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on AUDIO

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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

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The Editor's Corner

Overspecification

THOUGH Russian scientists achieved notable success in other fields, they never did much with magnetic recording. They haven't even claimed that it was invented there, presumably by Vladermarski Poulsenovich. But we often thought about a golden opportunity they missed for "getting in on the ground floor."

During the last world war, Armour Research Foundation was engaged in small-scale production of Model 50 wire recorders. At the time, these were considered as one of the modern miracles and there were never enough even for the highest priority military uses. Yet, our spirit of international co-operation was such that foreign allies could obtain higher priorities here than we could, and it came to pass that Russian agencies got authority to buy some Model 50 recorders.

But their red tape required a set of specifications. Russian specification writers in those days must have been very much like our own, and they really came up with something that would have gladdened the heart of a Hi-Fi copywriter. The response was flat from subsonic to ultrasonic frequencies; noise level inaudible; wow and flutter a microscopic fraction of a per cent; distortion likewise—all in a rugged portable package of about half a cubic foot in volume.

We were proud of our recorders, but they weren't *that* good. We notified the agency that our machines could not meet their specs. Apparently their attitude was one of "take it or leave it" and since we were quite busy, we "left it." There must be a moral to this story somewhere, but the practical effect was that we sent none of our machines to Russia.

Why not share your most interesting stories relating to Audio with others? Send them to "The Editor's Corner," c/o Marvin Camras, Armour Research Foundation, 10 W. 35th St., Chicago 16, Ill.

—Marvin Camras, *Editor*

PGA News

ADMINISTRATIVE COMMITTEE MEETING MINUTES

New York, N.Y., March 23, 1959

Members Present

B. B. Bauer
S. J. Begun
A. B. Bereskin
M. Camras
M. S. Corrington
J. K. Hilliard
H. F. Olson
F. H. Slaymaker
P. B. Williams

Members Absent

W. T. Selsted

Guests Present

M. Copel
M. Duckworth
R. A. Heising

Newly Elected Members Present

P. C. Goldmark
J. R. Macdonald

The meeting was held at the Waldorf-Astoria and it was called to order at 8 P.M.

1) Approval of the Minutes of the Last Meeting

By motion of Mr. Corrington, seconded by Dr. Olson, it was agreed to approve the Minutes of the previous meeting.

2) Result of the Election

The final results of the election, as reported by the Headquarters, were read by the Secretary as follows:

A. B. Bereskin—Chairman, 1959–1960
J. K. Hilliard—Vice-Chairman, 1959–1960

The newly elected Administrative Committee members are as follows:

P. C. Goldmark—1959–1962
J. R. Macdonald—1959–1962

The letter from the Headquarters informing of the results of the election is attached to these Minutes (Appendix A).

The Secretary pointed out, that, in accordance with the newly adopted Bylaws, a third member of the Administrative Committee shall have been elected to the Committee, while the newly elected Chairman would not be a member of the Administrative Committee during his term of office. However, this is in conflict with the Constitution. Considerable discussion ensued in which Mr. Heising participated. It was decided that the Constitution would govern and that a third member would not be declared elected until such time as the conflict between the Constitution and the Bylaws shall have been resolved.

By motion of Mr. Williams, seconded by Mr. Hilliard, it was resolved to instruct the Constitutional Committee to propose a revision of the Constitution and Bylaws so that they are consistent with each other. It was further resolved that this proposed revision provide for an Administrative Committee consisting of ten members, one of them being the Chairman, and the Chairman shall not have a vote, except in the case of a tie.

3) Awards Committee Report

Chairman Williams proposed the following awards:

PGA Senior Award—M. S. Corrington and T. Murakami
Achievement Award—D. W. Martin

By motion of Mr. Williams, seconded by Mr. Bereskin, this report was accepted unanimously.

4) Chapters Committee Report

Dr. Macdonald presented the report of the activities of the Chapters Committee, which is attached hereto (Appendix B).

5) Editorial Committee Report

Mr. Camras presented the report, which is attached hereto (Appendix C).

6) Program Committee Report

Dr. Begun presented the report.

7) Ways and Means Committee Report

Mr. Copel presented the report, which is attached hereto (Appendix D).

8) Finance Committee Report

Mr. Bauer presented the report, which is attached hereto (Appendix E).

By motion of Mr. Williams, seconded by Mr. Bereskin, it was agreed that the PGA would sponsor a meeting at the Audio Sessions of the NEC in Chicago, during the Fall of 1959.

Newly elected Chairman Bereskin has announced the following appointments:

Secretary-Treasurer—B. B. Bauer, 1959–1960
Chairman, Committee on Constitution and Bylaws—M. S. Corrington
Chairman, Chapters Committee—J. R. Macdonald
Chairman, Nominating Committee—F. H. Slaymaker
Chairman, Tapescripts Committee—A. B. Jacobsen
Chairman, Editorial Committee—M. Camras

By motion of Mr. Hilliard, seconded by Mr. Williams, the Administrative Committee approved these appointments unanimously.

B. B. BAUER
Secretary-Treasurer

APPENDIX A

A number of additional ballots have been received at Headquarters for the election of officers and Administrative Committee members for the PGA since I sent you the results of the ballot on February 5.

We have added this count to the one taken on February 5 and the final results of the election are as follows (arranged according to highest number of votes):

For Chairman and Vice-Chairman:

A. B. Bereskin
J. K. Hillard
S. J. Begun
P. B. Williams

For Administrative Committee Members:

P. C. Goldmark
J. R. Macdonald
W. B. Snow
R. W. Benson
I. Kerney
M. Copel

A total of 1286 ballots were returned to Headquarters by the membership.

In accordance with Article II, Section 7, Paragraph 4 of the PGA Bylaws as follows:

> "The Chairman thus elected shall not be a member of the Administrative Committee during his term of office and if he is serving an unexpired term at the time he takes office his membership shall terminate"

it would appear that Mr. Bereskin does not have to be an elected member of the Administrative Committee and that Messrs. Goldmark, Macdonald and Snow will be able to serve for a three-year term. Please let us know whether our interpretation of the Bylaws is correct and whether we can go ahead and notify them of their election. (Editor's Note: As discussed at the meeting, only Mr. Goldmark and Mr. Macdonald are declared elected until the Constitution and Bylaws are changed.)

GENE B. DUFFY
Assistant to the Technical Secretary

APPENDIX B

REPORT OF CHAPTERS COMMITTEE

One issue of "The PGA Chapter," vol. 4, no. 1, January 26, 1959 has been sent to all PGA Chapter Chairmen. Two letters (October 29, 1958, and January 8, 1959) were sent to all Chapter Chairmen urging them to keep the Chairman of the Chapters Committee informed of their meetings and thanking them for doing so.

Two letters (November 14, 1958, and February 24, 1959) were sent to the 25 Chairmen of IRE Sections with 24 or more PGA members but no PGA Chapter urging them to initiate the information of such a chapter. Negative results have been received from the Long Island Section but a Chapter has been formed in the Twin Cities Section. Formation of PGA Chapters in the Rome-Utica, Los Angeles, and New York Sections is currently under consideration.

Four summarizing writeups of Chapter news for use in the PGA Transactions have been sent to Marvin Camras.

No Chapter meeting response has been received from Baltimore, Boston, Cincinnati, Hawaii, Houston, Milwaukee, Philadelphia, Phoenix, San Antonio, or Washington.

J. ROSS MACDONALD
Chairman

APPENDIX C

REPORT OF THE EDITORIAL COMMITTEE

Objectives set by the Editorial Committee last year were:

1) To become "current" in issuing the journal.
2) To increase the number and quality of published papers.
3) To add features and departments for interest and balance.
4) To systemize the procedures for obtaining and processing papers.

Lack of publishable papers has been the perennial problem. After much negotiation we are now fortunate in being able to accept IRE convention papers, whereas in past years these were printed in the CONVENTION RECORD only. To obtain additional papers, we tried members of the Editorial Committee, chairmen of program committees, etc., but the most successful approach was direct correspondence with authors. This has been done almost entirely by the editor in the past, but it is hoped that other members of the editorial committee will help in the coming year. Currently we are caught up, with enough papers for two more issues. IRE headquarters now takes ten weeks or more to process an issue. We are trying hard to overcome the effects of this additional delay.

Beginning with the January–February 1959 TRANSACTIONS, we are including "The Editor's Corner," a place for editorials, stories, letters, guest comments, etc. The idea is to present sidelights on audio that are not found in formal articles and PGA news. In addition, we have tried to obtain guest editorials by authorities, and reviews of other audio journals, both domestic and foreign.

To keep our authors up to date, we printed postcards which are dispatched immediately upon receipt of a manuscript or other material. We also make some entry forms for papers and an "advertising" letter outlining the advantages to an author when he publishes in TRANSACTIONS.

In the coming year we would like to have enough papers to expand the journal and to make it more interesting. I would like to extend my thanks to editorial committee members and others who have helped to improve these TRANSACTIONS.

MARVIN CAMRAS
Editor

APPENDIX D

REPORT OF WAYS AND MEANS COMMITTEE

This report covers the period from March 1, 1958 to February 28, 1959.

1) Thirteen (13) institutional listings which expired during this period were disposed of as follows:

7 renewals paid for
5 were dropped (Audiophile, Electro-Voice, Fairchild, Sonotone and Telex)
1 is now due
—
13

2) Total receipts up to date $525.00.

3) In last year's annual report this Committee reported that the arrangement initiated and carried during the 1957 to 1958 period by which the Technical Secretary of IRE handles the renewal notifications for listings in the IRE TRANSACTIONS ON AUDIO did not work out satisfactorily. The same situation occurred this year. IRE only notifies the Committee when payment is received for renewal. If payment is not received, listing is automatically dropped, and the Committee only finds out when the next issue of TRANSACTIONS comes out, which usually is several months after the due date. The Committee therefore does not have the opportunity to send a personal letter to the subscriber. The Committee again recommends that, if possible, we revert to the old system whereby the Ways and Means Committee would handle notifications and renewals as well as new subscribers.

4) The records of the Committee are in good order.

MICHEL COPEL
Chairman

APPENDIX E

REPORT OF THE FINANCE COMMITTEE

The financial state of the IRE Professional Group on Audio is good and no change in financial policy is indicated. We have ended the year of 1958 with a balance in the treasury of $19,818.33, as compared with a balance at the end of the year 1957 of $15,976.79. The new balance is more than twice our current yearly expenses, which provides a good cushion against possible vicissitudes and places us in a good position to be of further service to the IRE and its members.

While the balance has been growing rapidly during the past several years this growth will slow down in the future. This is due to the recent change in the financial policy of the IRE. Instead of matching funds on basis of $1.00 per member the new policy is to reimburse the group for ⅓ of its publication expenses. As an example of the effect of the new policy, matched funds in 1958 were in the amount of $3981.00, indicating that 3981 members had paid their dues during that year. During 1958 our publication expense was $8562.61. Assuming that the publication expense will be $9000.00 in 1959, the support to be expected from the IRE is $3000.00.

One matter of concern is a drop in membership at the end of 1958, which we ended with 3736 paid members, as against 3920 on June 30, 1958; or a loss of 144 members. While this is a negligible loss in percentage, it is obvious that dynamic growth is lacking. Fortunately, the student member registration has increased during the year from 507 to 612 proving that there is an increasing interest in Audio among the upcoming generation of radio engineers. The Administrative Committee should give a careful thought to this matter and decide what should be done about it. Have we reached our proper level within the IRE, or should we attempt to further increase our membership, as by an intensified publication policy, services to members, prizes and awards, lectures and symposia, etc.?

I have prepared for your guidance an estimated budget for 1959. This budget is based upon the new form of matched funds, the present number of members, and the same income from advertising and sale of publications as last year. It assumes a slight increase in publication expense, and an increase of about $1000–$1200 in services of all sorts to the members. We have the money to do a lot of good, but, of course, we must be careful how it is spent.

BENJAMIN B. BAUER
Chairman

ESTIMATED BUDGET

From January 1 to December 31, 1959

Balance from December 31, 1958		$19,818.33
Estimated Receipts during Period		
IRE Matched Funds	$ 3,000.00	
Fees	8,084.00	
Advertising	675.00	
Sale of Publications	521.00	
Total Receipts	$12,280.00	
Total Balance and Receipts		$32,098.33
Estimated Expenses During Period		
Publications	$ 9,000.00	
Membership Service Charges	500.00	
Group Awards	500.00	
Editorial Administrative Exp.	500.00	
Others	900.00	
Total expenses	$11,400.00	$11,400.00
Estimated Balance as of December 31, 1959		$20,698.33

CHAPTER OFFICERS AND MEMBERSHIP STATISTICS

The IRE Professional Group on Audio includes the following twenty-two chapters. Paid members are as of December 31, 1958. Officers are listed from the latest information received by IRE headquarters to September, 1959.

