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Professional Group on Audio

A Group of Members of the I. R. E. devoted to the Advancement of Audio Technology

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TABLE OF CONTENTS

The Institute of Radio Engineers

IRE PROFESSIONAL GROUP ON AUDIO

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A NOTE ON NOISE IN AUDIO AMPLIFIERS*

H. J. Woll and F. L. Putzrath
RCA Victor Division
Camden, New Jersey

The limitations of background noise present a universal problem to designers of high-gain amplifiers. This technical editorial was invited for the purpose of summarizing these limitations for readers of TRANSACTIONS of the IRE-PGA. It happens that the conclusions are not in complete accord with another paper in this same issue.** The papers are being published simultaneously so that readers may have the benefit of both points of view.

— Editorial Committee

There are a number of noise sources in audio amplifiers such as:

a. Thermal noise in the input coupling circuit
b. Shot noise
c. Partition noise in pentodes
d. Flicker noise
e. Ballistics and microphonics
f. Pops
g. Hum

Hum will not be considered here although it is a serious problem and much can be said concerning its elimination. Ballistics, microphonics, and pops are matters of tube design and selection rather than circuit design and will not be considered either.

This paper will be primarily concerned with flicker, shot, and partition noise and the effects of the input coupling network. An attempt will be made to outline the requirements of the input circuit and to discuss the magnitude of second stage noise in conventional configurations.

Noise is generated in the resistive component of any impedance by the thermal agitation of the electrons. The magnitude of the thermal noise voltage of a resistance in temperature equilibrium is:

$$e = \sqrt{4kTBR} \text{ volts rms}$$

where

k = Boltzmann's constant
T = temperature in degrees Kelvin
R = resistance in ohms
B = bandwidth in cycles per second

A certain signal-to-noise ratio is available from the source and is determined by the signal strength and the magnitude of thermal noise. Using an amplifier, it can be approached but never exceeded. Noise figure is a convenient way of expressing the amount by which an amplifier deteriorates the signal-to-noise ratio available from the source. Noise figure may be defined as the available signal-to-noise ratio at the source divided by the available signal-to-noise ratio at the amplifier output. An equivalent definition is that noise figure is the ratio of the total noise power at the output of the amplifier to that component of the noise output power which is due to thermal noise in the source impedance. Noise figure is costumarily expressed in db.

Shot noise is generated because the plate current of a tube is not continuous, but consists of discrete charges which are numerous enough to very closely approximate a continuous current. Partition noise in pentodes is similar to shot noise but is caused by the division of current between the plate and the screen. Flicker noise is caused by local fluctuations of emissivity of the cathode. These sources of noise may be lumped and their effect duplicated by an equivalent voltage generator. This generator is commonly represented by a resistor, R_{eq}, in series with the grid of the tube and is chosen to be of such a value that the thermal noise voltage generated by it is equal to the sum of the shot, partition, and flicker noise voltages referred to the grid circuit. Thus:

$$R_{eq} = R_{shot} + R_{part.} + R_{flicker}$$

Shot noise and partition noise are independent of frequency and their equivalent noise resistances are generally considered to be:

$$R_{shot} = \frac{2.5}{g_m}$$

$$R_{part.} = \frac{20\, I_{screen}}{g_m\, I_{cathode}}$$

In the audio frequency band, flicker noise is much greater than either of the above two noise sources. It is a function of frequency and thus its equivalent noise resistance, $R_{flicker}$, is also. For any particular frequency characteristic an integrated $R_{flicker}$ can be found that is a constant.

The noise spectrum of a typical low noise triode with a coated cathode is shown in Fig. 1. In this tube, R_{shot} is 1500 ohms and $R_{flicker}$ integrated over a 12 kc flat bandwidth is 6000 ohms.

Consider the problem of determining the noise figure of an audio amplifier stage. Identical grounded-grid and grounded-cathode triodes with the same source re-

*Manuscript received January 29, 1954.
**R. Lee Price, "The Cascode as a low noise audio amplifier."

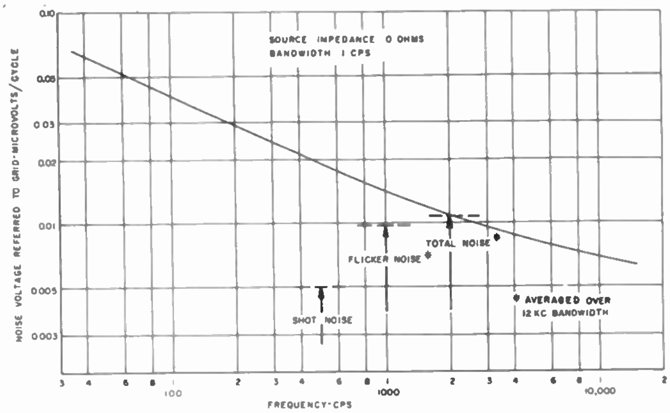

Fig. 1 – Triode tube noise vs. frequency.

sistance, R, are shown in Figs. 2 and 3. The tube noise is referred to the grid circuit and is represented by R_{eq}, which is a fictitious generator of voltage $e = \sqrt{4kTBR_{eq}}$. The tubes are then considered to be ideal amplifying devices. Since noise figure is:

$$F = \frac{\text{total noise power at the output}}{\text{that component of output noise due to thermal noise in the source}}$$

$$= \frac{4kTBR_s + 4kTBR_{eq}}{4kTBR_s} = 1 + \frac{R_{eq}}{R_s}$$

The interesting fact is that the noise figures of the grounded-cathode and grounded-grid stages are identical. Thus, although the input resistance of a grounded-grid stage might be under 500 ohms, a high source resistance

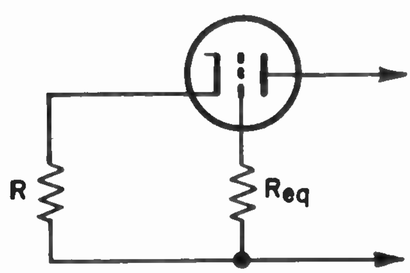

Fig. 2 – Grounded-grid equivalent circuit.

of perhaps 30,000 ohms is required to obtain a good noise figure just as in the case of the grounded-cathode connection.

It is to be noted that the above expressions represent *first* stage noise figures. The noise figure of an amplifier represents the deterioration of signal to noise ratio by *all* the stages in the amplifier. If the first stage gain is high, the succeeding stages do not contribute appreciable noise and the noise figure of the amplifier is about the same as that of the first stage. This is generally the case in audio amplifiers.

On the other hand, an amplifier with a grounded-grid input tube must be operated from a low source resistance to obtain appreciable first stage gain. Hence this amplifier is either operated from a high resistance source and suffers from noise contributed by the second stage, or is operated from a low resistance source and has a poor first stage noise figure, or some combination of the two.

Excepting ballistics, microphonics, hum, and pops, the following general conclusions can be drawn:

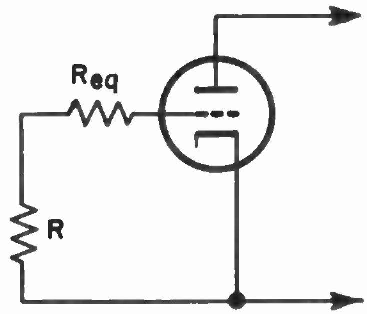

Fig. 3 – Grounded-cathode equivalent circuit.

Noise figure improves as the source impedance increases and zero db. noise figure can be approached in practice. (A low-noise triode with 22,000 ohm source resistance has a 1.0 db noise figure. See Fig. 4). As a result, efficient input transformers greatly improve the noise figure of a system with a low impedance source such as magnetic pickups and microphones.

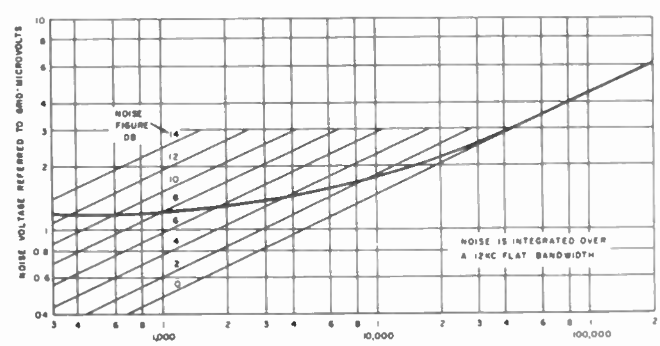

Fig. 4 – Noise vs. source resistance.

If the source impedance is low and bandwidth requirements or other conditions permit, the noise figure can be improved by paralleling input tubes increasing the g_m and thereby lowering R_{eq}.

It is to be noted that noise figure, *per se*, is meaningless. To express the performance of an amplifier, one must specify the noise figure with a given source resistance.

The noise figure of a stage is independent of the configuration, i.e., whether the stage is grounded-cathode or grounded-grid. Generally the gain of the first stage is high enough so that second and later stage noise make only a small contribution to the total.

Thus it can be generalized that cascaded grounded cathode stages at audio frequencies will give as good or better noise performance than other possible circuit configurations. In addition this configuration is most advantageous from a practical point of view – such as heater B+ supplies.

APPENDIX

A typical amplifier employing two cascaded grounded cathode stages is shown in Fig. 5. R_s and L_s are respectively the series resistance and inductance of

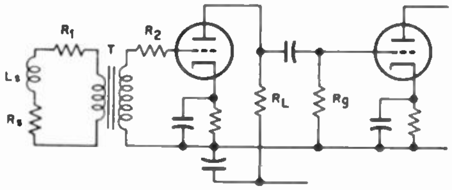

Fig. 5 – Typical two-stage amplifier.

the source. T is an input transformer with a turns ratio of n, a primary resistance of R_1, and a secondary resistance of R_2.

The equivalent circuit is shown in Fig. 6 where

$$L = n^2 L_s$$

$$R = n^2 R_s$$

$$R_t = n^2 R_1 + R_2$$

$$R_i = \frac{R_g R_L}{R_g + R_L}$$

The equivalent noise resistors of the two stages are represented by R_{eq^1} and R_{eq^2} respectively.

Examination of the noise figure of this amplifier will be made at point A. The noise figure at the input of the first ideal tube is:

$$F_1 = \frac{4kTB(R + R_t + R_{eq1})}{4KTBR} = 1 + \frac{R_t}{R} + \frac{R_{eq1}}{R}$$

The noise voltage due to the interstage coupling circuitry and the second tube is:

$$E_i = \sqrt{4kTB}\,\sqrt{\left(\sqrt{R_i}\,\frac{R_p}{R_p + R_i}\right)^2 + \left(\sqrt{R_{eq^2}}\right)^2}$$

$$= \sqrt{4kTB}\,\sqrt{\frac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq^2}}$$

where R_p is the plate resistance of the first tube. Dividing the above expression by the first stage voltage gain, G, this noise voltage is referred to the amplifier input so that the ratio of the ideal to actual noise powers due to the source and due to the interstage circuitry is:

$$F_2 = \frac{\dfrac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq^2}}{G^2 R_s}$$

The over-all noise figure of the two stages up to point A is then:

$$F = F_1 + F_2 = 1 + \frac{R_t}{R} + \frac{R_{eq^1}}{R} + \frac{1}{G^2}\,\frac{\dfrac{R_i}{\left(1 + \frac{R_i}{R_p}\right)^2} + R_{eq^2}}{R}$$

The above formula might be construed to mean that it would be desirable to let the source impedance have as high a resistive component as possible. However, an increase in source resistance must be accompanied by a corresponding increase in signal voltage, as is realized with an input transformer or with a "high impedance" magnetic playback head.

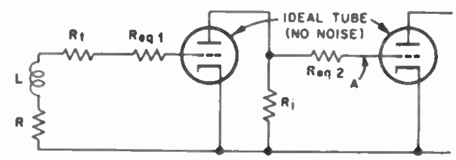

Fig. 6 – Equivalent circuit of two-stage amplifier.

A typical amplifier using a 12AY7 twin triode might have the following constants:

n = 28.3
L_s = 2 mh
R_s = 1 ohm
R_1 = 8 ohms
R_2 = 6000 ohms

$$R_{shot}\,1 = R_{shot}\,2 = \frac{2.5}{1660 \times 10^{-6}} = 1500 \text{ ohms}$$

$R_{flicker}1 = R_{flicker}2$ = 6000 ohms

$$\left.\begin{matrix} R_L = 100{,}000 \\ R_g = 1{,}000{,}000 \end{matrix}\right\} R_i = 100{,}000 \text{ ohms}$$

R_p = 25,000 ohms
G = 30

then:

$R_{eq^1} = R_{eq^2}$ = 7500 ohms
n^2 = 800
L = 1.6 hy
R = 800 ohms
R_t = 12,400 ohms

and

$$F = 1 + 15.5 + 9.4 + 0.02 = 25.9$$

or an equivalent ratio of 14.1 db.

From the above example it can be seen that the noise in this system is over 14 db worse than that which could have been obtained with an ideal amplifying device. A substantial noise contribution is made by the resistance in the input transformer, yet a negligible amount is contributed by the interstage coupling network and the second tube.

A preferred arrangement that would eliminate the noise contribution of the input transformer could be obtained by the use of a "high impedance" head. Thus:

$$n = 1$$
$$L_s = L = 1.6 \text{ hy}$$
$$R_s = R = 800 \text{ ohms}$$
$$R_1 = R_2 = R_t = 0 \text{ ohms}$$

other values as above.

Under these conditions

$$F = 1 + 9.4 + 0.02 = 10.4$$

or an equivalent ratio of 10.2 db.
Even here, the second stage noise contribution is negligibly small.

HOUSTON PGA CHAPTER

L. A. Geddes, Chairman

On Thursday, December 10, 1953, the Houston Chapter of the Professional Group on Audio held its second meeting of the season at the Haliburton Oil Well Company Auditorium in Houston, Texas. The meeting was well attended. Only standing room was available after 8:30 P.M. The attendance was made up of 6 PGA members, 12 IRE members, and 42 guests.

The program commenced with a sound film on the manufacture and testing of magnetic tape. The process described in the film was that used by Audio Devices Inc., of New York. It is similar to the production processes carried out by other manufacturers.

Following the film, a talk, "Binaural Sound Principles and Practices," was given by Harry Keep of Gulf Coast Electronics Company. The requirements of the system were outlined and was followed by a description of recording and playback equipment. A short summary of the present status of the art and the possibilities of binaural reproduction for home use, was also given.

A demonstration recording of a ping-pong game on binaural tape was reproduced through a Magnecord Recorder, McIntosh Amplifier, and two Bozak Loudspeaker Enclosures. The tape ably illustrated the principles and listening possibilities inherent in the Binaural System.

The latest Emory Cook "Sounds of our Times" twin track disc recordings were then played through the system. The level of the two Binaural Channels was altered from *zero* to *maximum* to illustrate the illusion of presence and spatial distribution obtained with this method of reproduction.

The meeting formally adjourned at 10:30 P.M. and was followed by an informal session of more recorded music on an experimental *Request-from-the-Audience* basis.

HOW MUCH DISTORTION CAN YOU HEAR?

E. M. Jones, Chairman
Cincinnati Chapter IRE-PGA

"How Much Distortion Can You Hear?" was the subject of the November 17, 1953 meeting of the Cincinnati Chapter IRE-PGA. A specially prepared series of recorded tones was played as a demonstration test of percentages of harmonic and intermodulation distortion which can be detected. The preparation of the tape recording was a cooperative effort of the executive committee consisting of E. M. Jones, The Baldwin Company; W. W. Gulden, Cincinnati and Suburban Telephone Company; J. Parke Goode, National Sound Service. They were assisted by D. W. Martin of The Baldwin Company. The demonstration also included several live A-B comparisons from the oscillators and the distorting amplifier which had been used to make the tape.

The test series consisted of 15 different sample tones having known harmonic or intermodulation distortion, compared with "undistorted" signals. In each case the distorted signal and "undistorted" signal were presented five successive times in randomized paired comparisons. For each time, the listener was asked to choose whether the "undistorted" signal came first or second. The percentage distortion and the correct answers were announced after each group of five comparisons.

The test tones were sinusoidal signals of approximately 100 and 1000 cps presented singly (for harmonic distortion) or combined (for intermodulation distortion). The distortion was artificially created by using as a voltage amplifier a push-pull pentode amplifier with feedback removed. The amount of distortion was controlled by varying the input signal while keeping the output constant by means of an output attenuator. The same signal, bypassed around the amplifier, served for the "undistorted" tone. However, when the two signals were combined for intermodulation distortion, the larger, low-frequency signal was still passed through the distorting amplifier. Thus the same harmonic distortion of the low-frequency signal would be present, and the only clue to the difference between the distorted and "undistorted" cases would be the intermodulation distortion. The tape recorder produced some measurable distortion, although recording and playback were well below the levels recommended. The power amplifier (described elsewhere in this issue) and the loudspeaker system (exponential HF and LF horns) were operated at a power of less than one watt. Acoustical measurements on the amplifier-loudspeaker output at this power showed negligible distortion.

The percentage distortion in the tones played was measured with a wave analyzer at the input to the power amplification system. For the single tones the percentage harmonic distortion was computed by dividing the square root of the sum of the squares of the second and third harmonics by the fundamental. For the combined tones, the intermodulation distortion was computed by dividing the square root of the sum of the squares of the 800, 900, 1100 and 1200 cps components by the amplitude of the 1000 cps signal (the 100 cps signal being four times as great).

It must be emphasized that the actual percentages of distortion reaching a listener could vary greatly above or below the computed values, because of unavoidable variations in frequency response of both the loudspeaker system and the auditorium. A listener might be located where there was considerable acoustic cancellation at 1000 cycles and reinforcement at the adjacent frequencies. For this listener the percentage intermodulation distortion would be greatly exaggerated.

Seventy-six test forms were turned in. A listener was considered able to detect the distortion if he guessed correctly at least four out of the five comparisons. With this definition, 19% of the listeners would "detect" the distortion by pure chance.

A summary of the results follows:

a. *Harmonic distortion of 1000 cps:* 87% of the listeners detected 11% distortion; 67% detected 4%; 42% detected 2.1%; 35% detected 0.9% distortion. "Undistorted" signal had 0.3% distortion in most cases.

b. *Harmonic distortion of 100 cps:* 42% of the listeners detected 10.5 distortion; 34% detected 4.3%; 43% detected 2.5%; 35% detected 1.7% distortion. "Undistorted" signal had .8% distortion.

c. *Intermodulation distortion of 100 and 1000 cps (on tape):* 85% of the listeners detected 22% distortion; 58% detected 11.5%; 37% detected 3.9%, and 28% detected 3.7% distortion. The "undistorted" signal had 1.5% intermodulation distortion.

d. *Intermodulation distortion of 100 and 1000 cps (live):* 42% of the listeners detected 5% distortion; 29% detected 2.2% and 21% detected 1.3% distortion. The "undistorted" signal had negligible intermodulation distortion.

