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The Editor's Corner

REPORT OF SPECIAL TECHNICAL COMMITTEE TO DETERMINE MARKET OPPORTUNITIES FOR THE TELEPHONE*

1) THE TELEPHONE is so named by its inventor, Mr. A. G. Bell, who sees for it a vast future as a means of personal communication by voice. He believes that one day they will be installed in every residence and place of business.

2) We note that Bell's profession is that of a voice teacher, and particularly a teacher of the deaf. He appears to have no direct experience with the telephone or any other form of communication, electrical or otherwise. Yet he claims to have discovered an instrument of great practical value in communication, which has been overlooked by the thousands of workers who have spent years in this field.

3) Bell's proposal to place his instruments in almost every home and business house (and this is the only way in which their potential may be realized) is fantastic in view of the capital costs of installing the endless numbers of wires and cables that would be demanded. The central exchanges alone would represent a huge outlay in real estate and buildings, to say nothing of the electrical equipment.

4) Bell expects that the public will use his instruments without the aid of trained operators. Any telegraph engineer will at once see the fallacy in this plan. The public simply cannot be trusted to handle technical communications equipment. In any home where there are children, to mention only one point, there would inevitably be a high rate of breakage and frivolous use of the instruments. Furthermore, when making a call the subscriber must give the desired number verbally to the operator. No one on this Committee would like to be that operator, and have to deal with persons who may be illiterate, speak with lisps or stammer, have foreign accents, or who may be sleepy or intoxicated when making a call.

5) While every telegram constitutes in itself a written record of what has been communicated, Bell's instrument uses nothing but the voice, which cannot be captured in any concrete form, and therefore there would be no record of what was said or agreed upon. We leave it to you to judge whether any sensible man of business would transact his affairs by such a means of communication.

6) Bell expects that the subscribers to this service will pay to have the instruments installed in their premises and will thereafter pay for each call made, with a monthly minimum even if no calls are made. We feel it very unlikely that any substantial number of people will agree to such an arrangement, in view of the telegraph offices which are now giving efficient round-the-clock service in every neighborhood and in the smallest towns, which charge only for actual messages sent according to length.

7) In conclusion, this Committee feels it must advise against any investment whatever in Bell's scheme. We do not doubt that it will find a few uses in special circumstances, such as between the bridge of a ship and the engine rooms, but any development of the kind and scale which Bell so fondly imagines is utterly out of the question.

Editor's Note: We are indebted to Peter W. Tappan of Warwick Manufacturing Company for the above report, whose author is unknown. We have at times wondered about the origin of ingenious and entertaining stories that are often told. There is rumor of the existence of a highly secret society having members of international fame, who prefer to remain anonymous. As a free public service and purely for the fun of it they create and disseminate some of these tales. Be that as it may, we traced the story from Pete Tappan to C. F. Hurc of the University of Wisconsin, Madison, who got it from Dr. Kenneth S. Colemen of AMF in Springdale, Conn., who said that the originator is unknown.

—MARVIN CAMRAS

* Received by the PGA, March 23, 1961.

PGA News

NEW OFFICERS ELECTED

Results of the balloting were announced March 21, 1961, at the IRE International Convention. The new PGA officers are:

Cyril M. Harris, Chairman of the Administrative Committee
H. E. Roys, Vice Chairman
Donald E. Brinkerhoff, Member of the Administrative Committee 1961–63
Frank A. Commerci, Member of the Administrative Committee 1961–63

PGA AWARD WINNERS ANNOUNCED

The Awards Committee has announced the following awards for 1960:

William B. Snow, PGA Achievement Award
Donald F. Eldridge, PGA Senior Award
William D. Roehr, PGA Award

Further details will be given in a subsequent issue of these TRANSACTIONS.

CHAPTER NEWS

Albuquerque-Los Alamos

The program for the Albuquerque-Los Alamos Chapter of PGA scheduled the following meetings during the 1961 winter-spring season:

"A Look at Tape Recorders," by Gerry Mayfield of The Audio Center, for February, held at The Audio Center.

"Tour and Discussion of KHFM Broadcasting Facilities," by David Annet, Chief Engineer KHFM; at Station KHFM in March.

"Tour and Inspection of Sandia Crest KGGM-TV Transmitter," by Braffet, Riggan, and Lemmon of KGGM-TV. The tour and picnic is in the Sandia Mountains for the May Program.

ANNOUNCEMENTS

PGA Clarifies Field of Interest

As approved recently by the PGA Administrative Committee, and further approved by the IRE Executive Committee at its recent meeting, Article 3, Section 1, of the PGA Constitution now reads as follows:

> The Field of Interest of the Group shall be the technology of communications at audio frequencies and of the audio-frequency portion of radio-frequency systems, including the acoustic terminations and room acoustics of such systems, and the recording and reproduction from recordings and shall include scientific, technical, industrial or other activities that contribute to this field, or utilize the techniques or products of this field, subject, as the art develops, to additions, subtractions, or other modifications directed or approved by the IRE Committee on Professional Groups.

Previously, the Field of Interest was recording and reproduction from recordings at audio frequencies.

Any Constitutional amendment approved by the Administrative Committee and by the IRE Executive Committee will go into effect unless 10 per cent of the group members file objections to it with the Executive Secretary of the IRE within 30 days of publication.

The preceding Constitutional amendment is in line with recent efforts on the part of the IRE to clarify and avoid conflicts within the fields of interest of professional groups. It will also take into account the fact that recording, even for audio, is now often done at high frequency.

Binders Now Available for TRANSACTIONS

Binders of Spanish-grain fabrikoid with gold lettering, ruggedly built, with mechanism for holding copies of the PROCEEDINGS and the TRANSACTIONS in place, are available. Copies are not damaged by their insertion; each individual copy will lie flat when the pages are turned, and the copies can be removed from the binder in a few seconds. These sturdy binders will protect your file of IRE publications from damage and loss.

TRANSACTIONS binders are gold lettered "IRE PROFESSIONAL GROUP TRANSACTIONS," 4-inch capacity, with twenty-four steel blades, in maroon fabrikoid. They are available at $3.00 for the standard binder. For a slight additional cost you may have your name, year, group, etc., imprinted. Please write to IRE Headquarters for further details.

Enhanced Stereo*

ROBERT W. BENSON†, MEMBER, IRE

Summary—**Practices utilized in producing a stereo recording are discussed relative to the performance of stereo-sound reinforcement systems. The reproduction of these recordings results in an enhanced stereo effect.**

THE current use of stereo recordings in the home has required the recording studio to make deviations from practices used in the original concept of stereo-sound reinforcement. In its beginning, stereo was first applied to sound reinforcement systems without the complexities of recording processes. In order to make stereo reproduction more acceptable to the consumer, it has been necessary to deviate considerably from those practices established in sound reinforcement systems utilizing the stereo concept. This paper is concerned with a discussion of those factors involved in making a suitable stereo reproduction and how the actual reproductions deviate from those which were involved in the concept of a stereo-sound reinforcement system.

Single Channel Systems

The first use of electronic amplification for providing sound reinforcement utilized a single-channel system. Although multiple microphones and loudspeakers may have been used, the electronic system had but a single amplification channel. All microphone outputs are combined to a single amplifier input and all loud speakers are combined to provide a single-channel output.[1] The sound arriving at the position of any listener in the audience is composed of that which reaches the listener by acoustic paths and that which reaches him through the aid of the electronic system. Although some sound intensity exists at each microphone location and sound is produced by each of the multiple loudspeakers, the nearest microphone to the performer and the nearest loudspeaker to the listener predominates. The time pattern behavior of sound arriving after being electronically amplified is a complex one, dependent upon the location of both the performer and of the auditor. In addition, a major part of the sound arrives at the position of the auditor due to the natural sound reinforcement provided by the walls and ceiling of the auditorium. The sound which arrives at the listener by pure acoustic means is the only sound which gives a clue to the position of the actual performer. Such sound reinforcement systems are satisfactory providing that they are used to supplement but not obscure the natural sound.

Dual Channel Systems

In the stereo-sound reinforcing system, the microphones become separated into groups, two or more independent amplifying systems are used, and two or more loud-speaking systems are used. The loudspeakers are located in positions corresponding with the location of the microphones. Such a system is illustrated in Fig. 1. The physical location of microphones and loudspeakers marked L are for the left channel and those marked R are for the right channel. In many systems, a phantom third channel is derived by mixing the output of an additional microphone placed in the center part of the stage, labeled C, with the input of both channels L and R. Such an arrangement is intended to give the effect of having a three channel system. Either the two independent channels or the two channels with a phantom center, give an indication to the listener of the position of each performer on the stage. Two factors are important in providing this information: the first factor being the time difference between the sound from the performer to a channel L microphone and out of a channel L loudspeaker to the listener, and the second factor being the sound from the performer to a channel R microphone and out of a channel R loudspeaker to

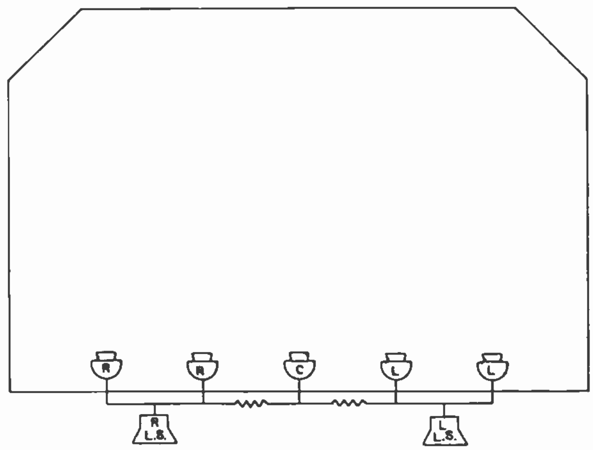


Fig. 1—Microphone placement for stereo-sound reinforcement system.

* Received by the PGA, March 8, 1961.

† Vanderbilt University, Nashville, Tenn.

[1] The system referred to is a single-channel system, commonly referred to incorrectly as a monaural system. Although if listened to by one earphone, it would be monaural, all sound reproduction in an auditorium is binaural, that is, the audience listens with both ears.

the listener. Time of arrival of signals will be dependent upon the location of both the performer and the listener. In addition, the intensity at the position of the listener will be dependent on both the location of the performer and the location of the listener. The two factors are therefore *time* and *intensity*. The sound which arrives at the position of the listener is composed of that provided through the electronic amplifying system and that which arrives by purely acoustic means. Both signals, that is, the pure acoustic and the electronically amplified signal have two components—the first being the direct sound reaching the auditor by passing through space without any reflections, and the second being that provided for with reflections from either a ceiling, wall, or combination of walls and ceiling. The time and intensity relationships between the reflected sound and the direct sound gives the listener in the audience an indication as to the size of the space in which he is listening. This is independent of whether the sound reaches him purely by acoustical means or through the electronic system. In order for the system to function properly, it is necessary that each microphone receive sound from each performer and that each loudspeaker produce sound which can be heard in every part of the auditorium. The term "heard" indicates that there will be a finite ratio between the intensities provided by the various channels. The proper delays in time must also be provided for so that the auditor can determine the possible location of the performer by comparing both time of arrival and intensity of sound. Again, such sound reinforcement systems have proven to be very satisfactory provided that the electronically aided sound is not of such intensity as to destroy the complete illusion of having natural sound. Reasonable loudness levels provided by such an amplification system can give the illusion that no sound reinforcement system is in effect at all.

Stereo Recordings

The stereo recording is an attempt to provide a more realistic reproduction of sound which is recorded in a given environment such as an auditorium. The major additional feature which the stereo or multichannel recording can provide over a single-channel recording is that of indicating the position of the various performers on the recording stage. In making a recording, it is possible to make several different choices in arriving at a suitable arrangement for providing reproduction in the home which will simulate or enhance the various factors influencing the spacial concept. The major choice left to the recording studio is the selection for the position of the microphones which are used to pick up the sound to produce the number of channels required. In current stereo practice, two channels are commonly used.

If the recording were to be made in an auditorium which has a stereophonic-sound reinforcement system installed, one procedure would be to utilize the outputs of the two independent channels and make a recording directly. Such a procedure will provide information on the recording which indicates the relative intensity at the various microphone locations and, in addition, the various times which the sound takes to travel from each performer to each independent microphone channel. The recording does not contain information regarding the location of the listener within the auditorium, nor does it contain either the reflected sound due to natural-sound reinforcement or reflected sound due to electronically-amplified-sound reinforcement. The effect of making such a recording, therefore, is to place the loudspeakers in the home in a position comparable to that of the microphone locations. In ordinary circumstances, this would place the listener essentially on the front of the stage rather than in a typical position for a listener within the auditorium.

Two other possible arrangments are used in making stereo recordings. The first of these is to move the location of the microphones further toward the audience. This means that the microphones picking up the independent channels will not only contain sound which reaches them directly, but also will contain sound which has been reflected from one or more surfaces within the auditorium. This choice of location for microphones places the listener in his living room in a place comparable to a seating position within the audience. When recordings are made by this technique, the additional reverberation present gives the auditor a sense of the size or volume of the auditorium, but unfortunately does not contain a sufficient discrimination in time and intensity for the determination of the location of various performers. In other words, this method of stereo reproduction emphasizes size instead of spacial location. Although this is a true stereo-type system, it is not the type of system which the consumer has been educated to expect. Since stereo-sound reinforcement systems give the listener both reflected and direct sound, giving an indication of the volume or size of the auditorium, it also maintains the time and intensity relationships for direct sound, giving the listener an indication of the location of the performer. The intensity feature is that which has been emphasized in commercial stereo recording.

If the second aspect of the stereo system is to be emphasized, the location of microphones is moved toward the performers rather than toward the listeners. This, in effect, puts the listener in his living room more in a location like that of the conductor of an orchestra rather than that of a typical listener in an audience. If only two microphones are to be used, the effectiveness of such a system in providing spacial clues is much like that provided by the stereo-sound reinforcement system but with a de-emphasis of reverberation characteristics of the auditorium. In recordings of this type, it is common practice to mix with both channels sound which is picked up by a microphone located within the audience area in order to convey the size of the auditorium.[2] In such a recording, the location of the various performers is indicated by both a ratio between the intensities and the times of arrival through the various channels. In addition, a background of reflected sound indicates the size of the recording studio or auditorium. Although such recordings appear to be realistic, and sound very much like the original production, they still do not emphasize the location of various performers. Since the consumer has been educated to expect an ability to locate each performer, techniques have resulted which produce an enhanced stereo affect. By using multiple microphones for each channel, it is possible to increase the ratio of intensities between one channel and another. A plan view of a recording studio with the location of microphones for channels L and R is shown in Fig. 2. In comparing the distance between any one

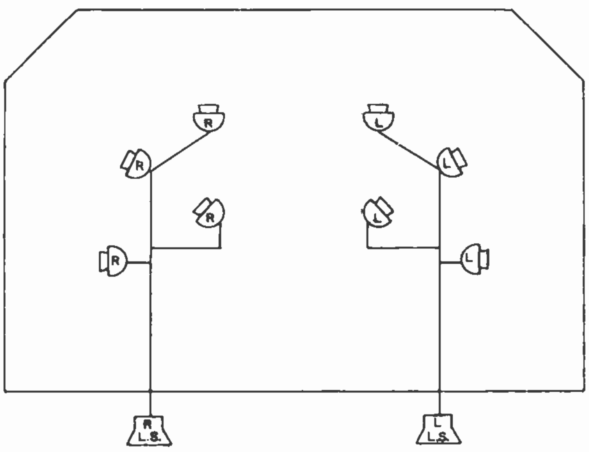


Fig. 2—Microphone placement for an enhanced stereo recording.

performer and a channel-L and a channel-R microphone, the ratios in intensity between the two channels is seen to be increased several fold over that which would have been provided in the simple two-microphone two-channel system. Furthermore, the time which the sound takes to travel to a channel-L microphone relative to the time required to arrive at a channel-R microphone, is changed considerably. Furthermore, since multiple microphones are used to pick up sound from performers near channel-L microphones, and multiple microphones are used to pick up sound from performers near channel-R microphones, the recording stage has been reduced electronically to a space which consists of essentially two points. When listening to such a recording reproduced in the home, it is possible to say that the piano is on the left because it is only heard out of the left speaker, and the bass drum is on the right because it is only heard out of the right speaker. The ratio in intensities is so great that there is no clue to the difference in times of arrival at the recording microphones. Although such a recording stituation creates a spacial illusion, it is not stereo-sound reproduction, but that which may be termed enhanced stereo—that which is provided by two points in space.

In order to provide the listener with sufficient information to locate a performer, it is necessary to provide both intensity and time of arrival information. If the ratio of the intensities is increased by the use of multiple microphones so that essentially each performer has his specific microphone or channel in which to record, then the time of arrival information is provided only by the difference between the two loudspeakers in the home and not by the position of the recording microphone. This means that the listener in his living room can tell where he is sitting but he cannot tell where the performer is located. By increasing the ratio of intensity, however, and reducing space to two points, the listener has the illusion that a performer is either sitting on the left or right and since this is an easier decision to make, he has convinced himself that such a system is a true stereo system.

[2] In some cases of studio recording, artificial reverberation is added.

Conclusions

In order to provide faithful sound reproduction, the sound reinforcement system utilized in auditoriums provides natural reinforcement of intensities with respect to both the location of the performer and the listener. Furthermore, the amplifying system is positioned in the auditorium to provide consistent time of arrival information for both direct and reflected sounds. In making a recording, it is difficult to choose a location for microphones with respect to the audience and the performers, which will provide all of the information provided by a natural-sound reinforcement system. Since the aspect of location of the performer has been stressed as being the most important feature of a stereo system, the easiest method to emphasize this characteristic is to increase the ratio in intensities between the two channels for a given performer location. In so doing, the natural environment has been destroyed in order to provide an enhanced stereo effect.

A New Stereophonic Amplifier*

NORMAN H. CROWHURST†, SENIOR MEMBER, IRE

Summary—A central feature of the new design of a stereo amplifier is an output transformer with original features that makes possible reduced cost and improved performance at the same time.

This paper discusses a varied possibility of design objectives for a stereo system, and explains the way in which the new output transformer functions. By variation in its method of use, or in choice of parameters, a whole range of amplifiers can apply advantages in different proportions or degrees.

The basic design of an output transformer, which is essentially inexpensive to make, provides for separation between "left" and "right" as well as crossover, and combining networks for mixed lows, if desired, without additional external circuits. It makes possible a new type of tone control, achieving high performance economically, using feedback, and/or improved matching between amplifier and loudspeakers over the entire frequency range as well as better separation and efficiency than the single-ended and push-pull transformer matrix can give.

