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TABLE OF CONTENTS

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Professional Group on Audio

IRE PROFESSIONAL GROUP ON AUDIO

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

IRE TRANSACTIONS® ON AUDIO

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The Editor's Corner

NOISE AND OTHER RANDOM PHENOMENA

LONG ago in a far away kingdom, his majesty was enjoying music from the royal stereo high-fi rig. In the middle of his favorite aria a crackling noise came out of the high-range super-tweeter, spoiling the whole program.

The king mentioned this annoyance informally at the following cabinet meeting and was surprised at the reaction. The minister of education said it was a sign of technical incompetence caused by antiquated teaching methods, underpaid instructors, and inadequate facilities which should be cured by government aid.

The minister of defense thought that it was a new form of "jamming" developed by enemy nations, and pointed out the need for an early warning radar screen, and a deterrent retaliatory force.

The minister of security suspected internal sabotage which should be held in check by more investigating committees.

The minister of commerce insisted that it would not have happened if adequate aid had been given to small business, to depressed areas, and to expanding the gross national product.

The minister of labor said that it was a manifestation of insecurity of the working force which could be remedied by fringe benefits and higher salaries.

The minister of health observed that we would need healthier taxpayers to finance the expanded budget.

These viewpoints made the king realize that what he needed was short-term help. He remembered his one year or 12,000 mile guarantee, and made this known to the president of the Stereo Atomic Space Automated High-Fi Company.

The president thought the fault lay in too low a protective tariff against cheap foreign sets. Nevertheless he assigned the problem to his vice president of engineering. The vice president muttered something about more basic research and that he needed new laboratories. But perhaps a brainstorming session would help, so he organized a meeting of his chief scientists.

The theoretical physicist said that since this was a noise problem, it required a general analysis of the statistical distribution of random phenomena, and information theory.

Another physicist thought that it could be set up on a computer, but he needed a larger one having solid-state electronics and thin film cryogenics.

The experimental physicist said it could be cured with masers, parametric amplifiers, and perhaps the Mossbauer effect could be worked in somehow.

The chief engineer said that their company's product was antiquated and should be ruggedized, transistorized, and miniaturized, preferably with molecular electronics and topology included.

Working staff members were called in, and had a conference of their own. The group leader maintained that in order to carry on any meaningful tests they needed a field free room with acoustic absorbing wedges, octave band noise analyzers, and automatic curve tracing equipment.

The design engineers looked for some way in which cathode followers and silicon rectifiers could be incorporated.

The test engineers wanted a program involving temperature cycling in a new environmental chamber, shake tables, life testing of components, and a facility for irradiating with neutrons and gamma rays.

The production engineers were in favor of automated checkout fixtures on the assembly line.

The newspapers heard of it, and increased their circulation with headlines such as: "Suspect Subversive Sabotage," "Education Stymies Space Race," "Study Science Lag," "Firms Vie for Aid."

The queen learned about the activities which were taking place, and mentioned to the king: "You know, at the time you were listening to your set, I was using the electric shaver. Could that have something to do with it?" A quick test showed that it could and it had.

Steps were taken to quiet down the incident, although the true facts were never revealed for fear of embarassing the dignitaries involved. Soon everything was forgotten, tranquillity reigned, and everybody lived happily ever after.

—MARVIN CAMRAS, *Editor*

PGA News

ADMINISTRATIVE COMMITTEE MEETING MINUTES

Chicago, Ill., October 9, 1960

Members Present

Hugh Knowles (Chairman)
A. B. Bereskin
Cyril M. Harris
Marvin Camras
Benjamin B. Bauer (Secretary-Treasurer)

Members Absent

Peter C. Goldmark
John K. Hilliard
Murlan S. Corrington
J. Ross Macdonald
Robert W. Benson
Michel Copel

Guests Present

William M. Ihde (Chicago Section Chairman)
John I. Sheetz (Cleveland Section Chairman)

Only three members of the Administrative Committee being present, in addition to the Chairman, there was no quorum. Therefore, any actions that normally require an official vote of the Administrative Committee would have to be subject to a letter ballot.

There has been no reply to date to the letter of April 6 to the IRE Headquarters from the Secretary-Treasurer about the question of the Constitutional amendment. It was agreed that the Secretary would follow up the Headquarters and arrange with the *Editor* for publication of the revised Constitutional amendment as soon as possible.

Pursuant to a recent request from Headquarters for a clarification and review (*Note:* Letter from Larry Cumming dated July 27, 1960), a discussion was held about the general scope of the field of interest of the group. It was pointed out that the present statement of scope was broad enough to include the desired activities of the group. However, the restriction about recording and reproduction as taking place at audio frequencies did not portray the present development of the art. By motion of Cyril Harris and seconded by Marvin Camras, it was agreed to delete the words "at audio frequencies" from the sixth line of Section I as currently constituted. This motion was passed unanimously, with the proviso that it would be submitted to the full membership of the Administrative Committee for a letter ballot.

In response to a request from Michel Copel, the Chairman of the Audio Technical Program at the 1961 IRE International Convention, it was agreed that all members of the Administrative Committee and all Committee Chairmen and Chapter Chairmen would be requested to send suggestions to Mr. Copel as soon as possible.

The Chairman pointed out that due to the early closing dates for various Conventions, such as the NEC and the WESCON, it became necessary to appoint program committees early in the Spring. It was suggested that if a General Program Committee Chairman can be found this year, he would be requested to appoint the committees for the 1961 WESCON and the 1961 NEC during the fiscal year ending March 31. In the event that a General Chairman is not found, then the National Chairman will appoint special purpose committees for these two meetings before the expiration of his term in office.

The report of the Awards Committee is as follows: (Refer to a letter from Chairman Bereskin of September 14, 1960, enclosed).

The report of the Nominating Committee is as follows: Benson and Harris have accepted nominations for the position of Chairman and Vice Chairman. The other holdover members of the Administrative Committee have refused to run. H. E. Roys' name has been placed in nomination for Chairman, and the Nominations Committee Chairman will seek his consent. The committee expects to complete all the nominations within the next two weeks, at which time the consent of the Administrative Committee will be obtained by letter ballot.

Benjamin B. Bauer
Secretary-Treasurer

CHAPTER OFFICERS AND MEMBERSHIP STATISTICS

The IRE Professional Group on Audio includes twenty-three chapters listed below. Paid members are as of December 31, 1959. Officers are listed from the latest information received by the IRE headquarters to December, 1960.

Chapter	Chairman	Vice-Chairman	Secretary-Treasurer	Paid Members*
Albuquerque-Los Alamos 1959–60	C. W. Hicks 1316 Guaymas Pl., NE Albuquerque, N. M.		Ben F. Sedlack (Secy.-Treas.) 801 Vassar Dr. NE Albuquerque, N. M.	44
Baltimore 1960–61	Louis R. Mills Recordings, Inc. 735 Deepdene Rd. Baltimore 10, Md.	James H. Jackson (Secy.) The Martin Co. Baltimore 3, Md.	W. Parrish Air Arm Div. Westinghouse Corp. Baltimore, Md.	74
Boston 1960–61	Donald J. Fritch Lessells & Assoc. 916 Commonwealth Ave. Boston 15, Mass.	Henry S. Littleboy Baird-Atomic Corp. 33 University Rd. Cambridge, Mass.	Charles L. Malme (Secy-Treas.) Bolt Beranek & Newman 50 Moulton St. Cambridge, Mass.	305
Chicago 1960–61	William Ihde General Radio Co. 6605 W. North Ave. Oak Park, Ill.	Peter Tappan Warwick Mfg. Co. 7300 N. Lehigh Chicago 31, Ill.	James F. Novak (Secy.) Jensen Mfg. Co. 6601 W. Laramie Ave. Chicago 38, Ill.	242
Cincinnati 1959–60	Wm. C. Wayne, Jr. 70 Pleasant Ridge Ave. South Fort Mitchell, Ky.	Clyde G. Haehnle Crosley Bdcstg. Co. 140 W. 9 Street Cincinnati, Ohio	John P. Quitter (Secy.-Treas.) 3837 Broadview Dr. Cincinnati, Ohio	52
Cleveland 1960–61	John Sheetz Ohio Bell Telephone 700 Prospect Ave. Cleveland 15, Ohio	Henry Hoff (same as Chairman)	Harry Dennis (Secy.) 1501 Euclid Ave. Cleveland 15, Ohio	47
Dayton 1960–61	Taulbee P. Mountz, Jr. Comm. & Nav. Lab. Wright-Patterson AFB Ohio	Stanley E. Weber WADC Wright-Patterson AFB Ohio	Albert P. Parker (Secy.) (Same as Vice-Chrm.)	47
Detroit Inactive				66
Hawaii 1959–60	D. H. DaShiell 224 Awakea Rd. Lanikai, Oahu, H.	Iwao Miyake Univ. of Hawaii Honolulu, Hawaii	Daniel L. Pang (Secy.-Treas.) 1809 Naio St. Honolulu, Hawaii	20
Houston 1960–61	James H. Carter The Wrye Co. 2410 W. Alabama Houston, Texas			47
Kansas City Inactive				34
Los Angeles Inactive				473
Milwaukee 1960–61	Otto Dobnick 1113 S. 28 St. Milwaukee, Wis.			56
North Carolina 1959–60	L. A. Cornett Western Electric Co. Lexington Rd. Winston-Salem, N. C.			39
Philadelphia 1960–61	T. A. Benham Haverford College Haverford, Pa.		D. Ridgely Bolgiano American Electronics Lab., Inc. 121N. 7 St. Philadelphia 6, Pa.	286
Phoenix Inactive				31
San Antonio 1959–60	Bill Case 122 W. White Ave. San Antonio, Texas			27

* As of December 31, 1959.

CHAPTER OFFICERS AND MEMBERSHIP STATISTICS (Cont'd)

Chapter	Chairman	Vice-Chairman	Secretary-Treasurer	Paid Members*
San Diego 1960–61	Lowman Tibbals USNEL San Diego 52, Calif.	Alfred Matthews Coronado Electronics Box 544 Coronado, Calif.		65
San Francisco 1960–61	Mort Fujii PO Box 3000 Mail Stop 331–2 Redwood City, Calif.	Charles Wilkins 2501 San Ramon Ave. Mt. View, Calif.	Stanley Oleson (Secy.-Treas.) 140 S. Margarita Ave. Menlo Pk., Calif.	221
Seattle Inactive				79
Syracuse 1960–61	E. O. Crow General Electric Co. NL Room 5 Syracuse, N. Y.	Philip B. Clark RD 3 Skaneateles, N. Y.	Dwight V. Jones (Secy.) General Electric Bldg. 7, EP, Rm. 217 Syracuse, N. Y.	56
Twin Cities 1960–61	Robert Sell Telex Minneapolis, Minn.	Richard F. Dubbe Minn. Mining & Mfg. 900 Bush Avenue St. Paul 6, Minn.	Joseph F. Dundovic (Secy.-Treas.) Nortonics Inc. 1015 S. 6th St. Minneapolis 4, Minn.	65
Washington 1960–61	Sachio Saito Natl. Bureau of Standards Washington, D. C.	J. Carlisle Hoadley Diamond Ordnance Fuze Labs. Washington, D. C.	Robert J. Carpenter Natl. Bur. of Standards Washington, D. C. (Secy.)	188

* As of December 31, 1959.

PGA MEMBERS HONORED BY IRE AWARDS

George W. Bailey, Executive Secretary of the IRE, has announced IRE awards which include the following members who have contributed to audio technology.

Peter C. Goldmark—The 1961 Vladimir K. Zworykin Prize Award—"For important contributions to the development and utilization of electronic television in military reconnaissance and in medical education."

Cyril M. Harris—1961 IRE Fellow Award—"For contributions to the science of acoustics."

Arthur C. Keller—1961 IRE Fellow Award—"For contributions to the recording and reproduction of stereophonic sound and to telephone switching."

Haldon A. Leedy—1961 IRE Fellow Award—"For contributions to electronic research management."

Richard M. Somers—1961 IRE Fellow Award—"For contributions to phonographic recording and dictating machines."

PGA CONSTITUTION REVISED

The IRE Executive Committee on November 14, 1960, approved the revision to ARTICLE IV, Sections 1 and 2 of the PGA Constitution which reads as follows:

Section 1. The Group shall be managed by an Administrative Committee consisting of the Chairman and nine additional members of the Group.

Section 2. The term of office of the elected members of the Administrative Committee, excluding the Chairman, shall be three years. The number of the members necessary to bring the committee to full strength shall be elected each year by the Group membership.

Section 1 now provides for an administrative committee of ten; formerly it had nine members. Section 2 clarifies the term of office and the number of members to be elected.

In accordance with Article XI, Section 1, of the Constitution these revisions are being printed in TRANSACTIONS so as to give the membership an opportunity to object if they see fit.

CHAPTER NEWS

Chicago, Ill.—On September 9, 1960, C. Eilers, of the Zenith Radio Corporation, spoke on "Receiver Design Considerations for Stereo FM Multiplex Broadcasting."

For the October 19 meeting, the PGA met jointly with the Chicago Acoustic and Audio Group, at the Universal Recording Studios, to hear Bruno G. Staffan, of the Hammond Organ Company, describe an "Organ Reverberation Device for Stereo High Fidelity."

Hall Effect Wattmeters*

D. P. KANELLAKOS†, MEMBER, IRE, R. P. SCHUCK‡, MEMBER, IRE, AND A. C. TODD||, SENIOR MEMBER, IRE

Summary—A transmission-type wattmeter can be formed utilizing the Hall effect to give wide-range multiplication over a frequency range extending from the low audio-frequency region to the SHF band. At the low frequencies, the Hall device can take the form of a thin wafer of indium antimonide mounted in series with the center leg of a cup core magnetic circuit. An audio-frequency wattmeter employing such a structure has been analyzed, constructed and tested. The wattmeter has a range of zero to 400 milliwatts when used in a 600-ohm circuit, and displays a frequency error of less than plus and minus 3.5 per cent over a frequency range of 100 to 6000 cps for a unity power factor load.

In the SHF range, a cavity and electric probe are arranged to provide excitation for the indium antimonide wafer in the manner first employed by Barlow. The cavity, about one wavelength long, is separated from the waveguide in which power flow is to be measured by a thin brass wall, and is coupled to the waveguide by a rectangular slot in the top wall of the waveguide. The semiconductor wafer is placed in the center of the cavity, at a position having a maximum in magnetic field and a minimum in electric field for a given cavity excitation value. The electric probe, which extends from the cavity to the waveguide, provides excitation current for the semiconductor element. Three tuning adjustments are required, two for the cavity and one for the probe. A wattmeter embodying this technique has been found to provide a sensitivity of one dc microvolt per milliwatt of transmitted power at a frequency of 9.40 Gc.

I. Introduction

Two separate transmission-type wattmeters utilizing the Hall effect have been designed, constructed and tested. The first type measures power in audio range and the second is employed in X-band power measurement. Although the same principle is used in both types, the design considerations are widely divergent. This necessitates an individual description of each type. The theory of operation, design, construction, test systems employed, and performance characteristics are presented along with suitable illustrations.

II. An Audio Frequency Wattmeter

A. The Hall Device as a Wattmeter Element

The output voltage v_0 developed by a Hall device, shown in Fig. 1, may be expressed

$$v_0 = R_h i_c b/d \tag{1}$$

where

v_0 is the open-circuit output of Hall element,
R_h is the Hall constant,
i_c is the control current,
b is the normal flux density of magnetic field through the semiconductor,
d is the thickness of Hall element.

If the magnetic circuit furnishing the magnetomotive force to the semiconductor has a linear magnetization characteristic (b-h characteristic), flux density b will be proportional to the Hall device excitation coil current i_e. Hence, the output voltage developed by the Hall device may be expressed

$$v_0 = k_0 i_c i_e \tag{2}$$

where k_0 is the over-all constant for the device, in volts/ampere².

A block diagram of the wattmeter is shown in Fig. 2. Let us assume that audio power is supplied by a sine-wave generator to a linear load, such that load voltage and current are

$$v_L = V_{\max} \sin(\omega t + \phi) \tag{3}$$

$$i_L = I_{\max} \sin(\omega t + \phi - \theta), \tag{4}$$

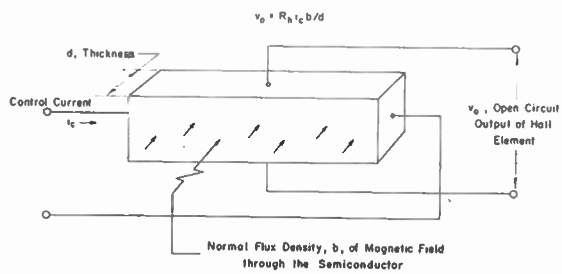

Fig. 1—Basic Hall generator.

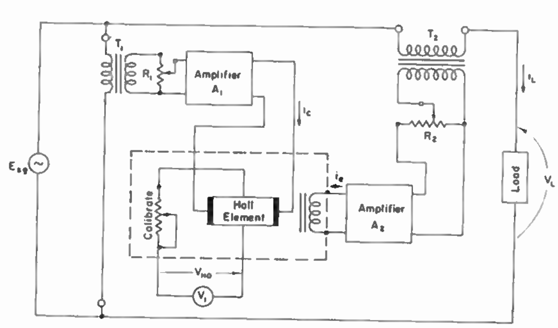

Fig. 2—Basic experimental wattmeter system.

* Received by the PGA, September 16, 1960. Presented at the Fourth Annual Joint Military-Industrial Electronic Test Equipment Symp., Chicago, Ill.; September 14–15, 1960. The work was done at Armour Res. Foundation of Illinois Inst. Tech., Chicago, Ill., as part of a program sponsored by the U. S. Army Signal Res. and Dev. Agency under Contract No. DA-36-039, SC-78269.
† Radioscience Lab., Stanford University, Stanford, Calif.
‡ Armour Res. Foundation, Chicago, Ill.
|| Hallicrafters Co., Chicago, Ill.

where

ϕ is the initial offset angle at $t=0$
θ is the impedance angle of the load.

Load voltage is monitored by potential transformer T_1, and its associated amplifier A_1; load current is monitored by current transformer T_2, and its associated amplifier A_2. The output stages of amplifiers A_1 and A_2 are assumed to be constant current devices. Furthermore, amplifier A_1 has a transconductance k_1/a_1 mhos and an input impedance of infinity; while amplifier A_2 has a current gain k_2/a_2 and an input impedance of zero. If the instrument transformers are assumed to be perfect, each may be replaced by its transformation ratio N_1 for the secondary-to-primary-turns ratio of transformer T_1, and N_2 for the primary-to-secondary ratio of transformer T_2.