Tabulation of the data according to average scores, or percentages of listeners getting five correct out of five attempts, gave similar distributions relative to the chance and perfect values.

It was concluded that most of the distortion percentages presented could be detected by a significantly large percentage of listeners in an auditorium, under the conditions of this particular experiment, i.e. simple, steady, sinusoidal tone presented immediately before or after a comparison tone that is identical, except having less distortion.

It is planned to add commentary to the tape used at this meeting for further use as a tapescript for IRE-PGA chapters.

PGA BRIEFS

Final approval was given by the IRE Executive Committee on January 5 for the establishment of IRE-PGA Chapters at both Cleveland, Ohio and Phoenix, Arizona.

Mr. Stanley K. Webster of the Beltone Hearing Aid Company gave a talk before the Cincinnati Chapter IRE-PGA on January 19, "An All Transistor Hearing Aid". The performance of the Beltone transistor hearing aids was described and the circuit designed by minimizing variations in transistor performance was explained.

We are informed that the Audio Section of the CONVENTION RECORD will be distributed to IRE-PGA members in the same manner as last year. This will be in addition to a regular May-June issue of TRANSACTIONS of the IRE-PGA. An effort will be made to obtain for later publication in TRANSACTIONS any papers which do not make the deadline for the CONVENTION RECORD. The May-June issue will contain the results of the IRE-PGA elections and audio news from the convention.

THE "VAGABOND" WIRELESS MICROPHONE SYSTEM*

Thomas W. Phinney
Shure Brothers, Inc.
Chicago, Illinois

SUMMARY – A cable-less microphone system has been developed. Designed for general public address use, the system utilizes induction coupling between the transmitter and receiver, and requires no license. The subminiature transmitter is completely contained in a stick type microphone housing. It consists of a microphone, a five-tube, printed circuit frequency-modulated transmitter, a self-contained antenna inductor, and batteries, and weighs less than one pound. The system gives excellent performance with operating areas up to 5000 square feet. The development of the system and of the subminiature transmitter is discussed and the performance of the induction system is evaluated.

I. THE FUNDAMENTAL PROBLEM – MAINTENANCE OF ACOUSTICAL COUPLING

Since the first use of electrically amplified sound, audio engineers have been faced with the problem of minimizing the pickup of extraneous sound by the microphone. The usual approach to the problem is to keep the microphone and the sound source in close proximity. This may be comparatively simple if the sound source to be amplified has a fixed location.

However, if the source is mobile, as is frequently the case, the problem of maintaining this proximity can be a very difficult one. In the case of an individual performer, several approaches to the problem are possible.

A microphone on a fixed stand can be used. This approach may fail unless the performer has had training in microphone technique. In many situations, the enforced immobility of the performer greatly reduces the utility of the system and the value of the performance.

Alternatively, a microphone can be mounted on a boom and a trained technician so manipulate the microphone as to follow closely the sound source. This approach is satisfactory for motion picture and television work, but completely unsuitable for general public performance from the standpoint of expense as well as for obvious aesthetic reasons.

A method frequently used in public address and entertainment work is to attach the microphone to the user or to have him carry it. The most serious defect in this technique is the encumbrance caused by the microphone cable. The user must move about, dragging the cable after him and avoiding entanglements.

A solution to the problem which eliminates these difficulties is the use of a miniature radio transmitter which can be carried or worn by the performer. It was for this purpose that the Shure "Vagabond" System was designed.

*Manuscript received December 29, 1953. Presented at the National Electronics Conference, Chicago, Ill., September, 1953.

II. VARIOUS APPROACHES TO THE WIRELESS MICROPHONE PROBLEM

When we set out to design a wireless microphone system, two basic avenues of approach lay before us. The first was to design a high frequency system with a large operating range which would adequately cover Madison Square Garden, or Soldiers Field. Such a system must necessarily be relatively high powered and hence have short battery life. It would definitely require a Federal License. This would greatly restrict, or even prohibit its use in the very applications which we wished to reach, namely, general public address work, and theater and night club entertainment. Furthermore, those frequency assignments which might be obtained are available on a non-exclusive basis and will be subject to greater and greater interference as time goes on. A final disadvantage of the licensed system was the strict technical requirements imposed by law. These requirements would make the design of a miniature transmitter more difficult and would certainly increase its cost.

The second approach available to us was to design an induction system which would have a relatively restricted operating range, but would require no Federal licensing and hence be available to any and all potential users. If such a system could be developed to give adequate performance, it would offer excellent operating economy through prolonged battery life, and it would have to meet only those technical requirements dictated by satisfactory operation. Since it was felt that a practical induction system could be developed, that approach was chosen for the "Vagabond" system.

III. THE FUNDAMENTAL DESIGN OF THE TRANSMITTER UNIT

Having decided upon a non-licensed or induction system, other fundamental design decisions had to be made. These were the choice of the physical form of the transmitter, the modulation system to be employed, and the operating frequency of the system.

The Physical Form of the Transmitter

Two physical forms were considered for the transmitting unit. One form consisted of a pocket-size case, containing the transmitter and batteries, which could be worn concealed about the person of the user, with a separate lapel-type microphone attached to the transmitter with a cable. The second form considered was that of a stick-type case which would contain the microphone, antenna, transmitter, and batteries.

The stick form was adopted, as it offered several advantages. Being completely self-contained, the stick mike could readily be handed from one person to another, or it could be placed in a microphone stand and used conventionally. The performer would not have to "dress" himself in the microphone. Having no interconnecting cables, the self-contained unit would not be subject to wear and tear and would require less maintenance. It was also felt that the stick form, being quite similar to several conventional microphones in appearance and use, would place performers at ease and hence would be more readily accepted. The only manifest disadvantage of the stick form was its inability to be concealed easily on the person. This was felt to be a minor consideration, while the ability of one microphone to be used by several performers in immediate succession represented an economic advantage as it would replace several wearable units.

Choosing the Modulation System

Two primary considerations led to the choice of frequency modulation for the "Vagabond" System. Since the coupling between the transmitting and receiving antennas in a wireless microphone system will vary greatly, some means of assuring constant over-all audio gain is imperative. A frequency modulation system was chosen as the simplest means of achieving this goal.

Secondly, the desire to obtain the best possible signal-to-noise performance, with a transmitter of limited power, again indicated the use of a frequency modulation system. A modulation index of at least five was set as a goal.

Choosing the Carrier Frequency

The carrier frequency chosen for the "Vagabond" system was a compromise of opposing factors.

To obtain reasonable performance from a miniature transmitting antenna and to accommodate the bandwidth required by the frequency modulation system, it was desirable to use the highest possible carrier frequency. Furthermore, a study of the literature on the occurrence of noise throughout the radio spectrum indicated that less interference could be expected at the higher frequencies.

On the other hand, the Federal Communications Commission regulation governing unlicensed transmitters limits the radiated field strength in inverse proportion to the carrier frequency, thus dictating the use of the lowest possible frequency.

A final consideration was the minimization of interference from other transmitters operating on the same frequency. This obviated the use of the Broadcast, Amateur, Police, and Loran bands which lie between 0.5 and 2.0 megacycles. In view of all these factors, a carrier frequency of approximately 2.1 mc was chosen.

IV. THE MECHANICAL AND ELECTRICAL DESIGN OF THE TRANSMITTER UNIT

The Transmitting Antenna

Since the feasibility of the entire stick design depended on obtaining a satisfactory miniature antenna, the development of such an antenna was undertaken. Calculation and experiment indicated that within practical limits the size of the transmitting antenna would not affect the performance of an induction system. Since any antenna could be reduced to an equivalent magnetic dipole, the only effect of size variation was to change the power required to produce the desired field strength. A satisfactory ferrite-core transmitting inductor was developed which was three inches long and weighed two ounces.

Transmitter Circuit Design

Having achieved a practical ferrite antenna, a transmitter circuit was developed to operate from a 30-volt hearing-aid battery and a 1.3-volt mercury cell. (See Fig. 1.) Five subminiature tubes were used. The transmitter circuit is divided into two sections: a two-tube audio section and a three-tube radio frequency section.

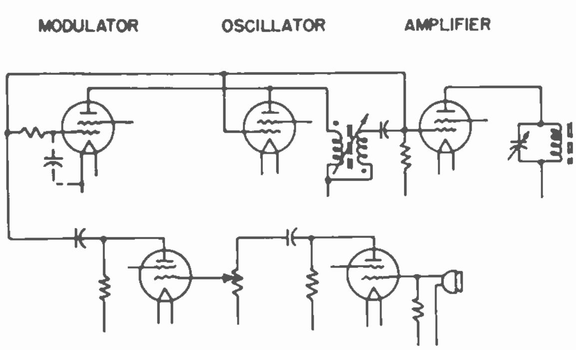

Fig. 1 – Transmitter schematic diagram.

Two tetrode voltage amplifiers are cascaded in the audio section to obtain a gain of 55 db at 1000 cycles. A miniature volume control between the two stages allows the gain to be adjusted for the desired degree of modulation. While the two-stage audio amplifier alone showed no tendency toward regeneration without any decoupling networks, motorboating at low frequencies occurred when it was connected to the reactance modulator, due to modulation of the plate supply voltage. To eliminate this motorboating, it was necessary to add a decoupling filter in the first audio stage and to restrict the low frequency response of the audio amplifier as well.

In order to obtain the best possible signal-to-noise ratio in the over-all system, an 80 microsecond pre-emphasis was used in the transmitter and a corresponding de-emphasis in the receiver. The problem of obtaining sufficient audio gain together with the necessary pre-emphasis was solved by providing a microphone cartridge of special design, with a response which very closely approximated the desired pre-emphasis curve. (See Fig. 2.)

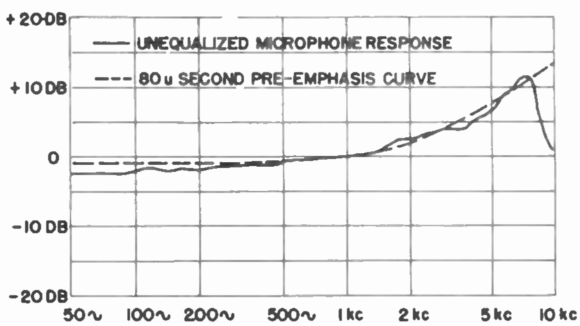

Fig. 2 – Microphone response curve.

An omnidirectional ceramic microphone cartridge was used. The choice of an omnidirectional unit was in conformity with present day trends in stick microphone design.

The radio frequency section of the transmitter consists of a self-controlled oscillator, a reactance modulator, and a radio frequency amplifier.

After trying many circuit arrangements utilizing only two tubes, the RF power amplifier stage was added. When only the two tubes were used, the oscillator necessarily had to operate at a higher power level, as its plate load was the antenna circuit. This arrangement required excessive modulator plate current to produce the two per cent frequency deviation required.

In order for a reactance modulator to cause a frequency shift of 1%, it must produce a reactive current which is approximately 2% of the magnitude of the circulating current in the oscillator tank circuit. If the oscillator tank circuit has a Q of 50, a typical value in this application, the reactance modulator must draw an average ac current equal to that of the oscillator and must be capable of 100% modulation of that current without distortion. This means that the modulator must draw three or four times as much plate current as the oscillator. This was obviously impractical, and it was necessary to use a low power oscillator followed by a power amplifier. A further advantage of such an arrangement was the elimination of carrier frequency shift because of antenna detuning caused by hand capacity.

A second circuit feature is the method of feeding signal and bias to the reactance modulator grid. Two operational difficulties are substantially reduced by this circuit arrangement. They are: first, the amplitude modulation of the oscillator due to variable loading by the reactance modulator, which produces distortion; and second, the carrier frequency shift which occurs as the batteries run down.

By obtaining the ac grid voltage for the reactance modulator from the secondary of the oscillator plate transformer, (connected out of phase with the primary), it is possible to use a single rc phase-shifting network and obtain a phase difference between the modulator grid and plate voltages which exceeds 90 degrees. The use of a modulator grid voltage which is more than 90 degrees out of phase with the plate voltage causes the reactance modulator to appear as a reactance in parallel with a negative resistance. This negative resistance will vary as the voltage on the modulator grid varies, and, with proper circuit adjustment, can be made to compensate for the variable loading of the oscillator by the modulator grid circuit. This minimizes the amplitude modulation of the oscillator and reduces the over-all distortion. Using the reactance modulator circuit shown, it has been possible to obtain excellent modulator performance together with adequate rf amplifier grid driving voltage at a reasonable total plate current. (See Fig. 3.)

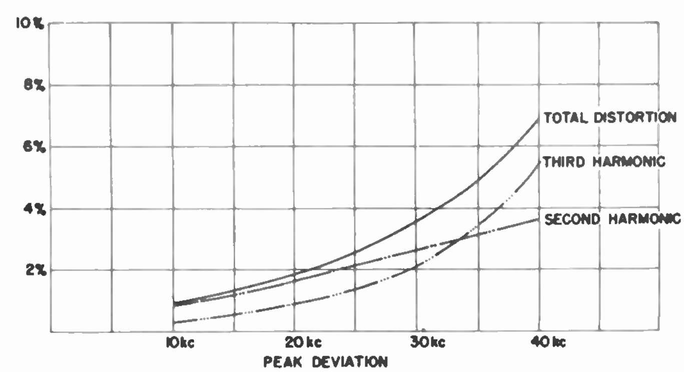

Fig. 3 – Transmitter distortion characteristic.

Since the use of any sort of automatic frequency control in the transmitter was out of the question due to the complex circuitry involved and the minute space available, it was necessary to design an oscillator-modulator circuit which would have sufficient frequency stability to meet the requirements of the system. This stability was achieved by means of the modulator grid bias circuit shown. The bias is derived from the self-rectified voltage developed by the oscillator and amplifier grids. As the plate and filament batteries run down in use, the transconductance of the reactance modulator will tend to be reduced and cause the carrier frequency of the transmitter to shift. However, the circuit arrangement used causes the dc grid bias of the modulator to be changed in such a fashion that its transconductance variation is greatly reduced.

Construction of the Printed Circuit Chassis

The entire transmitter circuitry, excluding the antenna and tubes, occupies a volume of about one cubic inch. The chassis casting, which contains eight condensers, eleven resistors, a volume control, five subminiature tube sockets, and a powdered iron core oscillator coil, is a cylinder one inch in diameter and 1.3 inches long. To obtain such extreme miniaturization, it was necessary to use rather unorthodox construction. (See Fig. 4.)

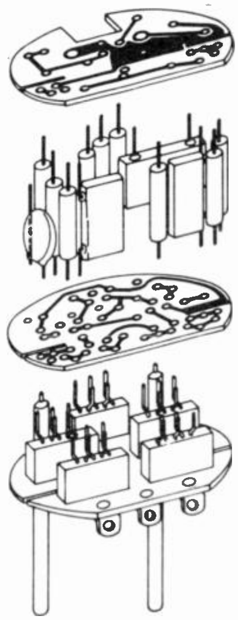

Fig. 4.

Two printed circuit plates form the fundamental structure of the transmitter chassis. They are made by a technique which produces a conductive pattern on both sides of a phenolic plate and on the inside surface of the holes through the plate. It is thus possible to produce cross connections between the two sides of the phenolic plate, automatically, and without the aid of eyelets. Since one printed circuit plate, with an area of less than one square inch, contains 54 holes for tube socket pins and component leads, the use of eyelets for making cross connections would be impossible. This is further emphasized by the fact that the center to center spacing of the tube socket pins is only 0.100 inch. A very important advantage of printed circuits of this type is the fact that it is possible to do all soldering on the upper side of the plate and still make connections to the pattern printed on the under side. In this assembly, the soldering of the 25 tube socket connections could be done in no other way.

Assembly of the chassis is begun by mounting the tube sockets in a phenolic plate which serves to position them and which carries three terminals for connection to the gain control. The lower printed circuit plate is then placed over the tube socket pins and soldered to them from the top side. Next, the various resistors and condensers are soldered to the top side of the lower plate. Finally, the upper printed circuit plate is fitted over the leads from the components, and the leads are cut off and soldered to the printed circuit. Again, connections are made to the printed pattern on the under side of the top plate by soldering on the upper side. (See Fig. 5.)

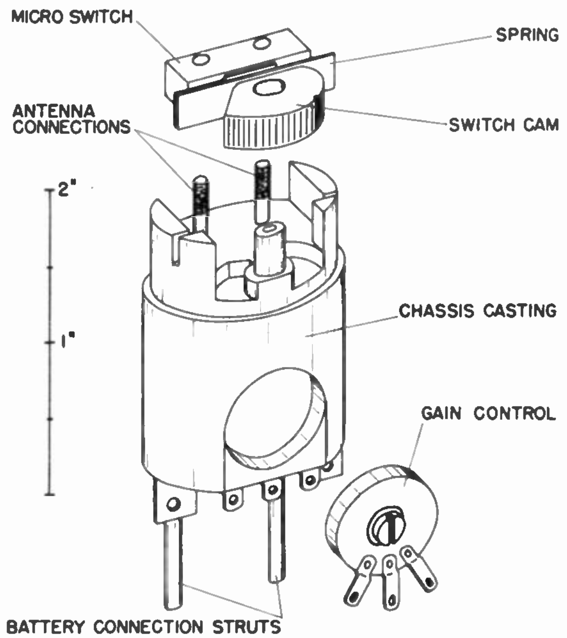

Fig. 5 – Transmitter chassis assembly.

After the printed circuits and components are assembled and inspected, the entire unit is placed in a mold and filled with a casting resin. Upon setting, this resin forms a rigid, moisture-proof mechanical assembly which cannot be damaged by vibration or shock, and which is readily attached to the other portions of the transmitter. An additional advantage of such an embedded circuit is the immobilization of all leads and components which eliminates any possibility of detuning due to mechanical motion.

After the chassis assembly is cast in resin, the gain control, filament circuit microswitch, and battery contact assembly are attached. (See Fig. 6.) The addition of the antenna and microphone completes the transmitter

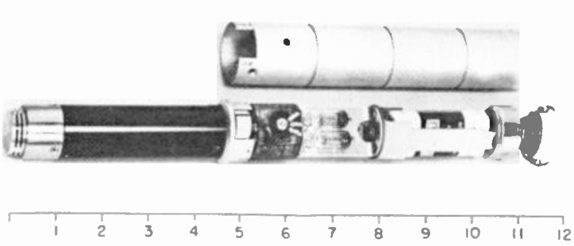

Fig. 6 – Complete transmitter chassis.

chassis assembly. A tubular case is attached to the transmitter chassis with three screws after the tubes are installed. The batteries are inserted in the bottom of the case and the battery cap is snapped into place. Connection to the junction between the filament and plate batteries is made automatically by the battery contact assembly.