Chapter	Chairman	Vice-Chairman	Secretary	Paid Members*
Baltimore 1959–1960	Ted N. Truske Westinghouse Elec. Air Arm Division P.O. Box 746 Baltimore 3, Md.	Louis R. Mills Recordings, Inc. 735 Deepdene Rd. Baltimore 10, Md.	James H. Jackson Guidance & Navig. The Martin Co. Baltimore 3, Md.	69
Boston 1959–1960	Donald J. Fritch Lessells & Assoc. 916 Commonwealth Ave. Boston 15, Mass.		Henry S. Littleboy Baird-Atomic Corp. 33 University Ave. Cambridge, Mass.	272
Chicago 1958–59	Robert J. Larsen Jensen Mfg. Co. 6601 S. Laramie Ave. Chicago 38, Ill.	Leonard Eckmann Gen. Tel. Labs. P.O. Box 17 Northlake, Ill.	William M. Ihde Gen. Radio Co. 6605 W. North Ave. Oak Park, Ill.	221
Cincinnati 1958–1959	William C. Wayne, Jr. 70 Pleasant Ridge Ave. S. Ft. Mitchell, Ky.	Clyde G. Haehnle Crosley 140 W. 9 St. Cincinnati, Ohio	John P. Quitter 3837 Broadview Dr. Cincinnati, Ohio (also Treasurer)	53
Cleveland 1958–1959	Kenneth Hamann Clev. Record. Co. 1515 Euclid Ave. Cleveland 15, Ohio			54
Dayton 1959–1960	E. O. Valentine Comm. & Nav. Lab. Wright-Patterson AFB Dayton, Ohio	Miles McLennan 304 Schenck Ave. Dayton, Ohio	Don Stouch Wright Air Dev. Wright-Patterson AFB Dayton, Ohio	46
Detroit Inactive				61
Hawaii 1958–1959	D. H. DaShiell 224 Awakea Rd. Lanikai, Oahu	Dr. Iwao Miyake Univ. of Hawaii Honolulu	Daniel L. Pang 1809 Naio St. Honolulu (also Treasurer)	20
Houston 1958–59	W. C. Wrye, Jr. 2410 W. Alabama Houston, Tex.			46
Kansas City Inactive				28
Los Angeles Inactive				377
Milwaukee 1959–1960	Tim Houle 1000 S. 56 St. West Allis, Wisc.	Otto Dobnick 113 S. 28 St. Milwaukee, Wisc.		51
Philadelphia 1959–1960	R. Michael Carrell RCA Camden, N. J.	T. A. Benham Haverford College Haverford, Pa.	Roy S. Fine 412 Spruce St. Haddonfield, N. J.	263
Phoenix Inactive				33
San Antonio 1958–1959	Bill Case Case Records & Sound San Antonio, Tex.			24
San Diego 1959–1960	Hal R. Brokaw U. S. Naval Station Code 2133 San Diego, Calif.	Lowman Tibbals USNEL San Diego, Calif.		56
San Francisco 1958–1959	D. L. Broderick Hewlett-Packard Page Mill Rd. Palo Alto, Calif.	Lambert Dolphin, Jr. Stanford Res. Inst. Menlo Pk., Calif.	L. W. Johnson Hewlett-Packard Page Mill Rd. Palo Alto, Calif.	205
Seattle Inactive				65

* As of December 31, 1958.

CHAPTER OFFICERS AND MEMBERSHIP STATISTICS (*Cont'd*)

Chapter	Chairman	Vice-Chairman	Secretary	Paid Members*
Syracuse 1958–1959	W. R. Chynoweth 17 Memory Lane N. Syracuse, N. Y.	Archie McGee, Jr. 205 Augusta Dr. N. Syracuse, N. Y.	G. Grenier General Electric Co. Auburn, N. Y.	64
Twin Cities 1959–1960	Robert L. Sell Audio Dev. Co. Minneapolis, Minn.	Richard F. Dubbe Minnesota Mining 900 Bush Ave. St. Paul 6, Minn.	J. F. Dundovic Nortonics Inc. 1015 S. 6 St. Minneapolis, Minn.	60
Washington D. C. 1958–1959	W. B. Bernard Bureau of Ships Const. & 18 St., N. W. Washington, D. C.	H. P. Meissinger Lab. for Elec. Engrg. 625 N. Y. Ave., N. W. Washington, D. C.	Sachio Saito NBS Conn. @ Van Ness Washington, D. C.	177
Albuquerque-Los Alamos 1958–1959	G. R. Bachand Sandia Corp. Albuquerque, N. M.	C. W. Remaley Sandia Corp. Albuquerque, N. M.	(same as Vice-Chairman)	35

* As of December 31, 1958.

NEED TECHNICAL PAPERS?

Tapescripts can round out your Chapter Program this year!

Since 1952, the IRE-PGA has provided tape-recorded talks with slides to Chapters and Sections. Several hundred presentations of tapescripts have been made. The most popular are listed first. Student Chapters have made particularly good use of this material. The wide range of subjects available offers interesting possibilities for your Chapter.

Tapescripts are loaned by the IRE-PGA tapescripts committee.

Andrew B. Jacobsen, *Chairman*
5618 E. Edgemont,
Phoenix, Ariz.

Order several for review as the cost is only the return postage on the material.

The best way to utilize tapescripts is to have a qualified person review the material and be prepared to answer questions that may come up. It is of the utmost importance that those presenting a tapescript run through the material from a purely mechanical standpoint to make sure they have the copies expected and that they have the technical equipment to reproduce the sound and picture.

Technical standards for tapescripts are as follows: Sound on $7\frac{1}{2}$-inch per second, $\frac{1}{4}$-inch tape, full track on 7-inch reels. The slides are 2-inch×2-inch cardboard mounts, double 35-mm slides. In a few cases where a limited number of copies are available, $3\frac{1}{4}$-inch×4-inch slides are used.

The following tapescripts are available.

"Magnetic Recording," by Marvin Camras, Armour Research Foundation of Illinois Institute of Technology. Discusses fundamentals of wire and tapes, heads, bias, circuits, equalization, present problems, and future developments. Thirty minutes.

"Phonograph Reproduction," by B. B. Bauer, Shure Brothers, Inc. Grooves and needles, fidelity and efficiency, pickup arms, and recording reproducing characteristics are some of the subjects discussed in this one-hour recorded paper.

"Method for Time or Frequency Compression-Expansion of Speech," by Grant Fairbanks, W. L. Everitt, and R. P. Jaeger, University of Illinois.

"Push-Pull Single-Ended Audio Amplifier," by Arnold Peterson and D. B. Sinclair, General Radio Company. A convention paper presented by Dr. Peterson, twenty-five minutes. $3\frac{1}{4}$×4 slides.

"The Electrostatic Loudspeaker—An Objective Evaluation," by R. J. Larson, Jensen Manufacturing Co., Chicago, Ill. The condenser loudspeaker, following development of new materials and methods, is now practical for high-frequency use in multichannel loudspeaker systems. Theoretical and mechanical design considerations illustrate the limitations at this stage of the art, including inherently high distortion at high output levels, and inability to withstand overloads. Principal advantages are low cost and efficient reproduction at extremely high frequencies. Demonstration will illustrate the operation of a typical electrostatic "tweeter" in combination with lower-frequency dynamic loudspeaker channels.

"Efficiency and Power Rating of Loudspeakers," by R. W. Benson, Armour Research Foundation, Chicago, Ill. The specification of the performance of loudspeakers, the measurement of the response-frequency characteristic and the directional characteristics of loudspeakers are discussed. The method of using a reverberation chamber to integrate acoustic power output and thus determine the efficiency is discussed in comparison to the more tedious method of analytical integration of measurements performed in free space. Power-handling capabilities of loudspeakers are determined by distortion measurements.

"Sound Survey Meter," by Arnold Peterson, General Radio Company. A convention paper. Twenty minutes. $3\frac{1}{4}\times 4$ slides.

"Microphone for High Intensity and High Frequencies," by John K. Hilliard, Altec Lansing Company. A convention paper. Twenty minutes. $3\frac{1}{4}\times 4$ slides.

"The Ideo-Synchronizer," by J. M. Henry and E. R. Moore, Boston Bell. A humorous satire on technical writing, specifications, and engineering. Good for mixed audience. Twelve minutes.

"An Improved Optical Method for Calibrating Test Records," by B. B. Bauer, Shure Brothers, Inc., Chicago, Ill.

"Electronically-Controlled Audio Filters," by L. O. Dolansky, Northeastern University, Boston, Mass.

"Energy Distribution in Music," by J. P. Overlay, Radio Manufacturing Engineers, Inc., Peoria, Ill. The manner in which the acoustic power encountered in music varies with frequency can be of interest in the selection of components to be used in audio reinforcement or reproduction systems. Energy distribution information available up to this time is not fully applicable for such a use because it concerns primarily the average energy distribution. This paper deals with the amplitude of fractional-second energy peaks, without reference to the rate of their occurrence. These peaks must be considered when distortion is of primary consideration; average power is useful only in predicting temperature rise (where applicable) of signal-handling components. The source material from which the distribution analysis is drawn consisted of recent commercial vinyl recordings played on a carefully equalized reproducing system. Ten various types of music are classified and a distribution curve is shown by breaking the spectrum into octaves with a band-pass filter.

"Bells, Electronic Carillons and Chimes," by F. H. Slaymaker, Stromberg-Carlson Co., Rochester, N. Y. Bells and chimes, unlike the more familiar string and wind instruments, produce tones in which the overtone structure cannot be expressed as a series of harmonics. The accuracy of tuning of the various overtones varies widely but the better-cast bell carillons, electronic carillons and tubular chimes do have very accurately tuned overtones. This paper describes the results of measurements on cast bells, electronic carillons and tubular chimes. The individual overtones of the three types of instruments will be demonstrated on a loudspeaker and data will be given on relative amplitude and decay rates of the various overtones. The reaction of "out-of-tuneness" will be discussed and explained. A new type of tone source for electronic carillons will be described and demonstrated. With this new tone source, a rod of carefully controlled rectangular cross section is used which can give in a single unified structure essentially the same overtones array as that of an accurately tuned cast carillon bell. The new rectangular bar tone source will be demonstrated.

"An Experimental Co-Channel Television Booster Station using Crossed Polarization," by John H. DeWitt, Jr., WSM, Nashville, Tenn. The special receiving and transmitting antennas and equipment developed to improve fringe and reception by co-channel booster methods are discussed. A report on the operation of an experimental installation at Lawrenceburg, Tenn., is given.

CALL FOR PAPERS

Technical Program

1960 IRE National Convention

March 21–24, 1960

Waldorf-Astoria Hotel and New York Coliseum, New York, N. Y.

Prospective authors are requested to submit all of the following information by the *deadline date of October 23, 1959.*

1) 100-word *abstract in triplicate*, title of paper, name and address
2) 500-word *summary in triplicate*, title of paper, name and address
3) Indicate the technical field in which your paper falls.

Aeronautical and Navigational Electronics
Antennas and Propagation
Audio
Automatic Control
Broadcast and Television Receivers
Broadcasting
Circuit Theory
Communications Systems
Component Parts
Education
Electron Devices
Electronic Computers
Engineering Management
Engineering Writing and Speech
Human Factors in Electronics
Industrial Electronics
Information Theory
Instrumentation
Medical Electronics
Microwave Theory and Techniques
Military Electronics
Nuclear Science
Production Techniques
Radio Frequency Interference
Reliability and Quality Control
Space Electronics and Telemetry
Ultrasonics Engineering
Vehicular Communications

Note: Original papers only will be considered, not published or presented prior to the 1960 IRE National Convention; any necessary military or company clearance of paper granted prior to submittal.

Address all material to:

Gordon K. Teal, *Chairman*
1960 Technical Program Committee
The Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

The "Null Method" of Azimuth Alignment in Multitrack Magnetic Tape Recording*

A. G. EVANS†

Summary—A number of methods for azimuth alignment were investigated. A technique for alignment which compares the output from two tracks of a multitrack tape provided a substantial improvement in alignment accuracy as compared to the methods which had been in use up to this time. A method for adjusting the lateral position of the head across the width of the tape was also developed which made use of the same basic principles as the "null method" of azimuth alignment.

MULTITRACK recording in general and stereophonic sound recording in particular have introduced many new problems which must be solved if the full potentialities of these systems are to be realized. The objective of the recording process is to store information so that it can be reproduced at some later time. The reproduced signal should be, as far as possible, a duplicate of what was originally recorded on the tape. To accomplish this, the reproduce head must be accurately aligned; that is, the gaps of the head should be adjusted for angular position (azimuth) and for lateral position with respect to the signals recorded on the tape.

The usual method for aligning the azimuth of the reproduce head makes use of a short wavelength signal recorded on an alignment tape especially for this purpose. The azimuth is adjusted to obtain maximum output from one of the heads when reproducing this signal. Simple application of this method to multitrack tapes does not produce sufficiently accurate alignment to insure faithful reproduction of the recorded signal. A better method for head alignment is required which will provide an accurate indication of the correct adjustment without the necessity of specialized equipment. The method should not be so complex as to be usable only in the laboratory. With this objective as a goal, a study was made of several methods of azimuth alignment and lateral head positioning techniques.

Single Track Recording

The technique used for azimuth alignment of single-track magnetic tape recorders when using a short wavelength signal recorded on an alignment tape is quite straightforward. The azimuth of the reproduce head is simply adjusted to obtain the maximum voltage output from the head. The test signal used for adjusting audio recorders is usually in the frequency range of 10 to 15 kc. The alignment tape is then removed and replaced by a piece of blank tape and the equipment set up to record a high-frequency tone. The azimuth of the record head is adjusted to obtain maximum output from the reproduce head, thus matching the azimuth of the record head with the reproduce head.

This method of alignment gives quite satisfactory results when applied to a single-track recorder since the loss in output at the alignment frequency will be no greater than the error which may have been made in reading the output indicator during the alignment. This is normally not greater than ± 1 db at the most, and is of no particular consequence to the over-all frequency response of the recorder. Fig. 1 is a plot of the output vs azimuth angle for a one-mil (0.001 inch) wavelength signal reproduced by a head scanning a one-hundred-mil wide recorded track on the tape.

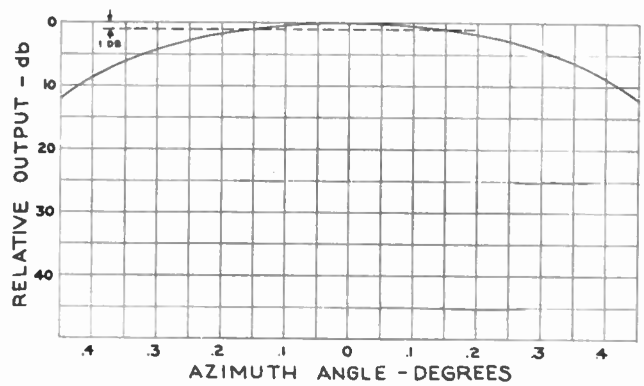

Fig. 1—Single-track recording. $\lambda = 1$ mil.

When reproducing short wavelength signals, there is usually some amplitude variation which makes it difficult to determine the exact center point of this curve. An amplitude variation of ± 0.5 db in output would not be unusual. With this degree of variation and assuming no error in meter readings, the error in azimuth angle could be as much as $\pm 0.15°$. For most applications, if the error in azimuth angle did not exceed $\pm 0.2°$, the loss in high-frequency output would not be objectionable. For other recording systems, the tolerances on the angle would vary approximately directly with the minimum wavelength and inversely with the track width.

