

RADIO TRONICS

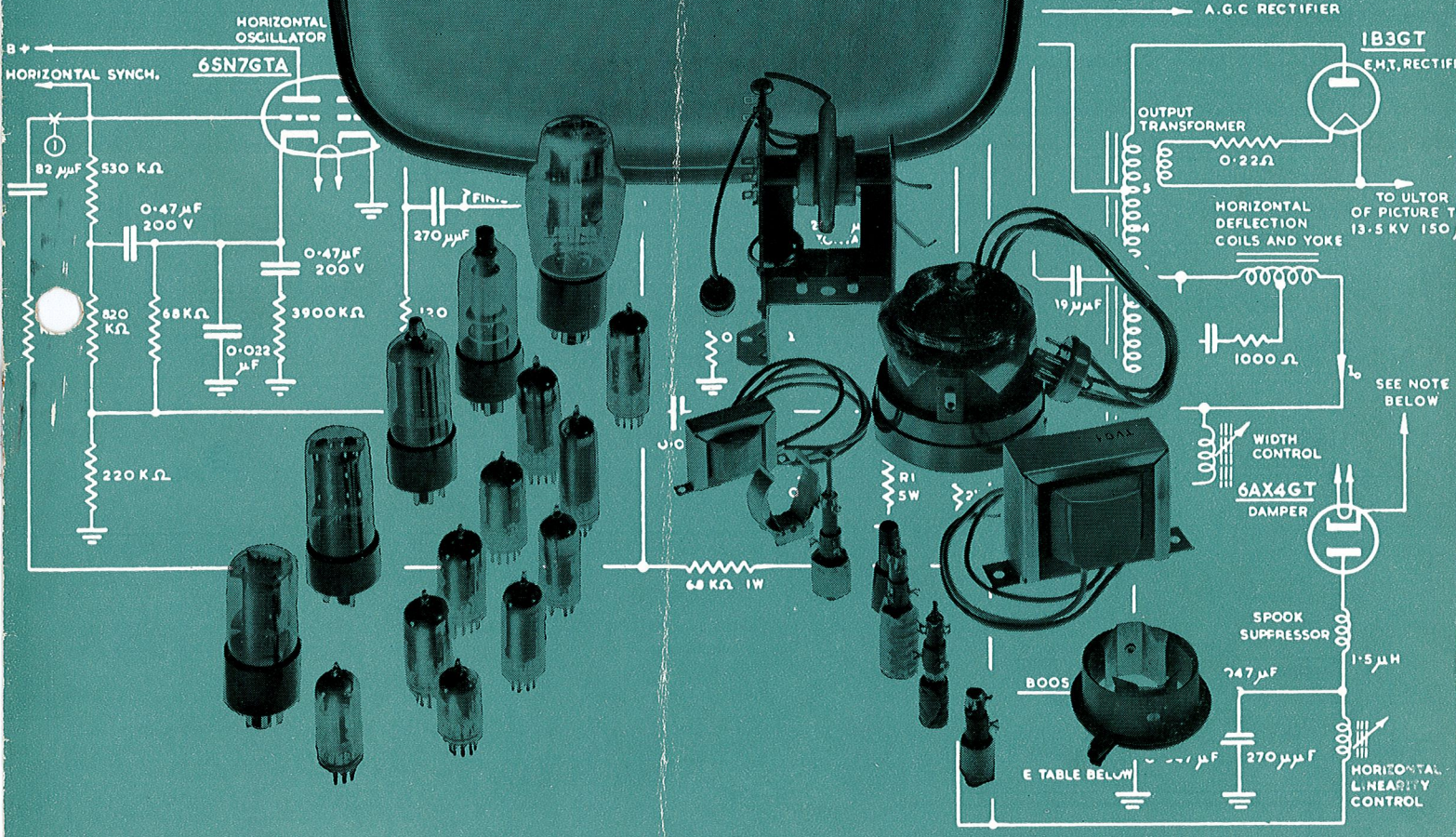
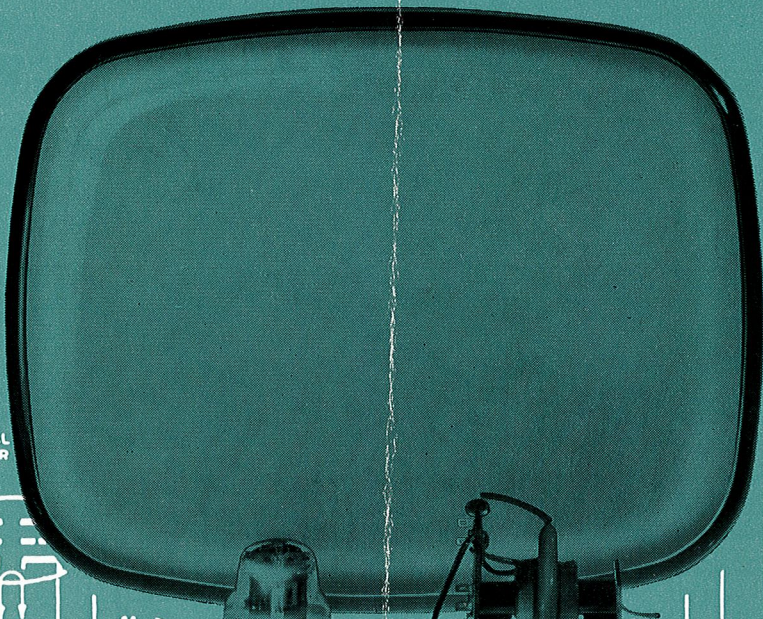
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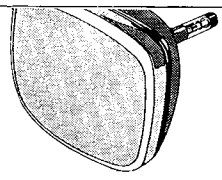
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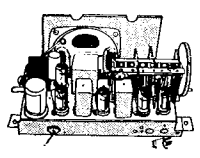
NO. 6



AMALGAMATED WIRELESS VALVE COMPANY PTY. LTD.



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- High Fidelity.*
- Television Engineering, Vol. 3—Waveform Generators.*
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"The Purpose of the Post-Equalizing Pulses in the Television Composite Video Signal"

By W. W. BURNS, A.S.T.C.
(TV Development Section, A.W.A.)

INTRODUCTION.

Although the literature which has been written concerning television is extensive, the explanation for the inclusion of the post-equalizing pulses into the composite video signal has generally been of a vague nature. The purpose of the pre-equalizing pulses has been fully treated in many technical articles, viz., to provide similar conditions from field to field at the start of integration of the vertical sync. block, in receivers which use this method for triggering of the vertical sweep generator. Patchett deals extensively with the difficulties encountered in Britain in obtaining good interlace without the inclusion of the equalizing pulses, pre and post.

The object of this article is to present a brief and simplified explanation of the desirability for the inclusion of the post-equalizing pulses into the composite video signal in regards to interlacing in television receivers. It is not claimed by the author that good interlacing cannot be obtained without the aid of equalizing pulses, but that their inclusion into the composite video signal simplifies the circuitry required in maintaining good interlace over the range of the vertical hold control.

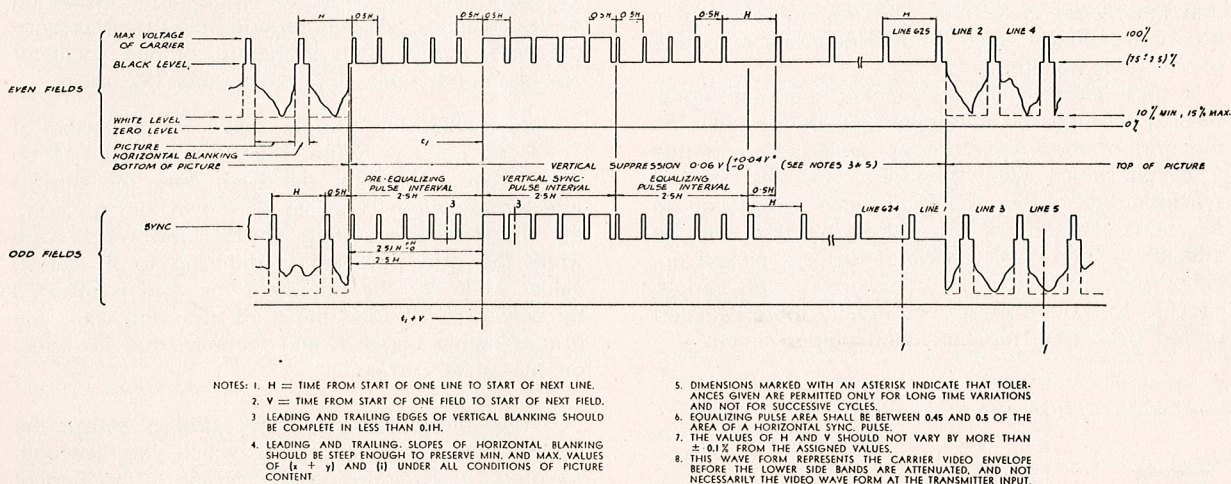


Fig. 1. Video Waveform (from ABCB drawing ZC-9-C sheet 1).

