IRE Transactions on AUDIO

Volume AU-6 **NOVEMBER-DECEMBER, 1958** **Number 6**

Published Bi-Monthly

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PUBLISHED BY THE

Professional Group on Audio

IRE PROFESSIONAL GROUP ON AUDIO

The Professional Group on Audio is an organization, within the framework of the IRE, of members with principal professional interest in Audio Technology. All members of the IRE are eligible for membership in the Group and will receive all Group publications upon payment of an annual fee of $2.00.

IRE TRANSACTIONS® on AUDIO

Published by the Institute of Radio Engineers, Inc., for the Professional Group on Audio at 1 East 79th Street, New York 21, New York. Responsibility for the contents rests upon the authors, and not upon the IRE, the Group, or its members. Individual copies available for sale to IRE-PGA members at $0.55; to IRE members at $0.85, and to nonmembers at $1.65.

PGA News

MINUTES OF THE FALL ADMINISTRATIVE COMMITTEE MEETING

Members Present

B. B. Bauer, *Secretary-Treasurer*
A. B. Bereskin
M. Camras
P. B. Williams

Members Absent

S. J. Begun
M. S. Corrington
J. K. Hilliard
H. F. Olson
W. T. Selsted
F. H. Slaymaker, *Chairman*

The meeting was held at the Hotel Sherman, Chicago, Ill., on Monday, October 13, 1958. Since a quorum was not present, a formal meeting of the Administrative Committee was not held.

1) The Secretary-Treasurer discussed the estimated budget for the year 1959 to be submitted to the IRE. This budget was approved by those present, a copy being enclosed herewith.
2) The Secretary-Treasurer discussed a recent visit by F. H. Slaymaker and communication from L. G. Cumming about changes in group financing and a decision to consider discontinuance of the CONVENTION RECORD. It appeared, however, that in order to benefit the Professional Groups, all the authors at the convention should be required to submit a manuscript, as before, for perusal by the Professional Groups' Editorial Committees.

   By motion of A. B. Bereskin, seconded by M. Camras, it was agreed to request the IRE to require a written version of papers before presentation at the IRE National Convention so that these papers may be available for publication by the Professional Groups.
3) M. Camras reported having sent to J. R. MacDonald all the procedures for the Chapters Committee.
4) The Secretary-Treasurer reported about a recent communication with the Chairman about the nominations and elections, and stated that he expected a momentary action from the Nominations Committee, which would have to be handled by letter.

B. B. BAUER,
Secretary-Treasurer

ESTIMATED BUDGET
JANUARY 1 to DECEMBER 31, 1959

Balance From December 31, 1958		$17,500.00
Receipts During Period:		
IRE Matched Funds	$ 3,000.00	
Fees	9,400.00	
Advertising	300.00	
Sale of Publications	1,100.00	
Surplus From Meetings	none	
Other Sources	none	
Total Receipts	$13,800.00	13,800.00
Total Balance and Receipts		$31,300.00
Expenses During Period:		
Publications	$ 9,000.00	
Membership Service Charges	500.00	
Group Awards	500.00	
Editorial Administrative Expense	500.00	
Others	900.00	
Total Expenses	$11,400.00	11,400.00
Balance as of December 31, 1959		$19,900.00

CHAPTER NEWS

J. Ross MacDonald, Chairman of the Committee on Chapters, would like to receive reports of meetings and other activities of local PGA Chapters. He will circulate such information in a Newsletter for chapter officers to help them plan interesting meetings. He will also see that your chapter news appears in the TRANSACTIONS. Please add Mr. MacDonald to your mailing list for chapter and section announcements. His address is Texas Instruments, 6000 Lemmon Avenue, Dallas 9, Texas.

ANNOUNCEMENTS

The Tenth Annual Convention of the Audio Engineering Society was held at the Hotel New Yorker, New York, N. Y., September 29 through October 3, 1958. Program papers are listed below. Preprints of many of these are available at $0.85 each from the Audio Engineering Society, P.O. Box 12, Old Chelsea Station, New York 11, N. Y.

Direct-Coupled Transistor-Tube Audio Amplifier for Radio and TV—*Palmer and Schiess*

Applying the Transistor in a Stereophonic Tape System—*Jones*

A 25-Watt High Quality Transistorized Audio Frequency Power Amplifier—*Minton*

Design of a Stereo Control Center—*Grace*

Psychological Factors Governing the Binaural Effect—*Cooley*

Broadening the Area of Stereophonic Perception—*Bauer*

Psychoacoustics Applied to Stereophonic Reproduction Systems—*Goldmark, Bauer and Hollywood*
Stereo as an Integral System—*Crowhurst*
Determining the Area of Stereo Perception—*Cohen*
Several Problems in Musical Acoustics—*Clark*
Specialized Equipment Used at Columbia University Studio for the Production of Tape Music—*Mauzey*
Sound Synthesis by the Use of Magnetized Arrays—*Frisch*
Musical Timbre Mutation by Means of the "Klangumwandler," A Frequency Transposition Device—*Ussachevsky*
Generation of Music by a Digital Computer—*Mathews and Guttman*
Possibilities for Combining Electronic Music, Tape Music and Musique Concrete with Traditional Music—*Luening*
From Stereo Tape to Stereo Disk—*Grundy*
A New Master Disk Recording Lathe for Stereo Disk Use—*Temmer*
A New Stereo Feed Back Cutterhead System—*Narma*
Checking the Axes of Operation of Stereo Cutter Channels—*Redlich*
Recent Developments in Stereophonic Disk Recording—*Davis and Frayne*
The Teldec Miniature Stereo Cutterhead System—*Temmer*
Some Thoughts on Geometric Conditions in the Cutting and Playing of Stereo Disks and Their Influence on the Final Sound Picture—*Bastiaans*
A Stereophonic Variable Reluctance Phonograph Cartridge—*Pritchard*
The Development of a High Quality Stereophonic Pickup Cartridge—*Stanton*
A Single-Element Stereo Cartridge—*Wood*
A Constant Displacement Stereo Cartridge—*Bachman*
The Development of a Double Rotating Coil Pickup—*Narma*
An Investigation of Magnetic Tape Drop-Outs—*Carson*
50 "Mylar" T—A Film Especially Designed for Use as a Magnetic Tape Base—*Barton*
Magnetic Tape Recording with Longitudinal or Transverse Oxide Orientation—*Dubbe*
The Noise in Magnetic Recording Which Is a Function of Tape Characteristics—*Smaller*
Signal-to-Noise Problems and New Equalization for Magnetic Recording of Music—*McKnight*
A New Device for the Reduction of Print-Through—*Radocy*
Evolution of a Successful Spring Driven Broadcast Quality Tape Recorder—*Travis*
Magnetic Recording of Audio Frequencies—*Beachell*
Electromagnetic Efficiency of Magnetic Recorder Heads—*Camras*
Factors Affecting High-Frequency Response at Low Tape Speeds—*Youngquist*
High Quality Reproduction of Magnetic Tape in a Cartridge—*Andrews*
A Magnetic Head for Stereo or Half Track on One-Eighth Inch Tape—*Warren*
Professional High-Speed Duplication of the New Four-Track Stereo Tapes—*Tinkham*
A Magnetic Disk Recording and Reproducing System—*Singer*
A Magnetic Head for Grooved Magnetic Recording Disks—*Warren*
Properties and Performance of Magnetic Rubber Recording Belts—*Fabing and Hartel*
The Automatic Plotting of Cartridge Response—*Germano*
Standards—Stepchild in the Laboratory—*White*
The ARP 2—A New Instrument for Sound Measurement—*Erath*
Determination of Absolute Recording Sensitivity of Magnetic Tape—*Daniel and Levine*
Selective Paging by Means of Audio Frequencies—*Phillipps*
Breaking the Industrial Sound Barrier—*Leibowitz*
Stereophonic Broadcasting by FM Multiplex Method—*Halstead and Burden*
Performance Characteristics of FM Multiplex Stereo Transmission—*Crosby*
Stability Considerations in High Fidelity Amplifiers—*Grenier*
Single Push-Pull Stage Amplifiers for Stereophonic Sound Reproduction—*Bauer, Bachman, Hollywood and Maerkle*
Designing a Multi-Purpose Stereo Pre-Amplifier—*de Miranda*
A Three-Channel Stereophonic Sound Reinforcement Mixing Console—*Erhorn*
An Audio Console Designed for the Future—*Angus*
Multi-Channel Stereophonic Mixer Console—*Miltenburg*
A Packaged Equipment for the Production of True Reverberation—*Franz and Skee*
The Development and Application of Synthetic Reverberation Systems—*Goodfriend and Beaumont*
A Definitive Loudspeaker System for Monitoring in Control and Audition Rooms—*Franz and Skee*
Two New Horns and Drivers Covering High Frequencies—*Levy, Matsuoka and Brociner*
Analysis of a Low-Frequency Loudspeaker System—*Tappan*
The Use of Subjective Measurement Scales in Audio Engineering—*Liebich*
Intermodulation Distortion Testing of Loudspeakers—*Malmsten*
Performance of Enclosures for Low Resonance High Compliance Loudspeakers—*Novak*
A New Wide Angle Direct Radiator Tweeter—*Petrie*
Three New High Efficiency Speakers for P.A. Use—*Levy, Sioles, Carlisle and Sharp*
A Novel Compact Flexible Stereo Speaker System—*Berlant*

A New Approach to Vented Cabinet Design—*Sioles and Brociner*

Development of a Full-Range Transducer Assembly for the Generation of High Intensity Sound—*Liebich and King*

Audio Papers for the 1959 IRE National Convention

The following are abstracts of papers which have been submitted for presentation on the program of the IRE National Convention in New York, March, 1959:

Three Channel Stereo Playback of 2-Tracks Derived from 3 Microphones, *Paul W. Kilpsch*—Three-channel stereo from 2-sound tracks is an established art with remarkably simple implementation. Experiments on 2-track tape recordings made with 3 microphones indicate that center events are retained in focus regardless of polarity of the tracks.

Where a center microphone is mixed equally into the two tracks and 3-channel playback is by the subtractive phantom system, the third microphone would be cancelled out. A phase-shifting network is employed to restore the center channel.

A Zero-Drift Direct Coupled Amplifier Utilizing a Clipper-RC Feedback Loop, *J. N. Van Scoyoc and E. S. Gordon*—A direct-coupled amplifier has been developed utilizing a combination clipper and RC feedback loop which is virtually dc drift free. Although the developed amplifier is one using vacuum tubes, transistors could be used equally as well. Several interesting possible applications of the amplifier exist.

In the theory of operation given, a direct coupled amplifier with a simple RC negative feedback loop is first treated. In this amplifier the gain is essentially unity for dc and increases as frequency is increased, finally reaching a plateau. Equations are developed and theoretical curves are plotted for this amplifier. Treated next is the actual amplifier configuration which utilizes a clipper circuit as well as an RC integrator in the feedback loop. In this circuit, also, the gain rises with frequency from unity at dc to a given plateau, but here the slope of the curve during the rise is extremely steep. Equations are developed and theoretical curves are given for the configuration. In addition, experimental curves are shown which agree quite closely with the theoretical curves. The experimental data are actually taken from a unit developed for the counting of aerosol particles; this application is next described.

The aerosol counting application is one involving the automatic counting and sizing of airborne particles over the very large dynamic range of approximately 4000 to one. The aerosol counter in point is an improved version which uses this amplifier. A publication on the original version is referenced. Simplified circuit diagrams of this amplifier are given and it is shown how the amplifier is ideally suited to the wide dynamic range required.

Another specific amplifier of the subject type has been designed which is adaptable to a number of additional possible applications. This design, the circuit of which is shown, has high output sginal and other characteristics appropriate for the possible applications.

One of the additional possible applications of the amplifier is its use as an actual dc amplifier with essentially zero drift. This application allows simplifications in chopping and detecting over conventional chopper amplifiers or chopper-stabilized amplifiers. It requires simply a single contact chopper at the input; output detection is inherent in the amplifier.

Another possible application of the amplifier involves the use of a single amplifier for the dc amplification of up to 20 separate dc input signals. In this system a motor-driven switch samples the multiple inputs and also applies the amplifier output to separate peak-reading indicating meters. No choppers are used in the amplifier proper. The device might be used very advantageously in multiple strain gauge or thermocouple measurement or recording.

A Frame-Grid Audio Pentode for Stereo Output, *J. L. McKain and R. Schwab*—A dual pentode using a single cathode, two separate *Framelok* grids and a twin-plate structure contained in one envelope is described. This new pentode, known as Type 6DY7, is a high-performance tube with superior characteristics of uniformity and stability obtained from its unique structure. Such factors as greater tube-to-tube characteristics uniformity, reduced characteristic spread, and less susceptibility to characteristic deterioration at high dissipations can be obtained.

This dual pentode offers extreme flexibility in application. Three basic configurations are: 1) sections operated separately (single-ended) giving 5 to 6 watts of audio power per section, 2) two sections in push-pull, Class AB_1 providing up to 20 watts output at less than 3 per cent distortion, and 3) two tubes in push-pull parallel.

A single tube can be used for two stereo output channels, or two tubes can be operated in push-pull for higher power requirements. The same advantages can be used for monoural audio systems.

The tube, therefore, offers the circuit designer a choice of usage not possible in presently available tubes and at cost advantages realizable through a reduction in the number of circuit components.

An Amplitude Modulation Stereophonic Sound Radio Broadcasting System, *H. F. Olson, D. S. McCoy, R. W. George, L. E. Barton, and H. C. Allen*—The advent of stereophonic sound reproduction by means of the magnetic tape, the phonograph disk record and the multiplex frequency modulation radio broadcasting systems makes it practically a requirement that stereophonic sound be extended to include the amplitude modulation system in the standard radio broadcast band in order to make stereophonic sound reproduction in the consumer entertainment complex complete. In this connection, it is particularly important and significant to point out that the amplitude modulation radio broadcasting system represents the largest element for the mass dissemination of sound reproduction in the consumer field as exemplified by the home, portable and automobile receivers. Consequently, to fill the existing void in stereophonic sound reproduction for consumer entertainment, amplitude modulation stereophonic sound radio broadcasting systems have been developed consisting of a two-channel stereophonic transmitter and a two-channel stereophonic receiver. In the amplitude modulation stereophonic sound radio broadcasting systems the two-channel stereophonic sound information is carried on the carrier and the sidebands within the same frequency band as the existing monophonic amplitude modulation radio system. It has been demonstrated that the amplitude modulation stereophonic sound radio broadcasting system is completely compatible with the existing amplitude modulation monophonic sound radio broadcasting system because the following combinations are possible, namely, stereophonic transmitter-stereophonic receiver, stereophonic transmitter-monophonic receiver, monophonic transmitter-stereophonic receiver, and monophonic transmitter-monophonic receiver.

The Single Stereophonic Amplifier, *B. B. Bauer and J. M. Hollywood*—The single stereophonic amplifier is briefly described. This amplifier handles both the left and right signals with the same pair of output tubes. The circuit is analyzed for the theoretical requirements for good separation and low distortion.

The separate amplification of two signals is obtained by adding a matrixing output transformer.

The turns ratio for ideal left and right separation is established for a circuit with finite plate and load impedance. Next, the effect of finite primary inductances on performance is determined.

The influence of negative feedback on separation and distortion is analyzed and stability conditions discussed.

Audio Applications of a Sheet-Beam Deflection Tube, *J. N. Van Scoyoc*—A number of unusual audio circuits have been developed which make use of a sheet-beam tube, type 6AR8. The 6AR8 tube is a miniature double-plate sheet-beam tube which incorporates a pair of balanced deflectors to direct the electron beam to either of the two plates and a control grid to vary the intensity of the beam.

This tube may be connected as a variable gain push-pull amplifier by connecting the input signal between the two deflectors and taking the output differentially between the two plates. When the tube is connected in this manner the amplifier gain is determined by the control grid voltage and may be varied over an 80-db range with negligible distortion.

The applications of this circuit include expansion and compression circuits, remote control of gain and mixing circuits, improved AVC circuits and phase inversion. A number of these circuit arrangements are given in some detail and other applications are outlined.

An Improved Method for the Measurement of Nonlinear Audio Distortion*

JAMES S. AAGAARD†

Summary—**The three most common methods of measuring nonlinear distortion in audio equipment are the harmonic method in which harmonics of a single sinusoidal input signal are measured; the SMPTE, or modulation, method in which the modulation of a high audio frequency by a low frequency is measured; and the CCIF, or difference frequency, method in which the beat note between two closely spaced frequencies is measured. These methods are discussed with particular reference to the behavior which may be expected in applications where the equipment under test includes pre-emphasis or de-emphasis networks, or where the distortion is symmetrical. It is shown that in a number of cases a more satisfactory method for use at the higher audio frequencies would be the measurement of the third-order component, rather than the second, produced in the CCIF method. A modification of the CCIF method using a sharp-cutoff low-pass filter is described which is capable of measuring both second- and third-order components. It is then shown that an instrument for this new method and the standard SMPTE method have many elements in common and that one instrument could be devised to make both forms of tests. The discussion is illustrated with the results of measurements made on a simulated distortion generator and on actual samples of audio equipment.**

Introduction

Any practical transmission system which may be constructed will inevitably introduce some distortion into a signal. Using the principles of Fourier analysis, the actual input may always be considered as the sum (or integral) of a number of sinusoidal components. If the system introduces no distortion, these same components, and no others, will appear at the output of the system, their relative magnitudes and phases unchanged. Distortion is indicated by the introduction of additional components, which is usually called nonlinear distortion, and by changes in the relative magnitudes and phases of those components present at the input. The latter effect is called linear or frequency distortion.

Although this study is limited primarily to nonlinear distortion, it will be shown that frequency distortion can have considerable effect on such nonlinear distortion. In order to narrow the scope still further, the analysis will be restricted to systems which transmit signals intended ultimately for the human ear—a video system, for example, might require entirely different performance criteria. And finally, implicit in the discussion is the assumption that a listener prefers sound which is a faithful reproduction of the original, thus there is some sort of correlation between audible quality and measured distortion.

* Manuscript received by the PGA, October 17, 1958. Presented at the National Electronics Conference, Chicago, Ill., October 13, 1958.

† Dept. of Elec. Eng., Northwestern University, Evanston, Ill.

Methods for the Measurement of Nonlinear Distortion

Methods for the measurement of nonlinear distortion may be divided into two basic classifications, *harmonic* and *intermodulation*. In the harmonic methods, which have been in use much longer than other methods, a single sinusoidal signal is supplied to the equipment under test. The output is then examined for components having frequencies which are integral multiples of the frequency of the input signal. The magnitudes of these additional components in comparison to the magnitude of the original signal provide a measure of the nonlinear distortion which has been introduced.

The difficulties which were encountered when attempts were made to correlate harmonic measurements with subjective estimates of quality prompted the development of the intermodulation methods.[1–3] These methods all require the application of at least *two* sinusoidal input signals, and the distortion is determined from the amplitudes of components of the output which are not harmonically related to either of the input signals.

Since all of the actual signals—music or speech—which pass through an audio system already contain a number of harmonic components, additional harmonics introduced by the system might not be expected to be especially objectionable. This is not to say that they would pass unnoticed, but rather that they would not be discordant. On the other hand, components which are not harmonically related to any of the original components might be expected to be quite objectionable. Of course, a nonlinear system will always produce both harmonic and intermodulation components and thus it might seem adequate to measure either. However, as discussed below, these components do not necessarily have a fixed relationship to each other, and therefore it seems more reasonable to measure directly the components which contribute to the unpleasantness of the reproduced sound. This, briefly, is the argument for the use of intermodulation methods of distortion measurement. Fig. 1 illustrates a harmonic method and two intermodulation methods.[4]

[1] J. G. Frayne and R. R. Scoville, "Analysis and measurement of distortion in variable-density recording," *J. SMPE*, vol. 32, pp. 648–673; June, 1939.

[2] J. K. Hilliard, "Distortion tests by the intermodulation method," Proc. IRE, vol. 29, pp. 614–620; December, 1941. Discussion, vol. 30, p. 429; September, 1942.

[3] H. H. Scott, "Intermodulation measurements," *J. Audio Eng. Soc.*, vol. 1, pp. 56–61; January, 1953.

[4] A. P. G. Peterson, "The Measurement of Nonlinear Distortion," General Radio Co., Cambridge, Mass. Tech. Pub. B-3; 1949.

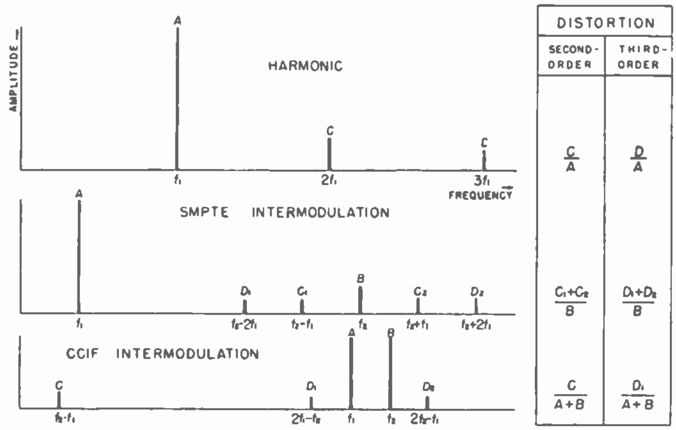

Fig. 1—Methods of measuring distortion.

