# RADIO and ELECTRONICS

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# The New Zealand Electronics Institute

Though this country is little more than a hundred years old, one of the most remarkable things about its development has been the progressive spirit in which current advances in engineering and applied science have been regarded. Our major undertakings, such as railways, telegraph and telephone services, roading, the provision of electric power, and public utilities generally, have always known the benefits due to the possession of the services of engineers and technicians of the highest calibre.

Because electronics is one of the newest branches of applied science, and itself has made such phenomenal strides since its beginnings early in the century, it is only recently that it has come to be recognised apart from the wider sphere of electrical engineering, as can be seen from the comparatively recent date of the formation of such bodies as the British Institute of Radio Engineers. Strangely enough, such recognition of electronics has come in this country several years later than was the case in England, and especially in the United States of America, and this in spite of the fact that its most important subdivision—that of radio—has been in constant development here for so long.

It is particularly gratifying, therefore, to see the formation of a New Zealand Electronics Institute. For too long now have radio men in this country felt the need of such a body, for, in spite of the comparatively large numbers engaged in the various technical aspects of the subject, so far no means have been provided for unifying their interests. The most important function of the newly-formed Institute will be to achieve professional recognition for those engaged in electronic work. To date, however, the only qualifications generally obtainable in this country (apart from membership of the I.E.E. and Brit.I.R.E.) are the Technical Certificate in Broadcasting and the Registered Serviceman's Examination. The latter is of too low a standard to be of much use to the man who wishes to be well qualified in electronics, while the former, though equal in standard to the theoretical portion of the P.M.G. First-class Wireless Operator's Licence, is too circumscribed in its application. Between these two examinations and the post-graduate course in electronics provided by the Engineering School at Canterbury University College, is a very wide gap indeed.

It is in bridging this gap that the New Zealand Electronics Institute will be able to play such an important part. As have the overseas institutions, so the Institute has a graded membership scale. Its constitution provides for an Admissions Committee, whose purpose will be to see that aspiring entrants are admitted to the grades most suitable to their attainments. If this Committee is composed of men of the highest standing in the profession, if before it commences to operate it acquires support in principle from the radio industry and from those Government Departments directly concerned, and, furthermore, if as qualifications for entry, it sets up a graded series of examinations bridging the gap between the Registered Serviceman's Examination on the one hand and the B.E. (Electrical) on the other, the worker in any branch of electronics will have available a series of qualifications recognised by all the interested authorities. Then there will be an incentive such as does not exist to-day for radio men to obtain knowledge of their work in organised channels. When this has become a reality, these men will obtain professional status according to their attainments, and with this will come all the advantages accruing to such status.

The New Zealand Electronics Institute already has the way partly paved towards this desirable end. The rest of the road may not be easy to build unless the Institute is adequately supported from the beginning, and unless its problems are tackled in the right way. The important thing at this juncture is that the Institute has been formed and a start made.

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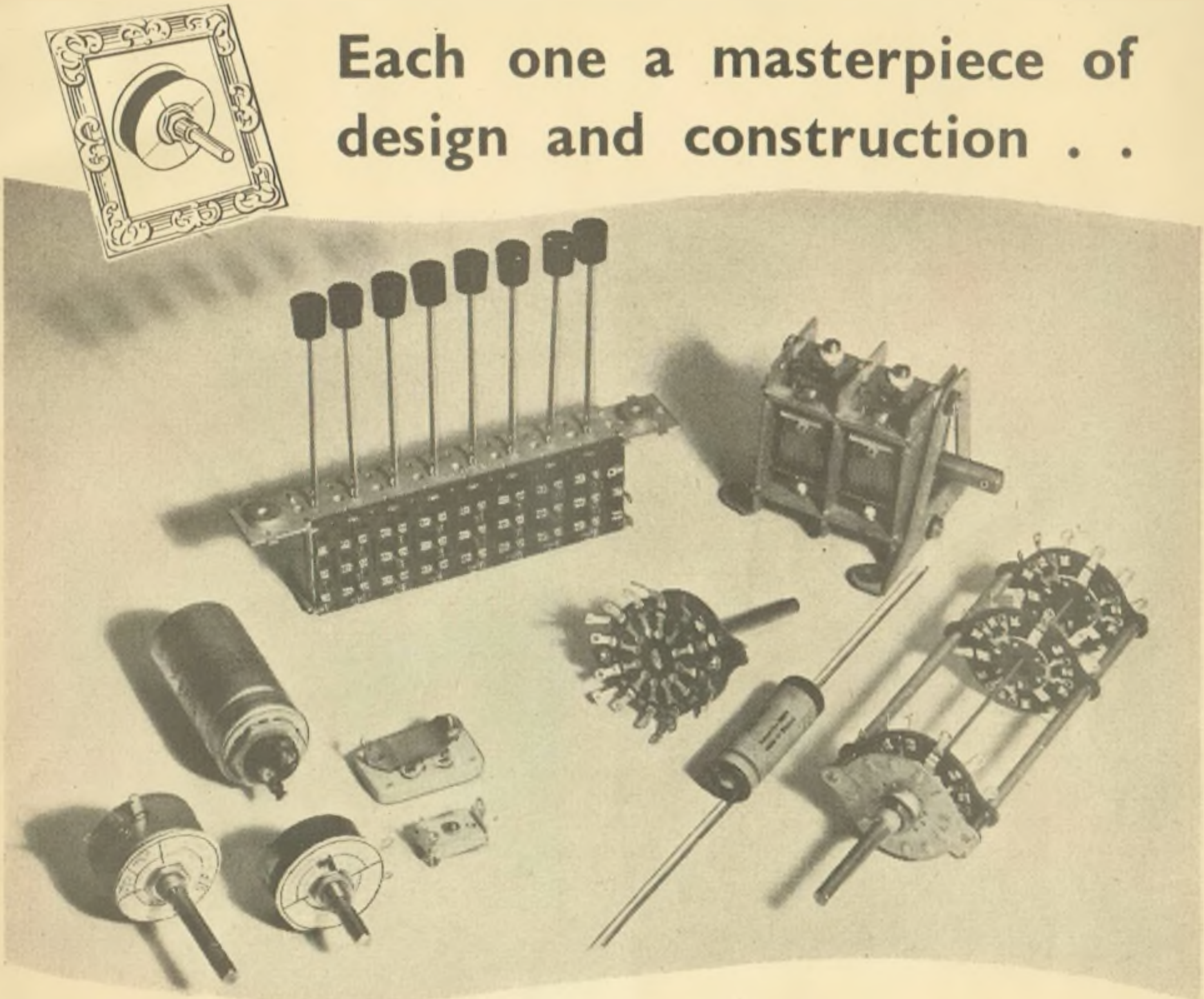

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range than the altitude will be lost in ground returns or other echoes. As the $H_2S$ aerial system is fitted to the under surface of the aircraft, it is apparent that no echoes can be received from above the aircraft, for the transmitter energy will be screened by the aircraft fuselage and wings.

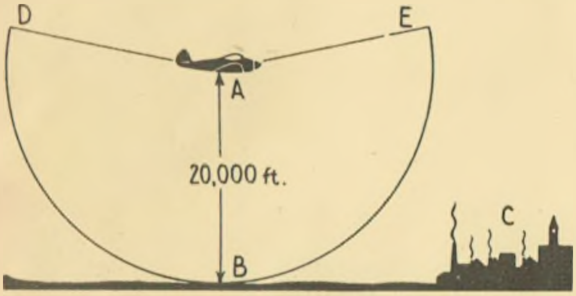


Fig. 1

This is the one disadvantage of Fishpond, but its other advantages warrant its use for proximity warning from any direction other than immediately above. Reference again to Fig. 1 will show that the warning area of Fishpond is that space enclosed by D.B.E.—the limit of detection above the aircraft being about 10 degrees above the horizontal position of the aircraft. As the scan of $H_2S$ is through a sweep of 360 degrees, Fishpond will give warning through the same sweep. Further, the whole of the $H_2S$ P.P.I. is "floating" on the gyrocompass, which means that irrespective of the heading of the aircraft, the top of the screen will always be north and the bottom south. This, of course, will have the same effect for Fishpond, which means that the exact direction of approach of another aircraft can be determined. There is, however, no means of indicating whether the approach is from above or below, but the fact that range and bearings can be found is very valuable information. For reasons related to radar bombing procedure, the transmitter pulse is usually not shown on the P.P.I. and the aircraft position corresponds to the first ground returns. In addition, the P.P.I. screen is only 5 in. in diameter, so that not much space is allowed in which to observe echoes between transmitter pulse and ground returns. For these two reasons, it was found necessary to use another P.P.I. tube in parallel with the $H_2S$ as the Fishpond indicator. Furthermore, the radial time base was "expanded" so that the entire screen area of the second P.P.I. would display only signals up to the maximum service altitude of the aircraft. Apart from this one feature, all other functions of the $H_2S$ P.P.I. are shown on the Fishpond P.P.I. The Fishpond indicator was fitted to the wireless operator's table, and it was his duty to maintain a careful watch on the screen and relay his observations to

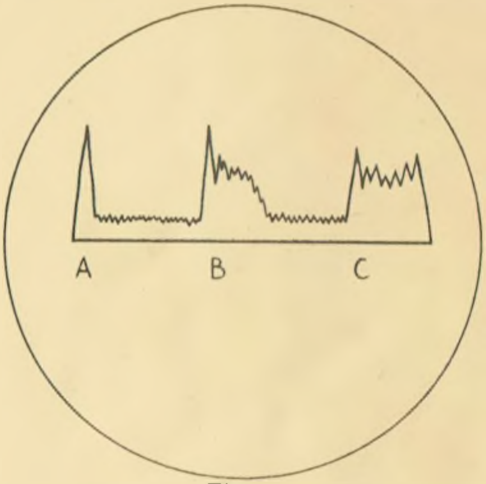


Fig. 2

the crew by means of the aircraft inter-communication set.


