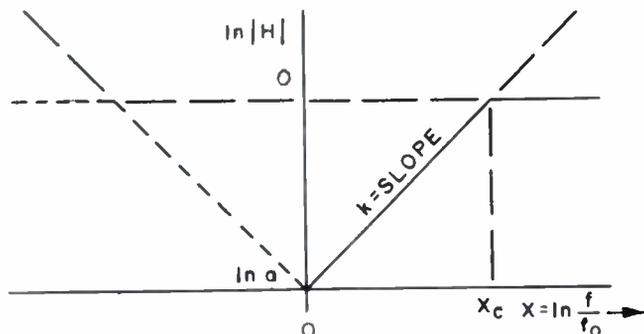


Fig. 3. Network Characteristic of the "V" Filter.

Since the input power spectrum and the characteristic of the "V" filter are symmetrical about the origin the output power is:

$$P_o = 2 \int_0^{\infty} \tilde{Q}_o(x) dx$$

where

$$\tilde{Q}_o(x) = |H|^2 \tilde{Q}_i(x)$$

or

$$\tilde{Q}_o(x) = \begin{cases} \frac{a^2}{\sigma\sqrt{2\pi}} \exp(2kx - \frac{x^2}{2\sigma^2}) & (0 \leq x \leq x_c) \\ \frac{1}{\sigma\sqrt{2\pi}} \exp(-\frac{x^2}{2\sigma^2}) & (x_c \leq x < \infty) \end{cases}$$

therefore

$$P_o = 2 \left[\int_0^{x_c} \frac{a^2}{\sigma\sqrt{2\pi}} \exp(2kx - \frac{x^2}{2\sigma^2}) dx + \int_{x_c}^{\infty} \frac{1}{\sigma\sqrt{2\pi}} \exp(-\frac{x^2}{2\sigma^2}) dx \right]$$

which can be written in terms of the probability integral

$$P_o = \frac{2}{\sqrt{\pi}} \left[a^2 \exp(2k^2\sigma^2) \int_{-\frac{\ln a}{\sqrt{2}k\sigma}}^{-\sqrt{2}k\sigma} \exp(-u^2) du + \int_{-\frac{\ln a}{\sqrt{2}k\sigma}}^{\infty} \exp(-v^2) dv \right]$$

The "V" filter can be used for the two different types of analysis mentioned previously. First of all it can be used to separate two power spectra of known dispersions and equal power by a proper choice of the slope k of the "V" filter.

It can be also used to determine the dispersion of an unknown power spectrum by the use of a reference spectrum or by normalization with respect to the input power.

The separation of two power spectra requires that a means of determining the optimum k be developed. If a is assumed fixed, the problem now is to select k so as to maximize the ratio of output power values for the two signals; namely, $P_o(k, \sigma_2)/P_o(k, \sigma_1)$. Since the expression for P_o cannot be developed into a closed form a graphical procedure was used to solve the problem.

The final expression shown above is a function of $k\sigma$. A numerically computed curve for this function, where \log is used to indicate logarithms to the base ten, is shown as Fig. 4 where $\log P_o$ is plotted versus $\log k\sigma$. For this curve the value of a was taken as 60 db attenuation at the central frequency. Specifying the values of σ_1 and σ_2 now determines the interval length $\log k\sigma_2 - \log k\sigma_1$ along the horizontal axis between the two points the difference of whose ordinates is to be maximized. This interval must evidently be located where the slope of the curve in Fig. 4 is largest which is in the region containing its point of inflection. More specifically, if the derivative of that curve is plotted graphically, as in Fig. 5, and seen to have a single maximum, the interval of specified length on the $\log k\sigma$ axis is located so that at its endpoints the ordinates of the derivative curve are equal. This is easily done by moving a straight edge of the required length parallel to the $\log k\sigma$ axis of Fig. 5 until its endpoints intersect the curve. The values of $k\sigma_1$, and $k\sigma_2$ so obtained determine the optimum k for the given σ_1 , σ_2 and a .

In a particular example considered with $\sigma_1 = 0.479$ and $\sigma_2 = 0.773$ the optimum value from Fig. 5 is $k = 3.06$. The ratio of the two outputs in this case is calculated to be 11.2 db.

The use of the "V" filter to determine the dispersion of an unknown spectrum is accomplished by taking a spectrum whose dispersion is known and plotting a curve of the ratio of power output versus dispersion with the known spectrum as a reference by using the expression derived in this section. The input powers of the unknown and reference signal are then made equal and the ratio of output powers can be used with this curve to determine the dispersion of the unknown signal.

Design of "V" Filter

In designing the "V" filter to check the computed results it was decided to design a stage for $k = 1$ and to cascade several of these so that it would be possible to analyze the variation of output power as a function of k . Using the method described by Baum² the transfer ratio for the filter was found to be:

$$\frac{E_2}{E_1} = K \frac{s^2 + \sqrt{2} s + 1}{s^2 + 10.1 s + 1}$$

This transfer ratio was synthesized in the form of a nonsymmetrical bridged "T" network which in its normalized form is shown in Fig. 6.

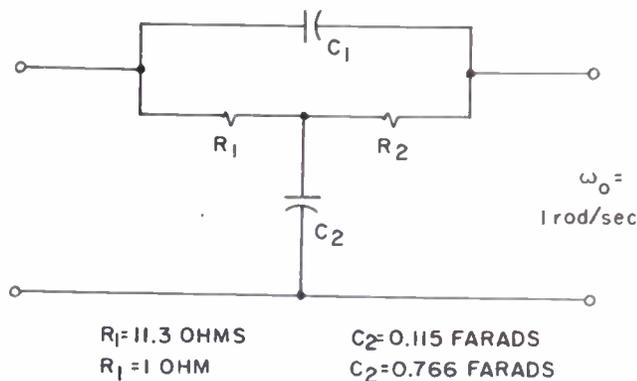


Fig. 6. Bridged "T" Form of "V" Filter for $k = 1$.

The final form of a single stage of the "v" filter is shown in Fig. 7. A cathode follower is used to drive the network and a stage of amplification is included. The filter network was scaled by increasing the impedance level to 5000 ohms and the central frequency to 5000 cps. Six stages were cascaded in the final form of the filter that was constructed.

Measurement Techniques

Measurements that were made to verify the theoretical conclusions made use of the equipment arrangements as shown in Figs. 8 and 9. The noise-shaping circuit shown in these block diagrams consisted of a 6AU6 pentode with an RLC parallel tuned circuit as its plate load. Varying the parameters of this tuned circuit produced the four spectra that were used in these tests. The General Radio 1390A Random-Noise Generator was used with the range switch set at 20 kc throughout these tests.

The arrangement of equipment shown in Fig. 8 was used to determine the power spectrum for the artificially generated signal from which the standard deviation was then computed. A sample of one of these power spectra is shown in Fig. 10. It should be noticed that the scale of the power axis is only proportional to power. No effort was made to determine absolute levels since the standard deviation would not be altered if the scale were simply multiplied by a constant. The Gertsch Driving Circuit indicated in Fig. 8 is a cathode follower using a 6AQ5 and is described on page 71 of reference 3.

Figure 9 shows the equipment arrangement with the "v" filter being used. The vacuum-tube voltmeters were used to indicate the rms values of the corresponding voltages.

In determining the values of k and a for the "v" filter the sine-wave response of the filter was plotted and the values determined graphically. Figure 11 shows the response of six stages. The

indicated computed points were determined by graphically adding the three Butterworth characteristics which were combined in forming the filter transfer function. Figure 12 is a chart which shows the values of a and k for each of the six stages of the "v" filter.

Stages	k	a (db)
1	0.85	17.0
2	1.50	26.3
3	2.42	38.5
4	3.00	50.5
5	3.67	55.3
6	3.91	65.2

Fig. 12. Values of k and a for "v" Filter.

Computations

The final expression for output power as developed in the Theory section was used in computing expected ratios of output power when the input powers were made equal. Since the data was in the form of rms voltages the square roots of the computed power ratios were calculated and used for comparison. The values of a , k and σ were the values actually measured. Care should be taken to convert the standard deviation to conform with the abscissa in terms of log to the base e as assumed in the derivation.

Results

The results of the measurements and computations are given in chart form in Fig. 13. The values corresponding to four stages of "v" filter were plotted and are shown as Fig. 14.

In the section entitled Theory an example was given for determining the optimum value of k for separation of two power spectra. A value for k of 3.06 was found with an expected ratio of 11.52 db separation. Actual measurements showed a 10.1 db separation.

Discussion

If it is desired to make use of the "v" filter to determine the dispersion of an unknown power spectrum by reference to a known spectrum a plot such as shown in Fig. 14 can be made and from this plot the value of the dispersion may be obtained from the knowledge of the ratio of output voltages.

As was stated in the initial development, the center frequency of the noise spectra are assumed to have the same value as that for the "v" filter. Since this might not be the case in actual practice it would be necessary for the center frequency of the filter to be automatically adjusted to coincide with that of the spectrum being analyzed. This would require the use of electronically controlled filters. The paper presented by Mr. Dolansky offers a description of the necessary techniques which would have to be followed. Preliminary work has been completed on necessary

alterations in the filter design but no working model has been assembled.

Acknowledgements

The authors wish to express their thanks to Professors S. H. Chang and H. L. Stubbs of the Northeastern University faculty for their continued encouragement and constructive criticism during the course of the work on this paper.

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3. Interim Technical Report No. 1, Contract No. AF19(604)-1039, Item I.

σ_ϵ	<u>Stages</u>											
	1		2		3		4		5		6	
	Meas.	Comp.	Meas.	Comp.	Meas.	Comp.	Meas.	Comp.	Meas.	Comp.	Meas.	Comp.
0.479	1.00	1.00	1.00	1.00	1.00	1.00	1.00	1.00	1.00	1.00	1.00	1.00
0.611	1.13	1.15	1.27	1.39	1.40	1.90	1.40	2.44	1.45	2.48	1.35	3.05
0.773	1.40	1.38	1.96	2.06	2.70	3.34	3.20	3.64	3.40	4.81	3.30	6.86
1.026	1.90	1.83	2.40	3.15	5.80	5.52	7.90	7.65	9.50	7.05	9.20	11.45

Fig. 13. Comparison of Computed and Measured Output Voltage Ratios as Functions of σ and the Number of Stages.

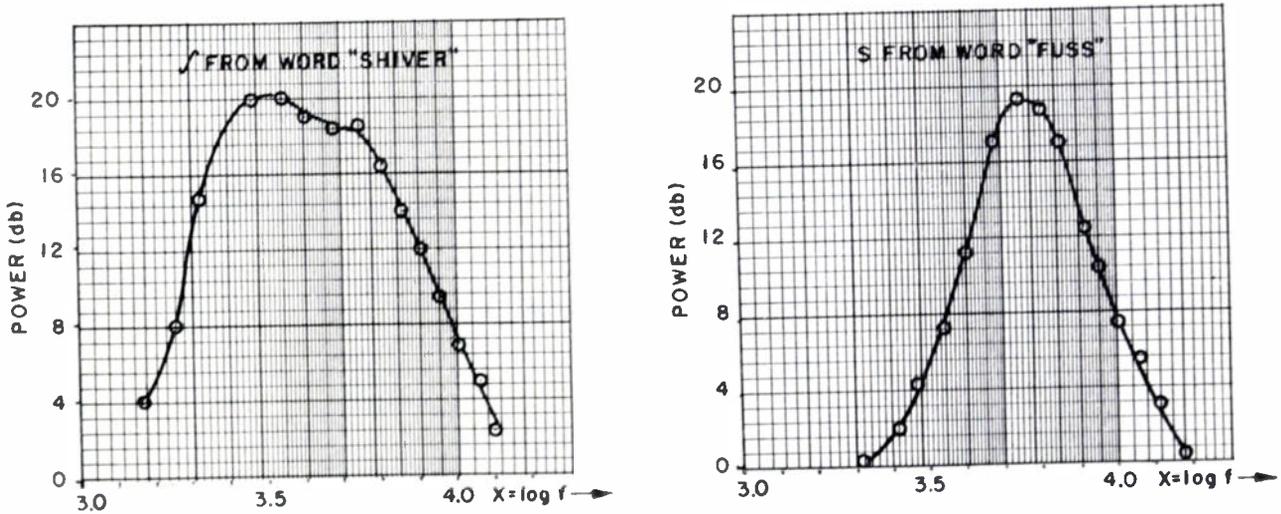


Fig. 1
Power Spectra for Turbulent Sounds

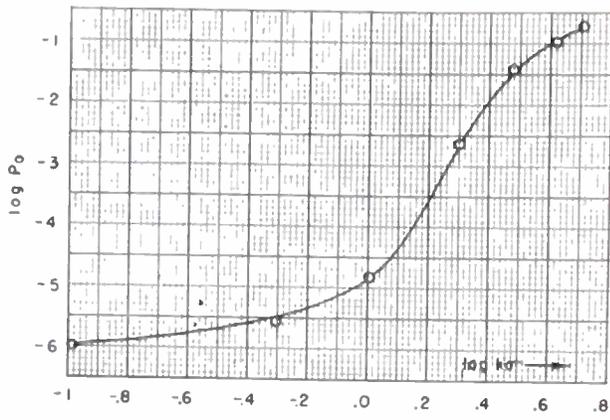


Fig. 4
Plot of $\log P_0$ vs. $k\sigma$.

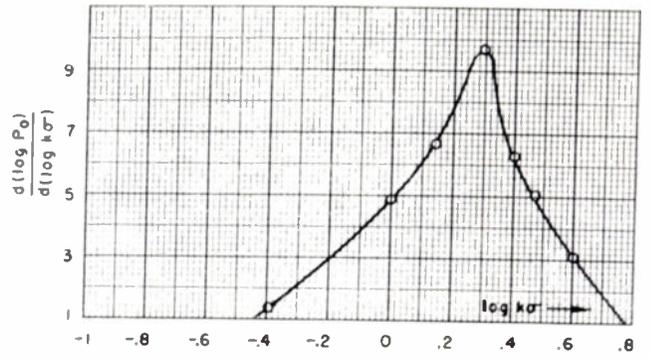


Fig. 5
Plot of Derivation of $\log P_0$ with Respect to $\log k\sigma$.

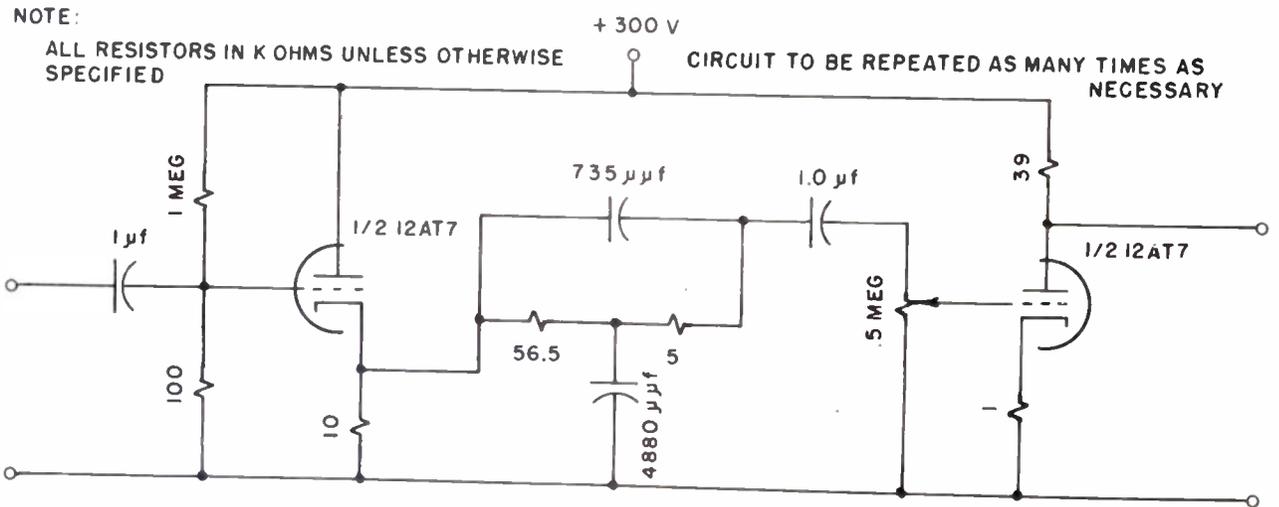


Fig. 7
Schematic Diagram of a One Stage "V" Filter.



Fig. 8
Apparatus for Measuring Power Spectrum of Artificially Generated Signals.

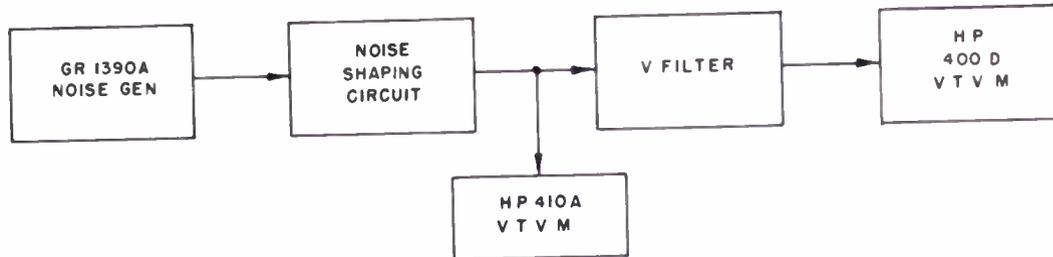


Fig. 9
Equipment Arrangement for Checking "V" Filter.

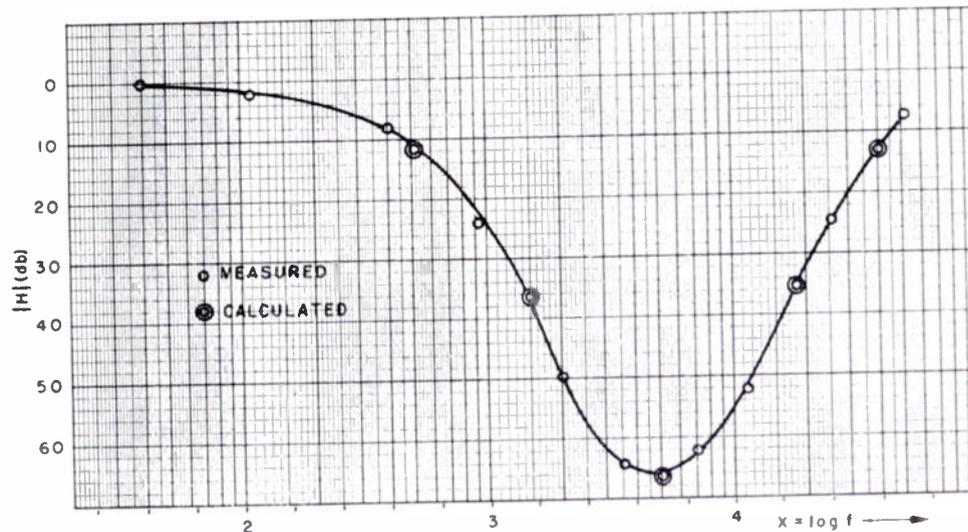


Fig. 11
Sine Wave Response of a Six Stage "V" Filter.

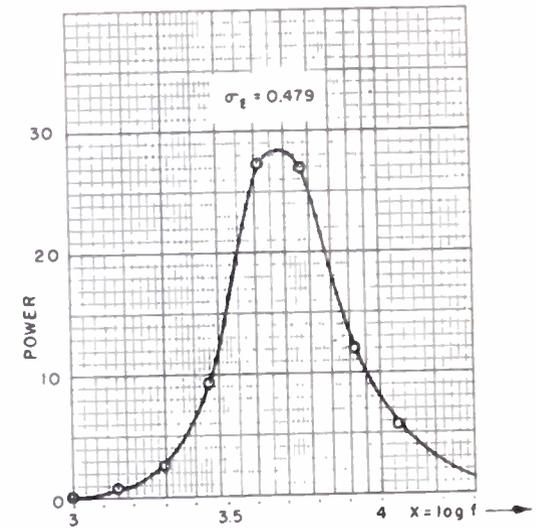


Fig. 10
Artificially Generated Power Spectrum.

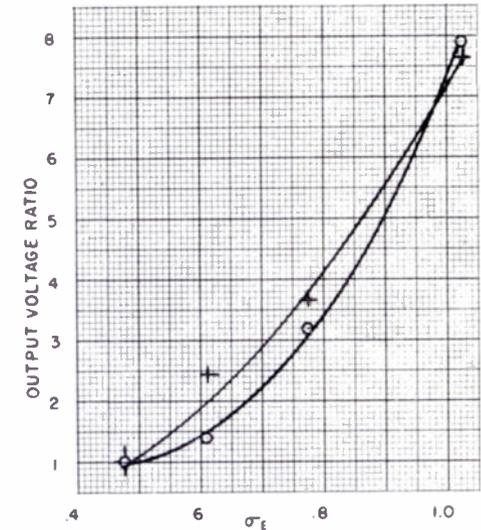


Fig. 14
Computed and Measured Response of a Four Stage "V" Filter to Artificially Generated Power Spectra.

CALIBRATION OF TEST RECORDS BY B-LINE PATTERNS

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ABSTRACT

Test records are often calibrated by measuring the width of the optical pattern formed when a sharp beam of light is reflected from the modulated grooves. An error has been found in this measurement owing to diffraction of light at the edges of the pattern, which results in a fuzzy ending and a general enlargement of the pattern width, especially at high frequency. A new optical method has been devised for calibration of test records by the use of interference-line patterns. Two sets of interference lines have been identified: (a) Uni-

formly-spaced lines which are related to the recorded frequency, the angular velocity of the record, and the wavelength of light—these have been called the A-lines; (b) lines with variable spacing which are related to the amplitude of modulation and to the wavelength of light—these have been called the B-lines. B-line patterns may be readily related to the theoretical width of the optical pattern free of diffraction error, with the resultant improvement in the accuracy of test-record calibration.

DESIGN AND PERFORMANCE OF A HIGH FREQUENCY ELECTROSTATIC SPEAKER

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ABSTRACT

An electrostatic speaker has been developed which provides a quality of high frequency reproduction not available with electromagnetic tweeters. The diaphragm consists of a thin plastic film bearing an evaporated metallic layer. The membrane is stretched around a semi-cylindrical perforated electrode on which ridges are embossed to provide clearance. The response varies less than ± 2 db in the frequency range between 8 and 16 kc. The azimuthal distribution pattern is excellent, owing

to the cylindrical geometry, and is essentially independent of frequency in the same range. The second harmonic distortion inherent in this type of speaker is maintained at a low value. An indication of the quality of high frequency reproduction is provided by oscillograms of the response to tone burst signals. The speaker is in quantity production and has been incorporated in several models of home reproduction instruments.

ELECTRONIC MUSIC SYNTHESIZER

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ABSTRACT

The electronic music synthesizer is a machine that produces music from a coded record. The coded record is produced by a musician, musical engineer or composer with a fundamental understanding of the composition of sound. The electronic music synthesizer provides means for the production of a tone with any frequency, intensity, growth, duration, decay, portamento, timbre, vibrato, and variation. If these properties of a tone are specified, the tone can be completely

described. The advantage of the electronic music synthesizer is that it can produce new and radical tone complexes for musical satisfaction and gratification. The new system does not displace the artist and musician of today. It does not take the place of talent combined with work. The electronic music synthesizer provides the musician, musical engineer and composer with a new musical tool with no inherent physical limitations.

PROPOSED CONTROLS FOR ELECTRONIC MASKING IN COLOR TELEVISION

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Summary

Electronic masking in the color television studio is accomplished by the use of crosscoupling networks. These networks yield three output signals, each of which is a linear function of the three input signals. The colors of the final reproduction depend upon the values of the coefficients in the linear equations relating output voltages to input voltages. A change in the value of any individual coefficient in these equations affects the hue, saturation, and brightness of most nonneutral colors. The effect of an incremental change is described for a given set of masking equations. Coefficient controls may be ganged to simplify the relationship between manual adjustments and visual effects. A block diagram is shown for a suitable circuit. A means for ganging the controls is also described.

Introduction

In earlier papers ^{1,2}, we have demonstrated that the color space reproduced on film does not coincide with the color space which can be reproduced on a color television picture tube. Because of the greater luminance output range of color film, scenes of high contrast can be better reproduced on color television if the luminance range is electrically compressed in the color television studio. One method of accomplishing this result employs gamma overcorrection and electronic masking.

Electronic masking has been suggested as a means of improving hues, saturations, and brightnesses of chromatic colors. Circuits^{3,4,5} for accomplishing electronic masking are known. The present paper is a further discussion of the application of electronic masking to color television. Attention is given to the problem of correlating manual masking adjustments with desired visual effects.

Masking Equation Coefficients

In an electronic masking unit, each output signal (red, green, or blue) is a function of all three of the red, green, and blue input signals. These functional relationships are essentially linear and can therefore be conveniently represented by the equations:

$$\begin{aligned}R_o &= a_{11}R_i + a_{12}G_i + a_{13}B_i \\G_o &= a_{21}R_i + a_{22}G_i + a_{23}B_i \\B_o &= a_{31}R_i + a_{32}G_i + a_{33}B_i\end{aligned}\tag{1}$$

The red output signal, R_o , is derived from the red, green, and blue input signals, R_i , G_i , and B_i , in the relative amounts of a_{11} , a_{12} , and a_{13} . The diagonal elements a_{11} , a_{22} , and a_{33} will be large and positive whereas the remaining six constants will usually be small and negative. It is seen that there are nine independent constants whereby the output signals can be adjusted as functions of the input signals.

Adjustments to improve picture quality must be made through the proper selection of values of these nine constants. If the neutral scale is to remain in balance, the sums of the three coefficients in each of the equations must be the same. It is convenient to set these sums equal to unity:

$$\begin{aligned}a_{11} + a_{12} + a_{13} &= 1 \\a_{21} + a_{22} + a_{23} &= 1 \\a_{31} + a_{32} + a_{33} &= 1\end{aligned}\tag{2}$$

Applying these equations to eliminate a_{11} , a_{22} , and a_{33} , the masking equations become:

$$\begin{aligned}R_o &= (1-a_{12}-a_{13})R_i + a_{12}G_i + a_{13}B_i \\G_o &= a_{21}R_i + (1-a_{21}-a_{23})G_i + a_{23}B_i \\B_o &= a_{31}R_i + a_{32}G_i + (1-a_{31}-a_{32})B_i\end{aligned}\tag{3}$$

Thus the requirement of proper gray scale balance is seen to reduce the independent parameters from nine to six. The masking unit shown in Fig. 1 maintains gray scale balance and has these six independent controls. Any desired masking improvement in the colors on the color picture tube must be obtained through the selection of proper values of these six constants.

There are thousands of colors which may be differentiated on a color television receiver. With no more than six independent parameters on the masking equipment, it is evident that the colors viewed on the screen cannot all be independently modified. A complete array of color changes on the picture tube will be associated with changes in the parameters, either singly or in combination. The question is, in what manner should these parameters be controlled to produce the most desirable pattern of color changes.

One possible arrangement would be to have each of the six parameters, the a 's in the equations, controlled by a separate control knob. Let us consider the pattern of color changes involved in the change in one of these parameters. Consider a_{12} , for example: Suppose that we make it more negative. The effect on numerous colors over a trilinear chromaticity plot of colors is illustrated qualitatively in Fig. 2. Reds are made more saturated. Cyans are also increased in saturation. Green colors are shifted in hue toward cyans. Magentas are shifted in hue toward reds. All of these effects are of somewhat comparable magnitudes; hence it would be difficult to associate with the change in this one parameter any simple pattern or any simple type of color change.

Association of Manual Controls with Visual Effects

As an alternative approach to the problem let us consider the color changes which we would like to have associated with our control knobs. Consider the array of reds from the neutral point up to the primary point as illustrated on the trilinear plot of Fig. 3. Suppose we have one control which would enable us to shift the hue of these reds toward yellow, as shown, without markedly changing their saturations, nor greatly influencing the hues or saturations of other colors. Turning the knob in the opposite direction would move the reds toward magenta. Suppose, likewise, that we had a second knob which, as shown in Fig. 4, could be used to cause the reds to become more saturated or less saturated, again without markedly affecting their hues and not greatly affecting other colors.

Such controls would make possible a simple association of control adjustment with resulting visual effects. If one saw a red color which he considered too saturated, or desaturated, or of slightly the wrong hue, adjustments could be made which would not be distorted by later changes in other controls. Such a solution to the problem also would appear attractive in terms of the number of knobs. Two knobs would be required for the

reds, one a hue control, and one a saturation control. A similar pair of knobs for the greens and another pair for the blues would involve six altogether. The number of knobs then matches the number of independent controls, six in each case.

Let us consider the possibilities of obtaining such a system. A careful examination of our basic equations (3), reveals that changes in the hue of the reds are affected most directly by the relative sizes of the coefficients, a_{21} and a_{31} . Similarly the hues of greens and blues are affected by a_{12} and a_{32} , and by a_{13} and a_{23} respectively. If a_{21} is made more negative, green will be taken out of the reds, thus moving them toward magenta. On the other hand, if a_{31} is made more negative, blue will be taken out of the reds thus moving them toward yellow. If one of the control knobs were made such that a_{21} increased in value as a_{31} became smaller and as the knob were turned in the opposite direction the opposite effects would take place, this would provide an excellent control on the hues of red colors.

The same controls will affect cyan colors as shown in Fig. 5. A change in reds toward yellow will move cyans toward blue. Similarly, a change in reds toward magenta will move cyans toward green. Thus the red hue control rotates the red-cyan axis. Similarly, the green hue control rotates the green-magenta axis, and the blue hue control rotates the blue-yellow axis.

Saturation control is a little more complicated. Saturations of reds, for example, can be increased by making constants a_{12} and a_{13} more negative. This in effect gives a greater red output signal for a given red input signal. Reds are made brighter and more saturated.

Saturations in reds are also affected by a_{21} and a_{31} . Making them more negative removes green and blue outputs from reds, thus making the reds more saturated. This also tends to make reds slightly darker.

Any of these changes have secondary effects. Changes in a_{12} and a_{13} will also affect the saturations of the cyan, as will changes in a_{21} and a_{31} . Changes in a_{12} and a_{13} will also affect hues unless a_{32} and a_{23} are similarly changed. We are of the opinion that the most practical way to control red saturation is to link the coefficients a_{21} and a_{31} . If these two constants are increased or decreased together, red saturations are controlled without major hue changes. Saturations of other colors are affected to a lesser extent. Similarly, a green saturation control may link coefficients a_{12} and a_{32} ; while a

blue saturation control may link coefficients a_{13} and a_{23} .

Fig. 6 shows a panel with the controls desired. The knobs in the upper row are associated with the six potentiometers of the masking unit shown in Fig. 1. The knobs* in the lower row provide hue and saturation control for reds, greens, and blues independently. The red-cyan axis hue knob controls coefficients a_{21} and a_{31} . Turning the knob in one direction increases a_{21} and decreases a_{31} , while turning it in the opposite direction has the reverse effect on these two coefficients. Two other controls similarly affect the hues along the green-magenta axis and along the blue-yellow axis. Saturation controls for the red axis involve the simultaneous increase or decrease of coefficients a_{21} and a_{31} . Similarly, saturation controls for the green and blue axes involve ganging of coefficients a_{12} and a_{22} , and a_{13} and a_{23} as shown. These illustrations are schematic and undoubtedly other arrangements can be devised.

Calculated Effect of Ganged Controls

To determine the probable effects of such a system of controls we made calculations for a number of colors. We assumed the television system shown in Fig. 7. The signals, R_1 , G_1 , and B_1 , from the film scanner are gamma overcorrected by an exponent of 0.67, and are then given the normal gamma correction of 0.45. Next comes the masking unit. The signals then go through the encoder and are transmitted to a standard color television receiver. The color picture tube is assumed to have a gamma exponent of 2.2.

In our calculations we first established the coefficients for the masking equations which would apply to Eastman Color Print Film, Type 5382 in this system. These coefficients were determined by a method of least squares for a number of Munsell

patches. Our procedure is further described in the appendix.

The resulting equations are as follows:

$$\begin{aligned} R_0 &= 1.9R_1 - 1.2G_1 + 0.3B_1 \\ G_0 &= -0.6R_1 + 1.9G_1 - 0.3B_1 \\ B_0 &= 0.1R_1 - 0.7G_1 + 1.6B_1 \end{aligned} \quad (4)$$

The coefficients are rather large and two of the nondiagonal coefficients are positive. The coefficients would be smaller if gamma overcorrection were not used.

To determine the effect of the red hue control, we added 0.40 to a_{21} and subtracted 0.40 from a_{31} . The chromatic changes are illustrated in the trilinear plot of Fig. 8. The colors along the red-cyan axis are changed in hue as predicted. There are minor changes in the remaining colors.

To determine the effect of our red saturation control, we subtracted 0.40 from a_{21} and a_{31} . The chromatic changes are illustrated in Fig. 9. Here the reds are increased in saturation. The side effects, however, are greater than for the hue shift controls. The side effects are increases in saturations of other colors, but to a lesser degree than for red. Hues, however, remain relatively constant. We made similar calculations covering changes in other pairs of coefficients corresponding to the green and blue axes. The effects are similar to those which have been described and are shown in Figs. 10, 11, 12, and 13. In these last six figures, the numbers associated with some of the arrows represent relative luminance values.

Conclusion

We have proposed that in the masking unit of a color television studio a simple means for associating manual adjustments and visual effects may be obtained. A block diagram of a suitable masking unit was shown. This unit automatically holds the neutral balance. A proposed control arrangement for this circuit was illustrated.

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1. W. L. Brewer, J. H. Ladd, and J. E. Pinney, "Brightness Modification Proposals for Televising Color Film," Proc. I.R.E., vol. 42, no. 1, pp. 174-191; January, 1954.
2. J. H. Ladd and W. L. Brewer, "Photographic Simulation of Color Television Brightness Modifications," Convention Record of the I.R.E. 1954 National Convention, vol. 2, part 7, Broadcasting and Television, pp. 110-113; 1954.

* In Fig. 6 are shown two rows of panel control knobs. The lines connecting the knobs represent belts running between pulleys on the shafts indicated. A plus sign at the end of a belt indicates that when the lower shaft is rotated, the upper shaft rotates in the same direction. A minus sign at the end of a belt indicates that when the lower shaft is rotated in one direction, the upper shaft rotates in the opposite direction. If all the pulleys indicated were rigidly connected to their respective shafts the system would be immobile. One way to prevent this is to make one pulley on each of the lower shafts free from the shaft. This pulley is engaged only when the lower control knob is depressed. It is then necessary to push in each lower knob before it is turned, in order to make desired adjustments.

3. R. P. Burr, "The Use of Electronic Masking in Color Television," Proc. I.R.E., vol. 42, no. 1, pp. 192-200; January, 1954.
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6. W. E. Tucker and F. J. Janda, "RCA Color Slide Camera TK-4A," Broadcast News, no. 77, pp. 52-57; January-February, 1954.
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Appendix

Masking Equation Coefficients

The coefficients for the masking equations (equations 4) were computed by the method of least squares for a sample of 13 Munsell patches distributed in color space. Since the answer depends upon the starting sample chosen, the coefficient values were rounded to the nearest tenth. The coefficients were established by the following procedure.

The 13 Munsell patches were each photographed on Eastman Color Negative Film, Type 5248. The illuminant was 3200 K, for which that film was designed. Prints were made on Eastman Color Print Film, Type 5382. The color television film scanner output signals from the Eastman Color Print Film reproductions of the original Munsell colors were measured on a color densitometer for which the red, green, and blue responses had been adjusted to approximate "ideal" practical responses already published.⁶ These synthetic scanner signals (R_1 , G_1 , and B_1) were routed as indicated in Fig. 7 to obtain the masking device inputs, R_1 , G_1 , and B_1 .

The corresponding masking outputs, R_0 , G_0 , and B_0 , were found by computing back from the desired color picture displays. For each Munsell color patch it was desired that the chromaticity (x_s , y_s) displayed on the color television picture tube be that which the patch itself would have if viewed directly under CIE illuminant C. However, it was desired that for each Munsell patch the luminance (Y_2) displayed on the picture

tube be computed by the equation

$$Y_2 = 0.30 R_2 + 0.59 G_2 + 0.11 B_2, \quad (5)$$

or

$$Y_2 = 0.30 (R_1)^{0.67} + 0.59 (G_1)^{0.67} + 0.11 (B_1)^{0.67} \quad (6)$$

This latter equation implies that the Eastman Color Print Film and the scanner were balanced for CIE illuminant C. The fractional exponent compensated for the greater-than-unity contrast of the film.

From the CIE tristimulus values (X_2 , Y_2 , and Z_2) for each display on the color picture tube, the corresponding phosphor input signals (R_* , G_* , and B_*) were computed, assuming standard NTSC phosphors. The equations⁷ are as follows:

$$R_* = 0.670 X_2 + 0.330 Y_2$$

$$G_* = 0.210 X_2 + 0.710 Y_2 + 0.080 Z_2 \quad (7)$$

$$B_* = 0.140 X_2 + 0.080 Y_2 + 0.780 Z_2$$

The 1/2.2-power of each phosphor input signal yielded the corresponding receiver decoder output signal. Transmission noise was neglected; hence each receiver decoder output signal was identical with the corresponding studio encoder input signal. By this procedure 13 sets of masking output signals were computed.

From the 13 sets of masking input signals and corresponding 13 sets of masking output signals, the coefficients in the masking equations were computed by the method of least squares.

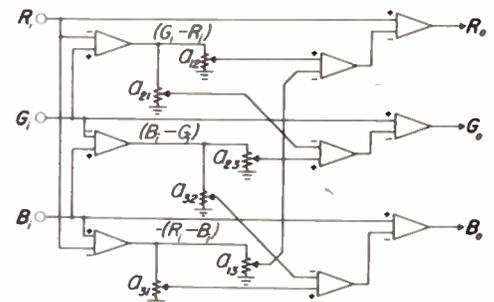


Fig. 1 Block diagram of color television studio masking circuit in which the six independent controls are suitable for ganging to simplify the relationship between manual adjustments and visual effects.

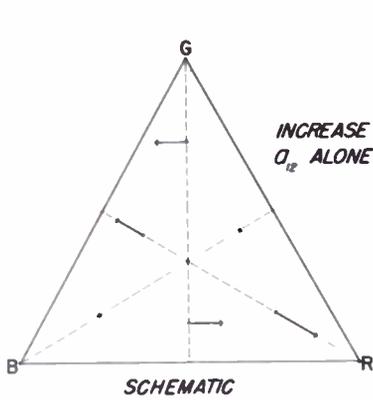


Fig. 2 Trilinear plot showing qualitatively the visual hue and saturation effects associated with a change in a single masking coefficient, a_{12} .

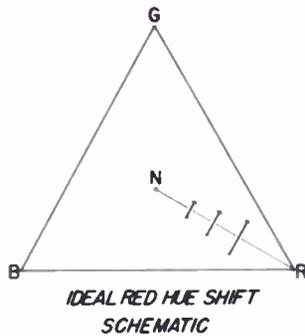


Fig. 3 Trilinear plot showing visual hue shift of reds alone. A manual control to do this would be desirable.

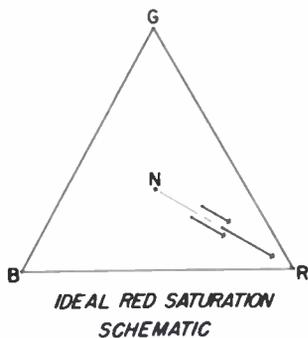


Fig. 4 Trilinear plot showing visual saturation increase of reds alone. A manual control to do this would be desirable.

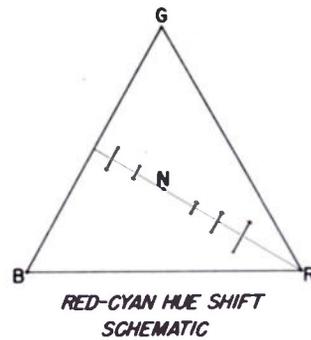


Fig. 5 Trilinear plot showing visual hue shift of cyan colors which occurs when red hues are shifted. The red-cyan axis rotates about the neutral point.

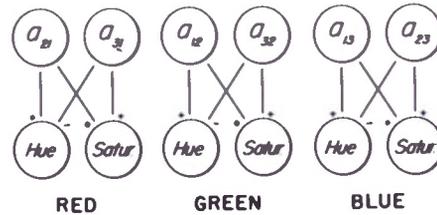


Fig. 6 Panel with two rows of color television studio masking control knobs. The knobs in the upper row provide direct manual adjustment of the six independent parameters shown in the masking circuit of Fig. 1. The knobs in the lower row provide manual adjustments which are associated with specific visual effects. The lines connecting the knobs represent belts running between pulleys on the shafts indicated. A minus sign indicates a crossed belt.

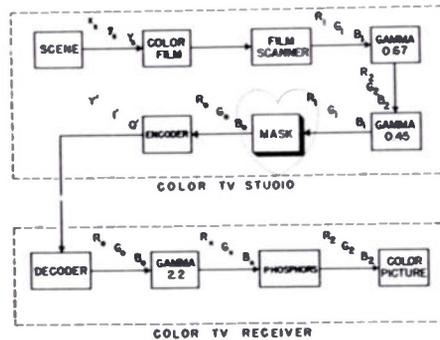


Fig. 7 Block diagram of color television film chain.

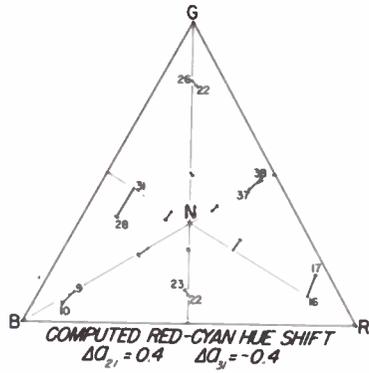


Fig. 8 Trilinear plot showing calculated visual effects from a 0.4 clockwise rotation of the red hue knob of Fig. 6. In this and the succeeding figures, the numbers associated with some of the arrows represent relative luminance values. In this and the succeeding figures, the initial positions of the upper row of control knobs of Fig. 6 are given by equations 4.

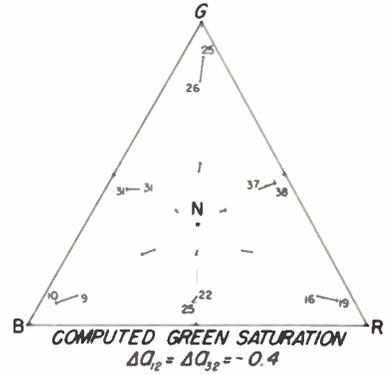


Fig. 11 Trilinear plot showing calculated visual effects from a 0.4 counterclockwise rotation of the green saturation knob of Fig. 6.

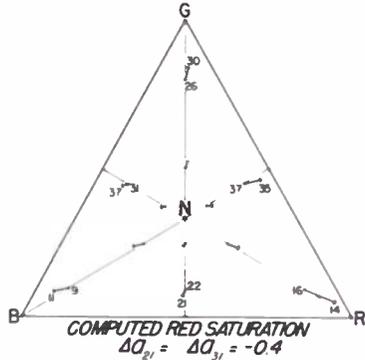


Fig. 9 Trilinear plot showing calculated visual effects from a 0.4 counterclockwise rotation of the red saturation knob of Fig. 6.

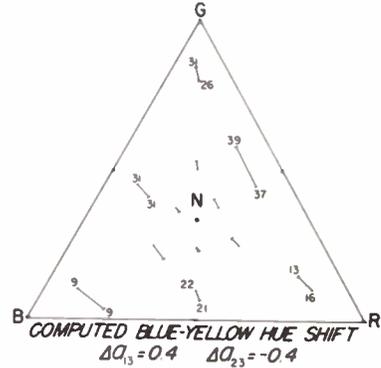


Fig. 12 Trilinear plot showing calculated visual effects from a 0.4 clockwise rotation of the blue hue knob of Fig. 6.

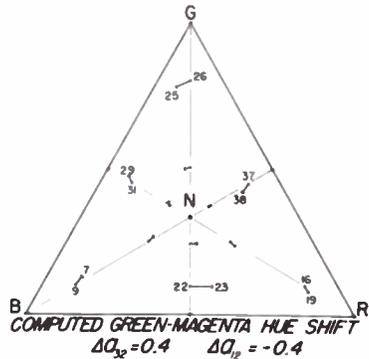


Fig. 10 Trilinear plot showing calculated visual effects from a 0.4 counterclockwise rotation of the green hue knob of Fig. 6

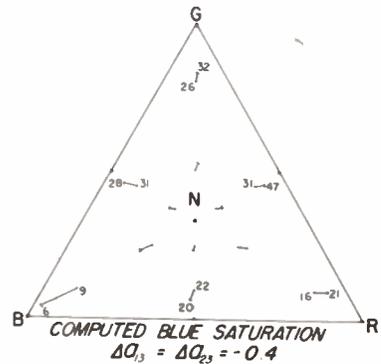


Fig. 13 Trilinear plot showing calculated visual effects from a 0.4 counterclockwise rotation of the blue saturation knob of Fig. 6.

EXPERIMENTAL EQUIPMENT FOR RECORDING AND REPRODUCING COLOR TELEVISION IMAGES ON BLACK AND WHITE FILM

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REQUIREMENTS OF A RECORDING MEDIUM

One of the most serious problems in the way of expansion of color television broadcasting is the lack of a convenient and economical recording scheme. Such a scheme must be satisfactory for the performance of several basic tasks. First, it must be capable of recording live color scenes for later use in color television broadcasting. It would be convenient if this first task could be accomplished with some simple device such as a mechanical movie camera. Second, it should have facilities for making kinescope recordings of good quality which can be later used for broadcasting purposes. A third requirement concerns the reproducing device. It should be capable of playing back either the recording of the live scene or the kinescope recording. It is, of course, always desirable to make the mechanisms involved as simple and as reliable as possible. This last requirement has become more important with the increasing complexity of color television equipment.

At the 1954 convention of the Institute of Radio Engineers, a general system for performing the aforementioned tasks was proposed.* At that time the system had not been tried. This paper is a report of the experimental work done and the results obtained with the system to date.

BASIC DESCRIPTION OF THE SYSTEM

The system as originally proposed, illustrated in figure 1, employed color separation and electronic switching in combination. The three color primaries were the C. I. E. standard all positive coordinates \bar{x} , \bar{y} , and \bar{z} . There were two columns of images on the black and white film. Each individual image represented a television field for its own particular primary. The column of images on the left side of the film strip of figure 1 consisted simply of successive fields of \bar{y} or luminance information. The column of images on the right side of the film strip consisted of alternate fields of \bar{x} and \bar{z} or chrominance information. That is, for one particular field, \bar{x} and \bar{y} information would be recorded but \bar{z} information would be ignored. In the next field \bar{y} and \bar{z} information would be recorded, but \bar{x} information would be ignored, and so on.

In reading the information out during playback three scanner tubes (or one scanner tube and an optical beam splitter) were to be employed. The scanner tubes were to have a horizontal line only, and the film was to run continuously with one complete image passing any given point each 1/60 of a second. Even though only two images were recorded simultaneously, three were simultaneously scanned, as shown in figure 1. Thus, one piece of color information would be 1/60 of a second out of step with the remaining luminance and chromi-

nance information. \bar{x} and \bar{z} would alternate field by field in this role. The luminance information \bar{y} was then read field by field by photomultiplier 1. The two pieces of chrominance information, \bar{x} and \bar{z} , were read by photomultipliers 2 and 3, which had to be switched electronically at a field rate. The film transport mechanism was to be synchronized by comparing appropriate field marks shown on figure 1 with the vertical synchronizing pulses. The electronic switch would be synchronized with appropriate field marks. It was originally proposed that since fields ($262\frac{1}{2}$ lines) instead of frames (525 lines) were being recorded, the images need be just as wide and one half as high as an ordinary 16mm. image to provide adequate definition. This is certainly true for kinescope recordings if the film transport mechanism is capable of placing the film accurately. For images recorded with a mechanical camera, the limiting definition capabilities of present day optical systems and fine grain films are such that both horizontal and vertical definition will be essentially determined by the characteristic of the television system. Therefore, there would be trivial losses in definition caused by reducing the height of the image by the amount discussed. The film speed compatible with these requirements and also with a 30 frame 60 field television system was 45 feet per minute.

The reasoning behind the choice of \bar{x} , \bar{y} , and \bar{z} in the originally proposed system instead of the more conventional red, green, and blue is as follows: Let us assume that the entire system was either linear or properly gamma corrected. If the gains of the two chrominance channels were identical, there would be no advantage in choosing either set of primaries except that, theoretically at least, better color accuracy can be obtained with all positive \bar{x} , \bar{y} , and \bar{z} than with all positive red, green, and blue primaries. However, let us suppose that there is a slight difference in gain between the two color channels. If red, green, and blue were the three primaries, then red and blue would be the electronically switched channels since it would be desirable to keep the primary with the most luminance information (in this case green) in the channel which will never have any fields missing. If there is a saturated red (or blue) patch in the picture, that red (or blue) patch would have slightly different brightness for successive fields but constant brightness for alternate fields. This would be quite likely to produce a small luminance flicker which might be objectionable. On the other

* Hughes, Wm. L., FEASIBILITY AND TECHNIQUE OF STORING COLOR VIDEO INFORMATION ON BLACK AND WHITE FILM. Presented at 1954 National Convention of the Institute of Radio Engineers.

hand, suppose that \bar{x} , \bar{y} , and \bar{z} primaries were used. Then red, green, and blue signals would be obtained from them by matrixing. However, luminance information for all three is obtained from the \bar{y} signal, and variation in chrominance channel gains would produce primarily a chrominance flicker which would be much less visible than a corresponding all luminance flicker.* It should probably be said at this point that later experimental investigation proved that this luminance-chrominance flicker problem was not as serious as originally expected, though these factors did influence the choice of primaries in the originally proposed system. Also, for other reasons to be discussed later there is a definite advantage in keeping as much luminance information as possible in one channel.

Before proceeding further it is expedient to consider two possible methods which might be used to place the information on a film in a mechanical camera. The most obvious way is to design a mechanical camera which will record sixty different pictures a second. A rotating color wheel would be used to alternately place \bar{x} and \bar{z} on one side of the film. Luminance information would be placed at a field rate on the other side of the film. Probably the major difficulty with this system is that the mechanical camera would be expensive and completely non-conventional. Fortunately another method may be used which allows the easy conversion of already existing conventional 35mm. movie cameras. In this method the camera will have a pull-down mechanism that operates 30 times a second instead of 60 so that four complete images (two luminance and two chrominance) can be recorded simultaneously. This camera will be set to move twice as much film per pull-down as the first camera. An optical system composed of an anamorphic (aspect ratio distorting) lens and a four-way beam splitter would make it possible for this type of camera to perform for color television functions analogous to those of a standard movie camera. The use of appropriate color filters in the beam splitter and panchromatic film in the camera is all that is necessary to expose film from which high quality color television images can be taken directly after development.

THE EXPERIMENTAL EQUIPMENT

To test the feasibility of the proposed system, it was decided that two basic units must be either constructed or otherwise obtained by the conversion of existing equipment. These two units are a complete playback system and a mechanical camera for making a suitable test film. It was the considered opinion of those involved in this project that if these two units could be made to operate successfully, the technical feasibility of the system would be reasonably well established.

