IRE Transactions

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 assessment of $2.00.

IRE TRANSACTIONS® on AUDIO

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Message from the New National Chairman

Harry F. Olson

It is a distinct honor for me to be elected National Chairman of the Professional Group on Audio of the Institute of Radio Engineers. Election to this office carries with it a great responsibility because this group is now the largest audio organization. The Chairmanship presents a challenge because of the record of progress established by the seven past Chairmen. In this connection, I would like to mention the successful term of the outgoing Chairman, Dr. Daniel W. Martin. You will be pleased to know that Benjamin B. Bauer has accepted reappointment as Secretary-Treasurer and that Professor Alexander B. Bereskin will continue as Editor-in-Chief for 1957–1958. These two appointive offices require a tremendous amount of work combined with considerable judgement. We are indeed fortunate that these offices are so competently filled. In addition, we have an able and experienced group on the Administrative Committee for the year 1957–1958.

The main function of the Professional Group on Audio is the dissemination of information on the technology of communication in the audio frequency range through the publication of the IRE TRANSACTIONS ON AUDIO and by holding audio sessions at the national and regional conventions. Progress in implementing and perfecting these objectives has been made with every passing year.

For many years PGA has organized more than one session at the IRE National Convention in New York. This year we had three sessions. In addition, PGA sponsors audio sessions at the National Electronics Conference and at WESCON. In this connection, it may be desirable for the PGA to conduct special meetings as has been done by other professional groups.

The PGA Chapters perform an important function as is evidenced by the fact that the chapters continue to grow in number.

The IRE TRANSACTIONS ON AUDIO is now a recognized journal of high standing. This is confirmed by the fact that the three papers selected for the W. R. G. Baker Award were published in these TRANSACTIONS. The PGA Awards system has stimulated further interest among authors in writing high grade papers for the TRANSACTIONS.

The IRE Affiliate Plan[1] is a new venture that recognizes the spreading of electronics in every walk of scientific and technological life. It enables the IRE to further its aim as a professional engineering society: to advance radio engineering and the related fields of engineering and science. The IRE Affiliate Plan will enhance the main function of the PGA, namely, the dissemination of information in the technology of communication in audio frequency range, because the affiliated societies are strong and well-established organizations of exceptionally high standing. The IRE Affiliate Plan in the Professional Group on Audio will, in effect, unite the entire acoustic and audio fraternity.

A decade or so ago, many felt that the field of audio was tapering off. Instead of showing a decreasing trend, developments in the manifold aspects of audio have expanded at a tremendous rate. This is confirmed by the fact that the total membership in the Professional Group on Audio is now about 4000. The expansion is due in part to the ever accelerated interest in sound reproduction as exemplified by disk and tape records in both single channel and stereophonic, in improved sound in radio broadcasting, in stereophonic sound motion pictures, and in television sound. A field which is expanding rapidly is the "new acoustics" in the general complex of communications involving analysis, synthesis, encoding, coding, and decoding of speech, the control of sound and noise by both passive and active transducers, and the production of music by various electronic means.

It is quite evident that the future for audio in all its varied phases is brighter than it has been in any past period. It is the responsibility of all of us to make certain that the Professional Group on Audio fulfills its aims and functions in this ever expanding field of audio.

[1] W. R. G. Baker, "The IRE 'affiliate' plan—a new venture in engineering society structure and service," PROC. IRE, vol. 45, p. 278; March, 1957.

PGA News

ADMINISTRATIVE COMMITTEE MEETING MINUTES

New York, N. Y.—March 19, 1957

Members Present

S. J. Begun
A. B. Bereskin
M. S. Corrington
W. E. Kock
D. W. Martin
H. F. Olson
F. H. Slaymaker
B. B. Bauer

Members Absent

A. B. Jacobsen
W. Goodale

Committee Chairmen Present

A. B. Bereskin
M. Camras
M. Copel
M. S. Corrington
W. E. Kock
H. F. Olson
P. Williams

Committee Chairmen Absent

A. B. Jacobsen

Newly Elected Members Present

H. F. Olson, *Chairman*
F. H. Slaymaker, *Vice-Chairman*
W. T. Selsted
P. Williams

The meeting was held in Suite 4N & P of the Waldorf-Astoria and was called to order at 7:00 P.M.

1) Affiliate Membership

Chairman Martin opened the meeting by announcing approval of the system of Affiliate Membership within the IRE Professional Groups. The following societies have been accredited within the interest of the Professional Group on Audio:

ASA—Audio & Ultrasonics
AIEE—Audio
SMPTE—Audio
AES—Audio.

It was agreed that a letter would be sent to these societies by the PGA Secretary or Chairman informing them of the Affiliate Plan. The Secretary is to prepare the draft of an article for publication by these societies explaining the Plan.

2) Report of Committees

Various Committee Chairmen reported on the activities of their Committees and the reports are appended herewith. It was moved by Kock and seconded by Slaymaker to approve these reports (with the exception of budgetary figures which were left for a later motion).

3) Advertising in TRANSACTIONS

A motion was made by Dr. Begun to announce that advertisements would be accepted by the Chairman of the Ways and Means Committee, and to consider the possibilities of hiring an Advertising Manager, possibly jointly with another Professional Group, to handle the selling of advertisements.

In view of subsequent discussion about the problems this action would involve, Dr. Begun withdrew this motion.

4) Discussion about Constitution and Bylaws

Considerable discussion took place about the proposed new Constitution prepared by M. S. Corrington. Several paragraphs were read and commented upon by various members. Chairman Martin decided to postpone this discussion for a special future meeting to be called by newly elected Chairman Olson—possibly coincident with the forthcoming meeting of the Acoustical Society in May. In the meantime all members were urged to send comments to the Chairman of the Committee.

At this point the meeting was turned over to Chairman Olson and the following matters were handled.

5) Appointments

Chairman Olson announced the following appointments:

For the year 1957–1958, B. B. Bauer, *Secretary-Treasurer*.

It was moved by Kock and seconded by Slaymaker to approve this appointment, with everyone voting in the affirmative.

The following Committee Chairmen were nominated:

Editorial Committee—A. B. Bereskin
Awards Committee—F. H. Slaymaker
Program Committee—P. Williams

Nominating Committee—W. E. Kock
Tapescripts Committee—A. B. Jacobsen
Ways & Means Committee—M. Copel
Constitution & Bylaws Committee—M. S. Corrington
Publications Review Committee—W. Selsted
Chapters Committee—M. Camras
Papers Procurement Committee—S. J. Begun.

Chairman Olson urged the newly appointed or reappointed Committee Chairmen to form their Committees as soon as possible.

6) Increase of Services to Members

No action appeared necessary.

7) Remuneration for Editor

This possibility was discussed, as several Professional Groups employ a paid editor. Professor Bereskin stated that remuneration was not required, especially since the University of Cincinnati agreed to donate secretarial services to the PGA.

8) Follow-Up on Student Papers Award

By motion of Corrington, seconded by Kock, it was unanimously agreed to refer this matter to the Awards Committee.

9) Budget

The following budget was presented for the fiscal year 1957–1958.

Editorial	\$8500.00
Awards	600.00
Tapescripts	300.00
Contingencies	500.00
Total	\$9900.00

By motion of Kock, seconded by Corrington, this budget was approved unanimously.

10) Matching Funds

It was agreed that matching funds were essential to continued operation of the PGA. By motion of Bereskin, seconded by Slaymaker, it was agreed to table the question of matching funds.

11) Sessions at the NEC

It was moved by Corrington, seconded by Selsted, and unanimously approved that PGA sponsor a Technical Session at the NEC in the Fall of 1957, and at the same time hold an Administrative Committee meeting coincident with that conference.

12) Publicity in the IRE PROCEEDINGS

There was discussion about the fact that PGA does not appear to get its share of publicity in the PROCEEDINGS. In view of the lateness of the hour, by motion of Bereskin, seconded by Kock, it was agreed to table this discussion for the present.

BENJAMIN B. BAUER
Secretary-Treasurer

PGA NATIONAL MEETING AT IRE NATIONAL CONVENTION

PGA Chairman's Report

Our principal PGA goals during the past year have been to extend our assistance to chapters, to improve PGA-sponsored technical sessions, and to improve our publication, IRE TRANSACTIONS ON AUDIO. As retiring Chairman of PGA, I am pleased to report progress to you in all of these areas and in others, too.

A new communication called "The PGA Chapter" is now being sent regularly from the Chapters Committee to the officers of local PGA Chapters. This contains up-to-date news on recent or future programs of other chapters, suggestions on operating a chapter, on program solicitation and arrangements, and announcements of national PGA news of local interest. Your Chairman has now visited nearly half of the PGA chapters currently active in twenty sections of IRE. These visits revealed a genuine professional interest in audio at the local level, which can contribute to PGA nationally if encouraged to develop.

The PGA program at the National Electronics Conference last fall, and the three audio sessions at this convention, have generated considerable interest and have been well attended. Yesterday's session on Home Measurements by Engineers on Their Hi-Fi Equipment had an overflow audience listening to the speakers by loudspeaker in another auditorium.

The IRE TRANSACTIONS ON AUDIO has continued to improve in content. Evidence of this is the high quality of the papers receiving PGA Awards, and the decision of IRE to give the first W. R. G. Baker Award to a group of three authors who produced a series of tutorial papers especially for our publication.

During the year PGA instituted a new Student Papers Award and supplied several articles on audio to the IRE STUDENT QUARTERLY, in order to acquaint student engineers with the opportunities and challenges in the audio field.

As part of the IRE Affiliate Plan, PGA has voted to accredit four other technical societies so that their members are now eligible to become affiliates of the PGA, and receive our publications, attend our technical sessions, participate in chapter activities, receive PGA Awards, or hold appointive office. This we hope will strengthen our ties with technical societies which share our scientific and engineering interests. Last summer PGA joined with other participating organizations and agencies in supporting the second International Congress on Acoustics in Cambridge, Mass.

PGA membership has increased steadily again this year, and our finances are in excellent condition, as will be reported in detail by our Secretary-Treasurer. We are studying a draft of a proposed revision of the PGA Constitution and By-Laws which will eventually be submitted to you for approval.

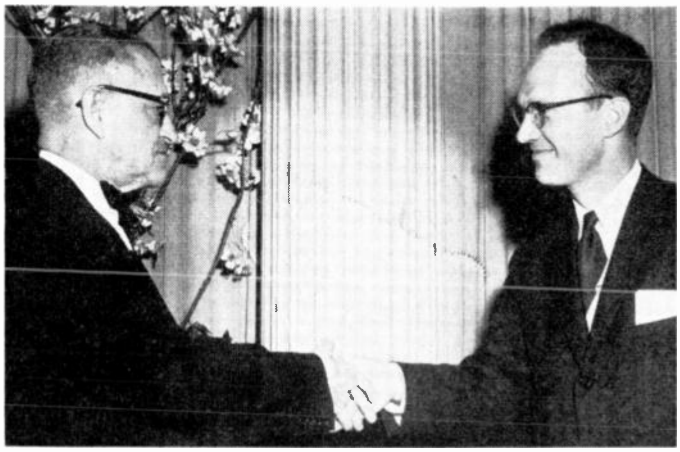
It has been a pleasure to serve the Professional Group on Audio as Chairman. Thank you for this opportunity. In closing I want to recognize the generous and helpful assistance during the past year of a number of fine people. First we should mention the great assistance given by IRE Headquarters, the Professional Groups Committee, and especially the IRE Publication Department. Also the following people deserve your hearty applause for their voluntary efforts on your behalf during the past year.

A. B. Bereskin and the Editorial Committee for TRANSACTIONS.
M. Camras for the PGA Chapter, and any Chapter Chairmen present (or their representatives).
P. Williams and the Program Committee.
F. Slaymaker for audio sessions at this convention.
M. Corrington on Constitution and Bylaws.
H. F. Olson and the Awards Committee.
W. E. Kock and the Nominations Committee.
A. B. Jacobsen, Vice-Chairman and Tapescripts.
M. Copel for Ways and Means.
S. Wilpon as representative on CONVENTION RECORD Committee.

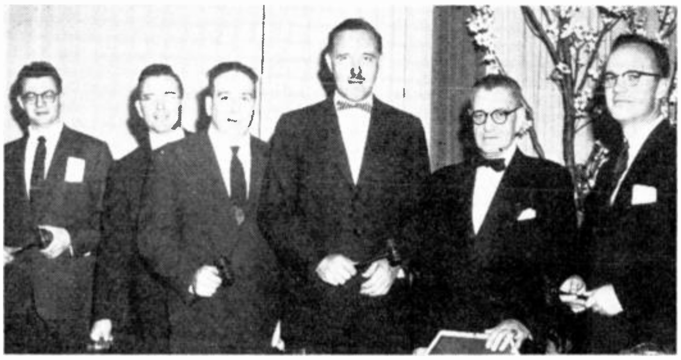
Last but by no means least, our ever helpful Secretary-Treasurer, B. B. Bauer, who will give his annual report to the membership.

1956-1957 Award Winners (left to right): R. J. Kircher, R. L. Trent, D. R. Fewer, W. R. G. Baker, A. B. Bereskin, Editor of TRANSACTIONS, H. F. Olson, and D. W. Martin.

W. R. G. Baker (left), Chairman of the Professional Groups Committee, is introduced by B. B. Bauer, Secretary-Treasurer of the IRE-PGA.

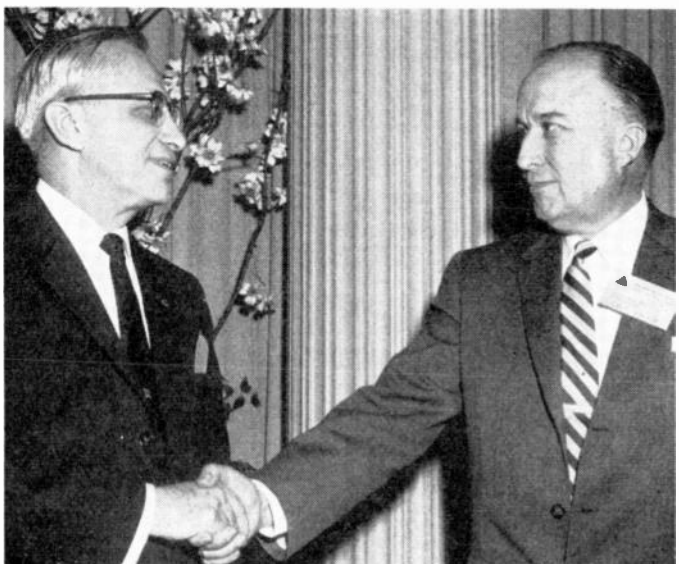


D. W. Martin presents a special gavel to W. R. G. Baker (left) from PGA in honor of his many contributions to the evolution of the Professional Groups System.



W. R. G. Baker awards gavels to former PGA Chairmen. (Left to right): B. B. Bauer (1951–1952), J. J. Baruch (1952–1953), M. Camras (1953–1954), W. E. Kock (1955–1956), Dr. Baker, D. W. Martin (1956–1957). Dr. Baruch received the gavel on behalf of L. Beranek (1950–1951), currently on a trip to Europe. V. Salmon (1954–1955) was unable to attend.

