the 2nd audio anthology

the 3rd audio anthology

the audio anthology

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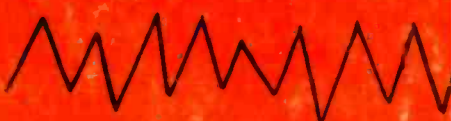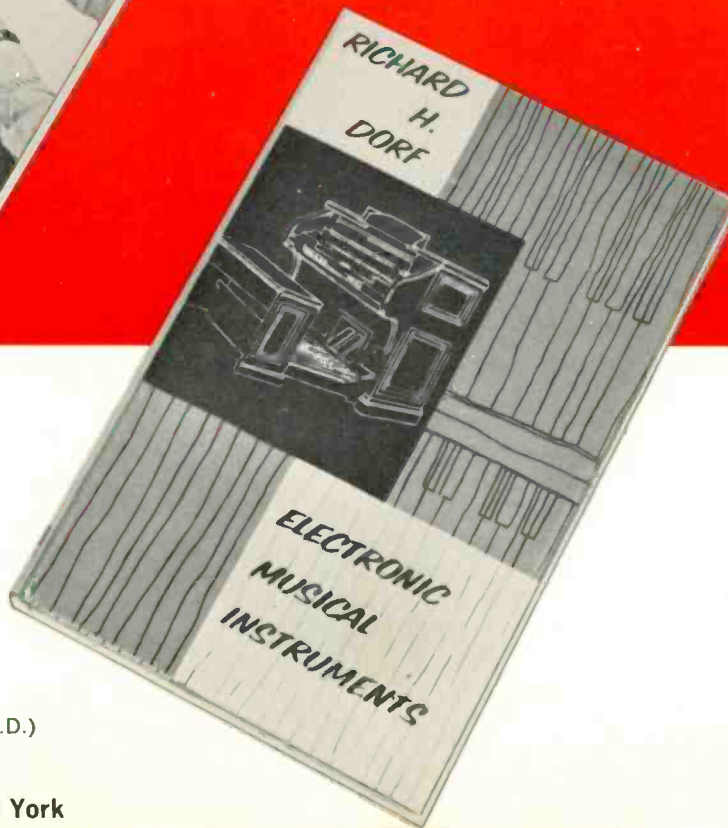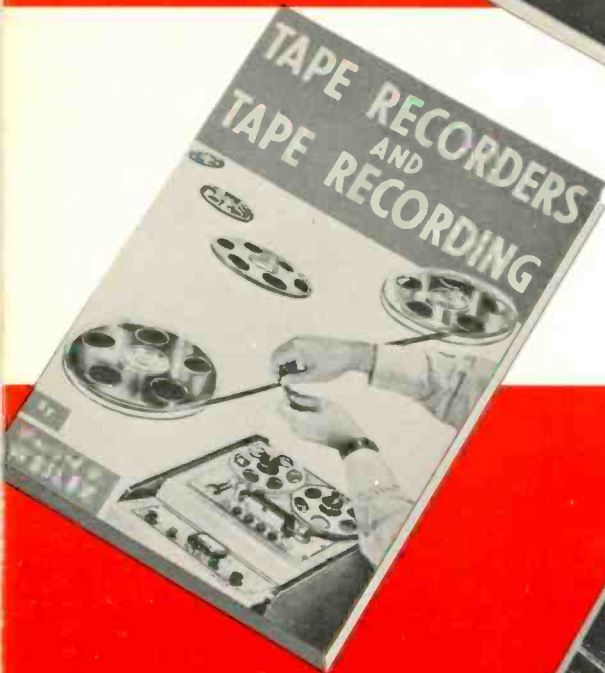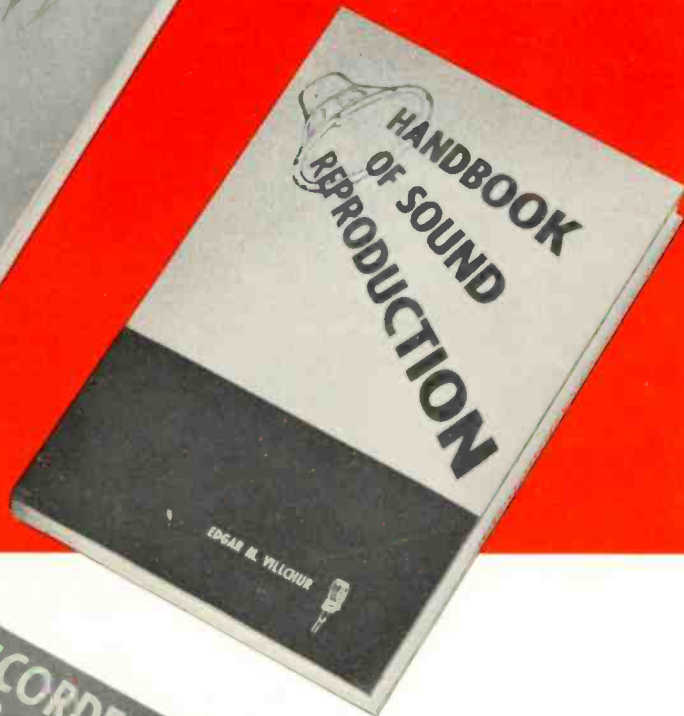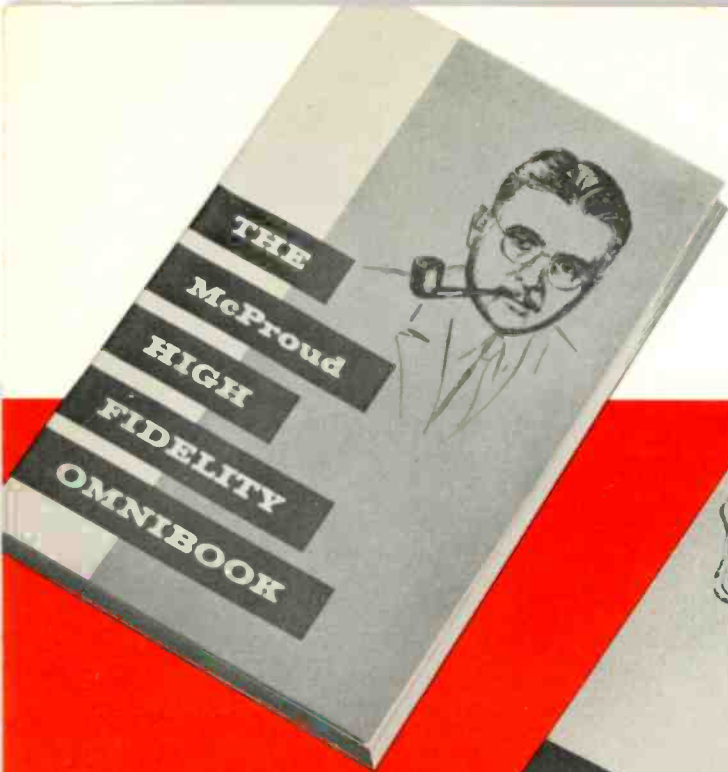
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the fourth audio anthology

COMPILED FROM

AUDIO

the *original* magazine about high fidelity.

From July 1955 to December 1957
by **C. G. McPROUD** Editor

RADIO MAGAZINES, INC.

MINEOLA, NEW YORK · 1958

CONTENTS

Foreword

For over eleven years AUDIO magazine and its predecessor, AUDIO ENGINEERING, have been considered the authority on all things pertaining to high fidelity — amplifiers, pre-amps, phono systems, tape recording and playback amplifiers, loudspeaker systems and enclosures, and so on — and many of the ideas and innovations first presented to readers in the pages of the magazine have become so well accepted by the serious audio experimenter to the extent that they have been incorporated in commercially built equipment. For example, to name a few, the "Standard" speaker system — originally presented in 1949, has been the pattern for dozens of factory-built models using the radiation from both sides of the cone in a room-corner housing; the loudness control — now *de rigueur* in all good amplifiers — first appeared in Æ in 1948; the Baxendall tone control was popularized in this country by several articles about it in the magazine, and now all the components are available to manufacturers in printed-circuit form for economical construction of the finished product.

The large majority of AUDIO's readers are hi-fi enthusiasts. Many of them wait from month to month for something new to construct — some apparently build every amplifier or preamp described. The three previous Audio Anthologies have each brought in a single volume most of the basic hi-fi material for the preceding 30-month period, and this Fourth Audio Anthology carries on this tradition.

The continued interest of our readers makes it possible — even pleasant — to strive to bring them the latest and most reliable of hi-fi information. And their continued clamor for this additional compilation of hi-fi articles has been our inspiration in its preparation. To all of these readers we respectfully dedicate this Fourth Audio Anthology.

C. G. McPROUD, *Editor*
AUDIO

Mineola, N. Y.
August, 1958

How to Plan Your Hi-Fi System

C. R. TIEMAN

The author offers a method for making comparative listening tests in order to evaluate performance of audio equipment on a quantitative basis as a logical means for choosing components.

WHEN ONE'S INTERESTS turn toward the acquisition of a high fidelity audio system, he is introduced to a comparatively new world of elusive values, and he most likely will find that to lay a sound plan for either constructing or assembling his system is a very interesting but involved task.

Aside from the barrage of claims and counter-claims of the equipment manufacturers, advice from all quarters is likely to appear to be in serious conflict. The beginner may be unable to plan effectively because the values of which so many speak glibly are subjective in nature and depend upon personal tastes and interests. Naturally, these different interests as expressed by different advisors can be mutually conflicting, and he has not been able to establish his own standards or recognize his own particular needs. The measures of system effectiveness are related to personal tastes.

The purpose of a plan is to provide the system which most nearly satisfies the listener's interests at a minimum cost. We all have heard of the fellow who, after spending much time, effort, and money to acquire a suitable system somehow fails to be satisfied; and after a while, has actually acquired enough equipment to assemble several systems. If your aim is not to "tinker," then some time spent in planning a system will pay off handsomely in the long run.

Regardless, whether we wish to design circuits or assemble a system from completed components, we first must establish some planning objectives or goals toward which to work. The second step is to choose individual components which will satisfy these objectives at a minimum of cost. The first and most important step is to set one's sights: if you aim too high, the budget suffers directly; and if you aim too low, the results will ultimately be unsatisfactory. Within the confines of the space here, we will spend most attention on the first step in planning, that of getting the objectives or goals outlined to satisfy personal tastes.

High fidelity means different things to different people, and a "good" system for one person may be a "bad" system for another. We may all have seen at least three kinds of enthusiasts which

Fig. 1. The Fletcher-Munson curves of equal loudness at the listener's ear.
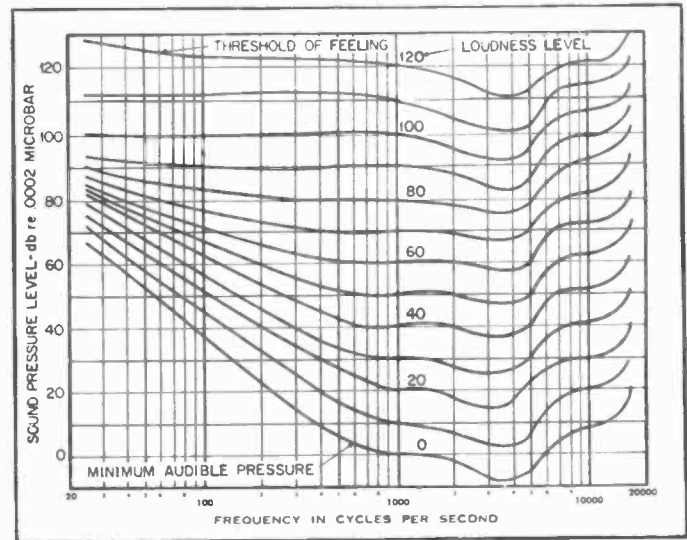
could be grouped about as follows: the "sound engineer," the "music critic," and the "interested listener." The "sound engineer" is the fellow who, above all, needs a variety of gadgets so that he can exercise complete control over the signal and compensate for any situation. In addition, he may require a multiplicity of inputs to his amplifier to give him flexibility in changing from tuners to microphones, or to any one of several recorders. He enjoys the thrill of being able to shape or modify the musical output to taste. Individuals in this category tend to emphasize the importance of the amplifier, preamplifier, and the auxiliary circuits; but so far as listening is concerned, frequently their needs are satisfied with an 8-inch speaker of moderate quality.

The second fellow is the perfectionist who demands the ultimate in performance and scrutinizes each individual unit to make sure that it is the best available within the state of the art. These people emphasize the importance of hearing every last note and overtone the music offers. The are impatient with the slightest noticeable distortion, and ask for accurate compensation for both the recording and the ear; some are concerned with the effects of the temperature and humidity of the room in which the music is played. Occasionally a few perfectionists tend to join a cult of "fanatics" who demand improvements beyond what the ear is able to hear. Another subclass of the "music critic" is the person who

feels that the symphony concert must be duplicated in the home in both tonal range and volume. He requires the 30- to 80-watt system, while the average single speaker home system will be well served with a 10- to 20-watt system.

The "interested listener" is the one who aspires to have a musical system of a quality that is much better than is afforded by the average commercial radio and TV equipment, but he borrows from the "sound engineer" and the "music critic" for what they can contribute to his listening pleasure. He wishes to use his system to satisfy his personal needs which may range from a dance party or background music, to occasional serious attention to musical masterpieces.

Importance of Individual Taste

Because high fidelity means different things to different people, the first thing that a hi-fi sales agent will want to discover when you visit his shop, is what your individual tastes are, and what you expect to do with the proposed system. Naturally, he will get to the point regarding the size of your proposed budget, for he can cite systems that can be assembled for less than $100, as well as those which will cost over $1,000.

To the salesman, one may quite innocently state that he would just like to have a simple system that accurately reproduces what the microphone picked up in the first place. To this remark, an experienced hi-fi "expert" may present a

disquisition on why this is very difficult, if not impossible. In addition, one expert proclaimed that some persons do not want to hear music as it originally sounded, but they prefer the modifications that are afforded by the electronic system. The guiding principle, they say, is: "let your ear be the judge," and if you get the impression that the original orchestra is present as the recording is played, then the system is "good"—this is the feeling of "presence." Some experts further assert that accurate quantitative analysis of sound reproduction for the listener is futile, because the listening pleasure derived from the sound system is largely subjective, and is purely a matter of personal taste.

Standard practice for evaluating system performance is to make a series of listening tests to ascertain the differences in the ways in which the same recording can sound from different systems. If the local hi-fi shop does not have some arrangement whereby different systems may be compared, then the beginner would be faced with some rather difficult choices, for it is virtually impossible in the final analysis to evaluate listening performance from advertising literature.

After a few listening tests, one can readily appreciate the advantages as well as the limitations of the subjective method of measuring listening pleasure or system performance. Qualities seem to be present in some systems that are not adequately described by the specifications. I found it virtually impossible at the outset to make any judgments on the basis of a few isolated demonstrations and without some coaching regarding what features were good or bad. The untrained ear can overlook a variety of desirable as well as undesirable features in a system. Without some experience, one can readily become confused when evaluating system performance by playing recorded music because one is exposed to a myriad of sounds in rapid succession that cover a wide range of frequencies and volumes, and a wide range of waveforms. Certainly, peculiarities or idiosyncrasies of the speaker, enclosure, amplifier, or turntable could be missed by the beginner; and poor performance could be judged as adequate. Experience and some instruction are, indeed, needed.

Because we rely on the ear to such a great extent, we must consider the ear as much a part of the entire system as the amplifier or the speaker. It is the ear that is either sensitive or insensitive to certain tones, or levels of volume. There is no point in paying attention to sounds that only the dog or the canary can hear, or that can be detected on an oscilloscope. The hi-fi equipment transforms the signals derived from a tuner or a recording into acoustic waves which excite the listener's ear. Hence, the first step in understanding one's needs and planning objectives is to learn something about what the ear actually hears.

The hearing characteristic of the average listener is shown in Fig. 1. These curves are called the Fletcher-Munson curves of equal loudness, and they illustrate how the ear—the physical termination of the hi-fi system—acts for various sound levels and various frequencies. Note particularly that as the loudness level of sound is decreased, the sensitivity of the ear at low frequencies changes with respect to the mid-frequency range: a sound at 0 db at 1000 cps is just as loud to the ear as a sound at 30 cps having an intensity 60 db greater. This is a factor of 1,000,000 in terms of acoustic power. However, the same two frequencies at the 100 db level both have the same loudness insofar as the ear is concerned.

This characteristic of hearing plays a very important part in the hi-fi system, for it means that at a reduced level of sound a given musical selection will appear to lose its low tones; conversely, boosting the volume will increase the apparent intensity of the lower tones.

For high fidelity reproduction we wish to maintain a balance between the highs and lows so that the sound resembles the original production to get the feeling of "presence." But the original, such as a symphony orchestra, may in the concert hall be at a level of 80 to 90 db. In the home, the level is normally reduced to a range from 60 to 70 db; hence, the home system must accentuate the lows by a substantial amount—almost 20 db for a 35-cps note—to compensate for the ear's characteristics.

Accentuating this loss of low tones with reduced volume is the reduction in performance of many components at low frequencies, and the difficulty of radiating such waves from speakers and their enclosures. Some speakers, for instance, may have a characteristic drop in power radiated below 100 cps, so that by the time the signal gets to 35 cps the radiated power is off by 10 to 20 db.

The rule "let your ear be the judge" is a fundamental one, but one that should be exercised with considerable caution and understanding of the way the ear reacts.

The basic need for control over the level of volume, and the amplification of the low frequencies in relation to the highs is certainly established experimentally if you conduct several listening tests. Adequate control over "selective" amplification is ordinarily provided by the "bass" and "treble" controls of the preamplifier or the first few stages of amplification.

Value of Listening Tests

The principal value of a series of listening tests lies in simultaneous comparison of different systems against each other. The facilities of several hi-fi shops are such as to allow one to synthesize a wide variety of systems simply by throwing a few switches. With such a facility, one may listen to systems ranging from the least to the most expensive.

For the purpose of planning, we recommend that one should approach the first series of listening tests in such a way as to establish his own preferences rather than attempt to make a selection of equipment. This author believes that component selection should be deferred until one has firmly in mind the standards of performance he feels are worth the cost in time and money. Only when one has an idea of the performance he seeks can he assemble a system in which the components make their full contribution and are still held to the minimum cost.

As a practical matter one is usually well acquainted with what might be called the "lower level" of performance because of familiarity with TV and radio. Based on this starting point, the tests described here were initiated to determine the best system performance available, and how various high fidelity systems compared with the best. Wide ranges of price and performance were found, and technique was sought to place them in a suitable line of succession for comparison. Assembled amplifiers ranging from $100 to $150 gave very high quality performance; speakers and their enclosures ranged from less than $40 to over $700, and the tonal range and generally pleasing quality of the sound varied widely. Turntables, on the other hand were almost universally of one make and model in the places visited, so there was no particular choice avilable in this item.

The techniques used to determine personal preference commensurate with the pocketbook was to make use of the several amplifiers recommended as best by the salesman, and then substitute speakers and enclosures to make up successive systems, because the speakers displayed the widest range of price and performance. We are, in effect, attempting to match the hi-fi equipment to the ear at best we can within budget restrictions.

Stressing the need for listening tests, and the inability of technical specifications to convey a measure of listening performance, one consultant asserted, "Now, in these multiple speaker units you will notice more depth of tone, a quality which we cannot adequately describe by our instruments, but which is, nevertheless present. That depth is a psychological effect caused probably by the sound coming from an area rather than from a point source."

Although not always the case, it is usually true that the systems with higher price tags gave a higher level of per-

formance. How much quality is gained by an additional outlay of money is important for the planner to have in mind before becoming obsessed with any one particular system, or with the struggle to achieve the ultimate.

After reviewing several synthetic systems, one begins to realize that high-quality audio reproduction can be achieved. In fact, some of the more advanced systems will give a performance that is virtually indistinguishable from the original. The second significant point is that the cost of a high-quality system is likely to be a little more than had originally been estimated. The choice of components that will lie ahead will be rather delicate because the mistakes can be costly.

Although the actual measures of performance of a hi-fi system are subjective, the person who has gained a limited amount of listening skill should be in a position to make comparative tests and to place his subjective reactions on a quantitative basis. If such crude measures can be made, then one can rate different systems. With these ratings together with the cost data, one can then construct the "cost-effectiveness" curve which can be an invaluable aid to system planning. In the beginning we recommend that the rater confine his attention to the general impression of "listening pleasure" or "presence" and take up more detailed refinements at a later time. Initial tests of this kind are sufficient to convice some "interested listeners" that the systems of moderate quality are adequate for the purpose, but others will lay aside all thoughts of intermediate steps and plan for the highest standards of performance that are possible.
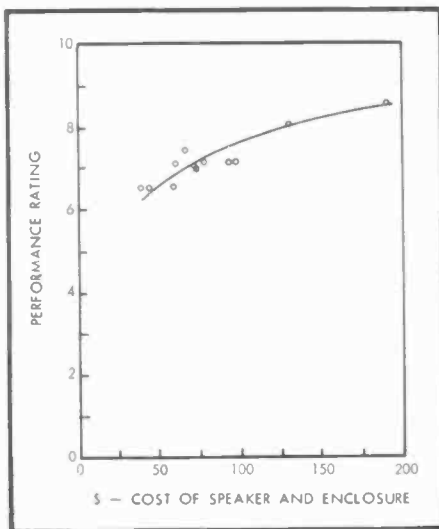

Fig. 2. Curve of cost vs. performance as constructed from the data in Table I, covering the test of the least costly systems.

Typical Test Results

The results of two tests are tabulated in Tables I and II, and are plotted in

TABLE I
COMPARISON OF SPEAKERS AND AMPLIFIERS — TEST No. 1

ENCLOSURE AND SPEAKER				TOTAL	PERFORMANCE RATINGS			
Type	Cost	Spkr No.	Cost	COST	L.F.	M.F.	H.F.	Overall
A	18.00	1	41.60	59.60	6	8	8	7
A	18.00	2	45.00	63.00	6	8	8	7.5
A	18.00	3	54.50	72.50	5	8	8	7
A	18.00	4	76.50	94.50	6	8	8	7.2
A	18.00	5	57.60	75.50	6	8	8	7.2
A	18.00	6	54.00	72.00	6	8	8	7
A	18.00	7	40.00	58.00	5	8	7	6.5
A	18.00	8	20.50	38.50	5	8	7	6.5
A	18.00	9	27.00	45.00	5	8	7	6.5
A	18.00	10	25.00	43.00	5	8	7	6.5
A	18.00	11	79.50	97.50	6	8	8	7
A	18.00	12	114.00	132.00	7	9	9	8
K	50.00	13	74.50	124.50	8	9	8	8.8

Williamson-type amplifier ($190.00), record changer ($68.00), and magnetic pickup not changed during test

Speakers tested included single-cone extended-range models, and two- and three-way integral models.

Enclosure types: A is bass reflex in kit form, unfinished plywood K is Karlson-type, in kit form.

graphical for in *Figs.* 2 and 3. These curves show a distribution of points up and down the scale of performance or effectiveness, plotted against the costs of the respective speakers and enclosures. The performance generally rises with cost. The total system costs could be de-

TABLE II
COMPARISON OF SPEAKERS AND ENCLOSURES — TEST No. 2

ENCLOSURE AND SPEAKER	TOTAL	PERFORMANCE RATINGS			
Type (Finished Unit)	COST	L.F.	M.F.	H.F.	Overall
(a) With amplifier and preamp at $190.00. Speakers include two- and three-way types					
BACK-LOADED HORN	244.50	7	9	9	8.5
BACK-LOADED HORN	276.00	7	9	9	8.5
CORNER HORN	300.00	10	9	9	9.3
CORNER HORN	720.00	10	10	10	10
INFINITE BAFFLE (Dual Spkrs)	100.00	8	9	9	8.3
INFINITE BAFFLE (Two-way)	126.00	5	8	9	7
(b) With amplifier and preamp at $203.00.					
INFINITE BAFFLE (5 Spkrs)	256.00	7	9	10	8
INFINITE BAFFLE (9 Spkrs)	450.00	9	10	10	9.5

Same changer ($68.00) and magnetic pickup used throughout.

rived by adding the amplifier and changer costs in each case. The most expensive system was $978, while the least costly was $296, but the lower figure could have been reduced by using less expensive amplifiers without affecting acoustical performance.

The distribution of points shown is peculiar to only one rater. If some other rater was to evaluate the same systems, he would probably place the points in different positions, but he would most likely get curves of about the same shape after averaging the points. In conferring with other listeners, however, the most expensive system was consistently rated as the top in quality. This unit was arbitrarily scored as 10, thereby establishing a standard at the top of the scale. At the lower level, the table model radio and a radio phonograph were scored at about 3 and 3.5 respectively, but these scores are not shown in the data or curves. These latter ratings set a lower limit or standard of "low-cost" perform-

ance. The region of high-fidelity begins with an over-all rating of about 6. The over-all rating was derived from three separate components, one to evaluate low-frequency response and freedom from distortion, one for medium frequencies, and the third for the highs.

The two curves show the extremes available for home listening. The combinations between these extremes are many, and the measures that can be taken to reduce the costs of the systems of highest quality are not exhausted in these two initial tests. For instance, one would like to see whether a system could be assembled having a performance between 9 and 10, but at a substantially lower cost than those listed in the tables. By using the "do-it-yourself" kits for the more complex systems, if they are available, such costs can be reduced; but one may be forced to forego the styling and fine finish of the speaker enclosure and the auxiliary cabinets. Cost reduction by a factor of two is a reasonable expectation.

At this point, one may have enough information to stir his enthusiasm, but not enough to be sure of his precise needs so that the system can be laid out. There is still another step the beginner may take before making any major decisions. This step is not essential for one who has made up his mind but is recommended for those who wish to develop better judgment. This step is to construct an "interim system" which will permit the experimenter to observe over a period of time the factors of importance to him which affect the performance of the system. With a judicious choice of components, one could apply any items purchased for the secondary system to the more advanced one. This experimental approach has the advantage of allowing one to study his re-
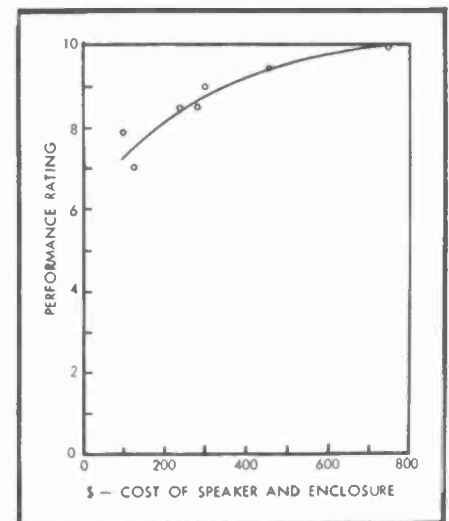
Fig. 3. Curve of cost vs. performance constructed from the data in Table II, covering the test of the most costly systems.

System Simplicity in Audio

R. G. CHAPLICK

The trend toward duplication—or even triplication—of controls in a high fidelity system is looked upon by this author as unnecessary, useless, and inconvenient. Furthermore, he tells you what to do about it.

THE MAJOR AIM of this article is to advocate "system simplicity" and to show that it can be employed to improve an audio system. Components recently featured in audio magazines show a trend towards increasing complexity. Although these gadgets are impressive looking, they require great skill to manipulate the myriads of knobs, My opinion is that many systems have overgrown and that the average audio enthusiast can achieve superior results by system simplification. Before system simplicity can be discussed intelligently, the object of an audio system should be known. My definition of a good music system is one that reproduces sound realistically—neither adding nor taking away, "Holding, as it were, a mirror up to nature." There must be no overemphasis of high or low frequencies. Concert-hall realism is not achieved by shaking windows with low frequencies or by hurting ears with high frequencies. I must add to the discussion of a good sound reproducing system, the plea that the listener attend as many live concerts as possible. Reproduction of sound may be pleasing but can never be more "real" than the actual performance. Moreover, the listener should have more opportunities to "keep his ear calibrated" by comparing the output of his system with "the real thing."

The current trend in the purchase of audio equipment is the selection of individual components by the audiofan who then assembles them into a system.
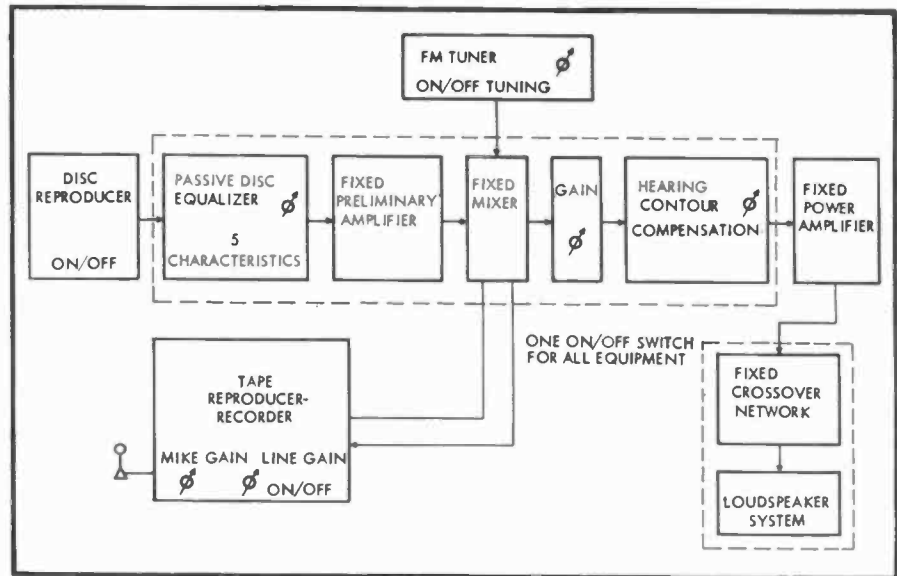

Fig. 2. Block diagram of ideal arrangement for an audio system.

Figure 1 is a block diagram of a system produced in this manner. This procedure has many disadvantages. First, there is no assurance that the components are electrically compatible. Second, there are too many controls which nullify or duplicate each other, Third, there are just too many controls. My conclusion after four years of buying or building components is that he who wishes to assemble a good system should obtain the guidance of a professional electro-acoustic engineer, one who can criticize objectively, who can test a system thoroughly, and who is not interested in selling any particular brand of equipment. The latter

requirement is the most important. Then, a new system can be engineered to meet any needed specification, or an existing system can be simplified and improved. In either case the consultant will be able to prevent many errors and to provide system engineering.

The ideal simplified system should be defined before I tell how my own system was simplified and improved. The general objective is to reproduce sound realistically at any level. The frequency response will cover the audible range, the power output will be adequate for the needs of the listening room, and the distortion will be inaudible. Controls and adjustments will be kept to a minimum, and where equalization is needed, fixed passive networks will be employed. From these generalities a block diagram (*Fig.* 2) of the ideal system can be made.

With the advice of the consultant, I devised the audio system described in the following paragraphs. Component circuit diagrams are omitted since conventional circuits are used. The number of controls was reduced to a minimum. Broadcast control room techniques were employed, and each unit was equipped with its own power supply and fuse. Individual parts of a unit were oriented for minimum hum. Standard telephonic techniques were employed in designing the wiring of the units. Source output
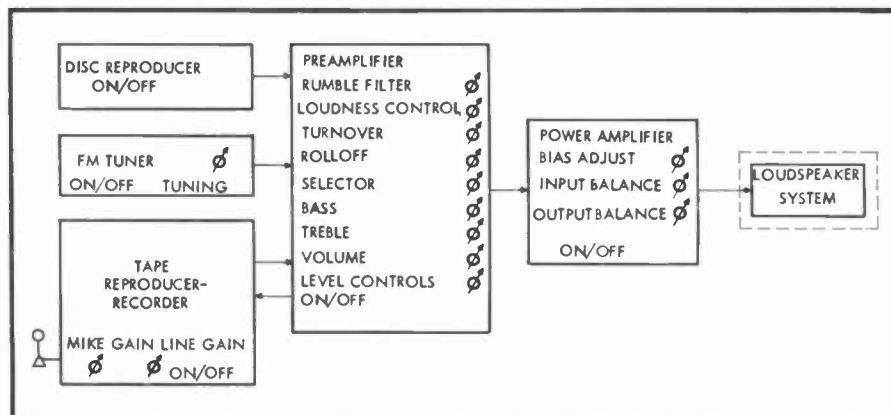

Fig. 1. Block diagram of system before simplification.

voltages were set at fixed valves to meet input requirements of the Control Unit. The stepped Hearing Contour Compensator reduces the level by an amount shown on the compensator setting. The single gain control is therefore usable over its entire rotation, not just the first ten or twenty degrees.

An audio system is divided into three parts: sources, sinks, and controls. In turn, each of the three parts has subdivisions whose specifications and method of simplification now can be discussed individually.

Sources

Disc Reproducer. Disc reproduction is designated exclusively for 33⅓ rpm LP discs in manual operation. Since all of of the program material which interests me is available on LP discs, only a single turntable speed is needed. I frequently wonder why turntable manufacturers do not produce a good single speed, 33⅓-rpm turntable. Elimination of the unwanted speeds, pulleys, and idlers is a form of simplification. A good quality magnetic cartridge completes the disc reproducer.

FM-tuner. The tuner has a.f.c. and a tuning indicator. Only two controls are necessary—the on-off switch and the tuning control. The output voltage was set at an average peak level of 0.5 v. rms. Equalized high-quality headphones can be plugged in for night listening.

Tape Reproducer-Recorder. This unit has NARTB equalization. Two gain controls are used in the "record" mode of operation: One in the "microphone" channel, one in the "line" channel. A VU meter is used. There is no gain control in "playback." Provision for both tape and input monitoring by headphone or loudspeaker is made. An Ampex 600 meets these requirements.

Sinks

Power Amplifier. The frequency response of the basic amplifier is flat within ± 1 db from 20 to 20,000 cps. When the adult human ear is shown to hear beyond 20,000 cps, then I'll start worrying about extending the frequency response. Fixed compensation for loudspeaker characteristics have been added. The amplifier is designed to furnish one-half the 25-watt maximum power at 0.6 v. rms input, allowing three decibels of reserve power for possible overswing. The internal generator impedance has been adjusted by test for optimum damping of the associated loudspeaker system. A pulse from an RC circuit was used in this test, and the amount of inverse feedback was varied until best damping was secured.

Loudspeaker System. A warble oscillator and a calibrated microphone were used to set the proper balance between high- and low-frequency loudspeakers and to set the equalization for loudspeaker and room acoustics. This equalization is permanent since room acoustics change very little from day to day.

Controls

A commercial preamplifier was reworked into a "Control Unit." Two of the sources listed above were mixed by means of a resistive network into the preliminary amplifier at the proper levels according to the output voltage of each. Because each source has its own power supply, its output signal can be removed by simply cutting off the power. Thus the selector switch could be eliminated. However, a switch was provided for tape playback to eliminate a feedback loop during recording. Tape recording may be monitored by headphones or by loudspeaker.

Preliminary amplifiers with separate high- and low-frequency controls for disc equalization have a common fault. Although the low-frequency control is intended to effect only the low frequencies it also effects the highs, and vice versa in the case of the high-frequency control. Therefore, unless accurate calibration has been made of all possible combinations of those controls, front panel markings are far from accurate. Disc equalization has been simplified to a one-knob control. Five equalization curves based on published curves of representative disc manufacturers are sufficient for all discs (old records being taboo). Circuits were designed to switch resistors rather than capacitors to minimize the effect of switching transients. The RC values were calculated and then corrected by frequency measurements to yield the proper equalization. The "Rumble Filter" was removed entirely since a transcription turntable is used and unreasonable "bass boost" is avoided.

The main feature of the control unit is the replacement of conventional "bass" and "treble" controls by a "Hearing-Contour Compensator." The Hearing-Contour Compensator improves the realism of high-quality sound reproduction by compensating for the difference in level between the music produced in the concert hall and that reproduced at a necessarily lower level in the living room. It compensates for the variations in human hearing sensitivity to sounds of different loudness. The variations in hearing have been measured and have been found quite uniform for persons of normal hearing. They are shown in the Fletcher-Munson Curves of equal loudness. The principle of the Hearing-Contour Compensator operation is based on a study of the differences between Fletcher-Munson Curves, rather than on contour at any one acoustic level. Averages have been selected based on a series of subjective tests which included listening alter-

nately to original sounds and then to the same sounds recorded and reproduced. Most of this testing was made with the orchestra of the Metropolitan Opera in New York and the U. S. Navy Band in Washington, D. C. The Hearing Contour Compensator performs in fixed calibrated steps of 0, 10, 20, 30 and 40 db. These figures indicate the difference in db between original and reproduced program levels. Appropriate attenuation is designed into the compensator. In case speech is reproduced, music equalization is completely wrong, since speech should be reproduced at about the same level as it was originally produced. For speech reproduction a switch is provided which retains the attenuation but removes the compensation.

It is important to digress at this point in order to speak again of the value of frequent listening to good orchestras under good concert hall conditions. Many "hifi" fans will find this an illuminating experience; a few may be discouraged and some will staunchly maintain that the orchestral sounds are much inferior to those reproduced by their "hifi" systems. This will prove the revised adage "be they ever so homely there are no ears like your own."

The number of controls in the system was reduced from twenty-one to ten.

CONTROLS

Before Simplification	After Simplification
FM tuning control	FM tuning control
Rumble filter	
Turnover ⎱	
Rolloff ⎰	Disc equalizer
Selector	
Volume	Gain
Loudness	Hearing Contour Compensation
2 level controls	
Bass	
Bias adjust	
Input balance	
Output balance	
Line gain	Line gain
Microphone gain	Microphone gain
5 on/off switches	4 on/off switches

This simplification of my audio system has had several broad results. I now have a positive knowledge of the average acoustic output of my speaker, whether at full volume or low, and the balance is correct for every level. There is no dependence on acoustic memory and no extreme overemphasis. Reproduced music sounds close to that I hear in the concert hall. The use of the Hearing Contour Compensator permits the use of the following operating technique. After a source is selected, a suitable listening-room loudness level is chosen. The Compensator is adjusted to the necessary setting (−10, −20 phms, etc.) to furnish this loudness level. The gain control is then rotated clockwise fully, fading in the desired sound smoothly with no acoustic shock to the listeners. •

Adequate Audio Power in the Home

JAMES MOIR

A discussion of the factors affecting the power required for satisfactory reproduction of typical program material and the methods of calculating it.

ESTIMATES OF THE AUDIO POWER required to produce adequate loudness from the domestic loudspeaker are characterised by a very wide divergence of opinion even among authorities, figures ranging from 100 milliwatts to 50,000 milliwatts (50 watts) having been quoted by different writers. It is interesting to examine the problem and to attempt to produce some reliable data. As a preliminary it is necessary to clear our ideas as to what is meant by the 'audio power' for it is evident that the same basic power may be expressed in several ways. Thus the same amplifier may be quoted as having an output of ten or twenty watts both figures being accurate statements of the performance.

Expressing the Power

In a mains frequency power circuit the supply voltage and current have the substantially sinusoidal waveform of *Fig.* 1 and without ambiguity the power dissipated as heat in a resistance load of R ohms will be given by $(0.707 \; V)^2/R$ where V is the peak value of the applied voltage. To eliminate the necessity of always multiplying the meter indication by 0.707, commercial meters used in the heavy engineering field are scaled to indicate, not the peak value, V, but the *rms* (root mean square) value $v = 0.707 \; V$. Within the usual engineering tolerances the value of voltage or current will be indicated quite accurately by ordinary commercial meters and the reading will be independent of the physical size of the meter.

The multiplying factor, 0.707 applies only to a sinusoidal waveform but in the communications field sine waves are generally confined to test equipment, speech and music signals having the much "spikier" waveform indicated by *Fig.* 2. There is no equivalent numerical factor relating peak and rms values that can be applied to such irregular waveforms and thus the output of an amplifier may be expressed either in terms of its peak power, V^2/R, or as rms power $(0.707 \; V)^2/R$ the later figure being the power dissipated as heat in a resistor of R ohms by a sinusoidal voltage having the
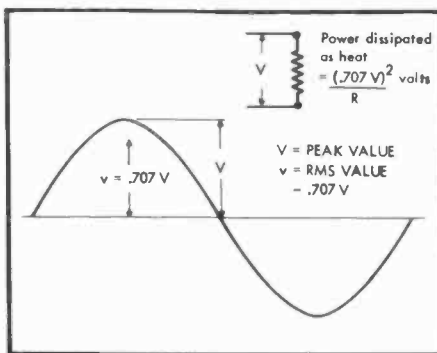

Fig. 1. Relation of peak and rms values of voltage for a sine wave.

same peak voltage as the speech wave. It should be appreciated that this is *not* the rms power in the speech wave but a figure which may be perhaps ten times higher.

On sinusoidal waveforms the rms power will only *be one half* (*ie* $(0.707)^2 = 0.5$) the peak power and thus the same amplifier may be rated in either peak power or rms power, the peak power figure being twice the rms power figure. As there is a fixed ratio between the two ratings there appears to be no good reason for departing from the practice of quoting the rms power output the standard practice in other engineering fields.

Measuring the Power

There need be no ambiguity in measuring the power output of an audio amplifier for sinusoidal test signals can be employed and special meters are not required, though it should be noted that the power specification is meaningless unless the distortion level is also quoted.

However our present interest is not in what power an amplifier *can* deliver but in what power it *does* deliver when used in the home. This is a much more troublesome problem, for speech and music waveforms are irregular, and have a high ratio of peak to rms power due to the intervals between words or phrases when no signal is present. Heating (a function of the rms voltage 0.707
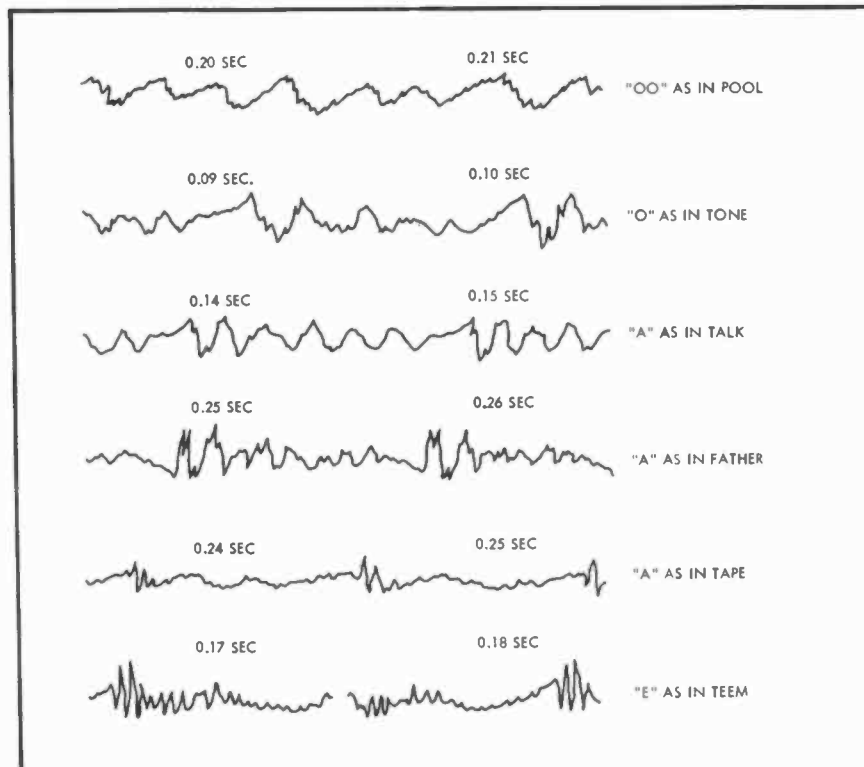

Fig. 2. Waveforms of typical vowel sounds. (From Fletcher, "Speech and Hearing.")

TABLE I

Preferred Maximum Sound Level
db above 10^{-16} watts/cm²

	Public Musicians			Programme Engineers		Engineers
	Men	Women		Men	Women	
Symphonic Music ..	78	78	88	90	87	88
Light Music	75	74	79	89	84	84
Dance Music	75	73	79	89	83	84
Speech	71	71	74	84	77	80

V) is of little consequence in either amplifiers or loudspeakers and in consequence it is more reasonable to measure the peak values of signal voltage and express the speech power in terms of its peak value, V^2/R.

The measurement of the peak voltage of such irregular waveforms is by no means easy. Pointer-type meters of any kind have movements of sufficient inertia to prevent them reading peak values and the indications may easily be in error by a factor of ten times. Large well damped meters of high nominal accuracy invariably have heavy moving systems and are particularly inaccurate when used to "measure" audio voltages. Measurements using pointer-type instruments of the programme voltage across, into a loudspeaker are therefore completely valueless. Three types of instrument are in current use for measuring sound power, the sound-level meter, the high-speed level recorder and the cathode-ray oscillograph.

The sound-level meter has the disadvantage of a pointer-type meter but as the mechanical constants of the meter are closely specified the error due to instrument inertia may be roughly estimated. A typical meter may give readings that are below true peak by 20 db, the error being small when the signal is steady and rising to 20 db on speech signals where the gaps between words and sentences may be comparatively long.

The high-speed level recorder employs a tube-operated servo system to drive the pointer and will generally indicate values that are 5–10 db below true peak readings.

The cathode-ray oscilloscope has no significant error due to inertia and can indicate true peak values on the most complex waveform, but care must be taken to operate with sufficient brightness to show up the faint high-speed traces characteristic of peaks of short duration.

Failure to indicate whether peak or rms power is being quoted and the use of unsuitable power measuring equipment undoubtedly accounts for differences of from 10 to 100 times in the amount of power thought to be necessary for domestic reproduction. This is a large error but even greater discrepancies can occur if the maximum loudness is not carefully specified.

What Constitutes Adequate Loudness

Difference of opinion as to what constitutes "adequate loudness" is responsible for considerable discrepancies between writers' estimates and the importance of clearing the air will be fairly obvious when it is realised that a difference of 10 db in specifying the maximum loudness level thought to be desirable will result in a change in the required amplifier output power of ten times. Published figures seem to indicate that the differences of opinion embrace a power range of something nearer 40 db (a power difference of 10,000 to 1) so it is absolutely necessary to have our thoughts clear on this point.

At first sight it appears reasonable to approach the problem by reviewing the volume ranges encountered in original speech and music on the assumption that "a perfect reproduction" will require the same volume range. The most difficult case, an original performance by a large symphony orchestra may involve a power ratio of 80 db (100 million to 1) but this range is generally only encountered for a few tenths of a second in several hours, a more frequently occurring range being nearer 74 db.

At the receiving end it is reasonable to assume that the listener should adjust his volume control to bring the *minimum* signal to somewhere near the room noise level and as an average value for the domestic noise level is about 40 phon it implies that peak levels in the region of 114 db (or phon) are required. Though this appears to be a very reasonable deduction, experience suggests that it is wise to make a check and this has been done both in England and in America. The B.B.C. have made a very careful study of the sound levels preferred by their monitoring staff and by the general public and Table I lists some of their data taken from a paper by Somerville and Ward.

In these tests the listeners were provided with a high-quality reproducer system of ample power handling capacity and were asked to set the loudness to the level they considered preferable. The acoustic level at a point about 18 inches from the listener's head was then checked with a standard type of sound-level meter. It is surprising to note that none of the listeners wished to have sound levels greater than 90 phon a re-

sult supported by similar tests in America which indicated a preference for levels about 8–9 phon lower than the B.B.C. results suggest.

Sound levels approaching 114 phon occur in concert halls and there is not the least evidence that these are anything but satisfying, but the available evidence does suggest that these levels are not optimum in the home. The reason for this difference is not clear, but in the writer's experience a level of 110 phon sounds "louder," though "smaller" and more oppressive in a small room than the same level in a concert hall.

A major discrepancy between the various estimations of "power required" may thus be attributed to the choice of maximum loudness thought desirable. An estimate based on the very reasonable assumption that concert-hall loudness levels are necessary in the home will suggest a power some at least 20 db (100 times) higher than another estimate based on achieving only the maximum preferred loudness level of 90 phon. As it will be seen from Table I that the general public only require a maximum loudness level of about 80 phon, a "logical" engineering estimate of the power necessary will be about 30 db (1000 times) higher than is really required.

This preference for lower levels in the home is providential because some consideration for the neighbours is necessary. In flats, terraced houses or houses built in pairs, a house-to-house insulation of 55–60 db can be achieved fairly easily by simple building techniques but science and the average builder are not yet in close touch, with the result that 45–50 db is the figure more usually achieved in semi-detached pairs of houses having a 9-in. party wall. Peak sound levels in the region of 110 phon will result in the neighbours enjoying *your* choice of programme at a level of 70–80 phon and while this may be just tolerable in the early evening when their own noise level is in the same region as your own it must become a little annoying to them when later in the evening their own noise level has dropped to something nearer 30 phon.

Acoustic Power Requirements

The next steps in the enquiry are to make an estimate of the actual acoustic

TABLE II

Maximum Loudness Levels produced by typical sound sources in domestic surroundings.

Small Upright Piano
Maximum in normal playing	— 72 db
Player asked to play a "loud" selection	— 82 db
Player asked to play "as loudly as possible"	— 90 db

Speech
Boy normal speech	— 60 db
Man " "	— 65 db

TABLE III

Acoustic Power required to produce given loudness levels in a room of 1540 ft³ and reverberation time of 0.5 sec. Computed from Eq. (7) of Appendix.

80 db	.00036	watt	(.36	milliwatts)
90 db	.0036	''	(3.6	milliwatts)
100 db	.036	''		
110 db	.36	''		
120 db	3.6	''		

power required to produce the loudness levels thought necessary, and then to examine the electro-acoustic conversion efficiency of loudspeakers for this will enable the electrical power requirements to be predicted.

The actual acoustic power required to produce acceptable loudness levels is very small indeed. A first approximation to the figure can be obtained by considering the data on the acoustic power required for normal conversation. The most reliable data, that of Sivian, Dunn, and White indicates that the instantaneous maximum power rises to about 700 microwatts (0.7 milliwatt) when making an impassioned speech to a large audience. About 5 per cent of speakers will produce powers five times higher than the figure quoted, making their acoustic output 3–5 milliwatts. Declamatory speech of this kind would be intolerably loud in domestic surroundings, rather suggesting that the maximum acoustic power required for any purpose is not likely to rise much above 5 milliwatts. Data is available on the acoustic output of most of the common instruments but it is not particularly useful as an indication of domestic requirements as all the figures refer to tests in which the instrument was played as loudly as possible. A concert grand, played loudly, has a power output of about 350 milliwatts but experience suggests that even a small upright piano can be intolerably loud in a small room. In my own room a small upright piano played by a moderately competent player produced the loudness levels shown in Table II and it is perhaps significant that normal playing gave maximum levels of 72 phon with a level of 90 phon reached when the player was asked to produce the absolute maximum output. It should be noted that readings were taken when the sound level was reasonably steady and the absolute peak levels are therefore likely to exceed the meter readings by only 4–8 phon.

Calculation of Sound Power Requirements

In the appendix it is shown that the acoustic power required to produce a sound level of 100 db can be computed from

$$P = .0000116 \ V/T \ watts$$

where V is the room volume and T is the reverberation time. Applied to one of my own rooms having a volume of 1540 cu. ft. and a reverberation time of

0.5 second it suggests that the power shown in Table III will be required for levels of 80–120 db, the power required for 100 db being computed from the equation directly, and being modified by a factor of ten for each 10 db change in level. The suggested maximum requirement of 90 db is reached with an acoustic power of only 3.6 milliwatts, a figure that is in substantial agreement with the power deduced from that produced by a human speaker at maximum output.

Objection has been raised to any formula that suggests that the power required is inversely proportional to the reverberation time, on the score that the bursts of energy in speech are so short that room reflections do not have time to reinforce the direct sound from the speaker. It has therefore been suggested that the power required should be computed on the assumption that the loudness is entirely due to the direct sound. The calculation is not difficult but it does require a knowledge of the polar diagram of the loudspeaker over the frequency range.

A sound wave leaving the speaker will diverge in the form of a solid cone with

TABLE IV

Electrical Power required to produce a loudness level of 80 db from three typical speakers.

A—17-in., 17,000 gauss magnet.
B—12-in., high-fidelity type, 14,000 gauss.
C—8-in., radio receiver type, 8,000 gauss.

Speaker	Sound Level db	Voice Coil Power mw	Electro-acoustic Efficiency, percent
A	80	9.5	3.8
B	80	55	.66
C	80	240	.15

the speaker at the apex but the angle of divergence will be a function of frequency, being greatest at low frequencies (180 deg. if the speaker is in the centre of one wall) and decreasing as the frequency increases until it is down to something near 25 deg. at 5000 cps. There is therefore some difficulty in fixing an effective average angle for the whole of the audio frequency range. Power, loudness and intelligibility are not linearly proportional to bandwidth, a fact that increases the difficulty in fixing an average angle for the whole frequency range. In spite of these difficulties it has been claimed that power requirements computed on the assumption that there is no gain in loudness from the reverberant sound, do give good agreement with measurement.

The earlier discussion suggests that the maximum acoustic power required in domestic surroundings is only in the region of 3–5 milliwatts but in the absence of data on the electro-acoustic efficiency of typical loudspeakers it is

difficult to translate the acoustic power requirements into electrical power to be provided by the amplifier.

Electro-acoustic Efficiency of Loudspeakers

There is very little published data on the conversion efficiency of loudspeakers, partly because of the difficulty of measurement but also because any single figure can be misleading and liable to misinterpretation. In these measurements to be described, the figure quoted as the efficiency was determined by measuring the electrical power input to a loudspeaker operating on ordinary programme in the normal living room and simultaneously measuring the loudness level in the room. Care was taken to observe steady values and from this data the acoustic power output was calculated. The efficiency is the ratio

$$\frac{Acoustic \ power \times 100.}{Electrical \ power}$$

With domestic approval a sound-level meter, oscilloscope and oscillator were set up in the dining room as shown in *Fig.* 3 and several listening and watching sessions enjoyed. As a first check some co-operative members of the family were asked to adjust the loudness to their liking and as it was found that the levels chosen were in good agreement with those obtained by the B.B.C. (Table I) it was assumed that nothing was seriously amiss. The procedure then employed for the power measurement tests was to set up the CRO and sound-level meter in close proximity to enable both meter and CRO to be viewed simultaneously and to mark the tube face each time the meter peaked to 80 db. After a few attempts it was possible to draw two parallel lines on the tube face defining the maximum deflections produced when the sound-level meter reached this figure. A Promenade Concert provided valuable test material, as it was possible to watch the meter on one phrase and check the CRO deflection when the phrase was repeated a second or so later. Music also has the advantage that complex tones are held for sufficient time to provide a steady deflection on the meter, thus eliminating any argument about the contribution of the
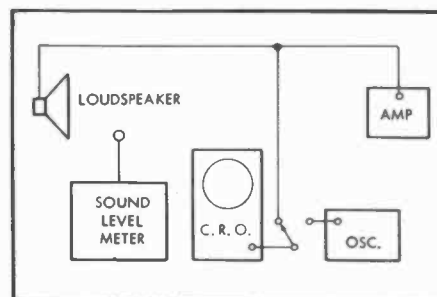

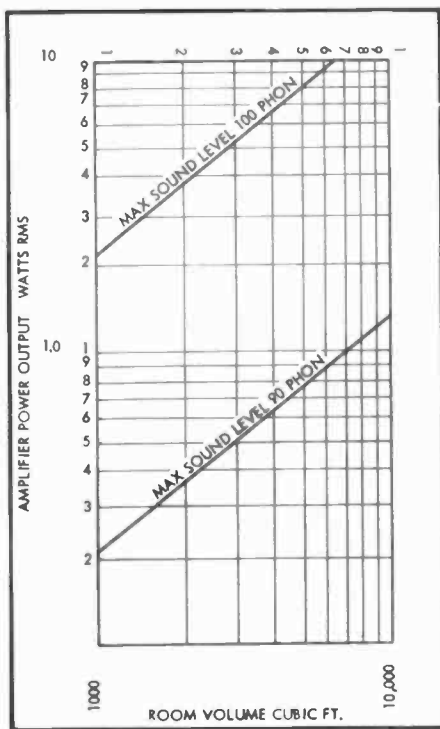Fig. 3. Schematic arrangement used for audio power measurements.

Fig. 4. Curves of power required for two sound levels in relation to room volume.

reverberant sound to the total loudness. Audience applause is equally effective for this purpose. The room was in semi-darkness and a bright trace employed to avoid missing sharp peaks of short duration.

Having defined the CRO deflection characteristic of a sound level of 80 db, the CRO was then switched to the calibrated oscillator and the rms voltage corresponding to the deflection noted. Three hours checking with three different loudspeakers provided some interesting data which is reproduced in Table IV.

As the input power to each of the speakers was adjusted to produce an acoustic level of 80 db in the room, it is assumed that the acoustic power produced is the same for all, a reasonable but not a precision conclusion, in view of the different frequency characteristics inherent in speaker units of such widely varying quality. Column 3 indicates the power in the voice coil computed on the assumption that the effective resistance of the voice coil is equal to its d.c. resistance. Column 4 contains figures for the electro-acoustic efficiency computed from the measured electrical input to the speaker on the assumption that the acoustic power output is given by Eq. (7) in the Appendix, corrected to a sound level of 80 phon.

Speaker A is a large 18-in. cone speaker having a 2½-in. voice coil working in a gap having a flux density of 17,000 gauss. Speaker B is a standard type of unit typical of the better quality 12-in. high-fidelity units, while speaker C is typical of the cheaper 8-in. units included in radio receivers.

Speaker B, typical of the units being used by most high fidelity enthusiasts only requires an input of about 55 milliwatts to produce a sound level of 80 db and a power of 0.55 watt to produce 90 db. If concert-hall levels of 110 db were required in domestic enclosures a power of 55 watts would be necessary but this speaker would have to call for help from at least four of its fellows if this power was to be handled.

Though a horn loaded unit was not tested it is known that electro-acoustic efficiencies of 20–40 per cent can be reached, enabling the concert hall level to be obtained for an input of about 1½ watts. As evidence of this, some recent measurements in a 700-seat theatre having a volume of 120,000 cu. ft. showed that the feature film was being regularly run with a maximum electrical input to the loudspeakers of less than one watt.

The 18-in. speaker is shown to have an efficiency twenty times that of the cheap radio speaker but this is insufficient to justify its use where cost is of importance, for acoustic power can generally be produced more cheaply by the combination of a small speaker and a large pentode, than by an expensive speaker and a small triode.

It is convenient to have available for ready reference curves relating to room volume, sound level, and electrical power required. *Figure* 4 provides this information based on the assumptions that

1. The acoustic power is computed from Eq. (7).
2. A loudspeaker efficiency of 1 per cent is obtained.
3. The optimum reverberation time relation of *Fig.* 5 is approximated in all cases.

In the majority of rooms above 2000 cubic feet the reverberation times of *Fig.* 5 are approximated, but in smaller houses current constructional methods appear to give a reverberation time of about half a second almost regardless of the furnishing scheme.

After reviewing the results obtained it appears that there is great opportunity for difference of opinion in estimating the power required to produce adequate loudness in small rooms. An experimenter measuring the power that gives him adequate loudness will find it to be in the region of 50 milliwatts if he uses a CRO, perhaps 5 milliwatts if he uses a high-quality rectifier voltmeter, and something less than 1 milliwatt if he has an rms-reading thermal meter. A devotee of Aristotle preferring meditation rather than experiment might be excused if he based his calculations on the assumption that the loudness level found desirable in concert halls would prove to be equally desirable in the home. He would then produce a figure approaching 40–50 watts, but if this was thought to be insufficiently impressive, he could with all honesty quote the same power as 80–100 watts peak, i.e. peak volts times peak current. A difference in estimate as great as 100 watts to .001 watt must be a record for an honest difference of opinion in the engineering field.

Though the reason is probably psychological the preference for reduced maximum loudness levels in the home is not understood and should form an interesting subject for further study.

APPENDIX

If it is assumed that "loudness" is related to the steady-state sound intensity the power required to produce any specified intensity can be computed from the standard exponential relation between sound-energy density and the time interval during which power is being supplied to the enclosure. The sound-energy density in ergs/cc at any time t secs. after the power is turned on, is given by

$$E = \frac{4P}{CS_a} \ (1 - C \cdot c^{Sat/4V}) \qquad (1)$$

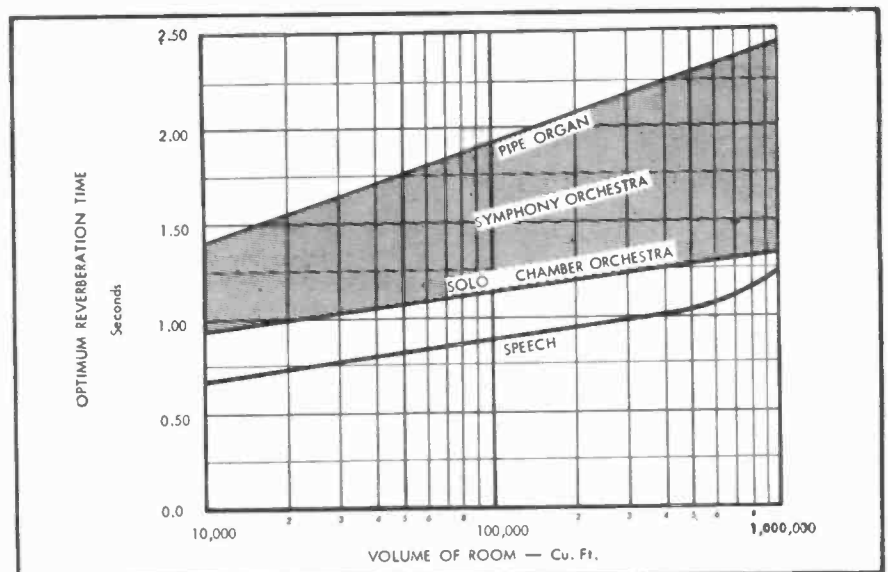Fig. 5. Curve showing optimum reverberation time in relation to room volume.

ADEQUATE AUDIO POWER

where

P = rate of emission of the source, ergs/sec.

C = velocity of sound, cm/sec.

S = total surface area of absorbing surfaces, sq. cms.

α = average coefficient of absorption of all surfaces.

V = total volume of room, cu. cms.

When steady state conditions are reached, theoretically after infinite time, but practically after T secs. where T is the reverberation time of the enclosure, the bracketed term is equal to unity and the sound energy density is given by

$$E = \frac{4P}{CS\alpha} \qquad (2)$$

It is more convenient to have a relation involving the reverberation time T and the volume of the enclosure V rather than S and α and this can be obtained from the normal Sabine relation for reverberation time $T = kV/S\alpha$, from which $S\alpha = kV/T$. Substituting kV/T for $S\alpha$ in Eq. (2) gives

$$E = \frac{4PT}{CkV} \qquad (3)$$

from which the source power in Ergs/sec. is given by

$$P = \frac{CkVE}{4T} \qquad (4)$$

If some standard intensity is adopted, the arithmetic is simplified and as 100 db is a convenient figure this will be inserted. It corresponds to a sound intensity of 10^{-6} watts/sq. cm. and a sound energy density of 3×10^{-4} ergs/cu. cm. Substituting this value in Eq. (4) and including all constants, the acoustic power in watts required from the source to produce a maximum intensity of 100 db is given by

$$P = \frac{3.4 \times 10^4 \times 16 \times 10^{-4} \times 3 \times 10^{-4}}{4 \times 10^7} \times \frac{V}{T} \qquad (6)$$

$$= 4.1 \times 10^{-10} \, V/T$$

or converting to ft. units

$$P = 1.16 \times 10^{-5} \, V/T = .0000116 \, \frac{V}{T} \, watts \qquad (7)$$

For any loudness level other than 100 db the power required will be doubled for each 3 db increase in intensity that is considered necessary. The threshold of pain is reached at an intensity level of about 120 db requiring a power 100 times that given by the equation and presumably fixing the absolute maximum value of power that *anybody* might ever consider necessary.

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HOW TO PLAN

quirements more closely and to develop a better understanding of actual rather than fancied needs; it may also serve to stimulate more interest in still better performance. Another experimental approach is to take advantage of the free home trials offered by local shops. As much as you may like a piece of equipment in the store, sometimes it may not "wear well" at home.

A little rummaging around in the attic in my own case produced components that were adequate for the interim approach. For the dance party requirement, there developed a need for power output—undistorted—that was well beyond the ability of the secondary system. The audience was found to absorb substantial amounts of acoustic power, and in addition, create an ambient noise level which had to be overcome by the speaker output. The lack of low-frequency tones in the secondary system became apparent because the lows often carry the rhythm needed for dancing. For listening to concert music, the volume level had to be reduced to keep peace in the family, so the loss of lows was again accentuated. The lack of extreme highs caused no comparable deep concern, so the extended high-frequency coverage will be included in the final system if it can be gained for a modest cost.

If you have been able to establish a standard of performance that the hi-fi system must eventually meet, then it is time to turn attention toward the second step in the planning cycle, that of selecting the individual components in such a way that the over-all cost is minimized without sacrificing the performance standard. In principle, at least, the beginner could extend the technique of listening tests so as to arrange the components to suit his need. We know of no better way to choose a speaker, but selecting a particular amplifier depends to some extent on how much the experimenter borrows from the "sound engineer" and the "music critic." Many would be satisfied with a simple substitution test using the chosen speaker and enclosure, and make use of a home trial. Selecting an amplifier in itself can be a detailed study the scope of this discussion; in fact, much has already been written on this subject.

Thus, if one can arrive realistically at some conclusions regarding his requirements in relation to what he can afford, the task of selecting the units for the system is reduced to manageable proportions; and one's limited energies and funds are not misdirected into unproductive channels. These two approaches are advocated as aids to planning: The comparative evaluation of system performance by actual listening tests, and the improvement of personal judgment through experience with an experimental or interim hi-fi system. Æ

Building Simplicity into the Hi-Fi System

ROSS H. SNYDER

Interconnection of the many elements of a hi-fi system demands some clear thinking if complete satisfaction is to be obtained. In particular, the method of connecting a tape recorder in the system for most convenient operation is outlined clearly by the author.

HIGH-FIDELITY COMPONENTS for home entertainment systems have been developed in such numbers and variety that rigorous application of good over-all systems' design has been almost impossible. The popularity of high-fidelity equipment has increased at such a pace that designers of components of all kinds have been faced with the necessity of building in a sort of makeshift universality which made it possible, in general, for the purchaser of an assemblage of these components to plug together a system which would function, but usually at something less than optimum.

Too Many Knobs

The commonest defect of an assembled system is multiplicity of controls, many of which perform duplicate functions. The likelihood of imperfect performance is great when such systems are operated by people whose interest is mainly in the music, not in the equipment upon which it is played. Typically, a radio tuner will have on its face a control for the selection of FM, AM, Phonograph, Tape, TV, etc. It will also have a volume control. Frequently such a tuner is selected especially because it has *relatively* few controls, and is connected into an elaborate audio control box, which will possess phonograph inputs, phonograph equalizer controls, power switch, volume

or loudness control (or both), and separate bass and treble tone controls. These control boxes are usually connected to power amplifiers which have, themselves, at least a gain control, and sometimes another set of tone controls and selectors. These are happily rare now that the basic "flat" amplifier is the norm. With a basic flat amplifier, adjustment of the main amplifier gain control is required only at the time of installation, and it is supposed to be set by the installer so as to provide correct gain for the audio control box with which it is used. There is a tendency for many users to adjust the power amplifier gain so that the control box volume control is rotated about one-third at comfortable room level. Those control boxes which contain loudness controls are usually contributing considerable "bass boost" at this rotation, and this effect is removed only when thunderous volume is being delivered into the living quarters by the equipment. The function of the loudness control should be to remove all artificial bass boost at a sound level equal to that which would be heard if the listener were in the room where the recording was made. Complete instructions on this adjustment are more and more being included in the Instruction Manuals which accompany high-quality equipment, and if the listener is so minded, the facility for proper functioning of his loudness control is at hand.

The Maximum Hum Control

But the handling of cascaded volume controls, one on the tuner and one on the control box, for example, is not so simple. The way is always open, if not for the high-fidelity enthusiast in the family who is responsible for the purchase and installation of the equipment, at least for other members of the family to adjust the equipment for maximum hum, or for maximum distortion. A tendency will be found, for example, to operate the volume control on the tuner at a medium setting, and then, upon interruption, for the listener to turn the volume down temporarily *at the audio control box.* Following the interruption, the listener may find it most convenient to raise the volume level again, *this time using the tuner control.* Thus, those amplifier tubes which follow the tuner volume control, may well be driven far into distortion, while the volume level of sound in the room is not particularly high, having been reduced by the control box knob. If the knob on the control box is a loudness control, this function will also be disturbed. On the other hand, if the procedure is reversed, and the level of sound reduced at the tuner, then later raised at the audio control box, the "maximum hum" situation will exist. The level of sound through those amplifier stages which exist after the tuner volume control, but before the audio control box volume control, may be sufficiently reduced as to be comparable to the internal hum level, and raising the amplification *with the control box knob* after *the unnecessary reduction* will result only in an *increase* of hum and unpleasantness. The remedy is, of course, to eliminate one of the volume controls, or at least to remove it to a screwdriver adjustment at the back of one of the components, and to leave in the hands of the listener only a single knob for the control of level. In this manner, adequate level in lines between components may be set at the time of installation, for the best compromise between low distortion and good signal-to-noise ratio. The difficulty of predicting what amplifier will be used with a tuner of given design is, of course, the reason for incorporating the control into the tuner in the first place. The thought was that it's better
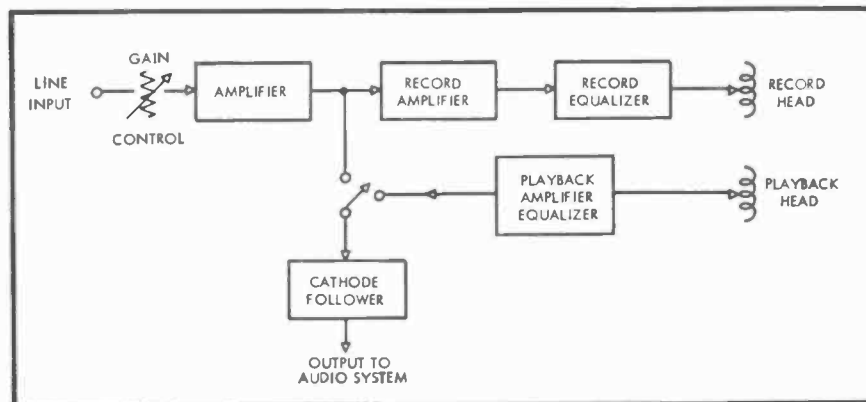

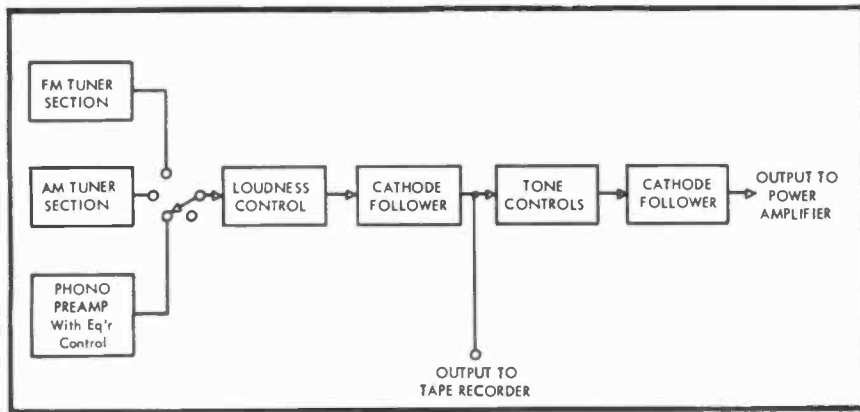Fig. 1. Block diagram of a typical high-quality tape recorder designed for home use.

Fig. 2. Block schematic of one model of deluxe FM-AM tuner with magnetic phono preamplifier and tone controls.

to provide more controls than necessary, rather than to eliminate one which may be needed. But it was a bad thought.

Volume controls, as an industry practice, probably should be left off radio tuners, unless these tuners also incorporate phonograph preamplifiers and tone controls, and are intended to serve as complete control centers as well as tuners. If unnecessary controls are provided, the installer ought to remove them. More than one maunfacturer follows good practice in this, and others should, for the greater convenience of users and for less opportunity for unpleasant sound in the home. Only those few listeners who wish to eliminate all functions from their systems excepting radio, and who wish no tone controls of any kind will need gain control on the basic tuner. Those may be of sufficient technical skill to devise a convenient volume control for themselves.

"How Many Selectors?"

The duplication of selector controls appears to be a harder problem. On those radio tuners which are designed as an adjunct to an audio control box, there is no justification for the provision of anything other than an AM-FM control, possibly with broad and sharp positions on AM, and AFC or no-AFC positions. Incorporation of selector positions for phonograph, tape, TV, and other sound sources may appropriately be provided only on tuners which also possess magnetic phonograph pickup preamplifiers, and are designed to function as combined tuners and audio control boxes. The ultimate simplification in control is probably provided by those tuners which have been designed for two-channel stereophonic service, and bring each of the two outputs, AM and FM, to separate jacks, for delivery to the audio control box. Such tuners may readily be connected into audio control boxes so that selection of phonograph, AM, FM, tape, etc., may be made on one knob on the audio control box, and there only. Realistically, no selector controls other than those for AM or FM should be incorporated into tuners which are designed as adjuncts to audio control

boxes, but only into the tuners which are designed as combined audio control centers and tuners. The very least which we should ask is that the knobs on the front panel of these tuners should be arranged so that if the a.c. switch and volume control are removed, the panel remains balanced and symmetrical. Many home high-fidelity system owners insist that all knobs be left on the panel, even though some are superfluous or rendered functionless, in order to preserve balanced appearance; it should be made possible to preserve both good appearance and good operation.

The Underfed Tape Recorder

As the high-quality magnetic tape recorder becomes a staple in the list of components in a high-fidelity home system, a provision for its incorporation in simple plug-in form becomes a necessity. The logical place for the incorporation of plugs which are intended to connect to the tape recorder input and output is in the audio control box, or into the tuner which is intended to function as an audio control center.

A block diagram of a typical high-quality magnetic tape recorder is shown in *Fig.* 1. In those recorders which use a common magnetic head as both record and playback, the selector switch is so ganged as to perform essentially the same function as that shown. Typically, the line input of the recorder presents a

high-impedance load, which may be bridged across a number of available points inside the audio control system, and requires no more than 0.5 volts rms to drive the recorder to maximum record level. This is a simple requirement to meet, although consideration for it has often been omitted from commercial components. The output from a typical magnetic recorder is of low internal source impedance, and relatively high level, which will adequately drive any of many possible points in the audio control system.

There are several typical arrangements for tape recorder connection. *Fig.* 2 illustrates a widely sold deluxe FM/AM tuner with magnetic phonograph preamplifier and tone control, designed to function not only as a tuner, but also as an audio control center. At least four input jacks are provided, one of which is intended for tape. An output is also designated for connection to the tape recorder. The signal which is delivered for recording to the tape recorder is, unfortunately, unsatisfactory. The user may reasonably be supposed to be listening, over his loudspeaker, at the same time he is producing a tape recording. The tuner selector may, then, be set for FM, for example, the loudness control adjusted to comfortable level. The output to the tape recorder, then, located *after* the gain control, is extremely low in level, and is exaggerated in bass, because of the effect of the loudness control at these low listening levels. Typical peak voltages obtained from this array will measure around 50 millivolts. Not only does this provide inadequate drive for the recorder, but it also produces a tape which is artificially heavy in bass. The output to the tape recorder could hardly have been located at a worse position. So far as proper level and proper equalization for the production of a flat tape recording are concerned, the tape recorder jack might better be connected between the loudness control and the selector switch. A cathode follower would be advisable, of course, as isolation, and in order to assure that the load of the tape recorder and the capacitance of the interconnecting cable would not affect the over-all performance of the system, or of the recorder.
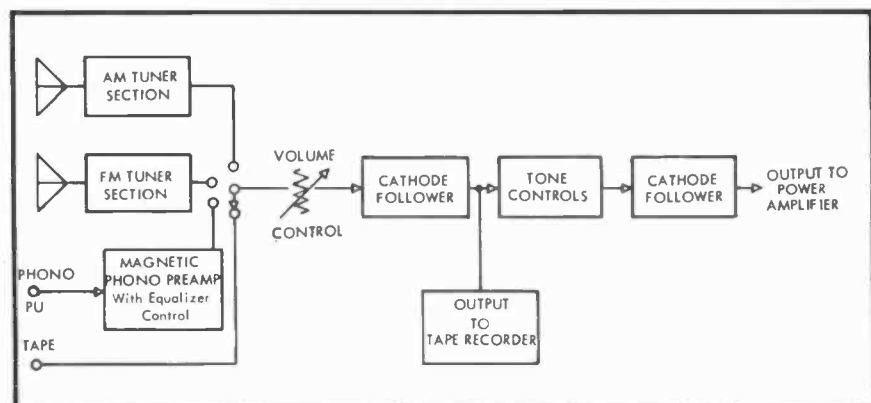

Fig. 3. Arrangement of popular-priced FM/AM tuner with phono preamp and tone controls.

The Case Of The Howling Tape Recorder

With the configuration of *Fig.* 2, however, still another source of unpleasantness for the listener is offered. Assuming that a recording has been made, the listener may now switch the selector control on his tuner to the fourth position, into which the output of the tape recorder has been plugged. Referring to *Fig.* 1, if the tape recorder selector switch has been left in the "record" position, a feed-back path is created from loudness control through cathode-follower to tape recorder input, to tape recorder output, to selector switch, to loudness control. The result is usually a loud howl. This effect can be avoided, of course, if the listener is careful always to turn his tape recorder switch to "playback" *before* he changes the position of his tuner selector control, but it would surely be good design to prevent so likely a cause for unpleasantness.

Figure 3 is of a lower cost tuner than that in *Fig.* 2. This unit does not have the fault of presenting to the tape recorder an artificially unbalanced signal, since the volume control is not a loudness control, but it does have the same fault as the tuner of *Fig.* 2 in being likely to be so used as to present a very small signal to the input of the tape recorder, due to the listener's having set his volume control for comfortable listening, rather than for adequate level for recording. This configuration also possesses, still, the possibility of feedback "howl."

Figure 4 outlines the configuration of a popular deluxe one-piece power amplifier and audio control box. In this case, the output to the tape recorder is of adequate *level*, but has been subjected to "tone control" whose purpose is primarily that of adjusting the sound for most comfortable listening, rather than for flatness. There is good reason for tone controls, of course. But a flat signal should nevertheless be presented to the tape recorder. Tone controls are for playback, and not for recording. It should be assumed that the listener will wish to adjust his tone controls, every time he listens, for conditions which exist at the moment, and which may not always be the same. To feed a signal through the same set of tone controls twice is to "double" the effect of the tone controls
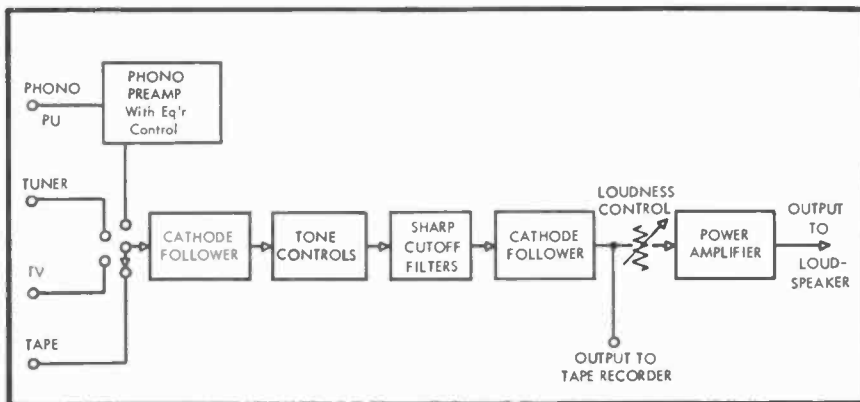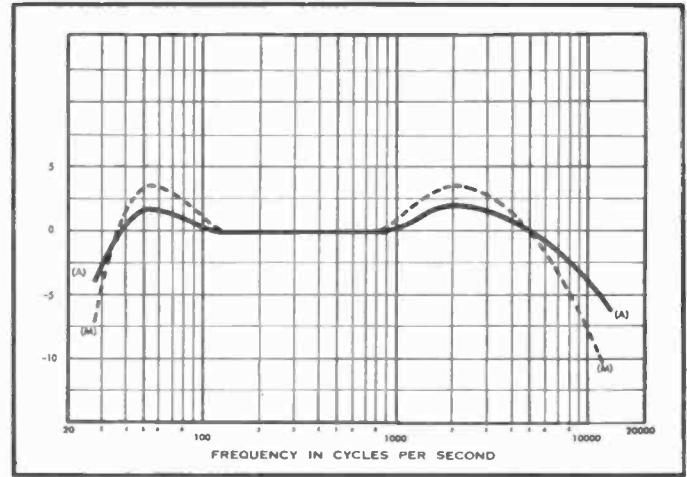

Fig. 5. Curves showing additive effect when signal passes twice through an amplifier which is not perfectly flat.

upon playback. On only one generation, with such a process, a 12 db-per-octave bass boost slope may be obtained, or a very sharp high-frequency cut-off be introduced.

Double Dipped Tone Controls

Even if the "tone controls" are set at the "flat" position, which often is not marked accurately on the control panel, this position usually is a little off true flatness. Typical of the "flat" position on tone control systems is the curve shown at (A) in *Fig.* 5. This curve, it is true, is "± 2 decibels from 50 cycles to 8,000 cycles." But, suppose the tape, which has been recorded to this degree of "flatness," is now played back, through the system shown in *Fig.* 4. Even though the tone controls be left *unaltered*, the playback curve differs from flatness by twice the amount of Curve A, forming Curve M. This, flat now only by ± 4 decibels from 50 to 8,000 cycles, is sharply rolling off at both low and high frequencies, with severe "bumps" in response. Small deviations from flatness in the tone control system, which are entirely negligible so far as the original function of the controls is concerned, now become major sources of unpleasantness, which may be blamed upon the tape recorder, even though, in this example, the tape recorder was assumed to be perfectly flat in frequency response.

If the tone controls had been set for only a little departure from the nominally "flat" position, the results would have been even worse.

Figure 6 illustrates a high-quality commercial audio control box which provides the tape recorder with a flat signal of adequate level. With this control box, the possibility remains for feedback, if the listener chances to select "tape" on the control box before switching the recorder to "playback," but all other considerations of good practice are observed. The general configuration of the commercial system in *Fig.* 6 is shown in *Fig.* 7. Great flexibility in the arrangement of loudness or volume controls, tone controls, sharp high or low cut-off filters, and so forth, may easily be designed without essentially changing this arrangement. Only the feedback problem remains.

High Fidelity Unlimited

Figure 8 shows an ideal configuration for incorporating a tape recorder into a high-quality home music system. Whether the arrangement for this connection is made in a tuner which is designed also as an audio control center, or in a deluxe audio control box is unimportant. The provision for placing the tape recorder in *series* with the circuit is the key to the removal of any possibility of feedback "howl." A low impedance-source signal to the tape recorder line should be provided—cathode followers work well. A jack should be provided for this output. If, then, *no connection* were normally provided between this and the jack which is to bring back the output of the tape recorder, the possibility for series insertion of the recorder exists. A simple jumper may be provided as standard equipment, to be removed when the tape recorder is installed. With this connection the tape recorder is either left on at all times when the system is being used, or the tape recorder may be provided with a means of automatically connecting its input to its output, directly, when the recorder is turned off. Such an arrangement is offered as standard equipment on some tape recorders which have been designed for home use, and is available as a factory modification on others.

Fig. 4. Block diagram of deluxe one-piece power amplifier and audio control unit.

Hi-Fi Surgery

When the owner of a high fidelity system, on which he may have spent many hundreds of dollars, buys his tape recorder, it is too late for corrective action by the manufacturer of his tuner or of his audio control box. If the machine is to function well as a unit, some sort of "corrective surgery" is going to be needed. This may range from simply unsoldering one connection from its present location, and soldering it to a new one, all the way up to the incorporation of an additional tube, and the changing of several wires. None of these procedures is beyond the skill of a typical hi-fi technical enthusiast, but probably ought to be undertaken only by his serviceman if the listener is one of the many thousands of newcomers to the field whose interest lies mainly in the music and not in the knobs and gadgets.

A tuner like that illustrated in *Fig. 2* might be modified in either of two ways. A single wire will be found which leads from the jack marked "output to tape recorder" to a certain pin on one of the tubes. This wire may be unsoldered from the tube, and transferred to the selector
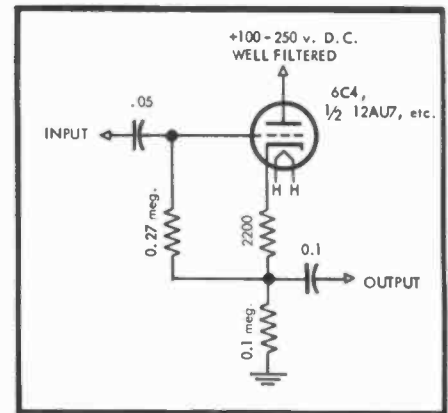
Fig. 9. Schematic of cathode follower stage which can be added to a tuner or control unit to provide suitable coupling to a tape recorder input.

switch, being soldered to that lug on the switch which represents the "rotor" or, if more convenient, to the "top end" of the loudness control. A more elegant solution would be to cut the chassis at a convenient point, to mount a new tube, such as a 6C4, in a tube socket, and to connect this newly added tube as a cathode-follower. The wire leading to the "output to tape recorder" jack would, then, be connected to the output of the cathode-follower, while the grid of the newly added cathode-follower would be connected to the rotor of the selector switch. Appropriate values for such a cathode-follower are shown in *Fig. 9*. Connection of the follower to the tuner's filament supply will probably present little problem, but the selection of an appropriate connection for the plate should be done most carefully. A point on the schematic diagram of the tuner should be found at which considerable "decoupling" has already been provided, and at which a large capacitor is already connected. The plate of any cathode-follower should be connected to a high-voltage d.c. source which is effectively "grounded" for audio signals. In most cases, it will be found that the follower will function well if its plate is connected to the same point as the high-voltage end of one of the plate resistors in a low-level audio amplifier stage.

If, in the case of a tuner like that in *Fig. 3*, no attempt is made to install a cathode-follower, care must be taken that the wire which carries the signal to the tape recorder is of the "low-capacitance" shielded type, and that the input impedance of the tape recorder is not so low as to "load" the volume control unduly. Otherwise, some distortion could occur, and if the wire were of high capacitance, the high-frequency response of the system could be impaired.

The audio control box illustrated in *Fig. 6* offers the possibility of simple wiring changes in order to effect the "series" configuration of *Fig. 8*. It is possible to lift the connection between the cathode-follower and the loudness
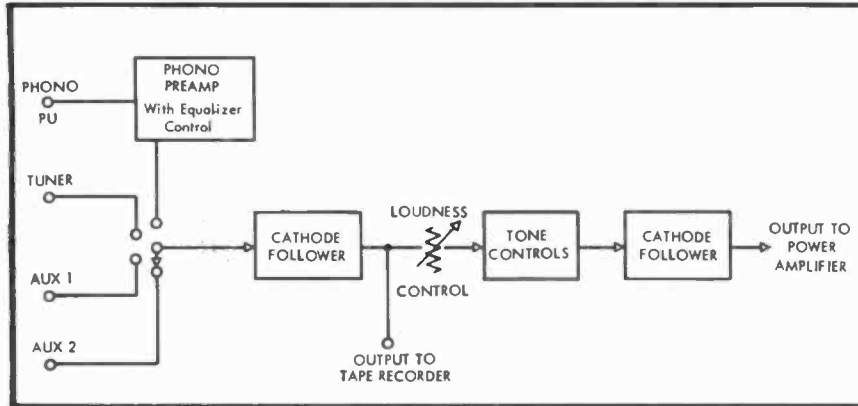

Fig. 6. Block schematic of high-quality commercial audio control unit.
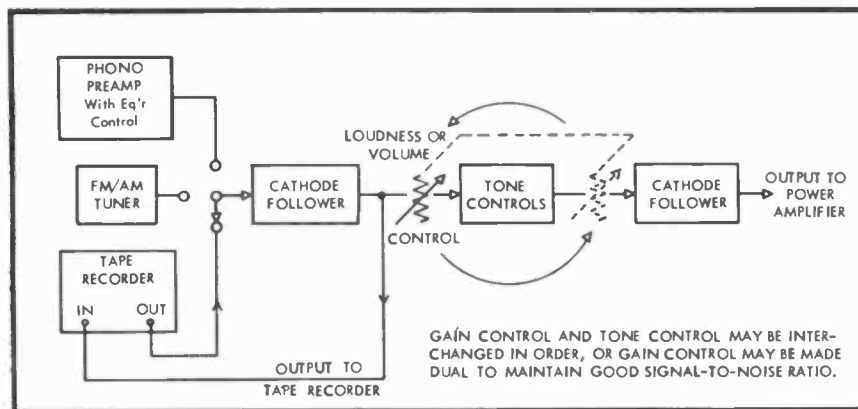

Fig. 7. General configuration for control box or tuner control system. Possibility of feedback is still present even with this arrangement.
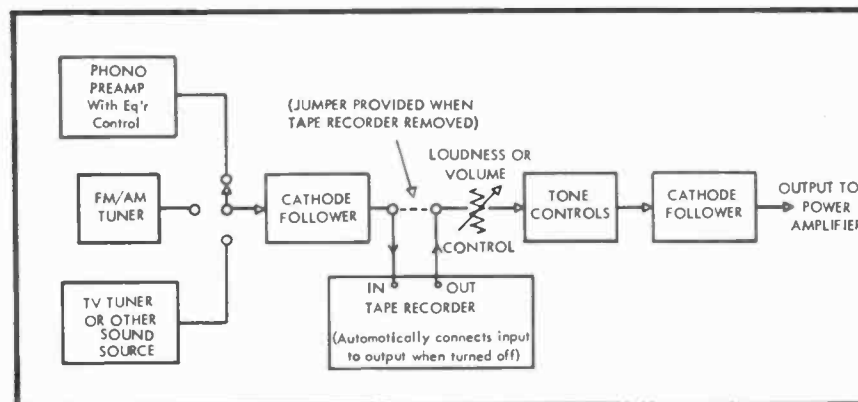

Fig. 8. Ideal configuration for incorporating a tape recorder into a high-quality home music system.

The Care and Treatment of Feedback Audio Amplifiers

W. B. BERNARD, CDR., USN

Anyone who has noticed a lack of stability in his Williamson-type amplifier may have wondered what caused it and how it could be corrected. The author gives the reasons and describes the methods taken to eliminate the troubles.

INVERSE OR NEGATIVE FEEDBACK has become widely accepted in the design of audio amplifiers. It may safely be said that it is incorporated in all output amplifiers of quality manufactured at the present time. This widespread use results from the benefits that can be produced by its application. This discussion will be limited mainly to the application of inverse voltage feedback to audio amplifiers. This type of feedback acts to reduce the output impedance of an amplifier in addition to reducing the distortion and noise produced. It also extends the frequency response of the amplifier.

The gain of an amplifier with voltage feedback is given by the equation $A_f = \dfrac{A}{1 - BA}$, where A is the raw gain of the amplifier (gain in the absence of feedback), and B is the percentage of the output voltage that is fed back. Distortion is reduced by the same proportion and, if the input signal is increased to make up for the loss of gain, the noise introduced by the amplifier is also reduced by the same factor. The loss of gain is a small price to pay for the benefits derived since voltage gain is easily obtained.

The output impedance of an amplifier with inverse voltage feedback is given by the equation $Z_f = \dfrac{R_p}{1 - \mu B A_1}$ R_p is the plate resistance of the output tube, μ is the amplification factor of the output tube, B is the portion of the output voltage fed back, and A_1 is the amplification of the amplifier between the point where the feedback voltage is inserted and the grid of the outut tube. In the case of pentodes and tetrodes where μ may be up to 200 or 300 it can be seen that a very small amount of feedback will give a tremendous reduction of output impedance. *Figure* 1 shows the result of applying 1/15 of the output of a 6L6 to the grid of the tube. The plate resistance is reduced from about 25,000 ohms
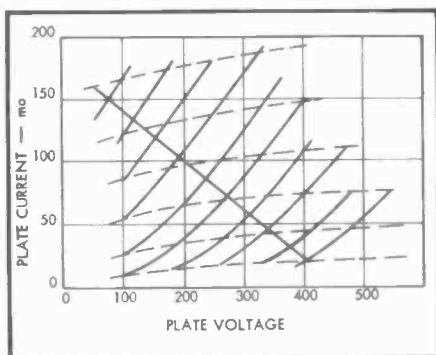

Fig. 1. Typical plate family of a 6L6 with 1/15 of its output fed back to the input.

to less than 2000 ohms. Observation of the curves will show that the power-output capabilities of the tube have not been diminished and a comparison with curves for a triode-connected 6L6 will show that it is much more linear, and regardless of the load placed on the tube the tetrode with inverse feedback has superior characteristics. Inverse feedback applied to the triode will make the plate characteristics more linear but they can do nothing to increase the power capabilities of the tube.

In practice the amount of feedback that is needed to reduce distortion reduces the output impedance to a satisfactory value. There are opinions which diverge from this view but they are seemingly in the minority and are divided between those who think that the usual amount of voltage feedback does not sufficiently reduce the impedance, and those who think that it reduces the impedance too much. From the standpoint of standardization both of these views create difficulties because it seems that all that it is reasonable to expect of a speaker manufacturer is that he will strive to produce a speaker which will give a uniform acoustic output over a given freuency range when the speaker is furnished a uniform voltage input.

If we are satisfied that the amount of inverse voltage feedback which we are going to apply will give us a usable output impedance—one fifth of the load

impedance or less—we may eliminate the consideration of over-all current feedback and the additional complications which it entails. We may still make use of negative current feedback inside the main feedback loop by such means as unbypassed cathode resistors. Having decided that negative voltage feedback is what we need in our amplifier we must consider how much we need and how we should apply it.

Feedback Methods

We may say that reducing the distortion to 1 per cent IM just before we drive the output grids to clipping level is a reasonable standard for high fidelity purposes. There may be some argument with this standard, but it is very close to what is generally realized in the better amplifiers today. In the usual circuits this calls for about 20 db of inverse feedback. In a properly designed amplifier the major portion of the distortion will be produced in the output stage; therefore, any useful feedback system will include the output stage. Internal feedback loops which do not include the output tubes should not be counted as being effective in reducing the total distortion. Such figures are most useful for advertising purposes.

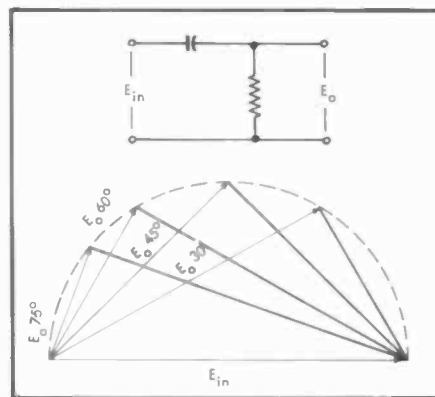At this point we may mention briefly two other feedback systems. The output

Fig. 2. Typical coupling circuit with one time constant and its circle diagram showing phase shift.

tube cathodes are sometimes returned to the ends of a secondary or tertiary winding on the output transformer which is so phased that the voltages produced in this winding oppose the grid to cathode signal voltages. This connection is very effective in reducing distortion and output impedance, but has the disadvantage that special output transformers are required. It is generally used with some type of over-all feedback system because if over 6 to 10 db of feedback is applied by this method the voltage required from the driver stage becomes difficult to furnish without encountering appreciable distortion in the driver stage.

The screen grids of the output tubes may be tapped up on the primary of the output transformer in the Ultra-Linear connection. This connection slightly reduces the output power available from a given set of tubes for a specified voltage supply condition, but it also allows the screen voltage to be increased so that this loss in power output capabilities may be recovered. This system is also usually used with some over-all feedback system since only 5 or 6 db of feedback may be applied thereby, and because it does not reduce distortion by the same factor that it reduces gain. It is useful, however, to reduce the output impedance of the tubes at very high frequencies when phase shifts in other parts of the circuit reduce the effectiveness of the over-all feedback loop for this purpose. This connection also requires a special transformer and such transformers are now available at about the same price as equivalent transformers without the screen taps.

Over-all voltage feedback is customarily obtained from the plates of the output tubes or from the secondary of the output transformer. Although some commercial amplifiers have been produced which take the feedback voltages from the plates of the output tubes, such systems have very serious disadvantages. First they do not remedy any defects in the output transformer response or distortion characteristics. Second, they place severe requirements upon the power-supply filtering and the balance between the halves of the primary of the output transformer if they are not to increase the hum in the output of the amplifier. This situation results because such a system acts to reduce, at the plates of the output tubes, any signal that is not present in the amplifier ahead of where the feedback voltage is introduced. If there is any hum voltage present at the center tap of the output transformer, the feedback will act to reduce the amount of hum at the plates of the tubes to less than the amount at the center. Therefore, there will be hum currents flowing in the halves of the output-transformer primar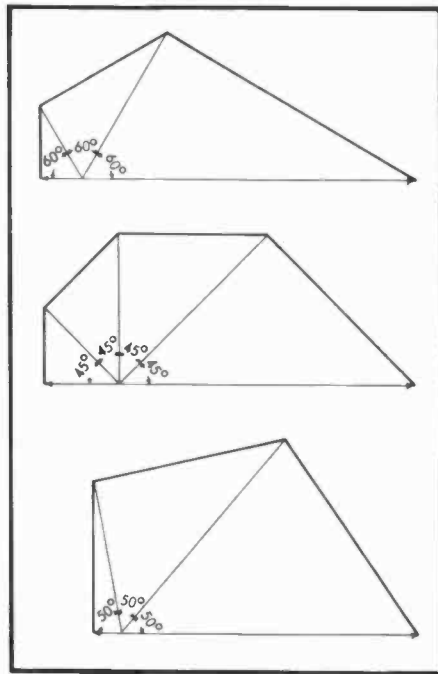y. These currents are out of phase but even so they will produce a voltage in the secondary of the transformer unless the currents are equal and the halves of the transformer primary are exactly balanced.

A special case of feedback is the connection of a network from the plate to the grid of a single stage. Such a system does not reduce the gain of the stage around which it is connected but reduces the gain of the previous stage by reducing the load resistance into which the previous stage works. Because such a system may offer an abnormally low load to the previous stage it may produce more distortion in the driver stage than it reduces in the output stage. In a high-quality amplifier it should not be used where the signal voltage exceeds a volt or two.

Finally we may consider the system where the feedback voltage is taken from the secondary of the output transformer. This has the advantage that in addition to reducing noise and distortion in the remainder of the amplifier it also reduces the distortion and the variation of frequency response caused by deficiencies of the output transformer. It has the disadvantage that the amount of feedback which can be applied may be more limited by the considerations of stability than is the case when the feedback voltage is taken from the primary of the output transformer.

Possibility of Oscillation

When we previously considered the feedback gain equation we were thinking of the mid-frequency situation where A is positive and B is negative thus making $-AB$ a positive number. At

Fig. 3. Circle diagrams for circuits having more than one time constant within the feedback loop.

frequency extremes A is no longer a positive real number. It becomes complex and may lie in any of the four quadrants. B may also change phase and magnitude with frequency. If AB becomes 1 or greater in magnitude and is a positive real number the amplifier will be subject to oscillation.

This situation may be avoided by insuring that the phase shifts around the entire feedback loop do not add up to 180 deg. until the quantity BA is less than one.

The difficulty of achieving the above requirement for stability will depend upon the design of the amplifier. At very low frequencies each RC coupling circuit may be considered as one time constant, and a resistor shunted by an inductance may also be considered one time constant. The output transformer may be considered as one time constant at low frequencies. At very high frequencies a resistor shunted by a capacitance is one time constant. At very high frequencies the output transformer is a much more complex device and if the speaker-system impedance increases so that the transformer may be considered to be operating unloaded it may be considered to be roughly the equivalent of two time constants.

Each time constant represents a phase shift which may reach a maximum of 90 deg. *Figure* 2 shows the circle diagram representing the action of one time constant. When the phase shift is 90 deg. the output voltage is zero so a feedback loop containing two time constants is stable because the gain approaches zero when the total phase shift approaches 180 deg. *Figure* 3 shows the results of having a larger number of time constants within the feedback loop. If we have three equal time constants within the loop the voltage gain around the loop will be reduced by a factor of 8, or 18 db, when the phase shift is 180 deg. In this case we could have about 12 db of inverse feedback and a safety margin of 6 db to allow for changes of gain due to aging of components and replacement of tubes. With four equal time constants the loss in voltage gain at 180 deg. phase shift is 4, or 12 db. This allows 6 db of feedback and 6 db for a safety factor. Since we may have as many as five time constants in an amplifier we must look for some solutions to the phase shift-gain problem if we are to apply the 20 db of feedback that we mentioned earlier.

Remedies

What remedies are available to reconcile the conflicting requirements of distortion reduction and of maintaining stability? The first and easiest method which we can adopt is to stagger the values of the time constants of the vari-

ous stages. Thus in the case of a three-time-constant circuit if we design so that one of the time constants is approaching 90 deg. when the other two are just past 45 deg., let us say one circuit at 80 deg. and two at 50 deg., we will have system which will allow the application of 20 db of feedback with almost 10 db of safety margin.

Since the Williamson type amplifier is a very popular type, let us study it in detail. An analysis of the circuit in *Fig.* 4 shows that it has three time constants at low frequencies—two *RC* coupling circuits and the output transformer; therefore we should be able to maintain low frequency stability with 20 db of feedback.

Because it is much easier to obtain response at low frequencies by the use of *RC* circuits than by means of transformers it is evident that we should have the longer time constants in the *RC* coupled stages. We should try to approach a 10-to-1 ratio between the time constant of at least one of the *RC* stages and the transformer primary inductance-plate to plate load combination. Such a long *RC* constant is difficult to obtain in the output stage of the Williamson since the output tubes are operated at a high plate dissipation which requires that the grid leak resistors be kept low in value to prevent the tubes from running away from the effects of ion currents or grid emission. The size of the coupling capacitors is similarly limited by the considerations of leakage and physical size. The time constant in the output grids is made about 1/40 sec. and the constant of the driver grid circuit is made about 1/8 sec. Exact calculations on the low-frequency characteristics of the amplifier are complicated by the fact that the primary inductance of the output transformer may change appreciably with a change of signal level; therefore, the time constant of the transformer will change with signal level.

