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CONTRIBUTIONS

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The Editor's Corner

NOTHING NEW IN AUDIO

BY THE TIME our car pool reached Elm Street, we had solved most of the world's important problems, and the subject of audio came up. Bill Sarnac, who always likes to argue, proclaimed that the hi-fi nuts give him a pain in the anatomy. "Every week: a major breakthrough. You investigate, and it's the same old thing in a different color package. There's nothing really basic in the last half-century. Binaural sound was demonstrated at the Paris exposition about seventy years ago. Tape recording in the 1890's. Dynamic and electrostatic loudspeakers are vintage of World War I."

Ed Lewis, as usual, took the positive side, and declared, "You sound like the patent commissioner of the last century, who recommended that the U. S. patent office be disbanded because just about everything had already been invented."

"We can look at it in two ways," said Mike Seda, who generally tries to be impartial. "There are new sciences and old sciences. In semiconductors they are discovering new things every day: thermoelectric generators, hall elements, transistors, masers, silicon rectifiers, tunnel diodes, solar cells. A good scientist has a lot of elbow room in these fields. But in audio—I don't know. We rehash the same old stuff, year after year."

Ed reflected, "Even so, an audio engineer can be just as enthusiastic about squeezing down that last tenth of a per cent amplifier distortion as a physicist about collecting a new particle from a cyclotron."

"In fact, I've noticed that many engineers will sidetrack things that are important," said Mike, "and concentrate on comfortable side-projects such as making flat amplifiers even flatter, recalibrating instruments that are already good enough, measuring inductances to four places, putting decals on the panels—."

"They are quite defensive about it," said Bill. "If you suggest something positive they tell you that it's no use doing anything further until the instrumentation is right."

Ed was thoughtful. "I think that the heart of the entire problem is that we are not yet able to measure the things that really matter in acoustics and audio. Room acoustics, for example, where we take the reverberation time; reminds me of biologists who will homogenize a colony of insects, then analyze the chemical mixture telling us they are 19 per cent carbon, 8 per cent nitrogen, and heaven knows what. The facts are true, and repeatable, and they can put them into equations. But they miss the real nature of the thing. And, what is worse, they are so busy studying these routine half-truths that they never get to the real truth. If we could only find measurements that correlate with human experience, we would open the door to real progress."

"Before we condemn the measurements we should go back even further," proclaimed Mike. "We can't measure something, unless we first know what we would like to measure, and I'm not sure we know that. When I was young and foolish, I used to think I wanted perfect realism in sound, and I was sure we'd have it some day. Now I am confused. I attended a learned session where a panel of experts dismissed perfect realism as impossible—I never could pin them down as to exactly why. I suspect it's the mixing of two sets of different acoustical environments, first in pickup and then in reproduction. Yet I've been at demonstrations put together by nonexperts. All they did was to connect together some good-quality commercial microphones, tape recorders, and loudspeakers. They placed the loudspeakers among live musicians; they played part of the time and then stopped, and the recorded music continued. Truthfully, I wasn't able to tell one from the other."

"These demonstrations are wonderful things," ventured Bill. "About fifty years ago Brunswick came out with "high-fidelity" shellac disks for wind-up phonographs—advertised as responding from 16 cycles to 16,000 cycles. Pure fiction, but they got eminent musicians to endorse the lifelike quality. In the twenties one of the large record manufacturers used to compare their piano recordings with a real piano. By coincidence they always chose a room with a noisy ventilating fan; skeptics used to say the exhibit carried a fan along, as it was the most important part."

"I'll admit," continued Mike, "that I'm not a musician. I wasn't familiar with the selection, nor with the room; and I don't know exactly how the real instruments should sound. It is as if someone would show me a real diamond and a fake. I might notice certain differences, but not being familiar, I'd probably choose the wrong one."

"From here we logically move into the realm of 'taste', " said Ed. "If the ten-cent jewelry looks prettier to you, then *for you* it is better than a real stone."

"It may be for a short time," answered Bill, "but after you live with it a while you might change your mind. The flashy qualities that had an impact when you were uninitiated are the ones you can't stand later on. That's why we have critics who are supposed to give us the benefit of their experience."

"However," said Ed, "don't overlook the possibility that music can actually become more pleasant when it passes through an electronic system. Musical instruments aren't magically perfect; they are just pieces of catgut, tin, and wood, glued together to make fiddles and whistles of different sizes and shapes. The reproducing system can blend them together, take the edge off harsh tones, increase the bass, and do many other things that the instrument makers wish they could build into their products."

"That is all right," answered Bill, "but if they want to doctor up the music, let them do it at the source. Here they can control the microphones, musicians, room effects, equalizers, and reverberation. And they have the benefit of the conductor's and the music director's judgment. On the other hand, if you are going to make your playback system a part of an instrument, then you have to set it differently for each instrument. It's not a iob that should be left for amateurs."

"For my part, I say: 'Get it perfect at the source. Then give me a replica of the original'."

"Now we are back to 'How are we going to do it'," said Mike. "The more I learn, the more complicated it gets; and I feel inadequate as a mere engineer. Somewhere there must be professors or advanced researchers with a lot more widom who can unravel these problems. Maybe the answers will be childishly simple when they find a new approach. Nuclear energy had them stymied fcr a long time; then suddenly it opened up."

At this point, we reached First Avenue, the end of the line. As Mike got out he summarized the feelings of the group: "You know, we may not have solved anything today, be we really did some solid thinking. See you tomorrow."

—MARVIN CAMRAS, *Editor*

Note: A series of articles in these TRANSACTIONS reviewing current research in acoustics and audio should be of special interest. The first of these, "Current Research in Communication Acoustics" by Roy A. Long and Vincent Salmon, appears on page 37 of this issue.

PGA News

CHAPTER NEWS

Note to Chapter Chairmen and Secretaries: Please send reports of meetings and other Chapter news to Bill Ihde, Chairman Committee on Chapters, General Radio Company, 6605 North Avenue, Oak Park, Ill.

Baltimore, Md.

Louis R. Mills of Recordings, Inc., presented a paper on "Modern Loudspeakers for Stereo-Design by Listening Approach" at a meeting on November 23, 1960.

Chicago, Ill.

A joint meeting of the Chicago Acoustic and Audio Group and the PGA Chapter was held at the Sunset Arms Hotel in Franklin Park, Ill., on November 19, 1960. Martin W. Basch of the General Radio Company, West Concord, Mass., presented a paper entitled "The Effective Use of the Level Recorder in Acoustic Measurements."

On Friday, January 13, 1961, following the IRE Section General Paper, Pete Tappan of Warwick Manufacturing Company spoke on "Improvement in Simulated Three Channel Stereo." This was also a joint meeting with CAAG at the Western Society of Engineers headquarters in downtown Chicago.

Theodore Prybst reported that attendance had increased to 51 at the September, 1960, PGA meeting, compared to 42 in September, 1959.

Philadelphia, Pa.

The following meetings have been scheduled for the 1960–1961 season, all held at WCAU Radio Station, City Line Ave., Philadelphia:

October, 1960 "Seresoid Modulation," a description of theory and application of seresoid modulation as used in FM transmitters, with demonstrations, by Bernard Weiss of Industrial Transmitter and Antenna.

January, 1961 "Power Amplifiers for High Fidelity," a review of the art, including demonstrations, by Bruce de Palma, formerly of Dynakit.

March, 1961 "High Fidelity Aspects of FM Receiver Design," with major emphasis on detector systems used, by Emery Fisher, Fisher Company, New York.

April, 1961 "Acoustic Suspension Loudspeaker Systems," a discussion of the AR-2 loudspeaker and similar models, by a representative of Acoustic Research, Mass.

San Antonio, Tex.

PGA programs reported were as follows:

December 8, 1961. "Stereo Geometry Measurements," presented with a demonstration by Paul W. Klipsch and Associates, Hope, Ark. The same presentation was made at Austin, Tex., on December 7.

March or April, "Transistorized Stereo Preamplifier," by Daniel Meyer of Southwest Research Institute.

San Francisco, Calif.

George Behklau, Charles Wilkins, and John Bennett, of Ampex Professional Products Division, spoke on "What's New at Ampex," at the PGA meeting on October 4, 1960.

Stan Hose, of Triad Transformer Corporation, Venice, Calif., presented a paper, "Design Considerations of Audio Transformers," on November 2, 1960.

Washington, D. C.

"Electrical Methods of Tone Generation and Control" was the subject of the meeting on September 12, 1960. Robert White of Kitt Music Company was the speaker.

"Sound in the Theatre" was discussed on October 18, 1960, by Harold Burris-Meyer.

ANNOUNCEMENTS

Help Wanted

Committee 30 on Audio and Electroacoustics is looking for people to work on:

Methods of Measurement of Distortion.
Methods of Measurement of Transmission Characteristics.
Methods of Measurement of Microphone Characteristics.
Methods of Measurement of Loudspeaker Characteristics.
Methods of Measurement of Recording and Reproducing Transducer Characteristics.

Please reply to Michel Copel, 156 Olive Street, Huntington Station, N. Y.

INTERNATIONAL ACOUSTICS CONGRESS

Physical, technical, physiological, and psychological acoustics will be covered at the Fourth International Congress on Acoustics, to be held on August 21 to 28, 1962, in Copenhagen. Sponsored by the International Commission on Acoustics, it is being organized by the Acoustical Societies of Scandinavia, a union of scientific bodies in Denmark, Finland, Norway, and Sweden. The Acoustical Society of Denmark will be in charge of local arrangements. An international exhibition of acoustical equipment will be arranged in connection with the Congress.

In view of the great number of contributions to the 1959 Congress, the Commission has announced that it may be necessary to limit the number of papers accepted for presentation in 1962. Preference will be given to new material of scientific interest, and papers which have already been published will not be accepted. Technical sessions will be held in the buildings of The Royal Technical University in Copenhagen.

For further information as it becomes available, write to the secretary of the Fourth ICA Congress, 10 Oestervoldgade, Copenhagen, Denmark.

SUMMARIZED ANNOUNCEMENT 1961 SPECIAL SUMMER SESSION

Moore School of Electrical Engineering University of Pennsylvania

The Moore School of Electrical Engineering of the University of Pennsylvania has announced a Special Summer Session on recent developments in the field of electrical engineering. Four two-week programs will be given from June 4 through July 15. Titles of the programs and the senior lecturers are as follows:

1) Modern Radar Techniques, Prof. Raymond S. Berkowitz.
2) New Devices for Amplification and Switching, Prof. Noah S. Prywes.
3) Communication Theory and Information Handling, Prof. Pierre L. Bargellini.
4) Logic, Switching Systems, and Automata, Prof. George W. Patterson.

The aim of the session is to provide a coordinated presentation of engineering developments of the past five years in these technical fields and thus to help bring engineers, scientists, mathematicians, and technical administrators abreast of new foundations and techniques in the fields closely related to their own.

Instruction will be intensive; participants will attend class 6 hours a day, 5 days a week, 60 hours in all. Teaching will be authoritative; the staff is being assembled from the forward ranks of industry. Prerequisite for enrollment is a bachelor's degree or its equivalent in engineering, mathematics, or science. Enrollment fee is "250 per program; an institutional rate of $125 is available to educators on request. Lodging in the University dormitories is available or, if the student prefers, accommodations can be booked in one of the nearby Philadelphia hotels. Further details may be obtained by writing to:

Prof. Morris Rubinoff, Coordinator
1961 Special Summer Session
Moore School of Elec. Engrg.
University of Pennsylvania
Philadelphia 4, Pa.

Dr. Robert Benson, formerly with Armour Research Foundation, is now a Professor of electrical engineering at Vanderbilt University, Nashville, Tenn. Dr. Benson has been active in PGA for many years, and is a member of the PGA Administrative Committee.

Current Research in Communications Acoustics*

R. A. LONG†, MEMBER, IRE, AND V. SALMON†, SENIOR MEMBER, IRE

Summary—**Despite the appeal of the visual arts, all persons turn naturally to sound for the rapid, thorough, and precise transmission of both practical and esthetic information. Communications acoustics deals with this flow of acoustic information, and with controlling noise that might interfere with it.**

Introduction

In all our waking hours we are engaged in generating, transmitting, storing, reproducing, and receiving acoustic signals involving people, machines, ideas, and nature. These activities form the subject matter of communications acoustics, and can be described in terms of both human and electronic mechanisms.

The electronic hardware used in communications acoustics forms the nucleus of an industry that continues to show steady expansion. In one small branch, the entertainment field, manufacture of TV sets in 1959 probably passed the five-million mark, running at least 20 per cent ahead of 1958. And radio, once considered moribund, accounted for about 13 million new receivers in 1959, at least 40 per cent above 1958 production. Not to mention that the phonograph, whose "death knell" was sounded by the advent of radio 35 years ago, shows annual sales increasing beyond the peak of its years of monopoly. Indeed, sales of electronic communications equipment are increasing faster than our gross national product. Nonmilitary sales probably amounted to $5 billion in 1959. Broadcasting, service, and distribution added at least $4 billion more.

The Amazing Ear

Acoustics enters into all these facets of the communications industry, because the ear is the court of last appeal. Let us summarize some disconnected facts about human hearing. Nerves can scarcely carry 1000 pulses per second, but we can hear to 18,000 cycles per second and beyond. With the assistance of a physiological automatic-volume-control mechanism, we can perceive an almost painful sound containing a million million times as much energy as the weakest discernible. The trained ear can make remarkably good guesses about the pitch of a tone, and a few favored persons have authentic absolute pitch. An especially attentive ear can act as a wave analyzer in identifying the primary constituents of a complex tone or noise. When we are immersed in a cocktail party milieu, we can "tune out" undesired conversation. No electronic device has completely reproduced this action of the ear.

The ear has a poor memory for loudness, and comparisons of the loudness sensation from two sources can be made only by switching back and forth between sources. However, this is one of the few characteristics in which electronic instruments greatly exceed the capabilities of the ear.

To be heard, a sound must first be generated. Humans and other animals communicate by an internally-generated voice. They also use means exterior to their bodies for producing and controlling musical sound. Studies on the use of human speech for conveying pragmatic information have, however, received the major share of recent attention.

Distilled Speech

Theory has indicated that we normally rely heavily on redundancy to ensure the unambiguous and precise transmission of intelligence. Thus, syllables and words that are garbled in transmission need not compromise the information. But suppose it were possible to abstract from speech those elements that carry the intelligence. If these essential elements were transmitted over a perfect communication link, presumably the message capacity of transmitting equipment would be greatly increased.

Research is aiming at this end. We already can analyze speech, transmit the analysis, and reconstitute the speech at the receiving end. Unfortunately, the process is still expensive and imperfect, and the received speech, while intelligible, is unnatural and unidentifiable. However, the expense of a transoceanic telephone call (or much later, an earth-moon call) will remain a strong impetus to the perfecting of such equipment.

