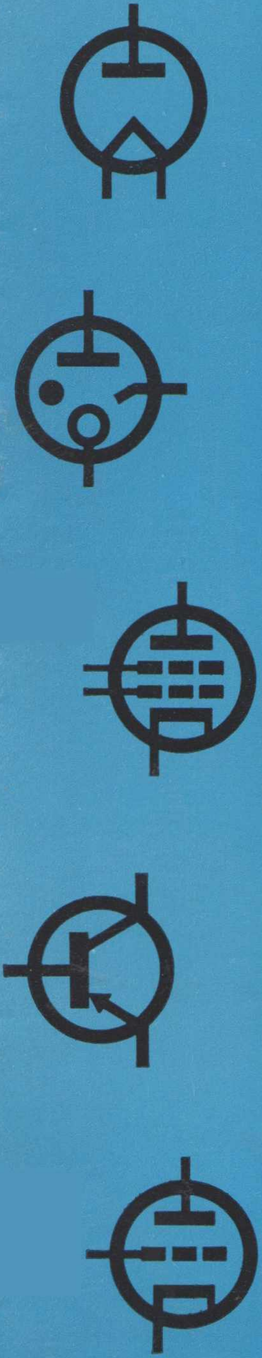


RADIOTRONICS

IN THIS ISSUE

Vol. 29, No. 7 July, 1964

7

AN AWV PUBLICATION

REGISTERED IN AUSTRALIA FOR TRANSMISSION BY POST AS A PERIODICAL.

HYBRID AUDIO AMPLIFIERS

By B. J. Simpson

Part 2: Another Solution

In December last, we introduced into these pages the concept of an audio amplifier using both valves and transistors, for reasons that we explained. Here we present a slightly different approach to the same problem, which has been based on further thoughts on the subject.

Introduction

Since the appearance of the article mentioned in the leading paragraph, we have been able to examine in further detail both the amplifier which was published earlier and some further thoughts on the matter. Readers will remember that what we did last December was to present a valve-equipped main amplifier in stereo form, together with a transistorized stereo control preamplifier unit. The preamplifier unit as published used a pair of input stages suitable for a magnetic cartridge, and it was explained that if a ceramic cartridge was required, the early stages of the preamplifier could be changed for a capacitive-input transistor stage developed earlier, and published in the October and November issues of this magazine last year.

The first hybrid amplifier published was rated for an output of 6 watts, and so is the alternative one to be described now. The distortion and other figures on both amplifiers are similar, except as described here. The main point of difference rests in just where in the amplifier chain one decides to change from valves to transistors, or vice versa. The decision is complicated by the fact that we would naturally endeavour to see that what we publish has as wide a field of application as may be possible, whereas an individual reader will doubtless have already narrowed down the factors on which he has to make a decision.

It must be pointed out early in the article that no attempt has been made to evaluate the comparative costs of the two different approaches described, at least in terms that are wide enough to be published here. From this point of view also, the individual reader, knowing exactly what he wants, will be in a better position to inform himself on this point, in relation to the facilities required.

It must also be mentioned that in providing the alternative suggestions contained here, we have also taken the opportunity of presenting a different arrangement of main amplifier. Whilst either of the main amplifiers may be used independently of the preamplifier with which we have connected them, the two preamplifiers, for reasons which will appear later in the discussion, are not interchangeable between the two main amplifiers.

The Dividing Line

As mentioned above, one of the big problems with a hybrid amplifier seems to be where to draw the line between the two types of active device, valves and transistors. In the first circuit published, a rather obvious line was placed between the main amplifier and the preamplifier. This meant that the main amplifier had no controls, except the on/off switch, and that all other control functions were embodied in the preamplifier. But whilst this may seem an obvious place to strike

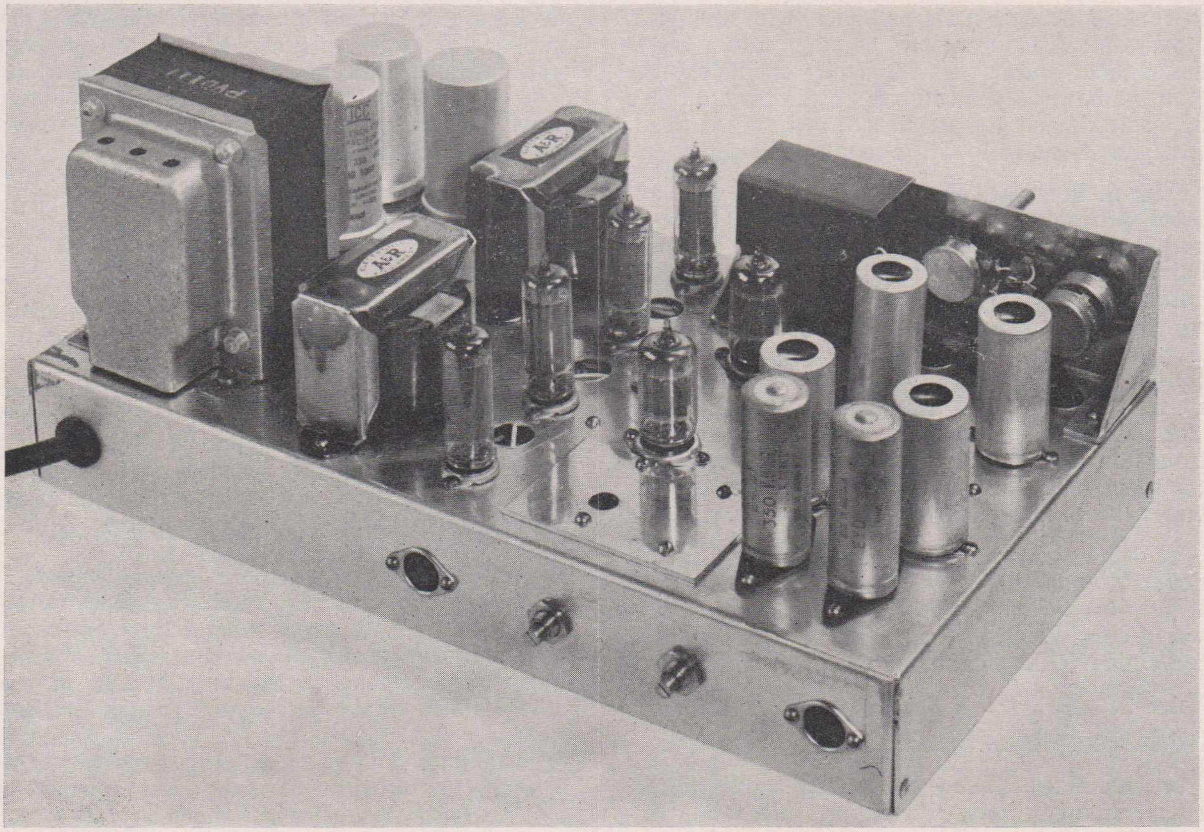


Fig. 1 — View of the valve section of the amplifier described here.

the line, further thought and an evaluation of other arrangements that would be possible, in relation to the likely facilities that the reader may require, makes it appear that the line of demarkation chosen may not be the best under all circumstances. There are so many factors involved that it was thought that the best thing to do was to publish the more likely ideas, and let the reader make his own decision.

One of the main reasons explained for using a transistorized preamplifier was that the comparatively unskilled builder should be able to put one together without encountering as many risks to success as may be present with a valve unit. The main difficulty, however, is in the first stage or first two stages of a preamplifier, where one is not only handling a low signal level, but is usually also applying frequency-selective feedback or using lossy networks to equalise the stage(s) in accordance with the RIAA replay characteristic. This section of the preamplifier will be referred to in this article as the "head amplifier," to distinguish it from the rest of the system.

In view of what has been said, we may very well consider that if we supply a transistorized head amplifier, then we have overcome most of the potential troubles. It then may be stated that the rest of the preamplifier chain may if required use either type of active device, with equal success. Another important consideration is that many people use crystal and ceramic cartridges which need a high load impedance with a flat frequency characteristic. In this type of circuit, it is probably true to say that on the face of it, the electron valve has a slight edge on the transistor. Against this, however, it must be remembered that work already described in these pages indicated that the use of a transistorized capacitive-input circuit with a crystal or ceramic cartridge in general produced a resulting output signal which was appreciably flatter, in terms of frequency response, than the output of the cartridge working into the normal high-impedance load of the order of 2 megohms or so.

The Alternative

It cannot be denied that one of the functions of journalism is to throw ideas into the melting pot, and so induce comment and discussion. It is in this spirit that we now present a different type of amplifier, but one which fills the same sort of purpose. With all the foregoing thoughts rattling around in the mind, together with others less important which will readily occur to the attentive reader, the second amplifier was intended to contain a stereo main amplifier, with tone and other controls incorporated, together with a valve preamplifier with a flat frequency characteristic. Now we have all the controls on the main amplifier instead, and we attach a transistorized head amplifier if we wish to use a magnetic cartridge.