One particular amplifier is discussed in detail, while a more general discussion shows possible application to more diverse design objectives.

Design Objectives

While the primary objective behind this development was economy—making it possible to achieve stereo of reasonable quality at considerably reduced cost—it was also agreed that realistic standards of quality must be met; perhaps it would be possible to achieve superior quality at lower cost. This can often happen, where a simplified, more logical approach replaces an older, more complex one. It has happened in the group of systems to be described in this paper. While there is one central feature that characterizes each amplifier of the group developed, its application is so flexible that a whole range of amplifiers has been designed to suit the entire range needs of a phonograph line.

The most expensive part of any audio amplifier has always been the output transformer. Some economy was effected in an earlier development[1] that used a circuit configuration similar to single-channel push-pull, but carried one stereo channel in each side of the push and pull. This also effected some economy in output transformer by utilizing a closed-cored, high-efficiency unit for the push-pull, or monophonic element, with a gapped, lower-efficiency unit for the single-ended, or stereo element.

The new development carries this economy much further. Efficiency is improved because the current for each channel has only to flow through one primary winding and one secondary winding, where the two-transformer matrix uses two windings in series for both primary and secondary (Fig. 1). Also the maintenance of good separation is less dependent on precise control of the number of turns in the various windings.

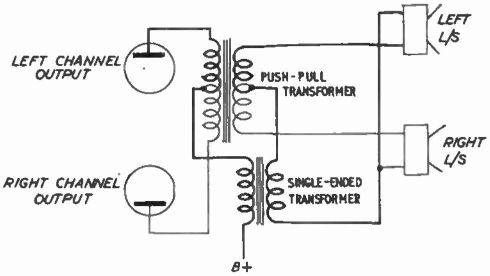


Fig. 1—The two-transformer matrix limits the economy that can be achieved because the signal current path for each channel flows through a winding on both transformers in series, both primary and secondary; current path for one channel is shown by heavy lines.

Requirements

In deciding what is acceptable performance, the question of separation and how well it is maintained at extreme frequencies, low and high, must be considered.[2] The character of separation is also important: most measurements of separation do not determine what the signal is that leaks from one channel to the other; it is assumed to be the left program leaking into the right channel, or vice versa; what is more important is the leakage of distortion components from one channel to the other.

To illustrate this difference: if the leakage between channels is pure program, some 12 to 15 db is probably adequate for almost all purposes; improvement beyond this would not noticeably improve the stereo effect. We set a minimum of 20 db as a target to insure a good margin. But if the leakage consists of distortion components, then 20-db crosstalk represents 10 per cent distortion!

Separation, at the low frequencies particularly, resolves into two kinds from the practical program viewpoint. The test method usually employed assumes the signal is present in only one channel, and must not be allowed to leak into the other. The practical program, except the type that has been "doctored for super

* Received by the PGA, January 20, 1961; revised manuscript received, March 27, 1961.

† CBS Labs., Stamford, Conn.

[1] B. B. Bauer, *et al.*, "A two-way stereophonic amplifier," *Audio*, vol. 42, pp. 19–20, 92; October, 1958.

[2] W. H. Beaubien and H. B. Moore, "Perception of the stereophonic effect as a function of frequency," *J. Audio Engrg. Soc.*, vol. 8, pp. 76–86; April, 1960; also in IRE Trans. on Audio, vol. AU-8, pp. 144–153; September–October, 1960.

stereo," has some of the signal present in both left and right channels. Aural separation is achieved by differences in the intensity and phase with which the signal corresponding to different program sources or instruments is contained in the two channels.

The conventional method of testing for separation merely determines by how much a system will degrade the proper amount of intensity separation. For example, if a piece of program has the intensity in one channel of 14 db above that in the other, and the system separation is 20 db: the leakage represented by signal content is 1/5th, and that added by the system (assumed in phase) is 1/10th, making a total of 3/10ths; so the 20-db system separation may degrade a program separation of 14 db to only 10 db.

But program can differ in timing, or phase, as well as in intensity, between the two channels. In a normally recorded program, both differences contribute at lower frequencies because sound intensity throughout a studio will not vary appreciably at these frequencies, and timing can only change by a fraction of a period from point to point.

The effect of incorrect timing can well be illustrated by reversing the phase of one channel at the loudspeaker. On some programs, all sense of stereo location is lost, but an exaggerated sense of spaciousness remains. In others, where a different recording technique was used, there may be little difference. But, however the program is recorded, the reversing phase of one speaker always gives an impression of bass deficiency. The only case where this may not occur is in the "super-stereo" type program, if all the bass is in one channel. This indicates that in most program material, both correct phase or time relationship and intensity differences are important at these frequencies.

Of course, coupling between channels will degrade phase as well as intensity difference, but in a different way. The distinction is mentioned here because it can explain some of the discrepancies between experimental tests conducted to determine the need for maintaining separation at the extreme low and high frequencies.

The Output Transformer

A new departure in output transformer design forms the central feature of the new line of amplifiers, with different aspects of its flexible range of attributes utilized for different applications. To understand the functioning of this transformer, it will be necessary to explain the properties of the quantity called leakage inductance, and we can illustrate the development of the new unit in terms of earlier applications.

Leakage Inductance

For filter design, leakage inductance combines properties of iron-cored and air-cored inductances. An air-cored inductance uses a very long magnetic path in air, linking with the coil [Fig. 2(a)]. An iron-cored inductance occupies this path with magnetic material of high permeability [Fig. 2(b)]. This greatly increases the Q of the inductor, but also introduces a nonlinear component, due to the nonlinear relationship between B and H in the magnetic core.

Leakage inductance is usually defined as a measure of the unshared magnetic flux due to imperfect coupling between the coils and is quantitatively dependent on the primary flux that does not link the secondary, and vice versa. While this is a correct definition, the concept of it as imperfect coupling leads to the notion that leakage flux is a small percentage of the main flux, and possesses the same nonlinearities the main flux does which is not true.

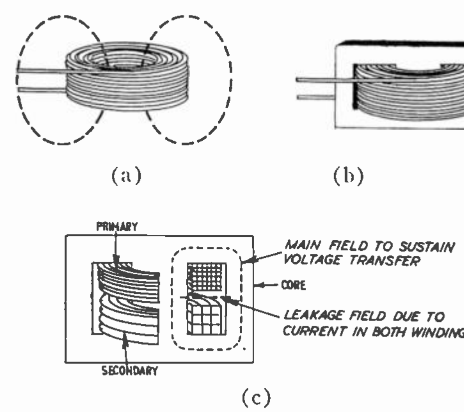


Fig. 2—Comparison of properties of (a) air cored, (b) iron cored and (c) leakage inductances, for use in filter design; the leakage inductance arrangement is shown in part section with the paths of main and leakage field indicated.

An alternative way to define leakage inductance avoids this difficulty. All the flux that stays in the core couples both primary and secondary in most designs. (In special designs magnetic material may be inserted in the leakage flux path, but these will not be considered in the present paper.) The leakage flux is induced to pass between the coils [Fig. 2(c)] by the combined current in both windings (principally due to load current drawn from the secondary and reflected to the primary) and is responsible for a voltage difference that is added vectorially to the voltage that would be induced in the windings by the main flux. So leakage inductance is the inductance between two coils occupying the same core due to a magnetic path between them, which path, usually, is wholly in air. Leakage flux is only such when it leaves the surrounding core. From this understanding, the properties of leakage inductance may be seen to differ from conventional inductance in two respects.

1) Although it has the properties of an inductance, it provides circuit isolation because two coils are involved, which may or may not be connected externally. The inductance is defined in terms of the

voltage induced in both coils, referred to the turns in one of them, by the rate of current change in both coils.

2) It has a Q considerably superior to an air-cored coil because the magnetic path length, although in air, is considerably reduced. At the same time, it does not, in itself, have the distortion-generating property of an iron-cored inductor. The leakage inductance is entirely due to magnetic path *in air*. The part of the magnetic path within the core material is not active in the leakage inductance element, but is part of the magnetizing current characteristic of the transformer of which leakage inductance is another element.

The magnitude of leakage inductance can be varied in several ways. In considering its magnitude, reference must be made to a specific winding. The same leakage inductance will have two distinct values, if the windings between which it appears do not have the same number of turns. Leakage inductance is proportional to the square of the number of turns in the reference winding and is governed also by the geometry of the windings.

Increasing turn length, the spacing between windings, or the dimension of the windings transverse to the leakage path, will increase leakage inductance [Fig. 3(a)]. Increasing the dimensions parallel to the leakage path, or multiplying the number of winding sections by division made parallel with the leakage path [Fig. 3(b) and (c)] will reduce leakage inductance.

An early use of leakage inductance for audio filter design yielded an efficient and compact low-pass filter, using the parameters generally known as m derived.[3] Later, the same principle was extended to produce a transformer with built-in crossover, for feeding low- and high-frequency loudspeakers with their respective ranges of frequency.[4] This use was the first to demonstrate a way of achieving high-pass as well as low-pass action through the use of leakage inductance (Fig. 4).

Fig. 4(a) represents a condition where the capacitor in shunt with the output winding operates in conjunction with the leakage inductance, which is effectively in series with the the transmission path to provide low-pass action. In Fig. 4(b), the position of L and C are transposed. This is achieved by using windings with the same numbers of turns, phased so the output voltage can be the drop across the leakage inductance between the two windings.

When stereo first became popular, the use of two winding assemblies on the same core as a means of economy was envisaged.[5] In this case, the leakage inductance between windings disposed on the two limbs of a core-type construction would provide potential separation between channels, and by correct phasing of the signal in the two channels, single-ended outputs could be used in such a way that the core loop has no resultant polarized (dc) magnetization.

[3] N. H. Crowhurst, "Leakage inductance," *Electronic Engrg.*, vol. 21, pp. 129–134; April, 1949.

[4] N. H. Crowhurst (Tannoy, Ltd.), "Improvements Relating to Electrical Transformers," Brit. Patent No. 734,346; July, 1955.

[5] N. H. Crowhurst, "Stereophonic Sound," John F. Rider Publisher, Inc., New York, N. Y., 1st ed.; 1957.

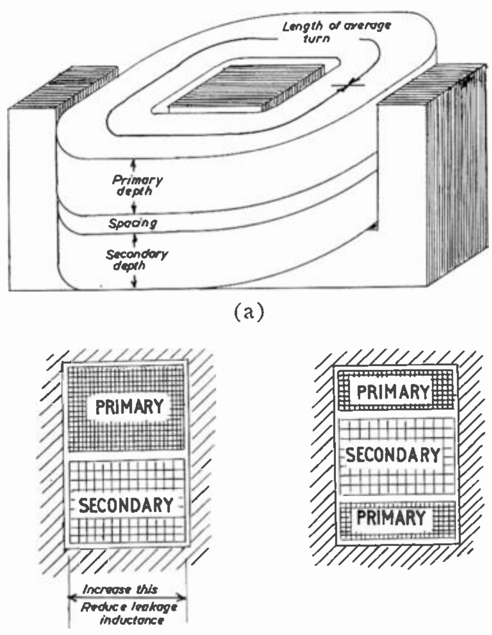


Fig. 3—Dependence of leakage inductance on the geometry of the transformer into which it is built: (a) dimensions shown here will each increase leakage inductance in proportion to themselves; (b) inverse proportion on this dimension; (c) the first simple step in sectionalizing to reduce leakage inductance.

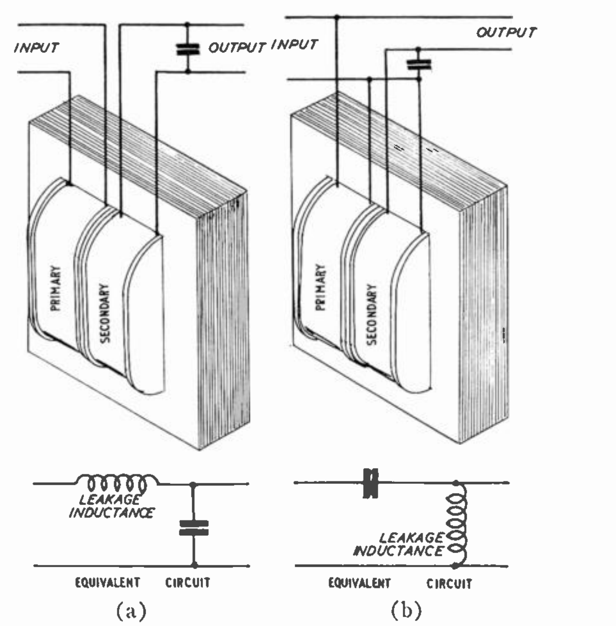


Fig. 4—Connections with double-wound transformer to utilize leakage inductance in (a) low-pass and (b) high-pass filter. Note that in (a) no electrical connection is necessary between input and output, while in (b) such connection is necessary and the turns in the windings must be equal.

The New Development

In a sense, the basic element of the new design combines the features represented in the literature.[4,5] But it does more than this; it gains some advantages due to the particular method of combination that do not pertain to either feature individually.

The basic configuration for the new transformer is shown in Fig. 5. The outputs of a push-pull stereo amplifier are accommodated on symmetrically disposed windings of a core-type transformer. Without the third winding, which incorporates the crossover function, the separation progressively improves as frequency rises due to the rising impedance of the leakage inductance between the two winding assemblies.

Basically, the third winding provides low-pass filter action between a mixture from both the main windings, which can be used to feed a common-bass unit. A capacitor connected across this third winding completes a two-element low-pass filter action. A full equivalent circuit of this joint leakage inductance action cannot conveniently be drawn with conventional circuit elements, nor can a simple equivalent circuit be constructed. The leakage inductance as measured from either main winding to the third winding, in conjunction with the leakage inductance as measured between the two main windings, do not conform to either loop or mesh configuration of a three-terminal network.

Below the crossover frequency, which is determined by the referred leakage inductance value for the turns used in conjunction with the capacitor, chosen to synthesize a filter having constant resistance properties, most of the energy is tightly coupled to the third winding. Above crossover, progressively less of the energy is coupled to this winding. But what there is, is strictly "sum" or monophonic signal.

The secondary windings on the main winding assemblies have the same number of turns as one half of the third winding. By appropriate phasing, the output voltages oppose at low frequencies, so that the left- and right-speaker terminals receive less and less signal below crossover. Above crossover, the bypassing action of the capacitor across the third winding, in conjunction with the leakage inductance between each main winding and the third winding, serves to couple the left and right speakers directly to their respective windings.

Immediately above crossover, up to a frequency where the leakage inductance between the left and right windings is directly adequate to provide separation, the presence of the third winding also serves to improve separation. The sum signal appearing across it is split in two and subtracted from the partially-coupled signals appearing across the main windings. The result is that each output is much more fully separated from the other (Fig. 6).

It should be noted that the main windings are not, in themselves, treble windings, although when separate common bass is used, their main function is to supply load current at treble frequencies. But the voltage across each main-winding secondary is tightly coupled to its respective primary winding. By grounding one end of each main-winding secondary, and the center tap of the third winding, the filter action, both low pass and high pass, can be achieved, while the main windings give full-range voltage for feedback purposes.

Notice that the entire load current for the common-bass output is delivered by the third winding. Except in the region just above crossover, the current for left and right is delivered directly from the main windings, although the third winding is in series. The capacitor across it bypasses the higher-frequency currents from the main windings, so voltage drop due to third-winding resistance is avoided. The only currents that cause voltage drop in the third winding are those below crossover, and this is then the active winding, feeding the common-bass speaker. Thus, efficiency can be maintained more easily than with the two-transformer matrix.

The circuit is more efficient, as well as costing less than either two separate transformers or a two-trans-

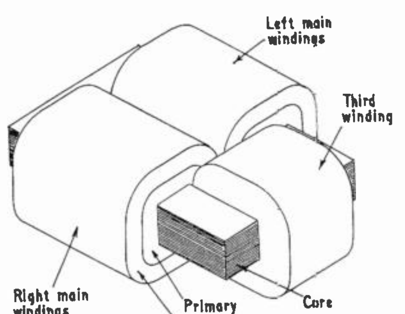


Fig. 5—Physical configuration of the single-assembly transformer.

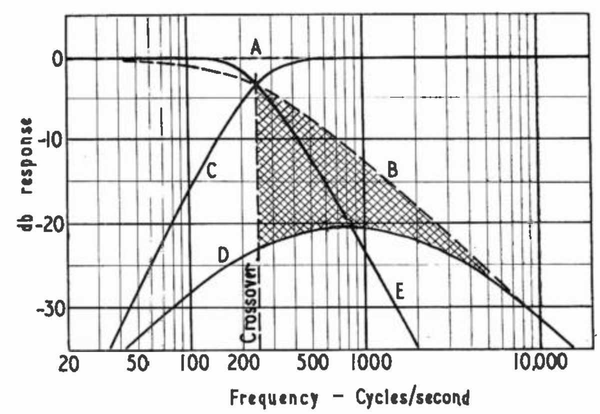


Fig. 6—Effect of the third winding and its capacitor on separation: A (dashed line), response in active channel without use of third winding; B (dashed line), response into inactive channel (breakthrough or crosstalk) without use of third winding; C (solid line), response into active channel with third winding and high-pass filter action; D, response into inactive channel (breakthrough or crosstalk); E, response into common-bass circuit.

former matrix, followed by separate LC low-pass filters to provide common bass. The common-bass filtering is provided in the magnetic circuit without resistive loss additional to normal transformer operation.

Spurious Effects

It proved important to avoid any internal resonance in the third winding of the transformer above the critically loaded one used for crossover. An early attempt used a simple winding with a center tap. Such a winding has a leakage inductance between its halves, which is very much smaller than the leakage inductance from the other windings (Fig. 7).

However the capacitor is connected—even if two separate capacitors are used to bypass each half separately—a resonance occurs between the leakage inductance from one half, regarded as the exciting winding, and the other, across which the capacitor appears as a virtual shunt load. This is a series resonant circuit that builds up a peak, injected into the opposite channel in the region of 12 kc. The disadvantage of this is that at 6 kc a signal from one single-ended amplifier contains a 12 kc distortion component that resonates in the opposite neutralizing winding to inject a high-amplitude, double-frequency cross-talk component.

The remedy was simple—bifilar winding with opposite ends of the two sections connected together to form the center tap. This makes the coupling so tight that no resonance occurs within the audio range, and where the new resonance might occur, the Q has deteriorated to much less than unity.