From a practical point of view it is not feasible to test every piece of equipment using all of the methods of measurement. Of course, this would not be necessary if some means could be found to convert distortion factors obtained by one method into those which could be obtained by other methods. This problem has occupied the attention of a number of investigators.[4–12] In brief, they have shown that such conversion is possible in some extremely simple cases but seems to break down for unknown reasons in cases of more practical importance. Actually it is not at all difficult to devise two systems which give indistinguishable results when measured according to the harmonic method and completely different results when measured according to the intermodulation methods, thus showing that such conversion is not possible in the most general case.

It is important, therefore, to attempt to determine the reasons why such analysis fails to give results which may be verified experimentally. It would appear that there are at least two.

In order to provide a means for making a mathematical analysis of distortion, it is convenient to assume a sinusoidal input to the system under consideration. As mentioned above, any actual input may be represented as the sum of such sinusoids, so that this approach involves no loss of generality. Thus, let

$$v_{\text{in}} = V \cos(\omega t + \theta). \tag{1}$$

[5] W. J. Warren and W. R. Hewlett, "An analysis of the intermodulation method of distortion measurement," PROC. IRE, vol. 36, pp. 162–173; February, 1958.

[6] A. Bloch, "Measurement of nonlinear distortion," *J. Audio Eng. Soc.*, vol. 1, pp. 62–67; January, 1953.

[7] D. E. Maxwell, "Comparative study of methods for measuring nonlinear distortion in broadcasting audio facilities," *J. Audio Eng. Soc.*, vol. 1, pp. 68–78; January, 1953.

[8] H. F. Olson, "Elements of Acoustical Engineering," (2 ed.), D. Van Nostrand Co., New York, N. Y., pp. 487–491; 1947.

[9] C. J. LeBel, "An experimental study of distortion," *J. Audio Eng. Soc.*, vol. 2, pp. 215–218; October, 1954.

[10] H. E. Roys, "Distortion in phonograph reproduction," *J. Audio Eng. Soc.*, vol. 1, pp. 78–85; January, 1953. See also *RCA Rev.*, vol. 14, pp. 397–412; September, 1953.

[11] H. Mueller, "Doppelton- und Klirrfaktormessung: Unterschiede zwischen den Ergegnissen und ihre Ursachen," *Telefunken Zeitung*, vol. 25, pp. 142–148; August, 1952.

[12] M. Pokrovsky, "La mesure de la distorsion non linéaire et les causes de divergence des résultats obtenus par les différentes méthodes," *L'Onde Electrique*, vol. 34, pp. 525–533; June, 1954.

This signal, passed through a system which introduces only frequency distortion, would then be of the form

$$v_{\text{out}} = A(\omega)V \cos[\omega t + \theta + \phi(\omega)]. \tag{2}$$

In a system producing only nonlinear distortion, the actual relationship between v_{out} and v_{in} may have various forms, but in any case of practical importance it may be expressed as a power series,

$$v_{\text{out}} = av_{\text{in}} + bv_{\text{in}}^2 + cv_{\text{in}}^3 + dv_{\text{in}}^4 + \cdots \tag{3}$$

If the transfer characteristic possesses relatively sharp corners, however, this series will not converge rapidly. Because of the great mathematical complexity involved, it is customary to use only the first few terms of the series. In an ordinary vacuum-tube amplifier, for example, it is found that the second-order term is by far the largest, and a good approximation to the actual behavior may be obtained by using this term alone. Unfortunately, however, in systems in which the second-order term is missing, for example, a push-pull amplifier, it does not follow that good results may be obtained from the third-order term alone. Since many systems do have a symmetrical characteristic of this type, it appears that the use of a few terms of a power series to calculate distortion products may lead to serious errors.

Furthermore, the mathematical representations given in (2) and (3) are based on the assumption that only frequency distortion or only nonlinear distortion is produced in the particular system under consideration. A system containing elements which produce both frequency and nonlinear distortion, however, is much more difficult to analyze mathematically. In many cases the causes of frequency and nonlinear distortion may be isolated from each other, and in these cases the two classes of distortion may be treated separately. An important type of system in which this is true is a system which includes pre-emphasis of certain frequencies. Examples of such systems are those involving the recording and reproduction of sound. In the manufacture of phonograph records, it is standard practice to pre-emphasize the higher frequencies when the recording is being made, and then to de-emphasize them when the recording is played. However, there is a possibility of nonlinear distortion being produced in the final stages of the recording amplifier or in the actual recording process after the signal has passed through the pre-emphasis circuit. Furthermore, the de-emphasis circuit will affect the distortion components in different manners depending on their frequencies. Also, additional distortion may be introduced in the playback amplifier after the de-emphasis has taken place.

Other systems employing pre-emphasis include FM broadcasting, motion-picture-film sound recorders, and magnetic tape recorders.

It is obviously impossible to perform a mathematical analysis which will embrace all possible combinations of nonlinear and frequency distortion. On the other hand, it would appear that the simplified analyses presented in the past are inadequate to explain observed results. Since the production of nonlinear and frequency distortion in the same piece of equipment can result in such a discrepancy, it would seem desirable to extend the analysis as far as possible in this direction. If there is still lack of agreement, then the assumptions underlying the analysis must be examined further.

These assumptions are:

1) The nonlinearity in the system may be represented by a power series consisting of only a few terms.
2) The system has no over-all frequency distortion, although the effect of the nonlinearity may depend on frequency.
3) The effect of the nonlinearity is relatively small.
4) The peak-to-peak amplitude of the input signal is the same for all methods of measurement.

The third condition permits the contribution of the nonlinearity at the frequency of the input signal to be neglected in computing distortion factors. The fourth condition, of course, is under the control of the operator and may therefore always be satisfied.

In accordance with the first condition, a nonlinear distortion characteristic similar to the first three terms of (3), but with $a=1$, will be assumed:

$$v_2 = v_1 + bv_1^2 + cv_1^3. \tag{4}$$

Whether or not this is an adequate number of terms will, of course, depend on the characteristics of the particular device which the series is supposed to represent.

The simplest manner in which the effect of the nonlinearity may vary with frequency corresponds to the impedance of a resistance-capacitance or resistance-inductance network. For example, the nonlinear element might be preceded by such a network, producing an increase in the signal amplitude at a rate of 6 db per octave, beginning at some fixed frequency. This is the type of characteristic used for pre-emphasis in the manufacture of phonograph records and in FM broadcasting.

However, caution is necessary when measurements of nonlinear distortion are made on elements of a system involving pre-emphasis in order to satisfy the second assumption. The reason for this assumption is primarily due to the fact that an audio transmission system is almost always intended to convert a particular sound pressure at a microphone into a particular sound pressure at a listener's ear with the ratio of these sound pressures being independent of frequency. Thus, if frequency distortion is introduced deliberately at some point in the system for a certain reason, it is usually corrected at some other point.

It seems reasonable, therefore, that if electrical measurements are made on one element of this system (rather than on the entire system), then external linear networks should be provided between test equipment and the element under test in order to produce the same frequency distortion normally produced by the other elements of the complete system.

The examples discussed below have been compensated in this way so that their over-all response-frequency characteristic is uniform. Case I consists of high-frequency pre-emphasis, a nonlinear element, and then high-frequency de-emphasis. Case II, included for comparison, is the same nonlinear distortion without any accompanying frequency distortion. Case III is similar to case I except that the pre-emphasis and de-emphasis are at the low frequencies instead of at the high frequencies.

The actual calculation of the formulas for distortion measured by the various methods in the general case of a second- and third-order nonlinearity between two sources of frequency distortion will not be given here, but the results are shown in Table I. It should be noted that the same nonlinearity will give different numerical values of distortion factor, even when the nonlinearity is independent of frequency, depending on the method of measurement used. Therefore, in Table II the various distortion factors of Table I have been normalized to unity for case II to facilitate a comparison of these methods.

Examination of Table II shows that for the detection of low-frequency second-order nonlinear distortion, the harmonic method is twice as sensitive as the modulation

TABLE I

VALUES OF DISTORTION FACTORS FOR NONLINEAR DISTORTION WHICH IS A FUNCTION OF FREQUENCY

Type of measurement and input frequencies		Case I (High-frequency pre-emphasis)	Case II (Uniform response)	Case III (Low-frequency pre-emphasis)
Second-Order Distortion Factors				
Harmonic	0.1	0.495 bV	0.50 bV	9.9 bV
	1.0	0.447	0.50	0.895
	5.0	1.293	0.50	0.517
Modulation	0.1 5.0	1.61	1.60	16.1
Difference frequency	0.1 0.12	0.253	0.25	0.422
	1.0 1.2	0.54	0.25	0.0905
	5.0 6.0	5.5	0.25	0.183
Third-Order Distortion Factors				
Harmonic	0.1	0.243 cV²	0.25 cV²	7.3 cV²
	1.0	0.223	0.25	0.671
	5.0	2.20	0.25	0.265
Modulation	0.1 5.0	0.97	0.96	97.0
Difference frequency	0.1 0.12	0.0955	0.0938	6.34
	1.0 1.2	0.229	0.0938	0.153
	5.0 6.0	3.6	0.0938	0.0993

TABLE II

VALUES OF DISTORTION FACTORS FOR NONLINEAR DISTORTION WHICH IS A FUNCTION OF FREQUENCY, NORMALIZED TO THE CONDITION WHERE THE NONLINEAR DISTORTION IS NOT A FUNCTION OF FREQUENCY FOR EACH METHOD OF MEASUREMENT

Type of measurement and input frequencies		Case I (High-frequency pre-emphasis)	Case II (Uniform response)	Case III (Low-frequency pre-emphasis)
Second-Order Distortion Factors				
Harmonic	0.1	0.99	1.0	19.8
	1.0	0.894	1.0	1.79
	5.0	2.59	1.0	1.034
Modulation	0.1, 5.0	1.005	1.0	10.05
Difference frequency	0.1, 0.12	1.01	1.0	1.686
	1.0, 1.2	2.16	1.0	0.362
	5.0, 6.0	22.0	1.0	0.73
Third-Order Distortion Factors				
Harmonic	0.1	0.972	1.0	29.2
	1.0	0.892	1.0	2.684
	5.0	8.80	1.0	0.958
Modulation	0.1, 5.0	1.01	1.0	101.0
Difference frequency	0.1, 0.12	1.02	1.0	67.5
	1.0, 1.2	2.44	1.0	1.628
	5.0, 6.0	38.4	1.0	1.059

method, while the difference-frequency method gives very little indication of the presence of such distortion. On the other hand, for third-order low-frequency distortion, the difference-frequency method is more than twice as sensitive as the harmonic method, and the modulation method is more than three times as sensitive as the harmonic method. It would thus appear that either the modulation method or the harmonic method is most useful for investigating the possibility of nonlinear distortion which increases as the frequency is reduced. The modulation method gives additional weight to the third-order component, which is probably a desirable feature.

In the case of high-frequency nonlinear distortion all methods give essentially the same results for third-order distortion as for second-order distortion. The modulation method fails to detect nonlinear distortion of this type; the harmonic method indicates it to some extent, while the difference-frequency method shows a large increase in distortion factor. Since the harmonic method would be at even more of a disadvantage if it were applied to one of the large number of systems with a sharp high-frequency cutoff, the only satisfactory method of investigating the possibility of nonlinear distortion which increases as the frequency is increased appears to be the difference-frequency method. However, it should be noted that this method, as usually applied (measurement of the second-order component only), is not suitable for push-pull amplifiers and other systems with a symmetrical transfer characteristic.

EXPERIMENTAL RESULTS

In an effort to obtain verification of the conclusions of the preceding section, a distortion generator was constructed which, it was hoped, would provide nonlinear distortion corresponding to the special cases under consideration.

Nonlinear distortion was produced in this generator by over-driving an ordinary triode amplifier stage operated at a fairly low plate-supply voltage (about 100 volts). Additional circuits were added so that this amplifier could be operated either single-ended or push-pull. Provision was also made for introducing pre-emphasis before the nonlinear amplifier and de-emphasis following it, the circuits being resistance-capacitance networks with a nominal corner frequency of 1000 cps. This would then correspond to $f=1$ in the tables. The maximum amount of pre-emphasis was limited to 30 db rather than the unlimited amount assumed in the analytical approach.

In order to permit listening tests to be made on the distortion generator, it was constructed so that it was possible to switch rapidly from one form of distortion to another and also so that the amount of each type of distortion could be adjusted independently. A diagram of the complete generator is shown in Fig. 2.

Tests were made with the generator adjusted to produce the same amount of nonlinear distortion, as measured by the modulation method, at 2 volts input for the case of no pre-emphasis (both single-ended and push-pull) and for the single-ended condition with high-frequency pre-emphasis. For the case of low-frequency pre-emphasis, both single-ended and push-pull, it was impractical to reduce the nonlinear distortion to this level since the modulation method is extremely sensitive to this form of nonlinear distortion.

It should be noted that in all cases, the value of input signal given is the root-mean-square value of a single sinusoid having the same peak value as the actual input signal.

The analysis of the preceding section indicated that for nonlinear distortion, which is primarily second order, the percentage distortion should be proportional to the amplitude of the input signal. The experimental data, shown in Figs. 3–5, confirm this result for moderate values of distortion factor. In these figures, a straight line is drawn which most nearly approximates the experimental points for the second-order difference-frequency method, and then straight lines for the other methods are constructed from the relationships of Table I. These also agree quite well with the experimental data.

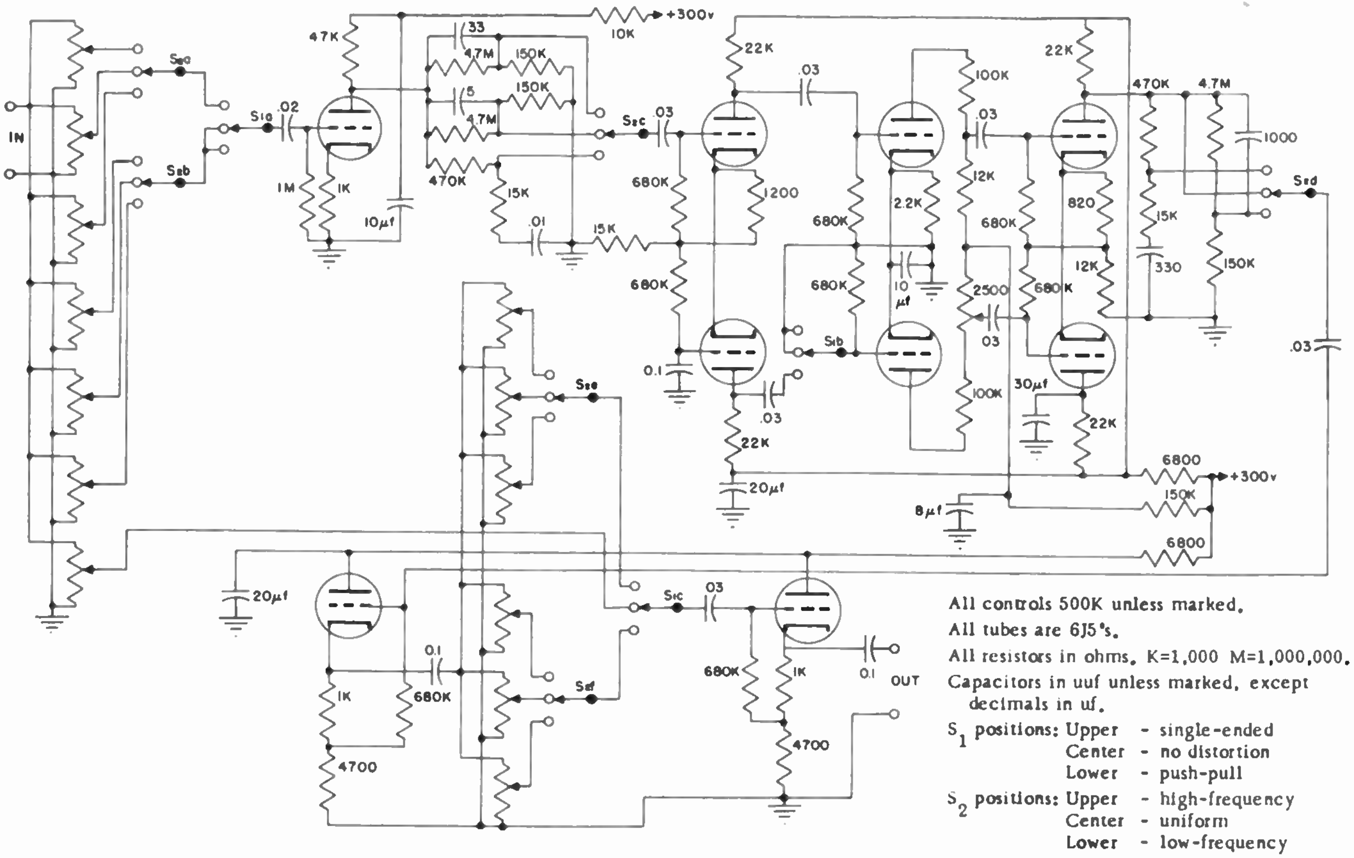

Fig. 2—Experimental distortion generator.

Push-pull operation, on the other hand, produced results which did not agree with the predictions based on the use of the third-order term only. The relationships of Table I indicate that the percentage of third-order distortion is proportional to the square of the magnitude of the input voltage, while the experimental data, plotted in Fig. 6, indicate that the distortion increases much more rapidly than this. Thus, a more careful consideration of the contribution of higher order terms is indicated.

In order to test the hypothesis that the higher order components were more significant in the push-pull case than in the single-ended case, the wave analyzer was used to measure individually the harmonics of a 1000-cps input signal. These results are plotted in Fig. 7.

Taking a 2.5-volt input as a typical value, a comparison of harmonic amplitudes from these figures with similar values obtained (but not plotted) from the single-ended amplifier was made. This comparison shows that in the push-pull case the fifth harmonic is approximately one quarter of the third harmonic, while in the single-ended case the third harmonic was only about 0.08 of the second and the fourth was less than 0.03 of the second. It is also interesting to observe that several of the curves turn downward as the input signal is increased, a result which is not predicted by the simple theory, but for which the presence of the higher order components would account.

In an actual system, of course, some of the coefficients in the series representation of the transfer characteristic would be of different sign, so that, for example, the third harmonic from the fifth-order term might partially cancel the third harmonic from the third-order term. Since the former is proportional to the fourth power of the amplitude of the input and the latter is proportional to the square of the amplitude, it is apparent how it is possible for the harmonics to decrease with increasing input amplitude in certain regions. Furthermore, this partial cancellation of nonlinear distortion components will occur to a different degree (at a particular input amplitude) depending on whether a harmonic or intermodulation component is being measured.

Obviously, then, a ratio between "second-order" components, measured by the harmonic, difference-frequency, and modulation methods and computed under the assumption that these components are due to an actual second-order term in the transfer characteristic, will be considerably in error if there is any appreciable contribution from a fourth-order term. Furthermore, the true second-order nonlinear distortion is directly proportional to input voltage, while the "second-order" nonlinear distortion due to the fourth-order term is proportional to the cube of the input voltage. That these effects will become of even more significance when terms higher than the fourth are present might also be expected.

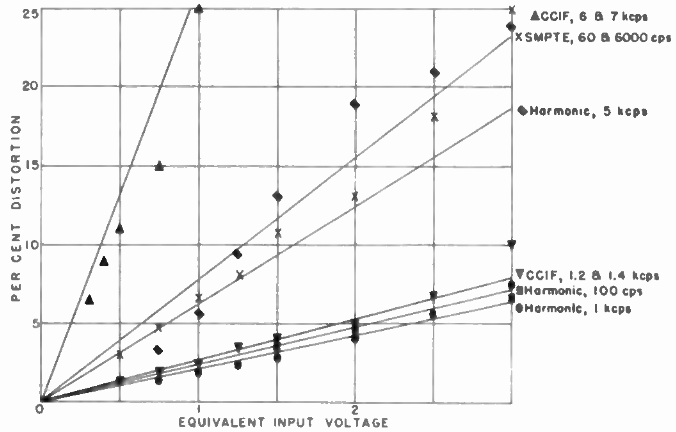

Fig. 3—Nonlinear (primarily second-order) distortion, involving high-frequency pre-emphasis. SMPTE—modulation method. CCIF—difference-frequency method.

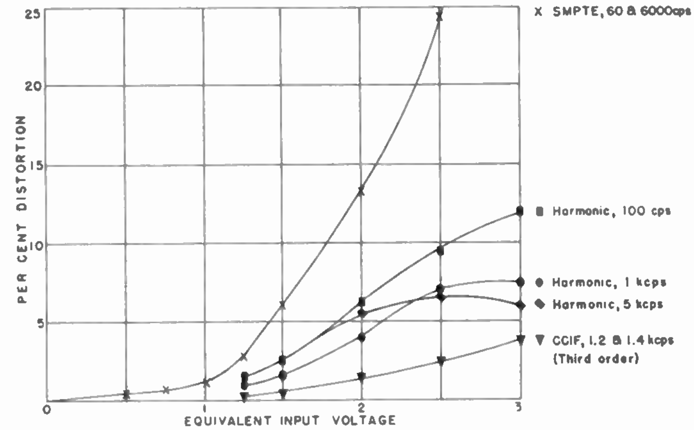

Fig. 6—Nonlinear (primarily odd-order) distortion, involving no pre-emphasis. SMPTE—modulation method. CCIF—difference-frequency method.

