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Hugh S. Knowles

A Message from the Chairman

That IRE members have a major interest in audio engineering is attested to by the rapid growth of the PGA into the largest professional group. Much credit is due the founders of the PGA who recognized this interest and set about serving the needs of this group. A major need was a publication which would, with minimum delay, publish both papers on important original developments of interest to professional workers and survey and tutorial papers of general interest to engineers having a peripheral or avocational interest in the audio field. Thanks to the liberal publication policy of each successive editor and the financial support of the PGA members, the Transactions on Audio has been conspicuously successful. Many successful PGA Chapters have been formed. Technical sessions, including presentation of many informative and frequently important papers on audio topics, have been organized. These are some of the more important favorable factors in our past history. The purpose of this note is to call your attention to some symptoms of trouble.

In the last two months, I have talked to many PGA members located in most of the active audio areas throughout the country. Two major impressions have emerged. The first is that the PGA has met a real need and should continue to do so. The second is that many PGA members, including some who were formerly active, have become apathetic. It is the extent of this apathy that is disturbing. Anyone who has been active in technical society work can readily visualize the very large amount of individual and collective effort that went into making the successful history of the PGA I have so briefly summarized. Scores of people did their bit. Many performed beyond their call of duty. A few were dedicated to their work. I find that even some of the dedicated are discouraged; tired, for the most part, by the very unreasonable effort required to get members to do even a small part of the bit they agreed to do. The symptom is readily discerned. The diagnosis and prognosis are difficult. One thing seems clear.

If the membership feels there are needs the PGA should continue to meet, then realism dictates that we recognize that these were met in the past only by the active support of the members.

Your national Administrative Committee, local officers of PGA chapters and their sponsors will attempt during the coming year to evaluate the needs of the PGA and to activate programs that will meet these needs. These people are at least as busy as you and I. They need your help. I am confident they will get it if you recognize the need. If you have a suggestion, take it up in your PGA chapter or write the editors of the Transactions, Marvin Camras, or me.

The Editor's Corner

DEFINITIONS[1]

It's in process—So wrapped up in red tape that the situation is almost hopeless.

A program—Any assignment that can't be completed by one phone call.

Expedite—To confound confusion with commotion.

Channels—The trail left by interoffice memos.

Coordinator—The guy who has a desk between two expediters.

Expert—Any ordinary guy with a briefcase, more than 50 miles from home.

Informed source—The guy you just met.

Reliable source—The guy who told the guy you just met.

Unimpeachable source—The guy who started the rumor in the first place.

To activate—To make carbons and add more names to the memo.

To implement a program—Hire more people and expand the office.

Under consideration—We're looking in the files for it.

Meeting—A mass mulling by master-minds.

Conference—A place where conversation is substituted for labor and thought.

[1] *Appliance Manufacturer*, April, 1959.

The New WLW AGC Amplifier*

R. J. ROCKWELL†

Summary—Since the inception of radio broadcasting, the control of audio level has been a constant problem. This paper discusses some of the shortcomings of presently-used circuitry in limiting and automatic gain control amplifiers and describes a new approach to the problem of level control, which provides both slow gain increase and decrease along with fast limiting action while maintaining wide frequency response and extremely low distortion under all operating conditions.

Introduction

RADIO broadcasting has for its objective the faithful reproduction in the listener's home of those sensations of sound to which they would be subjected at the scene of the broadcast. A completely faithful reproduction of sounds in the home is never realized in practice, and in many cases this would be actually undesirable. Probably one of the greatest defects in currently reproduced sounds is the lack of auditory perspective. A practical approach to the problem should take into account the degree of fidelity which will provide a reasonably satisfactory reproduction of various kinds of program material in any average home. For example, it is clearly undesirable to attempt to accurately reproduce a symphony orchestra with its full 70-db volume range in the listener's home; the minimum noise level in an average residence is about 25 db above 10^{-16} watts per square cm (the noise of an average whisper at a distance of 4 feet is 20 db) so that the peak audio power of a symphony orchestra reproduced with its full volume range of 70 db in the average listener's home would be 95 db above 10^{-16} watts per square cm. This sound intensity is equivalent to the noise reproduced by a riveter at a distance of only 40 feet.

The transmitted volume range is usually less than the volume range of the original program, because the audio control operator usually adjusts gain in such a way that the fortissimo parts of the program are reduced and the pianissimo parts are increased. There are several reasons why this is done; first, the listener does not wish to reproduce a wide volume range such as encountered in a symphony orchestra, since the fortissimo parts of the program would be entirely too loud when the pianissimo portions are made just audible above room noise. Second, the permissible volume range is limited to approximately 50 db by the broadcast and receiving equipment, even in the most favorable cases.

From the above, it should be clear that a transmitted volume range of approximately 50 db would be desirable for the reproduction of program material in the home.

* Received by the PGA, January 8, 1960.
† Crosley Broadcasting Corp., Cincinnati 2, Ohio.

For these reasons, Crosley Broadcasting Engineers have been vitally interested in amplifiers which would automatically convert a 70-db volume range to a 50-db volume range without noticeably altering the expression desired by the conductor.

Many previous automatic amplifiers have been designed by Crosley over the past 25 years and almost all commercial models have been studied and tested. It is believed that the amplifier described in this report is the first real accomplishment that has been made in this area; its action definitely excels that of an expert operator who slowly raises and lowers the gain as needed, returning the control to normal during extended quiet periods and while switching programs. This slow automatic control is followed by a fast-acting automatic attenuator to suppress occasional peaks.

Past History

To the best of our knowledge, WLW was the first broadcast station to use an automatic amplifier; this was as early as 1935. This is substantiated by our early patent position in this area.

Since that date, many improvements have been incorporated in our circuitry, and several commercial types have appeared on the market.

Among these commercial types are amplifiers incorporating circuits which are intended to delay the signal until gain correction is effected, amplifiers using both slow and fast gain reduction, those which convert the signal to a modulated and demodulated carrier so as to suppress "thump," etc.

Inherent in most of these types, as well as our older models, is the objectionable increase in background noise during quiescent portions of the program, followed by an objectionable fast-gain reduction as program is resumed, and/or other degrading and objectionable effects.

Since our recent venture into high-fidelity transmission with the new WLW Cathanode Transmitter, it has become necessary to improve all other associated equipment, particularly the AGC amplifier.

Redesign

From our own experience, and from tests conducted on commercial units, we realized that the first step in redesign was to analyze all shortcomings of presently available equipment. From this analysis, mandatory specifications and requirements were determined in order to be acceptable in the new WLW high fidelity system. These are shown in Table I.

TABLE I
DESIGN SPECIFICATIONS

Input Level
Adjustable, −20 dbm to 0 dbm

Output Level
Adjustable, −5 dbm to +15 dbm

Frequency Response
±0.5 db, 20 to 20,000 cps, for any condition of AGC action

Signal-to-Noise Ratio
Min. 70 db below normal output

Harmonic Distortion
Max. 0.5 per cent for any condition of AGC action

AGC Characteristics
Normal input signals cause no change in the slow-increase, slow-decrease, or fast-decrease circuits.
Low-level signals cause the gain to increase slowly, as needed.
High-level signals cause the gain to decrease slowly, as needed.

Slow-gain decrease range:	10 db
Slow-gain increase range:	10 db
Fast-gain decrease range:	20 db

Time Constants

Slow-gain increase attack time	To be determined by listening tests
Slow-gain increase release time	
Slow-gain decrease attack time	
Slow-gain decrease release time	
Fast-gain decrease attack time	
Fast-gain decrease release time	

Metering
Illuminated input VU meter permanently connected following input gain control.
Illuminated output VU meter permanently connected ahead of output gain control.
Illuminated matching meter. Normally zero left with center go-no-go tube current scale and increase-decrease DB scale from zero center.

Manual Controls
Input level control 20 db, in 1-db steps, unbalanced T.
Output level control 20 db, in 1-db steps, unbalanced T.
Electrical meter center (AGC expansion).
Electrical meter center (AGC compression).
AGC ON-OFF switch
Selector switch-tube check, AGC monitor.

Basic Circuit

The complete circuit diagram for the new WLW AGC amplifier is shown in Fig. 1, and photographs of the amplifier are shown in Figs. 2–4. The basic amplifier consists of two 6SN7 amplifier stages and a push-pull 6L6(5881) output stage operating Class A.

AGC Action

Between the first and second 6SN7 amplifier stages, V_1 and V_2, is an electronic bridge circuit incorporating a 6SN7 tube, V_2, whose grids are controlled by rectified signal. The variable resistance of this tube provides an automatic electronic variable attenuator. Three separately-acting 6AL5 rectifiers, V_8, V_9, and V_{10}, sense the signal level conditions and automatically control the grids of V_2.

Electronic Bridge Attenuator

The action of the electronic bridge attenuator is as follows: The output of the first amplifier stage, V_1, is fed into the top and bottom of the bridge, a voltage division occurs on the right side of the bridge through 68K and 22K ohm resistors R_1 and R_2 as one leg and 220K resistor R_3 as the other leg. Since R_1 and R_2 total lower resistance than R_{13}, the output voltage at the right side of the bridge will be fed mostly from the top input terminal. As positive dc is applied to the grid of the control tube V_2, its resistance decreases, thus reducing the voltage at the junction of R_1 and R_2 which lowers the output voltage at the right side of the bridge. Exactly the same effect occurs at the left bridge output terminal with respect to input voltage fed to the bottom of the bridge. Since, in a bridge circuit of this type, a linear increase of positive grid voltage produces an exponential reduction of audio output voltage, the control capability is greatly extended and flattened.

Slow Gain Increase-Decrease

It will be noted that two of the three signal rectifiers, V_8 and V_9, are connected in dc opposition.

Rectifier V_8 which produces negative dc control voltage is supplemented with a negative 13-volt threshold so as to bias V_2 to a quiescent static current, which results in a no-signal gain reduction of 10 db. As signal is increased from a very low value, this rectifier starts to cut off V_2, approaching complete cutoff at approximately 10 db below normal input, thus producing approximately 10 db slow gain increase. Just previous to cutoff, as signal level is further increased, the positive dc control rectifier, V_9, overcomes its threshold bias and, since this rectifier involves a series resistance of only 100K ohms, it takes precedence over the negative rectifier, V_8, whose resistance is over 1 megohm. The control tube, V_2, therefore becomes more conductive and slowly reduces gain, as shown in Fig. 5. It may be noted that further action of the negative rectifier circuit, above normal signal level, is clamped by a biased diode, so as to not interfere with the action of the slow-gain decrease circuit. The attack time of the slow-decrease circuit is approximately 20 msec, with a 30-second release time. The attack time of the slow increase circuit is 7 seconds, with a release time of 30 seconds. This characteristic was incorporated so as to simulate the action of a human operator who would cautiously increase a low-level signal yet rapidly decrease an excessive signal.

Fast Gain Decrease

In Fig. 5 it can be seen that the fast-gain decrease action parallels the slow gain increase-decrease action, but at a signal level about 4 db higher. The fast circuit utilizes the same 6SN7 control tube, V_2, as does the slow circuit, but is controlled from a short time constant rectifier (V_{10}). A high peak signal will cause the fast-

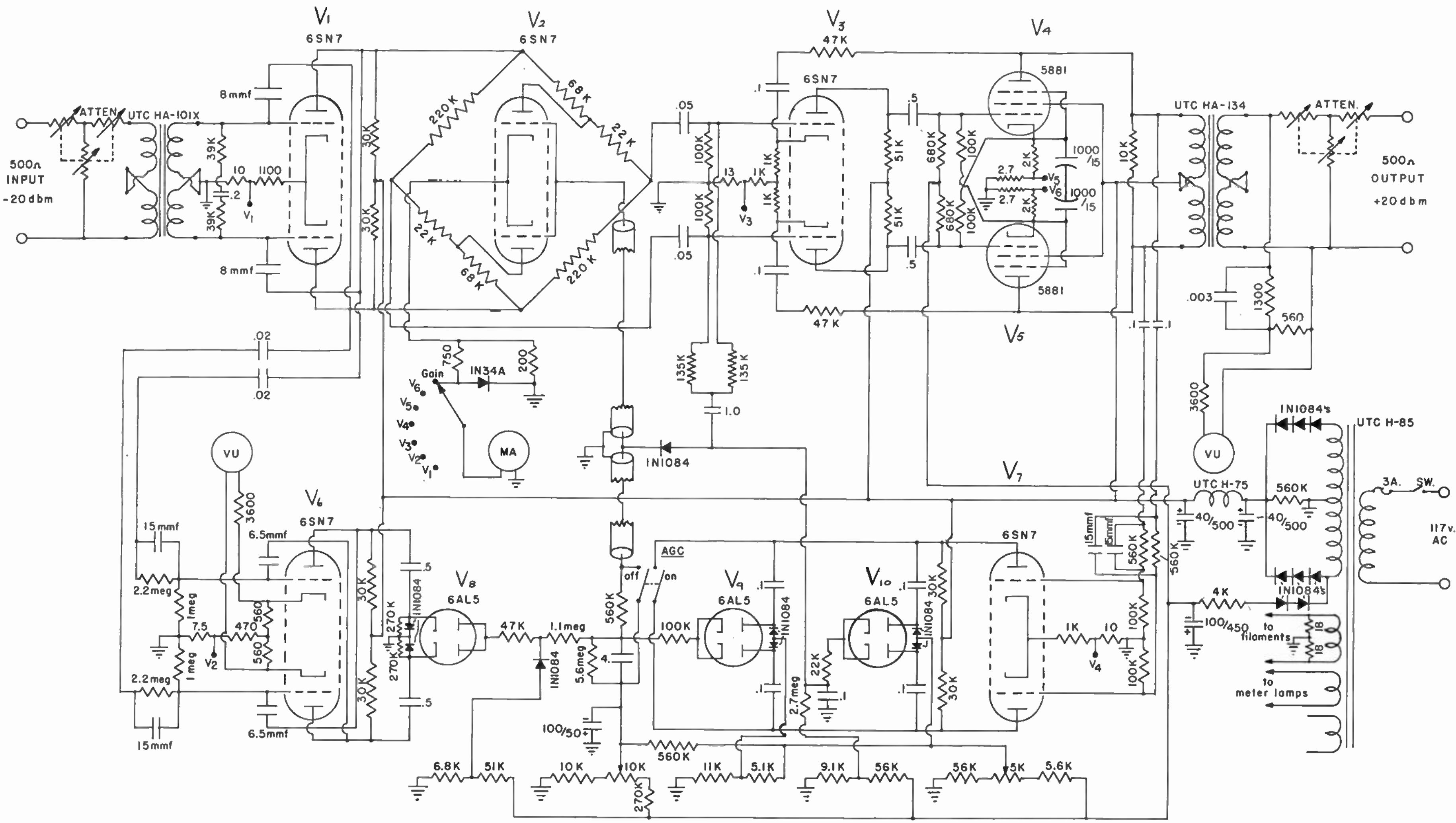


Fig. 1—Amplifier circuit diagram.

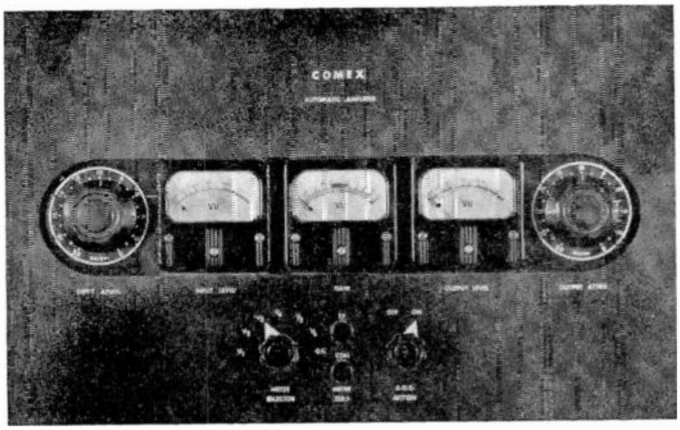


Fig. 2—Amplifier, front view.

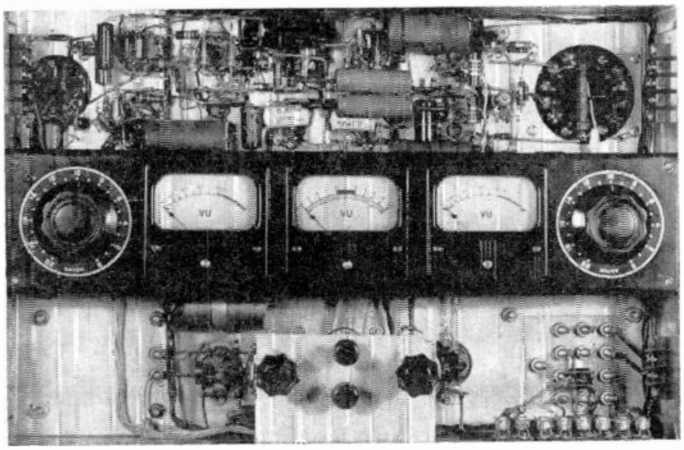


Fig. 3—Amplifier, front view with panel removed.

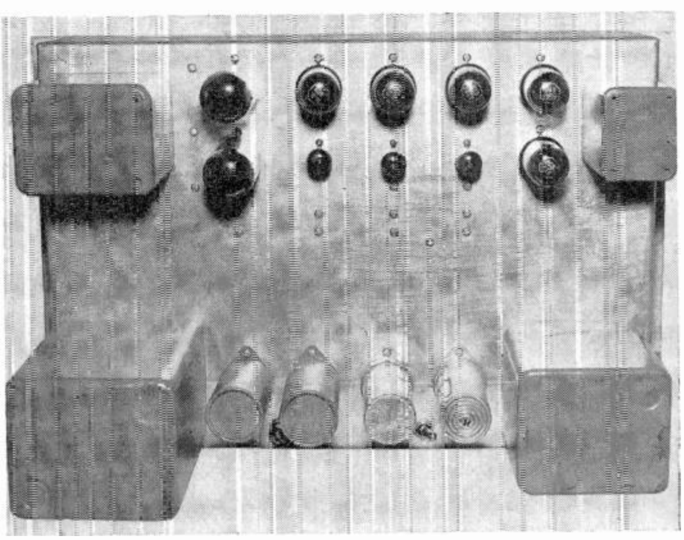


Fig. 4—Amplifier, rear view.

gain decrease circuit to take temporary control from either the slow-increase or slow-decrease circuits, but it is rapidly biased out of operation by the slow-decrease rectifier.