W. R. G. Baker (2nd from left) congratulates Award Winners (left to right): R. J. Kircher, R. L. Trent, and D. R. Fewer.



H. E. Roy (left) recipient of the IRE-PGA Achievement Award, is congratulated by H. F. Olson, Chairman of the Awards Committee and newly elected PGA Chairman.

D. W. Martin (left) turns over gavel to newly elected Chairman Olson as symbol of office.

PGA AWARD WINNERS

Winners of the IRE-PGA awards were announced in the March–April, 1957, issue of the IRE Transactions on Audio. Photographs of the winners and the certificate awarded them are shown here.

H. H. Kajihara

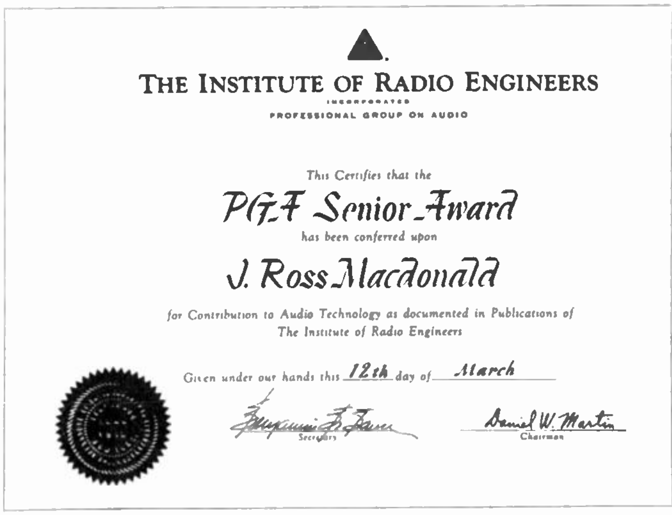
THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED
PROFESSIONAL GROUP ON AUDIO

This Certifies that the
PGA Award
has been conferred upon
Hitoshi H. Kajihara
for Contribution to Audio Technology as documented in Publications of
The Institute of Radio Engineers

Given under our hands this 12th day of March

Benjamin F. Baer, Secretary
Daniel W. Martin, Chairman

J. R. Macdonald



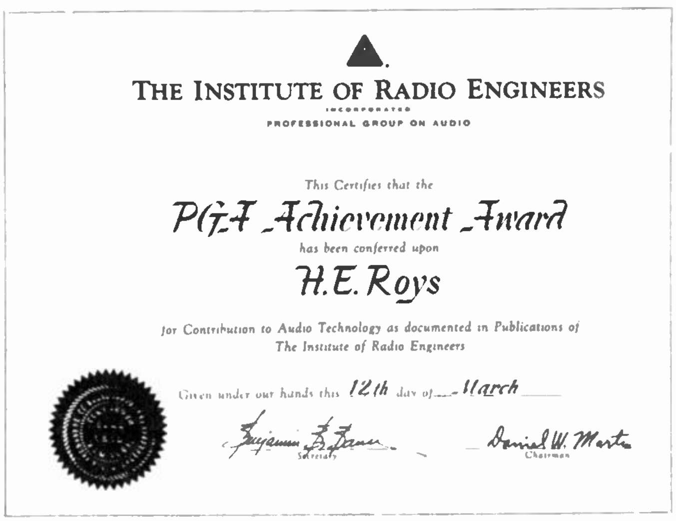
THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED
PROFESSIONAL GROUP ON AUDIO

This Certifies that the
PGA Senior Award
has been conferred upon
J. Ross Macdonald
for Contribution to Audio Technology as documented in Publications of
The Institute of Radio Engineers

Given under our hands this 12th day of March

Benjamin F. Baer, Secretary
Daniel W. Martin, Chairman

H. E. Roys



THE INSTITUTE OF RADIO ENGINEERS
INCORPORATED
PROFESSIONAL GROUP ON AUDIO

This Certifies that the
PGA Achievement Award
has been conferred upon
H. E. Roys
for Contribution to Audio Technology as documented in Publications of
The Institute of Radio Engineers

Given under our hands this 12th day of March

Benjamin F. Baer, Secretary
Daniel W. Martin, Chairman

PGA CHAPTER ACTIVITIES

Baltimore, Md.

Kenneth W. Betsh, Acting Chairman of the recently approved IRE-PGA Chapter in Baltimore, Md., reports that present plans of this chapter call for meetings in alternate months from September through May with election of officers at the May meeting. The March 13 meeting of the Baltimore IRE Section is being sponsored by the PGA Chapter. At this meeting Walter O. Stanton, president of Pickering, will discuss and demonstrate the features of Pickering phonograph pickups and electrostatic speakers.

Boston, Mass.

Richard S. Burwin, Doelcam Corp., 1400 Soldiers Field Road, Brighton, Mass., is the new Chairman of the PGA Boston Chapter. Please modify your list accordingly.

Chicago, Ill.

Robert J. Larson reports that Ralph H. Sprague and Ross Snyder of the Ampex Corporation presented a paper on Stereophonic Sound at the February 15 meeting of the professional group. The talk included a demonstration of 2-channel tape equipment followed by a discussion particularly suited for those interested in home applications.

Cleveland, Ohio

C. P. Germano reports the following meetings of the Cleveland Chapter:

October 4, 1956—"Tape Recording Clinic," with a demonstration of Ampex Stereo equipment by R. F. Dubbe of the Minnesota Mining and Manufacturing Co.

December 6, 1956—"Modern Ceramic Pickups," by J. F. Wood of Electro Voice, Inc.

Feburary 21, 1957—"Design of Acoustic Treatment of the Cleveland Public Auditorium," by J. L. Hunter, Professor of Physics at John Carroll University, and H. Mull, physicist with NACA. A demonstration was given of acoustic test equipment.

Dayton, Ohio

The following is a digest of a paper on "Acoustical Design and Properties of the C and N Lab's Anechoic Chamber," presented by Lt. John H. King and Arthur L. Peters, Jr., at the February 7 meeting of the PGA.

"Engineers and physicists now recognize an anechoic chamber as a necessity in a laboratory where accurate acoustical measurements are to be made on a regularly scheduled basis. A sound field generated in such a room therefore behaves essentially the same as it would in free space. That is, it follows the inverse square law wherein the sound pressure is reduced 6 db each time the distance from the sound source is doubled.

"Without such a room, measurements must be made using an outdoor facility provided with high towers isolated from surrounding reflecting surfaces. Measurements must be made under highly controlled conditions, that is, no wind, clear weather, normal temperatures, etc.

"Engineers of the Communication and Navigation Laboratory, Wright Air Development Center, who are continuously engaged in research, development and testing of speech communication equipment, drew up preliminary plans for an anechoic chamber early in 1950. Complete plans were then prepared by the Air Installations Office of Wright-Patterson Air Force Base. Final measurement work on the chamber was completed in February, 1953. An acoustical analysis of the properties of the anechoic chamber was made by the Acoustical Engineering Section of the Baldwin Piano Company, Cincinnati.

"The general design of the chamber was based upon the principles established by the Electro-Acoustic Laboratory of Harvard University, under the sponsorship of the Office of Scientific Research and Development of the National Defense Research Committee. Funds restrictions made it necessary to utilize a part of an existing building, limiting over-all space provided to $28 \times 24 \times 8$ feet.

"The inside of the walls was first covered with a one-inch thickness of cork. Over the cork was cemented a one-half inch thickness of insulating fiberboard. A coffering of fiberboard consisting of 16-inch squares, seven and seven-eighths inches deep, was built up to provide dead air space behind the wedges. To the coffering was attached furring on which 8×8 inch frames were installed to support the Fiberglas wedges.

"The wedge design adopted was the basic triangular wedge shape evolved from Electro-Acoustic Laboratory research. Wedges of 46 inches length, eight inches square at the base, were used on the walls. The over-all inside dimensions of the finished chamber are $17 \times 14\frac{1}{2} \times 9\frac{1}{2}$ feet. The floor is one and one-half feet above the tips of the lower wedges, leaving a working height of eight feet.

"The floor consists of rectangular sections of expanded-steel grating on $1\frac{1}{2}$-inch angle iron frames. The sections are supported by 2-inch diameter hollow steel posts, $1\frac{1}{2}$ feet above the tips of the lower wedges, stuffed and wrapped with Fiberglas Aerocor blanket in order to minimize reflection and air-column resonances. For very critical measurements, the experimental equipment can be supported from ceiling fixtures and the entire floor and its support can be removed.

Measurements Made

"Acoustical measurements made by engineers of the Baldwin Piano Company consisted of inverse square law, ambient noise, and wall-transmission loss measurements.

"The data showed that the chamber was entirely suitable for determination of the acoustic performance of headsets, microphones, and loudspeakers used by the Air Force. The chamber is affected by outside acoustical noises having intensities of over 75 db. The inverse square law measurements indicated variations of less than plus or minus 1 db in the frequency range of 100 to 10,000 cps.

"Considerable difficulty was encountered in the measurement work and many new techniques had to be devised and tried before suitable results were obtained."

Milwaukee, Wis.

William Alpert, Chairman of the Milwaukee PGA Chapter, reports a meeting held on January 14. At this meeting C. N. Hoyler of RCA Laboratories spoke on "New Adventures in Electronics," and gave a very interesting demonstration. This meeting was attended by 185 people.

Philadelphia, Pa.

A joint meeting with the Delaware Valley Chapter of the Acoustical Society of America on February 13 is reported by Mones E. Hawley of the Philadelphia Chapter.

The program consisted of a demonstration talk titled, "Speech, Hearing, and Music" and was presented by Floyd K. Harvey of the Bell Telephone Laboratories.

This talk featured physical concepts underlying the speech and hearing processes, analysis and synthesis of speech and music, narrow-band speech transmission, and automatic devices for the recognition of speech. The talk included several demonstrations on synthetically produced speech sounds and musical tones.

San Diego, Calif.

Larry La Zelle, Chairman of the San Diego Chapter reports a January meeting on "Instrumentation and Audio Test Equipment," and a February meeting on "A Report on the Los Angeles Audio Engineering Society Convention and Hi Fi Show."

Recent election results from this chapter indicate that George Somes of NEL is the new Chairman while Kenneth O'Neal of Neely Enterprises is the new Vice-Chairman. Larry La Zelle, Past Chairman, upon releasing his office, started a column "Audio Hum" in the *Engineering Bulletin of San Diego*.

San Francisco, Calif.

Don L. Broderick, Vice-Chairman of the San Francisco Chapter, reports a meeting on February 18 on "Transistor Amplifying Techniques" by Irwin Wunderman of the Hewlett-Packard Co. At this meeting Mr. Wunderman covered the basis of transistor action from a solid-state physics point of view and led up to the pertinent considerations for applying transistors as amplifying devices, including biasing arrangements, temperature stability, frequency response, and feedback.

Syracuse, N. Y.

On March 27, W. E. Stewart, of the Standard Register Co., Dayton, Ohio, presented a paper on "The Design of Hearing Aids." Mr. Stewart discussed the special problems involved in the design of both vacuum tube and transistor hearing aids and discussed the analysis of three years of transistor field failures.

WITH OTHER ACOUSTICAL AND AUDIO SOCIETIES

The January, 1957, issue of the *Journal of the Acoustical Society of America* contained numerous papers of interest.

W. Roth, Roth Laboratory for Physical Research, Hartford, Conn., deals with "Nondestructive Testing by Ultrasound: Objectives and Potentialities." Ultrasonic propagation is affected by elastic properties of material distribution of material and stress continuity. Since these are properties that determine structural strength of manufactured parts, ultra sound is useful for nondestructive testing. Variations in internal properties of structure can be detected, but often it is not possible to evaluate the effect that detected structural variations have on the ability of a part to function properly. This paper takes a critical view of the nondestructive testing field in the light of this objective.

"Acoustic Mapping Within the Heart" is an extremely interesting contribution by D. H. Lewis and G. W. Deitz, of the Philadelphia General Hospital, and J. D. Wallace and J. R. Brown, Jr., U. S. Naval Air Development Center, Johnsville, Pa. A cylindrical barium titanate element has been placed at the distal end of a specially designed catheter and used in studies of the heart sounds of dogs and humans. Heart sounds have been recorded with the element located in the four chambers of the heart as well as in the great vessels leading from the heart. X-ray photographs monitored the location of the catheter tip in each case. The sounds have been recorded on tape and simultaneously on a photographic galvanometer recorder; all studies were continuously monitored by electrocardiogram. Sound spectrographic analysis of the several sounds was made and analyzed.

B. Olney and R. S. Anderson of Olney and Anderson, Rochester, N. Y., titled their article "Acoustics of the Rochester (New York) War Memorial Auditorium."

This arena-type auditorium with a stage at one end has a volume of 3.24 million cubic feet, seats 5774 persons in stationary chairs and a total of nearly 10,000 with additional, portable chairs. Both the stationary and the portable chairs have upholstered seats. The acoustical hazard of concave rear wall and opposite wall containing the proscenium opening was ameliorated by special treatment. A combination structural acoustical ceiling was used. The design objective of a reverberation time of about 2.5 seconds in the unoccupied room at 500 cps was attained without adjustment. The results of reverberation-time measurements are shown and discussed. The public-address system is described and the subjective assessment of the auditorium is mentioned.

A. M. Liberman of Haskins Laboratories in New York and the University of Connecticut contributed "Some Results of Research on Speech Perception." Recent experiments with synthetic speech have succeeded in isolating some of the acoustic cues which underlie the perception of speech. This paper describes, and attempts to interpret, some of the research in that area.

G. V. Gersuni of the Pavlov Institute of Physiology of the Academy of Sciences of the U.S.S.R. wrote on "Concerning New Methods of Measurement of Hearing in Man." It is known that the main quantitative characteristics of hearing in man can be obtained by so-called psychophysical methods. In these methods the responses to sound stimuli are based upon the use of verbal instructions. This paper gives a description of methods of measurements of hearing based upon the use of other different responses to sound stimuli. A set of different conditioned responses was used (galvanic-skin reflexes, eyelid reflexes, electrocortical, and oculomotor reactions). The data obtained by these methods show the following: 1) absolute auditory thresholds and difference limens for frequency and intensity of pure tones can be measured with the same accuracy by these reactions as by verbal responses; 2) in certain cases conditioned reflexes subliminal to the verbal response in the range of 1 to 6 db may be detected; 3) changes of absolute sensitivity, attaining 25 to 30 db and dependent on the conditions under which the reactions take place, can be detected. The present data of hearing measurements obtained by means of different responses are considered as highly characteristic of the process of sound discrimination in man and animals.

