# TELEVISION RECEIVING EQUIPMENT

*BY THE SAME AUTHOR*

WIRELESS SERVICING MANUAL

ASSOCIATED
ILIFFE
TECHNICAL BOOKS

# TELEVISION RECEIVING EQUIPMENT

*By*

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M.I.E.E.

*With* 283 *diagrams and photographs*

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# Contents

# Preface

APPARATUS FOR TELEVISION reception is based on the fundamental principles common to all wireless communication and electronic apparatus, and its mode of operation therefore depends on the ordinary laws of electricity and magnetism. Just as telecommunications and electronics are often regarded as specialized branches of electrical engineering, so television may be considered a special branch of telecommunications.

Because it is a branch, and not an isolated subject, television must be approached with a background knowledge of wireless theory and practice. In this book the reader is assumed to have this background and to be familiar with the way in which ordinary broadcast receivers work. Unless he is well acquainted with the principles and practice of such sets he is unlikely to make much headway in television.

Certain parts of the television receiver, notably amplifiers, detectors and frequency-changers, are identical in principle with those employed in equipment for sound broadcast reception and differ only because the required performance is different in bandwidth and operating frequency. Other sections of the television receiver, such as deflecting circuits, have no counterpart in sound apparatus. The author has felt it desirable, therefore, to give much more detailed explanations of these "special" sections of the apparatus than of those other more ordinary circuits, for it is these special parts which are liable to be particularly difficult to the newcomer to television.

As its title implies, the book is confined to receiving apparatus. A great deal of its contents is, of course, applicable to transmitting apparatus because amplifiers and scanning circuits are as necessary there as in the receiver. They are treated here, however, only from the point of view of their application in receivers.

The first edition of this book was scheduled to appear in the autumn of 1939, but its publication was delayed by the war until the middle of 1940. There were then, of course, no television transmissions in this country, for the Alexandra Palace transmitter, which had been in regular operation since February 1937, closed down on the outbreak of war. In spite of this the book had to be reprinted five times during the war, because it was found that much of the material in it was as applicable to radar as to television!

There was no opportunity of revising the book until after the war. The author was then able to undertake this task and the second edition appeared in October 1946, and contained a great deal of new material. So rapid is the advance in television technique, however, that the author has again had to revise the book and the result is this third edition.

Again, a large amount of new material has been added and many chapters have been re-written in an attempt to improve and simplify the presentation. To this end most mathematical expressions have been removed from the text and placed in appendices where they are more readily accessible to the designer.

A new chapter on deflector coils is included and another on selectivity. This is becoming an important matter with the expansion of the television service and the adoption of vestigial-sideband transmission. Electromagnetic deflection is much more fully treated and includes "economy" circuits and fly-back e.h.t. systems, the chapter on synchronizing has been expanded, and methods of suppressing ignition interference are dealt with.

A minor change of nomenclature deserves mention. Previously, the pre-detector circuits of a superheterodyne were designated radio- and intermediate-frequency circuits, following a common practice. It was found, however, that this led to confusion in some cases because it implied that an i.f. amplifier was not a radio-frequency amplifier. Accordingly all such circuits are now termed radio-frequency and are treated together under this heading.

The term intermediate frequency is retained with its old meaning for circuits between the frequency-changer and the detector, but circuits operating at the frequency of the incoming signal (previously r.f. circuits) are now called signal-frequency circuits. Following the main chapter on "Radio-frequency Amplification" there are two short chapters, "Signal-frequency Amplification" and "Intermediate-frequency Amplification," in which the points of difference between the two are separately treated.

All these changes have involved a great deal of re-writing and so the opportunity has been taken of re-setting the book completely and adopting a new format.

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*Dorset House,*
*London, S.E.*1
*June*, 1950

NOTE

The necessity for reprinting this edition has permitted making a few minor corrections to the text.

# ABBREVIATIONS USED

| | | | |
|---|---|---|---|
| A | ampere | mH | millihenry |
| a.c. | alternating current | mm | millimetre |
| a.f. | audio frequency | mm/V | millimetres per volt |
| c.r. | cathode ray | msec | millisecond |
| c/s | cycles per second | mV | millivolt |
| db | decibel | MΩ | megohm |
| d.c. | direct current | pF | picofarad |
| e.h.t. | extra high tension | p—p | peak-to-peak |
| e.m.f. | electromotive force | *RC* | resistance-capacitance |
| °F | degree Fahrenheit | r.f. | radio frequency |
| H | henry | r.m.s. | root mean square |
| h.t. | high tension | s.f. | signal frequency |
| i.f. | intermediate frequency | V | volt |
| in | inch | v.f. | vision frequency |
| kc/s | kilocycles per second | W | watt |
| kV | kilovolt | Ω | ohm |
| kΩ | kilohm | μF | microfarad |
| m | metre | μH | microhenry |
| mA | milliampere | μsec | microsecond |
| mA/V | milliamperes per volt | μV | microvolt |
| Mc/s | megacycles per second | pF | picofarad |

CHAPTER 1

# General Principles of Television

ALTHOUGH TELEVISION IS considerably more complex than sound transmission and reception, it bears a close relationship to it and a large part of the apparatus employed is very similar. In the case of sound, variations of air pressure are converted by the microphone into changes in an electric current, and at the receiver the inverse function is performed by the loudspeaker. In television, changes in light intensity are converted into variations of an electric current by a photo-cell, or its equivalent, and at the receiver the current variations are converted back to light changes by a cathode-ray tube, or some alternative.

The essential difference between sound and a still picture is that while the former consists of a sequence of changes of air pressure with time, the latter consists of a series of changes of light intensity with position. A moving picture consists of a series of changes of light intensity with both position and time.

The most complex sounds can be dealt with easily by the microphone, because there is at any instant only a single value of air pressure at a given point. The nature of the sound depends on how the pressure varies from instant to instant and the microphone has only to deal with these changes.

In the case of a picture, however, the light intensity varies over its surface. The variations all exist simultaneously and occur in different positions. In addition, with a moving picture, the light intensity of any and every point will be likely to change with time. It is this question of the change of light with position which makes television so much more complicated than sound broadcasting.

It is not proposed in this book, which is intended to deal with cathode-ray television receivers, to consider transmitting problems in any detail. Transmission and reception, however, are closely linked and in order properly to understand the latter it is necessary to know at least the basic principles of the methods of converting the scene to be transmitted into its electrical equivalent.

The matter is by no means difficult to understand, but at first it is sometimes a little hard to see what goes on, so we shall tackle it by easy stages. First of all, assume for simplicity that the picture to be transmitted is a transparency, say, a piece of cinema film. If we place a lamp on one side of the film and a ground-glass screen on the other, we know that with a suitable arrangement of lenses the picture will be projected on to the screen, for the amount of light passed by any part of the film depends on the density of that part.

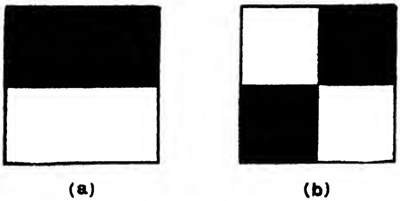



**Fig. 1.1—Two elementary " pictures " are shown here ; both have the same mean illumination**

The size of the projected picture on the screen depends on the optical system and physical layout of the parts and can be larger or smaller than the original. In the cinema it is very much larger.

Now consider the photo-cell. There are many types but we need not go into details here, and it is sufficient to say that it can be made to produce a current which is more or less proportional to the amount of light falling on its sensitive surface.

Suppose we project our picture on to this sensitive surface of the photo-cell. We shall clearly obtain a current proportional to the total light on the cell; this current can be used to operate a lamp or neon tube mounted so as to illuminate a screen. When we change the picture for a different one the amount of light reaching the photo-cell is likely to be different and so the lamp will be more or less brilliant. There will be no picture, however, on the receiving screen, for only changes in total illumination can be dealt with by this system; that is, variations of mean illumination with time.

Considering this a little further, suppose we adjust matters so that with a blank clear film at the transmitter the receiving lamp is at full brilliancy, and with an opaque film the lamp is just extinguished. Now suppose we insert a " picture " such as that of Fig. 1.1 (a), consisting of two rectangles of equal area, one white and one black. Obviously the amount of light falling on the photo-cell is one-half that which would be obtained with a completely white picture. The receiving lamp lights to half-brilliancy; call this grey.

Now remove this picture and insert that of Fig. 1.1 (b). The total light passing is obviously the same and so the photo-cell current is the same and the receiving screen still shows grey. The system is incapable of transmitting or reproducing detail.

Instead of using a single photo-cell and lamp, we might use four of each arranged as in Fig. 1.2. The light passing through each quarter of the picture can be made to fall on one cell only and similarly each lamp can be arranged to illuminate only one-quarter of the receiving screen.

It is clear that the system can now deal with the simple pictures of Fig. 1.1. It can be seen from Fig. 1.2 that squares 1 and 4 in the transparency pass no light, so that photo-cells 1 and 4 pass no current and lamps 1 and 4 do not light and squares 1 and 4 on the screen appear black. Squares 2 and 3, however, pass light and cells 2 and 3 pass current; consequently, the lamps 2 and 3 light and squares 2 and 3 on the screen are illuminated.

If the picture of Fig. 1.1 (a) were substituted, squares 1 and 2 would not pass light but 3 and 4 would do so. Photo-cells 1 and 2 would consequently pass no current, but cells 3 and 4 would; hence lamps 1 and 2 would not light, but lamps 3 and 4 would. On the screen, therefore, squares 1 and 2 would be black and squares 3 and 4 would be white.

Such a simple arrangement is only capable of dealing with patterns of four squares, but it will do this perfectly and any gradation of lighting between black and white in any square can be dealt with. Thus the squares need not be black and white, as we have assumed, but can be any depth of shade.

It is clear that more complicated pictures can be dealt with by providing more photo-cells, lamps and connecting wires. Theoretically, any picture, no matter how complicated and no matter how much detail it contains, can be dealt with by sufficient sub-division on these lines. If the elemental squares are small enough in relation to the size of the picture, the eye will not perceive them as such. The ordinary half-tone illustration, for instance, is composed of dots and they are clearly visible through a magnifying glass, but normally they are unseen.

It is not difficult to see that even if it were possible to employ the hundreds of thousands of photo-cells and lamps needed for high definition, it would be quite impossible to employ the equivalent number of connecting links between them. It is quite possible, however, to use only a single link by connecting the cells and lamps to it in sequence.

Let us go back again to the simple arrangement of four cells and consider Fig. 1.3. Here the arrangement is the same as in Fig. 1.2, save that there is only a single pair of wires for the connection of the cells and lamps. Switches $S_1$ and $S_2$ are provided and they are moved in synchronism so that $S_2$ is always on contact 1 when $S_1$ is on contact 1, on contact 2 when $S_1$ is on contact 2, and so on. If the arms move over all the contacts regularly and repeatedly, and the switches are worked sufficiently rapidly, the picture will appear on the receiving screen.

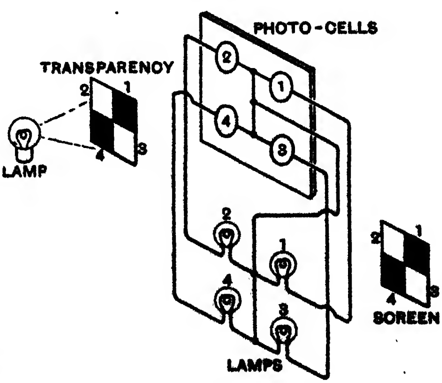



**Fig. 1.2—This diagram shows a simple television system capable of dealing with elementary "pictures"**

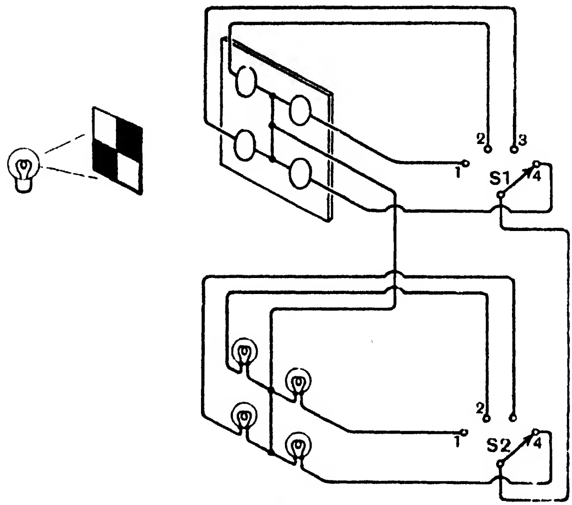



Fig. 1.3—A single pair of connecting wires can be used between transmitter and receiver if synchronous switching is employed

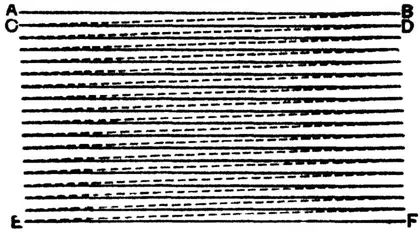
Although only one lamp will be alight at any instant, it will appear as though all were functioning because the eye cannot follow very rapid changes in illumination. If a lamp is switched off the eye continues to " see " it for a fraction of a second afterwards. This phenomenon is known as persistence of vision, and without it the cinema would be impossible. Here a succession of still pictures is thrown on the screen and the eye cannot perceive the gaps between them.

In the same way the scheme of Fig. 1.3 would work if the switches were operated sufficiently rapidly. Each rotation of the switch would correspond to one picture and about twenty-five rotations a second would be needed to give a good illusion of a steady picture. As compared with the direct connection of Fig. 1.2, the only discernible difference with a sufficiently rapid rotation of the switches would be a general loss of brilliancy. This is because each lamp now operates for only a fraction of the total time instead of continuously.

Needless to say such crude methods are not used in practice, and we now come to a workable system. Although it is not one which is used on the transmission side of the regular television service, it is still a useful one for experimental purposes. At the receiver it follows ordinary practice exactly.

The cathode-ray tube is discussed in a later chapter, but for the moment it is sufficient to consider it as a device with a screen, upon which a small spot of light can be made to appear and moved almost instantaneously to any required position. The brightness of the spot can also be controlled with equal rapidity.

**Fig. 1.4—The raster is traced out by a spot of light. Starting at A the spot travels to B, flies back to C, traces out CD, and so on. From F it goes back to A and starts again**

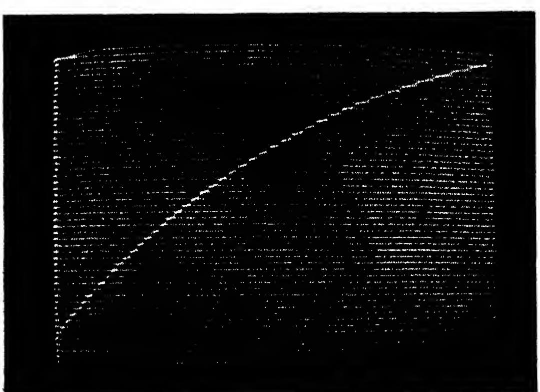



Now suppose we set up such a tube and make the spot move steadily across the screen, then fly back very rapidly to a position just below its previous starting point, and then commence again to move steadily across the screen. In this way it will trace out a series of lines such as AB, CD in Fig. 1.4, where the dotted lines represent the very rapid return strokes. In tracing out these lines it moves from top to bottom of the screen and when it has produced EF it flies back again to A and starts to trace out a new series exactly on top of the first.

If this process is continued sufficiently rapidly the eye will not perceive the spot but will see only a series of parallel lines. If the number of lines is increased sufficiently even these will begin to disappear and will be visible only on close inspection. The eye will see an illuminated rectangle on the screen and this is known as the raster. Actually, of course, at any instant there is no more than a single spot of light and the appearance of the raster is an illusion.

At the receiving end we set up a similar tube and arrange for its spot to traverse the screen in *exactly* the same way as in the transmitting tube. Not only must the spot always move at the same rate as that in the transmitter, but it must always be in exactly the same position relative to the raster which it builds up. A single photocell is used at the transmitter and its output is arranged to control the brightness of the spot on the receiving screen.

**Photograph of a typical raster. The number of lines is much less than that normally used for television, and the frame fly-back line is visible because it is not blacked out as it would be during the reproduction of a picture**



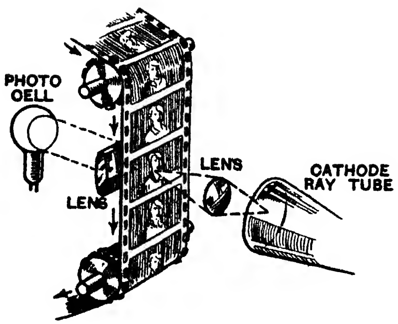



**Fig. 1.5—One method of transmission is illustrated here. The raster on the cathode-ray tube is projected through the film on to a photo-cell**

A transparency such as a cinema film is used and placed between the transmitting tube and the photo-cell as in Fig. 1.5. The raster as a whole is focused through the film on to the photo-cell. The eye sees the raster as a whole, but the photo-cell sees it as it is—as a single moving spot of light—and it sees it through the film. As it passes through parts of the film of different density the light falling on the cell varies accordingly and sets up a correspondingly varying current. This current controls the brightness of the spot on the receiving screen and as this spot is always in the same relative position as that at the transmitter the variations of brightness build up the picture.

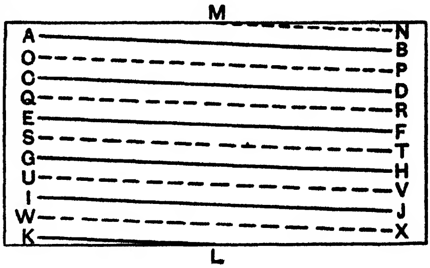
This particular method of transmission is not greatly used because it is limited to transparencies. Somewhat more complex arrangements are needed to deal with ordinary scenes, but the receiving arrangements remain the same. Every cathode-ray television receiver has a spot tracing out the series of lines which form the raster, and the brightness of the spot is controlled to give the light and shade and picture detail.

It is thus only necessary to convey on the link between transmitter and receiver the variations in current corresponding to the variations in light on the photo-cell, which variations are caused by the spot moving across the picture. In practice, however, something more is needed, for it is necessary to ensure that the transmitting and receiving spots move in synchronism. This is done by wasting a small amount of the picture, as by providing a black border, and using the intervals to transmit synchronizing signals.

Assuming a spot of infinitesimal dimensions, it is clear that the greater the number of lines which it draws the better will be the definition of the picture. In practice the spot is of finite size, so that there is a limit to the number of lines which can usefully be employed. This limit has not yet been reached with good cathode-ray tubes, but has been exceeded with poor ones.

The process of exploring the subject by the spot is known as scanning and the particular method described, in which the spot draws a series of lines each immediately beneath its predecessor, is called sequential scanning. Other methods are possible and are

**Fig. 1.6—With interlaced scanning the lines of successive frames are not superimposed, but are staggered. The lines of alternate frames are superimposed**




actually more widely used. The B.B.C. transmissions, for instance, use interlaced scanning.

This is very important and is only slightly more complicated than sequential scanning. The spot moves across and down the screen as before tracing out parallel lines. These lines, however, are no longer immediately adjacent to one another, but are separated by appreciable gaps. When the spot has reached the bottom it does not return to trace out another series of lines on top of the first, but to trace out a series in the gaps between the first. When it again gets towards the bottom it returns to trace out the third series on top of the first, and then a fourth on top of the second and so on.

This is illustrated in Fig. 1.6. The spot starts at A and traces out the line AB; then it flies back to C and traces out CD. In this way it traces out the lines EF, GH, IJ, KL. On reaching L it flies upwards to M and continues on MN; then to O, and traces out OP between AB and CD. It then goes on to cover QR, ST, UV, WX. From X it goes back to A, and the next line is AB, and so on.

There are thus two vertical traversals to each complete series of lines. One complete series is usually known as one picture and one half-series is called a frame, so that there are two frames to a picture.

Interlacing is adopted because it greatly reduces flicker for a given number of pictures a second. It is found that twenty-five complete pictures a second with sequential scanning will give no noticeable flicker provided that the brightness of the picture is low. In order to avoid flicker with bright pictures there must be about fifty per second. If the number of lines remains the same this means that the maximum frequency involved in the electrical circuits is doubled. Consequently, the space occupied in the ether by the transmitter is doubled and the transmitting and receiving equipment becomes much more expensive.

In order to avoid serious hum troubles the picture repetition frequency must be a multiple or sub-multiple of the mains frequency, in this country 50 c/s. Consequently, the only practicable frequencies are 25 c/s or 50 c/s. In America where the mains frequency is 60 c/s, the picture frequencies are 30 c/s or 60 c/s.

Interlaced scanning is used with a picture repetition rate of 25 c/s because this gives fifty frames a second. The frequencies involved in transmission are no greater than with sequential scanning and twenty-five pictures a second, but the flicker is the same as with fifty pictures a second, that is, it is unobservable. Actually, the result is not quite the same, for it is possible to detect interline flicker upon close observation, while if the eyes are moved rapidly while looking closely at the picture one set of lines seems momentarily to disappear. At the normal viewing distance, however, these effects are hardly observable.

CHAPTER 2

# The Television Signal

IN ORDER TO understand the operation of receiving equipment it is essential to be familiar with the waveform of the signal, for it is of quite different form from that encountered in ordinary broadcasting. The vision carrier is modulated by signals which correspond to the picture detail, the mean level of illumination of the picture, and the line and frame synchronizing pulses.

This is done by making one definite level of modulation correspond to black and another level to white. The picture signals always come between these two levels and the mean level of illumination is dependent on the mean level of modulation between these limits. At the end of every line and at the end of every frame there is a gap in the picture signal and during this interval the modulation is changed beyond black to provide the synchronizing pulses.

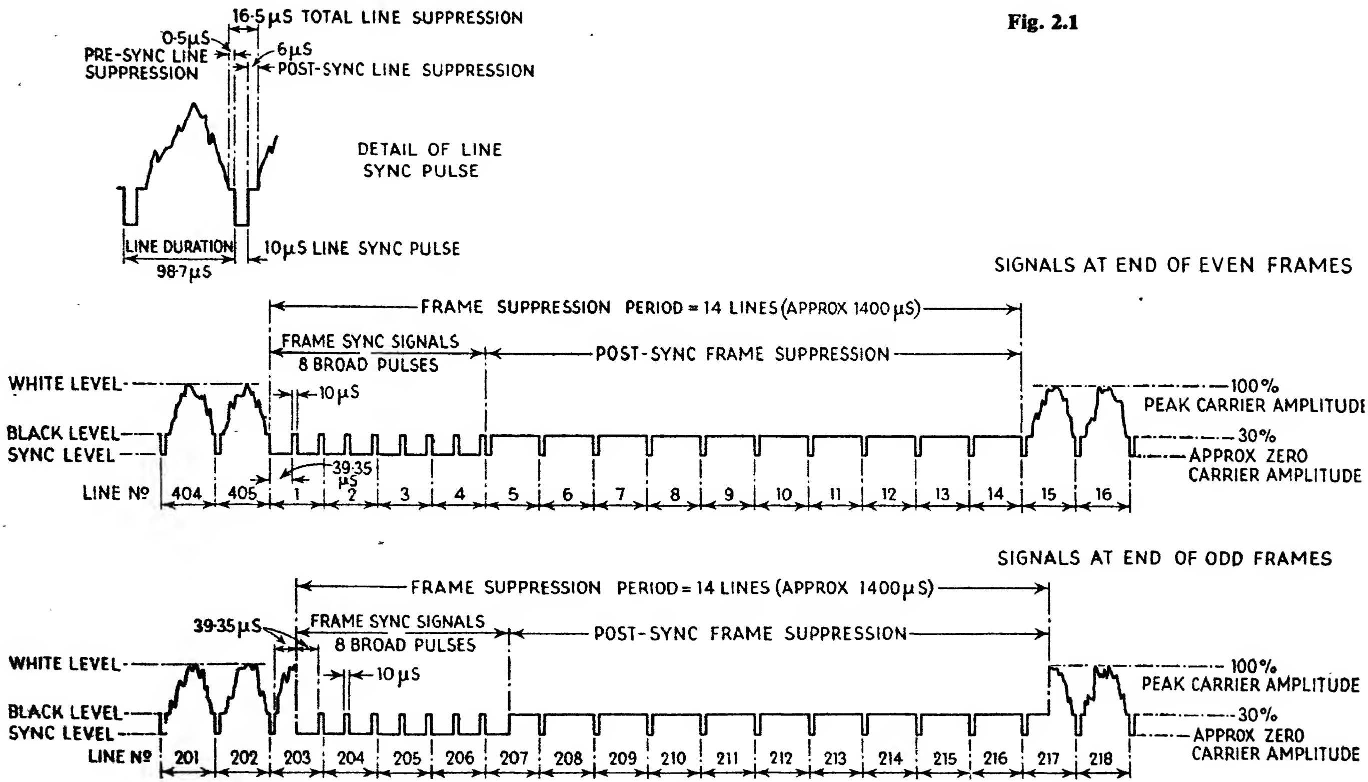
It should be noted that the term modulation has in television a slightly different meaning from that customarily applied to it in sound transmission. In the latter the r.f. carrier has a fixed mean amplitude and the modulation increases and decreases the amplitude at a rate depending on the modulation frequency. With 100 per cent modulation by a sine wave, the carrier amplitude is changing from zero to twice its mean value.

In television there is no fixed mean value to the carrier; only the maximum value is fixed. When we say that white corresponds to 100 per cent modulation, therefore, we mean that the carrier amplitude is at its maximum. Similarly, to say that black corresponds to 30 per cent modulation means that when transmitting a black signal the carrier level is 30 per cent of the maximum.

In the B.B.C. transmissions by the Marconi-E.M.I. system black corresponds to 30 per cent modulation of the carrier and white to 100 per cent; the range between zero and 30 per cent is reserved for the sync pulses. The sense of the modulation is thus positive. In America the same general arrangement is adopted, but with negative modulation; that is, with zero modulation corresponding to white.

There are 405 lines to each complete picture and twenty-five complete pictures a second with interlaced scanning, giving fifty half-pictures or frames, each of 202·5 lines, per second. There are $405 \times 25 = 10{,}125$ lines every second and so the duration of each complete line is 1/10,125 of a second; that is, 98·77 $\mu$sec. The picture ratio is four horizontal to three vertical, the ratio being reckoned on the active portions of the lines. (The ratio was originally 5:4, but it was altered to 4:3 in April 1950, to bring it into conformity with cinema practice.)

Fig. 2.1


16·5µS TOTAL LINE SUPPRESSION
0·5µS
PRE-SYNC LINE SUPPRESSION
6µS
POST-SYNC LINE SUPPRESSION
DETAIL OF LINE SYNC PULSE
LINE DURATION
98·7µS
10µS LINE SYNC PULSE
SIGNALS AT END OF EVEN FRAMES
FRAME SUPPRESSION PERIOD = 14 LINES (APPROX 1400µS)
FRAME SYNC SIGNALS
8 BROAD PULSES
POST-SYNC FRAME SUPPRESSION
WHITE LEVEL
BLACK LEVEL
SYNC LEVEL
10µS
39·35 µS
LINE Nº
404 405 1 2 3 4 5 6 7 8 9 10 11 12 13 14 15 16
100%
PEAK CARRIER AMPLITUDE
30%
APPROX ZERO CARRIER AMPLITUDE
SIGNALS AT END OF ODD FRAMES
FRAME SUPPRESSION PERIOD = 14 LINES (APPROX 1400µS)
39·35µS
FRAME SYNC SIGNALS
8 BROAD PULSES
POST-SYNC FRAME SUPPRESSION
10µS
WHITE LEVEL
BLACK LEVEL
SYNC LEVEL
LINE Nº
201 202 203 204 205 206 207 208 209 210 211 212 213 214 215 216 217 218
100%
PEAK CARRIER AMPLITUDE
30%
APPROX ZERO CARRIER AMPLITUDE

At the end of each line the modulation falls to black level and remains there for 0·5 μsec, it then falls to zero and remains there for 10 μsec. Then it rises again to black level and remains there for 6 μsec; the picture signal then recommences.

Of each line just over 10 per cent is occupied by the sync pulse and 6 per cent by black signals. Picture signals are transmitted during only 83·3 per cent of each line. Similarly, at the end of each frame 14 lines are devoted to sync pulses and black signals. The number of lines effective in the picture is thus 188·5 for each frame, and 377 per picture.

The waveform of the signal is shown in Fig. 2.1, and well illustrates the details just discussed. The waveform is, of course, that of the vision-frequency signal; that is, the waveform of the detector output, corresponding to the audio frequency of a sound signal. The object of this waveform is to make it easy to separate the sync pulses from the picture signal by a simple amplitude filter, for a limiter can be used to cut off all variations above a certain level and so remove all trace of the picture signal. It is consequently not difficult to ensure that the synchronizing is unaffected by changes in the picture content.

The gap of 0·5 μsec is inserted in order to make sure that the signal always returns to black level before the sync pulse starts. If it were not for this interval trouble would be experienced. Suppose, for instance, that there is a white edge on the right-hand side of the picture, then the modulation would be required to change from 100 per cent to 30 per cent instantaneously to be ready for the sync pulse. It cannot do this, and the effect would be to introduce an effective time-delay to the start of the sync pulse. The gap at black level is consequently inserted to give the modulation time to change to black before the sync pulse starts.

The duration of the sync pulse and the black level following it give time for the fly-back of the scanning spot, and also black out the spot during the fly-back, so that it is invisible. As there are fifty frames a second the frame scanning oscillator must function at 50 c/s and the line oscillator at 10,125 c/s (= 25 × 405).

For convenience of reference the times (correct to two decimal places) of duration of the various components of the signal are given in Table 2.1.

In the case of sound broadcasting one is accustomed to specifying the bandwidth of the receiver circuits in terms of the highest modulation frequency. The highest needed is, of course, the highest audible frequency, but the highest allowed for is often much less.

In television there is no simple direct correlation with frequency in this manner for the signal waveform resulting from scanning a picture is more of the pulse type and bears no obvious relation to frequency. It can, however, be expressed in terms of frequency and it is often convenient to do this for it is usually much simpler in amplifier design to think in terms of frequency than to deal directly with pulse waveforms.

TABLE 2.1

**VISION-SIGNAL WAVEFORM**

| *Line Detail* | *per cent* | *μsec* |
|---|---|---|
| Black level after scan | 0·506 | 0·5 |
| Sync pulse | 10·124 | 10·0 |
| Black level after sync pulse | 6·075 | 6·0 |
| Picture signal | 83·295 | 82·27 |
| Total | 100·0 | 98·77 |
| Total scan period | 83·8 | 82·77 |
| Total fly-back period | 16·2 | 16·0 |
| *Frame Detail* | *Lines* | *msec* |
| Sync pulses | 4·0 | 0·395 |
| Black level | 10·0 | 0·987 |
| Picture signal | 188·5 | 18·618 |
| Total | 202·5 | 20·0 |
| Total scan period | 188·5 | 18·618 |
| Total fly-back period | 140 | 1·382 |

One method of interpreting the picture on a frequency basis is to consider the active area as divided into alternate black and white squares of elemental size. For equal vertical and horizontal definition the elemental squares have a side equal to the width of a scanning line. The width of a line is the picture height divided by the number of active lines and the number of elements to the active portion of a line is the visible width of the picture divided by the line width. This is merely the product of the number of active lines and the aspect ratio; for the B.B.C. transmissions it is $377 \times 4/3 = 503$.

Reference to Table 2.1 shows that the duration of the active part of a line is 82·27 μsec, so that the duration of a picture element is 82·27/503 = 0·165 μsec. Now the ideal waveform for the chess-board pattern of alternate black and white squares is a rectangular wave like the one shown in Fig. 2.2 by the solid line. Such a wave can be regarded as being composed of the sum of a number of harmonically related sine waves of which the fundamental component recurs at the same frequency as the rectangular wave. This fundamental component is shown dotted in Fig. 2.2.

The time of one complete cycle of the sine wave is clearly equal to the duration of two picture elements, for the positive half-cycle corresponds to one element and the negative to the next. The

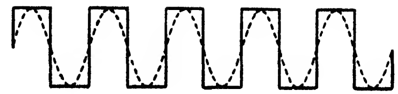


**Fig. 2.2—Rectangular waveform with a superimposed sine wave (dotted) of the same repetition frequency**

duration of a cycle is thus 0·33 $\mu$sec and this corresponds to a frequency of 1/0·33 = 3·0 Mc/s.

The fact that the wave is not transmitted during the synchronizing periods has been taken into account in computing the number of elements per line and it does not further affect the result.

In order to reproduce such a rectangular wave really well harmonics of the fundamental up to about ten are needed and so the maximum frequency would be 30 Mc/s. It is impracticable to use such a high frequency in television, however, and it is customary to take the maximum modulation frequency as that corresponding to the fundamental component of the rectangular wave. In fact, the maximum frequency transmitted is about 2·75 Mc/s.

This whole concept of frequency in television is rather artificial but it is convenient in practice because it is quite easy to measure the frequency response of apparatus. The frequency response alone, however, does not fully specify the performance of equipment and it should be taken as no more than a useful guide.

It is of interest to compute the width of a picture element as it would appear to anyone looking at a perfect reproduction of the elemental chess-board pattern. It is, of course, the line width and so is 1/377 of the picture height. With a 9-in tube, for instance, the picture is often 7·5-in wide and 5·65-in high, so that the line width is 5·65/377 = 0·015 in.

Because of the high modulation frequencies extending up to some 2·75 Mc/s television of the high definition variety cannot be transmitted on the normal broadcast bands. The carrier frequency must be higher than the highest modulation frequency and it is very desirable that it should be large compared with it.

The ordinary short-wave bands 10–50 m (30–6 Mc/s) are unsuitable for television because of their propagation characteristics. A considerably greater coverage would be obtained by direct ray than is possible at higher frequencies, but ranges of several thousand miles by sky-wave propagation would also be obtained. Selective fading would occur, however, and render such long-range television reception useless. This is easily realized when it is remembered how severe the distortion due to selective fading can be in sound reception when the modulation frequencies involved are perhaps only 5 kc/s.

Although too distorted to be resolved into a useful picture the signal would be there in sufficient strength to cause serious interference with local transmissions. As things are at present, therefore, it is necessary to choose frequencies which are not normally propagated by sky wave. It should be noted, too, that even if the short-wave band were suitable for television it is too much occupied by other services to be available for it.

TABLE 2.2

### TELEVISION CHANNELS

| *Channel* | *Allocation* | *Vision-carrier frequency* (Mc/s) | *Sound-carrier frequency* (Mc/s) |
|---|---|---|---|
| 1 | London (Alexandra Palace) | 45·00 | 41·5 |
| 2 | — | 51·75 | 48·25 |
| 3 | — | 56·75 | 53·25 |
| 4 | Birmingham (Sutton Coldfield) | 61·75 | 58·25 |
| 5 | — | 66·75 | 63·25 |

The British television band is 41–66·5 Mc/s and it is likely that the full band of 41–68 Mc/s allocated to television at the Atlantic City Conference will eventually be available.

Within this full band there is room for five television channels each comprising a vision channel and an associated sound channel. The present channel allocation is given in Table 2.2 and it will be seen that the sound channel is always 3·5 Mc/s below its associated vision channel. In all cases amplitude modulation is to be used for the sound channel. Channel 1, for London, differs from the rest in being a double-sideband transmission. For all the others single-sideband or, more properly, vestigial-sideband transmission will be used.

Sidebands up to 2·75 Mc/s lower in frequency than the carrier are fully retained. Above the carrier, sidebands up to 0·75 Mc/s only are fully retained and higher sideband frequencies are greatly attenuated. For the correct results with such a transmission the overall frequency response of the receiver should be 100 per cent (0 db) from 0·75 Mc/s to 2·75 Mc/s below the carrier frequency. It should fall off uniformly from 100 per cent at 0·75 Mc/s below the carrier to zero at 0·75 Mc/s above the carrier, passing through the point of 50 per cent (—6 db) response at the carrier frequency.

The response at the associated sound carrier 3·5 Mc/s below the vision carrier should be less than —30 db if interference from the sound is not to affect the picture. The attenuation needed at a frequency 1·5 Mc/s above the vision carrier depends on where the receiver is to be used since the attenuation needed clearly depends on the relative strengths of the wanted signal and of the sound signal of the adjacent channel at the point of reception. In fringe areas at least 50 db is likely to be needed, for the beating of the vision carrier with the adjacent sound carrier produces a 1·5-Mc/s beat and the resulting interference pattern on the tube is easily visible.

A receiver designed for the vestigial-sideband system of Channels 2–5 is also suitable for the London station on Channel 1, but it is not, of course, necessary to design a receiver for Channel 1 on these lines. Either double- or single-sideband reception can be used at will when the transmission contains both sidebands.

It seems probable, however, that double-sideband reception is only being retained for Channel 1 because any change would involve alterations to so many existing receivers. It is likely, therefore, that future receivers will all be of the single-sideband type.

CHAPTER 3

# The Cathode-ray Tube

AS APPLIED TO television, the receiving cathode-ray tube is invariably of the high-vacuum type, for the gas-focused tubes sometimes used for oscilloscopes are quite unsuitable. Essentially it consists of a source of electrons, means for forming the electrons into a narrow beam, a method of focusing the beam upon the screen, a deflecting system whereby the angle of the beam can be changed, and finally a fluorescent screen.

There are two main types of c.r. tube—the electrostatic and the electromagnetic. In the former focusing and deflection are accomplished by the application of suitable potentials to electrodes within the tube, whereas with the latter they are effected by passing currents through coils mounted outside the tube. Occasionally other tubes are met with; thus, sometimes electrostatic focusing is used, but electromagnetic deflection. Sometimes, also, electric deflection is adopted for the horizontal scan with electromagnetic for the vertical.

The tube consists of a long glass tube containing the electrode assembly and terminated at one end by a truncated cone and sealed at the other. The large end of the cone is closed by a sheet of glass which is made as flat as possible consistent with the necessary mechanical strength. The inside of this sheet is coated with a thin layer of fluorescent material and is called the screen.

The general arrangement of an electromagnetic tube is sketched in Fig. 3.1. For television purposes the overall diameter of the screen may lie between 5 in and 16 in, and the length of the tube is about two to three times the screen diameter.

The electron gun is contained within the narrow part of the tube at the end remote from the screen and is so called because its object is to shoot a narrow beam of electrons at the screen. It consists essentially of a cathode, grid and an anode. The cathode can be directly or indirectly heated, but the latter is now almost universal.

The grid is a grid only in name, for in form it is more like a metal cup with a small hole in the bottom and it is placed over the cathode and surrounds it so that electrons can only emerge through the hole. Originally known as the Wehnelt cylinder, this electrode was later called the cathode shield, but as it is the control electrode, it is now termed the grid in order to bring the terminology into line with that used for valves.

The arrangement of the gun is shown in Fig. 3.2, and in practice the cathode is only coated with emitting material at its tip, facing the hole in the grid. This electrode is maintained at a negative

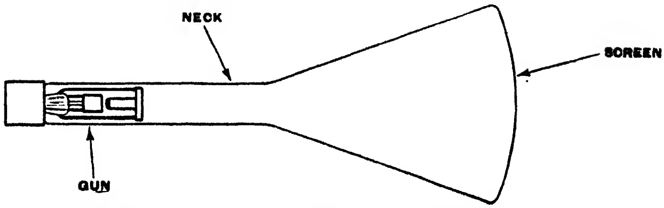



Fig. 3.1—General form of an electromagnetic c.r. tube

potential with respect to the cathode. Electrons emitted by the cathode leave it with a certain initial velocity and immediately come within the repelling field of the negative grid. They are constrained to fall back into the cathode and if the grid potential is sufficiently negative, they do so.

As the grid is made less negative, however, there comes a point at which a few electrons succeed in escaping through the hole in the grid. To a large extent it is their initial velocity which carries them through, but in line with the hole the repelling force of the grid is less than in other directions, and for two reasons. First, because of the hole itself there is no negative body in this line to repel them, only the field of the material surrounding the hole. Secondly, the positive anode outside the grid acts in some measure through the hole to attract the electrons.

As the grid is made less negative electrons escape in greater and greater numbers through the hole and are constrained by the electrostatic field around the hole to issue in the form of a beam. There is thus a beam of electrons leaving the grid and its density and, to some extent, its cross-sectional area are dependent upon the grid potential.

Outside the grid there is the anode or accelerator. This takes various forms but is usually based on a flat disc with a hole in the centre, sometimes there is a tubular extension around the hole, and often the inside of the glass wall of the tube, from the position of the anode nearly as far as the screen, is coated with a conducting material, such as a graphite compound. Such a coating is connected to the anode.

The anode is maintained at a potential of some 3,000–10,000 V positive with respect to cathode. The electrons emerging through the hole in the grid come at once under the influence of the anode

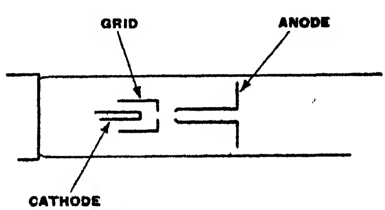



Fig. 3.2—General arrangement of the electrodes forming an electron gun

and their velocity is enormously accelerated. Some of them go straight to the anode and return to the cathode through the external circuit, others pass through the hole in the anode and continue to the screen in the form of a divergent beam.

On striking the fluorescent screen the electrons lose their energy. A part is expended in knocking out lower velocity electrons from the screen and the electrons of this secondary emission travel back to the anode, usually to the graphite coating of the tube walls. The rest of the energy is converted into light by the fluorescent material. On striking the screen the divergent beam causes a circular patch of light several inches in diameter.

In order to make use of the tube it is necessary to change the divergent beam from the anode into one which converges to a point at the screen. In other words, the beam must be focused. This can be done by means of a magnet which is arranged to provide a magnetic field parallel with the axis of the tube.

It is usual to employ an electromagnet in view of the ease with which the field strength can be adjusted. There is, however, now a decided tendency towards the use of the permanent magnet. With some types it is arranged to give approximately the correct field strength for focusing and its effect can in some measure be adjusted by sliding it along the neck of the tube. Occasionally some other adjustment of focusing is provided, however, and this may take the form of a control of the voltage applied to an electrode built into the tube, or the magnet may have a winding through which an adjustable current is passed.*

This last case corresponds very closely with the usual electromagnet, the only difference being that most of the field is provided by the permanent magnet and only a portion for adjustment purposes by the winding. The advantage over the pure electromagnet is a saving of energizing current and wire.

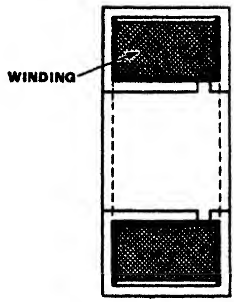
The usual method of controlling the effective field strength of a permanent focusing magnet, however, is by providing it with an adjustable air gap. The arrangement is constructionally similar to that of the electromagnet shown in Fig. 3.3, but the winding is replaced by the magnet and the air gap is adjustable.

An electromagnet can consist essentially of a coil wound on a former which slips loosely over the neck of the tube so that the axes of the coil and tube are coincident. It is necessary to have a uniform field and it is common to use an iron-clad coil. The iron encloses the coil except for a narrow gap through which the flux extends. An arrangement of this sort is sketched in Fig. 3.3.

The coil is placed around the neck of the tube beyond the anode (Fig. 3.4). Usually the rear end of the coil is an inch or two beyond the plane of the anode. The current through the coil is then adjusted for optimum focus. The current required depends on the position of the coil as well as on its design, but is often of the order of 200–600 ampere-turns. The size of the spot of light on the

* *Wireless World*, December 29, 1938.

Fig. 3.3—A typical electromagnetic focusing coil




screen depends on the position of the focusing coil in relation to the anode and decreases as the coil comes nearer to the screen.

The diverging beam of electrons from the anode enters the field of the focusing coil and electrons which are moving at an angle to the axis of the field are acted upon by the field. The force acting on these electrons changes their paths to new ones meeting in a point. By correct adjustment this point is arranged to occur on the screen.

The force acting on each electron depends on its charge, its velocity, and the field strength.* The velocity of the electron depends in large measure upon the anode voltage, consequently the strength of the magnetic field needed for focusing is a function of the anode voltage. Electrons which enter the field at an angle to its axis are not merely " refracted " to follow the desired convergent path beyond it, while in the field they are given a rotary motion about the axis and move in spiral fashion.

* "Magnetic Focusing," by H. Wood. *Wireless World*, April 2, 1937.

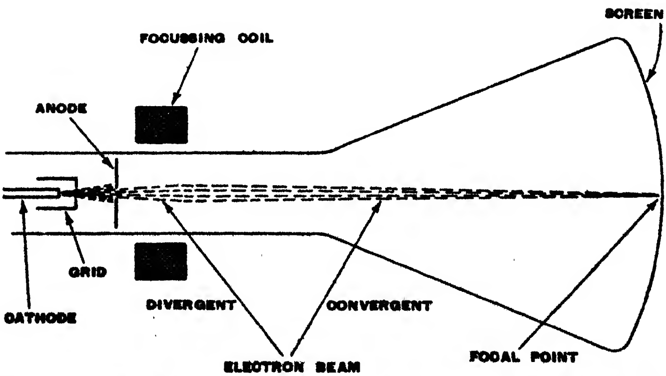



Fig. 3.4—Sketch of magnetic tube showing the position of the focusing coil

In practice, the electrons do not reach the screen at a point, but are still in the form of a beam of finite dimensions. The spot of light consequently has finite size and is, or should be, circular. The condition of optimum focus is the one giving a minimum diameter to this spot.

The reason for this is found to lie partly in imperfections of the magnetic field and partly in the mutual repulsion of the electrons forming the beam. This latter effect always gives the beam a tendency to spread, and the effect increases with the electron density of the beam. A smaller spot can be obtained by working with a high-velocity beam of low density than with one of low velocity but high density. In other words, high voltage and low current are better than low voltage and high current.

In practice, the magnetic field is not perfectly uniform and its axis may not coincide exactly with the axis of the beam issuing from the hole in the anode. Such imperfections affect the focus obtainable and the spot may not be circular. As the coil is external to the tube, it is, of course, always possible to adjust it for coincident axes. An alternative method is to keep the focus magnet centred about the tube neck and to centre the beam by a small magnet behind the focus magnet.

The brightness of the spot on the screen depends upon the fluorescent material, the velocity of the electrons in the beam, and the density of the beam; that is, apart from the screen material it depends on the anode voltage and beam current. With a given tube operated at a given anode voltage the brightness depends on the beam current which in turn depends on the grid voltage.

The picture signal is applied to the grid and so varies the brightness of the spot in accordance with the variations in the signal. The initial bias of the tube is adjusted so that black level in the signal corresponds with the current cut-off point of the tube, at which the spot is just extinguished. The picture-signal variations then drive the grid positive with respect to this point and so make its potential less negative with respect to the cathode.

The maximum signal which can be applied depends on the tube and usually depends on the focus required. As the negative grid voltage is reduced towards zero the beam current and brightness increase. It is usual, however, for a point to be reached at which the size of the spot begins to increase also, due to the greater density of the beam; this is equivalent to a loss of focus, but it cannot be corrected by adjustment of the focusing control. As it causes a loss of definition it sets the limit to the brightness obtainable.

When the tube is driven to this point on the white parts of the picture and the brightness is inadequate, it is necessary to increase the anode voltage. This increases both electron velocity and density, and hence brightness, but the spot size does not increase because the higher velocity of the electrons overcomes the tendency of the higher density beam to spread.

Deflection of the electron beam, so that it can build up the raster on the screen, is accomplished after focusing. Uniform magnetic fields at right angles to each other and to the undeflected beam are used and deflection is at right angles to the field.

A pair of large diameter coils with their planes vertical and their axes at right angles to the neck of the tube can be used and will cause a vertical deflection of the beam. In practice, such coils are not used, since they occupy too much space for it to be possible to use the two pairs needed for horizontal and vertical deflection.

The coils are usually wound on a rectangular former and then bent around the neck of the tube so that the horizontal sides of the two coils almost touch; these are used for the line or horizontal deflection. The frame or vertical deflecting coils are similarly wound and bent round the tube outside the line coils, the gaps between this pair being arranged at right angles to those between the other pair. The whole is often enclosed in a metal shroud of some kind of transformer iron. The arrangement is sketched in Fig. 3.5.

The deflecting currents passed through the coils are of saw-tooth waveform and if the fields produced are uniform the focus is unaffected. In practice, the fields are not perfectly uniform and have components which act along the axis of the beam and consequently have a focusing action on the beam. As this is already correctly focused the axial components result in a defocusing of the beam.

What happens with a poorly designed system is that towards the edges of the picture, where the deflection angle is large, the spot size increases and is often drawn out into a line. This naturally causes a marked falling off in the definition of the picture near the edges.

The difficulties of avoiding this defect increase with the deflection angle, so that for a given picture size it is better to use a long tube than a short one. The important distance is that between the centre of the deflecting coils and the screen. Not only does an increase in tube length at this point reduce the difficulties of coil design, but it also reduces the deflecting current needed. The current can also be reduced by using long deflecting coils, so that

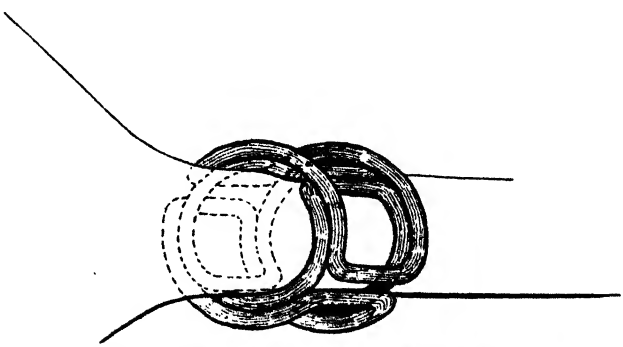


**Fig. 3.5—Typical arrangement of the deflecting coils**

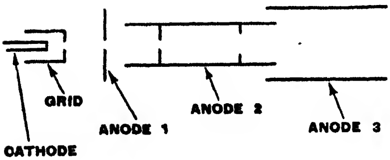



**Fig. 3.6—The average electrostatic tube has three anodes, or accelerators, arranged in the manner here shown**

the long tube scores in several ways over the short one; it is much more economical in coil design and in deflecting current. Nevertheless, the modern tendency is to use the short tube, largely because it can be mounted horizontally for direct viewing in a shallower cabinet.

The deflecting current needed with a given tube and coil system is directly proportional to the square root of the anode voltage. If it proves necessary to increase the anode voltage in order to obtain the required brightness, therefore, it is also necessary to increase the output of the time bases.

Turning now to the electrostatic system, we find that coils are not used, for the focusing and deflecting electrodes are built into the interior of the tube. Since no external coils are used and there is more internal apparatus the diameter of the tube neck is considerably larger than that of an electromagnetic tube.

The usual television tube has three anodes, two pairs of deflecting plates and the cathode and grid. The anodes or accelerators take various forms in practice and perform a two-fold function; they accelerate the electrons forming the beam and constrain them to follow paths which come to a focus on the screen. In the magnetic tube there is only one anode* which is included to accelerate the electrons, focusing being subsequently carried out by a magnetic field. In the electrostatic tube focusing and acceleration are carried out together in steps.

A typical electrode system is shown in Fig. 3.6, and it will be seen that up to the first anode the arrangement is similar to that of a magnetic tube. The potential of the first anode is usually quite low, however, only rarely exceeding about 400 V. The electrons emerge from the hole in $a_1$ as a divergent beam of moderate velocity and come within the field of $a_2$. This second anode is a cylinder maintained at a potential of some 600–1,200 V. It accelerates the electrons and alters the divergency of the beam.

What happens precisely depends on the design of the tube, but the general tendency is for the beam to be nearly parallel when it emerges from the second anode. It then comes within the third anode which is maintained at some 3,000–6,000 V; the electrons are further accelerated and the beam is made convergent.

It is the second anode which is the important one from the point of view of focusing and in practice its potential is made adjustable

* Some magnetic tubes have two anodes and are often spoken of as possessing a tetrode gun instead of the more usual triode. The first anode is operated at some 200–400 V and the second at 5,000–10,000 V.

and is quite critical for good focusing. Varying the potential affects the divergency or convergency of the beam at its point of entry into $a_3$ and so affects the distance the electrons must travel before their paths meet.

The action of these electrodes upon an electron beam is closely analogous to the action of lenses on a light beam and the subject is commonly called electron optics. On this basis the field between the grid and the first anode can be considered as a lens, and the fields between $a_1$ and $a_2$ and between $a_2$ and $a_3$ as second and third lenses. In the three-anode tube it is these last two which have the major effect.

As in the magnetic tube, the brightness of the spot and its size depend on the third-anode voltage, which should be as high as possible. The second-anode potential is critical for focusing, but the first-anode potential is in no way critical. The brightness increases with first-anode voltage, but an excessive voltage is likely to cause a reduction in tube life.* There is always likely to be a certain number of positive ions in the cathode region which travel to the cathode. Their effect on striking the cathode depends on their velocity which in turn depends on the potential of the first anode. A low potential is thus advisable in the interests of tube life.

In some of the latest electrostatic tubes a somewhat different electrode structure is adopted and the first and third anodes are operated at the same potential—3,000–6,000 V. The second anode is at a lower potential which is adjustable for focusing.

The picture signal is, of course, applied to the grid just as with a magnetic tube. It is, however, possible to apply it as well to the first anode. With this dual control an appreciable increase of modulation sensitivity is secured which can be quite valuable—especially with some of the older tubes needing a signal input of 60 V or so.

Immediately beyond the third anode come the deflecting plates. There are two pairs—one for the line and the other for the frame deflection. The frame plates are mounted horizontally and the application of a potential between them causes the beam to be deflected vertically. A potential of saw-tooth waveform with a periodicity of 50 c/s is used in practice.

The line deflector plates are mounted vertically and so cause a horizontal deflection of the beam. They are fed with a saw-tooth voltage wave of 10,125 c/s. The general arrangement of the plates and electrode system is sketched in Fig. 3.7.

One of the main disadvantages of electric deflection is that the deflector plates are not content with deflecting the beam, but act to some degree as electron lenses and so affect the focusing. For the optimum results it is necessary for the mean potential of each pair of plates to be constant and at the same mean potential as the electron beam between the plates.

* "Electrostatic Focusing," by O. Parr. *Wireless World,* April 2, 1937.

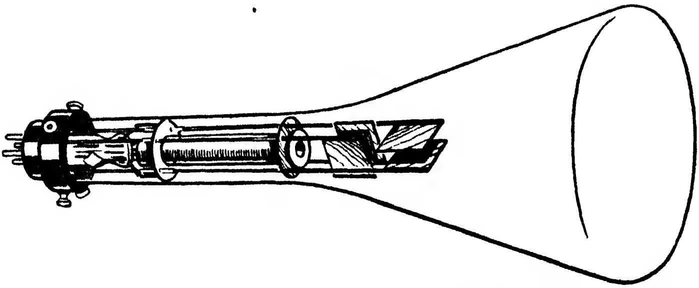


**Fig. 3.7—General arrangement of the electrodes of an electrostatic tube**

At one time, one plate of each pair was connected to the third anode and the saw-tooth potentials were applied to the other plates. The instantaneous potential of the inner plate then affects the deflection sensitivity of the outer pair and a rectangular raster is not obtained. The deflection sensitivity is inversely proportional to the third-anode voltage and from the point of view of the outer pair of plates the inner pair can be regarded as an extension of the third anode.

Now if one of the inner pair is connected to $a_3$ and the saw-tooth voltages are applied to the other, the instantaneous potential of this plate will vary about the third-anode potential. When it is below it, the sensitivity of the outer pair of deflecting plates will be abnormally high and when it is above it the sensitivity of the outer plates will be below normal. The practical result is that opposite sides of the raster are not of the same length—there is trapezium distortion.

In some tubes this effect has been overcome by special design, but in general it is necessary to apply the saw-tooth deflecting voltages in push-pull so that when one plate of a pair swings positive the other swings negative by an equal amount. At all times the mean potential of each pair of plates is then constant and trapezium distortion is avoided.

It is quite common to make the mean potential of the plates the same as the third-anode potential. It often happens, however, that the focusing varies considerably over the raster. It may be good at the centre of the picture, but very poor towards the edges. This defect can often be greatly reduced, if not overcome, by applying a bias voltage to the plates so that they are at the potential of the electron beam at the point where it passes between them. Depending on the design of the tube and its operating conditions the required bias may be positive or negative with respect to the third anode. The value required rarely exceeds 200 V and is usually considerably less.

It is usual to provide controls whereby one plate of each pair can be biased positively or negatively with respect to $a_3$ for the purpose of centring the raster on the tube. By providing such controls with a greater range of voltage and by providing a second pair for the other two plates, the mean potential can be varied as well.

The deflection sensitivity of a tube is usually given by an expression such as $900/V_{a3}$ mm/volt. This means that for every volt of potential difference between a pair of deflector plates the deflection of the spot on the screen is $900/V_{a3}$ mm. If $V_{a3}$ is 3,000 V, then the deflection is 0·3 mm. With a picture having a 10-in side, the deflection required is 254 mm, so that the deflecting voltage must be $254/0{\cdot}3 = 847$ V p–p.

Tubes vary in their sensitivity and the anode voltage also varies. In general, however, the deflecting voltages needed lie between 800 V and 2,000 V p–p, probably the commonest being 1,000–1,200 V.

It should be noted carefully that a small tube does not necessarily want a smaller deflecting voltage than a large one. If it is of the same general design it will probably need the same deflecting voltage. This is because the deflection angle remains unchanged, for a smaller screen diameter is usually accompanied by a proportionate decrease in length.

The operating voltages for a small tube are usually also of the same order as those of a large one, so that its use often effects no saving in the cost of the associated apparatus.

Ignoring for the moment any saturation effects in the fluorescent screen, the brightness of the picture is proportional to electron velocity and beam density, and is consequently proportional to the square of the anode voltage. The amount of light required for constant apparent brightness is proportional to the square of the tube diameter, since the picture area varies as the square of the tube diameter. We thus find that the anode voltage should be proportional to the tube diameter. If this could be adhered to the deflecting voltage needed would become proportional to diameter. A small tube would then offer a big saving in cost over a large one.

In practice, however, the rule cannot be followed because the spot size tends to increase as the anode voltage is lowered. For equally good definition, the spot size should be a constant fraction of the tube diameter and so should become smaller as the tube becomes smaller. To obtain this condition a higher voltage is needed with a small tube than with a large one, assuming similar tube design in each case.

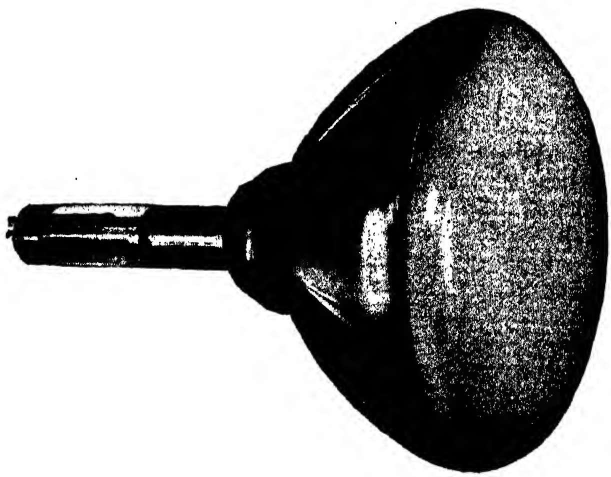
In general, a brighter picture can be secured for a given anode voltage from a magnetic tube than from an electrostatic. This is because it is easier to focus a bigger beam magnetically than it is electrostatically; consequently the tube can be designed for a bigger beam current for the same anode voltage.

With the modern 9-in magnetic tube 5,000 V gives sufficient brightness for all normal requirements and the picture is satisfactory for daylight viewing as long as bright light is not allowed to fall directly on the screen. At one time 5,000 V was considered sufficient for a 12-in tube, but now operation at around 9 kV is more usual.

There is a marked tendency towards the use of higher voltages and it is not uncommon to find 6–7 kV used for a 9-in tube (and sometimes even 9 kV) and up to 14 kV is commonly used with 15-in and 16-in tubes.

The fluorescent material used also affects the brightness and for a given material the thinner the coating used the brighter the picture. It is the inside of the coating which is struck by the electrons of the beam and which consequently fluoresces most brilliantly. The picture is actually viewed through part of the coating and this is why it looks brighter when one looks at the back of the screen.

The colour of the fluorescence depends chiefly upon the screen material and this is usually chosen to fluoresce as white as possible to give a black and white picture. These so-called white screens



**Ediswan 9-in electromagnetic tube**

are usually slightly blue. Other colours are possible and a yellow material giving sepia-tone pictures is sometimes adopted, while many so-called white screens have a slight greenish tinge.

The electron velocity has quite a marked effect upon the picture colour. A tube operated at a high voltage may give a good black and white picture, but as the voltage is reduced the colour is likely to change and acquire a brownish tinge.

It is clear, therefore, that it pays from the point of view of picture quality to use as high a voltage as possible for operating the tube; that is, up to the maker's maximum rating. Not only is a brighter picture secured but the picture colour is better and the spot size is likely to be smaller and so permit better definition. The use of a higher voltage is especially important with electrostatic tubes.

Instead of viewing the picture directly on the end of the tube, or through a mirror, a projection arrangement is sometimes used.

This is adopted when a larger picture is required than it is practicable to obtain with direct viewing. The tube used varies from about 2·5-in diameter to about 6-in, the smaller ones being employed to give a picture size of the order of 20 in by 15 in and the larger to provide a full-size cinema screen picture. A Schmidt optical system is now invariably used. This comprises a concave spherical mirror and a specially-shaped correcting plate, which is often made of plastic material. Depending on the mechanical layout one or more plane mirrors are often added and the picture is viewed on a flat screen, being usually projected on to it from the back. The smaller varieties of projection tube are operated at about 25 kV and the larger types at about 50 kV.

The picture size required depends entirely upon the distance from which it is viewed. For domestic use a picture much larger than 10 in × 7·5 in is too large. Experience shows that a picture of this size should be viewed at a distance of not less than five feet if the line structure is not to be prominent.

The minimum viewing distance for other tubes can be computed as it is directly proportional to picture size or tube diameter. The minimum distance for a 20-in × 15-in picture is thus about ten feet and is too great for the ordinary sitting-room, although it may be ideal for those possessing very large rooms.

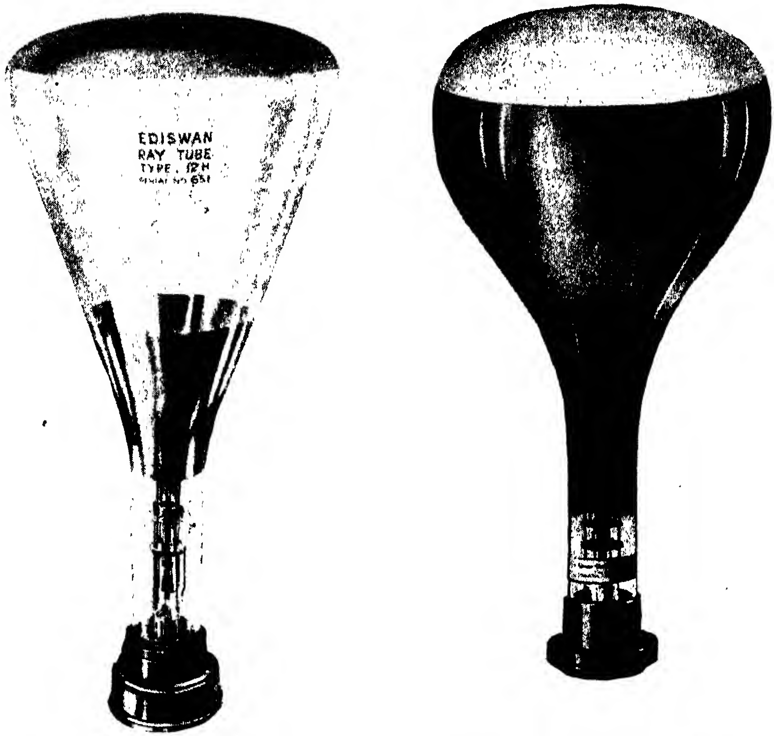
For those with small rooms a 12-in tube is unnecessarily large, and a 9-in or 10-in tube is extremely satisfactory. Smaller tubes are undoubtedly less desirable and although good pictures can certainly be secured, careful grouping of the viewers is necessary.

The general practice is to mount the tube horizontally so that its screen is viewed directly. As the total depth of the equipment must be somewhat greater than the length of the tube, it is necessary for the tube to be reasonably short if the cabinet depth is not to be excessive.

In the early days of television, tubes were undoubtedly long and a vertical mounting was consequently adopted, the screen being viewed through a surface-silvered mirror carried by the lid of the cabinet. The drawbacks to this were the cost of the mirror and the restriction of the viewing angle in the vertical direction; furthermore, a surface-silvered mirror tends to deteriorate fairly rapidly.

Shorter tubes are now used so that horizontal mountings can be adopted, giving direct viewing, with cabinets of reasonable size. All is not gain, however, for a short tube has a greater deflection angle than a long one for the same size of screen, and a greater output from the time bases must be provided. The electrical side is thus sacrificed to the mechanical.

It should be noted that electrostatic tubes are now obsolete for television purposes, save in sizes which are too small for most requirements. They are still useful in sizes up to 6 in, but for anything larger electromagnetic tubes are invariably employed. The usual sizes are 9 in and 12 in but some makers adopt 10 in and 14 in. There are also a few 15-in, 16-in, and even 20-in tubes.




**Types of electrostatic tube once widely used for television : Left, the Ediswan 12H ; right, the Cossor 3272**

Before concluding, some mention should be made of a common defect in electromagnetic tubes—namely, ion burn. No tube can be perfectly evacuated and there are always some ions present. Negative ions are accelerated towards the screen in a similar manner to electrons, but as their mass is much greater they are affected differently by the focusing and deflecting fields. In an electrostatic tube there is considerable focusing and deflecting action; the ions are then distributed more or less evenly over the screen and have little harmful effect. In a magnetic tube, however, there is little focusing or deflecting action and nearly all the ions fall on the middle of the screen. After a time they produce an ion burn on the screen. This usually appears as a circle of about 2-in diameter with quite a sharp edge and of a yellowish colour. It is only visible when the tube is in operation and it does not usually appear in under six months of normal operation.

Some tubes have been produced which incorporate an ion trap and which are thus free from the defect. The gun is set at an angle so that it does not point directly towards the screen. External permanent magnets are used to deflect the beam of electrons and bring it on to the axis of the tube so that subsequent focusing and deflection can take place as usual. The ions, however, are not deflected on to the axis by this field, and so do not reach the screen. The method is not much used, because it complicates the tube construction considerably.

Another method which is being increasingly adopted is to employ an aluminized screen. In this a very thin layer of aluminium is deposited on the inside face of the screen. The electrons penetrate this layer and energize the fluorescent screen in the usual way, but the more massive ions are intercepted by it. The method has a further advantage in that the face of the aluminized layer adjacent to the screen acts as a mirror and reflects the light from the inside of the screen outwards. Some of the light from the inside face of the screen is thus usefully employed instead of being wasted.

The gain in brightness depends on the voltage used and at low voltages the tube is actually poorer than the ordinary kind. The two types are usually equal in brightness at voltages around 4,000. It is usually considered that the benefit of aluminizing is not secured unless the tube is operated at 6,000 V or more, but some makers do use lower voltages. Whatever the operating voltage the great benefit of freedom from ion burn remains.

CHAPTER 4

# Cathode-ray Tube Voltage Supplies

APART FROM THE signal and deflecting voltages and currents, which are dealt with later, the cathode-ray tube requires certain fixed voltages for its operation. A supply of some 3,000–10,000 V is needed for the final anode, lower voltages for the other anodes, if any, and grid bias; there is also the heater.

Dealing first with this last element, most modern tubes have indirectly-heated cathodes. The heater is usually rated for 2, 4 or 6·3 V and this last is becoming increasingly common.

The tube-heater supply is arranged in much the same way as that of a valve. With a magnetic tube its cathode is usually within 100 V of earth and it is then common to connect it in parallel with the valve heaters and run them all from one winding on the mains transformer. They must, of course, then be all of the same voltage rating. Sometimes a separate winding is used for the tube and this is usually necessary when it has a different heater-voltage rating from the valves. In a.c./d.c. sets the tube heater is connected in series with the valve heaters and then all must have the same current rating; close regulation of the heater current is often needed and in addition a thermistor is often used to prevent excessive current when the set is first switched on and the heaters are cold.

Some older tubes have cathode and heater internally joined and a separate heater winding is then essential. Care must then be taken to join external circuits to the common heater-cathode terminal. If they are joined to the other heater lead serious hum will occur and will take the form of a dark band across the picture.

No heater supply difficulties occur with any normal tube when the negative of the e.h.t. (extra-high-tension) supply is earthed or earthy. This is invariably the case with magnetic tubes, but with electrostatic tubes it is sometimes desirable to earth the positive of the e.h.t. supply. The tube heater and cathode are then below earth by the amount of the e.h.t. voltage and the heater winding must be insulated accordingly.

With some tubes, especially directly-heated types, the filament or heater rating is based on current and the voltage required varies from tube to tube. It is then necessary to provide a transformer winding giving a somewhat higher voltage than the nominal figure for the tube, and to provide a variable series resistance so that the current can be adjusted precisely to the correct figure with the aid of an a.c. ammeter. Due allowance should be made for the resistance of the meter. Some types cause an appreciable voltage drop and the current will be higher when they are taken out of circuit.

Turning now to the high-voltage supply, magnetic tubes usually need only one voltage derived from this, for it is common to take grid bias from the receiver h.t. supply. The high-voltage unit is then extremely simple and commonly takes the form shown in Fig. 4.1.

A simple half-wave rectifier is used without smoothing equipment, for the current is so low that the voltage across $C_1$ is nearly equal to the peak a.c. input and the ripple is very small. The resistor $R_1$ is included merely to limit the current in the event of a short-circuit or a flash-over in the tube. A suitable value is 50,000 $\Omega$. $R_2$ is provided so that the capacitor $C_1$ is discharged after the apparatus has been switched off.

Without this resistor and given perfect insulation, $C_1$ would hold its charge indefinitely and even after a long period it would be capable of giving a dangerous shock. It is tempting to omit the resistance, because it draws current continuously and so makes the use of a large value for $C_1$ necessary. In practice, of course, $C_1$ is usually discharged by the c.r. tube even if $R_2$ is absent, for an indirectly-heated cathode goes on emitting for some time after its heater current is switched off. While it may thus be permissible to omit $R_2$ in a complete receiver, it must always be used in experimental equipment or where there is the slightest risk of the h.t. being switched on with the tube heater cold.

A suitable value for $R_2$ is 25 M$\Omega$ and it should be made up of several resistors in series to keep the voltage drop across each at a reasonable figure. In general, the voltage across any ordinary carbon or metallized resistor should not exceed 1,000 V and, preferably, should not be more than 500 V.

The value to be assigned to $C_1$ depends upon the total current and upon the amount of ripple tolerable. Nothing larger than 0·25 $\mu$F is likely to be needed and 0·1 $\mu$F will probably meet most requirements. The capacitor must be rated for continuous working at the peak value of the a.c. input; that is, 1·414 times the r.m.s. value.

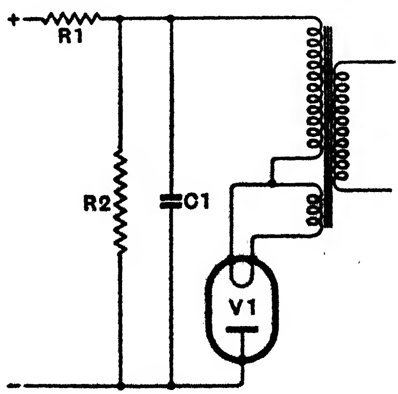



**Fig. 4.1—High-voltage supply unit for a magnetic tube**

The output voltage is lower than the peak value, but is near enough to it in practice for it to be satisfactory to choose the transformer winding on the basis of their equality. For an output of 6,000 V, therefore, one would choose a transformer winding of 4,250 V r.m.s.

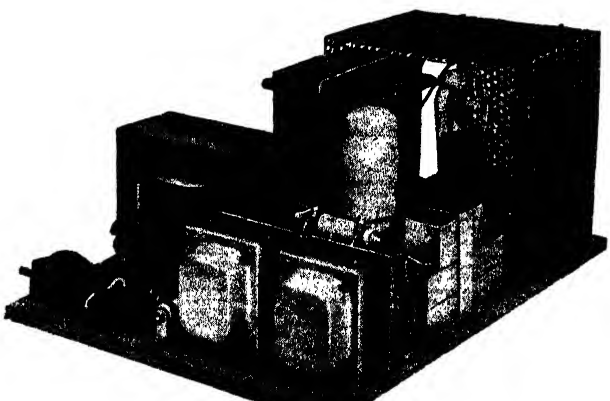
As the tube voltage is in no way critical, the temptation is to pick a transformer winding with a rating in round figures and take what output one gets. Thus one might decide on 2,500 V r.m.s.; the peak voltage is then 3,550 and the actual voltage on the tube might be 3,200–3,400 V. This is quite satisfactory until one comes to choose the capacitor and one then finds that these components are usually rated in 1,000-V steps.

One capacitor may be rated for 3,000 V and the next for 4,000 V. By working on the above basis, it often happens that it is necessary to use a higher voltage capacitor than would be needed with a different approach to the problem.

The cheapest course is usually to pick the capacitor and choose the tube voltage together. Thus, if the tube needs 3,000 V it will probably be satisfactory to choose a 3,000-V capacitor, and the transformer winding should then be 2,120 V r.m.s. The actual voltage on the tube will be about 2,800–2,900 V.

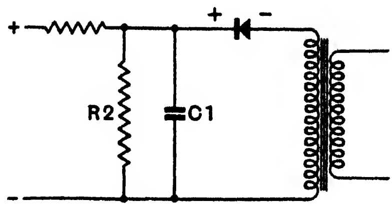
If a good quality capacitor is chosen there is little likelihood of trouble. Most difficulties in high-voltage units arise through the use of cheap mains transformers. Money spent here is well spent, for a poor component will soon break down. When the rectifier is non-conducting on the negative half-cycles of the input, the voltage across it is equal to twice the normal peak voltage of the transformer. Actually it is equal to the peak a.c. voltage plus the voltage across $C_1$, and it is called the peak-inverse voltage.

It not only appears across the valve but also between the filament winding and the core and primary of the transformer. With an



**Experimental power unit for an electromagnetic television receiver**

**Fig. 4.2—With a metal rectifier the peak-inverse voltage is removed from the transformer**




output of 6,000 V across $C_1$ the peak inverse is about 12,000 V and will quickly find out any weak spots in the transformer. It is by no means negligible at lower outputs, for at the normal to low voltage of 3,000 across $C_1$ it is 6,000 V.

A transformer made by someone who does not allow for this high voltage in its design may work well on open circuit and show no signs of a breakdown, but as soon as the rectifier is inserted trouble will start. It may break down at once, or only after a few hours or days, but break down it will sooner or later.

Even good quality transformers break down occasionally, of course, and the author had one with a winding of 1,750 V only which broke down after a year's use. In spite of the windings being spaced from the end cheeks and of a screen between them, a discharge between primary and secondary started and followed the surface of the cheeks. At length it broke down the insulation between some of the primary turns. The primary fuses then blew.

Conditions in the transformer are much better if the positive lead is earthed instead of the negative, for the peak inverse then appears only across the valve and not between transformer windings. The use of a metal rectifier removes the peak inverse from the transformer and with a negative earth it should be connected as shown in Fig. 4.2. When really high voltages are involved, and sometimes with only moderate voltages, it pays to use two rectifiers in the voltage-doubler circuit, for the transformer voltage is then halved.

The grid bias for the c.r. tube can be derived either from the high-voltage supply or from the receiver h.t. supply. When the former course is adopted the circuit is arranged on the lines shown in Fig. 4.3, the grid of the c.r. tube being returned to negative h.t. through a resistance.

The capacitor $C_2$ is necessary for smoothing purposes and should be of the order of 0·5 $\mu$F, although a smaller capacitance will often be satisfactory. The variable resistor $R_3$ controls the tube bias. The maximum bias voltage available is the full rectifier output multiplied by $R_3/(R_2 + R_3)$ and the tube anode voltage, relative to cathode, is the full output less the bias voltage.

If the supply is 3,000 V and a maximum of 100 V bias is needed, then $R_3/(R_2 + R_3) = 1/30$ or $R_2 = 29\ R_3$. If $R_3$ is made 0·5 MΩ then $R_2$ should be 14·5 MΩ. In practice, it is unnecessary to pick the values accurately, but they should always be selected so that ample bias is available to black out the spot under any conditions.

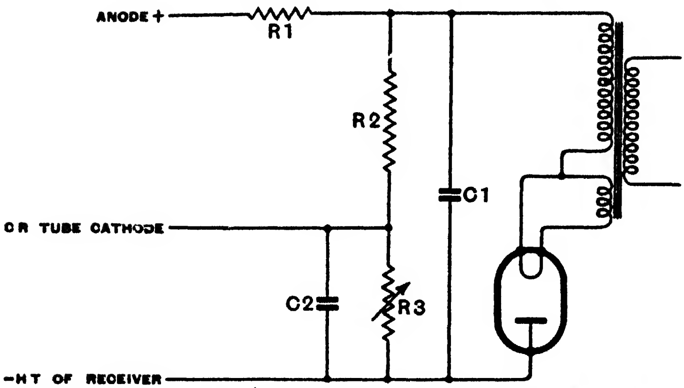



**Fig. 4.3—Method of obtaining the tube grid bias from the high-voltage supply**

When the tube bias is obtained from the receiver h.t. supply the circuit of Fig. 4.1 is used, and if desired, the negative lead can be connected to the *positive* of the receiver supply. The grid return lead is then taken to a potentiometer across the receiver supply.

This system has the advantage that there is no loss of voltage through the biasing arrangements, but the disadvantage that the tube will lose its bias if the receiver supply fails but the high-voltage supply is still operative. It is, therefore, safer to derive the h.t. and bias from the same source, as in Fig. 4.3.

