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PGA News

TAPESCRIPT COMMITTEE REPORT

A. B. Jacobson, Chairman of the Tapescript Committee, reports that he now has for distribution six additional tapescripts as follows:

"An Improved Optical Method for Calibrating Test Records," by B. B. Bauer, Shure Bros., Inc., Chicago, Ill. (Presented at the IRE National Convention, March, 1955 and published in the IRE TRANSACTIONS ON AUDIO, p. 137; September, 1955.)

"Electronically Controlled Audio Filters," by L. O. Dolansky, North Eastern University, Boston, Mass. (Presented at IRE National Convention, March, 1955 and published in the PROCEEDINGS OF THE IRE, p. 1580; November, 1955.)

"The Electrostatic Loudspeaker—An Objective Evaluation," by R. J. Larson, Jensen Mfg. Co., Chicago, Ill. (Presented at the NEC, November, 1955 and appearing in the IRE TRANSACTIONS ON AUDIO, this issue.)

"Efficiency and Power Rating of Loudspeakers," by R. W. Benson, Armour Research Foundation, Chicago, Ill. (Presented at the NEC, November, 1955 and published in the IRE TRANSACTIONS ON AUDIO, January, 1956.)

"Energy Distribution in Music," by J. P. Overley, Radio Manufacturing Engineers, Inc., Peoria, Ill. (Presented at the NEC, November 1955.)

"Bells, Electronic Carillons and Chimes," by F. H. Slaymaker, Stromberg-Carlson Co., Rochester, N. Y. (Presented at the NEC, November, 1955 and published in the IRE TRANSACTIONS ON AUDIO, January, 1956.)

The only cost to Audio Chapters for the use of these tapescripts is the return postage to:

Mr. Andrew B. Jacobson
2025 North 40 Street
Phoenix, Arizona.

WITH OTHER ACOUSTICAL AND AUDIO SOCIETIES

The November, 1955, issue of the *Journal of the Acoustical Society of America* contains among its 31 papers six which will interest many PGA members.

"Acoustical Properties of Carpet," by C. M. Harris. The acoustical properties of carpet are of interest in the design of auditoriums and studios, since the absorption which it contributes may be quite significant. However, little data exist on the acoustical properties of carpet and the scanty data available make no distinction among the various types of carpet. This paper describes measurements of the normal absorption coefficient and flow resistance for a wide variety of carpet samples.

"Fundamental Acoustics of Electronic Organ Tone Radiation," by D. W. Martin. The acoustical effects achieved and the techniques used in the radiation of organ tone are quite different from the conventional effects and practices of public-address sound systems. In organ music the sound sources and their images are widely distributed in space. This is especially true in those typical cases where the tone sources are enclosed in an organ chamber, or are installed in a highly reverberant environment. Another distinctive feature of organ tone radiation systems is the relative importance of the octave below 60 cps. This paper contains design principles for electronic organ tone chambers and describes several types of organ tone cabinets now in use.

"Design and Performance of a High-Frequency Electrostatic Speaker," by L. Bobb, R. B. Goldman, and R. W. Roop. An electrostatic speaker has been developed which provides a high quality of high-frequency reproduction not available with electromagnetic tweeters. The diaphragm consists of a thin plastic film bearing an evaporated metallic layer. The membrane is stretched around a semi-cylindrical perforated electrode on which ridges are embossed to provide clearance. The response varies less than ± 2 db in the frequency range between 8 and 16 kc.

"Probe Microphone Analysis and Testing at High Temperatures and High Intensities," by K. W. Goff and D. M. A. Mercer. A probe microphone has been developed suitable for measuring sound fields within such structures as altitude wind tunnels and jet engine test cells. This paper describes the methods of testing and analyzing the instrument. The microphone consists of a $\frac{3}{8}$-inch inside diameter probe tube with a porous metal tip, a condenser microphone, and a spiral resistive termination. It operates at sound-pressure levels up to 170 db with 2 per cent distortion, at ambient pressures down to 0.2 atmosphere, and with probe tip temperatures up to 900° F. The normal incidence response is flat within ± 3 db from 10 to 10,000 cps.

"Absolute Pitch," by A. Bachem. The author, who has had absolute pitch since the time of his earliest musical experiences, at the age of five, has spent much time in studying this musical phenomenon. Amused by the fact that the most extensive work on "absolute pitch" was conducted by psychologists who did not possess this faculty and that it was done on large groups of students, none of whom possessed this faculty either [for exception see O. Abraham, Sammelode d. Internat. Musikges. 3, 1 (1901) and G. Révész, Zur Grundlegung der Tonpsychologie (Leipzig)], he was convinced that a study by a possessor of absolute pitch on himself and on a large number of other possessors of absolute pitch might lead to different and better results.

"Relative Intelligibility of Speech Recorded Simultaneously at the Ear and Mouth," by H. J. Oyer. Monosyllabic words recorded at the lips and left ears of six speakers were fed to the headsets of 24 trained listeners at −12, −15, and −18 S/N ratios. Although the trend for intelligibility scores throughout the test is in the same direction for signals of both origins, decreasing S/N ratio is more destructive to the speech picked up at the lips.

The *Journal* contains its usual extensive references to "Contemporary Papers on Acoustics," by R. T. Beyer; and "Review of Acoustical Patents," by R. W. Young.

The October, 1955 issue of the *Journal of the Society of Motion Picture and Television Engineers* has an excellent study of the "Broadcast Studio Redesign," by L. L. Beranek. A review is made of psychoacoustic and audience-opinion information of recent and older sources from which criteria for studio design are drawn. New criteria are proposed for reverberation time for studios and auditoriums used for speech, music and general purposes. Examples of three types of existing studios that need revision are discussed. Their physical properties and acoustic data are presented and the manner in which they do not meet the criteria is discussed. Various alternatives for adapting these studios to meet the criteria are presented. Emphasis is placed on achieving the desired results with a minimum outlay of money.

B. B. Bauer

FIFTIETH ANNIVERSARY OF COOPERATIVE PLAN AT UNIVERSITY OF CINCINNATI

April 19 marks the start of a two-day conference and a week long exhibition "Panorama of Industrial and Scientific Development 1906–1956" in observance of the fiftieth anniversary of the establishment of the Cincinnati Cooperative Plan of Education at the University of Cincinnati. The theme for this observance is "Education and Industry at Work for Progress." It is interesting to note the coincidental fact that the *Chairman* and *Secretary-Treasurer* of the IRE Professional Group on Audio, and the *Editor* of the IRE Transactions on Audio, are all products of the Cooperative Plan at the University of Cincinnati.

W. E. Koch—*Chairman*, IRE-PGA, EE 1932, MS 1933 (University of Cincinnati), Ph.D. 1934 (University of Berlin)

B. B. Bauer—*Secretary-Treasurer*, IRE-PGA, EE 1937 (University of Cincinnati)

A. B. Bereskin—*Editor*, IRE Transactions on Audio, EE 1935, M.Sc. 1941 (University of Cincinnati).

PGA CHAPTER ACTIVITIES

Cincinnati, Ohio

On January 17, R. J. Larson, of the Jensen Manufacturing Company, presented a paper on "The Electrostatic Loudspeaker—An Objective Evaluation." The talk was concluded with an interesting demonstration of an experimental electrostatic loudspeaker.

Cleveland, Ohio

The order of preference of subject matter, based on 45 answers to the questionnaire mentioned in the Cleveland Chapter activities in the January, 1956 issue of the IRE Transactions on Audio is as follows:

1) Recording and Reproducing Techniques
2) Amplifiers and Pre-Amplifiers
3) Loudspeakers and Enclosures
4) Stereophonic Sound
5) Physics of Music and Hearing.

On December 8, Emory Cook, President of Cook Laboratories, gave a talk on "The Half-Mil Diamond for Microgroove Records." Demonstration of the equipment was arranged by courtesy of the Audio-Craft Company.

Houston, Texas

W. C. Wrye Jr., *Chairman* of the Houston PGA Chapter, reports an increase in the activity of the Houston Chapter. Interesting sessions have been held on "Electrostatic Tweeters," "Architectural Acoustics," "Music and Sound Synthesis," and "Multiple Loudspeaker Measurements." Other equally interesting sessions are being planned for the future.

Thumbnail sketch of Mr. Wrye: Communications, Horns, and Hi-Fi work with U. S. Air Corps and private industry since 1938. Spare time practicing lawyer. Free-lance high fidelity system design and installation. Design and construction of several large exponential horns. Operation of a Hi-Fi Emporium.

Philadelphia, Pa.

M. E. Swift, *Chairman* of the Philadelphia Chapter reports that the 1955–56 meetings have been received enthusiastically by all. For the April 1956 meeting, a program has been arranged, to be held in the Studios of WCAU-TV, Philadelphia, where Albert Preisman will talk in the general field of high-fidelity and present a demonstration which he gave in Constitution Hall, Washington, D.C., in 1955. At that time, Mr. Preisman, assisted by Irwin Stein, set up a 500-watt system with ten Jensen loudspeakers and associated amplifiers, four tape recorders, and various microphone arrangements. The National Symphony orchestra, under the direction of Howard Mitchell, played several numbers—one of which was recorded and then played back. A discussion of this "first," as well as preliminary measurements and technical features, will be the theme of Mr. Preisman's discussion.

Thumbnail sketch of Mr. Preisman: Currently associated with Capitol Radio Engineering Institute as Vice-President in Charge of Engineering, and also engaged as consulting engineer for the Military as well as civilian corporations. Since 1924 he has been active in electronic fields, having been associated with the Wagner Electric Corp., New York Edison Co., RCA Photophone, RCA Institutes, and Federal Telephone and Radio Corp. He is the holder of several patents, has made many contributions to leading technical journals, and is the author of *Graphical Constructions for Vacuum Tube Circuits.* He is a Fellow of the IRE, and a member of the AIEE, the American Society of Engineering, American Technion Society, Armed Forces Communications Association, the Acoustical Society of America, and Sigma Xi.

San Diego, Calif.

According to S. H. Sessions, *Chairman* of the San Diego Chapter, the quarterly-basis planned programs carried on for the past two years have been highly successful. The PGA sponsors approximately half of the regular section meetings. The local Bulletin of the San Diego Section carries a column "What's Doing on the PGA Front." The bulletin reaches about 1,000 members. The following are some of the interesting programs in which this chapter has cooperated during the 1955–56 season:

1) "Variable Damping in Loudspeakers," by Lowman Tibbals of the U. S. Navy Electronics Laboratory. An auditory comparison was made of the same type duplex loudspeaker in four enclosures.
2) "The Use of Negative Feedback in Audio Amplifiers," by Commander W. B. Bernard of the U. S. Navy Electronics Laboratory. This was a regular section meeting sponsored by the PGA.
3) Noise Symposium at the San Diego State College, October 4, 1955. This activity was jointly sponsored by the San Diego Section IRE-PGA, the San Diego Chapter of the Acoustical Society of America, the San Diego Section of the Institute of Aeronautical Sciences, and the San Diego State College. Active on the committee for arrangements of this Symposium were: Dr. Chesney Moe, R. S. Gales, Dr. R. W. Young, S. H. Sessions, P. E. Culbertson, F. X. Byrnes, and E. R. Hinz. Papers were presented on the following topics: "Noise Sources and Propagation," by Dr. Isadore Rudnick; "Noise Measurement Techniques," by Dr. R. W. Leonard; "Control of Noise at its Source," by R. O. Boe; "The Effect of Noise on the Development of High Speed Jet Aircraft," by G. L. Getline; "Effect of Noise on Man," by Dr. J. C. Webster; "Criteria for Annoyance," by Dr. L. L. Beranek, and "Personal Protection," by R. S. Gales.