At very high frequencies the problem of insuring stability is much more complicated because we have five time constants to deal with, two of which are tied to the characteristics of the output transformer. Even if it were easily possible there would be no advantage of making all the other time constants shorter than those of the transformer, since it is possible that the transformer alone could give 180 deg. phase shift before the gain around the feedback loop was reduced to one. It is therefore
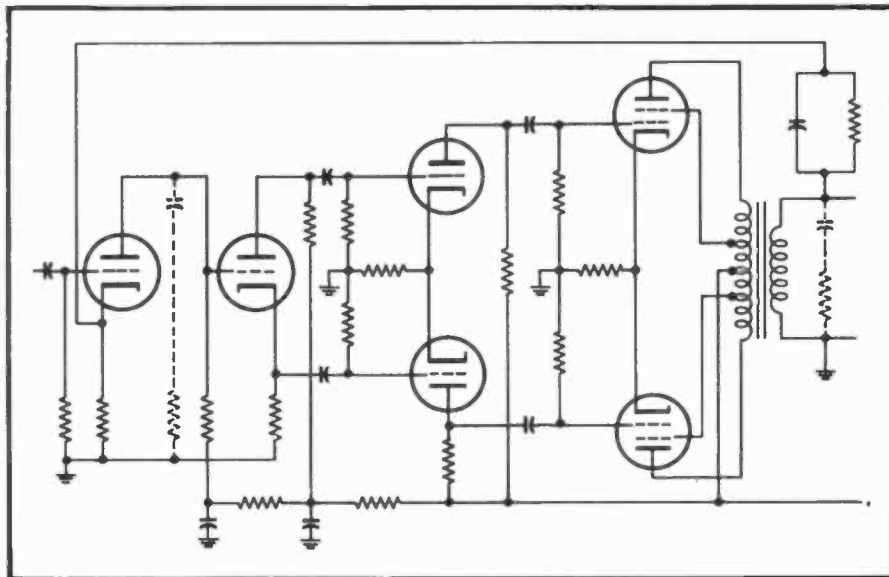

Fig. 4. Simplified schematic of a Williamson-type amplifier. Network shown dotted in plate circuit of first tube will change amplitude and phase response.

necessary to design the other stages so that the response of the whole circuit is down considerably before the resonant frequency of the transformer is reached. It is for this reason that it is necessary to have a transformer with good high-frequency response in order to obtain satisfactory operation in a feedback amplifier.

Although feedback will compensate for a great deal of loss within the band pass of an amplifier, an examination of the basic feedback gain equation will show that such a loss comes off the top of the amplification. That is, when the "raw" gain of the amplifier is reduced the over-all gain remains almost constant and the gain reduction due to feedback and with it the distortion reduction are diminished almost as much as the raw gain. For this reason the benefits of the feedback will be lost to about the same extent that the raw gain is lost. With such a limitation it is desirable to make

two of the *RC* time constants somewhat longer than those of the transformer and give special treatment to the other one. A step circuit connected from the plate of the input amplifier to ground, shown dotted in *Fig.* 4, can be added. Such a network will cause the first stage to have an amplitude and phase response as shown in *Fig.* 5. This response coupled with the response of the other two resistance-coupled stages can give a system that is just barely stable. A little margin of safety can be realized by making *B* a complex quantity with phase characteristics the opposite of those of *A*. This may be done by shunting all or part of the feedback resistor with a capacitor. In *Fig.* 4 all of the resistor is shown shunted.

Testing Procedures

Since very few individual experiment-

Fig. 5. Effect of network of Fig. 4 on amplitude and phase response.
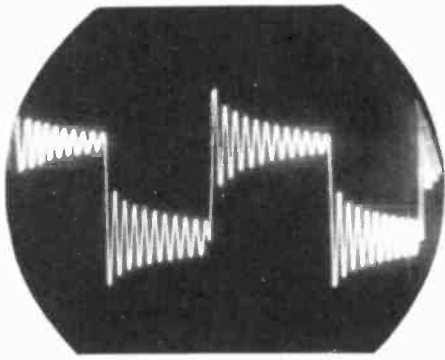

21
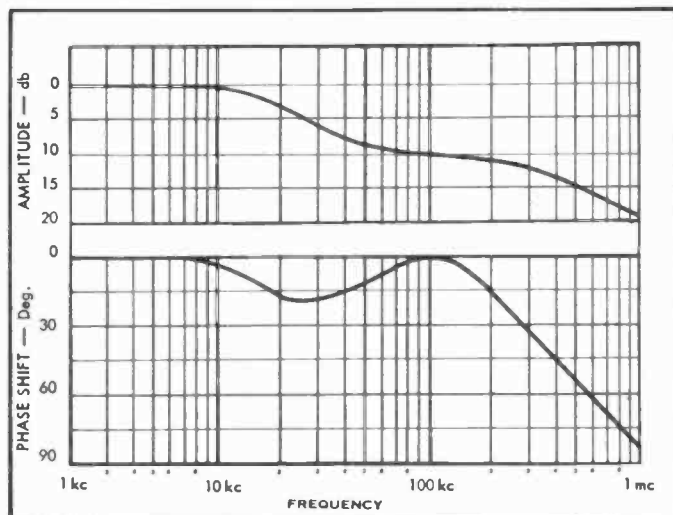
Fig. 6. 'Scope pattern showing ringing on 10,000-cps square wave.

ers or hobbyists who build audio amplifiers have phase meters or access to them and very few have audio oscillators with a range of 1 to 100,000 cps, the preceding information is of interest mainly for background purposes. With the minimum equipment which one should have available in order to adjust high-fidelity
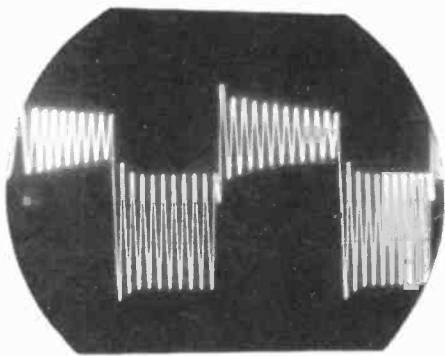

Fig. 7. When small capacitor is added across output circuit, oscillation increases.

systems the practical aspects of the stability problem can be worked out. This minimum of equipment consists of an audio oscillator which produces sine and square waves at frequencies up to 10,000 cps and an oscilloscope which will display these waveforms. Most oscilloscopes have response up to 100,000 cps and are thus adequate to reproduce a 10,000 cps square wave. An oscilloscope with a slow sweep speed of about two seconds is very useful for the investigation of the low-frequency response of the amplifier to transients, but essentially the same information can be obtained by watching a meter needle or a speaker cone when transients are fed into the amplifier. If only a sine wave oscillator is available a simple clipper can be built to change the sine waves to square waves.

Once an amplifier is finished it should be turned on with a resistance load connected to the terminals and the feedback loop disconnected. An oscilloscope should

be connected across the load and the audio oscillator connected to the input of the amplifier. The oscillator should be adjusted to furnish a small signal and the oscilloscope should be set up so that a trace of a convenient size is visible on the screen. Next the feedback network should be connected. If the amplitude of the trace on the 'scope screen is reduced you have everything hooked up correctly. If the amplifier goes into violent oscillation at some medium frequency it is necessary to reverse either the primary or the secondary leads of the output transformer. If instead of achieving a reduction of height of the trace when the transformer connections are correct you get very high frequency oscillations you have an amplifier which is unstable with the amount of feedback used and the amount of feedback should be reduced until steps are taken to increase the stability of the amplifier.

It may also happen that the amplifier is unstable at low frequencies which will be evidenced by motorboating which may either be spontaneous or be dependent upon being initiated by some transient. In this case also the amount of feedback should be decreased until the amplifier is stable so that means of increasing the margin of stability can be explored.

Taking the high-frequency troubles first; after stability has been restored by decreasing feedback a 10,000-cps square wave should be applied to the input of the amplifier. The wave form shown on the oscilloscope will be likely to have the appearance of *Fig.* 6 which shows violent ringing on the top of the 10,000-cps square waves. A .005 μf capacitor connected across the load resistor gives the waveform shown in *Fig.* 7. The capacitor lowers the resonant frequency of the output and thereby reduces the stability of the amplifier. *Figure* 7 shows that the amplifier is almost in continuous oscillation. When the step circuit
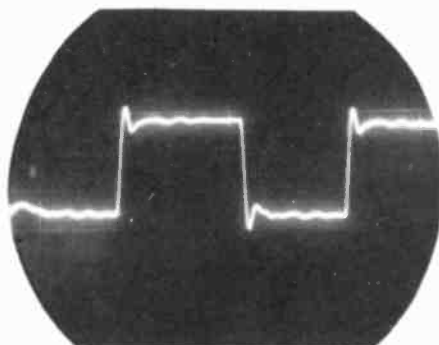

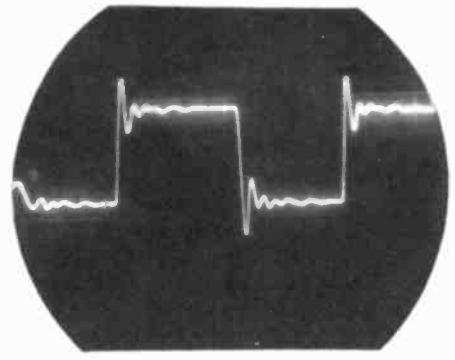Fig. 8. Addition of network of Fig. 4 reduces ringing appreciably.

Fig. 9. Capacitor across output emphasizes overshoot of amplifier, even with Fig. 4 network in place.

of 4700 ohms and .001μf is connected from the plate of the first tube to ground the waveform of *Fig.* 8 is produced. There is a slight overshoot on the leading edge of the wave. This overshoot is emphasized when a .05-μf capacitor is
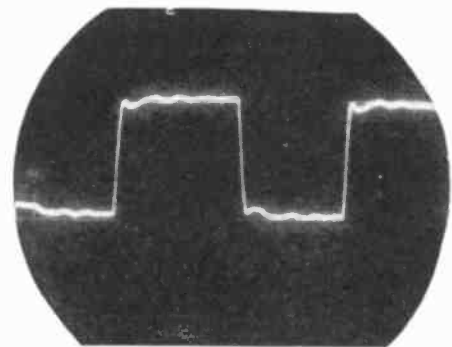

Fig. 10. Adding small capacitor across feedback resistor of amplifier changes pattern of Fig. 9 to this.

put across the load as shown in *Fig.* 9. A 150-μμf capacitor across the feedback resistor removes the overshoot as shown in *Fig.* 10. Capacitors up to .05μf make no appreciable difference when connected across the load.

The component values listed above may not be exactly correct in all cases, but they give a starting point and with most Williamson-type amplifiers with quality transformers the correct values will probably not be too much different from the ones listed. If the steps listed above do not cure your difficulties a 47-ohm resistor and 0.1-μf capacitor in series should be connected across the output terminals. This combination serves to load the transformer secondary at very high frequencies thus reducing the phase shift introduced by the trans-

former.

As the stability of the amplifier is increased the feedback may be increased until the desired amount has been reached. The margin of safety remaining may be estimated by connecting capacitors in the order of .002 to .02µf across the output terminals of the amplifier. If a capacitance of .005µf or greater across the terminals does not cause the amplifier to go into oscillation at some high frequency, the high-frequency stability is probably satisfactory.

It may be that, in the process of achieving high-frequency stability and increasing the feedback, low-frequency instability has appeared. Because most amplifiers have less potential phase shift and also because inferior transformers actually decrease the problem of attaining and maintaining low-frequency stability the low-frequency problem is not likely to be so acute. Generally the increase of coupling capacitors and grid leaks to the maximum desirable values will take care of the problems.

Figure 11 shows the effect of a transient upon an amplifier which is marginally stable. After more than a second the oscillations started by the transient have not nearly damped out. *Figure* 12 shows the improvement of low-frequency stability which was accomplished by increasing the time constant of the driver circuit grids and decreasing the time constant of the input circuit which is outside the feedback loop. *Figures* 13 and 14 show the improvement in overload recovery which were accomplished by the same changes. Once a stage within the feedback loop is driven beyond its dynamic range the feedback is no longer effective because there is little if any incremental amplification present. That is to say that additional input gives little or no additional output; therefore, if there is no gain there can be no gain reduction and consequently no distortion reduction. It is most desir-

able to prevent signals which will drive the amplifier beyond its dynamic range from reaching the amplifier input terminals. Despite much talk to the contrary there is little likelihood that the program material played through a high-fidelity system will include such signals since the program material has already been limited in amplitude and frequency range by the previous systems through which it has been processed. Both disc and tape recording systems have such limitations and, although a frequency modulation transmitter may have excellent transient and frequency response, it is likely that most of the program material will have been through some line or program amplifiers which have a response characteristic which is not better than that of the home equipment.

Speaker Distortion

It is possible that some of the program material will be beyond the capabilities of the speaker system to handle. It is most desirable to eliminate these signals before they reach the speaker since a speaker driven beyond its linear limits is a copious source of intermodulation. The limitation of the low-frequency response of a system can best be accomplished by installing a high-pass filter between the tone control amplifier and the output amplifier. This filter should cut off at a frequency no lower than 20 cps and preferably higher if the speaker system does not have an exceptional low-frequency response. Such a filter not only prevents program material which the speaker cannot handle from reaching the speaker, but it also prevents transients which may result from switching or from interference from overloading the amplifier. It also increases low-frequency stability in cases where the tone-control amplifier gets its plate power from the output amplifier.
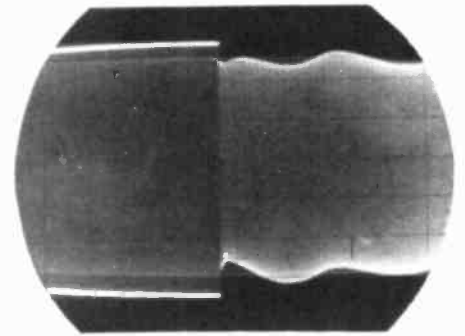

Fig. 13. Overload recovery of amplifier in condition shown in Fig. 11.
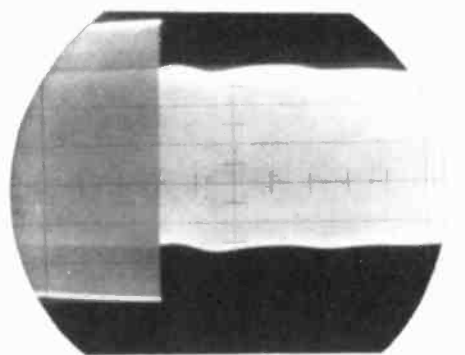
As a final check of stability the amplifier should be operated with each of the output tubes removed alternately to see if oscillation ensues. While one of the tubes is removed the amplifier should be driven to saturation at some

Fig. 14. Overload recovery is improved when changes are made as indicated in Fig. 12.

low frequency to see whether or not little bursts of high-frequency oscillation occur at some time during the low frequency cycle. As an acid test on my own amplifiers I repeat this test with the load removed, however anyone who does this should bear in mind that he is risking the output transformer should some high-amplitude oscillation result.

Although there are simpler amplifiers which will produce sufficient high-fidelity audio power to fill a living room, the Williamson amplifier or the circuits derived from it will give results which cannot easily be excelled. If your Williamson sounds bad it might be a good idea to check on its stability because there must be a great many of them in the condition of the amplifiers from which I made the "before" oscillograms. With just a little work they can be made as good as the amplifiers from which the "after" oscillograms were taken.   •
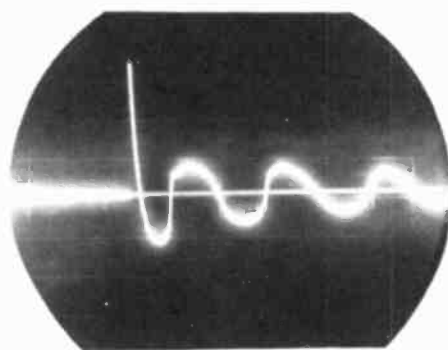

Fig. 11. Effect of transient on an amplifier which is marginally stable.
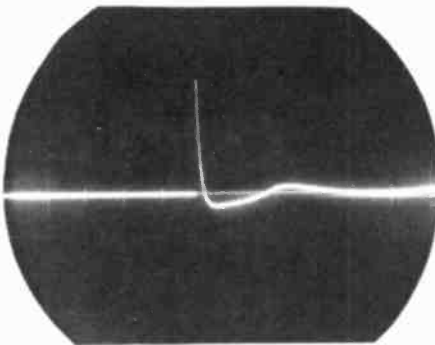
Fig. 12. Improvement over pattern of Fig. 11 is caused by changing time constant of driver grid circuit.

High-Quality
Dual Channel Amplifier

Cdr. CHARLES W. HARRISON, Jr.

A qualitative description of a preamplifier, high impedance R-C dividing network, and power amplifier that are intrinsically simple—yet capable of great performance.

A HIGH-QUALITY AMPLIFIER must be capable of passing rigid laboratory measurements, meet all listening requirements, and be simple and straightforward in design in the interest of minimizing performance degradation and eventual maintenance difficulties.

The circuits of the preamplifier, high-impedance dividing network, and power amplifier described in this paper are not fundamentally new; they represent a synthesis of well-known component circuits of recognized excellence.

In general, the playback system was evolved a "block" at a time after extensive experimentation and listening tests. Each unit had to "test" well, i.e., possess appropriate frequency response, adequate voltage or power output, low distortion and hum level, and then "sound" right when used as an integral part of the sound system. Any unit not meeting these criteria was rejected.

The preamplifier shown schematically in *Fig.* 1, consists of a type 6J7 input tube, followed by two type 6SN7 tubes.

"Local" feedback is effective in all stages except the first; however, it is to be observed that the feedback loop never encompasses more than two stages. Unconditional stability, low output impedance and the minimization of distortion is thereby assured. The type 6J7 input tube was selected because it is reliable and quiet in operation. It does not generate periodic "frying" noises and the hum level output is acceptably low. In addition the tube fits a standard octal socket having lugs of sufficient mechanical strength to support one end of a resistor or capacitor. The first stage serves exclusively as a voltage amplifier. No equalization is accomplished. It has been the writer's experience that most preamplifiers featuring a frequency-selective feedback circuit for equalization which connects to the cathode end of the bias resistor of the input tube generate an intolerable hum in any reproducing system capable of good bass response. This statement is sometimes true even when complicated d.c. heater supplies are employed. It appears mandatory that one employ a large bypass capacitor across the bias resistor. Preamplifiers utilizing

the method of "contact bias" are rejected because of the excessive intermodulation distortion developed in such circuits. (This bias method permits the direct grounding of the input tube cathode.) The distortion in the 6J7 stage is low even without feedback because the signal voltages rarely exceed 100 mv rms. If desired, the low-distortion input amplifier stage described later may be used, provided the entire bias resistor is heavily bypassed and the volume control is replaced by a resistor matching the pickup impedance.

Frequency correction of 6 db per octave below approximately 500 cps is accomplished by the passive R-C circuit shown between the 6J7 and first triode of the following 6SN7. The second triode furnishes some amplification and permits the application of negative feedback around the two stages associated with this tube. Following the volume control, a 36 position R-C equalizer appears. The maximum bass rise or cut is 12 to 15 db. At high frequencies the available rise is 3 to 5 db, and the cut is approximately 12 db. No interaction exists between the bass and treble sections of the equalizer.
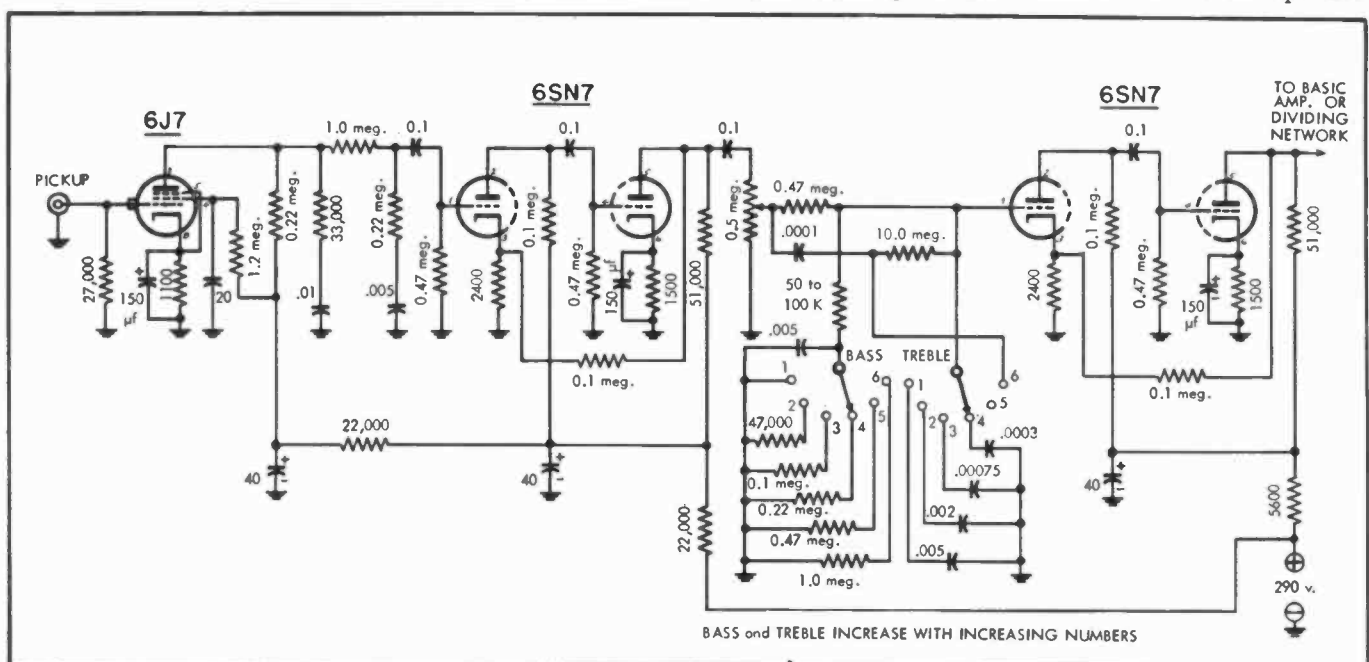

Fig. 1. Schematic of the preamplifier described by the author.

The resistor marked 50,000–100,000 should be selected on the basis of bass equalization required. Bass progressively increases as its value is reduced. The equalizer is followed by a two-stage amplifier, using a second 6SN7. Voltage-controlled feedback is applied around these two stages to minimize distortion and yield low output impedance. A cathode bypass capacitor is used in the output stage to eliminate degeneration at this point which would tend to raise the output impedance. If desired a cathode-follower output stage may be added to this preamplifier provided the power amplifier to be used in not high gain; otherwise hum problems are sure to be encountered. If feedback is not required around the first half of the 6SN7, the second half may be wired as a cathode follower. With slight circuit redesign, type 12AY7 low-noise dual triodes could be used in lieu of the 6SN7 tubes. If an FM tuner input is required, a two-position shorting-type switch should be installed adjacent to the volume control on the left.

The hum level of this preamplifier is extremely low. From experience the author can report that nothing is gained in this respect by the employment of d.c. on the tube heaters. It has been found that less than one-third of the equalizer positions available are needed in practice to compensate for the various recording characteristics in use.

This preamplifier does not feature built-in AES, NAB, RIAA (new orthophonic) response characteristics. The philosophy of precise preamplifier equalization which fails to take into account the frequency response of the pickup, power amplifier, and speaker to be employed is a mystery to the writer. System, rather than component engineering is required. As a practical example, suppose that the AES playback characteristic is specified for a given recording and that this response curve is built into the preamplifier. Excellent results will undoubtedly be obtained provided the pickup, power amplifier, and speaker are flat. Now let it be assumed that the speaker (high-frequency driver in a dual loudspeaker) is down 12.5 db at 12,000 cps, which is not at all unusual. The program material in this frequency region is now attenuated by some 12.5 db more than required by the AES playback curve. Percussion instruments will appear to be in the background, a condition not acceptable to a person who enjoys "high-highs." To approximate the AES play-back curve for a given sound system may actually require a preamplifier having essentially flat response above 500 cps; the high-frequency pre-emphasis used in recording being more or less offset by the high-frequency rolloff of the loudspeaker and pickup being employed. In addition to the factors mentioned above affecting preamplifier equalization, the influence of listening room acoustics must be given due weight.

A dual-channel amplifier has several advantages over a single amplifier for driving a dual loudspeaker. The use of a distortion-producing dividing network at high signal levels is avoided, as is the power-consuming attenuator normally required in the high-frequency channel to obtain bass and treble balance. The divided transmission system permits exact impedance matching between amplifiers and speakers and additionally permits one to obtain optimum generator impedance in driving the bass and treble speakers. This is generally impossible when a dividing network is interposed between an amplifier and dual loudspeaker. This scheme is a good way of achieving linear transmission of low frequencies (such as emanate from drums, gun shots, explosions, and thunder) together with linear transmission of high frequencies (such as emanate from triangles, castanets, cymbals, and tambourines), without severe modulation of high frequencies by the low frequencies.

The circuit diagram of an R-C dividing network employing cathode follower input and output stages is shown in *Fig 2*. Two type 12AY7 tubes are used; one in each channel. The values of capacitors and resistors shown result in an 800-cps crossover. If, for example, a crossover frequency of 500 cps is desired, the values of the filter capacitors should be multiplied by the ratio 800/500. The resistors do not change value. Similarly, multiplying the capacitor values by 800/1500 yields a crossover frequency of 1500 cps. Each R-C section of both filters should be adjusted to be down 1 db at 800 cps (for 800-cps crossover) by padding the appropriate capacitor and resistor so that the total attenuation for all three sections in cascade is 3 db. The low-frequency filter provides an attenuation approaching 18 db per octave above the crossover frequency, and the high-frequency filter provides an attenuation approaching 18 db per octave below the crossover frequency. By actual measurement on the R-C dividing network constructed by the author, the low-frequency filter is down 11 db at 1600 cps and the treble filter is down 11 db at 400 cps. Thus the attenuation afforded by the three-section R-C filters is 11 db in the first octave, the crossover frequency being taken as reference. The input impedance of the dividing network is extremely high. The output impedance of each channel is low, permitting the use of rather long cables to the bass and treble power amplifiers without deleterious effect on the high frequencies. Ten volts rms will not over-drive the dividing network. If the network is used in conjunction with power amplifiers like the one to be described in the following section the operating level need not exceed one-half volt. Thus essentially distortionless operation is assured.
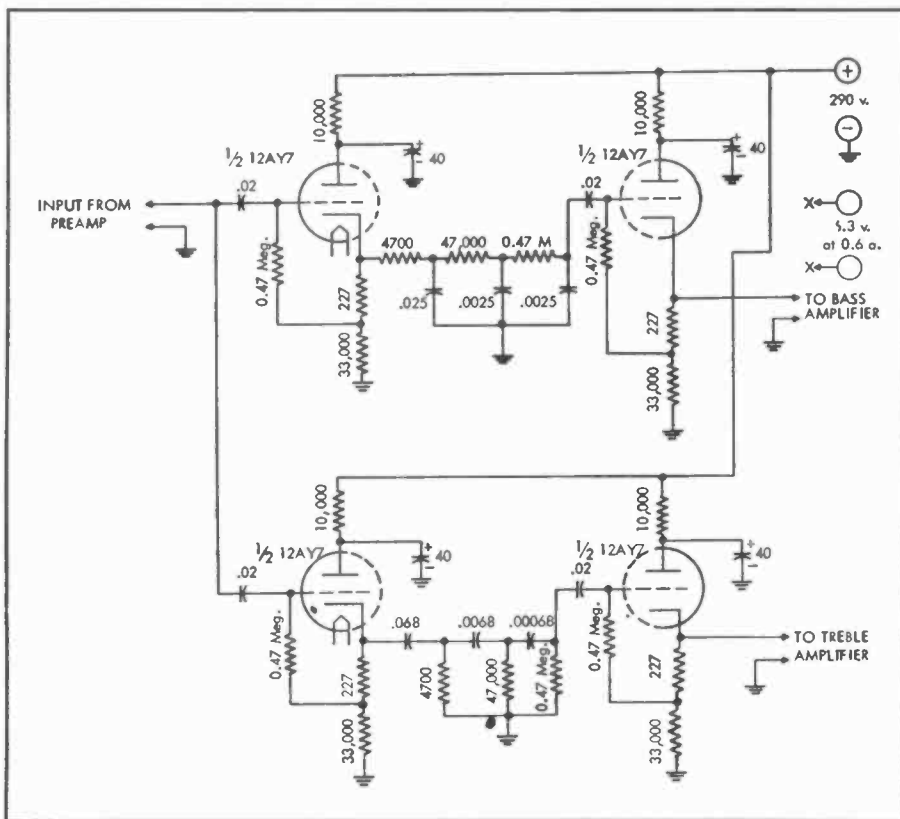

Fig. 2 Schematic of the high-impedance R-C dividing network between the amplifier and the inputs to the two power amplifiers.
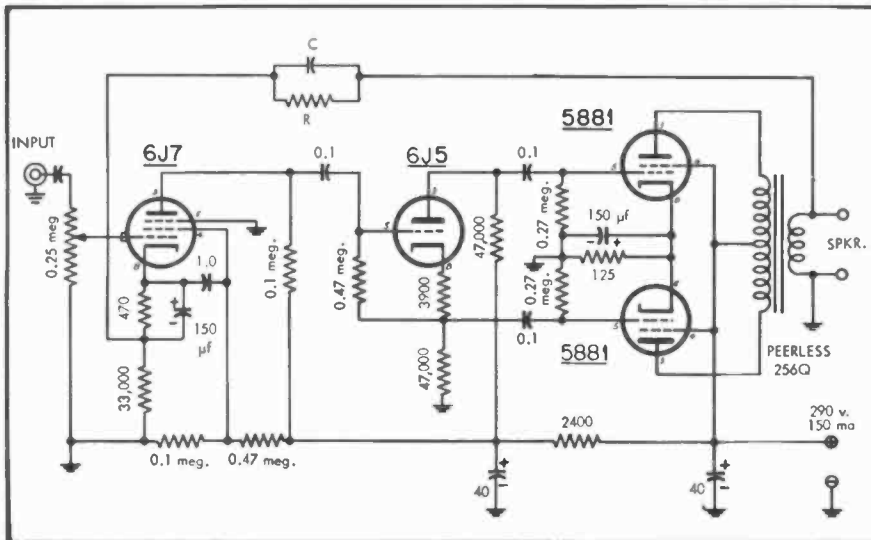
Fig. 3. Schematic of the power amplifier. Values of R and C in the feedback circuit are discussed in the text.

A schematic of the basic or power amplifier is shown in *Fig.* 3. The tubes employed are 1-6J7, 1-6J5 and 2-5881. Using tubes selected at random the amplifier is capable of delivering 10 watts at under 1 per cent intermodulation distortion; 12 watts at under 3 per cent. An 18-watt power output is available over the frequency range 20 cps to 140 kcs (by appropriate adjustment of the input voltage) without visible wave form distortion (estimated at under 3 per cent harmonic distortion). The amplifier is absolutely flat at 12 watts output from below 20 cps to 55 kcs for constant-voltage input, dropping to -2 db at 125 kcs; -5 db at 175 kcs and -6.5 db at 200 kcs. One half volt rms will drive the unit to full power output. It will deliver 12 watts for 0.38 volts rms drive. These performance data are based on the use of 10 db feedback.

The component values, i.e., resistors and capacitors associated with the 6J7 voltage amplifier, were selected to minimize intermodulation distortion. It was found desirable to use a voltage divider to obtain screen voltage and to bypass the screen to the cathode of the tube. The bias resistor is almost entirely bypassed; only a small portion of the total resistance being left unbypassed for the application of negative feedback.

The phase splitter, employing a 6J5, is an excellent method of coupling a single-ended plate circuit to a push-pull grid circuit. (A phase splitter, as well as a cathode follower, is defined for later usage as "one-half" stage.) This circuit is self-balancing, and distortion is low. Any unbalance effects at high frequencies are generally negligible.

The output stage features the use of a Peerless type 256Q 20–20 plus transformer. Note that the bias resistor for the push-pull type 5881 tubes has a value of 125 ohms. The 5881 tube is similar to Western Electric type 350B and are interchangeable. Both have "power filaments" in that 1.8 amperes at 6.3 volts is required for cathode heating.

Feedback is applied around the "2.5" stages; the required voltage being taken from the secondary winding of the output transformer. The values of R and C in the feedback circuit must be selected by test. The value of R controls the amount of feedback (usually expressed in db), and C controls the high-frequency ringing, i.e., for the purpose of damping out any small oscillations that may appear on the leading edge of a square wave. The equipment needed to determine the proper value of R and optimum value of C is: a vacuum tube voltmeter and a sine and square wave generator. It is customary to load the amplifier by a resistor equal to the nominal load impedance of the amplifier when choosing the correct values of R and C, rather than use the loudspeaker as load. Optimum generator impedance can be obtained by varying the value of R in the feedback path and conducting simultaneous listening tests. As R is increased the value of feedback is decreased.

This power amplifier is basically simple and utilizes the minimum number of stages required to do the job effectively. Although feedback is applied around "2.5" stages the amplifier is stable with feedback values up to at least 30 db. Many of the popular circuits of today feature the application of large values of feedback around "3.5" to 4 stages. This is an invitation to serious trouble. Marginal stability obtains and at some signal levels violent subsonic and supersonic oscillations may be generated. Even though these frequencies may not be heard, i.e., they fall outside the audio spectrum, the power delivering capability of the amplifier is largely consumed. Thus little "clean" power is available in the frequency range of interest. This principle is too frequently overlooked in practice. The power amplifier will deliver a clean signal over its entire frequency range even without feedback. This is not true of one well known circuit which utilizes 20 db of feedback. A sine wave input at 60 kc is likely to appear at the load terminals as a series of triangles!

The writer is of the opinion that an otherwise essentially distortionless amplifier does not require the application of large values of feedback. The use of 20, 40 or 90 db feedback is nonsense. Values of 10 to 15 db voltage-control feedback are adequate for two important reasons:

(a) *Instability tendencies are reduced.*
(b) *The experimentally observed bass loss in the frequency region of speaker resonance is minimized.*

It is interesting to note that the designers of theater sound equipment restrict the use of feedback to the 10 or 15 db level.

There may be protests to the effect that the equipment described in this article is not an "all triode" playback system. It would seem meaningless to insist on the exclusive use of triodes in amplifiers until records are available bearing the label "We guarantee all electronic equipments used in making this recording were fitted throughout with triode vacuum tubes." Note also that AM, FM, and TV stations will never measure up to the standards of the perfectionist who insists on the utilization of triode vacuum tubes in every tube application.
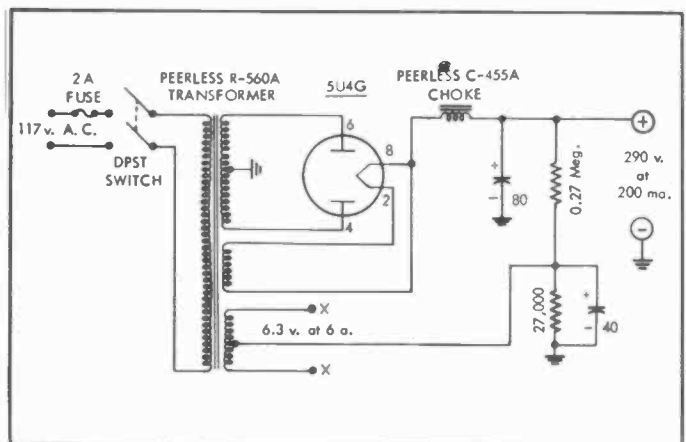

Fig. 4. Power supply schematic. Note that it is of conventional design.
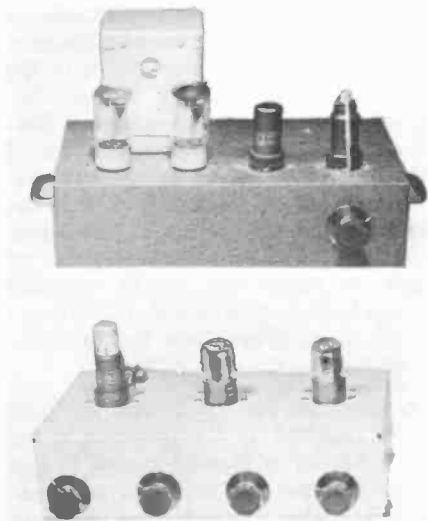
Fig. 5. Above, the preamplifier; below, the power amplifier. "Building block" construction makes for flexibility.
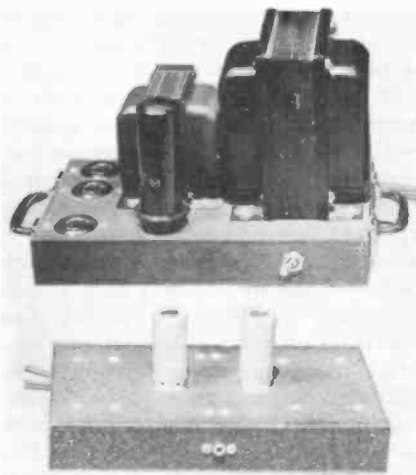
Fig. 6. Above, the power supply is a simple and neat construction; below, the dividing network chassis.

The power supply illustrated schematically in *Fig.* 4 is entirely conventional. It delivers 290 volts d.c. at 200 ma and 6.3 volts a.c. at 6 a. To minimize hum in the playback system, the heater winding is operated at a positive potential of about 29 volts, the center tap of the winding being heavily bypassed to ground. Although often omitted from commercial equipment, the bypass capacitor is a circuit element vital to the successful operation of this hum reduction scheme. Because of the relatively low d.c. voltages required for operation of the preamplifier, R-C dividing network and power amplifier, one may expect that 450-volt electrolytic filter capacitors, if used throughout the equipment, will have exceptionally long life.

The writer believes in building equipment with the best parts available. All coupling capacitors should be rated at 600 volts, and if 0.1 μf and less in capacitance should have a leakage resistance of at least 1500 megohms. The bass
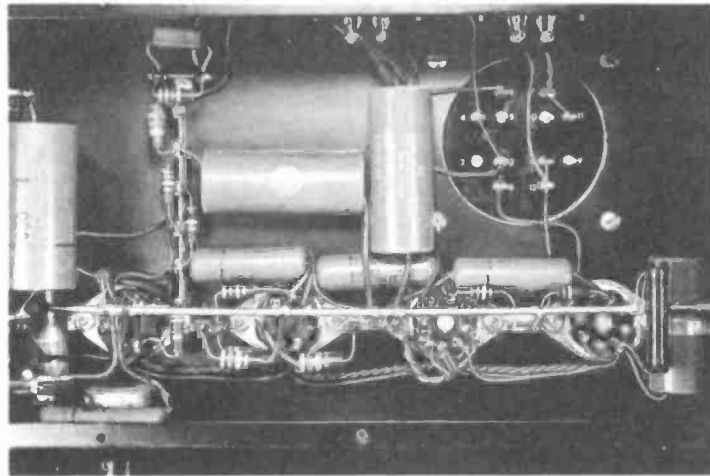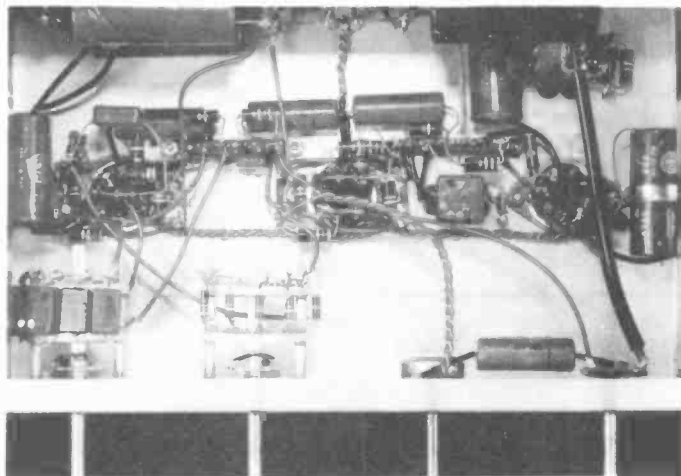

Fig. 7. Above, left, bottom view of the preamplifier with the base plate removed to show layout of parts and wiring. Fig. 8. Above, right, bottom view of power amplifier with base plate removed.
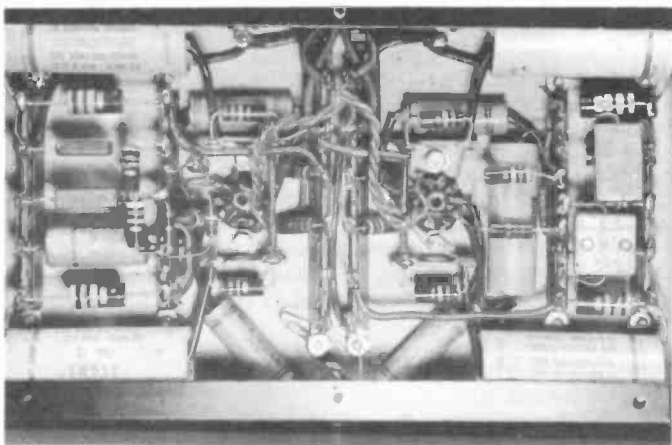

Fig. 9. Underside of dividing network chassis.

and treble controls in the preamplifier should be of the shorting type and feature silver contacts and steatite insulation. Capacitors used in the equalization circuits should be 5-per cent tolerance silver micas (except possibly in the largest sizes). Resistors in these circuits should be within 5 per cent of specified values, or better. Very precise values of resistance and capacitance are required in the filters of the dividing network. In the push-pull portion of the power amplifier the capacitors and resistors used should be selected for balance. The most reliable volume controls that can be obtained should be used, in log-taper form. In general, resistors rated at 1 watt dissipation are adequate, except in the following instances: The 33,000-ohm resistors in the dividing network are 2 watt types as is the 2400-ohm resistor in the power amplifier, and the 125-ohm bias

resistor in the power amplifier is a 10 watt type. It is a good idea to use wire-wound plate-load resistors in the low-signal-level stages. Although a Peerless transformer and choke are not essential circuit elements in the power supply, the writer built and tested the amplifier with a Peerless 256Q output transformer, and has no data on how another transformer type may operate in this circuit! Small broadcast-type connectors may be used conveniently for inter-chassis connections where audio voltages are involved. Four-conductor cables terminating in male plugs are useful for power connections and the chassis connectors being standard four-hole sockets.

The writer's present dual-channel playback system consists of a turntable, pickup and arm, a preamplifier (*Fig.* 1), an R-C dividing network (*Fig.* 2), two identical power amplifiers (*Fig.* 3), two power supplies (*Fig.* 4), and the dual loudspeaker described in an earlier paper.[1] The equipment corresponding to each schematic presented here was constructed on separate chassis as shown photographically in *Fig.* 5 and 6. This building-block technique was employed so that new innovations may be checked with minimum constructional labor. The preamplifier was built on an aluminum chassis having dimensions of $7 \times 12 \times 3$

inches. The arangement of parts may be seen in *Fig.* 7. If Vector socket-turrets are used the circuit can be built in a $5 \times 10 \times 3$ inch base. The employment of aluminum material that is not painted permits one to make the numerous low-resistance ground connections required by the circuit configuration. It is very important to keep ground leads short in high gain circuits. The dividing network and power supply fit nicely on a chassis measuring $5\frac{1}{2} \times 9\frac{1}{2} \times 1\frac{1}{2}$ inches. The orientation of parts in the dividing network is shown in *Fig.* 8. The power amplifier can be built on a $7 \times 12 \times 3$ inch black-crackle finish steel chassis with room to spare. A bottom view of this unit appears in *Fig.* 9. Since high signal levels obtain in this circuit a ground bus may be used (grounded to the chassis at each end) without development of hum difficulties. The writer finds the use of a ground bus a constructional advantage.

At present four parallel-connected bass drivers are in use in the bass section of the speaker described in reference 1. The driving-point impedance of the array is 2 ohms. Accordingly, the secondary of the output transformer in the bass amplifier is connected for this load,

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[1] Charles W. Harrison, Jr., "Coupled Loudspeakers," Audio Engineering, Vol. 37, No. 5, pp. 21, May, 1953.

and *R* and *C* selected for 10 db feedback and minimization of ringing, respectively. The value required for *R* is 330 ohms, and for *C*. 0.005 μf. The high-frequency driver in use is a Western Electric type 594A having an impedance of 24 ohms. A resistor of 48 ohms is connected across the voice coil so that the impedance of the parallel combination is 16 ohms. The output transformer of the treble power amplifier is connected for this load. In this case *R* is 1500 ohms for 10 db feedback, and the optimum value of $C = 1360$ μf[2].

A schematic for the 24 volt d.c. field supply required for the operation of the WE 594A driver is not included in this article because this driver is not generally available. Several other makes of high-frequency reproducers are available with permanent magnet fields, however.

Acknowledgment

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[2] For a load of 4 ohms, *R* is 1000 ohms and *C* is 1500 μμf.

BUILDING SIMPLICITY

control, and to re-route the wire which carries the output of the cathode-follower directly to the "output to tape recorder" jack. One of the "AUX" input jacks may have its wire disconnected from the selector switch, and connected directly to the "loudness" control. New labels should be made, showing this last jack as the "tone control input" or such other language as suits the user. With these wiring changes, the cathode-follower output of the selector switch is connected directly to the input of the tape recorder, whose output is then connected directly to the volume control, tone controls, power amplifier, and loudspeaker. The recorder may then be turned on whenever the system is in use, so as to feed all signals through to the amplifier and loudspeaker, or it may be equipped with a switch which connects its input to its own output whenever the recorder is turned off, so as to be effectively out of the circuit except when in use.

A careful examination of the schematic diagram of any equipment to which a tape recorder is to be connected will usually reveal a point in the circuit where level is high, and not subject to

variation with the volume control setting, and a point which has not been subjected to tone control. Usually, this point occurs immediately after the selector switch. If a cathode-follower happens to be in the circuit immediately following the selector switch, so much the better—the output to the tape recorder may then satisfactorily be connected here, rather than directly to the selector switch, provided the volume control was not placed earlier in the circuit than the point selected for the output to the tape recorder. Usually, this configuration will result in a system which can "howl" if the monitor selector on the tape recorder is set at "input" when the control box selector is set for "tape," but the system will be capable of making high-quality tapes, with adequate level, and with flat frequency response. "Howl" can be positively eliminated only by adopting the "series" type of connection for the tape recorder.

Stereo Monaural Companion Amplifier for the "Preamp with Presence"

LOUIS BOURGET

Built around a new phase-splitter circuit, this stereo amplifier will provide adequate power for the average installation with better than average performance throughout the entire audio spectrum. The phase splitter itself is worthy of notice, also.

MOST OF US have heard the impressive stereo tape demonstrations given at virtually every audio exhibit held within the last two years. Those of us who have long dreamed of owning a *really complete music system* may now take heart. The cost is relatively modest for the features provided and the entire system may be assembled progressively so that you may start with any existing source—say a phonograph player or AM/FM tuner—and eventually have a complete high fidelity system accommodating the following input sources:

1. Phonograph
2. AM/FM tuner
3. Monaural tape
4. Stereo tape
5. AM tuner—for stereo AM/FM broadcasts in metropolitan areas.

The stereo amplifier with dual speakers or speaker systems is also very effective when used in parallel from a single preamp for "spreading out" the sound source. The music appears to emanate from an area centered between the two speakers when they are in phase and operated at the same intensity level. The bass range is considerably improved —from better speaker coupling to the room air and an apparent filling in of room nodes.

Once you become accustomed to the versatility and superior sound distribution of a twin-channel system it is unlikely that you will settle for less.

The dual amplifier, shown in *Fig.* 1 with its power supply and the preamps, was designed to operate from two McProud preamplifiers ("Miniaturized Preamp with Presence")[1] which have been modified for playback of commercially recorded tapes. The amplifier incorporates some features usually found
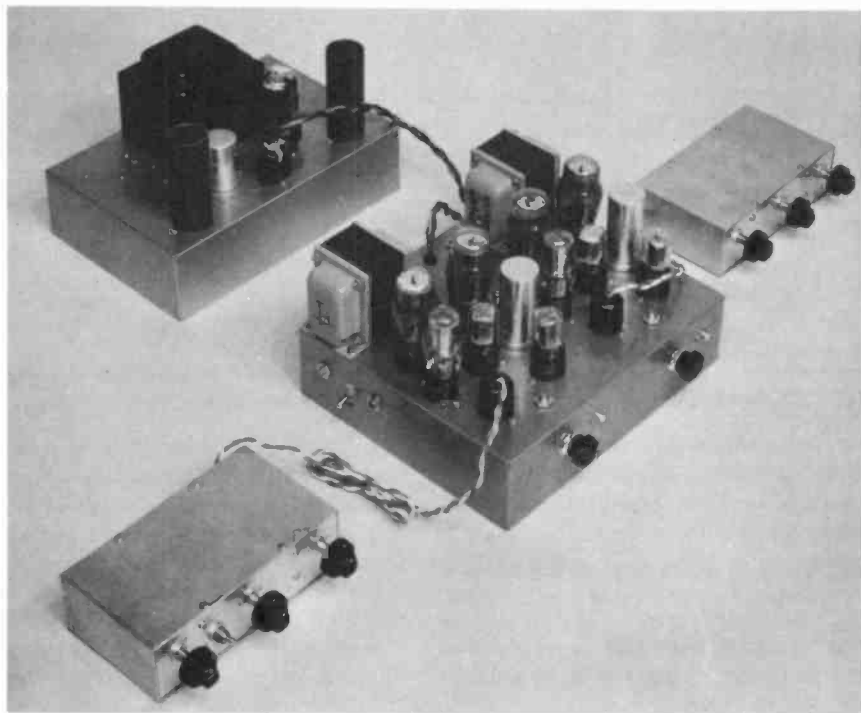
[1] AUDIO, *May*, 1955.

Fig. 1. The complete amplifier system, consisting of the stereo dual-channel power amplifier, two Miniaturized Preamps with Presence, and the power supply.

only in laboratory type equipment:

1. The output tubes are balanced for both dynamic and static conditions, sustaining full power delivery at low frequencies.
2. The phase splitter is balanced for both dynamic and static conditions. It is not frequency conscious. When balance is made at any audio frequency it will be correct for a frequency range wider than the audio spectrum.
3. Hum is of such low order that it becomes difficult to measure with accuracy.
4. Thermal hiss is low enough to permit the source material and preamp to act as the dominant influence without later stages causing masking of subtle high frequency detail.
5. The tube types used are moderate in cost and are operated under conditions which should give reasonably long life expectancy.

The amplifier power output (per channel) is based on the power requirements of the majority of existing speakers (or speaker systems) used in the home, to the extent that the speaker will reach excessive distortion limits ahead of the amplifier. Power beyond this requirement could well be a waste of money. To determine this power in watts for each half of the double amplifier led us to test a quantity of loudspeakers.

This study was made over a period of six months and included everything from eight-inch speakers to large three-way systems using fifteen-inch speakers for the woofer section. All of the speakers tested could be driven to excessive distortion levels before the amplifier capability was exceeded. It is interest-
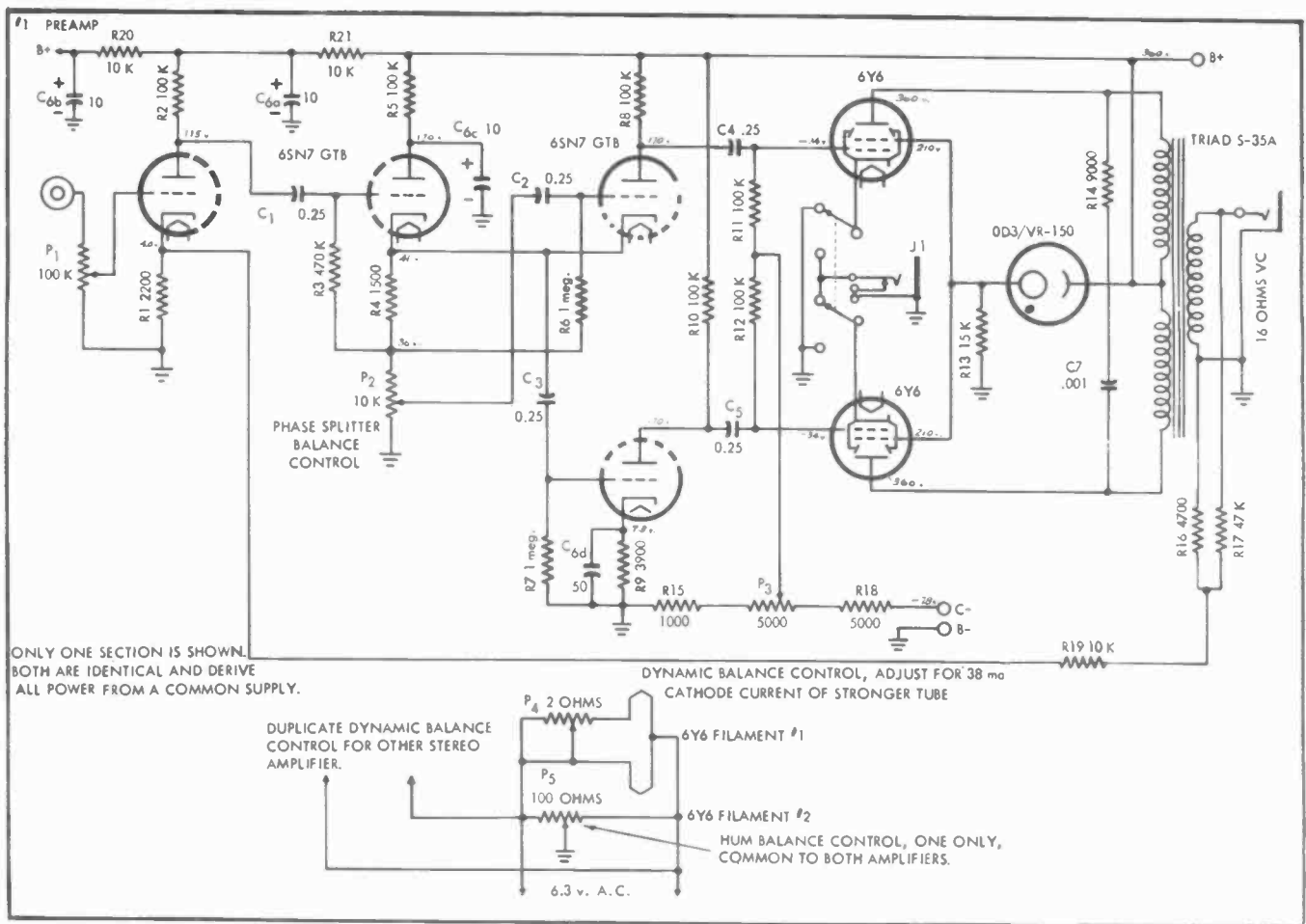
Fig. 2. Over-all schematic of one of the dual-channel amplifiers. Both sections are identical, and are built on a single chassis.

ing to note that ten *clean* watts when available down to 30 cps proved entirely adequate to reach one or more of the following conditions:

1. Limiting distortion of the speaker suspension system at low frequencies.
2. Excessive distortion with spurious frequency generation at the middle and high frequencies.
3. Loudness levels judged intolerable for home use.

The maximum power rating given by most speaker manufacturers is not intended as their recommended operating condition, but may preferably be interpreted in most cases as the danger level for the speaker mechanism. Fortunately the human hearing tolerance level is usually exceeded first except at extremely low frequencies and we are not often tempted to damage expensive reproducers.

To determine the power level at which suspension limiting takes place is fairly simple and makes use of a device frequently employed in the test laboratory. It consists of a dummy resistive load (of adequate power rating) mounted in a box with a quick changeover switch from voice coil to dummy load. Jacks are also provided for a calibrated oscilloscope and an a.c.-VTVM for rms val-

ues. These are connected in parallel so that they remain across the amplifier output on either resistive-load or voice-coil position. Comparison of the levels at which peak limiting occurs provides the answer at low frequencies. At middle and high frequencies a simple technique is used. The ear is remarkably sensitive to the apparent change in pitch due to frequency doubling or halving when overload point is reached for the loudspeaker from a sine-wave input source. Protective ear plugs are desirable here, as you may otherwise exceed the "threshold of pain" and this is as unwise as welding without goggles. The human hearing apparatus is also operating in a more discriminating manner when subjected to sound intensities well below the maximum tolerance level by the use of ear protective plugs.

All of this may seem a little beside the point in leading up to a description of the dual amplifier but if it were omitted many people might wonder about our manner of drawing such conclusions.

In this dual amplifier either side will deliver 20 watts before clipping. The extra power above 10 watts per side allows for tube aging and is considered an economy in terms of useful tube life at normal listening levels.

Voltage Amplifier and Phase Splitter

The voltage amplifier and three-tube phase splitter employ the newer 6SN7GTB which has been much improved for TV and governmental equipment. The circuit is shown in schematic form in *Fig. 2.*

The first stage is conventional except for the 0.1-meg. input potentiometers which are deliberately made less than the 0.5-meg. input commonly used in many amplifiers. This prevents the con-
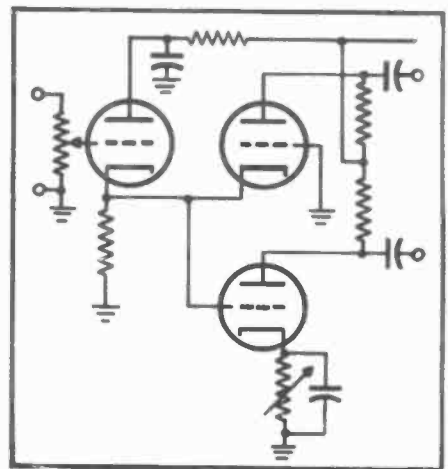

Fig. 3. Simplified schematic of the phase inverter circuit, which is capable of excellent balance throughout the entire audio spectrum.

trol from becoming a differentiating circuit at middle and lower settings due to the RC network, formed by stray capacitance from wiring and terminals, which normally cause spiking of square waves on many amplifiers. The "Preamp with Presence" (like all modern preamps) has low output impedance and no difficulties are posed.

The phase splitter is an improved variation of circuitry used by the writer since the late 1940's.[2] A cathode follower provides simultaneous audio signal voltage to the cathode of one driver and the grid of the opposing driver, as shown in the simplified schematic, *Fig.* 3. Note the use of a plate-voltage dropping resistor—well bypassed—in the cathode-follower plate circuit. This is important to establish operation of the follower on the same part of the dynamic characteristic as the phase opposed drivers. Also observe that the energized-grid- and cathode-circuit capacitances of the drivers (including strays) are effectively in parallel across the *low input impedance* of the cathode follower. This means that the shunt RC product is held to a small value and is virtually identical for the two drivers—a condition which makes it possible to obtain perfectly balanced driving voltages across *more* than the complete audio spectrum. The phase-splitter balance control is a 10,-000-ohm wire wound potentiometer in the cathode-to-ground circuit of the follower. Balance will be obtained with the arm set about 1000 ohms up from ground.

Balancing the Phase Splitter

If you have no test equipment available, the phase splitter may be balanced as follows: connect a temporary short lead from grid to grid of the 6Y6 final —pull out one of the 6Y6 tubes and balance for null on low level music. Remove the short and plug the 6Y6 back in its socket.

While much has been written about two-tube "self-balancing" phase splitters, some rather important defects are generally ignored or glossed over. The phase-inverted side causes the signal to go through one more tube than the "direct side." This usually leads to higher distortion and unequal phase rotation with attendant balance difficulties at frequency extremes. No one would be so optimistic as to expect high-quality performance from a final stage with one flat tube. Obviously the same thing applies to the drivers. When one tube is badly off in a pair of phase-opposed drivers, a self-balancing circuit only insures that the flat tube will be driven harder, with inevitable increase of distortion. The answer is simple. Tubes in

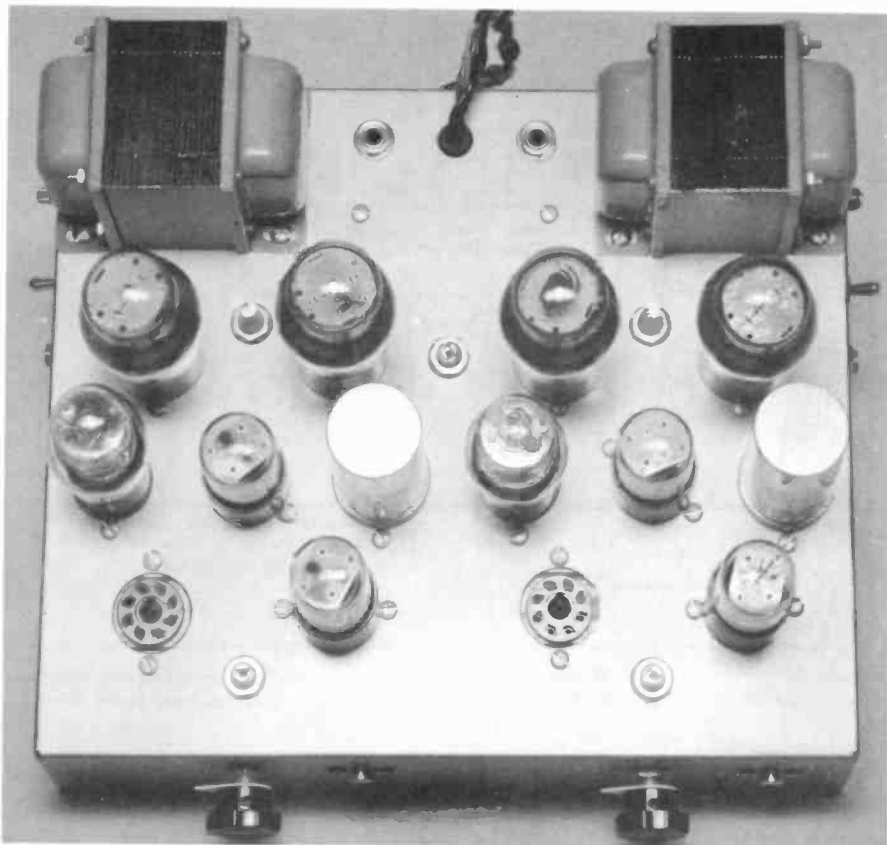[2] Patent 2,618,711 issued November 18, 1952.

Fig. 4. Top view of the amplifier chassis. The jacks at the rear are for the outputs, while power is fed to preamps from octal sockets toward the front.

high class equipment should be tested periodically and replaced when necessary. In this amplifier the use of the improved 6SN7GTB tubes at plate currents of only 1.3 ma in each triode section serves to insure long trouble-free performance from a circuit which may be set precisely for balanced operation.
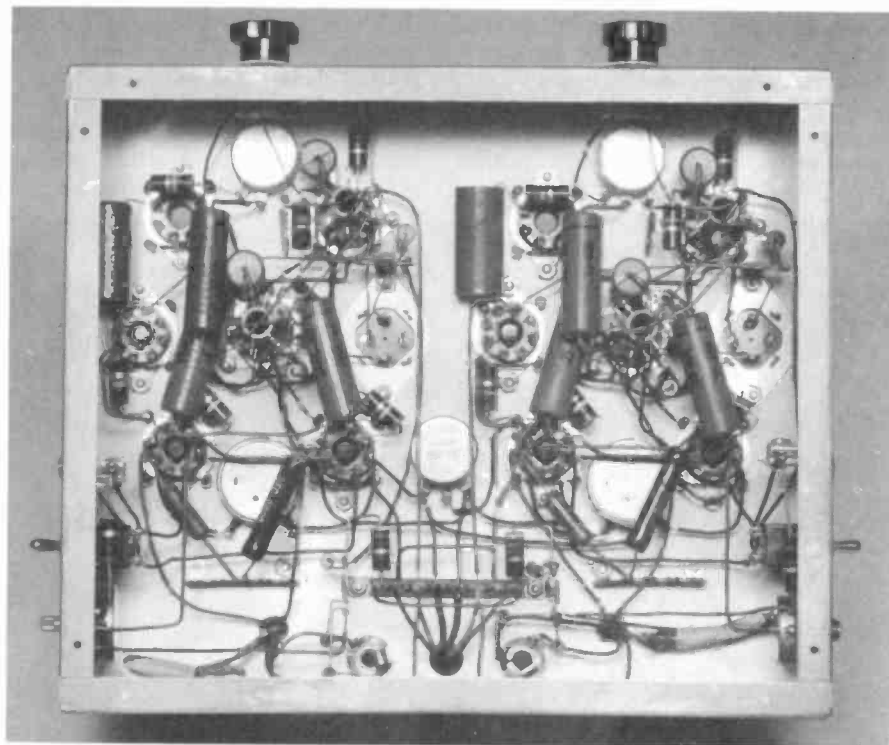

Fig. 5. Bottom view of amplifier chassis. Input controls are on front apron, with balance controls accessible from top of chassis just below them. Hum balance control is at center, individual bias controls are just above terminal strips. Filament emission balancing controls, plate current jacks, and toggle switch are on right and left aprons.
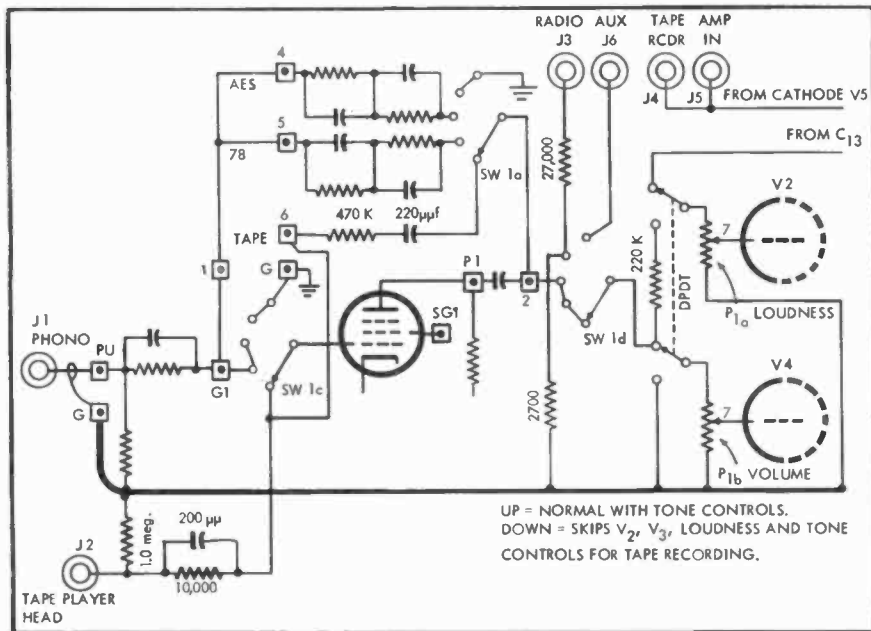
Fig. 6. Schematic of modifications to one of the preamps to permit direct connection from tape head as well as to eliminate tone and loudness controls from the circuit for recording.

The Output Stage

The choice of 6Y6 output tubes was made after testing many different beam tubes in both triode and pentode connection. These tubes have the advantage of high power output at moderate plate supply voltages. In addition, the optimum plate to plate load impedance is lower than for most beam tubes and permits better low-frequency performance from a given amount of iron and copper in the output transformer.

The 6Y6 output tubes are operated at 300 to 325 plate volts and an OD3/-VR150 gas tube is used as a *series* dropping device to maintain the screen voltages precisely 150 volts lower than the plate supply. The 15,000-ohm resistor from screen circuits to ground keeps

about 12 ma of gas-tube current flowing and stabilizes operation.

Improved low-frequency performance is assured by both dynamic and static balancing of the output stage. Most amplifiers—where balance adjustments are provided at all—permit balancing of only static cathode current values. This is usually arranged as either a variable cathode bias or grid bias circuit which permits reducing the plate current of the "high" tube to match the lower tube. Unfortunately this type of balancing generally leads to even poorer conditions of dynamic balance at medium and high level plate current excursions.

In the stereo/monaural amplifier, both output tubes in either push-pull pair have identical, fixed bias grid volt-

ages. A cathode current jack and switch permits comparing the cathode currents. The common bias control is adjusted to produce 38 ma of cathode current *for the lowest tube's plate current* with the filament control P_4 set for zero resistance. If the "high" tube is in the socket which has the filament control, it is only necessary to reduce the filament voltage *slowly* until 38 ma plate current is obtained. If the high tube is in the wrong socket, merely interchange the output tubes and proceed as described. Obviously we are balancing by means of reducing the emission of the "hotter" tube. Extensive testing has verified that this method results in improved dynamic balance and sustains the delivery of full power at low frequencies.

Much credit for the high performance-vs.-cost ratio of the amplifier must go to Triad's Model S-35A output transformer. In this circuit the transformer holds up remarkably well, down to 20 cps and costs about half of what you might normally expect to pay for these results. *Figure* 4 shows the chassis layout and *Fig.* 5 shows the underside wiring.

Equalization

The 9000-ohm wire wound resistor and .001-μf capacitor from plate to plate of each output stage, serves to neutralize any ringing tendency with the value of negative feedback employed. Feedback is taken from a voltage divider across the voice-coil winding of the output transformer and is otherwise conventional.

The amount of inverse feedback used is deliberately held to about 10 db. The amplifier is easily driven to full output from less than one volt of input signal so the two volume controls serve mainly as "level-setting" devices. These are linear 0.1-meg. controls and are set about one quarter of full rotation when used with the Miniaturized Preamp with Presence.

Preamp Modifications

While the Preamp with Presence kit is no longer available as a commercial unit, it is still possible to employ conventional construction practices and build the unit without the prefabricated etched wiring panel and the sheet-metal chassis parts, although there is more work involved. However, with a few modifications to the preamp circuit it may be made to operate directly from tape heads, and it is likely that other types of circuits could be modified similarly to obtain the same results. *Figure* 6 indicates the changes in the miniaturized preamp. The TAPE position replaces the FOR (foreign) phono position of the original circuit, and slight changes in the wiring of section C of *Sw₁* permits connecting both phono and tape head to in-
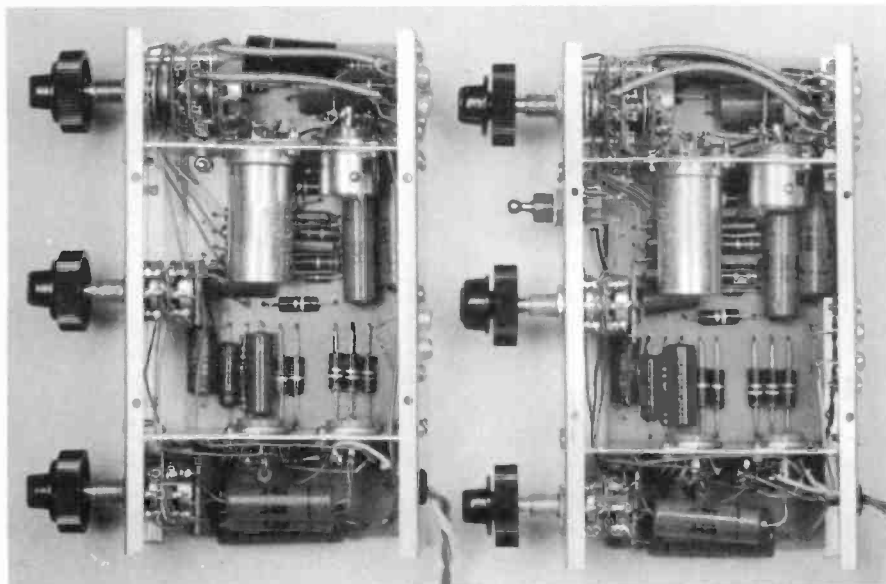

Fig. 7. View of two preamps showing modifications, including addition of toggle switch to the one at the right.

put jacks permanently with the switching selecting the input as well as changing equalization. Note that another phono jack has been added (J_6) and that the feed to the tape recorder has been changed to operate from the output of the entire unit—in parallel with the input to the power amplifier. A pair of resistors has been added to the radio input to reduce the signal level fed to the selector switch, and a DPDT toggle switch has been added to eliminate the tone and loudness controls when it is desired to record radio programs or dub from phonograph records. This change is incor-

Fig. 9. Underside of power supply chassis.

Fig. 8. Top view of power supply chassis.

porated in only one of the preamps unless the user plans on recording stereo in addition to playing back stereo tapes. *Figure 7* shows the two preamplifiers together, the one on the right having been modified with the toggle switch.

Power Supply

The power supply is designed around a 300-volt, 300-ma television power transformer which powers both ampli-
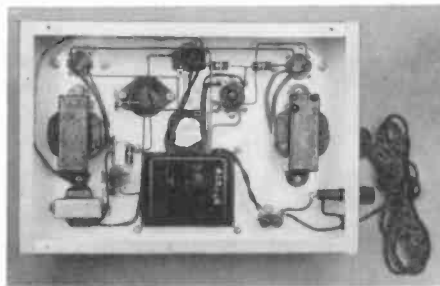
fiers through parallel filter systems. Separate low-resistance chokes and separate filter capacitors were employed to minimize common coupling in the filter system. Most TV transformers have more than one 6.3-volt filament winding. In our case, the two windings were carefully checked for correct phasing null with a pilot lamp and then wired in parallel. The bias supply uses a small filament transformer "backwards" to furnish 120 volts to a selenium rectifier and RC filter network. This provides almost instant bias as soon as the power switch is turned on and there is no danger of the tubes heating up ahead of the bias supply. *Figures 8* and *9* show top and bottom views of the power supply, and *Fig. 10* is the schematic.

While every audiofan or engineer is entitled to his prejudices, we feel that the versatility and performance of this amplifier makes it worthy of consideration for those who plan on having a complete music system.
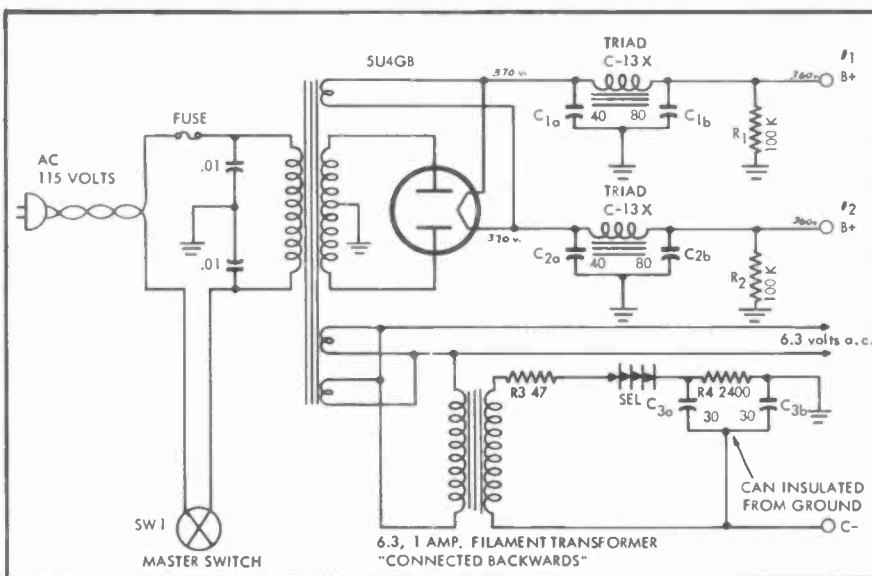
AMPLIFIER PARTS LIST

(Each amplifier section requires the following parts; hence two complete sets are required, except for the chassis.)

C_1, C_2, C_3, C_4, C_5	0.25 µf, 400 v. paper
C_6	10-10-10-50/450-450-450-25 electrolytic; Sprague TVL-4723
C_7	.001 µf, 1000 v, paper; Sprague 10TM-D1
J_1, J_2	Closed circuit phone jacks
P_1	0.1-meg potentiometer, linear; IRC Q11-128
P_2	10,000-ohm potentiometer, audio taper, IRC Q13-116
P_3	5000-ohm potentiometer, linear; IRC Q11-114
P_4	2-ohm potentiometer, wire wound; IRC W-2
P_5	100-ohm potentiometer, wire wound; IRC W-100
R_1	220 ohms, ½ watt
R_2, R_5, R_8, R_{10}	100 K ohms, 2 watts, 5%
R_3	470 K ohms, 1 watt
R_4	1500 ohms, ½ watt
R_6, R_7	1.0 megohm, ½ watt
R_9	3900 ohms, 1 watt, 5%
R_{11}, R_{12}	100 K ohms, 1 watt, 5%
R_{13}	15,000 ohms, 2 watts
R_{14}	9000 ohms, 10 watts, wire wound
R_{15}	1000 ohms, 1 watt
R_{16}	4700 ohms, 1 watt
R_{17}	47,000 ohms, 1 watt
R_{18}	5000 ohms, 5 watts
R_{19}, R_{20}, R_{21}	10,000 ohms, 1 watt
Sw_1	DPDT toggle switch
T_1	Output transformer, Triad S-35A
V_1, V_2	6SN7GTB
V_3, V_4	6Y6
V_5	0D3/VR-150
Chassis	10 × 12 × 3 in.
Sockets for 6SN7's are Vector 10MB12T (4 required)	

POWER SUPPLY PARTS LIST

C_1, C_2	80-40/475, electrolytic; Sprague TVL-2850
C_3	30-30/150 electrolytic; Sprague TVL 2422 (insulated from chassis)
C_4, C_5	.01 µf, 400 v. paper
L_1, L_2	Choke, 3 Hy, 160 ma, 75 ohms; Triad C13-X
R_1, R_2	100 K ohms, 2 watts
R_3	47 ohms, 2 watts
R_4	2400 ohms, 2 watts
Sw_1	DPST toggle switch
T_1	TV power transformer for 300/325 volts d.c. output at 300 ma; 6.3 volts at 12 a; 5 volts at 3 a.
T_2	Filament transformer, 6.3 v at 1 a.
V_1	5U4GB
Rect.	50-ma, 130-volts selenium rectifier
Chassis	7 × 11 × 3 in.

Fig. 10. Schematic of power supply.

Stereosonic Magnetic Recording Amplifier

ARTHUR W. WAYNE

Describing a specific amplifier designed for a Ferrograph Tape Deck, but
one which could be adapted fairly easily to accommodate any other type
of stereo deck with heads of similar impedances and drive requirements.

THE BASIC REQUIREMENTS of any magnetic recording system are few and simple. They are:

(1) A tape transport deck
(2) (a) A loudspeaker and (b) a box or baffle for it.
(3) An amplifier
(4) A reasonable amount of intelligence in the use of (1) (2) and (3).

Requirement (4) is easily disposed of, as it is obvious that every reader of AUDIO will more than satisfy it; and of the remaining three items, the only ones in reach of the ordinary amateur constructor are (2) (b) and (3). So far as (2) (b) is concerned, suggestions will be made in Appendix 2 for the construction of a resonant enclosure suitable for use with *one particular make of speaker only:* and, as there are few amateurs with the necessary facilities for acoustic determinations, where other loudspeakers are preferred, the maker's recommendations should be sought.

This leaves us with (3); and a strictly practical description of a commercial amplifier, intended for stereosonic or single-channel use at will, and eminently suited for amateur construction, follows.

With a genuine high-fidelity output of 15 watts per channel, rising to 25 watts peak, and a comprehensive tone-control system, it provides a quite useful amount of noise for the smaller P.A. operator as well as for the home.

The basic amplifier, the Shirley Laboratories Ltd. FS101, shown in *Fig.* 1 was deliberately developed with "listenability" in mind, a subtle facet of hi-fi, not always completely covered by contemporary design. Most modern amplifiers have approximately equal characteristics, but there is no doubt that, to paraphrase "Animal Farm," some amplifiers are more equal than others. Now, we engineers are a parochial lot, much given to blinding ourselves by science, and with a touching faith in figures: moreover, we labour under the extraordinary delusion, perhaps in company with the biologists, that these figures

tell the whole story. Even here we don't play fair, for we talk glibly about square waves and sine waves, and all the other sorts of waves, without explaining that these are functions, part of a general system of analysis of which our familiar audio problems are a very small part indeed. (Even the concept of a square—cubic?—wave in three dimensions seems a little difficult, and we do hear in three.)

The FS103 is designed to work with the "Ferrograph" type C88 stacked-head deck, now becoming available in the U.S.A.; and it has been demonstrated in conjunction with this deck at various Audio Shows in New York and elsewhere, where it appeared to arouse interest. For the amateur who wishes to experiment, Appendix 1 gives details of some possible modifications, one or two of which are in use on versions of the amplifier manufactured for specialized purposes. It is not proposed to discuss the theory of magnetic recording, as this has been fully covered in this journal and elsewhere.

Over-all Circuitry

In the over-all schematic *Fig.* 2, the figures and letters in the circles refer to the tag strips on the underside of the C88 deck. All function switching on Ferrograph equipment is provided on the decks themselves, which makes the task of the constructor considerably simpler than it would be if the switch units were incorporated in the amplifier. At the same time, it renders possible the provision of heavier and hence more reliable switch banks, those on the C88

being very substantial. The terminal strip locations are shown in *Fig.* 3, and the spare positions on the switches may be used for a variety of functions, as dictated by the will of the constructor. Where a letter and a digit appear in a circle, e.g. 3L, OU, this is to be taken that the letter indicates "L" for the left-strip and "U" for the upper strip, the digit referring, of course, to the number opposite the tag. The circuit description of the amplifier proper will be of one channel only, the left one in the diagram, the second channel being a mirror image of the first. The transpositions are obvious.

On replay, the input from the head is taken, through a standard co-ax socket, to T_3, the head-lift transformer, and via J_2 to the grid of voltage amplifier V_5, a low-noise pentode. The output from the anode of this valve is by the way of C_{16}, C_{17}, R_{18}, and J_1 to the top of P_2, the gain control. C_{17} and R_{18} supply a small amount of treble lift, the significance of which will be considered later, and R_{21} and C_8 are an RC bass lift network, providing most of the necessary compensation for the tape losses. Further amplification is by V_4, another low-noise pentode, the output from which is through C_{11} and the tone-control network P_1-P_2-R_{13}-R_{24}-R_{24A}-C_{12}-C_{14}-C_{15}-C_3-C_{13}. When the controls are at their mid positions, there is a boost of approximately 2.5 db at 50 cps. In theory, such a network should be fed from a low impedance source to avoid high frequency losses, but in fact, the difficulty does not occur, capacitor C_{15} compensating up to about 45 kcps. However, it is very easy to reduce the source impedance by the
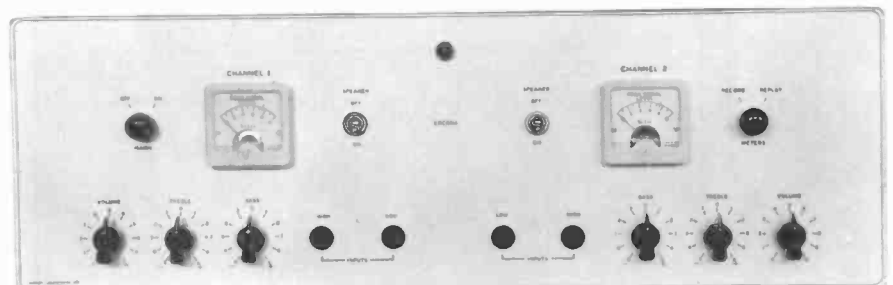

Fig. 1. Front panel arrangement of the completed stereo recording and playback amplifier.
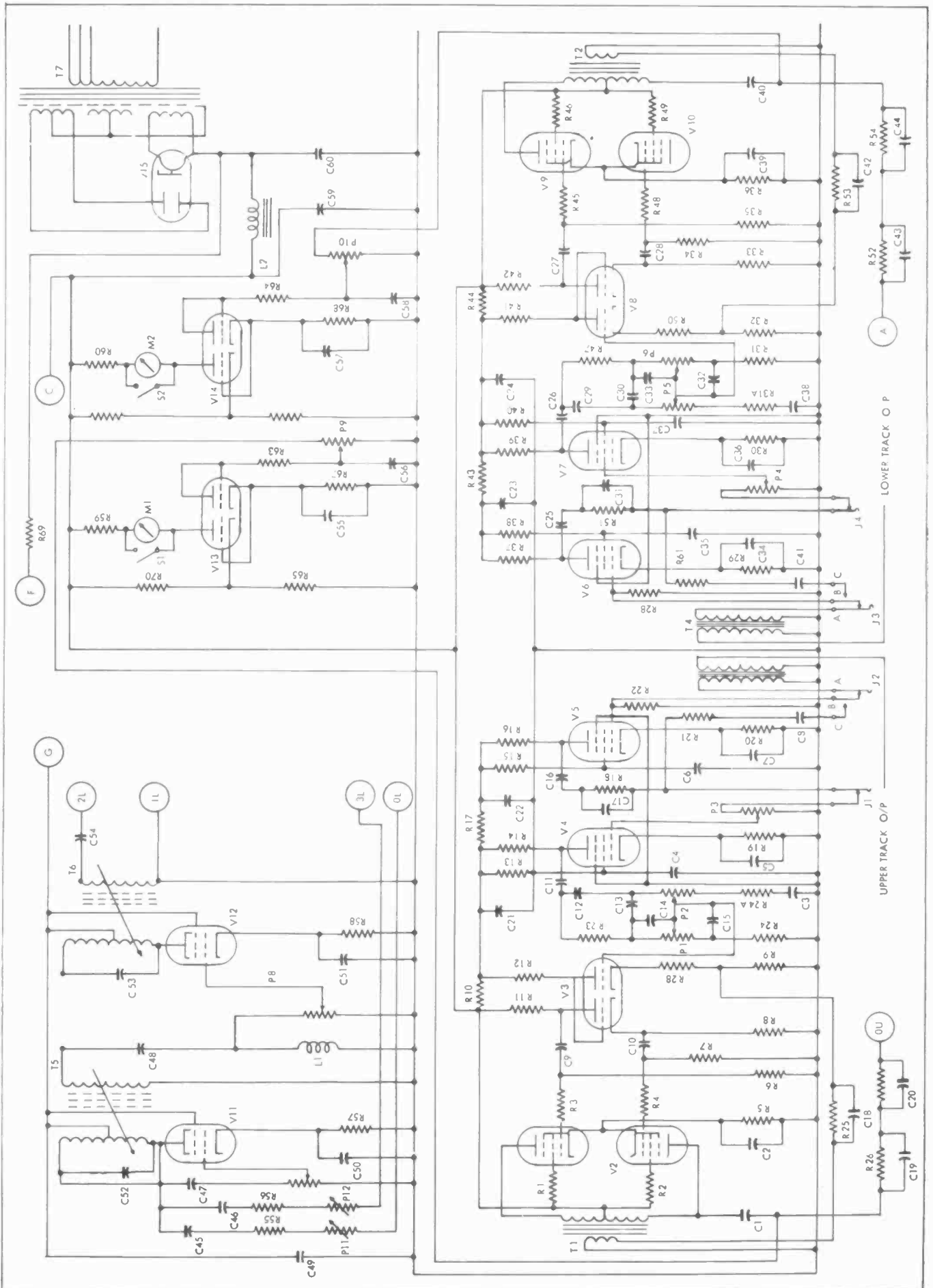
Fig. 2. Over-all schematic of the amplifier described.

simple expedient of connecting a 10-megohm resistor between the grid of V_4 and the junction of C_{11} and C_{12}. A 47,000-ohm resistor from the grid of V_4 to the arm of P_3 will tend to prevent any interaction with V_5. The maximum bass lifts and cuts are 18 db at 20 cps relative to 800 cps, the treble lift and cut at 20,000 cps being 14 db and 18 db respectively.

V_{3A} is another voltage amplifier, with feedback via potentiometer R_{25}, R_9 in the cathode circuit, phase correction being provided by C_{18}. Actually, C_{18} is more in the nature of an insurance against r.f. when the output tubes are viciously overdriven, it being quite superfluous under normal conditions. Additional feedback is obtained by the omission of a bypass capacitor for R_{28}. It is difficult to apply feedback over the whole of the amplifier because of (a) the provision of the two inputs at different levels and (b) the equalizing and tone-control networks; but the circuits of V_4 and V_5 are so calculated as to introduce negligible distortion in these stages. V_{3A} is d.c.-coupled to V_{3B}, the phase splitter, which operates with equal loads, R_8 and R_{11} in the anode and cathode circuits, thus providing the out-of-phase driving voltages for V_1 and V_2, feeds being via C_9 and C_{10} and the grid stoppers R_3 and R_4. For hair-splitting in addition to phase splitting, R_8 should be about 12 per cent lower in value than R_{11}, but in practice, very little difference in output voltage from the two sides will be observable. V_1 and V_2 are operated in class AB1 with common cathode resistor R_5, bypassed by C_2. Should it be desired to use the output valves in pure class A the anode-to-anode load must be altered to 11,000 ohms, R_5 dropped to 75 ohms, and C_2 omitted, when the valves will become self-balancing under most conditions, owing to feedback in R_5. R_1 and R_2 are screen-grid stoppers, to remove a possible source of unwanted r.f. oscillation on full drive, the stoppers being considerably more effective in this position than in the anode leads. Feedback to ensure valve balance may be attained by the inclusion of an undecoupled resistor of 150 ohms in the common H.T. feed between the junction of R_{10} and R_{11} and R_1 and R_2; but it is, however, unnecessary, unless the emissions of V_1 and V_2 are markedly different, in which case the valves should be changed, anyway.

T_1, the output transformer, must be chosen with care, as it is a very critical component, and only the best will do here. It should not be overlooked that it has to deliver up to 15 watts on continuous sine-wave drive, rising to 25 watts on peaks, and yet not lose its inductance under low-signal conditions; its coupling must be tight, its self-capacitance low, and its resonance points right outside both the audio and low r.f. ranges, this latter because of the inevitable stray bias appearing at the anode of the V_2, when recording : so it will be seen that no second-rate component from the surplus market will be satisfactory.

Power Supply

The power pack is perfectly normal, except for the provision of a 100-μf capacitor for smoothing. This is to prevent interaction between the channels at low frequencies, a point not to be forgotten if the constructor contemplates using existing stocks of different values off the shelf. The reservoir capacitor, too, must be chosen with care, that used in the commercial equipment being capable of handling a ripple current of 600 ma. As the total current drawn by the amplifier on RECORD is approximately 230 ma., it will be seen that the 600 ma. ripple requirement is not excessively high. The formulas for calculating both the impedance of the smoother and the ripple current in the reservoir are given in Appendix 1, as well as an alternative power supply section, to cater to the more impecunious reader.

R_{69} must be explained at this point. On the Ferrograph decks, a quick-release device is fitted in the form of a solenoid, the armature of which normally holds the switches and linkages "in" when operating. Depression of a small button on the deck control panel short-circuits the solenoid coil, so releasing the armature and stopping operations. R_{69} is the limiting resistor for the solenoid current, the minimum requirement of which is 30 ma., the coil resistance being 300 ohms.

On RECORD, movement of the deck control knob to that function automatically disconnects the heads from the input sockets, and joins the B+ lines to the oscillators, of which more anon. It also connects the anodes of the recording output valves to their respective heads, together with the bias inputs. On the amplifier, recording is done, in the case of low-level inputs, via J_2, which is a double circuit jack socket. Insertion of the jack changes over both the ground and live contacts, breaking the first, so disconnecting the bass equalizing chain

PARTS LIST

C_1, C_{40}, C_2, C_5, C_7, C_{33}, C_{36}, C_{39}	0.25 μf, 500 v. paper	
	25 μf, 25 v. electrolytic	
C_3, C_{38}	3200 μμf, mica or ceramic	
C_4, C_6, C_8, C_9, C_{10}, C_{11}, C_{16}, C_{25}, C_{26}, C_{27}, C_{28}, C_{35}, C_{37}, C_{48}, C_{50}, C_{51}, C_{54}	.05 μf, 500 v. paper	
C_8, C_{41}	.012 μf, 500 v. paper	
C_{11}, C_{29}	500 μμf, mica or ceramic	
C_{13}, C_{50}	30 μμf, mica or ceramic	
C_{14}, C_{33}, C_{36}, C_{57}	1000 μμf, mica or ceramic	
C_{15}, C_{22}, C_{45}, C_{46}	.01 μf, 500 v. paper	
C_{17}, C_{19}, C_{20}, C_{31}, C_{42}, C_{43}	200 μμf, mica or ceramic	
C_{19}, C_{44}	1500 μμf, mica or ceramic	
C_{21}, C_{22}, C_{23}, C_{24}, C_{49}	16 μf, 350 v. electrolytic	
C_{47}	2200 μμf, mica or ceramic	
C_{52}, C_{53}	3000 μμf, Silver Mica	
C_{59}	100 μf, 450 v. electrolytic	
C_{60}	60 μf, 450 v. electrolytic	
L_1	2.5 mh.	
L_2	10 Hy. 250 ma, 200 ohms	
P_1, P_2, P_3, P_4, P_5, P_6	2 megohms, audio taper	
P_7, P_8, P_9, P_{10}	50,000 ohms, linear, preset	
P_{11}, P_{12}	10,000 ohms, linear, preset	

R_1, R_2, R_{16}, R_{40}, R_3, R_4, R_{44A}, R_{31A}, R_{45}, R_{18}	47 ohms, ½ watt, 10%	
	4700 ohms, ½ watt, 10%	
R_5, R_{36}	130 ohms, 4 watts	
R_6, R_7, R_{35}, R_{15}	0.47 megohms, ½ watt, 10%	
R_8, R_{11}, R_{12}, R_{14}, R_{16}, R_{33}, R_{37}, R_{39}, R_{41}, R_{42}	0.27 megohms, high stability, 5%	
R_9, R_{32}	47 ohms, ½ watt, matched to within 5%	
R_{10}, R_{17}, R_{36}, R_{27}, R_{43}, R_{44}, R_{53}, R_{54}	47,000 ohms, ½ watt, 10%	
R_{13}, R_{15}, R_{38}, R_{40}	1.5 megohms, high stability, 5%	
R_{18}, R_{51}, R_{59}, R_{60}, R_{62}, R_{64}	0.1 megohms, ½ watt, 10%	
R_{19}, R_{30}, R_{50}	2200 ohms, ½ watt, 10%	
R_{20}, R_{59}	2200 ohms, high stability, 5%	
R_{21}, R_{55}, R_{56}, R_{61}	15,000 ohms, ½ watt, 10%	
R_{22}, R_{28}	1.0 megohm, ½ watt, 10%	
R_{23}, R_{47}	0.22 megohms, ½ watt, 10%	
R_{24}, R_{31}	22,000 ohms, ½ watt, 10%	

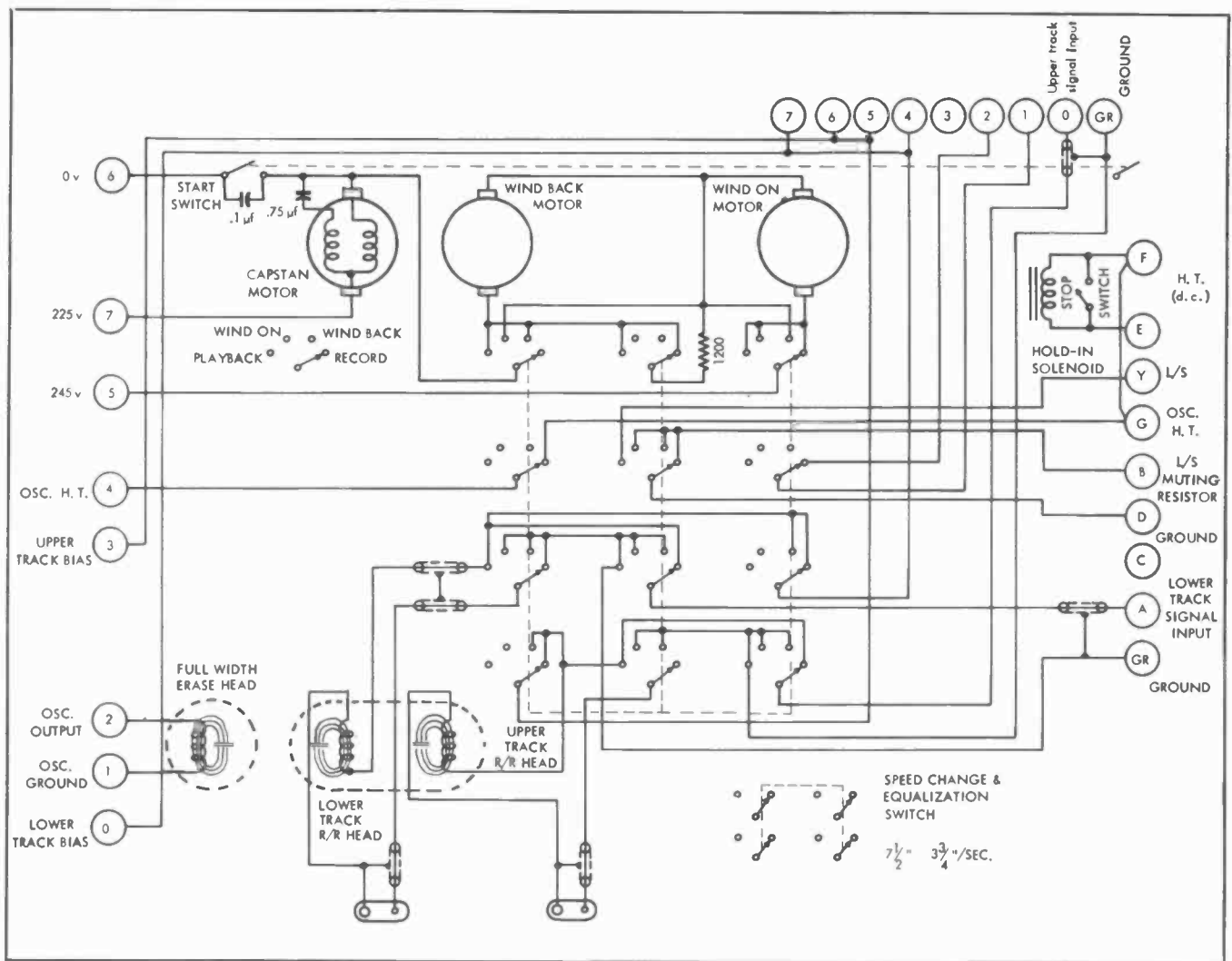|		
R_{25}, R_{52}	1000 ohms, high stability, 5%	
R_{57}, R_{58}	470 ohms, 1 watt, 10%	
R_{62}, R_{70}	0.33 megohms, ½ watt, 10%	
R_{65}, R_{66}	10,000 ohms, ½ watt, 10%	
R_{67}, R_{68}	10.0 megohms, ½ watt, 10%	
R_{69}	10,000 ohms, 15 watts, wirewound	
T_1, T_2	push-pull output transformer, 10,000 p-to-p/15.	
T_3, T_4	Wright & Weaire type 969	
T_5, T_6	Wright & Weaire type 579	
T_7	Power transformer. Primary as required; secondaries: 300-0-300 v. at 250 ma; 5 v. at 3 amps; 6.3 v. at 8 amps, CT.	
V_1, V_2, V_9, V_{10}	EL84	
V_3, V_8	ECC83 (12AX7)	
V_4, V_5, V_6, V_7	EF86 (Z-729)	
V_{11}, V_{12}	6V6	
V_{13}, V_{14}	ECC35	
V_{15}	GZ34	

Fig. 3. Semi-pictorial diagram of the underside of a Ferrograph tape deck showing the locations and designations of the terminal or tag strips.

R_2 and C_9, and transferring the grid input of V_5 from the secondary of T_3 to the tip of the jack. It will be noticed that the small treble lift capacitor, C_{17}, remains in circuit, when by rights it should be in use only on replay; but it was found to be of assistance when recording from certain sources deficient in top, so the original provision for disconnection has been discarded. Its effect may be countered quite easily by use of the treble tone control. High-level inputs, such as those available from tuners and crystal pickups, are fed into J_2, which is an ordinary single-circuit socket. When this is in use, the whole of V_5 and its associated circuit is completely isolated from the rest of the amplifier, so avoiding the possibilities of noise from this stage breaking through. The tone-controls are effective on RECORD as well as on REPLAY, and the value of this feature, unconventional though it may be, has been demonstrated again and again, the author's experience being that, once an operator has made use of this facility, he will not willingly revert

to the standard arrangement of no control on RECORD. The ability to suit the recording to both the acoustic surroundings and the material being transcribed, enables satisfactory tapes to be produced under conditions which would preclude the use of an amplifier with less flexible characteristics. The more conventional user has only to leave the controls on the "zero" positions to satisfy his fastidiousness.

Feed to the head is from V_2 through C_1 and R_{26} and R_{27} to present constant current conditions. A capacitor across the feed resistor as a treble equalizer is quite conventional English practice, but the division of both capacitor and resistor into two is rather unusual. It has been dictated by the necessity of ensuring corrections suitable for all tape speeds, without adding more complication than is essential. Its effectiveness is not to be doubted. At 3¾ i.p.s., C_{19} gives the required lift, C_{20} hardly having any effect at all, while at 7½ and 15 i.p.s., the combined effects of C_{19} and C_{20} together with the head losses at

each speed permit a response up to the theoretical maximum. In accordance with accepted standards as dictated by the physics of magnetic recording, nearly all the treble equalization is on RECORD and most of the bass correction on REPLAY: the characteristics of the tone controls give the recommended NARTB bass boost on RECORD, while C_{17} does the same for treble on REPLAY.

Modulation control is by M_1 and its associated circuitry, where, if at all possible, the meter specified should be used, as its ballistics are ideal. The arrangement is conventional for a sustained-peak-reading voltmeter, the delay on peaks being determined by the time constant of C_{55} and R_{67}. The values chosen appear to satisfy most conditions, but there is no reason why the constructor should not make alterations to suit himself. R_{61} and R_{65} bias the meter down to its zero position, and may be replaced by a potentiometer of 500K ohms, while P_9 is for setting the overload point on the scale. Directions for doing this will be given later. The meters

are in circuit on both RECORD and REPLAY, and in the latter position are used only to ensure balance between the two channels. Switches S_1 and S_2 short-circuit them when this operation is completed.

Bias Supply

The bias section comprises V_{11} and V_{12}, V_{12} being a slave oscillator controlled by V_{11}. A two-valve oscillator is the simplest way of providing the r.f. requirements, as the demands of the Ferrograph "C" decks in this direction are quite heavy; but other arrangements are quite feasible and two alternative circuits are shown in Appendix 1.

Recording bias to the heads is from the anode of V_{11} via C_{45}, R_{55}, P_{11}, etc., the rheostats being an essential part of the circuit, as each individual head has a bias requirement peculiar to itself. The values will be found on a label under the fly-wheel housing under the deck. Erase voltage is supplied by the separate oscillator V_{12}, and as it is not possible to ensure absolute matching of frequency between the two oscillators, it is controlled by a grid drive from V_{11}, via the secondary of T_5 and P_8, bias loading being provided by the 2.5 mh coil L_1. The oscillators and power pack are built on a chassis separate from the main amplifier.

This completes the description of the FS103, the construction of which should provide a few week-ends of amusement for the competent amateur. It is, in essence, a resonably simple piece of apparatus to make, needing patience, a certain amount of skill, and a fairly well-equipped workshop; but a hint or two as to assembly and what may be expected from the completed amplifier may not be out of place.

Fig. 4. Arrangement of components on terminal board which makes for simple and direct wiring.
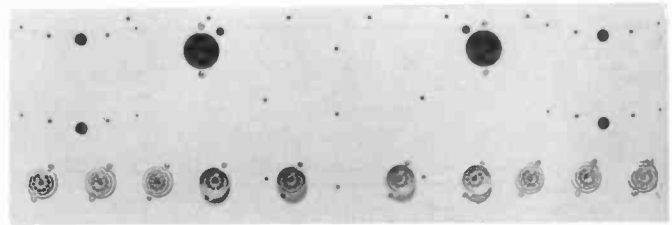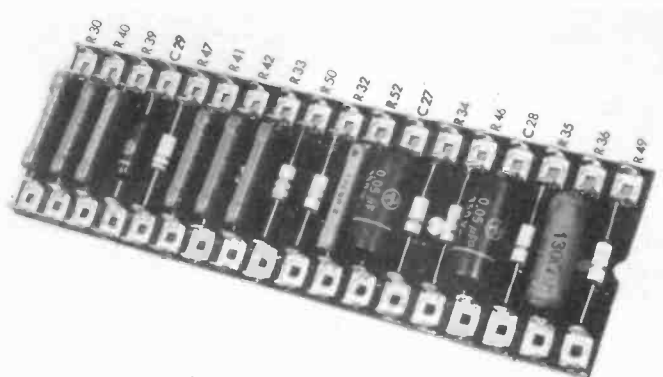

Construction

As seen in the parts list, certain resistors are 5 per cent high-stability units. In the commercial equipment, all these, with the exception of R_9 and R_{25}, are 1 per cent, but this is perhaps painting the lily. C_1 must, repeat must, be of very high insulation, as any leakage here will result in noisy recording; and R_{26}, R_{27}, C_{19}, and C_{20} should be chosen to within 5 per cent. C_{49} is to prevent a too rapid decay of oscillation when the RECORD switches are broken, and if switch clicks are objectionable, a 100-ohm resistor may be inserted in series with the B+ line. L_1 may sometimes be replaced by a 1000-ohm resistor. P_3, in view of the high gain—85 db or so[1]—must be absolutely above suspicion, and the author's unvarying choice is either the Morganite type "A" pot, or the Clarostat type "H," both of which are outstanding in the matter of silence. The group board in *Fig.* 4 was made up with Morganite type "S" resistors, and their neat appearance will be noted. They have excellent characteristics, particularly in their long term resistance to change, an important matter in matched amplifiers. *Figures* 5, 6, and 7 show the top and bottom views of the chassis in various stages of construction.