As is often true in the implementation of such

schemes certain changes from the basic plans became expedient as the project developed. One basic change made involved the choice of taking primaries. It will be recalled that initially standard C. I. E. primaries \bar{x} , \bar{y} , and \bar{z} were specified. It was later decided that the primaries for the first model would be all positive red, blue, and luminance information. The luminance primary would correspond to standard C. I. E. \bar{y} as closely as possible. It is true that this set of primaries does not give the theoretical color accuracy obtainable with the original primaries. Experience in color television broadcasting has proved, however, that all positive red, green, and blue primaries give perfectly satisfactory results from a subjective point of view. It would seem that such a choice of primaries would be reasonable for this system also. However a choice of primaries which we will symbolize by Y, R, and B (where Y = \bar{y}), has certain additional advantages. In this system, the green signal would be obtained by subtracting correct proportions of R and B from the Y signal and the anti-flicker feature in the presence of chroma gain differences would thereby be retained. In addition, the matrix required to obtain the green signal would be simple, since the other two channels would operate on a straight through basis. This represents a definite advantage over the \bar{x} , \bar{y} , \bar{z} system. Another advantage that might assume considerable importance is that the use of Y, R, and B primaries admits the possibility of direct gamma correction on the film. This makes possible the elimination of three complicated electronic chassis and also makes it possible to feed the three signals (after electronic switching of the chroma channels) directly to a standard N. T. S. C. encoder.⁷

Still another change in the experimental system is concerned with the film speed. This speed was increased from the originally proposed 45 feet per minute to a new figure of 54 feet per minute, which changed the individual image aspect ratio from 8:3 to 20:9. The reason for this increase in film speed is simple. Standard 35mm. film with negative perforations would be moved at the rate of 54 feet per minute if sixty individual sprocket holes passed any given point in one second. There is, then, a one to one ratio between images and sprocket holes, which means that conversion of existing standard movie cameras is greatly simplified. Also it was then possible to buy a 24 tooth sprocket wheel for

* Bailey, W. F., THE CONSTANT LUMINANCE PRINCIPLE IN N. T. S. C. COLOR TELEVISION. Proc. I. R. E. January, 1954. (Second color television issue), Pages 60-66.

⁷ Livingston, Donald C. COLORIMETRIC PROPERTIES OF GAMMA-CORRECTED COLOR TELEVISION SYSTEMS. Presented at 1953 National Convention of the Institute of Radio Engineers

a standard 35mm. simplex projector and mount it directly on the shaft of a commercially available Bodine synchronous motor-reducer (1800 rpm motor reduced 12:1) combination. When the motor was synchronized with the vertical synchronizing pulses, the combination made a simple film drive mechanism for test purposes.

Both of these basic changes in the system are illustrated in figure 2 which is a diagrammatic explanation of the modified system which was built. The field marks shown in figure 2 were not incorporated in the first experimental system. In the revised scheme these field marks would be useful in correcting the mode of electronic switching should it become reversed by accident; but they are no longer necessary to make the system work correctly, because both the drive mechanism and electronic switch are controlled by the vertical synchronizing pulse. Once the film is started correctly (there is a chance of error here) the switching will henceforth be correct for normal operation. In a commercial system, the field marks would be useful in that they could be used to insure correct starting of the film and fully automatic mode correction if something should happen during operation.

BLOCK DIAGRAM

A block diagram of the entire experimental system is shown in figure 3. A 5ZP16 is used for the fast decay scanner tube, and its surface is focused on three appropriate images on the film through the use of a special optical system. Three photomultipliers are placed so that each one receives light from a particular image on the film at any given instant. Each photomultiplier feeds a channel amplifier. The output of one channel amplifier (the Y channel) is fed through an aperture corrector which in turn feeds the output matrix. The other two channel amplifiers are fed to the two inputs of the electronic switch. Each of the two outputs of the electronic switch is fed to an aperture corrector and then to the matrix. Standard blanking is mixed with the video signals in the matrix chassis.

ELECTRONIC SWITCH

The circuitry of the entire system is quite conventional except for that of the electronic switch, which is given in figure 4. Provision is made for the separate adjustment of signal gain into each of the four keyer tubes through the use of four cathode followers (two 12AU7's). Each keyer tube is a 6J6 that is either biased to cutoff on one side because of extreme conduction of the other side or is allowed to operate as a simple low gain resistance-capacitance coupled amplifier on the first side when the second side is cut off. Low resistance potentiometers provide a means of obtaining d-c balance on the amplifier sides. The switching sides are

driven by an approximately 100 volt peak to peak square wave which has a frequency of 30 cycles per second. The square wave is obtained from a conventional Eccles-Jordan multivibrator. This multivibrator is driven by a Schmidt trigger circuit which is modified with an inductance -- damper diode combination in one plate so that it produces one very high pulse for each 60 cycle sine wave that is used to drive it. Since each pulse changes the Eccles-Jordan multivibrator mode, the combination is effectively a 2:1 frequency divider. Extensive decoupling must be used in the plate circuits of the keyer tubes, since the 30 cycle square wave must be passed with essentially no tilt. If this is not done, it is extremely difficult to balance the electronic switch to give no chroma flicker. Since no compensation is provided for high frequencies in the cathode followers and in the keyer tubes, it is necessary to insert overcompensated video amplifiers in each channel. These amplifiers serve to keep the signal at a usable level, since the keyer amplifiers have gains of much less than unity. They also serve to correct the high frequency roll-off caused by the keyer circuits. The entire unit has a frequency response of plus or minus one decibel from 20 cycles to 5 megacycles. D-c and a-c stability of the switch are good enough so that once careful adjustment is made, it is not necessary to adjust it again for several days, provided at least a 15 minute warm-up period is allowed when the switch is turned on.

OPTICAL SYSTEM

The optical system required for this system is quite unique. The image from the cathode ray tube face must be split into three images, and each of these three images must be placed on the film in a particular place. This optical system is illustrated in figure 5. Three individual 75mm. Carl Zeiss Tessar lenses are used. These lenses are displaced from the optical axis center as little as possible. The individual beams are bent through two 90 degree angles through the use of simple reflecting prisms. Two of the three lens-prism combinations are shown in figure 5. These two represent the chroma channels. The Y channel combination is not shown, but it is similarly constructed. The only difference is that the prisms displace the beam laterally rather than vertically. Registration is accomplished merely by adjusting the lenses. A photograph of this optical system is shown in figure 6. Once the individual images on the film have been scanned, it is necessary to transmit the gathered light from each image to the respective photomultipliers. This is accomplished through the use of polished optical lucite light pipes. A photograph of the entire photomultiplier pickup head is shown in figure 7. It should be remembered that no electronic circuitry is involved in the registration of the three images and excellent registry can be

obtained. If such a device is ever built commercially, and the lenses are provided with micrometer adjustments, registration should become a task that can be accomplished in a few seconds.

A photograph of the entire system plus monitoring equipment is shown in figure 8. It must be remembered that in this experimental system, all of the individual functions (channel amplification, aperture correction, switching, matrixing, etc.) were built into separate chassis to facilitate experimental work. In commercial equipment, the electronics could be greatly compressed.

TEST FILM

To test the experiment system adequately, it was necessary to make a suitable test film. This film was made in the following steps. A large test pattern was drawn with the vertical dimension compressed such that its aspect ratio was 20:9. Three identical photostats were made of this test pattern drawing. Papers of various appropriate shades of gray were pasted on the different photostat copies of the test pattern to produce colored bars and wedges, so that typical red, blue, and green patterns were obtained. A luminance pattern was not made because it was desired to test the system critically for flicker. Therefore if a regular green pattern is used and appropriate red and blue signals are not subtracted from it in the matrix, the most severe flicker features of the system will be present. Further, since saturated primaries would have the greatest tendency to flicker, the bars and wedges were chosen to be saturated red, green, yellow, and blue. The problem is particularly significant with the red and blue primaries. Small black and white negatives were made of these test patterns, and these negatives were carefully placed between plate glass in the proper orientation. Two green black and white negatives were placed one directly above the other. To the right of one green negative was placed a red negative. A blue negative was placed to the right of the other green negative. The plate glass with the negatives was then placed in a specially constructed integrating light box. The light box was simply a cube approximately three feet on a side and painted on the inside with a white diffuse reflecting paint. A hole was cut in one side of the cube with dimensions large enough to cover the four negatives mounted in the plate glass. Inside the cube and on the same side as the hole were placed twelve 500 watt number 2 photoflood bulbs. These bulbs were oriented so that they completely surrounded the hole, but they were shielded so that no direct rays from the bulbs could fall on the plate glass. Thus any light falling on the negatives had to go through at least one diffuse reflection. With the plate glass in place, the luminance variation over its entire surface was found to be less than one-half of one percent. This figure was considered satisfactory.

If there had been any significant luminance variation over the surface the Y channel on the film would have an inherent 30 cycle flicker regardless of the state of perfection of the electronic scanner it is used to test. Next a standard 35mm. Bell and Howell Eyemo Model K movie camera was modified to pull down two sprocket holes instead of four for each revolution of the shutter. When the camera was loaded and placed the correct distance from the illuminated negatives in the plate glass, a test film of high quality was made with it. Figure 9 is a photograph of the whole test film assembly. Figure 10 shows a sample of the resulting test film. The only things that must be done to make the camera capable of producing live films for use in the color scanner are: (1) Replace the lens with a proper beam splitting and anamorphous optical system, and (2) load the camera with panchromatic film.

PERFORMANCE

The general performance of the equipment is quite satisfactory.* As stated before, good registry is quite easy to obtain. With proper adjustment, there is no visible thirty cycle flicker. As a matter of fact, it is possible to put either individual color channel on a black and white monitor with no detectable flicker. The simple and economical film drive mechanism that was used for these tests gave a picture stability that can be described as about equivalent to that of a medium quality 16mm. projector. However, considerable improvement could be made. It was felt that this problem was one that could be handled easily by qualified engineers in that field. The film transport mechanism required for this application appears to be quite simple compared with that of a continuous motion projector for standard film because there is no moving optical system.

Some mention should be made of certain problems that arose in scanning a single line on the fast decay cathode ray tube. For moving film, an acceleration potential of 15 kilovolts and a beam current of 10 to 15 microamperes were used. The pictures obtained had good definition, and the noise content was below that normally seen in the pictures obtained from scanners using color film. This is due in part to the fact that the 5ZP16 is rich in blue and ultraviolet light and the photo-multipliers are characteristically sensitive in this region of the spectrum. Nevertheless, more light would not be detrimental. With these tube parameters it appears that at least five hundred hours can be obtained from any given tube, since the position of the line can occasionally be changed. An alternative is to scan the equivalent of

*In the oral presentation, a color slide of the test pattern as presented on a 156GP22 tricolor tube was presented to show overall color rendition.

approximately ten lines in a vertical direction. This puts a negligible vertical distortion in the reproduced picture while making it possible for the tube life to be increased probably to the life of the electron gun. Still another alternative is to put a standard raster on the tube and compress it optically. One feature of the system is that if the film is stopped, the electronic switch disabled, and a raster placed on the cathode ray tube instead of a line, the system will faithfully reproduce the frame. One difficulty with this procedure is that where the line was originally scanned, the x-ray darkening of the glass will be more significant as well as the fact that the light output of the phosphor may be slightly reduced. Therefore a slightly darkened line will appear in the picture when the film is not running. This defect is not present, of course, when the film is running. In any case it can easily be eliminated by using one cathode ray tube for still frames and one for moving film. If a compressing optical system is used for running film, the difficulty is not present.

POSSIBILITIES OF 30 CYCLE COLOR FRINGING

All of the films that have been used to test the system have been made from still subjects. The possibility has been suggested that, if it is attempted to record rapid motion, certain color fringing difficulties might occur. This possibility arises from the fact that four images (two luminance and two chrominance) are recorded simultaneously when the more economical camera system is used. Thus successive images in either column are either simultaneous or $1/30$ of a second apart. This presents no difficulty except on alternate fields when the luminance and one piece of chrominance information are simultaneous but the second piece of chrominance information is lagging by $1/30$ of a second. It is not yet known whether or not this problem will be serious in the system that has been described thus far. If it is objectionable, however, there appears to be a relatively simple way to overcome the problem. This technique requires a simple modification in the scanning procedure when the film is run in the playback system. This modification is illustrated in figure 11. A fourth photomultiplier is used, and four images are scanned simultaneously instead of three. For a given field, one luminance image, its adjacent chrominance image, and the chrominance image directly above that adjacent chrominance image are scanned simultaneously. It will be assumed for discussion purposes that these three images were recorded simultaneously. In addition, however, the chrominance image directly below the chrominance image adjacent to the luminance image will also be scanned. For this field, then, if chrominance photomultipliers 2 and 3 are selected by the electronic switch all of the luminance

and chrominance information will be simultaneous. For the next field, chrominance photomultipliers 2 and 4 will be selected by the electronic switch and luminance and chrominance information will still be completely simultaneous. This procedure is repeated for the next set of four images and so on. This system adds a little complexity to the optical system and requires an additional photomultiplier and channel amplifier. A simple study of electronic switching techniques will show, however, that the complexity of the switch required to perform this type of switching is not increased at all over that required by the original switching system.

FEASIBILITY OF KINESCOPE RECORDING

One technique of making recordings is to place a special mechanical camera in front of a dichroic display and synchronize the camera with the frame rate. It may be, however, that this technique, while theoretically possible, is not the most economical one. Figure 12 illustrates another possible method which may be more desirable. This kinescope recorder consists of two scanner tubes and a film transport mechanism much like the one required for the playback mechanism. One scanner tube is controlled with the luminance information and the other is controlled with appropriately switched chrominance information. Motion through the film transport mechanism is continuous, and the scanner tubes have no vertical sweep as before. A difficulty encountered with this type of system is that every other field of color information is thrown away for both red and blue. This is equivalent to decreasing vertical chroma definition. If it can be assumed, however, that a decrease in vertical chroma definition until it is equivalent to the existing horizontal definition in the chroma channels is allowable, then this unused color information should not present noticeable losses to the complete color picture. This system also inherently produces a possibility of $1/60$ of a second color fringing for fast motion. However, the short time involved coupled with low bandwidth chroma channels should not allow this problem to be serious. As a matter of fact, there is at least one piece of color television equipment available today which has this inherent possibility and yet has proved itself capable of excellent performance.

CONCLUSIONS

The experimental work done thus far seems to indicate that a film scheme for color television using black and white film with a combination electronic switching and color separation process is technically feasible. Further, the equipment involved for a complete system appears to be somewhat smaller in quantity and simpler in design than that required for a system using color film. Although they have not been proved, kinescope

recording possibilities with reasonable economy appear to be good.

It is well established, of course, that black and white film when used in a color separation process is an excellent way to record and store indefinitely high quality color pictures. In this particular system (which is devised specifically for color television) the running film costs at current prices for purchasing and processing film appear to have between a two and three to one cost advantage over 16mm. color film required to perform the same function. With proper operation, the image quality should be at least as good as that obtainable with conventional 16mm. color film. There is no cross coupling because of imperfect dyes. Also no electronic masking or over-gamma correction is required. The fact that no high degree of skill is necessary to run a black and white film processor is important to the operation of an ordinary television station. Whether or not television stations will be willing to operate a color film processor remains to be seen.

It is difficult to compare such a system with magnetic tape systems. At present, the only function in which the two systems might be competitive is kinescope recording, and there are not enough data available on either system for a fair evaluation.

The help and advice of a large number of people should be acknowledged. In particular it is necessary to thank Dr. Lyle Brewer and Mr. Jack Pinney of the Eastman Kodak Company for much assistance and advice in the handling of film. Dr. Lester Earls and Dr. Percy Carr of the Physics Department at Iowa State College gave much valuable advice in both the fields of colorimetry and optics. Many valuable suggestions were given by Mr. Paul Kristensen and Mr. Charles Otis of the Engineering Experiment Station and Electrical Engineering staff at Iowa State College. Finally, acknowledgement should be given to Messrs. Thomas Proctor, Leslie Westenburg, and LeRoy Anderson, students in Electrical Engineering, who actually built the equipment and who made many valuable suggestions for its improvement.

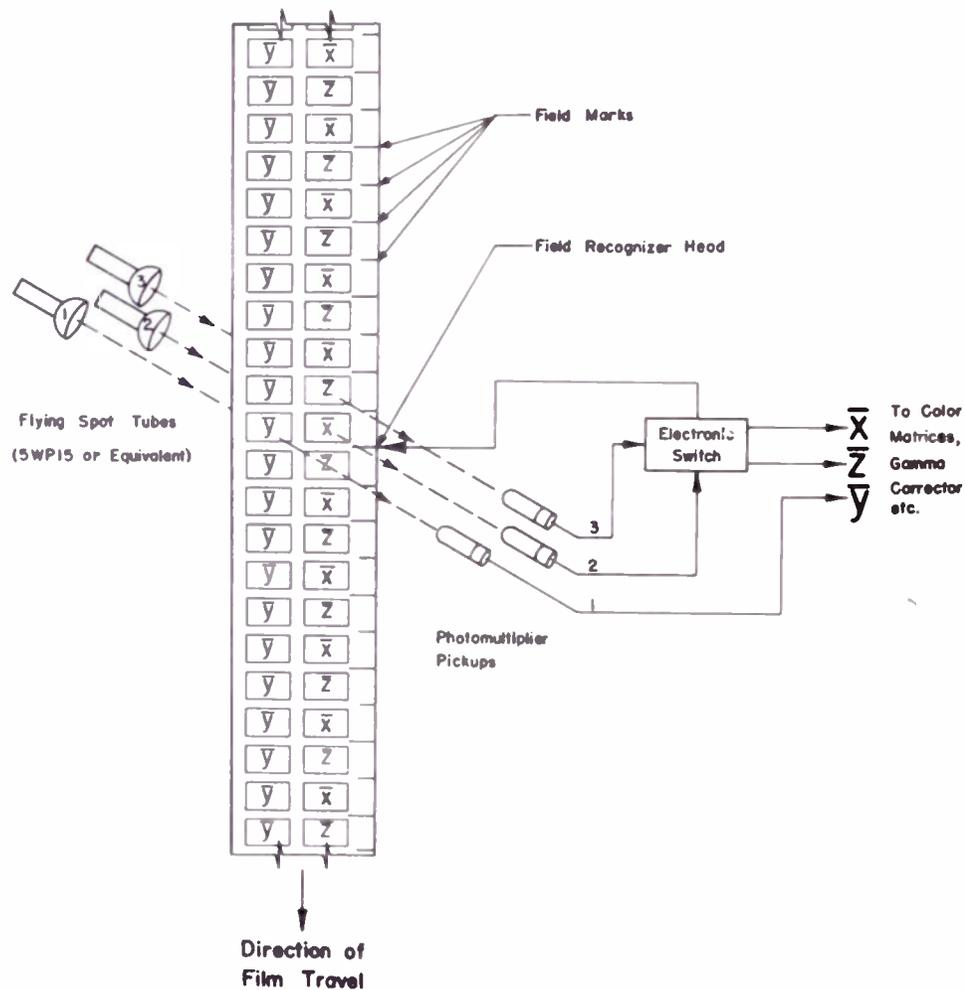


Fig. 1
Continuous film scanner for color television

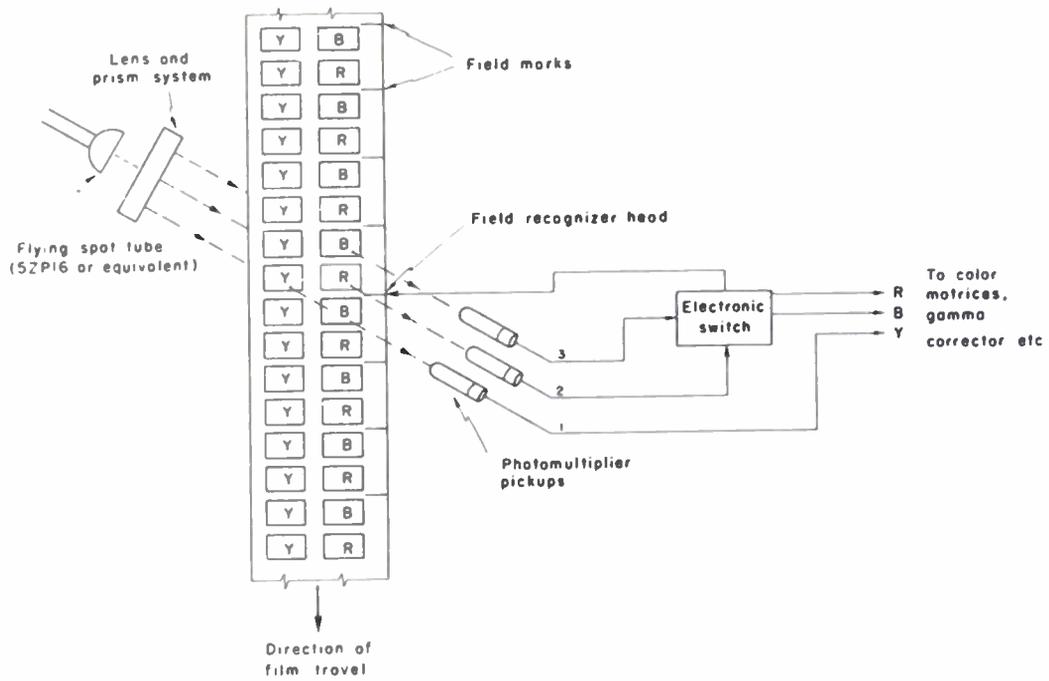


Fig. 2
Continuous film scanner for color television
(First type)

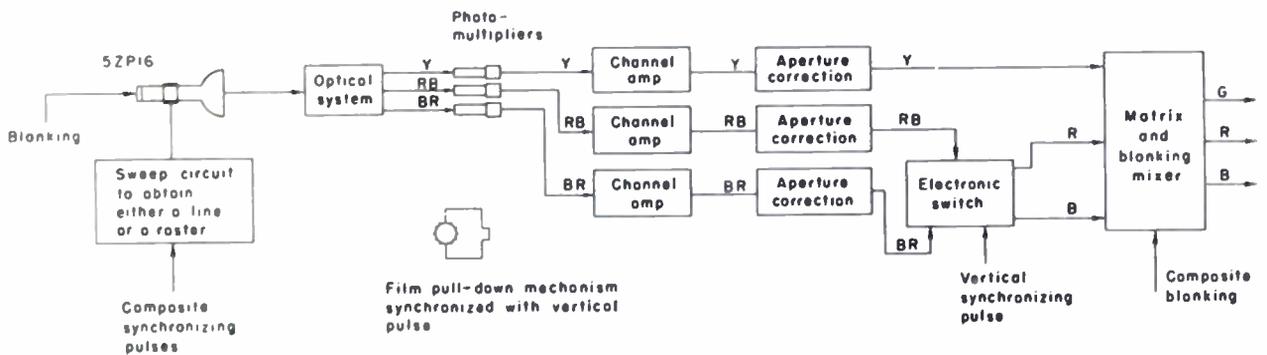


Fig. 3
Basic electronic system.

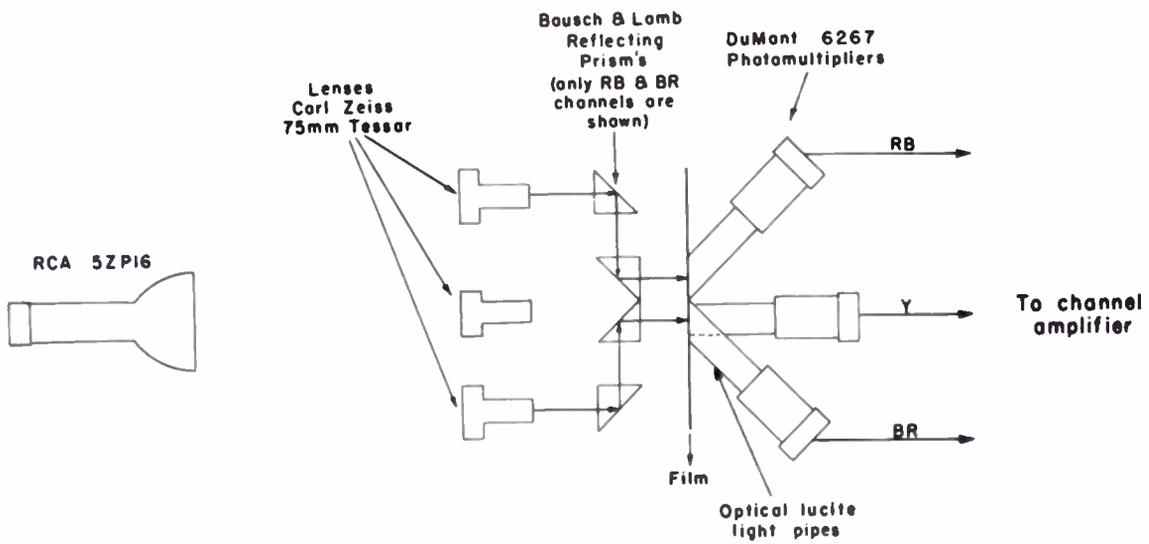


Fig. 5
Basic optical system

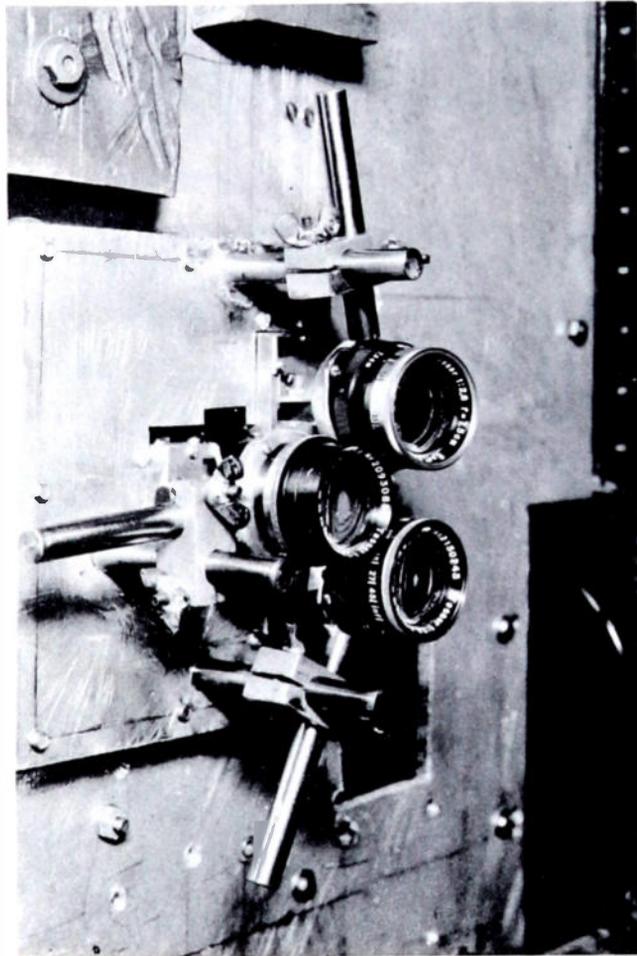


Fig. 6
Lens and prism system

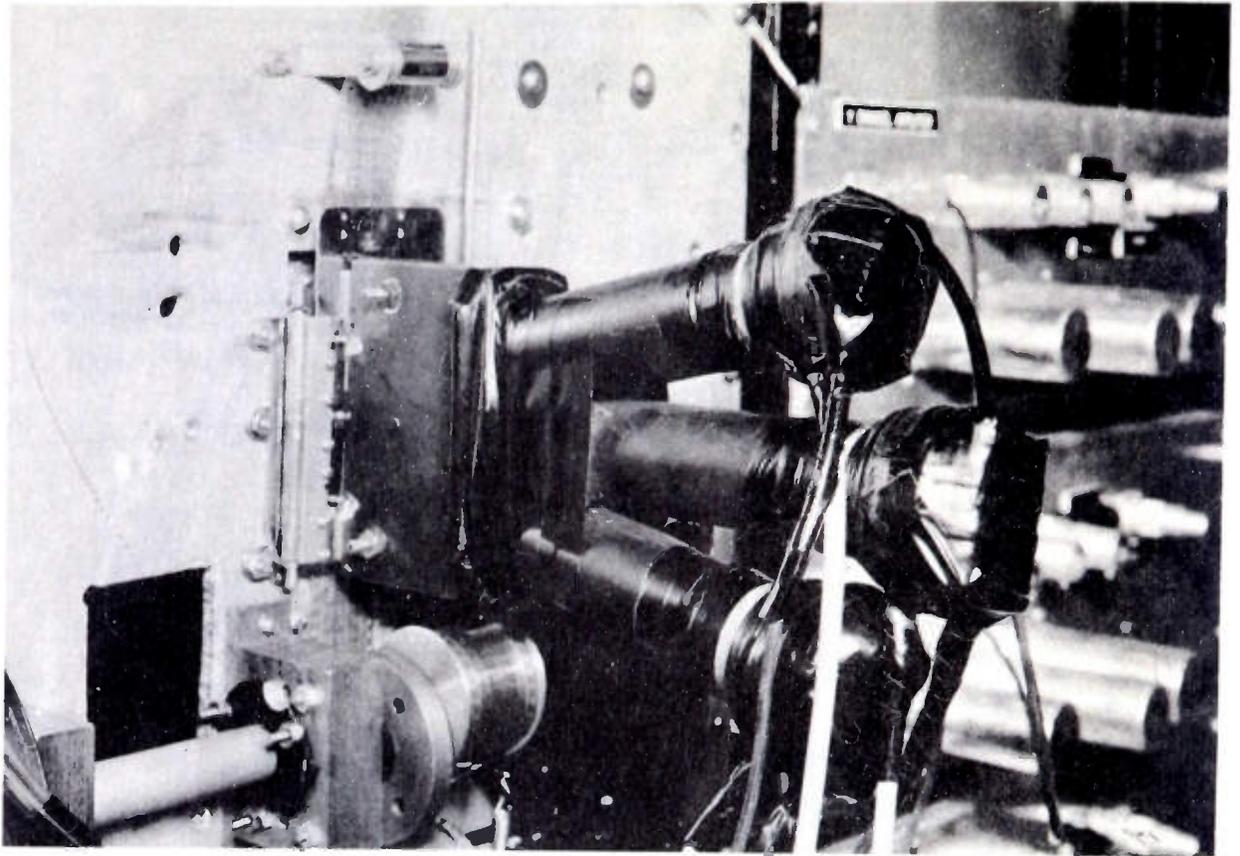


Fig. 7
Pick-up head

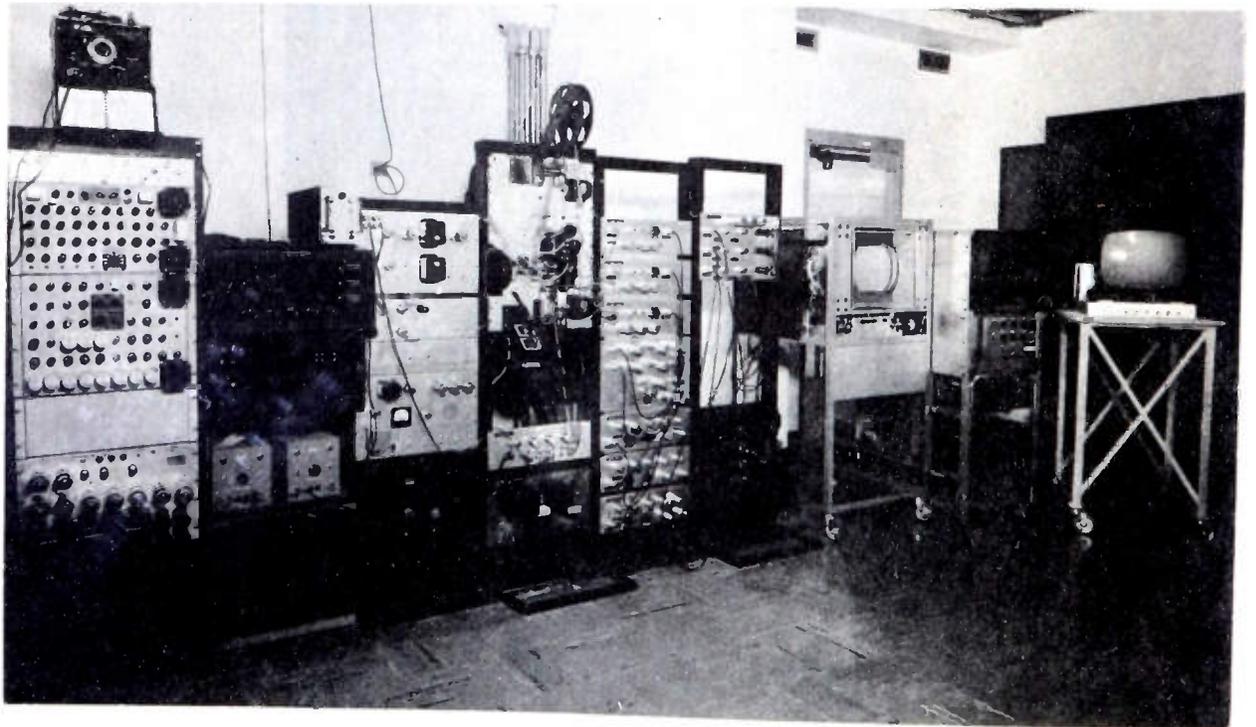


Fig. 8
Entire system

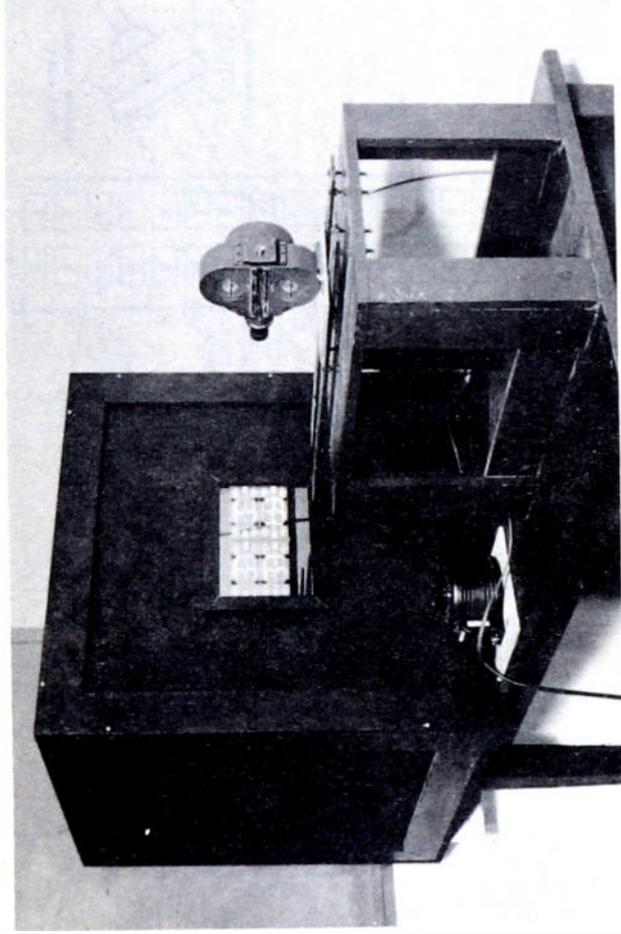


Fig. 9
Equipment for making test film

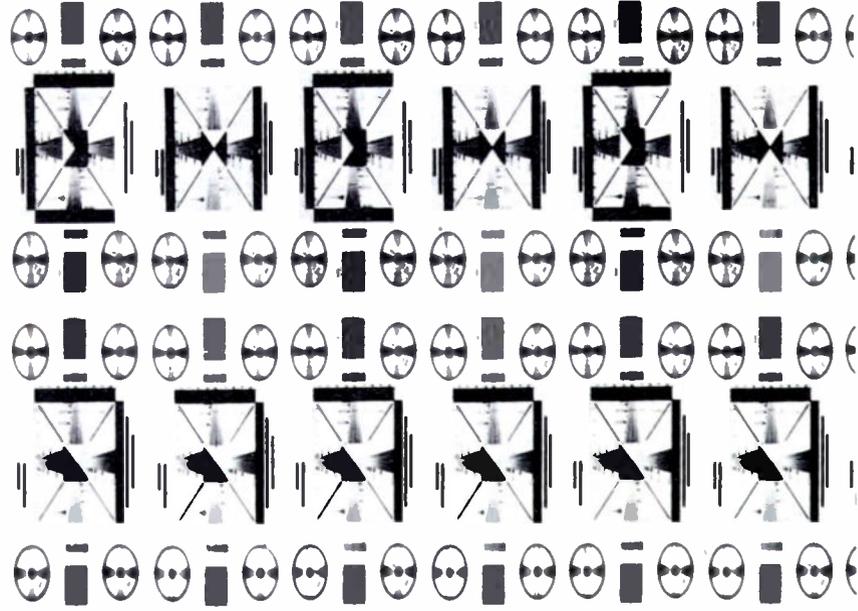


Fig. 10
Test Film

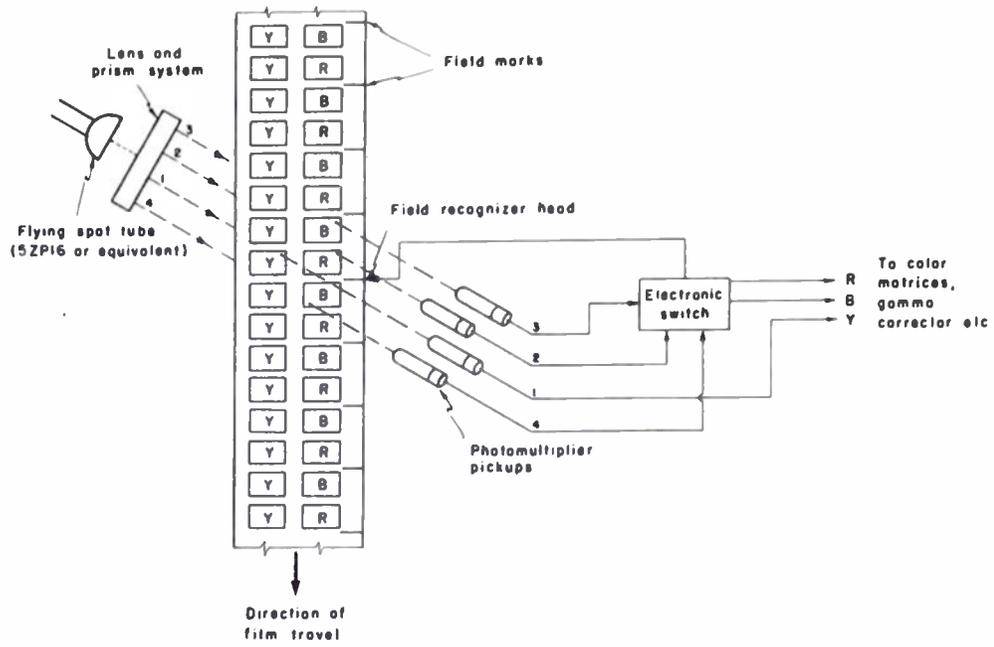


Fig. 11
Continuous film scanner for color television
(Second type)

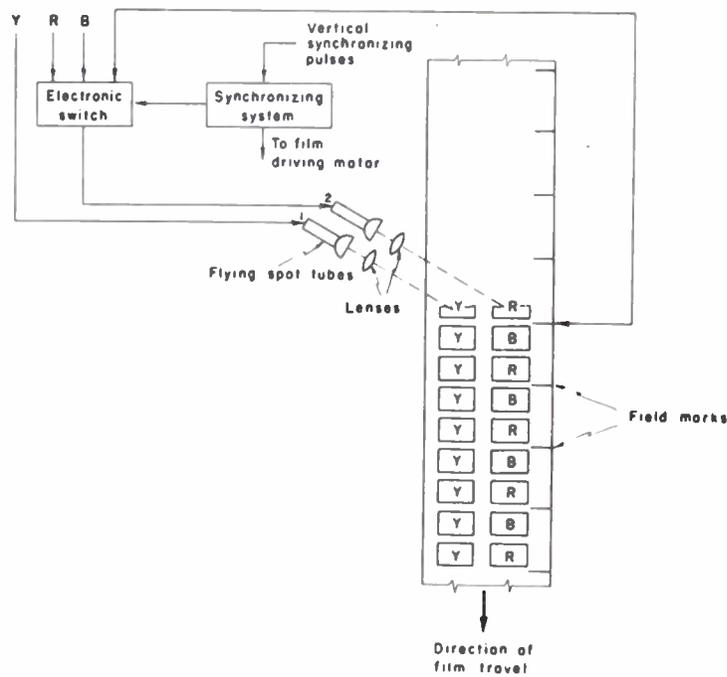


Fig. 12
Continuous motion kinescope recorder for color television

CATHODE-RAY VECTORGRAPH

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Summary

This paper describes a new Cathode-ray Vectorgraph developed primarily for color television instrumentation.

Introduction

The Cathode-ray Vectorgraph is an oscillograph developed primarily to simultaneously display the N.T.S.C. color television chrominance signals. In order that we may obtain a better understanding of the purpose and application of this instrument, we shall first consider some of the basic concepts of the N.T.S.C. color television system.

The Chrominance Signals

The location of any point in a plane of rectangular coordinates is defined in terms of its X and Y values. Hue and saturation of any color can likewise be defined in terms of the values along two chrominance or color defining axes. In color television these chrominance values are represented by two electrical signals known as I and Q. It is the phase angle of their instantaneous vector sum which determines the color that the receiver is to display. It is the amplitude of their instantaneous vector sum which determines the degree of saturation, that is, chromatic purity.

These chrominance signals are color camera information which has been processed in the unit immediately following the color camera. This unit, known as the Encoder, converts the red, blue, and green camera signals and the synchronizing signals into the composite color video signal. Matrices in the encoder combine appropriate amounts of the red, blue, and green camera signal to produce the two chrominance signals. The chrominance signals are recovered in the receiver where they are used to regenerate the red, blue, and green signals in another matrix circuit. They are also present in a test unit known as a Vector Decoder. The Vector Decoder and the Cathode-ray Vectorgraph are used as a unit to measure the chrominance components of the composite color television signal.

If the source information is a color bar test pattern, the chrominance signals will appear as in figure 1. This is a photograph of the I signal (above) and the Q signal (below). Each minor scale division horizontally represents 3 microseconds. The sequence of the color bars was green, yellow, red, magenta,

white, cyan, blue, black. The small pulse at the right hand end of each trace is the demodulated burst signal. The horizontal sync pulse does not appear because the chrominance signals for blacker than black are zero. The zero DC voltage axis is the number one minor division for the Q signal and the number 9 minor division for the I signal. The white bar, black bar, and horizontal sync pulse all fall on these lines indicating that the encoder was properly adjusted.

The oscillogram of figure 1, was made in the color studios of the Du Mont Television Network. Figure 2, is a block diagram of the method used to simultaneously display the I signal and Q signal. A TV line selector was used to synchronize a wide band Cathode-ray Oscillograph. I and Q signals were simultaneously fed into a High Frequency Electronic Switch operating in the triggered condition so that I and Q were alternately fed into the Y input of the oscillograph. Since the video information of all the lines of the bar pattern used was the same, adjacent lines could be used and the resulting chrominance signals would still represent the color information of a single line.

The Vector Display

When the chrominance signals are applied to the X and Y channels of an oscillograph, the instantaneous position of the spot on the face of the cathode-ray tube represents the instantaneous vector sum of the chrominance signals. The resultant display on the face of the Cathode-ray tube will appear as in figure 3. The center dot represents zero DC for both the I signal and the Q signal. This in turn represents black, white, and the intermediate gray scale. The Y or brightness signal carries all the picture information in this case. The small dot with a transition loop to the left of center and elevated above the horizontal by 33 degrees is the burst. The outer 6 dots represent the 3 primary colors and their negatives, that is, the 3 secondary colors. No Z axis signal was applied to the cathode-ray tube. The bright spots are the rest positions of the beam during the flat portion of each step of the chrominance signals. The transition lines are much fainter due to the rapid transition from one position of rest to another. The bandwidth of the X and Y channels of the Cathode-ray Vectorgraph used to make this photograph is approximately 600 KC.

Figure 4, is the identical display as reproduced by an oscillograph with amplifiers of 4 MC bandwidth. The larger diameter of the spots is caused primarily by the presence of some of the

subcarrier in the demodulated I and Q output signal circuits. The response of the amplifiers in the first instrument was sufficiently far down the subcarrier frequency -3.6 MC -to attenuate this residual. It is to be noted that the location of each spot can be determined equally well with the wide band or narrow band instrument. While the wide band instrument would be useful in the development of color equipment, it is felt that the greater cost makes its use prohibitive for station equipment.

Figure 5 is the scale which has been used on pilot run and engineering models of the Vectorgraph. It is a stationary illuminated clear plastic scale. The markings are standard tolerances specified for the chrominance signals. Accurate measurement of vector amplitude can be made by rotating the entire display until the spot in question falls on one of the calibrated axes. This rotation is accomplished electronically by rotation of the Vector Decoder phasing controls. In the case where the phase of one of the colors is off sufficiently to fall outside the tolerance box, the amount can be determined with a straight edge laid between the scale center and the compass rose.

Figure 6 shows idealized I and Q bandwidth curves. The encircled points are the F.C.C. specified points. As noted, the 3 DB point on the I characteristic is at 1.5 MC and the 3 DB point for the Q characteristic is at 500 KC. These limits have been imposed for reasons which cannot be covered in this paper. They are the primary factor which permits the use of an oscillograph with amplifiers of less than one megacycle bandwidth in this application. The oscillograms of the figure 1, were taken using instruments with more than 10 megacycles bandwidth hence it is fairly certain that the trace was a faithful reproduction of the original signal.

Figure 7, is a photograph of one of the rack

mounted instruments which were supplied to the Communications Products Division of Allen B. Du Mont Labs. These units had rotatable non-illuminated scales. Further field investigation has indicated that the rotation is not generally required, but that illumination is generally desired. Future Vectorgraph instruments will be supplied with non-rotating illuminated scales as standard unless otherwise specified.

General Purpose Usage

The characteristics of the Cathode-ray Vectorgraph have indicated that it may be useful as a general purpose oscillograph. The low relative phase shift of the amplifiers is of special interest. When the amplitude controls are set for equal sensitivity, relative phase shift is less than one degree, at least to 600 KC, and, in all probability, far beyond. For any combination of amplitude control settings, the relative phase shift will be less than 3 degrees below 600 KC.

The cathode-ray tube is the tight tolerance type 5AQP-. It is a flat face mono-accelerator type. Deflection plate alignment is held to within plus or minus one degree.

The instrument includes a modern highly linear vacuum tube sweep circuit. Incremental linearity is held to better than 10%. This circuit operates either driven or recurrent from the usual synchronizing sources.

Stability is assured by use of a Sola self-regulating transformer incorporated within the instrument.

Acknowledgements to R.T.Cavanagh and R.Deichert of the Du Mont Circuit Research Laboratory and J.Auld, W.Flynn and W.Triantefellow of the Du Mont Television Network for their assistance in preparing the photographs for this paper.

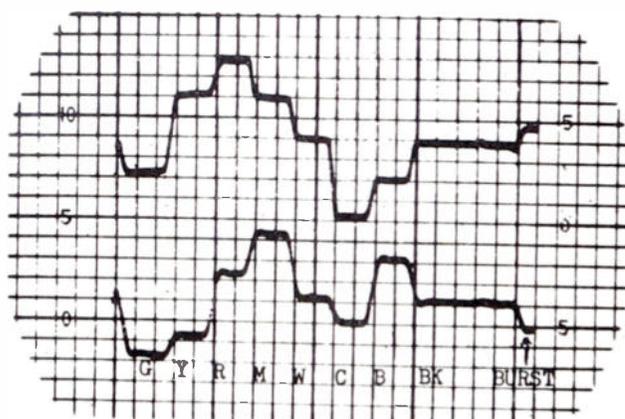


Fig. 1
I & Q Signals

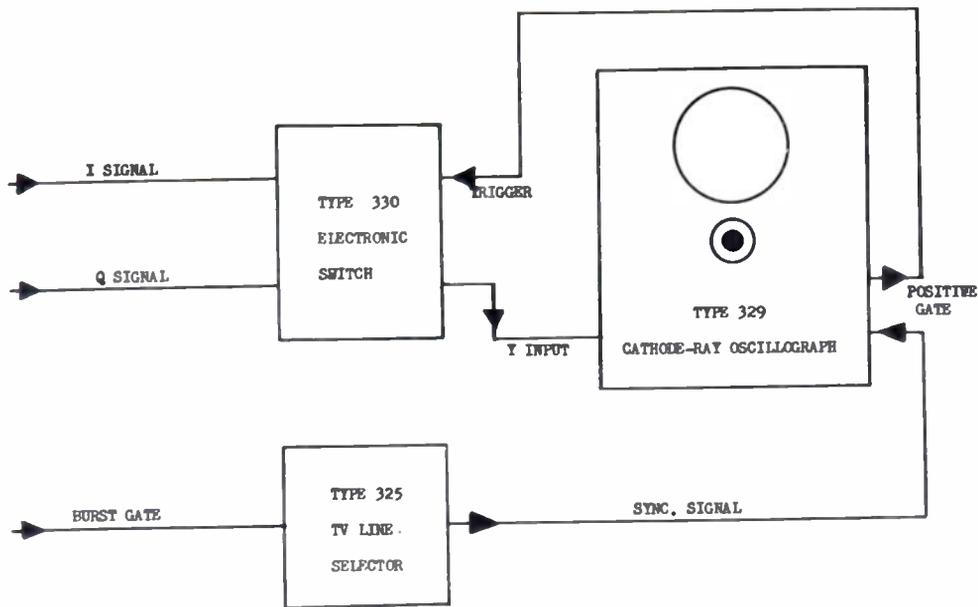


Fig. 2
(test set-up)



Fig. 3
Vectorgraph display
(600 KC Bandwidth)



Fig. 4
Vectorgraph display
(4 MC Bandwidth)

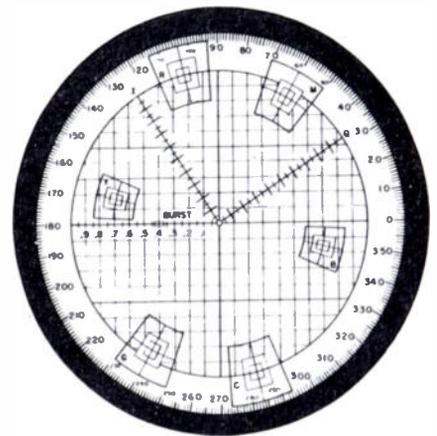


Fig. 5
Vectorgraph scale

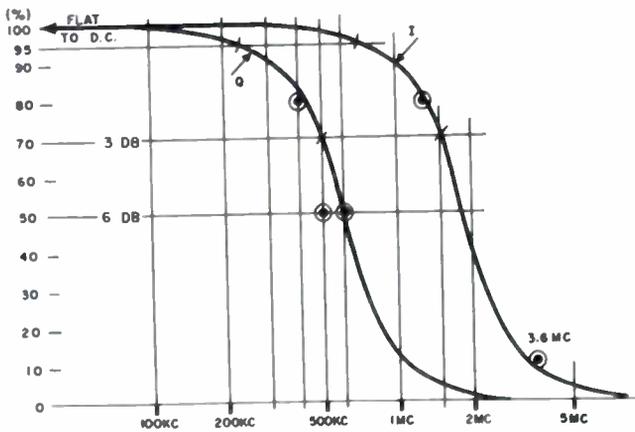


Fig. 6
I & Q Channels - Idealized Bandwidth Curves

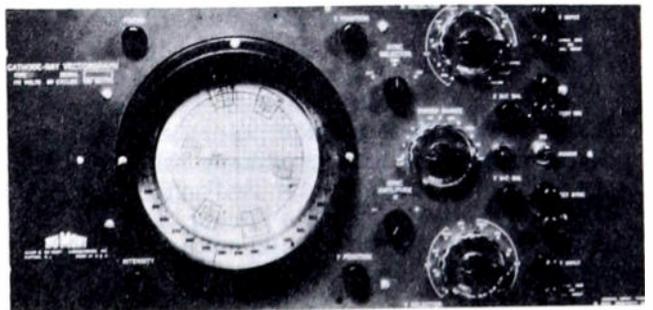


Fig. 7
Rack mounted Cathode-ray Vectorgraph

AUTOMATIC BALANCE CONTROL OF COLORPLEXERS IN COLOR TV

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Amityville, L. I., N. Y.