San Francisco, Calif.

D. L. Broderick, *Secretary* of the San Francisco Chapter reports that the 1955–56 sessions have been very successful and that the attendance has shown a marked increase. Following is a run-down of the Chapter's complete program for the year.

June 13, 1955: Subject: "Distortion," a tutorial lecture on both harmonic and intermodulation types of distortion, by Dr. B. M. Oliver, Director of Research, Hewlett-Packard Company.

September 12, 1955: Subject: "Audibility of Distortion," a tapescript program with demonstrations by R. A. Long, local PGA Chairman.

November 15, 1955: Subject: "Human Perception of Wow and Flutter," including demonstrations, by W. T. Selsted, Director of Research, Ampex Corporation.

February 21, 1956: Subject; "Application of Acoustical Engineering Principles to Home Music Rooms," by W. B. Snow, Consultant in Acoustics, Santa Monica, Calif.

The Electrostatic Loudspeaker—An Objective Evaluation*

R. J. LARSON†

Summary—**The electrostatic loudspeaker, following development of new materials and methods, is now practical for high-frequency use in multichannel loudspeaker systems. Theoretical and mechanical design considerations illustrate the limitations at this stage of the art, including inherently high distortion at high output levels, and inability to withstand overloads. Principal advantages are low cost and efficient reproduction at extremely high frequencies.**

Introduction

THE ELECTROSTATIC tweeter has recently appeared on the market after an absence of many years. This is the classic condenser loudspeaker which dates from as far back as 1926, when Colin Kyle filed his patent.[1,2] Other early work by Georg Seibt,[3] Adolph Thomas,[4] and Hans Vogt,[5] showed a good grasp of the principles involved and some ingenious methods of reducing the theory to practice. Unfortunately, these pioneers suffered from the lack of suitable materials for their loudspeakers. The age of plastics had not arrived. It was inevitable that the mechanical force between two charged electrodes be used as a means of creating mechanical vibrations and sound. The two other major methods of transduction, the piezoelectric effect and the magnetic effect, were also being investigated and because of certain advantages, the electromagnetic loudspeaker emerged as the most practical design. Very little commercial use was made of the electrostatic loudspeaker, and the principle was confined in its application to microphones.

With the increased interest in quality reproduction of sound brought about by frequency modulation broadcasting, the shortcomings of the single cone direct radiator type of loudspeaker became evident. As early as 1940, the two unit coaxial type of loudspeaker, pioneered by Jensen, became available to the home music listener. The coaxial loudspeaker allowed much better treble response, because the high-frequency signals were not forced to move the massive cone needed for good bass response.

For reproduction of extremely high frequencies, even the two channel loudspeaker was not adequate; so in 1950, Jensen marketed the first unitary three channel loudspeaker, the Triaxial, which is a great improvement over coaxial speakers.[6] Extension of the high-frequency limit is obtained by the use of a very light moving system and small spacing between the diaphragm and its sound chamber. Naturally, this close tolerance construction is reflected in the cost of the loudspeaker. At the higher audio frequencies, it becomes difficult to force a diaphragm to vibrate as a unit; the wave length of a 20,000 cycle signal is only about three-quarters inch in air so that multiple modes of vibration are possible in all but the smallest of diaphragms, when driven from a single voice coil. Horn loading of these small radiating surfaces is essential if adequate acoustic output is to be obtained, and the resulting tweeter is expensive. In this high-frequency range, the electrostatic principle becomes attractive, since mechanical force on a relatively large diaphragm may be applied over its entire surface instead of from a small voice coil. By use of the electrostatic principle, the diaphragm, no matter what its area, is forced to move as a unit. Thus, it would seem that the electrostatic loudspeaker is the perfect transducer. However, it will be shown that there are certain disadvantages to its use in its present state of development.

The electrostatic loudspeaker consists of two closely spaced electrodes, one of which is in the form of a diaphragm, which is allowed to move because of the mechanical forces applied to it by means of electrostatic charges. The theory of operation of the electrostatic loudspeaker may be obtained from various textbooks on applied acoustics, and is presented in a condensed form in the Appendix. The theory shows that the acoustic power output of the speaker is independent of frequency if it is driven from a constant current source, or will have a rising output characteristic when driven by a constant voltage source. The second harmonic distortion is shown to be a function of the ratio of peak ac signal, E_1, to dc polarizing potential, E_2; the expression for distortion being,

$$\%\ \text{2nd harmonic} = 50\,\frac{E_1}{E_2}. \tag{1}$$

Thus, to keep the second harmonic distortion below 10 per cent, the dc polarizing potential must be at least five times the peak value of the ac signal.

* Received by the PGA, November 10, 1955. Presented at the National Electronics Conference, November 3, 1955.

† Jensen Manufacturing Co., Chicago, Ill.

[1] U. S. Patent No. 1,644,387. Filed October 4, 1926.

[2] V. F. Greaves, F. W. Kranz, and W. D. Crozier, "The kyle condenser loudspeaker," Proc. IRE, vol. 17, p. 1142; July, 1929.

[3] U. S. Patent No. 1,753,137. Filed August 8, 1927.

[4] U. S. Patent No. 1,839,130. Filed March 7, 1929.

[5] U. S. Patent No. 1,948,637. Filed October 7, 1930.

[6] D. J. Plach and P. B. Williams, "A new loudspeaker of advanced design," *Audio Engineering*, vol. 34, p. 22; October, 1950.

Description

Electrostatic loudspeakers currently available make use of some of the newer plastic films as the dielectric; a typical arrangement, shown in Fig. 1(a), consists of a film of DuPont Mylar plastic which has been coated with a layer of aluminum or gold by a vacuum metallizing process. This film is placed over a back electrode which is made of punched metal. The holes in the back electrode allow the film to vibrate without the large stiffness which would be present if the back plate were solid. The holes reduce the electrostatic force on the diaphragm to less than that presented by a solid back electrode, so that a satisfactory compromise in the ratio of hole to metal area is sought. Fig. 1(b) shows an enlarged section through the loudspeaker drawn approximately to scale. Most of the acoustical radiation can take place only in the area over the punched holes, whereas the electrostatic forces driving the diaphragm are applied in the areas over the metal backing. Consequently, the greatest amount of radiation probably occurs around the circumference of each of the holes.

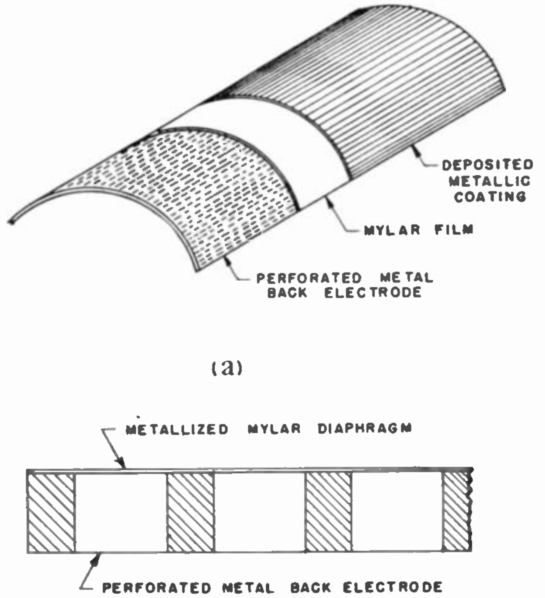


Fig. 1—(a) Electrostatic tweeter cross section. The thickness of the layers has been exaggerated to show the construction. (b) Electrostatic tweeter enlarged cross section. In this view, the parts are drawn to scale to show the relative thickness of the electrodes.

The thickness of the Mylar film determines the voltage breakdown point of the loudspeaker, and, to a certain extent, the efficiency. The dielectric strength of Mylar is 4,000 volts per mil in a 2-mil thickness. Most electrostatic loudspeakers use half or quarter-mil film, giving a breakdown voltage of 2,000 and 1,000 volts, respectively. Actually, breakdown occurs at lower voltages because of small inclusions or holes in the film. When the polarizing potential is first applied, these faults spark and produce small holes in the deposited coating. A permanent short circuit is not formed and once these faults have been cleared, there is no further breakdown in the film unless high audio voltages are applied.

A convenient size for an electrostatic loudspeaker which is designed to be used as a high-frequency unit is about 3 by 6 inches. The electrical capacitance of units of this size turns out to be about 5,000 micromicrofarads, or a reactance of about 30,000 ohms at 1 kc and 3,000 ohms at 10 kc.

The polar distribution characteristics of the electrostatic loudspeaker can be improved considerably by forming the radiating surface as a section of a cylinder. Fig. 2 shows the horizontal distribution pattern of a curved 3- by 5-inch unit at 7 kc and 10 kc. These are free field patterns and it is to be expected that the patterns will be considerably modified in the usual listening environment.

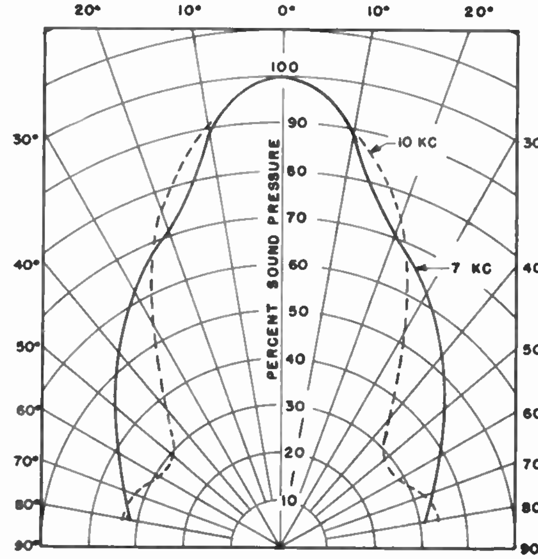


Fig. 2—Horizontal distribution of the sound pressure of an electrostatic tweeter under free field conditions.