Multitrack Recorders

The procedure described for single-track recorders was used to align multitrack recorders when stereophonic recording was first adopted as a standard practice. As long as these tapes were used only for stereophonic listening purposes, there were no serious com-

* Manuscript received by the PGA, May 27, 1959.
† Radio Corporation of America, Indianapolis, Ind.

plaints about azimuth angle alignment. When those same tapes were used to make monophonic recordings by combining the outputs of the tracks, the results were not satisfactory. The problem was traced to rather severe response variations at the high-frequency end of the audio spectrum caused by azimuth alignment errors. Errors which had not been objectionable in single-track recordings could cause complete cancellation of output from equal level in-phase signals from two tracks of the tape at any frequency above about 7 kc.

The tape equipment in use when this difficulty was discovered was a three-track machine with in-line heads. The tape speed was 15 inches per second (ips), and the track placement is that shown in Fig. 2. The most serious problem occurred when the outputs from the outer two tracks were combined. The following curves will be based on the combined output of these two tracks of the three-track tape. This particular equipment is used for the purpose of illustration only; however, the methods are applicable to any multitrack system.

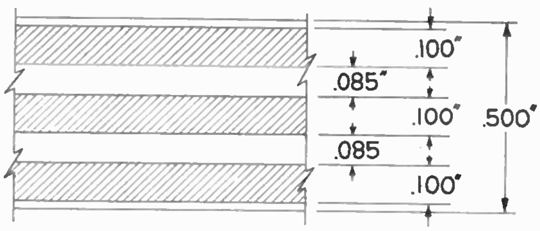

Fig. 2—Track placement. Three-track one-half-inch tape.

The degree of misalignment that can be tolerated in a multitrack recorder is a function of the minimum recorded wavelength, the track width and the spacing between tracks. In single-track recording, an error in azimuth angle of as much as $\pm 0.2°$ was found to be permissible without excessive high-frequency losses. However, using the combined outputs from the two outer tracks of the three-track tape machine, an error of $\pm 0.15°$ caused a loss of 3 db at 4 kc and complete cancellation of the output at 7.5 kc for signals of equal amplitude recorded in-phase. Even reducing the error to one-half, *i.e.*, 0.075°, would cause complete cancellation at 15 kc. To hold the error to $\pm 0.075°$ would require that the output be adjusted to within one-fourth db of the maximum output point using the single-track method of alignment. This, in turn, requires that the amplitude variations must be held to within plus or minus one-eighth of a decibel, which is a very severe requirement at short wavelengths for most audio recorders. There is no question that a better method of azimuth alignment must be found for this type of operation.

As an immediate solution to this problem, it was decided to align on the basis of the combined output signal from two channels while reproducing short wavelength in-phase signals. This procedure made possible a definite improvement in accuracy, but there are a number of reasons why it is not satisfactory for general use. A plot of the output vs azimuth angle for the combined outputs of the two channels is shown in Fig. 3. The curve exhibits a number of maxima near the point of zero azimuth angle, the one with the greatest amplitude corresponding to the point of zero azimuth error. The two adjacent points of maximum output (at an azimuth angle of approximately $\pm 0.15°$) differ in output by only 1 db from the correct maximum. This small difference in the outputs makes it necessary to rotate the head through a sufficient angle to encompass several of the maximum output points in order to determine with any degree of certainty that the final setting is on the correct peak in the output. This technique is reasonably satisfactory for laboratory applications where the time and care required to assure proper application of this method can be taken. In general use, this method was found to be somewhat of a problem since the busy recording engineer was not willing to take the time required to adjust the equipment by this method, especially because even to check and make sure that the equipment was properly aligned it was necessary to change the azimuth angle so that the output would vary through several peaks before he was certain that the equipment was adjusted to the correct maxima. Because of this problem, the method was considered only a stopgap measure and work was continued in an effort to find a better method of azimuth alignment.

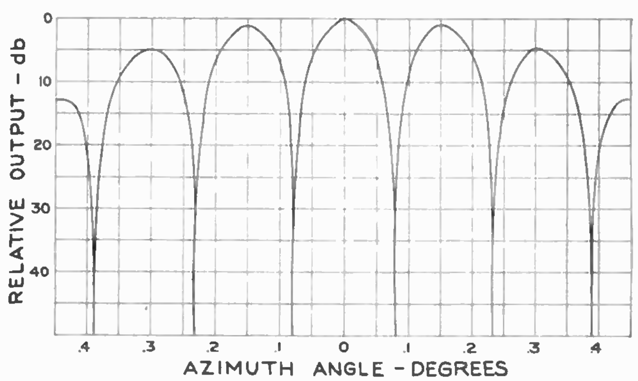

Fig. 3—Two-track recording. $\lambda = 1$ mil; outputs combined "in-phase."

The "Null Method"

A number of different methods for azimuth alignment were suggested, and each was investigated. These included such methods as attempting to align on the recorded bias and various techniques involving wave shapes other than sine waves which would have a low-frequency component for approximate alignment as well as a high-frequency component for exact alignment. None of these methods proved satisfactory, and all of them required some form of specialized equipment that would only be available in the laboratory.

The one thing that was particularly notable during this investigation was that, regardless of the type of wave shape used, the most accurate results were obtained when the signals from the two tracks were arranged so as to tend to cancel each other. This is the same technique used in the common-bridge circuits which are the basis of many electrical measuring systems. In this method, it is necessary to balance the signals with respect to both amplitude and phase for complete cancellation of the output voltage. The differences between the "in-phase" and "out-of-phase" conditions are best observed by writing the equations for the output voltage vs azimuth angle for each. Fig. 4 is a sketch showing the geometric relationships between two in-phase signals recorded in-line on the tape. The combined output from the two tracks is found by writing the equation for each track and adding these together. It is assumed that each track is recorded with the same signal amplitude and frequency so that the only difference between the two reproduced signals would be a phase difference determined by the track spacing, recorded wavelength and the azimuth angle.

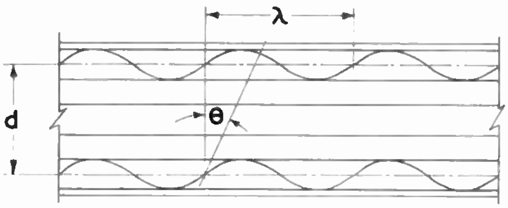

Fig. 4.

The equation for the amplitude of the output voltage e_0 when the two signals are combined so as to add when the azimuth angle is zero is

$$e_0 = \left| K \cos\left(\frac{\pi d}{\lambda} \tan\theta\right)\right| \tag{1}$$

where:

d is the distance between the track center lines,
λ is the recorded wavelength,
K is a constant proportioned to the recording level and the system gain

and

θ is the azimuth angle.

If either the phase of the recorded signal or of the output connection of one channel is reversed, the two signals will tend to cancel each other at zero azimuth angle. The equation for this condition is

$$e_0 = \left| K \sin\left(\frac{\pi d}{\lambda} \tan\theta\right)\right|. \tag{2}$$

The above two equations do not take into account the loss in output from each individual track due to the azimuth angle. To do this, it is necessary to correct (1) and (2) by the factor

$$K_1 = \frac{\sin \dfrac{\pi h \tan\theta}{\lambda}}{\dfrac{\pi h \tan\theta}{\lambda}} \tag{3}$$

where h is the width of the recorded track.

The product of (1) and (3) was used to compute the curve of Fig. 3, for the in-phase combination of one-mil wavelength signals. This curve has several maxima and minima near the point of zero azimuth which leads to confusion as to which is the correct one, and a slope of zero at zero azimuth angle which makes this point difficult to determine experimentally.

Eq. (2), being a sine function, has a very great slope near zero azimuth which makes it possible to determine this point with relatively great accuracy by experimental means. Reproducing a one-mil wavelength signal would result in the same number of maxima and minima found with the in-phase combination and the same care would be required to determine that the alignment was made to the proper minima. The relatively great rate of change of output for the out-of-phase signals in the region near the zero azimuth point makes it entirely practical to use a longer wavelength signal for azimuth alignment and still maintain the required accuracy of adjustment. A plot of output vs azimuth angle for a six-mil wavelength is shown in Fig. 5. Here there is one minimum point in the usual range of azimuth adjustment so there can be no confusion as to which is the correct point.

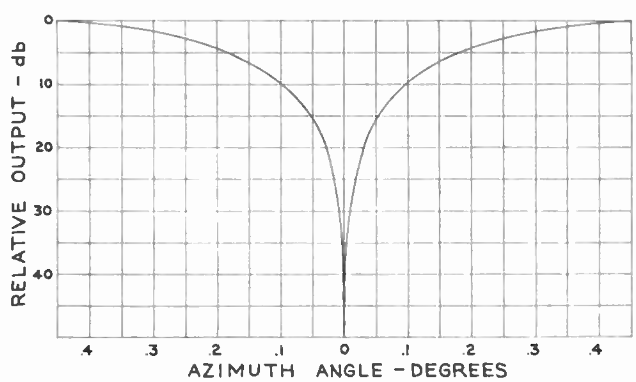

Fig. 5—Two-track recording. $\lambda = 6$ mils; outputs combined "out-of-phase."

This method for azimuth alignment has been called the "null method" because the azimuth is adjusted for a null in the combined output voltage. In practice, several advantages have been found for this technique. The longer wavelength signal can usually be recorded at a higher level than is practical with very short wavelength signals, thus reducing the problems caused by system hum and noise. The longer wavelength signals also tend to produce less amplitude variation which provides a greater degree of accuracy in the output measurement. It is not necessary to adjust the azimuth to determine if the setting is correct; a simple measurement of the combined out-of-phase output signals will determine the degree of azimuth misalignment since the voltage measured is roughly proportional to the error in angle. Working standards can be set as limits on the azimuth misalignment by simply specifying that the output voltage should be less than some predetermined value for any particular system. The ultimate degree of accuracy obtainable with this method depends on the signal-to-noise ratio of the system, the amplitude variation present in the output signals, and the tape guidance which can cause relative phase shift due to skewing of the tape.

The experience which has been gained in the use of this method indicates that in general the tape transport or the tape itself is the basic limitation of accuracy. It was found that the azimuth of the three-track recorder could be consistently adjusted to within $\pm 0.01°$. This represents an improvement by a factor of more than ten-to-one as compared to the usual single-track method. The uncertainties of the in-phase method are essentially eliminated with some improvement in accuracy as compared to the in-phase method with careful adjustment.

This discussion has referred to one particular track arrangement; however, it should be noted that it is equally applicable to any tape system with at least two recorded tracks and an in-line head arrangement with a common azimuth adjustment for the two tracks. It has been used to align two-track $7\frac{1}{2}$-ips and four-track $3\frac{3}{4}$-ips machines with equal success. Best results were obtained by reducing the wavelength of the alignment signal approximately in direct proportion to the tape speed of the machine being adjusted.

Initial Determination of Azimuth Angle

The engineer who starts out to record an azimuth alignment tape is faced with the rather difficult problem of determining when the signal is recorded to within the desired tolerance of zero azimuth, *i.e.*, that the recorded signal is perpendicular to the major axis of the tape. Various methods have been used to determine the azimuth angle accurately; these usually are optical measurements either made on the recording gap or by using carbonyl iron powder to observe the magnetic field at the tape surface. Either of these methods require accurate optical equipment that is not always available. One method that has been in use for many years for determining the true azimuth point is based on the fact that the azimuth error doubles when the signal is reproduced from the base side of the tape. The procedure for using this method consists of recording a short wavelength signal on the tape and adjusting the playback azimuth for maximum output. The tape is then turned over so that it can be reproduced from the base side and the azimuth of the reproduce head is again adjusted for maximum output. If a difference is found between these two settings, the azimuth of the record head is corrected by one-half of this difference and the process repeated until the azimuth is the same when reproduced from either face of the tape. This method is of limited value because of the inherent inaccuracy of the short wavelength method and the very low signal output obtained when reproducing a short wavelength signal with the recording spaced away from the reproduce head by the thickness of the base material.

The use of longer wavelength signals, higher recording level and the inherently greater accuracy of the null method of azimuth alignment would suggest that this method should be applicable to the above technique for determining the true azimuth on the tape. The procedure is exactly the same as for the single-track method. Accurate results can be obtained by this method without any special equipment being required. The technique is so easily applied that it is quite practical for setting the azimuth on equipment in the field when no alignment tape is available.

Lateral Positioning of the Head

The azimuth is only one of the several adjustments required on a head to reproduce a signal from the tape properly. One other adjustment that is very important in multitrack recording is the lateral position of the head across the width of the tape. This is especially true when very narrow tracks are used. Optical means are available for accomplishing this adjustment, but on some equipment this is not convenient and such equipment is usually not available in the field. The other alternative is to adjust the head by means of a recorded signal on the tape.

One method of adjusting the lateral position of the reproduce head is to adjust for maximum output when reproducing from a track recorded in the desired position. A fairly short wavelength signal should be used to reduce fringing effects if the accuracy of adjustment is important. At best, this method is tedious because there is usually some interaction between the position adjustment and the azimuth adjustment. A technique has been developed for determining the correct position of the head which makes use of the principles used in the null method of azimuth alignment. Two tracks are recorded on the tape so that they are equally displaced from the center line so that the head will scan a small portion of each track when it is properly centered, as shown in Fig. 6. These two tracks are recorded with the same amplitude, frequency and 180° out-of-phase. The track placement can be determined by use of carbonyl iron powder. The output voltage from the reproduce head will vary as it is moved across the tape and will be minimum when equal signals are reproduced from the two out-of-phase tracks. A signal with a wavelength several times that used in azimuth alignment should be

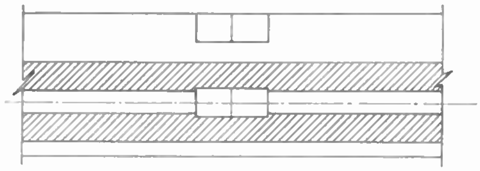

Fig. 6—Recorded track arrangement for lateral alignment.

used so that azimuth misadjustment will not introduce errors in this measurement. A curve of output voltage vs displacement is plotted for a 0.043-inch wide reproduce head in Fig. 7.

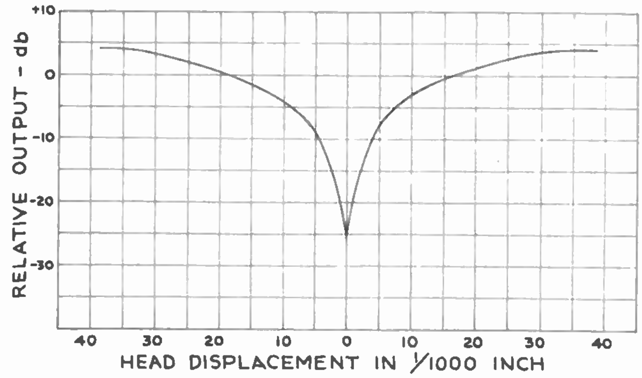

Fig. 7—Output vs displacement.