When electrostatic focusing is used, the high-voltage supply is more complicated and the basic arrangement is shown in Fig. 4.4. The rectifier circuit is the same, but following the reservoir capacitor $C_1$ there is a smoothing circuit comprising $R_1$ and $C_2$. The capacitor is usually about 0·1 $\mu$F and must be rated for working at the full voltage of the supply, while $R_1$ is of the order of 0·1–0·25 M$\Omega$. The resistors $R_2$ to $R_6$ are chosen to give the required voltages for the various anodes. If the total resistance is made of the order of 2 M$\Omega$ the current drawn by the tube is sufficiently small to be ignored in comparison with the potential-divider current. The resistance values can, therefore, be proportional to the voltages. Incidentally, the smoothing resistor $R_1$ should be in the unearthed h.t. lead; that is, it should be in the positive lead, as shown, when the negative is earthed, but in the negative lead when the positive is earthed.

At one time it was the standard practice to earth the positive. This was done largely because the procedure with oscilloscope tubes was followed, and a positive earth is there necessary as the deflector plates must be available for external connection.

In the case of television a positive earth has the advantages that with a valve rectifier the peak-inverse voltage does not appear in the mains transformer, and the four coupling capacitors between the time bases and the deflector plates need be rated for working at the time-base voltage only. The disadvantages are that the heaters of the c.r. tube and the d.c. restoring diode, and their wiring, must be insulated to withstand the full anode voltage of the tube, and the coupling capacitor from receiver to tube grid must also work at this voltage.

There is also a safety point to be considered. Long experience with receivers has given one the idea that heater, cathode and grid circuits are always at a potential quite near to earth and may be handled with impunity. When using a high-voltage supply with the positive earthed it is quite difficult to remember that it is the heater, cathode, and grid of the tube which are live. It is, therefore, essential that all these circuits be boxed up so that one cannot come into contact with them.

Even with a negative earth, of course, the high-voltage parts must be adequately protected so that accidental contact with them cannot be made. There is, however, less danger if the protection is not perfect, because one subconsciously expects the anode circuits to be live. With the negative earth, the four coupling capacitors must work at the full anode voltage, but the receiver coupling capacitor to the tube grid can be of low-voltage rating. With some circuits, in fact, it can be omitted. Special insulation of the heaters of the tube or d.c. restoring diode is unnecessary.

Before making a choice, it is also necessary to consider the effect of a breakdown in a high-voltage capacitor. With a positive earth a breakdown in the grid capacitor would apply the full third-anode voltage to the grid, with disastrous results to the tube. With a negative earth, a breakdown in a deflector-plate coupling capacitor would result in the appearance of the full voltage between the plate and the third anode and the other plates. A flashover might occur, but in all probability there would be no harmful effects on the tube.

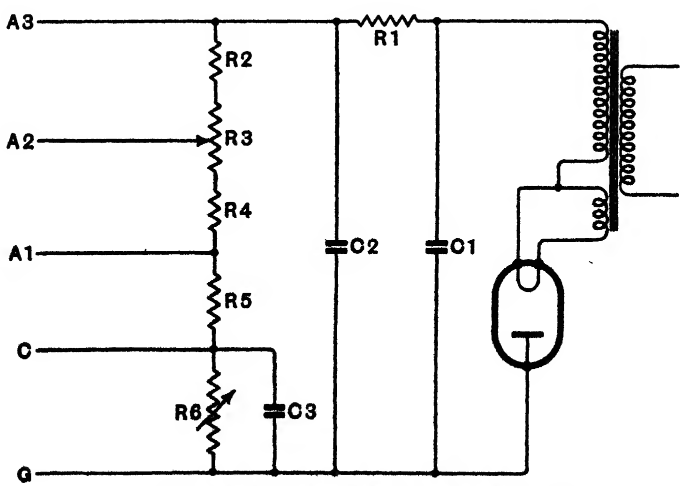



Fig. 4.4—E.H.T. unit for an electrostatic tube

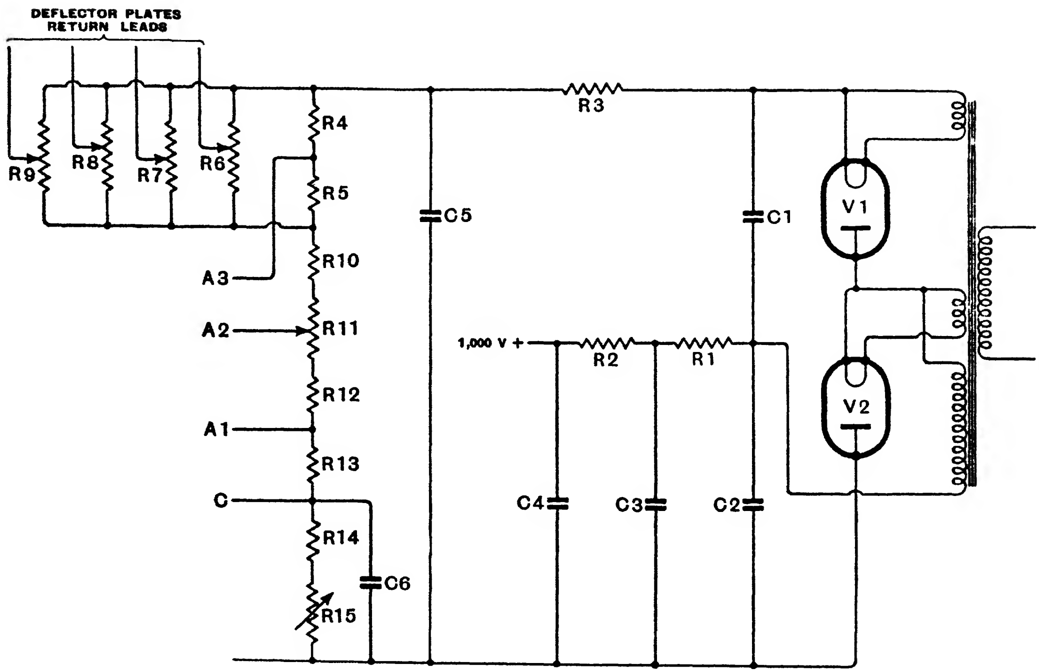



Fig. 4.5—Circuit of a power unit for an electrostatic tube; it also supplies h.t. for the time base

On balance, therefore, the author favours a negative earth and this system has the further advantage that it is sometimes possible to derive the time-base h.t. supply from the high-voltage unit. For electric deflection the time base generally needs about 1,000 V at some 20 mA.

Unless it is combined with the high-voltage supply, the whole vision equipment will need three power packs; one for the receiver with an output of the order of 100 mA at 250 V, one for the time base giving 20 mA at 1,000 V and one for the tube giving some 2 mA at 3,000–6,000 V. By adopting the voltage-doubler circuit for the high-voltage rectifier it is possible readily to combine the tube and time-base rectifiers.

The arrangement which the author has used most successfully is shown in Fig. 4.5. The transformer winding is 1,750 V at 30 mA, and the peak inverse is limited to 5,000 V. The two rectifiers, of the U17 type, are in the voltage-doubler circuit with $C_1$ and $C_2$. The full output, some 4,500 V, is used for the tube supply and is smoothed by $R_3$ and $C_5$.

The time-base supply is taken from the junction of $C_1$ and $C_2$, the voltage across $C_2$ being of the order of 2,000 V. The voltage across $C_2$ is less than that across $C_1$ because of the heavier current drawn from this point. This lack of balance might lead one to suppose that the arrangement would be in some way unsatisfactory, but in practice it works well and is quite reliable.

As the voltage across $C_2$ is greater than is needed by the time base, resistance smoothing is used with $C_3$, $C_4$, $R_1$ and $R_2$. If delayed switching of the h.t. supply is used, so that the mains are not connected to the transformer primary until the time-base valves have had time to warm up, the voltage ratings of the capacitors can be graded to suit the normal working voltages.

On the high-voltage side, four shift potentiometers are shown. These are $R_6$, $R_7$, $R_8$, and $R_9$ and enable any deflector plate to be biased positively or negatively with respect to the third anode. Two such potentiometers are normal and enable the raster to be centred on the screen.

The use of an extra pair, however, enables the mean potential of each pair of plates to be varied by the simultaneous adjustment of two controls while still keeping the picture central. As explained elsewhere, this bias often enables a distinct improvement in focusing to be obtained.

It is not necessary to use a special transformer for the e.h.t. supply, for it is possible to obtain it quite economically from the 350–0–350-V winding of the transformer which supplies the h.t. for the receiver. This can be done with a voltage-multiplier circuit* which is particularly well adapted to metal rectifiers.

A suitable circuit is shown in Fig. 4.6 attached to a normal h.t. rectifier circuit. This comprises the transformer T with a 350–0–350-V winding connected to the rectifier $V_1$; it produces an

* Television E.H.T. Supply, by A. H. B. Walker. *Wireless World*, April and May 1948.

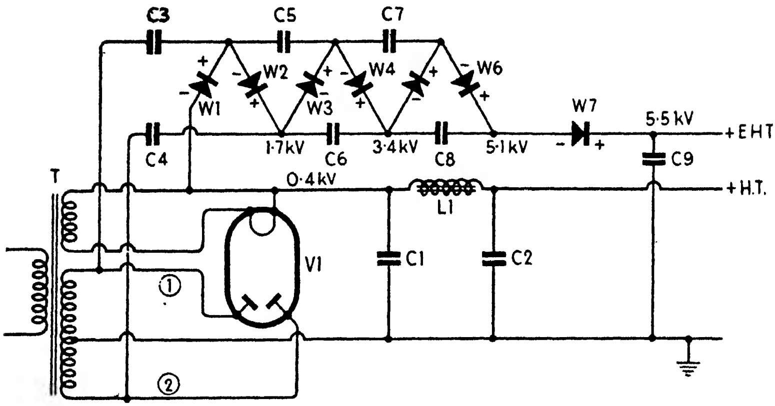



**Fig. 4.6—Circuit of voltage-multiplying system producing an e.h.t. supply of 5·5 kV from a 350–0–350-V transformer which also feeds the h.t. rectifier $V_1$ for the receiver h.t. supply of 400 V**

output of some 400 V at a current suited to the needs of the receiver, usually of the order of 200 mA. The components $C_1$, $C_2$ and $L_1$ are the usual reservoir capacitance and smoothing circuit and in no way affect the e.h.t. circuit.

This last is the upper part of the diagram from $C_3$ and $C_4$ onwards. The peak voltage on the half-secondary of T is $350\sqrt{2} = 500$ V. During the negative half-cycle " 1 " reaches a peak of 500 V below earth and this is applied in series with the 400 V from the h.t. supply to $C_3$ and $W_1$. The latter conducts and $C_3$ charges from the 900 V acting in the circuit to about 850 V.

During the next half-cycle " 1 " is 1,000 V positive with respect to " 2 " at the peak and this is applied in series with the 850 V across $C_3$ to $C_4$ and $W_2$ so that $C_4$ charges to nearly 1·85 kV—actually about 1·7 kV. On the next half-cycle the 1 kV across " 1 " and " 2 " is applied in series with the voltages on $C_3$ and $C_4$ to $C_5$ and $W_3$.

The whole process is continuous and $C_3$, $C_5$, $C_7$, $C_9$ all charge on the peak of one half-cycle while $C_4$, $C_6$, $C_8$, $C_9$ all charge on the peak of the next. The voltage gradually increases along the chain but the individual components can be of fairly low-voltage rating except for $C_9$. The capacitors $C_3$ to $C_8$ can all be of 0·1 $\mu$F and the voltage on none exceeds 1,000 V. The peak-inverse voltage on the two end rectifiers $W_1$ and $W_7$ is about 1 kV, but that on the others some 2 kV. The reservoir capacitor $C_9$ of 0·1 $\mu$F capacitance, must be rated for 6 kV.

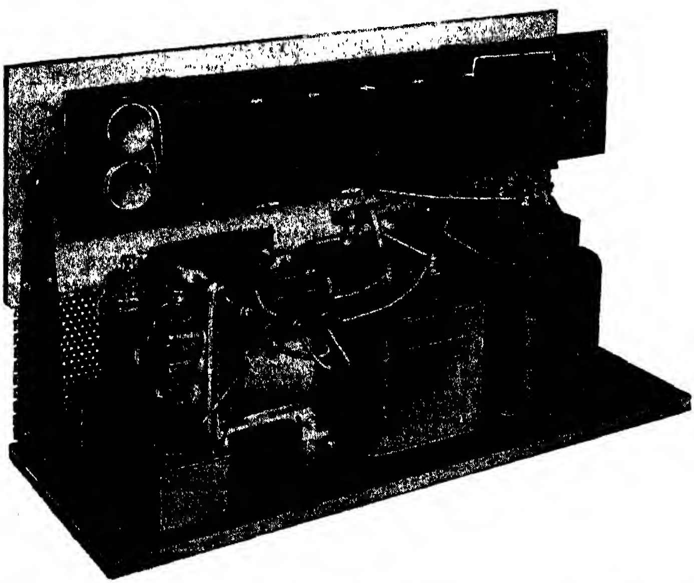
The system has been produced commercially as a unit comprising everything from $C_3$ and $C_4$ to $W_7$ inclusive and needs only connecting to the transformer and $C_9$ to produce an e.h.t. supply. This is the Westinghouse Westeht unit with an output current adequate fot the needs of a magnetic tube.

If a supply negative with respect to earth is required it is only necessary to reverse the connections to each rectifier and to return $W_1$ to earth instead of positive h.t. All voltages in the e.h.t. circuit are then 400 V less than the figures marked on the diagram and negative with respect to earth. This arrangement, with fewer multiplying stages, makes a particularly convenient way of obtaining 1·5–2 kV for an oscilloscope.

All 50-c/s e.h.t. supplies have one drawback, they need a reservoir capacitor of the order of 0·1 $\mu$F. This is bulky, weighty and expensive; of even greater importance is the fact that it stores a dangerous amount of energy.

The danger from a shock depends on the magnitude of the current, for how long it flows, and through which part of the body it passes. It is often said that a current of 25 mA is fatal in the right (or wrong!) place, but this is for a continuous flow. It is, in fact, quite possible for a 100-V supply to be fatal, but is rare for anyone to make good enough contact with such a low voltage for the current to reach a dangerous value.

With a capacitor of 0·1 $\mu$F charged to 5 kV it is unlikely that a fatal shock would result from accidental contact with it, but it is possible, especially if one is in a poor state of health. In any case the results are extremely unpleasant, as the writer can say from personal experience. The e.h.t. supply itself is not particularly dangerous because the circuit resistance is usually high and the continuous output is limited to a few milliamperes. In this respect,



**High-voltage unit supplying an electrostatic tube and its time base. The circuit is similar to that of Fig. 4.5**

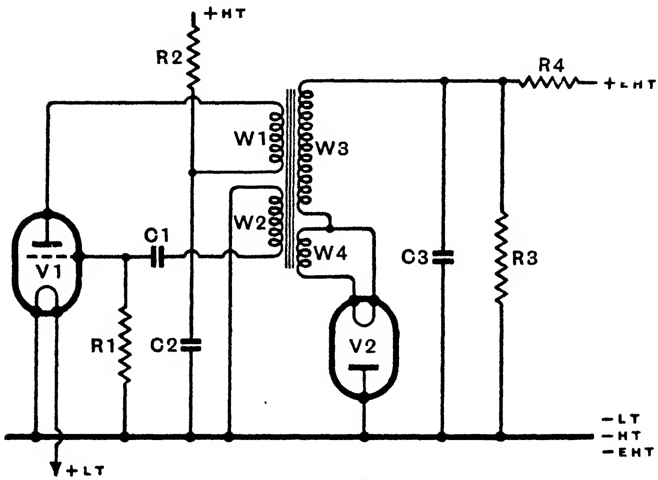



**Fig. 4.7—The circuit of an e.h.t. supply unit, which is suitable for battery operation, is shown here**

voltage multipliers are safer than a single rectifier and high-voltage transformer. It is a large reservoir capacitance which can be dangerous and it is a wise safety measure to fit a bleeder and then to make doubly sure by placing the blade of a screwdriver across the capacitor terminals before working on the set. If the bleeder has failed and the capacitor is charged, shorting it will probably destroy it, but this is better than a 5-kV shock from it. With metal rectifiers a bleeder is not necessary, for their back resistance is low enough to discharge the capacitor. It is essential with valve rectifiers.

High-frequency power supplies are much safer, because the reservoir capacitance can be proportionally smaller and so much less dangerous. At 10 kc/s, for instance, 0·001 $\mu$F is ample and the energy stored in it is only one-hundredth. Partly because of this and partly for other reasons such supplies are coming into general use and will probably be universal in the future.

One method derives e.h.t. from an r.f. oscillator, which is fed from the receiver h.t. supply. It is frequently used to produce voltages up to 50 kV for projection television, but is it not common for e.h.t. supplies under 10 kV although it is quite satisfactory for any voltage. An alternative arrangement is usually simpler at the lower voltages, however.

The circuits employed may vary considerably in detail but have the general form shown in Fig. 4.7, in which $V_1$ is an oscillator operated from the receiver h.t. and l.t. supplies. $C_1$ and $R_1$ are given values appropriate to a power oscillator for the frequency selected, and $C_2$ and $R_2$ are merely decoupling components.

The e.h.t. rectifier is $V_2$ and its filament is fed from the winding

$W_4$, while $W_3$ supplies the high voltage. The rest of the circuit is normal with a safety resistor $R_4$ and a bleeder $R_3$. The reservoir capacitor is $C_3$. If the frequency is above 50 c/s, the capacitance can be proportionately reduced. Thus, if the transformer is designed so that the frequency is 500 c/s, $C_3$ need rarely be above 0·01 $\mu$F. The windings $W_1$ and $W_2$ form the oscillator circuit, but the frequency obtained depends largely on $W_3$, since the self-capacitance of this winding will be important.

This general circuit shows an iron-core transformer and this is suitable for frequencies up to about 10 kc/s. In practice, this type of supply is usually operated at around 50 kc/s and an air-core transformer is used. High-Q low-capacitance windings are needed and the unit must be very carefully screened to avoid radiation.

This type of circuit, usually elaborated with voltage-stabilizing arrangements and a doubling or even tripling rectifier system, is mainly used in projection apparatus. It is not yet common for supplies of under 10 kV. It is, however, worth noting that the method is a useful one for obtaining comparatively low voltages when only batteries are available for the power supply. For this reason, the circuit of Fig. 4.7 is drawn to show a battery-valve for the oscillator.

An alternative system which is sometimes used is the pulsed-choke circuit. In this an inductance, or the primary of a step-up auto-transformer, is connected in the anode circuit of a valve which draws a fairly heavy current through it from the receiver h.t. supply. This valve is cut off at intervals, usually once every scanning line, by the line sync pulse, and the rapid change of current through the inductance develops a large voltage across it which is applied through a rectifier to a reservoir capacitance.

The principle of the method is identical with that of the line fly-back e.h.t. systems referred to below and is not often used because the scanning circuits can provide e.h.t. as a by-product much more economically. However, the circuit has certain advantages when supplies of over 6 kV are needed, for as the e.h.t. and scan circuits are quite independent there is greater freedom in design.

The line fly-back system is becoming common and seems likely to supersede all others for domestic television sets. It is simple and inexpensive and as the reservoir capacitance need not be more than 0·001 $\mu$F it is a relatively safe supply. It is possible only with electromagnetic deflection and it is treated in Chapter 6 under that heading because an understanding of the circuit requires a detailed knowledge of the characteristics and operation of the magnetic time base.

CHAPTER 5

# Electric Deflection

ALTHOUGH ELECTRIC DEFLECTION is now obsolete from the normal television point of view, it was widely used in the early days of television. The material of this chapter is retained, substantially unaltered, from earlier editions, however, partly for the sake of completeness, partly for its historical interest, and partly because it is of value in circumstances other than the normal ones. Experimenters, for instance, sometimes use 6-in, or smaller, electrostatic tubes for reasons of economy.

The deflecting plates of a cathode-ray tube must be supplied with voltages of saw-tooth waveform. One pair of deflecting plates must have a voltage waveform repetitive fifty times a second for the frame or vertical deflection and the other pair needs a voltage repetitive 10,125 times a second for the line or horizontal deflection. These figures are, of course, for the present B.B.C. transmissions.

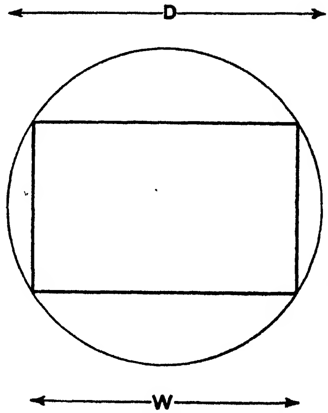
The amplitude of saw-tooth voltage needed is quite large and depends on the sensitivity of the tube and its anode voltage, and also upon the size of the picture. The sensitivity of the tube depends on internal factors such as the design of the deflecting system and its distance from the fluorescent screen; in other words, upon the deflection angle. The user can do nothing about this and has to accept what the designer of the tube gives him. In general, however, a long tube is more sensitive than a short one for the same size of screen, because the deflection angle is smaller for the same movement of the spot.

The makers quote a figure for the sensitivity of the tube and it is usually expressed in the form of $k/V$ mm per volt, where $k$ is a number and $V$ is the voltage applied to the final anode. The sensitivity is, of course, the distance in millimetres which the spot moves over the screen for one volt applied between the deflecting plates. The figure is for a steady deflecting voltage, but is the same for an alternating voltage of any waveform if the total amplitude is taken. This total amplitude is usually called the peak-to-peak value (abbreviated to p–p) and in the case of a sine wave it is twice the peak and 2·828 times the r.m.s. value.

A typical tube such as the Cossor 3272 has a sensitivity of $750/V$ mm per volt for the X-plates and $820/V$ mm per volt for the Y-plates, and the screen diameter is 305 mm (12 in). As the picture is wider than it is high it obviously pays to use the Y-plates for the horizontal deflection and this is, in fact, the usual practice.

Now the picture is rectangular and the screen is circular. If the whole of the picture is to be reproduced as large as possible its

**Fig. 5.1—Relation between tube and picture sizes**




corners must just touch the circumference of the screen as shown in Fig. 5.1. Denoting the screen diameter by $D$ and the picture width and height by $W$ and $H$ it is a matter of simple geometry to show that $H = D/\sqrt{1 + r^2}$ and $W = D/\sqrt{1 + 1/r^2}$ where $r = W/H$.

In practice, the standard of 4 : 3 has been adopted for $W : H$, so $r = 1{\cdot}33$, and the expressions $H = 0{\cdot}6\,D$ and $W = 0{\cdot}784\,D$ can be used. With a screen of 303 mm diameter, therefore, the picture size is 238 × 182 mm or 9·35 × 7·17 in.

As it usually happens that there is nothing of interest in the corners of the picture, it is not uncommon for the picture size to be increased somewhat, so that the corners of the picture are cut off. The extent to which one can go in obtaining a larger useful picture in this way is chiefly a matter of personal taste. In general, however, the amount of overlap allowed should decrease with the size of the tube.

In the author's view there should be no overlap with tubes larger than 12 in, a small overlap with 10 to 12-in tubes and more overlap still with 8 to 10-in tubes. With tubes of 6 to 8-in diameter the picture width should be equal to the screen diameter and with tubes smaller than about 5 in the diameter and picture height should be about equal. In this last case, there is naturally quite a big loss of picture area, but this is made up by the improved definition of what remains.

These suggestions are based on experience with average tubes and the recommendations with very small tubes are brought about by the fact that the spot size is usually *relatively* greater than with large tubes. If the spot size were proportional to the tube diameter, then it would be possible to include the whole picture on the screen with equally good definition. The spot size, however, is often little smaller with a small tube than with a big one, with the result that it is larger relative to the size of the picture. The definition consequently deteriorates and better results are secured by making the picture rather larger for the tube so that the spot size is relatively smaller.

A good picture size for a 12-in tube is 10 in $\times$ 7·5 in, or 254 $\times$ 191 mm. The tube sensitivity $S = k/V$ mm per volt, so that the deflection voltage required is $W/S = WV/k$ for the horizontal deflection and $HV/k$ for the vertical. Applying this to the Cossor 3272 tube we have for line—

$$\text{Deflecting volts p-p} = 254 \times 4{,}000/820 = 1{,}235$$

and for frame—

$$\text{Deflecting volts p-p} = 191 \times 4{,}000/750 = 1{,}020$$

assuming 4,000 V for the third anode.

The sensitivities of different tubes vary slightly, but the deflecting voltage rarely differs by more than a few hundred volts from these figures. Tubes of different size also need about the same deflecting voltage if they are of the same general construction because the deflecting angle is the same. With a smaller screen, for instance, the tube as a whole is usually shorter and the angle through which the beam has to be deflected remains the same. The deflectional sensitivity of small tubes is consequently lower than that of large ones under normal conditions.

We see that on the average the c.r. tube needs about 1,020 V p–p at 50 c/s for frame deflection and about 1,235 V p–p at 10,125 c/s for line. The saw-tooth waveform of the deflecting voltages must be as linear as possible over the scanning stroke and the fly-back must be completed within the time not occupied by the picture signals. In the case of the line scan, the time of the linear scanning stroke is 82·77 $\mu$sec and the fly-back time is 16 $\mu$sec. In the frame circuit the times are 18·62 msec and 1·38 msec respectively.

The scanning voltages are generated by a saw-tooth oscillator but the amplitude is usually only about 50–100 V. An amplifier is consequently needed and, as the deflecting voltages applied to the tube must be balanced to earth, this amplifier must be of the push-pull type.

The usual procedure is to use two triodes, one of which is fed from the saw-tooth oscillator and the other of which derives its input from the anode of the first in the manner of a paraphase circuit. Resistance-capacitance coupling is used and tne h.t. supply voltage must be high to obtain the necessary output. Roughly speaking, the h.t. voltage should be of the same order as the total peak-to-peak output required.

Amplitude distortion in the amplifier must be kept small, otherwise the wave shape of the scanning voltages will be affected. Frequency and phase distortion must also be small *and for the same reason.* In the time-base amplifier the $RC$ couplings cause frequency and phase distortion of the fundamental component of the saw-tooth waveform if the time-constant of the coupling capacitor and grid leak is not high enough. They also cause distortion of the

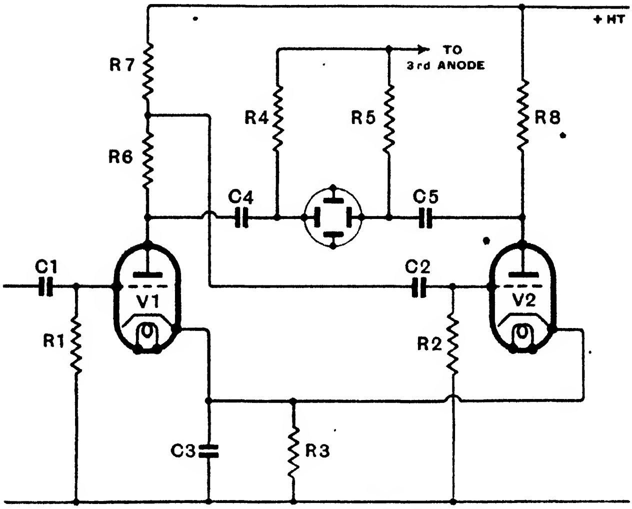


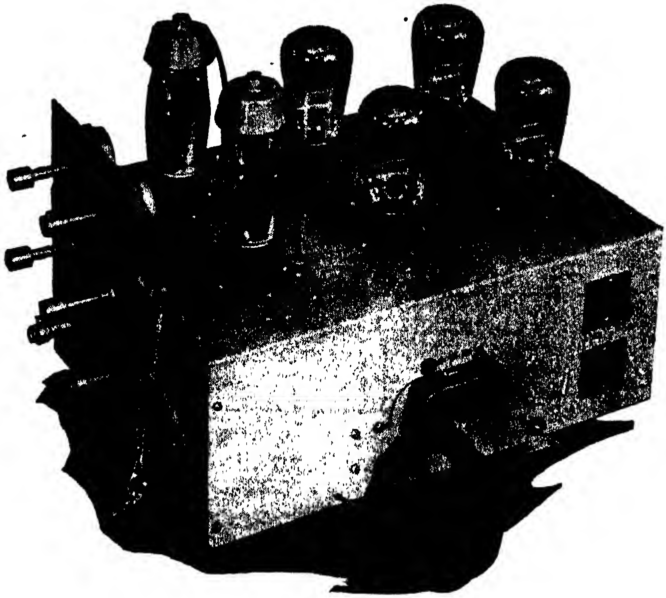

Fig. 5.2—Typical paraphase amplifier

higher harmonics of the fundamental if the time-constant of the coupling resistance and stray capacitances is too high.

The basic circuit of a time-base amplifier is shown in Fig. 5.2. The two valves have the same characteristics and are self-biased by $R_3$ in their cathode circuits. The coupling resistance $R_8$ is made equal to $R_6 + R_7$, which form the coupling resistance of $V_1$. The coupling capacitors to the deflecting plates $C_4$ and $C_5$ are made equal as are also the leaks $R_4$ and $R_5$. For equal output voltages from the valves, their inputs must also be equal, consequently only a portion of the output of $V_1$ is applied to $V_2$. Actually $(R_6 + R_7)/R_7$ must equal the amplification given by $V_2$, assuming that $R_2$ is much larger than $R_7$. There is a reversal of phase in each valve, consequently the output of $V_2$ is in opposite phase to that of $V_1$, and the instantaneous anode potentials of the two valves are equal and opposite with respect to the mean value.

In the case of the frame amplifier the high-frequency response is not important, because any practicable values of components in conjunction with any likely values of stray capacitance will cause a negligible loss of response at frequencies of the order of 1,000–2,000 c/s. In the line amplifier, however, the response must be maintained up to about 100,000 c/s and stray capacitances become very important. They place a limit to the values of coupling resistance and valve resistance which can be used.

It is unnecessary to enter into a detailed discussion of the design of time-base amplifiers here because it is carried out in exactly the same way as with any ordinary resistance-coupled amplifier for audio-frequency purposes. Having picked suitable valves and



Time base for electric deflection using gas-triode saw-tooth oscillators and triode amplifiers

estimated the stray capacitances the coupling resistances can be determined for the required maximum frequency response. The choice of grid leaks and coupling capacitances follows, being determined by the lowest frequency required. Then the valve operating conditions are selected for the required output.

It will generally be found that the mean anode potential of the valves is of the order of 500–600 V which is considerably above the rating of most small triodes. In spite of this the author has used the Mazda AC/P valves without any trouble and for long periods. This firm, however, lists a special valve for the work; it is the AC/P4 and has a top-anode connection and is rated for 700 V on the anode.

The makers give very full data for the use of this valve in time bases. As regards circuit values, of course, this data applies only to the AC/P4, but in other respects much of it is applicable to other valves, such as the AC/P. In the first place, individual valves vary in their characteristics with the result that if an amplifier is designed so that it will just give the required output disappointment may result when it is put into service. The output obtained may be appreciably above or below that expected, and in any case, will tend to fall off as the valves age.

Unless a considerable factor of safety is allowed, therefore, the use of a fixed tapping point for the input to $V_2$ of Fig. 5.2 is inadvisable. The best course is probably to make $R_7$ a potentiometer of rather higher value than the normal and feed $V_2$ from its slider. The amplifier can then be exactly balanced with very little trouble.

From the point of view of mains hum, however, it is usually better to take the input of $V_2$ not from a tapping on the coupling resistance of $V_1$ but from a tapping on the grid leak of $V_2$. This alternative method of feeding $V_2$ is shown in Fig. 5.3, and it is clear that a smaller proportion of the hum-voltage on $V_1$ is transferred to the grid of $V_2$. Owing to the input capacitance of $V_2$, however, this circuit cannot readily be used in the line time base amplifier, but it is quite suitable for the frame amplifier.

The same advantages can be secured for the line amplifier by adopting the circuit of Fig. 5.4, which is the one recommended by the valve makers for the AC/P4. The circuit is shown including a gas-triode saw-tooth oscillator for which the Mazda T31 would normally be used. Any other form of oscillator of suitable output can, of course, be used.

The first amplifier $V_2$ is conventional, but it will be noted that the paraphase valve $V_3$ is fed from the anode of $V_2$ through a small capacitor $C_8$. This capacitor with the input capacitance of $V_3$ forms a capacitance potentiometer across $R_{11}$. If the input capacitance of $V_3$ is $C$, then the input to $V_3$ is $C_8/(C + C_8)$ times the output of $V_2$. At low frequencies, however, the grid leak $R_{16}$ exercises an appreciable effect and the coupling becomes substantially $C_8$ and $R_{16}$. Any hum appearing on the anode of $V_2$ is greatly attenuated in its passage to the grid of $V_3$.

With full-wave rectification the hum frequency is chiefly 100 c/s

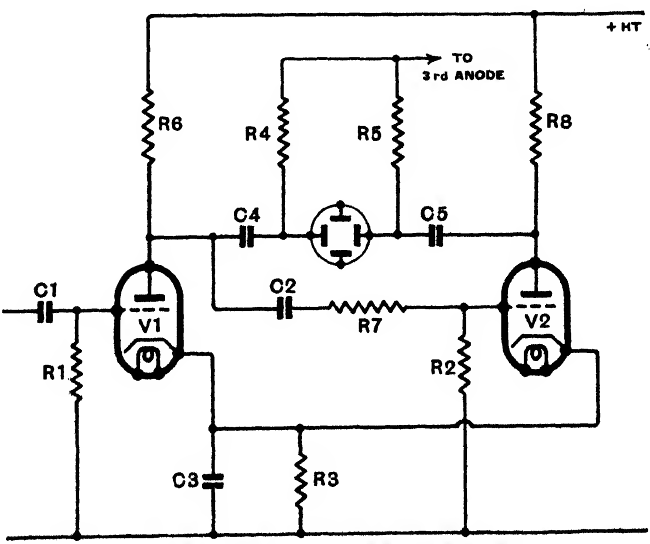
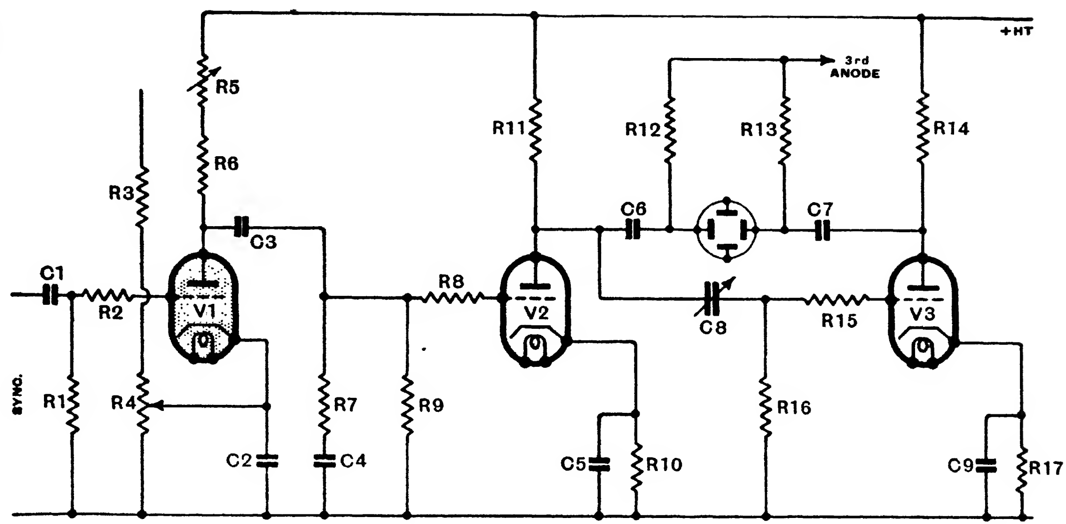



Fig. 5.3—Modified paraphase amplifier

**Fig. 5.4**—Self-compensating line time-base circuit (*courtesy The Edison Swan Electric Co. Ltd.*)




## TABLE 5.1

### Circuit Values for Fig. 5.4

$V_1$ = Mazda T31
$V_2 = V_3$ = Mazda AC/P4

$C_1 = 50$ pF
$C_2 = C_5 = C_9 = 0{\cdot}1\ \mu F$
$C_3 = 0{\cdot}005\ \mu F$
$C_4 = 0{\cdot}003\ \mu F$
$C_6 = C_7 = 0{\cdot}002\ \mu F$
$C_8 = 10$ pF max.

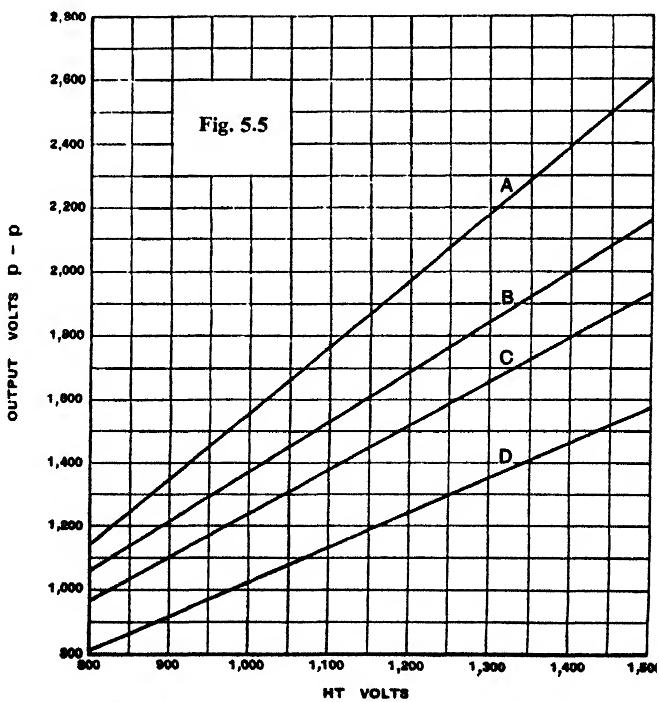
$R_1 = 50{,}000\ \Omega$
$R_2 = R_{10} = R_{17} = 5{,}000\ \Omega$
$R_4 = 3{,}000\ \Omega$
$R_5 = 0{\cdot}75\ M\Omega$
$R_6 = 0{\cdot}6\ M\Omega$
$R_7 = 400\ \Omega$
$R_8 = R_{15} = 500\ \Omega$
$R_9 = R_{16} = 1\ M\Omega$
$R_{11} = R_{14} = 90{,}000\ \Omega$
$R_{12} = R_{13} = 2\ M\Omega$

and up to 0·3 V r.m.s. ripple on the h.t. supply can be tolerated. With the usual half-wave rectification the ripple can be up to 0·2 V r.m.s.

The chief merit of the circuit is that there is a large degree of automatic compensation for variations in the characteristics of $V_3$. The input capacitance of this valve is due partly to the static input capacitance and partly to feedback through the grid-anode capacitance. This latter effect varies with the amplification. Consequently, if the gain of $V_3$ is lower than usual the effective input capacitance is smaller and a greater proportion of the output of $V_2$ is applied to its input. An abnormally high gain from $V_3$ means a larger input capacitance and hence a smaller input to the valve.

If the amplifier is set up with a valve of average characteristics for $V_3$ and $C_8$ is adjusted so that the outputs of $V_2$ and $V_3$ are equal, then the circuit is largely self-compensating for variations in the valve characteristics. The substitution of upper or lower limit valves will have little effect on the performance.

The effects of valve variations and amplifier balance are well brought out by the curves of Fig. 5.5, which show the total output voltage (peak-to-peak) of the amplifier plotted against h.t. voltage.




**Makers' rating for the output of Mazda AC/P4 valves under conditions explained in the text** (*courtesy The Edison Swan Electric Co. Ltd.*)

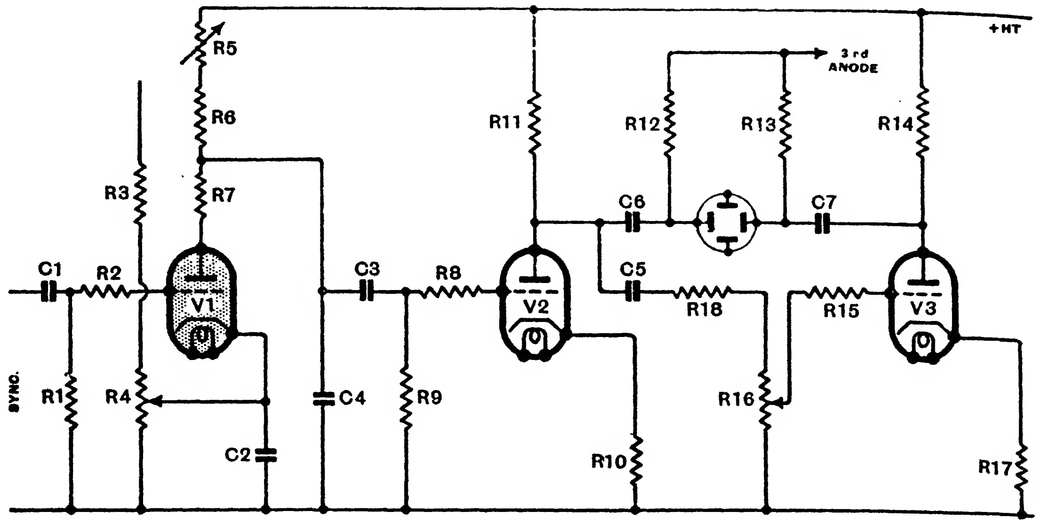



**Fig. 5.6—Recommended circuit for the frame time-base amplifier when using the Mazda AC/P4 valve**

## TABLE 5.2

### Circuit Values for Fig. 5.6

$V_1$ = Mazda T31
$V_2$ = $V_3$ = Mazda AC/P4

$C_1 = C_7 = 0{\cdot}1\ \mu F$
$C_2 = 50\ \mu F$
$C_3 = 0{\cdot}25\ \mu F$
$C_4 = 0{\cdot}5\ \mu F$
$C_5 = C_6 = 0{\cdot}05\ \mu F$

$R_1 = R_3 = 50{,}000\ \Omega$
$R_4 = 2{,}500\ \Omega$

$R_5 = 0{\cdot}75\ M\Omega$
$R_6 = 0{\cdot}4\ M\Omega$
$R_7 = 400\ \Omega$
$R_8 = R_{15} = 500\ \Omega$
$R_9 = 2\ M\Omega$
$R_{10} = R_{17} = 11{,}000\ \Omega$
$R_{11} = R_{14} = 0{\cdot}2\ M\Omega$
$R_{12} = 10\ M\Omega$
$R_{13} = 5\ M\Omega$
$R_{16} = 2\ M\Omega$
$R_{18} = 8\ M\Omega$

Curve A shows the output with two average valves and the correct balance adjustment, while curve B shows the effect of using extreme limit valves and the correct balance adjustment, the valves being individually self-biased. Curve C illustrates the reduction in output obtained by leaving the balance set for average valves, but other conditions as for curve B. It also shows the results for extreme limit valves correctly balanced but with common self-bias. Curve D shows the worst case of extreme limit valves operated with common self-bias and balanced for average valves.

With the automatic balance system recommended for the line time base, valves can be used without regard to their grading and the output will not be less than that indicated by curve B, but may be as high as that shown by curve A.

It was shown that a typical c.r. tube needs 1,235 V p–p for the line scan and 1,080 V for the frame. From Fig. 5.5, curve B, we see that a time-base h.t. supply of about 915 V is necessary. With this h.t. supply the frame time base will give the necessary output of 1,080 V p–p with conditions considerably worse than those of curve C. If individual bias is used, a fixed balance can be used and the recommended circuit is shown in Fig. 5.6. This has been slightly modified to permit the negative of the high-voltage tube supply to be earthed. When the positive is earthed $C_6$ and $R_{12}$ can be omitted and the deflector plate connected directly to the junction of $C_5$ and $R_{18}$; the value of $C_5$ should then be doubled so that it is equal to $C_7$, and $R_{16}$ and $R_{18}$ should each be halved, so that their sum is equal to $R_{13}$.

Suitable values for the time bases are given in Tables 5.1, (page 58) and 5.2 (page 60), and the h.t. supply can be chosen from Fig. 5.5. With a 1,250-V supply the current consumption of the line time base is 14·3 mA and of the frame 7·9 mA.

Power ratings of resistances and voltage ratings of capacitors are, of course, worked out in the usual way. It is worth noting, however, that the rating for the coupling capacitors to the deflecting plates of the c.r. tube will depend on whether the negative or positive terminal of the tube h.t. supply is earthed. When the positive is earthed the capacitors should be rated for working at the voltage of the time-base h.t. supply. When the negative is earthed these capacitors must be rated for working at the full voltage of the tube h.t. supply, usually some 3,000–6,000 V.

Instead of operating the time base at a high voltage and using a resistance-coupled push-pull amplifier, it is possible to work with an h.t. supply of 250–300 V only and use a single output valve by adopting transformer coupling to the deflecting plates. The amplifier circuit is shown in Fig. 5.7, and is extremely simple. The design of a suitable transformer, however, is quite difficult and the circuit is consequently not often used.

An output triode of the PX4 class taking some 50 mA anode current is usually adopted, although sometimes an output-type pentode is used. With a triode it is necessary that the transformer

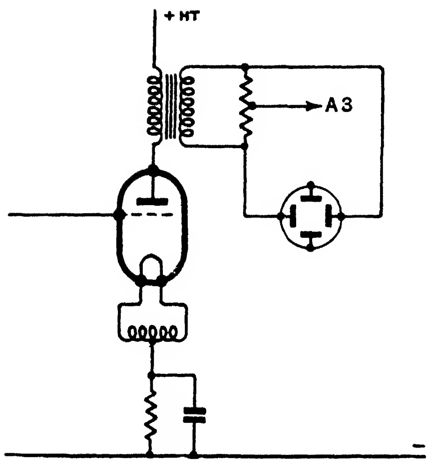



Fig. 5.7—A triode output valve can be used with transformer coupling to the c.r. tube

primary reactance at the fundamental saw-tooth frequency be large compared with the effective circuit resistance, which is to a first approximation equal to the parallel value of the valve a.c. resistance and the secondary load resistance transferred to the primary. The secondary capacitance must be kept low if any useful step-up is to be obtained in the transformer and the leakage inductance must also be low.

These requirements alone make the design quite difficult, but in addition the primary must carry quite a large direct current. Further, the core material is likely to cause amplitude distortion if all quantities are not suitably proportioned.

The difficulties are such that transformers are not often used, for it is considered easier and cheaper to use the extra valve and higher voltage, but low current, h.t. supply of the simple resistance-coupled amplifier. Those especially interested in the design of transformers are referred to the paper by D. M. Johnstone.* It should be noted that the problems of transformers for use with magnetic deflection are very different and are dealt with in another chapter.

While retaining resistance coupling in the amplifier, a two-valve arrangement of the type discussed is not always adopted. Instead of generating a low-voltage saw-tooth waveform and using a push-pull amplifier, the oscillator is sometimes designed to give an output of one-half the total deflecting voltage required and one deflecting plate is fed directly from it. A portion of the output is applied to an amplifier and the conditions are adjusted so that its output is equal in amplitude, but opposite in phase, to the oscillator output. The other deflecting plate is then fed from the amplifier.

This type of time base is dealt with in greater detail in Chapter 8 when discussing saw-tooth oscillators and there is no saving in

* "Notes on Design of Line Scanning Transformers," by D. M. Johnstone. *Journal of the Television Society*, June 1936.

valves over the more conventional arrangement, for if the output of the oscillator is to be linear and of large amplitude it is necessary to use a charging valve instead of a resistance.

When economy is required, however, it is not essential to do this and it is possible to use a time base containing two valves only. The oscillator is used to feed one deflecting plate directly and resistance charging of its capacitor is used. Owing to the large output the scan is not linear. A portion of the output is applied to an amplifier which feeds the other deflecting plate, and the characteristics of this amplifier are arranged to be non-linear in the opposite direction. The two distortions then cancel one another and a linear saw-tooth waveform is obtained between the two deflecting plates.

The general arrangement is shown in Fig. 5.8. The saw-tooth oscillator need not be the gas-triode shown but can be of any type. The capacitors $C_2$ and $C_3$ in series form the charging capacitance, which is charged exponentially through $R_4$ and $R_5$. At large output the waveform between A and earth is of the form shown in Fig. 5.9 (a). A portion of this voltage, that developed across $C_2$, is fed to the amplifier and phase-reverser $V_2$. This valve is operated in a non-linear condition so that the output between B and earth is of the form of Fig. 5.9 (b). Between A and B, the two voltages add together and give an output of the type shown at (c).

The conditions affecting the linearity of the final output are the output of $V_1$ in relation to the h.t. voltage, the input to $V_2$, the characteristics of this valve, its coupling resistance $R_8$, and its grid

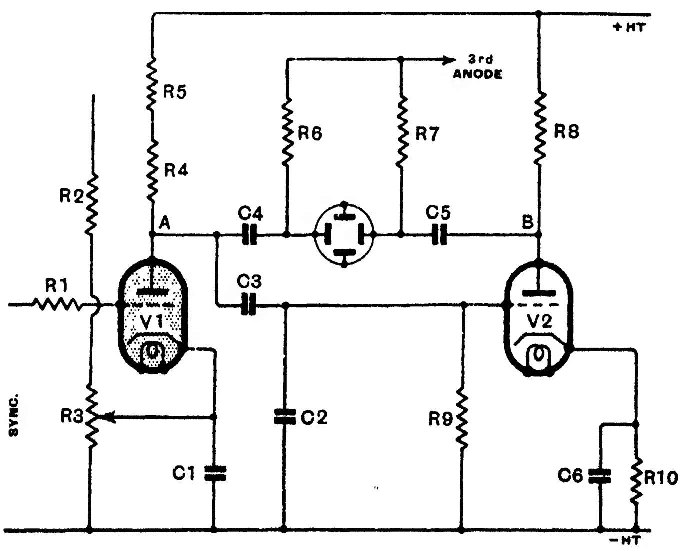



**Fig. 5.8—A time base in which $V_2$ is deliberately made to distort in order to compensate for previous waveform distortion**

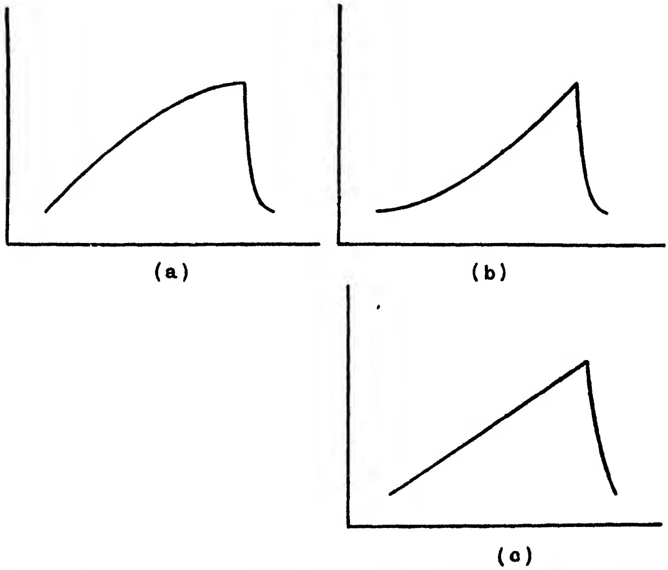


Fig. 5.9—The output of $V_1$ of Fig. 5.8 is of the form shown at (a); $V_2$ distorts it to the form (b), and the two together give (c)

bias. Careful design is necessary and the valve $V_2$ is ideally designed for the job, so that it has the correct shape of characteristic. Ordinary valves can be used, however, and the adjustment for optimum results is not difficult with the aid of an oscilloscope. The h.t. supply must be higher than with more conventional arrangements and in general should not be less than 1,500 V.

It should be pointed out that the output is not in true push-pull, for the outputs of the two valves are not at all times equal and opposite. There is consequently a change in the mean potential of the deflecting plates during the saw-tooth cycle. This is not usually very large, however, and does not seriously affect the focusing.

CHAPTER 6

# Electromagnetic Deflector Coils

WITH THE ELECTROSTATIC cathode-ray tube, deflection of the electron beam is accomplished by electric fields which are produced by applying suitable voltages to the deflector plates built into the tube. With the magnetic tube, however, deflection is accomplished by magnetic fields which are produced by passing suitable currents through coils arranged outside the neck of the tube. For such a tube, therefore, a current amplifier or generator, analogous to the voltage amplifiers of Chapter 5, is needed but, in addition, deflector coils must be employed. They are the magnetic counterpart of the deflector plates, but unlike the latter, they are external to the tube and not supplied with it.

The performance obtained with a magnetic tube depends very much on the deflector coils and the requirements of the current amplifier or generator depend on their characteristics. It is, therefore, necessary to treat them in some detail before the time-base circuits can be considered.

The design of deflector coils is nearly always carried out empirically. It is extremely difficult to calculate the performance of a given coil and the converse problem, which is the designer's problem, of calculating the kind of coil needed for a given performance, is practically impossible. In view of this, it is particularly unfortunate that there is very little published data on deflector coils.

The basic problem is to produce, as efficiently and economically as possible, two uniform crossed magnetic fields in the neck of the c.r. tube. The two fields can be produced in the same place so that the beam passes through both together, and this is the usual method, or they can be produced in sequence so that the beam passes first through one and then through the other. This second method is closely analogous to electrostatic deflection where the beam first passes between one pair of deflector plates and then between the second pair.

When a beam of electrons passes through a magnetic field its path is altered and it is deflected in a direction at right angles to the direction of the field and follows a curved path within the field. After passing through the field the beam follows a path which is at an angle to the original path. The magnitude of this deflection angle depends on the velocity of the electrons in the beam and upon the strength of the magnetic field. It is, in fact, proportional to the field strength and inversely proportional to the final anode voltage of the tube. If all the electrons have the same velocity and

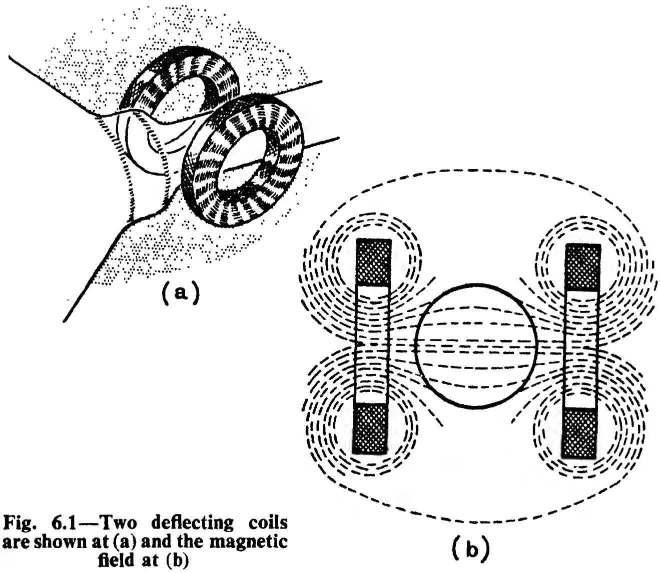


**Fig. 6.1—Two deflecting coils are shown at (a) and the magnetic field at (b)**

the field is perfectly uniform the beam will be bent into an arc of a circle but its cross-section will not be distorted. If the field is not uniform, two things happen. First, the path of the beam in the field is a curve which is not an arc of a circle and, secondly, different parts of the beam are deflected by different amounts. The cross-section of the deflected beam is then no longer circular and the spot on the screen is said to suffer from astigmatism.

It is always said that the aim in design is to produce uniform deflecting fields, but such fields are in reality physically impossible. The beam must pass from a region of no field into the field and then out of it again into a further region of no field. When entering into and emerging from the field the beam must pass through regions of non-uniform field because it is a physical impossibility for any field to stop abruptly. There must always be a transitional region in which the field strength falls off more or less gradually and in such a region the field cannot be uniform. The term " uniform field " must not be taken too literally, therefore.

Since completely uniform fields are impossible the deflection system cannot be completely free from distortion. The aim in design is therefore to make the fields sufficiently uniform for the distortion in the final raster to be so small that it is undetectable by eye.

The conventional way of representing a magnetic field is by lines which by their direction indicate the direction of the field and by their spacing show the strength of the field; the greater the spacing the weaker is the field. The representation of a uniform field is thus a set of parallel equally-spaced lines. Such a diagram, however, is only two dimensional whereas the field itself has three dimensions. It thus only shows the field on some chosen cross-section of the whole field, and to indicate the nature of that whole field a series of such diagrams is needed.

One very simple deflection system is shown in Fig. 6.1. Two large-diameter coils are placed one on each side of the tube neck (a). The magnetic field across a section through the axis of the coils has the form indicated at (b). Only the field within the tube neck, indicated by the circle, is of interest and useful, but the external field is important because it requires energy to produce it. This field serves no directly useful purpose and its energy is, therefore, wasted. An efficient deflector coil is one which produces a maximum field within the neck of the tube and a minimum outside it. The arrangement of Fig. 6.1 is much too inefficient to be of practical use.

If the field within the tube is examined it is clear that it is moderately uniform. The lines of force bow outwards, showing that the field weakens away from the centre. An electron beam entering such a field in the centre from below (i.e., coming upwards through the paper in the diagram) would emerge somewhere along a vertical line drawn parallel to the coils through the centre of the tube. The cross-section of the emergent beam would be distorted because the side of the beam farthest from the centre would be in a region of weaker field and deflected less than the side nearer the centre.

It is obvious that any required degree of uniformity of field can be secured by the simple expedient of increasing the dimensions of the coils. If the coils are made very large compared with the diameter of the neck of the tube, the field within the tube is only a small part of the total and the uniformity is high. This is not a practical solution, however, for the efficiency would obviously be very low and, moreover, there would not be sufficient space around the tube to accommodate two pairs of very large coils at right angles to each other for the two deflections.

The degree of uniformity of the field is affected by the spacing of the coils, however, in the manner indicated in Fig. 6.2. A section through the coils is shown at (a) and in (b), (c) and (d) the form of the field with various coil separations *d*. The coils are here indicated by the small circles. With wide spacing (b) the field is strongest in the centre of the system and the lines of force bow outwards. With close spacing (c) the reverse happens. The lines of force bow inwards and the field tends to be weaker at the centre than away from it. With the optimum spacing (d) the field is nearly uniform over a relatively wide area. This optimum spacing occurs when the coil separation *d* equals the radii of the coils.

However, coils of this form are too large to be practicable for television and they are never used. The simplest practical coils are of rectangular shape bent around the neck of the tube as shown in Fig. 6.3. The efficiency is obviously increased because nearly all the field produced inside the coils is in the tube and it is only the

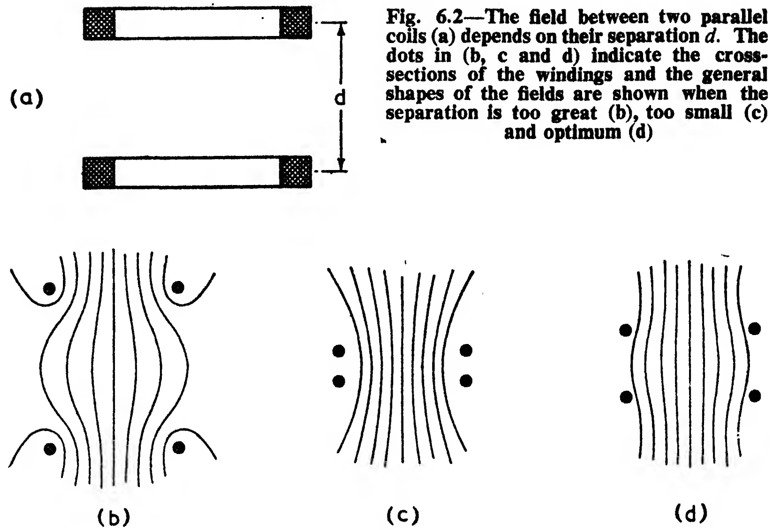



**Fig. 6.2—The field between two parallel coils (a) depends on their separation *d*. The dots in (b, c and d) indicate the cross-sections of the windings and the general shapes of the fields are shown when the separation is too great (b), too small (c) and optimum (d)**

field on their outside which is completely wasted. The coils are also smaller and it is easy to fit a second pair by wrapping them outside the first pair.

The uniformity of the field can be controlled in two ways. The first is by altering the gap between the adjacent sides of the two coils. As shown in (b) of Fig. 6.3 the field is bowed inwards like (c) of Fig. 6.2 and the side gaps are too small. As the gaps are increased the field becomes more uniform and passes through the form (d) to the bow outwards of (b) when the gaps become too great. This method of adjustment is very commonly used; in fact, it is always used when the bulk of the wire in the coils is concentrated into a small cross-sectional area.

The alternative method is a better one but is not used so much because the coils are more difficult to wind. The coils are placed with their adjacent sides touching and the distribution of wire in the cross-sections of the coils is graded. A cross-section through tube and coils is shown in Fig. 6.4; one coil is A and the other B and both are alike. The requirement for a uniform field is that the number of turns in the coil between OC and OD shall be proportional to cos $\theta$.

Coils of this kind are rarely used because of the difficulty of winding them. A close approximation to them is sometimes obtained by using three coils inside each other for each " coil " of the pair. This is indicated in Fig. 6.5, where one " coil " is made up of the sections 1, 2 and 3 while the other has sections 4, 5 and 6. Sections 1 and 6, 2 and 5, and 3 and 4 are alike, but sections 1, 2 and 3 in the one and 6, 5 and 4 in the other have increasing numbers of turns.

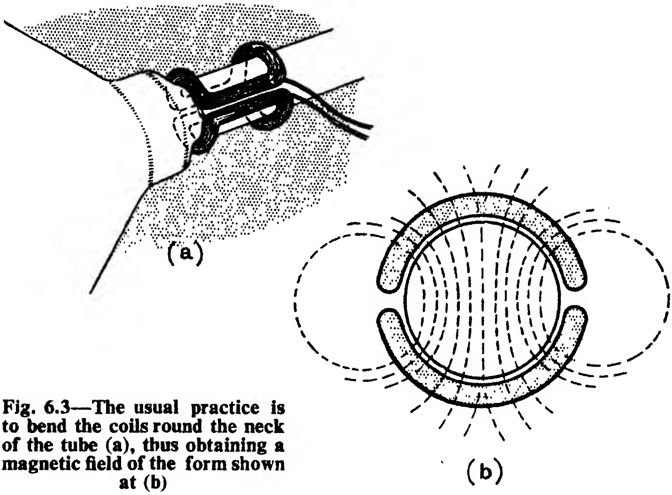



Fig. 6.3—The usual practice is to bend the coils round the neck of the tube (a), thus obtaining a magnetic field of the form shown at (b)

Only a single field has so far been considered. Two sets of coils at right angles are required to produce the crossed fields needed for a raster. Two sets of coils in tandem can be used so that the beam passes first through one set and then through the other. In this case the set of coils nearer the gun of the tube need not have nearly such a uniform field as the other set. The beam enters this set on the axis of the tube and leaves it anywhere on a horizontal plane through the axis, if the coils are the horizontally deflecting ones. It is, therefore, only necessary for the field to be uniform over a small area around this horizontal plane.

In the case of the second set of coils, remote from the gun, the beam enters anywhere on a horizontal plane through the axis, but leaves it anywhere within the tube neck—more exactly it leaves it

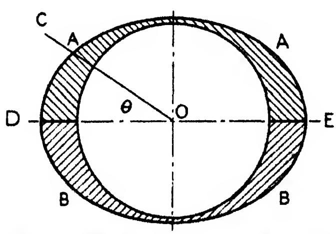



Fig. 6.4—A section through a coil assembly in which the turns are graded to give a uniform area. The turns between OC and OD are proportional to cos $\theta$

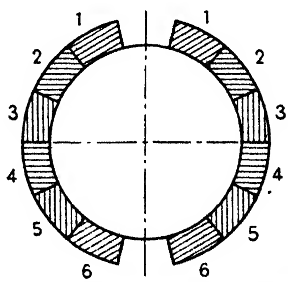



Fig. 6.5—This section through the coils shows an alternative way of obtaining a uniform field. Each " coil " is made up of three coils 1, 2, 3 in one and 4, 5, 6 in the other, having different numbers of turns

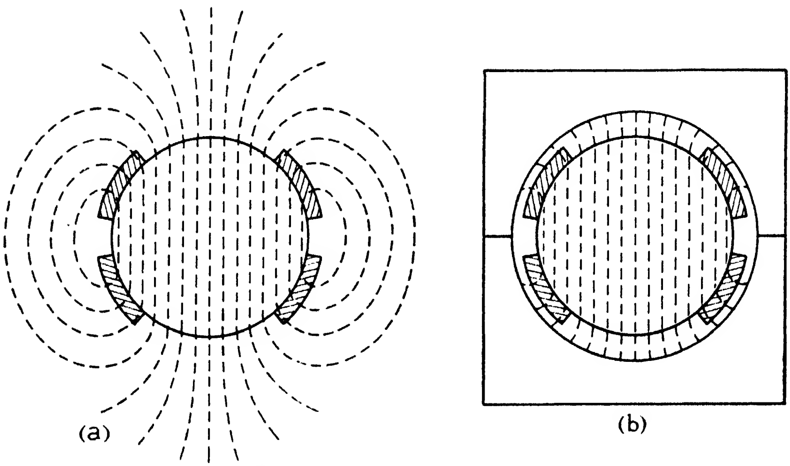



**Fig. 6.6—The general shape of the field produced by air-core coils is shown at (a) and the great reduction of external field which is obtained by the use of an external iron ring is depicted in (b)**

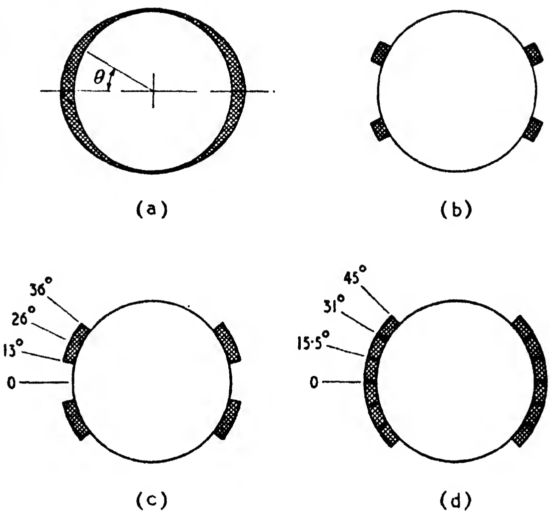
anywhere within a rectangle which is a scaled down version of the raster. This second field must be uniform over the whole of this area.

Usually the two sets of coils occupy the same place along the neck of the tube. If they are of the form of Figs. 6.4 or 6.5 the coils adjacent to the tube neck give the horizontal deflection and those for the vertical deflection are placed outside them. If they are of the simpler form shown in Fig. 6.6, the side limbs of the vertical deflection coils are placed against the neck of the tube in the space between the side limbs of the horizontal deflection coils.

In the interests of efficiency an iron circuit is normally provided outside the coils and takes the form of an iron ring. The efficiency improves as the inside diameter of this ring decreases and it is always kept to the minimum practicable value. As a result the form of winding shown in Fig. 6.5 is rarely used for, as the frame coils must lie outside the line coils and insulation between them is needed, the diameter of the iron ring must be greater than with interleaved coils. Normally, therefore, the width of any side limb of a coil is not allowed to exceed 45° so that the eight side limbs of the line and frame coils can just interleave around the tube.

Practical arrangements are shown in Fig. 6.7. At (a) the theoretically ideal cos $\theta$ distribution is repeated and at (b) is shown the other extreme with windings of square section; neither is often used. The usual arrangements are (c) and (d). In (c) the coils are uniformly wound and have about the maximum permissible dimensions for this type of winding. In (d) each coil is divided into three sections by changing the wire gauge while winding and each limb occupies 45° of the circumference. The 15·5° section has 25 per cent of the total turns, the middle 15·5° section has

**Fig. 6.7—(a) shows a graded winding of varying thickness and (b) a small uniform winding. The latter must be centred on a 26° angle (c) for a curved-face tube. A 45° winding with the turns centred on 26° is shown at (d); the turns are distributed by changing the wire size in steps**




36·5 per cent and the 14° section has 38·5 per cent; this last section is the inside one of the coil as it is wound.

These figures and the dimension angles of Fig. 6.6 are approximate only. The precise arrangement needed depends on the curvature of the tube screen. They are suitable, however, for the average curved-face tube.

With this usual form of construction, embodying a combined frame and line coil assembly, the two fields really combine to form a single composite field at any angle between the horizontal and the vertical. The angle depends on the relative intensity of the two individual fields and the strength upon the individual strengths.

The beam enters the composite field on the axis of the tube and at this point uniformity is really necessary only over a small area corresponding to the beam diameter. The beam emerges anywhere, however, and at the front end a high degree of uniformity is needed.

So far uniformity of field has been considered as an essential requirement and it is true that wherever the beam may be in the field it must be in a uniform field if its shape is not to be distorted and produce astigmatism. Such astigmatism causes deflection defocusing and the raster does not have a uniform focus over its area. Usually such defocusing is most marked in the corners of the picture.

However, a completely uniform field would often be unsatisfactory. It is right only if the centre of curvature of the tube screen lies at the centre of the deflector-coil assembly. If it is not, a non-uniform field must deliberately be produced in order to obtain a raster of rectangular shape. Usually, the tube screen is much flatter than would be obtained with such a radius of curvature and

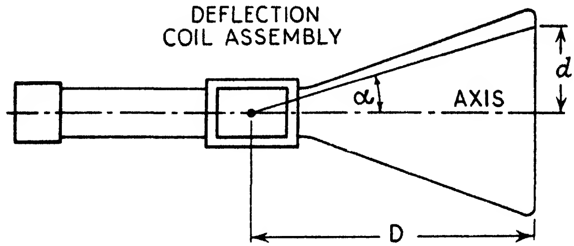



**Fig. 6.8—The geometry of a flat-faced tube is shown here. The displacement *d* of the spot is not proportional to the angle**

sometimes it is quite flat. By geometry it can be seen that a uniform deflection cannot be obtained with a uniform field.

The worst case, that of a flat-faced tube, is shown in Fig. 6.8. For a uniform field the deflection angle $\alpha$ is nearly proportional to the field strength. It is required that the displacement $d$ of the spot on the screen be proportional to the field strength. Actually $d = D \tan \alpha$ and there is approximate proportionality between $d$ and $\alpha$, and hence between $d$ and the field strength, only for small deflections.

To obtain a deflection distance on the screen proportional to the coil current the field must deliberately be graded over the cross-section of the tube neck so that it is strongest at the centre and weakest at the edges. The resulting non-uniform field must necessarily distort the shape of the beam and so cause defocusing. The importance of this depends on the beam diameter in the deflecting region. If it is small the amount of non-uniformity over its cross-section is also small, and it is only by arranging this that satisfactory results can be secured.

It is obvious, too, from Fig. 6.8, that some distortion of the spot will occur because the beam is not normal to the screen. Even at the screen the beam is of finite diameter and when it strikes the screen at any angle other than a right angle the spot must cease to be circular and become something like an angular section through a cylinder. This effect, however, is probably too small to be of much practical importance.

A rectangular raster is obtained only when the deflecting field is adjusted correctly to compensate for the "throw" distortion. When the correction is insufficient, as when coils designed for a normal tube are used with a flat-faced tube, the raster has hollow sides. If the correction is excessive, the sides are curved outwards.

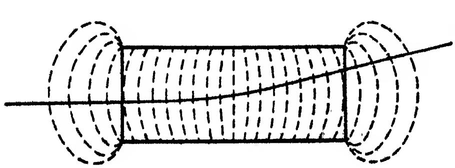


**Fig. 6.9—This section through the length of a deflection system shows the shape of the field in this plane and the path of the electron beam through it. The path shown is due not to this field but to the field at right angles to it of the other deflection system**

Strictly speaking, therefore, each type of tube needs a particular coil design. In practice, things are not very critical and usually a coil assembly designed for curved-face tubes is satisfactory with all but flat-faced tubes. The differences between different curved-face tubes are not usually sufficient to make special coils necessary for each.

So far only the field in a cross-section of the tube neck has been considered and that only in a section through the middle of the assembly. As the section is moved towards the ends of the coil assembly the field necessarily becomes less uniform. At the ends it may be very different from that at the centre and beyond the coils it necessarily becomes very non-uniform. Fortunately, its strength also falls off greatly here so that the effect of the lack of uniformity is not marked.

The field must bulge outwards from the ends of the coil as shown in Fig. 6.9. This bowing produces components of field acting along the beam which, therefore, have a defocusing action on it.

The shape of the field at the ends of the coil assembly depends very greatly upon how the end turns of the coils are disposed. Up to the present only the effect of the side limbs parallel to the neck of the tube has been considered. The wires connecting these side limbs usually pass around the tube neck at the ends as shown in

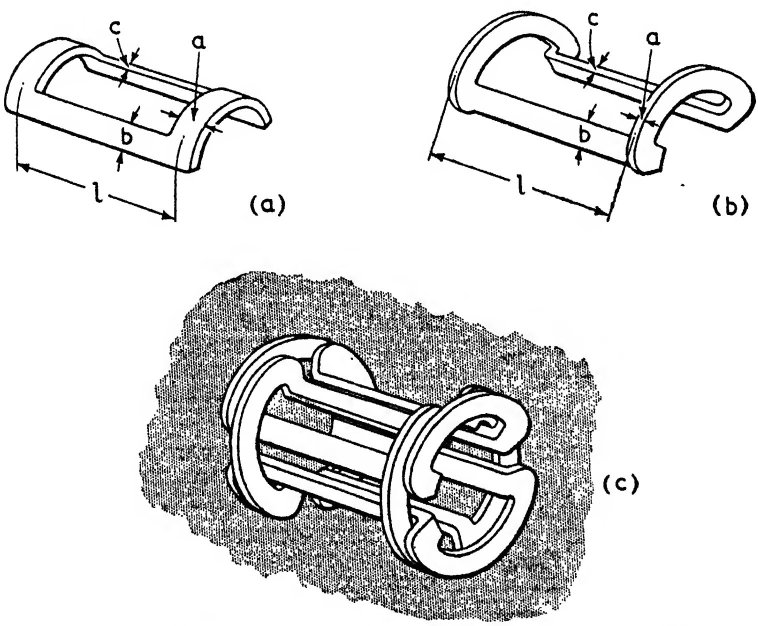



**Fig. 6.10—A thin slab coil bent to lie around the neck of the tube is shown at (a) and a similar coil with the ends bent up at (b). A greater effective length is secured with the latter. The way in which two pairs of such coils are interleaved for the two deflections is shown at (c)**

Fig. 6.3. They need not, however, lie actually touching the tube neck as shown here, but they can be bent away from it.

This is very common practice and is illustrated in Fig. 6.10. A simple slab coil bent to fit around the tube neck is shown at (a); this is one coil of essentially the same type as those of Fig. 6.3. The " bent-up end " coil is shown at (b) and the way in which two pairs are interlocked for the two deflections is indicated at (c).

With such bent-up end coils, the field inside the tube beyond the ends of the coil assembly is weaker and more uniform than with the simpler type, but they are usually rather less efficient. A greater length of wire is needed in " bent-up " ends and so the ends produce a bigger total field outside the neck of the tube. To offset this, at least partially, however, the effective length of the coil can be made slightly greater.

In Fig. 6.10 this length is *l* but since the dimension *a* of the bent-up end coil (b) is easily made a third or a quarter of that of the simple coil (a) *l* is rather greater. The deflection obtained is proportional to *l* so that this is advantageous.

The maximum value of *l*, however, is severely limited in practice. This is shown in Fig. 6.11; at (a) the effect of too long a coil assembly is indicated. The beam is shown dotted and with a deflection so that it should strike the screen at B; it is, however, intercepted by the shoulder of the tube at A. With a shorter coil assembly (b) and a greater deflection angle the beam just clears the tube shoulder.

This beam cut-off effect naturally occurs first in the corners of the raster where the deflection-angle is greatest; it is commonly termed " shadowing."

As the field extends beyond the coil assembly the effective length of the coil is greater than its physical length and so it is not possible to calculate the maximum permissible length just from the simple geometry of the tube. In practice, the overall length of the coil assembly is usually about 2 in.

Coils of this type can give a very satisfactory performance, but they are rather troublesome to wind. There are two satisfactory winding methods. One way is to wind the coil directly to its final shape on a special former which is afterwards removed. This is satisfactory

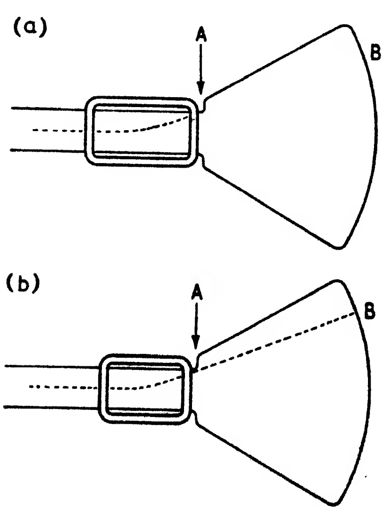



**Fig. 6.11—When the deflector coils are too long (a) the beam strikes the tube shoulder at A and is cut off, but with the correct length (b) it just misses the shoulder**

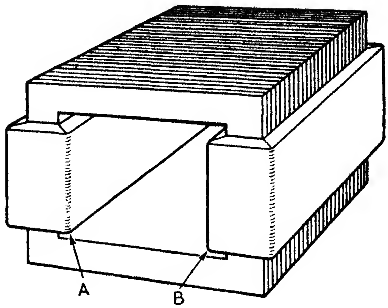



**Fig. 6.12—A form of deflector-coil assembly sometimes used consists of coils A and B wound around the limbs of a stack of rectangular laminations. A second pair of coils can be mounted on the horizontal limbs**

for low-inductance coils of up to about 300 turns, but it is too tedious for high-inductance coils which may need 2,000 turns or more. The alternative is to wind the coils as simple slab coils and afterwards to bend them in a jig to the required shape.

Deflector-coil construction is tedious rather than difficult and is well within the capabilities of the amateur. Full details of the winding and the necessary formers and jigs have been published.*

Although deflector coils of this type, with bent-up ends and an external iron ring, are usual there are several other kinds. In one variety the wire is actually wound around the iron of the iron circuit, which then forms a core to the coils. This is shown with one pair of coils only for clarity in Fig. 6.12. The two coils are connected in series or parallel, but in opposition so that their fluxes in the core are opposing. There is then a strong leakage field between the top and bottom limbs of the core and it is this which acts as the deflecting field. As shown the coils give horizontal deflection and the core has a rectangular window. The latter, however, can be square or circular.