Casting an entire circuit in plastic is certainly a case of "burning your bridges behind you," as it is virtually impossible to repair such a unit. However, any properly designed assembly which was good before casting, should be good after casting, and should remain good indefinitely. Component failure is very unlikely since the casting resin provides protection from corrosion and mechanical damage which are the chief causes of failure in very low power circuits.

It should be added, however, that a circuit which is to be cast must be designed to make allowance for the increased distributed capacitance caused by the dielectric properties of the casting resin. To simulate this dielectric effect during development work, sample assemblies were immersed in cottonseed oil, which closely duplicated the dielectric constant of the casting resin used. This proved to be an extremely useful experimental technique. Fig. 7 is a view of the completed transmitting unit.

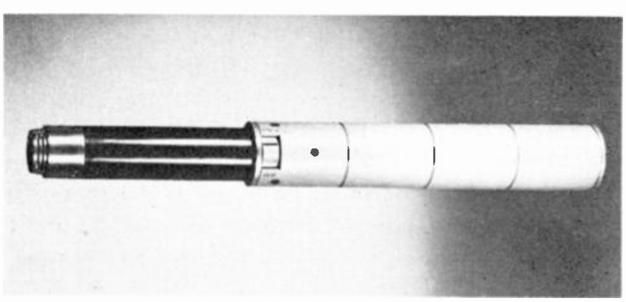

Fig. 7 — Over-all view of transmitter.

V. THE VAGABOND RECEIVER

The high performance superheterodyne receiver designed for the Vagabond System has a pentode, tuned radio frequency amplifier with a bandwidth of 150 kilocycles. A pentode mixer with a separate triode oscillator is used and automatic frequency control is provided by a pentode reactance modulator. The wide-band intermediate frequency amplifier employs two pentode amplifier stages, two cascaded triode limiters and a Foster-Seeley discriminator. The output of the discriminator is fed through a gated, cathode follower triode to the audio output terminals of the receiver. A triode DC amplifier is provided in the carrier-operated squelch circuit.

To aid in tuning the Vagabond transmitter, the receiver incorporates a double target tuning eye controlled by the outputs of the discriminator and the first limiter. By means of this tuning eye, it is possible to tune both the oscillator and antenna circuits of the transmitter. A three position switch on the receiver shifts the frequency of the receiver oscillator in 50 kilocycle steps to provide three operating channels. These may be used when more than one Vagabond System is to be operated simultaneously in the same vicinity or when strong interference is experienced on any one channel. Since the radio frequency amplifier has a 150 kilocycle bandwidth, no retuning of the receiver is required.

VI. CONCLUSION

The Vagabond System has been installed in a number of locations and some evaluation of the induction system can be made. It has provided ample quality for general public address and entertainment work with operating areas of 500 to 5000 square feet depending upon local interference conditions and the nature of the particular application. The audio response and signal-to-noise ratio have proved sufficient for radio broadcasting purposes in some studio installations.

The objective of developing a transmitting unit comparable in size and weight to current stick type microphones has been met. The transmitting unit weighs less than one pound and is 1¼ inches in diameter and 12 inches long. The life of a set of batteries is 40 operating hours. The battery cost is about five cents an hour.

Although the operating range of the induction system is limited, there are many applications where it has great value and where it alone is available.

VII. ACKNOWLEDGMENTS

The author wishes to express his gratitude to his fellow employees for the generous assistance, especially to Mr. John Knox, Assistant Project Engineer, Mr. J. S. Knechtsberger, Project Mechanical Engineer, and Mr. J. W. Medill, who developed the special microphone cartridge. Deep appreciation is also extended to Mr. B. B. Bauer, Vice-President — Engineering, and to Mr. E. V. Carlson, Development Engineer, for their many contributions to the project.

A HIGH EFFICIENCY – HIGH QUALITY AUDIO FREQUENCY POWER AMPLIFIER*

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The development of the power amplifier discussed in this paper was undertaken with the objective of providing a large amount of good quality audio power in a small package at relatively low cost. The size of an amplifier package and its cost are dependent largely on the efficiency of operation and the power sensitivity of the output stage. Beam power tubes satisfy these two requirements more readily than triodes. The quality of the audio power is improved by the use of push-pull operation and large amounts of feedback. The use of a suitable amount and type of feedback with beam power tubes eliminates the advantage of inherent low output impedance obtained with triodes.

In Class A or Class AB operation the dynamic characteristics of the two tubes may be matched to obtain a reasonably linear characteristic, and while feedback is desirable its use may sometimes be avoided. In Class B operation it is not possible to match dynamic characteristics to get linear operation, and therefore a large amount of feedback must always be used. The problems encountered in Class B operation are considerably more severe than those occurring in Class A and Class AB operation, and the question naturally arises as to whether this type of operation is worth the extra effort required. This question can be resolved rather conclusively by the following example:

Two 6*L*6 tubes operating Class A_1 push-pull with 30% efficiency, and having an allowable total plate dissipation of 40 watts, would be capable of developing an output power of $\frac{30}{70} \times 40 = 17.1$ watts.

The same two 6L6 tubes operating Class B_1, push-pull with 60% efficiency would be capable of developing $\frac{60}{40} \times 40 = 60$ watts. The possibility of getting 60 watts of output power in Class *B* operation for the same investment in power tubes which produced only 17.1 watts in Class *A* operation provides a real incentive to solve the severe problems encountered in Class *B* operation. The advantage of high efficiency operation also reflects advantageously on the size of the power supply necessary to operate the power amplifier.

The basic power amplifier circuit shown in Fig. 1 is capable of employing large amounts of feedback with good stability. This circuit has been operated with 36 *db* of feedback without showing any trace of instability. However, since the driving voltage required under these conditions is too great, the circuit is normally used with only 24 *db* of feedback. In this circuit the 12AX7 tube is used as a phase inverter – amplifier – driver stage. It is direct-coupled to the beam power output tubes and the

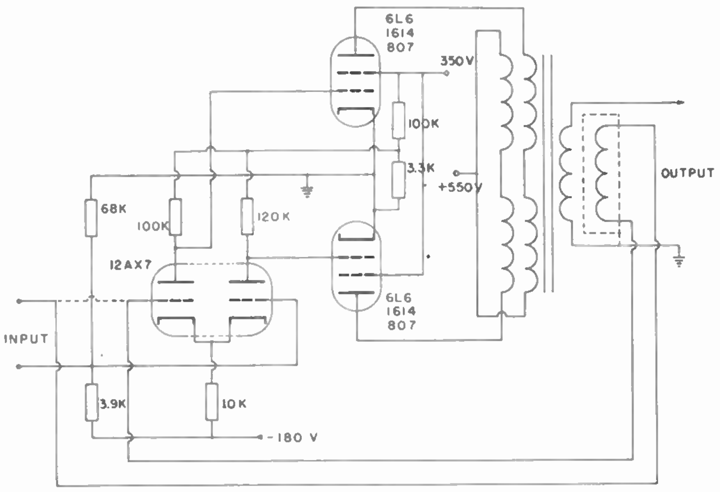

Fig. 1 – Basic Bereskin power amplifier circuit.

operation is essentially Class B_1 since the high impedance driver stage is incapable of driving the power tube grids positive. An added advantage of the direct-coupled driver is that it eliminates the possibility of blocking due to excessive input signal. The two output-tube cathodes are returned to ground and therefore any combination of screen and plate supply voltage may be used. The values shown on the diagram will keep both the screen and plate dissipation below the rated values for full signal Class B_1 operation with either the 1614 or the 807 tubes. The bias for the output tubes is supplied by the direct-coupled phase inverter – amplifier – driver and is normally adjusted to produce a zero signal plate current of about 15 ma per tube. The high value of cathode resistance employed makes the driver circuit fundamentally stable. A test with 6 different 1614 tubes and 12 different 12AX7 tubes, of different manufacturers and chosen at random, produced a zero-signal plate current variation ranging from 10 to 25 ma per tube. The full signal operation was substantially independent of the choice of 12*A*X7 and beam power output tubes. The feedback winding used is electrostatically shielded from the secondary but very closely coupled to it. The elimination of the electrostatic shield greatly reduces the amount of feedback that can be used successfully.

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One of the major problems associated with Class *B* operation is due to the energy stored in the leakage reactance between the two primary windings. A. P. Sah[1] showed how this stored energy gave rise to a conduction transfer notch which must be eliminated before Class *B* operation can be used successfully. Several different winding schemes will reduce the leakage reactance below the critical value but the most successful one is that of using a bifilar winding for the two primary sections.

The use of bifilar windings introduces new problems which were not previously important. One of these problems is that appreciable voltage may exist between the adjacent wires of a bifilar winding and sufficient insulation must be provided to withstand this voltage. Adequately insulated wires are now available commercially so this problem is seldom a serious one. A second and more significant problem is that appreciable capacitance exists between the adjacent wires in a bifilar winding and that charging current must be supplied to this capacitance before any voltage can be developed between the wires. This charging current must usually be supplied through the output stage tubes and is one of the major factors limiting the high frequency power-delivering capacity of an amplifier.

This problem can be understood more readily by examining the circuit of Fig. 2A. In this circuit the bifilar primary has been separated into two sections and the secondary has been sandwiched between them to keep the leakage reactance between the primaries and secondary at a low value. The four primary sections have been symmetrically interconnected and there is negligible dc voltage between these windings. If a peak signal voltage of 250 v is assumed on each of the primary sections, as shown by the vertical arrows, it is observed that this will give rise to a peak signal voltage of 500 v between all adjacent points on the bifilar winding as shown by the horizontal arrows. The undesirable feature here is that before this voltage can appear between the two primary windings the interwinding capacitance must be suitably charged, and the charging current must flow through one of the two tubes. An experimental transformer of this type, wound with adjacent-wire layer bifilar winding, using No. 28 Heavy Formvar wire, was found to have a capacitance of 0.045 microfarads between the two primary windings. A peak charging current of 1.5 amperes is required to charge this capacitance at 10KC with a sine wave voltage of 500 volts peak across the primaries. An ordinary tube used in a circuit of this type would normally supply only about 0.30 amperes peak so that at 2 Σkc the peak current capacity of the tube would be required to furnish the charging current for the interwinding capacitance. The amplifier's power-delivering capacity to a resistance load would therefore be down 3 db at 2 Σkc.

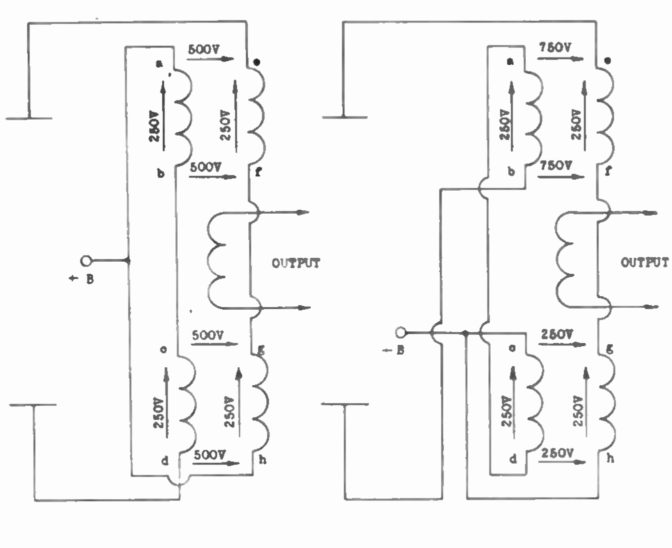

Fig. 2 – Primary interwinding voltage relations.

A natural step at this point is to consider the possibility of using a different interconnection of the primary sections to reduce the interwinding voltages, thereby reducing the charging current required for the interwinding capacitance. Fig. 2B shows a different connection of the primary sections. The voltage between the lower sections of the bifilar winding has been reduced to 250 v, but the voltage between the upper sections has been correspondingly increased to 750 v. The total interwinding charging current is the same as before but the insulation burden on the upper section is greater than it was before. As a matter of fact, further sectionalization and reconnection of the primary does not reduce the charging current problem and will, in most cases, increase the burden on the insulation.

A different type of interconnection is shown in Fig. 3. In this case half of the sectionalized bifilar

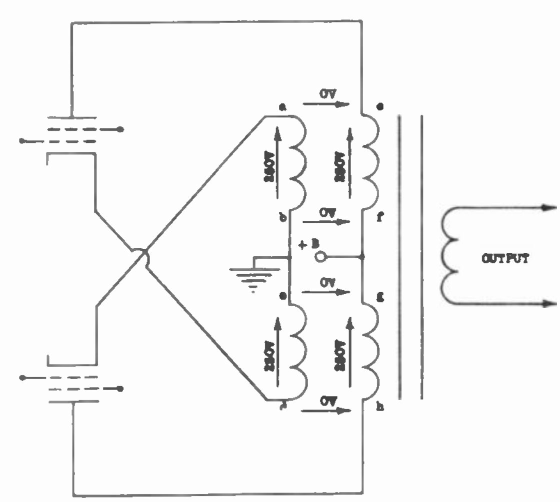

Fig. 3 – Basic McIntosh power amplifier.

[1]A. Pen-Tung Sah, "Quasi-transients in Class B audio-frequency push-pull amplifiers," *Proc. I.R.E.*, vol. 24, no. 11, pp. 1522-1541; November, 1936.

primary winding is connected in the plate circuit, as before, and the other half, with proper consideration for the signal polarities, is connected in the cathode circuit. Now points b, c, f, and g are all at zero ac signal potential. For signal conditions the same as those considered before, when a is instantaneously positive by 250 v with respect to b, e is positive by the same amount with respect to f. This means there will be zero ac signal voltage between points e and a. The same is seen to be true for points h and d and for all other adjacent points on the two bifilar windings. We now require zero charging current for the primary interwinding capacitance regardless of the value of this capacitance. Since point h is at the same ac signal potential as d, but at a much higher dc potential, it could be connected to the screen of the upper tube to supply a constant screen-cathode voltage for this tube. In a like manner point e could be connected to the screen of the lower tube. The screen and plate supply voltages in this circuit will be equal unless special circuitry is provided to make them different. This circuit also requires a grid drive voltage greater than 50% of the output transformer primary voltage. When this circuit is combined with a suitable driving circuit and feedback networks it becomes the McIntosh Power Amplifier[2] and is capable of delivering 50 w of high quality power over an exceptionally large frequency range.

A different solution to the problem is provided by the basic Sinclair-Peterson[3] circuit shown in Fig. 4A and its transformer coupled equivalent, using beam power tubes, in Fig. 4B. In the transformer coupled case both

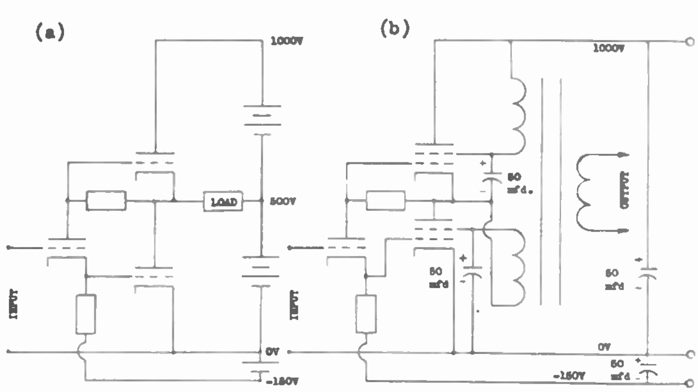

Fig. 4 – Basic Sinclair – Peterson power amplifier.

primary sections of the transformer work at all times so they need not be bifilar wound. These circuits present an appreciable burden to the phase inverter-driver for, if a peak primary or load voltage of 450 volts is assumed, it is seen that the phase inverter-driver plate supply voltage fluctuates between 1100 volts and 200 volts during the course of one cycle. Suitable points are available in Fig. 4B for supplying the screen voltages of the beam power tubes. A disadvantage of this connection is that the plate and screen supply voltages are forced to be equal unless special additional circuitry is used. The basic Sinclair-Peterson circuit has been incorporated in two complete power amplifier circuits which have been described in the literature[4,5]. In these circuits the screens are fed through suitable voltage dropping VR tubes to insure that the screen voltage will be less than the average plate supply voltage. The published data on these amplifiers indicate that their performance is substantially the same as that of the McIntosh amplifier mentioned previously. Neither the McIntosh nor the Sinclair-Peterson circuits use feedback around the output transformer.

The design of the transformer used in the circuit of Fig. 1 represents another solution to the problem introduced by the primary interwinding capacitance. In the general case of two isolated parallel circular wires an increase of the spacing between the surfaces from 10% to 20% of the diameter of the wire will reduce the capacitance between these wires by approximately 30%. An increase of this spacing from 10% to 100% of the diameter of the wire will reduce the capacitance by approximately 70%.

In the transformer winding even though we are not dealing with two isolated parallel wires but with many wires in a small space, the same general principles apply. Each wire will have capacitance to the two wires on each side of it in the same layer and also to the wires in the layers above and below it. The capacitance between wires in the same layer can be cut in half by transposing the two wires of the bifilar wire at every turn. The capacitance between wires in adjacent layers will not be modified by this process. In the non-transposed winding, assuming the same spacing between layer centers that exists between adjacent wire centers in a layer, and uniform dielectric material, the capacitance between the wires in the layers accounts for approximately two thirds of the total capacitance, while the capacitance between wires in adjacent layers accounts for one third of the total capacitance. Since transposition of the wires can be expected to cut in half the capacitance between wires in a layer without disturbing the capacitance between wires in adjacent layers, it should reduce the total capacitance to two thirds of its original value. The use of insulating materials such as cotton, varnish, and wax, and the accumulation of moisture will all tend to increase the capacitance. A mechanism has been developed which automatically transposes the bifilar winding at every turn so that this type of winding is no more

[2] F. H. McIntosh & G. J. Gow, "Description and analysis of a new 50 watt amplifier circuit," *Audio Eng.*, vol. 33, no. 12, pp. 9-11, 35-40; December, 1949.

[3] A. Peterson and D. B. Sinclair, "A single ended push-pull audio amplifier," *Proc. I.R.E.*, vol. 40, no. 1, pp. 7-11, January, 1952.

[4] A. P. Peterson, "A new push-pull amplifier circuit," *The General Radio Experimenter*," vol. 26, no. 5, pp. 1-7; October, 1951.

[5] H. W. Lamson, "A high power toroidal output transformer," *The General Radio Experimenter*, vol. 26, no. 6, pp. 5-8; November, 1951.

difficult to make than any ordinary bifilar winding.

An alternative to the transposed bifilar winding is a random wound bifilar winding. The random winding is not as consistent as the transposed winding but appears on an average to produce an increase in the capacitance of approximately 15%. Other than this the two windings are equivalent.