Hence, the Hall device currents are related to the load voltage and current by

$$i_c = V_{\max} N_1 k_1 \sin(\omega t + \phi + a_1) \tag{5}$$

$$i_e = I_{\max} N_2 k_2 \sin(\omega t + \phi - \theta + a_2). \tag{6}$$

Therefore, the output voltage of the Hall device is related to the load voltage and current by

$$v_0 = V_{\max} I_{\max} N_1 N_2 k_0 k_1 k_2 \sin(\omega t + \phi + a_1) \cdot \sin(\omega t + \phi - \theta + a_2). \tag{7}$$

Now

$$\sin x \sin y = \tfrac{1}{2}\{\cos(x-y) - \cos(x+y)\}. \tag{8}$$

Thus

$$v_0 = \frac{V_{\max} I_{\max} N_1 N_2 k_0 k_1 k_2}{2}\{\cos(a_1 - a_2 + \theta) - \cos(2\omega t - 2\phi + \theta + a_1 + a_2)\}, \tag{9}$$

the dc component of which is

$$\overline{v_0} = \frac{V_{\max} I_{\max} N_1 N_2 k_0 k_1 k_2}{2} \cos(a_1 - a_2 + \theta).$$

If the phase characteristics of amplifiers A_1 and A_2 are made identical, $a_1 = a_2$ and

$$\overline{v_0} = \frac{V_{\max} I_{\max} N_1 N_2 k_0 k_1 k_2}{2} \cos\theta \tag{10}$$

or

$$\overline{v_0} = V I N_1 N_2 k_0 k_1 k_2 \cos\theta \tag{11}$$

where V and I are rms values of load voltage and current, respectively. Hence, it has been shown that the average value of the Hall device output voltage will be proportional to the average power in the load under test.

B. The Wattmeter System

The wattmeter is composed of a current transformer (T_2), a voltage transformer (T_1), transistorized amplifiers (A_1 and A_2), an indium antimonide Hall generator (see Fig. 1), and a dc millivoltmeter.

The transformers offer isolation between the load and the amplifier circuits, are essentially lossless, have negligible phase shift, and are essentially distortionless over the desired operating frequency range of the system.

The amplifier circuits are transistorized and biased so that they act essentially as constant current sources to the Hall device. Stability, sufficient gain, negligible phase shift, and a flat-frequency response within the working range of the instrument were the principles around which the amplifier system was designed.

The operating conditions presented for the transformers and amplifier circuits were met to a sufficient degree so that (11) could be considered valid. The dc millivoltmeter, V_1, can be calibrated in terms of power, once a standard value of power at full-scale deflection of the millivoltmeter has been established. Scale changes can be accomplished by various means, one of which consists of varying R_1 and R_2 as shown in Fig. 2.

A schematic diagram of the wattmeter is shown in Fig. 3. It can be seen that the voltage and current monitoring channels employ amplifiers that are identical, except for the input and output circuit details. The amplifiers employ a type-2N383 transistor in a split-load phase inverter circuit to drive two type-2N383 transistors in a class AB push-pull grounded emitter arrangement. It may be observed that the phase inverter has been modified to present an equal source resistance for the base of each output transistor. The monitoring channels and their input systems were designed to permit the measurement of power levels of 1.0 to 400 milliwatts into a 600-ohm load over a frequency range of 20–20,000 cps. In the voltage monitoring channel, a UTC type-LS-151 bridging transformer has been arranged to present an input resistance of 16,000 ohms

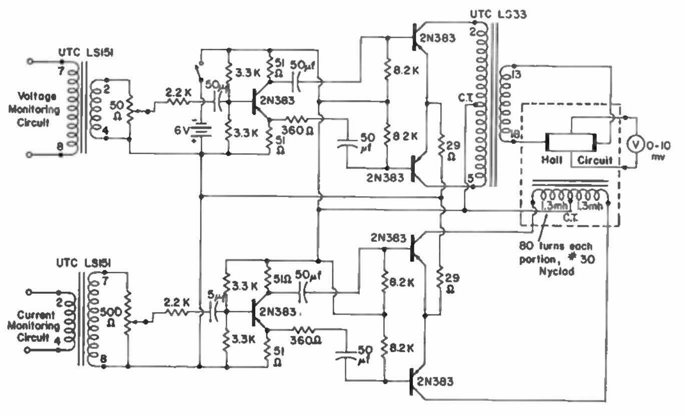

Fig. 3—Audio-frequency wattmeter circuit.

across the circuit under test and to provide a driving signal in-time-phase-with and proportional-to the test voltage for the voltage-monitoring amplifier. The 50-ohm potentiometer terminating the secondary of this transformer provides the means of voltage sensitivity variation during calibration.

In the current monitoring channel, a second UTC type-LS-151 transformer has been connected in reverse order to present an input resistance of about two ohms in series with the circuit under test, and to provide a driving signal in-time-phase-with and proportional-to the test-circuit current for the current-monitoring amplifier. In this case the transformer is terminated in a 500-ohm potentiometer, to provide a means for current sensitivity variation during the calibration of the wattmeter.

The voltage amplifier of the wattmeter employs a UTC output transformer type-LS-33 connected to give a collector-to-collector resistance of 250 ohms when the secondary is terminated in the 1.5-ohm Hall element. The current amplifier is terminated in the excitation coils of the Hall device. The output stages of both amplifiers were provided with 29-ohm resistors in the emitter circuits to prevent collector-current runaway in these stages.

C. Wattmeter Calibration

The wattmeter was calibrated at the 600-ohm level, using a HP-400C vacuum-tube voltmeter as the reference standard. The dc-output voltage produced by the Hall device was indicated on a Seimens and Halske precision millivoltmeter. Wattmeter output readings were obtained at intervals of one volt across a 600-ohm, 1 per cent noninductive resistor, over a range of 15 volts, at selected frequencies throughout the audio range. The results of the calibration tests are shown in Fig. 4 at frequencies of 100, 500 and 6000 cps. The performance of the wattmeter at other frequencies in this range is bounded by the 100- and the 6000-cps curves. It may be seen that the over-all deviation due to frequency variation is less than plus and minus 3.5 per cent, and that the slopes of the calibration curves are not constant. This small change in slope might be attributed to the slightly nonlinear operation of the voltage and current-monitoring amplifiers.

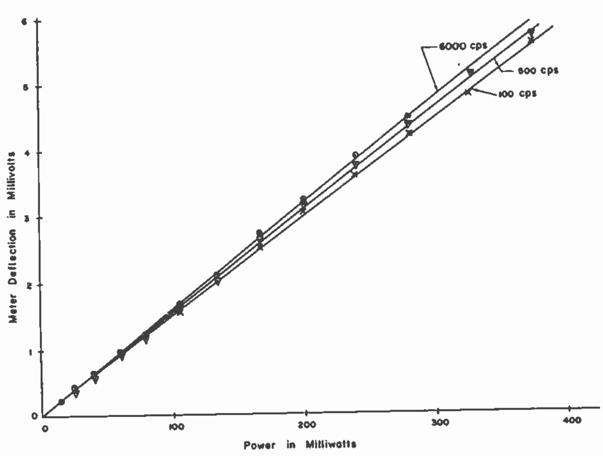

Fig. 4—Calibration curves of the wattmeter.

The final laboratory model is shown in Figs. 5 and 6. The instrument size is relatively small, measuring $10\frac{1}{4}$ inches $\times 5\frac{1}{4}$ inches $\times 7$ inches and weighing approximately 16 lbs. Although no attempt was made to miniaturize the device, it is still fairly compact. The size and weight were primarily determined by the transformers used.

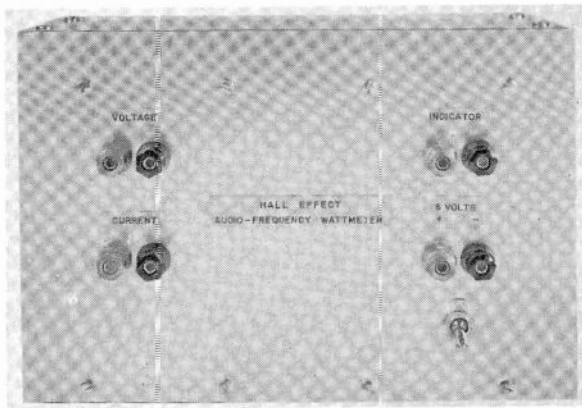

Fig. 5—Hall-effect audio-frequency wattmeter.

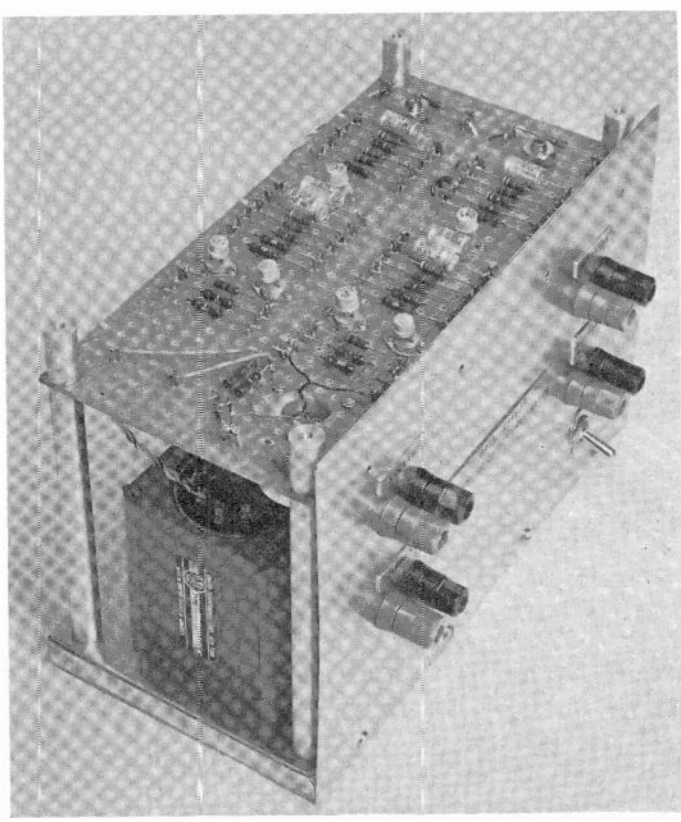

Fig. 6—Hall-effect audio-frequency wattmeter.

III. An X-Band Wattmeter

A. Theory of Operation

If a thin wafer of semiconductor material, as shown in Fig. 7, is placed in an electromagnetic field, a current density $\vec{i}$ proportional to the electric field component of the wave will flow in the y-direction and a magnetic flux density $\vec{b}$ proportional to the magnetic field component will flow in the x-direction, and, as a result of the interaction of the electric and magnetic currents inside the semiconductor, a Hall electric field, $\vec{E}_H$, will be produced in the z-direction. The magnitude of the Hall electric field is given by

$$E_H = R_H ib \tag{12}$$

where R_H is the Hall coefficient.

Now $\vec{i}$ and $\vec{b}$ are harmonic functions of angular frequency ω; therefore, the potential difference across the Hall element will contain a dc component and an ac component of angular frequency 2ω. When $\vec{i}$ and $\vec{b}$ are properly time phased, the dc compnonent of the Hall voltage will be directly proportional to the power flow through the semiconductor wafer. The above principle was used in the design and construction of an experimental wattmeter described herein. The wattmeter was similar to a model reported by Barlow,[1] but differs in the frequency of operation, and in the type of semiconductor material used as a sensing element.

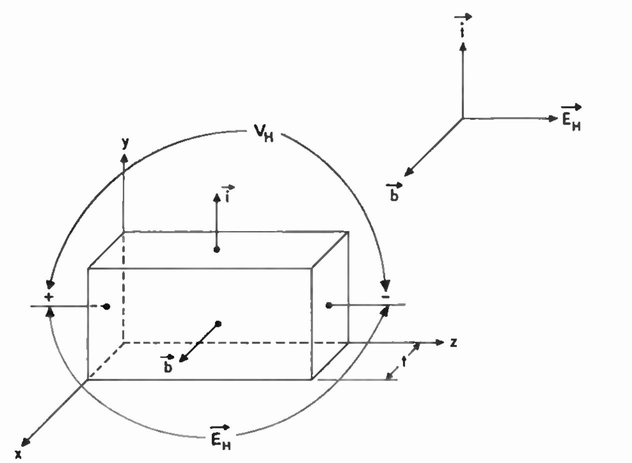

Fig. 7—The generation of the Hall voltage V_H. $\vec{i}$=current density in the material. $\vec{b}$=magnetic flux density through the material. V_H=voltage polarity for negative carriers.

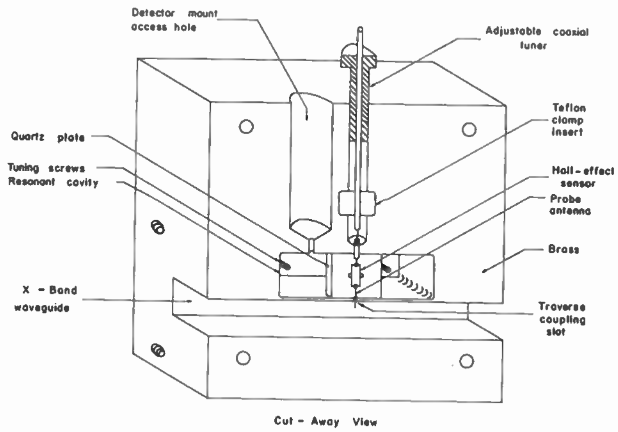

Fig. 8—X-band wattmeter.

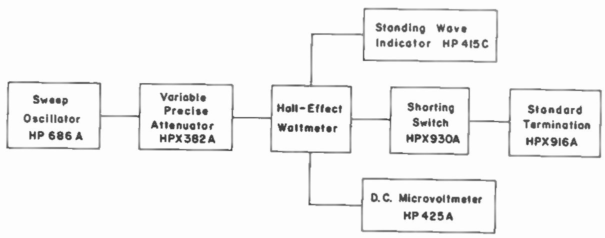

Fig. 9—Wattmeter test system.

B. Construction of the Instrument

The construction details of the wattmeter are shown in Fig. 8. The thin wafer of semiconductor material (indium antimonide) has been inserted in the center of a rectangular cavity placed directly above the main waveguide carrying the signal to be monitored. The cavity, about one wavelength long, is separated from the main waveguide by a brass wall about 0.015 inch thick. Magnetic coupling between the cavity and the waveguide is furnished by a 0.025×0.250 inch rectangular slot as shown in the figure. The semiconductor slab was secured in position at the center of the cavity by means of a small fused quartz plate. An electric probe, 0.010 inch in diameter, projects 0.125 inch into the main waveguide and furnishes the current density $\vec{i}$ for the semiconductor. Magnetic flux density $\vec{b}$ is furnished by the cavity through slot coupling to the main waveguide. Two tuning screws have been provided for cavity tuning. The cavity screw set also permits the field distribution within the cavity to be positioned to give zero electric field at the center of the semiconductor slab; thus the cavity provides magnetic excitation alone to the semiconductor. An adjustable coaxial line has been added to tune and phase the current density $\vec{i}$ through the semiconductor.

C. Experimental Results

The wattmeter was tested using the system given in Fig. 9. An auxiliary electric probe was employed as a detector to aid in the tuning of the cavity to the 9.40 Gc test signal frequency. In the cavity tuning process, the cavity screws were adjusted simultaneously to ensure proper centering of the electric field null within the cavity, as indicated by no Hall-effect output voltage when the electric probe circuit is far off resonance with the signal. After the cavity adjustments were made, the tuning screw of the electric probe was positioned to give maximum Hall output voltage for termination of the wattmeter in a flat load. Next, the matched load was replaced by a short circuit, and the cavity and

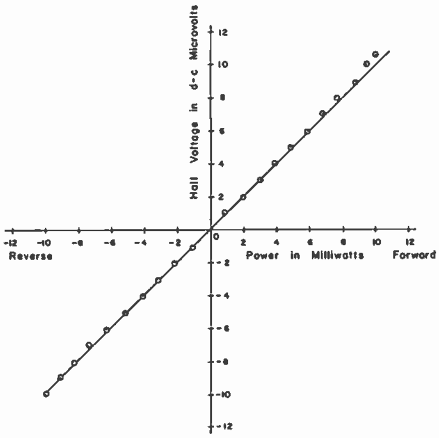

Fig. 10—Calibration curve of Hall-effect X-band wattmeter.

electric probe adjustments were moved slightly to reduce the Hall output voltage to zero. Again, the wattmeter was terminated in a flat load and the screws were adjusted carefully to give maximum Hall output voltage. The tuning process was repeated several times to attain proper phase characteristics of the cavity with respect to the electric probe circuit.

After the wattmeter tuning operation was completed, the performance characteristic of the device was obtained over a power range of zero to ten milliwatts. The results of this test are shown in Fig. 10. In this case a $0.005 \times 0.125 \times 0.375$ inch InSb crystal, having a resistivity of 5×10^{-3} ohm-cm, was used as the sensing element. It can be seen that the wattmeter characteristic is linear, and that the unit produces a dc output of about one microvolt per milliwatt of transmitted power.

On reversal of the direction of power flow, the wattmeter dc output polarity also reversed. The lack of symmetry of the calibration characteristic might be attributed to the small errors in the tuning of the wattmeter and to the residual rectifying action present at the junctions of the wire leads and the semiconductor.

D. Direct Power Measurement with the Hall Effect

A transmission-type wattmeter has been constructed by placing a $0.005 \times 0.167 \times 0.187$ inch InSb crystal directly in an X-band waveguide. The crystal is mounted on a quartz plate and positioned at the center of the waveguide, parallel to the side walls, and near the bottom wall. The dc output leads are brought out of a side wall in a plane parallel to the bottom wall. Preliminary tests indicate that the sensitivity of this wattmeter is about 0.1 microvolt per milliwatt of transmitted power.

E. Comparison Between Cavity-Type and Direct-Type Transmission Wattmeters

Since both the cavity-type and direct-type transmission wattmeters utilizing the Hall effect for power measurement in the X band have been described above, it should be pointed out that there are distinct operational differences. The cavity-type wattmeter described herein is much more sensitive than the one reported by Barlow;[1] however, the device is inherently a narrow-band instrument, requiring accurate tuning of a cavity and and electric probe circuit. A direct-type wattmeter was constructed in an effort to eliminate the cavity-tuning requirement. Initial tests have shown this type wattmeter to be a more broad-band instrument with a sensitivity about one order of magnitude below that of the cavity-type wattmeter.