The main chassis has been designed with an input sensitivity of only 30 millivolts for 6 watts output. This, together with the facts that the main unit has a flat frequency response and a high input impedance, means that if a crystal or ceramic cartridge is to be used, then only the main section is required, and is complete in itself. In other words, as far as piezo-electric pickups are concerned, the circuit is simply a valve amplifier. The four-transistor head amplifier can be powered from the main unit, and may be located either in the main unit, or at some remote point, such as under the base of the pickup arm.

It will be seen that this type of solution is not novel, having been used before in various guises. It does have some decided attractions, and is well worth considering. Some of these will appear as the circuit is described and various modifications are discussed.

Main Amplifier Section

As mentioned before, it seemed a pity not to take this opportunity of not only presenting the idea, but also to show a completely different main amplifier in order to make the discussion even more useful. An accompanying circuit shows one channel of the main unit, together with the power supply arrangements which are common to both channels. The main amplifier section is seen at the right of this circuit, and consists of two 6AQ5 and two 12AU7A valves, four all together. There are, of course, two main amplifier sections in each system.

The two 6AQ5s operate in a conventional manner with cathode bias, whereas in the earlier circuit an ultra-linear arrangement was used. The driver and phase splitter are completely different. These portions of the circuit arise partly from a study of all types of driver/phase splitter configurations, and an attempt to select a more suitable circuit for the home constructor than the commonly-used pentode/triode combination.

The latter type of circuit generally uses a triode/pentode valve in a direct-coupled arrangement, and is economical in that only one valve fills the functions of both driver and phase splitter. The direct-coupling in this circuit, and in any other similar circuit, is essential to minimise unwanted phase shifts in the feedback loop leading to instability. The direct coupling, however, helps to make the circuit rather sensitive to changes in circuit values and valve parameters.

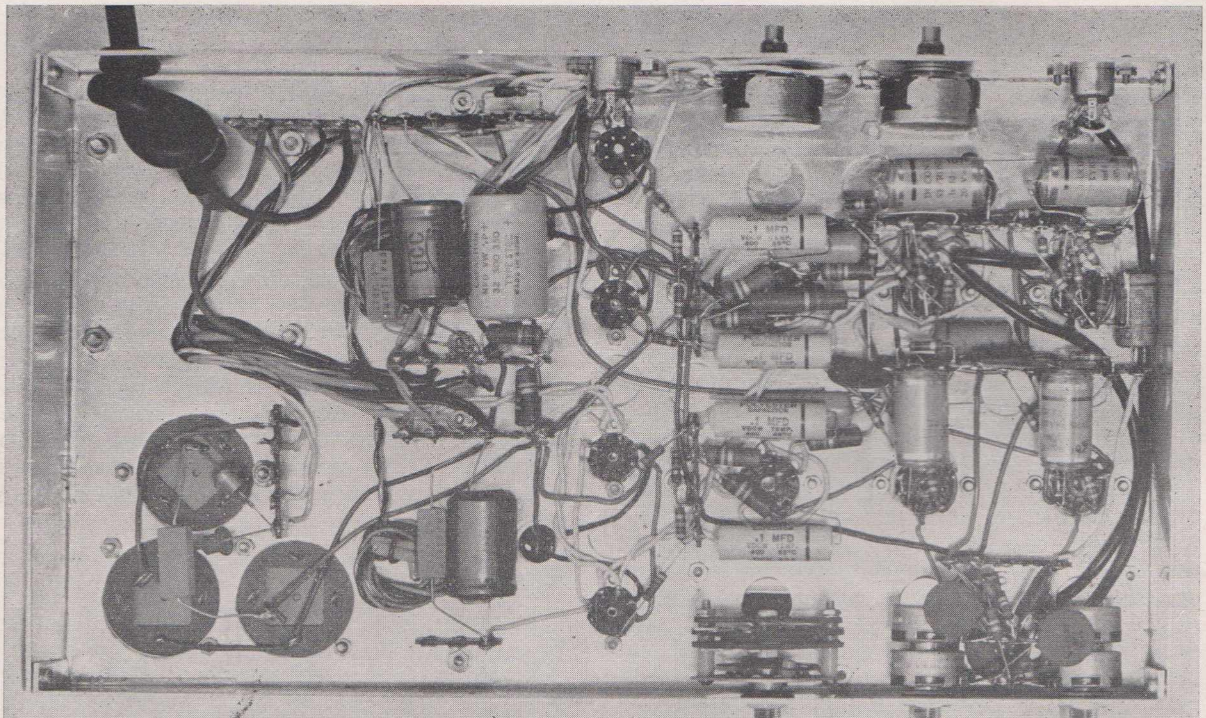
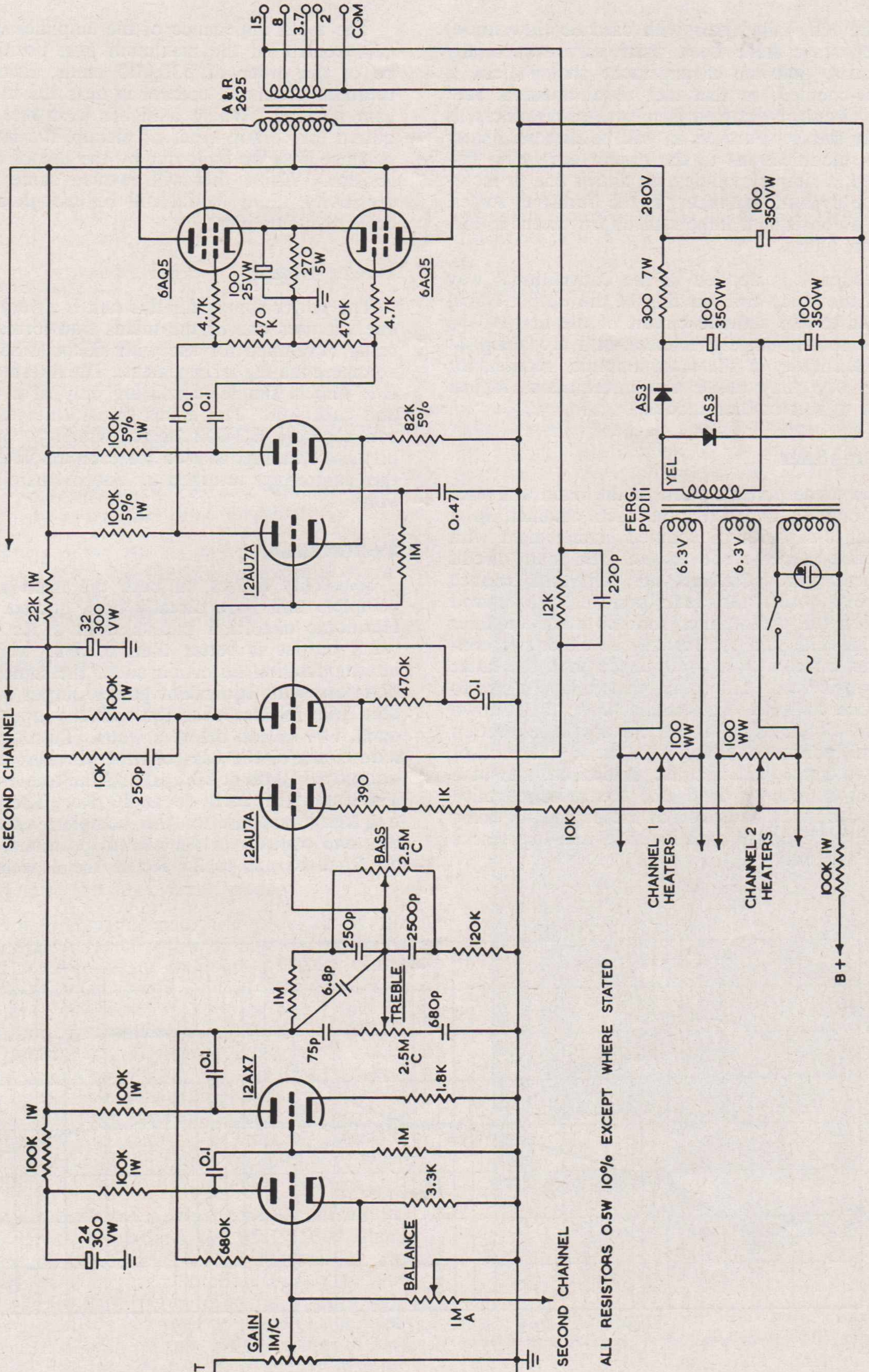


Fig. 2 — View of the underside of the amplifier.