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# RADAR

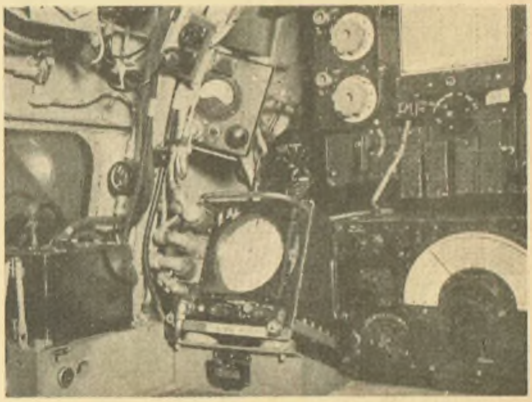
## FISHPOND

**Used as a radar proximity warning device on Britain's bomber aircraft, Fishpond assisted materially in the protection of crews and aircraft from enemy attack.**

With the rapid strides made by British scientists in the development of radar and the application of new techniques in radar warfare, many problems associated with strategic bombing were overcome, and the efficiency and accuracy of the bombing offensive against the enemy improved. Equipment was developed to give considerable aid to navigation and bomb aiming, but it was obvious that any apparatus which would assist in the protection of aircraft crews must, of necessity, be of the highest importance. It was then that radar, with its amazing versatility, was adapted to function as a proximity warning device to warn our bomber crews of the approach of enemy aircraft, and it was for this purpose that Fishpond came into being.

Fishpond is an integral part of $H_2S$ for, in effect, it utilises that portion of the information displayed on the $H_2S$ screen which can be used for no other purpose than height finding or proximity warning. We have outlined the $H_2S$ system in the September issue of *Radio and Electronics*, but some further information on this system is necessary before the fundamental principle of Fishpond is understood.

It is obvious that the first echoes to be received by the $H_2S$ receiver are those from the terrain immediately below the aircraft. These and other echoes of varying ranges and amplitudes combine to make a map-like presentation of terrain within the service range of $H_2S$. For the purpose of an explanation of Fishpond, we will assume that the beginning of the $H_2S$ P.P.I. radial time base is the exact centre of the tube, and that the transmitter pulse will show a bright spot at the start of the time base sweep. No more echoes will appear along the time base until such time as ground returns are received. These will appear as a circle on the P.P.I.—**the distance from the centre of the P.P.I. to the ground returns corresponding in range to the altitude of the aircraft.** This information is an important attribute of $H_2S$, for not only does it function as a radar altimeter, but this measurement is necessary for an accurate calculation of the ground range of a target. It will be realised that measurements of aircraft altitude and ground range are necessary to calculate the release point of the bomb. The $H_2S$ indicator has, in addition to the P.P.I., a range tube. This tube is a normal "A" scope C.R.T., and is used to line up the range markers of the set to measure the range of the ground echo. The dial which controls the marker is calibrated in altitude so that when the marker is set to the ground echo, the actual height of the aircraft is read directly from the dial.



**In the centre of the picture is the Fishpond Indicator fitted to the Wireless Operator's Table. All other equipment shown is for radio communications.**

Fig. 2 shows the echoes as seen on the $H_2S$ range tube and represents the transmitter pulse at A, the first ground returns at B, and a built up area at C. The aircraft is assumed to be at an altitude of 20,000 feet. It is obvious that no echoes can possibly be seen between A and B, as under normal conditions the first echoes which can be received are those from the ground. This point is the main feature in the operation of Fishpond, for any echo seen between A and B must of necessity be from another aircraft in the vicinity.

It is important to note that the proximity warning of Fishpond is restricted to the height of the aircraft, for it will be readily seen that any echo from another aircraft which is at a greater

# A COMPACT 4½ WATT GRAMOPHONE AMPLIFIER

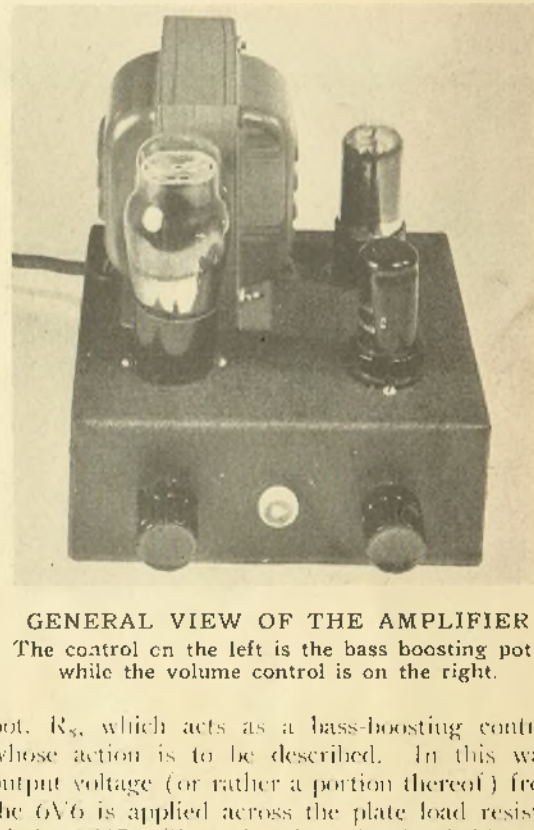
There was published in the April issue of this journal the design of a push-pull 6A3 amplifier capable of some ten watts output of high quality. However, not everyone wishes to build an amplifier as powerful or as expensive as this one. It is still recognised, of course, that for the best quality, push pull triodes are the thing to use, but the fact remains that such amplifiers are more expensive than those employing pentodes or beam tubes. If only small power is required, a single pentode or beam tube is the logical choice for the output stage since by the use of modern negative feedback circuits, the one time bug-bears of high distortion and poor loudspeaker damping may be removed. A beam power tetrode, such as the 6V6, with proper circuit design, can be made to deliver its maximum rated power output with very low distortion, and at the same time to damp out loudspeaker resonance as effectively as does a triode. Thus, although triodes in push-pull are for any power output the sine-qua-non as far as quality is concerned, the use of a single beam tube with negative feedback is a matter of practical economics where a power of no more than three or four watts is required.

## THE CIRCUIT

The present amplifier was designed with these considerations in view. It employs a well-tried circuit which is, however, not very often encountered. In most cases where the 6V6 is used as a single-ended output stage—namely, in the majority of domestic receivers—the circuit given here would require an extra valve to be used over and above the usual number. Thus, the usual five-valve receiver would became a "six," purely for the purpose of providing a good negative feedback circuit, and therefore better audio quality. This is obviously not a payable proposition for the set manufacturer, as it would make a considerable increase in his production cost, but does not prevent the circuit from being admirably suited to the purpose in hand.

The first stage is a 6SJ7 used as a resistance-coupled voltage amplifier. With the circuit constants shown, this tube has a gain of a hundred times, so that with no negative feedback it would be possible to load the 6V6 to full output with an input of only 0.125 volts. This amount of gain would be much too high for normal purposes were it not for the introduction of negative feedback over the 6V6 stage.

The feedback is introduced by connecting a 0.5 meg. resistor from the "high" end of the output transformer primary through $C_6$ to the plate end of $R_6$, which is the 100k. plate load resistor for the 6SJ7. $C_6$ is shunted by the 500k.



GENERAL VIEW OF THE AMPLIFIER
The control on the left is the bass boosting pot., while the volume control is on the right.

pot. $R_8$, which acts as a bass-boosting control, whose action is to be described. In this way, output voltage (or rather a portion thereof) from the 6V6 is applied across the plate load resistor of the 6SJ7. Since the plate circuit of this tube is coupled through to the grid of the 6V6, the feedback voltage is also applied to the grid of this tube. In this circuit, the degree of negative feedback is controlled by the relative values of $R_5$, $R_6$ and $R_7$. Since $R_5$ and $R_6$ are chosen from normal resistance-coupled amplifier practice, this leaves $R_7$ as the component actually exercising control over the feedback. In this case $R_7$ has been chosen to give a gain reduction factor of 3. This means that for a given output, three times as much input voltage must be supplied as compared with the same circuit with the feedback disconnected. It also means that for a given output, distortion is a third of that which would be present without feedback. From the above, it is clear that full output with the feedback applied is obtained with somewhat less than 0.5 volt input. Thus, in spite of the presence of quite heavy feedback, the amplifier still has an overall gain more than sufficient for most purposes.