IV. Conclusions

The instruments described in this paper were developed as part of a program sponsored by the United States Army Signal Research and Development Agency. The experimental models were intended to illustrate the application of the Hall effect in the measurement of power flow and to provide design data for use by the electronic instrument engineer. With appropriate modifications, the wattmeter arrangements could function at other frequency bands and at higher or lower power levels.

[1] L. M. Stephenson and H. E. M. Barlow, "Power measurement of 4 Gc/s by the application of the Hall effect in a semiconductor," *Proc. IEE.*, vol. 106B, pp. 27–30; January, 1959.

The Effects of Track Width in Magnetic Recording*

D. F. ELDRIDGE†, MEMBER, IRE, AND A. BAABA†

Summary—The effects of track width on various performance characteristics have been measured over a wide range of widths. Signal level, noise, and signal-to-noise ratio were determined for track widths from 1.1 mils to 92 mils. The effects of crosstalk, actual recorded track width vs head width, and tape guiding, are described. The experimental data are in good agreement with theory, and no serious practical limitations on the use of very narrow tracks were discovered. High-density audio and pulse recordings were made without difficulty. Digital bit densities of one million per square inch and above are shown to be possible.

Introduction

The purpose of this study was to determine, experimentally, the effects of track width on the performance of a recording system—over a wide range of track widths—and whether or not there were any practical limitations peculiar to very narrow tracks. Theoretically, the signal output per turn of head winding is directly proportional to the track width because signal fluxes from all portions of the track are in phase and add arithmetically. The tape-noise output per turn is theoretically proportional to the square root of track width because the noise signals from different portions of the tape have no phase relationship, and thus add vectorially. Therefore, the signal-to-noise ratio should also be proportional to the square root of track width. On this basis, if a 100-mil-wide track produced a signal-to-noise ratio of 50 db, a 1-mil track should produce 30 db. The possibility of obtaining usable signal-to-noise ratios with extremely narrow tracks is very attractive for numerous applications and, therefore, it was decided to obtain experimental verification of the theory.

Head Construction

A series of heads with eight different track widths was constructed to obtain sufficient data to construct a complete curve and to reduce the importance of variations in any particular head. Actual track widths were 92, 46, 23, 10, 5, 3, 2.1, and 1.1 mils. A record winding was placed on one leg and a reproduce winding on the other, for most heads. Two reproduce-only heads were also built to obtain more voltage output from the narrowest tracks.

A standard tape transport was used for the tests, and an adjustable guide was placed on each side of the head (see Fig. 1). No other modifications of the tape transport were required. To obtain basic crosstalk measurements, a special head-shield can was constructed which permitted micrometer adjustment of the head in the vertical direction (see Fig. 1).

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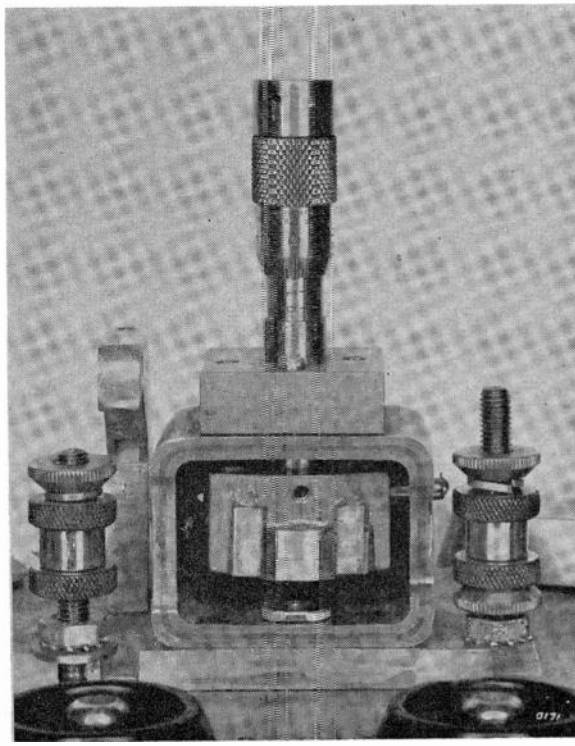


Fig. 1—Head assembly, showing tape guides and micrometer adjustment.

Experimental Data

A. General

The first data were obtained by recording with a full-track head and reproducing on the various narrower-track heads. Two types of signals were used for the level measurements: a 2000-cps sine wave recorded at +8 db at a tape speed of 30 inches per second, and a 2000-cps square wave (NRZ) recorded at saturation at 30 inches per second. Noise measurements were wide-band and were obtained with bias recorded on the tape, and with dc saturation recorded on the tape. No equalization or filtering was used. Both signal and noise were measured peak-to-peak on an oscilloscope, and on an average-reading rms calibrated voltmeter. The peak-noise measurements indicated the value of the occasional "spikes" rather than "average" peak noise.

The output signal levels, in volts per turn referred to the widest track, are plotted vs track width in Fig. 2. The dashed line indicates a 6-db (per octave of track width) slope. Although individual heads cause some deviation, the average slope is very close to the 6 db predicted by theory. The tape-noise output, in volts per turn referred to the 92-mil track, is also plotted vs track width in Fig. 2. The dashed line indicates the

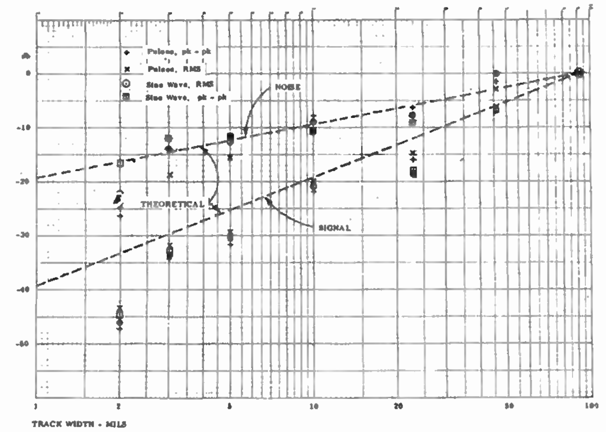


Fig. 2—Signal and noise vs track width, full track record.

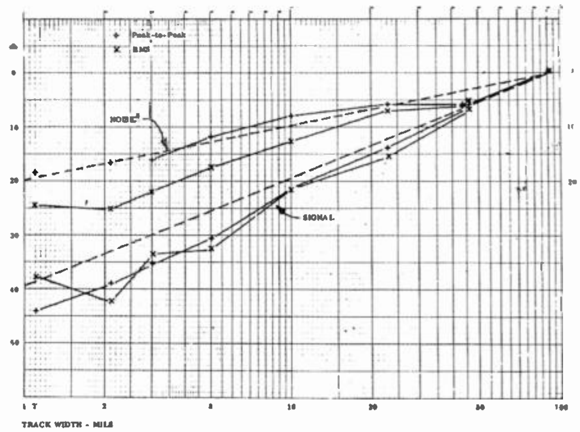


Fig. 4—Signal and noise vs track width, record/reproduce on same head.

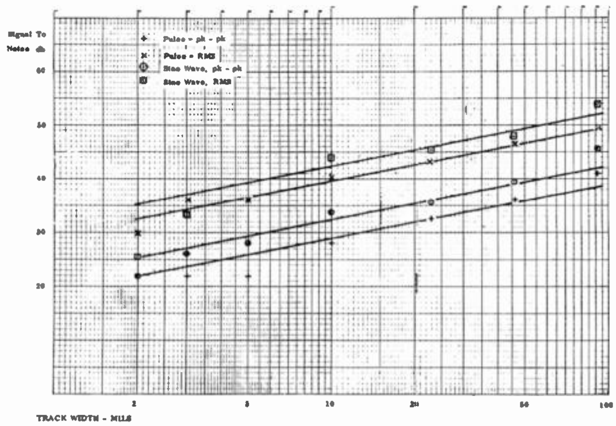


Fig. 3—Signal-to-noise vs track width.

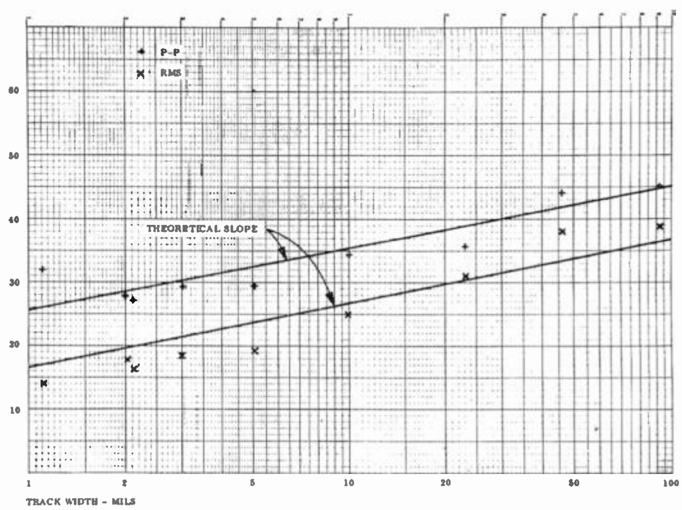


Fig. 5—Signal-to-noise ratio vs track width, record/reproduce on same head.

theoretical 3-db slope. Again, the average slope follows the theoretical curve. The computed signal-to-noise ratios are plotted vs track width in Fig. 3, and also follow the theoretical curve fairly well.

Next, signals were recorded and reproduced on the same head for each track width. As both the square-wave and sine-wave data followed the same slope in the previous data, square-wave recording only was used in these tests for simplicity. Because of the low output from the 2.1-mil head, a new head of the same size and a 1.1-mil head were constructed. To obtain greater accuracy, more turns were wound on these heads, which increased the tape-noise voltage well above the preamplifier noise level. The resulting signal, noise, and signal-to-noise ratios, are plotted vs track width in Figs. 4 and 5. The results are essentially identical with those obtained from the full-track recorded signal.

The signal-to-noise ratio at any track width is a function of the method of measurement. As mentioned earlier, the peak-noise measurements were based on the occasional "spikes," rather than on the average peak value. If the signal-to-noise ratio is calculated for the 1.1-mil head on the basis of the average (nonspike) peak noise, a value of 29 db is found. This value would be the signal-to-noise ratio if there were no imperfections in the oxide to produce spikes, and may be considered to be the basic, or error-free, signal-to-noise ratio. To investigate the character of the spikes, the oscilloscope trigger level was set near the peak of the spikes and the camera shutter opened for ten seconds. The resulting picture, shown in Fig. 6, indicates that the spikes are all nearly identical in shape and duration, and have the same shape as a very short recorded RZ pulse. The imperfections producing these spikes must therefore be less than one mil in length along the tape; if they were longer, the pulses would be longer. A microscopic examination of the oxide surface showed numerous imperfections which were one mil or less in diameter.

Theoretically, the width of the recorded track will be somewhat larger than that of the recording head, because of fringing of the record field. If the oxide is being saturated at a depth of 0.5 mil, saturation should reasonably be expected to occur 0.5 mil from each side of the head. To measure the track width, saturation recordings made with all heads were reproduced on a full-track head. The output voltage in volts per turn referred to the 92-mil track is plotted in Fig. 7. As the amount by which the width of the recorded track exceeds that of the record head should be independent of track width, it should be a larger percentage with the narrower tracks, and is so indicated by the deviation from the ideal curve in Fig. 7. The effective track width was computed from the reproduce voltage for the heads from 10 mils down, using the larger heads to establish the voltage per unit of track width. The calculated track width ranged from 0.7 mil to 1.3 mils larger than the head. The average excess track width is approximately 1.0 mil. It is interesting to note that an infinitely thin record head would produce a track 1.0 mil wide, if 0.5-mil oxide were recorded to saturation. Narrower tracks could be recorded if a thinner oxide were used.

B. Tape Guiding

The use of very narrow tracks is strongly dependent upon the tape-guiding accuracy. It was therefore desirable to obtain some measure of the accuracy of the guides used in these experiments.

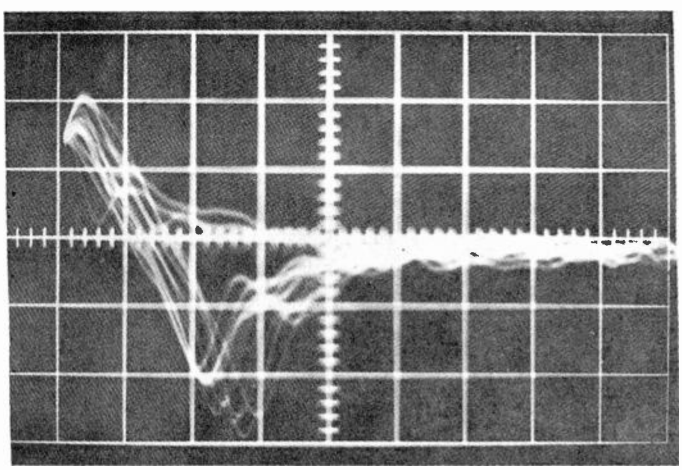


Fig. 6—Noise spikes (10-second exposure).

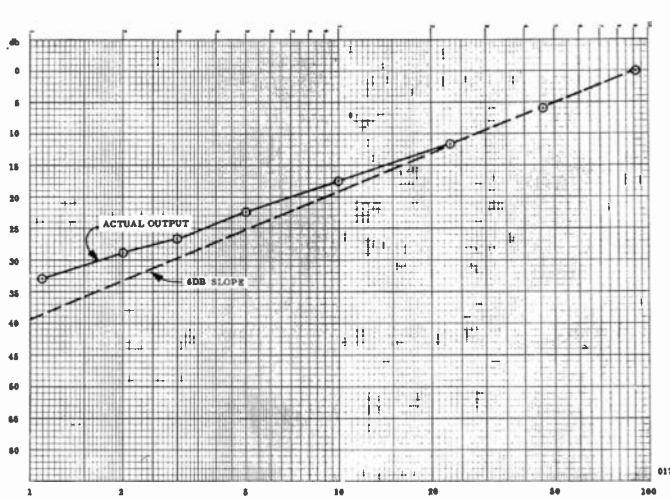


Fig. 7—Output voltage vs track width.

A 60-power microscope with a calibrated reticle was set up to view the edge of the tape as it passed over the head. The total side-to-side motion observed was ± 0.45 mil, consisting of a steady fluctuation of ± 0.3 mil with an additional peak of ± 0.15 mil which occurred every few seconds. The same results were found at 15 inches per second as at 30 inches per second.

Another measure of the guiding accuracy is to observe the amplitude stability of a signal recorded and reproduced on a very narrow track. Accordingly, both 1.0-kc and 15-kc square waves were recorded with the 1.1-mil head. The reproduced signals were photographed during a one-second sweep on the oscilloscope. This was also accomplished with the 92-mil head for reference (see Fig. 8). The amplitude variations are greater in the

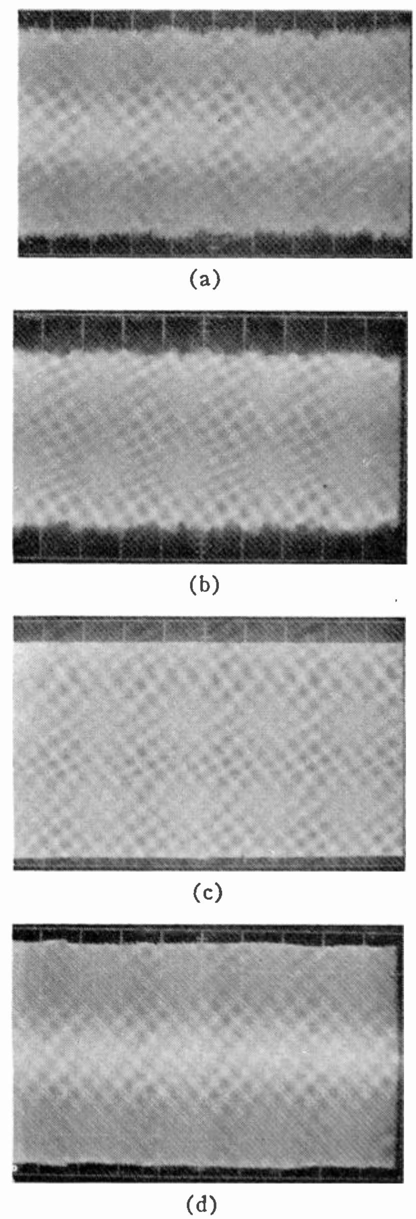


Fig. 8—Amplitude stability. (a) 1.1-mil head, 1 kc at 30 ips, 0.2 mv/cm. (b) 1.1-mil head, 15 kc at 30 inches per second, 0.2 mv/cm. (c) 92-mil head, 1 kc at 30 inches per second, 20 mv/cm. (d) 92-mil head, 15 kc at 30 inches per second, 20 mv/cm.

narrow than in the wide track, but are not sufficiently large to indicate any serious guiding deficiencies.

C. Crosstalk

The use of very narrow tracks does not in itself engender more serious crosstalk problems than occur with relatively wide tracks. Crosstalk problems do arise, however, when two or more tracks are placed very close together, because the reproduce head will read flux from adjacent tracks.

The ratio between the desired and undesired signals is a function of track width, track-to-track spacing, head arrangement, shielding, type of recording, and wavelength. Although a general investigation of all the factors mentioned would probably be worthwhile, it was believed to be outside the scope of this particular study. Data were obtained, however, on the amount of signal read as the head was displaced to the side of a single recorded track.

Sine-wave recordings were made at 1 kc, 3 kc, 15 kc, and 30 kc at 30 ips with a 2-mil wide head. The 2-mil reproduce head was mounted in the special shield can (see Fig. 1) to permit accurate displacement off the track. The reproduce voltage was measured on a wave analyzer to increase the signal-to-noise ratio and thus allow measurement of very small signals. Reproduce voltage is plotted vs head position in Fig. 9. The 1-kc and 3-kc curves follow a slope of approximately 65 d/λ when the head is off to the side of the track (d = displacement, λ = wavelength of recorded signal). One would expect the slope to be greater than the 55 d/λ realized when the tape is spaced outward from the head because of the differences in geometrical arrangement. The 15-kc and 30-kc data were rather difficult to obtain. The original data appeared to be rather inaccurate, and so the experiment was repeated. At the time of the repeated experiment, the 2-mil reproduce head had been modified with the addition of shielding, so the 1.1-mil head was used for reproduction. Recording was done with a 2-mil record head. The results from several lateral scans were averaged and are plotted in Fig. 9.