In a letter to the editor, R. L. Hanson and W. E. Kock of Bell Telephone Laboratories, Murray Hill, N. J., point out an "Interesting Effect Produced by Two Loudspeakers under Free Space Conditions." A little over two years ago, an effect was observed in the free space room of the Bell Telephone Laboratories at Murray Hill which involved feeding an audio signal to two loudspeakers, one of which had its polarity reversed. A person standing at equal distances from and facing the two loudspeakers received the impression that the sound was coming from a point inside or very close to his head, and as he moved to and fro sideways, the sound seemed to move about behind his head. Inasmuch as slight rotational movements of the head do not change the apparent source position, the effect is similar to the binaural effect obtained with the use of headphones, wherein nearby sound sources appear to be near or inside one's head and in the rear.

In a letter about "Sonic Images" by E. E. Suckling, State University of New York, there is an interesting description about a new method of taking pictures of various objects by use of ultrasound instead of light as illumination.

The February, 1957, issue of the *Journal of the Acoustical Society* contains twenty-one articles, the following being of special interest:

"Response of the Skin to Focused Ultrasound," by E. Bell and T. S. Argyris of Arnold Biological Laboratory at Brown University and the Department of Anatomy, Harvard Medical School, is a very interesting contribution. Skin of mice has been irradiated with focused ultrasound of one megacycle frequency. Skin in the growing phase of the hair growth cycle becomes ulcerated within two days following treatment, while skin in the resting phase of the cycle is relatively unaffected. It is concluded that the response of the skin to ultrasound depends upon the physiological state of the organ at the time of irradiation.

F. P. Burns of Bell Telephone Laboratories, Murray Hill, N. J., had an article entitled "Piezoresistive Semiconductor Microphone." A microphone using the piezoresistive effect of *n*-type germanium has been designed and constructed with a cantilever beam arrangement. Frequency response curves were taken in a sound field of 10 microbars with the microphone in a bridge type circuit and biased with 9 and 18 milliamperes. The microphone produces 10^{-11} watt in the flat low-frequency range and peaks to 10^{-9} watt at a resonance at 2570 cps with a total bias of 18 ma in a bridge type circuit. Other designs of transducers and different types of semiconductor material are discussed. The performance of a germanium rod microphone is compared with a standard carbon button transmitter. The ac power output for the rod and carbon microphone are computed to be 3.6×10^{-11} watt and 6.40×10^{-4} watt, respectively.

J. K. Hilliard and W. T. Fiala of the Altec Lansing Corporation, Beverly Hills, Calif., collaborated on "Condenser Microphones for Measurement of High Sound Pressures." A series of small condenser microphones and accessory equipment has been developed for the measurement of high sound pressure levels. The paper derives expressions leading to the evaluation of the nonlinear distortion of the microphone, describes the construction of the several elements of the microphone system, and presents performance data. Four interchangeable microphones, all employing clamped glass plate diaphragms, provide linear measurements over a pressure range of 1 to 10^8 dynes/cm^2.

A paper entitled "Ear-Insert Microphone" was authored by R. D. Black of Radio Corporation of America, Moorestown, N. J. It has been observed by several different technical groups that an acoustic transducer placed at the ear can be used as a microphone to pick up the voice of the wearer. This paper presents the results of an investigation to determine the various factors which make possible the use of a microphone in the ear.

M. Lawrence and P. A. Yantis, Department of Otolaryngology and Institute of Industrial Health, University of Michigan, wrote a paper "Overstimulation, Fatigue, and Onset of Overload in the Normal Human Ear." The experiments reported here compare the shift in audibility threshold and in onset of overload following one-minute stimulations with a 1000-cps tone at sensation levels of 20, 60, 80, 90, 100, and 110 db.

"Influence of Noise upon the Equivalence of Intensity Differences and Small Time Delays in Two-Loudspeaker Systems," was written by D. M. Leakey and E. Colin Cherry, both of the Imperial College of Science and Technology, London, England. The equivalence of intensity difference and time delay in two-loudspeaker systems is greatly affected by the presence of extraneous noise. An experiment is described in which this effect is measured with speech and wide-band noise. It is found that as the noise level is increased the equivalent intensity difference, for a given time delay, decreases, tending to zero for noise levels approaching the masking level. The results have an important bearing on the design of sound reinforcement and stereophonic sound systems.

The issues contain their usual extensive references to current publications in acoustics by R. N. Thurston and R. W. Young.

The July, 1956, issue of the *Journal of the Audio Engineering Society* has five papers, all of which will be of interest to members of the Professional Group on Audio.

E. W. Franck and E. Schmidt, Reeves Soundcraft Corp., Springdale, Conn., deal with "New Products and New Applications in the Magnetic-Tape and Film Fields." In the last two years, advances in the manufacture and application of magnetic products have been made. This paper covers the physical and electrical characteristics of current standard tapes and Mylar tapes of various gauges. A report is given on "print-through" characteristics of materials coated on base supports as thin as 0.5-mil Mylar. The use of striped magnetic film in motion pictures and in television is dealt with, as is the use of magnetic products for geophysical recording. Discussed, too, are strides which have been made in improving the friction characteristics of tape and film. These improvements, claim the authors, have solved some of the problems associated with telemetering and other instrumentation applications.

H. E. Roys of the Radio Corporation of America, Indianapolis, Ind., and E. E. Masterson of Sperry-Rand, Inc., Philadelphia, Pa., did an article "The Radial Tone Arm—an Unconventional Phonograph Pickup Suspension." A new method of supporting a pickup while reproducing a phonograph record is described. The arrangement permits the pickup to follow along a radial line, thus eliminating tracking error. Static friction is greatly reduced, so that the force required to pull the pickup across the record is small. In addition, the arrangement described here provides mechanical resistance that is effective in a lateral direction, and hence effective in damping tone-arm resonance. The damping is ineffective in a vertical direction, however, so that the pickup readily follows warped records.

"Tweeter Design Considerations" was contributed by G. W. Sioles of University Loudspeakers, Inc., White Plains, N. Y. This paper covers the general constructional and design characteristics of various types of tweeters and elaborates on the fundamental design problems of the moving-coil, horn-loaded tweeter. Also dealt with are air-chamber requirements, magnet and gap design, horn characteristics, etc. Performance, reliability, and cost considerations are discussed for different types of tweeters.

A. L. Seligson of the U. S. Naval Material Laboratory in Brooklyn wrote "Free-Field Technique for Secondary Standard Calibration of Microphones." The acoustic environment required for the performance of free-field secondary standard microphone calibrations is examined. The technique includes automatic compensation for variations in sound output level vs frequency of the sound source. Size and orientation of the standard and object microphones and mounting are considered with a view toward minimizing disturbances in the sound field, and resulting calibration errors, arising from reflections at high frequencies. Maximum and minimum working distances from sound sources of various dimensions necessary to maintain plane-wave free-field conditions are given for a variety of microphone types. The accuracy limits of the calibration method are indicated.

Comments on the paper, "On Stylus Wear and Surface Noise in Phonograph Playback System," were given by D. A. Barlow. The original paper by F. V. Hunt appeared in the *Journal of The Audio Engineering Society* in January, 1955. When a set of conclusions is reached in a study as fundamental as this, it is certain that particular factors have been accepted as a part of the working hypothesis essential to the formulation of conclusions which are open to challenge by another student of the subject. Mr. Barlow's studies, like those of Professor Hunt, are thorough and represent another view of the same subject.

BENJAMIN B. BAUER

Transistorized RC Phase-Shift Power Oscillator*

LAWRENCE J. GIACOLETTO†

Summary—**A transistorized RC phase-shift oscillator is very similar to the electron tube circuit. Two important differences are: 1) A transistor possesses active phase-shift elements which can be utilized as part of the total phase shift; 2) large current capabilities of the transistor make it possible to obtain significant power at reasonable resistance values. The small-signal analysis given herein is based upon a hybrid-π transistor equivalent circuit which includes the active phase-shift elements of the transistor. Data are given for some operating circuits. Frequency modulation of the transistorized phase-shift oscillator can be readily carried out by modulating the transistor active phase-shift elements.**

Introduction

The transistorized version of an RC phase-shift oscillator is very similar to the electron tube circuit.[1] One difference is the inherently greater phase shift encountered in a transistor as compared with an electron tube. The phase-shift circuit employed with the transistor must be suitably chosen and designed to accommodate and, where possible, to utilize the additional phase shift. Another difference is the larger current capabilities and, therefore, lower resistance levels of a transistor as compared with an electron tube. This makes it possible to obtain significant power from the transistor at reasonable resistance values and with good efficiency.

A transistorized phase-shift oscillator has been considered recently. In this development[2] the phase shift of the transistor is not accommodated in the design equations so that design is limited to low frequencies where the phase shift is negligible.

Discussion

The small-signal (linear) operation of transistorized RC oscillators is considered in Appendix I. The circuits considered in the Appendix are shown in Fig. 1(a) and 1(b). A hybrid-π equivalent circuit[3] is employed for the transistors, but the feedback elements normally present ($g_{b'c}$ and $C_{b'c}$) are omitted. When the load conductance, G_l, is large so that the voltage amplification is small, feedback will be of small significance. If this is not the case, a first-order correction for feedback effects can be introduced by adding $g_{b'c}g_m/G_l$ and $C_{b'c}g_m/G_l$ in shunt with $g_{b'e}$ and $C_{b'e}$. Since the analysis of the Appendix is based upon linear theory, it is only useful for determining the conditions for the initiation of oscillations. Analyses based upon the complete transistor characteristics are required in order to determine distortion, efficiency, etc.; trial and error experimentation is required to optimize these large-signal characteristics.

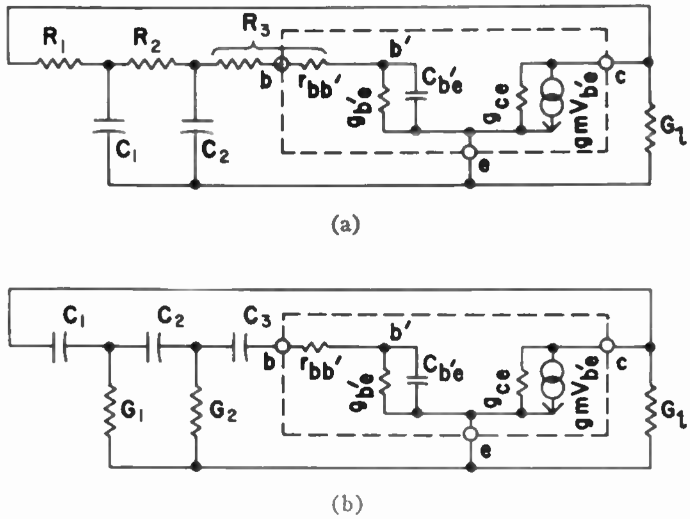


Fig. 1—Transistorized phase-shift oscillator circuit. (a) RC type high-frequency circuit. (b) C/G type low-frequency circuit.

An examination of the circuit of Fig. 1(a) indicates that $r_{bb'}C_{b'e}$ in the transistor serve as part of the RC phase-shift circuit. This type of circuit is well suited for operation at higher frequencies where the shunting effect of $g_{b'e}$ can be neglected. At low frequencies the shunting effect of $g_{b'e}$ becomes important and the circuit of Fig. 1(b) using a C/G phase-shift circuit is required. The $C_{b'e}/g_{b'e}$ time constant gives a rough indication of the crossover frequency, although the exact equations in the Appendix should be used for any specific case, since a large overlap in frequency is possible depending upon the various parameters involved.

Operating Circuits

A variety of circuits have been constructed. Detailed operating characteristics have not been determined, but the data tabulated below will serve to illustrate the performance.

RC Phase-Shift Circuit [*See Fig. 1(a)*]

Transistor: 2N104.
Collector supply voltage, V_{CC}: $-22\frac{1}{2}$ volts.
Collector current, I_C: -1 ma.
$R_1 = R_2 = 2400$ ohms. $R_3 - r_{bb'} = 2200$ ohms.
$C_1 = C_2 = 0.02$ μf.
$G_l = 10^{-4}$ mhos (resistor).
Base bias supplied by means of a shunt arrangement of 0.68 megohms and 0.25 μf connected between collector and R_1.

* Manuscript received by the PGA, December 4, 1956.

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[1] E. L. Ginzton and L. M. Hollingsworth, "Phase-shift oscillator," Proc. IRE, vol. 29, pp. 43–49; February, 1941.

[2] W. Hicks, "Transistor phase-shift oscillator," *Tele-Tech and Electronic Ind.*, vol. 15, pp. 55–56; July, 1956.

[3] L. J. Giacoletto, "Study of *p-n-p* alloy junction transistor from d-c through medium frequencies," *RCA Rev.*, vol. 15, pp. 506–562; December, 1954.

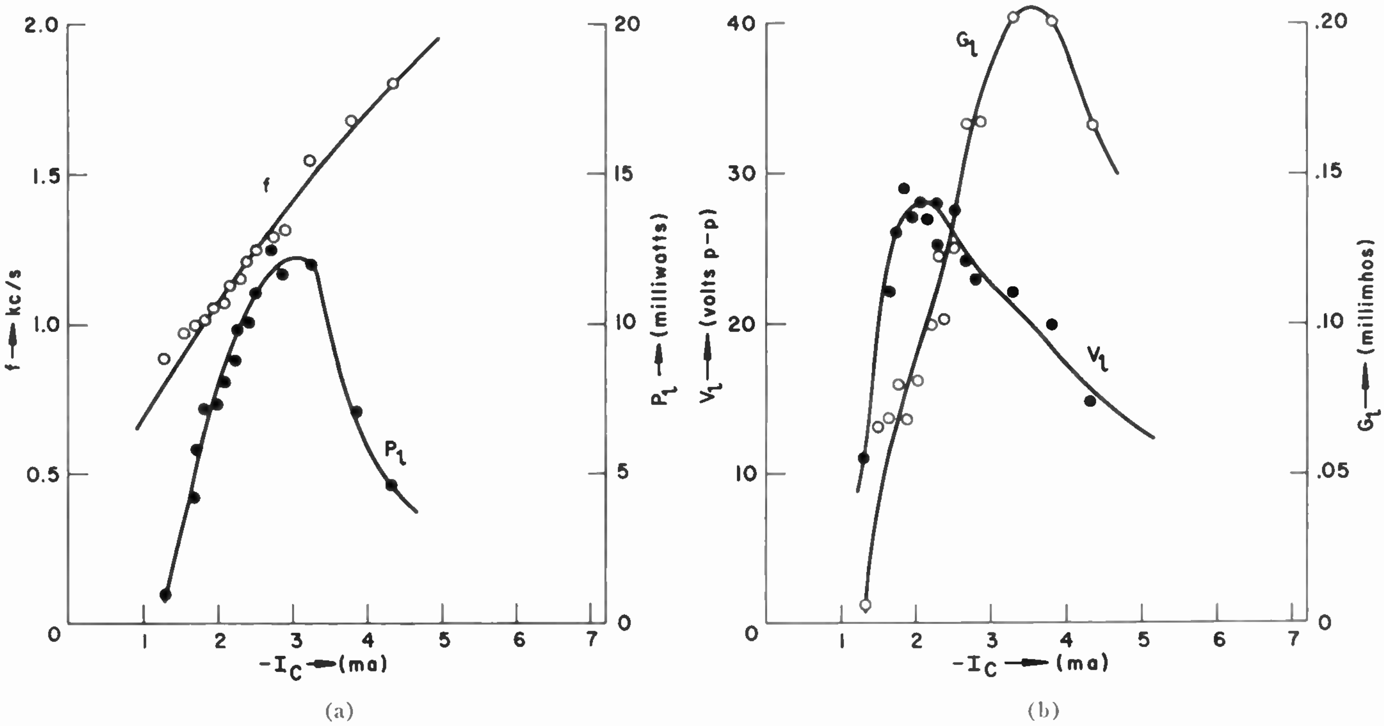

Fig. 2—Performance of C/G phase-shift circuit as a function of collector current. (a) Frequency and maximum power output. (b) Optimum load conductance and output voltage.