The difference between these coils and those previously described is not as great as it may at first appear. In both the side limbs of the coils are essentially the same. Instead of taking the wires which come out of one coil at A around the tube neck to join the wires entering the other coil at B, they are taken right away from the tube and round the outside of the iron to join the wires entering the other end of A. Inside the assembly the two systems are virtually the same.

Outside the assembly there are differences. The shapes of the end fields vary and the wire outside the iron produces no useful field. To reduce this external field a tight-fitting copper screen is sometimes fitted outside the assembly and this reduces the effective inductance of the external wires by eddy currents induced in it. The effect of such a screen, however, is frequency dependent and is greater at line than at frame frequency. It also greatly reduces the $Q$ of the coil and in some circuits this is important. The second set of coils, for frame deflection, is wound on the two horizontal limbs.

* *Wireless World*, February and March 1947. Reprinted under the title " Television Receiver Construction " (Iliffe.)

The great merit of this deflection assembly is its ease of construction. All coils are of simple shape and easily wound. If they are of high inductance they should be sectionalized to reduce the self-capacitance and this also makes it easy to grade the windings to secure the required shape of magnetic field. The core can be built from L-shaped stampings but it is necessary to take great care to secure well-fitting butt joints. Any appreciable air gap where the two L assemblies butt together distorts the raster asymmetrically. It is not usually necessary for the core to be a full stack of laminations and it is sufficient in most cases if it is built up of a number of $\frac{1}{16}$-in stacks spaced $\frac{1}{8}$-in apart.

The assembly is not a particularly efficient one for as the interior of the iron is square and the tube neck is circular a good deal of unnecessary field is produced. It is not much used for television in this form, but a modified version is not uncommon. This has coils on the top and bottom limbs for frame deflection but a pair of " bent-up end " or similar coils is placed inside the iron yoke for the line deflection.

The efficiency can be increased appreciably by using circular ring laminations instead of the square formation. The coils must then be wound on to the iron and a machine of the kind used for winding toroidal coils is almost a necessity.

Probably the most efficient coils of all have a similar ring core but the rings have internal teeth. The core is a stack of laminations of the shape shown in Fig. 6.13. The turns lie in the slots and the wire is either taken around the outside of the iron or around the neck of the tube to enter the equivalent slot on the other side. Thus either of the two basic forms of winding can be used with it. The toroid type demands winding in situ; the bent-up end type enables the coils to be wound and taped and then inserted in the slots, much as in armature winding.

The high efficiency of this type of coil assembly is due to the minimizing of waste field. The effective internal diameter of a deflector-coil assembly is roughly that of the iron ring and in this construction the internal teeth enable it to be brought very close to that of the tube neck.

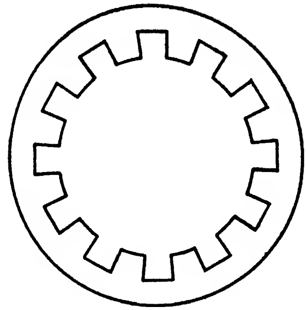
In spite of its efficiency it is not yet common in television sets, probably because the laminations are more expensive. They are certainly wasteful of material. In stamping them from sheet metal there is more material wasted than used unless a large number of different sizes are wanted.

It is probably cheaper to use a less efficient coil assembly and supply it with extra power, but where efficiency is really important its merits are worth close investigation. A considerable amount of information on coils of this type for radar purposes has been published.*

In order to design circuits to supply the deflector coils with

* "Iron-Cored Deflecting Coils for Cathode-ray Tubes," by A. Woroncow. *J. Instn elec. Engrs*, 1946, Vol. 93, Pt. IIIA, No. 10, p. 1564.

**Fig. 6.13—This form of lamination enables an exceptionally efficient coil assembly to be made**



saw-tooth currents it is necessary to know their characteristics. It is necessary to know the inductance and resistance of the coil and the current which it requires to produce a given deflection; for some circuits it is also necessary to know the $Q$ of the coil at a frequency of about 30 kc/s.

This information about a deflector coil is also sufficient to enable the efficiency of one type to be compared with that of another but it is not in a convenient form for doing so unless the two coils happen to have the same inductance. The author has found it convenient to express the characteristics of a coil by a figure of merit which is the value of $LI^2$ where $L$ is the inductance (mH) measured at 10 kc/s and $I$ is the peak-to-peak current (amperes) needed to produce a standard deflection on a standard c.r. tube operating at a standard final anode voltage. The figure is a measure of the energy needed to provide deflection and in consequence the smaller the figure the better the coil.

It would be better to measure the deflection angle rather than the deflection distance since the two are not necessarily exactly proportional to one another, but the angle cannot be measured directly. The author has adopted for his standard a deflection of 7·5 in on the screen of a Mullard MW22–14c tube operating at 5 kV; the angle is about 48°.

Highly efficient deflector coils usually have a value of $LI^2$ of about 1 mH–A², but the majority in common use have a figure nearer 1·3 mH–A². The theoretical limit is about 0·5 mH–A². Much higher figures are found with some designs.

A detailed investigation into the relative merit of various types of winding and iron circuit has been carried out* and it is of interest to quote a few of the figures. A bent-up end type coil of good design with no iron circuit was found to have $LI^2=2{\cdot}1$ whereas the same coil with a 1-in stock of 0·014-in laminations having a 42-mm circular window had $LI^2=1{\cdot}06$. The iron circuit thus doubled the efficiency of the coil. The quantity and grade of iron does not have a marked effect and figures around 1·3 mH–A² were obtained with a very meagre iron circuit. The use of an iron circuit with a 42-mm square window gave $LI^2=1{\cdot}1$.

* "Deflector-coil Characteristics," *Wireless World*, March, April and May, 1950.

With core-type coils (toroidal winding) using a square window much poorer figures were obtained. One coil gave $LI^2=3{\cdot}1$ mH–A$^2$ which could be reduced to 2·2 mH–A$^2$ with an external copper shield. The use of such screening is hardly to be recommended, however, for it is ineffective at frame frequency and it greatly reduces the $Q$ of the coil.

At frame frequency the $LI^2$ figure is not of first importance. If the coil can be matched to the valve, it is the value of $RI^2=LI^2\times R/L$ which is important. Since matching is not always possible it is advisable to use the two figures $LI^2$ and $R/L$. The $LI^2$ figure for a frame coil is usually about 30 per cent higher than for a line for the same deflection. As less deflection is needed for the frame scan the working $LI^2$ figures are about the same. The value of $R/L$ is usually 1·5–2 $\Omega$/mH.

For economy circuits it is necessary to know the $Q$ at 30 kc/s. This depends very much on the iron circuit and may be as low as 3. Without iron it may be around 15 in a typical case, but higher values are possible. A knowledge of the self-capacitance may also be required. This can vary greatly but is unlikely to be much under 100 pF and may be much greater. It is not usually important to know it accurately when the deflector coil inductance does not exceed 10 mH, which is the usual case.

On the line scan high peak voltages are developed across the coils and so great care must be taken to secure adequate insulation. Largely in order to minimize the insulation needed, the coils rarely have an inductance greater than 10 mH and are fed through a transformer. Even then the peak voltage can be as high as 1 kV. The self-capacitance of deflector coils is usually fairly high—50–200 pF—and the effect of this is greatly reduced by using low-inductance coils. The self-capacitance of coils of the type of Fig. 6.12 is especially high if the windings are not sectionalized.

In the case of the frame circuit the voltage is low and self-capacitance is not so important. High-inductance coils of 1–3 H are, therefore, practicable. Their use has the advantage of saving

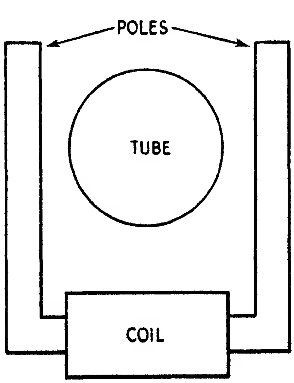



**Fig. 6.14—When the line and frame deflectors are in tandem, the frame assembly often takes this form. The poles are a stack of laminations of about ¼-in thickness. Sometimes they are fitted with shaped pole pieces**

a transformer, but interaction between the two sets of coils is more likely than if both are of low inductance.

If the coils are carefully constructed with their axes at right angles magnetic coupling between the coils is usually negligibly small. Capacitive coupling occurs, however, and the frame coils are liable to shock excitation by the line fly-back. They then carry an oscillation at their natural frequency which can distort the raster on the left-hand side. The effect is to make the lines bent on the left. Sometimes in bad cases there are several bends alternately up and down.

The effect can always be cured by screening the coils electrostatically. This is most easily done by taping the frame coils with strip tin foil; of course, insulating the point where the finish overlaps the start so that the screening does not form a short-circuited turn. The screening is earthed.

However, it is very rare indeed for such screening to be necessary and although some coupling exists it is usually of negligible magnitude. The use of low-inductance line coils with high-inductance frame coils is quite satisfactory in practice.

Tandem mounting of the coils is rarely adopted. It is normally used only when exceptional freedom from interaction between them is required, as when they are fed directly from oscillators rather than amplifiers. The line coils are always mounted on the tube neck and are of one of the types already described. They can be of somewhat simplified form because the uniformity of field needed is rather less than when both sets of coils are mounted together.

The frame deflector is placed an inch or so nearer the screen, usually slightly in front of the point A in Fig. 6.11. The coil itself is well away from the tube and the magnetic field is conveyed to the tube by iron poles mounted alongside it. The arrangement is shown in Fig. 6.14. It is simple to make but inefficient, because there is a large amount of waste field. The stack of laminations is usually only ½-in thick and so the effective length of the field is quite short. In addition, the centre of deflection is relatively near the screen.

The low efficiency is not as important in the frame circuit as it would be in the line, for in any case the frame circuit needs much less power. It is, however, a great inconvenience and the system is not used very much.

CHAPTER 7

# Electromagnetic Deflection Circuits

IN ORDER TO secure linear deflection of the beam in a c.r. tube, it is normally necessary for a deflector coil to be provided with a current which varies linearly with time during the scan. During the fly-back, however, it is permissible for the current to change back to its initial value in any convenient way. In practice, of course, perfect linearity of the scan current is rarely achieved and it is usually regarded as satisfactory if the departures from linearity do not exceed 5 per cent.

When there is no current in the deflector coils the scanning spot is, or should be, in the centre of the fluorescent screen. To obtain a centred raster, therefore, the deflection must be equal on either side of, and above and below, this centre. In its turn, this means that the current excursions in the deflector coils must lie equally on either side of zero; consequently, the deflecting currents must have no d.c. components.

These saw-tooth currents, which are needed in the deflector coils for scanning, are produced by methods which it is the aim of this chapter to describe. Valves are used in more or less complex circuits to control the currents, which are necessarily derived from the h.t. supply.

At least one of the valves is commonly called an amplifier and is often considered to be analogous to the output stage of an a.f. amplifier. The analogy is not a very close one, however, and in some circuits there is but little relation between the two. It is best, therefore, not to rely upon any resemblance which there may be between them but to consider the scanning circuit as something quite separate and distinct. There is then much less chance of being misled by certain similarities between circuits which do not extend to their modes of operation.

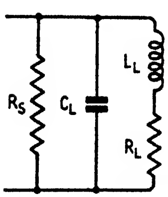



**Fig. 7.1 — Equivalent circuit of a deflector coil**

The proper starting point is the deflector coil. It has inductance $L_L$ and series resistance $R_L$ and there is also some capacitance $C_L$ across its terminals. In addition there are iron and dielectric losses which are best represented by a shunt resistance $R_S$. The equivalent circuit of a deflector coil thus has the form shown in Fig. 7.1. In general, $L_L$ and $R_L$ are important on the scan and $C_L$ and $R_S$ are negligible whereas on fly-back $R_L$ can often be ignored.

A certain current of peak-to-peak value $I_L$ is needed in the coil to produce the required deflection of the electron beam. For a given design of coil this current depends on the inductance and, as explained in Chapter 6, the characteristics of the deflector coil are best generalized by the values of $L_L I_L^2$ and $R_L/L_L$. The value of inductance which can be used is limited by mechanical considerations since there are limits to the size of wire which can conveniently be used. There are also certain limits set by the performance requirements of the associated circuits.

The inductance can be varied at will within these limits without affecting $L_L I_L^2$ or $R_L/L_L$, which are fixed only by the general design of the deflector coil, the deflection required and the operating voltage of the tube. Typical values are $L_L I_L^2 = 1{\cdot}3$ mH–A² for the line scan and $L_L I_L^2 = 0{\cdot}9$ mH–A² for the frame with $R_L/L_L = 1{\cdot}8$ Ω/mH for both.

Since the current varies linearly with time and has no d.c. component, its instantaneous value during the scan is very simply expressed as

$$i_L = I_L\left(\frac{t}{\tau_1} - \frac{1}{2}\right)$$

where $\tau_1$ is the scan period, 82·77 sec for the line and 18·62 msec for the frame. At the start of the scan the current is $-I_L 2/$, half-way through it is zero, and at the end ($t = \tau_1$) it is $I_L/2$.

This current flows through the resistance of the deflector coil and produces a voltage drop $i_L R$ across it. In the inductance, however, it produces a back e.m.f. $L_L I_L/\tau_1$; this is a constant voltage and does not vary during the scan. The total voltage across the terminals of the deflector coil is the sum of the two; it is $i_L R_L + L_L I_L/\tau_1$, and this represents the composition of a steady voltage $L_L I_L/\tau_1$, and a saw-tooth voltage of peak-to-peak amplitude $I_L R_L$. The ratio of the two is $\tau_1 R_L/L_L$ (if $R_L/L_L$ is in Ω/mH, $\tau_1$ must be in msec). For a typical deflector coil on the line scan it has the value $0{\cdot}0828 \times 1{\cdot}8 = 0{\cdot}149$, but for the frame scan it is $18{\cdot}62 \times 1{\cdot}8 = 33{\cdot}52$.

This is one factor which causes a marked difference between line and frame operating conditions. On the line scan the voltage drop across the series resistance is small compared with the inductive back e.m.f., whereas just the reverse is true on the frame scan.

On fly-back the change of voltage across the resistance is the same as on the scan but the inductive back e.m.f. is much greater because the rate of change of current is greater. The e.m.f., too, is of the opposite polarity, for the current is changing in the opposite direction. If the current change were linear, the voltage appearing across the inductance would be constant and greater than the scan voltage by the ratio of the scan to fly-back times; that is, 5·15 times as great for the line and 13·5 times for the frame.

These idealized waveforms are shown in Fig. 7.2 to scale (but note that the scale for the frame voltage is ten times that for the line) assuming a deflector-coil inductance of 6·5 mH. It is plain that on the line the inductive voltage greatly predominates over the resistive, whereas on the frame the reverse is true except during fly-back.

The scan conditions must be closely met in reality so that on the scan these ideal waveforms are very closely true in practice. On fly-back, however, the current is very far from linear and in consequence the inductive back e.m.f. varies considerably and reaches a peak value considerably greater than that shown. On the line coil, for instance, the back e.m.f. instead of being about 5 times that on the scan is usually around 8 times. The voltage with a linear fly-back is at its minimum possible value.

The current waveforms in the line and frame coils may be identical save for the different ratio of scan/fly-back times, the coils themselves may have very similar characteristics, and the magnitudes of the currents may be much the same; in spite of this, the voltage waveforms across the coils are quite different and the magnitude of the voltage is much greater for the line coils than for the frame.

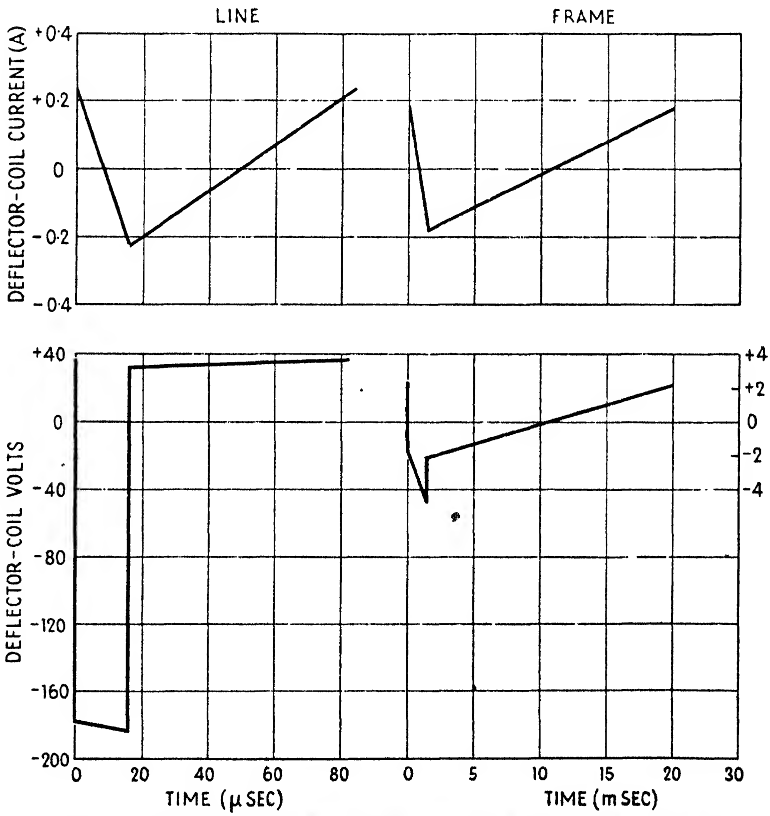
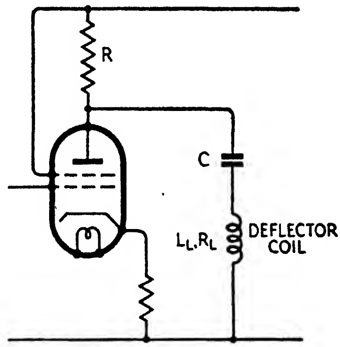



Fig. 7.2—Idealized current and voltage waveforms for a linear scan and fly-back. The inductance is assumed to be 6·5 mH and $R_L/L_L = 1{\cdot}8\Omega$/mH; $L_L I_L^2$ is 1·3 mH–A² (line) and 0·9 mH–A² (frame). Note that the voltage scale for the frame is ten times that for the line

**Fig. 7.3—Typical frame-scan output stage with a resistance-capacitance fed deflector coil**




The difference is solely due to the much greater rate of change of current in the line coil.

It is a very important difference. In conjunction with the fact, which arises out of it, that the circuit capacitance is important at line, but not at frame, frequency, it results in the circuits and valve operating conditions being quite different for the two. In the foregoing line and frame have been treated together in order to bring out this difference; from now on they have so little in common that they must be considered separately.

## Frame Scan

Because of the low voltage appearing across a frame deflector coil and because the circuit capacitance is usually of little account, it is feasible to employ a high-inductance coil of the order of 1–3 H. The limit to inductance is usually set by the extent to which it is safe to reduce the wire size. The winding space is limited and it is not practicable to increase the turns beyond a certain point without the wire becoming too thin for mechanical strength.

A high-inductance coil is normally resistance-capacitance or choke-capacitance fed from the valve. The former is shown in Fig. 7.3; with the latter the resistance $R$ is replaced by a choke.

During the scan the effect of the inductance of the deflector coil is negligible. The current through the coil and the voltage across it are thus substantially of linear form. If $L_L I_L^2$ remains at the figure of 0·9 mH–A² and $R_L/L_L$ at 1·8 Ω/mH, as they probably will do, then if $L_L$ is made 1 H, $R_L$ becomes 1·8 kΩ and the current is $I_L = 30$ mA p–p. The voltage across the coil is, therefore, 54 V p–p.

The current flows in the coupling capacitor and produces a voltage across it. This voltage is proportional to the integral of the current and can be split into three components. The first is a steady voltage equal to the mean anode voltage of the valve. The second is a linear voltage acting in opposition to the linear voltage across the coil. The third is a square-law voltage acting in the same direction as the coil voltage.

The voltage across the deflector coil and $C$ together, which is the anode-earth voltage of the valve, is thus the sum of constant, linear and square-law voltages. (See Equ. A.5.31, p. 365). The current in the coupling resistance $R$ is that due to the difference of this anode voltage from the h.t. supply voltage acting on $R$, and this current must necessarily contain a square-law component also. The current to be supplied by the valve is clearly the sum of the currents in the deflector coil and in $R$ and so also has a square-law term (Equ. A.5.32).

The magnitude of the square-law component of anode current depends on the value of $C$ in relation to the circuit resistances. If it is 20 per cent of the amplitude of the linear current, it is shown in Appendix 5, that $C = 56/(R + R_L)$. It is rarely possible to make the total resistance greater than 10 kΩ nor can it usually be much less than 3 kΩ so that the value of $C$ will be some 6 μF to 18 μF.

If the square-law current is to be only 1 per cent of the linear, the relation becomes $940/(R + R_L)$, so that the capacitance values must be 16·8 times as large. They then become 100 μF to 300 μF.

What this means is that if the valve operates linearly and is fed with a linear saw-tooth at its grid the coupling capacitor must have a value of several hundred microfarads if the current in the deflector coil is to be substantially linear. The component must be rated for several hundred volts, and it is consequently neither small nor cheap.

It is usual, therefore, to employ a capacitance of only about one-tenth this ideal and to arrange for the valve to supply the necessary square-law component of current. If this is not done the deflector-coil current ceases to be linear and takes an exponential shape. The result is a noticeable cramping of the bottom part of the picture.

The valve may in whole or in part itself generate the square-law component of current. If it is fed with a positive-going saw-tooth at the grid, the natural curvature of its characteristic is in the right direction for this. However, it is usually advisable to linearize the valve by current feedback from its cathode-bias resistor and to produce the square-law component in a special correction network preceding the valve or by a feedback circuit.

As the anode current of the valve increases during the scan the anode voltage falls and it reaches its minimum value at the end of the scan when the current is a maximum. It is necessary that the valve should be able to draw this peak current at the minimum anode voltage without the grid potential becoming too near zero, for if grid current flows the end of the scan will be sharply and seriously distorted. In a tetrode or pentode much the same thing will also happen if the minimum anode voltage falls too low. The visible effect is that the bottom of the picture is cut off and a bright line appears across the bottom.

The current to be supplied by the valve is the sum of the deflector-coil current and the current in $R$. Obviously $R$ should be as large as possible compared with $R_L$. The mean anode current must pass through $R$, however, and as a result the large voltage drop across a high value resistance demands the use of a very large h.t. voltage. Compromise is needed, therefore.

There is a certain minimum value for $R$ which is set by fly-back requirements. On fly-back the grid potential of the valve is usually changed as quickly as possible from its least negative to its most negative values. The various currents in the circuit do not follow at once because the time-constant of the inductance-resistance combination is too great.

The current in the deflector coil changes exponentially. At the start of fly-back the voltage on the anode of the valve jumps from its minimum value to its maximum and then falls off exponentially. It is necessary that the change be completed within the 1·38 msec allotted in the transmission and, as the current changes more quickly with a high value of resistance than with a low, this sets a limit to the minimum permissible resistance. It is shown in Appendix 5 that this limit is given by

$$R = 3{\cdot}26\, L_L - R_L$$

the units being k$\Omega$ and H. Taking the earlier values, this makes $R = 3{\cdot}26 \times 1 - 1{\cdot}8 = 1{\cdot}46$ k$\Omega$ as a minimum.

If $R$ is made too small the extreme top of the picture will be affected. Over a small range the scan will be very cramped and the top will appear to be folded over. The coupling resistor can be higher than the minimum value, but there is another limit, a maximum one, which must not be exceeded. This depends partly on the valve but chiefly on the h.t. supply available. If the maximum permissible value is less than the minimum permissible, of course, the circuit cannot be used at all. This is only another way of saying that the circuit requires a certain minimum h.t. voltage.

It is shown in Appendix 5 that the current swing to be provided by the valve is

$$i_p = I_L\,(1 + R_L/R)$$

Taking the earlier values, $i_p = 30\,(1 + 1{\cdot}8/1{\cdot}46) = 67$ mA. The current efficiency is poor, for the coupling resistance takes 55 per cent of the current output of the valve. It is better if the resistance is higher, of course, but then more voltage is lost.

It will usually be inadvisable to swing the anode current below about 10 mA ($= i_1$) and so the mean anode current must be around $10 + 67/2 = 43{\cdot}5$ mA. A pentode of the EL33 or 6V6 class is thus called for.

The maximum voltage across the circuit is

$$\begin{aligned} E_p &= i_1 R + I_L\,(R_L + R/2) \\ &= 10 \times 1{\cdot}46 + 30\,(1{\cdot}8 + 1{\cdot}35) = 109 \text{ V}. \end{aligned}$$

The h.t. supply voltage needed is this voltage plus the minimum permissible anode-cathode voltage of the valve plus the cathode-bias voltage. The minimum anode voltage for a pentode is usually 70–100 V and the bias voltage 5–15 V. The h.t. supply must, therefore, be 184–224 V according to the type of valve used.

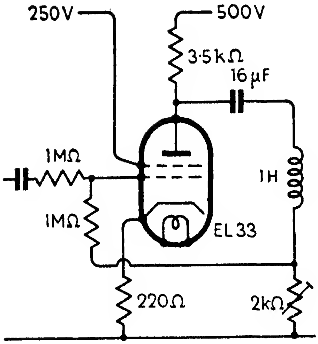



**Fig. 7.4—Complete frame output stage with resistance-capacitance coupling and linearity correction by negative feedback**

Nothing has been allowed in the above for component tolerances. When these are taken into account *R* will have to be rather higher, to ensure a sufficiently rapid fly-back with a low-tolerance resistor or a high-tolerance deflector coil, but the current will not necessarily fall much, if at all, since it must be chosen for opposite tolerances.

The result will be a need for a higher h.t. voltage and with a good design, allowing plenty of tolerance, an h.t. supply not far off 300 V is likely to be needed.

A typical practical circuit is shown in Fig. 7.4. Linearization is obtained by negative feedback from the 2-kΩ variable resistor which also acts as a height control. It is in series with the deflector coil and so carries the deflecting current. The h.t. supply of 500 V is greater than is needed for this stage but in the equipment in which the circuit was used a 500-V line had to be provided for other stages.

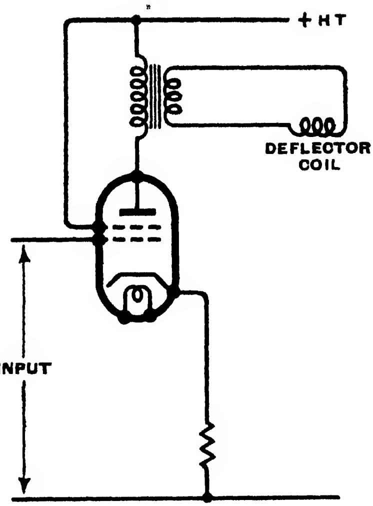
When the voltage and current requirements are excessive they can be reduced by replacing the coupling resistance by a choke. The inductance of the choke must be as large as possible and 50 H is about the minimum. It is impracticable to make it large enough to avoid its introducing distortion of the same nature as that caused by the coupling capacitance; as a result, the use of a choke considerably increases the amount of linearity correction needed.

The fly-back is usually speeded up considerably and can become too rapid, producing an undesirably large peak voltage. It is, therefore, often advisable to shunt the choke with a resistance of the order of 10–20 kΩ.

Choke-capacitance coupling is not much used because the choke costs nearly as much as a transformer and there are two components—the choke and the capacitor—which cause scan distortion instead of only one.

It should be noted that in the foregoing the use of a triode instead of a pentode modifies the circuit relations because its a.c. resistance becomes important. In some cases it can be taken into account by taking for the value of *R* the actual value of this resistor in parallel with the a.c. resistance of the valve. The triode is not often used, however, because the current needed is usually in excess of the

Fig. 7.5—Typical transformer-coupled frame output stage




capabilities of most types and the triode usually has a higher minimum permissible anode voltage.

The great merit of resistance-capacitance coupling is its simplicity for it needs no special components. It is entirely satisfactory as long as the h.t. supply is adequate. The general practice, however, is to use transformer coupling with low-inductance deflector coils of the order of 5–10 mH. Its greatest merit is perhaps its flexibility in the way of h.t. requirements.

The basic circuit of a transformer-coupled stage is shown in Fig. 7.5, and a step-down ratio of primary/secondary turns of around 15:1 is often used.

The effect of the transformer is most readily seen with the aid of one of its well-known equivalent circuits. There are several of them and the most convenient is shown in Fig. 7.6. In this $r_p$ and $r_s$ represent the resistances of the primary and secondary windings and the symbol $L_s$ stands for the secondary inductance of the real transformer. The coupling coefficient is $k$, and the shunt inductance $k^2L_s$ represents the secondary inductance in the equivalent circuit while $L_s\,(1 - k^2)$ represents the leakage inductance. The transformer shown in the dotted box is an " ideal " transformer which does nothing but provide a ratio of $n/k$ : 1.

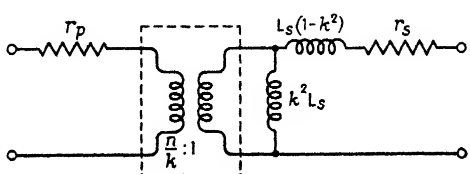



Fig. 7.6—Equivalent circuit of transformer

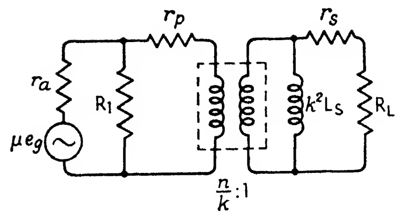



**Fig. 7.7—Equivalent circuit diagram of transformer-coupled stage effective for scan conditions**

Save that no account has been taken of winding capacitances or of core losses, this equivalent circuit represents the real transformer very closely.

The deflector coil is connected to the secondary so that its resistance and inductance are additive to the transformer secondary resistance and its leakage inductance. The ratio of $r_s$ to the leakage inductance is unlikely to be greatly different from the ratio $R_L/L_L$ and so the ratio of total resistance to inductance is of the same order. It was shown earlier that on the scan the deflector-coil inductance could safely be ignored and since the resistance/inductance ratio is still about the same it is permissible to neglect the deflector-coil and leakage inductances. The equivalent circuit of the whole stage can thus be brought to the simple form of Fig. 7.7 with reasonable accuracy.

The current in the deflector coil is, as before, assumed to be an ideal linearly-changing current of peak-to-peak value $I_L$, and its presence therefore necessitates a similarly linearly-changing voltage on the secondary of peak-to-peak value $I_L (R_L + r_S)$. This, of course, appears across the primary of the ideal transformer as a voltage $n/k$ times as great. The secondary inductance shunts the resistance and it must consequently draw a current which is proportional to the integral of the saw-tooth voltage. Its current, therefore, must contain linear and square-law terms and so the total current supplied by the secondary of the ideal transformer, and the current fed to the primary of either the ideal or the real transformer, must contain linear and square-law components if the deflector-coil current is to be linear.

This is exactly analogous to the effect of the coupling capacitance in the resistance-capacitance form of stage, but there the capacitance made a square-law voltage necessary. This had to be produced across a resistance and so necessitated a square-law current. Here the square-law current is a direct requirement and the voltage is linear on the secondary. However, in the primary the square-law current flows through $r_p$ and produces a square-law voltage across it. In both cases, therefore, the waveforms are substantially the same.

The calculation of the ratio of the square-law and linear components is quite easy and is treated in Appendix 5. It is not so easy, however, to determine what value of secondary inductance is needed for a given current ratio for to do it $r_s$ must be known and

$r_s$ depends on $L_S$. The difficulty can be got over by working in terms of $r_s/L_S$. With a component of given form $r_s/L_S$ is approximately constant, for if the number of turns is increased to obtain a larger inductance the wire size must be reduced to permit their accommodation in the fixed winding space and so the resistance goes up in proportion.

The value of $r_s/L_S$ depends on the design of the transformer and it can vary over a considerable range. A value as low as 1 Ω/H is possible but 5–10 Ω/H is more common.

The formula for $L_S$ is given in Equ. A.5.24, p. 363, and taking as an example $L_L = 6{\cdot}5$ mH, $R_L/L_L = 1800$ Ω/H, $r_s/L_S = 10$ Ω/H then for a ratio of square-law to linear currents of 0·2 the secondary inductance must be 1·46 H. If a 20 : 1 transformer ratio is wanted the primary inductance would have to be 400 times this, or nearly 600 H, which is impracticable.

The secondary resistance of the transformer has quite a marked effect on the inductance needed. If it could be made zero, then in the example $L_S$ would fall to 0·65 H. On the other hand if $r_s/L_S$ increases the value of $L_S$ goes up considerably. In fact, beyond a certain value it is impossible with any inductance to meet the requirements. It can be seen from Equ. A.5.24 that it is essential to have

$$\frac{r_s}{L_S} < 1 \Big/ \frac{\tau_1}{2}\left(1 + \frac{1}{a}\right)$$

where $a$ is the ratio square-law/linear currents and $\tau_1$ is the scan period. For $a = 0{\cdot}2$, we get $r_s/L_S < 18$, approximately. In practice, $r_s/L_S$ must be much less than 18 if the transformer inductance is not to be enormous.

In practice the primary inductance is usually 50–100 H. This is possible because a much higher ratio of square-law to linear currents is used and, in addition, the transformer ratio is sometimes reduced below the optimum for efficiency.

It is important to realize that it is quite impracticable to make the transformer inductance large enough to prevent the transformer from affecting the waveform considerably. It is essential to arrange for the valve to supply quite a large square-law current. The exact waveform needed is easily calculable from Equ. A.5.4 and, as an example, it is shown in Fig. 7.8 as a ratio $i_s/I_L$ (secondary current/deflector-coil current). The primary current waveform is the same. The assumed values are $L_L = 6{\cdot}5$ mH, $R_L = 11{\cdot}7$ Ω, $L_S = 0{\cdot}3$ H, $r_s = 3$ Ω, $k = 0{\cdot}995$, $\tau_1 = 18{\cdot}62$ msec.

The total current waveform is shown by curve C which is the sum of the linearly-changing current, curve A, and the square-law current curve B. Curve D represents the deflector-coil current. All these are shown with zero d.c. components. On the transformer primary, of course, a d.c. component is added and is the mean anode current of the valve.

It is interesting, and important to notice that the use of a large square-law component of current does not seriously increase the peak-to-peak current. The reason is that the shunt inductance of the transformer requires a negative linear current as well as the square-law current and this partially offsets the deflector-coil current.

Fig. 7.8 shows the linear and square-law currents supplied by the ideal transformer of Fig. 7.7 and their sum, the total current. In Fig. 7.9 the same currents are shown with a different grouping. Curve A shows the current drawn by the shunt inductance of the transformer, the sum of square-law and negative linear terms, and curve B the deflector-coil current, while curve C is the total. Curves B and C of Fig. 7.9 are the same as curves D and C of Fig. 7.8.

This fact, that the need for a large square-law current does not greatly increase the total current, is the one which makes transformer coupling possible at frame frequency. It is not practicable to make the inductance large enough for the square-law current to be negligible. If the square-law current needed by a normal transformer were simply additive to the deflector-coil current the total current would be so greatly increased that a large and uneconomical valve stage would be necessary to supply it. It is the fact that the transformer needs quite a considerable negative linear current which saves matters. The total peak-to-peak current needed is actually only $(1 + L_L/k^2L_S)$ times the deflector-coil current.

Reverting to Fig. 7.7, the voltage and current at the primary of the ideal transformer are $n/k$ and $k/n$ times the secondary voltage and current. The current flows through the resistance $r_p$ of the primary winding and so produces a voltage drop across it and the primary voltage is increased by this amount. It is usually fairly small.

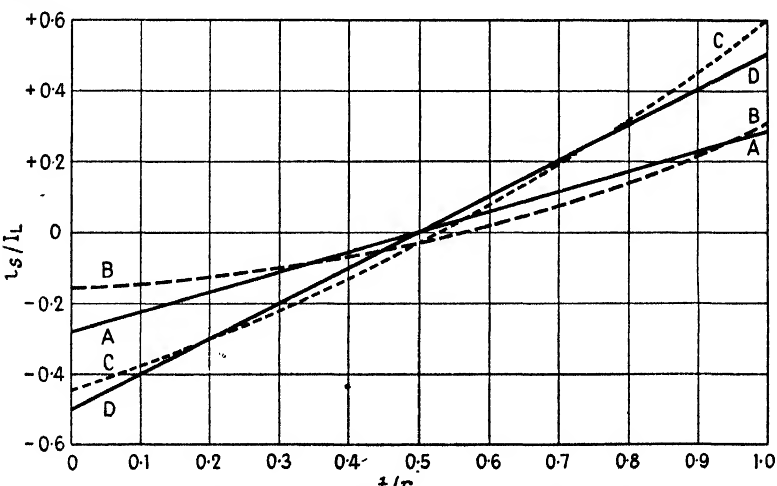



**Fig. 7.8—Current waveforms required for a linear scan with a transformer; A linear component, B square-law component, C total current, D deflector-coil current**

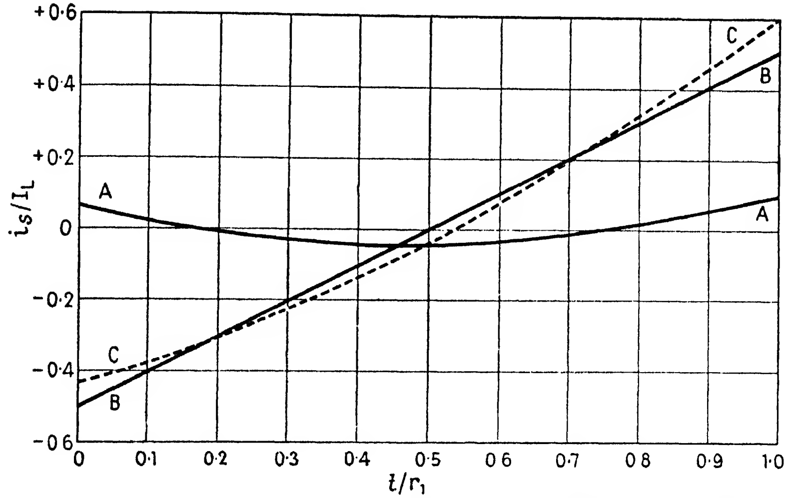



**Fig. 7.9—Currents corresponding to Fig. 7.9 but differently grouped; A transformer shunt current, D deflector coil current, C total current**

The shunt resistance $R_1$ represents core losses in the transformer and also any actual resistor which may be connected across the transformer. Such a resistance is usually needed with a pentode to prevent the fly-back from being too rapid, but is usually unnecessary with a triode.

Ignoring the small square-law component of voltage developed across $r_p$, the resistance $R_1$ draws a linear current which is additive to the primary current and must be supplied by the valve. It, therefore, reduces the ratio of square-law to linear currents which must be provided by the valve. The simplest way of taking it into account, with sufficient accuracy for most purposes, is to transfer it to the secondary as $R_1k^2/n^2$. Appearing in shunt with $r_s + R_L$, it reduces their effective value and so reduces the effective resistance/inductance ratio.

If the valve is a triode and $R_1$ is not present the valve has to supply only the primary current and the loss current of $R_1$ is avoided. It is easy to show that if the two valves have the same mutual conductances $g_m$, and the a.c. resistance $r_a$ of the triode is equal to $R_1$, (and, as will appear later, this last is necessary for the same fly-back times), the two valves need identical grid voltages.*

In practice the mutual conductance of a triode is generally

* Let $E_p$ be the primary voltage across the transformer, $i_p$ the primary current and $i_a$ the anode current. Then with a triode

$$\mu\, e_g = i_a\, r_a + E_p$$

and with a pentode

$$i_a = g_m\, e_g = i_p + E_p/R_1$$

But with the triode

$$i_a = i_p \text{ and } \mu = g_m\, r_a, \text{ therefore,}$$

$$g_m\, e_g = i_p + E_p/r_a$$

which equals the pentode expression if $r_a = R_1$.

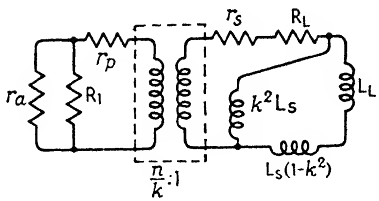



**Fig. 7.10—Equivalent diagram of transformer-coupled stage for fly-back conditions**

considerably lower than that of a pentode. This will result in the triode needing a bigger input voltage than the pentode but it will not affect the waveform of the voltage. It may, however, make it more difficult to obtain that voltage. The pentode, too, must supply a larger current.

The equivalent circuit of Fig. 7.7 does not hold on fly-back because the deflector-coil and leakage inductances have been omitted. Instead, the circuit of Fig. 7.10 is suitable. This is not an exact equivalent because $k^2L_S$ is connected to the wrong side of the resistances. This greatly simplifies calculation and the error is quite small. Equ. A.5.29 in Appendix 5 gives the relation between resistance and inductance.

This transformer-coupled stage comes the nearest of any to ordinary a.f. amplifier practice because the inductive back e.m.f. is negligibly small on the scan. The valve works into a substantially resistive load and to a first approximation it is possible to choose the transformer ratio in the ordinary way so that $(R_L + r_s)\, n^2/k^2$ is equal to the optimum load required by the valve. On fly-back, conditions differ from a.f. practice and it is necessary to watch that the output resistance of the valve is not too low for a quick fly-back. Even when the a.c. resistance of the valve is itself too low, however, it is usually possible to increase it artificially by negative current feedback from the cathode-bias resistance.

The optimum transformer ratio is not always used because it may lead to inconvenient current and voltage requirements for the valve. The use of a lower ratio, with greater current but lower voltage requirements, is sometimes desirable. The frame-scan stage cannot always be designed entirely on its merits. It is usually required to operate from an h.t. voltage which is fixed by the needs of the other stages.

The general commercial practice is to use a pentode when the h.t. supply is low,—200 V or so,—and a triode when it is relatively high,—300–350 V. Many designers prefer the pentode even then, however.

As a final example, let us consider a design for a deflector coil of 6·5 mH and 11·7 Ω requiring a current of 0·5 A p–p. This is representative of the requirements of a rather low efficiency coil with a tube operating at 5 kV or a good coil with a tube voltage of 9 kV. Let us also assume $r_s/L_S = 5\ \Omega/\mathrm{H}$, a unity ratio of square-law to linear currents, and a primary inductance of about 50 H.

The design equations give $L_S = 0{\cdot}233$ H and so $n = 14{\cdot}7$. Let us make $n = 15$ and $L_p = 52{\cdot}5$ H; then $r_p = 262{\cdot}5\ \Omega$ and $r_s = 1{\cdot}17\ \Omega$. The resistance reflected into the primary of the transformer is

$$R_p = 225\,(11{\cdot}7 + 2{\cdot}34) = 3160\ \Omega$$

The effective inductance is (assuming $k = 0{\cdot}995$)

$$L_e = 225\,\frac{0{\cdot}0065 + 0{\cdot}233\,(1 - 0{\cdot}99)}{1 + 0{\cdot}0065/0{\cdot}223} = 1{\cdot}93\ \text{H}$$

The effective resistance connected to the transformer primary is

$$R_T = 3{,}260 \times 1{\cdot}93 - 3{,}160 = 3{,}140\ \Omega$$

The primary current is

$$I_p = \frac{0{\cdot}5}{15}\left(1 + \frac{0{\cdot}0065}{0{\cdot}233}\right) = 0{\cdot}0343\ \text{A} = 34{\cdot}3\ \text{mA}$$

The peak primary voltage on the scan is

$$E_p = \frac{15 \times 0{\cdot}5 \times 12{\cdot}87}{2} + 262{\cdot}5\,i_o = 48{\cdot}1 + 262{\cdot}5\,i_o$$

If we guess 5 mA as the minimum anode current $i_o$, we get

$$E_p = 48{\cdot}25\ \text{V approx.}$$

The mean anode current of the valve will be roughly $I_p/2 + i_o$ or about 22 mA. A pentode is a likely choice and $r_a$ will be too high to need considering, so $R_T$ is virtually the resistance across the transformer primary, $R_1$. To allow for core losses and to give a tolerance $R_1$ would probably be around 10 kΩ in practice. Assuming this, the peak anode current of the valve is

$$i_{ao} = 0{\cdot}03933\left(1 + \frac{262{\cdot}5}{10{\cdot}000}\right) + 15 \times 0{\cdot}5 \times \frac{12{\cdot}87}{20{,}000} = 45{\cdot}1\ \text{mA}.$$

The h.t. supply must be 48·25 V plus the minimum permissible anode-cathode voltage (say 100 V) plus the cathode bias (say 10 V), —a total of about 160 V. This fits nicely into the requirements of a.c./d.c. technique.

## Line Scan

Because of the higher rate of change of current on both scan and fly-back, conditions in the line circuit are quite different from those in the frame. The series resistances of deflector coil and transformer can usually be neglected and it is the inductances which are important.

Transformer coupling is almost invariably used and, as will be shown later, the main effects of the transformer are to add inductance to that of the deflector coil and, of course, to provide a turns ratio. A transformer commonly has a step-down ratio from the anode of a valve to the deflector coil and also a step-up ratio from the anode to the e.h.t. rectifier. This is apt to be confusing at first and at present the e.h.t. circuit will not be considered; the transformer then becomes a simple one between the valve and the deflector coil.

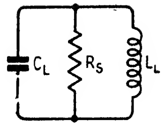



**Fig. 7.11—Equivalent circuit of deflector coil at line frequency**

This coil of inductance $L_L$ carries a linear current of peak-to-peak value $I_L$ during the scan of duration $\tau_1$ and so the back e.m.f. across it is $L_L I_L/\tau_1$ and is constant. Any shunt resistance $R_S$ must, therefore, draw a constant current $L_L I_L/\tau_1 R_S$ and any shunt capacitance $C_L$ must be charged to the voltage across the coil but draws no current. This basic circuit is shown in Fig. 7.11.

When the transformer is added the circuit takes the form shown in Fig. 7.12. This is an equivalent circuit and the transformer enclosed in a dotted box is an ideal one having only the ratio $n/k : 1$ as its electrical property. The leakage inductance is represented by $L_S\,(1 - k^2)$ and the shunting effect of the secondary by $k^2L_S$ where $L_S$ is the secondary inductance of the real transformer and $k$ its coupling coefficient.

The capacitance and resistance of Fig. 7.11 are transferred to the primary as $C$ and $R$ and include also the capacitance and resistance of the transformer. In Fig. 7.12 the deflector-coil current of peak-to-peak value $I_L$ flows through $L_S\,(1 - k^2)$ as well as $L_L$. The back e.m.f. on the ideal transformer secondary is, therefore,

$$E_S = \{L_S\,(1 - k^2) + L_L\}\, I_L/\tau_1$$

and this is greater than the back e.m.f. $E_L = L_L I_L/\tau_1$ on the deflector coil alone.

The shunt inductance $k^2L_S$ draws current because $E_S$ exists across it and the total secondary current becomes

$$I_S = I_L + \tau_1\, E_S/k^2L_S$$

and is greater than the deflector coil current. The product $E_L I_L = L_L I_L^2/\tau_1$ represents the deflector-coil volt-amperes. The product $E_S I_S$ represents the volt-amperes which must be provided by the secondary of the ideal transformer if $E_L I_L$ is to exist in the deflector coil.

The primary voltage is $E_p = E_S n/k$ and the primary current is $I_p = I_S k/n$ so that $E_p I_p = E_S I_S$ and $E_p I_p$ represents the volt-amperes which must be supplied to the primary of either the real

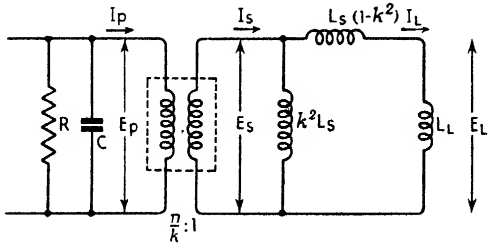



**Fig. 7.12 — Equivalent circuit of deflector coil and transformer**

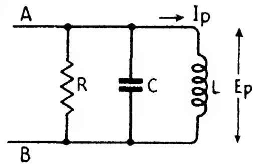



**Fig. 7.13—Equivalent circuit of deflector coil and transformer in the primary circuit**

or the ideal transformers. The efficiency of the transformer can be denoted by $\eta_T = E_L I_L / E_p I_p$ and it is shown in Appendix 5 that this is a maximum when

$$\frac{L_S}{L_L} = \frac{1}{\sqrt{1 - k^2}}$$

and then has the value

$$\eta_T = \frac{k^2}{(1 + \sqrt{1 - k^2})^2}$$

It commonly happens that the need for very high insulation and low self-capacitance in the transformer make it impossible to secure very tight coupling between the windings. As a result, the coupling coefficient $k$ may not exceed 0·98. With this value $\eta_T$ is only 0·665 so that the transformer is not a very efficient device. Efficiency figures of 0·5 to 0·7 are common.

Referring to Fig. 7.12 the three inductances on the secondary can be grouped as one and this one transferred to the primary by multiplying its value by $n^2/k^2$. When this is done, *and when $L_S/L_L$ has its optimum value*, the equivalent primary inductance becomes $n^2 L_L = L$. The circuit can then be reduced to that of Fig. 7.13 with the peak-to-peak scan current $I_p = \frac{k}{n} \frac{I_L}{\sqrt{\eta_T}}$ in $L$ and the voltage $E_p = E_L \frac{n}{k} \cdot \frac{1}{\sqrt{\eta_T}}$ across $L$.

Fig. 7.13 represents the actual state of affairs very closely. In deriving it the main errors involved are the neglect of the series resistances of the windings and of capacitance on the secondary. Since the latter comes straight across the deflector coil and not directly across the ideal transformer secondary it makes the network a complex one. Fortunately, with usual values its effect is very small for transformer ratios exceeding 3 : 1 and of only minor importance with somewhat lower ratios. The main effect of the winding resistances is to make the scan non-linear when the circuit is fed with a linear current or, conversely, to make it necessary to include a square-law current in the input to obtain a linear scan. The effect is the same as in the case of the frame scan, but very much smaller in magnitude.

The circuit of Fig. 7.13, representing as it does conditions at the primary of the transformer, is to be considered as connected in the anode circuit of a valve. The coil will therefore carry the mean anode current, whereas the actual deflector coil does not. This

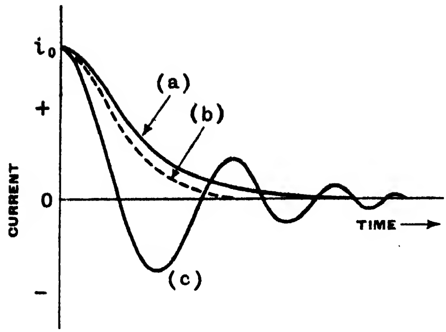



Fig. 7.14—The form of the current on fly-back is shown here for heavy damping (a), critical damping (b), and the oscillatory condition (c)

need cause no concern for a double-wound transformer removes this d.c. component and if an auto-transformer is used a blocking capacitor is inserted.

Considering Fig. 7.13, if B is joined to the anode of the valve and A to + h.t. the current at the end of the scan has its peak value and is flowing downwards in $L$ as shown. This is a conventional current; the electron current flows upwards. Call this peak value $i_o$. The back e.m.f. across $L$ makes terminal B negative with respect to terminal A by an amount $v_o$, and $R$ draws a current $v_o/R$. If the scan is linear $C$ is charged to $v_o$ throughout and takes no current.

At the end of the scan and the start of fly-back the valve is cut off suddenly and becomes inoperative. Fig. 7.13 then represents the whole circuit during fly-back. At the start of fly-back there is a current $i_o$ in $L$, and $C$ is charged to the voltage $v_o$. In order to understand line-scanning circuits of any type it is imperative to understand what happens under these conditions for it is a matter unavoidably common to all such circuits.

The current in $L$ continues to flow in the same direction but decreases in amplitude. It flows into $C$ and $R$, charging $C$ in the opposite direction to the initial charge on it. At first this merely reduces the charge on $C$, but after passing through zero, it builds up in the opposite direction. The voltage across $C$, therefore, starts with B negative to A, falls to zero, and rises with B positive to A. The voltage across $L$ is the same, of course, and depends on the rate of change of current. The resistance $R$ draws a current which increases as the voltage rises.

The precise form of the current and voltage variations depends on the value of $R$ in relation to the $L/C$ ratio. If $R$ is below a certain critical value the current dies away to zero at a decreasing rate. The voltage rises rapidly to a positive maximum and then dies away to zero. The general forms of the changes are shown by curves (a) in Figs. 7.14 and 7.15. If $R$ is increased to the value giving critical damping the waveforms are much the same, curves (b), but the fly-back is somewhat more rapid. If $R$ is increased still further, however, the oscillatory condition is reached, curves (c).

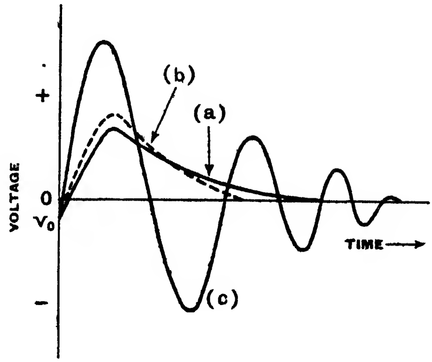



**Fig. 7.15**—The form of the voltage across the circuit on fly-back is shown here for the same three conditions as in Fig. 7.14

What happens is that $R$ does not absorb the energy from the magnetic field quickly enough. If $R$ were infinite the current would fall rapidly to zero, at which time the energy in the magnetic field would be zero and the original energy would have been entirely transferred to the electric field of $C$. The voltage across $C$ would then be a maximum. After this $C$ would start to discharge and to drive a reverse current through $L$ until after another quarter cycle the energy had been transferred back to the magnetic field. The process of energy interchange would go on for ever and the voltage and current would vary sinusoidally as continuous oscillations.

In practice $R$ is not infinite and it absorbs energy so that the voltage and current vary in the form of damped oscillations. Each successive maximum is less than its predecessor. It is sometimes convenient to indicate the amount of damping, not by the relation of $R$ to $L/C$, but by the amount of overshoot since this is readily observable by looking at the waveform with an oscilloscope. By overshoot is meant the ratio of the first negative current maximum to the initial value of current. With no damping the ratio would be unity; in practice, it can be made as high as 0·9, but with ordinary-grade components 0·3–0·6 is more probable.

It is obvious that oscillations such as this cannot be permitted for they would appear as gross irregularities of the scan. They would not only alternately compress and expand the scan in narrow bands but, through velocity modulation, they would cause vertical bright and dark bands to appear on the picture. Nevertheless, almost all scanning circuits are of oscillatory form.

The rate of change of current and the voltage across $C$ are both zero at the negative maximum of current. After this the current is changing in the same direction as the required scan current, and very slightly after the maximum it is changing at the proper rate so that the voltage across $C$ is the proper voltage for the scan condition. If at this instant the voltage could be clamped at this value the current would necessarily have to change linearly at the proper scan rate and all subsequent oscillation would be prevented.

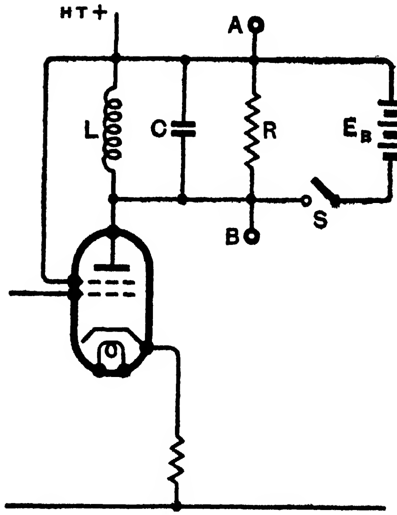



Fig. 7.16—Basic circuit of a damping diode system

In Fig. 7.16 this result is achieved by closing the switch S at the instant that the voltage across $C$ reaches the scan voltage and this puts the battery $E_B$ of the same voltage as on the scan across $C$. Thereafter, the voltage is fixed and the current flows linearly into the battery. The waveform is shown in Fig. 7.17. From the start of fly-back until the point A, which is only very slightly beyond the negative maximum, the circuit is free and is best as lightly damped as possible. At A the switch comes into action and the current decays linearly to B. The portion A B forms part of the scan and the driving valve in Fig. 7.16 remains cut off from the start of fly-back until B. At B it starts to drive a linear current into the circuit to finish the scan, B to C. After B the switch can be opened again.

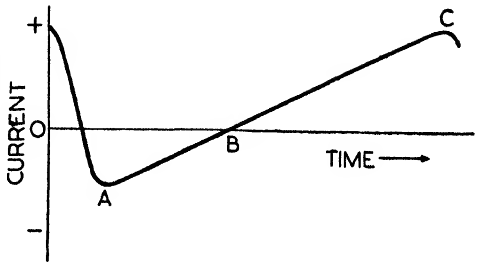
The peak current at the end of the scan is $i_o$. With no circuit losses, the overshoot is 100 per cent and the current at the negative maximum is $-\ i_o$, while at point A it is only a few per cent less than this. Therefore, one-half of the scan is provided from the energy stored in the magnetic field at the end of the previous scan, and $i_o = I_p/2$.

If the scan period is $\tau_1$ and the fly-back period $\tau_2$ the valve is conductive for $\tau_1/2$ only, or for the fraction

$$\frac{\tau_1/2}{\tau_1 + \tau_2}$$

of the complete cycle. This fraction has the value 0·42 for the British standards. If the valve provides a linear current of peak value $i_o$, its mean current while conducting is $i_o/2$ and its mean current over a whole cycle is $0{\cdot}21\ i_o = 0{\cdot}105\ I_p$.

**Fig. 7.17—With a damping diode the driving valve is cut off up to the point B, and the initial part of the scan AB is provided by the energy stored in the magnetic field at the start of the fly-back**




With no overshoot at all the valve would have to be conductive over the whole period $\tau_1$ or $\tau_1/(\tau_1 + \tau_2) = 0{\cdot}84$ of a cycle and deliver a peak current $i_o = I_p$. Therefore the mean current would be $0{\cdot}42\ I_p$.

The 4 : 1 reduction of current obtainable under ideal conditions with the switch circuit does not represent the limit. The energy stored in the magnetic field at A (Fig. 7.17) is not used in providing the initial part of the scan, it is merely released at the correct rate. It is turned into chemical energy in $E_B$ of Fig. 7.16 and serves to charge the battery. If this battery is a part of the h.t. supply to the valve, and it can be made this merely by joining + h.t. to the negative of $E_B$ instead of to the positive as shown, then the chemical energy is turned back to electrical energy and supplied to the valve during its conductive period to supply part of the energy for the scan.

If the whole system were completely loss free, including the valve, $E_B$ could be the whole h.t. supply. It would only then be necessary to supply enough initial energy to start the circuit working and it would then carry on indefinitely without any continuous power supply!

Energy losses in practice always prevent any approach to this ideal state of affairs. Nevertheless, it is possible to make use of both the basic ideas discussed. A properly designed diode circuit can replace the switch and a capacitor can replace the battery. The use of a large amount of overshoot then enables a considerable direct saving of current to be effected and the energy stored in the capacitor can be used for " h.t. boost."

When the circuit damping is low enough for there to be more than some 10 per cent overshoot the use of a diode switch circuit is essential. Both the current efficiency and the energy recovery then increase as the damping is reduced until the overshoot approaches its limit of 100 per cent. When a diode is used, therefore, it pays to make the damping as light as possible. In practice, no artificial damping is used with such circuits and the losses are due mainly to the iron circuits of the deflector coil and the transformer. For the highest efficiency, special precautions must be taken to minimize these losses.

It is by no means essential to use a diode switch, however, but when it is not employed the circuit must be so heavily damped that the overshoot is only about 10 per cent. With such small overshoot the subsequent oscillations are naturally very small for each is only 10 per cent of its predecessor. The positive maximum following the first negative maximum is only 1 per cent of the initial current.

It is, therefore, practicable to use a small overshoot and to let the valve start to drive shortly after the current maximum without having to use any switching circuit. When the overshoot is correctly adjusted the current after the negative maximum changes for a short period at nearly the right rate for the scan and the onset of the drive from the valve can be delayed to shortly after the second zero of current. The peak current of the valve is therefore about 10 per cent less than $I_p$ and the valve is non-conductive for about 1·5 times the fly-back period. The mean current is, therefore, roughly 0·34 $I_p$ instead of 0·42 $I_p$; this improvement is usually well worth having. The attainment of proper operation entails critical adjustment of both drive and damping.

It is, of course, possible to operate with still higher damping so that the current decay on fly-back is non-oscillatory and has the forms of (a) or (b), Fig. 7.14. The circuit adjustments are then much less critical. The valve is kept operative the whole time and the efficiency is reduced. Because the fly-back is slower with the heavier damping the maximum permissible $LC$ product is much smaller than with the other arrangements. The resulting low value of inductance makes the current requirements prohibitively large and this operating condition is consequently now hardly ever used.

Until recently, the most widely used arrangement has been the 10 per cent overshoot condition without a diode switch. The advent of a.c./d.c. sets, however, has necessitated the adoption of the switch circuits, for without them it is impracticable to obtain sufficient scan at the low h.t. voltage available in such sets.

On the scan with all circuits there is a maximum permissible value of $L$ which is set by the h.t. supply and the current. The h.t. supply voltage must not be less than the cathode-bias voltage of the valve plus the minimum permissible anode-cathode voltage of the valve plus the back e.m.f. across $L$ plus the resistive volt-drops minus any h.t. boost voltage derived from economy circuits. Neglecting the bias and boost voltages this is shown in Fig. 7.18. The back e.m.f. $E_p$ is constant so that the load line CD is vertical. If the resistive volt-drops, which have hitherto been ignored, are

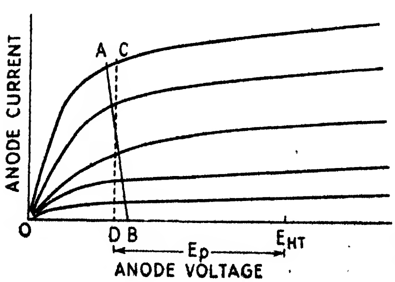



**Fig. 7.18—The anode-volts/anode-current curves of a typical pentode are shown here with the approximate CD and actual AB load lines under line-scan conditions**

taken into account it takes a slight slope and becomes AB. It is important that the point A should lie to the right of the knee in the valve curves. If it does not, serious distortion will result and manifest itself as a cramping of the right-hand side of the picture. With most pentodes the point A corresponds to about 70–100 V. In general, therefore, the sum of the bias and valve voltages is around 100 V.

It is important to note that this operating condition of a nearly constant anode voltage with the valve cut off during fly-back results in the anode dissipation being very low. Thus, the valve may take 100 mA mean current and the h.t. supply may be 500 V. The anode dissipation is not 50 W, as it would be in a class A a.f. amplifier, but 10 W only if the back e.m.f. across $L$ is 400 V so that the anode is at 100 V with respect to the cathode. But if the saw-tooth on the grid is removed the back e.m.f. disappears and the dissipation rises to the full 50 W. Under such conditions it is necessary to be careful to avoid removing the grid input for more than a very short period.

From the known characteristics of the deflector coil and the operating voltage of the tube the value of $L_L I_L^2$ is known. If the transformer efficiency is known $LI_p^2 = L_L I_L^2/\eta_T$. If $I_p$ is known from the valve characteristics and the amount of overshoot which can be used, $L$ can be obtained directly and so the transformer ratio. Then $E_p = LI_p/\tau_1$ and the h.t. voltage can be computed. Alternatively one can start with $E_p$ and determine $L$ and $I_p$ from this.

The value so calculated is the proper value for the assumed conditions but it may not be a permissible value for there is a maximum value for $L$ which is set by fly-back considerations. It depends on the capacitance $C$ and to a small extent on the amount of overshoot. Figures are given in Appendix 5 for the $LC$ products needed for several different degrees of damping. A somewhat larger $LC$ product is permissible with large overshoot than with small; the values range from 26,000 mH–pF to 16,900 mH–pF.

The value of capacitance obtainable depends chiefly on the design of the transformer and it is not easy to make the total less than 50 pF. Values up to 150 pF or more are likely if care is not taken; 75 pF is a reasonable figure to take. Therefore, $L$ must not exceed some 225–340 mH. With a deflector coil of 6·5 mH, this means the transformer ratio is limited to about 6 : 1 to 7 : 1. With a higher inductance coil, of course, the ratio must be lower.

As long as $L$ does not exceed the limit set by the capacitance there is a completely free choice for its value to suit the voltage and current requirements. If the desired value of $L$ is lower than the limiting value it is then permissible to increase $C$ and it is often advantageous to do so. In practice, however, it is not often that one can use $L$ much below the limiting value.

On fly-back the peak voltage reaches a value of over seven times the scan back e.m.f. and the latter is rarely less than 150 V and may

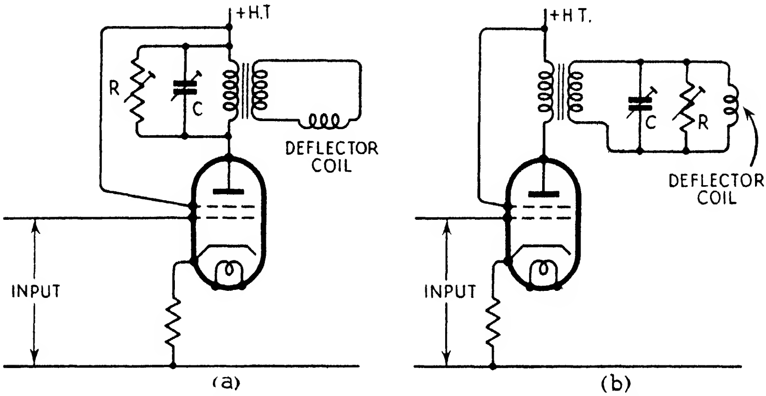



**Fig. 7.19—Line-scan output stage with linearity controls fitted to the transformer primary (a) and secondary (b)**

be as much as 400 V, depending on the h.t. supply. The peak voltage is thus of the order of 1–3 kV and it is usual to employ a valve with a top-cap anode connector; such voltages are too high for the ordinary valve base and holder. The voltage also appears in the transformer, hence the need for high insulation. This is accentuated in many fly-back e.h.t. systems for the transformer voltage is then often 6 kV or more.

It was mentioned earlier that when only very small overshoot is used the damping, and sometimes also the capacitance, need careful adjustment. Two ways of doing this are shown in Fig. 7.19; in (a) a variable resistor $R$ and a variable capacitor $C$ are connected across the transformer primary and in (b) they are joined to the secondary.

On the primary $R$ should be about 80 kΩ and $C$ about 50 pF, whereas on the secondary $R$ can usually be 5 kΩ and $C$ some 0·001 μF. Because of the leakage inductance their effect is not exactly the same on the secondary as on the primary, but the difference is usually very small. The value of resistance needed may vary considerably from these figures for the damping of the circuit depends very much on the core losses in the transformer and deflector coil as well as upon $R$.

The resistance must be rated to dissipate some 25 W and is usually best arranged as a variable resistor in series with a number of fixed resistors. The capacitor must withstand 3 kV if on the primary or 600 V or so if it is on the secondary. In practice, the components are usually fitted to the secondary.

The resistance controls the amount of overshoot, the capacitance the fly-back time, and the saw-tooth input to the valve governs the picture width. If the transformer and deflector coil are efficient, so that the valve can easily provide enough output for a picture of the required width, adjustment is very easy. The capacitance control can usually be dispensed with, the input is increased until the picture is about the right width and $R$ is adjusted so that the scan is linear on the left-hand side. The input and resistance controls are then adjusted in turn until the picture width and linearity are right.

When the valve must be driven hard to obtain a wide enough picture, however, the adjustments are much more critical. The current, viewed from the primary circuit, has the form ABC in Fig. 7.20 (a) on fly-back. The end of fly-back is the point B and beyond this the current is used for the start of the following scan. For an interval after B the valve is still cut off. The valve starts to conduct at D somewhat before the overshoot has died away completely. The valve current is not rising at its full rate, however, because of the curvature of the valve characteristics and between D and C there are both valve and overswing components of current. After C there is only the valve current. The two add together to form the required saw-tooth (b).

If the two currents are to join up correctly it is clearly necessary that each should be just right. The valve input controls the exact point D at which the valve starts to come into operation as well as the peak current at E. The resistance $R$ controls mainly the amount of overswing and the capacitance $C$ governs the precise end of fly-back.

Incorrect adjustment gives on the one hand a picture folded over on the left, or on the other a vertical white line an inch or so from the left-hand edge with the picture expanded to the left of this line. In the first case the damping and/or the capacitance is too great, making the fly-back too slow. In the second, the damping is not enough, the capacitance is too small, or the drive on the valve is too great.

It is usually best to start the adjustments with a picture about an inch less than the full width. Adjust $R$ for linearity on the left-hand side, then $C$, then $R$ again, and so on until really good linearity is obtained. Then increase the input to the valve to widen the picture

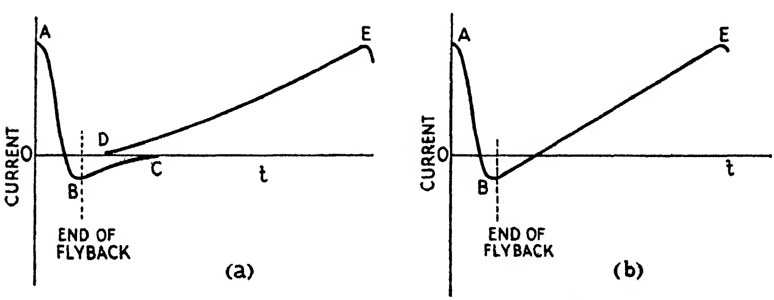



**Fig. 7.20—Under proper operating conditions the fly-back current has the form ABC and the valve current is DE in (a). The part BC forms the start of the scan and joined to DE gives the full wave (b)**

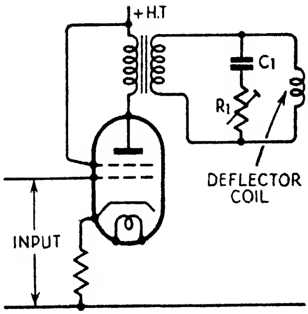



**Fig. 7.21—An alternative damping circuit which is very commonly used takes the form of a series resistor and capacitor on the transformer secondary**

and readjust $R$ and $C$. Continue until the required picture width is obtained. A point will eventually be reached at which it becomes impossible further to widen the picture while retaining the required fly-back time without bringing in a vertical white line or band near the left-hand side. This is the overload point of the valve and it means that the point D in Fig. 7.20 (a) has moved as far as it can do to the right if it is still to join up correctly with the overshoot.

An alternative damping circuit which is commonly used is shown in Fig. 7.21. Here a resistor $R_1$ and a capacitor $C_1$ are connected in series to the transformer secondary; $C_1$ is fixed and of 0·002 $\mu$F to 0·01 $\mu$F capacitance. Adjustment is carried out much as with the other circuit.

During the fly-back $R_1$ and $C_1$ in series are equivalent to resistance and capacitance in parallel; the equivalent parallel values are frequency dependent, however, and varying $R_1$ varies both together. The equivalence ceases to be valid towards the end of fly-back and it is then only a rough approximation to regard the circuit as equivalent to parallel $R$ and $C$.

In the author's experience this circuit is capable of giving slightly better results than the parallel arrangement provided that the core of the transformer is just right for it. The parallel circuit is definitely superior with some core materials and is suitable for a much wider range of materials.

Another modification which is sometimes made is the inclusion of an inductance of about 4 mH in series with $R_1$ and $C_1$ of Fig. 7.21. The aim is to provide a degree of damping which varies during the fly-back. The inductance, as it were, holds off the resistor during the first part of the fly-back, but allows it to have progressively more and more effect as the rate of change of current decreases.

Before leaving this heavily-damped form of circuit, some mention of scan linearity should be made. This is apart from difficulties which may occur on the left-hand side of the picture through the incorrect adjustment of the damping conditions. Because of the series resistance of the deflector coil and transformer secondary and the finite value of the transformer inductance a linear scan towards the right-hand side of the picture is obtainable only if the valve supplies a square-law component of current in addition to the linear current. The effect is the same as in the frame-scan circuit but the relative amplitude of the square-law current needed is usually much less.

In some cases the transformer inductance can be made large enough for the required square-law component to be negligibly small. When it is not, a negative-feedback circuit is one very good way of linearizing the scan. One way of doing this is shown in Fig. 7.22. A 16·5-Ω resistor is inserted in series with the deflector coil and the voltage developed across this is applied to the cathode of the penultimate valve. If the input is provided with a linear *negative-going* saw-tooth the deflector-coil current is maintained extremely linear except, of course, on the left-hand side of the picture. Here linearity is governed only by the damping conditions, for the output valve is cut off and the feedback is inoperative. This circuit, incidentally, includes an e.h.t. system which will be discussed later.

Turning now to the circuits having light damping and large overshoot the switch of Fig. 7.16 is commonly a diode and the basic arrangement is shown in Fig. 7.23. As before, $L$, $C$ and $R$ represent the constants of the deflection system and $R_1$ and $C_1$ form the bias circuit of the diode, replacing the battery of Fig. 7.16.

On fly-back both valves are cut off; $V_1$ is cut off by its grid voltage to produce the fly-back and $V_2$ is cut off by the fly-back voltage, for when the anode voltage of $V_1$ rises it carries with it the cathode of $V_2$, whereas its anode is held by the charge on $C_1$ at the voltage which corresponds to the scan anode voltage of $V_1$. At the end of the scan when the anode voltage of $V_1$ returns to its scan value $V_2$ becomes conductive again and can pass the negative current from $L$ which is needed to form the initial part of the scan. The bias

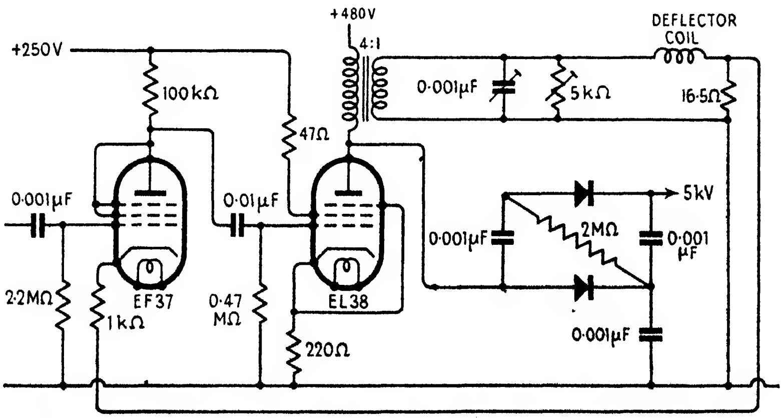



**Fig. 7.22—Line-scan circuit including output stage and pre-amplifier with overall negative feedback and voltage-doubler e.h.t. supply. It requires a negative-going saw-tooth input**

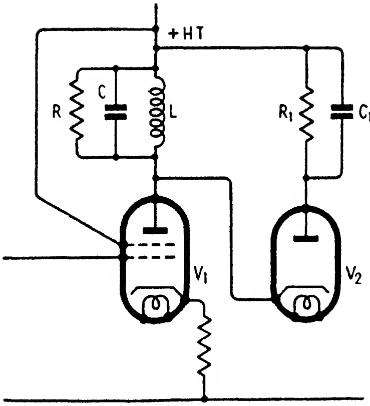



**Fig. 7.23—Basic diode-damping circuit**

voltage on $V_2$ is obtained from this current which flows into $C_1$ and builds up a voltage across it, the value of which depends on $R$.

As it stands, this circuit is not very satisfactory. The bias voltage across $C_1$ is substantially constant, whereas if the scan is linear the voltage on the anode of $V_1$ is falling linearly towards earth (voltage across $L$ increasing) because of the series resistance of the deflector coil and transformer. As a result, the voltage across $V_2$ increases during the scan and the current in $V_2$ must also increase. When $V_1$ is cut off, however, the current is decreasing. Consequently a linear scan cannot be obtained under these conditions.

To obtain a linear scan with this circuit $V_1$ must be conductive throughout the scan and must supply the surplus current needed by the diode. The efficiency is, therefore, low and this circuit is in practice of use only for providing two degrees of damping, light for the fly-back and heavy for the scan. The efficiency is of the same order as with the circuits without a diode, but critical damping adjustments are avoided. However, by connecting the diode to the transformer secondary the voltage across $R$ can be used as h.t. boost and much of the energy otherwise wasted in the diode circuit is recoverable. This is shown in Fig. 7.24 and the alteration is merely the connection of the diode bias circuit in series with the h.t. supply so that the self-bias voltage adds itself to the h.t. voltage. Under this condition $R_1$ is best removed and the transformer ratio adjusted so that the mean currents of the two valves are equal. However, this leaves no latitude for component tolerances and it is often wiser to use a larger step-down ratio so that the diode current exceeds the pentode current and to adjust $R_1$ to absorb the surplus.

For the most efficient operation and a linear scan it is necessary that the diode anode voltage should vary in a certain way. The maximum current in the diode and, hence, the maximum voltage drop across it, occurs at the start of the scan for it must then carry nearly the full current of the overswing. At this instant, therefore, the diode bias voltage should equal the peak voltage across the transformer

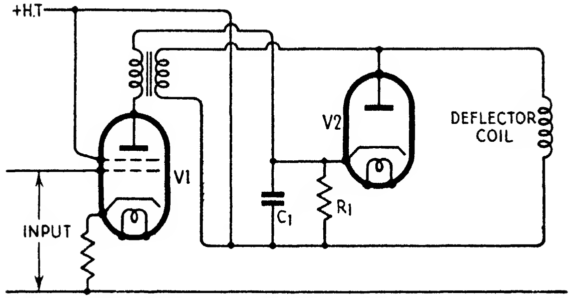



**Fig. 7.24—H.T. boost is obtained with this circuit for the diode self-bias is connected in series with the h.t. supply**

less than the diode drop. As the current falls linearly towards zero the drop across $V_2$ decreases and so the bias voltage should increase by the same amount. When the current drops to zero and $V_1$ takes charge diode current is no longer needed and the bias should increase to keep $V_2$ cut off. After $V_2$ is cut off the form of the bias voltage is unimportant as long as it is sufficient to keep the valve cut off and returns to the proper value for the start of the next scan.