ALL RESISTORS 0.5W 10% EXCEPT WHERE STATED

Fig. 3 — Circuit diagram of the valve section of the amplifier.

The long-tailed pair with cascode (low noise) driver stage have been used very successfully overseas, and the arrangement shown here is direct-coupled, so that that requirement is met. This circuit, by comparison with some others, is highly stable. Further, as will be discussed later when modifications to the circuit are being discussed, a simple modification allows one to incorporate dynamic balancing in the push-pull stages, with a consequent important improvement in distortion figures.

Feedback is applied in the conventional way from the voice coil winding of the output transformer to the cathode circuit of the first of the two cascoded stages, feedback at 1 Kc being 22 db. Sensitivity of the main amplifier section, for those who may desire the information, is just under 1 volt on our model.

Preamplifier

The preamplifier section of the main unit uses only one 12AX7 valve in each channel, in a resistance-capacitance coupled arrangement with feedback over the two stages. The input circuit to these stages contains the gain and balance controls, whilst the tone controls are placed between the preamplifier and the main amplifier section. The gain control may be either two concentric controls or a dual ganged unit, the latter being preferred. The same remarks apply also to the tone controls. The balance control as shown in the circuit is a simple linear single section control, portion of which is in shunt with each of the two input grid circuits. If desired, and at a loss of sensitivity of 6 db, the more effective method already described in these pages of using a dual ganged unit with log. and anti-log. tracks could be used.

The input impedance of the amplifier with the gain control in the maximum gain position will be of the order of 330,000 ohms, rising to 1 megohm when the control is near the minimum gain position. Where a higher load value is required for certain types of pickup, the input impedance may be built out by the use of a series resistor. Whilst this will involve some loss of sensitivity, there should still be enough input to drive the amplifier.

Power Supply

The power supply for this unit is a very simple one. It uses one of the mains transformers specially developed for use with silicon diodes in a voltage-doubling arrangement. The filtering of the B+ line is simple, consisting only of a resistor and capacitor. The mains transformer used has two separate 6.3-volt heater windings, and one of these windings is used for each amplifier, with the centre-taps returned to approximately +25 volts.

Performance

Sensitivity figures on both the main and the complete unit have already been quoted. Total Harmonic distortion measured at 1 Kc with 6 watts output is better than 0.25%, whilst the intermodulation distortion under the same conditions but with equivalent power output is better than 1%. Noise, measured with the input circuit open, is —70 db below 6 watts. Turning to the tone controls, the bass control provides a variation of ± 10 db at 50 cps, and the treble control a variation of ±10 db at 10 Kc. The overall frequency response of the complete unit, with the tone controls in the neutral positions, is ±1 db from 15 cps to 25 Kc at the 1 watt level,

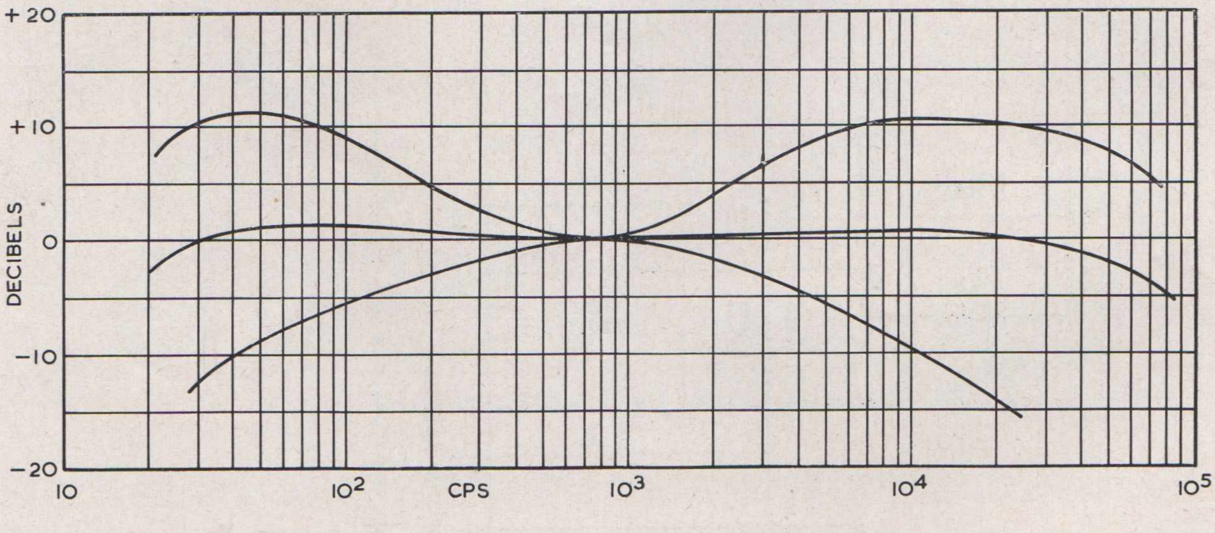
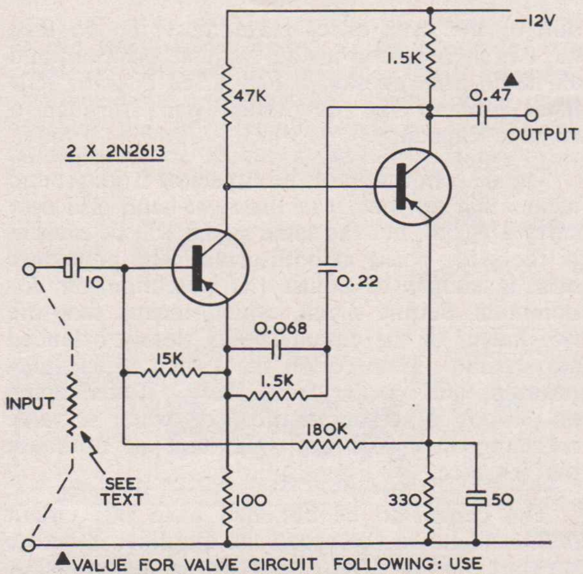


Fig. 4 — Response curve of the amplifier at the 6-watt level, and effect of the tone controls.



▲VALUE FOR VALVE CIRCUIT FOLLOWING: USE 10 TO 25 FOR TRANSISTOR CCT. FOLLOWING

Fig. 5 — Circuit diagram of one channel of the transistorized head amplifier.

3 db down at the 10 cps and 50 Kc points. At the 6 watt level, the frequency response overall is ±1.5 db from 25 cps to 40 Kc, 3 db down at the 20 cps and 60 Kc points.

Head Amplifier

The head amplifier chosen for use with this amplifier is a two-stage transistorized unit using the AWV low-noise 2N2613 transistors. These are arranged in a high-input-impedance, direct-coupled circuit, with frequency-selective feedback to give the RIAA replay characteristic. The voltage gain of the circuit is 20 times (26 db) at 1 Kc, so that a typical input of 5 millivolts will give an output of 100 millivolts, more than enough to load fully the valve main section.

Adequate dynamic range is provided in the head amplifier for any normal cartridge requiring a load impedance up to about 100,000 ohms. With the 12-volt supply as used here, the maximum input levels before clipping are 7 mv, 22 mv, 80 mv at frequencies of 30 cps, 100 cps, 1 Kc, respectively. The circuit is extremely stable, voltage variations of 25% up and down producing no appreciable change in performance characteristics. At lower supply voltages, however, the input level before clipping is reduced, falling to about one-third of the values quoted above when the supply voltage is lowered to 9 volts.

The measured input impedance on our models of this circuit is of the order of 120,000 ohms. This means that the circuit is suitable for any cartridge needing load impedances up to that value, 47,000 ohms being of course a fairly standard value; this is achieved by shunting the input with a 68,000-ohm resistor. Other values of load resistance can be obtained in a similar manner. This circuit, incidentally, is a very useful one, in that it can be used for various purposes. By changing the feedback circuit elements, equalization for tape replay can be incorporated, or a flat response stage for microphone use with a gain of 20 times can be obtained by making the feedback circuit 1.5K ohms in series with a 5 or 10 microfarad capacitor.

Distortion in the head amplifier is very low, and adds nothing to the main unit. Measured distortion at 1 Kc is below 0.1% total.

Construction

Construction of the main unit is very simple, following the photographs reproduced here. Whilst the unit is exceptionally stable, it would be wise to follow the layout used as carefully as possible to reduce noise. Layout should be used rather than screened leads in reducing noise pickup in the circuit, because due to the high impedances used in the preamplifier section, screened leads will lead to high-frequency roll off and a loss of highs.