One of the most difficult problems with which the color television industry is faced is that of chroma drift in the modulator section of a colorplexer. The colorplexer is regarded as the "Heart" of the compatible color television system which was adopted by the NTSC and the FCC. It is that unit which combines the simultaneous color information from red, green and blue video into a signal which will operate both black and white as well as color receivers.

If a colorplexer becomes unbalanced, it causes a color receiver showing a color picture to color everything in the direction of the unbalance. A black and white receiver will also show subcarrier in the form of dots on a black and white picture when there should be none. If a black and white picture is received on a color receiver, with the colorplexer unbalanced, the color receiver will show this as a tinted picture instead of in black and white. If the burst which supplies color sync information becomes contaminated by unbalance conditions, the receiver color sync circuits may not properly lock at the correct phase.

The broadcasting industry in particular, needs a colorplexer or color video translating device which will be precise and yet give trouble free service over long periods of time. Present color broadcasting occupies a relatively small part of the daily broadcast schedule. Thus it is desirable that the encoding or colorplexing equipment which is used intermittently, reach operating equilibrium as quickly as possible.

But all colorplexers drift! Some more than others. The average colorplexer takes a minimum of one hour of warm-up time and generally several hours to stop drifting. Even so, it has been found desirable during the course of a day to "touch up" for nulls of the background chroma components.

At present, frequent manual rebalancing, poses a personnel problem in connection with the maintenance and adjustment of colorplexers. To solve this adjustment problem, an automatic balance control has been developed. This automatic balance control in its various forms is sufficiently dependable so that after its installation only normal preventative maintenance routine is required by the station personnel. In fact, one colorplexer in our laboratory has been running for more than nine months without any attention being required for its balance after it was equipped with automatic balance control. With automatic balance control, approximately 25 seconds from a cold start is all that is required to bring the colorplexer into balance, instead of allowing the usual one to two hours for warmup. The design of the automatic balance control is such that it may be used with any colorplexer with no changes required in the colorplexer itself.

In all types of colorplexers, the modulator color channels must be able to generate a chrominance vector anywhere through 360°.

Typical colorplexers resolve themselves into four or five categories:

The first type has been called the Quasi Balanced colorplexer. This consists of an I or Q or R-Y, B-Y modulator with two balance channels; one to eliminate the chroma unbalance in the plates of the modulators; the other a video bucking channel, to eliminate the video components in the modulator plate, leaving only the chroma components.

A second type, which is by far the most common, is the double balanced modulator using 6AS6 or other double input tubes. A pair of tubes is required for each modulator channel. In the double balance modulator type, both the video and chroma subcarrier information are in split

phase, 0° and 180° . See block diagram figure 1. Sometimes encoding is done at R-Y, B-Y or with I and Q axes and a separate modulator for generating burst. When R-Y, B-Y modulators are used, burst is derived from the B-Y modulator channel, since the burst is at minus B-Y phase.

A third type is the conventional four diode bridge switch type. A fourth uses two diodes. A fifth type uses 6BN6 tubes. In these types both video and reference subcarriers may be fed in single ended.

All types of colorplexers have common features. They must encode a pair of vectors, normally I and Q, or R-Y, B-Y, (the chroma information) with the carrier suppressed in the absence of chrominance information. We are primarily concerned with chroma unbalance in the balance modulators so our discussion is confined to that portion of the colorplexer. Excursions of video from the color difference channels to the modulators are positive as well as negative around zero or black level. It is, therefore, necessary to have some form of black level reference for the modulators. This is generally accomplished by Wendt type keyed clamp black level setters which hold the grids of the modulator tubes to a reference black. The clamp tubes operate so that pulse keying voltages close the clamps during the reference period which generally is during horizontal blanking time. The clamps are then released during the remainder of the scanning line. In this way, should the average value of dc on the modulator grids attempt to move from the reference black level, it is restored or held once each line. The d.c. voltage at the bottom of the clamp sets the operating point of the modulators. Unfortunately, black level setters have dc drifts in themselves. Some of these drifts may be traced to: varying contact potential of the diodes; uneven conduction of the diodes; varying or uneven amplitudes of clamp pulses, high resistance leakage in the pulse coupling condensers to the clamps, or high resistance leakage in the video coupling condensers in the clamp circuits.

In addition, all modulators including the diode types, have drifts in themselves. For example, in the double balance modulator type using

two 6AS6's, the two tubes which carry push-pull video and subcarrier must be capable of sustaining balance for both the subcarrier and the video signals for long periods of time. The modulating elements, however, do not have very good long term stability. Changes in line voltage, heater and ambient temperatures, changes in values of resistors, differential aging of the tubes, contact potentials, as well as gas currents - all are present more or less.

The automatic balance control developed for using colorplexer used for encoding signals work as follows: Reference to Block diagram, Figure 2 it is seen if either modulator of a colorplexer tends to become unbalanced for any reason, some combination of unbalanced chroma quadrature components will be present in the output of the colorplexer. A sample of these unbalanced chroma components in the composite signal is taken during a time when there is never supposed to be any chroma information present. This may be derived anywhere during the blanking interval.

The most convenient time to sample the signal is during the sync period since the back porch normally has burst on it. An amplified sample of the unbalanced chroma components is fed to a series of gated amplifier tubes and then to a pair of amplitude-phase discriminator circuits. Reference quadrature voltages are also fed to the amplitude-phased discriminators. The references may be derived in a number of ways, either from a source of subcarrier or from the colorplexer. The dc outputs of the discriminators, which compare the reference signals against the quadrature unbalanced components of the colorplexer output are proportional to the magnitude and phase of the unbalanced components present during the keyed sample interval of the composite color video signal. Since the discriminators are sensitive to both phase and amplitude changes, the error voltages from these detectors may be fed back to their appropriate keyed clamp black level setters as dc voltages. If the reference phases are correct, a drift of a keyed clamp or a modulator or any component in the circuit which upsets the chroma balance is restored by a corresponding dc voltage in the opposite direction applied to the keyed clamp.

Because of the high loop gain in the automatic balance control servo system nulls resulting from the error signals can be as much as 70 db down from peak to peak signal values. Although a. c. amplifiers are used, the 3.58 mc information acts as a carrier giving high stability to the amplifier section of the ABC unit without DC drift problems. Because there is negative feedback present during the closed loop operation of the automatic balance control, it becomes possible to considerably expand the tolerances allowable for the normal operation of colorplexers. Thus, drift in the keyed clamps themselves, resulting from differences in the conduction of the diode clamps, a change in the height of the keyed clamping pulses, drift in different directions due to contact potentials, or G_m of the modulator tubes, become second order effects. Widely dissimilar characteristic modulator tubes may be replaced without matching a condition which would ordinarily make a colorplexer impossible to use. Drifts due to ambient temperatures or line voltages or heater voltages may be ignored when the automatic balance control makes the colorplexer a very reliable performer instead of its former unsteady self.

Two types of automatic balance control units have been developed as well as a complete encoder or colorplexer with the automatic balance control built in as part of its normal functioning. In some cases, it is possible to derive the reference quadrature voltages directly from the colorplexer itself. In this case, the small automatic balance control adapter shown in slide (3) may be mounted on the colorplexer itself and the correction voltages fed back with all the circuitry as part of the colorplexer.

A second type, with a 360° coarse phase shifter for any phase 3.58 mc subcarrier, develops its own reference quadrature voltages with circuits which are tuned .707 in amplitude above and below 3.58 mc resonance, giving -45° phase shift. In some cases the automat-

ic balance control may be operated as much as 100 feet away from the colorplexer which it is governing. Such an independent unit is shown in slide (4). Where switching transients or loss of signal or sync occurs, suitable memory circuits in the form of long time constant filtering of the dc error voltage path may be used to allow the balance to be maintained during the transient condition. Similarly, anti-hunt networks may be incorporated in the error voltage channels in order to remove instabilities resulting from initial surges of voltage that may result when a colorplexer is first turned on.

In all designs, the automatic balance control unit may be disabled, allowing manual control or for rebalancing. A typical circuit is shown in slide (5).

V1 is the gated amplifier tube. V2 also functions as a gated amplifier tube in order to allow a high order of discrimination between desired chroma information during the sampled time of the signal, as opposed to the picture time where there is normally chroma information which might leak through and provide spurious information to this error detector. V3 is a driver tube which drives a bifilar transformer, the output of which is fed to V4 and V5, phase and amplitude detectors. Note that no limiting is used before the phase detectors in order to have the feedback error voltages be proportional to the amplitudes of unbalance as well as their phase. V6 feeds reference 3.58 mc in quadrature which is derived from a tapped 360° delay line. A block diagram of an integrally designed ABC controlled colorplexer is shown in slides (6) and (7), a photograph in slide (8).

As an operational tool, the automatic balance control represents a major step forward in the design of colorplexing equipment. It allows the user of a colorplexer to treat it like most of the monochrome equipment now used in Telecasting.

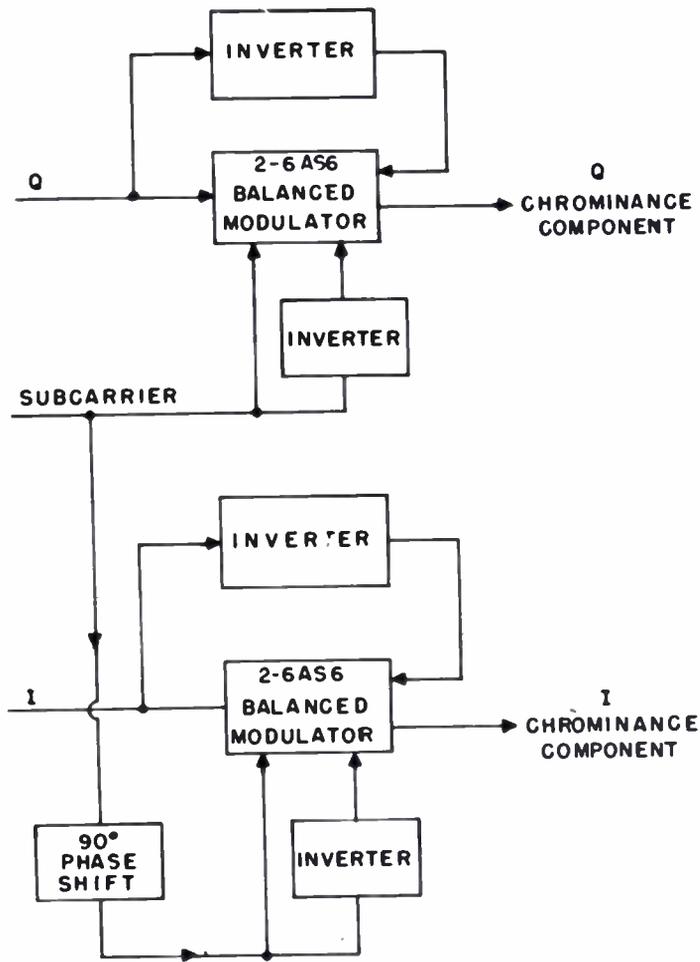


Fig. 1

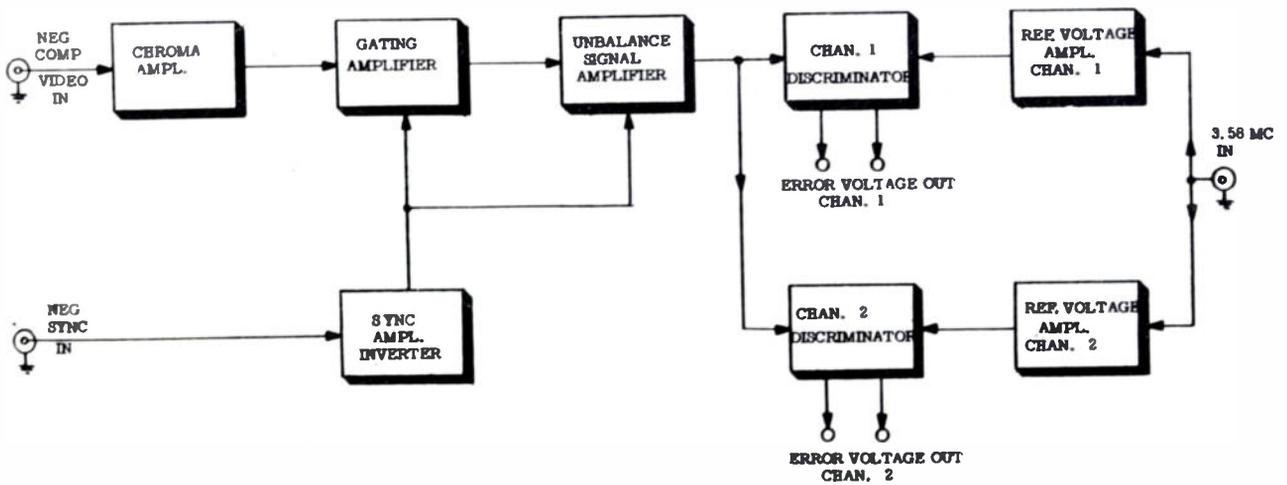


Fig. 2

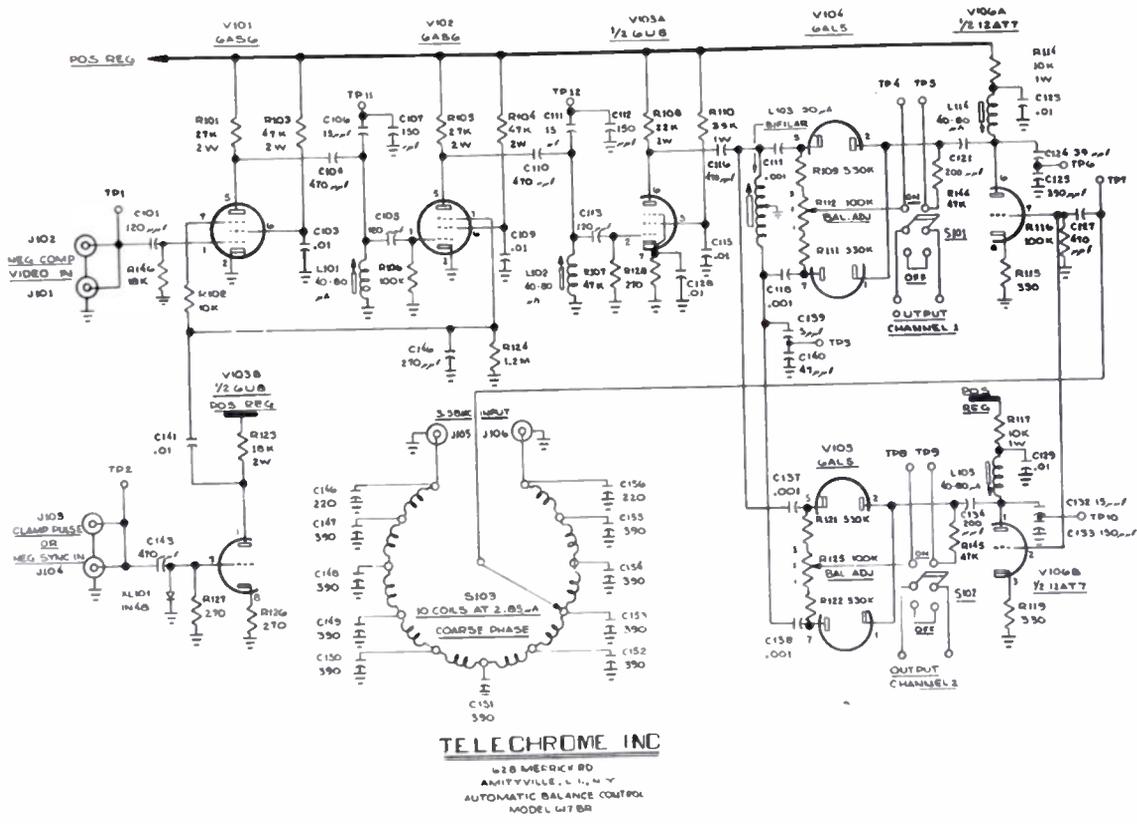


Fig. 5

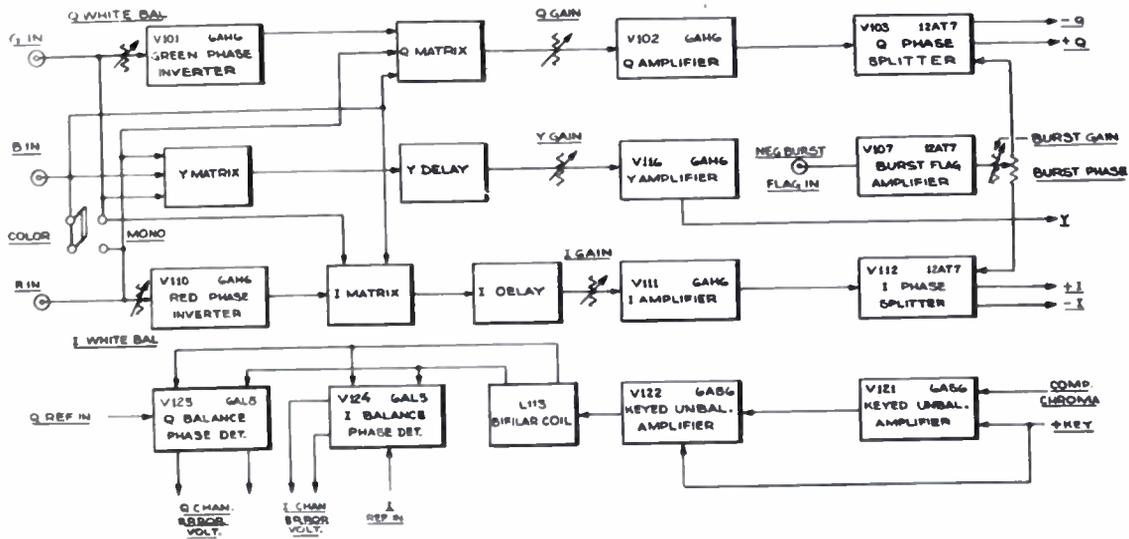
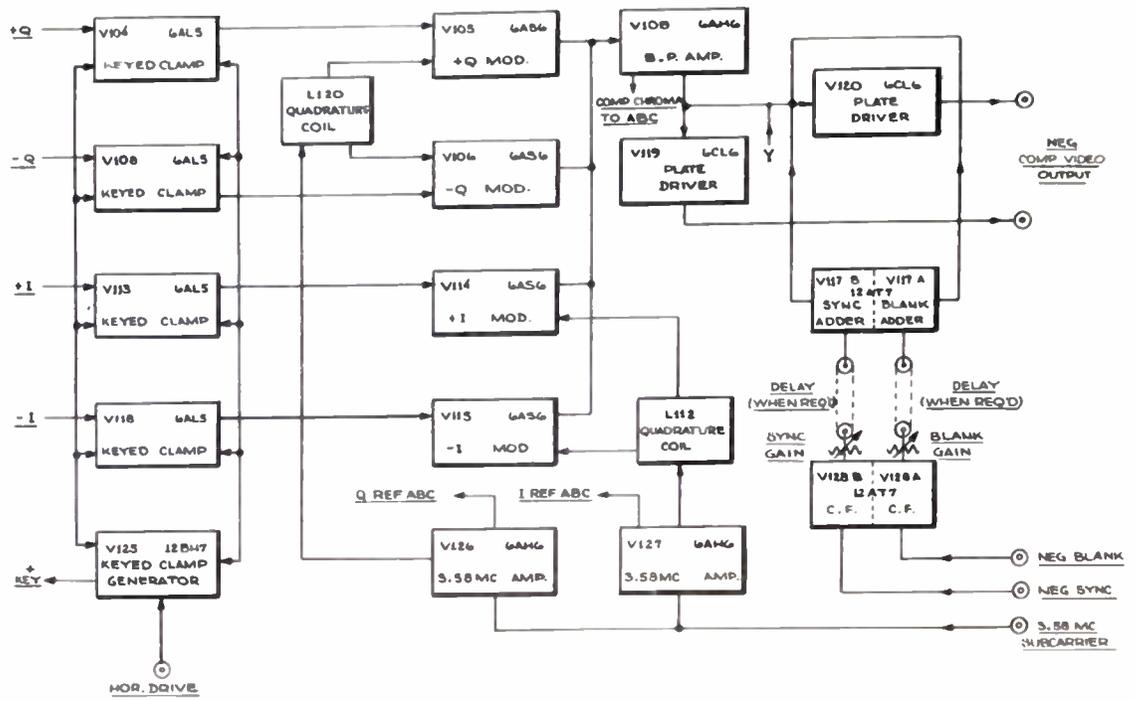


Fig. 6



TELECHROME, INC
 632 MERRICK RD.
 AMITYVILLE, L. I., N. Y.
 K.C. MOD AND ADDER SECTION
 MODEL 609R

Fig. 7

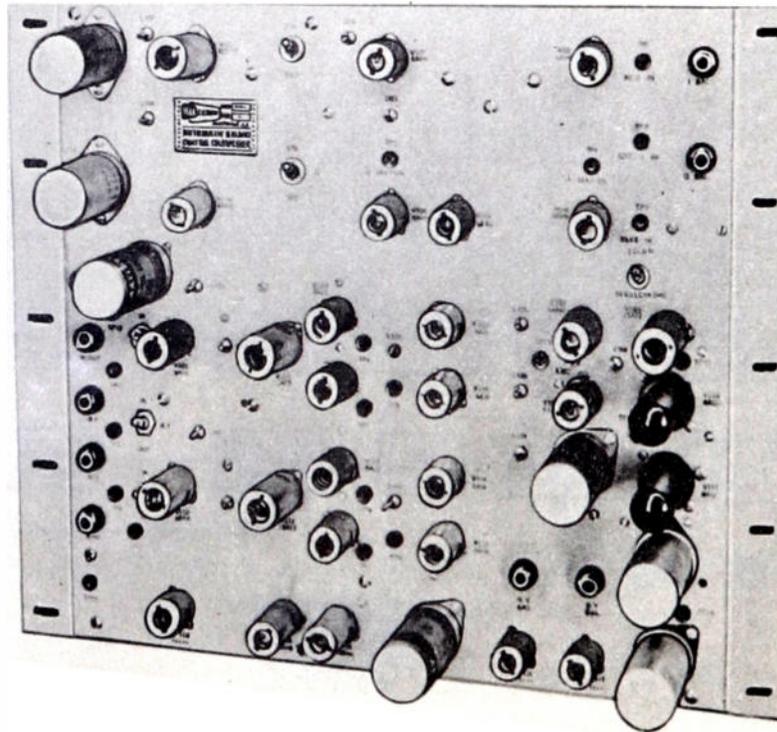


Fig. 8

TELEVISION IN EUROPE

H.A.S. Gibas

Garratt and Mumford have extensively reported about the history of television¹. A great number of inventions had to be accomplished before television could become the quality for public introduction. A part of these inventions come from Europe. Two well-known European inventors are Nipkow, who applied for a patent of a complete television system in Berlin already in the year 1884, and Baird, who contributed much to the development of television in Great Britain.

In Europe there exist different television standards. Great Britain opened the television service officially in the year 1936 with 405 scanning lines. This number of lines offers a good picture quality. With the high number of 819 the most scanning lines are used in France. Most other countries in Europe use the 625 lines standard.

The different standards are the source of several difficulties. In places lying in a zone where the reception of transmissions with two or more standards is possible, one needs receivers which are suitable for the reception of different standards. A further difficulty arises at the international program exchange. The standards of a television signal must be changed, when it is transmitted in countries with different standards. The program exchange between different countries in Europe is much practised. The rise of the number of viewers is especially great in times of extensive international program exchange.

Different television standards in Europe

The European standards differ in the first place in the number of lines. The principal standards are noted in the table below. The table contains, for comparison, also the American standard. One can see that the British standard has 405 lines, the French 819 and the Gerber standard, which is used by most European countries, 625 lines per image. The number of images per second is

25, for all standards in Europe. This is the reason, that the number of lines per second of the American and the Gerber standard differs by less than 1%. The 625 lines standard differs in several countries. In Belgium two standards are in use, the 625- and the 819-lines standard; but both differ from the standards in the neighbouring countries. The countries in Eastern Europe also use the 625 lines standard but with enlarged bandwidth. The conference of 1952 in Stockholm² had to solve a difficult problem. It had to allocate the channels of the television and FM-transmitters in the European zone. It concerned the following bands: Band I from 41 to 68 MC/sec, band II from 87,5 to 100 MC/sec and band III from 174 to 216 MC/sec. Band III was extended for France down to 162 MC/sec, and for several countries at the upper limit up to 223 MC/sec. The conference in Stockholm allocated the channels for 600 television transmitters. The allocation agreement in Stockholm was signed by the representatives of 21 countries. The European zone is formed by 31 countries. Portugal and the countries of Eastern Europe did not sign the convention. A proposition of Prof. Siforov of Russia, to cover Europe with a regular network of television transmitters, has not been accepted.

One could ask, what is the reason that so many different standards exist in Europe. Great Britain has had the 405 lines standard for 20 years. The results with this standard are satisfactory, and therefore Great Britain remained at this standard. After the war the communication authorities of the European countries investigated the possibilities of improving the picture quality of television reception. Different opinions about the requirements of good picture quality led to the different television standards. These different standards have already been discussed extensively and further discussion is useless. We in Europe are obliged to accept the fact of these different standards and to

	USA	British	Gerber	French
Lines per image	525	405	625	819
Images per second	30	25	25	25
Lines per second	15750	10125	15625	20475
Channel width MC/sec	6	5	7	13,15/14
Video modulation	Neg.	Pos.	Neg.	Pos.
Sound modulation	FM	AM	FM	AM

Table of the principal European television standards in comparison with the standard of the United States.

try to solve the problems which arise.

European television transmissions

Interest in television exists in all countries of Europe. The state of development of television is very different in these countries. In Great Britain a big network of television transmitters covers a large part of the country and reaches most of the population. The number of television receivers is about 4 million. In other countries commissions study the possibilities of introducing television. In most cases the obstacle in the way of a rapid introduction of television lies on the financial side. With a few exceptions there is no advertising at the broadcast transmissions in Europe. Many countries do not use advertisements at all, while others transmit only a few. The owners of television receivers have to pay a fee, which amounts to about 10 to 20 Dollars per year. This money is used to finance the programs and the upkeep and construction of television transmitters. Dr. Nestel of Germany indicated that at the end of the year 1954 the amount of 10 million dollars have been spent on transmitters, studios and the television network. At the same time there were about 100.000 viewers in Germany who could not as yet finance the German television expenditure. In other countries we find similar conditions, although the number of viewers is rising rapidly in countries with regular television programs.

Television in Belgium, Denmark, France, Germany, Great Britain, Italy, the Netherlands and Switzerland is in a state of good development. Their studios partly are technically and acoustically most modern. The hours of television transmissions in Europe are short in comparison with those in the United States. Great Britain is leading also in this respect, averaging 7 hours a day. But in other countries programs are on the air for only one or two hours a few evenings of the week.

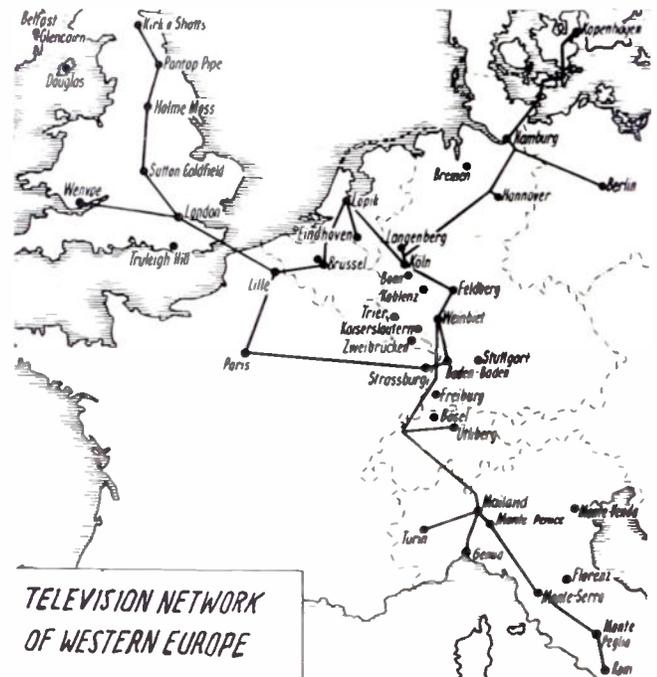
The possibilities of stratovision³ were investigated in Switzerland. The propagation of ultra short waves in mountains is difficult to foresee as reflections of mountains play an important role. The reflections also depend upon the growing of woods and plants on the hill sides. Therefore the propagation of waves and the coverage of television transmitters were investigated by a great number of measurements, in Switzerland as well as in other countries.

The quality of television programs is improving. The children's hour in Great Britain has special success. The children are very satisfied with their transmissions. It happened sometimes that the many television receivers working during the children's hour at sunset caused a shortage of electric current in some living districts in Great Britain.

Eurovision

The television networks of eight European countries are connected⁴. This network is shown in the figure below. The program exchange of Europe runs over this network. It is the beginning of the European network, which in later times can be connected with the networks of other continents. The international transmissions in Europe were named Eurovision. The international television transmissions are followed up with great interest; when a queen is crowned, the world championship of football is held, or other sporting or cultural events take place, than these can be seen on television screens in a great part of Europe. Plans are under consideration for the extension of these transmissions and programs. It is intended, during next summer, to transmit parts of the festival of Salzburg over the European television network. It is interesting to note that the number of television receivers sold is always large in times of international television program exchange.

The network of the Eurovision consists of microwave relays. The highest point of these microwave links is at the eastside of the Jungfrau in Switzerland at a height of about 12.000 feet. Here special provisions were necessary to protect the antenna system against snow and ice. The station lies high above the lower limit of glaciers in the Alps.



At the end of 1954 eight countries formed the television network Eurovision.

A particular problem in Europe is the conversion of the line number at the international program exchange. In the Eurovision such transformation points are in France and Holland. At these points the standards with 405, 625 and 819 lines are changed. The standard conversion was already investigated by Zworykin, Ramberg, Schröter and others⁵. In Europe we have the advantage that all standards are based on 25 images per second. The standard converter at the transformation point in Holland which was developed by Philips⁶ consists of a receiver and a television camera. The receiver produces a sharp image with long persistency. This receiver stands in front of a television camera, in which an image iconoscop is used. The difficulties, which are connected with this conversion method could be mastered with success. The loss of sharpness is small. The noise factor changes only a bit.

Part of the Eurovision network can work only in one direction; this means that a television program can run only in one direction. Some time is necessary to reverse the transmission direction. It is intended to provide the possibility of transmitting in both directions of the linkages. In Belgium a coordination and a control center of the European television exchange exist. The television companies of Europe are in close connection to make the television exchange more perfect and to study the possibilities of program exchange.

Television receivers

The development of the television receiver in Europe follows in principle the same line as in the United States. The dimensions of the pictures are smaller in Europe than in America. This is a question of the price and of the capability of production. The lower standard of living in many European countries demands low-priced receivers, this means receivers with a small picture tube. Nevertheless the largest picture tubes, which are produced in Europe in great quantities have 21". On the other hand the prices of television receivers show a tendency to decline, and the number of television receiver models with large picture tubes is rising.

The attempt of Philips to diminish the dimensions of a television receiver⁷ is interesting. The neck of the picture tube is bent and so the length of the tube is shortened. In this manner it is possible to build a receiver whose picture is 13" large, and which has a depth of 13".

Interesting problems arise in connection with the developments of receivers, which are suitable for the reception of transmitters with different television standards. In the zone in which a reception of Belgium, French and German transmitters is possible, the receiver must be able to receive four different standards⁸. In a special receiver this problem was solved in such a

manner, that the picture intermediate frequency amplifier has 38,9 MC/sec and a bandwidth of 4,25 MC/sec. So the French pictures cannot be received with the best possible quality. The sound intermediate frequency is 7 MC/sec and is produced in a second converter. The problem of the demodulation of AM- and FM-sound modulation is solved by a simple method. Positive and negative picture modulation is considered by applying the picture voltage either to the grid or to the cathode of the picture tube.

The determination of the intermediate frequency of television receivers has been investigated thoroughly. One is inclined in Europe to choose about 35 MC/sec for this frequency. The British Radio Equipment Manufacturers Association⁹ recommend a frequency of 34,65 MC/sec.

The possibilities of television in the moving motorcar were also investigated in Europe. Television receivers with projection tube and screen are seldom found for use in the home. The reason is certainly that picture tubes with large screen can be made easily to-day.

The Schmidt projector is used in the first place for the projection of large television pictures in cinemas¹⁰. This system is built in several cinemas in Europe. At Dr. Gretener in Zurich they are working on the very interesting Eidophor. The advantages of this cinema projector lie in the high light intensity and the large contrast of the picture. As in the United States also in Europe a great number of special measurement and service instruments for television purposes have been developed.

Industrial Television

In Europe many applications for industrial television exist. The underwater television was of good service in the search for the wreck of the British submarine "Affray" in 1951. The television camera identified it in a depth of 280 feet. Underwater television was also used to find the crashed Comet aircraft, which was lost near Elba in the spring of 1954. Television shows great advantages in medical education and at medical conventions. Televised surgical operations can be demonstrated in detail to a great number of spectators. There exist different types of television pick-up instruments in Europe for this special purpose. A television camera was also installed aboard a large whaling vessel in Great Britain. The captain of the ship can in this way observe exactly the happenings at the end of the ship. He need not leave the bridge, and can give orders easily and quickly.

Industrial laboratories in Europe worked on color television already before the last war. The war interrupted these developments. After the war we observed with great interest the development of color television in the United States and admired the great

effort which has been undertaken to come to an all-electronic compatible color television system¹¹. Pye demonstrated in 1949 color television according to the field-sequential system of Columbia. The picture quality was good and the demonstrations met with great interest and success at the public.

Emitron Television Limited, the producer of the well-known television pick-up tubes and Marconi are trying to find a way to simplify the television camera for color television. Marconi has developed a color television system following the N.T.S.C. technique, but for the British 405 lines standard. The British Broadcasting Corporation, E.M.I. and Marconi demonstrated in the year 1954 color television with success at several occasions¹². We suppose that the high price of color television, before all the high price of color television receivers, will prevent the introduction of color television in Europe in the near future.

This paper shows the advantage, if a big continent, as Northern America, has one television standard, and the difficulties which bring the differences of standards in neighbouring countries, as in Europe. On the other hand Europe is learning to master these difficulties. We suppose that in the near future the television networks of continents - for instance the networks of Northern America and Europe - will be linked together. At a later date there will certainly be one television network all over the world for the transmission of occurrences which are interesting for all nations. The standard conversion of the American 30 image standard and the European 25 image standards will become necessary. We are sure that the scientists will solve also this problem.

Numerous European technical journals report about the development of television in Europe. Some of these reports are listed in the Bibliography at the end of this paper¹³. Finishing the author is obliged to thank the many authorities, departments and firms in Europe, which have contributed to this report with technical material and information.

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ELECTRONIC ORGAN TONE RADIATION

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ABSTRACT

The principles of design for electronic organ tone chambers are outlined. The differences between the design goals for loudspeaker enclosures for organs and for other purposes are explained in fundamental terms. The construction of new organ tone cabinets for indirect radiation is described in detail. A few organ installation examples are given.

NOTE:

The complete manuscript of this paper, including the parts used in the condensed version presented at the Convention, has been published in the May-June 1955 issue of IRE TRANSACTIONS on Audio.

THE ROLE OF ROOM ACOUSTICS IN MUSIC LISTENING

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Summary

As sound travels from a source to a listener or to a microphone, its spectral and temporal characteristics are altered by the room through which the sound travels. In the case of recorded music, at least two rooms usually contribute to the quality of the sounds which the listener hears. The effects of the room on the music may be beneficial or detrimental to listening enjoyment. The evaluation of these physical effects in terms of listener preference is a subject of continuing interest to musicians, architects, and engineers concerned with design of concert halls and with the recording and broadcasting of music.

The topic of our symposium today is "Music, High Fidelity, and the Listener"; my remarks are directed at the role which room acoustics has to play in the transmission system which carries the music to the listener. I think you will agree with me at the outset that when we use the terms "high fidelity recording" and "high fidelity reproduction", we do not intend the words to be interpreted literally -- obviously we cannot wholly recreate, in a small apartment, say, everything that goes on in Carnegie Hall (even if we wanted to, and I know what my neighbors would say if I tried!).

Recording and reproducing a musical event is rather like taking a photograph - the photographer cannot possibly "reproduce" his subject in every detail, and he may choose to de-emphasize some features and emphasize others, intentionally introducing some kinds of distortion for purely aesthetic reasons. Distortions of this sort can greatly enhance the illusion which the photographer is trying to create, just as surely as his picture can be spoiled by undesirable distortions such as bad focus, coarse grained film, or the like. Indeed, it is these unpleasant effects that come immediately to mind when we

use the word "distortion" in the first place.

A music recording and reproducing system may well have electrical, mechanical, or magnetic elements which introduce spurious harmonic or in-harmonic tones, noise, and resonant phenomena, all of which seem to be disagreeable to a greater or lesser degree. Suppose for the sake of the present discussion that we assume that all such unpleasant distortions can be eliminated - or at least held within tolerable limits -- by careful application of existing audio engineering techniques such as quiet disk record materials, feed-back record cutters, magnetic recording, wide-range FM broadcasting, and the like -- in short, let us assume our recording and reproducing system is beyond reproach.

This perfect recording system is useless, however, until it is linked to a listener at one end and to a source of program material at the other --- and both of these links commonly involve rooms and room acoustics.

In recording music or in the production of various sound effects in motion pictures and radio, the recording engineer makes constructive use of the room characteristics in many familiar ways - a distant microphone may be used, for example, to add reverberation and make a small dead hall sound larger and the concert correspondingly more impressive - or the microphone may be brought in closer (or its directional response altered) to produce a more intimate effect, for a string quartet or a romantic dialog. As another example, suitable microphone placement has been shown to reduce the dynamic range required of a recording or broadcast system without any notably deleterious effect on the musical material.

Clearly, the effects produced by

a room can be manipulated to the advantage of the listener. But it is equally clear to all of us who have made recordings or attended concerts that rooms can also introduce unpleasant forms of distortion. In an effort to track down distortions of this type, many sorts of physical and analytic studies have been made, of rooms and in rooms --- reverberation time, decay modulations, transmission irregularity or frequency irregularity and space irregularity, pulse response, directional diffusivity, and many other physical concepts have been used to describe the properties of a room. But the interpretation of these physical measurements, in terms of what the listener likes or dislikes, is a more difficult problem, and one for which no wholly satisfactory solution has yet been found.

How then can we evaluate different acoustical situations -- how can we distinguish "good" rooms from "bad" rooms, or "good" locations from "bad" ones? As soon as we start using words like "good" and "bad", we imply that a listener is making an evaluation and rendering a judgement, so why not ask the listener directly?

It is not easy to get from a listener a reliable, consistent, and interpretable judgement, in the case of something so complex as what I shall call, for want of a better term, the "listening experience". The real meaning of this term "listening experience" was brought sharply home to me several years ago, when we first began broadcasting live concerts of symphonic music over wide-range FM in the Boston area. A friend of mine who is blessed (or cursed) with a "Golden Ear" was our friendliest but most vocal critic for many months. Even after the broadcast system had been painstakingly set up for a flat response of more than 15 KC (including the equalized condenser microphone used for single-point pickup) and a dynamic range of about 70 decibels above noise, for many months he always called me up the day after the broadcast, to register dissatisfaction about flaws in the broadcast the night before---occasional electrical interference, or slight imperfections in balance resulting from unexpected rearrangements in orchestra seating, and the like.

Then one day he failed to call. Fearing that something awful had happened to him, I called him. He

sounded hale and hearty -- yes, he'd heard the broadcast but, "you know", he said hesitantly, as if groping for words, "the music was so good I didn't have time to listen"!

This story gives us fair warning not to underestimate the psychological complexity of the overall "listening experience". More important, it gives us one positive hint as to what to expect if we ask a listener a direct question. We can expect to find some definite evidence of acoustic flaws, if these flaws are so aggressively unpleasant as to distract the listener's mind from the music, and consciously or sub-consciously interfere with his enjoyment of it. But if the flaws are not strong enough to over-ride his concentration on the music itself, he will probably be completely unaware of their existence, and therefore unable to give any meaningful information about them. I have seen an eminent composer and orchestrator listen with rapt attention and obvious enjoyment to a record in which the noise and distortion were almost painfully unpleasant to me, but he was so fascinated with the scoring of the fourth horn part that he was blissfully unaware of anything else.

These observations have been borne out in a number of studies which may be noted for reference purposes, even though a detailed discussion of the available evidence would be outside the scope of this paper. Helmut Haas¹ established for speech signals certain thresholds of "disturbance" produced by a single artificial echo as a function of echo level and time delay, although the time delays involved were not long enough for the listener to observe an actual discrete echo----- His results have been shown to be in good quantitative agreement with results published by Mason and Moir² on pulse response measurement and listener evaluation of motion picture theaters, and in still better agreement with measurements and questionnaire results obtained by Beranek, Bolt, and Doak³. In this latter instance, the key question asked of twenty-one listeners in a theater (during the playing of a standard SMPE test film) had no intended implication of pleasure or displeasure: it was "To what degree are you aware of the acoustics of the room?", and we had at that time no a priori evidence that such "awareness" need necessarily be construed as unpleasant. Yet subsequent questioning

of the listeners showed that all of them without exception interpreted "awareness of the acoustics" as meaning the intrusion of the room on their enjoyment of and concentration on the film.

It is of interest to note that Muncey and others in Australia⁴ have reported an extension of Haas's work to include string quartet music and electronic organ music as well as speech, using a magnetic tape recorder to produce a single echo in a free-field environment. The disturbance criteria for music were found to be somewhat less stringent than for speech. There was a hint that the complete absence of echoes was disturbing to the listeners in the music tests, but the test conditions were so different from any normal listening situation that this factor would be difficult to evaluate.

In contrast with the quantitative results which have been obtained in studies of speech and music under conditions where disturbing room effects are known to exist or are created artificially, a more extensive questionnaire test, carried out during a concert in a well-designed modern auditorium, failed to show anything but the overwhelming general satisfaction of the audience; evidently the enjoyment of the music wiped out any distractions which the room may have created.

In this brief review, we have noted the several parts that the room plays in music recording and in listening to music. In recording, the room is usefully employed for dramatic effects, but apparently the first approximation to good room acoustics is the absence of evident flaws; the second approximation, the good features of room acoustics which may actually enhance the enjoyment of listening, are apparently outside the capabilities of direct questionnaire techniques in their present stage of development. Quantitative measurement of these good features depends on a better understanding of the listening experience itself, as a psycho-physical phenomenon, and on a more sophisticated interpretation of the part which room acoustics plays in the interaction between the physical stimulus and the psychological responses of the listener.

Techniques other than the direct questionnaire have of course been widely explored in the evaluation of room

acoustics -- an outstanding example is the monumental investigation attendant upon the design and evaluation of the Royal Festival Hall in London. The report by Parkin et al⁵ is so comprehensive that I shall simply incorporate it by reference into this morning's proceedings. For subjective assessment of the properties of the Hall, there were test concerts, forums, and endless discussions and correspondence among an impressive assembly of musical and scientific experts, leading to a correspondingly variegated collection of descriptive terminology. In press comments alone, adjectives which appeared during the first eighteen months included the following:

"clarity (twenty times); brilliant (five times); beautiful, blend, definition, resonant (all four times each); alive, balanced, exhilarating, frightening, perfect, ruthless, shattering, subtle, superb, truthful, warm (all twice each); admirable, astonishing, charitable, charming, Chaucerian, cold, consolidated, dazzling, deplorable, eerie, exalting, exquisite, frank, full, glorious, hard, ice-clear, inspiring, intimate, lovely, magnificent, magnified, mellow, merciless, muddy, overbearing, ravishing, responsive, revealing, rich, ruinous, sharp, shimmering, shrill, sickening, sonorous, stunning, temperamental, toneless, touchy, tremendous, unrelenting, wonderful (all once each)."

The authors themselves note that this list is less than half as long as one published by Salmon⁶ to describe music from recordings. And they note later: "One difficulty with subjective assessments is the use by musicians of a large number of terms, and to bring order into the problem it has been necessary to translate some of their opinions into our own terms".

This is a succinct statement of the problem which seems to me to be next in line when it comes to improved methods for subjective evaluation of concert hall acoustics, recording systems, or other physical components which play a part in the overall psycho-physical interaction which I have referred to as a "listening experience".

Words constitute our most versatile technique for the description and classification of our sensory observa-

tions, and extremely sophisticated vocabularies have been developed to describe the characteristics of sounds, especially among practising musicians and hi-fi enthusiasts. Verbal scaling techniques are already being applied to the problem of establishing quantitative relationships in certain special applications⁷. Unless we have recourse to electrodes applied to the scalp or inserted in the auditory system, we shall almost certainly continue to stumble over the interpretations of words until we can put these interpretations on a quantitative basis. This is the principle obstacle which limits the usefulness of all our existing techniques for the evaluation of room acoustics and other components in music systems.

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ENVIRONMENTAL-FITNESS CONSIDERATIONS OF HIGH-FIDELITY AUDIO SYSTEMS

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Summary: Performance evaluations become progressively more difficult as the field of interest expands from isolated audio components to integrated systems and as rigorously objective measurements either become impracticable or are superseded by wholly subjective aural judgments. Yet if the final evaluation of overall sound qualities must be determined by aesthetic rather than engineering criteria, the latter still are significantly applicable to the fitness of any specific system to its specific environment. The present paper calls for extended and intensified study of present-day listeners' practical audio needs and psychological attitudes, and — anticipating the results of such study — suggests possible engineering approaches to the better "matching" of home high-fidelity systems to their actual "habitats."

Introduction

The subject of complete sound-reproducing systems, in its broadest aspects, is one of which most audio engineers are justifiably wary. Technical studies generally are restricted to professional studio installations or to specialized details of component integration, and the free-ranging discussion of home systems is left to merchandisers, semi-professionals, and amateurs — few of whom are particularly inhibited by any lack of reliable engineering data. From a narrow point of view this looks suspiciously like licensing fools to rush in where angels fear to tread, but, in larger perspective, it is both wiser and kinder to recognize that once an audio system has been installed in a layman's living room, what happens then pertains largely to sound-reproduction as an art rather than as a science.

Yet few conscientious engineers can abandon their creations to alien hands without a strong sense of disappointment and perhaps even of guilt. The higher they have set their own standards of performance, the more aggravating it becomes to find how far short of these standards most systems fall in actual home use — or misuse. However we may define "high fidelity," we all know that some kind of ideal in reproduced-sound quality is currently the subject of tremendous public interest, and that nowadays this ideal can be more closely approached by legitimately rated wide-range audio systems (as well as by the best recordings and broadcasts) than ever was possible in the past. Yet the gap between our high-fidelity ideal and our common sound-reproduction home practice grows steadily wider.

As one of the "fools who rush in" (in my case as an audio critic — i. e., an unofficial, self-appointed intermediary between engineer and listener), I don't pretend to have any pat solution

to a problem which patently is so complex that it well may be inherently insoluble. But at least I am sufficiently uninhibited to suggest a possible avenue of engineering approach to that problem. And I hope you won't be too shocked by my basing that approach on the unaccustomed, seemingly but not necessarily unscientific, notion of environmental-fitness considerations.

Terminology and Qualifications

Although the "environment" of an audio system usually is thought of (at least by engineers) in terms of the acoustical characteristics of the room in which the system is located, I'd like to employ it here in a larger sense which also takes into account the nature of the program materials, the types and habitual settings of the operating controls, and the response characteristics of listeners' aural sensibilities. The system's "environmental-fitness" then is determined by its degree of "suitability" not only to the listening room alone, but to all the various demands that are made on it. And obviously that suitability is measurable only in part by any technical means: in larger part it can be evaluated only in terms of "user-satisfaction."

Unfortunately, such unavoidably loose terminology is easily misinterpreted, and so I must emphasize immediately that I have no intention of raising the specters of "listener-preferences" or of implying that audio-performance standards be scaled down so as to match the lowest common denominator of popular tastes. On the contrary, I hold that tastes of any kind always are too personal, too subject to change, and too inexpressible ever to serve as guides — except on wild-goose chases! When I speak of user-satisfaction, I have in mind the basic human need for musical experience and the usually intuitive and inarticulate, but nevertheless very definite, ability of individuals to indicate by their reactions whether or not this need is rewardingly met. And since I believe that this need never can be fully satisfied (certainly not for normal sensibilities and over extended periods of time) by anything short of the very best that is possible today in sound-reproduction, my concept of environmental-fitness need not involve any compromise in the highest of fidelity standards, however these may be formulated.

To me, the goal is not one of designing and using home systems which will please their listeners (a relatively easy task on any short-term basis), but of designing and using those which will do full justice to the finest recordings and broadcasts available today, and with which their owners long can live happily.

Research Needs, Subjects, Locales, and Procedures

Now, I'm keenly aware that a discussion of environmental-fitness considerations raises many awkward questions which audio engineers hardly can be expected to answer. But surely it is not unreasonable to hope that they should at least lead the way in helping to formulate the most pertinently significant queries. For, as George A. Sarton¹ has shrewdly noted, "In science immense progress is made whenever the right question is asked, the asking in proper form is almost half of the solution." Yet it is right here that the audio profession as a whole tends to abdicate its present responsibilities.

Why, for example, are there so few contemporary re-examinations of the great basic studies of auditory perception, audible frequency and amplitude ranges of music, residence noise levels, etc. — most of which date from a decade or even two decades ago? Undoubtedly their fundamental conclusions are not likely to be altered radically, but surely they stand in need of extension and modification, if only as a result of the recent tremendous expansion in sound-reproduction, the vast increase in listening experience, the new popularity of symphonic (often highly "modern") music, and the consequent notably enhanced audio concern with both frequency-spectrum and loudness-level extremes.