There must be one ground point, and one point only, for the whole amplifier; and this point is where the co-ax inlets from the heads are located. The simplest

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[1] From the grid of V_4 to output.

method is to solder a bus-bar consisting of a piece of #16 s.w.g. wire along the tube sockets, taking all ground returns to it. The sections are screened from each other by a partition that divides the underneath of the chassis completely, while screens isolate each input side— that is T_3, V_5, C_{16}, R_{18}, C_{17}, R_{15}, R_{16}, C_6, R_{20}, R_{21}, C_7, C_8, R_8, and J_2 and the equivalents of Channel 2 from their respective amplifiers. The two co-ax shells are joined by a short length of #16 s.w.g. wire, so making them one from the ground point of view.

Substitutes for the values in the amplifying chains should not be used, as the basic amplifier was actually designed around those specified; and this injunction applies particularly to V_4 V_5 V_6 V_7, the Mullard low-noise pentode type EF86, for which the author has found no really adequate alternative. The oscillators are located on the power pack, which is a separate unit, and no special precautions need be taken here, except to see that the coils are under the chassis, to prevent undue radiation.

Performance

It is difficult to give a figure for permissible residual hum and noise, as it depends on what the individual defines as "permissible"; but some idea of the possibilities will be suggested by the fact that, on the commercial amplifier, given reasonably smooth mains, with the grid of V_4 short-circuited and with the ear held close to the speaker, it is literally impossible to detect a trace of hum even with the bass control at full boost; while with the heads connected and P_3 at full gain, the noise level is still much below the tape differential. At half gain, which is generally the maximum on REPLAY with a properly recorded tape, the background is nearly at vanishing point. There is no doubt that this desirable state of affairs is due in large part to the excellence of the EF86 as a low-noise amplifier.

The setting-up procedure is not difficult, neither need the test gear be particularly involved, at least for the amateur. An audio oscillator capable of a range of 50 cps to 50 kcps, a VTVM, and an

Fig. 5 Top view of the chassis with tube sockets mounted.

oscilloscope are the essentials. After making the usual tests and adjustments on the amplifiers in detail, the speakers can be phased by feeding a 100 cps signal into both channels simultaneously, and changing round the output leads to one of the speakers to the position that makes most noise: but the signal must be kept to a reasonable level, as the ripping of speaker cones off the spiders is not unknown. Next, the grids of V_6 and V_7 are short-circuited, the bass controls tuned to maximum and the volume controls to full gain, with the treble controls at about 12 o'clock. The interaction between the channels should be at least 100 db down, and if it is very much

Fig. 6 Under side view of chassis with filament wiring in place.
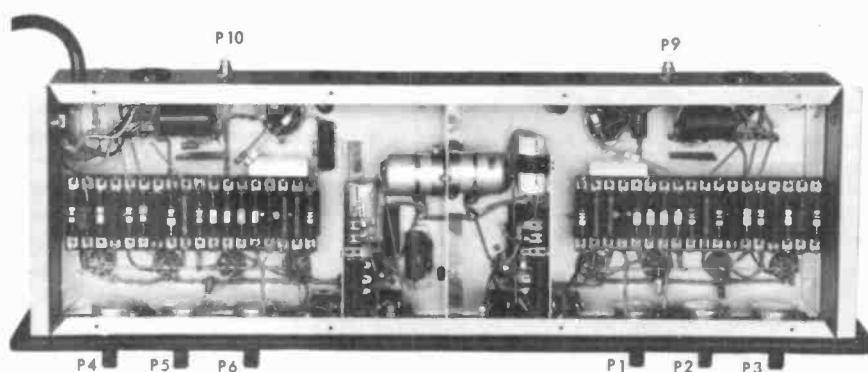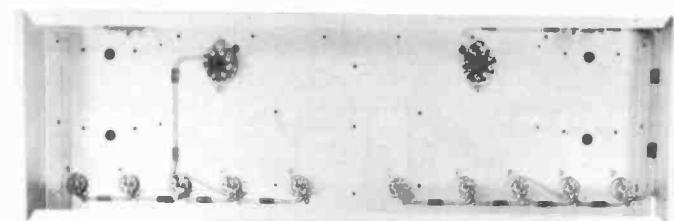

Fig. 7.
Bottom view of completed amplifier chassis.

different, the cause should be sought and cured before proceeding further. A possible source of trouble is a high power-factor or low capacitance C_{59}. If the first, a paper capacitor of 1 or 2 μf will sometimes overcome the trouble, while the cure for the second is obvious.

The connections to the deck may now be wired, and the hold-in solenoid tested, after the following alterations to the under-deck connections have been made. The link between F and G is removed, C on the right tag-strip is joined to 4 on the left tag-strip, 5 and 7 on this strip are joined, and the motor mains connected to 6 and 7. If all is in order here, a known good full-width tape should be played, using each channel separately by manipulation of the volume controls. If the results are satisfactory, the meters may be balanced. A 1000 cycle signal is injected into each

channel in turn, with the tone controls all at 12 o'clock and P_9 and P_{10} adjusted so that, with 40 volts across the recording networks, the meters stand at the 7½ mark, which represents peak signal level. Note that this is *not* a reasonable sine-wave recording signal, but the maximum permissible instantaneous peak. Amplifier balance is checked by feeding signals at various frequencies

into each channel in turn, and noting the relationships between the positions of the respective controls, for equal indications on the meters. The differences, if any, should be small, and large discrepancies tracked and cured.

The deck should now be switched to RECORD, and the oscillators checked for performance. With the oscilloscope connected between the top of P_7 and ground, the slider of this control is advanced to the maximum consistent with good wave-form, and T_5 is tuned, by means of its adjustable core, to 50 kcps. The classical means of calibration is to feed a signal of known frequency direct to the X plates, while the output from the oscillator is taken to the Y plates, the timebase meanwhile being rendered inoperative, and observing the resultant ellipse or Lissajous figure; but with the simpler oscilloscopes, it may prove difficult to attain sufficient stability, and it is far easier to use the oscilloscope in the normal way, filling the screen with, say, four waves from the known source, then disconnecting this and substituting the output from the top of P_7, adjusting T_5 until the same configuration appears. It takes about a minute to do. P_{11} and P_{12} are then set to provide the correct biases to the heads, after which, further adjustment of P_7 may be needed. The oscilloscope input lead is transferred to C_{54}, the slider of P_8 advanced about half-way, and the core of T_6 adjusted until V_{12} locks in. The slider of P_8 is now set as far as it will go without distorting the waveform, and a quick check made at the various positions already covered, as some slight further readjustment may be called for. This completes the preliminary setting up.

The chassis may be of brass, steel or aluminum, depending on the pocket and/or the metal-working skill of the constructor. The one shown in *Figs.* 5 and 6 is of 20-gauge steel, and measures $21 \times 10 \times 2$ in. It is fitted with a bottom cover, thus making a completely enclosed box. Octal sockets are used for the deck connections, and screening of the recording feed wires is unnecessary. Replay leads must be of good quality nonmicrophonic coaxial, and it is suggested that those supplied with the deck be replaced, as they are a little on the short side for custom installation. The ground connection from the deck chassis should be entirely separate from the head cable returns, and is to be taken to some point on the amplifier chassis remote from the co-ax inputs.

Stereosonic Magnetic Recording Amplifier

ARTHUR W. WAYNE

Concluding the description of a specific amplifier designed for a Ferrograph Tape Deck, but adaptable to accommodate any other type of stereo deck with heads of similar impedances and drive requirements.

In Two Parts — Part II

THE FIRST INSTALLMENT of this article described the amplifier without showing the derivations of some of the statements made. Here in the Appendix the author indicates the direction to take in case further delving into performance and operation is indicated or desired, suggests suitable loudspeakers for use with the equipment, and offers some helpful hints about operation.

APPENDIX 1

(1) Although the FS103 amplifier gives professional results, there is no reason why the experienced amateur should not do a little experimenting. After all, what is politely known as development work is really nothing more than the application of brute-force-and-ignorance methods to designs that look well on paper and satisfy the most esoteric manipulations of the slide-rule, but sound horrible when built, or just won't work at all.

One of the more interesting fields for experiment is in the matter of equalization, where many methods are in use, the applications of some being rewarding from the points of view of both interest and results. The twin-T feedback network is one of these, and the basic circuit is given in *Fig.* 8. The whole of the tone control network R_{33}, R_{14}, R_{24A}, P_1, P_2, C_{12}, C_{13}, C_{14}, C_{15}, and C_5 is omitted, and in its place, on RECORD, a feedback network is applied between anode and grid of V_4.

The formula for "resonance," when the impedance of the network is at its maximum, is

$$f = \frac{159 \times 10^6}{R \times C}$$

where R is in thousands of ohms and C is in thousandths of microfarads (picafarads) and the amount of feedback may be varied by tapping off the anode load R_{14}. With $R = 220K$ and $C = 68$ pf (.068 μf), the lift at around 10,000 cps is about 18 db: but in the author's opinion, it is too peaked, and more satisfactory aural results may be achieved with a bridged-T network. The reactance of CA gives the small bass boost, with RA responsible for the levelling-off.

Bass boost on replay is obtained by the substitution of a series feedback network consisting of 680K and 180 pf between anode and grid of V_4. However, the lift of about 25 db at 50 cps compared to 2,500 cps is perhaps too much for American replay standards, and 1.8 meg. will bring this down to approximately 18 db.

(2) The calculation of the coupling between the two channels, as represented by the impedance of the smoothing capacitor,

THE TRANSMISSION OF THE FEEDBACK NETWORK EXCEPT AT "RESONANCE" IS CONTROLLED ONLY BY THE REACTANCE OF CA AND THE RESISTANCE OF RA AND RB.

AT RESONANCE, THE IMPEDANCE OF THE NETWORK ALONE CAN BE VERY HIGH — "INFINITE ATTENUATION" — BUT IS MODIFIED BY THE SERIES RESISTORS AND ALSO BY THE SHUNT RESISTORS RC AND PART OF P3.

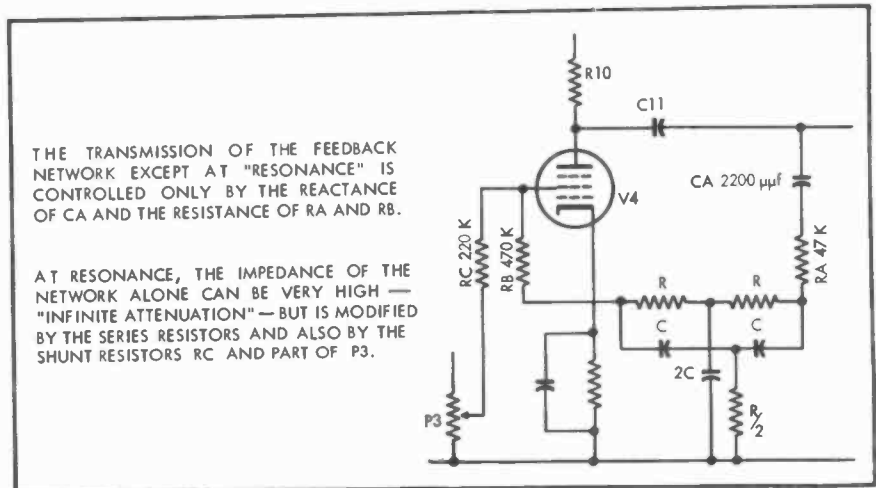Fig. 8. The transmission of the feedback network except at "resonance" is controlled only by the reactance of CA and the resistance of RA and RB. At resonance, the impedance of the network is given by the formula

$$ZC = \frac{10^5}{2\pi \times f \times C(\mu f)}$$

In the FS103, the smoothing capacitor = 100 μf. Common impedance =

$$\frac{10^5}{6.28 \times 50 \times 100} \text{ at } 50 \text{ cps} = 32 \text{ ohms approximately.}$$

Ripple current is approximately 1.4 times the load current. On RECORD, current = 230 ma; ripple current = 230 × 1.4 ma = 322 ma, which figure must be borne in mind when choosing the reservoir capacitor.

An alternative power circuit, which avoids the difficulties associated with high ripple currents, as well as being cheaper and lighter than that of the FS103 is given in *Fig.* 9. The capacitors should have a working voltage of 450 although 350 wv is permissible.
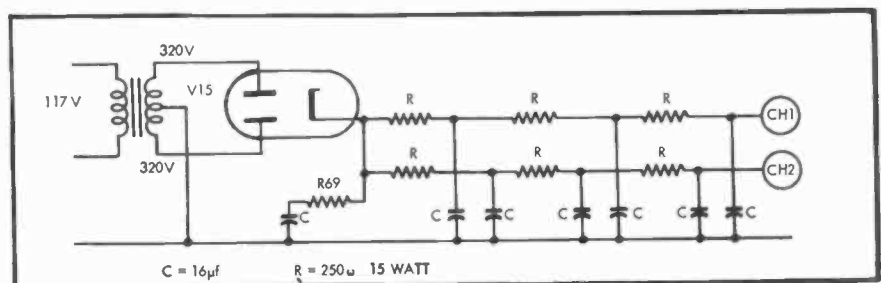
alone can be very high—"infinite attenuation"—but is modified by the series resistors and also by the shunt resistors RC and part of P_3.

(3) Two out of the many possible oscillator circuits will be found in *Fig.* 10. In the oscillator of (A), V_{11} is omitted altogether, and V_{11} is a Mullard EL34. P_7 is returned to ground through the Varite thermistor type V1011, adjustments being made as before. The EL34 is a power valve capable of a really remarkable r.f. output, and the thermistor stabilizes the drive to the grid, chiefly in the direction of bypassing it when the current increases beyond a predetermined level. In (B) of *Fig.* 10, V_{12} is again dispensed with, and a Mullard ECL82 is substituted for V_{11}. This is a combined triode-output pentode, the master oscillator being the triode section. Control is by grid-leak bias, and the circuit is largely self-regulating. As the amplitude of the oscillations increases, the grid capacitor charges and raises the negative bias, until a state of balance is reached in which the oscillations are the maximum
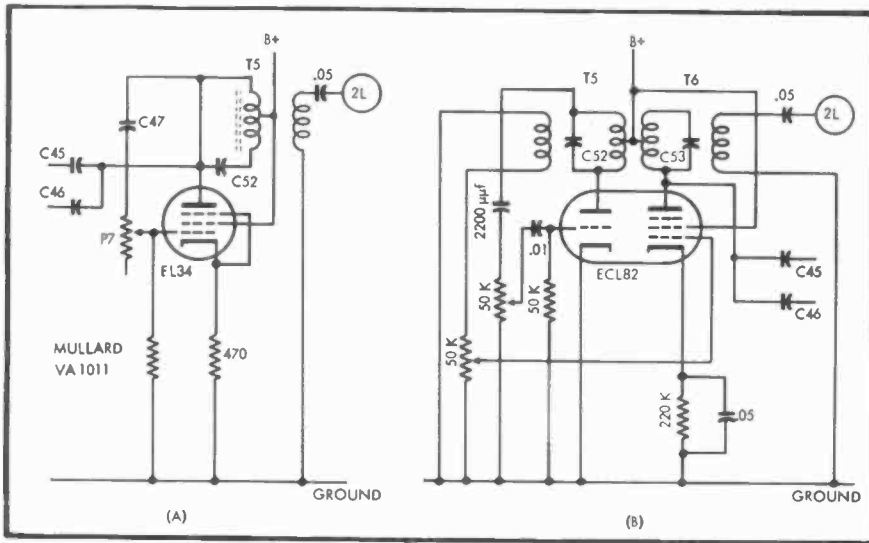

Fig. 9. Schematic of simple power-supply circuit which has a low ripple-current content.

Fig. 10. Two possible circuits suitable for the bias/erase oscillator.

possible, taking into consideration the setting of P_7. The pentode section of the ECL82 is arranged as the slave oscillator, but with both bias and erase taken from it, leaving the master oscillator free from external influences.

(4) Some constructors may consider fitting bias traps in the head feed circuits, to keep r.f. off the output plates. It seems a rather unnecessary refinement, as bias and signal do go together, but two circuits for the purpose are given in *Fig.* 11.

APPENDIX 2.

The suggested choice for loudspeakers for use with the amplifier is the Goodmans Axiom 22 or Axiom 150 Mk. 2. There are, no doubt, equally good speakers on the market, but the author has yet to hear them. Their response is wide enough to dispense with crossover systems and tweeters—which can introduce serious problems in phase shift—and it is characterized by quite silky smoothness. These speakers have only one fault—if the amplifier is not of the best, they proclaim it to the world unhesitatingly and unequivocally. A resonant enclosure of the dimensions shown in *Fig.* 12 gives good results, the separation and definition being excellent. Note that, when making power measurements, a resistor as load will not give a picture of the true output. The amplifiers are designed expressly to work into a loudspeaker load, and the dummy should consist of a
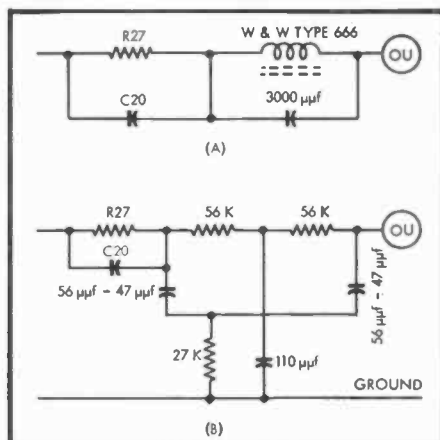

Fig. 11. Two types of bias-trap circuits which may be employed if considered desirable, although they are not absolutely necessary.

reactance of 15 ohms total impedance at 400 cps. An inductor of 3½ mh in series with 10 ohms is about right.

APPENDIX 3

On the operation of stereosonic reproducers.

An operator, using this type of equipment for the first time, will almost certainly try to achieve perfect balance between channels. Indeed, he is exhorted to do so, more than one writer on the subject stating it is mandatory that the gain and tone controls be ganged for the very purpose. This, in common with many other pontifical pronouncements by the engineering and hi-fi fraternity, is nonsense. To forestall righteous anger and condemnation, the author proposes to make a slight digression.

As was suggested earlier in this article, engineers are, on occasion, apt to make definitive statements about subjective matters, without always considering all the available evidence. If this be not so, how can one account for the changing fashions in the Hi-Fi world? Each new circuit is equated with the "real thing," and each subsequent one is so much better than the last; but it is also the "real thing," a sort of ultra-real realness. At one time, 10 watts was ample for the average living-room; now, according to one concatenate authority admittedly not overmuch given to understatement, 100 watts is the figure. And, as mentioned before, we aren't really honest about it. We use co-ordinate geometry as proof of our statements, and raise Fourier analysis to the dignity of a gospel; but a Fourier series merely happens to be a convenient tool in the manipulation of partial differentials, while, for statements about problems in which subjective perception is an important factor, tensors appear to be the appropriate discipline. Whichever way the matter is viewed, the figures are merely a manipulative convenience, and not statements of fact.

Now, the author is a very ordinary engineer, busily engaged in scratching a modest living in a competitive business; but he is, also, a professional musician of vast, literally vast experience. This is not to say that he is anything but a mediocrity, even in that profession, but his first public appearance was at the age of 7, and he is not going to say how long ago that was: (off the record, he would be a grandfather now if his children weren't so lazy!) And on the strength of long acquaintance with hi-fi in the raw, his advice to the amateur

using stereo for the first time is to give up listening with the slide-rule but use, instead, certain rather old-fashioned instruments, a couple of which can be found in most well-appointed homes. They are known as ears, and their discrimination is remarkable—in fact, they are the standard by which all the other instruments are, or should be, judged. If stereo sounds better with one channel slightly louder than the other, play it so. If the performer seems to be in the room, with treble up on track 1 and bass up on track 2, that's where the controls should be. If it sounds right, it *is* right, and don't let any long-haired back-room boys—including the author—tell you it's not. Your ears aren't perfectly matched, neither are the two halves of your room, nor your tastes with the next man's, and all the controls are for use, not ornament.

In conclusion, acknowledgments are due to Charles H. Frank Jr. of the Ercona Corporation, without whose encouragement —not to say vigorous prodding—the original FS103 would probably never been built.
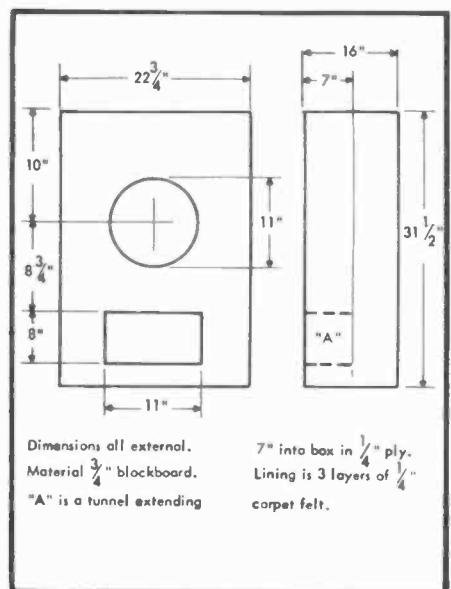

Fig. 12. Dimensions of a "resonant type" enclosure suitable for use with Goodmans Axiom 22 loudspeaker unit.

Errata to Part I

A few minor (?) errors crept into the drawing for *Fig.* 2 in Part I of this article, and at the end of Part II seems the most ideal place to bring them to the attention of readers who may have been particularly interested in this unit.

The jacks J_2 and J_4 were incorrectly drawn, and should have been shown as indicated here. The correct jack is typified by

Switchcraft MT-332C, although Mallory #A, 704A, 5, and 705 may be wired to produce the same results. The lettering refers to the original diagram.

The resistor in parallel with C_{40} in the lower left corner of the drawing is R_{17}; the voltage divider for the cathode of V_{11} consists of R_{42} at the top and R_{66} at the bottom; and the plate of V_{10} should be connected to the line leading from the primary of T_4 to capacitor C_{49}.

It is suggested that you make these corrections on the original drawing. Æ

High-Quality Treble Amplifier

Cdr. CHARLES W. HARRISON, JR., USN

The author describes a four-watt amplifier which employs a single-ended output stage, and which is intended for use as a driver for the tweeter of a two-way speaker system for home use. This unit will considerably decrease the cost of a two-amplifier system.

1. *Introduction*

IN A RECENT PAPER[1] the writer described a dual-channel playback system consisting of a dividing network and two identical amplifiers for driving the bass and treble sections of a dual loudspeaker. This arrangement is economical when speaker elements of approximately the same efficiency are employed, as for example, horn-type speakers for the reproduction of both the low and high frequencies. When the bass section is much less efficient than the treble system, as will be the case when direct-radiator dynamic loudspeakers are used for bass and a horn-type speaker for treble, it becomes entirely feasible to use amplifiers of considerably different power output ratings in a divided amplifier system. For example, if a direct-radiator bass speaker, in the appropriate baffle, has a conversion efficiency of 5 per cent, and the high frequency driver with horn has an efficiency of 50 per cent, 40 watts input to the woofer and 4 watts input to the tweeter will result in the radiation of 2 watts of acoustic power in each channel. One will be able to achieve low-frequency/high-frequency balance[2] under most circumstances, and simultaneously utilize the power capabilities of the bass and treble amplifiers. As a practical matter the proper setting of the volume controls on the amplifiers to obtain the most pleasing response must be determined experimentally by conducting listening tests in the room in which the dual loudspeaker is located.

The purpose of this note is to describe a simple 4-watt single-ended amplifier intended for use as a driver for treble speakers, such as the Western Electric 594A, Jim Lansing D-375, or Altec 288B, when used in a home music reproducing system, or small auditorium. The fact that the power output of the amplifier

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[1] Charles W. Harrison, Jr., "High-quality dual-channel amplifier," AUDIO, January, 1956.

[2] Balance between the low and high frequencies depends on such factors as the directive properties of the speakers, crossover frequency, speaker locations, room acoustics, and the spectral distribution of the energy in the program material. Consideration of these factors may dictate different power ratio requirements for a given system.
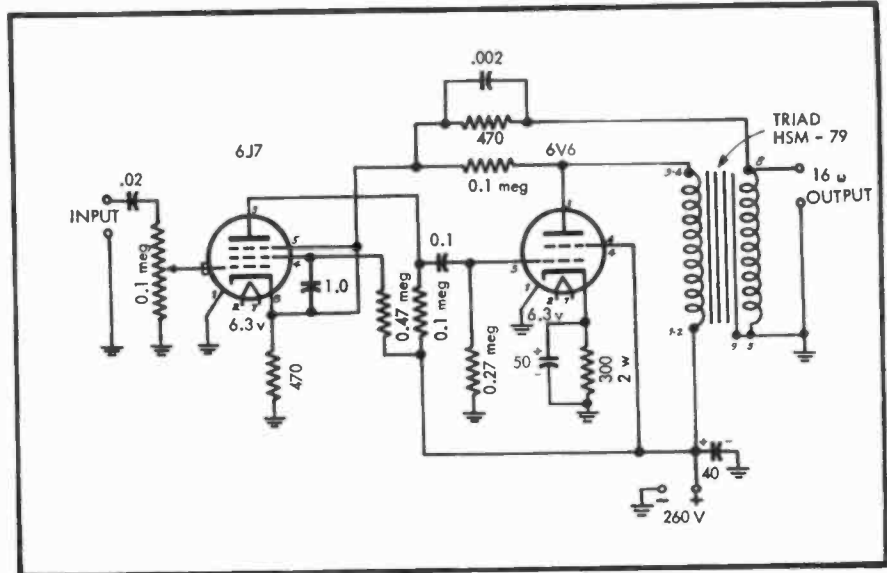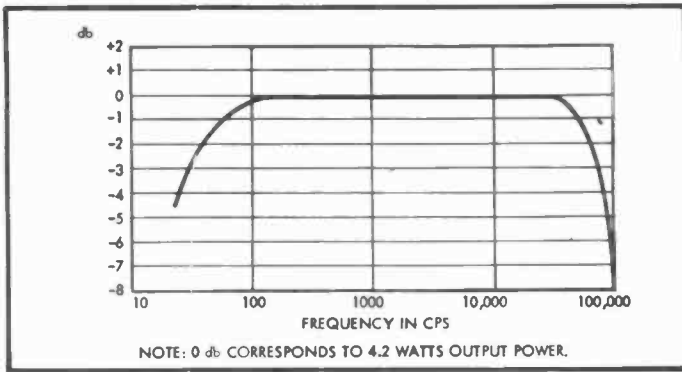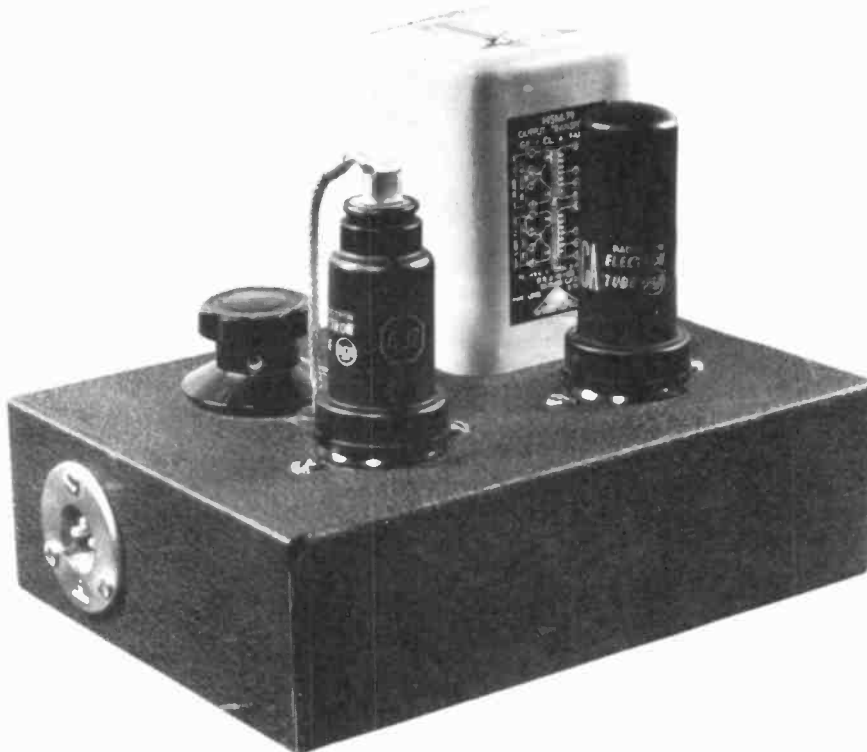

Fig. 1. Schematic wiring diagram of 4-watt treble amplifier.

cannot exceed 5 watts, regardless of the frequency and amplitude of the excitation voltage, insures that the tweeter diaphragm will not be fractured by the inadvertent application of low-frequency signals, or by the development of faults in the treble amplifier.

The Amplifier

The amplifier is built around the Triad HSM-79 hermetically sealed, high-fidelity output transformer. This transformer has a 5000-ohm primary designed to carry an unbalanced current of 40 ma, and secondary impedances of 16, 8 and 4 ohms are available. The guaranteed frequency response is within 1 db from 50 cps to 25 kcs.[3] The response is greatly improved particularly at the high end of the frequency spectrum, by the application of negative feedback around the

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[3] Some readers may feel that the low-frequency response of the Triad HSM-79 transformer is inadequate. The fact that the primary winding carries unbalanced d.c. makes it difficult to achieve high primary inductance. Thus one might anticipate degraded low-frequency response compared to the response of high-fidelity output transformer types designed for push-pull applications. If the amplifier is to be used in the treble channel of a dual-channel playback system, it is perfectly satisfactory —and even desirable—for the frequency response to begin falling off at approximately an octave below the crossover frequency. The "fusing" of the treble driver is enhanced by a rolloff in the bass response of the amplifier.

transformer. Two tubes are used in the amplifier—a 6J7 followed by a 6V6. The schematic is shown in *Fig.* 1. Two feedback paths are employed—one from the plate of the 6V6 to the cathode of the 6J7; the other path is from the secondary of the output transformer to the cathode of the 6J7. These paths are not independent, i.e., changing the circuit parameters in one path changes the effective value of feedback into the other path.

Performance Data

The performance data presented here was obtained from measurements made on an amplifier having circuit values shown in *Fig.* 1, with the following exceptions: (a) The .02-μf input capacitor was shorted. (b) A 270-ohm 2w resistor was used in the cathode circuit of the 6V6 output tube in lieu of the 300-ohm 2w resistor shown in the drawing. The measured grid bias was 12 volts. (c) The interstage coupling capacitor was 0.06 μf instead of 0.1 μf as shown.

A 16-ohm resistor was used to load the amplifier for all tests.

The component values employed in the feedback paths result in approximately 20 db loss in gain compared to the gain of the amplifier without feedback.

Figure 2 is the power curve of the amplifier. It was obtained by adjusting the input signal voltage at each fre-

Fig. 2. Amplifier power curve.

NOTE: 0 db CORRESPONDS TO 4.2 WATTS OUTPUT POWER.

Fig. 3. Photograph of completed amplifier.

watt except the cathode resistor in the 6V6 circuit. The output transformer must be connected in the circuit as shown to insure that the feedback is degenerative. *Figure* 3 shows the completed amplifier, and *Fig.* 4 shows the component arrangement.

The Power Supply

Many audio hobbyists possess a power pack that may be used to power the treble amplifier. The power supply described in a previous article[1] provides plate and filament voltages for both the bass and treble amplifiers in the writer's dual-channel playback system. A 10-watt resistor of 750 to 1000 ohms is required to drop the plate voltage to the correct value of 260 v. This resistor is shown in *Fig.* 4. The plate current of the 6V6 does not vary more than 1 or 2 ma from zero signal to maximum signal, so the voltage regulation of the power supply is not too important.

The power requirements of the treble amplifier are 6.3 v.a.c. at 0.75 a, and 260 v.d.c. at 45 to 50 ma. The schematic for a suitable, yet inexpensive power supply is given in *Fig.* 5. The transformer should have minimum ratings of 300 v.d.c. at 60 ma; 6.3 v.a.c. at 1 a, and 5.0 v.a.c. at 2 a. A 5Z4 is employed as a full-wave rectifier, and filtering is accomplished by use of a resistance-capacitance network. Such filters are recommended when the current drain does not exceed 50 ma. When choosing a plate transformer for use with RC filters it is important to remember that the power consumed in heating the filtering resistors must be provided by the transformer. If the power supply design is not carefully executed a transformer of

quency of measurement until barely visible waveform distortion occurred. The power output was then computed at that frequency. Thus *Fig.* 2 is in reality a curve showing power output as a function of frequency for constant distortion. 0 db corresponds to the power output of 4.2 watts. It is believed that approximately 3 per cent harmonic distortion in the amplifier can be detected by eye, when a good oscilloscope is used for viewing the output wave shape.

When the input signal voltage is adjusted so that the amplifier delivers 2 watts at 1000 cps, the amplifier is flat from 30 cps to 80 kcs. It is down 2.5 db at 20 cps and again at 100 kcs, tapering off to −8 db at 150 kcs and −12.5 db at 200 kcs.

The response of the amplifier to a 20,000-cps square wave is highly satisfactory; to a 10,000-cps square wave the response is perfect.

Constructional Details

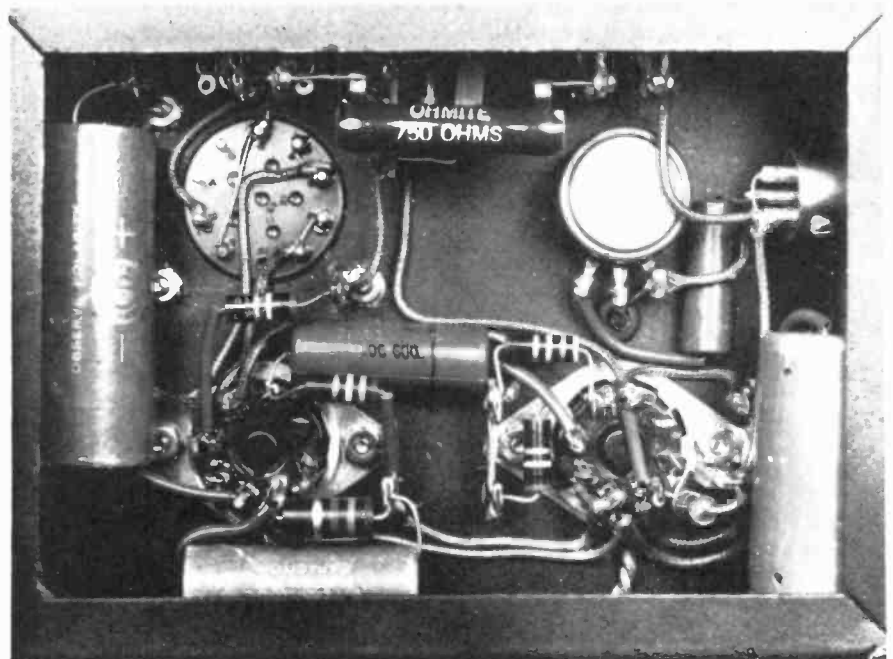The amplifier is easily built on a 5″ × 7″ × 2″ chassis. All resistors are 1

Fig. 4. Arrangement of components in the amplifier.

Amplifier Uses Cheap Output Transformer

NATHAN GROSSMAN AND WILLIAM HELLMAN

Pleasant sound does not necessarily mean highest fidelity. It can be obtained with inexpensive output transformers and by using feedback. The authors show designs for single-ended and push-pull jobs.

THE AUTHORS SET OUT to explore the possibilities of obtaining high fidelity from the average service-replacement output transformer produced by leading transformer manufacturers. These generally sell for about $3.00, with those intended for higher outputs running up to about $6.00. They generally have a primary inductance of 7 to 10 henries depending upon whether they are intended for single-ended or push-pull operation.

It is the low load presented by the primaries of these transformers which would ordinarily prevent their use in connection with quality amplifiers. The gain of the output stage depends upon the matching of the plate impedance of the tube used and the impedance produced by the inductance of the primary of the output transformer. Thus, a tube with a high plate impedance requires greater impedance from the primary inductance for good results. Moreover, to obtain the same response in the bass frequencies as in the middle frequencies also necessitates a high inductance, because the impedance of the inductance falls off proportionately to the decline in frequency.

To get the best results from these transformers then requires that they be used in conjunction with output tubes which require relatively low-impedance loads such as the 6B4 and the 6L6, which

operate satisfactorily with loads of 2500 ohms, and the 6Y6 and 50L6, which will do likewise with loads of 1500 to 2000 ohms.

Uniformity of response can be obtained by using inverse feedback in a proper circuit. For example, 6 db of feedback in the circuit of *Fig.* 1 will flatten out the response of such a transformer, when used with a 6L6 in the output stage, down to 100 cps. Twenty db feedback will flatten out the response down to 20 cps (see *Fig.* 2). These results are much like those obtainable from increasing the primary inductance by like factors.

Such a transformer when used in the circuit of *Fig.* 1 without R_5 and fed into a resistive load produced the following harmonics: at 100 cps:

Watts	2nd	3rd	4th	5th
1	8.5%	0.49%	0.28%	0.12%
2	12.0	0.68	0.50	0.10
6.5	20.0	2.00	2.80	1.60

At the same frequency but with the application of a factor of about 20 db of inverse feedback through R_5 from the secondary of the output transformer the following harmonics were produced:

Watts	2nd	3rd	4th	5th
1	0.80%	0.13%	0.01%	.03%
2	1.20	0.30	.065	.04
4	2.20	0.90	0.32	0.16

Under the same conditions but with an input frequency of 1000 cps the following harmonics were produced:

Watts	2nd	3rd	4th	5th
1	0.34	0.14	.02	
2	0.66	0.24	.05	
4	1.20	0.52	0.15	.04
6.5	1.6	0.90	0.32	0.14

From these tabulations several conclusions can be drawn. First, for the reproduction of speech and treble instruments this is a very fine amplifier. Second, that the harmonic distortion is reduced roughly by the factor of inverse feedback. Third, that the difference in the amount of distortion resulting at the two frequencies at which the measurements were made corresponds roughly to the factor of difference in gain response without inverse feedback at the two frequencies.

In an effort to follow up these conclusions and obtain further improvement a 6SJ7 was substituted for the 6SF5 in the driver stage of *Fig.* 1. The higher gain of the 6SJ7 would permit more inverse feedback. As the cathode was not bypassed a further improvement of 25 per cent in gain, and so also in the amount of inverse feedback, was obtained by connecting the return lead of the screen bypass capacitor directly to the cathode instead of the usual connection to ground. Since the screen is
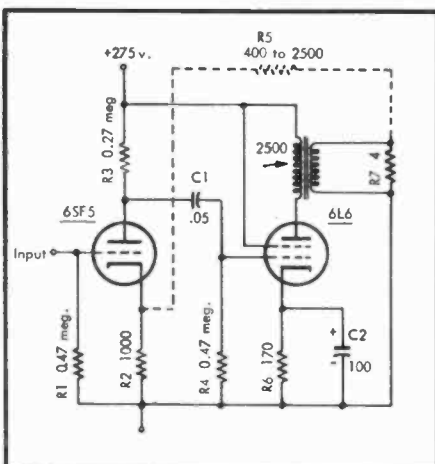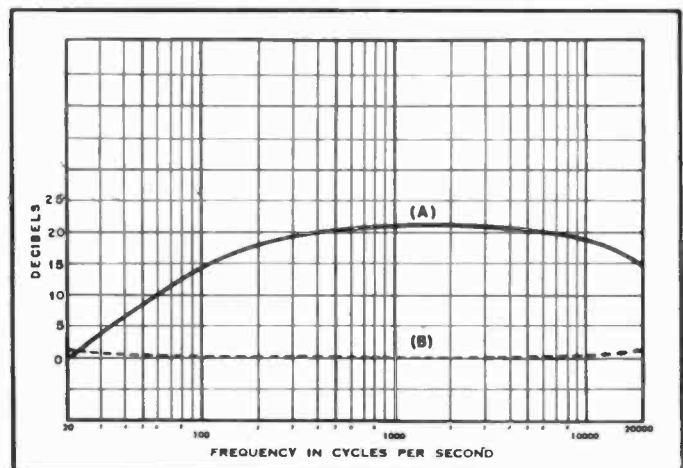

Fig. 1. The basic circuit employs feedback from the transformer secondary. Value of R_5 determines feedback.

Fig. 2. Curve (A) shows original frequency response of Fig. 1 amplifier without feedback. Adding 20 db of feedback gives response of curve (B).

really acting as the plate of a triode this change in circuitry avoids degeneration caused by permitting a.c. from the screen to pass through the cathode resistor.

It was further observed that to get the same amount of inverse feedback at 60 cps as at 400 cps it was necessary to increase the screen bypass capacitor from 0.5 to 8 μf.

By connecting the plate of the output stage through a resistor and capacitor to the cathode of the driver stage as shown in *Fig.* 3, a further increase in inverse feedback can be obtained. However, on checking the over-all response of this arrangement it was found that with a feedback of only 16 db in the middle frequencies there was no response at 20 cps and that above 800 cps there was a gradual loss which amounted to 10 db at 12,000 cps. This meant that there was positive feedback present and more of it at the treble frequencies. The squeals which emanated from the loudspeaker when rotating the dial of the FM tuner showed that there was oscillation. This approach was, therefore, abandoned.

A 6SH7, which has a higher gain than a 6SJ7, was then substituted, and two loops of inverse feedback were employed. The loop through R_7 in *Fig.* 4 served to flatten out the response from the output transformer and to reduce any tendency to oscillate. It also produced a sweeter "feel," much like that of a triode, which the 6L6 resembled after this reduction in its gain. A 6AU6 may be substituted for the 6SH7, but a 6BC5, 6CB6, or 6AG5 cannot be used, as these radiate badly.

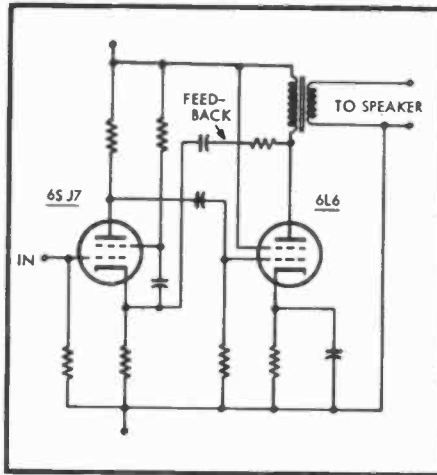The following factors of inverse feedback were obtained from the amplifier

Fig. 3. Obtaining the feedback path from the plate of the output tube gave higher feedback but poorer response.

shown in *Fig.* 4 while using a Stancor A-3830 output transformer:

Frequency	1st Loop	2nd Loop	Total
60	2.8 times	6.5 times	18.2 times
400	4.2 times	6.5 times	27.3 times

With this amplifier signals were heard below 20 cps and a slight loss of amplification was measured at 300 kc. The break-up of the sine wave on the oscilloscope (generally at 3 per cent total harmonic distortion) at 60 cps occurred at 3 watts, and at 400 cps at about 4 watts. Amplifiers which employ large amounts of inverse feedback show low distortion up to a point which is considerably below the ratings published in the tube manuals and beyond which there is a sharp and very great increase in harmonic distortion. Based upon the

above tables and compensating for the increase in feedback, it is estimated that at the 400–1000-cps point at just below 3 watts there should be less than 1/3 of 1 per cent harmonic distortion, and at the 60-cps point, after allowance for the lower frequency, 1.5 per cent harmonic distortion.

Where a tone-compensator stage is desired and where a variable-reluctance cartridge is to be used, the circuit shown in *Fig.* 5 is suggested. It may be necessary to increase the values of either the capacitor or the resistor, or of both, in the decoupling circuits at points A or B in order to overcome motor-boating, which can occur in this amplifier at a frequency as low as one half cycle per second.

A pair of 50L6's connected in push-pull were tried in the circuit shown in *Fig.* 5. An inverse feedback factor of about 7 and an undistorted output of about 3 watts was obtained at 60 cps. Less inverse feedback was needed for this result because the push-pull operation cancelled nearly all the 2nd and 4th harmonics and also the magnetizing effect of the d.c. in the windings of the primary of the output transformer. This latter raised effective input inductance. The 3-megohm volume control R_1 permitted the use of a broad-range crystal cartridge. Capacitor C_1 and resistor R_1 furnished some Fletcher-Munson compensation. Capacitor C_1 served to overcome losses in the shielded cable from the pick-up to the amplifier and also to afford some Fletcher-Munson compensation in the treble frequencies. Various values of C_1 should be tried until the most pleasing result is obtained. An FM tuner with a 1-volt output could be used to drive this amplifier to nearly
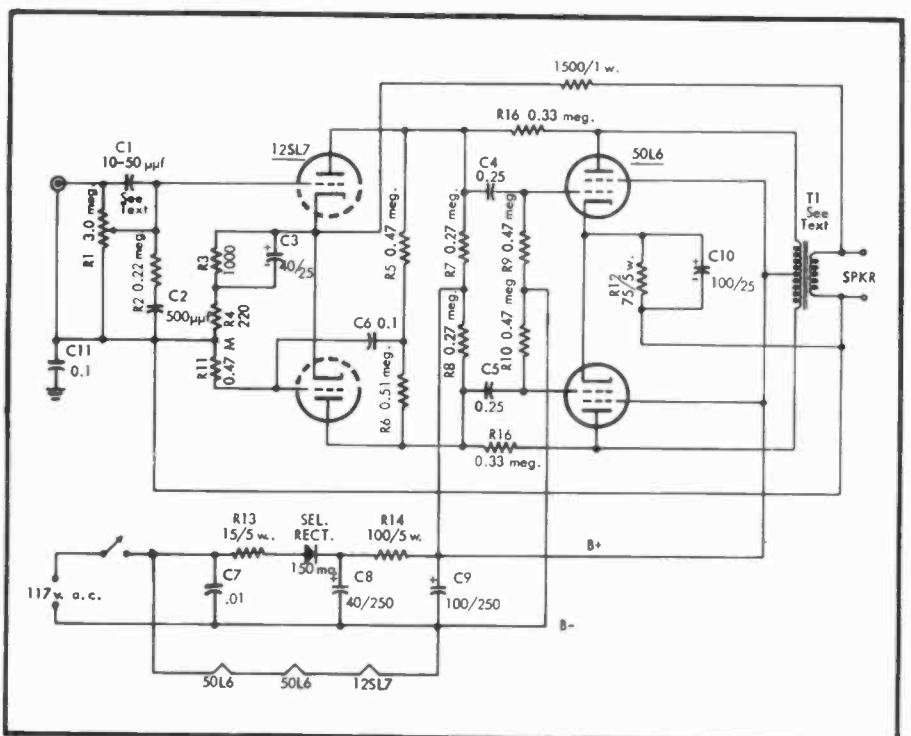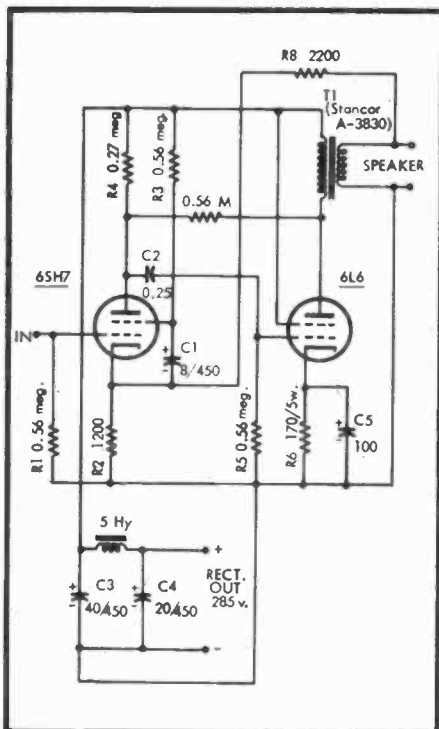
Fig. 4. (left) The final single-ended amplifier, with two feedback connections, one from the plate of the output tube to the plate of the 6SH7 and the other from the secondary of the transformer to the 6SH7 cathode. Fig. 5. (right) The push-pull version of the amplifier.

AMPLIFIER USES

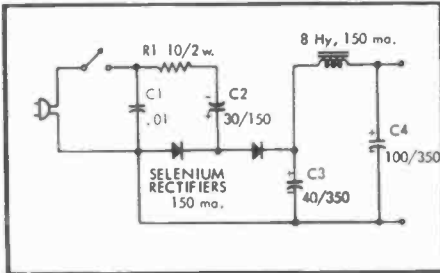full output. It may be necesary to change the polarity of the power line to reduce hum.

Fig. 6. A 135-volt power supply for the push-pull amplifier.

These amplifiers are suitable for home use where all that is needed is an amplifier capable of delivering 2 watts undistorted at 60 cps with a flat response from 20 to 20,000 cps. (Below 60 cps there is very little program material and most loudspeakers produce a great deal of harmonic distortion.) As a matter of fact, with the present higher-efficiency loudspeakers and loudspeaker housings an input to the loudspeaker of 2 watts bass is more than enough to reproduce symphony music in the average living room; and continuous operation at this level is sure to make the neighbors complain.

Although it was not tried, it is very probable that by raising the plate voltage in the circuit of *Fig.* 5 to 135, better than 5 watts undistorted power output can be obtained at 60 cps. If a transformer supplying 6.3 volts is available for the filaments, 6Y6's can be substituted for the 50L6's. The cathode-bias resistor of the output stage R_{11} should then be increased to 120 ohms. A B supply of 135 volts is shown in *Fig.* 6.

Several amplifiers were constructed in accordance with the circuits in *Fig 4*, in which the following transformers were used with excellent results: Stancor A-3825; Thordarson T22S60. A Thordarson 22S74 (costing about $5.00) was used in the circuit of *Fig.* 5 with 6L6's and the following changes: B+, 285 volts; primary impedance, 5000 ohms; R_{11}, 125 ohms; R_{15}, R_{16}, 400,000 ohms. With these changes the undistorted power output at 60 cps on the oscilloscope was 11 watts. A Stancor A-3830 was also used in a similar amplifier with the additional change of substituting a 12AX7 for the 12SL7. With the 12AX7 it was found necessary to remove the output transformer, which was not shielded, from the chassis and mount it on the loudspeaker frame, and also to connect a 100-μμf capacitor from the plate of the input half of the 12AX7 to ground to cut out high-frequency oscillation. Despite this, there was a subtle difference in "feel" between the 12AX7 and the 12SL7 in favor of the former.

TREBLE AMPLIFIER

marginal rating for a given job will be overloaded when RC filtering is employed.

The resistor labeled R in *Fig.* 5 must be chosen so that 260 v.d.c. is delivered to the amplifier. One may not be required. Positive heater bias of one-tenth the plate voltage is provided by the bleeder network. The bypass capacitor grounds the center-tap of the 6.3 v.a.c. winding to audio frequencies and must not be omitted if hum is to be minimized. gle-ended transformer makes it desirable to re-examine the utility of single-ended amplifiers for high-fidelity applications when high power output it not a requirement. The advent of tubes with high

Conclusion

The writer is of the opinion that a high-quality playback system requires the employment of a dual-channel amplifier. A high-level dividing network cannot operate satisfactorily unless each filter section is terminated in a pure resistance of appropriate value. The driving-point impedance of a speaker is complex and functionally related to frequency. It may be predominately reactive at some frequency in the pass band. This accounts in part for the fact that dual loudspeakers employing high-level dividing networks do not always sound right. Constructional details covering an excellent dual channel amplifier were presented in reference 1. The use of the treble amplifier described in the present article will reduce the cost of the system with no sacrifice in performance. When used to drive the writer's

speaker,[4] which is located in a room of modest size, the obtainable sound intensity level approaches the threshold of pain over the entire frequency range of the speaker.

The availability of a high-quality sin-transconductance, such as the Mullard EL-34 and Tung-sol 6550, which result in high power sensitivity and low drive requirements, makes it highly desirable for transformer manufacturers to produce a line of transformers capable of carrying the plate current in the primary winding and that are comparable in performance to those manufactured for push-pull application. It should be possible to obtain easily 10 watts of power output from a single-ended amplifier employing only two stages. The circuit is not complicated and there is no requirement for a phase splitter. Large values of negative feedback may be applied and the amplifier will remain unconditionally stable if the output transformer is properly designed. Admittedly, it is somewhat more difficult to design a good transformer for single-ended output stages than for push-pull application. But the Triad Transformer Corporation has made a good start in producing the model HSM-79 transformer, and this should serve as a challenge to other manufacturers in this field. ●
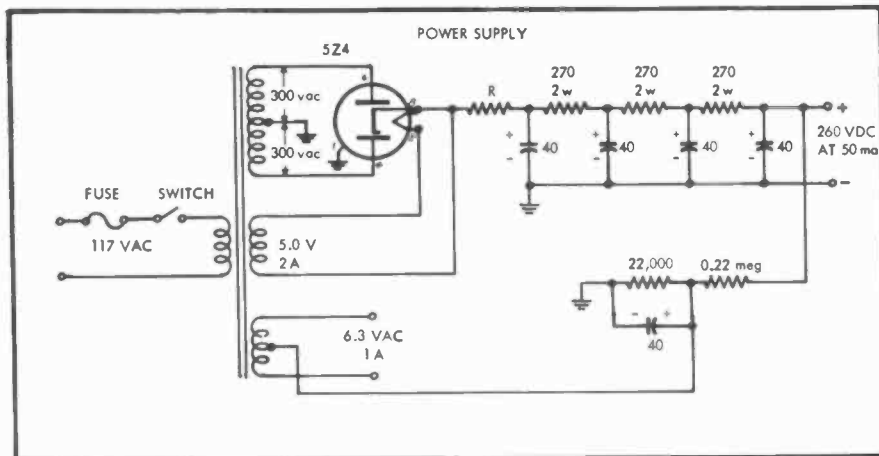
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[4] Charles W. Harrison, Jr., "Coupled loudspeakers," 3rd Audio Anthology, Radio Magazines, Inc. (1955) pp. 101–105.

Fig. 5. Power supply featuring RC filtering.

Effect of the Cathode Capacitor
on P-P Output Stage

ROBERT M. MITCHELL

Because of the fact that some amplifiers use a bypass capacitor across the cathode resistor in the push-pull output stage and some do not, the conclusions reached in this paper should be of considerable interest. Note, however, that these conclusions refer only to Class A amplifiers, while most modern amplifiers are designed to work up into the Class AB region.

WITH THE CONTINUED INTEREST in audio circuitry and design, it has become worthwhile for the audio engineer to re-examine some of the procedures that formerly were taken largely for granted. Among these procedures is the use of the by-pass capacitor in a Class A push-pull output stage. The purpose of this capacitor seems evident enough, yet the engineer is constantly finding circuits without this component and just as frequently finding circuits with it. To make the matter more puzzling still, he will read one author's admonition that its use is absolutely necessary, and another's that its use is specifically to be avoided.

In view of this ambiguity of opinion, it was felt that a test of the differences would be interesting. The results, which are not always as anticipated, are presented in this article, along with some evaluation of the outcome. The nature of this investigation necessitated the examination of some of the effects of balancing techniques in the output stage as well, and the results of this phase of the undertaking are also presented here.[1]

The Problem

Proponents of bypassed cathode operation maintain, among other things, that since the even-order harmonic terms which are generated in the output stage of necessity pass through the common cathode resistor, they therefore appear between grid and cathode of both tubes, are thus introduced as signal, and appear in the output. While it is quite true that these even harmonic terms do appear as a voltage drop from each grid to the

[1] The measurements described here were made at the laboratories of the United Transformer Company during the period when the author was director of those laboratories.
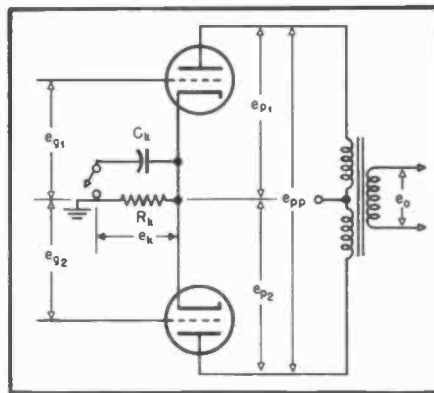

Fig. 1. Basic push-pull output stage arranged to switch capacitor in or out of circuit for measurements described.

common cathode, it is also true that since they appear as in-phase inputs at each grid, they are cancelled in the push-pull stage.

Another and more serious charge is that the even-order distortion terms of all kinds which pass through the cathode resistor may cross modulate with the input voltage and thus introduce additional intermodulation distortion terms which, not necessarily being in phase with each other, may appear in the output.

As a representative of the other ("un-bypassed") school of thought, Williamson states: (*Wireless World*, May 1947) "A feature of this arrangement is that the valves operate with a common unbypassed cathode bias resistor, which assists in preserving the balance of the stage under dynamic conditions," and in the August 1949 issue: "Due to the use of common unbypassed resistors for the push-pull stages, the amplifier is largely self-balancing to signal. . . ."

Here, then, are the two opposing views on the subject, and the reasons advanced for each by their respective proponents. Although the reasons advanced seem to be different, it is reasonable to assume that reduction of distortion is also the effect desired by Williamson, since this

is one of the results of a balanced stage.

The Initial Approach

Figure 1 shows the basic push-pull output circuit with signal voltages labeled and the bypass capacitor in question, C_k, shown in series with a switch which will allow it to be inserted or removed at will during the course of measurements. It should be noted that this is the common type of push pull circuit, and the entire discussion which follows is confined to this circuit and does not necessarily apply to the "single-ended" or "series d.c." type of push-pull circuit.

The actual amplifiers used in the test were the UTC W-10 Williamson amplifiers, since these conform almost exactly to Williamson's circuit. (See *Fig.* 2.)

First Tests

The first measurements made were of total harmonic distortion, using a General Radio Type 1932-A Noise and Distortion Meter, and a low-distortion oscillator. The input frequency was 50 cps and measurements were made at power levels from 4 watts up to overload. These initial tests were made on an amplifier using 5881's in the output stage, and gave very consistent results. The addition of the bypass capacitor always increased the distortion, regardless of power level. The increase was very slight, but nevertheless very definite. The total distortion was small, so it became very difficult to measure, especially since the distortion of the oscillator itself was of the same order of magnitude as that of the amplifier. Consequently, it was decided to measure the performance without feedback. This change would increase the distortion, of course, but would not affect the action of the capacitor. The results of this test were similar, with distortion increasing when the capacitor was used. (See *Fig.* 3). These results were obtained consistently after numerous checks and rechecks, in-

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cluding tests at higher frequencies (500 and 2000 cps). Several curves were drawn and the data was about to be assembled for write-up when it was decided to substitute some other type of output tubes and see if there was any difference. Accordingly, a pair of 1614's was substituted and the tests re-run. The results were as complete a reversal of the trend as could be imagined! Almost every test showed lower distortion with the capacitor *in* the circuit. A typical measurement is illustrated in *Fig.* 4. In view of such conflicting results, it was decided to re-measure with as many different tube types and amplifiers as possible. Consequently, the test schedule outlined below was evolved.

(1) Four different stock amplifiers with 1614 tubes were checked for total harmonic distortion on a distortion meter.

(2) The same four were checked for individual harmonic distortion components on a wave analyzer.

(3) The measurements of (2) were made with different degrees of current unbalance.

(4) One amplifier was checked with four different sets of output tubes, all of which are directly interchangeable in the UTC W-10. These four sets included two pairs of 1614's and one each of KT66's and 5881's.

(5) The amplifier of (4) was checked for intermodulation distortion with the three tube types mentioned above.

(6) The amplifier of (4) was examined for transient distortion by the square wave method, with differing degrees of current unbalance.

(7) All measurements were made with the 100-μf bypass capacitor switched in and out of the circuit and the comparisons made point-by-point on an A-B basis. Care was taken to prevent any transient disturbance during connection or disconnection of the capacitor from being included in recorded data.

Results

The outcome of the harmonic distortion measurements of (1) was very inconclusive at low frequencies. The addition of the capacitor either reduced or increased the distortion depending on which type of tube was used, and even varied among tubes of the same type. At the higher frequencies, however, the addition of the capacitor quite consistently increased the distortion.

This ambiguity of results pointed up the need for a more refined analysis, so the next step was to measure the individual harmonic components on a wave analyzer. A fundamental frequency of 50 cps was chosen, and the second and third harmonics were checked, with output tube currents adjusted to produce these four different conditions:

(1) Minimum unbalanced current

(2) Minimum 3rd harmonic in the output

(3) Maximum unbalanced current in one direction

(4) Maximum unbalanced current in opposite direction

Again the results were indefinite, with different tube types or different tubes of the same type giving different results, and no preponderance of results one way or the other. *Figure* 5 shows a graphic comparison of the results of this test for two particular pairs of output tubes.

An interesting, and rather unexpected, finding was that the capacitor made a greater relative change in distortion, the closer the system was to balance, with the greatest differences in the two conditions taking place when the currents were adjusted to produce a minimum of third harmonic. This is shown clearly in *Fig.* 5. The condition for minimum third harmonic was also found to occur very close to that for minimum second harmonic, so close in fact that the two were practically coincident.

For example, when the currents were adjusted for minimum unbalance or minimum third harmonic, addition of the capacitor produced changes of the order of two or four to one. When the currents were greatly out of balance, however, the addition of the capacitor caused changes of the order of only 10 per cent or less, although the distortion terms were much larger, of course. It was also found that the condition for minimum
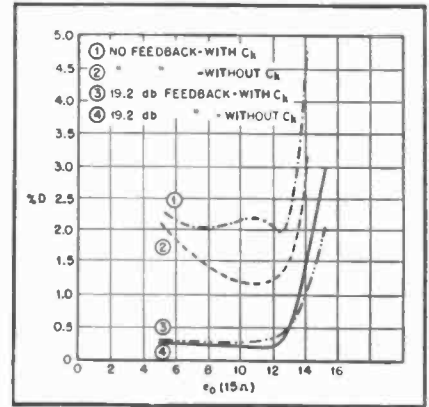

Fig. 3. Harmonic distortion curves on amplifier of Fig. 2. with and without bypass capacitor.

unbalanced current was generally *not* the condition for minimum harmonic distortion. This is not surprising, since the fact that the two tubes are in static balance (d.c. conditions) does not mean that they are also balanced dynamically (a.c. conditions). The purpose of the balancing arrangement in the "Williamson" amplifier is primarily to minimize the unbalanced d.c. current in the output transformer primary, and thereby increase the low-frequency response, while simultaneously reducing core saturation.

Intermodulation Distortion

The next step was to measure the intermodulation distortion of the amplifier, and since the method of measurement was not that most widely used, a brief discussion of the technique will be of interest.

Intermodulation distortion occurs when two or more frequencies interact so as to produce frequency components which are proportional to the product of the input frequencies. One result of such a relationship is the production of frequencies
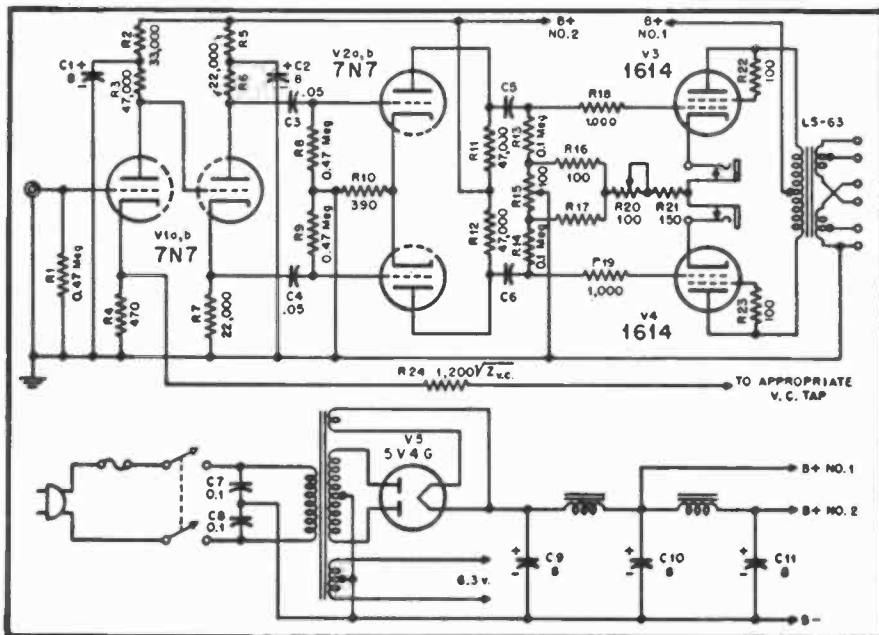

Fig. 2. Complete schematic of first amplifier used in these measurements.
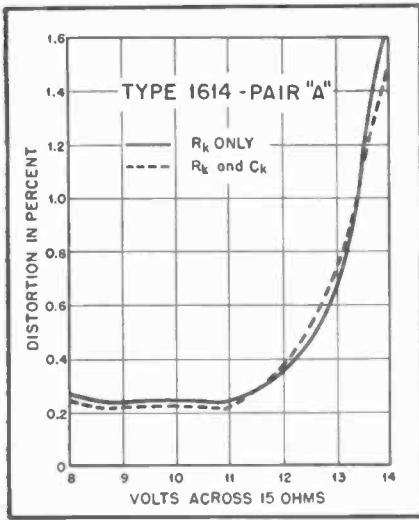
Fig. 4. Same measurements as Fig. 3 with different pair of tubes in the amplifier. These curves are plotted a different scale.

equal to the sum and difference of the two input frequencies, or of one input and harmonics of the other, etc. This action is the basis of our common system of amplitude modulation, where the production of such sidebands is a necessary feature of the process. In the commonly used (and most familiar) method of intermodulation tests, a high-frequency tone (the carrier) is modulated by a low-frequency tone, and the resulting sidebands detected and measured by a suitable circuit.

Although this method is an excellent indication of nonlinearity which generally correlates closely with listening tests, it was found that another method gave results which sometimes seemed to correlate even more closely. This second method of intermodulation distortion measurement also utilizes the fact that difference frequencies are produced. In this method, generally called the CCIF method, two high-frequency tones of equal amplitude and separated in frequency by a fixed amount are applied to the device under test. The output is measured by a wave analyzer tuned to one of the distortion frequencies such as $f_2 - f_1$, $2f_2 - f_1$, or $2f_1 - f_2$, and the distortion voltage is expressed as a percentage of the sum of the two input voltages. *Figure* 6 shows the two types of intermodulation in graphic form. One noticeable feature of the CCIF type is the location of the difference-frequency component $f_1 - f_2$. Since this component is so far removed from the two input frequencies, there is less tendency for the signal to mask the distortion as there might be in the SMPTE method, or in the CCIF method for the $2f_1 - f_2$ and $2f_2 - f_1$ terms. This difference tone is particularly objectionable when it falls in the region of maximum sensitivity of the ear (400 to 5000 cps), and is noticeable in amounts as small as a fraction of one per cent. It

must be pointed out, however, that due to the difference in method of expressing percentages in the two methods, equivalent percentages are *not* indicative of equivalent degrees of nonlinearity. The SMPTE values appear relatively high while the CCIF values appear relatively low for the same nonlinearity.[2]

In using the CCIF method of measurement two procedures are commonly used. One is to select two fixed input frequencies and measure the distortion as the power output is varied. The other is to select a fixed power output level and to vary the two input frequencies simultaneously, maintaining a constant difference frequency. The first is rather readily accomplished, whereas the second requires either a special oscillator such as the General Radio 1303-A, or two oscillators calibrated with sufficient ac-
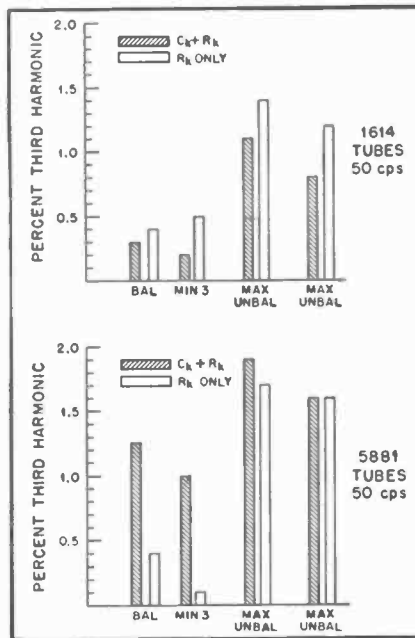

Fig. 5. Comparative distortions with different types of tubes in the output stage show different results.

curacy to enable the difference to be readily distinguished from the dial settings. The stock oscillator in the UTC laboratories is a decade-type oscillator with an accuracy of four places, and consequently is ideally suited to any such application.

Two such oscillators are used with the simple mixer-potentiometer circuit shown in *Fig.* 7 to provide control of frequency and output of either oscillator and the over-all voltage output. The isolating resistor networks prevent any interac-
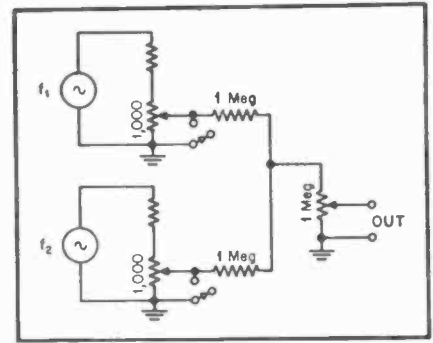

Fig. 7. Method of connecting two audio oscillators to provide CCIF type of IM signal without introducing intermodulation in the mixing circuit.

tion, and consequent intermodulation, between the two oscillators.

The CCIF intermodulation tests were performed using both methods mentioned above. When distortion was measured as a function of power level, the following relations were taken into consideration:

Since two frequencies are involved, a complex (nonsinusoidal) wave is produced, and the indication of an ordinary vacuum tube voltmeter is, therefore, not valid in determining the power level by the customary formula $P = E^2/R$. Furthermore, since one wave rides the other, the peak value of the two waves may reach a value equal to the algebraic sum of the individual waves, with the result that overload can occur for two frequencies when each is only one-half the amplitude required for overload by a single frequency. (See *Fig.* 8) Since each wave is only one-half the maximum amplitude, it can produce only one-quarter the power of a single maximum-amplitude wave. The power available from *two* such waves without any possibility of overload is, therefore, only one-half the power available from a single-frequency wave. The curves for the CCIF intermodulation tests are calibrated in terms of volts as read on an audio-frequency
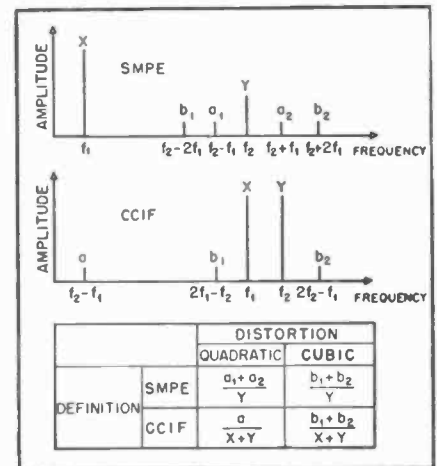
Fig. 6. Comparison of frequencies present in two different types of intermodulation measurements.

[2] A discussion of the two methods of intermodulation distortion measurement will be found in the General Radio Company Technical Publication B-3, *The Measurement of Non-linear Distortion*, by Arnold Peterson.
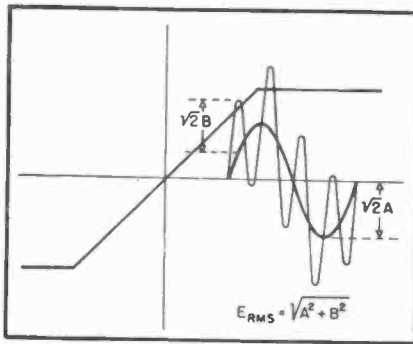
Fig. 8. Graphic representation of possible maximum voltage resulting from mixing two signals of slightly different frequency and of the same amplitude, as in the CCIF method.

Fig. 10. Effect of capacitor on CCIF intermodulation measurements when difference frequency is varied. (The three pairs of curves are plotted to different base lines.)

vacuum-tube voltmeter. In accordance with the above, maximum power level is equivalent to an output voltage reading of about 9.3 volts on these curves.

The curves of *Fig.* 9 were obtained by maintaining the indicated input frequencies constant, varying the input voltage up to overload and above, and measuring the first-order difference frequency. The tubes used were 1614's and the difference frequency was 400 cps. Notice that although the difference in distortion is slight, it is almost always lower when the cathode resistor is bypassed. The same results were obtained with 5881's and KT-66's.

The curves of *Fig.* 10 were obtained by keeping the input voltages constant and varying the input frequencies. This was done for three difference-frequencies, providing a considerable amount of range overlap as shown. The tubes used were KT-66's. Again it is seen that the capacitor effects a slight but definite improvement. (Note that each difference frequency has separate distortion ordinates in *Fig.* 10).

As a final check the amplifier square-wave response was observed for different amounts of current unbalance. With unbalanced currents up to 10 ma in either direction there was no discernible difference in the output wave shape as the capacitor was added or removed at fre-
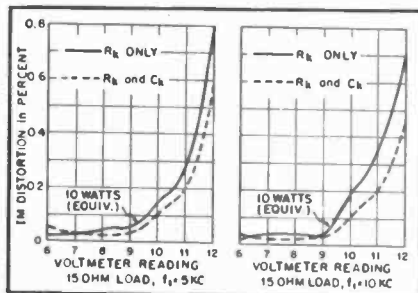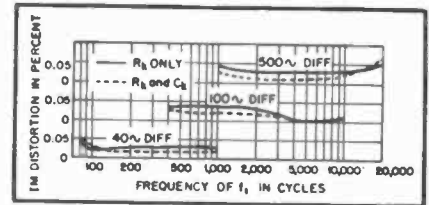
quencies from 20 to 20,000 cps. At low frequencies there was a noticeable rounding of the trailing edges as the unbalance exceeded 10 ma. (See *Fig.* 11) These results were obtained with any of the three types of output tubes.

From the foregoing experiments at least one curious result stands out: In a Class A amplifier the use of a bypass capacitor across the output cathode generally reduces the intermodulation distortion, although it may either decrease or increase the harmonic distortion.

The decision as to whether or not to use such a capacitor depends mainly on the magnitude of the distortion. If it is very small, then it may be safely left off, with no possible audible difference. If it is only moderately low, then the use of a bypass capacitor is advisable.

All of the foregoing applies to a Class A amplifier only. In the case of a Class AB amplifier the bypass capacitor is absolutely necessary if the amplifier is to perform within the modern limits of high fidelity performance.

CATHODE CAPACITOR

(Continued from preceding page)

Fig. 9. Effect of capacitor on CCIF intermodulation measurements at two different areas of the frequency spectrum.
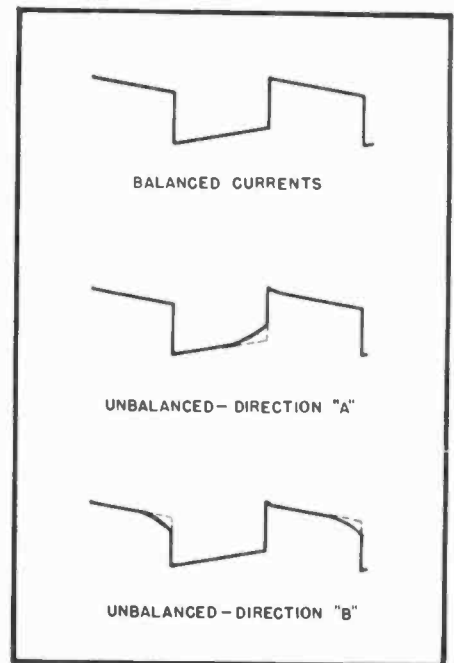

Fig. 11. Effect of current unbalance on square waves. Note that only the low frequency is affected, as indicated by rounding of trailing edge of wave.

What's All This About Damping?

N. H. CROWHURST

An engineering discussion of the elements entering into the effects of variable damping in an amplifier when the loudspeaker itself cannot be complemented accurately and completely.

IN RECENT MONTHS, much has been written about variable damping, ultimate damping, and various aspects of damping—principally concerning its application to the coupling between an amplifier and a loudspeaker. In an endeavor to clarify the general understanding of this subject, let us consider what damping means in a somewhat broader sense.

Let's start, for example, with damping as applied to musical instruments, where electronics does not enter into the picture at all. When a piano string is struck by the piano hammer, it continues to vibrate for a considerable period, especially if the check action is held off by holding the piano key down. This indicates that the Q of the resonant system is very high. It's true that considerable sound energy is radiated, but in comparison with the energy stored in the vibrating string the radiation is small because this energy is not radiated directly from the piano string.

In illustration of this fact, the writer well remembers listening to a piano which was not provided with the regular sounding board. This piano has been designed for use with electronic pickups, so the quality of sound could be entirely under electronic control. When this piano was played without the amplifier switched on, its music could only be heard by putting the ear close to the instrument. Just sitting in the same room with the piano, one would imagine that the musician was pretending to play it rather than actually depressing the keys. This shows that in the normal type of piano the principal radiation of sound comes from the sounding board, to which it is transmitted from the strings' supports.

Having realized this fact, consider how the vibration of the piano string may be damped. Application of damping to the sounding board has very little effect. It may be possible, applying some damping material to the sounding board, to considerably reduce the radiation of

sound, but it will not materially damp the vibration of the string. On the other hand application of the felt provided on the check action of the piano to the string itself, will damp the vibration of the string almost instantly. Touching the string with the finger while it is vibrating will also damp its vibration quite rapidly.

A number of other musical instruments could be similarly discussed. The principal things that we can learn from a consideration of these phenomena are two. First, a large surface is required to radiate sound into the air, because only in this way can satisfactory acoustic matching between the vibrating medium and the air load be achieved; a small vibrating element such as a string does not move the air, it rather cuts through it. Second to produce satisfactory damping, the damping agent must be applied at a suitable point sufficiently close to the vibrating medium itself. Although the vibrating medium is coupled to some extent to the sounding board, damping of the sounding board can only damp the movement of the string to the same extent as it is coupled to it. Because the coupling is what we would term in radio very loose, the damping that can be effected in this manner is extremely small.

Before turning to the discussion of loudspeakers and their damping, let us consider briefly two other analogies that will prove useful in helping to visualize the various components that make up our problem.

The first is a transmission line. A transmission line has a characteristic impedance. If the line is terminated by its correct matching impedance all the transmitted energy is absorbed when it reaches the receiving end, but if the line is not correctly matched some of the energy is reflected and travels back along the line. Correct matching of the transmission line can be considered as correct damping, because it will prevent reflections from occuring.

The other analogy that we can con-

sider is a transformer. The particular properties with which we are concerned are the primary inductance and the leakage inductances of the transformer, together with the secondary winding capacitance. For simplicity we will consider the transformer to be of 1:1 ratio. *Figure* 1 shows the equivalent circuit of a transformer, with the elements in which we are interested shown. The transformer can be a resonant circuit in several ways. The primary inductance can resonate with some capacitance in the primary circuit; similarly the same relative inductance can be resonated in the secondary; or the combined inductance can be resonated with capacitance part of which is in the primary and part in the secondary. All of these resonances are of the same basic type in which the inductance element being resonated is the primary inductance, but it is also possible to resonate the leakage inductance between primary and secondary with a capacitance either in the primary or secondary.

Consider the particular case of leakage inductance resonating with capacitance in the secondary circuit. Although the capacitance is physically connected in parallel with the transformer secondary winding its effect is very different from a similar capacitance connected in parallel with the primary winding. Short circuiting of the primary will result in maximum Q of the tuned circuit, because any resistance in series with the primary appears virtually as resistance in series with the tuned circuit. This is illustrated in *Fig.* 2. This kind of resonant circuit can be damped with either a resistance in shunt with the secondary winding, which provides shunt damping for the tuned circuit, or a resistance in series with the primary winding which provides series damping for the resonant circuit
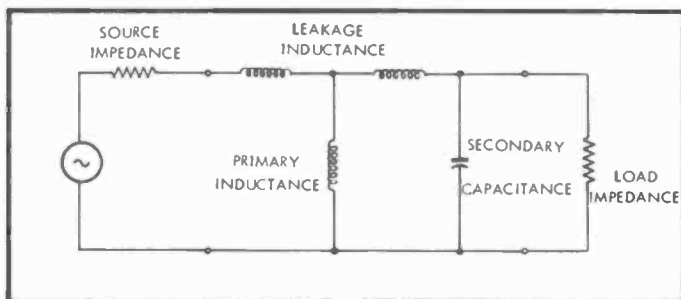
Fig. 1. Equivalent circuit of a transformer, to show possible resonances. For convenience the transformer is assumed to have 1:1 ratio.
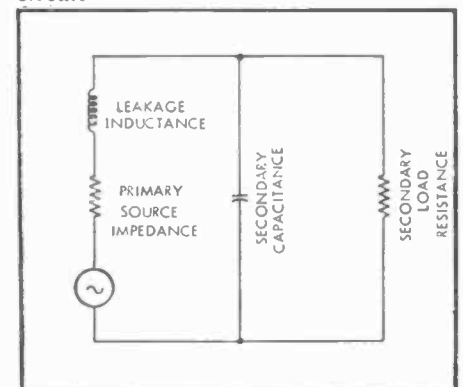
Fig. 2. Rearranged equivalent circuit for resonance between leakage inductance and secondary capacitance.
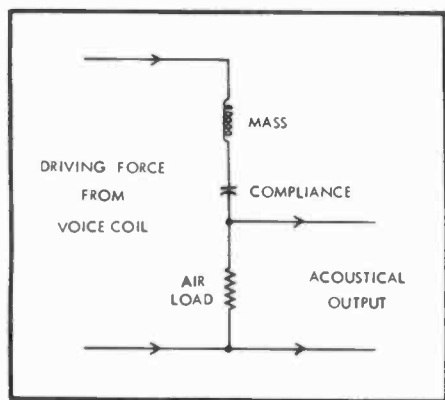
Fig. 3. Simplified equivalent circuit of acoustical action of a loudspeaker.

In this respect a transformer can be considered as somewhat similar to a quarter wavelength of transmission line, when the transmission line to which we referred previously is incorrectly terminated. If the termination value is too high, at a distance back along the line equivalent to a quarter wavelength of the frequency being transmitted the reflected impedance becomes the inverse of the terminating impedance, using the characteristic impedance as a mean.

Back along the line at consecutive quarter wavelength intervals, the impedance will change alternately between one that is high and one that is low compared with the characteristic impedance. Similarly, if the termination instead of being incorrect in resistive value is incorrect by being a reactance instead of a resistance, the apparent impedance measured at quarter-wavelength intervals back along the line will alternate between inductive and capacitive reactance.

If the line is loss-free, which corresponds with high Q conditions in a mechanical arrangement, then this transfer of impedance will go on indefinitely at quarter-wavelength intervals, without changing the relative magnitude of impedance at each half-wavelength interval. But practical lines introduce a certain amount of electrical loss, and for this reason each alternation away from the characteristic impedance of the line deviates by a decreasing amount from this value, and after a sufficiently long length of line, the measured impedance will become sensibly equal to the characteristic impedance. However, in such a line the loss becomes considerable and a relatively small amount of the transmitted energy will reach the receiving end.

The Loudspeaker

Having briefly discussed some of the analogies we can apply, now let's turn to the consideration of a loudspeaker. Basically, what we are concerned with in a loudspeaker is driving a column of air which communicates with the room so as to recreate a desired sound field.

Let us first consider the acoustical arrangement that achieves this. It consists of a loudspeaker diaphragm in contact with the air in front of it (which communicates with the room) and also

with the air behind it which is contained in the loudspeaker enclosure.

This can constitute a resonant system which can be analyzed by using the same mechanical terms for the elements of a resonant system as would be applicable to a vibrating piano string. The mass of this resonant system consists of the mass of the moving diaphragm and the voice coil attached to it, together with the mass of the small quantity of air adjacent to it, which can be regarded as having to move with it. The compliance consists of the compliance of the diaphragm surround and also the compliance of the centering spider, if one is used, together with the compliance of the air inside the enclosure behind the diaphragm. This latter will act basically as a compliance rather like the air inside a Helmholtz resonator. These are the reactive elements of the resonant system. The resistance elements provide damping and prevent it from having a natural vibration of its own in a well designed system. These are the viscosity of the air in the enclosure, the viscosity in the compliance of the surround and centering spider, and the radiation resistance coupled to the diaphragm by means of the air load which it drives to radiate energy into the room.

So far we have just considered an acoustical resonant system. Now we come back a stage further, to consider a mechanical-acoustical relationship. This rather corresponds—but in different proportions—with relationship between the vibrating piano string and the sounding board. The piano string is the basic driving force, but the sounding board is the element that radiates sound energy into the air. Similarly in a loudspeaker, the voice coil is the basic driving force but the diaphragm is the element that radiates the sound into the air. So the coupling between the voice coil and the diaphragm has to be considered as part of a loudspeaker system.

In the preceding discussion we considered the voice coil as if it were rigidly coupled with the diaphragm, but this idea ignores the fact that the diaphragm has to be constructed of some material which cannot be absolutely rigid. The diaphragm material itself has mass and compliance distributed over its entire surface area, and hence all parts of the diaphragm do not have to vibrate in an exactly similar manner.

This means that the diaphragm can behave somewhat after the fashion of a transmission line. A driving force is applied at the voice coil end of the diaphragm and is transmitted outwards toward the periphery where the surround is located. It's true that at the lower frequencies the time taken for the wave to be transmitted this distance corresponds with a small fraction of a wave period. But at higher frequencies an appreciable fraction of a wavelength is involved and various cancellation effects can arise which interfere with the frequency response of the loudspeaker.

For this reason a loudspeaker of good design has the surround viscously damped, either by choice of suitable configuration for the surround, or by application of some impregnation which will have the desired effect. This will prevent such cancellation effects due to transmission of sound waves to and fro radially in the diaphragm itself before they get transmitted to the air in contact with it.

There is another fashion in which the diaphragm can set up resonances of its own—by setting up local vibration patterns similar to Chladni figures in a vibrating plate. At higher frequencies the coupling of the air load to the diaphragm improves, and for this reason it is easier for the diaphragm to vibrate in a complex manner than to follow the driving force as a single element. In this case the diaphragm does not have to move as much air, because parts of it will be traveling inwards while the other parts are traveling outwards; this enables the air to oscillate sideways across the surface of the diaphragm instead of having to drive in and out as a single large mass.

These break-up resonances again depend upon the characteristic mass and compliance of the material of which the diaphragm is made. They can, however, be minimized by attention to the construction of the diaphragm, by the introduction of circular reinforcing indentations into the material, and by other means. However, as regards our damping feature, the important thing to notice is this: these resonances occur in the diaphragm material itself and cannot logically be damped out either by acoustical damping in the enclosure or by some form of damping in the voice-coil circuit, because they take place between the driving force provided by the voice coil and the loading force provided by the air. Nothing beyond these limits can affect the behavior of the break-ups.

Resonances

We have now considered the behavior of a loudspeaker, from the air column that it has to drive into the room back to the voice coil, which so far we have considered merely as a driving force. The basic resonance at this point consists of the mass of the diaphragm with all its appurtenances and some of the air which moves with it, together with the compliance of the surround and of the air in the enclosure, acting as a single resonant arrangement.

The natural frequency of this resonance is usually somewhere between 35 and 125 cps, varying according to the particular design of loudspeaker and the size and type of enclosure used. This is the resonance about which designers are concerned when they talk about damping applied in amplifiers. It should be noted that this is not necessarily the only resonance in a loudspeaker, but it is the principal one.

We now come to the point where the transformer analogy is useful. The voice coil and the loudspeaker magnet system constitute an electro-mechanical coupling unit, the purpose of which is to transfer electrical energy from the amplifier into energy in the voice coil.

This part of a loudspeaker is essentially similar in basic principles to an

electric motor, in which electrical power delivered to the terminals of the motor is converted into mechanical driving power at the motor shaft. A question of some importance in consideration of damping is the efficiency of this device as a motor. If the device were 100 per cent efficient, then whatever we do to the mechanical-acoustical part of the system will be completely reflected into the electrical circuit, as measured at the terminals of the voice coil. But it is well known that commercial loudspeakers range up to a maximum efficiency in the region of 20 per cent, some samples being no better than 5 per cent efficient. This means that it is impossible to damp a diaphragm electrically if it is inadequately damped mechanically or acoustically. It is rather like trying to damp the piano string by means of damping applied to the sounding board.

Figure 3. shows the basic view from the acoustical side of the resonant system we are considering. The mass, compliance, and radiation resistance may be regarded as three elements in a series resonant circuit. More energy can be radiated at the resonant frequency, because the diaphragm moves freely at this point. Below this frequency the system is compliance controlled and the energy radiated is restricted by the compliance, whereas above this frequency it is mass controlled and the energy radiated is restricted by the mass that the driving force has to move.

Transferring this circuit back to the electrical side, however, a transformation occurs similar to that noted in the transformer. The equivalent circuit looks like a parallel resonant arrangement in series with the electrical characteristics of the voice coil. If the voice coil is locked in position by wedging it, its electrical impedance is simply the voice coil resistance and inductance. However, when it is allowed to move in driving the diaphragm, some of the acoustical resonance gets reflected into the electrical circuit, and this appears as a resonant circuit in series with the voice coil inductance and resistance as shown at *Fig.* 4.

Only a fraction of the acoustical values get reflected into the electrical circuit, because of the limited efficiency of the electro-mechanical coupling arrangement. Suppose, for example, that the coupling efficiency is 20 per cent and that the effective Q of the mechanical-acoustical part of the set-up is around 10 after the loudspeaker has been mounted in its enclosure. This complete resonant arrangement will be reflected into the electrical circuit with its series-parallel transformation in the same proportions. So we will have a shunt-tuned resonant circuit having a Q of 10 in series with the voice coil resistance and inductance.

Now assume that we apply something electrically to this circuit to reduce the effective Q to the point of critical damping, so as to render it an aperiodic circuit. Looking back from the mechanical-acoustical side, only 1/5th of this damping effect will be reflected, so that while the Q of the electrical equivalent may

be reduced to unity, the Q in the electro-mechanical system will probably be reduced from its original value of 10 to about 8.

Of course, if the original acoustical-mechanical resonance had a Q in the region of 2 or 3 it is altogether possible that careful attention to the electrical circuit could adequately damp this resonance, but the value necessary would still be considerably greater than the value necessary to produce aperiodic damping, measuring the resonance as an electrical one in the voice coil circuit.

The analogy with the transformer is somewhat reversed: in the case we considered, what appeared to be a shunt-tuned circuit on the secondary of the transformer was transposed to an apparent series tuned circuit on its primary; in the case of the loudspeaker the equivalent mechanical-acoustical resonance looks like a series tuned circuit, which the electro-mechanical coupling transfers to appear as a shunt tuned circuit in the voice coil circuit.

This is not the only difference between the analogy of the transformer and the electro-mechanical coupling of the loudspeaker: in the case of the transformer its coupling efficiency will usually be reasonably high—probably better than 80 per cent; consequently damping can, in all probability, be applied with equal effectiveness either as a shunt resistance across the secondary or as a series resistance in the primary circuit. If however the transformer were deliberately made inefficient, for instance by using an arrangement where the primary and secondary coils were removed on different limbs of the magnetic core, the story would be different, and would perhaps be more analogous to the action of the loudspeaker. Then it will be found necessary to apply damping to a secondary resonant circuit *in* the secondary circuit, because the coupling to the primary circuit would become low and it would be impossible to damp the resonance adequately in the primary circuit. This is what usually happens in the electro-mechanical arrangement of a loudspeaker.

Conclusions

What conclusions can we draw from this discussion? First, that it is desirable to damp any resonant system in the "circuit" where the resonance occurs. This fact is well recognized in radio circuitry, but appears to have been overlooked somewhat in application to elec-

tro-acoustical mechanisms. The basic resonance in a loudspeaker is a mechanical-acoustical one, and there is really no substitute for taking adequate care of this resonance in the mechanical-acoustical part of the system.
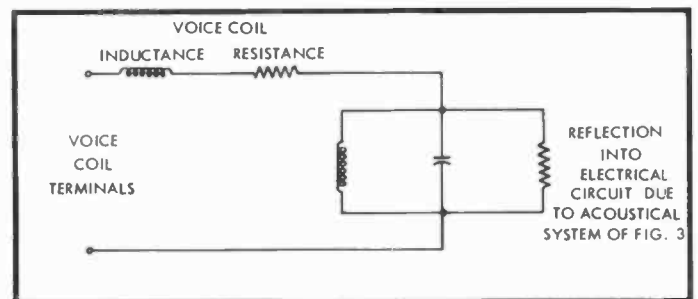
Secondly, if this resonance has not been adequately cared for in the proper place, attempts to provide electrical damping in the voice-coil circuit lead to conflicting requirements. The usual method of determining critical damping is by measuring the signal waveform across the voice coil, when some kind of transient is applied. It is true that selection of suitable damping will produce a condition where the waveform applied to the voice coil appears to have eliminated the resonance; but this has only provided critical damping for the equivalent electrical circuit reflected through the electro-mechanical coupling of the loudspeaker. This merely means that the driving current in the coil has now had the effect of resonance eliminated from it; but it does not eliminate the residual resonance that still may be in the acoustical-mechanical part of the loudspeaker.

The claim has been made that complete neutralization of the voice-coil resistance, by making the amplifier look like a negative resistance, can remove all distortion introduced by the loudspeaker itself. The foregoing discussion will show that such a method of damping cannot damp the break-up resonances that may occur at higher frequencies, if design of the loudspeaker has not taken care of this feature. Nor can it completely damp the fundamental resonance if the acoustical-mechanical system has too high a Q.

Further than this a loudspeaker may introduce distortion due to nonlinearity of its compliances, or due to nonlinearity of the law relating deflection of the voice coil to applied drive current. This nonlinearity can be due to nonlinearity of the magnetic flux in the voice coil air gap. If there are nonlinearities of this nature, the loudspeaker driving force will be nonlinear and the driving waveform can be distorted. Application of what has been called ultimate damping cannot cancel this kind of distortion.

In this article, the author's objective has not been to discount completely the advantages of variable-damping amplifiers but rather to clarify many of the claims that are currently being made and to enable readers to assess truly the value of variable damping and to approach the problem from a correct understanding.

Fig. 4. The reflected components due to the acoustical system have the same resonant frequency and Q as the acoustical system, but scaled down in value, due to the inefficiency of electromagnetic coupling with mechanical systems.

VOICE COIL

INDUCTANCE   RESISTANCE

VOICE COIL TERMINALS

REFLECTION INTO ELECTRICAL CIRCUIT DUE TO ACOUSTICAL SYSTEM OF FIG. 3

Electrical Adjustment in Fitting a New Output Transformer

If you have to replace an output transformer, it is a good idea to check up on its performance to make sure that you are getting as much out of your amplifier as you were before the change. Here are the steps to take.

NORMAN H. CROWHURST

THE OUTPUT TRANSFORMER is usually one of the more reliable components in an amplifier but occasionally one will go bad, developing a short circuit or open circuit, or maybe just shorted turns; in which case it becomes necessary to effect a replacement. Often, for various reasons, an exact replacement is not available: for example, by the time an output transformer goes bad it is quite probable that the particular amplifier is no longer being currently manufactured; however most transformer manufacturers make quite a range of output transformers from which it should be possible to select one having the right nominal ratings to suit the amplifier in hand.

But the fact that two output transformers have the same nominal ratings, in impedance ratio, power handling capacity, and frequency response, is no proof that they will behave equally well in the same amplifier. When a substitute transformer is connected into a modern feedback amplifier it may oscillate its head off or it may stay stable, but even when it is stable it is probable that the response and other aspects of the performance of the amplifier differ from the original. So it is well to make some checks and, if necessary, electrical adjustment to get the amplifier performing approximately according to its original specification.

If Amplifier Oscillates

If the amplifier oscillates with the replacement transformer, the first thing to do is get it stable. Try a small capacitor from plate to plate, say 100 micromicrofarads. If this makes no difference to the oscillation except possibly changing its frequency (if this can be observed), then remove the capacitor and try changing the value of the phase-shift capacitor already connected in the amplifier. This is the capacitor connected across one or other of the feedback resistors.

First try removing it and, if this does not stop the oscillation, try substituting values different from the original by one or two steps in the preferred value range. For example if the original capacitor was 68 micromicrofarads try a 100 or 150 and then try a 47 or 33. Usually one of these adjustments will make the amplifier stable so that measurements can be conducted.

Tests to Make

In most instances however, the new output transformer will not make the amplifier oscillate (perhaps it would be better if it did, because then the fact that something was different would be a little more obvious). Usually the difference in amplifier performance, caused by the change in output transformer, is a little more concealed. So set the amplifier up with a resistance load connected to the output, in place of the usual loudspeaker, and make the following measurements.

Distortion Characteristic

Set the audio oscillator to 1000 cps and if necessary use a 1000-cps filter between the oscillator and amplifier to remove any residual harmonics in the oscillator. Use a harmonic distortion meter to check the distortion present in the output from the oscillator after filtering. This must be less than the lowest amount of distortion you expect to measure from the output of the amplifier. Then measure the distortion in the output. See *Fig.* 1.

Most harmonic distortion meters provide a scope output so that the residual harmonic can be observed on a scope. Connect a scope to this. When the harmonic distortion meter is set to the calibrate position, so the scope displays the fundamental, adjust the scope time base so that a single sine wave is displayed on the screen. Then, when you switch over to measure harmonic, the trace on the scope will indicate the dominant orders of harmonic.

This can sometimes be useful in tracking down the cause of distortion, but a more useful aspect is that it checks whether the reading obtained is actual harmonic distortion, or hum. When the scope time base is set this way, so as to show only a single sine wave of fundamental, harmonic distortion will show up as a single trace on the screen, with a somewhat distorted waveform and having two or more cycles across the screen. On the other hand hum is evident by the fact that the trace is unsteady or a number of traces appear vertically displaced from one another.

From this it is possible to estimate how much of the measured distortion is hum and how much is harmonic. *Fig.* 2 illustrates this for a typical case: the
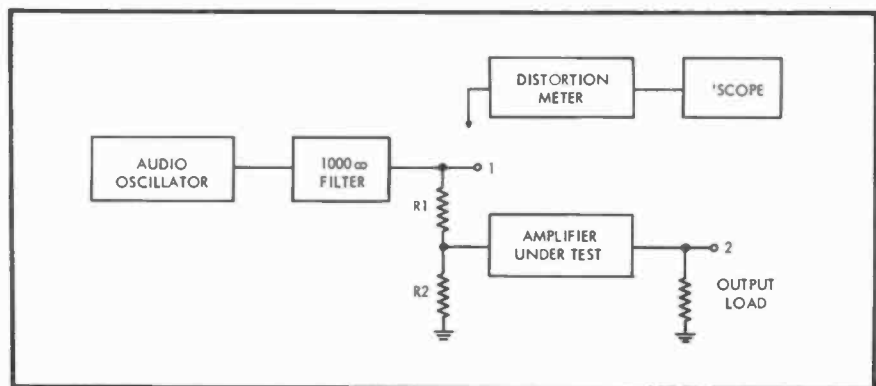

Fig. 1. The arrangement for making a distortion characteristic. First the waveform going in, at point 1, is checked, then the output waveform, at point 2. The resistors R_1 and R_2 form an input attenuator, so the level is not too low measure at point 1, but is right for the amplifier input at the junction of R_1 and R_2.
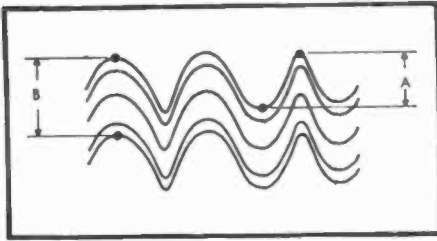
Fig. 2. How to interpret the waveform displayed on the scope, when the distortion meter is set to measure harmonic percentage: this waveform represents a residual with dominant third harmonic, plus higher order components, whose peak-to-peak amplitude is A; the multiple trace shows that a hum component is also present, of peak-to-peak amplitude B.

amplitude deviation of an individual trace, represented by A in *Fig.* 2, is the peak-to-peak voltage of harmonic distortion present in the residual output; the amount by which the whole trace fluctuates up and down, represented by B in *Fig.* 2, is the peak-to-peak hum voltage present in the residual output.

By estimating the relative components of each it is possible to deduce the actual harmonic distortion and hum voltage separately without the use of filters to actually separate them. Remember that the voltages combine, for measurement purposes, approximately on a root-mean-square basis. So, if the output reading is, say, 0.3 percent and the values of *A* and *B* measured on the scope trace are approximately equal, there will be about 0.2 percent of hum and 0.2 percent harmonics. The harmonic can be plotted in the form of a distortion characteristic as at *Fig.* 3.

Distortion

The next thing to check is the distortion at low frequencies, say 60 cps and, if possible, on down to 20 cps. Usually there will be difficulties in using a harmonic meter down at these frequencies, partly because it is difficult to get the oscillator output sufficiently free of distortion at these frequencies. However, fortunately the amount of distortion likely to be present under incorrect operation is considerably greater, so it is satisfactory to check distortion at the low-frequency end by looking at the waveform on a scope.

First check that the input waveform is a satisfactory approach to a true sine wave and then look at the output waveform (at full power output).

Another low-frequency defect that change of output transformer can set up is low-frequency instability or near instability. Instability will show up on the scope by a slow up and down movement of the trace. Near instability will cause the trace to "bounce" when input to the amplifier is changed or keyed.

Next make a check of distortion at high frequencies. Using a sinusoidal input,

watch the scope on the output as the frequency is swept way up to 20 or 30 kc at full output. See that the output waveform does not "fold over" anywhere over the frequency range from 20 to 20,000 cps (and higher, if possible).

Square Waves

The last but not the least check to make with a new output transformer installed is on the reproduction of square waves. A suitable square wave can be obtained from a number of sources. Some audio signal generators have a provision for changing from sine-wave to square-wave output. Another useful piece of equipment that will give square waves is the electronic switch used for providing two displays simultaneously on an oscilloscope screen: by turning the bias control over to one side and not connecting any inputs to the input terminals, the output becomes a square wave whose frequency can be adjusted by the frequency control on the electronic switch.

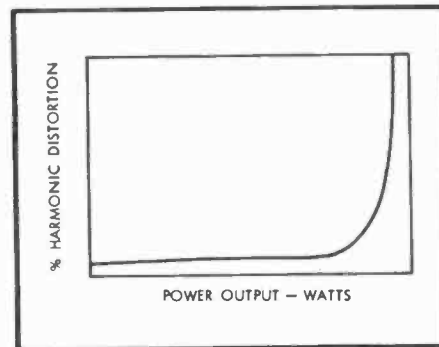First check the wave shape of the square wave going into the amplifier.

Fig. 3. Typical output/distortion characteristic, obtained with the method illustrated in Fig. 1.

A frequency of about 1000 cps is satisfactory for this purpose. Then check the waveform on the output, first with the resistance load still connected and then with a loudspeaker load. With a resistance load there should be no appreciable overshoot or ringing as shown in *Fig.* 4. With a loudspeaker load the waveform will definitely depart from square due to the reactances in the load, but there should not be any appreciable overshoot or ringing even under this condition.

Correcting Deficiencies

Having outlined the tests to be made we will now discuss what to do about rectifying any faults that show up under each test in turn.

First, if too much distortion shows up at mid-frequency: what usually happens in this case is that the distortion characteristic begins to rise earlier than the amplifier specification shows, as indicated at *Fig.* 5. If the original rating at which 1 percent distortion shows is 50 watts and the replacement transformer

drops the 1 percent point to, say, 47 watts, one has to decide whether or not this is acceptable.