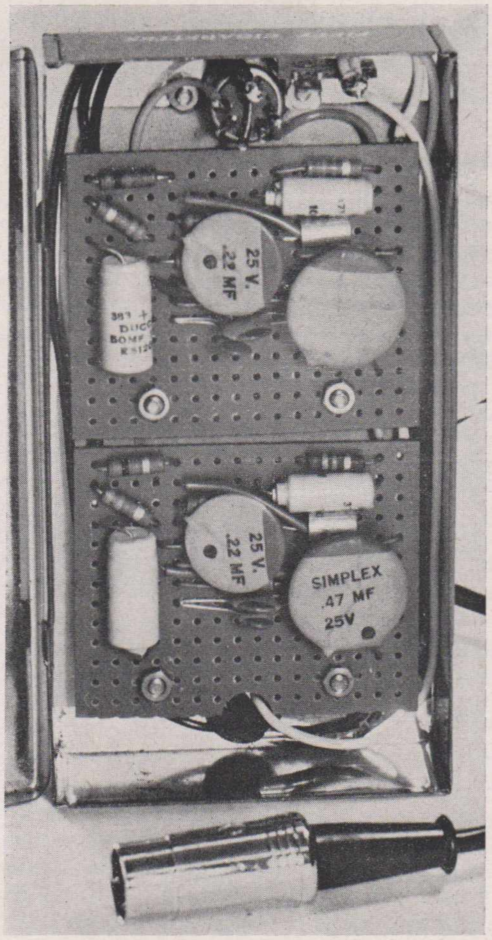


Fig. 6 — Photograph of the head amplifier, which in this case is housed in a "50" cigarette tin.

Provision has been made in the original model for a rotary mains switch and a three-position input selector, but these items are, of course, at the discretion of the builder. The chassis size used for the main unit is 8″ x 14″, and this makes a nice compact unit. The head amplifier is probably best constructed on a small section of matrix board, and enclosed in a small tin or can of steel or tinplate. A small flat tin of the type used to pack throat pastilles would be very suitable. As previously mentioned, the head amplifier may either be incorporated into the main unit, or used under the phonograph motor board.

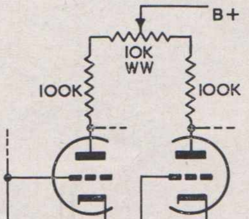
If selection of various inputs is required, it is merely a matter of providing a switch at the front end of the main unit with enough positions, remembering that automatic switching of the power supply to the head amplifier could easily be built into the scheme of things.

The feedback circuit of the main amplifier is arranged for the 15-ohm output. If speakers of other impedances are used, leave the feedback on the 15-ohm terminals of the output transformer, and connect the required terminals to the speaker sockets. The transformer used has provision for speakers of 2, 3.7, 8 and 15 ohms, the sheet enclosed with the unit detailing the various connections.

Balancing Circuit

One of the big advantages of a push-pull output stage is that in theory, even harmonics are cancelled out. It is clear, however, that this process can rarely be complete, due to small imbalances in the circuit. In very high grade circuits, measures are taken to adjust the dynamic balance of the output stage to achieve the lowest possible distortion figures.

A simple circuit of this kind can be used with the type of phase splitter used in this amplifier. A modified circuit of the long-tailed pair is shown here, in which there is a two-way switch with adjust/play positions, and a wirewound potentiometer. The idea is to adjust the signals through each side of the output stage to get best cancellation of the even order harmonics. To do this, the switch is placed in the "adjust" position, and an audio tone at, say, 1 Kc, is fed into the left-hand grid of the long-tailed pair, through a suitable capacitor.



The disconnection of the capacitor from ground means that not only can the right-hand grid now carry a signal, but the same signal will be present in the same phase at both grids. The procedure now is simply to adjust the potentiometer for minimum output signal, which means that the two halves of the circuit are as closely balanced as possible. Then return the switch to the play position, and go on from there. Re-balancing at periods of some months, or when a valve is changed, should keep the unit at optimum performance.

The benefit to be obtained from this circuit is dependent on how good the amplifier is before we start, and to what extent all the various tolerances add up. Typically, though, the use of the balancing circuit gave reductions in total harmonic distortion of up to five times in the main amplifier section of the unit.

The feature of the balancing control is put forward here mainly as a matter of interest, and perhaps an opportunity for those better equipped with instruments to experiment. In fact, of course, with the figures of distortion obtained with the standard configuration, there is little to be gained from a subjective or listening point of view in including a balancing system, as the difference obtained would not be heard.

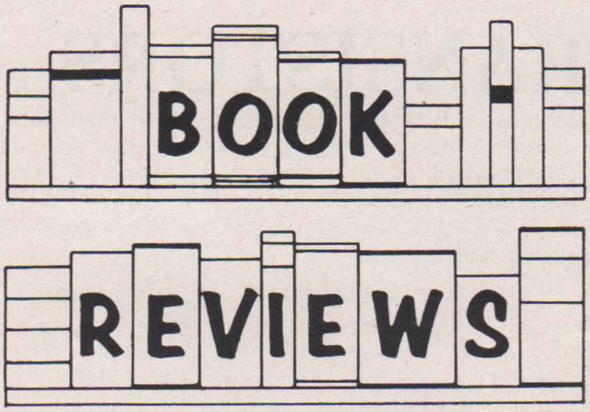
Head Amplifier Power Supply

No specific provision has been made for a power supply for the transistorized head amplifier, as readers will probably wish to introduce a few variations. The choice lies between batteries and a mains power supply. The current drain is so low that the use of a battery is in fact a very acceptable method of driving the unit, especially when the other advantages of battery operation are remembered. If this were done, then extra contacts on the amplifier on/off switch could be used to control the battery also.

If a special mains driven power supply is constructed, this could also be controlled from the main amplifier, and could in fact be built into it. In this event, the power supply should be very well filtered, preferably with a dynamic filter of the type so often mentioned in these pages. We published a basic circuit of this kind in November, 1962, which could be easily adapted for this purpose.

Except for readers who have had reasonable experience with amplifiers, it is not recommended at this stage that an attempt be made to take off dc power for the head amplifier from a point in the main unit. There is, for example, the possibility of constructing the head amplifier "upside down" for positive-active supply, and utilising the dc voltage developed across the cathode resistor of one of the output stages, with suitable dropping resistor and filter capacitor. Our own tests have shown that whilst this sort of thing can be done, it greatly increases the danger of instability in the system; it cannot therefore be recommended at this time.

* * *



"MATHEMATICAL TECHNIQUES IN ELECTRONICS AND ENGINEERING ANALYSIS," J. W. Head, Iliffe Books Ltd. 264 pages, including 65 text illustrations. Size $8\frac{3}{4}''$ x $5\frac{1}{2}''$.

The object of this book is to make available to engineers and other users of mathematics, widely applicable techniques which will make their work easier. The emphasis throughout is on how to **use** these techniques rather than how to **prove** them. Much of the text is based on the series 'Mathematical Tools' by Computer, published in "Electronic and Radio Engineer" and "Electronic Technology"; the subjects discussed are mainly those which have arisen naturally in the course of the author's work in the Research Department of the B.B.C. Engineering Division.

For the most part, the mathematics which is really useful and applicable is surprisingly elementary; the major need is for edequate formulation and specification of the problem to be solved, rather than for over-elaborate techniques of solution. A reader who understands the techniques discussed here is adequately equipped to overcome the fear of mathematics which holds so many in its grip and impedes their progress. He may sometimes need to consult a mathematical specialist to help him solve particular problems, but he will be able to understand his problems and grasp the essential facts about them so that the mathematical specialist need only be called in, if at all, to deal with specific and detailed points.

Although many of the techniques discussed were discovered and invented for the benefit of electrical technology, they can and should be widely applied in other fields, such as mechanics and even economics; for mathematics has a universal quality — the same types of equation or formula can arise in fields as distinct as statistics, acoustics and aircraft vibration. In particular, it is suggested that Heaviside's operational calculus, discussed in outline here, is simpler and more general than Laplace transforms or 'symbolic calculus' and makes the study of these unnecessary.

PRINCIPLES OF FEEDBACK DESIGN, G. Edwin and T. Roddam, Iliffe Books Limited. 238 pages, including 202 text illustrations. Size $8\frac{3}{4}''$ x $5\frac{1}{2}''$.

FEEDBACK is a principle of universal application. Without it the human body could not exist, nor could technology have developed to its present advanced state. It is employed in all systems of automatic control, and in modern amplifiers it ensures stability and reduces distortion. Consequently every electronic engineer should have a sound grasp of the principles of feedback design. A great deal has been written on the subject, but until the appearance of this book there has been nothing to bridge the gap between the elementary treatment of the subject in standard textbooks and very advanced works, so highly mathematical that they frighten the average reader — except for articles widely dispersed in the technical literature.