It will be noted that the fast dc surges from the short time constant rectifier are applied to the second ampli-

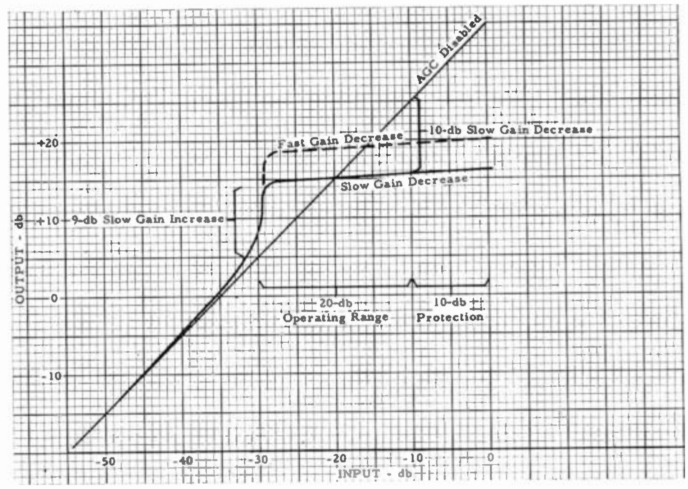


Fig. 5—Performance characteristic of WLW AGC amplifier.

fier grids through a 1-mfd condenser and 135K ohm dropping resistors; these positive pulses effectively neutralize the negative pulses resulting from fast plate-voltage reduction as the control tube increases plate current. This simple method of "thump" suppression eliminates the costly interstage transformer previously used to prevent parallel negative grid surges in conventional circuits; this condenser also provides the slow recovery time for the fast-decrease circuit. The attack time of the fast-decrease circuit is approximately 3 msec, and the release time if 1.5 seconds.

Frequency Response

Frequency response of this unit, under any condition of gain increase or decrease, is essentially flat from 10 cps to 30,000 cps, as shown in the curves of Fig. 6. These extreme specifications were met by careful neutralization of all high-resistance grid circuits and correction for input and output transformer characteristics.

Distortion

Several factors contribute to the very low distortion. Most conventional circuits use the rectified signal to control bias on a low-level stage. This is obviously a departure from proper bias, and reasonable distortion is only possible at very low signal levels. The gain control circuit, herein described, easily maintains distortion less than 0.5 per cent under all conditions of gain control up to 10 db above normal input, as shown in Fig. 7. All signal rectifiers are isolated from amplifier stages by isolation stages V_6 and V_7, so as to isolate rectifier distortion. It will be noted from the circuit diagram that the output amplifier's grid bias resistors return to opposite cathodes and that negative bias is fed directly to the grids through very high resistance (680K ohms), at high negative voltage. This novel arrangement provides nearly perfect balancing of anode current of unmatched tubes, thus reducing dc saturation of the output transformer, resulting in approximately 10 to 1 distortion re-

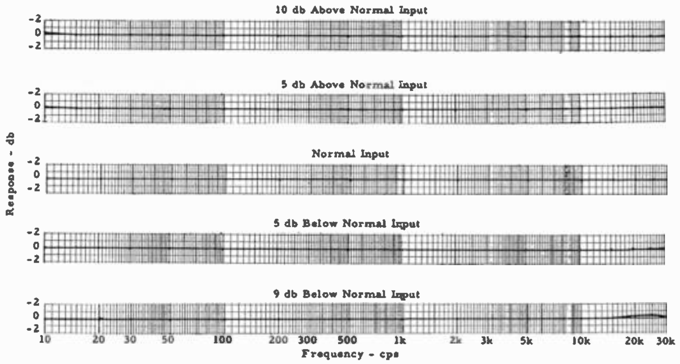


Fig. 6—Frequency response at various input levels.

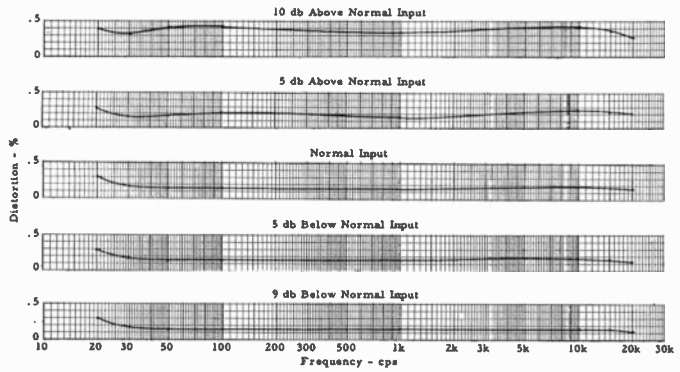


Fig. 7—Distortion at various input levels.

TABLE II

MEASURED PERFORMANCE CHARACTERISTICS

Input Level
Adjustable, −20 dbm to 0 dbm

Output Level
Adjustable, −5 dbm to +15 dbm.

Frequency Response
±0.5 db, 10 to 30,000 cps, for any condition of AGC action

Signal-to-Noise Ratio
80 db below normal output (AGC Off)
78 db below normal output (AGC On)

Harmonic Distortion
Less than 0.5 per cent for any condition of AGC action

AGC Action
Normal input signals cause no change in the slow-increase, slow-decrease, or fast-decrease circuits.
Low-level signals cause the gain to increase slowly, as needed.
High-level signals cause the gain to decrease slowly, as needed.

Slow-gain decrease range:	15 db
Slow-gain increase range:	9 db
Fast-gain decrease range:	24 db

Time Constants

Slow-gain increase attack time:	approximately	7	seconds
Slow-gain increase release time:	approximately	30	seconds
Slow-gain decrease attack time:	approximately	20	msec
Slow-gain decrease release time:	approximately	30	seconds
Fast-gain decrease attack time:	approximately	3	msec
Fast-gain decrease release time:	approximately	1.5	seconds

duction at 20 cps. Distortion values in the order of 0.25 per cent are obtained. Functionally, the cathode of a high-current tube will be more positive than the cathode of a low-current tube. With this circuit, the higher positive cathode voltage feeds through the grid resistor to the opposite low-current tube grid, thus increasing its current. Conversely, the cathode of the low-current tube biases the grid of the high-current tube to lower current. This compensating action is improved by using higher than normal cathode resistors in order to effect more dc correction.

Frequencies as low as 20 cycles per second were held within this very low distortion at plus 20 dbm output by using very high (2000 ohms) cathode resistors in the output stage and compensating with positive grid bias. This permit tube currents to balance closely, thus greatly reducing transformer dc saturation. Conventional 330-ohm cathode resistors resulted in 20-cycle distortion near 2 per cent with carefully selected output tubes and random tubes produced much higher distortion.

Operation

The equipment utilizing the circuits described will automatically perform the following functions:

1) Normal input signals cause no change in the slow-increase, slow-decrease, or fast-decrease circuits.
2) Low-level signals cause the gain to increase slowly, as needed.
3) High-level signals cause the gain to decrease slowly, as needed.
4) During conditions 1 through 3, occasional peaks result in fast-gain decrease with slow recovery until the slow circuits take precedence.

Setup

The setup for the amplifier is very simple and can be accomplished with a mid-frequency tone source, as follows:

1) Apply mid-range tone and set input meter to 100 per cent with input attenuator.
2) Set selector switch to position GN and AGC switch to "OFF" position and set center meter to zero with control marked "EX."
3) With AGC switch in "ON" position, again set center meter to zero with control marked "COM."

The amplifier is then ready for program use.

The presence of this amplifier in the WLW high-fidelity system is completely undetectable to the ear, yet chart recordings of modulation percentage show a definite higher modulation level as a result of its use, as shown in Fig. 8.

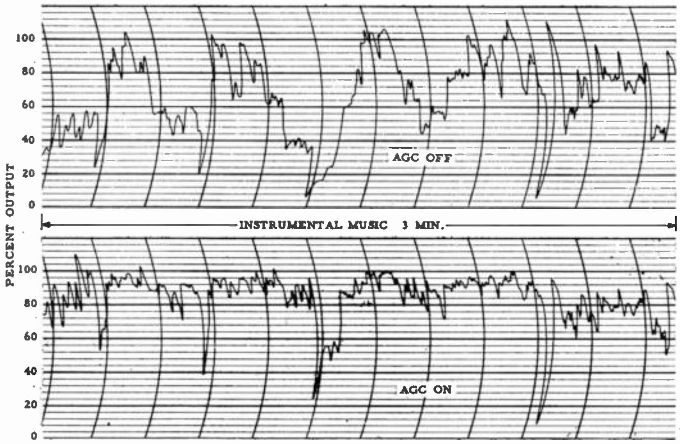


Fig. 8—Chart recording of rectified output.

ACKNOWLEDGMENT

The success of this project was, to a large degree, due to the contribution of Mr. E. B. Dooley, Staff Engineer, Crosley Broadcasting Corp. The first two paragraphs contain portions of memorandum from K. A. Norton to A. D. Ring in May, 1939, incorporated with Mr. Ring's permission.

The Distortion Resulting from the Use of Center-Tapped Transformers in a Class B Power Amplifier*

R. G. DE BUDA†

Summary—**The transformer-coupled Class B amplifier is investigated with regard to harmonic distortion caused indirectly by the self-resonance of the transformer primary. For undistorted Class B operation, the short-circuited flow of even harmonic unbalanced mode currents is essential. When these harmonics are at a frequency at which the transformer primary is close to self-resonance, the flow is impeded and harmonic plate voltages will appear. A slight circuit asymmetry or nonlinearity will permit the harmonic voltages to couple into the load, as harmonic distortion. By a similar effect negative feedback circuits can increase distortion by an amount for which formulas are given.**

INTRODUCTION

In this paper, the push-pull transformer-coupled Class B amplifier of Fig. 1 is discussed, which may, for instance, be used in the final audio stage of a high-level modulated RF transmitter; and a mechanism is investigated by which the transformer causes nonlinear distortion of the output signal.

These nonlinear distortions occur at certain narrow frequency bands at or near the upper end of the audio band, so they are certainly not related to low-frequency saturation effects of the transformer core. The Class B amplifier is assumed to have an ideally linear characteristic. On the surface it would appear that assuming ideally linear elements only, there can be no distortion. But this overlooks the fact that a Class B tube, regardless how "linear," has a sharp nonlinearity at the commutation from cutoff to the conducting state.

If the three-winding transformer in Fig. 1 has a finite leakage inductance between the halves of the primary, then the voltage wave induced across tube 2 by the current flowing through tube 1 will not exactly fit the voltage wave across tube 2 while tube 2 conducts. Thus, the commutation of Class B operation gives rise to quasi-transients which have been reported by Sah in 1936.[1]

* Received by the PGA, February 15, 1960. Presented at the IRE Canadian Convention, Toronto, Can., October 8, 1959.
† Canadian General Electric Co., Ltd., Toronto, Can.

[1] A. P.-T. Sah, "Quasitransients in class B audio frequency push-pull amplifiers," PROC. IRE, vol. 24, pp. 1522–1541; November, 1936.

Sah also gives expressions (rather complicated ones) for the wave shape of the quasitransients.

For our applications in broadcast work, we also want to know the frequency composition of these transients, and which part of their energy couples into the load, particularly in the presence of feedback.

To obtain usefully simple equations, and a better understanding of the effects involved, we tried different approaches, and found that it is possible to give a steady state theory of this distortion which requires considerably less algebra than the transient analysis. The method of approach must be carefully and suitably selected to achieve this, and some notes recording our approach are therefore in order.

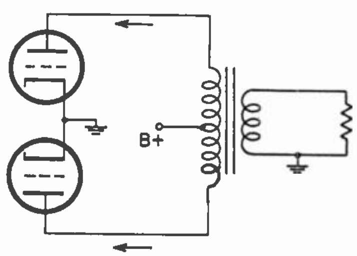


Fig. 1—Transformer-coupled Class B amplifier, where the transformer turn ratio is 1:1:n.

The Push-Pull Mode and the Unbalanced Mode

Because of the nonlinearity of the commutation effect, we are forced to consider our signals in time, and not in the frequency domain. All voltages and currents, unless otherwise stated, are therefore instantaneous values and denoted in the formulas by lower case e's and i's. The impedance Z of the transformer in the connection of Fig. 3 has a significant rise at frequencies near self-resonance of the primary; but since we work in time, not in frequency, we must consider Z as an operator, transforming one wave shape into another, and not just as a complex multiplicator.

The instantaneous currents and voltages which we shall use are, however, not the grid voltages, plate voltages and currents of the two tubes. We found that the theory of the distortion becomes much more precise and clear if we introduce as new parameters the symmetrical modes for which we shall use in this paper the names: "push-pull mode," indicated by the subscript p; and "unbalanced mode," indicated by the subscript u.

A signal of the push-pull mode is defined as half the difference between the corresponding signals in tubes 1 and 2; a signal of the unbalanced mode is half the sum. For instance: the push-pull plate current i_p and the unbalanced grid voltage e_u are defined respectively by

$$i_p = \tfrac{1}{2}(i_1 - i_2)$$

$$e_u = \tfrac{1}{2}(e_1 + e_2).$$

Similar defining equations may be written for all other quantities discussed. The signals of the tubes are then obtained by linear superimposition of the 2 modes. The justification for working with these modes, rather than with the actual signals, lies in their lack of coupling, which permits easy superposition. A current in tube 1 induces a voltage across tube 2, but a push-pull current causes no unbalanced voltage and vice versa; this follows from the symmetry of the circuit (however, there is a relation between push-pull and unbalanced current which we shall discuss later). It is possible to draw equivalent circuits, as shown in Figs. 2 and 3, which permit the study of one mode irrespective of the other. This is quite convenient; for instance, the circuit in Fig. 3, which is the equivalent circuit of the unbalanced mode, doubles as test circuit to measure directly the theoretically important impedance Z (of the short-circuited transformer) without the necessity of supplying a push-pull signal as well.

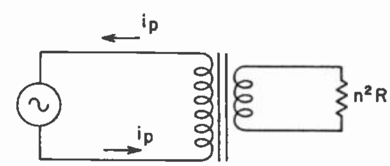


Fig. 2—Equivalent circuit of the push-pull mode, where the reflected load is $4R$.

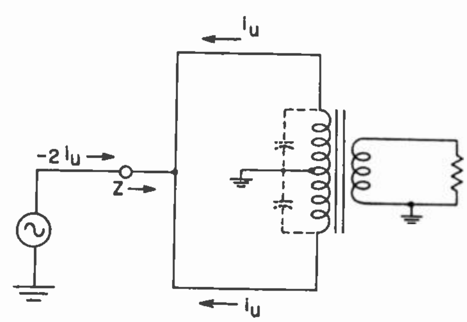


Fig. 3—Equivalent circuit of the unbalanced mode. The theoretically important impedance Z is the impedance of the transformer when connected as shown; it can be directly measured by bridge.

From inspection of Figs. 2 and 3 it is seen that the push-pull mode is the desired mode of operation. In this mode, the currents of the two generators are in series, and no current flows into the $B+$ supply, so that we may omit the center tap from the equivalent circuit (Figs. 1 and 2).[2] The transformer frequency response of this circuit is the main limitation, since it limits the frequency range at which power is delivered to the load,

[2] *Ibid.*, p. 1530, Fig. 4.

but this is a well understood process and we shall not deal with it here.[3,4]

The unbalanced mode of operation is something we would like to do without, but the presence of unbalanced currents is absolutely necessary for proper Class B operation. Since at any time current flows only in one tube, the unbalanced mode current certainly cannot disappear and must be positive. Further, in the tube which does not conduct, we have alternatively

$$i_u - i_p = 0 \qquad \text{while tube 1 conducts}$$

or

$$i_u + i_p = 0 \qquad \text{while tube 2 conducts.}$$

Since i_u is always positive, we can combine the two expressions into one equation,

$$i_u = |i_p|, \tag{1}$$

which is valid during the whole period. This shows that a certain amount of interaction between the modes may take place in order to ensure that (1) is fulfilled regardless of load. But if the tubes act as current generators (which is usually the case), the current is not affected by the load impedance and no interaction occurs.

We may specialize (1) for a sinusoidal signal. Then the unbalanced signal will not contain the fundamental, but will consist only of dc and all the even harmonics characteristic for a rectified sinewave $|\cos \omega t|$, for which expressions are available from the literature,[5] namely

$$|\cos \omega t| = \sum_{n=0}^{\infty} k_{2n} \cos 2n\omega t \tag{2}$$

$$\begin{cases} k_0 = \dfrac{2}{\pi} \\ k_{2n} = (-1)^{n+1} \dfrac{4}{\pi(4n^2 - 1)} \end{cases} \tag{3}$$

$2n$	0	2	4	6	8	$\cdots$
$k_{2n}[\%]$	$+63.6$	$+42.5$	-8.5	$+3.6$	-2.0	$+\cdots$

(4)

The Unbalanced Plate Voltage and the Impedance Z

The harmonic content of $|\cos \omega t|$ is quite undesirable, unless we can short-circuit these harmonic currents so that no unbalanced mode voltage appears on the tubes.

[3] F. E. Terman, "Radio Engineers Handbook," McGraw-Hill Book Co., Inc., New York, N. Y., sec. 5.7–5.10; 1943.

[4] P. A. Fessler, H. S. Lasher, H. W. Lord, and G. Walters, "Transformer Design Criteria and Data," General Electric Res. Lab., Schenectady, N. Y., Rept. No. RL-417; June 10, 1950.

[5] "Reference Data for Radio Engineers," ITT, New York, N. Y., 4th ed., p. 1022; 1956.

This is done in effect by the short-circuited primary of the transformer, if it is connected as shown in Fig. 3. We note that since, by definition, the unbalanced mode is in phase at both tubes, we can add in effect an ohmic connection between the two plates without affecting the operation of this mode. The push-pull mode can, of course, not exist in this connection; thus, we analyze the circuit here in terms of the unbalanced mode alone.

If the two halves of the primary of the transformer were ideally coupled, the unbalanced mode current would be completely shorted out and no unbalanced mode voltages would appear anywhere in the circuit. The same result can be obtained, of course, from the more popular concept of fitting wave shapes together. However, transformers are not ideal, and the transformer as connected in Fig. 3 will introduce a finite frequency-dependent impedance Z into the circuit, which acts only upon the unbalanced mode. (There will also be a finite impedance in series with the $B+$ supply, particularly at very low frequencies.) Thus, we see now that the seemingly more complicated method of describing the signals via the two modes gives us a powerful tool for quantitative analysis of the even harmonic distortion, which we state in the following theorem:

> If the tubes can be considered as current generators, the even harmonic voltage occurring at their plates can be obtained by feeding the (full wave) rectified signal current $-2i_u$ into the output transformer with both plate leads short-circuited, as shown in Fig. 3.

Coupling Between Modes

So far we have established the presence of even harmonics at the plates. However, these harmonics belong to the unbalanced mode, and they will not necessarily couple out into the load since the load is supplied only from the push-pull mode.

Normally there should be no interaction between modes, consequently no distortion at the load; but, as mentioned before, there are exceptions, and we must investigate them carefully. There are at least four different ways in which the unbalanced harmonics do reach the load. We shall discuss them in the sequence in which they are listed in Table I.