Operating frequency, f: 13 kc.
Output across 10 KΩ load resistor = 3 volts p-p.

For comparison purposes, if the following transistor parameters are *assumed*: $C_{b'e}=C_1=C_2$; $g_{b'e}=10^{-3}$ mhos; $r_{bb'}=200$ ohms; and $g_{ce} \ll G_l$, then the second of (2) indicates a computed operating frequency of 11 kc compared with the measured value of 13 kc.

C/G Phase-Shift Circuit [*See Fig. 1(b)*]

Transistor: 2N104.

Transistor characteristics at $V_{CE}=-6$ volts, $I_C=-1$ ma.

$r_{bb'}=480\Omega$, $g_{b'e}=0.38\times10^{-3}$ mhos, $C_{b'e}=5060$ $\mu\mu$f, $g_{ce}=6.6\times10^{-6}$ mhos, $g_m=32\times10^{-3}$ mhos.

$C_1=C_2=C_3=0.02$ μf.

Base bias supplied by means of a 2.5 megohm variable rheostat, R_f, connected between collector and base and a shunt combination of 2.7 K ohms and 10 μf connected in the emitter circuit.

Load arrangement consists of 8.6 henry choke with different values of G_l in shunt.

1) Performance with variable V_{CC}: $G_1=G_2=0.67\times10^{-3}$ mhos. (See Table I).

Using the transistor parameters given above together with $C_1=C_2=C_3=C=0.02$ μf, $G_1=G_2=G=0.67\times10^{-3}$ mhos, and $G_l=0.05\times10^{-3}$ mhos, the second of (6) indicates an operating frequency of 610 cps. However, for the same data the first of (6) indicates a gm requirement of 42.5×10^{-3} mhos. Since the gm available is only 32×10^{-3} mhos, oscillations are not possible as is verified by the last line of the above data.

2) Performance with variable I_C: $G_1=G_2=0.67\times10^{-3}$ mhos, $V_{CC}=-20$ v. Operating characteristics are plotted in Fig. 2. Values for G_l shown are for load conductance when adjusted for maximum power output, P_l. The collector current adjustment was obtained by varying the base-bias rheostat from 0.01 to 0.95 megohms.

3) Performance with variable G_l: $G_2=0.67\times10^{-3}$ mhos. $V_{CC}=-21$ volts (see Table II).

C/G Phase-Shift Circuit [*See Fig. 1(b)*]

Transistor: an experimental p-n-p power unit, $C_1=C_2=C_3=C$. Biasing arrangement consists of a resistor, R_f, connected between collector and base (see Table III).

Comments

The characteristics of transistors permit significant power output from the phase-shift oscillators. As the output power level is increased, distortion increases and a large signal analysis of the operation is required to determine operating conditions for minimum distortion. As the transistor current or load conductance is increased to produce greater output power, a point is reached where the transistor is no longer capable of supplying sufficient voltage amplification to maintain oscillations. Greater output power with good efficiency and waveshape can be obtained by increasing the collector supply voltage; the maximum value is set by collector breakdown or "punch-through."

Since the transistor supplies a portion of one section of the phase-shift circuit, it is possible to vary the oper-

TABLE I

V_{cc} (volts)	R_f (Meg Ω)	I_c (ma)	V_l (volts p-p)	G_l (millimhos optimum)*	P_l (milliwatts)	f_p (cs)
−21.0	0.58	−2.17	28	0.08	8.2	1100
−18.0	0.58	−1.85	24	0.08	6.0	1100
−16.4	0.58	−1.74	22	0.08	5.0	1100
−15.0	0.58	−1.52	20	0.08	4.0	1100
−12.3	0.58	−1.30	15	0.67	1.8	1100
−10.0	0.58	−1.19	13	0.05	1.1	1040
−5.6	0.58	No oscillations				

* Adjusted for maximum power output.

TABLE II

G_l (millimhos)	I_c (ma)	V_l (volts p-p)	G_l (millimhos optimum)*	P_l (milliwatts)	f (cps)
0.5	−2.18	30	0.10	11	1020
0.54	−2.18	30	0.10	11	1042
0.61	−2.18	29	0.10	10.5	1080
0.65	−2.23	29	0.10	10.1	1102
0.71	−2.28	28	0.10	9.7	1140
0.80	−2.39	27	0.10	9.0	1180
0.91	−2.45	22	0.10	6.0	1240
1.00	−2.50	13	0.05	1.0	1280

* Adjusted for maximum power output.

TABLE III

C (μf)	G_1 (millimhos)	G_2 (millimhos)	R_f (K ohms)	V_{cc} (volts)	I_c (ma)	V_l (volts p-p)	P_l (watts)	f (cps)
0.5	17.8	1.0	60	45	125	40	1	(not measured)
0.5	17.8	5.0	25	30	125	60	2.2	(not measured)
0.1	1.8	1.8	37	30	100	58	1.4	1350
0.1	1.0	1.0	42	30	100	57	1.3	930
0.1	3.7	3.7	35	30	100	56	1.3	2000

ating frequency by changing the transistor current and to a lesser degree by changing the transistor voltage. In the circuit of Fig. 1(a), since $C_{b'e}$ increases with increasing transistor current,[2] the operating frequency will normally decrease as the transistor current is increased. In the circuit of Fig. 1(b), since $g_{b'e}$ increases with increasing transistor current,[2] the operating frequency will normally increase as the transistor current is increased. At some intermediate frequency where both $g_{b'e}$ and $C_{b'e}$ contribute significantly to the phase shift, the operating frequency may tend to remain constant with operating current. To a first approximation the transistor parameters are independent of voltage,[2] and the operating frequency will remain constant. The small increase of $C_{b'e}$ and $g_{b'e}$ with decreasing collector voltage will cause the operating frequency to change in the same direction as for increasing current. Some frequency compensation can be obtained by balancing the two effects, voltage and current changes, against one another.[4]

For reasons explained above, frequency modulation of the phase-shift transistorized oscillator can be readily accomplished by modulating the transistor current. The extent of the frequency modulation can be increased by replacing C_1 and C_2 in Fig. 1(a) by separate transistors whose current is also modulated together with the original transistor. Similarly, G_1 and G_2 of Fig. 1(b) can be replaced by transistors. The outputs of the separate transistors can be combined for increased output power or can be used as individual phase-shifted outputs. Finally, R_1, R_2, and R_3 of Fig. 1(a) and C_1, C_2, and C_3 of Fig. 1(b) could also be replaced by transistors which, when suitably current modulated, would produce even greater frequency excursions.

[4] H. C. Lin, "Modulated Transistor Oscillators and Their Applications," in "Transistors I," RCA Labs., pp. 547–560; 1956.

Appendix

Conditions are given here which must be satisfied for oscillations to begin. The development is based on a hybrid-π small-signal equivalent circuit for the transistor in a common-emitter connection as in Fig. 1. Analysis can be carried out by a variety of techniques including matrix methods, simultaneous equations, etc. Probably the most straightforward method is to start at some point in the circuit (such as voltage, $V_{b'e}$ across $g_{b'e}$ and $C_{b'e}$) and work back step-by-step around the loop until you again arrive at the original starting point.

RC Phase-Shift Circuit [See Fig. 1(a)]

The simultaneous equations to be satisfied for the general case are as follows:

$$
\begin{aligned}
g_m + b_{b'e} &- \omega^2[(R_2+R_3)C_1 + R_3C_2]C_{b'e} \\
&- \omega^2(1 + R_3 g_{b'e})R_2C_1C_2 \\
&+ (g_{ce} + G_l)\{1 + (R_1+R_2+R_3)g_{b'e} \\
&- \omega^2[R_1(R_2+R_3)C_1 + R_3(R_1+R_2)C_2]C_{b'e} \\
&- \omega^2(1 + R_3 g_{b'e})R_1R_2C_1C_2\} = 0
\end{aligned}
$$

$$
\begin{aligned}
[1 + (R_2+R_3)g_{b'e}]C_1 &+ (1 + R_3 g_{b'e})C_2 \\
&+ (1 - \omega^2 R_2R_3C_1C_2)C_{b'e} \\
&+ (g_{ce} + G_l)\{R_1[1 + (R_2+R_3)g_{b'e}]C_1 \\
&+ (1 + R_3 g_{b'e})(R_1+R_2)C_2 \\
&+ [R_1 + R_2 + R_3 - \omega^2 R_1R_2R_3C_1C_2]C_{b'e}\} = 0. \qquad (1)
\end{aligned}
$$

There are 11 quantities in these equations. If 9 of these quantities are known, the remaining two can be determined. Usually, $G_l \gg g_{ce}$; in this case, g_{ce} can be neglected leaving 10 quantities. Since the negative terms in the above equations are those containing ω, it follows that for a given set of quantities there will be a lower limit to ω below which oscillations are not possible.

$$
\begin{aligned}
&g_m + g_{b'e} + (G_1+G_2)\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C_3}\right) + \frac{G_1C_{b'e}}{C_2}\left(1 + r_{bb'}G_2 - \frac{G_2 g_{b'e}}{\omega^2 C_{b'e}C_3}\right)\\
&\quad + (g_{ce}+G_l)\left[1 + r_{bb'}g_{b'e} + C_{b'e}\left(\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}\right) + \frac{r_{bb'}C_{b'e}}{C_1}\left(G_1 + G_2 + \frac{G_2C_1}{C_2}\right)\right.\\
&\quad \left. - \frac{G_2 g_{b'e}}{\omega^2 C_3}\left(\frac{1}{C_1}+\frac{1}{C_2}\right) - \frac{G_1 g_{b'e}}{\omega^2 C_1}\left(\frac{1}{C_2}+\frac{1}{C_3}\right) - \frac{G_1G_2}{\omega^2 C_1C_2}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C_3}\right)\right] = 0\\
&1 + r_{bb'}(G_1+G_2) - \frac{g_{b'e}}{\omega^2 C_{b'e}C_3}\left(G_1 + G_2 + \frac{G_1C_3}{C_2}\right) - \frac{G_1G_2}{\omega^2 C_{b'e}C_2}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C_3}\right)\\
&\quad + (g_{ce}+G_l)\left[r_{bb'} - \frac{g_{b'e}}{\omega^2 C_{b'e}}\left(\frac{1}{C_1}+\frac{1}{C_2}+\frac{1}{C_3}\right) - \frac{G_2}{\omega^2 C_{b'e}}\left(\frac{1}{C_1}+\frac{1}{C_2}\right)\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C_3}\right)\right.\\
&\quad \left. - \frac{G_1}{\omega^2 C_{b'e}C_1}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C_3}\right) - \frac{G_1}{\omega^2 C_1C_2}\left(1 + r_{bb'}G_2 - \frac{G_2 g_{b'e}}{\omega^2 C_{b'e}C_3}\right)\right] = 0.
\end{aligned}
\tag{5}
$$

Oftentimes it is convenient to choose $R_1=R_2=R_3=R$ and $C_1=C_2=C_{b'e}=C$. Here, (1) simplifies to

$$
\begin{aligned}
&g_m + g_{b'e} - \omega^2(4 + Rg_{b'e})RC^2\\
&\quad + (g_{c'e}+G_l)[1 + 3Rg_{b'e} - \omega^2(5 + Rg_{b'e})R^2C^2] = 0,\\
&(3 + 3Rg_{b'e} - \omega^2R^2C^2)\\
&\quad + (g_{ce}+G_l)R[6 + 4Rg_{b'e} - \omega^2R^2C^2] = 0.
\end{aligned}
\tag{2}
$$

If 9 of the 11 quantities in these two equations are known, the remaining two can be determined. Since all terms in the above equations containing $1/\omega^2$ are negative, there will be an upper value to ω for a given set of quantities above which oscillations will not be possible. If $C_1=C_2=C_3=C$ and $G_1=G_2=G$, the above equations can be simplified to

$$
\begin{aligned}
&g_m + g_{b'e} + 2G\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C}\right) + \frac{GC_{b'e}}{C}\left(1 + r_{bb'}G - \frac{Gg_{b'e}}{\omega^2 C_{b'e}C}\right)\\
&\quad + (g_{ce}+G_l)\left[1 + r_{bb'}g_{b'e} + \frac{3C_{b'e}}{C} + \frac{3r_{bb'}GC_{b'e}}{C} - \frac{4Gg_{b'e}}{\omega^2C^2} - \frac{G^2}{\omega^2C^2}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C}\right)\right] = 0\\
&1 + 2r_{bb'}G - \frac{3Gg_{b'e}}{\omega^2 C_{b'e}C} - \frac{G^2}{\omega^2 C_{b'e}C}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C}\right)\\
&\quad + (g_{ce}+G_l)\left[r_{bb'} - \frac{3g_{b'e}}{\omega^2 C_{b'e}C} - \frac{3G}{\omega^2 C_{b'e}C}\left(1 + r_{bb'}g_{b'e} + \frac{C_{b'e}}{C}\right) - \frac{G}{\omega^2C^2}\left(1 + r_{bb'}G - \frac{Gg_{b'e}}{\omega^2 C_{b'e}C}\right)\right] = 0.
\end{aligned}
\tag{6}
$$

Oftentimes (as is the case for electron tubes where $g_{b'e}=0$) $Rg_{b'e}<1$ and if further $g_{ce}<G_l$, the last equation above can be solved for the operating-radian frequency giving the usual equation

$$
\omega = \frac{1}{RC}\sqrt{\frac{3 + 6RG_l}{1 + RG}}. \tag{3}
$$

Using this solution and the same approximations, including also that $g_{b'e} \ll g_m$, the first of (2) can be used to solve for the g_m required for oscillations.

$$
g_m = G_l\left[\left(5 + \frac{4}{RG_l}\right)\left(\frac{3 + 6RG_l}{1 + RG_l}\right) - 1\right]. \tag{4}
$$

C/G Phase-Shift Circuit [Fig. 1(b)]

The simultaneous equations to be satisfied for the general case are

The usual equations applicable to the electron tube circuit can be obtained by assuming $r_{bb'}=0$, $g_{b'e}=G$, and $C_{b'e}=0$. For these assumptions together with $g_{ce} \ll G_l$, the last equation above can be used to give the operating radian frequency,

$$
\omega = \frac{G}{C}\sqrt{\frac{G_l}{6G_l + 4G}}. \tag{7}
$$

Using this value of ω and assumptions just enumerated, the first of (6) gives the value of g_m required for oscillations.

$$
g_m = G_l\left[29 + 23\left(\frac{G}{G_l}\right) + 4\left(\frac{G}{G_l}\right)^2\right]. \tag{8}
$$

ACKNOWLEDGMENT

K. R. Keller built the operating circuits mentioned above and obtained the data given.

Some Augmented Cathode Follower Circuits*

J. ROSS MACDONALD†

Summary—**Some direct-coupled augmented cathode-follower circuits are described suitable for power-tube grid drivers, coaxial cable drivers, high-input impedance isolation stages, active electronic filters, and frequency selective circuits. One of the circuits, using only $2\frac{1}{2}$ double triodes, has an input-output voltage transfer ratio of 0.995 or greater with no phase reversal, can supply up to 100 ma of positive current, has a frequency response essentially flat from zero up into the megacycle-second range, can have an input capacitance approaching or equal to zero, exhibits an input resistance greater than 10^{11} ohms for input swings of ± 100 volts or more using ordinary unselected receiving tubes, has an output resistance of 3 ohms, will handle a dynamic range of 630 volts peak-to-peak, and shows less than one part in a million total harmonic distortion at 20-volts rms output and only two parts in 10^5 distortion at 100 volts rms output.**

Introduction

THE CIRCUITS which are discussed in this paper have very high input impedance, low output impedance, wide dynamic range, extremely low distortion, frequency response from zero to the megacycle/second range, and input-output transfer ratios very close to or exceeding unity with no phase reversal. They are suitable for many applications. They may be used as grid drivers of high-power output tubes and will supply 100 ma or more of positive grid current in such service. They are suitable as isolation stages or buffers, particularly where their extremely high-input impedance characteristics are desirable. Their lack of phase reversal together with transfer ratios equal to or exceeding unity and their vanishingly small distortion as well as wide dynamic range makes them particularly useful in active electronic filters and frequency selective amplifiers.