GENERAL.

With the Australian standard composite video signal (Fig. 1), not only must each field contain 312.5 lines for good interlace but it is also essential that the forces producing vertical deflection must also be constant from field to field. This statement broadly means that the perpendicular distances traversed by the spot when being deflected from the top to the bottom and back to the top again of the picture tube screen, must be equal for each field.

This is dependent on a number of factors, among which is the stability of the grid drive to the vertical output valve. This driving voltage is usually a sawtooth voltage developed across a capacitor in the plate circuit of the vertical sweep generator. For the purpose of this explanation the waveform across

the capacitor may be regarded as a pure sawtooth waveform as shown in Fig. 2 (a).

Since this controls the vertical deflection; during the periods AB and CD the spot is being deflected vertically downwards for two successive fields, the respective retrace periods being BC* and DE*. Both scan and retrace voltages are constant as required. However, if an effect takes place which produces a difference between fields of

- (1) the positive peak amplitude,
- (2) the negative peak amplitude, or
- (3) both peak amplitudes together,

then the result could be pairing of the interlace. In some circumstances several lines on alternate fields could be displaced.

* In actual practice the retrace period of the vertical output stage is governed by the circuit constants. However, this does not invalidate the argument.

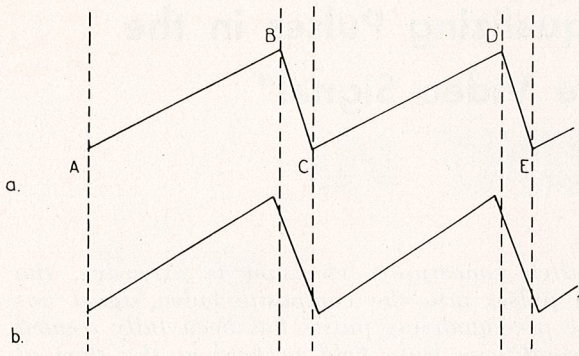


Fig. 2. (a) constant sawtooth voltage;
(b) irregular sawtooth voltage.

BLOCKING OSCILLATOR.

A blocking oscillator (Fig. 3) to generate a voltage of sawtooth waveform across C_4 will be considered as an example in illustrating the most important purpose of the post equalising pulses. It is assumed that the reader is familiar with the operation of a blocking oscillator when used to generate a voltage of sawtooth waveform.

It must also be borne in mind that any signal passing through the integrator will be passed onto the grid of the vertical sweep generator, and that a circuit which gives integration to the vertical synchronising block must give some integration to the horizontal (H) pulses. Also, the grid will be affected if any stray voltage pulses are picked up, either by radiation or by coupling through various circuits. Unwanted pulses are usually those radiated or fed back from the horizontal output circuit.

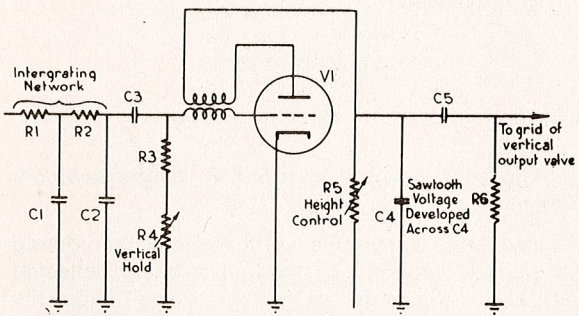


Fig. 3. Blocking Oscillator Circuit.

POST EQUALIZING PULSES.

The grid waveform of a blocking oscillator is shown in Fig. 4; during the active period it may take the shape as shown in Fig. 5 (continuous line) where line AB represents the cut-off potential of a valve. During the period represented by T_a the grid is active and plate current will only flow during this period. To consider a case approaching an extreme condition, and to simplify the explana-

tion, assume that triggering of the oscillator for both fields takes place at a time corresponding to the leading edge of the last vertical sync. pulse. Referring to the standard signal and assuming that

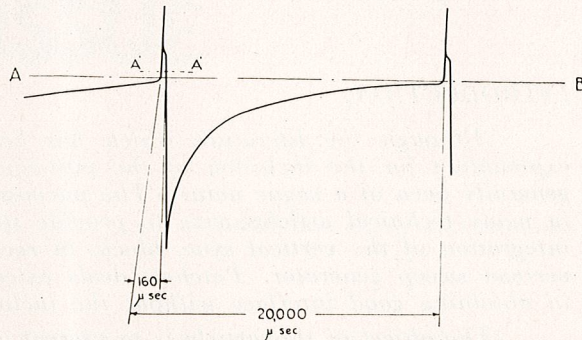


Fig. 4. Waveform of blocking oscillator.

the horizontal sync. pulses are retained at intervals of 1H; but no post equalising pulses; then in the even field, the H pulse occurs 32 μ secs, the second 96 μ secs, and the third 160 μ secs after triggering of the oscillator, while in the odd field the occurrence of these pulses after triggering of the oscillator would be 64 μ secs, 128 μ secs, and 192 μ secs.

Assume that the grid is active for a period of $T_a = 165 \mu$ secs, a figure which is considered to be quite good. Then for the even field the grid is affected by three H pulses as shown at Fig. 5 (a), the third pulse occurring on the downward slope while the grid potential is reducing to its cut-off value, while for the odd field the grid is affected by only two H pulses, the third occurring while the grid is highly negative and therefore has no effect on the plate current.

During the period while the grid is active, the capacitor C_4 (Fig. 3) across which the sawtooth waveform is generated, is discharging — the amount of discharge being dependent, among other factors, upon the grid potential. Therefore, if from field to field while the grid is active, the waveform of the potential varies due to H sync. pulses being fed through the integrator or by stray pickup, then the discharge of the capacitor will vary from field to field and cause pairing of the interlace. The pulses occurring while the grid is at a positive potential with respect to cathode, have very little effect, the grid being at a low impedance to the fed-through synchronising pulses, but it is while the grid has a high impedance, i.e. between zero volts and the cut-off bias, and still controls the plate current that any variation of the grid waveform in this region has a marked effect on the quality of the interlace.

This has been explained from a simple point of view. In addition to the above effect, the plate potential at the correct triggering point, i.e. at intervals of 20,000 μ secs, will not be constant, thus causing a time variation in the duration of each

field. When the post equalising pulses are inserted, then if there is any effect on the grid by these pulses, the conditions are identical from field to field and the voltage waveform across the capacitor

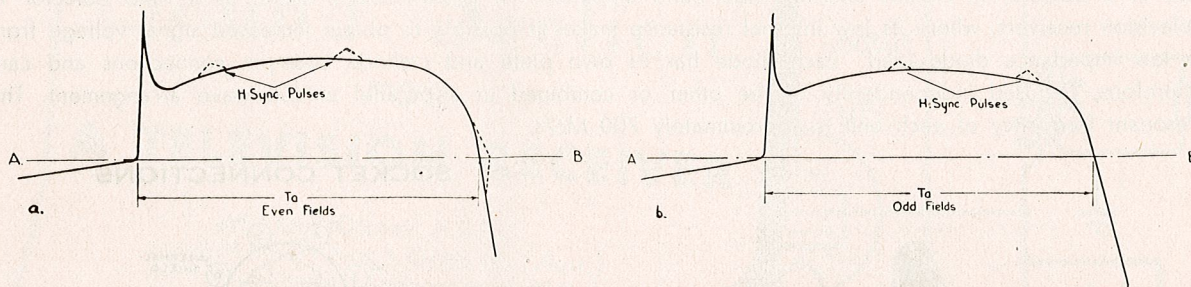


Fig. 5. Relative position of H pulses in odd and even fields during the active period of the blocking oscillator (i.e., $A-A^1$ of Fig. 4).

C_4 (Fig. 3) generating the sawtooth voltage which governs the vertical deflection of the spot will be the same for both fields.

Another point to be considered is that the post equalising pulses are terminated approximately 160 μ secs from the trailing edge of the last vertical sync. pulse, the first H pulse appearing approximately 192 μ secs and 224 μ secs in the odd and even fields, respectively, after the termination of the vertical sync. block. Therefore, if the grid of the vertical sweep generator is active for a longer period than 192 μ secs inequality between successive fields could once again occur. Generally, good results are obtainable if the active period of the grid is kept within 200 μ secs, as it is usual for the triggering point of the oscillator to be adjusted to take place at approximately mid position of the vertical sync. block.

EFFECT OF H PULSES FEEDBACK FROM THE OUTPUT.

However, the equalising pulses cannot correct for the effect of any H pulses feedback from the horizontal output stage either by circuitry or radiation. This problem has to be dealt with separately. The effect of these unwanted pulses is to cause a variation of the interlace as the vertical hold control is adjusted through the control range.