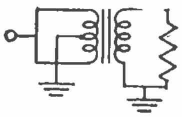
Circuit Unsymmetry

First, there may be insufficient unbalanced mode rejection by the transformer. For instance, the transformer may not be wound exactly symmetrically, or its stray capacitance or external strays may be unsymmetrical so that a small part of the unbalanced signal "leaks" into the load. In a conventional transformer, this effect will not be very pronounced, but if a center-tapped choke or an auto-transformer is used which permits a direct ohmic connection from one of the modulator tubes to the load, then coupling of the harmonics into the load will occur whenever $Z \neq 0$. Although the center-tapped reactor has a much better frequency re-

TABLE I

Distortion at the output of a transformer-coupled stage will occur if the two following conditions coincide:

1) The impedance Z of the shortcircuited transformer

$Z =$ 

becomes large;

2) The resulting unbalanced signal couples into the push-pull mode.

1. The impedance Z will be unavoidably large at certain self resonance frequencies of the transformer primary winding.
2. Coupling between modes can occur in any one of the following cases:
   a) If the transformer has not enough common mode rejection. This will not usually be serious, except where the modulation transformer is replaced by an auto transformer.
   b) If negative feedback is used to suppress the unbalanced voltage.
   c) If the plate resistance of the output tubes is low compared with Z.
   d) If the unbalanced voltage when added to the push-pull voltage exceeds the voltage range in which the tube is linear.

sponse than a three-winding transformer, the direct connection of the unbalanced mode to the load may more than offset this advantage with respect to harmonic distortion. The same reservations may be made with respect to an interesting transformer connection recently patented by Wahlgren.[6]

For similar reasons, unsymmetry of other circuit components or of the drive should be avoided.

Distortion Caused by Negative Feedback Circuits

A second type of interaction between the modes can occur as a result of the presence of feedback circuits.

To show this, we consider a push-pull amplifier similar to the one illustrated in Fig. 4.

The plate currents are shown as sum and difference, respectively, of their push-pull component i_p and their unbalanced (in-phase) component i_u. Similarly, the grid voltages are the sum of the (balanced) input signal $\pm e_i$ and two feedback voltages e_u and $\pm e_p$ which again represent the unbalanced and the push-pull components of the feedback signal.

When we are calculating the plate voltage, it too will naturally be split into its symmetrical components. Since the reflected load is $4R$, the push-pull component of the plate voltage will be $-2Ri_p$. The unbalanced current is short-circuited by the transformer except at those frequencies at which $Z \neq 0$. There, an unbalanced voltage $-2Zi_u$ will exist. The total plate voltage to ground is of course, the sum (or difference) of the above two terms.

The feedback voltages are developed from the plate voltages by a passive network which may be assumed

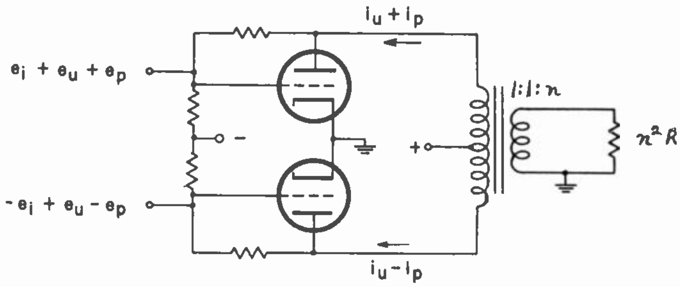


Fig. 4—Transformer-coupled class B amplifier with feedback, where

$\pm e_i$ = input signal

$e_u e_p$ = feedback voltages, defined by

$$e_p = \beta \cdot i_p \cdot 2R,$$
$$e_u = \gamma \cdot 2i_u \cdot Z.$$

to have circuit symmetry. Then the push-pull component of the feedback voltage can be derived only from the push-pull component of the plate voltage, and similarly for the unbalanced components. Introducing the (usually negative) feedback factors β and γ for the respective modes, we can write the following:

$$e_p = \beta \cdot 2i_pR \tag{5}$$

$$e_u = \gamma \cdot 2i_uZ. \tag{6}$$

In the circuit of Fig. 4, $\beta = \gamma =$ constant, but it is obviously possible to design more sophisticated networks for which the two feedback factors are no longer equal. For our analysis, we must treat them therefore as different parameters.

The feedback is used around a Class B amplifier which has a gain A during its 180° conduction angle. This amplifier may be a single-tube stage, as shown in Fig. 1 (for which A is, of course, the transconductance of the tube), or it may be a more complex Class B circuit.

In the conducting tube (or part of the amplifier), we have

$$i_p + i_u = + A(e_i + e_u + e_p). \tag{7}$$

In the other tube we have no current at that time. (It may be well to repeat here that we conduct our analysis in the time domain, because of the presence of nonlinearities, and that the e's and i's represent instantaneous values). 180° later, we have no current in the first tube, but in the second tube

$$- i_p + i_u = + A(-e_i + e_u - e_p). \tag{8}$$

In the second half period, we have $i_p < 0$, $e_i < 0$ and $e_p < 0$, and (1), (7) and (8) can be combined into (9).

$$\begin{aligned} i_u &= \left| i_p \right| \\ &= \tfrac{1}{2}A\left\{ \left| e_i + e_p \right| + e_u \right\}. \end{aligned} \tag{9}$$

The following derivation follows the development given in 1939 by Mayer[7] for the feedback around a single tube,

[6] W. W. Wahlgren, "Push-Pull Transformer," U. S. Patent No. 2-798-202; July 2, 1957.

[7] H. F. Mayer, "Control of the effective internal impedance of amplifiers by means of feedback," Proc. IRE, vol. 27, pp. 213–217; March, 1939.

except that special consideration is given to the nonlinearity in (9) in relation to the frequency response of Z.

When we close the feedback loop, (1) and (9) must be fulfilled together with (5) and (6). If we compare (6) with (9), we obtain

$$e_u = 2\gamma Z i_u = \frac{A\gamma Z}{1 - A\gamma Z} \left| e_i + e_p \right| . \tag{10}$$

Eq. (10) must be interpreted that e_u consists of those selected harmonics of $|e_i+e_p|$ which fall into the passband of a filter defined by $|A\gamma Z| \gg 1$.

Since the distortion term e_u is supposed to be small, we may conclude that $+i_p$, e_i and e_i+e_p change signs at about the same instant. Consequently:

$$\text{sign } i_p = \text{sign } e_i = \text{sign } (e_i + e_p)$$

$$= \begin{cases} +1 \\ -1 \end{cases} \quad \text{(switching function)} \tag{11}$$

$$i_p = i_u \text{ sign } (i_p) \tag{1'}$$

and (9) becomes

$$i_p = \tfrac{1}{2} A \{ \left| e_i + e_p \right| + e_u \} \text{ sign } (e_i + e_p)$$

$$= \tfrac{1}{2} A \{ e_i + e_p + e_u \text{ sign } (e_i + e_p) \} . \tag{12}$$

The last term in the bracket in (12) permits this interpretation: the unbalanced feedback voltage (6), $e_u = 2\gamma Z i_u$, modulated by sign i_p, is added to the grid signal as distortion. In the Appendix, the derivation of (13) and (14) for e_p and e_u is given.

$$e_p = 2\beta R i_p = \frac{A\beta R}{1 - A\beta R} \{ e_i + e_u \text{ sign } (e_i + e_p) \} \tag{13}$$

$$e_u = \frac{A\gamma Z}{1 - A\gamma Z - A\beta R} \left| e_i \right| . \tag{14}$$

If sufficient feedback is used, (14) simplifies to

$$e_u = - \frac{\gamma Z}{\gamma Z + \beta R} \left| e_i \right| . \tag{15}$$

From (13) we see that the distortion is given by the ratio of (e_u sign e_i) to e_i. If we would multiply (15) by sign e_i, to remove $|e_i|$ we would get a wrong result, because the operator Z does not act upon sign e_i. The proper procedure is to express the energy relation without knowing the actual waveshape, and obtain distortion percentage as follows: The energy of the distortion is not changed if the wave is multiplied by a switching function sign i_p. If e_i is a sinewave, the $2n$th harmonic of $|e_i|$ has the amplitude k_{2n} which is given by (3) and (4). If this is the only frequency at which $Z \neq 0$, the total distortion is given by

$$d = \frac{\gamma Z}{\gamma Z + \beta R} k_{2n} \tag{16}$$

and the distortion waveshape will be the $2n$th harmonic of $|e_i|$ modulated by sign e_i.

Ketchledge discussed a similar effect,[8] but he assumes implicitly an ideal transformer and restricts his derivations to almost linear tube characteristics, which seems to exclude Class B operation.

We may interpret our results as follows: The negative feedback γ tends to suppress any unbalanced voltage $-2Zi_u$. If $Z \neq 0$, γ will remove the corresponding harmonic from i_u. Since

$$i_p = i_u \text{ sign } i_p, \tag{1'}$$

the missing harmonic, after modulation by sign i_p shows up as distortion of i_p; consequently, the push-pull negative feedback β will try to remove this distortion, so that the unbalanced voltage at the plates reappears. In the resulting tug of war between β and γ, a compromise is reached, which represents (16).

Distortion Caused by Low Plate Resistance

A third case of coupling between modes is based on similar principles, except that here the amplification factor μ of the tube takes the place of the feedback.

Suppose that we use a single stage without feedback as amplifier, and that its tubes have a plate resistance which is too low to consider the tube as a pure current generator. Then the plate current will depend not only on the grid voltage but also on the plate voltage, and we may consider the terms e_p and e_u as the symmetrical components of the plate voltage divided by the amplification factor μ. In this case, we must substitute μ into (5) and (6) by

$$\beta = \gamma = - \frac{1}{\mu} . \tag{17}$$

The rest of the preceding analysis can be taken over, with A being the transconductance g, until (14) is reached. By use of Barkhausen's equation between g, μ and the plate resistance R_p,

$$g \cdot \frac{1}{\mu} \cdot R_p = 1 \qquad \text{(Barkhausen's equation)}, \tag{18}$$

(14) becomes

$$e_u = \frac{Z}{R_p + R + Z} \left| e_i \right| , \tag{19}$$

and the distortion

$$d = \frac{Z}{R_p + R + Z} \cdot k_{2n} . \tag{20}$$

Distortion Caused by Clipped Over-Voltages

A fourth undesirable effect of the modulation transformer resonance is an increase of the voltage swing. The unbalanced voltage adds to the push-pull voltage; this is not too serious, as long as the peak voltage is still

[8] R. W. Ketchledge, "Distortion in feedback amplifiers," *Bell Sys. Tech. J.*, vol. 34, pp. 1265–1285; November, 1955.

within the linear region of the tube. If it is not, clipping of the peaks will occur, resulting in additional harmonics, and these harmonics will couple out into the load. Also, the clipping will tend to produce intermodulation, which to the human ear is more obnoxious than the mere presence of harmonics.

To avoid this type of clipping, we must therefore extend the linear region of the tube beyond the voltage of the useful load. This, however, reduces the efficiency of the Class B stage, and it becomes apparent that it is not satisfactory merely to add a safety margin to the linear region. We must therefore try to express the peak voltage to be expected when unbalanced voltages are present, and design the amplifier accordingly.

The push-pull voltage and the unbalanced voltage at the plate are given by $-2Ri_p$ and $-2Zi_u$, respectively, in which $2R$ represents half the load reflected into the primary and Z is the impedance of the short-circuited transformer (Fig. 3).

If we consider a sinewave $i_p = I_0 \cos \omega t$ and assume that Z disappears for all frequencies except for the $2n$th harmonic, then we can express the total plate voltage by

$$V = 2RI_0 \cos \omega t + 2|Z| I_0 k_{2n} \cos (2n\omega t + \theta)$$

$$= 2RI_0 \left[\cos \omega t + \frac{|Z|}{R} k_{2n} \cos (2n\omega t + \theta) \right]. \tag{21}$$

In this equation, θ represents the phase shift caused by Z, and the factor k_{2n} is the coefficient of the rectified sinewave of (2) to (4).

If Z is real, the peak voltage will be exactly

$$V_{\text{peak}} = 2RI_0 \left(1 + k_{2n} \frac{|Z|}{R} \right) \tag{22}$$

If Z is complex, it is not possible to give a simple and exact expression for the peak voltage. Since we cannot tolerate a large increase in the required linear region, the term $k_{2n}(Z/R)$ must be kept small, and close bounds for the peak voltage may be derived. For instance, if Z is purely imaginary, we have $\theta = 90°$ and

$$V = 2RI_0 \left[\cos \omega t + k_{2n} \frac{|Z|}{R} \sin 2n\omega t \right]$$

$$V < 2RI_0 \left[\cos \omega t + 2nk_{2n} \frac{|Z|}{R} |\sin \omega t| \right] \tag{23}$$

$$V_{\text{peak}} < 2RI_0 \sqrt{1 + 4n^2 k_{2n}^2 \frac{|Z|^2}{R^2}}$$

$$< 2RI_0 \left[1 + 2n^2 k_{2n}^2 \frac{|Z|^2}{R^2} \right]. \tag{24}$$

Also, we have for all phase angles

$$V_{\text{peak}} \leq 2RI_0 \left(1 + k_{2n} \frac{|Z|}{R} \right). \tag{25}$$

The last equation is probably best suited to specify the frequency response of the coupling transformer in the test connection, Fig. 3: The linear region must extend by a factor $1 + k_{2n}(|Z|/R)$ beyond the loadline. It should be noted that k_{2n} decreases rapidly with increasing n. In our experience in high-power audio stages, it is usually the fourth harmonic which is more difficult to handle than the rest, since k_4 is still appreciably large (approximately 8 per cent), while the frequency response at four times the upper audio is already difficult to control in the transformer.

The expressions (22) to (25) assume that no unbalanced feedback is used. If there is, the peak voltage would become smaller, but as outlined before, unbalanced feedback is quite undesirable, so that it is unnecessary to give design equations. We note however that a modification which reduces γ as required by (15) may increase the voltage swing to a point where clipping in the tube may be worse than the distortion due to γ.

Conclusion

The flow of even harmonic unbalanced mode currents is essential for Class B operation. When these harmonics are at a frequency at which the transformer primary is close to self resonance, unbalanced harmonic plate voltages will appear. A slight circuit unsymmetry or nonlinearity will permit the harmonics to couple into the load, and indiscriminate use of negative feedback may increase distortion rather than reduce it. A proper understanding of these factors is essential for a good, distortion free, Class B amplifier design.

Appendix

Combining (12) with (5) gives

$$e_p = 2\beta R i_p = \frac{A\beta R}{1 - A\beta R} \{ e_i + e_u \operatorname{sign} (e_i + e_p) \} \tag{13}$$

$$e_i + e_p = \frac{1}{1 - A\beta R} \{ e_i + A\beta R e_u \operatorname{sign} (e_i + e_p) \}$$

multiply by sign $(e_i + e_p)$:

$$|e_i + e_p| = \frac{1}{1 - A\beta R} \{ e_i \operatorname{sign} (e_i + e_p) + A\beta R e_u \}$$

eliminate $|e_i + e_p|$ by (10) and replace sign $(e_i + e_p)$ by sign (e_i):

$$e_u = \frac{A\gamma Z}{1 - A\gamma Z} \cdot \frac{1}{1 - A\beta R} \{ e_i \operatorname{sign} (e_i) + A\beta R e_u \} \cdot$$

The last equation gives, after the necessary algebraic transformations,

$$e_u = \frac{A\gamma Z}{1 - A\gamma Z - A\beta R} |e_i|. \tag{14}$$

A Speaker System with Bass Back-Loading of Unusual Parameter Values*

PAUL W. KLIPSCH†

Summary—Work with corner horn speakers has led to a philosophy that there is a narrow range of optimum size limited by performance on the one hand and cohesion of bass and treble events on the other. In the design of a new speaker system of size smaller than these limits, the horn principle was abandoned for the bass section but continued to be used for the midrange-tweeter functions in the interest of lowest distortion. The bass section is of the enclosure type with a tubular back exhaust, corresponding in principle to the Bass Reflex but differing in the high inertance of the port. Adjusted empirically, the response of the new speaker resembles that of a three-way corner-horn reference system. The new speaker may be used against a wall, but as in all speakers affords best performance in a corner. It may be used for any of the two or three positions of a stereo array, including the bridged or derived center output.

Introduction

THE design of a new speaker system employing a direct-radiator type enclosure was undertaken. Midrange and tweeter units were of the time-proven horn-loaded compression type. The crossover-balancing network provides for individual adjustment of input to the midrange and tweeter, with the idea that a permanent setting for proper balance be made at the factory.

The bass unit was to be a noncorner design, but with the realization that any speaker is better if operated in a corner. Therefore, optimum performance in the corner was a design criterion. Since the size was to be limited to about 7 cubic feet and the unit was to sacrifice as little as possible in noncorner operation, the bass unit was to be a direct radiator. The problem was to determine optimum back-air-chamber and optimum port, slot or other opening, if any. The closed box resulted in deficient bass, even with a heavy-magnet 15-inch drive unit. A back port of much higher inertance than customary was evolved as offering nearly ideal response and extremely low distortion. After extensive tests for balance adjustments, comparison with the KLIPSCHORN[1] system as a reference indicated a closely similar response. With oscillator and program material producing 100-db intensity levels, distortion was extremely low. For example, an oscillator tone of approximately 100 cycles per second was adjusted to produce a 100-db intensity level at 6 feet; the same input in the range below 30 cycles produces a total output, fundamental and harmonic, at least 40 db lower.

In an "average" room, 100-db intensity at 50 cycles is produced by diaphragm excursion of the order of less than $\frac{1}{16}$ inch, indicating negligible Doppler distortion.

* Received by the PGA, January 21, 1960.

† Klipsch and Associates, Hope, Ark.

[1] Registered Trade Mark.

The Problem

While we acknowledge that the corner placement of a speaker offers many advantages,[2] there remains the need for a noncorner speaker system. Even in combination with corner speakers, the center channel unit in a three-channel stereo array has a strong claim to the right to be a noncorner unit. There is the apartment lessee who cannot remodel his parlor to eliminate a door in each of four corners.

The decision was to duplicate to the maximum degree possible the response of a standard corner horn speaker system in a structure suitable for use in or out of a corner. This is recognized as requiring a sacrifice in efficiency. The aim is to hold this sacrifice to a minimum. To this end, the normal environment was considered to be the corner.

It cannot be overemphasized that any speaker offers its best performance in a corner and the sacrifice of noncorner operation is costly in performance, floor space and price. Hence, the new design was considered primarily as a corner type, and its response was optimized for corner operation. The fact that it can be operated against a wall should not be a temptation to use it thus except in those cases where provision of a corner entails heroic measures. An exception is the use of the speaker as a center unit for three-channel stereo.

The unit has tentatively been named the CORNWALL and designated MODEL CW, connoting its applicability to either corner or wall use. But let it be reiterated that it, like all speakers, is primarily a corner unit, and performance criteria are based on a corner environment.

Experience with a corner-horn back-loaded bass system with horn midrange and tweeter elements and a balancing-crossover network, led to the choice of the same upper two units, with crossover frequencies of 1000 and 5000 cycles per second. The problem was confined to the design of a bass unit.