The study of room (as distinct from auditorium) acoustics gradually is being intensified, but as yet there are far too few papers, such as those by Kessler and Harris on this afternoon's symposium program, which are of directly practicable value to the home listener. And it is even more rarely that investigators are directly concerned, as Rosenblith is today, with the overall response characteristics of the human listener himself — the most complex transducer of all those with which we have to deal.

Indeed the most promising subject of new research well may be the contemporary "audiophile." For, although he is a relatively new phenomenon, at least in the numbers in which he exists nowadays, we now have access to an ever-growing fund of raw data on his nature, attitudes, practices, and native "habitats": both in the direct evidence of his comments, complaints, and queries published in the correspondence columns of the popular audio and record-reviewing press; and in the indirect evidence of that same press's editorials and articles, plus of course many recent books and pamphlets dealing with various aspects, technical and otherwise, of the "hi-fi craze."

Within more familiar engineering territory, one particularly happy example of the kind of investigation needed so badly today is W. R. Thurston's paper, *Testing and Adjusting Speaker Installations with the Sound-Survey Meter*², which I cite not merely for the value of its specific contents, but for the fact that all its data are based on overall sound-output measurements made under normal (living-room rather than laboratory

or anechoic-chamber) conditions. And, moreover, Thurston's procedure impresses me as setting a superb example to follow by other engineers willing to experiment on the handiest of all guinea-pigs — themselves and their own home systems. There are few audio engineers who are not also home listeners and presumably hi-fi fans too (at least to some degree short of the fanaticism of the more notorious amateur representatives of that breed!). So where can you better pursue the subject of "fitness" than with the suitability of your own system to your own environment?

Yours is an exceptional case, you protest? Perhaps, but all individual cases are exceptional in some respects and, from what I have seen and heard in engineer-friends' homes, I am strongly inclined to doubt that their typical environmental-fitness problems differ greatly from those of other audiophiles — or that engineers, in general, are notably more successful in solving these problems than many serious amateurs.

Nor should this be surprising. Moving from his laboratory to his living room, the engineer naturally tends to leave behind his objectivity along with the other tools of his profession. His greatest advantage over the technically untrained listener is his ability to analyze overall system response in terms of specific component-performance characteristics. Too often, at home, he may not make good (or indeed any) use of that advantage, but its very possession is likely to blind him to the fact that any clear analysis of this sort normally is impossible for the layman.

The latter, especially if he has had considerable listening experience, generally knows well enough when something is wrong with what he hears, but he seldom can — and as a rule makes no attempt to — discover where the fault lies: in his program materials, in one or more of his system components, in his own and his speaker locations in relation to each other and to his room's geometry, or in his own ears and mind.... Then, too, the lay listener's most decisive reactions probably are even more negative in character than those of most professionals. Both types of listener are more easily displeased than pleased, but the layman tends to be more acutely conscious of minor defects, and his annoyance over these flaws often is disproportionately prejudicial to his system as a whole.

Yet despite such differences, and indeed despite all the diversity of contemporary listeners' individual backgrounds, temperaments, equipment, and locales, I'm convinced that those most seriously interested in high-fidelity sound share many aural attitudes and ideals in common, and experience much the same difficulty in matching their particular systems to their particular needs. And risky though it may be to anticipate the results of the more formal studies for which I am calling here, my own observations point to three general areas in which environmental-fitness investigations seem especially needed and where they well may hold the promise of most immediate rewards.

Area No. 1

The first is, not surprisingly, further tactical improvements in system-components' functional suitability. I say "further," because this area is already well known and considerable progress currently is being made here, notably in the enhancement of home-equipment's visual appeal and use-convenience, and in the closer adherence to "professional" constructional and operational standards. But much remains to be done — perhaps particularly in effecting greater design economy, eliminating superfluous versatility, reducing essential controls to a minimum, and ensuring optimum "fool-proofing" throughout.

In theory, at least, many desiderata here are clearly enough recognized by engineers, but as yet the general public still has to be convinced that these are not merely ideal but quite practicable and indispensable. In actual merchandising and promotional practice such desiderata too often are given lip-service at best — and in consequence many true high-fidelity essentials are considered by laymen to be "refinements" or "elaborations" which only fanatical specialists really want or can afford. The safest examples to cite here are diamond pickup styli and high-quality turntables and pickup arms (rather than changers), but for myself I'd also add horn-loaded dual or multiple speaker systems, studio-type calibrated attenuators, and VU and FM-tuning meters....

I hesitate to touch on the more controversial details of system-control and loudspeaker problems, but just because so many controversies still rage in these domains, it must be obvious that their effective resolution demands far less heated and more illuminating investigations than we have had so far. Let me call attention only to the desirability of reconciling two currently conflicting schools of thought: one (to which I subscribe) claiming that the seeming need for "tone" and "loudness" controls is a sure sign that something is basically wrong or inadequate in the system itself; the other asserting that such controls are necessary for room-matching and quality-compensation purposes. Similarly, the former school finds a properly designed and located horn-loaded speaker system quite free from objectionable point-source effects; the latter yearns for multiple dispersed speakers, special sound-diffusion devices, or even quasi-stereophonic effects. Can either point of view be authoritatively justified — and how?

Area No. 2

Fortunately, the second area is less disputable, at least where ends rather than means are concerned, and possibly it may eventually supply some of the answers needed in Area No. 1. For it is the development of more effective overall system-performance test or "check" methods — to be conducted under normal rather than laboratory conditions, and as far as possible utilizing signal sources more closely akin to actual musical materials than steady-state pure tones.

Formidable as the difficulties here may seem, I'm convinced that some means must — and can — be found to overcome them. One encouraging sign is that we are gradually coming to be better supplied with appropriate signal-source materials. Besides many useful discs and tapes of steady-state or spectrum-swept frequencies, there are a few for Di-distortion testing, of which the Emory Cook N-A Beam Test disc³ is perhaps outstanding in that it permits immediate aural checks on the test results. Then there is the valuable Thermal Noise disc³ (also from Cook), in which switched comparisons between wide-range white noise and various types of restricted-range gray noise also may be evaluated, if less precisely, by ear. And best of all there now is available a wide variety of "demonstration" discs (both those specifically designed for hi-fi displays and those easily selected from the regular LP symphonic repertoires) with which audio systems may be subjected to the most exacting musical-performance checks.

With these last there is unfortunately no present way of making meaningful (instrumental) response measurements — yet under suitable conditions, and with properly qualified listeners, aural judgments can be made with a considerable degree of quasi-objectivity. I outlined a possible method in my "Engineering Listening" editorial a couple of years ago, but of course similar methods long have been more-or-less consciously employed by many critical listeners, and they well may be susceptible of further development and far wider application.

At any rate, a — if not the — prime criterion of high-fidelity reproduction is directly involved with a system's overall transient response and spectrum balance, and in my opinion one of the most pressing engineering needs still unmet today is that for better ways of making such response and balance measurements or evaluations. Perhaps what we really want is some kind of aural equivalent of a TV-broadcast test-pattern, with which John Q. Public himself can obtain a quick, accurate, and unambiguous answer to his vital question, "Is the overall performance of my audio system good, bad, or indifferent?" Surely some such ideal home-test means is not wholly impossible — but how much longer must we wait for some ingenious inventor to make it a reality?

Area No. 3

The third area is characterized neither by basic differences of opinion nor by any special technical difficulties — which makes its neglect in current practice all the more inexplicable.... For it is simply better protection against common performance deteriorations, the main sources of those noise, distortion, and unbalance troubles which are most readily recognized and always most strongly resented by lay listeners.

What shocks me most about nearly all the audio systems I hear in friends' homes is not so much their bad sound, as such, as it is their almost invariably sounding so much worse than they should

if they were in proper operating condition. Even with the best and most expensive of hi-fi systems, after they have been in home use for a few months, how often can you crank up the level control all the way and (with turntable or tape-transport running, but with no record playing) put your ear close to the speaker and hear no hum or rumble but only a smooth high-frequency hiss? How often can you play a modern wide-range recording and hear no evidence of motor unsteadiness or incorrect speed, none of stylus or head wear or misalignment, none of overloading, and none of unbalance between or among the various elements of a multiple speaker installation?

Yet it is exactly such defects, above all the rise of background noise, which are the first to be resented by even untrained listeners, and certainly it is the presence of distortion and spectrum unbalance which contributes most heavily to aural fatigue and consequent dissatisfaction with the system as a whole.

Unhappily, most home-system owners never have even heard of preventive maintenance: only a complete breakdown forces them to seek a serviceman — and then their chances of finding a competent audio specialist are slim indeed. Right here is where the whole audio business has fallen down disgracefully: designers and manufacturers for not meeting higher standards of ruggedness; dealers and servicemen for minimizing the need — or failing to supply the means — of regular expert maintenance; and my own profession as well, for audio critics and commentators have been as lax as others in failing to alert audiophiles to the early recognition of deterioration symptoms and the ways in which at least many simple faults may be corrected or emergency replacements be made.

Conclusion

Probably I am unduly biased by my personal orientation, but no matter from what direction I try to approach the problems of audio-system environmental-fitness I find the greatest barrier always is a lack of knowledge — on the part of professionals no less than on that of amateurs. J. Robert Oppenheimer⁵ diagnosed the general ailment of our not-so-golden age of technology when he sadly concluded, "It isn't the layman that's ignorant — it's everybody that's ignorant!"

As long as high-fidelity audio-system practice was confined to a narrow circle of recorders, broadcasters, and other specialists, its environ-

mental-fitness considerations could be reasonably well understood and controlled. Nowadays, however, comparatively large numbers of technically untrained laymen are forced to wrestle — for the most part blindly and helplessly — with far more recalcitrant home-system fitness problems. And one thing is sure: the home listener, no less than nature, "abhors a vacuum" and will continue to flounder toward "easy" pragmatic solutions until he is shown some better and more rational answers. High fidelity originally was an engineering ideal; today it has been taken over by a public which has perhaps more than its normal share of crackpots — and whatever responsibilities are abdicated by the engineers, the crackpots will be only too avid to seize.

The supreme challenge to audio profession and public alike is simply this: given the technical resources and skills available today, and given the incalculable active and potential interest in high-quality sound reproduction, why aren't high-fidelity home systems more common, more dependable, and far more rewardingly used? Probably there are many good reasons, but the best one I know is that the pertinent environmental-fitness considerations have yet to be fully explored, clearly grasped, and effectively acted upon.

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CAN A HIGH-FIDELITY SYSTEM BE DEFINED

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Summary

As yet, there has not been laid down a satisfactory operational definition for a high-fidelity system--one that depends only on specified measurements on the system in question. If our definition is to have significance, it must rank order systems in a way that will correlate with subjective evaluation. Thus an operational definition of a high-fidelity system can only be meaningful if it includes a rating of all variables that have a significant affect on the subjective evaluation of the system. For convenience, various possible factors that may be of importance have been divided into the following groups:

1. Response-frequency distortion
2. Non-linear distortion
3. Transient distortion
4. Phase distortion
5. Spatial distortion
6. Background noise level
7. Difference in level between original and reproduced sound

The problem is (a) to specify the variables in each group, and (b) to determine their relative importance by listener-preference tests. Various listener-preference tests are discussed. It is concluded that we do not have sufficient data at the present time to give a quantitative evaluation of all types of distortion that may be of importance to the subjective evaluation of a high-fidelity system. Therefore, it is not possible at this time, to give an operational definition of a high-fidelity system.

Much confusion has arisen in the field of the high quality reproduction of sound because the term high-fidelity system does not mean the same thing to each of us--the reason being that, as yet, there has not been laid down a satisfactory operational definition for such a system, one that depends only on specified measurements on the system in question. The definition of such a term is arbitrary in nature. But this does not mean that we can define, with significance, a high-fidelity system in any manner we choose.

If our definition is to have significance, it must rank order systems in a way that will correlate with subjective evaluation. Here the word system is used in the broad sense to include all components that influence the acoustic reproduction in one location of a sound source that has been generated in another. It includes, therefore, characteristics not only of the electroacoustic equipment, but of the source and listening rooms as well.

Suppose that a definition of a high-fidelity system is selected and that it is applied to a group of 10 systems to determine which of the group qualify as high-fidelity systems. If, for example, the definition classes 6 of the systems as high-fidelity, but listener-preference tests indicate that 3 of the non-qualifying group actually are judged to be of higher quality than those we have classed as being high fidelity, then it is apparent that the arbitrary definition was not well chosen. The criteria of rank order correlation with subjective judgements would thus rule out a definition that one hi-fi fan has suggested: "A high-fidelity system is one whose electrical components cost more than \$99.50."

Thus, an operational definition of a high-fidelity system can only be meaningful if it includes a rating of all variables that have a significant affect on the subjective evaluation of the system. The question is what are these variables and how important are they? If we can provide this answer, we can give a satisfactory operational definition of the term high-fidelity system. Here, then, is the objective of this paper. For convenience, various possible factors that may be of importance have been divided into the following groups:

1. Response-frequency distortion
2. Non-linear distortion
3. Transient distortion
4. Phase distortion
5. Spatial distortion
6. Background noise level
7. Difference in level between original and reproduced sound

Now the problem is (a) to specify the variables in each group, and (b) to determine their relative importance by listener-

preference tests. Unfortunately we have very little such material to discuss, for if one separates mere opinion, or opinion backed by meaningless listening tests, from carefully controlled experimentation, it turns out that the number of papers of importance in this regard are few indeed.

Greatest attention has been given to the first item of the above group, response-frequency distortion. Among the factors that might be considered here are:

- (1) General requirements as to frequency range.
- (2) Balance between upper and lower cut-off frequencies.
- (3) Rate of "roll-off" at the cut-off frequencies

Two studies are of particular importance in this connection. The nature of their experiments and their results will be stated briefly.

In 1945, Chinn and Eisenberg published a paper on a frequency-range preference study conducted with various groups of listeners.¹ The following is abstracted from their paper:

"Almost 500 subjects, in small groups, took part in the tests and all together, over 10,000 individual preferences were indicated. In addition to the "average" listeners, tests were undertaken with a group of professional musicians whose training presumably qualified them as critical listeners, and with a group of frequency-modulation listeners. A wide variety of program material, including popular, light-classical, and classical music, male and female vocals, and male and female spoken and dramatic speech, was presented at three tonal ranges. These were arbitrarily designated as narrow, medium, and wide. (The wide range was said to be essentially flat from 40 - 10,000 cps. The medium and narrow ranges were obtained by inserting filters into the electrical system having the following approximate passband characteristics: 70 to 7,000 cps and 150 to 4,500 cps respectively.) The trend was consistent throughout the investigation. Except for one series of tests that were made with live talent, all voice and music selections were produced from especially original recorded "masters" cut on cellulose-nitrate coated disks. The background noise, even during the reproduction of the records, was not detectable by the majority of the listeners. The measured distortion of the electrical portion of the system was extremely low throughout the frequency range. The loudspeaker unit was a dual unit of well-known manufacture, employing a folded horn for the low frequencies and a multicellular horn for the high

frequencies. The band-pass filters had the characteristics representative of the conditions that generally prevail in radio and recording equipment."

The main conclusions of the study are:

Listeners prefer either a narrow or medium tonal range to a wide one. However, the exact choice of bandwidth varies to some extent, within these limits, for different types of program content. Most listeners still prefer a narrow to a wide tonal range even when informed that one condition is "low-fidelity" and the other is "high-fidelity." No great difference in preferences were found between groups of different sex, age, education, and musical training; even professional musicians and frequency-modulation listeners having the same preferences.

An interesting study of another type was carried out by Olson who investigated listener preference for a restricted frequency range using a small live orchestra as a source². The listeners always heard the live orchestra--no electrical reproducing equipment was involved. In order to compare the full frequency range with a restricted one having an upper cut-off of 5,000 cps, an acoustical low-pass filter was inserted between the orchestra and listeners. The orchestra and filter were hidden by an illuminated shear cloth curtain. The results of tests, involving about 1,000 listeners, indicated a preponderant preference for the full frequency range. From this, one may conclude that if a restricted range is preferred in some cases, it is not because listeners have become conditioned to a restricted range and prefer it for that reason, nor because present musical instruments are improperly designed and would be more pleasing if higher frequency overtones were suppressed.

The above frequency range preference studies are not necessarily in conflict, they simply test two different situations. For example, there was considerable difference in the rate at which their response frequency curves dropped at cut-off. Until we have enough knowledge about this subject to enable us to extrapolate our conclusions from one listening situation to another, then each such study can be of help in contributing to our own understanding of the relative importance of frequency range in high fidelity for a specific condition. Thus it would be incorrect to assume that the extrapolated results of Chinn and Eisenberg necessarily apply to a situation where the wide-band condition has a 15,000 cps bandwidth; they may not necessarily apply to a situation if the distortion were an order of magnitude lower than was possible to obtain 10 years ago. Their results

certainly cannot be extrapolated to stereophonic reproduction, although some engineers have attempted to do this. On the other hand, it would be equally incorrect to assume that the results of Olson, taken with a live source, also apply a wide band single-channel reproducing system using transcription material as a source.

The above two studies differ in another very important respect--what we have called spatial distortion, that is, the difference between the original and reproduced sound that arises primarily as a result of differences in the acoustic radiation pattern of the source in the two cases. For example, consider a hypothetical single channel system that is perfect in every respect save one: At a specified distance from the loudspeaker in an acoustic free-field, the radiated sound pressure is uniform within an angle of 30° from the axis of the loudspeaker; beyond this angle the acoustic output is negligible. Now suppose a violin were used as a sound source. Even if the loudspeaker of the hypothetical system is placed in the same location in the room as the violin, it will not produce the same acoustic pattern in the room since its acoustic output would be confined within an angle of 30° in contrast to that of the violin which has a complex radiation pattern. As a result, spatial distortion is introduced by the reproducing system. Furthermore, the ratio of direct-to-reflected sound reaching the ears of a listener in the room would be distorted. Work of Maxfield and Albersheim seem to indicate that there are preferred limits for this ratio³. With reverberation time, it enters into a quantity they defined as liveness. In the example we have just considered with a violin as a source, the liveness at a fixed point in the room would be altered by the distorted acoustic radiation pattern.

So far, we have discussed a single instrument. If one considers a distributed source such as an orchestra, the resulting spatial distortion is even greater. Then the listener no longer has the ability to discriminate direction. This type of distortion can be decreased by increasing the number of reproducing channels. As the number becomes very large, one may approach true stereophonic conditions. This is essentially the situation that Olson tested since he used a live orchestra. The appeal of stereophonic reproduction is due to its significant reduction of this type of distortion. Early work in this field includes research by members of the Bell Telephone Laboratories, especially Fletcher, Steinberg, Wentz and Snow, who analyzed the subject of stereophonic sound in considerable detail^{4,5}. The renewed interest in this type of sound reproduction seems to

justify further experimentation in this direction.

Continuing with our discussion of factors included under the general heading of response-frequency distortion we now consider the question of so-called "balance." It has been stated and accepted for many years that the product of the low and high frequency limits of a reproducing system should equal a constant. Some Engineers use the value 500,000 while others prefer 600,000. Still others feel sure that they will be reasonably safe if they pick a value halfway between these. The origin of this concept is somewhat obscure--most engineers are not sure where the figures first arose. A little investigation has indicated that this concept originated in the broadcast industry more than twenty years ago, based partly on extrapolation of tests with pure tones and partly on the application of data of Snow for another type of experiment^{6,7}. Apparently no controlled listener-preference tests have been published to indicate the validity of the above rule. In fact, this relationship does not appear to be true--unless, of course, the per cent tolerance is extended to ridiculous limits, in which case it has no meaning. For example in one chart, which has been circulated widely, a shaded area for the product of the upper and lower cut-off frequencies is recommended for engineering purposes; it is stated that satisfactory aural balance will be obtained within this area. In this chart it will be noted that for an upper cut-off frequency of 8,000 cps a product as low as 320,000 is acceptable; in contrast, at an upper cut-off frequency of 15,000 a product as high as 1,200,000 is considered acceptable. Here then is a criteria many design engineers will appreciate--plenty of tolerance, about 100 per cent.

As indicated above, no controlled listener-preference tests have been reported in the literature on the subject of "balance". The only published results seem to be those which appear in a footnote of Olson's paper in which he says², "Some investigators have attributed considerable importance to balance in reproduced sound. Correct balance is said to obtain when the product of the upper and lower limits of the frequency range is 500,000 (cycles)². In order to obtain approximately the same balance for the restricted high frequency range condition, the frequency components below 100 cycles were eliminated or attenuated. A comparison of this condition with merely attenuating the high frequency range showed a greater listener preference for the latter condition. This is in spite of the fact that the latter condition is said by some critics to be improperly balanced. Under

certain conditions, there appear to be other factors which influence the balance, besides the arbitrary value for the product of the upper and lower limits of the frequency range."

Thus we may conclude a discussion of "balance" by indicating that there appears to be no evidence from listener-preference tests to substantiate the validity of applying the above balancing procedure to high-fidelity systems.

Another possible type of distortion is that which arises as a result of the difference in level between the original and reproduced sound. Since the overtone structures of musical instruments vary with level, reproduction of music at levels different than the original may introduce an unnaturalness. Furthermore, it may be accompanied by an undesirable reduction in dynamic range.

Somerville and Brownless of the British Broadcasting Company carried out a series of tests to determine listeners preferences for maximum sound levels for various types of music⁸. They set up a listening room in one of their studios so that it would have acoustical characteristics similar to those of the average living room. Recorded material was used as a source. Their results indicate that the preferred maximum sound levels decrease with age for men and women and are about the same for both; it is lowest for speech, higher for dance music, still higher for light music, and highest for symphonic music--the range from lowest to highest being on the average, about 6 db. The range of their data is about the same as Chinn and Eisenberg's, however, the preferred British levels were perhaps 5 db higher. This and other significant difference may result from a difference in listener preference between the American and British groups or from differences in the noise level of their program material, differences in overall transmission characteristics of the reproducing systems, or differences in non linear distortion. Thus, the types of distortion in our original list do not contain independent

variables. A change in one will affect a listeners preference with respect to another. For example, if the upper frequency limit of a high-fidelity system is increased then the amount of non-linear distortion that will be considered acceptable will be reduced. Therefore, in reporting listener-preference tests it is important that complete test conditions be stated so that the results may be compared with other studies.

Although the results of tests on all possible types of distortion have not been outlined here, it has been shown that we do not have sufficient data at the present time to give a quantitative evaluation of all types of distortion that may be of importance to the subjective evaluation of a high-fidelity system. Therefore, one concludes that it is not possible at this time to give an operational definition of a high-fidelity system.

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MAN, A SOMEWHAT NEGLECTED COMPONENT OF HI-FI SYSTEMS

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ABSTRACT

In recent years much progress has been made in assessing the transmission efficiency of communication systems. In most situations that are of interest to the hi-fi enthusiast, it is however not possible to specify the message that is to be transmitted. Under these circumstances, one might suggest that the most realistic yard-

stick for the performance of hi-fi systems is man's discriminative ability. This talk will, in the main, deal with man's hearing. It will also be concerned with the question: "To what extent do laboratory experiments on pure tones predict human reactions in more general listening situations?"

MAGNETIC TAPE AS A RECORDING MEDIUM

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There has been a steady improvement in the characteristics of magnetic recording tape over the past six years, partly as the result of the availability of new materials, and partly from more effective formulation with older ingredients. Since the performance of a magnetic recorder is, at least in part, dependent on the characteristics of the tape used, recorder results have improved in the following directions:

1. Base material of improved strength has permitted reliable operation under hitherto difficult conditions of humidity and temperature; and has encouraged use of a thinner base, allowing 50% more tape footage and recording time for a reel of given size. For the thinner base, it appears, that cellulose acetate is inadequate under summer conditions, and that only the new polyester Mylar should be generally suitable.

2. Improved oxide and the use of an orientation technique have produced:

- A. An 11 db increase in signal to noise ratio.
- B. A 15 db increase in signal to DC noise ratio.
- C. A 6 db increase in saturation output.
- D. A 31% decrease in erase current requirement.
- E. A profound improvement in relative response at the higher frequencies.

3. Improved coating and quality control techniques have improved uniformity of performance from one foot to the next and from one reel to another, to the degree indicated by the following:

	Tape	
	1949	1955
Uniformity at 7.5 mil wavelength:	2 db	0.25 db
Uniformity at 1 mil wavelength:	4 db	1.0 db

4. Binder formulation controls the coefficient of friction and the coating adhesion of tape. Early tapes were prone to poor physical properties, due to the lack of data on aging effects and inadequate knowledge of accelerated tests.

Adhesion to the base material was helped enormously by a method of presizing the base before the magnetic coating was applied. This opened new fields to the formulation chemist, and the binder was developed to incorporate new anti-tack compounds, anti-friction agents, and higher concentrations of magnetic oxide. Operation in hot humid climates without layer to layer sticking became possible, and friction was profoundly reduced.

5. The growing use of magnetic tape read-in and readout, and for multi-track instrumentation recording, has compelled intensive attention to the avoidance of physical defects on the surface since these lead to errors. It has been found that rolling and scraping techniques smear manufactured defects, over the tape surface rather than remove them entirely. Accordingly, there has been intensive attention to the production of tape entirely free from defects. The methods are indicated by the following production test results, secured with the defect counter described by Kramer:

No. of reels tested:	1200
Reel size	1250 x 1/4"
No. of permanent defects per reel:	.05

Permanent defects are those which are not removable, and exclude dust particles laying loosely on the surface, etc. Accordingly, it has proven practicable to supply tape which is guaranteed free from defects. However, the user must protect this quality by protecting the reel from dust at all times. Storage must be in the manufacturer's special dust-free packing.

MAGNETIC RECORDER-REPRODUCER DESIGN
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When magnetic recording was introduced into general use, it was first applied to sound recording. Since that time, the uses have become very widespread, but the basic principles which apply to sound recording are used in all fields. For use in the field most familiar to radio engineers, the design of a magnetic recorder must meet rather stringent requirements. These requirements are: (1) Excellence of performance, and (2) Reliability. The first of these is made up of a number of characteristics, such as frequency response, signal-to-noise ratio, distortion, flutter, ease of handling and threading, starting time, interchangeability of tapes, et cetera. Reliability, the second major requirement, is very important to the radio engineer since lost time, servicing, and general inconvenience are so expensive to the radio industry. Some of the factors which result in reliability are: basic design, conservatism of component ratings, ease of servicing, gentleness of tape handling.

All of the foregoing factors are important in the design of a magnetic Recorder-Reproducer for the radio engineer, and are covered in some detail in this paper.

Magnetic Recorder-Reproducer Design

In the design of a tape recorder-reproducer, the development program may conveniently be divided between the mechanical and the electronic components; the requirements relating to each may be conveniently divided into those which have to do with reliability, and those which have to do with performance.

Reliability

Reliability, as it relates to tape recorder design, may be considered in two ways. Short term reliability may be thought of as a function of outright component failure, such as burn-out of motors, failure of insulation in transformers or inductors, failure of tube filaments, and so forth. Such failures can only be minimized by design which minimizes the stresses, whether mechanical or electrical, on all components, and by careful choice of appropriate component elements. The components should never be operated at a large percentage of their maximum rating at any time, and assurance should be obtained that the ratings themselves are reliable. 600 volt capacitors, for example, would appropriately be used throughout equipment in which peak voltages do not exceed 400. Transformers should be rated to operate at considerably higher temperature rise than their operation will demand. This consideration applies, also, to motors, and similarly to all other components.

Long-Term Reliability

Long-term reliability factors are those concerned with failures which occur after extended periods and which are due to what may be termed normal wear and tear. Among these are the wear-out of heads due to tape abrasion, which may reasonably be expected to occur after the passage of approximately fifty

million feet of tape. Fast-forward and re-wind operations, contrary to initial supposition, need not be expected to contribute materially to head wear-out, even though the tape is not withdrawn from the heads during these high speed operations. Indeed, experience has established that head faces may not even be expected to become polished excepting by the passage of tape at normal velocities. This is due to the formation of an air film between tape and heads during high speed tape motion. The only reason, then, to design head structures so that tape may be pulled readily from the heads would be for convenience in editing operations. Heads should, of course, be designed, if possible, so that performance does not deteriorate in any way until the very end of normal useful life. Gradual deterioration due to abrasion may also be expected in such guides as are necessary to conduct the tape accurately past the heads. These should be of hard, abrasion-resistant material. Capstan wear will be rather rapid if the oxide surface of the tape contacts it. It is imperative, therefore, that the tape operate with the oxide away from the metal capstan. If this procedure is observed, capstan wear-out will eventually be due to bearing failure in the capstan assembly, and no gradual change in the speed of the tape will occur throughout normal life. Elsewhere in the transport mechanism, sleeve or ball bearings may be chosen for motors, idlers, guides, etc., and these may require attention before the machine exceeds its useful life.

The electronic components will require more attention on the part of the user than will the mechanical apparatus. The gradual, almost predictable deterioration of electrolytic capacitors and vacuum tubes may reasonably require of the user scheduled periodic replacements. Tubular paper capacitors and resistors are usually not subject to predictable long-term failure, and with the present quality of available components rarely need be a factor in reliability. Parks

reports that machines designed, in general, as described in the foregoing, have performed in network service, without back-up duplicate recorders, for more than 8,000 consecutive hours without sign of failure. Simple and suitable preventive maintenance procedures are followed¹.

Performance

We may conveniently separate performance considerations into those relating to the transport mechanism, and those relating to the electronics.

Transport Mechanism

Satisfactory performance of a tape recorder transport mechanism involves ease of handling and smoothness of tape motion. Appropriateness of size and weight for the application, and appropriateness of control functions are factors in ease of handling. A facility for remote control, for example, may be needed in machines designed for recording studio and radio broadcast use, whereas simpler lever controls are more appropriate for use in recorders whose application demands extreme portability, lightness, and compactness. Both types of machines find important uses in broadcast service. In broadcast operations, of course, it is extremely important that the machine be easily threaded in a minimum of time, since the equipment is often set up for operation during short time periods, such as station breaks, and since the whole nature of any broadcast operation is split-second timing. For this reason, also, starting time of the mechanism should be held to a minimum. Threading, indeed, should be so straightforward as to make it difficult if not impossible to thread the machine incorrectly. Push-button controls are most desirable where remote operation may be needed.

Among the design details which contribute to handling ease are those which have to do with the prevention of improper operation of the equipment, such as interlocking relays which prevent accidental erasure or breakage of the tape. These must be included in the design of any transport mechanism which is to be used in broadcast operations. In the event the machine is lever-controlled, it must be made as difficult as possible to damage the tape during operation of the machine, by the use of appropriate mechanical interlocks.

The second consideration in tape recorder performance, that of smooth tape motion, depends upon a number of interrelated details. Smoothness of tape motion is measured by flutter and wow content. To understand how these effects may be minimized, let us first consider a typical tape transport, as shown in Figure 1. The supply reel on the left feeds tape to a stabilizing idler assembly, from which the tape enters the head assembly. After leaving the heads, the tape passes through the capstan and is fed to the takeup reel. Figure 2 shows the

electrical equivalent of this mechanical system. The voltages produced by the components shown as V_S , V_C and V_T should be first considered. These represent the rotating fields within each of the three motors which drive, respectively, the reels and the capstan. V_T represents a voltage which corresponds to the constant velocity of the tape. If we consider those parameters which affect the value of V_T we will be considering those variables which will alter the velocity of the tape, producing flutter and wow.

Note, at the left, that V_S supplies current to the circuit through R_S , which represents the variations in torque created by irregularities in the torque motor, due to imperfect design and construction of the motor itself. C_S represents the combined inertia of the torque motor rotor, reel, and the tape on the supply side. LT_1 , LT_2 , LT_3 , LT_4 , and LT_5 represent the compliance of the various lengths of tape between the supply reel and C_I , which is the stabilizing idler assembly pulley, the erase head (represented by R_{H1}), the record head (R_{H2}), the playback head (R_{H3}), and the capstan, C_C . C_C represents the inertia of the flywheel on the capstan. CT_1 through CT_4 represent the mass of those lengths of tape between the elements from the stabilizing idler to the capstan. L_M represents the compliance of the magnetic field between the rotor and the stator of the synchronous capstan motor. C_T and R_T correspond to C_S and R_S on the supply side. LT_6 is the compliance of the tape between capstan and takeup reel. C_G and I_A represent the spring-loaded tape guides associated with the stabilizing idler. R_I represents the frictional loss in the ball bearings of this assembly. R_{H1} through R_{H3} represent the variable frictions between the head surfaces and the tape. R_C represents variable losses due to non-uniformity of the rubber in the capstan idler wheel.

The equivalent network thus set up could be considered, first, as to the way it responds to low frequency disturbances of various kinds, and, second, as to its response to high frequency disturbances, since the various kinds of flutter seem to divide themselves in approximately this manner. Insofar as tape motion is concerned, that portion of the tape located between C_I and C_C is the only important section. The most important consideration before us, when we refer to the electrical equivalent circuit, is that the values of V_{T1} , V_{T2} , and V_{T3} remain as nearly constant as possible. Let us consider, then, what sorts of disturbance would produce a voltage drop across LT_3 , LT_4 and LT_5 at a slow rate. Any change in the current flowing through these three circuit elements would cause a momentary change in V_{T1} through V_{T3} , and, correspondingly, momentary changes in tape velocity at the heads. If we assume for the moment that V_T originates in a source of indefinitely low impedance, so that changes in current will not affect its value, then the only changes in current through LT_3 , LT_4 and LT_5 at low frequencies will be due to changes in the values of R_I , R_S , and V_S . Changes at a low rate in R_I would be due to ball bearing

imperfections, such as runout, misalignment, or stiff lubricant. Changes in R_S would be those in torque due to design and manufacturing defects in the torque motor, such as "cogging," rotor and stator eccentricities, poor bearings, or bad winding distribution. The reel hub eccentricities are also considered as a part of R_S .

It becomes obvious from a study of the equivalent circuit that the greater the value of C_I with relation to R_I and R_S , the less will be the affect on V_T of these variables. The value of C_I for practical considerations should be only sufficiently high to attenuate changes in current which occur at a greater frequency than approximately one-half cycle, when considered as a portion of an RC filter, R being R_S and R_I in parallel.

High frequency disturbances from R_S and R_I would then be negligible.

Throughout the foregoing, it has been assumed that the impedance of the source of V_T is zero; this is not entirely attainable. In the practical case, the capstan, whose surface velocity is represented by V_T , is derived from a synchronous motor which has a compliance between the rotating field and the rotor, represented by L_M . Also associated with this motor, either mounted directly on the motor shaft or after speed reduction, is a flywheel, C_C . This combination is a resonant circuit of finite Q. Hence, changes in current through L_M will change the value of V_T , and if these disturbances occur at a rate approximating the natural resonance of C_C and L_M , the effects may be quite serious. One of the commonest causes of trouble in this portion of the circuit is R_C , which is the rubber capstan idler. Variations in rubber uniformity can cause serious changes in current (force) in L_M . The current changes in L_M due to the variations in R_C are normally many times greater than the total variations in current through components L_{T1} through L_{T5} , so that these latter variations, as regards slow frequency variations in V_T , may be neglected. For the same reason changes in V_T caused by current changes through L_{T6} due to changes in R_T may be neglected. It may be seen that the maintenance of a high degree of motional stability at low frequencies requires that the design be directed toward making the value of C_C as high as possible, L_{T3} through L_{T5} and L_M as small as possible, making R_S and R_I as constant as possible, L_{T2} as large as practical from the point of view of the size of the equipment, and, at the same time, making C_I of such value that the resonant frequency between it and its associated inductances is in a frequency range where no other periodic disturbances may be expected to occur. With suitable care in the selection of the above constants, the low frequency variations in V_{T2} and V_{T3} may be held to values as low as 0.03% r.m.s., on a commercial basis.

One of the reasons we have considered low

frequency flutter first is that it is much more important for an audio recorder, since the human ear and mind detect periodic speed variations in a manner which is much affected by the rate of flutter, and by the frequency of the tone being varied². See Figure 3. It will be obvious on examining these curves that flutter rates between 1 and 5 cycles per second are most easily detected, and therefore most rigorously to be excluded in the design of the transport mechanism. Whenever possible, shafts, idlers and other rotating components should be designed to turn at frequencies outside this region. It turns out in most practical designs, however, that rotational rates must often fall entirely within this region, so the components must be given extremely careful attention as regards perfection of bearings, eccentricities, etc.

From the considerations presented thus far, it might appear that it is always necessary to include a component corresponding to C_I . This is not, however, the case. C_I can be eliminated if the variations in R_S and V_S are greatly reduced. This can rather readily be accomplished by using a hold-back device of very smooth torque instead. Such a device may take several forms. A spring-controlled slip clutch, or a mechanical feedback type of constant brake would be appropriate. This is an especially desirable procedure where considerations of size and weight are of great importance, as in airborne or in broadcast portable recorders. Years of experience with quality torque motors have made clear that even the finest of these cannot satisfactorily be used without a precise and carefully designed, and probably necessarily expensive stabilizing idler. Constant tape tension throughout the reel is not necessary, however, for excellent performance. It is only necessary that tape tension be sufficient to provide adequate tape to head contact, and low enough to minimize head wear. Obviously such criteria allow for wide variations in average tension without degradation of performance.

For low frequency speed variations which are not audible, which might be referred to as "drift," the consequences are variations in program timing and minute variations from perfect pitch in the reproduction. To prevent these, it is essential that a synchronous capstan motor be used, and that when a speed reduction device is inserted between the capstan motor and the capstan shaft, no low frequency speed error is thereby introduced. Another consideration is the capstan shaft diameter, itself, which must be maintained within extraordinarily tight limits from machine to machine. In practice, the variations in tape velocity must be held within 0.2%, which is to say that the capstan diameter must be $\pm 0.1\%$ of design center, from machine to machine. If we allow the capstan to contribute half the allowable error, the tolerance on capstan diameter becomes 0.05%, independent of capstan size.

A consideration of the causes for high frequency flutter will center about the relations

among L_{T2} , L_{T3} , L_{T4} , L_{T5} , C_{T1} , C_{T2} , C_{T3} , C_{T4} , and R_{H1} , R_{H2} , and R_{H3} , it may be seen that the values of L and C which represent the tape between the stabilizing idler and the capstan form a resonant circuit which for diagrammatic convenience is shown as distributed elements. The resonant frequency of this length of tape depends upon the elasticity of the tape, its mass, and the distance between C_I and C_C . R_{H1} through R_{H3} (which represent variable frictional resistances between heads and tape) will tend to excite the tape into oscillatory motion in the vicinity of its natural resonance. This type of speed variation occurs generally at a frequency which is rarely detected by the ear when its magnitude is minimized by careful attention to the value of R_{H1} through R_{H3} . This means that a low coefficient of friction must be maintained, and that low pressures must be applied at each area of contact. One of the best ways to obtain uniform and low friction is to polish the head surfaces highly, and to obtain the necessary tape-to-head pressures, by smoothly "wrapping" the tape around the heads, rather than by providing pressure pads, or by the use of tortuous paths, such as are necessary when the heads are arranged in one straight line.

Whenever design considerations dictate a relation between L_M and C_C so that resonance is objectionable, it is necessary that one or the other of these elements be changed in value so as to move the resonant frequency or else the elements must be suitably damped.

If due consideration is given to all of the foregoing mechanical details, the resulting transport will have excellent reliability and performance.

Electronics

Consideration of the electronic portions of a recorder-reproducer may well begin with a discussion of head structures. Several elements must be considered simultaneously in the design of a high performance head assembly. Most important are those details which affect frequency response, signal-to-noise ratio, and useful life. Within the audio range, the uppermost frequency response of a magnetic playback head is primarily determined by its gap size as related to the velocity of the tape. In the design of the playback head, it is important that the size of the gap be small in comparison with the shortest wave length to be recorded and reproduced. Experience shows that in audio applications a gap of 0.00025 inches is a practical minimum, and yet allows satisfactory performance at 0.5 mil wave lengths (for example, response to 15 kilocycles at $7\frac{1}{2}$ ips). Figure 4 illustrates the interrelations of frequency response, signal-to-noise ratio, and useful life. A reduction in gap size, in the interest of increased frequency range, will either reduce useful output, and hence reduce signal-to-noise ratio, if the gap depth is held constant, or it

will reduce useful life if the gap depth is reduced so that additional turns may be added to compensate for the reduction in useful output. A reasonable choice among these parameters can result in a head which is responsive to half-mil wavelengths, with sufficient output for 60 decibels of signal-to-noise ratio at 15 ips tape velocity, and for a useful life of 50 million feet of tape. Thus, it may be seen when the design of the head is changed in order to favor frequency response over one of the other two related parameters, one or both of the others necessarily will suffer.

Signal-to-noise ratio as so far considered is that ratio which is imposed by the input noise level of the playback amplifier. If, however, shielding of the head assembly is inadequate, the hum level from the playback head in relation to the desired signal may well be the limiting factor on signal-to-noise ratio. In a machine designed for good mechanical characteristics it is extremely difficult to keep stray hum fields, originating in motors, solenoids, relays, etc., to an insignificant value. Therefore, it is almost always necessary to have a minimum of two mu-metal enclosures for the playback head, and preferably also for the record head. With carefully designed shielding of this type, it is possible to locate the capstan motor as close to the head assembly as may be desirable for mechanical reasons.

Record Heads

The problems associated with record heads are not nearly so exacting as those relating to playback heads, but these considerations do conflict. For a record head, by all odds the best choice of gap size appears to be in the vicinity of one-mil. A smaller gap provides inadequate bias penetration, and consequently distortion and irregular low-frequency response. A larger gap results in loss of definition and consequent degradation of high-frequency response. The information recorded on the tape is mainly a function of the field to which the tape is exposed immediately prior to leaving the gap area. The tape "remembers" only what it last "saw." A one-mil gap, then, may satisfactorily record half-mil wavelengths, although the playback gap must be smaller than this, for satisfactory response. These differences indicate the desirability of using separate record and playback heads.

Erase Heads

The erase head must be capable of developing sufficient flux density in the area occupied by the tape so that complete erasure of a previous saturated signal will result. Flux densities exceeding those necessary for saturation should, therefore, be provided. Complete erasure depends not only on the flux density, but also on the number of polarity reversals experienced by the tape in moving through the gap. This dictates either a large gap or a high erase frequency or

both. Since the bias frequency should be at least five times higher than the highest intelligence frequency, these requirements are compatible with a bias and erase frequency of approximately 100 kilocycles. This should be supplied by an oscillator of high power. In order that high intensity high-frequency fields be produced without overheating the erase head, the design of the head should be of high efficiency. The Western Electric double gap construction considerably improves erase head efficiency, and is recommended.

Signal-to-Noise Ratio

Assuming that the signal-to-noise ratio of the system will not be limited by hum output from the heads or by amplifier noise, the signal-to-noise ratio will ultimately be limited by the characteristics of the tape used. The output from a playback head of good life and good frequency response is low enough that such common expedients as well filtered DC plate and filament supplies should be used, as well as careful elimination of hum loops. Extension of signal-to-noise ratio to the limits imposed by the tape itself can be readily accomplished if these considerations are observed, and if appropriate equalization techniques are used.

Since there are a number of different frequency response losses in the record-reproduce process, these must be compensated by equalization, also. By suitable distribution of equalization between playback and record amplifiers, advantage may be taken of the equalization to improve signal-to-noise ratio. Referring to Figure 5, it may be seen that considerable pre-emphasis is used in the record process. The amount of pre-emphasis has been established by consideration of the energy distribution in typical audio signals. Studies made by Sivian, Dunn, and White³, and later confirmed by similar tests in laboratories of the Ampex Corporation, have established these curves as suitable compromises for the various tape speeds.

Figure 6 shows the complementary playback amplifier characteristic required to produce uniform overall frequency response for the various tape velocities. Some improvement in overall signal-to-noise ratio, especially in the high-frequency noise region, might result from modification of these characteristics. However, degradation of the desirable smooth overload characteristic will result when using certain types of tape. The best way in which to improve the high-frequency signal-to-noise ratio is to increase tape velocity. This results, of course, in increased operating costs.

A/B Facility

Since it has already been shown that a conflict exists between those parameters which describe an optimum record head and those which describe an optimum playback head, and since it has been shown that separate heads are therefore desirable, further advantage may well be taken of the situation, in the provision of separate record and playback amplifiers, with a switch facility to provide instantaneous monitoring either of the input to the recorder, or of the output of the playback amplifier. While the human ear is unsatisfactory as a meter of absolute effects, it is probably more sensitive than laboratory instruments at detecting significant relative variations. A continuous check on the performance of the recorder-reproducer, by the ear, the instrument which is in any case its ultimate judge, should be provided, in the form of an "A/B" input-output monitor selector switch. The gain of the record channel after the input gain control should be made as constant as possible, and the level therein should be monitored by a VU Meter of standard ballistic characteristics, in order that uniform tapes may be produced. It is important that no control should be readily accessible which will vary the relation between the VU meter and the record head current, in order that a ready reference on record level shall be preserved. For reasons of operating convenience, it may be desirable to provide a playback amplifier gain control, which necessarily would influence the reading of the VU Meter across the output of such an amplifier, but this, too, may possibly be dispensed with.

Conclusion

It has been the intention of this discussion to illustrate our contention that the design of a high performance tape recorder-reproducer of great reliability involves the simultaneous consideration of a large number of design details. If the procedures outlined are followed, a tape recorder of high performance and great reliability can be achieved.

¹Parks, Richard: "Flexible Broadcast Recorder Installation," Radio-Electronic Engineering Edition, Radio and Television News, December, 1954.

²Kellog, Edward W., "Comment On Flutter Standards," IRE PGA, July, 1954.

³Sivian, Dunn & White, "Absolute Amplitudes and Spectra of Certain Musical Instruments and Orchestras," J.A.S.A. January, 1931.

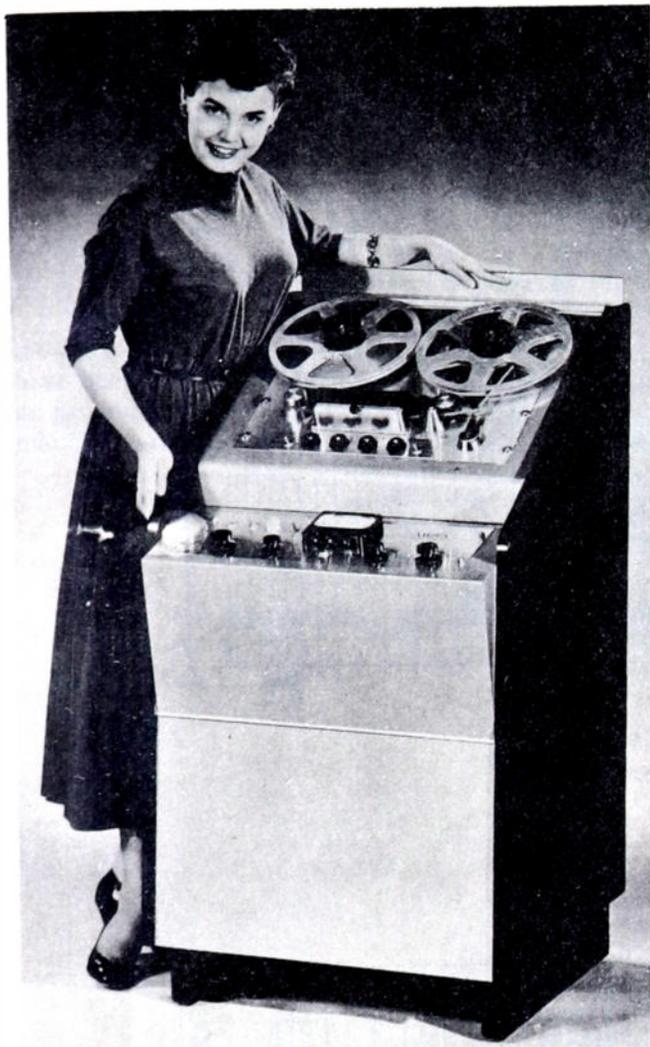


Fig. 1
Typical professional recorder using transport mechanism described in test.

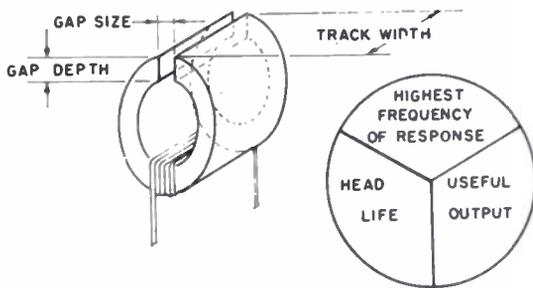


Fig. 3
Variation in the configuration of the head structure can give improvement in each of the qualities of a magnetic head only at the expense of one or both of the other desirable qualities.

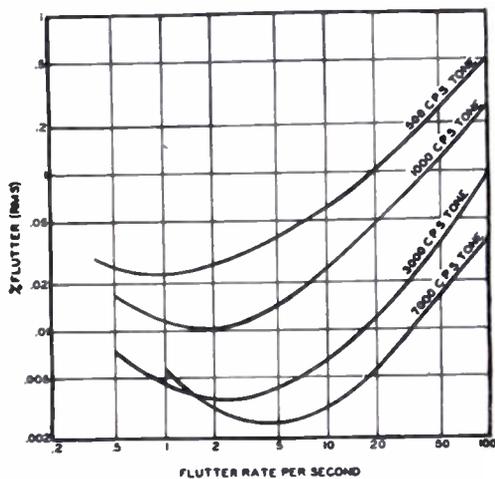


Fig. 4
Minimum perceptible per cent flutter for oscillator tones in small auditorium.

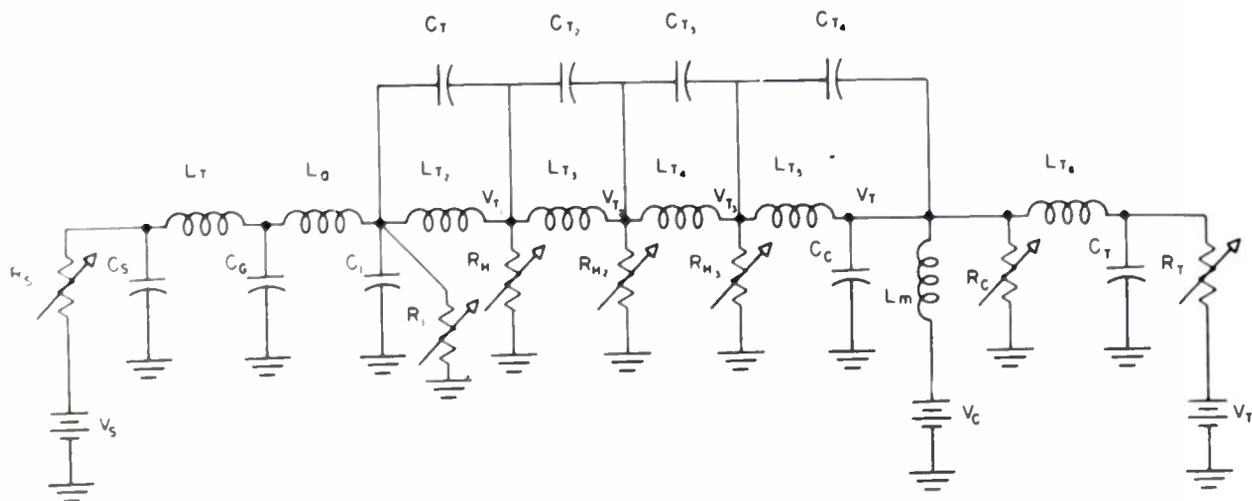


Fig. 2
Electrical equivalent of typical tape transport.