The introduction of space between the adjacent primary wires will tend to increase the leakage reactance between the two primary sections. The safety factor provided by the bifilar type of winding has so far been sufficiently large to avoid the appearance of the conduction transfer notch.

Fig. 5 is a schematic diagram of the output transformer used to make the tests discussed in the remainder of this paper, and Fig. 6 shows the coil buildup that was used. Fig. 6 is drawn in proper vertical scale but the horizontal scale has been modified to show the relative positions of the windings without showing the true coil width. This transformer winding was designed to be used with two grain-oriented Hipersil C Cores (Moloney ME-31 Hipercores or equivalent). The nominal impedance levels were intended to be 4, 8, and 16 ohms but the optimum levels obtained with this transformer were 4.63,

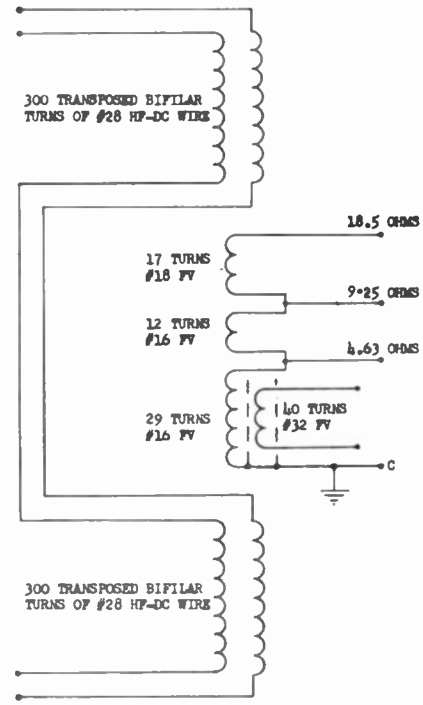

Fig. 5 – Output transformer winding arrangement.

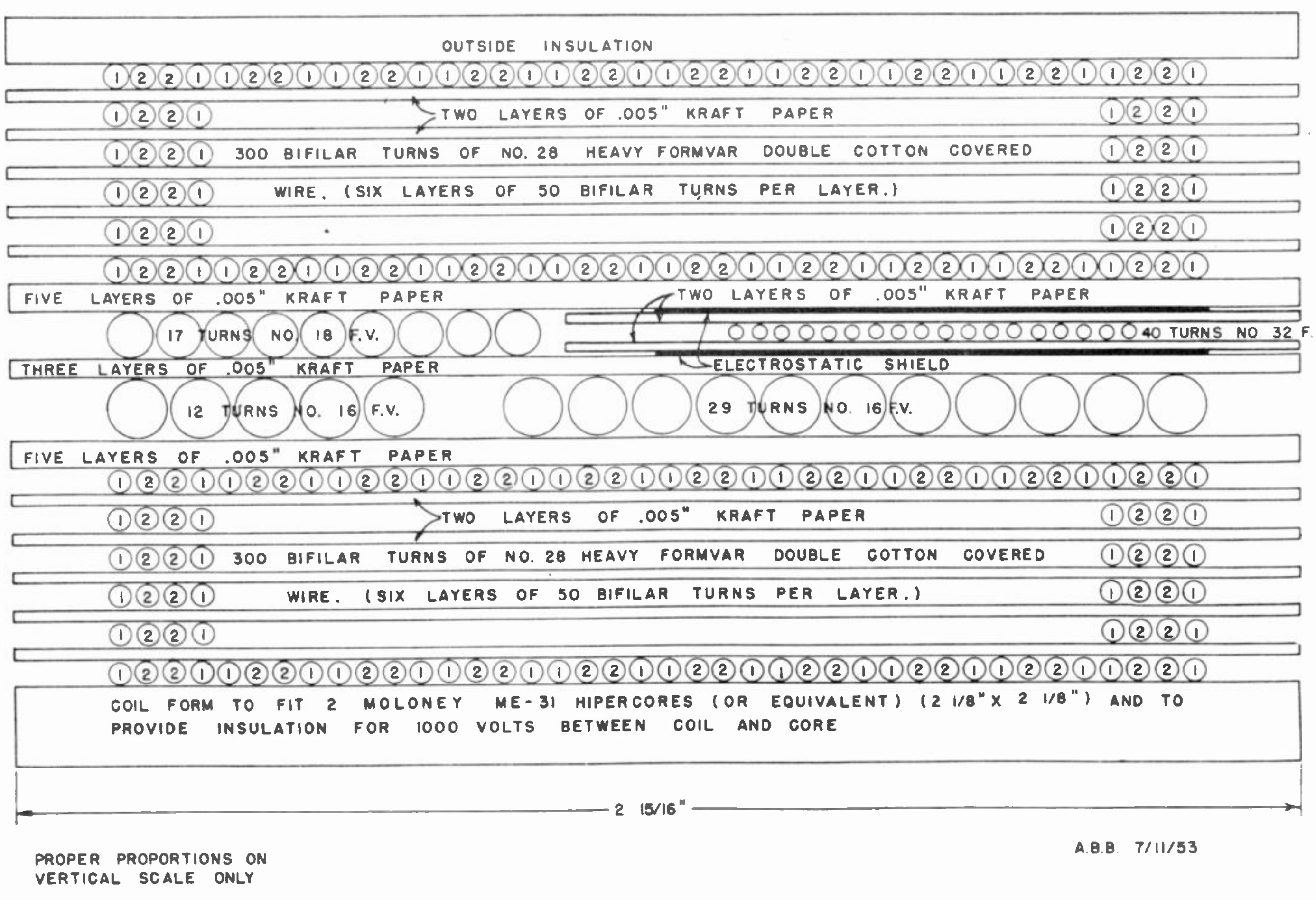

Fig. 6 – Output transformer coil buildup for Bereskin 50 watt 1614 tube power amplifier.

9.25, and 18.5 ohms. An optimum value resistor was used on the 9.25 ohm tap in all of the succeeding tests.

The transformer of Fig. 6 had a primary interwinding capacitance of 0.010 microfarads when it was vacuum impregnated with GE Type 9700 clear baking varnish and baked for the prescribed amount of time. This transformer was potted and has maintained the same value of capacitance since that time. Another transformer winding differing from this one only in the fact it had an ordinary layer bifilar winding had a primary interwinding capacitance of 0.015 mf. A third transformer winding in which No. 28 quadruple Formvar wire was used with the same spacing between wire centers and an ordinary layer bifilar winding had a primary interwinding capacitance of 0.012 microfarads. Additional paper was used between layers in this case to achieve the same spacing between wire centers in adjacent layers. This last winding has been found to have less tendency to pick up moisture from the atmosphere than the other two windings although potting eliminates this trouble in all cases.

In order to obtain the same low frequency power delivering capacity with an ordinary non grain-oriented core it is necessary to increase both the core cross section and the number of turns by about 25%. The combined effect of these increases is that the primary interwinding capacitance increases by approximately 50%.

The circuit of Fig. 1 was used with the transformer of Fig. 6 with transformer coupled input. The input transformer had an electrostatic shield and this shield was connected to ground. All filaments were operated from a common filament supply, one end of which was connected to ground. With well regulated screen and plate supply voltages, the residual hum in the output was 96 db below 50 w. For this particular amplifier, without any special attempt to obtain good balance, a ripple voltage of 42 volts inserted in series with the plate supply, or 9 v inserted in series with the screen supply, was necessary to bring the residual hum level in the output to 80 db below 50 w.

The power delivering capacity of this amplifier was tested by setting the input at the value necessary to produce 2% distortion in the output. Fig. 7 shows the results of this test. In the range below 30 cs the output was limited by the inability of the 1614 tubes to supply adequate magnetizing current to the transformer. Between 30 and 3000 cs the output was limited by peak clipping due to the inability of the 12AX7 tubes to drive the 1614 tube grids positive. Above 3000 cs the output was limited by the inability of the 1614 tubes to supply the charging current required by the primary interwinding capacitance. At the low end the amplifier has a drop-off rate of 9db/octave while at the high end it approaches a drop-off rate of 6 db/octave.

Most of the power in speech, song, and music is contained in the fundamental tones with frequencies below 3000 cs. The power levels of the higher frequency fundamental tones and of the harmonics of the lower frequency fundamentals drop off at a greater rate than the power-delivering capacity of this amplifier. It will be shown in the appendix that the power-delivering capacity of this amplifier is fully adequate for all audio frequency signals.

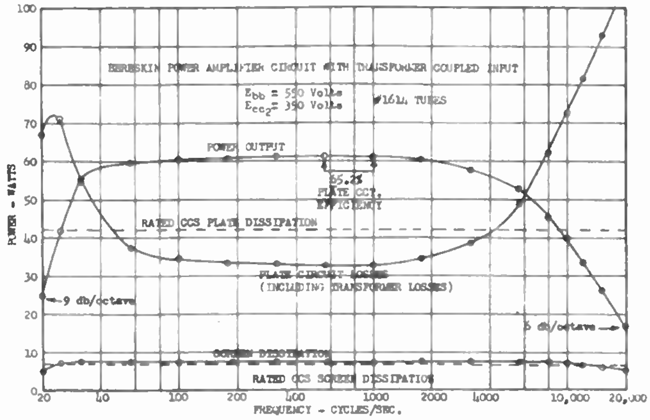

Fig. 7 – Two per cent distortion – power relations.

The amplifier develops its full power output of 60 watts over most of the middle frequency range with total plate circuit losses, including transformer losses, considerably lower than the rated CCS values. The screen dissipation exceeds the rated CCS value by about 7%, but this is only because the amplifier is being over driven to obtain the required 2% distortion. A reduction of 3 w in the output power, over most of the range, brings the harmonic distortion below 1% and the screen dissipation safely below the rated CCS value. The maximum plate circuit efficiency occurred in the 500-1000 cs. range and was 65.2%. This value includes the output transformer losses, and is ramarkably close to the ideal value of 78.5% for Class B operation, which does not include the transformer losses.

Since transformer coupled input is not usually available for amplifiers of this type a circuit including a preamplifier and power supply was developed and is shown in Fig. 8. This circuit was designed to deliver 50 w of high quality power over most of the middle frequency range. Since the output stage is very insensitive to ripple in the screen and plate supply circuits, very simple power supply filter circuits were adequate. Filter chokes were not necessary in either the plate or screen supply circuits. A single 5U4–G rectifier tube, operating within the manufacturer's ratings, was adequate to supply the power required by the plate circuits of the 1614 tubes. One 6X4 tube is used to supply the power required by the preamplifier and the screens of the 1614 tubes, while another 6X4 tube is used to supply the negative voltage required by the 12AX7 tube. The two 6X4 tubes could be replaced by a single rectifier tube with separate cathodes and plates, used as two single phase half-wave rectifiers, but the tubes available cost more than the two 6X4 tubes put together and, in addition, full-wave rectification in

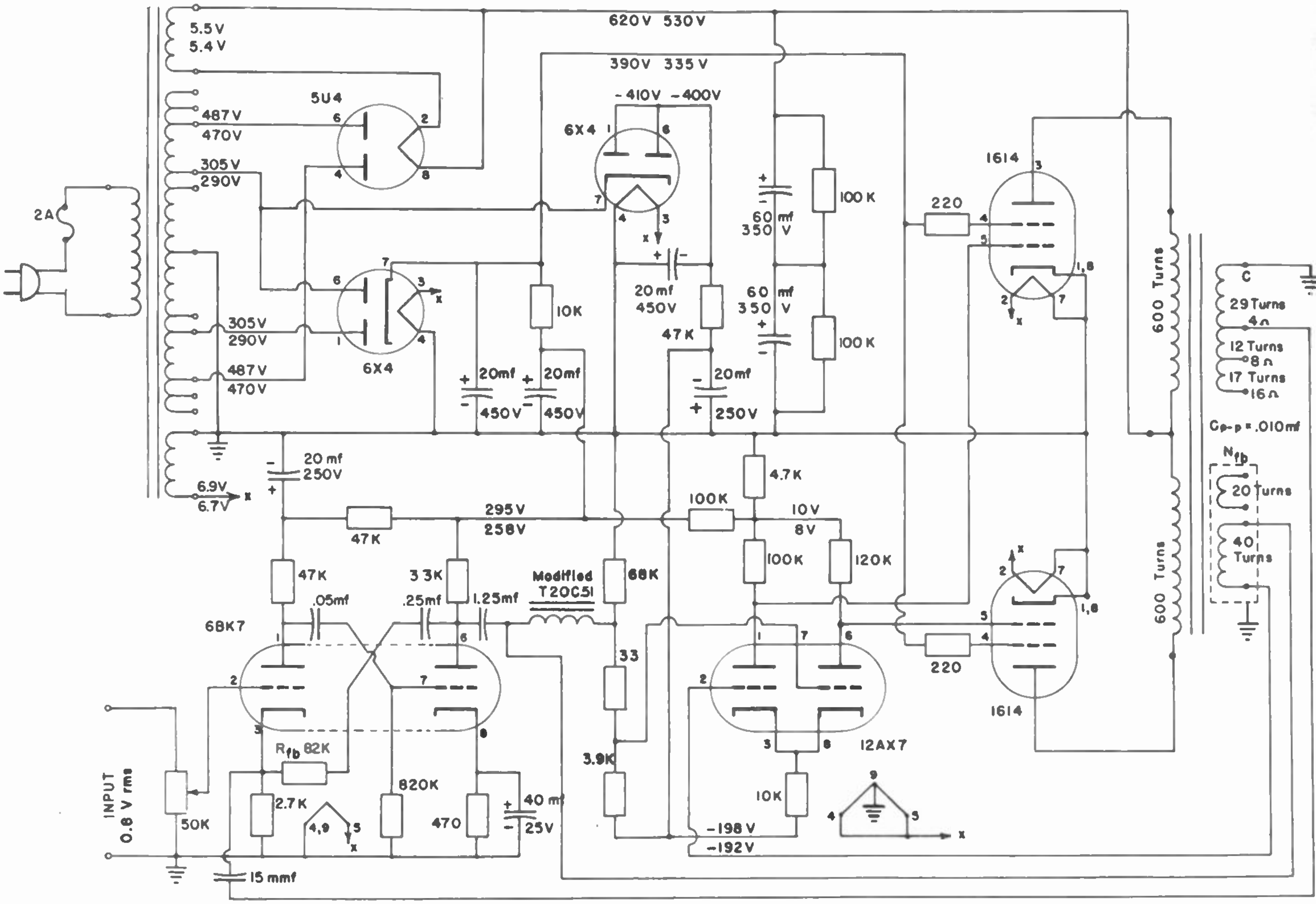

Fig. 8 – Bereskin 50 watt 1614 tube power amplifier.

the screen supply permits the use of a lower value of filter capacitance. The voltages indicated at various locations on the diagram are the measured no-load and full-load values.

The preamplifier consists of a two stage resistance-capacitance coupled amplifier with feedback between the second plate and the first cathode. This feedback provides good wave shape and low output impedance on the preamplifier. The preamplifier is coupled to the 12AX7 grid with a 1.25 μf capacitor and a modified Thordarson T20C51 choke. The modification consists in interleaving the laminations of this choke. A low dc resistance is necessary in this circuit because the 12AX7 has appreciable grid current when the grid voltage becomes more positive than -1 volt and this grid current must not be allowed to change the bias relations of the phase inverter. This coupling circuit has a low Q resonance between 10 and 15 cs.

Feedback from the secondary of the transformer is incorporated in a manner similar to that used before, but additional overall feedback has been incorporated from the 4.63 ohm tap to the first cathode in the preamplifier. It was found that a complex Bridged T network produced the best high frequency square wave response but that the square wave response was perfectly adequate when a simple 15 $\mu\mu$f capacitor was substituted in the overall feedback circuit. This capacitor has no effect on the low frequency response but reduces the tendency of the amplifier to ring slightly with sharp rise time square wave inputs. The photographs of Fig. 9 show the manner

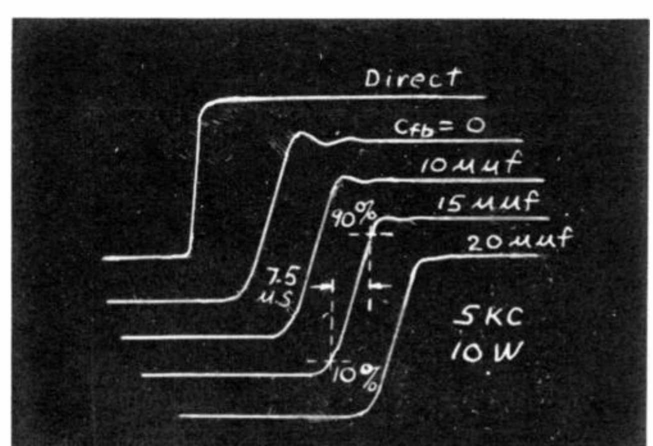

Fig. 9 – Leading edge analysis of 5 KC 10 watt square wave.

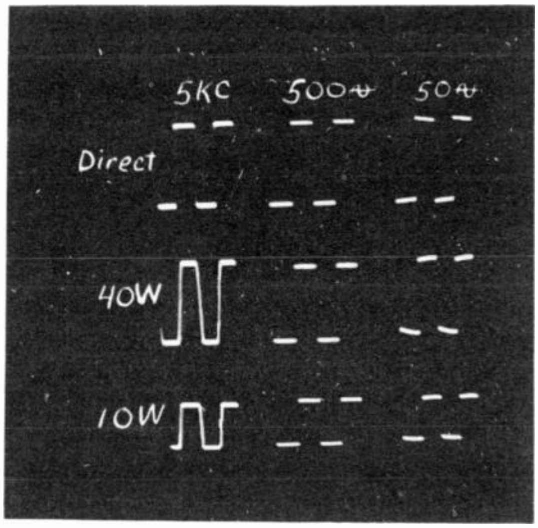

Fig. 10A – Amplifier square wave response.

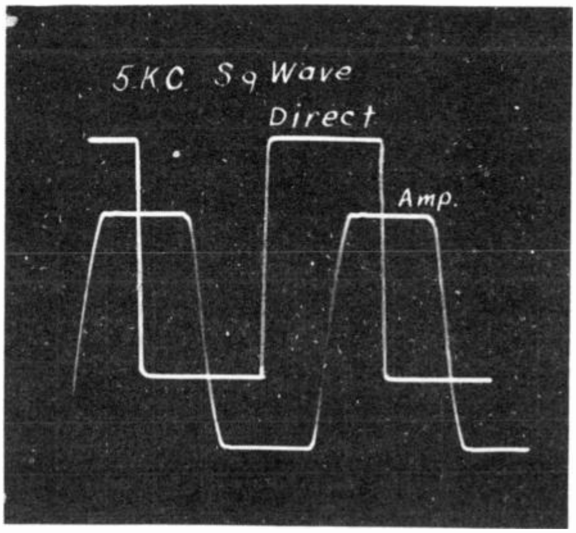

Fig. 10B – Enlarged view of amplifier response to a 40 watt 5 kc square wave.

in which the ringing, following the leading edge of a 10 w - 5 kc square wave, is modified by varying the value of this capacitor. The rise time of the leading edge is approximately 7.5 μs between the 10% and 90% points and the ringing frequency is approximately 100 kc. The complete square wave response at 50, 500, and 5000 cs. and power levels of 10 and 40 w is shown on Fig. 10A together with the output of the square wave generator connected directly to the CRO. A greatly expanded view of the 40 w - 5000 cs case is shown on Fig. 10B. The gain control settings were not changed during these tests.