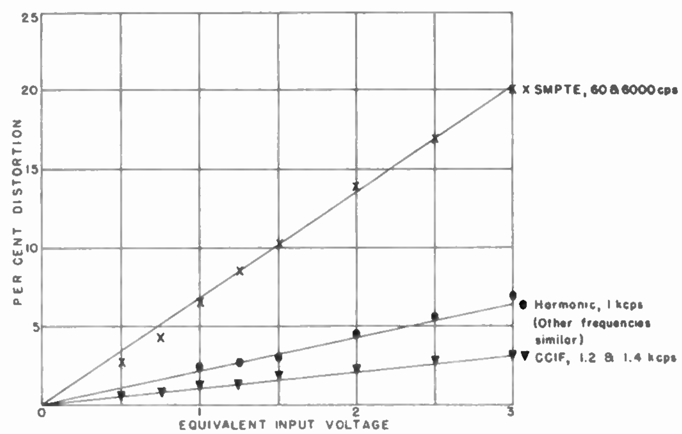

Fig. 4—Nonlinear (primarily second-order) distortion, involving no pre-emphasis. SMPTE—modulation method. CCIF—difference-frequency method.

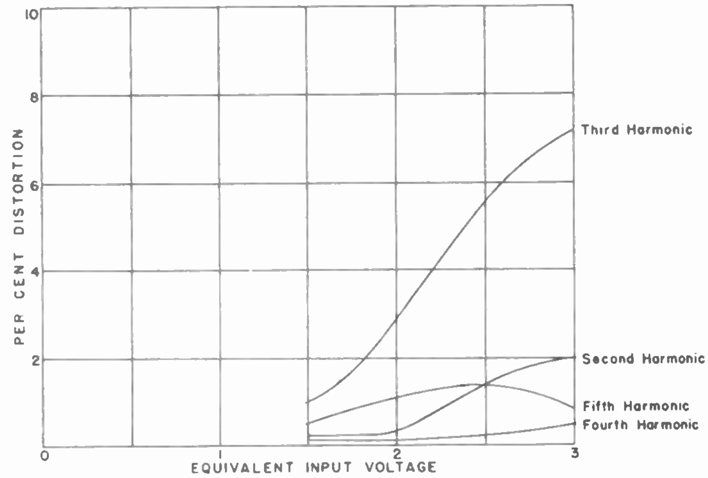

Fig. 7—Amplitudes of various components of the output from the distortion generator, push-pull operation. Input frequency 1000 cps.

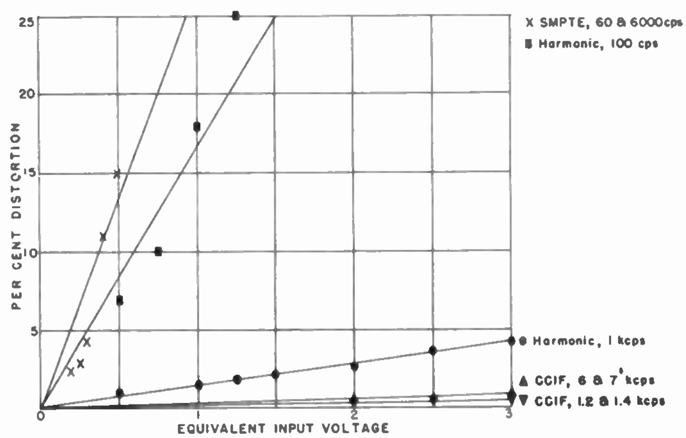

Fig. 5—Nonlinear (primarily second-order) distortion, involving low-frequency pre-emphasis. SMPTE—modulation method. CCIF—difference-frequency method.

The Choice of a Method for Measuring Nonlinear Distortion

It has already been mentioned that several of the common methods of measurement fail to indicate certain types of nonlinear distortion. In particular, the modulation method is quite insensitive to nonlinear distortion which only becomes significant at high frequencies, and the ordinary difference-frequency method, where only the component of frequency $\omega_2-\omega_1$ is measured, is insensitive to odd-order distortion. Although these limitations were apparent in the tests made on the distortion generator, tests were also made on several items of commercial equipment to see if the limitations were actually of consequence.

In addition to considerations based on the relative efficiencies of the various methods for detecting nonlinear distortion, there are also certain other considerations involved in the choice of a method for use. These include ease of operation of the necessary equipment, cost of the equipment, and suitability for use on a recording system which may introduce fluctuations in speed and therefore frequency.

Finally, and of major importance, is the correlation of electrically measured nonlinear distortion with subjective estimates of quality in various systems.

Considering all of these factors, it appears that the difference-frequency method would be very satisfactory if the use of a wave analyzer could be avoided and some convenient means provided to include at least the third-order component in the measurements. This can be accomplished by the simple expedient of using a low-pass filter following the system under test. However, it is necessary that the filter have a very sharp cutoff. For example, using test frequencies of 1000 and 1200 cps, the second-order component which is usually measured falls at 200 cps, while one of the third-order components is at 800 cps. Thus a filter is necessary which passes up to 800 cycles, while attenuating 1000 cycles and above by, say, 60 db. These are stringent requirements, but by no means impossible.

For purposes of test, such a filter was constructed with cutoff frequencies of 1000 cps and 5000 cps, using standard techniques. No attempt was made to investigate more sophisticated designs. The measured characteristics of this filter are shown in Fig. 8.

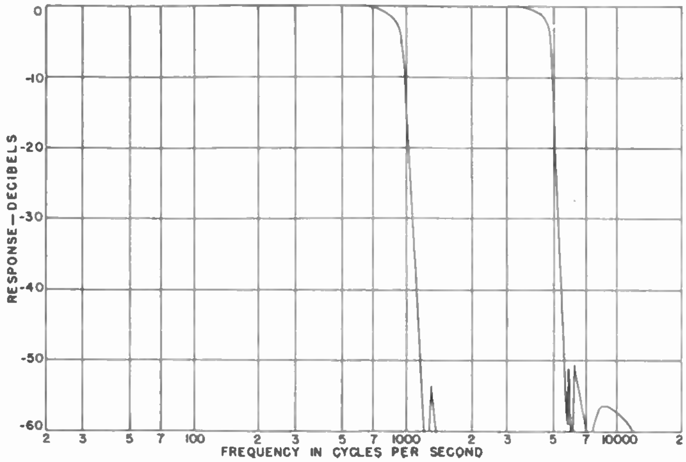

Fig. 8—Measured response-frequency characteristic of the sharp-cutoff low-pass filters.

Tests on Commercial Equipment

As an example of a system with high-frequency pre-emphasis, a high-quality magnetic tape recorder, a Magnecord model M90, was chosen. In addition to having nonlinear distortion which depends on the input frequency, such a machine might be expected to have primarily odd-order distortion. However, no attempt was made to adjust the electrical circuits for minimum even-order distortion. The amount of pre-emphasis which is provided is much greater at a tape speed of 7.5 inches per second than at 15 inches per second.

The test procedure consisted of recording 6000-cps and 7000-cps signals, each signal being set independently to give a reading of "0" on the recorder's vu meter. The tape was then played back and a wave analyzer used to measure the nonlinear distortion components at 1000 cps and 5000 cps.

In addition to this difference-frequency intermodulation test, a measurement was made using the modulation method with frequencies of 60 and 6000 cps and the same peak amplitude as used for the difference-frequency test.

At a tape speed of 15 inches per second, "second-order" difference-frequency distortion factor (at 1000 cps) was 0.42 per cent, "third-order" distortion factor (at 5000 cps) was 1.5 per cent, a low-pass difference-frequency measurement indicated 1.3 per cent, and the modulation method indicated 12 per cent. It is apparent that the symmetrical transfer characteristic of the system has considerably reduced the effectiveness of the ordinary second-order difference-frequency measurement, but that the amount of pre-emphasis is not sufficient to affect the modulation-method measurement to any great extent.

On the other hand, at a tape speed of 7.5 inches per second the "second-order" difference-frequency distortion factor was about 7.5 per cent, the "third-order" distortion factor was about 11 per cent, the low-pass difference-frequency distortion factor was 7.6 per cent, and the modulation-method distortion factor was only 9 per cent. This represents a very great loss in effectiveness for the modulation method. The difference-frequency measurements using the wave analyzer were particularly difficult to make at the slower tape speed. At both tape speeds, the second harmonic of the 6000-cps signal was less than 0.5 per cent, indicating the uselessness of the harmonic method for measuring high-frequency nonlinear distortion on this type of system.

Tests of a typical "high-fidelity" power amplifier emphasized even further the folly of using the ordinary difference-frequency method on a system with symmetrical distortion. This amplifier used push-pull 807 tubes in an "ultra-linear" output stage with about 20 db of inverse feedback over the entire amplifier. Such an amplifier might be expected to show a very pronounced corner in its transfer characteristic because of the large amount of feedback and also comparatively small even-order distortion because of the push-pull operation.

Again using frequencies of 6 and 7 kc per second, each one alone producing an output power of 10 watts, the low-pass difference-frequency distortion factor was 2 per cent while the second-order component alone measured only 0.02 per cent! The clipping was quite obvious on an oscilloscope, and a measurement using the modulation method indicated a distortion factor of 32 per cent.

Listening Tests

The remaining question which must be answered before a choice of a method of measuring distortion may be made concerns the correlation between measured distortion factors and audible quality. Unfortunately, since this is probably the most important consideration of all, it is also the most difficult to investigate. A completely satisfactory answer requires elaborate, carefully controlled experiments, far beyond the time or facilities available for this investigation. However, it was possible to make some listening tests with the distortion generator, and the results to be described represent the consensus of about ten persons listening to various samples of recorded music, although they should not be considered conclusive.

In the comparison of results obtained by listening tests and steady-state measurements, there is a fundamental difficulty involved—that of specifying the operating level of the program material. As mentioned previously, the uniform-response single-ended condition, the single-ended condition with high-frequency pre-emphasis, and the uniform-response push-pull condition were all adjusted to have the same distortion factor, measured by the modulation method at an equivalent sinusoidal input of 2 volts rms. At higher input levels, the two single-ended conditions still had nearly equal (but of course much higher) distortion factors while the push-pull condition produced distortion considerably greater than the two single-ended conditions. At input levels below 2 volts, the two single-ended conditions again gave nearly equal distortion factors while the push-pull distortion factor was much smaller. This makes a comparison between single-ended and push-pull operation a very difficult matter since the distortion factors are the same only at one particular input level.

The audio industry has standardized on the vu meter for the measurement of program levels, so a meter of this type was selected for the listening tests. The program level of the music was adjusted to give a "0" reading on the meter read in the standard manner.[13] By way of comparison, a sinusoidal signal set to this same level produced 1.85 volts rms at the input to the distortion generator, slightly less than the value for which the various distortion factors were adjusted to be equal.

(It was impossible to reduce the distortion factor of the single-ended condition with low-frequency pre-emphasis enough even to approximate the distortion factors of the other conditions, and the push-pull condition with low-frequency pre-emphasis could be reduced only to a distortion factor of about 29 per cent at 2 volts input. These conditions were included in the listening tests, nevertheless.)

Comparing the uniform-response single-ended operation and the single-ended operation with high-frequency pre-emphasis, the listeners' preference seemed to depend considerably on the type of program material. On music which contained a relatively large proportion of high-frequency components, there was a noticeable "muddiness" with the high-frequency pre-emphasis. Also, many persons commented that the system seemed to have poor high-frequency response, apparently due to the fact that the distortion generator was effectively clipping the high-frequency components from the signal. However, on material which did not contain much high-frequency energy the listeners' preference was definitely in favor of the high-frequency pre-emphasis condition over the uniform-response condition. This is rather surprising in view of the fact that both conditions have essentially the same amount of nonlinear distortion at the mid-range frequencies. One possible explanation is that the high-frequency de-emphasis, included to give the over-all system a uniform response to all frequencies, effectively attenuates sum-frequency distortion components resulting from the intermodulation of two low-frequency tones in the input.

At the program level used, there seemed to be little question that the condition of uniform-response and push-pull operation sounded poorer than the single-ended conditions. However, the case of push-pull operation with low-frequency pre-emphasis sounded better than either of the single-ended conditions, in spite of a much higher distortion factor at the selected input level.

In the literature[4,8,12] much has been made of the supposedly detrimental effects of nonlinear distortion which increases at high frequencies. This is presumably because two high-frequency components in the input signal may produce a beat tone at a much lower frequency which falls in the range of the greatest sensitivity of the ear. Since high-frequency tones are not particularly effective in masking low-frequency tones, even a very weak note would be audible, according to this theory.

On the other hand, in typical music the energy at high frequencies is much less than at low frequencies,[14] so that the difference-frequency beat tone produced by high-frequency intermodulation will be much weaker than the sum-frequency beat tone produced by intermodulation of the lower frequency components. This difference might easily be sufficient to more than offset the difference in masking effect in the two cases. Furthermore, it is likely that other tones are present in the music at the same time, so that the question of masking is not of particular importance.

The form of nonlinear distortion which is actually more objectionable, then, will depend on the severity of the increase in nonlinear distortion at high frequencies as well as on the energy distribution in the particular signal being considered. Obviously, the safest method is to keep all forms of nonlinear distortion to an absolute minimum, and so a method of measurement which will detect high-frequency nonlinear distortion would be desirable.

Besides the inadequacy of the modulation method in indicating the high-frequency nonlinear distortion, this method also gives an indication of excessive distortion factor when the nonlinear distortion is primarily at low frequencies. This does not correspond to the impression received by the ear. However, the modulation method is extremely well suited to measuring nonlinear distortion which is not combined with frequency distortion.

[13] American Standards Assn., "American Recommended Practice for Volume Measurement of Electrical Speech and Program Waves," New York, N. Y.; 1942.

[14] "Frequency Range and Power Considerations in Music Reproduction," Jensen Radio Mfg. Co., Chicago, Ill.; 1944. (See Fig. 6.)

Conclusions

If it is necessary to select only one "general purpose" method of nonlinear distortion measurement, the modulation method appears to be the logical choice. Not only is it the quickest and most economical, but the results seem to agree fairly well with audible impressions.

The harmonic method provides useful information if measurements are made at several different frequencies and if too much faith is not placed in measurements made at frequencies higher than one third of the upper frequency limit of a system.

The difference-frequency method, using only the term of frequency $\omega_2-\omega_1$, may be grossly misleading for a system with a symmetrical transfer characteristic. However, this method, or the low-pass version of it, is the only really satisfactory means for discovering nonlinear distortion near the upper frequency limit of a system.

When a rather thorough investigation of distortion is required, such as on a new system design or a strange piece of equipment, the best approach would seem to be a measurement by the modulation method, a measurement by the low-pass difference frequency method at about 1000 cps to see if the modulation-method measurement is possibly being unduly influenced by low-frequency nonlinear distortion, and a low-pass difference-frequency measurement at about 5000 cps to investigate the possibility of high-frequency nonlinear distortion. These frequencies could be modified appropriately for systems which cover somewhat different frequency ranges.

Measurements should be made at several input levels, and a plot of the distortion factor vs input will indicate to some extent whether the distortion is primarily second-order or is due to higher order terms. In evaluating the results allowance must be made for the fact that the modulation method gives a numerical value several (5 or 6) times the difference-frequency results.

In all cases, it is wise to supplement the distortion-measurement with oscilloscopic observation of the output waveform. Not only will this procedure give some indication of the shape of the transfer characteristic, but it will also reveal other defects, such as instability.

Since some of the items of test equipment required for performing modulation-method tests are the same as required for the low-pass difference-frequency method, it may be possible to combine these items in one piece of equipment which could perform both types of measurement. The following Appendix describes how such an instrument might be constructed.

It is further recommended in making tests of *parts of complete systems* that external equalization be introduced to compensate for any equalization incorporated in the equipment under test.

There does not seem to be any basis for conclusions regarding the relative audible effects of second and higher order nonlinear distortion, for the reason that it is impossible to compare the magnitudes of such distortion except under steady-state conditions. The results of this investigation support the observations of Maxwell rather than Peterson; that is, that surprisingly large amounts of system nonlinearity can be tolerated at both high and low ends of the spectrum.

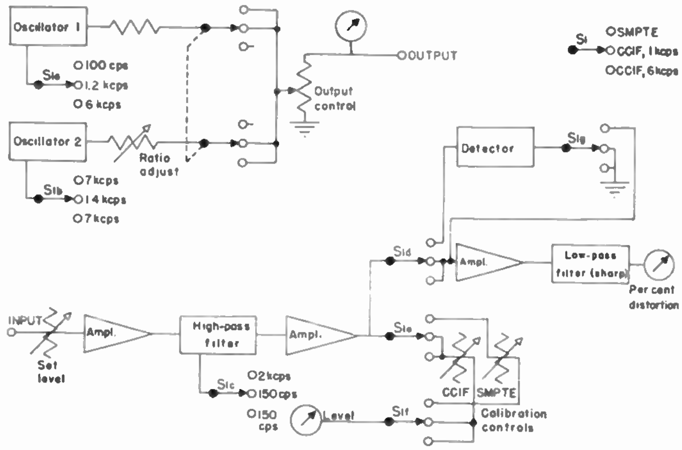

Fig. 9—Block diagram of a "universal" distortion analyzer. CCIF—difference-frequency method. SMPTE—modulation method.

Appendix

A Distortion Analyzer for both Difference-Frequency and Modulation Measurements

The modulation method of nonlinear distortion measurement requires two oscillators, a high-pass filter, a low-pass filter, a detector, and a voltmeter. The low-pass difference-frequency method described above requires two oscillators, a sharp-cutoff low-pass filter, and a voltmeter. In addition a high-pass filter is useful for eliminating hum which might be present. Since there is so much duplication of elements with these two systems, the possibility of combining them in one instrument is worth considering.

The oscillators used in both methods need not have exceptionally pure waveform, so that a simple oscillator circuit will suffice, with switching to change frequency for the different methods. Means for small adjustments in frequency for those frequencies used with the low-pass difference-frequency method should be provided. These would need only infrequent resetting.

The high-pass filter used with the modulation method may be a simple type, with a cutoff of approximately 2000 cps. The same type of filter with the cutoff shifted to about 150 cps will serve adequately for reducing the effects of hum on low-pass difference-frequency measurements.

The low-pass filter, on the other hand, is a very critical part of the low-pass difference-frequency method. However, this same filter may also be used in place of the simple low-pass filter usually used for modulation-method measurements.

The modulation-method detector may simply be switched out of the circuit when low-pass difference-frequency measurements are to be made, and the voltmeter is a standard element which may be common to both methods.

A block diagram showing the essential switching involved in such a "universal" instrument is shown in Fig. 9. For simplicity, separate voltmeters are shown for oscillator metering, setting input level, and reading distortion factors, functions which may be performed by switching the same meter.

ACKNOWLEDGMENT

The work described above was a part of a dissertation submitted to the Graduate School of Northwestern University in partial fulfillment of the degree, doctor of philosophy. It was performed under the supervision of Prof. R. E. Beam.

Modulation Noise in Two-Channel Disk Recordings*

DANIEL CRONIN†

Summary—**Part of a modulation-noise problem encountered in two-channel stereo disk recording is shown to have its source not in the recording channel but in the geometry of the record-playback system. A relation is given to show the order of magnitude of this effect and some of the possible means of improvement are indicated.**

An occasional subjective reaction to current stereo disk recordings is that they are more "hissy" than their monaural equivalents. On careful examination, this hiss often turns out to be largely modulation noise and can be shown to have an origin in the geometry of the recording-playback system.

For convenience, the system will be analyzed in terms of its lateral and vertical components which may be regarded as the sum and difference, respectively, of the "left" and "right" channels of the 45-45 system.

Referring to Fig. 1, all recording stylus motions, vertical, lateral, 45-45, or anything intermediate, take place in a plane through point S which is perpendicular to the line P-S. This imaginary plane is inclined from the vertical by the "tilt angle" ϕ.

It is presumably well known that a lack of correspondence between the recording and playback tilt angles ϕ_1 and ϕ_2 will lead to a minor loss in playback level of the vertical component and a perhaps more serious harmonic distortion production which will appear in the lateral channel.

Less attention has been devoted to the effect of the component of stylus motion which is in line with the centerline of the groove. Velocities in this direction will alternately add to and subtract from the average groove velocity. A playback cartridge with a tilt angle of zero (perfectly vertical sensitivity) would find the groove velocity modulated by $V_v \tan \phi_1$ where V_v is the velocity of vertical modulation.