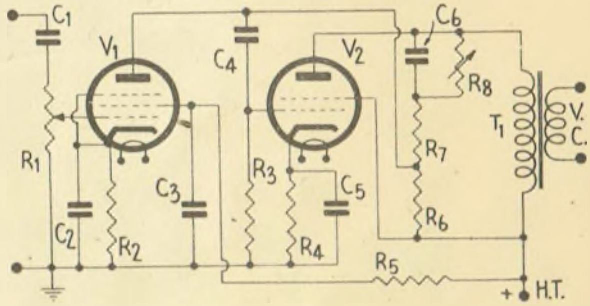
## BASS BOOST CONTROL

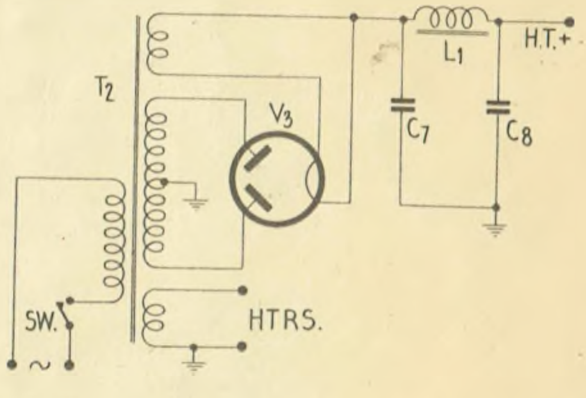
We have called $R_8$ a bass-boosting control rather than the somewhat overworked "tone-control" because the name exactly describes its function. The variable resistor $R_8$ is shunted across the condenser $C_6$, and has a variation of 0–500k. ohms. When the control is turned so that $C_6$ is short-circuited, the negative feedback is equally effective at all frequencies, since the feedback network now contains only resistors. This means that there is the same amount of feedback at all audio frequencies, so that as far as the valves are concerned, the response of the amplifier will be level from quite low frequencies to about 20,000 c/sec. But when $R_8$ is turned so that there is the full 500k. shunting $C_6$, the situation is somewhat different. In this case, $R_8$ is so high in value that it may virtually be regarded as an open circuit, leaving $C_6$ in series with $R_7$ and $R_6$. Now this network will pass high and medium audio frequencies with little or no attenuation, but at low frequencies there is a dropping off in the voltage output of the network at the point where $R_6$ and $R_7$ join. This dropping off is just as if $C_6$ were a grid-coupling condenser that is too small to pass enough low frequency voltage to the next stage, and occurs for exactly the same reason. In this case, the drop in frequency response causes less voltage to be fed back to the 6V6 grid at lower frequencies than is fed back at medium and high frequencies. The amplifier, therefore, has a higher gain for low frequencies than when $C_6$ is short-circuited by the control. The action of $R_8$ is therefore to leave unaffected the high- and medium-frequency response of the amplifier, but to give a controlable amount of extra low frequency gain. It is therefore a true bass-boosting circuit. This has the great advantage over normal "tone-control" circuits that whatever bass response is required, working the control does not attenuate the high-frequencies, nor does it allow the spurious distorted high frequency response to be produced.

typical of pentodes and beam tetrodes working without negative feedback.

## CONSTRUCTION

The layout of the amplifier is very compact, yet it is not over-crowded, and offers no mounting or wiring difficulties. The chassis measures 6½ in. x 6 in. x 2 in., the 6½ in. sides forming the front and back. As can be seen from the photograph, the 6SJ7 and 6V6 are placed along the right-hand side of the chassis; the power trans-




$R_1$ = 1 meg. pot.
$R_2$ = 500Ω.
$R_3$, $R_6$, $R_7$ = 500k.
$R_4$ = 250Ω.
$R_5$ = 100k.
$R_8$ = 500k. pot.
$C_1$, $C_4$ = 0.1µf. 600v.
$C_2$, $C_3$ = 50µf. 25v. Electro.
$C_5$ = 0.5µf. 600v.
$C_6$ = 0.001 mica.
$C_7$, $C_8$ = 16µf. 450v. Electro.
$T_1$ = Output Transformer, 5000Ω to voice-coil
$T_2$ = Power Transformer 385-0-385v. 80ma., 6.3v. 3 amp., 5v. 2 amp.
$V_1$ = 6SJ7.
$V_2$ = 6V6-GT.
$V_3$ = 80.
$L_1$ = 2000Ω speaker field.

former occupies the back left-hand corner and the rectifier the front left-hand corner. The three valve socket holes have their centres 1½ inches from the nearest sides. The three holes in the front on which are mounted the volume control (at the right), the input socket and the base-boost control are 1 inch from the top of the chassis. The input socket is in the dead centre of the front, while the two controls are each mounted 1½ inches in from their respective sides. The holes for the speaker plug and the power lead cannot be seen in the photograph, since they are in the back of the chassis. The former is directly behind the 6V6, i.e., 1 inch from the chassis top and 1½ inches from the side, while the latter is in a similar position, behind the power transformer.

## MOUNTING OF PARTS

In wiring up the amplifier, a combination of point-to-point and resistor strip mounting was used for the small parts. Most of the latter are mounted on a piece of resistor strip having nine lugs. The strip is mounted along the side of the chassis next to the two amplifier tubes. The only small parts in the amplifier circuit proper which are not mounted on it are the two potentiometers, the cathode resistors and bypass condensers of the two stages, the coupling condensers $C_1$ and $C_4$, and the 6SJ7 screen bypass condenser $C_3$. These are mounted as closely as possible to the valve sockets concerned.

The feedback network is mounted on the strip lugs nearest the 6V6 socket. Two wires, twisted together, are taken across the chassis from the feedback condenser to the bass-boost control. The manner in which this twisted pair is run is the only really important point about the layout of the wiring. If this pair, which carries audio voltage at a quite high level, is run close to the input wiring to the 6SJ7, the amplifier as a whole may oscillate at some super-audible frequency, causing overloading and distortion at very low signal outputs. However, by placing this lead in such a way that it does not approach the input wiring, this potential cause of trouble can be eliminated.

## OPERATION

A 0.1 $\mu f$. condenser has been wired into the input circuit so that where the amplifier is used to follow a radio tuner, no blocking condenser will be necessary external to the amplifier. If the amplifier is to be used with a gramophone pick-up only, this condenser could be omitted altogether. In order to use the amplifier with a low level microphone, an extra stage will be necessary, which could be built as a separate unit, power being derived from the amplifier itself. This amplifier is very simple to construct, and the one "bug" mentioned above is the only potential source of trouble. In spite of its small size and simplicity, the amplifier is capable of almost as good quality (within its power-handling capability) as the triode amplifier described in the April issue of *Radio and Electronics*. It should be remembered, however, that the quality of reproduction obtained will be directly proportional to the quality of the pick-up, tuner, and loud speaker used with it. If the best possible results are desired, it is false economy, even with a small amplifier such as this, to use anything but the best pick-up and speaker than can be afforded. If the latter are used, the quality obtainable will be surprising—certainly not very far behind that of a much more expensive amplifier having the same power output.

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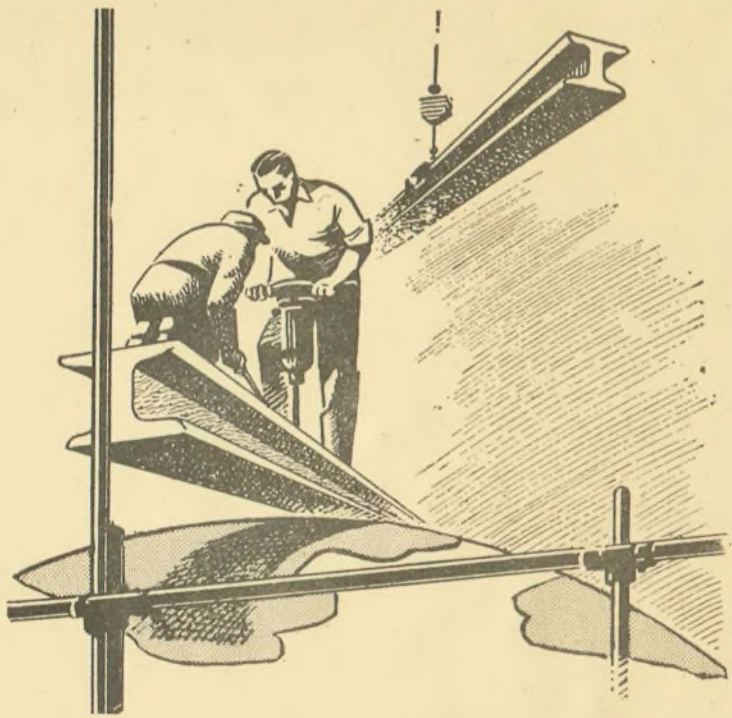

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# AN INTRODUCTION TO WIDE-BAND AMPLIFIERS

## PART I.

The need for really wide-band amplifiers first arose during the development of high-definition television. Since that need was first apparent, an enormous amount of work has been put into the design of special valves suitable for use in these amplifiers, and a great deal has been learned about the mechanism of this type of amplification. It is possible, with present-day tubes and circuits, to build amplifiers with almost any desired frequency or phase characteristic, even to the extent of obtaining uniform amplification

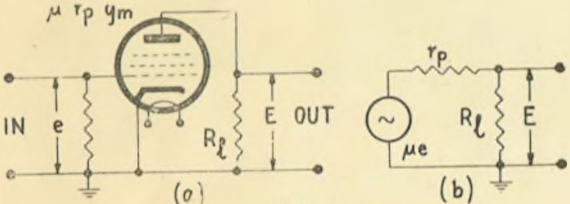


FIG. 1

over a range of 10 c/sec. to 10mc/sec—a range of a million to one! This is an extreme case, and even now would not be an easy design problem, but serves as an example of what can be accomplished.