The results of tests made to determine the best balance between the various types of feedback are shown in Fig. 11. In this figure curves 1 and 2 have inadequate feedback turns to correct for Class B operation at low power levels. The Bridged T overall feedback network produced 6 db feedback at operating frequencies so that curve 1 is lower than curve 2 at high power levels. Curve 3 has much less low level distortion than curve 2 because of the additional feedback turns but it requires more drive from the preamplifier and therefore has higher distortion than curve 2 at high power levels. Curve 4 uses approximately 6 db additional feedback in the preamplifier and its distortion is quite satisfactory at both low and high power levels. The 15 $\mu\mu$f overall feedback has no effect at these frequencies. The values of feedback turns and preamplifier feedback resistance used here represent a practical compromise between low input signal and low output distortion. Additional reduction in distortion could be obtained by increasing the number of turns on the feedback winding and reducing the value of the feedback resistor in the preamplifier. Both of these changes would increase the input voltage required to produce full output power. The conditions specified for curve 4 are the ones shown in the circuit diagram of Fig. 8 and are the ones used in all succeeding tests. Fig. 12 shows the curves of plate circuit losses (in-

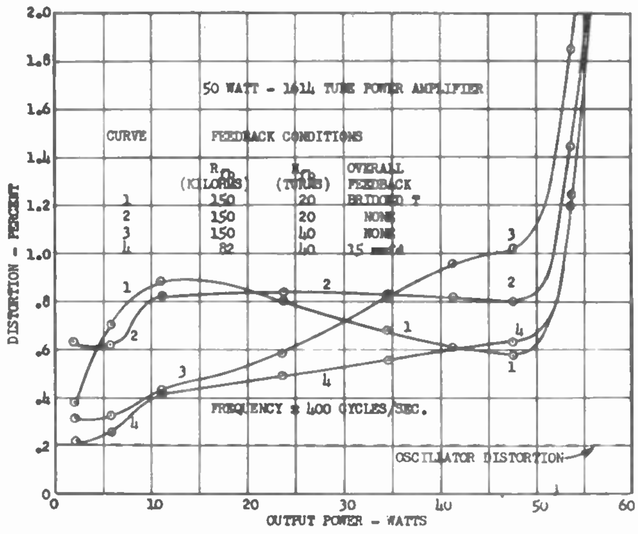

Fig. 11 – Feedback – distortion characteristics.

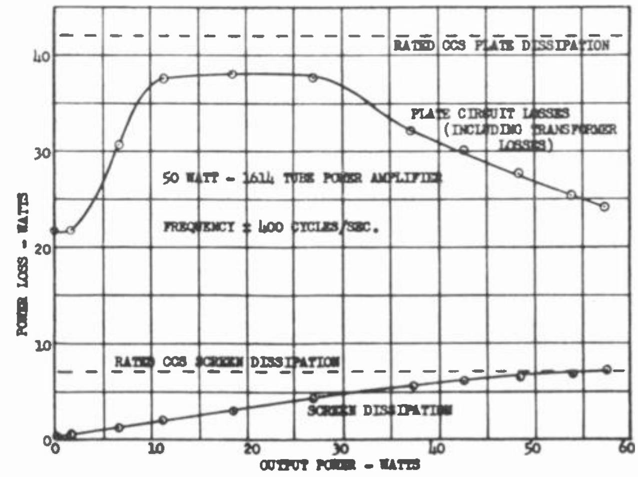

Fig. 12 – Power loss characteristics.

cluding transformer losses) and screen dissipation as the output power is varied. The plate circuit losses are less than the rated CCS values for all operating conditions. The screen dissipation becomes equal to the rated CCS value at the highest power levels shown but is less than the rated value at lower levels.

It can be seen from Fig. 11 that once the amplifier starts over-loading, the distortion increases very rapidly. For comparison purposes a 2% distortion level was considered a satisfactory standard value for determining the power delivering capacity of the amplifier. 1% distortion would have been adequate at 400 cs, where the oscillator distortion was only 0.2%, but not at 20 and 20,000 cs where the oscillator distortion was 0.8% and 0.6% respectively. Fig. 13 shows typical 2% distortion wave shapes at 20, 500, and 10,000 cs, for this amplifier.

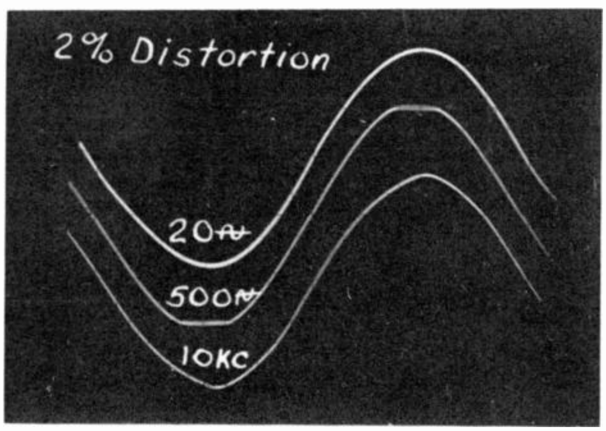

Fig. 13 – Two per cent distortion wave shapes.

The 2% distortion power handling capacity together with the corresponding plate circuit losses and screen dissipation are shown in Fig. 14. The highest plate

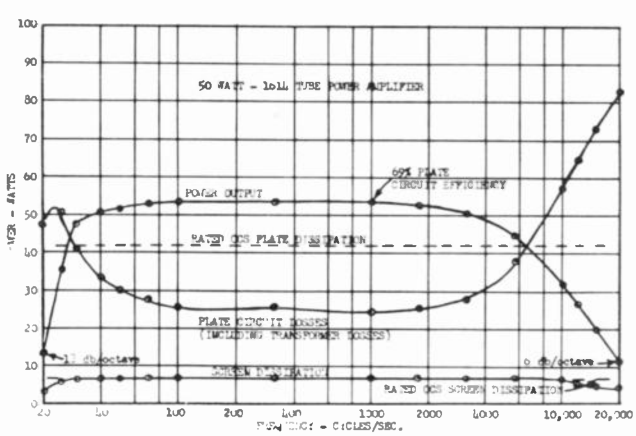

Fig. 14 – Two per cent distortion – power relations.

circuit efficiency, including transformer losses, occurred at 1000 cs and was 69%. The plate circuit losses are below the rated CCS value for the tubes alone and the screen dissipation is equal to the rated CCS value over most of the range. It should be emphasized that this curve represents the power delivering capacity of the amplifier and not the linearity of response with frequency variations. The frequency response characteristics, together with the power delivering capacity curve, plotted to a db scale, are shown in Fig. 15. It can be seen from

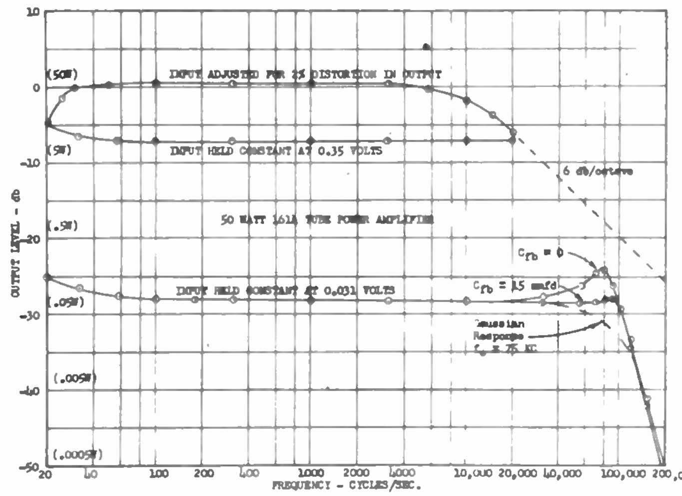

Fig. 15 – Frequency response characteristics.

this diagram that as long as the operating level is below the 2% distortion curve the response is perfectly flat between 100 and 20,000 cs. Below 100 cs the response rises slightly due to the series resonance in the impedance coupled circuit. The amount of this rise can be controlled by modifying the values of the coupling capacitor and choke. At low levels the response above 20,000 cs depends on the amount of feedback capacitance used. The curve for $C_{fb} = 0$ is seen to rise to a maximum at about 85 kc, and then to drop off very rapidly. The curve for $C_{fb} = 15$ micromicrofarads is seen to be almost perfectly flat to 95 kc, after which it also drops off very rapidly. A curve for Gaussian response with a -3 db point at 75 kc has also been shown for comparison purposes. Gaussian response would produce minimum rise time consistent with zero overshoot. It is seen that the ringing frequency of approximately 100 kc corresponds closely to the region of maximum deviation of the actual response characteristic from the Gaussian response. This also shows why the use of $C_{fb} = 15$ micromicrofarads reduced the ringing amplitude obtained with a square wave input signal.

Fig. 16 shows the results of an intermodulation distortion test using a 4:1 combination of 60 and 1500 cs. As is customary, the resulting distortion is plotted as a per cent of the smaller of the two signals. The distortion values shown here should be divided by 5 if they are to be compared with the harmonic distortion values discussed previously. The values shown are acceptable up to at least 112% peak-to-peak equivalent input. An intermodulation distortion test was also performed for a 4:1 combination of 60 and 15,000 cs. and the results are shown in Fig. 17. The purpose of this test was to find out if the high frequency roll off of the power handling capacity had any appreciable effect on the intermodulation distortion. The values obtained in this test were slightly higher than those shown in Fig. 22 but the performance is acceptable up to at least 105% peak-to-peak equivalent input.

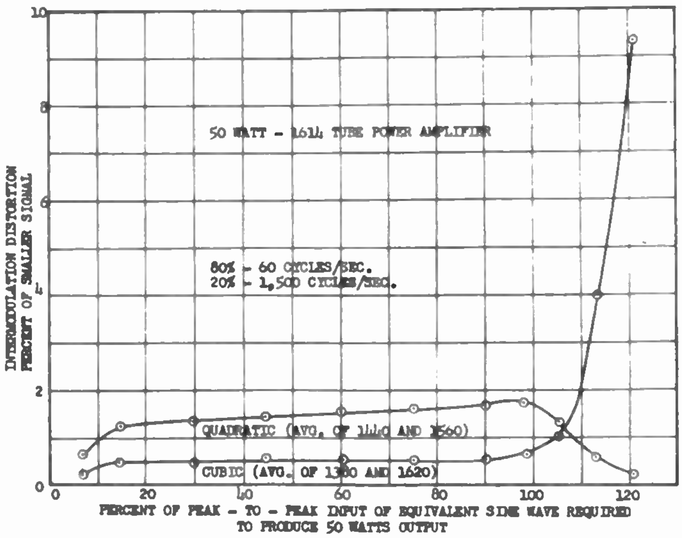

Fig. 16 – Intermodulation distortion characteristics.

Good transient response of a loudspeaker requires a low output impedance source. The manner in which the output impedance of this amplifier varies with frequency is shown in Fig. 18. The output impedance remains relatively constant with frequency and is about 10% of the nominal impedance of the tap. This output impedance was relatively independent of the current used to make the test.

A short circuited secondary test was performed on the amplifier to determine its susceptibility to accidental damage. The results of this test are shown in Fig. 19. The screen dissipation of the tube reaches 4.5 w at relatively low signal levels but remains below 5 w even with 120% of normal full load signal. With the same signal the plate circuit input approaches 120 w. This is considerably greater than the rated value but the plate structure is quite rugged and capable of handling these powers for short periods of time. The screen grid of the tube is quite fragile compared to the plate structure so it is fortunate in this situation that it only has to handle about 70% of its rated dissipation. This amplifier was operated for 10 minutes at the highest signal condition shown on the graph. At the end of the ten minute period the short circuit was removed and the amplifier operated normally. No damage could be detected in any part of the amplifier or its tube complement.

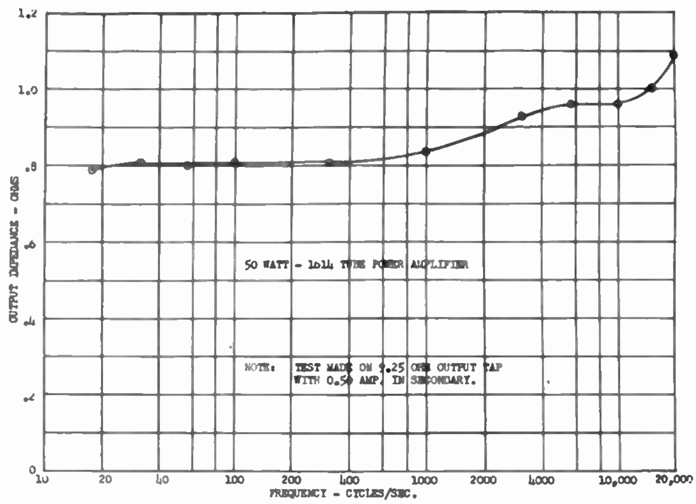

Fig. 18 – Output impedance characteristic.

The amplifier can also operate with full signal applied under open secondary conditions. The only noticeable change is that the output voltage rises by about ten per cent when the load is disconnected. This

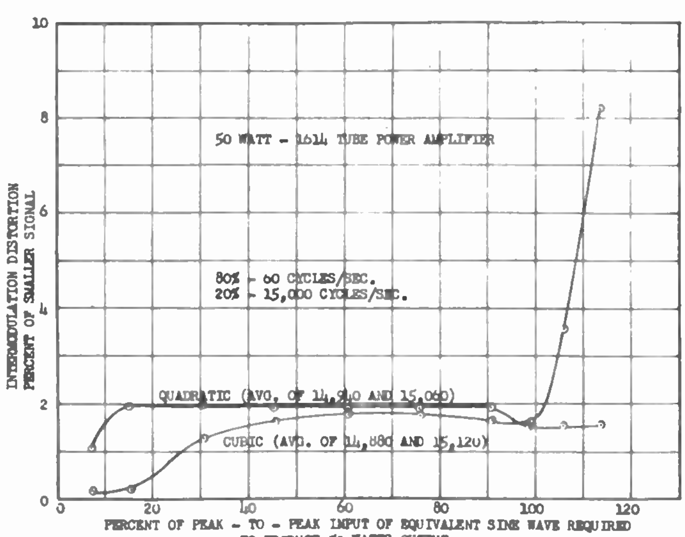

Fig. 17 – Intermodulation distortion characteristics.

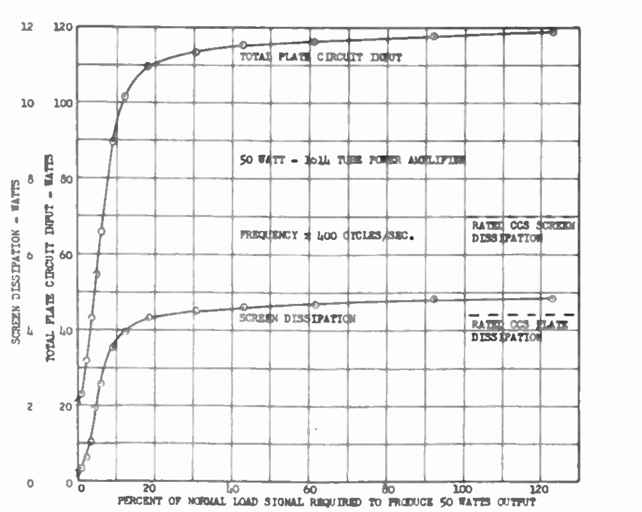

Fig. 19 – Short circuited output – power relations.

control is due to the close coupling used between the secondary and the feedback winding.

The residual hum in the complete amplifier is slightly greater than that measured in the basic amplifier with transformer coupled input. For the complete amplifier the residual 120 cycle hum is about 80 db below 50 w. The residual 60 cycle hum is about 66 db below 50 w. Due to its lower frequency, the 60 cycle hum is no more serious than the 120 cycle hum and both have been found to be completely negligible in most cases. The 60 cycle hum is picked up inductively by the unshielded modified T20C51 choke from the power transformer. If further reduction of the 60 cycle hum is desirable and the amplifier is operated remotely from other equipment, this choke can be oriented for minimum 60 cycle pickup. Approximately 20 db reduction has been obtained by this means but a clumsy mounting position was required for the choke. If the equipment is operated in proximity to other equipment whose relative position may change from time to time, a better solution would be to use a fully shielded choke. It should be emphasized, however, that in most cases further hum reduction is not necessary.

Fig. 22 is a photograph of a developmental model of the Bereskin 50 watt - 1614 Tube Power Amplifier. The output transformer is shown mounted on the chassis in its normal position but not potted.

APPENDIX

On the basis of the characteristics shown in Fig. 15, the 2% distortion power-delivering capacity of this amplifier is down approximately 6 db at 20,000 cycles while the low level response is flat to approximately 100,000 cycles. Theoretical and experimental investigations on both a qualitative and quantitative basis were made to determine the adequacy of this power-delivering capacity. It is the purpose of the appendix to discuss these investigations.

Sivian, Dunn, and White[6] and others have shown that most of the power in speech, song, and music is contained in the fundamental tones with frequencies below 3000 cs. The power levels of the higher frequency fundamental tones and of the harmonics of the lower frequency fundamentals decrease rapidly as the frequencies become greater than 3000 cs. Curve D in Fig. 20 is an adaptation of data presented by Sivian, Dunn, and White with regards to a 75 piece orchestra playing 4 different types of musical selections. In the original data the audio frequencies were divided into suitable but not equal bands, and for each of these bands the instantaneous peak power was recorded. This was done for each of the four musical selections and for many other sources of sound. The adaptation of this data consisted in taking the peak instantaneous power required in each of the four musical selections and calling this value 0 db. All four musical selections produced 0 db in the 250-500 cs band. The value plotted in any of the other bands was chosen from the musical selection in which this band had the minimum deviation from its own 0 db level. This curve is believed to be a close approximation to the most severe requirements for an audio frequency power system and has therefore been called the "MAXIMUM SEVERITY" composite. Curve E has been obtained by converting the

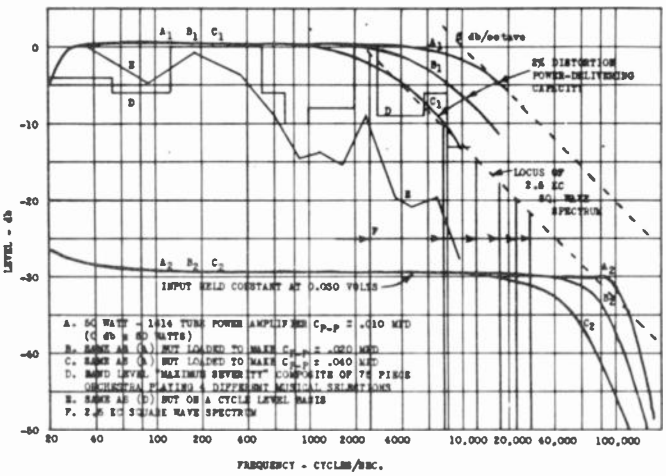

Fig. 20.

data represented by curve D to a cycle level. Sound is in general an integrated composite of many frequencies and these two curves show the relative contributions to be expected, under very severe conditions, from various frequency bands and ranges.