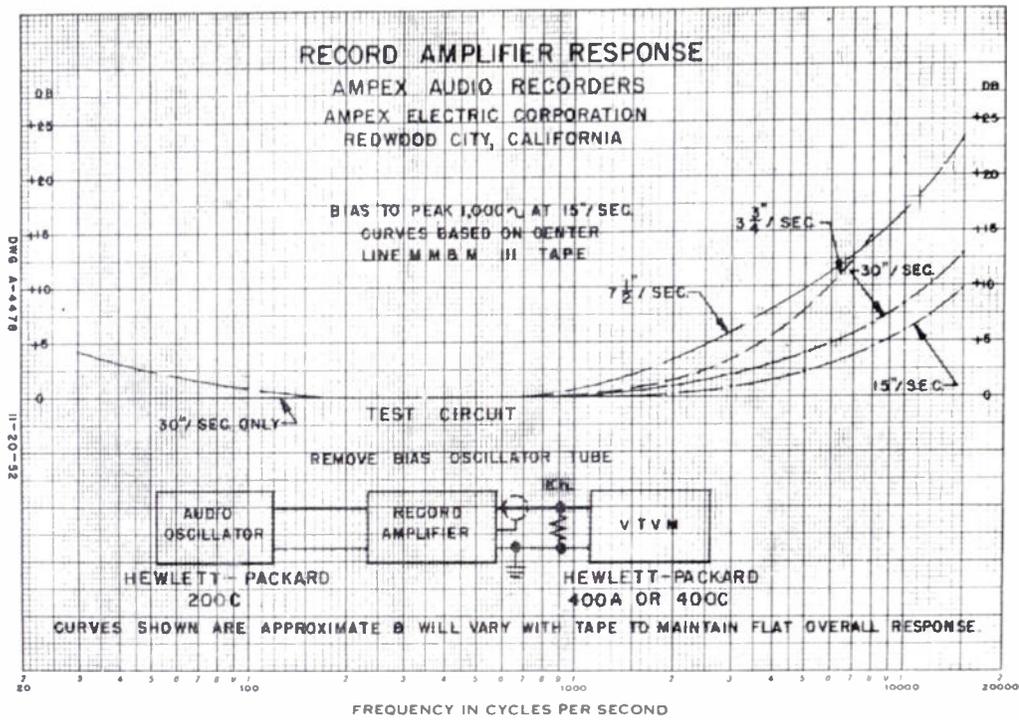


Fig. 5
 Record Amplifier Response.

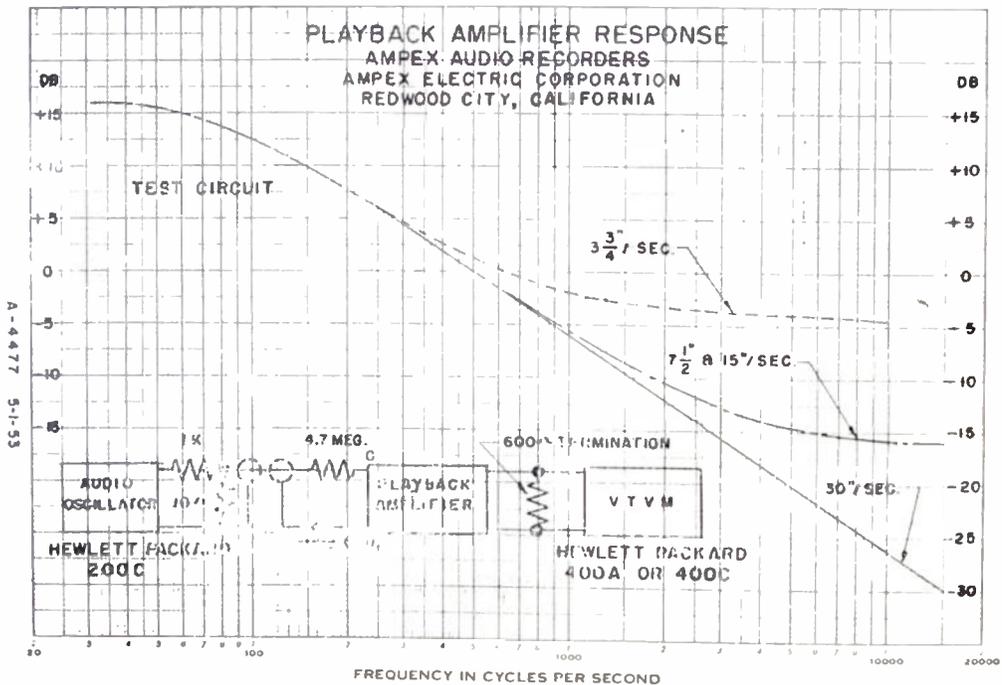


Fig. 6
 Playback Amplifier Response.

TAPE-RECORDING APPLICATIONS

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ABSTRACT

Standard designs are flexible enough for most uses of tape recorders. Special machines have been devised for unusual applications, such as pronouncing dictionaries, length-measuring machines, time compressors, dc and square-wave

recorders, artificial reverberation generators, musical instruments, memory devices, and video recorders. The construction and operation of typical devices are reviewed.

SUMMARY

Magnetic recording provides the scientist and engineer with a new tool which makes it possible to capture and store energy manifestations which can be recreated at will. Thus, when we speak of "tape life," we must consider not only the medium itself, but also the intelligence stored in the medium, which is equally important. With respect to the latter factor, it is necessary to consider (1) the manner in which it is affected and the degree to which it is affected during the passage of time and under various environmental conditions and (2) what can be done to preserve the information in its original form.

Magnetic recording tape, in the form with which we have been familiar over the past decade, has been called upon to perform man-sized feats while still in its infancy. Scientists and engineers were quick to recognize the potentialities of this medium and did not hesitate to make use of it; as a result, in many fields of physical science, magnetic tape, when used in conjunction with its associated apparatus, is an accepted laboratory tool. As more and more reliance was put on the validity and consistency of the intelligence recorded on the magnetic tape, questions arose concerning the life expectancy of the tape and of its contents.

Magnetic recording tape, which employs either plastic- or paper-base material, may have its physical structure and electrical properties undesirably altered by diverse environmental situations, which, unfortunately, we are frequently unable to control. An insurance policy is of little value as protection for magnetic-tape recordings of data and events which occur only once. The best insurance is to try to understand fully the factors which are detrimental to prolonging tape life and to avoid subjecting the tape to unfavorable environmental situations.

It will be the purpose of this paper to discuss these factors, to show how each affects magnetic recording tape and the magnetic patterns impressed thereon, and to suggest certain precautions which will minimize the number of undesirable changes which may be reflected in the medium and its contents as a result of improper handling and storage.

Of course, as in other, similar, situations, it is possible to go overboard and become overly cautious to the point where more damage than good is accomplished in attempting to preserve tape recordings. Look back a few years to the time when home movies were first becoming popular. Metal film-storage cans were made available to the public with special inserts, which the buyer was instructed to keep moist in order to prevent drying of films. It was not long before the idea lost popularity because films were being ruined by mildew!

What external forces influence tape life? Other than situations which cause partial or complete obliteration of the recorded material, temperature and humidity changes produce the most adverse effects. Radical temperature or humidity changes not only tend to alter the physical dimensions of the magnetic medium; they also are reflected in the intelligence recorded on the tape. One of the more common problems encountered with recorded tapes is termed "print-through." This condition exists when the magnetic pattern laid down on a portion of tape influences tape adjacent to it on the reel to the extent that a print of the pattern is impressed on the adjacent layers at a considerably reduced amplitude. Unfortunately, this situation is accelerated and magnified if the recorded reel of tape is subjected to a significant rise in temperature from that at which the recording was made. For several reasons, it is not always possible to keep a reel of tape stored at the temperature which prevailed when the recording was originally obtained. The storage problem is particularly difficult for recordings made at tropical and arctic temperatures. Also, a site selected for the storage of magnetic tapes will be designed to house not one but possibly thousands of reels of all sizes, and obviously it would be impractical to have individually climatized pockets for each reel.

Normally, the "ghost" signal created by the print-through process appears 55 to 75 db below 100 per cent program level, but it can become as high as 25 db below program level in cases in which the original sound was impressed on the adjacent layer at overload level. The minimum tolerable level is approximately 55 db. Ordinarily, in order to prevent this situation from arising, peaks of the material being recorded should be held below 2 to 3 per cent harmonic distortion. A policy at the Underwater Sound Laboratory, where virtually

all recording is done for scientific purposes - to document situations or processes and for subsequent data analysis - is to record 3 to 5 db below normal recording level, or approximately 9 to 11 db below overload when possible. Under these conditions, very few examples of print-through have been noted in spot checking several thousand reels of tape which have been stored since 1947. It has been found that the effects of print-through are greatly reduced if the recorded tapes are stored in temperatures which do not exceed 70° to 75° F.

The early German tapes, which had coercive forces in the order of 80 to 100 oersteds, did not contain properties conducive to print-through, but the later black-oxide tapes, which were first introduced in this country, had coercivities of over 350 oersteds and demonstrated the ability to print through from layer to layer quite readily. Each of these types of tape had certain disadvantages which brought about the formulation of the present oxides. The medium-coercivity red oxides (200 to 250 oersteds) present a compromise in that print-through is in the order of 5 to 8 db lower than that encountered with black oxides. In some cases, where print-through is not too severe, the undesired "ghost" signal can be virtually eliminated by means of surface erasure, but this elimination is accomplished at the expense of high-frequency components of the desired signal recorded on the tape.

The effects of temperature changes on the linear dimensions of samples of various magnetic tapes are illustrated in Fig. 1. It is interesting to note that uncoated cellulose-acetate backing expands as the temperature is raised, but that the oxide, together with its volatile binder, has a negative coefficient of greater magnitude. Therefore, all the coated

TYPE OF TAPE	Δ TEMP. °C (ORIGINAL TEMPERATURE 15°C)	ΔL (INCHES) PER Δ°C
ACETATE ONLY	32	+1.2 x 10 ⁻⁴
	44	+2.4
COATED ACETATE	28	-5.4 x 10 ⁻³
	39	-3.2
COATED PAPER	24	-3.5
	31	-3.4
COATED FILM	29	-3.7
	37	-3.1
COATED LAMINATED ACETATE AND PAPER	15	-3.3
	40	-2.3
IMPREGNATED	14	-1.7
	35	-1.6
COATED 2-MIL MYLAR	18	-1.1
	50	-0.8
COATED 1-MIL MYLAR	21	-1.2
	48	-0.9
COATED 0.5-MIL MYLAR	16	-1.5
	44	-2.1

Fig. 1 - Effects of Temperature on Length of Various Magnetic-Tape Samples

TYPE OF TAPE	ΔL (INCHES) PER Δ% RELATIVE HUMIDITY	
	21°C 60-97% RH	38°C 51-96% RH
ACETATE ONLY	13.5 x 10 ⁻³	18.7 x 10 ⁻³
COATED ACETATE	6.7	8.3
COATED PAPER	3.4	3.5
COATED FILM	3.4	5.6
COATED LAMINATED ACETATE AND PAPER	3.3	4.9
IMPREGNATED	1.5	5.5
COATED 2-MIL MYLAR	0.3	0.8
COATED 1-MIL MYLAR	0.3	0.7
COATED 0.5-MIL MYLAR	0.3	0.3

Fig. 2 - Effects of Humidity on Length of Various Magnetic-Tape Samples

acetate-base tapes contract as the temperature rises; this contraction is probably the result of evaporation of the volatile components. Tapes of DuPont polyester film base, however, tend to contract to a lesser degree. The impregnated tapes showed less contraction than did the coated acetate tapes. With the exception of the Mylar samples, all tapes exhibited a tendency to curl toward the oxide surface.

The change in length of the 1- and 2-mil Mylar tape samples as a function of change in temperature showed a definite decrease as higher temperatures were approached. Also, it was noted that after being subjected to a given temperature rise for a prolonged period the tape tended to return to its original physical dimension, although it never actually reached it. The initial length of all samples was 41 inches.

Perhaps the most important condition influencing cellulose-acetate magnetic tape during storage is that of humidity. Abnormal humidity conditions have the ability to alter the physical properties of the tape to the extent that dimensional instabilities are created which are reflected in the recorded data. In some cases, these changes are irreparable. Present-day Mylar tapes exhibit properties of remarkable physical stability when subjected to drastic changes in relative humidity. This is a relatively new improvement, however, and we are still confronted with the old problems when we consider the thousands of reels of recorded acetate-base tapes which are currently in storage.

In Fig. 2 are shown the effects produced on samples of various magnetic tapes as a result of changes in relative humidity. All tape samples, including the paper-base types, show expansion, but it may be noted that the Mylar tapes are affected to a much lesser degree. Also, note that the relatively high degree of expansion of the uncoated acetate base is considerably retarded by the application

of the oxide and binder. The impregnated tape is affected by humidity somewhat more than is the Mylar but much less than are the acetate and paper families. For these tests, the initial length of each sample was 66 inches at 60 per cent relative humidity.

It is possible for acetate-base tape to curl under various combinations of temperature and humidity. For example, samples were seen to cup at 70° F., 95 per cent relative humidity, and also at 110° F., 60 per cent relative humidity. At 110° F., 96 per cent relative humidity, tape became wrinkled, cupped, and slippery to the extent that there was no surface adhesion between tape and capstan. High humidity produces by far the most devastating effects on cellulose acetate tapes, and Mylar is not entirely immune to its effects.

The "spider-web" effect present in reels of tape which have been subjected to varying conditions of humidity is familiar to many users of magnetic tape. This effect indicates partial deformation of the tape itself and is usually enhanced if the tape has not been wound properly on its reel. Either the tension on the tape was insufficient, or an uneven wind occurred which left edges of single layers extending above adjacent layers at random intervals.

An excellent illustration of this condition is presented in Fig. 3. The inner portion of the reel was uniformly wound with sufficient tension on the tape to prevent such deformation, but the outer portion was wound unevenly and with insufficient tension. It is important that the tape be properly wound before being placed



Fig. 3 - Effects of Humidity on Properly Wound and Improperly Wound Magnetic Tape

in storage in order to avoid such damage. Tape in the condition shown in Fig. 3 is difficult to restore by means other than a process of rewinding evenly and then aging under the conditions which existed when the tape was recorded. Winding and rewinding deformed tape under increased tension is sometimes a practical, but not the ideal, means of eliminating cupping, since the process will actually stretch the uncupped portion to the dimension of the distorted section. Normally, for musical and vocal recordings, this effect may pass unnoticed, but in the treatment of recorded scientific data, where deviations of fractions of a cycle per second are observed with extreme interest, such phase distortions would be intolerable.

Magnetic-tape recordings of underwater sounds made with tape which was originally in good physical condition can create erroneous impressions when reproduced at a later date, if the tape is permitted to become deformed during storage. Periodic loss of high-frequency energy as the cupping of the tape causes it to lose contact with the reproduce head creates an impression quite similar to that produced by actual ship propellers and may, in effect, indicate the presence of a ship which actually was not there when the original recording was made. Rewinding under excessive tension in order to relieve this condition may set up stresses in the tape during storage which result in layer-to-layer adhesion. Of course, if the tapes are reproduced on a tape transport where the capstan precedes the magnetic-head assembly, the effects of adhesion are greatly reduced, but few machines are so constructed. The new narrow-width reels are a great help in achieving uniform winding.

When it is considered that rises in temperature cause acetate-base tape to shrink and that changes in relative humidity cause it to expand, it is easy to visualize the contortions through which the tape goes and the accompanying distortions of the intelligence impressed on the tape. Tests have revealed that conditions of 40 to 60 per cent relative humidity and temperatures between 70° and 75° F. are perhaps the most nearly ideal for reducing the effects of temperature and humidity on magnetic tapes which are to be stored for a great length of time.

At the Underwater Sound Laboratory, where recording is used as a scientific tool, all original disk and tape recordings are preserved. Since the formation of the Recording Branch in 1942, the library of original recordings pertinent to the field of underwater acoustics has grown to include over 5000 disks and over 4000 reels of tape of all sizes.

A portion of the present storage room may be seen in Fig. 4. These disks and tapes have not always been stored under ideal conditions,



Fig. 4 - View of Section of Tape-Storage Facilities at the Underwater Sound Laboratory

but reasonable values of temperature and humidity have been maintained when possible. Many of the recordings have been made in submarines in temperatures exceeding 100° F. at extremely high relative humidities. Others were recorded in sub-freezing temperatures. These reels of tape have shown little tendency to distort physically once they were allowed to become acclimated gradually to the storage conditions. From these thousands of original recordings, random samples have been taken over the years for checking for any ill effects of storage.

Figure 5 shows how the unmodulated groove noise of acetate disk recordings has been affected during storage since 1942. The high surface noise of some of the earliest samples is a result of surface drying. Note, however, that the unmodulated groove noise of

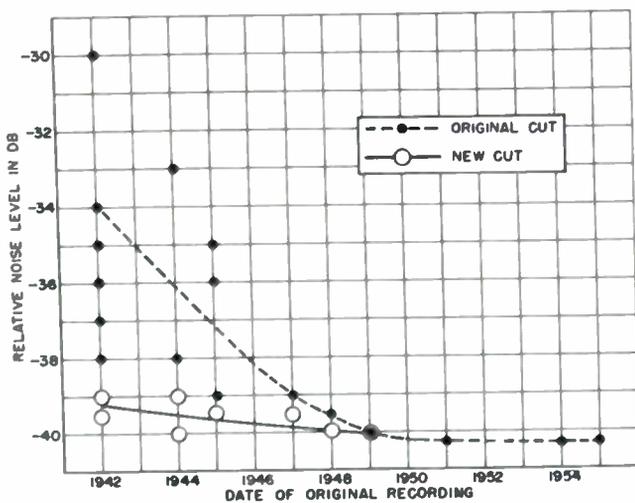


Fig. 5 - Unmodulated Groove Noise

fresh cuts made on these disks approached currently obtainable noise levels. This information is presented to show that surface or groove noise of disks does increase with age even under favorable storage conditions, but the uncut portions remain relatively unaffected as is evidenced by the low noise level of new cuts made in these areas.

Figure 6 illustrates the effects of storage upon magnetic-tape noise—that is, noise from portions of tape which have been subjected to high-frequency erasure but to no other signal. In addition, this graph also shows an interesting bit of history. The 1945 figures were obtained from samples of coated and impregnated German tapes of that period. These tapes, especially the coated tape, have become somewhat brittle, but it may be noted that after re-erasure on a modern machine the noise level was not improved, a fact which indicates that the nine years of storage have had little effect in that respect. These early tapes employed vinyl copolymer bases with vinyl acetate binders which were probably not formulated

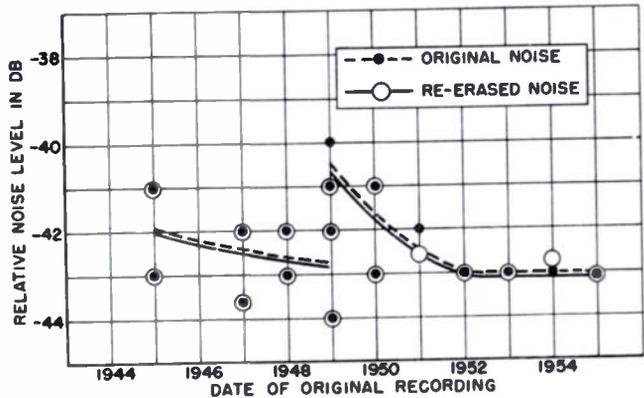


Fig. 6 - Unmodulated Tape Noise

properly to withstand the rigors of climatic variations. Apparently, the Germans did not give too much consideration to the longevity of tape at that time. It is understood that their current policy is to re-record every five years. This procedure, of course, alters the original material.

In 1947, the first American tapes, consisting of black oxide on calendered kraft paper, were appearing. These tapes, as in the case of the German variety, exhibited low noise levels when checked—noise levels which did not change when the same samples were re-erased with high-frequency energy. Both the German tapes and these early American tapes exhibited low signal-to-noise ratios, and it is interesting to note that when the broad-band signal energy originally recorded on the tape was compared with tape noise, no change had taken place in this ratio over the years. The left-hand portion

of the curve indicates these German and early American tapes.

As shown by the graph, 1949 is a year of interest, since it marked the advent of the medium-coercivity red-oxide tapes on cellulose-acetate backing. Note that the black-oxide tapes exhibit approximately the same properties as for the previous year but that the new red-oxide tapes showed higher noise levels. As the years progressed and better oxide formulations appeared, the noise levels of the tapes were reduced to their present status. In all cases, however, when noise levels of the original sample were compared with re-erased noise levels, it was apparent that no increase had occurred during storage, and similarly, the broad-band signal-to-noise ratio had not been altered. It should be emphasized that the points on this chart represent numerous observations, and in many instances the values obtained from the measurements of several reels coincide.

One informative observation was made when stored reels of tape which contained sections of paper leader and timing tape spaced at intervals throughout the reel were being checked. On the portion of magnetic tape immediately preceding the paper timing tape, partial erasure of high-frequency energy had taken place. In Fig. 7 are shown the relative reduction in a 15-kc signal and the degree to which lower frequencies were similarly affected by this phenomenon. Investigation

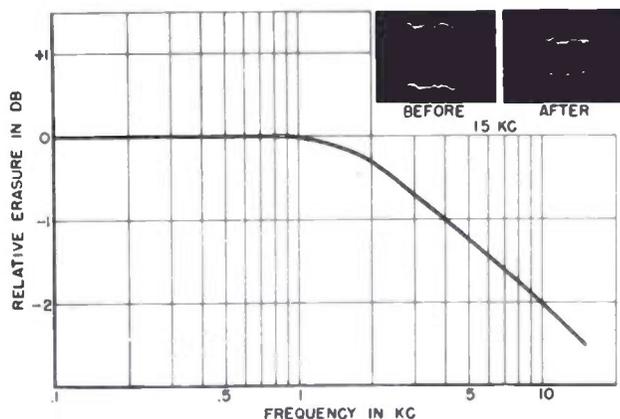


Fig. 7 - Erasure of Recorded Energy as a Result of Contact with Paper Timing Tape

revealed that the paper tape accumulated a static charge during passage over the magnetic-head assemblies. This static charge was then dissipated through the adjoining magnetic tape as the paper tape and plastic tape came into contact on the take-up reel, and partial erasure resulted. Modern plastic leader and timing tapes should greatly reduce this effect.

It should be mentioned that information concerning levels of definite frequencies was

available over the years because of the practice of accompanying all scientific-data recordings with carefully made calibration recordings which permit quantitative treatment of data.

Where magnetic-tape recordings are used for scientific applications, the tape is frequently subjected to conditions placing demands upon its physical strength and electrical properties which far exceed those encountered in normal broadcast and commercial use. At the Underwater Sound Laboratory, thousands of loops of tape are made annually for purposes of spectrum analysis. In order to obtain maximum definition in the analyzed data, it is frequently necessary to run these loops continuously for four hours or more. This application raised a question as to how well the recording medium and its contents would stand up under continuous motion over rollers and heads. Tests performed with loop recordings of individual sine-wave frequencies were first conducted at the Laboratory on a controlled basis early in 1951. After several hours of continuous replay, a significant accumulation of oxide on the reproduce head was noted, and there was an appreciable reduction in high-frequency energy. Because of the vast strides made in the development of new magnetic tapes during the ensuing years, it became apparent that certain improvements had been achieved; therefore, the tests were recently repeated with gratifying results. Figure 8 shows no noticeable decrease in high-frequency energy for 5400

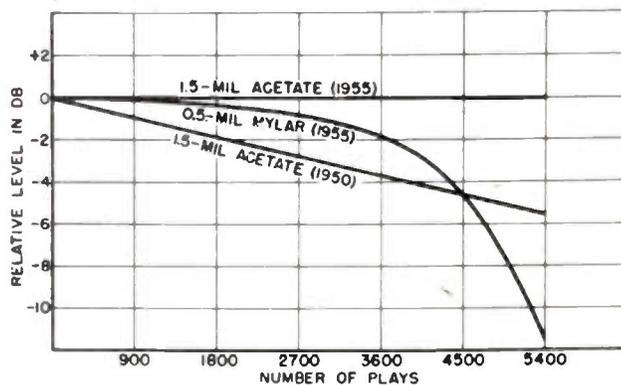


Fig. 8 - Reduction in Level of a 15-kc Signal as a Result of Continuous Replay

consecutive replays of all current types of tape except for 0.5-mil Mylar tapes. Note the loss encountered with the 1950 tapes, however. This information is encouraging, since tape recordings are analyzed at the Laboratory for data in the region of 100,000 cycles per second. Also, no excessive amount of oxide particles was seen on the reproduce head, a fact which indicated improvement in bonding methods. In Fig. 9, may be seen oscillograms of a 15,000-

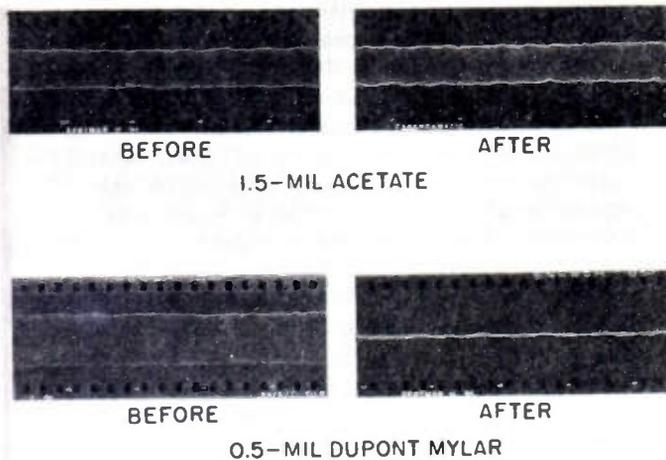


Fig. 9 - 15-kc Signal Before and After 5400 Replays--
1955 Tapes

cps signal, before and after 5400 continuous replays, as recorded on 1.5-mil acetate- and 0.5-mil Mylar-base tapes. Tapes of half-mil thickness will probably tend to print through quite readily during storage, since it was noted that a 10,000-cps signal reproduced through the back of the tape reduced only 23 db in amplitude, whereas 1.5-mil tapes showed no such tendency.

Current magnetic tapes are actually more durable than many persons are inclined to believe. In all the samples of reputable tapes analyzed, only on the early black-oxide plastic tapes did the oxide show any tendency to scrape off, and in this case it was the result of tape adhesion on the supply reel. The base materials appear to be in excellent condition and exhibit adequate tensile strength, which indicates very little weakening of the components structurally. It is known that laboratory samples of cellulose acetate show no ill effects after 12 years. Reels of recorded magnetic tape have been placed on top of voltage regulators, in trucks near FM transmitting equipment, and aboard high-speed ships, where they are subject to rather severe vibration, with little or no apparent ill effect either to the tape or to the information recorded on it. Recorded tapes have been dropped from heights of 15 feet onto solid concrete floors and have been heated to the point where they were uncomfortable to hold, with no serious physical or electrical effects.

Only in the case of recorded reels of tape which had been submerged for several hours in salt or fresh water were any serious effects permanently incorporated in the recorded intelligence, and these effects occurred at the high-frequency end of the spectrum in the form of rather severe amplitude variations. The variations were reduced as the tapes were dried and rewound, but they did not disappear entirely. It is not recommended that this practice be adopted for recorded tapes, but because of the

wide variety of circumstances under which recording work is done at the Underwater Sound Laboratory it was necessary to go to extremes to establish the precautions which must be observed when recorded magnetic tapes are handled.

Although the present storage room at the Laboratory has been lined with fire-retarding material, it was necessary to consider the threat of fire originating externally, and for this reason some rather extensive burn tests were conducted. It was found that the cardboard box in which a reel is stored provides very good fire protection for the reel of tape. Individually boxed reels of tape were partially placed in an intense flame for a period of 10 to 15 minutes and then examined. The result of one such test is shown in Fig. 10. The box itself did not

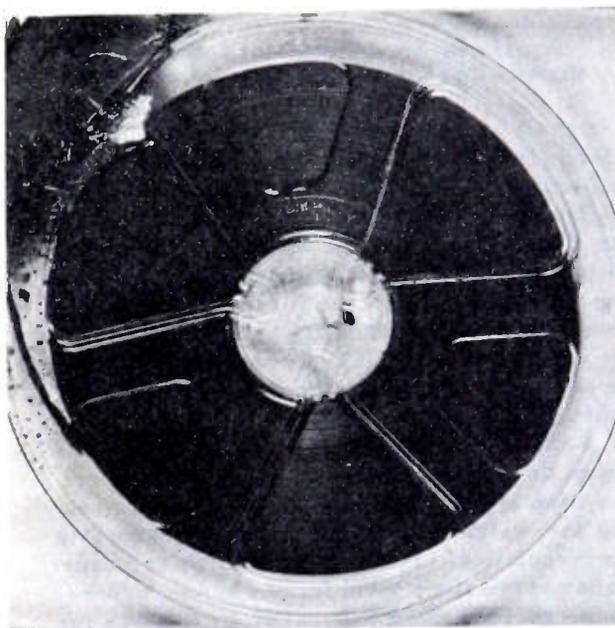


Fig. 10 - Effect on Boxed Recording Tape of 15-Minute
Exposure to Direct Flame, Cover Removed

ignite, but pressure built up inside to the extent that smoke was puffing out of the corners of the box. The reel started to melt but did not burn, and only the outermost layer of tape had been seriously affected. With the exception of metal reels and cans, all items directly connected with the spooling and packaging of magnetic tape are combustible and will burn with varying degrees of intensity once they are ignited.

Since the recordings made at the Laboratory are valuable, not only as records of past events but also as tools for future scientific research, plans were prepared for the construction of a special vault which could be climatically controlled by means of separate temperature- and humidity-control devices. The interior of

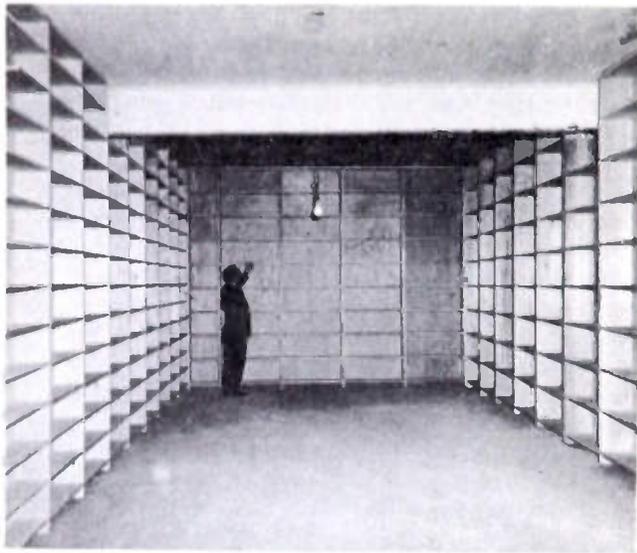


Fig. 11 - Interior of Climatically Controlled Magnetic-Tape-Storage Vault Under Construction

this vault may be seen in Fig. 11. It is entirely above ground level and is constructed of 8-inch cement walls; the ceiling and the floor are also of cement. A thin plaster surface coats all but the floor. The vault is 50 feet in length, is 16 feet wide, and has a 10-foot ceiling. There are no openings into the vault other than one door and the air-circulating vents connected to the dehumidifier, which is located externally, as are all other temperature- and humidity-controlling devices.

Storage shelves line the perimeter of the room, and approximately three inches of space has been left between the back of the shelves and the walls in order to provide for adequate air circulation. The bottom shelves are four inches above floor level for the same purpose. These shelves alone provide storage space for approximately 18,000 seven-inch reels and 2,600 ten-and-one-half inch reels of 1/4-inch tape. Additional rows of shelves will be installed in the center of the room for the purpose of storing stocks of unused tape. It is planned to store some original disk recordings in this vault also, since optimum storage conditions for these are practically identical to the best conditions for magnetic-tape storage.

The dehumidifier which controls the climatic condition of the vault can be adjusted to maintain almost any desired combination of temperature and humidity. This unit is shown in Fig. 12. It is a desiccant type, which employs two 50-pound piles of silica gel. The blowers in the dehumidifier circulate air in the vault at the rate of 400 cubic feet per minute, and the unit is capable of removing 7.5 pounds of water per hour. The temperature of the vault will be maintained at from 70° to 75° F., and

the relative humidity will be confined to the limits of 40 to 60 per cent.

It is not difficult to draw conclusions from the information discussed in this paper. Neither is it difficult to make recommendations. The difficulty lies in setting up conditions, which, based on the conclusions, will fulfill the recommendations. Climatically-controlled storage facilities are costly. Other storage rooms, which have less than desirable temperature- and humidity-control facilities, should be considered as temporary measures. If the data recorded on the magnetic tape are sufficiently important scientifically to warrant retention at all, it is certainly obvious that we will want the stored information to remain unchanged during storage; we should therefore take all precautions to protect it and the medium upon which it is stored.

Where funds and facilities are not available for adequate protection of tape during storage, the next best procedure is to avoid creating conditions which would result in introducing undesirable effects into the tape and the information on the tape. Do not use tape which already has been severely affected by humidity as indicated by the familiar "spider-web" effect. Before attempting to record, rewind the tape under slightly increased tension in order to reduce this effect, and allow the tape sufficient time to become acclimated to the

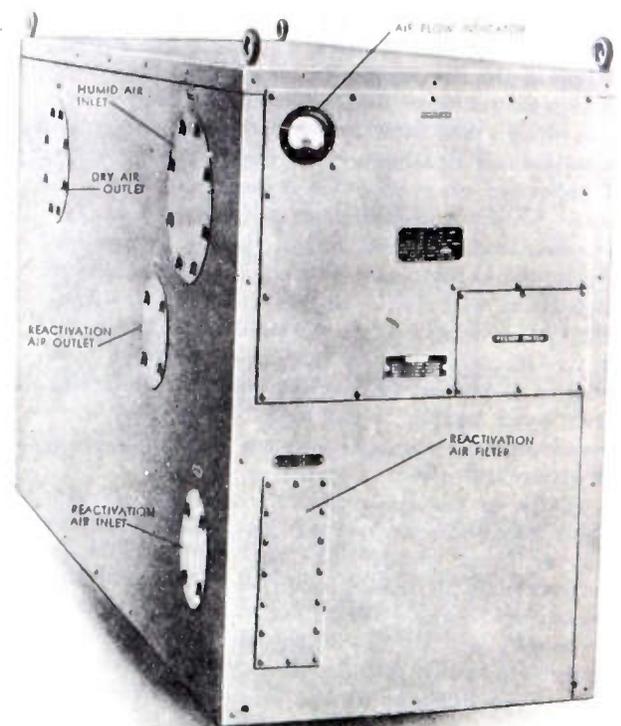


Fig. 12 - Automatic Dehumidifier for Magnetic-Tape-Storage Vault

conditions under which the recordings are to be made. Take extra precautions during the recording process to insure that no continuous signal overload occurs, which, when supplemented by increases in temperature at a later time, may cause layer-to-layer print-through in the reel of tape. Be sure that the tape winds evenly during both record and rewind operations. Discard warped reels and check clutch tensions on the tape transports frequently. Repair idlers which have worn bearings. The latter items affect the tension and skewing of the tape during

winding. Store the recorded reels of tape in their individual boxes, on end, and avoid exposure to dust. Tape which has been run recently through a tape transport will often attract dust particles in much the same manner as does a vinylite pressing.

As a final word, scientific use of magnetic tape recordings demands the utmost and offers little compromise. The intelligence recorded on the tape today must be mirrored in the intelligence obtained when this tape recording is reproduced years hence.

THE FUTURE OF MAGNETIC RECORDERS

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The principle of magnetic recording was invented over 50 years ago but the art did not develop to any great extent until relatively recent times, and certainly the greatest development has occurred since the end of World War Two.

The early wire recorders called telegraphones suffered from serious limitations in that no vacuum tubes were available to amplify the weak output from the reproduce head. Consequently, it was not until the middle thirties that any considerable satisfactory magnetic recorder was produced.

Since the end of World War Two advances in both components and systems have been extremely rapid. One need only look at the amount of business done in magnetic recorders in 1945 as compared to 1954. In the former year the total business in magnetic recorders, essentially wire, there being no tape recorders at that time, probably amounted to a few hundred thousand dollars; while in 1954 the total tape recorder business was over 120 million dollars. That the future of magnetic recording is very bright cannot be denied. It is anticipated that 1955 will see an increase of about 15% in dollar value and it does not appear that there will be any leveling off of this increase within the foreseeable future.

In trying to predict the future of this art one may break the matter down into two areas; namely, components and systems. Certainly they go hand-in-hand but by this analysis the problem is somewhat simpler.

Components can be grouped into essentially the following: transducers, media and drive mechanisms. Transducers, of course, refers to what we commonly call the head. We are all familiar with the straight forward type of record reproduce head as used on standard audio type of recorder. Perhaps not so familiar are some of the limitations of this type of head such as the fact that its output, when used as a reproduce head, is essentially a function of frequency. A further difficulty is the requirement that the gap which contacts the tape be small, smooth and straight.

A few years ago we all thought we were doing very well if we could record and successfully reproduce one thousand cycles per inch. As of today, a number of machines are available commercially which will do two thousand cycles per inch and these have certainly not approached the ultimate resolution of the medium.

Improvements in the operation of this type of head do not appear to be very practical. However, improvements in its manufacture are certainly coming. At the end of the war practically no one produced a quality head of this type for sale as a component. However, today there are a number of manufacturers in this field and others are planning to enter it. Manufacturing techniques are being continuously improved to the end that a cheaper head is produced, yet has higher quality. Other types of heads such as the magnetic amplifier, the perpendicular bias field and the boundary recording type have seen considerable development in the past few years. It is to be expected that still further development will be seen in these types of heads to the end that increased resolution, increased signal-to-noise ratio, and narrower tracks will be possible.

In the area of media, we had ten years ago essentially only one tape which was an American-made copy of material produced in Germany during the war. Today there are quite a number of tape manufacturers producing a tremendous variety of products. A large amount of tape has been produced having one and one-half times as much on a spool as was formerly available. This seems to me to indicate a trend although it seems that the ultimate has almost been reached.

One of the shortcomings of tape recording, as compared to that practically ancient art of wire recording, is that for a given quantity of recording on tape, a greater cubic content is required. Certainly the advances made in recent years toward thinner base tapes and increased resolutions has lead to diminishing of this undesirable factor. In fact, a tape is available today which has 2400 feet on a seven inch spool in-

stead of the usual 1200 feet.

A further development in tapes has been to increase their output, minimize their distortion and improve their frequency response. Certainly the manufacturers of tape are not overlooking the possibility of still greater improvements but they are undoubtedly hampered by the large number of existing machines which cannot accept a very great change in tape parameters. It seems reasonably certain that a tape could be manufactured today having marked improvements over those presently available, but such tape would require a relatively new system of both recording and driving means. Consequently, the market for such tape would be relatively limited, at least at the beginning.

One of the difficult factors in the driving mechanisms, using the extremely thin base materials, is that the transient condition of starting and stopping may very well exert excessive mechanical strains in the medium. The development of new types of drives is proceeding almost as rapidly as any other area in this business. In particular, the availability today of drive motors operating at such speeds that practical sizes of capstan can be used to drive the tape at normal speeds results in an exceeding constant drive having relatively low flutter and wow. Further development of the reel drives, such as those in employing electrical or mechanical servo systems, points to the direction of further stability and simple operation of machines.

In the area of systems development certainly the greatest progress has been made in the non-audio or instrumentation recorders. In fact so many different systems have been developed that this progress almost staggers the imagination. Of particular interest are those systems now used for recording telemetered data, in computing machines, and TV. The most spectacular of these is certainly TV.

It is relatively apparent that considerable development effort is being expended in this area and it is indicated in the trade press that one of these

systems is now being used on a limited operational basis. It is to be anticipated that the ultimate application of TV recording techniques will be such that for many applications photographic recording will be essentially eliminated.

The method of recording data as frequency modulation is undergoing still further development and may be expected to result in increased bandwidth and improved accuracy in the foreseeable future.

Further development on the pulse ratio type of recorder system should undergo the same progress. Multi-track recorders such as those used in computing machines have seen improvements in both linear and transverse packing density as well as other parameters. Many of these machines operate on a high speed start-stop basis, and machines are currently or will shortly be available which accelerate a piece of tape from stop to 100 inches per second in the order of one millisecond. It is to be expected that further improvements will be made in this direction.

In summing up one might predict the day when almost everything we do is controlled or at least influenced in some way by this growing industry of magnetic recording. Already our telephone calls are sometimes recorded magnetically even to where our bills are magnetized spots on a piece of tape - we are bombarded with magnetically recorded music in our favorite restaurant - the sound in a local movie comes from a magnetic track - freight shipments on the railroads are speeded by tape recorders - the local police and fire departments all use such units - and the list goes on for page after page with no end in sight.

What will the list look like next year or ten years from now? Certainly with the increased resolution, improved signal-to-noise ratio, better media, new types of recording and reproducing systems along with still simpler and more reliable systems we can look forward to a bright future in this magnetic recording art.

REMARKS BY RALPH BOWN, CHAIRMAN OF SYMPOSIUM ON
SPURIOUS RADIATION AT IRE CONVENTION, MARCH 23, 1955

Ladies and Gentlemen:

I would like to open this Symposium on Spurious Radiation by reminding you that this session is sponsored by the Professional Groups on Broadcast and Television Receivers and Broadcast Transmission Systems. Perhaps no better example need be cited of how the professional group system contributes to the vitality of the Institute than to note that two of the most fundamental groups have joined to initiate this discussion of a pressing and difficult present and future problem of radio engineering.

Being naturally a simple and direct sort of fellow, when I sat down to think about the subject of Spurious Radiation I first wanted to know just what I was going to think about. As a radio engineer I was quite familiar with the word "radiation." But just what did "spurious" mean? Going to the dictionary I found these meanings:

1. Bastard, of illegitimate birth;
2. Not proceeding from the true source, not genuine, counterfeit, false.

The part of this which struck me as most applicable was the word "illegitimate" - illegitimate radiation. Just what does illegitimate mean?

The one dictionary definition said "unlawful" and another said, "not authorized by good usage."

Here it seemed to me I had hold of words which fitted the case. Spurious radiation is radiation not authorized by good usage.

My next question was, what is the history of spurious radiation - when did it originate? One might say it originated back in the days when the second radio transmitter and the second radio receiver got tangled up with the first transmitter and receiver. But bear in mind that there was no good usage in that day, and certainly no authorized good usage. The word "spurious" would hardly have applied. Moreover in those early days of passive and insensitive receivers, radiation was presumably only a property of transmitters. As receivers have become more sensitive and transmitters more numerous, Interference, with a capital "I" has become and will continue to be a major problem of radio engineering.

At this point we must distinguish between natural interference, as from lightning and other atmospheric phenomena, and man-made interference. The former has long been known to radio engineers as strays or static. The term static is now of doubtful utility since it has been adopted by

the public as a name for any nuisance signal bothering radio or TV reception. Spurious Radiation does not include natural interference, but only the man-made variety.

Spurious Radiation is thus a property of apparatus and systems, and so, theoretically, is capable of engineering control through design and operating specifications and practices. The question then becomes one of what is "authorized good usage." Some philosophical discussion of this point may be appropriate as an introduction to this symposium.

When I arrived at this juncture in making up my notes for this discussion, my secretary had just come back from the library with the official IRE definition of Spurious Radiation she was able to find. It is stated to be "any emission from a radio transmitter at frequencies outside of its communication band." I think you will agree that this did not help very much for our purposes today because the kinds of spurious radiation we are going to talk about may come from a transmitter, a receiver, or any other device which radiates energy beyond the bounds of "authorized good usage."

Let us look at these three words again - authorized good usage. Each of them is significant.

"Usage" my dictionary calls "customary procedure or action." In other words, it is what we normally do.

"Good" my dictionary says is "sufficient or satisfactory for its purpose." So we may say good usage is what we normally do that is sufficient and satisfactory. Obviously good usage will change with time and circumstance and obviously it is essentially practical in nature.

Let us now open the dictionary to the word "authorized." It means "possessed of or endowed with authority." And authority, in turn, is said to be "legal or rightful power."

I do not wish to become tiresome or fine-spun about this semantic exercise, but it seems clear to me that certain ideas flow out of this language which are important and fundamental.

- a) The basic standard of comparison is good usage - that which is customary and satisfactory. In other words, our judgments should be practical and workable rather than visionary or theoretical.

- b) Working standards of comparison should be endowed not only with legal but with rightful power. They should be not only practical and workable, but morally right. That is to say, they should appeal to our conscience and our ethical sense.

If in our discussion and argument we would accept and keep these principles in mind, perhaps we would save time and arrive at sound conclusions.

When I came to this point in my thinking I looked back over my notes and concluded that there were at least three other fundamental considerations which we should have in mind in discussing spurious radiation.

The first one is that there are many circumstances in which radio is the only feasible method of communication. These include not only mobile stations on land, sea and air, but also those cases of isolation as in the arctic or as in the mountains where communication by continuous electrical conductors is just not feasible.

The second circumstance relates to the fact that in many cases the communication is endowed with special responsibility for safety of life or other very important human values. This relates not only to disaster calls from mobile units, but perhaps also such cases as alternate routing or back-up services for major public services handling great volumes of communication, including those of national defense.

Evidently these two categories of radio communication are endowed with a special character which might put them in a different class than services that are concerned essentially with amusement or convenience.

The third and final point, it seems to me, is the economic equation. In any case of con-

flict involving spurious radiation - and one may question whether the term spurious is applicable unless there is conflict - it is necessary to recognize that there is a dollars and cents equation which needs to be balanced so as to produce as nearly as practical the minimum overall cost to society and/or the most equitable distribution of costs among those elements of society which must bear them.

I would like now to recapitulate the points I have tried to make as being fundamental to a discussion of spurious radiation - a term which I like to define as meaning "radiation not authorized by good usage." I find I can collect the points into three packages, as follows:

1. Good usage implies practical and workable judgments with standards of practice which have the force of moral soundness.
2. Circumstances in which radio is either the only feasible method of communication, or is deeply involved in individual or public safety, should be credited with this fact.
3. The ultimate solution will always involve some reasonable economic balance if it is a good engineering solution.

Questions

1. Does the FCC now have power to forbid use of receivers causing interference to other services?
2. Is it or is it not a good idea to have the FCC specify radiation limits for all receivers and require each one sold to carry its label of type test approval?
3. As a practical matter, how can FCC type testing of receivers be avoided?

CONTROL OF RADIO INTERFERENCE IN CANADA

Mr. G.C.W. Browne
Controller of Telecommunications
Department of Transport, Ottawa.

The large geographical area of Canada as compared with the size of its population has always made it economically difficult to provide radio broadcast signals of adequate strength in many areas, and this condition has emphasized the need to keep down the level of radio noise and interference. The importance of this was recognized since the early days of broadcasting, and as early as 1928 an interference service was set up in what is now the Department of Transport.

From the inception of radio communication the Department administered the laws governing the use of radio, and in 1936 legislation was passed empowering the Government to prohibit or regulate the use of any machinery, apparatus or equipment causing or liable to cause interference to radio reception.

In 1941, the regulations made under these powers took on very nearly their present form, prohibiting the use of any electrical or other apparatus which causes or is liable to cause interference to reception of a radio signal of 500 microvolts per meter or greater, provided the interference can be suppressed at a cost of less than fifty dollars. In addition the Minister of Transport may order suppression of interference, even though the cost exceeds fifty dollars. A further provision prohibits the use of devices which emit radio frequency energy intentionally and for a purpose, that is, whose radiation is essential for the purpose, rather than incidental. Such devices may only be used with the written permission of the Minister of Transport.

All this, of course, is in addition to the legislation whereby the Department licenses radio transmitters, and the use of radio communications in general. Transmitter harmonics and spurious emissions are of course controlled in the terms of the licence of the individual transmitter and are in general required to be at or below the levels recommended by the International Telecommunication Union.

To enforce the provision that apparatus shall not be used which causes interference to reception of a normal signal, the Department maintains some fifty cars fitted with interference investigating equipment, at the major centres across Canada.

All complaints of radio interference are investigated, and while the onus of having it

suppressed lies with the owner of the interfering apparatus, the Department does advise as to the best means of suppression. Each investigator is equipped with a kit of plug-in and other types of suppressors, to ascertain by test the most efficient and economical means of suppressing the interference.

In the past, interference control has been largely in the "use" stage but recently the Radio Act has been amended to give powers to the Government to make regulations to also control the sale or offering for sale of any electrical or other apparatus liable to cause radio interference.

Regulations to this effect are under study, and one of the principal bases for these are the Interference Standards developed by the Canadian Standards Association, with the co-operation of the Department of Transport. The following is a brief resume of these Standards:

Standard C22.4 No. 101 - "Radio Noise Measuring Instruments and General Methods of Measurement". This Standard specifies the use of quasi-peak type measurements for radio noise, with second detector time constants of less than three milliseconds charge time and 600 milliseconds discharge time.

Standard C22.4 No. 105 - "Tolerable Limits and Special Methods of Measurement of Radio Interference from Electrical Appliances and Equipment".

This Standard calls for conducted noise to be less than 500 microvolts, and radiated noise to be less than 100 microvolts per meter at 30 feet. Equipment is exempted, however, if the noise does not occur for a total of more than ten seconds in any one hour period, nor more often than thirty times in any one hour, in normal operation.

Conducted noise is measured by using R.F. isolating chokes in the power supply, and by loading each side of the line by 150 ohms resistance to ground, via a coupling condenser. The measuring instrument forms a part of the 150 ohms. The 500 microvolt limit refers to the actual R.F. voltage on the line, and not merely to the portion across the meter.

Standard No. 105 also provides for the use of sampling procedures so that measurements need only be made on a limited number of items which are produced in large quantities.

Standard C22.4 No. 106 - "Tolerable Limits of Radio Interference from Radio Frequency Generators - Industrial, Scientific and Medical". Radiation shall not exceed 15 microvolts per meter beyond 1000 feet, except harmonics of the Industrial, Scientific and Medical (I.S.M.) frequency bands, which shall not exceed 25 microvolts per meter at this distance. At R.F. powers above 5 kilowatts, up to 10 microvolts per meter at one mile may be permitted.

Standard C22.4 No. 104 - "Tolerable Limits and Special Methods of Measurement of Radio Interference from Vehicles using Internal Combustion Engines".

The limits are as follows:

500 kilocycles to 100 megacycles, 35 microvolts per meter at 15 feet from the side of the car on which the distributor is situated; vertical polarization.
100 to 400 megacycles, 100 microvolts per meter at 50 feet from distributor side of car, horizontal antenna, $7\frac{1}{2}$ feet above ground.
400 to 1000 megacycles, 1500 microvolts per meter, same measurement method as 100 - 400 mcs.

All values are quasi-peak.

There are also Standards for noise from power and other lines, trolley cars, etc., but time will not permit detailing these.