It has been shown that the distortion to be expected in an electrostatic loudspeaker is a function of the ratio of ac signal to the dc polarizing potential. In usual amplifier circuits, voltages on the order of 250 are available; this limits the ac peak signal to about 50 volts for 10 per cent second harmonic distortion. This distortion may appear to be excessive; however, when the electrostatic loudspeaker is used as a tweeter with a crossover frequency of 7 kc, the second harmonic of 7 kc occurs at 14 kc, which is near the upper limit of hearing. The celebrated Fletcher-Munson curves show the ear to be about 5 db less sensitive at 14 kc than at 7 kc at usual listening levels so that effective distortion is only about 6 per cent. This effective percentage of distortion will decrease even further at higher input frequencies. In order to take advantage of the ear's reduced sensitivity to high frequencies, it is necessary to employ a network which cuts off the electrical signal to the tweeter at all frequencies below about 7 kc.

A further advantage of using a crossover network lies in the energy distribution curve of typical orchestral music.[7] The output of an orchestra in the region above 7 kc is about 8 db below the average level in what would be the woofer range of a two way system. Thus, if the tweeter sine wave input is limited by distortion considerations to be 50 volts, the actual system input on music may be as much as 8 db higher, or 125 volts. It is common practice in two and three way dynamic speaker systems to rate the entire system at, say, 30 watts music and speech input, whereas the tweeter by itself may be capable of withstanding only 5 watts of sine wave power.

An unusual effect is obtained when the input signal is allowed to overload the electrostatic loudspeaker. A linearity curve (Fig. 3) shows that the output is proportional to the input up to a point where the peak of the ac input signal approaches the dc polarizing potential.

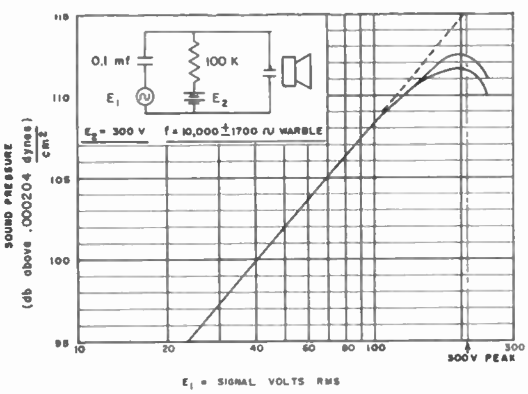


Fig. 3—Linearity curve for an electrostatic tweeter. The double value near the top of the curve shows that the output decreases as a function of time in this region.

When this happens, the sound pressure falls off abruptly and, furthermore, decreases with time. If this overload condition is allowed to persist, a semipermanent decrease in sensitivity of the loudspeaker is produced. Full output may be restored by reversing the connections to the loudspeaker, but this is inconvenient in most installations. The condition of reduced sensitivity will continue for several days, but if left unenergized the loudspeaker will regain its sensitivity eventually.

The Mylar used as a dielectric in these loudspeakers is capable of retaining a charge for long periods of time; like many dielectrics, it is capable of dielectric absorption, and after being charged will be found to have a reversed charge after it has been shorted out for a short period of time. It is thought that during the peak signal periods in which the polarity on the loudspeaker is driven in the opposite direction to the polarizing potential, that this reversal takes place in the dielectric and effectively reduces the dc polarizing potential. With a constant ac signal applied, the loudspeaker becomes overloaded more and more easily as the dc level drops lower and lower; this accounts for the rapid decrease in output as a function of time. It is interesting to note that after extreme overload conditions, the sensitivity becomes greater than with newly constructed units when the connections are reversed. Here, the absorbed charge aids the polarizing potential. The condition is short-lived, however.

When making frequency response measurements of complete pieces of equipment employing the electrostatic tweeter, it is necessary to restrict the output level to the point below which the tweeter is not overloaded. Sine wave voltages will overload the tweeter at levels which are safe for music because of the 8 db safety factor mentioned above.

Fig. 4 shows the sound pressure of a typical unit when measured on its axis at a distance of two feet. A polarizing potential of 250 volts and a constant signal voltage of 50 volts rms was used. The microphone used was a Western Electric 640AA equalized for normal incidence of sound waves. The upper frequency limit of the electrostatic loudspeaker is approached when the mass reactance of the diaphragm and its air load equals the stiffness of the moving system.

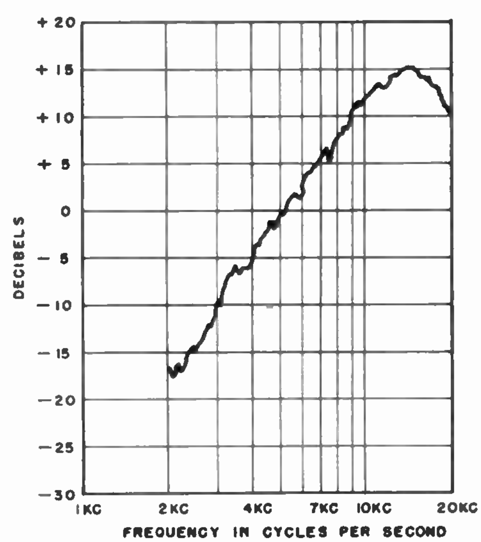


Fig. 4—Axial sound pressure response of an electrostatic tweeter.

By considering one hole of the punched metal backing electrode as an elemental radiator and neglecting the mutual impedance effects of neighboring radiators, the frequency of the first vibratory mode works out to be about 35 kc for currently available tweeters. Laboratory measurements have confirmed this figure, although frequency response measurements become difficult when the wavelength of the sound approaches the dimensions of the microphone to speaker spacing. It is felt that the results of these experiments are too inconclusive to be presented with this paper.

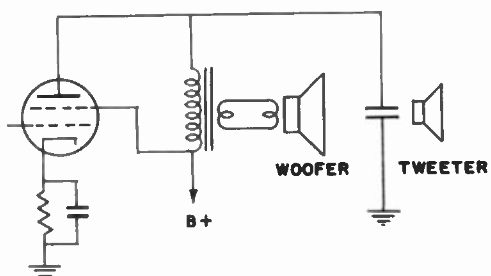
Application

The foregoing has shown that the electrostatic loudspeaker is most efficiently employed as a high-frequency

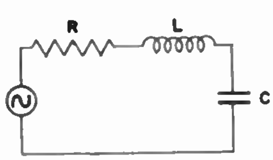
[7] H. Fletcher, "Some physical characteristics of speech and music," *Bell System Tech. Jour.*, vol. 10, p. 349; July, 1931.

unit, or tweeter. A crossover frequency of at least 7 kc is needed in order to match the output level of most cone type woofers in a two way system. The electrostatic tweeter should find its greatest application in equipment having a good electrical output in the region above 7 kc. This category includes high quality record players, tape recorders, frequency modulation radio receivers, and television sets.

In order to obtain the largest possible voltage swing for the tweeter, the usual arrangement is to connect it to the plate of the output tube. The available voltage at the tube plate depends on the load impedance used and the plate supply voltage. For the simplest connection, the tweeter may be connected directly from plate to ground as shown in Fig. 5(a). This connection provides the necessary dc polarizing potential with a superimposed ac signal from the output tube. The impedance seen by the tweeter looking back into the amplifier is the transformer primary impedance in parallel with the output tube plate resistance. With a pentode and no negative feedback, the plate resistance is usually so high that the effective source impedance is only that provided by the transformer and loudspeaker load.



(a)



(b)

Fig. 5—(a) Electrostatic tweeter connected directly to the plate of an output tube. (b) Equivalent circuit.

The equivalent circuit is shown in Fig. 5(b). A rudimentary form of crossover network is formed by this connection. By resonating L and C at the crossover frequency, the acoustic output of the tweeter may be made fairly flat in the region above crossover.

In applications in which the listener is likely to be fairly close to the loudspeaker system, it may be found that low-frequency signals tend to make the tweeter emit audible rattling and buzzing sounds or that the distortion is too high for critical applications. The response of the tweeter is down at least 30 db in the low-frequency region, but large low-frequency signals cause the stretched diaphragm to flutter. For this reason, and for reducing distortion, it is desirable to limit the low-frequency energy presented to the tweeter. A simple network for accomplishing this is shown in Fig. 6. The inductance, L, is chosen to resonate with the capacitance, C, of the electrostatic tweeter at crossover. Inductances in the range of 100 millihenries are typical. Air core inductances are recommended because of their lower losses at the higher audio frequencies and their constant inductance, regardless of signal level applied. A path for dc polarizing potential is provided by R_1.

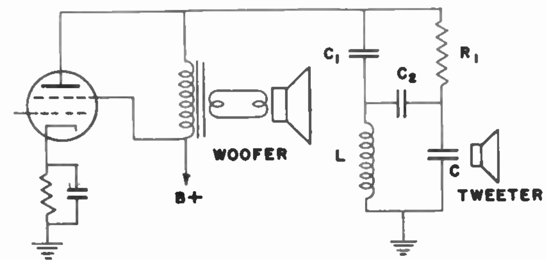


Fig. 6—Electrostatic tweeter connected to an output tube through a crossover network.

It is not necessary to provide the low-frequency loudspeaker with a rolloff network since the high-frequency response may be shaped by the design of the loudspeaker itself. The output transformer may be designed to favor good bass transmission, since it is not used to feed the tweeter. To obtain the smooth transition between the woofer and tweeter, the two units should be designed to complement each other, both as to level balance and for obtaining the required rolloff in the low-frequency unit at the crossover frequency. Fig. 7 shows a composite curve of an 8-inch low-frequency speaker and an electrostatic tweeter.

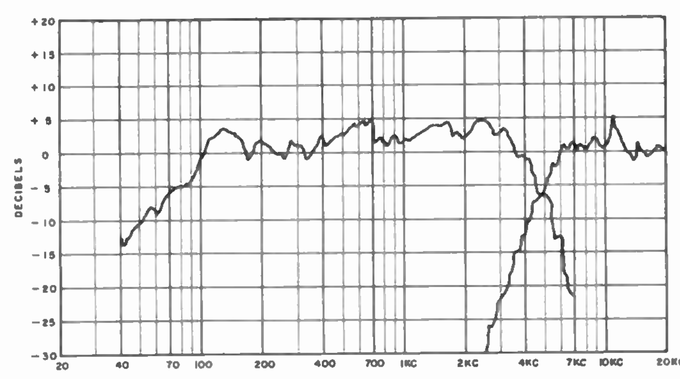


Fig. 7—Acoustical crossover curve showing the output sound pressure of an 8-inch dynamic woofer and an electrostatic tweeter with crossover network as shown in Fig. 6.

Conclusion

In evaluating the electrostatic loudspeaker, the following summary of advantages and disadvantages may be helpful. Advantages are: low cost; smooth, peak-free acoustic output; good polar distribution; and low transient distortion. Disadvantages are: high second harmonic distortion; necessity in most applications for complex crossover network; electrical shock hazard unless suitably protected; high impedance, not convenient to locate remotely; requires source for dc polarization; and sensitivity decrease upon overload.