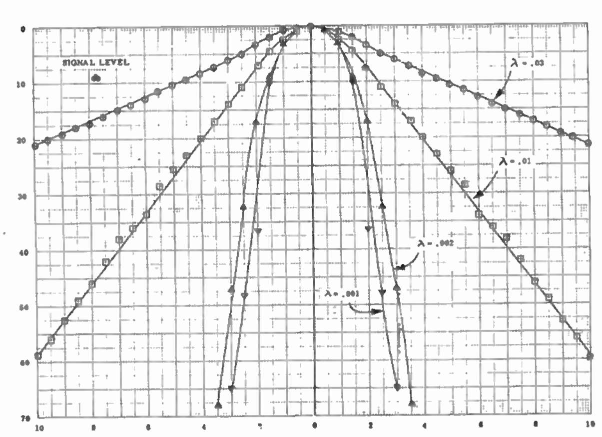
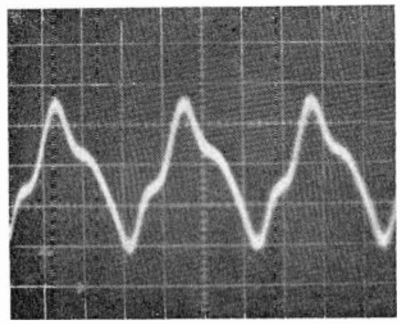
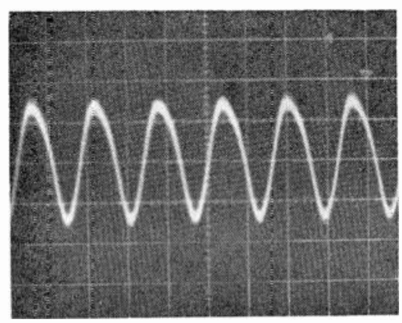


Fig. 9—Basic crosstalk data, 2-mil recorded track, 30 inches per second.

Because of measurement difficulties, no across-the-track scan was made for recorded pulses. NRZ pulses were recorded with the 1.1-mil head in seven adjacent tracks, spaced 2 mils center-to-center. The pulses from the center track were reproduced with the 1.1-mil head, and crosstalk was found to be about 20 db below the signal on each track. Fig. 10 shows the reproduced pulses at 1000 and 2000 pulses per inch (bit densities of 500,000 and 1,000,000 per square inch). A powder pattern of this recording was "developed" with a suspension of pure iron particles and is shown in Fig. 11.



(a)



(b)

Fig. 10—Reproduced pulses (500,000 and 1,000,000 bits per square inch). (a) 1000 bits per inch, 500 tracks per inch, 30 inches per second, 20 μsec/cm, 200 volts/cm, (b) 2000 bits per inch, 500 tracks per inch, 30 inches per second, 20 μsec/cm, 200 volts/cm.

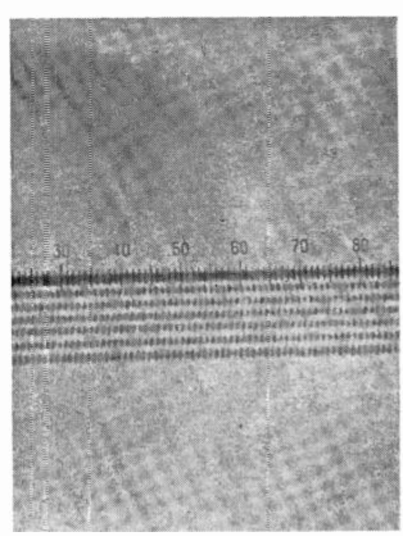


Fig. 11—Powder pattern, 500,000 pulses per square inch (small divisions on scale equal 0.001 inch).

Some Possible Applications of Narrow-Track Recording

There are numerous possible applications of very-narrow-track recording, a few of which will be mentioned herein.

First, there is *monitoring*. There are many types of signals such as voice communications, which do not require extremely high signal-to-noise ratios, but do require the recording of large amounts of information over an extended period of time. In such applications, the usefulness of very narrow tracks is obvious. As an example, a machine could be built using a track spacing of 200 per inch, *i.e.*, 3-mil tracks and 2-mil spaces, with 14-inch reels, and 1-mil Mylar-base tape running at 15/16 inches per second. If ½-inch tape were used, 100 channels of information could be recorded continuously for 24 hours. A reel of 2-inch wide tape could provide a total recording time of over a year.

Second, there is the *recording of high-accuracy analog information*. In direct recording, the accuracy or signal-to-noise ratio is limited by the characteristics of the medium. In time-division recording systems such as FM or PWM, the signal-to-noise ratio is limited by both the motion stability of the tape transport and the characteristics of the medium. The only recording means which can be made insensitive to both tape-transport motion instability and faults of the medium is digital recording. The analog information must be sampled at a rate more than twice per cycle of the highest frequency, and each sample must then be converted to a numerical value. This numerical value is in binary form and may be recorded as on-off information. Therefore, the amplitude characteristics of the medium do not affect the result, unless a complete loss of a digit occurs. Motion instabilities may be circumvented because the original data was sampled at a precise and constant rate. Therefore, even if the samples arrive in the reproduce electronics somewhat irregularly, they may be stored momentarily, and fed out at a constant rate.

The major drawback of this system for either audio or instrumentation application is the large number of pulses per second required. For example, to obtain 0.1 per cent accuracy (60 db SNR) requires 10 bits, or pulses, per sample. At only 2 samples per cycle, 20 pulses will be required for each cycle of the highest frequency desired.

The recording of 20-kc signals then requires a pulse rate of over 400,000 per second. To record this rate with commercially-available equipment, one would have to use 9 or 10 tracks of a 150-inch per second digital tape transport. The bit density on present machines is about 5000 to 7000 pulses per square inch. With very narrow tracks, bit densities up to 100 times as high may be utilized, thus reducing tape speed and allowing more channels on the tape. For instance, at a bit density of 200,000, fourteen channels of 20-kc information could be recorded on 1-inch tape at 30 ips, with 0.1 per cent accuracy. To achieve 0.012 per cent accuracy requires only 30 per cent more capacity per channel.

When high bit densities are used, the problem of dropouts becomes more severe. With analog recording, however, dropouts would not have as serious an effect as with digital recording for bookkeeping purposes. In analog recording there should be, in most cases, a fair degree of correlation between successive numerical values. Hence, if one particular number is missing entirely, it could probably be reconstructed from the values of the numbers which preceded and followed it. Another solution to the dropout problem is to use redundant recording—record the same information on two or more tracks spaced some distance apart. The information may then be recovered from either track and a dropout will rarely occur on both tracks simultaneously. It is also probable that the quality of the oxide coating will continue to improve until tape can be produced which is essentially dropout-free.

A third possible application is in wide-band recording. A stationary head with a very high track density could be used instead of a rotating head. At a bit density of 500,000 per square inch, there would be 1,000,000 bits per lineal inch on 2-inch-wide tape. At 100 inches per second, the bit rate would be 100 megacycles. One advantage of a stationary head system is the possibility of using high tape speeds to achieve wide bandwidths. (Present rotating-head machines cannot easily be speeded up to increase bandwidth.) Also, the wear problem is less severe than with a rotating head.

One means of using high track density for wide-band recording is to commutate the signal into successive heads. Each head would record a pulse, the amplitude of which is proportional to the amplitude of the signal at the sampling instant. The signal frequency would be a maximum of just less than one half the bit rate. More refined encoding techniques could be used at the sacrifice of signal frequency to obtain better signal-to-noise ratios.

Conclusions

Theoretically, the reproduce signal (from a tape recorder) is proportional in amplitude to the track width, and the noise and signal-to-noise ratio are proportional to the square root of track width. These relationships have been verified experimentally for track widths from 0.092 inch to 0.0011 inch. If the data are extrapolated, the track width at which the signal-to-noise ratio becomes unity is approximately one micro-inch. Although it is of little practical value, the extrapolation does indicate that we are presently very far from complete use of the medium. The recorded track width was found to be approximately 0.001 inch larger than the record head, because of fringing. The accuracy of the simple tape guiding system was found satisfactory even with the 0.0011-inch wide head.

Measurements of basic crosstalk indicate that no serious problems exist at short wavelengths, but that considerable shielding may be necessary at long wavelengths. A track density of 500 per inch was recorded and reproduced, with a signal-to-noise ratio of about 20 db, using pulses. Bit densities of five hundred thousand to one million bits per square inch were realized.

The use of very narrow tracks should have many possible applications in future equipment. Among those which appear feasible are high-capacity monitoring, high-accuracy analog recording and wide-band recording with a nonrotating head.

Theory of Motional Feedback*

EGBERT DE BOER†

Summary—**In motional feedback the *mechanical vibrations* of the loudspeaker cone are the source of the feedback voltage. Feedback then improves the over-all response characteristic and reduces the total distortion. The theory of this method is presented here in a simplified, though enlightening, way. The treatment is based on an unorthodox theorem on impedance conversion by feedback.**

I. Introduction

When an amplifier drives a loudspeaker, the application of negative feedback affects only the purely electrical part of the system. Aside from the damping of the mechanical resonance of the loudspeaker, distortion occurring in the electromechanical transduction process is hardly reduced. If, however, the feedback voltage depended on the acoustic sound pressure or on the diaphragm's motion, the situation would be different; then, distortion of the loudspeaker would also be reduced by the feedback. In "motional feedback" the required dependence is partly realized, the feedback voltage depending on the movement of the voice coil. Any movement of the loudspeaker diaphragm in the air gap produces an induction voltage in the voice coil. If this voltage can be separated from the voltage required for driving the loudspeaker, the feedback voltage will solely depend on the diaphragm velocity. We then can expect a partial reduction of transduction distortion. The separation of induction voltage and driving voltage is easier when the former is high, *i.e.* when it is in the region of the fundamental resonance frequency. That is the reason why this type of "motional feedback" is most effective at low frequencies, where, incidentally, it is most needed because of large cone excursions.

Many recent amplifiers (mostly American) have been provided with damping control along much the same principle. The separation of induction and driving voltages is performed with help of a bridge circuit as shown in Fig. 1. Suppose the bridge is in equilibrium for dc, *i.e.* for an impressed dc voltage between terminals 1 and 3, no voltage appears between terminals 2 and 4.

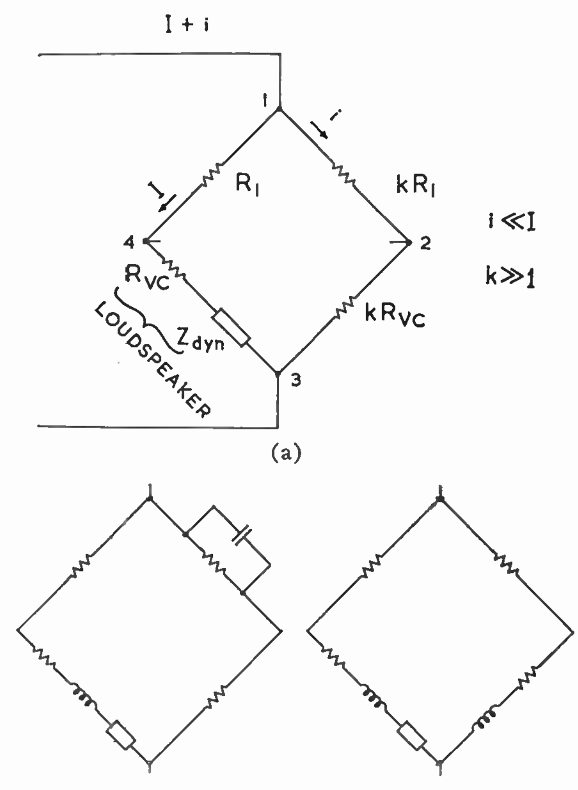


Fig. 1—Bridge circuit used to extract the "motional" voltage.

* Received by the PGA, September 28, 1960.
† Physical Lab., E.N.T. Clinic, Wilhelmina Hospital, Amsterdam, Netherlands.

Suppose also (and this is important) that the bridge stays in equilibrium when it is driven by ac but that the diaphragm is not allowed to move. To accomplish this the voice coil has to be blocked and the bridge may include some additional elements about which more will be said later.

As soon as the diaphragm is allowed to move, a voltage is induced in the voice coil. A fraction of this voltage will then appear between terminals 2 and 4. This output voltage can be used as feedback voltage. The application of voltage feedback tends to make the feedback voltage independent of frequency. Since the feedback voltage is proportional to the diaphragm velocity, the latter quantity is equalized; resonances are damped and a part of the distortion (not including that due to a non-homogeneous magnetic field across the air gap) is reduced.

The discussion of the operation of motional feedback has been very confusing (see Bibliography). The system has been described as a combination of negative current feedback and positive voltage feedback, a viewpoint that served well to obscure all rational thinking. Obviously the authors did not realize that for ultimate performance only the net result matters.

In this article the author proposes to show that the principle can be understood surprisingly well when a few restrictive approximations are made. The derivation will be centered around a fairly unorthodox, but very enlightening, theorem of feedback. This theorem, in which the major emphasis is on net impedances in the circuit, sheds a light on many related phenomena. In fact, it has also been very useful in normal feedback situations. Here we will only apply it to the case of motional feedback. A quick glance will show that with this theorem a theoretical discussion can be presented with quite a low number of formulas.

Instead of dealing with distortion as such, the discussion will be focused on the linear aspects of the system. The latter will be represented in the form of equivalent circuits. The system is thought of as an amplifier loaded with a certain impedance. The proper placing of the (imaginary) connection between these two is most important. The whole system, from its electrical input to its ultimate acoustical output, is considered as an electrical filter. Negative feedback then should improve the response as well as any (small) distortions. This attitude toward the problem will clear up all obscurities at once.

II. Loudspeaker as Filter

This discussion of motional feedback necessarily comprises the subject of electro-acoustic transduction. To facilitate this, we resort to electromechanical analogies. Mechanical forces (or pressures) will be portrayed as electric voltages, mechanical velocities as currents. Mechanical and acoustical reactions to movement are represented by analogous electrical impedances. The response of the mechanical system then follows from the electric analog.

When we assume that the diaphragm is oscillating with a frequency $\omega/2\pi$, a force must be exerted on it which compensates for inertia and stiffness as well as frictional resistance. Hence the main elements of a loudspeaker, effective at low frequencies, are the diaphragm's mass, stiffness and resistance. These main elements are shown in Fig. 2 as an inductance (mass) M, the capacitance (compliance) C and resistance R. In the analogy employed, voltage stands for acoustic sound pressure and electric current for velocity, *i.e.*, ω times the amplitude of sinusoidal motion. The remaining elements in Fig. 2 represent the reaction to the diaphragm, due to the radiated sound field. It is mostly inertial, shown by the inductance L_a. The resistance R_a represents the radiated sound power. As such it is indispensable, but as far as response is concerned it is not very important. We propose to leave R_a out from here on. At resonance, the current (*i.e.*, cone velocity) is restricted mainly by friction as represented by the resistance R.

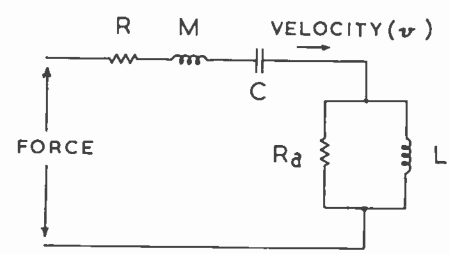


Fig. 2—Electrical analog of the loudspeaker cone.

The reader may have noted that electromagnetic damping has been left out of the discussion. This is because neither the driving amplifier nor the electromagnetic properties of the loudspeaker unit participates in the constraints between purely mechanical forces and velocities. To incorporate it, we will consider the actual transformation of mechanical and electrical quantities in an electrodynamic loudspeaker. At this point it is well to regard the network of Fig. 2 as a symbolism that depicts some relation between forces and velocities, as set by purely mechanical properties. We will now see how this is reflected back on the electric side. Assume the voice coil executes a sinusoidal vibration with an instantaneous velocity $v \cos \omega t$. The velocity is symbolically represented in Fig. 2 by a current having numerically the same value. Electrically, however, the motion produces an induction voltage e in the voice coil. The instantaneous voltage is proportional to v: $e = K_1 \cdot v$. Inversely, an electric current i through the voice coil will cause a force $f = K_2 \cdot i$ to be exerted on the diaphragm.

Hence, a current v in the electrical analog appears electrically as a voltage e, while a pressure p (proportional to the force f) in the analog is actually caused by a current i through the voice coil. The quantities K_1 and K_2 depend on the magnetic field strength and the length of wire embedded in the air gap.

Now, the electrical analog actually represents a restraint imposed on pressure and velocity. When everything is expressed electrically, this restraint appears as a relation between induction voltage e and voice coil current i. Careful consideration shows that the latter constraint can be represented by an electric network that is dual to that of Fig. 2, currents v being replaced by voltages e, and pressures p by currents i. This network is depicted in Fig. 3(a).[1] The relation between e and i imposed by mechanical properties of the loudspeaker is expressible as an impedance. We conveniently call the impedance of Fig. 3(a) the "motional impedance" and denote it by Z_{dyn}.

We have to consider this impedance as being, in effect, in series with the impedance of the (blocked) voice coil. Ultimately the whole voice-coil impedance Z_{vc} is connected to an amplifier, represented in Fig. 3(b) by a voltage source e_0 and an internal resistance R_i.

The network obtained could be analyzed in a straightforward way if it contained a quantity that could serve as the output. As said earlier, we desire to know the ultimate acoustic sound pressure. To be more specific, we want the sound pressure at a specific point in the listening area. Since the spatial distribution of the radiated sound field is determined solely by the geometry of the diaphragm and the loudspeaker's surroundings, we might conveniently take the sound pressure near the diaphragm as the output variable.

The sound field is, at least for low frequencies, nearly spherical, so that a flat frequency characteristic of the sound pressure near the cone implies a flat frequency characteristic at any point in the listening space. In Fig. 2 the sound pressure at the cone is represented by the voltage across R_a. If we leave out this resistance, it can also be taken as $j\omega L_a$ times the current v through L_a. Converted into the terms of Fig. 3 we find that the desired sound pressure is proportional to $j\omega$ times the voltage across Z_{dyn}.