The 50 Watt – 1614 Tube Power Amplifier lends itself readily to the control of its power-delivering capacity without appreciably affecting its low level frequency response or its middle frequency distortion characteristic. Curve A_1 in Fig. 20 is the 2% distortion power-delivering capacity of the amplifier. Curves B_1 and C_1 are experimental 2% distortion power-delivering capacity curves for this same amplifier when it was loaded with .005 microfarad and .015 microfarad capacitors respectively between each of the 1614 tube plates and B+. For curve A the total primary interwinding capacitance is .010 microfarads, for Curve B it is .020 microfarads, and for Curve C it is .040 microfarads. Curves A_2, B_2, and C_2 are the corresponding low level frequency response characteristics obtained with a constant input voltage. The 400 cs harmonic distortion for all three cases was identically the same as that of Curve 4 in Fig. 11. The loading capacitors were connected with a rotary switch which made it possible to switch from any one condition to any other condition without going through the remaining condition.

Several full orchestra passages were recorded from LP records to tape with a Magnecord Tape Recorder operating at 15 inches/second. The tapes were cut into

[6]Sivian, Dunn, and White, "Absolute amplitudes and spectra of certain musical instruments and orchestras," *Jour. Acous. Soc. Amer.*, vol. 11, no. 3, pp. 330-371; January, 1931.

four foot lengths which were spliced into rings so that they could be played over and over with a one second dead interval in between to permit switching of the amplifier capacitance load. The Magnecord output was used as an input to the power amplifier and was also connected to the horizontal deflection system of a cathode-ray oscilloscope. The vertical deflection system of the cro was connected to the power amplifier output and the various gain controls were adjusted for a 45° trace on the face of the cro. Distortionless operation was characterized by a straight diagonal trace with slight tendency towards an ellipse due to the higher audio frequencies. Middle frequency distortion was characterized by horizontal extensions at the tips of the diagonal lines. High frequency distortion was characterized by small loops, similar to musical half note marks, at the tips of the diagonal lines.

For each section of tape the input level to the power amplifier was adjusted so that the amplifier just failed to clip peaks at the highest level passage of this section when the condition of Curve A was used. Switching the amplifier to the conditions of Curves B and C produced correspondingly larger indications of high frequency distortion. Reductions in input level ranging between ½ and 1 db were necessary to eliminate the distortion for the condition of Curve B. Reductions in input level ranging between 2 and 3 db were necessary to

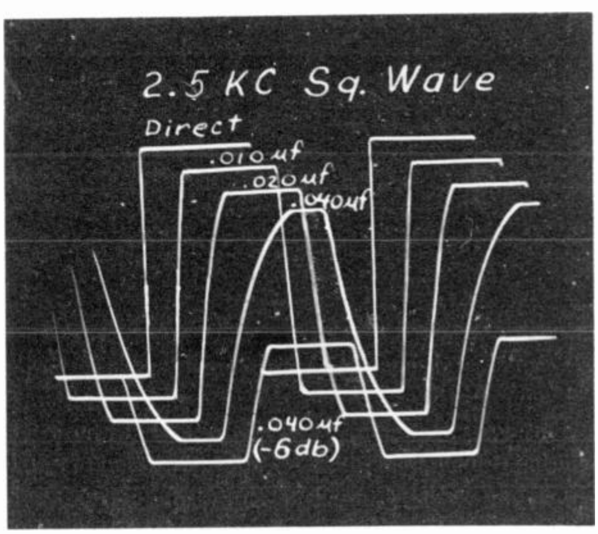

Fig. 21 – Effect of capacitance loading on square wave response of amplifier.

eliminate the distortion for the condition of Curve C. In all cases the distortion shown by the condition of Curve A was of the middle frequency peak clipping variety. Listening tests made at full level for these conditions showed in general that a slight difference could be detected between the conditions of Curves A and C but there was no conclusive agreement as to which represented the better operating condition. In the region of 2500 cycles/second Curve B is seen to be approximately 1 db below Curve A and Curve C is seen to be approximately 3 db below Curve A. This same region is seen to correspond to humps in both Curves D and E. These humps are probably exaggerated by the manner in which the "Maximum Severity" data was adapted from the original data but there is ample evidence that the 2000 to 2800 cycle/second band makes an appreciable contribution to the power content of the composite signal. Beyond this band Curves A_1 and B_1 safely override the band level "Maximum Severity" composite curves. All three of the Curves A_1, B_1, and C_1 are seen to drop off at a lower rate than the cycle level "Maximum Severity" composite in the high frequency range.

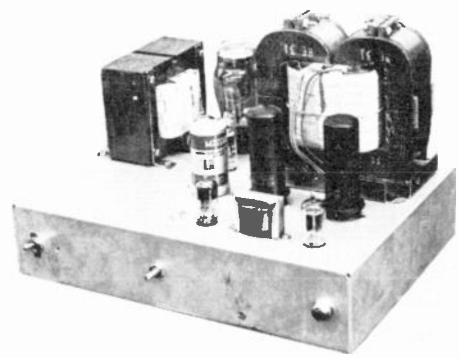

Fig. 22

This amplifier was used in listening tests with several combinations of full frequency range equipment and LP records containing full orchestra passages. In some of these tests the input level was adjusted to just fail to produce peak clipping during the most severe full orchestra passages with the condition of Curve A. In the remainder of the tests a high level was maintained but this level was below the peak clipping region. The transition between conditions A, B, and C was noiseless and in no case was the audience able to detect any difference between the three conditions.

The spectrum of a 2500 cs square wave, having a total power of 0 db, is shown up to the 11th harmonic by the arrows marked F in Fig. 20. The components above 27.5 kc have been left out to avoid confusion in the diagram but their effect can be understood by recognizing that only odd harmonics are present and that the spectrum level of a square wave drops off at a 6 db/octave rate. Several of the harmonics of this wave are in the audible frequency range and the drop-off rate of the harmonics does not differ too much from the band level and cycle level "Maximum Severity" composite characteristics. The behaviour of the amplifier conditions to a full level square wave input signal could be used for comparison purposes to interpret the high frequency power delivering capacity of other power amplifiers.

The oscillograms in Fig. 21 were obtained by the application of a 2.5 kc square wave to the amplifiers corresponding to Curves A, B, and C in Fig. 20. The input signal used in the upper 4 waves was equal to that required to produce 50 w output at 250 cs. The power present in the output was slightly less than 50 w because of the sloping sides of the output waves. The input signal used for the lowest wave was 6 db below that used for the other curves. The top curve is used for reference and represents the application of the output of the square wave generator directly to the cro with the same gain control settings that were used in all of the other tests.

The rise times between the 10% and 90% points are tabulated below:

C_{P-P} (microfarads)	Signal Level	Rise Time (10% to 90%) (microseconds)	(% of period)
.010	0 db	22	5.5
.020	0 db	44	11.0
.040	0 db	92	23.0
.040	-6 db	36	9.0
(direct from sq. wv. gen. to CRO)	0 db	2.5	.63

The wave for condition (C_{P-P} = .040 mfd) shows a relatively large amount of rise time for full signal operation but this rise time is reduced to less than half its original value when the signal level is reduced by 6 db. All of the waves are characterized by an extremely small amount of ringing which is an indication of good low level frequency response.

It appears from the tests discussed previously that the condition of Curves B (C_{P-P} = .020 mfd) represents a transition range where it is possible to detect incipient loss of high frequency program material with instruments but the loss involved is still insufficient to be detected from listening tests. Although further investigations will be made, it appears reasonable at this point to assume that an amplifier will have adequate high frequency power-delivering capacity if it can reproduce a 2.5 kc square wave at full signal level with a rise time of less than 40 microseconds between the 10% and 90% points of the wave. To avoid the possibility of audible intermodulation components due to the combination of ringing and high audio frequencies, the ringing amplitude should be relatively small and the ringing frequency should be relatively high.

It has been shown that the amplifier discussed in this paper has adequate high frequency power-delivering capacity for all normally encountered audio frequency signals. The middle and low frequency power-delivering capacity and the transient response are all excellent. The harmonic and intermodulation distortion are uniformly low and could be further reduced at the expense of requiring additional input signal. All of these performance features have been achieved with a structure which is compact and relatively inexpensive and capable of withstanding unusual abnormal conditions, such as full signal input with short circuited load, without damage to the amplifier itself.

THE CASCODE AS A LOW NOISE AUDIO AMPLIFIER*

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The improvement of signal-to-noise ratio has always been a major problem to engineers engaged in the development and design of high quality audio equipment. Faced with the limitations of the maximum signal level obtainable from transducers together with the minimum accompanying amplifier noise level, improvements in signal-to-noise ratios over the years have been made slowly by improvements in transducers and amplifiers. For example, a microphone with a maximum output signal level of -55 dbm obviously cannot achieve a signal-to-noise ratio greater than 60 db with an amplifier whose equivalent input noise level is -115 dbm.

Noise level is a particularly serious problem in recording. The development of modern professional quality tape recording has greatly increased the signal-to-noise ratio which is generally available in recorded form.

Even here noise is still a problem, for even when it is 50 to 60 db below the signal, a careful listener can detect it. The output from multi-channel and narrow track heads is restricted by the available track width and the resulting signal to noise ratio is then usually limited by the amplifier input stage noise level. Most of the recent work on the improvement of signal-to-noise ratio has been in the direction of increased signal output by

*Manuscript received February 9, 1954; See also H. J. Woll and F. L. Putzrath, "A note on noise in audio amplifiers" in this issue. These two papers are being published simultaneously so that readers may have the benefit of both points of view. — Editorial Committee.

the development of higher output tapes and heads. Some work also has been done on the decrease of noise level by the development of tapes with better uniformity, bias oscillators of improved waveform, and magnetic pickup heads with reduced microphonism or magnetostrictive noise. The improvement which can be obtained by these methods, however, is limited by the tube noise of the input stages of high gain amplifiers. Most of the recent work in reducing the equivalent input noise level of audio amplifiers has been confined to refinements in the design of existing circuits and components. Some degree of improvement in input stage noise is sometimes obtained by the use of selected tubes, the selection being made either by the tube manufacturer who gives the selected tubes a special type number, or by the user on a trial and error basis. It is obvious, however, that tube selection is not a completely satisfactory answer to the low noise amplifier problem.

In the present development, a completely different approach was used. Techniques from outside the audio field were reduced to their basic form and then modified to give the desired results at audio frequencies. Work done during the war by Wallman and others resulted in the development of the grounded cathode triode grounded grid triode, or "Cascode" amplifier circuit. Wallman has shown[1,2] that this amplifier gives the lowest noise figure obtainable at the present state of the art. It has been used with considerable success in the improvement of the noise figure of high frequency circuits used in radar, television, and many other RF applications. More recently, it has been used to reduce the noise level and to increase the dynamic range of video input amplifiers[3,4] for television cameras.

The first use for the grounded cathode triode in series with a grounded grid triode was in a dc amplifier in a voltage stabilizer.[4,5] The low noise feature of the present circuit is, however, entirely unconnected with this original use. An investigation of the basic form of the "cascode" circuit given by Wallman[5] and shown in Fig. 1 shows that it is not the exclusive property of the high frequency worker. In fact, the elimination of tuning and neutralizing circuits which are unnecessary in audio applications, results in considerable simplification over the forms usually seen.

As shown in Fig. 1, the basic "cascode" circuit consists of a grounded cathode triode in series with a grounded grid triode. Additional coupling elements are not needed between the two triodes and the same dc flows through both tubes. This "cascode" connection combines the advantages of the grounded cathode triode with those of the pentode while minimizing their disadvantages.

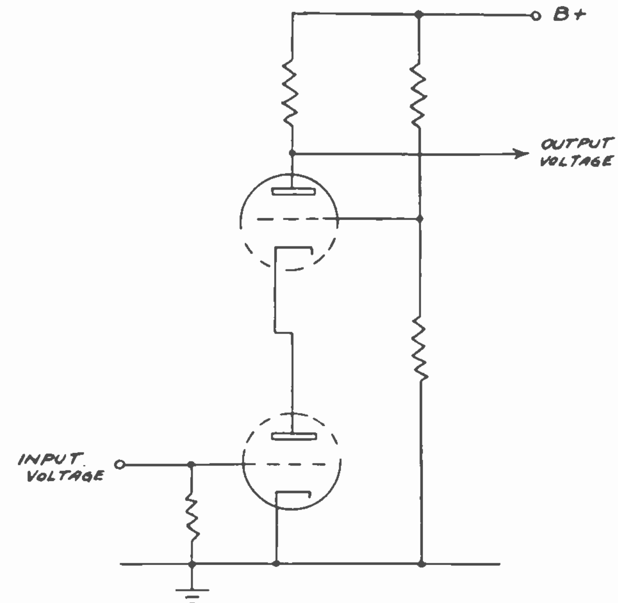

Fig. 1 – Basic form of cascode circuit.

THEORY OF OPERATION

The operation of the audio frequency "cascode" amplifier can perhaps best be undersood by considering briefly the relative advantages and disadvantages of grounded cathode pentodes, grounded cathode and grounded grid triodes.

The grounded cathode pentode amplifier has the advantages of high gain, and low input capacity due to the isolation of the plate circuit from the control grid by the screen grid. For these reasons, pentodes are often used in audio input stages in spite of their disadvantages of a noise level which is three to five times higher than that of a triode. The grounded-cathode triode has moderate available power gain with an output conductance of $1/r_p$ and a low equivalent input noise level.

The grounded grid triode has a large input conductance and a rather low available power gain.

The ideal triode input circuit should have all the improvement in input stage noise figure over the pentode that is theoretically possible, with the contribution of second and later stages to the noise figure no greater than with pentode input stages of the same bandwidth. Therefore, the stage gain should be of the same order of magnitude as that of a pentode. In addition, the circuit should be stable, with low input capacity and effective isolation of the input circuit from the output circuit.

Out of the nine combinations possible with two triode input tubes, using various combinations of grounded cathode, grounded grid, and grounded plate, the only

[1]Valley and Wallman, "Vacuum Tube Amplifiers," McGraw-Hill Book Company, New York, pp. 656-660; 1948.
[2]Wallman, Macnee, and Gadsden, "Low noise amplifier," *Proc. I.R.E.*, vol. 36, no. 6, pp. 700-708; June, 1948.
[3]K. B. Benson, "Modified preamplifier improves movie telecasts," *Electronics*, vol. 26, no. 12; pp. 166-169; December, 1953.
[4]"Application note on Type 6198 Vidicon television camera tube," published by the Tube Department, RCA.
[5]Valley and Wallman, ibid., p. 440.

circuit which meets all of the above requirements is the grounded cathode triode in series with a grounded grid triode or "cascode" circuit.

With this circuit arrangement, the upper tube has a fixed dc grid potential at ac ground which tends to hold the lower triode plate potential fixed, but still permits its current to flow in a load resistor. If e_{p_1} were really held constant, the current gain of the lower triode would be g_m and the voltage gain from e_{g_1} to e_{p_2} would be $-g_m R_L$ as in a pentode. Thus, the behavior is similar to a pentode, with the advantage that no screen current with its accompanying partition noise is present. The grid of the upper tube, being bypassed to ground, has an effect on stage behavior similiar to that of a screen grid. The output conductance of the first tube is of the same order of magnitude as the optimum source conductance for the second tube, so that the full available power gain of the grounded cathode triode is utilized. The cushioning effects of the two space charges are in series, no physical coupling resistances are needed, and effective isolation of input and output circuits and low input capacity also result. The reduction of capacitive loading of the input grid circuit due to the "Miller Effect" which is present in conventional triodes makes possible higher interstage coupling circuit gain.

It has been pointed out[3,6] that the use of the series connected cascode circuit results in considerable reduction of cross modulation distortion as compared with pentode circuits.

Fig. 2 shows the equivalent circuit of the "cascode" amplifier using two identical triodes. The gain of this amplifier may be derived as follows. Around the closed loop of Fig. 2 we obtain:

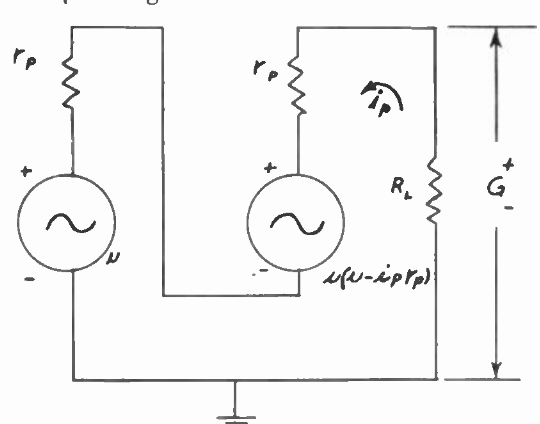

Fig. 2 – Equivalent circuit of cascode amplifier.

$$\mu + \mu\,(\mu - i_p r_p) = (2r_p + R_L)$$

$$\text{Gain} = i_p R_L = \frac{\mu\,(1 + \mu)}{1 + \dfrac{r_p}{R_L}\,(2 + \mu)}$$

[6]R. M. Cohen, "Use of new low noise twin triode in television tuners," *RCA Rev.*, vol. 12, no. 1; March, 1951.

When analyzing this circuit, it should be remembered that since the same current flows in both triodes, the two tubes are not independent of each other but must be considered as a single stage. It has been pointed out[7] that tube noise is best represented by a constant current generator between plate and cathode.