Research suggests another approach: use ordinary speech to control a reconstituted voice that is impersonal and perfect. The major difficulty is in the regional and individual variation of accents. If you are a careful speaker, you can "dial" a phone number by speaking into a digit-recognizing device adjusted to your voice. But if you are from New England, don't expect a Texan to get the right number! However, the convenience of such a device makes its development almost inevitable. On the other hand, voice-operated typewriters are much further in the future.

Speech production is essentially the modulation by the vocal chords and oral passages of a steady flow of air from the lungs. This natural process suggests similar mechanical devices, and in particular, airstream modulators for producing loud sounds. The idea is an old one, extending to the late 19th century. Recently, interest has been renewed, and several modulated airstream loud-

* Received by the PGA, January 30, 1961. Reprinted from the *SRI J.*, vol. 4, 1st Quarter, 1960.
† Stanford Research Inst., Menlo Park, Calif.

speakers have been devised. However, standing squarely in the way of a fundamental understanding of the processes involved is our ignorance of the general solution of the equations of pulsating compressible flow. Because these equations also represent processes that occur during combustion in rockets and jet engines, their solution is all the more important. The person who first succeeds will become a bright star in the firmament of mathematical physics.

Resonance

When sound is transmitted within a closed space (a "room" to the acoustical scientist), its effect on the ear depends on the characteristics of both room and source. Every room has a set of resonances peculiar to it, and the ear hears the direct sound plus that from all the resonances that are excited. Each resonance occurs at a certain frequency, and each resonance is spaced from its neighbors by an approximately constant frequency difference.

This theory, developed in the last decade, has an interesting application, based on the fact that the room is harder to "excite" between resonances. When a public address system is used in a room, and its volume control is turned up too far, the system may then "sing." This phenomenon is caused by sound from the loudspeaker getting into the microphone, and then getting amplified for another round trip around the system. Such singing is often controlled by the resonance characteristics of the room. To eliminate the singing, a device is introduced between the microphone and the loudspeaker which changes all frequencies by such an amount that we shift from the easily excited resonance to the more-difficult-to-excite region between resonances. This has been done on a laboratory scale, and the system operates as predicted, making possible much greater sound reinforcement before "singing." Considerable engineering will be needed before a commercial device can be produced for a portable public address system, but we should see this soon.

In the transmission of sound out of doors, we are often faced with the problem of shielding against undesired sound. A solution frequently suggested is the planting of trees and shrubbery. Careful and detailed tests have revealed consistently that the reduction in sound intensity is discouragingly small unless a dense forest about 100 feet deep is planted. Nevertheless, even limited plantings have a useful effect in visual screening of the source of sound. If a sound source is not particularly prepossessing in appearance, then visual screening works psychologically to reduce the aural distraction and annoyance. Thus, plantings can be useful, but not for the reason usually given.

Sound Storage

Of primary interest to the communications industry is any instrument that can store sound for later reproduction. By using sound to produce patterns of mechanical or magnetic analogs, we have familiar phonodisk and phonotape recordings. One of the goals is to give the listener the sense of being transported into the presence of the original sound. One of the best means of accomplishing this is by binaural (*not* stereo) recording.

In this process, sound is picked up by two microphones placed approximately an ear's width apart, then reproduced into earphones worn by the listeners. (Stereo involves two microphones set wider apart and is played back through loudspeakers.) In addition to experiencing the extreme naturalness of the reproduced sound, the listener retains a sense of direction of the source. Finally, if several conversations are recorded at once, the listener can "tune" to the one selected, then repeat the playback while tuned to another. This immediately suggests important uses for binaural recording. Court reporters can use it as "audio notes" to provide an adjunct to their shorthand notes. Similarly, conferences can be taped and transcribed with relative ease. Noise studies of moving sources can be conducted by first making binaural recordings in the field. Then, on repeated playback to a "sound jury," subjective judgments of loudness, annoyance, and distraction can be obtained under controlled and repeatable conditions in the laboratory. Although complete binaural recording systems are not now sold as such, and research to determine optimum operation has not been accomplished, much can be done with equipment that is now available. The next decade should see binaural recording firmly ensconced as a useful scientific tool.

Whether or not sound is stored on phonotape or phonodisk, the commercial success of the final product depends on having rapid, inexpensive means of duplication. The cost of duplicating a 12-inch LP record in large quantities, including material, is considerably less than 50 cents each, and the process takes 150 seconds or less. However, in duplicating phonotape, the cost of a seven-inch reel of tape alone is over a dollar. Added to this is the cost of duplication time; this amounts to about 5 minutes per reel. Before tape can compete economically, this problem must be solved. A vigorous and sustained research program on improved means of duplication will be necessary to exploit the many desirable characteristics of magnetic tape.

Noise Problems

Noise, defined as undesired sound, is the unwelcome participant in all communications processes. It can mask the intended signal, or by its presence can distract, annoy, or even deafen. Quantitative research on noise loud enough to deafen has been underway for only ten years. The research effort was triggered by changes in the attitude of the courts toward work injury cases. Ordinarily, compensation is awarded for loss of ability to earn a living. In hearing loss cases, some awards were on a new basis, that of decreased ability to enjoy life. It has been estimated that if all the possible suits in New York State alone were settled on this basis, the workmen's liability insurers would have a $2 billion bill.

Although research has helped establish tentative measures of sound intensities above which ear damage can occur, three major problems remain unsolved. Of greatest value to industry would be a rapid means of detecting individuals who may be especially susceptible to ear damage, so that they may be kept from aurally hazardous jobs. At present, this is discovered after the fact, and only by periodic audiometric examinations. A second problem is estimating the degree of damage cumulated in noise exposure from previous employment, military service, and age. Several hypotheses are being tested, but we must still rely to a great extent on the history of the person. A third problem is that of estimating ear damage from impact noise that occurs so rapidly that the ear's automatic volume control does not have time to function. Recent work on cats has provided some answers; humans must be tested in a more nondestructive fashion.

The subjective annoyance and distraction from noise still defies objective measurement, because it is only indirectly related to loudness. Consider your distraction on hearing your name in a *sotto voce* whisper! The annoyance problem is becoming acute in this jet age, as airport runways grow longer, aircraft are more powerful and noisier, and homes and schools are built near airports. The psychological aspects are probably greater than the physical. Survey teams have long since learned never to ask whether a noise is annoying. This question alters the reaction, and makes it invalid. Much work has been done on correlating community reaction with noise characteristics, time of day, occurrence rate, association of noise with job, and association with danger. Some tentative measures of expected complaint reaction have been evolved, but are not of general application. The acoustical specialist will need assistance from many scientific disciplines in obtaining generally valid measures of community reaction to noise.

The noise problem is particularly bad in motels built near airports. The difficulty is usually one of isolation and not absorption of noise. These two problems require completely different attacks, materials, and constructions in their solution. In actuality, most motel occupants need only a quiet place to sleep. For those not suffering from claustrophobia or necrophobia, an underground sleeping vault has been suggested. A more acceptable solution may be the use of a dynamic sound absorber that introduces a cancelling tone to reduce the loudness of unwanted low-pitched sounds. Placed near the head of the sleeper, such a device should be a useful supplement to the sound isolation afforded by a construction using tightly-sealed, heavy, double walls with their inner and outer surfaces connected only resiliently. Another alternative is to place a noisy fountain in the motel courtyard to provide an "acoustic blanket."

Theory indicates that the amount of noise isolation afforded by walls depends on their weight. Present research is directed toward achieving acoustic insulation that at least equals this theoretical value over a wide range of frequencies. This goal seems reasonably close, but the next, that of beating the "weight law," will take much more than the present desultory research. If the weight-law limit can be greatly exceeded, there will be a made-to-order market waiting in aircraft (and later spacecraft), where there is a high premium on low-weight contruction.

The control of noise is best effected by reducing it at its source. Some noises are under the direct control of the operator—*e.g.*, sonic booms. In truck noise, operator technique is almost as important as a good muffler. But in most industrial noise problems, the machines must be designed for quiet operation. If this is attempted after the machine has been built, the results are often expensive and unsatisfactory. Design of machines for quiet operation ordinarily involves well-known physical facts which have only to be recognized and applied. However, in the really difficult problems such as jet and rocket engine noise, sound is so intimately associated with performance that it is a real research task to reduce noise without compromising performance.

Playing Back

If our original sound has survived detection, transmission, and storage reasonably well, it is ready to be recreated by some sort of playback equipment. In the entertainment field, stereo equipment is presently popular. This recording-reproduction system, when used properly, aids in creating the illusion called *presence*—that is, the listener feels an intimate rapport with the soloist who seems to be moving toward him from the mass of sound. Much of the presence realized on phonodisk and phonotape arises from judicious use of the *precedence* effect. This effect implies that if one sound arrives at the ear from two sources, the first signal to arrive will capture the attention of the ear, even if the second sound is more intense. Research has delineated the relation between the permissible intensity change and the delay between the two signals. Thus, if a soloist is given a close-in microphone, we hear the soloist first, because the other microphones for the accompaniment are usually more distant. Hence the soloist captures the ear, and the precedence effect is used to obtain presence.

A more recent application of the precedence effect is in ambiophonic recording, which may eventually appear on the market. If we record music meant to be played in a reasonably reverberant hall in the usual fashion (say stereo), the hall reverberant sound will be rather subdued. We would like to be able to adjust this reverberation to the living room in which the sound is reproduced. To accomplish this, we make an additional (and simultaneous) recording taken from a microphone placed about 30 feet beyond the main ones. The stored output from this is played back from loudspeakers placed toward the rear of the listeners. By controlling the volume of sound from these speakers, we control the intensity of the "stereo reverberation" introduced. At the same time, since sound from the main speakers in front of us is heard first, by the precedence effect we ascribe to these speakers all the sound. Choral music is especially effective with ambiophonic recording. The next decade may see some of the technical problems of the method solved.

Even after a sound has been reproduced, it must be heard. We would like some objective measure of the acoustical quality of the room in which we listen. At present, there is considerable ferment about this problem, because subjective judgments are still the only way of assessing the value of the objective determinations. Indeed, acoustical scientists are often accused of listening to the sound rather than the music. In one study of the acoustics of concert halls, even those auditoriums with gross acoustical faults were sometimes accepted by musicians, but not by acoustical scientists. Again, the psychoacoustic aspects probably outweigh the physical. This will continue to be a most fruitful field for study.

Hearing Aids

The ear can suffer such great hearing loss that hearing aids are necessary. Transistors have made possible small hearing aids, the cosmetic appeal of which has attracted many who needed an aid long ago.

Most aids are fitted on the basis of a hearing loss curve. This is usually determined by the intensity of a minimum audible tone, with the intensity compared to that of an arbitrary and accepted norm. However, such a procedure has little relation to the performance of the ear under normal conditions of hearing. To make the necessary measurements is difficult and is often beyond the capabilities of many who fit hearing aids. A most heartening development in mounting a logical attack on this problem has been the recent adoption of an integrated approach. A team consisting of at least an otologist, an audiometrist, and an electronics engineer can diagnose the medical condition, measure the hearing, and select the proper hearing aid. What is still needed is a rapid means of measuring hearing deficiencies for the conditions under which we hear. Speech and noise signals can be used, but a test facility giving rapid and consistent results would be welcomed by all. At present, the electronics man selects aids, using his painfully-won experience with them, and then fits them by trial and error. The coming decade will see improvements that will make the integrated team approach a still more useful one.

Conclusion

The foregoing has indicated that the worker in communications acoustics must have capabilities in electric circuit theory, vibration theory, electronics, music, psychology, architecture, and medicine. All too few of our universities offer this comprehensive background in a well-planned curriculum. But only on such broad-gauge training can effective interdisciplinary research be based. One can only hope that the next decade will include the development of adequate training programs along with an attack on the challenging research still remaining in communications acoustics.

Acoustical Measurements on a Home Stereo Installation*

JOHN K. HILLIARD†, FELLOW, IRE

Summary—**This paper describes a two-way 500-cycle crossover loudspeaker system using the infinite baffle principle. A 300-cubic-foot volume is used. Two identical systems are spaced 15 feet apart for stereo reproduction.**

A description of the living-room architectural acoustics is included. There are no parallel walls or surfaces, and the splay of the side walls is a minimum of one inch per foot. Reverberation decay charts are provided for different frequencies to indicate the effect of the splayed walls. Sine wave and warble tone curves are shown and sine wave single-frequency levels are plotted for the width and length of the room.

LOUDSPEAKER enclosures and room acoustics are a very important part of any high-quality music system. It is well known that with an infinite amount of enclosed space separating the front and rear radiation, adequate radiation impedance of low-frequency speakers can be obtained down to the lowest frequencies desired.

A living room was designed and built to provide all of the essential features needed to obtain the maximum performance available, especially for two-channel stereo reproduction.

The living-room floor plan is shown in Fig. 1, along with dimensions. There are no parallel walls or surfaces. The splay of the side walls is a minimum of one inch per foot in the horizontal direction.

The side walls are 40 per cent glass and 60 per cent wood. The wood panels are cedar boards 10 inches wide, and the back bracing is of a random design so as to avoid panel resonance.

This angular living room has exposed-beam construction. The carrying beams come into the ridge beam at a 30° angle.

The floor has 75 per cent of the area covered with tufted wool carpet, with rubber undermat. The balance of the floor area is slate.

The loudspeaker enclosures are mounted on each side of the chimney, which has a total width of 15 feet and a height of 18 feet. The chimney is composed of cinder block, with stone facing. Each enclosure is 4 feet by 4 feet by 18 feet. The three inside walls are concrete, and the wall facing out is one inch plywood, with random diagonal bracing. The loudspeaker system is mounted on a hinged door 4 feet by 4 feet, held in position by screws.

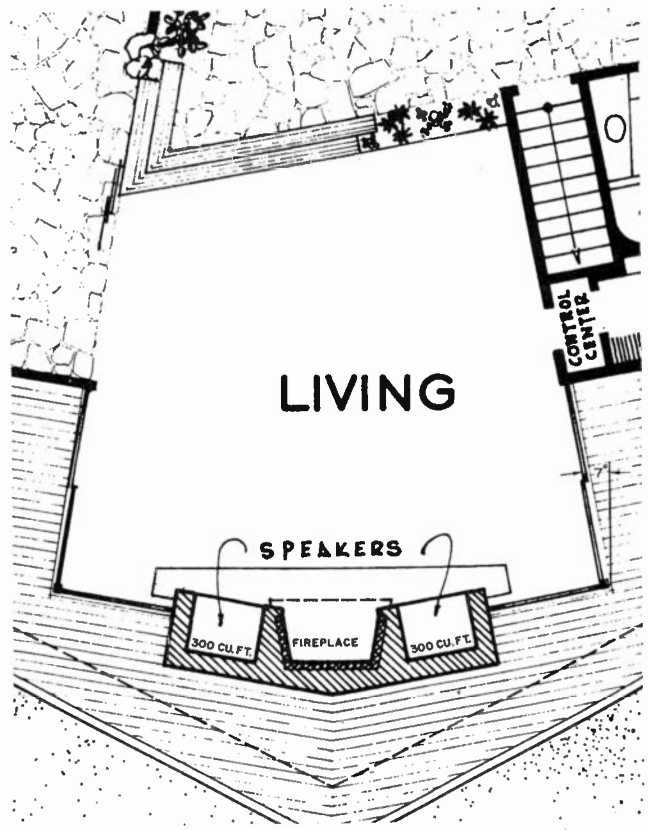

Fig. 1—Living-room floor plan.