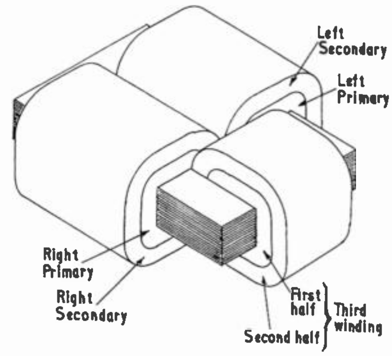


Fig. 7—When the third winding is center-tapped in conventional manner (shown here physically), the leakage inductance between its halves is considerably smaller than that from either main winding to the third winding.

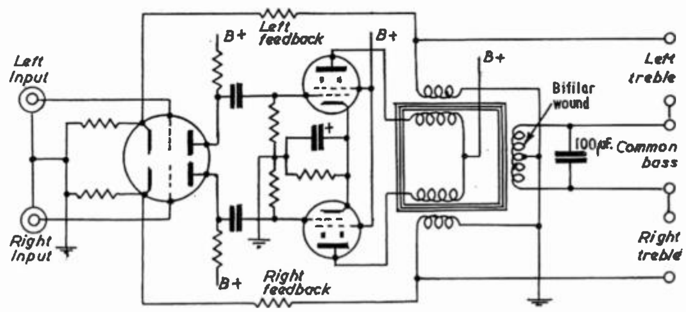


Fig. 8—Schematic of amplifier utilizing the single-assembly output transformer to provide common bass, left- and right-treble outputs, and full-range feedback to drive stage.

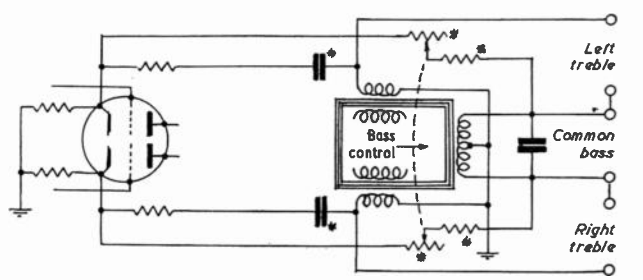


Fig. 9—Partial schematic to show the six extra components needed to add a very effective two-channel-bass control to the circuit of Fig. 8.

The Amplifier

Variations of the transformer have already been applied in several different amplifiers, but one application in particular shows how this unique circuit can produce advantages in several directions, some of which were not envisaged at the outset of this project.

First developed was a straight amplifier, to which straight feedback was applied, yielding lower distortion and improved separation (Fig. 8). This amplifier yielded enough gain to give full output with only one voltage stage, using a ceramic pickup, and with sufficient margin to allow for turning up the wick on weak records, or to accommodate the people who do not think it's loud unless it's distorted.

Tone Control

To provide tone-control facilities would require either an extra stage to provide for the loss needed to obtain boosts as well as cuts, or the insertion of reverse-type controls in the over-all feedback. Economy suggested trial of the latter method.

But inverting a bass tone control, to be terminated by the input-stage cathode resistor, is not easy. Then it was realized that the transformer contains its own bass filter, already used for the common-bass output. By augmenting or attenuating the feedback obtained from the appropriate half of the third winding, and attenuating the low-frequency feedback from the main winding by using a suitable value of capacitor in series with its feedback resistor, we have a full-bass tone control (Fig. 9).

This is where the circuit gives a very useful bonus. The low-pass filter that separates bass from mid-range uses a two-reactance filter, enabling a sharper slope boost and cut than is possible with conventional circuits using only one reactance (usually a capacitor) for this purpose. This enables the level of frequencies below crossover (chosen as 250 cycles, which is the center of the musical scale) to be varied quite drastically without noticeable change in gain above this point in mid-range. The tone-control action is superior to many circuits that cost very much more in terms of components and gain.

To complete the tone control, using a similar method for separating the higher frequencies, first a maximum lift is inserted by connecting a resistor and capacitor across the cathode resistor (Fig. 10). Then a high-pass resistor and capacitor pick off frequencies above the point where the lift commences, to provide the adjustable element of feedback above this point. A simple variable resistance, bypassing the fixed one for mid-range gain, completes the treble tone control.

The resulting amplifier produces frequency response, distortion and separation characteristics that represent considerable improvement over its predecessors in a comparable price range; it also has extremely effective tone controls, without the need for extra stages, and because of the fewer components, it effects quite a cost-saving too. The new development makes it easy to improve performance and cut cost at the same time.

Other Variations

This is only one of a group of amplifiers developed, using the same central principle, with variations. A less expensive version omitted the third common-bass winding, but retained feedback, using it for volume control function including compensation that makes it effectively a loudness control. This produces lower distortion and better bass than was possible in this cost bracket previously, without any increase in cost.

For the really high-quality applications, many variations are possible. If the tone control function is separated to its more conventional location in a preamplifier a greater amount of feedback is used in the power amplifier section to reduce distortion almost to vanishing point. If the common-bass feature is not desired, a change in connections enables two full-range loudspeakers to be used (Fig. 11).

There are other possible advantages of the new transformer and its associated circuitry. For example, by using the common-bass winding in reversed polarity, with full-range units, the impedance matching from the tubes can be made to suit the impedance characteristic of a loudspeaker with its rising value in the bass (Fig. 12).

Above crossover, the main windings only provide separated left and right outputs. Below crossover, the fairly sudden coupling of the third winding adds turns in series with each output so that the impedance match is for a higher value. This enables an amplifier to deliver greater power at the low frequencies where it is sometimes needed without sacrificing the damping that feedback can give, or limiting the properly matched power available for the mid-range and higher frequencies.

Use of this development is not confined to tubes or to a single-ended operation. It may provide its most effective economy to single-ended circuits, but it retains its other advantages, including economic, with circuits using two separate channels with push-pull outputs, or using transistors instead of tubes.

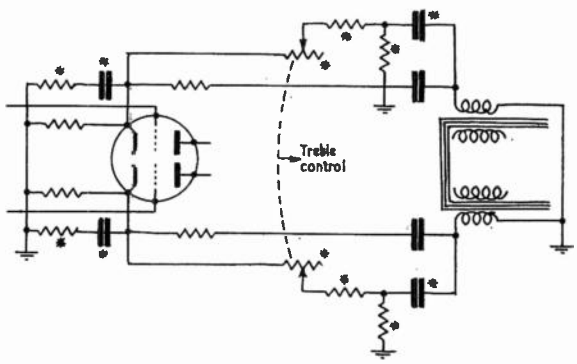


Fig. 10—Circuit of a two-channel treble control, using the additional components marked with asterisks; this circuit is partial, added to the basic circuits of Figs. 8 and 9.

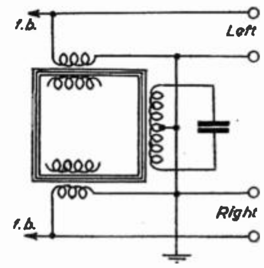


Fig. 11—Change in output connections needed when two full-range speakers are to be used instead of common bass.

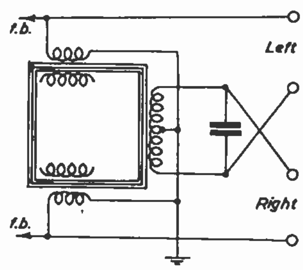


Fig. 12—A further change in connections that can be used to improve matching of the amplifier to the loudspeaker impedance over the whole frequency range.

Conclusion

The new transformer principle which forms the heart of the new development provides a number of advantages, which may be used in various combinations or degrees in individual amplifier designs:

1) An efficient means of combining mixed lows and retaining separate left and right at frequencies above a crossover built into the design of the transformer.
2) Improved separation for the degree of circuit complexity or precision involved, both in the range immediately above crossover and the extreme high frequencies.
3) The provision of convenient take-off points which may be used for full-range feedback, or for the inclusion of feedback bass tone control, with superior performance; the addition of circuit to include treble control is relatively simple.

Three-Speaker Stereo

At the cost of new defects, the defects of two-speaker stereo may be partly overcome by adding a third speaker midway between the other two and feeding this speaker an equal mixture of the left and right signals. This arrangement appears to have been invented by Steinberg and Snow[4] and was first experimented with by them in the early thirties. In recent years the arrangement has been endorsed and expounded on at great length by Klipsch[5–10] and others.[11,12] Following Klipsch's terminology, the author will henceforth refer to this arrangement as 2-3 stereo (two channels, three speakers), and to the conventional two-speaker array as 2-2 stereo. The letter T will be added where appropriate to indicate that the left and right speakers are toed in (*e.g.*, 2-3T stereo).

Naturally, with 2-3 or 2-3T stereo, it is important to feed the center speaker an in-phase mixture of the two channels rather than an out-of-phase mixture, for otherwise central sources would not be reproduced by this speaker and its whole purpose would be defeated. It is also important that this speaker be in phase with the other two. The optimum signal level to be fed this speaker may depend on the recording, the speaker spacing and orientation, the room acoustics, and the whim of the listener, so that a variable control is desirable; but it has been found that at least for flank-to-flank spacings of ten to twenty feet, a reasonable level to start with is minus six db. That is, a signal in either channel alone produces a signal across the voice coil of the center speaker that is six db below that across the corresponding flanking speaker; with one volt across the left speaker, and nothing across the right, there will be half a volt across the center, assuming all speakers to be identical. It should be noted that with this choice of center level, a central source (monophonic signal) will be reproduced with equal intensity by all three speakers.

Let us now examine the advantages of 2-3 stereo over 2-2 stereo. First, it is often pointed out that the center speaker can compensate for a recording with exaggerated separation which leaves a hole in the middle. If the center level is high enough to do that, however, it is then high enough to ruin the separation of more normal recordings, so that a control is necessary. The same effect, incidentally, can be achieved at much less expense with a blend control. It is also pointed out frequently that the center speaker can fill in the hole in the middle resulting from an extremely wide speaker spacing. While this is certainly a valid usage, under normal domestic conditions an acceptable substitute would be to put the speakers closer together. It seems to the author that the real value of the center speaker, seldom mentioned, is in the definition and stability that it gives to centeral sources. With the listener on axis, these sources appear less enlarged and their location more precisely defined. Intuitively it seems natural that this should be so, and the intuitive judgment stands up under closer analysis. The magnitudes of the three blurring factors discussed previously are all lessened by the presence of the center speaker. For a given head rotation, the intensity and time differences at the ears more closely approximate those due to a small source than when the center speaker is absent. Frequency response differences at the ears, caused by response differences in the two channels or in the flanking speakers, are lessened when the center speaker is added. Similarly, frequency response differences at the two ears, resulting from differential time delays, are also lessened.

For the listener off axis, similar benefits accrue, especially if the flanking speakers are toed in. Center sources appear better focused for the above reasons, and because no intensity-time compensation is needed for sound from the center speaker. Also, the center speaker helps to "anchor" central sources, preventing them from wandering as much when the listener moves to a location where the intensity-time compensation is faulty.

Unfortunately, these advantages are obtained at the cost of a considerable loss of spread or separation. Because flanking sources are reproduced not only by their respective flanking speakers but also by the center speaker perhaps six db down, their apparent positions are shifted perhaps a quarter of the distance to the center, for an on-axis listener. One may, of course, increase the speaker spacing to compensate for this, if the room dimensions permit, but doing so aggravates some of the undesirable effects discussed previously. It is interesting to note, however, that in some rather carefully controlled listening tests reported by Moore,[13] *on-axis* listeners preferred a 2-3 system over a 2-2 system with the same ten-foot spacing, in spite of the reduced spread, when the level of the center speaker was minus six db.

[4] J. C. Steinberg and W. B. Snow, "Auditory perspective-physical factors," *Elec. Engrg.*, vol. 53, pp. 12–17; January, 1934.

[5] P. W. Klipsch, "Stereophonic sound with two tracks, three channels by means of a phantom circuit (2PH3)," *J. Audio Engrg. Soc.*, vol. 6, pp. 118–123; April, 1958.

[6] P. W. Klipsch, "Three-channel stereo playback of two tracks derived from three microphones," IRE Trans. on Audio, vol. AU-7, pp. 34–36; March–April, 1959.

[7] P. W. Klipsch, "Wide-stage stereo," IRE Trans. on Audio, vol. AU-7, pp. 93–96; July–August, 1959.

[8] P. W. Klipsch, "Circuits for three-channel stereo playback derived from two sound tracks," IRE Trans. on Audio, vol. AU-7, pp. 161–165; November–December, 1959.

[9] P. W. Klipsch, "Experiments and experiences in stereo," IRE Trans. on Audio, vol. AU-8, pp. 91–94; May–June, 1960.

[10] P. W. Klipsch, "Signal mutuality in stereo systems," IRE Trans. on Audio, vol. AU-8, pp. 168–173; September–October, 1960.

[11] T. G. Dyar, "Three speaker stereo," *Audio Engrg. Soc.*, Preprint No. 121; October, 1959.

[12] J. M. Eargle, "Stereophonic localization: an analysis of listener reactions to current techniques," IRE Trans. on Audio, vol. AU-8, pp. 174–178; September–October, 1960.

[13] H. B. Moore, "Listener ratings of stereophonic systems," IRE Trans. on Audio, vol. AU-8, pp. 153–160; September–October, 1960.

Perhaps the worst flaw of the 2-3 system, and one that, in the author's opinion, is probably most responsible for the poor opinion some people have of the arrangement, is noticed when the listener is off axis and the flanking speakers are not toed in. If the listener moves to the right, far-left sources move toward the center because of the precedence and intensity effects discussed previously. If the center speaker is only six db down, the listener does not have to move very far to the right before left sources seem to come essentially from the center speaker, and the apparent spread has been halved. It is probably this phenomenon that accounts for Moore's poor *off-axis* results with the 2-3 system as compared with the 2-2 system. With a proper toe-in arrangment, however, this defect may be greatly mitigated.

In summary, properly executed 2-3T stereo represents a significant improvement over 2-2T stereo, but suffers from a reduced spread.

Xtereo

In conventional 2-3 stereo, a fraction of the sum of the two channel signals is fed to the center speaker, while the flanking speakers receive the same signals as in 2-2 stereo. Hence, the sum content of the program has been increased, while the difference content is unaffected. Thus, another way of explaining the reduced spread is to say that the ratio of sum to difference has been increased, thereby rendering the program more monophonic in character.

In the system to be proposed, the ratio is electrically decreased to compensate for this effect. Now, the basic idea of electrically decreasing the ratio of sum to difference is not new, as it has been discussed in the literature[14-17] and is currently used commercially by Zenith to enhance the apparent separation afforded by one-piece stereo consoles wherein the speakers are only two or three feet apart. To the author's knowledge, however, the technique has never been knowingly applied to a three-speaker system; and it will be seen in the discussion that follows that the principal mechanism of its action is somewhat different in this case. In order to have a name for this new system, let us call it Xtereo, or 2-3TX stereo.

What we would really like to have is complete independence of the signals to the three speakers. That is, we would like to be able to feed far-left source signals to the left speaker alone, dead-center source signals to the center speaker alone, and far-right source signals to the right speaker alone. Unfortunately, of course, we cannot do this without three genuine channels to start with. But let us see how close 2-3T comes to this, and how much closer we can get with 2-3TX.

We may regard the signals to the three speakers as coming from three pseudochannels, having certain signal-to-crosstalk ratios. The ratios for each pseudochannel are those between the voltage, produced by a signal in the pseudochannel, across the intended speaker; and the voltages, produced by that same signal, across the other two speakers. In a true three-channel system (3-3), these ratios may be infinite (at least in theory). In 2-3 systems, on the other hand, they cannot all be infinite. To determine the ratios, let us assume that the three speakers are identical (including their impedances). Let us call the voltage in the left channel V_l and that in the right channel V_r. Further, let us call the voltages across the three speakers L, C, and R, respectively. In a conventional 2-3 system, then, $L=V_l$, $R=V_r$, and $C=k(V_l+V_r)$; where k is some number between zero and one, such as one half. When $V_r=V_l$, we have a central source or signal in the center pseudochannel alone. The voltages across the three speakers are then $L=V_l$, $C=2kV_l$, and $R=V_l$. The signal-to-crosstalk ratios of the center pseudochannel are thus $C/L=C/R=2k$. For example, if $k=\frac{1}{2}$, then $C/L=C/R=1=0$ db [see Fig. 1(c)]. This does not seem like much of a ratio when compared with normal communications practice, but it must be borne in mind that a large ratio is not needed for a central source because the crosstalk tends to be imaged in the center anyway. After all, for a 2-2 system, $C/L=C/R=0$!

When there is a signal in the left pseudochannel alone, $V_r=0$ and the voltages across the three speakers are $L=V_l$, $C=kV_l$, and $R=0$. The left signal-to-crosstalk ratios are thus $L/C=1/k$ and $L/R=\infty$. Obviously, L/R is perfect, but L/C leaves much to be desired if

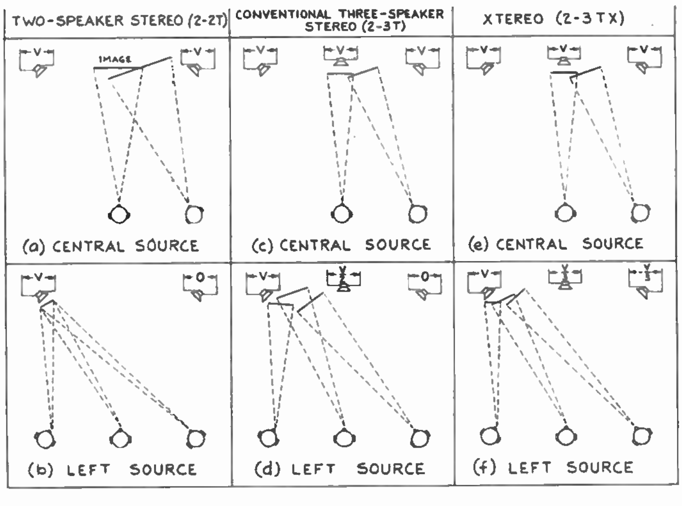


Fig. 1—Comparison of two-channel stereo-speaker arrangements.

[14] T. T. Sandel, *et al.*, "Localization of sound from single and paired sources," *J. Acoust. Soc. Am.*, vol. 27, pp. 842–852; September, 1955.

[15] H. A. M. Clark, *et al.*, "The 'Stereosonic' recording and reproducing system," IRE Trans. on Audio, vol. AU-5, pp. 96–111; July–August, 1957.

[16] P. A. Stark, "A continuously variable stereo dimension control," *Audio*, vol. 43, p. 22; July, 1959.

[17] R. Glasgal, "A dimension control for stereo," *Electronics World*, vol. 65, pp. 56–57 and p. 116; April, 1961.