The electrostatic loudspeaker shows promise for further development. In its current form it is useful only at comparatively high frequencies, with the possibility that program sources presently available will not provide enough high-frequency energy to make its inclusion worthwhile in many pieces of equipment. Future improvements in materials could change this situation, however.

Appendix

Derivation of the frequency response characteristic of an electrostatic loudspeaker:

From energy considerations, the mechanical force required to move a charge of voltage, E, a small distance, dy, is

$$f = \frac{1}{2} E^2 \frac{dC}{dy} \tag{2}$$

where dC is the change in capacitance between the plates. For a parallel plate capacitor,

$$C = \frac{KA}{4\pi y} \quad \text{and} \quad \frac{dC}{dy} = -\frac{KA}{4\pi y^2} \tag{3}$$

where A is the area of one of the plates and y is the spacing between them. K is a constant, depending on the system of measuring units and the dielectric material used. Thus

$$f = -\frac{KAE^2}{8\pi y^2}. \tag{4}$$

For proper operation of the electrostatic loudspeaker, a steady dc potential, E_2, must be superimposed on the peak ac signal voltage, E_1, applied to the plates:

$$E = E_1 \sin \omega t + E_2. \tag{5}$$

The mechanical impedance of the moving system is stiffness controlled in the audio frequency range, and therefore

$$f = kx \tag{6}$$

where k is the spring constant and x is the displacement from the position of rest, which is assumed to be small when compared with the electrode spacing, y. Combining (4), (5), and (6), and making use of a trigonometric identity gives the displacement as a function of signal voltage:

$$x = \frac{KA}{8\pi k y^2}\left(2E_1E_2 \sin \omega t - \frac{E_1^2}{2} \cos 2\omega t + \frac{E_1^2}{2} + E_2^2\right). \tag{7}$$

The acoustic sound pressure is a function of the diaphragm velocity, and

$$v = \frac{dx}{dt} = \frac{KA}{8\pi k y^2} (2E_1E_2\, \omega \cos \omega t + E_1^2\omega \sin 2\omega t). \tag{8}$$

The second term of this expression is a component of twice the frequency of the input signal and will produce second harmonic distortion in the output sound waveform. However, if $E_2 \gg E_1$,

$$v = \frac{KAE_1E_2\omega \cos \omega t}{4\pi k y^2}. \tag{9}$$

For frequencies above approximately 1,000 cycles, the radiation resistance presented to the diaphragm by its air load is practically constant, so that the acoustic power output of the loudspeaker is

$$P = r_a v^2 \tag{10}$$

$$P \cong r_a(E_1E_2\omega \cos \omega t)^2. \tag{11}$$

So the power output is proportional to the input frequency squared for a constant voltage source. This output will rise 6 db per octave from about 1,000 cycles upward. The measured sound pressure at some point will depend on the directional characteristics of the loudspeaker, and may show as great as 12 db per octave upward slope on the axis. The electrical impedance of the electrostatic speaker is almost entirely capacitive. If a constant current source is used,

$$E_1 = IZ = \left|\frac{I}{\omega C}\right|, \tag{12}$$

so that

$$P \cong r_a \left(\frac{IE^2}{C} \cos \omega t\right)^2, \tag{13}$$

the power output is independent of frequency.

A 3,000-Watt Audio Power Amplifier*

A. B. BERESKIN†

***Summary*—A 3,000-watt audio power amplifier has been developed using the Bereskin power amplifier circuit described at the 1954 IRE National Convention and in the March–April 1954 issue of the IRE Transactions on Audio. Solutions were found for some interesting problems that arose in this connection. A unit capable of delivering more than 3,000 watts with less than 2 per cent distortion over a 400–6,000 cycle frequency range was developed. The design procedure and test data on the final unit will be discussed.**

Introduction

A NUMBER of successful design variations of the Bereskin Power Amplifier[1] have been produced since the circuit was conceived. These have included various power and frequency ranges, in particular a 3,000 watt unit which presented a number of problems leading to what are believed to be interesting solutions. It is the purpose of this paper to discuss these problems and their solutions and to present experimental test data on the completed unit.

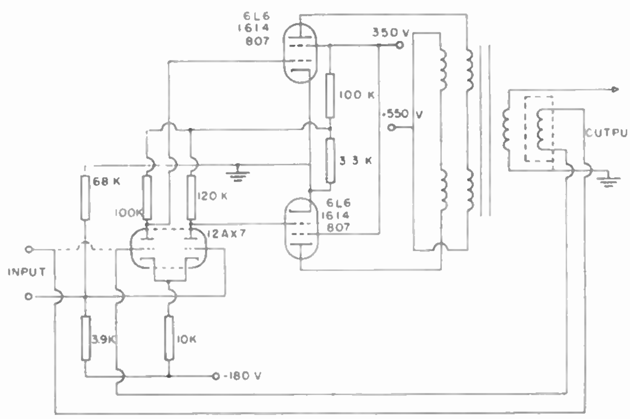


Fig. 1—Basic Bereskin power amplifier circuit.

Basic Circuit

Fig. 1 shows the basic circuit of the Bereskin Power Amplifier. In this circuit two beam power tubes are connected in push-pull with their cathodes at common ground potentials. The screens are fed from any suitable power supply which need not be derived from the plate power supply. The screen and plate supply voltage may therefore be chosen independently to best suit the particular application.

The dual triode acts as a direct coupled phase inverter and driver supplying enough bias and drive for Class B_1 operation of the beam power tubes. Class B operation of the output tubes requires that the output transformer primaries be bifilarly wound in order to avoid the conduction transfer notch. A feedback winding, closely coupled to the bifilar primary and to the secondary, and statically shielded from them, is connected in series with the input to the grid of the left-hand section of the dual triode. Good coupling between the bifilar primary and the secondary is assured by dividing the primary into two sections and sandwiching the secondary between these two sections.

This type of circuit has produced stable operation with 40 db of feedback, but this amount of feedback requires an excessive amount of driving voltage so that a value closer to 25 db is generally used. Due to the large amount of feedback used, the circuit is relatively insensitive to screen and plate supply ripple and regulation. The input signal may be either transformer- or impedance-coupled to the phase inverter-driver. Resistance-capacitance coupled input has been found to be relatively unsatisfactory due to the large dc resistance introduced in the grid circuit of the input triode. Grid current flow in this resistance tends to produce dc bias unbalance in the output tubes.

An appreciable amount of capacitance exists between the bifilar wound primaries of the output transformer. This capacitance tends to limit the high-frequency power delivering capacity of the amplifier. This undesirable effect of the primary interwinding capacitance is reduced by operation at low plate signal voltage and high plate signal current. The capacitance itself can be reduced by approximately one third by transposing the bifilar winding at each turn.

Design Problems and Solutions

The specifications for the power amplifier proper required that it be capable of delivering 3,000 watts in the frequency range between 400 and 4,000 cycles per second with a distortion not to exceed 10 per cent. The amplifier developed was able to deliver the required power with less than 2 per cent distortion.

Investigation and analysis of available tubes showed that a pair of 4-1000A tubes operated Class B_1 push-pull would be capable of delivering slightly in excess of 3,000 watts within the rated plate and screen dissipation if operated with 5 kv plate supply voltage and 1.25 kv screen voltage.

Information of this type is not usually available in the tube data supplied by the manufacturer. In order to develop the permissible peak plate currents without

* Received by the PGA, October 20, 1955. This paper will also appear in the 1956 IRE Convention Record. The 3,000-watt audio power amplifier was developed under contract AF33(616)-2320.

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[1] A. B. Bereskin, "A high-efficiency, high-quality audio-frequency power amplifier," Trans. IRE, vol. Au-2, pp. 49–60; March, 1954; IRE Convention Record, Part 6, Audio and ultrasonics, pp. 18–24.
A. B. Bereskin, "Fifty-watt amplifier for high-quality audio," *Electronics*, vol. 27, pp. 160–164; October, 1954.

TABLE I

POWER DELIVERING CAPACITY CALCULATIONS ($E_{bb}=5$ kv, $E_{c2}=1.25$ kv)

$E_{c2}=1$ kv, $e_{c1}=0$ v		$E_{c2}=1.25$ kv, $e_{c1}=0$ v		
e_b (volts)	i_b (amp)	e_b (volts)	i_b (amp)	W_{max} (watts)
600	.930	750	1.252	2,660
800	1.130	1,000	1.520	3,040
1,000	1.270	1,250	1.710	3,210
1,200	1.340	1,500	1.805	3,160

positive grid voltage drive it is necessary to operate the tubes with screen voltages considerably in excess of the values specified for Class AB_2 and B_2 operation. This is generally permissible if the various insulation and power dissipation limitations are not exceeded.

Insulation limitations must be determined either from manufacturer's data or by test. For this particular tube EIMAC specifies in pamphlet Number 3, "Pulse Service Notes," that the 4-1000A tube may be operated with 2.5 kv screen voltage and 15–30 kv plate supply voltage in pulsed operation. While the operation involved here is not of the pulsed variety, the insulation available is believed to be entirely adequate for the purpose.

The determination of plate and screen dissipation requires transposition of the available plate and screen characteristics to values corresponding to the desired value of screen voltage. This can be done if it is remembered that the voltage field pattern and the current distribution will not be altered if all interelectrode voltages are either raised or lowered by the same relative amount. The current values, however, will take on new values in accordance with the 3/2 power law.[2] (EIMAC suggests that the use of the 4/3 power provides more accurate information.) If an adequate amount of feedback is used then it can be assumed that a sinusoidal input signal will result in sinusoidal plate current and plate voltage variation.

The manner in which the deliverable power can be computed is shown in Table I above. Here the first two columns were obtained from the $e_{c1}=0$ v, $e_{c2}=1$ kv data supplied by the manufacturer. Columns 3 and 4 represent computed data for $e_{c1}=0$ v, and $e_{c2}=1.25$ kv. Column 3 was obtained by multiplying the values in column 1 by 1.25. Column 4 was obtained by multiplying the values in column 2 by $1.25^{4/3}=1.347$. Column 5 represents the maximum ac power that can be developed in Class B_1 push-pull operation for $e_{c2}=1.25$ kv when e_{bmin} and i_{bmax} correspond to the values in columns 3 and 4 respectively and a plate supply voltage of 5 kv is used.

Maximum average plate dissipation, with sine wave signal, will occur when the plate is driven 2/3 of the way to zero volts, corresponding to a peak ac signal voltage of 3,333 volts. For the condition of row 2 in Table I this results in a peak plate current of 1.27 amperes and an average plate current, for the two tubes, of .805 amperes. The dc input power is therefore 4,025 watts while the ac plate power developed is 2,115 watts. The maximum average plate dissipation is therefore 1,910 watts. The condition of row 1 would not supply the required 3,000 watts of output power while the conditions of rows 3 and 4 would have plate dissipation conditions exceeding the rated value of the tubes.

Maximum screen dissipation occurs at maximum drive. Table II, opposite, shows manner in which information can be set up to compute the expected screen dissipation and other important quantities assuming sufficient feedback is used to force a cosine variation of plate current and plate voltage.