This may be understood by once again referring to Fig. 5 where the H pulses as shown may be considered, not as fed through the integrator but as pulses due to feedback from the horizontal output stage.

In the position as shown, when the grid is active, differences occur from field to field and poor interlace will result. When the hold control is varied so that triggering of the vertical sweep generator takes place previous to the last vertical sync. pulse, then the positions of these H pulses vary in time relative to the active period of the grid. This results in variation of the charge and discharge of the capacitor forming the sawtooth voltage and thus causing variation of the interlace.

It may be added that if the equalising pulses were not inserted, then H. pulses fed through the integrator would also cause a variation of the interlace with rotation of the vertical hold control.

The purpose of the post equalising pulses has been made illustrating their effect with regard to a blocking oscillator. The same reasoning applies with respect to other vertical sweep generators such as multivibrators, etc.

EXAMPLE WITH NO EQUALIZING PULSES INSERTED.

Referring once again to Fig. 5, consider the active period of the grid as 165 μ secs. The odd field is being affected by two H pulses. On the even field the third fed-through H pulse appears just before the time that the grid is about to become inactive; its effect is to keep the grid active for a longer period than 165 μ secs, e.g., say 165.1 μ secs. The capacitor in the plate circuit of the blocking oscillator is charging and discharging, while the grid is inactive and active respectively every 20,000 μ secs.

If the discharge is considered as being linear (reasonably justified) then the discharge between fields varies by 0.1 in 165 or approx 0.06%. In a raster, if the lines and the spacing between the lines are considered to be of equal width, then these lines are 1/1250 of raster height h . For pairing, a variation from field to field of 1/1250 h or 0.08% is all that is required. Thus in the above example a condition exists approaching pairing.

CONCLUSION.

With practical vertical sweep generating circuits as used in present-day television receivers the inclusion of the equalising pulses, pre and post, are a desirable part of the composite video signal to facilitate the maintaining of good interlace over the full range of the vertical hold control.

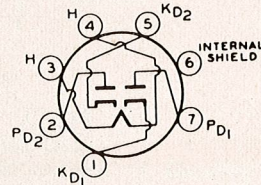
REFERENCE.

G. N. Patchett—“A critical review of synchronising separators with particular reference to correct interlace”. Journal of the British Institution of Radio Engineers, 14th May, 1954, p. 191.

MINIATURE TWIN DIODE

The Radiotron 6AL5 is a miniature twin diode which, because of its high perveance, is suitable for use as a detector in circuits utilising wide band amplifiers. It is particularly useful as a ratio detector in television receivers, where its low internal resistance makes it possible to obtain increased signal voltage from a low impedance diode load. Each diode has its own plate and cathode base-pin connections and can, therefore, be used independently of the other or combined in a parallel or full wave arrangement. The resonant frequency of each unit is approximately 700 Mc/s.

SOCKET CONNECTIONS



(bottom view)

- Pin 1 Cathode of Diode No. 1
- Pin 2 Plate to Diode No. 2
- Pin 3 Heater
- Pin 4 Heater
- Pin 5 Cathode of Diode No. 2
- Pin 6 Internal Shield
- Pin 7 Plate of Diode No. 1



GENERAL DATA.

Heater voltage 6.3 volts
 Heater current 0.3 amp.

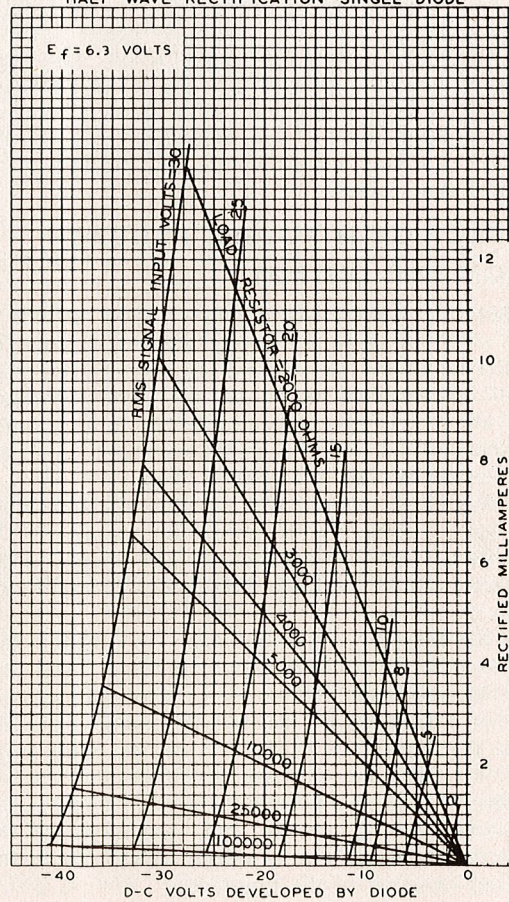
Maximum Ratings:

Peak inverse plate voltage 330 volts
 Peak plate current per plate 54 mA
 D.C. output current per plate 9 mA
 Peak heater-cathode voltage:
 Heater negative with respect to cathode 330 volts
 Heater positive with respect to cathode 330 volts

Typical Operation as Half-Wave Rectifier:

(The two units may be used separately or in parallel)
 A.C. plate voltage per plate (r.m.s.) ... 117 volts
 Minimum total effective plate-supply impedance per plate 300 ohms
 D.C. output current per plate 9 mA

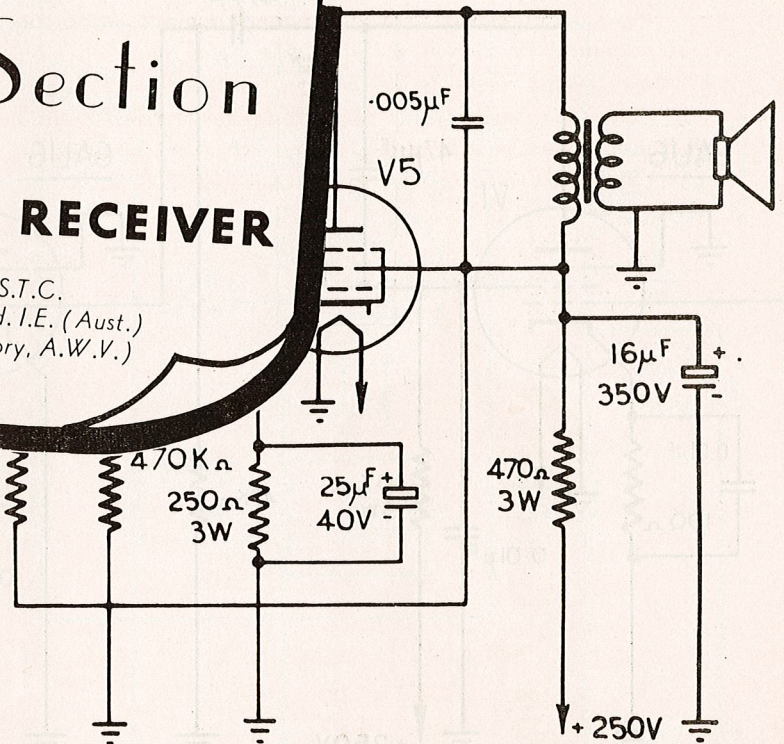
AVERAGE CHARACTERISTICS HALF-WAVE RECTIFICATION-SINGLE DIODE



Sound Section FOR A TELEVISION RECEIVER

by R. Darnell, A.S.T.C.
A.M.I.R.E. (Aust.), Grad. I.E. (Aust.)
(Applications Laboratory, A.W.V.)

A typical sound i-f and audio amplifier circuit is presented with a discussion of its performance. Coil and circuit details are given as well as the alignment procedure and test specifications.



INTRODUCTION.

The use of frequency modulation for television sound was originally adopted in 1941 because it allows a higher signal to noise ratio in the very high frequency transmitting band. However, the advent of the intercarrier system has made F-M sound even more desirable because it simplifies the separation of the video signal and the sound signal at the video detector. The United States and continental European standards specify F-M sound but Great Britain and France use A-M sound. Australia has followed the former lead combining the improved signal to noise ratio with the advantages to receiver manufacturers of intercarrier receivers. These advantages are that fewer valves are required and that since the intercarrier frequency is dependent on the spacing between the transmitted sound and video carriers alone, the need for highly stable and non-microphonic local oscillator operation is removed.

The ratio detector has been most commonly used as an F-M detector for television receivers in the past although other methods of detection are now rapidly gaining favour overseas. Much has been written about ratio detector transformers and their operation.^{1,2,3} The main reason for their popularity has been their ability to suppress amplitude modulation of the sound carrier without the use of a limiter. However, as this suppression is not normally considered sufficient the valve driving the ratio detector transformer is usually arranged to act as an amplifier at low levels but as a limiter for larger input signals.

INTERCARRIER BUZZ.