If identical frequency components should be present in the two channels, only harmonic and intermodulation distortion would result. In the more general case, where

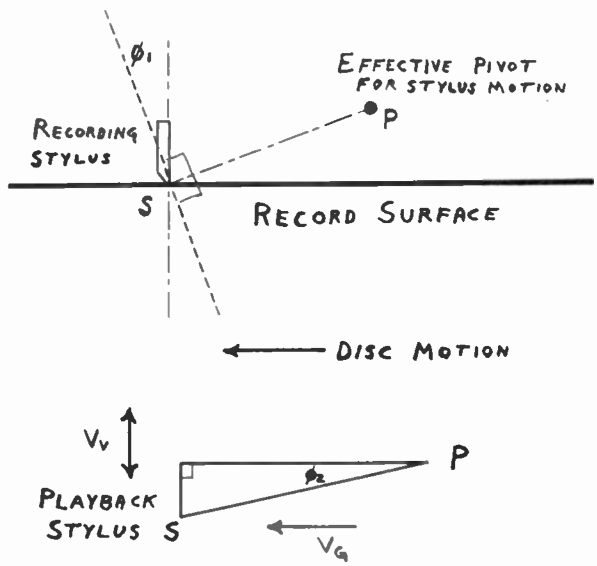

Fig. 1—Record-playback geometry.

they are dissimilar, the frequency modulation of the lateral component by the vertical would be essentially random and would result in modulation noise. For small values of $\phi_1 - \phi_2$, this noise output will be determined by the angular error, the vertical velocity, and the lateral "loudness" (RIAA de-emphasized velocity).

The percentage of frequency modulation will be

$$\text{FM (per cent)} = \frac{V_v(\tan \phi_1 - \tan \phi_2)}{V_g} \times 100 \qquad (1)$$

where V_g is the groove velocity.

Groove velocity of course is a function of the record diameter but can be eliminated if we can assume a ratio of V_v to V_g. Taking the heel angle of the cutting stylus as 45° (which is usual), the vertical velocity should not exceed the groove velocity as an upper limit since that would result in heel contact on downward excursions and the stylus would be embossing or deforming the lacquer rather than cutting it. On the other hand, there are commercial pressures tending to keep the recorded

* Manuscript received by the PGA, October 28, 1958; revised manuscript received November 26, 1958.
† Chief Engineer, Bell Sound Studies, Inc., New York 36, N. Y.

velocities as high as possible. If we assume the available velocity is 100 per cent used at the inner diameter and the same level is recorded at the outermost diameter, then our peak vertical velocity at the outer diameter (where we have the best chance of getting high-quality reproduction) will be 50 per cent of the groove velocity at that diameter.

If the tilt angles are in the order of 30° or less (manufacturers indicate that ϕ_1 is between 20° and 30°), the rough approximation:

$$\tan \phi_1 - \tan \phi_2 \cong \tan (\phi_1 - \phi_2)$$

is close enough for our purposes. Another useful approximation in this range is

$$100 \tan \phi \cong 2\phi,$$

where ϕ is in degrees.

Substituting into (1), we arrive at the rather convenient relation that

$$\text{FM (per cent peak)} = \phi_1 - \phi_2 \quad (\phi \text{ is in degrees}).$$

The implications of this are the same as for Corrington's data[1] and are in addition to the case analyzed by Corrington. Some observers have felt that the consumer might not notice the harmonic and intermodulation distortions, but most people will notice noise when they hear it.

Obviously, ϕ_2 should be the same as ϕ_1, but this will be very difficult to achieve in a changer mechanism where the records keep lifting the arm up to higher angles with each successive record. Warped records and unusual variations in record thickness might also be disturbing in those tone arms whose vertical pivot is far forward on the arm, and it might be preferable to have the vertical pivot as far removed from the stylus as possible to minimize variations in ϕ_2.

There is still another problem in specifying ϕ_1: there is little indication that the different manufacturers are getting together on a single specification for this. There is also some indication that this angle is not a constant for any one recorder but may be a function of frequency, the effective pivot point tending to move closer to the stylus for higher frequencies. Since the highest velocities are usually recorded in the upper half of the frequency spectrum, a weighted average ϕ_1 should be taken, allowing a heavier weighting to the higher frequencies.

In so far as possible, the ϕ_1 agreed on should be as low as possible since the harm caused by a given angular error is directly proportional to ϕ.

[1] M. S. Corrington and T. Murakami, "Tracing distortion in stereophonic disc recording," 1958 IRE National Convention Record, pt. 7, pp. 73–81.

Electromagnetic Efficiency of Heads in Magnetic Recording*

MARVIN CAMRAS†

Summary—**A simple method is described for obtaining a figure of merit for response of magnetic recorder due to core losses and electrical characteristics.**

FREQUENCY response of a head is determined by: 1) mechanical factors such as gap size, straightness, alignment contact with the tape, etc., and 2) electromagnetic factors such as eddy current and hysteresis losses in the core, resistance and dielectric losses in the coil, resonances, etc.

Ordinary response curves from a tape give an over-all picture which does not tell which factors are important and which are minor or negligible. We can eliminate the mechanical factors by using a small external probe to introduce flux into a head to be studied. The response determined in this manner measures the electromagnetic efficiency of the design. Mechanical factors, if desired, are then determined by "difference" from the over-all response using a tape record source.

In the tests the probe and circuitry shown in Fig. 1 is convenient. It is an air core coil with 10 turns of no. 50 Formex wire, 0.060 inch by 0.025 inch by 0.010 inch in size, energized with 70 ma of current held constant for all frequencies from 30 to 20,000 cps. The induced voltage in a perfect head rises proportional to frequency, so that a 6 db per octave correction was applied in all cases to give comparisons with a flat response curve.

A number of heads was tested by this method. The results, arranged in order from best to poorest response, are given in Table I.

Column 2 of Table I shows that high impedance heads, designed to operate directly into the amplifier grid, supply about 2 to 11 mv. Low impedance heads

* Manuscript received by the PGA, October 28, 1958. Presented at the Convention of the Audio Engineering Society, New York, N. Y., October 1, 1958.
† Armour Res. Found., Ill. Inst. Tech., Chicago 16, Ill.

giving about 0.5 mv should be used with matching transformers, which will step up their output voltage. Some heads, such as B, actually have a slight rise at 10 and 20 kc due to resonance.

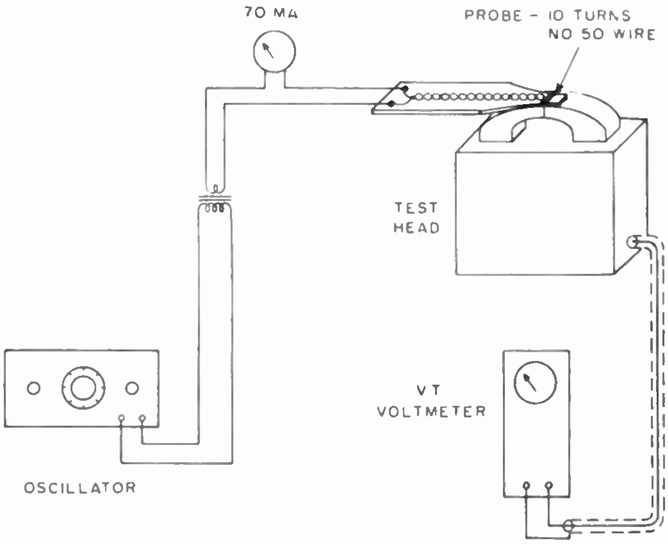

Fig. 1—Probe used for head tests.

The above results lead to interesting conclusions:

1) Ferrite heads have negligible core loss at audio frequencies. Their frequency response is a function of gap and contact effects.
2) Heads with cores built of 0.006-inch laminations have negligible loss at 10 kc and a slight loss at 20 kc. Loss at erase and bias frequencies is appreciable, but not serious. As with ferrite heads, their response is determined mainly by gap and other mechanical factors.
3) "Single lamination" heads of the type used in many home recorders have important losses of about 7 db at 10 kc, and about 11 db at 20 kc. These may be more important than gap effects. Since the loss occurs on record and again on playback, the over-all response of a typical head is 14 db down at 10 kc, and 22 db down at 20 kc as a result of core loss alone.
4) Single lamination heads of the type where the gap is maintained by soldering can be improved by using spot welded construction and thinner core material. Imbedding in solder should particularly be avoided.

TABLE I

Head	Output at 1 kc millivolts	Relative output at 1 kc, db	Relative output at 10 kc, db	Relative output at 20 kc, db	Remarks
A	0.45 mv	0	0.0 db	0.0 db	Experimental head with core made of Ferrite.
B	3.4	0	0.5*	1.30*	¼-inch wide, made with a stack of 32 laminations. Used on a high-grade professional recorder.
C	2.1	0	0.0	−0.6	0.190-inch wide, stack of 24 laminations. Commercial design no longer available.
D	11.0	0	0.2*	−2.2	0.93-inch wide, stack of about 15 laminations. Widely used high-quality head.
E	3.2	0	−0.7	−2.8	0.050-inch wide, stack of 8 laminations for multichannel operations. Same manufacturer as head D.
F	5.5	0	−4.0	−8.0	"Single lamination" no solder or welds used at gap. Experimental head.
G	10.0	0	−5.5	−8.5	"Single lamination" experimental spot welded construction.
H	0.56	0	−4.5	−9.3	"Single lamination." Method of fastening not determined. Used on a European make of tape recorder.
I	0.47	0	−7.0	−11.2	"Single lamination" half-track (stereo) head, soldered construction. Used on a medium-priced professional recorder.
J	3.0	0	−8.2	−10.4	"Single lamination" lapped-clamped construction. This head is popular for home recorders.
K	0.44	0	−9.0	−12.0	"Single lamination" full-track playback head, soldered construction. Used on a medium-priced professional recorder.
L	6.3	0	−14.0	−18.5	"Single lamination" half-track soldered construction. Used on a home recorder.

* Rise due to resonance.

Correspondence

Compatible Method of Recording and Reproduction of Stereo Sound*

While interesting, the method for recording stereophonic sound, suggested by Lamberty,[1] has several weaknesses which must be considered:

1) Far from eliminating the "complicated and critical equipment" necessary to demodulate the FM subcarrier type of stereo disk, the method proposed by Lamberty requires two low-pass filters, a high-pass filter, and a band-pass filter, all of which must be rather sharp and therefore consist of several sections each. Such filtering becomes somewhat unwieldy.

2) The high-pass filter in the FM subcarrier system is not too critical since the demodulator, if properly designed, tends to discriminate against the amplitude variations in the main channel. Since this type of FM demodulator with its attendant limiting is absent in Lamberty's system, the filters must provide all of the channel separation desired.

3) Since the oscillator frequency is to be recorded on the disk, all intermodulation distortion must be eliminated to prevent this signal from interfering with both the main channel and the second channel. Since intermodulation distortion is present in relatively large amounts in all recording and playback heads and cartridges, a certain amount of interference must be expected.

4) Since the second channel has a very narrow frequency response as compared to the main channel, a severe unbalance will be heard between the two channels, and the stereo illusion will therefore be largely destroyed.

5) This system is equally sensitive to signal and noise, so the second channel has to be recorded at a high level in order to obtain an acceptable signal-to-noise ratio. However, since most, if not all, playback cartridges are incapable of reproducing properly the ultrasonic range, especially at the high velocities which would have to be used to achieve a high snr, a playback problem results with more distortion and possible ruining of the disk as by-products. The FM subcarrier method of stereo disk recording makes it possible, because of the very nature of frequency modulation, to attain acceptable snr even at very low recording level.

6) Quite often the stylus "skips" a portion of a high-frequency signal recorded on the disk by sliding and bouncing over it. In the system suggested by Lamberty, which after all is highly sensitive to amplitude variations, this skipping will be highly objectionable and distracting. A system based on frequency modulation is sufficiently insensitive to amplitude variations to minimize this type of distortion.

7) The stereo recording system suggested by Lamberty requires higher quality equipment than the FM system. For example, to achieve a response up to 15,000 cps in both channels, the former system requires equipment substantially *flat* up to about 32,000 cycles. Such equipment is as yet hard to find. On the other hand, the FM subcarrier system, to achieve approximately the same response, needs equipment to reach only about 28,000 cycles *but* it need not be flat up to that frequency. A certain amount of treble loss and possible peaking are permissible since the amplitude variations with frequency are not detected by an FM detector. For example, the peak above the audio range often present in cartridges due to needle resonances will appear only in the system suggested by Lamberty, and not in the FM system, as a peak in the response of one channel.

Needless to say, any such distortion and deterioration of frequency response as described above is highly undesirable. It is the writer's opinion that no stereo recording system should be adopted which is not capable of quality at least comparable to present monophonic reproduction. The presently accepted 45°–45° orthogonal groove stereo disk seems fully capable of eventually achieving that goal at a reasonable price.

PETER A. STARK
New York 28, N. Y.

* Received by the PGA, November 17, 1958.

[1] B. J. Lamberty, IRE TRANS. ON AUDIO, vol. AU-6, pp. 89–90; July–August, 1958.

Contributors

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Index

to

IRE Transactions

on

Audio

Volume Au-6, 1958

IRE Professional Group on Audio Combined Index for 1958

Compiled by D. W. Martin

Use of the Index

The combined indexes which follow are of three types.

The first type is simply a compilation of Tables of Contents of IRE TRANSACTIONS ON AUDIO and the Audio portion of the CONVENTION RECORD.

Next is an Author Index.

The third type is an Analytic Subject Index.

The volume system of indexing began with the first issue of 1953. Volume AU-6, issues 1, 2, · · · 6, contain all IRE-PGA publications for the year 1958 and the Audio part of Section 7 of the 1958 CONVENTION RECORD. In the Tables of Contents, this section is numbered CR-6-7, and inserted chronologically between issues AU-6-4 and AU-6-5, IRE TRANSACTIONS ON AUDIO.

In both the Author Index and the Analytic Subject Index references are made to publications by the issue and page number in the volume (*e.g.*, AU-6-2, 32).

The Analytic Subject Index lists titles under the appropriate classifications in the series shown below. In many cases valuable material is published under titles which cannot be fully descriptive of all the material which the paper covers. It is for this reason that some titles are listed under various classifications, some of which may seem inappropriate to the title itself. It is hoped that this will increase the probability of finding quickly most of the information available under a particular classification.

Classification of Subjects

1. IRE-PGA
   - 1.1 General
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   - 1.3 National and Regional Meetings
   - 1.4 Chapters
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   - 1.6 Transactions
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2. Bibliographies, Reviews, Standards, Tapescripts
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3. Sound Systems
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   - 3.2 Stereophonic, Binaural, and Spatial Effects
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4. Microphones
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5. Amplifiers
   - 5.1 General
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   - 6.4 Horns, Enclosures, Baffles
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7. Disk Recording and Reproduction
   - 7.1 General
   - 7.2 Disks
   - 7.3 Recording
   - 7.4 Pickups and Tone Arms
   - 7.5 Pre-emphasis and Postequalization
   - 7.9 Special Mechanical Recorders
8. Magnetic Recording and Reproduction
   - 8.1 General
   - 8.2 Tape and Wire
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value falling between 1 and 2 meters. We shall adopt a distance of 2 meters between the measuring microphone and the baffle of the loudspeaker, for cone types, or the plane of the horn mouth, for horn types.

B. Instrumentation

It does not come within the scope of this paper to describe in detail the electroacoustic instrumentation necessary to perform the tests which are mentioned in the next section; we shall limit ourselves simply to a few characteristics that it is desirable to impose on the instrumentation to be used, so as to confer to the results a sure scientific value.

1) Electrical Sources:

a) Oscillator: It must give, by continuous variations, all the frequencies between 20 and 15,000 cps with a sufficient precision (less than 2 per cent error), and with a harmonic distortion of less than 0.5 per cent. Between the same (frequency) limits, its calibration curve must present variations of less than ±0.5 db about the mean value. The marking of the dial should conform with the recommendations of the CCIF, that is equilinear from 0 to 100 cps, and logarithmic above that frequency.

b) Power amplifier: If the power available from the above instruments is not sufficient to test the loudspeakers under normal conditions, it is naturally necessary to utilize a power amplifier whose response curve should be linear within ±2 db between 20 and 15,000 cps, and whose harmonic distortion seems to have to conform approximately to the specifications given in Table I, showing the results that we have obtained on an instrument of good quality. Its noise level should naturally be as low as possible, and the gain adjustment should be continuous.

TABLE I

Frequency cps	Harmonic distortion in per cent for several output powers			
	1 watt	2 watts	5 watts	10 watts
40	2.5	3	3.5	4
60	2	2.5	3	3
100	1.5	1.5	2.5	3
400	1.5	1.5	2.5	2.5
1000	1.5	1.5	2.5	2.5
5000	1	1.5	2.5	2.5

c) Microphone and microphone preamplifier: The microphone used for the tests must be as small as possible and have a well defined geometrical form (preferably spherical), so as to allow the calculation of diffraction corrections, the theory of which was originated by Lord Rayleigh.[8] The over-all response curve of the microphone and preamplifier must be linear within ±2 db between 20 and 15,000 cps and its nonlinear distortion must be very low (less than 0.5 per cent).

d) Recorder: The logarithmic recorder should be advocated, but the use of a linear instrument, or a cathode ray oscillograph, for example, remain naturally possible. In any event, the over-all time constant must be reduced to a minimum. It is also desirable that the recorder permits the measurement of level variations of at least 40 db, without the necessity to readjust the sensitivity.

2) Instrumentation for the Measurement of Distortion ("Harmonic") and Intermodulation: Harmonic distortion measurements can be made with a distortion meter or a ("wave") analyzer. Recently, however, a very interesting automatic method of measurement was developed in the U. S.; it is based on the use of a series of 14 high-pass filters highly selective which cut off the fundamental in the microphone circuit.[9]

With respect to the measurement of intermodulation, it seems preferable to utilize a wave analyzer.

III. Performance Characteristics to be Measured

To control the acoustic quality ("performance") of a loudspeaker by objective measurements, several tests must be performed, each one showing a certain quality or deficiency of the equipment under study.

The steady-state response curve as a function of frequency was one of the first characteristics to be studied, but it was soon found that, although necessary, this was insufficient. Therefore, even before World War II, in order to have a pretty good idea of loudspeaker properties, the German Committee on Acoustics advocated five tests permitting the definition of the following elements:

a) Steady-state response curve as a function of frequency.
b) Harmonic distortion.
c) Directivity.
d) Efficiency.
e) Maximum power handling.

More recently, certain test methods were standardized in the U. S.[10] These consisted of the first four tests enumerated above, but added the measurement of impedance as a function of frequency, and the maximum power handling as indicated by the loudspeaker manufacturer. It is then necessary to specify the exact definition of its "maximum power handling," which seems to be what people now tend to call the nominal power of the equipment under test; it is therefore sufficient to verify whether it conforms to the manufacturer's rating (see project of standard specification CCTV No. 267:

[8] Lord Rayleigh, "On the bending of sound waves on a spherical obstacle," *Proc. Roy. Soc. London;* 1903.

[9] H. F. Holson and D. F. Pennie, "An automatic nonlinear distortion analyzer," *RCA Rev.*, vol. 12, pp. 35–44; March, 1951.

[10] "American recommended practice for loud speaker testing," *Amer. Standards*, no. C; November 6, 1942.

the nominal power is the electrical power dissipated between the input terminals of the loudspeaker, at 400 cps, for which the harmonic distortion reaches the value of 3 per cent).

The present tendency, which affirms itself more and more, is to add to the five tests specified above, a test of transient response, for which the test methods are still quite varied.[11] Therefore, to summarize, the list of information to be obtained would concern itself with the following six points:

a) Frequency response characteristic (steady state).
b) Directivity.
c) Harmonic and intermodulation distortion.
d) Impedance characteristic.
e) Efficiency.
f) Transient response.

For each of these tests, we shall give a few details with respect to the methods, and the variable parameters of the measurement, but without formulating imperative clauses, since this work is not a specification, but there are elements that are specified, in each particular case, as a function of the tests described below.

A. *Frequency Response Characteristic*

This test is the one that was the first to be studied by various authors[12,13] and which now is still the most frequently performed. The absolute response of a loudspeaker, at a given frequency and at a given point, is defined as being the relation between the acoustic pressure P at that point and frequency, and the square root of the electric power W applied at its input terminals:

$$\text{Absolute response} = \frac{P}{\sqrt{W}} \quad \frac{P\sqrt{Z}}{E}$$

if Z is the impedance, and E the voltage across the (voice coil) terminals since $W = E^2/Z$. To obtain the absolute response as a function of frequency, it is necessary to supply the loudspeaker under test with a constant electric power, which is not possible, because the impedance Z varies appreciably with frequency. Therefore, it is necessary to calculate $\sqrt{Z}/E$ for each frequency and multiply it by the measured acoustic pressure, in order to plot the characteristic of absolute response as a function of frequency. In practice, this absolute characteristic is obtained while maintaining a constant voltage across the loudspeaker voice coil; it is defined at each frequency by the variation of acoustic pressure measured at a certain point, and expressed in decibels re $p_0 = 2 \times 10^{-4}$ baryes (microbars):

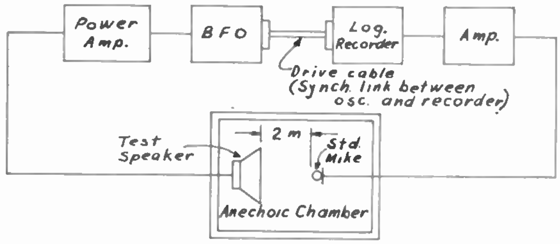


Fig. 5—Block diagram of the setup generally used to obtain response characteristics of loudspeakers.

$$N_{\text{db}} = 20 \log\left(\frac{p}{p_0}\right) = 20 \log\left(\frac{p \times 10^4}{2}\right).$$

Several methods can be used to obtain this characteristic:

1) Point by point method,
2) Semiautomatic method using pure sine waves, or
3) Entirely automatic method using: a) pure sine waves, or b) complex noise having a continuous and uniform spectrum (white noise).