The value of zero signal tube plate current was arbitrarily chosen at .100 ampere to produce 50 per cent of rated plate dissipation. This represents a reasonable compromise between severity of Class B bias and quiescent plate dissipation.

Fourier analysis applied to the data of Table II yields the following information:

$I_b=.980$ ampere
$I_{c2}=.118$ ampere
Screen dissipation = 148 watts
Plate circuit input power = 4,900 watts
AC power developed = 3,040 watts
Plate dissipation = 1,860 watts
Plate efficiency = 62 per cent

This last tabulation indicates that the operating conditions chosen will yield the required power output within the plate and screen dissipation ratings of the tubes chosen. Of course the correct operating conditions are rarely obtained on the first choice and a few incorrect operating conditions may have to be investigated before the correct one is determined.

There is no dc voltage between the two primaries, but an instantaneous peak voltage of 4,000 volts appears between adjacent points of these windings at full signal. Commercially available polyvinyl chloride, Teflon, and Kel-F insulated wires have been found to have adequate insulation strength for this purpose. Polyvinyl chloride insulated wire is not satisfactory because of its high dielectric constant (6.5), which would produce a high primary interwinding capacitance and thereby interfere with the high-frequency power delivering capacity, and its high power factor (0.10) which would produce excessive insulation temperature rise. Both Teflon and Kel-F had acceptable dielectric constant and power factor. Kel-F insulation had the advantage of higher dielectric strength, listed as high as 2,500 volts per mil, high resistance to cold flow, and lower cost. At the time of this development polystyrene foam insulation was being discussed in the literature but was not commercially available. This insulation would have a marked advantage from the point of view of dielectric constant which would approach 1.00.

[2] EIMAC Tube Reference Data on 4–65A tube, p. 6.

TABLE II

Maximum Signal Performance Calculations
($E_{bb}=5$ kv, $E_{c2}=1.25$ kv, $e_{b\min}=1$ kv, $i_{b\max}=1.52$ amp)

	At operating value of E_{c2} (1.25 kv)			At nearest available curve value of E_{c2} (1.0 kv)				At operating value of E_{c2} (1.25 kv)	
θ (degrees)	e_b (volts)	Composite i_b (amp)	Tube i_b (amp)	e_b (volts)	Tube i_b (amp)	e_{c1} (volts)	i_{c2} (amp)	e_{c1} (volts)	i_{c2} (amp)
0°	1,000	1.520	1.520	800	1.130	0	.350	0	.471
22.5°	1,310	1.405	1.405	1,050	1.043	−17	.120	−21.2	.161
45.0°	2,180	1.072	1.072	1,745	.798	−44	.040	−55.0	.054
67.5°	3,470	.581	.581	2,780	.432	−78	.015	−97.5	.020
90.0°	5,000	0	.100	4,000	.074	−140	0	−175	0

Its other characteristics would require further investigation.

The wire used for the bifilar primaries was #22 (7-30) wire with .014-inch wall of Kel-F insulation. Twisted samples of this wire were tested with 20 kv peak 60 cycle power without breaking down. The manufacturer specifies that 100 per cent of wire of this type with .008-inch wall is subjected to an Insulation Flaw ("Spark") Test with an impressed voltage of 7,500 volts rms. Samples of the .008-inch wall wire must also pass the manufacturer's test of a four-hour immersion in tap water, with the wire ends left out of the water, with a subsequent application of 5,000 volts rms for one minute between the conductor and the tap water. All of these tests represent appreciably greater dielectric stress than that encountered between the two adjacent wires of the bifilar winding. No trouble has been experienced due to the lack of insulation in the bifilar winding.

Preliminary calculations indicated that the high-frequency power delivering capacity would begin to fall off at frequencies slightly below 4 kc, so it was decided to take advantage of the reduction in primary inter-winding capacitance obtained by transposing the bifilar winding at every turn. Subsequent performance tests showed that the high-frequency power delivering capacity specifications would have just barely been met without this refinement but were quite adequately met with the refinement.

The output transformer was designed to be used with two Moloney MA-306 grain oriented C cores. The winding buildup for this transformer is shown in Fig. 2

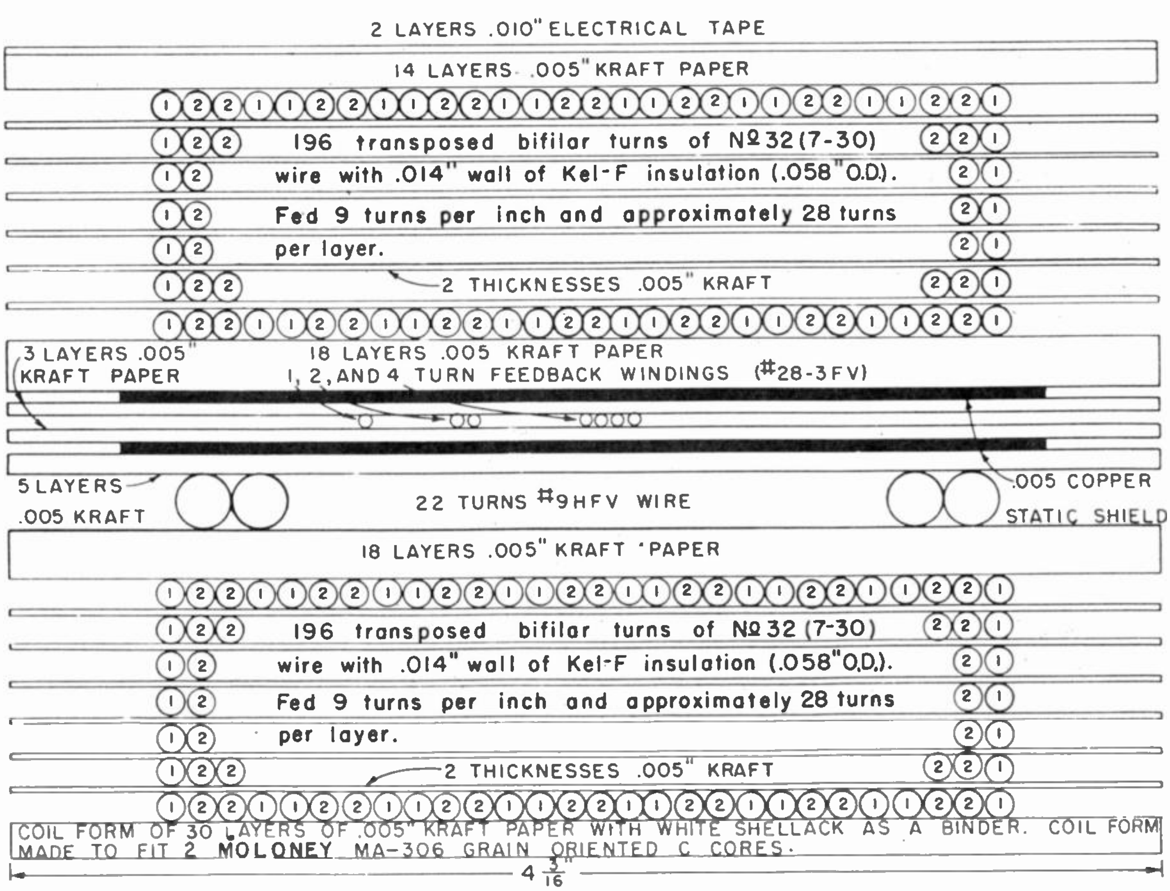

Fig. 2—Output transformer coil buildup for 3,000-watt power amplifier (proper proportions on vertical scale only).

and the complete circuit diagram is shown in Fig. 3.

In order to supply the grid driving voltage required by the 4-1000A tubes without introducing excessive grid circuit resistance, the double triode was replaced with two 6AU6 tubes. This provided the additional convenience of being able to provide dc balance by adjustment of the 6AU6 screen voltages. The choke in the impedance coupled input is a Thordarson T20C51 choke modified by full interleaving of the laminations. The voltage limits of the plate and screen supply voltages are also shown in Fig. 3.

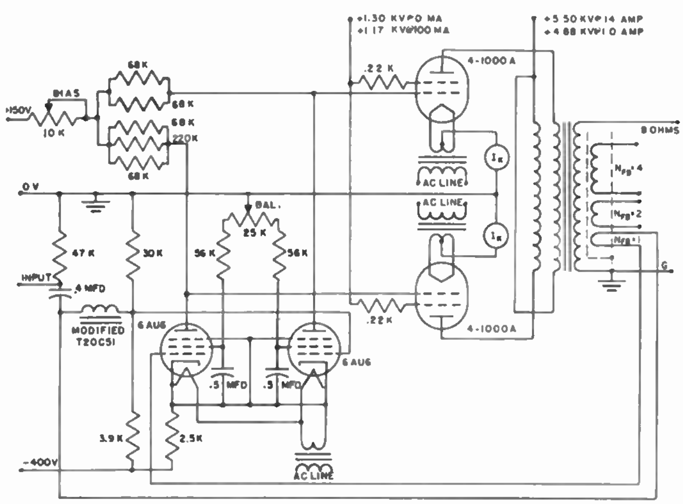

Fig. 3—3,000-watt audio power amplifier.

Performance

The effect of various amounts of feedback on the residual hum and on the full signal input voltage required is shown by the curves in Fig. 4. Operation of this amplifier was stable with the maximum possible feedback of 30 db corresponding to the full seven turns

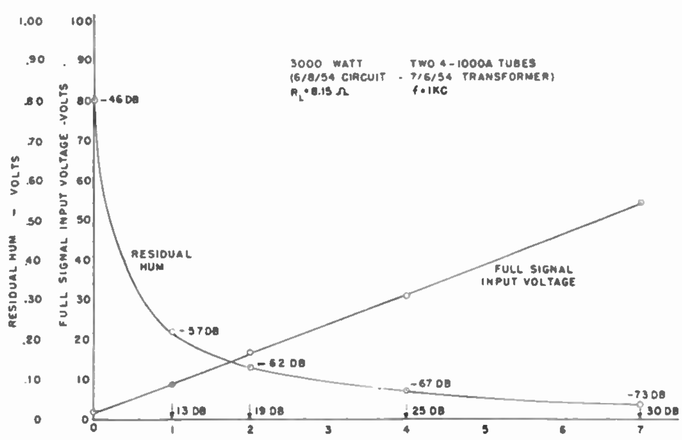

Fig. 4—Feedback analysis characteristics.

in series. A single feedback turn corresponding to 13 db provided an adequate hum level 57 db below 3,000 watts and only required 8.7 volts input voltage for 3,000 watts output. The remaining tests were performed using the single turn feedback winding.

The 1 kc power loss and distortion characteristics are shown in Fig. 5. The plate dissipation curve in this figure was obtained by subtracting the transformer losses from the total plate circuit losses. The plate dissipation exceeds the rated value by 2 per cent in the 1.4 kw output region while the screen dissipation remains below the rated value up to full power output.