The book commences with a thorough examination of the application of negative feedback to simple amplifiers, of both the valve and transistor types. An essential part of this problem is to be able to determine the response of an amplifier before it is built; the authors show how this can be done both easily and quickly. The second half of the book deals with more general problems including signal flow diagrams, a discussion of the analytical approach and the use of feedback amplifiers as filters. The final chapter discusses in broad outline a miscellany of other feedback problems with the object of showing how a knowledge of basic principles can assist in the understanding of the behaviour of a wide range of closed loop systems.

MEASURING $r_{bb'}C_c$ AND f_T OF RF AMPLIFIER TRANSISTORS

THOMAS ROBE

(RCA Electronic Components and Devices)

Introduction

This Note describes the methods used by RCA for determining the $r_{bb'}C_c$ and f_T of rf amplifier transistors. Test methods are described, schematics of the test circuitry are presented, and the usefulness of these parameters in rf circuit design are briefly explained.

General Information

The $r_{bb'}C_c$ and f_T parameters are useful for determining the high frequency power gain and noise figure performance of a transistor amplifier. They may also be used to determine the upper limit on the frequency at which the transistor can be made to oscillate. In addition, f_T is a basic parameter for choosing transistors for wideband amplifier applications.

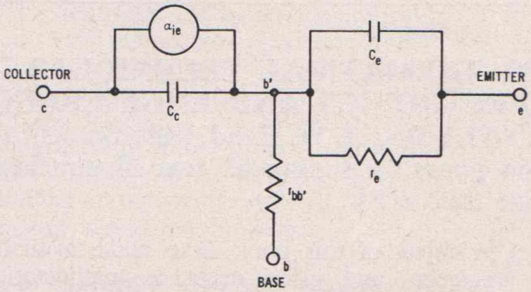


Fig. 1

Nielsen's equation (1) for transistor noise-figure shows $r_{bb'}$ to be of prime importance.

$$\text{NF} = 1 + \frac{r_{bb'}}{R_g} + \frac{r_e}{2R_g} + \frac{(r_{bb'} + R_g + r_e)^2}{2R_g r_e \beta_o}\left[1 + \left(\frac{f}{f\alpha}\right)^2\left(1 + \beta_o\right)\right] \qquad (1)$$

The maximum frequency (f_{max}) at which a transistor can oscillate is that frequency where its power gain is unity. This is given by:

$$f_{max} \cong \sqrt{\frac{f_T}{25r_{bb'}C_c}}$$

Below f_{max}, the unilateral power gain (MAG) under matched conditions increases at approximately 6 db per octave. At any frequency (f) in the 6 db/octave region, the maximum available power gain is found from the relation:

$$\text{MAG} \cong \left(\frac{f_{max}}{f}\right)^2 = \frac{f_T}{25r_{bb'}C_c f^2}$$

A knowledge of the product $r_{bb'}C_c$ and a good estimate of device collector transition capacitance C_c can give a good idea of the value of $r_{bb'}$ for use in the noise-figure equation. At high frequencies, the equation also indicates the effect of frequency cutoff ($f\alpha$) on the noise-figure. Since $f\alpha \cong f_T$, and if f_T is known, it is possible to calculate the increase in NF at high frequencies.

Measurement of $r_{bb'}C_c$

Measurement is accomplished by applying a known value of rf voltage between the collector and base of a common-base amplifier (emitter open-circuited) and measuring the attenuated rf

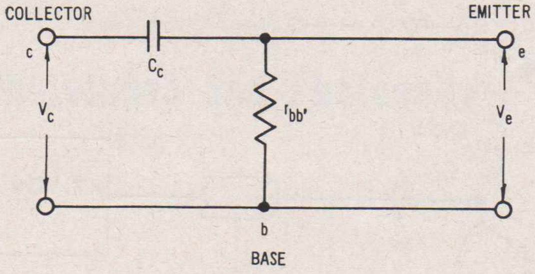


Fig. 2

voltage at the emitter to get the v_e/v_c ratio. The equivalent circuit of the transistor when connected for this measurement, is derived from the common-base T equivalent circuit shown in Figure 1. With the emitter ac open-circuited, it is simplified to a basic voltage divider as shown in Figure 2.

The ratio v_e/v_c is measured from this voltage divider :

$$\frac{v_e}{v_c} = \frac{r_{bb'}}{r_{bb'} + \frac{1}{sC_c}} = \frac{sr_{bb'}C_c}{sr_{bb'}C_c + 1}$$

If the frequency of the applied signal is such that $sr_{bb'}C_c << 1$, the ratio is :

$$\frac{v_e}{v_c} = 2\pi f r_{bb'}C_c = > r_{bb'}C_c = \frac{1}{2\pi f}\frac{v_e}{v_c}$$

If v_c is 0·5 volts (rms) at 31·9 Mc or 0·4 volts (rms) at 40 Mc, $r_{bb'}C_c$ is equal to v_e in millivolts (rms) and can be read on an rf vacuum tube voltmeter such as the Boonton Model 91CA.

A schematic of the test circuit is shown in Figure 3. Effectively, the 10,000-ohm resistor in the emitter circuit presents an ac open-circuit to the emitter, while still providing a dc current path.

Measurement of f_T

Gain-bandwith (f_T) can be determined by measuring h_{fe} at a frequency where h_{fe} is decreasing at 6 db/octave, and then using equation :

$$f_T = h_{fe} \times f_{measurement}$$

A good frequency for this purpose is 100 Mc, then $f_T = 100 \times h_{fe}$.

Since h_{fe} is a current gain measurement, the test circuit should be a good current amplifier. This implies an infinite source impedance compared to the input impedance of the transistor, and zero load impedance compared to the output impedance of the transistor.

A schematic of the f_T test circuit is shown in Figure 4. The series resistance at the input simulates the constant current source, the 10-ohm resistor at the output a short-circuit load. It is necessary to use good high-frequency resistors due to the shunt capacitance across the resistor terminals at high frequencies. Placing two resistors in series as shown, rather than using one of higher value, tends to reduce the end-to-end shunt capacitance.

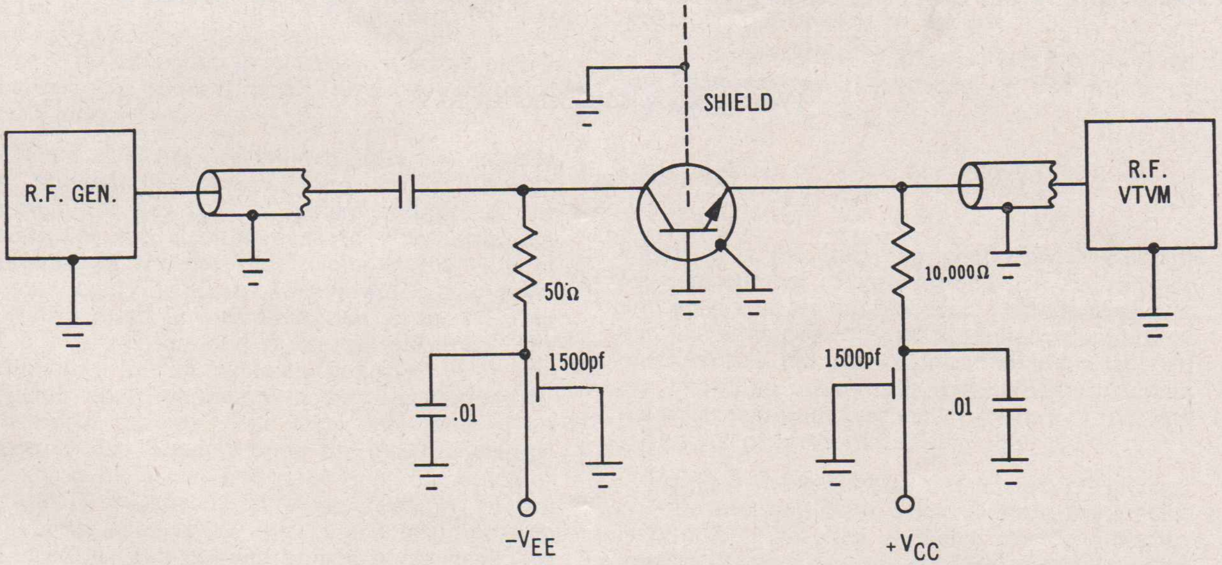
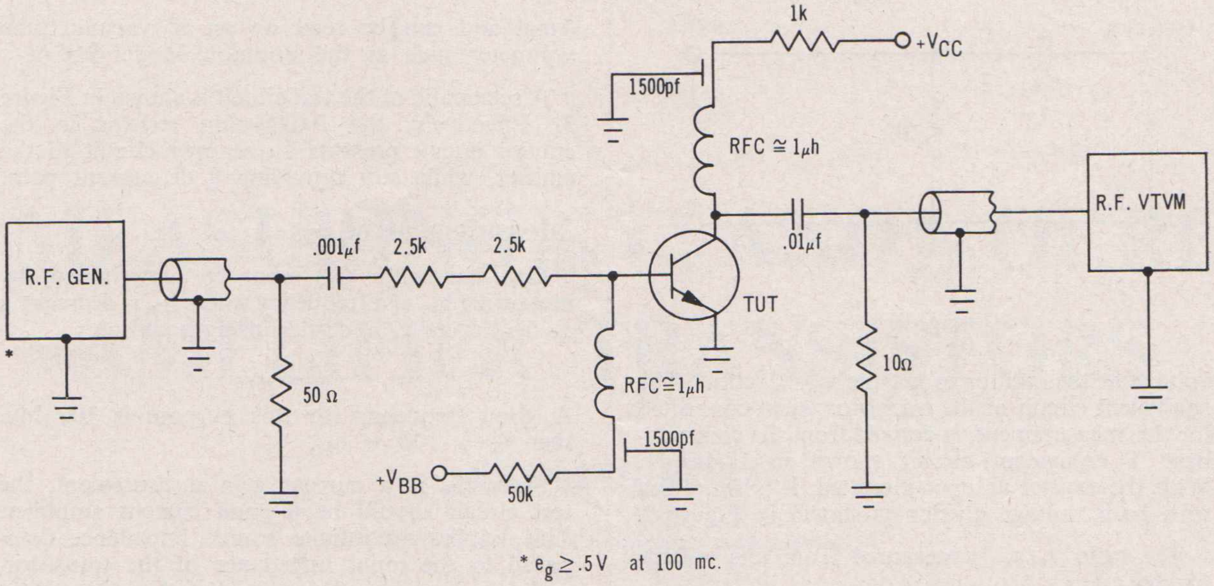