The reproduce head can be positioned to within approximately 0.001 inch either by actually measuring the output voltage or by listening to the output signal. The fact that a minimum of output represents the correct adjustment does make this method useful for production use and for the service of equipment where it is desirable that a single measurement indicate if the equipment is properly adjusted. In this type of application, there would be no particular disadvantage in requiring a specially recorded tape for head positioning.

Conclusion

The null methods for azimuth alignment and head positioning provide accurate means for adjusting a reproduce head to scan the information on a recorded tape properly. These methods are particularly applicable where routine checks are necessary to assure that the equipment is properly adjusted, such as service or manufacturing operations on audio tape recorders, since only a single measurement of output is required to determine if the equipment is properly aligned. These methods should be equally useful in any type of multitrack tape recording systems where the relative phase between tracks is of importance. Some variations in track arrangement would probably be required for each particular application, but the basic technique would be unchanged.

Phase Shift in Loudspeakers*

W. RICHARD STROH†

***Summary*—A simple method of measuring the phase characteristic of a loudspeaker is described and typical phase curves are given for moving-coil and electrostatic loudspeakers. Distortion in correlation functions measured with an electrostatic loudspeaker is described and related to the phase characteristic of the speaker.**

Introduction

It has been customary for many years among both designers and users of loudspeakers to consider the amplitude response the principal figure of merit. The phase characteristic has been neglected, on the grounds that the ear is relatively insensitive to phase distortion. Though there may be a measure of truth in this statement, there is certainly a large component of ignorance; in fact, recent evidence suggests that the ear is more fastidious about phase than had previously been supposed.[1] On the other hand, if a speaker is used in certain kinds of physical measurements—correlation measurements, for example—there is no doubt that phase distortion will damage the performance as readily as amplitude distortion. The phase characteristic is therefore a legitimate object of the engineer's curiosity. A simple method for measuring the phase characteristic is described below, and results are given which are typical of two types of speakers, the moving-coil and the electrostatic.

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† University of Rochester, Rochester, N. Y. (The research described in this paper was done while the author was at the Acoustics Res. Lab., Harvard Univ., Cambridge, Mass.)

[1] S. S. Stevens, Ed., "Handbook of Experimental Psychology," John Wiley and Sons, Inc., New York, N. Y., p. 1024; 1951.

Let us recall the requirements which a system must meet if it is to transmit arbitrary functions of time with no distortion other than a translation along the time axis, or simple delay: the amplitude response of the system must be independent of frequency and the phase shift through the system must be proportional to frequency. If the first of these is not satisfied, the system is said to exhibit frequency distortion; if the second is not met, we say that delay distortion is present. Let us define an axial transfer function $b(\omega)$ for a loudspeaker as the ratio of the complex sound pressure $P(\omega)$ at some point on the principal axis to the complex voltage $V(\omega)$ across the voice coil (it may occasionally be more convenient to use voice-coil current):

$$b(\omega) = P(\omega)/V(\omega). \tag{1}$$

This may also be written

$$b(\omega) = B(\omega)e^{j[\phi(\omega)-\omega\tau_a]}, \tag{2}$$

where $B(\omega)$ is real and represents the ordinary axial pressure response function, $\phi(\omega)$ is real and represents the phase characteristic of the speaker, and τ_a represents the air-path delay between the speaker and the observation point. If this transfer function is known, then the pressure $p(t)$ corresponding to any input $v(t)$ is given by

$$p(t) = \int_0^\infty g(t')v(t-t')dt', \tag{3}$$

where $g(t)$ is the impulse response and is related to $b(\omega)$ by

$$g(t) = \int_{-\infty}^{\infty} b(\omega)e^{j\omega t}d\omega. \tag{4}$$

An ideal system is characterized by constant B and a ϕ proportional to ω. Under these conditions we have

$$\begin{aligned} B(\omega) &= B_0, \\ \phi(\omega) &= \omega\tau_0 \qquad (5) \\ \phi(\omega) - \omega\tau_a &= \omega(\tau_0 - \tau_a) = -\omega\tau_1 \\ b(\omega) &= B_0 e^{-j\omega\tau_1} \\ g(t) &= B_0\delta(t-\tau_1) \\ p(t) &= B_0 v(t-\tau_1). \qquad (6) \end{aligned}$$

For all other functions $b(\omega)$, $p(t)$ will *not* be a delayed replica of $v(t)$.

Method of Measurement

It is clear that if we wish to predict in detail the effect of the loudspeaker's transfer characteristics on a particular measurement, we need detailed knowledge about both $B(\omega)$ and $\phi(\omega)$. Direct continuous measurement of $B(\omega)$ is relatively easy; all one needs is a free field, a suitable variable-frequency source, a microphone whose amplitude response is known to be flat, and an appropriate recording apparatus.

Continuous direct measurement of $\phi(\omega)$ requires, in addition to the above-mentioned apparatus, an accurate phase meter, a microphone whose phase characteristic is proportional to frequency or is at least known, and some means of eliminating the air-path phase shift $\omega\tau_a$. Wiener's pioneer phase measurements[2] were made by simply recording the phase angle between voice-coil voltage and microphone output voltage; his data therefore include the air-path term which is so large that it makes the fluctuations caused by the speaker look quite insignificant.

If one is satisfied with knowledge of only the phase slope $d\phi/d\omega$, then one can use the modulation phase shift method of Nyquist and Brand.[3] The system under test is excited by an amplitude-modulated voltage. The phase difference between the sinusoidal input modulation envelope and the envelope of the output is proportional to $d\phi/d\omega$, where $\phi(\omega)$ is the phase shift through the system at the carrier frequency. The modulation frequency must be small compared to that of the carrier. Measurements made on loudspeakers by this method have been reported by Ewaskio and Mawardi.[4] The air-path delay appears as a constant term in the result, and is easily eliminated; however, if there is a region of constant or nearly constant nonzero phase shift in $\phi(\omega)$, its presence will not be revealed by this method.

Apparatus

The directly-measured values of $\phi(\omega)$ reported below were obtained by using an electrical delay in the parallel channel to offset the air-path delay. The system is depicted in Fig. 1. The electrical delay is provided by a low-pass filter delay line, which is variable in 0.05-msec steps up to a maximum of 25 msec.[5] The delay line is not perfect; for normal use, one accepts a compromise between flatness of the amplitude response and straightness of the phase characteristic. For these experiments, a variable phase-correcting circuit was added to the line, and adjusted so that at all line settings used, the over-all phase characteristic differed no more than ±10° from the ideal.

The microphone used was a Western Electric Type 640-AA. The resonance frequency of this microphone lies in the neighborhood of 7 or 8 kc. At frequencies well below resonance (up to about 1000 cps) the output voltage of this microphone is in phase with the sound pressure. As resonance is approached, the voltage lags the pressure by an increasing amount, which reaches 90° at resonance. Data on this phase shift in the 640-AA have been published by Wiener.[6] In the system of Fig.

[2] F. M. Wiener, "Phase distortion in electroacoustic systems," *J. Acous. Soc. Amer.*, vol. 13, pp. 115–123; October, 1941.
[3] H. Nyquist and S. Brand, "Measurement of phase distortion," *Bell Sys. Tech. J.*, vol. 9, pp. 522–549; July, 1930.
[4] C. A. Ewaskio and O. K. Mawardi, "Electroacoustic phase shift in loudspeakers," *J. Acoust. Soc. Amer.*, vol. 22, pp. 444–449; July, 1950.
[5] W. R. Stroh, "A 25-Millisecond Electromagnetic Delay Line," Acoustics Res. Lab., Harvard Univ., Cambridge, Mass., Tech. Memo. No. 41; December 5, 1957.

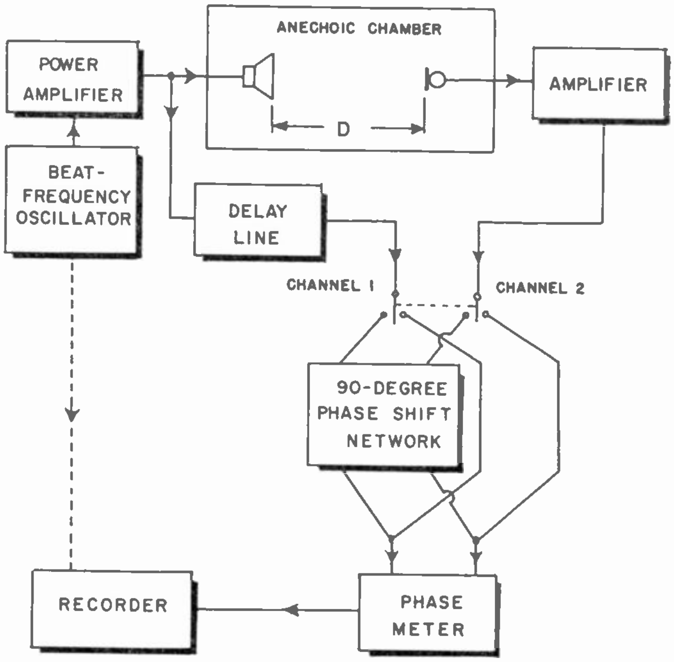


Fig. 1—Block diagram of apparatus used to measure the phase characteristic of a loudspeaker.

1, microphone phase shift was compensated by a phase lag, coinciding with Wiener's data up to 4000 cps, placed in Channel 1.

A 90°, frequency-independent phase shift network was included in the system for two purposes; occasionally it was used to shift the zero of the phase meter from end-of-scale to center-scale; it was also used to offset certain loudspeaker phase shifts, as described below.

The useful frequency range of the system was from about 50 cps to 4000 cps. These rather narrow limits were set chiefly by the delay line; the other elements of the system would accommodate perhaps several additional octaves, although an extension of the upper limit would require more careful attention to phase correction of the microphone (or the use of a microphone with a much higher resonance).

The delay line is, of course, not the only device that could be used to cancel the air-path delay. Since the testing is done exclusively at single frequencies, one could use any phase shifting agent which could be linked mechanically or electrically to the oscillator in such a way as to produce a shift proportional to frequency. The device should, however, allow variation of the proportionality factor either continuously or in small increments, and should provide a very accurate proportionality at any one setting.

Measuring Technique

The chief difficulty encountered in making phase measurements proved to be determining the correct setting of the delay line. Obviously, this should be $\tau = D/c$, where D is the distance from the speaker to the microphone, and c is the speed of sound. The flat microphone diaphragm provides an unambiguous starting point for measuring D, but at what point on the speaker should the measurement terminate—at the apex of the cone, at the rim, or at some intermediate point? If the measured value of D is in error by two-thirds of an inch, the corresponding value of τ will be in error by 0.05 msec. At 4000 cps, this corresponds to a phase error of 72°. The best answer to this question seems to be that one should choose that setting of delay which gives the phase function nearest to that expected from theoretical considerations. For example, one expects the phase function of a moving-coil loudspeaker to approach zero as the frequency increases, as explained below. If the delay is set to give this high-frequency behavior, one finds that the index for D lies somewhere between the apex and the rim of the cone.

We have tacitly assumed above that the axial phase function should be independent of the value of D. If the speaker behaved as a rigid piston, if the microphone were placed precisely on the speaker axis, and if the test chamber were completely anechoic, this assumption would be true. In practice, none of these conditions is met, but the assumption still holds pretty well. Curves taken at different values of D have slightly different fine structure, but the gross excursions are similar in magnitude and form, and occur at the same frequencies; the general trend is the same.

It is perhaps not superfluous to call attention to the fact that this invariance of $\phi(\omega)$ with D holds right up to the surface of the radiator. This may be inferred from the expression for the pressure on the axis of a circular piston of velocity $U_0 e^{j\omega t}$ operating in an infinite baffle:[7]

$$p = 2\rho c U_0 \sin \frac{k}{2} \left(\sqrt{D^2 + a^2} - D\right)$$

$$\cdot \exp j\left[\omega t + (\pi/2) - (\omega D/2c)\left(\sqrt{1 + (a^2/D^2)} + 1\right)\right] \quad (7)$$

where a is the radius of the piston. No matter how small the value of D may be, the last term of the exponent remains proportional to frequency so that its effect can always be cancelled by a suitable electrical delay. The 90° shift represented by the factor $e^{j\pi/2}$, being independent of frequency, is not cancelled; it is important in interpreting the measured phase curves in terms of the known properties of the source and will be referred to below.

Typical Measured Phase Functions

In Figs. 2 and 3 are shown curves of the amplitude function $B(\omega)$ and the phase function $\phi(\omega)$ for two loudspeakers typical of those in general use today. The first is a Western Electric Type 755-A; the second is an electrostatic loudspeaker constructed in the laboratory. Measurements were made on a number of other speakers but in all cases the performance was not essentially

[6] Francis M. Wiener, "Phase characteristics of condenser microphones," *J. Acous. Soc. Amer.*, vol. 20, p. 707; September, 1948.

[7] H. Stenzel, "Leitfaden zur Berechnung von Schallvorgängen," Julius Springer Verlag, Berlin, Germany, p. 62; 1939.

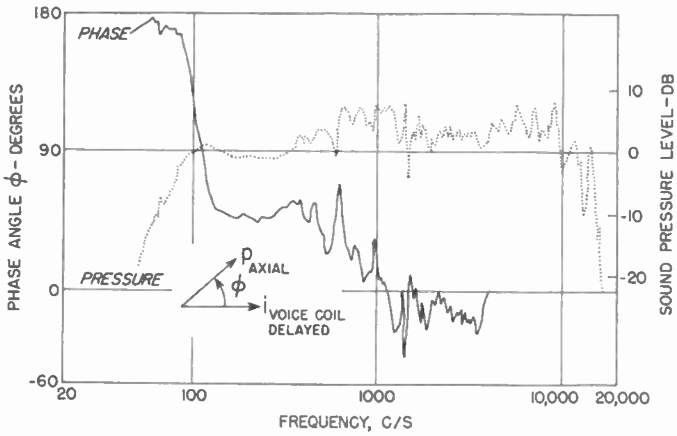


Fig. 2—Phase angle between delayed voice-coil current and axial sound pressure, and axial sound pressure level for constant voice-coil voltage for Western Electric Type 755-A moving-coil loudspeaker in a 1-foot[3] closed box.