## NOT ONLY FOR TELEVISION

Although television was the spur which forced the development of wide-band amplifiers, it is by no means the only branch of the science of radio which makes use of them. For instance, had wide-band technique not been in an advanced state of development at the beginning of the recent war, it would have been quite impossible for radar (or any other system using narrow pulses) to have been brought to its present state of perfection. However, it is not with the idea of allowing everyone to become his own radar designer that we present this article. A knowledge of wide-band amplifiers is of very great assistance to anyone who is interested in audio amplifiers or oscilloscopes, for instance. The audio amplifier itself is the direct ancestor of all video amplifiers, for it is itself a wide-band amplifier, and a study of the technique of video amplifiers can give the audio enthusiast a far greater insight into the possibilities and limitations of audio amplifiers. Nowadays, such things as pulse-modulation and the rapid testing of audio amplifiers by means of square-waves must be interpreted in the light of wide-band amplification. Resistance-capacity audio oscillators and the design of oscilloscope circuits require a knowledge of this subject, if not its direct application.

Enough has now been said to indicate that the subject of this article is not necessarily of interest solely to the television or radar engineer, and, as the subject is in itself an interesting one, it is felt that a large number of our readers will be well repaid by at least a simple exposition of it.

## AUDIO AMPLIFICATION

The basis of our discussion is the resistance-coupled audio amplifier, for, as mentioned above, this circuit is the ancestor of all wide-band amplifiers. Fig. 1 (a) shows the basic circuit of such an audio stage. For purposes of discussion, we have drawn all the tubes in these diagrams as pentodes. The simple stage of Fig. 1 (a) has been shown to be entirely equivalent at low and medium audio frequencies to the even simpler circuit of Fig. 1 (b). Here, the valve has been replaced by its plate resistance rp, in series with an imaginary generator whose output voltage is $\mu$ times the input voltage e. As in the actual circuit, the output voltage E is taken from the load resistance Rl, so that from the equivalent circuit of Fig 1 (b), it is readily seen that the output voltage E is equal to

$$\mu e \times \frac{Rl}{Rl + rp}$$

This follows from the simple voltage divider action of rp and Rl. Thus, the gain of the stage at low and medium frequencies is: —

$$G = \frac{E}{e} = \frac{\mu e}{e} \cdot \frac{Rl}{Rl + rp} = \frac{\mu Rl}{Rl + rp}$$

which is the well-known formula, applicable to any tube, either triode or pentode. Before proceeding with the argument, it is necessary to point out a useful simplification of this relation. This simplification applies to the case where rp is very much greater than Rl, so that in practice it can be used only in referring to pentode stages.

If the above condition holds, namely $rp >> Rl$, the formula simplifies to:—

$$G = \frac{\mu Rl}{rp} = gm.Rl.$$

Thus, in a pentode resistance coupled stage, the gain is directly proportional both to the mutual conductance of the tube and to the value of the load resistor. This is the state of affairs for the circuit of Fig. 1, at low and medium frequencies, say, below 500 c/sec.

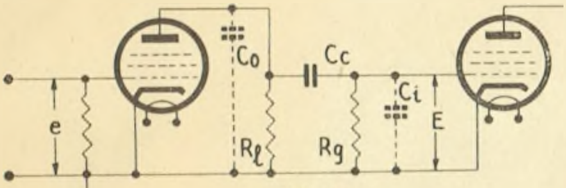

FIG. 2

## LOW AND HIGH FREQUENCIES

Unfortunately, the simple state of affairs shown by Fig. 1 does not occur in practice, either for very high or very low frequencies. The reason for this can be seen from Fig. 2. This circuit shows the original stage resistance-capacity coupled to the next stage. Since the output voltage of the first stage is really the voltage delivered to the grid of the following stage, it has been shown at that point on the new circuit. In addition to the coupling components Cc and Rg, there are shown two new capacities, Co and Ci. These are respectively the output capacity of $V_1$ and the input capacity of $V_2$.

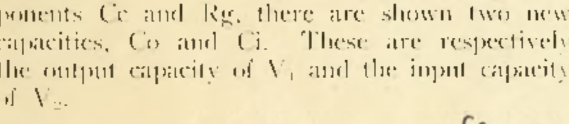

FIG. 3

### High Frequencies:

Now Cc has practically no effect on the operation of the amplifier at high frequencies, because its reactance is so low at these frequencies that it can be regarded as a short circuit. However, the capacities Co and Ci are shunted across Rl and Rg respectively, so that each will have the effect of partially bypassing high frequencies to earth. The higher the frequency, the more effective is the bypassing effect of these capacities. In this way it can be seen that the whole coupling network between the plate of $V_1$ and the Grid of $V_2$ can be regarded at high frequencies as exactly equivalent to Fig. 3 (a). In this, Rt represents the value of Rl and Rg in parallel (Cc having no effect) and Ct is the sum of Co and Ci. This simplification having been made, it can be seen that at medium frequencies, the small capacity Ct has no effect on the output voltage (and therefore on the amplification of $V_1$), because its reactance is so high compared with the value of Rt that no shunting occurs. However, as the frequency is increased, the reactance of Ct becomes lower and lower in comparison with the value of Rt, and an appreciable bypassing effect comes into play. This progressively decreases the output voltage as the frequency is raised, while ultimately a point is reached where there is no output voltage at all, and the stage gain is zero.

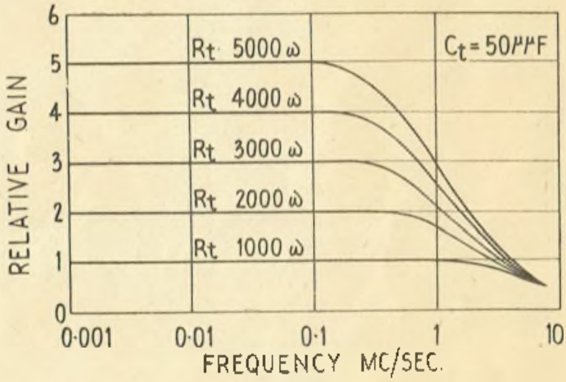

FIG. 4

### Low Frequencies:

As previously mentioned, Co and Ci are too small to have any effect on the low-frequency operation of the stage. Thus, the only extra factor introduced into Fig. 2 on the low frequency side is the coupling condenser Cc. Thus, the equivalent circuit of Fig. 3 (b) is valid at low frequencies. Here, the reactance of Cc becomes progressively higher the lower the frequency, so that Cc and Rg act as a voltage divider, and the grid of $V_2$ receives only a portion of the amplified voltage developed across Rl. In any given cases, the falling-off at low and high frequencies discussed above, takes the forms given in Figs. 5 and 6, so that these can be considered to be the high and low-frequency response curves of any resistance capacity coupled amplifier stage. No specific frequencies have been shown on the diagrams, which will be useful when it comes to discussing the circuit values used in practice. Without further consideration, however, it is obvious that the points at which the frequency response starts to drop off will be determined in any particular amplifier stage by the actual values of Co, Ci, Rl, and Rg.

## SCOPE OF DISCUSSION

The above discussion is quite general in its application, and refers to *any* resistance-coupled stage, either triode or pentode. So far, the causes of the falling-off of the response curve both at high and low frequencies have been presented, without any reference to measures that may be taken to combat these defects. In audio amplifiers using triodes, the high-frequency effect does not come into play until well above the audio range, and, except in unusual circumstances, may be disregarded. In pentode audio stages, however, the load resistors sometimes used are high enough in value for the tube capacities to come into action at as low as 5000 c/sec., and means must be found for avoiding this if high-frequency response better than this is desired. Again, referring only to audio amplifiers, the low-frequency response can be made quite adequate with either triode or pentode stages by making the coupling condenser Cc large enough in relation to Rg.

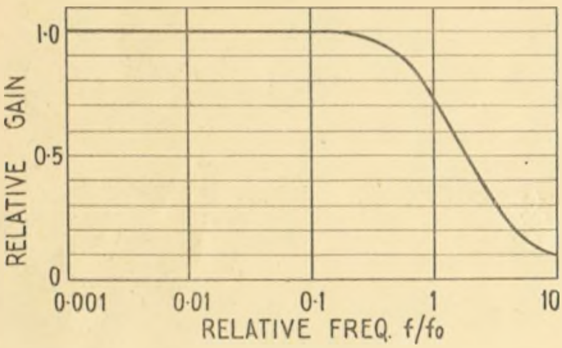


**FIG. 5**

However, when it is desired to extend the frequency range of an amplifier to both higher and lower frequencies than are normal for the audio range, special means must be found for doing this. First of all, we will consider the limitation at the high-frequency end, and one of the ways that can be used for obtaining a wide response band.