Now when the resistances are lumped together, and radiation resistance and voice-coil inductance are neglected, it is relatively easy to see that the over-all response is of the form

$$F(j\omega) = \frac{K \cdot (j\omega/\omega_0)^2}{1 + a(j\omega/\omega_0) + (j\omega/\omega_0)^2} \cdot$$

Here ω_0 is the resonance radial velocity determined by the total inductance and capacitance. The constant K involves many factors among which is the sensitivity as determined by the efficiency. The constant a determines the form of the response curve. A high value of a means a high damping. Some response curves are given in Fig. 4. Two extreme cases are easily visualized. Suppose first that the amplifier has a very large internal resistance. Then the only damping of the loudspeaker arises from friction, the resonance is very pronounced, a is low. On the other hand, an amplifier with a low R_i, coupled to a loudspeaker with very low R_{vc}, behaves completely differently. The value of a is high, and damping may be so large that the response drops below a frequency larger than the resonant frequency.

In the above discussion the required output variable was, apart from a factor $j\omega$, the voltage e developed across Z_{dyn}. Since Z_{dyn} comprises the energy-storage elements that cause the resonance, it is evident that the total resistance presented to Z_{dyn} determines the ef-

[1] We could have started, of course, by regarding the network of Fig. 3 as the electrical analog of the acoustic system. This is, however, somewhat less enlightening than the approach used here.

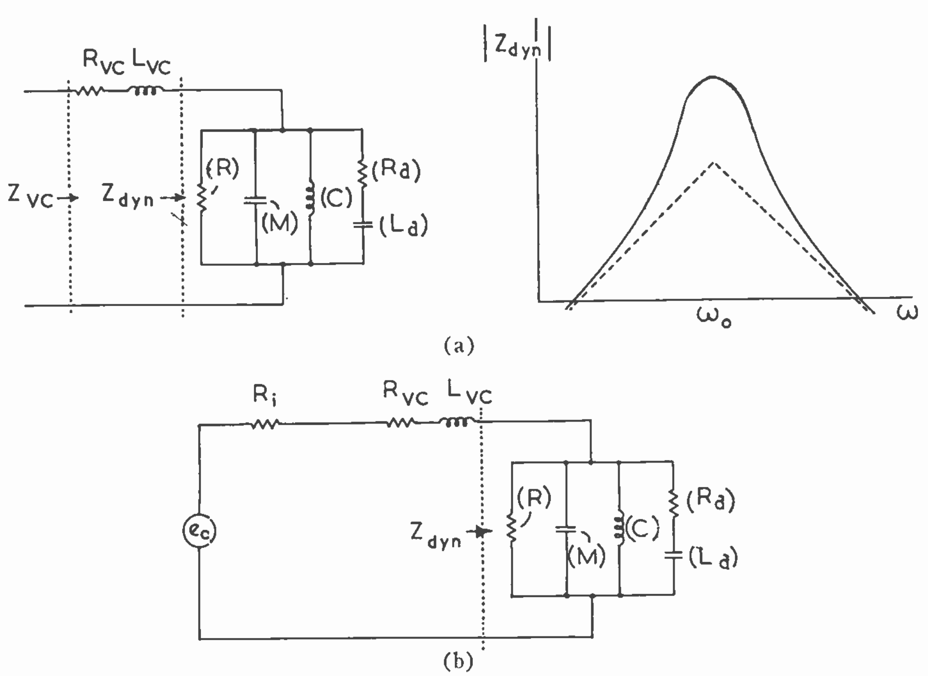


Fig. 3—Electrical impedance of the loudspeaker.

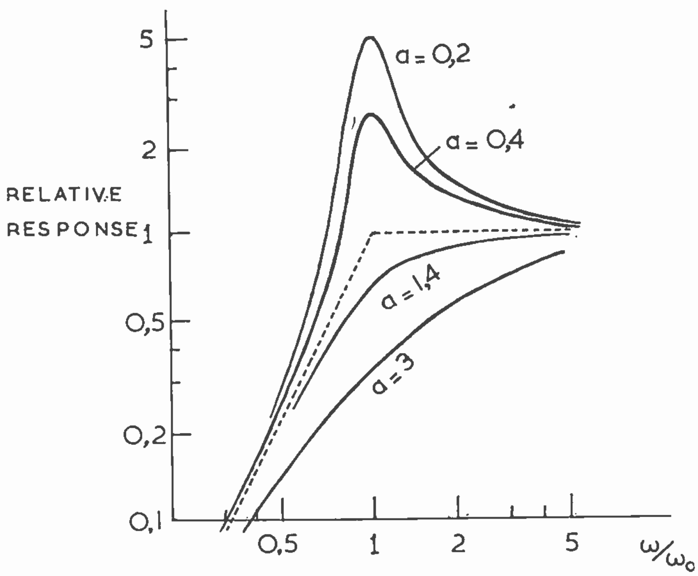


Fig. 4—Typical response characteristics.

fective damping. We remember that the radiation resistance R_a and the voice-coil inductance have been neglected. If we adhere to this description, we see that it is most advantageous, though somewhat unorthodox, to regard the whole system as an "amplifier" that has only Z_{dyn} as a load impedance. The driving "amplifier" now includes the voice-coil resistance R_{vc}. This has the additional advantage that motional feedback is now equivalent to ordinary voltage feedback, assuming the feedback voltage to be proportional to the output voltage as developed across Z_{dyn}. It is well known that negative feedback changes the internal impedance of the amplifier with respect to the load, as quite distinct from changing the level of amplification.

Hence, motional feedback, which is in our system equivalent to voltage feedback, will cause the damping of the load to be increased. We remember that the load impedance Z_{dyn} has the characteristics of a parallel resonant circuit. Its impedance level is sketched as a function of frequency in Fig. 3(a). This function will now be flattened by feedback. The effect will be most pronounced around the frequency of resonance. In the ultimate acoustic response, which involves a factor $j\omega$, this brings about an increased value of a, and consequently a less pronounced resonance peak. It is remarkable that the over-all level of amplification (at larger frequencies) is not affected.[2] Too-large feedback will cause the lower frequencies to be extremely reduced, without leaving a trace of the former resonance. This is an often discussed phenomenon, which is easily explained by Fig. 4.

[2] This should be well distinguished from the behavior of a normal feedback amplifier. Suppose an amplifier has an over-all response like one of the curves of Fig. 4. If voltage feedback is applied, the level of amplification is decreased, the cut-off frequency is lowered, and the resonance will become *more* pronounced. That motional feedback produces an entirely different effect is due to the fact that the feedback voltage is essentially obtained from a *band-pass* system, the ultimate acoustic response requiring the factor ω.

III. Basic Theory

In the Introduction, the bridge of Fig. 1 has been described as the method to extract the voltage e across Z_{dyn}. If the bridge is properly adjusted, the output voltage is under all circumstances proportional to the voltage e. We saw that the electrical environment of Z_{dyn} is the all-controlling factor as regards response. This is stressed by considering the amplifier, *including bridge and voice-coil resistance*, as a driving system, loaded only by the motional impedance Z_{dyn}.

In a way the diaphragm has become the final link in the amplifier chain. We can now represent the amplifying system as a Thevenin source, a voltage source e_0 with an internal resistance R_i. Note that R_i includes not only the internal resistance of the original amplifier, but also parts of the bridge and finally the voice-coil resistance.

In order to simplify the discussion we first want to consider one extreme form of the system, namely, that in which the internal resistance is so large that it can be taken as infinite. The representation of the amplifier is then a current source i, loaded by Z_{dyn} [Fig. 5(a)]. To justify this representation, two conditions have to be fulfilled. The first is that the amplifier itself behave as a current source. The second is that the right-hand resistors in the bridge of Fig. 1 be so large that their presence can be neglected under all circumstances. We choose this extreme form as the one which deserves motional feedback most urgently. This is not as unrealistic as it seems since the driving system here considered includes the voice-coil resistance. Hence it amounts to considering an amplifier that damps the loudspeaker hardly at all, because, *e.g.*, the loudspeaker has a low efficiency. The treatment of motional feedback will now become surprisingly simple, as we will see presently. We recall that a "current-source" amplifier drives an impedance Z_{dyn}. The feedback voltage is taken directly from the output terminals (*i.e.*, across Z_{dyn}). The reduction in feedback voltage, brought about by the bridge, is neglected. The "current-source" amplifier is described by its transconductance g, the output current for one volt of input voltage:

$$i = g \cdot e_i. \tag{1}$$

We will now prove a theorem that is basic in the discussion.

Theorem: A current-source amplifier, with a transconductance g, will, after the application of full voltage feedback, behave as a current-source amplifier with the same transconductance g, but with an internal resistance 1/g in shunt.

The proof is extremely simple. With feedback, the effective input voltage e_i is the difference of the actual input voltage e_{in} and the output voltage e_{out}:

$$e_i = e_{in} - e_{out}. \tag{2}$$

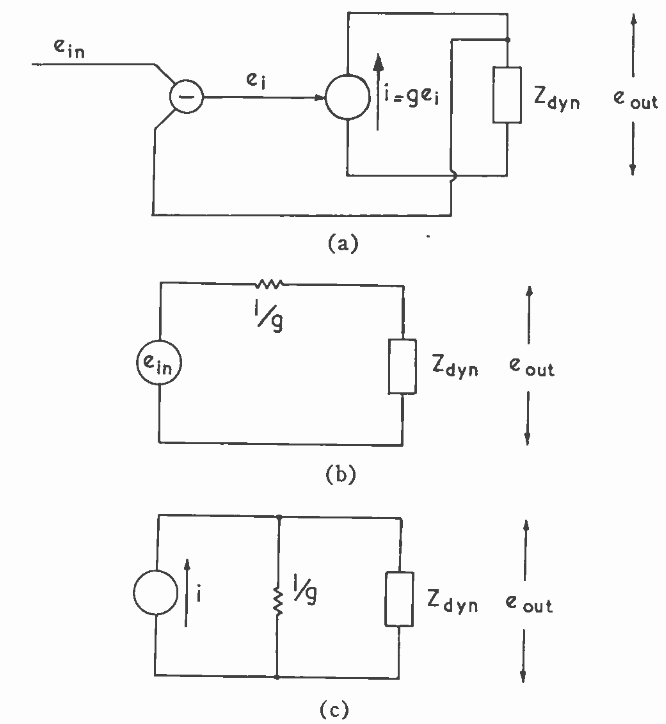


Fig. 5—Illustrating basic feedback theorem.

Combination of (1) and (2) gives:

$$e_{\text{out}} = e_{\text{in}} - \frac{1}{g} i.$$

This is represented by the Thevenin circuit shown in Fig. 5(b) which is easily converted into the Norton equivalent shown in Fig. 5(c). It is hardly surprising to find that the transconductance of the "amplifier" has not been changed, since the short-circuit output current necessarily stays the same; the internal resistance, however, has from being infinite become finite.

This theorem has a central place in the discussion of motional feedback. In the worst possible case, where the amplifier cannot exert influence on damping at all, it is quite easy to find the required amount of feedback to give critical damping ($a=\sqrt{2}$). Better, the required minimum sensitivity of the amplifier is found. Needless to say, the inevitable losses in the feedback path have to be compensated for by a corresponding increase in g.

The reduction of distortion is the same as the reduction of over-all amplification. The latter is found by considering the ratio in which the total load to the current source is changed. Easily we find as the feedback factor:

$$b = 1 + gZ_{\text{dyn}}. \tag{3}$$

The distortion reduction is effective only in the frequency region, where Z_{dyn} reaches appreciable values, *i.e.*, around the resonance frequency. The feedback disappears where $g \cdot Z_{\text{dyn}}$ approaches zero, where we also cannot expect any reduction of distortion.

IV. Refinements

The above derivation is valid under two restrictive assumptions. The first of these is the requirement that the right-hand branches of the bridge (Fig. 1) have infinite resistance. The second condition is that the amplifier behave as a pure current source with respect to the load. In order to gain insight into more practical cases, these restrictions have to be removed. We will show that the two restrictions are equivalent and that their removal consequently can be carried out in one step.

The first restriction is the least important. Referring to Fig. 1, it amounts to putting $k \gg 1$. This is equivalent to minimizing the output power lost in the shunt branches of the bridge. As such, the restriction can be easily satisfied. More important, however, is the restriction on the amplifier, which requires its action as a current source. A real amplifier has a finite internal resistance R_i, which under normal circumstances effects considerable control over the loudspeaker's fundamental resonance. Fig. 6(a) shows such an amplifier, with its associated circuitry. For the sake of uniformity it is represented as a current source $i = ge_i$, with the internal resistance R_i in parallel. Figs. 6(b) and 6(c) show how R_i can be absorbed by the right-hand branches of the bridge.

We see again a pure current-source amplifier that drives a load via a bridge circuit, but the condition $k \gg 1$ on the latter can no longer be satisfied. Due to the inclusion of R_i, k may eventually become smaller than unity. The bridge is still balanced, so that the voltage between terminals 2 and 4 is proportional to the voltage across Z_{dyn}. However, the full current i is no longer flowing through Z_{dyn}; an appreciable part is taken up by the right-hand branches of the bridge.

The problem can again be treated by the technique of the equivalent circuit with respect to the impedance Z_{dyn} as the load. Motional feedback is then again equivalent to voltage feedback. The equivalent circuit with respect to Z_{dyn} can be represented as its Thevenin or its Norton equivalent. We choose the latter representation. The internal resistance with respect to the load Z_{dyn} is

$$R_i' = (k' + 1)(R_1 + R_{vc}), \tag{4}$$

in which k' now includes the effect of the amplifier's original internal resistance R_i.

The short-circuit current (flowing when $Z_{\text{dyn}} = 0$) is,

$$i' = i \cdot \frac{k'}{k' + 1}.$$

Finally we get the circuit of Fig. 7(a). The current source has acquired a new transconductance g':

$$g' = g \cdot \frac{k'}{k' + 1} \cdot$$

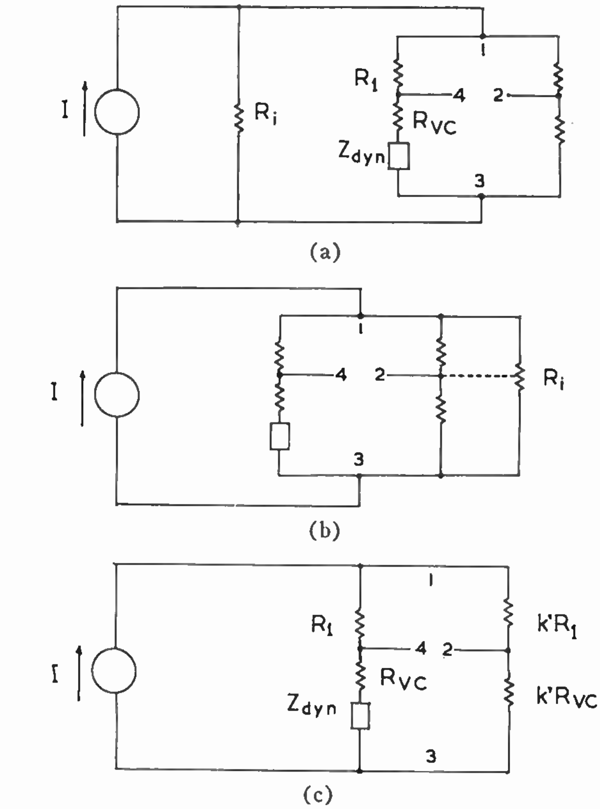


Fig. 6—Absorption of internal resistance of amplifier.

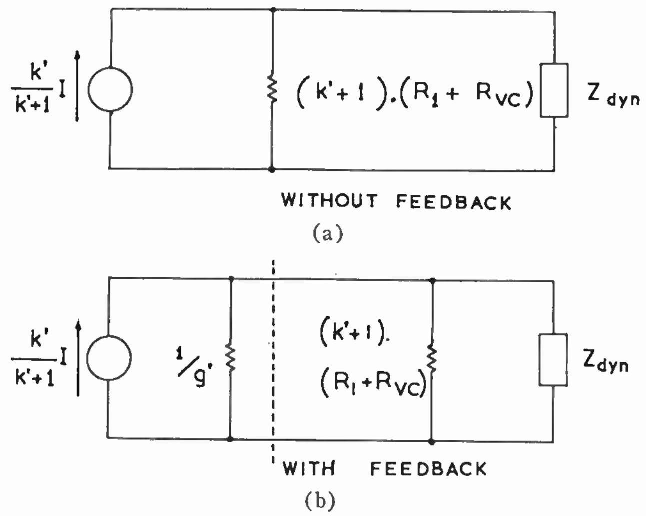


Fig. 7—Net effect of motional feedback.

To this circuit we can apply our fundamental theorem. We consider the parallel combination of Z_{dyn} and R_i' as the load to the current source i'. Voltage feedback then results in the appearance of a shunt resistance $1/g'$. [See Fig. 7(b)]. Note that the shunt resistance is larger than before, since $g' < g$. We finally obtain in parallel with Z_{dyn} the following resistances:

1) R_i', comprising the voice-coil resistance and the combination of the bridge resistance R_1 and the internal resistance R_i of the original amplifier;

2) the resistance $1/g'$ appearing as the result of the feedback.

The net effect is that the original damping resistance R_i' by the application of feedback is reduced by the factor

$$\frac{1}{\dfrac{1/g'}{R_i' + 1/g'}} = 1 + g\cdot k'(R_1 + R_{vc})$$

with

$$k' = k\cdot\frac{R_i}{R_i + k(R_1 + R_{vc})}.$$

The derivation shows that the treatment of the more general case can be reduced to the simple case. The process includes repeated conversions into Norton-type equivalent circuits. The final result is obtained by applying a basic theorem on the conversion of impedances by the application of feedback. Due to its exceptional simplicity, this theorem allows one to obtain considerable insight into a seemingly complicated situation. Incidentally, it does the same in more conventional applications of feedback.

At this point we conclude the theoretical treatment of motional feedback. Assuming the loudspeaker constants known, one can, with help of the discussion above, obtain the required sensitivity g of the amplifier to be used. In the concluding section some remarks are made that may aid in matters of practical design.

V. Miscellaneous Remarks

1) One should not forget that the voltage developed between terminals 2 and 4 of the bridge is actually a fraction of the voltage across Z_{dyn}. The attenuation is given by

$$e_4 - e_2 = \frac{R_1}{R_1 + R_{vc}}\cdot e_{dyn},$$

where e_{dyn} is the voltage across Z_{dyn}. The attenuation has to be compensated for by a corresponding increase in the over-all sensitivity as expressed by the required value of g.

2) Given the fact that motional feedback operates only around the resonance frequency, one should apply normal feedback to the amplifier in order to reduce its distortion. In general, voltage feedback will be more advantageous, since it aids in the damping. Under (6) we will describe a method of combining over-all voltage feedback, operative at all frequencies, with motional feedback.