A Belgian linen grill cloth is used having a 50 per cent opening, and extends the full height of 18 feet. The lower portion, over the baffle, hinges in such a manner that the entire 18 feet of height is in one piece.

The loudspeaker system utilizes two 15-inch low-frequency cone speakers having a natural resonance of 25 cycles, and a sectoral horn and driver for the high-frequency power. The crossover is a 12-db-per-octave, 500-cps M, dividing network.

Experience with high-quality theater and reinforcement systems has provided information to warrant only one crossover and it should be at a frequency around 500 cps. This low crossover is dictated by the fact that the 15-inch speaker provides uniform acoustic power only up to this point. Above 500 cps, the acoustic power decreases and the angle of radiation becomes smaller.

The high-frequency horn is a sectoral type, 21 inches in length, and has a mouth area of 150 square inches.

* Received by the PGA, February 7, 1961.
† Ling-Altec Research Division, Ling-Temco Electronics, Inc., Anaheim, Calif.

The distance or horizontal separation between the acoustic centers of the low-frequency and high-frequency units is 14 inches, which is little more than one millisecond time delay. Numerous tests over a period of years on high-quality loudspeakers have revealed that a delay of less than two milliseconds is not detected. More than this amount is not recommended.

The low-frequency speakers are mounted as close together as possible and the high-frequency horn is located above, since vertical stacking yields the best horizontal distribution. See Fig. 2.

The room has a volume of 10,000 cubic feet, and the reverberation time over the range of 30–6000 cps is shown in Fig. 3. The individual decay curve for 100 cps is shown in Fig. 4, and for 3500 cps in Fig. 5. The nonparallel walls and absorption of the random braced panels contribute to a smooth decay.

The response of the system for sine wave input is shown in Fig. 6, and the response to warble tones having a 40-cps change at the rate of 32 per second is shown in

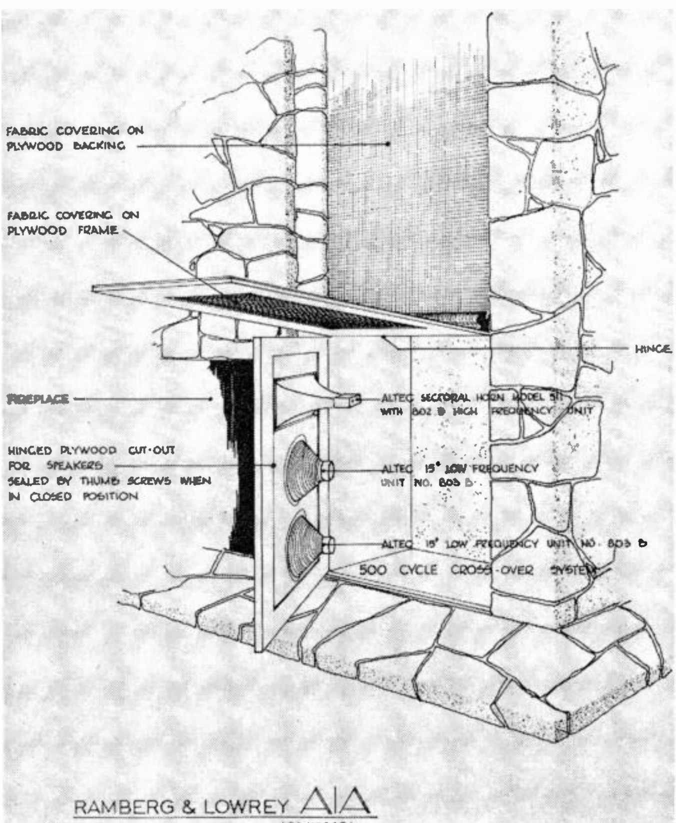

Fig. 2—Speaker array.

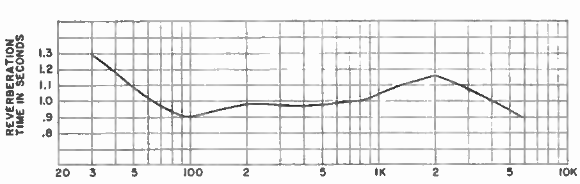

Fig. 3—Reverberation curve.

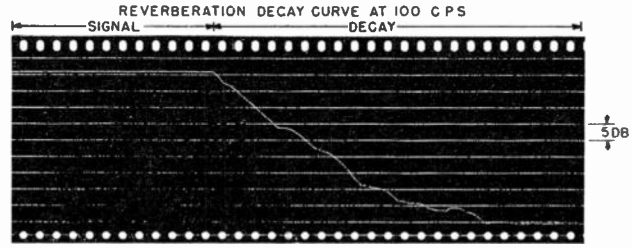

Fig. 4—Decay curve for 100 cps.

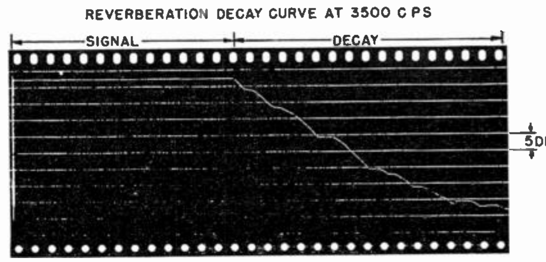

Fig. 5—Decay curve for 3500 cps.

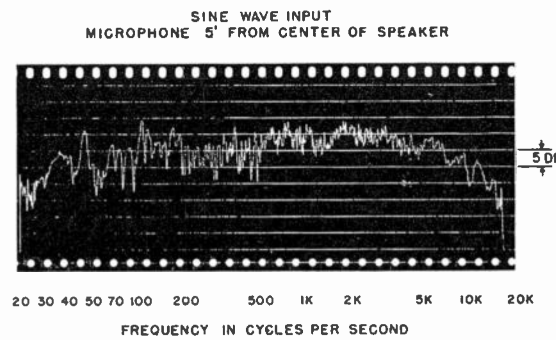

Fig. 6—Sine wave response.

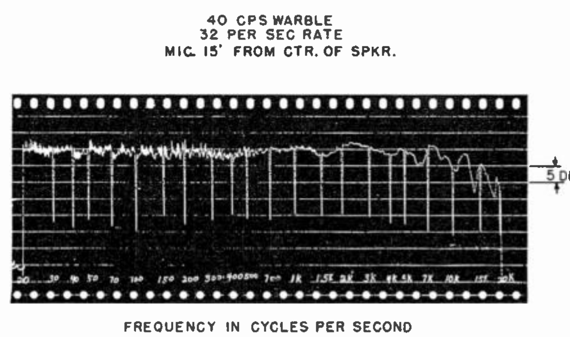

Fig. 7—Warble tone response.

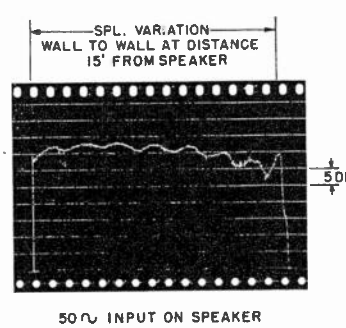

Fig. 8—50-cps variation.

Fig. 7. The variations for a pure tone of 50 cps across the room are shown in Fig. 8.

The generally smooth decay in reverberation time is attributed to splayed walls, random bracing of wood panels, and distributed absorption.

The use of a very large enclosed volume of air behind the speakers has resulted in the achievement of results comparable to horn loading at extremely low frequencies. The combined area of two 15-inch cone loudspeakers has provided an equivalent piston of 250 square inches. This has reduced the necessary amplitude of cone motion for adequate acoustic output with low distortion.

Plane-wave tube measurements indicate an electroacoustical efficiency of 40 per cent for the low-frequency units in the frequency range of 50–400 cps.

Conclusions

These measurements confirm the expected improvement in smoothness of reverberation decay resulting from splayed walls. The use of a large sealed enclosure provides an acoustic radiation impedance permitting efficient operation to the very low frequencies.

Significant information on frequency response using sine wave excitation can be obtained when optimum architetctural acoustics properties are provided.

Intermodulation Distortion Meter Employing the Hall Effect*

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Summary—This paper presents a method of measuring intermodulation distortion produced by a nonlinear system operating in the audio-frequency spectrum. The design of the measuring device is based upon the Hall principle. Transistorized circuitry is used throughout. Both the circuitry employed and the theory of operation are discussed. The instrument is simple to operate and permits the measurement of intermodulation distortion at any frequency on a point-by-point basis within the range of 400 to 20,000 cps. Distortion measurements may be made in two ranges; zero to one per cent and zero to five per cent.

I. Introduction

At PRESENT, there are three standard techniques for the measurement of amplitude distortion in audio amplifiers; these involve:

1) The measurement of the individual output wave components of the amplifier under test by means of a waveform analyzer, when a single-frequency sine wave is employed as the test signal.

2) The measurement of the total rms harmonic distortion in the output wave of the amplifier under test by means of a signal-elimination filter and rms voltmeter, when a single-frequency sine wave is employed as the test signal.

3) The measurement of the difference-frequency (or the sum-frequency) component in the output wave of the amplifier under test by means of a band-pass filter and rms voltmeter, when two single-frequency sine waves are employed as the test signal.

The total rms distortion measurement type 2) is easy to perform at any audio frequency, using commercial equipment designed for this purpose. The measurement of the individual output signal components by means of the waveform analyzer is a slow, tedious task, rarely undertaken outside the research laboratory. Intermodulation distortion task 3) is readily performed using commercial equipment designed for this purpose; but it is seldom performed at more than two sets of frequencies.

It is the purpose of this paper to describe a simple method that will permit the measurement of intermodulation distortion at any frequency in the audio spectrum with the same ease as experienced in the measurement of total rms harmonic distortion. The simplicity of design was achieved by employing the Hall effect along with appropriate circuitry. The theoretical aspects of the problem and the operation of the instrument are presented.

* Received by the PGA, October 31, 1960. The work described in this paper was performed as part of a program sponsored by the U. S. Army Signal Res. and Dev. Agency under Contract no. DA 36-039, SC-78269. The experimental model was constructed by R. H. Fors and J. E. Hutter.

† Hallicrafters Co., Chicago, Ill.

‡ Armour Research Foundation of Illinois Institute of Technology, Chicago, Ill.

II. Theory of Operation

If a fixed low-frequency sine wave current with a frequency of 75 cps, for example, is applied to the excitation coil of the Hall device, and an arbitrary, single-frequency sine wave signal, adjustable throughout a frequency range of 400 to 20,000 cps, is applied to the Hall element, the output signal developed by the Hall device will be

$$v_0 = kI_1I_2 \cos \omega_1 t \cos \omega_2 t, \tag{1}$$

where:

k = the over-all constant of the device (including the Hall constant)

I_1 = the crest value of the current at fixed angular frequency ω_1

I_2 = the crest value of the current at arbitrary angular frequency, ω_2

By expanding the expression in the usual manner,

$$v_0 = \frac{kI_1I_2}{2}\left[\cos (\omega_2 + \omega_1)t + \cos (\omega_2 - \omega_1)t\right]. \tag{2}$$

It may be noted that the difference between the frequency of the two sideband components is constant at twice the frequency of the fixed generator, that is,

$$\frac{\omega_2 + \omega_1}{2\pi} - \frac{\omega_2 - \omega_1}{2\pi} = \frac{2\omega_1}{2\pi} = 2f_1. \tag{3}$$

III. Intermodulation Distortion Measurement System

Fig. 1 presents the block diagram of the intermodulation system, and Fig. 2 illustrates the circuit diagram. The system is composed of the following stages: an internal oscillator having a fixed frequency of 75 cps, an external variable audio-frequency oscillator, a balanced modulator, a low-pass filter with a cutoff frequency of approximately 150 cps, a dual-purpose amplifier (broadband 20 to 20,000 cps; frequency selective at 150 cps,) and a detector circuit.

The per cent distortion is read directly on a front panel meter which monitors the signal level of the difference frequency of the two sideband components generated by the balanced modulator after the signal has passed through a nonlinear network (amplifier under test) and a frequency selective network.

Two switching circuits are provided. The first one has three positions, permitting the measurement of either 0–1 per cent D, 0–5 per cent D, or I_c; where per cent D indicates per cent intermodulation distortion and I_c is the control current supplied to the Hall element. The second switching arrangement has two positions: one for calibration in which broad-band amplification is employed, and the other for per cent D measurement whereby the frequency selective amplifier and its associated circuitry is used.

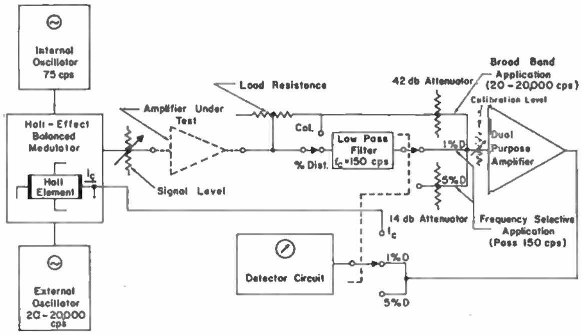

Fig. 1—Intermodulation distortion test system, block diagram.