## EXTENSION OF HIGH FREQUENCY RANGE

A mathematical analysis of the equivalent high-frequency circuit, Fig. 3 (a), is quite easy to perform, and leads to a very simple and useful conclusion, as follows:—*The high-frequency response drops to 0.707 of the mid-band response at the frequency where the reactance of Ct equals the value of Rt.*

For example, suppose that in a particular case Ct = 50 $\mu\mu f$. and Rt = 50,000 ohms. The reactance of 50 $\mu\mu f$. is equal to Rt (i.e., 50,000 ohms) at a frequency of 63 kc/sec. Thus, this stage would have a response equal to 0.707 of the mid-band response at 63 kc/sec. Since 0.707 of the maximum voltage output represents a drop of 3 db., we can say that the stage in question has a response flat within 3 db. up to 63 kc/sec.

This simple relationship indicates that one way of extending the high-frequency response of a resistance-coupled stage is to use a lower value of Rl, giving a lower value of Rt. In the case mentioned above, if the latter was reduced to a value of 10,000 ohms, the point for a 3 db. drop in response would be raised to 317 kc/sec.! This appears at first sight to give a very easy solution to the high-frequency problem. Within limits, this is true, but these limits are most important.

Firstly, if triodes are being used, the extent to which the frequency range may be extended by this means is severely limited, for a triode

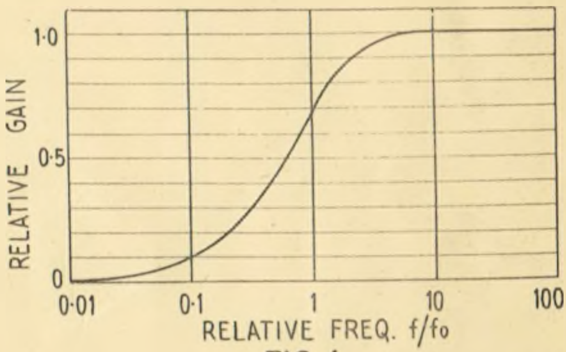


**FIG. 6**

must not be used with a load resistance of much less than twice its plate resistance if excessive distortion is to be avoided. The average small triode has a plate resistance in the region of 10,000 ohms, so that a load resistor of less than 20,000 ohms could not be recommended. In passing, it should be mentioned that, with triodes, Ci is much higher than with pentodes, owing to Miller effect. For these two reasons triodes are

(*Continued on page 48.*)

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# A VACUUM-TUBE VOLTMETER FOR D.C. MEASUREMENTS

**This instrument is easily constructed, accurate and inexpensive. It uses a 0-1 ma. meter movement, yet has a sensitivity of 0-2 volts full-scale on its lowest range.**

Most vacuum-tube voltmeter circuits published in the literature specify the use of a micro-ammeter, which is costly, delicate, and sometimes difficult to replace. With this in mind, *Radio and Electronics* has designed a V.T.V.M. which uses only a 0-1 ma. meter movement, and yet has a maximum full-scale sensitivity of 2 volts. Higher ranges may be provided very easily, and the electrical design is such that very great overloads are impossible. It is hoped, therefore, that the present design will fill a rather long-felt want for servicemen and amateur enthusiasts. Although presented here in its simplest form—for D.C. measurements only—the circuit can form the basis of a quite elaborate meter for A.C. and D.C. measurements, with any desired frequency and voltage range. As the circuit stands, it may be used, without essential modification, as a very high-range ohm-meter with a mid-scale reading of five megohms. The input resistance of the meter is almost infinite on the 0-2 volt range, and is five megohms on all higher ranges.

## THE FUNDAMENTAL CIRCUIT

The full circuit, including power supply, is shown in Fig. 2, but, for simplicity, the basic measuring circuit is given in Fig. 1. $V_1$ is essentially a triode amplifier with a resistive load, $R_2$, and an unbypassed cathode resistor, $R_1$. These resistors are so proportioned that the bias on $V_1$ is a little more than sufficient to prevent the flow of grid current from contact potential in the tube. In other words, the bias is arranged so that, with no input voltage, the tube is working near the top of the straight portion of its mutual characteristic. Thus, if a negative potential is applied to the grid of $V_1$ the plate current decreases. The voltage drop across $R_2$, therefore, decreases, and the plate voltage increases towards the potential of the H.T. + rail.

$V_2$ is a tube similar to $V_1$ in characteristics, and with a load resistor $R_3$ equal to $R_2$. $R_4$ and $R_5$ in series provide a means of biasing $V_2$ so that its plate voltage is the same as that of $V_1$. The variable section is necessary in order to compensate for slight variations in valves and resistors, so that $R_5$ forms the zero-setting control. Since with no input voltage to $V_1$, the plate voltages of $V_1$ and $V_2$ are adjusted to be equal, there can be no current through the meter M, which will read zero. When the above-mentioned negative voltage is applied to the grid of $V_1$, its plate voltage increases. Since the grid of $V_2$ is earthed, no alteration in grid voltage can occur, so that the plate voltage on this tube remains constant under all conditions. Thus, the application of a negative input voltage to $V_1$ results in the plate voltage of $V_1$ rising above the plate voltage of $V_2$: a potential is thus developed across the meter, and current flows through it and is measured. In the complete instrument, the meter shunt $R_6$ is employed as a convenient means of producing the correct full-scale meter reading on the 0-2 volt range. $R_6$ is necessary because, with a 250-volt power supply, as used here, the full scale sensitivity was approximately $1\frac{1}{2}$ volts with the meter used. The latter had a resistance of 50 ohms, but any 0-1 ma. meter that is likely to be used will give approximately the same sensitivity. $R_6$ will be further discussed in connection with calibration.

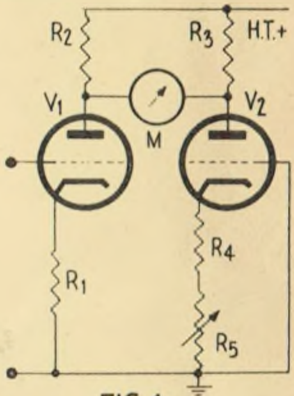


FIG. 1

## POWER SUPPLY

For a meter such as this to be successful, a regulated power supply is essential, but, apart from this, the power supply circuit used is slightly unconventional in that no smoothing choke is used. It is possible to dispense with this because, first, the current drain is quite light, secondly, because the balanced voltmeter circuit renders ineffective any hum voltages applied to the plate circuits, and finally, because of the smoothing effect of the V.R. tubes. The only conventional filtering used is the $16\mu f.$ 450v. electrolytic condenser $C_1$. The transformer used is of the small 40 ma. instrument variety. The value of $R_7$ is such that the V.R. tubes draw approximately 20 ma. The current drain of the two measuring tubes is only 5 ma. with no input, dropping slightly as input is applied to $V_1$. It should be noted that the complete power supply is "floating." There is no direct connection to the metal the test leads are both at potentials above ground. It is important also that $C_2$ should be of 600-volt rating, for, should it break down, the voltmeter chassis will assume the potential of the positive test lead.

On the circuit diagram, $R_9$ has been shown as incorporated in the probe used for the negative test lead. There is a slight advantage in this on account of the filtering action to R.F. that might be induced in the lead under some circumstances, but the probe construction is not

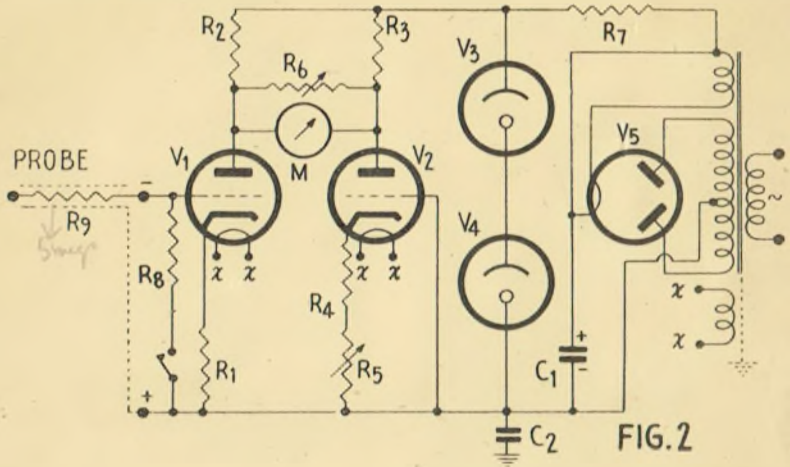

FIG. 2

$R_1 = 500\omega$.
$R_2$, $R_3 = 50k$.
$R_4 = 250\omega$.
$R_5 = 500\omega$ w-w pot.
$R_6 = 200\omega$ w-w pot.
$R_7 = 4500\omega$ 5w. w-w.
$R_8 = 200k$. (for 0-50v. range).
$R_9 = 5$ megs.