Experience with bass horns shows that a certain optimum size is required for a given maximum wavelength, regardless of efficiency. The direct radiator, on the other hand, may be reduced in size, within limits, by a sacrifice in efficiency while retaining a given rela-

[2] Superiority of response by corner speaker placement has been recognized since 1925 at least. Quantitative examination of the benefits for both monophonic and stereophonic applications are offered in other papers by P. W. Klipsch: "Corner speaker placement," *J. Audio Engrg. Soc.*, vol. 7, pp. 106–109, 114, July, 1959; "Wide-stage stereo," IRE Trans. on Audio, vol. AU-7, pp. 93–96; July-August, 1959.

tive bass response. A wall type horn would be about double the bulk of a corner horn, or say 30 cubic feet. The only approach is therefore that of a direct radiator, but it was decided to derive any possible benefits from reversed-phase port loading or other back radiation.[3]

The simple open port of calculated size for "optimum" bass response had been for several years the subject of repeated experiments by the author, always with the result that the bass response remained less than calculated and the distortion too high. An elongated or tubular port seemed attractive. Increased inertance is available.

It should be emphasized that the simple direct radiator, even with the area of a 15-inch diaphragm, requires large excursions and velocities to produce adequate power in the lowest octave, and these large low-frequency velocities produce considerable modulation distortion. Subjective tests show that as little as 0.35 per cent RMS frequency-modulation distortion is very irritating.[4] This amount of distortion would be produced by a 12-inch loudspeaker (10-inch cone), performing an amplitude of 0.21″ at 50 cps.

By utilizing the back wave through an additional coupled radiation area, the diaphragm velocity may be reduced by a factor of ½, and distortion (including both frequency and amplitude modulation) likewise reduced.

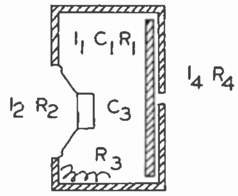
Empirical Approach

Experiments with a tunnel port[5] produced an excess or boomy bass when first tried, whereas the closed-back box offered deficient bass. The reduction of tunnel area in progressive amounts resulted in the correct response in the audible range. The tunnel length was held constant. The final adjustment of low bass range was made by listening comparison with a reference corner unit. The resultant response curve is not flat, but is allowed to droop slightly in the bass range below 80 cps. It is possible with the experimental structure to obtain a substantially flat indoor response curve, but the boomy quality of the reproduction of the plucked bass viol and a medium bass drum was obviously exaggerated.

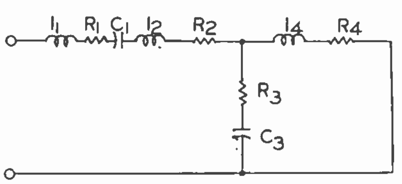
The magnitude of this droop is less than 10 db. Experiments with contemporary closed-back speakers show a greater droop than this and yet the speakers exhibit a "boomy" quality. This appears to be due to a response rise near 200 cycles which accounts for the sensation of bass. High diaphragm velocities in the 60–100 cycle range also produce the frequency and amplitude modulation of higher frequencies which in turn appears to enhance the bass sensation, somewhat in the manner of the beats between pipes to produce the "resultant" sub-bass used in some pipe organs.

Comparison with the reference corner speaker was conducted with the aid of a filter[6] in which the response in each octave could be raised or lowered over a 9 octave range. Deviations from "flat" were indicated and adjusted first on the midrange and tweeter levels, and then in the bass unit. Bass unit adjustments consisted in the port slit width for the 40–80 cycle range, and balance-network modifications in the 500–1000 cycle range.

In Fig. 1 is shown the ported box structure and equivalent circuit. It appears that this equivalent circuit is applicable to the present structure, with the exception that I_4 is increased over the values considered correct for "bass reflex" speakers.



(a)



(b)

Fig. 1—(a) One form of slot-back-loaded box.
(b) Equivalent circuit, where
I_1 = Inertance of suspended system
I_2 = Inertance of front air load
I_4 = Inertance of slot structure and back-radiation load.
C_1 = Compliance of suspension
C_3 = Compliance of cavity
R_1 = Resistance (dissipative) of cone suspension system.
R_2 = Resistance, front radiation
R_3 = Resistance, sound deadening material in cavity
R_4 = Resistance, back radiation
All values of R are variable with respect to frequency.

The "bass reflex"[2] system is theoretically adjusted so the resonant system of the cavity and port is antiresonant at the frequency at which the suspended system is resonant, or perhaps tuned slightly to one side to equalize the two resultant response peaks. But since the frequency of suspension resonance may vary over half an octave with different excursion amplitudes, the port adjustment is always crude. In the present system,

[3] Dickey, Caulton, and Perry, *Radio Engrg.*, vol. 16, pp. 8–10, 22; October, 1936. See also, H. F. Olson, "Elements of Acoustical Engineering," McGraw-Hill Book Co., Inc., New York, N. Y., 2nd ed., pp. 154–155; 1947. As originally disclosed, the port consisted of one or more tubes of considerable length, offering a higher inertance than the "Bass Reflex" port. The "Bass Reflex" name was a trade-mark applied by Jensen to this system. The trade-mark name has apparently been allowed to become public domain.

[4] P. W. Klipsch, "Subjective effects of frequency modulation distortion," *J. Audio Engrg. Soc.*, vol. 6, p. 143; April, 1958.

[5] This is to be distinguished from the slit port used in other speaker systems, *e.g.*, the Klipsch SHORTHORN and REBEL. The tunnel is a tube of considerable length and greater inertance than a mere port or slit of substantially zero length.

[6] "AUDIO BATON," manufactured by Blonder-Tongue Laboratory, Inc., Newark, N. J. To call this a handy tool would be an understatement. Individual calibration is suggested.

the same parameters apply, but are vastly different in magnitude, the gross difference being in I_4 which is several times the conventional value. The "second resonance" occurs in the low infrasonic range.

Addition of sound-absorbent material (never used in the "SHORTHORN"[7] speaker system series) was found necessary in the rectangular box structure to eliminate response anomolies in the 300–1000 cycles region. The ohmic peak at resonance is ordinarily lowered by addition of dissipation; in the present design, the voice coil impedance peak was raised instead of lowered by the addition of the sound-absorbent material. There have been other anomolies in the relation between electrical impedance and sound pressure measurements in the 30–60 cycle region in corner horns as well as in direct radiators. There is something of a shadow over interpretations based solely on voice coil impedance measurements. Currently, routine testing of the big corner horns includes sound pressure output which does not always correlate with the impedance measurements.

Performance

Fig. 2 shows the voice coil impedance curve with its final low-frequency peak at 8 cycles. No fundamental radiation occurs at this frequency.

Fig. 3 is a set of curves of a closed box of the same size as the CW, indicating the importance of corner placement,[8] and indicating the severe bass loss and aggravated peak-to-trough ratio caused by isolating a speaker from important reflecting surfaces. One could go so far as to say that any speaker design involving legs would be more in the interest of appearance than performance.

This set of curves was run (necessarily) indoors; two microphones were used to reduce the standing wave effects. Details may be found in the original paper.[8]

The CW is essentially this same box with the addition of a tunnel port at the rear and the sound absorbent material. The response curve of the new system is shown in Fig. 4, with curve (3) of Fig. 3 added for comparison. The curve of Fig. 4 was run in the same manner as those of Fig. 3, except that absolute pressure levels were used whereas relative levels only were used in the curves of Fig. 3.

Application

The speaker system has been designed for either corner or noncorner application. Thus, it can serve as a single monophonic speaker, or as any speaker or speakers in a two- or three-channel stereo system. While the center channel of a three-channel stereo system has been felt to be less demanding (some writers advocating a bass cutoff as high as 300 cps), recent experiments indicate a very considerable improvement in over-all effect with an extension of the bass range to 100 cycles, and a further improvement with extension to 50 cycles. Thus, the new speaker should be ideally suited for center channel use as well as being qualified for the flanking channels.

To achieve the near-precise tonal balance of the reference corner horns, the crossover-balancing network is

[7] Registered Trade Mark.

[8] P. W. Klipsch, "Corner speaker placement," *op. cit.*

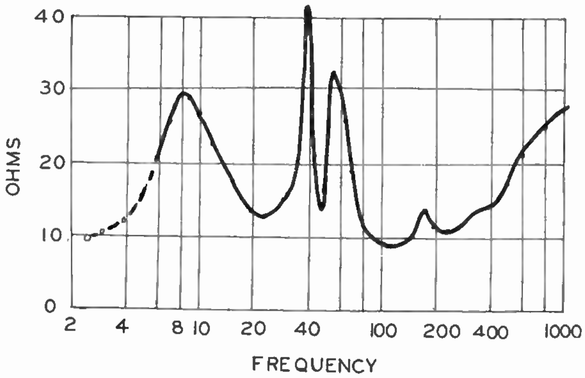


Fig. 2—Measured motional impedance of bass driver voice coil in the "MODEL CW" speaker system. The dc resistance of the voice coil is approximately 3.6 ohms.

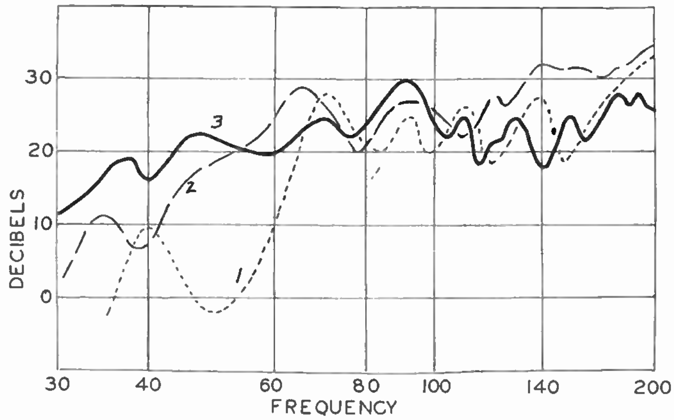


Fig. 3—Response of 6.7-cubic-foot closed box in 3 locations: Curve 1 box on 14-inch legs, four feet from each wall at a corner; Curve 2 box on floor, same location as 1; Curve 3, box on floor diagonally in corner. The curves show obvious successive improvements below 80 cycles, but less obvious and more important is the increased smoothness in the second octave—100 to 200 cycles. Technique described previously.[4] Courtesy *J. Audio Engrg. Soc.*, vol. 7; July, 1959.

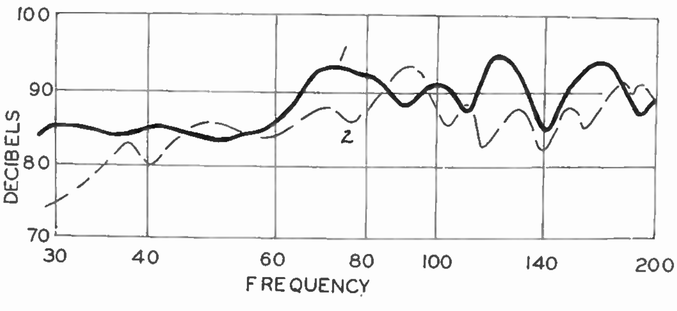


Fig. 4—Response of CORNWALL in corner (solid curve) and of closed box in corner (dashed Curve, 2; same as from Curve 3 of Fig. 3). The improved loading is evident in greater smoothness. The same order of magnitude of deterioration of response will occur in the new speaker, as with any other, as it is removed from a corner. Curve run in same room and conditions as Curve 3 of Fig. 3.

modified to the extent that response within a given element range, for example, the woofer, is tailored to narrow limits. In one instance, the response is modified approximately 2 db at 600 cycles. This required an extra inductor, a rather expensive component for so small a response change, but the close comparison with the reference speaker could not otherwise be achieved.

The system is closely compatible with the reference corner horns, and any feasible combination of the new type with the corner horns will be more closely compatible than any combinations ever achieved in the past. By the same token, their response accuracy makes the new units very excellent in their own right, singly, in combination with each other, or in combination with the reference corner horns.

Experience with a corner-horn back-loaded series of speakers indicated that relatively little bracing of wood panels was necessary. The CW differs in the larger expanses of panels and the parallel surfaces. The first experimental model in ½-inch plywood (quite adequate for the SHORTHORN) produced several panel resonances in the CW. The eventual structure required ¾-inch plywood plus extensive bracing, with some critical locations for the bracing.

While the structure appears to be quite simple, the rigidity required, the placement of bracing, and application of damping material seems to remove this item from the proper realm of supply in kit form. For these reasons the MODEL CW will be offered as an integrated system in proper balance and tested, with a choice of two drive systems.

The standard model will be three-way, using the same drive components employed in the reference corner-horn system. For an economy application, a carefully chosen extended-range 12-inch driver will be used. Thus, a user may start with a relatively low priced unit and later upgrade it by successive steps to the standard three-way unit.

Conclusion

Rather violent revision of the parameters applying to ported enclosures has resulted in a speaker system which compares favorably with a reference corner-horn system over a wide frequency range. Increased power output is possible compared with direct radiators mounted in total enclosures of traditionally low efficiency.

Because useful radiation occurs from both front and back in the first octave,[9] diaphragm excursion for a given sound power output is smaller than for a single direct radiator in a tightly closed box. Doppler distortion is negligible even at power levels sufficient to produce 100-db sound pressure levels at a 10-foot distance in an average room. Removal of the second-order resonance from the audible spectrum holds distortion resulting from infrasonic disturbances to negligible magnitudes. Exploration of the far infrasonic response reveals an impedance peak at 8 cycles where the back port finally "unloads" the diaphragm. At this frequency, distortion products are radiated, but this frequency is two octaves below 32 cycles, where steep attenuation of microphone, tape and disk response will preclude energy reaching the voice coil.

The feature providing planned upgrading appears to be the first in which a choice of eventual type is offered between a universal wall-or-corner speaker or the reference corner-horn system.

[9] From 40 to 80 cycles.

Appendix I

The "equivalent circuit" of Fig. 1(b) is valid only as follows: The suspension constants C_1, I_1 and R_1 are approximately constant and independent of frequency within the frequency range over which the diaphragm acts as a rigid piston. The radiation parameters I_2, R_2, I_4 and R_4 are complicated (and complex) functions of the ratio of piston diameter to wavelength or radiation area to wavelength. The capacitance C_3 is approximately constant for frequencies corresponding to wavelengths larger than four times the largest cavity dimension; where the box dimensions become larger than ¼ wavelength, the reactance varies between wide limits. The resistance R_3 seems to be negligible at very low frequencies and increases with frequency. The value of the circuit is to help visualize the multiple resonances but not to evaluate quantitatively the resultant performance. The manner of variation of the different parameters may be studied in any standard treatise on acoustics. The quantities I_2, R_2 were first studied, and reported, by Lord Rayleigh in 1877; graphic solutions by Stewart and Lindsay, about 1930, and Olson and Massa, in 1934.

Obviously the "equivalent" circuit is not a tool or device to be set up and used as a means of evaluating the acoustic system.

Appendix II

The voice coil impedance was explored into the infrasonic range and is shown in Fig. 2. The peak of 30 ohms at 8 cycles represents a no-load acoustic condition where only dissipative damping of the suspension limits the impedance. Any sound radiated at this frequency consists solely of distortion, but the sound output is negligible, besides which frequencies below 30 cycles would properly be heavily attenuated.

The existence of three peaks suggests an additional degree of freedom beyond what the "equivalent circuit" of Fig. 1, (C), would produce, but the complicated relationships between the acoustic load parameters and frequency may account for the third major peak. The second-order peak at 180 cycles appears to be a characteristic of the driver unit, as is the rise beyond 250 cycles.

A Plotter of Intermodulation Distortion*

E. F. FELDMAN† AND B. RANKY†

Summary—For use mainly in audio systems, the instrument plots on a CRT screen or chart recorder the amplitude of the difference-frequency tone vs the lower excitation frequency in CCIF intermodulation distortion measurements. The unit furnishes two swept equal-amplitude tones with a selected audio difference frequency to excite the tested system. Readout is on a modified audio-frequency spectrum analyzer which remains tuned to the difference frequency. The automatic excursion provides considerable test time economy in quality measurements of such devices as loud-speakers which exhibit wide variation in intermodulation with excitation frequency. Point-by-point techniques, in addition to being tedious, run the risk of missing critical conditions.

With slight modification, the unit is also capable of plotting the lower third-order distortion component at $(2f_1-f_2)$ vs excitation frequency.

A swept spectrum analyzer and means for single-line plots of frequency response curves are also integral components of the equipment. Switching facilities permit alternate displays on successive scans of any two of the modes of operation. For example, acoustic output and IM distortion vs frequency of loud-speakers may be plotted alternately for rapid comparisons.

Introduction

QUANTITATIVE evaluation of the quality of sound-reproducing equipment and linear networks generally requires the measurement of distortion. Harmonic analysis, although easily performed, is insensitive to distortion near the upper end of the useful band. There is conclusive evidence that listeners are more critical of intermodulation than similar relative degrees of harmonic distortion components.[1,2] Ratings of equipment are often in terms of both harmonic and IM characteristics.

One of the more widely used distortion tests is the so-called CCIF or difference tone method. The CCIF method requires a two-tone excitation to the network under test and an amplitude measurement of the difference frequency component. Use of the difference tone permits exploration up to the upper useful limit of the frequency band of interest. Intermodulation is taken as the ratio of amplitudes between the difference frequency component as compared to the sum of the amplitudes of the two desired output frequencies, as illustrated in Fig. 1. A related intermodulation distortion measurement of increasing interest involves determination of the third-order distortion for the two equal-tone excitation. The third-order distortion components occur simultaneously with the sum and difference terms. They are related mathematically to the cubic and higher-order odd exponent terms in the power series representation of a system transfer function. As shown in Fig. 1, the third-order components are at frequencies $(2f_1-f_2)$ and $(2f_2-f_1)$, where f_1 and f_2 are the excitation frequencies. Generally, the third- and higher-order components are smaller in magnitude than the sum or difference terms which are of second order. However, in systems which have symmetrical transfer functions, the difference term and other even-order nonlinearities tend to be suppressed. Examples are push-pull amplifiers and tape recorders with symmetrical magnetization curves.