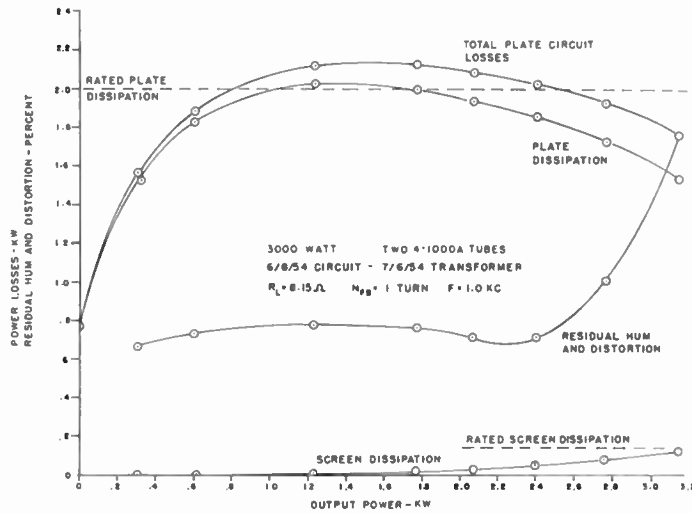

Fig. 5—Power loss and distortion characteristics.

Residual hum and distortion remains at about 0.8 per cent over most of the operating region, rising rapidly to 1.4 per cent at 3,000 watts output. Additional reduction in the residual hum and distortion could have been obtained by using more feedback but the performance was already much better than that required by the specifications.

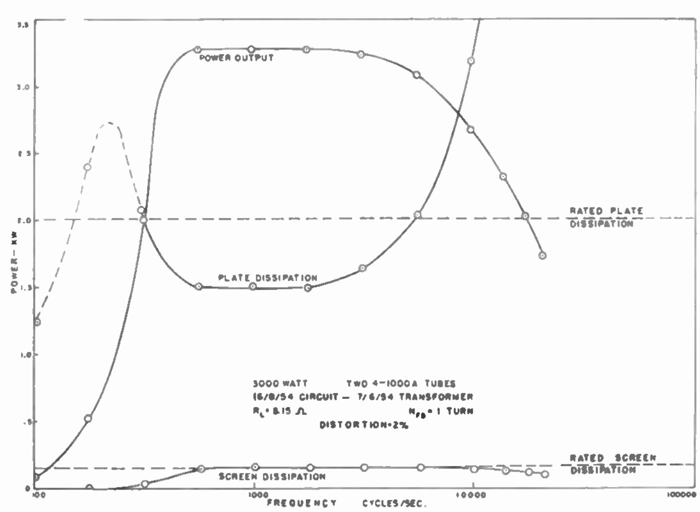

Fig. 6—Two per cent distortion power relations.

The power delivering capacity of the amplifier is shown in Fig. 6. At each of the frequencies shown in this figure the input was adjusted to produce 2 per cent distortion in the output. The amplifier is capable of delivering in excess of 3,000 watts within the plate and screen dissipation ratings of the tube over the required frequency range of 400 to 4,000 cycles per second.

The frequency response characteristics are shown in Fig. 7. The upper curve in this figure is the 2 per cent distortion power delivering capacity plotted to a db scale. The lower curve is the low level frequency response characteristic obtained by maintaining constant input voltage while varying the frequency. This characteristic deviates by less than 1 db from the middle frequency response over a frequency range of 100 to 38,000 cycles per second. The low frequency peak in response is due to series resonance between the input capacitor and the modified T20C51 choke.

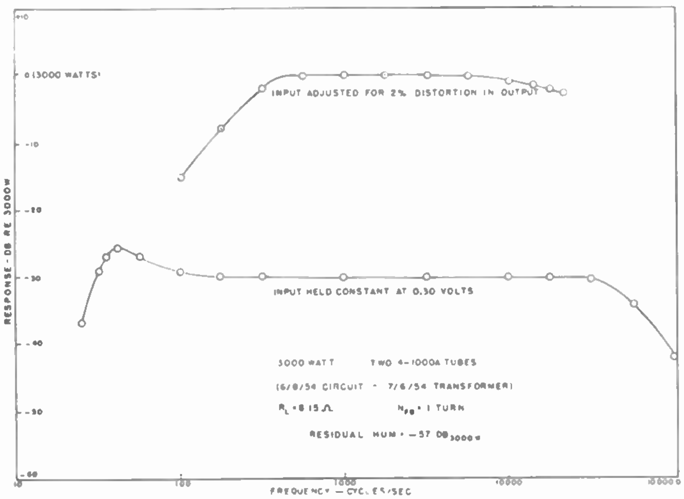

Fig. 7—Frequency response characteristics.

Fig. 8 is a photograph of the completed amplifier, exclusive of power supplies showing the chassis arrangement of the various components and the blower used for forced draft cooling of the tubes and transformer. The bottom of the chassis was enclosed and the blower provided positive air pressure inside the chassis. The air was allowed to escape through the air system sockets for cooling the 4-1000A tubes and through suitably located holes for cooling the transformer.

A characteristic of this type of amplifier is that the transition from very low power level to very high power level is made in a single step. The relative size of the two 6AU6 tubes used to drive the two 4-1000A tubes is clearly seen in Fig. 8.

The transformer, complete with mounting hardware, weighs 26.5 pounds, while the complete amplifier shown in Fig. 8 weighs 54.5 pounds.

Only one unit of the amplifier under discussion was required. If another unit had been required certain minor modifications would have been introduced in the transformer construction. The primary and secondary turns would be increased by approximately 40 per cent and the primary wire size would be changed to #24 (7-32) with a .014-inch wall of Kel-F insulation. The secondary wire size would have been changed to #10 Heavy Formvar. The changes indicated should reduce the 1 kc-2 per cent distortion core loss from 204 watts to 114 watts while increasing the copper loss from 20 watts to 40 watts. The increase in copper loss is not serious and the decrease in core loss is highly desirable. The increase in turns should move the low-frequency end of the power delivering capacity curve to the left so that this curve would cross the 3,000 watt level at about 280 cycles per second. The increase in number of turns would tend to increase the primary interwinding capacitance but this would be overcome to a large extent by the reduction in the wire size so that the high-frequency power delivering capacity would not be expected to change appreciably.

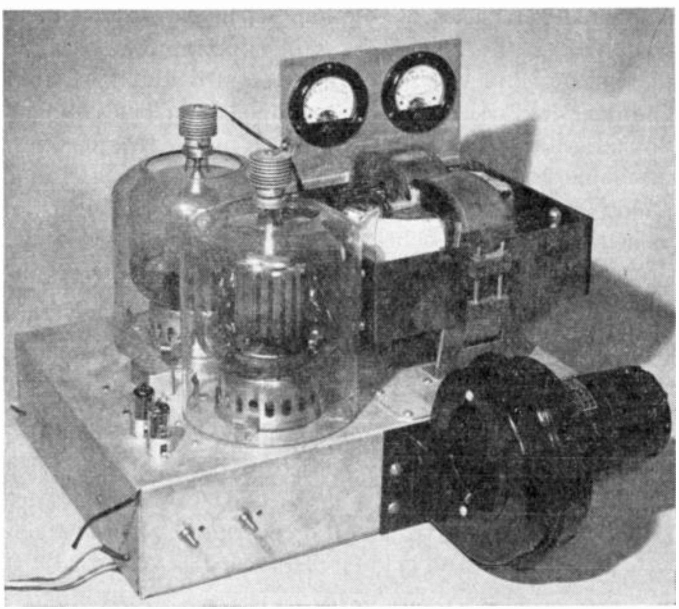

Fig. 8—3,000-watt audio power amplifier.

If the specifications had called for operation at considerably lower frequencies than those required, the transformer would have been designed with a much larger magnetic path cross section. Modifications in the number of turns and wire size would also have to be made.

The amplifier developed satisfied all of the design specifications and showed very good correlation between design data and final performance.

Triode Cathode-Followers: A Graphical Analysis for Audio Frequencies*

T. J. SCHULTZ†

Summary—**Graphical methods are presented with which one may determine the operating path upon the plate-current characteristic curves for several commonly used feedback circuits. These include the ordinary RC-coupled triode stage without a cathode bypass capacitor and two forms of the cathode follower: one in which the grid is returned to ground, the other in which it is returned to a tap on the cathode bias resistor. Once the operating path is established, the familiar computations which determine the gain, the 2nd and 3rd harmonic distortion, the dissipated and delivered power, etc. for the common triode stage may be used here. Excellent agreement is found between predicted and measured results.**

THERE ARE well-known graphical constructions whereby, when a triode and the circuit constants are chosen, one can predict the behavior of a conventional plate-loaded amplifier with respect to quiescent voltage and current, gain, maximum allowable plate swing, 2nd and 3rd harmonic distortion, dissipated and delivered power, etc.[1] This paper describes a similar procedure for cathode followers at audio frequencies, using methods closely parallel to those of the familiar triode construction. It will be sufficient to establish the operating path for each case, for once this is found the distortion and power computations are the same as for the triode. All of the following constructions use as examples one half of a 12AU7 as the tube of the circuit.

The Triode Amplifier

I. We begin by reviewing familiar construction for a triode amplifier with by-passed cathode resistor (Fig. 2). The equations governing the circuit behavior are found by summing the voltage drops in loops taken around the plate and grid circuits respectively:

$$e_b = E_B - i_b(r_p + r_k) \qquad (1)^2$$

$$e_c = -e_k = -i_b r_k \qquad (2)$$

where these quantities and others to be used are defined in Fig. 1. These two equations may be plotted on the plate characteristic curves of the tube. For (1), the construction consists of simply connecting with a straight line the point on the horizontal axis ($i_b = 0$) which represents the plate supply voltage E_B and the point ($e_b = 0$) on the vertical axis at which $i_b = E_B/(r_p + r_k)$. This is the familiar *load line* (for resistive loads) along which the operating path lies. It establishes the values of e_b and i_b for any grid voltage e_c. Eq. (2) is