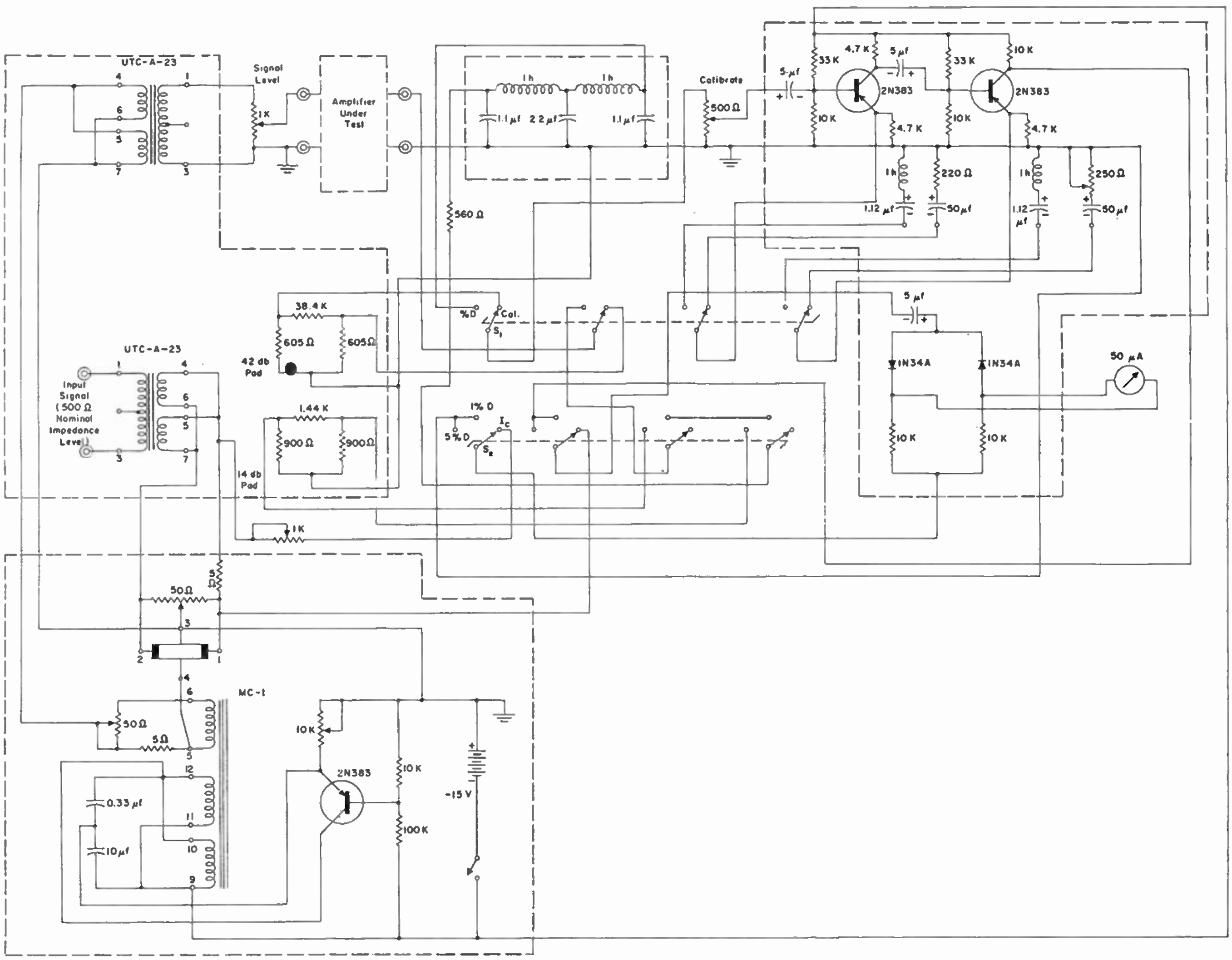

Fig. 2—Intermodulation distortion meter.

IV. Balanced Modulator

The balanced modulator was tested for carrier suppression by utilizing a HP 300A harmonic wave analyzer. Suppression characteristics were checked from 400 to 16,000 cps. From 400 to 2500 cps, the carrier level was negligible when compared to the sideband components. Between 2500 and 16,000 cps, the suppression ranged from −48 db to −25 db. The modulator utilized an Ohio Semiconductor MC-1 Halltron Magnetic circuit.

V. Frequency Selective Networks and Their Characteristics

Two networks are used to provide the necessary frequency selectivity of the device. The first network is a low-pass filter consisting of four constant-k half sections connected in a twin-π lumped parameter configuration. The second network is a selective amplifier composed of two cascaded common emitter stages.

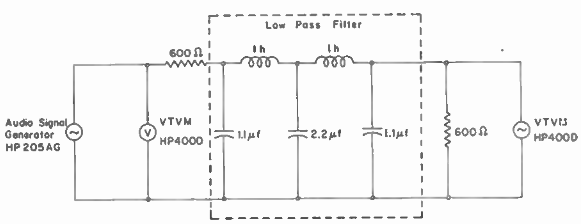

Fig. 3—Filter test system.

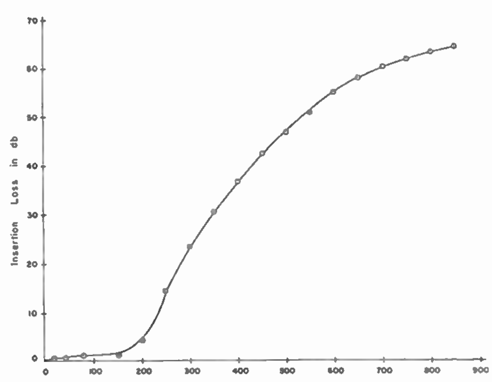

Fig. 4—Filter performance characteristics.

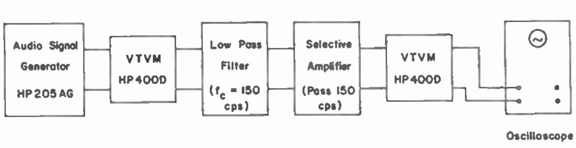

Fig. 5—Combined filter-selective amplifier system.

The filter was designed to work into a 600-ohm load (the approximate input impedance of the amplifier) and have a cutoff frequency of approximately 150 cps. Fig. 3 shows the test system used to obtain the filter characteristics illustrated in Fig. 4.

Selectivity is achieved in the amplifier unit by utilizing a series LC circuit resonant at 150 cps in the emitter leg in each stage of amplification. Fig. 5 illustrates the test system used to determine the output characteristics of the combined filter and amplifier circuits. Figs. 6 and 7 show, respectively, the frequency selectivity characteristics and the overload performance.

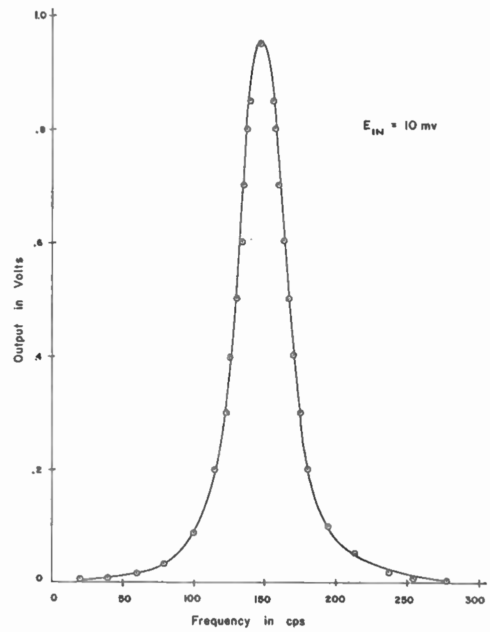

Fig. 6—Selectivity characteristics.

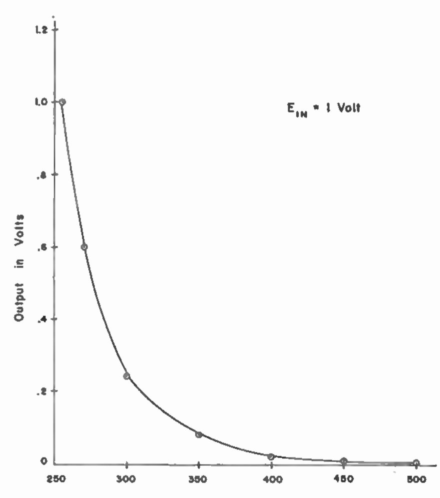

Fig. 7—Selective amplifier overload characteristics.

VI. Broad-Band Amplifier Performance

The broad-band amplifier was designed so that the voltage amplification at 150 cps is the same as that of the selective amplifier; this is quite important since the degree to which they match in amplification performance is a major factor governing the accuracy of the distortion reading obtained. Fig. 8 illustrates the frequency-response curve of the broad-band amplifier.

If switch S_1 is placed in the per cent distortion (or frequency selective) position and S_2 in the one per cent position, the 42-db pad is bypassed and a full-scale meter deflection would indicate one per cent distortion, *i.e.*, one per cent when compared to the total reference signal as previously calibrated. To broaden the scope of the instrument, a 14-db attenuation pad has been placed prior to and in series with the amplifier. S_2 is utilized in placing this pad in the circuitry, and in this position a full-scale meter deflection indicates a distortion of five per cent. By properly calibrating the meter face, distortion measurements can be observed from zero to five per cent in two separate ranges.

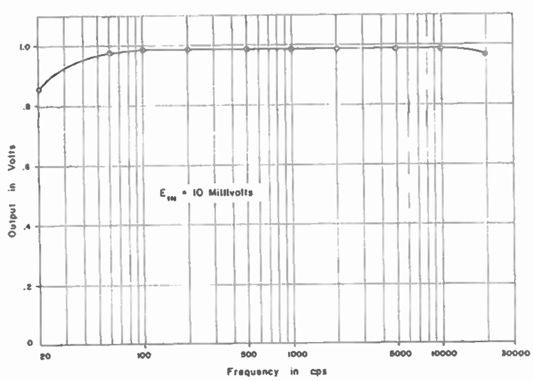

Fig. 8—Broad-band amplifier characteristics.

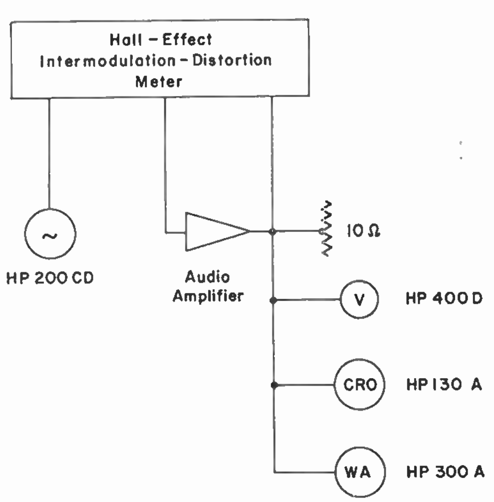

Fig. 9—Distortion meter test system.

In addition to the one and five per cent distortion measurement ranges, switch S_2 utilizes the detector circuit, at full-scale meter deflection for a relative measurement of the control current I_c, which is essentially constant throughout the usable frequency range of the instrument.

VII. Experimental Behavior

The test system of Fig. 9 was employed to compare the performance of the Hall-Effect Intermodulation-Distortion Meter with that of a Hewlett-Packard model 300 Harmonic Wave Analyzer, when the audio amplifier under test was operated at an output level of one watt and ten watts. The frequency-response characteristic of the audio amplifier given in Fig. 10 shows that the amplifier has a flat gain characteristic over a frequency range of 15 cps to 45 kc. At the ten-watt level, the behavior of the amplifier is similar, except that the amplitude distortion becomes excessive at frequencies below 30 cps. The results of the distortion measurements are presented in Fig. 11. It may be observed from a comparison of the difference-frequency intermodulation distortion values that the instruments are substantially in agreement. Figs. 12 and 13 show the completed instrument.

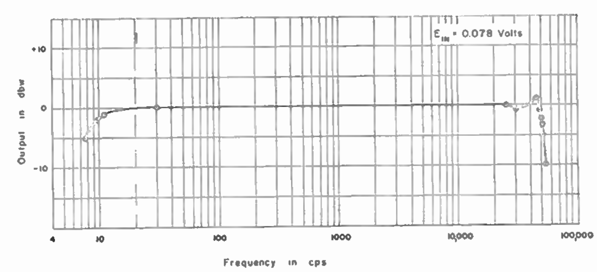

Fig. 10—Frequency response of audio amplifier.

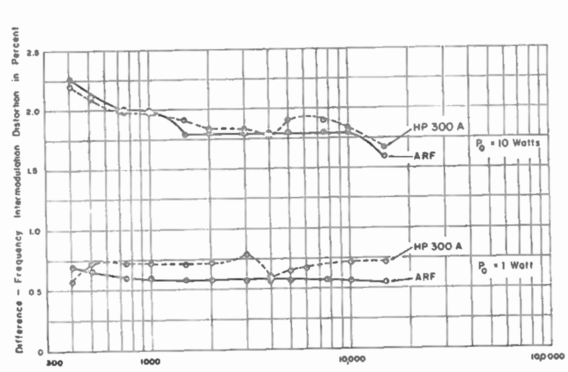

Fig. 11—Intermodulation distortion measurement curves.

VIII. Conclusions

A method for the measurement of the difference-frequency distortion of an audio amplifier has been presented. The method has been embodied in a simple experimental distortion meter employing only three transistors and a Hall-effect modulator. The difference-frequency intermodulation-distortion of a wide-range audio amplifier was measured by means of the experimental distortion meter, and the measurements were verified by means of a commercial harmonic wave analyzer. The distortion values were found to be in agreement, within the usual limits of accuracy for this type of measurement. The speed and ease of measurement afforded by the Hall-Effect Intermodulation-Distortion Meter should make it attractive in all instances in which the measurement of difference-frequency component is considered a sufficient indication of audio-frequency amplifier amplitude behavior.

Fig. 12—Hall-Effect Intermodulation Distortion Meter, front view.

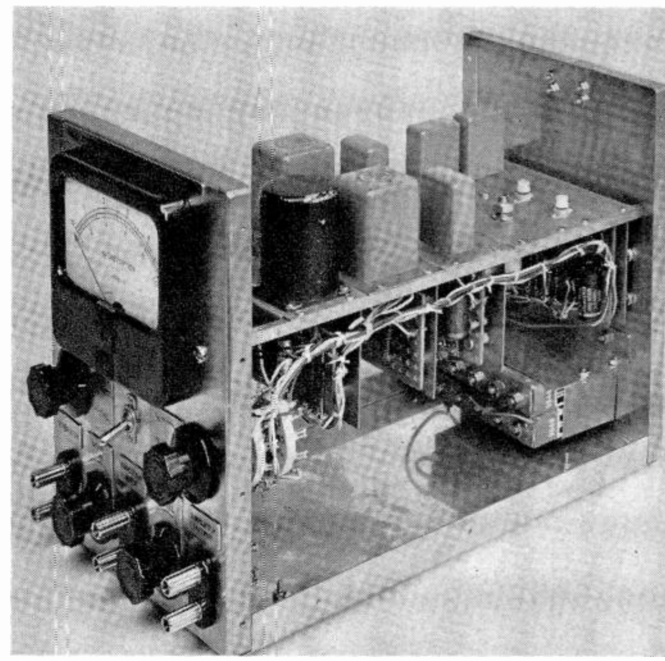

Fig. 13—Hall-Effect Intermodulation Distortion Meter, open side view.