The conservative approach in checking a new or unknown element is certainly to check for both even- and odd-order distortion components. Some preliminary conclusions can be gleaned from a harmonic analysis with a single-tone input. The second-, third-, and higher-order harmonics can be easily ascertained so long as the input frequency is several octaves below the upper frequency limit. It has been shown, however, that no valid general comparisons between intermodulation and harmonic distortion measurements can be made except for the simplest equivalent network configurations.[3]

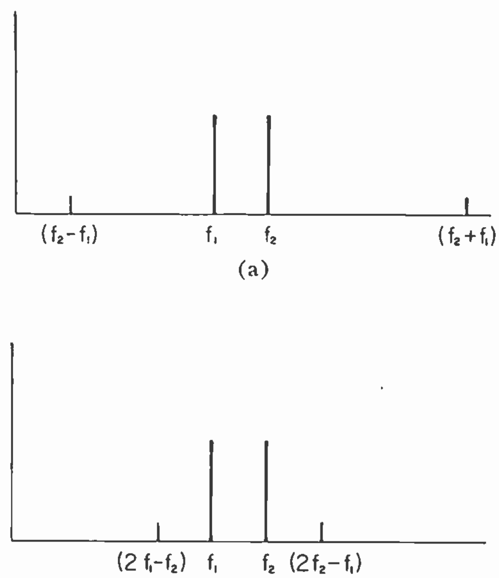


Fig. 1—Spectrum representations. (a) Sum and difference intermodulation components shown. (b) Third-order intermodulation distortion at $(2f_1-f_2)$ and $(2f_2-f_1)$.

* Received by the PGA, March 28, 1960. Previously published in the 1960 IRE International Convention Record, pt. 7, pp. 55–63.

† Panoramic Radio Products, Inc., Mount Vernon, N. Y.

[1] A. P. G. Peterson, "Intermodulation distortion," *Gen'l Radio Experimenter*, vol. 25; March, 1951.

[2] H. H. Scott, "Intermodulation measurements," *J. Audio Engrg. Soc.*, vol. 1, pp. 56–61; January, 1953.

[3] C. J. LeBel, "An experimental study of distortion," *J. Audio Engrg. Soc.*, vol. 2, pp. 215–218; October, 1954.

In addition to the problems of measuring each of several types of distortion, there is also the need to evaluate audio devices at various excitation levels and at many frequencies through the audible band.[4] Modern recording and broadcast facilities have pre-emphasis and shaping characteristics which tend to make distortion a distinct function of frequency. Loud-speakers, hearing aids, and other electro-mechanical devices have irregular frequency response characteristics and typically exhibit wide variations of distortion level with both frequency and level.

Convenient and accurate instruments are required for performing the rather sophisticated tests required in complete distortion measurements. This paper describes a comprehensive test set which plots intermodulation distortion as a function of frequency on a cathode ray tube, as well as serving as an automatic harmonic analyzer and a frequency response curve tracer.

Instrumentation

The use of scanning spectrum analyzers and companion sweep frequency generators for harmonic distortion measurements and response curve plotting has been widely accepted for some time.[5] With such a spectrum analyzer and a two-tone signal generator, intermodulation distortion has heretofore been measured on a point-by-point basis. A typical log-log Fourier distortion analysis resulting from a two-tone excitation is shown in Fig. 2. The difference term, third-order distortion components and higher-order harmonics, and IM are all shown here as pips which can be compared to the two main tones along the decibel amplitude calibration. If intermodulation distortion is to be measured for other input frequencies, the two input tones must be adjusted for each test point, intermodulation component levels of interest noted, and a graph drawn. While such analysis is generally instructive, the time required for IM-vs-frequency plots is considerable. In production testing, the effort would most likely be considered too costly.

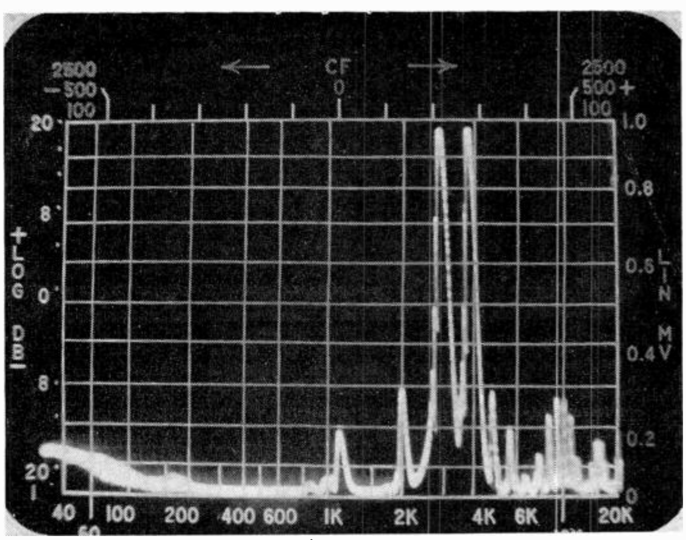


Fig. 2—Broadband spectrum analysis of amplifier output showing distortion caused by a two-tone input. The difference frequency component at 1.2 kc, third-order distortion components, harmonics, and higher-order components are measurable individually. Amplitude scale is two decade log. Frequency scan is logarithmic as calibrated along bottom of scale.

The new test set incorporates a slave two-tone generator and a sweeping spectrum analyzer as the basic elements in IM plotting. The two tones sweep through the audio spectrum maintaining a constant difference frequency. The analyzer is conditioned to remain tuned to the difference frequency. In such tests, the analyzer CRT horizontal deflection is proportional to the lower excitation frequency so that the readout represents difference frequency component amplitude vs lower excitation frequency. With the continuously sweeping generators, there is no inadvertent omission of critical input conditions. In production test, visual inspection of the CRT suffices to check IM characteristics. For permanent records, a photograph of the display or an adjunct chart recorder is used. Templates for go-no-go acceptance checks are readily inscribed on the CRT.

The elements of the equipment as arranged for CCIF difference-frequency plotting are shown in a functional block diagram in Fig. 3 (p. 127). The heterodyne spectrum analyzer is adjusted with two front-panel controls so that the swept oscillator excursion corresponds to the required band of excitation frequencies. The analysis limits for this instrument are 20 cps and 22.5 kc. Taking any instant during an excursion interval, the swept local oscillator of the analyzer is generating a frequency equal to (IF-f_1), where f_1 is the frequency shown to be the lower excitation component. f_1 is also the frequency at which the spectrum analyzer would be tuned at that instant in its normal operating mode. In the two slave sweep generators, this frequency is converted in balanced modulator stages such that output frequencies f_1 and $f_2=(f_1+\Delta f)$ are derived. The two generator outputs are linearly combined in a resistive network and adjusted such that their amplitudes are equal prior to injection into the tested device.

As shown in Fig. 3, the crystal oscillator output of one of the two slave sweep generators corresponds in frequency to the intermediate frequency of the analyzer plus the difference frequency, Δf. This frequency is used in the input converter as the substitute fixed local oscillator so that the analyzer remains tuned to Δf. Functionally, this simply requires a switching provision which disconnects the swept local oscillator from the analyzer balanced mixer and impresses the fixed crystal oscillator output in its place.

[4] J. S. Aagaard, "An improved method for the measurement of nonlinear audio distortion," IRE Trans. on Audio, vol. AU-6, pp. 121–130; November-December, 1958.

[5] E. F. Feldman, "An automatic sonic spectrum analyzer and curve tracer," 1955 IRE National Convention Record, pt. 10, pp. 10–16.

As the two test tones sweep with constant difference frequency through the selected excursion, the analyzer responds to the amplitude of the difference frequency component. In the output of the network under test, the two desired tones are present in addition to all distortion components. A bandpass filter is interposed between the tested network and the input to the analyzer so that overload resulting from the relatively large tones is avoided. The bandpass filter extends the linear dynamic range of the measurement beyond the normal limits of a heterodyne spectrum analyzer. In the unit which has been developed, residual distortion is better than 80 db down, making it compatible with high-quality audio networks and IM measurements. It is quite feasible to design a similar unit with a much greater linear dynamic range.

Selection of a particular difference frequency for IM plots is governed by the type of device being examined. It is desirable to have as low a difference frequency as possible, thus extending the useful test range from somewhat above the difference frequency to the upper limit of the network response. A low Δf also tends to keep the two exploring tones from diverging in amplitude. The difference frequency should correspond to a relatively sensitive point on the network frequency response curve to avoid artificially optimistic IM readings. The first equipment of this type was designed mainly for a 90-cps difference frequency. 90 cps is a particularly desirable value because it falls between the 60 cps and 120 cps hum frequencies which may very well be greater in amplitude than low-level intermodulation which is to be measured at the output of a vacuum-tube amplifier, for example. The selectivity of the spectrum analyzer is adjustable to 10 cps bandpass providing excellent rejection of the unwanted components. The high degree of selectivity is valuable in loud-speaker tests where the microphone often picks up extraneous broadband background noise.

Changing the difference frequency requires substitution of another crystal in one slave sweep generator and a suitable bandpass filter for the new Δf ahead of the analyzer. 2300 cps was selected for a tweeter production-line IM test setup with this equipment.

Comparison with Acoustic Output

Operation of the equipment for plotting the difference-frequency component alone shows the intermodulation variation with frequency but does not derive a quantitative IM rating. In another mode, the test set facilitates comparison between intermodulation amplitude and single-tone acoustic output to establish per cent IM. Readings are taken along the calibrated two-decade decibel axis of the cathode ray tube overlay. Fig. 4 shows the plots of a loud-speaker acoustic output vs frequency and 90-cps difference frequency vs the same frequency band. A variable attenuator, set at 20 db in the signal path, between the network output and the analyzer (see Fig. 5) is automatically switched out during IM plots so that 20 db is to be added to the apparent deflection differences between the two curves. It should be noted that the method employed does not take into account the second-tone output level and that the IM readings obtained are 6 db on the pessimistic side. For this speaker, the difference-tone component is largest when the exploring tones are in the region of the peak amplitude response. In Fig. 6 are screen photos showing superimposed IM and amplitude responses for some shaped response amplifiers.

Built-in switching circuits alternate the mode of operation of the test set between the acoustic output and IM plotting. The analyzer must "look at" the lower excitation frequency rather than the difference tone for frequency response tracing. This is accomplished by restoring the swept local oscillator output to the analyzer input mixer and bypassing the bandpass filter. Thus, the composite output signal is injected into the analyzer. With analyzer selectivity set for a value considerably less than the difference frequency, there is negligible interference because of the presence of the higher exciting tone in the analyzer input stages. A signal alternator is included in the test-set control panel to switch these several functions on successive scans. The schematic in Fig. 5 describes the function selection circuitry including the bi-stable attenuator.

The technique employed in plotting third order vs frequency is illustrated in Fig. 7. As in the previous IM plot, the heterodyne analyzer remains tuned to a fixed frequency. Two sweep generators move across the spectrum such that a constant lower third-order term is maintained. The synthesis is derived as follows.

Let the lower third-order term be defined as a constant

$$\Delta f = 2f_4 - f_3, \tag{1}$$

where f_4 and f_3 are the lower and upper excitation frequencies, respectively.

$$\therefore\ f_4 = \frac{f_3}{2} + \frac{\Delta f}{2}. \tag{2}$$

It is seen by (2) that the two input tones are harmonically related except for a constant offset frequency, $\Delta f/2$. One of the slave sweep generators in the difference-tone plotting arrangement, as shown in Fig. 3, derives $(f_1+\Delta f)$ where f_1 is the frequency axis calibration point of the spectrum analyzer. It is convenient to use the same signal as the upper excitation tone for the third-order configuration. Thus, the upper frequency is

$$f_3 = (f_1 + \Delta f). \tag{3}$$

From (2),

$$f_4 = \frac{f_1 + \Delta f}{2} + \frac{\Delta f}{2}.$$

$$\therefore\ f_4 = \frac{f_1}{2} + \Delta f. \tag{4}$$

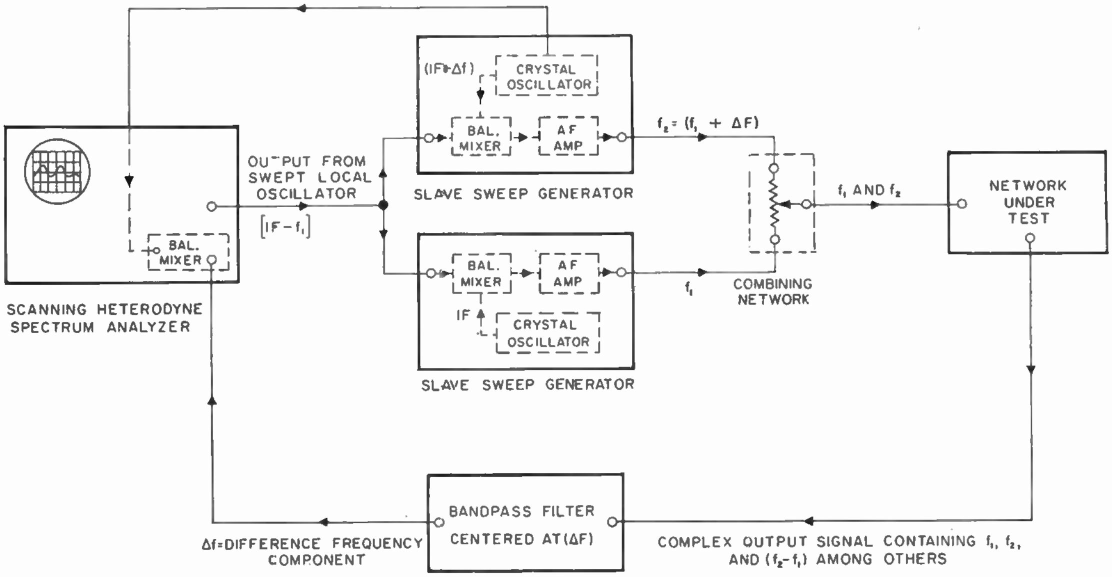
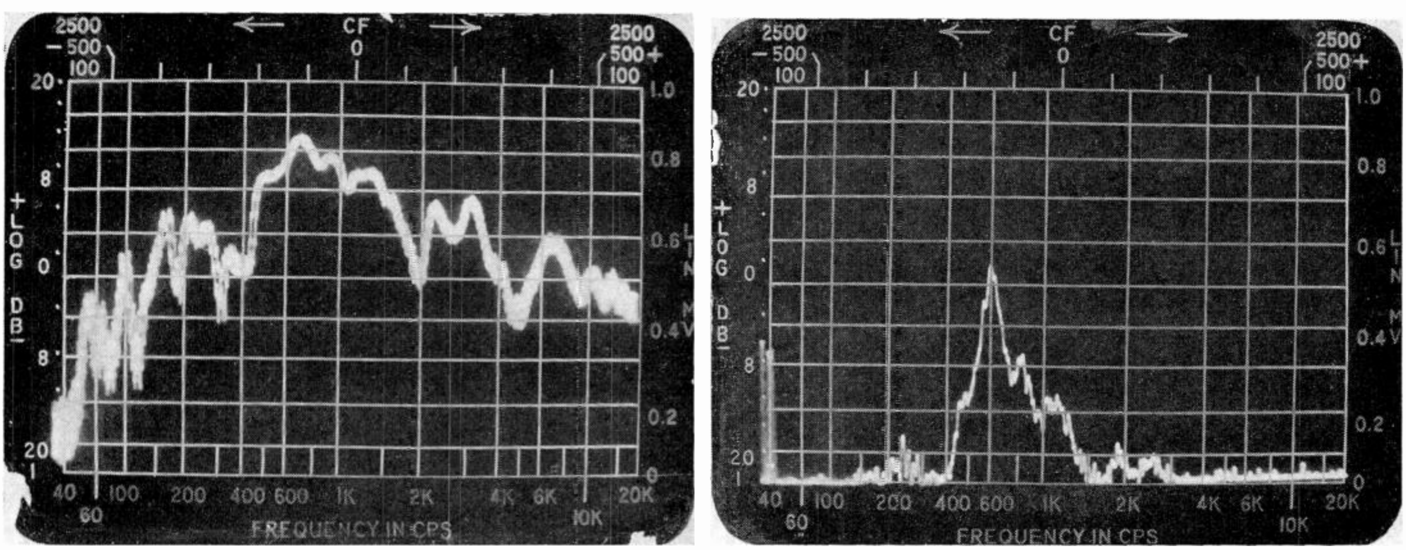


Fig. 3—Functional block diagram for difference tone (Δf) plotting vs lower excitation frequency, f_1.



(a) (b)

Fig. 4—Swept CCIF difference tone IM analysis of low-priced 10-inch loud-speaker. Linear frequency sweep from 500 cps to 5500 cps, log amplitude scale. (a) Amplitude vs frequency response (acoustic output) of speaker. (b) IM vs frequency plot along same frequency scale. A 90-cps difference was maintained between the two sweeping generator frequencies. Note that IM is greatest in region of peak response.

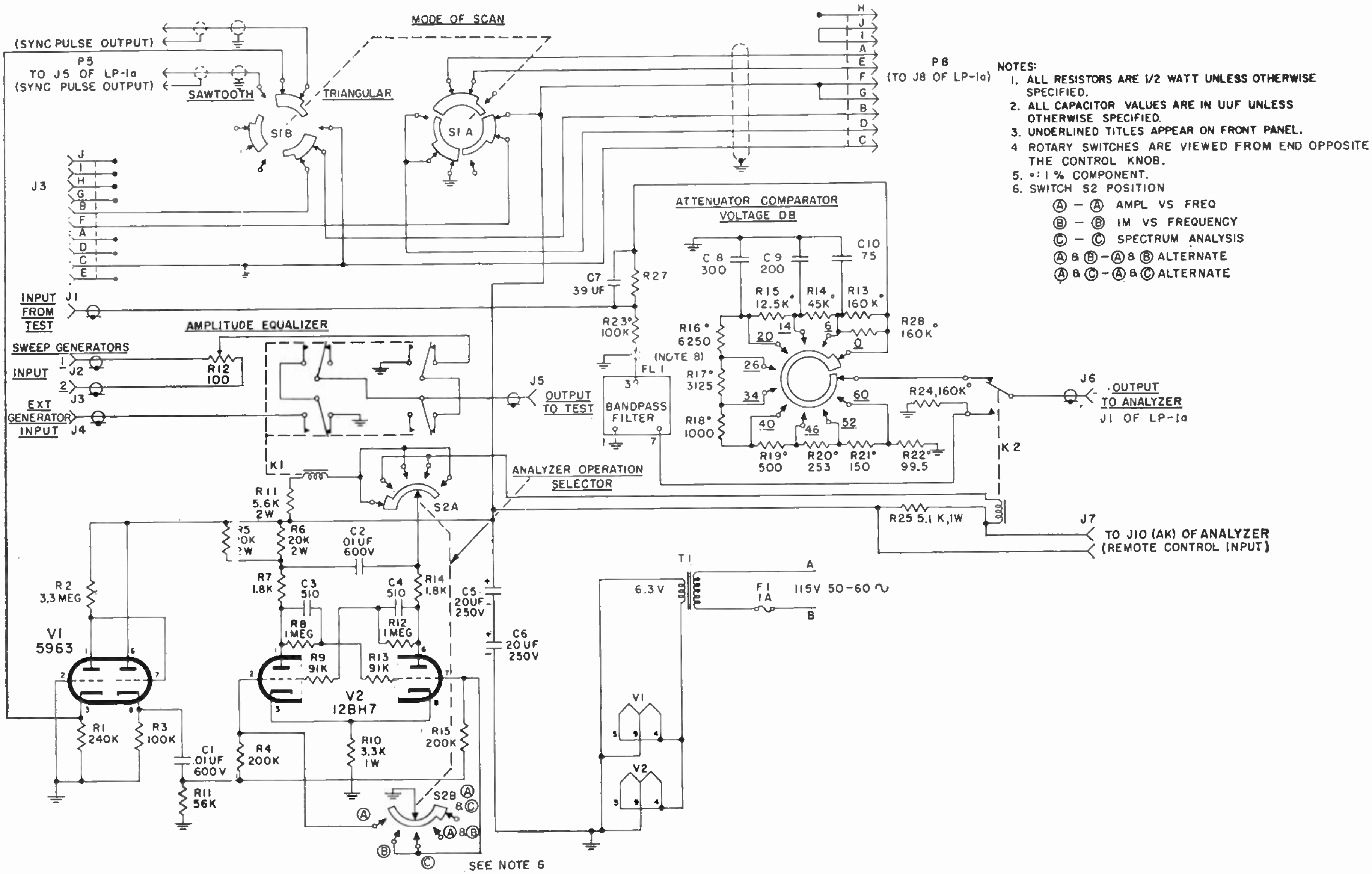


Fig. 5—Function control panel of test set.