If there are no other detrimental effects from the transformer, the difference between 47 and 50 watts is only 0.3 db, which no one is ever going to be able to detect audibly. On the other hand, it is just possible that a check is required for some reason to show that the amplifier still performs to specification. In this case steps are necessary to raise the output to the full 50 watts.

First check that the transformer ratio is correct for the impedance transformation it is supposed to produce. The formula for this has been published a number of times, also charts to facilitate the calculations. The voltage ratio can be readily obtained by measuring the primary and secondary voltages with an a.c. voltmeter using a steady signal of a 1000 cps going in from the audio signal generator.

If the ratio checks as being correct, the next step is to determine where the loss of power is occurring. First check that the voltage reaching the plates of the output tubes is according to specification. If the voltage actually reaching the plates is lower than the specified value, but the B+ is correct, this indicates that the primary resistance of the transformer is higher than the original.

The way to overcome this is to boost the B+ supply a little. This is not too easy to do as a rule because amplifiers usually push the B+ supply pretty well to the limit of the components used. Most modern amplifiers use a capacitor input filter and a few more volts can usually be achieved by increasing the reservoir capacitor value. This however will often exceed the rectifier dissipation so that it is necessary to double up on the rectifier.
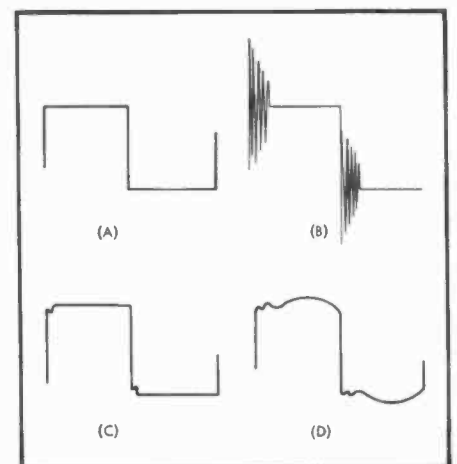

Fig. 4. Typical waveforms associated with square wave testing: (a) the input wave; (b) a typical wave when there is excessive ringing; (c) a waveform like this, with a resistance load on the output, is acceptable, and should be the target; (d) when the loudspeaker is connected, the waveform should be something like this.

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If, for example, a 5U4 rectifier is used, it would be advisable to use two 5U4's in parallel with twice the reservoir capacitor. This means an additional filament supply for the second 5U4, which will have to be provided by means of a separate filament transformer—using resistance in series if a 6.3 volt transformer is used with the 5 volt filament, to adjust the actual filament voltage to exactly 5 volts. This is shown in *Fig.* 6. Also watch that the voltage rating of electrolytics is not exceeded in making this change. Use new components with higher rating, if necessary.

If the plate voltage is not lower than the schematic indicates and the B+ is normal, it may be that the resistance of the secondary winding is too high, causing loss of some of the power. This *can* be checked by transferring the resistance load from secondary to primary. Obtain a plate-to-plate resistance of the rated loading value and sufficient dissipation and connect this across the primary side and again make measurements of power. This will check whether the loss is in the transformer windings or not.

If it is not easy to make this check because suitable resistances are not at hand, it is probably hardly worth going to the trouble of getting resistances, because there is little you can do about it except obtain another transformer. Do not try boosting the B+ voltage in this case, because you will be over-running the output tubes and, as stated earlier, the loss of 3 watts in 50 is quite unimportant anyway.

Inadequate Inductance

If distortion shows up at low frequencies the most probable explanation is that the output transformer has inadequate magnetic core to sustain the low-frequency amplification without running into saturation. However, as explained in the previous article, distortion at low frequencies is not always due to saturation of the core.

It can also be due to inadequate inductance of the primary. This can be
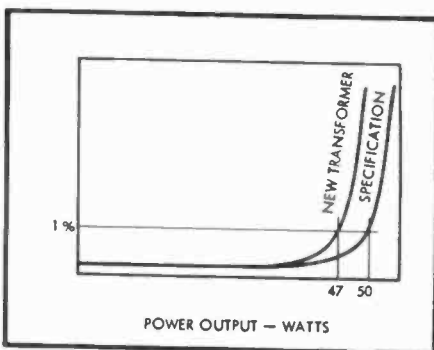
Fig. 5. Typical relationship between the published specification for distortion characteristic, and the measured figures, when the new output transformer has resulted in loss of efficiency, due to some cause that should be identified.

checked by the method described in the previous article, and if the waveform displayed proves that the trouble is due to deficient inductance rather than saturation, it is possible that a change in output loading may improve the performance at the low-frequency end. It will, of course, deteriorate the maximum output power throughout the whole frequency range.

The instability or near instability is also due to difference in primary inductance. It can sometimes be rectified by changing coupling capacitor values in the amplifier. The only way to work out positively what correction is required here would be to calculate out the stability criteria—equivalent to complete design! Trial and error approach is usually quicker.

If the output transformer comes with 4-, 8-, and 16-ohm taps try connecting an 8-ohm load first to the 4-ohm tap and then to the 16-ohm tap, to see whether the distortion changes at low frequencies, as compared with the distortion turnover or clipping point at middle frequencies. If so, it is possible that use of the amplifier with deliberately incorrect loading of this nature may produce better results.

However, before deciding to do this, check the distortion characteristic with the changed load at the middle frequency: you may find it better to live with the low-frequency distortion, or you may try for a better transformer.

High-Frequency Distortion

Next, if you get distortion at high frequencies, it is probably due to some kind of resonance in the output transformer *that should not be there.* The straight resonance between primary shunt capacitance and leakage inductance may modify the frequency characteristic and the performance on square waves, but it should never prove sufficient to cause distortion of the fold-over variety at the higher frequencies.

Where this occurs it is probably due to primary-to-secondary capacitance introducing a resonance. This is unlikely to occur in a transformer with a secondary impedance no higher than 16 ohms, but some output transformers provide for constant-voltage-line distribution with a higher impedance output—say 500 ohms. Effective capacitance between the primary and this winding can cause trouble in a modern feedback amplifier.

Usually some point on the secondary has to be grounded as part of the feedback arrangement. The point to be grounded will depend upon the phase of signal. Usually the capacitance between primary and secondary causes distortion because the wrong point on the secondary is grounded. If we ground a different point on the secondary, however, the feedback will be in the wrong phase, so
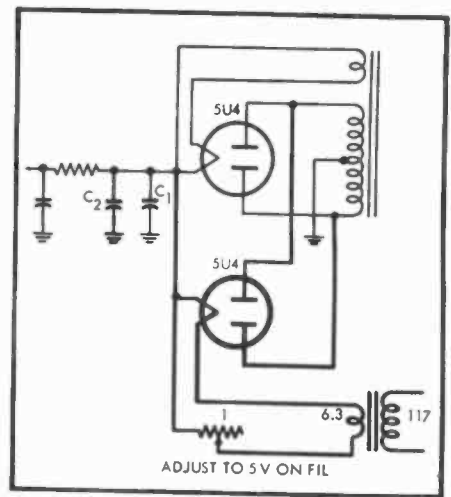
Fig. 6. Additions necessary to typical amplifier to boost B+ voltage. The extra components are shown in heavy lines. The extra capacitor should have the same value as the original C_1. Check that ratings are not exceeded.

the solution is to change the connection from the primary: connect the plate leads to the opposite output tubes; in this way the secondary connection can also be reversed in phase and this will usually obviate the effect. In an Ultra-Linear output stage, the screen leads must also be changed when the plate leads are reversed.

Ringing and Overshoot

The commonest and most important effect that the change of an output transformer can have on the performance of an amplifier shows up on the square-test. You may wonder why this test was reserved till last. The reason is that if any adjustments have to be made to satisfy the other conditions the square-wave performance will also be modified. If adjustments were made at the outset to correct the performance on square waves and then other adjustments prove necessary to eliminate distortion at low or high frequency it would only be necessary to conduct the square-wave test all over again with further adjustments. This was reserved till last so that when this test is made and the circuit modified as necessary the amplifier is complete.

Ringing or overshoot on the output under this test shows that the amplifier has a tendency to peak at some supersonic frequency. The remedy is the same as that described for oscillation at the beginning of the article, but more careful adjustment is necessary to eliminate overshoot or ringing than just to keep the amplifier stable.

First try the effect of different capacitors across the primary of the output transformer. If any capacitance at all across the primary increases the overshoot or ringing then don't connect any.

If the connection of capacitance across the primary tends to minimize the overshoot or ringing then choose a value that produces a minimum overshoot without causing a rounding of the wave.

Usually the matter cannot be completely rectified just by capacitance on the primary of the output transformer, so the next step to try is adjustment of the phase-shift capacitor in the feedback loop. Where any capacitance at all across the transformer primary exaggerates the effect, the phase-shift capacitor alone will have to be relied upon to produce the best compromise. Where capacitance across the transformer primary improves the situation, various values can be tried in both positions to see which combination produces the minimum overshoot or ringing.

In some instances it will not be possible to produce a satisfactory waveform by any combination—always the overshoot or ringing seems to be somewhat excessive. Under these circumstances the only solution is to change the feedback
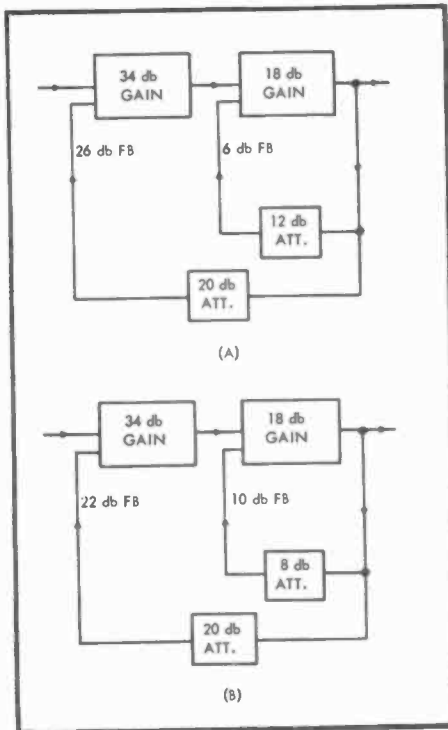

(A)

(B)

Fig. 7. Block schematics illustrating how the distribution of feedback can be modified in a two-loop feedback amplifier: in the original amplifier (A) the inner loop gives 6 db feedback, with an attenuation in the feedback of 12 db across a gain of 18 db; this gives a resultant gain of 12 db; the over-all gain is controlled chiefly by the 20 db attenuation in the over-all feedback; with $34 + 12 = 46$ db total forward gain, this loop gives 26 db feedback. By increasing the inner-loop feedback to 10 db (reducing the attenuation in the feedback path from 12 db to 8 db), the output section with feedback now drops to 8 db gain (B), and the over-all gain is now only $34 + 8 = 42$ db. As a result, the outer-loop feedback is now 22 db instead of 26 db. These figures are approximations based on the formula $A = 1/B$, which is not too accurate here; in practice there would be a change in gain of 3 db feedback on the inner loop and of about 2.7 db on the outer loop, leaving a net change of about 0.3 db (lower). A slight reduction in outer-loop feedback could offset this.

arrangement itself. It may be necessary to reduce the over-all negative feedback. However this is a little undesirable because it will result in slightly greater distortion, because the over-all feedback reduces the distortion present in the output. But if the amplifier employs two feedback loops as many modern amplifiers do, it may be possible to introduce a compromise by increasing the feedback in the inner loop and reducing the feedback on the outer loop.

To achieve this the feedback resistors in the outer loop need not as a rule be changed, because increasing the feedback in the inner loop alters the effective gain in the outer loop and so automatically reduces the feedback in the outer loop by approximately the same amount as the inner loop feedback is increased. *Figure 7* illustrates this. Do not try to vary this more than about 3 to 6 db from the original operating condition. Even this much may sacrifice some of the available power output, due to the fact that the operating levels in the amplifier are disturbed and the same maximum levels cannot be achieved to produce the full output of the amplifier.

It may be well in conclusion to stress the significance of an error that has been made more than once on this job. A transformer can fail to give good results because *it is too good for the amplifier.* If the amplifier was originally built round a low cost transformer, it will seldom work as well with a higher quality job. So don't make the common mistake of concluding the transformer is not up to spec. After some careful adjustments in the manner outlined, you will have a better amplifier. ●

Which Tube Shall I Use?

GEORGE FLETCHER COOPER

The author presents a lucid yet simplified description of the use of tube characteristic curves for the selection of the proper tube, from the standpoint of distortion, for a given application.

I AM ALWAYS READING, and for that matter writing, articles on how to design this and that, especially in the way of audio amplifiers. My staff of trained statisticians estimates that if all the audio amplifiers which have been designed were set to operate at full output simultaneously, one in four of the population would be off to have their eardrums pierced. One thing, however, is scarcely mentioned: although I explain why I take feedback round from end to beginning, and my friend X tells you why he uses only local feedback, we rarely explain why we chose to use a particular tube in a particular place. Oliver Heaviside said somewhere, "Even Cambridge mathematicians deserve justice," so I must rapidly plead that we do usually tell you why we chose our output stage tubes. My own personal choice for the 20–100 watt output stage is the EL34, even if it is a little harder to get than old faithfuls like the 6L6, just because I have found it very easy to drive. Maybe some of the newer tubes are even easier, but at the moment I have no reason to try to find out.

The tube at the front end of the amplifier may choose itself, too, especially in portable equipment. My own personal feeling is that microphony is best treated by using the tube you like in its ruggedized version and providing some mechanical anti-shock system, while hum is rarely troublesome if the heater line is lifted up to about 20 volts above ground and is kept balanced. I am tempted to rectify the heater supply rather than use special tubes, but so far I have never been that desperate. As a result I am pretty free to choose the tubes right through the amplifier. There are some special considerations to be taken into account which will not affect most readers: when you are working on a whole range of equipment, spares inventories must be remembered or you find you need more space for spare tubes than for the operating installation.

Designers of studio-frequency amplifiers usually adopt a figure of merit based on the ratio of transconductance to capacitance. Just how you combine input and output capacitance depends on the interstage coupling circuit. This is discussed in Terman and I do not propose to look it up and copy it down. A more sophisticated designer will go on to worry about grid-plate capacitance, the usual limiting factor in designing narrow-band r.f. amplifiers. And then —but we are, for now anyway, audio designers. What are we to do?

The basic problem of the audio designer is to get gain without distortion. As you all know, you can trade gain for distortion by using negative feedback and you can also trade distortion for gain by using positive feedback. Not surprisingly, the bigger the deal the bigger the headache. Ideally each stage of an amplifier could then be used to maximum effect, but this really does involve quite a design job.

Tube Data

What we need in making our choice of tube type is some fairly simple criterion. In the old days the tube makers gave us a short set of tube data, curves and a couple of circuits all on a single small sheet of paper: now no self-respecting manufacturer would send out data sheets weighing less than the tube itself. What we need to do, therefore, is to find some way of summarizing this information into a form suitable for easy comparison between tubes.

Most of the vital statistics of a tube seem to be included, for our purposes, in a single curve. This is the curve of transconductance against bias. From this curve we can derive a whole mass of other information, and we can also just plot the curves for several different tubes on the same sheet of paper without getting into too much of a muddle. For some reason the textbook writers have never taken to this rather simple approach so that it is not nearly as widely known as it deserves.

First of all, let us assume that we are working with pentodes. Then the characteristics of the tube are, for all practical purposes, independent of plate load. We write down the plate current I_p as a function of grid voltage:

$$I_p = I_o + \alpha e_g + \beta e_g^2 + \gamma e_g^3 + \dots \quad (1)$$

In this expression e_g is measured from the normal working point and represents the input signal. This corresponds to the practical arrangement in which the cathode is biased positive by the drop in the cathode resistor, the bias is held constant by a large decoupling capacitor and the grid is returned through a high resistance to ground. I_o is just the plate current with no signal applied. To find the transconductance we differentiate, giving

$$\frac{\alpha I_p}{\alpha e_g} = g_m = \alpha + 2\beta e_g + 3\gamma e_g^2 + \dots \quad (2)$$

Of course we normally consider the transconductance to be the limiting value for very small signals, so that actually α is the transconductance given in tube data.

In Eq. (2) we have written down:

$$\frac{dI_p}{de_g} = g_m$$

This is just the same as writing down

$$I_p = \int g_m \alpha e_g$$

provided we put in the proper limits. We know that if $e_g = e_{gc}$, the cut-off voltage, we must have $I_p = o$ and if $e_g = o$ the plate current is I_o. Then

$$I_o = \int_{e_{gc}}^{o} g_m \alpha e_g$$

As I hope you remember, this is just the area under the $g_m - e_g$ curve, the area shown shaded in *Fig.* 1. Therefore we can tell the price we must pay in plate current for any particular transconductance, and thus for any particular gain. The area under the curve, if it is of the form shown in *Fig.* 1, is most
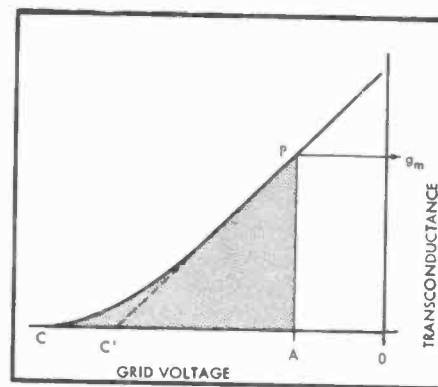

Fig. 1. A typical curve of transconductance against grid voltage. At P, the working point, the grid voltage is taken to be zero, AO being the voltage of the cathode above ground. C is the cutoff point. The area shaded is the standing plate current.

easily found by calculating the area of the triangle C'AP and then either counting squares or making a rough estimate of the area of the small wedge on the left. Remember, when doing this, that the average tube tolerances are quite large and do not, I beg you, try to work to 1 per cent, or even 5 per cent. It just doesn't mean a thing!

What else can we find out from this graph? Well, let us look again at Eq. (1), and assume that $e_g = e \cos \omega t$. It is, by the way, always a good thing to use $\cos \omega t$ rather than $\sin \omega t$ in harmonic calculations, because then there are no minus signs to make the expressions more awkward. Since $e_g = e \cos \omega t$, we have

$$e_g{}^2 = e^2 \cos^2 \omega t = \frac{1}{2} e^2 (1 + \cos 2\omega t)$$

and

$$e_g{}^3 = e^3 \cos^3 \omega t = \frac{1}{4} e^3 (3 \cos \omega t + \cos 3\omega t).$$

Equation (1) therefore becomes:

$$I_p = I_o + \frac{1}{2}\beta e^2 + \left(\alpha + \frac{3}{4} \gamma e^2 \right) e \cos \omega t$$
$$+ \frac{1}{2}\beta e^2 \cos 2\omega t + \frac{1}{4} \gamma e^3 \cos 3\omega t + \dots .$$

$$(1a)$$

This equation is correct as long as we are justified in neglecting anything above the third power of e_g. Already, as you see, the steady component is affected by the γ term and the fundamental is affected by the β term. These interactions are actually intermodulation effects in which the signal mixes with its own harmonics to produce other harmonics. The more terms we take, the more likely the editor is to say he doesn't like mathematics!

If we simplify Eq. (1a) rather more by assuming that γ is zero, which means taking only second harmonic into account, and then turn to Eq. (2), we have

$$g_m = \alpha + 2\beta e_g. \qquad (2a)$$

Now let us differentiate this, giving

$$\frac{dg_m}{de_g} = 2\beta$$

This means that β is a measure of the slope of the transconductance characteristic. The ratio of second harmonic to fundamental in the plate current is

$$\frac{1}{2}\beta e^2 / \alpha e = e\beta/2\alpha$$

Let us look at Fig. 2, which is really only a part of Fig. 1 redrawn. The signal drives the grid alternately to the right and the left of A with a total range of $2e$ (i.e. the peak-to-peak voltage). At P the transconductance is α, but the range over which the transconductance varies is $2g$. The slope of the transconductance curve is thus $2g/2e = g/e$. But this, we have just seen, is 2β. Hence we have $g = 2\beta e$ and the ratio of
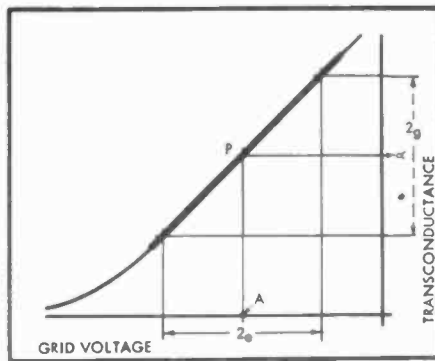

Fig. 2. A signal drives the grid to a peak distance of e volts on either side of A, the reference point. The transconductance varies from $\alpha - g$ to $\alpha + g$.

the second harmonic to the fundamental becomes $g/4\alpha$.

It is, of course, very easy to find this just by looking at the tube characteristic.

We want to have the largest gain, that is the largest value of α, for the smallest value of distortion, which in this case means the smallest value of $g/4\alpha$. If we take gain/distortion as a figure of merit we have $\alpha^2/4g$, or since we are always comparing tubes and need not keep dividing everything by 4, we have α^2/g as a gain/distortion figure of merit.

Practical Example

At this point we should, I suppose, look at some typical tubes. Skimming through a tube handbook I have picked out three tubes and have replotted their characteristics on the same scale, together, in Fig. 3. Choosing a working point at a bias of -1.2 volts and assuming a swing of ± 0.4 volts, we have for these three:

	α (transconductance)	$2g$ (see Fig. 2)	α^2/g
1. 6AU6	3.5	1.4	17.5
2. 6BA6	4.0	1.0	32
3. 6BS7	2.0	0.55	14.5

It would appear, then, that the 6BA6 has the highest figure of merit, and that the low-g_m tube, the 6BS7, although it produces less distortion than the 6AU6, does not pay off because it is better to accept the distortion for the sake of the extra gain.

One thing will be noted, however. The 6BA6 curve is everywhere above the 6AU6 curve and in fact the current taken by the 6BA6 is about twice that taken by the 6AU6. We really should carry out another check of this sort, based on equal plate currents. Then we look at our particular problem and decide whether economy in current is important. Very often it is, even if only because it becomes easier to decouple the early stages from the power-supply disturbance originating in the output stage. In portable equipment every milliampere adds ounces to the weight of the smoothing components.

It is worth while also working out what the distortion actually is: for our three tubes under the conditions given we have:

$$g/4\alpha \times 100 = \text{2nd harmonic distortion in per cent}$$

1. 6AU6 $(0.7/14) \times 100 = 5.0$
2. 6BA6 $(0.5/16) \times 100 = 3.12$
3. 6BS7 $(0.275/8) \times 100 = 3.44$

This shows us that we are favoring the 6BA6 for both distortion and gain. It also gives us some numbers to compare with the distortion in other parts of the amplifier, so that we can decide if this is our critical point or if we should be spending more time worrying about another stage.

Triode Considerations

We have now seen how to assess a pentode stage for its second harmonic distortion. Suppose, however, we are triode users. The answer then is, I'm afraid, it all depends on the tube maker. I have found one who provides me with curves of transconductance vs. grid voltage for three different plateloads, for the 12AT7, anyway. Sometimes you only get curves of μ and r_p. When that happens you have to do rather more work.

The gain of a triode stage is

$$\mu R_L/(r_p + R_L)$$

so that the effective transconductance is $\mu/(r_p + R_L)$. It is tedious but not exhausting to tabulate μ and r_p for different values of e_g, then add R_L to each r_p and then work out this effective transconductance. To bring triodes into our general net we must do this. I think it is worthwhile, just because one unified approach does save quite a lot of thought and effort in the long run.

While we are talking triodes let us deal with the effect of a local feedback loop. This is long-hair language for leaving the cathode resistance without decoupling. (By the way, just what is the difference between the long-hair ap-
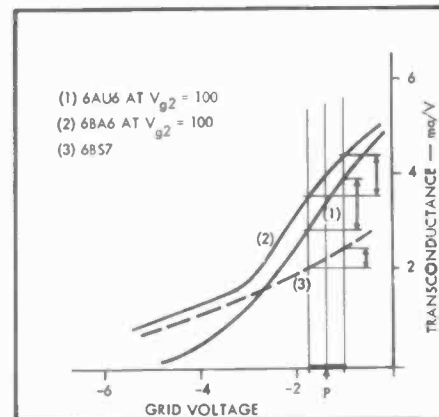

Fig. 3. Comparison of three tube types operating at -1.6 v. bias with a grid swing of ± 0.4 volts.

$100 = 200$ ohms, so that $g_{mf} = 5.0$ ma/v and we can plot this point. By working out a reasonable number of values in this way you can quite easily sketch in the curves shown in *Fig.* 4. You can do this for a pentode on the ordinary transconductance curve, and for a triode on the effective transconductance curve described earlier in this article.

There is a rather interesting thing which appears if you draw out this set of curves very carefully, very large. When you work out the figure of merit for a swing from $e_g = o$ to $e_g = -2.5$ for the tube alone and for the tube with a 100-ohm cathode resistor, there is a small advantage in favor of the tube alone. It is less than 10 per cent better, which is not very much, of course. But the feedback does not quite reduce the distortion in the same proportion as it reduces the gain. Just why this is so is a rather complicated question which I hope to discuss in another article.

Another point about the $R_K = 100$ and $R_K = 200$ curves of *Fig.* 4 is that they are not, like the tube transconductance characteristic, straight lines. Of course the practical tube transconductance itself would have some curvature, but adding the feedback seems to introduce some extra curvature. The result is, in fact, to add a third harmonic term. Physically this is produced by a mechanism of the following kind: we apply a cosine wave to the grid and produce some second harmonic in the cathode current. Because the cathode resistor is not decoupled the cathode voltage contains a second harmonic term. Between grid and cathode, then, we have a signal containing both fundamental and second harmonic. The tube now acts as a mixer, to produce terms of the $(2f + f)$ and $(2f - f)$ kind, of which the first is, ob-
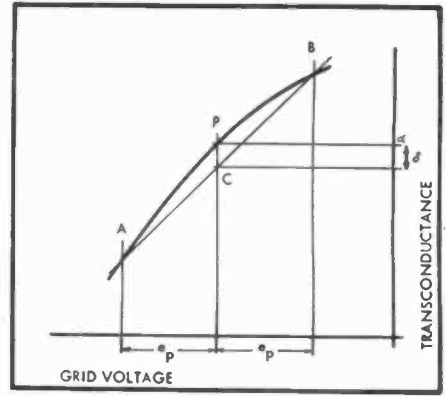

Fig. 5. From the deviation of the transconductance curve from a straight line —the chord AB—we can find the third harmonic distortion.

viously, the third harmonic.

The transconductance curve contains information about this third harmonic. I am not going through the mathematics in detail, because it is rather lengthy and can be regarded as an exercise for the enthusiast. The practical result requires us to refer to the idealized characteristic of *Fig.* 5. Having chosen our working point P and the maximum signal amplitude $\pm e_p$ we draw the chord AB across the transconductance curve. At P the transconductance is α. We measure the distance PC which we call δ. For any input of less than e_p, say e volts, the third harmonic distortion is

$$\frac{3\delta}{\alpha} \cdot \left(\frac{e}{e_p}\right)^*$$

I am not going to apply this to the curves of *Fig.* 4, because it takes pretty accurate drawing. Looking back to our second harmonic expression $g/4\alpha$ we see that if $\delta = g/12$ the second and third harmonics will be equal. You need to look pretty closely to check on the third harmonic.

I think we can now get a rough idea of when local feedback will really be profitable. Most tubes tend to have transconductance characteristics which sag below the straight line I have been drawing so glibly. A little cathode feedback produces a curvature in the opposite direction. By careful choice of the cathode resistor you can get a pretty good linear characteristic and then, going push-pull, balance out the second harmonic. I remember that this worked out very well indeed with the 5763, though I have no figures at hand now to show the improvement.

Some readers may feel that all this is just a paper exercise. Maybe so, if you can afford to buy the wrong tubes. But if you want to get the best performance out of something you design yourself it is worth-while to sit down for a few hours and think before you put your money on the counter. Æ

WHICH TUBE?

(Continued from preceding page)

proach and the egg-head?). We know that the gain of the stage becomes

$$\mu R_L / [R_L + r_p(\mu + 1)R_K]$$

and we have already decided that $\mu / (R_L + r_p)$ was the effective transconductance, which we will call g_m. So the gain becomes:

$$R_L / \left(\frac{1}{g_m} + \frac{\mu + 1}{\mu} R_K\right)$$

Now μ is always big enough for us to take $(\mu + 1)\mu = 1$. After all we shall probably use 20 per cent tolerance components for R_K. We thus have a new transconductance, the effective feedback transconductance g_{mf}, which is given by

$$\frac{1}{g_{mf}} = \frac{1}{g_m} + R_K$$

Let us look at *Fig.* 4. To make the work easier I have drawn a straight line g_m characteristic and I have marked the g_m axis in the values of $1/g_m$. Thus at $e_g = o$ the g_m is 10 ma/v, and $1/g_m = 100$ ohms. If we consider the effect of a cathode resistance of 100 ohms the value of $1/g_{mf}$ is clearly, at this point, $100 +$
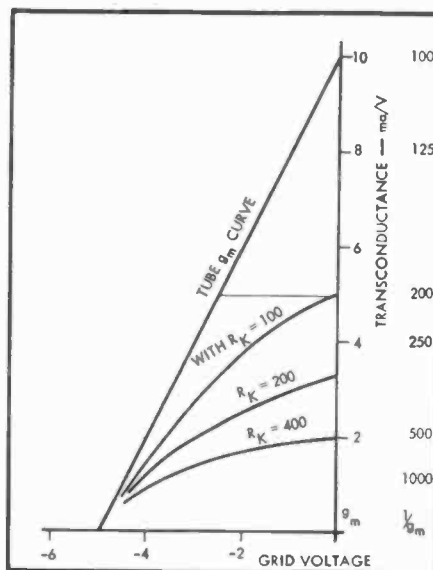

Fig. 4. The construction of these curves is described in the text.

Understanding Intermodulation Distortion

MANNIE HOROWITZ

The author explains the meaning of the term "intermodulation distortion," and describes methods of measuring it. Anyone comparing amplifier specifications is likely to encounter the term, and many want to know what it means.

I N PRE-WAR DAYS, when a music lover referred to high fidelity, he would discuss the frequency response of his amplifier and the associated equipment.

Just after the end of the war, high fidelity achieved a broader meaning. The frequency response was still important. Harmonic distortion was, however, the significant factor in determining how good an amplifier really was.

Amplifier designers were not satisfied with this for long. They found that there was little correlation between frequency response, harmonic distortion, and the listener's approval or disapproval of a particular high fidelity setup.

Some of the experts turned then to phase distortion. This type of distortion exists when it takes longer for an audio signal of one frequency to pass through an amplifier than a signal of another audio frequency. It was soon found that this type of distortion had to be extremely bad to be discernible during the playing of musical passages.

It was soon found that intermodulation distortion (abbreviated IM) was closely related to the degree of unpleasantness of sound reproduction to the human ear. Various methods were devised to measure the IM distortion factor in amplifiers. Acceptable standards of measurement were set up by at least one organization—the Society of Motion Picture and Television Engineers.

Non-Linear Tube Characteristics

Just as in the case of harmonic distor-

Fig. 2. A sine wave applied to a non-linear portion of the tube characteristic.

tion, IM is due to the nonlinear characteristics of the vacuum tube. This nonlinearity is shown by the curves which describe the operation of these tubes. If a curve for the 12AT7 were plotted, assuming a load resistor of 30,000 ohms in the plate, the resultant nonlinearity
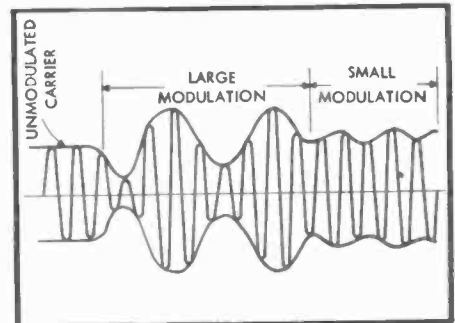
Fig. 3. Wave form of a modulated signal.

would be obvious, as in *Fig.* 1. It can be seen that a change of −2 volts from the operating point, −3 v., to −5 v. in grid potential causes a change of 2.4 ma in plate current, while a change from −3 v. to −1 v. causes a plate-current change of 2.7 ma. If the curve were linear, a grid voltage change of 2 volts either way would indicate a plate current change of 2.7 ma either way. If a 4-volt peak-to-peak sine-wave signal were ap-
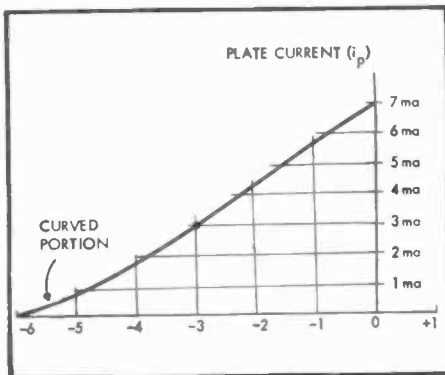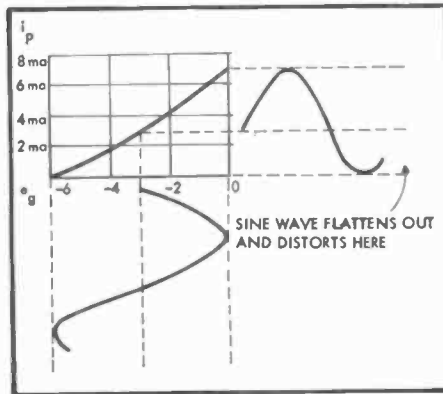

Fig. 1. Typical grid-voltage/plate-current characteristic curve. A plate load resistance of 30,000 ohms and a supply of 300 volts is assumed for the 12AT7.
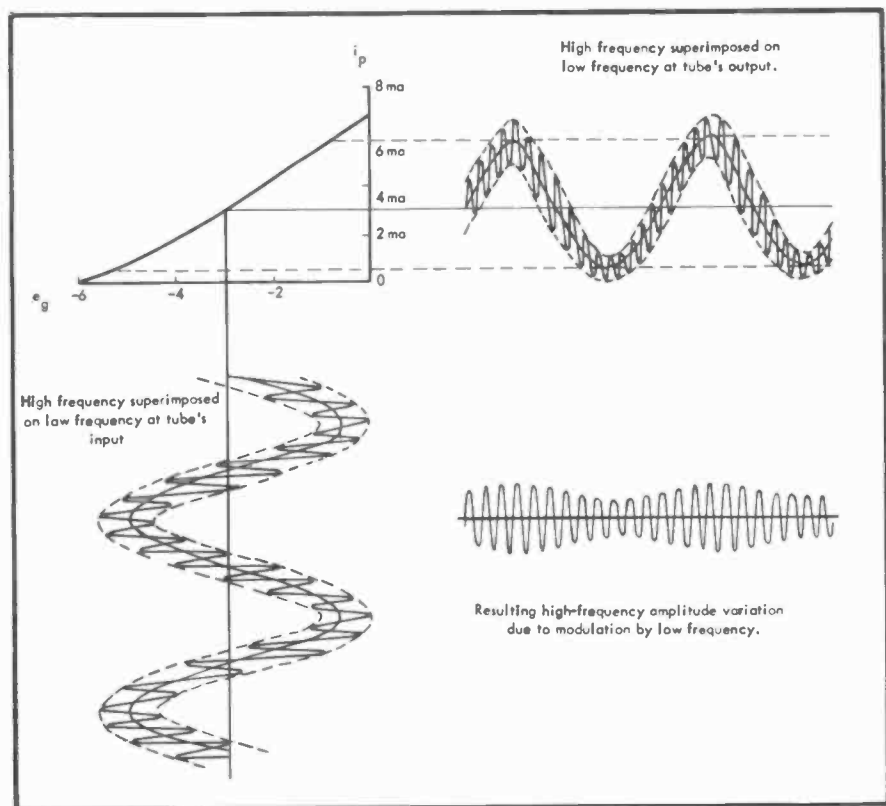

Fig. 4. Result of feeding two signals of different amplitude to the curved portion of the 12AT7 curve.

plied to the grid circuit of this tube, one half would be amplified more than the other half, as shown in *Fig. 2*. This distortion of the original shape of the wave of this signal is due to the nonlinear characteristic of the tube.

Harmonic Distortion

A wave of any shape—square, sawtooth, or even the distorted wave due to the nonlinearity of the tube characteristic—is made up of the sum of many sine waves. The distorted wave above contains not only its original frequency, known as the fundamental, but also numerical multiples of that frequency, which are the harmonics. Thus if a 350-cps wave were somehow distorted, it would consist not only of the fundamental 350-cps wave, but also of components of 700, 1050, 1400 cps, and so on. These added frequencies, referred to as the second, third, and fourth harmonics respectively, usually have smaller amplitudes than the fundamental. Adding all these harmonics together, in the proper amplitude proportion, will give the original distorted waveform.

A pure sine wave consists only of the fundamental, with no harmonics. Therefore it is said to have 0 per cent harmonic distortion. A distorted sine wave contains a certain amount of these harmonics. The percentage of harmonics in any wave determines the harmonic distortion, which is expressed in per cent. Obviously, an amplifier with a minimum of added harmonics due to nonlinearity is preferred to one with a large number of these generated components. Since these harmonics were not present in the original sine wave fed into the unit, it is undesirable for an amplifier to create them for the finished output.

Modulation

To the radio man, modulation is not a new concept. The radio station sends out a modulated signal (*Fig. 3*).

When an amplitude-modulated signal is analyzed mathematically, it can be seen that it consists of a high-frequency carrier, such as 1,000,000 cps with an audio signal, such as 400 cps, changing the strength or amplitude of this carrier. The result is a 1,000,000-cps wave varying 400 times a second in amplitude. When the variation of the 400-cps is great in amplitude, the peaks of the 1,000,000-cps carrier are greater; when the 400-cps modulating signal is low in amplitude, the carrier varies to a smaller degree. This is the method of transmitting audio waves by radio through the use of high-frequency carriers.

It can also be found that there are new frequencies created due to this variation. Not only are the 1,000,000 cps and the 400 cps being transmitted, but there are also sum and difference frequencies present. Thus, due to this modulation, four frequencies are present—1,000,000, 400, 1,000,400, and 999,600 cps. These latter

two frequencies are known as the *sidebands*.

This same principle of modulation with sidebands is once again used in every superheterodyne radio receiver. The 1000 kilocycles (1,000,000 cps = 1000 kilocycles) arriving from the radio station is mixed with 1455 kilocycles (kc) created by the local oscillator in the radio. The result is the creation of the sum frequency, 2455 kc, and the difference frequency, 455 kc. Only the 455-kc sideband is amplified by the i.f. amplifier with the 2455-kc sideband being discarded. This process of mixing of the two signals by the first detector in the radio is accomplished because of the nonlinear action of this first tube. If this tube were perfectly linear as far as its input voltage-output current characteristics were concerned, there would be no mixing and no 455-kc sideband.

Intermodulation Distortion

Extending this theory of modulation to audio equipment, the mechanics of intermodulation distortion become obvious.

In music, there is always more than one frequency present. Assume in the simplest case, that there are only two frequencies available—100-cps and 5000-cps. If the amplifier were perfectly linear, there would be only two frequencies coming out of the unit—100 and 5000 cps—neither one of which would be distorted or mixed in any fashion. However, if the amplifier were not perfectly linear—as is usually the case—the 100 cps and the 5000 cps would mix, modulate each other, and there would be the addition of the sum and difference frequencies, namely 5100 cps and 4900 cps. The amount of these sum and difference frequencies present would constitute the percentage of intermodulation distortion.

However, this distortion goes one step further. Since the amplifier is non-linear, there is also harmonic distortion present. Thus not only are there 100 cps and its harmonics such as 200, 300, and so on; not only are there 5000 cps and its harmonics such as 10,000, 15,000, and so on; but there are also the sum and difference frequencies of these harmonics present to add more to the intermodulation distortion.

This process finally ends with the side bands, their harmonics and side bands, and the harmonics and sidebands of every conceivable combination outlined. The intermodulation distortion is a check on the percentage of all these undesirable frequencies present in the output of an amplifier due to the existence of the two original signal frequencies.

This is not as unwieldy as it might originally seem. The higher harmonics usually have small amplitudes and may be considered negligible. This by itself

would narrow down the intermodulation distortion components considerably. These components can be further narrowed down when it is observed that higher order sidebands and harmonics are outside of the audio range of 20 to 20,000 cps. For this discussion, it will be satisfactory to use the two fundamentals and their sidebands as the sole factors contributing to intermodulation distortion. In the meantime, it should not be forgotten that this all encompassing feature of IM is that it gives excellent correlation with listening tests.

The exact process for the creation of a variation in the amplitude of the high-frequency wave due to the modulation by a lower frequency can be seen in *Fig. 4*. Here, the high frequency is superimposed on the lower frequency in the grid input circuit. Due to the nonlinearity of the grid input voltage to plate output current characteristic, there is a variation in the high-frequency amplitude in the plate circuit of the tube. The resultant amplitude of the high frequency component is shown below the output wave as a modulated signal. The analysis of this signal is indicated under MODULATION above.

Testing for Percentage IM Distortion

There are two general standard methods used for measuring intermodulation distortion. The first, known as the SMPTE method (Society of Motion Picture and Television Engineers) mixes two sine waves in an amplifier. A low frequency, ranging from 50 to 100 cps is mixed with a high frequency of about 7000 cps, and fed together into an amplifier. The ratio of the amplitudes of the lower frequency to the higher frequency is fixed at 4:1. The resulting degree of modulation determines the percentage of intermodulation distortion present in the amplifier under test.

A second method in use is the CCIF test (International Telephonic Consultative Committee). Here, two signals of equal amplitude are fed into the amplifier under test. The two signals have a small difference frequency such as 100 cps. Thus these signals can be 7000 and 7100 cps or 8300 and 8400 cps, and so on. The resulting difference frequency of 100 cps due to non-linearity is the measure of the percentage of intermodulation distortion.

Both methods of measurement are useful in their own spheres.

Amplifiers exhibit different amounts of nonlinearity at different parts of the frequency band. Since the CCIF method measures the low-frequency distortion component due to the distortion in the higher frequencies, the nonlinearity at these higher frequencies of the audio spectrum is observed here.

The SMPTE method utilizes a strong

low-frequency signal and a weak high-frequency component. The difference frequency is of a high order of magnitude. Thus, this method will indicate the effect of low-frequency nonlinearity on a high frequency.

Equivalent Sine Wave Power

This derivation assumes the use of the SMPTE method where two signals with amplitude ratios of 4 to 1 are used.

Assume a signal of 1 volt is superimposed on a signal of 4 volts. The peak voltage applied would then be 5 volts. (See *Fig.* 5.) Since power is proportional to the square of the voltage ($P = E^2/R$), the equivalent power output is
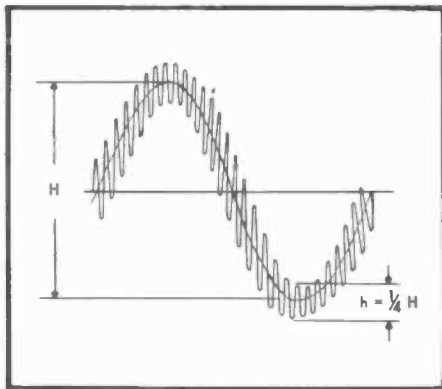
Fig. 5. Waveform of a 1-volt high-frequency signal superimposed in a 4-volt low-frequency signal. This is the usual form of intermodulation test signal.

proportional to $(5/4)^2$ times the power at the 4-volt output.

The real power output is not the peak power. Both frequencies deliver their individual amounts of power to the amplifier's output. The true power output is actually the sum of the powers delivered by each frequency component. In this case, the output power is proportional to $(4 \text{ volts})^2 + (1 \text{ volt})^2$.

However, distortion refers to the peak power output which is proportional to 5^2. The ratio of the peak power to the actual power is $5^2/(4^2 + 1^2) = 25/17$, or 1.47. Thus to find the peak power, the actual power indicated on the meter when making the intermodulation test is multiplied by 1.47. Amplifiers are rated at this peak power, commonly called "Equivalent sine-wave power." This refers to the power in a sine wave signal whose peak voltage equals the peak voltage of the IM signal.

Typical IM Analyzer

Figure 6 shows a theoretical schematic of an IM analyzer and the method by which it operates.

(A) shows two signals having an amplitude ratio of 4:1 combined and fed into the audio amplifier. Coming out of
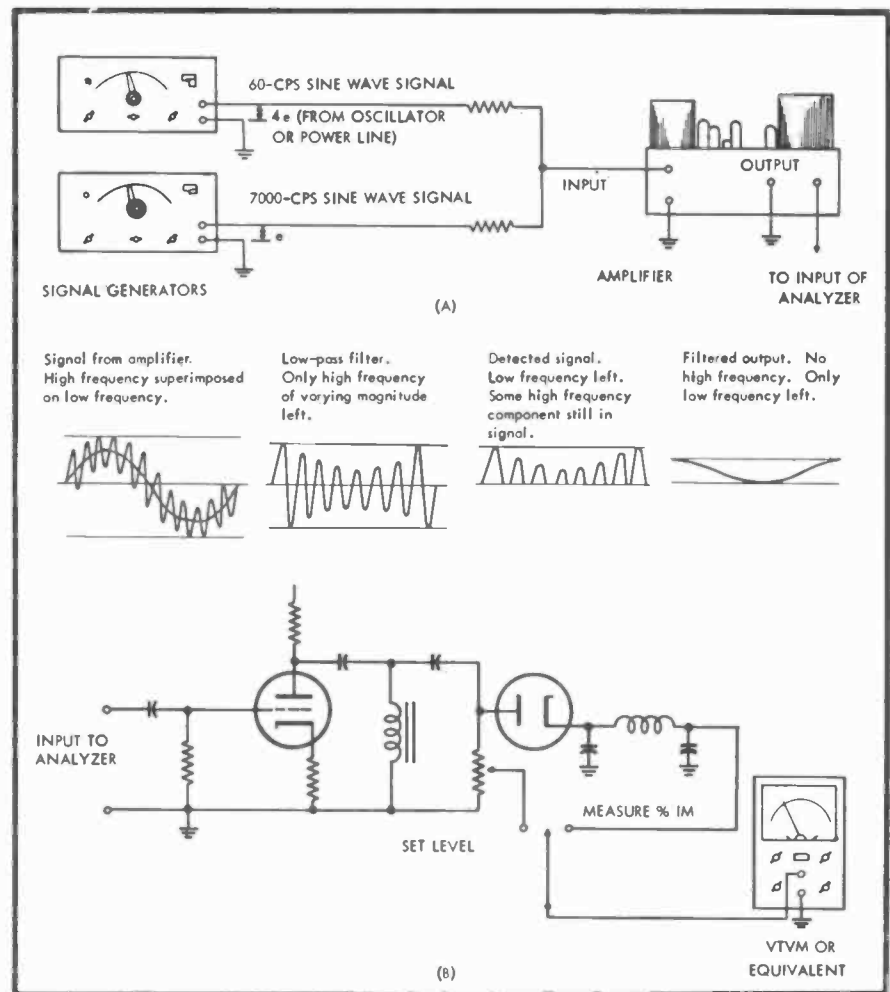
Fig. 6. (A) Arrangement of generators and amplifier for making IM tests. (B) Simplified schematic of IM analyzer. The percentage of IM distortion is equal to 100 x (output voltage/set-level voltage).

the audio amplifier are the two signals modulated, with the amplitude of the high-frequency signal varying in accordance with the low frequency. The high-pass filter in the analyzer eliminates the low-frequency component and passes only the high frequency which has a low-frequency amplitude variation caused by the distortion in the amplifier. This modulated signal is used to set the reference voltage level for a vacuum-tube voltmeter. The modulated signal then passes through a detector similar to that found in a radio receiver. The resultant signal is the low-frequency component which originally modulated the high frequency. The low-pass filter bypasses any of the high frequency left after detection, with the result that only the

modulating low-frequency component is left. This component is a measure of the actual amount of intermodulation distortion created by this amplifier. Feeding this signal to the VTVM, and comparing this with the original amplitude of the modulated signal, indicates the percentage of intermodulation distortion.

To specify IM distortion by itself is not enough. The method used for testing is significant. IM distortion measured by the SMPTE method below about 2 per cent cannot readily be detected by the ear. Valves measured by the CCIF method can not be related directly to those obtained by the SMPTE method. To describe fully the distortion present in an amplifier, both harmonic and IM distortion tests should be made.

The Sad Tale of a Half-Watt Resistor

OR

What a ten-cent component can do to your hi-fi ambitions.

WALTHER RICHTER

Recently an amplifier available in kit form had been assembled, and it was decided to check the actual performance against the specifications. This particular power amplifier is highly regarded by Hi-Fi enthusiasts, and had received quite favorable reviews and test reports. The amplifier has a rated output of 50 watts, and the distortion at this output is specified as less than one per cent.

An intermodulation analyzer was connected to the amplifier, and an attempt was made to drive it to full output. However, it showed that even with 40 watts output the intermodulation distortion exceeded the specifications. Various attempts were made, such as improving the balance between the plate currents of the output tubes by providing separate bias for them, all to no avail. As a matter of fact, in the course of a week's experimentation with the amplifier, the performance kept on deteriorating until at the end of this period the distortion was approximately nine per cent. The situation was indeed becoming desperate.

Up to this time the investigation had been confined to the measurement of distortion, since this was considered as the most important characteristic of the amplifier. Now it was decided to measure some of the other characteristics, in the hope that such investigations might throw a light on the deterioration of the performance. The first additional characteristic to be checked was the required input voltage for full output. The specifications stated that an input voltage of 1.5 volts is needed to obtain full power output. Lo and behold, instead of 1.5 volts, a signal of 3 volts was required to obtain full output. Now in an amplifier with a large amount of negative feedback, the gain from input

to output is usually essentially determined by the feedback network, rather than by the actual gain of the amplifier. An examination of the diagram showed that feedback was obtained from the 16-ohm output terminal of the transformer through a series combination of a 1000-ohm and a 47-ohm resistor to the cathode of the input stage. To produce 50 watts in a load of 16 ohms requires 28.3 volts, and with a series combination of 1000 and 47 ohms this would feed back approximately 1.23 volts to the cathode of the input stage; the excess of 1.5 volt over 1.23 volt would then represent the actual input voltage required between cathode and grid of the amplifier.

The large discrepancy in the input requirements immediately threw suspicion on the feedback network. Measurement of the two resistors showed that the 1000-ohm resistor was well within tolerance limits, but that the 47-ohm resistor was around 100 ohms; replacing this resistor with a new 47-ohm resistor not only brought the input requirement down to the specified value, but brought the intermodulation dis-

tortion to approximately 0.3 per cent, therefore well within the specifications.

Not the Whole Story

The story could have ended here. But a little meditation showed that there was perhaps more to be learned from this incident. The change of the resistance value from 47 ohms to approximately double this value increased the feedback to about twice the original value; this was a perfectly logical explanation for the increased input requirements. However, as far as distortion is concerned, an increase in feedback should, if anything, decrease the distortion, rather than increase it (provided of course that the amplifier remains stable with the increased feedback, which it did). There was only one possible explanation, and one that is hard to swallow, namely that one of these two resistors had become nonlinear. Since it was the 47-ohm resistor which had changed its value, naturally suspicion centered on this resistor. The suspicious resistor was set up in a Wheatstone bridge circuit, as shown in *Fig.* 1. Note that the resistance placed
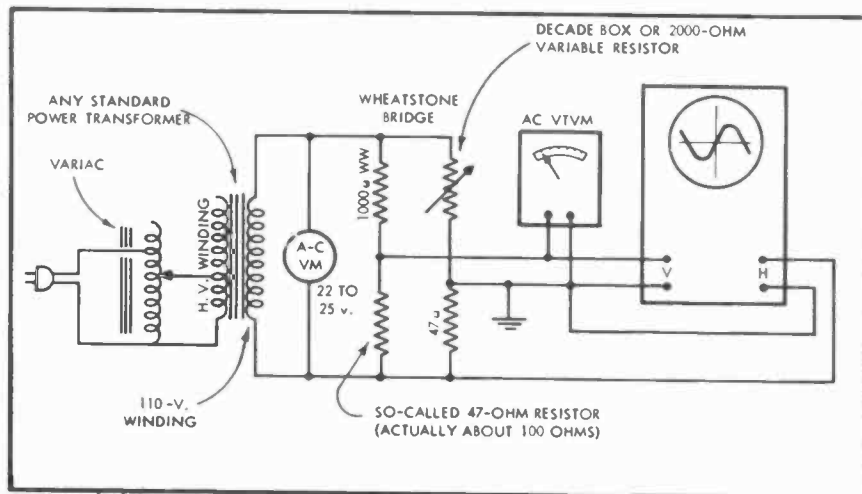

Fig. 1. Test circuit used to demonstrate non-linearity of the offending resistor.

in series with it is exactly of the same value as that found in the amplifier circuit, and that the bridge voltage was made equal to the a.c. voltage which appears across these two resistors in the amplifier under full-load conditions. The unbalance of the bridge was observed on a cathode-ray oscilloscope as well as on a vacuum-tube a.c. voltmeter.

This being an a.c. bridge, exact balance is, of course, possible only when the resistive as well as the reactive components are in balance. However, with a frequency of 60 cps, and the relatively low resistance values in the bridge, stray capacitance and inductance have only a minor effect; if due to these influences a small reactive unbalance remains, it will show itself on the scope as an ellipse, and it can usually be balanced out by placing a variable capacitor across one of the four arms. But in this test, the unbalanced voltage could not be brought below approximately 80 millivolts, and the trace observed on the screen of the cathode-ray tube showed that the remaining unbalanced voltage was a 180-cps voltage, in other words a third harmonic, which could not be balanced out. This means that with a 60-cps sinusoidal current flowing through this so-called resistor, there appeared across it not only a 60-cps voltage of approximately 2 volts, as one would expect, but additionally a 180-cps voltage of approximately 80 millivolts, which is 4 per cent of the 60-cps voltage. A resistor with a built-in harmonic distortion of 4 per cent.

Just to make sure, that this startling result was not by any chance caused by a faulty method of analysis, the simple series combination of 1000 ohms and the suspicious resistor were connected once more to the intermodulation analyzer, interposing an amplifier of known and very low distortion between the analyzer and the series combination. The voltage applied to the two resistors in series was considered as input voltage, the voltage across the defective resistor was considered as output voltage. This simple network now showed an intermodulation distortion of 15 per cent. (Evidently between taking the resistor out of the circuit and completing these measurements, further deterioration had taken place.) In the April, 1948, issue of the *Proceedings of the IRE*, W. J. Warren and W. R. Hewlett published a paper entitled "An Analysis of the Intermodulation Method of Distortion Measurement". In this paper it is shown that if the distortion is due to a non-linearity of the transfer characteristic of a network or an amplifier, a well defined relation exists between harmonic and intermodulation distortion. The paper shows that if the non-linearity is caused by a third harmonic, the ratio between intermodulation distortion and harmonic distortion is approximately 3.8. The measurement with the Wheatstone bridge method indicated a harmonic distortion of 4 per cent, which would result in $4 \times 3.8 = 15.2$ per cent intermodulation distortion, according to this paper. With this degree of agreement between the results obtained by two entirely different methods, there can be no reasonable doubt about the non-linearity of the resistor.

Over-all Effects

For the high fidelity enthusiast the implications presented by this investigation are positively frightening. In his eternal quest for the elusive goal, perfection, he is forever replacing good components with better components. If the advertisements for the new Super-Triple-X amplifier state that it provides 99.44 per cent perfect reproduction, he cheerfully plunks down 250 dollars or so, and discards his previous amplifier, which had 1 per cent distortion and was therefore only 99 per cent perfect. Whether he personally can hear this difference, is probably of less importance than the fact that he can now brag about his new equipment, and feel superior to the poor boobs who haven't the new Super-Triple-X. So what, if a resistor goes bad in the Triple-X and the distortion goes up to 1, 2, or 3 per cent? You still can brag and feel superior about "having the best" (which naturally means the most expensive). This writer can, of course, not speak for others, but he knows definitely that he can not hear the difference between 1 and 2 per cent distortion, as a matter of fact, suspects that it would take 4 or 5 per cent before he would notice it, and a good deal more before it would become disagreeable. He considers himself quite fortunte, not to be cursed with a so-called "Golden Ear," but makes good use of his "Tin Ears" to enjoy thoroughly his library of 800 or so LP records (mostly classical), many of which undoubtedly have more than a modest 3 per cent distortion built right into them. So he can not help but feel a little bit sorry for those who actually *can* hear 1 per cent distortion, because this evidently not only puts severe limitations on the amount of program material available to them, but puts them at the mercy of every component in their system, from 10-cent resistors to 40-dollar output transformers, all of which must be in perfect condition. It is, however, highly probable and fortunate that most of us do not possess the kind of ears that detect with agony a deviation of 2, 3, or even 5 per cent from perfection, otherwise the enjoyment of recorded music would be almost an impossibility.

Compensation for Amplitude-Responsive Phono Pickups

R. H. BROWN

Because of different internal impedance and a different method of generating the output voltage, compensating circuits used with capacitance, crystal, and ceramic pickups differ from those used with magnetics. Before designing networks for crystal and ceramic types, however, make sure that they need compensation—most are designed with correct characteristics to feed into "flat" amplifier.

RECORD WEAR soon becomes a matter for grave concern to many owners of high-fidelity phonograph equipment. When the crisp clarity of a favorite recording gives way to a comparative dullness aggravated by high-frequency distortion, components of the playback system are sometimes suspected of deterioration before record wear is discovered to be the cause of the difficulty. Some hi-fi fans would be unwilling to face up to the cost per play based on the number of times a record can be played before high-frequency loss and distortion due to wear become appreciable. An increase in record life of six times, or better, is possible with use of a capacitance pickup, for capacitance pickups are available which require somewhat less tracking force than some magnetic cartridges and have less than one-sixth as much dynamic mass as the stylus tip.

The compensation required for a capacitance pickup, or any other amplitude-sensitive type, is considerably different from that required for the more popular velocity-sensitive pickups. Consider the problem of obtaining correct compensation for the RIAA recording charac-

teristic shown by the solid line, A, in *Fig.* 1. This characteristic approximates that shown by the dashed line, B, which is constant velocity up to 50 cps, constant amplitude from 50 cps to 500 cps, constant velocity between 500 cps and 2120 cps, and constant amplitude beyond 2120 cps. The dotted curve, C, shows the output that would be obtained from an RIAA test record with an uncompensated amplitude-responsive pickup. Each of these curves has been arbitrarily adjusted to zero level at 1000 cps. The inverse of C is the frequency characteristic of an appropriate amplitude compensation circuit for RIAA recordings. This characteristic has a 12.5 db treble boost to compensate for the constant-velocity recording between 500 cps and 2120 cps, and it also has a 6-db-per-octave
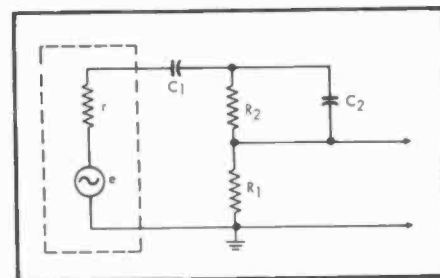

Fig. 2. Amplitude compensation circuit. e is open-circuit output voltage of generator supplying signal to compensation circuit. r is internal series resistance of generator output. C_1 is bass pre-emphasis compensation capacitor. C_2 is crossover control capacitor. R_1 and R_2 provide treble boost.

cut below 50 cps to compensate for the constant-velocity recording below that frequency. One half of the maximum treble boost is required at 1028 cps, which is the frequency at the center of the constant-velocity portion of the recording characteristic. By making a similar analysis for the other important recording characteristics one obtains the data given in the first four columns of Table 1.

Figure 2 shows a schematic of a circuit which will provide correct compensation for an amplitude sensitive pickup. In this circuit, e represents the audio output signal of the pickup, or a subsequent amplifier stage. With a capacitance pickup e comes from the associated oscillator-converter, or a subsequent stage; r represents the internal resistance *in series* with e. For the Weathers oscillator-converter r is about 45,000 ohms. C_1 in connection with the resistance $(r + R_1 + R_2)$ provides 6-db-per-octave compensation for bass pre-emphasis. The maximum treble boost ratio G is equal to $(r + R_1 + R_2) / (r + R_1)$, from which it follows that $R_2 / (r + R_1)$ is equal to $(G - 1)$. Values of this parameter for the important record-

TABLE 1
AMPLITUDE PICKUP COMPENSATOR DESIGN PARAMETERS

Curve	G	B db	fm cps	T=(r+R1+R2)C1 microseconds	R2/(r+R1)	4K	J	R2C2 microseconds
London	6.00	15.6	1225	1590	5.00	.800	1.452	189
AES(old)	6.27	15.9	1000	≥ 10,000	5.27	.758	1.462	233
RIAA	4.24	12.5	1029	3180	3.24	1.233	1.376	213
LP	3.18	10.0	892	1590	2.18	1.835	1.265	213

G = Maximum treble boost ratio. B = Maximum treble boost in decibels, 20 Log G. fm = Frequency at which treble boost is half maximum. T = time constant for bass pre-emphasis compensation. K = (r+R1)/R2. J = √(3+4K)/(1+4K). For r, R1, R2, C1, and C2, see Fig. 2.
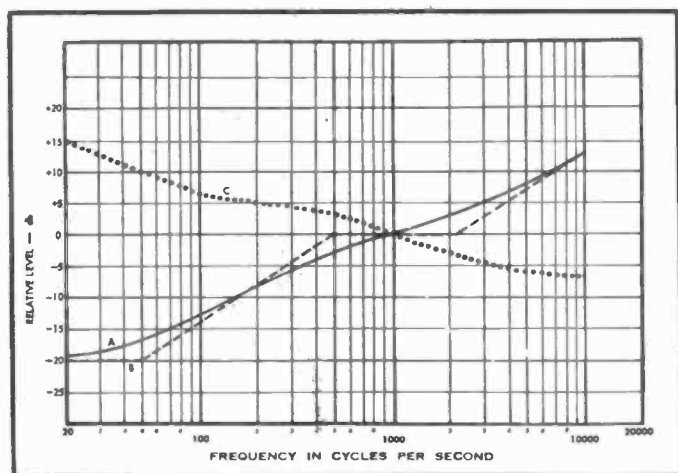

Fig.1. A—RIAA recording characteristic (velocity presentation). B—velocity characteristic approximated by A. C—output from an uncompensated amplitude - responsive pickup on an RIAA test record.

ing characteristics are given in column five of Table I.

C_2 controls the frequency at which the treble boost reaches half its maximum value. To obtain a design equation for C_2 one may express the parallel combination of R_2 and C_2 in terms of its series equivalent and then determine the value of C_2 for which the total impedance in series with e is midway between the low-frequency value $(r + R_1 + R_2)$ and the high-frequency value $(r + R_1)$. Using the abbreviation K for $(r + R_1) / R_2$, one obtains for the condition on C_2 which provides that one-half
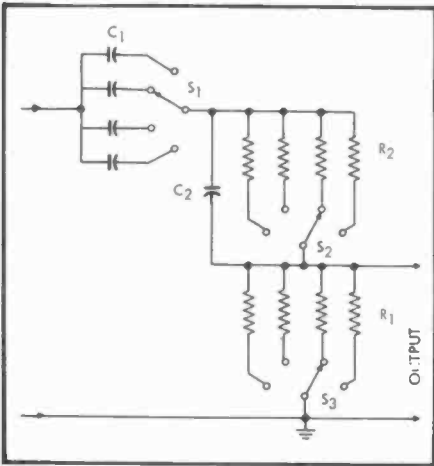

Fig. 3. Basic amplitude compensator. S is a three-gang, four position, shorting-type selector switch. See Table II for values of C_1, C_2, R_1, and R_2.

of the maximum treble boost occur at the frequency f_m

$$R_2 C_2 = \frac{1}{2\pi f_m} \sqrt{\frac{3 + 4K}{1 + 4K}}$$

By abbreviating the radical with the symbol J one obtains the following simple expression for the time constant of R_2 and C_2 in terms of f_m:

$$R_2 C_2 = \frac{J}{2\pi f_m}$$

Values of J and $R_2 C_2$ are tabulated in columns seven and eight of Table 1.

Table II gives representative values for compensation circuit components as determined from the design parameters of columns four, five, and eight of Table I. A fixed value for C_2 gives the compensation circuit of lowest cost and also the greatest ease of construction. A design should be chosen for which R_1 is always at least three times r to avoid loss of signal within the source of e.

Complete Circuit

A suggested compensator circuit is shown in *Fig 3*. In this circuit S_1 controls the bass pre-emphasis compensation. S_2 and S_3 together control the

treble compensation. If any resistance is placed across the output of the compensator, the values of the resistors connected to S_3 must be chosen so that their parallel combinations with this load resistance are equal to the design values for R_1. To avoid a high-frequency loss of more than one db at 15,000 cps, the total of the capacitance to ground (1) at the output of the source of the signal fed to the compensator, (2) in the compensator itself, (3) at the input of the load connected to the compensator, and (4) in any connecting cables, must not be greater than $5(r + R_1)/rR_1 \mu f$ (time constant of five microseconds). Where short leads are possible the compensator could in some cases be connected directly between an amplifier input and the oscillator-converter of a capacitance pickup. The insertion loss of the compensator is given by B in column two of Table I.

A completely isolated amplitude compensator with level control and cathode follower output is described in *Fig. 4*. The output from this circuit is ample to drive power amplifiers requiring a high-level input. The decoupling filter in the circuit allows B + to be taken from an associated power amplifier without motorboating difficulties. Provision for tone controls could be made by adding another stage of amplification, or by

using a high-mu twin triode with the tone control circuit between the two amplifier sections and the compensator circuit between the input connection and the level control. For the latter arrangement the values of the resistances connected to S_3 would need to be redetermined. It is evident that the capacitance pickup offers advantages of simplified compensation and preamplification circuitry in addition to the advantage of greatly reduced record wear.

Ceramic Pickup Compensation

The ceramic (and crystal) pickups are also amplitude-responsive, but in most instances they are designed to work into specific load impedances which result in a close adherence to the current standard recording characteristic, the RIAA. However, it may be that the ser wishes a different curve from his equipment. The same circuits that work with the capacitance pickups will also work with ceramics, but it must be remembered that the latter already have built into them the rolloff characteristics of the RIAA curve. Thus while the low-frequency response may be corrected as shown, the user should remember that there is already some rolloff, and that if he wishes less rolloff than this he will need to provide high-frequency boost; conversely, for more rolloff, he will need to provide additional losses.

The same circuitry will obtain with both types of pickups, but it is only important that the "built-in" rolloff be kept in mind. Considerable experimentation may be required to end up with the desired result, but the principles are the same and the final response can be tailored to suit without much trouble.

TABLE II
AMPLITUDE COMPENSATOR CIRCUIT COMPONENT VALUES

Curve	C_2 μμf	R_2 Kilohms		$R_1 + r$ Kilohms		C_1 Microfarads	
		Design Value	Nearest 5% RMA Component	Design Value	Nearest 5% RMA Component	Design Value	Nearest 5% RMA Component
London	750	252	240	50.5	51	.00526	.0051
AES (old)		310	300	58.8	56	.027	.03
RIAA		284	270	87.4	91	.00858	.0082
LP		310	300	143	150	.00351	.0036
London	500	378	390	75.6	75	.00351	.0036
AES (old)		465	470	88.3	91	.018	.02
RIAA		425	430	131	130	.00572	.0056
LP		465	470	214	220	.00234	.0022

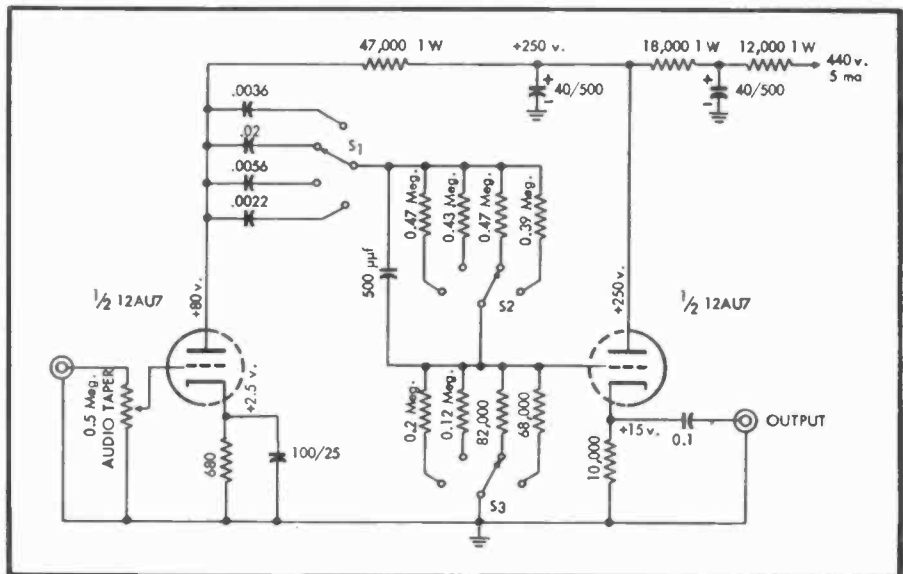Fig. 4. Compensator preamplifier for amplitude-responsive pickups. S should be a three-gang, four-position shorting switch. Compensator switch positions are: (1) LP; (2) RIAA; (3) AES; (4) London. All components in the compensator circuit (connected to terminals of S) should be ± 5 per cent.

A Versatile Bass-Treble Tone Control

CHARLES T. MORROW

This bass-treble tone-control circuit, containing only resistors, capacitors, and switches, provides control not only over the amount of equalization but also over the frequency at which the response starts up or down.

BASS-TREBLE EQUALIZERS or tone controls are appropriately used in a high-fidelity amplifier to compensate for loudspeaker deficiencies, or, if a more specialized equalizer is not available, to compensate for recording characteristics. They are occasionally useful when a radio program is improperly equalized at the studio. They may be used as a substitute for loudness controls, although the author has never been tempted to try to compensate for ear characteristics.

In general, it is more important to control the frequency at which boost or attenuation begins than the maximum equalization at the high or low end of the spectrum. Most tone controls incorporated into commercial amplifiers control primarily the maximum equalization.

The circuit of *Fig.* 1 has been in use in the author's home amplifier for the past two years and provides both types of control. Four two-gang rotary switches are used—preferably of the shorting type. The circuit also uses fifteen resistors and eight capacitors exclusive of the input coupling capacitor. For simplicity, the diagram omits the 10 megohm resistors that are shunted across the eight capacitors to minimize switching transients. Two switches provide choices of nominal 3-db frequencies: OUT, 300, 500, or 700 cps for bass, and 3000, 5000, 7000, or OUT for treble. The other two switches provide corresponding choices of 6-db-per-octave droop, droop to a depressed 6-db plateau, OUT, boost to a 6-db elevated plateau, or 6-db-per-octave boost. The insertion loss is approximately 26 db, which is made up by a preamplifier. Other 3-db frequencies could be chosen or added by connecting other resistors to the switch contacts, and other maximum equalizations could be obtained by connecting other capacitors.

The various equalization curves for the circuit of *Fig.* 1 are shown in *Fig.* 2,

and it can be seen that enough choice is possible for most ordinary situations, although there might be some virtue in extra switch positions for elevated and depressed plateaus at 12 db, or a change to a plateau between 6 and 12 db. The reader may prefer to select his own 3-db frequencies. For this reason, after the operation of the specific equalizer circuit is discussed, formulas will be given, and non-dimensional curves for bass and treble separately. From these the reader should be able to design his own equalizer. A brief derivation involving complex algebra is given in Eq. (6), which the reader may ignore if he chooses.

Bass Equalization

Examination of the left half of the circuit of *Fig.* 1 shows that it consists of a resistive voltage divider with a loss of approximately 26 db, and a set of capacitors for bass boost. The 3-db points are varied by shunting the resistive voltage divider with resistors. In general, the

diagram shows nominal values for commercial components.

For a selection of a 3-db frequency of 300 cps and a 6-db-per-octave bass droop, the bass equalizer reduces to the circuit of (A) in *Fig.* 3, where the total resistance $R_1 = 0.1$ megohms approximately, $k = 1/20$, and $f_1 = 300$ cps.

If a 6 db depressed plateau is chosen, the bass equalizer reduces instead to the circuit of (B) in *Fig.* 3b, which includes a capacitive voltage divider which starts to take over in the region of the nominal 3-db frequency and has 6 db more loss than the resistive divider.

If "out" or flat response is chosen, the circuit of (C) in *Fig.* 3 is applicable (or (D) if switch connections are made according to the dotted lines of *Fig.* 1). The capacitive divider in this case has the same loss as the resistive.

If an elevated 6 db plateau is chosen, the circuit of (E) in *Fig.* 3 is applicable. A capacitive divider with 6 db less loss than the resistive begins to take over in the region of the nominal 3-db frequency.
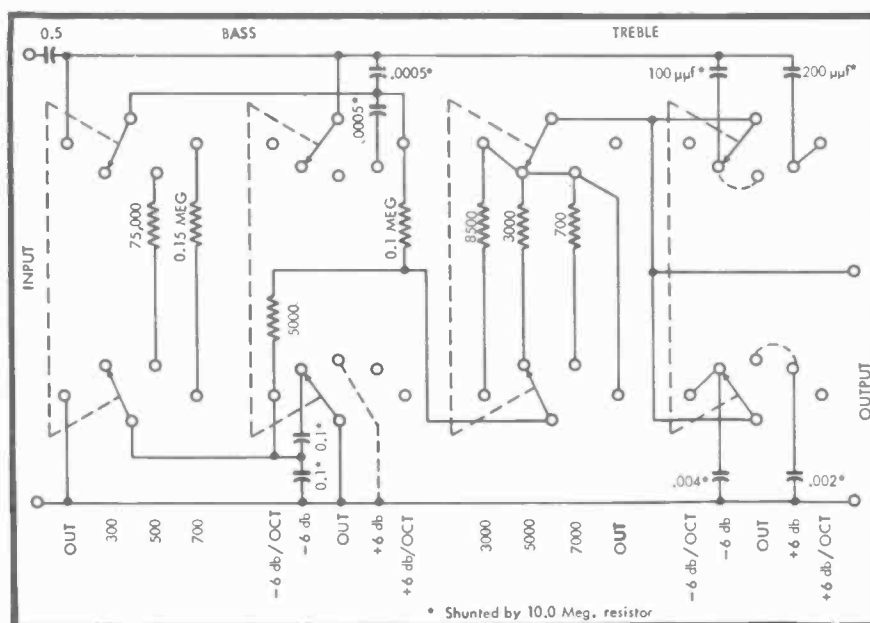

Fig. 1. Schematic of amplifier employing the tone correction circuits described in the accompanying text.
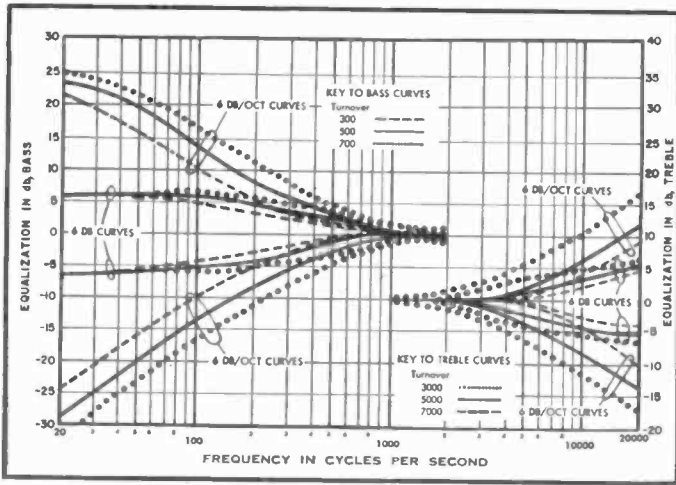
Fig. 2. Basic characteristics of the various equalization curves provided by the circuit of Fig. 1.

Finally, if a 6 db/octave boost is chosen, the circuit of (F) is applicable, in which the capacitive reactance is equal to kR_1 at the 3-db frequency and becomes the controlling load impedance at lower frequencies, decreasing the insertion loss.

Use of a resistive shunt on the resistive divider raises the 3 db frequency to 500 or 700 cps by effectively decreasing R_1 while C_1 remains constant. The effect at low frequencies is the same as if the individual resistors of the divider were divided by the same factor.

Treble Equalization

Operation of the treble equalizer is similar except that the capacitors are in shunt with the voltage divider resistors rather than in series, and the resistors used to control the 3-db frequency are in series with the output lead rather than in shunt with the voltage divider.

For example, if a 6-db-per-octave droop is chosen, the treble equalizer reduces to the circuit of (A) in Fig. 4, where R_1 is the total resistance of the unshunted resistive divider. This is merely a circuit with a mid-frequency transmission k and a high-frequency droop corresponding to that of a resistance R_2 working into a capacitive load C_2.

If a 6-db depressed high-frequency plateau is chosen, the circuit of (B) in Fig. 4 is applicable. In the region of the nominal 3-db frequency, a capacitive voltage divider with approximately 3 db more loss than that of the resistive divider begins to take over.

For "out" or flat response, the circuit of (C) is applicable [or (D) if switch connections are made according to the dotted lines of Fig. 1]. In the latter case, the capacitive divider has the same insertion loss as that of the resistive.

For a 6-db elevated plateau, the circuit of (E) in Fig. 4 is applicable, in which the capacitive voltage divider has approximately 6 db less insertion loss than that of the resistive divider.

For a 6-db-per-octave boost, the circuit of (F) is applicable. The voltage due to current through the capacitor C_2/k is closely equal to that due to current through the upper resistor at the 3-db frequency f_2.

Discussion

Approximate formulas for the gain of the simplified circuits as a function of frequency are given in Eqs. (1) through (8) in case the reader wishes to design his own equalizer. The gain (1A) is plotted as a function of f/f_1 and f/f_2 in Figs. 5 and 6, for the case of $k = 1/20$ and plateaus at 6 db.

It will be evident that the capacitors in Figs. 3 and 4 were chosen so that the 3-db points for Figs. 3A and 4A would be exactly f_1 and f_2, and so that the attenuations of the capacitive voltage dividers would be exactly $k/2$, k, or $2k$ as called for. For small values of k, the expressions for calculating the capacitances become quite simple.

Formulas for Bass Equalization

2A. $$|A| = \frac{kR_1}{\sqrt{R_1^2 + R_1^2 f_1^2/f^2}}$$
$$= k \frac{1}{\sqrt{1 + f_1^2/f^2}} \quad (1)$$

Simplifying the remaining equations,

2B. $$|A| \approx k \frac{\sqrt{1 + f_1^2/4f^2}}{\sqrt{1 + f_1^2/f^2}} \quad (2)$$

2C. $$|A| \approx k \frac{\sqrt{1 + f_1^2/f^2}}{\sqrt{1 + f_1^2/4f^2}} \quad (3)$$

2F. $$|1| \approx k \frac{\sqrt{1 + f_1^2/f^2}}{\sqrt{1 + k^2 f_1^2/f^2}} \quad (4)$$

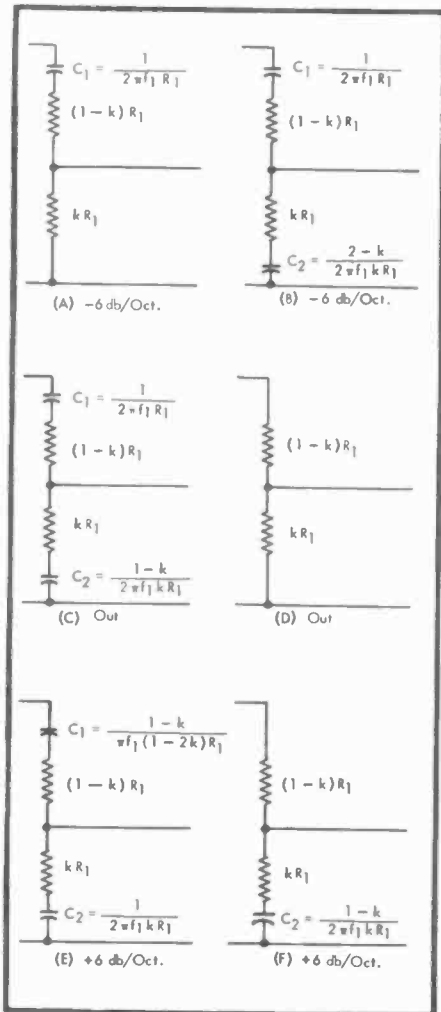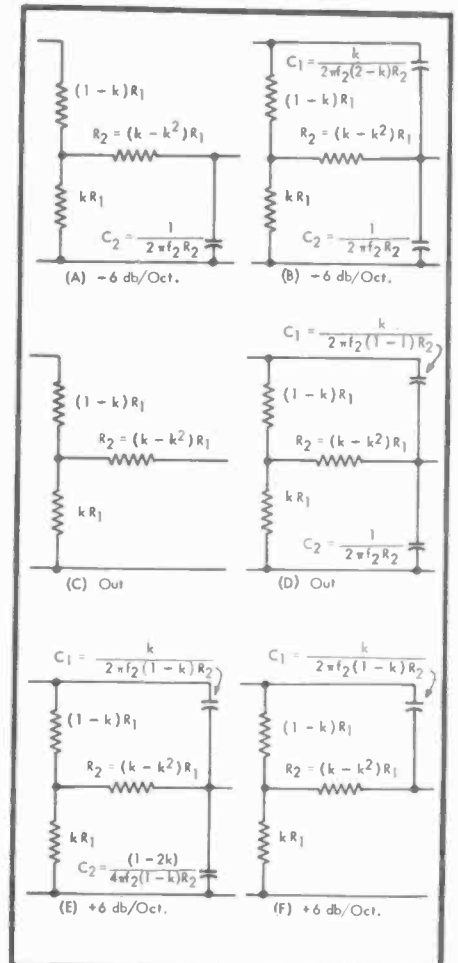Fig. 3. Circuit configurations and formulas for bass equalizer section.

Fig. 4. Circuit configurations and formulas for treble equalizer section.

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Record Speed and Playing Time

A tongue-in-cheek analysis of the possibilities of high-quality reproduction at 16 2/3 rpm in comparison with the present 33 1/3 and 45 rpm discs—but an analysis, nevertheless, which has a strong background in good engineering.

CHARLES P. BOEGLI

SEVERAL YEARS AGO, lovers of recorded music went their way in a state of blissful misery (if it can be defined by a combination of such seemingly contradictory terms) with heavy and fragile shellac records of at best poor fidelity, and an occasional better-than-average vinyl disc to provide a sort of oasis. The 33 1/3-rpm LP record, the 45-rpm disc, new and better pickups and amplifiers, and so on, appeared almost overnight and at last the music lover could listen to sound as well as (and in some cases instead of) music.

But there is a law of compensation, and the price had to be paid for all this new enjoyment. Sometimes one wonders whether it was worth it, for beside the cost of new equipment there was a surtax, as it were, on the audio fan's peace of mind. Nowadays one has no sooner bought a speaker than a newer, better, cheaper one appears; and to purchase the latest pickup seems to insure the announcement of a superior one next month. Even turntables are vulnerable; ominous whisperings about 16 2/3-rpm records continue to be heard and will not be stilled.

Happily, mathematical laws sometimes aid the peace of mind of the audio fan. For example, consider 16 2/3-rpm records. Whether they will ever replace the "old" 33 1/3-rpm discs depends upon what sort of benefits can be realized from them, and what sort of expense must be endured to enjoy these rewards. Certainly, we cannot assume that by a simple halving of speed the amount of music on a record side can be doubled, for, as we shall see later, the picture is complicated by other factors. For example, there is the problem of pickups.

Even at the present state of the art, pickups for 33 1/3-rpm discs are not yet perfect, and cause much more record and stylus wear than they should. A great deal of work remains to be done; the one-mil tip should track with a force of around one gram and have a dynamic mass less than one milligram[1]. Admittedly these figures do not

appear unreasonable, as they would have ten years ago, but they remain to be accomplished. The problem is multiplied by smaller styli and it doesn't appear that we will have thoroughly satisfactory pickups with 1/2-mil styli for some time to come. In view of the probable reluctance of music lovers to endure once again the rigamarole of replacing all their equipment, it seems reasonable that if 16 2/3-rpm discs are to make their appearance very soon, they will have to be cut for a one-mil stylus.

The question then comes down to whether much longer playing time can be realized with present equipment simply by reducing the speed at which the disc turns. Fortunately, a good answer to that problem is easily obtained. Suppose we let n equal the number of grooves per inch cut on the record. With newer methods of recording this quantity will of course vary across the disc but that will have no effect, as we shall see later. There is a tacit assumption, however, that the "average" n will be the same irrespective of the speed at which the record turns, and this is not a bad assumption if the same stylus is used for all the speeds. The total number of grooves on the disc is then $(R - r)n$, R being the outer radius and r the inner radius at which grooves are cut. This quantity, of course, is also the number of revolutions the disc must turn to be played, and if s is the rotational speed in rpm, the total playing time

$$t = \frac{(R-r)n}{s} \text{ minutes} \qquad (1)$$

Now, the inner radius is related to s; obviously, the lower the speed, the greater will r have to be to avoid excessive tracing distortion at the center of the disc, even with a perfect pickup. We may get an idea of the relationship from Roys,[2] where it is shown that for a one-mil stylus the intermodulation

[1] F. V. Hunt, "On stylus wear and surface noise in phonograph playback systems." J.A.E.S. January, 1955, p. 2.

[2] H. E. Roys, "recording and fine-groove technique." AUDIO ENGINEERING September, 1950, p. 11.

distortion reaches 10 per cent on a 45-rpm disc at an inner diameter of about 4.9 inches, and on a 33 1/3-rpm record at a corresponding measurement of around 6.8 inches. It seems logical that this distortion is attained approximately at a fixed linear speed, so that if we assume

$$r = \frac{4.9 \times 45}{2s} = \frac{110}{s}$$

we shall have the desired relationship. Putting $s = 45$ in this equation, we get $r = 2.45$ inches ($d = 4.90$ inches) and for $s = 33 \ 1/3$, $r = 3.31$ inches ($d = 6.62$ inches), so the equation agrees pretty well with experiment.

On the other hand, Columbia has specified a minimum diameter of 4¾ inches for their discs ($r = 2.375$ inches) and we must assume they knew what they were doing, so we should modify the above equation to permit this smaller diameter. It turns out then that

$$r = \frac{79}{s} \qquad (2)$$

for commercial discs meeting the specifications of Columbia, and this is the equation we shall substitute into Eq. (1), yielding the result

$$t = \frac{(R - 79/s)n}{s} = \frac{nR}{s} - \frac{79n}{s^2} \qquad (3)$$

As a matter of interest, let us find the s at which t will be a maximum. We have

$$\frac{dt}{ds} = -\frac{nR}{s^2} + \frac{158n}{s^3}$$

which we set equal to zero to find the s for a maximum t. This gives

$$s = 158/R = 316/D \qquad (4)$$

for maximum playing time, where D is the maximum diameter of the record. Note that the quantity n has dropped out and does not appear in Eq. (4).

Now, for a 12-inch disc, $D = 11.5$ inches (allowing for the lead-in grooves) and $s = 316/11.5 = 26.6$ rpm.

For a 10-inch disc, $D = 9.5$ and $s = 32.2$ rpm.

For a 7-inch disc, $D = 6.5$ and $s = 47$ rpm.

These speeds agree surprisingly well with those in commercial use. For 10-

and 12-inch discs 33 1/3 rpm is a good speed, particularly since it was standardized long before LP discs made their appearance. For 7-inch discs, 45 rpm is not far from the ideal value.

With an "average" n of 225 lines per inch, a 12-inch record which will play 22.8 minutes at 33 1/3 rpm will be good for only 13.5 minutes at 16 2/3 rpm, before distortion becomes too great. And the situation for the 10-inch disc is even worse, for here the comparison is between 16 minutes at 33 1/3 rpm and no time at all for 16 2/3 rpm! What this means, of course, is that if it is permissible to record to a minimum inside radius of 2.375 inches for a 33 1/3-rpm disc, the corresponding dimension for a 16 2/3-rpm record is 4.75 inches, equal to the outer radius of a 10-inch recording. The distortion is too great as soon as the playing starts.

So it looks as though before the 16 2/3 rpm disc becomes much of a threat to higher speeds, smaller styli will have to be developed and this in turn will require a great deal of work on pickup design in general.

It doesn't seem likely to happen for a number of years, anyway.

TONE CONTROL
(*Continued from page 69*)

Formulas for Treble Equalization

3A. $$|A| = k \frac{1}{\sqrt{1 + f^2/f_2^2}} \tag{5}$$

3B. $$|A| \approx k \frac{\sqrt{1 + f^2/f_2^2}}{\sqrt{1 + 4f^2/f_2^2}} \tag{6}$$

3C. $$|A| \approx k \frac{\sqrt{1 + f^2/f_2^2}}{\sqrt{1 + f^2/4f_2^2}} \tag{7}$$

3D. $$|A| \approx k \frac{\sqrt{1 + f^2/f_2^2}}{\sqrt{1 + k^2 f^2/f_2^2}} \tag{8}$$

If something other than a 6-db plateau is desired, the appropriate formulas may be obtained from (2), (3), (6) and (7) by changing the factor 4 under the square root. The ratio of capacitances should be changed in proportion to the square root of this number.

The bass and treble equalizations are very nearly independent. Thus, if f_1 and f_2 are close together, the corresponding deviations from the attenuation k—as obtained, for example, from *Figs.* 5 and 6—are directly additive on a decibel basis.

Recording Characteristic Simulator

This simple device will make it possible to check response curves of phonograph preamplifiers in a minimum of time and with reproducible accuracy—completely eliminating the need for adjusting signal levels as you change frequency.

S. K. GHANDI

A SIMPLE METHOD of testing the frequency response of an audio system is to feed into it a series of audio-frequency signals, whose amplitudes vary according to a particular recording characteristic. The over-all system response is correct when the output signals are of constant amplitude. Feeding in these signals of different amplitudes is normally considered to be a nuisance, and the audio enthusiast usually ends up by purchasing a test record having the various frequency bands already recorded at the appropriate amplitudes. This entails some disadvantages:

a) A good test record costs money. (With the price comparable to that of a regular long-playing record, music is always a much better buy.)

b) The test record is usually cut for one particular characteristic, while the audiofan has provisions for at least four.

c) After a few runs of the record (and the record is often borrowed by friends having "permanent" osmium tipped needles), the surface noise becomes so great that meter indications on the output are often swamped by the record noise.

With the above points in mind, the

Fig. 2. Straight-line approximation (dotted) and actual transition curve for a "corner" frequency f_0 of 1000 cps.
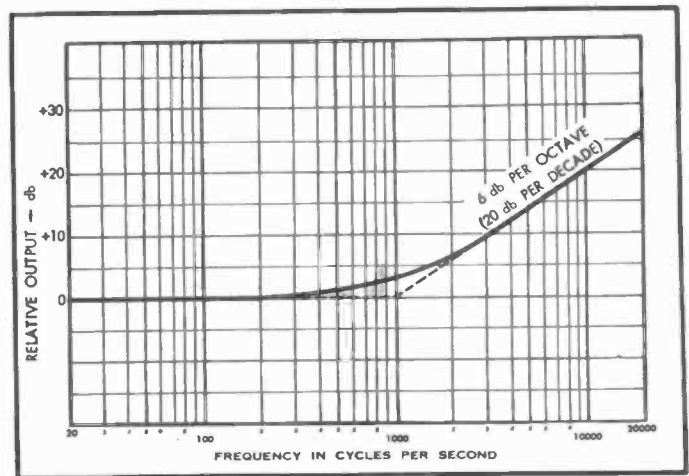
writer set out to construct a recording characteristic simulator, of the type suggested by D. T. N. Williamson.[1] Such a device is essentially a network. One end of this network is fed from a signal generator such as the Heathkit Model AG-8, which covers the audio band without too much variation in signal output (± 1 db), while the other end is fed into the preamplifier input. An a. c. vacuum-tube voltmeter is put across the output terminals of the power amplifier, with

[1] D. T. N. Williamson, "High-quality amplifier modifications." *Wireless World*, May 1952, pp 173–176.

a resistive load instead of the speaker; the system compensation is correct if the meter shows no variation as the signal generator is swept over the audio-frequency band.

Certain restrictions had to be placed immediately.

a) The device had to use no inductances, and a minimum of other components.

b) The device had to be passive, and operate with a reasonably low output impedance so as to avoid hum pick-up in the output leads.
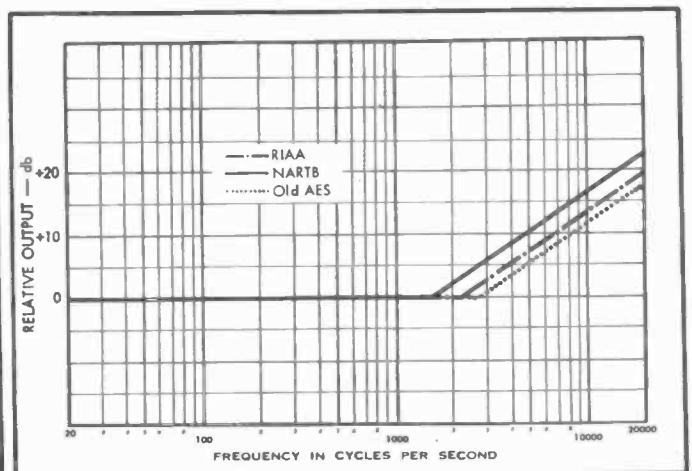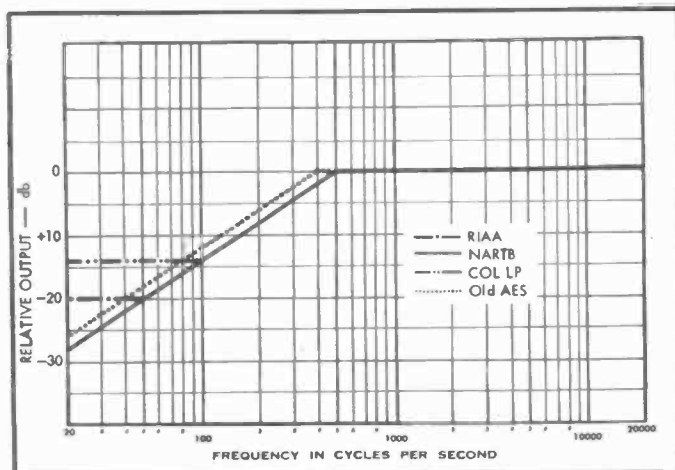
c) The input impedance had to be

Fig. 1. Recording characteristics, before being rounded off as they are in actual practice. Left, low frequency, and right, high frequency.
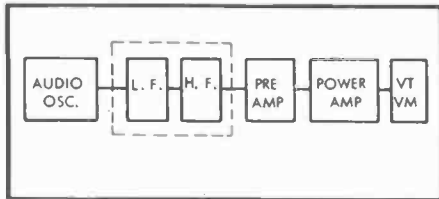
Fig. 3. Block schematic of manner in which the recording characteristic simulator is used in making measurements.

Fig. 5. (A), corners at f_1 and f_2 are obtained with circuit of Fig. 4. When R_1 is eliminated, lower corner is no longer present, as at (B).

high enough so as not to load the signal generator.

d) As many recording characteristics as possible were to be handled by the device. While the unit to be described was restricted to microgroove records, (the writer has only three 78-rpm discs) certain simple changes will allow it to be used for standard records also.

After studying the various lists of manufacturers and their recording characteristics, it was seen that the low-fre-
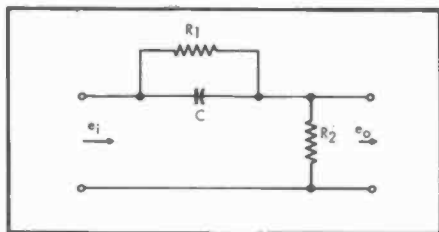

Fig. 4. Configuration of each of the two sections of the simulator.

quency end was covered by three such characteristics:

a) RIAA. (Record Industry Advisory Association), which is flat up to 50 cps, then rises at 6 db per octave to 500 cps, where it again flattens out.

b) COL. (Columbia), which is flat up to 100 cps, then rises at 6 db per octave to 500 cps, where it flattens out.

c) NAB. (National Association of Broadcasters), which has a steady rise at 6 db/octave to 500 cps, where it flattens out.

d) Old AES. (Audio Engineering Society), which has a steady rise at 6 db per octave to 400 cps, where it flattens out.

The high-frequency end was covered by three characteristics:

a) RIAA, flat up to 2120 cps and then rises at 6 db per octave.

b) NAB, flat up to 1590 cps, and then rises at 6 db per octave.

c) AES, flat up to 2500 cps and then rises at 6 db per octave.

While the characteristics are most conveniently described in the above terms, it is to be understood that Mother Nature simply won't stand for sharp kinks in the response curves. Actually, a smooth transition occurs at all the "corner" frequencies, and the above description may be considered to be a straight line approximation of what ac-

tually happens. (A) and (B) of Fig. 1 show the straight line approximations of the recording characteristics at low and high frequencies. The 0 db reference axis is chosen quite arbitrarily in all cases, since this is entirely dependent on the position of the gain control. Figure 2 shows a straight line approximation at a typical corner frequency, together with the actual transition. It should be pointed out that the behavior of all minimum-phase-shift networks using only resistances and capacitances may be described by a series of straight lines which are either flat, or rising or falling at slopes that are multiples of 6 db per octave.

Network Design

The design of the network for the recording characteristics is taken up in two sections, and the sections cascaded. Care is taken to see that the second section does not load the first, so that the over-all response is the combined response of the two networks taken separately. This is done by putting the low-frequency network before the high-frequency network as in Fig. 3. At low frequencies the second network has a very high impedance, and by the time its impedance falls (at high frequencies), the output impedance of the first network is considerably lessened. In this manner, loading is avoided at all frequencies.

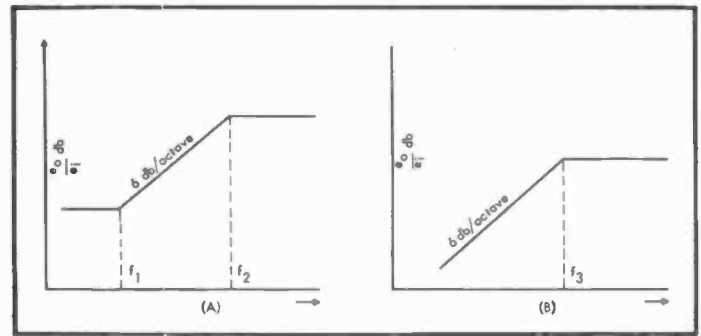The basic network used is shown in

Fig. 4. (A) of Fig. 5 shows the straight line approximation of the behavior of this network. The corner frequencies are given by

$$f_1 = \frac{1}{2\pi CR} \qquad (1)$$

$$f_2 = \frac{1}{2\pi C \left(\dfrac{R_1 R_2}{R_1 + R_2} \right)} \qquad (2)$$

where C is in microfarads, and R_1 and R_2 are in megohms. (B) of Fig. 5 shows the behavior when R_1 is removed from the network. In this case,

$$f_3 = \frac{1}{2\pi CR_2} \qquad (3)$$

The low-frequency networks may now be designed. The COL and RIAA networks will be as in Fig. 4, while the NAB and AES networks will have R_1 missing.

It is impossible to design the high-frequency network with only one corner (f_1) since this would require an indefinite rise at 6 db per octave with increasing frequency (again, Mother Nature won't stand for this). Sooner or later, the curve must flatten out. Consequently, while the network of Fig. 4 is still used, f_2 is selected so as to be at least a decade beyond the audio range, so that its influence is not felt in the operating range (20–20,000 cps). Figure 6 shows the actual network.

While the actual computed values are given in Fig. 6, the closest available

SWITCH POSITIONS (LF — HF)

1 — RIAA - RIAA
2 — COL - NARTB
3 — NARTB - OLD AES
4 — OLD AES - NARTB
5 — NARTB - NARTB
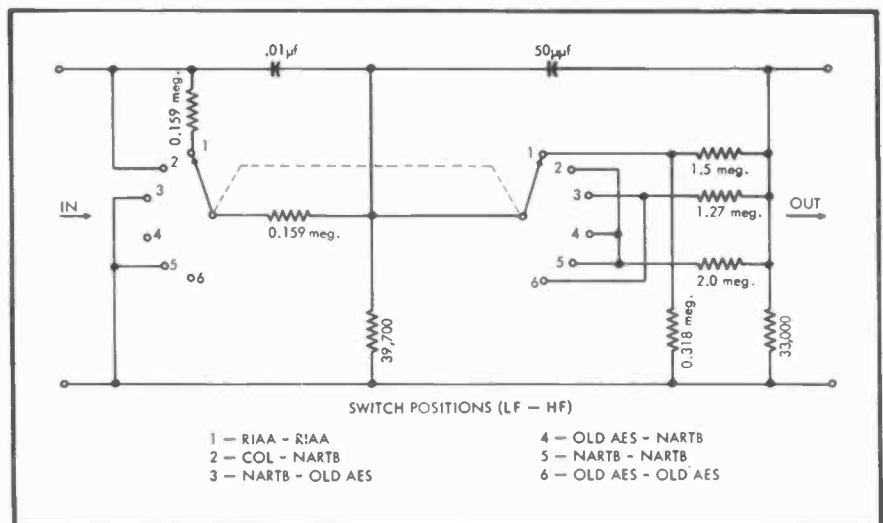6 — OLD AES - OLD AES

Fig. 6. Complete schematic of the recording characteristic simulator

Transistor Action

PAUL PENFIELD, Jr.

The physical principles underlying transistor action are discussed, and the basis of operation for a number of junction devices reviewed. No mathematics is required to understand this intuitive explanation.

TRANSISTORS ARE NOW so common that engineers and radio hobbyists are beginning to make wide use of them in audio applications. If the designer of transistor circuits is to be anything more than a "tinker" he will want to know something about the physical principles underlying the devices he is using. While it is true that good circuit design can be done by using the "black box" technique, in which the actual device is replaced by an equivalent circuit for computational purposes, nevertheless the astute engineer can make better use of physical transistors if he has a clear understanding of the physical principles involved in transistor action.

Much literature is available for those readers with considerable mathematical training; however a clear and simple explanation of transistor action in intuitive terms is not available. Either such explanations bring in much extraneous material and bore the reader, or else skip over the important parts, or else arrange the material in a fashion which prevents the reader from seeing similarities in the various junction devices, which similarities can be used to advantage in the explanation. No good intuitive explanation of junction transistor action is available.

This article has, I hope, hit a mean between too much background material and too little, and has arranged the explanations in such a way that similarities between various junction devices become immediately apparent. In connection with the former statement, the author assumes that the reader has some intellectual awareness (not an unwarranted assumption for the readers of AUDIO) and is not unwilling to accept some of the preliminary statements concerning modern physical theories without proof. If the reader has had some experience with transistors, so much the better. And if he has made an attempt in the past to understand transistor action, also so much the better.

Part I—Physical Fundamentals

With this short introduction, we can proceed to the business at hand. The article is divided into seven sections, which should be read sequentially. At the end of each section is a short review of the important points covered.

It should not astound the reader to find that modern physicists believe all matter to be made of "atoms," each composed of a central body, the "nucleus," and one or more "electrons" which may be thought of for the purposes at hand as revolving around the nucleus. Each electron has associated with it a negative charge of value denoted by "e." The atom as a whole is electrically neutral, the nucleus having a positive charge of e times the number of electrons revolving around it.

It will simplify our explanation if we consider the electrons in any atom divided into two categories: "bound electrons" and "valence electrons." It is the valence electrons, ranging in number from zero up through seven, that determine some of the chemical properties of elements.