$k \geq \frac{1}{2}$ [see Fig. 1(d)]. Similarly, the right ratios are $R/C = 1/k$ and $R/L = \infty$, and the same comments apply.

Let us now calculate the same ratios for the Xtereo system. In this system, the voltages across the three speakers will be $L = V_l - dV_r$, $R = V_r - dV_l$, and $C = k(L+R) = k(1-d)(V_l + V_r)$, where d is some number between zero and one half. In other words, we add a fraction d of the right channel signal in reverse phase to the left channel, and vice versa. Incidentally, this is equivalent to multiplying the sum signal by $1-d$ and the difference signal by $1+d$.

When there is a signal in the center pseudochannel alone, $V_r = V_l$, and hence, $L = (1-d)V_l$, $C = 2k(1-d)V_l$, and $R = (1-d)V_l$. The center signal-to-crosstalk ratios are thus $C/L = C/R = 2k$. These are exactly the same as in the conventional 2-3 system, and central sources are hence reproduced equally by both systems [see Fig. 1(c) and (e)].

When there is a signal in the left pseudochannel alone, $V_r = 0$ and the voltages across the three speakers are $L = V_l$, $C = k(1-d)V_l$, and $R = -dV_l$. The left signal-to-crosstalk ratios are thus $L/C = 1/k(1-d)$ and $L/R = -1/d$. Similarly, the right ratios are $R/C = 1/k(1-d)$ and $R/L = -1/d$.

Obviously, if $d = 0$, we have a conventional 2-3 system. By making d greater than zero, we are able to trade excessive flank-to-center crosstalk for some flank-to-flank crosstalk. For example, suppose that $k = \frac{1}{2}$, so that a center source is reproduced equally by all three speakers. Then with a conventional 2-3 system, $L/C = R/C = 1/k = 2 = 6$ db, while $L/R = R/L = \infty$. If, with the Xtereo system, we set $d = \frac{1}{3}$, then $L/C = R/C = 1/k(1-d) = 1/(\frac{1}{2})(\frac{2}{3}) = 3 = 9.5$ db and $L/R = R/L = -1/d = -3 = 9.5$ db. Thus, we have reduced the flank-to-center crosstalk from -6 db to -9.5 db at the expense of increasing the flank-to-flank crosstalk to the same magnitude [see Fig. 1(d) and (f)]. Listening tests have shown that this restores most of the spread lost through the introduction of the center speaker.

Now, it may at first glance appear that 9.5 db is not a very good left-right separation figure, and that because the right speaker is farther from the left speaker than is the center speaker, the left-right crosstalk should be less than the left-center crosstalk. It should be noted, however, that the crosstalk produced by a flanking source in the opposite speaker is out of phase with the signal produced by that source in its own speaker. For listening positions close to the central axis, out-of-phase crosstalk shifts the apparent source much less than in-phase crosstalk, and for an accurately centered listener can actually increase the spread compared to that with no crosstalk. Thus, the effect of the flank-to-flank crosstalk, which is out of phase with the desired signal, is considerably less for the *average* listening location than the effect of the in-phase flank-to-center crosstalk even though the crosstalks are equal in magnitude. For side listening positions, of course, where the listener is much closer to one flanking speaker than to the other, the phase of the crosstalk makes very little difference.

At any rate, if one is dissatisfied with the results obtained with these values of k and d, he need merely change the values to suit himself. Table I illustrates some of the signal-to-crosstalk ratios that can be obtained with various values of k and d. The system obviously cannot completely duplicate the results obtainable with three real channels, but the essential point is that, in general, nonzero values of k and d can be found that will yield a fidelity of geometry superior to that obtainable with any other two-channel system in present use.

TABLE I
SIGNAL-TO-CROSSTALK RATIOS AS A FUNCTION OF k AND d

System	2-2	2-3				2-3X					
k	0	0.35	0.45	0.50	0.71	0.35	0.45	0.50	0.50	0.71	1.0
d	0	0	0	0	0	0.178	0.31	0.20	0.33	0.41	0.50
C/L and C/R, db	$-\infty$	−3.0	−1.0	0	3.0	−3.0	−1.0	0	0	3.0	6.0
L/C and R/C, db	∞	9.1	7.0	6.0	3.0	10.8	10.2	8.0	9.5	7.6	6.0
L/R and R/L, db	∞	∞	∞	∞	∞	15.0	10.2	14.0	9.5	7.7	6.0

SPEAKER CIRCUITS FOR TWO AMPLIFIERS

Of course, the three speakers or speaker systems may be driven by three separate amplifiers, but a more economical method is to use only the normal two amplifiers. There are only two basic circuits for connecting the three speakers to two amplifiers without an additional transformer or choke. These are the series and shunt methods illustrated in Figs. 2 and 3. The correct relative speaker polarities, and correct relative amplifier polarities for a sum signal are shown in the figures, as is the relative impedance of the center speaker as a function of k (assuming equal speaker efficiencies). The load impedance which is presented to each amplifier varies somewhat with the signal present in the other amplifier, and the values for pure-sum and pure-difference signals are shown in the figures. The value of load impedance seen by one amplifier when there is no signal in the other amplifier lies roughly midway between the sum and difference values and depends slightly on the damping factor of the other amplifier. The optimum nominal load impedance rating of each amplifier presumably lies about midway between the sum and difference values. In any event, the load impedance variation with normal values of k is not likely to affect the system performance significantly, as evidenced by the fact that amplifiers are extremely tolerant of the ill-behaved impedance vs frequency characteristic of most speakers.

A commonly used variation of the shunt circuit is shown in Fig. 4. This permits the use of speakers all having the same impedance regardless of the value of k, by driving the center speaker from appropriate taps on the output transformers. The ratio of the impedance of the center-speaker taps to the impedance of the flanking-speaker taps is shown as a function of k. A similar variation often used is shown in Fig. 5. This permits the amplifiers to be in phase for a sum signal instead of out of phase as in the other versions of the shunt circuit, when feedback from the transformer secondary is used.

Two more circuits, using a center-tapped choke, are shown in Figs. 6 and 7. These permit speakers of identical impedances to be used when $k=\frac{1}{2}$. The circuits of Figs. 4 and 5 are to be preferred for this purpose because of the expense and performance limitations of the choke, but in adding a center speaker to an existing installation, the proper taps are not always available on the output transformers.

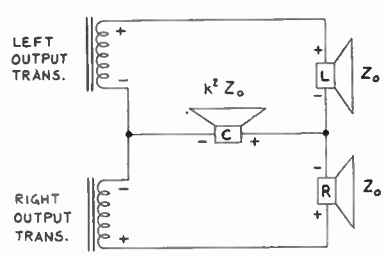


Fig. 2—Basic three-speaker two-amplifier series circuit. Z load (difference signal) $=Z_0$. Z load (sum signal) $=(1+2k^2)Z_0$.

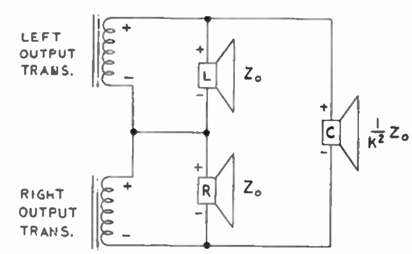


Fig. 3—Basic three-speaker two-amplifier shunt circuit. Z load (difference signal) $=Z_0$.

Z load (sum signal) $= Z_0/(1+2k^2)$.

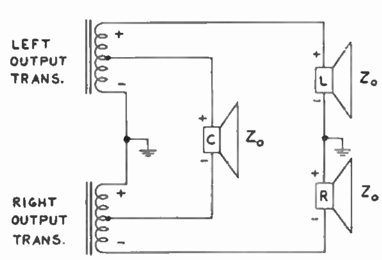


Fig. 4—Variation of shunt circuit.

$$\frac{\text{Impedance of center-speaker tap}}{\text{Impedance of flanking-speaker tap}} = k^2.$$

Equivalent Z load referred to flanking-speaker tap = same as in Fig. 3.

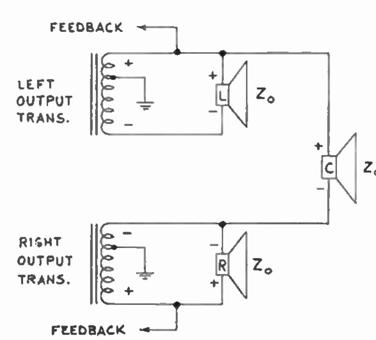


Fig. 5—Another variation of shunt circuit.

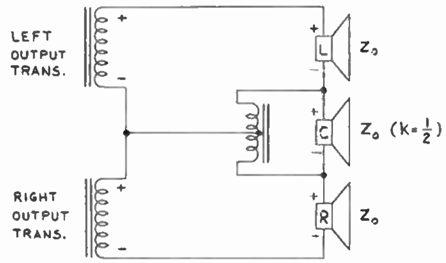


Fig. 6—Series circuit using a center-tapped choke. Z load (difference signal) $=Z_0$. Z load (sum signal) $=3Z_0/2$.

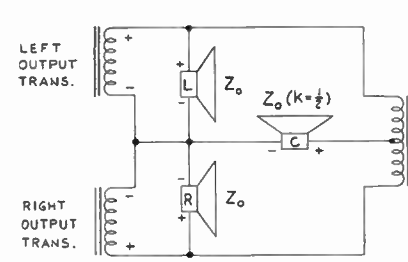


Fig. 7—Shunt circuit using a center-tapped choke. Z load (difference signal) $=Z_0$. Z load (sum signal) $=2Z_0/3$.

Circuits for Decreasing the Sum-to-Difference Ratio

There are many ways of electrically reducing the ratio of sum content to difference content in the stereo program, in order to achieve the desired value of d. A number of these will be outlined in the following discussion. Some of the methods to be described were devised by A. R. Thompson of Warwick. In every instance, it is of course possible to calculate the appropriate parameter values to achieve the desired value of d. It is usually simpler, however, to adjust these values to obtain the desired measured signal-to-crosstalk ratios at the speaker terminals.

1) An obvious method, which was used for a different purpose by Blumlein back in the early thirties or before,[15] is to matrix the two channels to obtain actual sum and difference voltages. The sum voltage is then attenuated (or amplified less than the difference voltage) and the resulting signals are rematrixed to yield new left and right signals.
2) The signal source is connected to the amplifiers in such a manner that they are driven *out* of phase by a monophonic or sum signal, or if this in inconvenient, a phase inverter is added to one amplifier. The divider network shown in Fig. 8 is then connected between the two amplifiers prior to the output stages. This attenuates the sum signal but leaves the difference signal unaffected. The resistors R need not be actual resistors but may be the output resistance of the preceding stage.
3) With the signal source connected to the amplifiers in such a manner that they are driven *in* phase by a sum signal, the cathodes of two corresponding tubes in the two amplifiers are tied together and returned to ground through a common unbypassed resistor. The tube gain for the difference signal will then be the same as if the cathode were completely bypassed, but the gain for the sum signal will be reduced. Depending on the gain reduction and bias requirements, the circuit of Fig. 9(a) or that of Fig. 9(b) may be more appropriate. It is important that this method and the preceding be applied only to portions of the amplifiers outside feedback loops, for the feedback would drastically alter the results.
4) With the signal source connected as in the preceding method, and using amplifiers wherein negative feedback is applied to the same point in each amplifier, these points are bridged by a resistor. The resistor reduces the feedback for the difference signal, thereby increasing the gain, but does not affect the sum signal.
5) Feedback is applied from some stage in each amplifier to a preceding stage in the other amplifier. This feedback will be positive for in-phase signals and negative for out-of-phase signals, or vice-versa, depending on the number of phase reversals that take place within the loop.
6) Examination of the basic series speaker circuit Fig. 2, reveals that a signal in one amplifier generates a voltage not only across its own speaker and the center speaker, but also, in reverse phase, across the opposite speaker. The magnitude of this crosstalk is negligible if the source resistance of the opposite amplifier is high compared to the speaker impedance (low-damping factor) as with pentodes without feedback, but may be significant if the damping factor is large. The same comments apply to the circuit of Fig. 6. With the circuits of Figs. 3–5 and 7, on the other hand, the crosstalk is maximized by a low damping factor.

This crosstalk is equivalent to a decrease in the sum-to-difference ratio. In fact, as shown previously, the ratio of the voltage across the proper flanking speaker to that across the opposite speaker is equal to $-1/d$. The value of d that may be obtained by this method is usually less than is desired. It may be further increased by any of the methods previosly described, or by altering the efficiency and impedance of the center speaker to maintain the effective value of k constant, but increase d. For example, in Fig. 2 the impedance of the

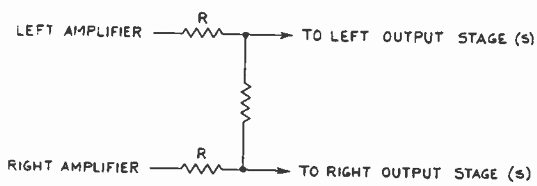
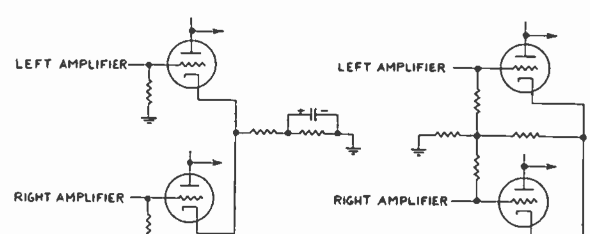


Fig. 8—Voltage-divider sum attenuator.



(a) (b)

Fig. 9—Cathode-coupled sum attenuators.

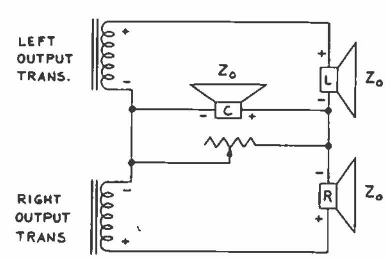


Fig. 10—Circuit for varying k and d simultaneously.

center speaker would be increased to obtain the desired flaking crosstalk, and then its efficiency would be decreased to restore the proper center level.

A useful circuit for amplifiers with high damping factors and matched speakers is shown in Fig. 10. This circuit employs a variable control, for the level of the central speaker, which simultaneously varies the value of d. When the control resistance is set at zero, the center speaker is silent and $d=0$, so that conventional 2-2 stereo is obtained. When the control resistance equals the speaker impedance, $k=\frac{1}{2}$ and $d=\frac{1}{3}$, yielding $C/L=C/R=0$ db and $L/C=L/R=R/C=R/L=9.5$ db. Ideally, the control impedance should match that of the speaker at all frequencies, but satisfactory results will usually be obtained with a resistor if the speaker impedance variations are not too severe.

Conclusions

Reproduction of the source size and location with 2-2 stereo and conventional 2-3 stereo is far from perfect. A closer approximation to 3-3 stereo may be obtained with the system described. There are many convenient ways of connecting the center speaker and obtaining the desired signal-to-crosstalk ratios. These same techniques might be applied with advantage to wide-spaced systems of three or more actual channels as used in movie theatres and large sound reinforcement systems.

Acknowledgment

The author wishes to acknowledge the guidance, encouragement, and suggestions furnished by E. S. White and J. E. Bridges.

Transient Distortion in Loudspeakers*

R. J. LARSON†, MEMBER, IRE AND A. J. ADDUCCI†, ASSOCIATE MEMBER, IRE

Summary—**The response of a loudspeaker to sudden starts and stops of its input signal is analyzed both theoretically and experimentally. Transient distortion occurs when the acoustic output level does not change as suddenly as the input signal. Waveforms of loudspeaker response to various input signals are shown, and a method for plotting a continuous transient response curve is described. The curves indicate a correlation exists between a speaker's steady-state frequency response and its transient performance.**

It was found that little correlation exists between the transient performance of a loudspeaker and musical listening tests. Two explanations are given. One discusses how the psychoacoustic performance of the ear tends to make it insensitive to the shape of the wave envelope of a tone burst. Another relates how echoes in the usual listening room tend to mask the hangover transient of the loudspeaker.

A RECENT advertisement in a high-fidelity magazine boasted of a speaker with excellent transient response. Accompanying it was a photograph of a tone burst at a certain frequency. Impressive as this picture was, just what did it really tell the reader about the quality of the speaker? Actually, very little, for up to the present, transient distortion in loudspeakers has been one of the most rarely measured factors in the reproduction of sound. Not just the layman, but even the loudspeaker engineer is not familiar with all of its implications. Just what is transient distortion? Can it be measured accurately? What constitutes good or bad transient response? The purpose of this paper is to try to answer some of these questions, and to strive for a clearer understanding of the effect of transient distortion on the listening quality of loudspeakers.

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† Jensen Manufacturing Co., Chicago, Ill.

Definitions

Transient distortion, qualitatively, is the degree to which a device fails to respond accurately to abrupt changes in input-signal level.

A loudspeaker is a device which attempts to convert complex electrical waveforms into their equivalent acoustical signals at the ear of the listener. The objective usually is accomplished, more or less successfully, by conversion of the electrical signal first into mechanical motion; it is in the mechanical moving system that much of the transient distortion is generated. Rather than a simple mechanism containing mass, compliance, and damping resistance in lumped quantities, the complete loudspeaker equivalent mechanical circuit contains an unbelievable complex of small resonant circuits. A glance at the Rocky Mountain frequency-response curve of a typical loudspeaker should confirm the existence of these dozens of small resonators.

If we consider each of these resonant circuits individually, it will be found that the principles of transient analysis as applied to resonant electrical circuits will apply. A more complete discussion is contained in the Appendix, but we may summarize here that, in general, these resonant circuits will require a certain time to reach full amplitude of movement after the input signal is applied, and also that they will continue to oscillate after the signal is turned off. Those resonances with a high Q (small damping) will show the effect to a greater degree than the low Q ones, and it can be shown that any circuit with a Q greater than $\frac{1}{2}$ will oscillate after the stimulus is removed.[1]

The mechanical arrangement of a loudspeaker does not account for all of the peaks in its response curve, however. The acoustical coupling through the air between the loudspeaker cone and the ear contributes its own resonances, damped and undamped. Reflections from the loudspeaker cabinet and the walls of the listening room produce multiple-transmission paths for the sound waves, resulting in selective distortion analogous to radio-signal transmission.

Some frequencies are transmitted more readily than others and if the selective effect is sharp enough, transient ringing or hangover is produced just as though from an underdamped mechanical system.

Before leaving the subject of the loudspeaker frequency response, we should like to mention that these small mechanical resonances are desirable parts of the design. Theoretical frequency response of a loudspeaker with only a simple, one-degree-of-freedom moving system, predicts only a limited frequency range of operation. Secondary cone resonances, properly controlled, are the secret of successful loudspeaker designs.