Proper operation of the circuit is not possible unless the bias voltage is controlled and one way of doing this is sketched in Fig. 7.25. The main part of the diode bias is still provided by $C_1R_1$ but a variable voltage is provided by the transformer T in conjunction with $C_2R_2$. When $V_1$ conducts, a voltage is induced in the secondary of T which cuts off the diode. The secondary of T with $C_2$ and $R_2$ forms a heavily-damped resonant circuit and the voltage swing in it gives approximately the right control voltage for the diode.

In practice, a damping diode is not usually connected to the primary of the transformer because its heater supply becomes difficult.

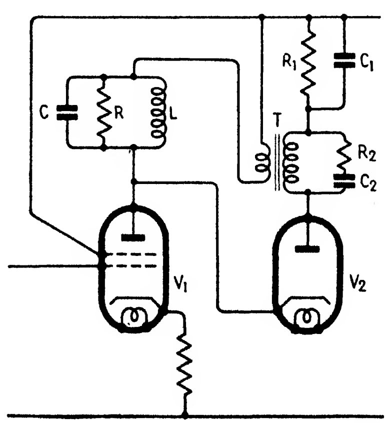



**Fig. 7.25—Controlled diode-damping circuit**

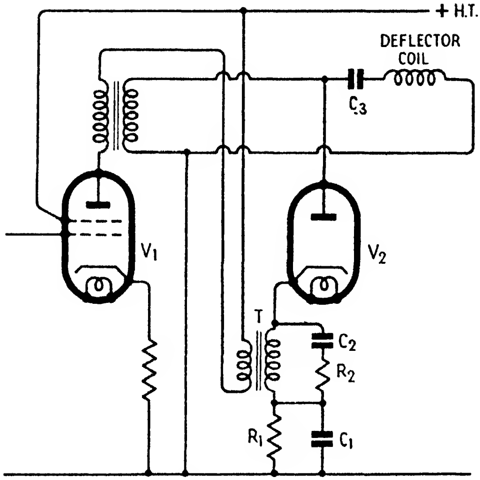



**Fig. 7.26—Practical circuit of controlled diode system**

The cathode voltage varies from about 100 V to perhaps 2 kV above earth during the cycle and to avoid heater-cathode insulation troubles heater and cathode must be joined together. As a result, the heater supply becomes expensive. This can be avoided by connecting the diode to the transformer secondary. The voltage variations can then be upon the diode anode instead of the cathode, for the polarity of the transformer secondary voltage can be made the opposite of the primary.

One arrangement is shown in Fig. 7.26. The transformer is poled so that the diode anode swings negatively when the pentode anode swings positively and vice versa. The capacitor $C_3$ is usually required to prevent the raster being displaced sideways. Without it the mean diode current flows through the deflector coil and transformer secondary in parallel and divides between them in inverse proportion to their resistances. To avoid scan distortion $C_3$ should be large—at least 0·5 $\mu$F.

For proper operation the adjustments are critical. The input to $V_1$ and its cathode bias must be correct and $R_1$ and $R_2$ should be adjustable. In addition, it is desirable to have the inductance of T adjustable by a dust-iron core.

It is important to note that these switch circuits cannot control the energy stored in the leakage field of the transformer. At the end of the scan the driving valve is supplying its maximum current to the transformer primary; the transformer magnetizing current and the deflector-coil current are at a maximum. Part of the magnetic field of the primary does not link with the secondary, however, and part of the magnetic field of the secondary does not link with the primary. This has been expressed hitherto by the transformer leakage inductance $L_s$ $(1—k^2)$.

It is obvious, however, that a diode switch circuit on the secondary cannot control the release of the energy stored in that part of the field of the primary which does not link with the secondary. As a result, energy cannot be recovered from the leakage inductance by a secondary diode and the efficiency is reduced. In addition, the energy in the leakage inductance is dissipated in oscillatory fashion in a series resonant circuit comprising $L_s$ $(1 - k^2)$, the primary capacitance $C$ and the diode resistance. This circuit is unfortunately coupled to the deflector coil by the diode resistance with the result that the oscillatory current is transferred in part to the deflector coil. It produces a pattern of light and dark bars on the left-hand side of the picture. The trouble can sometimes be overcome by connecting a series-resonant circuit in shunt with the diode and tuning it to the natural frequency of the oscillations. It does not suppress the oscillations but the tuned circuit effectively short-circuits the diode at this frequency and largely removes the coupling.

It is, however, better to keep the leakage inductance small. This increases the natural frequency of the subsidiary resonant circuit and so its damping tends to increase, and it may even become non-oscillatory. Also, as the energy stored in it is reduced, the amplitude of the current on its release is also reduced and its visible effect on the picture may become negligible. Furthermore, the reduction of leakage inductance improves the transformer efficiency and so the efficiency of the whole circuit. It is, therefore, very important to keep the leakage inductance at its minimum practicable figure consistent with the requirements of low self-capacitance and high insulation.

Referring to Fig. 7.26, the mean cathode voltage of $V_2$ is positive to earth. For a given overshoot, the mean diode current in $R_1$ increases as the transformer ratio is increased and the mean cathode voltage decreases, the power in $R_1$ being constant. At one particular ratio the mean current of $V_2$ equals the mean current of $V_1$. When this happens $R_1$ can be removed and replaced by the effective resistance of the output stage. The voltage across $R_1$ is then added to that of the h.t. supply as h.t. boost. This is shown in Fig. 7.27. The operation of the circuit and, in particular, the linearity, are critically dependent on the transformer ratio for it is essential that the mean currents of the two valves should be equal. The mean currents will depend on the precise values of the circuit constants and especially on the damping imposed by the core material of the transformer, since this affects the overshoot. It is often advisable, therefore, to make the transformer ratio higher than the critical value so that the mean diode current exceeds the mean pentode current and to bleed the surplus current away in a resistance. This is shown by $R_1$ in Fig. 7.27 and it should be adjustable to provide current matching. The higher diode current means a smaller boost voltage, of course, and so lower efficiency.

Many modifications of this circuit are possible and one is shown in Fig. 7.28. Here the transformer T is replaced by the inductance $L_1$ and two capacitors $C_1$ and $C_2$ of the order of 0·05 $\mu$F. Both $L_1$ and $R_1$ are adjustable as linearity controls.

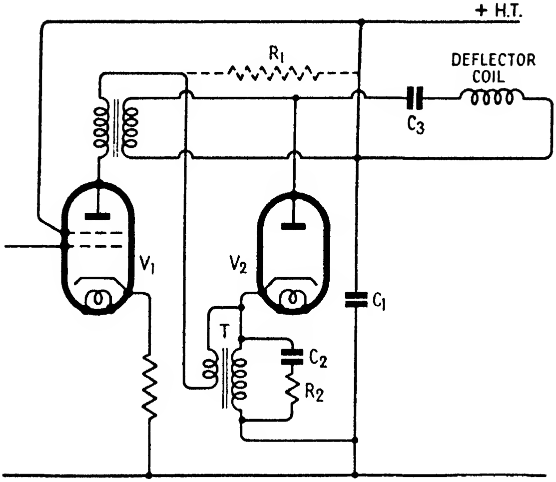



**Fig. 7.27—Circuit for giving h.t. boost as well as current saving**

In these linearity circuits the diode cathode is above earth by the full h.t. supply voltage plus the boost voltage, so that the heater-cathode insulation must withstand it or a separate heater winding be used. Sometimes this can be overcome by arranging to insert the boost voltage in series with the cathode of the pentode, but it is not always so convenient to do this.

In the U.S.A. a triode is often used instead of a diode and it has the advantage that it is easily controlled by a voltage on the grid and this voltage can be derived through an *RC* network. The circuit is shown in Fig. 7.29. It functions exactly like the diode circuit but instead of needing a transformer or inductance-capacitance circuit to control the current it is done by $R_1C_1$, $R_2C_2$. The valve used is a 6AS7 and there is no British equivalent.

Whether the valve is a diode or a triode it must be able to carry a peak current of the order of 250 mA with as low a voltage drop as possible and it must withstand up to about 1 – 1·5 kV when non-conductive. An ordinary 500-V h.t. rectifier is quite suitable.

These current saving and h.t. boost circuits are an important development, especially when large scan power is needed. It must be understood, however, that the improvement in efficiency with them is dependent upon the amount of overshoot obtainable with the basic resonant circuit formed by the scanning components. Unless the overshoot exceeds about 30 per cent they are hardly worth while.

Now the amount of overshoot depends chiefly on the power losses in the deflector coil and transformer. Using ordinary grades of iron with laminations of 0·014-in, core and dielectric losses will restrict the overshoot and it is unlikely that much more than 30

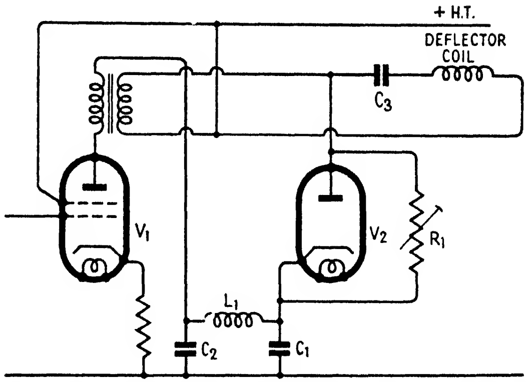



**Fig. 7.28—Simple boost circuit which does not require a special transformer for controlling the diode**

per cent will be readily obtainable. With care in design 50 per cent overshoot may be possible.

If an attempt is made to use these " economy " circuits with an overshoot of this order, difficulty is likely to arise because the efficiency of the deflection circuit is critically dependent on the losses of components. For success it is necessary to control these losses closely and this may be difficult.

In general, therefore, the aim should be to achieve an overshoot of some 90 per cent with the highest loss components likely to be used. No matter how much lower the losses may be in individual components, the overshoot cannot possibly reach 100 per cent and is unlikely to exceed 95 per cent. The variations of efficiency between one receiver and another can then be kept small.

Such large overshoot demands the use of special core materials. Some designers use dust-iron, others prefer Caslam or Ferroxcube. Materials of this nature are really necessary if full advantage is to be taken of these diode circuits.

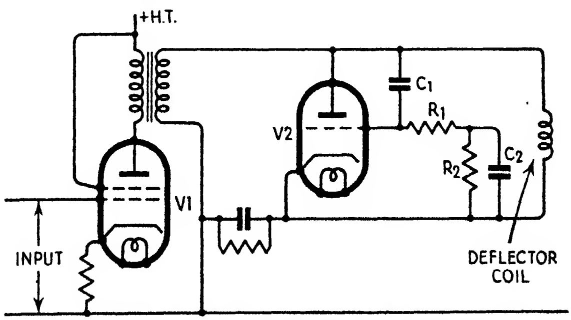



**Fig. 7.29—This diagram shows a current-saving circuit embodying a triode in place of the diode**

Apart from core losses the transformer needs careful design if it is to be satisfactory. The requirements of a large coupling coefficient and low self-capacitance are conflicting and a solution is made more difficult by the necessity for high insulation.

The need for high insulation seriously complicates the transformer design. It is not at all difficult to find insulating materials which will withstand the voltage without puncturing, the difficulties nearly all arise from surface leakage. It is probable that the most compact and efficient design would be one in which the windings were embedded in a suitable insulant, such as polyethylene, by a moulding process. It is rather impracticable, however, for anything but large-scale quantity production.

With the usual fabricated process of construction it is essential to provide very long leakage paths if the transformer is not to break down in a short time. If moulded bobbins can be used, or bobbins turned from *solid* material, such long leakage paths are often conveniently obtained by deeply grooving the bobbin. The path length is then increased by twice the depth of the groove. This increased length of path, however, is not necessarily obtained if the bobbin is fabricated by winding paper or other sheet material on a mandrel. If it is afterwards very thoroughly impregnated, so that it becomes virtually a solid, it may be satisfactory, but there is always a risk of a leakage path developing within the material along the faces of the layers composing it.

It should not be thought from this that laminated bobbins are unsuitable. They can be quite satisfactory. However, it is not safe to assume that the leakage path between the bottoms of two adjacent slots is up the sides of the slots and over the outside face—it may well be directly through the wall between them.

The need for high insulation helps in attaining low self-capacitance, but it makes it much more difficult to obtain low leakage inductance. It becomes almost impossible to use the many interleaved primary and secondary sections which are the normal way of securing tight coupling between the windings of a transformer. About the most that is practicable in this way is to divide one winding, usually the primary, into two sections with the other between them. The use of three concentric bobbins is one of the easiest ways of doing this.

It is usually advisable to choose insulating materials which will withstand a good deal of heat. In transformers designed for use in heavily damped circuits core losses in the transformer are relied upon to provide much of the damping. The core then gets quite hot and it is necessary that the insulating material in contact with it should be able to withstand it. For this reason, a wax dip for the windings should not be used unless it is known that the core is designed to run cool.

One frequent difficulty is the avoidance of mechanical noise, in the form of a whistle, from the transformer. It is almost impossible to prevent the transformer from producing it and the sound can be avoided only by so boxing the transformer or set or both that the sound is contained within it. One common way of doing this is to enclose the transformer in a metal case lined with sponge rubber. It must float in the rubber if the screening is to be effective. As the ventilation will be very poor the transformer losses must be kept to a minimum.

So far, but little has been said about the magnitude of the voltage peak on fly-back. The voltage varies in the form of a half-sine wave and reaches a peak value of between 7 and 8 times the scan back e.m.f., the exact value depending on the damping.

The peak voltage is commonly around 2–3 kV and it can be utilized to provide the e.h.t. supply for the cathode-ray tube by adding a rectifier circuit. Sometimes a voltage-multiplying rectifier is used to develop an output of 5–6 kV; more often, a single rectifier only is employed and the voltage is increased by winding the primary of the line-scan transformer as a step-up auto-transformer. For unusually high voltages both the auto-transformer and the voltage-multiplying rectifier can be used together.

A typical arrangement is shown in Fig. 7.30. The normal transformer primary is AB, but additional turns BC are provided to step up the voltage further to the rectifier $V_2$. The peak input to $V_2$ is the peak anode voltage of $V_1$ multiplied by the ratio of turns AC to turns AB. This peak voltage at the anode of $V_2$ is conveniently called $V_m$ and the mean e.h.t. voltage across $C$ may be termed $V_{HT}$.

As long as $V_m$ is less than $V_{HT}$ the diode is non-conductive and the current for the tube is supplied from the reservoir capacitance $C$. This naturally drops the voltage, but if $C$ is large enough it drops it by a negligible amount and it is quite accurate enough to assume the voltage to be constant. On fly-back the voltage on the anode rises rapidly and when it reaches $V_{HT}$ the diode conducts. If it is a low-resistance diode, the input voltage is very nearly clamped to $V_{HT}$ and rises very little more; it never reaches $V_m$, the value it would attain if the diode were absent. The current in the inductance of the circuit, which would otherwise have flowed into the circuit capacitance to charge it to $V_m$, now flows into the reservoir capacitance $C$ and as $C$ is large it does not raise its voltage appreciably.

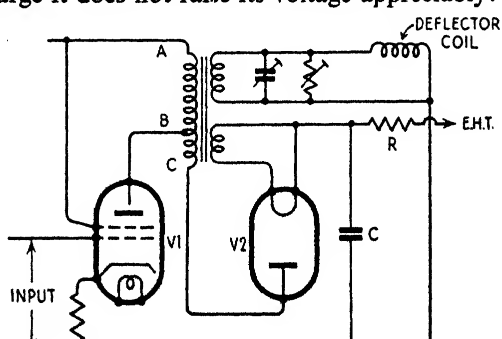



**Fig. 7.30—In this e.h.t. circuit the transformer primary is turned into an auto-transformer to step-up the voltage pulse and increase the output voltage**

After a short interval the deflector-coil current drops nearly to zero, actually to the value needed by the shunt damping resistance. There is then no surplus left for $C$ to absorb and the input voltage falls, cutting off the diode.

The diode conducts for quite a small fraction of the fly-back time and during the rest of the cycle is non-conductive. The charge conveyed to $C$ during this short interval must, in the equilibrium condition, equal the charge lost by $C$ due to the tube current during the full cycle.

If the tube current varies, as it does with changes in the mean brightness of the picture, the e.h.t. voltage varies also. Thus, if the current increases a greater charge must be conveyed to $C$ during the conductive period of the diode. This can only be achieved by conduction starting earlier when the coil current is greater. The voltage is then lower, however, and so $V_{HT}$ must fall to permit conduction to start earlier.

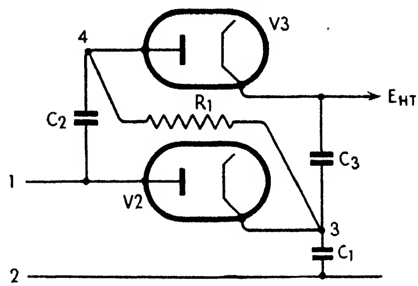
Such changes of e.h.t. are reflected in alteration of picture size and focus because both depend on the final anode voltage of the tube. The focus change is not usually serious and the effect on picture size is much more noticeable.

Since energy is removed from the scan circuit when the diode is conductive the amount of overshoot is affected by the e.h.t. circuit. In other words, it damps the circuit, but not quite in the same way as a simple resistance. The damping effect is small and is negligible with heavily-damped small-overshoot circuits. With these, the effect of the regulation of the e.h.t. supply is to give the same percentage variation of picture width as of picture height with changes of picture brightness. The aspect ratio is unaffected.

With lightly-damped circuits of large overshoot, however, the damping imposed by the e.h.t. supply is not negligible and may affect the overshoot by about 5 per cent. This results in the overshoot being affected by the picture brightness and, hence, the deflector-coil current changes also. This change is in the right direction to reduce the alteration of picture width brought about by the change of voltage. As a result a given change of tube voltage causes a smaller change of picture width with a lightly-damped scanning circuit than with a heavily-damped one. The frame scan, however, is unaffected and so there is a change of aspect ratio. There is a possibility that this could be reduced when h.t. boost is used, by feeding the frame-scanning generator from the boost voltage, for this also will be affected slightly since its magnitude depends on the amount of overshoot. It is unlikely that the change would be sufficient, however.

The magnitude of the voltage obtainable is limited by the circuit capacitance and, in particular, by the self-capacitance of the winding AC, Fig. 7.30. It was shown earlier that there is a definite limit to the capacitance across AB for the proper fly-back time. The effective capacitance across AB rises rapidly as the turns BC are

**Fig. 7.31—The voltage-doubler portion of Fig. 7.21 re-drawn with valves instead of metal rectifiers**




increased and great care in transformer design is needed if a step-up of more than 2 : 1 is needed.

It is the usual practice to operate the filament of the rectifier valve from a winding on the transformer, which must be insulated to withstand the full e.h.t. voltage. Miniature diodes with low filament power are available, so that it does not impose much damping on the circuit. The adjustment of the turns on this winding is difficult because the filament voltage cannot readily be measured. It is usually done by comparing the filament brightness with that of another valve of the same kind heated from a known voltage.

When the filament is heated from the transformer in this way it is clearly inadmissible to control picture width in the usual way by adjusting the input to $V_1$. What is often done is to include a variable inductance in series with the deflector coil for this purpose.

When an adequate step-up cannot be obtained in the transformer, a voltage-multiplying rectifier can be used and metal rectifiers are particularly convenient since they need no filament supplies. A voltage-doubler is shown in Fig. 7.31 and provides output of 5 kV without any step-up from the anode of the valve when the peak voltage on the valve is around 2·7 kV. The voltage-doubler has the reputation of giving poor regulation. It has acquired this, however, in mains rectifier service where the impedance of the source is low and the regulation depends mainly on the rectifier system itself. In fly-back e.h.t. circuits the source impedance is very high, usual 4–6 MΩ, and the rectifier system has relatively little effect on the regulation.

The operation of the voltage-doubler is easier to understand if the circuit is drawn with valves instead of metal rectifiers, if only because most people find it easier to remember which are the conducting and non-conducting directions. The circuit so drawn is shown in Fig. 7.31.

The input voltage at terminals 1, 2, is of pulse form and is shown in Fig.7.32 (a). When the conditions have settled down to the steady state $C_1$ becomes charged so that 3 is positive with respect to 2 by nearly the peak value of the input. $V_2$ and $C_1$ form a normal half-wave rectifier.

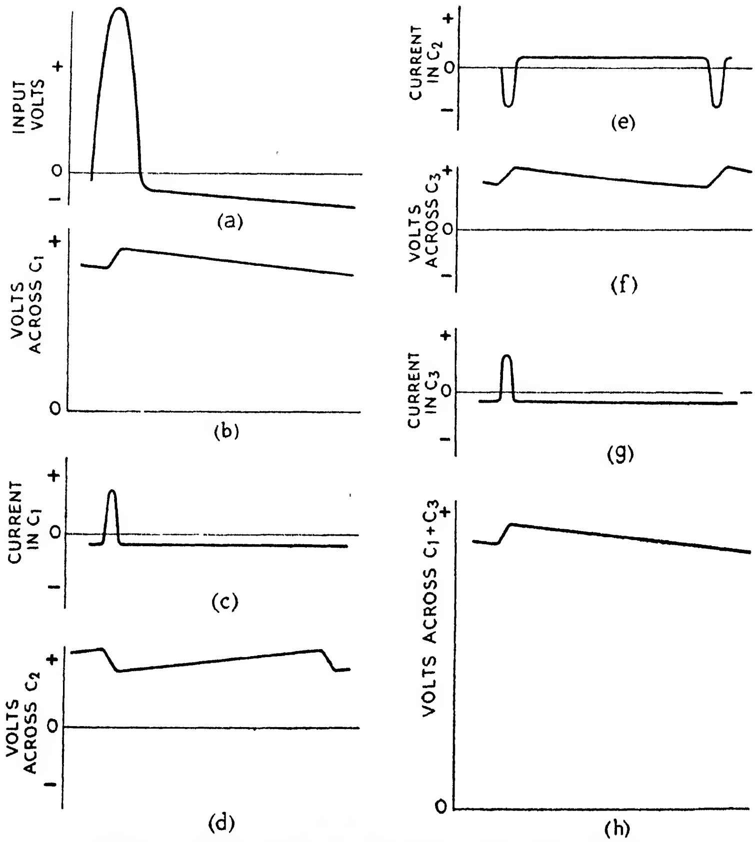



**Fig. 7.32—Waveforms at various points in the circuit of Fig. 7.21**

Assume $C_1$ to be charged nearly to the peak value of the input. Following the peak the input voltage drops and the cathode of $V_2$ is held more positive than the anode by the charge on $C_1$; the valve then becomes non-conductive. After this $C_1$ discharges slowly through its load circuit and the voltage across it falls, and it becomes less than the peak value of the input.

When the next pulse comes along, therefore, it makes $V_2$ conductive and $C_1$ charges through $V_2$ until the voltage across it is nearly equal to the peak value. There is thus a rapid charge of $C_1$ during each pulse and a slow discharge between pulses. The voltage across $C_1$ and the current into it thus vary in the way shown at (b) and (c) in Fig. 7.32.

Now consider the rest of the circuit just after a pulse and take $V_3$ non-conductive, $C_3$ charged but to a lower voltage than $C_1$ and $C_2$ charged to a lower voltage than $C_3$. Because the input voltage has dropped substantially to zero, terminals 1 and 2 can be considered as joined to a first approximation and it is then clear that $C_2$ charges through $R_1$ from $C_1$. The last component thus supplies the charging current for $C_2$ as well as the current to the load.

The capacitor $C_3$ is isolated because $V_3$ is non-conductive and discharges to supply load current. As far as the load is concerned $C_1$ and $C_3$ are in series and the output voltage $E_{HT}$ is the sum of the individual voltages across them. During the interval between pulses $C_1$ and $C_3$ are both discharging and $C_2$ is charging and the voltage across $C_3$ falls below that across $C_2$. This does not make $V_3$ conductive, however, for the anode 4 is positive to 2 only by the voltage on $C_2$ whereas the cathode 5 is positive to 2 by the sum of the voltages on $C_1$ and $C_3$.

When the next pulse comes along $V_2$ becomes conductive as we have seen and it virtually joins points 1 and 3. Then 4 is positive to 1 and 3 by the voltage on $C_2$ and 5 is positive by the voltage on $C_3$; but the former is the greater of the two, so $V_3$ becomes conductive and $C_3$ charges from $C_2$ through $V_3$.

During the scanning period $C_2$ charges from $C_1$ through $R_1$. During the peak of fly-back $C_1$ charges from the source, and $C_3$ charges from $C_2$ since these two capacitors are then virtually in parallel. At all times $C_1$ and $C_3$ in series supply the load. The waveforms are sketched in Fig. 7.32.

In practice it is found that all three capacitors can be alike and about 0·001 $\mu$F while $R_1$ can be 2 M$\Omega$. The rectifiers need to withstand a peak inverse of little more than the peak input, instead of twice this figure as with a sine-wave input. If the rectifiers are rated, as usual, on a sine-wave basis, therefore, one can safely use them for nearly double their rated output in this circuit. The capacitors, too, need be rated only for the peak input voltage.

It is to be noted that with metal rectifiers no bleeder is needed across the output, for this action is automatically performed by the back resistance of the rectifiers.

As already mentioned this form of voltage-doubler is shown in Fig. 7.22 connected to the line-scan circuit. The circuit is substantially that used in the *Wireless World* Television Receiver* and it is important to remember that the values shown apply only if the deflector coil described for that set is used. The use of a less efficient coil would result in the scan being inadequate, while a more efficient coil would lead to an e.h.t. supply below 5 kV when the picture width was properly adjusted.

It has already been mentioned that the regulation of e.h.t. flyback systems is not good. In spite of this, they are very widely used because they are inexpensive and relatively safe. The reservoir capacitance required is small and can store only a limited amount of energy. Consequently, accidental contact with the e.h.t. supply is not nearly so dangerous as with a 50-c/s supply, where the reservoir

* "Television Receiver Construction." Iliffe.

capacitance is about 200 times as great. This advantage exists in other high-frequency e.h.t. systems, of course.

The regulation depends mainly on the relation between the energy needed per cycle for the tube and the energy existing in the scanning circuit and provided for scanning purposes. The former is proportional to the power—the product of tube current and final anode voltage—while the latter is proportional to $LI_p^2$. The use of an inefficient deflector coil and transformer, by making $LI_p^2$ large, improves the regulation. This is often the only way of improving it and it can be an expensive way.

When metal rectifiers are used, however, it is possible to obtain better regulation by choosing their voltage rating carefully with regard to the voltage existing in the circuit. Beyond a certain point the back resistance of these rectifiers falls very rapidly with increasing voltage. Therefore if the back voltage on the rectifier is chosen so that at full load it is just less than the critical value, any rise of voltage on lower loads brings the rectifier beyond the critical point and its reduced resistance limits the amount of the voltage rise.

CHAPTER 8

# Saw-tooth Generators

EVERY TELEVISION RECEIVER has two saw-tooth generators, one for the frame scan and one for the line. They are usually voltage generators; that is, they produce saw-tooth voltage waveforms. Their outputs are applied to the amplifiers described in Chapter 5 if electric deflection is used or to those treated in Chapter 7 if, as is usually the case, electromagnetic deflection is employed. Sometimes, however, saw-tooth current generators are used; the currents then feed the deflector coils directly and the amplifiers of Chapter 7 are not needed.

The use of current generators has not found great favour so far but it has always had some adherents. Before the war it was used in this country by Baird Television and it was adopted in several German sets. It has also been used in the U.S.A. and is employed in at least two current British sets. Nevertheless, the vast majority of television sets have saw-tooth voltage generators with amplifiers, and so the voltage generators will be treated first.

A saw-tooth voltage is invariably generated by alternately charging and discharging a capacitor. If a capacitance $C$ and a resistance $R$ are connected in series with one another and with a battery of voltage $E$, the capacitance charges up until the voltage across it equals the battery voltage. At the instant of switching on there is no voltage across the capacitance and the current $i$ through the resistance is $i = E/R$. When the capacitance is fully charged the current is zero and the voltage across it is $E$.

At any time $t$ after switching on, the current $i = \frac{E}{R}\epsilon^{-t/CR}$ and the voltage $e$ across the capacitance is $e = E - iR = E\,(1 - \epsilon^{-t/CR})$. The curve connecting $e/E$ with $t/T$ where $T = CR$ is shown in Fig. 8.1, to two different scales. It is clear that the rise of voltage is not linear but exponential. Assuming a linear amplifier, the scanning spot will consequently move at an uneven rate across the screen—it will slow up as it moves to the right or downwards, as the case may be. As a result, the right-hand side and the bottom of the picture will be cramped.

It is, however, clear that for values of $e/E$ up to about 0·1 the curve is very nearly linear. If the capacitor is not allowed to charge up to more than a small fraction of the applied voltage, the change of voltage across the capacitor is sufficiently linear for practical purposes. When the amplifier is linear the saw-tooth amplitude should not be allowed to exceed about 5–7 per cent of the applied

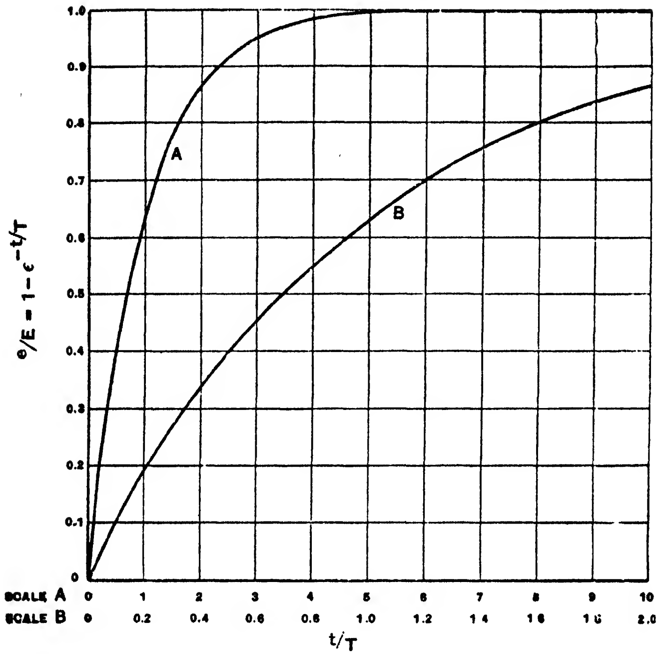



**Fig. 8.1—These exponential curves show the way in which the voltage across the capacitance rises**

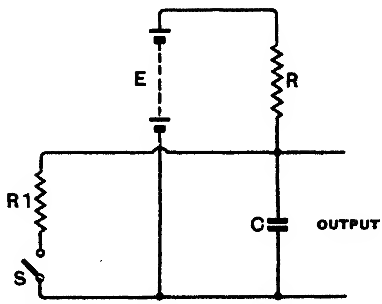
voltage, but when it is not very linear it can be about 10–15 per cent. This is because amplifier distortion is usually in the opposite sense to the non-linearity of the charging circuit and the two distortions tend to cancel.

The basic saw-tooth oscillator circuit is shown in Fig. 8.2. The capacitance charges through the resistance $R$ and the voltage across it rises. After a certain interval of time the switch $S$ is closed and, as $R_1$ is small compared with $R$, the capacitance rapidly discharges through $R_1$. After a much shorter time interval $S$ is opened again and the capacitance starts to charge again.

It is easy to see that if the opening and closing of $S$ are performed regularly the capacitance will be alternately charged and discharged. If the parts are properly proportioned and the switch timing is correct, the voltage across the capacitance will change in the manner shown in Fig. 8.3 (a). On the other hand, if the capacitance is allowed to charge up to too high a voltage, the saw-tooth will be non-linear and of the form shown at (b).

If we decide to charge up to 7 per cent of the supply voltage, Fig. 8.1 shows that $t/T$ should be about 0·07, or $CR = 14{\cdot}25\ t$. In the case of the line scan stroke the time of the stroke is 82·77 $\mu$sec, so that $CR = 1{\cdot}175 \times 10^{-3}$ F–Ω $= 1{,}175$ $\mu$F–Ω. If we make

**Fig. 8.2—A simple saw-tooth voltage generator is shown here. The charge and discharge of the capacitance are controlled by the switch S**




$C = 0{\cdot}005\ \mu\text{F}$, then $R$ would be about 240,000 Ω. For the frame scan the time of the stroke is approximately 18·62 msec, consequently $CR = 0{\cdot}265\ \text{F–}\Omega = 265{,}000\ \mu\text{F–}\Omega$. In practice $R$ might be 2·65 MΩ with $C = 0{\cdot}1\ \mu\text{F}$.

Such values are not uncommon with time bases for magnetic tubes, for the h.t. supply is then low and the saw-tooth amplitude is a fairly large fraction of it. With time bases for electric deflection, the h.t. voltage is usually large, but the saw-tooth amplitude is of the same order as with the magnetic system. The $CR$ products required are consequently larger.

The discharge of the capacitance corresponds to the fly-back and the time required for this depends on the product $CR_1$. As the fly-back time must be much smaller than the scan time, $R_1$ must be much smaller than $R$.

Now in practice $S$ and $R_1$ of Fig. 8.2, are replaced by a valve with which may be associated various components, and even other valves. It is necessary at this point to distinguish between two entirely different kinds of time base. One does not function in the absence of sync pulses, the other generates the saw-tooth waveform independently of the sync pulses, which are used merely to keep it

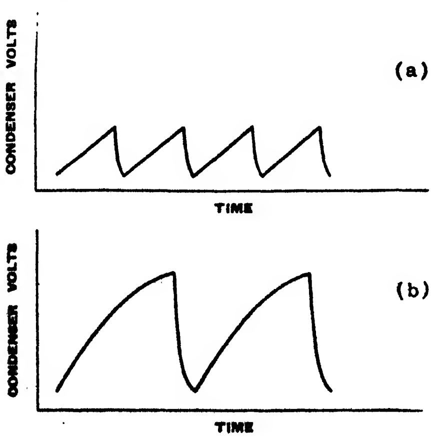



**Fig. 8.3—When only a small amplitude is developed the rise of voltage is nearly linear (a), but when the amplitude is large it becomes markedly exponential (b)**

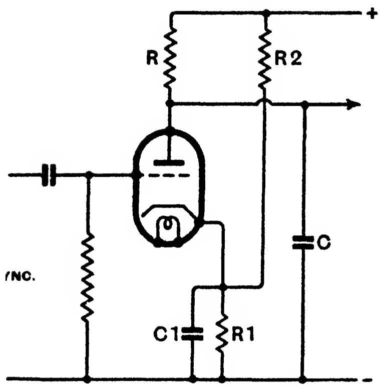



**Fig. 8.4—A practical saw-tooth generator, using a hard valve controlled by the sync pulses in the transmission, is shown here**

running at the right frequency and phase. The former is sometimes called the distant-controlled time base and is rarely used at the present time, the latter is in one form or another almost universal.

In spite of this, we shall discuss the distant-controlled generator first, because it is the simpler. The circuit is shown in Fig. 8.4, and the valve is biased by $R_1$ and $R_2$, so that it does not pass anode current until the anode voltage has risen appreciably above the normal maximum saw-tooth amplitude required.

Starting at the beginning of a line, $C$ charges through $R$ and the voltage across it rises, and hence the anode voltage of the valve. At the end of the line the sync pulse comes along and reduces the grid potential nearly to zero. The valve is consequently conductive and it passes anode current and so discharges the capacitance. The discharge is usually not complete, but this is not very important.

At the end of the sync pulse the grid goes negative again and the valve ceases to conduct. Consequently, the capacitance starts to charge again for the next line.

It may be undesirable to drive the grid of the valve positive on account of grid current, and the valve must, therefore, be of a type taking a fairly heavy anode current at zero grid bias if it is to discharge the capacitance in the time of duration of the sync pulse. A fairly high negative grid bias will then be needed for anode current cut-off, and consequently a large amplitude of sync pulse. Generally about 10–30 V of sync pulse is needed.

It will further be seen that it is essential that all the sync pulses be alike. If one pulse is a little smaller than usual the capacitance will not be as completely discharged as usual, so that it will start to charge for the next line from a slightly higher residual potential than normal. This next line will thus be slightly displaced to the right. Similarly, if one pulse is stronger than usual, the capacitance will be further discharged than usual, and the following line will be displaced to the left.

Extremely good sync separation is needed if satisfactory results are to be secured. Moreover, it is important to retain the rectangular shape of the pulses. If this is not done the difficulties of obtaining an adequate fly-back are increased.

One great disadvantage of the distant-controlled time base is that it only functions when the sync pulses are present. Development is thus only possible during a television transmission, unless a local pulse generator producing the same waveform is available, which is rarely the case. It must be realized, also, that when the transmission ceases the raster collapses so that to avoid damage to the tube it is essential to include some safety device. This might take the form of providing an initial heavy bias on the tube to black out the spot. When the time base is functioning its output can be applied to a rectifier and smoothing circuit to develop a positive voltage for application to the tube to offset the initial bias. If the time-base output fails, the extra bias disappears, and the initial bias is left to black out the spot and to save the tube from damage. Of course, if e.h.t. is taken from the line fly-back the voltage automatically disappears.

The ordinary time base does not suffer from this defect, since it generates the saw-tooth waveform independently of the signal. The sync pulses do not operate the time base, but only control the precise timing of the successive waves.

This type of circuit can be made much more independent of the form of the synchronizing pulses and so can operate much more regularly under conditions of severe interference. The simpler forms of generator do not necessarily do so, however.

One widely used saw-tooth oscillator of this type is shown in Fig. 8.5. A gas-filled triode is used and it functions quite differently from an ordinary hard valve. When the valve is conducting the grid exercises little or no influence on the anode current. It is not possible to stop the anode current by driving the grid negative, as with a hard valve, but only by removing the anode voltage.

If the grid is kept at a negative potential and the anode voltage

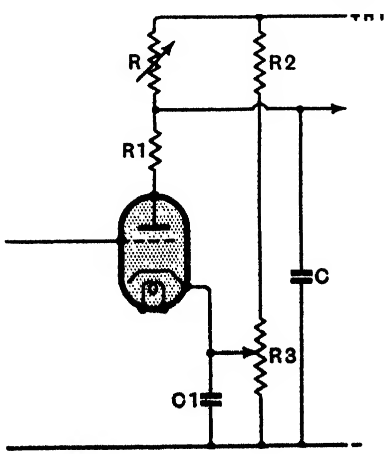



**Fig. 8.5—This diagram shows the conventional oscillator using a gas triode or thyratron**

is raised from zero, the valve remains non-conductive until a certain critical anode potential is reached. The gas then ionizes and the valve conducts and remains conductive until the anode voltage is reduced below another lower critical value of only a few volts.

The upper critical value depends on the grid bias. If the negative bias is increased, then the critical anode voltage is higher and vice versa. Moreover, if the anode voltage is above the lower critical value but below the upper, the valve will become conductive if the grid voltage is made less negative by a suitable amount.

Referring again to Fig. 8.5, the method of operation will now be clear. The capacitance charges through $R$ from the h.t. supply until the voltage across it reaches the upper critical value of the valve. The gas then ionizes and the valve is conductive so that the capacitance rapidly discharges through $R_1$ and the valve. The voltage across the capacitance thus falls rapidly and when it reaches the lower critical value the gas de-ionizes and the valve becomes non-conductive. The cycle then begins to repeat itself.

The resistances $R$ and $R_3$ control the amplitude of the saw-tooth waveform and its repetition frequency. It is generally said that $R$ is the frequency control and $R_3$ the amplitude. Their precise effect, however, depends upon whether the oscillator is synchronized or not.

We are at the moment considering the unsynchronized case, and it is clear that $R$ affects the frequency and has little effect upon the amplitude. An increase in the value of $R$, for instance, means that it takes longer for $C$ to charge up to the striking potential. The frequency is thus lower, but as the striking potential is unchanged the amplitude remains the same.

A variation of $R_3$ changes the grid bias and alters both frequency and amplitude. This is easy to see when the effect of an alteration, say an increase, of bias is considered. At the increased bias the valve requires a higher anode voltage to strike, so that the amplitude of the output is increased. It takes a longer time for the voltage across $C$ to rise to this higher value, however, so that the frequency is lower.

An unsynchronized time base will not run regularly enough for television purposes and synchronizing is always employed. The synchronizing pulses in the transmission are filtered and suitably shaped, as described elsewhere. These pulses are applied to the grid of the gas-triode so that they change the grid potential in a positive direction.

Now consider the state of affairs in a properly running synchronized gas-triode oscillator. The capacitance $C$ charges up through $R$ as before. The voltage does not now rise to the striking value as determined by the fixed bias used, because a short time before this the sync pulse comes along and lowers the grid potential and hence the striking potential. The time at which the discharge commences is thus definitely fixed by the sync pulse, and as the discharge occurs earlier than it would do without the pulse the amplitude is smaller and the frequency higher than with the unsynchronized time base.

Once the sync pulse has initiated the discharge nothing will stop it, and the amount by which the capacitance is discharged and the time occupied by the discharge are substantially independent of the shape or amplitude of the sync pulse. The shape of the commencement of the pulse is important, however.

The action will be clear from Fig. 8.6, where the solid lines represent the saw-tooth waveform in the absence of synchronizing. The application of synchronizing pulses of the form (a) causes the gas triode to strike earlier and the corresponding waveform is shown dotted. It is clear that the amplitude is lower and the frequency higher. In the example shown there are five cycles of synchronized waveform to four of unsynchronized.

A distorted sync pulse of the type illustrated at (b) will give as good results as the rectangular waveform of (a) because the leading edge of the pulse is unchanged. The distorted form (c), however, is bad and may lead to poor synchronizing. The leading edges of the pulses are bent and the grid voltage of the gas triode consequently does not rise sufficiently to strike the valve until a small time interval after the beginning of the pulse.

This in itself would not be important, but if all pulses are not identical in amplitude, the relative times at which the grid voltage reaches the value needed to start the discharge will vary. As a result, the anode voltage of the valve at the striking point will vary and hence the length of the lines.

The effect is particularly bad if the pulse distortion occurs before the pulses are separated from the vision signal, for then the size and shape of the pulse may depend in considerable degree on the

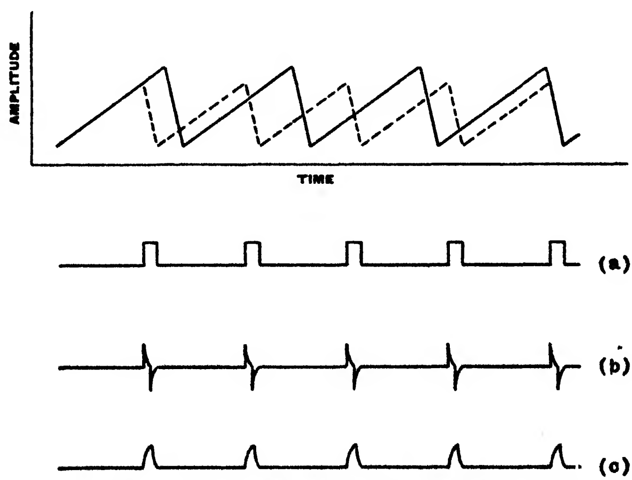



**Fig. 8.6—The top drawing shows in solid lines the waveform of an uncontrolled oscillator and in dotted the waveform when the sync signals are applied to the valve. Undistorted sync pulses are shown at (a) and distorted forms at (b) and (c)**

picture signal immediately preceding it. The effect causes a movement of the lines to the right when there is a white edge on the extreme right of the picture.

Returning to Fig. 8.5, the resistance $R_1$ is included to limit the discharge current to a safe value for the valve. The internal resistance of a gas triode is so low that the current may damage the valve if the capacitance voltage is large. In general, $R_1$ does not exceed 1,000 ohms and is often smaller. It does, of course, increase the fly-back time, but not seriously.

When the valve is synchronized, the resistance $R$ controls the amplitude of the output and $R_3$ has little effect. Too large a change of either, however, will move the natural frequency of the generator so far from the synchronized value, that synchronization will be lost. The amount of latitude in the setting of these controls depends on the amplitude of the sync pulses and increases with the amplitude.

A large amplitude of sync pulse, however, demands much greater perfection of pulse shape and regularity. The conditions, in fact, approximate to those of the distant-controlled generator, with the difference that if the sync pulses fail the generator will still continue running, but at a widely different frequency and amplitude. It is probably best to regard this method of operation as making the time base a distant-controlled generator in which a gas triode is used instead of a hard valve in order to safeguard the c.r. tube in the event of a failure of the sync pulses.

If the sync pulses are of very small amplitude, the settings of the controls will be very critical and small changes in the supply voltages and the temperature of the components will be liable to throw the circuit out of synchronism. The general course, therefore, is to adopt a moderate value of sync pulse, which is large enough to keep the generator in step in spite of all normal changes in voltages and circuit values and is yet not large enough to demand excessive perfection of the sync pulse.

There are many kinds of gas triode available, for they are used for purposes other than television. All are not suitable for television, however, especially the kind in which the gas is mercury-vapour. These valves are subject to large changes with temperature. Various inert gases are used in the television types, notably helium or argon, although sometimes neon is adopted.

The grid circuit of a gas triode is important and the resistance used in it must not be too low, if regular and stable operation is to be secured. A value of the order of 30,000 Ω or more is usually advisable. In the case of the frame generator, a resistance of this order can be inserted directly in series with the grid lead, but the line scan generator is more difficult.

If such a resistance is inserted in series with the grid the sync pulses will be deformed because of the relatively high time-constant of this resistance with the input capacitance of the valve. The difficulty can be overcome by adopting the arrangement shown in Fig. 8.7. This is recommended by the makers of the Mazda T41 thyratron.

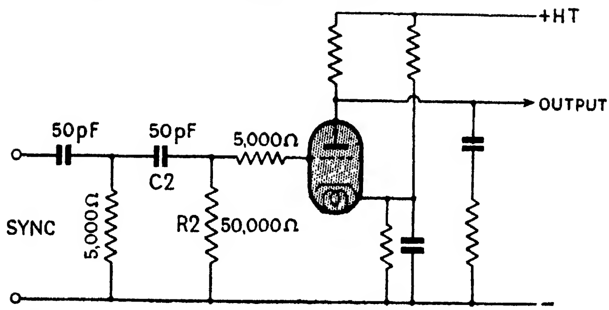



**Fig. 8.7—The oscillator is often fed through a differentiating circuit**

It is not, however, by any means essential to use a gas triode for the saw-tooth oscillator. It is quite common for a hard valve to be employed. In all cases the aim is to devise a circuit which will enable a hard valve to simulate the action of a gas triode.

There are several reasons why many designers prefer hard valves to gas triodes. In the first place the hard valve costs less. In itself, however, there is probably no great advantage in this, for the hard valve needs more associated components and sometimes two hard valves are required to replace one gas triode. In general, the gas triode has a shorter life than the hard valve and so needs more frequent replacement.

In its early days the gas triode acquired a reputation for erratic operation which it probably has not yet outlived. It is true that it fails at high frequencies and is there greatly inferior to the hard valve, but the 10 kc/s needed for the line scan is certainly within the capabilities of a suitable type of gas triode.

It is used in British equipment to some degree but is not as common as the hard-valve type. It seems never to be used in the U.S.A. for television.

The most commonly used hard-valve time base is the blocking oscillator, the basic circuit of which is shown in Fig. 8.8. The circuit looks much the same as that of a reacting detector and it is, in fact, a grid-leak biased oscillator. It has developed from the r.f. sine-wave oscillator but although the circuit is the same its mode of operation is very different.

An r.f. oscillator is under some conditions liable to squegger. It oscillates as required for a short period of time during which grid current charges *C* so much that the valve becomes biased beyond cut-off. Oscillation then ceases while the charge leaks away through *R* and recommences when it has leaked away sufficiently. The output then consists of short bursts of oscillation, at a frequency determined by the inductance and stray capacitance, separated by quiescent intervals during which there is an exponentially varying voltage across *C*. There are usually something like five to twenty cycles of r.f. oscillation in each burst.

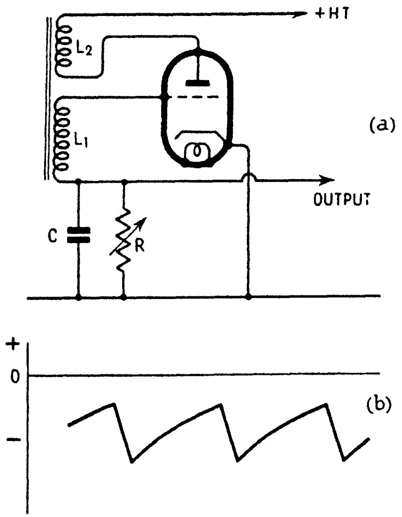



**Fig. 8.8—Basic circuit of the blocking oscillator (a) with the waveform across C (b)**

The blocking oscillator is essentially a squegging oscillator in which the blocking action is completed in one half-cycle of oscillation. In order to secure this action it is necessary that the tuned circuit formed by the coils and the stray capacitance should be of low $Q$. The circuit, however, must be capable of vigorous oscillation when the valve is conductive and so a high $L/C$ ratio is used and very tight coupling between the anode and grid coils. The valve must also have a high mutual conductance, particularly in the positive grid region. The coils are usually wound transformer fashion with a laminated iron core since this enables tight coupling to be secured and the core losses are helpful in obtaining a low $Q$.

The detailed action of the blocking oscillator is very complex. So much is this the case that in spite of its great wartime use in radar no exact analysis has yet been made; the mathematical difficulties are almost insuperable. A fairly detailed, but approximate analysis has been made* but, in practice, design is usually carried out on a trial and error basis. This is actually quite a satisfactory method, for the circuit is so simple, and for television the circuit constants are by no means critical.

In view of the complexity of its detailed operation only a somewhat simplified explanation will be attempted here. It is best to start by assuming that the capacitance $C$ of Fig. 8.8 is initially so charged that the valve is cut off. This means that the grid of the valve is negative with respect to the cathode and, in practice, it can easily be as much as 100 V below cathode. The voltage to which $C$ is charged is actually of the order of one-third of the h.t.

* "Blocking Oscillators," by R. Benjamin, B.Sc. (Eng.). *Journ. Instn elect. Engrs*, Vol. 93, Part IIIA, No. 7, p. 1159.

supply voltage. As the valve is cut off the anode is at the potential of this supply voltage.

The capacitance discharges through $R$ and the grid becomes less negative but nothing else happens until it reaches the cut-off voltage of the valve, which may be $-$ 5 to $-$ 20 V according to the type of valve and the supply voltage. The valve then starts to draw anode current which flows through $L_2$ and sets up a back e.m.f. across it which drops the anode voltage below that of the supply. It also induces a similar e.m.f. in the grid winding $L_1$ in such a direction that the grid potential becomes less negative.

This further increases the anode current and the rising current makes the grid potential change still further in the positive direction. The action is regenerative and the valve is very quickly driven into the positive-grid region with very heavy grid and anode currents. The anode current at its maximum is usually about 1·5–2 times the grid current and the two together can be as much as 1 A. Currents of more than 0·2–0·3 A are neither usual nor necessary in television practice, but higher currents are common in radar, where the circuit is used rather as a pulse than a saw-tooth generator.

Small receiving valves stand up to these heavy currents quite well. Although the peak current is so heavy it flows for such a short time that the mean current is only a few milliamperes.

The turns ratio between $L_2$ and $L_1$ is often 1 : 1, although sometimes a step-down ratio to the grid is used. With a 1 : 1 ratio the e.m.fs across the windings are equal. The anode-cathode voltage is the supply voltage less the back e.m.f. across $L_2$; the grid-cathode voltage is the back e.m.f. across $L_1$ less the voltage across $C$. The grid current flows into $C$ to charge it and the voltage across it approaches the e.m.f. across $L_1$ in magnitude.

The precise voltages depend upon the circuit constants. It is unusual for the grid to become more than 50 V positive to the cathode and in television operation 5–10 V is more usual. The capacitor is charged to 50–100 V so that the back e.m.f. across $L_1$ is of the order of 55–110 V.

This back e.m.f. is produced by the rising anode current and is opposed by the rising grid current. When the rate of increase falls off, the back e.m.f. decreases and the grid potential is reduced. In its turn this reduces the anode current. When the current starts to fall the back e.m.fs reverse and the grid is rapidly driven negative. The action is again regenerative and the valve is cut off.

Because of the transformer and valve capacitances the circuit is oscillatory and the grid of the valve swings more negative than the output terminal while the anode swings above + h.t. If the damping is heavy little more than a half-cycle of oscillation is obtained and the waveforms on grid and anode are like those sketched in Fig. 8.9 while the voltage variations across $C$ have the form shown in Fig. 8.8 (b).

If the damping is less the grid voltage varies in the manner shown in Fig. 8.10. This does not necessarily do any harm, for the

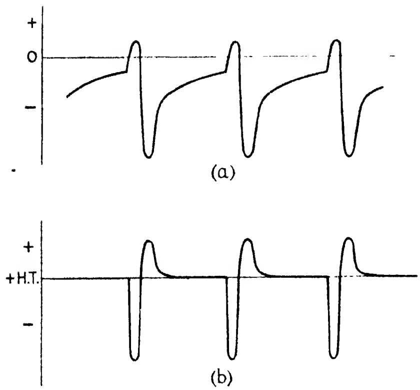



**Fig. 8.9—The grid (a) and anode (b) voltage waveforms of a blocking oscillator are shown here**

capacitor voltage may still be like Fig. 8.8 (b), since the oscillatory voltages are across $L_1$. However, if the damping is so light that the first positive half-cycle swings beyond the current cut-off point the action is profoundly affected.

Regeneration occurs again and in the limit the circuit functions as a self-biased class C sine-wave oscillator and there is no blocking action. An intermediate condition is possible, however. Then the positive cycle initiates another regenerative cycle which leaves the capacitor more fully charged than the first. This occurs in several succeeding cycles and $C$ is charged a little more each time until, finally, the positive swing fails to bring the grid above the cut-off point. This is the squegging mode of operation which is now rarely used.

Squegging is very unlikely to occur when the period of one half-cycle of the natural frequency of the transformer is long enough to permit $C$ to be charged to a voltage which is a considerable fraction of the peak e.m.f. across $L_1$. It is probable when the period is so short that $C$ can charge only to a small fraction of it. In the television case using an iron-core transformer of 500 turns or more for each winding squegging or sine-wave operation are very unlikely to occur. The damping is heavy enough, and the half-period of oscillation is long enough, to prevent anything but the proper blocking mode. If air-core coils resonant at 2 Mc/s or more are used it is unlikely that the true blocking mode will be obtained. Sine-wave operation is probable, but if the damping is properly adjusted squegging can usually be obtained without much difficulty.

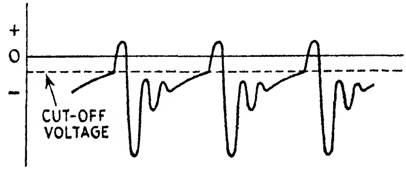



**Fig. 8.10—Blocking oscillator grid voltage wave when the circuit has rather a high Q**

This squegging mode is usable as a time base but it is very rare to find it adopted now. It generally results in a longer fly-back period, because $C$ is charged in a succession of half-cycles and so is being charged for less than one-half of the fly-back period instead of the whole of it. Because of the high frequency of oscillation the circuit radiates badly and needs careful screening.

In the true blocking oscillator, as applied to television, the half-period of oscillation is made rather less than the required fly-back time and so the oscillation frequency is fairly low—of the order of 50–100 kc/s for the line scan. It is quite possible to use the transformer designed for the line scan also for the frame, but this is rarely done because the fly-back is then unnecessarily fast and the peak current into the large capacitance $C$ is very great. The frame-scan transformer usually has thousands of turns, whereas the line-scan component only has hundreds.

As a matter of interest it is possible by suitable transformer design to obtain a fly-back time of about 5 $\mu$sec only and to keep this approximately constant despite very large changes in the values of $C$ and $R$. Saw-tooth repetition frequencies of, say, 10 c/s to 20 kc/s can be obtained by varying $C$ and $R$—the fly-back period being roughly constant.

More or less rectangular pulses can be obtained across resistors inserted in the anode and cathode circuits. Negative-going pulses of a duration equal to the fly-back time can be obtained across a resistor of 500 $\Omega$ inserted between $L_2$ and + h.t. Positive-going pulses can be obtained across a 50-$\Omega$ resistor inserted in the cathode lead of the valve. They can have an amplitude of up to about one-fifth of the h.t. voltage and appear at very low impedance. Pulses as short as 1 $\mu$sec can be obtained by special design of the transformer, but so far have no television application.

When the transformer and $C$ are constant and the repetition frequency is varied by changing $R$, or the voltage to which $R$ is returned, the mean anode current is proportional to the repetition frequency. This is obvious because each current pulse has the same duration and amplitude and there is one for each cycle of operation. An anode-current meter can thus be calibrated in frequency. The important factor from the television point of view, however, is that at a high repetition frequency it is necessary to check that the mean current ratings of the valve are not exceeded. Thus, if a blocking oscillator draws 2 mA at 1,000 c/s it will draw 20 mA at 10,000 c/s, the transformer being the same in both cases.

The blocking oscillator is not difficult to synchronize. Pulses with sharp leading edges are needed, as in most other cases, and must be positive-going if applied to the grid. A negative-going pulse can be applied to the anode, however, for it is transferred to the grid as a positive-going pulse by the transformer. A positive-going pulse is sometimes fed to the grid by placing a small resistance in series with $C$, but more usually the pulse is fed directly to grid or anode, according to its polarity.

A pulse amplitude of the order of 10 V is usually desirable and it is usually necessary to employ a buffer valve between the sync separator and the blocking oscillator. This is to prevent the large voltages developed in the stage from reacting back on the separator circuits. This is especially important in the case of the line-scan circuits since if the line-frequency voltages get back to the sync separator they will be passed to the frame saw-tooth generator and prevent proper interlacing.

Low-capacitance diodes are useful as buffers but, as the capacitance of even the best is not negligible, screened tetrodes and pentodes are better. Strictly, the need for buffers is not confined to blocking oscillators for all saw-tooth generators are liable to react back on to the synchronizing circuits. However, blocking oscillators do generally produce the most "backwash" and so make buffers more necessary.

The linearity of the saw-tooth produced by the circuit of Fig. 8.8 is very poor because the peak-to-peak amplitude of the saw-tooth is nearly equal to the total voltage acting on discharge; that is, the capacitor discharges nearly completely each cycle. A very great improvement is obtained by returning the discharge resistor $R$ to + h.t. instead of − h.t. The voltage in circuit is then increased by that of the h.t. supply and the linearity is correspondingly improved.

As an example, if the h.t. supply is 300 V, the cut-off voltage of the valve is − 20 V and $C$ charges to 100 V, then in the circuit of Fig. 8.8 when the generator is running unsynchronized the saw-tooth amplitude is about 80 V. The voltage acting on discharge is only 100 V so that, referring to Fig. 8.1, $e/E$ is 0·8 and $t/T$ is 1·6. The curvature of the scan stroke is very marked.

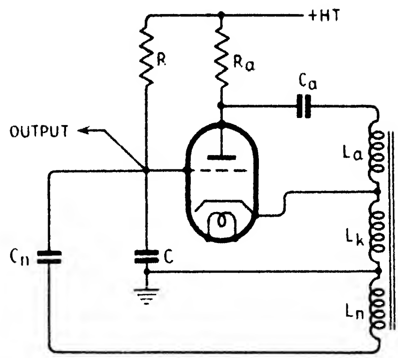
Now if $R$ is returned to + h.t. the voltage acting is 400 V. The amplitude is the same so that $e/E$ is only 0·2 and $t/T$ is 0·19. The discharge is much more linear and may or may not be adequate for television, depending on the characteristics of the subsequent amplifier.

For the same discharge period the $CR$ product must be greater when $R$ is taken to + h.t. than when it is connected to − h.t. In the above example it must be eight times as great.

If the generator is synchronized the saw-tooth amplitude is less because the sync pulse necessarily triggers the generator before $C$ has discharged to the cut-off bias of the valve. In the particular case quoted above one might well use a 30-V sync pulse arranged to carry the valve well above cut-off, say to zero grid volts. It should then occur when $C$ has discharged to − 30 V instead of − 20 V and so the amplitude is reduced to 70 V.

This simplest form of blocking oscillator is, in the author's experience, the most satisfactory of all for frame frequency because it is much easier to secure good interlacing with it than with other arrangements. It is not always satisfactory at line frequency, however, because then a certain proportion of the oscillatory voltage

**Fig. 8.11—Cathode-coupled neutralized blocking-oscillator circuit**




across the inductances appears also across $C$. If the circuit is redrawn to include the valve capacitances and transformer interwinding capacitance it will be found to be that of a Hartley oscillator in which $C$ completes the tuned circuit. The oscillatory currents in the tuned circuit necessarily flow through $C$ and their effect on the saw-tooth is negligible only if $C$ is very large compared with the other capacitances. In practice, the effect begins to be noticeable when $C$ is less than 0·01 $\mu$F, and is serious when it is under 0·001 $\mu$F.

It is possible to overcome it by modifying the arrangement of parts and using an extra winding on the transformer. This is shown in Fig. 8.11. The anode coil is shunt-fed by $R_a$ and $C_a$, but this is not essential; a separate anode winding with series feed can be used. The three parts of the coil $L_a$, $L_k$ and $L_n$ have equal turns. Ignoring $L_n$ and $C_n$ for the moment the circuit is a shunt-fed cathode-coupled blocking oscillator. The charging capacitor $C$ is in the grid circuit and appears to be well isolated from the oscillatory circuit. It is, however, coupled to it by the grid-anode valve capacitance $c_{ga}$ and the grid-cathode capacitance $c_{gk}$. The performance is, in fact, little different from that of the more conventional circuit. However, it is possible to neutralize this coupling by adjusting $C_n$ properly. When all windings are alike and are very tightly coupled the optimum condition is $C_n = c_{gk} + 2\,c_{ga}$. The circuit can be satisfactory with values of $C$ as small as 100 pF, but the adjustment of $C_n$ is critical.

In television practice the saw-tooth voltage across $C$ is not always used for scanning. It is necessary to produce this saw-tooth, for it is an essential part of the operation of the blocking oscillator, but it is not necessary to use it. The blocking oscillator is very commonly used primarily as a pulse generator to discharge an entirely separate $RC$ circuit across which the useful saw-tooth is developed.

The simplest circuit of this nature is shown in Fig. 8.12. The grid capacitor $C_1$ of the blocking oscillator is here shown next to the grid with the grid leak $R_1$ returned to cathode. The arrangement of Fig. 8.8 can be used equally well, however, and the grid

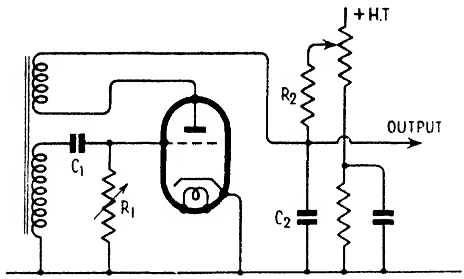



**Fig. 8.12—In this blocking-oscillator circuit the valve is used to discharge the capacitor $C_2$**

leak may be returned to either + or − h.t. These are all minor variations of the circuit.

The saw-tooth which is used is now developed across $C_2$. When the valve is cut off $C_2$ charges from the h.t. supply through $R_2$. When the valve conducts it draws its anode current from $C_2$ and so discharges it. The action of the blocking oscillator itself is affected by this only in so far as it no longer has a fixed h.t. supply but one which drops rapidly during the conduction period.

It is important to have the time-constants so proportioned that $C_2$ does not charge fully before $C_1$ has discharged to the cut-off voltage of the valve, but this is automatically secured if conditions for a moderately linear wave across $C_2$ are realized.

A more common arrangement is to use a tetrode or pentode with the control and screen grids acting as the grid and anode of the blocking oscillator and the anode as a discharge electrode for the time-base capacitor. One form of this circuit is shown in Fig. 8.13. The screen grid, which acts as the blocking-oscillator anode, is fed from a potential divider $R_3$, $R_4$ across the h.t. supply. The charging circuit is $R_2$, $C_2$. In this case $R_2$ is shown as variable to control the saw-tooth amplitude. As the resistance is reduced $C_2$ charges more quickly and so charges to a higher voltage in a given time. In an actual receiver, such a control is usually labelled Picture Height or Picture Width.

An alternative form of control which gives much the same results

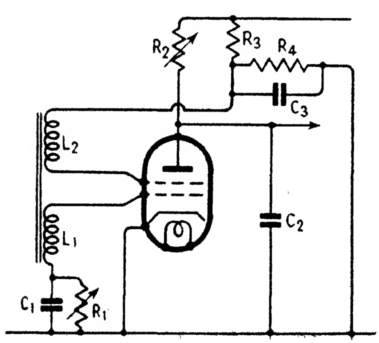
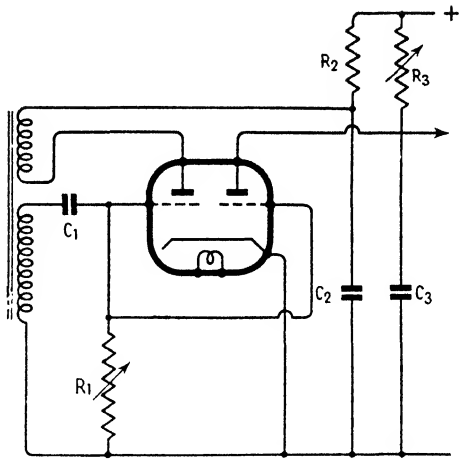



**Fig. 8.13—Tetrode blocking oscillator in which the anode-cathode path of the valve discharges $C_2$ while the grid and screen form the oscillator electrodes**

**Fig. 8.14—In this version of the blocking oscillator one valve is the oscillator and the other acts to discharge C₃**




is a fixed resistor fed from a variable voltage. This is shown in Fig. 8.12 where the fixed charging resistor $R_2$ is fed from the potential divider.

The two forms of control are usually interchangeable. The variable resistor has the advantage of giving a slightly more linear saw-tooth because the full voltage is always used. However, the potential-divider method lends itself particularly well to the grouping of the controls at a convenient operating point, since its three leads only carry direct current and can be made any desired length.

Another form of blocking-oscillator generator is shown in Fig. 8.14. It is one which is rarely found in this country, probably because double-triodes have only recently appeared here, but is very common in the U.S.A. One half of the valve acts as a blocking oscillator in the manner already described, the other discharges $C_3$ when it becomes conductive. In this case $R_2$ and $C_2$ are made large so that the voltage across $C_2$ does not change appreciably; $C_3$ charges through $R_3$ when the valves are cut off. When the blocking oscillator conducts so does the other half of the valve, since the two grids are tied together.

Sometimes the grid and anode of a tetrode are used as the blocking-oscillator electrodes, the sync pulses being fed to the screen grid. Best operation is then usually obtained with a very low screen voltage.

There are a number of other circuits related to the blocking oscillator. These circuits are really varieties of the blocking or squegging oscillators and are not often found now. One is given in Fig. 8.15 and can operate in two different ways. In the first mode $RC$ is much larger than $R_1C_1$ so that the voltage across $C$ remains substantially constant. The operation is then much the same as in the ordinary blocking oscillator but $C_1$ is charged by both anode and grid currents. The saw-tooth appears at the

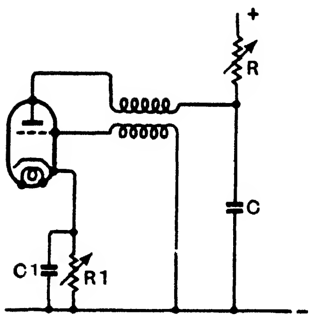



**Fig. 8.15—Cathode charging form of blocking oscillator**

cathode. It is negative-going and is very non-linear unless $R_1$ can be returned to a point highly negative to − h.t. The resistance $R_1$ must be fairly low, under 10 k$\Omega$, if the circuit is to function.

The second mode of operation occurs when $R_1C_1$ is much larger than $RC$. The cathode voltage then remains substantially constant and the saw-tooth is taken from $C$. The valve is non-conductive and $C$ charges through $R$ thus raising the anode voltage until the valve starts to draw current. Regeneration occurs and $C$ is discharged. Since anode control of a valve is much less effective than grid control the circuit is much less precise in its action than the ordinary blocking oscillator.

Fig. 8.16 shows a circuit which is sometimes useful in a time base for electric deflection where an h.t. supply of 1,000 V or so is usually available. It has the disadvantage of needing a separate heater supply since the cathode is at high voltage to earth. It is quite an old circuit now* but is really a modification of the circuit of Fig. 8.15 in its cathode-output mode of operation. As $C$ charges through $R$ the cathode potential is carried down and eventually approaches the grid potential which is set by $R_1$. The valve conducts, regeneration occurs, and $C$ is discharged. At the end the valve is left with a very low anode-cathode voltage and a highly negative grid voltage.

One characteristic of the circuit is worth mentioning. It is very

* "New Time-Base Circuit," by Desmond MacCarthy. *Wireless World*, October 4, 1935.

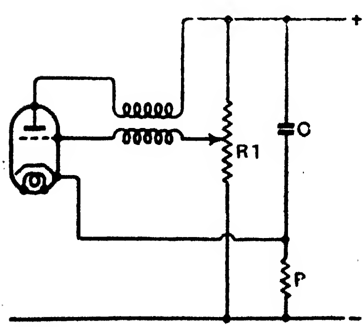



**Fig. 8.16—This diagram shows a modification of the circuit of Fig. 8.15**

largely free from the coupling between the oscillatory and the charging circuits which results in an oscillatory voltage appearing across *C* at high frequencies. Actually, *C* is coupled to the coils through the anode-cathode and grid-cathode capacitances but as the anode and grid undergo oscillatory voltage variations in opposite phase the two couplings tend to cancel each other. If the ratio of the two voltages is made, by adjusting the transformer turns ratio, the inverse of the capacitance ratio there is theoretically no interaction—the circuit is, in fact, self-neutralized. As the two capacitances are in practice nearly equal a 1 : 1 ratio in the transformer is usually satisfactory.

Blocking-oscillator transformers differ considerably in design and sometimes ratios as high as 4 : 1 are used. In the author's experience, however, they are far from critical and a great deal of latitude is permissible as far as television is concerned. For the frame circuit two windings each of 3,000 turns of No. 40 or 41 gauge enamelled wire are suitable, the core being a 1-in stack of Magnetic and Electrical Alloys, Silcor No. III, Lamination No. 74, 0·014 in thick. For the line scan, windings of 500 turns of No. 34 enamelled wire are suitable with a ¼-in stack of the same laminations.

For the valve a triode of 10 kΩ nominal a.c. resistance with a mutual conductance of 2–3 mA/V is suitable.

So far all the hard-valve saw-tooth generators discussed have been of the blocking- or squegging-oscillator type but it is by no means essential to use coils in a hard-valve time-base generator. A valve is needed for the charge and discharge of the capacitance, and this valve must be provided with adequate positive feedback to make its action continuously repetitive. This feedback can be obtained with coils in the manner already described or by means of resistance coupling between the anode and grid circuits. In this latter case, however, it is usually necessary to interpose a valve in the feedback chain to secure the necessary reversal of phase.

One of the best of these circuits has been fully described by Puckle,* and it is shown in Fig. 8.17. *C* and *R* form the charging

* *Journal of the Television Society*, June 1936.

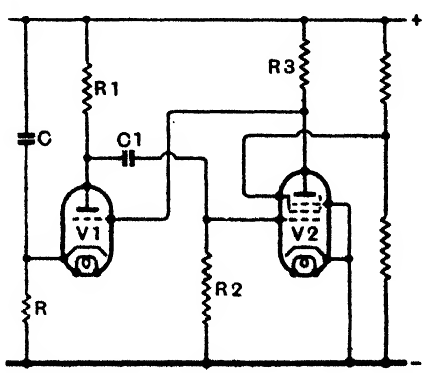



**Fig. 8.17—A hard-valve time base using two valves is shown here**

circuit with $V_1$ for the discharge valve; $V_2$ is the phase-reversing valve. Consider the operation when $C$ is uncharged. $V_2$ is passing anode current, for its screen is maintained at a fixed positive potential by the potentiometer, the anode is returned to positive h.t. through $R_3$ and the grid bias is about zero or very slightly negative. The grid of $V_1$ is at the same potential as the anode of $V_2$ for it is connected directly to it.

At the instant we are considering $C$ is uncharged, so that there is no voltage across it. Consequently, anode and cathode of $V_1$ are at the same potential, which is that of the h.t. line. The grid is considerably negative with respect to cathode, however, because it is joined to the anode of $V_2$ and this point is considerably negative with respect to positive h.t. $V_1$ is thus non-conductive.

Now $C$ charges through $R$ and the voltage across it rises, making the cathode negative with respect to the anode and also making the cathode approach the grid in potential. As far as $V_1$ is concerned this is the same as the anode going positive and the grid becoming less negative.

When the potential across $C$ is large enough $V_1$ starts to pass anode current. This current sets up a voltage drop across $R_1$ and the anode becomes slightly negative with respect to positive h.t. By means of $C_1R_2$ this change of anode potential is communicated to the grid of $V_2$, making it more negative. The anode current of $V_2$ decreases and the voltage drop across $R_3$ consequently falls; that is, the anode voltage of $V_2$ rises and carries with it the grid potential of $V_1$.

The rise in grid potential of $V_1$ naturally increases the anode current and hence the voltage drop across $R_1$, and the grid is driven still more positive. The action is cumulative and as soon as $V_1$ starts to draw current its grid is driven violently in a positive direction and the capacitance $C$ is rapidly discharged.

At length $C$ can no longer supply the current and the voltage across $R_1$ starts to fall; the grid potential of $V_2$ consequently becomes less negative, the anode current rises, and the grid of $V_1$ is driven negative. This further reduces the anode current of $V_1$, and the valve is rapidly driven back into the non-conducting condition. The capacitance $C$ then starts to charge again and the whole cycle repeats itself indefinitely.

The resistance $R_1$ is in the discharge path and so should be kept of low value—1,000–2,000 $\Omega$. The circuit is a very reliable one and is easy to get working. In common with most hard-valve generators it works better with a large output than with a small, so it is usually employed with a pentode valve instead of the resistance $R$. As will be explained later, this permits a linear saw-tooth output to be obtained which approaches the h.t. voltage in amplitude.

This arrangement is not often used in television equipment on account of the cost of the large number of valves. Moreover, the circuit of Fig. 8.17 suffers from the disadvantage that the cathode of $V_1$ is fluctuating at saw-tooth potential. If the heater were connected to negative h.t., as usual, nearly the full h.t. voltage would appear between heater and cathode and the insulation would almost certainly be inadequate.

In practice, it is necessary to connect heater to cathode and to provide a separate heater winding on the mains transformer for this valve. This winding and the wiring to it will fluctuate in potential with respect to earth by the full saw-tooth output. The winding must consequently have a low capacitance to earth and the wiring must be carried out with care.

The mains transformer and wiring capacitances affect the saw-tooth frequency, since they are substantially in parallel with $C$. At line frequency they must be small if unavoidable changes are not to upset the generated frequency seriously.

With care in design no serious difficulty will be encountered from this cause, but the necessity for separate heater windings for the discharge valves is undoubtedly a drawback.

Another similar circuit which is free from this defect and for which a double-triode can be used instead of two separate valves is shown in Fig. 8.18. It is a form of cathode-coupled multivibrator.

Here $V_2$ is the discharge valve. When $C$ is uncharged $V_2$ passes no anode current, for its grid is biased negatively by the voltage drop across $R_3$ set up by the anode current of $V_1$ and also by a charge accumulated on $C_1$. The valve $V_1$ passes current because it has a higher anode potential.

The capacitance $C$ charges and $V_2$ at length passes current which flows through $R_3$ increasing the voltage drop across it and so changing the grid potential of $V_1$ negatively. The anode potential of $V_1$ rises and the change of anode potential is communicated to the grid of $V_2$ as a positive pulse through $C_1R_2$. This accelerates the discharge of $C$. If the positive pulse is large enough, grid current flows in $V_2$ and charges $C_1$.

When $C$ can no longer supply the anode current of $V_2$, the current falls, reducing the current in $R_3$ and letting the grid potential of $V_1$

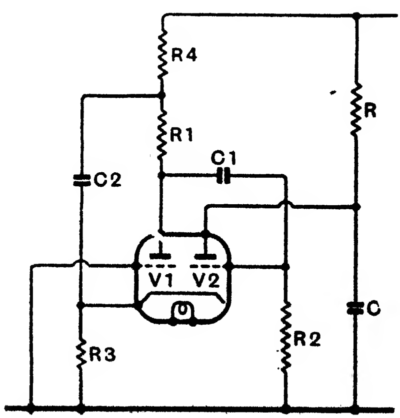



**Fig. 8.18—In this form of oscillator it is possible to use a double-triode**

return towards its normal value. The anode potential of $V_1$ thus falls, and the grid potential of $V_2$ consequently returns to its normal value and anode current ceases. The cycle then recommences with the charging of $C$.

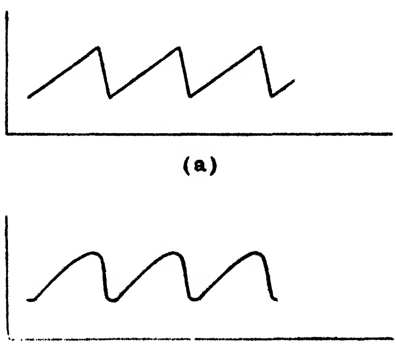
The mean potential of the grid of $V_2$ is negative with respect to the negative h.t. line, because of the charge accumulated on the grid during the positive pulses applied to it.

This circuit works very well indeed provided that too small an output is not required. The minimum output for reliable operation appears to be about 40 V p–p, so that the h.t. supply should be of the order of 700 V if a reasonably linear waveform is needed. The h.t. voltage can, of course, be lower if the exponential wave then obtained is corrected at a later stage.

All generators appear to have a minimum output for satisfactory operation and it is usually of the order of 20–40 V p–p. More critical choice of circuit values is needed for small output than for large if a good waveform is to be secured. What usually happens as the output is reduced is that the effective fly-back time increases until it is eventually of the same order as the scan time. Instead of abrupt transitions from scan to fly-back and vice versa, the change becomes gradual. The waveform, in fact, becomes rather like that of Fig. 8.19 (b) instead of as at (a).