The two first methods are now almost obsolete and only the third is generally used:

a) Use of pure frequencies. It is recommended to utilize a continuously variable oscillator and to obtain the characteristic for the 25–15,000-cps band. The variation in frequency must be done at constant speed (3 to 4 minutes are necessary to cover the 25–15,000-cps band),[14] and the rotation of the oscillator variable condenser must *track* with the paper chart.

The microphone must be placed on the loudspeaker axis under the general test conditions specified in the preceding section. Fig. 5 shows the block diagram of a setup currently utilized for this purpose, and Fig. 6 shows the frequency response characteristics of several loudspeakers obtained in this manner. The method is the one most used at present; it gives the most accurate results, but unfortunately requires the use of an anechoic chamber, because free-field measurements are inconvenient, and conditional upon too many factors (wind, rain, noise, etc.).

b) Use of white noise. This method had been utilized more than ten years ago,[15–17] and more recently,[18] but it is more ticklish and less accurate than the preceding method. Its only advantage is that, in principle, it per-

[11] L. L. Beranek, "Acoustic Measurements," Chapman and Hall, Ltd., London, Eng., 914 pp.; 1949.

[12] L. G. Bostwick, "Acoustic considerations involved in steady-state loudspeaker measurements," *Bell Sys. Tech. J.*, vol. 8, pp. 135–138; January, 1929.

[13] L. Wolffet and A. Ringel, "Loudspeaker testing method," PROC. IRE, vol. 15, pp. 363–376; May, 1925.

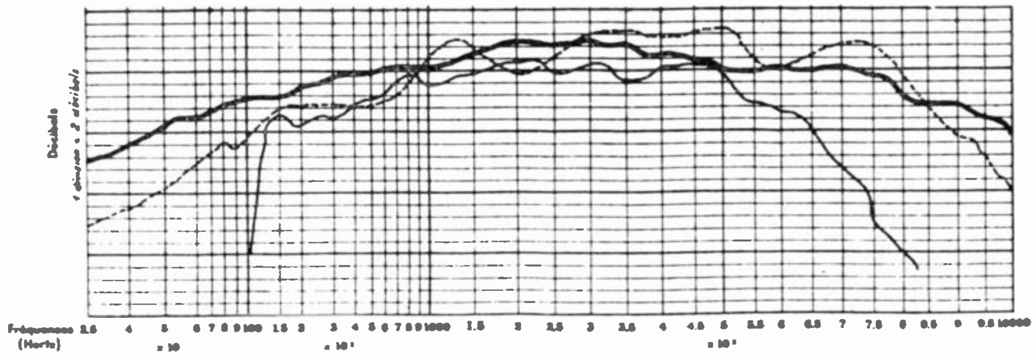
[14] P. S. Veneklasen, "Physical measurements of loudspeaker performance," *J. Soc. Motion Pict. Engs.*, vol. 52, pp. 641–655; June, 1949.

[15] F. H. Brittain and E. Williams, "Loudspeaker reproduction of continuous spectrum input," *Wireless Eng.*, vol. 15, pp. 16–20; January, 1938.

[16] H. G. Freygang, "Ueber ein neues Verfahren zur Ausmessung von Schallfeldern in Innenraümen vermittels eines kontinuirlischen Spektrums," *Akust. Z.*, vol. 3, p. 80; 1938.

[17] L. B. Hallman, Jr., "Selecting loudspeakers for special operating conditions," *Electronics*, vol. 12, pp. 22–24; November, 1938.

[18] B. Olmey, "Experiments with the noise analysis method of loudspeaker measurement," *J. Acoust. Soc. Amer.*, vol. 14, pp. 79–83; July, 1942.



——— horn loudspeaker with exponential horn
——— electrodynamic loudspeaker with exponential cone
- - - - electrodynamic loudspeaker with straight cone
(Ordinate = decibels, 2 db per division; abscissa = frequency cpe)

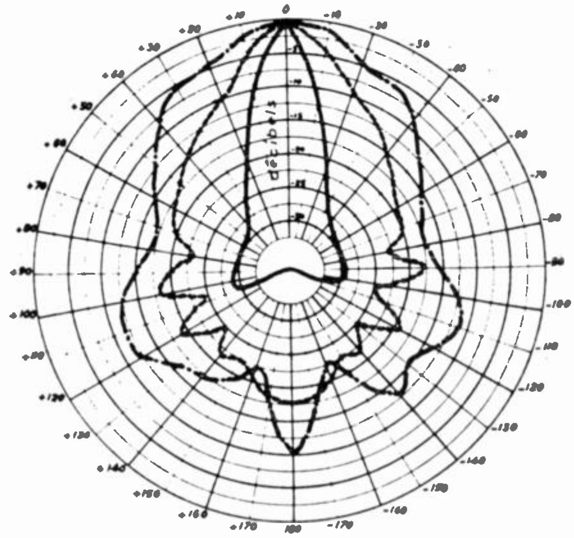
Fig. 6—Response characteristics of three loudspeakers of different types taken under the conditions specified in the text.

mits the use of any room (however, the reverberation time should not be too long); it requires the use of a frequency analyzer, which is always delicate to calibrate. Tests recently made at the Acoustic Department of CNET showed that the results obtained using such a source were comparable to those yielded by the usual method.

In any case, the voltage E used to take the frequency response characteristic must be such that it does not overload the loudspeaker under test; it would be desirable to set this voltage so as to apply rated power to the loudspeaker at the reference frequency.

B. Directivity Characteristic

The directivity of a loudspeaker characterizes at a given frequency its change of response as a function of angle (direction of transmitted sound). The characteristic should be taken at several frequencies: we propose 125, 400, 2000, and 5000 cps, which is not, however, intended to be limiting. It is also recommended to effect a continuous measurement, which means that the output sound pressure of the loudspeaker under test will be recorded while it is rotated slowly and continuously (complete rotation in about 3 to 4 minutes), for a fixed position of the microphone. It is desirable to represent the results obtained in the form of a polar diagram, as shown in Fig. 7. The test conditions have already been defined in the preceding section, and the input voltage at the loudspeaker should be the same as that selected for the frequency response characteristic. If it is not feasible to obtain the output sound pressure for continuous rotation of the loudspeaker, it is recommended to vary the angle from 0 degrees to 360 degrees in 15 degree increments.



-·-·-·- Elliptical cone electrodynamic loudspeaker
- - - - - - Circular cone electrodynamic loudspeaker
——— Horn loudspeaker with exponential horn

Fig. 7—Directivity patterns of various types of loudspeakers at 5000 cps.

C. Nonlinear Distortion (Harmonic Distortion and Intermodulation)

The nonlinear distortion of a loudspeaker is essentially characterized by a change of the waveform of the acoustic output with respect to the applied electrical signal amplitude.

At present, two methods are utilized to show the nonlinear distortion of a loudspeaker. In the first (harmonic distortion), a sinusoidal electrical voltage of known frequency, as clean as possible, is applied at the input of the loudspeaker, and the amplitude of all the harmonics and subharmonics existing in the acoustic output are measured with a distortion meter or analyzer, as shown in Fig. 8. The second method (intermodulation) is a more recent conception, and consists of applying to the loudspeaker under test two sinusoidal voltages of frequencies not harmonically related; the amplitudes of all the "sums" and "differences" existing in the acoustic output are then measured with an analyzer. Fig. 9 shows the block diagram of a setup often used for this purpose.

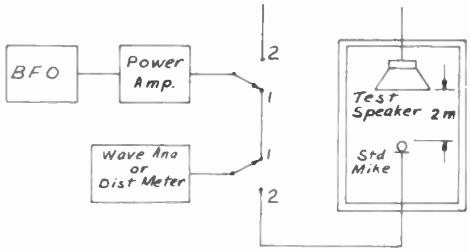


Fig. 8—Block diagram of the setup generally used to measure the harmonic distortion of loudspeakers. Position 1: measurement of harmonic distortion of driving voltage. Position 2: measurement of harmonic distortion of loudspeaker.

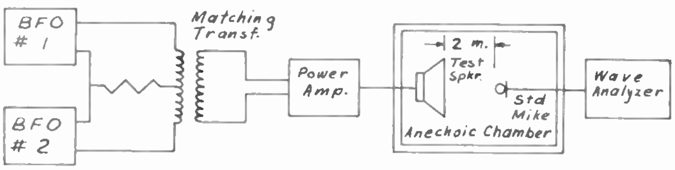


Fig. 9—Block diagram of the setup generally used to measure the intermodulation distortion of loudspeakers.

1) *Study of the Harmonic Distortion:* Harmonic distortion is, in practice, determined by the ratio of harmonic distortion:

$$D = 100\sqrt{\frac{(p_2^2 + p_3^2 + p_4^2 + \cdots)}{(p_1^2 + p_2^2 + p_3^2 + \cdots)}}$$

where $p_1, p_2, p_3, \cdots, p_n$ are the sound pressures of the fundamental sound (test frequency), and its harmonics, produced by the loudspeaker under test. The test frequencies should be chosen preferably at the resonant frequency of the loudspeaker cone, 400 and 2000 cps (for cone loudspeakers), and at 125–400 and 2000 cps for the others.

The test conditions were specified in the preceding section; it is also recommended to take the measurements, at each frequency, for three different power inputs, one of which is the nominal power handling capacity of the loudspeaker, and one is lower and the other higher than the nominal power.

Table II gives an indication of per cent harmonic distortion obtained for an electrodynamic loudspeaker having a nominal power-handling capacity of about 5 watts (resonant frequency about 40 cps).

TABLE II

Frequency cps	Per cent harmonic distortion for various powers			
	1 watt	2 watts	5 watts	10 watts
40*	8.4	43.8	66.5	80
60	3.6	4.15	7	11
100	2.1	2.5	4.8	6.1
400	1.3	1.65	3.2	4.7
2000	2.9	3.1	4.7	6.8

* Cone resonance.

simultaneously applied to the loudspeaker for this test are F_1 and F_2 ($F_2 > F_1$), the loudspeaker will reproduce them correctly, but "sum" and "difference" frequencies will be created (if the loudspeaker is not absolutely linear).

An analysis will show, for example, the frequencies $F_2 - F_1$, $F_2 + F_1$ and in general, $nF_2 \pm mF_1$. Often, frequencies of the first ($F_2 \pm F_1$) and second order ($2F_2 \pm F_1$ and $|2F_1 \pm F_2|$) are the more important, and the rest can be neglected. The main problem consists in determining for which two fundamental frequencies the test will be made, and what will be the voltage ratio of the two signals; on this point opinions are divided, some authors[20] advocate 40, 60, or 100 cps for F_1, and 1000, 7000, and 12,000 cps for F_2, both signals having the same amplitude; others advocate a constant $F_2 - F_1$ difference and explore the acoustic frequency band as completely as possible; finally, others suggest a ratio of 4 between F_1 and F_2.[21] We propose the following, with due consideration given to the various authors:

a) F_1 = resonant frequency of the loudspeaker cone (or 40 cps for driver units), $F_2 = 1000$ cps.
b) F_1 = resonant frequency of the loudspeaker cone (or 40 cps for driver units), $F_2 = 400$ cps.
c) $F_1 = 100$ cps, $F_2 = 1000$ cps.
d) $F_1 = 400$ cps, $F_2 = 7000$ cps.

For tests a) and b), the voltage ratio of the two signals will be 1; it will be 4 for tests c) and d); the common voltage will be that which will be determined when taking the frequency response; it will also be the same for F_1 for tests c) and d). The per cent intermodulation is calculated, by convention, by the following formula:

$$\text{per cent intermodulation} = 100\sqrt{\frac{(P^2_{2-1} + P^2_{2+1} + P^2_{(2\times 2)-1} + P^2_{(2\times 2)+1} + \cdots)}{(P_1^2 + P_2^2)}}$$

2) *Study of Intermodulation:* The first measurements of this type seem to have been made on a carbon microphones intended for use in telephony,[19] and, to our knowledge, only in 1941 was a similar method used in the study of loudspeakers.[20] If the two sinusoidal signals

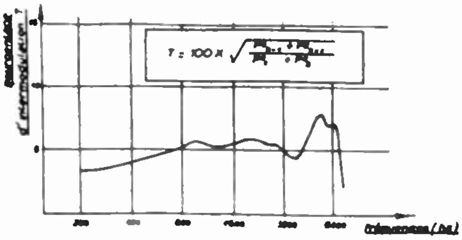
where $p_1, p_2, p_{2-1}, p_{2+1} \cdots$ are the sound pressures produced by the loudspeaker under test, for F_1, F_2, F_{2-1}, $F_{2+1} \cdots$.

As an example, Fig. 10 gives the intermodulation curve of a cone loudspeaker, and shows the graphic representation often used.

[19] E. Meyer, "Ueber eine einfache Methode der automatischen Klanganalyse und der Messung der Nichtlinearität vol Kohlenmicrophonen," *Elektr. Nachr. Tech.*, vol. 5, pp. 398–403; October, 1928.

[20] J. K. Hilliard, "Distortion tests by the intermodulation method," Proc. IRE, vol. 29, pp. 614–620; December, 1941.

[21] W. J. Warren and W. R. Hewlett, "An analysis of the intermodulation method of distortion measurements," Proc. IRE, vol. 36, pp. 457–466; April, 1948.



(Ordinate = per cent intermodulation distortion)

Fig. 10—Intermodulation distortion curve of a loudspeaker (the two signals were of equal amplitude, with a constant frequency difference of 100 cps; the frequency shown as the abscissa is equal to $F_n = \sqrt{F_1 F_2}$).

D. Impedance Characteristic vs Frequency

In a circuit which includes resistance, inductance, and capacitance, Kirchhoff's law gives

$$E = I\left(R' + j\omega L' + \frac{1}{j\omega C'}\right) = Z_E I,$$

where

$$Z_E = R' + j\omega L' + \frac{1}{j\omega C'},$$

Z_E being the impedance of the circuit under consideration.

In the case of a loudspeaker, to the electrical impedance Z_E is added a kinetic (motional) impedance Z_m (motional impedance) due to the fact that the system can vibrate and, in general,

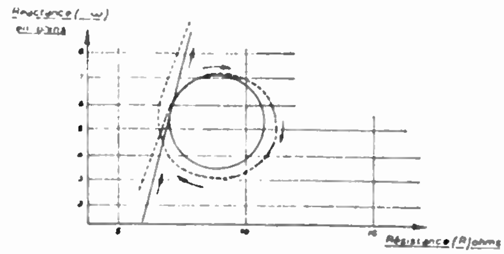
$$Z_m = R_1 + jL_1\omega.$$

The total impedance of the loudspeaker is equal to the sum of Z_E and Z_m; the two parts of this impedance are $R = R' + R_1$ and $j\omega L = j\omega L' + jL_1\omega$, since the capacitance $1/jC\omega$ can be neglected in most of the cases. The impedance which will be measured by one of the methods currently utilized (impedance bridge or other apparatus), is the total impedance Z:

$$Z = Z_E + Z_m.$$

It is useful, in some cases, to measure the two parts, real and reactive, of this impedance, in particular when one is looking for the precise resonant frequency, and when one wants to plot the Kennelly circle, which is obtained by plotting R as a function of ωL, for various frequencies. R and ωL represent the difference in resistance and reactance measured for a same signal ω, respectively, when the cone is free to vibrate and when it is blocked mechanically.

Fig. 11 shows such a curve: the circular shape is only due to the mechanical resonance of the system, since if the latter did not exist, one would simply obtain the straight line slightly inclined (which would be vertical if the electrical resistance R' did not cross very lightly with the frequency because of Kelvin's effect).



→ direction of increasing frequency
—— theoretical curve
- - - - experimental curve

Fig. 11—Voice coil reactance $L\omega$ of an electrodynamic loudspeaker as a function of its resistance R. (Kennelly's circle.)

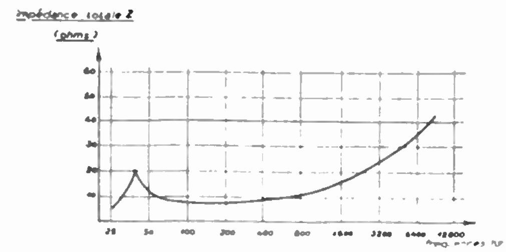


Fig. 12—Total impedance (electrical impedance + motional impedance) of an electrodynamic cone loudspeaker as a function of frequency.

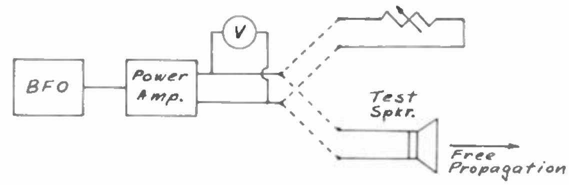


Fig. 13—Block diagram of setup used to measure the total impedance Z of loudspeakers. ($Z = R$ pure resistance, when the voltage E read on the voltmeter is the same for either position of the switch.)

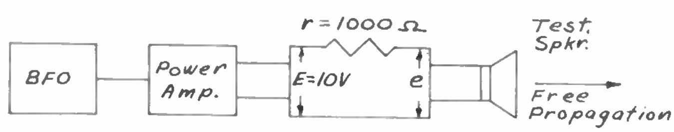


Fig. 14—Block diagram of a setup used by Veneklasen to measure the total impedance Z of loudspeakers (since $Z \ll 1000$ ohms, we have:

$$Z = \frac{e}{i} \approx \frac{e}{\frac{E}{1000}} \approx 100e.$$

It is therefore only necessary to measure e.)

Fig. 12 illustrates an example of a loudspeaker total impedance Z as a function of frequency, and Figs. 13 and 14 show the principles of two setups enabling the measurement of this impedance.

Fig. 14 allows the rapid calculation of Z, with an accuracy sufficient for the measurement, as already defined in the preceding section. If it is found that the impedance characteristic as a function of frequency changes with the power applied to the loudspeaker, it is recommended to obtain the characteristic for several values of applied power.

E. Efficiency Characteristic

The efficiency η of a loudspeaker at a given frequency is defined in general as the ratio of the total radiated acoustical power to the applied electrical power.

$$\eta \text{ (per cent)} = \frac{P_a}{(P_e)} \times 100.$$

To obtain the efficiency characteristic as a function of frequency, it is therefore necessary to determine, for several frequencies, the electrical power applied to the loudspeaker P_e ($P_e = E^2/Z$ as indicated in Section III-A) and the total radiated acoustical power.

If the measurement of Pe does not offer any difficulties, it is not the same for Pa; therefore we shall dwell upon the evaluation of the latter.

1) *Reverberation Chamber Method:* This was first advocated in Germany by Meyer and Just in 1929,[22] and consists of measuring the sound energy in the room, after having ascertained that the sound is diffuse and uniformly distributed. The classical theory of reverberation gives for the energy intensity for each point in the room:

$$E_d = 4\frac{Pa}{Ac}$$

where A is the absorption of the room and c the speed of sound. Since

$$E_d = \frac{P^2}{\rho c^2}$$

where P is the acoustic pressure at the point of measurement, and ρ is the air density, we have for Pa

$$P_a = \frac{Ap^2}{4c\rho}.$$

In this case, it is desirable that the reverberation time of the room be sufficiently high (>5 seconds); it is also necessary to operate not with sine waves, but with warble frequencies, or with $\frac{1}{3}$ octave frequency bands (noise), to avoid the formation of standing waves. We advocate the adoption as measuring frequencies, the frequencies chosen for the directivity characteristic, 125, 400, 2000, and 5000 cps. If warble tones are used, we propose a modulation equal to ± 10 per cent of the above frequencies, at a rate of 20 cps; if white noise is utilized, it would be recommended to use $\frac{1}{3}$ octave wide bands, centered about the frequencies indicated above, since the $\frac{1}{3}$ octave bandwidth seems more and more standardized, internationally, in acoustic measurements.

2) *Free Field Method:* The test conditions of measurement are the ones defined in Section II. The method consists in measuring the sound pressure at a great number of points of a sphere, the center of which is occupied by the loudspeaker; theoretically it is necessary that the radius of this sphere be large in relation to the wavelength, so that the sound pressure measured at each point be in phase with the particle velocity. The total acoustic power P_a is then given by

$$P_a = \left(\frac{1}{\rho c}\right) \iint p^2 dS$$

integrating between 0 and 2π or 4π, according to whether the semispherical radiation is considered (case of the infinite baffle), or the entire sphere (general case).

This method assumes that the measuring microphone is very small, in order not to disturb the acoustic field. This method seems more rigorous than the preceding one, and the results cannot be contested if the measurements are conducted carefully. The frequencies at which the measurements are made were mentioned above; in this case warble tones are not necessary, but if it is desirable one can always utilize white noise. For the purpose of information, Table III gives several results of efficiency measurements made according to this method, at various frequencies, on several loudspeakers of different sizes.