Fig. 3—NOTE: It is important that a metal shield be inserted between the collector and emitter leads ; this reduces stray signal feed-through. Some high-gain units may oscillate in this circuit due to the inductance in the base lead and the lead which connects the case to ground. It may be necessary to place a shorting wire between the base lead and the case lead right at the transistor header to stop such oscillations.



* $e_g \geq .5$V at 100 mc.

Fig. 4

If the transistors to be measured have relatively high input impedances at 100 Mc (near 300Ω), a correlation unit of the type should be obtained. The 100 Mc h_{fe} of this unit can be determined on a G.R. 1607-A Trans Imittance Bridge. If such equipment is not available, correlation units with measurements can normally be obtained from the transistor manufacturer.

With the correlation unit in the socket and the bias set to the desired value, adjust the generator voltage to give the known h_{fe} reading on the rf VTVM. Use a convenient scale of the VTVM, normally 10-30 millivolts full scale.

If the input impedance of the transistor at 100 Mc is negligible compared with the source impedance, a short can be placed in the socket between the base and collector pins and the signal generator set to give 1 or 10 millivolts on the VTVM. The short is then removed and transistors to be measured are placed in the socket at the proper bias. The readings in millivolts divided by the reference of 1 or 10 millivolts is the desired h_{fe}.

(With acknowledgements to RCA)

Phototubes and Photocells

3: GAS-FILLED PHOTOTUBES

Construction and Principles of Operation

Most gas-filled phototubes have the same general construction as vacuum phototubes except that an inert gas such as argon at a pressure of approximately 0.1 millimeter of mercury is introduced before the final sealing of the tube. Ionization of the molecules of the inert gas results in amplification of the primary photoemission. This amplification provides an important advantage over the vacuum phototube for applications in which the primary photocurrents are small and it is necessary to minimize external amplification. With gas-filled tubes, amplification factors of from 5 to 10 become quite practical; even higher amplification can be used under carefully controlled conditions.

When electrons are emitted from the cathode by photoelectric action, they are accelerated through the gas by the applied voltage. If the energy of the electrons exceeds the ionization potential of the gas (15.7 volts in the case of argon), collision of an electron and a gas molecule can result in **ionization,** that is, the creation of a positive ion and a second electron. The probability of an ionizing collision in a gas depends upon the energy of the electron and the density of the gas. The mean free path of an electron also depends upon the electron energy. In argon the mean free path is of the order of 1.4 to 3 millimeters at a gas pressure of 0.1 millimeter of mercury. (For a general discussion of electrons in a gas and related phenomena, see references 1 and 2.) Not every collision results in an ionization. As the voltage on the phototube anode is increased above the ionization potential, the amplification of the photocurrent increases. At 90 volts, an electron averages 3 ionizing collisions while traversing the gap between cathode and anode. This degree of ionization results in an eight-fold amplification (2^3) of the photocurrent.

However, in practice, the actual amplification is greater than that which results from ionization because secondary effects become more important as the voltage on the tube is increased. The most important of these effects is the release of secondary electrons when positive ions strike the photocathode. Other effects of minor importance are the release of secondary electrons by metastable atoms produced by electron excitation in the gas, ionization by positive ions, and electron emission from photons created in the gas.

The combination of these effects produces an amplified current i described by the following equation:

$$i = i_0 \frac{e^{\alpha d}}{1 - \gamma(e^{\alpha d} - 1)} \qquad (25)$$

where i_0 is the initiating photoelectric current, α is the number of ions formed per electron per unit length across the tube from cathode to anode, and γ is a lumped constant (nominally the number of secondary electrons emitted from the cathode per impacting positive ion, but actually including the other minor and secondary sources of regenerative current in the tube).

Eq. (25) shows that, as the voltage is increased on the tube and both α and γ increase, a point is ultimately reached at which the denominator approaches zero as a result of the combined effect of the primary and secondary mechanisms. At this point, a state of uncontrolled current is reached, and the current increases to the limit of the circuit or until a **glow discharge** sets in, which may result in permanent damage to the photocathode.

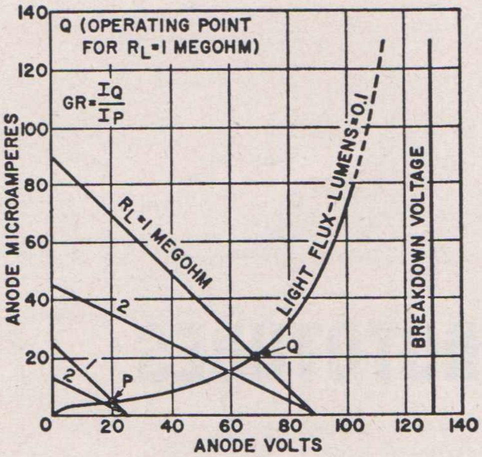


Fig. 25—Current-voltage characteristic for a gas-filled phototube illustrating gas-ratio (GR), load lines, operating point, and breakdown voltage.

Properties of Gas-Filled Phototubes

Current-Voltage Characteristics

Fig. 25 shows the increase in anode current of a gas-filled phototube as the voltage is increased. Most commercial gas-filled phototubes are designed to operate with a 90-volt supply. The intersection of the load line and the anode-current characteristic defines the operating point. The ratio of this current to the current at 25 volts (with the same load) for a specified light flux (usually 0.1 lumen) and a specified load (usually 1 megohm) is referred to as the **gas-ratio** or **GR.** The **breakdown voltage** is that voltage at which, with no light on the tube, an uncontrolled discharge occurs. This voltage is well above the 90-volt maximum operating voltage to provide for stable performance.

Variation of Current With Light Flux

A series of current-voltage curves for various values of light flux is shown in Fig. 26 for type 918. As the level of light increases, the current increases more than linearly. This relationship is illustrated more specifically in Fig. 27, which shows the current developed as a function of light flux for a gas-filled phototube. The nonlinear behaviour is caused by positive-ion space charge. The field strength near the cathode is low because of the cylindrical construction. The mobility of positive ions is much less than the mobility of the electrons; at a current of approximately 20 microamperes, the accumulation of positive-ion space charge is sufficient to distort the cylindrical field and increase the electrical gradient near the photocathode. This increased gradient provides a more efficient field distribution for the production of multiple ionization than when the bulk of the voltage drop is concentrated near the anode, as in the case of the undistorted cylindrical field. For applications in which linear response is required over a wide range of light levels, it is best to use a vacuum phototube.