A group of atoms of the same kind (that is, the same number of electrons in each atom) forms material that is known as a single "element." Hydrogen and oxygen are examples of elements. In addition, atoms can combine with atoms of other elements in certain ways to make "molecules" which in turn form material that is known as a "compound," to distinguish it from an element.

Material present in the world is often classified generally speaking as "solid," "liquid," or "gaseous." One important type of solid is known as a "crystal." Crystalline substances are characterized by the fact that their individual atoms or molecules are arranged in a definite mathematical pattern. The forces which act to hold together a crystal are exceedingly strong. One such force arises from the "covalent bond," which is a configuration of two valence electrons, one from each of two atoms, between which the bond is located. This configuration happens to be quite stable. Note that two electrons are required to form this bond.

With respect to electrical conduction properties, solids can be classified as either "conductors," "insulators," or "semiconductors," with surprisingly little ambiguity. In conductors, the valence electrons are quite free to move about the material without much opposing force. On the other hand, in insulators, the electrons are *not* free to move about, hence cannot flow to form a current. Typical conductors are copper, silver, aluminum, brass, etc., including most other metals. Typical insulators are wood, paper, mica, glass, cloth, etc. An example of a semiconductor, of which there are many known, is crystalline germanium. In order to explain semiconduction further, we'll look at the germanium crystal structure.

Germanium is the most-used material for making junction devices. A germanium atom has four valence electrons. It can form a stable crystal structure by forming four covalent bonds, with its four neighboring atoms. The configuration, that is the crystal lattice, is in three-dimensional space, and is known as the diamond structure, because crystalline diamond has the same form. Often the structure is represented in two-dimensional space by rows and columns of germanium atoms, as in *Fig. 1*. Since the three-dimensional distribution of atoms is quite hard to picture, we will not attempt to draw it here.

Since all four valence electrons are used up, there is none left over to contribute toward a conduction current. Thus one might at first think that crystalline germanium is an insulator. However, two means exist to produce current-carrying, or "conduction," electrons within a sample of germanium crystal. First, thermal agitation of the atoms[1] can at room temperature be sufficient to knock a few electrons out of their covalent bonds. Not many, but a few. This situation is shown diagrammatically in *Fig. 2*. And in addition, if the specimen is illuminated with light[2], the light energy of the photons can disrupt a normal covalent bond. These two means of producing conduction electrons prevent crystalline germanium from being an insulator.

Now let's think about what happens to the conduction electron and the bond it left. The electron may merely drift away through the material. Since the

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[1] Remember that, when viewing things from an atomic level, temperature is merely a measure of the rate at which particles are "bouncing around"—the higher the temperature, the faster the atoms, which can move somewhat within their specified position in the crystal lattice, jiggle around.

[2] Remember that ordinary light can be thought of as little packages, or "photons," of energy.

covalent bonds in the lattice can accommodate only two electrons, it cannot become a permanent fixture at any one spot in the lattice. Or else it may immediately fall back into the bond it left. In general, the electron is removed with such energy that it drifts away from the spot where it was. The bond, on the other hand, is now lacking an electron. A bond in this state is called a "hole." A surprising feature of the lattice is that this hole can move in roughly the same fashion as an excess electron. Its movement, of course, consists of having an electron from a nearby bond jump into the original bond, thus moving the hole to the spot where the electron came from. The hole, being the lack of an electron, possesses a positive charge equal to e. For the purposes of transistor physics, the hole may be thought of as a particle with a positive charge e, and with characteristics similar to those of a conduction electron.

A hole can re-combine with an electron by the simple process of coming close enough so that their electric attraction will cause the electron to "fall into the hole," to put it crudely. Sometimes this process is accompanied by a release of energy in the form of a photon; more often it is not. (Of course if the hole and electron could not recombine, the crystal would eventually fall apart from lack of covalent bonds. Needless to say, this doesn't happen.)

Since the concept of the hole as a current carrier is paramount in the discussion that follows, the reader should fix in his mind the following facts about the hole: (1) It may be created by somehow drawing an electron away from the covalent bond, (2) It and an accompanying electron may be simultaneously created by thermal agitation within the crystal, (3) It and an accompanying conduction electron may be created by an incident photon, (4) The hole may be considered as a positively charged particle when thinking of its current-carrying abilities, (5) A hole and a conduction electron will re-combine if
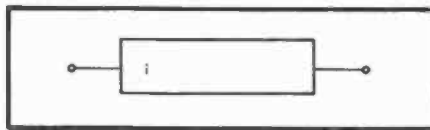

Fig. 3. A piece of intrinsic germanium with two leads, one at either end, like a common resistor. It can be used as a small photocell.

they happen to meet, and (6) A flow of holes in one direction just as much constitutes current as a flow of electrons in the opposite direction.

With what we know about a germanium crystal already, we can see that a single piece of germanium, made with two leads, similar to a resistor, (see Fig. 3) could perhaps perform some useful functions. For example, the temperature dependence of the "intrinsic current," that formed by thermal agitation, could be utilized in making the device act as a thermometer, with a conductivity that would decrease with increasing temperature. Fortunately, more reliable and more sensitive electrical thermometers, such as thermistors, are available.

However, the device is used as a photocell. Incident light produces electron-hole pairs, which, if a voltage is applied across the device, increase the current flowing. The so-called Germanium Photoresistor (type 1N189) is an example of commercial use of photoconductivity. (Actually, for reasons which we won't go into here, n-type germanium, as described below, is used rather than intrinsic germanium in this photoresistor.)

From this section the reader should understand in an intuitive sort of way the difference between insulators, conductors, and semiconductors. He should remember that germanium forms a stable crystal lattice in which all valence electrons are used up, but that conduction particles (electrons and holes) can be formed even at room temperature by thermal agitation, and also by incident photons. Holes can be treated in much the same way as electrons—as real par-

ticles. A flow of holes in one direction just as much constitutes current as a flow of electrons in the opposite direction.

Part II—Impurities in a Germanium Crystal

The useful properties of semiconductors do not end with the pure crystals. With controlled amounts of special impurities, useful devices can be made.

Remember that in the last section we were talking about a pure sample of germanium. This was a semiconductor because the germanium had four valence electrons, all of which formed covalent bonds. If, however, one of the germanium atoms is replaced by an atom with only three valence electrons, such as indium, there will be a hole automatically formed in the lattice. The impurity atom does not break up the lattice structure —instead it fits in as well as it can, forming a hole. Of course, the crystal as a whole is still electrically neutral—the indium atom has one less positive charge in its nucleus. But nevertheless a conduction particle—namely, a hole—has been formed in a crystal which otherwise had none, except for occasional thermally or light-caused pairs. The situation is represented in Fig. 4.

A crystal of germanium "doped" with indium atoms (say one for every fifty million or so germanium atoms) can carry current and therefore is a better conductor than pure germanium. Because the current carriers are almost exclusively positively-charged holes, it is known as "p-type" germanium. The indium atoms are known as "acceptors" because they form bonds which accept electrons from nearby bonds, forming holes. Note that holes have been introduced without forming corresponding conduction electrons.

Similarly, a crystal can be doped with an element with five valence electrons, such as antimony, to form "n-type" germanium. The antimony atoms are called "donors" because when they fit
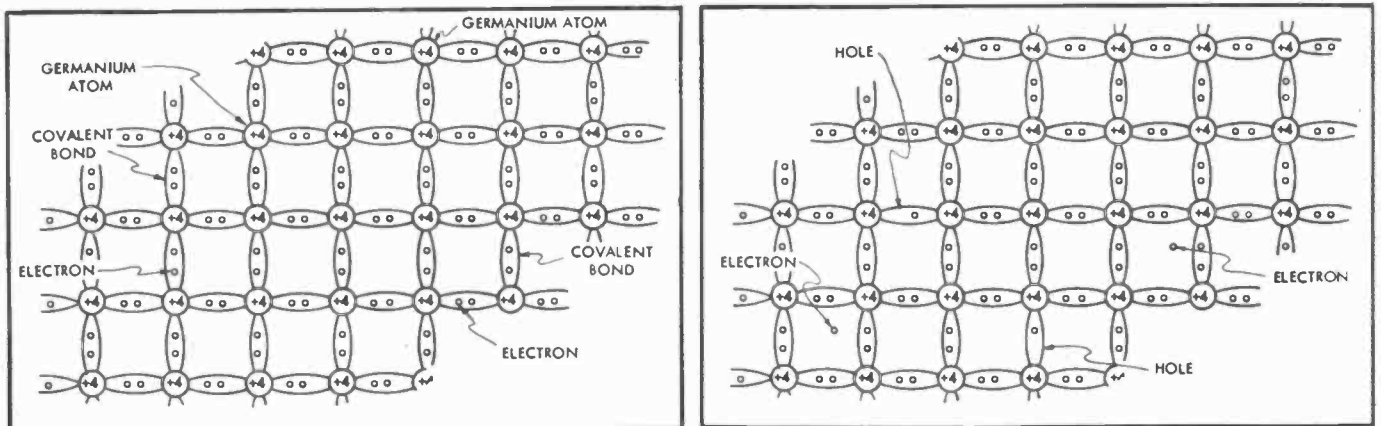

Fig. 1 (left). Representation of pure germanium crystal. Each germanium atom forms four covalent bonds with its four adjacent neighbors. Fig. 2 (right). Intrinsic germanium. Note the temperature-caused holes and electrons.
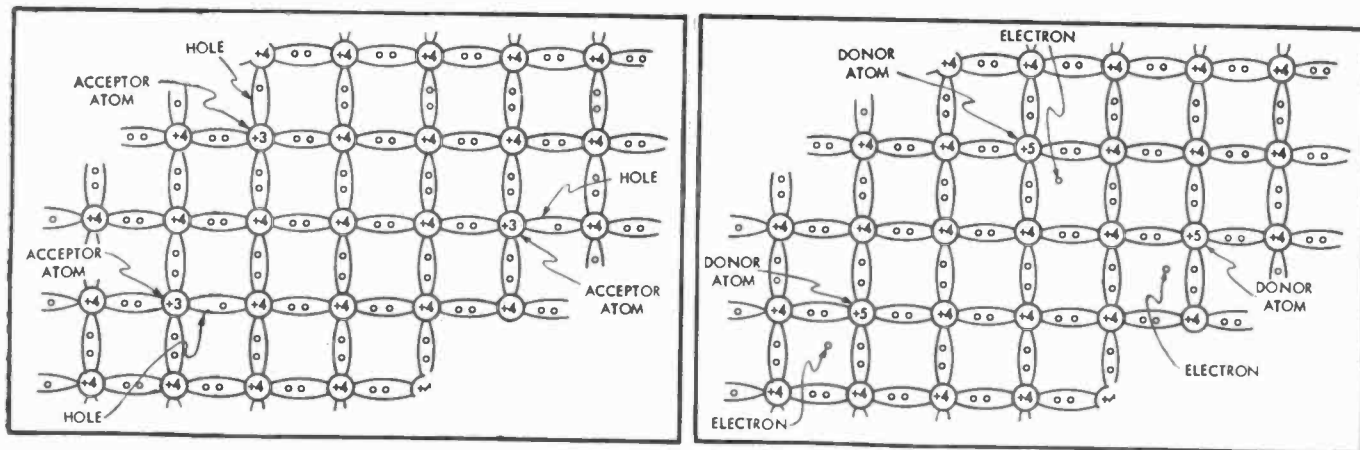
Fig. 4 (left). The addition of acceptors to otherwise pure germanium creates holes without creating excess electrons.
Fig. 5 (right). The addition of donors to otherwise pure germanium creates conduction electrons without corresponding holes.

into the lattice there is an extra electron left over which is free to act as a current carrier. This is represented in *Fig. 5*. At very low temperatures (much below room temperature), the carrier introduced is attracted to the impurity atom, because they have opposite charges. However, thermal agitation shakes off these impurity carriers relatively easily.

Pure germanium, free to conduct only because of thermally-generated carrier pairs, is said to possess "intrinsic conductivity," as opposed to "n-type conductivity" (predominantly by means of excess electrons) or "p-type conductivity" (predominantly by means of holes).

The role played by the three types of germanium, p-type, n-type, and pure, is very important in transistor physics. The reader will want to remember from this section that: (1) in n-type germanium, formed by the introduction of donor atoms, the principal current-carrying particle is the electron, and the remaining donor atom in the lattice structure has a local positive charge, which however, does not succeed in "trapping" an electron and keeping it tied down at normal temperatures; (2) in p-type germanium, formed by the introduction of acceptor atoms, the principal current-carrying particle is the hole, and the remaining acceptor atoms in the lattice structure have a local negative charge, which however, does not succeed in "trapping" a hole and keeping it tied down at normal temperatures; (3) suitable juxtaposition of n-type, p-type, and intrinsic areas produces useful devices.

Part III—Action at a Junction

If we have a crystal of germanium which is half p-type and half n-type, the surface separating the two areas is known as a "p-n junction." On one side of the junction we see acceptor atoms with their local negative charge distributed throughout the area, and holes also distributed. On the other side are immovable donor atoms and many

conduction electrons wandering about. Each side is at first glance electronically neutral—with equal positive and negative charge. Right near the junction there will be some diffusion of electrons and holes, with some re-combination taking place. As a result the remaining acceptor and donor atoms set up a small electric field, and the equilibrium condition of the crystal is that further diffusion of electrons and holes be stopped by a small electric field localized right at the junction.

Although the existence of this localized electric field means that the two sides of the crystal are at a slightly different electric potential, the reader should not jump to any conclusions such as that of the junction being replaced by a battery, or anything so foolish. The junction of course cannot supply power to an external resistor, and furthermore the potential difference between the two sides is a function of the temperature, and in addition can be varied by applying external power, as we shall see later.

In order to understand the rectifying action at a junction, consider the piece of crystal with a p-n junction in it, with leads attached to each end of the crystal, on either side of the junction, as shown in *Fig. 6*.

Normally, enough electrons and holes have diffused together so that quite near the junction there are no carriers (i.e. electrons or holes) present—and more carriers will not come near the junction because of the small localized electric field set up, as explained earlier. Now, if a battery is connected so that its positive terminal is con-
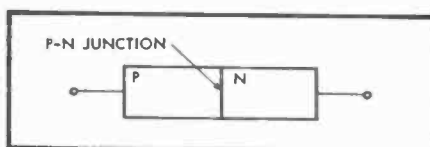
nected to the p-region, and its negative terminal to the n-region, holes in the p-region will be driven away from the end of the crystal by the action of the battery, and more holes will flow into the crystal from the battery (that is, some bonds near the end of the crystal will lose one electron). Similarly, at the other end of the crystal, electrons are being driven away from the end by the action of the external battery, and more supplied to the crystal. With the externally-caused electric field in such a direction to push the holes and electrons toward each other, the crystal is said to be biased in the "forward" direction. When the electrons and holes reach the center of the crystal, they pass right through the junction, and in general travel a small ways into the other half of the crystal, whereupon they combine with carriers of opposite sign, thereby vanishing. But since more holes and electrons are continually being supplied by the battery, continued current flows through the device.

On the other hand, if the battery leads were reversed, so that the positive terminal went to the n-type germanium, and the negative terminal to the p-region, the action would be such as to draw the electrons and holes within the crystal away from each other—that is, toward the ends of the crystal. Clearly very little current can flow in this situation, since the electrons and holes cannot recombine easily.

Thus we see that this device, known as a "junction diode," can pass current easily in only one direction. The common 1N91 is an example of such a rectifier. Crystal rectifiers, made both from germanium and silicon, are in limited use already, and are expected to replace vacuum tube rectifiers in many applications as soon as the price falls a bit more.

The forward current in these devices is limited by the I^2R losses within the germanium, and also by the fact that the junction electric field never is com-

Fig. 6. A p-n junction is the surface separating p-type from n-type germanium. Here is a junction diode, using one p-n junction.

pletely eliminated by the externally-applied field. The reverse current that flows is produced mainly by imperfections in the crystal construction, or else by thermally-generated carrier pairs. If a thermally-generated pair occurs near enough to the junction so that there are no other carriers present, the hole will be attracted by the p-region, and the electron will move toward the n-region, and their motion will constitute current. And if a reversely-connected junction diode is illuminated with light, incident photons will produce carrier pairs, increasing the current. The effect is made use of in photo-diodes, as explained in the next section.

The principles the reader should retain from this section are: (1) the surface between n-type and p-type material is known as a p-n junction, and a two-terminal device employing a p-n junction is known as a junction diode. (2) At thermal equilibrium with no external voltage applied, a slight electric field is set up across the junction which keeps the holes on one side and the electrons on the other. (3) If a diode is biased in the forward direction, the holes and electrons are pushed by the external power source toward the junction, near which they re-combine. (4) If a diode is biased in the reverse direction, the holes and electrons are pulled away from each other and away from the junction, so little current flows. (5) When a diode is reversely biased, any carriers, whether hole or electron, which are placed near the junction will flow toward the end of the diode.

Part IV—Some Other Junction Devices

The last statement in the last section is extremely important and is fundamental to an understanding of transistor action. "When a diode is reversely biased, any carriers, whether hole or electron, which are placed near the junction will flow toward the end of the diode." If the reader understands nothing at all from the last sections but this, he's still ready to proceed.

In the last section we described the action of a junction diode. Now let's take that same diode, and establish a reverse bias on it by connecting an external battery with its positive terminal on the n-region, and its negative terminal on the p-region. The reverse current is now due only to thermally-generated carrier pairs created in the vicinity of the junction. However, if we shine light on the junction, more hole-electron pairs will be formed, and consequently more current will flow through the device. When connected and used in this way, the device is known as a "junction photo-diode"—the 1N188 is an example of a germanium photo-diode. It is possible under good condi-

tions to achieve a yield of nearly 1—that is, one hole-electron pair for every light quantum hitting the diode. The device is thus seen to be a practical, very small, sensitive photocell.

However, illumination and thermal agitation are not the only ways to introduce holes or electrons near a reversely-biased junction. Consider the case of a three-region piece of germanium—with two p-regions at the ends, and a small, narrow n-region in the middle. See *Fig. 7*. Suppose each region is brought out to a terminal, and that between the middle and right regions a reverse bias is applied—by applying the positive terminal of a battery to the middle region, and the negative terminal to the end. Thus one junction is reversely biased. And little current will flow. Now, however, we shall connect a small battery between the middle region and the left end—
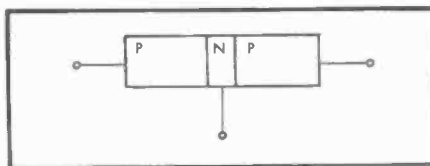

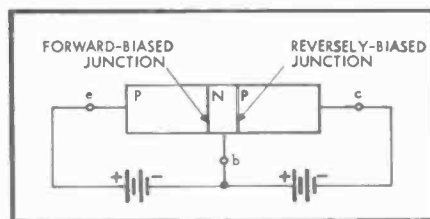Fig. 7. A two-junction device, with each of the three regions brought out to a terminal.

Fig. 8. Normal biasing of the p-n-p junction transistor. The emitter junction is biased forward, the collector junction backward.

this time in the forward direction. What will be the result of this connection, shown in *Fig. 8*?

At first glance, one might be tempted to treat the two junctions separately, and say that the one will remain non-conducting, and the other will conduct. However, this is not the effect observed. Instead, holes that enter the middle n-region from the forward-connected left-hand junction will see only the other junction ahead of them —and will act just like any carrier introduced in the region of a reversely-biased junction—they will flow through the junction and out the other end of the crystal. A few, to be sure, will re-combine with the electrons within the n-region, but the vast majority, especially if the n-region is thin, will proceed through both junctions and thus pass right through the crystal.

If the reader hasn't guessed it by

now, the three-terminal device we have been talking about is a "p-n-p junction transistor." The end terminal which emits the holes into the middle region is known, appropriately enough, as the "emitter." The other end terminal, which collects all the holes which the emitter injects, is known as the "collector." The middle region is called the "base." The theory given above for the operation of the transistor is known as "transistor action"—the control of current through a reversely-biased junction by means of current injected near the junction by another electrode (in this case another junction).

Because of its importance let's go through it again: First, a reverse bias is set up between the base and the collector—that is, across the collector junction. The only collector current which flows (if the transistor is shielded from light) is due to thermally-generated hole-electron pairs created near the collector junction. Now, however, a forward bias is set up between the base and the emitter—that is, across the emitter junction. Thus, much current flows through the emitter. The question becomes, "what happens to the emitter current once it reaches the base?" First, a small portion of the emitter current is due to electrons which flow across the emitter junction from the base—these recombine with holes somewhere within the emitter. Secondly, some of the holes that enter the base re-combine with electrons within the base. These two together constitute the current which flows through the base lead. However, if the transistor is properly designed, the vast majority of holes pass right through the base into the collector, and serve to increase the collector current. Of course, superimposed upon the emitter current might be some sort of fluctuating signal which requires amplification.

But now the question may arise, "so what?" We just saw that the collector current is always (in the normal operating region) less than the emitter current. Is that amplification? Well, it's not too hard to see that, no matter what the collector-to-base voltage is, so long as the collector junction is biased reversely, the collector current is determined almost completely by the emitter current. In other words, a large resistor in series with the collector which changes the collector voltage when the collector current changes, will not appreciably affect the amount of the collector current, which will still be determined by the emitter current alone. Thus our input signal, at a very small voltage, can be increased to several times this voltage—in other words the transistor connected this way will amplify.
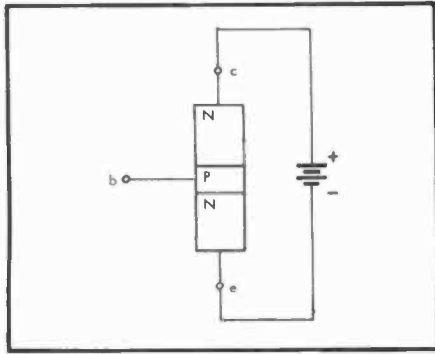
Since the base terminal is common

Fig. 9. Common-emitter biasing method for a p-n-p junction transistor.

to both the input and the output of the simple amplifier, it is called a common-base, or grounded-base configuration. We will see in the next section that in another configuration the device can act as a current amplifier.

This section is, of course, the most important section in the article. The sections before this merely served to introduce certain concepts used here. The following three sections will further describe transistor action, and will describe a few more commercially available junction devices of interest. Out of this section the reader should have learned: (1) A photo-diode is merely a reversely-biased diode whose current is controlled by incident light. The incident photons produce electron-hole pairs near the junction. A small, practical, sensitive photocell is the result. (2) A p-n-p junction transistor is merely a three-terminal device having two p-n junctions "back-to-back," separated by a small n-region. The collector current is determined, in the grounded-base configuration, by the emitter current only—not by the collector' voltage. A small, efficient, amplifying device is seen to result. (3) Transistor action is merely the control of current through a reversely-biased junction by means of injecting proper carriers in the vicinity of the junction from another electrode.

Part V—Grounded-Emitter Operation

We saw in the last section that transistor action is merely the control of current through a reversely-biased junction by means of current from another source deposited near the junction. In a p-n-p junction transistor with the base lead common to both the collector and emitter circuits, a majority of holes coming from the emitter pass right through the base region into the collector. Since the collector current is less than the emitter current, the transistor does not amplify current, although we saw that it would amplify voltage. Let us call the fraction of emitter current which *does* reach the collector α.

α will normally be just a trifle less than one—for the sake of example,

let's say that α = 0.95. If we call the emitter current i_e, the collector current i_c, and the base current i_b, we immediately see that

$$i_c = \alpha i_e \qquad (1)$$

and

$$i_b = (1 - \alpha) \; i_e \qquad (2)$$

and therefore

$$i_c = \left(\frac{\alpha}{1-\alpha}\right) i_b = \beta \; i_b \qquad (3)$$

where β is called the "grounded-emitter current gain," and

$$\beta = \frac{\alpha}{1 - \alpha}$$

so when α = 0.95, β = 19.

If we consider the base as the input terminal, and the collector as the output terminal, we can see from Eq. (3) that the collector current is many times greater than the base current, so when the transistor is operated in the grounded-emitter configuration, that is with the emitter common between the input circuit and the output circuit, current amplification takes place.
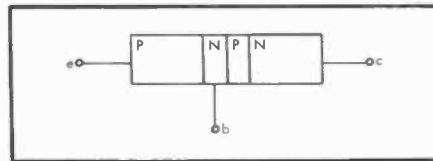

Fig. 10. An n-p-n junction transistor behaves in exactly the opposite way as a p-n-p junction transistor. All polarities are reversed.

Furthermore, by the same sort of reasoning as was employed in the last section, this current can be made to flow through a large load resistor, so voltage amplification occurs as well.

However, this mathematical derivation may not be at all convincing to the reader, so we'll go through the grounded-emitter stage again, from a physical viewpoint.

Consider the case when a battery is placed directly between the collector and the emitter, with the base left unconnected, as in *Fig.* 9. Of course, for the p-n-p junction transistor under consideration, the collector should be connected to the negative battery terminal, and the emitter to the positive. In this case, the collector junction is biased in the reverse direction. In addition, the potential of the base is slightly higher than the potential of the emitter. For if it were not, emitter current would flow, one twentieth of which would remain in the base to charge it up to the point where no more current will flow. This is the condition of the transistor for equilibrium. Now consider the case when an electron appears in the base region for one reason or another. Perhaps it was introduced in through the base lead, in which case it represents base current. The potential of the base region is

lowered somewhat by the presence of the electron, with the result that the emitter injects holes into the base to try to raise its potential up to the point of equilibrium. For each electron in the hole region, the emitter injects twenty, or $1/(1 - \alpha)$ holes—one of which recombines with the electron, and 19 of which flow through the base into the collector. Thus a very small amount of current through the base can control a rather large amount of current through the collector. In fact, 1 milliampere through the base can control 19 milliamperes through the collector, or $\alpha/(1 - \alpha)$ milliamperes. Thus the current gain is seen again to be $\alpha/(1 - \alpha)$. This is, of course, the same result we achieved two paragraphs ago by considering the device mathematically.

The reader should recognize the fact that this "transistor current multiplication" is merely another manifestation of transistor action, and is quite equivalent to the statement about transistor action made at the end of the last section.

The reader should note the following pertinent points arrived at in this section: (1) In the grounded-emitter configuration, the transistor is capable, to a first approximation, of a current gain of $\alpha/(1 - \alpha)$. (2) This transistor current multiplication (often referred to as "hook multiplication") is merely another manifestation of transistor action—and thus is entirely equivalent to the former statement of transistor action.

Part VI—Other Two-Junction Devices

In this section we will discuss two more two-junction devices—both of which rely for operation on transistor action.

Besides p-n-p junction transistors, n-p-n junction transistors exist as well. See *Fig.* 10. Transistor action is exactly the same in these n-p-n units, except that all battery polarities and current directions must be reversed. For example, instead of injecting holes into the base, the n-p-n emitter injects electrons. The n-p-n transistor is exactly the same, to a first approximation, but opposite in polarity to a p-n-p transistor.

Let us consider an n-p-n transistor operating grounded-emitter—that is, with only one battery connected between the collector and the emitter, with the collector positive[3] as shown in *Fig.* 11. As a first approximation, we stated in the last section that no current would flow. As a matter of fact, however, thermally-generated electron-hole pairs will be created near the reversely-

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[3] From his knowledge of transistor action the reader should be able to verify in his own mind that the collector junction will be biased reversely with this connection.
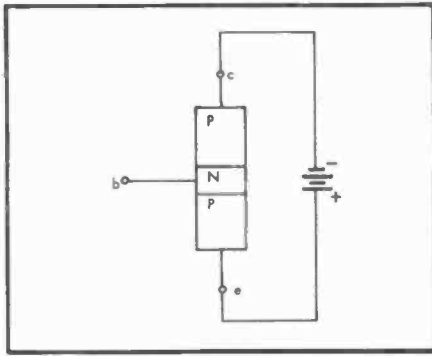
Fig. 11 A grounded-emitter n-p-n junction transistor. This is correct biasing for use as a photo-transistor.

biased collector junction, the electron of which will be pulled by the positive terminal of the battery into the collector, and the hole of which will proceed into the base. In order to counteract the presence of the hole in the base, the emitter must inject 20, or $1/(1-\alpha)$ electrons into the base, one of which will re-combine with the hole, and 19 of which will pass through into the collector. Since this is so, the collector current will be twenty times the rate of creation of carrier pairs due to thermal agitation alone. In short, the problem of thermal current in the grounded-emitter stage is greater than the problem when the transistor is fed with the base common. "Bias stabilization" under widely varying temperatures is often a severe problem—although there are techniques for reducing the effect considerably.

So hook multiplication is a problem when thermal current is considered. However, it can be used to advantage in certain cases. Consider, for example, the same n-p-n transistor with its base disconnected, and normally-biased as in *Fig.* 11, and with the reversely-biased collector junction illuminated. For every light-caused electron-hole pair formed, twenty electrons will flow through the collector circuit. Thus, in effect, the photo-diode current is multiplied by the hook multiplication ratio—$1/(1-\alpha)$. This device is known as a "photo-transistor," and is often described as a photo-diode with a built-in amplifier. Texas Instruments type 800 is typical of modern photo-transistors.

From this section the reader should retain the following points: (1) n-p-n junction transistors operate in exactly the same way as p-n-p junction transistors, except that all battery polarities are reversed. (2) Grounded-emitter transistors with the base left open-circuited exhibit hook multiplication of thermally-generated current. (3) Photo-transistors use the inherent hook multiplication of transistors to advantage in producing a more sensitive photocell than the photo-diode. For some pur-

poses they can be thought of as a photo-diode with a built-in amplifier.

Part VII—Three-Junction Devices

A device can be made which is analogous to the photo-transistor in the same way that an ordinary p-n-p junction transistor is analogous to a photo-diode. For this operation, some current is injected by a fourth element placed quite near the collector junction. This element serves the same purpose as the emitter of a normal junction transistor, and so in the composite device is called the emitter. What was formerly the collector plays the role of the base, so it is now known as the base. What formerly was the emitter now becomes the collector.

The device, known as a "p-n hook transistor" is shown in *Fig.* 12. Federal Telecommunication Labs makes an experimental model, type CP-611. The device can be most easily understood by considering it connected grounded-base. In this connection, the three ele-
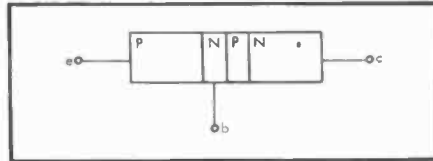

Fig. 12. A three-junction device, three of whose sections are brought out to terminals. This is the hook transistor.

ments at the right (as in *Fig.* 13) form a hook multiplier—the same way that an n-p-n transistor normally would. Emitter current injected at the left passes into, and 95 per cent (or α) of it through, the base region. The portion which passes through the collector junction in the middle finds itself in what looks like the base of an n-p-n junction transistor, so biased that hook multiplication will occur. For each hole so present, 20 or $1/(1-\alpha)$ electrons will be drawn from the collector region to the far right, 19 of which will again pass through the reversely-biased junction in the middle into what is called the base of the composite hook transistor. If the base is grounded, these will flow out of the base, in which case the "collector current" will be many times the "emitter current." In fact, if the region at the left has a normal current gain of α_1, and the three elements at the right taken together have a current gain of α_2, (both less than unity), the ratio of collector current to emitter current will be $\alpha_1/(1-\alpha_2)$, or approximately β_2.

Note that the base was grounded in the last discussion. The device in this configuration possesses a current gain greater than one—something which a normal junction transistor does not. It should be noted that care must be

taken in designing circuits around the hook transistor, since it, like the point contact transistor, which also can have a current gain greater than 1, is unstable in certain configurations. In fact, too much resistance in the base circuit can make the device unstable.

However, Federal Telecommunication Labs reports that in their transistor, the over-all current gain from emitter to collector is very much a function of the collector current, dropping down to practically 1 for low-current operation. For this reason, circuit design problems may be less severe than otherwise expected.

Another possible device similar in form to the hook transistor may find use someday. We shall call it the hook photocell. If a hook transistor is arranged in some stable arrangement, with the base removed from ground by means of a series resistor, and then the center, reversely-biased junction is illuminated by a light source, a current multiplication will occur which is somewhat more than that due to one hook multiplication alone. Thus the device could be more sensitive than the photo-transistor. Whether a device of this sort will ever find much practical use remains to be seen, but it is mentioned here so that it may be considered a logical extension of the practical devices. The problem of a very large thermally-generated current would limit the usefulness of the hook photo-cell drastically.

Attempts to make a five-terminal transistor using two hook multiplications within the same crystal will probably be doomed to failure, for the injected carriers must be placed square in the middle of the middle region of the hook transistor for such a device, and the problems of building such a device out of a single crystal are quite difficult, as the reader may be able to see. This is not to say that useful five-terminal devices will not be made using junctions and transistor action—but they will probably have two or more terminals attached to one region, as present tetrode transistors and double-base diodes do.
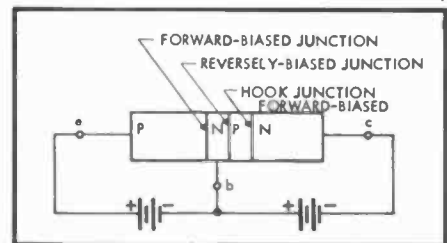
Out of this section the reader should

Fig. 13. Correct biasing for the hook transistor. This is the same as the p-n-p junction transistor. Only difference is that the collector has a "built-in hook mechanism."

Transistor Preamps

H. F. STARKE

A discussion of practical considerations involved in the design of preamplifiers for phonograph reproduction—with their attendant low-frequency boost—and flat amplifiers for microphone applications.

TRANSISTOR PREAMPLIFIERS may be classified broadly according to their various applications — in fact, it might be better to say that they *must* be so classified because (1) frequency discrimination networks are, in general, somewhat more difficult with transistors than with tubes and (2) the input resistance varies enormously with circuit configuration, operating current, and transistor alpha (α). The present account will be concerned mainly with two general types of amplifiers: those having flat responses and intended as dynamic microphone preamps or as impedance transformers, and those designed as preamps working from an inductive pickup into an otherwise flat main amplifier.

Values have been published for the new standard record curve (RIAA) to the nearest millibel.[1] A straight line between the extremes at 30 and 15,000 cps has a slope of very nearly 4 db per octave and if this line is raised a trifle it does not depart more than 1.6 db from the RIAA at any point (*Fig.* 1). Although it must be seriously doubted whether the most critical observer could distinguish between the two, the point is largely of academic interest because, as it happens, an amplifier with a slope of 4 db per octave throughout the entire band would be more difficult to design

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[1] "The Proposed AES Disc Standard," (RIAA curve). *J.A.E.S.*, Jan. 1954.

than one having RIAA response with its slight 1,000-cps flattening. This is because resistive-reactive networks are characterized by slopes of 6 db per octave—obviously too much for a nine-octave band.

In general, it is a mistake to interpose a resistive element between the transistor input and the generator for the purpose of obtaining frequency adjustment if a primary objective is that of providing the highest possible signal-to-noise ratio. The special case of a preamp designed for working from an inductive pickup, however, may be considered as a practical exception only to the extent that the designer would like to mitigate the severity of the input impedance problem at the expense of some degradation of the signal-to-noise ratio.

One form of the phono preamp is shown in *Fig.* 2 in which the following features may be recognized:

(1) No frequency networks between pickup and input.
(2) High-frequency rolloff by controlled input resistance.
(3) Low-frequency boost by means of a negative feedback loop from collector to collector. This technique is, of course, similar to the plate-to-plate loop of the vacuum-tube amplifier.
(4) Bass flattening by adjustment of the interstage blocking capacitor.

With an inductance of 0.52 Hy, the cartridge used during the development of this amplifier requires an input resistance of 6200 ohms for RIAA re-

sponse when used with a vacuum-tube amplifier. This drops very little for the transistor amplifier because the d.c. resistance of the cartridge is less than 500 ohms.

If the crossover point appears to require adjustment, this can be most readily accomplished over a reasonable range by changing the operating current of the first stage. It will be noted that the base divider for this stage consists of equal values of resistance in order to keep the parallel impedance of these elements as high as possible. The adjustment of operating current should accordingly be made by changing the emitter resistor rather than the 1 to 1 ratio of the base-divider elements. If the required current change turns out to be considerable, it will be necessary to modify the load—the general objective being to maintain a collector voltage between 1.5 and 2 volts in this stage from a supply of 6 to 7 volts.

The low-frequency boosting network with a crossover of approximately 500 cps works from a source consisting of the second-stage collector load (20,000 ohms) in parallel with the output impedance of that stage, $R_c(1-\alpha)$, and works into a load made up of five impedances in parallel:

(1) First-stage output impedance.
(2) First-stage collector load.
(3) Second-stage upper base resistor.
(4) Second-stage lower base resistor.
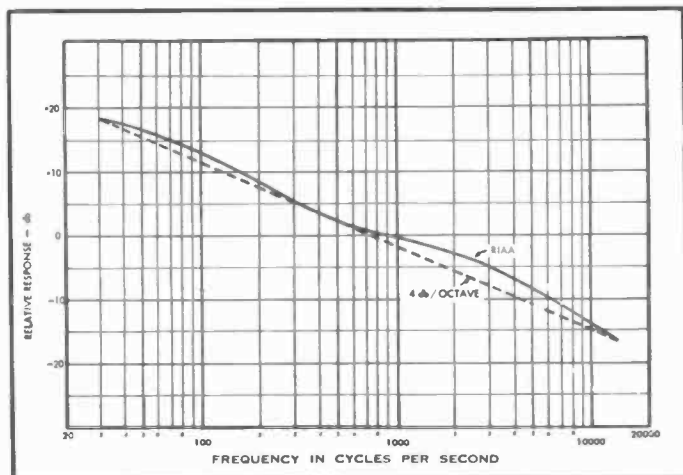(5) Second-stage input impedance.

Fig. 1. RIAA playback characteristic (solid line) plotted against a straight-line response curve having a slope of 4 db per octave (dotted line).
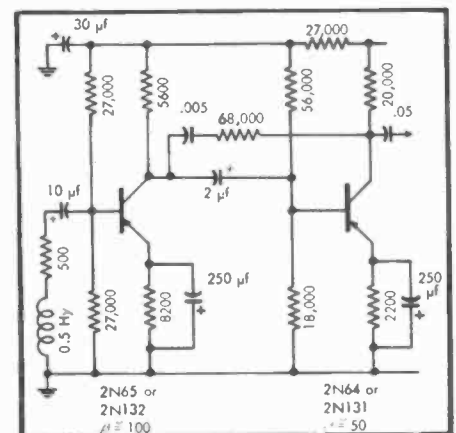

Fig. 2. Transistor phono preamp in initial stages of design.

It will therefore be recognized that this network cannot be predicted to the same exactitude that will be found in vacuum-tube circuits unless at least the alphas of the two transistors are known and their values of R_c and R_b are not too far from average for the operating current. Also, if a substantial change is made in the first-stage operating current for the purpose of correcting high-frequency rolloff, the corresponding change in first-stage output impedance will have some effect upon the low-frequency-boost network although the converse effect is much less. Both input and output impedances vary inversely with current, of course, and when the first-stage current is reduced the slopes of both high and low networks are increased. For a given change, however, the effect is more pronounced for the LR network because R in this case is a single impedance (the input impedance of the first stage) while in the other case the first-stage output impedance is only one of several impedances in parallel.

Accordingly, if the first frequency run on the completed amplifier deviates from the standard by more than an acceptable amount in one or two places, the preferred order of adjustment would be:

(1) High-frequency rolloff.
(2) Low-frequency boost.
(3) Bass flattening.

The standard calls for a 3 db flattening at 50 cps and most of this is obtained from the rather low value of interstage coupling—the remainder being due to the emitter bypasses (250 µf) which have enough impedance at this frequency to drop the gain 0.6 or 0.7 db. More or less flattening is readily obtained by altering the value of the coupling capacitor with little effect upon the rest of the curve.

Practical Considerations

Since a two-stage amplifier with phase reversals in each stage has a final output in phase with the amplifier input, the input base divider should be decoupled. The filter serves the further purpose of dropping the supply voltage to the first stage and—as used here—leaves only the second-stage collector supply at the higher voltage. An earlier version of this form of preamp used 6 volts to both stages and resulted in a design which gave adequate performance in gain, noise level, and frequency response, but which was deficient in the matter of maximum undistorted output voltage. This has been corrected in the present amplifier by raising the second-stage supply and holding the emitter resistor to a reasonable value.

It will be recognized that no particular importance attaches to the value of 22½ volts other than the fact that some may

elect to use a battery in view of the low current demand and this is a value that provides 5 or 6 volts of output. On the other hand, if the supply is obtained from a 250- or 300-volt plate supply in the main (vacuum-tube) amplifier by means of a divider, there is no particular reason for setting the tap higher than 30 volts. The bleeder current in this divider should be at least four or five times the total current taken by the transistor preamp so that the voltage at the tap will not rise unduly if the preamp is switched off separately.

The 1000-cps voltage gain of this amplifier is approximately 52 db and if this is somewhat in excess of what is required, the first stage collector load may be tapped down without much effect upon over-all frequency response provided the low boost network is left at the collector.

In terms of comparative physical size, the 250 µf emitter bypasses stand out as the largest components physically, on the list. If an amplifier of this type is to be built into the pickup arm, it accordingly becomes desirable—because of both size and weight—to eliminate these capacitors. This leads to the form of preamp shown in *Fig. 3* which uses the type of regulation circuit in which the base divider feeds from the collector instead of from the battery. Compared to the
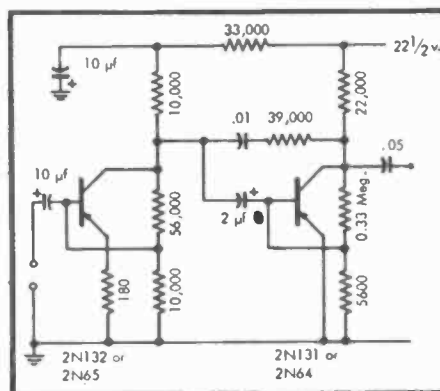

Fig. 3. Amplifier of Fig. 2 is modified to avoid use of large emitter bypass capacitors.

first circuit, this form has somewhat less regulation and (because of the negative feedback from collector to base) lower stage gain. The feedback also has the effect of reducing both the input and the output impedance of each stage apart from the rather low value of the resistor from base to ground. This latter can be increased only at the expense of achieving a poorer regulation factor.

The over-all combination proved to have an input impedance somewhat too low for the inductive pickup with a first-stage transistor having a current gain of 50. As an alternative to increasing the impedance by reducing the operating current—a procedure discussed

in connection with *Fig. 2*—a small resistor was placed in the first-stage emitter. Although either scheme appears to be workable and would probably result in approximately the same reduction in gain (other things being equal) it should be noted that the slight reduction in battery current must be balanced against maintaining or slightly improving the regulation factor.

The gain of 40 db at 1000 cps—although 12 db lower than the gain of the previous amplifier—is still adequate for the purpose. Since the second-stage collector supply is the same, the maximum output is also the same: 15 or 16 dbv. The impedance of the low-frequency-boost network is not much more than one-half that of the corresponding network of *Fig.* 2 because it works between the lower impedances resulting from the negative feedback. The 2-µf capacitor for bass flattening has sufficient effect with no augmentation from emitter bypasses—again because of lower impedance.

Adding Other Curves

For the benefit of those who would like to provide, by means of switching in other networks, a choice of frequency responses, the following practical difference between transistor and vacuum tube circuits may be noted:

Although, in general, a reactive impedance must equal the sum of the resistive impedances on either side at the crossover frequency, it is common practice in vacuum tube circuitry to disregard the source impedance (usually a plate load shunted by the plate resistance) and consider only the much higher following impedance—i.e., a grid leak shunted by the grid impedance. In direct contrast, the transistor circuit—because of the choice of circuit configuration or the use of feedback—may show a preceding impedance higher than, the same as, or lower than the following impedance and both must be taken into account.

At 15,000 cps, audio-frequency transistors will show some loss of gain due to frequency cutoff of alpha. On a flat amplifier, such as will be described in connection with dynamic microphone preamps, this can result in a response that is down 3 or 4 db at 15 or 20 kc in two stages. On the RIAA type of response, the effect may be noticeable as a convexity (facing up) of the portion of the curve between 2,000 and 15,000 cps. In brief, if the response is down by the correct amount at 15,000 cps, it will be 3 or 4 db too high at one octave lower or, conversely, if it is down by the correct amount at 7 or 8 kc, it will be 3 or 4 db too low at 15,000 cps. To determine how much alpha cutoff correction is required, it is of course in-

formative to run the curve to 20,000 cps despite the fact that the standard stops at 15,000 because—unlike the other end of the band where the 6 db/octave response due to low-frequency boost is severely modified by bass flattening—the slope of the high-frequency end of the RIAA curve is rather well defined.

The simple, low-impedance network to accomplish this correction will be discussed in connection with the flat-response amplifiers. For the present, it may be noted that the circuit of *Fig.* 3 will require less alpha cutoff correction than that of *Fig.* 2 because, as might be expected, the presence of negative feedback independent of frequency reduces the drop.

Frequency responses for these amplifiers have not been included for the simple reason that the conformity of an amplifier of this type to RIAA—or to any of the older curves for that matter—is largely either a question of patience for the individual experimenter or of tolerances to the commercial designer. It seems safe to say that the desired gain (32 to 35 db minimum) and frequency response can be obtained with transistors having common-emitter current gains of at least 30 or 35 although the preferred range would appear to be between 70 and 100. The choice between the two circuits is a matter of deciding whether or not the large emitter bypasses are too much to pay for the extra stability. In this connection it may be worth noting that the quantity S (stability factor) as now used is not particularly informative. In the first place, since a larger number denotes a more unstable circuit, the correct term would appear to be "instability factor." Of more importance, however, is the fact that, with transistors having current gains of 15 to 150 available for various purposes, the S-factor alone does not give a clear indication of circuit stability in terms of the improvement effected over the use of a unit having the same current gain in an unregulated circuit. For example, transistor A with a current gain of 100 is used in a circuit giving an S-factor of 20—an improvement ratio, obviously, of 5-1. Transistor B with a current gain of 30 is used in another circuit having constants resulting in an S-factor of 10 or an improvement of 3 to 1. But because the latter is a lower S-value, the erroneous conclusion can be made that B is the better circuit. Since the factor is intended to describe the stability of the circuit and not the stability of the transistor, it would be more informative to quote the ratio illustrated above and call it "regulation factor." In this notation unity describes, of course, an unregulated circuit and the higher numbers are the more stable circuits.

The two phono preamp circuits show

values as follows: (B is the common-emitter current gain)

Figure	Stage	B	S	B/S
2	1	100	2.9	34.5
2	2	70	7.3	9.6
3	1	50	2.6	19.2
3	2	50	14.2	3.5

This matter of regulation and designing circuits for a reasonable temperature range cannot be dismissed lightly. Too many designs have appeared in the recent past showing unregulated circuits although their designers must be aware of the fact—since no one has tried to keep it secret—that transistors are temperature responsive. In the present instance, a rise of 10 or 15 degrees Centigrade would produce, on an unregulated phono preamp, a frequency response only distantly related to RIAA or any of its forebears. With rising temperature (and rising collector current) transistor impedances will drop and if the degradation of frequency response is too great to accept, the designer must improve regulation to the point where the deviations are tolerable for the temperature range over which the equipment is intended to be usable.

Flat Preamplifier

Preamps to be used with dynamic microphones and having flat responses must generally be designed for a specific application with due regard to microphone impedance, line impedance, physical size and the input characteristics of the main amplifier. In the situation most prevalent at present, the latter will be a vacuum-tube amplifier with either a high-impedance (direct to grid) input or a line-to-grid transformer. To obtain the advantages of operating the line at a higher power level (i.e. higher than the level of the unaided microphone) and with a fairly wide choice as to impedance, it is necessary to build the transistor preamp either into the microphone case or in a small cylindrical housing which can be interposed between the cable plug and the microphone receptacle. This is practically the same, it will be noted, as saying that the transistor preamp may be regarded as an impedance transforming device which could conceivably replace the transformer now used for this purpose—the important difference being that gain may be obtained along with the transformation at little or no cost as far as signal-to-noise ratio is concerned.

For these purposes, the actual gain obtained in the input stage should be considered as secondary to obtaining the required impedance because it is always possible to add a second stage without an excessive increase to the size, weight and battery drain of the preamplifier.

For example, a 50-ohm microphone will require a common base stage and if this must work into a low-impedance line (say 250 or 300 ohms) because of frequency and line capacitance considerations the stage gain will be less than 10 db since the gain of the common base connection is derived from the ratio of the load to the source.

It is a mistake to suppose that a "cathode follower" (i.e. grounded collector) type of input stage can be used for these applications as a sort of universal input device capable of working from a few tens of ohms to a few tens of kilohms with the real gain derived from the following stages. Those who think along these lines exhibit only a slavish adherence to vacuum-tube circuitry in the face of the demonstrable fact that low (and approximately equal) noise factors are possible for all three common-electrode arrangements only when impedances are matched or at least approximately so. They would like, in short, to enjoy the enormous advantages of transistors without paying the necessary (and not unreasonable) price of learning to think of the transistor as a power device.

Impedance Adjustment

Fortunately for present purposes, it is possible—as already intimated—to cover a very wide range of input impedances by the selection of transistor alpha, circuit arrangement, and operating current. The selection, furthermore, may be made to show some overlap in the transition from one circuit to another although the region of the overlap will usually favor, for one reason or another, the common emitter. A few illustrative examples may serve to clarify this point, for the benefit of those who have not hitherto given much thought to this problem.

Let us assume that the designer, in addition to the choice of circuit, has available the following:

(1) Maximum operating current: 2 ma.
(2) Minimum operating current: 50 μa.
(3) Maximum transistor beta: 90 (types 2N65 and 2N132).
(4) Minimum transistor beta: 22 (types 2N63 and 2N130).

It will be understood that these maximum and minimum beta values are not absolute but merely representative of the types indicated. Also, the beta at 50 μa will be typically about 50 per cent of the 1-ma value and the following estimates are based upon this reduction.

The minimum possible input resistance (grounded base, short-circuit load) is:

$$R_i = R_e + R_b(1 - \alpha) = R_e + R_b \frac{1}{1 + \beta}$$

At 2 ma:

$$R_i = 12 + 400 \ (.011) = 16 \ \text{ohms}$$

For maximum with grounded base (current 50 μa, beta 11):

$$R_i = 500 + 1200 \ (.083) = 600 \ \text{ohms}$$

The input resistance of the grounded emitter is:

$$R_i = R_b + R_e \ (\beta) \ (1 + \beta)$$

For the minimum, the current is 2 ma and the beta 22:

$$R_i = 400 + 12 \ (23) = 676 \ \text{ohms}$$

For the maximum ($I_e = 50$ μa; $\beta = 45$):

$$R_i = 5000 + 500 \ (46) = 28,000 \ \text{ohms}$$

The final values have been rounded because the R_b quoted for each case is simply the most probable value and the actual R_b may differ since this parameter, for a large number of units, has its own distribution.

The important parameters, then, for the control of input resistance are alpha (or beta) and base resistance. For either the equipment designer who must be prepared to accept a reasonable parameter dispersion or the experimenter who must perforce take his chances as to the exact values of the few units he is willing to purchase, the common emitter circuit would appear to be preferable because:

(1) The minimum R_i for a given beta and emitter current (as, for example, the 676 ohms quoted above) can be still further reduced by the use of collector-to-base feedback.

(2) This type of feedback will reduce the range of input resistance caused by a given range of transistor current gain and base resistance.

(3) Negative feedback is extremely difficult to apply to the grounded-base circuit because of its characteristically low input resistance and the severe losses in gain—lower than the grounded emitter to begin with—caused by the shunting of the load. It will be understood that we are speaking not of the case where the designer considers it no particular hardship to use 22½- and 45-volt B batteries, but of the more practical case where the desired results are obtained with a minimum of supply power.

Variation of Operating Parameters

In elaboration of the foregoing, a digression at this point may be permissible. While it is true that the phono preamp already described uses a 22-volt supply (for the output stage only) it must be recognized that this was done for one specific purpose: that of providing approximately the same maximum output voltage as may be obtained from the vacuum-tube preamp without resorting to the use of an output transformer. As far as its performance otherwise is concerned, the transistor preamp can be built with a 4-cell, 3-cell, or possibly even a 2-cell battery.

The higher voltage, as already noted, is of course available from the power supply of the main amplifier. This is also true of many other transistorized units intended to be used with vacuum-tube equipment already on hand. We are in a period of what might be called "hybridization" in which the transistor is relegated to the role of performing only those functions where it is demonstrably preferable, for one reason or another, to the vacuum tube. In many of these cases design problems would actually be simplified if the entire equipment were transistorized. Since it appears likely that some equipments now line operated will take the form of battery portables by the full exploitation of transistor capabilities, it may be recommended that designer and experimenter alike begin to channel their thinking away from 45-volt B batteries and toward low-voltage operation.

The phrase "without resorting to the use of an output transformer" which appears above has a similar explanation. High-quality transformers, in addition
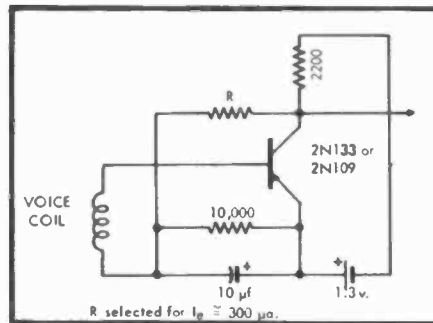

Fig. 4. Simple form of one-stage amplifier for use with dynamic microphone.

to being rather costly, are usually much larger and heavier than the rest of the components put together, including the transistors. Consequently, if the end result can be achieved without using transformers, it seems probable in the long range viewpoint that their use will decline in miniaturized applications. In fact, it may be recalled that part of the present discussion will be devoted to an examination of the feasibility of using transistors as impedance transforming devices.

To return to the main subject: Of relatively less importance is the fact that the upper limit on R_i for the common-emitter circuit can also be extended by the use of external emitter resistance feedback because these higher impedances (above 25,000 ohms) are also available with the common-collector circuit at normal operating currents and also because the emitter type of feedback gives virtually no improvement in restricting the range of input resistance. The grounded-collector input resistance is extremely sensitive to output loading and values ranging from something like 10,000 ohms to several hundred thousand

ohms may be obtained with loads not in excess of 20,000 or 30,000 ohms.

"Flat" Amplifier Circuits

For the flat amplifiers we will describe first a unit intended to work from a high-quality dynamic microphone into a tape recorder. Many low- and medium-priced recorders are designed to operate from crystal microphones and consequently are not equipped with line input transformers. If the owner of such a recorder wishes to use a microphone of better quality and also to be able, on occasion, to place the microphone at some distance from the recorder (very difficult with the high-impedance crystal microphone because of line capacitance) he will find a satisfactory solution to his problem in a transistor preamp at the microphone.

Let us assume the following factors:

Microphone impedance: 500 ohms.
Microphone power: 95 db below 1 watt/microbar.
Microphone response: down 3 db at 50 and 12,000 cps.
Amplifier input: direct to grid.
Microphone cable: 50 ft of single conductor at 70 μμf/ft.
Battery: one or two cells.

Figure 4 is typical of the simplest sort of one-stage amplifier that might be seriously considered for the purpose. With the omission of the output blocking capacitor (since this would be included with the tube-amplifier input) the circuit shows three resistors, one capacitor, one transistor, and one cell. The combination of load and collector current leaves 0.6 to 0.85 volts at the collector according to whether a mercury or a carbon cell is used. By connecting the base divider to the voice-coil return instead of directly to the base, the feedback normal to this type of regulation is made practically zero and the power gain of the stage is approximately 17 db. This does not sound very impressive (unless viewed in relation to the paucity of components) but it may be instructive to take a close look at just what we have before leaving it.

The power from the microphone, at 500 ohms, is given as −95 dbw per microbar. This is 3×10^{-10} watts for a sound pressure level of 74 db. The same microphone with a built-in transformer to raise the impedance to 25,000 ohms will deliver 5.5 millivolts to an open grid for the same pressure. The cable in this case, it should be noted, must have a capacitance no greater than approximately 525 μμf for response to 12,000 cps.

The original microphone power (3×10^{-10} watts) plus 17 db (the preamp gain) is 1.5×10^{-8} watts which is 5.75 millivolts in 2200 ohms. The maximum

tolerable capacitance across the latter is nearly 6000 μμf. The net practical comparison, then, is that we have the same voltage at the grid with this extremely simple preamp, working from the 500-ohm mike over a line more than 12 times as long, as we would have from the high-impedance mike.

With a transistor alpha of 0.975, the input resistance of this stage in the absence of feedback is approximately 3500 ohms and a mismatch of this order may yield a noise factor 1 or 2 db higher than the minimum value possible for a given transistor. To the perfectionist this may sound a trifle alarming but this particular microphone has an inherent signal-to-noise ratio of 68 db in a 74 db sound field and it is consequently of little practical importance in this particular case whether the noise factor of the transistor is 4 db or 7.

To a microphone which is down 3 db at 12,000 cps, it matters little whether or not the transistor is down 1.5 or 2 db at 20,000 cps because of frequency cut-off of alpha. There is accordingly no need for a compensating network and this is true, in general, for most preamps for dynamic microphones although this factor deserves more attention if for some special purpose considerable gain with several stages is built into the preamp.

For the second case involving a microphone preamp, let us assume that the owner of the tape recorder has a low impedance mike of 50 ohms and finds that when he tries to use it the recorder gain must be forced to the point where the quality of the resulting recordings is very dubious because of hum and microphonics—the latter being due to the tape reel motor. A similar situation can be encountered (although somewhat less annoying) if the mike is of medium (250 ohms) impedance.

We will therefore assume:

Microphone impedance: 50 ohms.
Microphone power: – 79.4 dbm/microbar.[2]
Microphone response: 50 to 10,000 cps.[2]
Amplifier input: direct to grid.
Preamp supply: one cell.
Microphone cable: two-conductor.

Although the length and capacitance of the cable are relatively unimportant because of the very low impedance, the two-conductor cable is usually supplied to balance out hum pickup. This in turn calls for a balanced input at the amplifier and if the necessary input transformer has not been provided the only makeshift solution is to operate with an unbalanced line and reduced cable length. This difficulty disappears with the transistor preamp because the hum

[2] Manufacturer's data on Model 556S (Shure Bros.) Cardioid.

picked up by an unbalanced line of almost any reasonable length will be, in effect, down by an amount equal to the preamp gain.

The input impedance of 50 ohms indicates a common-base input stage with a transistor having rather high current gain (100) so that the $R_b(1-\alpha)$ portion of the input resistance may be held to 10 or 15 ohms thereby allowing an emitter resistance of 35 or 40 ohms for a current of 600 or 700 μa. *Figure 5* shows a two-stage amplifier with a power gain of approximately 28 db and an output impedance of 2200 ohms. The corresponding output voltage for an input power of 1.15×10^{-11} watts (-79.4 dbm) is 4 millivolts or -48 dbv. This compares to -60.5 dbv (open circuit) for the same type of microphone at 35,000 ohms.

For any microphone supplied with a

R1 : 39,000 to 68,000   ($I_e \approxeq 600$ μa.)
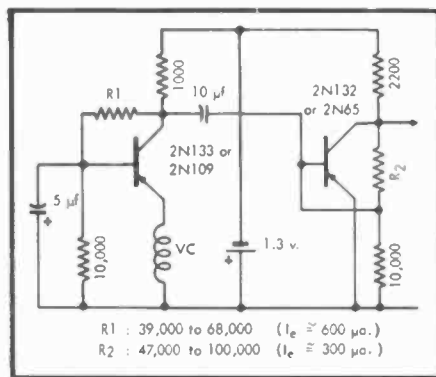R2 : 47,000 to 100,000   ($I_e \approxeq 300$ μa.)

Fig. 5. Two-stage microphone amplifier with a power gain of 28 db and the comparatively low output impedance of 2200 ohms.

two-conductor cable the preamp designer is under no particular compulsion to package the cells with the preamp because the change to the unbalanced line makes the second conductor available for supplying power. Alternatively, if the designer elects to change to a single-conductor cable with the battery in or near the preamp, the switch usually supplied for grounding a single-impedance mike can be made to serve as an ON/OFF switch for the battery. If the preamp is built into a cylindrical housing plugged into the mike receptacle, the size of such a package will of course be considerably smaller than that of the housing now commonly used for the cable transformer. The diameter, in fact, need be no longer than whatever is necessary to accomodate the connectors; with a length of two or three inches according to the number and type of cells and the mechanical arrangements for their replacement.

For the benefit of those who are unduly impressed by statements which still appear on some rating sheets for junction transistors to the effect that:

(1) The transistor should not be in-

serted into the socket with the power on or (2) Switching transients should be avoided; (3) Capacitor discharge surges should be avoided; (4) The socket should be designed so that, upon insertion of the transistor, the collector makes contact last—it may be noted that statements of this kind represent little more than survivals from similar statements appearing on ratings for contact types. With currently available junction types having avalanche voltages usually well in excess of 50 volts, switching tests conducted at voltages well below this value must be run into the hundreds of thousands before significant changes in major parameters may be ascribed to switching alone.

To return to *Fig.* 5: the capacitor from base to ground of the input stage may be described somewhat beyond the mere statement of its value. At first sight, it might be supposed that this value should be several hundred microfarads to avoid attenuation of low-frequency response in view of the 50-ohm input. Since such a capacitor would be by far the largest component in the preamp, it is fortunate for purposes of miniaturization that such is not the case. Although called, by precedent, a "bypass" condenser (since it bypasses the lower bias resistor) it shows up in an equivalent circuit as an impedance in series with R_b and, as such, in the common base circuit its impedance is reduced by the same factor $(1-\alpha)$ as R_b. Consequently, the effectiveness of this capacitor in relation to low-frequency cutoff is increased as if its value were multiplied by $1/(1-\alpha)$ or $(\beta+1)$.

For the marginal case where only a small amount of gain is required, the first stage alone will give a power gain of approximately 11 or 11.5 db with a corresponding voltage gain (50 ohms to 1000 ohms) of 18 or 18.5 db. With a sound pressure level of 74 db at the microphone, the preamp output in this case is -78 dbv. This gain for this stage is, of course because common-base gain must be obtained from $R_L/4R_g$ and the load has been reduced to 1000 ohms to provide the 50-ohm input with one cell. The addition of the second stage as shown results in an over-all power gain of approximately 28 db—the output voltage varying according to the value chosen for the collector load and line impedance.

Design of Microphone Preamp

Let us now assume that the microphone on hand is of the "high impedance" class. As used with dynamic microphones, this can mean anything from 20,000 to 40,000 ohms, although the value most often encountered is 25,000. We will therefore assume:

Microphone impedance: 25,000 ohms.

Microphone voltage (open circuit): –60 dbv.

Microphone power (into 25,000 ohms): 10^{-11} watts/microbar.

Frequency response: 50 to 10,000 cps.

Amplifier input: direct to grid.

Microphone cable: single-conductor.

Figure 6 is representative of an amplifier adjusted to an input impedance of 25,000 ohms. Because the resistance in this case depends predominantly upon the product of β and R_e, it is necessary to obtain the major portion (say, 20,000 ohms) by operating at a rather low emitter current. Alternatively (or supplementarily) the product can be raised to the required level by increasing R_e with an external resistor. However, whether the 25,000 ohms is obtained by reducing the operating current, by increasing the external emitter resistance or by a combination of the two, the gain will be about the same for a given beta in the first stage. The actual gain realized is not of overwhelming importance because another stage will be required in any case because with a high source impedance and a low load impedance the voltage gain will be considerably less than the power gain.

For most general-purpose a.f. transistors, maximum beta occurs at an emitter current of 1.0 or 1.5 milliamperes and, although the peak is very broad on a linear current scale, the value will be down about 35 per cent at 100 μa and 50 per cent at 50 μa. As in the case of the phono preamp, in lieu of measuring R_e, R_b and β at several operating currents and computing the resulting impedance, a simpler and more direct method is the familiar one of supplying a known current signal through a resistor much larger than the expected R_i and measuring the ensuing voltage at the input. Since a 2 to 1 mismatch loses only 0.51 db in power gain and the noise factor will not worsen appreciably until the mismatch becomes gross, any value between 15,000 and 40,000 should be acceptable.
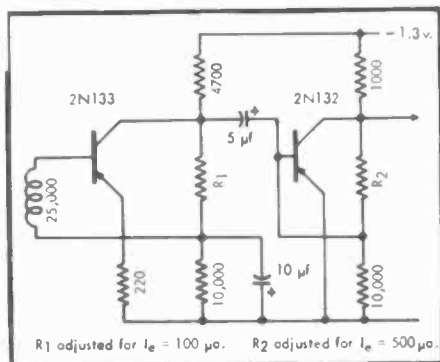
Fig. 6. Transistor microphone amplifier with an input impedance of 25,000 ohms.

With a type 2N133 having a current gain of 50 to 100 μa in the first stage and a 2N132 showing a gain of 70 at 500 μa in the second stage, the power gain is approximately 40 db, assuming average base resistance. The voltage gain (25,000 ohms to 1000 ohms) which is of greater immediate interest if the output must work into a transformerless tube input is less according to:

$$VG(\text{db}) = PG(\text{db}) + 10 \log R_L/4R_g = 26 \text{ db}$$

If some other output impedance is selected to meet some particular requirement, the second-stage current should be readjusted if possible to drop about 0.5 volts across the collector resistor. Similarly, if the first-stage current must be changed from the indicated value to meet the input impedance requirement, the collector resistor of that stage should be changed. The object in both cases is to obtain the highest possible regulation factor from the single cell without operating at a collector voltage too close to the knee of the collector characteristic.
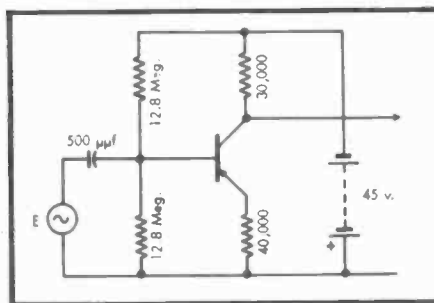
Fig. 7. Common-emitter input stage with a very high input impedance.

High-Impedance Sources

Most difficult of all is the situation where the amplifier is required to work from a high-impedance capacitive source. A good example of a particularly difficult case is that of the Western Electric 640-AA condenser microphone. This unit, with a capacitance of 50 μμf, is commonly worked into a cathode follower of at least 200 megohms for sound-pressure measurements of 20 cps with little or no correction and if used for ordinary a.f. purposes to –3 db at 80 cps must still see 40 megohms. Since these impedances are not available with germanium transistors, we may at least examine briefly the much more common case of the 500 μμf pickup or microphone.

The common-base connection is, of course, completely useless and for either the common collector or the degenerated common emitter a major cause of the difficulty is the fact that the bass potential must be fixed by the divider which cannot be placed in the low side of the source as in some of the circuits
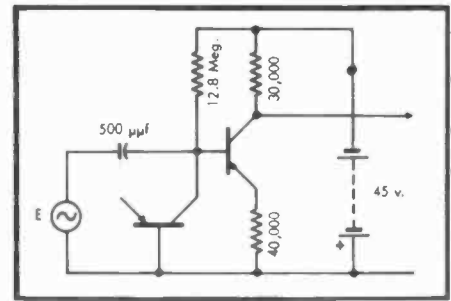
Fig. 8. Circuit of Fig. 7 modified by use of diode as lower base resistor to provide temperature compensation.

already discussed because there is no d.c. current path. In the equivalent circuit, the divider elements are in parallel with each other as well as in parallel with the input.

Although in this instance the use of an input transformer appears to be the obvious and low-cost solution, there still remains the problem of primary resonance and the necessary correction of the frequency response in subsequent stages of the amplifier. Here again the transformer may be rather large if the resonant point is made at least an octave below the lowest frequency of interest.

There appear to be only two possibilities of obtaining the required impedance (3.2 megohms for –3 db at 100 cps) without the use of a transformer. Neither is particularly encouraging—they are included in this account only to illustrate the difficulties involved.

Figure 7 is a common-emitter input stage with a rather large emitter resistor and a 45-volt supply. The input impedance of the transistor is approximately βR_e or 6.4 megohms for $\beta = 160$ and the same impedance would be available with a common collector by taking the output from the emitter and bypassing the collector resistor. There is a practical difference in favor of the common emitter, however, because here the collector load may operate into almost any reasonable value for the following-stage input while the common collector must work into something considerably greater than 40,000 ohms to avoid excessive shunting of the emitter resistor upon which the input impedance depends.

The latter is now 3.2 megohms but the cost is almost prohibitive. The beta of 160 (at approximately 55 μa) would certainly represent some selection but even more serious is the very poor regulation factor under conditions of collector voltage and current such that a very good regulation factor is strongly indicated. The entire circuit, in fact, is scarcely better than if we had used a betta of 80 in the first place and omitted the base divider entirely.

Figure 8 is similar except that the lower base resistor has been replaced by a diode. If this diode has an impedance

of the order of 12 megohms and a back current, at approximately 22 volts, equal (or nearly so) to the cutoff current of the transistor the circuit will be temperature compensated—provided also that the temperature coefficients of the two currents are similar. The collector diode of another transistor would appear to be the only device capable of meeting these requirements and while the purist is likely to be disdainful of the notion of using a transistor for such a simple function —with an unused electrode at that—it nevertheless remains a possible solution to the general problem. The matching of cutoff currents would certainly be tiresome and vexatious—virtually impossible to the experimenter with a limited number of units at his disposal although, at the equipment design level, it is not improbable that the stage could be designed to an acceptable tolerance on input impedance *vs.* temperature without resorting to an excessively large number of current brackets.

Figure 9 is a two stage amplifier with a power gain of 49 db from a 500-ohm source to a 5600-ohm load and a frequency response flat within 0.5 db from 20 to 20,000 cps. Although constructed with a 5-volt supply (four mercury cells) at approximately 800 μa, a similar design operating with one or two cells at somewhat lower gain is not particularly difficult. As in the case of the phono preamp the large emitter bypasses may be eliminated by changing to the type of regulation in which the base divider feeds from the collector at some sacrifice in stability and gain.

Compensation Methods

The network for high-frequency correction (to compensate for alpha cutoff in both stages) is in the second stage emitter in the form of an extra emitter resistor with a small bypass capacitor. Although there are severe practical limitations to this type of network as far as frequency range is concerned, it is quite adequate to the present purpose of providing flat response to 20,000 cps. The actual values will, of course, vary according to highest frequency of interest, transistor and circuit parameters, alpha cutoff, and so on, and are consequently more difficult to compute than to deduce experimentally. At the risk of appearing obvious, we may outline the following routine:

(1) Run a response curve of the uncorrected amplifier out to 1.25 or 1.5 f_H (f_H is the high end of the desired band).

(2) Insert an emitter resistor (without bypass) of such a value that the 1000-cps gain of the amplifier is now less by an amount 2 to 3 db greater than the loss at 1.25 f_H in the uncorrected curve.
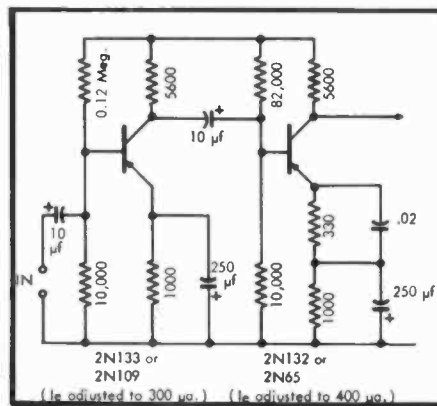

Fig. 9. Two-stage microphone amplifier with a power gain of 49 db between a 500-ohm source and a 5600-ohm load.

(3) Add the bypass, computed to have the same impedance at 1.25 f_H as the resistor.

In the final curve, slight second-order curvatures will be discernible with magnitude depending upon the required bandwidth and the alpha cutoff of the particular transistors used. This is associated with the usage limitation mentioned previously and is due to the fact that, while alpha cutoff has essentially the same characteristic as an *RC* network, merging rapidly into a 20 db/decade slope, the frequency characteristic of a stage using a selective emitter network is a sigmoid with asymptotic limits at both ends. *Figure* 10 is a plot of stage gain as a function of external emitter resistance for a stage of reasonably average characteristics and loading. In

Fig. 11, the corresponding attenuation from maximum gain has been plotted against frequency for a network having a crossover point of 25,000 cps. Since the maximum slope at any part of this curve is only 8 db per decade, a casual inspection might lead one to suppose that the device is not particularly suitable but (1) more slope may be obtained by increasing the value of the emitter resistor and (2) the actual corner of the alpha/frequency characteristic is somewhat more rounded than its *RC* analogy because of dispersion of charge carrier transit time, non-planar geometry, and so on. This effect mitigates the severity of the original problem of correcting the response to 20,000 cps although, for a band appreciably wider than this, it would be preferable to use r.f. transistor types.

For considerable amounts of treble boost, it may be more practicable to use an emitter network in more than one stage in order to avoid attenuating more than 15 db in any single stage. For the completely transistorized tape recorder amplifier in which both bass boost and treble boost may be required, a considerable excess of mid-frequency gain will be necessary and, if the over-all gain must be fairly high, a promising stage line-up would be:

1st stage: Wide open to secure good noise factor and controlled input impedance.

2nd stage: Collector to base network for low-frequency boost.

3rd and 4th stages: Emitter networks for treble boost.

In the event that the low-level stages must be restricted to three, it may be feasible to incorporate the treble boosting into the first and third stages—leaving the second free to handle the full swing of the low boost. The deciding factors are the relative amounts of mid-
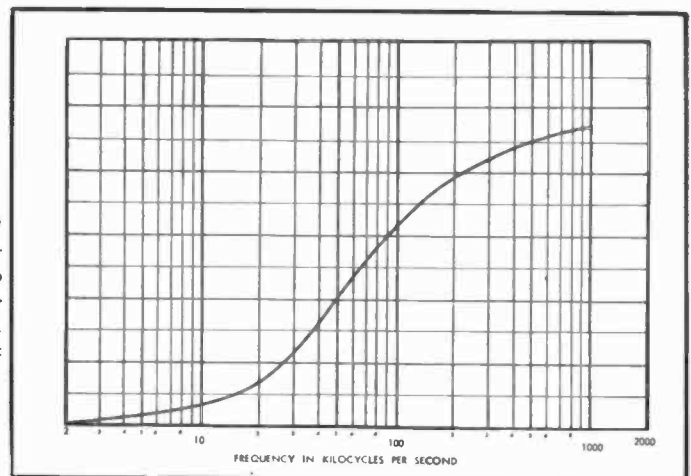
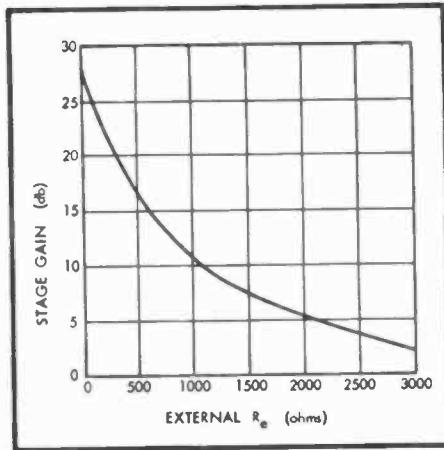Fig. 11. Response of corrective emitter network to compensate for normal high-frequency droop of transistor.

Fig. 10. Plot of stage gain vs. external R_e for a typical RC-coupled stage.

frequency gain and end-frequency boosting required.

Conclusion

By way of conclusion, we may be permitted to make some general remarks—particularly with respect to the noise factor.

The hallmark of a good transistor is its ability to withstand moderate increases of collector voltage and/or temperature without showing a substantial increase in the noise factor. These are also the units having:

(1) 1000-cps noise factor below 10 db.

(2) A noise power spectrum better than that of the average vacuum tube having an oxide coated filament or cathode.

(3) The same (or slightly better) integrated noise factor over a band of 16 to 20,000 cps as the spot noise factor at 1000 cps.

(4) Wider tolerance to input mismatching without a significant increase in the noise factor.

However, for the purposes described in this account, the potential transistor user should not place an abnormal emphasis upon the desirability of securing units having factors of 4 or 5 db as against those having factors of 7 or 8 db. It is a demonstrable fact that a subjective listening test conducted to provide a direct comparison between systems differing in noise factor by as much as 3 db must be set up with virtually no time lag between the two—as by direct switching—in order to hear the difference and if the lag is as great as one-half minute, it is extremely difficult, if not impossible, to distinguish between them. ●

RECORDING CHARACTERISTIC

standard values were used in the actual unit. Five per cent components were used throughout, and tests on the various arrangements indicated the network behavior to be within ±1 db of the theoretical recording characteristics.

Construction

Mechanical construction of the unit is quite straightforward. The entire wiring is constructed on a two-circuit, six-position switch, and mounted in an aluminum box $2\frac{1}{4} \times 2\frac{1}{4} \times 4$ in. The input terminals are a set of banana plugs, spaced at $\frac{3}{4}$ in. so as to plug directly into a signal

Fig. 7. Internal construction of the unit, showing how all components are mounted on switch.

generator. The output is taken from a pair of banana jacks spaced at the same distance. *Figure* 7 shows an inside view of the unit. ●

TRANSISTOR ACTION

have gathered the following information: (1) The hook transistor uses hook multiplication to multiply the collector current of an otherwise-normal transistor. In this manner an over-all current gain greater than 1 results. (2) In time a hook photocell may be developed which affords greater amplification of light-generated current than even the photo-transistor.

Summary

If the reader has been able to follow the arguments leading to an explanation of transistor action, and has followed the explanation of the various devices, he now has an intuitive feeling for the physical behavior of junction devices, which will help in designing

circuitry to use these junction devices in. Described in this article were: Germanium photo-resistor, Junction diode, Junction photo-diode, Junction transistor, Junction photo-transistor, Junction hook transistor, and Hook photocell.

Various semiconductor devices were not described at all, both because of the lack of space, and because in some of these devices the exact theory of operation is not very well known. Not described at all include the following: Point-contact diode, Point-contact photocell, Point-contact transistor, Co-axial transistor, Point-junction transistor (in which one element is a point-contact and the other is a junction), Surface-barrier transistor, Field-effect transistor, Semiconductor relay, Intrinsic region junction transistor, Double-base diode, Junction tetrode, Photo-voltaic cell, Fieldistor, Symmetrical transistor, Zener reference diode,

Thermistor, Photo-conductive cell, or Analog transistor.

If the reader wants to do further reading in the very interesting field of transistor action and the physical foundation of semiconductor devices, he is referred to any one of the many fine books on transistors now available, or to the three references given here, which in the author's eyes cover the field quite well. The first is now a classic explanation of transistor action.

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Transistor Tips and Techniques

PAUL PENFIELD Jr.

A short introduction to the mysterious world of transistors, for audio fans and engineers who haven't yet done any experimenting in the field. Practical tips are given, to supplement theoretical material already available.

O F ALL THE RECENT INVENTIONS in the electronics and audio fields, the transistor will undoubtedly have the greatest effect. Such great promise is held for these little "miracles of matter" that their use is expected to become widespread before long.

Most professional engineers have by this date had experience of one kind or another with transistors, possibly in the audio field. However, the chances are that hobbyists, experimenters, serious audiofans, and even some engineers have yet to try their hand at transistor audio applications. For these people, this article is written and dedicated, to make their entrance into this strange new field as painless as possible.

One difference is immediately apparent between transistors and vacuum tubes: their small size. Small-sized tools sometimes are required for constructing transistor circuits. A pair of tweezers is a handy tool, and long slim-nosed pliers are easier to work with than conventional chain-nosed pliers. However, ordinary soldering irons, contrary to popular belief, usually do as good a job of soldering transistors as the lower-power soldering pencils. Making a quick, hot joint is preferred to making a slow, not-so-hot joint, which may burn out the transistor. The amount of "re-tooling" necessary for experimenters to start work in transistors is quite small. The budget-minded experimenter will do well to get along on his present tool supply.

Transistor Components

Audio engineers and audiofans should have no trouble adapting themselves to transistors. Since use of transistors will be widespread in the near future, serious enthusiasts have no choice but to get some experience in with the units. If this article will help to bridge the gap into the unknown for some of the readers, it will serve its purpose. Experimentation with transistors is, incidentally, much easier than experimenting with vacuum tubes. No expensive, bulky high-voltage power supplies are required, and there are no filaments to heat. All of the circuitry directly contributes toward the function of amplification.

No radically new and different types of components are required for use with transistor circuits. Resistors, capacitors, transformers, batteries, plugs, sockets, coils, meters, fuses, etc., perform the same functions in transistor circuits as they do in vacuum tube circuits. However, because of the low voltages and impedances present in transistor circuits, different values for these components will be appropriate.

Coupling capacitors must normally be quite high in value—up to and even beyond 50 μf at times. Fortunately the voltage requirements are low, and electrolytic capacitors are available with these ratings. Tantalum capacitors (see *Fig.* 1), although appropriate in value, are quite expensive. Conventional low-voltage units, however, are quite reasonable.

Transistor sockets are available—in most cases the standard subminiature in-line tube sockets, such as the Cinch-Jones type 2H5, will serve well. (See *Fig.* 1) "Transistor sockets" are available commercially, but cost a trifle more.

Transistor circuit power supplies will generally be batteries. Conventional batteries do quite well, and can be selected for the purpose on the basis of voltage and expected current load, the same way batteries for any other purpose are selected. Hearing aid batteries or mercury cells may be useful. Some companies have "Transistor batteries" which are nothing inherently different from other batteries. For experimental purposes, on the bench, storage batteries do as well as any. Line-operated power supplies are generally complex, because of the amount of filtering necessary, but, if properly designed for the correct voltages and current ratings, do as well as batteries. Use of ordinary laboratory-type power supplies should be avoided, however, because of the high no-load voltage, which may easily cause transistor break-down.
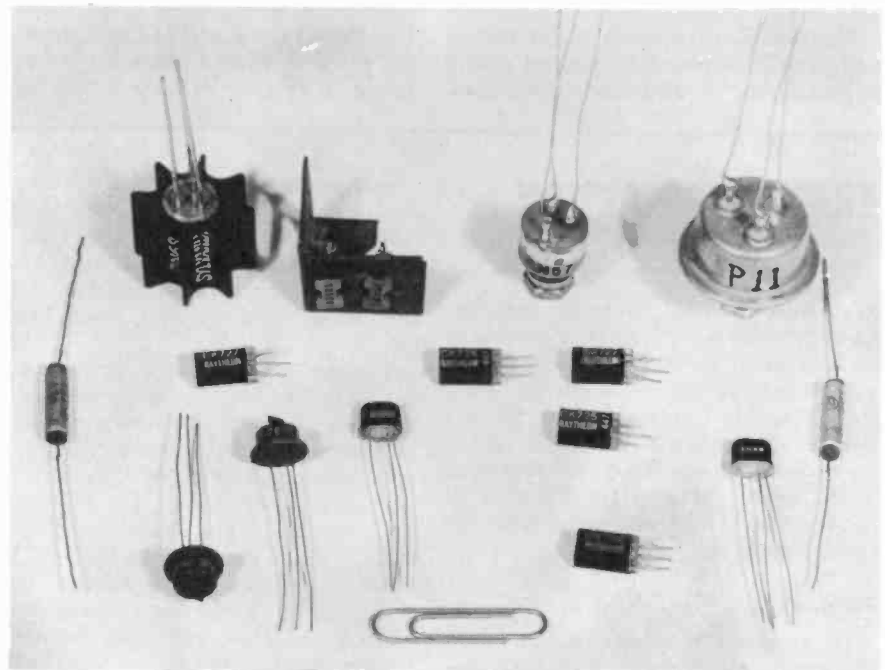

Fig. 1. Various transistors and components. The five Raytheon transistors have been adapted to fit into subminiature tube sockets, two of which are shown mounted in a universal power transformer mounting bracket, upper left. Two tantalum electrolytic capacitors are at extreme right and left positions. The rear row shows power transistors, the foreground common low-power transistors.
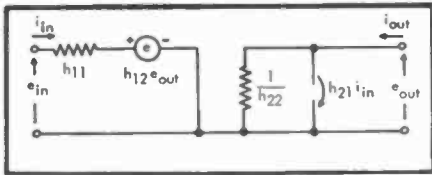
Fig. 2. Most commonly-used small-signal equivalent circuit for low-power transistor stages.

Transistor transformers have appeared on the market; however these transformers generally sacrifice quality for small size, and so where fidelity is important should not be used. Unfortunately at this writing a line of high-fidelity transistor transformers is not available. In most cases, luckily, the use of transformers can be avoided altogether. When this is not possible, obtaining a suitable transformer can be a real problem.

High-impedance microphones, such as crystal mikes, are less suitable for transistor work than low-impedance models, such as dynamic types, inasmuch as the latter may generally be coupled directly to a transistor amplifier, without the use of a step-down transformer.

Other components, such as loudspeakers, meters, resistors, fuses, potentiometers, switches, and terminals are either normal types, or else miniaturized versions of their vacuum-tube counterparts. No special difficulty should be encountered by use of standard parts.

Simple Audio Amplifiers

Well, we've now got the transistors, and the necessary transistor tools and components, so the question becomes, "what can we do with them?" The design of simple audio amplifiers is not difficult.

It is customary for design engineers, when making transistor audio amplifiers, to use au equivalent circuit of the type shown in *Fig.* 2, which is much like the vacuum-tube small-signal equivalent circuit shown in *Fig.* 3. Using the transistor equivalent circuit, however, is a bit more complicated than using the vacuum tube circuit, and can be avoided for a while, until the reader gets the "feel" of transistors, at which point it can be taken up with the least pain. The industry is now standardizing on the so-called "h-parameters"[1], as used in the equivalent circuit of *Fig* 2, and with which the serious audiofan is well advised to familiarize himself. However, for the first plunge into transistors, the load-line technique, so useful in vacuum tube work, will prove adequate.

Graphical analysis is necessary in any event to determine the bias at which the transistor operates, but it can tell more as well. For example refer to *Fig.* 4,

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[1] Refer to Shea: *Transistor Audio Amplifiers*. Wiley, New York, 1955.

which is a typical pentode plate characteristic. The load line XY is drawn, determined by the supply voltage X and the load resistor. The bias, or the "quiescent point" is determined by the intersection of this load line with the grid bias line chosen, of course. As the grid voltage varies about its quiescent value, the point of operation moves between points R and S, as shown. Here, of course, the a.c. load line is the same as the d.c. load line, as is usually the case for vacuum-tube voltage amplifiers.

Now, however, look at *Fig.* 5, which is a typical audio transistor grounded-emitter characteristic. Notice the similarity in shape, except that the determining parameter of the family is now base current, instead of grid voltage. Shown is a d.c. load line XY, determined again by the supply voltage X and the load resistor. This line, together with the value of quiescent base current determines the quiescent point Q. It could be argued from analogy that a small change in base current would produce a change in the collector voltage in exactly the same way, following the line XY, but this is not so. In most practical transistor amplifiers, the a.c. load line and the d.c. load line are not the same.
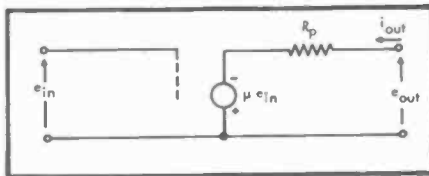

Fig. 3. Conventional grounded-cathode vacuum tube small-signal equivalent circuit, shown for comparison.

The load presented by the succeeding stage is not negligible, as is usually the case in vacuum tube circuits of this type.

An a.c. load line, corresponding to the a.c. resistance that the transistor "sees" at its terminals, must be drawn, looking something like the line RS in *Fig.* 5. Now a given base current change produces a smaller voltage change at the output than before we took account of the different a.c. load.

The quiescent point may, of course, be chosen anywhere within the ratings of the transistor; however good engineering practice usually dictates a choice such that on high signals, the a.c. load line will go into saturation S and cut-off R simultaneously—that is, RQ = QS. This point can be determined by observing the graph.

It may be handy to know such things as voltage gain, current gain, input and output impedances, power gain, etc., at this point. Indeed it is essential to know these quantities. The graphical analysis here can give only the current gain. The rest can be had only by investigating the equivalent circuit, and cannot be told from the load-line graph.

At this point the author will beg off the subject of amplifier design, because of the splendid array of material already available on the subject. The bibliography at the end of this article lists several books dealing in part or in whole with transistor audio applications. Instead, the author will present material not found in these theoretical books, but still highly important to anyone working with transistors for the first time. The more practical aspects of amplifier design and of handling transistors, are of more than passing interest to the experimenter who wants to keep from burning out his transistors.

Transistor Protection

Much has been made of the high life expectancy of transistors, implying that they can be permanently wired into circuits just like other "reliable" components, such as resistors and condensers. It is true that transistors are much longer-lived than vacuum tubes, and even other components, but this alone is not justification for permanent wiring, especially in breadb ard and experimental designs. There are powerful reasons why transistors should be modified to plug into sockets.

First, heat from a soldering iron can permanently damage a transistor in no time flat. Even experts sometimes overheat connections, and this practice is fatal to transistors, especially if the leads were cut short. Also, the leads are quite flexible, and when bent usually bend and break right next to the body of the transistor, leaving you with an electrically good, but still unuseable transistor, and a very frustrated feeling. Of course these remarks apply more to experimental and breadboard circuits— in assembly-line manufacturing operations the transistor should be welded in place by its leads.

Using a standard subminiature tube socket, or other transistor socket, precludes the disappointment of damaging
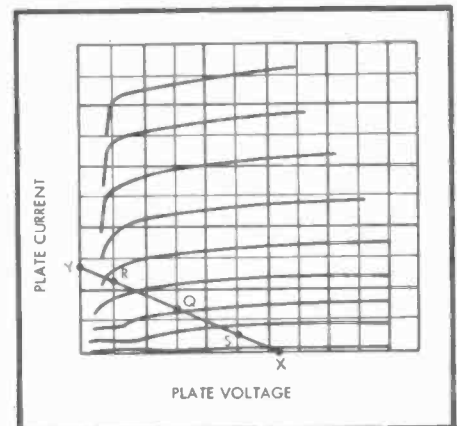

Fig. 4. Load line on a pentode plate characteristic family. Note that the a.c. load line and the d.c. load line coincide.

a transistor during construction. Transistor leads should be cut off to a length of about a quarter of an inch, with the base lead left slightly longer than the other two.[2] This was done with five of the transistors pictured in *Fig.* 1. One end of the socket should be painted red, or some other color, to distinguish it from the other end. This end can be used for the collector lead, to prevent plugging in the transistor backwards. Most transistors either have a distinguishing dot next to the collector lead, or else have a larger separation between collector and base leads.

Mounting the transistor socket can be somewhat of a puzzle in building transistorized equipment. A retaining ring is supplied with the sockets; however the mounting requires an oblong hole. Fortunately, several types of universal mounting brackets (such as intended for power transformers, loudspeakers, etc.,) have elongated holes of the correct size, and these may often be used as an inexpensive, convenient socket mounting. The socket may be held in place either with the retaining ring, or with a touch of plastic cement. *Figure* 1 shows two sockets mounted in a power transformer universal mounting bracket.

Several common vacuum tube practices must be forgotten when working with ·transistors—for example, battery polarity. With p-n-p junction transistors (the most common type at present), the collector lead must be supplied from the negative lead of the battery—not the positive. Forgetting this polarity and connecting a transistor backwards is asking for trouble. Sometimes it will not harm the transistor (especially if large current-limiting resistors are used in the collector lead), but more often the transistor goes off to transistor heaven with a sometimes visible puff of smoke. N-p-n transistors are connected just the reverse from this, with the collector lead positive.

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[2] This is to prevent accidently connecting the emitter and collector, but leaving the base free, which connection might be dangerous to other components.
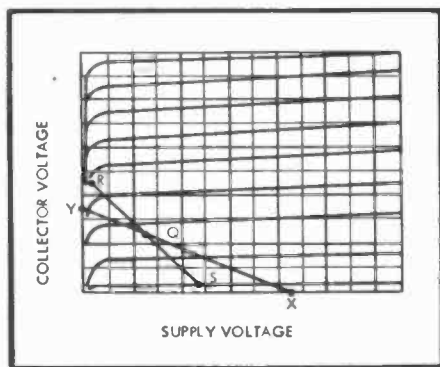

Fig. 5. Load lines on a grounded-emitter transistor collector family. Note that the a.c. load line and the d.c. load line do not, in general, coincide.

tive. It is always wise, especially if working with both types, to check each circuit each time the power is applied. Special care should be taken when both p-n-p and n-p-n units are in the same circuit, as for instance in complementary circuits. People familiar with vacuum tube practice may be a little careless, for in vacuum tube operation, reversing the B-plus does no harm except keep the circuit from operating. But in transistor work, the price paid for carelessness is higher.

Similarly, care should be taken with all polarized components—sometimes a little thought is needed to remember just what the correct polarity is. Some typical polarized components likely to give trouble : milliameters, electrolytic capacitors, diodes, voltmeters.

When working on a breadboard, it is wise to protect the transistor continually with a fuse. A rating of about three-quarters of the maximum recommended collector current is best. Slow-blow fuses, sometimes used to protect motors and other equipment which pass transients, are definitely not to be used in
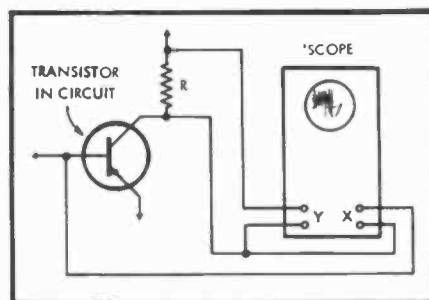

Fig. 6. Method for continually monitoring the load line of a transistor stage on an oscilloscope.

protecting transistors. The fastest-blowing fuse is the best, for by the time a slow-blow fuse gets around to blowing, the transistor is already shot. And don't be tempted to use a higher rating if you continue to blow fuses. Fuses are at present a bit cheaper than transistors, although there is every indication that this will not always be so.

One arrangement that is sometimes useful with low-power transistors is shown in *Fig.* 6. Continual monitoring of the load line (collector current *vs.* collector-to-base voltage) is done by displaying the quantities directly on an oscilloscope. The voltage across a known sampling resistor R is fed to the vertical amplifier, and the collector-to-base voltage is fed to the horizontal amplifier. The value·of the sampling resistor must be large enough to produce a reasonable-sized pattern, but not large enough to interfere with amplifier operation. By considering the value of this resistor, and by using a voltage calibrator, the pattern on the face of the scope can be calibrated to read directly in volts and
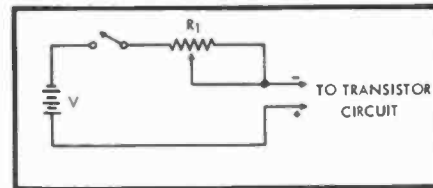

Fig. 7. Method to prevent transients from harming transistors in experimental set-ups.

milliamperes. Once that is done, the instantaneous power dissipation (which must never exceed the manufacturer's ratings) is merely the product of the two at any point on the trace, expressed in milliwatts.

This monitoring can also be used to detect large transients in the circuit. When the power is turned on or off, large transient currents may be drawn by the large capacitors. Transients as such do no damage, unless they exceed the manufacturer's ratings, which is often the case.

One method of eliminating trouble caused by transients altogether is to employ the circuit shown in *Fig.* 7. Here the battery is switched on while a large resistor R_1 is in series with it. Then this resistance, in the form of a rheostat, is removed with the power on. A safe value for this resistance is

$$R_1 = \frac{V}{I_{max}}$$

where V is the battery voltage, and I_{max} is the maximum recommended collector current for one transistor.

Of course, after the audiofan has had transistor experience, he can forget some of the safety precautions outlined here. But one unfamiliar with transistor characteristics will do well to protect his transistors, lest he have to dig into the sugar bowl for the wherewithal to replace the transistors he needn't have burned out.

Transistor Testing

Commercial transistor testers, which check many of the transistor parameters, cost so much as to preclude their use by audio enthusiasts at this date. However, a simple test given here will usually determine if a transistor is completely shot, although borderline cases, or partially burned out transistors cannot be tested reliably.

The test is quite simple; it merely consists of testing each junction separately, to see if each is a good rectifier, and then checking for transistor action. The only tool needed is an inexpensive multimeter, or a VTVM with a resistance range. For the majority of low-power transistors available now, the forward resistance of each junction should read roughly 1000 ohms or less, depending on the type instrument used, and the range it is set in. The backward re-

sistance should read more than 50K for both junctions. Now immediately upon checking the reverse resistance of the collector junction, move the test lead from the base terminal to the emitter terminal. A lower resistance should register, indicating more current, by a factor of perhaps ten or more. This may at first seem rather startling, but a little theory and a moment's thought should convince the reader that this unexpected result is merely one manifestation of transistor action. If this increase in current does not occur, the transistor is not displaying transistor action, and is thus no good.

The values of resistance to be expected in this test cannot be given accurately, because they vary from one transistor type to another, and also are quite non-constant—that is, will read differently on different ranges of the same instrument. However, by comparison with new transistors, the reader can work out a simple test procedure for himself.

Testing Transistor Amplifiers

Normal testing procedures apply for transistor audio amplifiers as apply for their vacuum tube counterparts. Frequency response, transient response, distortion, noise, hum, etc., all can be checked in the normal way. However, a couple of points are tricky and therefore worth noting here about using normal test equipment:

First, care should be taken to be sure that any signal generator used does not have d.c. superimposed on the a.c. signal output. This of course will disturb the transistor biases. Also, some generators have a low-resistance path between the output terminals, which again can foul up the input-stage biases. Using a 50 or 100 μf blocking capacitor is recommended—watch polarity here.

Typical low-level transistor amplifiers should have a frequency response flat over the entire audio range from 20 to 20,000 cps. cycles. High-power transistors generally lack in high-frequency response, and poor low-frequency response may be due to too low a value coupling and emitter-bypass capacitor.

Simple amplifiers should not have more than about 1 per cent total harmonic distortion, except for the output stage, which may be quite a bit worse, especially if it is operated Class B. Feedback can be used, of course, to reduce the distortion in a final design, as in vacuum tube amplifiers.

This article has tried to give some practical suggestions and tips for making an entrance into the world of transistor circuitry. These practical tips and techniques should be supplemented by good solid theoretical material. If the audiofan is going to be anything but just a "tinker" he must have some idea of the physical principles behind tran-

sistor action, and the methods engineers use to attack problems of transistor circuit design. Fortunately, books are available which adequately tell the story for every level of technical background, from the experienced engineer to the novice. All the English-language books available at this writing are noted here.

R. F. Shea, *Transistor Audio Amplifiers*. Wiley, New York, 1955. The only book devoted exclusively to transistor audio applications, this book is at present the "bible" of the field. Aimed at college students and practicing engineers alike, this book covers most facets of amplifier design with succinct thoroughness. The methods used are explained in detail to enable the reader to apply these methods on future transistors, which may be quite different from present-day models. The sections on power transistors are of necessity general, because the development of power transistors is so new. This book will for some time deserve a place on every transistorized audiophile's bookshelf.

R. F. Shea, *Principles of Transistor Circuits*. Wiley, New York, 1953. The same writer's previous book covers all sorts of transistor circuitry in such a manner that the material will probably not be out of date for some time to come. This book is a good introduction to transistors for practicing engineers, but not for the beginner.

A. Coblenz and H. L. Owens, *Transistors: Theory and Applications*, McGraw-Hill, New York, 1955. This book covers some areas not handled by Shea, such as manufacturing techniques, specialized semiconductor devices, and silicon devices. As far as the audio sections go, the authors use the less-favorable "r-parameters" in developing formulas for amplifier design. Furthermore, in a few places the assumptions the authors make lead them to confusing, or downright misleading statements. Shea's treatment is to be preferred as far as audio amplifier design goes; however Coblenz and Owens' book does have other worthwhile features.

L. M. Krugman, *Fundamentals of Transistors*, Rider, New York, 1954. Claimed to be aimed at the technician and amateur, this book in reality is so well-written that it will serve the needs of experienced engineers as an introduction to the field. A minimum of formal mathematics is needed, but this does not seem to restrict the material unnecessarily. Unfortunately, this book came out before the b parameters were in common use; therefore has the disadvantage of using r-parameters throughout. However, as a succinct introduction to the field, for students, engineers, and technicians, this book is good.

R. P. Turner, *Transistor Theory and Practice*. Gernsback Publications, New York, 1954. This book is on a slightly lower technical level than Krugman's. It will fill adequately the needs of those without much technical training, and without much mathematical prowess. More sophisticated readers will resent many of Mr. Turner's assertions "out of the blue," but this is characteristic of writing aimed at Mr. Turner's audience.

L. E. Garner, Jr., *Transistors and Their Applications*, Coyne Publications, Chicago, 1953. This is a very elementary book aimed at beginners not only in transistors, but in electronics and mathematics as well. Probably not suited for the readers of AUDIO.

28 Uses for Junction Transistors, Sylvania Electric Products, Inc., New York, 1955. A collection of 28 elementary circuits, most of them not in a form to be used immediately without more design work. This should serve as an inspiration and "idea book" for experimenters. 12 of the 28 circuits are audio amplifiers.

Raytheon Transistor Applications, Raytheon Manufacturing Co., Newton, Mass, 1955. A collection of 53 short articles, many reprinted

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Transistor Preamp for Low-Output Pickups

ANTON SCHMITT

A low noise, transistorized preamp for flat amplification of pickup signal affords greater fidelity in sound reproduction, and can be built with assurance of successful operation
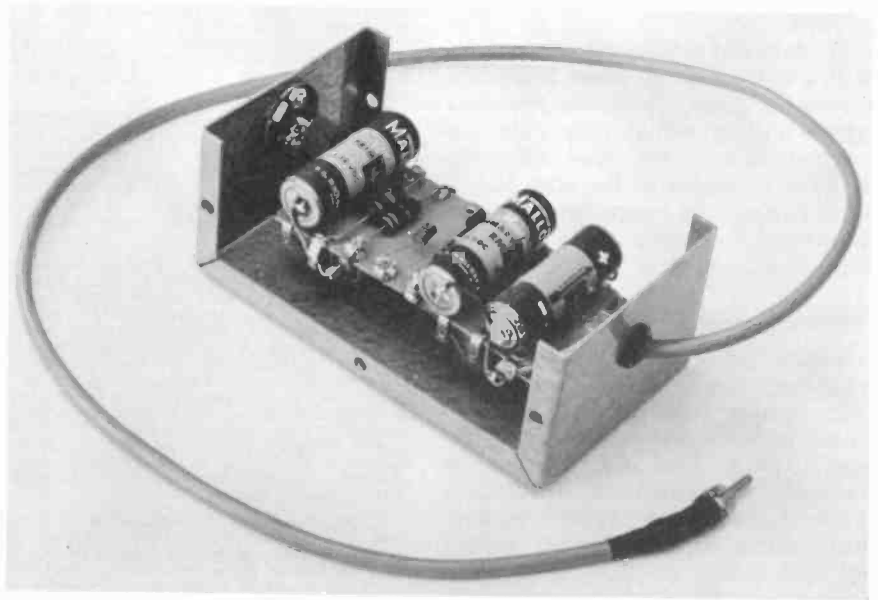

Fig. 1. Simple transistor preamp provides gain necessary to permit connecting low-output pickups to normal preamplifier which incorporates suitable equalization curves.

IN RECENT YEARS the quest for better sound reproduction has led to the development of a number of improved types of phonograph pickups. One such high-quality electromechanical transducer now enjoying a widening use is the moving-coil pickup, in which the coil is arranged either to rotate in the magnetic field like a d'Arsonval meter, or simply to move within the field. Its advantages include (1) low effective mass and a high compliance, (2) a resonant frequency above audibility, (3) low impedance to minimize hum pickup difficulties and (4) a highly linear response—all of which are important factors in the production of clear, "quality" voice and music signals.