It is usually easiest to secure a good waveform at small output with a gas triode, but even then it is usual to generate about 30 V p–p. No difficulty normally arises when electric deflection is used, because the amplifier following the generator needs an h.t. supply of the order of 1,000 V and this is also used for the generator itself. Consequently, even with resistance charging, about 50–70 V output can be obtained with good linearity and any of the generators described can be made to give good results.

Because the h.t. supply is only about 200–300 V when electromagnetic deflection is used, it is hardly possible to obtain a very linear saw-tooth across the capacitor with any of the foregoing circuits. It is, therefore, usually necessary to obtain some form of correction in the following circuits. This may be done purely by



(a)

(b)

**Fig. 8.19—With incorrect operating conditions the good waveform (a) degenerates to the rounded shape (b)**

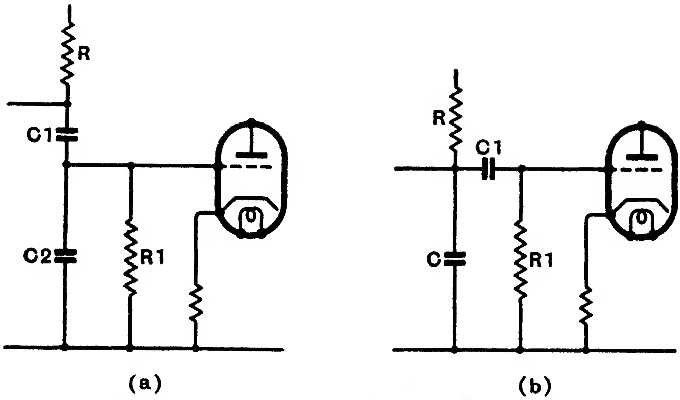



**Fig. 8.20—When an exponential waveform is wanted a large voltage can be generated and only a portion of it applied to the amplifier (a) ; alternatively the grid leak $R_1$ can be made of low value (b)**

circuitry or by utilizing the inverse curvature of a valve characteristic or by a combination of both. In addition, correction for deficiencies in the amplifier itself is often needed.

It is usually desirable to make the saw-tooth generated as linear as possible in order to simplify the correction arrangements. It is important to remember, therefore, that the coupling to the following valve can itself considerably increase the unwanted curvature of the saw-tooth.

Two common arrangements are shown in Fig. 8.20. The first (a) is normally employed only when it is necessary to generate a saw-tooth of greater amplitude than the following valve can handle. Ignoring $R_1$ for the moment, $C_1$ and $C_2$ in series form the charging capacitance but only the fraction $C_1/(C_1+C_2)$ of the total voltage is applied to the following valve. The presence of $R_1$ increases the curvature. This is easily seen in the limiting case when $C_1$ becomes infinitely large, for the circuit then behaves as if $R_1$ were absent and $R$ had a value $RR_1/(R+R_1)$ and $C_2$ charged from a voltage only $R_1/(R+R_1)$ of the actual voltage.

Similarly in Fig. 8.20 (b) if $C_1$ is very large the presence of $R_1$ increases the curvature of the saw-tooth in the same way. If $C_1$ is not very large then the time-constant $C_1R_1$ itself introduces considerable distortion of the wave.

The resistance $R_1$ should always be as large as possible if the distortion is to be kept low. When the following valve is an output valve, however, $R_1$ may be limited to 0·5 MΩ. It almost certainly will be for the line scan, but for the frame a value up to 2 MΩ is sometimes permissible. For low distortion, therefore, $C_1$ of Fig. 8.20 (b) will not be less than 0·01 μF for the line scan and 0·5 μF for the frame. (See Appendix 3.)

On rare occasions it may happen that the generated saw-tooth is too linear. The curvature is then readily increased by reducing $R_1$.

Although distortion compensation by curvature of the valve characteristics can be extremely useful, in practice it is wise to rely on it as little as possible, for it renders the equipment liable to go out of adjustment with the ageing of the valves. There is consequently a tendency, and a good one, to make the individual stages as linear as possible.

The amplifier can be made more linear by the use of negative feedback and the control of the feedback affords a convenient adjustment for the picture size. The greater input demanded by the amplifier, however, makes the output of the saw-tooth oscillator still more curved in form. Further curvature in the same sense is usually introduced in the coupling to the deflector coil.

As will be shown later, it is possible by using an extra valve to generate an extremely linear saw-tooth. The distortion of the deflector-coil coupling still remains, however, but it is possible to remove this by overall negative feedback in the amplifier. For such feedback to be effective it is necessary for the saw-tooth wave at the input of the amplifier to be highly linear. This method involves separate and independent linearization of the generated saw-tooth and of the amplifier. More often a common linearizing circuit is used.

One such circuit is shown in Fig. 8.21, in the form in which it was used in one Murphy Radio receiver. Although a gas triode is used it can be employed with other saw-tooth oscillators. Bias for the valve is obtained in a conventional manner from $R_5$ and $R_6$ with $C_4$ acting as the cathode by-pass capacitance.

The resistance $R_1$ is of low value and acts to safeguard the valve by limiting the discharge current. $R_2$ and $R_3$ form the charging resistance. The charging capacitance is split into two parts, $C_1$ and $C_2$, and across $C_2$ is developed the usual saw-tooth voltage of exponential form, while that across $C_1$ is nearly the same.

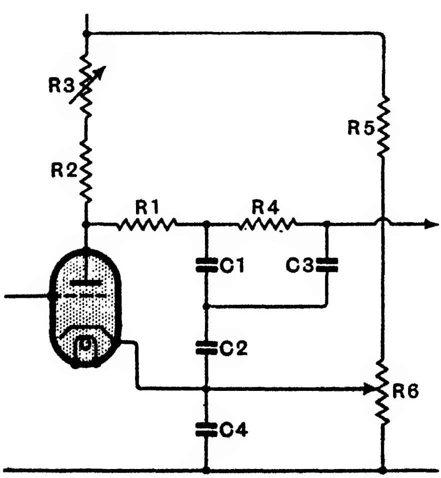



**Fig. 8.21—This circuit includes correction for the exponential waveform of the oscillator**

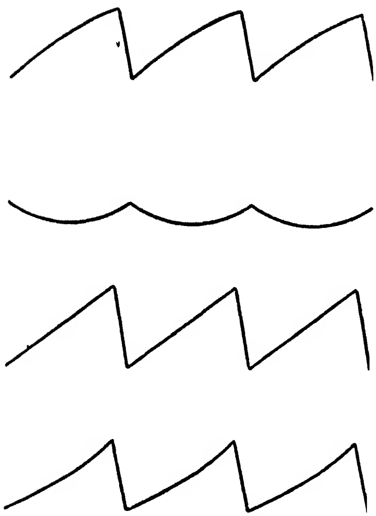

(a)
(b)
(c)
(d)

**Fig. 8.22—The exponential wave is shown at (a) and the correcting wave across $C_3$ of Fig. 7.18 at (b). Combined, these give a linear wave (c). A case of over-correction is shown at (d)**

A subsidiary charging circuit $R_4$ $C_3$ is provided, however, and connected in shunt with $C_1$. The output is the sum of the voltages across $C_2$ and $C_3$ and can be nearly linear.

Consider what happens, starting when the capacitances are partly charged and the valve strikes. $C_1$ and $C_2$ rapidly discharge through the low resistance path afforded by $R_1$ and the valve, and the voltage across them falls very rapidly to its minimum. $C_3$ also starts to discharge, but through the high resistance $R_4$. At the end of the fly-back, when the voltages across $C_1$ and $C_2$ are very small, there is still a relatively large voltage across $C_3$.

The valve now becomes non-conductive in the usual way and $C_1$ and $C_2$ begin to charge through $R_2$ and $R_3$. The voltage across $C_1$, however, is less than that across $C_3$, so that $C_3$ still continues to discharge, its discharge current helping to charge $C_1$. During the first part of the scan, therefore, the voltages across $C_1$ and $C_2$ are rising and that across $C_3$ is falling. The total voltage across $C_2$ and $C_3$ is thus rising less rapidly than that across $C_1$ and $C_2$.

At length the voltages across $C_1$ and $C_3$ become equal and the discharge of $C_3$ ceases. After this the voltage across $C_1$ predominates and $C_3$ starts to charge. Now the output is the sum of the voltages across $C_2$ and $C_3$ and can be much more nearly linear than either of these voltages alone.

The waveforms in the different parts of the circuit are sketched in Fig. 8.22. At (a) is shown the voltage across $C_2$ in its usual exponential form, and at (b) the voltage across $C_3$. The output voltage, which is the sum of these voltages, is shown at (c). Only the general form is shown here and the drawings are not to scale. By using incorrect values over-compensation is easy and a waveform such as (d), or even one with a double-bend on the scan, can be

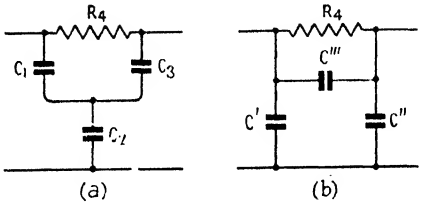



**Fig. 8.23. Alternative networks of identical performance; (a) is the star form and (b) the delta**

secured. The " over-corrected " waveform (d) is actually very useful for it can compensate for distortion in the deflector-coil coupling.

Circuit values must usually be found experimentally, but a good basis for a start with the line time base is given by $R_4 = 0{\cdot}25\ \text{M}\Omega$, $C_1 = 0{\cdot}0015\ \mu\text{F}$, $C_2 = 0{\cdot}05\ \mu\text{F}$, and $C_3 = 0{\cdot}002\ \mu\text{F}$.

The charging capacitance and correction network of Fig. 8.21 are re-drawn in Fig. 8.23 (a) with the same reference letters for the components. The circuit (b) is sometimes a more convenient alternative and is electrically identical. According to the star-delta theorem the star of capacitors in (a) is identical with the delta arrangement of (b). For equivalence

$$C' = \frac{C_1C_2}{C_1+C_2+C_3};\ C'' = \frac{C_2C_3}{C_1+C_2+C_3};C''' = \frac{C_1C_3}{C_1+C_2+C_3}$$

Still another arrangement which has similar, but not identical, properties is shown in Fig. 8.24. Here $C_1$ is the charging capacitance and $R_1$, $R_2$ and $C_2$ form the correcting network.

All these circuits can be regarded as ones which integrate a saw-tooth wave and then add the integrated saw-tooth to the original saw-tooth in suitable proportions. The integrated saw-tooth is parabolic in form, like the wave of Fig. 8.22 (b). The integration is not, of course, perfect and the output wave is never perfectly linear. These correction circuits are, however, extremely useful in practice.

Another method of correction which is very widely used for the frame scan involves negative feedback from the anode of the frame output stage. Because it involves the amplifier characteristics it is essential to be familiar with them before it can be understood. They are treated in Chapter 7.

The arrangement is shown in Fig. 8.25 and $R$, $C_1$ and $C_2$ form the charging circuit, the discharge valve, which is usually a blocking oscillator, being connected to A. Feedback occurs from the anode

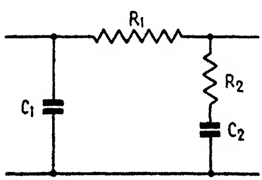
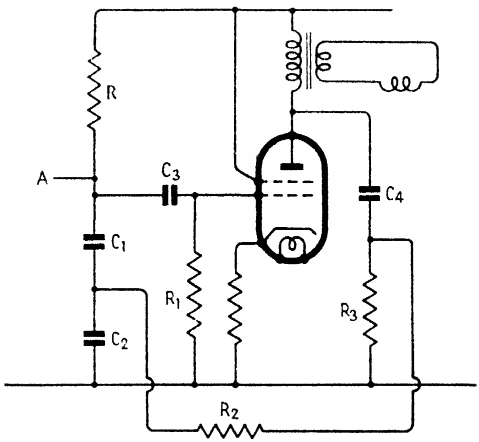



**Fig. 8.24—Another form of correcting network is shown here**

**Fig. 8.25—Commonly-used correction circuit in a frame time base**




of the valve through the network $C_4$, $R_3$ and $R_2$; typical values are:—$C_4$=0·1 $\mu$F, $R_3$=100 k$\Omega$, $R_2$=10 k$\Omega$, $C_1$=0·05 $\mu$F and $C_2$=0·02 $\mu$F. $R_2$ is usually adjustable over a range of perhaps 5 k$\Omega$ to 20 k$\Omega$ as a linearity control.

For a linear deflection-coil current, and allowing for transformer distortion, the voltage wave on the anode must be a combination of a negative-going linear saw-tooth and a negative-going square-law wave, while on fly-back there is a large positive-going pulse. The anode-current and, hence, the grid-voltage waves are of similar form but are positive-going and without the fly-back pulse.

The combination $C_4$ $R_3$ has the form of a differentiating circuit so that the scan-voltage wave across $R_3$, so far from changing negatively with increasing rapidity, as does the anode voltage, does so with decreasing rapidity. The voltage fed back through $R_2$ opposes the charging of $C_1$, $C_2$ but does so most at the start of the scan and least at the end. In effect, negative feedback is secured but the amount of feedback progressively decreases from the start to the finish of the scan.

Instead of using *RC* networks and/or feedback circuits to achieve overall linearity, it is possible but often less economical to linearize each stage. The simplest and best method of linearizing a saw-tooth generator is with the aid of an extra valve, although in one case at least this function can also be performed by the discharge valve.

As already explained, when a capacitance is charged through a resistance the voltage across it rises exponentially because the current is not constant. The current is equal to the voltage across the resistance divided by the value of the resistance, and the voltage is equal to the supply voltage minus the voltage across the capacitance. This last increases with time as the capacitance charges and the voltage across the resistance, and hence the current, fall.

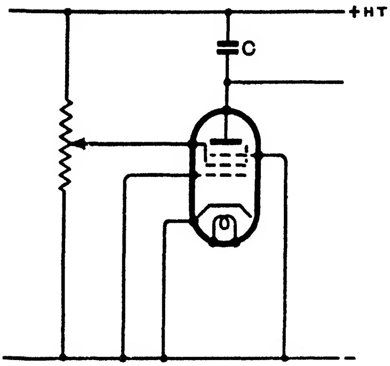



Fig. 8.26—This diagram illustrates the use of a valve instead of a charging resistance

If the capacitance voltage is to rise linearly with time the charging current must be constant. With a constant supply voltage, therefore, the resistance must not be constant but a function of current. It must pass the same current irrespective, within wide limits, of the voltage applied to it. Such a constant current device has an infinite a.c. resistance.

One of the earliest methods of obtaining such a characteristic was to use a saturated diode for the charging resistance, the current being controlled by varying the filament current. The saturated diode is not used for television purposes, however, because it is necessary to use a valve with a directly-heated cathode run from direct current. Moreover, modern cathodes are unsuitable.

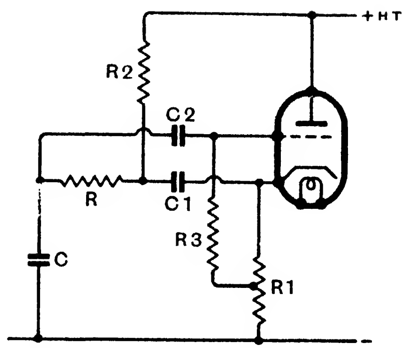
An alternative to the diode is a tetrode or pentode. With such valves the anode current is nearly independent of the anode voltage, provided that it exceeds a certain value. With the early screen-grid tetrodes this value is some 10–20 V more than the screen voltage, say, 70–100 V. With modern tetrodes and pentodes it is of the order of 30–50 V.

Provided the anode voltage is not allowed to fall below this figure, the anode current is nearly independent of the voltage, but can be varied by changing either the control grid or the screen-grid voltage. In practice, the valve is usually operated with zero control grid voltage and the screen voltage is varied to control the charging current

A typical arrangement is shown in Fig. 8.25, and is extremely simple. The screen-feed potentiometer controls the charging current and is hence the frequency control of the generator, taking the place of the variable resistance shown in the earlier circuits. The discharge valve is connected across the capacitance.

It will be noted that owing to the need for a fixed screen voltage, when once it has been adjusted to the right value, it is necessary for the valve to have its cathode at a fixed potential. Consequently, it cannot be conveniently used with generators in which the discharge valve also has a fixed cathode potential.

**Fig. 8.27—Cathode-follower arranged to linearize the voltage across the capacitor C**




It can be used readily to replace $R$ of Fig. 8.17, but with other generators the circuits must normally be modified so that the discharge valve has its cathode fluctuating at saw-tooth frequency.

Partly because of this drawback and partly because of the cost of the extra valve, the constant-current charging system is rarely used in modern television receivers. It is, however, very good and enables a linear output of about 50–100 V less than the h.t. supply to be obtained directly from the generator.

It is possible, however, to use negative feedback for linearizing the saw-tooth and one very satisfactory scheme is shown in Fig. 8.27. The valve is used as a cathode-follower,* so that the change of cathode voltage is nearly equal to the voltage change across $C$. This is the time-base capacitance and the discharge circuit is connected across it; this is not shown in Fig. 8.27 since we are now concerned only with the linearizing action.

To a first approximation the voltage across $C_1$ can be considered as constant. Then $C$ charges through $R$ from the voltage across $C_1$. The changing voltage is communicated to the grid through $C_2R_3$, the only purpose of which is to enable the correct mean grid potential of the valve to be obtained. As the grid potential changes positively the anode current rises and hence the cathode potential.

The cathode voltage is in series with the voltage across $C_1$ as far as the charging circuit is concerned. Consequently as $C_1$ charges the effective charging voltage increases. The voltage at the cathode is nearly equal to that across $C$, consequently there is nearly constant voltage across $R$ and the charging current is nearly constant. Hence the voltage across both $C$ and $R_1$ is nearly linear.

When the parallel value of $R_1$ and $R_2$ is equal to the a.c. resistance of the valve, the circuit behaves as though it were a simple $RC$ charging circuit with an applied voltage equal to one-half of the h.t. voltage plus $(\mu + 2)/2$ times the voltage across $C_1$. As $\mu$ may be 20 and the voltage across $C_1$ may be 150 V, the total effective voltage may be 1,600–1,700 V for an h.t. supply of only 300 V.

* The circuit is sometimes known as the "bootstrap."

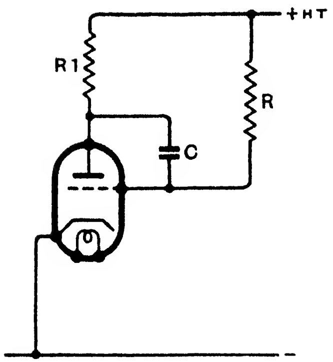



**Fig. 8.28—Basic circuit of the Miller integrator**

Another method of obtaining a linear saw-tooth is by means of a Miller integrator and this has the great advantage that it can be combined with a transitron-type oscillator to give a single-valve linear saw-tooth oscillator. This is one of the war-time radar developments which has great television possibilities.

The basic circuit of the integrator is shown in Fig. 8.28 and $R_1$ is usually very small compared with $R$. Consider the conditions with $C$ charged to the full h.t. voltage and the valve drawing current. Because the valve is drawing current, the anode potential is negative with respect to + h.t. Because $C$ is charged to the full h.t. voltage, the grid is negative with respect to the anode by this full h.t. voltage. Therefore, the grid is negative with respect to cathode by the amount of the voltage drop across $R_1$. The valve is working with a negative grid and grid current does not flow in spite of the grid return $R$ being taken to + h.t.

The capacitance $C$ discharges through $R$ and $R_1$ and the voltage across it falls. The grid-cathode potential becomes less negative and the anode current rises. This makes the anode potential more negative with respect to + h.t. The falling voltage across $C$ is nearly equalled by the rising voltage across $R_1$. The total voltage across $R$, and hence the current in $R$, is nearly constant, and so the voltages across $C$ and $R_1$ change nearly linearly.

It is not difficult to show that the circuit behaves as though $C$ were $(1 + A)$ times its actual value and the voltage acting in the circuit were $A$ times the h.t. supply, where $A$ is the voltage amplification of the valve with $R_1$. By using a pentode for the valve, $A$ can be made large and exceedingly good linearity obtained. Thus a valve with $g_m = 6$ mA/V can be used with $R_1 = 10$ kΩ; $A$ is then 60 and with a 300-V h.t. supply the linearity is as if 18,000 V were acting in the circuit.

The maximum output obtainable is, of course, limited by the actual voltage, for it is necessary that the valve should act as a linear amplifier. In general, the maximum linear output is about 50 V less than the h.t. voltage. When combined with a transitron it is less.

The complete linear saw-tooth generator is shown in Fig. 8.29. Starting with $C$ fully charged, the suppressor $g_3$ is at about zero potential and the screen-grid $g_2$ is at its most positive. The discharge starts and the action is as described above, being modified only slightly by the effect of $g_2$ and $g_3$. The valve acts substantially as a linear amplifier, but its amplification is reduced somewhat by the fact that as the grid potential changes in the positive direction, the $g_2$ current increases and so the $g_2$ voltage falls. This gives a species of negative feedback and the effect is substantially that of a pentode amplifier without a by-pass capacitance on $g_2$. The potential of $g_3$ remains nearly constant because the product $C_1R_3$ is small compared with $CR\,(1 + A)$.

As the anode current is larger than the screen current and $R_1$ and $R_2$ are of the same order of magnitude, the voltage drop across $R_1$ is greater than that across $R_2$. As $C$ discharges, the anode voltage falls more than the $g_2$ voltage. Consequently a condition is at length reached in which the anode can no longer draw the full proportion of the total cathode current.

It must be remembered that in a valve of this kind the greater part of the electron stream passes through $g_2$. Of the electrons which do pass it those of low velocity fall back to it and, with those intercepted directly by the grid wires, form the $g_2$ current. The higher velocity electrons come within the field of the anode and pass to it.

When the anode potential falls, a greater number of electrons passing the screen fail to pass beyond the field of $g_2$ because the pull of the anode is weaker. Since more electrons fall back to the screen, the screen current increases and the anode current falls.

The falling anode voltage thus causes a rising $g_2$ current and a falling $g_2$ voltage. Now this voltage change is communicated to $g_3$ through $C_1$, even if only to a small degree at first, and the negative potential on this electrode further increases the $g_2$ current. This makes $g_2$ and $g_3$ both change still more negatively. The effect is cumulative and there is a rapid change over to the fly-back condition with anode current cut off by a highly negative $g_3$ and the $g_2$ current a maximum and voltage a minimum.

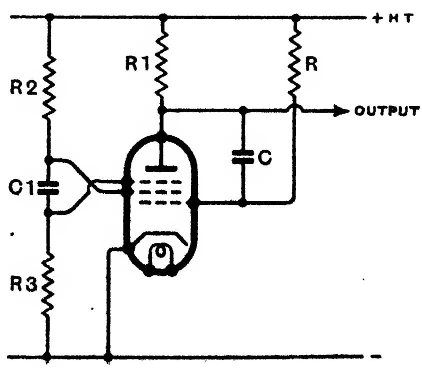



**Fig. 8.29—Complete linear saw-tooth oscillator combining the Miller integrator and the transitron**

As anode current is cut off there is now no voltage drop across $R_1$ and as the voltage across $C$ is less than the h.t. voltage, because $C$ is partially discharged, the grid $g_1$ becomes positive with respect to the cathode. Grid current flows and $C$ is rapidly charged from the h.t. supply through $R_1$ and the $g_1$-cathode path of the valve.

When $C$ is charged the grid-cathode potential falls back to zero and this reduces the cathode current and hence the $g_2$ current. Therefore the $g_2$ voltage rises and $g_3$ becomes less negative. The anode potential is now high, and it starts to draw current. The $g_2$ current thus falls still more, which makes $g_3$ become still less negative. Again the effect is cumulative and there is a rapid transition to the initial condition of $g_3$ at zero volts, $g_2$ at its maximum voltage, $g_1$ negative and the valve as a whole acting as a linearizing amplifier for the discharge of $C$.

While the detail of the action may appear very complex, the circuit is very easy to get working and is very reliable. Valves of the TSP4 and EF50 class are suitable, and at frame frequency suitable values for Fig. 8.29 are:—$R_1 = R_2 = 10\,\text{k}\Omega$; $R_3 = 0{\cdot}5\,\text{M}\Omega$; $R = 4\,\text{M}\Omega$; $C = 0{\cdot}02\,\mu\text{F}$; $C_1 = 0{\cdot}002\,\mu\text{F}$. An output of 40–50 V at the anode is obtainable for an h.t. supply of 220 V. A variable output is easily secured by making $R_1$ a potentiometer and taking the output from the slider. $R$ should be variable as a frequency control.

The output is a negative-going saw-tooth instead of the more usual positive-going. This is about the only disadvantage of the circuit, because most amplifiers for electromagnetic deflection demand a positive-going input. An output of this form can be obtained from the cathode by using the circuit of Fig. 8.30 but its amplitude is only about one-half.

Suitable values for the line scan are:—$R_1 = R_2 = R_4 = 10\,\text{k}\Omega$; $R_3 = 0{\cdot}5\,\text{M}\Omega$; $R_5 = 1\,\text{M}\Omega$; $R_6 = 20\,\text{k}\Omega$; $R = 4\,\text{M}\Omega$; $C = C_1 = 100\,\text{pF}$; $C_2 = 0{\cdot}1\,\mu\text{F}$. If a control of amplitude is needed

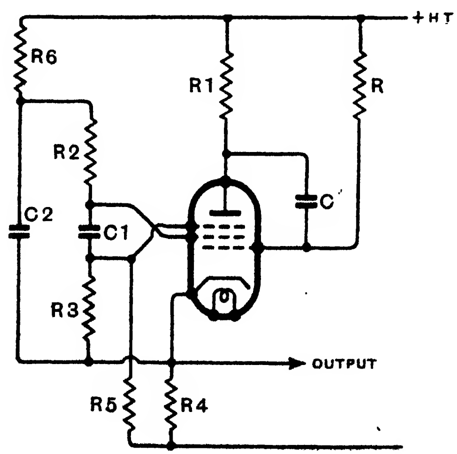
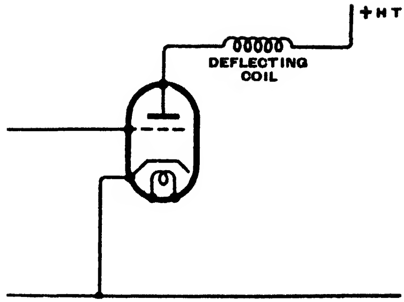



**Fig. 8.30—Linear oscillator arranged to give a positive-going saw-tooth output at the cathode**

**Fig. 8.31—The basic circuit for generating a saw-tooth current waveform**




$R_4$ can be made variable, and in general, $R$ should be variable as a frequency control.

Synchronization is best effected by a negative-going pulse applied to $g_2$ or $g_3$ and is conveniently obtained by joining the anode of the valve supplying the pulses to $g_2$. In the case of Fig. 8.30, however, a positive-going pulse applied to the cathode can also be used, but the action is rather less reliable since the circuit can also be triggered, but with some delay, by a negative-going pulse on the cathode.

It should be pointed out that while the circuit is in most respects very satisfactory indeed and delivers a beautifully linear output it does, in the author's experience, suffer from a certain regular irregularity. Because of this, difficulty has been found in obtaining satisfactory interlacing when using it for the frame-scan generator. The fact, too, that it demands only a small amplitude of sync pulse is not an unmixed blessing, for it makes more thorough screening and decoupling necessary if the frame oscillator is not to be tripped by unwanted pulses. Such pulses can easily find their way into the circuit with only a trace of stray coupling to the output circuit of an electromagnetic line-scan time base. For the line scan, however, it is very satisfactory indeed.

Turning now to apparatus for electromagnetic deflection, the usual procedure is to generate a saw-tooth voltage waveform in one of the ways just described and apply it to an amplifier which provides a saw-tooth current output. Such amplifiers are dealt with in Chapter 7.

It is, however, possible to generate a saw-tooth current directly, and the basic circuit for doing this is shown in Fig. 8.31. A low resistance triode is used and is operated at about zero grid bias and with some 250–300 V on the anode. It is a distant-controlled time base and the sync pulses are applied to the grid in negative phase At frame frequency a sync pulse amplitude of 50–100 V may suffice, but at line frequency it must be much greater.

Take the conditions at the end of a sync pulse. The grid is so negative that there is no anode current. The pulse ends and the grid potential returns to zero. The anode current does not immediately rise to its final value because of the inductance in the anode circuit.

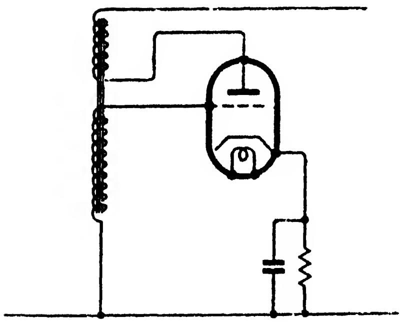


**Fig. 8.32—The use of a self-oscillator for generating a current waveform is illustrated here**

The time-constant $L/R$, where $L$ is the coil inductance and $R$ is the resistance of coil and valve, is such that during the scan time of one line (or frame) the current rises to only about 7 per cent of the final value to which it would rise after a long period. The change of current during the time of one line is thus nearly uniform and a close approximation to a linear scan stroke is obtained.

At the end of the line the sync pulse comes along and drives the grid of the valve negative, thus interrupting the current. The magnetic field collapses and a back e.m.f. is generated across the coil which drives the anode of the valve positive. At line frequency this back e.m.f. is of the order of 1,000–3,000 V, depending on the fly-back time, the coil inductance and the current.

Because of this large anode potential during the fly-back, a very large amplitude of sync pulse is needed to keep the valve non-conductive. A power triode, for instance, may have an amplification factor of 5 and the sync pulse amplitude required is not likely to be less than one-fifth of the anode potential. Thus up to 600 V of sync pulse amplitude may be needed.

It would be very difficult and uneconomical to obtain such an amplitude and in practice this form of time base is not used. A modification of it which is self-running and which only relies upon the sync pulses to keep it in step is quite practical, however, and will give a good performance. The system is one which was used in some of the pre-war Baird Television receivers. It has also been used in Germany and the U.S.A. After a period of disuse, it is gaining in popularity in various modified forms.

In the basic circuit of Fig. 8.32 the coils shown actually form the deflector coils. The operation is similar to that of the arrangement just described, but the grid voltage is provided by induction from the anode circuit. Start with the valve beyond current cut-off. It is kept there by a large negative voltage on the grid provided by the collapse of the magnetic field on the fly-back. When the anode-coil current has reached zero the back e.m.f. disappears and with it the negative grid voltage.

The grid potential then returns towards zero to a value provided by the cathode-bias resistance. Anode current starts and rises at a rate determined by the inductance and resistance. The changing anode current sets up a back e.m.f. across the anode coil which acts in opposition to the h.t. voltage and lowers the anode potential. At the same time the induced e.m.f. in the secondary, or grid, coil drives the grid positive.

The action is cumulative and the rising grid potential tends to increase the rate of change of anode current and so make the grid potential more positive. The action, however, is complicated by grid current, for the input resistance of the valve falls as the positive grid potential increases and so places an increasingly heavy load on the grid coil. In fact, it acts rather as a diode switch such as that described on page 99.

With careful design a good approximation to a linear scan can be obtained. At length, the rate of change of anode current begins to fall off and the grid potential becomes less positive, thus further slowing up the rate of change of current. The action is again cumulative and the anode current begins to fall. The induced e.m.f. on the grid is then reversed in sign and it drives the grid negative.

Once the fly-back is initiated the anode current falls very rapidly and the back e.m.f. on the anode reaches some 2,000 V as before. This time, however, there is an induced e.m.f. of the same order on the grid driving it negative. While the anode still goes positive by something of the order of 2,000 V, the grid goes negative by a similar amount, so that the valve is unquestionably beyond anode current cut-off.

When the rate of change of coil current becomes less towards the end of the fly-back period these voltages become smaller and they become zero when the current ceases to change. The cycle then recommences.

In practice synchronization is effected by applying a negative pulse to the grid. This pulse, arriving before the fly-back would start of itself, lowers the grid potential and hence the anode current and the grid is rapidly driven very negative by the induced e.m.f.

For frame deflection Baird Television used the arrangement shown in Fig. 8.33. This is essentially the same as the arrangement just discussed, but there is an additional coil $L_3$. This is included to balance out the direct current through $L_1$ and $L_2$. There is a mean anode current through $L_2$ and a mean grid current through $L_1$ which tend to balance one another in their effect on the core, but do not do so completely since the ampere-turns are unequal. The coil $L_3$ is provided, therefore, so that the balance can be completed. As far as the mean currents are concerned the ampere-turns of $L_1$ are equal to the ampere-turns of $L_2 + L_3$. This balancing of the mean currents is necessary to obtain the picture in the centre of the tube screen. Without it, the steady field due to the mean current would deflect the picture off the tube, for the coils shown are actually the deflector coils. The choke *Ch* is included in order to prevent $L_3$ from having any effect on the saw-tooth wave. The

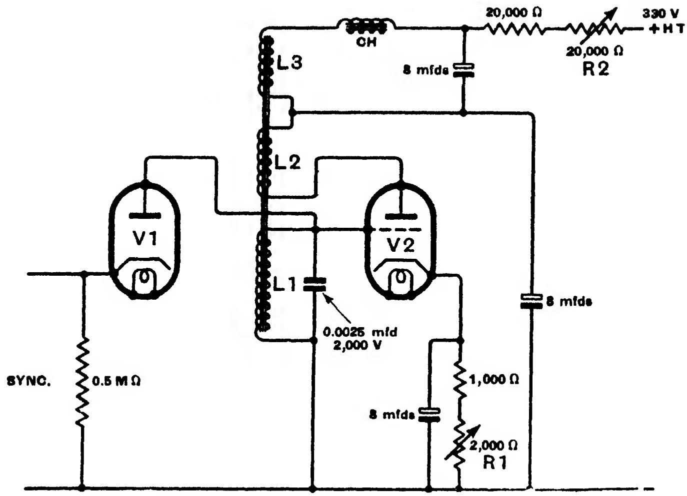



**Fig. 8.33—The Baird frame oscillator takes the form shown here**

valve $V_2$ is a triode such as the Cossor 41MP or Mazda AC/P, and $R_1$ and $R_2$ are the frequency and amplitude controls.

The sync pulse is applied through a diode $V_1$, actually the Mullard 2D4A. This is necessary to obtain an adequate fly-back time. If the diode were not used the coil $L_1$ would be shunted by the impedance of the sync circuits which is usually about 7,500 Ω or so. The rate of change of current in the coils would then be greatly decreased and the fly-back time far too long.

Actually, it is probable that $V_1$ could be dispensed with in the case of the frame time base, especially if the sync pulse were fed to the grid through a very small capacitor. The diode has one great advantage, which is that the large negative pulse which appears on the grid on the fly-back drives the diode anode negative and makes it non-conductive. This pulse, therefore, is prevented from reaching earlier circuits, except via the diode anode-cathode capacitance. Interaction between the time bases is thus reduced and better interlacing secured.

At line frequency the arrangement is fundamentally the same but differs in detail. The circuit is shown in Fig. 8.34. In this case the oscillator and deflector coils are separate, so that there is no need to provide balancing coils for the mean currents. $L_1$ and $L_2$ are wound on a closed iron core and the deflector coil $L_3$ is fed from a tapping on $L_1$ through a large capacitor. $L_1$ thus constitutes a step-down ratio auto-transformer.

Grid bias is derived through the flow of grid current through $R_1$ and $R_2$, the latter of which is adjustable to form a frequency control. No amplitude control is provided. The valve $V_3$ is a diode, actually the Mullard UR1C h.t. rectifier; it conducts only when the anode potential falls below zero. It is, in fact, a damping diode.

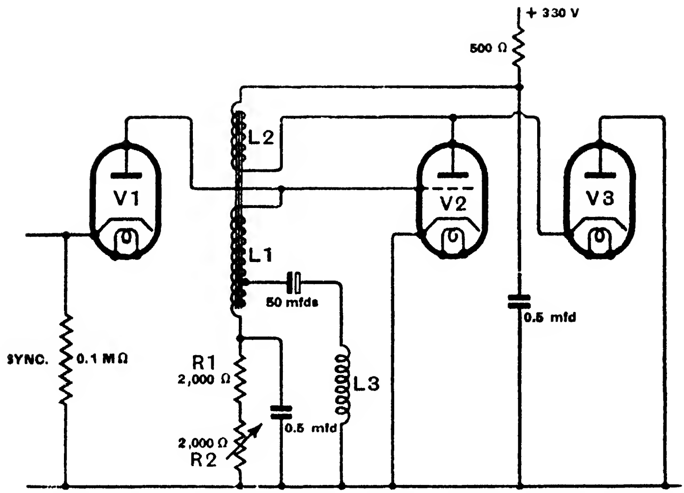



**Fig. 8.34—The line-scanning oscillator used by Baird is rather different from the frame circuit**

The valve $V_1$ is a Mullard 2D4A as in the frame time base, and $V_2$ is a Cossor 41MP. The two time bases together consume about 40 mA at 330 V and are thus extremely economical. The output is sufficient to scan a 9-in tube.

It may be remarked that it is necessary to provide a separate 20-V winding on the mains transformer for the heater of $V_3$, and this winding must be insulated for 2,000 V.

So far as the anode circuit is concerned these circuits must function in a similar manner to the anode circuit of an amplifier and the details given in Chap. 7 must necessarily apply. The differences are brought about chiefly by the fact that the grid current in the transformer provides part of the scanning current and influences the linearity. The action on the scan is very much like an amplifier with a damping diode, the grid-cathode section of the valve acting as a damping diode.

The great merit of the circuit is its economy. Its main disadvantage is that it is difficult to design for there are so many inter-related factors which are almost impossible to pre-determine. The best method of attack is probably to design the anode circuit on the lines of an amplifier, but allowing for higher circuit capacitance because the grid-cathode capacitance of the valve multiplied by the square of the grid/anode turns ratio of the transformer is effectively in the anode circuit.

The grid circuit turns needed depend on the valve and there should usually be at least as many as on the anode coil. The valve used must be a power triode and of a type which will withstand some 40 mA peak grid current as well as the high voltages.

Before concluding this chapter some remarks should be made on

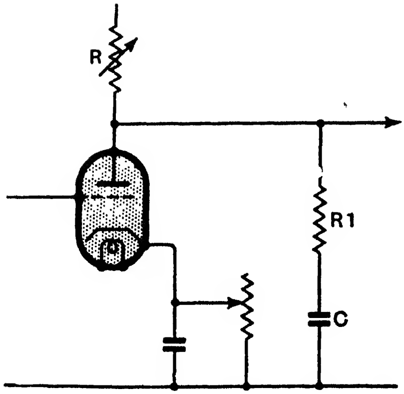



**Fig. 8.35—A resistance $R_1$ is sometimes inserted in series with the charging capacitance**

an addition which is sometimes made to the saw-tooth voltage generator. This consists of the insertion of a resistance—usually 500–3,000 Ω—in series with the charging capacitance.

This arrangement is shown in Fig. 8.35, and the resistance is $R_1$. On the charge stroke $R_1$ makes little difference; there is a small and nearly constant voltage drop across it and the output is the voltage across $C$ less this constant voltage.

On the fly-back, however, it makes a profound difference. The voltage across $C$ falls in the usual way, except that it now takes a little longer for $C$ to discharge because the total resistance of the discharge path is increased by the presence of $R_1$.

If $R_1$ is fairly high in comparison with the internal resistance of the valve, the voltage developed across it is almost equal to the capacitance voltage. The output is taken across $R_1$ and $C$ and is consequently the sum of the voltages across each. As it is the discharge current which is responsible for the appearance of a voltage across $R_1$, and it is the capacitance voltage which is responsible for the discharge current, the voltage across $R_1$ opposes that across $C$. As they are nearly equal, the output voltage drops instantaneously very nearly to zero.

The capacitance continues to discharge and the output remains nearly zero. When the discharge is completed, however, $C$ starts to charge again through $R$ and the direction of current through $R_1$ is reversed, and there is consequently a sudden rise in potential at the output terminals, preceding the scan stroke.

The waveform produced is sketched in Fig. 8.36. The capacitance discharge starts at A and the output voltage falls to B because of the voltage developed across $R_1$. The discharge is completed during BC and the output is nearly zero and substantially constant.

The discharge is completed at the point C and recharging starts. The sudden rise CD is caused by the voltage drop across $R_1$. If the charging current is not constant, but falls off as the voltage across $C$ rises, the voltage across $R_1$ will also fall. This will make

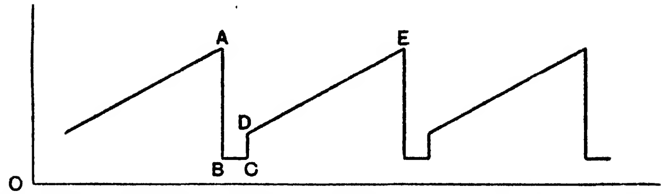



**Fig. 8.36—The use of a resistance in series with the charging capacitance introduces a pulse into the waveform**

the output rise still less rapidly. Consequently, $R_1$ will accentuate any non-linearity of the scan caused by an exponential charging current.

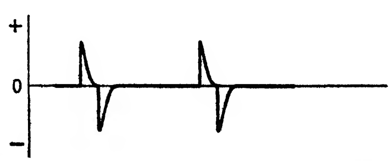
It is often said that this circuit is used to generate the combination of a pulse and saw-tooth which is the voltage wave across the $L$ and $R$ of a deflector coil with a linear current. It is said that this grid-voltage wave is necessary for a linear scan. This is, however, only true if a triode output valve is used; with the usual pentode a pulse component in the grid-voltage wave is not necessary. The circuit is actually used to cut off the output valve more quickly on fly-back.

No saw-tooth generator is sufficiently regular in its action to be allowed to run free, it must always be locked in some way by the synchronizing pulses in the signal. The normal procedure is to separate the pulses from the picture signal with a limiter, then to separate the line and frame pulses from each other and finally to apply the pulses to the line and frame saw-tooth generators so that they initiate the fly-backs. The separation process is treated in Chapter 16.

In the case of the line time base the leading edges of the pulses are usually well retained and are as sharp as it is possible to make them. They have the general form shown in Fig. 8.37 in which the leading edges are positive-going. The action of the pulses is much the same with all generators but is most easily seen with the blocking oscillator, Fig. 8.8.

Assume that the waveform of Fig. 8.37 is developed across $L_1$ of Fig. 8.8, say by induction from another winding. The grid-cathode voltage is then the sum of the saw-tooth voltage across $C$ and the sync voltage, and it appears as shown in Fig. 8.38 (a). Here the dotted line represents the voltage at which the valve starts to draw current and which must be reached before the regenerative fly-back action can occur. The pulse shown in (a) is just too small to

**Fig. 8.37—Typical pair of line sync pulses**



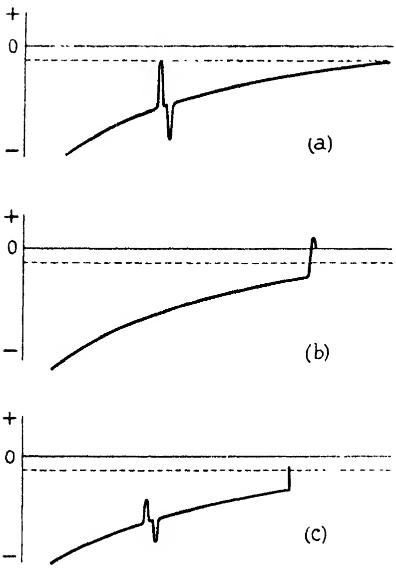



**Fig. 8.38—Line sync pulses superimposed on the saw-tooth wave of a blocking oscillator. At (a) the pulse is not large enough to trip the oscillator, but at position (b) the pulse is amply large enough to do so. At (c) the pulse half-way along is too small to have any effect**

reduce the grid-cathode voltage sufficiently to trigger the valve. To do this it must obviously either be of larger amplitude or occur later in the cycle. The second case is shown in (b); here only the leading edge is shown for in practice the rest of the pulse is masked by the fly-back wave, Fig. 8.10.

If the sync-pulse amplitude is very small it clearly can trigger the time base only if it occurs just before the oscillator is ready to trip naturally. This means, in practice, that the free-running frequency must be very close to the locked frequency. It means that the " Hold " control is critical in its setting and requires frequent adjustment, for there is little tolerance to accommodate small circuit changes and mains-voltage fluctuations. On the other hand, the circuit is not very sensitive to any spurious signals which may accompany the sync pulses.

If the sync-pulse amplitude is large, the time base will be triggered quite early on in its natural cycle. The saw-tooth amplitude will be reduced considerably and the locked frequency will be much higher than the free-running. The " Hold " control will have little effect on the lock over quite a wide range of settings and, since it affects the charging current, it will do little more than alter the amplitude. In this condition the lock is very hard and there is a great tolerance for minor changes. However, the time base is much more likely to be tripped by any spurious signals.

If the circuit is adjusted for a hard lock it necessarily means that the sync-pulse amplitude is a good deal greater than the critical amplitude needed just to trigger the time base at the same point with a critical setting of the " Hold " control. If there is a spurious pulse of a little smaller amplitude than the sync pulse and a little earlier in time, it will not affect matters with a weak lock, because it will not bring the valve to the triggering point. With a hard lock, however, it will probably still be sufficient to trip the time base, which will fire early on the spurious pulse.

At short distances from the transmitter, where interference is small and circuit noise is negligible, there is no difficulty in obtaining good clean sync pulses free from any spurious signals and it is possible to use a very hard lock. At long distances circuit noise alone may be sufficient to trip the time base critically and it is good practice to use a fairly light lock. It should only be just hard enough to cover mains-voltage changes and minor alterations in valves and components. Occasional adjustment of the " Hold " control will then be necessary. It is sometimes helpful to restrict the high-frequency response of the sync channel somewhat, as by adding shunt capacitance to a coupling. Although this is disadvantageous in reducing the sharpness of the leading edges of the pulses, it often so restricts interference that there is a net gain.

During the frame pulses, sharp pulses occur at half-line intervals (see Fig. 16.18). If the sync-pulse amplitude is too great the line time base will be tripped at half-line intervals during the frame flyback. In itself this is not very important, for the scanning spot is blacked out while it occurs, but it upsets the regular operation of the time base. In particular there is a shift in the mean d.c. level in the circuits. What happens precisely depends on the circuit time-constants, but usually the top quarter-inch or so of the picture is shifted to one side.

The effect is slightly less likely to occur when the saw-tooth across *C*, Fig. 8.8, is nearly linear than when it is markedly exponential, but this is not of great practical importance. Its presence or absence depends chiefly on the adjustment of the " Hold " control. For maximum immunity from it this control must be nearly at one end of the " Hold " range; the other end will give a maximum of top displacement. This can be seen from Fig. 8.38 (c), where a half-line pulse is shown half-way along the saw-tooth. The amplitude is such that the proper pulse only just trips the time base and the pulse at half-time has no effect.

If the pulse amplitude is increased, clearly the time base will eventually trip at half-time. If instead the charging rate is altered the effect is that of sliding the two pulses to the right along the curve, and again, the half-line pulse will eventually trip the time base.

It would be quite possible to avoid the effect by eliminating the half-line pulses. One way of doing this would be to add a saw-tooth to the line sync pulses and apply the resultant to a limiter adjusted to pass only the proper full-line pulses. The added complication of this is not usually considered worth while and, in effect, the saw-tooth generator is adjusted to respond only to the full-line pulses by what is virtually the same mechanism.

The synchronizing action has been explained in the above with

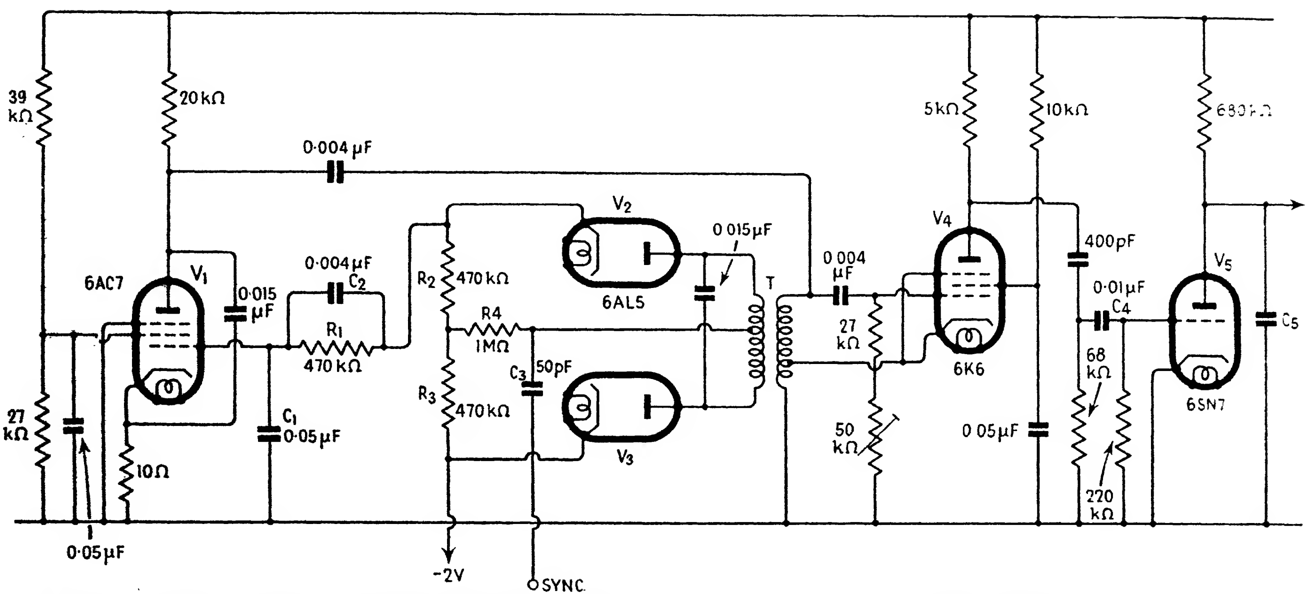



**Fig. 8.39**—Circuit of fly-wheel sync system including a sine-wave oscillator $V_4$ to control the saw-tooth generator $V_5$. The values of components are suited to the American scanning frequency, but should not differ greatly for the British standards

reference to the blocking oscillator, with which the sync pulses and saw-tooth wave actually are added on the grid. It is, however, identical with other forms of generator even when the two waveforms are not directly added, because the voltage required on the sync electrode to trip the time base does usually depend on the saw-tooth amplitude even if this voltage is actually on a different electrode.

In the frame time base most of the above applies equally, but interference and noise are usually much less troublesome; also, there is no equivalent of the half-line pulse trouble. Nevertheless, there is often considerable difficulty in securing a sufficiently regular action for good interlacing. Poor interlacing may occur through the use of an integrating type of frame sync-pulse separator as explained in Chapter 16 but even when great care is taken to ensure that the sync pulses are nearly perfect good interlacing may still not be obtained.

The usual cause is a line-frequency signal in some part of the saw-tooth generator. It may not necessarily be on the sync electrode, it may be on one of the others, for it is usually possible to synchronize on any electrode, even if not very well. The line scan output valve normally has a voltage pulse on its anode of 2–3 kV on each line fly-back, and in fly-back e.h.t. circuits a pulse of 5–10 kV may be developed. This pulse coincides with the frame sync pulse in even frames, discounting minor delays of a few microseconds, but half a line before it on odd frames.

If it is present on the frame saw-tooth generator, therefore, it may trip it before the arrival of the proper frame pulse. The tolerable amount of line signal in a frame time base depends on the type of generator but is always quite small. It should never be allowed to exceed one-tenth of the frame-pulse amplitude and should preferably be smaller.

This normally means that well under one volt is the maximum line-frequency signal permissible in the frame generator circuits and as pulses of up to 10 kV may exist in the line circuits it is clear that careful screening is necessary. Even when this is provided there is always some coupling between the deflector coils, and line-frequency pulses are usually present on the anode of the frame output valve. From there they may find their way back to the generator through the anode-grid capacitance of the valve or through a feedback network.

Although it is the usual practice, it is not necessary to apply the sync pulses to the saw-tooth generators in order to secure synchronism. Instead, what is usually termed "fly-wheel" synchronizing can be used. With this, the locally-generated wave is compared with the sync pulses in a discriminator circuit. This produces a unidirectional voltage depending on the phase difference between the two and of polarity depending on which way they are out of phase. This voltage is used to control the saw-tooth generator and it acts to bring the difference between the two to zero. It is analogous to the automatic frequency control system which is sometimes used on the oscillator of a superheterodyne.

There is a considerable time-constant imposed on the system so that it responds to no individual sync pulse but only to the combined effect of a large number, because of this it is very little affected by interference and set noise, and the system is well suited to the conditions of long range reception. The saw-tooth generator used with it often starts with a more-or-less sine-wave oscillator of the *LC* type. This runs at line frequency and is maintained in synchronism by the control voltage from the discriminator acting on a reactance valve. Its output is used to generate pulses which in turn control the saw-tooth generator itself.

Of the various forms of fly-wheel circuits the one commonly termed the " sine-wave system " is reputed to provide the greatest immunity from interference and a typical circuit is shown in Fig. 8.39. A pentode $V_4$ is used as a sine-wave oscillator; it is a cathode-tap Hartley oscillator using the screen grid as an anode and operates at the line frequency. Current to the true anode is cut off during a portion of each cycle and so the waveform on the anode is more or less a sine-wave with its positive half-cycle very much flattened. This wave is differentiated by the *RC* network coupling $V_4$ and $V_5$ and appears on the grid of $V_5$ as a positive-going pulse.

This valve $V_5$ is the saw-tooth generator and is cut off except when the positive pulse arrives on its grid. Grid and anode currents then flow. The former provides sufficient charge on $C_4$ to keep the valve cut off except during the pulse and the latter discharges $C_5$.

The sine-wave is applied to the two diodes $V_2$, $V_3$ and also to the reactance valve $V_1$. This behaves as a reactance in shunt with the oscillator tuned circuit and the value of the reactance depends on the grid voltage. Hence, by varying the grid voltage of $V_1$, the oscillator frequency can be altered.

The oscillator voltage is applied to $V_2$ and $V_3$ in push-pull so that the two diodes conduct equally on alternate half-cycles when no other signal is present. The currents in their load resistances are equal and the resulting voltage drops are added in opposition and so cancel to provide zero voltage across $R_2$ and $R_3$.

The sync pulses, however, are also applied to the diodes through $C_3$, in the same phase to both. Again, the total output voltage is zero in the absence of the sine-wave signal.

When both are present, however, the balance is upset unless the sine-wave input voltage is passing through zero when the sync pulse arrives. At any other time the input to one diode will consist of the sync-pulse voltage plus some sine-wave voltage but the input to the other will be this same sync-pulse voltage minus the same sine-wave voltage. If the sync-pulse amplitude is greater than the value at that instant of the sine-wave voltage, both diodes will conduct, but because their inputs are unequal their currents will be unequal also and the voltages across $R_2$ and $R_3$ will not balance.

There will then be a difference voltage across $R_2$ and $R_3$ which is applied through the integrating network $R_1$, $C_1$, $C_2$ to the reactance valve. This alters the effective reactance of the oscillator circuit in such a way as to alter the phase and frequency of the sine-wave so as to nullify the difference between it and the sync pulses.

If the phase is such that when the sync pulse arrives the sine-wave voltage is making the anode of $V_2$ positive and the anode of $V_3$ negative, $V_2$ conducts more than $V_3$ and a positive voltage is applied to the grid of $V_1$. On the other hand, if the error is opposite and the sine-wave makes the anode of $V_2$ negative and the anode of $V_3$ positive, then $V_3$ conducts more than $V_2$ and the output to the grid of $V_1$ is negative.

The immunity from interference arises chiefly from the high time-constant of the integrating circuit. Because of this the circuit responds only slowly and, if disturbed, may take an appreciable time to lock-in on the regularly recurring sync pulses. Impulsive interference and noise is essentially irregular and over any lengthy period there will probably be as many interference pulses tending to pull the circuit out one way as there are tending to pull it out the other. Their net effect, therefore, tends to zero.

The second factor which affects the immunity to interference, and which is responsible for the reputation of this circuit as being the best type of fly-wheel circuit, is the use of a sine-wave oscillator. This type of oscillator is inherently more stable than the relaxation type, such as the blocking oscillator.

Other circuits without the sine-wave oscillator are widely used, however, and are considerably simpler. One method is to compare the sync pulses with the saw-tooth scanning wave in a double-diode discriminator similar to the one employed in the sine-wave circuit. The saw-tooth is usually obtained by integrating the pulse voltage from the line output transformer. The mean output voltage of the discriminator is applied as grid bias to the blocking oscillator which generates the saw-tooth.

These circuits have been described in the American literature* and their widespread use in that country appears to be due more to the fact that receivers must there be capable of operating without adjustment on any one of several local transmitters than to there being a large amount of interference. Apparently the sync pulses produced at different stations are not sufficiently alike for the ordinary circuits to be satisfactory.

* "Automatic Frequency Phase Control of Television Sweep Circuits," by E. L. Clark, *Proc. Inst. Radio Engrs.*, May 1949, p. 497.
"Locked Oscillator for Television Synchronization," by Kurt Schlesinger, *Electronics*, January 1949, p. 112.
"Modern Television Receivers" by Milton S. Kiver, *Radio & Television News*, Part 21, January 1950, p. 45; Part 22, February 1950, p. 43; Part 23, March 1950, p. 50.

CHAPTER 9

# Vision-frequency Amplification

IN THE TELEVISION receiver the vision-frequency amplifier corresponds to the audio-frequency amplifier of a sound receiver, but differs from it in several important particulars. The amount of amplification is usually much lower and it is required to give a voltage output rather than a power output. The frequency response needed is very much wider in range, and phase distortion can be very important, but amplitude distortion is usually a secondary consideration.

As applied in a television receiver v.f. amplification rarely possesses any really difficult problems. This is not the case in the transmitter, however, for there the need is for high amplification with a frequency response extending down to zero, and high-gain d.c. amplifiers are notoriously difficult! The low gain required in a receiver makes the use of more than one stage unnecessary in most cases and because of this the employment of direct coupling is often practicable. Even when it is not, the presence of the synchronizing pulses with the vision signal proper so breaks up the latter that the use of a.c. couplings is permissible.

In this chapter the problems of v.f. amplification will be discussed only in so far as they apply to receiving apparatus. Multi-stage amplifiers, such as are used in transmitting-camera channels, will not be considered. Most of the matters discussed here apply to the transmitting case, of course, but they do not fully cover it.

The purpose of the v.f. amplifier of a television receiver is to amplify the output of the detector to a sufficient level to modulate the beam of the c.r. tube. The voltage needed for this depends on the design of the tube; in early patterns it was often about 60 V p–p but modern designs usually need only 20–30 V p–p. This is the vision-signal amplitude corresponding to a change of modulation depth from 30 per cent to 100 per cent in the r.f. signal.

The synchronizing pulses are normally preserved throughout and applied to the c.r. tube with the vision signal, so that of the output of the v.f. amplifier 70 per cent corresponds to the vision signal and 30 per cent to the sync pulses. If a tube needs an input of 21 V p–p, therefore, the v.f. stage must give a total output of $21/0{\cdot}7=30$ V of which 9 V is sync pulse.

The tube does not itself require power, for its input resistance is high enough to be ignored. It has, however, an input capacitance which is often around 15 pF and which must be taken into account.

The attainment of an adequate output and adequate linearity is chiefly a matter of choosing a suitable valve for the output stage and applying to it the correct electrode potentials. The current needed to produce the output voltage across the coupling impedance depends on the value of this impedance, however, and so the choice of valve and operating conditions cannot be made until the impedance is known. It depends on the type of circuit used and upon the high-frequency performance required.

The first step is, therefore, to consider the performance of v.f. couplings at high frequencies. There are two alternative lines of attack; one is based on steady-state characteristics and the other on the transient response. Since the transmitted waveform is specified in terms of the amplitude–frequency response the former is usually considered the simpler. Unfortunately, the phase–frequency response requires to be known also and this is not at all easy to measure. Then when the responses are imperfect, as they always are in practice, it is not at all easy to determine the effect of the imperfections on the picture.

The transient response is much easier to interpret in terms of picture quality and is not too difficult to measure with cathode-ray apparatus. Its main disadvantage lies in certain mathematical difficulties of circuit analysis. However, the circuits used in television *receivers* are fairly simple and transient analysis is very useful.

In the transmitter frequencies up to 2·75 Mc/s are retained and so it is necessary for the response of the receiver to be substantially uniform up to this frequency. In itself, there is no point in maintaining a uniform amplitude–frequency characteristic at higher frequencies, but it is sometimes done because it often enables the phase characteristics to be improved at frequencies below 2·75 Mc/s. Unless phase-equalizers are used, a sharp frequency cut-off is invariably accompanied by large phase distortion.

It is not necessary here to analyse the characteristics of the various circuits used, for the procedure is well-known and quite straightforward. Instead, only the results of the analyses are presented—mainly in graphical form.

The basic circuit from which all other v.f. couplings have grown is the simple resistance coupling of Fig. 9.1. Its defects all arise from the inevitable stray capacitance in shunt with the coupling resistor $R$. This capacitance is made up mainly of the anode-cathode capacitance of the valve plus the grid-cathode capacitance of the following valve or c.r. tube plus incidental wiring capacitances. Its total value is often around 30 pF and it is represented in Fig. 9.1 by $C$.

The way in which the amplification varies with frequency is shown by curve A of Fig. 9.3. This curve gives the response in decibels relative to the response at zero frequency in terms of the product $fCR$, since this makes it of universal application. To illustrate its use, and the kind of performance obtainable from simple resistance coupling, suppose that a response of $-1$ db at 2·75 Mc/s

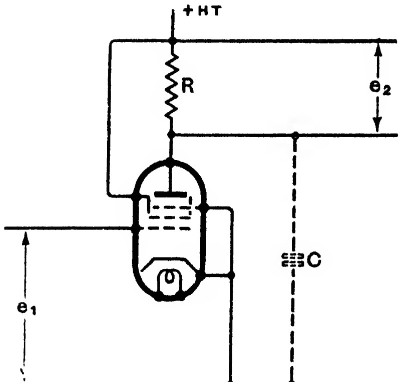



**Fig. 9.1—This diagram shows the basic circuit of a resistance-coupled v.f. stage**

is required and that $C$ is 30 pF. For $-1$ db, curve A shows $fCR=80$; therefore $R=80/(30\times2{\cdot}75)=0{\cdot}97$ kΩ.

Assuming that the valve is of the pentode type, which is usual, so that its a.c. resistance is very large compared with $R$, the amplification is $g_mR$. This holds also for other circuits to be described later. It should be noted, however, that cathode feedback may reduce the effective value of $g_m$.

If $g_m=6$ mA/V, a common figure, the amplification in this example becomes $6\times0{\cdot}97=5{\cdot}82$ times. This may be sufficient, for if the output required is 30 V p–p, an input (and hence a detector output) of 5·15 V p–p will be needed and this is not too large for the detector to supply.

However, if 30-V p–p output is needed with a resistor of 0·97 kΩ, a peak-to-peak current change of $30/0{\cdot}97=31$ mA must occur in the resistance and the valve must supply it. A certain minimum current must be allowed in the valve and its peak current would, therefore, have to be around 40 mA while the mean current, which depends on the picture content, would probably be about 20 mA.

Alternative couplings allow a higher value of $R$ to be used and consequently make the stage gain higher. It is often more important, however, that they proportionately reduce the current needed by the v.f. stage.

Before proceeding to discuss these alternative couplings it is instructive to consider the simple resistance circuit from the transient viewpoint. When the grid-cathode voltage of the valve is suddenly changed from one value to another the output voltage does not change immediately but takes a certain time to reach its new value. The law of change is shown by curve A of Fig. 9.6 in terms of a stage of unity gain.

The output reaches 95 per cent of its final value when $t/CR=0{\cdot}003$. In the previous example $CR$ was $30\times0{\cdot}97=29{\cdot}1$, so that $t=0{\cdot}003\times29{\cdot}1=0{\cdot}0873$ μsec. This means that with a picture 10 in wide the scanning spot moves $10\times0{\cdot}0873/82{\cdot}77=0{\cdot}01$ in while the output is changing from its initial to within 5 per cent of its final value.

Thus an ideally abrupt transition from black to white is reproduced as a blurred edge of 0·1 per cent of the picture width.

It is of interest at this point to consider the mechanism involved. Suppose that the grid voltage of the valve is suddenly changed so that the current through it increases instantaneously by the amount $I$. If $R$ were absent this current would flow in $C$ and the voltage across $C$ would rise linearly and be $E=It/C$. (It should be noted that it is the anode end of $C$ which is free to move in potential; the anode voltage falls nearer to earth potential, but the voltage across $C$ increases in magnitude.)

Now if a given voltage $E$ is required with a given capacitance $C$, time and current are inversely related. If the voltage is required to change very quickly a large current is needed. To raise the voltage it is necessary to convey a given charge into the capacitance; the quicker this must be done, the greater must be the rate of flow of charge and, hence, the current, for current is nothing but the rate of flow of charge.

The relation $E=It/C$ expresses a fundamental limit which no practical circuit can reach. To change the voltage $E$ across a capacitance of $C$ farads in $t$ seconds it is necessary to supply a current $I$ amperes for that period.

Now when there is a resistance in shunt with $C$ the voltage starts to rise at the same rate as if $C$ alone were present, but as soon as some voltage appears $R$ starts to draw current and so robs $C$. The rate of change of voltage, therefore, decreases. Initially all the current flows into $C$, finally it all flows in $R$.

The theoretical limit on performance is reached when $R$ is infinite, for then all the current flows into $C$, and the most rapid charging possible for a given current change occurs. A shunt resistance is necessary, however, to put a finite value to the change of voltage if the current change is maintained indefinitely. It is also needed to pass the mean anode current.

The object in the more complex circuits is to prevent the resistance from robbing the capacitance of current while it is charging and then to effect a rapid diversion of current from the capacitance to the resistance when it is nearly fully charged. Viewed on a frequency basis, the object is to make the circuit impedance more nearly independent of frequency.

The circuit which is most widely used for v.f. amplification is shown in Fig. 9.2. The capacitance $C_3$ is rarely present as a component, but consists merely of the self-capacitance of the inductance $L_1$. There are several possible relations between the components. One commonly used relation is $L_1/CR^2=0·5$, but this leads to a small hump in the frequency-response curve.

The condition for a maximally-flat response is $L_1/CR^2=0·414$ and $C_3/C=0·354$. If $C_3$ is negligibly small, the $L_1/CR^2$ value holds unchanged but the response is not quite so good in that the maximum $CR$ value is slightly smaller.

The performance is shown by curve B of Fig. 9.3 and for a

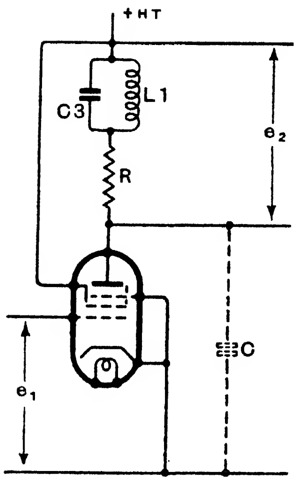



**Fig. 9.2—The shunt-corrector circuit has an inductance $L_1$ in series with the coupling resistance ; $C_3$ is often merely the self-capacitance of $L_1$**

response of $-1$ db, $fCR=240$. The circuit is just three times as good as simple resistance coupling under this condition and for $C=30$ pF and $f=2{\cdot}75$ Mc/s, $R=2{\cdot}91$ kΩ. As a result, the current of the stage is one-third and the gain three times as great. If $C=30$ pF, clearly $C_3=10{\cdot}62$ pF.

The transient response is indicated by curve B of Fig. 9.6 and is of quite a different character from the simple coupling. The rise is much quicker, but the output voltage overshoots its final value and oscillates about it. It is important to consider what the visible effect of this on the picture will be. Taking a 10-in wide picture and a $CR$ value of $30\times2{\cdot}91=87{\cdot}3$, from Fig. 9.6 the final output is reached at $t=0{\cdot}0016\times87{\cdot}3=0{\cdot}15$ $\mu$sec, a displacement of the canning spot of $10\times0{\cdot}14/82{\cdot}77=0{\cdot}017$ in.

The voltage does not stop there, however, but overshoots about 7 per cent at $t/CR=0{\cdot}0023$. This corresponds to a further spot displacement of $(0{\cdot}0023-0{\cdot}0016)$ $0{\cdot}017/0{\cdot}0016=0{\cdot}0074$ in. At this point the spot will be 7 per cent brighter than it should be.

The voltage then falls and undershoots by 3 per cent at $t/CR=0{\cdot}0033$, a further displacement of $(0{\cdot}0033-0{\cdot}0023)$ $0{\cdot}017/0{\cdot}0016=0{\cdot}0106$ in.

The effect of this is for any transition in the picture to be followed by a series of bands corresponding to the oscillations of the circuit. A black vertical edge is thus followed by a white band and this in turn by alternate grey and white bands.

The effect can, and does, occur in practice to a marked degree and may be very unpleasant. However, it would be very wrong to think that it does so merely as a result of a single v.f. coupling. The effect as it appears on the picture is the sum total of similar effects occurring throughout the chain from the transmitting camera to the receiving tube. We are concerned here only with the receiver and have to assume a perfect transmission. Consequently, we assume that any such striations appearing on the picture are a receiver fault. This is not necessarily true, for the transmission may not be perfect and reflections during propagation can cause similar effects. However, even if the fault lies in the receiver it is by no means necessarily due to the v.f. couplings. In fact, the r.f. circuits are a more probable location for it.

It is doubtful if in practice the 7 per cent overshoot of curve B of Fig. 9.6 would produce visible distortion of the picture. The reason is that no input wave to the v.f. stage ever does produce the sudden change of grid voltage which has been assumed in computing the curves of Fig. 9.6. The grid voltage takes a finite time to effect its change and this time is of the same order as the rise time of the v.f. stage output circuit itself. This does not prevent the response still being of an oscillatory nature, but it reduces the amplitude of the oscillations.

By adopting different circuit relations it is possible to make the response non-oscillatory. This is obtained if $L/CR^2=8/27$ and $C_3/C=1/8$. The response is then given by curve A of Fig. 9.7.

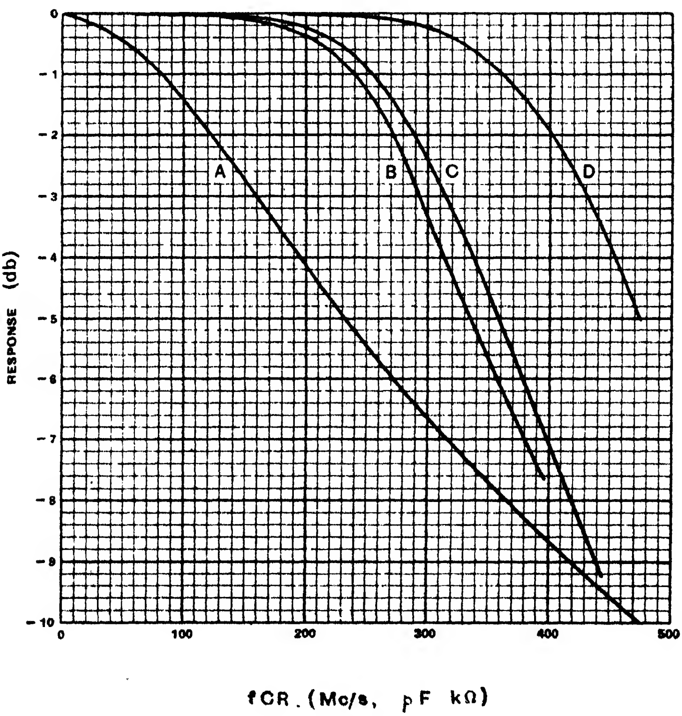



**Fig. 9.3—Frequency-response curves of the four circuits discussed in the text : A, resistance-capacitance coupling ; B, shunt-corrector ; C, series-corrector ; D, series-shunt-corrector**

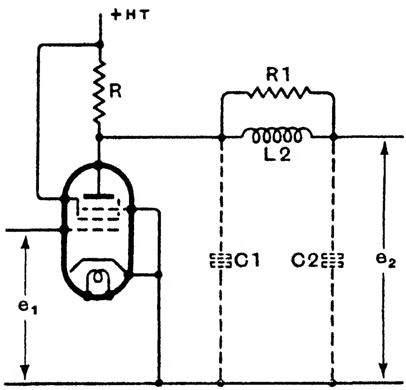



**Fig. 9.4—The series-corrector circuit takes the form shown here. The resistance $R_1$ is not always needed**

Another circuit which is quite often used is shown in Fig. 9.4 and has the configuration of a low-pass filter. The circuit capacitance is divided by $L_2$ into two parts $C_1$ and $C_2$, and the value $C$ which appears in the performance curves is this sum (i.e., $C=C_1+C_2$). It is assumed that the self-capacitance of $L_2$ is small enough to be neglected, but this is not always true unless a carefully-designed coil is used.

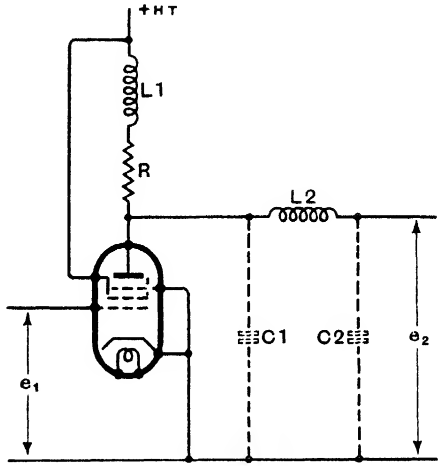
Curve C of Fig. 9.3 gives the response for the maximally-flat condition. Except when otherwise mentioned $R_1$ is absent. The performance is only very slightly better than with the shunt circuit of Fig. 9.2 and the circuit has the disadvantage of requiring the circuit capacitances to be arranged in a particular manner. It is not always easy to maintain the correct ratio $C_1/C_2$ without adding capacitance.

The transient response is given by curve C of Fig. 9.6 and it will be seen that the overshoot is greater than with the shunt circuit. Also, the start of the curve is quite different. Instead of the output starting to rise at its greatest rate and then slowing off, it starts to rise very slowly and it is not until $t/CR$ is about 0·0004 that it approaches a rapid rise. This is a characteristic of all 4-terminal networks (that is, circuits having different input and output terminals) and there is no particular harm in it. It means, in effect, that the signal as a whole is delayed slightly; the signal takes a small finite time to pass through the circuit.

This circuit also can be used in a critically damped condition to avoid an oscillatory response and the response then is given by curve $C_1$ of Fig. 9.7. If the proper ratio of $C_1/C_2$ cannot be used the critically-damped condition can still be obtained by adding the resistance $R_2$ and changing the value of $L_2/CR^2$. The proper values are shown by the curves of Fig. 9.8. Curve $C_2$ of Fig. 9.7 shows the response for $C_1=C_2$ and it is not very different from the optimum condition.

The circuits of Figs. 9.2 and 9.4 can be combined into the more

**Fig. 9.5—The series-shunt corrector is really a combination of the circuits of Figs. 9.3 and 4**




complex arrangement of Fig. 9.5 with a substantial improvement in performance. The maximally-flat response is given by curve D of Fig. 9.3 and the transient response by curve D of Fig. 9.7. For a response of $-1$ db $fCR$ is 360, which is just 50 per cent better than with the circuit of Fig. 9.2 so that $R$ can be 50 per cent higher and 50 per cent more output voltage obtained for a given current. The transient response is only very slightly worse.

The circuit has the disadvantage of requiring the capacitance to be split in a particular ratio and it is more critical in requiring correct circuit values. In the critically-damped condition the response has the form of curve D of Fig. 9.7.

The component relationships for all these circuits are given in Table 9.1 and as an example of their use the values for a response of $-1$ db at 2·75 Mc/s with the circuit of Fig. 9.5 will be derived assuming $C=30$ pF.

From curve D, Fig. 9.3, $fCR=360$, so $R=360/(30\times 2{\cdot}75)=4{\cdot}36$ k$\Omega$ and $CR^2=30\times 4{\cdot}36^2=570$. From Table 9.1, $L_1=0{\cdot}0625\times 570=$

TABLE 9.1

**V.F. COUPLING RELATIONS**

| *Circuit* | $L_1$ | $L_2$ | $C_1$ | $C_2$ | $C_3$ | $R_1$ |
|---|---|---|---|---|---|---|
| *Flattest Frequency Response* | | | | | | |
| Fig. 9.2 | $0{\cdot}414CR^2$ | — | — | — | $0{\cdot}354C$ | — |
| Fig. 9.4 | — | $0{\cdot}66CR^2$ | $0{\cdot}25C$ | $0{\cdot}75C$ | — | — |
| Fig. 9.5 | $0{\cdot}143CR^2$ | $0{\cdot}583CR^2$ | $0{\cdot}404C$ | $0{\cdot}596C$ | — | — |
| *Critical Damping* | | | | | | |
| Fig. 9.2 | $0{\cdot}296CR^2$ | — | — | — | $0{\cdot}125C$ | — |
| Fig. 9.4 (1) | — | $0{\cdot}375CR^2$ | $0{\cdot}888C$ | $0{\cdot}111C$ | — | — |
| (2) | — | $0{\cdot}5CR^2$ | $0{\cdot}5C$ | $0{\cdot}5C$ | — | $R$ |
| Fig. 9.5 | $0{\cdot}0625CR^2$ | $0{\cdot}391CR^2$ | $0{\cdot}2C$ | $0{\cdot}8C$ | — | — |

35·6 $\mu$H, $L_2$=0·391×570=223 $\mu$H, $C_1$=30×0·2=6 pF, $C_2$=0·8×30=24 pF.