TABLE III

Frequency cps	Per cent efficiency of loudspeakers of various diameters					
	12 cm	16 cm	21 cm	24 cm	28 cm	34 cm
300	0.015	0.014	0.04	0.067	0.107	0.091
1000	0.16	0.09	1.83	1.3	2.48	0.93
2000	2	1.83	2.18	0.51	1.71	
(3000)						(7.27)
4000	1.2	0.55	1.84		2.092	
8000	0.013	0.021	0.086	0.35	0.104	

The values of efficiency obtained are always very small (<7 per cent); they are higher for exponential horn loudspeakers, and can reach 30 per cent in some cases, at some frequencies.

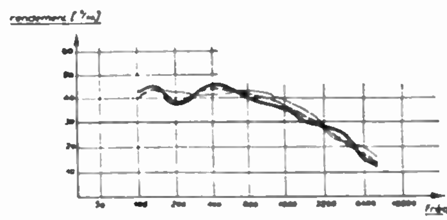
3) *So-Called "Motional Impedance Method":* This method[23] was developed before that of the reverberation chamber, and goes back originally to Kennedy's work; it consists of obtaining two impedance measurements, one when the loudspeaker functions normally, under the conditions prescribed above (determination of R and Z), the other when the vibrating system is entirely blocked, with the same test conditions as for free motion (determination of R' and Z_E), utilizing the notations defined above in Section III-D on the impedance characteristic. The efficiency is then expressed by the relation:

$$\eta \text{ (per cent)} = \frac{(R_1)}{R} \times 100.$$

R is given directly by the first measurement, and $R_1 = R - R'$, R' is given by the second measurement.

[22] E. Meyer and P. Just, "Messung der Gesamtenergie von Schallquellen," *Z. Tech. Phys.*, vol. 17, pp. 309–316; 1929.

[23] A. E. Kennelly and G. W. Pierce, "The impedance of telephone receivers as affected by the motion of their diaphragms," *Proc. Amer. Acad. Art. Sci.*, vol. 48, p. 111; 1912.



——·—— motional impedance method
·········· free field method
———— reverberation chamber method
(Ordinate = per cent efficiency)

Fig. 15—Efficiency characteristic of a horn loudspeaker with an exponential horn, measured by the above methods.

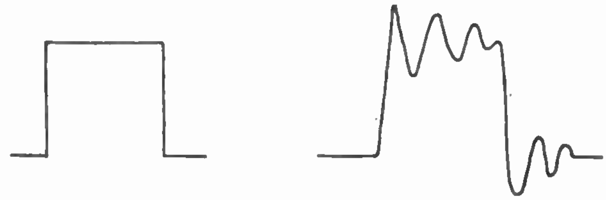
When the efficiency of the loudspeaker is low, R and R' are very close to one another, and a small error in their measurement results in a much more important error in the determination of the efficiency; also, this method assumes evidently that there are no mechanical losses in the diaphragm or in the suspension system, and it neglects the losses due to the viscosity of the air. It is also possible, for the determination of R_1, to perform the measurement of R_1 in a vacuum;[24] in that case, however, the load on the diaphragm is not normal, and the losses can be different from what they are normally.

As an example, Fig. 15 shows the efficiency characteristic of an exponential horn loudspeaker, measured by various methods.

F. Transient Response Characteristic

It seems that it is MacLachlan[25] who first had the idea to use square waves in the study of loudspeakers. In 1946, Shorter[26] preferred to use trains of sine waves for the study of transient response; in his opinion, this method was more realistic with respect to the true functioning of loudspeakers than that using pulses; he even preconized a three-dimensional graphic presentation, in which he used for coordinates, the frequency, the decay time after sudden cutoff of the wave train, and the value of the decay in decibels. Figs. 16 and 17 show the oscillograms obtained by each method.

The great interest in the transient response of a loudspeaker comes from the fact that, when its applied electrical signal is suddenly interrupted, generally some of its elements continue to vibrate, and several milliseconds elapse before the diaphragm comes back to a position of complete rest; there is then some analogy with the phenomenon of reverberation in a room in which a sound is suddenly interrupted. It is well known that, taking into account the reverberation theory and its limitations, the reverberation time is a primordial factor in appraising the acoustic quality of a room; it is therefore conceivable that the transient response of a



(a) (b)

Fig. 16—Example of the reproduction of a pulse by an electrodynamic cone loudspeaker. (a) Wave shape of the electrical pulse applied to the loudspeaker voice coil. (b) Wave shape of the electrical signal produced by a microphone located in front of the loudspeaker.

(a) (b)

Fig. 17—Example of the reproduction of a sinusoidal wave train by an electrodynamic cone loudspeaker. (a) Wave shape of the sinusoidal voltage applied to the loudspeaker voice coil. (b) Wave shape of the electrical signal produced by a microphone located in front of the loudspeaker.

loudspeaker is equally very important, and several authors judged it indispensable to define the acoustic quality of a loudspeaker.[26,27] Others, on the contrary, such as Hentsch,[28] think that the steady state characteristics are sufficient in this respect, provided that they include all the elements playing an important part in the reproduction of the transients, which is, in their opinion, generally the case.

It was shown,[29] at least for quadripoles, that phase is not independent from the steady state frequency response characteristic, and recent experiments by Ewaskio and Mawardi[30] have confirmed these hypotheses, except in the case of multiple loudspeakers, which was predictable *a priori*, since in this case, the low and high frequency sounds are generated by different units. It is concluded from these experiments that the variations of decay time as a function of frequency agree exactly with the variations of the steady state frequency response characteristic, and that the latter should be sufficient to judge the sound reproduction quality of loudspeakers in normal transient usage. However it might be, some authors think that the curves of amplitude and phase as a function of frequency are independent from each other, at least within certain limits;[27] we share this opinion, since we have observed that two loudspeakers, having steady-state response curves little different from each other, could yield sound reproduction of relatively different quality.

[24] P. K. Turner, "Some measurements of a loudspeaker in vacuum," *J. IEE., London*, pp. 591–610; May, 1931.

[25] N. W. Maclachlan, "Loudspeakers," Oxford University Press, Oxford, England; 1934.

[26] D. E. L. Shorter, "Loudspeaker transient response," *BBC Quart.*, no. 3; October, 1946.

[27] G. Guyot, "Étude sur les haut-parleurs en régime transitoire," *Rev. gén. Électr.*, vol. 57, pp. 245–253; June, 1948.

[28] J. C. Hentsch, "La fidélité des haut-parleurs dans la reproduction des régimes transitoires," *Tech. Mitt.*, vol. 29, p. 201; June, 1951.

[29] H. Backaus, "Ueber die Bedeutung der Ausgleichvorgängen in der Akustik," *Z. Tech. Phys.*, vol. 13, pp. 31–46; 1932.

[30] C. A. Ewaskio and O. K. Mawardi, "Electroacoustic phase shift in loudspeakers," *J. Acoust. Soc. Amer.*, vol. 22, pp. 44–48; July, 1950.

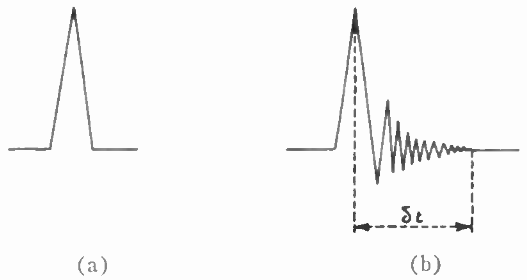


Fig. 18—Wave shape and determination of the time duration of a transient produced by a loudspeaker. (a) Approximate wave shape of the pulse applied to the loudspeaker voice coil 10 to 15 times a second. (b) Wave shape of signal produced by a microphone located in front of the loudspeaker.

We shall not go further, in this paper, into the theory of transient tests, but for the measurement of transient response we recommend the use of pulse signals of very short duration (Dirac pulses) (impulse function) which meet the condition $\int_{-\infty}^{+\infty} V(t)dt = A > 0$ and such that $V(t)$ be always null, except when $t = t_0$.

In practice, it seems that one could use pulses of duration of the order of 1/1000 to 1/10,000 second, without deeply affecting the results obtained.[27]

There is no advantage in choosing, as repetition rate for these signals, a frequency lower than 10 or 15 cps, as it seems that the frequencies in the loudspeaker are almost always damped after about 10 or 15 periods.

The general test conditions shall be naturally those specified in Section II, but it is then necessary to utilize wide band microphone preamplifiers (20–100,000 cps), which are relatively easy to realize, and it is desirable that the microphone itself have as wide a band-pass (response) as possible, which is much more difficult to obtain. Preferably, the phenomenon will be read on a scope, by a photo of pulse waveform picked up by the microphone, to determine the duration of the transient phenomenon $\delta(t)$ in accordance with Fig. 18. Very many subjective tests still would have to be performed to know exactly what is the correlation between $\delta(t)$, the loudspeaker quality of reproduction, and its steady state frequency response, but it seems that it would be advantageous to have the duration of the transient $\delta(t)$ as short as possible to realize a good quality loudspeaker. Some authors[30] indicate a maximum duration of 1 millisecond, but it is quite possible to accept longer durations (maybe up to several 1/100 second, the time constant of the ear being of the order of 100 milliseconds).

Finally, let us point out a method actually preconized and tried by the "British Standard," which has the great advantage of being very simple, but which necessitates a great deal of measurements, if one desires to explore the entire frequency spectrum reproduced by the loudspeaker. It consists of energizing the loudspeaker under test, steady state, successively at the various frequencies for which one desires to carry out the tests, and measuring, under conditions similar to those which were defined in Section II, the sound pressure thus produced, then recording the decay curve of this pressure as a function of time, after a sudden interruption of the signal applied to the loudspeaker.

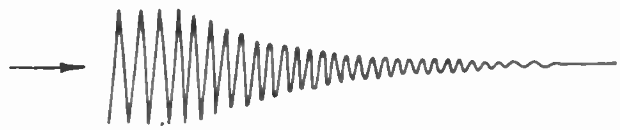


Fig. 19—Decay curve of the sound pressure produced by a loudspeaker, as a function of time, after sudden interruption of the signal (test signal = 85 cps). Resonant frequency of the loudspeaker cone = 85 cps. Decay time = about 300 milliseconds.

It is evident that the decay time becomes more important when the test frequencies are closer to the electrical or mechanical resonant frequencies of the loudspeaker.

Fig. 19 gives an example of such a curve, corresponding to a pronounced resonant frequency of a loudspeaker. This procedure is naturally purely quantitative, but it permits, however, to obtain interesting information with respect to the loudspeaker quality tested for the reproduction of transients.

IV. Conclusions

The object of this paper was to specify the conditions of acoustic measurements and the methods which are generally utilized for the study of loudspeakers. These measurements, which are performed either by official (Government) laboratories to verify conformance with specifications, or by manufacturers for development and quality control of their product, should be made according to specifications. We have tried to outline the main points of such specifications which we are already using, partly at least.

It is to be presumed that, at future international conventions on acoustics, the problem of standardization and strict regulation of such measurements will be discussed, as it has been since 1947, for the measurement of sound transmission equivalents for partitions and ceilings, and for the measurement of shock waves, and, as it is at the present time, for the measurement of absorption coefficients in reverberation chambers. This paper would then naturally be a contribution to the discussion which would start at that time. It is in this spirit, and with a view to serve as a concrete basis to future discussions, that this paper was written.

One problem, somewhat similar, is at present before the Commission Electrotechnique Internationale (CEI): standardization of measurements to be performed on radio receivers. Questions connected to those treated in this paper were raised. When the question of whether to include the loudspeakers as one of the component equipments to be measured was raised, it was unanimously decided as necessary by the members of the writing subcommittee. Certain paragraphs of this paper enabled the French delegation to prepare a beforehand proposal for the meeting of the CEI at Montreux (November, 1951), and to present a more detailed proposal at La Haye (September, 1952), which will be examined by Committee 29 (electroacoustics) at the next meeting of the CEI at La Haye in June, 1953.[31]

This simple example shows the importance of the problem of the standardization of measurements which should be performed on loudspeakers. If we have been able by this work to bring some new light to this problem, while pointing out the difficulties and precautions inherent in the practical application of the provisions herein submitted, we shall have achieved one of the main objectives that we had assigned ourselves.

BIBLIOGRAPHY

[1] Wolff, L. "Sound Measurements and Loudspeakers Characteristics," PROCEEDINGS OF THE IRE, Vol. 16 (December, 1928), pp. 1729–1741.

[2] Beers, G. L. and Belar, H. "Frequency Modulation Distortion in Loudspeakers," *Journal of the Society of Motion Picture Engineers*, Vol. 47 (April, 1943), pp. 207–221.

[31] Since then, several meetings have been held, and a document prepared by the U. S. delegation was to be examined in July, 1958.

[3] Beers, G. L. and Belar, H. "Loudspeaker Frequency Response Measurements," Jensen Manufacturing Company, Technical Monograph No. 1, (1944).

[4] Hopkins, H. F. and Stryker, N. R. "A Proposed Loudness Efficiency Rating for Loudspeakers and the Determination of System Power Requirements for Enclosures," PROCEEDINGS OF THE IRE, Vol. 36 (March, 1948), pp. 315–335.

[5] Roessler, E. "Zur Entstehung nichtlinearer Verzerrungen im Lautsprecher," *Funk und Ton*, Vol. 4 (November, 1950), pp. 549–553.

[6] Hubner, R. "Neue Lautsprecher Prüfmethoden," *Radio Service*, Vol. 11 (January/February, 1951), pp. 2107–2110.

[7] Salmon, V. "Efficiency of Direct Radiator Loudspeakers," *Audio Engineering*, Vol. 35 (August, 1951), pp. 13–14.

[8] Heimann, W. "Ueber die Bestimmung des Wirkungsgrades von Lautsprechern," *Elektrische Nachrichten-Technik*, Vol. 9 (1932), pp. 302–308.

[9] David, P. "Dans Quelle Mesure l'Étude d'un Hautparleur en Régime Permanent Permet-Elle de Prévoir son Comportement en Régime Transitoire?" *Onde Électrique*, Vol. 17 (June, 1938), p. 309.

[10] Helmbold, J. H. G. "Oscillographische Untersuchungen von Einschwingvorgängen bei Lautsprechern," *Akustische Zeitschrift*, Vol. 2 (1937), p. 256.

[11] Moir, J. "Transient and Loudspeaker Damping," *Wireless World*, Vol. 56 (May, 1950), pp. 166–170.

[12] Corrington, M. S. "Transient Testing of Loudspeakers," *Audio Engineering*, Vol. 34 (August, 1951), pp. 9–13.

Design of Transistor RC Amplifiers*

RAY P. MURRAY†

Summary—**Some of the basic factors involved in the design of transistor RC amplifiers are considered. Particular emphasis is placed on operating point stabilization and its relation to such factors as gain, battery drain, and distortion. The stabilization factor employed here is a measure (in terms of per cent) of the stabilization contributed by the stabilization circuitry. Only the fundamental common-emitter connection is discussed.**

VARIATION in transistor characteristics is one of the greatest problems encountered in the design of transistor circuits. Failure to take these variations into account will generally result in an unsatisfactory circuit design. For example, let us suppose that an unstabilized audio amplifier (Fig. 1) is assembled and that R_B and R_C are adjusted so as to give best performance (gain and fidelity) with a particular transistor at room temperature. Now if this amplifier is duplicated using the same values for R_B and R_C and the same type transistor, in many cases the performance will be unsatisfactory for one or both of the following reasons: 1) incorrect operating point due to dissimilarity of transistors of the same type, and 2) incorrect operating point due to changes in transistor characteristics caused by temperature variation.

* Manuscript received by the PGA, March 15, 1958; revised manuscript received, June 23, 1958.

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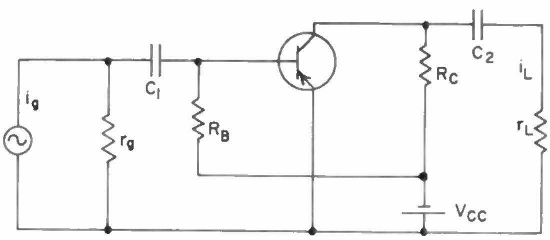


Fig. 1—Unstabilized amplifier.

Those who are new to the transistor field sometimes underestimate the importance of the above effects until they find that a circuit may become completely inoperative when the transistor is replaced with another of the same type, or when an amplifier built in the cool of the evening just won't operate in the warmth of the afternoon.

Here we consider two methods of dealing with the problem of variation in characteristics. The first applies to an unstabilized amplifier (Fig. 1) in which the effects of temperature and parameter variation due to production spread are handled by proper choice of operating point and circuit resistors. The second method utilizes a stabilized circuit (Fig. 2) in which dc negative feedback tends to make the circuit performance less sensitive to changes in transistor characteristics.

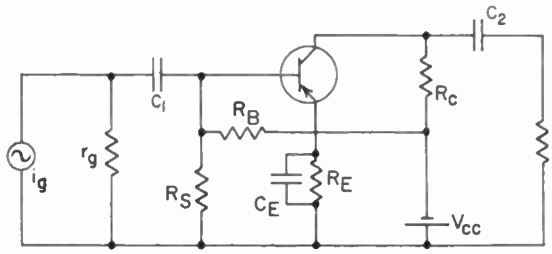


Fig. 2—Stabilized amplifier.

Preliminary Considerations

In the beginning it is well to realize that it is difficult if not impossible to produce a "paper design" of a complete transistor amplifier system in which the usual performance factors of gain, nonlinear distortion, phase shift, frequency response, and temperature characteristics are met with a high degree of accuracy. After initial testing, the first "paper design" may be modified considerably by adding inverse feedback circuitry to meet distortion specifications, addition of equalization circuitry to produce the desired frequency response, and possibly the addition or reduction in the number of stages in accordance with gain requirements, etc. The goal here will be the first "paper design."

Input and output stages have requirements (*e.g.*, equalization, noise figure, and impedance levels) that are peculiar to the terminal devices, and therefore this paper deals with an intermediate amplifier. First one must decide whether it is to be a voltage, current, or power amplifier. An inspection of the transfer characteristics of Fig. 3 shows that for low-level transistors, the collector current is more nearly proportional to the base current than to the base voltage. Thus we conclude that it is best to consider the low-level transistor as a current-operated device, and therefore the intermediate RC amplifier will be a current amplifier regardless of the system input and output. The effect of temperature on the transfer characteristics is also shown in Fig. 3. The transfer characteristics of an audio power transistor (Fig. 4) show that it can best be considered as a voltage-operated device and hence is usually driven by a low-impedance source.

Gain

Stage gain is a clear-cut term in vacuum-tube amplifier circuitry, but with transistors the effect of one stage on another makes the use of the term "stage gain" somewhat undesirable, unless one specifies insertion gain, transducer gain, etc. For our purposes here we are not so much interested in the gain itself as in the factors that affect the gain. We will consider a "working circuit" (Figs. 1 and 2) in which the single stage is only a part. Here the following transistor stage is represented by resistor r_L, and the preceding transistor stage is represented by a constant current generator with internal

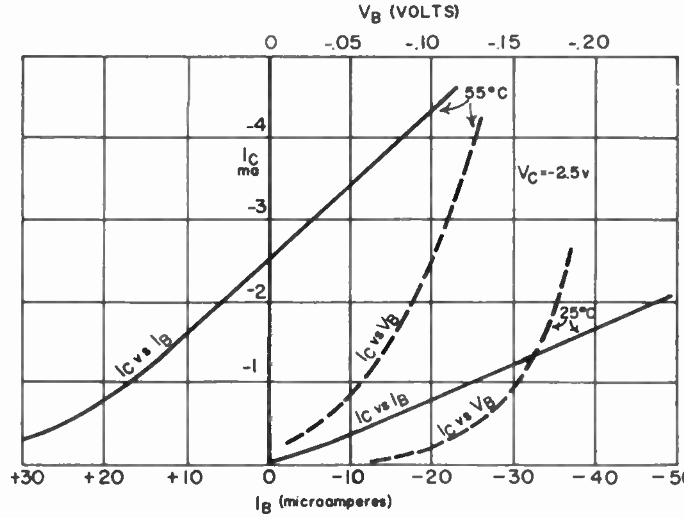


Fig. 3—Common-emitter transfer characteristics for a low-level transistor.

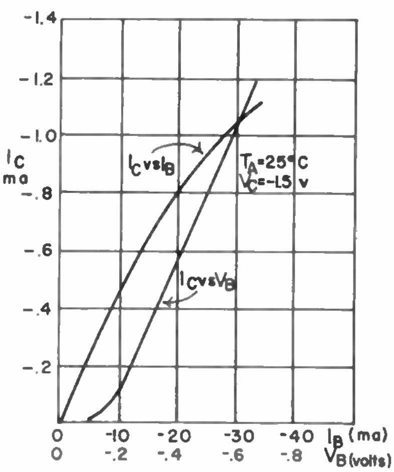


Fig. 4—Power transistor transfer characteristics.

resistance of r_g.[1] For gain calculations the input circuit may be simplified by replacing the shunt resistances with their parallel equivalent, R_g, which we shall call the *source resistance*. In Fig. 1,

$$R_g = \frac{r_g R_B}{r_g + R_B} \tag{1}$$

and in Fig. 2,

$$R_g = \frac{r_g R_B R_S}{r_g R_B + r_g R_S + R_B R_S}. \tag{2}$$

In Figs. 1 and 2 the current gain

$$G = \frac{i_L}{i_g} \tag{3}$$

[1] r_g would normally be the same as the collector coupling resistor for the preceding stage if the output resistance of the preceding transistor is much greater than the coupling resistance.

may be considered to be governed by three factors: the *input circuit attenuation factor*, A_{in}, which accounts for the loss of signal current in r_g, R_B, and R_S, the transistor current amplification factor, β, and the *output circuit attenuation factor*, A_{out}, which accounts for the loss of signal current in R_C.[2]

$$G = A_{in}\beta A_{out} \tag{4}$$

$$A_{in} = \frac{1}{1 + \dfrac{r_{in}}{R_g}} \tag{5}$$

where r_{in} is the ac input resistance of the transistor.

$$A_{out} = \frac{1}{1 + \dfrac{r_L}{R_C}}. \tag{6}$$

For Figs. 1 and 2,

$$G = \left(\frac{1}{1 + \dfrac{r_{in}}{R_g}}\right)(\beta)\left(\frac{1}{1 + \dfrac{r_L}{R_C}}\right). \tag{7}$$

For high gain, both attenuation factors should be high, that is, approach one. For the input circuit, this requires that r_g, R_B and R_S (if present) be high compared to the input resistance of the transistor. For the output circuit, R_C should be high compared to r_L. There are other requirements on the circuit resistors (R_S, R_B and R_C) so compromise is necessary, but it is well to keep in mind that any reduction in resistance of any of these resistors means a reduction in gain. The circuit gain is also affected by temperature changes since r_{in} and β may both change with temperature.