Methods of Measurement

One method of measuring mechanical transient distortion is to use the back EMF developed by the motion of the voice coil in the magnetic field of the speaker after a signal has been removed. This method is sometimes called the shock test.[2] Although it has an advantage over direct acoustical pickup of the loudspeaker's sound pressure since it eliminates room effects, its chief disadvantage is that the speaker cone is only loosely coupled to the voice coil at the high-frequency range of the loudspeaker's operation. Thus, high-frequency transient waveforms do not appear as electrical signals at the voice coil. Damping of the low-frequency cone motion can be controlled by the internal resistance of the amplifier, but high-frequency transients are unaffected by the amplifier for this reason.

A second method, and the one used in our experiments at Jensen, measures the sound pressure produced by the motion of the cone when an interrupted sine-wave audio signal is fed to the voice coil. A truer picture of the speaker's transient performance is given, even though some acoustical-path resonances may be present. These are reduced as much as possible by making the measurements in a large anechoic chamber with the microphone close to the speaker.

Experiments

To study these transients, photographs were taken using an oscilloscope and a Polaroid Land camera. A sine-wave signal, interrupted by means of a multivibrator keyer, was fed to the voice coil of the speaker through an amplifier having zero internal impedance. The output of the microphone was then fed to the vertical plates of the oscilloscope. With proper synchronization of the 'scope sweep, a continuous display of the transient response of the speaker could be made and photographed. Fig. 1 shows the transient response of a typical loudspeaker sample.

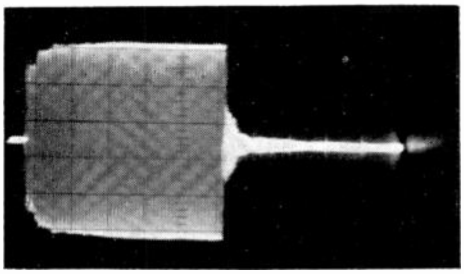


Fig. 1—Typical tone burst of a loudspeaker.

Note that although the starting transient is not outstanding, close inspection of the picture does reveal the starting transient superimposed on the audio signal. The signal lasts for 56 msec, then is abruptly cut off and what follows is the hangover transient of the speaker. This hangover transient was chosen as the quantity to be measured in the determination of the transient response of a speaker. Experimental observation of dozens of loudspeaker transient waveforms showed that a large hangover transient was usually accompanied by a large starting transient. It will be shown later that the hangover transient is the more important to measure, and, fortunately, the easiest.

To obtain a continuous curve of transient distortion vs frequency, the following modification of the method of Shorter[3] was used. The dial on the audio oscillator which produced the basic frequency of the interrupted audio signal was mechanically coupled to the chart drum on a semiautomatic curve tracer. The frequency was slowly changed by a motor. Fig. 2 is a block diagram of the measuring apparatus. A translucent mask having a single, narrow, vertical slit and, at its base, a horizontal slit perpendicular to it, partially obscured the face of the cathode-ray tube on the 'scope (see Fig. 3). At the beginning of each run, the Y position control of the 'scope was adjusted so that the en-

[1] J. F. Novak, "Performance of enclosures for low-resonance high-compliance loudspeakers," IRE Trans. on Audio, vol. AU-7, pp. 5–13; January–February, 1959.

[2] G. A. Briggs, "Sound Reproduction," Warfedale Wireless Works, Bradford Road, England, p. 122; 1953.

[3] D. Shorter, "Loudspeaker transient response," *BBC Quarterly*, vol. 1; October, 1946.

velope of the signal was visible in the horizontal slit. This was done at 400 cycles with the microphone calibrated to a standard reference level.

By adjusting the X-position control on the oscilloscope, the envelope of the transient response could be placed under the vertical slit and the time in milliseconds from the signal cutoff to the vertical slit could be determined. During the run, the envelope amplitude, as observed through the vertical slit, was kept at a certain height by adjusting the signal to the vertical plates with the hand control. This hand control is also mechanically connected to the recording pen, causing it to move and thereby tracing a line on the moving paper graph. Increasing the signal on the vertical plates caused the pen to register a decrease in the measured transient at that point and decreasing the signal caused the pen to register an increase. In other words, as the frequency changes, the signal and transient envelopes constantly change with respect to each other. Adjusting the gain control so that the transient envelope is kept at a constant height in the slit causes the pen to record the resulting curve of the transient output of the loudspeaker.

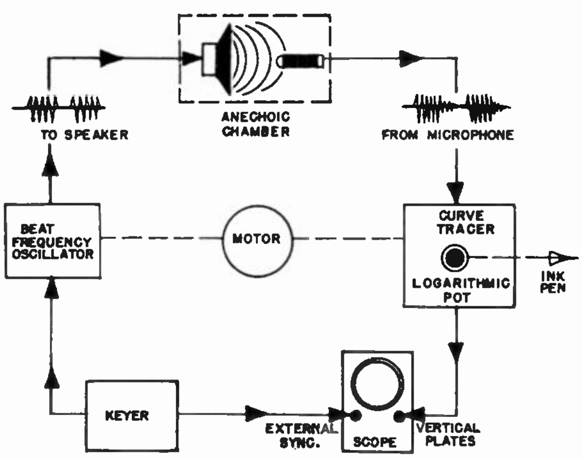


Fig. 2—Diagram of transient measuring apparatus.

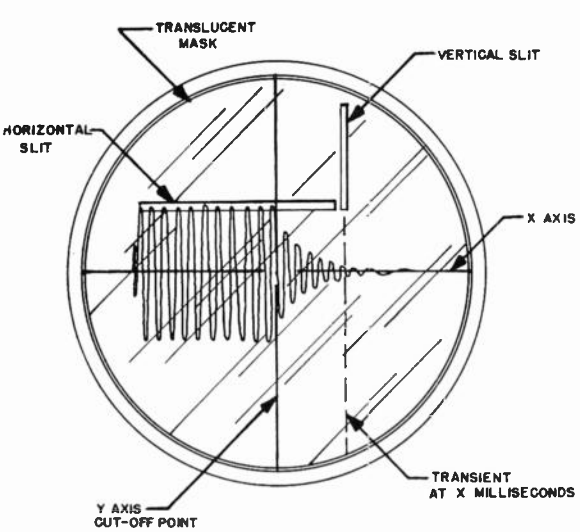


Fig. 3—Tone burst seen through mask.

For comparison purposes, a sound-pressure curve was run in the same manner except that the horizontal slit was used and the height of the signal envelope kept constant in this slit throughout the run.

Fig. 4 is a sample curve of the type made in this transient measurement, and shows the response of a horn-loaded speaker. The curve includes the steady-state, characteristic and the transient response 4 and 10 msec after the end of the sine-wave burst. The horn is designed to cut off around 1000 cycles and it will be noted that there is a large transient below this frequency. Fig. 5 shows the response of this speaker at 510 cycles which is the frequency corresponding to the worst transient response as denoted by the curve. The X axis in the photograph is divided into 10 msec divisions, so that the 4- and 10-msec output can be compared with the response curve at 510 cycles.

The possibility of using noise bursts for transient measurements was explored. It was thought that by using white-noise bursts instead of sine wave, the resultant

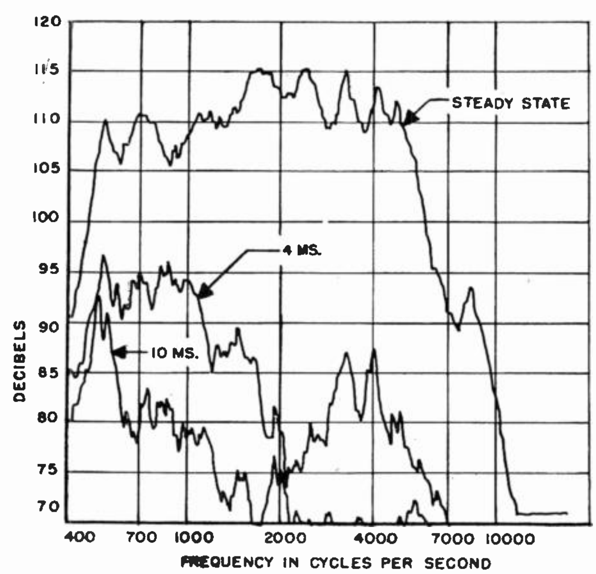


Fig. 4—Loudspeaker response curve with transient curves at 4 and 10 msecs.

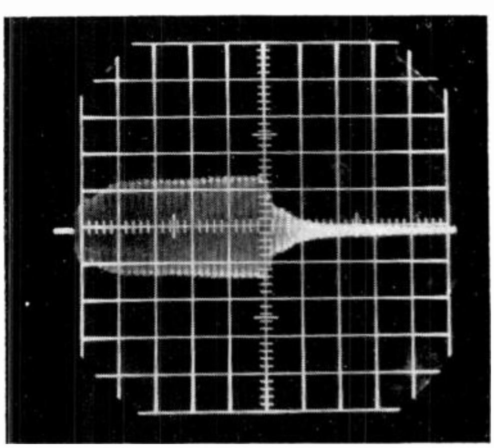


Fig. 5—Speaker response at 510 cycles.

transient output of the speaker would contain the entire audio spectrum in one display. However, when it was tried, it was found that only the relatively large, low resonant frequency transient of the speaker was visible (see Fig. 6).

When loudspeaker samples were found which showed especially pronounced hangover transients at certain frequencies, an additional experiment was performed. The microphone was used as an exploring probe over the surface of the loudspeaker cone. It was found that in many cases the underdamped resonance producing the transient was radiated from localized areas of the cone. The cone edge compliance corrugations were noteworthy offenders in many cases.

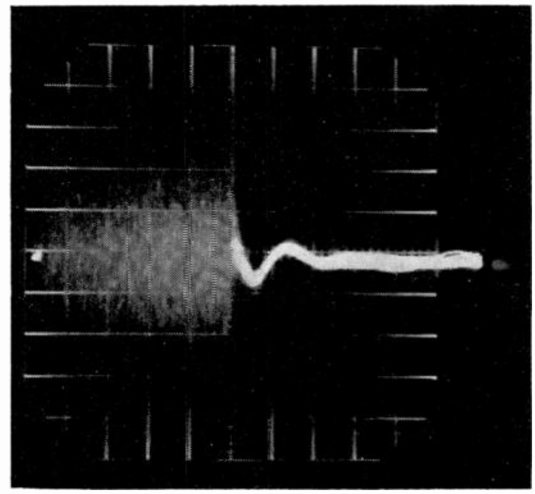


Fig. 6—Speaker response with noise burst input.

Listening Test Results

In the course of these experiments a number of speakers were obtained with large hangover transients. When these speakers were compared with others having relatively low transient distortion on musical listening tests the results were inconclusive. Preferences were expressed for various samples, but apparently for characteristics of the frequency response, there was little correlation with transient performance of the samples.

As mentioned before, loudspeakers which exhibited poor transient response at a certain frequency usually showed some aberration in the steady-state frequency response curve at the same frequency. Where listeners disliked certain loudspeaker samples, it seems likely that their judgments were influenced by the accompanying response shape rather than by the resulting transient response.

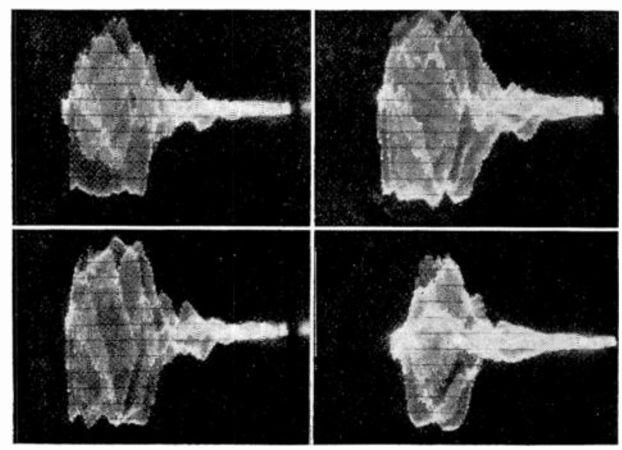


Fig. 7—Four trumpet notes that sound alike.

Schaeffer at the Gravesano Experimental Studios[4] recently published an excellent paper, in which he showed that for certain types of musical signals the shape of the wave envelope was unimportant to the ear. In his experiments, he cut as much as 50 msec off of the beginning of piano notes by means of tape editing, and found that the loss was undetectable, even though it would appear that the start of the wave is the most important part. Also, a recording was made of a trumpet playing a series of eight staccato notes, all of which sounded the same to the ear. However, when an oscillogram was made of these notes, no two of the waveforms resembled each other. Fig. 7 shows four of the trumpet notes.

One possible explanation for these effects may lie in the gestalt theory, which shows that every sound event is analyzed by the ear/brain computer in its entirety rather than in the time sequence. The gestalt theory is based on the principle that physical, psychological, and biological events do not occur through the summation of separate elements or sensations or reflexes, but through formed patterns of these integrated units which function singly or in an interrelationship.[5] The ear localizes the sound event, making a synthesis, and the more familiar it is with the cause and ending of a sound, the more compressed is the synthesis. This would explain the weird effect of playing a recording of a cymbal crash backwards. The ear cannot discern the cause of the sound and is completely baffled by its ending. In this case, the ear seems to analyze the sound as if it were stretched out or less compressed than the forward cymbal crash.

Applying the theory to loudspeaker wave envelopes, we conclude that the ear is more interested in the over-all dynamics of a sound event rather than to irregularities in its waveform, especially when the sound is familiar, such as a piano note. It would appear, therefore, that irregularities in starting and hangover transients are not detected by the ear.

[4] P. Schaeffer, "Note on time relationships," *Gravesano Review*, vol. 5, no. 17, pp. 50–77; 1960.

[5] W. Köhler, "Gestalt Psychology," Liveright Publishing Corp., New York, N. Y.; 1945.

A listening test was performed to evaluate this application of the gestalt theory and to investigate its various effects on the perception of transient distortion. A tape recording was made of a tone burst from a loudspeaker having the longest hangover transient of any of the samples measured. The output of the keyed oscillator was also recorded. Random lengths of tape containing the transientless oscillator pulse, the loudspeaker reproduced pulse, and a reversed section of the loudspeaker output tape on which the transient preceded the pulse, were spliced together. This tape was then played back in a listening room having reverberation time somewhat less than the average living room, and through a loudspeaker known to have no hangover transients at the frequency under consideration. The tape was played back at half the recording speed, thus doubling the hangover transient time. Twelve people, consisting of both trained and untrained listeners, acted as observers. It was found that most of the subjects could not tell the difference between the keyed oscillator pulse and the loudspeaker pulse; however, every subject noticed the backward transient pulse and identified it as having a slur in front and appearing longer in duration than the other pulses.

To find out if this inability to distinguish the clean-break pulse from the transient pulse was due to psychoacoustics or room echoes, the experiment was repeated using earphones. Here the effect was less pronounced and all subjects were able to hear the transient at the end of the loudspeaker pulse. As an additional check, a microphone was placed in the listening room and its output observed on the 'scope while the test tape was running. Surprisingly, a hangover transient from the room appeared on each of the types of pulse and having about the same order of magnitude as the transient of the horrible example loudspeaker (see Fig. 8).

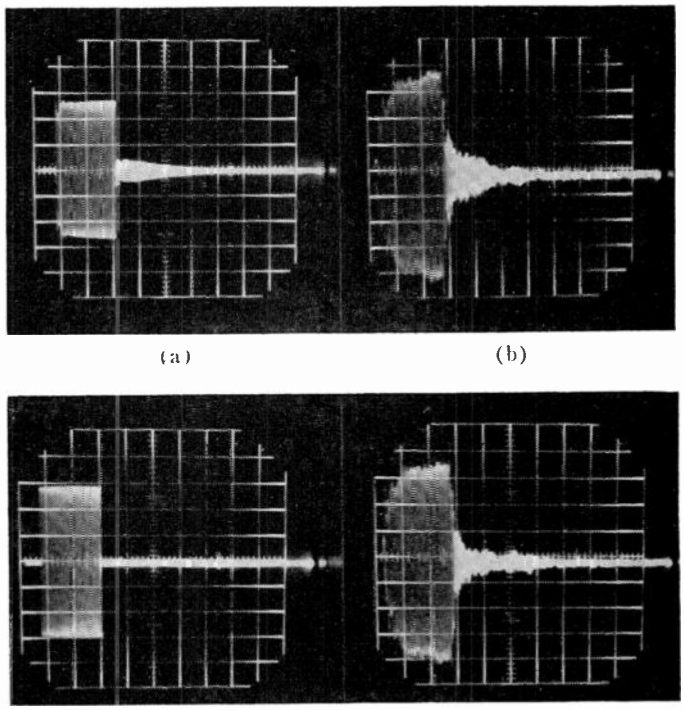


Fig. 8—(a) Loudspeaker pulse on tape. (b) Same pulse in listening room. (c) Transientless oscillator pulse. (d) Same pulse in listening room.

Conclusions

As a result of these experiments, we have reached two conclusions. The first is that the transient performance of a loudspeaker is reflected in its steady-state response curve. Those speakers having the smoothest frequency response curves also exhibited the least hangover and the smallest starting transients. Also, these speakers usually, but not always, sounded better. The second conclusion is that the amount of hangover transient output of most speakers is small in comparison to the echoes of musical listening rooms having even the shortest reverberation times. Even so, the gestalt theory indicates that at least for familiar sounds, such as music, the ear will tolerate gross distortions of the wave envelope of a degree greater than those transients produced by the worst loudspeaker samples.

Even though our experiments lead to the conclusion that no serious correlation exists between transient distortion and listening tests, we do not want to give the impression that transient measurements in loudspeakers should be disregarded as worthless. On the contrary, there is much evidence that transient measurements may be considered valuable for the detection and correction of poor sounding speakers. It seems likely that transient measurements will become a standard laboratory technique in the development of loudspeakers.

Appendix

Whenever a stimulus is applied to a mechanical system, there is a time delay before the system reproduces the applied stimulus. Also, when the stimulus is abruptly removed, the system returns to its rest position after a certain time has elapsed. What happens to the system during this time delay in the build up and decay of the reproduced stimulus in a mechanical system can be referred to as the transient or transitory response of the system. In loudspeakers, the stimulus would be an ac audio signal. The amount of time delay along with motion of the diaphragm in the build up and decay in the reproduction of the audio signal at different frequencies would determine the transient distortion of the speaker.