TV horizontal sweep oscillator radiation is not covered by these Standards, but is a major source of public complaint. In fact, in certain cities in Canada this interference is so acute that even powerful local stations are over-ridden. Because of the nature of the interference it would appear that a limit of about one-fifth the limit for appliances would be appropriate. A similar limit, about one-fifth that for

appliances, also appears appropriate for radiation from the heterodyne oscillator of TV receivers using other than a 41.25 megacycle intermediate frequency.

Before closing I should perhaps make a few comments on the problems of applying and administering such limits. As we all are aware, the practical application of such limits tends to be rather complex and we therefore feel that every effort must be made to simplify procedures as much as possible. Measurement of radiation at VHF frequencies tends to be affected by a number of variables which are difficult to control, and serious consideration is being given to relying, initially at least, on measurements of conducted noise only, which are much more manageable and easy to make. This of course would apply only to appliances and possibly to TV receiver interference. Such measurements of conducted noise from TV receivers would include measurements with the antenna terminals grounded.

Another simplification which appears possible in the case of appliances and TV receivers, at least in the initial stages, is to apply the limits only to noise generated on frequencies in the Standard Broadcast band, and on TV channels 2 to 13. Since all of these are essentially household appliances, they are normally located much closer to Standard band broadcast and TV receivers, than to receiving facilities in any other service. Also, suppression for Standard Broadcast and TV frequencies will generally be at least partially effective at other frequencies.

In the application of these Standards, and the development of further Standards, we shall continue to seek the co-operation of the industry, as in the past.

THE TECHNICAL CONSIDERATIONS UNDERLYING THE REGULATION
OF SPURIOUS RADIO EMISSION: A STUDY UNDERTAKEN FOR
THE FEDERAL COMMUNICATIONS COMMISSION BY THE JOINT
TECHNICAL ADVISORY COMMITTEE.

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The title of my talk, and its relevance to our world of communications, have been fully set forth to you by Dr. Bown. The concern of the Federal Communications Commission with this matter, and the justification for that concern, have been clearly presented to you by Commissioner Webster. Let me therefore, by way of introduction, speak only to the position of the Joint Technical Advisory Committee in relation to this morning's topic.

The Joint Technical Advisory Committee was established seven years ago by the IRE and the Radio-Electronics-Television Manufacturers Association, to serve as a source of competent and unbiased technical information in the radio and allied fields (insofar as competence and lack of bias are humanly achievable). It was the intention of the sponsor organizations that the services of the JTAC should be available to public bodies concerned with our field of interest. The Federal Communications Commission, with its tremendous responsibility for the regulation of the use of the radio spectrum in the public interest, is of outstanding importance as a public body of this sort.

The composition of the JTAC reflects its sponsorship; of the eight members, half are appointed by the President of IRE, and half by the Director of Engineering of the RETMA. The Chairman of the JTAC is appointed alternately by the sponsor bodies, and serves for a one-year term. The Chairman and the members receive no compensation for their services, but the sponsor organizations provide meeting space, adequate secretarial support, and the funds to cover the costs of such publication as may appear appropriate. So much for introduction.

On December 3, 1952, Mr. Walker (then Chairman of the Federal Communications Commission) wrote to Dr. Bown asking that the

Joint Technical Advisory Committee undertake a study of the technical factors underlying the regulation of spurious radio emission. The Committee recognized the importance of the FCC request and agreed to proceed with it, and established a subcommittee for the conduct of the study.

To develop an understanding of the nature of the problem has been itself a major task. The members of the JTAC and the members of the subcommittee have spent a great deal of time on the direct question and on related questions in the interval since Chairman Walker's request. A report discussing the technical factors is now in preparation, and it is my purpose this morning to present to you a summary of the present form of that report. Incidentally, I should like to make it quite clear that since the report has not yet been officially approved by the JTAC, and is therefore subject to change, my remarks here represent only my own expectation as to its final content.

The report takes note, among others, of two particularly important technical facts: first, the generation (or for that matter even the amplification) of radio frequency energy is in general accompanied by some measure of spurious radio emission; second, a reduction in the amount of spurious radio emission accompanying any generation of radio frequency energy usually increases the cost of the apparatus generating the energy.

In preparing its report, it has been continuously apparent to both the JTAC and the subcommittee that it was the desire of the members of the FCC (and their obligation under the law, as well) to regulate spurious radio emission in such fashion as best to serve the public interest.

To regulate in the public interest----

Let us look at a specific example, in order to try to fix the nature of the public interest. With respect to present practices, I believe the example I shall use to be entirely hypothetical. Assume, however, that tv receivers using intermediate frequencies of the order of 40 Mc had the general custom of placing the oscillator below the signal for the reception of channels 7 through 13. When receiving channel 10, therefore, the receiver local oscillator would lie in the band of frequencies near 150 Mc which has been heavily used for certain public services and for taxicab radio and things of this sort. Now, if radiation from the tv receivers is great enough to require drastic increase in transmitter power for the 150 Mc services to operate effectively, the cost of the added transmitter power, or, alternatively, the cost of giving up the services will have to be borne in the end by the public. On the other hand, the cost of reducing the radiation from the tv receivers to the point where it causes no disturbance whatsoever, even to seriously underpowered services in the 150 Mc region, will also come home to the public in terms of a substantial increase in the prices which must be paid for the television receivers. Now, in such a case, where does regulation in the public interest lead us?

Obviously, there is no simple answer. Probably, in my hypothetical example, reasonable regulation would turn out with a permitted amount of spurious radio emission from the tv receivers very much greater than the least perceptible amount of power, but on the other hand very much less than the power which receivers might radiate if no limit at all were to be established.

The objective of regulation, seen in the light of the important related technical factors, appears to be the making possible of the amount and kind of each of several services which the public wants at the minimum total cost to the public.

In practice, the problem of establishing limits for permissible spurious radio emission from the various classes of emitters and at the many frequencies of interest in the radio spectrum is an extremely difficult job. However, approximate values can be developed relatively easily, and may then be submitted to open discussion by the various affected interests. If the importance of developing limits which are close approximations to the ideal numbers is great enough to justify the effort, these limits can always be found.

In the studies leading to the JTAC

report, apparatus producing spurious radio emission has appeared to justify separation into two classes: first, the class of apparatus which is operated under license to a governmental agency such as the Federal Communications Commission; second, the class of apparatus which is not normally regarded as requiring a license for its operation. The most important difference between these two classes of apparatus comes up with respect to the time at which control of the spurious radio emission of the apparatus is most conveniently exercised. In the case of licensed apparatus, control may be exercised at any time that it becomes necessary, since on any occasion when an excessive and troublesome amount of spurious radio emission is traced to such a piece of apparatus, the licensee is under direct obligation to use his best efforts to correct the complaint, and he is in the general case competent to act. With non-licensed apparatus, however, control of the spurious emission is readily exercisable as a practical matter only at the time of manufacture. Television receivers are examples of this class of apparatus, and here it is obvious that as a practical matter control of the spurious radio emission must be exercised by the designer and the producer of the merchandise rather than by the thousands of individuals who purchase the merchandise at the retail market.

It is probable that under present legislation the FCC authority with respect to non-licensed apparatus extends only to its use and not to its manufacture. However, if the Commission were to define these kinds of non-licensed apparatus which are capable of radiating sensible amounts of spurious radio emission as certifiable apparatus, a substantial step forward might be taken. The Commission could then invite manufacturers proposing to build such apparatus to propose limits for permissible radiated power and to propose methods of securing conformance by the industry with those limits. The Commission could then invite comment on the proposals from users of the frequencies where disturbance might be expected and the Commission could then determine limits in the light of all of the considerations brought before it. If the Commission would then accord to a piece of certifiable apparatus which had in fact been duly certified the same privilege accorded to a licensed piece of apparatus; namely, that of being a duly authorized emitter, the purchaser of the television receiver or whatever other apparatus was involved might be given assurance that all requirements applicable to his apparatus at the time it was made had been complied with, and that he would have every reason to expect that he would be able to enjoy undisturbed operation of the apparatus which he purchased.

IRE AND THE MEASUREMENT OF SPURIOUS EMISSION

Ernst Weber, Polytechnic Institute of Brooklyn
Chairman, Standards Committee, IRE

Recent discussions of the problem of spurious emission from radio and industrial equipment have naturally led to strong interest in measurement methods of spurious emission. Since standardization of methods of measurement has been one of the foremost professional contributions by the IRE technical committees it is appropriate to relate here the pertinent activities within IRE. But even before doing that, it might be well to point out some of the technical aspects in order to put the whole problem into its proper perspective.

The spurious emission at radio frequencies of any radio, television, or industrial equipment is extremely dependent upon environmental factors. A reasonably accurate measurement of spurious emission represents therefore an extremely difficult problem. It might be best to illustrate the complexities in terms of the testing of receivers, be it broadcast receivers or television receivers.

Major sources of interfering electric and magnetic fields and power line interference are local oscillators, sweep transformer and output tube elements and circuits, high voltage rectifier and associated circuits, and others. The distribution of the fields constituting spurious emission depends upon the structure of the equipment as well as upon the surrounding building walls and materials.

Any measurement of spurious emission from equipment needs to be carried on under conditions which are indicative of the possible interference levels in the actual use of the equipment. Yet, to be meaningful, the measurements must be conducted under standardized environmental conditions. This requires therefore correlation of the standard measurement method with the practical interference experienced. To date no general such correlation has been established but data are being accumulated at a rapid rate.

Early approaches to the measurement problem favored the standardization of open field conditions. Of course, the properties of ground and the presence of the measuring equipment itself affect the radiation pattern. Most frequently, measurements were carried on at a distance of about 100 feet, which is about one wavelength at the frequency of 10 MC. Going to a distance of 15 feet as has been proposed, brings us to the edge of the near field or induction zone and will likely give results different from that obtained for the larger distance. At any rate, the field strength measurements, usually made with commercially available field strength meters, did not correlate well with the interference experience besides having the inconvenience of any outdoor measurement even in so-called temperate climate.

A logical alternative is the screen-room technique which is now favored for frequencies up to about 100 MC. It permits measurement of the induction field at distances up to one-sixth of the wavelength as well as measurement of conductive interference. Correlation with open field measurements might be established by replacing the equipment under test by a standard signal generator for which the free space field strength can be computed. However, the screen room is a "live" room with reflection patterns affected by the constructional details of the equipment under test as well as the test equipment itself.

For this reason, there has recently come under very serious consideration the construction of so-called dark rooms which are essentially screened rooms covered on the inside with ideally absorbing material to simulate essentially free space. Unfortunately, absorbing material of reasonable cost is conveniently useful above 400 MCs, thus restricting dark room techniques to UHF and higher frequencies. Data are yet quite sparse and attempts to find absorbing materials which could be used economically down to about 100 MC have not been successful to date. There is hope, however, that this development, if properly supported, can supplement the live-room technique. Obviously, at this point we shall have to pay attention to cost. It might be feasible to create certain primary standard sites which could be used for the calibration of secondary installations of less elaborate design.

We must also keep in mind, that in the various frequency ranges, different kinds of pick-up devices need to be employed. In addition, the detecting device must be designed to properly detect according to the wave form of the spurious emission, which might vary from a sinusoidal oscillation to an impulse type depending on the device producing the spurious emission.

With this as background, I can assure you that the IRE technical committees have been keenly aware of the need for standardization of measurement methods for spurious emission. But obviously, the lack of direct correlation and the inability of analytical treatment of the problem requires extensive experimental data for any serious endeavor of standardization. The Receiver Committee of IRE has been particularly active and has recently published a standard for measurement of interference output of television receivers in the range from 300 KC to 10 MC employing screen-room techniques.

Spurred by the increasing urgency for standard methods of measurement of spurious emission, the Standards Committee of IRE created

in July 1954 an Ad Hoc Committee on this particular subject to coordinate all efforts bearing upon standardization of spurious emission measurements in eight specific technical committees. It became evident rather quickly that coordination alone was not sufficient to assure quick progress. By action of the Board of Directors a new committee was established on "Radio Frequency Interference" and this committee has just recently held its first meeting taking reports from the subcommittee chairmen covering, respectively, the following topics:

- 27.1 Basic Measurements
- 27.2 Definitions
- 27.3 Radio & TV Receivers
- 27.4 Radio Transmitters
- 27.5 Industrial Electronics
- 27.6 Recording Equipment
- 27.7 Mobile Comm. Equip.
- 27.8 Carrier Current Equip.
- 27.9 Community Antennas
- 27.10 Test Equipment

All the subcommittee chairmen have had extensive experience with technical committee work of main committees in these respective fields. They bring to their assignment a broad professional background and can count on support by their respective organization.

The Subcommittee on Basic Measurements has started the preparation of a general standard for measuring spurious emissions to cover radio transmitters, radio receivers, industrial electronic devices and other devices capable of such emissions. Because of its wide scope, the standard will incorporate a variety of methods and will describe the standard arrangements for open field tests, screen room tests, as well as dark room tests. It is expected that this basic standard will be pushed to completion in a comparatively short time, perhaps within the year. However, it must remain flexible to include more precise techniques as they are developed and to include additional factors as experience with experimental set-ups will indicate.

In addition, the Standards Committee of IRE has actively coordinated its activities with task forces appointed by the Engineering Depart-

ment of RETMA which were established in August 1954. It is intended to avoid duplication and to supplement the efforts of the two organizations in their respective spheres of interest.

With respect to spurious emission from industrial electronic equipment, close coordination has been established with the respective committees of AIEE, particularly through Subcommittee 27.5 of the new Committee on Radio Frequency Interference.

All standardization efforts are also correlated with ASA in order to assure as far as possible a single standard for all professional and industrial uses.

Even further than this, IRE has undertaken active participation in, and cooperation with, international standardization activities, in particular with Technical Committee 12 of the International Electrotechnical Commission, or briefly IEC. Committee 12 has the subject Radio Communication and its first subcommittee deals with measurements. At the forthcoming meeting of IEC in London in July of this year, IRE will be represented by its Standards Coordinator, Mr. Axel Jensen, and one of the important topics will be the discussion of an international standard on the measurement of RF oscillator radiation from television and FM receivers. Our own comments have been submitted through the U. S. National Committee.

In similar manner, IRE has undertaken direct participation in the activities of the International Radio Consultative Committee, or C.C.I.R., through the Standards Committee. It has two representatives on the CCIR Executive Committee of the U. S. National Delegation.

In summary, IRE has been actively interested in spurious emission measurements for a considerable time. It has recently, because of the urgency of the situation, taken leadership in creating standard methods through its new technical committee on Radio Frequency Interference; it has actively participated in similar endeavors with other national organizations, as well as with appropriate international bodies.

FCC LOOKS AT SPURIOUS RADIATION

Commissioner E. M. Webster

Mr. Chairman, fellow members of the Panel, members of the IRE and guests: As indicated by the title of my statement, I shall endeavor briefly to delineate the problems relating to spurious radiation from the viewpoint of the Federal Communications Commission.

The emission of spurious radiations by communication and other electrical equipment is a problem of long standing, and it appears that it will remain with us for a long time to come. The optimum use of radio and other forms of communication which are subject to interference from this source depends upon the maintenance of effective measures of interference control.

Proper administration dictates such measures of control be taken as are consistent with the public interest. In many areas of concern, perhaps in most, it does not appear feasible to take measures which will prevent individual cases of interference from arising. In these areas the goal should be to take such steps as are practicable to reduce the number of interference cases to manageable limits, and at the same time assure that proper consideration is given to problems of manufacture, to economics, and to proper operating procedures for the offending equipment and for the systems of communications which are affected. It seems that this may best be accomplished by first encouraging voluntary control by the manufacturer and user, and second by local, state or Federal regulation.

Engineers engaged in the design, development, manufacture, installation, operation or maintenance of communications equipment or electronic devices have long been cognizant of various aspects of the problems of spurious radiation. However, I sometimes wonder whether some have been sufficiently concerned about the interfering aspects of the problem. To design, manufacture and operate equipment which is needlessly capable of causing interference to the various authorized radio services does not appear to be in accord with our delegation to "encourage the larger and more effective use of radio in the public interest"

The Commission has been seriously concerned with most phases of this problem for a long time. It has proposed or adopted rules which specify limits for non-essential radiations from both communications and non-communications equipment. Prior to the time when it became feasible to establish specific limits, our rules contained general requirements that equipment design be in accordance with "good engineering practice", or

with "the state of the art", in order to call attention to the necessity for giving serious attention to these problems in the design and use of communications equipment.

In some cases it has been less difficult for the Commission and the Industry to resolve problems relating to the establishment of transmission standards for communications services than it has been to cope with some of the problems inherent to the control of unwanted radiations from diathermy machines, industrial heaters, arc welders, and a miscellaneous group of devices which we have called "restricted radiation devices".

The need for national and international regulation of emissions from radio stations became apparent as far back as the early part of 1927 when the question was considered in connection with the International Radio Telegraph Conference of Washington.

The years following 1930 saw a steady increase in the use of diathermy machines and electronic devices, with a consequent increase in interference which they caused to radio services. Of particular concern were the safety services such as aviation, marine, police and fire. This prompted the Commission to initiate studies looking towards the regulation of such devices.

International consideration of the subject developed when the delegates participating in an Inter-American Conference held in Havana, Cuba, in 1937, made an administrative agreement looking toward regulation of all non-communication apparatus which generates radio frequency energy as an essential to its operation.

The Commission, in 1938, had completed extensive studies of the problem of interference caused by the operation of restricted radiation devices, diathermy machines, induction heaters or furnaces, vacuum tube bombardiers, and other industrial electronic devices of higher power. In November of that year it adopted what is now Part 15 of the Rules and Regulations, relating to a limited category of restricted radiation devices. This was the first rule making action of the Commission which looked toward the possible solution of overall interference problems created by the operation of unlicensed and non-communications devices.

Subsequently, at an International Conference in Santiago, Chile, in 1940, the delegates agreed that their respective countries should adopt measures to alleviate interference caused by radio frequency equipment which may radiate energy capable of interfering with the reception of regularly authorized transmissions.

In the latter part of 1943, the radio industry set up an organization known as the Radio Technical Planning Board, to work with the Commission in an attempt to solve many of the involved problems relating to frequency allocations, including the problem of non-communications radio frequency devices. On the basis of testimony and recommendations of representatives of the Radio Technical Planning Board and other interested parties, the Commission recognized the need for diathermy and other industrial devices and allocated three frequencies for their general use and development in its report of May 25, 1945.

Rules Governing Industrial, Scientific and Medical Equipment were adopted by the Commission on May 8, 1947. The International Radio Conference held at Atlantic City during the same year, recognized the fact that the uncontrolled operation of radio devices used for industrial, scientific and medical purposes could seriously interfere with vital communications, and made the first provision for the allocation of frequencies for such purposes on an international basis. The three frequencies allocated were slightly different from the three adopted by the Commission, but the Commission's Rules, Part 13, were later modified to conform thereto.

The Commission's present Rules Governing Restricted Radiation Devices, Part 15, provide in substance that any apparatus used for communication over very short distances which generates a field not exceeding 15 microvolts per meter at a distance of 157,000 feet divided by the frequency of operation in kilocycles is not subject to the other rules of the Commission, provided that no objectionable interference results to the reception of authorized radio signals.

Recognizing the inadequacy of present Part 15 with respect to the regulation of the many thousands of unlicensed "low-power" communication devices, and several million "unintentional" radiators such as carrier current systems, radio receivers, ignition systems and electrical appliances, the Commission on April 13, 1949, issued a notice of proposed rule making setting forth the administrative and engineering factors to be considered in amending the rules.

Studies relating to the problem of radiation from TV and FM receivers were undertaken, and in the fall of 1950 joint meetings between members of the Commission's staff and representatives of the Radio-Television-Electronics Manufacturers Association (RETMA) were held to discuss the oscillator radiation problem. I shall leave to Dr. Baker to relate the action taken as a result of these meetings.

In December 1953, the Commission requested the Joint Technical Advisory Committee (JTAC) to study the question of spurious radiation to the end of developing methods and standards for dealing with practical problems which the Commission faces in its regulatory functions. The JTAC accepted the task and have coordinated their activities with the RETMA and the IRE in setting up committees to cover the broad field of interest.

On April 14, 1954, the Commission issued Notice of Further Proposed Rule Making In the Matter of Amendment of Part 15 of the Commission's Rules Governing Restricted Radiation Devices. This proposal would govern the operation of both incidental and restricted radiation devices. The first named category includes devices in which the production of radio energy is unintentional or incidental, such as electric power, lighting and ignition devices. Restricted radiation devices embrace carrier current communication systems, signal generators, beat frequency oscillators, radio receiver oscillators, and other low powered radio frequency generators. Comments have been received on this proposal but rules have not as yet been adopted by the Commission.

A second matter which is currently the subject of rule making has to do with the type acceptance and type approval programs for communications equipment. These programs are designed to simplify our licensing processes by setting up lists of accepted or approved equipment to which reference will be made by applicants for radio station licenses. Type approval, which is required for certain marine safety services and for the citizens' radio service, is based on tests performed by or under the direction of the Commission's Laboratory. Type acceptance is based on certified test data submitted to the Commission by the manufacturer. Such data include measurements of various forms of spurious radiation, in addition to other matters required by the rules. These rules are under study by a special Panel of the RETMA and certain user groups, in an effort to have some of the Commission's proposed rules modified before final adoption.

SYMPOSIUM ON SPURIOUS RADIATION

QUESTION AND ANSWER PERIOD

Ralph Bown (Chairman): It is not often that engineers find a member of the Federal Communications Commission as a sitting duck in front of their guns so I beg that you do not concentrate your fire on Commissioner Webster. However, this does not mean that questions cannot be asked which involve him and as a matter of fact I think that perhaps a good way to start off the discussion is to ask Commissioner Webster if he would be willing to respond to a question which he provided himself. Perhaps he didn't expect to have to answer it himself but I would like to ask him if he wishes to say a few words about a question which reads as follows: "What can be done to make the play of free competition work towards a reduction of spurious emissions or responses in equipment without intervention by Government?"

E. M. Webster: Mr. Chairman before I reply, if anyone has a heater in the room, whether it would cause interference or not, I wish he'd bring it up.

I asked that question or at least I put that question down on a list, and as you can tell it has to do with the role of the manufacturer. Now I have never been a manufacturer and maybe that's what makes me competent to answer. But I have one or two thoughts in mind in regard to that question. In the first place, I'm giving my personal opinion. You know we have six other Commissioners and they have some personal opinions too and I try to be rather careful at times when it comes to giving my own personal opinion.

I have heard this subject discussed for I don't know how many years, but it has been a live one, and it seems to me that there's something wrong somewhere when it takes so long to come to conclusions. The engineering industry -- the engineers, the radio engineers, the electronic engineers in my opinion can do most anything. You've all heard the statement that the impossible we do tomorrow, and in my experience with radio engineers, that's exactly what happens. I take my hat off to the communication engineers, the electronic engineers, and what they've accomplished. It just seems to me that there's something wrong when it is said the engineer hasn't solved this problem. I think he has solved it. I think there's another element in this that has to be looked at. It's not in my opinion a truly engineering problem. There is a policy-executive problem that arises in the midst of it, and that involves the executive and policy levels of the industry. Now the policy and executive groups of the industry don't hesitate one minute to come down to Washington to discuss problems with the policy and executive group of our Government -- in this case the Commission.

I have a great deal of respect for Dr. Baker and what he said. I think that there's a lot of meat in his paper and I want every Commissioner to read it. Personally, I may not agree with everything that he has said but that's immaterial for the moment. He has talked about things I think the executives should talk about and I put the Commissioners in that category. They are not the engineers of the Commission -- they are on the executive level with the President, Vice-President and so forth of companies and they are the people that ought to talk together about this problem. The engineers of the industry shouldn't come down and talk to the Commissioners. They should come down and talk to our chief engineer and our staff. It's the Presidents and the Vice-Presidents that I want to see come down to Washington and talk to us about policy.

With any electronic apparatus we end up as its being a system. You have a transmitter on one end and a receiver on the other and it just seems to me that it's too bad that the law only covers one end of it. We write volumes -- you know what our rules look like -- we write volumes about one end of the system, and there isn't very much about the other end of the system. That is left to industry. And in my opinion industry, as Dr. Baker has intimated, has a very great and high responsibility for what it should do on that end for the unsuspecting purchaser. And it isn't always the television buyer who's unsuspecting -- I have been out through the country and I have found company after company that has told me that the equipment they bought six months ago is no good. They put it on the shelf and bought somebody else's. The manufacturer in that case was not acting in the public interest.

Another thing that I'd like to see -- Dr. Baker talks about the cost. I'm alive to that, that's a big economic problem that you're faced with. But nobody comes down to us and says not only should you look at the engineering in conjunction with the industry -- (that it is a joint responsibility of the Government and industry to look at this from an engineering point of view) but also let's all get around the table and talk about the cost. If that point is raised, immediately they will tell you that that is a competitive angle, that we can't talk about the cost because it reveals competitive material. But if you're going to talk about this problem it just seems to me that you've got to get down in the dirt and talk about every single portion of it. That is what's been worrying me over a long period on this particular subject.

Now as I said in the beginning, I'm not a manufacturer and everything that I've said can be shot full of holes by a manufacturer. I don't

know that that answers the question. I've thought of the answer sitting here after listening to these papers. I didn't come up here with any preconceived idea of what the answer should be but I think that the public should know the facts about this interference problem and I don't think the public does today. I'd like to see this aired publicly so that everybody knows what the difficulties are -- and it's more than engineering. I don't think the public knows the difficulties on the economic side, if there are difficulties.

Of course, we're always having trouble and probably most of this is on account of the non-conformists. You know that most laws are made for that purpose; not to take care of the man who will conform and do right, but for the man that's trying to cut the corners and do what he shouldn't do. I think that is probably part of our problem, to bring that man in line. I know that RETMA has its difficulties. I hear everyday from members of RETMA the difficulties they have, the disputes they have among themselves. But it all boils down to economics and I'd like to see these Presidents; I'd like to see a panel up here in front of you engineers, let's say composed of Mr. Folsom or even General Sarnoff himself or Price of Westinghouse or Galvin from Motorola or Commander McDonald from Zenith; maybe Stanton from CBS, he's a manufacturer too. Not that they're doing anything wrong, because I think that they're doing the very best they can. They're turning out good equipment. But I think those people or people in that class could give you some very illuminating material to chew on. That is my position in regard to that particular question.

Mr. Crayton (General Electric Company): I would like to question what technical term should be used. The point is raised that people don't know what they're talking about. I'd like to refer to the term spurious emissions and to the term extraneous emissions. Spurious and extraneous mean very much the same thing.

Ralph Bown: I think I would like to attempt to answer that one myself if I may. The term spurious has been used this morning primarily because that is the way in which the title of the Symposium was stated. There are obviously synonyms which are useful, such as extraneous. I think you will find that this word is used primarily as a synonym rather than as a substitute or for a different meaning. It seems to me that this is something which would get down to too fine a distinction for us to attempt to wrestle with in a group of this size.

Question from Floor: I think in the electrical industry a great deal of public education has been accomplished by giving some kind of Underwriters Laboratory seal of approval. Could a similar process be adopted for the control of spurious radiation? Can there be some discussion of the matter by this panel?

W. R. G. Baker: Of course there is a great deal of difference between the Underwriters Laboratory and any certification laboratory that could be established either by the industry, the FCC or jointly. The Underwriters Laboratory has real teeth in its ability to enforce. There are some states which would permit the sale of a product either in the white goods line or any hard goods consumer line without having an Underwriters tag on it. A company that doesn't use the Underwriters approval takes serious liability risks if somebody's injured or something else happened with an appliance no matter what it may be. The certification laboratory, then established by the industry or by the Commission or jointly, would have to have some means of enforcing. And that is a very difficult problem when you are enforcing a product that is as complicated as television or radio, where you're building many many thousands a day. And how far do you carry the checking of the quantity and all the rest of it. The Underwriters Laboratory is a distinct advantage to the appliance manufacturer but I think this is an entirely different animal. We studied the certification problem in great detail and so far have not come up with what seems to be a workable answer.

Ralph Bown: I'm going to turn to Art Loughren for the question which he handed to me himself, and ask him if he would like to attempt a few remarks on it. It's a very pertinent question and it reads this way: "Why can't apparatus producers be expected voluntarily to conform to suitable spurious radiation emission standards?"

A. V. Loughren: First let me make the comment that it wasn't part of my expectation that I be caught with my own weapon. But I think there is a difficulty to which the question is directed that is basic to a free enterprise economy. The maker of a product to be offered for public sale in a free enterprise economy is concerned with satisfying the requirements of the market and if he fails to satisfy these, his product doesn't sell. He may have some other requirements to satisfy also. The case of the Underwriters listing was mentioned a moment ago and this is enforced not by the purchaser ordinarily but by Governmental agencies usually on a local level acting to protect the purchaser. In the case where a manufacturer, not faced with such a requirement as the Underwriters requirement, adds cost to his product to make an improvement which will not be sought after by the purchasers, then his competitor who has not added cost in this same fashion can sell the competing product cheaper. If the difference in value, from an overall public interest point of view, is not a difference which is apparent to the individual purchaser at the time that he is making up his mind which of two competing products to buy, then there is no reason why the purchaser should be expected to buy the more expensive and socially more desirable product. The manufacturer who

puts the added cost into his product may find himself out of business. It seems to me that the manufacturer who would make the socially more suitable product needs some measure of assistance to insure that the socially desirable thing be done by his competitors as well. It seems to me it's undesirable to restrict the freedom of the market place in any fashion other than to achieve a needed social end. Usually when some small group attempts to say this is better for the customer and therefore we must make him have it, that small group may run the risk of being completely wrong. I speak here from some experience because I was on one occasion a member of a small group that was quite convinced that the public ought to have what we call high fidelity equipment and I'm sure that I hurt the sales of my own company the year when I and a few others put on an internal campaign to get some of the company's merchandise into this class. The public didn't think much of what we said the public ought to have. So here was a small group that did something wrong and the public turned us down. There are dangers in reducing the freedom of the market place but there are times when perhaps it simply must be done and this may be a case of that sort.

Ralph Bown: Perhaps I shouldn't allow myself to go scott free. Since I'm in a position to do so I'll try not to. Like Commissioner Webster, I'm not a manufacturer either, at least my personal job has never been in manufacturing. I feel that I would like to reinforce somewhat the remarks which he made and I would like to say it this way. This matter of spurious radiation is not a thing which can be licked and taken care of by engineers alone. Certainly in the most difficult forms in which it comes up, namely where you have apparatus which is widely sold to the public on a very competitive basis, the engineers can only go so far and, even though they know the right answers, they are not necessarily in a position to apply them. I believe this means that the policy and executive levels of the industry must look at and deal with this question in a responsible manner. The engineers I'm sure stand ready to help them to any degree possible but they can't do it all alone. Is this a fair statement of it, do you think Commissioner Webster?

E. M. Webster: I think so.

R. F. Guy: Mr. Chairman I'd like to suggest to a gentleman that represents the Commission that they might make a contribution which would be helpful to arriving at a common solution and that is, to give more publicity to the interference problem. Service by service perhaps indicate the seriousness of interference of various types which we can classify as extraneous or spurious, so that they can be publicized to a much greater degree than they are now. We hear these problems of interference spoken of in more general terms, I might say in generalities.

It would help to have more specific information. In one case in point I know that JTAC requested from the Commission analyses of interference complaints which the Commission possibly compiled but so far as I know that has never been published in a manner available to our public use. I suggest that that sort of stuff be publicized in a manner so that we can appraise the problem with them.

E. M. Webster: May I comment just a second. I certainly do agree with that and I think the Commission should do a great deal more than it does. I have been one of those in the Commission since I have been a Commissioner that has suggested from time to time that we do more and give more publicity to things. This gives me a chance to say that unfortunately we don't have the staff, the funds and so forth to do things. I made that very strong statement the other day before the Congressional Committee. I felt that the Commission was not doing the job it should be doing because we did not have the facilities to do it. One of them in my opinion is to give more publicity to things. Of course on the other hand there are companies that come running right in as soon as you give publicity to something thinking that you are taking a crack at their particular equipment although you don't name it. You have that other side of the picture too, but I agree thoroughly that there should be more publicity given.

Ralph Bown: We're approaching our closing time and I would like to ask Mr. Browne whether he or Mr. Coffey who accompanies him might have any comment which they would like to make on the discussion which has occurred this morning.

G. C. W. Browne: Thank you Mr. Chairman, I don't think there is anything that I can comment on from our standpoint. I am not a Commissioner nor have I six other Commissioners as Commissioner Webster has with him. I do have a deputy minister to whom I report and he is the executive head of our department and he of course reports to the political head who is a member of the Cabinet. That is the setup in Canada as most of you probably know. I have attempted to outline in my talk what we have done so far in Canada and we, of course, as I stated, would be glad to cooperate with the industry in working out these many problems that are still to come. I'd say that we enjoyed their cooperation in the past although we do hope that things will get moving a little faster from now on so that we may setup the remaining standards that are yet to be dealt with. Thank you very much.

Ralph Bown: Thank you. I am not going to ask George Sterling or Ed Allen who so kindly consented to come up and sit with us to say anything this morning. I wish to thank them for coming. I would like to thank in the name of the audience the members of the panel who have given these discussions and I would like to declare the meeting adjourned.

A DEVELOPMENTAL POCKET-SIZE BROADCAST RECEIVER EMPLOYING TRANSISTORS

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ABSTRACT

This paper describes a pocket-size developmental AM broadcast receiver which utilizes eight junction transistors. Its performance is comparable to that of conventional personal receivers. Emphasis has been given to developments which contribute to stability with respect to temperature, battery voltage, and variations among transistors. The superheterodyne circuit employed uses a single-transistor frequency converter to perform the functions of both mixer and oscillator. Refined detector and automatic-gain-control circuits and an audio amplifier embodying further develop-

ment of the principle of complementary symmetry are incorporated. Reduction in physical size and battery requirement, as compared to conventional receivers, is substantial.

The circuits are described in detail and certain aspects of components and of physical arrangement, which contribute to the small size, are discussed. Detailed performance data are also included.

(The full paper appears in the June 1955 issue of PROCEEDINGS OF THE I. R. E.)

PROGRESS IN FERRITE COMPONENTS
FOR TELEVISION AND RADIO RECEIVERS

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Summary

The reported improvements in ferrite components for television and radio receivers are based on better materials, novel circuits, or the utilization of normally unwanted properties. An example of each type is given.

A new high efficiency ferrite (Class W-02) perceptibly out-performs all others for color television flyback transformers. A ferrite nearly as efficient (Class W-03) is considered for monochrome television flyback transformers. The performance criteria of these materials and the methods of testing and evaluation are outlined.

A simple compact omnidirectional ferrite antenna circuit provides a possibility for inside automobile antennas and miniaturized aircraft antennas.

Ferrite beads and cups acting essentially as resistors for UHF only and as shields are incorporated in discoidal feed-through capacitors. Acting as integrated filter units of small size, they have a minimum of 40 db insertion loss.

A. Introduction

This report describes improved ferrite components for television and radio receivers. The knowledge of these improvements may be advantageous in the development of military equipment if these components are used in the same class of application and in the same frequency range.

Progress in the use of ferrites may be based on the following principles:

- a) Better materials,
- b) Novel circuit arrangements, or
- c) Utilization of normally undesirable properties.

An example of the application of each principle will be discussed.

B. Improved Ferrites (Classes W-02 and W-03)

This is an exemplification of the first principle: better materials. It should be understood that there is no better or best ferrite, unless the application is specified in all pertinent points. Ferrite "A" may be the best available material, or even the ideal material for purpose "a", but it may be of no avail for purpose

"b", where material "B" is the best while "A" is worthless.

A ferrite for antenna rods operating at very low power levels in the broadcast range requires an initial permeability of between 100 to 300, a low temperature coefficient of μ_0 , a high Q, a low temperature coefficient of Q, and negligible aging effects. In the case of a flyback transformer the above parameters are irrelevant. The sought-for conditions are: high flux density, low losses at large flux densities, zero or negative temperature coefficient of the losses at large flux densities, low B_T and high incremental μ at large flux densities, under heavily biased conditions and at elevated temperatures.

In high energy recovery systems presently used with television flyback transformers, it is true, a frequency spectrum as that depicted in Figure 1 is encountered. However, the ferrite used in the flyback transformer is a non-linear element, and in spite of linearization by air gaps, it inherently will not distinguish between the frequency components shown. The ferrite is operated in the range of "irreversible" domain wall movements. It will respond to the momentary force dH/dt as determined by the saw tooth deflection conditions. The frequencies of interest, then, for the core material will be the basic sweep frequency of 15.75 Kcps and of 55 Kcps, the latter being established by the half cycle free oscillation during the retrace.

The power handling capacity of any magnetic device is given by $B^2 f^2$. The specific losses (same volume) P are nearly proportional to B^2 . Assume two materials: (1) with maximum usable B_1 and with losses P_1 at B_1 , and (2) usable up to $B_1 \times \sqrt{2}$ and having only $P_1/2$ losses at B_1 . There are at least two possibilities to gain from this better material:

(a) By using only $1/\sqrt{2}$ the cross-section of the core material (2) as compared with the original material (1). Then the magnetic losses are maintained, but the mean length of the winding is reduced, decreasing the copper losses; or

(b) To maintain the same core cross-section, but use less turns. This again results in a reduction of copper losses. This second approach may sometimes be impractical because of prohibitive per turn voltages.

Depending on the design goals, additional gains can be made with the better material, particularly with respect to copper losses and distri-

buted capacitance. The higher B material should have the same or higher effective permeability to draw no more or to draw less magnetizing current.

Figure 2 depicts the normal magnetization curves for Class W-01 and Class W-02 ferrites, the latter being perceptibly superior. In both cases the magnetization curves are shown for 25°C and 100°C. There is a safe margin for practical design purposes when 100°C magnetization curves are used, since the temperature of a well designed flyback transformer will not exceed 95°C. When using the Class W-02 material, having much lower losses, the temperature in the magnetic material will be from 10° to 20° lower. Figure 2 indicates that the W-02 ferrite, even at 100°, is superior to W-01 ferrite at 25°.

The advantage of W-02 material, moreover, becomes more obvious in Figure 3. Here the specific losses P, measured in $\mu\text{Wcm}^{-3}\text{cps}^{-1}$, for three fixed flux densities of 500 Gauss, 1000 Gauss and 2000 Gauss, are plotted against frequency up to 400 Kcps. At 16 Kcps the losses in W-02 are less than one-half of the losses in W-01. Even at 50 Kcps W-02 has about half the losses of W-01. The curves shown in Figure 3 were obtained from measurements using pulsed conditions with a duty cycle of about 2% to avoid masking of the results by heating effects.

Figure 4 compares the losses of W-02 and W-01 material in a B (in Gauss) vs. P (in $\mu\text{Wcm}^{-3}\text{cps}^{-1}$) diagram. Whereas W-01 has a slightly negative temperature coefficient of the losses (B being constant), the temperature coefficient of W-02 is practically zero. It is essential that the temperature coefficient of losses be zero or even negative. If not, the heating process in the flyback transformer will not stabilize, and the material will develop excessive heat. Cases are known in which the ferrite had a large positive temperature coefficient of losses which resulted in so much heating in the flyback transformer that the Curie point was approached and a complete "slump" occurred. Checking of temperature coefficient of flyback transformer ferrites at high flux densities is an essential feature of production batch testing.

As explained in several papers (See Bibliography Nos. 1-6), d.c. bias is present in the magnetic core of the flyback transformer. Depending upon the mode of operation, the efficiency of the circuit, and the additional drain taken off the flyback circuit, the magnetizing current is more or less unsymmetric about the zero axis of the normal magnetization curves.

The following curves (Figures 5, 6 and 7) pertaining to the incremental permeability Δ/μ were measured with Class W-03 ferrite, which is a cheaper grade of Class W-02 ferrite, W-03 having as a lower limit a 20% lower μ_{max} and 20% higher losses. In other respects both ferrites are intrinsically the same. A pair of representative U-pieces of the shape U-2218 was used for these

measurements. The mean ferrite length of a pair of these U-cores is 6.54 inches.

To minimize the number of diagrams and to avoid confusion by showing too many curves on one illustration, only a few data obtained at 100°C are plotted. Despite the fact that the test conditions disfavor the lower loss W-03 ferrite (it gets less warm) in comparison with W-01 ferrite, the advantage of the new material comes clearly to light in the figures to be discussed now.

Two cases of biasing conditions are distinguished:

(a) A constant current bias (Figures 5 and 6) as parameter. A constant current bias corresponds essentially to the flyback transformer operation.

(b) A proportionate d.c. bias that is a certain constant percentage of the current swing (Figure 7). This premise is more closely approximated in a pulse transformer, where the larger the duty cycle, the higher the percentage of d.c.

One of the major problems in flyback transformer design is the avoidance of "slumping" after warm-up. The "slump" is caused by diminished incremental permeability affecting the mutual inductance of the flyback transformer. Here an analogon to the low frequency limit of a transformer exists. Since the loading deflection yoke inductance is practically invariant with temperature and the shunting mutual inductance of the flyback transformer decreases with temperature or flux density, there will be a certain temperature or flux density where both inductances become equal. Then a "slump" of 50% occurs (if all other conditions remain constant). The higher the operating flux density, the lower is this "slumping" temperature.

Unfortunately, since practically no flyback transformers are compensated for d.c. bias, the mutual inductance in question cannot be made 20 or 30 times larger than the yoke inductance. The reason is twofold:

(1) The leakage inductance would be too high and decrease the efficiency, and

(2) The d.c. bias in a flyback transformer shifts the operating point into the flat saturation part of the normal magnetization curve, resulting in small incremental permeability and small ΔB . To remedy this situation an air gap has to be introduced. Because of the non-linearity of the underlying relations, the optimum air gap is best determined experimentally as a function of temperature and flux swing.

In Figure 5 the incremental permeability vs. air gap is contrasted for W-03 and W-01 ferrite. The parameters of the curves are d.c. bias of 0, 1, and 2 Acm^{-1} . Constants of the diagram are $T = 100^\circ\text{C}$ and $B = 2000$ Gauss. For a bias of 2 Acm^{-1} for Class W-03 ferrite, the maximum Δ/μ is 240 at an air gap ratio of $\Delta l/l = 2 \times 10^{-3}$. Class W-01 ferrite reaches a maximum Δ/μ of only 145.

Selecting from Figure 5 an optimum air gap of 8 mil for Class W-03 ferrite at 1 Acm^{-1} bias and at 100°C , in Figure 6 Δ/u vs. B for a constant air gap of 8 mil is plotted for 100°C for both kinds of ferrites, d.c. bias being the parameter. *

When replacing a ferrite core of the old material with one of the new, the new material is not fully utilized. A new design is then recommended. Thus, an older flyback transformer using a W-01 core was equivalently replaced by a new one with a W-03 core having 65% of the ferrite volume and about 68% of the copper volume. Leading manufacturers using W-02 cores for color television have reduced component cost by utilizing the increased efficiency of the W-02 ferrite.

Figure 7 provides curves significant of the previously mentioned case (b); namely, proportionate bias. In order to cover severe cases, measurements were made going to an extreme case of unsymmetry. This is in accordance with the always helpful engineering principle that a complex situation becomes more transparent if the limiting cases are considered. To investigate the effects of proportionate bias on the magnetization curve, a diode was inserted in the driving circuit. Thus, only half cycle waves were applied to the core, resulting in a d.c. bias of $1/\pi$ of the peak current. Although in practice this extreme case will seldom be encountered except in some pulse transformers, the curves as obtained still show the pronounced superiority of the new material as compared with the old. All pertinent constants are denoted in the graph.

C. New Circuits (Omni-Directional Ferrite Antenna)

There are numerous cases in which the merit of ferrite rod antennas, namely, small size and freedom from electrostatic interference (if properly shielded), is defeated by the directivity of this type of antenna. In order to make an omni-directional ferrite antenna (for horizontal magnetic polarization) the following two principles were combined:

(a) There must be a 90° difference with respect to time and space when two figure eight dipoles are combined into one omnidirectional antenna.

(b) In critically coupled circuits the voltages in each circuit are 90° out of phase with each other.

The resulting configuration is sketched in Figure 8. The two antenna rods cross each other at 90° . They are positioned in such a way that the coupling coefficient $k = 1/Q$.

A portable radio was satisfactorily converted with this input circuit. There was excellent reception for all stronger stations when the radio was placed on the dashboard inside of an automobile. However, since the modified radio did

not have an R-F amplifier stage (as the built-in car radio had, used for comparison), weak stations were poorly received. The addition of a ganged preamplifier would remedy this situation by improving the S/N ratio at the mixer grid. The advisability of ganging four tuned circuits is a question for the set manufacturer, not for the ferrite producer.

D. Utilization of Normally Unwanted Properties ("Ferri-Cap")

Caused by domain wall resonance, ferrites have very low permeabilities and sometimes extremely high losses in the UHF range. Ferrites are, therefore, often considered useless in this frequency range. Furthermore, some ferrites, in spite of magnetic excitation, become capacitive in the UHF range. The Q's are often less than one. In other words, a bead of ferrite on a conductor represents essentially a zero impedance for d.c. and a predominantly ohmic impedance (with an inductive or capacitive phase angle) at UHF. With the proper selection of ferrites a small bead can represent 50 to 100 ohms impedance at UHF. If, then, a low pass filter incorporating a discoidal feed-through capacitor (having, let's say, about .5 ohms in the UHF range⁷) is built with a ferrite bead, an insertion loss of at least 40 db can be expected. Such a combination of ferrites and dielectrics for UHF filters is shown in cross-section in Figure 9, and is called a "Ferri-Cap". The ferrite bead and cap, together with the metallic eyelet, surround and shield the dielectric disc and the center lead, thus preventing radiation coupling that might otherwise induce energy in the feed-through capacitor and its leads.

Figure 10 compares the physical appearance of the "Ferri-Cap" (constituting an integrated small size filter unit) with a "standard" UHF low pass filter taken from a UHF tuner. This filter consists of a tubular capacitor (of the same capacitance value as the "Ferri-Cap") and a choke of .223 μH . The insertion loss vs. frequency obtainable with these units is contrasted in Figure 11. It becomes apparent that the "Ferri-Cap" has at least 40 db insertion loss above 200 Mcps, whereas the conventional filter (because of parallel resonances in the tubular feed-through capacitor) exhibits at certain frequencies up to 16 db less insertion loss than the "Ferri-Cap".

Obviously, the "Ferri-Cap" is superior to "standard" filters, not only with respect to filtering performance, but also with respect to space requirements and ease of assembly. The "Ferri-Cap" should prove advantageous not only in UHF tuners, but also in military equipment.

D.C. currents up to 400 mA or external d.c. fields up to 500 Gauss will reduce insertion loss of the "Ferri-Cap" by not more than 1 db through the whole VHF and UHF range.

* The measurements were made with sinusoidal flux.

The writer wishes to express his appreciation for the work and contribution of Mr. L. Beaudoin (Example A), Mr. B. Budny (Example B), and Mr. R. Wickline (Example C).

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Spectral Analysis of Driver Current and Flux in TV Set No.11

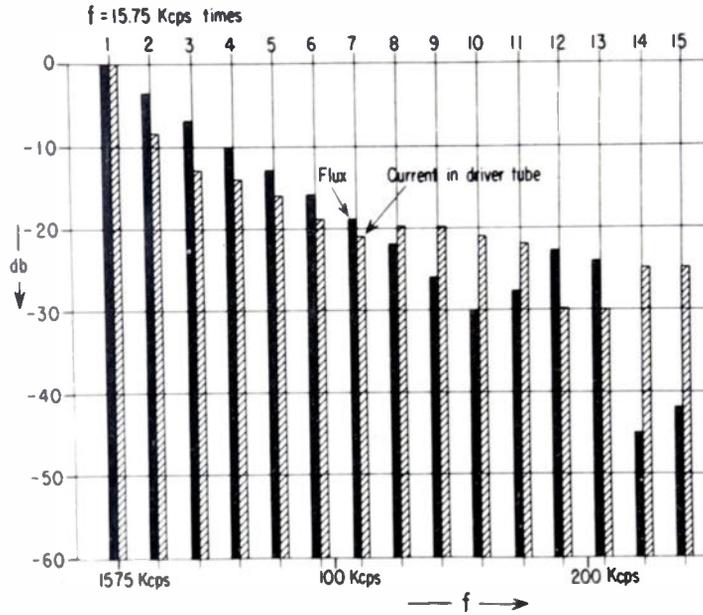


Fig. 1
Frequency Spectrum Measured on a Flyback Transformer.

Normal Magnetization Curves

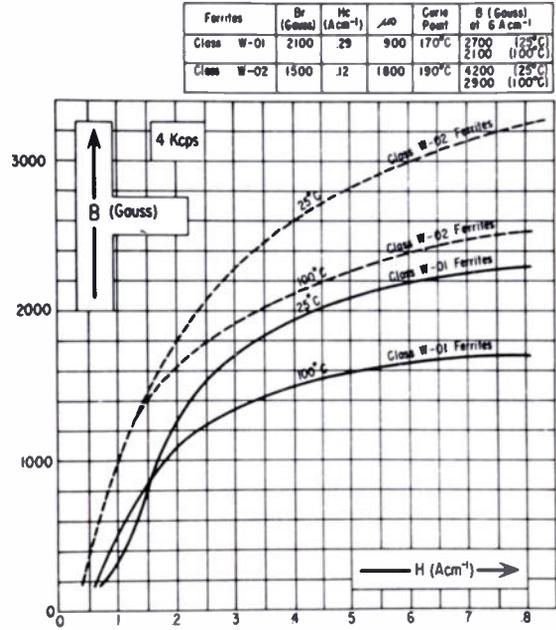


Fig. 2
Normal Magnetization Curves for Class W-01 and W-02 Ferrites.

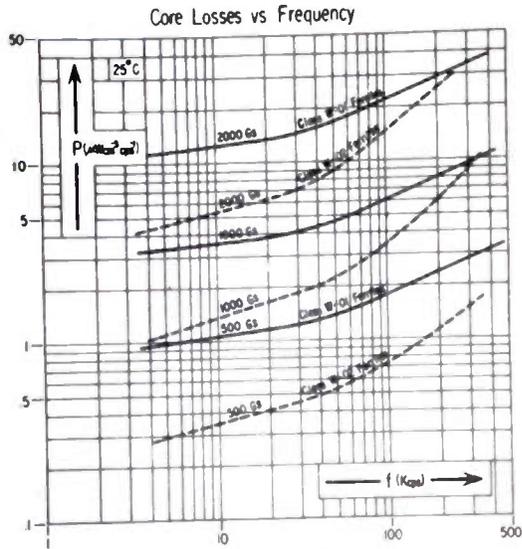


Fig. 3
Frequency Dependency of Hysteresis Losses for Class W-01 and W-02 Ferrites.