The circuits to be discussed occupy a somewhat intermediate position between ordinary cathode followers and operational amplifiers. An unmodified cathode follower has relatively high input impedance, an unloaded ac input-output transfer ratio generally less than 0.98, an appreciable input-output dc offset, a fairly wide dynamic range, an output impedance of several hundred ohms, and frequency response up to the megacycle/second region. On the other hand, an operational amplifier generally has ac and dc voltage amplification very close to or greater than unity, very low nonlinear distortion, an output impedance of a few ohms or less, and a frequency response ranging from a few to a few hundred kilocycles.

The PACFD

Previously, a parallel augmented cathode-follower driver (PACFD) has been discussed [1] and incorporated as an output-tube grid driver in a high-quality audio amplifier [2]. This circuit, shown in Fig. 1, is more complex than an ordinary cathode follower, has an output resistance of 5.6 ohms, an input-output transfer ratio of 0.986, and can supply up to 200 ma of positive driving current without excessive distortion. Table I presents expressions for the mid-frequency input-output transfer ratio, G, and output resistance, r_0, of several of the circuits with which we shall be concerned. The arithmetic amplifications g_1 and g_2 which appear in some of these formulas are [3]

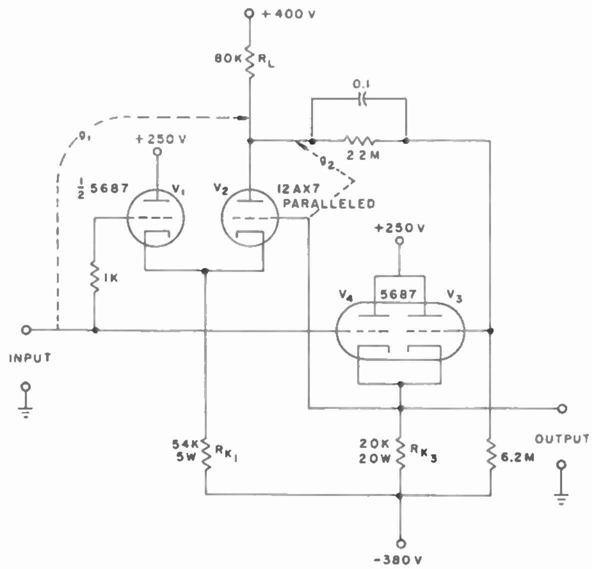

Fig. 1—The parallel augmented cathode-follower driver (pacfd) circuit.

TABLE I

Circuit Designation	Input-Output Transfer Ratio, G	Output Resistance, r_0
CF	$\mu/[\mu'+r_p/R_K]$	$r_p/[\mu'+r_p/R_K]$
PACFD	$\dfrac{\mu_3 g_1+\mu_4 r_{p_3}/r_{p_4}}{[\mu_3 g_2+\mu_3'+\mu_4 r_{p_3}/r_{p_4}+r_{p_3}/R_{K_3}]}$	$\dfrac{r_{p_3}}{[\mu_3 g_2+\mu_3'+\mu_4 r_{p_3}/r_{p_4}+r_{p_3}/R_{K_3}]}$
ACF-1	$\dfrac{\mu_3 g_1}{\mu_3 g_2+\mu_3'+r_{p_3}/R_{K_3}}$	$\dfrac{r_{p_3}}{\mu_3 g_2+\mu_3'+r_{p_3}/R_{K_3}}$
ACF-2	See text	See text

$$g_1 = \left[\frac{\mu_1}{\mu_1' + r_{p_1}/R_{K_1}}\right] \cdot \left[\frac{\mu_2'}{1 + r_{p_2}/R_L + (\mu_2' r_{g_1}/R_L)/(\mu_1' + r_{p_1}/R_{K_1})}\right], \quad (1)$$

$$g_2 = \mu_2[1 + r_{p_2}/R_L + (\mu_2' r_{p_1}/R_L)/(\mu_1' + r_{p_1}/R_{K_1})]^{-1}, \quad (2)$$

* Manuscript received by the PGA, January 9, 1957.
† Texas Instruments, Inc., Dallas 9, Tex.

where $\mu' = \mu + 1$, the subscripts refer to tubes V_1 and V_2, and loading of V_3 on V_2 is neglected—usually a good approximation. Note that the algebraic amplification corresponding to g_2 is negative.

The ACF-1

The PACFD circuit of Fig. 1 is not an optimum design in terms of number of tube halves employed and minimization of dc offset. It can be simplified and improved, as in Fig. 2, by eliminating the direct signal path through V_4 and by either omitting this tube half or connecting it in parallel with V_3. The resulting circuit with the two sections of a 5687 tube in parallel will supply as much positive driving current as the PACFD and can be adjusted (*e.g.*, by varying R_L, R_{K_1}, etc.) so that there is no dc offset for a given quiescent dc input operating level. We shall designate the augmented cathode-follower circuit of Fig. 2, using only a single output tube half for V_3 by the acronym acf-1. As Table I shows, its transfer ratio and output resistance are not much different from that of the PACFD; thus, its output resistance is still of the order of 5 to 6 ohms for the values shown in Fig. 2. The acf-1 is not particularly new; a similar arrangement has been used for some time in connection with frequency-selective amplifiers[1] where the frequency-discriminating circuit is inserted in the feedback line a of Fig. 2.

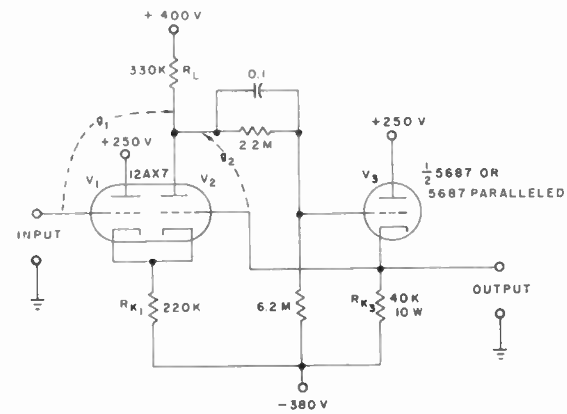

Fig. 2—The augmented cathode-follower number 1 circuit (acf-1).

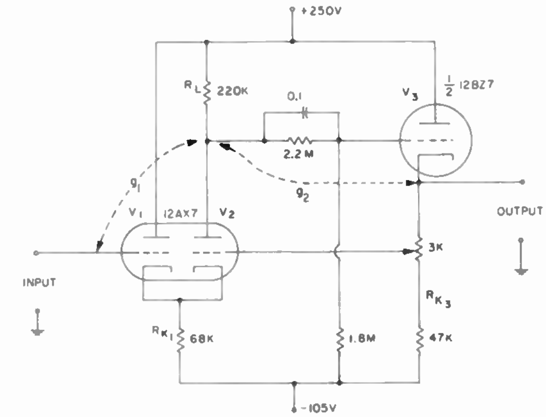

Fig. 3—The acf-1 circuit with voltage amplification control.

The acf-1 has a transfer ratio G slightly less than unity when connected as in Fig. 2. It may be readily modified, however, to have a G of exactly unity or a greater amplification almost as large as g_1. Increased amplification is produced most simply by tapping the feedback line a down on the output cathode resistor R_{K_3} as shown in Fig. 3. The resulting negative feedback reduction is equivalent to a decrease in g_2 with g_1 remaining constant. As Table I shows, such a decrease increases both G and the output resistance r_0. A useful feature of the circuit is that amplifications of the order of 1 to 10 are still achieved with quite low output impedance and with no phase inversion. These features make the acf-1 well suited for the active element in active RC filters [3, 5, 6]. Note, however, that when line a is tapped very far down on R_{K_3}, the voltage divider between V_2 and V_3 may have to be adjusted, and the quiescent dc output level begins to exceed that of the input appreciably. This result is often of no consequence; it need not exist, of course, with ac coupling between V_2 and V_3.

The acf-1 circuit of Fig. 3 was designed specifically for an active filter [6] where it was important that the transfer ratio be essentially independent of supply voltages, that amplifications slightly greater than unity be achieved, and that the output impedance be low. The measured r_0 for minimum G was 4.75 ohms at audio frequencies and 5.6 ohms at 0.5 mc. With a 12AT7 in place of the 12BZ7, r_0 increased to 7.75 ohms at low frequencies. The differential-signal transfer ratio G could be varied with the potentiometer shown between 0.97 and 1.07, and it varied by less than 0.1 per cent when the dc supply voltages were varied by ± 10 per cent.

The ACF-2

The Circuit

The acf circuits of Figs. 2 and 3 may still be considerably improved, although as they stand they are suitable for a wide variety of applications. Although their dynamic range is relatively large, it is limited by the quiescent voltage which may be applied to the tubes without exceeding their ratings and by the magnitude of the negative supply voltage. In addition, the main feedback loop is not effective in reducing nonlinear distortion generated in the input cathode follower V_1, although, being a cathode follower, its internal feedback helps keep such distortion low. Nevertheless, no matter how much the main loop feedback may reduce distortion in the rest of the circuit, the final limiting distortion will be that of V_1.

A first step is further distortion reduction, and further increase of the dynamic range is, therefore, the improvement of these quantities for V_1. First, we may replace R_{K_1} by a constant-current triode,[2] which has

[1] See Valley and Wallman [3], pp. 384–408.

[2] See Valley and Wallman [3], p. 432.

the effect of greatly increasing R_{K_1} and of improving the dynamic range, particularly for large negative signals. Next, to reduce distortion further and to extend the dynamic range for positive signals without exceeding the quiescent 300-volt plate-voltage rating of these tubes, we may employ plate driving of V_1 as shown in Fig. 4. We shall designate this circuit as the acf-2. Note that in order to extend the full amount of feedback down to zero frequency, two avalanche-breakdown silicon diodes are used for part of the divider between V_2 and V_3. In their breakdown region, these diodes have very low ac or differential resistance, but a dc voltage drop nearly independent of current. Each diode in Fig. 4 is selected to have a drop of about 100 volts.

In addition to the constant-current tube V_5, the circuit of Fig. 4 differs from that of Figs. 2 and 3 by the presence of the cathode follower V_4 which drives the plates of V_1 and V_3 with a signal very nearly equal to that at the input. Thus, for example, when the level at the input is increased by 100 volts, the plate and cathode of V_1 both rise by nearly the same amount, as do the grid, cathode, and plate of both V_2 and V_3. It is therefore evident that as long as none of the tubes are saturated or cut off, the effective dynamic operating points of V_1, V_2, and V_3 will be virtually independent of signal. Such independence means, in turn, that these tubes cannot generate appreciable distortion. Since there is a signal path through V_1, V_2, and V_3 from the input to the output, the output will be a virtually undistorted replica of the input. Furthermore, it is clear that although all the dc levels can be arranged so that no quiescent plate voltage exceeds 300 volts, the instantaneous plate voltage of V_1, V_2, and V_3 can increase, because of a positive input signal, until V_4 is saturated. This behavior allows a very wide, undistorted dynamic range to be obtained without exceeding tube ratings. Note that the acf-2 can be constructed using only $2\frac{1}{2}$ double triodes.

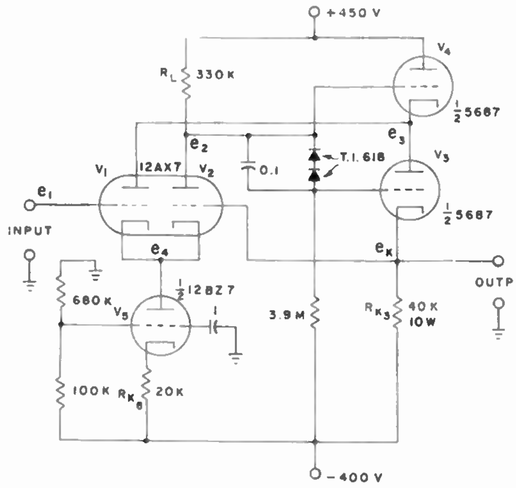

Fig. 4—The acf-2 circuit.

Circuit Operation and Transfer Ratios

Below are given some of the algebraic results of an approximate analysis of the mid-frequency equivalent circuit of Fig. 4 which shows that e_2, e_3, and e_4, and e_K will all follow the input e_1 closely. However, these results can be described in words as follows. The input signal to V_3 will be essentially e_2, since no appreciable loss of signal occurs across the diode-capacitor combination. Further e_3 will be nearly equal to e_2, since V_4 is a cathode follower. Thus, e_K will also be nearly equal to e_2. Since e_1, e_K, and e_3 are all equal to or almost equal to e_1, e_4 will also almost equal e_1. Finally, the negative feedback between V_3 and V_2 will act in such sense that e_2, which equals e_4 plus the amplified difference between e_4 and e_K, will, in turn, be nearly equal to e_1. Because of the cathode follower action of V_3, e_K must be smaller than e_2; e_2 will thus be slightly greater than e_4 and may even exceed e_1. Note that driving the plate of V_3 with the signal e_3, which is almost as large as e_2, makes e_K closer to e_2 than would be the case if the plate of V_3 were not so driven. If there were no loss between e_2 and e_K, it would be easy to show that than e_2 and e_K would also equal e_4.