If, during any portion of a field, the ratio of the video carrier level to the sound carrier level falls below a value of about 2, excessive amplitude modulation of the sound carrier will occur. This is evident in the audio system as a buzz at the field frequency rate. The sound carrier is normally attenuated about 26 db relative to the video carrier in the i-f amplifier. It is already down 7 db because of the 5 to 1 video to sound power ratio of the transmitter. This gives a total of 33 db difference which means that the sound carrier level is normally 2.2% of the video carrier level. Now the minimum white level has been fixed at 10% in the Australian system

and provided this level is adhered to by the transmitting stations at all times, the attenuation of the sound carrier in the i-f amplifier has to be decreased by 7 db before the level of the sound carrier is 5%, giving a ratio as defined above of 2:1. Hence, if the attenuation of the sound carrier in the i-f amplifier is less than 19 db, intercarrier buzz may become evident when the picture contains a white portion. Another form of intercarrier buzz is described in the next section.

SOUND TAKE-OFF TRANSFORMER.

The 5.5 Mc/s frequency modulated beat note which becomes the sound carrier is present in the

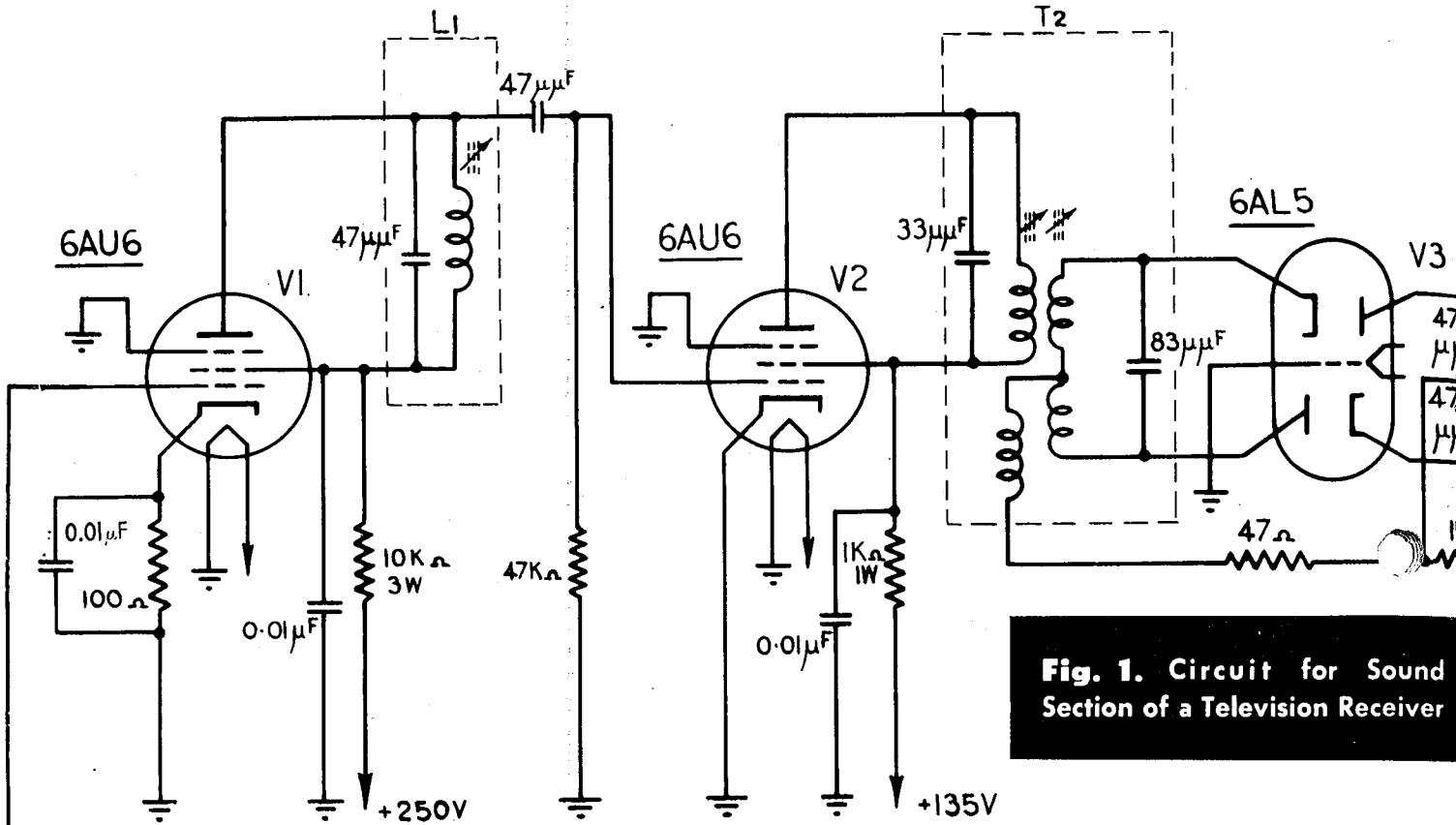
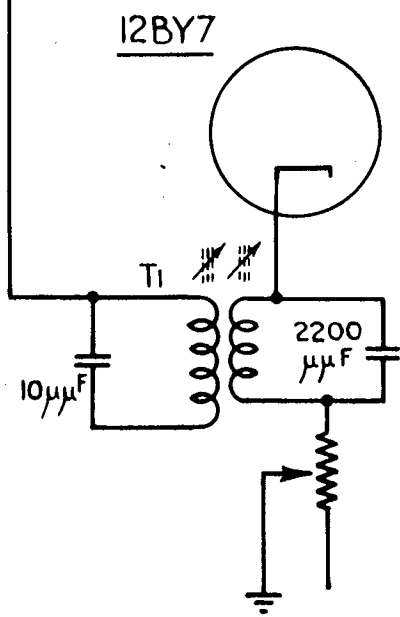


Fig. 1. Circuit for Sound Section of a Television Receiver



output of the video detector and can be separated from the video information by a tuned circuit either in the detector output circuit or in the video amplifier plate or cathode. Taking the sound carrier from the video amplifier can cause the following trouble. If the a.g.c. system of the receiver is not good enough, the synchronising information could cause the video amplifier to become cut off both at the line frequency and at the field frequency. If only the tips of the synchronising pulses cause this cutting off, the picture will remain synchronised but the sound carrier will be interrupted and will present intercarrier buzz and

possibly a line frequency whistle if the audio system is good enough. This trouble is more likely to be apparent where simple a.g.c. systems are used. The circuit described here is designed to work in a receiver which has an adequate a.g.c. system.

The take-off point for the circuit described here is a high impedance winding tuned to 5.5 Mc/s

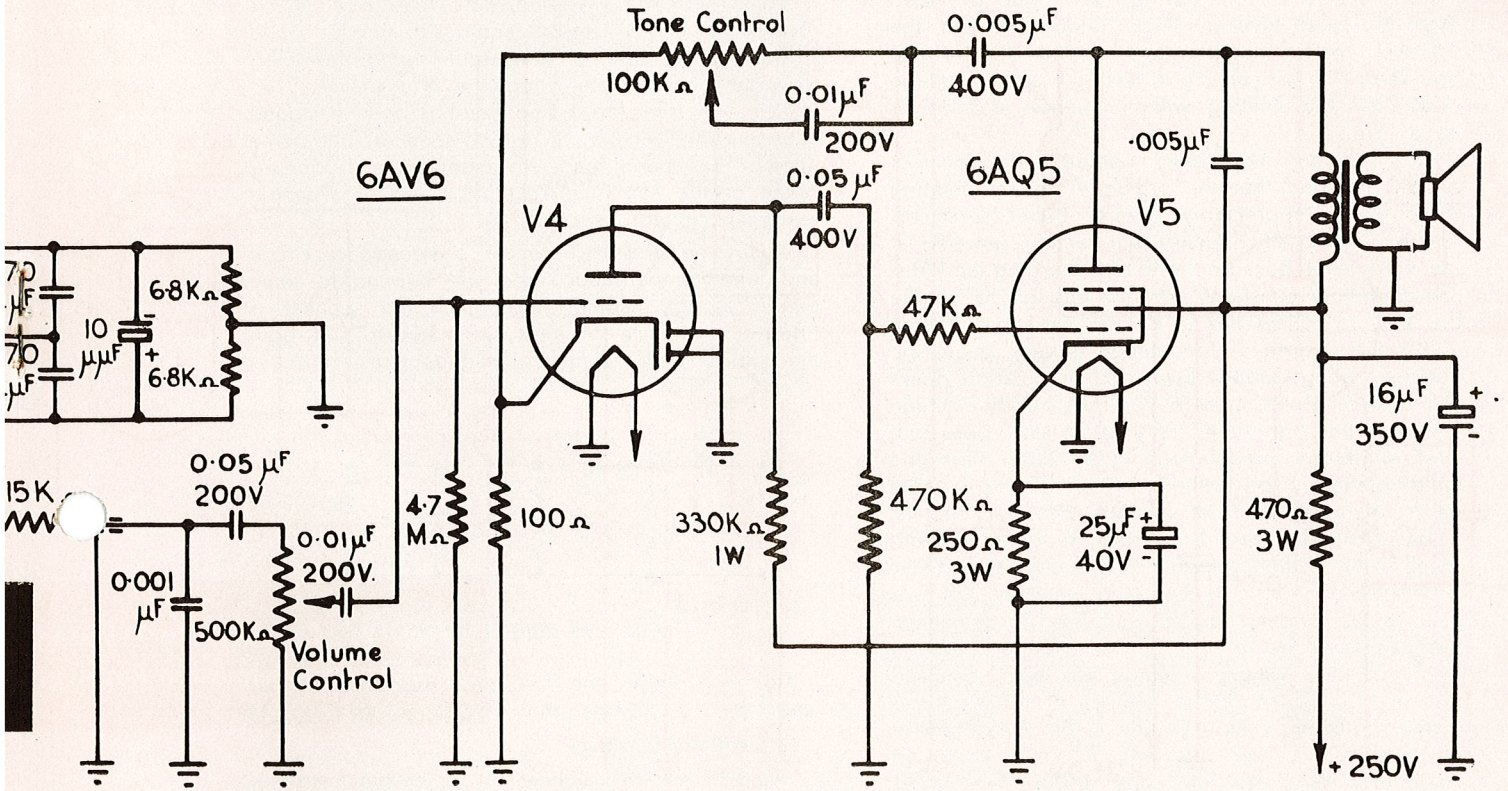
and coupled to the trap in the video amplifier cathode.⁴ This system does not utilise the gain of the video amplifier but provides a step up of two times from the video amplifier grid. This step up could be obtained from a tapped transformer in the video detector output circuit, but it is more convenient to use the cathode trap and save complicating the detector circuit.