Time- Or Frequency-Response Characteristics

The time of response of as gas-filled phototube, unlike that of a vacuum tube, is limited by the secondary effects associated with gas amplification. Fig. 28 shows the frequency-response characteristics of gas-filled phototubes having cesium-antimony photocathodes and silver-oxygen-cesium

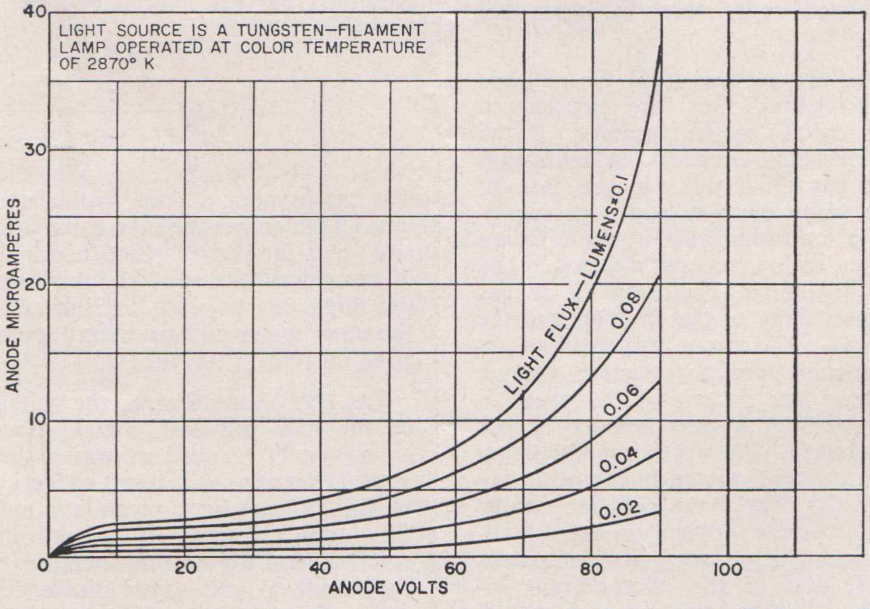


Fig. 26—Current-voltage characteristics for various light levels for a type 918 gas-filled phototube. Some nonlinearity with light is observed at maximum currents.

photocathodes. Because response becomes increasingly poor above 10,000 cycles per second, applications are limited to the audio range. Gas-filled phototubes are widely used in pickups for sound reproduction, both in theatres and in 16-millimeter sound systems. The frequency-response characteristic shown in Fig. 28 was obtained by passing light through a toothed wheel driven by a variable-speed motor and then through a fixed aperture and onto the photocathode. The teeth were so shaped that in combination with the aperture they produced a sinusoidal variation of the light flux.

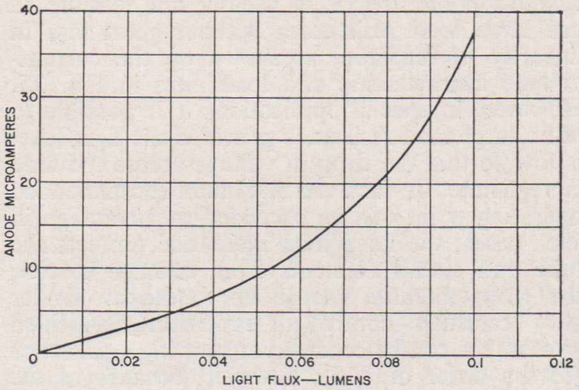


Fig. 27—Anode current (at 90 volts, zero series resistance) in a gas-filled phototube as a function of light flux showing the increasing nonlinearity at high levels of light flux.

The loss in high-frequency response is chiefly the result of the transit time of the positive ions involved in the gas-amplification process. The loss of frequency response becomes more severe as the gas amplification is increased. On the other hand, when an argon-filled phototube is operated at a voltage below the ionization point for argon, it behaves very much like a vacuum phototube with no gas amplification and little loss in frequency response.

At normal operating voltage for a gas-filled phototube, the transit time of the positive ions is less than 10 microseconds. Cumulative effects of the regenerative process cause slightly longer delay times for part of the current. However, a small component of the current is delayed by too great a factor to be the result of positive-ion transit-time effects. Ordinarily, only a small percentage of the total current shows this effect. The slight falling off of the frequency-response curves (Fig. 28) near 1000 cycles per second and less which results from this effect increases as the gas amplification is increased.

Fig. 29 shows this component of the delayed current for a special gas-filled phototube designed for use in studying the mechanism of delay in gas amplification.[3] The very slow component of the gas-amplified photocurrent results from secondary-electron emission by metastable atoms. The transit time of metastable atoms (of the order of 10^{-4} second) is governed by diffusion time and is not affected by the electric field.

Noise

The gas-amplification process is not entirely noise-free because the gas ratio for an individual photoelectron is a statistically variable quantity. However, additional noise resulting from the gas-amplification statistics is only a fraction of that which results from the random emission of electrons from the photocathode. The **Equivalent Noise Input** for a gas-filled phototube is nearly the same as for a vacuum phototube having equal photocathode sensitivity provided the vacuum type is followed by a noiseless amplifier having a gain

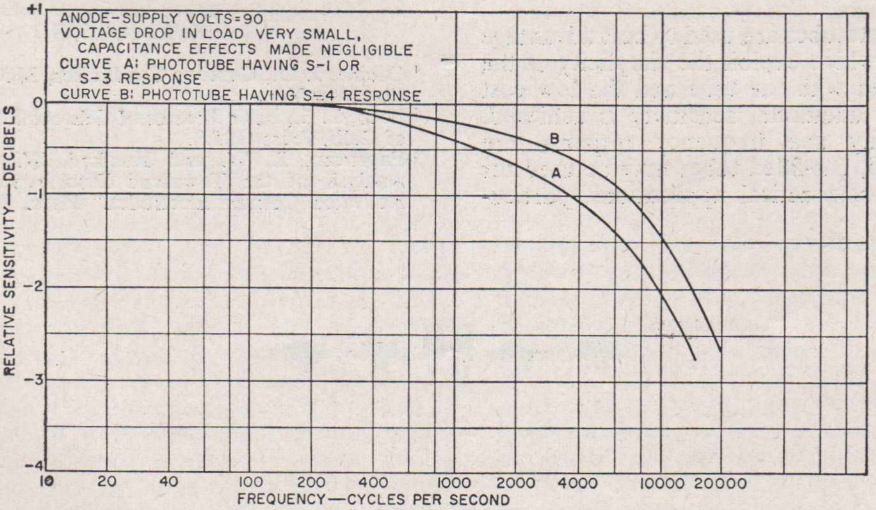


Fig. 28—Frequency response of gas-filled phototubes: (a) Response of a tube having S-1 spectral response (Ag-O-Cs photocathode); (b) Response of a tube having S-4 spectral response (Cs-Sb).

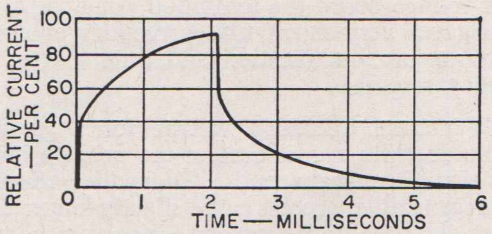


Fig. 29—Time response to a square wave flight emphasize the lag resulting from secondary for a special gas-filled phototube designed to effects of metastable argon atoms (Ref. 3).

equal to the gas ratio of the gas tube and a bandpass limited by the frequency-response characteristic of the gas-filled phototube. The principal advantage of the gas amplification in realizing low equivalent-noise input is to reduce the value of the load resistance for which the resistor noise is equal to the thermionic shot noise of the tube.

Environmental Factors

The previous discussion of the effects of temperature on vacuum phototubes applies also to gas-filled phototubes. Because the number of gas molecules does not change with temperature (even though the pressure does), the gas amplification does not vary appreciably with temperature.

In other respects (shock, vibration, humidity) the behaviour of the gas tube is similar to that of the vacuum phototube. However, because the positive ions bombard the photocathode during operation, the life and the stability of a gas-filled phototube are not as good as the life and stability of vacuum phototubes operated at the same current.

Application Considerations

Gas-filled phototubes are used to best advantage in applications which exploit the simplicity of the circuit associated with the tube and the low cost with which the additional sensitivty is achieved. Because linearity and frequency response are reasonably good, gas-filled tubes may be used for a wide variety of practical applications, particularly when the more precise characteristics of the vacuum phototube are not needed. The most important use of these tubes is in motion-picture sound-on-film sensor systems for theatre and home projection equipment.

It is especially important not to exceed the absolute maximum voltage and current ratings of gas-filled phototubes; excessive voltages can cause damage from ionization effects, and excessive currents can result in loss of sensitivity.

Because the gas-filled phototube does not have the flat current-voltage characteristic of the vacuum phototube, it is usually not feasible to use large load resistances without great loss in linearity of response, as shown by the current-voltage characteristic and load lines in Fig. 25. However, in special applications it is possible to use a large load resistance provided the light level is low so that the drop in voltage across the load is negligible. In fact, the maximum recommended operating point can be exceeded to advantage in such cases; the large load resistance protects the tube and circuit elements from damage in case the glow potential should be exceeded. Under very carefully controlled conditions, gas-filled phototubes can be operated at very high gas ratios (of the order of 100); however, because of the inherent instability of the tube under these conditions, such operation is normally not recommended.