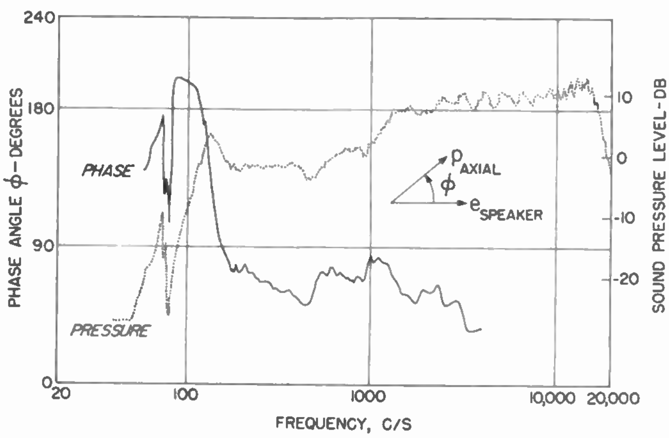


Fig. 3—Phase angle between delayed loudspeaker voltage and axial pressure, and axial sound pressure level for constant input voltage for 18×18-inch electrostatic loudspeaker in an 11-foot[3] closed box.

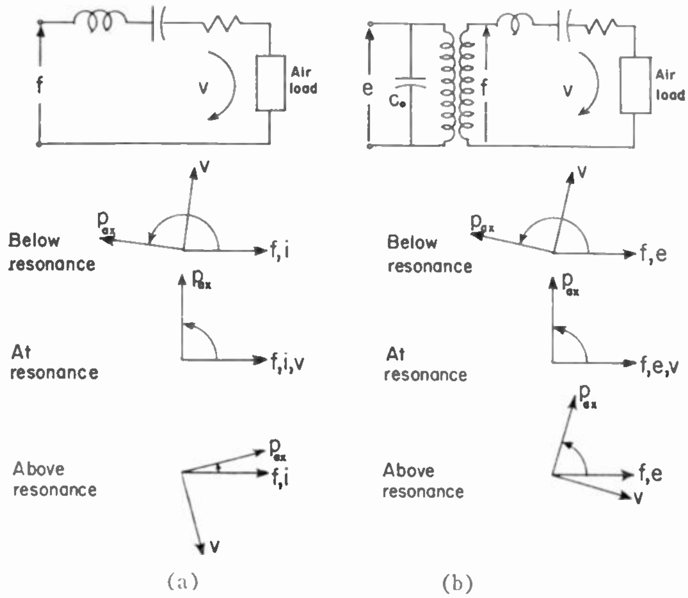


Fig. 4—Equivalent circuits for determining the phase angle between delayed loudspeaker current or voltage and axial sound pressure. (a) Moving-coil loudspeaker. (b) Electrostatic loudspeaker.

different from the typical samples presented here. It did not seem necessary to include additional curves.

The general form of the phase function for a moving-coil loudspeaker may be inferred from a simple equivalent circuit representing the mechanical properties of the speaker and its air load.[8] Such a circuit is shown in Fig. 4(a). It is assumed that the force on the voice-coil is in phase with the voice-coil current at all frequencies; this current (rather than voice-coil voltage) is the natural phase reference for the system. At the principal resonance, the cone velocity is in phase with the voice-coil current, and by (7) the axial pressure, if the air-path delay is cancelled, will lead the velocity by 90°, so the pressure will be found to be 90° ahead of the voice-coil current. At frequencies below the principal resonance, the mechanical stiffness becomes the controlling factor, resulting in an axial pressure which leads the current by phase angles up to 180°. At frequencies above the principal resonance, in the useful range of the speaker, the air-load reactance gradually diminishes, while its resistance at first increases with frequency, then levels off and becomes independent of frequency. The mass of the moving system now controls the phase relations in the speaker itself so that the velocity of the cone lags the current by angles up to 90°, and the axial pressure gradually comes to be in phase with the voice-coil current.

These gross features are well illustrated in Fig. 2, which shows the performance of the 755-A speaker mounted in a closed box having a volume of approximately 1 ft³. Of course, the phase curve in Fig. 2 exhibits a fine structure which is not explained by the simple analysis just given. This fine structure, which begins to appear at frequencies in the neighborhood of 200 to 400 cps, is undoubtedly the result of cone break-up and the resulting nonuniform motion of the cone. We have not sought a detailed explanation of these fluctuations. It may be noted, however, that each prominent phase excursion coincides with a peak or notch in the amplitude curve.

The expected phase behavior of an electrostatic loudspeaker may be inferred from an equivalent circuit, Fig. 4(b), which is essentially the same as that shown in Fig. 4(a). The circuit as shown is single-sided, but applies equally well to a push-pull or symmetrical structure if the circuit elements are properly interpreted.[9] The static electrical capacitance C_0 of the unit appears to the left of the ideal units transformer. The force f driving the mechanical system is in phase with the voltage e, so this voltage is the natural phase reference. If this reference is chosen, the capacitance C_0 will have no effect on the phase relationships of the system.

The behavior of this circuit may be predicted if suitable typical values of the elements are introduced, such

[8] L. L. Beranek, "Acoustics," McGraw-Hill Book Co., Inc., New York, N. Y., p. 191; 1954.

[9] F. V. Hunt, "Electroacoustics," John Wiley and Sons, Inc., New York, N. Y., p. 201; 1954.

as those given by Janszen, Pritchard, and Hunt.[10] From these data it appears that the speaker's behavior is quite similar to that of the moving-coil type. The principal resonance occurs at the lower end of the useful frequency range; below this frequency, the diaphragm is stiffness-controlled. Above this frequency, the mechanical input impedance has a rather large resistive part due to the air load and a positive reactance due partly to the mass of the diaphragm but mostly to the air load. As the frequency increases, the air load reactance diminishes, but the reactance due to the relatively small mass of the diaphragm gradually increases, so that at very high frequencies the motion is mass-controlled, as in the moving-coil speaker. Thus we expect the phase characteristic of this speaker to be qualitatively similar to that of the moving-coil speaker, but quantitatively different in the middle range of frequencies, where the mechanical impedance is largely resistive.

The results, shown in Fig. 3, bear out this prediction. These curves were made with an 18×18-inch push-pull electrostatic speaker constructed in the laboratory. At the upper-frequency limit of the phase-measuring system, the phase characteristic is still hovering about 90°, which suggests that the mechanical impedance at this frequency is still predominantly resistive. The extrema in both the phase and the amplitude curves at about 80 cps were traced to a resonance of the metal stationary electrodes of the speaker. Probably the swing of the phase curve above 180° is also associated with this resonance.

Effect of Loudspeaker Transfer Function on Measured Correlation Functions

The skeptical reader may feel that the rather unusual phase curve shown for the electrostatic speaker is a result of an erroneous measurement rather than a real phase shift. Doubt on this point should yield, however, to the evidence provided by correlation measurements. If the system shown in Fig. 1 is used to measure crosscorrelation functions of Gaussian noise in octave bands, the resulting curves should have the form of autocorrelation functions centered on the value of delay corresponding to the air-path delay. That is, the curves should be symmetrical about this point and their maxima should be unity. Instead, the curves are not symmetrical; their maxima fall at different values of delay and the maxima fall short of unity. Since all the elements in the system except the loudspeaker perform nearly as they should, we may confidently blame the loudspeaker for most of these shortcomings.

An extreme case of distortion in the measured correlation function is shown by the solid curve of Fig. 5. Instead of being symmetrical, the curve is almost antisymmetrical. Since we have seen how the electrostatic

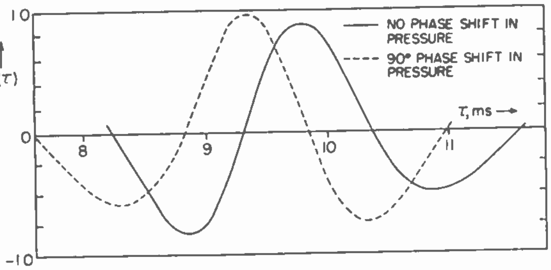


Fig. 5—Normalized crosscorrelation function of loudspeaker voltage and free-field axial sound pressure for octave-band noise, 300–600 cps. Speaker: 30×36-inch electrostatic unit.

speaker introduces a phase shift of approximately 90° in this frequency range, it is natural to try introducing ahead of the correlator the 90° phase shifting network shown in Fig. 1. This has the expected result of restoring the correlation function to approximately symmetrical form, as shown by the dashed curve of Fig. 5. Here is positive evidence that a phase shift of the order of 90° exists over part of the speaker's range; this evidence is independent of the foregoing, and therefore corroborates it.

It is interesting to speculate a little further about the effects of a 90° frequency-independent phase shift. It can be shown that such an operation in the frequency domain is equivalent to a Hilbert transform in the time domain.[11] In general, a Hilbert transform converts a symmetrical function of time into an antisymmetrical one, and vice versa. In particular, the transform of a single rectangular pulse centered on the point $t=0$ is antisymmetrical and has a positive pole at the leading edge of the original pulse and a negative pole at the trailing edge. This gives a rough idea of the distortion such a pulse would experience in traveling through a transducer which effected a phase shift of approximately 90°. This effect was investigated experimentally; a square voltage was applied to an electrostatic speaker and the resulting sound pressure was observed. It was found to consist of a series of alternating positive and negative spikes. Here is another confirmation of the existance of a rather surprising phase behavior in the electrostatic loudspeaker.

Phase Measurements on Multiple-Speaker Systems

The utility of a phase-measuring technique, such as that described above, in the construction and installation of multiple-speaker systems, is self-evident. To be sure, the phase performance of the speakers may be obscured by the phase crimes committed by the various dividing networks in common use, but these elements can easily be evaluated separately and the effects of speaker placement and phasing may be determined. Undoubtedly such information would be useful in the design of stereo systems.

[10] A. A. Janszen, R. L. Pritchard, and F. V. Hunt, "Electrostatic Loudspeaker," Acoustics Res. Lab., Harvard Univ., Cambridge, Mass., Tech. Memo. No. 17; April, 1950.

[11] S. Goldman, "Information Theory," Prentice-Hall, Inc., New York, N. Y., Appendix V; 1953.

A Two-Watt Transistor Audio Amplifier*

WILLIAM D. ROEHR†

Summary—For low distortion, power transistors should be driven from a low impedance source. Thus an emitter-follower driver has definite appeal. A further advantage is that the driver transistor may be direct coupled to the output transistor.

This paper describes the basic design method for this circuit and illustrates its performance. Important considerations such as stability, transistor interchangeability, frequency response, and distortion are discussed and typical measurements shown.

The features of this circuit are: good stability and the facts that transistor parameters are noncritical, no bias adjustments are required, frequency response is flat over the audio range, and distortion isl ow.

Introduction

To PRODUCE an amplifier with a reasonable power gain and low signal input voltage requirements, the emitter-follower–common-emitter configuration as shown in Fig. 1 was chosen. This choice permits the power transistor to be driven from a low-impedance source, which minimizes distortion, and easly lends itself to direct coupling.

Circuit Description

The functions of the resistors will be discussed. R_1 and R_2 form the bias network for the entire amplifier. Their values will be determined by R_6, the required stabilization, and the desired value of quiescent current in Q_2. They should be chosen so that $V_{B_1} \approx I_{Q_2}R_6 + 0.4$ for germanium transistors. The 0.4 volt accounts for the drop across the base-emitter junctions. As will be shown, with proper design, I_{CO} in Q_2 will have negligible effect upon quiescent point stability but I_{CO} in Q_1 is important. Thus, knowing the maximum expected I_{CO} at maximum operating junction temperature, the permissible shift in V_{B_1} may be computed for a given stability; hence, the values of R_1 and R_2 are determined.

Since the emitter current of Q_1 equals the base current of Q_2 plus some additional drain through R_5, maximum values of Q_1 emitter current will occur when the lowest gain transistors are used in Q_2. With knowledge of this, R_3 and R_4 are chosen so that the minimum V_{CE} is about 1 volt under peak signal conditions. This divider will significantly increase the stability by operating Q_1 at the lowest possible voltage, thus resulting in less power dissipation and less I_{CO}.

R_6 largely determines the stability of the system and is the sole determinant of the tolerable variation in current gain of Q_2. It should be chosen so that the voltage drop across it is large compared to the changes in base-to-emitter voltage required to keep emitter current constant over the expected temperature range for the various types of transistors proposed for Q_2. Generally, a 1.5-volt drop will insure excellent stability when germanium transistors are used.

R_5 serves to improve the circuit with regard to I_{CO} in Q_2. I_{CO} here will cause a few millivolts shift in V_{B_2}. This causes little change in Q_2 current because of the stabilizing action of R_6; however, a drastic change in the current of Q_1 results. R_5 causes the current in Q_1 to be higher than that required to drive Q_2 to cutoff. Thus, some loss in current caused by I_{CO} in Q_2 will not limit the power output. R_5 will lower the input impedance of Q_2 and cause a loss of power gain through Q_2 but it can be shown[1] that this is slight as long as R_5 is not smaller than h_{ie}. Over-all power gain will remain virtually unchanged because the input impedance of Q_1 will normally be much greater than the shunt of the bias resistors.

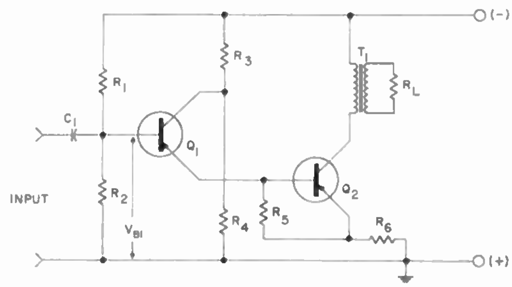


Fig. 1—Basic circuit.

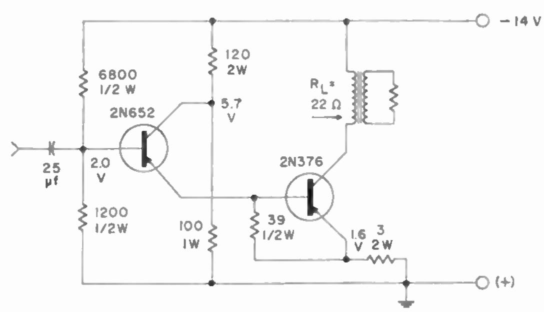


Fig. 2—Two-watt audio amplifier.