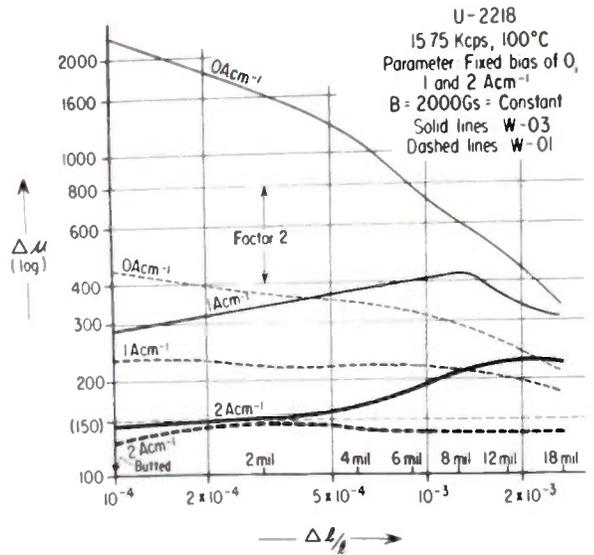


Fig. 5
Incremental Permeability vs. Normalized Air Gap for Class W-01 and W-03 Ferrites. Constant d.c. Bias as Parameters. $T = 100^\circ C$.

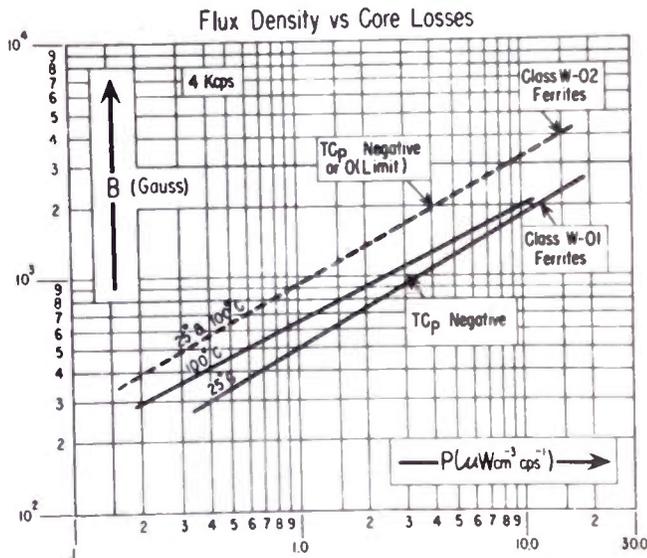


Fig. 4
Hysteresis Losses as Function of Flux Density for Class W-01 and W-02 Ferrites.

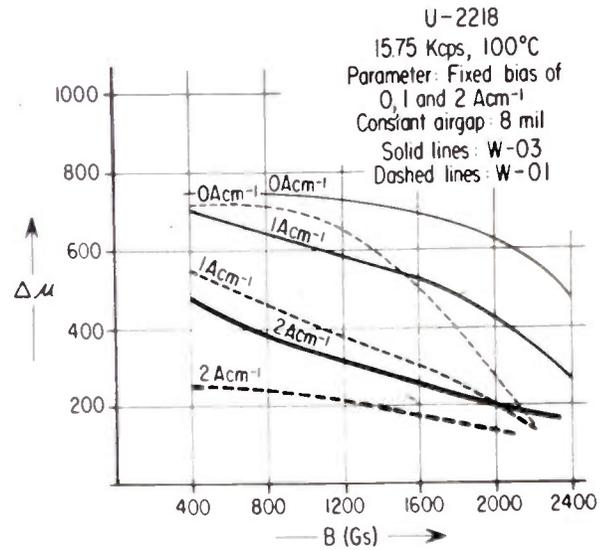


Fig. 6
Incremental Permeability vs. Flux Density for Class W-01 and W-03 Ferrites. Constant d.c. Bias as Parameters at $T = 100^\circ C$.

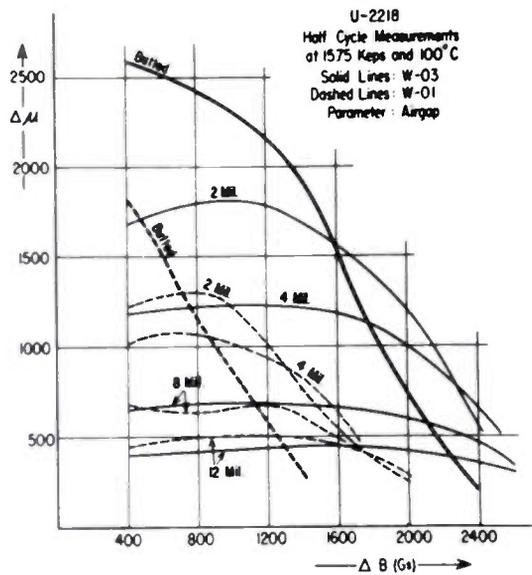


Fig. 7
Incremental Permeability vs. Flux Density for
Class W-01 and W-03 Ferrites.
Proportionate $(1/\pi)$ d.c. Bias at $T = 100^\circ \text{C}$.

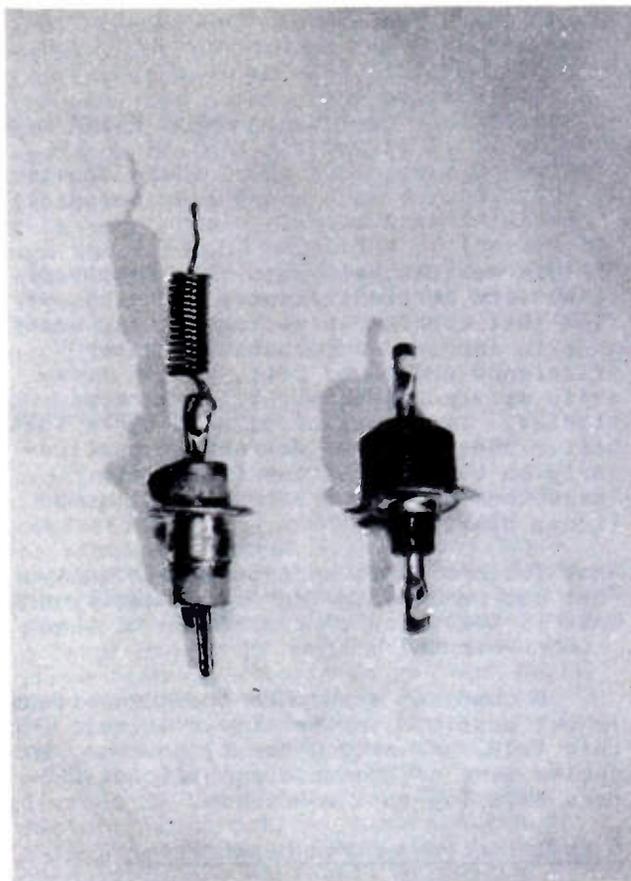


Fig. 10
Photograph Contrasting a "Ferri-Cap" and a
"Standard" UHF Filter.

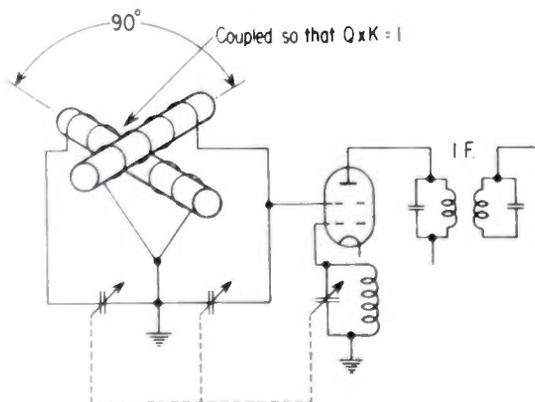


Fig. 8
Semischematic Diagram of an Omnidirectional
Ferrite Antenna.

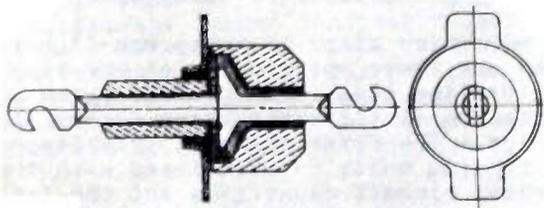


Fig. 9
Cross-Section of a "Ferri-Cap".

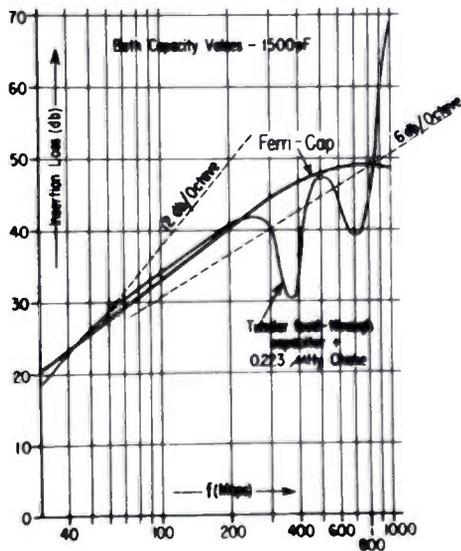


Fig. 11
Insertion Loss Vs. Frequency for a "Ferri-Cap"
and a "Standard" UHF Filter.

WHAT PRICE - HORIZONTAL LINEARITY

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If we look back upon the history of television horizontal sweep circuits, we find that considerable progress has been made in improved components, greater efficiency and lower cost, but in linearity we are now essentially where we started. Yet, most of us will agree that better linearity is desirable - particularly in our "top of the line" sets. These considerations pose several questions; namely,

What factors limit horizontal linearity, What can be done to improve linearity and What is the price or complexity of these improvements?

A complete answer to these questions is not possible in the time available for this talk. We will attempt, however, to review some of the most significant factors relating to the problem.

Classification of Non-Linearities

As an aid to the discussion, let us classify several general types of sweep non-linearity in terms of performance and relate these to the parameters or effects which cause the difficulty.

The first general type which may be defined is that which possesses a different sweep speed between the first and last halves of the sweep period. This type of non-linearity is primarily due to improper adjustment of drive and bias on the damper tube which in turn are dependent upon the voltage drops caused by diode resistance during the diode conduction period, improper tapping of the diode on the transformer, excessive power losses or variations during retrace, appreciable variations in transformer or yoke inductance, changes in diode bias by drift or misadjustment of the linearity circuit, and changes in the output tube average current. In general, it will be found that this type of non-linearity can be minimized at the design center by properly adjusting the damper tap position on the output transformer. Variations of the above-mentioned parameters with line voltage, age, and drift, however, may result in significant contributions by this type of non-linearity.

The second type of non-linearity which may be classified is that which appears in the middle of the scanning period. This change is primarily related to the relative times at which the diode stops conducting and the output tube starts conducting. Any circuit variations which cause the relative conductions of these tubes to vary tends to cause the sweep to slow down toward zero velocity for a short period of time or, to a lesser degree, to speed up in the mid-scan region. This is particularly critical where the starting time of conduction of the output tube is very nearly identical to the time where the damper tube goes out of conduction. Any small change which causes the output tube conduction to occur later or the damper tube conduction to end earlier results in the greatly magnified distortion of sweep linearity. As the conduction periods appreciably overlap, this type of non-linearity becomes far less critical because of the inherent feedback built into the system for this adjustment. Of the parameters which affect this type of non-linearity, it has been found that the screen voltage and the bias voltage on the output tube are particularly significant. Variation of either voltage tends to change directly the relative time of output tube conduction. As shown later, these parameters may be stabilized, but even so, represent the most critical variables for this type of non-linearity. Other parameters which may affect mid-scan linearity are variations in the grid drive waveform, variations of the output tube transconductance, variations in the damper bias, variations in the damper position on the transformer, and variations of power loss during retrace period. Most of the latter type of parameter are fairly easily controlled and tend to remain relatively stable with time and voltage variations.

A third class of sweep non-linearities are those which occur at the start of the line scan. Two general forms of variation of the sweep occur during this period. The first and most troublesome is ringing which is associated with the various circuit capacities and the leakage reactances of the horizontal output

transformer which are triggered by the flyback. Generally speaking, the major forms of ringing are those associated with the leakage reactance between the yoke and output tube plate and the leakage reactance which exists in the high voltage overwind. Another major source of ringing exists when the two halves of the yoke are capacitively or dissipatively unbalanced. The latter may be controlled by properly center tapping the output transformer and connecting to the middle of the yoke to equalize capacities or by using proper capacity balancing across the hot side of the yoke. Minimization of ringing originating in the transformer may be effected by reducing leakage reactance to as low a value as possible and by using a high permeance damper tapped as near to the yoke as possible.

The second form of non-linearity is generally a short exponential change in the sweep current. The latter type of non-linearity is affected by incomplete cut-off of the output tube as well as stored energy and time delay in the transformer ferrite core, transformer core saturation, yoke Q, and by imperfections in the linearity circuit which cause the damper to delay conduction for a short period after the negative current maximum.

A further general type of non-linearities are variations which occur near the end of the scan. Four main effects contribute to this type of non-linearity including variation in the grid drive waveform, output tube knee conduction, resumption of conduction of the damper tube at the end of sweep and saturation of the transformer iron. Of these variations, the iron saturation may be controlled by modification of the air gap or core size of the iron while the grid drive waveform may be generated from passive circuitry such that both types of variation may be minimized. Damper conduction near the end of the sweep, however, represents a more serious problem. This resumption of the damper conduction is caused by the resistance drop in the yoke and transformer during the latter part of the sweep causing the voltage waveform across the damper to have a downward sawtooth shape of sufficient magnitude such that the damper conduction does resume. Correction of this condition requires change of the tap position of the damper on the horizontal output transformer, change of the damper bias, reduction of the sawtooth voltage drop

across the yoke or more effective linearity circuitry. In general, this requires that the damper tap position on the transformer be made different from the yoke tap point, thus aggravating the sweep ringing problem. An ultimate approach to this difficulty is to adjust the circuit so that the damper conducts throughout the whole scan period. In this case, sharp changes are minimized and the feedback mechanics of the damper tube conduction are obtained to minimize linearity variations. This adjustment, however, represents very poor efficiency in circuit operation.

Minimization of Variations

As noted previously, the most critical parameters with respect to linearity are those which affect the conduction time of the output tube, particularly screen voltage and bias voltage variations. It has been found that when screen voltage varies by a given amount, the change can be compensated to a large degree by varying the bias in proportion to the screen voltage variation so that the start of conduction is maintained at a constant time. Ideally, the amount of bias change should be equal to the screen voltage change divided by the screen amplification factor. In practice, compensation of this type may be incorporated in the circuit by use of a well by-passed cathode resistor with the output tube. Though wasteful of power, this circuit also minimizes the effect of transconductance changes in the output tubes with life and appears desirable in a reasonably stable deflection system.

Another method of minimizing variations in the sweep linearity is to permit both output tube and diode conduction to continue throughout the whole scan period. It may be shown that under this condition of adjustment that the damper circuit represents a feedback system which attempts to maintain a constant voltage across the yoke. Neglecting damper and yoke resistance, this would result in perfect linearity. When resistors are included, however, it will be found that the scan has a sawtooth velocity component because of the yoke losses. When the output tube and damper do not conduct simultaneously, the feedback action does not exist and maintenance of linear sweeps requires absolute control of the current being supplied to the yoke as well as all of the parameters which affect this current. Output tube conduction during the first part of the sweep

serves also to help damp transformer ringing. Again it is found that this adjustment greatly stabilizes the sweep linearity but at a large cost in power.

Another place where the performance could be appreciably improved is the design of improved yokes. Several factors are possible in this line. First, if the yoke can be made more efficient by shaping to fit over the bell of the CR tube such that less current is required for a given yoke inductance, then less total power must be handled by all the various components in the sweep circuit which tends to permit greater freedom in the handling of parts to minimize variations and also minimizes the change in the various parts due to heating and saturation. For a given resistance, the smaller current also minimizes the problem of the damper coming back into conduction near the end of the scan period. Another change which might be made in the yoke to improve performance is the reduction of the resistive component which causes the sawtooth voltage waveform across the damper. The total inductance of the yoke might also be optimized in any given circuit by adjusting its value so that the damper may be connected directly across the yoke in order to minimize the ringing problem.

Another method of improving the stability of sweep linearity is by use of improved linearity circuits in order to compensate the resistance drop in the yoke. Among possibilities along this line are the use of a saturable series reactor in the yoke circuit which can produce a compensating negative resistance effect that can both minimize the normal sawtooth velocity variation in the sweep and also minimize the tendency of the damper to come back into conduction. The sawtooth component may also be decreased by use of flux absorption techniques in which a small auxiliary yoke is inserted beneath the main yoke and its output connected to a proper time constant load which will be charged by the flyback pulse and produce current during scan time corresponding to a bucking sawtooth, thereby giving a more linear flux.

The drive circuitry must be reasonably accurate and stable in order to maintain the output linearity with present type sweeps. Usually it has been found that the drive circuit is particularly critical with B+ voltage variations and that this results in appreciable changes in the sweep linearity and width. It thus appears that

some form of feedback is necessary from the output circuit back to the drive circuit if a reliable sweep is to be produced.

Perhaps the most serious linearity problem is that of ringing. A number of compensation techniques are well-known, such as balancing the yoke capacity and damping circuits. Other possibilities include improving damper permeance or tapping the bottom of the yoke up somewhat on the horizontal output transformer in order to establish a ringing bridge circuit across the yoke. Output transformer ringing may also be reduced by the addition of separate damper diodes together with appropriate biasing arrangements to the plate of the output tube or to the top of the high voltage overwind.

No linearizing techniques are particularly useful if the compensation is subject to drift or change of the same order as the defect for which it is used. Nor is it useful if normally expected variations in other parts of the sweep over-shadow the compensation means. Therefore, we must be concerned not only with our sweep linearity in the laboratory but also in the field where variations of output tube transconductance, varying line voltage or drifts in components must be considered. The effect of these considerations is that we must discount somewhat the degree of improvement to be obtained from many of the linearizing techniques.

Experimental Results

Experimental studies show that most major causes of non-linearity may be reduced by the various mechanics discussed to a value of $\pm 5\%$. The major limitations which still stand out are variation of output tube gm and drive changes - both are quite critical and large changes are to be expected for each in practical receivers.

To obtain this degree of performance, it is necessary to adjust the system so that a considerable amount of power is required in comparison to the ideal minimum power condition. In effect, the reductions in non-linearity variations are primarily attributable to the feedback mechanics of the output tube cathode resistor and of the damper conduction throughout the whole scan so that the added power is inherent if these particular forms of stabilization are to be maintained.

As an example of the application of some of these corrections, consider the sweep shown on slide 1 which is typical of current practice for large tube sizes. Now let us compare to this sweep a re-designed version shown in slide 2 in which the following improvements have been incorporated:

1. The yoke resistance is halved by use of larger wire size.
2. The damper permeance is increased by use of two tubes in parallel.
3. The damper tubes are adjusted for full scan conduction.
4. A linearity circuit is incorporated.
5. A by-passed cathode resistor is added to the output tube for bias stabilization.

Slide 3 compares the performance of the two circuits. Though the modified circuit reduces the amount of non-linearity from 27% to 6%, it is apparent that the improvement has been gained at a considerable increase in circuit complexity and power input.

Conclusions

It is concluded from this study that the sweeps can be appreciably improved in linearity provided they are adjusted to handle sufficient power, i.e., that the damper be made to conduct throughout the whole scan cycle and that other feedback mechanisms such as cathode bias in the drive tube be incorporated. This particular adjustment results in improved

stability because of the feedback provided with this adjustment. Adjustment for minimum power input will not give this feedback and the linearity becomes extremely critical with all components. Though the linearity may be improved by restricting the tolerances of some components, other parameters such as tubes cannot be controlled in the practical case unless feedback is used. Among the most critical of these parameters which are beyond control in the simple sweep are the various supply voltages and the transconductance of the output tube. It was found that the improved sweeps still show wide linearity variation as the result of changes in the output tube transconductance and the grid drive voltage. It is, therefore, concluded that the usual type sweep is not satisfactory for more than approximately $\pm 5\%$ linearity. It is also to be noted that considerable cost has been added because of the high power input required for reasonable linearity. In effect, this study shows that the present industry practice has reached the point of diminishing return and that further improvement beyond $\pm 10\%$ linearity becomes increasingly expensive, though somewhat better linearity is definitely possible.

Finally, it may be concluded that there still exists a definite need for invention in the field of better horizontal sweep circuitry. Since the most serious variables of present circuits are those which cannot be controlled on an absolute basis, it appears as though some form of feedback mechanism is desirable in such new approaches to the problem.

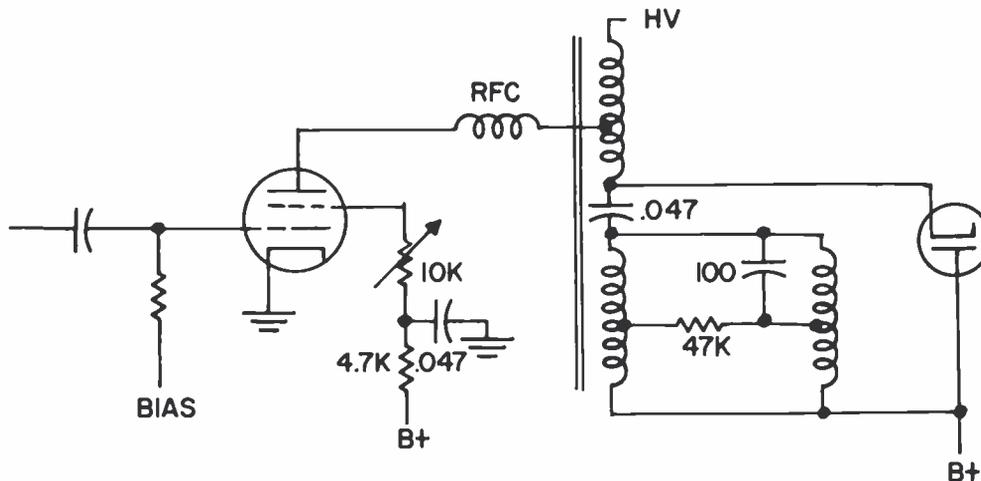


Fig. 1

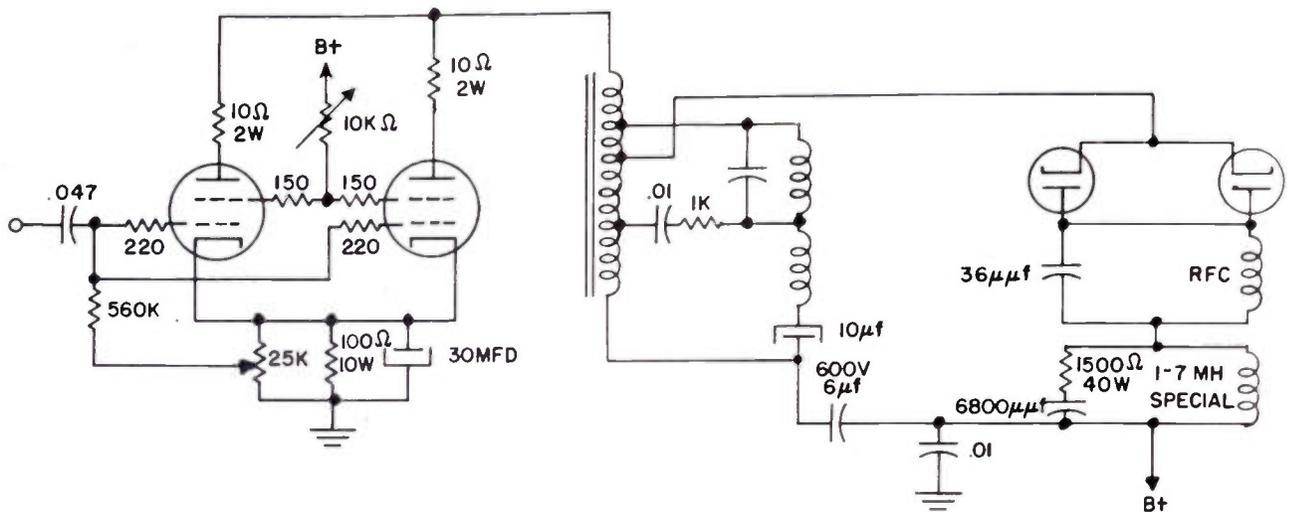


Fig. 2

<u>ITEM</u>	<u>NORMAL SWEEP</u>	<u>IMPROVED SWEEP</u>
YOKE Q @ 70KC	13.7	22.5
YOKE CURRENT-ma p-p	850	850
DAMPER CURRENT- PEAK	500	500
DAMPER CONDUCTION PERIOD	70% SCAN	100% SCAN
DAMPER MINIMUM CURRENT DURING SCAN- ma	0	100
OUTPUT TUBE PEAK CURRENT -ma	360	600
OUTPUT TUBE CONDUCTION PERIOD	65% OF SCAN	100% OF SCAN
AVERAGE CURRENT	150 ma	300 ma
POWER INPUT	37.5 WATTS	80 WATTS
PERCENT NON-LINEARITY	±13%	±4%

Fig. 3

A HIGH DEFINITION MONOCHROME TELEVISION SYSTEM

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SUMMARY

A system for doubling the maximum number of horizontal and vertical picture elements of a televised scene is described. This system operates with conventional sweeps and is compatible with present day monochrome receivers.

The additional picture definition is obtained by making more efficient use of the available bandwidth. Redundancies between a given picture element in successive fields and between adjacent elements in a given field are measured at the transmitter, and this information is transmitted to the receiver in a compatible manner. This information is used to determine whether the receiver will use large or small picture elements to display the picture information.

Introduction

At the time that the monochrome television standards were formulated, it was suggested that about 75 lines per inch of picture height would yield an adequately fine line structure at normal viewing distances.¹ These 525 line standards provide approximately 70 lines per inch of picture height on a 12 inch picture tube. The standards leave something to be desired when larger picture tubes are used. As the size of picture tubes has increased, the viewer has been forced to view the picture from a greater distance in order to avoid resolving the objectionable line structure.

A 1,000 line picture is desirable for 24 or 27 inch television receivers. This resolution would allow the viewer to enjoy these larger receivers in his living room because he would be able to sit closer to the picture. The extra detail afforded would be very pleasant.

In order to double the vertical and horizontal resolution of the present picture, an 18 Mc television channel would be required if present methods of transmission were to be used. This solution is not practical because of compatibility considerations and the difficulties associated with wide band circuits.

It is desirable to obtain the increase in resolution by modifying the present system so as to obtain a high definition system that is compatible with the present system. In order for a system to be compatible, its transmitted signal must produce a satisfactory picture on an unmodified conventional receiver. The modified compatible receiver must also be able to receive

a satisfactory picture which is transmitted by an unmodified conventional transmitter.

The requirement of compatibility imposes restrictions on the modified high definition system. It must use the same vertical and horizontal sweep rates and have the same basic scanning pattern. It must utilize the same frequency channel in a nearly identical manner.

The scanning rate and scanning pattern impose a field repetition rate of 60 times per second. This field rate is desirable because it allows high screen brightness to be used without producing large area flicker.

The bandwidth available allows the transmission of approximately 4.2 Mc of video information. This limits the maximum number of elements of information that can be transmitted in a given time. More picture detail may be transmitted in a given bandwidth by increasing the time required to complete a picture. The vertical interlace used in the present system is a first step toward increasing the time required to complete a picture while retaining the field repetition rate of 60 times per second.

The vertical resolution of the present system may be increased by increasing the vertical interlace ratio. It is also possible to increase the horizontal resolution by breaking each line into a plurality of dots which are presented in one field and then filling in the area between these dots in another field by presenting an interlaced line of dots. The spacing between these dots is important and the use of this dot interlace will be described with reference to elemental picture areas.

It is well known that the present television system provides a picture containing approximately 490 active scanning lines with a horizontal resolution of approximately 450 picture elements. The elemental picture areas may be obtained by dividing the picture into 490 horizontal strips and dividing each of these strips into 450 rectangular sections. A portion of the picture divided into these areas which will be referred to as large picture elements is shown in Figure 1. The present television system when utilizing its maximum bandwidth is ideally able to display any desired brightness in each of these elemental areas during the two fields or one thirtieth of a second required to complete a picture.

Interlaced System

These large conventional picture elements may be subdivided vertically and horizontally so that four small picture elements are obtained from each large picture element as shown in Figure 2. In order to obtain higher resolution in a modified system, the electron spot diameter is made equal to one half the diameter required to fill the large picture elements and the raster is slightly displaced vertically. The electron beam is gated as it is scanned so that the electron spot fills only one small picture element in each large picture element during two fields. During the next two fields, the raster is shifted and the electron beam is gated so that another set of small picture elements is filled in. At the end of eight fields all the small picture elements are filled in and a single picture is completed. The numbers in the small picture elements of Figure 2 indicate the field of the eight field cycle in which each element is scanned. The transmitter and receiver must be properly synchronized so that they scan these picture elements in the correct order. More detail has been obtained using this interlace scan than with the conventional scan. However, eight fields instead of two are required to complete a picture.

Flicker Considerations

If an entire picture was presented by scanning the small picture areas, the local flicker obtained would be objectionable when the picture is closely viewed. This local flicker results because the persistence of vision is shorter than the time required to complete a picture.

The maximum amount of information that may be displayed in a single picture has been increased to four times that possible in the conventional system by using the small picture element display over the entire picture area. When displaying an actual picture, the information content obtained by using the small picture elements will be only slightly greater than that obtained by using the conventional system. This results from the large redundancies present in most pictures.

The proposed system will obtain increased resolution by taking a longer time to transmit a picture in the same bandwidth. Redundancies will be used to choose an intelligent method of presenting the picture information by considering the structure of the picture. Let us first discuss the redundancies present in televised pictures and then attempt to take advantage of them to reduce the local area flicker.

Redundancies in Television

It has been found that the information transmitted in television contains redundancies in time and in position. Each picture element consists of any one of many levels of brightness. The television system is capable of reproducing

any combination of these brightnesses in each picture. At present, each picture element is treated as a complete surprise. Each element is not a complete surprise, however. This is evident from viewing a motion picture film each frame of which corresponds to a television picture frame. Two consecutive pictures do not differ greatly and it is therefore possible to look at one or two consecutive pictures and predict quite accurately what the next picture will contain. The more picture elements you are able to predict the greater the redundancy. In the case of transmitting a still picture, all the pictures transmitted after the first one are redundant because no new information is obtained.

Redundancy also exists in a single frame. There is a large correlation between adjacent picture elements. The probability of adjacent picture elements having identical or nearly identical brightness levels is very high in most pictures. E. R. Kretzmer² of the Bell Telephone Laboratories has measured this correlation by a simple and ingenious process. He placed two identical slide transparencies of a typical scene such as the model named "Slinky" shown in Figure 3, in an accurate positioning assembly and measured the light transmitted through these two slides in cascade. The transmitted light varied from peak value when the slides were in exact register to a minimum asymptotic value when the slides were widely displaced in opposite directions. The excess transmission component over the asymptotic value which corresponds to the desired autocorrelation was normalized and plotted as a function of the distance and direction that the slides were shifted from their in-register position. The results for this picture are shown in Figure 4. The separation between picture elements for the slide size used is approximately 7.5 mils horizontally and 5 mils vertically.

A relation between autocorrelation and lower bound redundancy has been developed by P. Elias³ and states that the lower bound redundancy is roughly equal to the negative of one-half the \log_2 of the drop in autocorrelation for one Nyquist interval shift. This Nyquist interval corresponds to a shift of one picture element. The horizontal correlation for "Slinky" results in a lower bound redundancy of approximately three bits of information. A picture element which contains any one of 64 levels of brightness contains $\log_2 64$ or six bits of information. If we were able to remove this redundancy of three bits of information, it would be possible to obtain a three bit reduction in channel capacity. Using an ideal coding scheme, a 50 per cent reduction in channel bandwidth could be theoretically obtained.

The proposed high definition system will not use the redundancies to gain more efficient use of the frequency channel by coding. Such a coded transmission could not be directly utilized by a present receiver and therefore would not be

compatible. The redundancies will be used to secure compatibility and to reduce the local flicker.

The local area redundancy can be utilized to reduce this flicker. The brightness of all four small picture elements which constitute a large picture element will be identical in many cases because of the large correlation between adjacent picture elements. These areas of the picture may be reproduced using the low definition scan without degrading the picture resolution. Only those areas in which the brightness of the four small picture elements differs will be reproduced using the small picture elements. A large portion of the picture will be reproduced using the large picture elements and therefore will have the standard 30 cycle frame repetition rate thereby reducing the total flicker.

The redundancy in time allows the use of the lower picture repetition rate and can also be used to reduce the local area flicker. The eye cannot resolve very fine detail in moving objects because of a physiological limitation. The reduction of resolution in moving objects is approximately 50 per cent⁴. Moving objects may be reproduced using the large picture elements without suffering much loss in usable picture detail. The conventional repetition rate of 30 frames per second will provide good motion continuity for these moving objects. The flicker will be reduced by further reducing the area in which this flicker occurs.

Information Analysis

In the high definition system, the video information for the entire picture is obtained at the transmitter by scanning the small picture elements as previously described.

The receiver will display this information on the large picture elements or on the small picture elements depending upon the motion and detail present in each area. The size of the picture element to be used in each area of the picture will be determined by an analysis of the video information. This analysis may be performed either at the receiver or at the transmitter. It is desirable to perform this analysis at the transmitter because it is not economically feasible to incorporate this equipment in each receiver. A separate signal must therefore be transmitted to tell the receiver which picture element size to use.

This analysis and the transmission of this signal will be explained later with respect to the proposed compatible system.

Experimental Investigation

Equipment was built to demonstrate the presentation of a picture using the principles outlined. In order to facilitate the comparison of the large and small picture element scans and to ease circuit complexity, the conventional

field rate was retained but the number of scanning lines was reduced by a factor of approximately two. The picture obtained can be thought of as representing one quarter of the normal picture. The order of small picture element presentation shown in figure 2 was chosen because its simplicity eased circuit requirements. A block diagram of this equipment is shown in Figure 5. The broken line in the center of Figure 5 separates the equipment representing the transmitter from that representing the receiver.

Transmitter

The electron beam of the flying spot scanner is scanned in a 245 line raster by 60 cycle vertical and 7350 cycle horizontal deflection amplifiers which are controlled by the sync generator. The light from the scanner is focused by a lens onto a slide transparency. The light passing through the slide is converted into an electrical signal by the photomultiplier. This signal is amplified by a video amplifier having a 2.5 Mc bandpass, passed through the detail analyzer and sampled by the first sampler switch at a 2.47 Mc rate. This 2.47 Mc signal is accurately synchronized to the horizontal sync pulse so that the first sampler switch samples the same portion of the raster each time it is scanned. The signal obtained from the sampler is equivalent to the signal that would be obtained from the video amplifier if a stationary fine mesh wire screen was placed on the face of the flying spot scanner. The sampled video output of the sampler switch is passed through a low pass filter which limits the transmitted bandwidth to 1.25 Mc and therefore provides a continuous video signal by reproducing the envelope of the sample pulses.

A raster shifter which shifts the raster at the end of two fields is controlled by the sync generator so that the transition is made during vertical retrace time. The raster is shifted by means of additional vertical and horizontal coils on the neck of the flying spot scanner. The result is to shift the location of the video samples with respect to the televised image. This result is equivalent to that obtained by shifting the imaginary fine wire screen so that the open areas of the screen coincide with the number one areas of Figure 2 during field one, the number two areas during field two and so forth until the eight field cycle is completed.

Figure 6 illustrates the necessity for video sampling. The video information from a given line is sampled at the positions indicated by the dark pulses during one field and then sampled at intermediate positions during a subsequent field. A 2.5 Mc sampling rate is used in each field and the information may be transmitted over a 1.25 Mc channel. The information obtained in these two fields can be reconstructed to obtain the equivalent of 2.5 Mc horizontal resolution. If sampling were not used, the same information would be transmitted in both fields resulting in 1.25 Mc resolution.

Receiver

The continuous video signal from the low pass filter is sampled by the second sampler switch which is synchronized to sample in phase with the first sampler switch. The output of this switch consists of dot impulse video which is passed through a video amplifier having a 12 Mc bandwidth to the cathode of the kinescope.

The kinescope raster is scanned and shifted in synchronism with the flying spot scanner raster. A dot pattern is obtained on the screen of the picture tube when the high definition scan is used. The position of the dot shifts from field to field as shown in Figure 2.

The spot enlarger is energized by the detail analyzer if a large picture element scan is desired. The spot enlarger defocuses the electron beam by changing the potential of the focus electrode. The dots blend together so as to cover the entire picture area in each two fields. The small picture element scan is used in those areas which contain high detail while the large picture element scan is used in areas of low detail. The amount of detail is obtained by making a comparison of the information from two consecutive small picture elements. If they have the same video level, it is assumed that there is no high detail information in that area and the large picture elements are used. If the video levels are different the small picture elements are used.

Photographs of pictures reproduced using this equipment illustrate the possibilities of the system. We have not as yet obtained photographs to show the use of the large and small picture elements in a single picture. It should also be pointed out that the integration provided by the camera eliminates the flicker problem, so that the photograph is more pleasing than the actual picture.

Figure 7 is a photograph of a typical scene reproduced in a conventional manner on a 245 line raster. Figure 8 shows the same picture which has been sampled in the horizontal direction to obtain a dot pattern having approximately 300 dots per line. This picture contains the information transmitted during two fields over a 1.25 Mc bandwidth and represents the detail obtainable with the large picture element scan although the spot size has not been enlarged. Figure 9 shows the same picture using the small picture element scan which was transmitted during eight fields over a 1.25 Mc channel.

The relative resolution capabilities of these scans is clearly illustrated by Figures 10, 11 and 12 which are photographs of the center portion of a test pattern. Figure 10 shows the result obtained using the conventional line scan on a 245 line raster. Unfortunately, the video bandwidth was not limited to 1.25 Mc in this figure, resulting in increased horizontal

resolution capabilities. Figure 11 shows the result of breaking the 245 line raster into dots. This picture was transmitted during two fields over a 1.25 Mc channel. The horizontal resolution is theoretically reduced to 70 per cent of the normal line scan value by breaking the line up into dots. The reason for this is illustrated in Figure 12. If the vertical stripe pattern to be reproduced is in register with the samples the full resolution is realized. If the samples are displaced between the stripes, no lines are resolved. This position uncertainty is similar to that obtained in vertical resolution considerations and the effective resolution of 70 per cent of the maximum value therefore applies to the horizontal dot resolution.

Figure 12 shows the same picture reproduced using the small element scan which was transmitted during eight fields over a 1.25 Mc bandwidth. The small element scan provides twice the vertical and 1.4 times the horizontal resolution obtainable with the conventional scan when operating with the same channel bandwidth.

Improvements on Preliminary System

The simple dot scanning presentation shown in Figure 2 has two outstanding disadvantages; it produces a serious flicker and results in a non-compatible system.

Consider the presentation of narrow white vertical and horizontal bars which are surrounded by a dark background. These bars are shown superimposed on the present small picture element raster in Figure 14. The horizontal bar is excited only during fields 2 and 6 producing a 15 cycle per second flicker. The vertical bar is excited only during consecutive fields 7, 8, 1 and 2 producing a 7.5 cycle per second flicker. This flicker is particularly serious when the small picture elements are displayed on a short decay phosphor. The rate of flicker of the horizontal bar may be doubled using the dot presentation order shown in Figure 15. A horizontal and vertical bar are shown superimposed on this raster. The horizontal bar is excited during fields 2, 4, 6 and 8. The vertical bar is excited during all eight fields. By exciting interleaved parts of these bars at a higher rate, the flicker is confined to local areas rather than resulting in the bar appearing and disappearing.

The simple dot scanning presentation shown in Figure 2 results in a non-compatible system. The samples are shifted together in a vertical or horizontal direction in order to reach the desired picture elements. The picture on a conventional receiver would shift vertically and horizontally at 15 and 7.5 cycle per second rates respectively. If this sampling pattern were used in the high definition transmitter. This difficulty may be eliminated by using the sampling pattern shown in

Figures 15 and 16. By staggering the picture elements excited in a single field vertically and horizontally, the average position of the samples does not shift from field to field. The picture on a conventional receiver will not shift but some local area flicker will be obtained in areas of high resolution.

Proposed System

It is not convenient to use a simple parallel scan to cover these interleaved picture elements. They may be easily covered however by adding a vertical sinusoidal undulation having a peak to peak amplitude of one small picture element to the standard scan as shown in Figure 16. The phase of the sinusoidal undulation will determine which small picture elements are covered. The peaks of the sinusoidal displacement during field one will occur in the number one elements in the first row while the troughs of the sinusoidal displacements will occur in the number one elements in the second row. The camera tube is scanned in this manner and its video output gated on at the peaks and the troughs of the displacement sinusoid in order to obtain the information from the number one elements. The sampling frequency will therefore be twice the displacement frequency. Figure 17 shows the block diagram of a compatible transmitter operated in this manner.

The sync generator controls the horizontal and vertical deflection generators which scan the camera tube in a conventional 525 line raster. The sync generator will also provide the phase selecting switch with 4 Mc signals phased at 0, 90, 180 and 270 degrees. The phase selecting switch selects one of these signals depending upon which of the eight sets of small picture elements is to be scanned. This 4 Mc signal controls the high frequency deflection amplifier which provides the vertical undulation by exciting an auxiliary vertical deflection coil consisting of approximately eight turns. The 4 Mc signal is also doubled and fed to the gate or sampler switch which samples the video at the peak and trough of the deflection sinusoid. This sampled video is filtered by a 4 Mc low pass filter and impressed upon a delay device such as a video tape recorder⁵. The video must be delayed by exactly two fields between each of the following: the recording head, pickup head A, pickup head B, pickup head C and pickup head D. There will be therefore an eight field or one picture delay between the recording head and pickup head L. The information from these heads corresponds to the identical picture element of two consecutive pictures. If the video levels are identical, there is a redundancy in time and an indication that no movement of the subject in that area has taken place during 2/15 of a second. This analysis will be made by the movement comparator which may consist of a differential amplifier. An output will be obtained if the video levels are different.

The video from pickup heads A, B, C and D corresponds to the brightness of the four adjacent small picture elements which constitute a large picture element. An output will be obtained if all four elements do not have the same brightness thereby indicating the presence of high definition information. If there is no movement and high detail information is present, an output will be obtained from the coincidence circuit indicating that the small picture elements should be used at the receiver.

This additional signal, which we shall call the redundancy signal, may be transmitted in a compatible manner by a single sideband suppressed subcarrier similar to the NTSC color subcarrier. A reference phase may be transmitted as a burst during horizontal retrace time. A subcarrier in phase with the reference would cause the receiver to use the small picture element display. A subcarrier out of phase with the reference would cause the receiver to use the large picture element display.

The subcarrier frequency may be chosen so as to place the subcarrier at the uppermost part of the video band with its signal sideband components being lower in frequency. Only one sideband is necessary because only two opposite phases are of interest. The vector representing the phase need only rotate in one direction to reach these two positions with equal facility.

The video signal from pickup head D which has a 4 Mc bandwidth and the conventional sync signals will also be transmitted.

The receiver shown in Figure 18 contains the conventional FF and IF amplifiers and a video detector. The sync signals are separated in a conventional manner. The 4 Mc burst will be used to synchronize a 4 Mc reference generator. This reference generator provides a 4 Mc vertical deflection for the small picture element scan which is identical to that used at the transmitter. This signal is frequency doubled and used to sample the video at the peaks and troughs of the 4 Mc vertical deflection as previously described with reference to the transmitter. The 4 Mc reference signal will also be used to demodulate the redundancy subcarrier. In the case of the two gun cathode ray tube shown, this redundancy signal may be used to control a paraphase amplifier which chooses whether a small spot gun will be used to excite the small picture elements or whether a large spot gun will be used to excite the large picture elements.

Flicker Reduction

Flicker is not only a function of the repetition rate and the size of the area which is excited. It is also a function of the percentage of time that light is emitted from an area. Figure 19 taken from a paper by F. W. Engstrom⁶ shows the effect of the degrees opening of a sector disk on the frequency at which flicker

became just noticeable. These curves indicate that flicker may be further reduced by the use of long decay phosphors similar to those used in radar tubes. The reduction in flicker is obtained by increasing the percentage of time that light is radiated from each picture element. The drawback of using long decay phosphors is the smearing of moving objects due to the tail of light left behind them.

The advantages of both the long and the short decay phosphors may be obtained at the expense of picture tube complexity. A tube similar to those used in color television may be constructed with interleaved short and long decay phosphors and a means for separately exciting them. Figure 20 shows a shadow mask tube having interleaved long and short decay phosphor dots. Two electron guns are provided in the neck. The large size spot gun excites the short decay phosphor in the large picture elements while the small size spot gun excites the long decay phosphor in the small picture elements. The small spot gun is provided with a pair of electrostatic deflection plates to add the small vertical deflection to enable the beam to reach the desired picture elements. Figure 21 shows a portion of the faceplate of this tube. The phosphors are deposited in interleaved dots. Each small picture element is subdivided into an area of long decay phosphor and an area of short decay phosphor.

The left portion of Figure 21 illustrates the high definition scan using the small spot beam on long decay phosphor, while the low definition scan using the large spot beam on short decay phosphor is illustrated in the right portion. During field one, the number one long decay elements are excited in the high definition portion while the number 1, 3, 5, and 7 short decay elements are excited in the low definition area. During field two, the corresponding interlaced areas are excited. During field three, the number three long decay elements are excited in the high definition portion while the odd number elements of the low definition portion that were excited during field one are re-excited.

Competibility

This receiver may be used to receive a signal transmitted from a conventional transmitter. The receiver circuitry may be arranged so that the reference generator will be inoperative if a 4 Mc burst is not transmitted. The video will not be sampled and a conventional line scan will be obtained. The receiver may be designed to use the large spot beam in this case. The receiver will therefore function as a conventional receiver when receiving a conventional transmission.

A conventional receiver may receive a signal transmitted by the high definition transmitter shown by Figure 17. The redundancy subcarrier will tend to produce a very fine dot pattern on the screen. This pattern should only cause a

slight picture degradation because it will be largely attenuated by the conventional receiver. Some picture degradation will result from superimposing the video information obtained from the small transmitted picture elements on the large picture elements of the conventional receiver. A local area flicker will result in areas of high detail. This flicker will only be obtained in a small portion of the picture, however, because of the high correlation of the brightness value of adjacent picture elements which was previously discussed.

Although it is desirable to be able to select the size of each picture element, it will only be possible to select the size of each group of three or four consecutive elements using the redundancy subcarrier. This results from the limitation of the subcarrier bandwidth which will be determined primarily by cross talk considerations.

Conclusion

A picture having twice the vertical resolution and 1.4 times the horizontal resolution of a conventional picture has been obtained using the dot interlaced technique described. Observers agreed on the desirability of this increased resolution when viewing a picture which was reproduced using the small picture element scan over the entire picture area. They found the local area flicker objectionable, however.

Future experimental work will be directed toward confining the flicker to those areas having high definition information. The possible advantages of using long and short decay phosphors will be investigated. Methods of improving the compatibility of the system are also being studied.

Acknowledgements

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DWG. 52-D-4274

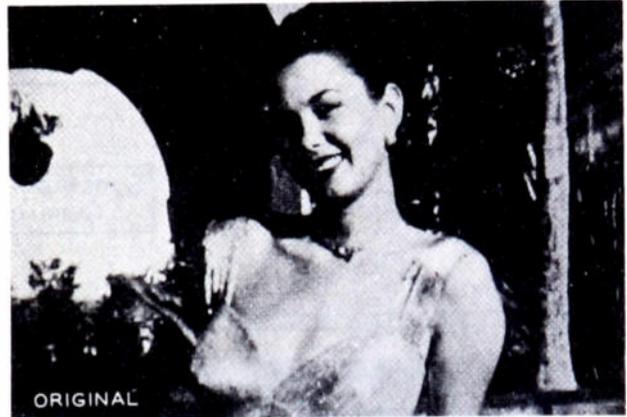
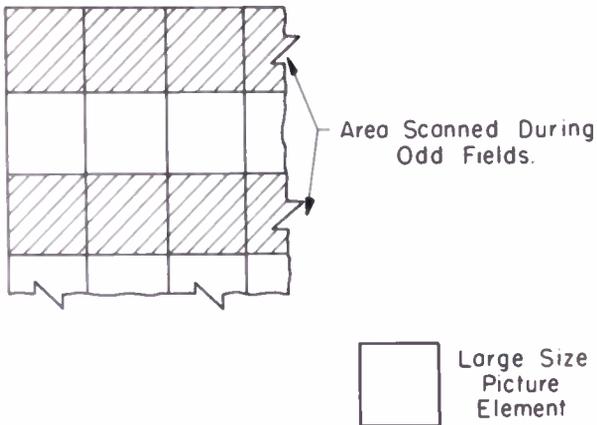


Fig. 3

Fig. 1
Low definition Scan

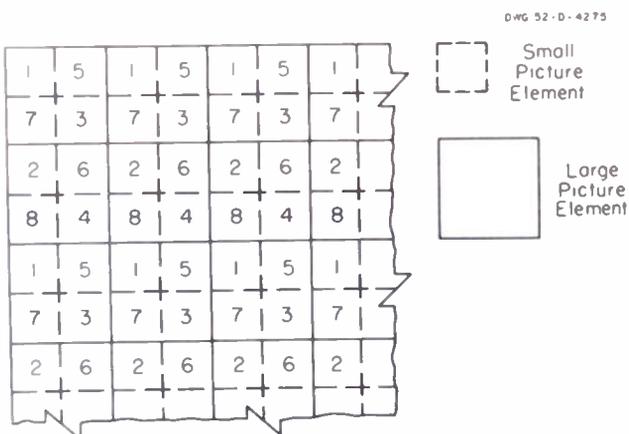
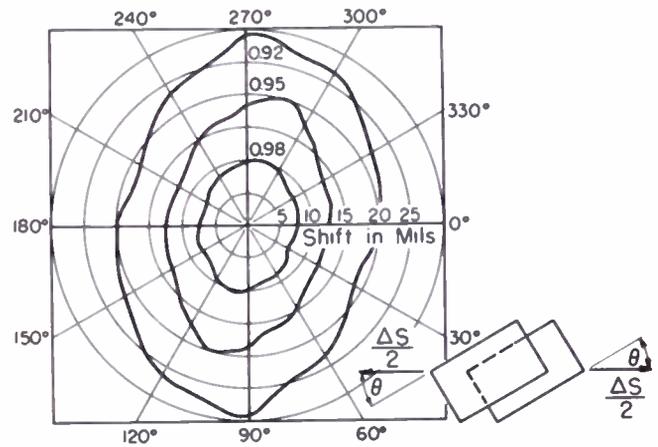


Fig. 2
High definition Scan



Courtesy of E R Kretzmer, Bell Telephone Laboratories.