Although algebraic expressions for the various transfer ratios which may be defined for Fig. 4. have been worked out taking into account the loading of V_3 and V_4 on V_2 and that of V_1 on V_4, the results are exceptionally complicated and will not be given. Even when such loading is neglected, algebraic expressions for the transfer ratios are still long. We shall give those for $G_2 = e_2/e_1$ and $G = e_K/e_1$, because of the light they throw on the distortion of the circuit. The main error occasioned by neglecting loading comes from omitting the effect of the plate current of V_1 on the cathode current of V_4. Since the former will generally be at least ten times smaller than the latter for the circuit of Fig. 4, the error resulting from its neglect will be small.

We have the following approximate results

$$G_2 \cong \frac{\mu_1\mu_2'}{\mu_1'\mu_2 M - \mu_2' N + \mu_1'\left(1 + \frac{r_{p_2}}{R_L}\right) + \mu_2\left(M\frac{r_{p_1}}{R_{K_1}} + \frac{r_{p_1}}{R_L}\right) + \frac{r_{p_1}}{R_L}\left(1 + \frac{R_L}{R_{K_1}} + \frac{r_{p_2}}{R_{K_1}}\right)}, \tag{3}$$

$$G \cong MG_2, \tag{4}$$

where

$$M = \frac{\mu_3'\mu_4 + \mu_3}{\mu_3'\mu_4' + \mu_4' r_{p_3}/R_{K_3} + r_{p_4}/R_{K_3}}. \tag{5}$$

and

$$N = \frac{\mu_3'\mu_4 + \mu_4 r_{p_3}/R_{K_3} - \mu_3 r_{p_4}/R_{K_3}}{\mu_3'\mu_4' + \mu_4' r_{p_3}/R_{K_3} + r_{p_4}/R_{K_3}}. \tag{6}$$

For simplicity, we have written the effective resistance of the constant current tube, $r_{p_5}+(\mu_5+1)R_{K_5}$ as R_{K_1}. For the circuit of Fig. 4, R_{K_1} is of the order of 2 megohms. The quantities M and N are less than unity. Eq. (3) shows that G_2 may exceed unity; even so, G will be slightly less than unity. Note that the parameters of V_4 enter the expression for G only through M and N. Further, for Fig. 4, these quantities may be approximated quite closely by

$$M \simeq \left[\frac{\mu_3}{\mu_3' + r_{p_3}/R_{K_3}}\right]\left[\frac{1 + 2/\mu_4}{1 + 1/\mu_4}\right]$$
$$\cong \left(\frac{\mu_3}{1+\mu_3}\right)\left(1 + \frac{1}{\mu_4}\right); \tag{5a}$$

and

$$N \simeq [\{1 + r_{p_3}/(\mu_3' R_{K_3})\}\{1 + 1/\mu_4\}]^{-1}$$
$$\cong \left(1 - \frac{1}{\mu_4}\right). \tag{6a}$$

Thus, to a good approximation only μ_4 enters M and N and then only as a second order term. Since μ for a tube is relatively independent of operating point, the effect of the parameters of V_4 on G is, therefore, exceedingly small and their changes arising from wide excursions of the input signal e_1 will have a negligible effect on G. We may also note that (3)–(6) indicate that increasing the μ's and g_m's of the various tubes will make G closer to unity.

The transfer ratio G may be made unity or greater by tapping down the feedback line in the manner of Fig. 3. Some of the other transfer ratios of the circuit may alternatively be made unity by inserting a resistor R_s in series with R_L, connecting the voltage divider for the grid of V_2 at the junction of R_s and R_L, and connecting the grid of V_4 to the junction of R_s and the plate of V_2. The method is illustrated in Fig. 9. Table II shows some experimental results for various transfer ratios measured on the acf-2 with $e_1=10$ volts at 10^3 cps. The first row for $R_s=0$ shows that $G_4=e_4/e_1=0.9975$, $G=0.995$, and that G_2 is slightly greater than unity. These ratios are nearly independent of signal magnitude over a wide range and show that a 100-volt increase in the grid voltage e_1 of V_1 would lead to a 99.75-volt cathode rise and a 94.1-volt plate rise; as far as V_1 is concerned the 100-volt input signal looks, therefore, like a 0.25-volt grid bias decrease and a cathode-to-plate voltage decrease of only 5.6 volts.

Table II shows that increasing R_s from zero to 21K causes $G_3=e_3/e_1$ to become unity, while an R_s of 82K makes G_4 unity while causing $G_2'(=e_2'/e_1)$ and G_3 to be appreciably greater than unity. Here, e_2' is the signal at the plate of V_2, while e_2 is that at the juncture of

TABLE II

R_s	$1-G_4$	$1-G_3$	$G_2'-1$	$1-G_2$	$1-G$
0	0.0025	0.059	0.0039	−0.0039	0.0050
21K	0.00178	~0	0.066	0.00068	0.0050
82K	~0	−0.182	0.260	0.0118	0.0050

R_s and R_L. The quantities $|1-G_i|$ given in Table II were determined by direct measurement of the difference between the two corresponding signals (*e.g.*, e_1-e_4) using a Ballantine type 310A ac voltmeter. The places in Table II marked ~0 are not shown as exactly zero, because they were limited to a minimum nonzero value by the exceedingly small nonlinear harmonic distortion components present in the output signal as compared to the input e_1, even when R_s was adjusted to make their fundamentals exactly equal.

Because of the feedback present in the acf-2, loading the output tends to increase G_2 to compensate for such loading. For example, the connection of a 10K resistor across the output increases (G_2-1) to 0.024 and increases $(1-G)$ from 0.0050 and 0.0053 when measurements are made with $e_1=10$ volts. The other transfer ratios do not change appreciably. Note that the acf-2 constructed with a type 5687 tube as in Fig. 4 can supply up to 100 milliamperes of positive current to a load. Its negative current is, of course, limited by the quiescent current in R_{K_3}. Increased positive current handling capacity could be achieved, if desired, by replacing the two sections of the 5687 tube by more powerful tubes such as the 6BX7-GT double triode.

Output Resistance, Frequency Response, and Capacitance Cancellation

Measurements of the acf-2 of Fig. 4 yielded an output resistance of 3 ohms and a frequency response into the megacycle/second range. At 0.5 mc, $1-G$ had only increased from its midfrequency value of 0.005 to 0.026. It may be noted that shielding the high-impedance portions of the circuit and driving the shield with the output signal e_K should allow a frequency response less than 3 db down to 10 mc or beyond. This cancellation of stray capacitance is effective when the input and output signals are very nearly equal as they are here; it may also be applied to cancel input capacitance such as that of a coaxial cable at the input [7]. Perfect cancellation is possible over a limited range of frequencies by adjusting G and/or G_3 to the proper values.

Linear Operation Range

Fig. 5 shows curves of the dc linearity of the acf-2 with either two diodes in the voltage divider chain as shown in Fig. 4 or with only one such diode. Here we plot $\Delta E = E_{in} - E_K$ vs E_{in}, where the capital letters denote dc voltages. ΔE was measured directly with a Fluke Type 801 dc vtvm. The nonzero slopes of the central portions of these curves arise from the deviation of G from unity and are consistent with $1-G=0.005$. It

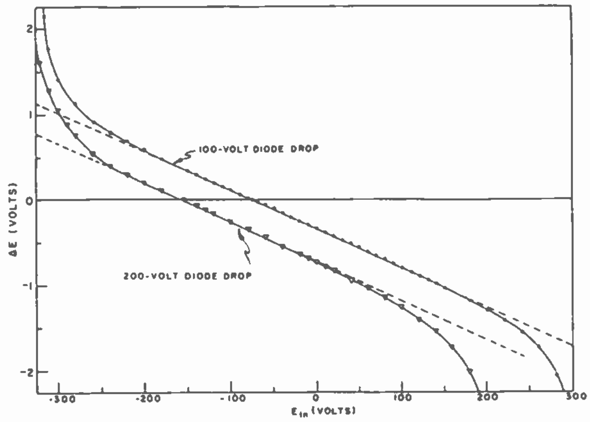

Fig. 5—Linearity curves for the acf-2.

is felt that the few visible deviations from the central straight lines are due to experimental error. For these regions, the device is so linear that deviations from linearity should not show up visually even on this magnified scale.

The deviations from linearity at the ends of the curves arise from the onset of positive or negative clipping. Even the extreme deviations shown at the ends of the curves represent only about one per cent departure from linearity. Note that with one diode, $E_K = E_{in}$ for $E_{in} = -75$ volts, while for both diodes, this point is reached at $E_{in} = -155$ volts. This point may be varied and brought to $E_{in} = 0$, if desired, by changes in the values of R_{K_5} and/or R_L. Under these conditions, there would be no dc offset at $E_{in} = 0$ and offset at any other value of E_{in} would be produced just by the slight departure of G from unity. Experimentally, this condition may be achieved by making $R_{K_5} = 27$K for the one-diode circuit, and 39.5K for the two-diode circuit. These values of R_{K_5} apply when E_{in} is derived from a low impedance source.[3] It is found that a decrease in R_{K_5} somewhat increases the maximum undistorted output swing. With the one-diode circuit, the maximum span is increased from the $+290-(-290) = 580$ volts obtained with $R_{K_5} = 20$K to $+290-(-340) = 630$ volts with $R_{K_5} = 13$K. This swing of 630 volts is further increased to $+350-(-325) = 657$ volts if the positive supply voltage is changed from $+450$ to $+510$ volts. These swings were found to correlate well with measurements of the maximum ac output signals possible with the circuit. For these measurements, the dc bias applied to the input was adjusted to place the quiescent operating point in the center of the linear operating region shown in Fig. 5. By this means, positive and negative clipping of the ac signal could be made to occur together. With $R_{K_5} = 20$K, the one diode circuit did not show clipping until $e_K > 206$ volts rms, equivalent to 583 volts peak-to-peak.

Input Resistance

We have already mentioned how input capacitance at the input of the acf-2 may be cancelled. For many applications, it is desirable to have exceedingly high input resistance as well. When the grid of tube V_1 of Fig. 4 is left floating completely free, the output level is found to be -65 volts. This value is determined by the grid-cathode bias of V_1 at which positive and negative grid currents cancel.[4] Since the output follows the input, we may infer that the input grid also floats at a level of about -65 volts. For infinitesimal input signals applied around -65 volts which do not appreciably alter the grid-cathode bias of V_1, the input resistance is essentially infinite, since the grid current is zero. However, in the present circuit, one must remember that the cathode potential of V_1 follows its grid potential very closely; hence, we may expect that appreciable input signals may be applied before the V_1 grid-cathode bias changes enough to lower the input resistance greatly.

We have measured the input resistance r_i in two different ways. First, it may be found from measurements of the output level E_K when a given dc voltage is applied to the input directly and then through a large resistor. We have used resistors of 10^9, 10^{10}, 10^{11}, and 10^{12} ohms for this purpose. The input resistance may be calculated quite accurately from the inferred drop across the high input resistor. Secondly, it may be calculated from a capacitor discharge time. Here, a small high-quality capacitor is connected from the input to ground, charged to a given positive or negative voltage, and the source of charging voltage disconnected. The voltage level at the output is then observed as the capacitor discharges through r_i. The time, T, for the difference between the output level at the start of discharge and the final quiescent level to decrease to e^{-1} of its original value is then observed and r_i calculated from the result and the known value of capacitance. Mylar, mica, and air capacitors ranging from 10^2 to 10^4 $\mu\mu f$ have been used for such measurements. It has been found that the two methods yield comparable results.

We find for the circuit of Fig. 4 that $r_i \simeq 2 \times 10^9$ ohms when $+100$ volts is applied to the input and 4×10^{10} ohms for -100 volts. For the measurement of dc potentials and charges from a very high impedance source, it is desirable that the input grid float at or nearly at zero potential when left free. By adjusting R_{K_5}, we can readily change the floating point from $E_K = -65$ volts with $R_{K_5} = 20$K to $E_K = 0$ volts with $R_{K_5} = 24$K. Now, when E_{in} is $+100$ or -100 volts r_i is found to be 1.6×10^{10} and 7×10^{10} ohms, respectively. The results for smaller input swings are the same or higher. Comparable results are obtained when a single 100-volt-drop diode is used instead of the two of Fig. 4.

Although an input resistance of 10^{10} ohms or greater

[3] If the input-output dc voltage offset is made zero for a given input level and at the same time, the input-output voltage transfer ratio is made exactly unity, then the dc offset will remain zero for any input with the dynamic range of the circuit. By proper adjustment of R_L, R_{K5}, and the amount of the signal e_2 eventually fed back to the right grid of V_2, it should be possible to achieve the above result. Even when functioning in this ideal fashion with unity gain and zero dc offset, the other advantageous characteristics of the circuit should remain unimpaired.

[4] See Valley and Wallman [3], p. 418.

is high enough for many applications, it was found that even higher values could be obtained in the following manner. The maintenance of an extremely high input impedance over a wide input voltage range depends, as we have noted, on the grid-cathode bias of V_1 remaining at or nearly at its grid-current cancellation value independent of the actual input level. By making $G_4=1$ by means of the series plate resistor R_s mentioned earlier, the ac and differential dc voltage transfer ratio from input to the cathode of V_1 become unity and changes of input level should have no effect whatsoever on the grid-cathode potential of V_1. Such operation should, therefore, give increased input impedance over a wide range.

This idea was investigated and did indeed yield higher r_i values. It was here necessary to set R_s to some value greater than zero, then adjust R_{K_5} to make $E_K=0$ with the input grid floating. The floating point is a fairly sensitive function of V_1 heater current under these conditions, and it was found desirable to regulate the heater voltage for the entire circuit, to render E_K more stable with respect to time with the input floating. Increased stability was also achieved by replacing V_5 with one-half of a 12AX7 instead of the 12BZ7 shown in Fig. 4. Such replacement did not appreciably affect r_i.

With $R_s=50\text{K}$ and $E_K=0$ with input floating, r_i was found to be 2×10^{11} and 1×10^{11} ohms for $E_{in}=-100$ and $+100$ volts, respectively. For $R_s=75\text{K}$, $r_i=5\times10^{11}$ ohms for swings of both -100 and $+100$ volts. The above values of R_s were insufficient to make $G_4=1$, but caused it to be closer to unity than with $R_s=0$. The floating point was quite stable under these conditions. It was found possible to further increase R_s to obtain $r_i\simeq10^{12}$ ohms with fair stability of the floating point. However, when R_s was further increased to make $G_4=1$, the floating point was quite unstable, r_i was difficult to measure, and even unstable negative resistances could be observed. It is worth mentioning that the above results do not depend critically upon the choice of the 12AX7 used for V_1 and V_2; practically identical results were obtained with several different 12AX7's and 12AD7's. These results indicate that an input-output resistance transfer ratio exceeding 10^{11} is obtainable with the acf-2. Since it responds linearly over a wide input signal range, the acf-2 would be very well suited for the input stages of a wide-range, extremely high input resistance ac-dc vacuum tube voltmeter. Using input capacitance cancellation, its effective input capacitance could also be held to a fraction of a micromicrofarad over a relatively wide frequency range.