DESCRIPTION OF CIRCUIT.

The circuit is shown in Fig. 1.

Sound I-F Amplifier Stage.

This stage is a single tuned 6AU6 amplifier fed from the secondary of the sound take-off transformer and has a gain of 40. By using a higher inductance and smaller tuning capacitor the gain of this stage could be almost doubled but considerable detuning is then experienced when the input level changes. This detuning is caused by input impedance changes in the following (limiter) stage as the level changes and can be quite high. As the gain requirement of this stage is not extremely high, as will be seen later, it is desirable to reduce this detuning effect at the expense of some stage gain as has been done in this circuit. The detuning for a change in input from 3-1 mV to 100 mV was found to be 80 Kc/s whereas for a higher impedance circuit tuning with 12 µµF it was 200 Kc/s.



When measurements are made without the take-off transformer connected a d.c. grid return must be provided for this stage—say a 100,000 ohm resistor. The total bandwidth must be about four times the maximum deviation frequency which for the Australian system is 50 Kc/s thus giving a requirement of 200 Kc/s. If a vacuum tube millivoltmeter with a diode probe suitable for working at a frequency of 5.5 Mc/s is available, the bandwidth of this stage can be measured directly by noting the change in input frequency between the points where the output measured at the plate (with the V.T.V.M.) is 3 db down. If such a meter is not available a diode detector as shown in Fig. 2 can be connected to the plate of the following stage. (The 220 ohm resistor which damps out the effect of the ratio detector primary could be larger, say 1000 ohm. The lower value was chosen because such a detector is used in the video i-f amplifier alignment.)⁵ The input frequency can be varied as before to find the bandwidth between 3 db points measured with a direct voltmeter at the detector output or a sweep generator input can be used, the detector output being observed on an oscilloscope and the bandwidth determined by use of a marker generator. These measurements should be done at a low input level, say 1 mV, to eliminate limiting at the grid of the following stage. The stage in the amplifier described here has a bandwidth of 200 Kc/s.

Limiter Stage and Ratio Detector.

As mentioned previously the limiter stage acts as an amplifier at low input levels. The level at which limiting commences depends on the grid base of the valve. The more remote the cut-off, the higher will be the input level at which limiting commences. The limiting level can be reduced by shortening the grid base. This is achieved by reducing the screen voltage of the valve. To enable the valve to act as an amplifier for low levels the screen voltage in this case is kept fairly high at 120 volts. The valve then begins to limit at an input level of from 0.6 to 1 volt at its grid. The high gain of the preceding stage should ensure that the input is always maintained well above this level except in very extreme conditions.

Limiting is achieved by using a coupling time constant in the grid circuit which is long compared with the carrier frequency but very short compared with any amplitude modulation likely to be superimposed on the carrier. The 47 μF and 47,000 ohm combination used has a time constant of 2.2 microseconds. A bias is developed by grid current which flows on positive peaks of the input signal and causes the coupling capacitor to charge. The charging time constant is the capacitance times the input impedance of the valve during this grid current flow and is hence smaller than the discharge time constant of 2.2 microseconds. The bias developed is just large enough to clamp the positive peaks at the zero bias level. The plate

current of the valve then varies from that for zero bias to cut-off current during each cycle provided that the input level is large enough to cut the valve off on the negative swing. Then for a valve operating with a smaller grid base the cut-off voltage is smaller and limiting will commence at a lower input level.

Now, for any amplitude modulation of the input signal, provided that the period of the modulating frequency is appreciably greater than 2.2 microseconds, the positive peaks will remain clamped at the zero level because both the charging and discharging time constants are fast enough to enable the bias to follow the amplitude variations. Then, unless the degree of modulation is very large, the negative swing during downward modulation peaks will still carry the grid beyond its cut-off voltage and so the amplitude modulation is suppressed. Any amplitude modulation which does pass the limiter stage will be further reduced by the variable diode damping inherent in ratio detector circuits.

The ratio detector transformer is quite conventional with a high impedance primary and bifilar wound secondary. The coupling between the primary and secondary has been adjusted for maximum sensitivity and the ratio, r , of the secondary voltage to the tertiary winding voltage has been adjusted to unity by choice of the number of tertiary turns giving optimum A-M rejection. The measurement of the ratio r is described in the Radiotron Designer's Handbook on page 1102. The circuit showed an improvement in sensitivity of 4 db and in A-M rejection of 20 db when the spacing between windings was changed from 0 to $\frac{1}{8}$ " and r altered from 1.4 to 1.

The measurement of A-M rejection was done in a manner different from that normally used,⁶ but one which is very simple if there is available a signal generator which can be set to produce either frequency or amplitude modulated output. The results, of course, are relative and cannot be compared with those obtained using other methods. With the circuit correctly aligned (the alignment procedure will be described later) feed a 5.5 Mc/s signal, frequency modulated to a deviation of 15 Kc/s (which is approximately 30% of the maximum deviation), to the grid of V_1 and adjust the input level to some point well below the limiting level and note the audio power (50 mW is a convenient level). Now switch from F-M to A-M and modulate to a depth of 30% and note the drop in audio output power. This will give the A-M rejection figure as mentioned above. A figure of 27 db was obtained for this circuit at an input level of 450 microvolts. As the input will normally be well above the limiting level the figure will be much greater than this in practice.

Audio Amplifier.

Television transmissions do not have the 5 Kc/s bandwidth restriction as in radio broadcasting and can utilise the full audio bandwidth. This means that a good audio section in a television receiver can produce some good quality sound on normal

programmes. However, the receiver manufacturer is concerned with keeping down the cost of his receiver as much as possible and has to limit his sound section appreciably. It does seem futile, however, to waste the extra bandwidth that is available. With this in view, the audio amplifier has been designed economically with a 6" x 9" speaker but some attempt has been made to extend the frequency range of the amplifier a little by use of a simple feedback circuit.

It must be stressed here that the feedback circuit and tone compensation used will depend to some extent on the cabinet into which the speaker is fitted and adjustment must be made to suit both the cabinet and the maker's requirements.

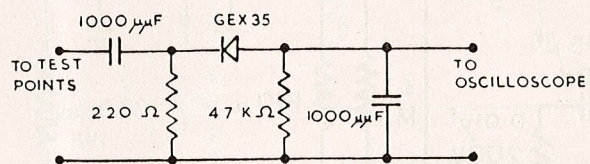


Fig. 2. Detector probe.

Figure 3 shows the frequency response of the amplifier:—

- (i) without feedback;
- (ii) with feedback control set for maximum treble;
- (iii) with feedback control set for minimum treble.

Curve (ii) is the interesting curve. It will be seen that at 10 Kc/s the response is down 11 db relative to 500 c/s. This is too far down for good quality reproduction but can be effectively lifted by altering the de-emphasis network which precedes the audio amplifier. The use of pre-emphasis and de-emphasis is briefly explained here but is generally well known. Since the noise introduced in the early stages of the television receiver is mainly concentrated in the higher audio frequency region and the normal high frequency components in a typical programme are of much lower magnitude than the lower frequency components, the higher frequencies can be boosted relative to the low frequencies in the transmitter (pre-emphasised) and attenuated again in the receiver (de-emphasised) with the subsequent advantage that the noise introduced in the receiver is attenuated in the de-emphasis network with the wanted high frequency components resulting in a much improved signal to noise ratio. The time constant of both the pre-emphasis and de-emphasis networks in the Australian system is 50 microseconds.