$C_1 = 16\mu f.$ 450v. Electro.
$C_2 = 1\mu f.$ 600v.
$V_1$, $V_2$ = 6AC7/1852, screen and suppressor tied to plate.
$V_3$ = VR105/30.
$V_4$ = VR150/30.
M = 0-1 ma. moving-coil meter.
Power Transformer = 300-0-300v. 40 ma.

chassis at any point, but the negative rail of the power supply is earthed to the chassis through $C_2$. The purpose of this is to allow both input leads to be insulated from the chassis, so that the meter may be used with both test leads at potentials above ground, there being then no possibility of shocks, from the chassis of the meter being "up."

## CONSTRUCTION

Since the instrument is used for D.C. readings only, layout is not very important, and any convenient chassis arrangement may be used. The valves should be totally enclosed so that contact with the metal 6AC7 envelopes is impossible. The shell of these valves should be connected not to chassis, but to the negative power supply rail, as otherwise a high voltage would be developed between shell and cathode when absolutely essential, although it is advisable. However, it *is* essential for the shielding of both test leads to be thoroughly insulated, since they will often be at high positive potentials when the meter is in use. $R_6$ and $R_5$ should preferably be mounted on insulating sheet, since, although the terminals of these potentiometers will normally be insulated from their case, this insulation is not necessarily intended for working at voltages up to 600, and a breakdown to chassis of any component in the meter must be avoided at all costs.

## CALIBRATION

The accuracy of the meter can be as great as that of any instrument used as a calibration standard. The first step in the calibration is to adjust the 0-2 volt range. A source of voltage and a calibrating meter is required. The voltage source can be two No. 6 dry cells in series, and the calibrating meter should be as good a one as can be obtained. The procedure is as follows:

(1) Connect a 2000-ohm potentiometer across the voltage source, and the calibrating meter between the moving arm of the potentiometer and the positive battery terminal.

(2) Connect the negative V.T.V.M. test lead to the moving arm of the potentiometer and the positive lead to the positive terminal of the battery.

(3) Turn the potentiometer until the calibrating meter shows zero volts.

(4) Turn on the V.T.V.M. and adjust $R_2$ until the meter reads zero. (About three minutes' warm-up period should be allowed.)

(5) Turn up the input voltage with the 2000-ohm potentiometer until the calibrating meter reads exactly two volts.

(6) Adjust $R_a$ until the V.T.V.M. reads exactly full scale.

The 2-volt range is now completely calibrated and $R_a$ will need no further adjustment. Calibration of the higher voltage range (or ranges) is carried out by providing, as before, a known source of voltage, read on the calibrating meter, and by adjusting the value of $R_x$ until the appropriate reading is indicated on the V.T.V.M. In the case of the prototype constructed in our laboratory, only two ranges were incorporated, the high one reading 0-50 volts. Thus, for the meter to read 50 volts full scale, one-twenty-fifth of the input voltage must be applied to the grid when $R_x$ is switched into circuit. Since $R_4$ is 5 megs., $R_x$ should therefore have a value of 208,300 ohms. Since this is a non-standard value, it will have to be made up by using two standard values in parallel. In this case, 1 meg. in parallel with 250k. gives a resistance of 200k. This value is not correct, but with the tolerance of 10 per cent. normally found with carbon resistors, any such combination will range from 190k. to 210k., so that the correct value is possible to obtain by using stock resistors. All that has to be done is to find three resistors of nominal values of 5 meg., 1meg., and 250k., which, when connected in the appropriate manner will give a voltage division of one-twenty-fifth. Calibration of the range is therefore purely a matter of trial and error until the correct voltage is indicated by the V.T.V.M. This scheme is far better than using a carbon potentiometer for $R_x$, since the latter's setting is more subject to change than is the value of the fixed resistors.

Ranges of 0-250 volts or even higher could be provided by choosing the correct resistors for the position of $R_x$, in conjunction with a known voltage source for calibration purposes.

## USING THE METER

The meter may be used for any purpose where extremely high input resistance is necessary. It is most useful for checking actual voltages at high impedance points of a circuit, such as measuring A.V.C. voltage directly, in cases where trouble is suspected in this part of a set. With it, actual voltages between grid and cathode of self-biased valves may be measured, as can actual voltages at the plates and screens of resistance-coupled audio stages. It will therefore show up faults that a 1000-ohm per volt meter will not indicate, and should form a very useful adjunct to the service bench.

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# SOME NOTES ON THE DESIGN OF AMATEUR-BAND RECEIVERS

## Part 2

### Image Response

In amateur superheterodyne receivers, one point which is often overlooked is that of image response. In any set which has one stage of tuned R.F. amplification ahead of the mixer, image response is ordinarily negligible on the 80-metre band, noticeable on the 40, and quite serious on 20 metres, assuming the usual I.F. in the vicinity of 465 mc/sec. In order to give really good image rejection on the 20-metre band, two stages of R.F. are necessary, and have become standard practice in the best communications receivers. It is not suggested that amateur receivers should necessarily be built to conform with the most stringent commercial requirements, but it is plainly advantageous that the closer they approach these requirements the better, other factors such as cost and ease of construction having been given due consideration. Notwithstanding, the problem of image response is well worth careful consideration, particularly when the congested state of the amateur bands is taken into account, for image responses add directly to the number of heterodyne whistles which can be heard in tuning over the band. In addition, signals completely outside the desired wave-band may be heard by means of image response, and amount to increased congestion from the viewpoint of an operator searching for true signals.

Since image response is a direct function of the ratio of the intermediate frequency to the signal frequency, it may be minimised by two methods—by increasing the selectivity at the signal frequency, or simply by raising the I.F. Unfortunately, I.F. transformers tuned to frequencies higher than 465 kc/sec. are difficult to obtain, and the use of a high I.F. such as 1500 kc/sec. severely limits I.F. gain and adjacent channel selectivity.

Thus, the line-up suggested above as being suitable for the best signal-to-noise ratio is not very good from the point of image response, since it includes no R.F. stage. However, if there is added an R.F. stage which uses a 6AC7/1852, the image response will be greatly improved and the signal-to-noise ratio degraded very little owing to the low noise properties of this valve. From the point of view of good signal-to-noise ratio and good image response, therefore, an excellent line-up for an amateur receiver would be: 6AC7 R.F., 6SN7 infinite-impedance mixer and oscillator, EF38 first I.F. and 6SK7 second I.F. This receiver would have very much better DX capabilities than many commercial communication receivers, and would be admirable for use on the 80, 40, and 20-metre bands. On 10 metres it would also be entirely adequate, except that image response might be somewhat poor on this band.

### GAIN CONTROL

The question of gain control in communication receivers is one which, though outwardly simple, hides quite a number of problems.

#### (1) Strong Signal Distortion:

With all types of gain control, manual or automatic, sufficient range of control must be available, and at the same time a minimum of distortion must be introduced on strong signals. These requirements are mutually contradictory, since even modern variable-mu tubes have a limit to the signal that can be handled without excessive distortion. If the receiver is a multi-stage affair, there is a great danger of overloading the last I.F. stage, unless the control is staggered in such a way that this stage receives only a portion of the total control voltage. Although this principle is fairly well known in its application to A.V.C. systems, it should be realised that it applies equally to manual control circuits where grid bias variation is used. In this case, one-half to one-third of the total control voltage is the right proportion to apply to the last I.F. stage, the full control being given to all preceding stages. In the case of a receiver in which it is desired to use screen-voltage control, staggering is even more important, and may necessitate a quarter or less of the total control being applied to the last I.F. stage, or even leaving it un-controlled.

#### (2) Signal-to-Noise Ratio:

This factor, too, can suffer if the gain control is poorly designed for a high-sensitivity receiver. This is because all ordinary R.F. amplifier and mixer tubes suffer an increase in noise level as the grid bias is made more negative. Unfortunately, this degradation occurs more rapidly than might be expected if the R.F. stages are controlled, and can result in a decrease of as much as 18 db. in signal-to-noise ratio for a signal increase of a hundred times. One way out of this difficulty is to control only the I.F. stages, but if this is done a poor control characteristic results, in combination with greater difficulty over strong-signal distortion.

### (3) Cross Modulation:

Where a weak desired carrier lies adjacent to a strong undesired one, cross modulation is likely to occur if gain control takes place in the I.F. stages only. This indicates that an improvement under these circumstances can be effected by controlling only the R.F. stages. Unfortunately, doing this is in exact opposition to requirement (2) for best signal-to-noise ratio. However, communications receivers often provide separate R.F. and I.F. gain controls for this very reason. The operator can then choose his method of control to suit the reception conditions. Of course, the common method of controlling R.F. and I.F. tubes simultaneously, either manually or by A.V.C., is a compromise between these two conflicting factors. It may be mentioned at this point that in a receiver using a very low-noise "front end," gain control in one or two R.F. stages may not degrade the signal-to-noise ratio to any noticeable extent.