3) To measure Z_{dyn}, the bridge of Fig. 1 is to be built and adjusted. It should be driven in such a way that the current through the loudspeaker is constant, independent of frequency. The voltage developed across Z_{dyn} can be derived from the terminals 2 and 4. The loudspeaker should be placed in a closed box, provided with absorbing walls to reduce standing waves.

4) In these measurements on Z_{dyn}, the influence of the voice-coil inductance is considerable. For the *measurements*, this difficulty can be decreased by padding of the right-hand bridge elements so as to improve the balance of the bridge in this respect.

In Fig. 1 some possibilities are shown. One should not forget that the voice-coil inductance and its loss factor are not independent of frequency. It might be necessary to increase the complexity. In the *actual application* of the bridge, this method cannot be applied. This is because the amplifier's internal resistance R_i comes in parallel with the bridge, thus spoiling the correction. A better method is to connect a corrective impedance in parallel with the loudspeaker.

5) When the loudspeaker is part of a multichannel system, the bridge can be placed either in front of the crossover filter or in the low-frequency channel. Perhaps the first method is the most convenient one. All loudspeaker inductances have to be balanced out in order to make the total impedance constant except for the "motional" voltage developed by the low-frequency channel. This all puts a considerable strain on the accuracy and constancy of the filter elements.

6) An important problem arises in connection with the unbalanced bridge. From the starting point on, it has been assumed that the bridge is in dc equilibrium. If this is not the case, two possibilities exist:

a) A part r of R_{vc} is effective in series with Z_{dyn} (Fig. 8).
b) The bridge is unbalanced the other way; the resistance r is then to be considered negative. Since the system is now likely to oscillate, this situation ought to be avoided.

The first case, exemplified by Fig. 8, provides a useful possibility for combining normal with motional feedback. In the source-load configuration studied above (Fig. 5) the resistance r is to be considered in series with Z_{dyn}. The feedback factor, expressed in 3) then does not go to zero at frequencies where Z_{dyn} goes to zero. Hence, voltage feedback is present over the whole frequency range. In the resonance region this is partly effective as motional feedback. In many cases this is the only method for combining motional feedback with voltage feedback around the output transformer. This is due to the fact that the secondary winding of the output transformer is floating with respect to earth.

7) The conventional method of providing variable damping by unbalancing the bridge is not correct. A better way is to provide a pure voltage feedback from the right-hand section of the bridge, and to use a potentiometer to vary the feedback toward motional feedback. Fig. 9 shows the circuit to be used.

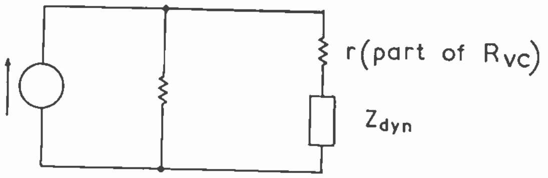


Fig. 8—Unbalanced bridge.

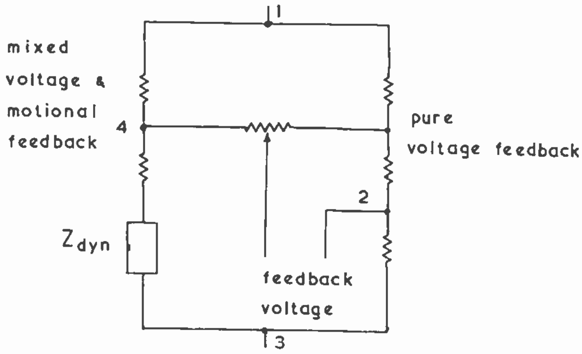


Fig. 9—Method of varying the damping.

8) A major question involves the stability. If the amplifier cuts off at a frequency above the loudspeaker's resonance frequency, motional feedback has little effect, except on the dangerous possibility of oscillation. That is because the amplifier then has a phase shift of about 90° around the resonance point. With feedback, the risk os oscillation is considerably increased. Damping cannot be expected, since the amplifier behaves as a reactance after the application of feedback (g is nearly imaginary!) and a reactance cannot absorb power. This stresses the requirement that the amplifier has to have excellent low-frequency performance.

9) Inspection of Fig. 4 shows that the distortion reduction to be expected from motional feedback is only effective around the resonance frequency. If the resonance is very pronounced, considerable feedback results. The advent of modern, highly-damped loudspeakers will cause the effect of motional feedback to be deceptively small. This constitutes a basic limitation of the principle of motional feedback. A gain in this respect can be expected when the motional feedback is deliberately made too large. It is then necessary, of course, to correct in the preamplifier for the resulting dropping response.

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A Stereophonic Transistor Preamplifier*

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Summary—**Designed to drive two transistor-power amplifiers previously described, this preamplifier contains all the controls necessary for stereophonic and monophonic programs. The first two of the three stages form a complementary circuit readily adaptable for the various input sources, which provides the necessary gain and equalization for low-level phonograph and tape inputs, guard voltages to reduce the capacitive effects of the input cables, and impedance transformation for high-level inputs from tuners and other tube circuitry. Loudness control is achieved by the combined effect of variable negative feedback and a novel tone-control arrangement. The frequency range, including the power amplifiers, extends from below 6 cycles to above 50 kilocycles per second.**

I. Introduction

THE stereophonic transistor preamplifier to be described here was primarily designed to supplement the negative-impedance power-amplifier which was the subject of a previous paper.[1]

Fig. 1 shows a complete stereophonic system, using two of the twelve-watt amplifiers presented there. As apparent from this block diagram, the preamplifier consists essentially of two input amplifiers with the necessary switching arrangement to select the various input sources, stereo controls for mode of operation and balance adjustment, loudness and tone-control circuitry, and finally, two output stages designed to drive the power amplifiers. These various portions will now be discussed.

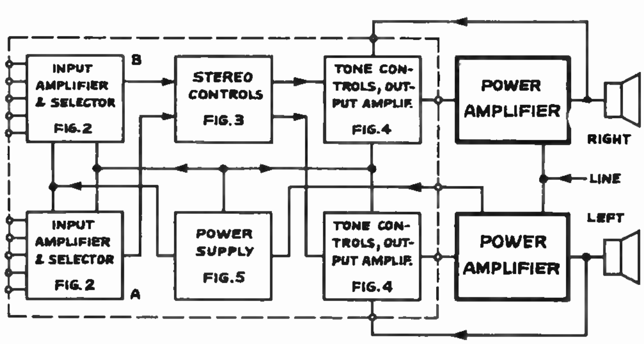

Fig. 1—Stereophonic amplifier system.

II. The Input Amplifiers

The input amplifier is a flexible two-transistor arrangement whose performance readily adapts to the different requirements for the various input sources.

* Received by the PGA, November 10, 1960.

† Hughes Semiconductor Div., Newport Beach, Calif.

[1] W. Steiger, "Transistor power amplifiers with negative output impedance," IRE Trans. on Audio, vol. AU-8, pp. 195–201; November–December, 1960.

The signals from magnetic phonograph cartridges and tapeheads require low-level amplification and playback equalization. Most magnetic cartridges have a fairly high inductance and consequently demand a relatively high load impedance at higher frequencies. A resistive load in the order of 50 to 100 K ohms is in general satisfactory, but it is equally important that capacitive loading (primarily by the cable) is kept low enough so that no resonance occurs at audible frequencies. The input amplifier was therefore designed to furnish guard voltages for the phonograph and tape cables.

AM-FM tuners and television sets on the other hand deliver high-level signals. In this case, the input amplifier functions merely as an impedance converter, raising the low impedance at the stereo controls to the order of 100 K ohms. Although tuners with cathode follower output could drive the amplifier directly, the impedance conversion, needing no additional components, has several advantages: replacement of the tuner output capacitor by a large electrolytic value is not necessary; the amplifier can be connected directly to the sound discriminator of a TV set, or to most vacuum tube circuitry; the arrangement of the program selector switch becomes comparatively simple.

The circuit of the input amplifier uses two directly coupled complementary transistors and is shown in Fig. 2. The component values for the complete preamplifier are given in Table I.

In the tape and phono positions, negative feedback by R_{10} from the collector of Q_2 to the emitter of Q_1, restricted to dc by C_2, stabilizes the operating points of both transistors. Playback equalization and the necessary rise of input impedance at higher frequencies are obtained by separate ac feedback networks (R_8, R_{11}, C_3, C_4 for phono, and R_9, R_{12}, C_5 for tape).

Under the assumptions that $h_{re1}=h_{re2}=0$, $h_{fe1}\gg 1$, and $1/h_{oe1}\gg h_{ie2}+Z_1$, it can be shown that the input amplifier in its low-level mode has the following characteristics.

Voltage gain:

$$v = \left[\sigma + \frac{h_{ib1} + Z_1}{h_{fe2}(Z_2 \| Z_L)}\right]^{-1}.$$

Input impedance:

$$Z_{\text{in}} = h_{fe1}[h_{ib1} + Z_1 + \sigma h_{fe2}(Z_2 \| Z_L)].$$

Output impedance:

$$Z_{\text{out}} = Z_2 \left\| \frac{h_{fe1}(h_{ib1} + Z_1) + Z_G}{\sigma h_{fe1} h_{fe2}}\right.,$$

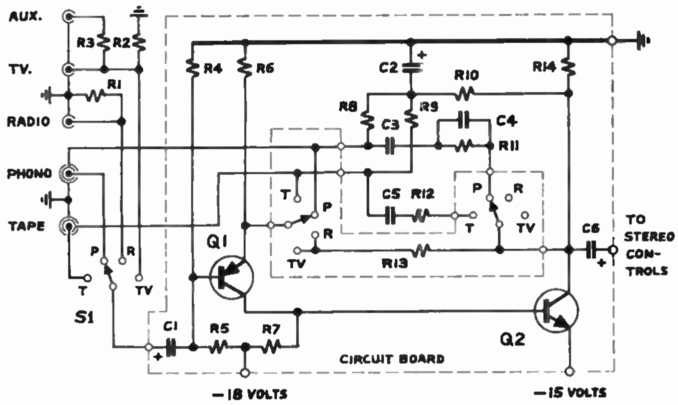

Fig. 2—Input amplifier.

TABLE I

PARTS LIST FOR STEREOPHONIC PREAMPLIFIER

R_1, R_2, R_3		470	K
R_4, R_5		220	K
R_6, R_{10}		15	K
R_7		6.8	K
R_8, R_9, R_{30}		47	Ohm
R_{11}, R_{12}, R_{18}, R_{20}, R_{22}, R_{23}		10	K
R_{13}, R_{28}, R_{29}, R_{33}		1	K
R_{14}		3.3	K
R_{15}, R_{16}, R_{21}, R_{32}		2.2	K
R_{17}, R_{25}, R_{27}, R_{34}		4.7	K
R_{19}		22	K
R_{24}		100	K
R_{26}, R_{31}		100	Ohm
P_1, P_3	Linear	20	K
P_2	Dual, see text	2×25	K
P_4	Logarithmic	50	K
C_1		5	μF, 12 V
C_2		100	μF, 12 V
C_3, C_8		47	nF
C_4, C_5		10	nF
C_6, C_{12}		10	μF, 12 V
C_7		4.7	nF
C_9		22	nF
C_{10}		0.47	μF
C_{11}, C_{13}		0.22	μF
C_{14}		0.1	μF
C_{15}		100	μF, 50 V
C_{16}, C_{17}, C_{18}		250	μF, 25 V
Q_1, Q_3		*P-N-P* transistor, 2N324, 2N508, or similar high-gain unit.	
Q_2		*N-P-N* transistor, 2N169, 2N1304, or similar high-gain unit.	
D		Hughes silicon diode 1N859	
S_1		Rotary switch 6×4 positions	
S_2		Rotary switch 4×9 positions	

where

σ = ac negative feedback ratio, determined by the feedback network,

Z_1 = Circuit impedance seen by the emitter of Q_1,

Z_2 = Impedance resulting from the parallel connection of R_{10}, R_{14}, h_{oe2}, and the feedback network,

Z_G = Source impedance,

Z_L = Load impedance.

The voltage gain v of the two stages is chosen to be 200 at midfrequencies for phono and at high frequencies for tape. For pickups with comparatively high output, more negative feedback may be applied by increasing R_8. The driving voltages for the inner shields of the cables are derived from the emitter of Q_1.

In the radio and TV positions, the small resistor R_{13} provides full feedback for all frequencies down to dc, which results in a voltage gain of approximately unity. Due to the very high current gain of the combination, however, the input impedance is essentially determined by the bias resistors R_4 and R_5 of Q_1.

III. Stereo Controls

The signals from the two input amplifiers are then fed to the stereo control arrangement shown in Fig. 3, which consists of the mode switch S_2 and the balance potentiometer P_1.

In the monaural positions either channel A or B or the sum of the two can be routed to the left or the right channel or to both in any desired proportion.

For stereophonic operation, S_2 has six positions, three each for normal and reverse stereo. The reverse positions simply interchange the two input channels. A choice is possible between full stereo, two-thirds and one-third stereo. This allows correction for exaggerated recordings or too widely-spaced speakers. In these cases, the stereophonic effect is reduced by cross-feeding some of each channel's signal into the other channel.

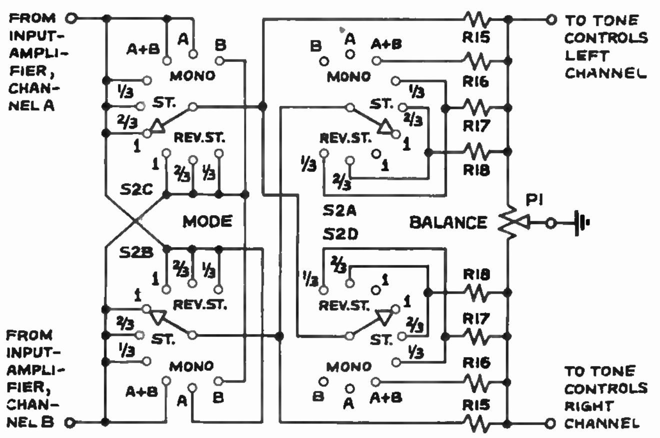

Fig. 3—Stereo controls.

IV. Tone and Loudness Controls

At midfrequencies, the signal transfer through the tone control circuit (Fig. 4) is essentially determined by the ratio of R_{19} and R_{21}. Treble boost or cut is adjustable by potentiometer P_3.

In the arrangement presented here, the low frequencies have two ways to reach the base of transistor Q_3, namely through R_{20}, R_{21} and P_2, and, bypassing the loudness control, through R_{20}, P_4 and R_{22}. This permits more effective use of the available bass frequencies. The second path gains in relative influence as P_2 is turned down and thus helps to make it a loudness control rather than a volume control.

However, the loudness control characteristics are primarily achieved by frequency dependent negative feedback from the power-amplifier output. This feedback is determined by the network consisting of C_{13}, C_{14}, R_{27} through R_{30}, and by the position of P_2 whose lower portion is part of the feedback loop. If the loudness control is set to very low values, maximum feedback occurs, chosen approximately 20 db at midfrequencies and 10 db at high frequencies, but not extending to low frequencies. As the loudness control is turned up, this equalization contour is gradually flattening out.

By interrupting the feedback, a switch, labeled presence, permits accentuation of the midfrequencies if desired.

It is true that this feedback, being derived from the output of the power amplifier, reduces the negative output impedance. However, since the low frequencies are not fed back, the negative impedance remains unaffected in the range of its importance, the region about the speaker resonance frequency.

It is desirable that the loudness-control potentiometers are only slightly logarithmic, since the variable feedback itself causes a curved characteristic. Were it not for the feature that the feedback can be switched off, a linear taper would be satisfactory.

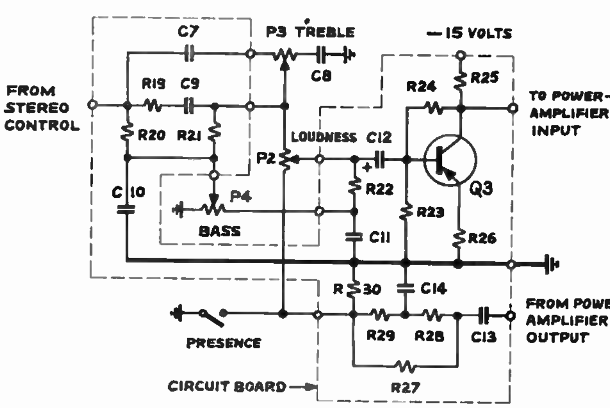

Fig. 4—Tone controls and output amplifier.

V. Output Amplifiers and Power Supply

Transistor Q_3 (Fig. 4) is arranged in a straightforward stage with moderate gain, designed to drive the power amplifier. Some negative feedback stabilizes the operating point and improves linearity.

The input impedance of the power amplifier is some 3500 ohms. R_{25} is chosen about $\sqrt{2}$ times this value, and the collector voltage of Q_3 (5 volts) is approximately three-tenths of the supply voltage. This represents an optimum condition, as can be seen from the following.

Consider a transistor in common-emitter connection, having a collector resistor R_C, a supply voltage E, and a capacitively-coupled load resistor R_L. The bias conditions for optimum class A operation, that is, clipping occuring symmetrically, follow from simple load-line considerations. If the saturation voltage is neglected and no emitter resistance is present, collector voltage and current have to be

$$U_C = \frac{E}{2 + R_C/R_L},$$

$$I_C = U_C\left(\frac{1}{R_C} + \frac{1}{R_L}\right).$$

For this operating point, the power gain, taken as the ratio of power delivered to the load R_L and power consumed by the transistor input, can be expressed in the form

$$PG = \frac{2\eta h_{fe} qE}{kT},$$

where

$$\eta = \frac{1}{2(3 + 2R_L/R_C + R_C/R_L)}$$

is the circuit efficiency, that is, the ratio of maximum output power $P_{L\max}$ to dc supply power $E \times I_C$. (The transistor input impedance was approximated by $h_{fe}kT/(qI_C)$, and h_{oe} was assumed to be much smaller than the load admittance).

Evidently, both power gain and efficiency assume maximum values if

$$R_C = \sqrt{2}\, R_L.$$

The efficiency is then

$$\eta_{\max} = \frac{3}{2} - \sqrt{2} = 8.6 \text{ per cent},$$

and the maximum output power

$$P_{L\max} = \frac{(3 - 2\sqrt{2})E^2}{4R_L} = 0.043E^2/R_L,$$

which in the present design amounts to 2.7 milliwatts. The power amplifier requires 0.1 to 0.3 milliwatt for full output.