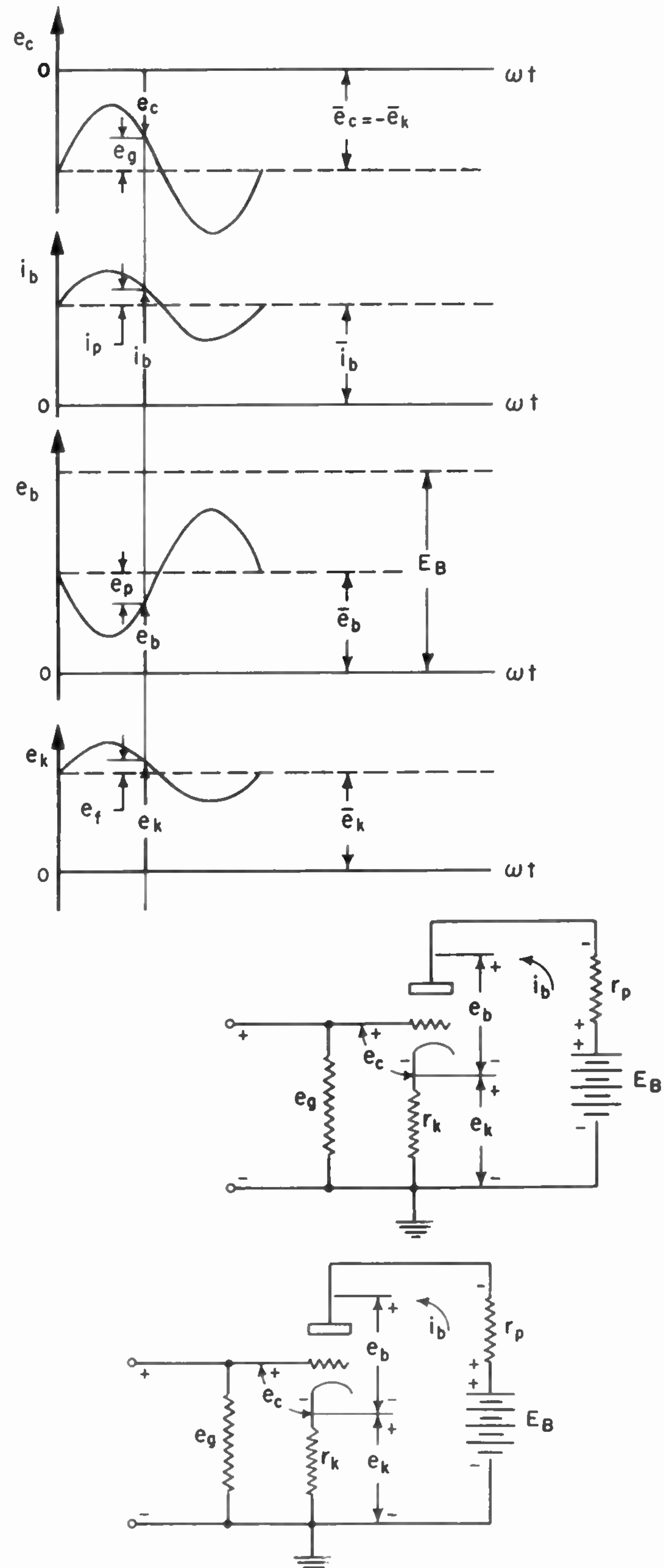

Fig. 1

* Received by the PGA, November 21, 1955.
† Douglas Aircraft Co., Inc., Santa Monica, Calif.

[1] a) Cruft Electronics Staff, "Electronic Circuits and Tubes," McGraw-Hill, 1947, New York, N. Y., pp. 275–282; 1943.
b) F. E. Terman, "Radio Engineer's Handbook," McGraw-Hill, New York, N. Y., pp. 377–382; 1943.
c) F. Langford-Smith, ed., "Radiotron Designer's Handbook," 4th ed., 1952, distributed by RCA, pp. 486–487, 548–550, 559. This book also contains an excellent treatment of cathode-followers: pp. 316–330, 390–401.

[2] Usually $r_p \gg r_k$ and r_k may be neglected.

plotted point by point as follows: given r_k, find from (2) the value of i_b necessary to give the e_c appearing as the parameter on each characteristic curve, and plot a point on the characteristic at this value of current (see Fig. 2). Connecting these points yields the *cathode-bias* curve (not necessarily a straight line) which establishes the voltage drop across r_k for any value of i_b. This voltage appears in series with the grid signal. Note that because of the capacitor C, whose reactance is small compared to r_k at audio frequencies, a bias for the grid will be established by the *average* i_b and will not fluctuate with signal variations in the plate current. Thus when we apply a signal to the grid, total grid-cathode voltage e_c will swing with signal e_s about the $-\bar{i}_b r_k$ bias.

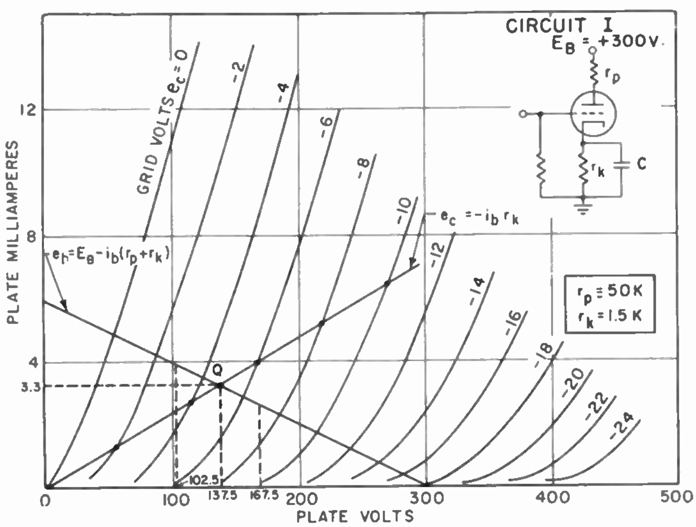

Fig. 2

The intersection of the load line with the cathode bias line locates the quiescent (or Q) point for the circuit and represents the point at which the voltages and current of the circuit are self-consistent in the absence of a signal. These values of voltage and current are the same as the average values in the presence of reasonably small signals. The voltage drop across the tube e_b appears along the voltage axis between the origin and the Q-point; the voltage between the Q-point and E_B represents the sum of the drops across the resistors in series with the tube. The plate current i_b is read on the vertical axis from the origin to the Q-point. When a rapidly varying signal is applied at the grid, the operating point moves along the load line [curve of (1)], always occupying a position corresponding to the instantaneous value of total grid voltage e_c; this determines the corresponding instantaneous values of i_b and e_b and we may then compute gain, distortion and power distribution by the procedures in the references previously given.[1]

In the illustrated example, Fig. 2, $r_p = 50k$ ohms and $r_k = 1.5k$ ohms. The intersection of the two plotted equations occurs at Q where $\bar{i}_b = 3.3$ ma; the grid bias is -5v and e_b, the voltage-drop across the tube itself, is 137.5v. Since r_k is usually much smaller than r_p, we can regard the potential e_b as the voltage from plate to ground. These are the steady conditions in the absence of signal.

Let us now apply at the grid a signal having a peak value of 2v. The operating point moves from Q in alternate directions along the load-line between positions limited by the characteristic curves $e_c = -3$v and $e_c = -7$v (*i.e.*; $e_c = -5 \pm 2$v); the corresponding extremes of plate voltage are 102.5v and 167.5v. The gain may then be computed as $(167.5 - 102.5)/(2 \times 2) = 65/4 = 16.25$.

Since r_k is by-passed by a capacitor it does not enter into the operation of the circuit at signal frequencies and, once the Q point was established, the operating path should therefore strictly have been plotted from a modified equation, (1)′, $e_b = E_B - i_b r_p$; but since $r_k \ll r_p$ there is usually a negligible difference between the curves of (1) and (1)′. We may compare the graphed with the measured values for this circuit as follows:

TABLE I

	Measured	Graphed
Gain	16.2	16.25
i_b	3.33 ma	3.3 ma
$\bar{e}_k = -\bar{e}_c$	-5v	-5v

A limit is imposed on the permissible grid swing by the requirement that the grid never go positive; *i.e.*, the operating path must not extend beyond the $e_c = 0$ curve.

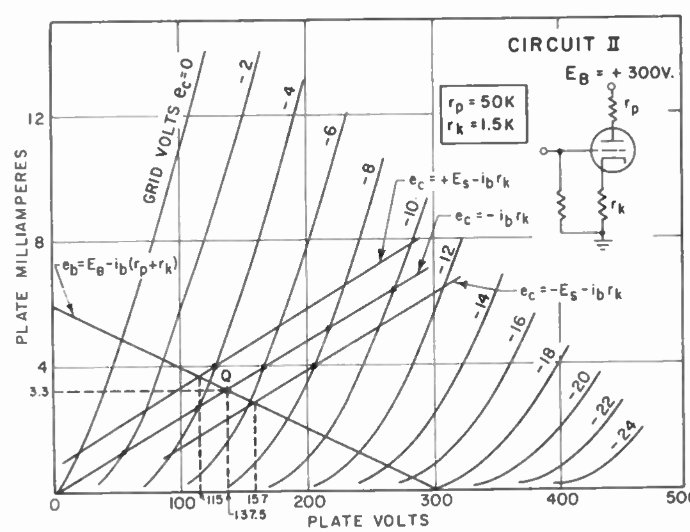

Fig. 3

II. For the case in which the by-pass capacitor C is omitted (Fig. 3) the procedure is similar except that in establishing the extremes of the grid swing we must now consider the voltage changes at the cathode caused by the signal current flowing in r_k.

The Q point is established as before from (1) and (2) and will be, for identical values of r_p and r_k, the same as in case I.

Now if a constant dc voltage were applied at the grid we could account for it with an additional term in (2), plot equations as before and regard their intersection as the new operating point. Moreover if the applied voltage at the grid, instead of having a constant dc value, were changing continuously, it would be possible in principle to repeat this process finding a separate intersection and operating point for each value through which the grid voltage passes; this would define a locus which establishes the complete path of operation. In practice, however, it is sufficient to locate only the end points, since the entire path will lie along the load line. Let us now do this for the 2v applied signal of example 1 to show the effect of removing the cathode capacitor. To find the extremes of the operating path we imagine that we have connected between grid and ground, successively in both positive and negative polarities, a battery whose voltage equals the peak applied grid signal, E_s. We plot in each case a new equation, using i_b and the characteristic curves as coordinates:

$$e_c = \pm E_s - i_b r_k \tag{3}$$

This equation states that for each value of i_b we find the new extremes of total grid voltage by moving along a constant i_b line, in each direction from the load line, a distance E_s, measured on the plate characteristic curves themselves. For example in Fig. 3, the point on the cathode-bias line at $\bar{i}_b=4$ ma, and $e_c=-6$v is moved each way along the "$i_b=4$ ma-line" to points where $e_c=-4$v and -8v, corresponding to a grid excursion E_s of ± 2v. This would be done for all points on the cathode-bias line. The actual construction procedure can be done more simply by establishing any pair of these grid voltage extremes (those already illustrated above, for example) and drawing lines through them parallel to the cathode-bias line already established.

The intersections of these new curves with the load-line then give the extremes of the grid-cathode voltage excursion (the ends of the operating path) and the corresponding values of i_b and e_b. The output voltage in this case is reduced by negative feedback in the cathode circuit, for the gain is found to be $(157-115)/(2\times 2)=42/4=10.5$, compared with a measured 10.4 and with the previous gain of 16.2. The removal of the capacitor has caused a loss in gain of about 3.7 db. The same limitation on permissible grid swing applies in this case as in the previous one.