Third-Order Distortion and Cross Modulation in a Grounded Emitter Transistor Amplifier*

HELMUT LOTSCH†

Summary—**The nonlinear modulation characteristic of a transistor is approached by the first terms of a Taylor-series formula and the output voltage is calculated for a suitably-chosen ac-input voltage. For higher frequencies and different working points, the influence of the diffusion capacity and the emitter admittance are illustrated by experimental results. It is shown that—as in electronic tubes—the third-order distortions (harmonic distortion, distortion of modulation, alteration of the degree of amplitude modulation, degree of cross modulation) are closely connected, but unlike the situation in electronic tubes, the distortions disappear at a special working point. The influence of the base resistance and the internal resistance of the signal source are discussed in relation to the appearance of this zero point and the frequency dependence.**

In an RF amplifier a transistor is modulated, for example, only by a part of the resonant circuit voltage because of the low input resistance in a grounded emitter circuit. Although this voltage transformation is accounted for, the transistor is inferior to the electronic tube with respect to cross modulation. To reduce the cross modulation in a transistor RF stage, two possibilities are described for correcting the distortions with a fixed or controlled working point using a predistortion or a push-pull modulation. In this case the different influences of a resistance in the base lead, or of a resistance blocked for HF but not for LF in the emitter lead, are discussed.

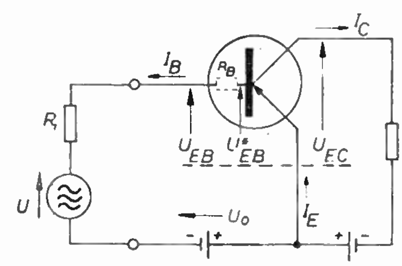

Fig. 1—Transistor circuit, indicating positive directions of alternating currents and voltages.

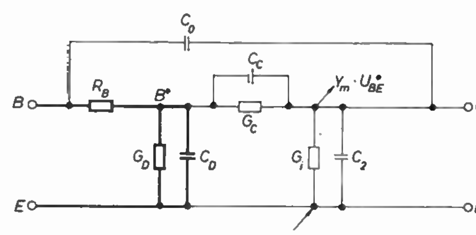

Fig. 2—Equivalent circuit of a transistor in grounded emitter circuit.

I. Equivalent-Circuit Diagram and Modulation Characteristic of a Transistor

The transistor presented as a four-terminal network is represented by an equivalent-circuit diagram containing complex resistances and one or several ac generators.[1–3] As with the grounded cathode circuit in tube amplifiers, the grounded emitter circuit is the most important one. If the output of the transistor is short-circuited ($R_a = 0$ in Fig. 1), the distortions are caused only by the nonlinearity of the short-circuit current-amplification factor. Because the output is short-circuited, there are no reactive effects from output to input and the reduced equivalent circuit diagram (Fig. 2) is adequate. To calculate the nonlinear distortion for LF, the diffusion capacity C_D may also be neglected, but for HF it is important. This diffusion capacity enlarges the input admittance for HF and at the same time diminishes the effective modulation voltage at the interior base-lead B^*. Diffusion capacity C_D and emitter admittance G_D are dependent on the working point of the transistor.[4,5]

Considering low frequencies, Boltzmann's law shows that the base current I_B depends approximately exponentially on the voltage U_{EB}^* at the emitter barrier layer.[6,7]

$$I_B = I_0 \exp \frac{U_{EB}^*}{U_T}; \tag{1}$$

$$I_0 = I_B \quad \text{for} \quad U_{EB}^* = 0,$$

$$U_T = \frac{kT}{e}, \tag{2}$$

$$(U_T = 25.74 \text{ mv} \approx 26 \text{ mv at } 23°),$$

* Received by the PGA, December 5, 1960; additions received, January 30, 1961.

† Cathode-Ray Tube Dept., Telefunken, Ulm, Germany.

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

[2] G. Meyer-Brötz, "Die Vierpolparameter des Flächentransistors in den drei Grundschaltungen," *Telefunken-Zeitung*, no. 111, pp. 21–28; March, 1956.

[3] W. Moortgat-Pick, "Ersatzschaltbild und Verstärkungseigenschaften des Flächentransistors," *Arch. Elektr. Übertragung*, vol. 13, pp. 33–48, 82–89; January, February, 1959.

[4] W. Minner, "Die Vierpolkenngrößen des HF-Transistors OC 615 für den Frequenzbereich von 30 bis 135 MHz," *Telefunken-Röhren-und Halbleitermitteilungen*, no. 590 147, pp. 1–8.

[5] R. Olschewski, "Die Messung der dynamischen Transistor-Kenngrößen bei Hochfrequenz," *Telefunken-Röhrenmitteilungen für die Ind.*, no. 561 014, pp. 1–5.

[6] W. Engbert, "Die Kennlinien und Ladungsträgerverteilungen des Legierungstransistors," *Telefunken-Zeitung*, no. 114, pp. 277–287; December, 1956.

[7] H. J. Thuy and R. Wiesner, "Halbleiter-Bauelemente, ihre Physik und technische Entwicklung," *Elektrotech Z-A*, no. 15, pp. 473–480; August, 1959.

where

k = Boltzmann's constant
T = temperature in °K
e = electronic charge.

Because of the ohmic voltage drop $I_B R_B$ at the base resistance R_B, the voltage U_{EB} at the exterior-base contact of a practical transistor must be higher than U_{EB}^* and from (1) it follows that

$$I_B = I_0 \exp \frac{U_{EB} - I_B R_B}{U_T}. \tag{3}$$

The original exponential law is linearized by the base resistance R_B (Fig. 3).[8,9]

The current amplification factor $\beta = dI_C/dI_B$ gives a relation between the collector current I_C and the modulating base current I_B; it depends on the working point of the transistor.[10] In the derivation, it is assumed that, for every working point, the current amplification factor β may have another value, but in its immediate vicinity β is taken approximately constant.

$$I_E \sim I_C \sim I_B = I_0 \exp \frac{U_{EB} - I_B R_B}{U_T}; \quad U_{EC} = \text{const.} \tag{4}$$

To calculate the nonlinear distortions, the modulation characteristic is approached by a Taylor-series formula around the fixed working point (Appendix); the first differentiations to the base voltage U_{EB} follow from (4):

$$I_B \exp \frac{I_B R_B}{U_T} = I_0 \exp \frac{U_{EB}}{U_T}, \tag{5}$$

$$\frac{dI_B}{dU_{EB}} = I_B' = \frac{I_B}{U_T + I_B R_B}, \tag{6}$$

$$\frac{d^2 I_B}{dU_{EB}^2} = I_B'' = \frac{U_T I_B}{(U_T + I_B R_B)^3}, \tag{7}$$

$$\frac{d^3 I_B}{dU_{EB}^3} = I_B''' = \frac{U_T^2 - 2U_T I_B R_B}{(U_T + I_B R_B)^5} I_B. \tag{8}$$

Considering the internal resistance of the signal source R_i (Fig. 1), we obtain

$$I_B = I_0 \exp \frac{U - I_B(R_i + R_B)}{U_T}. \tag{9}$$

The derivation of the following equations is made for the case $R_i = 0$. Considering a final internal resistance of the signal source R_i, R_B is to be replaced by the sum $(R_i + R_B)$ and U_{EB} by U (Fig. 1) in the corresponding equations.

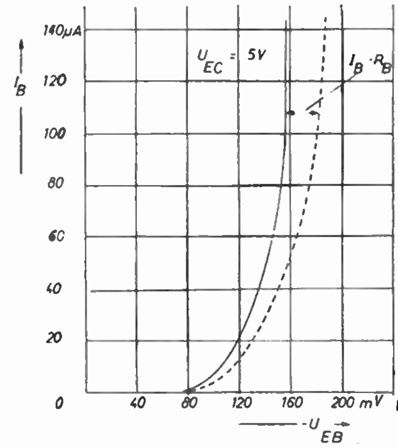

Fig. 3—I_B U_{EB} characteristic. Solid line = original curve. Broken line = linearized curve.

II. Calculation of Third-Harmonic Distortion

In order to calculate the harmonic distortion, a nonmodulated signal is assumed between emitter and base, as follows:

$$u_{EB} = \hat{U}_1 \cos \omega_1 t. \tag{10}$$

The relative third-harmonic distortion, expressed by the amplitude ratio of third-harmonic A_3 to the fundamental A_1, is approximately equal to the nonlinear harmonic-distortion factor k_3, and using the mentioned simplification (Appendix), (43) yields the desired distortion.

$$\frac{A_3}{A_1} = \frac{|I_B'''|\hat{U}_1^2}{24 I_B'} \sim k_3, \tag{11}$$

or with regard to (6) and (8),

$$\frac{A_3}{A_1} = \frac{\hat{U}_1^2 U_T}{24}\left(\frac{|U_T - 2I_B R_B|}{(U_T + I_B R_B)^4}\right) \sim k_3. \tag{12}$$

Eq. (12) will be discussed further in Section V.

III. Distortion of Modulation and Alteration of the Degree of AM

In a selective RF amplifier with an AM input signal between emitter and base,

$$u_{EB} = \hat{U}_2(1 + m \cos \omega_n t) \cos \omega_2 t, \tag{13}$$

the modulation is distorted and the degree of modulation is alterated, because the modulation characteristic of a transistor is nonlinear. The relative harmonic distortions of the modulation are expressed by the following definition:

$$d_n = \frac{\text{amplitude with angular frequency } n\omega_n}{\text{amplitude with angular frequency } \omega_n}.$$

[8] C. W. Mueller and J. I. Pankove, "A p-n-p triode alloy junction transistor for radio-frequency amplification," *RCA Transistor 1*, pp. 189–201; March, 1956.

[9] G. Meyer-Brötz and K. Felle, "Die nichtlinearen Verzerrungen im Transistorverstärker," *Elektron. Rundschau*, no. 11, pp. 297–301; October, 1957.

[10] W. M. Webster, "On the variation of junction transistor current amplification factor with emitter current," Proc. IRE, vol. 42, pp. 914–920; June, 1954,

At small-signal amplitudes $\hat{U}_2$, (45) yields

$$d_2 = \frac{3}{16}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} mU_T\hat{U}_2^2, \qquad (14)$$

$$d_3 = \frac{1}{32}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} m^2U_T\hat{U}_2^2. \qquad (15)$$

The new depth of modulation follows also from (45):

$$M = \frac{1 + \frac{3}{8}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} U_T\hat{U}_2^2\left(1 + \frac{m^2}{4}\right)}{1 + \frac{1}{8}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} U_T\hat{U}_2^2\left(1 + \frac{3m^2}{2}\right)} m. \qquad (16)$$

The relative alteration of the depth of modulation,

$$\frac{M - m}{m} = \frac{\frac{1}{4}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} U_T\left(1 - \frac{3m^2}{8}\right)}{1 + \frac{1}{8}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} U_T\hat{U}_2^2\left(1 + \frac{3m^2}{2}\right)} \hat{U}_2^2. \qquad (17)$$

At small-signal amplitudes and at depths of modulation m small with respect to unity, we obtain

$$\frac{M - m}{m} = \frac{1}{4}\cdot\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} U_T\hat{U}_2^2. \qquad (18)$$

Eqs. (14), (15), and (18) yield the nonlinear effects mentioned above. They will be discussed further in Section V.

IV. Calculation of the Cross Modulation

A nonmodulated and an AM signal are assumed between emitter and base:

$$u_{EB} = \hat{U}_N \cos \omega_N t + \hat{U}_S\cdot(1 + m_S \cos \omega_{nS} t) \cos \omega_S t. \qquad (19)$$

Eq. (19) with the new nomenclature $\hat{U}_N$ and ω_N for the desired signal, and $\hat{U}_S$, m_S, ω_{nS}, and ω_S for the interfering signal, fits (40).

We shall assume that the output of the transistor is tuned to the angular frequency ω_N. Thus, only terms of angular frequency ω_N and of frequencies in its immediate vicinity are taken into account. Using the simplification mentioned in the Appendix and neglecting the harmonics of modulation, we obtain the following expression for the ac-collector current from (47):

$$i_C \text{ prop } \hat{U}_N I_B'\left(1 + \frac{1}{2}\cdot\frac{I_B'''}{I_B'} m_S \cdot \hat{U}_S^2 \cos \omega_{nS} t\right) \cos \omega_N t, \qquad (20)$$

with the degree m_k of cross modulation

$$m_k = \frac{1}{2}\,\frac{|I_B'''|}{I_B'} m_S\hat{U}_S^2, \qquad (21)$$

or with regard to (6) and (8),

$$m_k = \frac{1}{2}\,\frac{|U_T - 2I_BR_B|}{(U_T + I_BR_B)^4} m_SU_T\hat{U}_S^2. \qquad (22)$$

For electronic tubes it is usual to represent the properties of cross modulation by the dependence on the working point of the effective interfering voltage U_S for one per cent cross modulation. For a transistor we obtain from (22) a corresponding relationship:

$$U_S = \frac{0.1}{\sqrt{m_SU_T}}\,\frac{(U_T + I_BR_B)^2}{\sqrt{|U_T - 2I_BR_B|}}. \qquad (23)$$

V. Discussion of Theoretical Results in the Case of Cross Modulation

A comparison of (12), (14), (15), (18), and (22) shows that these third-order distortions, as in electronic tubes, are closely connected, and that if one of these is measured, the others are known. Neglecting the base resistance R_B or considering small base emitter currents, these equations are very simple:

$$\frac{A_3}{A_1} = \frac{1}{24}\frac{\hat{U}_1^2}{U_T^2} \qquad (24)$$

$$d_2 = \frac{3}{16} m\cdot\frac{\hat{U}_2^2}{U_T^2}, \qquad (25)$$

$$d_3 = \frac{1}{32} m^2 \frac{\hat{U}_2^2}{U_T^2}, \qquad (26)$$

$$\frac{M - m}{m} = \frac{1}{4}\,\frac{\hat{U}_2^2}{U_T^2}, \qquad (27)$$

$$m_k = \frac{1}{2} m_S \frac{\hat{U}_S^2}{U_T^2}, \qquad (28)$$

and agree with the corresponding equations of Akgün and Strutt.[11,12]

Fig. 4 shows the interfering voltage for one per cent cross modulation calculated by (23) for the example $R_B = 104$ ohms. $U_T = 26$ mv and $m_S = 100$ per cent. At very small currents this interfering voltage amounts to 0.1 U_T with (28); with increasing base current it arises first slowly, then faster and faster. At the working point (base current),

$$I_{B0} = \frac{U_T}{2R_B}; \qquad (29)$$

this function has a pole, according to (23). At this point, there are no third-order distortions and the third differentiation (8) changes the sign. After passing the pole, the function decreases to about $U_S = 10$–11 mv, but

[11] M. Akgün and M. J. O. Strutt, "Nichtlineare Verzerrungen einschließlich Kreuzmodulation in Hochfrequenz-Transistorstufen," *Arch. Elektr. Übertragung*, vol. 13, pp. 227–242; June, 1959.