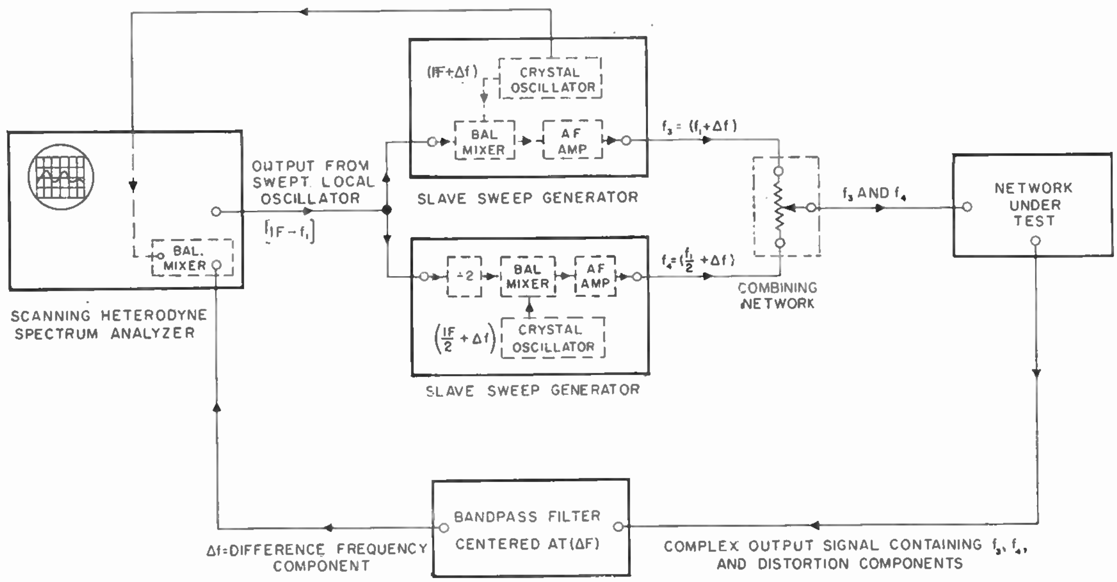
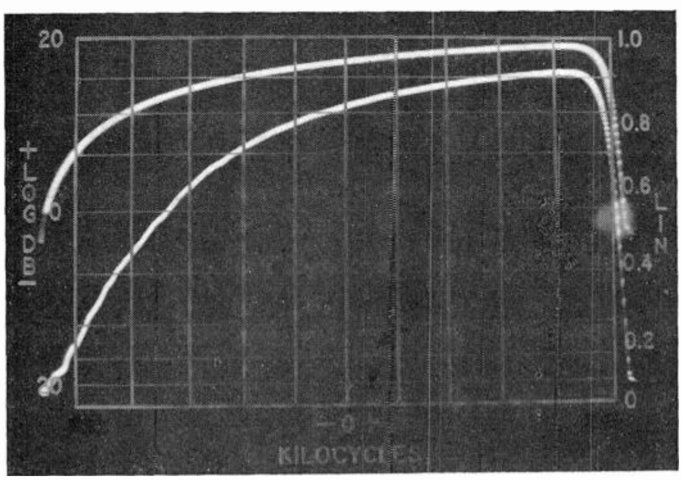
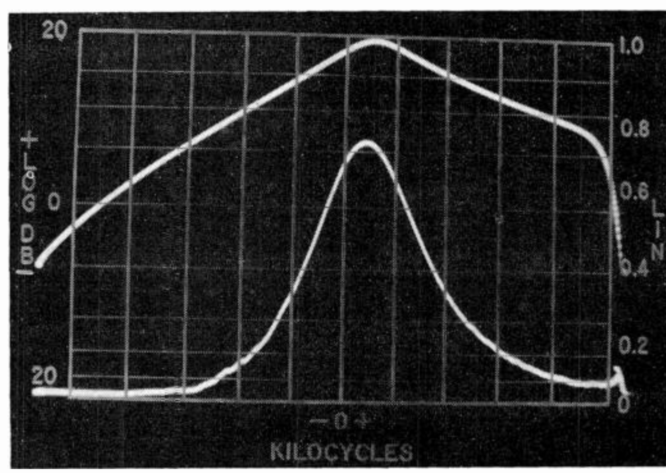


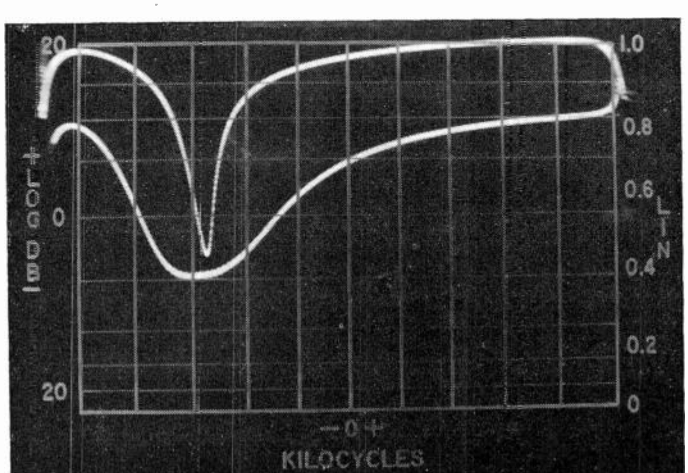
Fig. 6—Functional block diagram for third-order distortion plotting, $\Delta f=(2f_4-f_3)$.



(a)



(b)



(c)

Fig. 7—Difference frequency distortion and amplitude response plots for audio amplifiers with shaped response characteristics. Linear scan 40-db log amplitude scale. Add 26 db to the frequency response deflections when comparing with IM. (a) Increasing gain with frequency. Upper curve is response; lower curve shows IM which increases more rapidly than output level. (b) Band pass response, upper curve; steeply peaked IM characteristic, lower curve. (c) Response with band rejection slot, upper curve; IM, lower curve.

The lower frequency is half the upper plus a constant equal to the lower third-order frequency to which the analyzer is tuned. For this case, the distortion plot is vs the upper excitation frequency offset by a constant. By holding the fixed distortion frequency, one generator sweeps at twice the rate of the other through the selected spectrum segment. The difference frequency constantly varies throughout the scan. At one point, the two distortion modes coincide, causing an ambiguous readout. At coincidence,

$$2f_4' - f_3' = f_3' - f_4' = \Delta f \qquad (5)$$

where f_4', the lower excitation frequency $= 2\Delta f$. For the 90-cps offset, the 180-cps and 270-cps input frequencies generate equal lower second- and third-order distortion frequencies. To avoid this point, the useful band may be specified as beginning at $2\Delta f$ if desired. In other respects, third-order measurements are similar to the difference-tone operation of the system.

An alternate means of obtaining the third-order plot is to substitute a frequency doubler for the binary divider in the sweep generator shown in Fig. 6. The necessary crystal oscillator frequency is then twice the intermediate frequency plus the offset, Δf. The advantages of the doubler over the divider are that the frequency calibration relates to the lower tone as in the difference tone method and that twice as much sweep width is obtained. A reliable broadband doubler is much more costly to construct than a two divider, however.

An important design requirement for third-order tests is low second-harmonic content of the lower-frequency component. Twice the lower frequency is removed from the upper term by Δf, the offset frequency. Difference-frequency distortion caused by interaction between the second harmonic of the lower input frequency and the upper input frequency would be sensed by the analyzer and be indistinguishable from the third-order products.

Scanning Velocity

Scanning velocity is defined as the product of the scan repetition rate and the sweep width of the instrument. Although it is desirable to employ high scanning velocities so that the cathode ray tube plots will be easily visualized, the maximum permissible scan velocity is governed by the delay characteristics of the network under test and the analyzer IF filter bandwidth. Excessive scan velocities can cause erroneous results. There is an apparent frequency translation in the direction of scan as compared to the frequency calibration for zero scan velocity. In addition, responses are incorrectly plotted, especially in the regions of steeply sloped responses such as when tracing through a high Q resonant point of a loud-speaker.

The test set includes provision for adjustable sweep widths and scan repetition frequencies. A bidirectional linear scanning mode is employed so that direct visual comparison between the low-to-high and high-to-low scans can be made. If these two plots do not superimpose, the operator simply reduces the scan speed. Selection of slower speeds than those which correspond to virtual superimposition merely increases the test time without any commensurate improvement in accuracy.

In some typical loud-speaker difference-tone intermodulation measurements, with a 90-cps difference frequency, a 10-second scan time was found to be adequate when the sweep width was restricted to a few kilocycles. In this case, the time delay in the spectrum analyzer which was set for a 10-cps IF filter bandwidth was greater than the delay resulting from the loud-speaker. When the difference frequency was increased, so that a broader analyzer IF could be used, more rapid scans were feasible. However, the useful testing range is reduced with higher difference frequencies, and important regions of interest may be omitted. Depending upon laboratory or production-line test requirements, a compromise is made among scan time, selectivity, and useful scanning bands. Very often a critical frequency region of a response characteristic is determined after some experimentation so that each production-line sample need not be tested in the entire audio range. Typically, the test set can be used for sweep widths up to 5 kc centered in the audible spectrum. For tests of more than 5 kc bandwidth, the center frequency control is adjusted accordingly. The bidirectional scan intervals may be set as required between a fraction of a second and over one minute per scan.

"A manual" tuning control is also furnished. When it is used, the automatic scanning is switched out and the operator may select any point within the sweep width range for stable tuning. This control adjusts the swept local oscillator frequency in the analyzer, which in turn determines the frequencies of the two slave signal generators. The same correspondence between X-axis deflection and frequency obtains for the "manual" as for the automatic scanning operation. This permits the operator to examine critically, for example, maximum and minimum points located during the rapid scans. By rocking the "manual" control about any desired point, he can define small spectrum segment response.

As shown in Fig. 5, three basic modes of operation are selected with the "Analyzer Operation Selector" switch. These are: "A," frequency response curve tracing; "B," IM vs frequency, "C," and spectrum analysis. In the spectrum-analysis position, external signal generators are used to derive input signals for harmonic analysis or insertion of frequency markers. (See Fig. 8.) The final two positions of the selector switch set up alternate operation between "A" and "B" or between "A" and "C." The switching between functions is accomplished by triggering the relay switch with a pulse derived at the end of each scan interval. The bi-stable switching provision permits free selection of sweep speeds.

The spectrum analyzer is also equipped with a broadband logarithmic scanning mode, as shown in Fig. 2, which is directly calibrated along the bottom of one cathode ray tuber overlay. The log sweep repetition rate

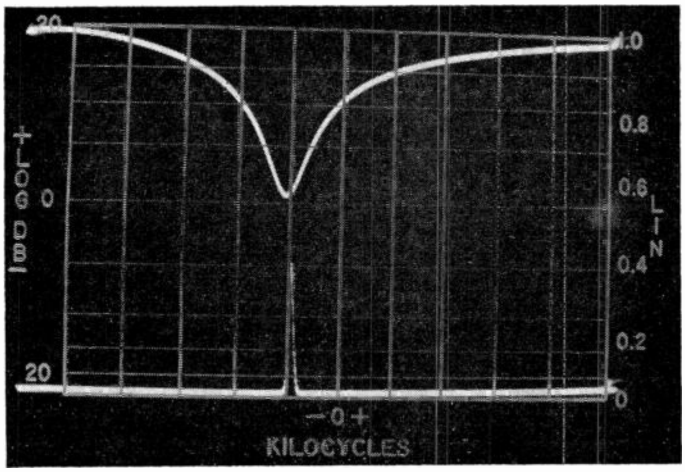


Fig. 8—Frequency marker pip derived from external signal generator is displayed with response curve on alternate scans.

is fixed at one per second and is well suited for rapid harmonic analysis. For frequency response tracing, the broadband log sweep is valuable for preliminary plotting, so that the area of concentration for the slower magnified linear scans can be located rapidly without searching through the entire band.

Fig. 9 pictures the Audio System Test Set in a relay rack-style cabinet.

Acknowledgment

The authors wish to acknowledge the significant contributions of W. I. L. Wu and L. J. Malmsten to this work.

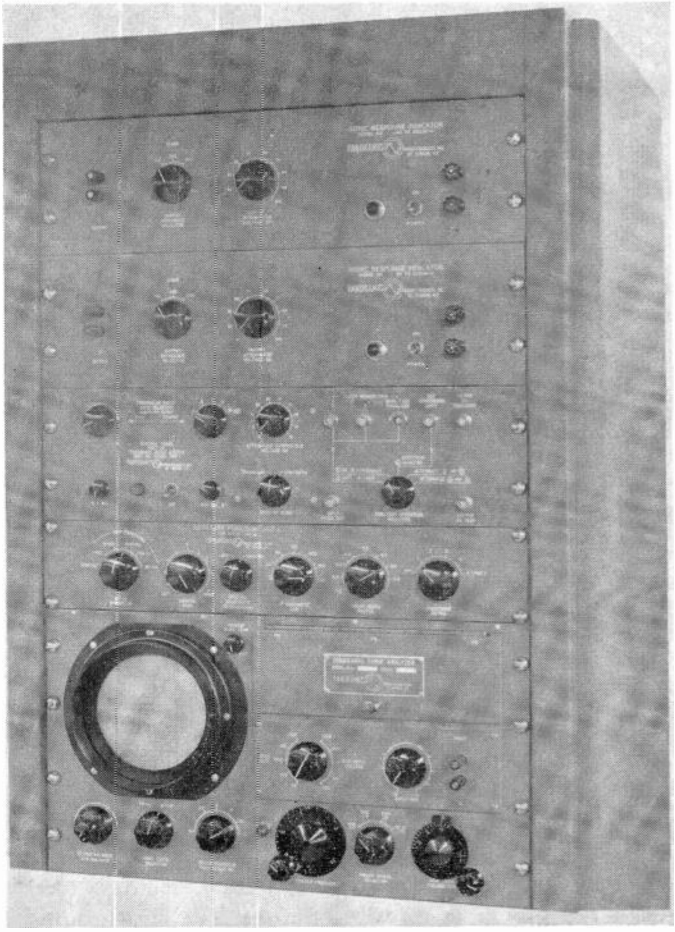


Fig. 9—Photo of the Audio System Test Set.

Calculation of the Gain-Frequency Characteristic of an Audio Amplifier Using a Digital Computer*

DONALD E. BRINKERHOFF†

Summary—**Electronic circuit designs are most commonly made by "analog" (or "breadboard") methods. This is particularly true in arriving at the desired gain-frequency response characteristics. The mathematical approach to circuit analysis has suffered from the distinct disadvantage of involving the solution of large systems of simultaneous equations with complex, frequency-dependent coefficients often requiring days to solve by hand calculations.**

The ability of digital computers to solve large systems of simultaneous equations at high speed and without mistakes allows the audio engineer to solve many problems faster by using classical methods than by use of the breadboard technique.

Introduction

THE USE of digital computers in solving engineering problems is becoming widely accepted practice. The advantages of this electronic device, however, are being exploited more effectively by engineers outside the electronics field. The few applications of digital computers to electronic circuit design found in the literature[1,2] refer to switching and timing circuits where transistors or tubes are alternately biased at cutoff and saturation—definitely not an audio application!

One explanation for the relative lag in the application of digital computers to electronic design is the fact that an electronic circuit is itself essentially an analog computer. A large quantity of data can be obtained on many parameters quickly and inexpensively by merely twisting the knobs on a group of decade boxes. However, in a mechanical problem, for instance, more time and expense are usually involved in changing the mass, compliance, or damping of a dynamic system.

A mathematical treatment of the equivalent network of a electronic circuit produces large systems of simultaneous equations with complex, frequency-dependent coefficients which often require days instead of minutes to solve by hand calculation. As a result, the average electronic engineer has developed a "feel" for the analog approach and has all but forgotten the classical methods.

The ready facility of digital computers to perform innumerable arithmetic operations at a high speed and without mistakes makes it now feasible to solve large systems of equations. Therefore, matrix algebra can be profitably revived from a theoretical science to a useful engineering tool, enabling audio engineers to return to more accurate and classical methods of analysis.

Purpose

In this paper, we will discuss the calculation of gain vs frequency of a transistor audio amplifier used in the output stage of an automobile receiver. We will also investigate the effect of varying some of the parameters on the over-all gain and response of the stage.

Automotive styling requirements result in varied acoustic load characteristics presented to the loudspeaker.[3] The frequency response of the receiver as well as the loudspeaker must therefore be carefully tailored to insure the proper tone balance for a high-quality automobile receiver. Most of the tailoring of the audio response of the receiver is naturally done in the audio stage.

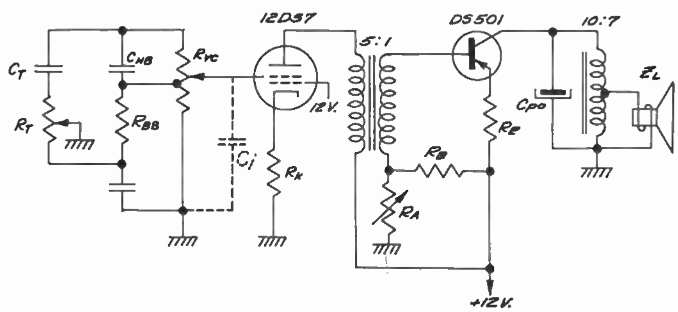


Fig. 1—Schematic diagram of audio stage.

Fig. 1 shows a schematic diagram of a typical audio output stage. A Delco DS501 power transistor is used in a transformer-coupled common emitter circuit. The driver in this case is a low-voltage tube used in hybrid auto receivers. We will limit this discussion to the portion of the audio amplifier from the driver grid to the load. In general, the design objective for this part of the stage is a response as flat as possible in the bass and middle range with a suitable roll-off of the high response to suppress electrical noise and adjacent-channel interference on the AM band.