Intuitively it can be reasoned that the mass of the speaker diaphragm, the stiffness of the diaphragm and annulus, and the total mechanical resistance of the system will affect the transient response of the speaker.

If we assume that the diaphragm of a speaker is displaced from its equilibrium position by a dc voltage applied to its voice coil and then allowed to move freely, its displacement at any time t is[6]

$$d = A_0 e^{-rt/2m} \sin (t\sqrt{s/m - r^2/4m^2} + \theta), \qquad (1)$$

where

A_0 = the maximum amplitude of the diaphragm from its equilibrium position.
r = total mechanical resistance,
s = the stiffness of the annulus and diaphragm,
m = the mass of the system,
θ = phase angle,

which can be resolved to

$$d = A_0 e^{-kt} \sin (t\sqrt{s/m - k^2} + \theta),$$

$$= A_0 e^{-kt} \sin (\omega' t + \theta)$$

where

$$\omega' = \sqrt{s/m - r^2/4m^2}$$

$$k = \text{damping constant} = \frac{r}{2m}.$$

Since the sine term of (1) alternates between +1 and −1, maximum displacement occurs periodically. The amplitudes of the sinusoid are given by

$$A = \pm A_0 e^{-kt},$$

and the curve of the damped vibrations lies between the envelope

$$A = \pm A_0 e^{-kt}.$$

If we take the ratio of successive amplitudes where any given amplitude will be represented by

$$A_n = A_0 e^{-kt},$$

the succeeding amplitude is

$$A_{n+1} = A_0 e^{-k(t+2\pi/\omega')}$$

$$= A_0 e^{-kt} e^{-k(2\pi/\omega')},$$

then

$$\frac{A_n}{A_{n+1}} = e^{k(2\pi/\omega')} = e^{r\pi/m\omega'}$$

$$\Delta = \log_e \frac{A_n}{A_{n+1}} = 2.3 \log_{10} \frac{A_n}{A_{n+1}} = \frac{\pi r}{m\omega'}$$

$$\Delta = \frac{\pi r}{\sqrt{s/m - r^2/4m^2}} = \frac{2\pi r}{\sqrt{4ms - r^2}}. \qquad (2)$$

[6] M. Rettinger, "Practical Electroacoustics," Chemical Publishing Co., Inc., New York, N. Y.; 1955.

The quantity Δ is known as the logarithmic decrement. When

$$\frac{s}{m} > \frac{r^2}{4m^2},$$

the motion of the moving diaphragm is a damped sinusoidal curve (Fig. 9). When

$$\frac{s}{m} = \frac{r^2}{4m^2},$$

critical damping occurs and the motion is no longer periodic. When

$$\frac{s}{m} < \frac{r^2}{4m^2},$$

the value of the expression under the square-root sine of (2) becomes negative. The physical interpretation of this imaginary quantity is that motion is "aperiodic" or overdamped. Fig. 10 shows the curves resulting when a

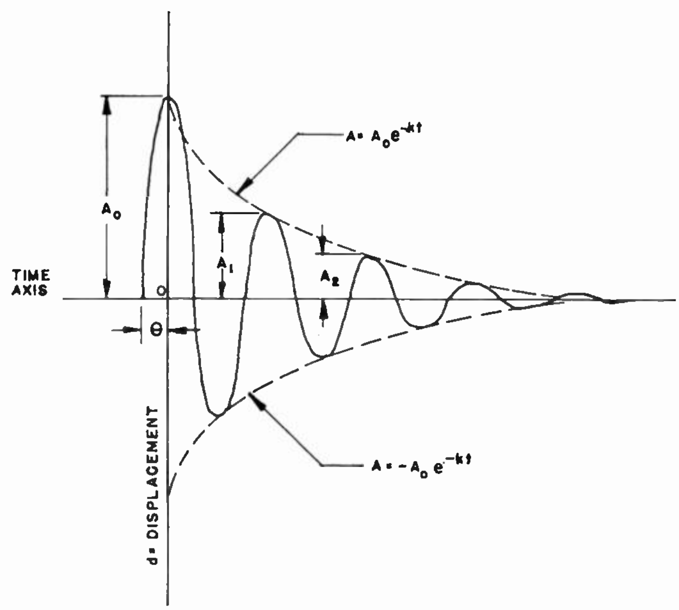


Fig. 9—Damped sinusoidal motion.

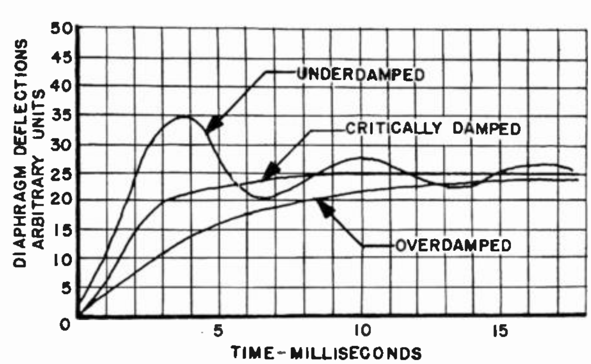


Fig. 10—Underdamped, critically damped, and overdamped motion.

dc voltage of 25 units (arbitrary) placed on the voice coil of the diaphragm under the three conditions of underdamped, critically damped, and overdamped.[7]

When a periodic force caused by an audio ac signal applied to a voice coil vibrates the diaphragm of a loudspeaker, three effects are produced. First, on the initial excitation a form of oscillation occurs which almost instantly dies away in the manner shown by the curve of Fig. 9. This oscillation is sometimes called a "starting transient" and is a damped free vibration which is superimposed on the forced vibration resulting from the ac signal. After this starting transient has disappeared and the applied signal is still active, there is a sustained vibration of the diaphragm or steady-state response equal to the frequency of the applied ac signal. Finally, when the signal is removed, "hangover transients" occur which have the same frequencies as the natural frequencies of the speaker, and decay exponentially as shown in Fig. 11. In order to prevent these transients from interfering with the (transient) signal, the damping of the loudspeaker must be large. If it is small and the frequency of the signal is close to a natural frequency of the speaker, "beats" or "booming" of the speaker are likely to occur.

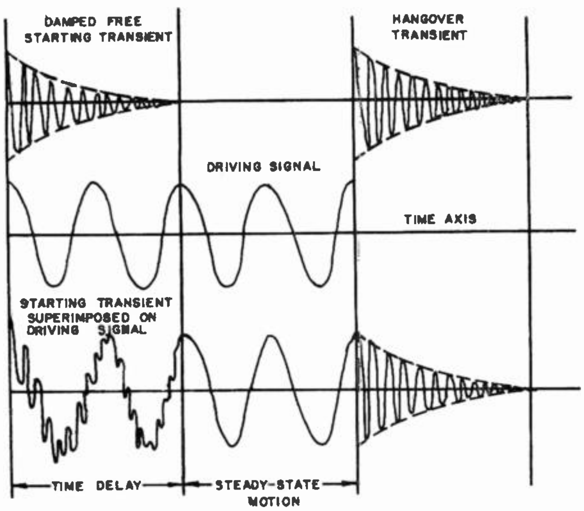


Fig. 11—Motion of loudspeaker diaphragm.

[7] *Ibid.*, p. 167.

It is important to stress that transient distortion does not occur only at one or two natural frequencies of a speaker. On the contrary, it exists along the entire range of the audio spectrum. An examination of the response curve of a loudspeaker shows many natural frequencies in the audible range, and the transitory response may include, at first, vibration at each of these frequencies then spreading out to the rest of the spectrum with a rate of decay different for each frequency and generally slowest at the lowest natural frequency.

Eq. (1) gives us the transient vibration of the diaphragm. The complete solution, however, would have to include the steady-state vibration which persists as long as the driving signal is applied to the voice coil. In the case of lightly damped systems

$$\frac{r}{2m} < n,$$

the complete solution is[8]

$$A = A_0 e^{-kt} \sin (n_1 t + \theta) + \frac{F_0}{\omega Z} \sin (\omega t - \phi), \qquad (3)$$

where $n_1 = \omega'$.

The first term of the equation or the transient vibration which arises from the initial application of the driving signal rapidly dies down because of the exponential damping factor as illustrated in Fig. 9. The second term is the steady state which persists as long as the signal is applied. When the signal is removed, the steady-state term at once disappears and the motion is converted into free vibrations with a frequency of

$$\frac{n_1}{2\pi},$$

or the natural frequency of the speaker. In the early stage of the vibration, which follows the application of the driving force, the transient and steady-state motions coexist and if the two frequencies n_1 and ω differ only slightly, the transient distortion will be large due to the superimposition of the transient and steady-state vibrations.

[8] A. H. Davis, "Modern Acoustics," G. Bell and Sons, Ltd., London, Eng., p. 11; 1934.

A Low-Noise Microphone Preamplifier*

ALEXANDER B. BERESKIN†, FELLOW, IRE

Summary—**This paper describes a low-noise two-transistor preamplifier which has been developed for use with microphones. For a source resistance of 1000 Ω, a noise figure of 1.3 db has been achieved. A corresponding middle-frequency gain of 40 db, bandwidth of 30 kc and output impedance of 175 Ω resulted.**

IN amplifying the low-level signals produced by ribbon and dynamic microphones, it is important to provide good linearity and frequency response and low distortion. It is equally important that random noise, microphonics, and hum pickup be kept at a minimum.

Excellent microphone preamplifier circuits using vacuum tubes have been developed. In these circuits considerable attention must be devoted to the filament supply and tendency towards microphonics in the low-level stages. Selected types of tubes and selected tubes of a type are frequently used to insure the best results. An equally important problem is the random noise contributed by the low-level stages and the source-impedance level necessary to maximize the SNR of the system. An impedance level of about 100,000 Ω, as seen by the grid circuit of the first stage, appears to result in the optimum SNR. At these impedance levels it is difficult to shield the system against power-frequency signal pickup. If the microphone is located at an appreciable distance from the preamplifier, it is first necessary to transform the signal from the microphone impedance to approximately a 250-Ω level, to transmit the signal over the required distance, and then to transform it to the 100,000-Ω level required at the grid of the first amplifier. An alternative method is to locate the first amplifier in the microphone barrel or stand and to transform the signal directly to the 100,000-Ω level. The disadvantage of this process is that anode and filament supplies must be brought in through the microphone cable. These disadvantages are not insurmountable since very excellent microphone preamplifier systems have been developed.

The purpose of this investigation was to examine some of the problems involved in the use of relatively low-cost commercially available transistors for the design of a low-noise microphone preamplifier. Only room temperatures were to be considered and it was desired to have a circuit that would fulfill the requirements and still be simple enough to be incorporated in the barrel of a microphone. Since microphonics would be absent and hum pickup could be eliminated by adequate shielding, it was expected that impedance levels could be chosen for optimum noise figures.

The basic circuit developed is shown in Fig. 1. This circuit uses two low-noise RCA 2N1010 *n-p-n* transistors. The two transistors used were selected on the basis of low-noise figure from a group of four units purchased. The first stage is a regular common-emitter stage with a 10-K collector load resistance and direct coupling to the second stage which is a common-collector stage with a 10-K emitter resistor. A suitable bias and feedback resistor is connected between the second emitter and the first stage. Neglecting the power supply and the input and output coupling capacitors, the circuit uses two transistors and three resistors. A 5-μf electrolytic capacitor is used to couple the input signal to the first base and a 25-μf electrolytic capacitor is used to couple the output signal to the output terminal. The power supply[1] shown is one developed by the author for use in low-level transistor amplifiers. In addition to being transformerless, it is compatible with almost any other equipment in use since accidental grounding of either ac line does not change the dc-output voltage of this power supply. Of course, any other suitable source of dc voltage would work equally well.

The amplifier was tested for voltage gain and frequency response by connecting a low-impedance source in series with various resistors to the input terminals of the amplifier. The "voltage gain" in this case is defined as the ratio of the amplifier output voltage to the source voltage. The variation of voltage gain and the frequency at which the response is 3 db below the middle-frequency gain is shown as a function of the source resistance in Fig. 2. In particular, it should be noted here, that a source resistance of 1000 Ω yields a −3-db frequency response of 30 kc and a voltage gain of 90 (39 db). The variation in voltage gain corresponds to an amplifier input resistance of 2000 Ω.

With suitable shielding the hum level was made negligibly small compared to the random-noise voltage in the amplifier. This noise voltage was measured in three different manners:

1) A X100 amplifier[2] was interposed between the

* Received by the PGA, March 20, 1961.
† Elec. Engrg. Dept., University of Cincinnati, Ohio.

[1] A. B. Bereskin, "A transistorized stereo preamplifier and tone control for magnetic cartrides," *Proc. Natl. Electronics Conf.*, vol. 15, pp. 111–118; 1959. Also, IRE TRANS. ON AUDIO, vol. AU-8, pp. 17–20; January–February, 1960.

[2] A. B. Bereskin, "A transistorized decade amplifier for low level audio frequency applications," *Proc. Natl. Electronics Conf.*, vol. 13, pp. 189–197; Also, IRE TRANS. ON AUDIO, vol. AU-5, pp. 138–142; September–October, 1957.

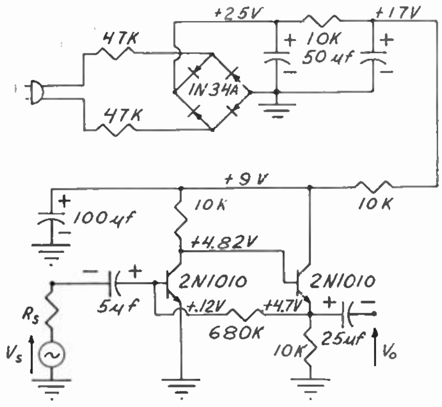


Fig. 1—Microphone preamplifier circuit.

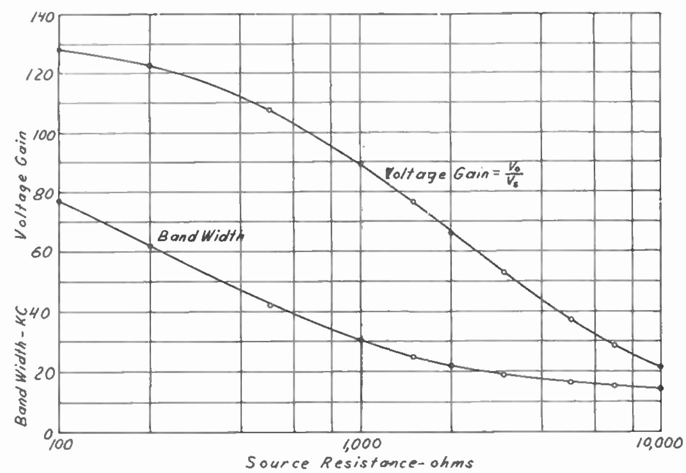


Fig. 2—Gain and bandwidth characteristics.

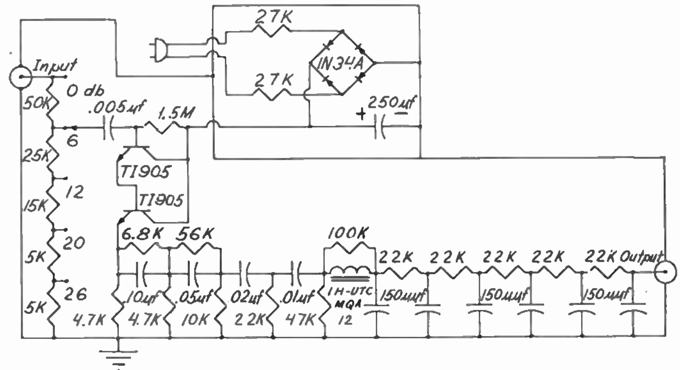


Fig. 3—Filter circuit used in case C measurements.

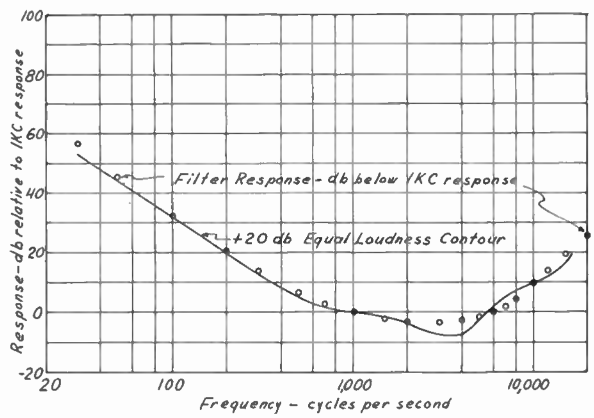


Fig. 4—Response characteristic for the filter of Fig. 3.

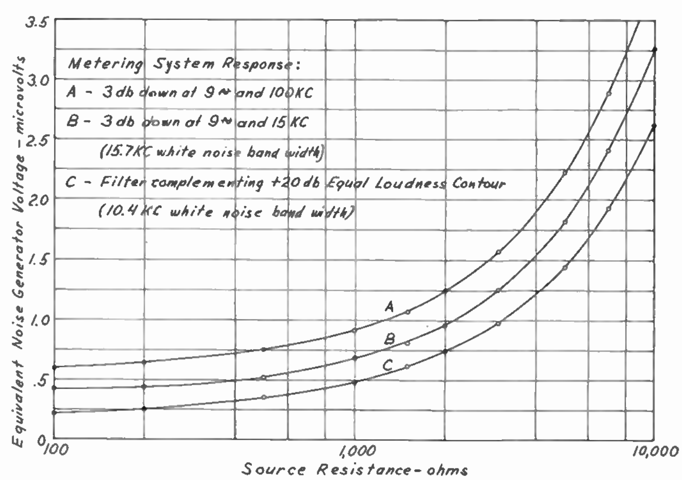


Fig. 5—Equivalent source-noise voltage.

output of the microphone amplifier and the input of a Ballantine Model 320 true RMS-VTVM. The metering system in this case was down 3 db at 9 cycles and 100 kc.

2) Same as 1) except that a low-pass filter was interposed between the X100 amplifier and the VTVM. In this case, the metering circuit was down 3 db at 9 cycles and 15 kc and the equivalent white-noise bandwidth was 15.7 kc.

3) Same as 1) except that a filter complementing the +20-db Fletcher-Munson equal-loudness contour was interposed between the X100 amplifier and the VTVM. The circuit diagram for this filter is shown in Fig. 3 and the filter response along with the +20-db equal-loudness contour is shown in Fig. 4. The white-noise bandwidth for this system was 10.4 kc referred to the 1-kc response.

The equivalent source-noise voltage obtained by these three methods is shown in Fig. 5 as a function of the source resistance used. In general the noise voltage is seen to increase both with bandwidth and with source resistance.