[12] M. Akgün and M. J. O. Strutt, "Cross modulation and nonlinear distortion in RF transistor amplifiers, IRE Trans. on Electron Devices, vol. ED-6, pp. 457–467; October, 1959.

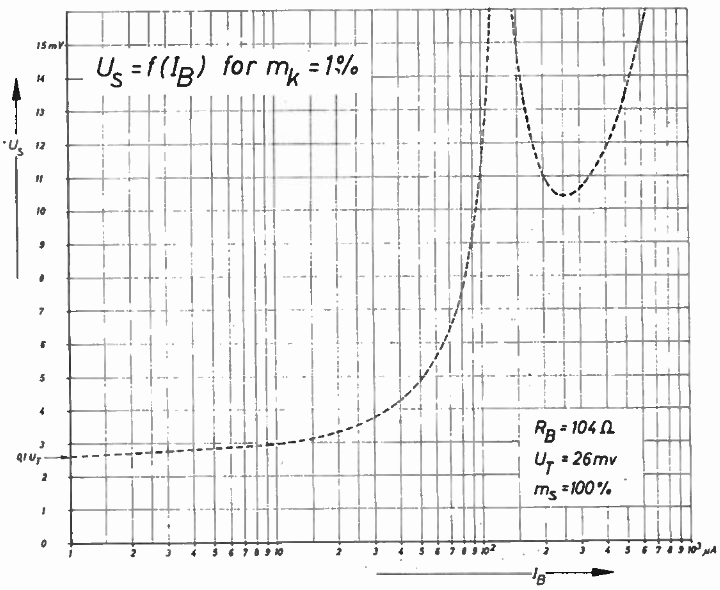

Fig. 4—Calculated dependence of interfering voltage for 1 per cent cross modulation on base current.

with an increasing base current it rises again very quickly because in the numerator of (23) the product $I_B R_B$ rises more and more. The pole of the function of interfering voltage and the zero point of the function of cross modulation, is to be explained by the linearization of the modulation characteristic owing to the base resistance R_B (Fig. 3). Therefore, the first differentiation of the modulation characteristic, the function of transconductance, has a point of inflection where the third differentiation disappears.[13]

In my own measurements of Telefunken alloy-junction transistors, OC 612 and OC 613, which are replaced by the secondary type AF-101, the interfering voltage for one per cent cross modulation was measured up to a value smaller than 7 mv to avoid overmodulation. Eq. (12) is also adequate to low frequencies and we obtain with (22)

$$m_k = 12k_3. \tag{30}$$

The function of interfering voltage for one per cent cross modulation and different working points (emitter currents) can also be calculated from the measured third harmonic AF-distortion factor by means of (30). Figs. 5 and 6 show a comparison between interfering voltages for two different transistors, calculated with (30) from the measured distortion factor k_3 at 1 kc and (23).

Considering the final internal resistance, of the signal source, we replace R_B by the sum $(R_i + R_B)$ in (12), (14), (15), (18), (22), and (23). With the same character of curve, the functions (22) and (23) appear to be similarly scaled down, considering a final internal resistance of the signal source (compare Figs. 5 and 6). At higher emitter currents the absolute value of the second minimum in the function of interfering voltage (23), and the second maximum in the function of the degree of cross modulation (22), is not influenced by the internal resistance of the signal source. These properties are confirmed by the measurements of Akgün and Strutt.[11,12]

At higher frequencies (for instance, medium wave range), the distortions of a transistor are determined by the signal voltage at the interior base contact B^* (Fig. 2) because of the partition dependent on frequency and working point between base resistance and parallel connection of emitter admittance and diffusion capacity. The RF signal (desired and interfering voltage) at the exterior base contact B is only effective by a part at the interior base emitter B^*E. Therefore, the voltage at the exterior base contact, referred to the same effective signal at B^*E, increases with the frequency. In Figs. 5 and 6 the curves measured at RF are also plotted. At very small emitter currents these curves coincide with the ones which are calculated by (23) and with the measured AF-distortion factor k_3; however, they rise sooner with increasing emitter current. This result is confirmed by Fig. 7, where the interfering voltage for one per cent cross modulation is plotted dependent on the frequency of the interfering signal for three working points (emitter currents). At small emitter currents, the interfering voltage is nearly independent of frequency. At higher emitter currents, it increases with the frequency of the interfering signal, because the diffusion capacity and emitter admittance also increase with the emitter current.

It seems that the base resistance R_B is favorable, because it linearizes the modulation characteristic and thereby diminishes the distortions; however, it diminishes the RF amplification at the same time.

[13] H. Lotsch, "Untersuchung des Kreuzmodulationsverhaltens von HF-Transistoren," *Elektron. Rundschau*, no. 8, pp. 290–294; August, 1959.

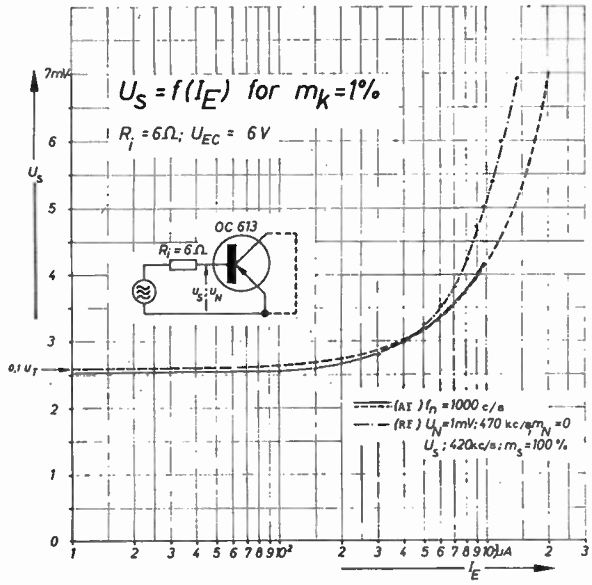

Fig. 5—Comparison between the effective interfering voltage measured by RF (–·–), and by AF (——); and the calculated one (– – –) for 1 per cent cross modulation at voltage modulation.

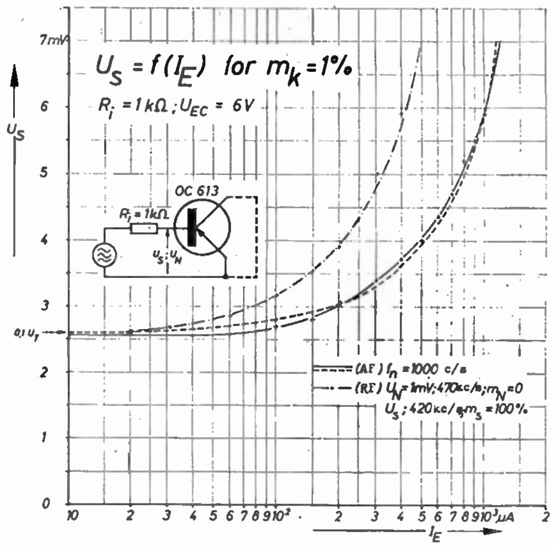

Fig. 6—Comparison between the effective interfering voltage measured by RF (–·–), and by AF (——), and the calculated one (– – –) for 1 per cent cross modulation with an internal resistance of signal source (1/$k\Omega$).

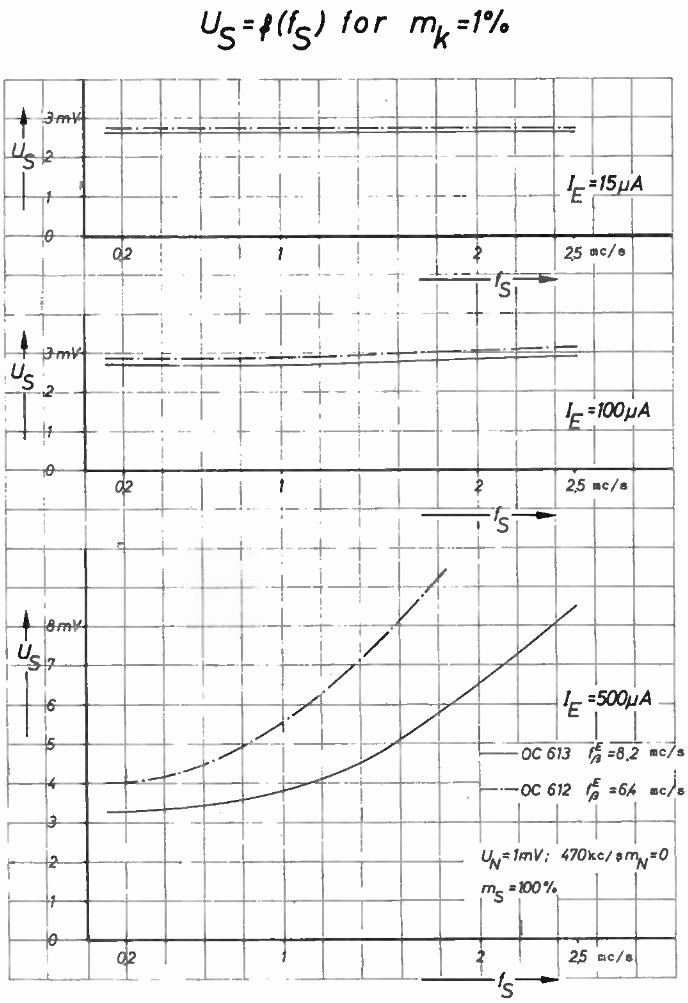

Fig. 7—Effective interfering voltage for 1 per cent cross modulation vs frequency of the spurious signal. $U_{EC} = 6$ v; voltage modulation.

In their measurements, Akgün and Strutt[11,12] have determined the dependence of the degree of cross modulation on the working points with a constant amplitude of the interfering signal; from these are calculated the interfering voltage for one per cent cross modulation, depending on the working points. Thus, the function at higher amplitudes of interfering signal, as the pole, could be found out. In Fig. 8 a curve measured by these authors and a theoretical one according to (23) are compared. Both functions show the same character; the difference in the ordinate is caused by the mentioned partition of RF voltage between base resistance and parallel connection of diffussion capacity and emitter admittance.

Considering an internal resistance of signal source with (22) and (23), one assumes that at very high values of internal resistance of signal source (current modulation), and with emitter currents that are not too small, no cross modulation occurs within the working range of practical interest.

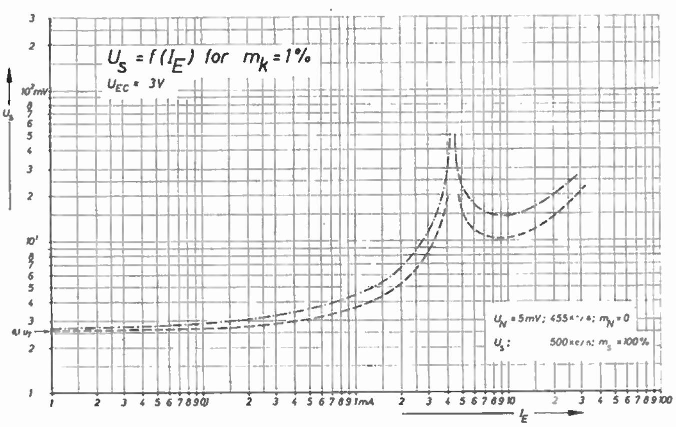

Fig. 8—Effective interfering voltage for 1 per cent cross modulation vs emitter dc current; —·— measured by Akgün and Strutt;[11] - - - calculated by (23).

VI. Reduction of Cross Modulation by Using Push-Pull Modulation

The present investigations show that only a few millivolts of a 100 per cent modulated effective interfering voltage causes one per cent cross modulation within a transistor, while in electronic tubes this value is about 100 times larger. Because of the relatively low input resistance in grounded-emitter circuits, the base is only modulated by a part of the resonant circuit voltage. Although this voltage transformation is taken into account, the transistor is still more susceptible to cross modulation. Using methods similar to that in vacuum-tube amplifiers to displace the working point into regions of decreased amplification, such as decreased currents for an automatic gain control in a transistor amplifier, the transistor has to work with a larger input signal, and thus becomes more sensitive to cross modulation.

In the following, a method is described by which the cross modulation of a given transistor can be diminished greatly without deteriorating its quality of amplification. This consists of push-pull modulation by using the demodulated AF current.[14,15]

Eq. (41j) shows that the ac collector current i_C contains the following component:

$$i_C = \cdots + \frac{I_B'' m_S \hat{U}_S^2}{2} \cos \omega_{nS} t \cdots . \tag{31}$$

This is a poor AF current, containing only the frequency $\omega_{nS}/2\pi$ of modulation. Within the output circuit of the transistor, this component is removed completely by the electric-wave filter; therefore it has no efficacy.

In order to compensate the cross modulation, the desired signal is modulated by this component of the output circuit current. At the base current determined by (29), there is no cross modulation, since the function of interfering voltage for one per cent cross modulation has a pole, and the third differentiation (8) changes its sign. Compensation at a fixed working point and its vicinity consists in displacing this pole into the working point. Therefore, a resistance R_{EZ} (Fig. 9) is inserted in the emitter lead, and is bypassed by a small condenser C_3. The capacity of C_3 is large enough so that the RF currents cause no appreciable voltage drop at the resistance R_{EZ}, but small enough so that the voltage drop caused by the AF currents (31) is hardly attenuated.

$$u_3 = \hat{U}_3 \cos \omega_{nS} t \approx R_{EZ} \beta I_{BnS} \cos \omega_{nS} t. \tag{32}$$

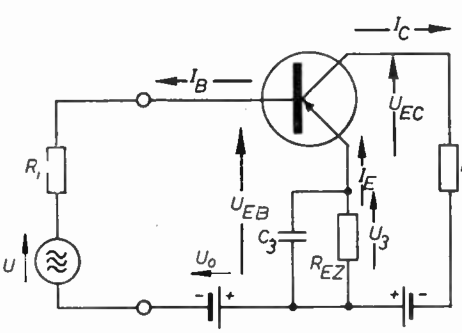

Fig. 9—Transistor circuit, indicating positive directions of alternating currents and voltages, in the case of push-pull modulation.

The resistance R_{EZ} is simultaneously wired within the base circuit of the transistor; the AF voltage is opposite the input signal (Fig. 9). The whole modulation signal at the base of the transistor is composed by (19) and (32):

$$u_{EB} = \hat{U}_N \cos \omega_N t + \hat{U}_S \cdot (1 + m_S \cos \omega_{nS} t) \cos \omega_S t - U_3 \cos \omega_{nS} t. \tag{33}$$

The approximate degree m_{kg} of cross modulation in which the push-pull modulation is considered, follows from (49):

[14] E. Hudec, "Kreuzmodulation und Eingangsrauschspannung," *Elek. Nachr. Tech.*, vol. 20, pp. 123–135; 1943.