A Practical Design

Example of a completed circuit showing typically measured voltages is shown in Fig. 2. Resistance values are in ohms. For full power output, the signal source must be capable of supplying 1.2 volts at low distortion into the 1000-ohm input impedance. Power gain is then 31 db.

* Manuscript received by the PGA, July 6, 1959.
† Motorola, Inc., Phoenix, Ariz.

[1] W. D. Roehr, "Characteristics of Degenerative Amplifiers Having a Base-Emitter Shunt Impedance," Motorola Semiconductor Div., Phoenix, Ariz., Application Note No. 12; December 10, 1958.

Stability

I_{CO} can be easily simulated by using a high-voltage battery in series with a large resistance chosen to produce the desired constant current. The results of I_{CO} simulation are shown in Table I for the circuit of Fig. 2. Power output and distortion were essentially unaffected by the I_{CO} values shown in the table. Stability is seen to be adequate. Using Shea's stability definition[2] ($S_I = \Delta I_E / \Delta I_{CO}$), S_I is about 5 for Q_1 and 4 for Q_2. The values of I_{CO} chosen were thought to be representative of actual values encountered when operating this amplifier at an ambient temperature of 60°C. I_{CO}, however, is not the only undesirable result of high temperature operation. The forward voltage drop of the emitter-base junction decreases,[3] causing additional emitter current. Fig. 3 illustrates the results of a temperature check showing this effect. 2N652 and 2N376 transistors were used in the circuit. Temperatures shown are ambient temperatures. The power transistor junction temperature was approximately 25°C higher because of the heat dissipator which was used. The junction of Q_1 was only about 10°C higher since the can was fastened to the chassis by a fuse clip. Performance over the temperature range of −25°C to +60°C is adequate.

TABLE I
I_{CO} Stability

Collector Current	No I_{CO}	I_{CO} = 120 μa in Q_1	I_{CO} = 2 ma in Q_2	I_{CO} = 120 μa in Q_1 and 2 ma in Q_2
I_{Q1} (ma)	10.6	11.3	7.7	8.5
I_{Q2} (ma)	510	550	520	560

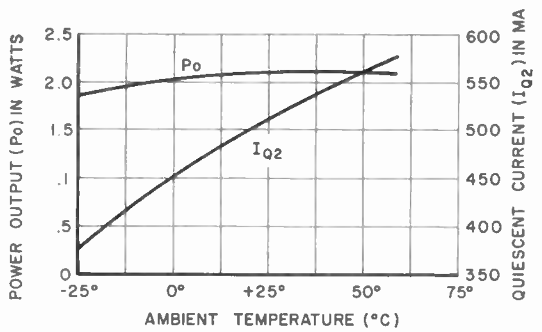


Fig. 3—Output transistor quiescent current and power output as a function of temperature.

At 60°C, however, some power transistors were used which would not produce full power output. This was traced to excessive I_{CO} in Q_2, which reduced the current through Q_1 to a value insufficient to provide full drive. Accordingly, Fig. 4 was prepared using I_{CO} simulation to illustrate this effect. The figure shows the expected current through Q_1 and the effect of I_{CO} on power output when various values of R_5 are used. In actual use, of course, some compensation would occur at high temperatures, because I_{CO} in Q_1 and changes in the base-emitter conductance of the transistors would increase the current through Q_1. Other characteristics of the amplifier are virtually unchanged, since R_5 is varied over the range shown. With the circuit in Fig. 2 which was used for all other data in this report, I_{CO} should not exceed 3.0 ma at 12 volts V_{CE} and at the highest junction temperature of interest. As the data show, with $R_5 = 20$ ohms, it would be possible to permit I_{CO} to reach 7.0 ma with no adverse effects. However, lower I_{CO} in the power transistor permits a sizeable reduction in the dissipation required of Q_1. The effect of R_5 on the current through Q_1 is shown in Fig. 5.

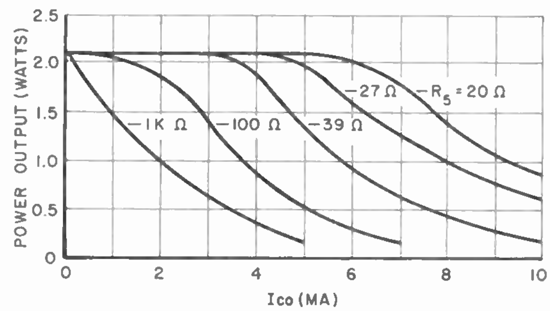


Fig. 4—Effect of output I_{CO} and R_5 upon power output.

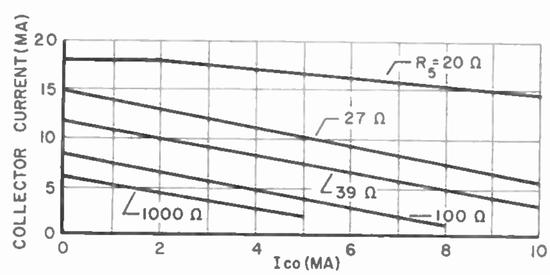


Fig. 5—Effect of output transistor I_{CO} and R_5 upon the driver collector current.

Interchangeability of Transistors

Several transistors having a wide range of current gain were used in both stages. The current gain of the driver need only be over 75 to have no measurable effect upon performance; therefore, type 2N652 was chosen. Fig. 6 shows the effect of the dc current gain of Q_2 upon the collector currents of both transistors. From this curve it appears that a minimum of 60 is required. Several Motorola types were used for the power transistor. Type 2N376 is a reasonable design center and is used in all the data which follow.

Frequency Response and Distortion

Harmonic distortion is shown in Fig. 7. Values shown are typical. Plus or minus 20 per cent variations could be expected from this curve. The frequency response at a low-reference level is shown in Fig. 8. Plus or minus 1 db variations at 20 kc should be expected. The effect

[2] R. F. Shea, *et al.*, "Transistor Circuit Engineering," John Wiley and Sons, Inc., New York, N. Y., ch. 3; 1957.

[3] R. Greenburg, "Base to Emitter Voltage as a Function of Temperature," Motorola Semiconductor Div., Phoenix, Ariz., Application Note No. 2; August 22, 1957.

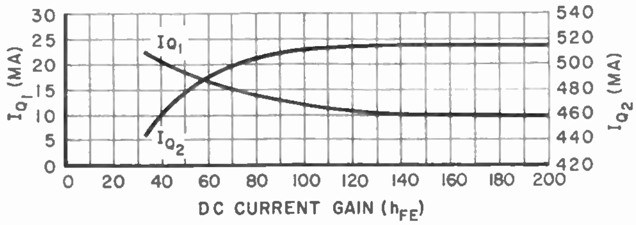


Fig. 6—Collector currents as a function of output transistor current gain.

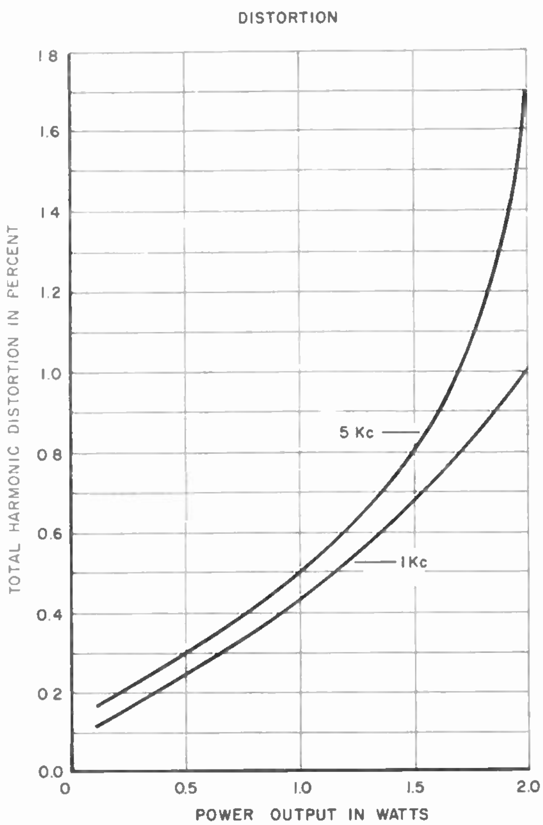


Fig. 7.

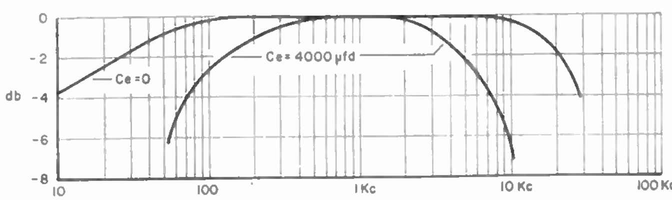


Fig. 8—Frequency response 0 db = 0.5 watt.

of bypassing, or rather attempting to bypass, the emitter of the output stage is shown in the lower curve. The bandwidth becomes too narrow and even 4000 μf is not large enough to allow adequate low-frequency response to be obtained. About 18 db more gain results and distortion is increased by about a factor of six. In some cases, partial bypassing might produce a good design, but was not judged practical for high-fidelity applications. Signal source impedance for these measurements was 10,000 ohms. Response improves and distortion increases slightly as the source impedance is lowered. Distortion of the generator used measured 0.1 per cent.

The most significant performance curve is shown in Fig. 9, which illustrates the power output at a constant total harmonic distortion of 1 per cent. The half-power point is reached at 10 kc, which is excellent performance. Low-frequency distortion is not shown as it is solely a function of the transformer used. A choke was actually used in this amplifier for response measurements as a suitable transformer was not available.

The output impedance, looking back from the transformer primary, is approximately 310 ohms. Therefore, to provide damping, a feedback voltage taken from the output transformer should be applied to the stage preceding the driver.

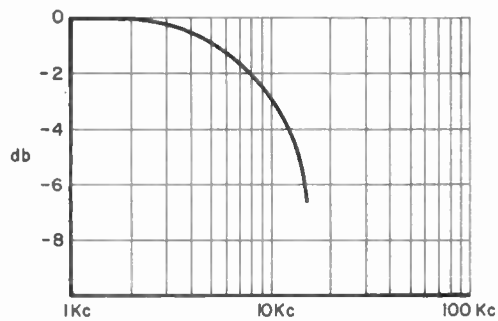
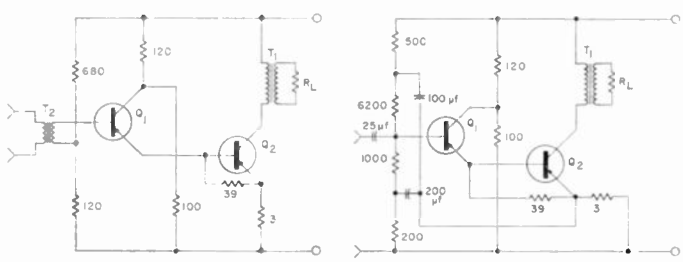


Fig. 9—Power response at 1.0 per cent total harmonic distortion 0 db = 2.0 watts.



(a) (b)

Fig. 10—Possible modifications.

Circuit Variations

Steps could be taken to increase the gain of the system. An obvious approach is to use transformer coupling into the input. By matching to the high input impedance of Q_1, which is approximately 15,000 ohms, an additional 12 db of gain could be realized. However, some changes in gain will now appear as transistor parameters vary and distortion will be a function of the source resistance. Stability with regard to I_{CO} in Q_1 could be improved because now the bias resistors, R_1 and R_2, can be made very small since they no longer shunt the input. Fig. 10(a) shows how this could be done. Practically speaking, the cost of a high quality input transformer would probably exclude this scheme from high-fidelity applications.

Another method of raising the gain by reducing the shunt effect of R_1 and R_2 is to utilize the resistance multiplying circuit[1] for Q_1 as shown in Fig. 10(b). The feedback voltage must be taken from the emitter of Q_2, however, because Q_1 will not tolerate any additional loading. This system will provide about 6 db more gain at the expense of 50 per cent more distortion and some variation of power gain with transistor parameters. If the signal source has a high impedance, this approach may have merit.

Conclusion

The supporting data in this report show that a single-ended two-watt amplifier, which is very noncritical of transistor parameters and which has performance acceptable for high-fidelity applications, is practical. High-temperature I_{CO} in the power transistor must be controlled, however, or performance is seriously degraded.

Nonlinear Distortion Reduction by Complementary Distortion*

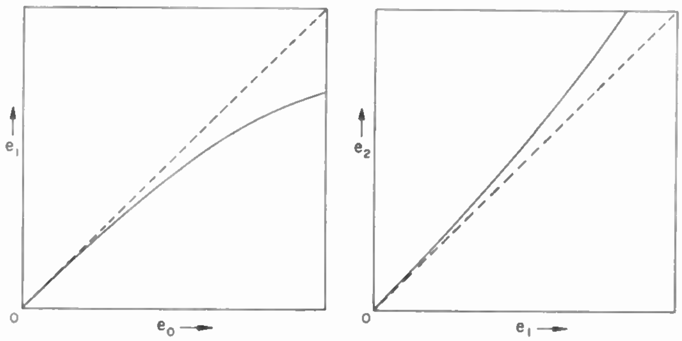
J. ROSS MACDONALD†

Summary—**Nonlinear distortion produced in a given circuit can be reduced by pre- or postdistorting the signal applied to or from the circuit. Such complementary distortion cannot reduce the original distortion to zero in practice because of distortion of distortion, but it can result in greatly reduced output distortion over a limited amplitude range. General results for the design of pre- or postdistortion circuits are given, and the mathematical results are illustrated by comparing the total harmonic distortions obtained with pre- and postdistortion corrections of increasing complexity applied to a simple nonlinear circuit.**

Introduction

THE correction of an undesired frequency distortion, such as a droop in loudspeaker output at low frequencies, by means of a complementary response in the applied signal is well known and often used. Somewhat less well known and understood is the corresponding technique for reducing nonlinear distortion. It is usually stated or implied[1,2] that nonlinear distortion such as that arising from the response characteristic of Fig. 1(a) can be cancelled by passing the distorted signal through a circuit having the complementary response characteristic of Fig. 1(b). For example, it is often expected that if the distortion arises only from a square-law term, it may be completely cancelled by subsequent transmission through a network yielding square-law distortion of equal magnitude but opposite sign. The present work shows that complete cancellation is impossible because of distortion of the original distortion and that over-all distortion reduction is only possible over a limited range of input signal amplitude.