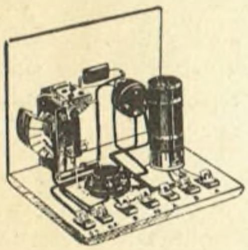
### (4) Mistuning of R.F. and I.F. Circuits:

In spite of the low grid-plate capacities of modern R.F. pentodes, the gain in such stages causes the Miller effect to be appreciable. A detuning effect, therefore, occurs as the gain is controlled. The detuning in any one case will be a function of the L/C ratio of the input circuit of and the selectivity in this circuit. In medium and low selectivity arrangements, the effect will not be serious, but where many stages are controlled, it can be so. The standard cure for this trouble is to leave unbypassed a small portion of the cathode bias resistors in the stages concerned, but the receiver suffers a slight reduction in gain through negative feedback if this is done.

The above four points apply equally to manual or automatic gain control. Those which follow are concerned more especially with the latter.

### (5) Effectiveness of A.V.C.:

By this is meant the degree closeness with which the receiver approaches the ideal of constant output for varying input. First, "simple" A.V.C. without voltage delay is ruled out completely, as this system decreases the sensitivity for weak signals, and does not give a very flat characteristic. For communication receivers a large delay voltage

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# TUBE DATA

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## TYPE ECH-35 TRIODE HEXODE

### General.

Type ECH-35 is a frequency changer valve comprising a triode oscillator and hexode modulator on a common cathode—the triode being mounted vertically below the hexode part.

The signal is applied to the innermost grid of the hexode which has variable mu properties. The oscillator grid is internally connected to the third (injector) grid of the hexode, which is isolated from signal grid and anode by the second and fourth grids which are connected together and comprise the screen.

### Short Wave Operation.

Type ECH-35 is particularly well suited for short wave operation because of its high conversion conductance of 650 $\mu$a/v at full gain and remarkably low frequency drift; the latter remaining low when the valve is under A.V.C. control. Mains voltage fluctuations have also a small effect only and when the oscillator tuned circuit is connected in the triode anode lead a mains fluctuation of 10 per cent. will produce a frequency drift of less than 1 K.C. at 20 mc/s The change in oscillator frequency on biasing the hexode to cut-off is less than 3 K.C.'s at 20 mc/s with a tuned circuit capacity of 50 $\mu\mu f$.

### Conversion Slope.

The high conversion slope results from the high slope of the hexode from grid 1 to anode and the sharp fall of this slope with increasing negative voltage on grid 3. Another result is that only a low oscillator A.C. voltage is required, amounting to 8v. R.M.S. for optimum performance. The conversion slope remains high over a wide range of oscillator voltage; at 5v. it is approximately 580 $\mu$a/v and falls but slightly from the maximum value at double the optimum voltage. This wide tolerance simplifies the design for wide wave ranges since a considerable change of oscillator output has little effect on the conversion slope and amplification. (See Fig. 3.)

The optimum value of 8v. R.M.S. (200 $\mu$a in a grid leak of 50,000 ohms), represents the most favourable compromise between noise, heterodyne whistles and conversion gain.

### Lower Oscillator Frequency.

In the hexode there is no electron coupling between the signal grid and the third grid, but

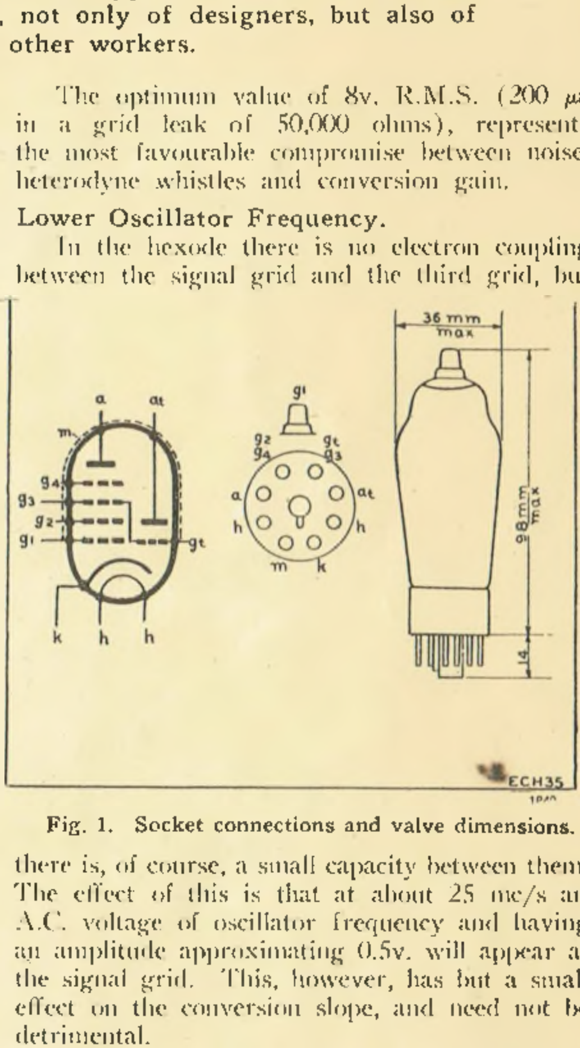


**Fig. 1. Socket connections and valve dimensions.**

there is, of course, a small capacity between them. The effect of this is that at about 25 mc/s an A.C. voltage of oscillator frequency and having an amplitude approximating 0.5v. will appear at the signal grid. This, however, has but a small effect on the conversion slope, and need not be detrimental.

The A.C. voltage at oscillator frequency appearing at the signal grid through the capacity to the third grid increases or reduces the conversion slope, according to whether the oscillator frequency is higher or lower than the signal frequency. It is better to operate the oscillator higher in frequency than the signal.

### Triode Characteristics.

The triode section anode current versus grid volts curves are shown for anode voltages of 150 and 100 in Fig. 5.

## ELECTRICAL DATA.

**1. HEATER**—Indirect heating, A.C. or D.C.

| | |
|---|---|
| Heater voltage | 6.3v. |
| Heater current | 0.3 amp. |

**2.—CAPACITIES:**

**(a) Hexode.**

| | |
|---|---|
| Input Cap. ($g_1$ — cath.) | 5.0 $\mu\mu f$. |
| Output Cap. (p — cath.) | 10.0 $\mu\mu f$. |
| C (p — $g_1$) | 0.003 $\mu\mu f$. |

**(b) Triode.**

| | |
|---|---|
| C (g — cath.) | 9.0 $\mu\mu f$. |
| C (p — cath.) | 5.0 $\mu\mu f$. |
| C (p — g) | 1.6 $\mu\mu f$. |

**(c) Between Hexode and Triode.**

| | |
|---|---|
| C (triode grid — hexode $g_1$) | 0.3 $\mu\mu f$. |

**3. OPERATING CHARACTERISTICS OF HEXODE AS MIXER:**

**(a) Fixed Screen Voltage.**

| | | |
|---|---|---|
| Plate Voltage | 200v. | 250v. |
| Screen Voltage | 100v. | 100v. |
| Cathode Bias Resis. | 215 ohms | 215 ohms |
| Osc. Grid Leak | 50k. | 50k. |
| Osc. Grid Current | 200 $\mu$a. | 200 $\mu$a. |
| Min. Grid Bias | — 2v. | — 2v. |
| Plate Current | 3 ma. | 3 ma. |
| Screen Current | 3 ma. | 3 ma. |
| Conversion Conductance | 650 $\mu$a/v. | 650 $\mu$a/v. |
| Plate Resistance | 0.9 meg. | 1.3 meg. |

**(b) Screen Supply Through Potentiometer.**

| | | | |
|---|---|---|---|
| Plate Voltage | | 250v. | |
| Potentiometer Res. $R_1$ (Figs. 2 and 4) | | 24k. | |
| Potentiometer Res. $R_2$ (Figs 2 and 4) | | 33k. | |
| Cathode Bias Res. | | 215 ohms | |
| Osc. Grid Leak | | 50k. | |
| Osc. Grid Current | | 200 $\mu$a | |
| Grid Bias | — 2v. (min.) | — 23.5v. | — 31v. |
| Screen Voltage | 100v. | — | 145v. |
| Plate Current | 3 ma. | — | — |
| Screen Current | 3 ma. | — | — |
| Conversion Conductance | 650 $\mu$a/v. | 6.5 $\mu$a/v. | 1.5 $\mu$a/v. |
| Plate Resistance | 1.3 meg. | 3 meg. | 4 meg. |

**(c) Condition for Best Sig./Noise Ratio.**

| | | | |
|---|---|---|---|
| Plate Voltage | | 250v. | |
| Potentiometer Res. $R_1$ | | 47k. | |
| Potentiometer Res. $R_2$ | | 33k. | |
| Cathode Bias Res. | | 310 ohms | |
| Osc. Grid Leak | | 50k. | |
| Osc. Grid Current | | 200 $\mu$a | |
| Grid Bias | — 2v. | — 19v. | — 23v. |
| Screen Voltage | 70v. | — | 100v. |
| Plate Current | 1.5 ma. | — | — |
| Screen Current | 1.6 ma. | — | — |
| Conversion Conductance | 450 $\mu$a/v. | 4.5 $\mu$a/v. | 1.5 $\mu$a/v. |
| Plate Resistance | 2 meg. | 5 meg. | 6 meg. |

**4. TRIODE CHARACTERISTICS AS OSCILLATOR**

| | | | |
|---|---|---|---|
| Plate Voltage | 100v. | 150v. | — |
| Supply Voltage (Through Resistance) | — | — | 250v. |
| Plate Supply Resistance | — | — | 45k. |
| Plate Current (Oscillating) | 3.3 ma. | 8 ma. | 3.3 ma. |
| Mutual Conductance | 2.8 ma./v. | 3.8 ma./v. | 2.8 ma./v. |
| Amplification Factor | 24 | 24 | 24 |

Due to the high slope of the triode the starting and maintenance of oscillations will not present any difficulties and the feed-back coil may be rather loosely coupled.