Due to the shortcomings of the audio amplifier there is already some attenuation of the high frequencies which provides part of the de-emphasis required and which can be allowed for by using less de-emphasis in the R-C filter. A 15 microsecond time-constant has been used giving a lift of 5 db at 10 Kc/s over that which would be present using a 50 microsecond de-emphasis network. This effect can be seen in Figure 4 which shows three curves each obtained by measuring the current in

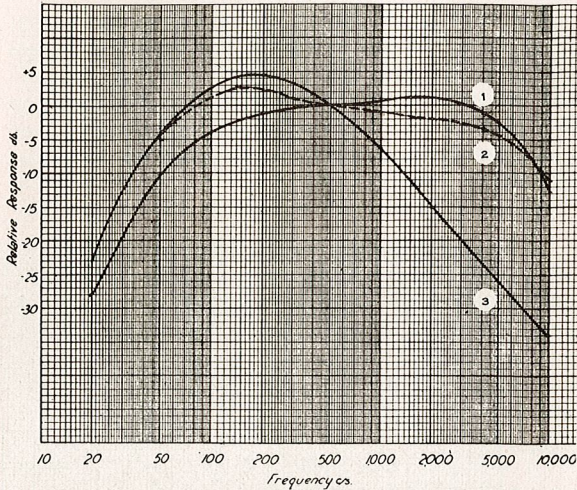


Fig. 3. Frequency response of audio amplifier
 (i) without feedback
 (ii) for maximum treble
 (iii) for minimum treble.

the voice coil of the speaker and plotting it as a function of the frequency used to modulate the 5.5 Mc/s input signal.

ALIGNMENT PROCEDURE:-

Connect a signal generator to the grid of V_1 and feed in a 5.5 Mc/s unmodulated carrier. Connect a meter of impedance at least 10,000 ohms per volt to read the direct voltage developed across either of the two 6,800 ohm load resistances. An input voltage of about 5 millivolts should produce a direct voltage of 3 to 4 volts across each of the resistors. Tune the coil L_1 for maximum d.c. output. Then tune first the primary and then the secondary of T_2 for maximum d.c. output. Now connect the meter across the de-emphasis capacitor and, using

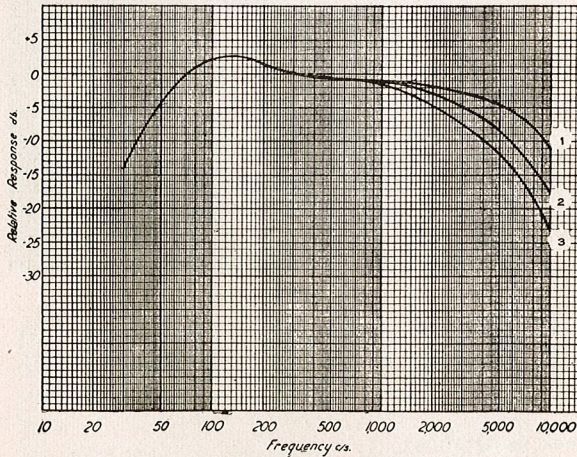


Fig. 4. Overall response showing effect of de-emphasis
 (i) no de-emphasis
 (ii) 15 μ secs de-emphasis
 (iii) 47 μ secs de-emphasis

the most sensitive range of the meter, tune the secondary carefully for zero direct voltage on the meter. (Note that the output at this point will pass from positive to negative or vice-versa rather quickly as the secondary is tuned slightly from one side of the correct position to the other.) The circuit is now aligned correctly.

To align the sound take-off transformer T_1 the 5.5 Mc/s signal is fed to the grid of the video amplifier through a 1000 μ F capacitor and each winding tuned for maximum voltage across either of the diode load resistors.

In aligning both L_1 and T_1 the input level must be below that required for limiting to commence.

MEASUREMENTS.

The measurements recorded here were made both as a check on the performance and as a guide to readers who may wish to check the circuit themselves. Many of the measurements are made without the feedback connected mainly because of simplicity, but also to provide a means of comparison for those who wish to modify the feedback circuit.

Sensitivities For Each Stage.

Measurements of the input to each stage for 50 mW audio output were taken without the feedback connected. These are listed below:—

Sensitivity	Input to Grid of	Signal
900 mV	6AQ5 (V_5)	1000 c/s
14 mV	6AV6 (V_4)	1000 c/s
18 mV	6AU6 (V_2)	5.5 Mc/s; $\Delta f = 15$ Kc/s
450 μ V	6AU6 (V_1)	5.5 Mc/s; $\Delta f = 15$ Kc/s

Feedback at 1000 c/s is 7 db with feedback control in maximum treble position and 12 db for minimum treble.

Input Versus Output Curve.

The limiting action of V_2 can be seen in Figure 5 which shows the ratio detector output plotted as a function of the input to the grid of V_1 . The output was actually measured at the 6AQ5 plate on a power output meter but the audio signal was attenuated in 10 db steps at the volume control each time the output reached 50 mW and the 10 db added to all subsequent readings. The limiting level can be obtained readily from this curve. Also, if a high impedance a.c. voltmeter is not available the curve can be used to determine the maximum audio output voltage from the ratio detector.

This is done as follows:—

- Input to V_4 grid for 50 mW output = 14 mV
- Increase in ratio detector output over this level = 36 db from curve
- Hence maximum audio output from ratio detector = 14 mV + 36 db = 880 mV

The actual voltage at which the second stage begins to limit cannot be measured accurately by feeding a signal into the grid because the different source impedance will effect the result. As well

as this, not many signal generators have sufficient output for this purpose. However, the level can be obtained from the curve as follows:—

$$\begin{aligned} \text{Gain of } V_1 \text{ from previous measurements} &= \frac{18 \times 10^{-3}}{450 \times 10^{-6}} \\ &= 40 \end{aligned}$$

Input to V_1 at which limiting is fully effective
= 45 mV from curve

Therefore, the input to V_2 at which limiting is fully effective
= 45×40 mV
= 1.8 volts

This level could be reduced by working V_2 at a lower screen voltage, as mentioned previously, but this would result in reduced sensitivity.

The maximum voltage output from the sound take-off transformer will normally be less than 100 millivolts and hence the need for the first stage is apparent where good sound reception is required particularly in weak signal areas.

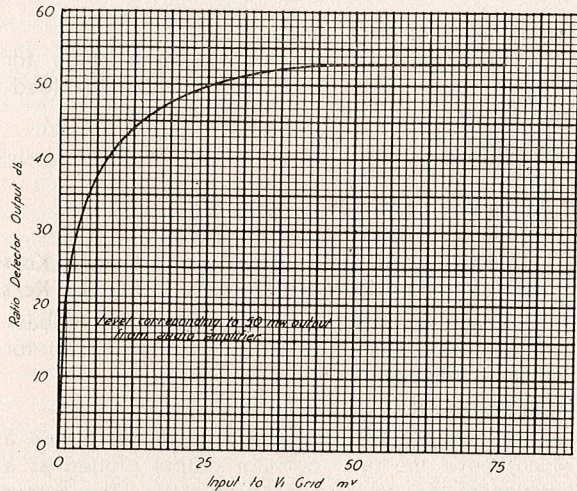


Fig. 5. I-F input versus A-F output (deviation 15 Kc/s; feedback disconnected).

Distortion Measurements.

For all distortion measurements the signal generator used was externally modulated by a low distortion oscillator. The distortion introduced in the modulation process was found to be reasonably small. Each measurement was done with the feedback control set for maximum treble to produce the worst conditions. The input was 100 mV in each case but the distortion was found to be independent of input level. The distortion was measured on a distortion meter connected across the voice coil.

Figure 6 shows the distortion as a function of deviation. The output power of the audio amplifier was limited to 50 mW by the volume control during the test to ensure that the increase in distortion measured was due to the ratio detector circuit alone.

The maximum deviation of the transmitted carrier is 50 Kc/s. At this deviation the distortion is less than 2% ensuring good reproduction of "peaks". The distortion is still below 10% for deviations up to 125 Kc/s thus ensuring good reproduction even when the ratio detector is detuned up to about 100 Kc/s.