It is not essential to use inductance in order to extend the frequency response for it is possible to do it by means of *RC* circuits. One such arrangement is shown in Fig. 9.9 and consists essentially of a frequency-discriminating potential divider preceding the valve. Put rather crudely, the action is that as the load impedance falls at high frequencies, because of the presence of *C*, the input to the valve increases because the impedance of $C_1R_1$ falls also. The expression for amplification is

$$\frac{e_2}{e_1} = g_m R \times \frac{R_2}{R_1 + R_2} \times \frac{\sqrt{1 + \omega^2 T_1^2}}{\sqrt{(1 + \omega^2 T^2)(1 + \omega^2 T'^2)}} \qquad 9.1$$

where $T = CR$, $T_1 = C_1R_1$ and $T' = \dfrac{(C_1 + C_2) R_1 R_2}{R_1 + R_2}$

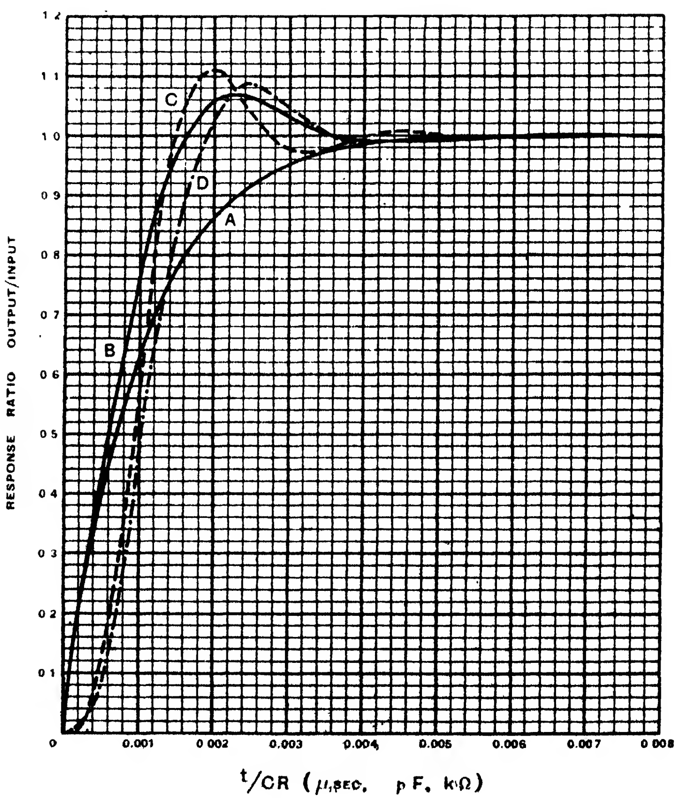



**Fig. 9.6—Transient-response curves of four circuits in which the component relations are adjusted for the flattest frequency response: A, resistance-capacitance coupling ; B, shunt-corrector ; C, series-corrector ; D, series-shunt-corrector**

Provided that $1 \gg \omega^2 T'^2$ at all frequencies required, correction is complete when $T_1 = T$ and then $\frac{e_2}{e_1} = \frac{g_m R R_2}{R_1 + R_2}$

In general, however, $\omega^2 T'^2$ is not negligible and the compensation is only approximate. The circuit is useful on occasion, but usually leads to lower amplification than inductance correcting systems and it is consequently not very widely used.

It should be noted that if $C_1 R_1 = C_2 R_2$, $T_1 = T'$ and the potential divider is independent of frequency. This is a useful result when constant attenuation is required.

So far no bias resistance has been shown in the circuits. Such a resistance is usually employed and connected in the cathode lead as shown in Fig. 9.10. The amplification as shown in the preceding curves is then multiplied by the factor

$$\frac{\sqrt{1 + \omega^2 T_c^2}}{\sqrt{(1 + g_1 R_c)^2 + \omega^2 T_c^2}} \qquad 9.2$$

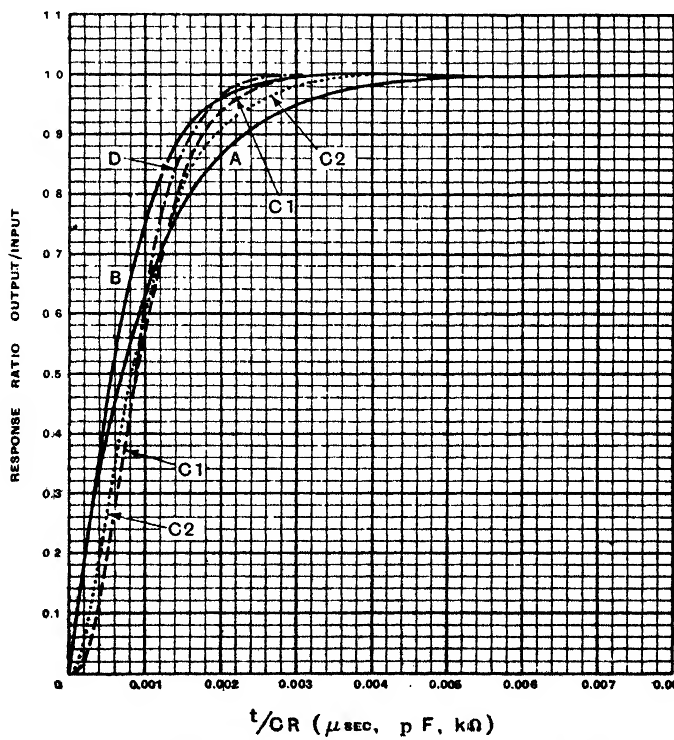



**Fig. 9.7—Transient-response curves for critical damping: A, resistance-capacitance coupling; B, shunt-corrector; C1, series-corrector with $C_1 = 8C_2$, $R_1 = \infty$; C2, series-corrector with $C_1 = C_2$, and $R_1$ finite; D, series-shunt-corrector**

where $T_c = C_cR_c$ and $g_1$ is the mutual conductance of the valve *acting as a triode with grid and anode strapped.* The formula is approximate only since it neglects several minor factors. When $\omega^2T_c^2 \gg (1 + g_1R_c)^2$ the expression reduces to unity and the stage gain is unaffected. This condition can be realized for alternating currents by making $C_c$ large enough, but at zero frequency the expression is always $1/(1 + g_1R_c)$ and the stage gain is reduced.

This corresponds to a reduction in the d.c. component of the signal and to avoid it it is customary to do without the capacitance $C_c$ and to use an unbypassed resistance $R_c$. The gain is then the figure calculated previously multiplied by $1/(1 + g_1R_c)$: there is a loss of gain but there is no frequency discrimination. In practice, $g_1$ may be about 9 mA/V and $R_c$ about 50 ohms. Consequently, the gain is reduced by a factor of the order of 0·69.

It should be noted that this circuit can also give high-frequency correction. With resistance coupling the time-constants should be about equal for the anode and cathode circuits; that is, $CR$ in Fig. 9.1, should equal $C_cR_c$ in Fig. 9.11. This usually leads to a capacitance for $C_c$ of the order of 0·001 to 0·002 $\mu$F.

It is now necessary to consider the low-frequency end of the v.f. spectrum. The vision signal proper is broken up by the sync

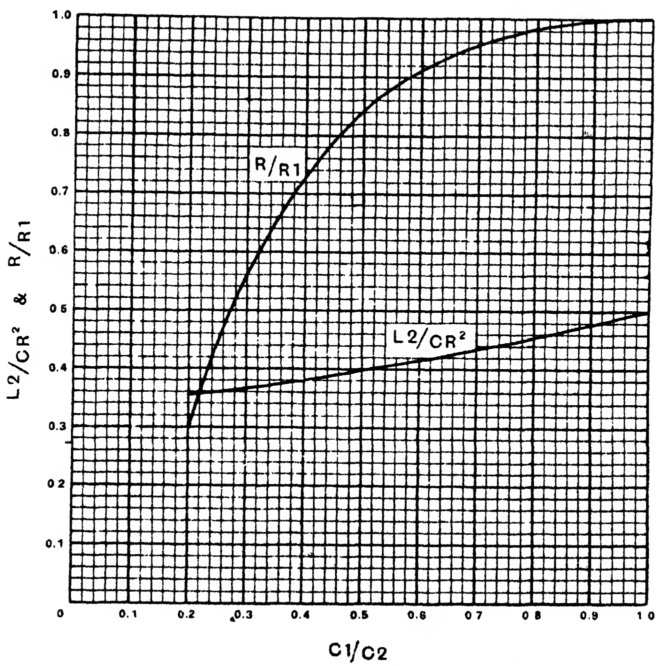



**Fig. 9.8—Component relations for critical damping in the series-corrector when $C_1$ is not 8$C_2$**

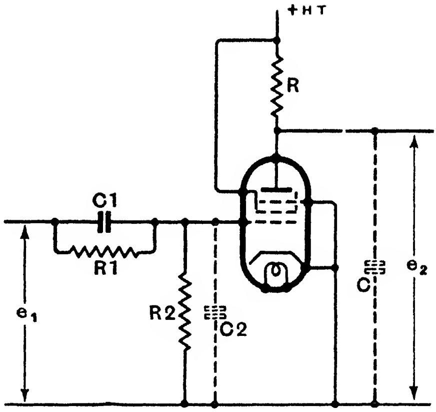



**Fig. 9.9—Correction can be obtained by means of a frequency-discriminating potential divider**

pulses of 10 $\mu$sec duration and the longest interval between sync pulses is 88·77 $\mu$sec. It is only this duration which governs the choice of time-constant in a coupling circuit *CR*, Fig. 9.11. It is shown in Appendix 3 that if the loss of voltage at the end of a pulse is restricted to 2 per cent the time-constant *CR* must be 50 times the pulse duration.

In this case, the time-constant must be 4,400 $\mu$sec approximately as a minimum; that is, $CR=0{\cdot}0043$ $\mu$F–M$\Omega$. Normal values would be $R=0{\cdot}47$ M$\Omega$ and $C=0{\cdot}01$ $\mu$F. Values of this order are sufficient in nearly all receiving cases to pass the a.c. component of the vision signal with negligible distortion.

When the signal passes through a coupling of this kind, however, the so-called d.c. component of the signal is lost. The signal waveform as it appears in the detector output with its d.c. component is shown in Fig. 9.12 (a) for a black line, a white line and a black line with a white bar in the middle. The signals are shown as negative-going on the picture signal. It is to be noticed that the

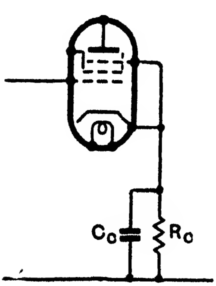



**Fig. 9.10—Occasionally correction is obtained by shunting the bias resistance by a small capacitance**

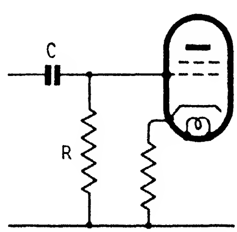



**Fig. 9.11—In a coupling circuit of this type the time-constant CR must be large enough for C not to charge appreciably on the signal of longest duration**

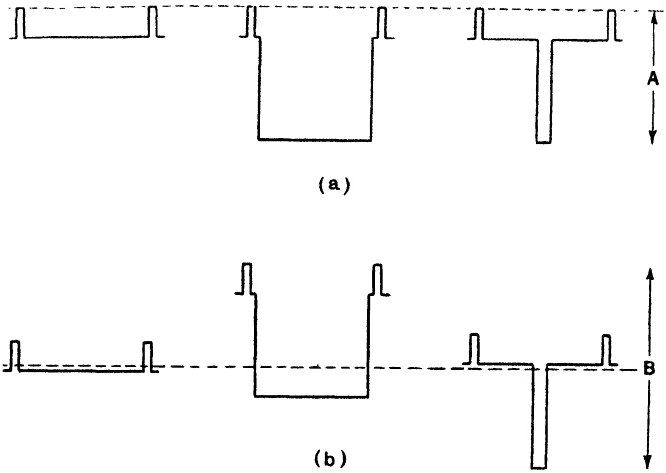



Fig. 9.12—This sketch illustrates three typical waveforms at (a), and the same signals with the d.c. component removed at (b)

tips of all the sync pulses, which correspond to zero output from the detector, all rest on the same base line and that in consequence the black level in all three lines shown is the same.

The same three lines are shown in Fig. 9.12 (b) after passing through a coupling which removes the d.c. component and in the steady state after any disturbances due to passing from one signal to another have disappeared. The waveforms settle down around the zero reference line so that they enclose equal areas on each side of it.

The two forms of signal (a) and (b) obviously will affect the c.r. tube differently. In practice, this may not be very serious for the mean brightness of normal pictures does not vary a great deal and, apart from certain shading effects which may occur, correction can be obtained by adjusting the brightness control. However, the output of the v.f. stage usually feeds the sync separator also and here it is essential for the d.c. component to be present.

Because of the presence of the sync pulses in the signal it is possible to restore the d.c. component of the signal by the very simple process of connecting a diode in shunt with the grid leak, *R* in Fig. 9.11. It must be so connected that it conducts on the sync pulses. With an input having positive-going sync pulses in Fig. 9.11 the diode is connected with its cathode earthed and its anode to the grid of the valve, whereas if the signals are negative-going on the sync pulses it is reversed and has its anode earthed and its cathode joined to the grid of the valve.

In any *RC* coupling of this type the capacitor gains charge on one portion of a wave and loses charge on the other. In the steady state the gain and loss per cycle are equal. When only *C* and *R* are present the time-constant is independent of the polarity of the signal and equilibrium demands a shift in the mean voltage level so that on the short synchronizing pulse a higher voltage is acting on the capacitor than during the long line-scan periods.

When the diode is present and conducts, its effect is to reduce the time-constant very greatly. The capacitor can then change its charge quickly during the sync pulse and so only a small voltage is required.

The action is not perfect but is amply good enough for television purposes. The circuit is more fully analysed in Appendix 4. When the d.c. restoring diode is used, the time-constant $CR$ is usually made larger than the value given in the foregoing and common values are 0·1 $\mu$F and 1 M$\Omega$.

When the signal has positive-going sync pulses on the grid of a valve the diode becomes unnecessary for the grid-cathode path of a valve can operate as a d.c. restoring diode if the bias is removed. The circuit then has the form of Fig. 9.13.

It might be thought that, where several couplings are involved in an amplifier, it would suffice to apply d.c. restoration in the final circuit where the d.c. component is actually wanted. Except when the signal level is very small, however, it usually pays to use d.c. restoration after each coupling. The reason is that if this is not done each stage must be designed for twice the signal-handling capacity. Referring to Fig. 9.12, it is clear that although the change of amplitude over a short period cannot exceed A, the limits of voltage excursion on the valve over a long period reach B and that B is nearly twice A. As the output valve must be fairly fully loaded for economy, it is obviously important to retain the d.c. component, for a smaller valve can be used when it is present than when it is absent.

There is a difference, too, between the two conditions as regards

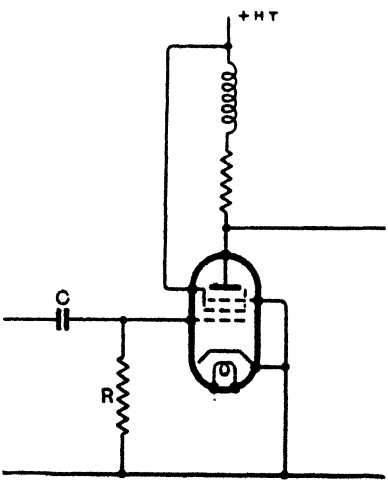



**Fig. 9.13—In some cases d.c. restoration can be secured by omitting negative grid bias**

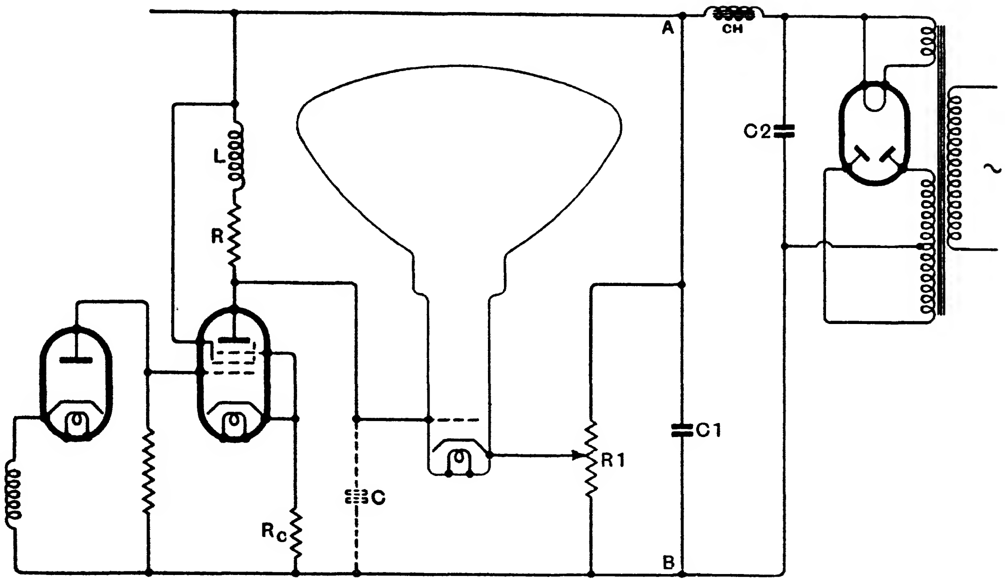


Fig. 9.14—When direct coupling is used between v.f. stage and the tube the circuit often takes this form

the bias point of the valve. Without the d.c. component, the valve must be biased to the mid-point of its grid-volts–anode-current curve just as in the case of an a.f. valve. With the d.c. component, however, the valve should be biased only to the point of grid current cut-off (that is, about − 1 V) or, if the valve itself acts as a d.c. restorer, zero bias should be used. The signal then swings the grid only negatively from this point and full use is made of the valve. Of course, if the signal is in opposite phase, the bias must be made abnormally high—about twice the value used for a.f. amplifier practice.

The modern practice is to use direct coupling between the detector and the v.f. stage and to bias the valve at about −1 V or towards current cut-off, according to whether the sync pulses in the detector output are positive- or negative-going. The picture signal itself, of course, changes in the opposite direction from the sync pulses.

A.C. couplings normally occur only after the v.f. stage, and often only in the sync separator circuits. When electric deflection is used in the c.r. tube, it is not uncommon to have the positive of the e.h.t. supply earthed. The tube cathode and grid are then several thousand volts below earth and an a.c. coupling from the v.f. stage is essential. A d.c. restoring diode is then needed and can usually be operated from the same highly-insulated heater winding as the tube.

With modern electromagnetic deflection the negative of the e.h.t. supply is invariably earthed and it becomes possible to use direct coupling between the v.f. stage and the tube. This is the usual practice and the use of d.c. restoration is often confined to the synchronizing channel.

Direct coupling is not without its problems, however. Few, if any, usually arise in the detector-v.f. stage coupling. It is in the v.f. stage-c.r. tube coupling that they occur and they are of two kinds—performance problems and tube safety problems. The first arise because perfect amplification of the d.c. component is not achieved on account of the frequency dependence of the impedance of the h.t. supply system. Consider Fig. 9.14, which shows a v.f. stage directly fed from the detector and directly coupled to the c.r. tube. The impedance of the h.t. supply measured between A and B obviously varies with frequency. At high frequencies it is only the reactance of the capacitance $C_1$, at some low frequency there is parallel resonance between $Ch$, $C_1$, and $C_2$ and the impedance is resistive and high; at lower frequencies the impedance is inductive, and at zero frequency it becomes resistive again. It is then composed of the resistances of the choke, rectifier and mains transformer.

Now the h.t. supply forms a part of the anode circuit of the v.f. valve, so that the amplification will tend to rise at the frequencies for which the h.t. supply impedance is no longer negligible. To some extent this will be counteracted by the fact that the impedance is common to both screen and anode circuits. When the h.t. supply impedance rises the voltage developed across it is applied to the screen and so gives a form of negative feedback. This particular effect is absent if the screen is fed from a voltage-stabilizer circuit, as is sometimes done.

These effects can only be avoided completely by using neon voltage stabilizers for both anode and screen supplies or by building out the h.t. supply to constant impedance. Both courses are expensive, but the latter is sometimes adopted in transmitting equipment. Fortunately, such refinements are quite unnecessary for the receiver in normal circumstances.

The variations of impedance of the h.t. supply usually occur well below 50 c/s and are of little importance provided that the impedance at zero frequency is kept to such a value that the d.c. component is amplified at the proper level. Since the resistance of the h.t. supply system is usually unknown its effect cannot be calculated. It would in any case be rather complex for it depends also on the feedback in cathode-bias and screen-feed resistors.

It is to be noted that the effect does not necessarily entail a reduction of the d.c. component of the signal, it may produce an increase. This will usually have no effect on synchronizing for the signal passes through an a.c. coupling to these circuits in many cases. It will, however, result in the changes of mean brightness of the picture becoming excessive and because of this it is said to accentuate the bad effects of interference caused by reflections from aircraft.

Turning now to the problem of tube safety the drawback to direct coupling between the anode of the v.f. stage and the grid of the tube is the risk of damage to the tube. In Fig. 9.14, the h.t. supply is likely to be about 250 V and the no-signal voltage drop across $R$ may be 75 V giving an anode potential (and c.r. tube grid potential) of + 175 V with respect to the negative line. The c.r. tube requires a negative grid bias of, perhaps, 40 V so its cathode is then returned to a point + 215 V above the negative line.

This is satisfactory during normal operation, but suppose the v.f. valve fails. A broken heater or heater lead, or a broken screen connection, will give zero anode current and hence no voltage drop across $R$. The tube grid will become 250 V positive, or $250 - 215 = +35$ V with respect to its own cathode. There is no doubt that this is extremely bad for the tube, and it may seriously damage it.

Then consider the conditions even when there is no defect and the set is switched off. The h.t. voltage falls to zero very quickly and hence the tube bias also. The tube h.t. supply falls much more slowly, however, owing to the higher time-constant of its smoothing circuits. The conditions are not ideal ones, but they are not necessarily very harmful. One would expect, however, that the tube life would be somewhat shorter than if the tube bias and h.t. were derived from a common source, or if delayed switching were used to remove the tube anode voltage before the bias.

When switching on, the receiver h.t. supply usually comes on before the tube voltage, either because delayed switching is used or because there is an indirectly-heated rectifier or, in modern sets, because it comes from the line fly-back. With a directly-heated rectifier for the receiver h.t. supply the full voltage appears almost at once and the tube grid will be positive with respect to its cathode by at least 35 V until the v.f. valve warms up and takes current. An indirectly-heated rectifier for the receiver supply is obviously indicated.

An alternative now commonly adopted is to apply the v.f. signals to the cathode of the c.r. tube instead of to its grid. The basic circuit of Fig. 9.13 still applies if the grid and cathode tube connections are interchanged and the detector diode is reversed. This last is necessary because changing the tube connections reverses the polarity of the picture and the diode must also be reversed to obtain a positive picture.

This method is a good one and is quite safe, because a failure in the v.f. valve merely results in the cathode of the tube being carried positively. The main disadvantage is that the v.f. gain is lower, because the valve must be biased nearly to cut-off in the absence of a signal—instead of only to avoid grid current—on account of the reversed polarity of the signal. This means a higher cathode bias resistor and hence more feedback across it.

One factor which must be taken into account is the permissible voltage between heater and cathode of the c.r. tube. This is commonly limited to 100 V and it is often inconvenient to use a separate heater supply for the tube. The anode of the v.f. stage is often 150–180 V above earth and so the tube cathode cannot be joined to it without exceeding the heater-cathode rating of the tube.

A way out is shown in Fig. 9.15. The tube cathode is connected to the v.f. stage through a frequency-discriminating potential divider $R_3$, $C_2$, $R_4$, and a compensating network $R_2$, $C_1$ is placed in the anode circuit of the v.f. valve. Usually $R_1$ and $R_2$ are about equal in value and $R_3$ and $R_4$ are also about equal and high in comparison. Thus $R_1$ and $R_2$ may be 3·5 kΩ and $R_3$ and $R_4$ 100 kΩ.

The mean cathode voltage of the tube becomes the mean anode voltage multiplied by $R_4/(R_3+R_4)$ and is usually about one-half the anode voltage—some 75–90 V. By properly relating the time-constants it is possible to make the response of the circuit independent of frequency but in view of the complication introduced by the screen feed this is not often attempted. It is usual to make $C_1$ and $C_2$ both large—8–16 $\mu$F—so that $R_2$ and $R_3$ are effectively short-circuited for all vision frequencies. Conditions are then much the same as with a direct connection to the anode except that at d.c. the valve load becomes $R_1$ and $R_2$, with screen feedback from $R_2$ and the gain to the anode is increased. Only a fraction of this is applied to the c.r. tube, however, and by suitably adjusting the values of $R_3$ and $R_4$ it is possible to make the gain to the tube

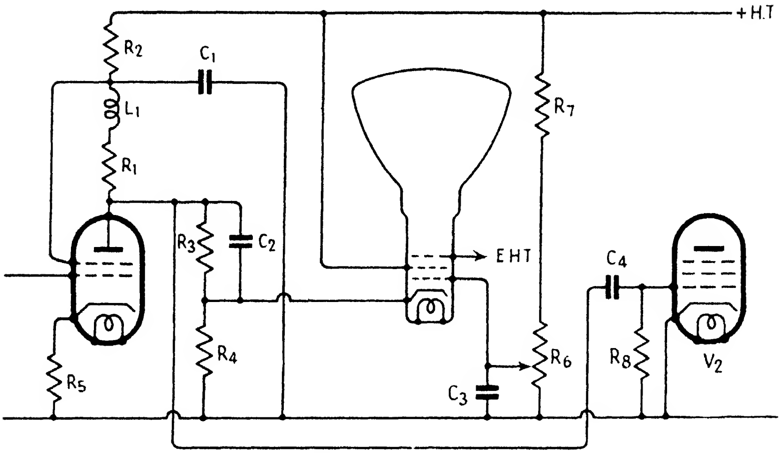



**Fig. 9.15—Direct connection between the v.f. stage $V_1$ and the cathode of the c.r. tube is shown here. The inputs required by the tube and the sync separator $V_2$ are in the same phase**

cathode equal to the gain at higher frequencies, thus overcoming the disability mentioned earlier.

In Fig. 9.15 a sync separator valve is also shown. It is a.c. coupled to the v.f. stage by $C_4$ and $R_8$ and as the sync pulses are positive-going, d.c. restoration is obtained without a diode by the grid-cathode path of the valve.

The type of valve used for the v.f. stage depends very largely upon the c.r. tube employed. Some types need an input of about 20 V p–p only for full modulation whereas others need as much as 60 V p–p. As pointed out earlier the output of the v.f. valve must be greater than these figures because they refer only to the voltage changes between black and white, but the valve has also to handle the sync pulses. The total output of the v.f. valve for an average tube, therefore, must be about 30 V p–p and for an insensitive tube close on 90 V p–p.

The valve used in the v.f. stage must be able to provide a change of anode current (mA) equal to the required total output voltage divided by the load resistance (kΩ). The peak anode current must be about 20 per cent greater because the valve must pass a reasonable current when its grid is at its most negative value if it is to function with reasonable linearity. For an output of 30 V, therefore, and a resistance of 3 kΩ, a current change of 10 mA is needed. Adding 20 per cent, the peak current must be 12 mA.

If the valve feeds the grid of the cathode-ray tube this must be the no-signal anode current and the valve will need a small bias of perhaps −1 V only. If it feeds the cathode of the tube, this represents the peak-white current, and with no signal the valve must be biased to pass, say, 2 mA. This demands quite a large cathode resistor and as a result the stage gain will be low on account of the feedback from it. This can be reduced by using a lower value resistor and bleeding extra current through it by connecting an extra resistor between + h.t. and cathode.

The type of valve used is usually a high-$g_m$ r.f. pentode. Most such valves are rated for operation at cathode currents up to 15 mA and are eminently suitable. Cases do arise, however, where something bigger is needed, principally when the tube needs an unusually large input. Thus, if the tube wants a 90-V input, the peak current must be 36 mA if the coupling resistor stays at 3 kΩ. A suitable valve for this case is the EF55, but failing this an output pentode may have to be used. This may well have higher capacitances and so require a reduction in the value of the coupling resistor which, in turn, will make a higher current necessary.

The foregoing method of estimating the requirements for the valve is a rough and ready one, but is very useful as a preliminary guide. Having selected a suitable valve, its operating conditions can be determined accurately by drawing the load line on the anode-volts–anode-current curves in the usual way. Owing to the very low value of load resistance usually adopted, however, this is by no means essential and it is possible to work with good accuracy from a single grid-volts—anode-current curve for the screen voltage to be used.

The accuracy obtainable depends upon the a.c. resistance of the valve over the working range and is high if the anode current is substantially independent of anode voltage. The procedure is to take the valve curve and to determine from it the maximum change of anode current obtainable with reasonable linearity.

The required output voltage is known by the tube which will be used, so that the minimum value of coupling resistance can be determined straight away. If this resistance, in relation to the circuit capacitances, enables the required frequency response to be obtained all is well, but if it does not, then a larger valve will have to be used.

Curvature of the valve characteristic tends normally to reduce the contrast range in white parts of the picture when the signal is fed to the grid of the c.r. tube. When the signal is fed to the cathode, however, it usually has little effect on the picture because the sync pulses fall on the bottom bend of the characteristic. The only effect, therefore, is to reduce the relative amplitude of the sync pulses in the output. This is a point to watch because it makes good sync separation more difficult to obtain.

This is affected also by the v.f. stage gain, for the detector is not linear and tends to reduce the effective amplitude of the sync pulses below the proper 30 per cent level. It does this because the rectification efficiency falls off at small inputs. The detector must not be operated at too small an input, therefore, and it will be if the v.f. stage gives too much gain. An amplification of around ten times is usually satisfactory, but on theoretical grounds even lower

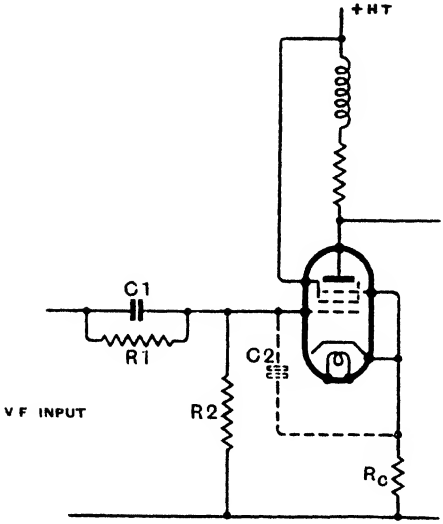



Fig. 9.16—This diagram illustrates the use of an input potential divider to a v.f. stage

gain is sometimes desirable. The attainment of this presents difficulties, because pentodes of low mutual conductance do not usually take sufficient anode current to give the required output.

One way out is to use an ordinary v.f. stage with a gain of eight to twelve times and apply only a fraction of the detector output to its grid. This can be done by the potential divider shown in Fig. 9.16. The proportion of the detector output applied to the valve is $R_2/(R_1+R_2)$.

The capacitor $C_1$ is necessary to correct for the stray capacitance across $R_2$. The potentiometer ratio is independent of frequency if $C_1R_1 = C_2R_2$ provided that $C_2$ is in parallel with $R_2$. In practice, it is not, for the cathode-bias resistance $R_c$ introduces complications. The effective value of $C_2$ is thus somewhat less than the grid-cathode capacitance of the valve. It is necessary here to distinguish between the grid-cathode capacitance and the grid-earth capacitance usually quoted; the latter is the sum of the capacitances between grid and cathode, heater, screen, and suppressor.

An advantage of this circuit is that by increasing $C_1$, the high-frequency response can be improved and may correct for a deficiency elsewhere. The circuit is basically the same as that of Fig. 9.9.

Sync separation sometimes complicates the v.f. stage. Occasionally the separator is fed directly from the detector, but the signal level at this point is usually too small. When the signal is fed to the cathode of the c.r. tube there is usually no difficulty for the signal is in the right phase for feeding to the separator and the very simple arrangement of Fig. 9.15 can be employed. This circuit, or something akin to it, is usually the best for its performance is as good as the alternatives and it is the simplest.

Of the alternatives one which was considerably used in the past

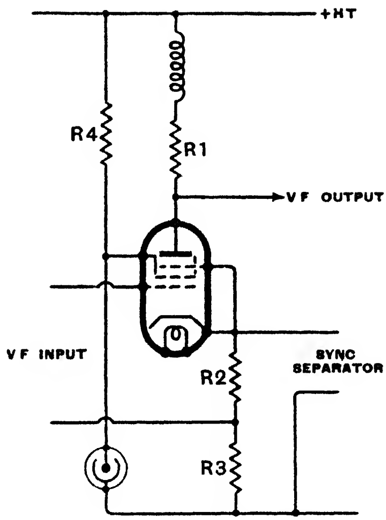



**Fig. 9.17—It is not uncommon to feed the sync separator from the cathode of the v.f. stage**

is shown in Fig. 9.17. It is intended for grid feed to the tube and enables direct coupling to the sync separator to be employed.

The output to the tube is taken from the anode circuit in the usual way. The output to the sync separator, however, is taken from the cathode circuit and is developed across $R_2$ and $R_3$. The input is applied as shown and it is important to note that no part of the detector circuit can be earthed without short-circuiting $R_3$. This means that the whole detector circuit including its input circuit is floating at cathode potential. As a result, it is somewhat more difficult to maintain r.f. stability; the use of a small capacitance across $R_3$ is helpful from this point of view, but it must be quite small if the shape of the sync pulses is to be preserved. The output at the cathode is in the same phase as the input to the grid and in opposite phase to the output at the anode.

For use with the Mazda SP42, the makers recommend a screen-cathode potential of 95 V with $-$ 1·0 V grid bias. The no-signal anode and screen currents are 16·0 mA and 4·0 mA respectively. The resistances should be:—$R_1 = 4{,}000\ \Omega$; $R_2 = 50\ \Omega$; and $R_3 = 1{,}000\ \Omega$. If greater cathode output is required $R_2$ should be 67 Ω and $R_3 = 2{,}000\ \Omega$; the screen-cathode potential should be raised to 100 V.

A stabilized screen supply is essential if direct coupling to the sync separator is used. It stabilizes both anode and screen currents and hence the cathode potential, because the anode current is substantially independent of anode potential with a tetrode or pentode.

As the screen current flows through the cathode resistances the action as an amplifier between grid and cathode is quite different

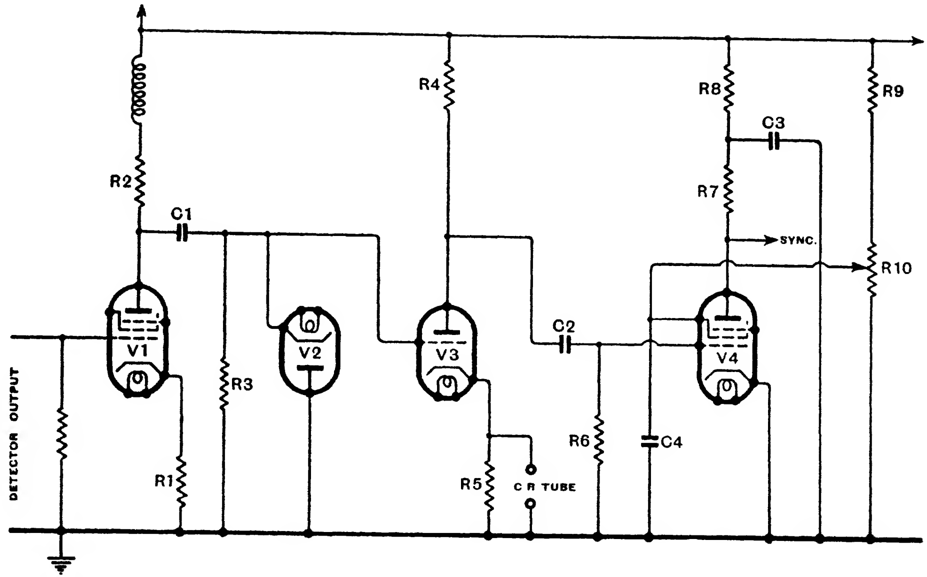



Fig. 9.18—A very satisfactory output system is shown here. $V_1$ is a v.f. stage, $V_2$ a d.c. restorer, $V_3$ a phase-splitter, and $V_4$ the sync separator

from that as an amplifier between grid and anode. Viewed between the grid and cathode circuits the valve behaves substantially as a triode.

One point that is often overlooked is that there is feedback through the grid-to-screen-grid interelectrode capacitance. At first this may seem absurd, for the screen is maintained at a fixed potential relative to earth by the stabilizer. It must not be forgotten, however, that the cathode potential is fluctuating with respect to earth and from the point of view of feedback through the grid-screen capacitance this is the same as if the cathode potential were fixed and the screen potential fluctuating. Owing to the feedback, the effective input capacitance is likely to be considerably higher than one expects.

Sync separation is dealt with in detail in Chapter 15 and it will be clear that the larger the vision signal which is applied to the separator the easier it is to remove the picture content. For this reason it is desirable to apply the full output of the v.f. stage to the separator rather than a smaller signal derived from some other part of the equipment.

The most perfect separator is a valve working as anode-bend rectifier supplied with a vision signal of negative phase (sync pulses positive). This phase is opposite to that needed by the grid of the c.r. tube so that when the signal must be applied to the grid it often pays to include a unity-gain phase-reversing stage. This can take the form of a triode arranged like a phase-splitter for a push-pull amplifier, giving equal outputs in opposite phase at anode and cathode.

The stage is shown in Fig. 9.18 together with the v.f. and sync separator valves. $V_1$ is the v.f. stage with the usual corrected coupling. $V_2$ is the d.c. restoring diode and $C_1$ and $R_3$ can be 0·1 $\mu$F and 1 M$\Omega$ respectively. If desired, the c.r. tube can be connected across $R_3$.

The output of $V_1$ is applied to the phase-splitter $V_3$ which can have $R_4$ and $R_5$ of about 7,500 $\Omega$ each. It is not, however, essential for these resistances to be equal, but when they are, the output of $V_3$ at the anode is equal to that at the cathode.

The initial bias of $V_3$ is provided by the voltage drop across $R_5$ and is so high that the valve is near the current cut-off point. The signal, however, always drives the grid in a positive direction so that this condition is correct.

There is heavy negative feedback along $R_5$ and the valve characteristics consequently become exceedingly linear and the valve will handle an input of 40–50 V p–p easily. The output at the anode is equal to that at the cathode, but of opposite phase, and is about 80–90 per cent of the input.

The output at the anode is fed to the sync separator $V_4$, d.c restoration being effected by the grid-cathode circuit of this valve; $C_2$ and $R_6$ can well be 0·1 $\mu$F and 1 M$\Omega$ respectively. Instead of feeding the c.r. tube from the input to $V_3$ it can be fed from the

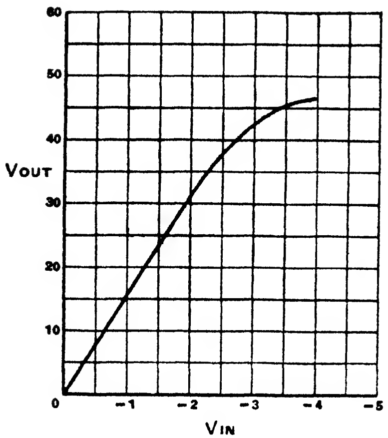



**Fig. 9.19—The curve above shows the amplitude characteristic of the circuit of Fig. 9.18**

cathode output of this valve, the tube grid being connected directly to the valve cathode.

This relieves the coupling between $V_1$ and $V_3$ of the input capacitance of the tube and so permits the use of a somewhat higher value for $R_2$. Roughly speaking, the consequent increase in the gain of $V_1$ compensates for the slight loss in $V_3$.

It should be noted that while it may be considered dangerous to tube life to connect its grid directly to the anode of a valve no such objection arises to the cathode connection. It is clear that if the h.t. supply to the receiver fails or the heater of $V_3$ breaks, the anode current of this valve ceases and the cathode falls to earth potential. In other words, the change of tube grid potential is negative, not positive.

The performance of the circuit of Fig. 9.18 is illustrated by the curves of Figs. 9.19 and 20. For the measurements the following circuit values were used:—$R_1 = 37{\cdot}5\ \Omega$; $R_2 = 3{,}200\ \Omega$; $R_3 = R_6 = 1\ \text{M}\Omega$; $R_4 = R_5 = 7{,}500\ \Omega$; $C_1 = C_2 = 0{\cdot}1\ \mu\text{F}$; $V_1$ = SP41; $V_2$ = D1; $V_3$ = 354V; $V_4$ = SP42.

The relation between the input to $V_1$ and the anode or cathode output of $V_3$ is shown in Fig. 9.19. An output of 45 V p–p is obtainable for some 10 per cent second-harmonic distortion. This is for 250 V h.t. supply and a no-signal current consumption for $V_1$ and $V_3$ of 24 mA. The voltage amplification is 19·5 times.

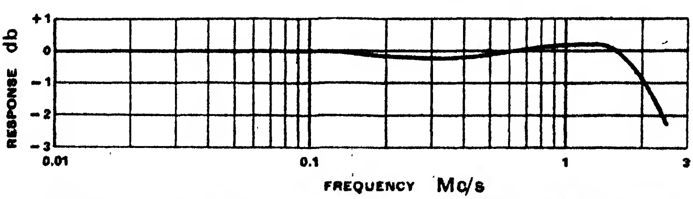



**Fig. 9.20—The overall frequency response is shown by the curve above**

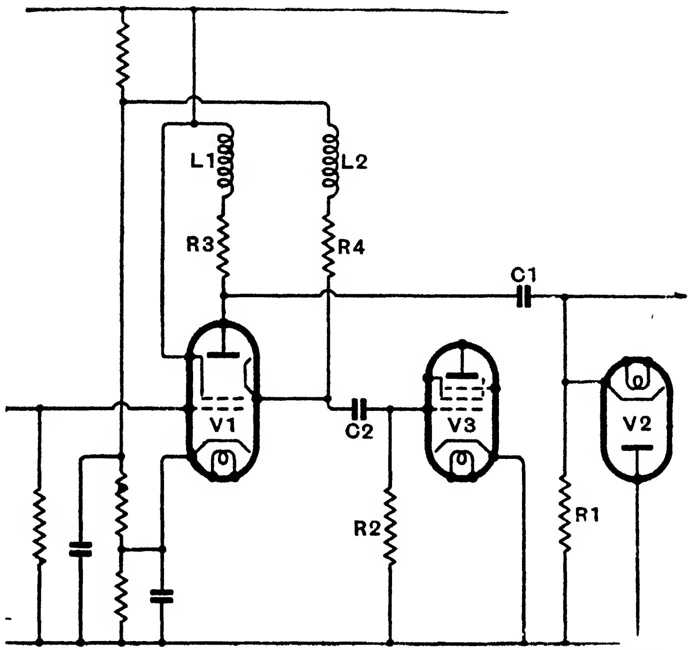



**Fig. 9.21—When using a secondary-emission valve the sync separator can be fed from the auxiliary cathode**

Fig. 9.20 shows the frequency response, the correcting inductance being 144·5 $\mu$H. At 2·5 Mc/s the response is $-2{\cdot}3$ db.

A somewhat similar result can be achieved without the need for the phase-splitter by employing an electron-multiplier valve such as the Mullard TSE4 or EE50 for the v.f. stage. Rather higher gain results because the mutual conductance is about 14 mA/V, and an output of the same phase as the input can be taken from the auxiliary cathode. These valves are now obsolete, however, and at the time of writing there is no equivalent valve on the market.

The arrangement is shown in Fig. 9.19, where $V_1$ is the v.f. stage. The output at the anode is in positive phase and is fed to the c.r. tube through $C_1$ and $R_1$ with the d.c. restoring diode $V_2$. The sync separator is fed from the auxiliary cathode in negative phase through $C_2R_2$. A correction circuit in the auxiliary-cathode lead is necessary to obtain the correct frequency response in the anode circuit; hence the inclusion of $L_2$.

CHAPTER 10

# Radio-frequency Amplification

IN THE STRAIGHT SET all pre-detector amplification is carried out at the frequency of the incoming signal, but in the superheterodyne the bulk of the amplification is effected at a different frequency, which is usually lower than that of the incoming signal and which is termed the intermediate frequency. This intermediate frequency is produced by the frequency-changer stage and it is usual to precede it by some amplification and this is, of course, at the signal frequency.

The superheterodyne, therefore, generally has amplification at both signal and intermediate frequencies, whereas the straight set has amplification only at the signal frequency. Both signal and intermediate frequencies lie within the radio-frequency spectrum and both are properly called radio frequencies. This term is sometimes used as a synonym for signal frequency, but this implies that the intermediate frequency is not a radio frequency and this is manifestly untrue.

In actual fact, the only differences between a signal- and an intermediate-frequency amplifier are minor ones brought about by the different mid-band frequencies. It is perfectly possible to use an i.f. amplifier as an s.f. amplifier and vice versa, if the mid-band frequencies happen to be right for the particular application.

In this chapter radio-frequency amplification is discussed and it should be clear that, unless specifically stated to the contrary, everything said applies equally to both s.f. and i.f. amplifiers. The points in which the two differ and matters peculiar to either are discussed in later chapters.

In essence, the pre-detector circuits of a television receiver do not differ from those of an ordinary broadcast set, but in detail they are quite different. The main reason for this is the enormously greater bandwidth needed; instead of being 8–20 kc/s, it is 2–6 Mc/s. Such a bandwidth can only be obtained with low amplification per stage, so that instead of having one or two r.f. valves only there are usually three to five r.f. valves in a television receiver.

There are many different arrangements possible for the r.f. amplifier intervalve couplings, such as tuned anode, staggered tuned anode, and coupled tuned circuits. Each of these has its own advantages and disadvantages and all are to be found in present-day commercial equipment.

The tuned anode circuit is the simplest and as applied to a television r.f. amplifier it takes one of the forms shown in Fig. 10.1. Strictly speaking, only (b) is a tuned anode circuit and (a) is a tuned grid. However, when $C_1$ is large their performance is the same. No

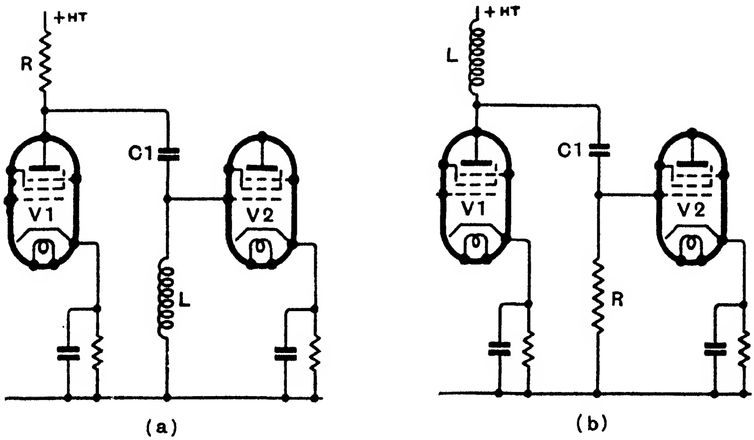



**Fig. 10.1—A simple tuned grid intervalve coupling is shown at (a) and the tuned anode circuit at (b)**

tuning capacitance is shown because in practice the valve and stray circuit capacitances are employed. When $C_1$ is large enough it can be ignored and the equivalent circuit of one stage drawn as shown in Fig. 10.2; it is assumed here that the series r.f. resistance of the coil can be neglected. At resonance the impedance of the load circuit is equal to $R$, and as long as $R$ is much smaller than the valve resistance $r_a$, the stage gain is

$$A = g_m R \qquad 10.1$$

where $g_m$ is the mutual conductance of the valve (mA/V) and $R$ is in kilohms.

It can be shown that circuit capacitance and resistance are closely related to the bandwidth. Calling the resonance frequency $f_r$ there are two frequencies, one higher $f_2$ than resonance and the other lower $f_1$, at which the response is the same but less than at $f_r$. These frequencies are such that $f_r = \sqrt{f_1 f_2}$. We call $f_2 - f_1 = n$ the bandwidth.

The ratio of response at $f_r$ to that at $f_1$ or $f_2$ we denote by $S_n$. Then

$$nCR = 159\sqrt{(S_n^2 - 1)} \qquad 10.2$$

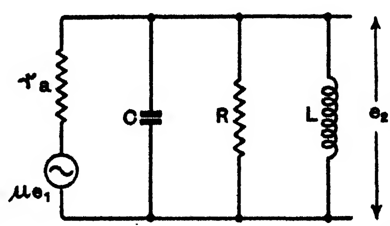



**Fig. 10.2—If $C_1$ is large this circuit is equivalent to both those of Fig. 10.1**

Combining (1) and (2),

$$A = \frac{159 g_m \sqrt{(S_n^2 - 1)}}{nC} \qquad 10.3$$

(The units are Mc/s, kΩ, mA/V, and pF).

It is clear that for a given value of $S_n$ the stage gain is inversely proportional to the bandwidth and to the capacitance $C$, and is directly proportional to the mutual conductance $g_m$ of the valve.

The importance of the various factors can be most clearly realized by an example. For a drop in response of 1 db, $S_n = 1{\cdot}122$, and if the bandwidth is 3 Mc/s, $CR = 27$ pF–kΩ.

The minimum value of capacitance is made up of the output capacitance of $V_1$ (Fig. 10.1), plus the input capacitance of $V_2$ plus the self-capacitance of the coil plus stray wiring capacitances. The coil capacitance is usually about 5 pF and it is wise to allow as much again for the wiring capacitances; this gives 10 pF, to which must be added the valve capacitances.

A typical valve which is eminently suited to broadcast sound receivers is the 6J7. It has $g_m = 1{\cdot}225$ mA/V and input and output capacitances of 7 pF and 12 pF respectively. The value of $C$ is thus $10 + 7 + 12 = 29$ pF; consequently, $R = 27/29 = 0{\cdot}932$ kΩ. The stage gain is $0{\cdot}932 \times 1{\cdot}225 = 1{\cdot}14$ times.

The stage does amplify, but only just. Practically speaking it is useless. Better results can be obtained with "Acorns"; the Osram ZA1, for instance, has capacitances totalling only 6 pF and a mutual conductance of 1·1 mA/V. With this type of valve a better circuit layout can be obtained and the stray capacitances reduced to about 3 pF, while with great care it is possible to reduce the coil capacitance also to 3 pF. This gives $C = 12$ pF and $R = 2{\cdot}25$ kΩ, so $A = 2{\cdot}47$. This is a distinct improvement but is again impracticable. To obtain a gain of 5,000–6,000 times nine or ten stages would be needed and the cost, especially with Acorn valves, would be prohibitive.

It is because reasonable gain cannot be secured with ordinary valves that special types have been developed for television purposes. The Mazda SP41, for instance, has $g_m = 8{\cdot}5$ mA/V with capacitances totalling about 20 pF. This gives a total circuit capacitance of 30 pF, so that $R$ is 900 Ω and the gain is 7·62 times. This is much better, but even now the gain cannot be said to be at all high.

If an amplifier is built using tuned anode couplings of this type and all circuits are tuned to the same frequency in the usual manner, the overall resonance curve is the product of the resonance curves of the individual circuits, assuming feedback effects to be negligible.

If the overall bandwidth is constant, therefore, the bandwidths of the individual stages must be increased as the number of stages increases and consequently the stage gain decreases. As a result, the total gain at first rises as the number of stages is increased, reaches a maximum, and then *decreases*. For a given overall bandwidth, and with given valves and circuit capacitances, there is a maximum value of overall amplification which cannot be exceeded. The effect is not confined to single-circuit couplings but occurs in any amplifier in which the stage bandwidth must be increased with the number of stages. The practical importance of the effect depends upon the magnitude of the gain of the basic single stage and of the total gain required. With the values normally encountered in television practice the limit is rarely reached and the effect means only that one or two more stages are needed than with alternative arrangements. As an example, suppose that an amplification of 10,000 times (80 db) is required with a 3-db bandwidth of 3 Mc/s and that $g_m=6$ mA/V and $C=20$ pF. From Equ. (10.3) the basic stage gain is 15·9 times or 24 db. If this could be maintained in all stages 80/24 stages would be required. As fractional stages cannot be used, this means four stages and a gain of 96 db or three and a gain of 72 db.

By trying a few values in Equ. (10.3) it is soon found that five stages are needed for a gain of 78·7 db. The method is as follows:—For an overall response of −3 db at the edges of the pass-band the response per circuit with five stages must be $-3/5=-0{\cdot}6$ db. Therefore, $s_n^2=1{\cdot}148$ and the gain per stage is $15{\cdot}9\sqrt{0{\cdot}148}=6{\cdot}12$ or 15·74 db. Hence, five stages give 78·7 db.

This drawback of a falling stage gain with an increasing number of stages can be completely overcome by a procedure known as stagger tuning. When it is fully applied no two circuits are tuned to the same frequency but the individual resonance frequencies are grouped about the mid-band frequency.

With an odd number of circuits one, and one only, is tuned to the geometrical mid-band frequency. With an even number of circuits none is tuned to this frequency.

When the scheme is fully applied the overall amplification for a 3-db bandwidth is simply

$$A_T=\frac{159g_m m}{nC} \qquad 10.4$$

where $m$ is the number of stages and assuming that all stages have the same $g_m$ and $C$ values.

In practice it is rare for stagger tuning to be fully applied. Usually an amplifier is, as it were, divided into a number of sub-amplifiers each comprising two or three stages forming a stagger-tuned group. Thus, if there is a total of six tuned circuits in an amplifier for which the required band-edge loss is 3 db it might well be built up from three similar pairs of stagger-tuned circuits, each pair being designed for a 1-db loss. Alternatively it could be built from two groups of three stagger-tuned circuits each being designed for a 1·5-db loss.

The reason for such grouping is a practical one; as the number of circuits in a group goes up so does the number of different resonance frequencies and, consequently, the number of different frequencies which must be provided in the signal generator used

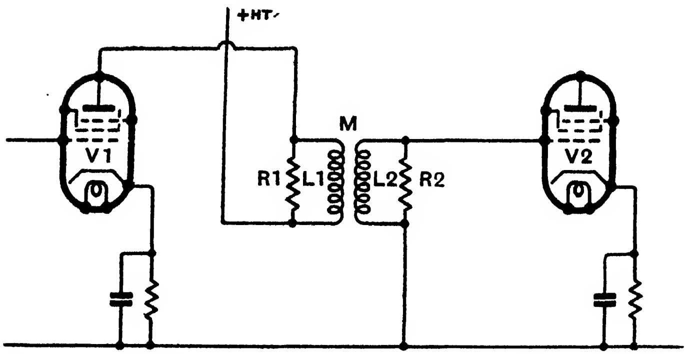



**Fig. 10.3—Coupled circuits are often used for the intervalve coupling and form a band-pass filter**

for alignment. The number of frequencies is the same as the number of tuned circuits in a stagger-tuned group.

Thus, with six circuits with full stagger tuning six alignment frequencies are needed, but with six circuits arranged as two three-circuit groups only three are wanted. The number drops to only two if there are three two-circuit groups.

Design equations for groups of up to four circuits with optimum stagger are given in Appendix 6. One characteristic has already been mentioned but it is so important that it deserves especial emphasis. It is that the gain per stage for a given bandwidth is independent of the number of stages.* With all other forms of coupling the gain per stage drops as the number of stages increases if the overall bandwidth is kept constant.

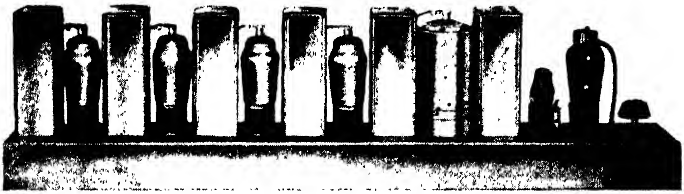
This being so, the average amplification per stage of an optimum stagger amplifier is given by Equ. (10.3). It should be pointed out that it is an average gain and not necessarily that of any individual stage of an actual amplifier.

Nowadays stagger tuning is widely used, but it is not the only coupling adopted. Still higher gain per stage—at least for a limited number of stages—can be obtained with coupled circuits of the band-pass type as shown in Fig. 10.3. Whether or not this higher gain is actually realized depends on quite a number of factors, and it can easily be less than with stagger tuning. As usual, it is important to keep circuit capacitance at a minimum and the optimum condition is with the damping applied to one circuit only, usually the secondary. This means that in Fig. 10.3, $R_1$ is not used, only $R_2$.

This method of damping is rarely used, however, for although it gives the highest gain, the circuit becomes very difficult to align, and it goes out of alignment easily through the effects of vibration and humidity.

The most reliable results are secured with the damping and

* " Performance of Coupled and Staggered Circuits in Wide-Band Amplifiers," by D. Weighton *Wireless Engineer*, October 1946, Vol. 21, p. 468.



**An experimental amplifier terminating on the right with a detector and sync separator**

coupling so adjusted that each stage has individually a flat response, and design data for this condition are given in Appendix 6. From the equations given there it can be seen that the amplification can be written in the form $A = g_m CR/2\sqrt{(C_1 C_2)}$. For a given total capacitance $C_1 + C_2$, the product $C_1 C_2$ is a maximum when $C_1 = C_2$ and hence the amplification is a minimum. It follows that it is most advantageous to use coupled circuits when there is a large difference between the primary and secondary capacitances. The coupling to a diode detector is a common example.

The relative advantages of the three forms of coupling depend also upon the drop in response permissible at the edges of the pass-band. It is not safe to conclude that coupled circuits are always the best. In general, the total capacitance is higher with coupled circuits because of the self-capacitance of the extra coil. In addition, it is difficult to avoid the use of capacitance trimmers which still further increase the total. The net result is that in many practical cases stagger tuning results in higher amplification.

With single-circuit couplings it is fairly easy to arrange a variable inductance—the coil can be constructed as a variometer or, more usually, a form of spade or iron-dust core tuning can be adopted. It is less easy to do this with coils coupled by mutual inductance, however, for adjustments to the inductance of one coil are liable to affect the inductance of the other as well as the coupling of the two. Such a variable inductance need increase the circuit capacitance by very little.

From the point of view of selectivity the performance of the three methods of coupling is well illustrated by the curves of Fig. 10.4, which are for an r.f. amplifier with a mid-band frequency of 13 Mc/s. Each curve is for two stages and A represents the case of resonance-tuned circuits and B stagger tuning; C is for coupled pairs of circuits. This last curve shows by far the highest selectivity because there are double the number of tuned circuits. The advantage of stagger tuning over resonance tuning is well brought out, however. If matters are arranged so that the sound intermediate frequency falls at 9·5 Mc/s, the resonance tuning gives an attenuation of 12·4 db only for each pair of circuits, while stagger tuning gives 17·5 db.

If the sound intermediate frequency is at the higher of the two possible frequencies, that is, 16·5 Mc/s, neither system gives as much attenuation, but stagger tuning is still the better. In general, however, it is not essential to provide sufficient selectivity for the avoidance of interference from the sound signal in the i.f. couplings themselves. If necessary, other methods can be adopted for avoiding interference, such as a wave-trap in the r.f. circuits tuned to the sound intermediate frequency.

Radio-frequency amplifiers used in television have mid-band frequencies ranging from about 6 Mc/s to 70 Mc/s. Above 40 Mc/s an r.f. amplifier becomes a signal-frequency amplifier but below it it must be an i.f. amplifier. For reasons which are given later a common choice of intermediate frequency is 13 Mc/s, so that television r.f. amplifiers are usually at either 13 Mc/s or at a signal frequency of 45 Mc/s or more.

The bandwidth and gain depend on very many factors. Modulation frequencies up to 2·75 Mc/s are radiated so that the ideal bandwidth for double-sideband reception is 5·5 Mc/s. At short ranges where interference and noise are not important this is quite easy to achieve, but even then it is only worth while if the rest of the equipment is up to the same standard. In particular the size of

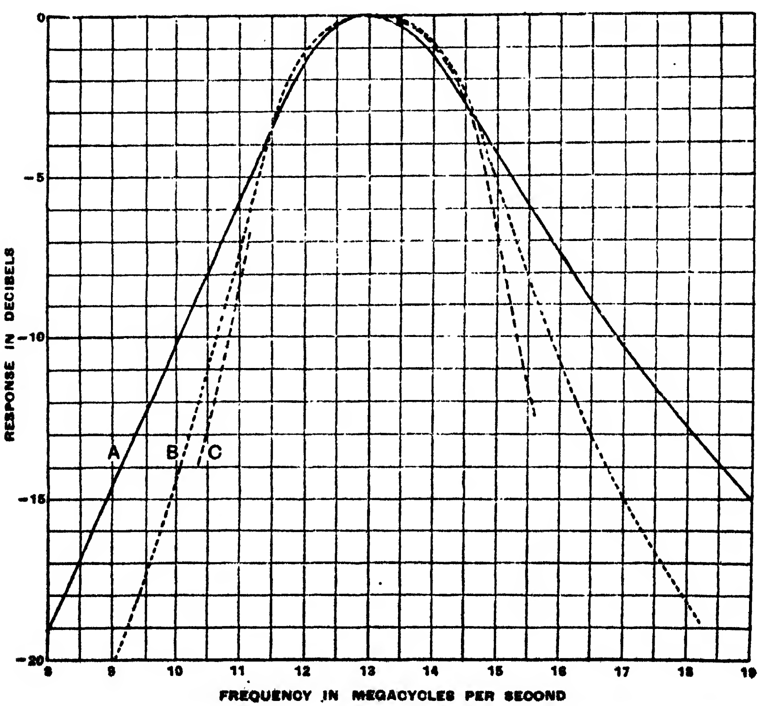



**Fig. 10.4—Curve A shows the performance obtainable with tuned anode coupling, curve B with staggered tuned anode, and curve C with coupled pairs of circuits**

the scanning spot on the c.r. tube must be small enough to utilize a frequency of 2·75 Mc/s.

At longer ranges where noise and interference become important a narrower bandwidth may, by reducing them, lead to a better picture. Then there is the question of cost, and commercially this is very important. Probably most double-sideband receivers have a bandwidth of little more than 4 Mc/s.

The gain needed depends on many factors. Speaking very roughly the minimum signal on the first grid that need be catered for is one of about 0·5 mV. If the detector needs an input of 5 V, then an amplification of 10,000 times is required. However, if v.f. amplification is not used, something nearer ten times this amplification will be needed.

In a superheterodyne the gain is split between the s.f. and i.f. amplifiers and there is little difficulty in obtaining stability. With a straight set the whole gain is obtained at one frequency and stability problems are more serious. Even then there is not usually any serious trouble.

The author has found it particularly satisfactory to arrange the r.f. valves in line with the couplings between them. A metal chassis is used and each coupling is contained in a screening can which is slightly larger and taller than the valves. With the alternation of valves and cans in a straight line, the cans automatically provide screening between the valves and nothing more is usually needed. Matters are often greatly helped by the use of single-ended valves of the EF50 type, since shorter leads are obtainable.

Direct pick-up of short-wave signals on the grids of the valves is, of course, possible, but is unusual. Should this trouble occur, valve screens can be fitted, taking care that the capacitance to the grid circuit is not appreciably increased, or the whole amplifier can be provided with a metal box.

A layout of this type lends itself well to the attainment of short leads. It is sometimes advisable to return all the by-pass capacitors associated with a valve to one *point* on the chassis to which all earthing electrodes such as the suppressor grid and metallizing are taken separately. This is clearly shown in Fig. 10.5. With tuned-anode coupling it is sometimes a little difficult to know whether to return the coil to the earth point of the valve which precedes it or which follows it. The coil really belongs to both and the only solution would be to use one point only for every earthing lead in the amplifier.

This is impracticable and it usually suffices to return the coil to the earthing point of the valve which follows it. No such point arises with two-circuit intervalve couplings, of course, for the primary is considered as belonging to the first valve of a pair and the secondary to the second.

The same effect can be obtained with single-circuit couplings by using a very tightly coupled double-wound transformer. A coupling like Fig. 10.1 would then have a *circuit* like that of Fig. 10.3 with

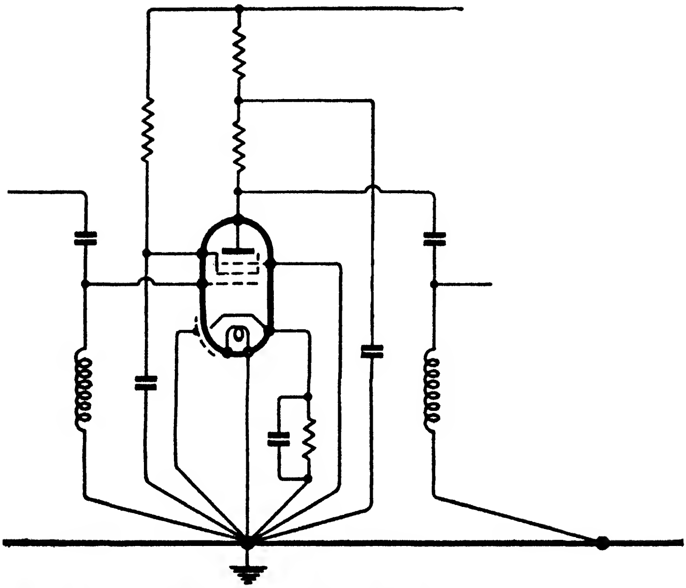


**Fig. 10.5—This diagram illustrates the arrangement of the wiring for a single-point earth connection**

$R_1$ removed, but electrically it would behave like the single coil. Both windings have the same number of turns and one is wound as closely as possible over the other, with only the necessary insulation for the h.t. supply between them.

This is often the best way of putting single-circuit couplings into practical effect, for the ability to return the grid and anode circuits to their respective valve cathodes helps considerably in achieving stability. There is one danger, however, for unless the coupling between the windings is very tight there may be a spurious response at a higher frequency.

At frequencies of the order of 13 Mc/s decoupling and by-pass capacitors of 0·01 $\mu$F capacitance are large enough. The mica-dielectric type are the best but rather expensive. Paper capacitors can usually be employed without trouble; however, if exceptionally high gain is wanted decoupling becomes more important and they may not then be good enough. At 45 Mc/s, a capacitance of 100 pF is enough. Paper-type capacitors can be used, but the mica-type are usually better.

If the r.f. valves need the same screen and anode supply voltages, as is often the case, separate decoupling is unnecessary and a common capacitance can be used as in Fig. 10.6. This not only saves components, but is convenient since the anode resistor can lie across the anode and screen sockets of the valveholder, where it is readily accessible and yet anchored securely.

The heater wiring is a prolific source of feedback troubles. It is unsatisfactory to run the heater wiring in the usual way and have only the centre tap on the mains transformer secondary earthed. With such wiring it is necessary to connect a capacitor to earth from each valve heater. It is better, cheaper, and simpler to earth one side of each heater directly to the chassis. Of course, if the transformer has a centre tap this must be left unconnected or one-half of the transformer winding will be short-circuited. Even when this method of wiring is adopted, it is sometimes necessary to connect a capacitor across each valve heater and sometimes small chokes in the heater leads are needed.

So far, the greatest stress in the r.f. amplifier has been laid upon response. Frequency response is important because if it is inadequate good definition will not be secured in the final picture. Phase distortion is also important because it represents a time-delay. In general, however, it does not enter greatly into the calculations when the resonance curve is smooth and reasonably symmetrical, and double-sideband working is adopted. This is because the phase shift on one side of resonance, which affects one set of sidebands, is positive while that on the other side, which affects the other set of sidebands, is negative. The two shifts in opposite directions tend to cancel one another. This is not the case with single-sideband operation; with this, phase shift may be very important.

Reverting to the double-sideband system, however, a peaky resonance curve will cause distortion. Its form will be a blurring

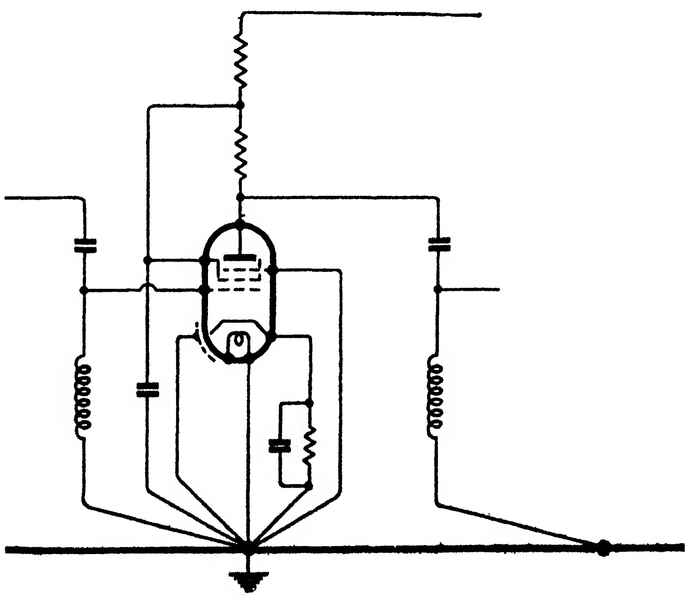


**Fig. 10.6**—With some valves common screen and anode decoupling components can be used

of the picture if the carrier is tuned to the peak. On the other hand, if the carrier is not tuned to the peak the distortion is likely to take the form of multiple contours. Thus, to the right of a black object in the picture there will be a white band, sometimes followed by a dark band, and then another white one and so on.

Their presence is usually a sign that one or more circuits is inadequately damped or that regeneration is occurring. When the r.f. circuits are moderately sharp the condition can readily be produced by adjusting their tuning while watching the picture. This effect is not always a disadvantage. Where the bandwidth is limited the apparent definition can be improved by deliberately introducing it in a small and controlled degree. It tends to improve definition, for by the contrast of the distortion band it sharpens the edges of objects in the picture.

Contrary to the case of sound reproduction, amplitude distortion is not of primary importance in television. A degree of distortion which would be intolerable in the case of sound is hardly noticeable on a picture. Owing to the use of positive modulation in this country, amplitude distortion occurs normally only on the white parts of the picture, and the effect is to reduce the contrast between parts of the picture which are nearly white but of slightly different tone values. Severe distortion will completely remove the contrast in the white region and a white object on a nearly white background will become indistinguishable.

Where a fairly high degree of vision-frequency amplification is used, the signal level in the r.f. amplifier is low enough for amplitude distortion to be negligible here. With only a moderate amount of v.f. gain the last r.f. stage may have to be carefully designed, especially if the c.r. tube requires a large input. If the tube is fed directly from the detector, however, so that the use of v.f. amplification can be avoided, the last r.f. valve becomes extremely important.

The input needed by a c.r. tube varies from about 10 V p–p to as much as 60 V p–p. Most tubes need about 20 V; this is actual picture signal corresponding to 70 per cent of the modulation of the signal. Adding the 30 per cent occupied by the sync pulses, the total receiver output must be nearly 30 V p–p.

The detector efficiency is usually about 50 per cent only, so that if the tube is fed directly from the detector the last r.f. valve must give an output of 60 V peak, or 120 V p–p. It has already been shown that the impedance of the coupling at intermediate frequency is likely to be about 2,500 Ω. With these values, the current must be $60/2{,}500 = 0{\cdot}024$ A $= 24$ mA p–p.

The anode current of the last r.f. valve must consequently exceed 12 mA and, in general, should be not less than 15 mA for this case. Few r.f. pentodes are rated for more than 10 mA anode current so that it is not easy to obtain sufficient output from an r.f. amplifier to permit the c.r. tube to be fed directly from the detector without intermediate v.f. amplification. It is possible to do it, however, by employing an output pentode.

The first essential is an adequate standing anode current, and the second is low interelectrode capacitances. These will normally be considerably higher than those of an r.f. pentode, so that the load must be reduced and some of the advantage of the higher anode current is lost.

The chief difficulty is brought about by the grid-anode capacitance. If possible, a pentode with a top-grid connection should be used. One such valve is the Marconi and Osram N43, but it is often possible to use a duo-diode output pentode and ignore the diodes. Such a valve is better than its counterpart without the diodes because its grid-anode capacitance is lower on account of the top-grid connection.

Even with such valves the input circuit is difficult because of the feedback through the grid-anode capacitance. This feedback makes the input capacitance very large and gives a negative input resistance at certain frequencies. The high capacitance means that the coupling impedance to the last r.f. valve must be of low value to preserve the bandwidth and the penultimate r.f. stage consequently gives only low gain. The negative input resistance means that the resonance curve will be distorted even if actual instability can be avoided.

The author has used the N43 successfully at 10 Mc/s, the major effects of the negative input resistance being avoided by using a 50-Ω grid-stopper. Regeneration was by no means negligible, however, and the tuning and damping of the early circuits needed very careful adjustment. Once done, however, the amplifier proved entirely satisfactory and reliable and was used for some years in experimental work.

If a large detector output is needed with a large bandwidth, the N43 may be inadequate, and a valve taking 60 mA or more may be needed. Such valves usually have a high grid-anode capacitance and a comparatively low mutual conductance. Great difficulties then arise in the penultimate stage on account of the large effective input capacitance; it may even prove necessary to use an output pentode at this point in order to drive the last r.f. valve! The design then becomes very uneconomical.

As an alternative to using a large pentode in the r.f. stage there is the possibility of using two r.f. pentodes. Two valves in parallel will not give twice the output of one, because the circuit capacitance will be greater and the impedance must be proportionately reduced. In fact, if the r.f. valve capacitance were the only capacitance in the circuit, two valves would give no greater output than one, for as the capacitance would be doubled, the resistance would have to be halved and double the current would produce the same voltage across it.

In practice, the valve capacitance is not the only capacitance and there is then a gain to be obtained with two valves. The input capacitance of the detector may be 5 pF, and the self-capacitance of the coil also 5 pF; add, say 3 pF for wiring and the total is 13 pF, plus the output capacitance of the r.f. stage. The anode-cathode capacitance of an r.f. pentode is about 7 pF, so that the total capacitance for one output valve is 20 pF, and for two is 27 pF.

For the same bandwidth the load resistance varies inversely as the capacitance so that the output with two valves is related to that with one in the ratio $2 \times 20/27 : 1$ or $1{\cdot}48 : 1$. Instead of doubling the output by using two valves in parallel it is increased by about 50 per cent only. These figures only apply to the particular capacitances quoted, of course, and the advantage of using two valves in parallel increases when the r.f. valve capacitances are smaller compared with the other capacitances.

If the requisite output can be secured by using two valves in parallel it is better to do this than to use a single output-type pentode. The grid-cathode and anode-cathode capacitances of two r.f. pentodes will be little, if any, larger than those of a single output pentode and the grid-anode capacitance will be much smaller. Feedback troubles are not likely to be important, therefore, and the design of the penultimate stage becomes easy.

The advantages of feeding the c.r. tube directly from the detector are that no special precautions for tube safety are needed, there is no necessity for d.c. restoration circuits, and sync separation becomes much easier. The disadvantages are the difficulty of obtaining adequate output from the last i.f. valve, the need for higher amplification at intermediate frequency for a given overall gain, and an increase in any tendency to i.f. harmonic interference because of this extra gain.

It is generally considered that the disadvantages outweigh the advantages, and direct feed from the detector to the c.r. tube is rarely used in commercial receivers. Almost invariably one vision-frequency amplifier is used between detector and tube. In practice, the direct feed from a diode detector is only adopted when the tube is of a type requiring an abnormally small input.

The use of a v.f. stage for the output is also more economical in anode current consumption. Assuming the same load resistances and a detector efficiency of 50 per cent, an r.f. stage must develop across its load four times the voltage change that a v.f. stage need produce for the same input to the tube. The power consumption of an r.f. stage is thus likely to be about sixteen times that of a v.f. stage. In practice, the difference will be greater, for it is usually possible to make the load on a v.f. valve higher than that on an r.f. stage.

At the present time and with the average tube the use of direct feed from the detector is hardly feasible. In special cases it may be desired to adopt it, however, and so its particular problems have been discussed in some detail.

So far it has been assumed that double-sideband working would be adopted. This is not essential, however, and it is quite possible to use a single-sideband system in the receiver even when the transmission contains both sidebands. It is, of course, essential when single-sideband transmission is used.

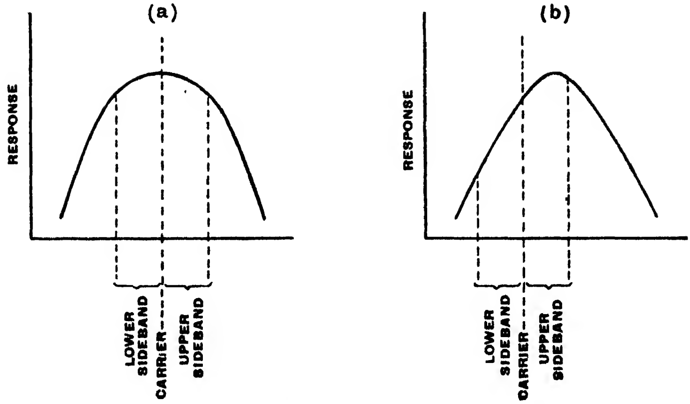



**Fig. 10.7—These sketches illustrate the tuning for double- (a) and single-sideband (b) working**

With the normal double-sideband system the receiver is tuned so that the r.f. carrier falls in the centre of the resonance curve of the r.f. amplifier as illustrated in Fig. 10.7 (a). Both sidebands are passed by the amplifier in equal proportions, and the bandwidth required is obviously equal to twice the highest modulation frequency. With single-sideband working the receiver is tuned as shown in Fig. 10.7 (b) so that the carrier falls on the sloping side of the resonance curve, at a point where the response is $-6$ db. At high modulation frequencies one sideband is passed fully by the amplifier and the other is greatly attenuated, so that true single-sideband conditions exist. At low modulation frequencies there is so little difference between the frequency of one sideband and its opposite number on the other side of the carrier that they cannot be separated properly and no attempt to do so is made. Such frequencies fall on the side of the resonance curve and are passed in nearly equal intensity.

At low modulation frequencies double-sideband working is used, therefore, and there is a gradual changeover to the single-sideband condition as the modulation frequency rises. When both sidebands are present the detector output is double that obtained with only one sideband, thus the carrier is mistuned to a point $-6$ db down on the side of the resonance curve to equalize the output.

The advantage of single-sideband working is that the bandwidth can be halved, with the result that the gain from each r.f. valve is doubled. As a partial offset to this the receiver must be built to have twice the normal gain to counteract the loss caused by the mistuning.

Phase distortion in the r.f. circuits becomes more important and the tuning is much more critical. It is very important that the carrier be placed at the right point on the resonance curve. If it is too near the pass-region the high-frequency response will be reduced and the picture definition will be poor. If it is too far from the pass-region, the high-frequency response will be excessive and serious distortion will result. The effect is usually known as plastic and the picture appears in relief, all boundaries being greatly exaggerated.

In the U.S.A. single-sideband reception is universal for the very good reason that single-sideband transmission is used. It is usually called vestigial-sideband transmission because it is a compromise between true single- and double-sideband operation. It is the former for high modulation frequencies and the latter for low.