* Received by the PGA, March 4, 1960; revised manuscript received March 11, 1960. Previously published in the 1960 IRE INTERNATIONAL CONVENTION RECORD, pt. 7, pp. 73–79.

† Delco Radio Division, General Motors Corp., Kokomo, Ind.

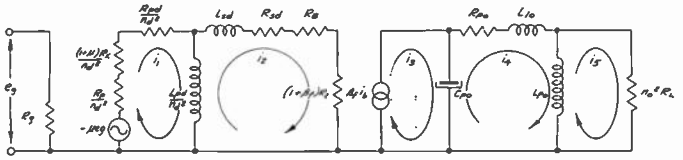
[1] J. Alman, P. Phipps, and D. Wilson, "Design of a basic computer building block." *Proc. 1957 Semiconductor Symp.*, 119–124.

[2] G. H. Goldstick and M. Kawahara, "Application of the NCR data processor to the synthesis of a digital computer building block," 1959 IRE NATIONAL CONVENTION RECORD, pt. 4, pp. 204–217.

[3] B. A. Schwarz and D. E. Brinkerhoff, "Some observations on reproduced sound in an automobile," *J. Audio Engrg. Soc.*, vol. 6, pp. 58–63; January, 1958.

Procedure

Our first step is to draw an equivalent ac circuit of this stage to provide a model for our mathematical analysis. At this point, a decision must be made as to how exact an equivalent circuit is to be used. In hand calculations, many approximations are usually made to simplify calculations. On a digital computer, however, once the program is prepared, an exact solution can be calculated almost as rapidly as an approximate one. Factors which obviously have negligible effect on performance are, of course, eliminated to simplify programming.



R_g = Grid resistor
R_p = Plate resistance
R_k = Cathode resistor
R_{pd} = Pri. res.-driver trans.
L_{pd} = Pri. inductance-driver trans.
L_{ld} = Leakage ind.-driver trans.
n_d = Turns ratio-driver trans.
n_o = Turns ratio-output trans.
R_{sd} = Sec. res.-driver trans.
R_B = Base resistor
R_E = Emitter resistance (internal and external)
B_f = Large signal current gain at frequency (f.)
C_{po} = Output pad
R_{po} = Pri. res-output trans.
L_{po} = Pri. ind.-output trans.
L_{lo} = Leakage ind.-output trans.
R_L = Load resistance.

Fig. 2—Equivalent circuit of audio stage (from driver grid to load).

Fig. 2 shows the equivalent circuit of the output stage. A five-mesh network is required to represent the stage. Since the characteristics of the driver tube are similar to those of a triode, an equivalent constant voltage generator is used to represent the source. The constant current generator type equivalent circuit[4] is used to represent the transistor. In the equivalent circuit of the transformers, we have reduced to unity turns ratios, and have neglected distributed capacitances, coupling capacitance, hysteresis and eddy current losses. We have also simplified the problem by substituting a resistance load for the electrical analog of the loudspeaker load. Note that this circuit has 16 parameters which can be varied to obtain the desired performance. Using the method of loop analysis involving the repeated application of Kirchhoff's Voltage Law, we obtain five simultaneous equations in five unknowns. It should be pointed out that analysis of the volume control with its tone compensation requires three more meshes. This part of the circuit has been treated in a separate program.

Fig. 3 shows the five equations derived from the network. Also shown is the equation for over-all gain of the stage. The five equations must be solved simultaneously to get i_5 in terms of e_g to obtain the over-all transfer function of the stage.

Since practically all of the coefficients in the five equations are frequency-dependent, the equations must be solved for each of several frequencies required to describe adequately the gain-frequency characteristics. This is the task that is made feasible by a digital computer.

All manufacturers of electronic digital computers have available machine programs for the solutions of systems of simultaneous equations. Most of these prepared programs, however, require extra attachments to a medium-speed basic computer such as machine floating decimal and an index register. If this equipment is not available, a general equation for i_5 in terms of e_g can be derived by solving the simultaneous equations once by hand calculation using symbols for the coefficients, as shown in Fig. 4. The matrix form for i_5 is set up and solved by the use of determinates, producing the equation shown on the lower right.

Substituting the original coefficients of the simultaneous equations produces the somewhat messy equation for i_5 shown in Fig. 5. After collecting real and imaginary terms, the equation is programmed in the computer to be solved at 18 frequencies across the audio band for each set of circuit values. The program then calls for the solution of the gain equation:

$$G_{\mathrm{db}} = 10 \log_{10} \frac{i_5^2 N_0^2 R_L R_g}{e_g^2},$$

where

N_0 = turns ratio of output transformer
R_L = load resistance
R_g = grid resistor
e_g = voltage at grid,

giving the over-all gain of the stage in decibels. A table of logarithms to the base ten is stored in the memory for use in converting the input-output power ratio into gain in db.

The effects of variation of transistor current gain, driver transformer, and output transformer design can be quickly determined by the use of this computer program.

Results

Fig. 6 shows a family of curves of stage gain vs driver transformer primary inductances at low frequencies. These data quickly indicate the optimum range of values from a cost-performance standpoint.

[4] A. W. Lo, R. O. Endres, J. Zawels, F. D. Waldauer, and C. C. Cheng, "Transistor Electronics," Prentice Hall, Inc., New York, N. Y.; 1957.

$$1)\quad \frac{(\mu+1)R_k+R_p+R_{pd}+j\omega L_{pd}}{n_d^2}i_1-\frac{j\omega L_{pd}}{n_d^2}i_2=\frac{\mu e_g}{n_d}$$

$$2)\quad -j\frac{\omega L_{pd}}{n_d^2}i_1+\left[i\frac{\omega L_{pd}}{n_d^2}+j\omega L_{sd}+R_{sd}+R_B+(1+\beta_f)R_E\right]i_2=0$$

$$3)\quad \beta i_2-li_3=0$$

$$4)\quad \frac{-1}{j\omega C_{po}}i_3+\left[\frac{1}{j\omega C_{po}}+R_{po}+j\omega(L_{Lo}+L_{po})\right]i_4-j\omega L_{po}i_5=0$$

$$5)\quad -j\omega L_{po}i_4+(j\omega L_{po}+n^2R_L)i_5=0$$

$$G=\frac{p_0}{p_1}=\frac{i_5^2n_0^2R_LR_g}{e_g^2}$$

Fig. 3.

$$1)\quad A_1i_1+B_1i_2+Oi_3+Oi_4+Oi_5=K_1$$

$$2)\quad A_2i_1+B_2i_2+Oi_3+Oi_4+Oi_5=0$$

$$3)\quad Oi_1+B_3i_2-li_3+Oi_4+Oi_5=0$$

$$4)\quad Oi_1+Oi_2+C_4i_3+D_4i_4+E_5i_5=0$$

$$5)\quad Oi_1+Oi_2+Oi_3+D_5i_4+E_5i_5=0$$

$$i_5=\frac{A_2B_3C_4D_5K_1}{(A_1B_2-AB_1)(D_5E_4-D_4E_5)}$$

Fig. 4.

$$_5=\frac{\left(\frac{\mu e_g}{n_d}\right)\left(\frac{j\omega L_{pd}+L_{po}}{n_d^2C_{po}}\right)\left(\frac{\beta_m}{1+jf/f_{\beta C}}\right)}{\left\{\frac{R_p+R_k(\mu+1)+R_{pd}+j\omega L_{pd}}{n_d^2}\left[i\omega\left(\frac{L_{pd}}{n_d^2}+L_{sd}\right)+R_{sd}+R_b+\left(1+\frac{\beta_m}{1+kf^2}\right)R_e\right]+\frac{\omega^2L_{pd}^2}{n_d^4}\right\}\left\{-\omega^2L_{po}^2-\left[j\left(\omega L_{po}+\omega L_{lo}-\frac{1}{\omega C_{po}}\right)+R_{po}\right](j\omega L_{po}+n_o^2R_L)\right\}}$$

$$G_{\text{db}}=10\log_{10}\frac{p_0}{p_{1n}}=10\log_{10}\frac{i_5^2n_0^2R_LR_g}{e_g^2}$$

Solve gain equation for the following frequencies: 50, 75, 100, 200, 300, 400, 600, 800, 1000, 1500, 2000, 3000, 4000, 5000, 6000, 8000, 10,000 and 12,000 cps.

Fig. 5.

Above a primary inductance of six henries we have diminishing returns in stage gain, while below two henries a small reduction in cost is accompanied by a large reduction in performance. A few seconds of machine calculation time produced data of this type which would require hours to measure in the actual circuit (not including the time required to build the six driver transformers considered here).

Fig. 7 shows a similar family of curves of stage gain at low frequencies vs primary inductance of the output transformer which can be used to select the correct value for the performance required.

Fig. 8 shows a family of curves of stage gain vs leakage inductance of the output transformer at high frequencies. Note the response rise caused by resonance between the leakage inductance and the output pad. These curves show that a wide range of leakage inductances can be tolerated over the frequency range of interest in AM receivers.

Fig. 9 shows a family of over-all response curves using a range of transistor current gains. This shows that the transistor has a negligible effect on tone balance in the circuit under consideration—only the stage gain is affected.

Fig. 10 shows a family of over-all response curves using various values of the output pad. This is one curve that is easily obtained by the "decade box" method. Note that an output pad at this point produces a quick roll-off in the 6 to 10 kc range without appreciably attenuating the "middle high" range which is important in adding "presence" to the program material. After we have selected the desired values, Fig. 11 shows the calculated over-all response of the stage using the set of values chosen compared to the measured response on the actual circuit using the same set of values. The calculated and measured response shows fairly good agreement throughout the audio range.

Fig. 12 shows the phase shift of the stage calculated from the real and imaginary terms of the i_5, e_g ratio.

Fig. 13 shows the computer equipment used on this problem.

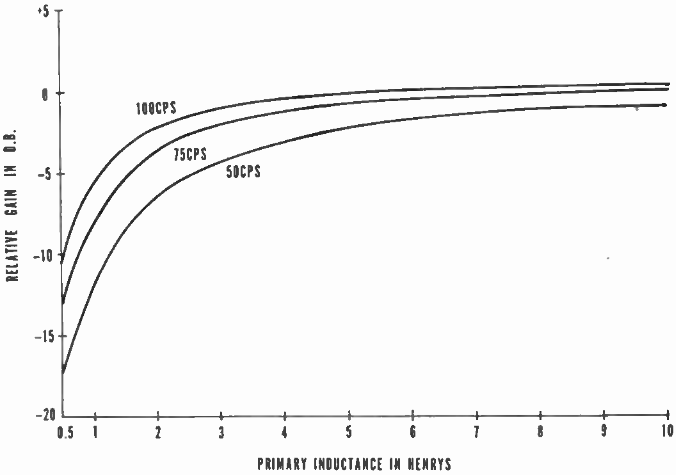


Fig. 6—Stage gain vs primary inductance of driver transformer.

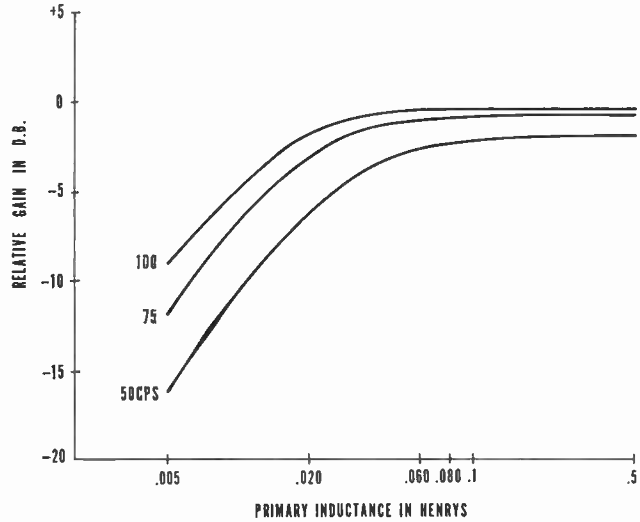


Fig. 7—Stage gain vs primary inductance of output transformer.

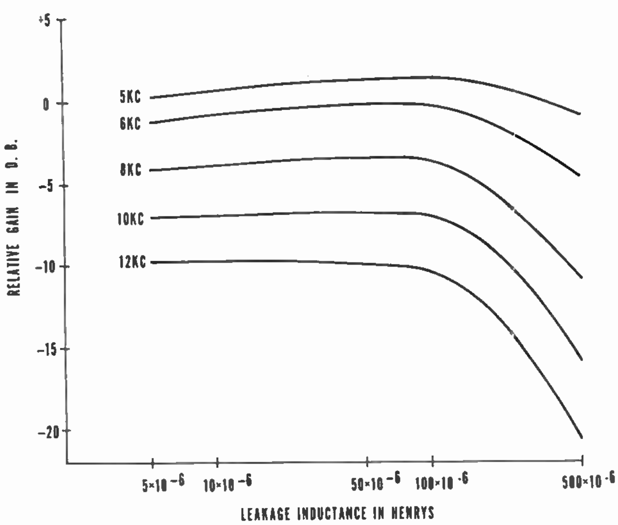


Fig. 8—Stage gain vs leakage inductance of output transformer.

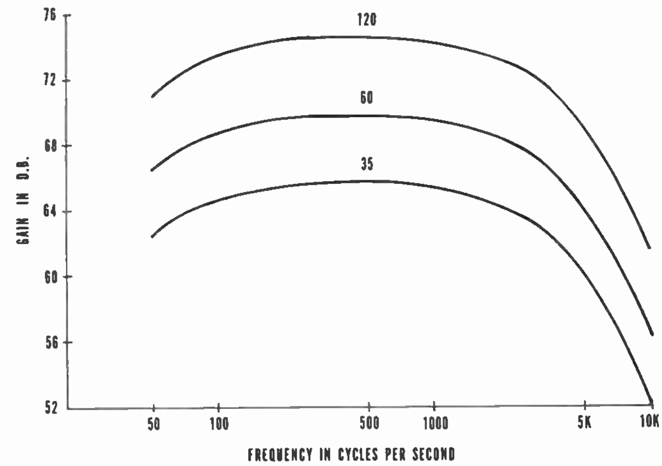


Fig. 9—Stage gain vs transistor DS 501 power transistor; gain from 12DS7 grid to load.

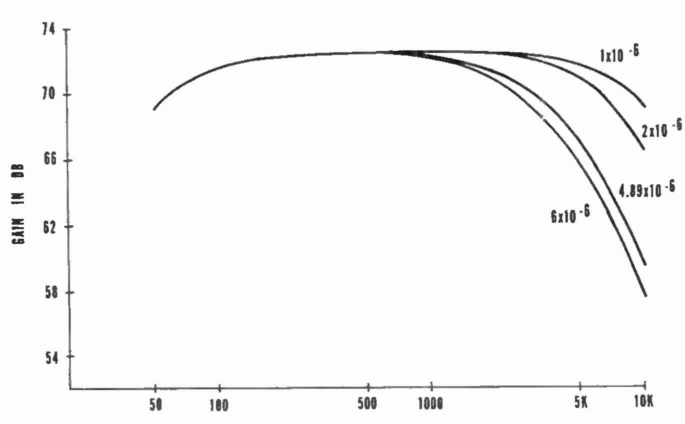


Fig. 10—Stage gain vs output pad.

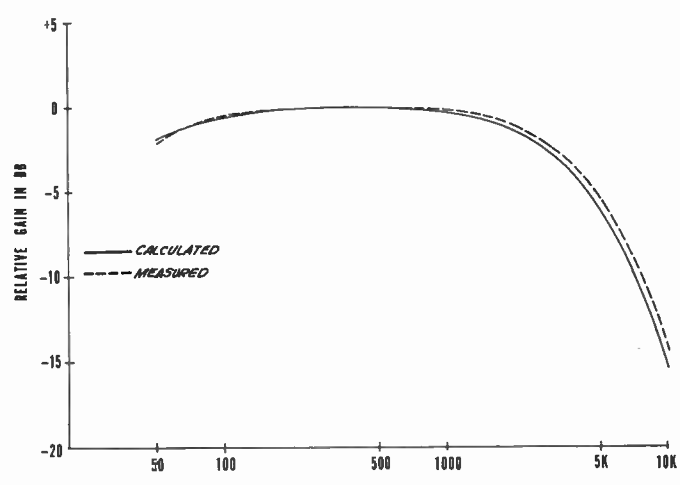


Fig. 11—Calculated vs measured response of transistorized audio output stage.

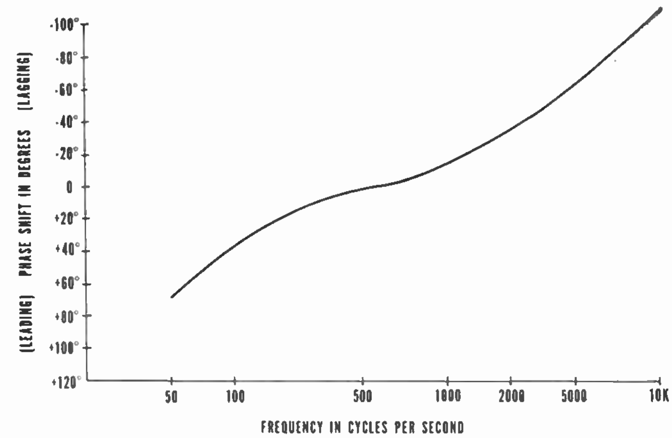


Fig. 12—Phase shift vs frequency for transistorized output stage.

Applications and Extensions

In the foregoing program, we have merely accelerated the standard design technique with the high-speed programmed computation provided by a digital computer. This program, however, indicates the possibilities of more sophisticated programs where the input consists of the performance specifications—in this case, the desired frequency-gain characteristic. The computer could then be programmed to scan the circuit parameters and automatically search out all possible combinations which will produce the required performance. These combinations can then be studied for other characteristics such as linearity, power handling, and cost. A few minutes of calculating time can produce a detailed analysis which would require months of laboratory work to obtain.

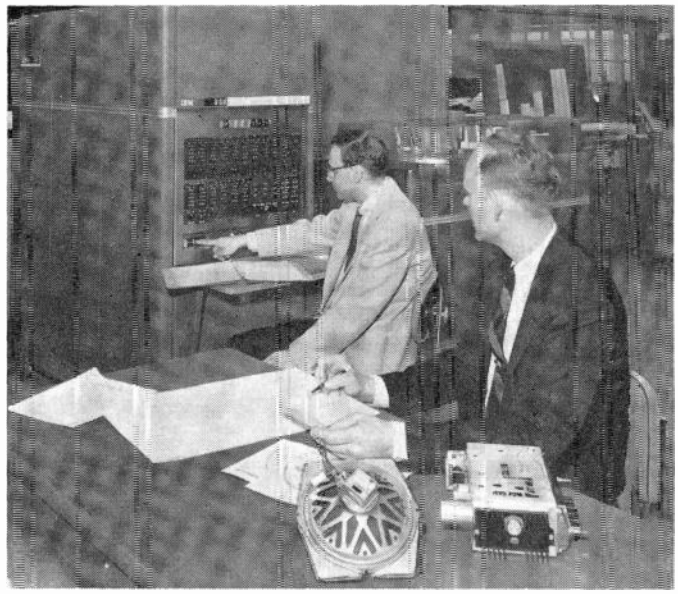


Fig. 13—Checking computer input data on a Delco Radio receiver problem are C. F. Springer (left), Engineering Programmer, and the author (right).