The supply voltages are furnished by the simple rectifier and filter circuit shown in Fig. 5. The rectifier receives its ac voltage from the power-amplifier transformer.

VI. Performance

The frequency response of the stereophonic amplifier system is illustrated in Fig. 6. These measurements apply to the high-level inputs and include the power amplifier with a resistive 8-ohm load and the output impedance adjusted to minus 4 ohms.

The two sets of curves demonstrate clearly the action of the loudness control. The curves passing through zero decibels at 1000 cps represent the response when the loudness control is turned fully clockwise. The lower curves were obtained with the loudness control turned back by 40 db at 1000 cps.

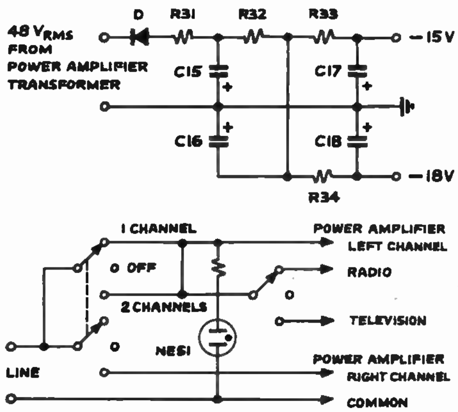

Fig. 5—Power supply.

In each case, the range of the tone controls is displayed by the curves for maximum boost and maximum cut of the bass and treble frequencies. With the tone controls in their center positions, the 3-db points of the response curve are at 6 cycles and 50 kilocycles for full loudness. At minus 40-db loudness, they move to beyond 3 cycles and 100 kilocycles.

The distortion of the preamplifier was measured at 1000 cps and at an output of 200 microwatts, the average level required by the power amplifier to deliver 12 watts into an 8-ohm load. The total harmonic distortion remained below 1 per cent, even in the unfavorable situation where one input amplifier, fed by a low-level source, drives both channels (positions *A* or *B*, monaural).

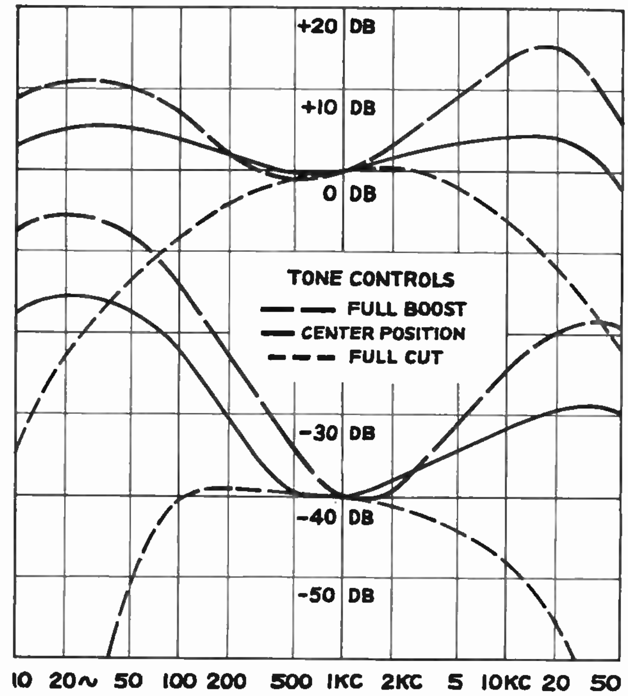

Fig. 6—Frequency response. High-level inputs, including power amplifier.

The sensitivity (input required for rated output) at 1000 cps is 7 millivolts for the low-level inputs, and 1.5 volts for the radio and TV inputs. If necessary, the high-level inputs can easily be made more sensitive by increasing the negative feedback resistor R_{13} in the input amplifier.

Polarity, Phase and Geometry*

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Summary—**Following the teaching of the early Bell Telephone Laboratories' experiments in auditory perspective, wide speaker spacing is needed to realize accuracy of geometry. This results in substantially random phase, so that polarity is relatively unimportant. This is supported by Lissajous figures of two-channel stereo signals.**

It is still good practice, however, to observe polarities, if for no other reason than to permit monophonic reproduction over a stereophonic array. Where the stereo signals contain a strong monophonic component, correct system polarity is better than random polarity. In some stereo situations, bass is improved by correct polarities.

* Received by the PGA, July 20, 1960. Presented in part at Albuquerque IRE Section, November 17, 1960; Austin IRE Subsection, December 7, 1960; and San Antonio IRE Section, December 8, 1960.
† Klipsch and Associates, Hope, Ark.

Introduction

Snow,[1] in 1953, wrote, "It is also good practice to observe a poling convention through all channels, including the microphones and speakers, although the channel spacings are so wide that only very low frequencies can be considered at other than random phase in one channel compared to another."

The work of Bell Telephone Laboratories from 1931 to 1933 in various forms of "auditory perspective" employed spacing of the order of 40 feet. Snow pointed out that the sound source cannot be wider than the

[1] W. B. Snow, "Basic principles of stereophonic sound," *J. SMPTE*, vol. 61, pp. 567–589; November, 1953.

speaker array. Recent practice of 3- to 6-foot spacing offers compactness as its only merit. Narrow microphone spacing by use of the stereo microphone[2] should not be confused with the narrow microphone spacing employed for binaural effects. The stereo microphone may prove to offer the advantages of three spaced microphones, including accuracy of stereo geometry.[3] The work of Snow thus needs to be interpreted in terms of the various recent concepts.

As a tool or at least a heuristic medium, the Lissajous[4] figures will be resorted to.

Lissajous Figures

The Lissajous figure produced by impressing one stereophonic signal on the vertical deflection and the other on the horizontal was first witnessed by this writer in late 1956,[5] at which time it was regarded as a curiosity. Later it was used as means to prove that a claimed stereo recording was actually monophonic. Bauer and Sioles[6] used the figures to draw important conclusions relative to stylus motion, nature of signal content, and other features of a stereo recording. Observing order and polarity, the vertical plates might indicate the left channel, the horizontal, the right; up and to the right are "positive" pulses with respect to a poling convention, and the amplitudes are adjusted so that equal channel voltages will produce equal deflections. Then, a single line at 45° represents a normal monophonic signal; a line at 135° would be a monophonic signal with one-channel polarity reversed. A narrow pattern with 45° slope would suggest a monophonic signal with slight amplitude or phase errors. A "small" phase error would be, say, one or two degrees, as distinct from a polarity reversal or precisely 180°. A perfectly random situation would be a pattern that is substantially symmetrical in general outline, with no significant major or minor axes.

Examples

This writer recorded a symphony orchestra in 1955. Microphones were 22 feet apart, about 15 feet in front of and 15 feet above the podium. Lissajous figures of the two-channel recording display a complete randomness, as in Fig. 1. Listening tests reveal that polarity reversals of any speaker or channel are not detectable aurally. This applies to two-channel playback and also to two-track with bridged center speaker playback. The indifference of the "center channel" to $A+B$ or $A-B$ recombination is significant in indicating the completely random phase conditions. The independence of polarity was not affected by speaker spacing.

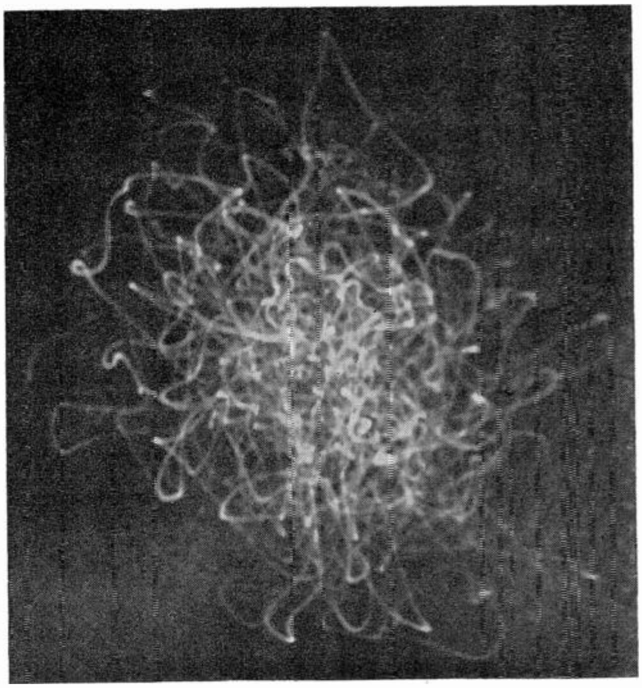

Fig. 1—Complete randomness is shown by the Lissajous figure from a tape recording of a symphony orchestra with omnidirectional condenser microphones 22 feet apart.

The "derived center channel" was found to be effective in "focusing" the soloist near the center of the "playback stage." Properly adjusted, the focus was such that observers on axis judged the singer to be a trifle left of center; observers off axis tended to judge that the soloist was slightly displaced in the direction that the observer was off axis.

By extreme contrast a monophonic signal shows a tilted line, and polarity reversal of one of a pair of speakers is evident by an approach to cancellation as the speakers are brought closer together—this is particularly noticeable in the bass range. A derived center channel requires a sum mixture signal. Use of the null produced by a difference signal is a convenient test for balance. Obviously then, for playing monophonic program material over a stereo system, "a poling convention" should be observed. In fact, this is one suggested way of checking polarity.

Between the two extremes of truly random phases and the zero phase-shift of monophonic lie an infinitude of variations.

Fig. 2 shows the same orchestra, this time playing softly behind a soprano soloist. The soloist was about 5 feet left of center and the same two microphone locations applied. The Lissajous figure was of a sustained high note which was well above the loudness level of the orchestra. The vibrato cycle with its frequency modulation gives rise to the phase shifts which are not random, as in the case of an ensemble of instruments, but may be considered random for purposes of sound reproduction.

[2] Two directional microphones in a single housing: an excellent example is the Telefunken SM-2 which, with 180° spacing of the cardioid patterns for a 90° pickup, rendered substantially the same accuracy of stereo geometry as did the best spaced-microphone technique.

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

[4] Lord Rayleigh, "Theory of Sound," The Macmillan Co., New York, N. Y., pp. 28–30; 1877 (reprinted 1937).

[5] J. M. Eargle, personal communication.

[6] B. B. Bauer and G. W. Sioles, "Stereophonic patterns," *J. Audio Engrg. Soc.*, vol. 8, pp. 126–129; April, 1960.

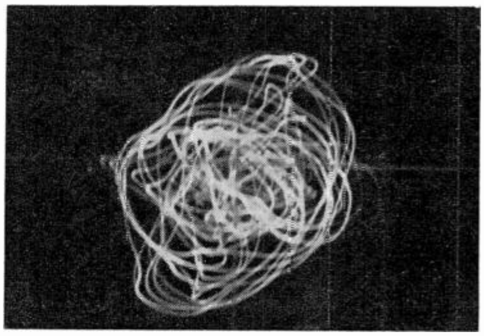

Fig. 2—The same orchestra as background for a soprano soloist, with the same microphone spacing as for Fig. 1. Lack of a sloping major axis indicates randomness of phase.

Here again, polarity of speakers is immaterial, and a derived center channel may be played with either a sum or difference mixture signal with no audible difference.

A center microphone output might be regarded as a "monophonic component" which could be tested for by its cancellation in a difference combination signal for a center derived channel. Such a center microphone output, mixed into two sound tracks, introduces a component which is visible in the Lissajous figure. Similarly, a "stereo microphone" with its cardioids spaced at 180° exhibits an overlap region in the center of the angular pattern. The result is similar to that produced by a center microphone in a spaced microphone array. Fig. 3 shows the pattern produced by such a microphone. At this recording session, the subject matter was a "stage band." The microphone was suspended 13 feet above the stage floor, the cardioids set at 180° and in front of band-to-microphone spacing chosen such as to subtend approximately 90°.

The Lissajous figure is typical of that observed; the major axis is at 45° and the ratio of major to minor axis is about 2:1.

The derived center channel of a derived three-channel array had to be played at some 6-db lower level than for recordings made with two spaced microphones.

Geometric accuracy of the various microphone configurations were restudied,[7] including three spaced microphones with electrically independent channels, and the various uses of two electrical sound tracks with "derived" third channels. The same order of magnitudes of accuracy of geometry was found in each; the only notable difference was that the center derived output using the stereo microphone was always too loud and needed about 6-db extra loss for balance. And, this derived center channel was particular about being $A+B$ or a sum signal rather than a difference. With widely spaced flanking speakers, polarity of flanking units did not seem to be particularly significant in any of the recordings, although some observers seemed to detect polarity effects when the derived center channel was cut out. Another group of recordings involved deep bass. A theater organ with two pipe lofts about 50 feet apart was recorded with two microphones about 40 feet apart and each microphone about 15 feet from its respective loft. Playback of this recording was indifferent to polarity changes; in spite of the deep bass, the two stereo signals were substantially random. Consider a single source such as a 16 foot tibia in the right loft; the left microphone output would be some 10 db below that of the right microphone so that, regardless of phase, no cancellation could occur, and the spacing of 40 feet was 2.5 half wavelengths.

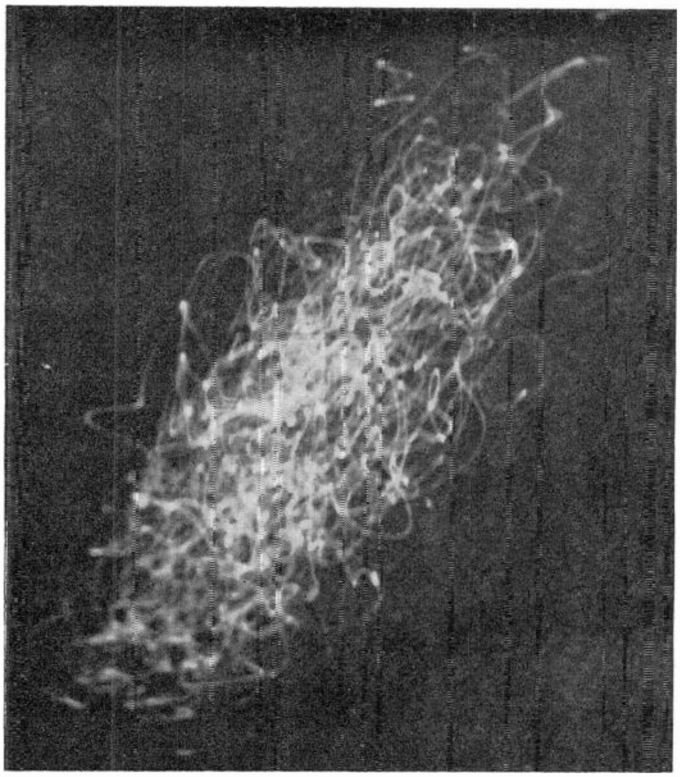

Fig. 3—Lissajous figure produced by recording a jazz band with a Telefunken Stereo microphone. The overlap between cardioid patterns produces a monophonic component indicated by the elliptical pattern. Maintaining correct polarities results in the 45° slope of major axis.

In this case, the derived center playback channel is not noticed, which is as it should be. By itself, the center channel is indifferent to a choice of $A+B$ or $A-B$, which again is as it should be.

Many of the disk recordings currently extant seem to display only a little "polarity effect." An error in wiring a pickup head existed for several months without discovery. When corrected, no noticeable improvement evolved. Had the playback system been two-channel on a narrow array, it is possible that the error would have been discovered by listening. As it was, playing a stereophonic test record[8] was the tip-off and this type of record is as necessary as the tone records were for monophonic.

[7] P. W. Klipsch, "Stereo geometry tests," presented at Acous. Soc. Amer. Convention, San Francisco, Calif., October 22, 1960: submitted to *J. Acous. Soc. Amer.*

[8] "Electronics World," Stereo/Monophonic Test Record, Ziff-Davis Publishing Co.

The fact that one may wish to play monophonic program material over a stereo system is alone sufficient to demand proper polarities. The improved response of stereo program material containing strong monophonic components should be an inducement. Actually, something like nine tenths of the stereo program material is indifferent to polarity which is perhaps the reason that this writer postponed "phase" worries until weightier problems had been settled.

Maintaining Polarity

Maintaining polarities in a tape system is particularly easy. If everything that is not grounded is regarded as "plus," the polarity takes care of itself up to the amplifier output terminals. From there color-coded wire may keep things right side up as far as the speakers. The speaker system may or may not be polarized but, if not marked, the manufacturer can give a schedule by which the driver units can be traced back through the network to the input terminals, which may then be coded.

The disk phonograph is not so easy, but with four terminal pickup heads and a test record, one pair of leads may be reversed, if necessary, to get things right. And if, in the plurality of possible reversals a mistake is made, it isn't fatal.

Conclusion

Polarity is a special case of phase. Polarity can be only 0° or 180°, whereas phase may be any value between 0° and 360° or multiple thereof. Therefore, where line reversal of a speaker is involved the author prefers Snow's "polarity" term rather than the current misnomer "phase."

Polarity is mainly of significance when monophonic signals are being reproduced, or where the stereo signals contain a strong monophonic component. These cases are sufficiently important to force agreement with Snow that it is "good practice to observe a poling convention. . . . "

Appendix

Terminology relative to a center speaker bridged across two channels of stereo to form a combination channel is still a matter of controversy. Since the effect of a third channel is simulated about as closely as with three electrically independent channels,[3,7] the term "derived center channel" is used here, but with the concession that a better terminology may arise.

Acknowledgment

The author wishes to acknowledge use of the "Symposium on Auditory Perspective"[9] by the staff of the Bell Telephone Laboratories. This group of papers contains most of the basic knowledge available twenty-seven years later. The fact that attempts to violate the principles included there have failed would lead to a conclusion that this symposium contains principles as inviolable as the law of conservation of matter and energy.

[9] "Symposium on Audio Perspective," *Elec. Engrg.*, vol. 53, pp. 9–32, 214–219; January, 1934.

Correspondence

Modification of the Magnecord Professional Tape Recorder*

This modification was undertaken in an effort to reduce mechanical flutter in the reproduced signal of a *Magnecord* PT63AH tape transport mechanism. It took the form of adding a stabilizing flywheel driven by the tape motion, as shown in Figs. 1 and 2.

Since the upper circumference of the stabilizing flywheel is three quarters of an inch above the highest point of the machine front panel, the machine was mounted on an extension panel designed and notched for relay rack mounting as in Fig. 3. This notched panel has become a permanent fixture to the machine since it provides clearance for the flywheel. As a result of this extra clearance space required for the flywheel, this machine, being impractical for remote location work, has become the permanent relay-rack mounted recorder.