In some gas-filled phototubes, there is a slight tinting of the glass envelope opposite the photocathode. This tinting, which does not materially affect tube operation, is caused by sputtering of cathode material as a result of ion bombardment during the processing and ageing of the tube. Further tinting may occur during long and especially severe operation of the phototube.

References

[1] Cobine, J. D., **Gaseous Conductors,** McGraw-Hill Book Co., New York (1941).

[2] Loeb, L. B., **Basic Process of Gaseous Electronics,** Univ. of Calif. Press (1955).

[3] Engstrom, R. W. and Huxford, W. S., "Time Lag Analysis of the Townsend Discharge in Argon with Activated Cesium Electrodes." **Phys. Rev., 58** (1940).

NEWS AND NEW RELEASES

2N2876

A new vhf power transistor, the 2N2876, has been developed by RCA. This device is a triple-diffused, planar transistor having an interdigitated emitter-base geometry. A combination of high current capability, 2.5 amperes maximum collector current, and high typical gain bandwidth of 200 Mc is attained by providing 0.40 inches of emitter periphery in a device area of only 0.0015 square inches.

A minimum power output of 10 watts at 50 Mc with an efficiency of 70%, and a power gain of 7 db is guaranteed. Performance at higher frequencies is controlled by a power gain test at 150 Mc. This transistor is also 100% tested for secondary breakdown to assure operation in Class A. The 2N2876 extends RCA's transistor power-frequency capability by an order of magnitude, and is finding extensive use in solid state transmitters, both as an output stage and as a driver for varactor multiplier chains.

NEW PHOTOTUBES

Two new multiplier phototubes, one for missile applications and the other for ruby-laser detector systems, have been introduced by RCA Electronic Components and Devices. The 4441A is the first in a new series of RCA multiplier phototubes for which tube ruggedness and reliability are assured by 100% environmental testing. This 10-stage, head-on type is designed to withstand severe environmental conditions for equipment mounted on the structures of missiles launched by high-thrust engines. It is intended for industrial and military nuclear radiation and low-level light detection - and - measurement applications. The 4441A utilizes a special photocathode connection which assures continuous contact with the cathode when the tube is subjected to rough usage.

The spectral response of the 4441A covers the range from about 3000 to 6500 angstroms. Maximum response occurs at approximately 4400 angstroms. The tube, therefore, has high sensitivity in the blue and less sensitivity in the red region of the spectrum.

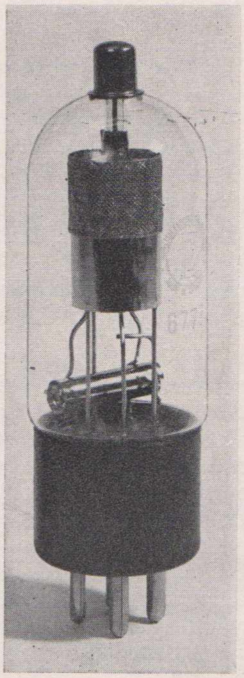
The other new multiplier phototube, the 4459, is a 12-stage, head-on type with a spherical faceplate. This tube, a multialkali-photocathode version of the 7850, is intended for use in near-infrared ruby-laser detector systems. It also may be employed in flying spot scanning equipment, photometry and other applications which require low dark current as well as high sensitivity over the visible and near-infrared regions of the spectrum. High blue sensitivity and other characteristics of the new 4459 make it particularly useful in scintillation counters.

The spectral response of the 4459 covers the range from about 2900 to 8000 angstroms, with maximum response at approximately 4200 angstroms. The response extends beyond the visible region into the blue region on the one end and well into the red and near-infrared regions on the other end.

2N3228 SCR

A new low-cost, all-diffused silicon controlled rectifier, the first unit specifically designed and rated to operate directly from a typical household (USA) 117-volt line, was announced by RCA Electronic Components and Devices. This 5-ampere 2N3228 now makes "proportional power control" commercially possible in several new product areas, according to Ben Jacoby, Manager, Industrial Semiconductor Market Planning RCA Industrial Tube and Semiconductor Division. Proportional control of heat, light and the speed of fractional horsepower motors are prime application areas ideally suited for low-cost power control devices, Mr. Jacoby added.

The light-industrial equipment and appliance manufacturers now have an electronic power-control device that will make the convenience of proportional power control economically feasible for waffle irons, frying pans, light dimmers, power tools, food mixers, blenders, cameras, toys and many other applications. No longer will the housewife have her food mixer or blender stall at its low-speed setting. Adequate power at all speed settings is only one of the benefits derived from SCR power-control devices.



NEW COMPACT EEV THYRATRONS

EEV has introduced two new hydrogen thyratrons, the 6587 and the 6777, which are designed for use in compact equipments where space is at a premium. These tubes are intended for pulse operation at high repetition rates and both incorporate hydrogen reservoirs for long and reliable life. The 6587 is mechanically smaller than, and electrically superior, to the 5C22 and is also tested to a more exacting specification. It also embodies a tetrode type electrode structure, a feature of the EEV 8503 (FX290), which has been shown to enhance long life and reliability characteristics.

6146A/8298A

A new beam power valve which provides more power in new equipment designs and extended tube life in renewal use has been announced by RCA. The new 6146B/8298A will give extended life when used in existing 6146, 6146A and 8298 sockets, according to the announcement. This beam power valve permits higher plate dissipation (35 CW ICAS watts max.) for increased plate current. It also has all the advantages of improved performance made possible by the RCA "Dark Heater." At normal heater ratings, capabilities of the 6146B/8298A are: 85 watts CW output (ICAS) at 60 Mc and 50 watts CW output (ICAS) at 175 Mc. In fixed station use, 6.3 volts is the recommended value for the new valve's "Dark Heater." In mobile service, the valve operates efficiently over a range of heater voltages from 5 volts to 8 volts.

EEV C1149/1

EEV has introduced a new pulse tetrode C1149/1 for use in radar modulators. This valve is the outcome of a prolonged series of tests in the laboratory and at sea.

THE PROBLEM . . . to develop a reliable tetrode which would not only withstand the severe vibrational environment found in a fishing trawler, but also the various forms of malfunctioning which can sometimes be expected in high voltage equipments. The life was to be 3000 to 4000 hours. EEV engineers solved the problem by extending the design of the C1133, which has become known

all over the world for use in marine radars manufactured by various companies.

THE SOLUTION . . . first, the internal electrode assembly was strengthened to reduce the chances of inter-electrode short circuits under vibration. Second, a successful method was developed for forming a satisfactory oxide cathode immune to flaking and peeling. By a modified processing procedure, uniform emissivity was obtained all over its surface and peak emission values were raised.

At this stage, the C1133 was re-named the C1149. The third and last improvement, which has now been completed, was to increase the bulb diameter slightly to promote longer life, replace the tungsten pins at the base with pins of a more pliant nature and thereby avoid glass breakages. Up to the present, ten samples have survived 20,189 hours of operational life under the most arduous conditions without one single failure. The C1149/1 is now specified to withstand intermittent vibrations up to 5g from 20 cycles per second to 1500 cycles per second and each valve is subjected to a minimum shock test of 200g. It has a pulse output power of 330Kw, a maximum anode dissipation of 60 watts at 20Kv and a pulse current of 18 amperes.

EEV HIGHLIGHTS LONDON AIRPORT RADAR

During the period March 1963 to February 1964, trials have been conducted at London Airport with a daylight-viewing radar display incorporating a high-brightness storage tube type E702A, designed and developed by EEV. The prototype equipment, developed by The Marconi Company, together with the original tube, has been running continuously day and night for 8,000 hours with only minor control adjustments and is still functioning satisfactorily.

The bright display, a feature of the E702A, enables air traffic controllers to see easily and clearly, under the high ambient light conditions experienced in control towers, the position of aircraft on the last 10 miles of the approach to land. The trials have proved so successful that serious thought is being given to the installation of fully engineered equipments at London and other major airports throughout the country. The E702A is a 5-inch diameter tube with a typical brightness of 2,150 ft-lamberts. Its stored image can be erased by applying a positive pulse to the storage element and if the pulses are controlled, erase time can be varied for a maximum effective persistence of 10-15 minutes.

* * *

Editor .... **Bernard J. Simpson**

Radiotronics is published twelve times a year by the Wireless Press for Amalgamated Wireless Valve Co. Pty. Ltd. The annual subscription rate in Australasia is £1, in the U.S.A. and other dollar countries $3.00, and in all other countries 25/-.

Subscribers should promptly notify Radiotronics, P.O. Box 63, Rydalmere, N.S.W., and also the local Post Office of any change of address, allowing one month for the change to become effective.

CLOISTER PRESS (W. SHORT), 45 George Street, Redfern


AWV