The present program does not, of course, represent a complete analysis on the audio stage. Such characteristics as harmonic and intermodulation distortion vs power output can best be determined by measurement on a breadboard. The above does illustrate, however, that there are many parameters that can now be analyzed faster by numerical methods, and indicates some of the possibilities of this new tool to the audio engineer.

Photo-Sensitive Resistor in an Overload-Preventing Arrangement*

J. RODRIGUES DE MIRANDA†

Summary—The signal from a pre-amplifier or tuner is supplied to the input terminals of the power amplifier by means of a voltage-dividing network, the element in parallel to these input terminals being a photo-sensitive resistor of new design. Facing this photo-resistor in a lightproof enclosure is a neon lamp connected to the output voltage of the amplifier. Because a neon lamp ignites at a fixed voltage, the value of the photo-resistor resistance will be decreased as soon as this voltage is reached. Consequently, the input signal to the amplifier will be decreased as well, making overload substantially impossible.

Properties of the CdS Photo-Resistor

One of the more recent photo-sensitive devices is the sintered cadmium sulphide (CdS) photo-resistor. The properties of these resistors and some of their applications have been described extensively elsewhere. Therefore, I will confine myself here to some general information.

As is the case with all nonconductive solids, practically all electrons of CdS (in dark) are bound to nuclei of the atoms which make up the crystal. However, when radiation (light) falls on the substance, the energy of this radiation is absorbed by some of the outer electrons, with the result that their binding force becomes so much smaller that they behave as free electrons. Electricity transport is now possible, the CdS has become conductive. The conductivity will depend on the amount of radiation absorbed, as can be understood as follows:

After some time, the "free" electrons are recaptured; if radiation continues to fall on the material, an equilibrium is reached where just as many electrons are recaptured as are set free. With stronger radiation, the equilibrium will move towards a higher average number of free electrons.

We thus see that CdS can be used as a variable resistor, the value being controlled by the amount of light falling on it. The variation in resistance can be quite considerable: a typical execution of the photo-resistor has a "dark" resistance of 100 megohm and a "light" resistance of 100 ohm; a variation ratio therefore of $1:10^6$. Another property which is of importance for the subject of this paper is the speed of variation.

The variation of resistance against time as a result of a certain illumination follows a law which is very complicated (because of the action of so-called electron traps) but resembles somewhat an exponential law. The initial rising is very steep, the steepness depending on the amount of impinging light. When the light is removed, the resistance increases, and here again the initial increase is very quick whereas the final "dark" value is reached only after a relatively long time.

By setting the working range of the resistance values appropriately, we can obtain, within this range, a relatively quick drop in resistance and a relatively slow rise. We can say that we make use of the fact that radiation sets the electrons free almost immediately, whereas, when the radiation disappears, the electrons will not recombine with the ions all at the same moment and immediately, so that a considerable time is required to reach the "dark" value. The time of decrease and increase of the resistance is shown in Figs. 1 and 2, respectively. (Note: the time-scales are different.) The curve of Fig. 1 has been taken with relatively weak light, as is obvious from the relatively high final resistance.

Application of a Photo-Resistor as a Volume Control

From the above it can be seen that, among other applications, the CdS cell can be readily applied as a volume-control device for an amplifier by combining, with a small lamp the current through which can be regulated. The circuit diagram of a simple device of this kind is shown in Fig. 3. The time which is needed to decrease the resistance to 1/10th of a certain value depends on the initial value, as is shown in Fig. 1. It takes about 1 msec for the resistance to drop from 10 megohms to 1 megohm, but another 5 msec to reach the value of 0.1 megohm. The recovery time, including the tail, is much longer and can, depending on the treatment of the material during production, be anywhere between 0.1 second and 2 seconds (see Fig. 2). It will be clear that if the photo-resistor R_2 is illuminated, the input signal drops noticeably as soon as the value of R_2 becomes smaller than that of R_1.

In Fig. 3, the manual control is obtained by means of variation of R_3, which controls the amount of light falling on R_2. Resistor R_3 can obviously be at a considerable distance from the photo-resistor, a distance not restricted by factors influencing audio-frequency considerations. As a matter of fact, an additional advantage of a CdS volume control would be that the controlled resistor can be located at the most favorable point from a circuit point of view, while the controlling

* Manuscript received by the PGA, May 16, 1960. Presented at the Audio Engineering Society, New York, N. Y. October 5, 1959, and reprinted with their permission from *J. AES*, vol. 8, no. 3, July, 1960.

† Radio Apparatus Development Labs., N. V. Philips, Gloeilampenfabrieken, Eindhoven, Holland.

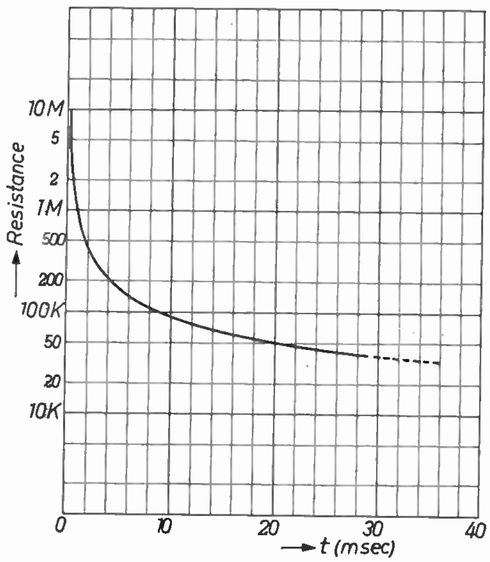


Fig. 1.

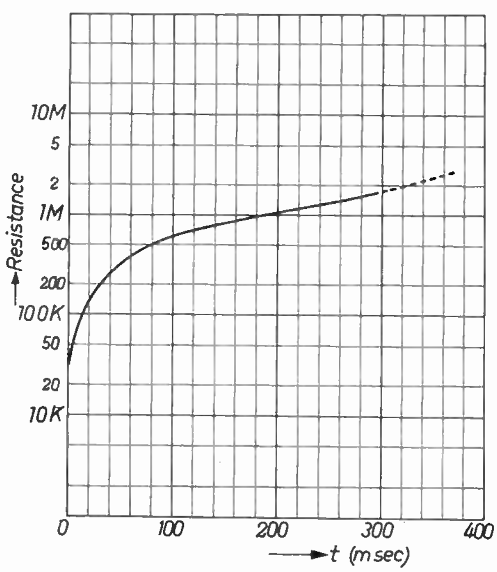


Fig. 2.

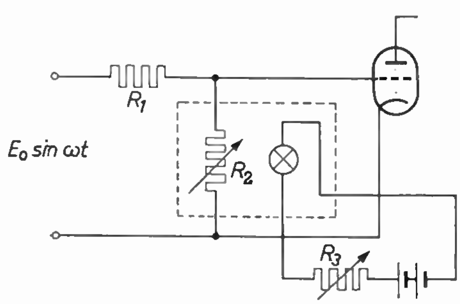


Fig. 3.

resistor R_3 can be mounted at a place for most convenient handling. More detailed information on a manual volume control of this type can be found in the literature.[1]

The CdS Photo-Resistor in an Amplitude-Limiting Device

We come now to another application of this CdS photo-resistor, which application will be the main subject of this paper, *viz.* an amplitude-limiting device. Some of the properties of the CdS photo-resistor make an element of this kind very suitable for preventing overload of an audio amplifier. The characteristics of the device to be described are comparable to those of arrangements usually referred to as peak limiters. There might be a difference of opinion about some of the required properties of a limiter, but I believe that the following characteristics will always be considered as advantageous.

1) The device shall have no influence whatsoever below a certain predetermined output level.
2) Above this threshold, it shall decrease the output very rapidly below the overload level.
3) When the signal drops below the value which necessitates the functioning of the device, it shall slowly become inoperative.

The demands 2 and 3 can be described as a short attack time and a long decay time, respectively. A photosensitive resistor like a CdS cell used in the properly chosen resistance range has by its own nature these properties 2 and 3, as will be clear from the account given of the phenomenon of photo-conductivity.

We can make use of these properties in the following way: Control of the output of an amplifier can be effected by the volume control at the input, *e.g.*, in Fig. 3 by changing the value of R_2. And, if this resistance is a CdS cell, a lowering of the output level may be obtained by illuminating this cell. In order to get a limiting action, we have to lay a connection between the output-power or voltage and the illumination of this cell. It was found that this link can be realized in a very simple way by using a neon lamp, ignited by the output voltage.

Quasi-Static Conditions

As is well known, a neon lamp has a fixed, very constant ignition voltage and an extinguishing voltage which has a slightly lower value. If the photo-electric CdS cell and the (small) neon lamp are built together in a lightproof enclosure, the CdS cell being connected as R_2 in Fig. 4, the full input voltage will be applied to the first grid of the amplifier as long as the neon lamp is not ignited, *i.e.*, as long as the output voltage does not exceed a certain predetermined value. As soon as this

[1] N. A. de Gier, W. v. Gool, and J. G. v. Santen, "Photoresistors made of compressed and sintered cadmium sulphide," *Philips Tech. Rev.*, vol. 20, no. 10, pp. 277–308; July, 1959.

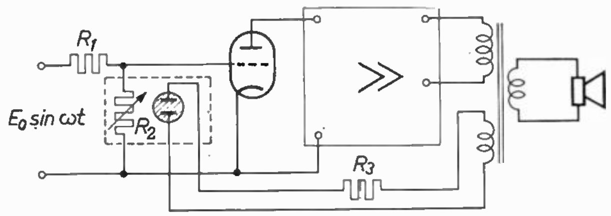


Fig. 4.

value is reached and the lamp ignites, the resistance of the CdS cell drops. As a consequence, the input voltage decreases to a value which corresponds to an output voltage just a little above the ignition voltage such that the light will just keep burning. If the initial input voltage increases still further, the output voltage rises only very little. In other words, a rigid limit is set to the output voltage, as is illustrated in Fig. 5.

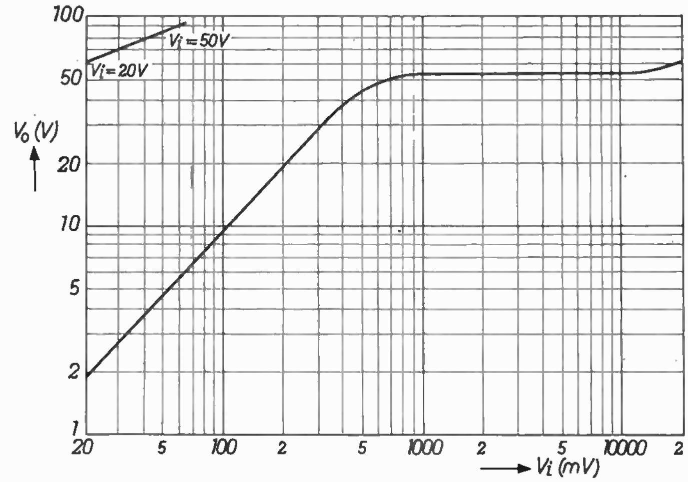


Fig. 5.

Rapid Changing Conditions

The decrease of the input voltage occurs very quickly because of the almost immediate ignition of the neon lamp and the rapidly following decrease of resistance of the CdS cell. The return to the original setting (when the input voltage to the complete system is decreased) goes relatively slowly, for reasons explained earlier. The behavior of this limiting device under "rapid changing" conditions, such as occur during a music program in which rapid peaks are followed by weaker passages, is almost completely determined by the properties of the CdS cell and not, as is the case with other limiting devices, by RC elements. If a sudden peak reaches the output terminals, the control action takes place in a couple of milliseconds, before the ear can detect any distortion. The amplifier continues to work with a turned-down volume control, as it were. If no further peaks follow, the regulation becomes noneffective, the volume is "turned back," relatively slowly, to its original setting. Therefore, this device operates as an almost ideal limiter: no influence whatsoever before overload starts to take place. Rapid action is taken analog to turning down a volume control when overload starts. When overload conditions no longer exist, this action is slowly returned to normal.

Contributors

Donald E. Brinkerhoff (A'54–M'60) was born on December 6, 1921 in Bryant, Ind. He graduated from Purdue University, Lafayette, Ind., in 1943 with a B.S. degree in electrical engineering.

D. E. Brinkerhoff

During World War II, he served as an officer with the U. S. Army Signal Corps. After completing the Army Officers Electronics School at Harvard University and the Massachusetts Institute of Technology Radar School, both in Cambridge, Mass., he was appointed an instructor in electrical communication engineering at the M.I.T. Radar School. He also served as a member of the Army Ground Force Board II, where he was responsible for new radar equipment evaluation. He joined Delco Radio in 1945, as a specifications engineer, and in 1952 he assumed his present position of engineer-in-charge of the Acoustical Engineering Laboratory at Delco Radio Division, General Motors Corporation, Kokomo, Indiana. His present responsibilities include loud-speaker and associated audio-acoustical system design and development.

He has taken graduate work in math and physics under the Purdue University off-campus Graduate Program at Delco Radio.

Mr. Brinkerhoff is a member of the Acoustical Society of America, the Audio Engineering Society, and the Kokomo Engineering Society. He is chairman of the subcommittee on Automotive Acoustics under the Committee on Electro-Acoustics and Audio Devices in the Acoustical Society of America.

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Rudolf G. de Buda was born in Vienna Austria, on January 21, 1924. He attended the University of Vienna, where he received the degree of Dipl. Ing. in electrical engineering (communications) and the Ph.D. degree in mathematics in 1948 and 1949, respectively.

R. G. de Buda

He went to Canada in 1951, and joined the Canadian General Electric Company, Toronto, first as power transformer design engineer (two patents), then in the Electronic Equipment and Tube Department, where one of his assignments is in development of high-power broadcast circuits.

Dr. de Buda is a member of the Association of Professional Engineers of Ontario.

Edward F. Feldman (S'52–A'53–M'56–SM'58) was born in New York, N. Y. on August 11, 1928. He received the Bachelor of Electrical Engineering degree from the Cooper Union School of Engineering, New York, in 1949. In 1956, he received the degree of Master of Electrical Engineering from the Polytechnic Institute of Brooklyn, Brooklyn, N. Y. Since then he has been attending New York University Graduate School of Engineering on a part-time basis, pursuing post-graduate electrical engineering studies.

E. F. Feldman

In 1949, Mr. Feldman joined the New York City Board of Transportation, where he was engaged in a modernization program of railway signal circuitry. He has been on the engineering staff of Panoramic Radio Products, Inc., of Mount Vernon. N. Y., since 1951. He has been active in the development of automatic test instrumentation such as spectrum analyzers, sweep frequency generators, telemetry calibration equipment, and several special-purpose military systems, and has pending several patents.

Mr. Feldman is a member of Tau Beta Pi, Sigma Xi and Eta Kappa Nu.

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

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J. Rodrigues de Miranda (SM'58) was born on December 24, 1905, in Amsterdam, Netherlands. After graduation from the Technical High School at Delft, he joined a wholesale radio components firm in Amsterdam in 1927 as manager of the technical department. In 1938, he joined Philips' Industries in Eindhoven, where he worked in the electroacoustic department and the patent office before beginning his current association with the Radio Apparatus Development Laboratory. He has been in charge of the Laboratory since 1959.

J. R. de Miranda

Mr. Rodrigues de Miranda is a member of the Radio Genootschap, the Koninklijk Instituut van Ingenieurs, the Geluidsstichting, and the Audio Engineering Society.

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Bela Ranky (SM'54) was born in Budapest, Hungary, on January 18, 1910. He received the M.S. degree in electrical and mechanical engineering from the University of Technical Sciences, Budapest, in 1932, and attended the Technical University of Berlin, Charlottenburg, Germany, in 1933 on a one-year scholarship.

B. Ranky

He joined the Hungarian Post Telegraph and Telephone Administration in 1934, and became chief engineer of Radio Budapest Broadcasting in 1941. After World War II, he was engaged in broadcast studio design especially related to magnetic tape recorders in France. He came to the United States in 1953, and, since the following year, has been employed at Panoramic Radio Products, Inc., Mount Vernon, N. Y., as a design engineer. Major areas of activity have included development of sweep generators, telemetry test instrumentation, and special products.

Mr. Ranky holds several European patents and one U. S. patent, and has several U. S. patents pending.

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Ronald J. Rockwell (A'25–M'31–SM'43–F'45) was born in Omaha, Neb., on February 4, 1904. He received the B.S.E.E. degree in 1927, and his Professional E.E. degree in 1946, both from the Iowa State College, Ames.

R. J. Rockwell

In 1920, Mr. Rockwell designed, built, and operated the first broadcast station in the Western part of the country, WNAL, Omaha. From 1927 to 1928, he was in charge of the condenser microphone laboratory of the General Electric Company in Schenectady, N. Y., where he developed "camera model" condenser microphones used by NBC studios for a number of years. In 1928, he entered into his own consulting practice, designing and installing KTHS, Hot Springs, Ark., and KLRA, Little Rock, Ark. In 1929, he went to Cincinnati, Ohio, as engineer in charge of audio circuit and speaker development for the Crosley Corporation. In 1936, he was transferred to the Broadcast Branch of the organization and was made Director of Engineering; in this capacity, he had charge of all technical activities of the Crosley stations, which then included WLW, WSAI, and W8XAL. When the Broadcasting Branch of the Crosley Corporation was separately incorporated as the Crosley Broadcasting Corporation in 1946, he was made a Vice President of the new corporation and was placed in charge of its engineering department, which includes responsibility for five maximum-power television stations, WLW-T, WLW-D, WLW-C, WLW-I, and WLW-A, 50,000-watt radio station WLW, and six 200,000-watt transmitters for the Voice of America.

Mr. Rockwell has been a Registered Professional Engineer in the state of Ohio since 1944, and throughout his career has served on numerous government and industry-sponsored boards and committees. He is also a Fellow of the AIEE and a member of the SMPTE, the Acoustical Society of America, Eta Kappa Nu, and Tau Beta Pi. He holds numerous patents in the field of electronic circuitry.

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INSTITUTIONAL LISTINGS

The IRE Professional Group on Audio is grateful for the assistance given by the firms listed below, and invites application for Institutional Listing from other firms interested in Audio Technology.

Charge for listing in six consecutive issues of the TRANSACTIONS—$75.00. Application for listing may be made to the Technical Secretary, Institute of Radio Engineers, Inc., 1 East 79th Street, New York 21, N.Y.