Fig. 4
Autocorrelation contours for "slinky"

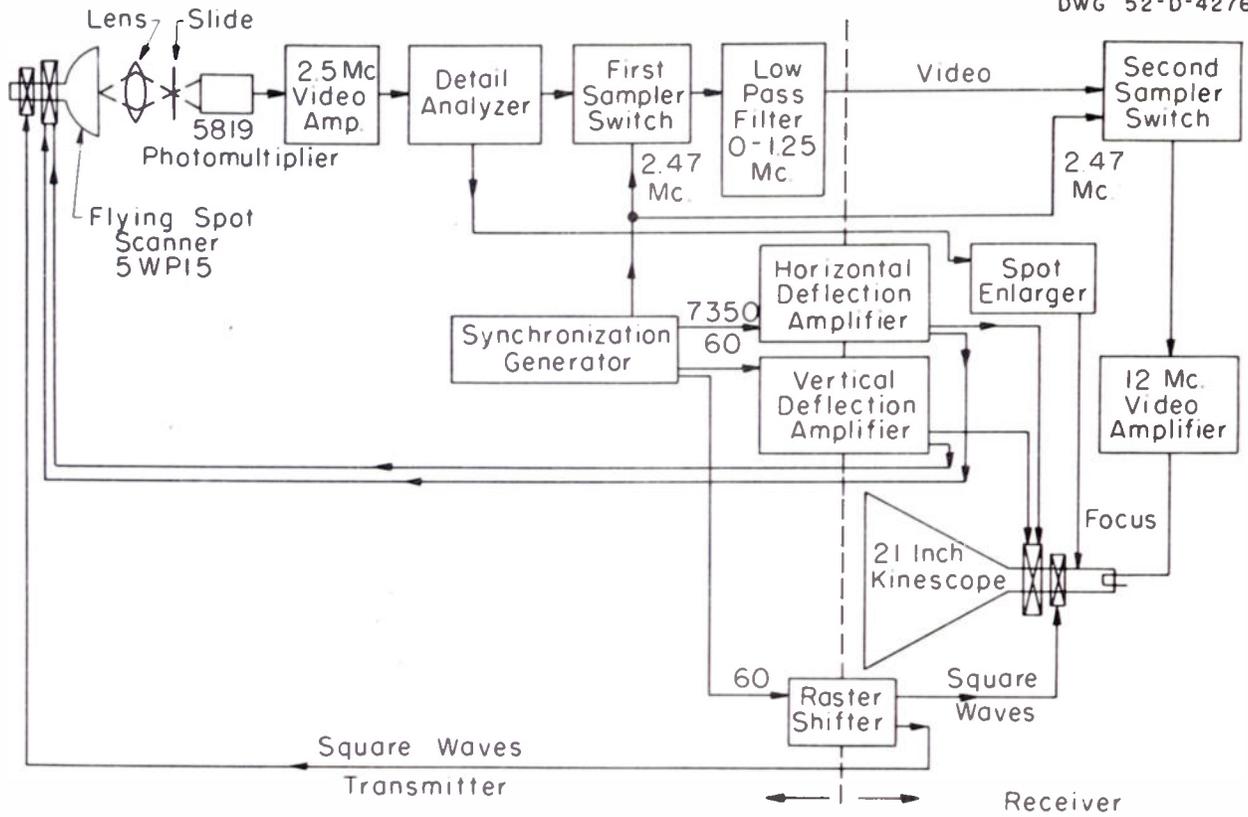


Fig. 5
Block diagram

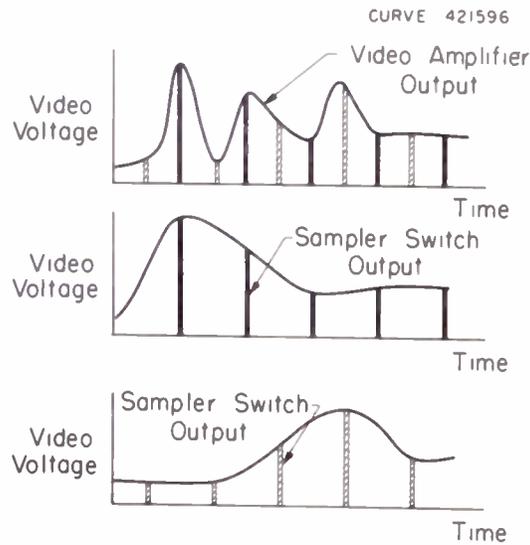


Fig. 6
Video Sampling



Fig. 7

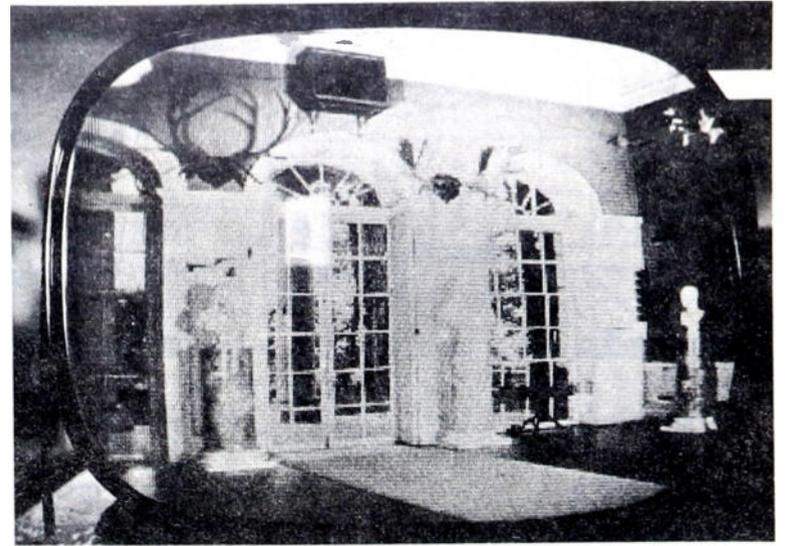


Fig. 8

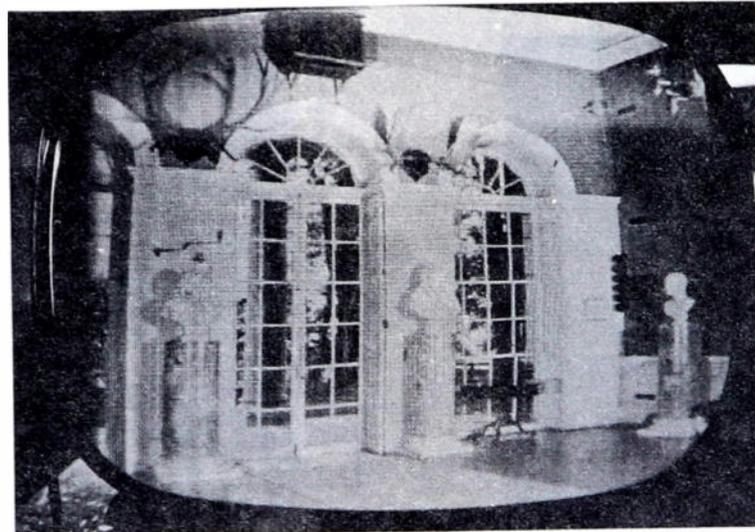


Fig. 9

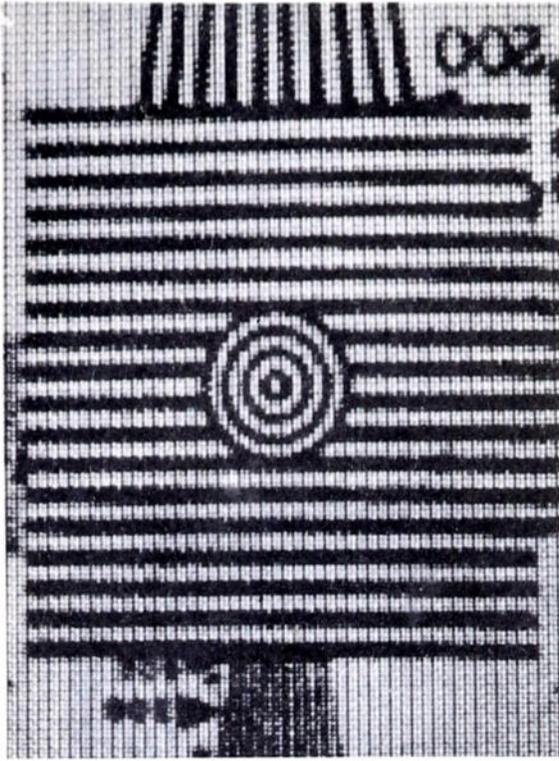


Fig. 12

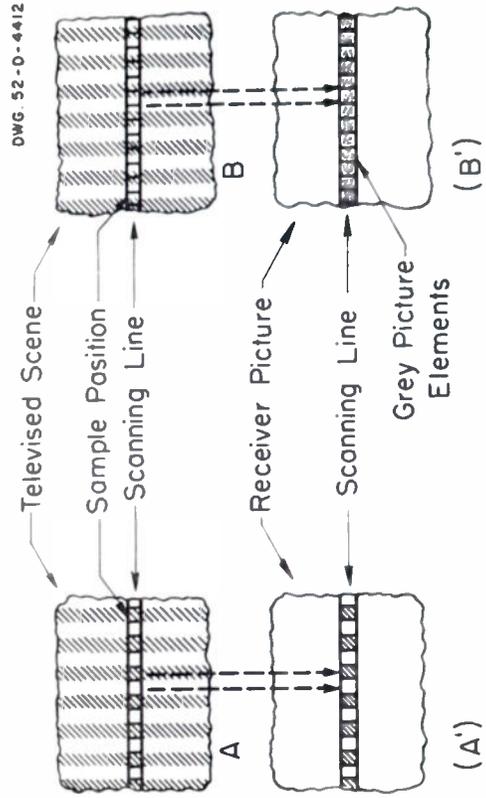


Fig. 10

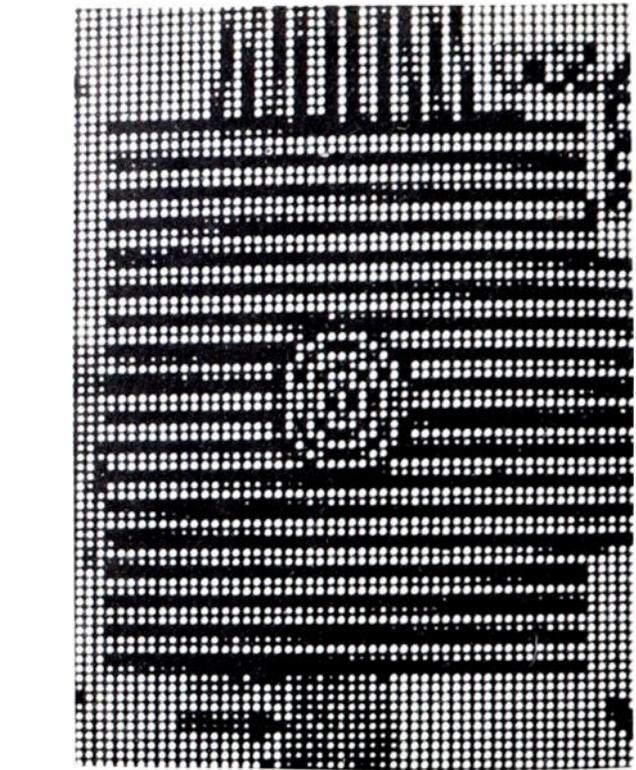
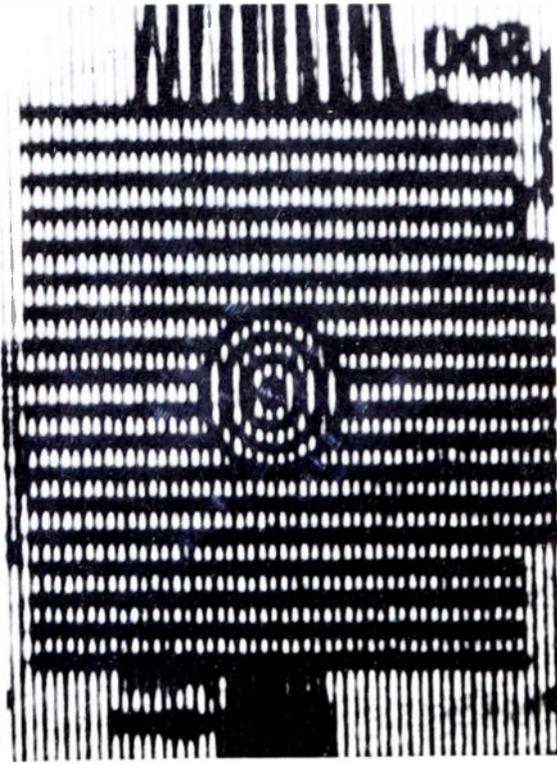


Fig. 11

Fig. 13
Relationship of position of samples of
horizontal resolution

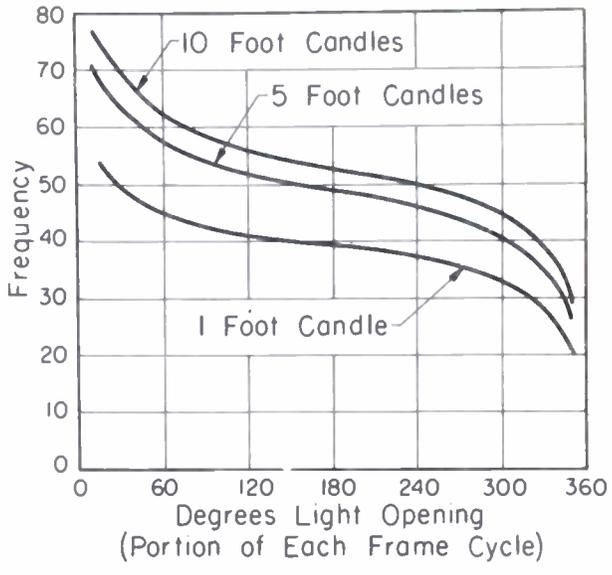


Fig. 19

Degrees opening of sector disk vs. frequency for just-noticeable flicker.

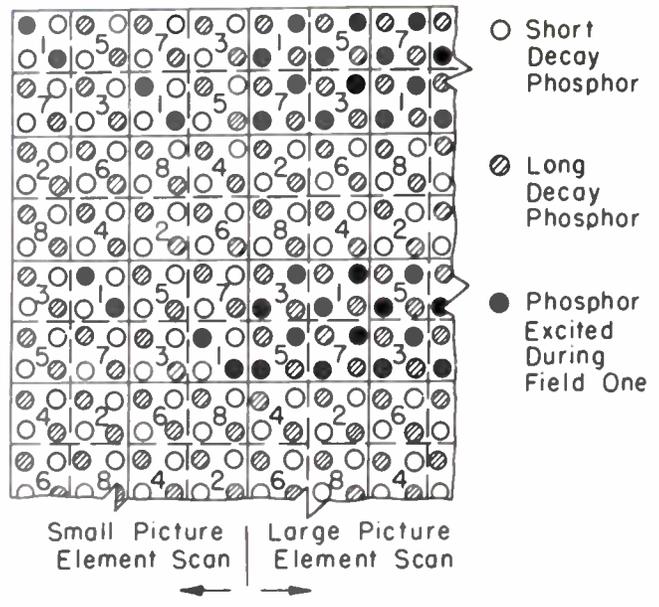


Fig. 20

Two-beam cathode ray tube

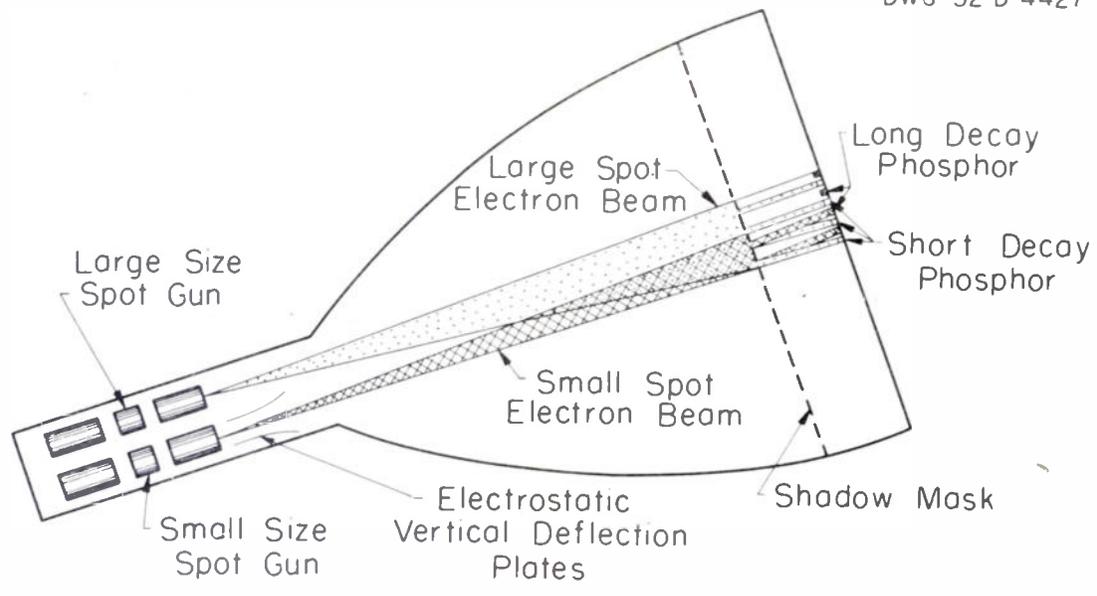


Fig. 21

Portion of faceplate of display device

DETERMINATION OF THE OPTIMUM DEMODULATION ANGLES IN COLOR RECEIVERS

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Now that the basic soundness of the compatible color television system has been demonstrated successfully, factors of economy in design have taken on a new importance. Color receivers in which the color difference signals and the monochrome signal are added on the picture tube offer substantial cost advantages. The use of the newly developed beam deflection demodulator tube leads to a further saving since this tube makes both a positive and a negative output signal available¹.

It will be shown that this tube type is not used to its best advantage in the conventional circuit. A substantial increase in the available picture tube drive can be obtained without increasing the circuit complexity materially, if the principles described in this paper are used.

The 6AR8 beam deflection tube operates as an electronic double throw switch driven by the reference carrier. The beam current is modulated by the chrominance signal. The currents from the two output plates consist of the chrominance signal multiplied by opposite phases of the reference carrier. In this way, positive and negative color difference signals are obtained in the same tube. In the conventional equiband circuit, shown in block diagram in Figure 3, the two demodulator tubes are made to detect positive and negative R-Y and B-Y components. The two negative components are added to form the G-Y signal.

Maximum demand considerations show that a much larger output of B-Y signal is required than for the other two signals. This requirement on the B-Y presents an even greater problem due to the fact that the detector efficiency is low for B-Y signals. Consequently, overloading of the B-Y demodulator limits the output capabilities of the system. A very substantial increase in

the available output signal can be obtained by distributing the load more evenly. The principles involved will first be explained with some simplified reasoning and subsequently put on a more general basis.

The output requirements for the B-Y demodulator can be reduced by inserting some B-Y signal at low level into the input of the Y amplifier. This is accomplished by interconnecting the demodulator plate and the Y input with a large resistor. This resistor does not load the B-Y output appreciably. The R-Y demodulator is rephased in order to obtain a minus B-Y component to cancel the positive one in the Y channel. The G-Y matrix should also be modified for the same reason.

The B-Y component in the Y channel is subject to the delay of the amplifier, but since it is a wide band amplifier, it has a negligible delay with respect to the rise time of a narrow band color difference component. A further improvement can be achieved by also feeding some R-Y back to the Y amplifier. This is shown in block diagram in Figure 4.

At this point, a more general analysis of the type of signals obtainable from a properly phased demodulator is in order. The general expression for a signal detected at a certain reference carrier angle ϕ_1 and with a certain demodulator gain K is given by equation (1) and simplified in equation (2). The constants θ_R , θ_G , θ_B , r_0 , g_0 and b_0 have values specified by the color television standards².

$$E = K \left[r_0 R \cos(\phi - \theta_R) + g_0 G \cos(\phi_1 - \theta_G) + b_0 B \cos(\phi - \theta_B) \right] \quad (1)$$

$$E = r_1 R + g_1 G + b_1 B \quad (2)$$

It can be shown that the constants r_1 , g_1 and b_1 can be freely chosen by selecting the angle and the gain within the limitations set by the expression

$$r_1 + g_1 + b_1 = 0 \quad (3)$$

The restriction of equation (3) means that the demodulated signal must be zero on neutral shades or if $R = G = B$. Signals which agree with the restriction in equation (3) will be designated as color difference signals. The values of the constants r_1 , g_1 and b_1 can be determined graphically by drawing the curves for

$$\begin{aligned} r &= r_o \cos(\phi - \theta_R) \\ g &= g_o \cos(\phi - \theta_G) \\ b &= b_o \cos(\phi - \theta_B) \end{aligned} \quad (4)$$

as is shown in Figure 2. The inverse operation of finding the angle and gain if the output signal is known can be accomplished with the aid of the curves

$$\frac{b}{r} = \frac{b_o}{r_o} \left[\cos(\theta_R - \theta_B) - \sin(\theta_R - \theta_B) \tan(\phi - \theta_R) \right] \quad (5)$$

and

$$\frac{r}{b} = \frac{r_o}{b_o} \left[\cos(\theta_R - \theta_B) + \sin(\theta_R - \theta_B) \tan(\phi - \theta_B) \right] \quad (6)$$

Any color difference signal can also be obtained as the weighted sum of another pair of color difference signals.

Any combination of R, G, and B can be formed by adding the output of two demodulators to Y. Complete freedom exists, therefore, in the selection of the new common channel signal L.

$$L = r_2 R + g_2 G + b_2 B$$

The following normalization will be used:

$$r_2 + g_2 + b_2 = 1$$

R-L, G-L, B-L, and L-Y will be color difference signals because of this normalization. The

signals R-L, G-L, and B-L are obtained from the demodulator tubes. Two of these signals are detected directly and the third is formed by the addition as in the case of G-Y in the conventional system.

We will next investigate which pair of signals should be detected independently. The signals, A, B, and C, which a synchronous demodulator detects at three different angles are indicated in Figure 1. Since only two independent components can be derived from a quadrature modulated signal, a relationship must exist between A, B, and C. This relationship is given by

$$A \sin \alpha + B \sin \beta + C \sin \gamma = 0 \quad (7)$$

$$A = -B \frac{\sin \beta}{\sin \alpha} - C \frac{\sin \gamma}{\sin \alpha} \quad (8)$$

The A derived from the negative plates of the beam deflection demodulator tube is larger than the A detected directly if

$$\frac{\sin \beta}{\sin \alpha} < 1 \quad \text{and} \quad \frac{\sin \gamma}{\sin \alpha} < 1$$

or

$$180^\circ - \beta < \alpha < \beta \quad 180^\circ - \gamma < \alpha < \gamma \quad (9)$$

for

$$\beta > 90^\circ \quad \text{and} \quad \gamma > 90^\circ$$

Equation (9) means that the pair of color difference signals with the smallest angle between them should be demodulated for the largest output from the negative plates. In the conventional system, the smallest angle occurs between the B-Y and R-Y vectors.

The modified system can be explained with an example. If a picture tube requiring equal drive is used, the following output voltages are needed in the R-Y, B-Y system (Figure 3). E_v is the peak video drive.

$$B-Y = 1.78 E_v$$

$$R-Y = 1.40 E_v$$

$$G-Y = 0.82 E_v$$

(10)

By using $M = 1/3 R + 1/3 G + 1/3 B$ as a common

channel signal, the output voltages are equalized (Figure 4).

$$B-M = 1.33 E_v \quad R-M = 1.33 E_v \quad G-M = 1.33 E_v \quad (11)$$

The angles at which these signals are demodulated are shown in Figure 5. They can be found with the aid of the curves in Figure 2. For example, $b = g$ for the R-M signal which is seen to occur at 239.5° clockwise. It is seen that in this case the angle between the R-M and G-M vectors is the smallest and this pair should therefore be detected. The required amount of color difference signals fed back to the input side of the wide band amplifier is determined by the following equations:

$$M-Y = -0.48 (G-M) - 0.19 (R-M) \quad (12)$$

$$M-Y = 0.48 (B-M) + 0.29 (R-M) \quad (13)$$

Which of the two equations (12 or 13) is used depends on the polarity of the Y signal at the point of addition, which depends in turn on the number of stages in the amplifier. For an even number equation (12) is used; for an odd number equation (13).

The described system has several advantages.

- 1) The maximum output requirements in the color difference channels have been reduced from $1.78 E_v$ to $1.33 E_v$ or by 25 percent. A lower supply voltage can therefore be used.
- 2) The demodulator tube current swing is reduced even more to 45 percent. This means that the same picture tube drive can be obtained with a demodulator tube having 45 percent of the current capability required for the conventional system.
- 3) The R-M signal is almost the same as I. An I Q receiver can be designed by feeding the R-M demodulator a wide band chrominance signal and the G-M demodulator a narrow band signal.

Some picture tubes require different drives on the guns due to inequalities in the colorimetric phosphor efficiencies. The described system of feedback matrixing can be adapted to these requirements. A value for the signal L is chosen according to the specifications of equations (3) and (4).

At the reference carrier angle, ϕ_R , a signal equal to r_3 (R-L) is detected and at ϕ_G , a g_3 (G-L) signal. The factors r_3 and g_3 indicate the resultant relative picture tube drive voltage in the two channels. The L signal is applied in the proper ratios with a voltage divider (see Figure 6).

$$r_3 (R-L) = r_3 (g_2 + b_2) R - r_3 g_2 G - r_3 b_2 B \quad (14)$$

$$g_3 (G-L) = g_3 (r_2 + b_2) G - g_3 r_2 R - g_3 b_2 B \quad (15)$$

A given angle ϕ_R does not completely specify the L signal. The ratio

$$\frac{r_3 g_2}{r_3 b_2} = \frac{g_2}{b_2}$$

is, however, determined by ϕ_R . In the same way, r_2/b_2 is determined by ϕ_G . The two angles together determine all the parameters. The angles ϕ_R and ϕ_G are restricted to a certain range by the fact that the color components in the L signal must be positive. The peak value of L would be larger than necessary if one or more of these components were negative. The R-L signal can range between R-G and R-B and the G-L signal between G-R and G-B.

$$77.6^\circ < \phi_R < 150.7^\circ$$

$$193.5^\circ < \phi_G < 257.6^\circ$$

Within these ranges, g_2/b_2 increases with decreasing ϕ_R and r_2/b_2 increases with increasing ϕ_G . For a given ϕ_R , an increase in r_2/b_2 will cause $g_2 + b_2 = 1 - r_2$ to decrease. Since $r_3 (g_2 + b_2)$ is fixed by ϕ_R , the value of r_3 increases with increasing ϕ_G . Analogously, g_3 increases with decreasing ϕ_R . The effect of ϕ_R on r_3 and of ϕ_G on g_3 is smaller. Increasing ϕ_R and decreasing ϕ_G in the proper ratio leaves the ratio of r_3 and g_3 constant while increasing b_3 . By a method of successive approximations using the graphs in Figure 2, a rapid calculation can be made of the optimum angles for matching a given drive ratio.

The transient response of the system can be evaluated. The response is the same as for a receiver without a feedback matrix in which a spurious signal is introduced in the Y channel consisting of a negative L-Y component followed by a positive L-Y component of equal amplitude after a time delay equal to that of the amplifier. The resulting spurious component consists of the differentiated minus L-Y signal or a minus L-Y signal passed through a network with a linearly rising frequency characteristic.

These differentiated components form a pulse. The amplitude of this pulse is small in comparison to the L-Y step from which it is derived, if the time delay in the wide band amplifier is smaller than the rise time of the L-Y chrominance signal. The amplitude of the pulse is approximately 0.2 times the step, if the amplifier delay is 0.2 microseconds and the chroma bandwidth is 600 kc. The maximum M-Y step occurs for a green to purple transient. The

step can be $0.5 E_V$. The amplitude of the pulse can therefore be $0.1 E_V$. The spurious pulse darkens the green to purple transients and it increases the brightness of the reverse transients. The effect is however smaller than the darkening of the transients between complementary colors due to the gamma correction.

Conclusion

It has been shown that a much more efficient use can be made of the demodulator tubes by adapting the circuit to the demand requirements of the color standards. The application of the described principles is not restricted to receivers using beam deflection tubes.

References

- 1) Electronics, May, 1954, p. 148 - Robert Adler and Charles Heuer, "Beam Deflection Tube Simplifies Color Decoders."
- 2) Proc. I.R.E., January, 1954, p. 66 - George H. Brown, "Mathematical Formulations of the NTSC Color Television Signal."

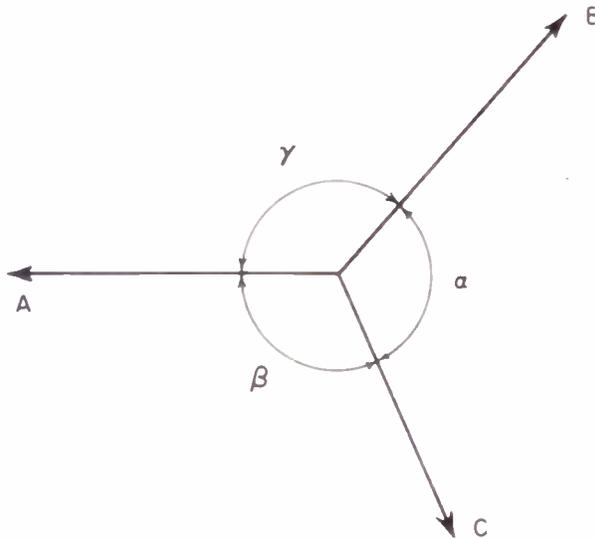


Fig. 1
Vector Diagram of the Output of a Synchronous Demodulator at Different Angles

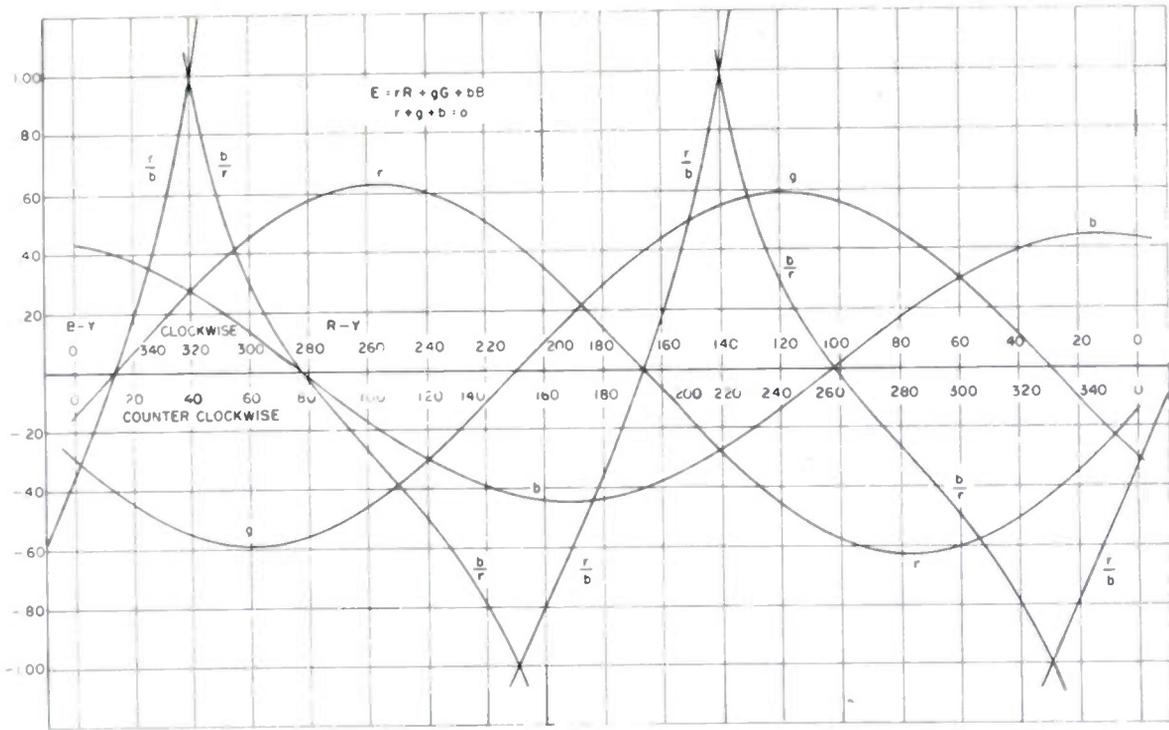


Fig. 2
Graphical Representation of the Chrominance Demodulator Output

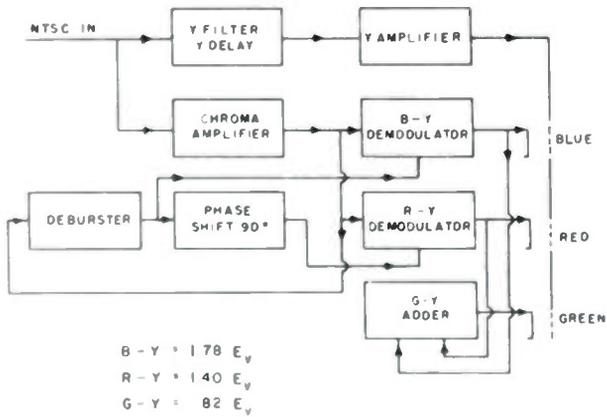


Fig. 3
Block Diagram of a Conventional Equiband Receiver

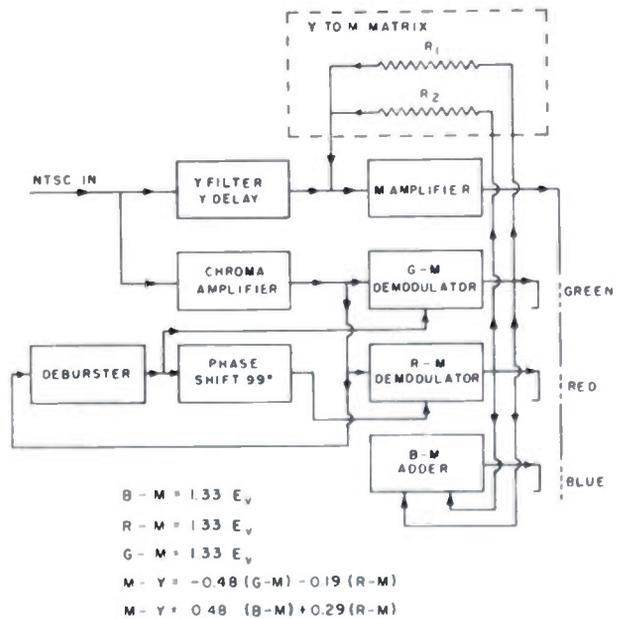


Fig. 4
Block Diagram of the New System

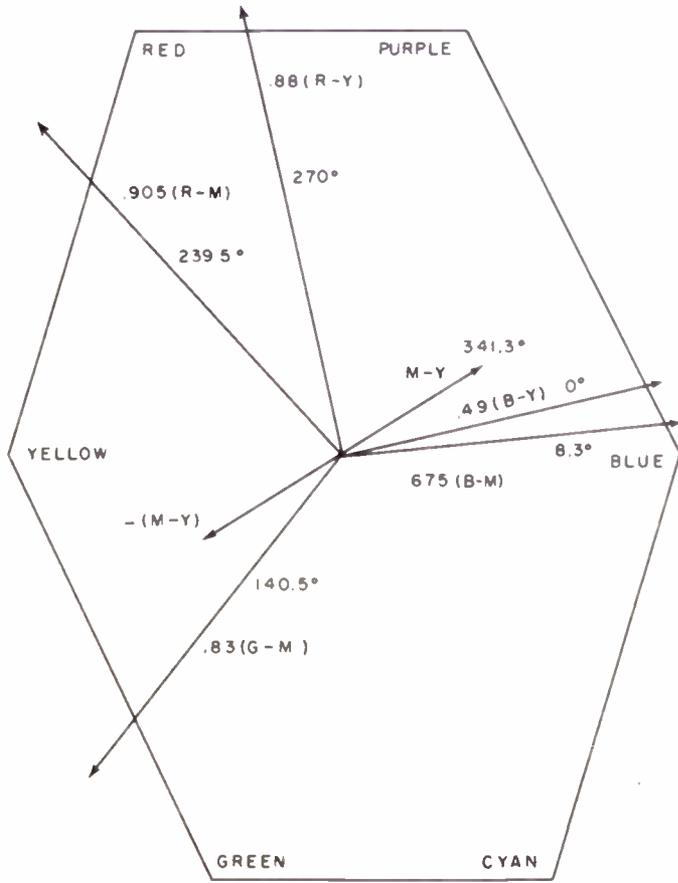


Fig. 5
Vector Diagram of the Demodulator Output and
the Maximum Demand Hexagon

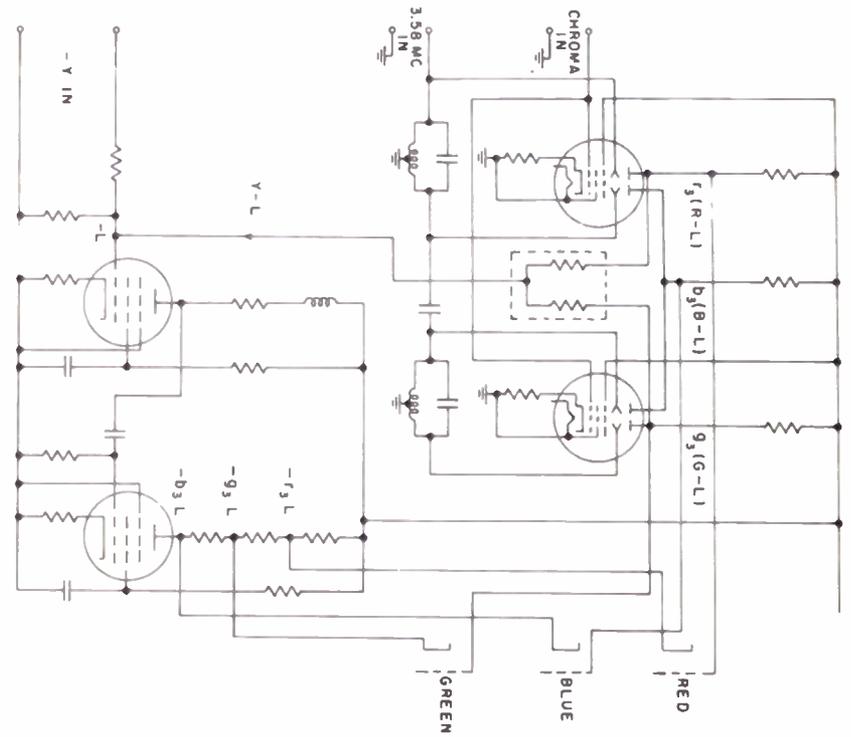


Fig. 6
Circuit Diagram of the New System for a Picture
Tube Requiring Different Drive on the Three Guns

A COLOR PROJECTION RECEIVER

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SUMMARY

This report presents some results of a program of work directed to the use of projection display assemblies for color television receivers, discusses the applicability of these assemblies, and explores briefly the important technical problems which were encountered in the studies.

The problem of providing adequate and assured accuracy of registration of the several images in a color projection assembly previously widely regarded as not subject to acceptable solution, appears to have been brought under control. Other considerations affecting the suitability of projection assemblies for use in color television receivers such as cost, convenience, pictorial performance, and reliability may now deserve renewed study; some discussion of these considerations is presented.

INTRODUCTION

In the realm of aural broadcasting, we have today a universally accepted set of transmission standards for AM broadcasting (and, to meet more stringent requirements, generally accepted standards for FM broadcasting); to complement these transmission standards, we have in the moving coil loud-speaker a device whose universal acceptance is effectively established by the fact that about 1,000,000,000 of these have been put into service. A reader with a long memory may recall how trivial a part the moving coil loud-speaker played in the early days of radio communication and even of radio broadcasting; a period of many years was required before it became clear that the moving coil loud-speaker was better suited to the needs of broadcast receivers than were any of the competing devices.

For black-and-white television, we have had for some years a fairly complete understanding of the nature of good signal transmission practices, and the standards adopted for this purpose by our Federal Communications Commission reflect this understanding well. The practice of displaying the received signals upon a directly viewed cathode ray tube, employing magnetic deflection and magnetic focus, is now virtually universal. Yet here again, the device which appears at present to be firmly established in the field was very slow in being recognized; mechanically scanned arrangements (including some miracles of ingenuity) received first attention; projection arrangements and directly viewed arrangements, electrostatic focus and mag-

netic focus, electrostatic deflection and magnetic deflection all received long and serious consideration before the general acceptance of the presently used method developed.

Color television is today, in a sense, at the point in its cycle corresponding to the position of black-and-white television about 1940, or of sound broadcasting in the middle 1920's. Color television has been provided with a set of standards for the broadcasting of its signals (and studies of the nature of human vision support fully the view that these standards are adequate). But, to pursue the parallel with sound broadcasting, no agreement exists in our industry with respect to a suitable equivalent to the loud-speaker. Not only is there no agreement among the skilled managers, merchandisers, and engineers of our industry with respect to this matter, but the measure of uncertainty is so serious that there have not yet been numerous enough and effective enough offerings to the public of different kinds of devices to establish any record of public preference which might assist us in selecting the color display unit most likely to command continued public support.

The Hazeltine organization, like many others interested in the television field, has given much study to the color television display arrangements of various kinds that appear to have some prospect of early application to household receivers. The presentation of some aspects of this study with respect to projection assemblies is the purpose of this report.

IMAGE REGISTRY

The generation of an acceptable image in full color from a set of three independent, simultaneously produced primary color images, requires that the primary color images be made to appear to coincide. Display arrangements employing multi-gun color tubes and display arrangements employing separate projection tubes appear to face this difficulty squarely in one form or another. In the multi-gun picture tube, the coincidence problem takes the form of a requirement that the several electron beams converge to a single point on the phosphor screen, with the further requirement that the convergence be uniformly acceptable over the entire screen area. Stability of convergence behavior with the passage of time, variation in supply voltage, variation in signal level, and the hazards of transportation represents a further highly desirable characteristic. In the case of the projection assembly employing three separate tubes, the coin-

idence problem is perhaps more clearly expressed as one of image registry. In the center of the picture, registration accuracy without regard to stability is easily obtained simply by differential centering adjustments on the several primary color images, and its stability with variations in time, line voltage etc. affords a direct and reliable indication of the prospects of a projection device for giving stable performance. Uniformity of registry over the entire picture area, once the central registry has been stabilized, appears to require the adherence to suitable tolerances in both the optical and the electronic elements of the three-tube projection assembly.

To illustrate somewhat better the nature of the several requirements imposed by registry considerations, and the portions of the over-all receiver design upon which those several requirements impinge, a color projection receiver may conveniently be thought of as subdivided into three portions:

In the first portion, there is the chassis (subdivided if desired) containing the circuits which handle the signal, the power supply circuits, the circuits for generating scanning currents and the circuits for furnishing the high voltage to the picture tubes. In addition to these, the chassis provides a source of focussing and centering current to the picture tubes.

Second, the receiver includes a color projection assembly with its three picture tubes. For each of these, there is provided within the assembly a deflection yoke, a focussing arrangement with provision for adjustable centering of the image, and the portion of the optical system which is individual to each picture tube. Further, the projection assembly provides the rigid and permanent mechanical support for these several pieces, the optical arrangements for superposing the three pictures in coincidence, and the means of adjustment for optical focus, central registry, and area registry. It appears that, if due regard is paid to mechanical integrity, thermal behavior, and protection against excessive external magnetic fields, these adjustments can be made upon the color projection assembly as a receiver component, and their adequacy relied upon thereafter for a period limited only by the need for replacement of a picture tube or other component of the color projection assembly.

The third division of the receiver comprises the cabinet with its provision for mounting the chassis and the color projection assembly; the cabinet also carries the projection screen and any mirror or mirrors which may be used to fold the light path and thus bring about very attractive cabinet depth dimensions. It appears to be desirable that the cabinet provide, by construction

or by adjustment, for reasonably accurate optical path length. It seems likely that reasonable precautions in the cabinet structure will provide a stable path length and that the correct initial value for the path length can be obtained either by the provision of an inexpensive adjustment in the projection assembly mounting or by the maintenance of tolerances not appreciably closer than those common in good cabinet construction,

Returning to the chassis for a moment, one further point should be noted. It is probably advantageous to center the several primary color images by means of magnetic deflection, which may for example, be derived from small distortions of a focussing magnetic field. The use of a magnetic field, from whatever source, for centering and establishing of registry for the three component images requires that the strength of this magnetic field and the strength of the electric field which accelerates the electrons in the cathode ray tubes shall continue to bear the correct relationship without disturbance from changes in values of components, supply voltage, thermal conditions, etc. Provision for this requires circuits in the chassis to regulate the high voltage as a function of the magnetic field strength, or vice versa. The arrangements for doing this must recognize that a five percent change in magnetic field requires a ten percent change in high voltage to produce the same trajectory; account must also be taken of any magnetic saturation which may be present in the relation between magnetizing current and effective magnetic field.

Observation of successful performance in a small group of developmental projection assemblies, over a limited period of time, and analysis of the causes of inadequate performance of many earlier forms, gives reasonable assurance that the requirements are now fully comprehended and have been effectively met.

USE OF PROJECTION DISPLAY ASSEMBLIES IN COLOR TELEVISION RECEIVERS

With adequate and permanent accuracy of registry reasonably assured, attention may be properly directed to the commercial and technical suitability of projection assemblies in color television receivers. In comparison with other display arrangements available now, or likely to be available in the near future, the differences in commercial and technical suitability are for the most part included in the following list. Each of the items listed is the subject of brief discussion in a subsequent paragraph:

- a. Cabinet depth.
- b. Replacement cost in event of picture tube failure.

- c. Image coincidence: whose problem?
- d. Nature and extent of color purity problem.
- e. Prospective commercial availability in large quantities.
- f. Maintenance of contrast under conditions of high ambient lighting.
- g. Ease of control of color characteristics.
- h. Light output and directivity considerations.
- i. Need for special circuitry.
- j. Inherent contrast range.
- k. Dust collection.
- l. The relative backgrounds of experience with several devices.

In the light of the studies in the Hazeltine laboratories, the first seven considerations in the foregoing list appear likely to show up differences which favor the projection type of display. The remaining considerations appear to be of either uncertain significance or of significance representing net advantage to other display forms. The relative weights of the several considerations, as they will affect a retail purchaser, are not at all clear at the present time. The discussions on each of the items which are presented in the succeeding paragraphs, supplemented by observation and study of the receivers employing projection display assemblies and of receivers employing other forms of displays may help to establish a useful commercial perspective.

a. Cabinet depth: Color television receivers using projection methods may be housed in relatively shallow cabinets. The ease with which the light path may be folded has made it readily possible to construct sample receivers producing a picture of approximately 240 square inches with a total cabinet overall depth of only 24-1/4 inches. By slight redesign the picture area can be increased to 280 square inches without an increase in cabinet depth. In comparison, receivers employing 19-inch or 21-inch shadow-mask tricolor tubes appear to require overall depths of the order of 31 inches which exceeds the standard doorway opening.

b. Replacement cost in event of picture tube failure: Tricolor picture tubes of the presently available types represent a large initial investment, and correspondingly a large replacement cost. Further, the life of a tube of this sort is terminated when failure occurs in any one of its three electron guns. In contrast, the projection tubes have individual initial costs (and presumably replacement

costs also) of the order of 1/15th of those of the tricolor tube; further, even though three of the projection tubes are employed, the failure of any one requires the replacing of only that one. The development of substantial customer responsiveness to this difference appears to be a real commercial possibility.

c. Image coincidence: whose problem? It appears entirely possible to produce a color projector display assembly in which all adjustments affecting registry are made upon the assembly before any attempt is made to install it in a receiver. It appears further that a design may be produced in which the permanence of these adjustments may be relied upon. If such a projector assembly is supplied to a receiver manufacturer, the set manufacturer himself does not encounter any registry problem. In the case of the tricolor tubes presently available, the corresponding problem of convergence has not thus far received comparable treatment.

d. The color purity problem: The convergence problem which is experienced with the three-gun tricolor tube is in some respects analogous to the registration problem with the projection arrangement. There is no similar comparison however with respect to the problem of color purity. In the case of the three-gun tricolor tube, adjustment of the tube operating conditions must be such that the "red" electron gun excites only red phosphor, etc.; this adjustment is a specific requirement in the setting up for proper operation of a receiver using one of these tubes. In the case of the projection apparatus, the "red" electron gun is in a tube which can produce only red light on the projected image and the color purity problem simply does not exist.

e. Prospective commercial availability: The important consideration here is that which will arise in the event that public demand for color television receivers shows a sudden and major growth. If this situation should develop, the expansion of production facilities for color display devices may become a real limitation to progress of the television industry. Since the cathode ray tubes of the projection assembly are small, simple, monochrome projection tubes with no unusual internal construction features, rapid expansion of manufacturing capacity for these appears possible; as for the optical and mechanical elements of the projection assembly, production capacity in very large quantity appears to be either already in existence or capable of establishment within rather reasonable periods of time. It is not clear that a comparable statement can be made for the forms of tricolor tube which have so far become available.

f. Maintenance of contrast under high ambient lighting: The projection screen must receive

light from a relatively small angular region represented by the projector; it must distribute this light over a somewhat larger angular region to the viewers. Light which strikes the screen from some other direction may be largely directed against black surfaces within the cabinet and therefore prevented from "washing out" the shadows of the picture. No comparable technique is available to directly viewed picture tubes, although the use of a dark face plate does of course give some measure of comparable effect.

g. Ease of control of color characteristics: With a three-tube projection assembly, the primary colors need not be those of the unmodified phosphors. Filters can be placed where they will affect the light from only one of the tubes, thus making it possible to modify the effect of the phosphor. In practice, this offers the advantage of permitting a wider choice of phosphors, and may well lead to either greater efficiency of operation or, alternatively, the ability to use effective primary colors of more nearly ideal characteristics than appear practical with directly-viewed tricolor tubes.

h. Light output and directivity considerations: The total available light output from the currently available shadow-mask tubes appears to be somewhat greater than that available from the projection assemblies of the sort with which we have worked. In a television receiver of the projection type, it is desirable to use a projection screen which is at least moderately directive in its distribution of the available light. The advantages gained by so doing are a substantial increase in image brightness in the favored portion of the viewing area and a likewise substantial increase in the ability of the apparatus to maintain good image contrast under conditions of high ambient lighting. Image brightnesses in the directions representing the preferred viewing positions can be made to equal or exceed those available from the other devices. The corresponding disadvantage, of course, is the readily

perceptible loss of brightness when an observer moves away from the normal viewing area.

i. Need for special circuitry: In a receiver employing a projection color assembly, it appears desirable to provide regulation of the picture tube high voltage relative to the centering magnetic fields, or vice versa. The accuracy of regulation required appears to be somewhat greater than that which has been found necessary for use with spherical-face shadow-mask tubes. As a partial or perhaps even complete offset to this, no convergence components nor circuitry are required with the projection assembly.

j. Inherent contrast range: With respect to contrast range shown by a color display device when measured in a dark room, the projection assemblies with which Hazeltine has worked thus far exhibit a lesser overall range than do the currently available spherical-face shadow-mask tubes. It appears likely that improvements in optical design which can readily be incorporated in production models will substantially increase this inherent contrast range.

k. Dust collection: Projection assemblies employ more surfaces upon which the collection of dust can impair the optical performance than is the case for directly-viewed arrangements. It is important therefore that care be given in a practical design to some reasonable measure of protection against dust entry. It appears probable that such protection can be obtained without significant additional cost.

l. Extent of general experience with the several forms of display devices: The total amount of experience among members of the industry with the use of tricolor picture tubes of the shadow-mask type is obviously many times greater than the corresponding measure of experience with color projection display arrangements. Of itself, this is a disadvantage for the projection assemblies, since only the background of substantial experience builds up confidence in the suitability of a device.