Fig. 4 shows the lower idler wheel removed from the front panel of the machine and a hole punched in the panel at that point to clear the bearing race of the proposed flywheel and stabilizer drum. At first, it was decided to fit the mandrel shaft and the stabilizer drum together from two pieces, by expanding the drum with heat and allowing it to contract on the mandrel shaft. This was entirely unstaisfactory however, since concentricity could not be maintained or regained. A second shaft and drum was machined from a solid piece of cold-rolled steel and turned between centers on the lathe, and, while the small diameter and length of the mandrel shaft tended to spring slightly, very light cuts were taken and micrometer measurements made after each cut to insure that tapering was not taking place.

The machined shaft and drum was fitted into the ball bearing races and a dial indicator device was set up to measure the total run-out in thousandths of inches. There was no perceptible movement of the dial indicator pointer when the stabilizer drum was rotated slowly by hand.

A portion of the machine superstructure had to be cut away as in Fig. 5 to provide clearance for the upper lip of the flywheel, and the entire erase-bias oscillator chassis structure was moved two and one half inches to the rear. Additional bracket supports were fitted to the superstructure, which hold the bias-erase oscillator three inches immediately to the rear of its former position.

The flywheel and drum parts are shown in Fig. 6, and the assembly operation in Fig. 7. The first stabilizer drum was quite small, and it took an excessively long period of time to bring the flywheel up to speed, and the wrap of the tape around the drum was not sufficient to maintain flywheel speed.

* Received by the PGA, July 20, 1960; additional material received, November 23, 1960.

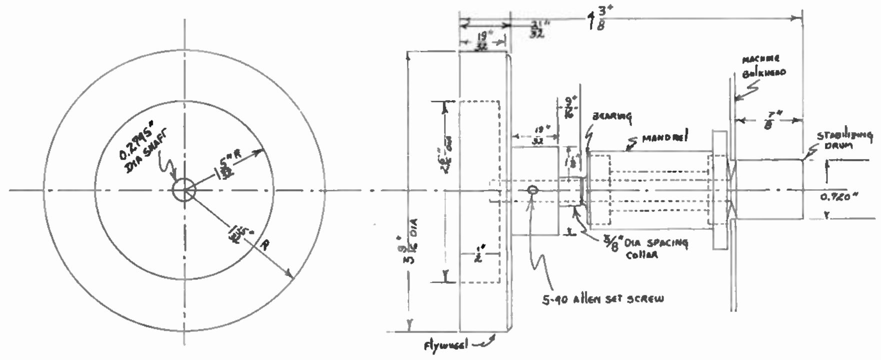

Fig. 1—Assembly of stabilizing flywheel. The space between the mandrel and the bulkhead of the machine is for a 1/32-inch spacing washer to provide clearance for the sealed surface of the front bearing. The bearings are push-fitted into the race openings. The 5–40 Allen set screw in the flywheel secures it to the shaft. The 3/8-inch dia. spacing collar between the flywheel and the rear bearing raceway is to provide upper lip clearance of the flywheel to the machine superstructure.

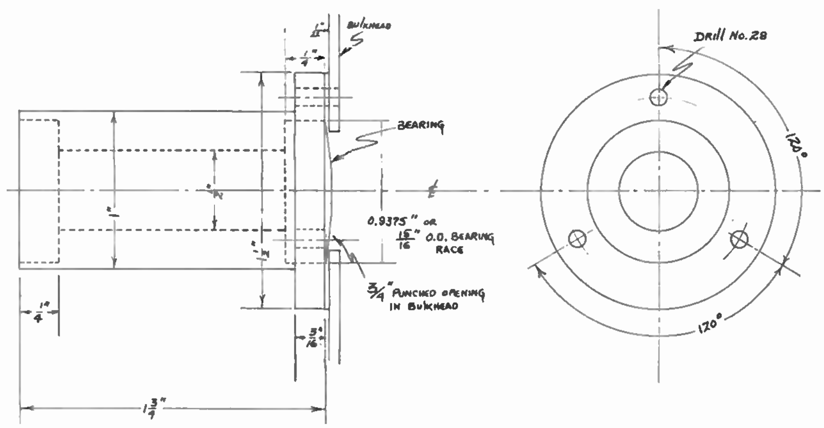

Fig. 2—Mandrel and bearing race detail. The three equally-spaced mounting holes in the rim of the bearing mandrel are drilled No. 28, clearance drill for 6-32 machine screws.

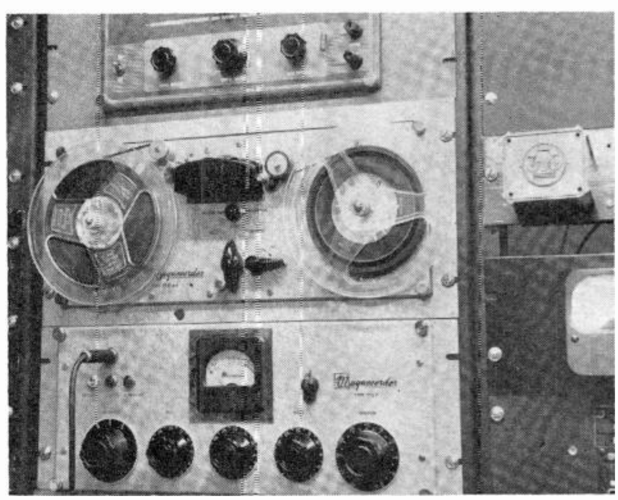

Fig. 3—An over-all view of the completed modification project on a *Magnecord* PT63AH tape transport mechanism. The stabilizer drum can be seen in position formerly occupied by the lower idler wheel. The stabilizer drum in this photograph was the first one made. It proved unsatisfactory because the small diameter of the stabilizer drum did not provide enough wrap of the tape to bring the flywheel up to speed nor to maintain the speed. A second larger drum was made with a greater surface for tape wrap, allowing flywheel speed to be maintained with a lower percentage of slip between the tape and the drum surface.

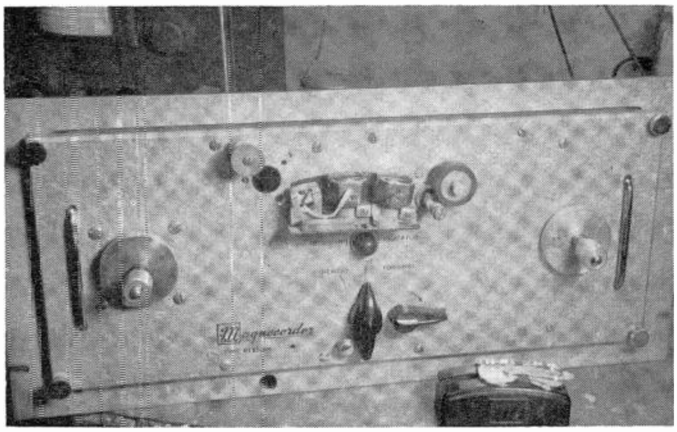

Fig. 4—A front view of the *Magnecord* LT3AH tape transport mechanism during the modification process. The punched hole for the stabilizer bearing raceway, plus the three mounting holes, spaced 120 degrees apart, can be seen. The mounting hole nearest the three head record-reproduce assembly is countersunk for a flathead screw, to provide clearance for the head assembly cover. The other two bearing mount screws are binder heads to match the other screw heads that appear on the front panel. The notched relay rack mounting panel extension is shown here as a permanent fixture to the machine because of the overhang by the stabilizer flywheel.

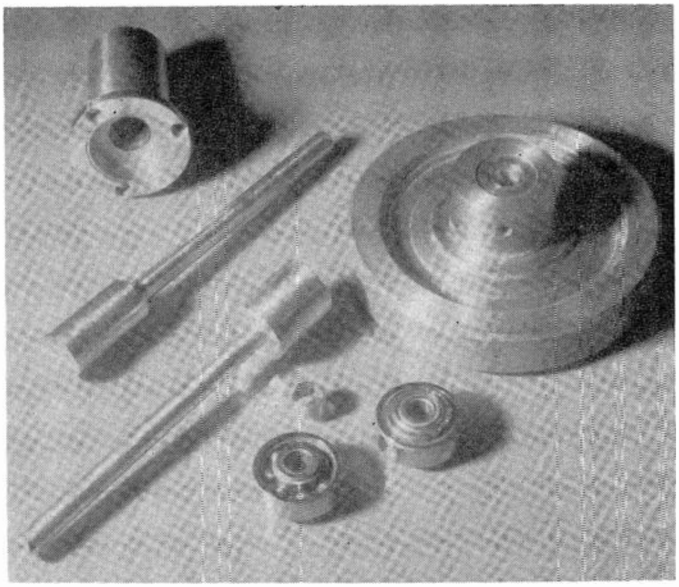

Fig. 6—A closeup view of the stabilizer flywheel and drum assembly parts. The bearings support, shown in the upper left-hand corner, was machined from a solid piece of cold-rolled steel. A hole was drilled through the center for mandrel shaft clearance. The bearing support was then mounted on a mandrel and turned between centers for the outside shoulder and the inside bearing race surfaces. In the photograph, the upper mandrel shaft and stabilizer drum was first made by joining two pieces with heat, but proved unsatisfactory since concentricity could not be maintained in the process. The lower shaft and drum was machined from a solid piece and turned in a lathe between centers where it maintained very close tolerances. The little collar is placed on the mandrel shaft between the rear bearing race and the flywheel proper, as a spacer. The flywheel was machined all over from a solid bronze casting, drilled and tapped for a 5-40 Allen set screw.

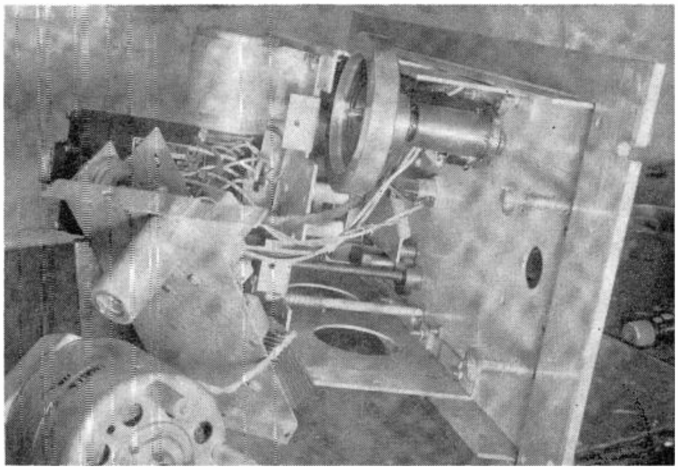

Fig. 5—A closeup view of the *Magnecord* PT63AH modified machine showing the erase-bias oscillator dismounted, but with its extension bracket, plus the dismounted rewinding motor. The cut-out portion of the machine superstructure can be seen where the upper lip of the flywheel extends above. This picture shows the completed stabilizer flywheel assembly just after completion of the modification job.

Fig. 7—The author mounting the flywheel on the mandrel shaft of the modified *Magnecord* PT63AH tape transport mechanism. The extension bracket on the erase-bias oscillator can be seen. The rewind motor assembly was dismounted during the modification process to permit access into the machine superstructure.

When the second shaft was machined, the size of the stabilizer drum was increased to that of the original idler wheel (0.720 inch). This permitted sufficient wrap of the tape to bring the flywheel up to speed, and to maintain speed even at the 15-inch per second tape speed, and still there was adequate energy stored in the flywheel to provide a stabilizing influence on the tape motion.

In the absence of a flutter meter, I used the following method of determining instability in the tape motion of the machine: Record a sinusoidal tone of some midrange audio frequency, say 3000 cps, and then, without disturbing the audio generator, play back the tape and "zero beat" the tape-recorded signal against that of the generator. By virtue of this comparison, minute flaws in stability show up quite nicely and are easily detected by the unaided ear. The human ear, under certain circumstances, is an excellent comparing device capable of detecting very small deviations of two tones very near the same cyclic rate within the audio hearing range. The differential between the two tones takes the form of "beats." The beats become slower as the two tones approach coincidence, or the same frequency. Flutter and wow can be easily separated and identified, although, due to a lack of instruments, the percentage is unmeasurable.

Before the modification, there was appreciable unsteadiness in the tape motion, as indicated by this comparison process, principally resulting from the method of drive through the use of intermediate friction drive wheels between the motor shaft and the capstan flywheel. In this connection, it is interesting to note the very-low-flutter figures obtained with the *Ampex* endless-belt drive systems, where the intermediate drive wheels are eliminated.

Even in view of the drawbacks of the drive system under consideration, a most gratifying result was immediately obvious by reason of the addition of the stabilizing flywheel described. The same midaudio frequency tone from the generator and from the recorder was so closely coincident and so nearly equal in stability that there was difficulty in determining from which source the tone emerged. Switching back and forth from the generator output to the recorder output, it was impossible to detect instability. Only when the combined outputs from the generator and the recorder were listened to was there any discrepancy between the two. Compared to results before the modification, this was considered about all that could be accomplished in the way of tape-motion stabilization without completely rebuilding the machine to encompass another system of drive.

In conclusion, significant improvement in the stability of this machine for smooth tape motion was noticed immediately after the job was completed. This was particularly so in the case of recording and reproducing constant tones. However, ordinary wide-range music listening tests were not appreciably improved, except at the 15-inch per second speed. For a relatively minor modification, the results indicated a major improvement in the steadiness of tape motion, and a very worthwhile experiment in machine shop practice for someone so inclined and so equipped.

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Contributors

Albert Baaba was born in Chicago, Ill., on February 4, 1932.

He was a radar technician in the United States Air Force from 1950 to 1952. In 1956, he joined the Manufacturing Division of Ampex Corporation. He transferred to the Ampex Corporate Research Division, Redwood City, Calif., in 1958, and has engaged in research on various aspects of magnetics and magnetic recording, most recently in the position of project engineer.

A. Baaba

❖

Egbert de Boer was born in Amsterdam, The Netherlands, on July 8, 1928. He studied at the University of Amsterdam, receiving the Sc.M. degree in physics in 1953. In 1954 he joined the Physical Laboratory, Otorhinolarynogology clinic of the Wilhelmina Hospital, Amsterdam, where he did research on cochlear microphonics and on auditory signal analysis. In 1956 he received his Doctorate in Physics and Mathematics from the University of Amsterdam.

E. de Boer

In 1957 he did research on communication biophysics at the Massachusetts Institute of Technology, Cambridge, on a Fulbright Travel Grant. In 1958 he returned to The Netherlands, and is currently engaged in research on auditory masking and audiology.

Dr. de Boer is a member of the Nederlandse Natuurkundige Vereniging (Physical Society), the Nederlandse Vereniging voor Audiologie (Audiological Society), and the Nederlandse Keel-, Neus-en Oorheelkundige Vereniging (Society of Otorhinolaryngology).

❖

Donald F. Eldridge (A'50–M'55) for photograph and biography, please see page 68 of the March–April, 1960, issue of these Transactions.

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Demetri P. Kanellakos (S'55–M'60) was born in Tripolis, Greece, on October 26, 1932. He received the B.S. and M.S. degrees in electrical engineering in 1956 and 1957, respectively, from Illinois Institute of Technology, Chicago, Ill.

From 1956 to 1958 he served as an instructor of electrical engineering at Illinois Institute of Technology teaching various courses, and at the same time he pursued further graduate studies and did research at the same institute. From 1958 to 1960 he was with the Armour Research Foundation of Illinois Institute of Technology, where he worked in the microwave department on circular waveguide components, the shielding effectiveness of electromagnetic enclosures, the application of the magneto-resistive and Hall effects to microwave electronic circuit devices, and parametric amplifiers. Since August, 1960, he has been employed by the Radioscience Laboratory of Stanford University, Stanford, Calif., as a research associate working in the field of ionospheric wave propagation.

D. P. Kanellakos

Mr. Kanellakos is a member of Eta Kappa Nu and Sigma Xi.

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Paul W. Klipsch (A'34–M'44–SM'45), for photograph and biography, please see page 106 of the May–June, 1960, issue of these Transactions.

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Roland P. Schuck (S'57–M'59) was born on December 10, 1931, in Chicago, Ill. He received the B.S. degree in electrical engineering from the University of Illinois, Urbana, in 1958.

He served in the U. S. Navy as a communications technician for the National Security Agency. Since 1958, as a staff member at Armour Research Foundation, of Illinois Institute of Technology, Chicago, he has been engaged in a variety of electronic research projects in the field of electronic instrumentation. His most recent work has been the investigation of the Hall effect as applied to the advancement of the state-of-the-art of electrical measurements and communications.

R. P. Schuck

Mr. Schuck is an associate member of the AIEE. He received his certificate for Professional Engineer-in-Training, state of Illinois, in 1958.

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Werner Steiger (M'58–SM'60), for photograph and biography, please see page 236 of the November–December, 1960, issue of these Transactions.

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Alva C. Todd (A'38–M'47–SM'54) was born in Ligonier, Ind., on July 30, 1917. He studied at Indiana University, South Bend, the George Washington University, Washington, D. C., and Purdue University, Lafayette, Ind., and received the B.S., M.S., and Ph.D. degrees in electrical entineering from the latter.

A. C. Todd

After serving as a Lieutenant at the U. S. Coast Guard laboratory during World War II, he joined the electrical engineering staff at Purdue University. In addition to teaching assignments and thesis direction, he was responsible for the organization and operation of the electronics instrumentation laboratory. In 1953, he became Director of Engineering at the Fournier Institute of Technology, Lemont, Ill. In addition to his administrative responsibilities, he taught courses in communication engineering, antennas and waves, and network theory. He joined the Farnsworth Electronics Co., Ft. Wayne, Ind., in 1955 and served as a consultant to the microwave and radar departments, and in 1956 entered private practice, specializing in electronic instrumentation and control. He joined the staff of the Armour Research Foundation of the Illinois Institute of Technology, Chicago, in 1957, and was responsible for a wide range of programs, including the development of circular electric waveguide components and special-purpose microwave antennas, and a broad study aimed at advancement of the state-of-the-art of electronic instrumentation techniques. He joined the Hallicrafters Co., Chicago, Ill., in 1960, and is in charge of applied research in the military division.

Dr. Todd is a member of AIEE, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi. He served as the Faculty Adviser of the IIT student chapter of AIEE and is currently the Chairman of the Electronics Committee of the Chicago chapter of AIEE. He is a registered professional engineer in Indiana and Illinois.

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