[15] H. Lotsch, "Übersicht über die nichtlinearen Verzerrungen in Transistorstufen, einschließlich der Kreuzmodulation," *Arch. Elektr. Übertragung*, vol. 14, pp. 204–216; May, 1960.

$$m_{kg} = \frac{1}{2I_B'} (I_B''' m_S \hat{U}_S^2 - 2I_B'' \hat{U}_3). \tag{34}$$

With (6), (7), (8), (32), the amplitude I_{BnS} of the AF current (41j) or (31), (34) is transcribed as follows:

$$m_{kg} = \frac{1}{2} \frac{m_S U_T \hat{U}_S^2}{(U_T + I_B R_B)^4} \cdot \left(U_T - I_B\left(2R_B + \frac{\beta R_{EZ} U_T}{(U_T + I_B R_B)}\right)\right). \tag{35}$$

The effective interfering voltage for one per cent cross modulation follows from (35):

$$U_{Sg} = \frac{0,1}{\sqrt{m_S U_T}} \cdot \frac{(U_T + I_B R_B)^2}{\sqrt{\left| U_T - I_B\left(2R_B + \frac{\beta R_{EZ} U_T}{U_T + I_B R_B}\right)\right|}}. \tag{36}$$

Eq. (29), being available for the pole of the function of interfering voltage without push-pull modulation, shows that, in the denominator of (36) and in the numerator of (35), the product $I_B R_B$, can be approximately neglected within the range $I_B = 0$ to I_{B0}.

$$U_S = \frac{0,1}{\sqrt{m_S U_T}} \frac{(U_T + I_B R_B)^2}{\sqrt{| U_T - I_B \beta R_{EZ} |}}. \tag{37}$$

The pole displaced by the resistance R_{EZ} in the emitter lead is approximately at base dc current according to (37):

$$I_{B0g} = \frac{U_T}{\beta R_{EZ}}. \tag{38}$$

With decreasing emitter current, the current amplification factor β decreases also.[10] When the pole is displaced to smaller emitter currents, the resistance R_{EZ} increases more than in an inverse proportion because of βR_{EZ}. This result is verified by experiment Fig. 10. In order to simplify these measurements, they were made with different emitter currents. The theoretical derivation is based on nonlinearity of the short-circuit input resistance, that is, the base current.

The second increase of the function of interfering voltage at higher emitter current, as in (23), is given by the quadratic increase of the numerator (36), so that it is scarcely influenced by the resistance R_{EZ}.

Considering the internal resistance R_i of the signal source (9), R_B is replaced by the sum $(R_i + R_B)$ in (35) and (36). A comparison of (23) and (36) or (22) and (35) shows that a resistance in the base lead (internal resistance of the signal source) and a resistance in the emitter lead have very different influences in relation to the properties of cross modulation of a transistor. A resistance in the base lead only compresses the scale in the abscissa, while the character of the function is retained; but a resistance blocked at HF, but not at LF in the emitter lead, only displaces the pole and the zero point, respectively, within the range $I_B = 0$ to I_{B0} and has little effect at higher currents.

For a controlled RF stage, it would be very advantageous, if displacing the working point would accurately shift the pole of the interfering voltage, but the dependence of the current amplification factor β on the working point gives some trouble. A diode poled in the range of free transmission as a variable resistance R_{EZ} in the emitter lead[15] improves the behavior of cross modulation of a transistor amplifier rather considerably (Fig. 11). A comparison with Fig. 10 shows that with a given interfering voltage, the distance between both slopes of the curve near the pole is enlarged. The resistance R_{EZ} in the emitter lead also influences the modulation of the desired AM signal. This influence, however, is negligibly small.[15]

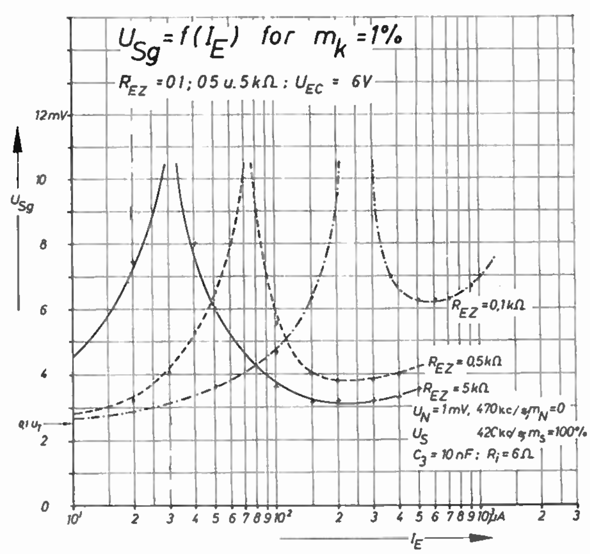

Fig. 10—For a transistor-type OC 613; measured effective interfering voltage for 1 per cent cross modulation in the case of push-pull modulation.

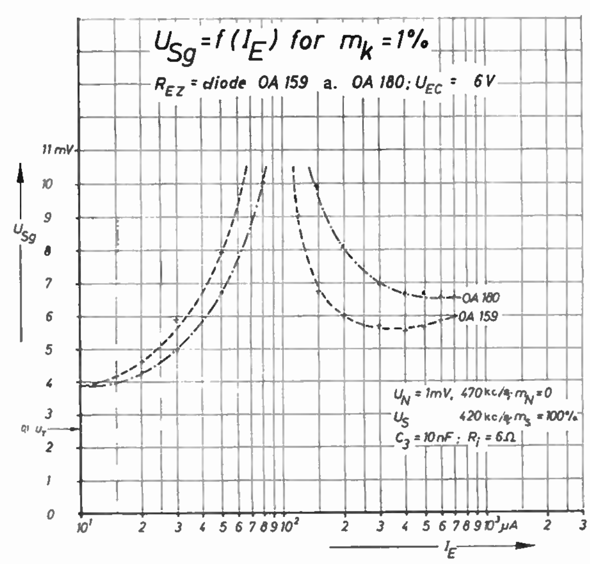

Fig. 11—For a transistor-type OC 613; measured effective interfering voltage for 1 per cent cross modulation in the case of the push-pull modulation (with a diode in the emitter lead poled in the range of free transmission).

VII. Reduction of Distortions by Predistortion

The automatic gain control causes certain troubles in a transistor RF amplifier. In order to avoid these problems, Cantz,[16] using the method of a controlled attenuator,[17] has developed an interesting circuit (Fig. 12) in which the distortions are diminished together. The diode works with the control current $I_R = I_{Di}$ in the range of free transmission and has a very high resistance when uncontrolled, but a very small resistance when fully controlled, so that the partition of voltage within the range of control changes considerably. In order to illustrate compensation of distortion, the fully controlled method of operation is discussed because it gives the best results. The resistance R_{Di} of the diode is small compared to the resistance Z_L and the input resistance of the transistor. Therefore, a sinusoidal source current I_{An} causes a distorted voltage (Fig. 13) at the nonlinear resistance of the diode. This predistorted voltage $u_{Di} = u_{EB}$ is corrected by a "replica" at the $I_C - U_{EB}$ characteristic of the transistor. Because of the property of an exponential law, such a compensation for arbitrary working points of the diode and transistor is available within the entire range in which the exponential law holds with satisfactory exactness. The measurements (Fig. 14) illustrate the quality of this control circuit. However they show also that due to the partition of voltage, the maximum voltage gain is smaller. In order to reduce the control power, Cantz[16] has also suggested a circuit in which a transistor working as a dc amplifier substitutes for the diode.

At very high frequencies the assumed conditions do not apply and the RF transconductance of the transistor,[3] which depends on frequency here, has to be considered.[18]

Investigations of cross modulation in transistor amplifiers by Holmes,[19] Akgün and Strutt,[11,12] and Nieveen van Dijkum and Sips[20] have been published. These authors obtained the same experimental results, but their theoretical calculation is more or less different.

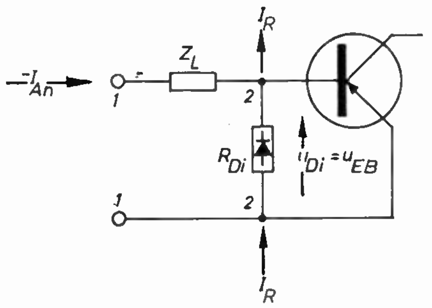

Fig. 12—Schematic circuit of the controlled attenuator.[16]

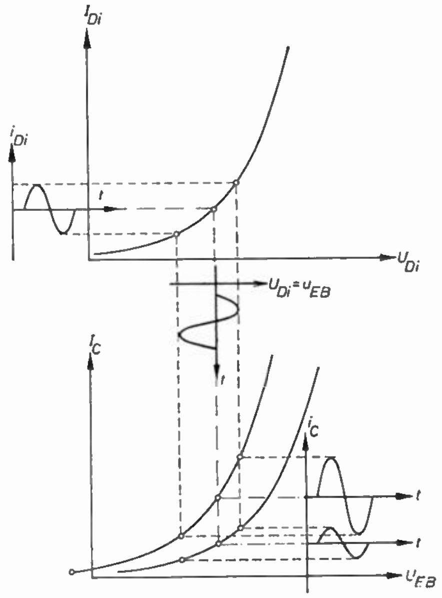

Fig. 13—"Replica-diagram" of a diode and a transistor characteristic for illustrating compensation of distortion.[16]

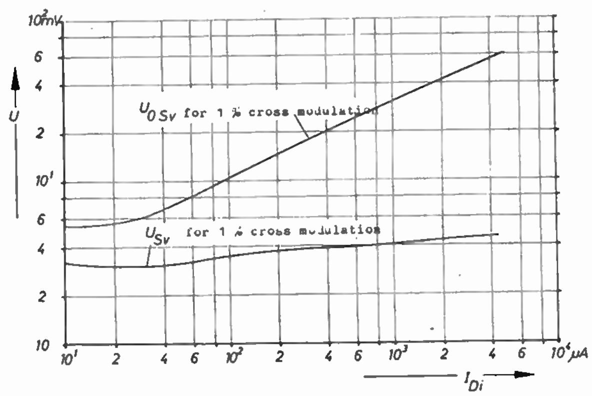

Fig. 14—Effective interfering voltage for 1 per cent cross modulation in the circuit of Fig. 12. [U_{0Sv} = interfering voltage at the input of the attenuator (both points 1); U_{Sv} = interfering voltage at the input of the transistor (both points 2).]

Appendix

The collector ac current i_C is proportional to the base ac current and therefore dependent on the instantaneous voltage u_{EB} at the base of the transistor and on the fixed working point U_{EB0}. This function is approached by a Taylor-series formula at the working point U_{EB0}, and its immediate vicinity.[21]

[16] R. Cantz, "Eine Schwundregelschaltung mit Diode für Transistorempfänger," *Die Telefunken-Röhre*, vol. 35, pp. 31–42; September, 1958.

[17] E. Klotz, Deutsches Patent No. DRP 690 807; January 15, 1940.

[18] W. Minner, "Die Steilheit des HF-Legierungs- und Drifttransistors als Funktion der Frequenz und des Arbeitspunktes und ihre Ableitungen," *Arch. Elektr. Übertragung*, vol. 14, pp. 411–420; September, 1960.

[19] D. D. Holmes, "Cross-modulation in transistor-radio-frequency amplifiers," *RCA Transistor I*, pp. 422–430; March, 1956.

[20] H. J. Nieveen van Dijkum and J. J. Sips, "Cross-modulation and modulation distortion in AM receivers equipped with transistors," *Electronic Applications*, vol. 20, no. 3, pp. 107–127; 1959/1960.

[21] G. A. Spescha, and M. J. O. Strutt, "Theoretische und experimentelle Untersuchung der Verzerrungen in Niederfrequenz-Flächen transistor-Vierpolen," *Arch. Elektr. Übertragung*, vol. 11, pp. 307–320; August, 1957.

$$i_C \text{ prop } \left(I_B'(U_{EB0})u_{EB} + \frac{I_B''(U_{EB0})}{2!}u_{EB}^2 + \frac{I_B'''(U_{EB0})}{3!}u_{EB}^3 + \cdots\right) \quad (39)$$

$$I_B' = \frac{dI_B}{dU_{EB}}; \qquad I_B'' = \frac{d^2I_B}{dU_{EB}^2}; \qquad I_B''' = \frac{d^3I_B}{dU_{EB}^3}$$

are the first derivatives of the base current at the working point U_{EB0}. A nonmodulated ($\hat{U}_1$; ω_1) and an amplitude-modulated ($\hat{U}_2$; m; ω_2; ω_n) signal are assumed between emitter and base.

$$u_{EB} = \hat{U}_1 \cos \omega_1 t + \hat{U}_2(1 + m \cos \omega_n t) \cos \omega_2 t. \quad (40)$$

With (39) and (40), we obtain the instantaneous components of the collector ac current:

$$i_C \text{ prop } \left\{\cos \omega_1 t \left[I_B' + \frac{I_B'''}{4}\left(\frac{\hat{U}_1^2}{2} + \hat{U}_2^2\left(1 + \frac{m^2}{2} + 2m \cos \omega_n t + \frac{m^2}{2} \cos 2\omega_n t\right)\right)\right]\hat{U}_1 \right. \quad (41a)$$

$$+ \cos \omega_2 t \left[I_B' + \frac{I_B'''}{4}\left(\hat{U}_1^2 + \frac{\hat{U}_2^2}{2(1 + m \cos \omega_n t)}\left(1 + \frac{3m^2}{2} + 3m\left(1 + \frac{m^2}{4}\right)\cos \omega_n t + \frac{3}{2}m^2 \cos 2\omega_n t + \frac{m^3}{4}\cos 3\omega_n t\right)\right)\right](1 + m \cos \omega_n t)\hat{U}_2 \quad (41b)$$

$$+ \cos 2\omega_1 t \left[\frac{I_B''}{4}\right]\hat{U}_1^2 \quad (41c)$$

$$+ \cos 2\omega_2 t \left[\frac{I_B''\hat{U}_2^2}{4}\left(1 + \frac{m^2}{2} + 2m \cos \omega_n t + \frac{m^2}{2}\cos 2\omega_n t\right)\right. \quad (41d)$$

$$+ \cos 3\omega_1 t \left[\frac{I_B'''}{24}\right]\hat{U}_1^3 \quad (41e)$$

$$+ \cos 3\omega_2 t \left[1 + \frac{3m^2}{2} + 3m\left(1 + \frac{m^2}{4}\right)\cos \omega_n t + \frac{3}{2}m^2 \cos 2\omega_n t + \frac{m^3}{4}\cos 3\omega_n t\right]\frac{I_B'''}{24}\hat{U}_2^3 \quad (41f)$$

$$+ \cos \omega_1 t \cos \omega_2 t [1 + m \cos \omega_n t] I_B''\hat{U}_1\hat{U}_2 \quad (41g)$$

$$+ \cos \omega_1 t \cos 2\omega_2 t \left[1 + \frac{m^2}{2} + 2m \cos \omega_n t + \frac{m^2}{2}\cos 2\omega_n t\right]\frac{I_B'''}{4}\hat{U}_1\hat{U}_2^2 \quad (41h)$$

$$+ \cos 2\omega_1 t \cos \omega_2 t [1 + m \cos \omega_n t]\frac{I_B'''}{4}\hat{U}_1^2\hat{U}_2 \quad (41i)$$

$$+ \cos \omega_n t \left[\frac{I_B''m}{2}\right]\hat{U}_2^2 \quad (41j)$$

$$\left. + \cos 2\omega_n t \left[\frac{I_B''m^2}{8}\right]\hat{U}_2^2 + \cdots \right\}. \quad (41k)$$

In order to calculate the different distortions, the generally valid equation (41) is specialized.