Negative feedback is commonly regarded as the great panacea for distortion. Nevertheless, there are instances where its application for nonlinear distortion reduction is inconvenient or impossible. Such instances often occur at the beginning or end of a signal transmission system. In the audio field, it is difficult to generate an error signal to correct any nonlinear distortion arising in a record pickup. Loudspeaker nonlinear distortion, at the opposite end of the system, is usually more important because of its greater magnitude. Because of reverberation and phase shifts, it is not generally practical to apply negative feedback between the sound output of a loudspeaker and its driver. On the other hand, feedback derived from a separate winding on the voice coil will be imperfectly related to the actual sound output. In this instance, where negative feedback is impractical or inefficient, complementary nonlinear distortion can greatly reduce the distortion present in the speaker output.



(a) (b)

Fig. 1—(a) Typical input-output characteristic for a nonlinear circuit. (b) Input-output characteristic complementary to that of (a).

There are two ways by which complementary distortion correction may be applied. The usual way, which will be designated postdistortion, is that illustrated in Fig. 1. Here the complementary distortion acts on the originally distorted signal. Comparable but not identical results can be obtained, however, if the signal is first in-

* Manuscript received by the PGA, July 6, 1959.
† Texas Instruments Incorporated, Dallas, Tex.
[1] P. A. Reiling, U. S. Patent No. 2,293,628, issued August 18, 1942.
[2] G. Guanella, U. S. Patent No. 2,776,410, issued January 1, 1957. A means alternative to complementary distortion correction is described in this patent.

tentionally predistorted and then passed through the original distorting circuit. Correction of loudspeaker nonlinear distortion is an instance where pre- but not postdistortion is applicable.

The practical application of complementary distortion requires that the original distortion dependence on amplitude be known over the amplitude range of interest. If the input-output response characteristic is very irregular or has strong discontinuities in slope or value, it cannot be well represented by a rapidly convergent power series, and complementary distortion correction will not be practical. It will therefore be assumed that the input-output characteristic is a smooth function and can be represented by a power series of the form

$$e_1 = \sum_{m=1}^{M} a_m e_0{}^m, \tag{1}$$

where M may be finite or infinite, e_0 is the input signal and e_1 the output. The zero-order, or dc term, has been omitted for simplicity.

In general, the a_m coefficients will be functions of frequency, and the complementary distortion circuit will then also have to be frequency dependent to yield maximum distortion reduction over the amplitude and frequency ranges of interest. In practice, however, the coefficients may be frequency independent over much of the range. While it is possible to incorporate frequency-sensitive nonlinear correcting elements in the complementary distortion circuit to compensate for the frequency dependence of the original distortion, such complications will not be pursued further herein.

The coefficients of the input-output-characteristic power series must be known to allow design of the complementary distortion circuit. Haber and Epstein[3] have given equations which allow these coefficients to be calculated from the results of harmonic distortion measurements together with measurements of the polarity of the harmonics. As shown below, these coefficients may then be used to determine the corresponding coefficients in the power series describing the pre- or post-complementary distortion circuit. Finally, the resulting nonlinear characteristic can be realized in a practical circuit using diodes and resistors and other components and techniques well known in the analog computer art. It should be emphasized that both pre and post-distortion techniques are also applicable when the aim is not as linear amplification as possible but instead a close approximation to some more complicated functional relationship between input and output, such as, for example, square-law output with minimum linear and higher-than-second-order output terms.

Mathematical Analysis

For postdistortion, (1) may be used to represent the characteristic of the device or circuit whose distortion is to be reduced by a subsequent complementary distortion circuit. The characteristic of the latter may then be written

$$e_2 = \sum_{n=1}^{N} b_n e_1{}^n. \tag{2}$$

Here the b_n's must be determined in terms of the a_m's to minimize resultant distortion. For convenience, in the predistortion case (1) will be used to represent the initial complementary distortion while (2) will then describe the original distorting device or circuit. Thus, in this case, the b_n's are assumed known and the a_m's are to be determined as functions of them.

Substituting (2) in (1) yields

$$e_2 = \sum_{n=1}^{N} b_n \left[\sum_{m=1}^{M} a_m e_0{}^m \right]^n \equiv \sum_{s=1}^{NM} c_s e_0{}^s, \tag{3}$$

where the c_s's are new coefficients whose values, determined from (3), appear in the second column of Table I. Now for zero output distortion in e_2, we desire $e_2 = c_1 e_0$. To obtain the values of a_m or b_n which make the higher order c_s's zero, we can set these c_s's to zero and solve them individually to obtain the desired a_m's for predistortion or the b_n's for postdistortion. This procedure becomes very arduous as the order increases, and a preferable method is to use reversion of series.[4]

On setting $e_2 = c_1 e_0$ equal to (2) and solving for e_0, we obtain

$$e_0 = \frac{1}{c_1} \sum_{n=1}^{N} b_n e_1{}^n. \tag{4}$$

If now (1) is reverted to yield e_0 in terms of e_1, one obtains (see Appendix) an infinite series like (4) and comparison of terms yields the postdistortion b_n's directly in terms of the a_m's. The results up to fifth order are given in the third column of Table I. Note that when less than an infinite number of correction terms are used, there will remain residual distortion which, however, may be much lower than that originally present.

A similar procedure can be carried out for predistortion. Eq. (4) may be written as

$$c_1 e_0 = \sum_{n=1}^{N} b_n e_1{}^n.$$

Next, this series may be reverted to yield e_1 in terms of $(c_1 e_0)$ and the result compared with (1). Using $c_1 = a_1 b_1$, the equations of column 4 of Table I are obtained. Columns 3 and 4 of Table I are the basic results of the present work. Although expressions connecting the coefficients have only been stated to the fifth order, higher-order terms may be readily obtained from the corresponding known expressions tabulated in the reversion of series method.[4]

[3] F. Haber and B. Epstein, "The parameters of nonlinear devices from harmonic measurements," IRE Trans. on Electron Devices, vol. ED-5, pp. 26–28; January, 1958.

[4] H. B. Dwight, "Tables of Integrals and Other Mathematical Data," 3rd Ed., The Macmillan Co., New York, N. Y., p. 11; 1947.

TABLE I

EXPRESSIONS FOR c_i, b_i, AND a_i

i	c_i	Postdistortion, $c_i = 0$ b_i	Predistortion, $c_i = 0$ a_i
1	a_1b_1	b_1	a_1
2	$a_2b_1 + a_1^2b_2$	$-\left(\frac{b_1}{a_1^2}\right) a_2$	$-\left(\frac{a_1^2}{b_1}\right) b_2$
3	$a_3b_1 + 2a_1a_2b_2 + a_1^3b_3$	$\left(\frac{b_1}{a_1^4}\right)(2a_2^2 - a_1a_3)$	$\left(\frac{a_1^3}{b_1^2}\right)(2b_2^2 - b_1b_3)$
4	$a_4b_1 + (a_2^2 + 2a_1a_3)b_2 + 3a_1^2a_2b_3 + a_1^4b_4$	$\left(\frac{b_1}{a_1^6}\right)[5a_2(a_1a_3 - a_2^2) - a_1^2a_4]$	$\left(\frac{a_1^4}{b_1^3}\right)[5b_2(b_1b_3 - b_2^2) - b_1^2b_4]$
5	$a_5b_1 + [2(a_1a_4 + a_2a_3)]b_2 + [3(a_1^2a_3 + a_1a_2^2)]b_3 + [4a_1^3a_2]b_4 + a_1^5b_5$	$\left(\frac{b_1}{a_1^8}\right)[6a_1^2a_2a_4 + 3a_1^2a_3^2 + 14a_2^4 - a_1^3a_5 - 21a_1a_2^2a_3]$	$\left(\frac{a_1^5}{b_1^4}\right)[6b_1^2b_2b_4 + 3b_1^2b_3^2 + 14b_2^4 - b_1^3b_5 - 21b_1b_2^2b_3]$

The above reversion procedure shows that for either pre- or postdistortion correction to yield $e_2 = c_1e_0$ exactly, an infinite number of complementary correction terms will be required. For postdistortion, for example, each correction term acts on the original distortion to create higher-order distortion terms which, in turn, require the presence of higher-order correction terms to eliminate them and so on. Further, the larger the number of correction terms, the smaller in general the amplitude range over which reduced distortion is obtained. Nevertheless, a finite number of pre- or postcorrection terms can effect a very significant improvement in nonlinear distortion over a finite and important amplitude range.

EXAMPLE

The square-law distortion case, being simplest, will be used to show how complementary distortion may be applied for over-all distortion reduction. This case is useful also as an example since it can be solved directly, as shown in the Appendix. For predistortion, we shall take $N=2$ so that $e_2 = b_1e_1 + b_2e_1^2$, while $M=2$ for postdistortion yields $e_1 = a_1e_0 + a_2e_0^2$. As an illustration, we shall investigate in both cases how the residual total harmonic distortion (THD) varies with normalized amplitude when $c_2 = 0$ only (one correction term), when $c_2 = c_3 = 0$, and when $c_2 = c_3 = c_4 = 0$.

For both pre- and postdistortion, the procedure is to substitute the values of a_m or b_n which make the desired c_s's zero into the higher-order c's and thus evaluate the residual, nonzero distortion. The results of such a calculation with no distortion terms omitted are summarized in Table II. For convenience, we have set $a_1 = b_1 = 1$ in the results of Table II. The quantities a_1 and b_1 are merely scale factors such as amplification factors, and no significant generality is lost by taking them unity. It should be emphasized that these results are pertinent only to the square-law distortion case, as shown by the appearance of only a_2 and b_2. More complicated results would be obtained for cubic distortion or for a combination of quadratic and cubic distortion.

TABLE II

EXPRESSIONS FOR NONZERO c_i'S IN VARIOUS PRE- AND POSTDISTORTION CASES

	$c_2=0$		$c_2=c_3=0$		$c_2=c_3=c_4=0$	
	Pre	Post	Pre	Post	Pre	Post
c_3	$-2b_2^2$	$-2a_2^2$	0	0	0	0
c_4	b_2^3	$-a_2^3$	$5b_2^3$	$5a_2^3$	0	0
c_5	0	0	$-4b_2^4$	$6a_2^4$	$-14b_2^4$	$-14a_2^4$
c_6	0	0	$4b_2^5$	$2a_2^5$	$14b_2^5$	$-28a_2^5$
c_7	0	0	0	0	$-20b_2^6$	$-20a_2^6$
c_8	0	0	0	0	$25b_2^7$	$-5a_2^7$
c_9	0	0	0	0	0	0

Next, the results of Table II together with

$$e_2 = \sum_1 c_s e_0^s$$

may be used to obtain the harmonic distortion terms in each case. If one takes $e_0 = A \cos \omega t$ and expresses all powers of $\cos \omega t$ as harmonic terms, one finally obtains the finite series of harmonics pertinent to each case. For example, for $c_2 = 0$ and predistortion, the result is

$$e_2 = e_0 - 2b_2^2e_0^3 + b_2^3k_0^4, \tag{5}$$

or

$$e_2 = [A - (3b_2^2A^3/2)] \cos \omega t + (b_2^3A^4/2) \cos 2\omega t - (b_2^2A^3/4) \cos 3\omega t + (b_2^3A^4/8) \cos 4\omega t, \tag{5}$$

where dc terms have again been omitted.

In general, we may write

$$e_2 = \sum_{s=1} c_s e_0^s \equiv \sum_{r=1} h_r \cos r\omega t \equiv A \sum_{r=1} g_r \cos r\omega t, \quad (6)$$

where the h_r's are the harmonic amplitudes and $g_r = h_r/A$. Next, let $x = Aa_2 \equiv Aa_2/a_1$ and $x = Ab_2 \equiv Ab_2/b_1$ for post- or predistortion, respectively. Here x is a normalized distortion amplitude variable, as shown by the result

$$\begin{aligned} e_1 &= a_1 e_0 + a_2 e_0^2 = a_1 e_0[1 + (a_2/a_1)e_0] \\ &= a_1 A \cos \omega t[1 + x \cos \omega t]. \end{aligned} \quad (7)$$

In the present case, x clearly measures the relative importance of the original square-law or second-harmonic distortion term. For small x, in fact, x equals twice the original, uncorrected THD. The normalized harmonics, g_r, can now be expressed entirely in terms of x and the results are summarized in Table III.

Rather than compare the predictions of these different cases by comparing individual harmonic behavior graphically, the THD's of the various cases will be compared as functions of x. We may write

$$\text{THD} = \left[\frac{\sum_{r=2} h_r^2}{\sum_{r=1} h_r^2}\right]^{1/2} = \left[\frac{\sum_{r=2} g_r^2}{\sum_{r=1} g_r^2}\right]^{1/2}. \quad (8)$$

Note that g_1 may reach zero in most of the cases of Table III. In such cases, the THD will be unity and there will be no fundamental component present. The THD is sometimes written as

$$\text{THD}' = \left[\sum_{r=2} h_r^2/h_1^2\right]^{1/2}, \quad (9)$$

and would be infinite at the points where h_1 or g_1 were zero. The distinction between the two forms is only important for high values of THD anyway, and this is generally not the region of most physical interest.