The recommended value of grid leak is 50,000 ohms and the grid condenser 50$\mu\mu f$. These values are suitable for all wave ranges. To keep frequency drift to a minimum and to facilitate tracking of the oscillator, it is recommended to insert the tuned circuit in the anode side of the feed-back coil in the grid circuit of the triode.

In Fig. 2 is shown a typical circuit for ECH-35. For parallel feed of the oscillator as shown, the feed resistor with an anode supply of 250v. should be 30,000 to 45,000 ohms and the blocking condenser 100 $\mu\mu f$. On the broadcast band it may be advantageous to return the feedback coil to the padding condenser, to keep the oscillator more nearly constant. On short-wave the oscillator

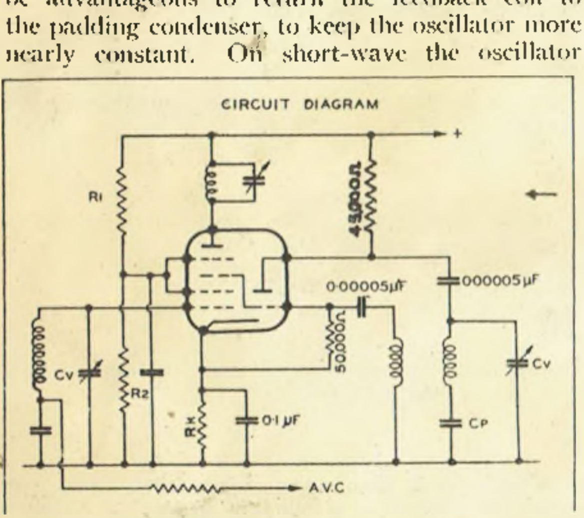


**Fig. 2. Type ECH-35 valve. Typical circuit diagram.**

voltage may be held constant by inserting a resistor of about 100 ohms in series with the feedback coil. This resistor operates with the triode input capacity to produce a damping which increases rapidly with frequency.

### Screen Voltage Supply and A.V.C.

As the mixing section of the valve is a hexode, without a suppressor grid, it is preferable for the screen voltage to be obtained from a voltage dividing network rather than from a series voltage dropping resistor. The variation of screen current with control grid voltage is shown in the graph of Fig. 4. When the conversion slope of the valve is greatly reduced by the biasing voltage applied to the control grid, as may result from the action of the A.V.C. circuits on strong signals, the current through the screen is greatly reduced, and, if supplied through a series voltage dropping resistance, the screen voltage may rise until it approaches that of the anode, under which condition secondary emission may take place from the anode causing a drop in the anode cathode internal resistance which would reduce the selectivity of the tuned circuits fed from the anode.

Fig. 4 shows, by means of two load lines, the variation in screen voltage with control grid volts when the screen is fed through the voltage dividing network $R_1$, $R_2$ shown in Fig. 2 having the values indicated in Fig. 4.

The variation of conversion slope with control grid voltage (A.V.C.) can be controlled by varying the size of the resistors $R_1$ and $R_2$ in the voltage dividing network. Fig. 7 shows the dependence of hexode anode current as a function of the control grid voltage for different values of screen grid voltage (plate voltage 200-250 volts).

### Conversion Slope & Oscillator Grid Current.

Fig. 3 depicts the dependence of conversion slope, and anode internal resistance, on the oscillator grid current through a grid leak of 50,000 ohms, with fixed screen voltage. A fairly flat peak is obtained with a grid current of 200 $\mu a$. The conversion slope falls slightly for higher values.

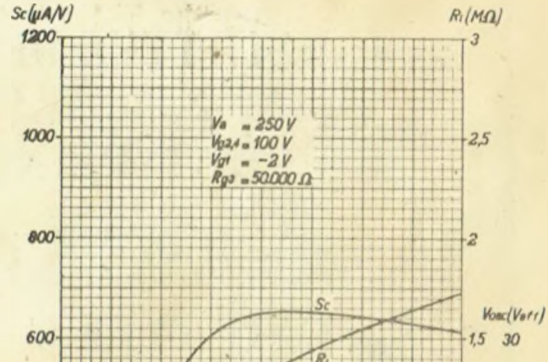


Fig. 3.

### Cross Modulation and Hum Modulation.

In any valve with a curved characteristic—that is, in any valve suitable for control of

# LOUDSPEAKE

## *Units, completely built-up, are available as follow*

### • UNIT TYPE S

Incorporating an R.N.Z. 8 in. P.M. speaker in walnut veneer case with volume control. Designed for general paging and low level indoor sound distribution. Power handling capacity 4 watts. Dimensions: Front, 12 in. x 12 in.; depth, front to rear, 7½ in.

### • UNIT TYPE SFS 8

Incorporating an R.N.Z. 8 in. P.M. speaker in metal case with single directional flare. Designed for general paging and low level sound distribution. Power handling capacity 5 watts. Dimensions: Flare, 13 in. x 13 in.; front to rear, 13 in.

### • UNIT TYPE SFS 10

Similar to Type SFS8 but with R.N.Z. 10 in. P.M. speaker. Power handling capacity 7 watts. Dimensions: Flare, 13 in. x 13 in.; front to rear, 15 in.

### • UNIT TYPE DFS 8

Similar to Type SFS8 except that double flares are provided for bi-directional sound distribution in opposite directions. Dimensions: Flares, 13 in. x 13 in.; length overall, 21 in.

### • UNIT TYPE DFS 10

Equivalent to Type DFS8 but incorporating 10 in. speaker. Power handling capacity 7 watts. Dimensions: Flares, 13 in. x 13 in.; length overall, 24 in.

### • UNIT TYPE NU 8

Searchlight type with spun aluminium casing and flare incorporating R.N.Z. 8 in. P.M. speaker. Pedestal mounted and adjustable for directional distribution of sound. Weather-proofed for outdoor use. Power handling capacity 6.5 watts. Dimensions: Flare diameter, 14 in.; overall height from base of pedestal, 15 in.; depth, front to rear, 12 in.

### • UNIT TYPE G

High-pressure exponential horn speaker incorporating a Grampian pressure unit in a cast aluminium throat with spun aluminium flare. Designed and weather-proofed for outdoor sound projection. With an exponential cut-off characteristic at about 300 c.p.s. the unit is particularly suitable for speech projection. Power handling capacity 10 watts. Dimensions: Flare diameter, 21 in.; length overall, 38½ in.

# RADIO CORPORATION O

80 Courtenay Place
Wellington

# R EQUIPMENT

S

## • UNIT TYPE PR 10

Incorporating an R.N.Z. 10 in. P.M. speaker in a large exponential flared metal casing. Designed for general outdoor use (speech and music) with moderately directional distribution. Power handling capacity 10 watts. Dimensions: Flare, 27 in. x 27 in.: depth, front to rear, 24 in.

## • UNIT TYPE PR 12

Similar to Type PR10 but incorporating a 12 in. Goodman P.M. speaker. Power handling capacity 15 watts. Dimensions: Flare, 27 in. x 27 in.; depth, front to rear, 26 in.

## • UNIT TYPE HR

Infinite baffle type in walnut veneer case with volume control and incorporating R.N.Z. 10 in. P.M. speaker. Designed for general paging and indoor sound distribution. Power handling capacity 6.5 watts. Dimensions: Front, 15½ in. x 18 in.; depth, front to rear, 7¼ in.

## • UNIT TYPE HRL 10

Similar to Type HR but with larger infinite baffle. Designed for high quality indoor sound distribution and musical appreciation work. Power handling capacity 10 watts. Dimensions: Front, 23 in. x 25½ in.; depth, front to rear, 11¾ in.

## • UNIT TYPE HRL 12

As Type HR10 but incorporating a Goodman 12 in. P.M. speaker. Power handling capacity 15 watts.

## • UNIT TYPE BR 10

Bass reflex baffle in a handsome "console" style cabinet incorporating R.N.Z. 10 in. P.M. speaker. Designed for high-quality studio or auditorium use. Power handling capacity 10 watts. Dimensions: Height, 36 in.; width, 25 in.; depth, 15 in.

## • UNIT TYPE BR 12

Similar to Type BR10 but incorporating Goodman 12 in. P.M. speaker. Power handling capacity 15 watts.

**Further details and price list available upon application**

# F NEW ZEALAND LIMITED

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