Distortion[3]

In low-level RC amplifiers, there are two main sources of severe distortion. First, the input circuit is quite nonlinear as shown in Figs. 5 and 6, and secondly, distortion may occur in the output circuit as a result of the operating point being on a very nonlinear portion of the characteristic. The distortion caused by nonlinear input impedance can be minimized by proper choice of operating point and by driving the transistor from a high-impedance source. Thus the same circuit requirements for high gain are also desirable to swamp out the effects of the nonlinear input impedance. Selection of an operating point (see Fig. 6) where the input impedance does not vary severely with collector current is desirable and quite necessary in case of a fairly low impedance source.

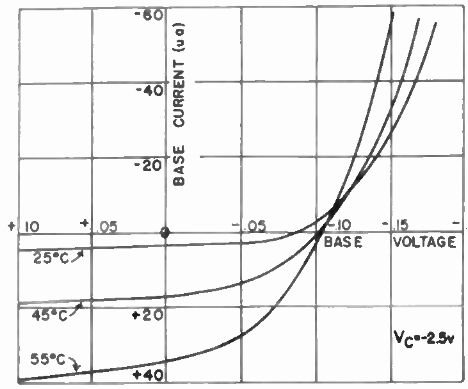


Fig. 5—Common-emitter input characteristics.

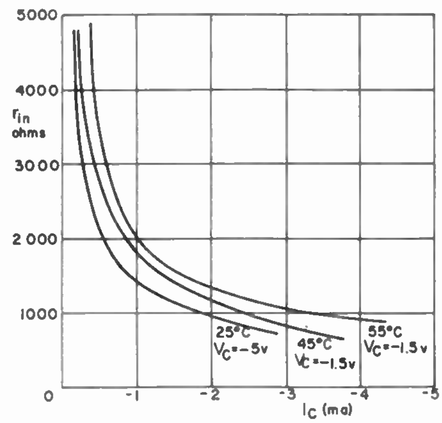


Fig. 6—Variation of common-emitter input resistance with collector current.

Frequency Response

The high-frequency performance of audio RC amplifiers is generally determined primarily by the frequency characteristics of the transistor. The input impedance, being capacitive, goes down at high frequencies, tending to increase the gain. The output impedance, also capacitive, goes down at high frequencies and tends to reduce the gain. However, the amplitude and phase shift associated with β at high frequencies is usually the most significant factor in determining the high frequency performance. Thus high-frequency performance is obtained by the choice of a transistor with a high cutoff frequency. Transistor specifications generally list the α cutoff frequency, f_α, whereas the cutoff frequency of importance in the common-emitter amplifier is f_β. For a first approximation, the transistor selected for an RC amplifier should have an f_α greater than $(\beta+1)$ times the highest frequency of interest as far as the amplifier

[2] The shunting effect of the transistor output resistance has been neglected here since in most cases

$$R_C \ll \frac{r_c}{\beta + 1}.$$

See, A. W. Lo, R. O. Endres, J. Zawals, F. D. Waldhauer, and C. Cheng, "Transistor Electronics," Prentice Hall, Inc., Englewood Cliffs, N. J., ch. 3; 1955.

[3] A more complete discussion of distortion will be found in: L. P. Hunter, "Handbook of Semiconductor Electronics," McGraw-Hill Book Co., Inc., New York, N. Y., sec. 11; 1956.

is concerned.[4] Thus for an amplifier to be flat to 15,000 cps, a transistor with a β of 49 ought to have an f_α of about 1 mc.

The low-frequency performance of an RC amplifier is determined primarily by the coupling and by-pass capacitors. For the emitter by-pass capacitor (C_E of Fig. 2), the reactance at the lowest frequency of interest is given by[5]

$$X_{CE} = \left(r_e + \frac{R_g + r_b}{\beta + 1}\right)\sqrt{\frac{1}{P^2} - 1} \qquad (8)$$

where r_e, r_b and β are transistor parameters, R_g is the source resistance (parallel combination of R_B, R_S and r_g), and P is the low-frequency attenuation factor due to incomplete by-passing of R_E. In terms of the transistor input resistance

$$X_{CE} = \frac{r_{\text{in}} + R_g}{\beta + 1}\sqrt{\frac{1}{P^2} - 1} \qquad (9)$$

if $(\beta+1)r_L R_C/(r_L+R_C) \ll r_c$. Since r_{in} and β are functions of temperature and operating point (see Figs. 6 and 7), a conservative design requires the use of the highest value of β and the lowest value of r_{in}.

For Fig. 2, it can be shown that the reactances of C_1 and C_2 are given by

$$X_{C1} = \left(r_g + \frac{r_{\text{in}} R_S R_B}{R_S R_B + r_{\text{in}} R_S + r_{\text{in}} R_B}\right)\sqrt{\frac{1}{P^2} - 1} \qquad (10)$$

and

$$X_{C2} = (R_C + r_L)\sqrt{\frac{1}{P^2} - 1} \qquad (11)$$

where P_s is the low-frequency attenuation factor due to the capacitor involved. If there are N such low-frequency calculations in a system design, then

$$P = (P_s)^{1/N} \qquad (12)$$

where P_s is the low-frequency attenuation factor for the system.

The Temperature Problem

The resistance of the emitter-collector path of a transistor decreases with rising temperature, so that in a transistor circuit the collector current, and/or the collector voltage, may change considerably for only a few degrees' increase in temperature. The extent of this temperature effect depends largely on the temperature of operation, being rather insignificant at low temperatures but quite severe at high temperatures. For germanium transistors, small temperature changes become noticeable at room temperature and may have radical effects for temperatures as low as 40°C. The temperature effects do not become significant for silicon transistors until the temperature has increased appreciably above room temperature.

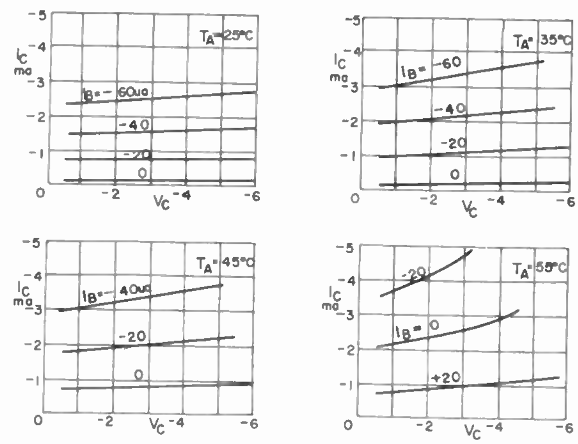


Fig. 7—Effect of temperature on output characteristics.

Some of the effects of temperature on the transistor characteristics can be seen from the output characteristic curves of Fig. 7. In addition to an upward shift of all curves to higher values of I_C (reduced resistance of emitter-collector path), the spacing between curves (indication of β) increases appreciably as does their slope (indicative of the output resistance).

The effect of collector dissipation ($V_C I_C$) is to raise the junction temperature above the ambient temperature, and thus the curves bend upward for high values of dissipation.[6] In Fig. 7, the range of I_C and V_C is limited to small values, and thus the bending effect is not very great. At 25°C the effect of temperature increase is not nearly as great as at 55°C, so the upward bending due to collector dissipation is more pronounced on the high temperature graph. This bending would not show on curves taken for constant junction temperature.

Although some of the reasons for the temperature effects are well known, no accurate relation is available for representing I_C in terms of temperature. A relation which is satisfactory for one transistor type may not be acceptable for another. Graphs or equations which relate I_{CO} to temperature are a help, but they fall short of telling the complete story.[7] It is suggested here that the most satisfactory method of dealing with the temperature-dependent characteristics is to use two sets of characteristic curves, one for some low temperature (*e.g.*, 25°C) and another set corresponding to the maximum temperature of operation. If, in obtaining the two sets of graphical characteristics, we take into account the extremes of β and I_{CBO} allowed for a certain type transistor, the problem of parameter variation due to production spread is handled along with the temperature problem.[7]

[4] R. F. Shea, "Transistor Circuit Engineering," John Wiley and Sons, Inc., New York, N. Y., chs. 2 and 7; 1957.

[5] R. P. Murray, "Emitter bypassing in transistor circuits," IRE Trans. on Audio, vol. AU-5, pp. 71–72; May–June, 1957.

[6] For low-level audio transistors with no heat sink, the junction temperature increases approximately 0.3°C per milliwatt of junction dissipation.

[7] I_{CO} is the current that flows in the collector circuit when the emitter is opened, and I_{CBO} is the collector current with the base opened.

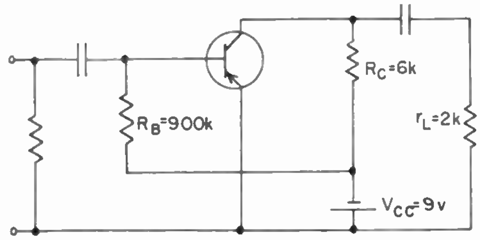


Fig. 8—Unstabilized RC amplifier in which the effect of temperature has not been taken into account.

We have seen how temperature affects some of the transistor characteristics. Now let us see how the operating point of an RC amplifier is affected by the change in characteristics. Refer to the unstabilized amplifier circuit of Fig. 8 which uses a transistor with characteristic curves of Figs. 9 and 10.

Since the base bias current ($I_B \cong V_{CC}/R_B = -10\ \mu a$) is not affected by temperature, the operating point for any temperature will be at the intersection of the dc load line and the transistor curve corresponding to a base bias of $-10\ \mu a$.

Note the low temperature operating point in Fig. 9. As the temperature rises, the $I_B = -10\ \mu a$ bias curve rises, and thus the operating point shifts up and to the left along the dc load line.

As shown in Fig. 10, the high temperature operating point is in such a position that small changes in base current produce no change in collector current. The amplifier is inoperative (gain equals zero) at 55°C.

The Unstabilized RC Amplifier

From the standpoint of battery drain and collector dissipation, it is desirable to select an operating point corresponding to the smallest permissible values of collector voltage and current. A higher level operating point may be chosen for many reasons, some of which are linearity, gain, noise figure, impedance levels, and temperature stabilization. Here we will choose the lowest level operating point that will permit proper operation over the specified temperature range. Thus the signal to be considered in the design will correspond to the maximum signal that the amplifier is expected to handle.

The most significant specifications for the intermediate amplifier are signal level and input and output impedance levels. Let us assume the maximum values of the signal voltage and current in r_L to be V_{sig} and I_{sig}. Now if r_L is small compared to R_C, the variations in collector voltage and current will also be V_{sig} and I_{sig}. This condition on r_L and R_C can generally be met if the temperature variation is not severe and if a high value of V_{CC} is available; however, it will not be met for wide temperature variations, and a modification of the first design will be required or else the amplifier will not handle the specified signal level.

In the unstabilized RC amplifier, the operating point is allowed to shift (with temperature) along the dc load

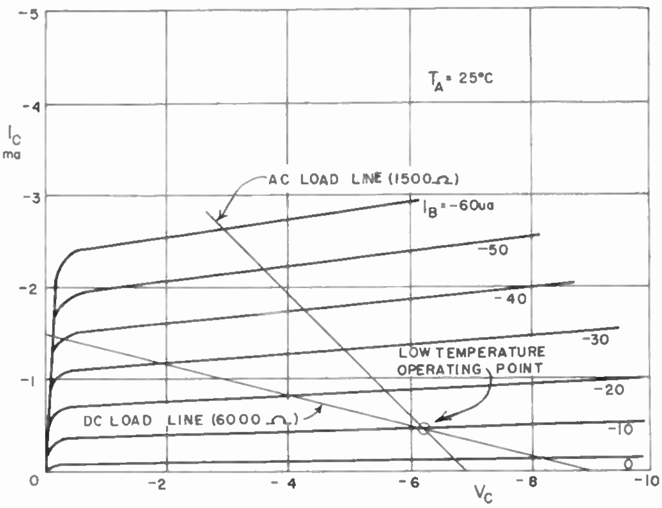


Fig. 9—Output characteristic for 25°C.

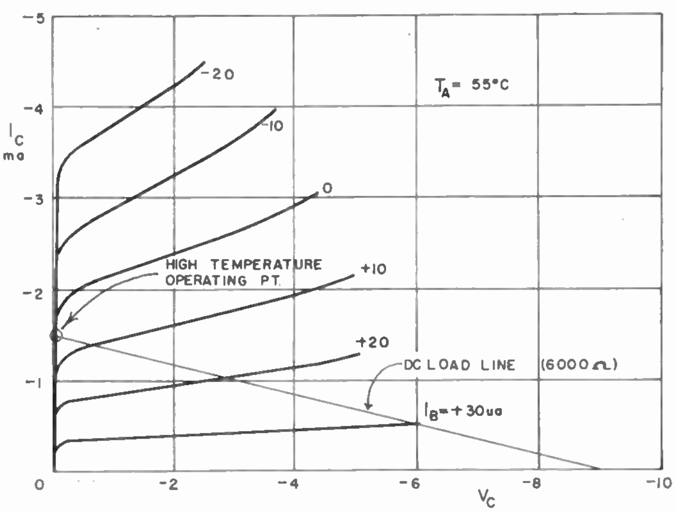


Fig. 10—Output characteristic for 55°C.

line (see Fig. 11). The factors that limit the extent of allowable operating-point shift are the signal level, the minimum allowable collector voltage at the left end of the dc load line (high temperature graph), and the minimum allowable collector current at the right end (low temperature graph).

The smallest allowable operating point collector current will occur at the low temperature.

$$I_{C1} = I_{min} + I_{sig} \tag{13}$$

where I_{C1} is the low temperature operating point collector current and I_{min} is the lowest permitted value of I_C. To choose I_{min} requires an inspection of the low temperature output, transfer and input characteristic curves. From the output curves (Fig. 11) we see that I_{min} must be at least equal to I_{CBO} if we are to allow for operation below 25°C. From the transfer curve (Fig. 3) we see that nonlinearity at low values of I_C is not great and frequently may be neglected. However, the nonlinearity of the input characteristics (Figs. 5 and 6) must not be overlooked as severe distortion may result

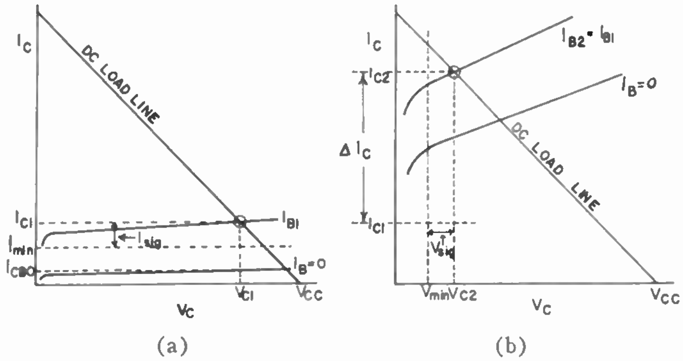


Fig. 11—Shift of operating point in a properly designed unstabilized RC amplifier. (a) Low temperature graph. (b) High temperature graph.

from this nonlinearity. As mentioned previously, the effects of the input circuit nonlinearity may be swamped out by driving the amplifier from a high-impedance source, *i.e.*, high impedance with respect to r_{in}. The lower limit on the collector current can be determined only by considering the input characteristics along with the source resistance, R_g, and the amount of allowable distortion. If the source resistance is comparatively low, it may be necessary to employ an unby-passed resistor in series with either the emitter or the base to swamp out the nonlinearity of the emitter junction.

Once I_{min} and I_{C1} have been determined, we go to the low temperature output characteristic curves [Fig. 11(a)] and determine the value of the base bias current required to give I_{C1}. Since I_C is not completely independent of V_C, the choice of I_{B1} is not exact, but if chosen near $V_C=2/3V_{CC}$, the approximation will be quite accurate at least for transistors with high collector resistance.

Making use of the fact that the voltage between emitter and base is very small compared to V_{CC}, the base bias resistor may be calculated as

$$R_B = \frac{V_{CC}}{I_{B1}}. \tag{14}$$

Now consider the high temperature operating point. The smallest value of the operating point collector voltage is

$$V_{C2} = V_{min} + V_{sig} \tag{15}$$

where V_{min} is determined by the nonlinearity of the high temperature output characteristics [Fig. 11(b)] at low values of collector voltage. The high temperature operating point is on the curve $I_B=I_{B2}=I_{B1}$ at V_{C2}. The value of I_{C2} can now be read from the graph and the dc load resistance (equal to R_C) calculated as

$$R_C = R_{dc} = \frac{V_{CC}-V_{C2}}{I_{C2}} \tag{16}$$

and the dc load line drawn in on the low and high temperature graphs.

The actual value of signal voltage and current in r_L will be less than the original value assumed for V_{sig} and I_{sig} by a factor that depends on the relative values of R_C and r_L. A redesign based on the first can be used to increase the signal level that can be handled. The increase in signal-handling capacity will be at the expense of reduced gain and higher power consumption.

Illustrative Example I

Purpose

1) To illustrate the design of an unstabilized RC amplifier that can be made to perform properly over a specified temperature range.
2) To observe the effects of signal level and input nonlinearity on distortion.

Given

Transistor with characteristics of Figs. 3, 5, 6, 12, and 13; circuit diagram of Fig. 1; maximum temperature of operation of 45°C; and

$$r_L = 1000 \text{ ohms} \qquad V_{sig}=0.5 \text{ volt}$$
$$r_g = 5000 \text{ ohms} \qquad I_{sig}=0.5 \text{ ma.}$$
$$V_{CC} = -9 \text{ volts}$$

Solution

1) From Fig. 12, we see that $I_{CBO}=-0.14$ ma near $V_C=-6$ volts, but the 25°C curve of Fig. 6 shows that the input resistance rises sharply for small values of I_C. If I_C is not allowed to go below about 1 ma, the variation in r_{in} is not great, and the input circuit distortion would be held to a low value. However, here we will deliberately choose I_{min} a little lower than dictated by fidelity so as to illustrate the effect of input circuit distortion. Choose $I_{min}=-0.5$ ma and then,

$$I_{C1} = I_{min} + I_{sig} = -0.5 - 0.5 = -1 \text{ ma.}$$

2) From Fig. 12, we see that a base current of about -23 μa is required for an I_C of 1 ma. Thus

$$R_B = \frac{V_{CC}}{I_B} = \frac{-9}{-23\times 10^{-6}} = 391{,}000 \text{ ohms.}$$

3) From Fig. 13, we observe V_{min} to be about -0.5 volt, and thus

$$V_{C2} = V_{min} + V_{sig} = -0.5 - 0.5 = -1 \text{ volt.}$$

At the point, $V_{C2}=-1$ volt and $I_B=-23$ μa, $I_{C2}=-2$ ma and therefore,

$$R_C = \frac{V_{CC}-V_{C2}}{I_{C2}} = \frac{-9+1}{-0.002} = 4000 \text{ ohms.}$$

4) The dc and ac load lines may then be drawn on Figs. 12 and 13 and the low and high temperature operating points indicated.

The amplifier described in the foregoing example was constructed and tested with the following results:

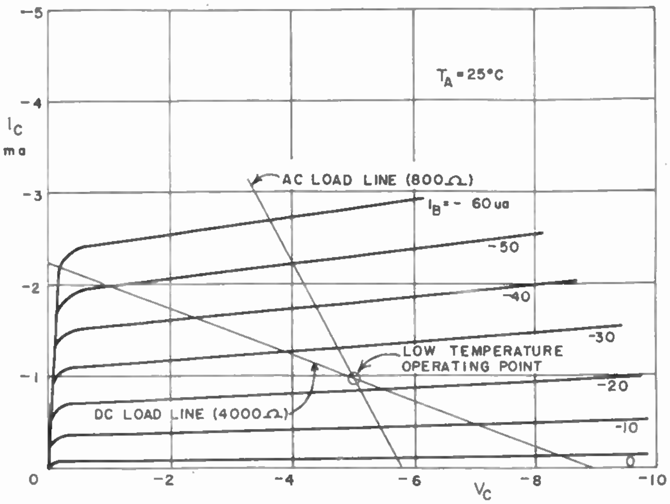


Fig. 12—25°C output characteristic for Example I.