As an experimental check on the action of the grounded grid triode in increasing the gain and decreasing the noise of the lower triode, the gain and noise figure of the arrangement of Fig. 3 was compared with the circuit of Fig. 4, all other circuit conditions being

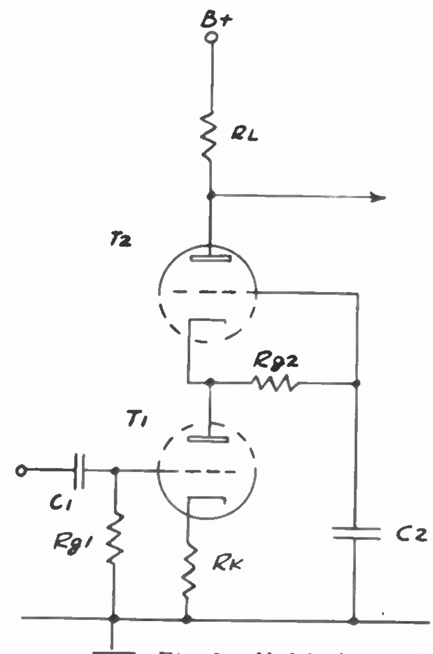

Fig. 3 – Modified cascode amplifier.

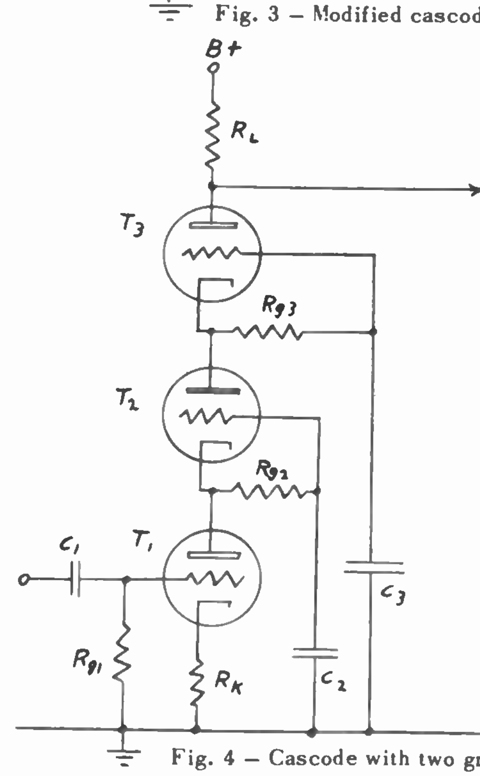

Fig. 4 – Cascode with two grounded grids.

[7]Valley and Wallman, Ibid., pp. 562-563.

equal. The addition of the third grounded grid triode in series with the other two resulted in a substantial further reduction of noise and increase in stage gain, thus demonstrating the effectiveness of the grounded grid tube.

PRACTICAL APPLICATIONS

In practical applications of the basic circuit of Fig. 1, it is soon found that due to variations in tube characteristics, the bias on the grounded grid triode must be adjustable to maintain it at a slightly negative potential with respect to its cathode. A modification which overcomes this critical adjustment is shown in Fig. 3. This modification consists of the use of grid leak or contact potential bias on the upper triode to replace the voltage divider of Fig. 1. The value of grid to cathode resistor is chosen so as to maintain the grid-cathode bias potential at approximately one volt negative. AC ground potential is maintained on g_2 by condenser C_2. With this biasing method, tube variations have very little effect on the biasing of either triode.

Figs. 5 and 6 show two possible methods for applying negative feedback to this amplifier stage. In Fig. 5, feedback is applied around one stage only. Fig. 6 shows

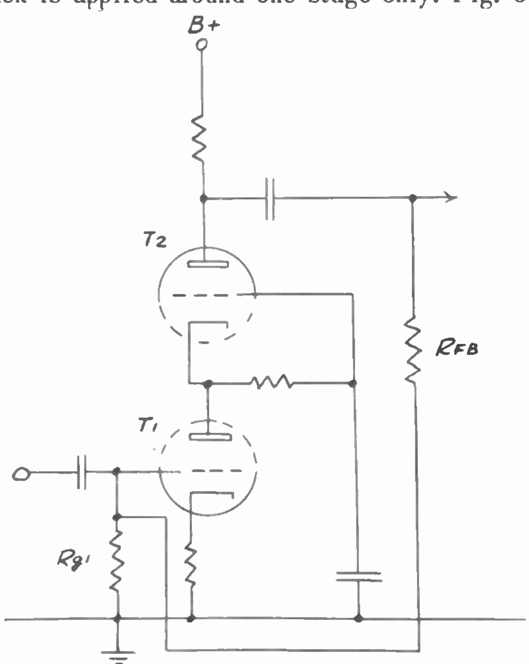

Fig. 5 – Cascode with feedback over one stage.

negative feedback applied over two stages. R_k also provides cathode bias and degeneration for the input stage. The use of negative feedback apparently has little effect on the noise figure of the amplifier. Feedback does, however, tend to reduce microphonic noise and increase the available dynamic range for a given distortion level.

It is important in the construction of a low noise amplifier of this type to use only the highest quality components and to exercise extreme care in parts location and layout so as to avoid undesired noise pickup due to leakage currents, stray magnetic and electric fields, etc. It has been the author's experience that it is advisable to use low noise resistors such as the deposited film type. Resistances composed of carbon granules usually generate noise considerably in excess of thermal agitation noise, due to fluctuation in contact resistance between adjacent granules. It is also helpful if the resistors can be kept in fairly stable thermal equilibrium by avoiding high ambient and operating temperatures.

As is the case with other types of low level amplifiers, it is advisable to operate the filaments from dc to minimize hum problems.

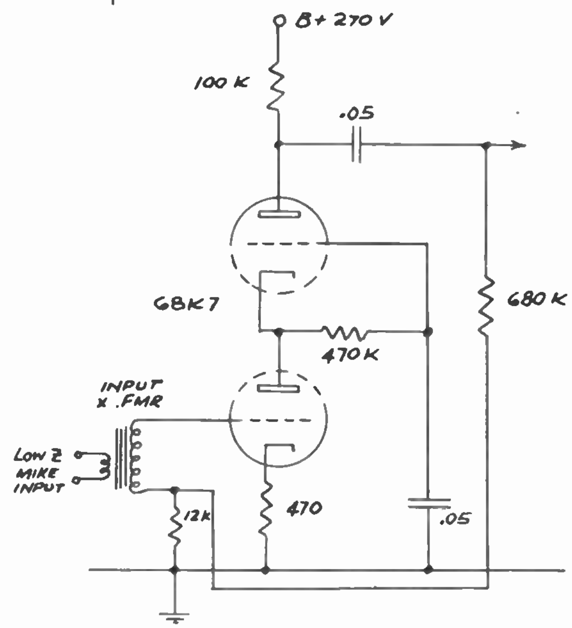

Fig. 6 – Cascode with feedback over two stages.

PERFORMANCE

In evaluating the performance of the cascode circuit at audio frequencies, comparative measurements were made between the circuit of Fig. 7 and several commercially available low noise amplifiers. Noise generator techniques were used as well as absolute level measurements and we were able to correlate results between the two methods quite closely. The noise figure measurements were made using a temperature limited diode noise generator.[8] With the "cascode" amplifier, noise figures as low as 1.5 were obtained which is 1.8 db from the thermal agitation noise in the input grid. The equivalent input noise level as obtained by dividing the output noise by the amplifier gain was found to be approximately -127 dbm over a 15 kc band which agrees quite well with the noise figure measurement. Input levels up to -20 dbm were possible without exceeding 1% total harmonic distortion.

These measurements showed an improvement in

[8] Terman and Pettit, "Electronic Measurements," 2nd Ed., McGraw-Hill Book Co., New York, pp. 372-373; 1952.

noise figure of from two to twelve db between the "cascode" and the conventional types of low noise amplifiers. The exact amount of improvement will depend on the

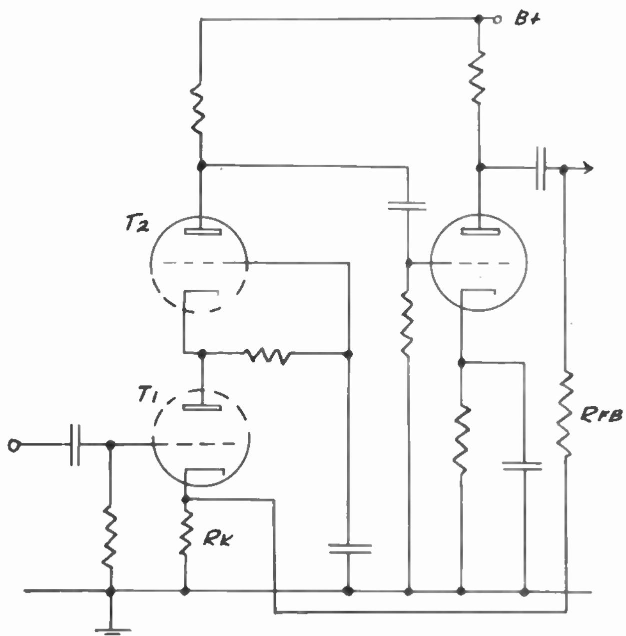

Fig. 7 – Microphone amplifier using cascode circuit.

type of tubes used and the frequency characteristic involved. Tube noise in the audio range consists of low frequency noise due to cathode "flicker," medium frequency noise due mainly to microphonics, and uniformly distributed noise due to "shot effect".

Pentodes have an additional noise known as "partition noise" which is due to the random fluctuation of the distribution or partition of cathode current between plate and screen and which generally overrides the other types of noise just mentioned. The final selection of tubes will depend to a large extent on which of these types of noise it is desired to minimize.

An additional practical consideration is the fact that with the "cascode" circuit, relatively large stage gains may be obtained with medium mu tubes which are apparently less critical with respect to microphonics than are the high mu triodes and pentodes ordinarily used in amplifier input stages. No increase in the number of circuit components and parts is required over a pentode stage and as a general rule, the double triodes which are suitable for cascode use are less expensive than are low noise pentodes.

In recent months, several new types of dual triodes specifically designed for cascode circuit applications have appeared. These types include 6BK7, 6BQ7, 6BZ7 and GL-6386. In addition to these types, good results have also been obtained using the more conventional types of dual triodes such as the 12AU7.

CONCLUSION

The work which has been done in adapting the "cascode" amplifier to audio frequency service has resulted in the successful development of a stable, low-noise, high-gain input stage, having low distortion, which requires only a minimum number of component parts and which does not require the use of expensive or especially selected tubes. An improvement in noise figure of from two to twelve db has been obtained with this circuit as compared with conventional input amplifier circuits.

The "cascode" audio amplifier is being used in the new Magnecord M-80 series of tape recorders with considerable success. Consistently lower noise has resulted than with the special low noise tubes used previously and it has not been necessary to resort to an undue amount of tube selection to maintain amplifier noise specifications in production.

The author wishes to acknowledge the valuable assistance and suggestions given by Professor A. H. Wing of Northwestern University and Mr. W. F. Boylan of Magnecord, Inc, which contributed greatly to the success of this development.

AN ALL-TRANSISTOR HEARING AID*

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The device to be described uses three *p-n-p* alloy-junction transistors. It is well known that, in the manufacture of this type of transistor, it is necessary to have pure germanium to 1 part in a billion. Producing purities of this order were unknown up to three years ago. In this state of purity, the crystal lattice structure has four valence electrons. Controlled amounts of "impurity" elements are than added to the germanium. If antimony having five valence electrons is added, the material becomes *n* type, i.e., the current flow takes place due to a movement of electrons. If gallium or indium having three valence electrons is added, the material becomes *p* type, and current flow is described as being due to a movement of positive holes.

In the *p-n-p* junction type transistor the *n* material forms the base connection and the layers of *p* material on each side provide the terminals called emitter and collector. As with vacuum tubes, there are three basic methods of operation, namely; – grounded base, grounded collector, and grounded emitter. For hearing aid application, the grounded emitter is more useful than either of the others, because greater gain is obtained, and only one battery supply is necessary.

The microphone used is a balanced magnetic armature type. It has a sensitivity of approximately 10^{-10} watts at 1000 cps when excited by a sound field of 1 microbar (74 db re.0002 dyne/cm^2). This type of microphone has an impedance of about 1000 ohms, which is an excellent match for the transistor input circuit used. Because the microphone is magnetic it is much more stable with respect to temperature and humidity than the Rochelle salt microphones which were used previously with high-impedance input circuits of the vacuum-tube type. The microphone is capacitively coupled to the base of the transistor with a 4 microfarad electrolytic condenser.

Much progress has been made recently in the design of very small and very high quality electrolytic condensers for transistor circuit applications. The particular condenser used is a tantalum type hermetically sealed in a metal can. These condensers are practically impervious to effects of temperature and humidity, at least over the range anticipated in hearing aid use.

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Many claims and counter-claims have been made regarding transistors and their ability to perform properly and uniformly. In some cases charges have been made against the transistor which have resulted from limited understanding of circuit application information. As with vacuum tubes there are many different types with different characteristics. If the circuit engineer becomes familiar enough with transistors to develop circuitry which takes full advantage of the transistor characteristics, the transistor can be controlled as carefully as vacuum tubes. It may be temperature-stabilized so that, over a range of temperature from 40° to 120°F, the change in current will be very small. Thus, over a wide range of temperatures, a stable control of noise characteristics can be achieved, plus a stable control of battery operating current and battery cost. This stabilization can also provide a control of less than 1 db change in gain over the temperature range of 40° to 120°F.

The transistor in the first stage of the hearing aid is temperature-stabilized by means of a voltage-divider network, which maintains a rather constant voltage at the base of the transistor. Further stabilization is accomplished by proper choice of the emitter resistance. The value of the resistors used in the voltage-divider network are dependent upon the characteristics of the transistor, and upon the supply voltage used.

It is necessary, for example, to make the current through the bleeder resistors large as compared with the normal base current of the transistor. There is also a critical ratio for the center point of the divider connected to the base. This ratio changes with different transistor characteristics, and also with different supply voltages.

This particular circuit was designed for Raytheon's CK.718 PNP type junction transistor. The emitter resistor value and the voltage divider ratio and value may be predetermined in design for a given supply voltage, so that it is not necessary to select transistors for the circuit. The collector circuit is transformer coupled to the next stage. The reflected primary impedance is 20,000 ohms while the secondary impedance is 1000 ohms. Particular care and attention must be given to the transformer design, so that the dc primary resistance will be suitable for the type of transistor used and for the supply voltage, in order to give the proper dc voltage at the collector.

The transformers used in the Beltone Concerto instruments are magnetically shielded in mu-metal cans in order to eliminate electrical hum pickup caused by fluorescent lights, power lines, motors and such devices. The mu-metal cans serve also as container into which the transformer is potted, so that the effects of moisture and mechanical shock are practically eliminated. The mu-metal cans even improve the efficiency and frequency response.

The problem of tone control design is quite different in a transistor circuit than in vacuum tube hearing-aid design. Between the secondary of the first transformer and the second transistor a series resistor, shunted by a capacitor, has been added to provide the tone control action. Operation with the tone-control switch in the shorted position gives the "flat" response or "full-tone" condition. When this switch is opened, the parallel resistor-capacitor network accentuates the passage of high-frequency tones and gives the effect of suppressing low tones. This noise suppressor is intended primarily to enable the user to adjust the frequency response of the instrument to his environmental needs. When the user is in a noisy location, such as a restaurant or office, or exposed to street traffic noise, he will probably prefer the number two, or suppressor, position. In this way many of the annoying background noises will be eliminated. On the other hand, when he is in a quiet environment such as in his own home he will prefer to switch to the number one position in order to get the full range of tonal quality available in the instrument.

The volume control is a 10,000 ohm potentiometer controlling the input signal to the transistor. Because impedance reflected from the output stage to the input stage has an influence on transistor circuits (that is not present with tube design), the value for the emitter resistor changes with different stages. The second stage is also temperature-stabilized by use of a proper voltage-divider network and emitter resistance. The output or collector of this second transistor is again transformer-coupled to the output stage. Control of the dc current in the power-output stage is provided by the selection of the proper base resistor for the particular supply voltage and transistor being used.

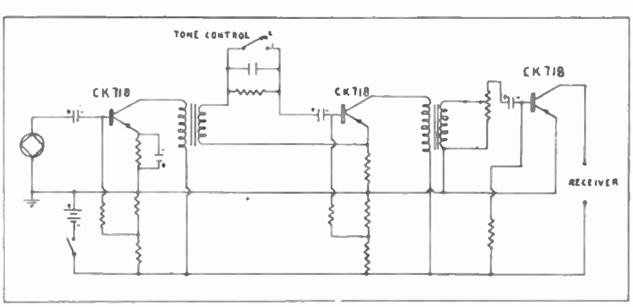

Fig. 1 – Triode tube noise vs. frequency.

The hearing aid "receiver" or earphone is magnetic. It is connected directly in series with the collector circuit of the output stage. This transducer has a nominal impedance of 1000 ohms. Again particular attention must be given to the design of the coil, so that the DC resistance of the receiver winding will be proper for the transistor and for the supply voltage used.

Great flexibility is provided in the basic circuit (See Fig. 1). More than one supply voltage may be used. When this is done, the model is given another number or color code. Reference to Fig. 2 will disclose the extremely wide range of maximum acoustic power levels available in the Concerto instrument.

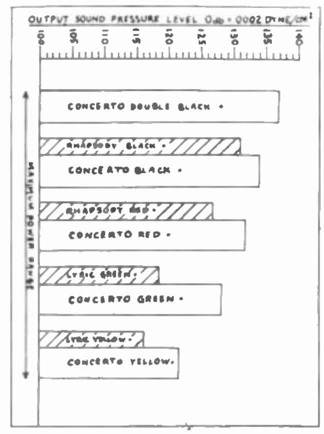

Fig. 2 – Grounded – grid equivalent circuit.

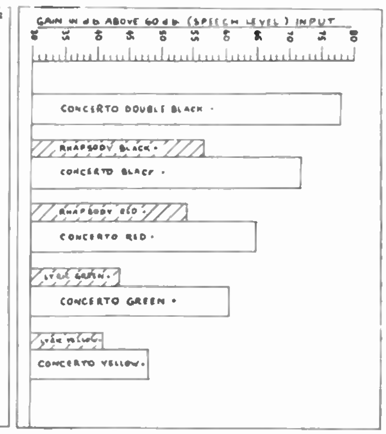

Fig. 3 – Grounded – cathode equivalent circuit.

It should be mentioned that in all our previous experience with vacuum tube hearing aids, it has been necessary to choose an instrument which has enough gain and power available, when the batteries are discharged to "end-point." The gain and power with a fresh battery was considerably more than that available when the battery had reached end-point. It has always been necessary, therefore, to choose a fitting which would have enough reserve gain and power to provide for a reasonable drop in battery voltage.

A conservative estimate of loss in acoustical power between a fully charged "B" battery, and one which has reached normal end-point of ⅔ voltage, is a loss in maximum power level of six decibels. At the same time this drop in battery voltage was responsible for a loss of at least 12 db, conservatively speaking, in gain. Fig. 2 shows the power available in the various color-code combinations of the Concerto model. Here the design has incorporated stair steps at acoustic power level starting at 122 db for the Yellow Dot, going to 128 db for the Green, 132 db for the Red, 134 db for the Black, and 137 db for the Double Black Dot. This may be compared in Fig. 2 with the power available with reasonable battery life in the Lyric-Rhapsody line of vacuum tube aids.