Until recently in this country only double-sideband transmission has been used, but single-sideband reception has been frequently employed, largely because the problem of rejecting the adjacent sound channel is greatly eased if only the vision sidebands remote from it need be retained.

However, vestigial-sideband transmission is used for the Birmingham station and is to be adopted for all other stations. This does not help the sound-channel rejection problem, since it is the sidebands adjacent to the sound channel which are retained. It is, however, almost certain that all future receivers will be of the single-sideband type.

CHAPTER 11

# Selectivity

IN TELEVISION RECEPTION the problem of selectivity has hitherto been confined to the avoidance of interference from the accompanying sound channel. It has not been at all a serious one. The position is becoming rather different now for in the reasonably near future the sound channel of an adjacent television station may appear only 1·5 Mc/s above the wanted vision carrier.

The question of selectivity in television reception is a somewhat peculiar one and it is because of this that it has been given a chapter to itself and so separated from the purely amplification problems. In sound reception it is conventional to obtain amplification and selectivity together, the tuned circuits needed to couple the valves for amplification also providing the selectivity. In television it is quite impracticable to secure freedom from interference from the sound signal in this way, except in one particular case. It is nearly always necessary to introduce special rejector circuits.

It is not always realized that with a given form of circuit a specification of the bandwidth automatically specifies the selectivity also. The two are quite inseparable. Thus if a stagger-tuned amplifier with a bandwidth of 3 Mc/s at − 3 db is designed, the response at other frequencies farther from its pass-band is fixed and cannot be altered without introducing other circuits. Different forms of intervalve coupling will lead to different relations between the bandwidth and the response away from the pass-band, but the difference is not very great unless a large increase in the number of circuits is involved.

Since selectivity and bandwidth are inextricably tied together the selectivity is independent of the mid-band frequency, except for some second-order effects of small practical importance. The use of the superheterodyne principle with a relatively low intermediate frequency confers no advantage in selectivity, as long as conditions are such that the bandwidth at signal frequency is not greater than is needed. When rejectors are included, however, the superheterodyne does give some advantage.

A single pair of stagger-tuned circuits gives a response

$$1/S \approx \sqrt{1 + 4\Delta f^2/n^2} \qquad 11.1$$

where $\Delta f$ is the frequency difference from resonance and $n$ is the 3-db bandwidth. For instance, if $n = 6$ Mc/s and the band is centred on the vision carrier, the response at the sound carrier, for which $\Delta f = 3{\cdot}5$ Mc/s, is $\sqrt{[1 + (2 \times 3{\cdot}5/6)^2]} = 1{\cdot}54$ or $- 3{\cdot}73$ db. If single-sideband reception were adopted, picking

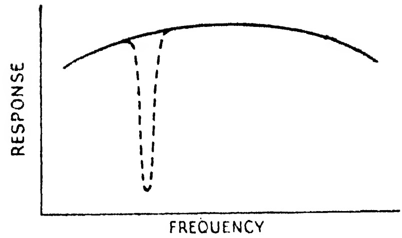



**Fig. 11.1—The solid-line curve indicates the response of a wideband amplifier, while the dotted curve shows how the response is modified by rejector circuits**

the sidebands *remote* from the sound carrier the bandwidth would be 3 Mc/s for — 6 db and the sound carrier would be 5 Mc/s from mid-band. A bandwidth of 3 Mc/s at — 6 db corresponds to one of 1·73 Mc/s for — 3 db for a pair of circuits, and so the sound-channel attenuation becomes $\sqrt{[1 + (2 \times 5/1{\cdot}73)^2]} = 15{\cdot}4$ db.

As the number of circuits is increased the selectivity per pair decreases because the bandwidth per pair must be increased to keep the overall bandwidth unchanged. The overall selectivity increases, however.

It is obvious that in double-sideband reception the attenuation of the adjacent sound channel is quite inadequate, and this applies also to single-sideband reception of the sidebands adjacent to the sound channel. The selectivity with three or four pairs of circuits is quite sufficient when single-sideband reception of the sidebands remote from the sound carrier is adopted, however. Hitherto this method has found considerable favour in receiver design and if a set does not include special sound-channel rejectors it is fairly safe to conclude that it is of this type. It is, of course, a method suited only for the reception of the London station and so is unlikely to be much used in the future.

The use of one or more rejectors tuned to the sound channel modifies a response curve like the solid line of Fig. 11.1 to the form shown by the dotted line. It puts a crevasse in the curve and so enables very high attenuation of a particular signal to be secured. It does not help much in general selectivity.

By using a number of rejector circuits with staggered resonance frequencies, however, it is possible to improve matters over a wider range of frequencies, but at frequencies between the resonances of each pair of rejectors the response always tends to come up to that of the main amplifier. The main use of rejector circuits is to eliminate interference from signals which are very close in frequency to the edges of the pass-band. They are not of much use in improving the general selectivity. It follows that the aim in design should be to obtain as much selectivity as possible in the main amplifier.

Stagger-tuned circuits give a degree of selectivity which is fixed by the bandwidth and the number of circuits used and these in turn are fixed by the number of stages of amplification needed to provide the required gain. Coupled circuits give higher selectivity for the same bandwidth and number of stages, merely because double the number of resonant circuits is involved.

It is probably best from the selectivity point of view to adopt over-coupling in some stages of coupled pairs. This requires less heavily-damped circuits and a steeper-sided response curve can be obtained. The peaks in the curve are taken out in the usual way by suitably damped single-circuit couplings. The main disadvantage of this scheme is that it requires very careful handling if the necessary frequency and phase responses are actually to be obtained in practice. In particular, circuit alignment is not easy.

General selectivity has not so far proved a serious problem in television and it remains to be seen whether it will become one under the new conditions. Hitherto, the main problem has been sound-channel rejection and this will still remain unabated.

The simplest form of rejector is a series-resonant circuit connected across one of the amplifier intervalve couplings. It is shown in Fig. 11.2 where $L$, $C$ and $R$ represent the coupling circuit and $L_1$, $C_1$ and $R_1$ the rejector. The idea behind the rejector is that away from its resonance frequency its impedance is so high that it has no effect on the intervalve circuit, whereas at its resonance its impedance is so low that it forms a virtual short-circuit to that circuit.

Needless to say, the performance in practice falls far short of this ideal. It could be obtained only with an infinite $Q$ for the rejector circuit. Even if this could be obtained, and there are circuits which simulate it, the response curve of the rejector would be so sharp that it would be difficult to keep its nose on the signal.

Since $R_1$ is always much smaller than $R$, the ratio of the response at mid-band to that at the rejection frequency (usually designated $f_\infty$ for short) is simply $R/R_1$. The resistance $R_1$ is simply the r.f. resistance of the coil $L_1$ ($Q_1 = \omega L_1/R_1$). One is more accustomed to thinking in terms of $Q$ than of $R_1$, and the ratio can be written $R/R_1 = RQ_1/\omega L_1 = RQ_1\sqrt{C_1/L_1}$. The ratio $C_1/L_1$ is in theory independent of frequency and $R$ is not dependent on it; $Q_1$ tends to be proportional to the square root of frequency for a coil of given physical dimensions. Therefore, the rejection efficacy of the circuit tends to increase with frequency.

At frequencies above and below $f_\infty$ the rejector becomes respectively inductive and capacitive and the reactance affects the main coupling. To minimize this it is necessary to make $L_1$ very large compared with $L$ and in turn this means making $C_1$ very small compared with $C$. As $C$ is only 20–30 pF this means that $C_1$ must be well under 1 pF.

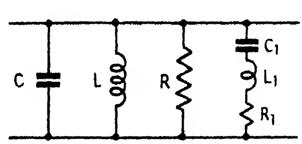



**Fig. 11.2—The components L, C and R represent a typical intervalve coupling while $L_1$, $C_1$ and $R_1$ form the simplest type of rejector**

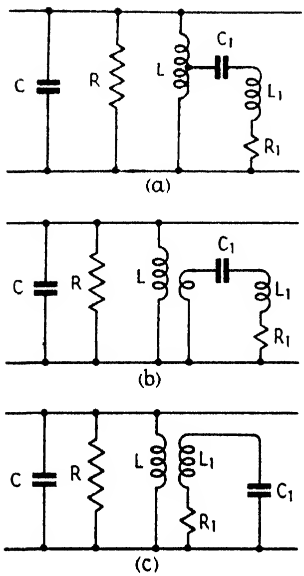
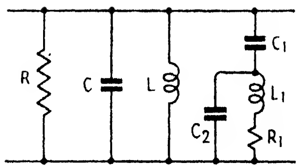



**Fig. 11.3—The rejectors shown here are all equivalent to the basic circuit of Fig. 11.2. At (a) the rejector is tapped down the intervalve coil L; at (b) it is coupled to it by a small coupling winding; and at (c) the coils L and $L_1$ are coupled directly together**

This is impracticable, for $L_1$ will have a self-capacitance of some 5 pF and so the real circuit will not be at all like that of Fig. 11.2. The difficulty can be got over by using more reasonable values for the rejector components and connecting it to a tapping on $L$ as shown in Fig. 11.3 (a). Alternatively, a coupling coil can be used as at (b) or the rejector coil itself can be coupled to $L$ in the manner shown at (c). These circuits can easily be shown to be identical with that of Fig. 11.2 save for the actual values of components used in the rejector itself. Experimentally the arrangement of Fig. 11.3 (b) is the most convenient; $C_1$ can arbitrarily be given a convenient value, say, 25–100 pF, and $L_1$ chosen to resonate with it at the required frequency, allowing a little for the inductance of the coupling coil. The required compromise between high rejection at $f_\infty$ and minimum effect in the pass-band is secured by varying the coupling to $L$. The lower capacitance values of 25–50 pF would normally be chosen for use at signal frequency and the higher of 50–100 pF for use at intermediate frequency.

It was said in the foregoing that the actual circuit of Fig. 11.2 is unusable because of the self-capacitance of the coil. This leads us to a particularly useful rejector circuit, for by deliberately adding a suitable capacitance in shunt with $L_1$ the circuit is given an extra and valuable property. The modified circuit takes the form indicated in Fig. 11.4 and $L_1$ and $C_2$ are chosen to resonate at the frequency corresponding to the lower edge of the pass-band. At this frequency, if $Q_1$ is high, the dynamic resistance of $L_1R_1C_2$ is very high. It is so high that even if $C_1$ were very large it would have a negligible influence on the main coupling circuit $LCR$.

**Fig. 11.4—This form of rejector has special advantages when the rejector frequency is very close to and lower than the edge of the pass-band**




At a lower frequency $L_1C_2$ becomes inductive and resonates with $C_1$ to form a series-resonant circuit of low impedance across the main coupling. Within the pass-band the circuit as a whole is capacitive. The effective capacitance varies with frequency but cannot exceed $C_1$, so that if $C_1$ can be very much smaller than $C$ the effect in the pass-band is negligible. The circuit is a very useful one in practice and gives an exceptionally sharp cut-off at the edge of the pass-band.

An approximate analysis of the circuit is given in Appendix 7 and it is instructive to consider the sort of performance which is likely to be obtained at signal frequency. The coupling is likely to have a capacitance of 20–30 pF with a shunt resistance of the order of 2 kΩ. Therefore, $C_1$ should be perhaps 2–3 pF only while the dynamic resistance $R_D$ at parallel resonance should be much higher than 2 kΩ and the effective series resistance $R_S$ at series resonance should be as small as possible. Certainly it must be considerably less than 2 kΩ.

The frequency difference between the sound channel $f_\infty$ and the adjacent edge of the vision-channel pass-band is $\Delta f = 0{\cdot}75$ Mc/s. For London $f_\infty = 41{\cdot}5$ Mc/s. From Equ. A.7.2, $C_1/C_2 = 0{\cdot}0362$. As a trial let us arbitrarily make $C_1 = 2$ pF, then $C_2 = 55{\cdot}4$ pF. This is, perhaps, a little on the high side for the frequency, but not unduly so. $L_1$ is, of course, chosen to resonate at 41·5 Mc/s with these capacitances and will usually be made adjustable by a dust-iron core or copper slug for trimming.

From Equ. A.7.5, $R_S = 53{,}000\ (1 + 766/Q_1^2)/Q_1$ and from Equ. A.7.7, $R_D = 62Q_1$. The value of $Q_1$ depends chiefly on the quality of the components used; $Q_1 = 100$ is easily possible and would make $R_S = 570\ \Omega$ and $R_D = 6{,}200\ \Omega$. The attenuation at $f_\infty$ is 10·9 db. A change in the value of $C_1$ will not help much for it proportionally alters both $R_S$ and $R_D$. The only thing to do is to increase $Q_1$. There is no doubt that $Q_1 = 200$ is possible and it is quite probable that a higher value is commercially practicable.

Doubling $Q_1$ doubles $R_D$ which becomes 12·4 kΩ, while $R_S$ drops to 270 Ω. The attenuation is 17·4 db. This is an extremely useful amount and two or three such rejectors fitted to successive r.f. couplings should provide all the sound-channel rejection needed.

In the case of Birmingham $f_\infty = 58{\cdot}25$ Mc/s. Keeping $C_1$ at 2 pF, $C_2$ becomes 78 pF, which is rather high. It is probably better in this case to make $C_1 = 1$ pF and $C_2 = 39$ pF. With $Q_1 = 200$, $R_S$ becomes 514 Ω and $R_D = 14{\cdot}7$ kΩ. The attenuation is 11·8 db.

Even at the higher frequencies, therefore, the circuit permits a

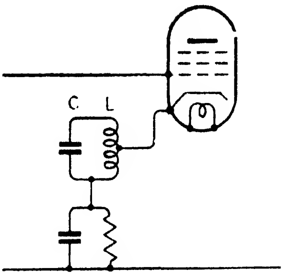



**Fig. 11.5—This cathode-circuit rejector operates by introducing negative feedback at the resonance frequency of LC**

very useful degree of attenuation to be obtained. In the i.f. amplifier of a superheterodyne it is also useful. The conditions are a little different, however, for $C_1$ must usually be larger and it then affects the pass-band more. Thus if $f_\infty = 9{\cdot}5$ Mc/s, $C_1/C_2 = 0{\cdot}164$. It will probably be undesirable to make $C_2$ less than 50 pF and then $C_1 = 8{\cdot}2$ pF. If it is necessary to use a compact single-layer coil, $C_2$ must be 100 pF and then $C_1$ is required to be $16{\cdot}4$ pF and will exercise a considerable effect on the performance of the amplifier in the pass-band. It will be necessary to alter the damping and stagger frequencies to correct for it. Experimentally, it has been found that with three rejectors of this type a measured attenuation of 40 db is obtainable.

It should be noted that this circuit only functions as described when the sound signal is lower in frequency than the vision. This is the case at signal frequency but at intermediate frequency the sound channel can be either higher or lower than the vision, depending on whether the local oscillator is higher or lower than the signals. The circuit is only useful in the latter case.

Another form of rejector which is commonly used in commercial practice is shown in Fig. 11.5 and consists of the circuit $LC$, tuned to the sound channel, in the cathode lead of a valve. At resonance it has a high impedance and causes heavy negative feedback in this stage, thus reducing the gain at the sound frequency. The circuit is effective, but the valve to which it is fitted must not have its electrode potentials varied for gain control, since this would vary the rejection.

Because of the grid-cathode capacitance of the valve the cathode impedance affects the input impedance of the valve, and hence the effective constants of the vision-channel coupling. The input resistance can be negative at some frequencies.

It is found that if $C$ is adjustable for tuning the rejector there are some values for which the stage oscillates. These values are appreciably removed from the correct setting for rejection and need not cause any trouble when everything is correctly adjusted. They complicate the process of adjustment, however.

It is worth noting that with some methods of splitting the sound and vision signals into their separate channels, the circuits used also act as sound-channel rejectors. This is further discussed in Chap. 17.

CHAPTER 12

# Signal-frequency Amplification

EVERY TELEVISION RECEIVER HAS at least one circuit operating at signal frequency, even if it is only an aerial coupling circuit between the aerial feeder and a frequency-changer. Most sets include some s.f. amplification, however. In the case of a straight set there are usually four or five s.f. stages, but in the superheterodyne the normal practice is to use only one s.f. stage, although occasionally two may be found. The bulk of the amplification is obtained at intermediate frequency and there are often three or more i.f. stages.

The design and characteristics of wideband r.f. amplifiers have been treated in Chapter 10 on a general basis. Here it is necessary to deal with matters peculiar to the signal-frequency side. There are really only two essential differences between s.f. and i.f. amplifiers in television and they are really differences of degree rather than of kind and brought about by the difference of frequency. They are ones which occur not so much because the one operates at signal frequency and the other at intermediate frequency as because the one operates above, say, 15 Mc/s, and the other below.

A 45-Mc/s amplifier is here an s.f. amplifier. If television were broadcast on 100 Mc/s a 45-Mc/s amplifier would be an i.f. amplifier but it would still have all the attributes of the s.f. amplifier as treated here. An s.f. amplifier for 100 Mc/s would be a very different thing.

The first essential difference between television s.f. and i.f. amplifiers is that the valves have quite a low input resistance at signal frequency, whereas the input resistance is high enough to be unimportant at normal intermediate frequencies. The second difference lies in the greatly increased difficulty of avoiding unwanted couplings between the stages.

The input resistance comes about largely through feedback effects caused by the inductance of the internal cathode lead of the valve in conjunction with the grid-cathode capacitance. The electron transit time in the valve also plays a part, however. With a given valve the resistance is inversely proportional to the square of frequency, so that at 45 Mc/s it is only about one-twelfth of that at the usual intermediate frequency.

The actual magnitude of the resistance depends largely on the valve design, but it can be considerably affected by the circuit and its physical form. Some of the early high-$g_m$ r.f. pentodes had an input resistance of only 500–1,000 Ω at 45 Mc/s. Resistances of 4,000–5,000 Ω are now more normal and by proper circuit design they can be pushed up to 10,000 Ω or so.

In a wideband amplifier the shunt damping resistance on the coupling has a value of 1–3 kΩ in most cases. The input resistance of the valve comes across the coupling and it forms part of the damping resistance. If the input resistance of the valve is $R_{in}$ and the coupling demands a damping resistance $R$ for the correct bandwidth, then the effect of the valve is to increase the resistance needed to

$$\frac{R\ R_{in}}{R_{in}-R}$$

Thus suppose that the coupling requires a damping resistance of 2 kΩ and that the valve input resistance is 4 kΩ, then the resistance actually fitted for damping will be $(4 \times 2)/(4-2) = 4$ kΩ. In practice, it may be somewhat higher still because of other minor circuit losses, and in the aerial coupling circuit it will be much higher because of the damping effect of the aerial resistance.

Under present conditions the input resistance of the valve will nearly always be higher than the required circuit damping resistance. This was not always so. With the earlier valves it often happened that the valve resistance was the lower of the two. This was very unsatisfactory because it dropped the gain obtainable very considerably and by making the bandwidth unnecessarily great it made selectivity almost non-existent. Possibly it was largely for this reason that most of the older television receivers were superheterodynes.

Valve input resistance as such now causes no serious difficulty—at least up to 45 Mc/s. Unfortunately, however, it is dependent on

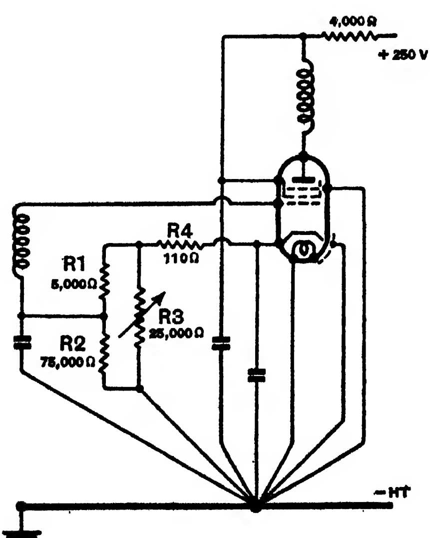



**Fig. 12.1—This diagram illustrates the connection of a gain control to an s.f. or i.f. stage**

the mutual conductance of the valve and if special precautions are not taken it varies over a wide range in a stage in which the valve electrode potentials are varied for gain control. The input capacitance of the valve also varies and as this is quite a large proportion of the grid-circuit tuning capacitance the tuning also varies. The gain control thus affects the tuning and damping of the circuits connected to the grids of the valves controlled by it.

In a stage operating at intermediate frequencies, usually 13 Mc/s or less, the change of input resistance is usually negligible because the lowest value of input resistance is too high to have appreciable effect on the damping. The change of input capacitance, however, still occurs and is just as important, so that the special measures needed in an s.f. amplifier to avoid the changes are still necessary in an i.f. amplifier.

One method of reducing the effect to small proportions is to arrange the gain control so that it biases the suppressor grid negatively as well as the control grid. The precise suppressor bias needed naturally depends on the particular valve used, but is of the order of ten to fifteen times the control grid bias.

One simple way of obtaining this bias is shown in Fig. 12.1, where the values are suitable for the Mazda SP41. The full bias developed across the cathode resistance is applied to the suppressor grid, but only the fraction $R_1/(R_1 + R_2)$ is applied to the control grid.

The performance is illustrated in Fig. 12.2. Curve A shows the change of input capacitance with control grid bias, while curve B shows the results with a control grid bias of 0·06 the suppressor grid bias. A change of input capacitance of 3·25 pF is reduced to 0·15 pF by the use of suppressor bias.

The input resistance as a function of grid bias is shown by curve B of Fig. 12.3, while curve A shows the mutual conductance. When

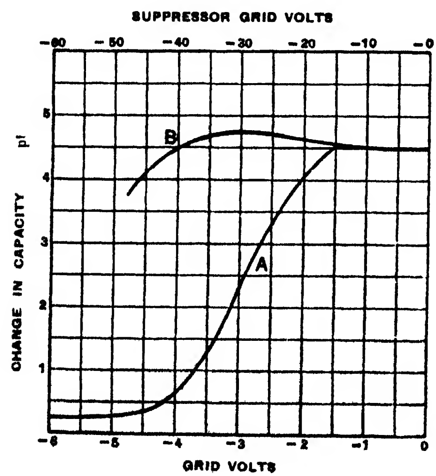



**Fig. 12.2—Curve A shows the change of input capacitance with grid bias, and curve B the change when both control and suppressor grids are biased**

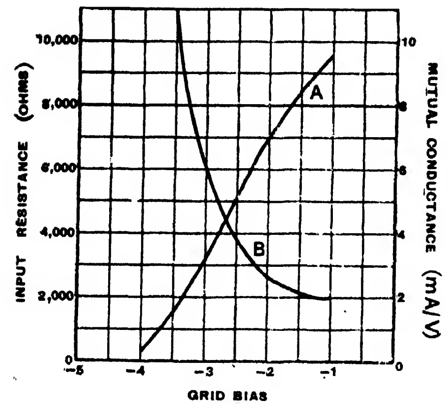



**Fig. 12.3—The relationships between grid bias and mutual conductance (curve A) and input resistance (curve B) are shown here**

suppressor bias is adopted the input resistance stays nearly constant at the value corresponding to the minimum bias adopted. If this is − 1·5 V, the input resistance is just over 2,000 Ω.

An alternative method of avoiding serious changes of input impedance with grid bias is to include an unbypassed resistance of some 30–50 Ω in series with the cathode lead of the valve. Unfortunately, the resulting negative feedback reduces the effective mutual conductance appreciably, but if this can be tolerated the method is simple and effective.

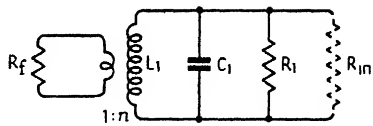
In the case of a valve of the EF50 type the input capacitance varies from 10 pF to 7·5 pF when no precautions are taken and the input resistance from 4,000 Ω to 9,800 Ω as the bias is changed from − 2 V to − 4·5 V to alter the mutual conductance from 6·5 mA/V to about 1 mA/V. By including a 32-Ω cathode resistor shunted by a 50-pF capacitor, the input capacitance changes from 9 pF at − 2 V bias to 8·9 pF at − 4·5 V, while the input resistance varies from 19,000 Ω to about 14,400 Ω. The maximum mutual conductance is hardly changed by the inclusion of the cathode components.

Although the resistance change is still quite large the magnitude of the resistance is so much increased that it is not important. The method is thus a particularly useful one, because it is so simple.

As the high-$g_m$ valves used in television are not of the variable-mu type a wide range of gain control by grid bias variation is inadvisable. It is, therefore, necessary to control more than one stage—usually two or three. In a straight set the first two stages would normally be controlled, but in a superheterodyne the s.f. stage and the first i.f. stage would be chosen.

The first s.f. circuit of the receiver is the aerial feeder-to-first valve coupling whether this valve is an s.f. amplifier or a frequency-changer. The usual circuit has the basic form shown in Fig. 12.4, where $R_f$ is the feeder characteristic impedance, usually about 70 Ω. The tuned circuit has inductance $L_1$, capacitance $C_1$, and a total

**Fig. 12.4—The basic input circuit showing the aerial or feeder impedance $R_f$, the valve input resistance $R_{in}$, and any additional damping resistance $R_1$**




damping resistance $R$ which is determined by bandwidth considerations as described in Chapter 10 and Appendix 6. Neglecting coil losses, the circuit is damped by a shunt resistance $R_1$, the valve input resistance $R_{in}$ and the transferred feeder impedance $R_f$ and the combined effect of these three must equal $R$. The transformer is taken to have an effective voltage transfer ratio of 1 : $n$ and this is not necessarily equal to the physical turns ratio, but is roughly equal to it if the coupling is very tight.

It is easily seen that at resonance the grid voltage is

$$e_g = e_f n \frac{R_1 R_{in}/(R_1 + R_{in})}{n^2 R_f + R_1 R_{in}/(R_1 + R_{in})} \qquad 12.1$$

and it is easily shown that this is a maximum when

$$n = \sqrt{R_1 R_{in}/R_f (R_1 + R_{in})} \qquad 12.2$$

and then

$$e_g = n e_f/2 \qquad 12.3$$

From the point of view of damping we have

$$\frac{1}{R} = \frac{1}{R_{in}} + \frac{1}{R_1} + \frac{1}{n^2 R_f}$$

If $n$ has the optimum value of Equ. (12.3), we get

$$R_1 = \frac{2RR_{in}}{R_{in} - 2R} \qquad 12.4$$

As an example, suppose $R_f = 70\ \Omega$, $R_{in} = 10{,}000\ \Omega$, $R = 2{,}000\ \Omega$ then from (12.4), $R_1 = 6{,}660\ \Omega$ and from (12.2), $n = 7{\cdot}56$. The voltage applied to the grid of the valve will be 7·56 times that appearing across the properly terminated feeder. This is a step-up of the same order of magnitude as the amplification of a valve stage and is well worth having.

For television purposes, there is usually little advantage in departing from the value of $n$ given by (12.3). However, this is not always the case, for the signal-to-noise ratio can be improved somewhat by adopting a smaller value of $n$ and increasing $R_1$ to keep the damping constant. This is a feeder coupling somewhat above the optimum for sensitivity.

It is not often worth-while to do this in television sets, for the improvement which can be realized is rarely greater than 1 or 2 db. As signal-noise ratios corresponding to at least 30 db are needed, it can be seen that the improvement is likely to be rarely detectable. This does not mean that signal-noise problems can be neglected. They are important in setting a limit to the maximum usable amplification and so to the maximum range at which satisfactory reception is possible.

The term noise is, perhaps, inappropriate to television, but it is well-established and convenient. It is, of course, carried over from sound technique and used for the background hiss which appears on the picture as a dirty background.

The noise arises from two causes: thermal agitation of electrons in all conductors and valve noise. So far as the receiving equipment is concerned, the first source of noise is the aerial and even if the receiver were perfect its presence here would set a limit to the weakest signal which would be useful and hence to the amplification which could usefully be employed.

No receiver is perfect, however, and the first circuit, coupling the aerial to the first valve, introduces noise, as does the valve itself and subsequent valves and circuits. The importance of the noise depends entirely on its magnitude relative to that of the signal at the same point. If the noise at a given point is 100 $\mu$V it is of no importance at all if the signal at the same point is 100 mV or more. It will usually be of negligible importance if the signal is 10 mV, but it will certainly be noticeable if the signal is only 1 mV, while if it is 100 $\mu$V the noise will certainly result in a very poor picture.

It is clear, therefore, that the noise generated in the later stages of the receiver, where the signal is relatively strong, will be of little importance. In the ideal receiver the conditions would be such that only the aerial noise would be important. This is not always practicable, and the limit should then be set by first-stage noise. To achieve this it is necessary that the gain per stage should not be below a certain minimum figure which depends on the type of valve and circuit.

A reasonably accurate expression for the signal voltage needed in the aerial, neglecting feeder losses, for a given output signal-to-noise voltage ratio $S$ is

$$e_S = 0{\cdot}126\, S\sqrt{r_a N \Delta f} \qquad 12.5$$

where $e_S$ is in $\mu$V r.m.s., $r_a$ is the aerial impedance ($\Omega$), $\Delta f$ is the bandwidth (Mc/s) and $N$ is the noise factor.

This last is given by

$$N = x\,(2 + y) + (1 + x)/y + T_a/T \qquad 12.6$$

where
$$x = R_n \Big/ \frac{R_1 R_{in}}{R_1 + R_{in}}$$
$$y = \frac{R_1 R_{in}}{R_1 + R_{in}} \Big/ n^2 r_a$$
$$T_a/T = \text{ratio } \frac{\text{aerial temperature}}{\text{ambient temperature}}$$

in degrees Kelvin.*

* Degrees Kelvin = degrees Centigrade plus 273.

The ambient temperature is usually taken as 290° K; the aerial temperature $T_a$ is often considered to be different, usually higher to allow for external noise, such as inter-stellar noise, which reaches a maximum on the present television frequencies.

Inter-stellar noise appears to originate in the Milky Way and is of sufficient intensity to make it hardly worth-while to go to extremes in reducing receiver noise. The lengths to which it pays to go in a centimetre-wave radar set are quite inappropriate in television. This does not mean that all reasonable precautions should not be taken to reduce receiver noise to a minimum; it means only that extraordinary measures are unlikely to be of much use.

When the coupling is adjusted for optimum sensitivity and $T_a/T = 1$, $N = 2\,(1 + 2x)$. With the coupling adjusted for the optimum signal-noise ratio and $T_a/T = 1$, $n$ should have the value given by Equ. (12.3) divided by $4\sqrt{1 + 1/x}$ and then

$$N = 1 + 2x + 2\sqrt{x\,(1 + x)} \qquad 12.7$$

No mention has so far been made of the term $R_n$ which appears in $x$. This is a fictitious resistance which is considered to be in series with the grid of a perfect valve and which produces the same noise as the actual valve. By a simple extension of this idea it can include also noise generated in the second valve and the coupling to it..

If $R_{n1}$ is the noise resistance of the first valve and $R_{n2}$ is that of the second and the two valves are coupled by a circuit having a resistance $R_2$ (determined by bandwidth considerations), the effective noise resistance $R_n = R_{n1} + (R_2 + R_{n2})/A^2$ where $A = g_{m1}R_2$ = voltage amplification of the first stage.

Under conditions of optimum sensitivity we then have

$$N = 2\,(1 + 2x) = 2\left[1 + 2\,\frac{R_{n1} + (R_2 + R_{n2})/A^2}{R_1R_{in}/(R_1 + R_{in})}\right] \qquad 12.8$$

and clearly for second-stage noise to be negligible we should have

$$A \gg \sqrt{\left[\frac{2(R_2 + R_{n2})}{\dfrac{R_1R_{in}}{R_1 + R_{in}} + 2R_{n1}}\right]} \qquad 12.9$$

As an illustration of the use of these equations assume that bandwidth requirements necessitate a total damping on each of the first two circuits of 2,000 Ω; this is about the right value for a bandwidth of 3 Mc/s. With a valve such as the EF50, $g_m$ is about 6 mA/V and $R_n$ about 1,400 Ω, while $R_{in}$ may be 10,000 Ω. We take $r_a$ = 70 Ω, and $S$ = 31·6 (30 db).

From Equ. (12.4), $R_1$ = 6·6 kΩ
(12.2), $n$ = 7·5
(12.8), $N$ = 3·4
(12.5), $e_s$ = 106 μV
(12.3), $e_g$ = 400 μV
(12.9), $A$ = 12 ≫ 1

By working out (12.8) with $A$ very large the reduction of $N$ amounts to 0·1 db only, so that the amplification obtained is actually sufficient to make second-stage noise negligible.

If the second stage is the frequency-changer of a superheterodyne $R_{n2}$ will be greater than the 1,400 Ω assumed above. It is unlikely to be less than three times this figure and might well be five times. This would make $R_{n2} = 7{,}000\ \Omega$, and the noise factor would be 0·1 db greater. This, again, is negligible.

If the frequency-changer were the first valve, however, the signal-noise ratio would be much poorer. Assume the amplification of the mixer is 2; that is, about one-third that of the same valve as an amplifier and that its noise resistance is 4,200 Ω. The noise factor of the input circuit and valve alone becomes 10·4. The first i.f. stage with $R_2 = 1$ kΩ, $R_{n2} = 1 \cdot 4$ kΩ adds noise which brings $N$ up to 11·6, so the second stage adds nearly 0·5 db. This is because of the low gain of the mixer. For the same signal-noise ratio, the input must be 1·54 times as great as when an s.f. amplifier is used, or some 750 $\mu$V on the first grid.

This is rather an optimistic case for, in practice, mixers often have much higher noise resistances than the figure assumed above.

It will be clear from the foregoing that under normal conditions only the first-stage noise is important. In a superheterodyne an s.f. stage should be used before the mixer when reception in any area of low field strength is required. It is not necessary, however, in areas of medium to strong field strength—say, when the field is more than 200 $\mu$V/m—provided that the frequency-changer is carefully designed for minimum noise.

When reception at extreme ranges is required every little counts. At least one s.f. amplifier will be needed and it is necessary to pick the first valve carefully for minimum noise resistance. A triode is inherently quieter than a multi-electrode valve because there is no partition effect; that is, the cathode current flows only to the anode and is not divided between screen and anode.

An earthed-grid (cathode-input) triode first s.f. stage is the easiest way of using the valve, but a neutralized triode stage is still quieter. It is more difficult to get into satisfactory operation, however.

In the limit, such a stage must be mounted directly at the aerial so that the feeder loss comes after the first stage instead of before it. This is extremely inconvenient because of the need for weather-proofing the stage and for providing its power supply and it is likely to have large advantage only when the feeder is very long.

Normally with a well designed receiver if it is found that the signal-noise ratio is inadequate the first thing to do is to check that the feeder loss is not excessive. It should not normally exceed 2–4 db. The next thing is to improve the aerial as much as possible in the space available. It is here that the biggest improvement can be made in most cases. Only after this will the use of special valves and a pre-amplifier at the aerial be considered.

So far little has been said about the constructional form of an s.f. amplifier. This is quite important for the stability depends very largely upon it. Feedback through the grid-anode capacitance of the valves is not usually sufficient to cause any serious difficulty in the case of the vision channel because the stage gain is low. It can be very important in the sound channel, however, for the narrower bandwidth permits much higher gain per stage.

With a careful layout of components surprisingly little screening is necessary and it is sometimes possible to build quite a high-gain amplifier without any deliberate interstage screening or even using screened coils. This is only possible if the layout more or less follows the circuit diagram so that the valves and coils are in order. It comes about largely because of the low stage gain in a wideband amplifier, for the many stages needed automatically ensure a good separation of the parts between which the gain is high.

Stability is likely to be much better if the amplifier is built on a long, narrow and deep chassis than on a wide one. Feedback from output to input is likely to be considerably less. This is because the inverted-tray type of chassis acts very like a waveguide to couple output and input. As a waveguide, however, it is operating well below any propagating mode and gives an attenuation which increases as the dimensions of its cross-section decrease. The r.f. chassis should thus be narrow, deep and long with all valves in line.

When single-tuned couplings are used, whether they are staggered or not, one's first reaction is to employ single coils in the tuned-grid or anode circuit because of their mechanical simplicity. While this is satisfactory at the relatively low frequency of an i.f. amplifier, the author has found that a considerable improvement in stability is obtainable at television signal frequencies by using double-wound coils.

Two windings are used, one for the anode circuit and the other for the following grid. They are as tightly coupled as possible and with single-layer coils one should be wound directly over the other with only a thin layer of paper between for insulation. The advantage is that it is possible to return both anode and grid circuits directly to the cathodes of their valves and so to reduce circulating currents in the chassis.

This will be clear from Fig. 12.5, where (a) shows a tuned-anode circuit. The tuned circuit is $L$ tuned by $c_{ak}$ and $c_{gk}$; as these capacitances are chiefly in the valves they must be physically separated and no matter where $C_d$ is returned the currents in them must flow through the chassis. In (b) the double-wound transformer separates the return leads for the currents and the leads can be taken directly to the appropriate cathode-return points, thus keeping the currents out of the chassis.

If $L_1$ and $L_2$ are identical and are 100 per cent coupled the circuit is electrically identical with a single coil. If the coupling is not 100 per cent, and it never can be, there is a difference. It really forms a coupled pair and will possess two resonance frequencies. The second and unwanted one usually comes at a much higher

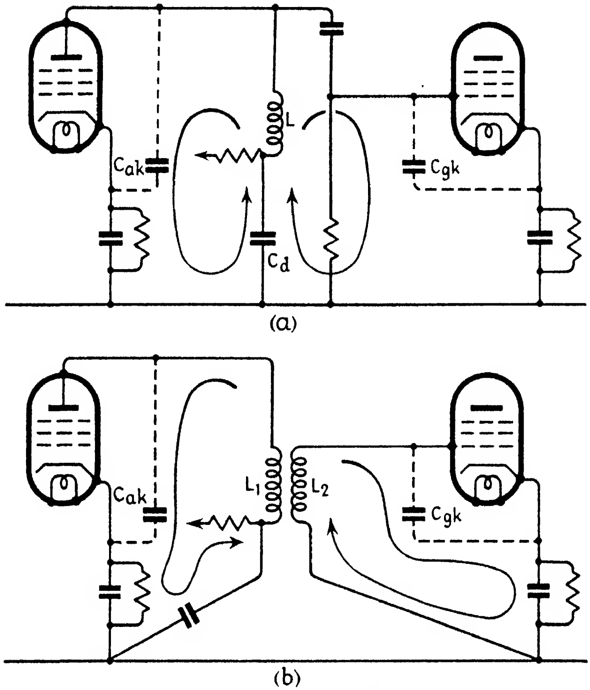



**Fig. 12.5—These diagrams illustrate the way in which the use of a double-wound transformer coupling (b) enables circulating currents to be kept out of the chassis. With a single coil only (a) the circulating currents inevitably flow in the chassis**

frequency than the signal channel and it is consequently so heavily damped by the low input resistance of the valve at this high frequency that it gives but little response.

The chassis and screen materials are important and must be of high conductivity. It is usually best to use copper, but if one has plating facilities it is satisfactory to employ steel copper-plated, or better still, silver-plated. Although inferior to copper, aluminium is often satisfactory, and has the advantage of lightness. It does suffer from the great drawback of being a difficult material to which to make a really sound and reliable electrical contact. With copper or silver it is easy to solder all chassis connections.

The heater wiring is a common cause of feedback which causes instability. In the author's experience it is desirable to connect one side of each heater directly to the chassis and to run the bus-bar for the other heater connections very close to the chassis. In high-gain amplifiers a capacitor of some 0·001 $\mu$F may be needed across one or more of the valve heaters and in addition small r.f. chokes in series with the unearthed leads may be necessary.

It is often easier to achieve stability with the single-ended valves of the EF50 type than with the kind having a top-grid or anode connection, for the mechanics of the design are easier. It must never be forgotten, however, that the valveholder contact for the centre spigot *must* be earthed for it provides a large part of the grid-anode screening.

In general, the suppressor grid ($g_3$) must be earthed, not connected to cathode. If the latter connection is adopted, instability is probable. This can easily be seen from Fig. 12.6. In (a) the grid-cathode capacitance $c_{g_1k}$, the anode and suppressor capacitance $c_{g_3a}$ and the cathode by-pass capacitor $C_k$ form a star network. This has an equivalent in $\Delta$ form (b) in which $C_1$ between grid and anode has the value $C_1 = \frac{c_{g_1k}\, c_{g_3a}}{c_{g_1k} + c_{g_3a} + C_k}$. The finite value of $C_k$ results in an increase of the effective grid-to-anode capacitance. The effect can be avoided by returning the suppressor grid to the chassis instead of to the cathode.

In general, no serious difficulty should be encountered with three or four s.f. stages. With more stages the difficulties increase rapidly.

The superheterodyne is easier than the straight set from the stability point of view, for the gain is split between two different frequencies, but it has its own peculiar problems. In the present television band it is perfectly possible to obtain all the amplification needed for television with a straight set, so that it is never *necessary* to use a superheterodyne.

Up to the present there have been no tuning problems in British television. As Alexandra Palace has been the only station, it has been necessary only to make the s.f. circuits trimmable to 45 Mc/s or to frequencies staggered closely about that figure.

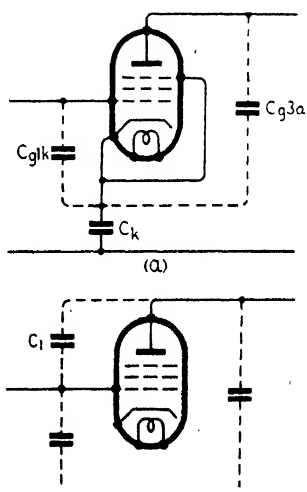



**Fig. 12.6—When the suppressor grid is returned to cathode the various circuit capacitances shown in (a) have an equivalent form (b) from which it can be seen that the effective grid-anode capacitance is increased**

Since it is important to keep the circuit capacitance as small as possible the use of capacitance trimmers would cause a considerable drop in gain and it is invariably the practice to use inductance trimming. The coils are usually single-layer wound and provided with a movable core. This is sometimes of dust-iron, in which case the inductance increases as it is inserted into the coil, or it is a metal slug which reduces the inductance as it is inserted.

Since the losses in wideband circuits are in any case heavy such a slug can be made of brass. But if it is to be used in a low-loss circuit, such as one in the sound channel, or a sound-channel rejector, it must be of copper or of silver-plated brass.

The trimming range obtainable by these methods is quite small and is often only about 10 per cent. As normally applied, therefore, they are unsuitable for tuning a set over the television band.

At the time of writing nothing is known about the methods which will be used to cater for the different television stations of the future. The plan for operating these stations appears to be for them all to relay the same programme; in any case, in most areas only one station will be receivable. There is no need, therefore, for a set to be tunable over the full band. The user's requirements are met if his set can easily be altered to tune to a different frequency if he should happen to move to a district served by another station.

From the manufacturing point of view it is highly desirable that all sets should be alike. The simplest way of covering the band would probably be to provide as wide a trimming range as is economically practicable and to supplement this by tapping the coils. A change of frequency could then be covered by re-trimming plus a possible change-over of one wire per coil. However, it is possible to obtain a frequency ratio of 20–25 per cent from either slugs or dust-iron cores with careful design, so that by using a composite core, half metal-slug and half dust-iron, it should be possible to cover the whole television band. Changing frequency would then be merely a matter of circuit realignment.

The superheterodyne has here one big advantage over the straight set—only three circuits need be altered to change frequency, whereas the straight set will have at least nine and probably 12–15, including those in the sound channel. In the U.S.A. the superheterodyne is universal, but there conditions are different. Some 6–12 channels are provided for selection by the user and quite a number of them are on frequencies well over 100 Mc/s, where the straight set would be unsuitable in any case. It is not safe, therefore, to judge by American practice.

CHAPTER 13

# Intermediate-frequency Amplification

As ALREADY EXPLAINED an i.f. amplifier differs from a television s.f. amplifier mainly in the matter of its mid-band frequency. As this frequency is usually considerably lower the performance is rather less dependent on the details of the mechanical design and it is rather easier to achieve stability. There is not a great difference, however, and it is wise to design the amplifier on the lines discussed in Chapter 12 in connection with s.f. amplifiers. Decoupling and by-pass capacitors should be somewhat larger and usually about 0·01 $\mu$F.

The most important matter in which i.f. amplifier design differs from an s.f. amplifier is the choice of the mid-band frequency. The equations in Appendix 6 show that the gain per stage for a given bandwidth is independent of the mid-band frequency. In fact, it does depend on it, but to such a small degree that it is quite negligible for any otherwise satisfactory frequency. The gain tends to decrease very slightly as the frequency is lowered.

The advantages of the superheterodyne over the straight set are two. First, it is much easier to cater for the reception of a number of different stations since only the oscillator and a few s.f. circuits need be altered for tuning. Secondly, because the amplification is split between the s.f. and i.f. amplifiers it is much easier to achieve high amplification with stability.

The first advantage has been hitherto of no importance at all because there has been only one television station but in the future it may become a deciding factor. The second advantage is somewhat illusory, for although unwanted feedback may not result in instability as it would do in a straight set it has other effects which can seriously affect the picture quality.

If these troubles are to be avoided in the right way—by eliminating the feedback—the superheterodyne demands just as much care in design and construction as the straight set. If the set has to cover a range of signal frequencies it is necessary to eliminate feedback, but if only one signal frequency need be catered for, it is possible to dodge the effects of the feedback by a careful choice of intermediate frequency.

Theoretically, the intermediate frequency can lie anywhere between 3·5 Mc/s and 40 Mc/s. This is a wide band and there are two factors which will influence the choice. The lower the frequency used, the easier it is to avoid feedback effects within the i.f. amplifier —effects which may cause instability or serious distortion of the resonance curve.

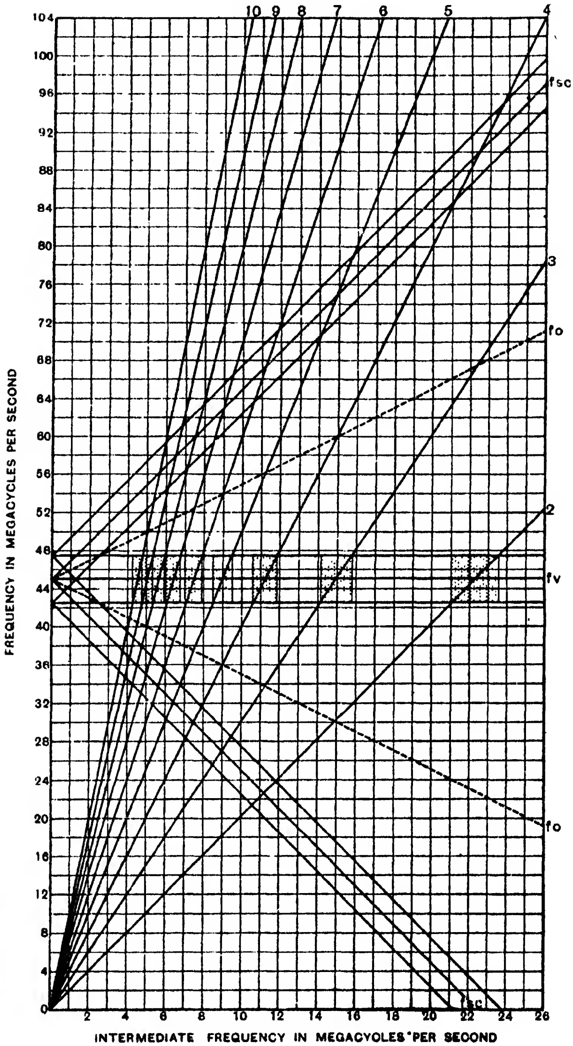



**Fig. 13.1—This diagram shows the intermediate frequencies at which i.f. harmonic interference is likely to occur when the vision carrier is at 45 Mc/s**

On the other hand, the lower the frequency the more difficult it is to filter out i.f. currents from the detector output. If such currents are fed back from the detector to the input circuits of the receiver serious interference occurs. The effect is the same as i.f. harmonic interference, which is responsible for many tunable whistles in broadcast receivers.

If a receiver were set up with a variable intermediate frequency and having a certain amount of feedback present, it would be found that at high intermediate frequencies there are quite wide bands without interference, intermarked by bands of very severe interference. As the intermediate frequency is lowered the interference bands increase in number, but decrease in intensity and the clear bands get fewer and narrow. At length, with low frequencies, there are no clear bands and there is interference at all frequencies, but of considerably lower intensity.

The reason for this is that it is harmonics of the intermediate frequency which cause the trouble. At a high frequency the second harmonic causes the interference and there is then a wide gap until the third harmonic comes into play; then another gap and the fourth is active. As the frequency is lowered the interference becomes of smaller magnitude because harmonics of higher order are involved.

The change of intermediate frequency needed to change from one harmonic to the next gets smaller as the order of harmonic involved increases. At length the gaps cease because as soon as the frequency has moved out of the sphere of one harmonic it is within the sphere of the next.

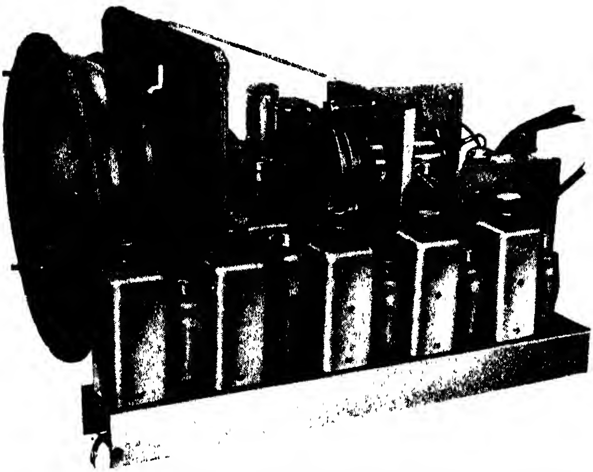
The diagram of Fig. 13.1 makes this clear. It is due to L. H. Bedford* and clearly shows which intermediate frequencies are likely to lead to interference and which are not. The vertical scale shows signal frequencies and the horizontal intermediate frequencies. The horizontal lines mark out the signal frequency with its sidebands; the bandwidth shown should be that of the i.f. amplifier and in the diagram it is shown as 5 Mc/s. The interference-free bands decrease with an increase of bandwith.

The diagonal lines radiating from the corner represent the i.f. harmonics, and the points of intersection of these lines with the outer horizontal lines give the intermediate frequencies between which interference will be found if the feedback in the receiver is sufficient. The interference bands are shown shaded in Fig. 13.1, and it will be seen that below 6 Mc/s the bands overlap. At 6 Mc/s there is a very narrow clear band; there is a wider one just below 7 Mc/s, another at 8·2 Mc/s, and a fairly wide one centred on 10 Mc/s; 13 Mc/s is the centre of a wide band and 18·5 Mc/s is the centre of a very wide one.

The interference can be avoided in two ways—by adopting adequate screening and filtering so that the feedback is eliminated, or by choosing an intermediate frequency at which the interference is not found. The former is the more elegant course and is essential if the receiver has to be tunable over a range of signal frequencies.

If the set has to receive only one station, however, it is quite legitimate to pick the intermediate frequency for the avoidance of interference. No special screening is then needed and a simple detector filter will suffice.

* *Journal of the Television Society*, March 1937.



**In this view of a magnetic television receiver, the s.f., frequency-changer and i.f. stages are in the foreground**

Up to the present it has been the usual practice to adopt this course and the favourite intermediate frequency is probably 13 Mc/s. It is a good choice, for it is in the centre of a fairly wide band which is free from feedback troubles, consequently, there is no fear of interference occurring if the actual frequency used turns out to be a little different from what one wanted. It is high enough to make filtering the detector output reasonably easy and it is not so high that it is necessary to take into account electron transit time effects in the i.f. valves. Moreover, it is not difficult to stabilize a high-gain ampl fier at this frequency.

In commercial practice 13 Mc/s is a common choice and a higher frequency is rarely, if ever, used. Lower frequencies, such as 10 Mc/s or 8·25 Mc/s, are sometimes adopted. In the absence of special circumstances which introduce other factors affecting the choice, 13 Mc/s is on theoretical grounds about the lowest frequency which it is safe to adopt and this agrees with the author's experience.

It is worth noting that the choice of frequency does depend on the receiver bandwidth. Referring to Fig. 13.1 it will be obvious that increasing the bandwidth decreases the width of the clear bands, for moving the horizontal lines depicting the limits of the pass-band farther apart increases the width of the shaded areas and so reduces the clear spaces between them.

With a really wideband receiver an intermediate frequency of 18·5 Mc/s is the safest choice, but even then it is probably better to keep to 13 Mc/s and to increase screening and filtering to remove interference. This is because it is easier to avoid interference from the sound channel when the frequency is low than when it is high.

The carrier frequency of the i.f. signal should always coincide with the centre of one of the clear bands of Fig. 13.1. With the usual double-sideband working this will also be the mid-band frequency of the amplifier. With single-sideband working, however, the i.f. carrier should still come to the centre of the clear band; that is, one edge of the i.f. pass-band should coincide with the centre of a clear band.

Thus, suppose that the intermediate frequency proper is 13 Mc/s. For double-sideband working the bandwidth may be 5·5 Mc/s and the pass-band will be made 10·25 Mc/s to 15·75 Mc/s. This band will overlap the areas of interference in Fig. 13.1, but this usually causes no trouble if the carrier lies in the centre. For single-sideband working the bandwidth may be 2·75 Mc/s and the pass-band will be made 10·25 Mc/s to 13 Mc/s, or 13 to 15·75 Mc/s, the carrier being placed at 13 Mc/s in either case.

In Fig. 13·1, diagonal lines $f_o$ show the oscillator frequencies for any intermediate frequency, and the other lines $f_{sc}$ show the bands from which second-channel interference may be experienced. With the oscillator at a higher frequency than the signal, it must be at 58 Mc/s for an intermediate frequency of 13 Mc/s. Second-channel interference can occur from signals on frequencies in the neighbourhood of 71 Mc/s and its possibility is relatively small.

The sound accompaniment to television on 41·5 Mc/s will produce an intermediate frequency of 16·5 Mc/s and so will appear on the less selective side of the vision i.f. amplifier.

Using the lower oscillator frequency it is set at 32 Mc/s and second-channel interference is possible from signals around 19 Mc/s (15·8 metres). Such interference is quite probable if care is not taken in the design of the s.f. circuits. The sound appears as a frequency of 9·5 Mc/s, on the more selective side of the i.f. amplifier. Better oscillator stability is usually obtainable at the lower frequency and for this reason it is often preferred.

It should be noted that it is important to provide adequate s.f. selectivity for the avoidance of second-channel interference even if there happen to be no stations which are likely to cause such interference. There are two reasons for this. The first is that if the input circuits are responsive to the second-channel frequency they will be responsive to noise in this band, including aerial and receiver noise. The signal-noise ratio may then be anything up to 3 db worse than it need be.

The second reason is a little more complicated. Although the intermediate frequency may have been selected in relation to the signal frequency to avoid i.f. harmonic interference, it does not follow that this relation will hold as far as the second channel is concerned. I.f. harmonics falling in the second channel will pass to cause interference if the s.f. selectivity is inadequate.

Referring to Fig. 13.1, 13 Mc/s is a clear band and is normally chosen. If the oscillator is higher than the signal frequency the second-channel band is centred on $45 + 26 = 71$ Mc/s. Careful examination of the diagram will show that 13 Mc/s still comes in a clear band between the 5th and 6th harmonics, but it is a very narrow one and there is very little factor of safety.

On the other hand if the oscillator is lower in frequency than the signal, the second-channel band is centred on $45 - 26 = 19$ Mc/s and this is a region which is quite clear of any harmonics. From this point of view, it is safer to use the lower of the two possible oscillator frequencies. It must be remembered, however, that external second-channel interference is much more likely from signals on 19 Mc/s than on 71 Mc/s, so in either case it is necessary to provide sufficient pre-selection to avoid second-channel interference.

CHAPTER 14

# Superheterodyne Frequency-changers

THE FREQUENCY-CHANGER of a television superheterodyne differs in no essential from that of any other receiver. It is in one respect somewhat simpler because the need for ganging signal- and oscillator-tuning circuits has not arisen.

The important factors are to obtain a high degree of frequency stability from the oscillator, low noise and high conversion amplification. The importance of oscillator stability is sometimes overlooked for it arises only through the presence of the sound channel. It would be unnecessary if the receiver had to produce an output from the vision signal only and if the direct selectivity of the i.f. amplifier were adequate.

This case is quite a practical one under some conditions. Thus, if a single-sideband receiver is used to select those sidebands of the London station which are remote from the sound channel its selectivity can be adequate without special sound-channel rejection circuits. An entirely separate receiver might be used for the sound channel. Oscillator stability would then be unimportant for a frequency tolerance of as much as $\pm$ 0·2 Mc/s would probably have little discernible effect on the picture. With the oscillator at 32 Mc/s this is a stability of only 1 part in 160.

In most practical cases, however, sound-channel rejectors are used and it is then necessary for the stability to be good enough to keep the sound i.f. signal within the bandwidth of the rejectors. When a combined sound and vision receiver is used it must also be good enough to keep it within the bandwidth of the sound channel.

Since a sound-channel bandwidth of 30 kc/s is all that is necessary to make full use of the transmission it is easy to see that very high oscillator stability is needed. In practice, the bandwidth is always made more than 30 kc/s. It is usually 50–100 kc/s. The latter allows a permissible drift of $\pm$ 35 kc/s, so that the oscillator stability must be to roughly 1 part in 1,000 when it is at 32 Mc/s. If the alternative frequency of 58 Mc/s is chosen the stability must be to 1 part in 1,600.

Many, if not most, commercially produced superheterodynes include a trimmer so that the user can correct for any drift. He does this by treating it as a tuning control and adjusting it for the best sound signal.

Under normal conditions oscillator drift arises mainly from temperature changes. Voltage changes can also produce it, but are less important in most cases. Since television equipment is used chiefly for entertainment purposes the ambient temperature in which the set is operated can be taken as substantially constant at around 65–70° F. No one is likely to sit for long watching a programme in a room temperature less than 60° and the occasions in summer when the room temperature exceeds 80° are rather few.

The one case when in normal use there might be a large change of ambient temperature does not occur very frequently. It is during a really cold spell, when one might come into an unheated room with a temperature around freezing and switch set and heating on together. Even then the change of room temperature is not likely to exceed 30° or so.

The temperature changes which cause drift of oscillator frequency are chiefly those which occur in the set due to the power dissipated within the set itself. They are of two kinds. The first is in the oscillator valve itself, the second in the oscillator components.

The valve capacitances change appreciably with temperature and so affect the oscillator frequency. The heat comes from the cathode and is unavoidable. All that can be done is to ventilate the oscillator valve as well as possible in order that it may reach a stable operating temperature quickly. With good ventilation the warming-up period can be confined to 5–10 minutes at most. In addition the use of a low $L/C$ ratio in the oscillator circuit minimizes the magnitude of the drift by making the valve capacitance, which is subject to temperature drift, only a small part of the total capacitance.

The major part of oscillator drift occurs, in practice, through changes in the components apart from the valve and this can be overcome completely if enough trouble is taken. A poor layout of components and inadequate ventilation are the usual causes.

In the average television set something like 200 W is being dissipated continuously within the cabinet. The temperature must rise until the rate at which heat is lost from the cabinet equals the rate at which it is being supplied to it. The cabinet is nearly always of wood or plastic and is a poor conductor of heat; as a result, heat is lost chiefly to the surrounding air. If the ventilation is not good the whole interior of the cabinet will reach a high temperature and oscillator drift will be large. Moreover, it will take a long time to reach equilibrium, perhaps an hour or more, and so the tuning will need frequent readjustment during a programme.

Adequate air inlets and outlets must be provided. A great deal can be done by properly placing the oscillator components in the receiver. The best course is to place them at the lowest part of the set at a lower level than any heat-producing parts and immediately over an air inlet. This is not always easy and a common mistake is to lay out a receiver chassis well, and then to mount it in the cabinet on top of a power pack !

When care is taken in the design and construction it will usually be found that the oscillator drift is negligible. If it is not, temperature compensation is possible, by using a suitable mixture of capacitors having positive and negative temperature coefficients.

The main drawback to this, from the point of view of the amateur, is the difficulty of determining the proper values. It can only be done experimentally and it takes some hours to determine the effect of each small change. It is a good scheme commercially, however, and may well enable a higher temperature in the set to be satisfactory.

Any of the conventional frequency-changer circuits can be used. The triode-hexode is simple and gives good results. A typical circuit which the author has used successfully with the oscillator at 58 Mc/s is shown in Fig. 14.1. The values shown on the diagram are suited to a Mazda ACTH1 valve. The resistor $R_1$ is sometimes needed to suppress parasitic oscillation and should have a value of about 4 Ω.

The most important factor in the avoidance of parasitic oscillation is the use of very short leads to the oscillator tuning capacitor $C_1$, which can conveniently be of about 12–15 pF maximum capacitance. Excessive amplitude of oscillation should also be avoided and the makers' figure of 9 V should not be exceeded. This corresponds to a current of about 150 $\mu$A through the grid leak $R_2$.

The amplitude obtained depends largely upon the coil and especially upon its diameter. A coil of 5 turns of No. 22 enamelled wire wound 8 turns per inch on a $\frac{3}{4}$-in diameter former is suitable and gives the correct amplitude when $C_2$ is 10 pF. A coil of smaller diameter will usually necessitate a larger value for $C_2$ if the amplitude is to be maintained.

This is not as good, however, for a small value of $C_2$ means that the tuned circuit is loosely coupled to the valve. Consequently,

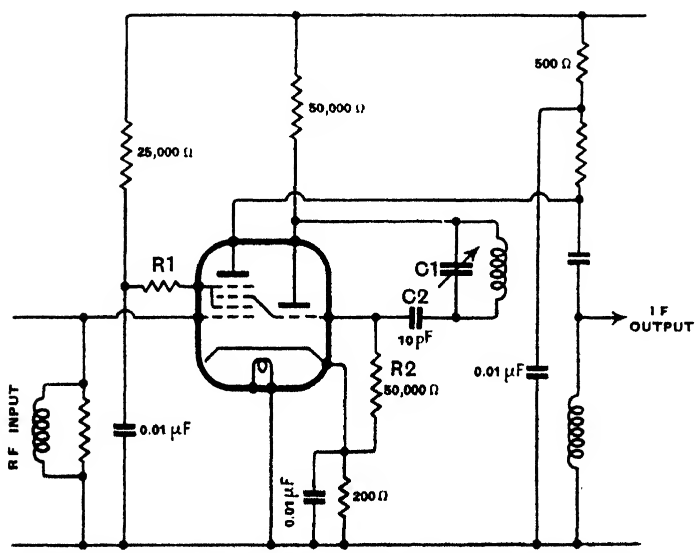



**Fig. 14.1—This diagram shows a typical triode-hexode frequency-changer**

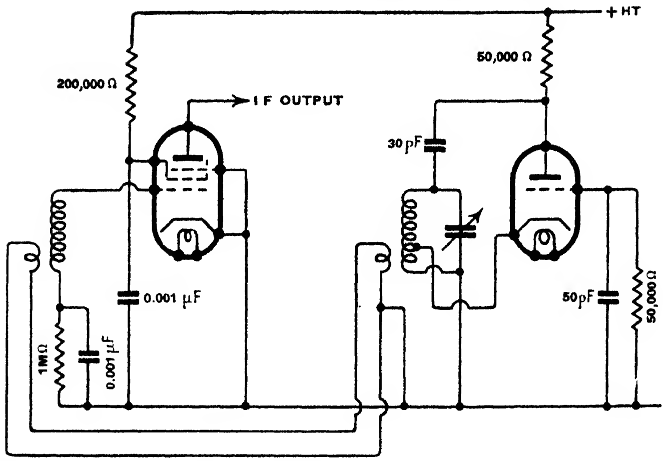



**Fig. 14.2—An r.f. pentode can be used as a mixer with a separate triode oscillator**

the stability is likely to be somewhat higher than with a larger capacitance.

The stage gain of the frequency-changer is computed in the same way as for an i.f. stage, save that the conversion conductance of the valve is used instead of the mutual conductance in the design formulæ. This is of the order of 0·75 mA/V only for many frequency-changers.

In addition, the output capacitance of a triode-hexode is usually higher than that of an r.f. pentode. The coupling impedance following this stage is thus lower, still further reducing the gain. In a typical case, it may be only 1,500 Ω giving a stage gain of only 1·12. Normally one cannot reckon on the frequency-changer contributing to the amplification, and it is sufficient if it does not attenuate.

Greater amplification can be secured, however, by using an r.f. pentode as the mixer with a separate triode oscillator. The pentode can be biased nearly to current cut-off, as is usual with such frequency-changers, but the makers of the Mazda SP41 recommend the use of bias obtained by grid rectification.

As shown in Fig. 14.2, a grid leak and capacitor are used with values of 1 MΩ and 0·001 μF respectively. The screen potential is obtained by a dropping resistance for which a suitable value is 0·2 MΩ. The optimum heterodyne voltage on the grid is then 2·25 V peak and the conversion conductance is nearly 3 mA/V.

As the output capacitance is the same as in an i.f. stage, a normal value of coupling resistance can be used and appreciable gain secured from the stage. Actually, the gain is almost exactly one-third of that given by the same valve as an amplifier.

The recommended oscillator circuit using the P4 valve is also shown in Fig. 14.2. The performance obtained naturally depends very largely upon the attainment of the correct amplitude of oscillation on the grid of the mixer and this must be found experimentally since small variations in the length of the leads will affect it appreciably.

In this respect the triode-hexode is easier to use and the fact that its gain is lower is not necessarily a disadvantage, since it comprises only a single valve assembly as compared with the two separate valves of Fig. 14.2. It is true that the triode-hexode really consists of two valves, but as they are in a single envelope they cost little more than one ordinary valve.

Counting all the electrode assemblies in one envelope as a valve, we can obtain more amplification from one r.f. stage and a triode-hexode frequency-changer than from a two-valve frequency-changer only. The two-valve arrangement usually gives appreciably better frequency stability. In the first place, a valve of higher mutual conductance can be used for the oscillator and it can consequently be coupled more loosely to its tuned circuit. Secondly, many fewer

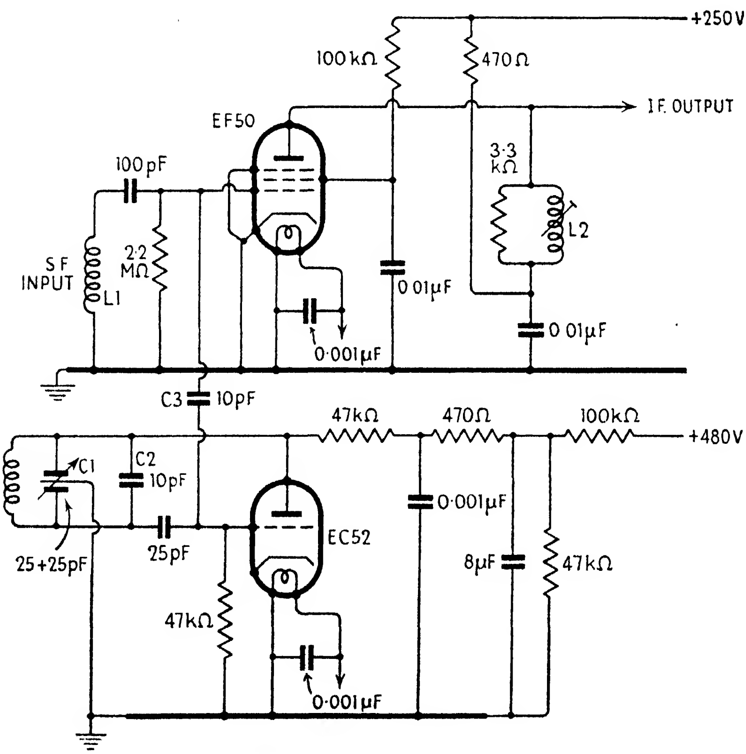



**Fig. 14.3—Circuit of a two-valve frequency-changer designed for high oscillator stability**

oscillator volts are needed on the control grid of a pentode mixer than on the injector grid of a triode-hexode and, again, looser coupling can be used.

The noise introduced by the two-valve circuit is usually considerably less. The noise resistance of a triode-hexode is usually over 25 kΩ whereas that of the two-valve arrangement may be 5 kΩ or so only.

Because of these points the author has come to prefer the two-valve circuit and an arrangement which he has used very successfully is shown in Fig. 14.3. A Colpitt's oscillator is used with a high-$g_m$ triode and a split-stator tuning capacitor $C_1$ of 25 pF for each section. In addition a 10-pF fixed capacitor $C_2$ is used across the circuit to increase the total capacitance. The valve is of a type having low interelectrode capacitances. The oscillator is coupled to the mixer by the 10-pF capacitor $C_3$. The oscillator coil is closely wound on a $\frac{3}{8}$-in diameter polystyrene former, the turns being cemented in place. For 32 Mc/s, 15 turns of No. 20 enamelled wire were used.

In order to minimize instability due to voltage fluctuations the oscillator h.t. supply should be taken from the most stable point in the set. This is usually after the first smoothing choke and is also the highest voltage point. Little or no protection against mains-voltage fluctuations is secured, but adjustments to the gain or focus controls have less effect on the oscillator frequency.

In the set in which the circuit of Fig. 14.3 was used a 480-V supply was available and so the oscillator was fed from this through the potential divider shown. It could, of course, be fed from the 250-V line by changing the resistor values appropriately.

One factor is very important. Thorough screening of the frequency-changer is necessary to prevent radiation of the oscillator frequency. For the same reason an s.f. stage before the frequency-changer is necessary even in the few cases where it is not essential for the attainment of the required signal/noise ratio.

CHAPTER 15

# The Detector

ANY OF THE DETECTORS used in sound receivers can be employed for television purposes with suitable changes in the circuit values. In practice, however, the diode detector is almost invariably adopted.

As usual in television the practical difficulties are connected with the stray circuit capacitances and, in order to obtain the required frequency response, the resistance values have to be about one-hundredth of those commonly used in broadcast equipment. For the efficiency of rectification to remain unchanged, the diode resistance would have to be proportionately lower.

This is impracticable, however, but special diodes of much lower resistance than usual are actually employed. Their resistance is not low enough for the normal efficiency to be secured, and it is often in the region of 50 per cent instead of being about 90 per cent.

The basic detector circuit is shown in Fig. 15.1, where $L$ represents the input tuned circuit at signal or intermediate frequency. No tuning capacitor is shown since this is usually provided by the stray capacitances.

The detector shown at (a) gives an output in positive phase and that at (b) in negative phase. That is to say, the output increases as a positive or negative potential respectively for increasing modulation depth of the input.

The circuit (a) is of the correct form for feeding the grid of the c.r. tube directly or through an even number of v.f. stages, while (b) is used with an odd number of v.f. stages. As there is

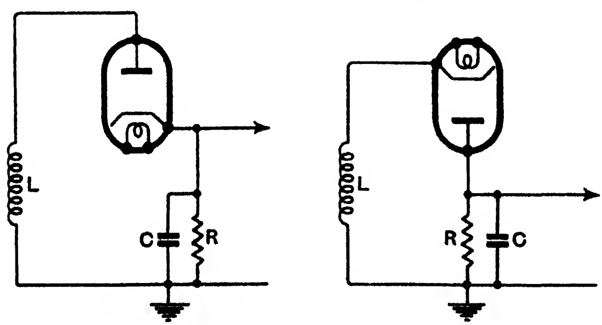



**Fig. 15.1—The connections of a diode detector are shown for an output in the positive sense at (a) and in the negative at (b)**

usually one v.f. stage only, circuit (b) is normally adopted. Of course, when the v.f. signal is applied to the tube cathode instead of the grid the conditions are reversed. Thus, with one v.f. stage circuit (a) is then used.

For a given diode the efficiency increases with the values assigned to $C$ and $R$, but unfortunately these values are severely limited by the necessity for maintaining the modulation frequency response to 2–3 Mc/s. The choice of values is best made on the same lines as in a v.f. stage.

In the computation of frequency response some doubt often arises as to how the diode resistance is to be taken into account. It is sometimes felt that the time-constant of the circuit is not $CR$ but $CR_1$ where $R_1$ is the value of $R$ in parallel with the diode resistance. This is by analogy with the case of a triode amplifier.

Actually, there are two different time-constants effective according to whether the modulation is increasing or decreasing. For increasing modulation depth the time-constant is $CR_1$ and for decreasing modulation it is $CR$. It must be remembered that the diode conducts only on the peaks $E$ of the r.f. input signal and in doing so it builds up a potential across $C$ which has an average value of $\eta E$, where $\eta$ is the efficiency of rectification. Suppose $E$ is 10 V peak and $\eta$ is 0·5, then the average voltage across $C$ is 5 V. Then on each cycle of input the valve does not conduct until the instantaneous voltage exceeds about 5 V.

Now if the input is suddenly increased to 15 V, say, the capacitor voltage will rapidly rise to 7·5 V, because the charging current flows through the low resistance diode. If the input suddenly drops to 5 V, however, then the output cannot so readily follow it.

With a normal input of 10 V, there is already 5 V across $C$, so that if the input drops to this value the diode is non-conductive until the capacitor voltage has fallen by discharging through $R$.

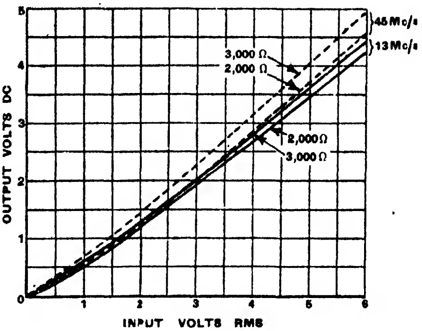
The practical effect of this is that one can have a better response for increasing modulation depth than for decreasing. In other words, if there is a vertical black bar in the picture and the definition is being limited by the detector, the right-hand side will be sharper than the left-hand.

To be on the safe side in design the circuit values should thus be chosen on the basis that the frequency response is dependent on $CR$. The values can then be chosen as for a v.f. stage, remembering that $C$ must not be too small since the input signal is applied through it to the diode.

The resistance $R$ usually has a value of 2,000–5,000 Ω and if a capacitor $C$ is used additional to the stray capacitances its value does not exceed 10 pF. In view of the low value of $R$, a low resistance diode is necessary if the efficiency is to be reasonable. Its interelectrode capacitances must also be low. Ordinary diodes of the type used in sound receivers are quite unsuitable.

Midget diodes are now available and are very suitable for television. One such is the Mazda D1, for which input-output curves

**Fig. 15.2—Input-output curves of a diode detector at two frequencies are given here**




for two different values of $R$ are given in Fig. 15.2. The values of $R$ are 2,000 and 3,000 Ω and are used with a total value for $C$ of 40 and 26·6 pF respectively, giving a constant value $CR$ = 80,000 pF–Ω. The curves are shown for input frequencies of 13 Mc/s and 45 Mc/s.

The anode-cathode capacitance for these curves is 1·5 pF. The efficiency decreases with a larger value and increases with a smaller. The valve capacitance is of the order of 1·5 pF, and so the capacitance cannot be reduced below this figure; every effort should be made to avoid external capacitance between anode and cathode, however.

In practice, the detector circuit is rarely as simple as in Fig. 15.1, for it is usual to include some form of low-pass filter. This is very necessary as there is quite a large radio-frequency voltage developed across $C$ which must, as far as possible, be prevented from reaching later circuits. This r.f. voltage is principally at the input frequency, but contains components at its harmonics.

In the straight set, only the fundamental is of major importance, but in the superheterodyne harmonics which fall near the signal frequency are of at least equal importance. It is essential, in the first case, for stability, that no appreciable part of the voltage across $C$ can find its way by stray couplings to the input circuits. In the second case it is equally important for the avoidance of interference patterns on the picture that no appreciable part of certain harmonic voltages across $C$ can find their way back to the input.

The usual circuit, therefore, is of the form shown in Fig. 15.3. The obvious thing to do is to design the filter on conventional lines as a low-pass filter terminated by the resistance $R$, which is the diode load resistance. The filter should then have a similar resistance in shunt with $C_1$.

If this is included, however, the detector efficiency will be low because the diode load will be effectively $R/2$. This resistance is ignored, therefore, and the effective diode output resistance is assumed to give about the right termination for the filter.

According to filter theory $L_1 = L_2 = R/\pi f$, $C_2 = 2C_1 = 2C_3 = 1/\pi f R$, where $f$ is the cut-off frequency. Suppose $C_3$ ( $= C_1$) = the

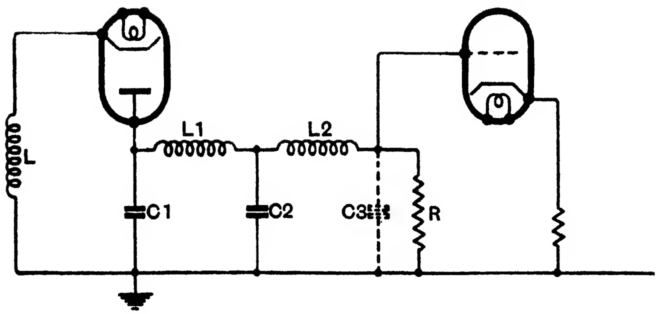



**Fig. 15.3—The connection between the detector and the v.f. stage is usually made through a low-pass filter**

input capacitance of the v.f. stage = 15 pF and $f$ = 2·5 Mc/s, then we have the following values for the components—

$$C_1 = C_3 = 15 \text{ pF} \qquad R = 4{,}240\ \Omega$$
$$C_2 = 30 \text{ pF} \qquad L_1 = L_2 = 540\ \mu\text{H}$$

The snag lies in the self-capacitance of the coils. It is hardly possible to make it less than 5 pF, and in view of the inductance value, with a normal compact design it is more likely to be 10 pF. With 5-pF self-capacitances, $L_1$ and $L_2$ become parallel tuned circuits resonating at 3·08 Mc/s, and with 10 pF at 2·175 Mc/s.

The attenuation is very large at this resonant frequency and it obviously invalidates the filter design. With the 10-pF self-capacitances there is a so-called frequency of infinite attenuation at 2·175 Mc/s, which is below the cut-off frequency of 2·5 Mc/s. With lower self-capacitance conditions are better and the resonant action improves the sharpness of cut-off.