The noise figure of the amplifier, referred to the open-circuit thermal noise of the source resistance, is shown for cases B and C in Fig. 6. The curve for case B is representative of the performance that might be measured in most cases while the curve for case C is more representative of what would be heard by a human audience since it takes into account the frequency variation of human hearing.

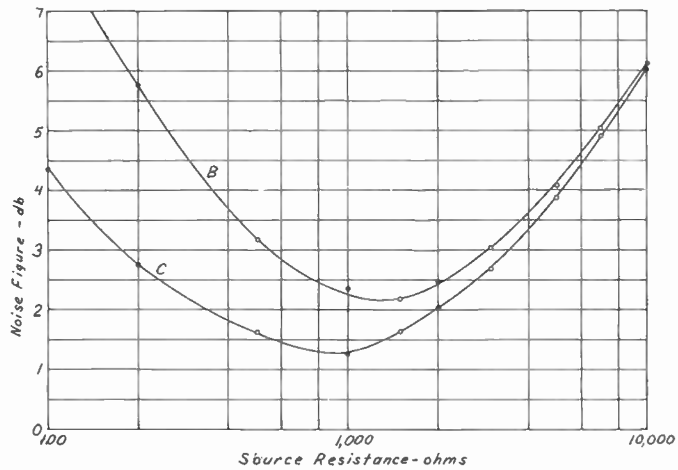


Fig. 6—Noise-figure characteristics.

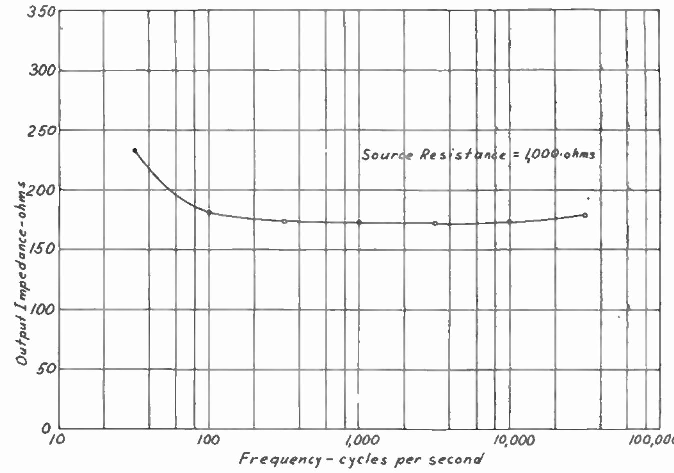


Fig. 7—Output-impedance characteristic.

For case B the curve bottoms at about 1300 Ω while for case C the curve bottoms at about 900 Ω. A nominal value of 1000 Ω would come close to providing minimum noise figure in both cases. Values of source resistance close to 1000 Ω have also been reported by other investigators.[3] The noise figure for case C does not deteriorate quite as rapidly at lower values of source resistance as that for case B but, in any case, a 150-Ω source resistance would have a noise figure 2 db poorer than the optimum value. For case B this would be 4 db poorer than the optimum value.

The circuit was designed to have low output impedance to facilitate its connection to other equipment or for transmission of the signal over appreciable distances. The curve in Fig. 7 shows that for a source resistance of 1000 Ω the output impedance is approximately 175 Ω over most of the audible frequency range.

As mentioned previously, it was not intended that this amplifier be operated at temperature extremes. It was, however, considered important to determine just what the permissible temperature range would be. The upper curve in Fig. 8 shows the variation in response as a function of temperature for the amplifier of Fig. 1. In this curve 0 db represents a voltage gain of 100. Satisfactory operation was obtained up to 40°C. For temperatures in excess of 40°C the performance deteriorated rapidly. A slight increase in the upper temperature limit of operation is obtained by connecting a suitable thermistor between the base of the first transistor and ground. The lower curve in Fig. 8 shows that the temperature limit of operation has been increased to about 48°C by using a thermistor having 16-K Ω resistance at 25°C and a −4.4 per cent/°C-temperature coefficient of resistance. Additional improvement in operation at high temperatures could be obtained by more sophisticated stabilization techniques.

[3] J. J. Davidson, "Low noise transistor microphone amplifier," 1957 IRE NATIONAL CONVENTION RECORD, pt. 7, pp. 162–168.

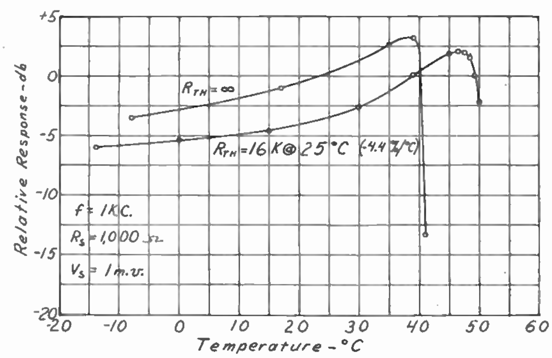


Fig. 8—Temperature-response characteristic.

CONCLUSION

At the present time, dynamic and ribbon microphones are available with either 50,000-Ω or less than 250-Ω impedance levels. For the amplifier described in this paper, at a source-impedance level of 150 Ω, the noise figure has deteriorated 2 to 4 db depending on the type of filter used in the noise measurement. A substantial improvement in performance would result if the microphone manufacturers were to supply the microphones with 900- to 1100-Ω impedance levels for use with transistor preamplifiers. At this impedance level the signal could be carried over a substantial distance to a remote preamplifier without picking up appreciable power-frequency hum. It would also be possible to locate this extremely small preamplifier in the barrel or stand of the microphone, to raise the signal level by approximately 40 db and then to transmit it over an appreciable distance at a 175-Ω level. The microphone cable would need to supply only 1 ma at 9-v dc for the operation of the preamplifier.

It is shown in this paper that the performance of the transistor microphone preamplifier is as good as the better vacuum-tube units, if proper source-impedance levels are maintained, and that other substantial advantages are present.

Cathode Followers and Feedback Amplifiers with High-Capacitance Loads*

THOMAS L. GREENWOOD†, SENIOR MEMBER, IRE

Summary—Investigation, both theoretically and experimentally, was made into the mode of operation of cathode followers and feedback amplifiers with high-capacitance loads. The capacitance loaded cathode follower compares unfavorably with the resistance loaded cathode follower, particularly regarding operation with fluctuating input signals. Input overdrive of a capacitance loaded cathode follower causes production of transient voltages in the output circuit. Threshold of input overdrive is considerably lower than in resistance loaded cathode followers, especially at high audio frequencies. Also, harmonic distortion is increased at high frequencies and high-frequency response drops off. Experimental analysis of the voltage relations between the input and output circuits was made, and guides for design were developed. It was found that a symmetrical E_g-I_p curve is the most important factor in development of a cathode follower for driving high-capacitance loads where transient distortion is to be avoided.

THE cathode follower is a useful device for coupling a high impedance source to a lower impedance load. Its properties are due to feedback, which operates in a near-ideal manner if the load is a pure resistance with a relatively shunt reactance. However, where the load includes a relatively low shunt reactance, the feedback conditions are modified and the device no longer behaves as desired.[1–3]

This discussion is concerned with the problem of maintaining high cable driving capability of cathode followers, with low distortion of fluctuating input signals, over the audio frequency range.

The limit of input signal voltage for low distortion under steady-state conditions is governed by plate-current cutoff and grid-current threshold points. However, sudden changes of the input signal amplitude often causes a low-frequency transient signal in the output circuit, the duration of which is determined by the RC time constant of the output circuit. This effect usually occurs at a much lower input voltage than is necessary to reach plate current cutoff or grid current threshold under steady-state conditions. Capacitance in the load magnifies these effects.

* Received by the PGA, February 27, 1961.

† George C. Marshall Space Flight Center, NASA, Huntsville, Ala.

[1] F. Langford-Smith, "Radiotron Designers Handbook," published by the Wireless Press for Amalgamated Wireless Valve Co. Pty. Ltd., Sydney, Austral.; reproduced and distributed by RCA Victor Div., Harrison, N. J.

[2] W. T. Cocking, "Cathode follower dangers," *Wireless World*, vol. 52, pp. 79–82; March, 1946.

[3] H. Goldberg, "Internal impedance of cathode followers," PROC. IRE, vol. 33, pp. 778–782; November, 1945.

RESISTANCE LOADED CATHODE FOLLOWER

Fig. 1 shows the circuit of a triode vacuum tube operated as a grounded plate amplifier, or cathode follower, with its load impedance in the cathode side of the circuit. Circuit parameters are suitable for operation in the audio frequency range. E_0 is taken from the cathode through a coupling capacitor if it is desired to block the dc from the load. The grid is biased a few volts negative relative to the cathode, and R_g is usually in the order of megohms depending upon the input impedance desired. When a sinusoidal input signal is applied, negative feedback occurs in the grid-cathode circuit, so that E_0 is slightly lower peak-to-peak amplitude than E_{in}. The difference in the vector quantities E_{in} and E_0 is applied between grid and cathode. With a resistance load, there is no phase shift between E_{in} and E_0, therefore E_{gk} is the

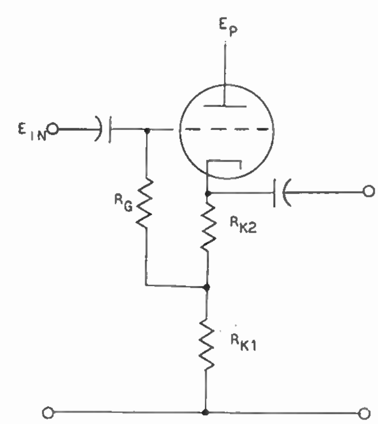


Fig. 1—Cathode follower schematic circuit.

algebraic difference between E_{in} and E_0. To illustrate: $R_g = 10$ megohms, $E_{in} = 10$ v, $E_0 = 9$ v, $E_{gk} = 1$ v. Therefore, the current in $R_g = 10^{-7}$ a. This current is the result of applyiny E_{in} of 10 v, and therefore the source of E_{in} sees a resistance of 10 v/10^{-7} a $= 100$ megohms instead of 10 megohms. The feedback, in effect, multiplies the R_g value in proportion to the effective amplification factor of the vacuum tube. The exact expression is

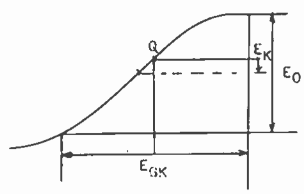
$$R_{in} = \frac{R_g + R_{k1}}{1 - k}, \quad \text{where} \quad k = E_0/E_{in}.$$

The output impedance of the cathode follower is reduced by positive feedback to a fraction of the R_p

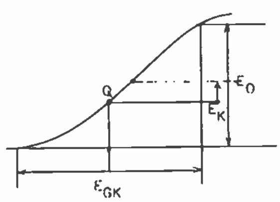
value. The approximate expression is $R_0 = R_p/(\mu+1)$. In applications where it is desired to attain lowest R_0, it is necessary to operate the vacuum tube so that R_p is low and μ is high. This condition requires fairly low, grid bias, low plate voltage, and high plate current.

Cathode followers are often designed without consideration of transient characteristics, the main objective apparently being highest gain. This means operation too near the upper end of the E_g-I_p curve, which usually results in the average I_p shifting downward upon application of an input signal with gradually increasing amplitude. E_k is proportional to I_p and the term E_k will be used hereafter in this discussion to denote instantaneous cathode voltage.

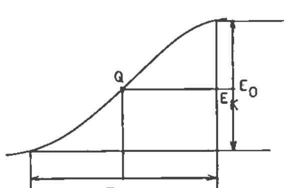
This condition is illustrated in Fig. 2(a), where the grid bias is set at Q, and symmetrical sine waves applied. Positive peaks cause grid current and flattening of output positive voltage peaks, while negative-peak excursions remain on the linear portion of the E_g-I_p curve. The results are that the average cathode voltage is shifted to a lower value, and a transient is produced.



(a)



(b)



(c)

Fig. 2—Grid bias conditions.

If the grid bias is set too high, as at Q in Fig. 2(b), and the same input amplitude applied as in Fig. 2(a), then the negative peaks would extend into the nonlinear portion of the E_g-I_p curve, while positive peaks would extend only into the linear upper portion of the curve. This would result in the average cathode voltage being shifted to a higher value, and a transient is again produced.

If the grid bias is set at Q in Fig. 2(c), and the same amplitude input applied as in Fig. 2(a) and 2(b), then both positive and negative peaks would be flattened in the same degree, but by a lesser amount than in either Fig. 2(a) or 2(b), and the average cathode voltage would remain the same as its no-signal value. A higher amplitude input wave could possibly be applied before the average cathode voltage shift becomes intolerable.

Fig. 3 shows comparative tests with cathode followers having the characteristics described above. The top oscillograph trace shows conditions illustrated in Fig. 2(a), and the middle trace shows conditions of Fig. 2(c).

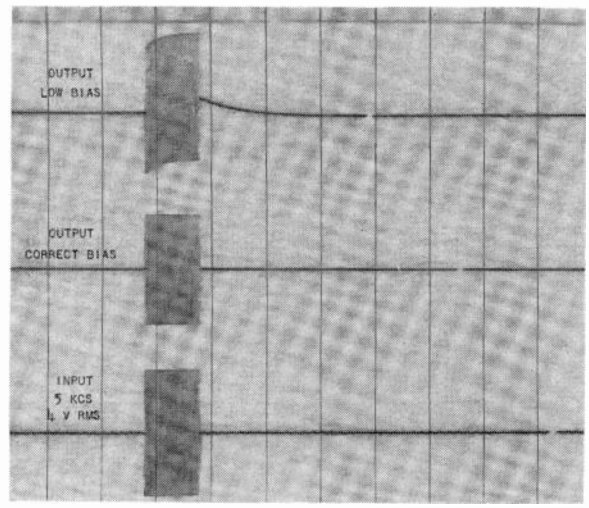


Fig. 3—Oscillographic recording of comparative tests.

The input signal is a 50 cycle wave train gated on and off at zero crossings, applied to the two cathode followers in parallel, identical except for grid bias values. The maximum input amplitude for transient-free operation of the "low bias" circuit was 2 volts rms, while the maximum for the "correct bias" was 8 volts rms. Distortion of steady-state sine-wave inputs of the same amplitude was the same for both bias conditions.

Capacitance Loaded Cathode Follower

When capacitance is added in shunt with the resistance load of a well-designed cathode follower, that is, one with good transient response, such as Fig. 2(c), and a fixed frequency input voltage is increased upward from zero, the average cathode voltage starts shifting upward at a critical input amplitude. This is due to a number of factors which will be discussed in turn.

A capacitive reactance load causes a phase shift between input and output, so that the familiar E_g-I_p curve becomes an ellipse.[4]

When the input and output voltages of a capacitance plus resistance loaded cathode follower are applied to the X and Y axes of a CRT oscilloscope, the familiar ellipti-

[4] M. von Ardenne, "On the theory of power amplification," Proc. IRE, vol. 16, pp. 193–207; February, 1928.

cal figure will be displayed. The output voltage with no input would appear as a spot in the center of the screen. Increasing input amplitude would cause a small ellipse to appear surrounding the center of the screen. The major axis would nearly coincide with the E_g-I_p curve previously obtained with a resistance load. Further increase of input amplitude would cause the ellipse to grow in size, until the extreme ends extend into nonlinear portions of the E_g-I_p curve. At the same time the major axis would shift away from the previous E_g-I_p curve, becoming steeper. With most tube types, the top part of the ellipse becomes pointed. This characteristic causes distortion of the wave at the cathode, making the positive side peaked and the negative side flattened. Thus, harmonic distortion is generated, and the average charge on the output coupling capacitor increases to a value where the area under the positive and negative voltage curves are equal. This is shown in Fig. 4. Average cathode shift in such a capacitance-plus-resistance loaded cathode follower occurs at a lower input voltage than in a resistance loaded cathode follower. This effect is similar to the "low-bias" condition shown in Fig. 2(a). Further increase of input amplitude would cause the average cathode voltage to reach a maximum, then shift downward to a value below the no-signal value.

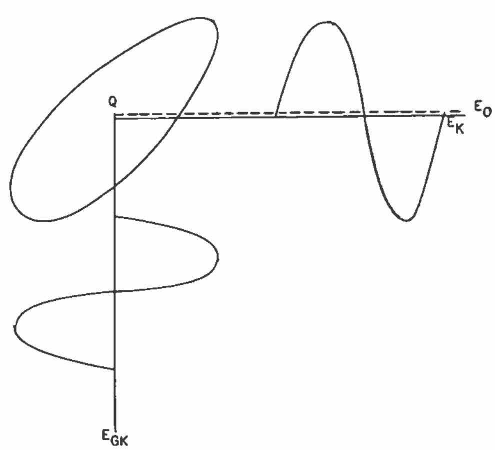


Fig. 4—Vacuum tube characteristic with reactive load.

Increase in E_{gk} bias would cause the average cathode voltage to be shifted upward to a higher value before starting to shift downward with increasing input amplitude. Decrease in E_{gk} bias would cause only a downward shift with increasing input amplitude. The bias finally selected in a practical design would be a compromise depending upon maximum input overdrive voltage desired, and the value of capacitance in the load. Overdrive conditions begin gradually and it is necessary to select an arbitrary value of average cathode voltage variation to be tolerated. In this discussion the allowable variation of average cathode voltage is ±2 per cent. The input voltage which results in more than ±2 per cent variation is called the overdrive voltage.

Another effect of the shunt capacitance load is the effect on high-frequency response above the middle range, according to the formula[5]

$$\frac{E_0}{E_{in}} = \frac{1}{\sqrt{1 + (R_0\omega C)^2}}.$$

This formula is not applicable at low frequencies where the response drops off due to high reactance of coupling capacitors. At high frequencies, the E_0/E_{in} relation would be modified so that $E_{in} - E_0 = E_{gk}$ would be increased for a given input voltage. This E_{gk}, which is the effective grid driving voltage, would have to be decreased by reducing E_{in}, to prevent instantaneous cathode voltage extending into nonlinear portions of the E_g-I_p curve and causing shift of the average cathode voltage.

A third effect caused by capacitance loading is the variation in E_{gk}, due to phase shift between E_{in} and E_0, which is proportional to frequency, according to the formula

$$E_{gk} = E_{in} \sin \phi,$$

where ϕ is the phase difference between E_{in} and E_0. Thus, E_{gk} is increased at high frequencies, for a given E_{in}, and requires that E_{in} be reduced to prevent the instantaneous cathode voltage from extending into nonlinear portions of the E_g-I_p curve with consequent variation of the average cathode voltage. This is illustrated in Fig. 5 which shows E_{in}-E_0 relations at several frequencies with E_{in} at threshold of overdrive input voltage.