Nonlinear Distortion

The exceptional linearity of the acf-2 suggested that it might be difficult or impractical to measure the exceedingly small amounts of nonlinear distortion to be expected for signals of medium amplitude. Such distortion has been measured, however, and the results are presented in Figs. 6 and 7. Fig. 6 shows how the total harmonic distortion of an ordinary cathode follower and of the acf-2 depend on output signal with no added load. The cathode follower used one-half of a 5687 tube with 250 volts applied to the plate and a cathode resistor of 40K connected to the -400 volt supply. The frequency of the applied signals was 10^3 cps.

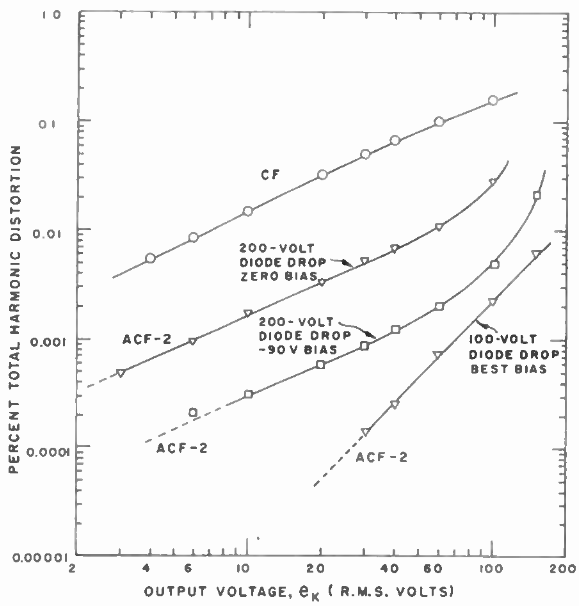

Fig. 6—Per cent total harmonic distortion of an ordinary cathode follower and the acf-2 vs output signal.

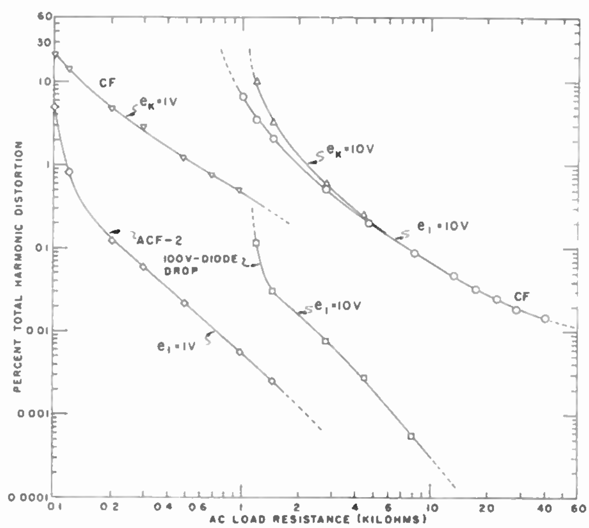

Fig. 7—Per cent total harmonic distortion of an ordinary cathode follower and the acf-2 vs total ac load for fixed signal magnitudes.

These low distortion measurements were made by the bridge method shown in the block diagram of Fig. 8. Since the input and output signals of the acf are essentially in phase at low frequencies and nearly equal, one can subtract a portion of the input equal in magnitude to the fundamental component of the output from the output and have left only the distortion components present in the output. Such subtraction occurs in the general radio bridge transformer of Fig. 8. This transformer has shielded input and output windings, and, by floating and grounding the shields as shown, the

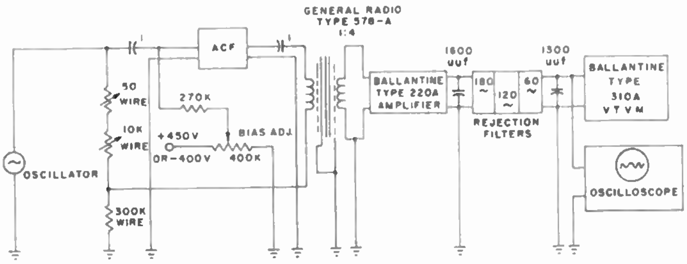

Fig. 8—Bridge circuit for measuring total harmonic distortion of a device having an input-output transfer ratio of unity or less.

large in-phase signal applied to both sides of the primary is essentially eliminated from the secondary. For perfect fundamental cancellation in the output, a capacitative balance (not shown) was often necessary, as well as the resistive balance produced by adjustment of the 50 ohm and 10K resistors. It may be noted that instead of balancing a reduced replica of the input against the output, it is possible to increase G to unity by the method of Fig. 3 and balance the output directly against the input. Both methods give equivalent results, although the circuit feedback is slightly reduced by the $G=1$ adjustment.

Since the lowest distortion components measured were of the order of microvolts, it was necessary to reduce pickup and hum as much as possible. The Ballantine amplifier is battery operated, has amplifications of 10 or 100, and has low-internal noise generation. Residual hum components and high-frequency noise are sufficiently reduced by the shunting capacitors and parallel-T rejection filters shown. Before each use, the circuit of Fig. 8 was carefully calibrated as a function of frequency from the transformer input to the ac voltmeter. For the larger distortion components, the calibration could be checked and distortion measurements carried out by connecting the voltmeter or amplifier terminals directly to the bridge output normally connected to the transformer.

It was usually found using the above measuring apparatus and oscilloscope presentation that the distortion components were predominately second or third harmonics. Since we measure the rms value of the distortion and that of the fundamental plus harmonics, their quotient gives the total harmonic distortion, D_h, directly. Note in the log-log presentation of Fig. 6 that the slope of the cathode follower distortion and that of parts of the next two lower curves is very nearly unity. Such a linear dependence of D_h on fundamental signal magnitude is exactly what one would expect for pure second harmonic distortion, and it was verified by means of the oscilloscope that the distortion was, in fact, of just this character for these curves. On the other hand, the lowest curve is slightly steeper than the square-law dependence to be expected for pure third-order distortion. Here the bias was progressively adjusted from +30 to +90 volts to place the quiescent operating point of the acf-2 at the position on the dynamic transfer characteristic that gave symmetric operation, thus greatly reducing even-order distortion generation. Note that the total harmonic distortion indicated by the lowest curve is less than one part in a million at 20 volts output and only two parts in a hundred thousand at 100 volts output.

Fig. 7 shows the dependence of the distortion of the various circuits on total loading (the parallel combination of added load and the 40K R_{K_3}) for different fixed signal magnitudes. Measurements were carried out at 10^3 cps, and the added load was applied in series with a 15 μf oil-filled capacitor. Since the acf-2 holds e_K essentially equal to e_1 down to the point where negative clipping occurs, the acf-2 curves marked $e_1=1$ or 10 volts are equivalent to similar curves with e_1 replaced by e_K. The comparison between the cathode follower and the acf-2 is fairest when the $e_K=$constant cathode-follower curves are compared with the acf-2 curves, since for constant e_1 loading reduces the output signal of the cathode follower quite appreciably so that negative peak clipping is not approached as rapidly as with $e_K=$constant. The rapid rises in the curves around loads of 1K and 100 ohms come from the approach of negative peak clipping, since the cathode current in both the circuits was approximately 10 ma. It will be noted that in both Figs. 6 and 7 the acf-2 has of the order of a hundred times less distortion than the simple cathode follower.

The ACF-3

Finally, in Fig. 9 we show a simplified acf circuit which we shall designate the acf-3 and which uses only two double triodes. This circuit has an output impedance of several hundred ohms, appreciably higher than that of the other augmented cathode followers. Its main advantages are that its ac and dc input-output transfer ratio can be made exactly unity (with the resulting exceedingly high input resistance already discussed in connection with the acf-2), and it can have essentially zero dc offset over a wide range. It is thus an impedance converter from very high-input impedance to moderately low-output impedance with unity transfer ratio and wide-frequency response extending from dc to tens of megacycles with driven shielding.

By adjusting the series resistor R_s, e_2' may be increased to a value which makes the transfer ratio $G_4=e_4/e_1$ or $G_5=e_5/e_1$ exactly unity. The resistor R_0 is then adjusted to make the dc drop across it exactly equal to the grid-cathode bias of V_1. Then the dc output will also be exactly equal to the input. Since V_5 is a constant-current tube having very high differential resistance, the dc current through R_0 will be held nearly constant and independent of the input signal magnitude or level. Further, since the transfer ratio G_4 is unity for differential changes, the grid-cathode bias of V_1 is also virtually independent of signal level; thus, the dc offset itself remains zero, independent of level over a wide range. It will be noted that R_s cannot be adjusted to

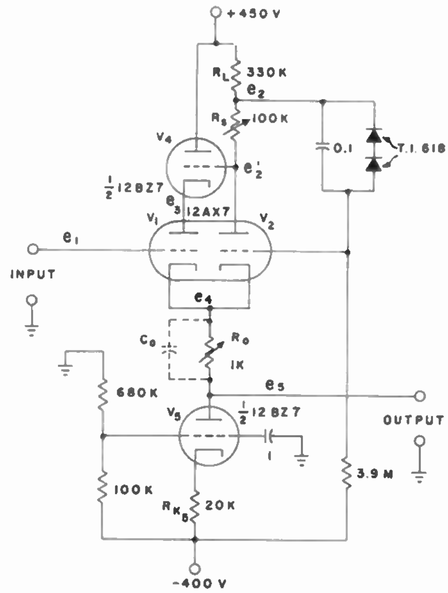

Fig. 9—The acf-3 circuit.

make both G_4 and G_5 simultaneously unity, because of the ac drop across R_0. Since R_0 is so much smaller than the differential resistance of V_5, this drop will be exceedingly small, however, and can be even further reduced if necessary by means of the large capacitor C_0.

For most applications, it will be best to make $G_5=1$ and rely on the fact that G_4 will then be exceedingly close to unity so that the grid-cathode bias of V_1 will still remain nearly constant keeping the dc offset very nearly independent of level. Note that when dc offset is of no consequence, R_0 may be omitted. Finally, it is worth mentioning that for both the acf-2 and the acf-3, the resistances R_L, R_s, and R_{K_5} can all be adjusted to values which will make the output level zero whether the input grid is floating or grounded. Under these conditions, this grid will float at ground potential, there will be no input-output dc offset, and the zero offset will be independent of source resistance.

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Emitter Bypassing in Transistor Circuits*

RAY P. MURRAY†

***Summary*—The emitter bypass capacitor has probably become as commonplace in transistor circuitry as its vacuum tube counterpart, the cathode bypass capacitor. An equivalent circuit analysis of the common-emitter amplifier yields a simple relation that may be used for the determination of the approximate value of the capacitor in terms of the required low-frequency response and the circuit and transistor parameters.**

THE USE OF a resistance in the emitter lead of low-level audio amplifiers is quite common practice. This resistance may have one or more of the following purposes: 1) provide negative feedback at the signal frequency, 2) provide dc negative feedback to stabilize the operating point, and 3) increase the input impedance. If the resistance is unbypassed, it may provide all three of the above functions; however, in many cases, the only desired purpose is to provide operating point stabilization and, in order to prevent loss of gain, a bypass capacitor is employed as shown in Fig. 1.

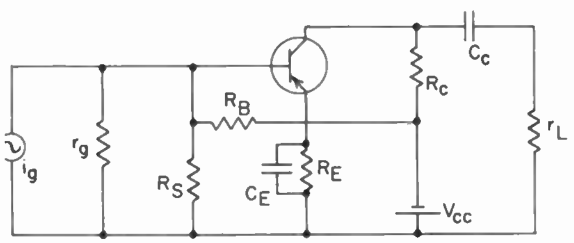

Fig. 1—Stabilized common-emitter amplifier. The signal source is considered to be another transistor amplifier and is simulated in this circuit by a constant current generator.

The proper size of C_E may depend on R_E, R_S, R_B, r_g, the frequency of operation, and the transistor parameters. In the following, we will show the effect of R_E on the current gain of the circuit and will derive a formula for the computation of the reactance of C_E.

r_g of Fig. 1 would normally be the collector coupling resistance of the preceding amplifier, since the output resistance of the transistor is very high. In the equivalent circuit of Fig. 2, r_g, R_S, and R_B have been replaced by their parallel equivalent R_g, and r_L and R_c have been replaced by their parallel equivalent R_L. In addition, C_E has been omitted for the present, to determine the effect of R_E on the current gain. The three terminals in Fig. 2 represent the terminals of the transistor.

Ordinarily the sum of r_e and R_E is very small compared to the sum of the other resistances in the collector circuit, and we may consider the impedance effects of r_e and R_E to be negligible in the collector circuit. This

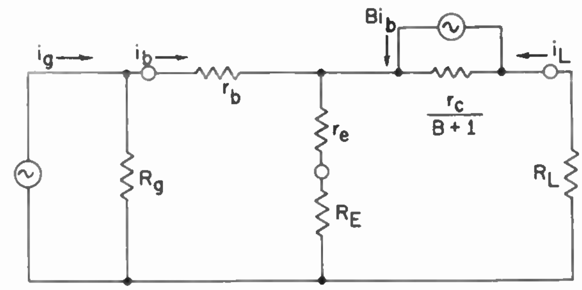

Fig. 2—Equivalent circuit of amplifier in Fig. 1 with C_E omitted.

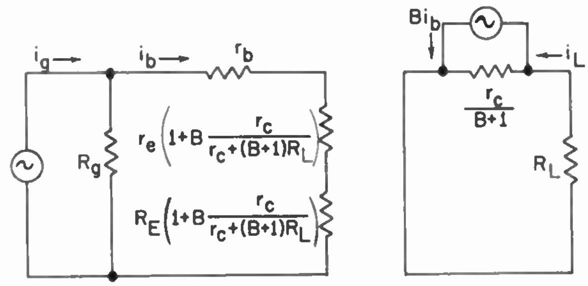

Fig. 3—Simplified equivalent circuit

$$(r_e + R_E) \ll \left(R_L + \frac{r_c}{\beta + 1}\right).$$

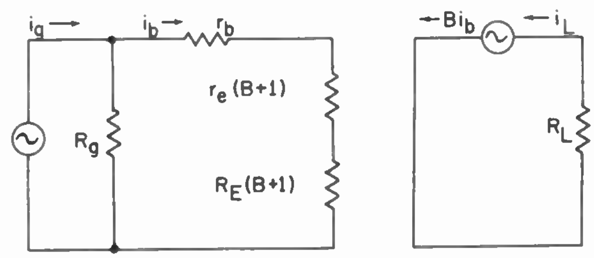

Fig. 4—Additional simplification of equivalent circuit which is valid when $r_c \gg (\beta+1)R_L$.

leads to the simplified equivalent circuit of Fig. 3 in which the effect of the collector current upon the input circuit is taken into account by the modified values of r_e and R_E.

We may simplify this circuit even further as shown in Fig. 4 if r_c is much greater than $(\beta+1)R_L$.