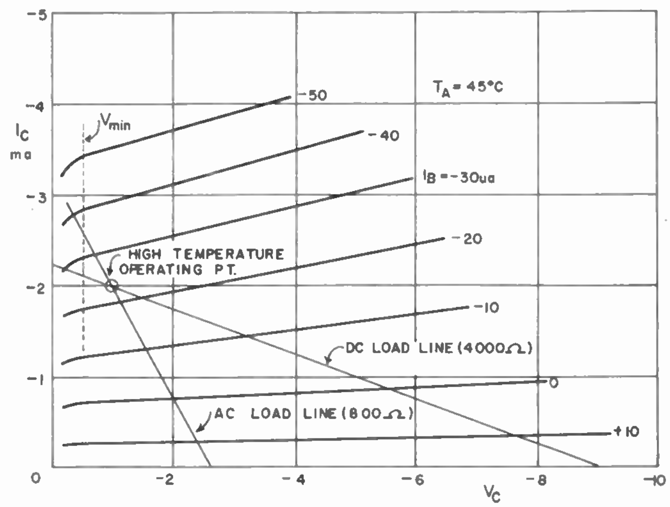


Fig. 13—45°C output characteristic for Example I.

$T_A = 25°C$	$T_A = 45°C$
$I_{C1} = -1$ ma	$I_{C2} = -2$ ma
$V_{C1} = -5$ volts	$V_{C2} = -1$ volt
$I_{B1} = -23$ μa	$I_{B2} = -23$ μa
$I_g = 18$ μa (max)	$I_g = 15.6$ μa (max)
$I_L = 500$ μa (max)	$I_L = 500$ μa (max)
$G = 28$	$G = 32$
Distortion = 5.8 per cent	Distortion = less than 1 per cent
Battery drain = 1 ma.	Battery drain = 2 ma.

Note the 5.8 per cent distortion at 25°C as compared to less than 1 per cent at 45°C. The reason for this is obvious when one inspects the input characteristic of Fig. 6. The input resistance for $T = 45°C$ and $I_C = -2$ ma is more nearly constant than for $T = 25°C$ and $I_C = -1$ ma. To demonstrate this input circuit nonlinearity, the generator resistance, r_g, was increased to 500,000 ohms and the distortion at 25°C decreased to 3.6 per cent.

Recall that the design was based on the assumption that the collector signal current and voltage were the same as the signal current and voltage in the load, r_L. It can be determined from the low temperature graph (Fig. 12) that the maximum output signals to be handled according to the stated value of I_{min} are $V_{sig} < 0.4$ volt and $I_{sig} < 0.4$ ma. Accordingly, the distortion at 25°C was measured with $V_{sig} = 0.3$ volt and found to be 1.6 per cent with $r_g = 500{,}000$ ohms and 3.4 per cent with $r_g = 5000$ ohms. The remaining 1.6 per cent distortion may be attributed to the nonlinearity of the transfer and output characteristics.

It is possible to achieve satisfactory performance from an unstabilized amplifier circuit. However, if the gain, distortion or battery drain are of primary importance, then the signal level, temperature variation, and production spread of parameters must not be great.

The Stabilized RC Amplifier

In the unstabilized RC amplifier, no attempt is made to check the increasing collector current as the temperature rises. The circuit is merely designed to take the variation in operating point into account at the expense of reduced gain (due to a low value of R_C) and high battery drain at the high temperature. The change in I_C in the unstabilized amplifier is designated as ΔI_C in Fig. 11.

In the stabilized circuit, the collector current is allowed to rise only a fraction of ΔI_C. In Fig. 14, ΔI_C has been divided into two parts: $S\Delta I_C$ and $(1-S)\Delta I_C$, where S is a circuit stabilization factor.[8] Zero stabilization corresponds to the unstabilized amplifier, and unity or 100 per cent stabilization indicates that I_C would not change as the temperature rises to its maximum value. From the standpoints of gain and battery drain there may be an optimum value of S for a particular circuit.

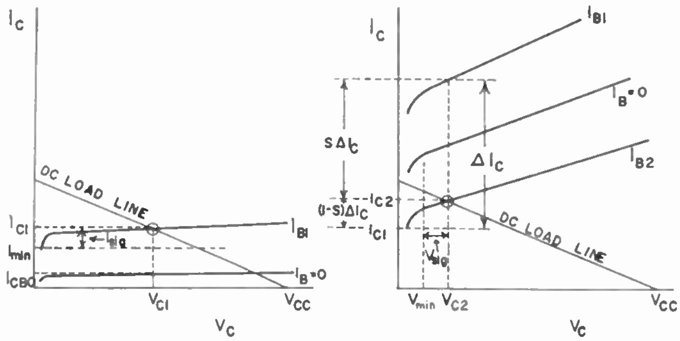
The most commonly employed stabilization circuitry is shown in Figs. 2 and 15. Here dc negative feedback from the emitter resistor causes the base bias current to be dependent on the collector current. Note that the effect of I_C in R_E is to produce a "source" voltage in the base circuit which subtracts from V_{CC}. Thus as I_C increases with temperature, the voltage across R_E increases and reduces the base bias current. The bias current may even reverse direction at high temperatures.

It can be shown that the stabilization factor for the circuit of Fig. 15 is given by[8]

$$S = \frac{\beta}{\beta + 1 + \left[R_E\left(\frac{1}{R_S} + \frac{1}{R_B}\right)\right]^{-1}}. \tag{17}$$

So if it is desired that the variation in I_C be held to a small value, a high degree of stabilization requires a large value of R_E and small values of R_B and R_S. But both of these requirements have some adverse effects:

[8] R. P. Murray, "Systematic design of transistor bias circuits," *Electronic Indus. and Tele-Tech*, vol. 16, pp. 75–77, 147–148; November, 1957.



(a) (b)

Fig. 14—Shift of operating point in a stabilized RC amplifier. (a) Low temperature graph. (b) High temperature graph.

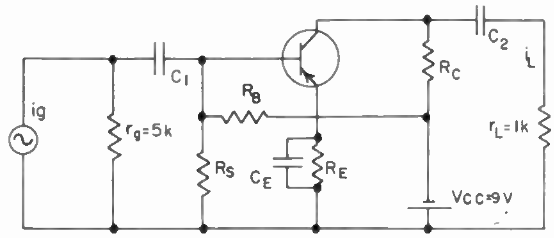


Fig. 15—Stabilized common-emitter amplifier of Example II.

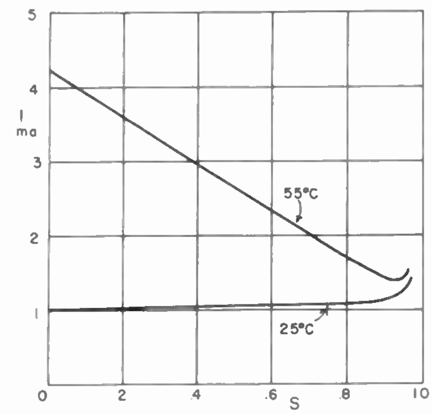


Fig. 16—Effect of S on battery drain. Applies to Example II.

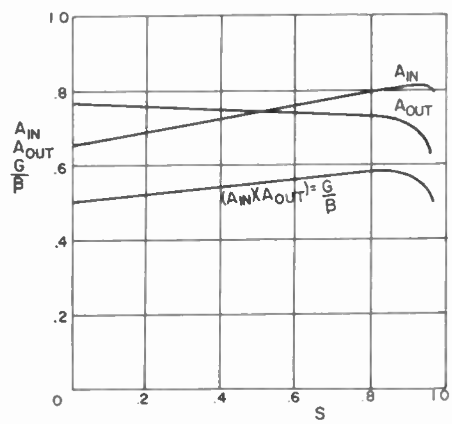


Fig. 17—Effect of S on gain. Applies to Example II.

small values for R_B and R_S reduce the gain, increase the battery drain, and increase the distortion caused by the nonlinearity of the input circuit. A high value of R_E means that for a given V_{CC}, R_C will have to be reduced accordingly since

$$R_{dc} = R_C + R_E \tag{18}$$

if we make the approximation that $I_E \cong I_C$. A reduced value for R_C means less gain.

Fig. 16 shows the computed battery drain as a function of S for the circuit of Fig. 15. Note that the battery drain at the low temperature is essentially equal to I_{C1} (I_{C1} does not change with S) until the effect of the bleeder current in R_S becomes significant with respect to I_{C1}. On the other hand, the high temperature battery drain is greatly affected by S since low values of S indicate that I_{C2} will be high. A minimum in high temperature battery drain occurs when the decreasing collector current (due to increasing S) is offset by an increasing bleeder current in R_S. Thus from the standpoint of battery drain, S should be chosen slightly less than that value at which the minimum in high temperature battery drain occurs.

From the standpoint of circuit gain, the optimum value for S occurs when the increase in gain due to the use of a high R_C and low R_E is offset by the decrease in gain due to the shunting effect of R_B and R_S. Fig. 17 shows the two attenuation factors and the total gain (in terms of β) for the amplifier of Fig. 15.

Another factor to be taken into account in the choice of S is the collector dissipation. Low values of S mean high dissipation at high temperatures.

We now turn to the circuit design. In Fig. 14 choose V_{C2} and I_{C1} in accordance with the nature of the transistor characteristics and signal level as before. Next, in accordance with the foregoing discussion, we select a stabilization factor or operating-point current on the high temperature graph. Then

$$R_{dc} = \frac{V_{CC} - V_{C2}}{I_{C2}}. \tag{19}$$

We must now decide how much of R_{dc} to allot to R_E. If R_E is chosen low, then the output attenuation factor will be high due to a high R_C, but both R_S and R_B will be low and reduce the input attenuation factor. The proper value for R_E can be found by maximizing the circuit gain with respect to R_E. Using the approximation that $V_B \cong 0$, the equations for the emitter-base circuit (Fig. 15) are

$$V_{CC} = I_S(R_B + R_S) + I_B R_B \tag{20}$$

$$0 = I_C R_E + I_B R_E - I_S R_S. \tag{21}$$

Eliminating I_S from (20) and (21) and solving for R_B gives

$$R_B = \frac{V_{CC} - (I_B + I_C)R_E}{\frac{R_E}{R_S}(I_B + I_C) + I_B}. \tag{22}$$

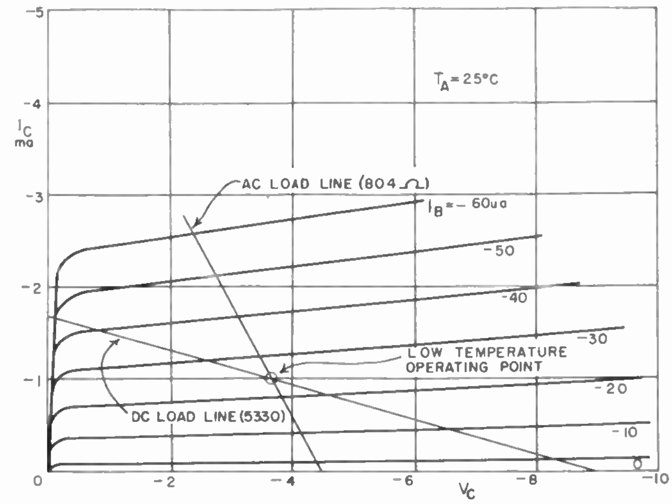


Fig. 18—25°C output characteristic for Example II.

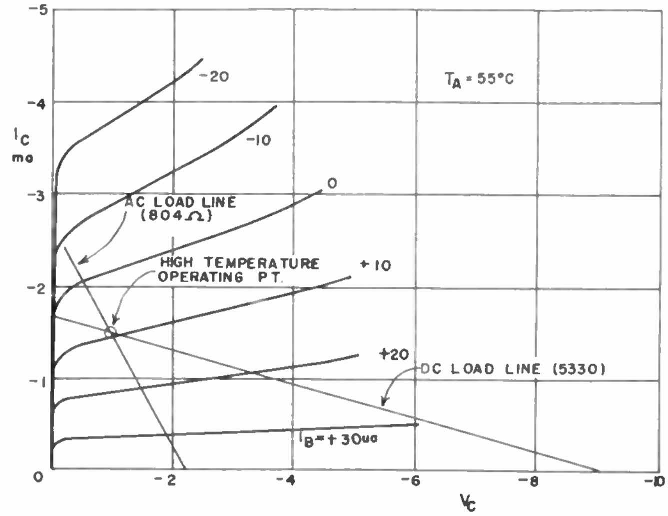


Fig. 19—55°C output characteristic for Example II.

Making two equations from (22) for the two conditions of temperature, and eliminating R_B, we get

$$R_S = \frac{V_{CC}R_E[(I_{C2} - I_{C1}) - (I_{B1} - I_{B2})]}{V_{CC}(I_{B1} - I_{B2}) - R_E(I_{B1}I_{C2} - I_{B2}I_{C1})} \tag{23}$$

and in terms of S,

$$R_S = \frac{V_{CC}R_E[(1 - S)\Delta I_C - \Delta I_B]}{V_{CC}\Delta I_B - R_E(I_{B1}I_{C2} - I_{B2}I_{C1})} \tag{24}$$

where $(1-S)\Delta I_C$ and ΔI_B are the changes in I_C and I_B as the temperature goes from low to high (see Fig. 14). Solving for R_B

$$R_B = \frac{V_{CC}[(1 - S)\Delta I_C - \Delta I_B)}{I_{B1}I_{C2} - I_{B2}I_{C1}}. \tag{25}$$

The gain of the circuit of Fig. 15 is given by (7). Substituting (18), (24), and (25) into (7) gives the gain as

$$G = \frac{\beta}{\left(1 + \dfrac{r_L}{R_{dc} - R_E}\right)\left[1 + r_{in}\left(\dfrac{1}{r_g} + \dfrac{\Delta I_B}{R_E[(1 - S)\Delta I_C - \Delta I_B]}\right)\right]}. \tag{26}$$

Now assuming R_{dc} to be constant (determined by choice of S), taking the derivative of the gain (26) with respect to R_E, and equating to zero we get

$$R_E = \frac{R_{dc} + r_L - \sqrt{(R_{dc} + r_L)\left[r_L + \left(\dfrac{(1 - S)\Delta I_C}{\Delta I_B} - 1\right)\left(\dfrac{1}{r_{in}} + \dfrac{1}{r_g}\right)(R_{dc}r_L)\right]}}{1 - r_L\left(\dfrac{(1 - S)\Delta I_C}{\Delta I_B} - 1\right)\left(\dfrac{1}{r_{in}} + \dfrac{1}{r_g}\right)}. \tag{27}$$

We now have relations for the four circuit resistors and three capacitors as follows:

R_E from (27)	C_E from (9)
R_C from (18)	C_1 from (10)
R_S from (24)	C_2 from (11).
R_B from (25)	

The actual resistances of the circuit resistors will not be these calculated values, but the nearest available values. Some of these resistance values are more critical than others. For example, it would be better to choose R_E nearest to its calculated value and take up the slack in R_C.

Illustrative Example II

Given

Circuit of Fig. 15, using transistor described by curves of Figs. 3, 5, 6, 18, and 19. Maximum temperature of 55°C and 30 cps for the low frequency at which the response is down 3 db.

$r_L = 1000$ ohms $\quad V_{sig} = 0.5$ volt

$r_g = 5000$ ohms $\quad I_{sig} = 0.5$ ma.

$V_{CC} = -9$ volts

Solution

1) As in 1), 2), and 3) of Example I, we have

$$I_{C1} = -1 \text{ ma}$$
$$I_{B1} = -23\ \mu\text{a}$$
$$V_{C2} = -1 \text{ volt.}$$

2) From Fig. 19, choose an appropriate value for I_{C2}, keeping in mind the effect of S on gain and battery drain (Figs. 16 and 17). Here we will choose $I_{C2}=-1.5$ ma which corresponds to $S=0.843$. Then,

$$R_{dc} = \frac{V_{CC} - V_{C2}}{I_{C2}} = \frac{-9+1}{-0.0015} = 5330 \text{ ohms.}$$

3) From (27), (18), (24), and (25) and using $r_{in}=1500$ ohms, we have

$$R_E = 1245 \text{ ohms}$$
$$R_C = 4085 \text{ ohms}$$
$$R_S = 22{,}000 \text{ ohms}$$
$$R_B = 95{,}600 \text{ ohms.}$$

4) If the gain is to be 3 db down at 30 cps, this corresponds to $P_s=0.707$. If there are a total of three low-frequency calculations in the entire circuit, we have from (12) $P=0.89$. From (9)–(11) we get

$$X_{CE} = 38.1 \text{ ohms} \quad (\text{From Fig. 19, } \beta = 71)$$
$$X_{C1} = 3270 \text{ ohms}$$
$$X_{C2} = 2600 \text{ ohms}$$

and at 30 cps, these reactances correspond to capacitances of

$$C_E = 139 \ \mu f$$
$$C_1 = 1.62 \ \mu f$$
$$C_2 = 2.04 \ \mu f.$$

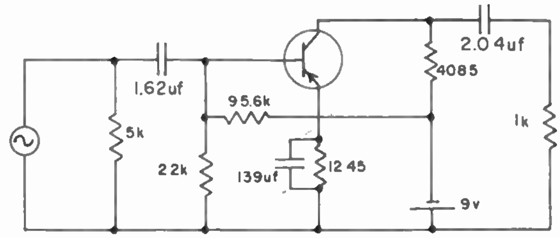


Fig. 20—Stabilized amplifier of Example II.

The amplifier described in the foregoing example, shown in Fig. 20, was constructed and tested with the following results:

$T_A=25°C$	$T_A=55°C$
$I_{C1}=-0.93$ ma	$I_{C2}=-1.44$ ma
$V_{C1}=-4$ volts	$V_{C2}=-1.2$ volts
$I_{B1}=-22$ μa	$I_{B2}=+9$ μa
$I_g=13$ μa (max)	$I_g=9.38$ μa (max)
$I_L=300$ μa (max)	$I_L=300$ μa (max)
$G=23$	$G=32$
Distortion = 3.4 per cent	Distortion = 1.86 per cent
Battery drain = 1 ma.	Battery drain = 1.5 ma.

As in Example I, the distortion at 25°C may be reduced by choosing a slightly higher value for I_{C1}, and the distortion at both temperatures may be reduced by increasing the resistance in the base-emitter circuit. The signal handling capability may be increased by a redesign in which V_{C2} and I_{C1} are increased.

Correspondence

Binaural Speaker Listener Tests*

While touring some of the more recent electronics and audio exhibits, some binaural and stereophonic sound setups were heard with considerable interest. One, in particular, of a wide-band recording of locomotive yard sounds was rather startling to one who has been away from "HIGH FIDELITY" for some time.

A rather intriguing point was then brought up and debated rather inconclusively. A two-speaker-binaural vs one-speaker-monoaural comparison-listening test was being used, as with most of these types of exhibit demonstrations. Is this actually a fair basis for comparison? In the binaural setup, not only have we two point sound sources, but the wattage level is distributed between two banks of speakers. A speaker's distortions nose dive as its wattage input is halved. Would a fairer comparison test be made if in switching from binaural to monoaural, the single-channel signal were to be fed into *both* speaker channels, or at least, to *two* equivalent speaker banks, side by side?

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* Received by the PGA, June 5, 1958.

Contributors

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Mr. Chavasse is professor at the Ecole Nationale Supérieure des Télécommunications, lauréate of the Institut de France (Académie des Sciences), member of the International Electrotechnique Commission and of the I.S.O., past president of the Electroacoustic Group of the Société des Radioélectriciens, and founder and general secretary of the Société des Acousticiens de Langue Française (1948).

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Réné G. Lehmann was born on January 7, 1921, in Paris, France. He studied at the Faculté des Sciences of the Université de Paris, and obtained the degree of licencie-es-sciences (Master's degree) in physics in 1943. He has held the position of engineer with the Department of Acoustics of the Centre National d'Etudes des Télécommunications since 1945. He is a member of Committee 29 of the International Electrotechnique Commission and Committee 43 of the International Standards Organization.

Mr. Lehmann is one of the charter members of the Société des Acousticiens de Langue Française, organized in 1948, and since that time has been its associate secretary. He is also secretary of the Electroacoustic Group of the Société des Radioélectriciens.

❖

M. Copel

Michel Copel (M'53–SM'57) was born on March 20, 1916, in Paris, France. He received the B.S. degree in 1935 from the University of Paris, and the E.E. degree in 1937 from the Conservatoire National des Arts et Mêtiers, Paris. Mr. Copel also attended New York University, New York.

From 1942 to 1946 he was engaged in the design and development of military loudspeaker equipment as chief design engineer of University Loudspeakers, Inc. He was senior engineer at Dictograph Products, Inc., from 1946 to 1948.

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❖

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Ray P. Murray (M'57) was born on November 14, 1920, in Summerfield, Kan. He attended Kansas State College, Manhattan, Kan., from 1938 until 1940, at which time he entered the United States Navy. He served in a wide variety of assignments dealing with electronics and electronics training until his release in 1946. He then returned to Kansas State College to complete his training and received the B.S. degree in electrical engineering in 1947. In 1954, he received the M.S. degree from Brown University, Providence, R. I.

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