The Cathode Follower

III. When the plate resistor is removed, the circuit becomes that of a cathode follower (Fig. 4). This corresponds to the "B" configuration of a previously published paper[3] which provided design curves for two common cathode follower circuits. The r_p term drops from (1) so that the governing equations for the circuit are now

[3] T. J. Schultz, "Triode Cathode-Followers for Impedance Matching to Transformers and Filters," TRANS. IRE, vol. AU-3, No. 2, pp. 38–37; March–April, 1955.

$$e_b = E_B - i_b r_k \tag{4}$$

$$e_c = - i_b r_k \tag{2}$$

and the no-signal condition is found from the intersection of the plotted lines: $\bar{i}_b=2.6$ ma, $\bar{e}_k=13.0$v. Note that e_k, the voltage from cathode to ground, is represented on the graph between the operating point and E_B; e_b, the voltage drop across the tube, is represented as before between the origin and the operating point.

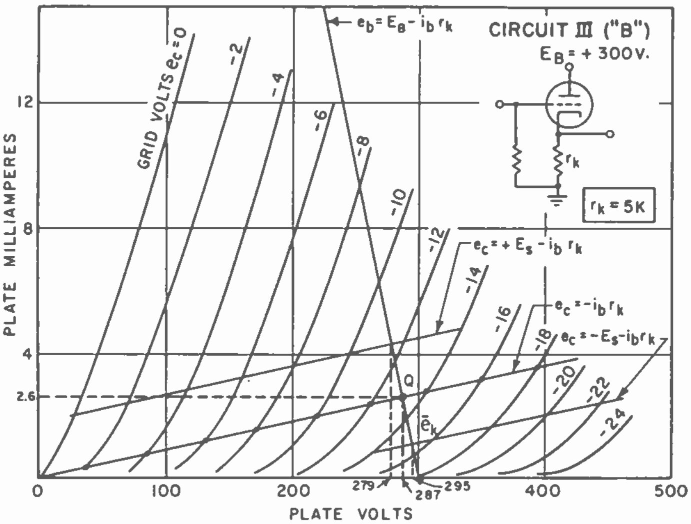

Fig. 4

Again we use the imaginary battery $\pm E_s$ to establish the operating extremes with (3). In this case there will usually be a higher grid bias and we can afford a greater peak grid swing. Let us choose ± 10v. Following the same construction procedure, we move Q along the $i_b=2.6$ line to points $e_c=3.0$v and $e_c=23.0$v and through these points draw curves parallel to that of (2).

These new lines intersect the load-line to determine the extremes of the operation path, with corresponding extremes of cathode (*i.e.*, output) voltage of 21v and 5v (measured from the +300v point). The gain is computed as $(21-5)/(2\times 10)=0.8$, compared with a measured 0.78, a severe decrease of gain from the values for either of the plate-loaded conditions. The measured $\bar{i}_b$ was 2.7 ma.

It should be noticed that some 2nd harmonic distortion may be expected here since the excursions to either side of the Q point are unequal. The limit on permissible grid excursion is usually imposed by increasing distortion as the tube nears cutoff on the negative swing.

IV. In order to reduce the bias on the tube without decreasing the load resistor and thus allow greater signal excursions without distortion, the grid is sometimes returned to a tap on the cathode resistor instead of to ground (Fig. 5). This is configuration "A" of the empirical curves of a previous reference.[3] This construction is slightly more complicated but is the same in principle as in the previous examples.

The governing equations in this case are

$$e_b = E_B - i_b(r_1 + r_2) \tag{5}$$

and

$$e_c = - i_b r_1 \tag{6}$$

The intersection of these curves gives the no-signal condition: $\bar{i}_b = 7.5$ ma and $\bar{e}_k = 82.5$v, of which the fraction

$$\frac{r_1}{r_1 + r_2} = \frac{1}{11} = 7.5\text{v}$$

is effective as bias on the tube; the remaining 75 volts represents $\bar{i}_b r_2$.

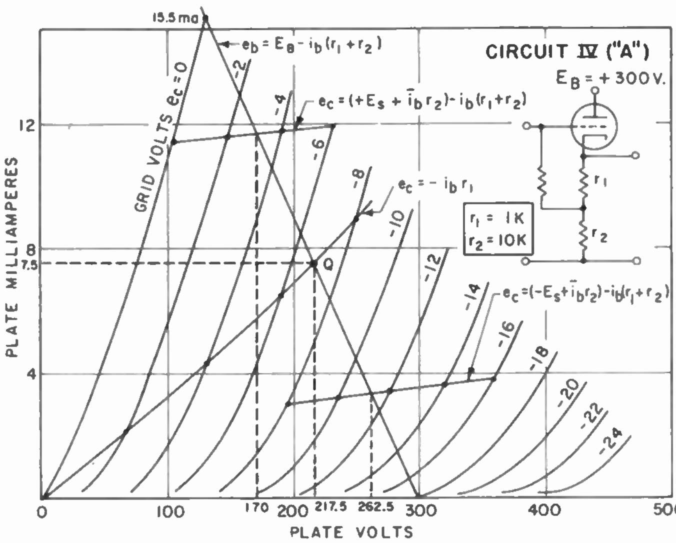

Fig. 5

In determining the extremes of the operating path we must notice two differences which show up for the first time in this case: 1) there appears as a constant term in the grid loop, in addition to the applied signal, the average voltage drop across r_2; and 2) the resistance across which the feedback is generated is different from that which establishes the bias on the tube. Hence, the *signal equation* [analogous to (3) of the previous cases, for we again employ the fictitious battery] is

$$e_c = (\pm E_s + \bar{i}_b r_2) - i_b(r_1 + r_2) \tag{7}$$

Eq. (7) will yield curves which are *not* parallel to that of (6); moreover, in this case (6) itself is not a straight line. For this reason it will not be possible to follow the procedure of cases II and III where we first plotted the cathode-bias line (2) and then found the grid-voltage extremes with signal lines parallel to and appropriately spaced from it. Now we must plot point by point not only the cathode-bias line (6) but also both "signal" lines (7) using in the latter the value (75v) of $\bar{i}_b r_2$ found above for the quiescent condition. Again the intersections with the load line establish the extremes of output voltage and plate current. In this case, the input signal was chosen with a peak value of 50v for which the graphical construction predicts that the cathode voltage at the end points (appearing on the graph between the operating point and +300v) will be 130v and 37.5v. This would give a gain of $(130-37.5)/(2\times 50) = 0.925$, compared to the measured 0.93. The graphed $\bar{i}_b$ was 7.5 ma, compared to measured 7.8 ma.

The limit on permissible applied signal may be imposed by increasing distortion either as the tube approaches cutoff on the negative swing or as the grid attempts to go positive with respect to the cathode.

Added Loads

To deal with added resistive loads in any of these circuits, it should be assumed that capacitive coupling is employed so that the Q-point is not altered by the load. If this is not the case, then the load may be lumped together with the appropriate resistor for the previously described construction. When the circuit is loaded, the Q-point is determined in the usual way as described above, the presence of the load being disregarded. However, the load line along which the operating point moves is affected by the load impedance as well as by the original circuit constants; it passes through the same Q-point but with a slope which is found by substituting in the appropriate position in the equations a term which represents the parallel impedance of the load and the paralleled circuit element. For instance, *for the purpose of determining the slope only,* in circuits I and II (1) becomes

$$e_b = E_B - i_b(Z_1 + r_k)$$

where Z_1 represents the combined impedance of r_k and the load in parallel; (2) and (3) are unchanged. In circuit III (3) becomes

$$e_c = \pm E_s - i_b Z_2$$

and (4) becomes

$$e_b = E_B - i_b Z_2$$

where Z_2 represents the combined impedance of r_k and the load in parallel. In circuit IV (5) becomes

$$e_b = E_B - i_b Z_3$$

and (6) becomes

$$e_c = (\pm E_s + \bar{i}_b r_2) - i_b Z_3$$

where Z_3 represents the combined impedance of $(r_1 + r_2)$ and the load in parallel.

Again, these substitute equations are used only to determine the slope of the ac path of operation, which passes through the Q-point as does the dc load line.

If Z_1, Z_2 or Z_3 happen to be reactive the operating path will be elliptical and the simple graphical constructions discussed in this paper will be inapplicable.

Contributors

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A. B. Bereskin (A'41–M'44–SM'46) was born in San Francisco, Calif. in 1912. He received the Electrical Engineering degree in 1935 and M.Sc. in Engineering degree in 1941, both from the University of Cincinnati.

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Before joining the faculty of the University of Cincinnati Mr. Bereskin worked for the Commonwealth Mfg. Corp. and the Cincinnati Gas and Electric Co. During the years 1944–45, while on leave of absence from the University of Cincinnati, he worked for the Western Electric Co. He is also active in consulting engineering and is a registered professional engineer in the State of Ohio.

In the IRE Mr. Bereskin is a Senior Member, a member of the Education Committee, Chairman of the Region IV Subcommittee of the Education Committee and Institute Representative at the University of Cincinnati. He is Past Chairman, Vice-Chairman and Treasurer of the Cincinnati Section and is at present Editor of the IRE Transactions on Audio. He is also a member of the AIEE, Sigma Xi, and Eta Kappa Nu.

Mr. Bereskin has published work on audio power amplifiers, video amplifiers, regulated power supplies, and power factor meters. He has also done work in the fields of special RC oscillators, frequency selective amplifiers, low jitter multivibrators, special stabilized power supplies, and transistor audio power amplifiers.

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R. J. Larson

R. J. Larson (S'51–A'52–M'55) was born in January, 1926, at Galesburg, Ill., and graduated from secondary school in Minneapolis, in 1944. After two years' service with the Navy, he attended Knox College for a year, transferring to Northwestern University from which he graduated in 1951, with a degree of Bachelor of Science in Electrical Engineering.

While an undergraduate, he was employed by the Jensen Manufacturing Company as a student engineer under Northwestern's Cooperative Program. After twenty-two months on active duty with the Navy during the Korean conflict, he returned to Jensen, where he is presently engaged in loudspeaker development work.

Mr. Larson is a member of the Institute of Radio Engineers, the Chicago Acoustical & Audio Group, and was elected to Tau Beta Pi and Eta Kappa Nu while at Northwestern. His hobbies include music and amateur radio, using the call letters W9UYU.

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T. J. Schultz

T. J. Schultz was born in 1922 in Jefferson City, Mo. where he attended grade and high school. He attended the Eastman School of Music of the University of Rochester in 1940–41 the Electrical Engineering Departments of the University of Missouri in 1941–43 and the University of Texas in 1943–44. He attended the U. S. Naval Academy in 1944 and remained there to teach electrical engineering.

Mr. Schultz attended Harvard University in 1946–47 and received the S.M. degree. He was a research physicist with the Naval Research Laboratory in 1947–48 but returned to Harvard for the years 1948–54, during which time he did research, taught, and obtained a Ph.D. degree.

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