In order to determine the effect of R_E on the current gain, let us define P as the ratio of the current gain with R_E present to the gain with R_E equal to zero. Since the output current, i_L, is directly proportional to i_b, we may determine P by

$$P = \frac{i_b \text{ with } R_E \text{ present}}{i_b \text{ with } R_E = 0}. \tag{1}$$

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From Fig. 4, the values of i_b with and without R_E are

$$i_b = \frac{i_g R_g}{R_g + r_b + (r_e + R_E)(\beta + 1)} \tag{2}$$

$$i_{b|R_E=0|} = \frac{i_g R_g}{R_g + r_b + r_e(\beta + 1)}. \tag{3}$$

Thus, the ratio of the current gains is

$$P = \frac{R_g + r_b + r_e(\beta + 1)}{R_g + r_b + (r_e + R_E)(\beta + 1)}. \tag{4}$$

As an example, let us consider the amplifier of Fig. 5. Substitution of the parameter values in (4) gives a value of $P = 0.0298$. Thus the effect of R_E is to reduce the current gain to only 2.98 per cent of its value when $R_E = 0$.

To prevent the large loss in current gain, R_E may be bypassed with capacitor C_E. In most cases the required reactance of C_E will be much lower than the resistance of R_E, and the approximate equivalent input circuit will be as shown in Fig. 6.

From the circuit of Fig. 6, the absolute value of i_b is

$$|i_b| = \frac{i_g R_g}{\sqrt{[R_g + r_b + r_e(\beta + 1)]^2 + (\beta + 1)^2 X_c^2}} \tag{5}$$

and the ratio of the current gains from (3) and (5) is

$$P = \frac{R_g + r_b + r_e(\beta + 1)}{\sqrt{[R_g + r_b + r_e(\beta + 1)]^2 + (\beta + 1)^2 X_c^2}}. \tag{6}$$

Solving for X_c

$$\boxed{X_c = \left(r_e + \frac{R_g + r_b}{\beta + 1}\right)\sqrt{\frac{1}{P^2} - 1}}. \tag{7}$$

Using the same circuit values as in Fig. 5 and specifying that the current gain at 30 cps be down to 90 per cent of its midfrequency value, (7) gives a value of $X_c = 40.3$ ohms and therefore $C_E = 131$ microfarads.

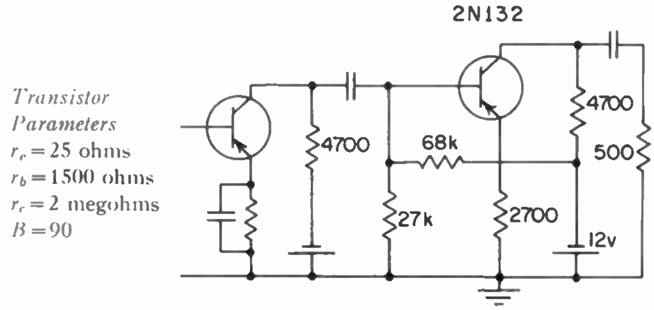

Fig. 5—Amplifier with unbypassed emitter resistor.

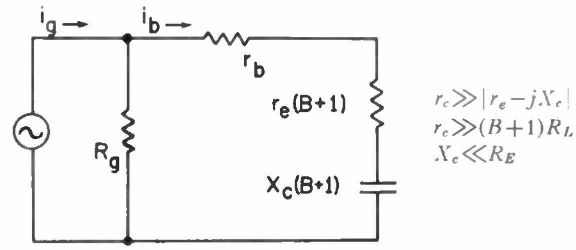

Fig. 6—Approximate equivalent input circuit with C_E present.

Parameter Variations

Of the quantitites in (7), r_e, r_b, and β may be considerably different from the values listed in the transistor specifications. Normally these parameters are specified by the manufacturer as typical or average values at a given temperature such as 27°C and for a given operating point such as $I_e = 1$ milliampere and $V_c = 6$ volts. In some cases, the manufacturer's specification sheet includes information concerning the variation in parameters due to production spread, temperature, and operating point.[1] For a conservative choice of C_E, the values employed in (7) should be based on the minimum values for r_b and r_e and the maximum value of β. However, in some cases, it will be found that the variations in r_b and r_e tend to offset the variation in β.

[1] A listing of many transistor specifications and information on parameter variation is found in R. F. Shea, "Transistor Audio Amplifiers," John Wiley & Sons, Inc., New York, N. Y., ch. 2; 1955.

Effect of Heat on Piezoelectric Properties of a Ceramic Element*

AUSTIN BROUNS†

Summary—**Ceramic piezoelectric elements are affected by elevated temperatures to an extent which bears upon their use in microphones and phonograph pickups. Sensitivity and capacity measurements before and after heat tests illustrate this effect.**

An element, subjected to a constant temperature for one hour in the range 20°C to 100°C, suffers a permanent sensitivity loss which is negligible below 50°C and moderate (3 to 5 db) between 50°C and 80°C. Above 80°C the sensitivity loss becomes increasingly severe until at 105°C the element is essentially depolarized.

Repetition of the heat treatment increases the permanent sensitivity loss, but after 3 or 4 repetitions, the loss does not increase appreciably.

The element capacity is found to increase under all heat tests.

Introduction

Ceramics have come into general use as piezoelectric transducer elements, partly because of their temperature stability. For example—the barium titanate ceramic depolarizes only at 120°C and thus is still operative at a temperature beyond that at which Rochelle Salt decomposes (55°C). Nevertheless, it has been observed that conventional ceramic elements are affected by temperature to an extent which may be important in certain applications. The purpose of this paper is to describe the results of tests conducted to study the effect of temperature on the properties of ceramic elements currently used in phonograph reproducers and microphones.

The piezoelectric element described here consists of two ceramic wafers with an intermediate brass shim.

The element's dimensions are roughly:

length—0.7 inches
width—0.07 inches
thickness—0.03 inches.

The ceramic used is primarily barium titanate ($BaTiO_3$), but also contains small amounts of zirconium and other additives. These additives influence the sensitivity and temperature stability. The data presented here may be considered as indicative of what may be expected in commercial elements rather than representative of barium titanate elements as a class. Some other studies with a more technical treatment have been reported by Mason.[1]

Units covered by this study are used as bender elements in microphones and phonograph pickups. In normal use they may be operated at slightly elevated temperatures over long periods of time, due to climatic conditions, room environment, or adjacent electronic equipment. Higher temperatures may be expected for only short periods of time, such as in a closed automobile on a hot day or near a radiator. A brief résumé, including some environmental effects, was given in a previous paper by Bauer,[2] the present paper being a more complete description of the tests conducted in connection with the development of ceramic pickups.

Description of Tests

The elements were maintained at a specific temperature in a constant temperature oven. After removal from the oven two properties were measured; voltage sensitivity and capacity. Voltage sensitivity is a measure of the output of the ceramic element when it is flexed by a 60 cycle alternating force. In this test, the element is laid on two supporting edges, one near each end of the ceramic piece, and is driven in the middle by a third edge which contacts the element from the top. The driving edge is attached to the coil of a speaker type driver.[3] Capacity was taken between the outside surfaces and was measured on a General Radio Bridge.

Eight groups of 3 elements each were used in the test. Initially the voltage sensitivity and capacity of all units were measured.

Temperature conditions were the following: Temperatures at intervals of $5\frac{1}{2}$°C and ranging from room temperature (27°C) to 110°C were selected. Each group of elements was maintained at one particular temperature for one hour and then removed from the oven. Voltage sensitivity and capacity were measured as soon as the element had reached room temperature after removing it from the oven and also after a recovery period of 1 day and 24 days. For certain elements, the high temperature conditions were repeated at intervals of about one week for several weeks, and appropriate measurements were made.

Results of Tests

Curve a of Fig. 1 shows the average output in decibels of 3 ceramic elements after 1 hour at temperature indicated, followed by $\frac{1}{4}$ hour at room temperature. Curve b of Fig. 1 shows the output loss in decibels of same elements after one week additional recovery period.

* Manuscript received by PGA, February 1, 1957.
† Shure Brothers, Inc., Evanston, Ill.
[1] W. P. Mason, "Aging of the properties of barium titanate and related ferroelectric ceramics," *J. Acous. Soc. Amer.*, vol. 27, p. 73; January, 1955.
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[3] For a more complete description of a similar device, refer to B. B. Bauer, "Piezoelectric ceramics," *Radio and Television News* (*Radio-Electronic Eng. ed.*); August, 1948.

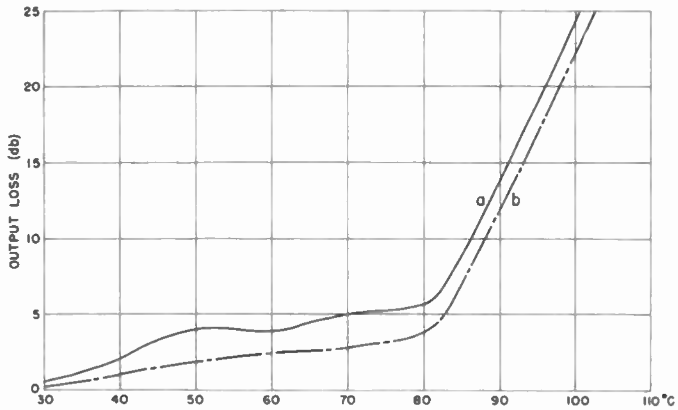

Fig. 1—Reference output = 26 volts/newton.

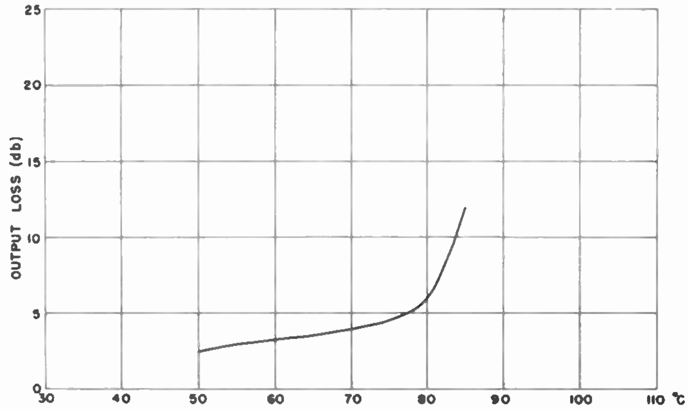

Fig. 2.

The average output loss in decibels of ceramic elements after 5 cycles is shown in Fig. 2. Each cycle consisted of 1 hour at temperature indicated, followed by 1 week recovery period.

Curve a of Fig. 3 shows the gain in capacity, in per cent, of 3 ceramic elements after one hour at temperature indicated, followed by $\frac{1}{4}$ hour at room temperature. The capacity gain in per cent of the same elements after 24 days additional recovery period appears in curve b of Fig. 3.

These ceramic elements showed the greatest recovery within 24 hours after returning them to room temperature. A longer period produced only slight further increase in output. As curve b in Fig. 1 indicates, they never fully recover their voltage sensitivity and above 80° C the permanent loss is rather severe. At 105°C so far as detectable output was concerned, the elements were depolarized.

Repetition of the heat treatment resulted in further permanent loss except that the loss apparently reached a limit after four or five repetitions (Fig. 2).

The heat tests described above produced a permanent increase in capacity of approximately 15 per cent.

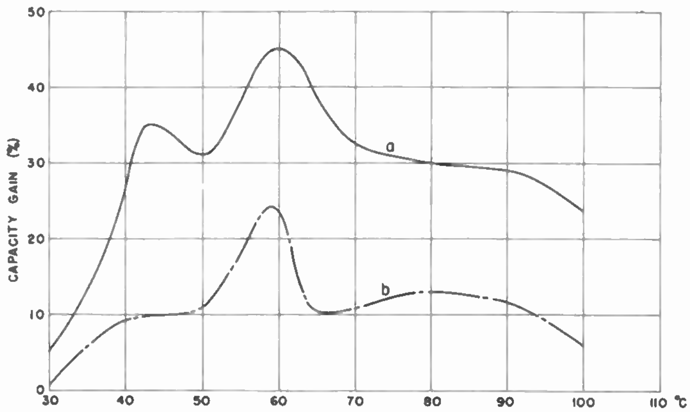

Fig. 3—Reference capacity = 600 $\mu\mu$ farads.

A high capacity is not necessarily undesirable; in a transducer device it moderates loss due to lead capacitance and thus partially compensates for sensitivity loss.

Acknowledgment

The author wishes to express his appreciation to B. B. Bauer for his helpful suggestions in the preparation of this paper.

Correction

Benjamin B. Bauer, author of the paper, "Compensation Networks for Ceramic Phonograph Reproducers," which appeared on pages 8–11 of the January–February, 1957, issue of these Transactions, has requested that the following correction be made to his paper.

In (6), on page 10, the denominator of the second term in brackets should read $C_1C_2R_2/(C_1+C_2)$.

Contributors

Austin Brouns was born in Osakis, Minn., on November 15, 1931. He attended the University of Chicago, Ill., where he graduated in 1952 with the degree in liberal arts.

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Dr. Giacoletto was associated with the Collins Radio Company during the summers of 1937 and 1938 and with the Bell Telephone Laboratories in 1940. From 1941 to 1945, he was on military duty with the Signal Corps Engineering Laboratory concerned with development activities in the fields of radio, navigational, and meteorological direction-finding equipments. He returned to inactive status as a Major in the Signal Corps Reserve, in May, 1946. From June, 1946 to June, 1956, he worked on electronic and semiconductor devices as a research engineer with the RCA Laboratories, Princeton, N. J. Since then, he has been with the Scientific Laboratory, Ford Motor Company, Dearborn, Mich. as manager of the electrical department.

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J. Ross Macdonald (S'44–A'48–SM'54) was born in Savannah, Ga., on February 27, 1923. He received the B.A. degree in physics from Williams College and the S.B. degree in electrical engineering from M.I.T. in 1944. After a semester of teaching at the M.I.T. Army-Navy Technical Radar School, he was commissioned an ensign in the U.S. Naval Reserve in 1944. Upon completing radar courses at the Harvard and M.I.T. radar schools, he served as a technical radar officer. Upon release to inactive duty in 1946, he returned to M.I.T. and received the S.M. degree in electrical engineering, in 1947. During this period, he carried out research on storage tubes for the M.I.T. digital computer project. After a year's further graduate study in the M.I.T. Physics Department, he was awarded a Rhodes Scholarship from Massachusetts for study at Oxford University. Dr. Macdonald received a D. Phil. degree in physics from Oxford, in 1950, for theoretical and experimental work on ferromagnetic phenomena.

J. Ross Macdonald

In 1950, Dr. Macdonald joined the Physics Department of the Armour Research Foundation and there carried out and directed work in theoretical and experimental physics until 1952. He then spent a year's leave of absence at the Argonne National Laboratory of the A.E.C. working on solid-state physics problems. He is presently Director of Solid State Physics Research at Texas Instruments Incorporated, Dallas, Texas. In addition, he is also serving as Clinical Associate Professor of Medical Electronics at Southwestern Medical School of the University of Texas in Dallas.

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❖

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Mr. Murray is a member of Phi Kappa Phi, Sigma Tau, and Eta Kappa Nu.

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