One drawback to the moving-coil pickup is the low output signal it generates, usually of the order of 0.5 to 5 millivolts. This is appreciably lower than the 15- to 20-millivolt output of the usual magnetic or variable reluctance pickups. The higher voltage levels ordinarily are required at the high gain inputs of most preamplifiers. For this reason, most moving-coil pickups are designed to work into a matching step-up transformer, which makes possible the signal voltage output of 15 to 20 millivolts needed to override hum and noise in the preamplifier input stage.

While presently available input transformers are markedly improved over earlier versions, the writer's observation has been that even the best of these units fails to satisfy the stern requirements of the "no compromise" school of audiofans. To deficiencies in the input transformer may be traced such disappointing phenomena as hum pick-up, peaked high end response, "spread" or inadequate bass (comparable to that resulting from improperly damped speakers) and an unnatural coloration of program material.

An obvious approach to the problem would be through the elimination of the input transformer. While producing greater fidelity, such a solution unfortunately entails the employment of maximum or near-maximum settings of the gain control of the preamplifier—a practice likely to produce an undesirable

noise level in even the best modern preamplifier.

Why not, then, substitute for the input transformer a cascode-stage preamplifier? Offhand, would not the 6BQ7 so used appear to offer possibilities? After a number of experiments, the writer found that the cascode amplifier was not very satisfactory for this particular application. The limitations of the cascode circuit for use in phono preamplifiers have been pointed out by Marshall.[1]

Transistor instead of Transformer

An approach recently made by the writer entailed the development of a simple and basic low-noise, high-gain transistorized preamplifier, designed to replace the input transformer usually used with low-output moving-coil pickups. It can be built in a few hours, and costs even less than a top-quality input transformer. The results of both instrument and ear tests of this unit have been most gratifying.

Prior to the development of the device it was decided that the finished product,

[1] Joseph Marshall, "The grounded ear," *Audiocraft*, Dec. 1955.

to merit attention, would have to satisfy the following requirements:

1. The gain must equal or exceed that of readily available commercial input transformers.
2. No noise level above that of "negligible" could be tolerated.
3. Hum level must also be negligible—if not eliminated.
4. Sound reproduction quality, by subjective listening tests, must be superior to that obtainable by the use of an input transformer.
5. Frequency response must be linear from 10 to 40,000 cps.
6. High-end response must be clean and smooth, neither "peaked" nor "wiry"; low-end response must be "tight" and clean.
7. Circuitry must be simple. Components needed—especially the designated transistors—must be readily obtainable and low or moderate in cost.

The transistor preamp described here successfully meets these criteria. A number of units are now in use, and have been commended by seasoned and highly critical listeners.

Construction

This preamp requires only a few parts and the circuit, *Fig.* 2, is simple. Since direct coupling is used between cartridge and transistor, the limiting factor for

bass response is the input resistance of the regular preamplifier. With the 47,-000-ohm load generally used for the more common magnetic cartridges as the input resistance, low-frequency response will be down less than 2 db at 10 cps. High-frequency response will be down less than 2 db at 75,000 cps.

Resistor R_1 and capacitor C_1, in conjunction with the battery in the emitter branch, automatically provide proper operating potential between the collector and the emitter. The degenerative action of unbypassed resistor R_2 in the emitter circuit reduces distortion, smooths overall frequency response and increases input resistance. An input resistance of about 5000 ohms is suitable for such cartridges as the Electrosonic and the Fairchild. An increase in the value of the resistor R_2 raises the input resistance, but lowers the gain.

The collector load resistor R_3 affects both the gain and the amount of thermal noise. The collector current is constant at 200 microamperes for values of R_3 from zero through 8200 ohms. Beyond 10,000 ohms, the base current increases from a low of 10 microamperes, and the collector potential becomes too low. In this condition, there is a loss of gain and an increase in distortion. With a load of 6800 ohms, gain is 20 db and noise is barely audible. Increasing the load to 8200 ohms increases the gain but also increases thermal noise.

Three mercury batteries are used: one is in the emitter, and two are used in series to supply the collector potential. In *Fig. 1*, they are shown soldered in place, since at the time of construction the battery holders currently available were not on the market. With a measured drain of 200 microamperes (with pickup plugged in), battery life in continuous service is estimated at over one year. Battery life could be extended if an off-on switch were installed, or the pickup removed from the arm when not in use. Several units have been in continuous operation for a year without battery replacement.

Despite battery operation, difficulty with induced hum may be experienced unless proximity to power transformers and turntable motors is avoided. Returning all grounds to a point at the input
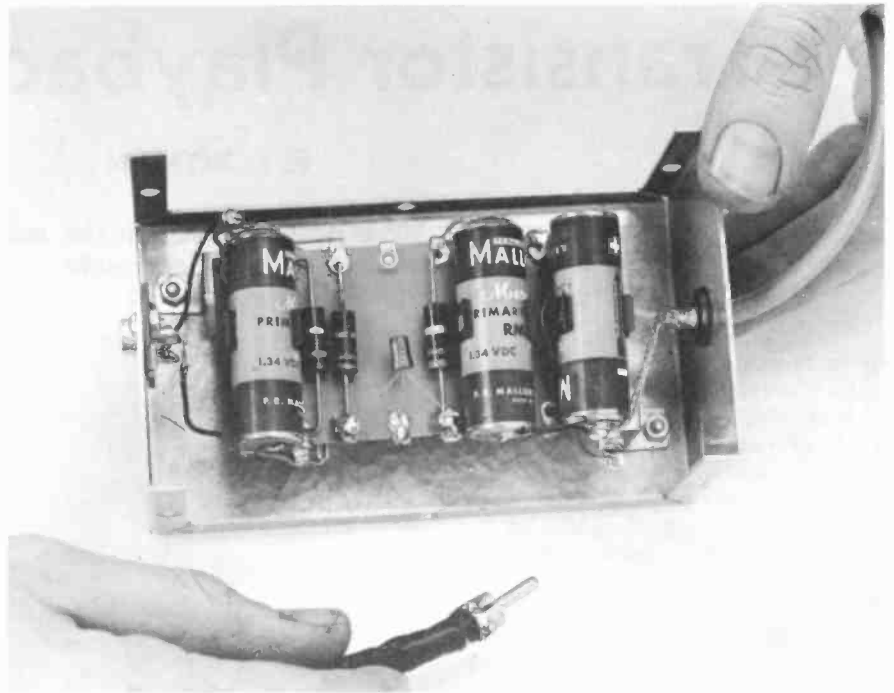
Fig. 3. Close-up of internal construction. Battery clips specified in parts list eliminate need for soldering batteries in place.

to the shield will minimize any induction hum problem.

All of the low-noise transistors currently available have functioned satisfactorily in this circuit, although occasional individual units have been excessively noisy. It is recommended that a transistor be checked for noise before clipping the leads and soldering in place. In soldering, pre-tinning the leads and lugs and brief applications of heat should permit making satisfactory connections without risking damage to the transistor. Reversed battery polarities can be ruinous to transistors and caution should be practised—although in the experiments accompanying the development of the present circuit, polarities were reversed on a number of occasions without apparent damage to the transistor, probably because of the low potentials present.

Among the transistors the writer found satisfactory in this application were the Raytheon 2N133, and the RCA 2N77, 2N109, and 2N105. In the preamp shown, the Raytheon 2N133 was used. The transistor is soldered in place with

the red dot facing the batteries which supply potential to the collector—the lead soldered on the terminal lug nearest the two batteries in series is the collector lead.

The preamp was built in a $3'' \times 2\frac{1}{8}'' \times 5\frac{1}{4}''$ gray hammertone aluminum box. The resistor strip is a Miller 450. A two-foot, single conductor shielded microphone cable serves as the output lead, entering the box through a grommet. Component values are given in the accompanying parts list. *Figures* 3 and 4 show two additional views of the completed preamp.
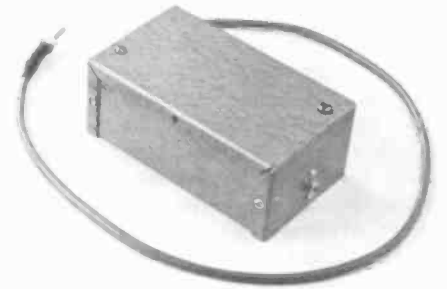

Fig. 4. External view of completed preamp.

PARTS LIST

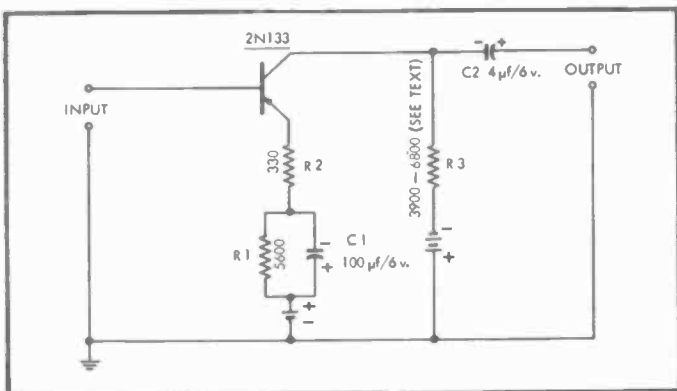|	
C_1	100 µf, 6 volts, electrolytic
C_2	4 µf, 6-volts, electrolytic
R_1	5600 ohms, 1 watt
R_2	330 ohms, ½ watt*
R_3	3900 to 6800 ohms, 1 watt (see text)
1	Transistor—Raytheon 2N133, RCA 2N77, 2N105, or 2N109
1	Acme Battery Holder, #153
1	Aluminum box, $2\frac{1}{8} \times 3 \times 5\frac{1}{4}$
1	Miller #450 Resistor Mounting Strip
3	Mallory RM-12R mercury batteries

Fig. 2. Schematic of the transistor preamp.

A Transistor Playback Amplifier

F. E. WYMAN

The author describes a new approach to the method of equalization for a tape amplifier along with the complete unit which—when constructed in duplicate—serves as an ideal stereo playback "front end."

THE PROBLEM OF magnetic tape playback amplifier design is one of the more challenging problems encountered in the audio field. The present state of the art contains some very satisfactory vacuum tube playback amplifiers as exemplified by the performance of several very expensive professional recorders now in production. It is the purpose of this paper to demonstrate that the performance of these amplifiers can be matched by the application of transistors to the problem.

The acquisition of one of the popular tape transport mechanisms with its stacked stereophonic playback head offered the author the opportunity of indulging in a growing desire to attempt to design a satisfactory transistor playback amplifier. The particular unit at hand was complete with the manufacturer's single channel record and playback amplifier, but in this unit the playback electronics possessed an intolerable amount of 60-cps hum. In addition, the location of the power transformer in proximity to the pickup head added an additional amount of hum.

The most desirable course of action appeared to be the design of a two-channel battery-operated transistor pre-amplifier that would be small enough to fit in the storage space to the rear of the transport deck. The unit shown in *Fig.* 1 is the result of the successful conclusion of this project. The design allows the inexpensive unit to function as a single channel or stereophonic tape playback device with performance characteristics closely approximating the much more expensive professional machines. In addition, the arrangement is such that the normal single channel record function, by means of the original electronics, is retained.

Specifications and Design Considerations

The specifications set forth at the outset were severe and admittedly a bit arbitrary in view of the fact that the possibility of attainment was unknown. It seemed unwise to set the goals lower than the performance of the best grade vacuum tube amplifiers and to this end the following initial specifications were used as a target:

1. Signal-to-noise ratio, measured

Fig. 1. The completed two-channel transistor playback amplifier.

from the 3 per cent tape distortion signal level to amplifier noise, of at least 55 db.

2. Output level, for zero level signal, of one volt.

3. Equalization to allow playback flat to ± 2 db from 50 to 15,000 cps from a pre-equalized tape.

4. Distortion low enough that the 3 per cent distortion point due to the magnetic process could be established in all cases.

5. Low output impedance to accommodate a practical length of cable without loss of high frequencies.

Having defined a goal it is well to examine the parameters that must be dealt with. The playback head in this case was a Dynamu stacked stereophonic model and therefore it was possible to do all the design work and testing with an available Dynamu 8001 single track head. This procedure kept the new tape recorder free for normal use until design and construction were complete. *Figure* 2 shows the published frequency *vs.* output characteristic of this head in curve (A). It is typical of almost any head likely to be encountered and shows the characteristic 6-db-per-octave rise from low frequency to a point near 3000 cps where various losses take over and reverse the rising trend.

The over-all electronic system of any satisfactory tape recorder must operate on this characteristic and produce an output characteristic that is flat as nearly as possible over the desired frequency range. When extending the NARTB standard into the field of home recorders, it is customary to follow the practice of compensating for the region above the peak output in the record amplifier. Thus it remains for the subject amplifier to provide a characteristic falling at the rate of 6 db per octave from the lowest frequencies to about 3000 cps and a flat characteristic on to the highest frequency desired.

Figure 2 also shows, in curve (B), the published frequency *vs.* head impedance characteristic for the same head. Again this is typical of all heads, varying only in value and the exact location of the peak. The important fact is that heads that are of the class available to
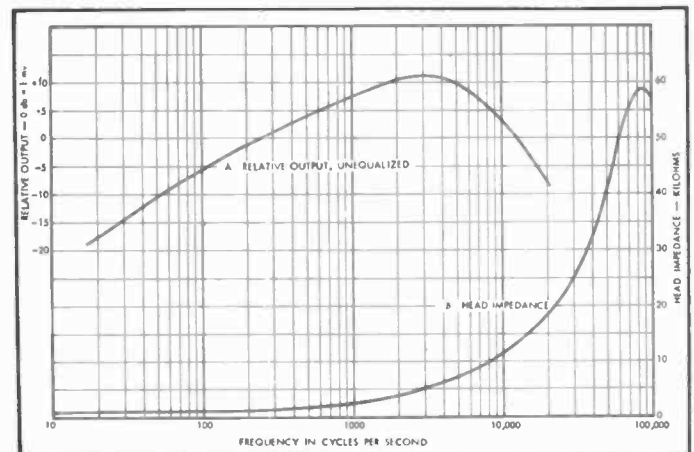
Fig. 2. Curves of typical Dynamu 8001 head. (A) Relative output level vs. frequency, and (B) impedance vs. frequency.

home constructors are of relatively high impedance. They must work into an amplifier of even higher input impedance or suffer a serious loss at the higher frequencies. When one uses conventional vacuum tubes for the input stage, the input impedance will automatically be high and only with the highest impedance heads will simple precautions have to be taken. The input impedance situation is alarmingly different with transistors. The flood of transistor articles in recent literature has made users aware that transistors are essentially low-impedance current-sensitive devices. In any transistor configuration that will give voltage gain, the input impedance will be somewhere in the range of a few ohms to a few thousand ohms. Furthermore, any given parameter is dependent on all other parameters. In the particular case at hand if the output loading is modified for some design consideration of the second stage, the input impedance of the first stage will change, usually with embarrassing results in the downward direction.

The Circuit

The circuit shown in *Figure* 3 is that of one channel of the complete stereophonic amplifier. As in so many developments considerable experimentation and adjustment of component values took place until the desired performance was obtained. At this point the device was put to work and no further effort exerted to further optimize the circuit. Two copies of the amplifier have been made and were found to perform essentially the same as the original, hence it would appear that sufficient stabilization has been incorporated to render the circuit relatively free of individual transistor variation.

RCA 2N105 transistors were used throughout simply because they were readily available and appeared to possess satisfactory characteristics. A review of all manufacturers' data would indicate that there were other transistors with even lower noise figures which could be used to advantage. Nevertheless, using only the one type and picking the individual transistors at random gave very satisfactory results.

The first stage, *Fig.* 3, is the common collector configuration. It has an input impedance of 0.15 megohms and hence does not load the playback head or cause any attenuation of high frequencies. As a result of this circuit configuration the voltage gain is slightly less than unity but the output impedance is low, thus this stage has the effect of performing an impedance transformation. This is perhaps the most satisfactory means of transforming the high impedance of the head to the low impedance of the second stage, but the use of a second transistor operating at low level contributes to a
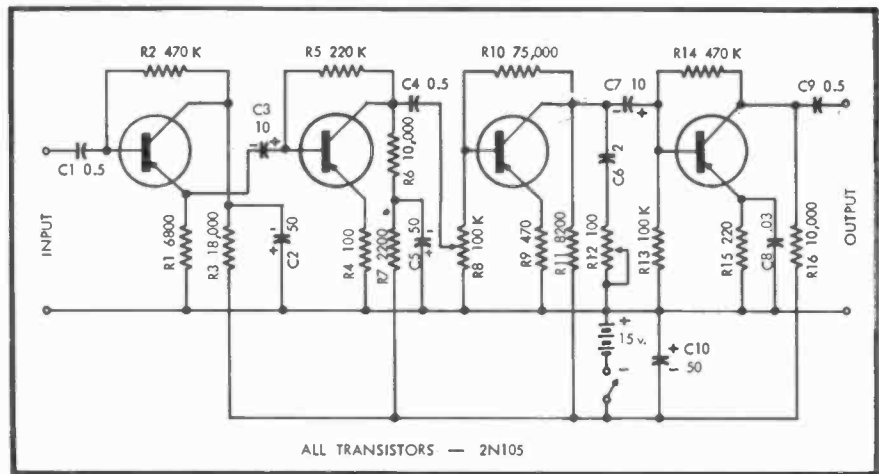

Fig. 3. Schematic of a single channel of the transistor amplifier.

greater noise output. R_1 in the emitter circuit is the load resistor, R_2 sets the base current and the operating point, C_1 blocks an additional d.c. path and thus protects the operating point, and C_2 couples the output to the next stage. It is essential that the designated polarity of this latter capacitor be observed.

The second stage is the more conventional self-biased common-emitter configuration operating at a source voltage of about − 8 volts obtained by means of the bypassed dropping resistor R_7. Resistor R_4 in the emitter circuit is common to the input and output circuit and thereby introduces considerable degeneration and reduces the voltage gain of the stage to about eighteen. At first look this seems to be wasteful of gain but it is by the use of degeneration that the low second and third harmonic distortion is obtained.

The third stage is again a common-emitter stage operating from the full 15-volt battery supply. It is biased by

the combination of R_8 and R_{10}. Equalization for the characteristic 6 db per octave rise of the playback head is accomplished in the output of this stage by means of R_{12} and C_6. This method is typical of vacuum tube amplifier applications but the value of resistance may seem alarmingly low. In fact, when the amplifier is adjusted for the playback characteristic shown in *Fig.* 5, R_{12} has the value of about 40 ohms and C_6 has an equal value of reactance at the turnover point of approximately 2000 cps. Due to the low impedance property of transistors the 40-ohm value is in order. R_8 is an Ohmite miniature potentiometer and serves as a volume control. Readers accustomed to volume controls in voltage operated devices such as vacuum tubes are almost sure to feel that a mistake has been made in the circuit diagram, though the proper use of a current divider in a current-operated device is designated.

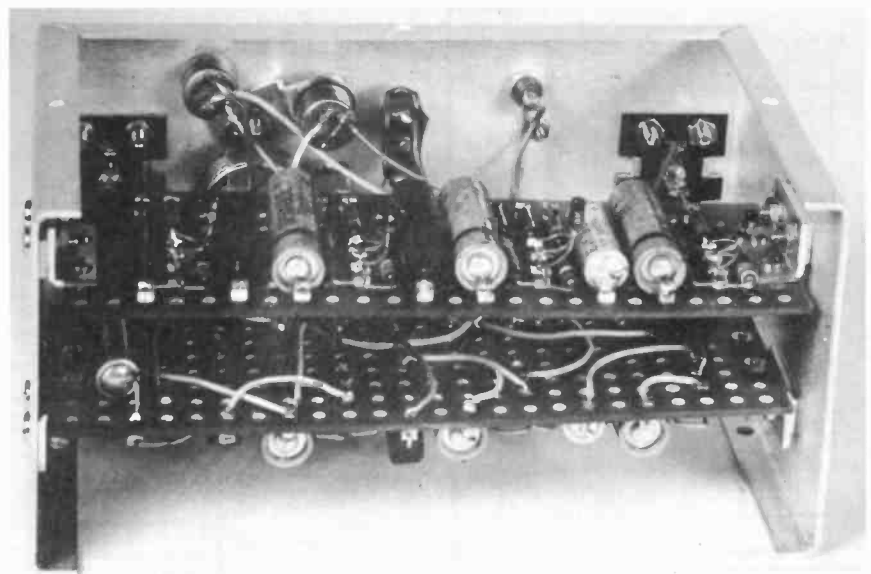After the application of equalization,

Fig. 4. Internal view of the amplifier. Each channel is constructed on its own "chassis" —assembled from Alden perforated terminal boards.

the midrange and high frequencies have been attenuated to a low value and must again be amplified by a fourth stage in order to meet the design specification of one volt output for normal tape level. This stage is also a common-emitter type with emitter-circuit feedback to allow the relatively large output signal to be obtained with satisfactorily low distortion. It differs from the previous stage in that it works into the high impedance of the first grid of a following amplifier. This allows considerably higher gain in spite of the feedback and calls for a different operating point along with a much higher value of R_{14}. Secondly, the emitter resistance R_{15} is partially bypassed by C_8. Each stage from the second on suffered from a slight high-frequency attenuation due to a low value of beta cutoff. This attenuation accumulated to about 3 db loss at 15,000 cps and the .03 μf capacitor controls the degeneration due to R_{15} sufficiently to bring the response up to flat at the high end.

Construction

The construction is clearly shown in *Fig.* 4. The unit pictured contains two identical channels. The gain control R_8 is a two-gang potentiometer common to both channels and an additional miniature potentiometer has been included in each channel so that small differences in gain or differences in output of the two heads can be compensated for. On-off switches have been provided for each channel so that battery life can be conserved when operating from a single track. The miniature U 10 battery is

mounted in an external clip for easy replacement. Probably a more satisfactory voltage source would be a pair of 7½ volt "C" batteries, Burgess type 5540, since each channel draws about 5 ma.

The author found it very convenient to use Alden perforated terminal boards and the appropriate lugs that can be riveted at any point, but any form of terminal board on which to mount the components should be satisfactory. As far as could be determined from experience with a breadboard prototype and the two pictured amplifiers there is nothing critical about placement of components or lead dress.

The transistors, RCA type 2N105, come with long flexible leads and it is convenient to clip them to the desired length and solder directly onto the mounting lugs. This eliminates the bothersome task of mounting the sockets and the questionable contact between the socket and the small leads. Transistors, unlike vacuum tubes, have long life and barring a catastrophe should last for an indefinitely long time. Of course, because of their nature, it is necessary to observe some simple precautions in soldering them into the circuit. If the short lead is grasped between the fingers and quickly soldered with a hot iron to a well tinned lug, the operation can be completed before any warmth can be felt on the fingers. Under these conditions there is no danger of heating the junction inside the small case to a temperature that will destroy it. While on the subject of temperature it is well to observe that the manufacturer places the rating of 50° C. (122° F.) as the maximum operating temperature for the 2N105 transistors. This will require some care when placing the amplifier near heat generating electronic components in an enclosed space. In the pres-
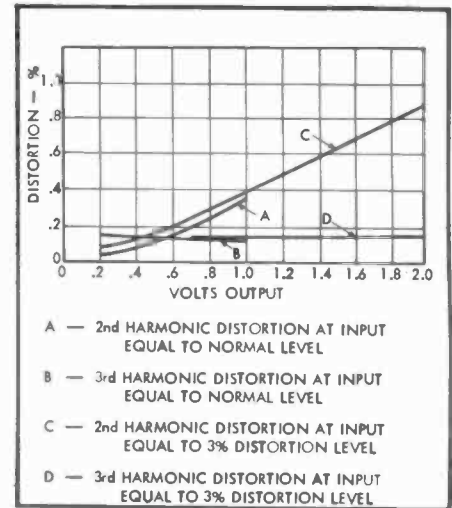

Fig. 6. Curves showing output voltage vs. distortion. (A) and (B) show 2nd and 3rd harmonic distortion respectively at input equal to normal level. (C) and (D) show the same distortions at an input of a level suitable to provide output at 3-percent distortion.

ent example it seemed satisfactory to mount the amplifier on the back side of the tape deck in a well ventilated place away from tubes and transformer.

Performance

At the inception of this project the author had considerable misgivings as to the wisdom of using transistors in playback amplifier service. As the work progressed the fears gradually disappeared but an unusually large amount of testing was done. Since it is feared that
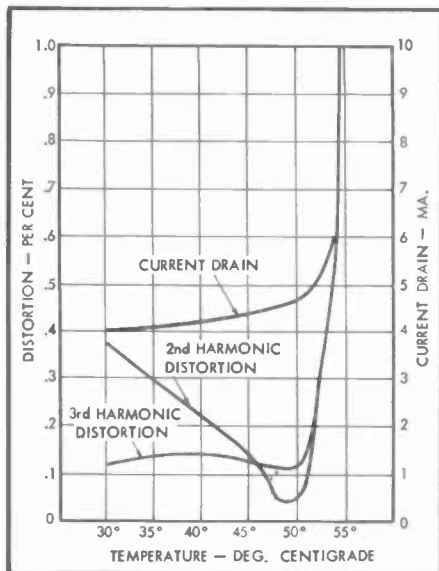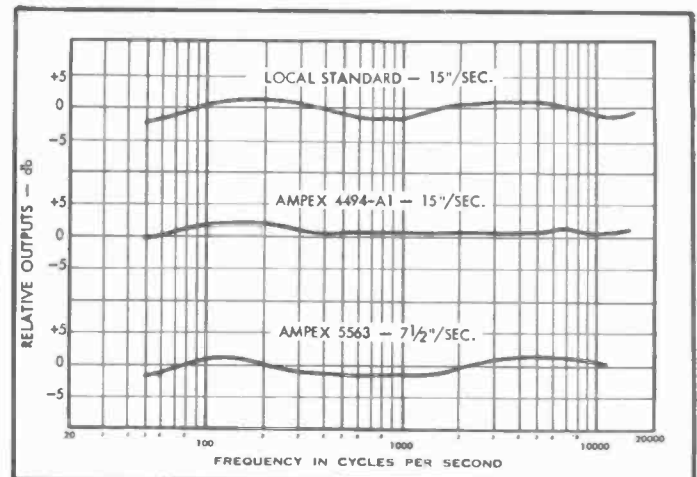

Fig. 5. Curves of distortion and current drain vs. temperature for a 0-level 400-cps signal.

Fig. 7. Output vs. frequency response as measured from two Ampex standard tapes and one "local standard."

some of the readers will have similar doubts the performance and results will be reported in considerable detail.

Perhaps the first requirement of such an amplifier is that it must play back a pre-equalized standard tape in a flat manner. This amplifier was intended to be used with a machine operating at 7½ ips but an adequate design should perform well at 15 ips with little or no equalization adjustment. Three pre-equalized standard tapes were available and all were used. They consisted of an Ampex number 5563 at 7½ ips, an Ampex number 4494-A1 at 15 ips, and a very carefully prepared local standard at 15 ips. *Figure* 7 shows output *vs.* frequency from these three tapes as measured on a Hewlett Packard 400C voltmeter. Curve A is the result of adjusting the equalization for the best compromise with the local tape. Curve B was taken from the Ampex 15-ips tape without additional adjustment. Curve C was taken from the 7½-ips tape after slight readjustment. Without readjustment the results at 7½ ips would have been satisfactory. It can be seen that all tapes play back to ±2 db which is flat enough for the best professional recorders.

With the amplifier properly equalized, noise measurements were made. Since it was the amplifier that was under test and not the over-all system, the head was replaced with a 300-ohm resistor to represent the resistive component of the head impedance. The noise was measured on a Hewlett-Packard voltmeter and compared to the output due to a signal equal to the 3 per cent tape distortion level. The noise was found to be 52 db below such a signal. This in itself is good but an examination of the frequency content of the noise revealed it to contain a large component of very low frequency. This is a well known characteristic of transistors and since even the best speakers could not reproduce such frequencies it was decided to define an equivalent noise ratio. The low-frequency component was eliminated by means of a sharp-cutoff high-pass filter set at 40 cps. Within the meaning of this definition the 3 per cent distortion signal-to-noise ratio was 59 db which makes it comparable to the best professional units.

Transistors without adequate bias stabilization are known to exhibit objectionable temperature characteristics. As the temperature rises the operating point shifts and distortion may increase without bound or in extreme cases the amplifier may cease to amplify at all. Careful temperature tests were made over the range from room temperature to 54 deg. C. and the results are presented in *Fig.* 5. A 0-level 400-cps signal was fed into the amplifier and the second and third harmonic distortion measured as the temperature was slowly raised. Current drain was observed to

Fig. 8. Curves of distortion vs. frequency. (A) shows 2nd harmonic distortion; (B) 3rd.
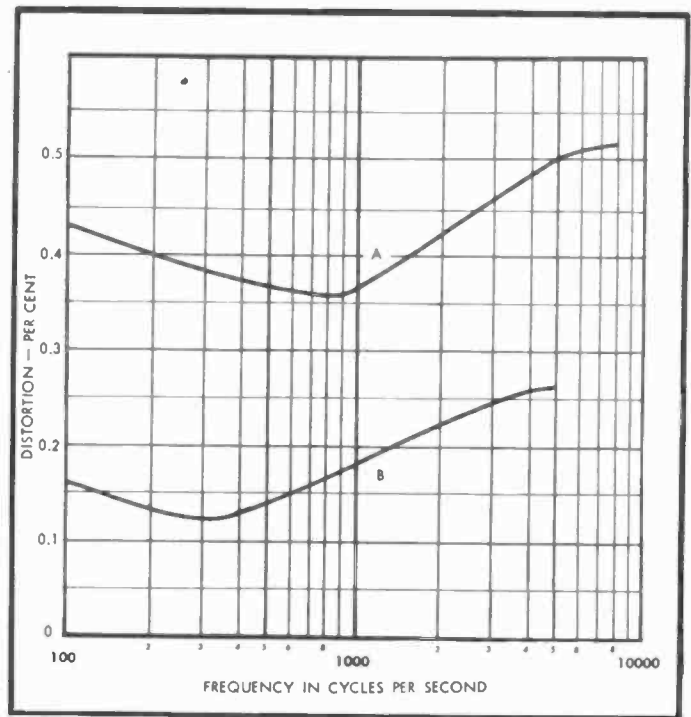
increase from four to five milliamperes while the amplification remained constant to within ±0.25 db. Within the manufacturer's operating limit of 50° C. the observed distortion remained low.

Distortion *vs.* output level is of great interest to Hi-Fi enthusiasts and the results of such measurements are given in *Fig.* 6. Curves (A) and (B) give the second and third harmonic distortions respectively for a 400-cps 0-level signal with the gain of the amplifier varied to give outputs up to one volt. Curves (C) and (D) show the same for a 3 per cent tape distortion signal. *Figure* 8 shows the second and third harmonic distortion as a function of frequency. Since modern final amplifiers will give listening level output with much less than one volt input, it is safe to say that the harmonic distortion of the subject amplifier will always be much less than 0.4 per cent at all frequencies. Thus the distortion compares favorably with the best professional amplifiers.

The final specification has to do with the output impedance and the ability to drive a length of cable without appreciable loss of high frequencies. The measured value of output impedance turned out to be 900 ohms which is within the range of average cathode followers. To discover how much cable capacitance could be tolerated a capacitance decade box was shunted across

the output and the capacitance adjusted to give a 3-db attenuation at 15,000 cps. A total capacitance of 1000 μμf was required. This would represent at least ten feet of the highest-capacitance shielded cable.

The final test can be reported only in subjective terms. About all the good quality program material available to the author has been auditioned before many people. In all cases the amplifier received enthusiastic acclaim.

Conclusion

It has been repeatedly mentioned that the goal was to equal the performance of the best vacuum tube playback amplifiers The section on performance has demonstrated that this goal has been attained. The question remains then, what are the advantages of this amplifier? No such claims are made. It simply offers freedom of design, a new way of obtaining the highest performance. In some applications the use of transistors would be indicated while in others the opposite would be true. The one case that comes readily to mind is that of a playback-only unit. The compact size and the freedom from 60-cps hum problems make a transistor playback amplifier especially attractive.

Transistor Tone Control Circuits

F. D. WALDHAUER

The author describes a circuit thoroughly and discusses its derivation in a manner which should increase understanding of transistor circuit operation.

IT IS NOW A PRACTICAL MATTER to utilize transistors in all of the audio circuits of a high fidelity system. In addition to the advantages of small size and lack of heat brought about by the high efficiency of the transistor, excellent long-term reliability, low noise, and ruggedness coupled with complete absence of microphonics may be obtained. In power amplifiers, the absence of an output transformer permits a much greater amount of feedback than is usable in conventional vacuum-tube amplifiers for a given stability margin. Hence, feedback, in conjunction with the intrinsically low distortion characteristics of transistors, may be used to produce power amplifiers having low distortion at power levels exceeding 50 or 100 watts. At the other end of the system, where noise limits the dynamic range, transistors may be advantageously employed in preamplifiers for tape playback heads or low-output phonograph cartridges with noise characteristics closely limited, as in the case of only the best vacuum-tube amplifiers, by thermal noise itself. The lack of a heater with its required rectifier and filter makes possible this high-quality performance even in relatively low-cost equipment.

We shall discuss both low-level and high-power circuits in future articles. At present, we shall be concerned with control circuits, and, in particular, a dual-tone-control circuit employing transistors. Dual-tone-control circuits, in which the bass and treble signals are boosted or attenuated relative to mid-frequencies by two separate controls, have long been used in vacuum-tube control amplifiers. An early transistor tone control circuit described by the present author[1] has recently been given some attention[2]. The circuit to be described here offers several distinct advantages over this early transistor circuit.

Objectives

Aside from the usual audio-circuit objectives of low noise and distortion,

[1] Lo, Endres, Zawels, Waldhauer, and Cheng, "Transistor Electronics", Prentice Hall, Englewood Cliffs, N. J., p. 187–189.
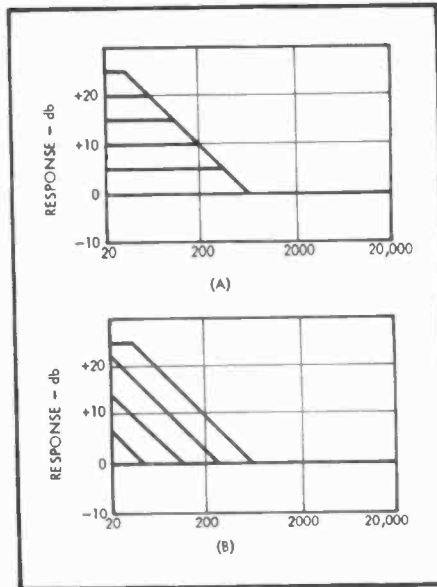[2] H. R. Lowry, "All transistor hi-fi amplifier", Radio News, Nov. 1956.

Fig. 1. Frequency response asymptotes for bass-boost characteristics illustrating (A) variable shelving, and (B) variable turnover frequency characteristics.

reliability and stability with respect to environment, several further requirements may be imposed upon a transistor tone-control circuit to secure practical, economical, and functionally desirable characteristics. The total range of treble and bass boost and cut must be large and, what is more important, the way in which the frequency response varies with bass and treble control setting must be carefully controlled. From both the economic and functional standpoints, continuously variable controls are desirable. Inductors are to be avoided on the basis of economy, hum pick-up, and possible transient response (by virtue of parasitic capacitances usually accompanying inductors). The tone-control circuit should be capable of being set accurately to the flat response position, and should operate smoothly with respect to rotation of the controls.

Response Characteristics

Two basic types of response characteristics are possible in a simple dual-tone-control circuit. The first type, termed *variable shelving* is illustrated at (A) in *Fig.* 1 for a series of asymptotic bass-boost curves. This type of response is characterized by a turnover frequency which is fixed for any position of the control. The response flattens off at a value determined by the control setting. The second, or *variable turnover* type of response is illustrated at (B) in *Fig.* 1, and is characterized, as the name indicates, by a turnover frequency which varies in accordance with the control setting. The two types of curves may have the same maximum boost or cut characteristics, but the operation for intermediate positions of the controls is markedly different. With a small amount of boost, the variable-turnover characteristic emphasizes the extreme frequencies while the variable shelving characteristic, at the same control setting, emphasizes the middle high or low frequencies.

The variable-turnover type of control appears to be the more desirable of the two, and is coming into increased favor as its characteristics become more widely known.

Advantages of the New Circuit

The advantages of the new circuit over both the author's early transistor circuit and the more common vacuum-tube circuits may be summarized as follows. First, the present circuit may be arranged to give a variable-turnover type of frequency-response variation. As a direct result of this, an accurately flat position of the tone controls may always be achieved even without use of close tolerance components. This comes about by virtue of the fact that in the flat position, the turnover frequencies are moved out of the audio-frequency range entirely. This is to be contrasted with the variable shelving control in which variation in standard tolerance components may produce irregularities of response in the maximally flat position of over 3 db. Another advantage of this circuit is the reduction in the number of components as compared with conventional circuits. In addition, the circuit is readily "designable"; that is, individual components may be altered to secure a desired result without adversely affecting the operation of the remainder of the circuit. Finally, a considerably greater range of control may be obtained with this circuit than with the previous one.

A Practical Circuit

A practical version of the tone control circuit, with circuit values for obtaining somewhat more than ±20 db
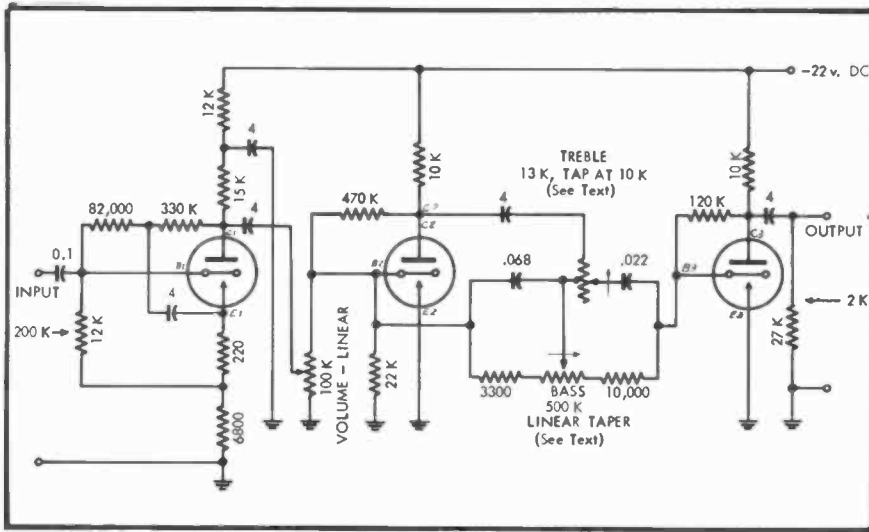
Fig. 2. Circuit schematic of amplifier incorporating, in the second stage, the new tone-control circuit. The first and third stages provide amplification as well as suitable impedance levels for both the tone-control stage and external circuits.

range of bass and treble control, is shown in *Fig*. 2. The second stage is the tone control, while the first and third stages provide amplification as well as suitable source and load impedances for both the tone control and externally connected circuits. The transmission characteristics of the entire circuit for various settings of the bass control are shown in *Fig*. 3, and similar curves for the treble control are shown in *Fig*. 4. The bass and treble characteristics are completely independent. The total response for any condition of the bass and treble controls thus may be obtained by adding the curves of *Figs*. 3 and 4.

It is noted in passing that the volume control located between the first and second stages has a logarithmic or audio attenuator characteristic even though it is itself a linear tapered potentiometer. This will be explained in the subsequent article, in which the present tone control circuit will be incorporated in a complete preamplifier and control
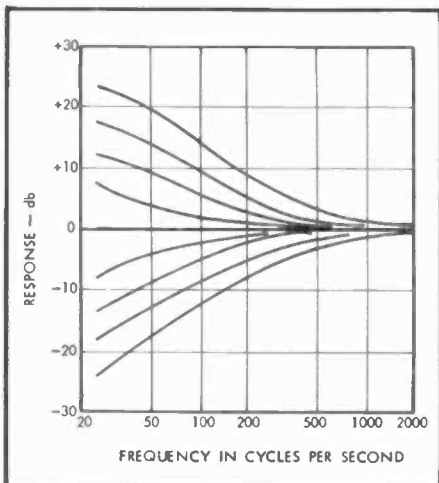
circuit, having, among other things, record equalization for both magnetic and FM type pick-ups. The over-all circuit shown in *Fig*. 2 is designed to provide 1 volt output with 0.25 volt applied to the high-impedance input, and is thus suitable for use with a tuner or other fairly high level audio source. For those who may wish to construct the circuit of *Fig*. 2, it is noted that all parts are standard except possibly for the 4-μf coupling capacitors which may be either tantalum or aluminum electrolytic capacitors, and the treble-control tapped potentiometer, which has a total resistance of 13,000 ohms, 3000 ohms on the clockwise side and 10,000 ohms on the counter-clockwise side of a tap located at the midpoint of rotation[3]. The transistors should have common emitter-current gains, α_{cb}, of 100, 50, and 50, ± 33 per cent for the first, second, and third stages, respectively, with noise factors of 10 db or less. Several transistor types normally meet these specifications, such as the 2N109, 2N175 of RCA, the 2N190 of G.E., etc.

The output impedance of the last stage is about 2000 ohms, which is suitable for feeding a fairly long output cable. Distortion is reduced by virtue of local negative feedback on each of the three stages. These and other features will be discussed in the subsequent article.

Principles of Operation

The manner in which the circuit of *Fig*. 2 achieves the results shown in *Figs*. 3 and 4 will now be described. At (A) in *Fig*. 5, a transistor feedback circuit is shown in which d.c. collector

[3] A standard IRC potentiometer, No. Q17118X, may be substituted for this potentiometer, although the total treble control range will be greater than necessary.

current is supplied by the resistor R_D, and d.c. stabilized base bias current is supplied through the resistor R_B. If Z_F is low in impedance compared with R_B with which it is in parallel, and Z_L is small compared with R_D, with which it is in parallel for audio signals, we may approximate (A) of *Fig*. 5 with the circuit of (B). A signal current flowing into the base of the transistor will be amplified and a much larger current will flow into the collector of the transistor. This current will divide at the junction of Z_F and Z_L. The manner in which this current divides depends upon the fact that the signal voltage at the base is negligibly small, so that the ends of Z_L and Z_F remote from the collector are essentially at ground potential. Thus the voltages across Z_F and Z_L are equal and the currents are in inverse ratio of the resistances; in other words, if I_L is the load current and I_F is the feedback current, I_L/I_F is equal to Z_F/Z_L. If the gain from base to collector of the transistor is high (e.g., 50 times) the base current may be small relative to the feedback current. The input current, which is equal to the sum of the base current and the feedback current, is then essentially equal to the feedback current. Hence, the current gain, K_i, from input to load of the circuit of (B) is given approximately by I_L/I_F, and

$$K_i \approx \frac{Z_F}{Z_L} \qquad (1)$$

This relation holds if there is appreciable feedback.

The above relation is the essence of the transistor feedback-tone-control stage. In order to produce a tone control, we must now find a resistor-capacitor configuration which allows us to vary the ratio of Z_F/Z_L as a function of frequency. Note that the current gain is what is sought here. The reason for this is that the impedance of the input circuit of a transistor, in particular that

Fig. 3. Measured response of circuit of Fig. 2 for various bass-control settings.
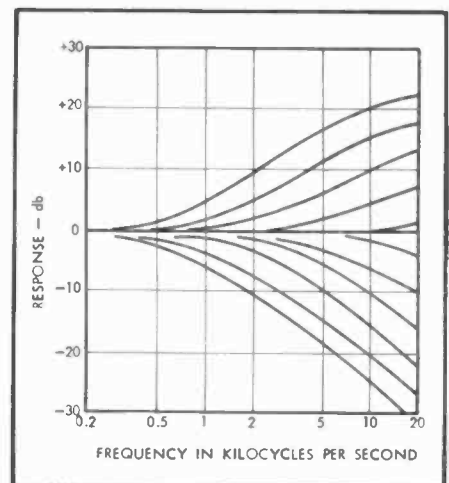

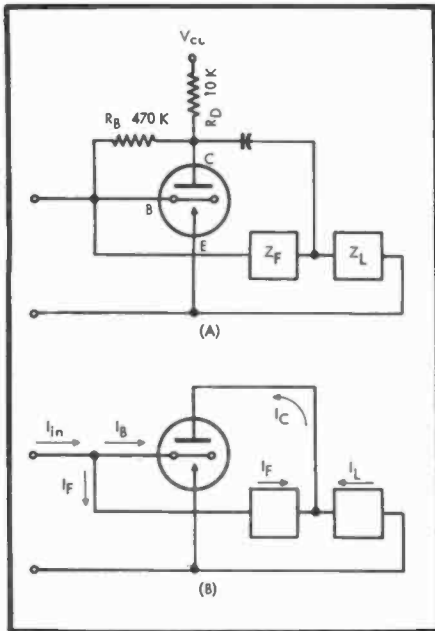Fig. 4. Measured response of circuit of Fig. 2 for various treble-control settings.

Fig. 5. Transistor feedback circuit for obtaining stabilized current gain approximately equal to Z_F/Z_L.

of the following stage, is quite low, and it is therefore more convenient to control the current flowing into the base than it is to control the voltage from base to emitter. The impedance Z_L will be the series combination of a control impedance and the very low input impedance of the following transistor stage.

The particular resistor and capacitor configuration developed for the tone control of *Fig.* 2 produces the desired variable turnover characteristics. This configuration may be adapted for loss-type controls for either transistors or vacuum tubes as well as for the feedback control described here. We note parenthetically that if the tone-control network itself is removed from *Fig.* 2 and connected so that point B3 is driven by a vacuum-tube cathode follower, point B2 grounded, and the grid of a *following* stage connected to point C2, a satisfactory variable-turnover type of tone control for vacuum-tube circuits is formed.

If Z_F and Z_L of *Fig.* 5 are replaced by a pair of capacitors C_F and C_L (whose impedance varies inversely with frequency) the current gain will remain flat with frequency as long as the conditions of Eq. (1) hold, since the fre-
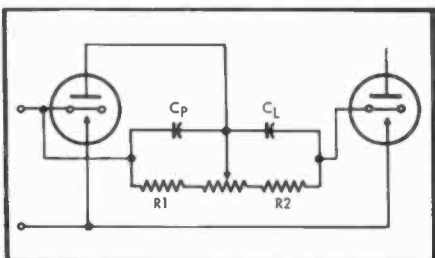
Fig. 6. Basic circuit for the bass control.

quency dependence of Z_F cancels that of Z_L. By proportioning the capacitors suitably, and designing the remaining circuit properly, Eq. (1) may be made to hold over the entire audio band. In particular, Z_L must not get too small relative to the input resistance of the amplifier stage following the tone control, since the high-frequency response will be about 3 db down at the frequency at which the capacitive reactance of Z_L is equal to the third-stage transistor input impedance. Furthermore, at low frequencies, the capacitive reactance of C_F will rise as the frequency is lowered until the low-frequency response is 3 db down at the frequency at which the capacitive reactance of C_F is equal to the total resistance in shunt with it. This total resistance is made up of the parallel combination of three components: First, the bias resistance shown at (A) in *Fig.* 5; second, the collector-to-base resistance of the transistor itself, which is of the order of two megohms; third (and most esoteric), the d.c. load resistor multiplied by the current gain of the transistor, which is of the order of one-half megohm if the transistor current gain is 50 times.

Thus we now have a capacitive network which is essentially a *capacitive* collector-current divider, sending part of the collector current back to the input as feedback (where, if the feedback is large, it is essentially equal to the input current to the stage) and part into the load as useful load current. It should be emphasized that while the impedance of the network falls as the frequency increases, the current gain is flat with frequency. We now wish to use a pair of potentiometers in order to vary the ratio of load to feedback currents at bass and treble frequencies.

For bass control, the ends of a potentiometer are connected as shown in *Fig.* 6, with the slider connected to the collector. As the slider is moved to the right, Z_F remains capacitive and rising as the frequency is lowered, while Z_L becomes resistive and therefore constant with frequency. Thus, according to Eq. (1), the bass signals are increased relative to midband signals. Conversely, as the slider is moved to the left, the denominator of Eq. (1) increases while the numerator remains constant as the frequency is lowered, thereby producing bass attenuation. In each case, as the resistance is varied, the turnover frequency, or frequency at which the response departs from its midband value, varies, thereby producing the desired variable turnover characteristics. The resistors R_1 and R_2 limit the maximum bass turnover frequency to a suitable design value, such as 800 cps.

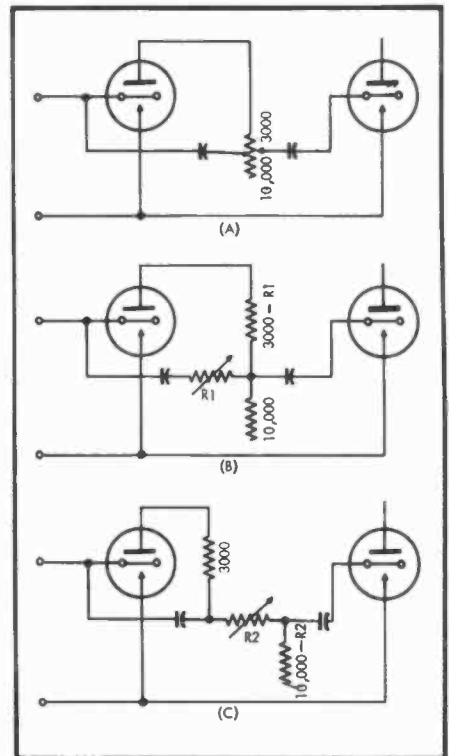For control of the treble signals, resistance may be inserted in series with

Fig. 7. (A) Treble control circuit. (B) Equivalent for (A) under treble-boost conditions. (C) Equivalent under treble attenuate conditions.

Z_F for treble boost, or with Z_L for treble attenuate, as may be seen easily by inspection of Eq. (1). For the case of treble boost, Z_F will comprise the series combination of the fixed feedback capacitor and the series resistance. The turnover frequency will be given by equating this resistance to the reactance of the feedback capacitor. Treble attenuation is obtained by removing the resistance in series with the feedback capacitor and inserting resistance in series with the load capacitor. The turnover for treble attenuation is the frequency at which the inserted resistance equals the reactance of the load capacitor. The series resistance must appear only in one divider arm at a time in order to secure variable turnover characteristics. The arrangement shown at (A) in *Fig.* 7 using a tapped potentiometer performs this function. As the slider is moved upward, series resistance appears in the feedback lead, but not in the load circuit as shown at (B) in *Fig.* 7, thereby giving treble boost; conversely, as the slider is moved below the fixed tap, series resistance appears in the load circuit but not in the feedback circuit as indicated at (C) of *Fig.* 7. The resistance of the potentiometer between the collector and the current division point is negligible compared with the output impedance of the transistor collector, and may be ignored. The combination of the circuits of *Figs.* 6 and 7 gives the configuration shown in the second stage of *Fig.* 2.

Design of the Tone-Control Circuit

The design of a linear amplifier normally starts with the establishment of the d.c. operating point, which in the case of transistors is usually determined by the signal level to be accommodated in the collector circuit (as opposed to tube circuits where the operating point must be set at an inefficiently high voltage- and current-operating point in order to secure reasonable gain and low distortion). The first stage was designed for a transistor having any value of α_{cb} (common-emitter current amplification factor) between 65 and 130. The design range of α_{cb} for the second and third stages is between 33 and 66. The circuits are designed to operate without degradation of performance up to a temperature of 120° F. Collector-to-base degenerative d.c. feedback is used to achieve this stable operation[4]. A design method for R-C coupled amplifiers of this type is given in the literature[5] and is of little more complexity than that involved in the use of "Resistance Coupled Amplifier Charts" for vacuum tubes.

For the tone control stage itself, the values of the feedback and load capacitors as well as the resistance of the potentiometers must be determined. The problem of finding an optimum set of capacitor values is somewhat complex. Suffice to say that the output capacitor should be just large enough to permit full output from the output stage at low frequencies when the signal swing at the collector of the tone control stage is at its maximum value (as determined by the power supply voltage and variations in transistors due to tolerances and temperature effects). This procedure maximizes the high-frequency response while assuring adequate low-frequency signal handling capability. The feedback capacitor should have the smallest value consistent with good low-

[4] F. D. Waldhauer, U. S. Patent No. 2,750,456.
[5] Reference 1, p. 171, Eq. (5–29a); p. 175, Eq. (5–43).
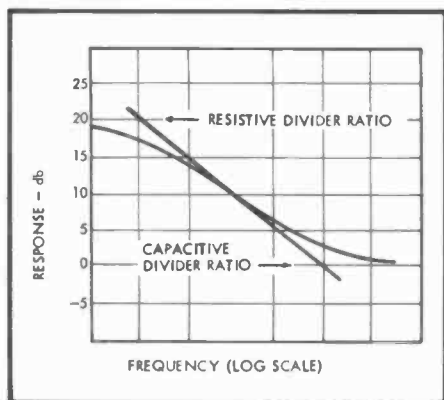
Fig. 8. Bass boost response curve illustrating the manner of operation of the bass control.

frequency response in relation to the resistance it faces, as described above, in order to secure maximum gain or, actually, minimum loss, from the stage. The circuit of Fig. 2 balances these factors in a direction of low distortion and wide range of control at the expense of gain. By substituting the values of Fig. 2 in Eq. (1), it is seen that the tone-control stage actually has a loss of about 10 db. The gain of the other two stages makes up for this loss and provides the additional required gain.

Having determined the capacitor values, we may now find the resistances of the tone-control network. The total treble-control resistances on each side of the fixed tap are chosen to give the minimum desired treble turnover frequency, as would be obtained with the treble control in the extreme positions of boost or attenuate. This frequency is taken as 800 cps. The resistance on the boost side of the treble control is made equal to the reactance of the feedback capacitor at this frequency. The resistance on the attenuate side is equated to the reactance of the output capacitor.

The bass-control circuit includes a pair of resistors connected to the ends of the bass-control potentiometer. These serve to limit the amount of shunting of the feedback and output capacitors, and thereby limit the bass-turnover frequency to a value not exceeding the desired maximum. This maximum was taken again as 800 cps, so that the values of the limit resistors are about equal to the values of the resistances on either side of the fixed tap of the treble control.

The choice of a bass-control potentiometer would be as simple as that of the treble control if we had a potentiometer which shunted only one capacitor at a time. If, for example, bass boost is desired, a resistance should be placed in shunt with the load capacitor C_L, and the resistance shunting the feedback capacitor should be negligibly high. If a potentiometer is connected as shown in Fig. 6, the resistance is not removed from its position shunting the feedback capacitor, so that at some very low frequency, the presence of this resistance will cause the response to return from its rising characteristic (as frequency is decreased) to a flat characteristic. Hence, a certain amount of bucking of the boost and attenuate portions of the bass control exists. This may be viewed in another way which may be enlightening. The current divider at the collector of the transistor is essentially a capacitive divider at midfrequencies. At extremely low frequencies, the reactance of both capacitors becomes very high so that we have essentially a resistive divider made up of the resistances of the potentiometer on either side of the slider. The response will therefore be controlled by the capacitive divider
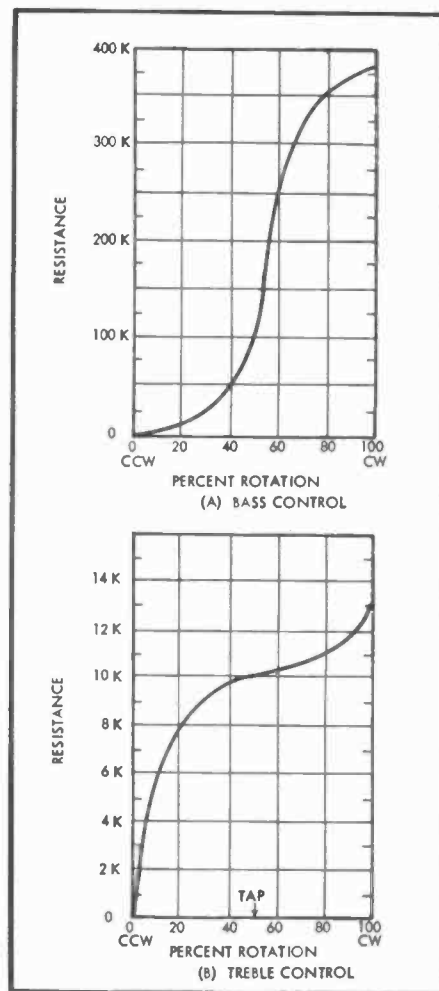
Fig. 9. Curves of resistance vs. rotation for smoothest distribution of the tone control action over the range of the controls.

at midfrequencies and by the resistive divider at very low frequencies; in between, an asymptotic slope of 6 db per octave joins these two levels, as illustrated in Fig. 8. By use of a very high resistance potentiometer, the amount of bucking can be made negligibly small. If the resistance is made too high, however, the tone control action is crowded toward the extremes of rotation of the potentiometer. A compromise is indicated, in which a certain degree of bucking is tolerated in return for smooth tone control action. In the case of the circuit of Fig. 2, reasonably smooth control is achieved with a slight departure from true variable-turnover characteristics. As may be seen in Fig. 3, this departure is small, and in a direction to give less slope as the amount of boost or attenuate is lessened.

Potentiometer Requirements

The requirements for a tone-control circuit closely approaching the objectives stated above have met except possibly in one respect; namely, that of smoothness of tone control action. The ideal control in this respect is probably

A Transistor VU Meter

A simple transistor amplifier will increase a VU meter's sensitivity so that it can be used in many applications.

PAUL PENFIELD, JR.

Fig. 1. A typical modern VU meter. Most meter manufacturers make one or more models of VU meters.

WHENEVER VOLUME IS CONTROLLED for some purpose other than immediately hearing it, some sort of volume indicator is necessary. Recording machines, P. A. systems, and broadcast facilities all need fairly precise control over volume and some sort of indicator.

Of the several ways to observe volume, the best is with a special type of meter —the well-known *VU Meter*, or *Volume Unit Meter*. Less desirable ways include a "magic eye" tube, a series of flashing neon lights, or a conventional-type meter. The reason the VU meter is the best is that it is made specifically for the job.

Professional equipment almost invariably incorporates a VU meter, rather than the less desirable schemes mentioned above. However, the cost of the meter, plus the fact that it is relatively insensitive, combine to keep its use in "home" recorders and even high-fidelity equipment small. While the device described here cannot reduce the cost, it does do away with the problem of sensitivity.

The VU Meter

For those readers not familiar with VU meters, they are a type of voltmeter made especially for reading volume. Program material contains varying levels all the time. Conventional meters do not agree with one another in readings taken of the steep wavefronts contained in sound, because they have different types of suspension systems. So although they might agree well on steady tones, the mechanical differences make them respond differently on transients. On the other hand, VU meters were designed just for speech and music. The ballistics and electrical properties are now standardized so that any two VU meters will always read the same—on any type program material. The most important features of the standard are summarized in the box below.

Most important, though, is the fact that readings taken on VU meters correspond very closely with audible distortion in amplifiers, recorders, and so on. That is, the point of first audible distortion will occur at the same VU-meter indication regardless of the type of program material. Other indicating devices are not as good in this respect. On, say, a power meter, distortion might first be-

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Specifications for VU Meters

So that all VU meters will respond the same, the design has been standardized. Here are the most important specs from the American Standards Association C16.5-1942.

FREQUENCY RESPONSE: Flat to within 0.2 db from the 1000-cps value between 35 and 10,000 cps, and within 0.5 db between 25 and 16,000 cps.

IMPEDANCE: Internal resistance of the meter is 3900 ohms. For correct ballistics it must work from an external impedance of 3900 ohms.

BALLISTICS: Sudden application of a 0-VU level sine wave voltage will make the pointer reach 99% of 0-VU between 0.27 and 0.33 seconds, and will cause the pointer to overshoot to between 1.0 and 1.5%.

OVERLOAD CAPACITY: The VU meter can withstand peaks of 10 times 0-VU voltage for 0.5 seconds, and a continuous overload of 5 times 0-VU voltage.

SCALE: Two standard scales are specified of which scale A is pictured in Fig. 1. Scale B has the 0-100 markings on the top.

STANDARD CIRCUIT: A standard circuit, including an external attenuator, is specified for using the meter to read from a 600-ohm line.

HARMONIC DISTORTION: The VU meter will not introduce more than 0.3% harmonic distortion to a 600-ohm line when connected in the standard circuit.

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come noticeable at a certain indication for speech, at another indication for chamber music, and so on. With the VU meter, this difference is much less.

Steady-tone measurements can also be read on the VU meter, just as on any a.c. voltmeter. But the VU meter is the *best* meter to use for monitoring audio programs because it was designed for the job.

Of course such a good meter has disadvantages. First, it has a big magnet, and should be kept away from steel. Use aluminum or wood panels instead. And second, its sensitivity is low. With the recommended auxiliary equipment, it has a full-scale sensitivity of about 1.2 volts r.m.s with an internal resistance of 7500 ohms. This, as panel meters go, is mighty insensitive. While this is no problem in radio stations, large public-address systems, and the like, there is often in an audio system no point where the signal is sufficiently strong at the correct impedance to use the VU meter direct. There is no place on most home tape recorders, for example, where one can be attached.

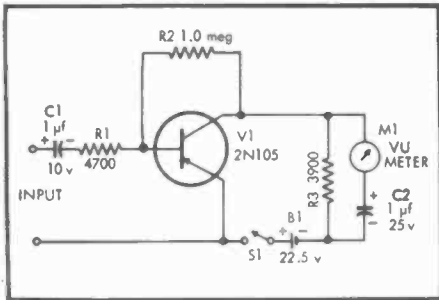An amplifier will boost its sensitivity, without losing any of the advantages of

Fig. 2. The simple transistor amplifier which converts the VU meter from an insensitive instrument to one that responds full scale on five microamperes signal current.

the meter. A transistor is used in the amplifier described here.

The Amplifier

No unusual parts are required for the VU meter amplifier. Any VU meter. any size or make, will do. The Triplett 327-TC VU meter comes in an attractive plastic case (*Fig.* 1) into which the amplifier can fit easily. The whole unit using this meter is made self-contained. Of course this is not necessary—if the finished meter is to be mounted on a panel, there is no point to the miniaturization techniques.

However, using the 327-TC the construction is quite simple. The parts fit around the meter proper as shown in *Fig.* 3. The only trick is providing for the input terminals, and an on-off switch. To do this, one can use a miniature Telex closed-circuit jack, #8570, and plug, #9231. The inner contact of the jack was bent as shown in the right of *Fig.* 4, so that it closes the circuit
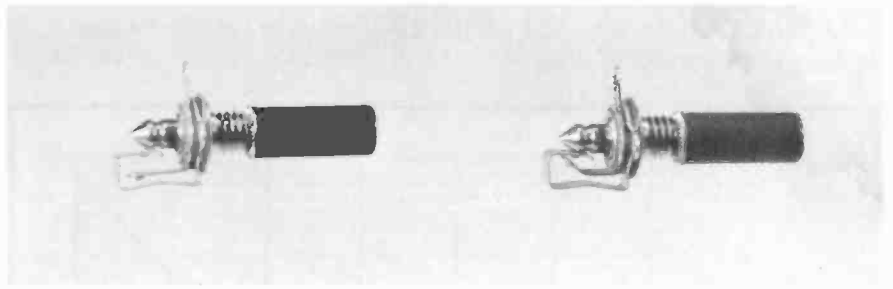
Fig. 4. Bending the inner contact of the jack changes it to a normally-open model.

when the plug is *in*, rather than when the plug is out. The jack fits perfectly in the hole in the Triplett case without any work whatever. The circuit was made to turn the device on by inserting the plug; with the plug out, no current is drawn from the battery.

Obviously other solutions to this problem are just as good—this one is offered merely as one example.

The transistor circuit is quite straightforward. For correct ballistics the meter must operate from a source of about 3900 ohms impedance. This is approximated quite closely by using a load resistor (R_3) of just that value.

Figure 2 shows the schematic. 22.5 volts is needed as a battery to provide sufficient swing for voltage peaks in the speech or music waveform.

The input resistor R_1 is set at 4700 ohms as a compromise between ultimate sensitivity and hum pickup. In practice an external resistance greater than this must be added anyway, to decrease the sensitivity, and to prevent excessive loading on the circuit tapped into, as will be explained later. There is nothing particularly significant about its value.

The base bias resistor supplies current to the base from the collector circuit in an attempt to improve the stability. It helps (as opposed to taking it directly from the battery) some 15 percent or so, although the device is still somewhat temperature-sensitive. However, the 2N105 transistor has quite a low cutoff current, as well as good uniformity from one transistor to the next. so there's less need for good d.c. stability measures.

The input capacitor C_1 merely blocks d.c. on the input circuit, and so any size capacitance which will pass the lowest frequency of interest may be used. If the meter is hooked to a circuit that has. for example, B+ on it, an additional high-voltage capacitance in series with the input will be necessary.

Connecting the Meter

The transistorized VU meter can be connected to the audio system in a variety of places—speaker lines, inside amplifiers, etc. Do *not* connect it across the recording head, however, since the supersonic bias is apt to get through as well as the program material, throwing the readings off.

Anywhere inside the amplifier will do —any high-level single-ended grid or plate, for example. If possible, use the grid of the phase splitter or the final stage. Use a plate only when necessary. however, since this means extra trouble in the form of a required capacitor in series with the input lead to block the d.c. component. The input capacitor C. has a low voltage rating, in the interest of making it small physically, and so is easily overloaded.

Generally a resistor will be required in series with the input. The complete device operates full-scale on five microamperes. Thus only in unusual circumstances will the device be too insensitive to use. The resistor necessary to draw only five microamperes for any given signal voltage, computed to take account of the input resistance of the amplifier, is shown in *Fig.* 5. Experimental points agree closely, within component tolerance errors.

To adjust the sensitivity of the VU meter, put some kind of program material on, or preferably a steady tone.
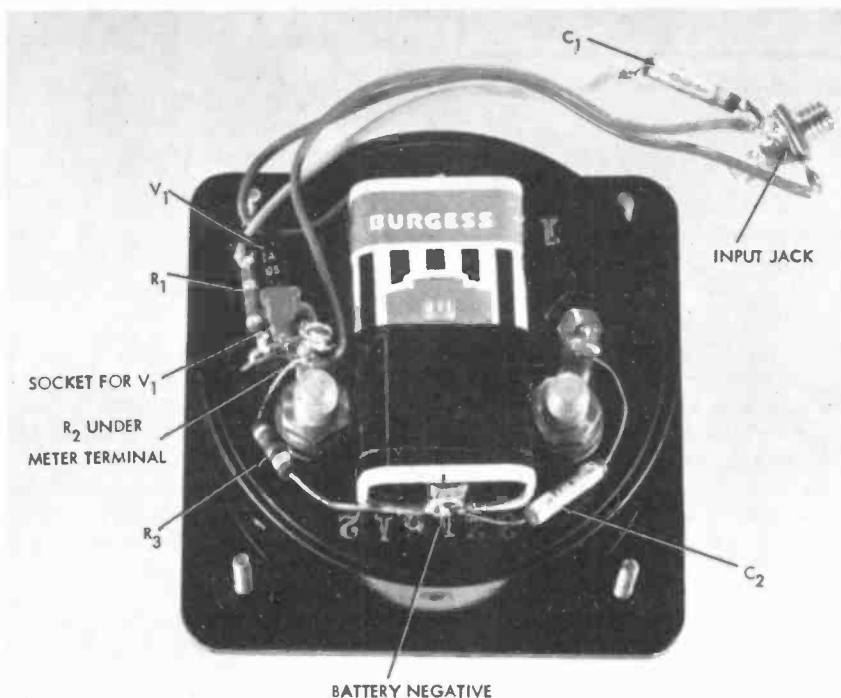

Fig. 3. The parts placement using the Triplett 327-TC meter and housing. The small battery and other parts fit inside the housing.

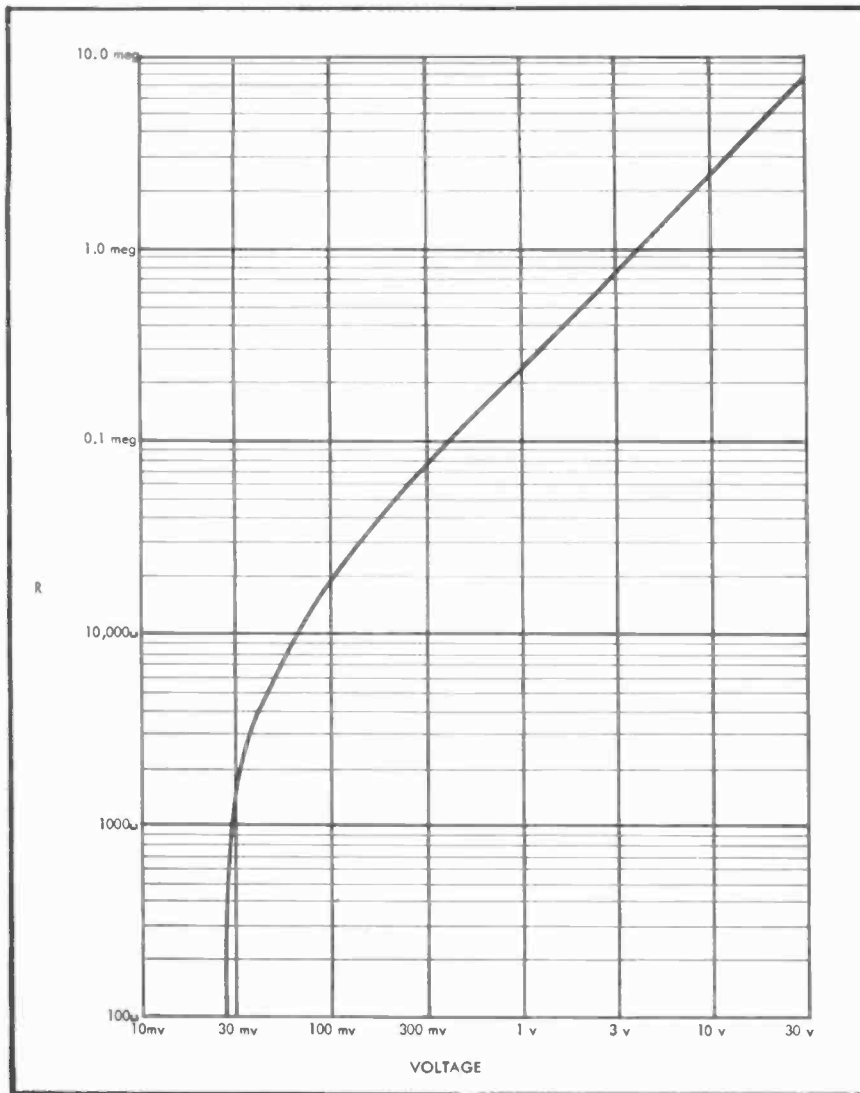# TRANSISTORIZED VU METER

(Continued from preceding page)

Fig. 5. Use this graph to find approximately how much series resistance is needed for any expected a.c. voltage.

and adjust the series input resistor until the desired sensitivity is achieved. A calibrated attenuator ahead of the meter may be useful in some cases where it is necessary to read a wide variation in volume.

There are many places where the transistorized VU meter can help out by providing an easy-to-read indication of volume with high sensitivity. Tape recording is the most obvious, but others include P.A. systems, high-fidelity rigs, and so on.

REFERENCES

O. Berliner, "Uses and abuses of the VU meter." AUDIO, Nov., 1955.
H. M. Tremaine, "Power level and volume indicator meters." RADIO & TELEVISION NEWS, Nov., 1950.
H. A. Chinn, "The measurement of audio volume." AUDIO ENGINEERING, September and October, 1951.

PARTS LIST

B_1 22.5 v. battery; Burgess U-15 or other.
C_1 1µf, 10 v. miniature electrolytic.
C_2 1µf, 25 v. miniature electrolytic.
M_1 VU Meter, any size, any make.
R_1 4700 ohms, ½ watt.
R_2 1 megohm, ½ watt.
R_3 3900 ohms, 5%, ½ watt.
S_1 On-off switch, s.p.s.t. May be incorporated as part of input jack.
V_1 RCA 2N105 Transistor.   ●

TONE CONTROL

(Continued from page 101)

one in which the flat position is obtainable over an appreciable range of the rotation to allow accurate setting of this important reference condition, and in which the turnover frequency varies exponentially with rotation. This would give equal db increments of boost or attenuation for equal increments of arc.

We may, then, set forth the requirements for the bass and treble potentiometers for ideal smoothness of operation. The desired curves of resistance *vs.* rotation for the bass and treble controls is shown in *Fig.* 9. It should be noted that rather wide deviations from this curve are permissible with very little degradation in smoothness. The linear taper used

in *Fig.* 2 for the bass control, for example, deviates quite widely from the desired characteristic, resulting in a wider flat position and some crowding of the tone control action toward the extreme ends of rotation of the potentiometer. The smoothness of action of this control, however, compares favorably with dual tone controls of most commercially available equipment.

The remaining circuits required for a complete control amplifier will appear in part two of this article, along with noise and distortion measurements and an analysis of the first stage of the circuit of *Fig.* 2.   Æ

Distortion in Tape Recording

Common sense, careful thinking, and a set of accurate measurements will enable anyone to choose an operating point, which will give the best over-all quality from his tape recorder. The author tells you how.

HERMAN BURSTEIN and HENRY C. POLLAK

MORE AND MORE audio fans, especially in areas having one or more "good music" FM stations, are making off-the-air tape recordings. Often the program source is live—symphony, chamber music, instrumentalist, singer, or choral group—while at other times the source consists of a first rate disc or tape recording. In either case, many owners of tape recorders have numerous opportunities to capture musical moments worth preserving, either indefinitely or until a better rendition comes along. Moreover, some recordists make tapes of their own singing or instrumental playing, which they are eager to hear for pleasure or improvement.

Unfortunately, the recording does not always sound "clean" in playback. It may lack the effortless, silky quality of the original source. Due to distortion, it may have a more or less grating quality, either constantly or only during loud passages. This situation is not confined to amateur recordings. Sometimes professional recordings contain objectionable distortion.

Distortion, presuming none in the source, may be due either to a fault in the tape recorder or to an excessive amount of signal applied to the tape. The latter is of concern here, that is, distortion resulting from high signal levels, and it shall be assumed that the tape recorder heads and electronics (amplifiers and bias oscillator) are in proper condition.

Although in a direct sense over-recording—that is, the desire for a high signal to noise ratio—may be blamed for distortion, in a basic sense the desire for wide frequency range, perhaps unnecessarily wide, may also be partly at fault. This can be true in two ways. First, in order to maintain good response out to 15,000 cps or so at a speed as low as 7.5 ips, the amount of high-frequency preemphasis required in recording may be sufficient to cause tape overload at treble frequencies. Above 7,500 cps, where most of the boost occurs, there would be virtually no audible harmonic distortion inasmuch as the harmonics fall outside most persons' hearing range as well as outside the recorder's pass

band, which cuts off sharply beyond 15,000 cps or earlier. However, in any nonlinear system there would still be intermodulation products generated by interaction between two high frequencies or between a low and a high frequency; many of these products would be within range of the ear and the recorder.

The desire for extended high-frequency response can also be responsible for distortion by virtue of the required bias setting. Over the bias range customarily used, an increase in bias generally causes distortion to fall, while a decrease in bias generally causes distortion to rise. However, increased bias also results in greater attenuation of high-frequency response. The desire to maintain high-frequency response well beyond 10,000 cps at low tape speed may lead to bias reduction, thereby resulting in greater distortion at a given recording level.

The following discussion seeks to throw light on:

1. The relative changes in harmonic and intermodulation distortion as input level is varied.
2. The relative changes in harmonic and intermodulation distortion as bias is varied; determination of bias for minimum distortion.
3. Variation among tapes with respect to intermodulation distortion.
4. Method of setting bias so as to yield the optimum combination of high signal-to-noise ratio, wide frequency range, and low distortion.

It should be made clear that the measurements described in the following discussion are not definitive in the sense of providing exact values under given recording conditions. Rather, they are broadly indicative of what happens. The values may fluctuate as the test is repeated at a different time, on a different machine, with a different tape, at different temperature or humidity, and so on. However, the tests have been repeated sufficiently to indicate reliably the general nature of the observed phenomena.

The measurements underlying the following discussion were made on two professional tape recorders in the $2,000 class, operating at 15 or 7.5 ips, and using a commercial high quality tape. The machines have separate record and play-

back heads, permitting immediate plotting of results. Test equipment consisted of an audio oscillator, an oscilloscope, a sensitive a.c. VTVM, a harmonic distortion tester which measures the total signal content after the fundamental has been filtered out, and an SMPTE type IM tester which, using 60 and 6,000 cps respectively in 4:1 ratio, measures the extent to which the high frequency is modulated by the low frequency.

Variation of Distortion With Input Level

Invariably, tape recorder specifications make no mention of IM distortion, referring only to harmonic distortion. Tape recorders have a VU meter or other type of recording level indicator to show when recording level is such as to produce 1 or 2 or 3 per cent harmonic distortion. However, as *Fig.* 1 reveals, when harmonic distortion is still at relatively innocuous levels, below 3 per cent or so, IM distortion can be disruptive—20 or 30 per cent or more.

The measurements in *Fig.* 1 were made on a machine operating at 15 ips with bias set approximately at optimum, in the manner described later. The 0 db reference input level for measuring IM distortion was equated to that for harmonic distortion by adjusting these input levels for equal peak-to-peak readings on an oscilloscope.

Figure 1 indicates that IM distortion begins to rise much earlier than harmonic distortion, and that the rate of increase is far greater for IM distortion. After IM distortion has reached about 4 or 5 per cent, it rises very precipitously. It may be observed, therefore, that in the effort to add a few db to signal-to-noise ratio, the recordist runs the risk of trading a slight decrease in noise for a large increase in IM distortion.

For the purposes of the measurements underlying this discussion, the recorder was adjusted so that its VU meter indicated 0 when IM distortion was approximately at the maximum level considered tolerable for high fidelity purposes, say about 2 or 3 per cent.

In actual use, however, the recorder should be adjusted so that the VU meter indicates 0 for a signal perhaps 8 or 10 db *below* that which causes maximum

allowable distortion, because on transients the pointer of the VU meter may lag 8 db or more behind peak signal level. If in actual use, the meter were calibrated to read 0 for a steady-state signal which produces 2 or 3 per cent IM distortion, allowing the needle to hit 0 when recording program material would often bring the reading into the region of extreme distortion, albeit briefly. Therefore, it is necessary to allow a margin in adjusting the VU meter. Even so, unless the recordist uses discretion, based on the nature of the music he is recording, fortissimo portions of a musical work, or at least the attacks, can be marred by the breakup and fuzziness symptomatic of distortion, even though the VU meter indicates only 0.

The recordist is forced into a choice among three alternatives: (1) to accept occasional high distortion in exchange for an improved signal-to-noise ratio; (2) to make some sacrifice in signal-to-noise ratio (which means relatively more hum, tube noise, and tape hiss) in exchange for low distortion throughout the recording; (3) to ride gain, reducing input level during loud passages, which means exchanging dynamic range for low distortion throughout a recording. The last alternative implies ability and willingness to compare the program source against a score and accurately anticipate changes in level.

The recordist's decision on the course to follow will be influenced by the tape recorder he is using and purposes for which it is employed. If it is a quality machine with a high signal-to-noise ratio, he may well follow the expedient yet satisfactory course of setting recording level just low enough so that peak passages are recorded at a level of distortion which, at least for a brief period, has no appreciable effect upon the listener. On the other hand, if the machine's signal-to-noise ratio is inferior, the preferable course may be to accept some obvious distortion during peaks for the sake of keeping background noise comfortably low throughout the recording. The program source can also influence the decision. For example, a relatively high input level might be used to record the spoken voice because in this instance a considerable amount of distortion can usually go unnoticed. On the other hand, one might have to exercise considerable more restraint in setting gain for an organ or piano in order to obtain a pleasing similarity to the original.

Variation of Distortion With Bias

Figure 2 indicates the effect of bias current on distortion, using two relatively high input levels. It must be taken into account that as bias varies so does the amount of signal recorded on the tape. In short, tape output as well as distortion varies with bias. However, we are only interested here in how distortion varies with bias. Therefore it is necessary to hold tape output constant. For this reason, the input level was constantly adjusted to maintain a fixed indication on the VU meter *in playback*. Curves 1 and 2 are based on a playback indication of 0 db on the VU meter. Curves 3 and 4 result from levels 3 db higher. At the 0 VU playback level, with bias set for minimum IM distortion, the harmonic distortion test signal was matched to the IM test signal by comparing peak-to-peak playback amplitudes on the oscilloscope.

Figure 2 reveals that: (1) IM distortion once again varies much more than harmonic distortion; (2) Distortion does not indefinitely continue to decline as bias is increased, but rises again, and this rise is sharper in the case of IM distortion; (3) The higher the input level, the more critical is the bias setting for minimum distortion; thus, in order to find the minimum-distortion bias with ease, it is merely necessary to use a very high input level. (4) A rise in input signal level produces the least increase in distortion when bias is set for minimum distortion.

From the above it can be concluded that to the extent the recordist seeks to maximize signal-to-noise ratio by turning up gain, the more important it becomes that he adjust bias properly for the particular tape he is using. Otherwise he may get much more distortion, especially IM, than is acceptable.

(An interesting phenomenon is displayed by the left portion of the curves in *Fig.* 2. If bias current is reduced enough below the normal working range, distortion drops again. Inasmuch as a reduction in bias current serves to improve high frequency response, it might seem that one might profitably operate in the area of extremely low bias current. However, there is good reason for not doing so. The reduction in distortion achieved by using very low bias current is most striking for high input levels. At low input levels, however, distortion re-
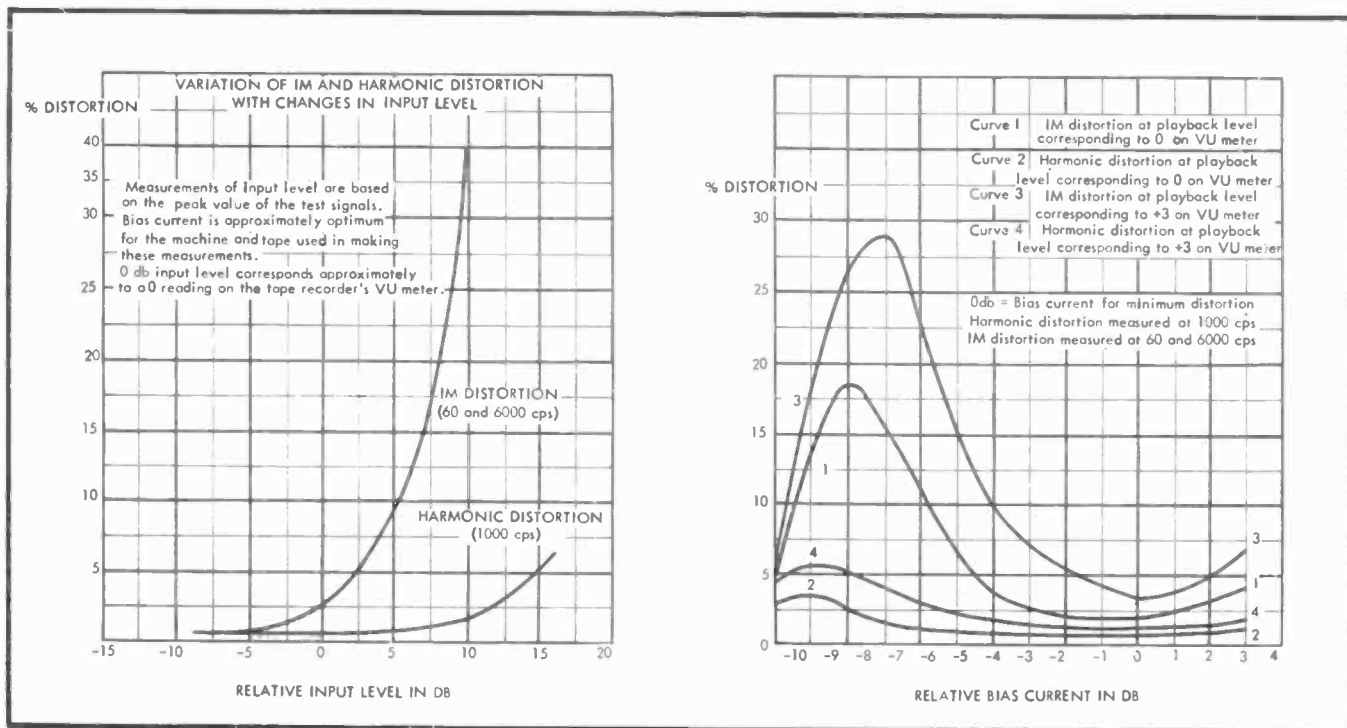

Fig. 1 (left). Variation of IM and harmonic distortion with changes in input level. Fig. 2 (right). Variation of IM and harmonic distortion with changes in bias current.

mains higher than when operating in the normal bias range. Furthermore, the amount of recorded signal drops at low bias values, so that to maintain the same amount of tape output requires considerably greater power from the output stage supplying the record head.)

Variation in Distortion Among Tapes

Using a relatively high recording signal, several popular brands of tape were compared with respect to IM distortion. Input level was varied so that each tape produced the same output level as read on the VU meter during playback. Bias was adjusted for each tape until minimum distortion was obtained. Following were the results.

Tape	Minimum IM Distortion	Relative Bias Setting
A (reference)	7.6%	0.00 db
B	9.0	.75
C	11.0	− .50
D	10.0	0.00
E	3.5	−1.00

It is interesting to note that the bias setting for minimum distortion varied only moderately from tape to tape, while the amount of distortion varied considerably more. However, these findings would not be sufficient on which to base the choice of a tape. It would be further necessary to consider the tape's frequency characteristics at the bias current resulting in minimum distortion, the *shape* of its output versus bias curves for different frequencies, its noise properties, and so on.

Determination of Optimum Bias Current

Let us assume that on the basis of curves such as in *Fig.* 2, the bias current for minimum distortion has been ascertained, using a given machine and a particular tape. However, depending upon the tape speed and upon the brand and kind of tape (regular, high output, longplay, etc.), high frequency response may be inadequate at this bias current.

As previously stated, treble response goes down as bias is increased. This is a wavelength effect. Inasmuch as a given frequency results in a shorter wavelength at reduced tape speed, the problem of poor treble response due to high bias current is most serious at the lower speeds such as 7.5 and 3.75 ips. Consequently at these speeds, in order to maintain satisfactory response, it is probably necessary to use less bias than the amount permitting minimum distortion. This means greater distortion for a given amount of tape output, or less output for the same distortion (lower signal-to-noise ratio), or a compromise between the two.

Figure 3 indicates the procedure to be used in determining optimum bias current. It is assumed that the tape recorder provides ready means for varying bias current and for varying treble preemphasis in recording. It is further assumed that playback equalization is fixed (in accordance with the NARTB standard for 15 ips). Curves 1 and 2 in *Fig.* 3, representing variation of IM distortion with bias, have been redrawn from *Fig.* 2. 0 db bias represents bias current for least distortion.

When the tape recorder represented in *Fig.* 3 is operating at 15 ips, Curves 3 and 4 respectively show how response at 400 cps and at 15,000 cps varies with bias; input level was kept low enough to avoid any possibility of saturation. 400 cps is used as a reference frequency, not being affected by equalization used in the record preamplifier. When 0 db (minimum distortion) bias current is used, response at 15,000 cps is 1.5 db higher than at 400 cps. In order for frequency response to be perfectly flat at 15,000 cps, it is necessary either to increase the amount of bias current to 1.4 db or reduce the amount of treble preemphasis. Since a rise in bias current would increase distortion, the desirable step is to lower the treble boost.

Thus it can be seen that at a speed as high as 15 ips, at least for the machine and tape represented in *Fig.* 3, one can set bias for minimum distortion and yet maintain response out to 15,000 cps. (It should be noted that a final determination of the amount of treble preemphasis required would depend upon a frequency-response run. Possibly, if response at 15,000 cps is kept flat, there would be excessive boost at lower treble frequencies. Thus in order to achieve the flattest possible response over the treble range as a whole, it may be necessary to accept response which is a few db down at 15,000 cps.)

Now let us consider the situation where the tape recorder represented in *Fig.* 3 operates at 7.5 ips. Curve 5 shows the 15,000-cps response at 7.5 ips as bias is varied. At minimum distortion bias, 15,000 cps response is about 10 db below 400 cps. Possibly this situation can be improved by increasing the amount of treble boost in the record amplifier. On the other hand, increasing the treble boost may cause appreciably greater tape overload in the upper treble range. Let us therefore assume that Curve 5 is based on the maximum amount of treble boost which may be safely used, taking into account the typical distribution of musical energy over the frequency range;[1] any additional treble boost would increase the likelihood of distortion.

Consequently, in order to maintain response out to 15,000 cycles at 7.5 ips, it is necessary to reduce bias. Curves 3 and 5 intersect at approximately −3.6 db bias; at this reduced bias, flat response out to 15,000 cps can be had. However, as bias is reduced to −3.6 db, IM distortion rises from 3.5 to 8.5 per cent for the signal level represented by Curve 1. On the other hand, by sacrificing 3 db in signal-to-noise ratio—that is, reducing signal level to the proportions represented by Curve 2—IM distortion can be kept at only 3 per cent when bias is −3.6 db.

[1] See the article by Herman Burstein, "Tape Recording Equalization," *Radio & Television News*, February 1956.

Fig. 3. Determination of optimum bias current.
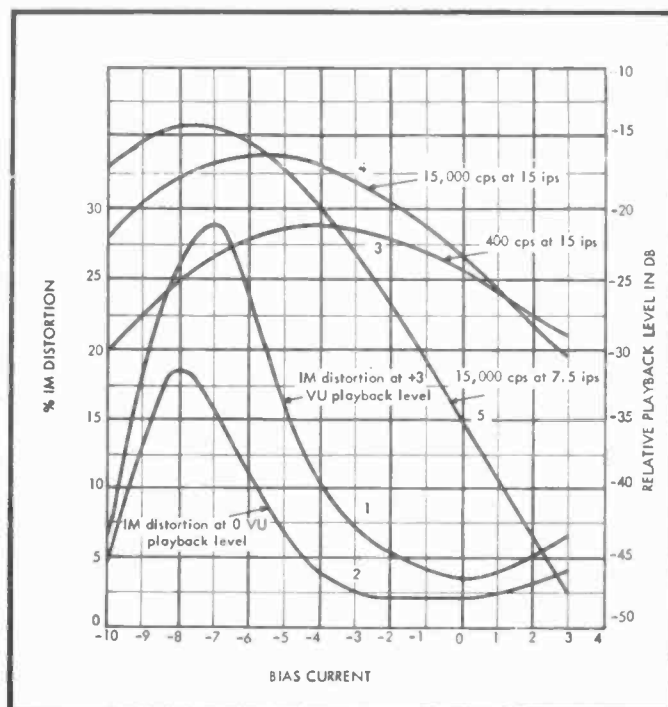
A Time-Delay Commercial Suppressor

RONALD L. IVES

One method of solving the age-old problem of cutting out the commercials while leaving the musical program untouched. This may be killing the goose that lays the golden eggs, since it is the commercial that makes the music possible, but so long as only a small percentage of the listeners practice "commercial killing" the broadcasting industry is not likely to fall apart completely.

BACK IN THE "good old days" of radio broadcasting, announcements were short, simple, and terse, such as "This is WAAM, the I. R. Nelson Co., One Bond St., Newark, N. J." Today, the *vox locutor* has degenerated into the *vox loquacitor*, and programs of classical music are too often interrupted to bring us the rasping tape-recorded voice of Madame Perra, who this week is plugging Barrelform Girdles. These, we learn while getting up to shut the blasted thing off, come in all sizes, from 38 "petite" to 80 "grande dame."

The problem of passing instrumental music, while squelching unwanted chin music and singing commercials is quite complicated. One device, consisting of a flashlight and a photoelectric control on the receiver, is quite effective on TV, but is not of much use on an ordinary broadcast receiver, as there is no clue to the end of the announcer's gabfest. Shutting the receiver off when Abdul the Necromancer starts plugging used Volks-

wagens is very effective, but on forgetting to turn the receiver on again, we miss the wanted time, weather, and news announcements.

An electric discriminator, sometimes called the VOCK (voice-operated commercial killer), has appeared in several different forms, and can be so adjusted that it will distinguish between most music and most voices. It can be set to differentiate between Walter Winchell and Gabriel Heatter, but cannot distinguish between "HMS Pinafore" and the "Burpy-Cola Song." It also takes the surprise out of the *Surprise Symphony,* and develops acute schizophrenia with piano solos. In addition, it eliminates both the wanted announcements and the lectures on how to borrow your way out of debt with the Stoneheart Finance Company. Additional problems are introduced by the differences between languages; a VOCK cannot be made to work as well for Spanish as for English, and fails completely for Burmese, Bantu, and Yucateca.
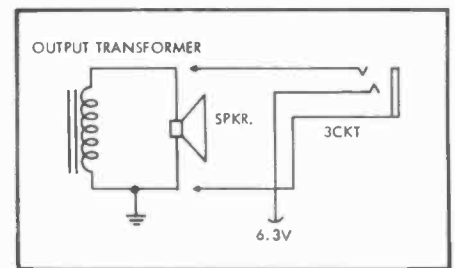
Fig. 2. Receiver connections for time-delay suppressor.

Some study of the nature of commercials, made with a radio receiver, a stopwatch, and some blood-pressure medicine, shows that the length of the commercial at most stations is fairly constant, and seldom exceeds 80 seconds. Thus, a device which will shut off the audio output of the receiver for 80 seconds plus a margin of safety, will eliminate the major part of most commercials. Such a device is easily made from a self-holding relay, a push button, and a time-delay unit, necessary power being supplied by the receiver. The circuit of such a time-delay commercial suppressor is shown in *Fig. 1* and the method of connecting this device to the receiver is shown in *Fig. 2*.

How it Works

In operation, when the push button is pressed the relay closes, shorting the speaker terminals, and also shorting the push button through the contacts of the time-delay relay, which are normally closed. The heater of the time-delay relay, being in shunt to the coil type relay, begins to heat. After a finite time interval (here 90 seconds), the time-delay contacts open, opening the current supply to the relay, which opens and de-energizes the time-delay unit, removing the short from the speaker. The switch is available for use when the announcer is extremely longwinded. Any
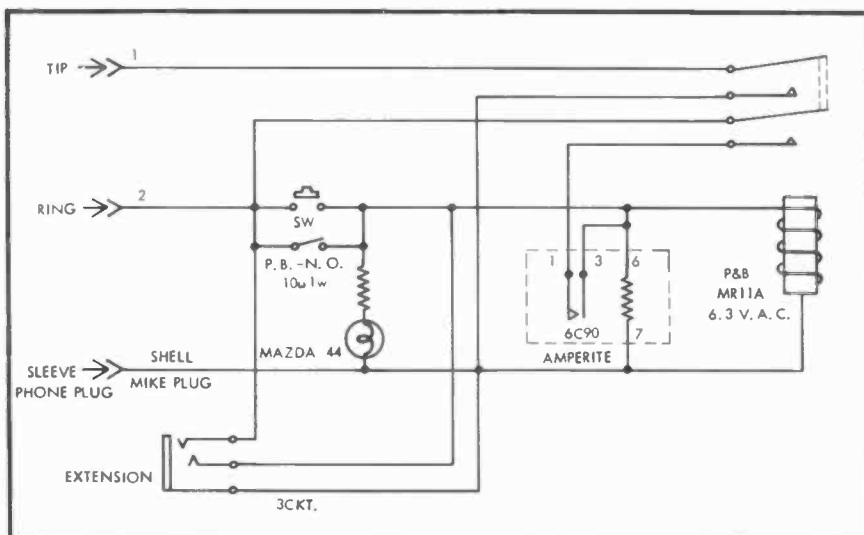

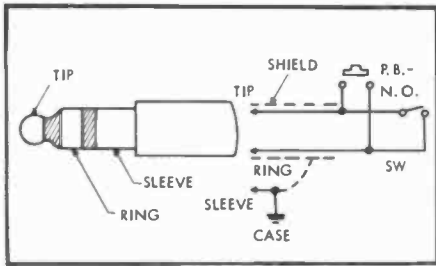Fig. 1. Schematic of time-delay commercial suppressor.

Fig. 3. Schematic of extension control.

number of extensions may be used, and extension lead length may be as much as 50 feet, if desired. The circuit of the extension is shown in *Fig. 3*. Shielded cable may be necessary from the extension to the main control box if high sensitivity a. f. equipment is used in its vicinity.

The general appearance of the time-delay commercial suppressor is shown at the right side of *Fig. 4*; the extension, designed to mount on the wall next to the telephone is shown at its left. Several alternate designs are possible, and present no technical problems, there being no high frequencies or critical adjustments involved in its construction.

In an experimental form assembled by the author, the housing for the main suppressor is a 4 × 6 × 2 in. SeeZak expandable chassis, with the corner bosses replaced by machine screws, as are the sheet metal screws holding the top and bottom plates in place. The main cord

connection, on the rear, *Fig. 5*, is an Amphenol three-terminal mike fitting, terminal 3 being connected to the shell. The other end of the connecting cord, which is a two-wire shielded microphone cable, terminates in a three-circuit plug (Switchcraft #267), which engages the receiver jack (*Fig. 2*). The front jack, for the extension, is a narrow-gauge type (Switchcraft JJ-033) to engage a like three-circuit plug (Switchcraft PJ-068). The use of unlike connectors here is intentional, as components are not interchangeable.

The case for the extension is an aluminum box chassis 1⅜″ by 1⅜″ by 2¼″ (IMB-MOO). The connecting cord is run through a grommet on one end, and the shield is anchored to a soldering lug held under the switch.

Internal construction of the main control is shown in *Fig. 6*. The bracket for mounting the time-delay-unit socket is made from part of a SeeZak rail; the pilot light is mounted in a Dialco #705 socket, with the web bent at right angles and soldered to a threaded spacer. The bulb passes through a grommet in the top plate, and a screw through this top plate holds the socket in place. No special care, other than good workmanship, is needed in assembly.

Although not electrically necessary, a bottom plate on the case is most desirable to keep grubby fingers out of the mechanism, and to permit mounting four rubber feet on the bottom. These prevent skidding and scuffing. The bottom
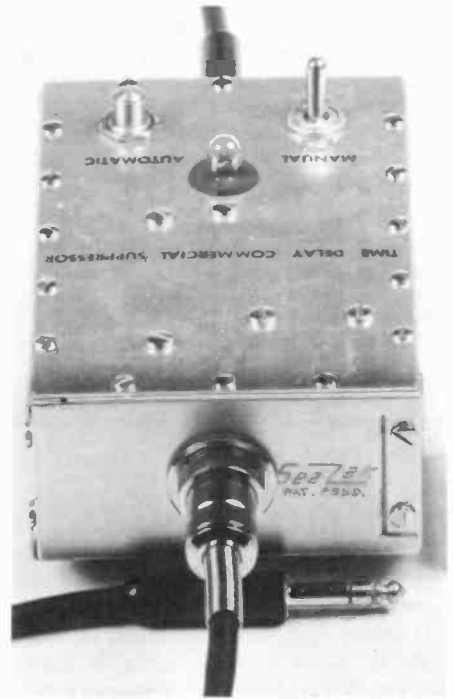
Fig. 5. Rear view of suppressor chassis shows use of Amphenol plug for connection from amplifier.

screws are 4-40 binding heads, tapped into the case rails.

In actual use, the radio is left on until the announcer winds up with "—and now, friends, we bring you—." Press the button, and enjoy 90 seconds of blessed silence. If the announcer talks overtime, you may catch the final "—Del Rio, Texas, and don't forget to put the dollar
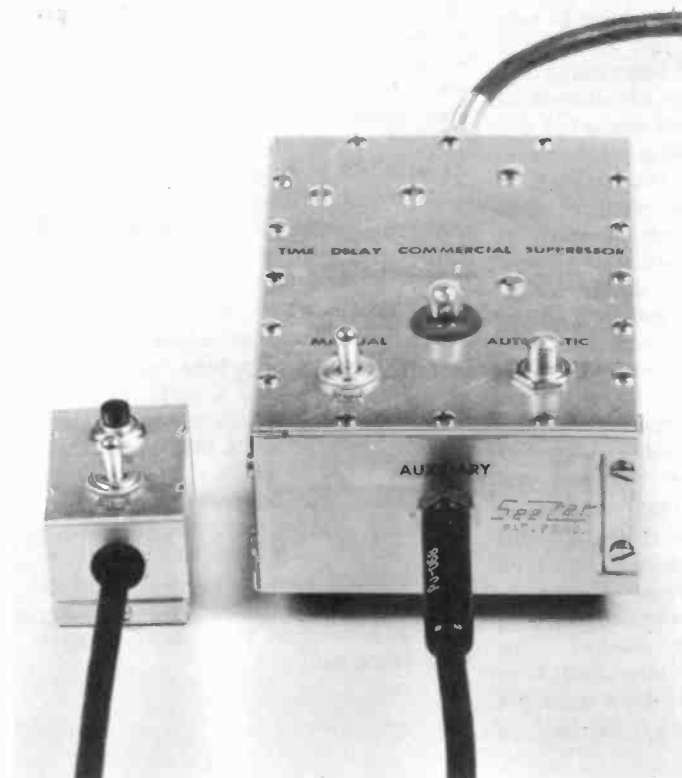
Fig. 4. External appearance of commercial suppressor (right) and the extension control unit.
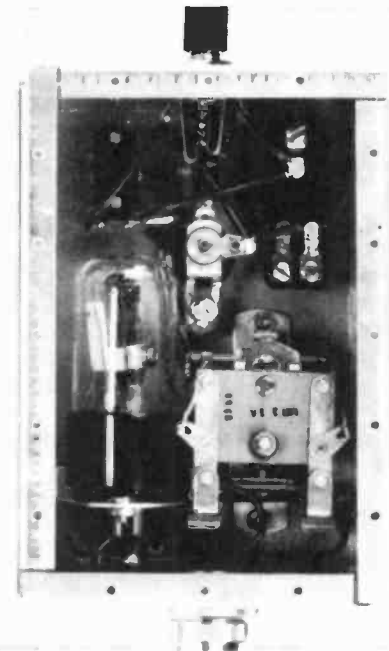

Fig. 6. Interior view of main control chassis. Amperite time-delay unit is the tube-like device at lower left.

in the envelope"; but the intervening vocabulation, extolling the supposed merits of "real, genuwine, simulated diamonds," has been harmlessly dissipated as millicalories of heat, instead of raising your blood pressure by millimeters of mercury.

Parts List

Main Control
1 SeeZak expandable chassis, 4 × 6 × 2, with bottom plate
4 ⅞" rubber feet, with mounting screws

1 Amphenol 91MC3F female cable connector
1 Potter & Brumfield MR-11A relay, DPDT, 6.3 v. a.c. coil.
1 10-ohm 1-watt carbon resistor
1 Dialco 705 socket
1 #44 pilot lamp
1 Amperite 6C90 time-delay unit
1 Switchcraft JJ-033 jack, 3-cct.
1 Toggle switch, SPST
1 Unimax MXJ-1 push button, SPDT

Main Cord
1 Amphenol 91MC3M cable connector
1 Switchcraft 267 phone plug

Desired length of 2-wire shielded mike cable (limit 50 ft.), Belden 8414

Extension
1 Aluminum box chassis, 1⅜ × 1⅜ × 2¼ in.
1 Switchcraft PJ-068 phone plug
1 Rubber grommet, to fit mike cable
1 Switchcraft 201 push button, SPST, N. O.
Desired length of 2-wire shielded mike cable (limit 50 ft.), Belden 8414

TAPE DISTORTION

It would seem that a reduction of only 3 db in signal-to-noise ratio is little enough to exchange for frequency response good to 15,000 instead of 7,500 cps. However, there are two counter views: (1) Few if any tape recorders have decibels to spare in the matter of signal-to-noise ratio. Whereas ratios of 70 db, 80 db, and better are commonly found in preamplifiers and power amplifiers, a tape recorder is doing extremely well if it gets up to 55 db. The designer of such a tape recorder fights hard for every last decibel or two in striving for a figure of 55 db, and a sacrifice of 3 db is consequently not unimportant. (2) Operating at −3.6 db bias puts the tape recording process into a region where a slight miscalculation as to input level produces a large difference in IM distortion. On the basis of *Fig.* 2 (or 3) at 0 db bias a 3 db miscalculation in level increases IM distortion only 1.5 per cent, but at −3.6 db bias the same miscalculation raises distortion by 5.5 per cent.