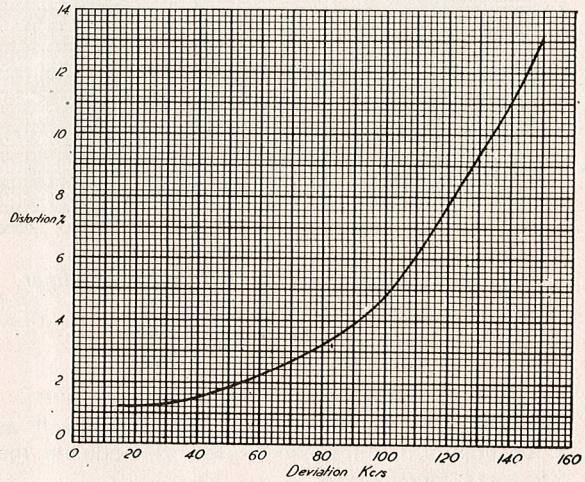


Fig. 6. Distortion versus deviation (input 100 mV, output 50 mW, modulating frequency 1 Kc/s).

Figure 7 shows the distortion as a function of output power measured at the plate of the 6AQ5. The deviation was maintained at 15 Kc/s during these measurements and the output power controlled by the volume control. At a level of 2 watts the distortion was only 2.5% which, combined with the distortion at large deviations still ensures quite good reproduction.

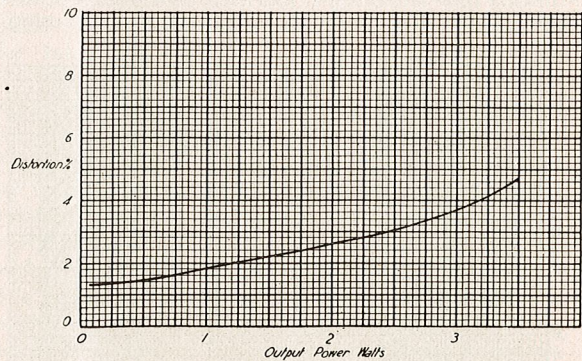


Fig. 7. Distortion versus output power (input 100 mV, deviation 15 Kc/s, modulating frequency 1 Kc/s).

Peak Separation.

The peak separation can be measured by one of three different methods. The first and simplest is to observe the audio output from the ratio detector on an oscilloscope while increasing the deviation of the 5.5 Mc/s input signal. When the

audio output waveform becomes flattened on both peaks the deviation is equal to half the peak separation. Using this method, the peak separation measured was 200 Kc/s.

If this method cannot be used due to lack of a suitable signal generator, the response of the ratio detector can be plotted in the following manner:—

Feed a 5.5 Mc/s carrier into the grid of V_1 at a level above the limiting level—say 100 mV. This high level is used to eliminate the effect of the first stage. Measure the direct voltage across the two diode load resistors in series. Connect a battery equal to this voltage across these two resistors of such a polarity that the voltage will be maintained constant. Now vary the input in 20 Kc/s steps from 5.3 Mc/s to 5.7 Mc/s and note the direct voltage across the de-emphasis capacitor each time. By plotting the output obtained in this manner against the input frequency the response curve of the ratio detector will be obtained and the peak separation can be measured. The response obtained is illustrated in Figure 8. It will be noted that the peak separation is 225 Kc/s.

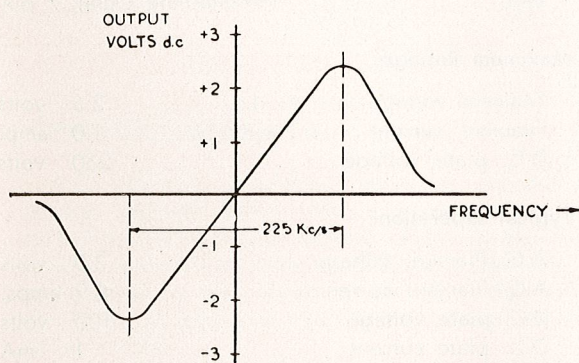


Fig. 8. Ratio detector response showing peak separation (10 mV input to grid V_1).

The third method is to feed a sweep generator into the grid of the first stage with a frequency deviation of at least ± 200 Kc/s at a centre frequency of 5.5 Mc/s. By observing the voltage across the de-emphasis capacitor on an oscilloscope and feeding in separate markers to the amplifier, the peak to peak separation can be measured.

It is interesting to note that as the signal input level is increased up to the limiting level, the diode damping of the secondary will be heavier and the peak separation will increase. However, above this level the damping, and hence the bandwidth, of the ratio detector will remain almost constant.

DETAILS OF COILS.

All formers $\frac{1.7}{6.4}$ " outside diameter.
All cores Vitriplex.
Cans $\frac{3}{4}$ " square.

Radiotronics

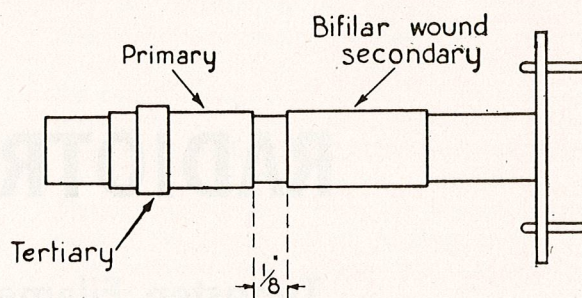


Fig. 9. Ratio detector transformer.

Sound Take-Off Transformer—T1.

Primary	6 turns	18 S.W.G. Tinned copper wire.
		Length of winding $\frac{3}{8}$ ".
		Tune with 2,200 $\mu\mu\text{F}$.
Secondary	60 turns	30 S.W.G. Cotton covered enamelled copper wire.
		$\frac{1}{8}$ " Pie winding.
		Tune with 10 $\mu\mu\text{F}$.
Spacing	$\frac{1}{4}$ "	Both windings tunable.
L_1	65 turns	34 B & S enamelled copper wire — close wound — tunable.
		Tune with 12 $\mu\mu\text{F}$ capacitor.

Ratio Detector Transformer—T2.

		(See Figure 9.)
Primary	63 turns	34 B & S enamelled copper wire — close wound.
		Tune with 33 $\mu\mu\text{F}$ capacitor.
Secondary	20 turns	Bifilar wound.
		One winding using 33 B & S nylon covered copper wire and the other using 34 B & S bare copper wire.
		Tune with 82 $\mu\mu\text{F}$.
		Spacing $\frac{1}{8}$ ".
Tertiary	8 turns	33 B & S nylon covered wire wound $\frac{1}{8}$ " from outside end and directly on primary.

ACKNOWLEDGMENTS.

In conclusion I wish to acknowledge the work done by Mr. E. Pieter who assisted in the development of this circuit and carried out the various measurements and to thank him and other members of the Applications Laboratory staff who have assisted in the preparation of this material. I also wish to thank Mr. J. Patterson, of Bondi, who supplied the coils used in this circuit.

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2. Radiotron Designer's Handbook, p. 1095.
3. K. R. Sturley, Wireless World, November, 1955.
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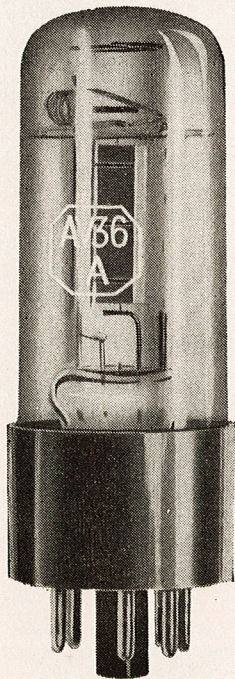
RADIOTRON AV36

Tungsten Filament Control Diode

DESCRIPTION.

The Radiotron type AV36A is a control diode having a pure tungsten filament and is intended for use in a.c. voltage regulator circuits. It supersedes the Radiotron AV33 over which it shows many improvements. Besides being more robust, this new diode contains a plate-to-filament short circuiting device which operates if the filament opens. This will prevent an excessive rise in the regulator output voltage (see application data for further details).

All tubes are now given a stabilising run. This factor, together with the larger filament diameter compared to that of the AV33, gives longer life, higher stability, and enhanced reliability. Thermal inertia and hence response time are similar to that of the AV33.



GENERAL DATA.

Mechanical:

Mounting position	Any
Overall length	$3\frac{7}{8}$ " max.
Seated height	$5\frac{5}{16}$ " max.
Diameter	$1\frac{3}{16}$ " max.
Bulb	T9
Base	Intermediate Octal, 7 pin

Maximum Ratings:

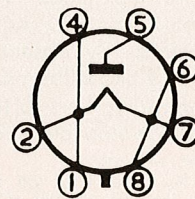
Filament voltage (a.c. or d.c.)	3.5 volts
Filament current	1.0 amp.
D.C. plate voltage	250 volts

Typical Operation:

A.C. filament voltage	3.0 volts
A.C. filament current	0.9 amps.
D.C. plate voltage	100 volts
D.C. plate current	1 mA
D.C. plate voltage for 80% of saturated plate current	20 volts

Socket Connections:

(Bottom View)



Pins 1, 2 and 4 Filament

Pin 5 Plate

Pins 6, 7 and 8 Filament

To reduce contact resistance each end of the filament is brought out to three base pins, all of which should be used.