In Fig. 3 is shown the extremely wide range of fitting available in the various color dots. Here again, the design has purposely incorporated stair-step progressions of acoustical gains which are available and reference to Fig. 3 will show the comparable gain which was available with the Lyric-Rhapsody (Vacuum-tube) line, considering normal battery drop. The gain and power data are shown at 1000 cps for simplicity and for comparison to previous instruments for which similar data have been published. All of the gain and power data shown for the Concerto Model in Figs. 1 and 2 are based upon the use of a standard microphone and a T-3 receiver.

hearing aid to work at top efficiency with the maximum performance in gain, power, battery economy, and minimum distortion, it is necessary to design for a predetermined battery voltage. If it is desired to provide temperature stabilization, it is necessary to use a completely different stabilizing circuit at one voltage than for another. Moreover, the input and output impedances of the transistor change radically with a change in supply voltage. Completely different circuitry is required for optimum performance at different supply voltage. Resistance tolerances are held to plus or minus one-percent in the instrument. This is necessary for optimum transistor performance.

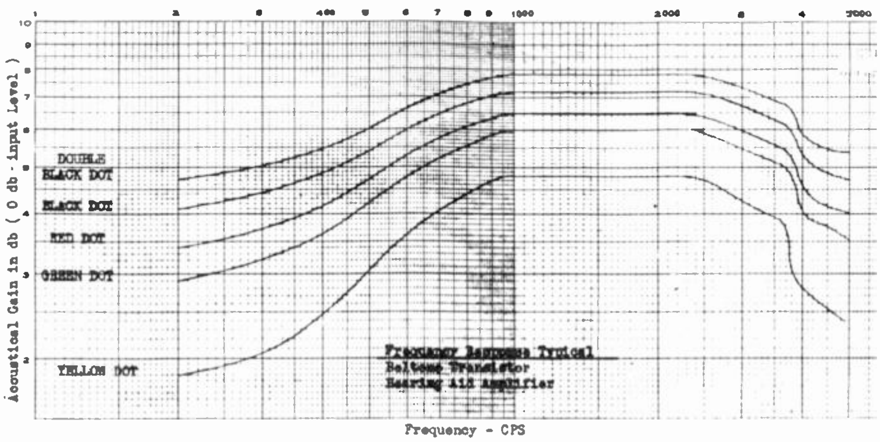

Fig. 4 – Noise vs. source resistance.

The yellow and green dot instruments have circuits which have been designed for 1.25 volt operation. This provides a low-gain, low-power instrument; and a medium-gain, medium-power instrument. The red, black, and double-black dot instruments are designed for a supply voltage of 2.5 volts. This provides a high-gain, high-power instrument; a very high gain, high power; and a special instrument for extreme (high loss, high tolerance) cases.

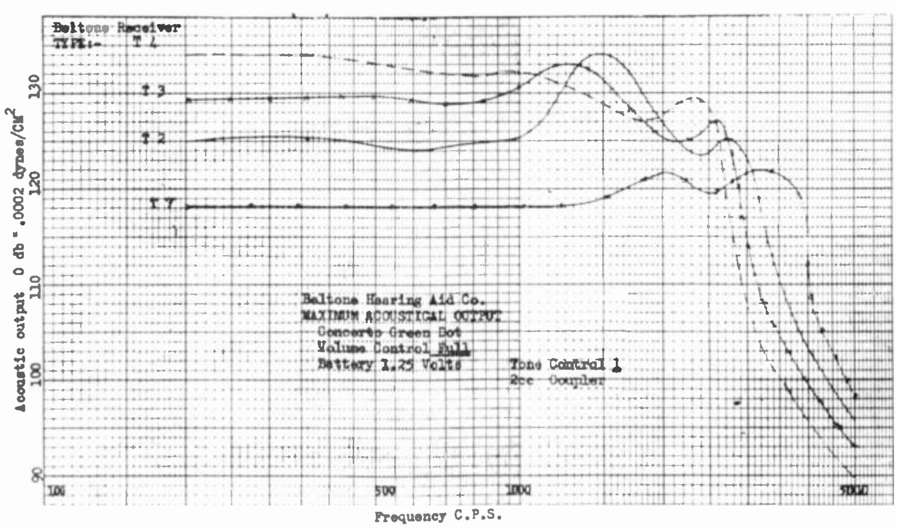

Fig. 5 – Typical two stage amplifier.

Additional flexibility is available in the maximum output, as well as the frequency-response, by selection of one of several receivers (earphones), each with different characteristics. Many factors influence the circuit design of a transistor hearing aid. In order for a transistor hearing aid to work at top efficiency with the maximum performance...

The transistor completely eliminates the necessity for a "B" battery. This ends the constant expense of buying "B" batteries and the nuisance of changing them frequently. Battery costs for the Concerto model, used by a person with average hearing loss, are as low as $2.00 for a full year. Because most users of vacuum-tube aids with average losses are accustomed to paying $40.00 to

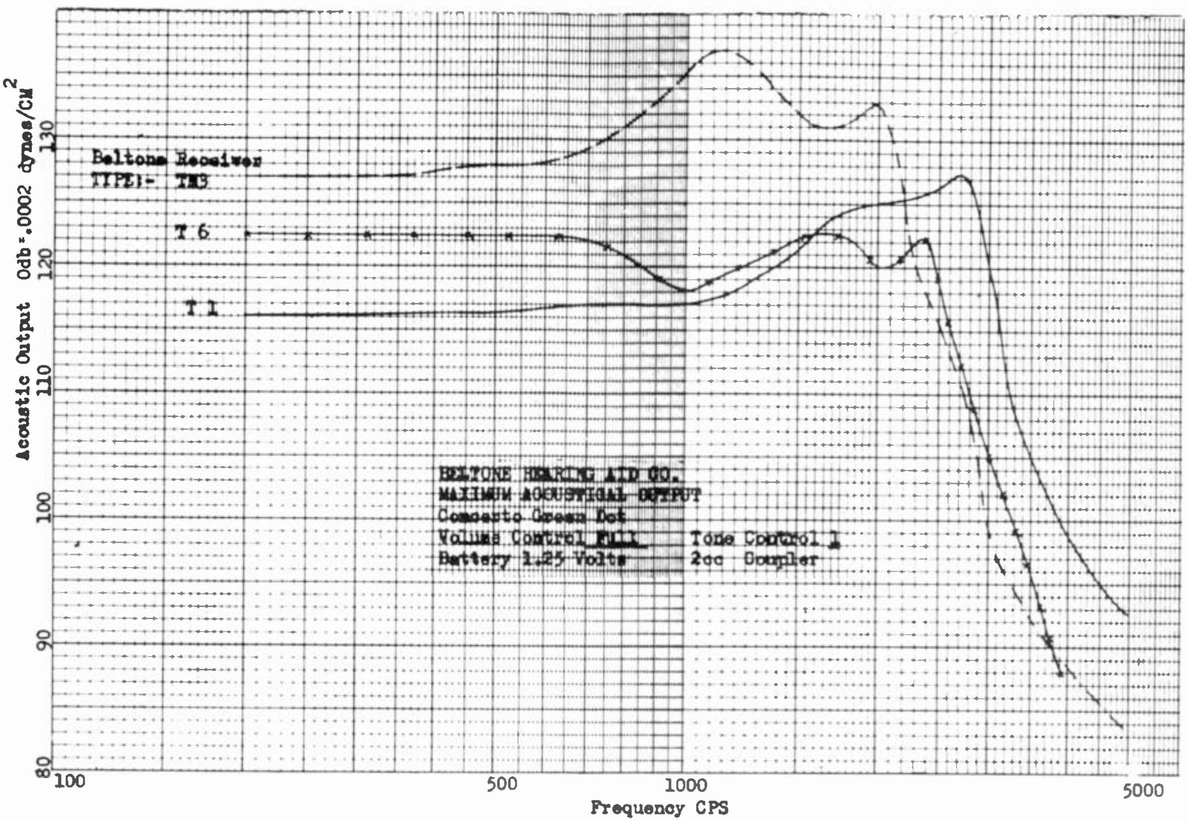

Fig. 6 — Equivalent circuit of two stage amplifier.

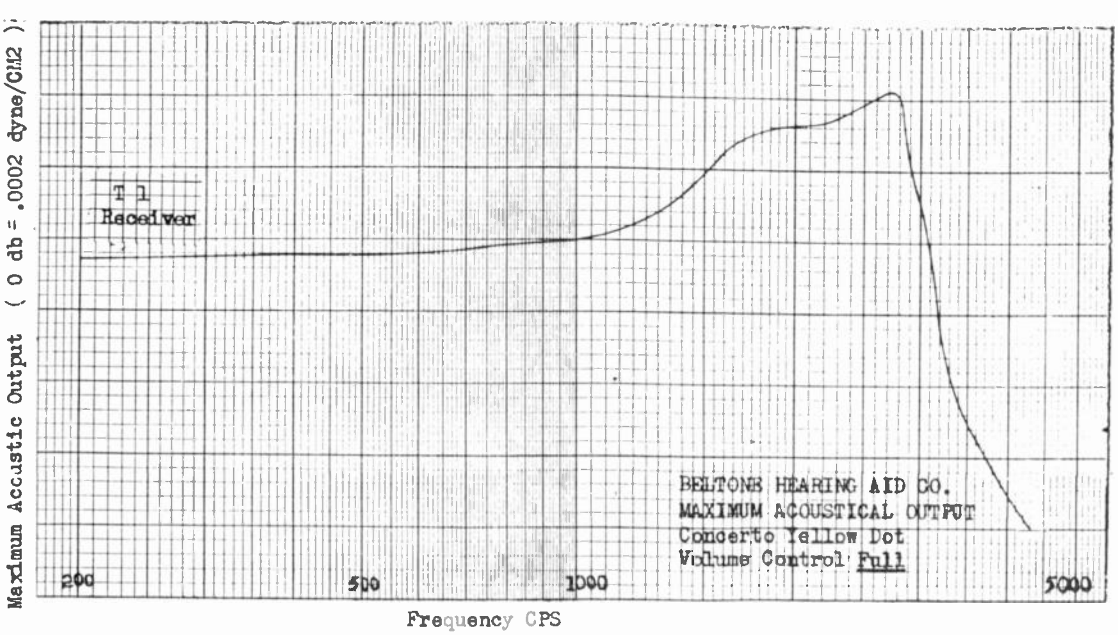

Fig. 7.

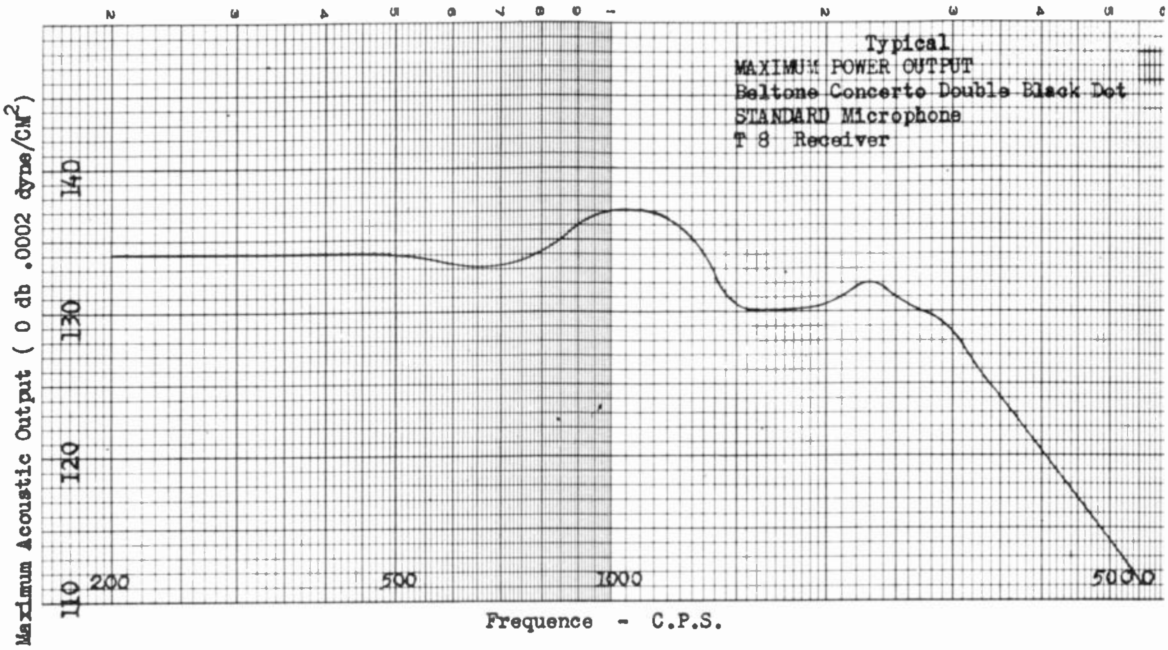
Typical
MAXIMUM POWER OUTPUT
Beltone Concerto Double Black Dot
STANDARD Microphone
T 8 Receiver
Maximum Acoustic Output (0 db .0002 dyne/CM^2)
110
120
130
140
200
500
1000
5000
Frequence - C.P.S.

Fig. 8.

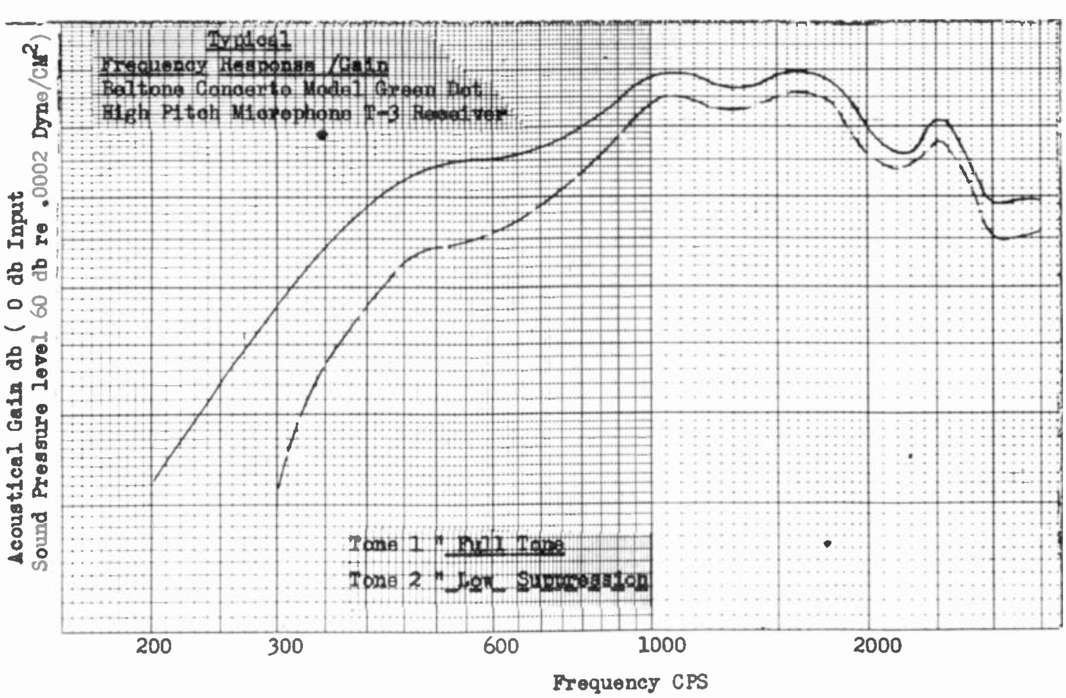
Typical
Frequency Response /Gain
Beltone Concerto Model Green Dot
High Pitch Microphone T-3 Receiver
Tone 1 " Full Tone
Tone 2 " Low Suppression
Acoustical Gain db (0 db Input
Sound Pressure level 60 db re .0002 Dyne/CM^2)
200
300
600
1000
2000
Frequency CPS

Fig. 9.

\$60.00 a year for batteries, this tremendous saving is one of the transistor's most powerful appeals. Even people with very severe losses can be helped to hear for less than one-cent an hour.

The instrument uses a mercury type battery which has a practically constant voltage discharge curve. This gives the instrument uniform gain and power throughout the life of the battery. The voltage of the mercury battery, when it is completely discharged, drops off very rapidly (in a matter of a few minutes).

The total current drain of the 1.25 volt circuit is approximately 1.2 milliamperes. This gives more than seven-hundred hours of life from a battery which has a retail cost of about 30¢. In the case of the very high power instrument where the 2.5 volt battery is used, the total current drain is approximately 2.4 milliamperes, and this gives an average of 160 hours life from a battery costing 60¢ retail.

Reference to Fig. 4 will show the relative frequency response and gain of the various Concerto instruments using the standard microphone and T-3 receiver. Note the wide frequency range and the stair-step levels of sensitivity starting at 48 decibel gain for the Yellow Dot, and increasing to 78 decibel gain for the Double Black Dot.

The Concerto model instrument has the microphone mounted on the front cover. This simplifies replacement in the field. The Concerto model is available with four different front combinations:

1. Standard Front—This permits the use of any one of three microphones. It includes the tone control, but it does not have the Phone Clarifier nor the Fashioner. Note: The telephone clarifier is particularly applicable to transistor circuitry because the coil used for telephone pickup matches into the transistor circuit with a much greater efficiency than was possible when a coil of this type was matched into a vacuum tube circuit.

2. A front cover is available which has both the microphone and the Phone Clarifier.

3. A third front cover is available which uses the Fashionear only. It does not have the Phone Clarifier, but has an internal microphone and the Fashionear microphone socket, so that the Fashionear may be plugged in. When the cord is removed from the front cover, the internal microphone is reconnected.

(The same microphone is used in the Fashionear as is mounted internally in the instrument. Thus the same performance, gain and frequency response are available, whether the internal microphone or the Fashionear microphone is used.)

Figs. 5 and 6 show the wide range of flexibility in the fitting, when the various receiver types are used. All of the curves shown in Figs. 5 and 6 are Green Dot Concerto with a "standard" microphone.

Fig. 7 shows the frequency-response and maximum power of the Yellow Dot Concerto using the midget T-1 receiver. Fig. 8 shows the power output curve of the Double Black Dot with a T-8 receiver.

Fig. 9 is the frequency-response of the Green Dot instrument with a "high-pitch" microphone and a T-3 receiver.

Experimental work is in progress which indicates that eventually grown-junction *n-p-n* silicon transistors will be feasible for use in hearing-aid applications. This may well lead to completely new circuit approaches, and perhaps to a combination of *p-n-p* and *n-p-n* in complementary circuits.

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