In view of the above two considerations, a recordist or tape machine designer equipped with the necessary test instruments might decide that at 7.5 ips he cannot afford, in terms of distortion and/or noise, the luxury of response more or less flat to 15,000 cps. Instead he may decide on a compromise course, shifting to a bias current intermediate between 0 and −3.6 db. Thus, for example, his choice might cost him only a 1 db reduction in signal-to-noise ratio and a reduction in flat response from 15,000 cps to 10,000 or 12,000 cps. At the same time he would have better protection against the consequences of overrecording than if he used −3.6 db bias.

In order to find this optimum bias point, it would be necessary to draw a number of curves similar to Curve 5 in Figure 3, showing the effect of bias current variations on several frequencies

such as 9,000, 10,000, 12,000 cps, and so on. Input level should be kept 20 to 30 db below maximum recording level to avoid saturation. Then for each frequency curve one can evaluate, along the lines indicated in Fig. 3, what flat response out to this frequency signifies in terms of increase in distortion and/or reduction in signal-to-noise ratio because of departure from 0 db current. Based on these evaluations, the bias current can be selected which reflects the individual's concept of the optimum combination of frequency response, distortion, and signal-to-noise ratio within the capacities of a particular machine.

Conclusions

It has been pointed out that IM distortion can be a serious problem in tape recording, especially if one attempts to cut close to the line in maximizing signal-to-noise ratio; that adjustment of bias current can be quite critical if distortion is to be kept to a minimum at high recording levels; that departures from this critical bias point can exaggerate the consequences of excessive recording levels; and that, if the necessary test equipment is available, a definite procedure can be followed to determine first the bias current for minimum distortion and secondly the bias current which at speeds below 15 ips provides the most satisfactory compromise among the requirements of low distortion, wide frequency response, and high signal-to-noise ratio.

A number of judgments are required in determining maximum recording level and optimum bias current. How wide need frequency range be in order to give essentially satisfactory results? How much IM distortion is tolerable? How much for a split second? How much for a few seconds? How much for half an hour?

These of course are subjective judg-

ments. Consequently the determination of maximum recording level and optimum bias current is not a hard and fast procedure.

The writers have heard a number of professional master tapes, one or two generations removed from the original, which, according to indications of a properly calibrated VU meter in playback, were recorded at excessively high levels; the VU pointer frequently kicked to full scale instead of staying below 0. Yet many of these seemingly over-recorded tapes nevertheless sounded clean to the ear. Although IM distortion was undoubtedly present in substantial degree, perhaps it was occurring in such short bursts as not to be disturbing; or perhaps the nature of the musical selection was such as to mask the effects of distortion. On the other hand, the writers have listened to master tapes seemingly recorded at conservative levels, yet less clean-sounding than desirable. Possibly other factors than recording level and bias setting intervened between the original source and good reproduction. At still other times the writers have listened to recordings velvety smooth except for a relatively high background of hum, noise, and tape hiss. They would gladly have accepted more distortion for less background distraction.

The above observations point up the fact that top quality tape recording is both a technique and a craft. It is advisable to have a technical grasp which enables one to adjust a tape recorder, if feasible, so as to make the most of its capabilities with respect to distortion, frequency range, and signal-to-noise ratio. At the same time, one must have the craftsman's touch, which is based on experience, qualitative judgment, and—the best instrument of all in audio work—an acute ear.

Baffles Unbaffled

E. J. JORDAN

Part 1. The author presents a thorough discussion of the design, construction, and performance of the various methods of mounting loudspeakers, ranging from the flat baffle to the most elaborate enclosures.

JUDGING FROM the number and nature of queries that have been received regarding loudspeaker enclosures, it has been realized that a great deal of confusion and misconception exists. This is not surprising when one considers the presentation of much of the information on this subject. We have, on the one hand, the highly technical discourses coupled with the advanced mathematical treatment that is necessary for the basic understanding of acoustics, and, on the other, empirical and often inaccurate information which, being more readily absorbed by the reader, receives wide publication and when one asks "what type of enclosure shall I use?" one is immediately confronted with a diversity of answers governed often by personal prejudice and sales talk.

An attempt is being made in this paper to clarify this position. In due course, all the principal types of enclosure will be discussed, leading finally to the description of new developments
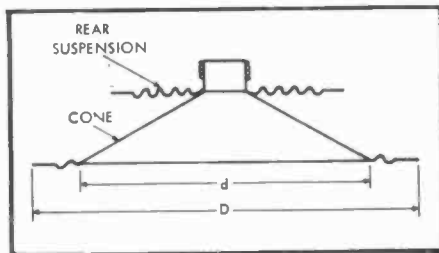
Fig. 1. Typical loudspeaker cross section. D = speaker diameter; d = piston diameter of cone; $\pi d^2/4$ = piston area.

that have been made. The aim throughout is to provide a good fundamental outline of the principles involved in each type of enclosure and to discuss fully their merits and demerits, these being demonstrated, where possible, by impedance curves, and to include the necessary formula and guidance for the home constructor to enable him to determine readily which type of enclosure is most suited to his particular requirements, and to design and build such an enclosure.
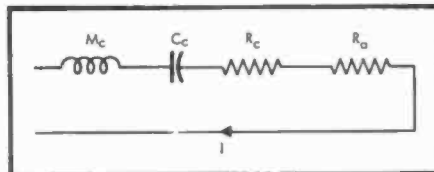
Fig. 2. Loudspeaker analogy. R_c = friction loss in cone; M_c = mass of cone; C_c = compliance of suspension; R_a = radiation resistance.

So that we may form a foundation for future discussion, the following elementary facts are set out.

A loudspeaker cone, when reproducing those frequencies where the wavelength is large compared to the cone diameter, behaves approximately as a rigid diaphragm of area πr^2 where r is the radius of the cone base. (*Fig.* 1). This is known as the piston area of the cone. If the cone is vibrating in free air each surface of the cone will cause alternate compressions and rarefactions in the adjacent air layers which will in turn be radiated as sound waves. Where the distances between surfaces is short compared with a wave length, the air compressed on one surface will flow into the rarefaction occurring simultaneously on the other, and the total radiated sound energy will be negligible. A loudspeaker cone under these conditions can be regarded from some points of view as analogous to an electrical circuit which takes its simplest form as in *Fig.* 2.

This is a series resonant circuit where the current in the circuit is analogous to the cone velocity, and the power developed in R_a is analogous to the radiated sound power—and it will be seen that for a given value of R_a this will reach its maximum at the resonance of M_c and C_c, i.e. the bass resonance of the loudspeaker where the Impedance $Z = R_c + R_a + j \left(W \ M_c - \dfrac{1}{W \ C_c} \right)$ simplifies to $Z = R_c + R_a$ and the current rises to a high value limited only by these resistances. Below this frequency the reaction of C_c rises rapidly and the current through R_a falls correspondingly. If the loudspeaker is unbaffled it takes very little energy just to push air from the front to the back and vice versa,

and we say that R_a is low. Consequently the radiated power is low, although at resonance the velocity and amplitude of the cone will be very high. (*Fig.* 4).

It will be seen throughout this paper that velocity characteristics are indicated by voice-coil impedance curves. This is for convenience and is justified if the base of the velocity scale is represented by a line drawn at that impedance presented by the voice coil if the voice coil were clamped. This is shown by the following :–

Z_t = Total impedance of voice coil

Z_m = Motional impedance

Z_c = Clamped impedance of voice coil (d.c. resistance at low frequencies)

v = Velocity

Now $Z_m \approx v$

$Z_m = Z_t - Z_c$

$Z_t - Z_c \approx v$

At the frequencies we are considering

$Z_c \doteq$ d.c. resistance $= 12$

$Z_t - 12 \approx v$

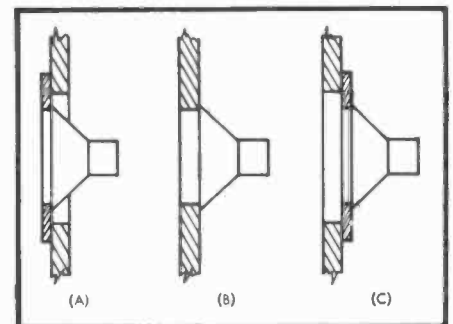Then $v = 0$ if we draw the velocity base line at $Z_t = 12$

Fig. 3. Common methods of mounting loudspeakers in a wall.

The velocity curve is therefore an indication of the radiation from the cone. It is not necessarily an indication of the total radiation from the combined cabinet and loudspeaker system, but is nevertheless very useful in determining the action of an enclosure and in this respect is more useful than pressure response curve, since these vary greatly with microphone position. It is for these reasons that velocity

curves have been used throughout.

We have seen that to secure good radiation at the low frequencies, the radiation from the rear of the cone must be prevented from cancelling that from the front, and we shall discuss the various means of doing this.

The Flat Baffle

This is the simplest method of loudspeaker mounting. Ideally the dimensions of the baffle should be infinite, but a very close approximation to this is achieved by mounting the loudspeaker in a wall, e.g. the partition wall between two rooms. This method is used quite frequently and the following points should be borne in mind: (1) It is rather important to mount the loudspeaker on a sub-baffle as shown at (A) in *Fig.* 3. The sub-baffle should be of substantial wood or chipboard ¾ to 1 in. thick. With the loudspeaker mounted as in (B) or (C) it will be seen that an air column is set up in front of the cone, the length of which is equal to the thickness of the wall. This air column will have a natural resonant frequency. If for example the wall is 2½ in. thick then this frequency will be approximately 968.4 cps. As a general rule the thickness of the baffle should not exceed one-tenth of the diameter of the opening, unless it is bevelled.

(2) The position on the wall also calls for some consideration. It is preferable to mount the speaker near a corner since this increases the air loading on the cone and improves the bass radiation. The reason for this will be

Fig. 4. Voice-coil impedance of loudspeaker in free air (dotted curve) and mounted in wall (solid curve).

seen later. The height of the speaker above the floor is largely a matter of personal taste with regard to the high-frequency distribution; e.g. if the speaker is placed near the floor there may be excessive absorption of the high frequencies due to furniture, carpets, and so on, although the low frequencies will have the advantage of, effectively, a corner position. It is usually preferred to have the speaker at ear level when one is seated in the normal position. Often it is better to sit slightly off axis, especially when listening to orchestral music, since the inevitable beaming of the high frequencies, however slight, cause one to be unduly aware of the point source
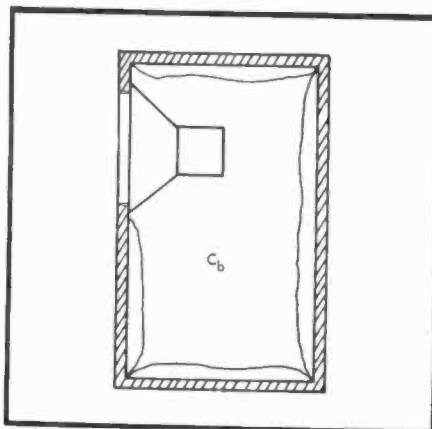
Fig. 5. Cross section of closed box.

which may spoil the realism of the reproduction.

It will be appreciated that much of the foregoing will apply to all forms of speaker mounting.

Small Flat Baffles

Often wall mounting is not possible and relatively small rigid baffles are used. In general, the back-to-front cancellation occurring with an unbaffled speaker occurs also with a small baffle except at a lower frequency.

Considering firstly a circular baffle with the speaker mounted centrally, then the minimum radius r required to maintain the radiation down to a given frequency f is given by :-

$$r = \frac{\sqrt{3\lambda}}{2\pi} \text{ or } r = \frac{311}{f} \text{ ft.}$$

Where λ = wavelength at frequency f.

There is no point in making f lower than the bass resonance of the loudspeaker since, as we have seen, below this frequency the radiation falls rapidly due to the increase in the stiffness reactance (*Fig.* 2). The use of a circular baffle, however, is not recommended since the path length from front to back of the loudspeaker is the same in all directions and standing waves
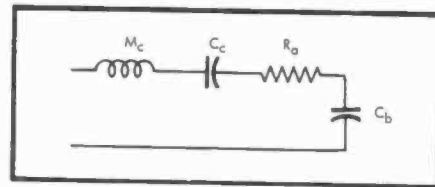

Fig. 6. Analogy of the closed box mounting.

will be set up causing a series of peaks and troughs in the loudspeaker frequency response. The preferred shape of a small baffle is an irregular one, e.g. a square with the loudspeaker mounted off-center, in which case the minimum dimensions should be determined as for the circular baffle where r = the distance from the centre of the speaker to the nearest edge. More usual, however, is the rectangular or square baffle, and the minimum dimensions here are given by :-

$$l = \frac{\lambda}{2} \text{ or } \frac{565}{f} \text{ ft.}$$

where l = the length of the smallest side of the rectangle or the side of the square.

If, for example, we have a loudspeaker with a bass resonance at 60 cps and we require to mount the speaker on a square baffle, the optimum length of the side is given by $l = \frac{565}{60}$ or 9.42 ft. Such baffles must be rigid and if made of wood this should be of thickness not less than one tenth of the baffle hole diameter.

The analogous electrical circuit for a loudspeaker mounted on an infinite or finite baffle of optimum size is similar to that shown for a loudspeaker in free air (*Fig.* 2) except that the baffle produces a large increase in R_a and a small increase in L_a. We should therefore expect to find a corresponding decrease in the cone velocity and the resonant frequency. This is shown in *Fig.* 4.

It is sometimes not realized that small loudspeakers acting as treble units should be mounted on a baffle large enough to ensure full radiation down to about half the crossover frequencies. This is necessary since the crossover does not occur sharply, but there is some over-lap.

If, for example, the crossover is at a 1,000 cps, then the baffle size for the treble unit should be $\frac{565 \times 2 \times 12}{1000} = 13.6$ in. square. If the baffle is rectangular, this should be its smaller dimension.

Considering again the low frequencies from the point of view of the commercial set manufacturer, the optimum baffle size is very often far too large even when frequencies as high as 100 cps are the lower limit and a flat baffle of any size or shape has little aesthetic
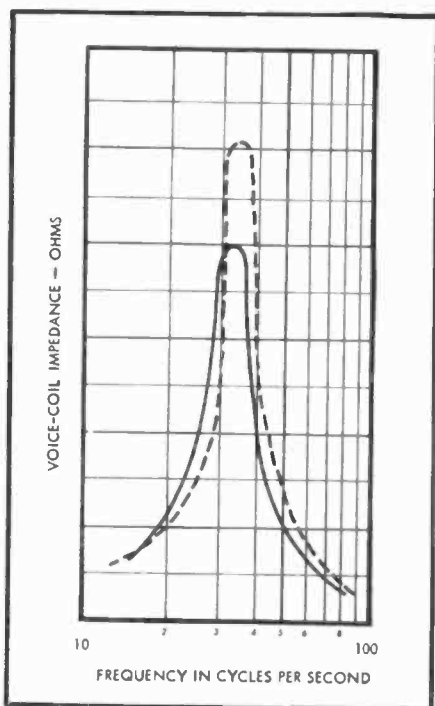
SPEAKER	BASS RESONANCE — CPS					
	Off Baffle	1 cu. ft.	2 cu. ft.	3 cu ft	4 cu. ft	5 cu ft.
15"	25	105	85	77	70	65
12"	35	116	85	75	70	65
12"	55	120	95	85	78	72
10"	52	105	82	74	70	68
9"	25	116	77	65*		
8"	75	108	92	85	80	78

* The "Q" of this resonance become too low to measure beyond here

Fig. 7. Typical resonant frequencies of various sizes of loudspeakers in closed boxes with volumes from 1 to 5 cu. ft.

appeal, so the baffle takes the form of an open backed cabinet.

The limiting frequency in this case may be calculated using the formula for circular baffles where r is the distance from the center of the loudspeaker cone to the nearest point on the rear boundary of the enclosure; the sides and top forming part of the baffle.

A cabinet of this nature, however, has an air column resonance, the frequency of which approximates to

$$f_1 = \frac{6780}{l + .85R} \text{ cps}$$

where l = depth of cabinet

$$R = \sqrt{\frac{A}{\pi}}$$

A = area of open back

when all dimensions are expressed in inches. The above expression will be discussed more fully in a section devoted to tuned pipes.

In addition to f_1 there will be a number of harmonically related resonances f_2, f_3, f_4, etc. where $f_n = nf_1$, n being any whole number. These resonances give rise to the unnatural booming quality that is characteristic of many commercial receivers. Often this introduces an artificial bass which compensates for the fall in true bass due to the insufficient baffle area.

To achieve this, the bass resonance of the loudspeaker must be carefully chosen. It is usually undesirable to have this higher than the frequency given by the above expression. Maximum bass accentuation will be had when the resonance frequency of the cone is equal to f_1. If, however, this is excessively boomy, a speaker having a lower cone resonant frequency should be used. This will also have the advantage of extending the bass range. Where the cabinet contains the auxiliary radio apparatus the formula given for f_1 will not be accurate, but may still be used as a guide. If the cabinet is very small (e.g. extension loudspeaker cabinets, etc.) f_1 may be of the order of 400 to 500 cps. In this case, it has been found that the most pleasant results are to be had by the use of a speaker with a cone resonance between 70 and 90 cps.

The True Infinite Baffle has the advantage of providing full radiation down to the cone resonance of the loudspeaker without the introduction of other resonances above this frequency, and is therefore suitable for high-quality reproduction.

Small Flat Baffles are suitable for use with high-quality loudspeakers only if the dimensions are large enough to fully justify the speaker. The term "small" is used in the relative sense, and it will be found most speakers would require a baffle that was very large in order to provide good radiation down to their resonant frequency of the cone.

Open Backed Cabinets are usually very convenient, but have very little to recommend them from the acoustic point of view. The inherent resonances make them unsuitable for use with high-quality speaker systems.

The Corner Position

We have said earlier that it was preferable to mount the loudspeaker near a corner of the room. This is universally true for all methods of loudspeaker

MATERIAL	DENSITY gms/cu. cm.
Concrete	2.6
Brick	1.8
Dry Sand	1.5
Chipboard	0.81
Oak	0.72
Plywood	0.67
Mahogany	0.67
Pine	0.45

Fig. 8. Densities of materials used in construction of speaker enclosures.

mounting, since at low frequencies the bass radiation will be increased in a manner readily appreciated if we consider firstly a small source of sound in an open space.

The radiation from this source will be of equal intensity at a given distance in all directions, i.e. spherical. If now a large flat wall is placed near the sound source then the total radiation will be concentrated into a hemisphere and its intensity will then be doubled. Similarly if a second wall is placed near the sound source at right angles to the first the total radiation will be concentrated into one quarter of a sphere, then its intensity is four times greater. Again a third wall at right angles to the other two will increase the intensity eight times.

A loudspeaker standing in the corner of the room may at medium low frequencies be regarded as similar to the second case, and approaching the third case as the frequency falls to a point where the wavelength is much greater

than the height of the speaker above the floor.

The Closed Box

One method of preventing back to front cancellation is, of course, to enclose the rear of diaphragm completely, thus achieving what is effectively an infinite baffle (*Fig. 5*). The enclosed volume of air, however, may be regarded as an elastic cushion which, when the loudspeaker cone is displaced inwards, is under compression, and when the cone is displaced outwards, is in rarefaction. In either case the enclosed air will attempt to return to its normal state, and in so doing will apply a stiffness force to any movement of the cone from its position of rest.

We have seen in *Fig. 2* that the bass characteristics of a loudspeaker are determined largely by its frequency of resonance which, in turn, is governed by the mass of the cone and the stiffness of the suspension (M_o and C_c). We now see that the enclosed volume of air adds a further stiffness which is shown as an additional series capacitance C_b in the analogous circuit (*Fig. 6*). The effect of this, of course, is to raise the resonant frequency.

Since the value of C_b varies with the volume of the box, the larger the box, the more extended will be the bass response for a given value of cone resonance. Alternatively, for a given bass extension, the lower the cone resonance, the smaller will be the box required.

It is difficult to give a formula showing this last relationship since it is not sufficient to know only the cone resonance, but the corresponding values to M_o and C_o must also be known, and usually they are not. However, the table of *Fig. 7* is offered as a guide.

When constructing enclosures of this and other types which will be described subsequently, two important points must be borne in mind:—

(1) At frequencies whose wavelength is comparable to the internal dimensions of the enclosure, reflections between inside faces will occur which will be additive at some frequencies and cancelling at others, thus causing irregularities in the response. It is therefore necessary
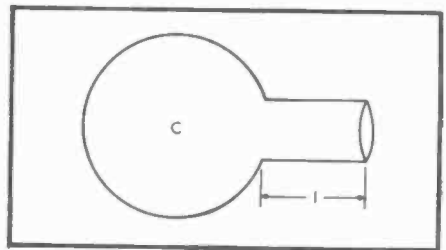

Fig. 9. Diagram of a Helmholtz Resonator.

to reduce these standing waves, as they are called, to a minimum. This may be done by lining the enclosures heavily with some soft absorbing material, such as felt or cotton wool. There seems to exist in some minds the fallacy that this treatment of the enclosure decreases the top response. This is probably due to the standing waves which occur in an unlined enclosure imparting a hard apparent brilliance to the reproduction, and lining the enclosure removes the coloration. It must be remembered that the high-frequency radiation comes mainly from the front of the cone and this is not affected in any way by the enclosure which should *always* be lined. (2) The material (usually wood) of which the cabinet is made, possessing both weight and stiffness will therefore resonate at one or more frequencies, and in so doing will (a) behave as a diaphragm, and become quite an efficient radiator at that frequency; (b) modify the air loading on the cone at that frequency, both of which provide unwanted coloration in the reproduction. Partially or completely enclosed cabinets are more prone to this defect than open backed cabinets because of the high pressures often built up in the former. The difficulty is overcome to a great extent by constructing the enclosure of as rigid and dense a material as possible. Concrete is excellent for this purpose, and is sometimes used. A little more practicable is the use of sand filling, where sand is sandwiched between two thin wooden laminations the thickness of the sand being of the order of 1 to 2 in. If solid wooden walls are used, these should be at least $7/8$-in. plywood or material of equivalent density. A table of comparative stiffnesses and densities is given (*Fig. 8*).

The Vented Box

So that the enclosure does not raise the resonant frequency of the loudspeaker cone an opening or vent may be included and to understand the effect of this we must first consider the Helmholtz Resonator. This resonator, named after a German physicist, consists of a cavity having a communicating duct to free air as shown in *Fig. 9*. This arrangement will have a resonant frequency, and it functions as follows:

As we have seen, an enclosed volume of air will behave as an elastic cushion. We now consider the air in the duct as a piston having mass and being capable of moving backward and forward.

If this air block is displaced or excited, it will oscillate on the elasticity of the air in the cavity. The frequency of this oscillation depends upon this elasticity or stiffness and the mass of the air in the duct. These values are
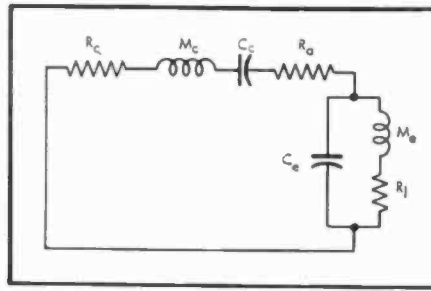
Fig. 10. Analogy of the vented box.

determined by the dimensions of the resonator and the resonant frequency f_o is given by

$$f_o = \frac{c}{2\pi} \sqrt{\frac{A}{l'V}}$$

where $A =$ the cross section area of the duct in sq. ft.
$l' =$ the length of duct plus correction in ft.
$v =$ the volume of the air cavity in cu. ft.
$c = 1130$

In practice the duct may be sometimes omitted; l' then becomes equal to the thickness of the resonator wall.

Some of the air adjacent to the ends of the duct moves with the mass of the air in the duct, and the effective length of the duct is greater than the actual length. This calls for a correction factor.

It has been shown by Rayleigh that the effective increase in length at each end $\Delta l = \frac{8}{3\pi} R$ where l and R are of the same denomination, and R is the radius of the duct.

Therefore effective length $l' = l + \frac{16R}{3\pi} = l + 1.7R$.

If the section of the duct is not circular, the above is still true if $R = \sqrt{A/\pi}$ where $A =$ the cross sectional area of the duct.

Returning now to loudspeaker enclosures, a vented box containing a loudspeaker will behave as a Helmholtz resonator and when the loudspeaker strikes the resonant frequency of the box the motion of the air in the vent will be at a maximum and in phase with that of the cone, in which case, since the air in the enclosure is being brought under compression or tension by both the loudspeaker cone and the air piston in the vent simultaneously, its effective stiffness rises and the cone velocity is reduced at this frequency.

Another way of looking at this is to consider the electrical analogy where the vented box is represented by a parallel tuned circuit in series with the loudspeaker series circuit we have seen previously. This is shown in *Fig. 10*. Again the cone velocity is represented by I and at resonance the impedance

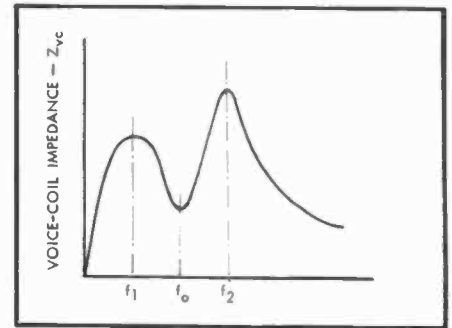of the parallel tuned circuit becomes high and I is reduced. We will call this

Fig. 11. Impedance curve for the vented box or bass-reflex enclosure.

frequency f_o. Since the cone velocity falls at f_o, it is an antiresonance.

Below the resonance of the enclosure the vent air mass moves in anti-phase to the cone, i.e. as the cone moves inwards the air compression forces the air in the vent outwards, and vice versa. Under this condition the volume of air in the enclosure is mainly employed transferring the motion of the cone to the vent air mass; consequently this mass can be considered for most purposes as being added directly to the mass of the cone. The latter thereby exhibits a resonance at a frequency lower than its unbaffled value.

This condition is represented in the circuit shown in *Fig. 10* where below the resonance the dominating reactance of the parallel circuit is M_e and the resulting series circuit consisting of R_1, M_c, C_c, R_a, M_e and R_1 has a resonant frequency we shall call f_1, where the cone velocity rises.

Again, above the resonance of the enclosure the vent air mass becomes too inert to move readily at these higher frequencies and the vent will behave as though it were blocked. The vented box thus becomes effectively a closed box and, due to the stiffness of the enclosed air the loudspeaker cone exhibits a resonance at a higher frequency than its unbaffled value.

This may be represented again in *Fig. 10* where, above its resonance the dominating component of the parallel section is C_e and the resulting series circuit R_c, M_c, C_c, R_a, and C_e has a resonant frequency we shall call f_2. A typical impedance curve showing this is shown in *Fig. 11*.

It will be realized from the foregoing that at resonance f_o the radiation from the port is in phase with that from the front of the cone, and at frequencies below this the phase of the port radiation changes and is 180 deg. out of place at f_1 and the lower frequencies. This has little effect when the area of the port is small compared to the cone pis-

ton area. If, however, the port area is comparable to the piston area then the radiation from the port, at frequencies around the cabinet resonance f_o, will add to the radiation from the cone and substantially compensate for the fall in output from the cone due to the reduced velocity at this frequency. At frequencies approaching f_1 and below, the port radiation, under these conditions, will tend to cancel the radiation from the cone and consequently below f_o the output falls rapidly. In these circumstances, velocity curves are not an indication of the over-all radiation efficiency but they are used to indicate the nature of the air loading applied to the cone.

Thus we have seen that a loudspeaker mounted in vented box will show two resonant frequencies f_1 and f_2, in between which is an anti-resonance f_o. The expression for the resonant frequency f_o of a vented box is the same as that for a Helmholtz resonator, which, including corrections, becomes:

$$f_o = \frac{C}{2\pi} \sqrt{\frac{\pi R^2}{V \left(l + \frac{16}{3\pi} R \right)}}$$

where v is the volume of the box
$A =$ the area of the vent
$R = \sqrt{\dfrac{A}{\pi}}$
$l =$ length of tunnel or thickness of wood and all dimensions are of the same denomination.

Evaluating constants and expressing dimensions in inches we have:—

$$f_o = 3{,}824 \sqrt{\frac{R^2}{v(l + 1.7R)}} \text{ cps}$$

from which $v = \dfrac{1.463 \times 10 \ R^2}{f_o{}^2(l + 1.7R)}$ in ins.

The vented box then has two advantages over the previous systems. (1) The stiffness of the enclosed air is relieved by venting, thus the amplitude of the resonance f_2 due to this stiffness acting on the cone is reduced.

(2) Since f_1 is below the normal cone resonance, the bass response from the speaker is extended down to this frequency if the port area is small.

In practice, the design of a vented box is a process of trial and error. It is not usual for such an enclosure to have a duct extending from the vent so the effective duct length equals the thickness of the cabinet material plus end correction. The "tuning" of these enclosures is accomplished by varying the area of the vent and in many cases facilities for doing this are given to the user. Needless to say, there is no attempt to tune these enclosures to a predetermined frequency, but rather to tune them until they sound right which is after all a very commendable ambition. As a guide, however, as the vent area is decreased the lower resonant frequency of the system f_1 will decrease both in frequency

and amplitude, whereas the upper resonant frequency will increase in amplitude. *Fig.* 12 shows this effect.

Obviously, it is very much a matter of
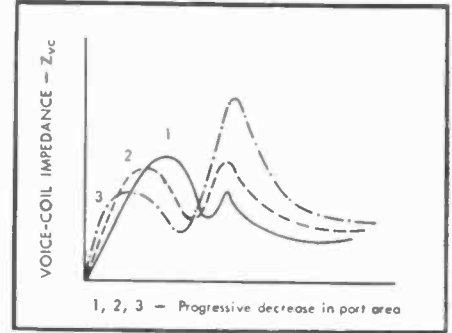
Fig. 12. Varying port area affects the impedance curves for bass-reflex enclosures.

opinion which of these is to be preferred. the curve showing the most extended bass also has the most prominent resonance at f_2. It is interesting to note that if the vent were closed up completely f_1 would disappear and we should be left with a very prominent f_2 which is of course the characteristic we have shown for the closed box.

Part 2. Concluding the presentation of a thorough treatise on the performance of loudspeakers in different types of mountings, from the flat baffle to the most elaborate enclosures. This installment commences with a discussion of the reflex cabinet.

THE REFLEX CABINET is a special case of the vented box where the area of the vent is fixed and normally equals the piston area of the cone, and the resonant frequency of the enclosure is made to coincide with that of the cone. The term "reflex" is based on the fact that at its resonant frequency the radiation from the back of the cone appears at the vent inverted due to the resonator action described previously. The design of the reflex cabinet is greatly simplified since now we know the frequency to which we are going to tune the enclosure and we can use previous formulas for its dimensions.

In order to keep the dimensions as economic as possible a duct is often employed so that a smaller cabinet can be used for a given resonant frequency. The duct length is limited by two factors:

(1) It is convenient to have the duct extending into the enclosure (*Fig.* 13), so the volume of the duct must be added to the volume of the enclosure; consequently if the duct length is increased a point will be reached when the decrease in the required cavity volume is being

equalled or even exceeded by the increase in the volume of the duct. The optimum duct length for a minimum volume occurs when the length of the duct $l = \dfrac{2160}{f} - 1.7R$ where both dimensions are in inches.
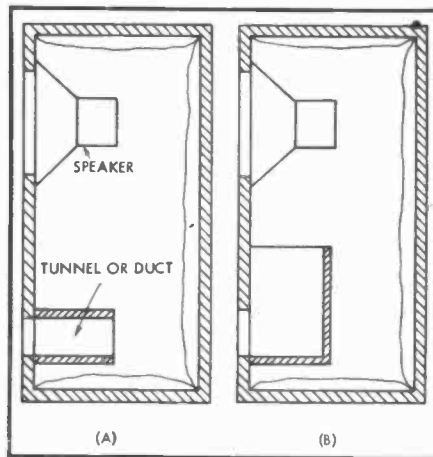

Fig. 13. Cross section of reflex cabinets. (B) shows how the tunnel may be folded.

(2) It is necessary for the duct to be short compared with a wavelength at frequencies around resonance, and should not exceed $\dfrac{1130}{f_o}$ inches.

The total volume with the duct is given by:

$$v = \pi R^2 \left[\frac{4.66 \times 10^6}{f^2 (1 + 1.7R)} + 1 \right]$$

where all dimensions are expressed in inches.

The advantage of the coincident resonances in a reflex cabinet is that the loudspeaker cone receives maximum damping at its resonance which results in

(1) Reduction of cone velocity at resonance but an increased radiation efficiency around the frequency due to vent radiation.
(2) Increased power handling capacity.
(3) Improved transient response due to increased damping at resonance.
(4) Reduced distortion due to reduced cone amplitude.

The resonances f_1 and f_2 are spaced either side of the cone resonance and are equal in amplitude. Impedance curves comparing similar speakers (A) on an infinite baffle and (B) in a reflex cabinet are shown in *Fig.* 14.

Further Types of Vented Enclosures

The Acoustical Corner Ribbon Enclosure. This is an enclosure of the double reflex type as shown in *Fig.* 15. The analogous circuit is also shown. A system of this nature will exhibit three impedance resonances alternating with two anti-resonances. A sketch showing comparison as stated is shown in *Fig.* 16. The progressive increase towards a flatter and more extended bass is noticed.

The BBC Enclosure. An enclosure based on a similar principle was introduced by the BBC in 1935 except that the mass M_1 was formed by a blanket of felt as shown in *Fig.* 17.

The RJ Enclosure. Recently introduced in the U.S.A. was an enclosure known as the RJ. This was another

Fig. 15. Cross section and analogy of Acoustical Manufacturing Co's. corner ribbon enclosure.

Fig. 14. Impedance curves of reflex enclosure vs. infinite baffle.

form of vented enclosure of very small dimensions in which a low resonant frequency was achieved by using a long duct of very small cross-section. Due to the very resistive nature of such a duct, the lower-frequency resonance normal with this type of enclosure had a greatly reduced amplitude. Normally, as we have seen, this would result in the higher-frequency resonance having prominence approaching closed box conditions. This, however, was overcome by reducing the loudspeaker aperture take the form of a narrow slot. A number of designers have used a slot in this manner so we will consider it a little

more closely.

Slot Diffusion. This involves mounting the loudspeaker not facing into a circular hole as is normally done, but into a narrow vertical slot. The initial advantage claimed for this was an improved treble distribution. A second advantage of using a slot is that the reduced area in which the cone is radiating increases the mass and friction loading on the cone and thereby lowers the resonance both in frequency and amplitude. The analogous circuit of the RJ enclosure circuit is shown in *Fig.* 18 where R_s and M_s are the slot components.

Arguing as before, the lower resonant frequency f_1 will be substantially that of R_c, M_c, C_c, R_a, R_s, M_s, M_e, R_{a_2} in series. The upper resonant frequency f_2 will be R_o, M_c, C_c, R_a, R_s, M_s, C_e in series.

Since the combined impedance of M_e and R_{a_2} is much higher than that of M_s and R_s, the latter will have a greater effect on f_2 than f_1. Thus the use of slot loading tends to overcome the excessive prominence of f_2 when f_1 is heavily damped by using small resistive ducts.

The great disadvantage of slot loading is in the fact that the cavity formed between the cone and the faces of the slot together with the slot itself, form a Helmholtz resonator which, depending upon dimensions, will be highly active in the middle registers. In addition, serious irregularities in this range will occur due to standing waves existing between the surface of the cone and the inside surface of the slot face. Consequently the use of a slot at frequencies above 200 to 300 cps is rather unsatisfactory.

Various alternative methods of applying the series mass and friction components have been used in this country. A simple yet very effective system patented by Murphy Radio Ltd. was to apply the loading to the rear of the loudspeaker, in the form of a corru-

gated cardboard drum (*Fig.* 19) where the "tubes" formed by the corrugations provided the required loading. The disadvantage of this arrangement was that the drum tended to exhibit resonances normally encountered in a tube whose dimensions equalled those of the drum.

Mass Resistive Enclosure. This enclosure, developed more recently by Goodmans Industries Ltd., whilst following in the broad principle the above systems, had, however, several refinements. Consisting basically of an enclosure with a small vent, the dimensions of these were so chosen in conjunction with the constants of the cone to be used therein that the lower resonance f_1 was made to occur at 20 cps, this being considered the lowest frequency that would be required. This meant that the resonance of the enclo-
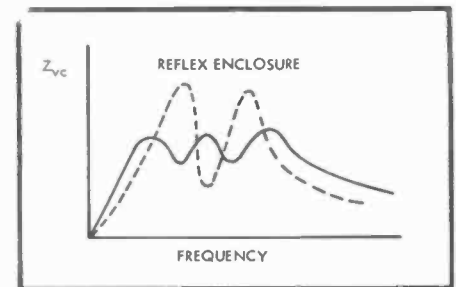

Fig. 16. Response curve (solid line) of enclosure of Fig. 15 compared with reflex cabinet.

sure occurred at some higher frequency, the value of which was of no consequence, and so for the upper resonance f_2 where the only concern was its amplitude.

The dynamic impedance of the enclosure was also considered with its relationship to that of the loudspeaker. This is necessary since if this impedance is too high attenuation will occur at some bass frequencies, and if too low the cone velocity may become excessive and introduce distortion.
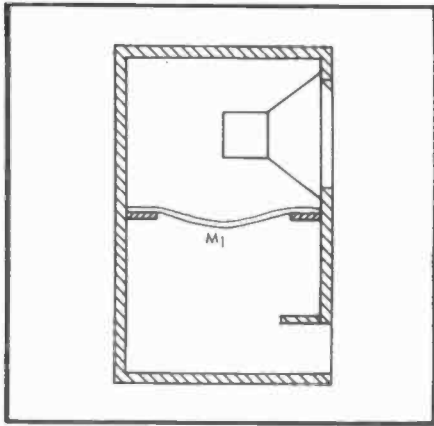
Fig. 17. Cross section of British Broadcasting Corporation enclosure.
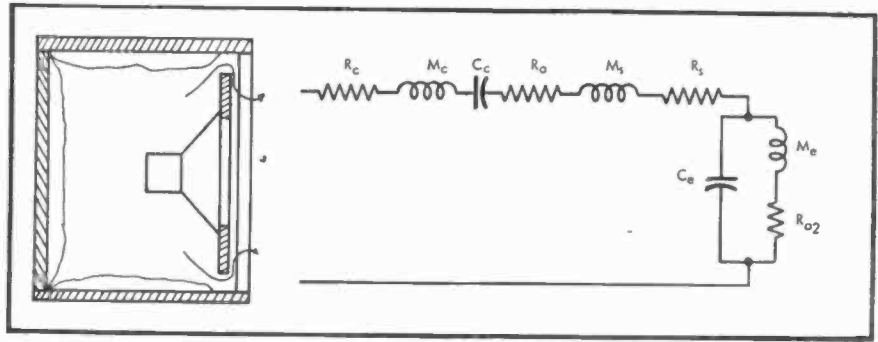

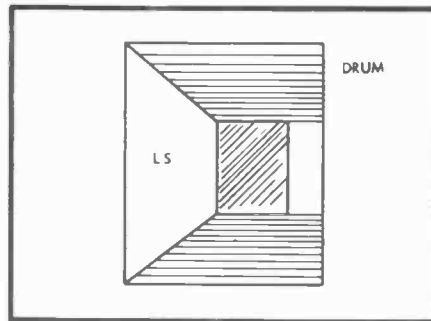Fig. 18. Cross section and analogy of RJ enclosure.

Fig. 19. Arrangement of corrugated cardboard drum surrounding loudspeaker to increase resistive loading.
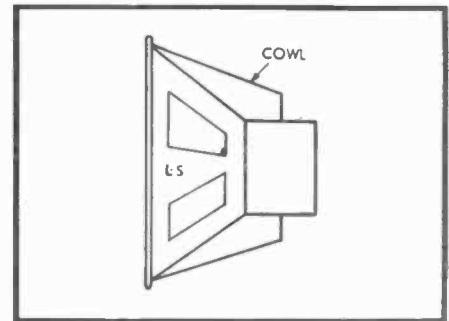

Fig. 20. Another loading device involves the use of a cowl around the loudspeaker mechanism.

A rigid cowl of certain dimensions was introduced to the rear of the loudspeaker as shown in *Fig.* 20. Since the sides of the cowl were not parallel with the sides of the cone standing waves were avoided, and by making the area of the outlet (formed by the annulus between the magnet and the cowl) large compared to the volume existing between the cowl and the cone, its efficiency as a Helmholtz resonator was low. Also since almost the entire length of the cowl was filled by the cone and chassis assembly there was insufficient uninterrupted length to maintain pronounced tube resonance. The additional cone loading was determined by the area of the annulus.

The principal advantage of these types of enclosure over the reflex cabinet are

1. Great reduction in size
2. Better loading down to lower frequencies, although this type of enclosure is not tuned to the cone resonance of the loudspeaker, better damping is obtained by the highly frictional nature of the small vents used.
3. Upper resonance f_2 is not so prominent.

The disadvantage was that although middle- and upper-frequency coloration could be greatly reduced, it could not be entirely removed, and when such an enclosure was used with a really high-quality loudspeaker having a very smooth frequency characteristic, the coloration due to the cowling could be detected. In addition to this, at frequencies immediately above the lower resonance f_1 the output falls due to the restraint imposed on the cone by the cowling.

We have now discussed a few of the types of enclosure used for mounting loudspeakers, and at this stage a brief summary of our findings so far might be considered.

1. *The True Infinite Baffle* consisting

of a loudspeaker mounted in a wall is capable of providing excellent reproduction down to the bass resonant frequency of the cone. There are no additional resonances or coloration above this frequency; there is no appreciable radiation below this frequency due to the mechanical reactance of the cone changing from mass to stiffness. There is no additional air loading at the cone resonance, and excessive cone velocity at this frequency may cause distortion.

2. *The Finite Flat Baffle* possesses the

same characteristics as the wall mounting when the dimensions are sufficiently large. Otherwise a loss in bass radiation occurs below a frequency determined by the dimensions.

3. *The Open-Backed Cabinet*, although the most convenient form of enclosure, is acoustically unsound due to inherent resonances, although these may be used to augment a falling bass characteristic due to insufficient baffle area.

4. *The Enclosed Box* is very good provided the loudspeaker cone has a sufficiently low resonance; otherwise, added stiffness due to the enclosed air causes excessive rise in this resonance.

5. *The Vented Box* overcomes the inherent disadvantage of the closed box and will extend the bass response of almost any loudspeaker quite appreciably. In the case of loudspeakers having very low cone resonances, i.e. between 20–25 cps, this extension is rather pointless. The upper resonance f_2 is present in
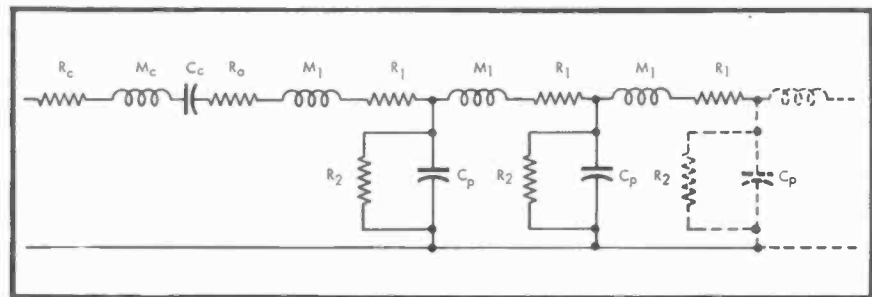

Fig. 21. Analogy of infinite pipe loading on loudspeaker. R_c, M_c, C_c, and R_a are loudspeaker components; M_1 is distributed mass, C_1 is distributed compliance, R_1 is friction at walls of pipe, and R_2 is absorption at walls of pipe.

this type of enclosure and is usually pronounced when using normal loudspeakers.

6. *The Reflex Enclosure* has the very great advantage over all previous types by providing the loudspeaker cone with maximum air damping at resonance. The concomitant advantages are outlined in the description. It loses some of this advantage when compared with wall mounting at frequencies above the

cone resonance due to the upper resonance f_2, although this may not be sufficiently serious to turn the scale. The reflex enclosure is usually more practical to construct.

7. *The systems* involving additional mass loading on the cone, when carefully designed, hold all the advantages over the previous systems, but fall down by having irregular middle and upper frequency characteristics, and it is very doubtful whether it is better to have very smooth bass response if reproduction above 200 cps is colored.

PIPES—FOLDED AND TAPERED; LABYRINTHS AND HORNS

As before, we are concerned with what happens to the back radiation of a cone. So far we have pushed it into the next room, shut it up in boxes and squeezed it out through vents. We shall now consider a few further modes of attack.

The Infinite Pipe

If the rear of a loudspeaker were connected to an infinitely long pipe all the back radiation would disappear to infinity or be absorbed on the way, as was the case for the true infinite flat baffle. The pipe, however, would provide a much greater air loading on the loudspeaker cone. The infinite baffle could very approximately be replaced by a pipe whose length was very much greater than a wavelength of the frequency being considered.

An example of such a pipe has been met where the loudspeaker was mounted on a baffle fitting flush with the fire grate, so that the back radiation went up the chimney. (The extension of this story which says that a bird's nest in the chimney provided the correct resistance termination is discounted.) Pipes of this nature may be regarded as alternate air masses and air compliances and if the pipe were of constant cross section, these would be distributed evenly along its length. The analogous circuit for this is given in *Fig*. 21.

The Absorbing Pipe or Labyrinth

Usually it is impracticable to make the pipe long compared to the wavelengths of the lowest frequencies involved, and problems of pipe resonance are encountered. As will be seen later, an open pipe will exhibit resonances, due to reflections from the open end, at all frequencies where the length of the pipe approximates to any whole number of half wavelengths, i.e., if the length of the pipe is equal to half the wavelength of frequency f_o then the pipe will resonate at f_o, $2f_o$, $3f_o$—nf_o where n is any whole number. In order to reduce these resonances, the walls of the pipe are arranged to have a high coefficient of absorption and to

present as much friction to the air movement as possible.

Again considering the analogous circuit, at frequencies where the pipe is short compared to a wave length, C_p vanishes and the circuit may be redrawn as at (A) in *Fig*. 22. The total impedance, Z_p, of the pipe components is almost entirely resistive. This is demonstrated by the impedance curves where the loading due to the pipe has reduced the amplitude of the cone resonance, but its frequency is unchanged, as shown at (B) of *Fig*. 22. Once again there is little to be gained from evaluation of component values since there are too many factors normally unknown to the constructor. We will say, however, that a folded labyrinth will be effective down to that frequency where
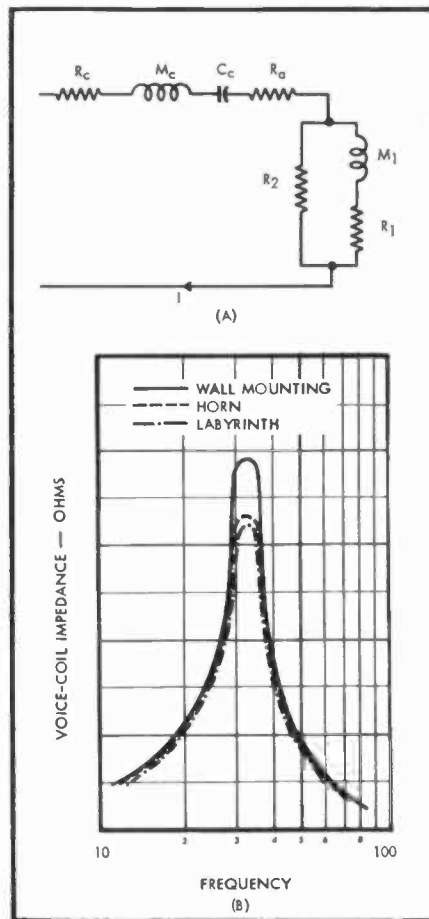
Fig. 22. Analogy and response curve for absorbing pipe.

the length of the pipe is one quarter of the wavelength. The cross-sectional area of the pipe should be equal to the piston area of the cone.

The advantage of the labyrinth, over the infinite baffle, provided it is sufficiently long, is that it provides an increased air load on the cone. In addition, it provides complete absorption of the back radiation without the introduction of any resonances of its own and without modification of the speaker resonance. This resistive loading is also

responsible for the principal disadvantage of the labyrinth, i.e. its low efficiency. From the analogy it will be seen that the cone velocity (I) is limited to a large extent by the friction component (R_1). The absorption factor is too low (R_2 is too high) to have very much effect at very low frequencies. However, the excellent air loading at very low frequencies makes possible the use of sufficient bass boost in the amplifier to compensate for the fall in efficiency without the danger of undue distortion.

As was indicated previously, this is only true if the length of the labyrinth is not less than one-quarter wave length of the lowest frequency that will be fed into the speaker. The cone resonant frequency should not be considered as the limiting frequency, since by use of the above methods and with suitable bass lift good reproduction may be had from the speaker at frequencies well below its resonance.

The Tuned Pipe

We have mentioned on several occasions that a pipe will exhibit resonances when its length approaches a quarter wave length. We will now examine this more closely. Consider *Fig*. 23.

Assume that one end of each pipe is connected to a piston vibrating at frequency f. Assume further that the piston has just moved forward to produce a pressure corresponding to point A on the curve. The distribution of pressure in pipes of various lengths is indicated by the curve.

Pipe P_1 is short compared to the wave length at frequency f and therefore the pressure difference along its length is small. If the far end of P_1 is open, the contained air will behave as a mass at this frequency, and if P_1 is closed at the far end the enclosed air will behave as a stiffness, either case providing an impedance to the motion of the piston.

Pipe P_2 is a quarter wave length of the frequency f, and at the instant shown the pressure varies from maximum to zero along its length; a quarter of a cycle later than this instant the point of maximum pressure would have
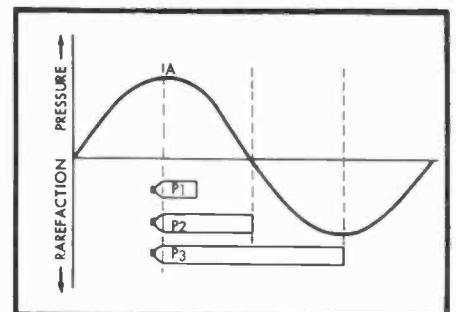
Fig. 23. Distribution of pressure in pipes of various lengths.

reached the far end of the pipe. If this end were open the pressure would be expelled and a rarefaction would be reflected back to arrive at the piston just as this had moved outwards, thus adding to the partial vacuum against which the piston was already working. Thus at this frequency the load on the piston would be increased, in other words an anti-resonance occurs.

If, however, the far end of the pipe were closed, the pressure wave arriving there could not be expelled and, therefore, would be reflected back to arrive at the piston as the latter moved outwards. This would tend to cancel the partial vacuum against which the piston was working, and the load on the piston would be decreased, i.e. a resonance occurs.

In the case of the open pipe, resonances will occur at all even numbers of quarter wave lengths, and anti-resonances will occur at all odd numbers of quarter wave lengths. For the closed pipe the reverse is true.

We shall now see how these actions are applied to loudspeaker enclosures. One system is to use an open pipe whose anti-resonance is made to coincide with the cone resonance. If this frequency is f, then the length of the pipe is:

$$l = \frac{1130}{4f} - 1.7\ R \text{ ft.}$$

With l expressed in inches,

$$l = \frac{3390}{f} - 1.7\ R$$

Where $R =$ the radius of the pipe or $\sqrt{A/\pi}$ when A is the cross sectional area.

This enclosure then exhibits a very similar impedance characteristic to the reflex cabinet described earlier. It has the advantage of being rather more simple to construct. The disadvantage of tuned open enclosures is that, as we have seen, they exhibit resonances at all the harmonic frequencies. These may be reduced by the use of an absorbent lining as we have described previously.

An alternative to this is to use a closed pipe tuned to a frequency a little lower than the cone resonance and thus use the resonance of the pipe to overcome the stiffness reactance of the cone at this frequency, thereby extending the bass response. The formula is that given previously. A satisfactory value for f in this case would be about 20 per cent lower than the cone resonant frequency. An enclosure of this type exhibits resonances at all the odd harmonics of its resonant frequency.

The analogous electrical circuit for this type of enclosure is the same as that for the infinite pipe shown earlier in *Fig.* 21 except in the case of the tuned pipe the value of R_1 is much lower and R_2 is very high.

Many attempts have been made to prevent the troublesome reflection oc-
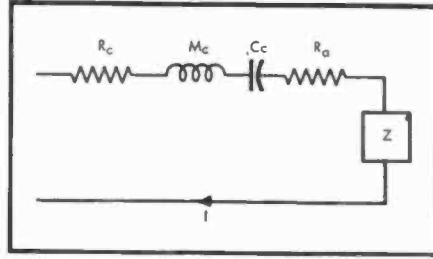
Fig. 24. Equivalent circuit for horn loading on a loudspeaker. Z is the impedance of the enclosure.
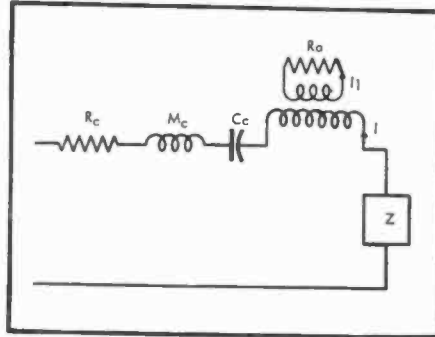

Fig. 25. Analogy of horn loading to show matching of generator to load by means of a transformer.

curring in a short open pipe by the addition of some friction materials to the open end. It is well known in electrical terminology that a circuit of the form shown in *Fig.* 21 is a transmission line which will present a certain characteristic impedance to the input circuit. If, however, the line is very short but its far end is terminated in an impedance equal to the characteristic impedance, the line will possess identical characteristics to a transmission line that is infinitely long. The line is then said to be correctly terminated.

This being so, it was not unnatural to imagine that the open end of the tube could be correctly terminated. The difficulty here, however, was that unless the tube was either completely open or completely closed, it would tend to resonate on the Helmholtz principle making its performance as a tube difficult to determine.

The various systems described in the section on vented enclosures for reducing the upper resonance may be applied to the tuned pipe. Any system increasing the mass loading on the one will decrease the harmonic resonances with respect to the fundamental. Once again, of course, the problem of middle and upper frequency coloration is introduced.

One further system that has found favour is the reduction of one of the harmonic resonances by virtue of the speaker position; for example the third harmonic resonance which is the most prominent in a closed quarter wave

tube may be removed by positioning the speaker one-third of the length from the closed end. The speaker is then at a node of the third harmonic resonance and this is then not excited. This principle is used by the Decca Record Company in their corner enclosure and gives excellent results. A further feature of this enclosure is that the pipe is tapered. This serves to lower the efficiency of the resonance and extend it over a wider band of frequencies and is a principle incorporated in the Voigt corner horn.

Incidentally, it is interesting to note that since the speaker is nearer the closed end of this enclosure the motion of the air in the pipe will, at resonance, be in anti-phase with that of the cone. The formula given will not apply in this case since the speaker opening in this position will modify the characteristics of the tube.

Horn Loading

At frequencies where the dimensions of the loudspeaker cone become very small compared with a wave length, the radiation resistance falls. Above the cone resonance, however, the laws governing the velocity of the cone are such as to compensate for this to some extent, but nevertheless the efficiency of the speaker is considerably reduced at these frequencies. All this may be seen by considering the equivalent circuit (*Fig.* 24).

At very low frequencies R_a becomes small compared with the impedance of the remaining components and consequently V and the power in R_a is decreased although above resonance the increase in I due to the falling impedance of the inductive elements partially compensates for this. It is well
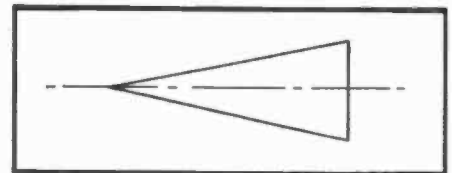

Fig. 26. Cross section of conical horn from throat to mouth.

known that the maximum power delivered from a generator to a load occurs when that load equals the impedance of the generator. If necessary, the load can be matched to the generator by a transformer. Applying this to the analogy we have *Fig.* 25.

The higher impedance of the transformer primary matches more accurately that of the remaining components so V is increased. In the secondary, I_1 is much higher than I in the previous condition, so that the power in R_a is increased.

Acoustically, such a transformer is

analogous to a horn (*Fig.* 26). The pressure build-up due to the limited radiating area at the throat of the horn will present a high impedance to the cone of the loudspeaker and reduce its velocity in exactly the same way as it does in a tube.

At the mouth of the horn due to its larger area, the impedance to the sound wave will be much lower; the pressure will therefore be lower and the velocity high. This ties up with our pressure-voltage, current velocity, analogy.

There are two factors which determine the lowest frequency at which a horn will maintain its efficiency. These are (1) the rate of flare; (2) the diameter of the mouth.

(1) There is no formula showing an optimum rate of flare for a conical horn for a given low-frequency extension. Since the efficiency falls steadily over a wide band of frequencies one can only state that the lower the rate of flare, the better.

Most horn designs are based, not on the conical, but on the exponential horn and here it is possible to arrive at an optimum value, as we shall see.

(2) The mouth diameter of any horn must be equal to at least one quarter of the wavelength of the lowest frequency it is required to radiate, otherwise standing waves will be set up due to reflections from the open end in the manner we have shown for tuned pipes.

The Exponential Horn

It has been found that for a given low-frequency extension, size for size, the exponential horn is very much more efficient than a conical horn. Its use, therefore, is almost universal.

An exponential horn is one whose cross-sectional area increases exponentially with the distance from the throat (*Fig.* 27), and the cross-sectional area A at any distance d from the throat is given by

$$A = a\varepsilon^{md}$$

where a = throat area

$\varepsilon = 2.713$

$$m = \frac{4\pi f}{c} = \frac{f}{927}$$

if all dimensions are in inches. The very great disadvantage of the horn is its size. In order to demonstrate this, let us design a horn suitable to load a 12 in. loudspeaker down to, say, 30 cps which is, after all, quite a modest figure.

Mouth Diameter $D = \frac{c}{4f} =$

$$\frac{13,560}{4 \times 30} = \quad 113 \text{ in.}$$
$$= \quad 9.5 \text{ ft.}$$

Area of Mouth $A = \frac{D^2}{4} =$

$$\frac{113^2}{4} = 10,020$$

Area of Throat $a =$
$$\text{piston area of 12-in. cone}$$
$$= 78.5 \text{ sq. ins.}$$

$A = a\varepsilon^{md}$ (d is the unknown)

$$\frac{A}{a} = \varepsilon^{md} \quad m = \frac{f}{927} = \frac{30}{927}$$

$\text{Log}_\varepsilon A - \text{Log}_\varepsilon a = md$

$$d = \frac{\text{Log}_\varepsilon - \text{Log}_{\varepsilon a}}{m}$$

$$= (\text{Log}_\varepsilon 10,020 - \text{Log}_\varepsilon 78.5)927$$

$$= \frac{(9.2122 - 4.3631)927}{30}$$

$$= 150 \text{ in.}$$

$$= 12 \text{ ft.6 ins.}$$

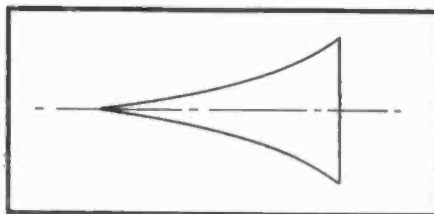Needless to say, these dimensions are

Fig. 27. Cross section of exponential horn from throat to mouth.

quite impracticable for most people.

One way of overcoming this is to flick through the pages of some of the magazines devoted to this subject, and one will find designs and advertisements for horns fully capable of maintaining a high radiation efficiency down to 20 cps constructed within a space of about five cubic feet.

We have no comments to offer.

Commercial Folded and Corner Horns

Horns of this nature have come into prominence in recent years, particularly in the U.S.A. and they are a compromise between performance and size. The Klipsch corner horn is accepted universally as one example of this. First introduced by Klipsch, the corner horn, as its name implies, is a folded horn with the final fold formed by the corner and the walls of the room (*Fig.* 28).

Using this artifice, a horn may be made to radiate down to 40 cps without being excessively big. On average, however, the lower frequency limit of these horns is about 60 cps, and one should beware of claims setting this frequency at a much lower figure. Above this frequency, of course, these horns are very efficient indeed and are capable of very impressive reproduction. Indeed, so great is the radiation at these frequencies compared with the middle and upper frequencies that in some designs the bass may be excessive.

There are many different designs of these horns. Usually the rear of the cone is horn loaded and the upper frequencies are radiated from the front in the usual way. In this case, the bass radiation from the front of the cone is small compared with that from the mouth of the horn so the question of cancellation does not arise.

An alternative system is to enclose the rear of the speaker in a small bass chamber and use the mass reactance of the horn loading in the front face to overcome the stiffness reactance of the chamber.

In the well known Voigt corner horn the rear of the loudspeaker is loaded by a tuned tapered pipe, and the position of the loudspeaker in the pipe is so chosen that the acoustic impedance presented to the cone over a wide bond of bass frequencies provides optimum loading for the voice coil. A good radiation efficiency is secured by means of the pipe resonance, which is distributed over a wider range of frequencies by virtue of the taper.

The front of the cone is loaded by a small flare which increases the high frequency efficiency and improves distribution. A schematic diagram of the enclosure is shown in *Fig.* 29.

Summary

1. The Labyrinth is designed to absorb all the back radiation from the cone without introducing any resonance effects. Excellent for maintaining air load on cone down to very low frequencies resulting in a very smooth extended bass response. Its efficiency is low but resistive control of cone velocity allows bass boosting to be used without distress.

2. The Tuned Pipe extends bass response and can provide similar loading conditions to the reflex cabinets on the loudspeaker cone at resonance. It has the advantage of simple construction, but the bass response is not as smooth due to harmonically related resonances.

3. The Horn. The full size horn is without any doubt the ideal method of cone loading, combining high radiation efficiency with very low distortion due to low cone velocity. The only disadvantage is the immense size required to reproduce really low frequencies.

4. Commercial Folded Horns. These are very efficient in their working range. Pros and cons vary from model to model but in general the middle bass may tend to be excessive, and the extreme bass poor; nevertheless capable of impressive results.

FRICTION LOADED ENCLOSURE

We have now covered fairly fully the features of performance and design of most of the recognised methods of mounting a loudspeaker, and we are now able to view these as a whole, pick out the better systems and review their respective advantages and disadvan-

Fig. 28. Back-loaded folded horn in horizontal cross section.

WHERE ℓ = EFFECTIVE LENGTH
A = MOUTH AREA
a = THROAT AREA

Fig. 31. First analogy of friction loaded enclosure.
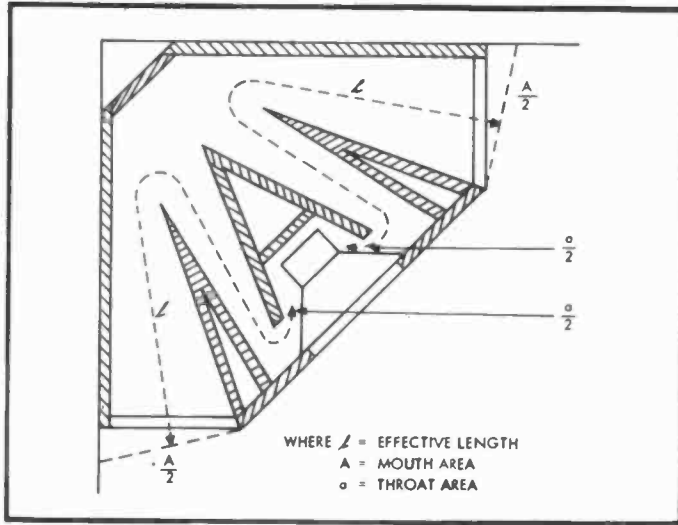
tages on the basis of our foregoing discussions.

The Full Horn. Acoustically this is the ideal method of loudspeaker mounting, it provides excellent air loading on the cone, is devoid of self-resonance and possesses a high radiation efficiency down to any desired frequency, being limited only by the horn dimensions. The disadvantage of the horn is the very great size required to be effective down to very low frequencies.

The Absorbing Labyrinth. This again presents excellent resonance free air loading on the loudspeaker cone, and in this respect is comparable to the horn. It is effective down to any desired frequency, being limited, like the horn, by its dimensions. Unlike the horn, however, the disadvantage of mounting a loudspeaker in a labyrinth is the low efficiency of the system; although this may be partially compen-

sated for in the amplifier. A labyrinth capable of good absorption down to very low frequencies is still rather big.

The Reflex Cabinet. The advantage of the reflex cabinet is that an improvement in the radiation efficiency is maintained at very low frequencies by resonant phenomena. At the same time, excellent damping is applied to the loudspeaker cone at its resonance where it is most required. A further point in its favour is that it is relatively simple to construct.

The bass response from a reflex enclosure will have an efficiency somewhat higher than that from a labyrinth, and for a given bass extension, will be smaller, although it still makes a rather dominating piece of furniture in the drawing-room.

The response will not be so smooth as for a labyrinth due principally to the upper of the two resonances common to this type of mounting. If very much bass boost is applied the reflex enclosure will tend to sound boomy.
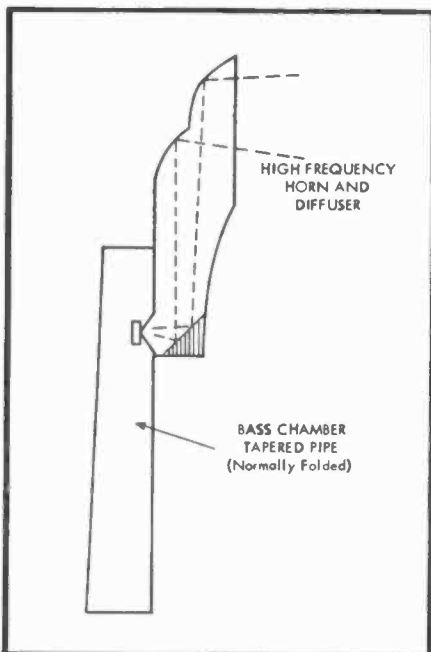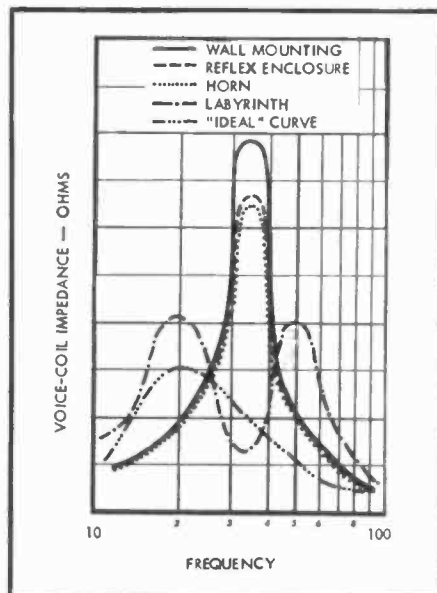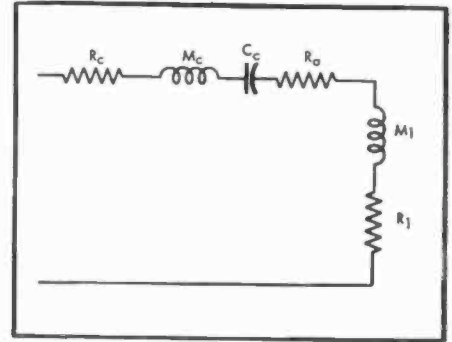
The Wall Mounting or Infinite Baffle. The main virtue of this is the absence of resonances above the loudspeaker resonance, the damping due to air loading is low at resonance, and although this makes for higher efficiency, excessive cone movement at resonance may cause distortion. Typical impedance curves for identical speakers mounted in these four types of enclosure are given in *Fig.* 30.

With the exception of the efficiency attributable to a full-size horn, it was considered, after a great deal of research into the design of all types of enclosure that it should be possible to design an enclosure possessing all the advantages and none of the disadvantages of the various systems. Such an enclosure must have the following qualities :

(1) Resistance controlled mass loading down to 20 cps.

(2) Complete absence of resonances above this frequency.

(3) A general efficiency at least as high as a reflex cabinet and fully maintained down to 20 cps.

(4) Small size.

Such enclosures have now been developed and in order to understand the principle upon which they work, we will develop the argument that led to their design.

In order to satisfy requirements (2) and (3) the cone velocity must increase progressively as the frequency is lowered to 20 cps. Therefore, the enclosure must load the cone in such a way as to bring the effective cone resonance down to this frequency. There must also be a high resistive component in order to satisfy requirement (1).

In the analogy, these conditions are fulfilled by the velocity curve shown dotted in *Fig.* 30 and the corresponding analogous circuit shown in *Fig.* 31 where series inductive and resistive elements are added to the cone circuit.

If we are to apply the impedance

HIGH FREQUENCY HORN AND DIFFUSER

BASS CHAMBER TAPERED PIPE (Normally Folded)

Fig. 29. Schematic diagram of Voigt Corner Horn.

WALL MOUNTING
REFLEX ENCLOSURE
HORN
LABYRINTH
"IDEAL" CURVE

VOICE-COIL IMPEDANCE — OHMS

FREQUENCY

Fig. 30. Impedance curves of identical speakers mounted in various ways.

M_1R_1 to the cone, it is imperative that we also include a shunt capacitance to prevent this impedance from reducing the cone velocity excessively at the higher frequencies, in other words, to decouple the impedance at these frequencies. (*Fig.* 32).

It will now be seen that this is analogous to a special case of the vented box in which the dimension must be considered in conjunction with the mass and stiffness of the loudspeaker cone so that the lower impedance peak f_1 is of a predetermined frequency and magnitude. The approach to our ideal curve is now limited by the upper resonance f_2 and the anti-resonance f_0.

Considering first the latter, we know that this is due to the rise in impedance of the parallel circuit C_e, M_1R_1 at resonance, and since we are now only concerned with the lower resonance f_1, the anti-resonance f_0 (not being tuned to the loudspeaker) serves only to introduce a trough at some frequency above f_0.
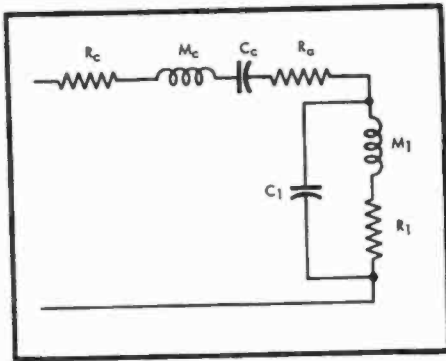
In order to reduce the magnitude of this trough a shunt resistance may be included in the circuit which will have the effects of reducing the "Q" of the parallel section and providing a limiting value for its impedance at resonance. This is shown in *Fig.* 33.

This resistance will also reduce the magnitude of the remaining resonances and it would be possible, by a suitable choice of components, to make the entire circuit almost completely aperiodic. In practice it was found, however, that an increase in the radiation efficiency of the system at extremely low frequencies was an advantage which could be had if the lower resonance was not completely damped, but was controlled by a sufficiently high frictional component in order to prevent excessive cone excursion.

In the final design this was achieved by carefully determined component values which, in addition, are such as to prevent an optimum acoustic impedance to the face of the cone. This is very

important since if this impedance is too low then the cone velocity will be excessive resulting in undue distortion. Alternatively too high an impedance will cause a loss of bass radiation when the acoustic impedance of the enclosure is at its optimum, good radiation is secured with a cone velocity well inside the limits of the speaker. For these conditions it has been found that the dimensions are considerably smaller than those of vented or reflex enclosures. The optimum port area is considerably less than that of cone piston so radiation from the port may be neglected. The friction component R_f is relatively high.

The resistance R_f is analogous to a highly frictional air leak in the enclosure. A convenient way of forming this is to have a number of very narrow slits in one or more of the enclosure walls.

The choice of narrow slits will be appreciated when it is realized that any aperture will have both frictional-resistive and mass-reactive components, and it is very important in our case to ensure that the resistance is high compared to the reactance. The expression for the impedance of a narrow slit is given by

$$Z = \frac{2.23 \times 10.^{-5}d}{w^3 1} + j\frac{6\delta\omega}{51w}$$

Where d = depth of slit

w = width of slit

l = length of slit

δ = density of air

$\omega = 2\pi \times$ frequency

From this it can be seen that as the width of the slit is reduced its resistance increases as the cube of the reactance.

An Axiom 150 Mk II was used in conjunction with this enclosure and the impedance curves are shown for this compared with those of the reflex cabinet and a true infinite baffle when housing speakers identical to the above. The evidence is fairly conclusive. The effect of closing the slits removing R_f is also shown in *Fig.* 34.

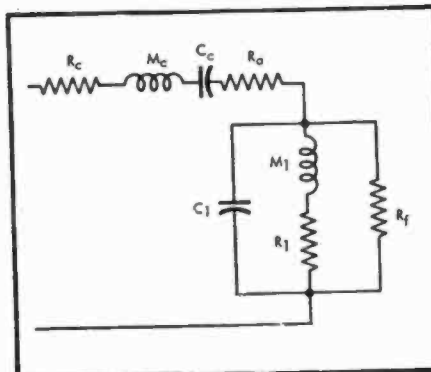Listening tests comparing this enclosure with other types was also encour-

aging. It proved to have a firm bass response without any trace of resonance or coloration. The bass radiation was somewhat better than that from the reflex cabinet at middle bass frequencies and considerably better at the very low frequencies, thereby imparting a warm, well-balanced quality to the reproduction. Tests with an oscillator showed that a strong, pure 20 cps fundamental note could be radiated without excessive cone movement. Transient curves taken showed a very short clear time characteristic of non-resonant conditions. This is more interesting when one realizes that the volume of this enclosure is about half that of a correctly designed reflex cabinet for the same speaker.

In addition to the qualities mentioned this type of enclosure has the following advantages.

1. It is simple and cheap to construct.

2. The dimensions of the enclosure (corresponding to c_1 in the analogous circuit) are not extremely critical and may be varied up to ±10 per cent if necessary for styling. The dimension of the port and slits, however, are critical.

3. The enclosure can be of any shape and the port and slits can be placed in any position relative to each other and to the speaker provided they do not approach the latter too closely so as to produce some degree of back to front cancellation.

4. The resonant frequency of the loudspeaker is not critical provided it is not higher than 50 cps.

Many of these enclosures have been constructed for various speaker systems, and all exhibit the same excellent bass characteristics.

There are of course many different

Fig. 34. Impedance curves of friction loaded enclosure in comparison with other types of mounting.

Fig. 32. Second analogy of friction loaded enclosure.

Fig. 33. Third analogy of friction loaded enclosure.

Ported Loudspeaker Cabinets

JAMES MOIR

A thorough understanding of the author's presentation will enable anyone to design and construct a bass-reflex cabinet which will provide improved performance over conventional "boxes with holes in them."

ORTED OR REFLEX CABINETS are deservedly popular as loudspeaker mountings at the present time, their special merits being the extension of the low-frequency range that may be obtained in a relatively small volume, coupled with an appreciable reduction in the amplitude distortion generated by the loudspeaker. The theory, construction, and operation is an interesting study and it is proposed to comment on some of the aspects in which present theory and practice appear to be at variance.

The first major advantage is the increase in the low-frequency output that is obtainable from a reflex cabinet when compared to the output obtainable from the same speaker unit mounted in a flat baffle, or in many of the alternative enclosures. The increase in output is the result of several contributory factors.

(a) Utilization of the acoustic power output from both sides of the cone.

(b) The close association of two radiating surfaces vibrating in the same phase.

(c) The addition of an Helmholtz resonator to the acoustic system.

Some of the many possible forms are illustrated in *Fig.* 1 from which it will be appreciated that the characteristic feature of all ported cabinets is the addition of an Helmholtz resonator coupled to the rear of the cone, the resonating elements being the acoustic capacitance of the box volume and the acoustic inductance of the mass of air contained in the port and tunnel. At, and in the vicinity of resonance, there is a considerable movement of air through the port and the energy radiated as sound from the port may exceed that radiated from the front of the diaphragm by a factor of several times. If the phase of the radiation from the port is the same (within ± 90 deg.) as that from the front of the diaphragm the total sound output will be increased. It may be shown that the combination of acoustic elements is such that the backward wave from the speaker diaphragm is reversed in phase and thus appears at the port opening *in phase* with the radiation from the front of the cone. The exact mechanism of the phase reversal will not be pursued at this point for the agreement between calculated and measured values of some of the elements in the acoustic phase changing path is poor. Actual measurements of the relative phase of the sound pressure at the port and diaphragm confirm the qualitative theory however.

Though the radiation from the port is in phase with that from the diaphragm in the vicinity of resonance it deviates considerably both above and below the resonant frequency. As the resonant frequency is usually chosen to be near the bottom end of the audio range, the deviation from phase identity below the resonant frequency is not of great consequence. Above the resonant frequency the phase difference can also reach 180 deg. and as this would reduce the total sound output it is necessary to attenuate the high-frequency radiation from the port by adding absorbent material to the interior of the enclosure. A qualitative comparison between the sound output with and without a ported cabinet is given by *Fig.* 2 from which it will be seen that some worthwhile gain is obtained over about one octave above and below the resonant frequency but the effective sound output at very low frequencies is actually reduced by the addition of the acoustic resonator.

Design Procedure
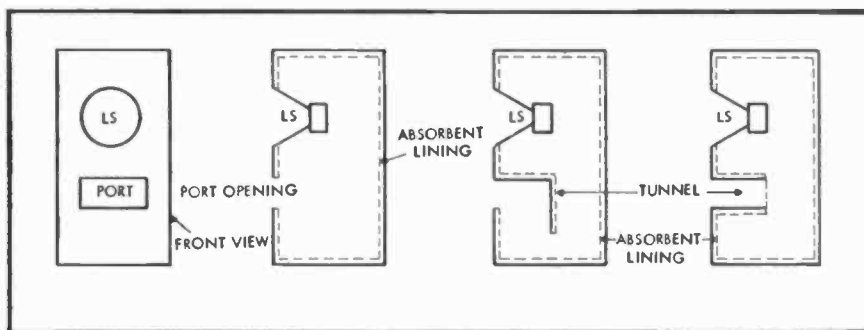
The first problem to be met when ap-

Fig. 1. Typical arrangements of ported cabinets showing two possible positions of the tunnel or duct.
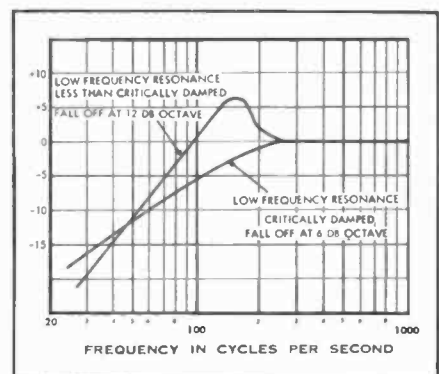

Fig. 2. Effect of damping of low-frequency resonance on acoustic output of a cone loudspeaker.

proaching the design of a ported cabinet is that of choosing the resonant frequency for which the enclosure is to be designed. A full discussion of the reasons governing the choice would require more space than is available and it will be shortened to the point of saying that the resonant frequency of the enclosure is usually chosen to be the same as the resonant frequency of the speaker cone. The acoustic coupling between the resonant enclosure and the resonant mechanical system of the cone and surround is assumed to be such that the electrical impedance/frequency curve of the speaker voice coil will have "maximum flatness." A typical sort of result is illustrated by *Fig.* 3 from which it will be seen that the over-all impedance/frequency curve exhibits the double humped form characteristic of coupled electrical circuits, certainly a major advantage, for as previously pointed out,[1] a flat impedance/frequency curve results in minimum amplitude distortion from the amplifier output stage.

The acoustic resonant frequency of the enclosure is controlled by the physical dimensions of the enclosure volume, the port area, and the tunnel length interpreted as in *Fig.* 4. The relation may be approached either by calculating the equivalent electrical circuit elements or by a more direct approach involving the physical dimensions only. The former gives a clearer insight into the basic process and is invaluable in any investigation but the latter method is shorter and is quite adequate for the enclosure designer. Several analyses have been made but the one most closely in agreement with measurements is that due to Planer and Boswell.[2] Their work leads to an expression for the resonant frequency,

$$f_r = \frac{2150}{\sqrt{V(1\sqrt{A} + L/A)}} \ cps \quad (1)$$

where

V = box volume

[1] J. Moir, "transients and loudspeaker damping." Wireless World, May, 1950.

A = port area
L = effective tunnel length

with all dimensions in inch units.

A critical comparison of the calculated and measured resonant frequencies of a dozen or more enclosures indicated that while none of the published design equations were in perfect agreement with practice, the expression quoted consistently gave the best agreement.

To the enquiring mind, marginal disagreements between theory and practice are often of greater interest than complete agreement so the subject will be pursued in an endeavor to account for the discrepancies. The factors entering into the design equations are enclosure volume V, port area A, and tunnel length L, and as the effective values of these may differ somewhat from their physical values they will be considered in turn. In the simple case where no tunnel is employed there would not appear to be any great margin for error in determining box volume V, though the literature is a bit inconsistent in deciding whether the volume of any absorbent lining should be deducted from the chamber volume to obtain the effective volume. Qualitative considerations suggest and experiment confirms that the volume of permeable linings such as fibreglass or hair felt should not be deducted from the casing volume but that allowance should be made for the volume of the more impermeable materials, such as insulation board, cane fibre or asbestos fibre tiles.

As far as can be ascertained prior literature is completely in error in dealing with the effective volume of an enclosure that includes a tunnel, the unanimous and apparently reasonable decision being that the tunnel volume should be subtracted from the internal volume in order to obtain the effective volume. This outlook would seem to be based on the simplifying assumption that the air in the tunnel takes no part in the compression and expansion cycle which charac-

[2] Planer and Boswell, "Vented loudspeaker enclosures." Audio Engineering, May, 1948.
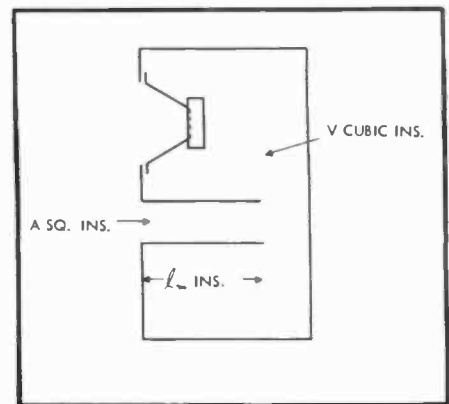

Fig. 4. Parameters important in the design of a ported cabinet.

terises the acoustic regime in the volume V but merely undergoes translation along the tunnel. A little thought will suggest that this assumption is probably untenable but any doubts were resolved in a relatively simple manner.

Experimental Determination

A ported enclosure was constructed in which the tunnel volume represented some 30 per cent of the enclosure volume and of such a shape that the tunnel could be added either on the inside or outside of the box as indicated in *Fig.* 5. This artifice maintains the tunnel length substantially constant but allows the tunnel volume to be removed from the cabinet volume. The resonant frequency of the enclosure was then measured (using a precision low-frequency oscillator) with the tunnel in both positions. In neither example tested was there any indication that the position of the tunnel had any significant effect upon the resonant frequency although the test method employed was capable of detecting a frequency shift of less than one tenth of that expected from calculations based on the normal assumption.

The actual experimental verifications were carried out by two competent engineers well versed in the conventional theory and quite skeptical about the writer's preliminary suggestion that the accepted theories were in error. We may say with some confidence that the tunnel volume should *not* be subtracted from the enclosure volume to obtain the effective volume of the enclosure.

Speaker's Volume

Prior literature is also quite unanimous and apparently in error about the correction to be made for the volume occupied by the speaker unit, specifying that the effective volume of the speaker unit is that shown in solid at (A) in *Fig.* 6. Once again some preliminary theorizing suggested that the effective volume of a
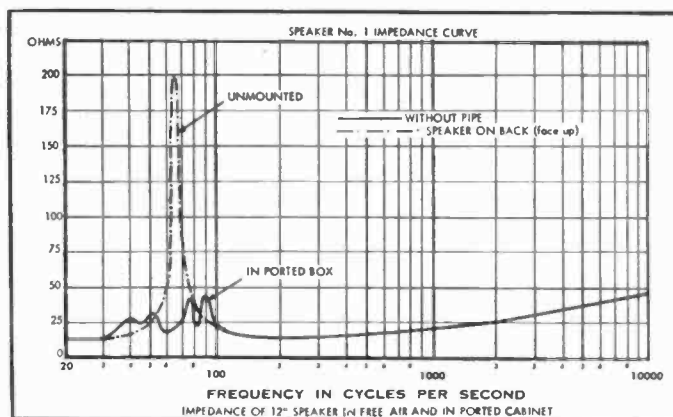

Fig. 3. Impedance of a typical 12-inch speaker in free air and in a ported cabinet.
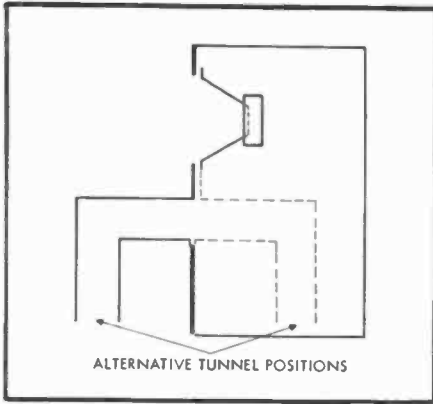
Fig. 5. Cabinet arrangement for experimental determination of effective volume of tunnel.

speaker unit is in fact only that of the iron parts and does *not* include the volume enclosed by the conical diaphragm. An enclosure divided into two half sections by a partition in the form of a thin infinitely flexible and massless diaphragm behaves as a single volume, for the diaphragm offers no obstruction—either resistive or reactive—to the movement of vibrating air particles in the vicinity of the partition. The air volume enclosed by the conical diaphragm is similarly tightly coupled to the volume of the enclosure.

Once again any doubts were resolved by a simple experimental attack. Measurements of resonant frequency of an enclosure of normal volume are insufficient to determine any change due to the insertion of a speaker unit but the use of an enclosure having an internal volume of little more than one cubic foot permits the change in resonant frequency to be accurately determined. The resonant frequency of the small enclosure was determined with and without the speaker unit, the effective volume of the speaker unit being determined by calculation from the two experimentally determined frequencies. It was confirmed by inserting wood blocks of known volume into the empty enclosure to bring the resonant frequency up to that of the enclosure with the loudspeaker.

The effective volume of the speaker determined by these two methods differed from an estimate of the volume of the iron parts, as shown at (B) in *Fig.* 6, by less than 3 per cent and bore no relation to the volume enclosed by the speaker outline.

The foregoing discussion enables the effective volume of an enclosure to be determined leaving the port area A and the effective length of the port or tunnel L to be determined. Unless a port of slit shape is adopted the effective area is the same as the physical area and it thus presents no difficulty in its determination.

The effective length of the tunnel may, and generally does, differ appreciably from its physical (i.e. measured) length. Helmholtz resonator theory presupposes that the air in the tunnel undergoes a translatory motion along the tunnel but that the air particles immediately outside the tunnel ends are stationary. This is clearly an oversimplification but an accurate mathematical determination of the effect of the air movement outside the tunnel is a difficult process that has exercised many investigators because of its importance in determining the effective length of an organ pipe.

Movement of the air outside the ends of the tunnel will clearly increase the mass of air in resonant motion and result in the effective length of tunnel being greater than the physical length. Rayleigh has proposed to allow for these "end effects" by adding an end correc-
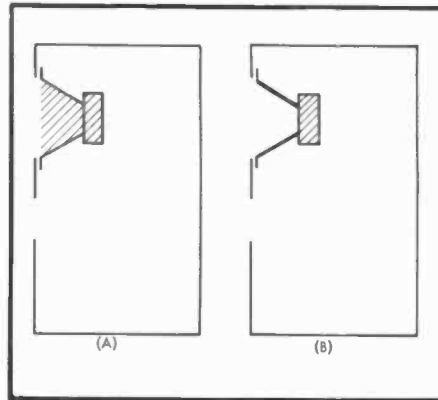

Fig. 6. Effective volume of loudspeaker unit.

tion $l_c = 0.4D$ to the measured length of tunnel and his proposal is confirmed at least for measurements of engineering accuracy. D is the diameter of the port if circular or the diameter of the circle having the same area as the port where the port is non-circular. The effective length of tunnel L to be used in equation (1) is therefore the measured length l_m plus the correction l_c. Where the port is the chamber wall thickness only the effective tunnel length will differ from the measured tunnel length by an appreciable amount for the correction length l_c, being a function of port area only, becomes greater than the physical length. With the modifications discussed, the equation presented by Boswell and Planer appears to predict the value of enclosure resonant frequency with an error of less than 2 per cent when any simple form of construction is employed.

The design procedure based on the Planer and Boswell equation is presented in the form of a single set of curves in *Fig.* 7, which covers the design of any size of speaker unit in any size of enclosure and with any value of enclosure resonant frequency.

Other Variables

The resonant frequency is, however, a function of (among other things) the enclosure shape. Thus when a spherical resonator with a circular opening is employed, the resonant frequency is determined almost entirely by the volume V and port area A and may be accurately calculated. At the other extreme a chamber in the form of a long narrow pipe has a resonant frequency which is deter-

Fig. 7. Design relations of ported enclosures.
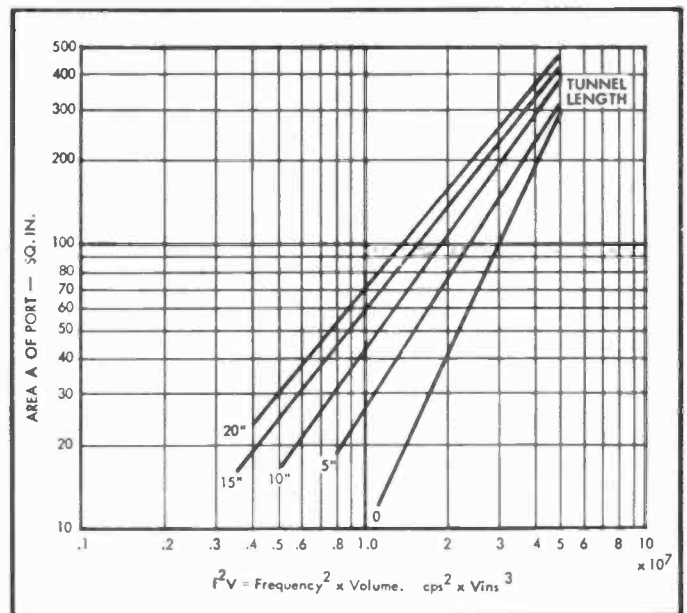
mined almost entirely by the pipe length, the volume V and the port area A having second order effects only. In practice cabinets do not stray too far from a cubical shape for which Planer and Boswell equation for f_r is quite adequate when the inevitable drift in the speaker resonant frequency is taken into account.

Experience suggests that almost any addition to the cabinet interior affects the measured resonant frequency, such apparently minor factors as the relative position of port and speaker or the provision of an isolating shelf between port and speaker shifting the resonant frequency by a few cps. If it is desired to build a cabinet having an accurately determined resonant frequency some final adjustment to enclosure volume or port area must be made after the unit is completed.

The gain in power output at the lower end of the frequency range is determined by, amongst other things, the Q of the enclosure, a factor that it is impossible to calculate with any pretensions to accuracy (Q for a cabinet has the same meaning as the Q for an electrical circuit, being the ratio of stored energy to dissipated energy per cycle). Energy is stored in the enclosed air volume and dissipated as sound or in frictional or viscosity losses in the cabinet and lining structure. The fraction dissipated as sound power is small, structural dissipation accounting for the majority of the losses. Structural losses in an airtight enclosure are largely due to flexion of the chamber walls and are therefore affected by the material used in the construction, but over a fair range of woods experience seems to indicate that the Q falls between 3 and 6. A high value of Q leads to "cabinet hangover" any low-frequency transient having a low-frequency tail oscillation added to the original, giving a soft and rather flabby character to the reproduction. The enclosure Q cannot be reduced to unity or there would be no advantage in using an enclosure, and so the final value must be a compromise to suit personal tastes of the user.

Acoustic Damping

Damping may be added in many ways, the most popular being the use of hair felt, fibreglass or some similar absorbent attached to the walls, though more recently it has been realized that the absorbent material is largely ineffective if mounted on the walls where the air particle velocity normal to the surface is substantially zero. Some consideration of the reason for including the damping will indicate the best position for mounting it in the enclosure.

The added damping really has two duties to perform, it decreases the Q at the resonant frequency to the desired value and it provides sufficient absorp-

tion at frequencies above twice the resonant frequency to attenuate the sound output from the port. This is essential if severe interference between sound from the front of the cone and the sound from the port is not to occur. Maximum attenuation to sound energy in the maximum number of modes of enclosure oscillation is provided by a single sheet of absorbent material suspended from the front left and rear right top corners and fastened down to the rear left and front right bottom corners. Maximum attenuation at the basic resonant frequency is given by a sheet of absorbent material across the port for at this point the velocity of the air particles is a maximum.

The thickness and character of the absorbent used across the port is considerably more critical than when it is suspended inside the cabinet. Absorbent material suspended inside the cabinet is effective as a high resistance shunted across the parallel resonant system formed by the box volume and port volume whereas an absorbent diaphragm across the port is effective as a low resistance in series with the effective inductance of the port. Thus an enclosure requiring 15 sq. ft. of half-inch felt attached to the walls will be damped to the same degree as the basic resonant frequency will be by a single thickness of calico across the port.

Both methods of damping may be employed usefully, a length of felt or fibreglass sheet being suspended in the cabinet to deal with the higher frequency modes of resonance, while the basic resonance is dealt with by absorbent material inserted in the port. Provision of a shelf having a depth of one-half that of the cabinet immediately adjacent to the port between port and loudspeaker considerably attenuates the high-frequency output from the rear of the cone, and is generally to be recommended.

A primary requirement for any loudspeaker enclosure is that it should not add too much coloration of its own creation to the outgoing signal. Some idealists might even suggest that it should add no coloration to the signal but this may be misdirected effort. Every room surface, including the floor, door panels, furniture, each hollow vase, and even the brick walls contribute their own quota of coloration to every sound reproduced in the room. Thus it would appear more reasonable to consider the amount of coloration added by the speaker cabinet against the background of coloration added by the room and its contents. If this is done it will usually be unnecessary to worry about a speaker enclosure of normal construction until the piano is removed from the room.

Adequate freedom from cabinet vibration can usually be secured by the use of half-inch plywood or one of the recon-

structed wood boards for the body, stiffened either by a triangular corner block of 4-inch side, or half a dozen one-inch square struts across from wall to wall. As a third alternative half a dozen $3 \times \frac{3}{4}$ stiffening strips screwed "edge on" across the walls in a random pattern can be recommended. Cross bracing strips or 3-inch deep stiffeners both serve as an excellent support for absorbent material, spacing it well away from the walls.

If the ultimate in vibration-free housings is required, a double walled construction of $\frac{3}{8}$-inch plywood spaced $\frac{1}{2}$ inch apart and having the space packed with dry sand is excellent. An enclosure built from $\frac{1}{2}$-inch plywood with half-inch insulation board glued to the plywood under pressure is less troublesome to construct in some ways, but gives excellent results.

Shape of Cabinet

Enclosure shape is important, though that importance is not revealed by any design equation. Shapes in which one dimension exceeds the others by a large factor are generally to be avoided. Thus, pipes with or without adequate internal damping have always proved disappointing. Irregular-shaped interiors appear to have some acoustical advantage that is not revealed by current practice in measurement, an advantage that is possessed by the triangular corner cabinet. Corner mounting is generally to be preferred both from the acoustical and domestic points of view, for no other shape permits such a large number of cubic feet to be so inconspicuously concealed.

While internal irregularity has its advantages the opposite is true of the exterior. Each external edge and corner produces irregularities in the primary high-frequency response due to diffraction at the surface discontinuity, and though these irregularities are masked to some extent by the generally reverberant sound they should be borne in mind when considering alternative enclosure designs.

The position of the port with respect to the speaker is not highly critical but there is some slight theoretical advantage in placing the port near to the speaker opening. Klapman has shown that the effective radiation resistance presented to a diaphragm is directly proportional to the number of diaphragms if they are all closely associated in space. Close spacing has the disadvantage of increasing the high-frequency radiation from the port unless precautions are taken to prevent it, but if such precautions are taken the balance of advantage is marginally in favor of close spacing of port and speaker openings.

The subject of loudspeaker housings is one of considerable complexity, but it is of such importance as to justify ex-

tended consideration. The present contribution is taken from the chapter on Loudspeakers in a book "High Quality Sound Reproduction" shortly to be published by Chapman & Hall of London, in which the subject of ported cabinets and other forms of speaker mounting are considered in greater detail. All of the illustrations in this article are from the same book.

BAFFLES UNBAFFLED

ways of forming the friction resistance R_f. A convenient method adopted was to make an enlarged port and cover part of it with closely spaced wooden slats having previously determined the required dimensions. Again the slits could be replaced by an aperture covered with a material of suitable porosity such as silk or felt supported between two layers of a ridged metal grid. It was important to prevent the material vibrating with sound pressure.

Further investigation along this line has resulted in the development of an Acoustical Resistance Unit which may be installed directly in a rectangular aperture of the proper size to provide the correct loading with a minimum of constructional work. Such a unit is shown in *Fig.* 35. It consists of a frame on which is mounted a decorative expanded metal grille, backed by a screen which is flocked to a closely controlled pattern so as to provide the proper acoustical loading.

Since this enclosure is capable of reproducing down to 20 cps and is free of resonances above this frequency the bass response is unaccentuated and perfectly natural. If, however, some accentuation is required this may be applied by means of the bass boast control on the amplifier, in which case the excellent air loading applied to the diaphragm will keep distortion to a minimum. This is considered more desirable than producing accentuation in the enclosure itself since with the exception of a fullsize horn the latter cannot possibly provide the same carefully controlled accentuation as can an amplifier having a well designed boss lift control working in conjunction with a loudspeaker system with a high degree of aperiodic control down to zero frequency.

From its mode of operation this enclosure was named the Friction Loaded Enclosure, and we do feel that it represents a notable advance on previous types.
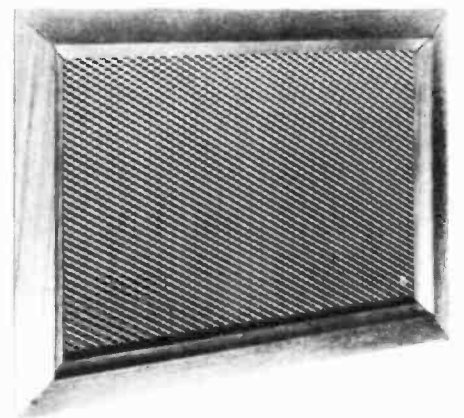
Fig. 35. Commercial form of Acoustical Resistance Unit designed to mount directly into rectangular opening in cabinet of specified dimensions.

The "CW Horn"

A Constant-Width Folded Exponential Loudspeaker Horn

A loudspeaker enclosure of modest size which achieves the degree of realism necessary for high-quality sound reproduction. The secret?—a folded exponential back-loading horn which gives smoothness and an extended low-frequency range to its complement of one or two 8-inch speakers.

D. P. CARLTON

THE CONSTANT-WIDTH FOLDED EX-PONENTIAL HORN which is described herein has been developed in a series of intermittent experiments which have continued since early 1942. My interest was aroused in 1939 by Combs[1] article on a homemade exponential horn. It was built to back-load an 8-inch speaker. I assisted in the construction of several Combs horns and my colleagues and I were impressed by the unusually fine, low-frequency response, the over-all clarity and crispness of the tone, the unusual sensation of presence, and the

[1] E. E. Combs, Jr., QST December 1939.

high efficiency of the cone and back-loading horn combination. The goal of my experiments has been to develop a smaller enclosure with a less complicated horn maze which would retain and perhaps surpass the fine qualities of the "Combs Horn." The result is shown in *Fig.* 1.

Out of the experiments has come a folded exponential horn designed with a constant width between its two parallel sides. It has a rectangular cross-section the height of which varies exponentially, along its length, from the throat to the mouth. This type of horn has been given the name "CW Horn". In such a horn, once the length and rate of flare have

been chosen, the width may be changed from horn to horn without altering the exponential features.

The basic unfolded form of the "CW Horn" is shown in *Fig.* 2. The values for L, W, h_t and h_m are found in the next paragraph. The horn must be folded for two very practical reasons. The most obvious reason is to fit the horn into a box of convenient size and shape. The second reason is equally important. The horn itself transmits only the lower range of frequencies. However, by properly folding the horn the driving speaker may be used as a direct radiator for the important middle range of frequencies.

Fig. 1. Vertical mounting of small unit described makes a suitable high-quality speaker for the smaller room, as shown at left. For different decor, the horizontal mounting, right, may be preferred.
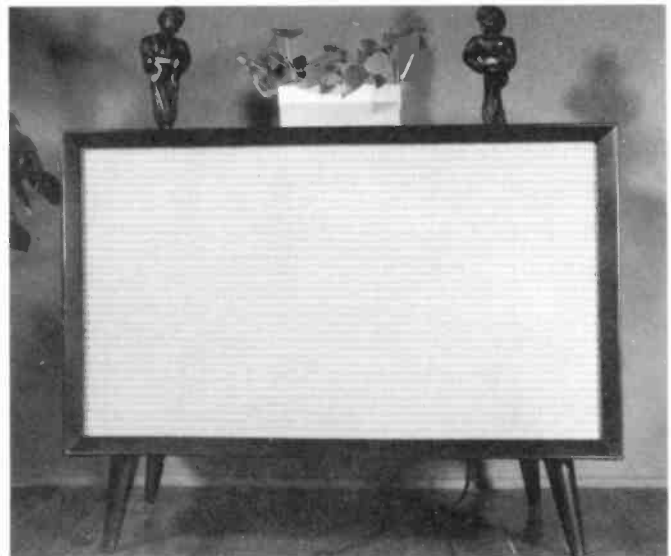
Figure 3 shows the folded exponential horn incorporated in this enclosure has a length, L, of 6 feet and a constant width, W, of 17 inches. The cross-section is rectangular. The height of the cross-section varies exponentially from 1¾ in., h_t, at the throat to 10 in. h_m, at the mouth. The corresponding cross-section areas are 30 and 170 square inches. According to McLachlan[2] a horn of these dimensions would have a low-frequency cut-off of 25 cps. This horn was deliberately designed so as not to limit the low-frequency of any 8-in. speaker that might be used with it.

In the present version of the CW Horn, a length L was selected so that when acting as an open organ pipe it would have a fundamental frequency f_r of about 90 cps to avoid any horn resonance near the principal hum frequencies of 60 and 120 cps. Using the formula, $f_r = 1100/2L$, (where f_r is 90 cps and 1100 feet per second is the velocity of sound in air), the calculated value of L is 6 feet.