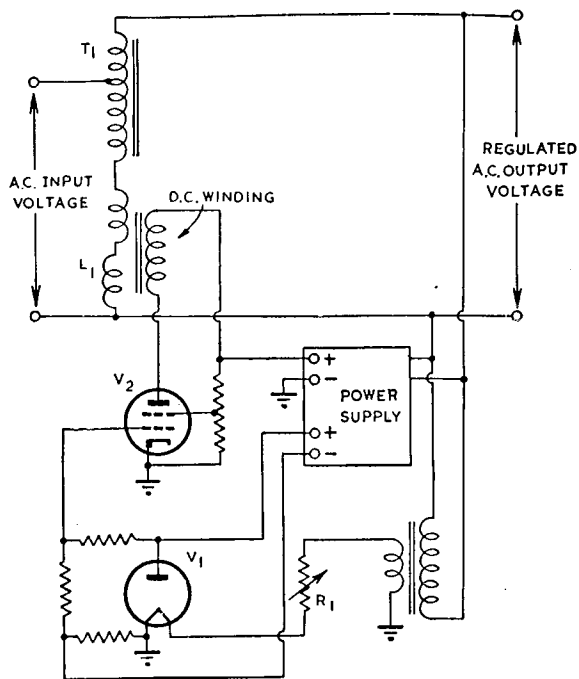
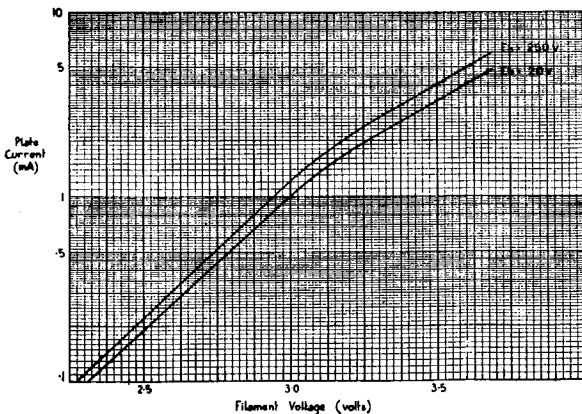


Fig. 1. Typical Circuit.

APPLICATION DATA.

A.C. voltage regulators generally include a sensing element which detects voltage deviations from the desired output voltage. The output of this sensing element is, after amplification, applied at some point in the circuit to minimize the initial deviation, so returning the output voltage to the correct value.

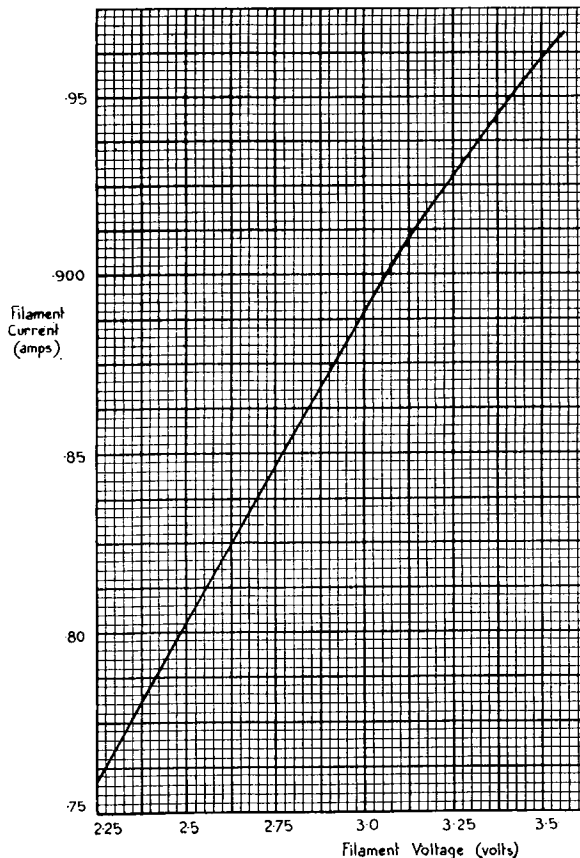
Temperature limited diodes having a pure tungsten filament are very satisfactory sensing elements because of their rapid change of saturated plate current with changing filament voltage.



If the filament of an ordinary control diode open-circuits, so causing its plate current to vanish, it can be seen from the above explanation that the output voltage must rise to the maximum extent allowed by T1. This is very undesirable in certain applications and, in the AV36A control diode, is avoided. The built-in switch shorts the plate-to-filament automatically when the filament opens so causing the output voltage to drop to the lowest extent allowed by T1. No damage due to excessive regulator voltage can thus occur.

Typical regulators of this type will maintain the output voltage accurately to 0.5% over a 10:1 range in load current, for input voltages varying 20% from normal mains voltage. The waveform distortion introduced can be made less than 5%. The AV36A is insensitive to frequency changes. This is an advantage in critical applications.

Control diodes having heavier pure tungsten filaments than the AV36A are often useful because of their longer response time. The AV36 B and C are intended to cover such requirements.



TECHNICAL LIBRARY

R.C.A. PHOTOSENSITIVE DEVICES and CATHODE RAY TUBES CRPD-105

This 24 page technical publication contains information on 109 R.C.A. tubes including single-unit, twin-unit and multiplier phototubes; flying-spot tubes, monitor, projection, transcriber and view-finder picture tubes.

A noteworthy feature of this publication is the layout in chart form of a brief description, use, physical characteristics, maximum ratings and operating conditions of the various tubes. Easy to read socket-diagrams are also included for all the tubes. This style of presentation of data facilitates the selection of the required tube for an application.

The sections on **phototubes** and **camera tubes** include a number of spectral sensitivity curves while the **cathode-ray tube** section includes interesting descriptions and characteristics of a number of fluorescent screen phosphors.

This manual would be of particular value to the circuit engineer concerned with development and research.

(J.A.H.)

TELEVISION ENGINEERING— PRINCIPLES AND PRACTICE VOLUME 3 WAVEFORM GENERATORS

by

S.W. AMOS, B.Sc. (Hons.), A.M.I.E.E.

and

D. C. BIRKINSHAW, M.B.E., M.A., M.I.E.E.

Published by Iliffe & Sons Ltd., London,
Philosophical Library, N.Y.

This volume has been written to describe in detail the fundamentals of waveform generators. It commences with the requirements of a TV camera chain.

The first oscillators discussed are the sine wave oscillators in their various forms, as these are frequently used as a standard from which are clipped, integrated or differentiated—a wide variety of shapes are evolved.

Circuit requirements are shown for bi-stable, monostable and astable (free running) generators—both anode and cathode coupled. Requirements for triggering or synchronising these units are illustrated. The work ends with a discussion of saw tooth, blocking oscillators and parabolic generators.

Radiotronics

This volume was originally prepared for the instruction of B.B.C. television engineers, and it follows that it is a worthy source of reference for all engaged in TV transmission.

Typical circuits are shown, although values are not given. However, the text includes all relevant formulae for the calculation of circuit values.

(J.W.E.)

“HIGH FIDELITY” A Practical Guide

Published by McGraw Hill Book Co. Inc., New York, this 310-page book was written by CHARLES FOWLER, the publisher of High Fidelity magazine.

Here is a book that tells you in clear, non-technical language the secrets of high fidelity.

Whether you are a beginner planning your first hi-fi system or a veteran with many years' experience and intentions of adding to your present outfit, this book will be a source of sound guidance and expert knowledge to you.

Mr. Fowler's book is divided into 14 chapters, beginning with a fascinating introduction to sound itself, what it is, how it is produced and how it is affected by room acoustics. Some of the other and more important chapters deal with loudspeakers, dividing networks, speaker enclosures, amplifiers, tuners, pick-ups, record changer and turntables, selecting, matching and budgeting equipment.

If you are looking for a book that tells you what brand-name equipment to buy, then forget about this one. Instead it tells you what to look for when purchasing sound equipment, what can be expected of certain size speakers and amplifiers and the type of units that Mr. Fowler's many years of experience have shown to be best.

Instead of telling you which unit to buy, this book lists the various types of loud-speakers, amplifiers, enclosure boxes, turntables and pick-ups, tells of their advantages and disadvantages and leaves the reader to make his own choice, because Mr. Fowler, like many other hi-fi authorities, knows that high fidelity depends a lot on what the individual wants.

I found the last chapter to be of particular interest. In this chapter, "Selecting, Matching and Budgeting Equipment," Mr. Fowler has gathered together a list of questions and answers to help the reader to determine certain fundamentals about the type of equipment best suited to his requirements.

The second part of this chapter considers individual pieces of equipment, and is designed to help in the selection of features you may want. The third section concerns finance and how much you should spend to get a balanced system.

Summarizing this book shows you the way to heighten your listening enjoyment with better reproduction in your home.

(B.A.B.)

June, 1957