TABLE III

NORMALIZED HARMONIC AMPLITUDES AS FUNCTIONS OF x FOR VARIOUS CASES

No Corrections		$c_2 = 0$		$c_2 = c_3 = 0$		$c_2 = c_3 = c_4 = 0$	
Case	A	B Pre	C Post	D Pre	E Post	F Pre	G Post
g_1	1	$1 - \frac{3}{2}x^2$	$1 - \frac{3}{2}x^2$	$1 - \frac{5}{2}x^4$	$1 + \frac{15}{4}x^4$	$1 - \frac{5x^4}{16}(28 + 35x^2)$	$1 - \frac{5x^4}{16}(28 + 35x^2)$
g_2	$\frac{x}{2}$	$\frac{x^3}{2}$	$-\frac{x^3}{2}$	$\frac{5}{8}x^3(4 + 3x^2)$	$\frac{5x^3}{16}(8 + 3x^2)$	$\frac{35}{16}x^5(3 + 5x^2)$	$-\frac{35}{16}x^5(6 + x^2)$
g_3	0	$-\frac{x^2}{2}$	$-\frac{x^2}{2}$	$-\frac{5}{4}x^4$	$\frac{15}{8}x^4$	$-\frac{35}{16}x^4(2 + 3x^2)$	$-\frac{35}{16}x^4(2 + 3x^2)$
g_4	0	$\frac{x^3}{8}$	$-\frac{x^3}{8}$	$\frac{x^3}{8}(5 + 6x^2)$	$\frac{x^3}{8}(5 + 3x^2)$	$\frac{x^5}{32}(84 + 175x^2)$	$\frac{x^5}{32}(168 + 35x^2)$
g_5	0	0	0	$-\frac{x^4}{4}$	$\frac{3}{8}x^4$	$-\frac{x^4}{16}(14 + 35x^2)$	$-\frac{x^4}{16}(14 + 35x^2)$
g_6	0	0	0	$\frac{x^5}{8}$	$\frac{x^5}{16}$	$\frac{x^5}{16}(7 + 25x^2)$	$-\frac{x^5}{16}(14 + 5x^2)$
g_7	0	0	0	0	0	$-\frac{5}{16}x^6$	$-\frac{5}{16}x^6$
g_8	0	0	0	0	0	$\frac{25}{128}x^7$	$-\frac{5}{128}x^7$
g_9	0	0	0	0	0	0	0

The dependence on x of per cent THD calculated from (8) with the help of Table III is shown in Fig. 2. The cases shown pertain to predistortion only. There is no difference between the pre- and postdistortion THD's in cases B and C. The limiting slopes for small x are, reading upwards, 4, 3, 2, and 1. Note that F yields minimum THD for $x<0.31$, while D is better than BC for $x<0.21$, and A (no correction) is better than BC for $x>0.53$. In general, the more correction terms present, the smaller the value of x below which the corrected THD is less than the uncorrected value. Nevertheless, when x is small, several correction terms can greatly reduce the over-all distortion. For an uncorrected THD of 2 per cent, three-term predistortion correction (case F) reduces the THD to about 0.0011 per cent.

Finally, Fig. 3 shows a comparison between the pre- and postdistortion cases for $c_2=c_3=0$ and for $c_2=c_3=c_4=0$. The quantity R is the ratio of the postdistortion THD to that for predistortion. These curves show that postdistortion results in somewhat lower THD than predistortion when two correction terms are used, but that the reverse is the case when three correction terms are used. In the region of considerable distortion improvement however, say for $x\leq 0.2$, there is no significant difference between the pre- and postdistortion corrections. Thus, which of the two methods to use in practical cases can be determined solely on the basis of applicability or simplicity.

Appendix

Given simple square-law distortion of the form $e_1=a_1e_0+a_2e_0^2$, we can revert this finite series directly to obtain

$$e_0 = -\frac{a_1}{2a_2}+\left[\left(\frac{a_1}{2a_2}\right)^2+\frac{e_1}{a_2}\right]^{1/2}. \qquad (10)$$

If $|4a_2e_1/a_1^2|<1$, expansion yields

$$e_0 \cong (1/a_1)e_1-(a_2/a_1^3)e_1^2+(2a_2^2/a_1^5)e_1^3\cdots. \qquad (11)$$

Comparison with (4) yields $c_1=a_1b_1$ and the values of b_2 and b_3 given in the postdistortion column of Table I (with $a_3\equiv 0$).

The series expansion in (11) on which the reversion solution is based is only convergent for $|4a_2e_1/a_1^2|<1$. This condition may be written

$$4\left|\, a_2(e_0/a_1)+a_2^2(e_0/a_1)^2 \right| < 1, \qquad (12)$$

or

$$\left|\,(a_2A/a_1)\cos\omega t+(a_2A/a_1)^2\cos^2\omega t\right| < \tfrac{1}{4}. \qquad (12a)$$

Now the only maxima of the left-hand side, considered as a function of (ωt), which will satisfy the inequality is that obtained for $\cos\omega t=0$. Thus, the most restrictive condition following from (12a) is

$$|x+x^2| < \tfrac{1}{4}. \qquad (12b)$$

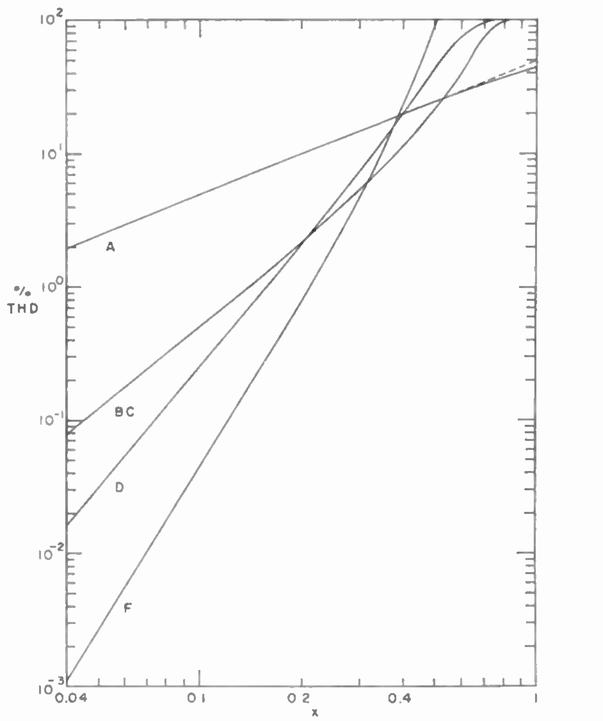


Fig. 2—Log-log plots of per cent total harmonic distortion in various cases (defined in Table III) vs the normalized distortion variable x for original square-law distortion only.

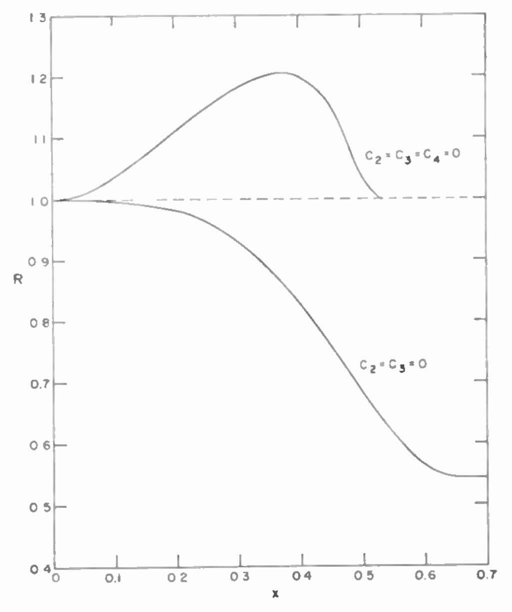


Fig. 3—Total harmonic distortion with postdistortion divided by that with predistortion vs x.

Replacing the inequality by an equality and solving for x yields $x=(\sqrt{\tfrac{1}{2}}-\tfrac{1}{2})=0.207$. This result shows that in the present case the reversion solution with an infinite number of power-law correction terms is convergent for $x<0.207$. For smaller x, an infinite number of complementary distortion terms may be used to correct completely (in principle) for the original distortion. When $x\geq 0.207$, complementary distortion correction with an infinite number of correction terms is impossible, but Fig. 2 shows that some improvement with a finite number of terms is still possible for x somewhat greater than 0.207.

In the present simple case, there is an alternative to the reversion solution which will allow complete distortion correction for a different range of x. We wish to obtain $e_2=c_1e_0$, with all higher-order terms zero. From (10), this equation may be written

$$e_2 = c_1\left[\frac{-a}{2a_2} + \left\{\left(\frac{a_1}{2a_2}\right)^2 + \frac{e_1}{a_2}\right\}^{1/2}\right]. \tag{13}$$

If c_1, a_1, and a_2 are known, the output e_1 from the nonlinear circuit whose nonlinearity is to be corrected may be passed through an analog computer which operates on e_1 in accordance with (13) to produce the undistorted output $e_2=c_1e_0$. Eq. (13) can only be applied when the radicand is not less than zero. This condition leads to

$$(4x \cos \omega t)[1 + x \cos \omega t] \geq -1. \tag{14}$$

The most restrictive condition is $\cos \omega t=-1$, leading to $4x-4x^2=1$. The solution of this equation is $x=\frac{1}{2}$, the maximum value permitted.

It should be emphasized that a closed-form reversion of the type illustrated by (10) is only possible in the simplest cases. In general, the series-reversion method must be used with a finite number of correction terms, leaving finite residual distortion. Further, as the complexity of the distortion to be corrected increases, it is likely that the range over which effective distortion reduction can be produced will diminish.

Acknowledgment

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Correspondence

Nonlinear Temperature-Compensating Devices for Transistors*

In a recent letter,[1] this author pointed out the possible error in predicting transistor circuit temperature instability by considering I_{co} variations only. This letter was concluded with the statement that linear stabilizing techniques were expensive in terms of gain loss. There appears a general reluctance to use nonlinear methods, possibly because of the difficulty of analyzing the performance of nonlinear elements as compensating devices.

Lin and Barco[2] have published results of the application of nonlinear elements to the stabilization of RF transistor circuits. The application to audio amplifiers is not as easy because of the practical difficulty of "by-passing" the nonlinear elements for audio frequencies. It is possible, however, to use and analyze the performance of nonlinear elements such as point contact and junction diodes as well as thermistors.

The conventional amplifier of Fig. 1(a)

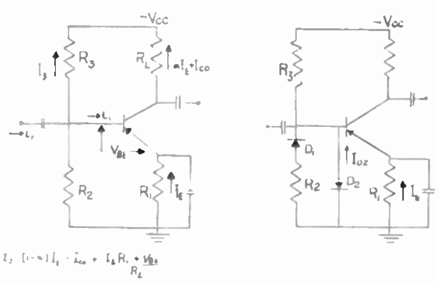

Fig. 1—Temperature-compensated amplifier.

can be modified by the use of junction diodes of the form of Fig. 1(b). The function of diode D_1 is to lower the base voltage at the same rate as V_{BE} falls with increasing temperature. The function of diode D_2 is to cancel the I_{co} of the transistor through R_3. Since D_1 is forward biased, its effective resistance is low and the value of R_2 can be decreased slightly to maintain the same bias. D_2 is reverse biased and presents a high resistance to ground, which does not interfere with the ac or dc operation of the circuit.

To analyze this circuit, we may assume the emitter current (or any other bias parameter of interest) to be a function of the temperature terms

$$I_E = f(V_{BE}, I_{co}, V_{D_1}, I_{D_2}). \tag{1}$$

This expression can be differentiated with respect to temperature:

$$\frac{dI_E}{dT} = \frac{\partial I_E}{dV_{BE}} \cdot \frac{dV_{BE}}{dT} + \frac{\partial I_E}{\partial I_{co}} \cdot \frac{dI_{co}}{dT} + \frac{\partial I_E}{\partial V_{D1}} \cdot \frac{dV_{D1}}{dT} + \frac{\partial I_E}{\partial I_{D_2}} \cdot \frac{dI_{D2}}{dT}. \tag{2}$$

The partial differential terms can be evaluated by taking a suitable mesh equation and differentiating with respect to each term. If this is done it can be shown that

$$D_v = \frac{\partial I_E}{\partial V_{BE}} = -\frac{1}{R_1}$$

$$D_i = \frac{\partial I_E}{\partial I_{co}} = \frac{G_1}{G_3 + G_2}$$

$$D_{vD} = \frac{\partial I_E}{\partial V_{D_1}} = \frac{R_3}{R_1(R_2 + R_3)}$$

$$D_{iD} = \frac{\partial I_E}{\partial I_{D_2}} = -\frac{G_1}{G_2 + G_3}.$$

Eq. (2) can now be integrated over the temperature range of interest to obtain the change in I_E

$$I_E(T_2) - I_E(T_1)$$
$$= \frac{G_1}{G_2 + G_3}[I_{co}(T_2) - I_{co}(T_1)] - \frac{1}{R_1}[V_{BE}(T_2) - V_{BE}(T_1)] - \frac{G_1}{G_2 + G_3}[I_{D_2}(T_2) - I_{D_2}(T_1)] + \frac{R_3}{R_1(R_2 + R_3)}[V_{D_1}(T_2) - V_{D_1}(T_1)]. \tag{3}$$

In the event that the compensating diodes were perfectly matched to the characteristics of the transistor, the compensation would be nearly perfect. On the other hand, the total change in I_E can be calculated no matter what the characteristics of the two diodes are. It is generally necessary to determine the temperature characteristics of the diodes experimentally. More complicated networks can be analyzed in the same fashion. A graphical method of designing a given temperature-current characteristic has been outlined by Keonjian and Schaffner[3] which can be used to "fit" a compensating network to wide variety of temperature dependent devices.

To illustrate the saving in signal power, it can be seen in Fig. 1 that R_2 and R_3 provide a shunt path for signal currents. The attenuation is obviously

$$\beta = \frac{i_{in}}{i_T} = \frac{R_2R_3/(R_2 + R_3)}{h_{11} + R_2R_3/(R_2 + R_3)} = \frac{1}{1 + \dfrac{h_{11}|D_v|}{D_i}}.$$

It is apparent that the more stable we attempt to make this circuit by decreasing D_i, the greater will be the loss in signal power.

The curves in Fig. 2 illustrate the possi-

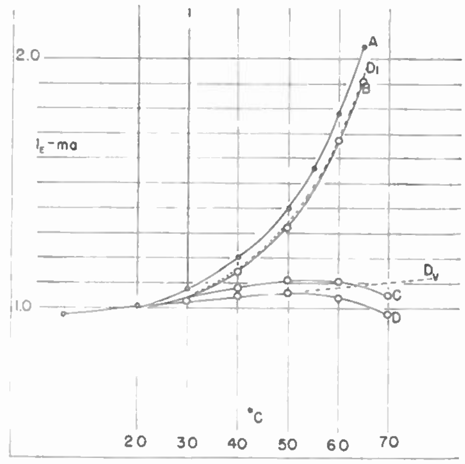

Fig. 2—Nonlinear stabilization. (A) Linear compensation $D_i = 15$; $D_v = 0.001$. (B) Effect of D_1. (C) Effect of D_2. (D) Combined effect of D_1 and D_2.

bilities of this type of compensation. The original circuit had a D_i of 15 and a D_v of 0.001 with a resultant attenuation of

$$\beta = 0.91.$$

To achieve this result by means of linear methods would require a D_i of approximately 1 and a resultant attenuation of 0.3.

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