The arrangement of the partitions, *Fig.* 3 shows the driving chamber to

[2] McLachlan: *"Loud Speakers,"* p. 183. Oxford Press.
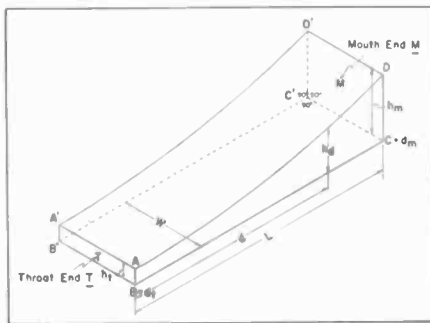
Fig. 2. Development of the exponential horn follows recognized principles.

which the speakers are to be attached as the folded exponential horn. Although the external dimensions are modest, the enclosure gives exceptional performance when used in high-quality home music systems. It should be noted that the present configuration of the parts makes use of almost the total cubic content of the cabinet and produces an inherently rigid structure. The material required to build the box can be cut from a single 4×8 ft. sheet of plywood with less than 5 per cent waste, and the use of a constant width for the folded horn results in remarkable ease of assembly.

Excellent performance may be obtained from the enclosure with one 8-inch cone speaker, preferably one with a rather low natural resonance and a cone surround sufficiently flexible to permit wide excursions of the cone at low frequencies. The low-frequency response will be well down into the lower 40-cycle per second range and it will be real, not the result of doubling. The use of a pair of 8-inch cones will produce a better over-all performance than a single cone, especially in the low-frequency spectrum. When desired, the high-frequency range may be increased by the use of a tweeter.

There are those who ask the question: "Why not use a 10, 12 or 15 inch speaker to drive the horn instead of 8-inch speakers?". The case for the 8-inch cone arises from the fact that because of its shorter radius it will be stiffer and stronger than the larger cones. In operation, it will begin to function as a piston at a higher frequency than the speaker with the larger cones. There also will be less tendency for the smaller cone to buckle or otherwise lose its shape in driving the horn. The smaller, stiffer cone is better adapted to handling the
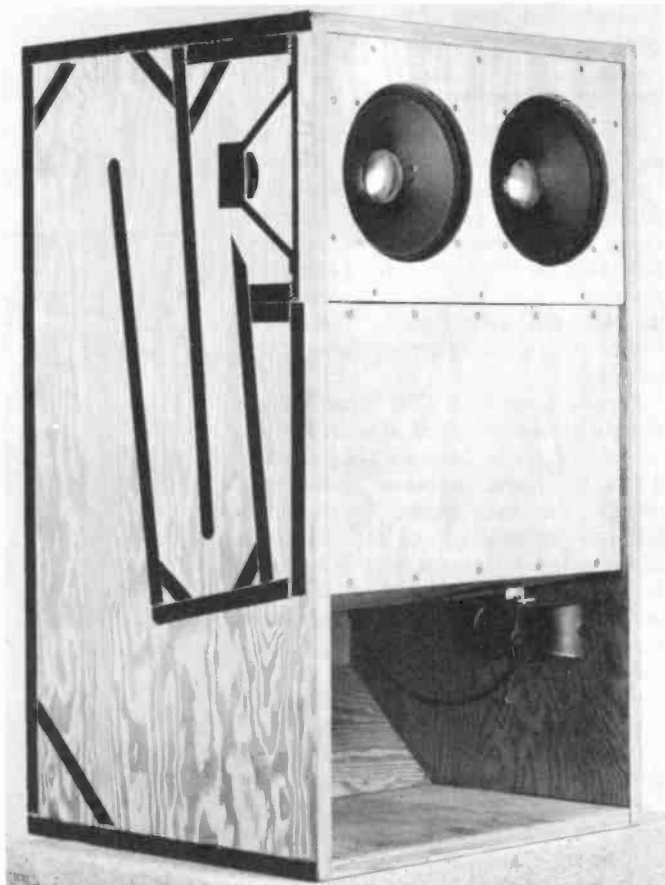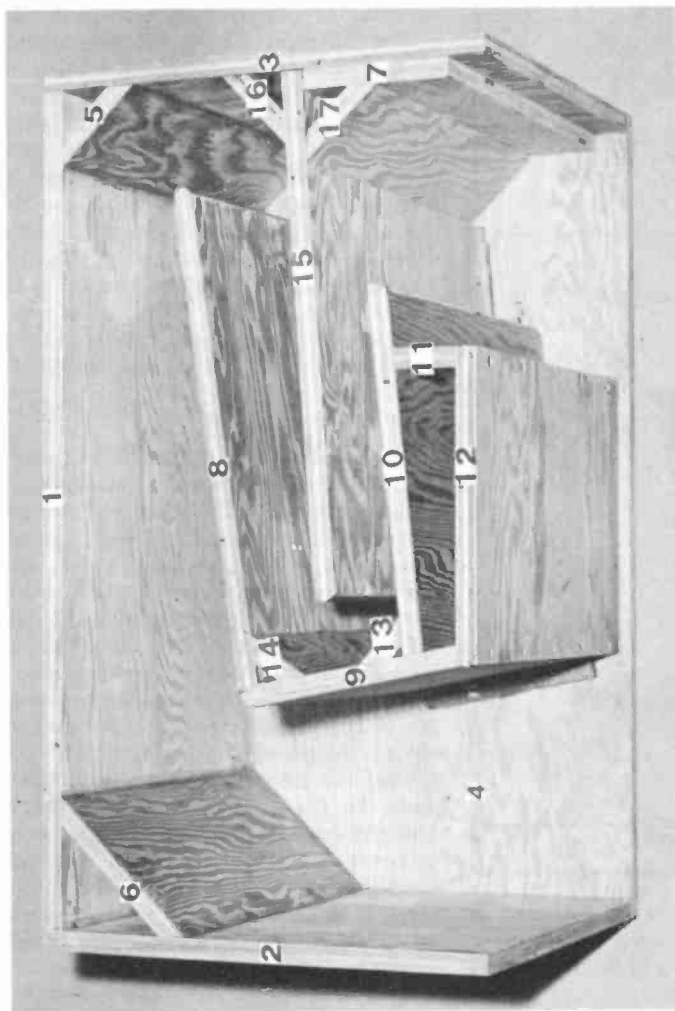

Fig. 3 (left). Internal construction of the folded CW Horn before mounting the second side. Fig. 4 (above). Finished CW Horn with two 8-inch Jim Lansing speakers and a University tweeter. Black tape on side indicates internal construction.
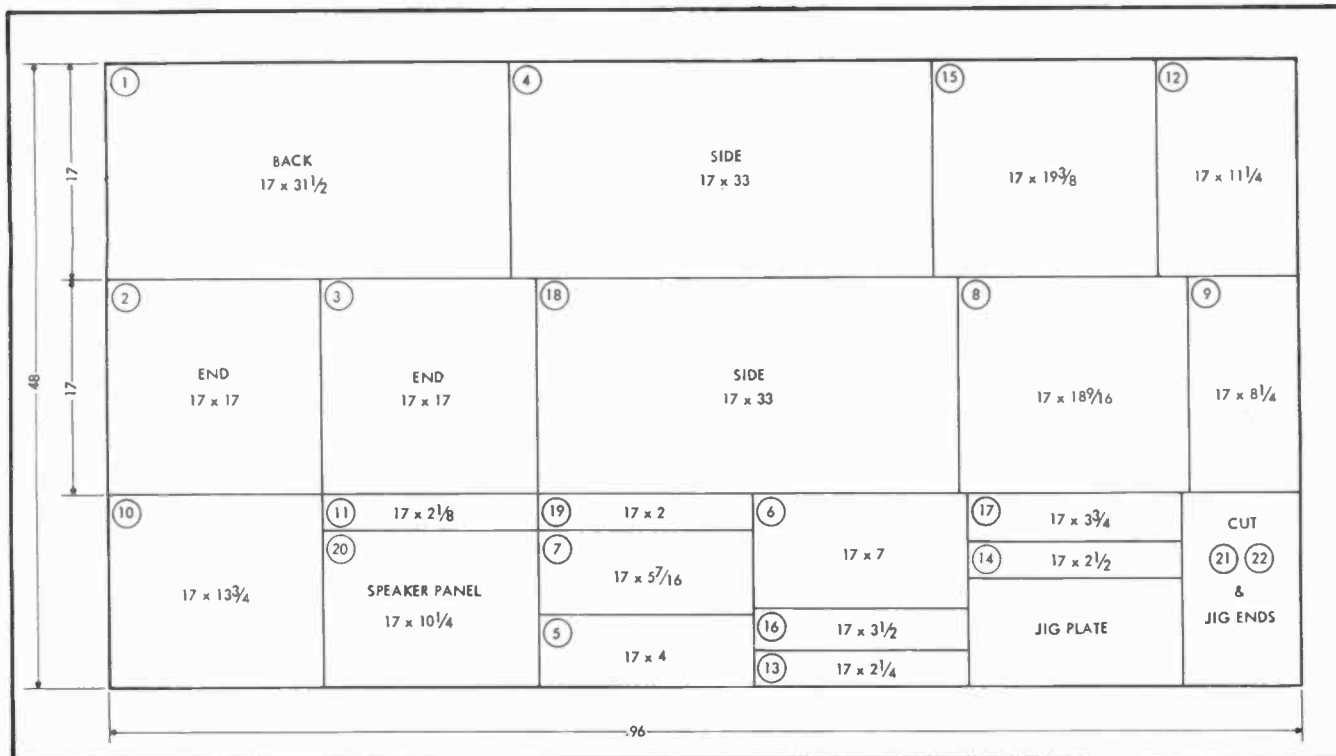
Fig. 5. Layout for cutting a single sheet of ¾ inch plywood into the necessary individual pieces for construction of the CW Horn.

high pressures developed by the long cone-travel required to produce low-frequency sound. The 8-inch cone will also give a superior mid-range response and eliminate the need for a mid-range speaker.

Selection of Speaker

In choosing 8-inch cones, there are many excellent hi-fi units upon the market. Permoflux, Stromberg-Carlson, University, Electro-Voice, Phillips, Jensen, Altec, Lorenz, Stephens, Jim Lansing, and the new PanaSonic are among the more prominent manufacturers. For tweeters you may turn to University, Electro-Voice, Stephens, Jensen, or Lorenz.

Figure 4 shows a CW Horn with a mounted complement of speakers. Upon the side of the enclosure a projection of the horn maze, speaker cavity and speaker has been shown to convey a better understanding of the utilization of the space within the box. Where two cone speakers are used it is necessary to mount the tweeter in the mouth of the horn. When using a single cone speaker and a tweeter both may be mounted on the panel. The orientation of the tweeter is dictated by the choice of the builder as to whether the horn is to be operated in a horizontal or vertical position.

Over the long period of development of the "CW Horn" numerous models have been built. A conscious effort was made to keep the over-all dimensions modest, always reasonably close to those of the described enclosure which, when made of ¾ in. plywood, has external
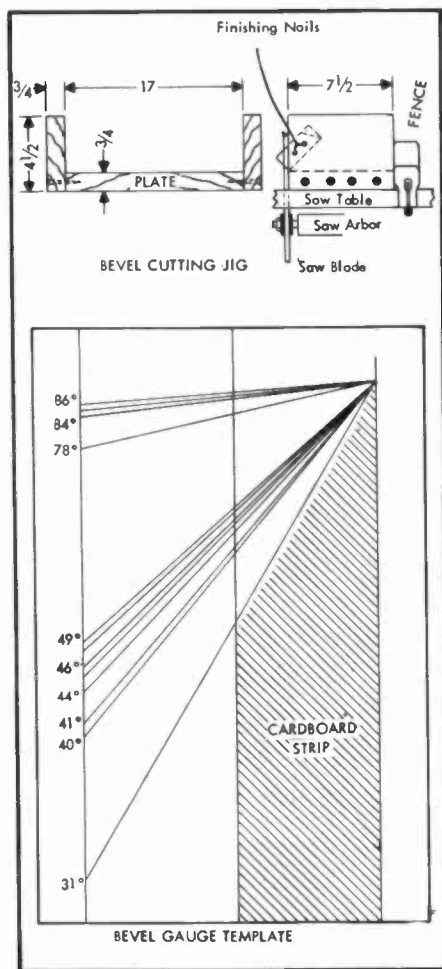
Fig. 6. (A) Layout of the jig designed to facilitate cutting the various bevels. (B) Bevel-gauge template for the angles used.

dimensions of 33 in. high, 18½ in. wide and 17 in. deep. Many varieties of 8-inch cone speakers and a few 10-inch cone speakers have been tried. Different varieties of material in various thicknesses, different lengths of horn, rate of flare, sizes of throat and mouth have been tried. The speaker enclosure which combines the most acceptable size, economy of structural materials and reasonable cost of electronic components with the best over-all performance is the one herein described. Some 50 to 60 of these horns are now being used by local hifi enthusiasts. Make one, equip it properly, and listen! You will marvel at the performance of this enclosure which certainly belies its size and cost.

Construction

The construction of the "CW Horn" is an excellent project for the amateur builder, especially if he has access to a power table saw and a power jointer. The plywood sheet should be cut into the 22 rectangular parts required as shown in *Fig.* 5. The cutting of the bevels where indicated becomes a simple job. Those who have no power equipment may be able to persuade the manager of the neighborhood planing mill to cut up the plywood sheet and make the bevels. The box can be made with hand tools. In this case, use fillets of soft, straight grained lumber, which is easy to bevel with a hand plane. A medium-coarse, flat bastard file, 10 or 12 inches long, is a good tool for making bevels on plywood where a relatively small amount of material is to be removed.

The photographs and construction de-

tails should furnish sufficient data to enable anyone possessed of a reasonable degree of skill to make the "CW Horn" with little or no difficulty. One skilled amateur, equipped with proper tools, built an enclosure in 4 hours; most builders will require 8 or more hours.

Perhaps each builder will see different ways to make the box; however, there are pitfalls that may lead to wasted material and unnecessary expense; therefore, for the first enclosure it is advisable to follow the orderly plan which experience has shown to be good, and which will be found in the paragraphs that follow.

Procedure

Before the required materials are purchased, a choice should be made as to whether to build with screws or 8d finishing nails; there is actually little or no preference.

Cut the plywood sheet (A-D exterior grade is preferable) exactly as directed in the cutting layout, *Fig.* 5 and number each piece properly as it is cut. There will be 22 rectangular pieces, 20 of which have the constant-width dimension of 17 inches. The two exceptions are parts 21 and 22 which eventually become part of the speaker panel seat.

A number of the pieces must be beveled with one or more of the angles indicated at (A) in *Fig.* 6. A simple template which may be used to mark the pieces or set a machine can be provided by using a strip of cardboard 6 to 8 in. long and ¾ in. wide, one end of which can be marked by use of the bevel gauge template, *Fig.* 8, and then the cardboard cut off at the required angle.

The bevel cutting jig of *Fig.* 6 is pictured in *Fig.* 7. It is a proven device for cutting the bevels accurately and *safely* with a power saw. It is especially useful with the bevels varying from 31 to 49 deg. Those bevels which vary from
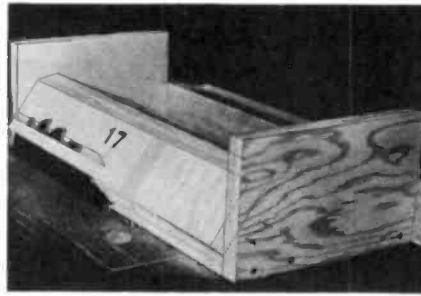
Fig. 7. Photograph of the jig shown in Fig. 6.

78 to 86 deg. can be readily and safely cut on the power jointer.

Having completed the preliminary details of cutting up the plywood sheet, making the beveling jig, and cutting the bevels, preparation for the construction of the box begins with parts 4 and 18 which are hereinafter called the top and bottom "sides," respectively. During construction the horn mouth is usually kept to the left, and the horn is built with the right side down—hence the terms top and bottom "sides."

Lay the bottom side (4) on the bench, smooth surface down. Draw on its upper "rough" surface the data shown in *Fig.* 8 with the exception of the fillets 5, 6, 13, 14, 16, and 17. Locate accurately points A, B, C, D, E, F and G. Draw the lines AB, CD, EF and AG. These lines control the positions of parts 8 through 14 and 15 through 17. Spot the position of the holes to be drilled for the screws or nails which are to hold the box together.

Builders may find it desirable to make a complete full scale drawing of *Fig.* 8 on tracing paper or brown wrapping paper and by means of carbon paper use it to transfer pertinent data to the top of part 4.

Lay part 18 on the bench, smooth surface up, and place part 4 on top of it

with the smooth surface down. Clamp these two pieces together and drill the screw or nail holes, marked on part 4, through both pieces simultaneously.

An airtight horn is essential. Therefore, use glue generously and plenty of nails or screws. A small can of Reardon's or Walton's water putty will be found very useful for filling in voids which are often found in plywood. These materials do not shrink upon drying and both are excellent putties.

Assembly

Experience has shown that the actual assembly of parts may be divided into several well-defined phases which are as follows:

Phase 1: Fasten parts 1, 2 and 3 together with plenty of glue and nails and in turn fasten part 4 to their lower edges. Check with a square for rectangularity and then fix parts 5, 6 and 7 in place as in *Fig.* 9. The front edge of part 7 will eventually be part of the speaker panel seat while the back edge will control the position of the right end of part 15.

Phase 2: This involves parts 12, 11, 9, 10, 13, 8, and 14 (*Fig.* 9), which are assembled in that order. The use of temporary cleats M, N, R, S, T, and U bradded directly to the upper surface of part 4, will be a great help in keeping these pieces in their proper positions. Alternatively, this group of parts may be assembled on a bench or another piece of plywood in a duplicate set of cleats and the assembled parts set into place as a whole on the upper face of part 4 after Phase 3 is completed, *Fig.* 3.

Phase 3: This involves assembly of parts 15, 16, and 17 in place. Temporary cleats O and P, *Fig.* 9, are used to keep the left end of part 15 in its proper position.
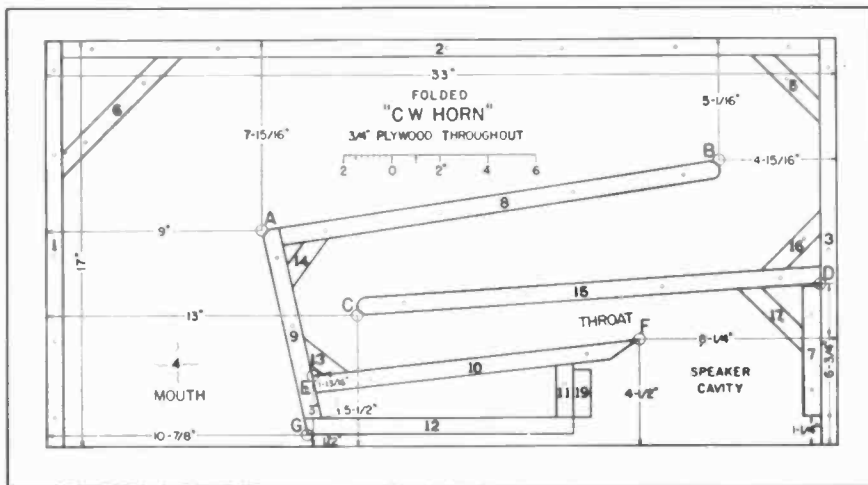
Fig. 8. Assembly of the horn, with pieces numbered in accordance with Fig. 5. Layout dimensions are shown.

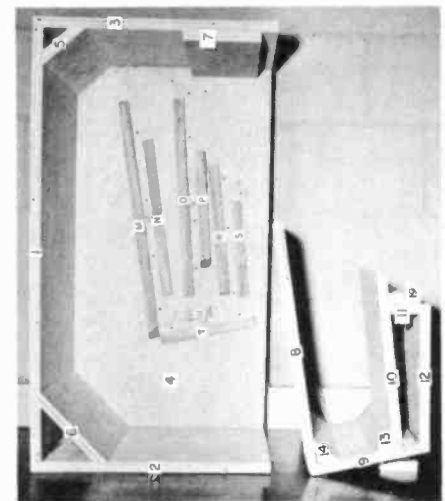Fig. 9. Method of simplifying the assembly by use of temporary locating strips.

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Phase 4: Lay into the maze the woofer and tweeter leads stapling them to part 10 near the throat edge so that they will not fall back into the horn. The lead to the amplifier should be led out through the center of the back through a hole in which the cord fits snugly.

Phase 5: At this point, the top edges of the pieces assembled on part 4 should be checked. Their upper edges must lie in a single plane. Irregularities should be smoothed out to insure that when part 18 is applied, a continuous air-tight seal will result. Apply glue generously to the upper edges of these pieces and then fasten the upper side, part 18, into place with screws or nails.

Phase 6: Complete the box by fitting parts 19, 21, and 22 which together with part 7 form the speaker panel seat. Make sure that the seat is flat. Apply the necessary gasket material to assure an air-tight seal when the speaker panel is fixed in place.

Phase 7: Bolt the speakers to the speaker panel using gasket material on their rims to assure airtight seals. Connect the speakers, being careful that the cones are properly phased. Fasten the speaker panel into place with screws and the enclosure is ready to use. The mechanical design of various loudspeakers will dictate the design of the speaker panel. It may be necessary to use thinner plywood for this part (20) in order to avoid the possibility of the speaker rims or speaker panel projecting past the front edges of the box. It is a simple task to mount a pair of 8-inch speakers with their rims outside the panel. When the rims must be mounted on the inside, as is necessary with Jim Lansing speakers, make two panel boards exactly the same length and width with thicknesses of $\frac{1}{4}$ and $\frac{3}{4}$ inch. On the thicker one cut two holes which will be tangent at the center of the panel and have radii equal to the radius of the speaker rim. On the thinner one cut two holes with $3\frac{1}{2}$ in. radius making sure that when the two pieces are fastened together the holes in the two different panels will be concentric. The $\frac{1}{4}$-in. piece mounts the speakers, the $\frac{3}{4}$-in. piece provides an effective seal on the panel seat.

The enclosure as it has been developed was designed to fit into a cavity in an existing radio-phonograph console. It can be used in a console especially designed for it or to fit into a cavity in the wall of a home. When specifically providing a cavity for one of the CW Horns allow a space of $\frac{1}{2}$ in. or more completely around the sides, ends, and front, in order that the enclosure may rest on two $\frac{1}{2} \times 1 \times 16$ in. strips of felt or rubber. This provides a measure of isolation which is important in preventing acoustic feed-back.

THE CW HORN
(Continued from preceding page)

In situations where grill cloth is not needed it is well to tack plastic or glass —not metal—window screen over its front to prevent the box from becoming infested by insects. Caution: in designing grill cloth frames avoid encroachment upon the mouth area of the horn. An excellent hardware for making snap-on-and-off grill frames is to be found in Johnson 108–75A banana plugs which are available in most electronic supply houses.

The enclosure built of fir plywood and held together by nails or screws, the heads of which are visible, needs decoration to produce an acceptable piece of furniture. Certainly most any builder can develop a satisfactory solution to his specific problem. There are many fabrics, plastics and wood veneers which may be used. The box may be laid on its side if you wish. Development of a satisfactory conversion is an interesting project. Two such units at opposite ends of a large equipment cabinet serve as a stereophonic system when connected to a two-channel amplifier; when paralleled, they provide a wide "monaural" sound source.

List of Materials

4' × 8' sheet of ¾-in. plywood.*
½ lb. of glue (Weldwood or Cascomite).
1 Cross #6 or #8—1½" wood screws or 8d finishing nails (1 lb.).
1 Box of 1-in. wire brads.
⅛ lb. of 4d finishing nails.
Gasket material: 1/16- or ⅛-rubber or cork for speakers and speaker panel. Sheet 12" × 36".
☆ A-D Exterior grade preferred.

List of Parts

		Bevel:
1—17" × 31½"	Back	
2—17" × 17"	End	
3—17" × 17"	End	
4—17" × 33"	Side	
5—17" × 4"	Fillet	45°–45°
6—17" × 7"	Fillet	45°–45°
7—17" × 5-7/16"		
8—17" × 18-9/16"		85°
9—17" × 8¼"		78°–85°
10—17" × 13¾"		31°–84°
11—17" × 2⅛"		84°
12—17" × 11¼"		78°
13—17" × 2¼"	Fillet	40°–44°
14—17" × 2½"	Fillet	46°–49°
15—17" × 19⅜"		86°
16—17" × 3½"	Fillet	41°–45°
17—17" × 3¾"	Fillet	45°–48°
18—17" × 33"	Side	
19—17" × 2"		
20—17" × 10¼"	Speaker Panel	
21—8¾" × 1¼"	Panel End Support	
22—8¾" × 1¼"	Panel End Support	

The Aperiodic Loudspeaker Enclosure

E. de BOER

The author analyzes the requirements of a resonance-free woofer enclosure and then proceeds to design a practical cabinet that fulfills these requirements to a maximum degree.

THE REALIZATION of adequate reproduction of low frequencies is a well-known trouble of high fidelity. This difficulty is mainly due to the low radiation efficiency of cone-type loudspeakers at these frequencies. More specifically, when the dimensions of the radiating member are small compared to the wavelength of the emitted sound, a poor match exists to the air load. This not only causes trouble in obtaining a flat frequency response but a severe limit is imposed upon the maximum power as well.

The only way to meet both design objectives is to increase the radiating area. Since this is not feasible beyond a certain limit, the loudspeaker enclosure is designed to improve the match between diaphragm and air load. It can be stated quite generally that an improved match over a large frequency range can only be attained by a system occupying a large volume. In a small-sized cabinet, where resonance is used to reinforce the bass end, the improvement is present over only a very narrow frequency range.

One can conclude from this that for a given maximum volume one has to concentrate either upon obtaining flat frequency response with limited power capability or upon an improved power match resulting in a jagged frequency response curve. Many solutions are actually situated midway between these extremes. The only one of these extremes which can be called satisfactory is the first one. The author believes that there may be interest in an enclosure which has been designed to obtain a smooth frequency response, the power requirements being completely put aside.

It has recently been shown[1] that under certain circumstances a small enclosure is theoretically capable of a flat frequency response toward some 20 cps. Provided with a large-diameter loudspeaker such a system can produce sufficient power for normal listening conditions. A similar type of reasoning
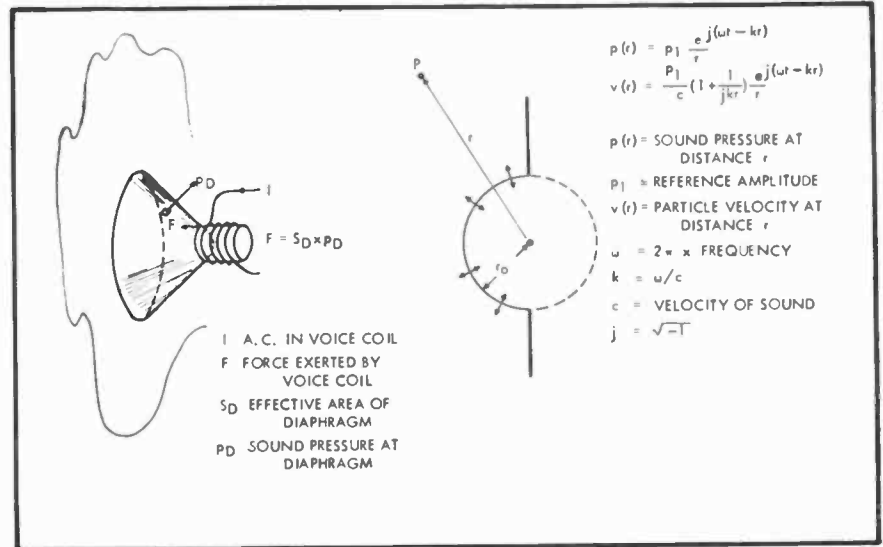
has led to the Acoustical Suspension System.[2] These systems have in common that they demand a rather critical choice of parameters of the loudspeaker as well as of the enclosure.

The author has developed a small-sized enclosure which can be used with a normal type of loudspeaker and is not critical. The design objective has been to approximate the behaviour of the loudspeaker in an infinite baffle with the requirement that the fundamental resonance be critically damped. This has been realized by adding two adjustable parameters to a system resembling a bass-reflex system.

Electrical Analogue of a Loudspeaker

Let us discuss first the performance of a loudspeaker on an infinite baffle. For reasons of simplicity we first assume that the diaphragm has no mass and is freely suspended. When a constant current I passes through the voice coil, a force F is exerted on the diaphragm. This force can be balanced only by acoustic forces, so the sound pressure p_D on the diaphragm is constant with frequency.

For low frequencies, the radiating sys-

tem can be converted into a pulsating sphere radiating into the left half-space without introducing too large an error,[3] as shown in *Fig.* 1. From the given formulas for the sound field it can be seen that the amplitude of the sound pressure is inversely proportional to the distance. Hence our hypothetical loudspeaker (fed by a constant current I) has the property that the pressure response at a given point in the radiation space is independent of frequency.

This simplified theory shows why an actual loudspeaker may have a reasonably flat pressure response even when

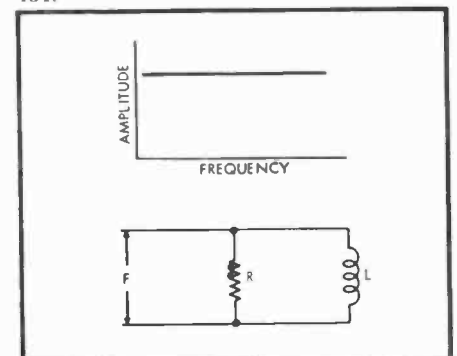[3] F. V. Hunt, "*Electroacoustics.*" Fig. 5.7 (page 158) and related text on page 151.

Fig. 1. Hypothetical loudspeaker in an infinite wall.

$$p(r) = p_1 \frac{e^{j(\omega t - kr)}}{r}$$

$$v(r) = \frac{p_1}{c}\left(1 + \frac{1}{jkr}\right)\frac{e^{j(\omega t - kr)}}{r}$$

p (r) = SOUND PRESSURE AT DISTANCE r

p_1 = REFERENCE AMPLITUDE

v (r) = PARTICLE VELOCITY AT DISTANCE r

ω = 2π x FREQUENCY

k = ω/c

c = VELOCITY OF SOUND

j = $\sqrt{-1}$

I   A.C. IN VOICE COIL
F   FORCE EXERTED BY VOICE COIL
S_D EFFECTIVE AREA OF DIAPHRAGM
p_D SOUND PRESSURE AT DIAPHRAGM

$F = S_D \times p_D$

Fig. 2. Electrical analogue of the system of Fig. 1.

[1] E. J. Jordan, "Loudspeaker enclosure design" *Wireless World*, Jan. and Feb., 1956.

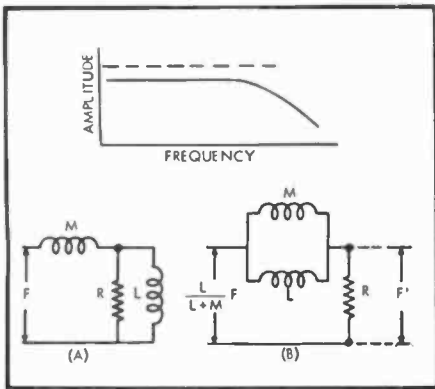[2] E. M. Villchur, "Revolutionary loudspeaker and enclosure" AUDIO, Oct., 1954, and July, 1955.

Fig. 3. Loudspeaker with mass (M) reaction included.

the dimensions are smaller than the wavelength.

For a discussion of the power capability we are interested in the amplitude A of the diaphragm excursions. The diaphragm velocity $v_D = j\omega A$ is related to the force F by the concept of mechanical impedance z_M defined as

$$z_M = \frac{F}{v_D}$$

From the formulas given in *Fig.* 1 it follows that

$$z_M = S_D.\rho c\, \frac{kr_o}{kr_o - j}\ \text{where}\ k = \omega/c \quad (1)$$

For low frequencies z_M is nearly proportional to frequency; hence the amplitude A is nearly inversely proportional to the square of frequency in order to maintain a constant sound pressure. This is the reason sound power at low frequencies is severely limited.

In order to visualize more clearly the operating conditions of actual loudspeakers we will refer to an electrical analogue of the mechanical system. Any mechanical force will be represented by an electric voltage, and a velocity by a current. A mechanical impedance is then transformed into an electrical impedance, which may be numerically equal to it.

As can be verified easily, impedance[1] can be represented by the electrical impedance of a resistance and an inductance connected in parallel as in *Fig.* 2.
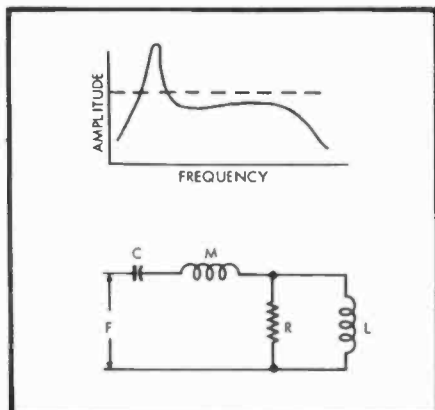
Fig. 4. Loudspeaker with effect of finite compliance (C) included

We refer to this impedance as the radiation impedance of a circular orifice. In the case of our hypothetical loudspeaker this impedance is connected directly to a voltage source representing the constant force F. The voltage across the configuration of *Fig.* 2 now represents also the sound pressure on the diaphragm.

For the spherically symmetrical radiator of *Fig.* 1 it represents the sound pressure at a given point in the radiation space as well. So from our electrical analogue it is easily verified that the pressure response of the loudspeaker will be flat. The radiated power is represented by the dissipation in R. The inductance L represents a certain mass of air vibrating in the neighborhood of the diaphragm without dissipation (radiation mass).

We have now acquired the skill to study the performance of actual loudspeakers from the electrical analogue. The mass, stiffness, and so on of the diaphragm are added to the analogue and the pressure response will still be given by the voltage across the radiation impedance of *Fig.* 1. The addition of these extra, though unavoidable, elements cause the frequency response to deviate from the ideal one.

First we assume that the diaphragm has a non-vanishing mass but still infinite compliance. Since the force F has to be divided between mass and air load the mass reactance appears in series in the electrical circuit. See (A) in *Fig.* 3. By way of the equivalent configuration of (B), the pressure response is easily calculated. Here F represents the mechanical force and F' the part of it that is effective in radiation. One can see that the introduction of mass reduces the efficiency and causes the higher frequencies to be reduced in strength.

Now we add the feature of a finite compliance. This appears in *Fig.* 4 as a capacitor (proportional to the compliance), again in series with the radiation impedance. The resulting configuration gives rise to a series resonance which actually lies in a frequency region where the damping by R is very small. This means that this fundamental resonance will not be materially damped by acoustic radiation.

The damping arises mainly from mechanical friction and from the electromagnetic coupling to the driving amplifier. The friction can simply be represented as a series resistor. The electromagnetic damping can also be studied from the electrical analogue but the derivation is somewhat more involved. Suppose for a moment that the voice-coil resistance is zero and that the amplifier is a pure voltage source. The motion of the diaphragm will then be such that the induction voltage developed in the voice coil just balances the impressed voltage. The diaphragm velocity
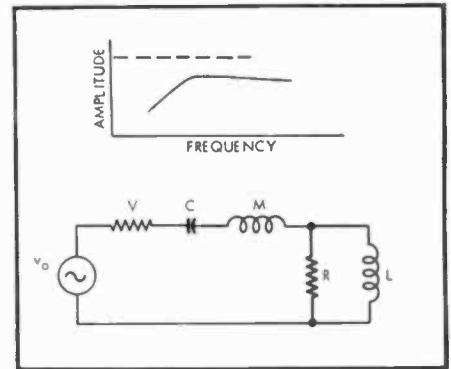

Fig. 5. Velocity-controlled loudspeaker.

is completely controlled by the amplifier and the driving system is represented in the electrical analogue *Fig.* 5, by a pure current source. From this it is apparent that the low-frequency response drops 6 db per octave. This is a well known property of high-efficiency speakers driven by an amplifier with low internal impedance. The actual situation will be somewhere between those pictured in *Figs.* 4 and 5. Since amplifiers are frequently designed so as to deliver a nearly constant output voltage, the driving system is preferably depicted as in *Fig.* 6, consisting of a current source v_0 shunted by a resistor Q. The velocity v_0 is the velocity attained by the diaphragm when the total ohmic resistance would be zero. The current consumption by Q is a measure of the actual voltage drop over the internal resistance of the amplifier and the voice-coil resistance, v. Hence the numerical value of Q is inversely proportional to this total resistance. From *Fig.* 6 it is apparent that the damping is largest when Q is large, that is, when the total resistance in the voice-coil circuit is low.

Loudspeaker in Enclosure

The electrical analogue serves as an important tool toward understanding the action of more complicated systems. When a loudspeaker is placed in an enclosure the diaphragm has to exert additional forces. Hence the elements due to the enclosure appear in series with the diaphragm impedance of *Fig.* 4.
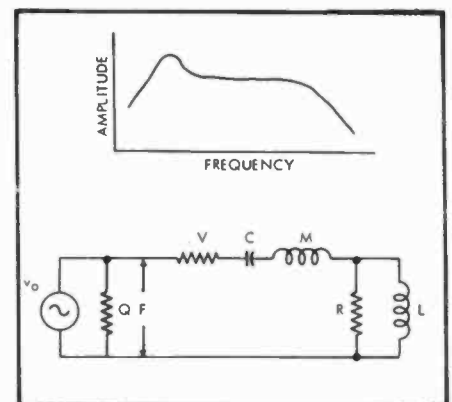

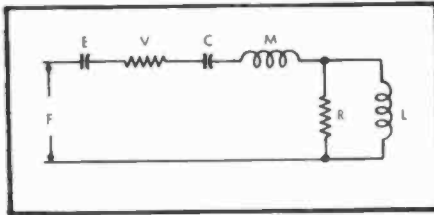Fig. 6. Loudspeaker with electromagnetic damping.

Fig. 7. Loudspeaker in completely closed box (E represents stiffness of enclosed volume of air).

Because of the finite compressibility of air a completely closed box acts as a stiffness, at least for low frequencies. The analogous capacitor E appears in series with the diaphragm stiffness C, thus increasing the resonance frequency (*Fig.* 7.) When the enclosure is provided with a port, the situation becomes much more complicated. Under the assumption that the port radiates without interaction with the diaphragm, its impedance is again represented by an inductance L' and a resistance R' in parallel. In order to find the proper places of this impedance and the enclosure's stiffness we note that the air flux issuing from the rear side of the diaphragm can choose between two ways. Part of the flux leaves the enclosure via the port, the rest is effective in building up sound pressure inside the box. This two-way choice can be represented by a parallel connection of the capacitor E and the port's radiation impedance as is illustrated in *Fig.* 8. The parallel configuration can be defended as well by noting that the pressure inside the box must be the same on all places. Thus the port is driven by the same sound pressure as the enclosure's stiffness. In the electrical analogue both impedances must then be connected across the same voltage, hence they must be connected in parallel.

The circuit of *Fig.* 8 gives the generally accepted electrical analogue of what is easily recognized as a bass-reflex system. In two respects, however, the circuit is deficient. First, the interaction between diaphragm and port is neglected. Proper introduction of this effect makes the circuit much more complicated. In the second place the radiation of the rear side of the diaphragm is 180 deg. out of phase with that of the front side. In our qualitative discussion we assume that both effects do not have a large effect on the various pressures and ve-

locities appearing in the circuit. For the acoustical response, however, we may state that the sound pressure at large distance is approximately represented by the difference of the voltages across L and L'.

The performance of the bass-reflex system will be analyzed only briefly. At the free-air resonance frequency of the loudspeaker nearly the whole force F appears across the parallel section. When this section resonates at the same frequency, full radiation is retained while at the same time a large mechanical impedance is presented to the diaphragm. This advantageous situation is only present over a small frequency range, however.

At lower frequencies the parallel section becomes inductive and the series section capacitive, so that a series resonance develops. At frequencies higher than the parallel resonance, the parallel section becomes capacitive and the series section inductive so that a second series resonance of the whole circuit appears. At these two resonances the diaphragm velocity becomes maximal. The resulting maxima of the induction voltage developed in the voice coil cause the well known double hump of the voice-coil impedance which is specific to the bass-reflex system. The opposite polarities of front and rear waves cause the lower resonance to be reduced in amplitude. A typical theoretical response is shown in *Fig.* 8.

Damped Enclosures

The bass-reflex system provides us with an easily realizable configuration of mechanical impedances due to the enclosure. Over a restricted frequency range an improved match between diaphragm and air load is obtained by way of the port radiation. This occurs, however, at the expense of the response at other frequencies.

In our design we will stick to this type of configuration, though the requirements are completely different. The impedance of the enclosure will be used solely to control the diaphragm's motion. The port will not be used as an alternative radiator of sound. This is obtained by making the port area exceptionally low. We will try to modify the system
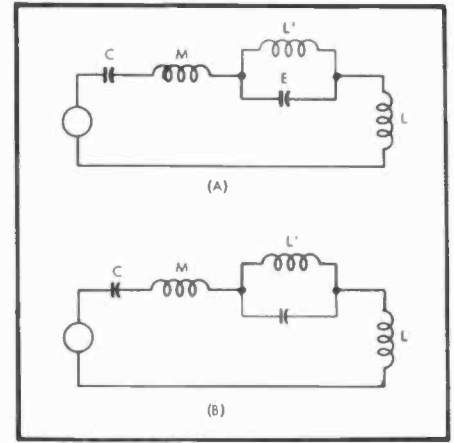

Fig. 9. Origin of main resonances.

so that smooth frequency response is obtained from the radiation of the diaphragm. The unavoidable resonances can only be attacked by the application of damping.

It has been shown[1] that under certain conditions a very small value of the resistance R' can serve this purpose effectively. Such a value can be realized by introducing a port provided with a thick though porous resistive cloth. The resulting enclosure is claimed to have a flat response towards some 20 cps.

Due to the fact that one resistance is used to damp two resonances and one anti-resonance, the system is quite critical, however. In addition, the residual radiation of the damped orifice tends to impair the lower frequency. For these reasons the present author has tried to find a solution which does not suffer from these disadvantages.

Let us focus attention to one of the series resonances of a typical bass-reflex enclosure (provided with a port that is too small). At the frequency of the upper resonance the mass-like impedance of the diaphragm resonates with the stiffness-like impedance of the enclosure. The latter is given almost exclusively by the stiffness of the enclosed volume of air. This series resonance is indicated schematically by the thick line in (A) of *Fig.* 9. This resonance can be
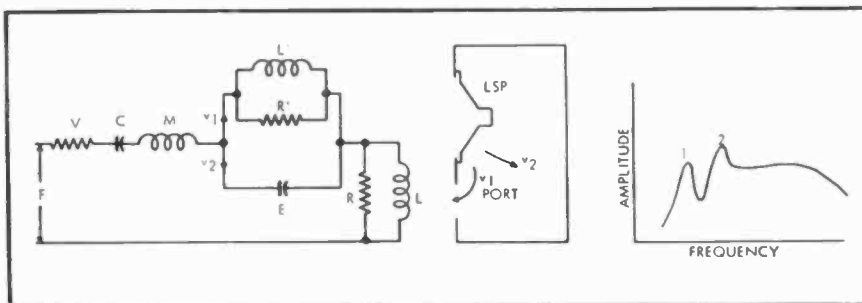

Fig. 10. Loudspeaker in "aperiodic" enclosure.

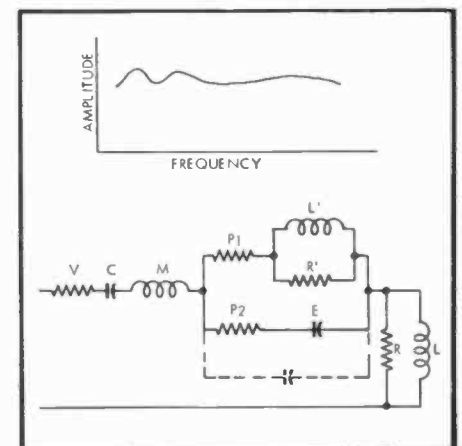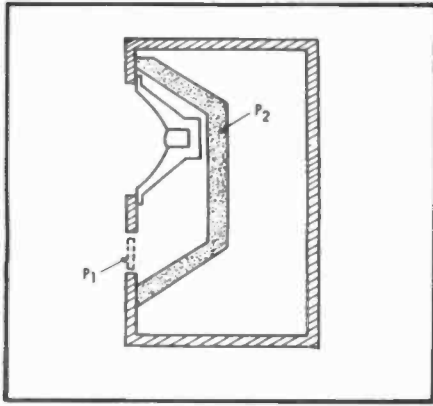Fig. 8. Loudspeaker in box provided with additional orifice.

Fig. 11. The aperiodic enclosure.

damped by a series resistor inserted anywhere in the indicated path.

The other series resonance is due to the stiffness of the diaphragm impedance at low frequencies being balanced by the mass-like character of the enclosure. The current path of this resonance is indicated at (B) in *Fig.* 9. Both resonances can be damped nearly independently by inserting two resistors, one in series with E, the other in series with L' (see *Fig.* 10).

At intermediate frequencies the re-

sonance of the parallel section is met. This resonance tends to cause a dip in the frequency response. It is, however, damped by the two resistances P_1 and P_2. With a proper choice of parameters, correct adjustment of P_1 and P_2 automatically reduces this anti-resonance to the wanted degree.

These parameters have been determined experimentally by way of the electrical analogue. It appeared that the choice of the parameters L' and E is not critical so that a rather low value of E can be chosen. This results in a very low enclosure volume which is, of course, a very attractive feature of this system.

There remains the question whether the obtained system can be realized acoustically. The electrical analogue shows that the resistors P_1 and P_2 have to be inserted in series with L' and E. As regards P_1, the acoustic air flux has to pass through P_1 before developing sound pressure across the port impedance. Referring to *Figs.* 8 and 10 it follows that an acoustic resistance has to be placed in the port.

The placement of the element corresponding to P_2 is somewhat less obvious. The air flux v_1 toward the port (*Fig.* 8)
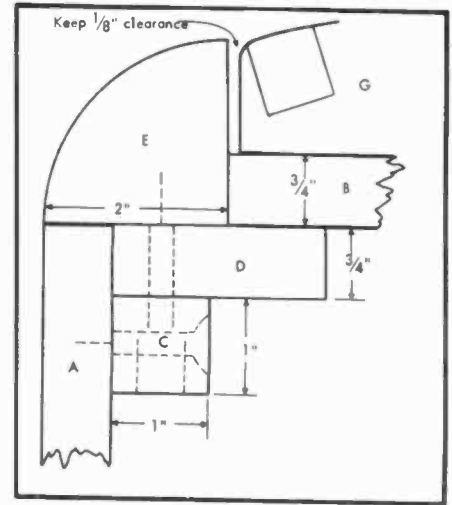

Fig. 13. Detail of front corner construction.

must not be hindered, whereas the flux v_2 has to pass through a resistance.

This situation can be realized by stretching a resistive cloth over the rear of the loudspeaker, so that both the loudspeaker and the port are isolated from the main volume. The obtained enclosure is shown schematically in *Fig.* 11. It has been checked experimentally that this enclosure operates as expected and that the small cavity behind the loudspeaker (see the dotted capacitor in *Fig.* 10) does not introduce a noticeable extra resonance.

Construction of the Enclosure

The external appearance of the enclosure is much like that of a bass-reflex system. The dimensions are chosen quite differently, however. The volume is less than one half the normal value, and the port area is much smaller than the diaphragm. By way of a resistive cloth the interior volume has been divided into two parts (*Fig.* 11). The smaller cavity contains the loudspeaker and opens into the port, the larger one consists of the remaining part of the volume. This dividing member serves to reduce the upper resonance, thus eliminating the most objectionable cause of boom.

The port, here having the form of a narrow slit, is covered by a thin layer of similar material, in order to damp out the lower resonance.

The damping of the resonances can be adjusted by changing the thicknesses of the layers. This adjustment being one of the major stages in the construction, the construction must be such that the interior of the enclosure is easily accessible. When properly adjusted, the system is capable of a smooth frequency response extending to one half octave below the speaker's free-air resonance. The lowest frequencies are somewhat attenuated because of the cancelling action of the residual sound radiated by the
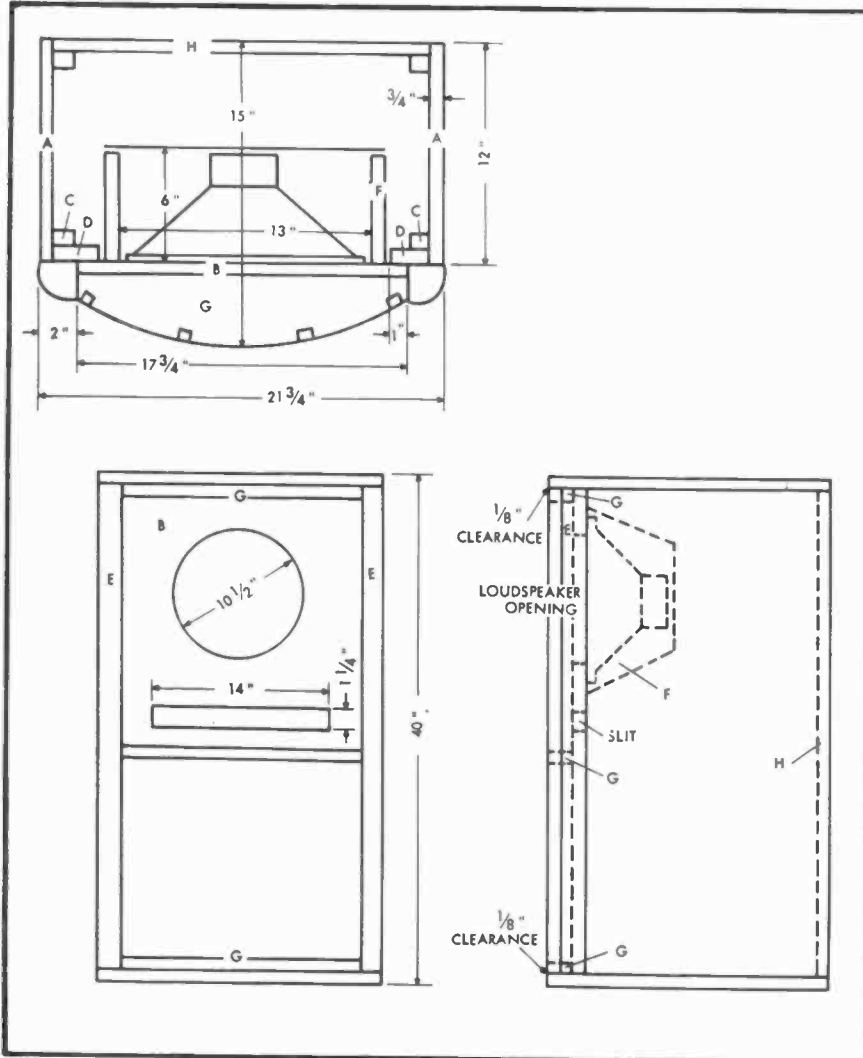

Fig. 12. Constructional data for aperiodic enclosure.

slit. This defect is easily overcome by applying some bass boost in the amplifier.

Full constructional data of a free-standing model of the enclosure is given in *Fig.* 12. With a few modifications the enclosure can easily be converted into a corner model. The construction is intended for use of a 12-inch bass-range loudspeaker. Before ordering wood be sure that the critical dimensions indicated by the thick arrows in the figure are sufficient to accomodate the speaker. We suppose that all panels are pre-sawed to the correct dimensions and that the vertical posts *E* are dressed to a quarter of a cylinder in a cabinet shop. These vertical posts are used to allow the front panel to be shifted a few inches backward. As shown in *Fig.* 13, the space thus created between the posts accommodate a wooden frame *C* that bears the decorative grille used to hide the loudspeaker from view. In this way an excellent external appearance can be attained even by the unskilled home worker. The limited space does not permit us to describe the construction in all details. We will confine ourselves to a few general remarks. See *Fig.* 14.

Before assembling the parts, the side panels *A* are provided with 1″ × 1″ strips along the edges, and with strips *C* and *D* along the front-side edges. The latter strips are used to bear the vertical posts *E* and the front panel *B*. Strips placed on the panels at irregular intervals and in random directions serve for bracing purposes. Needless to say that all joints are glued and screwed in such a way that no heads of screws are visible from the outside. After addition of the vertical posts *E* the enclosure is erected by mounting the front and backside panels provisionally, that is, by means of screws one size too small. Top
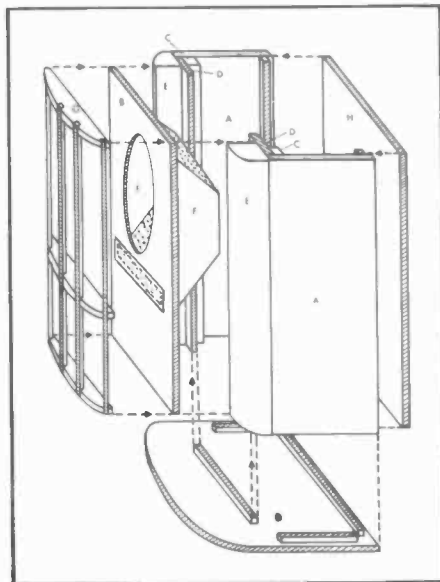
and bottom edges are checked to be exactly level and the corresponding panels are brought into place. All joints are again glued and screwed. After drying overnight the front and backside panels (which have been fastened by screws only) are removed to permit completing the front panel. Two wooden clamps *F* are used to house the perforated metal plate which bears the main resistive member. The terminal wires of the loudspeaker are led out of the cage through holes in these clamps so that they cannot touch hard surfaces when vibrating. The clamps are provided with rubber strips along their protruding edges in order to prevent vibrations of the metal plate. A two-inch layer of plastic foam is tightly sewn to the plate before the latter is secured to the clamps. A similar plate of perforated metal is fitted over the front side of the slit. For the first test run this is not equipped with damping material.

The enclosure is now assembled and tested by feeding it with pure tones. The voice-coil impedance can be determined from the voltage across it when the loudspeaker is fed by a nearly constant current. The test run reveals the usual two resonances of which the upper one appears to be heavily damped. Now the thickness of the absorbing material on the perforated plate is increased until the resonance has about disappeared. The lower resonance as well as the minimum in the impedance curve are still clearly visible (and audible). For arriving at the proper thickness of the material to be placed over the slit, it is necessary to install the wooden frame bearing the decorative grille. This can be screwed on the front panel just by removing the back side panel and screwing from the inside. The layer has to be about ¼ in. thick. When the adjustment has been completed, the upper resonance has completely disappeared, the lower one being still noticeable on measuring, but sufficiently damped by the audio amplifier. All screws, especially those of the loudspeaker, are once more tightly

secured and the enclosure is lined by 1 in. felt. The front and back panels are now mounted by large screws. A sturdy plug, fitted in such a way that no leakage of air can occur, serves as electrical connection.

Before concluding, a few additional remarks. Since this system has been provided with a decorative grille, it is not very suitable for all-range use. It is preferably used as the bass section of a multi-channel system. In this connection it is not superfluous to mention that the efficiency of the bass section has been reduced by the applied dampings, so that introduction of attenuator pads in the other channels is necessary. *Figure* 15 depicts a three-way system with crossover frequencies of 400 and 3000 cps. The attenuators shown are capable of some 10 db of damping thereby providing a nearly constant impedance.

In the author's system the midrange speaker and the tweeter (direct radiators) are housed in small cabinets of which the rear sides are replaced by one-inch layers of plastic foam. This permits the loudspeaker to be oriented in such a way as to improve the diffusion of sound without causing trouble by backward radiation.

The enclosure's dimensions given apply to use of a 12 in. woofer. For larger sizes all dimensions can be increased in the same proportion as the cone diameter. The interior volume is not at all critical so that a somewhat smaller increase may be sufficient. ●
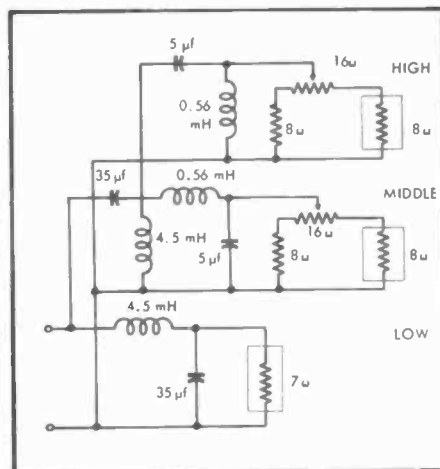

Fig. 14. Exploded view of aperiodic enclosure.

Fig. 15. Three-way crossover network.

The "Standard" Speaker System

C. G. McPROUD

The forerunner of most of the present-day back-loaded horns was first described in these pages eight years ago this month. Mentioned occasionally as our "Standard" speaker system, it has aroused so much interest among those who missed the first publication that it is here reprinted with only minor changes.

PAR FOR THE COURSE in hi-fi is the desire for change, be it amplifiers, tuners, phono equipment, or—more likely—speaker cabinets. A healthy condition, and one to which we subscribe heartily if we are sure that we are able to better ourselves or our music reproduction by so doing. In addition, when changes are made the equipment that is replaced is often passed on to some newcomer to hi-fi, who starts the cycle over again with his continued upward improvement. One of advantages of component hi-fi is that the equipment usually lasts for many years, but that if changes become necessary they may be made simply by replacing the individual section. However, the speaker cabinet described here has not been changed (except to add a super-tweeter which was not even on the market when the cabinet was first built) since 1948. This particular enclosure was described in these pages in January and February, 1949, and reprinted in the 1st Audio Anthology. However, so many inquiries have come in about what we consider the "standard of comparison" amongst loudspeakers that we are reprinting the original article almost exactly as it appeared eight years ago. At that time it was the first corner cabinet that employed both front and back waves from the low-frequency cone.

It will be seen that this enclosure provides for a television set with the picture tube located between the two speaker sections. The original design used a 12-in. tube, and since then the favored tube size has increased to 21 inches. The basic idea could well remain the same, even though the tube enclosure would have to be enlarged. On the other hand, it is thought probable that many would prefer to mount the high-frequency unit(s) in this same space and eliminate the superstructure. In any case, we again present this article—with slight modifications and editorial changes—for those of AUDIO's more recently acquired readers.

* * *

With television firmly established as a home entertainment medium, (in many homes) a *complete* installation must necessarily contain TV facilities, without sacrificing the superb quality desired for radio and phonograph reproduction. Therefore, solving two problems at once, the TV installation has been combined in a cabinet with the loudspeaker in a form which results in high-quality reproduction, a reasonable compactness, and a piece of furniture which is an eye-appealing addition to a modern living room.

Basic Design

It has been fairly well established that the most efficient location for a loudspeaker is in the corner of a room. The most familiar example of this arrangement is represented by the Klipschorn, which consists of a two-way speaker system with both high- and low-frequency units being horn loaded. The cabinet work for the Klipschorn is extremely complicated, and certainly not one which the amateur woodworker should attempt. Some constructors have mounted several medium-quality cone speakers on the two sides of an obtuse enclosure which was placed in corner and have used this arrangement with excellent results. The corner location is optimum from the standpoint of loading on the speaker, since the radiation is over only half the angle of that from a speaker mounted on a flat wall. With a number of ordinary cones, the result is a means for moving a rather large volume of air without the necessity of having a large cone excursion of a single unit. Thus, better low-frequency response is obtained with speakers which individually would not perform so satisfactorily.

The writer had long used a standard two-way speaker, with the woofer in an 8-cu. ft. bass-reflex box of conventional design, and while the reproduction quality has been considered excellent, the low-frequency output did not compare with that of a good theatre-system. Thinking from this point, the next step appeared to be in the direction of a corner speaker, yet utilizing the reflex action of a vented cabinet. Basically, therefore, the new design could still be described as a bass-reflex cabinet, but it occupies the corner of a room and is arranged so that the vents are loaded by a horn comprised of the walls and the sides of the cabinet enclosure. The plan view of the cabinet is shown in *Fig.* 1, with the vent openings A-A' along the sides. Thus the vents are loaded by the straight-sided "horn" between the wall and the cabinet.

Experience has shown that loading of the vents should be accompanied by a similar loading on the direct-radiating side of the low-frequency speaker, so the front of the cone is provided with

The two-way speaker system in a modern corner cabinet, with a 12-inch TV tube in the optimum location for the best illusion. Minor changes would permit the use of larger tubes, if desired, or its space could be occupied by the high-frequency speaker section, eliminating the superstructure.
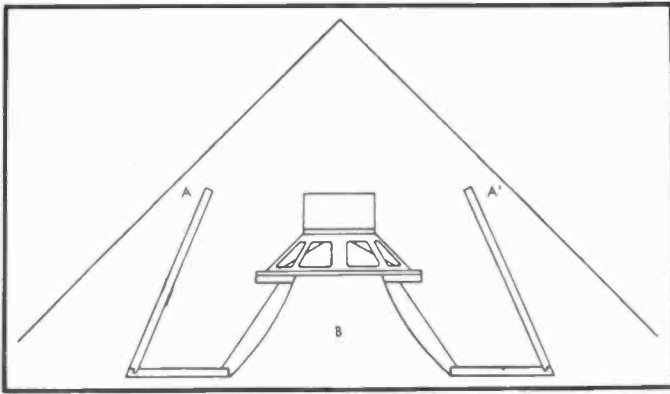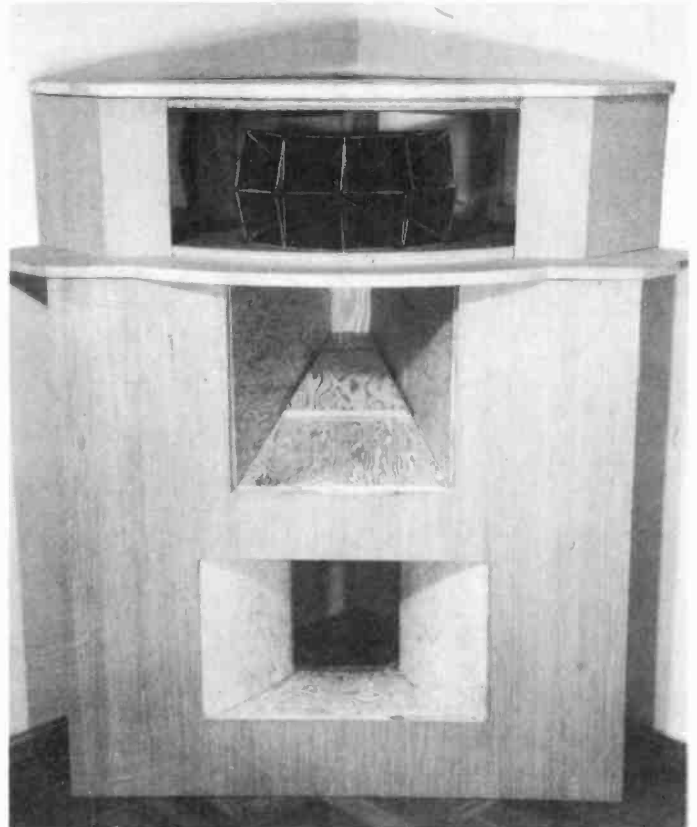
Fig. 1 (above). Plan view of the new design, showing vents at either side between the cabinet and the wall and the horn loading provided by the wall and the cabinet sides.

Fig. 2 (right). Front view of the cabinet during construction, showing the woofer "well" and the TV tube enclosure.

another horn section, B, thus equalizing front and back loading and increasing the radiation efficiency. A top for the cabinet provides an air seal by means of gaskets between it and the wall, and the floor provides the other side of the horns on the the vented ports. The entire cabinet is open to the back, and utilizes the room corner, although if desirable for use in other locations, a false corner could be constructed to provide the necessary back.

After determining the basic design, any necessary variations can be made to accommodate TV, as has been done in this case. The picture tube is simply enclosed in a wood housing, and doors in the cabinet front cover the screen when it is not being used. The superstructure, which can be seen in *Fig.* 2, houses the multicellular high-frequency horn and unit, and the super tweeter, and the space behind it is large enough to accommodate the TV receiver chassis. With such a construction, the picture tube is between the speaker sections, and the illusion of sound coming from the picture is considerably better than if the speaker were either above or below, or at the side of the screen.

Development of the practical aspects of the construction is controlled by the units selected for both high- and low-frequency speakers.

Good speaker performance depends on a number of factors. Among these is a high gap flux, which should be as great as possible. A high field strength makes for good damping as well as the maximum of efficiency. Another important factor is the relative weights of the cone itself and the voice-coil structure. It is considered good practice—for good low-frequency reproduction—to have these two weights as nearly equal as possible. It is also important to have as low a resonant frequency as can be obtained readily.

Construction Details

Getting down to a specific design, therefore, the cabinet takes the shape shown in *Fig.* 1, for a cross section at the plane of the low-frequency cone, and at (A) of *Fig.* 3 at the plane of the center of the TV picture tube. The top of the low-frequency cabinet has the outline shown in the solid line at (B), with the superstructure shown by the dotted lines. The top is 39 inches from the floor, and the corners of the top meet the side wall 36½ inches from the corner. Allowing for the volume of the speaker well and speaker end of the tube enclosure, the net volume of the cabinet is 8.5 cu. ft. This does not include the
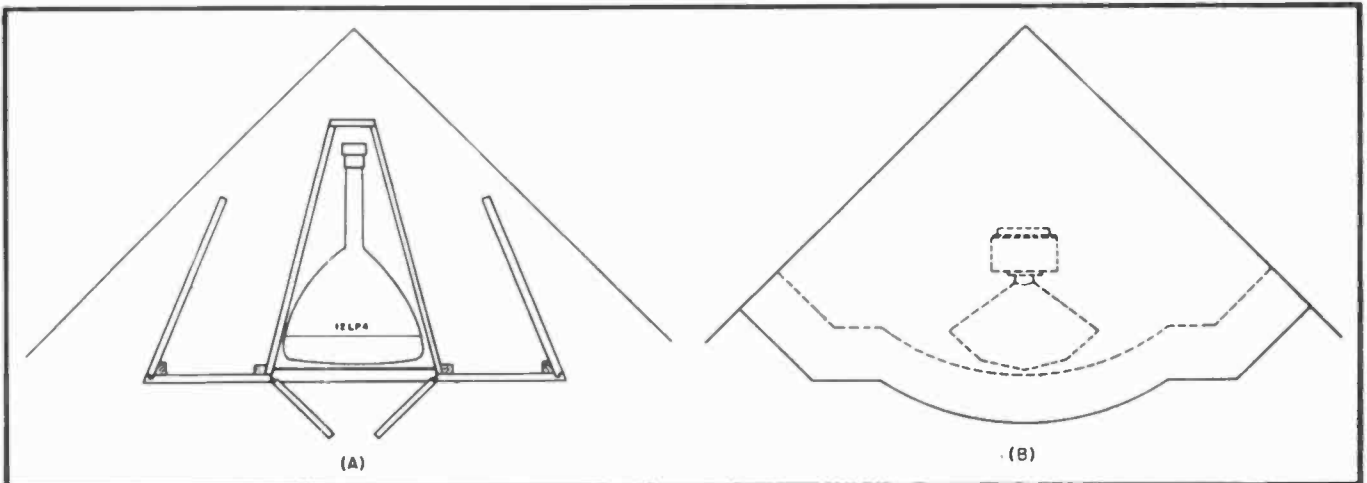

Fig. 3. (A) Cross section of the cabinet at the plane of the center of the picture tube to show location of tube enclosure. (B) Plan of the top (solid lines) and of the superstructure (dotted lines).
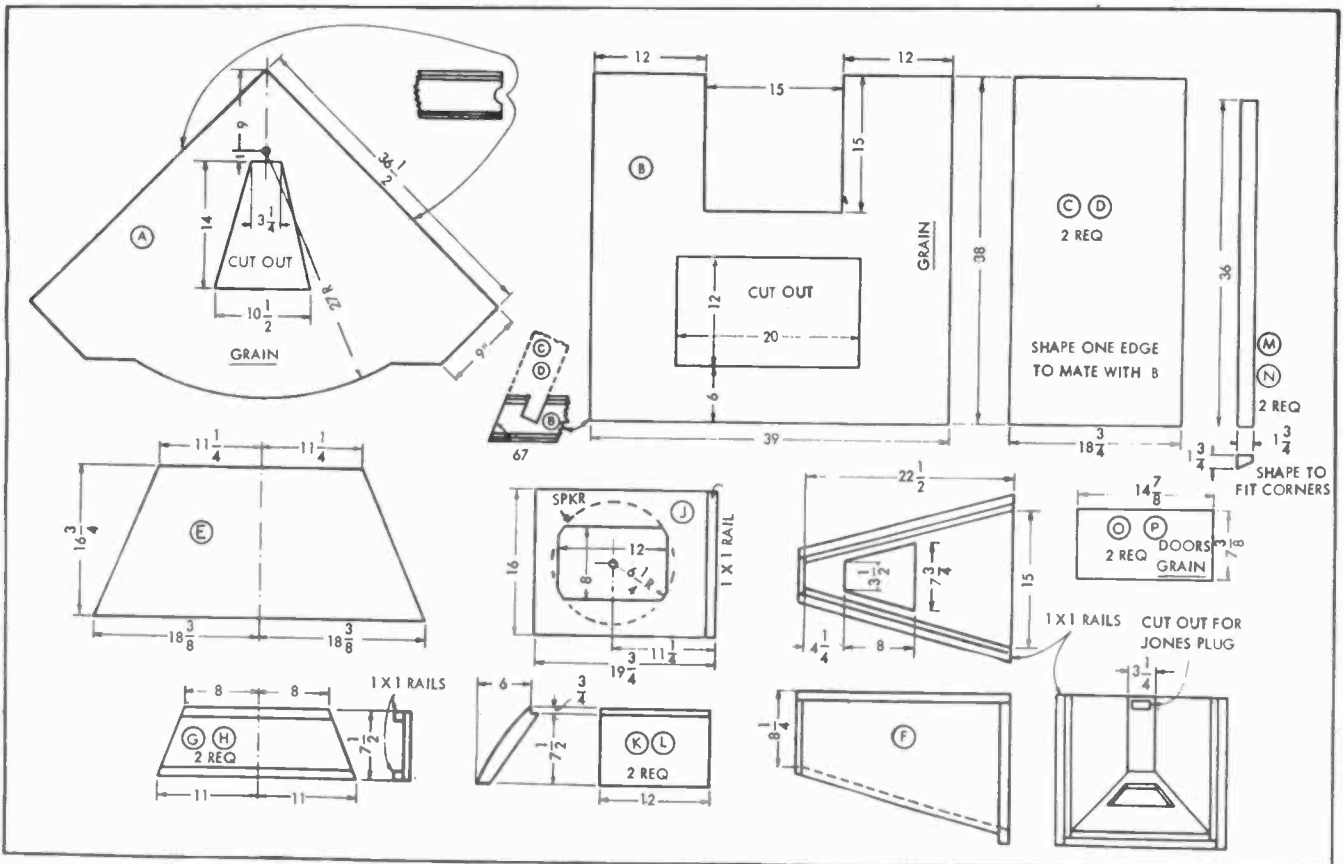
Fig. 4. Details of the pieces which comprise the lower cabinet, speaker well, and tube enclosure. Latter is for 12-in. tube, now almost obsolete.

vent horns, which are usually added into the volume when vent pipes are used on the reflex ports.

Figure 4 shows the major parts used in the assembly of the low-frequency portion of the speaker, together with the housing for the picture tube. This housing is scaled for a 12LP4, which is obsolete by now. If a larger tube is to be used, the housing should be arranged to be large enough. However, it is felt that the majority will not combine TV with the speaker anyhow, and will simply omit the tube housing or else place the high-frequency horn at this point, with a suitable housing around it. *Figure* 5 shows one model with the high-frequency horn in a space at the top.

If the TV tube is to be located in the speaker cabinet, the method will be found applicable, even though some dimensions must be changed. It will be noted that there is a hand hole in the bottom of the tube enclosure, with a removable cover which serves two purposes: It mounts the deflection yoke, and thus permits adjustment of the TV receiver with the tube removed from the cabinet; and it also permits anchoring the cabinet to the corner of the room, if desired, by means of a pair of chains and two turnbuckles. The cabinet is placed close to the corner, and with the turnbuckles open to their maximum, chains are looped over a hook mounted on the floor right in the corner. Then the turnbuckles are tightened

up, thus locking the cabinet into the corner with the top tight against the wall. The quarter-inch semicircular groove along the back edges of the top provide space for a gasket to make an airtight seal. When the hand-hole cover is replaced, the structure is airtight except for the vents.

The wood selected for the top and front of the cabinet should be a suitable match (or contrast) for the furniture used in the room where the speaker is located. For solid construction, ¾ in. material is recommended, with veneered hardwood being used for the top (A) and the front (B). The doors (O) and (P) should be veneered on both sides. The bottom (E), sides or wings (C) and (D), speaker baffle (J), and the tube enclosure (F) can be of less expensive fir plywood, also ¾ in. thick. The tube enclosure is a part of the acoustic chamber, which accounts for its seemingly over-solid construction.

The sides, (K) and (L), of the speaker well are shaped from two-inch white pine, and should be fitted to the opening in the panel. The speaker baffle is drilled with eight holes, and T-nuts for mounting the speaker are installed on the front before the "horn" is assembled. In addition to the pieces shown, a number of ¾ × ¾ strips will be needed for corner reinforcement. Parts (M) and (N) are for the acute corners at the front of the cabinet.

The details of the superstructure will

be described in the following pages, and the parts are not shown in *Fig.* 4. However, it might be well to plan on another veneered piece nearly as large as the top (A), since the grain should run parallel with the front of the cabinet. The two tops will cut readily from one panel of hardwood veneer.

Since this speaker is supposedly "functional," no attempt is made to disguise its appearance. The front of the low-frequency cone is visible in the speaker well, or horn, being protected by a screen of expanded metal. In the writer's cabinet the inside of this horn is finished in dark blue lacquer, as are the sides of the cabinet and the edges of the two tops. (Thin ribbons of wood are now available for covering the edges of plywood or veneered paneling. These are already glued and may be applied quite readily to the edges prior to finishing, thus giving a uniform wood surface throughout.) The front and the top, together with the superstructure, are bleached oak, as is the tube mask. Lacquer covers the jointing of the speaker well to the panel, as well as the non-veneered edge of the top. If a uniform hardwood appearance is desired, the edges should be veneered—a job best done by the cabinet maker who cuts the pieces out. One caution is necessary—make sure that the top will fit the corner tightly. *Not all rooms have 90-deg. corners.*
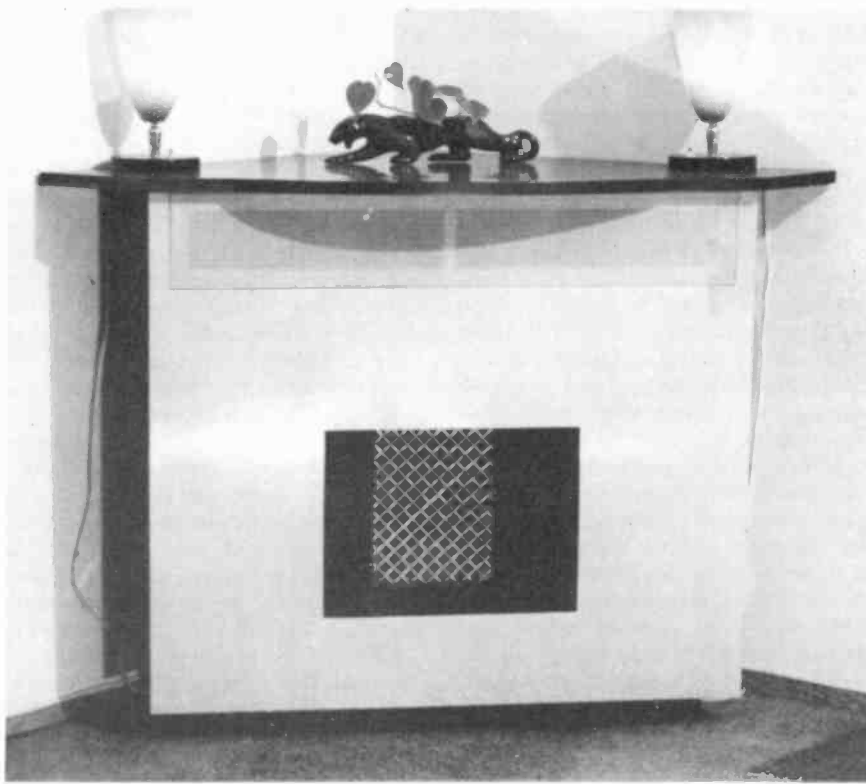
Fig. 5. Modified version built by Harrison Associates, New York. High-frequency units are behind grille at top of cabinet.

Assembly

Once all the pieces have been cut out, the next step is that of assembly. Since some of the operations appear to be tricky, it is well to follow a certain procedure to avoid having to put the last few screws in with an offset screwdriver. The first step is to assemble the speaker well, which approximates a short exponential horn. Parts (G) and (H) are mounted on part (J), using the shaped sides as spacers. Remember to put the T-nuts in place on the baffle before attaching the other parts. All joints should be glued, preferably with casein glue, and secured with 1¼—12 flathead wood screws, countersunk. This assembly should then be attached to the front panel, also with glue and wood screws. The shaped sides, (K) and (L), are then fitted into place, also with glue and screws. Every joint in the cabinet is made with both wood screws and glue except that between the top and lower section. The top is attached only with screws, so it may be removed to enable the cabinet to pass through a 30-in. door.

After the speaker well is completed, the bottom is attached to the front, using a ¾-in. strip at the joint. The front extends clear to the floor, to eliminate the extra construction necessary for a recessed base. Actually, however, after living with the unit for over four years without a recessed platform mounting, the person in charge of furniture in our house prevailed upon us to raise the cabinet by using three pieces of 2 × 2

on the bottom, set back 2 in. from all sides, and with the front corners mitered. The bottom is thus inset, since the sides also extend to the floor. After the bottom is attached to the front, it is also secured to the speaker baffle.

Next the corner braces are attached to the front, and the strips along the lower edges of the sides are screwed in place, ¾-in. up from the edge. The sides are then fitted into the groove in the front panel, and all joints screwed together. The tube enclosure is next mounted to the front, and supported at the back with a cross brace. The entire structure should now resemble that shown in *Fig.* 6, which also shows the ½-in. square furring strips for the sound-deadening lining.

At this point, the doors should be fitted, using ⅜-in. Soss invisible hinges which are mortised into the front and the doors. These are the least obtrusive of any hinges available, and while they are a little difficult to mount, the final appearance warrants the extra effort.

Electrical Connections

Some provision must be made to introduce the signal and an a.c. line to the unit, since it will not be readily accessible once the cabinet is mounted in place. The power circuit is necessary for the TV chassis, as well as for a possible outlet for a lamp or clock as an ornament on top of the speaker. Since the speaker is designed to work from a radio-phono system housed elsewhere, the speaker signal must also be fed in. This is done at a small panel located just inside the lower right corner of the cabinet. One three-way male receptacle is used for speech, and a two-way male twistlock receptacle is used for the a.c. line. The speech circuit goes to
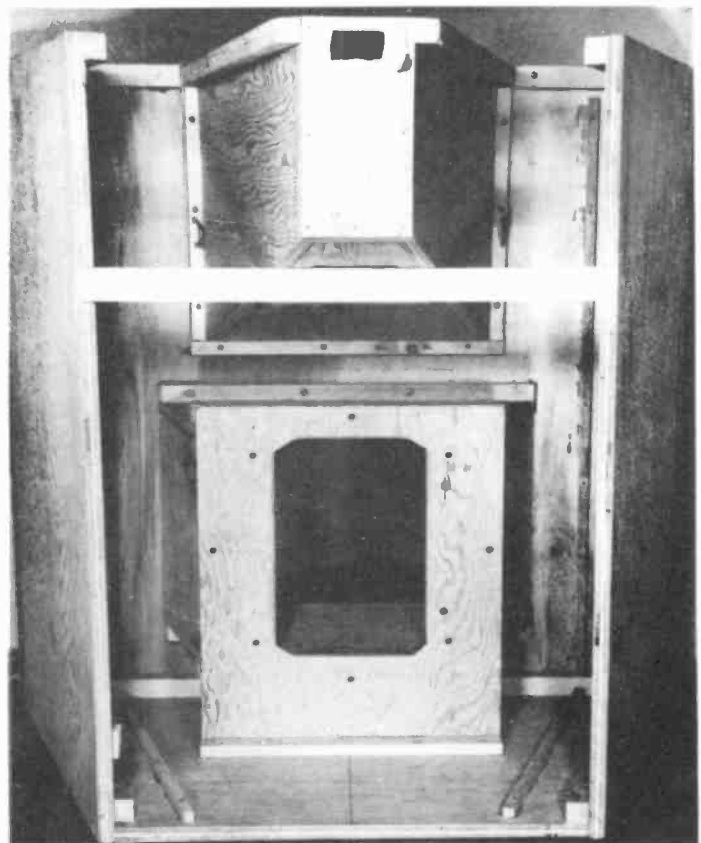

Fig. 6. View of the rear of the lower cabinet showing method of assembling the various sections.

a switch which selects radio-phono in one position, TV in another, and turns the speaker off in the third—both inputs being properly terminated when not in use. The switch is mounted on the right wing just under the top, and the high-frequency horn attenuator along side of it. The output of the switch then goes to the dividing network, mounted on top of the speaker well, and thence to the two speakers units. Access to the high-frequency unit is had through an 8-terminal Jones receptacle, which also receives the speaker output from the TV receiver and carries the a.c. line up to the superstructure. This receptacle is mounted at the back of the tube enclosure, and permits removal of the top without unsoldering any connections. The electrical circuits are shown in *Fig.* 7 for a crossover frequency of 900 cps.

Preliminary Finishing

After the lower section is completely assembled, it should receive its first finishing operation. To protect the surface of the wood, the interior and the bottom should be given a primer coat, depending on the finishing method selected. All cracks in the exterior should be filled with plastic wood, and the rear corners of the speaker well should be rounded out with fillets of the same material. After thorough sanding, the sides and the speaker well should receive a coat of Firzite, which is an excellent filler for plywood, and the front and top surfaces should be filled or primed, as desired. Final finishing of the hardwood exterior

should wait until the superstructure is completed in order that the two sections match as well as possible. Since most of the work on the lower section is now complete, the padding may be tacked in, using large-headed roofing nails to prevent tearing out. Ordinary rug padding, such as Ozite, appears to be satisfactory for this purpose, although rock wool or Fibreglas is recommended by some constructors. The possibility of the fine glass shredding around a speaker cone argues against the use of either of the latter insulating materials, and the Ozite appears to do a satisfactory job of deadening without this risk. It is desirable to use two thicknesses over the larger areas, though the furring strips provide a good absorptive covering since there is an air space behind the padding. A still better padding has been introduced since the original cabinet was built. This consists of a 3/8-in. layer of felt applied with linoleum paste over the entire inside surface of the enclosure, making sure that the adhesion is complete throughout. This type of felt may be obtained from Ingalls Electronics Co., 30 W. Putnam Ave., Greenwich, Conn.

The Superstructure

All of the cabinet above the top lettered (A) in *Fig.* 4 is called the superstructure. It is primarily an ornament, since it serves no function except to enclose the high-frequency unit and horn and the TV chassis. This section is permanently attached to the top, and when the speaker is to be moved, the

top and superstructure are detached from the lower cabinet, since the complete assembly will not go through a standard door. The parts for the superstructure are shown in *Fig.* 8.

The actual measurements will depend upon the high-frequency horn and unit selected, as well as the TV chassis, and their placement in the superstructure. The cabinet was originally designed to house a Tech-Master 630-type kit, available at a considerable saving over the completed model, and still a superb TV set. Since some modifications had to be made, it seemed desirable to start from the blank chassis and build the entire receiver—the true experimenter's viewpoint. As work progressed, it was learned that while the 630 chassis would fit in the space, it was doubtful if the writer could lift it up and put it there without the aid of an overhead crane. Therefore the TV chassis was cut apart —with the r.f., i.f., video and audio sections and the power supply on one part and the deflection circuits on the other. The two parts are interconnected by a 13-wire cable, and two separate shielded pairs—one for sync and one from the AGC winding on the width-control coil. Feed to the picture-tube socket comes from the r.f.-etc. chassis, and that to the yoke and focus coil comes from the deflection chassis. One other refinement is necessary—a very heavy cable must be used to connect the two chassis grounds together. This was not done for a long time after the original construction, but a slight "S" curve in all vertical lines was finally traced to an a.c. drop in the ground lead of the connecting cable because this lead also carries the relatively high filament current. A heavy ground strap cured the trouble.

Since it was not considered desirable to have screws showing on the top, the fastening consists of two 5/16-in. rods, threaded on both ends, which extend from the center deck and engage two T-nuts which are set into the top panel and fastened with flat-head wood screws. Wing nuts are threaded onto the lower ends of the rods and peened on so as to serve as handles, as shown in *Fig.* 9. The cotter keys keep the rods from dropping out of the hole in the upper framework, so there is no difficulty in engaging the T-nuts when the top is put in place.

The triangular cut-out is fitted with a perforated metal grille to serve as ventilation for the TV receiver, which draws some 300 watts, and consequently needs free circulation of air. The two side panels are assembled of 3/8-in. veneer, of the same wood as the tops and the front; they are simply screwed to the grille support from inside. The TV controls are brought through these panels—the channel switch, fine tuning, picture, focus, brightness, and sound-volume controls at the right, and the
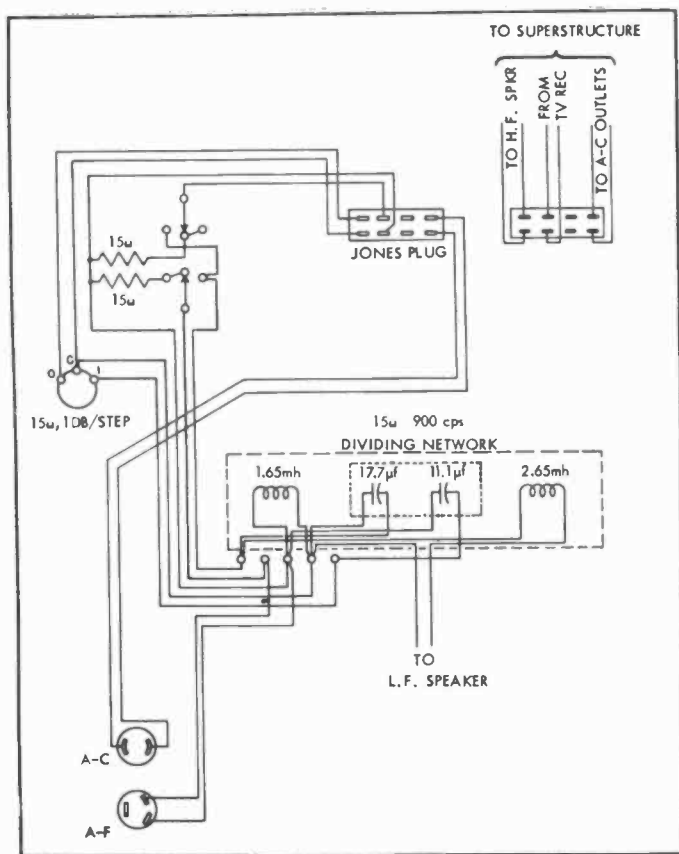

Fig. 7. Wiring diagram of the lower cabinet. Plug and cable at upper right are a pigtail on the superstructure.
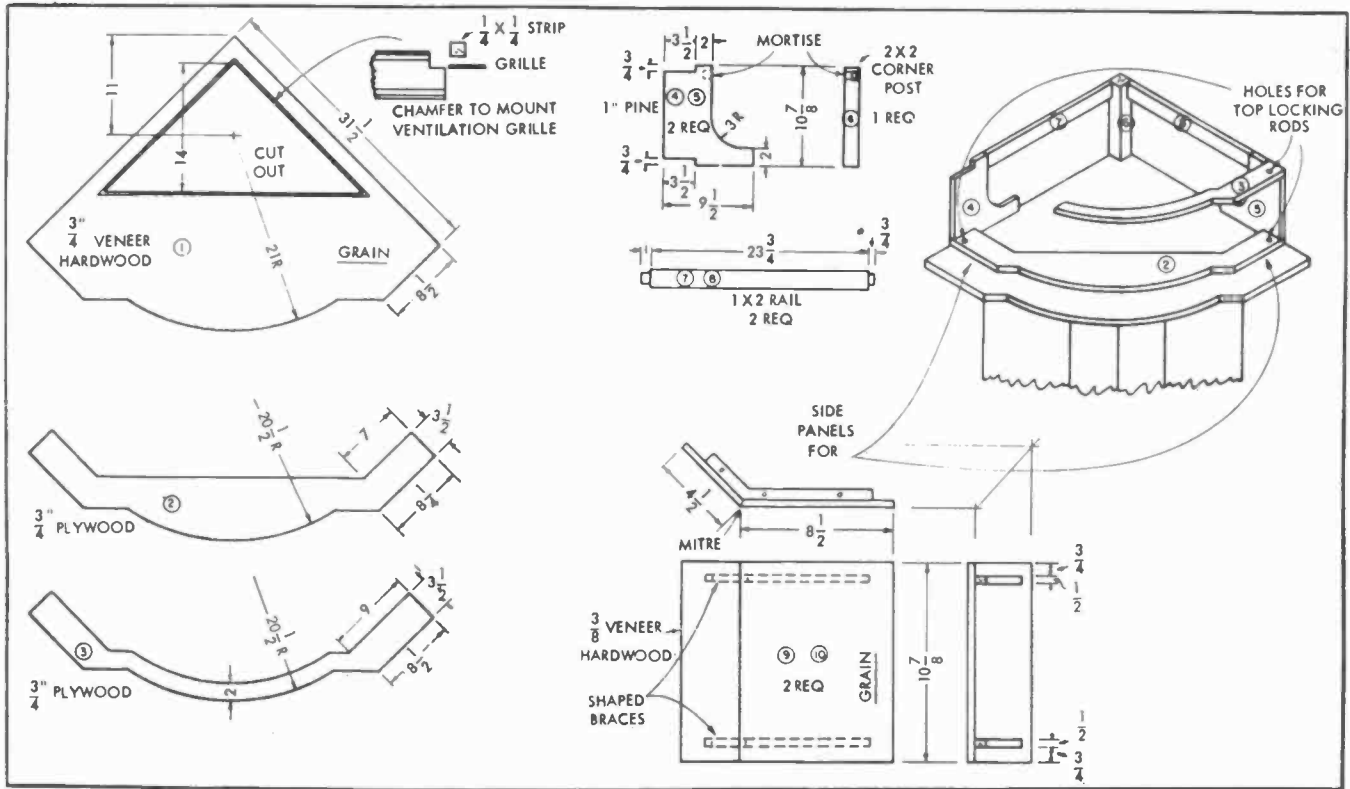
Fig. 8. Details of the various parts comprising the superstructure. Triangular cutout in part (1) is for ventilating grille.

hold controls, which only vary d.c. voltages, are extended from the chassis with long leads and with the shafts extending down through the bottom panel where they are accessible in the left port-horn opening.

It is desirable to include an a.c. outlet on the top so as to accommodate a lamp or clock. It is equipped with a length of wire to plug into a dual outlet on the TV tube cover, which also mounts the high-frequency speaker unit, and a male plug which receives the output of the TV receiver. All connections to the top section are carried through an 8-terminal Jones plug and receptacle, the latter being mounted on the back of the tube housing.

The screen in front of the high-frequency horn consists of another piece of perforated metal, covered with a piece of plastic grille cloth. This effectively hides the high-frequency horn, although both the perforated metal and the horn should be painted a light color so as to avoid the appearance of a dark mass behind the cloth. Although it is claimed that the speaker is functional and not too much effort is expanded to disguise its appearance, the grille in front of the high-frequency horn was added as a concession to appearance. However, if the listener is able to see two separate speakers of a multispeaker system, he is almost certain to feel that the sound is coming from two separate places. If the speakers are covered—even as little as in this cabinet—the two sources blend together perfectly.

H-F Speaker Mounting

It is necessary to provide access to the tube compartment, so the high-frequency horn and unit are quickly demountable. Jumbo banana plugs were mounted on the front corners of the horn, and on the mounting block for the unit, and jacks were set into the framework for the front pair, and into the tube cover for the rear ones. The electrical connections for the high-frequency unit are carried on the latter two, and the entire h.f. speaker may be

Fig. 9. Method of fastening the top to the superstructure without exposed screws to mar the appearance.

lifted bodily from the jacks when necessary, without the need for watching phasing. The a.c. outlets and a receptacle for the TV output are on the same tube cover, and all connections are carried through a short jumper cable to the male Jones plug.

Needless to say, the high-frequency speaker must be phased correctly before its final position is determined. This is best done by feeding a tone at crossover frequency to the speaker and reversing the high-frequency leads if necessary to obtain the greatest output from the entire system, preferably measuring the output by a microphone and another amplifier with a volume indicator at its output. Then the high-frequency speaker is moved backward and forward until the maximum output is obtained. If the additional equipment is not available, put a tone at the crossover frequency on the system, and listen to first one speaker and then the other, moving the ear up and down in a plane parallel with and about 18 inches from the front of the cabinet. If the speakers are correctly phased, there will be a continuous tone heard from one speaker to the other. If not, there will be a null somewhere between them. At the crossover frequency (900 cps for the constants shown in *Fig.* 7) it should not be possible to detect any difference between the two speakers as the head is moved up and down. Try moving the high-frequency speaker backward and forward until there is no difference between the two sound outputs, then check by
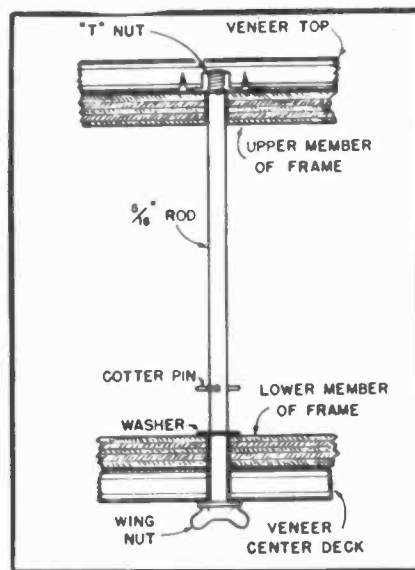
listening to male voices on the complete system. Optimum results can be obtained without instruments if necessary, and the lack of them need not deter anyone from assembling a satisfactory system. It is suggested, however, that after deciding upon the correct position and phasing for the high-frequency speaker, this position be marked carefully. Then, live with the system for a week or so before making the mounting permanent. Try shifting the unit back and forth while listening to a familiar record or program. Make sure that the optimum position is determined before the job is considered complete.

TV Receiver Placement

The entire superstructure was planned to house the Tech-Master kit receiver, since this model has an excellent reputation for performance. Other models can undoubtedly be fitted into the space if desired, and the particular method of placement is left to the ingenuity of the individual constructor.

Connections to the picture-tube socket should be made through a 5-prong socket and plug, and those for the focus and deflection coils should be made with an octal socket and plug. It is necessary to carry a ground connection to the brackets which mount the two coils so as to have a ground for the outside coating of the tube, as well as for protection during adjustments. For the high-voltage connection, it is suggested that a banana jack be installed on the Bakelite insulating strip where the lead normally leaves the high-voltage compartment. This permits the chassis to be removed for servicing without unsoldering any connections.

The focus coil and the deflection yoke are mounted on the hand-hole cover in the tube compartment, using the original bracket modified to mount on the sloping surface. The tube is centered in the compartment, with padding all around for protection. The leads are brought up through notches along the side of the tube compartment cover. A plastic mask is mounted on a $\frac{3}{8}$-in. oak veneer front panel, providing both protection for the face of the tube and masking for the picture tube.

Woodwork Finishing

In the cabinet shown, the front and tops are oak veneered, treated in the blond "rift" finish. This is a simple operation, consisting solely of painting the well-sanded oak surface with one coat of Firzite, allowing it to dry for about five minutes, and then wiping it off. The white remains in the grain of the wood and gives it an attractive finish. After the paint dries for about 24 hours it may be waxed, or else several coats of shellac may be used, rubbing each down with steel wool. Since there are so many methods of finishing wood, this part of the work may well be left to the discretion of the builder.

The speaker well, the sides, and the edges of the tops are lacquered with two coats of ensign blue Larcoloid, which gives a glossy finish from a brush coat.

This particular treatment may not appeal to everyone, but it is modern and attractive, and requires a minimum of work. Since the entire design is essentially modern, it is felt that the surface treatment should also be modern.

After the cabinet is completed, it can be permanently installed as previously outlined. Two rings are mounted at the sides of the tube compartment. To these are attached 8-in. lengths of chain, which are connected in turn to 4-in. turnbuckles. Another length of chain is attached to the other end of each turnbuckle, and looped as closely as possible over a large screw hook inserted in the floor right in the corner of the room. The turnbuckles are then tightened up, sealing the top against the wall. Originally it was planned to use a steel cable, but the difficulty of attaching the hooks and turnbuckles to the cable ruled it out after a few trials. The chain is much simpler. It may be desirable to deaden the chain with a cloth sleeve over it, or by lacing venetian-blind cord through the links. The gasket used for the top is a 5/16-in. braided clothesline, tacked on the ends and glued into the groove for its entire length. The sides of the cabinet are spaced from the wall by ordinary door stops—adjusted to the required 1½ in. by selecting the point at which they are attached to the cabinet. Actually, there does not seem to be a noticeable difference in performance if the cabinet is not perfectly airtight to the corner.

Performance

Subjectively, this speaker system seems to be "the answer," in the opinion of the writer and of many others who have heard it. It gives the feeling of a wide source of sound, as would be expected since the separation between low- and high-frequency speakers is approximately 33 inches from center to center, and the very low frequencies come from the side vents. The over-all width of the speaker from wall to wall on the plane of the front is 65 inches. This wide-source effect is pleasant in the extreme and until it is experienced, the listener may doubt its advantages. A similar effect may be obtained for a trial by connecting two or three speakers to the output of an amplifier, and placed about this distance apart.

The efficiency of the low-frequency speaker is evaluated by comparing the setting of the high-frequency attenuator with that used when the identical components were assembled in a 7½ cu. ft. bass reflex cabinet of conventional design, as they were prior to building this unit. With the old cabinet, balance was obtained with 6 db in the h.f. attenuator, while only 4 db is required with the

corner speaker. This indicates that the low-frequency speaker is 2 db more efficient in this cabinet than it was with the bass-reflex box.

Performance of this system compares favorably with medium-size theatre systems, and it has "presence"—the intangible characteristic of realism which gives the desired feeling that the performer is actually in the room.
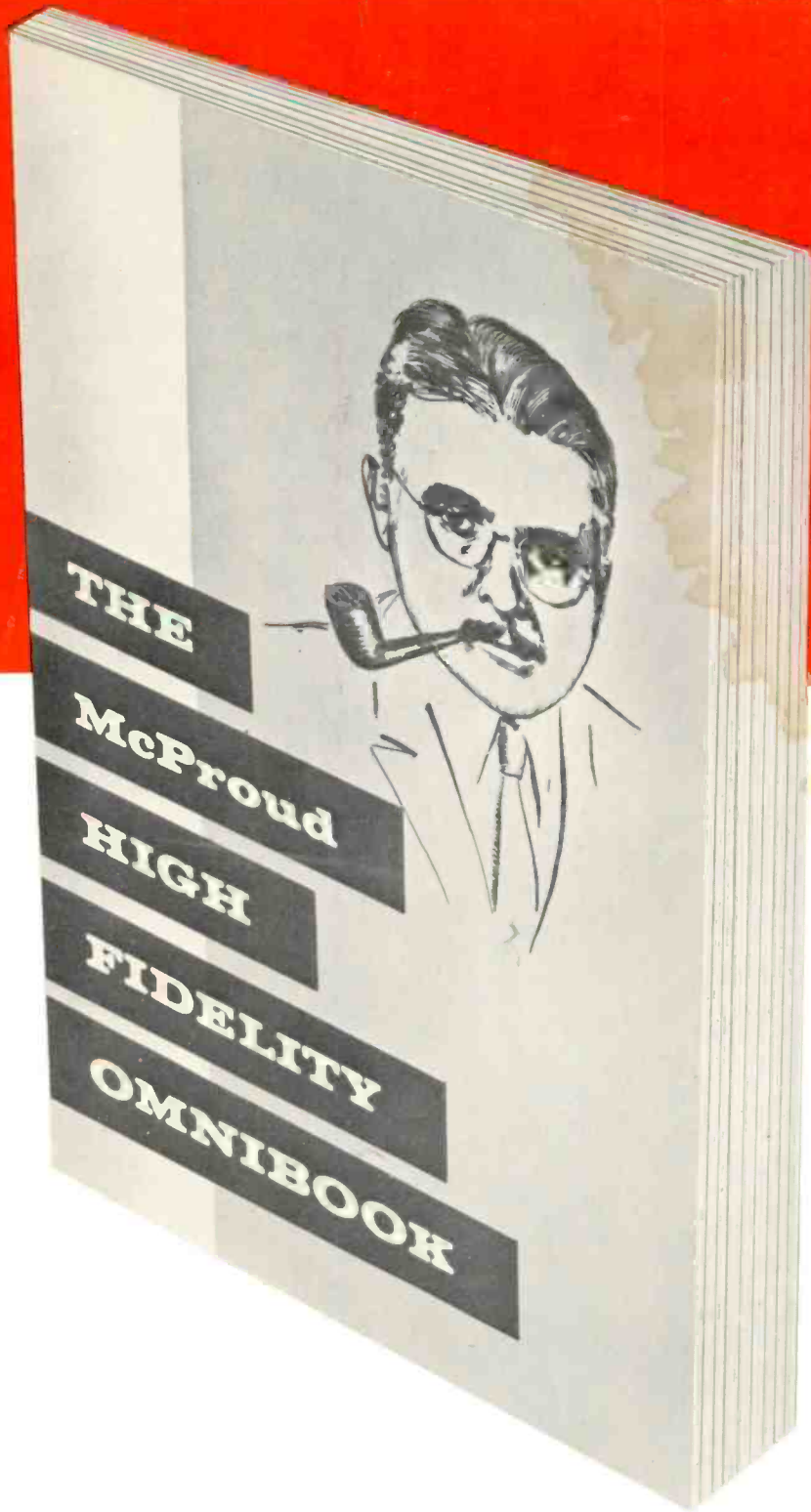
Late Modification

Since this enclosure was first built in 1948, no changes have been made in the original design except for the addition of a "super-tweeter" high-frequency unit and its associated dividing network and level control. The speaker itself is mounted on the left side and the level control on the right, being located slightly above the center of the narrow panel of parts (9) and (10) shown in *Fig.* 8. The 4000-cps dividing network is enclosed in a small metal case and is mounted on the back of the tube-cover plate, which covers the cutout in part (A) of *Fig.* 4. This location is convenient, but would depend somewhat on the size and type of midrange unit employed.

Altogether, the design is sufficiently flexible to permit the use of cone tweeters, horns of various sizes, or any other high-quality components which may be selected. For example, the model shown in *Fig.* 5 employs a 1 × 4 multicellular horn and a super-tweeter, both being located behind the grille shown at the top of the front panel. One modification of this type of cabinet consists in placing a 90-deg. "V" behind the woofer with the two panels extending from the bottom to the top and placed as close to the speaker unit as possible. This is quite effective if the front horn is eliminated, with the woofer being mounted directly on the front panel.

We have not seen fit to make any changes from the original (except for the super-tweeter) and to date we do not believe that we have heard any other complete speaker system designed for the home that excels this one. Those that have come close have almost always been of the same general type of design, but this may be attributed to a certain degree of coloration to which we are accustomed and which we now think of as being "the way it should sound." However, most others who have heard this particular system have gone away with the feeling that this one was outstandingly good—in fact, one listener was described as "having stars in his eyes."

The system has stood the test of time —eight years, so far—and we have no hesitancy in recommending its construction to those interested in making their own. We believe that the results will more than justify the time and money expended. ●

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