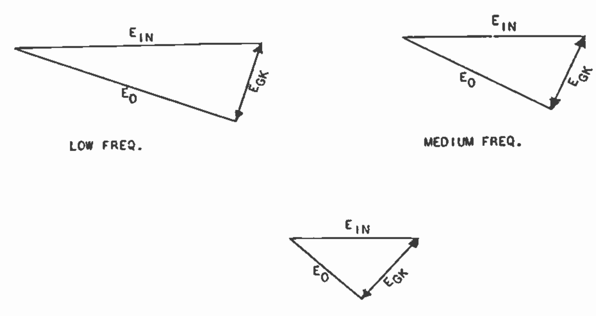


Fig. 5—Input-output voltage relations.

A fourth effect is the nullification of distortion reduction properties of the cathode follower. This is caused by failure of the negative feedback to operate in the normal manner. With full feedback in operation, the E_g-I_p curve is linearized to some extent and harmonic distortion is reduced. With capacitance loading, at high frequencies the distortion was found to be equal to that obtained in a class A power amplifier using the same circuit parameters.

[5] V. C. Rideout, "Active Networks," Prentice-Hall, Inc., New York, N. Y., p. 82; 1954.

Because all of the effects mentioned above cause the difference between E_{in} and E_0 to be increased at high frequencies, thus increasing E_{gk}, the over-all effect is to require that E_{in} be reduced in order to prevent the instantaneous cathode voltage from extending into nonlinear portions of the E_g-I_p curve, with consequent average cathode voltage shift. The ratio of $E_{\text{overdrive}}$ for capacitance loaded cathode followers (E_0), to $E_{\text{overdrive}}$ for resistance loaded cathode followers (E_r), may be expressed approximately as follows:

$$\frac{E_c}{E_r} = \frac{1}{\sqrt{1 + \left[\left(\frac{R_0}{5}\right)^2 \omega C\right]^2}}.$$

Multistage feedback amplifiers often include a cathode follower in the input or output circuit, or both. Effects of capacitance loads on such amplifiers are, in general, the same as for single cathode followers. Overdrive input voltages are inversely proportional to the gain employed and to the amount of capacitance loading. Fig. 6 shows test results on a three-stage amplifier, having an input cathode follower to provide high input impedance, a pentode intermediate amplifier, and an output cathode follower to provide low-source impedance, with feedback between the two cathode followers.

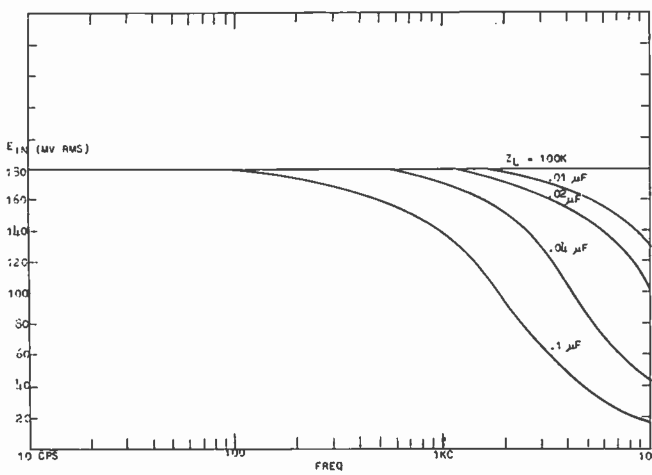


Fig. 6—Input overdrive characteristic of feedback amplifier.

Developmental Work

The demand for higher overdrive capabilities in cathode followers for high capacitance cable loads resulted in experimental designs for this service. Experiment has shown the desirability of low R_0 in a cathode follower for driving a capacitance load. The source impedance of a cathode follower consists of the parallel combination of $(R_{k1}+R_{k2})$ and $R_p/(\mu+1)$, but due to $(R_{k1}+R_{k2})$ being much higher than $R_p/(\mu+1)$, the expression $R_0=R_p/(\mu+1)$ may be used without serious error. Therefore, a vacuum tube with low R_p and high-amplification factor would be the most desirable type for a cathode follower. However, it was found that in considering different tube types progressing from low μ to high μ, the amplification factor increased at a lesser rate than the plate resistance, so that the tube types with lowest plate resistance also had the lowest R_0 when used in a cathode follower circuit. Another factor to be considered is that the high μ types, with high plate resistance, require higher plate supply voltage than low μ, low plate-resistance types. Two common-tube types, 12AU7 and 6BX7, were selected for the above reasons. Circuit component values were selected according to Class A power amplifier design practice. That is, plate circuit load resistance at least twice R_p and grid bias approximately half the cutoff value. Bias is derived from the voltage drop across R_{k2} which was made variable to compensate for slight differences in tube characteristics. It was found that the ratio R_p/R_{k1} could be varied over a range of 1 to 0.2 without significant change in operating characteristics, provided that the grid bias was adjusted for low E_k shift for each value of R_{k1}. Test results are shown in Figs. 7–10. Fig. 7 shows the overdrive characteristics of the cathode follower using a type 12AU7 tube with two sections in parallel. It may be seen that the limit of input voltage without transients being produced is about 14.5 volts rms for the resistance loaded condition, at all frequencies. For the 100,000 Ω plus 0.04-µf load, the performance is seriously degraded at frequencies above about 500 cps. For the 100,000 Ω plus 0.1-µf load, the performance is very much degraded above 100 cps. Fig. 8 shows relative frequency response for different load conditions. $R_0=215$ Ω for this arrangement.

Fig. 9 shows test results on a cathode follower using a 6BX7 type with two sections in parallel. The limit of input voltage is about 22.5 volts rms for the pure resistance load condition. With 100,000 Ω plus 0.04-µf load, the limit begins to drop off above 1000 cps. Fig. 10 shows the relative frequency response for different load conditions. $R_0=120$ Ω. It may be seen that the 6BX7 model is superior to the 12AU7 model, as far as transient-free cable driving capability is concerned. Nonlinearity measurements showed a maximum value of $1\frac{1}{2}$ per cent at high frequencies for both models. This is due mainly to odd-order harmonics and a small amount of even-order harmonics which were not cancelled because of reduced feedback at high frequencies. It may be seen that the figure of merit, high-overdrive threshold, is inversely proportional to R_0. The absolute value of input voltage at overdrive threshold is a function of the E_g-I_p curve linearity of each tube type, and is not readily subject to exact analysis. R_0 was measured at 1000 cps with an ac bridge with the input circuit of the cathode follower terminated by the same capacitance as in normal working conditions. The average E_k was measured with a high-resistance dc voltmeter.

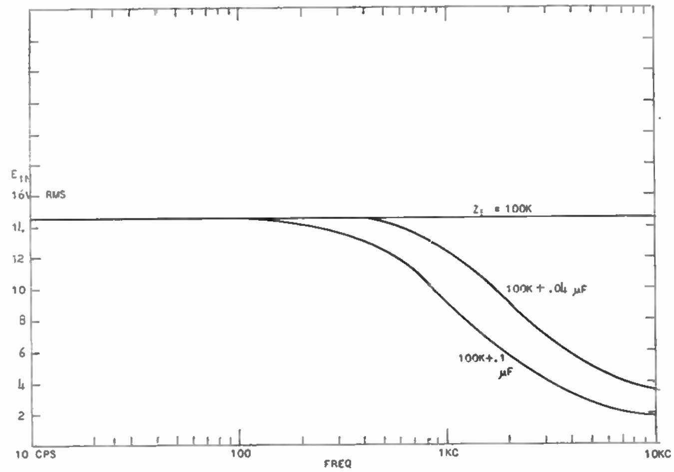


Fig. 7—Input overdrive characteristic of 12AU7 cathode follower.

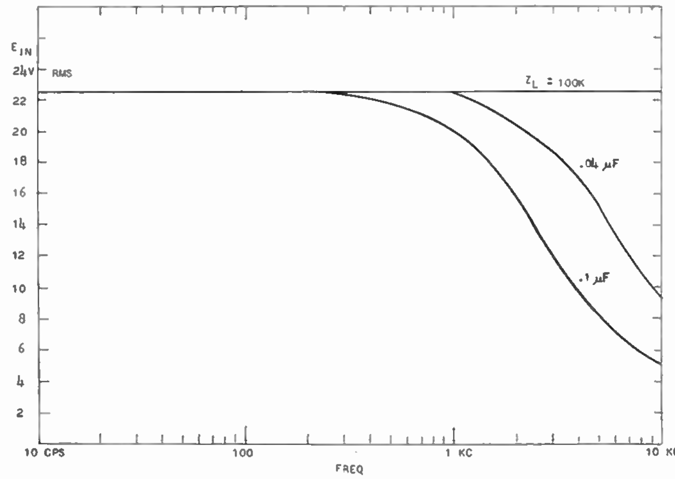


Fig. 9—Input overdrive characteristic of 6BX7 cathode follower.

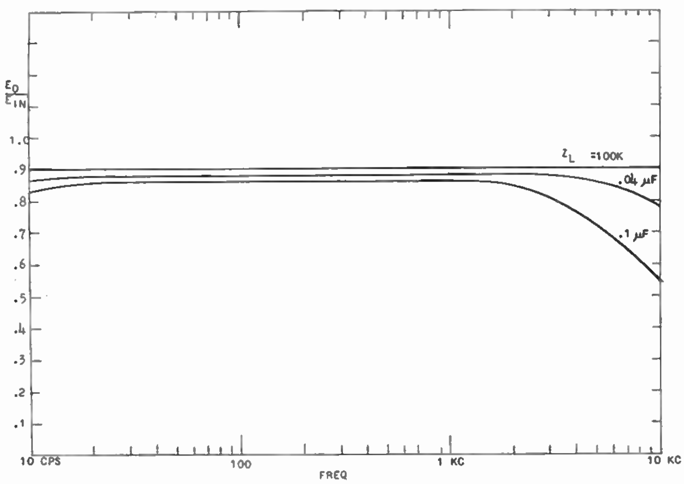


Fig. 8—Frequency response characteristic of 12AU7 cathode follower.

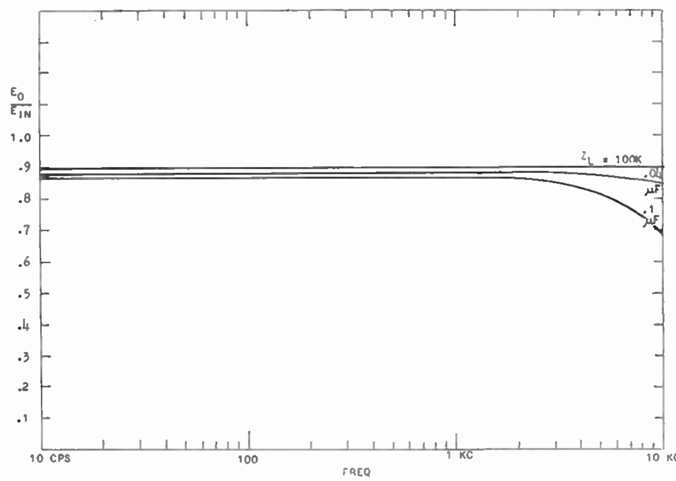


Fig. 10—Frequency response characteristic of 6BX7 cathode follower.

Design Considerations

Evaluation of the information derived from experimental investigation of cathode follower operation leads to the following guides for design of cathode followers for high capacitance load driving capability:

1) Selection of a vacuum tube with low R_p and low or medium amplification factor.
2) Selection of cathode resistance load, R_{k1}, which is comparable to that suitable for a Class A power amplifier, one to four times R_p.
3) Selection of grid bias, by variation of R_{k2}, for symmetrical voltage variation around the average no-signal cathode voltage.
4) R_0 should not be greater than a few per cent of X_c of the load at the highest frequency of interest, unless a moderate reduction of the overdrive limit can be tolerated. Examination of Figs. 7–10 shows that if R_0 is as much as 10 per cent of X_c, the overdrive threshold is seriously reduced.

Conclusions

Experimental evidence shows that cathode followers with high-capacitance loads are subject to the following effects:

1) Input overdrive threshold is greatly reduced at high frequencies.
2) High-frequency response deteriorates.
3) Harmonic distortion (nonlinearity) is increased at high frequencies.
4) Phase shift between input and output is greatly increased, especially at high frequencies.
5) Transient response at high frequencies deteriorates.

These effects can be minimized by proper design based on attainment of low output impedance (R_0).

Contributors

Anthony J. Adducci (A'60) was born in Chicago, Ill., on August 14, 1937. He received the B.S. degree in physics in 1959 from St. Mary's College, Winona, Minn., and did additional study at the Illinois Institute of Technology, Chicago.

A. J. Adducci

He joined the engineering staff of Jensen Manufacturing Company, Chicago, Ill., in February of 1960, where he is currently working in the Loudspeaker Development Department. One of his projects consisted of research work into the nature and effects of transient distortions in loudspeakers and correlations between transient measurements and listening tests.

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Robert W. Benson (M'52) was born in Grand Island, Neb., on January 21, 1924. He received the B.S.E.E., M.S.E.E. and the Ph.D. degrees from Washington University, St. Louis, Mo., in 1948, 1949 and 1951, respectively.

R. W. Benson

From 1948 to 1954 he was a Research Associate with the Central Institute of the Deaf, St. Louis, Mo.; and from 1954 to 1960 he was with the Armour Research Foundation, Chicago, Ill., where he was Assistant Director of the Physics Division, having previously served as supervisor of the acoustics section. Currently he is a Professor of Electrical Engineering at Vanderbilt University, Nashville, Tenn. He has been active in all phases of acoustics including electroacoustics, physiological acoustics and room acoustics, as well as working in the field of ultrasonics.

Dr. Benson is a Fellow of the Acoustical Society of America, a member of Sigma Xi, a member of the editorial board of *Noise Control*, an editorial consultant to *Industrial Research*, a member of CHABA, and a consultant to the American Academy of Ophthalmology and Otolaryngology.

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Alexander B. Bereskin (A'41–M'44–SM'46–F'58) was born in San Francisco, Calif., on November 15, 1912. He received the E.E. degree and the M.Sc. degree in engineering from the University of Cincinnati, Ohio, in 1935 and 1941, respectively.

A. B. Bereskin

He has done work in the fields of special RC oscillators, frequency-selective amplifiers, low-jitter multivibrators, special stabilized power supplies, and transistor pulse amplifiers. He has published work on vacuum-tube and transistor audio power amplifiers, low-level transistor audio amplifiers, video amplifiers, regulated power supplies, and power-factor meters. He is presently Professor of electrical engineering at the University of Cincinnati. He is also active in consulting engineering and is a Registered Professional Engineer in Ohio.

Professor Bereskin is presently Director of IRE Region 4 and a member of the Administrative Committee of the IRE-PGA. He was formerly Chairman of the IRE Professional Group on Audio and Editor of the IRE Transactions on Audio. He was a member of the Education Committee, Chairman of the Region 4 Subcommittee of the Education Committee, member of the Professional Groups and the Sections Committees, and Institute Representative at the University of Cincinnati. He is Past Chairman, Vice-Chairman and Treasurer of the Cincinnati Section. He is also a member of the AIEE, Sigma Xi, Eta Kappa Nu, and Tau Beta Pi.

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Norman H. Crowhurst (SM'53) was born in Essex, England, on November 3, 1913. He graduated from South East London Technical College, England, in 1935, and subsequently served as Senior Lecturer in electrical engineering at the same college and also at Beckenham Technical College, Kent County.

N. H. Crowhurst

He also became Chief Engineer of Tannoy Lmited, London, England, in 1935, where he was responsible for expanding the activities of that company into all phases of audio. In 1945 he became a free-lance consulting engineer in England and commenced his writing activities in order to make available to the industry the techniques in audio which he had developed. In 1953 he emigrated to this country, where he became a citizen in 1960. He was employed as a Project Engineer at Fairchild Recording Equipment Company, Long Island City, N. Y., where he was responsible for developing the Perspecta system of stereophonic sound for motion pictures. Since 1955 he has been a free-lance consultant engineer for various companies and is currently working for CBS Laboratories, Stamford, Conn., in the area of audio and electroacoustics.

Mr. Crowhurst is a member of the IEE, the British Sound Recording Association, the AES, the SMPTE, and the Acoustical Society of America.

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Thomas L. Greenwood (A'53–SM'55) was born in Prosper, Texas, on August 12, 1908. He attended Washington University, St. Louis, Mo., from 1950 to 1951, and the University of Alabama, Huntsville Branch, from 1951 to 1953.

T. L. Greenwood

He has had 32 years experience in electronics and communications, including eight years as a Merchant Marine Radio Officer; four years as a radio broadcast technician; eight years as Chief Engineer of Radio Station, WMOB, Mobile, Ala., and three years in the Communications Department of the Gulf, Mobile, and Ohio Railroad. From 1951 to 1960 he supervised activities in guided-missile instrumentation development at the U. S. Army Recorder and Electronics Laboratory, Redstone Arsenal, Huntsville, Ala. At present he is with the George C. Marshall Space Flight Center, NASA, Huntsville, Ala., engaged in the development of instrumentation and data handling systems.

Mr. Greenwood is a member of the Rocket City Astronomical Association.

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Robert J. Larson (S'51–A'52–M'55) was born in Galesburg, Ill., on January 7, 1926. He received the B.S.E.E. degree from Northwestern University, Evanston, Ill., in 1951.

While an undergraduate, he was employed by the Jensen Manufacturing Company, Chicago, Ill., as a student engineer under Northwestern's Cooperative Program.

R. J. Larson

After 22 months on active duty with the Navy during the Korean conflict, he returned to Jensen, where he is presently engaged in loudspeaker development work.

Mr. Larson is a member of the Acoustical Society of America and is a Registered Professional Engineer.

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Peter W. Tappan (M'56) was born in New York, N. Y., on September 26, 1928. He received the B.S. degree in physics in 1952, and the M.S. degree in 1958, both from the Illinois Institute of Technology, Chicago.

P. W. Tappan

From 1951 to 1956 he worked in the Physics Department of the Armour Research Foundation, Chicago, Ill., performing research on such varied projects as an X-ray intensification system, special recording heads, an electric piano, and high-powered public address systems. In 1956 he joined the Warwick Manufacturing Corp., Chicago, where currently he is responsible for acoustical and audio research; he also has worked on the design and development of speaker systems, phonograph pick-ups, and stereophonic equipment.

Mr. Tappan is a member of the Acoustical Society of America.

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