Third-Harmonic Distortion

$$u_{EB} = \hat{U}_1 \cos \omega_1 t; \qquad \hat{U}_2 = 0 \quad (42)$$

$$i_C \text{ prop } \hat{U}_1\left[\left(I_B' + \frac{I_B'''\hat{U}_1^2}{8}\right)\cos \omega_1 t + \left(\frac{I_B''\hat{U}_1}{4}\right)\cos 2\omega_1 t + \left(\frac{I_B'''}{24}\hat{U}_1^2\right)\cos 3\omega_1 t\right]. \quad (43)$$

At a small modulation voltage in (43), $(I_B'''/8)\hat{U}_1^2$ can be neglected relative to I_B'.

Distortion of Modulation and Alteration of the Degree of AM

$$u_{EB} = \hat{U}_2 \cdot (1 + m \cos \omega_n t) \cos \omega_2 t; \qquad \hat{U}_1 = 0 \quad (44)$$

$$i_C \text{ prop } \hat{U}_2\left[I_B' + \frac{I_B'''}{8}\hat{U}_2^2\left(1 + \frac{3m^2}{2}\right) + \left(I_B' + \frac{3}{8}I_B'''\hat{U}_2^2\left(1 + \frac{m^2}{4}\right)\right)m \cos \omega_n t + \frac{3}{16}I_B'''m^2\hat{U}_2^2 \cos 2\omega_n t + \frac{1}{32}I_B'''m^3\hat{U}_2^2 \cos 3\omega_n t\right]\cos \omega_2 t. \quad (45)$$

At small modulation voltage in (45),

$$\frac{3}{8}I_B'''\hat{U}_2^2\left(1 + \frac{m^2}{4}\right)$$

can be neglected relative to I_B'.

Calculation of Cross Modulation

$$u_{EB} = \hat{U}_N \cos \omega_N t + \hat{U}_S(1 + m_S \cos \omega_{nS} t) \cos \omega_S t. \quad (46)$$

Only the frequency $\omega_N/2\pi$ and its immediate vicinity are considered:

$$i_C \text{ prop } \hat{U}_N\left[I_B' + \frac{I_B'''}{4}\left(\frac{\hat{U}_N^2}{2} + \hat{U}_S^2\left(1 + \frac{m_S^2}{2} + 2m_S \cos \omega_{nS} t + \frac{m_S^2}{2}\cos 2\omega_{nS} t\right)\right)\right]\cos \omega_N t. \quad (47)$$

In (47) the expression

$$\frac{I_B'''}{4}\left(\frac{\hat{U}_N^2}{2} + \hat{U}_S^2\left(1 + \frac{m_S^2}{2}\right)\right)$$

can be neglected relative to I_B'.

Reduction of Cross Modulation by Using a Push-Pull Modulation

$$u_{EB} = \hat{U}_N \cos \omega_N t + \hat{U}_S(1 + m_S \cos \omega_{nS} t) \cos \omega_S t - \hat{U}_3 \cos \omega_{nS} t. \quad (48)$$

Putting (48) in (39) and considering only the frequency $\omega_N/2\pi$ and its immediate vicinity, we obtain for the collector ac current, if we neglect, as in (47)

$$\frac{I_B'''}{4}\left(\frac{\hat{U}_N^2}{2} + \hat{U}_S^2\left(1 + \frac{m_S^2}{2}\right)\right) \text{ relative to } I_B',$$

$$i_C \text{ prop } \hat{U}_N I_B' \left[1 + \frac{1}{I_B'}\left(\tfrac{1}{2} I_B''' m_S \hat{U}_S^2 - I_B'' \hat{U}_3\right) \cos \omega_{nS} t\right] \cos \omega_N t. \quad (49)$$

LIST OF SYMBOLS

A_1; A_3 = amplitude of the fundamental and third harmonic in the ac collector current.
B = exterior base.
B^* = interior base.
C = exterior collector.
C_D = diffusion capacity.
C_3 = capacitor, which at RF currents bypasses the resistance in the emitter-lead.
d_2; d_3 = relative second- and third-harmonic distortions in the modulation.
E = exterior emitter.
f_S = frequency of the spurious signal.
f_β^E = cutoff frequency of a transistor in a grounded emitter circuit.
G_D = emitter admittance.
I_{An} = sinusoidal source current
I_B = base current.
I_B'; I_B''; I_B''' = first, second, and third derivatives of the base current with respect to the base voltage.
I_{B0} = base dc current (the third-order distortions are cancelled).
I_{B0_g} = base dc current (considering the push-pull modulation, the cross modulation is cancelled).
I_{BnS} = amplitude of the AF component of the base current.
I_C = collector current.
I_{Di} = diode current.
I_E = emitter current.
I_R = control current.
i_C = instantaneous value of the emitter ac current.
k_3 = nonlinear harmonic-distortion factor.
M = new degree of AM after alteration of the degree of AM.
m = degree of AM.
m_S = degree of AM of the spurious signal.
m_k = degree of cross modulation.
m_{kg} = degree of cross modulation in the case of push-pull modulation.
R_B = base lead resistance of the transistor.
R_{Di} = resistance of the diode.
R_{EZ} = extra resistance in the emitter lead.
R_a = extra resistance of the output circuit.
R_i = internal resistance of the signal source.
t = time.
U_{EB} = emitter-exterior base voltage.
U_{EB}^* = emitter-interior base voltage.
U_{EC} = emitter-collector voltage.
U_{Di} = voltage drop at the diode.
$\hat{U}_N$ = amplitude of the desired signal.
U_S = effective interfering voltage for 1 per cent cross modulation.
$\hat{U}_S$ = amplitude of the spurious signal.
U_{Sg} = effective interfering voltage for 1 per cent cross modulation in the case of push-pull modulation.
U_{Sv} = effective interfering voltage for 1 per cent cross modulation at the input of the transistor in the case of predistortion.
U_{EB0} = dc current for fixing the working point.
$U_T = kT/e$.
U_{0Sv} = effective interfering voltage for 1 per cent cross modulation at the input of the controlled attenuator in the case of predistortion.
$\hat{U}_1$ = amplitude of the nonmodulated signal.
$\hat{U}_2$ = amplitude of the AM signal.
$\hat{U}_3$ = amplitude of the AF component.
u_{EB} = instantaneous value of emitter-base voltage.
u_{Di} = instantaneous value of the voltage drop at the diode.
u_3 = instantaneous value of the AF component.
Z_L = resistance of the controlled attenuator.
β = current amplification factor in grounded emitter circuit.
ω_N = angular frequency of the desired signal.
ω_S = angular frequency of the spurious signal.
ω_n = angular frequency of the AF modulation.
ω_{nS} = angular frequency of the AF modulation of the spurious signal.
ω_1 = angular frequency of the nonmodulated signal.
ω_2 = angular frequency of the AM signal.

Contributors

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John K. Hilliard (A'25–M'29–SM'43–F'52) was born in Wyndmere, N. D., on October 22, 1901. He received the B.S. degree in physics, and did graduate work in electrical engineering at the University of Minnesota, Minneapolis. In 1951, he was awarded an honorary Doctor of Science degree in engineering from Hollywood University.

He was with the MGM studios for fourteen years. Here he developed recording and reproducing equipment for film and tape, and designed microphones and loudspeakers for theaters. He was also a project engineer on radar in the Radiation Laboratory at the Massachusetts Institute of Technology, Cambridge.

In 1943 he joined Altec Lansing Corporation, and in 1960 became Vice President and Director of Ling-Altec Research Division of Ling-Temco Electronics, Inc. Projects have included high-intensity sound environmental equipment, such as microphones for nuclear blasts, high-speed boundary layer measurements, jet and missile noise, fatigue of metals, microphones to pick up heart sounds, communication equipment for the Bell Telephone System, magnetic antisubmarine warfare equipment, and underwater microphones for very low frequencies.

Dr. Hilliard is a member of the Acoustical Society of America, the Audio Society, the Society of Motion Picture and Television Engineers, Eta Kappa Nu, the Armed Forces Committee on Hearing and Bioacoustics, and the Institute of Environmental Engineers. He is an Acoustic Consultant of the Brain Institute at the Medical School of the University of California at Los Angeles.

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B. F. LOGAN, JR.

Benjamin F. Logan, Jr. (M'57) was born in Coahoma, Tex., on June 6, 1927. He received the B.S.E.E. from Texas Technological College, Lubbock, in 1946, and the M.S. degree from Massachusetts Institute of Technology, Cambridge, in 1951. He is currently working toward the Eng. D.Sc. degree in electrical engineering at Columbia University, New York, N. Y.

During his studies at M.I.T., he was a research assistant in the Research Laboratory of Electronics investigating characteristics of high-power electrical discharge lamps. Later he was engaged in analogue computer development at the Dynamic Analysis and Control Laboratory, M.I.T., where he remained until 1955. He was then employed by Hycon-Eastern, Inc., Cambridge, Mass., where he was concerned with the design of air-borne power supplies. He joined Bell Telephone Laboratories, Murray Hill, N. J., in the summer of 1956, as a member of the Visual and Acoustics Research Department, where he currently is concerned with the processing of speech signals.

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Roy A. Long (S'48–A'50–M'56) was born in Cutler, Calif., on August 9, 1925. He received the B.S.E.E. degree from Stanford University, Stanford, Calif., in 1950.

During World War II he served with the U. S. Army Signal Corps as a radio technician and instructor. In 1949 he joined the staff of Stanford Research Institute, Menlo Park, Calif., as a research engineer in the Communication and Propagation Laboratory of the Engineering Division. He participated in research on single-sideband communication, radio propagation, speech communication, frequency standards and synthesizers, Doppler radar, and audio and acoustical devices.

Mr. Long is a Fellow of the Audio Engineering Society, the Acoustical Society of America, the Scientific Research Society of America, and the Golden Gate Sapphire Group (an association of professional sound-recording engineers). He has been Chairman of the local chapters of the PGA and AES, and President of the Golden Gate Sapphire Group.

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Helmut Lotsch was born in Bretten, Germany, on February 26, 1933. He received the Dipl. Ing. degree in electrical engineering (communications) in 1958 from the Institute of Technology of Karlsruhe, Karlsruhe, Germany.

As a student he was employed at Siemens and Halske, Karlsruhe; Standard-Elektrik-Lorenz, Pforzheim; Robert Bosch GmbH, Stuttgart; and Fernseh GmbH, Darmstadt (Germany). In 1957, he joined Telefunken GmbH in Ulm, Germany, where he was engaged in work on nonlinear distortions of a transistor amplifier-stage. At present, he is employed in the Cathode-Ray Tube Department, working towards the development of a TV picture tube with low drive signals.

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V. SALMON

Vincent Salmon (SM'46) was born in Kingston, Jamaica, on January 21, 1912. He received the B.A. and M.A. degrees in physics from Temple University, Philadelphia, Pa., in 1934 and 1936, respectively, and the Ph.D. in theoretical physics in 1938 from the Massachusetts Institute of Technology, Cambridge.

In 1939 he joined the Jensen Manufacturing Company, Chicago, Ill., doing research on electroacoustic transducers. In 1949 he joined Stanford Research Institute, Menlo Park, Calif., where he is a Senior Physicist and Manager of the Sonics Section. At the Institute he has specialized in industrial acoustics—the use of sound for testing and processing.

Dr. Salmon is a Fellow of the Acoustical Society of America (Biennial Award, 1946; Chairman, Technical Committee on Audio Engineering and Electroacoustics; and Chairman, 1960 San Francisco meeting). In the IRE he has been Chairman of the PGA, and of the 1959 National Ultrasonics Symposium. He is a Fellow of the Audio Engineering Society, and is past Western Vice President, Editor of the *Journal*, and Chairman of the local chapter. He is past president of the SRI chapter of the Scientific Research Society of America and of the Golden Gate Sapphire Group (an association of professional sound recording engineers). He has given courses in electroacoustics at Stanford University. He is a Registered Professional Engineer in Illinois and California.

Manfred R. Schroeder was born in Ahlen, Germany, on July 12, 1926. He received the degree of Diplom Physiker in 1951, and that of Dr. rer. nat. in physics in 1954, from the University of Goettingen, Germany.

M. R. SCHROEDER

In 1954, he came to the United States to join the technical staff of Bell Telephone Laboratories, Murray Hill, N. J. His work there has encompassed a variety of phases in the field of acoustics. He has been engaged in study and advancement of speech bandwidth compression and vocoders, artificial reverberation and stereophony, the theory of sound fields in rooms and new measurement methods for architectural acoustics, a method for reducing feedback in public address systems, compatible single-sideband transmission and nonsynchronous time multiplex for mobile communications. In the area of psychoacoustics, he has studied monaural phase perception and other properties of the human hearing mechanism.

He is the author of many technical articles and has been a speaker at a number of international professional conferences. He served on the National Stereophonic Radio Committee, an advisory group to the FCC, and is a member of the Armed Forces —National Research Council Committee on Hearing and Bio-Acoustics.

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He joined Illinois Institute of Technology, Chicago, in 1942, where he served as Instructor and Laboratory Supervisor in electronics until 1946. From 1946 to the present, he has been a member of the Electronics Division of Armour Research Foundation, Chicago, where he holds the position of Engineering Advisor. Much of his work has been in the field of instrumentation and circuit design. He has had experience in the design and development of magnetic devices and magnetic amplifiers, and has been active in the development of special amplifiers and modulation systems for recording and measurement purposes. Recent work has been in the application of semiconductor devices to the field of measurement and control.

Mr. Van Scoyoc is a member of Sigma Pi Sigma and Sigma Xi.

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