At frequencies much higher than their resonant frequency, however, $L_1$ and $L_2$ cease to behave as inductances and it is only their self-capacitances which are important. The circuit then ceases to be a filter and becomes a capacitance attenuator; the actual " filter " circuit is then as shown in Fig. 15.4.

With self-capacitances $C'$ and $C''$ of 5 pF the attenuation is roughly 30 db at very high frequencies and about 20 db when they are 10 pF. Such attenuations are inadequate and it is usually better to choose values rather differently.

Exact design becomes unnecessary and much smaller inductance

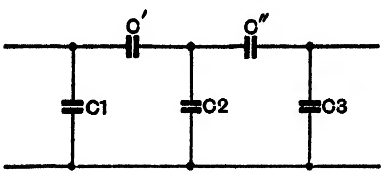



**Fig. 15.4—At high frequencies a low-pass filter is liable to become a capacitance attenuator**

values are used. The idea is to choose inductances such that they resonate with their self-capacitances at frequencies where high attenuation is important.

In a superheterodyne the intermediate frequency is important because it is desirable that the relatively large voltage of this frequency across $C$ should not be permitted to reach the v.f. valve. The signal frequency is also important, for any leakage of currents of neighbouring frequencies is likely to cause interference.

In general, the important frequencies are 13 Mc/s and 45 Mc/s. Assuming a coil capacitance of 5 pF an inductance of 30 $\mu$H will resonate at 13 Mc/s and give very high attenuation at this frequency. At 45 Mc/s the self-capacitance can be lower, say 3 pF, and the inductance can be 4·2 $\mu$H.

By making $L_1$ about 30 $\mu$H and $L_2$ some 4·2 $\mu$H very good filtering can be achieved in practice. The attenuation curve is not smooth and as there are two points of very high attenuation which we utilize so there are others of low attenuation.

Over the range of modulation frequencies the coils have little effect and $C_1$, $C_2$, and $C_3$ are effectively in parallel. $C_3$ is the input capacitance of the v.f. stage, say, 15 pF, and $C_2$ can be 10 pF. $C_1$ includes stray capacitances of, perhaps, 5 pF and an added capacitance of some 10 pF, making a total of 15 pF. In choosing $R$, therefore, the circuit capacitance should be reckoned as 40 pF, and so $R$ will normally be about 2,000 Ω only. Of course, if a sync separator is also fed from the detector output, its input capacitance will appear in parallel with that of the v.f. stage and the value of $R$ will have to be reduced.

The input resistance of a diode detector is usually given as $R/2\eta$ where $\eta$ is the efficiency; that is, the ratio of the d.c. output voltage to the peak r.f. input. It is doubtful whether this is more than an approximation when the capacitance across $R$ is so small and the anode-cathode capacitance is appreciable as in the case of television.

To a first approximation, however, the input resistance can be taken as $R/2\eta$, or about 2,100 Ω with the D1 diode when $R$ is 2,000 Ω. This must be taken into account when deciding the damping needed for the input tuned circuit, and in many cases no additional damping will be needed.

In some cases, and especially when the intermediate frequency is low, a push-pull detector is used on the lines shown in Fig. 15.5. The main advantage is that even harmonics of the input frequency are largely balanced out and less filtering is needed. There is also no loss of input voltage across the input capacitance of the filter, as with a single diode, so that the rectification efficiency is higher.

The total input voltage required, however, is twice that of the single diode, ignoring the higher efficiency for the moment, and the input i.f. coupling is more difficult to design. It must be a transformer and it is quite difficult to obtain the requisite inductive coupling without excessive self and mutual capacitances between the windings.

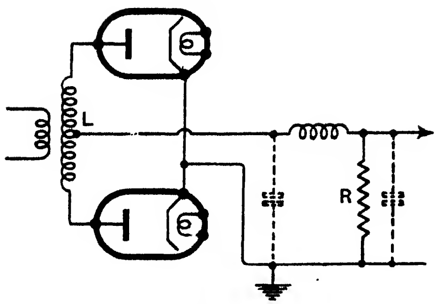



**Fig. 15.5—A push-pull or full-wave detector is sometimes employed**

Owing to the omission of the input capacitor to the filter and the simple filter which can be used the total capacitance in the output circuit is less than with a single diode. In a typical case it might be 25 pF instead of 40 pF. Consequently, $R$ can be higher; 3,200 Ω instead of 2,000 Ω.

The efficiency of each rectifier is higher owing to the absence of a voltage loss across the output capacitance and because the load resistance is higher. It may approximate to 70 per cent. The input resistance across each half of $L$ is thus of the order of 3,200/1·4 = 2,300 Ω approximately. The total input resistance across $L$ is thus 4,600 Ω.

With a single diode the peak input $E_1$ needed for an output $E_2$ (d.c.) is $E_1 = E_2/\eta$, whereas with the push-pull detector it is $E_1 = 2E_2/\eta$. This voltage must be developed across the input resistance so the input power is $P = E_1^2/2R_{in}$.

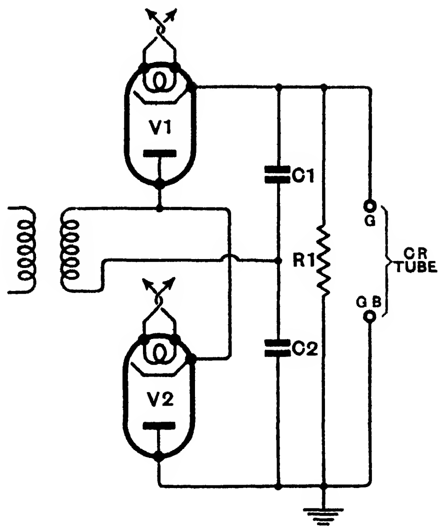
For a single diode we have $P = E_2^2/2\eta^2 R_{in}$, and for the push-pull detector $P = 2E_2^2/\eta^2 R_{in}$. In the former case we find that $\eta = 0{\cdot}465$ and $R_{in} = 2{,}100$ ohms in a typical case, while for push-pull $\eta = 0{\cdot}7$ and $R_{in} = 4{,}600$ Ω. Consequently, the powers for half- and full-wave rectification are likely to be $E_2^2/910$ and $E_2^2/1{,}130$ respectively.

The push-pull stage is thus more efficient, provided that advantage can be taken of its higher input resistance. This can only be done if the capacitance on the input transformer secondary can be reduced proportionately to the increase of $R_{in}$. If this cannot be done, the single diode is likely to be the more efficient, taking into account the input circuit as well as the detector.

The push-pull detector is not the only full-wave detector that can be used. It is quite possible to employ the voltage-doubler as shown in Fig. 15.6. In most cases this would only be adopted when the detector feeds the c.r. tube directly and the circuit is consequently arranged to give an output in positive phase. With a single v.f. stage, the connections to each diode would, of course, be reversed. No filtering is shown, but this is used in practice and is included in the output lead.

The output $E_2$ is approximately $2E_1\eta$ and the input resistance is $R_1/4\eta$. $C_1$ and $C_2$ are of equal capacitance and with strays total

Fig. 15.6—The voltage-doubler detector is occasionally useful




about 15 pF apiece. As far as the load circuit is concerned they are in series, so that the total capacitance is likely to be about 32·5 pF as compared with 40 pF for one diode. $R_1$ can thus be about 2,500 Ω.

The value of $\eta$ will be slightly higher in consequence, say, 0·5, making $R_{in} = 1{,}250\ \Omega$. The input power will be $P = E_1^2/R_{in} = E_2^2/4\eta^2R_{in} = E_2^2/1{,}250$. The efficiency is thus higher than with either of the other detectors.

Where correct matching can be carried out the improvement is not great and is not enough to justify the extra valve. There are cases where its use is worth while, however.

The chief one is where the detector feeds the c.r. tube directly and the output of a single detector is insufficient with the largest convenient valve in the last i.f. stage. The relative efficiency of the voltage-doubler is greatest when the circuit capacitance is high.

For instance, if the capacitance is such that a damping resistance of 1,250 Ω is necessary for the requisite bandwidth, the other detectors will compare very unfavourably with the voltage-doubler. But if the capacitance is low, as is often the case, then there is much less to choose between them.

The single diode meets the requirements of the average case admirably and it is consequently the most widely used.

Theoretically it is possible to combine the diode detector and the v.f. stage by operating a pentode as a grid detector. There are many disadvantages in so doing, however. In the first place, the input capacitance of the valve, some 15 pF, takes the place of the diode anode-cathode capacitance of about 1·5 pF, so that the efficiency is low unless the grid capacitor is large.

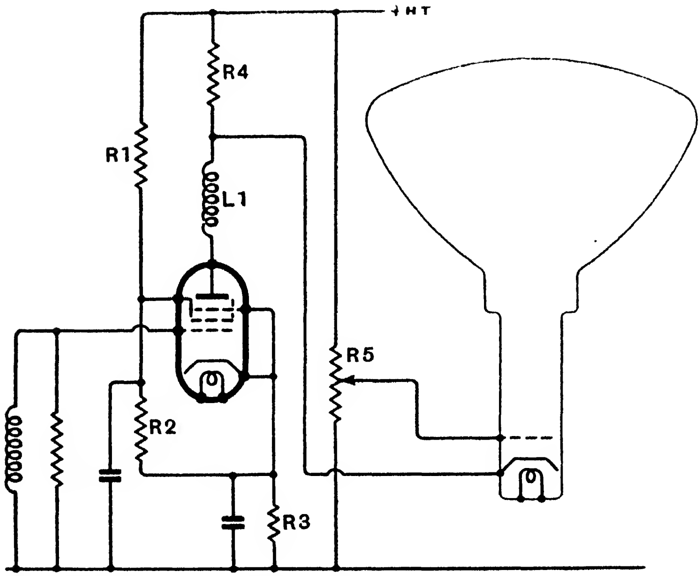



**Fig. 15.7—With an anode-bend detector the cathode of the c.r. tube can be fed directly from the detector anode**

The grid capacitor cannot exceed about 20 pF, however, if the frequency response is to be reasonably good. The grid leak should, of course, be about 2,000 Ω only, the same as for a diode.

The second disadvantage is that the full r.f. signal is applied to the pentode and the filtering must be performed in the anode circuit. More careful screening is thus needed on account of the probably greater amplitude of r.f. signal at this point, but the main drawback is that the signal loads the valve so that a smaller v.f. output is obtainable for a given anode current.

The maximum v.f. output is likely to be about one-half of that from the same valve used as a straightforward v.f. amplifier. This detector is thus rarely used.

The anode-bend detector, although not common, is sometimes adopted and its circuit values should be chosen in the usual way. The output is in negative phase and can consequently be fed to a single v.f. stage through an *RC* coupling, d.c. restoration being obtained automatically in the grid circuit of the v.f. stage.

In spite of the output being in negative phase, it can be fed directly to the c.r. tube if the input is applied to the cathode instead of the grid. This arrangement is shown in Fig. 15.7.

The anode circuit coupling components are chosen in the usual way relative to the stray capacitances. $L_1$ is a filter choke which can also be made to give correction for the stray capacitances, the circuit being equivalent to that of Fig. 9.4.

The screen and cathode potentials are chosen to give the correct rectifying action and the values of $R_1$, $R_2$ and $R_3$ thus depend on the valve selected. In general, the screen voltage should be about the normal value for the valve as an amplifier, and the cathode voltage sufficient to bring the valve nearly to the current cut-off point in the absence of a signal.

So far nothing has been said about correcting the frequency response of the diode load. It is common practice to include an inductance in series with the load resistance for this purpose. Its value is chosen just as in a v.f. stage to suit the non-conductive conditions of the diode. Thus, in the basic circuit of Fig. 15.1 the inductance is chosen to suit $R$ and $C$ just as if these were the components of a v.f. stage.

CHAPTER 16

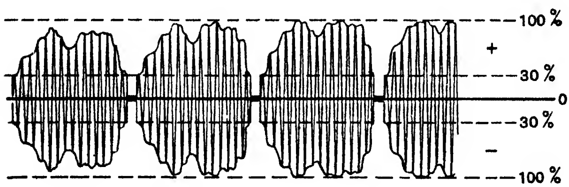
# Sync Separation

EXAMINATION OF THE waveform of the television signal shows that the carrier amplitude is varied between 30 per cent and 100 per cent of its maximum value for the picture signal and falls to zero regularly once every line for the synchronizing pulse. Other pulses are arranged to occur between frames for the frame synchronizing.

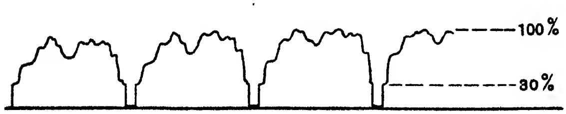
For successful operation it is necessary to separate the sync pulses from the picture signal so that it is impossible for the time bases to be affected by the latter. Some form of amplitude limiter is invariably adopted for this and is followed by circuits which separate the line and frame pulses by virtue of their different durations.

The arrangements used for sync separation vary greatly. It is possible to dispense with any special filter and arrange for the saw-tooth oscillators of the time base themselves to effect the separation of synchronizing and picture signals. On the other hand, as many as four valves are sometimes used solely for this purpose. Probably the most common arrangement is to use one or two valves.

Not only must the separator effectively discriminate against the picture signal, but it must meet certain definite requirements as to frequency response. It must provide output pulses in the correct form for the line and frame time bases and it must not introduce coupling between the two.




(a)




(b)

**Fig. 16.1—The r.f. signal is of the form shown at (a) while after detection the waveform is like (b)**

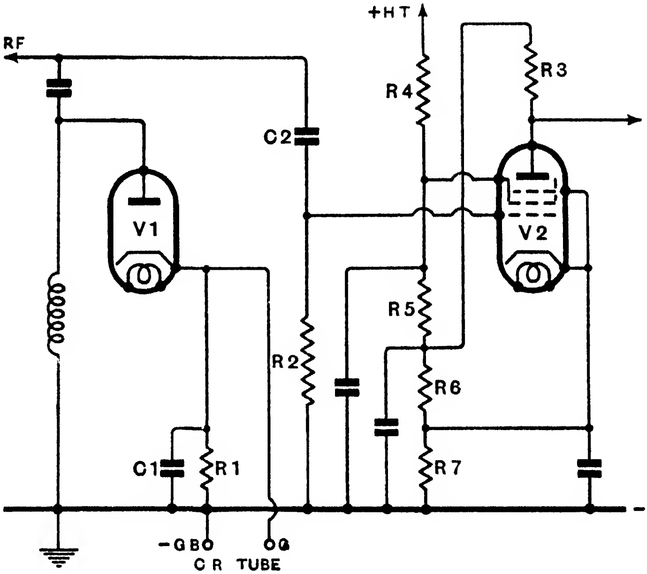



**Fig. 16.2—This diagram shows a method of sync separation employing an r.f. signal input**

The pulses can be separated directly from the r.f. signal. This method is not much used, but is included for the sake of completeness. The r.f. signal is of the form sketched in Fig. 16.1 (a), and the rectified output of the detector is shown at (b). This output is applied to the c.r. tube and is the voltage developed across the detector load resistance $R_1$ of Fig. 16.2. The r.f. signal is also applied to $V_2$ through the capacitor $C_2$, and the separated sync pulses appear across the anode-circuit resistance $R_3$. The other resistances $R_4$ to $R_7$ are merely to provide the correct steady potentials for the valve.

The valve is a screened tetrode or pentode and is operated with some 20–60 V applied to the screen and about 4–10 V supply for the anode. Under these conditions the anode current cuts off for only a few volts negative grid bias, but rises steeply as the bias is reduced. Saturation sets in at about $-1$ V grid potential, and the current tends to stay at a constant value as the grid potential becomes more positive.

The operating conditions are sketched in Fig. 16.3. The initial bias is chosen so that the valve is slightly beyond current cut-off. The negative peaks of the signal then only drive the grid further negative and cause no change of anode current. The signal amplitude is chosen so that with 30 per cent modulation the positive half-cycles just come beyond the upper bend in the characteristic. The picture modulation thus falls on the flat upper portion of the curve

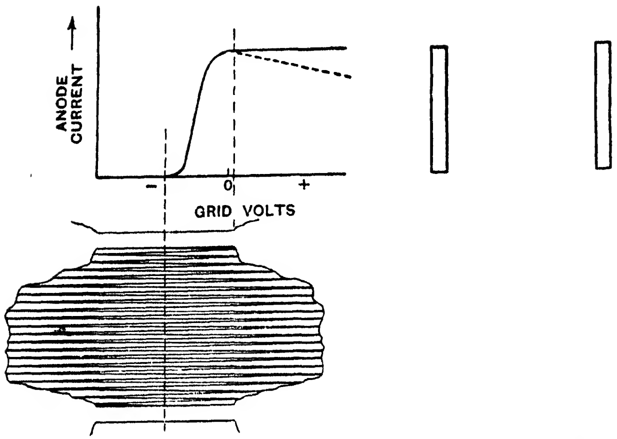



Fig. 16.3—The valve characteristic needed for an r.f. sync separator is shown here together with the input and output waveforms

and, ideally, causes no change of anode current. The sync pulses thus appear in the output free from any picture signal, as shown on the right of Fig. 16.3.

In practice, results are not perfect. The envelope of modulation is not solid but indicates only the peaks of the r.f. waveform. The separator limits the peaks to a constant amplitude and in so doing it distorts the waveform. The current waveform on each half-cycle is no longer a half-sine wave, but becomes of the form shown in Fig. 16.4, where (a) represents the case when the amplitude is only just greater than the value needed to bring it to the top of the bend in the valve curve. When the amplitude is large, however, the form is that shown at (b) and the mean current is obviously greater.

With this form of separator, therefore, the output is affected in some degree by the picture content. The effect can be considerably reduced by using a valve with a characteristic of the form shown dotted in Fig. 16.3.

Typical characteristics are shown in Fig. 16.5. They are for the Cossor MS/Pen with an anode supply of 5·05 V. Used with 27 V for the screen, curve A applies and the normal grid bias should be

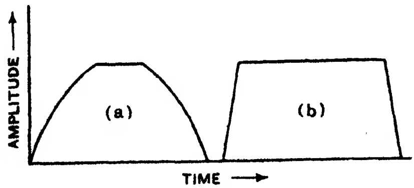



Fig. 16.4—A defect of the r.f. separator is that it distorts the r.f. waveform and the separation of the pulses is not completely independent of the picture content

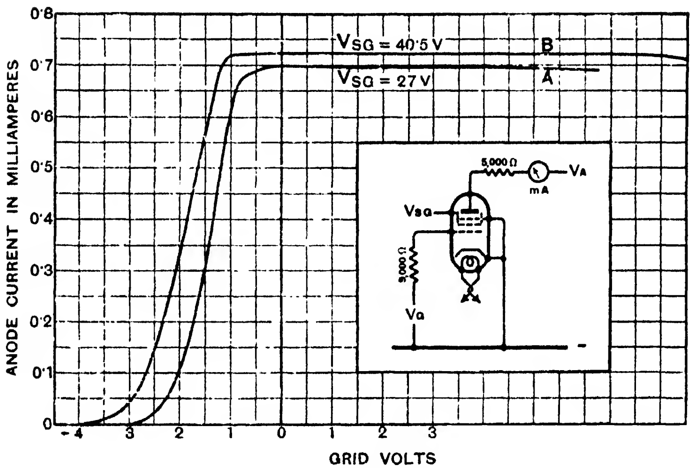



**Fig. 16.5—Curves of the Cossor MS/Pen for two values of screen voltage are shown here**

about $-$ 3·0 V. For good separation the signal input should not be less than 10 V peak on 100 per cent modulation.

The circuit is not often used in this form because it requires a larger r.f. input than is often available, it has a fairly low input resistance owing to the grid current flow, and it has an appreciable input capacitance. Furthermore, the separation is not perfect. It is, however, worthy of note that it is easy to obtain a sync pulse output in either phase merely by altering the grid bias.

As shown, the sync pulses appear in the form of positive voltage changes across the output resistance $R_3$ of Fig. 16.2, and this is in the form normally required by a time base. Negative pulses can be obtained by placing the carrier at the upper bend in the curve instead of the lower, so that the whole of the positive half-cycles of the input fall on the flat top and the negative half-cycles are utilized. With curve A of Fig. 16.5, zero grid bias would be used instead of $-$ 3·0 V.

This type of amplitude filter can be, and often is, operated with a v.f. input instead of an r.f. input. If the signal at the detector is sufficiently large the separator is connected across the detector output as shown in Fig. 16.6; the operation is clearly illustrated in Fig. 16.7, and it will be seen that little or no use is made of the lower bend in the curve. The separation is effected by the upper bend, and a flat curve for positive voltages is necessary.

Referring to Fig. 16.6, the resistance $R_2$ is important. If this resistance is not included the diode load $R_1$ is shunted by the input resistance of $V_2$ over the whole of the picture signal. As the grid may be 10 V or more positive, the input resistance may be only 1,000 Ω or so. The result is heavy attenuation of the picture signal. The inclusion of $R_2$ limits this action and it should be as

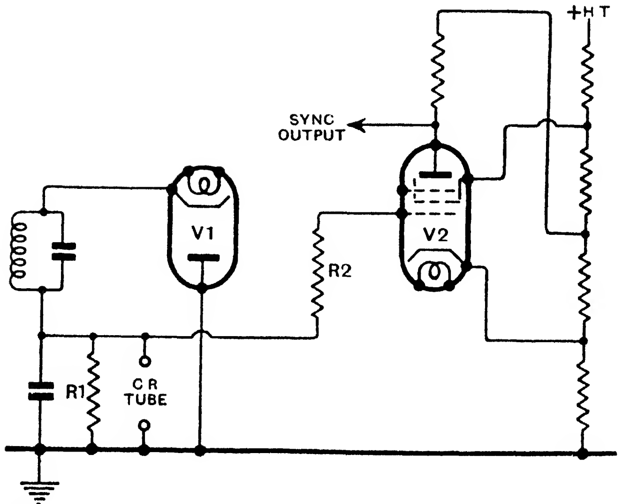



**Fig. 16.6—This diagram illustrates the use of a sync separator having a double bend in its characteristic with a v.f. input**

high as possible compared with the input resistance of $V_2$ and also considerably higher than $R_1$.

As $R_1$ is of the order of 2,000–5,000 Ω in most circumstances a good value for $R_2$ is 10,000 Ω. In general, it cannot be much higher than this if the requirements of frequency response are to be met. This resistance also helps the limiting very considerably. What happens is that $R_2$ and the input resistance of $V_2$ form a voltage divider across $R_1$, and the input to $V_2$ is only the voltage developed across its own input resistance. If $R_2$ is 10,000 Ω and the input resistance is 2,000 Ω, the grid volts of $V_2$ are only one-sixth of the volts across $R_1$. During sync pulses, however, when the grid of $V_2$ is negative its input resistance approaches infinity and the full voltage is applied to the valve.

This particular filter is a good one and operates very well indeed. Its drawback is that it requires a v.f. signal of at least 10 V p-p (for 100 per cent modulation of the carrier), and this is often greater than the normal detector output. When the c.r. tube is fed directly from the detector, it is probably the best amplitude filter to use. The phase of the output pulses is positive when the picture signal in the detector output is positive-going.

If the detector connections are reversed, so that the picture signal is negative-going, the same limiter can be used if its bias is changed to about $-1$ V only. It then removes the picture signal by anode current cut-off and it is no longer necessary to use a low anode voltage on the valve. The series grid resistance $R_2$ is also no longer needed. As the phase of the input signal is reversed the output now consists of negative-going pulses.

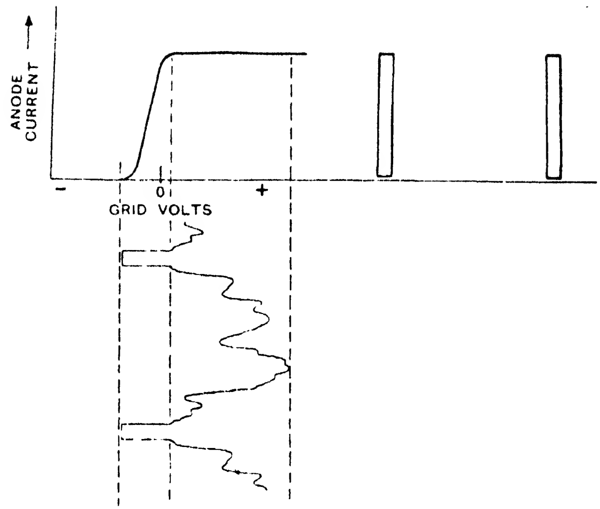



**Fig. 16.7—The operation of the separator of Fig. 16.6 is indicated in this drawing**

This arrangement is very reliable and satisfactory. The separation is somewhat better than when working on the upper bend, because the picture signal drives the valve beyond current cut-off. With upper-bend working equal results are only obtained when the curve has a flat top and this is rarely exactly the case, although a good approach to it can be obtained.

When more than a moderate degree of v.f. amplification is used the detector output is insufficient for satisfactory sync separation to be effected at this point. The separator must then follow the v.f. stage and certain difficulties then arise.

If proper sync separation is to be obtained it is essential for the d.c. component of the signal to be retained. Direct coupling between the detector and v.f. stage can be, and almost invariably is, adopted. Provided that there is no by-pass capacitor across the bias resistance and that the screen grid is not decoupled, the d.c. component of the signal will be reproduced in the anode circuit reasonably well. It will not be reproduced perfectly, however, unless a neon voltage stabilizer is used for the h.t. supply or the smoothing circuits are built to have a constant impedance at all frequencies down to, and including, zero. Nevertheless, in spite of the lack of perfection the d.c. component is retained sufficiently well for most practical purposes.

The need for retaining the d.c. component of the signal was pointed out in Chapter 9, and methods of doing it were there discussed in some detail. The matter is so important in sync separation, however, that some repetition is advisable.

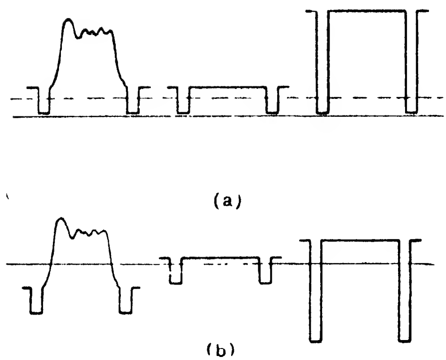



Fig. 16.8—Three typical signal waveforms are shown at (a), and the same signals with the d.c. component removed at (b)

The importance of the d.c. component can be realized from a study of Fig. 16.8, which shows at (a) three different picture waveforms with the full d.c. component. At the left is the signal during one line of typical picture, in the middle one black line, and at the right one white line. The bottoms of the sync pulses all rest on the same base line and it is possible to draw a line (shown dotted) through all to represent the cut-off point of a limiter, the line being such that the picture modulation always lies above it and that the amplitudes of the sync pulses below it are all the same.

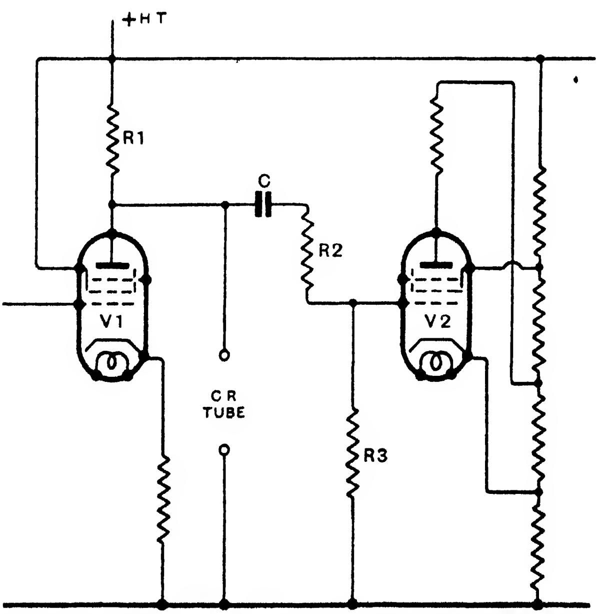



Fig. 16.9—The sync separator shown here does not work because the capacitor C removes the d.c. component of the signal

At (b) are shown the same three waveforms with the d.c. component removed and after several lines of the same waveform so that steady conditions have been reached. The steady condition is that the areas enclosed by the waveform on each side of the refernece line are equal. It is clear that it is now no longer possible to draw through all the sync pulses a straight line which will in every case leave all the picture signal on one side of it. It is, therefore, quite useless to feed a separator with a signal which has its d.c. component lacking.

Now if the v.f. stage is followed by a sync separator, such as that of Fig. 16.6, with resistance-capacitance coupling between the two, the performance will be quite unsatisfactory because the capacitor will remove the d.c. component. The circuit is shown in Fig. 16.9, and is unusable as it stands.

It is not possible in practice to omit $C$ and $R_3$ so that the grid of $V_2$ is joined directly to the anode of $V_1$ through $R_2$, because the stability of the h.t. supply is not good enough. Apart from this it would be quite a practicable course and it would only be necessary to choose the values of the resistances in the voltage-divider so that all the potentials applied to $V_2$ were more positive than usual by the amount of the no-signal anode voltage of $V_1$.

In practice, small fluctuations in the h.t. supply voltage are sufficient to move the characteristic of $V_2$ by seriously large amounts relative to the anode potential of $V_1$. The operating portion of the curve of $V_2$, for instance, may be shifted so that the sync pulses no longer fall across it or so that the picture signal falls on it.

Direct coupling must be ruled out and it is consequently fortunate that it is possible to " fake in " the d.c. component of a signal after a resistance-capacitance coupling. The arrangement is shown in Fig. 16.10, and it will be seen that a diode $V_2$, which is preferably

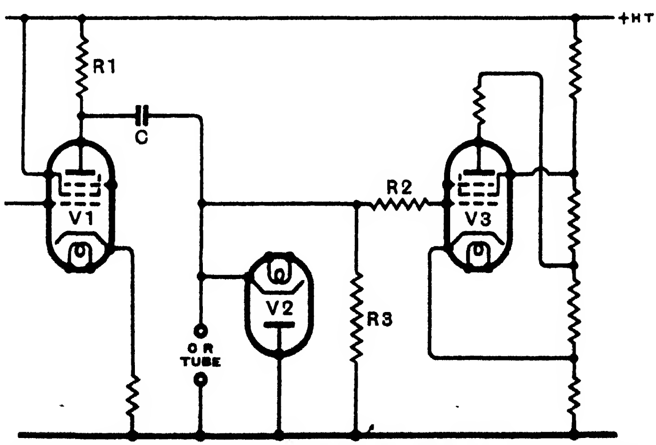



**Fig. 16.10—The circuit of Fig. 16.9 can be adopted if a d.c. restoring diode $V_2$ is added as shown here**

of the low-resistance type, is shunted across the grid leak $R_3$.

During the picture signal the diode cathode is driven positive and $V_2$ is non-conductive. During each sync pulse, however, the diode cathode goes negative and the diode conducts and charges the capacitor $C$. As a result of this current and the charge on the capacitor, the cathode of $V_2$ is left at a *positive* potential with respect to the earth line.

Now the amount of current, and consequently the cathode voltage after the sync pulse, depends upon how far the sync pulse has driven the cathode negative, and as can be seen from Fig. 16.8 (b), this in turn depends on the picture content. The sync pulse following a black line only drives the cathode negative by a small amount and the final positive voltage of the cathode during the following line is consequently small. The sync pulse following a white line, however, drives the cathode much more negative and it is correspondingly more positive during the following line.

The voltage developed across $R_3$ in this way is nearly sufficient to bring the bottoms of the sync pulses to the same level. There is some irregularity, it is true, but it is actually quite small. During lines, of course, the capacitor $C$ discharges through $R_3$ and the time-constant $CR_3$ must be properly chosen. Normal values are 0·1 $\mu$F for $C$ and 1 M$\Omega$ for $R_3$.*

There is still one difficulty with the arrangement of Fig. 16.10. The grid of $V_3$ is driven positive during the picture signal when $V_2$ is non-conductive. Grid current flows and tends to charge $C$ negatively. If $R_2$ were absent there would be, in fact, two diodes in parallel connected anode to cathode and cathode to anode; one would always be conductive and they could be replaced with a low value resistance, so far as performance is concerned.

Grid current in $V_3$ seriously affects the performance. Unfortunately, the grid current cannot be avoided and its effects must be minimized by making $R_2$ as large as possible. Reasonable values are $C = 0{\cdot}5\ \mu$F, $R_2 = 0{\cdot}5$ M$\Omega$, and $R_3 = 1$ M$\Omega$. Quite good results can be secured, but the d.c. restoration is imperfect and the high value of $R_2$, together with the input capacitance of $V_3$, makes the frequency response poor. This in turn makes the synchronizing liable to be affected by white objects on the extreme right of the picture. This particular circuit is usable but cannot be regarded as satisfactory for present-day needs.

However, if the output of the v.f. stage is negative-going on the picture signal, as it is when the signal is fed to the cathode of the c.r. tube, all the difficulties disappear. The separator valve limits by anode-current cut-off and so works under the best conditions. The signal at its grid is positive-going on the sync pulses and so the valve can provide good d.c. restoration without any special diode, for the grid-cathode path itself acts as a diode.

The circuit takes the form shown in Fig. 16.11 and is not only

* For a more detailed description of d.c. restoration see Chapter 9, and Appendix 4, page 356.

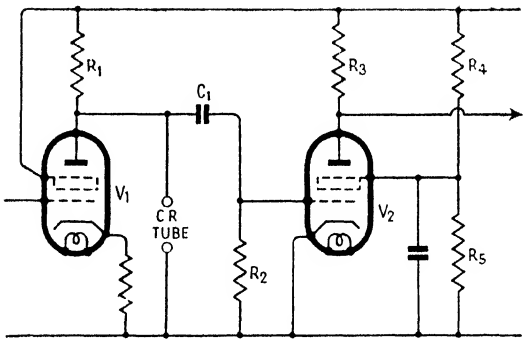



**Fig. 16.11—The basic circuit of Fig. 16.9 can take this form when the output of $V_1$ is positive-going on the sync pulses. It is one of the best and most reliable of sync separator circuits**

the simplest but is probably the best of all separator circuits. The coupling components $C_1$ and $R_1$ are usually 0·1 $\mu$F and 1 M$\Omega$ and the valve $V_2$ is generally a high-$g_m$ pentode, although this is not essential. The output coupling resistance $R_3$ can be as high as 50 k$\Omega$ and a negative-going pulse output of 80 V is usually obtainable. The pulse shape is better with a lower value of resistance but the output is less.

The cut-off point of $V_2$ is adjusted by the screen voltage which, in turn, depends on the values of $R_4$ and $R_5$. Adjustment is best made by observing the output waveform on an oscilloscope and varying $R_5$ and/or $R_4$ until all traces of the picture signal disappear and only the sync pulses remain. The adjustment should be carried out with $V_1$ providing a much smaller output than will normally be used, certainly not more than one-half of the normal output and preferably less. This is necessary to allow for variations of contrast-control setting, valve variations, and mains-voltage fluctuations.

It is not, however, essential to use tetrodes or pentodes for separating the sync pulses from the picture signal. Diodes can be used and there are two basic methods shown in Figs. 16.12 and 16.13 for the case when the v.f. signal appears across the detector load resistor.

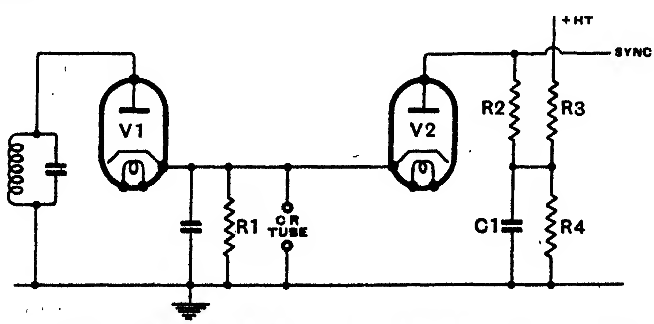



**Fig. 16.12—This diagram shows a diode sync separator following a diode detector**

With the first method, Fig. 16.12, $V_1$ is the detector and $R_1$ its load resistance, across which is connected the diode $V_2$ in series with the resistance $R_2$. A positive bias is applied to the anode of $V_2$ from the voltage-divider $R_3R_4$ across the h.t. supply; $C_1$ is of large capacitance.

With no signal $V_2$ passes current because of its positive anode voltage and there is consequently a voltage drop across $R_2$ which is nearly equal to the bias voltage. When a signal is applied $V_2$ remains conductive until the voltage across $R_1$ becomes about 1 V more positive than the bias. $V_2$ then becomes non-conductive and remains so no matter how much the voltage across $R_1$ rises. The voltage across $R_2$ consequently falls to zero.

The initial bias is adjusted to be of the same order as the sync pulse amplitude across $R_1$. The diode then conducts during most of the sync pulse but ceases to conduct on the picture signal. Current flows through $R_2$ on each sync pulse and develops a voltage across it. The phase of the sync pulses across $R_2$ is, of course, negative, the same as that of the pulses across $R_1$. If the detector output is in the opposite phase, the diode $V_2$ must be reversed, and a negative bias applied to its cathode.

The chief drawback to this circuit is the possibility of picture signals being passed to the output through the anode-cathode capacitance of $V_2$. This capacitance is not likely to be less than 1·5 pF and may be considerably higher. If $R_2$ is of high value, the picture signal voltage across it at high frequencies will be equal to that across $R_1$ multiplied by $c_{ak}/(C + c_{ak})$ where $c_{ak}$ is the anode-cathode capacitance of $V_2$ and $C$ is the total capacitance across $R_2$. Say $C = 10$ pF and $c_{ak} = 2$ pF, then the voltage across $R_2$ is one-sixth of that across $R_1$.

This represents very little attenuation and it means that although the separator may work well when the picture involves no very sudden changes of current, it may be ineffective under certain conditions. If the picture contains a white object on the extreme right-hand side, for instance, a very rapid change of amplitude from full white to black occurs at the end of the line and only slightly before the sync pulse. The change of current is in the same direction as the pulse and it will set up a voltage across $R_2$ because of the capacitance between anode and cathode of this valve, and this voltage is likely to be of the same order as that of the sync pulse. It is thus quite likely that this unwanted voltage will trip the time base before the sync pulse comes along 0·5 $\mu$sec later.

In practice, it is found that this type of diode separator is inclined to give trouble of this type. It can be used, of course, and will give good results if carefully designed. The capacitance across $V_2$ must be kept as small as possible, $R_2$ must be made as low as possible consistent with the required output, and the stray capacitance across $R_2$ should be as large as can be tolerated from the point of view of frequency response.

The alternative diode separator is shown in Fig. 16.13, and is

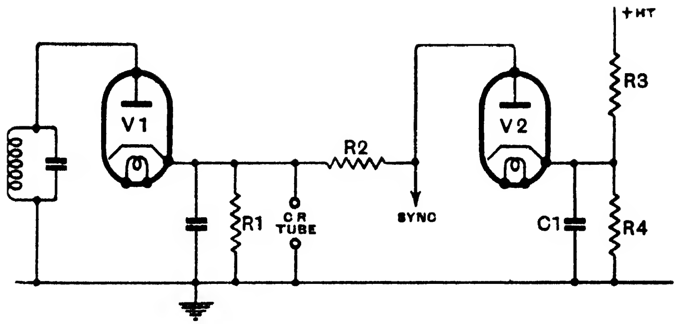



**Fig. 16.13 – A separator in which the diode $V_1$ conducts on the picture signals**

very similar, but $R_2$ and $V_2$ change places. The cathode of $V_2$ is biased positively so that $V_2$ is non-conductive until the voltage across $R_1$ rises above a certain value. The idea is that $V_2$ is non-conductive during the sync pulses, and these are consequently taken straight off $R_1$ through $R_2$, but that it short-circuits the sync output terminals during the picture signal.

In practice the short-circuit is by no means complete and the picture signal output at the sync terminals is the voltage across $R_1$ multiplied by $r_a/(R_2 + r_a)$. Now the diode resistance $r_a$ may be 1,000 Ω and $R_2$ about 10,000 Ω; so the output at the sync terminals is about one-eleventh of that across $R_1$. This is not large attenuation, but is sufficient for many purposes and it is free from the capacitance transference trouble of the alternative arrangement. In practice, it usually works better.

When a diode is used as a sync separator it is common not to rely entirely on the diode, but to follow it with a tetrode or pentode arranged to have a double bend in its characteristic as already described. The imperfections of the diode are then greatly reduced.

Direct-coupled diodes are sometimes used after a v.f. stage in spite of the difficulties brought about by variations of the h.t. supply voltage. The circuits of Figs. 16.12 or 16.13 can be used with the diode connected directly to the anode of the v.f. valve. The values of $R_3$ and $R_4$ are, of course, changed so that the potential of their junction is more positive by the no-signal voltage of the v.f. valve than it is in those circuits where the separator follows a detector.

These arrangements are open to the same objection as that of a tetrode or pentode separator with direct coupling to the anode of the v.f. valve, namely, that the operation is likely to be seriously affected by fluctuations in the h.t. supply voltage. The diode is rather less critical than other valves in its operating conditions and it is possible to use it in this way. Of course, if a stabilized h.t. supply is used there is likely to be little difficulty.

The difficulties caused by fluctuations of the h.t. supply can often be greatly reduced by using automatic bias on the diode instead of fixed bias derived from a potential divider across the h.t. supply.

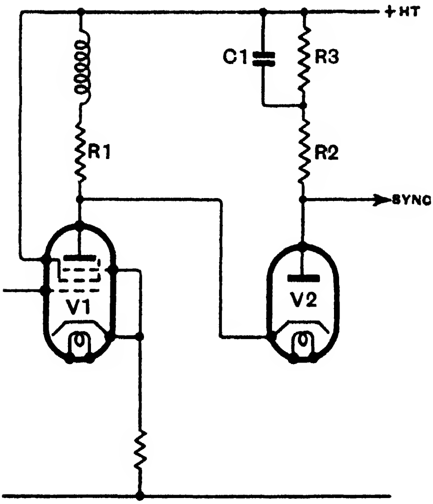



**Fig. 16.14—Automatic bias can be secured for a diode by means of this circuit**

One arrangement of this nature is shown in Fig. 16.14. In the absence of a signal the diode is conductive and its anode is at nearly the same potential as the anode of the v.f. valve. In the absence of the capacitor $C_1$, the diode would always be conductive no matter what form the signal variations took and the diode anode would follow in potential the pentode anode with only a slight difference in potential between them, perhaps 0·1–1 V.

When $C_1$ is in place, however, the potential across $R_3$ cannot change rapidly and if the time-constant $C_1R_3$ is large enough there will be a negligible change in voltage across $R_3$ during the time of the picture modulation of one line. It is easier to follow the action if the voltages are considered with respect to positive h.t. than if negative h.t. is taken as the datum, and this course will consequently be adopted in what follows.

Assume a total output of 30 V p–p from $V_1$, of which 9 V is sync pulse and 21 V picture modulation. The no-signal anode voltage may well be 100 V negative with respect to positive h.t. For the sync pulse, therefore, the anode voltage varies from − 100 to − 91 V, and for the picture signal from − 91 to − 70 V. In the absence of a signal the diode anode is at slightly less than − 100 V.

Now when a signal comes along the anode of $V_2$ rises in potential, but because of the high time-constant of $C_1R_3$, the voltage across $R_3$ cannot follow the change straight away. The diode anode consequently becomes negative with respect to its cathode and it becomes non-conductive. During the whole of the picture portion of the line $C_1$ is discharging through $R_3$ and the potential of the diode anode is rising in value. At the next sync pulse the anode potential of $V_1$ falls to its no-signal value and there is a flow of current through $V_2$ to recharge $C_1$. This current flows through $R_2$

also and sets up a voltage across it which forms the separated sync pulse.

After a few cycles an equilibrium condition is reached and the quantity of electricity which $C_1$ loses during each picture portion of a line is equal to the quantity gained during the sync pulse. The values assigned to the components must be such that the diode remains non-conductive over the whole picture portion of each line. That is, taking the figures given above as an example, as the black level is $-91$ V for the anode of $V_1$, the potential across $R_3$ must not become less than $-91$ V during the picture portion of a line.

The calculation of the best values of components is not particularly easy and it is usually simplest to determine them by experiment. Usually $R_2$ can be 5,000–10,000 Ω and $C_1$ about 0·5 μF. $R_3$ is made variable and adjusted for correct sync separation.

It should be noted that the variation in potential across $R_3$ appears at the sync terminal as well as the sync pulses. Further, the regularity of the voltage changes across $R_3$ is upset during the frame sync pulses, for then the diode is conductive for a longer time than it is non-conductive. During each series of frame pulses, therefore, the voltage across $R_3$ rises.

These effects are shown in exaggerated form in Fig. 16.15. The signals for the last two lines of a picture followed by two lines of framing pulses are shown at (a) and represent the voltage changes on the anode of the v.f. valve. The voltage output at the diode anode is of the form shown at (b). The slight rise in potential during the lines is due to the fall in voltage across $R_3$ and the reduction in pulse amplitude during the frame pulses is caused by the rise in voltage across $R_3$. Automatic bias is less perfect than fixed bias, but there is no doubt that it is the more suitable when the h.t. supply voltage is liable to fluctuation.

There is, however, one case where fixed bias can be used with less risk of trouble. This is when an anode-bend detector is used instead of the usual diode detector and following v.f. amplifier. This is shown in Fig. 16.16. The phase of the signal output is the

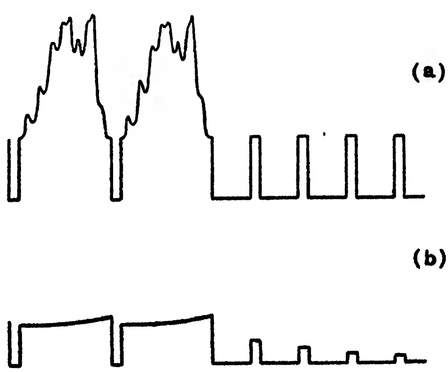



**Fig. 16.15—The voltage changes on the v.f. valve anode for line and frame pulses are shown at (a) and the separated pulses at (b)**

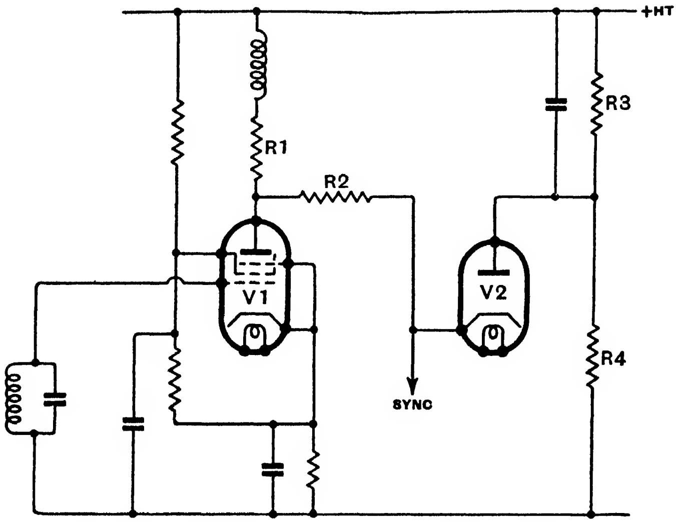



**Fig. 16.16—A directly-coupled diode separator following an anode-bend detector**

opposite to the usual, for the anode potential of $V_1$ moves in a positive direction on the sync pulses and in a negative on the picture modulation. A positive picture is secured by applying the signal to the cathode of the c.r. tube instead of to the grid.

The diode $V_2$ is the separator and is biased from the voltage-divider $R_3R_4$. The diode conducts on the picture signal and becomes non-conductive on the sync pulses, thus functioning in the manner described in connection with Fig. 16.9.

If in the absence of a signal $V_1$ is biased to the anode current cut-off point, there is no voltage drop across $R_1$ and the diode bias voltage across $R_3$ need only be small. In fact, it can be the same as if the separator were coupled directly to the output of a diode detector.

An increase in h.t. supply voltage increases the diode bias by the same percentage. Thus, if the diode bias is 3 V and the h.t. voltage rises by 10 per cent the diode bias increases to 3·3 V. For a bias of 3 V the normal amplitude of sync pulse across $R_1$ would be of the order of 6 V, so that good separation would still be obtained with the increased bias.

Taking these figures and remembering that a diode conducts until the anode is at about $-1$ V, the normal amplitude of sync pulse output would be $6 - (3 + 1) = 2$ V. When the h.t. voltage rises 10 per cent the output is $6 - (3{\cdot}3 + 1) = 1{\cdot}7$ V. If the h.t. falls 10 per cent the output is $6 - (2{\cdot}7 + 1) = 2{\cdot}3$ V. A change of supply voltage of $\pm$ 10 per cent thus causes a change of sync pulse amplitude of $\pm$ 15 per cent in this case, but has no other harmful effect.

Now it should be noted that a rise in h.t. voltage causes a rise in the screen voltage of $V_1$ and also in its initial grid bias. These two changes tend to counteract one another and to maintain the valve at the current cut-off point. They may not do so exactly, however.

It is instructive to consider at this point just why this type of circuit fails following a v.f. stage. Assume that the circuit is that of Fig. 16.14, but with an extra resistance between the junction of $R_2$ and $R_3$ and negative h.t. so that this resistance and $R_3$ form a voltage-divider across the h.t. supply. Let the h.t. supply be 250 V and the no-signal anode potential of $V_1$ be 150 V with respect to the negative line. For 30 V p–p output the anode potential of $V_1$ varies from 150 V to 159 V for the sync pulses. The diode $V_2$ should cease to conduct when the potential rises above, say, 155 V. The diode anode potential should be 154 V, allowing for the inherent diode bias of 1 V. The sync pulse output will be nearly 5 V.

Now if the h.t. voltage rises 10 per cent it becomes 275 V and the diode anode potential rises to 169·4 V. If the screen voltage of $V_1$ is kept constant, as with a neon stabilizer, the anode current is nearly constant, and the no-signal anode potential becomes $150 + 25 = 175$ V. The diode cathode is now never negative with respect to its anode and the sync pulse output falls to zero.

If the screen voltage of $V_1$ is not stabilized, the rise in screen potential causes an increase in anode current, which makes the voltage drop across $R_1$ greater, and so the anode voltage of $V_1$ does not rise as much. The increased current through the bias resistance increases the grid bias and tends to offset this. In practice, the conditions are better than with a stabilized screen supply, but they are not good.

It is not always easy to stabilize the h.t. supply, for it is usually about 250 V and the process of stabilization involves a loss of voltage of perhaps 100 V. What is sometimes done, therefore, is to use a valve for the v.f. stage which needs a screen potential of about 100 V only and this screen potential is stabilized. As the anode current of a tetrode or pentode is substantially independent of anode voltage, the anode current is automatically stabilized. Anode and screen current together form the cathode current which becomes unaffected by supply voltage variations and the grid bias developed across the cathode resistance is thus maintained constant.

The sync separator is now fed from the cathode of the valve. The voltage developed across the bias resistance is, of course, usually inadequate, for it is always less than the signal input to the valve, so an additional resistance is included. The arrangement is shown in Fig. 16.17; the c.r. tube is fed from the anode and the sync separator from the cathode, the output voltage being that developed by the cathode current across $R_1 + R_2$. The input is applied as shown and not between grid and earth. With the latter connections the initial bias would be too high and the output across $R_1 + R_2$ could not exceed the input.

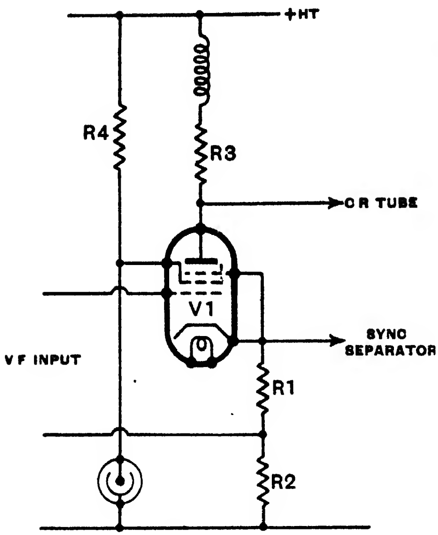



**Fig. 16.17—A v.f. stage with stabilized screen voltage is shown here; the c.r. tube is fed from the anode and the sync separator from the cathode**

An advantage often claimed for this circuit is that it relieves the picture signal circuits of the input capacitance of the sync separator. In practice, however, this is not true. As regards its cathode circuit the valve is acting as a triode and there is feedback through the capacitance between the control and screen grids. The effective input capacitance of the valve may be even higher than would be the case with the sync separator connected in parallel with its input.

This may be a little difficult to see at first, because the screen is maintained at earth potential to alternating currents, but it must not be forgotten that the input circuit is not at a fixed potential but is floating at cathode potential. Suppose the gain between the input terminals and the sync output is five times. Then if the grid potential changes 1 V with respect to cathode, the cathode changes 5 V in the same direction with respect to earth. The grid potential has thus changed 6 V with respect to earth and screen.

The voltage effective in driving a current through the control-grid to screen-grid capacitance is thus 6 V instead of 1 V and the current will be the same as that produced by the input of 1 V acting on a capacitance six times as large as the true capacitance. We thus say that the effective capacitance has been increased six times.

The effect is exactly the same as the well-known Miller effect in a triode, but it is apt to be overlooked because of the different point of connection of earth. The use of this form of cathode coupling increases the effective capacitance between control and screen grids by $1 + A$ times where $A$ is the ratio of the voltage across $R_1 + R_2$ to the voltage between grid and cathode. This is, of course, on the basis of pure resistance coupling; the presence of the inevitable stray capacitance across $R_1 + R_2$ will modify the magnitude and phase angle of the feedback somewhat.

If unity gain is obtained between input and cathode output circuits, the additional input capacitance due to feedback is equal to the control-grid–screen-grid capacitance. This is not likely to be less than 10 pF and is of the same order as the input capacitance of the sync separator itself. It is, therefore, as good to connect the separator directly to the detector and it is simpler because no stabilization of the h.t. supply is needed.

The conditions are different if the input to the v.f. stage is applied between grid and earth. The gain of the v.f. stage is then reduced by negative feedback and the cathode output voltage is less than the input. Capacitance feedback effects become small, however, and the effective input capacitance is reduced. The effect is similar to that in a cathode-follower.

Now whatever type of sync separator is used its output must be fed to the time bases. A direct connection cannot be used because there is usually a steady voltage which must be removed and interaction between the line and frame time bases must be avoided. With most saw-tooth oscillators grid current flows on the fly-back and sets up voltage pulses across the impedance in the grid circuit.

If the pulses set up by the line time base reach the frame time base it is liable to be synchronized by these pulses instead of by the synchronizing pulses proper. As the frame frequency is not a sub-multiple of the line frequency correct interlacing will not be obtained.

The best way of avoiding the interaction is to use two amplitude filters, one for frame and the other for line, so that the coupling circuits can be quite separate and distinct. When diodes are used for separation this is quite inexpensive for a duo-diode can be employed. When a tetrode or pentode is adopted the anode circuit is sometimes used to feed only one time base, and a resistance is included in the screen circuit so that the output for the other time base can be taken from this point. At one time Cossor used a special tetrode for this purpose with a split anode. There were really two anodes which could be connected independently to the two time bases. The use of separate synchronizing channels is not essential, however, for good synchronizing can be secured with a common amplitude limiter if the proper coupling networks are used after it.

A common practice is to feed the line time base from the sync separator through a differentiating circuit and the frame time base through an integrating circuit. Essentially, a differentiating circuit is a resistance-capacitance coupling which does not respond well to low frequencies, while an integrating circuit is one which does not respond well to high frequencies.

At this point it is essential to have a good idea of what happens when a square-topped wave is applied to such circuits.* The sync

* See also Appendices 2 and 3.

signals in the output of a perfect limiting stage are shown in Fig. 16.18; at (a) there are four lines of which the first two are normal picture lines at the bottom of the picture and the next two are lines containing the frame pulses for even frames. The different form of the signals at the end of odd frames is shown at (b).

The important thing to notice is that the voltage rises regularly at the end of every line, even during the time of the frame pulses, for the latter are broken up expressly to ensure this. Secondly, it is the leading edges of the pulses which actually trigger the time base.

Now if these waves are put through a differentiating circuit of very small time-constant only the vertical changes of voltage will be reproduced and the maintained horizontal voltages will disappear. The waveforms of (a) and (b) will be changed to the forms sketched in (c) and (d) respectively. There will, of course, also be a loss of amplitude, which is not shown.

The conditions shown in (c) and (d) are such that the pulse voltage falls substantially to zero in the time $t$ of one line pulse, 10 $\mu$sec. If the pulse has an amplitude of $E$ volts and the constants of the differentiating circuit are $R$ and $C$ then the voltage at the end of the pulse will be $e = E\epsilon^{-t/RC}$. If we arbitrarily make $e = 0{\cdot}011E$ then $t/RC$ is 4·5 and $RC = 2{\cdot}22 \times 10^{-6}$ F–$\Omega$ or 2·22

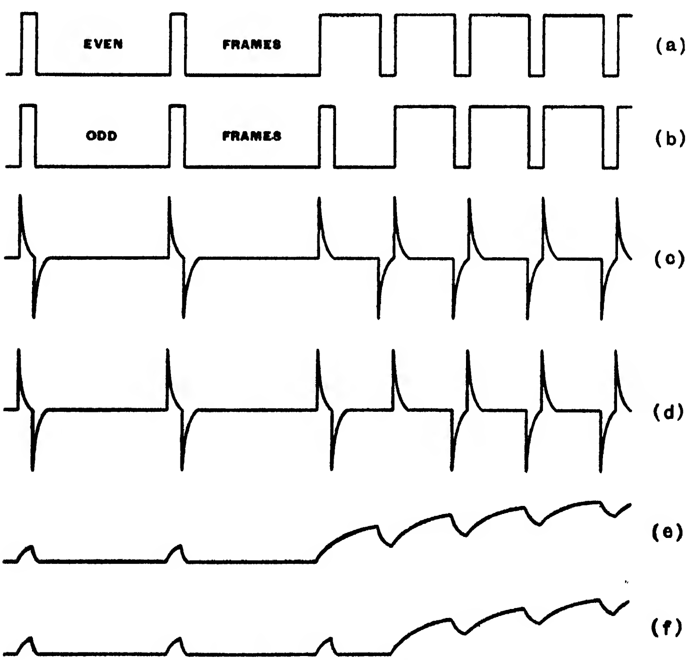



**Fig. 16.18—Line and frame pulses for even and odd frames are shown at (a) and (b). The effect of a differentiating circuit is illustrated at (c) and (d), while that of an integrating circuit appears at (e) and (f)**

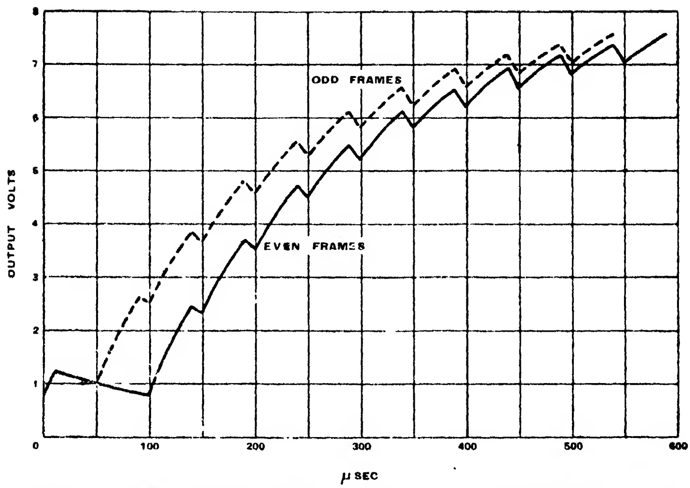



**Fig. 16.19—The response of an integrator to the frame sync pulses at odd and even frames is shown here for an input pulse of 10 volts amplitude**

pF–MΩ. In practice, the resistance might well be 50,000 Ω and the coupling capacitance should then be 44·4 pF, or say, 50 pF.

In the case of the frame circuit a series resistance and shunt capacitance are used to form an integrating circuit. This type of circuit responds only slowly to sudden changes of voltage so we find that the vertical voltage changes in the waveform are distorted and the horizontal maintained voltages are reproduced. The general effect of an integrating circuit is shown at (e) and (f) in Fig. 16.18 for even and odd frames respectively.

The line pulses are distorted and greatly reduced in amplitude. This is essential in order to prevent the frame time base from being fired by one of the last few line pulses. The frame pulses now appear in rounded form and with serrations corresponding to the line pulses.

This is shown more clearly by Fig. 16.19 in which the shape of the pulse is accurately indicated for both odd and even frames. The curves are for an integrator with 20-kΩ resistance and 0·001-μF capacitance with an input pulse amplitude of 10 V.

The peak output on a line pulse is 1·2 V, so that the frame time base must be arranged to trip at a higher level than this. Suppose this level to be 2 V, then it is clear that the time base will trip at 73 μsec and 127 μsec for odd and even frames after the onset of the last line pulse. The difference of timing is 54 μsec instead of 50 μsec. The line spacing is thus in the ratio 54 : 46 instead of 50 : 50.

If the triggering voltage is changed to 2·5, on odd frames this voltage is reached at 87 μsec, before the first serration, but on even frames it is reached at 154 μsec, *after* the first serration. The

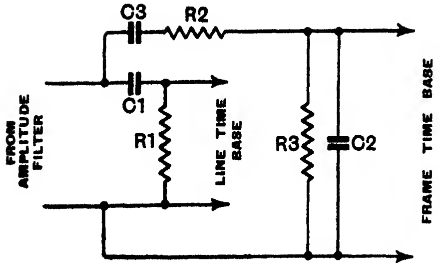



**Fig. 16.20—Typical differentiating and integrating circuits are shown here**

difference is 67 $\mu$sec and the line spacing is in the ratio 67 : 33 and there will be noticeable pairing of lines.

Matters are still worse at higher levels. Thus, suppose the tripping level is 6·5 V, on odd frames the time is 335 $\mu$sec and on even it is 387·5 $\mu$sec. The difference is 52·5 and the ratio is 52·5 : 47·5. This is good, but observe that if the tripping level changes to 6·6 V, the times become 337 $\mu$sec and 417 $\mu$sec. The difference is 80 $\mu$sec and the ratio is 80 : 20 and bad pairing results.

In practice matters are made worse by stray couplings from the line time-base output to the pre-integrator circuits for this introduces a line-frequency component, which may be of quite large amplitude, into the waveform.

For an integrator to give reliable interlacing it is necessary for the frame time-base tripping voltage to be below the first serrations. If the output of Fig. 16.19 is put through a slicer (double-limiter) limiting at 1·5 V and 2 V, then the maximum variation of timing will be about 52–55 $\mu$sec and the maximum pairing will be 55 : 45, which is not serious. Such a narrow slicing limit is hardly obtainable, however, and in practice it would be necessary to use at least three times the input. With 50 V into the integrator the limiting levels would be 7·5 and 10 V, a slicing width of 2·5 V, which is much more reasonable. With all limiters and slicers it is necessary to use as big an input as possible.

With extreme care in design it is quite possible to get good interlacing with an integrator, but great care must be taken in the avoidance of coupling between the time bases themselves. The combined circuit of differentiator and integrator, as often employed, is shown in Fig. 16.20. Here $C_1$ and $R_1$ form the differentiator and may have values of 50 pF and 50,000 Ω respectively.

$R_2$ and $C_2$ form the integrator and usual values are 20,000 Ω and 0·001 $\mu$F. $C_3$ and $R_3$ are to filter out any steady voltage from the amplitude filter and should be large. $C_3$ can usually be 0·1 $\mu$F and $R_3$ about 100,000 Ω.

The use of an integrator is not essential, however, and a particularly clever method of obtaining frame pulses with straight leading edges has been described by K. S. Davies* and is shown in

* *Journal of The Television Society*, December 1937.

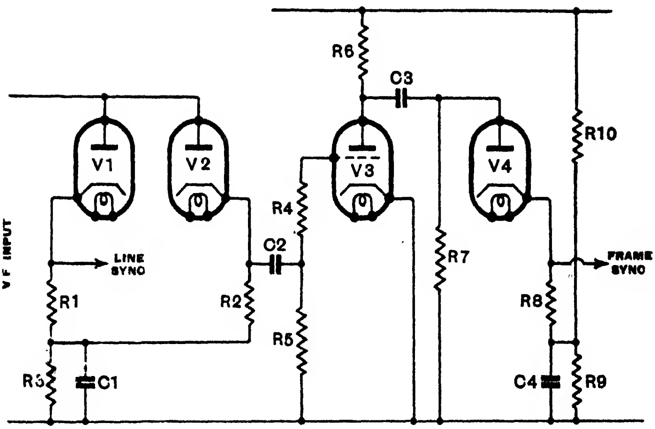



**Fig. 16.21—This type of sync separator is capable of giving a frame pulse with a straight leading edge**

Fig. 16.21. The two diodes $V_1$ and $V_2$ are amplitude filters for the removal of the picture signal; each has its own load resistance $R_1$ and $R_2$, and they are self-biased by the circuit $R_3C_1$. This portion of the equipment operates in the manner already described and sync pulses appear across $R_1$ and $R_2$. Those across $R_1$ are taken to the line time base through the usual differentiating circuit.

The voltages across $R_2$ are the same and two diodes are used only to avoid coupling between the two time bases. The phase of the output is positive so that the waveform across $R_1$ and $R_2$ approximates to that of Fig. 16.18 (a) and (b).

The output of $V_2$ is applied to the triode $V_3$ through the resistance-capacitance coupling $R_5C_2$. This valve is operated with zero grid bias so that the grid circuit operates under d.c. restoring conditions, and in effect the d.c. component is to a large extent retained. The voltages across $R_6$ are thus substantially a copy of those across $R_2$, but are amplified and reversed in phase. The voltages are then passed through a differentiating circuit $C_3R_7$ with a carefully chosen time-constant.

The effect of a small time-constant was illustrated in Fig. 16.18, and is unsuitable in this case. Actually the time-constant is made equal to the time of one framing pulse, 40 $\mu$sec.

The negative pulses appearing across $R_6$ are shown at (a) and (b) of Fig. 16.22, and the effect of passing them through a differentiating circuit of 40 $\mu$sec time-constant is illustrated at (c) and (d).

We saw that when the time-constant was small each line pulse was followed by another of equal amplitude and opposite phase. With this value of time-constant, however, the leading edge of the pulse causes a large pulse in the output, which is in this case negative, and because the loss of voltage during the pulse time is small, the following positive pulse is also small.

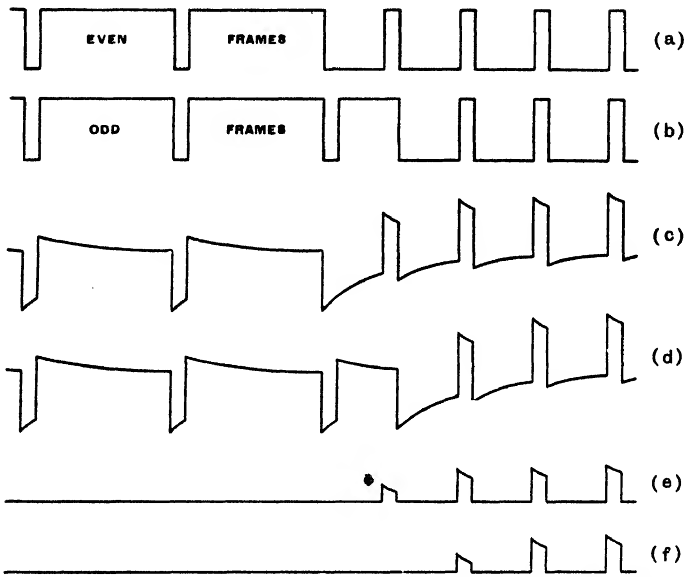



**Fig. 16.22—The sync pulses for even and odd frames are shown at (a) and (b) and the effect of a critical value of time-constant at (c) and (d). The result of using a further amplitude filter is illustrated at (e) and (f)**

During the longer frame pulses, however, the capacitance loses more of its charge and the voltage changes at the end of the pulses swing the potential positive. The intervals between the negative frame pulses thus appear as positive pulses of the same duration as line pulses.

These " interval pulses " are conveniently termed inverse frame pulses and it will be noticed that even the first is of greater amplitude than the positive pulse after each line sync pulse. The latter can consequently be removed by a limiter, leaving only the frame pulses.

This is the function of the diode $V_4$ in Fig. 16.21. It is biased by $R_9$, $R_{10}$, and $C_4$ so that it is non-conductive until the anode becomes more positive than the positive peaks after the line pulses. It becomes conductive on the more positive inverse frame pulses and so allows these to appear across $R_8$. The voltages across this resistance are then of the form shown in (e) and (f) of Fig. 16.22.

It is clear that it is only necessary to see that the amplitude is sufficient for the time base to fire on the first inverse frame pulse for accurate timing to be secured. All pulses have sharp leading edges and the first pulses at the ends of the frames are displaced by one-half line in alternate frames.

The primary requisite for success with this system is sufficient output from $V_3$ for the following limiter $V_4$ to work without its adjustment being too critical. The positive peaks against which $V_4$ must discriminate have an amplitude of about 20 per cent of the total pulse amplitude across $R_6$ while the minimum amplitude of inverse peak pulse is about 50 per cent. There must thus be sufficient voltage difference between the 20 and 50 per cent levels for the diode to change over from the conducting to the non-conducting conditions with reliability. Variations between valves and fluctuations in the bias voltage must be taken into account.

The precise cut-off point of the diode may well be taken half-way between the two levels, or at 35 per cent of the initial pulse amplitude. Suppose we allow 1·5 V between the 20 per cent level and the cut-off point of the diode. Then the sync pulse amplitude across $R_6$ must be 10 V. The positive peaks on the diode after the line pulses will be 2 V and on the inverse frame pulses about 5 V minimum. The diode cuts off at 3·5 V, so that the frame pulse output will be about 1·5 V.

Owing to the inherent diode bias of about 1 V, the actual bias developed across $R_9$ must be about 4·5 V. These operating conditions should afford reasonable tolerance for practical variations, but a larger input to $V_4$ would be better. The time-constant $C_3R_7$ must be 40 $\mu$sec and is well met with $C_3 = 0{\cdot}001\ \mu F$ and $R_7 =$ 40,000 Ω.

Other circuit values are straightforward, coupling resistances being chosen for the maintenance of a sharp leading edge to the sync pulses, just as in a line circuit. $C_2$ and $R_5$ can have d.c. restoration values, say, 0·1 $\mu$F and 1 MΩ, and $V_3$ must give sufficient gain to develop at least 10 V p–p of sync pulse across $R_6$ from the voltage across $R_2$.

It may be remarked that it is not necessary to employ a diode for the frame limiter $V_4$; any other type of limiter can be used. The arrangement of Fig. 16.21 gives positive-going frame pulses in the output. If the time base requires negative-going pulses for synchronizing, and some do, the simplest way is to use a pentode-type limiter in place of $V_4$, for this will give the required phase-reversal as well as the limiting action.

One separator of this type is shown in Fig. 16.23. The input terminal is joined to the cathode of the c.r. tube and to the v.f. stage anode, so that the sync separator is fed with the same input as the tube—an input which is negative-going on the picture signal. D.c. restoration is effected by $C_1R_1$ in conjunction with the grid-cathode path of $V_1$. This valve also acts as the main sync separator, being operated with a fairly low screen voltage and being conductive only during the sync pulses. A fairly high-$g_m$ valve, such as the SP41 or EF50, is best.

The output is developed across $R_2$ of 10–20 kΩ and supplied through $C_2R_3$ to the grid of $V_2$. The coupling has a time-constant of about 40 $\mu$sec so that the grid voltage has the form shown in Fig. 16.22 (c) and (d); $C_2$ can be 200 pF with $R_3$ 0·2 MΩ. The valve $V_3$ is the frame pulse separator and is heavily biased so that it is normally cut off but becomes conductive on the tips of the inverse frame pulses. A 6J7 or similar valve is quite suitable.

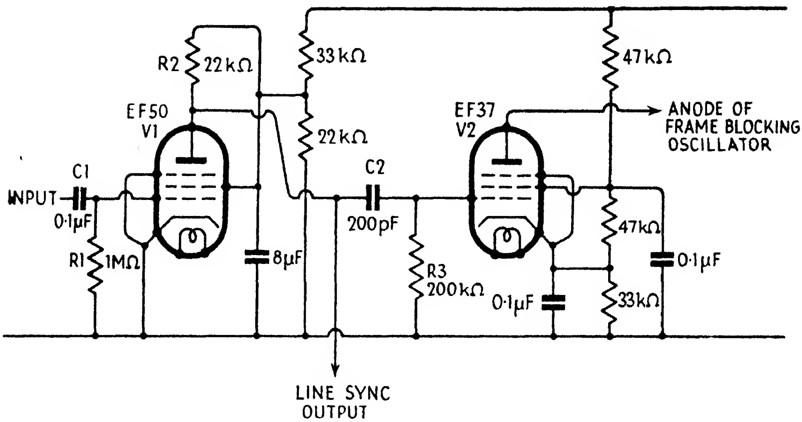



**Fig. 16.23—An inverse frame-pulse separator $V_2$ using a pentode. $V_1$ is the main sync separator and provides line sync pulses as well as feeding $V_2$**

Another method of frame-pulse separation is shown in Fig. 16.24. $V_1$ is the main sync separator and functions just like $V_1$ of Fig. 16.23. The positive-going sync pulses at the anode are applied to the first diode $D_1$ in $V_2$. Between pulses this diode is conductive, anode current is cut off in $V_1$, and $C_2$ charges to the h.t. voltage. $D_2$ is cut off because its anode is returned to a point less positive than the anode of $D_1$ and therefore one which is negative to the cathode.

On every sync pulse $D_1$ is cut off for the duration of the pulse because the anode voltage of $V_1$ and $D_1$ drops below the cathode potential of $D_1$. During the pulse $C_2$ discharges through $R_3$. The time-constant is chosen in relation to the anode voltage of $D_2$ so that the cathode voltage does not fall sufficiently during a line pulse for $D_2$ to become conductive. After every pulse $C_2$ is recharged again through $D_1$.

During the longer line pulses the voltage across $C_2$ falls sufficiently for $D_2$ to conduct. When this happens current flows through $D_2$ and also through $L$ which is a winding on the frame blocking-oscillator transformer. This current develops an e.m.f. in $L$ which is induced into the other windings (not shown) and trips the time base. The cathode circuit of the diode is, in practice, given a time-constant of about 55 $\mu$sec, typical values being 250 pF and 22 kΩ.

Although the circuit may at first sight appear similar to an inverse frame-pulse separator it does not actually utilize the back edge of the pulse and is more nearly related to an integrating circuit.

The question of the frequency response must now be considered. The signal waveform of one white line followed by one black line is shown by the solid lines in Fig. 16.25. It will be seen that after the sync pulse, lasting 10 per cent of the line time, there is an interval

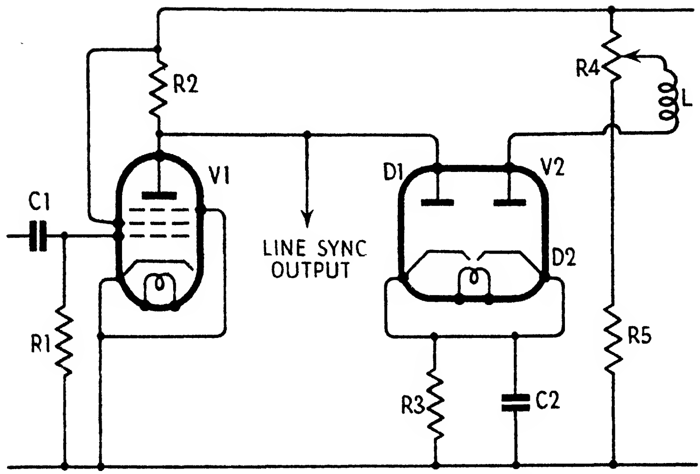



**Fig. 16.24—A frame-pulse separator depending on a time-constant $C_2R_3$, about equal to the frame-pulse duration**

at black level of 5 per cent of the line time. After the picture signal and before the next sync pulse there is another interval of 0·5 per cent of the line time.

For proper sync separation it is necessary for the frequency response of the circuits *preceding* the separator to be good enough to allow the signal to change from full white to black in 0·5 per cent of the line time. When the circuits are not good enough, the performance is affected in a manner roughly indicated by the dotted lines of Fig. 16.25.

At the time of the sync pulse following the white line the voltage has not reached black level. At the beginning of the pulse, the rate of change of voltage increases, but the voltage does not fall below black and affect the amplitude filter until some time after the start of the pulse.

After a black line, the voltage falls at once, but as the sync

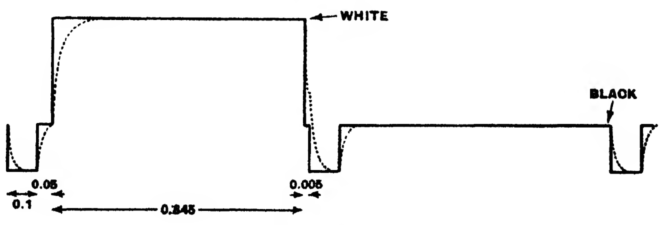



**Fig. 16.25—The solid line shows the waveform corresponding to a white and a black line, while the dotted curve illustrates the effect of an inadequate high-frequency response**

separator usually starts to work at some distance below black level, there is still a time delay. The delay, however, is smaller than after a white line.

It is thus clear that an inadequate frequency response affects the synchronizing by introducing a time delay in the effective part of the sync pulse. Moreover, when the inadequate response occurs before the main limiter the magnitude of the time delay depends upon the picture content at the end of the lines; that is, on the extreme right-hand side of the picture.

The visible effects will be a shifting of the lines sideways according to the picture content. The effect is quite noticeable in practice and rather irritating. It is often called "pulling on whites."

An inadequate frequency response *after* the limiter is much less serious, for all sync pulses are affected equally. The only result, therefore, is a slight delay in the commencement of the line fly-back. It should be noted, however, that the rounded pulse is not as good as the square-faced type for triggering many time bases, and furthermore, the effective time delay will vary if, for any reason, the amplitude of the sync pulses delivered to the time bases varies. It is, therefore, a wise plan to aim at maintaining the square waveform.

The first requirement is that a change from white to black in the signal must be capable of being substantially completed in 0·5 per cent of the line time or some 0·5 $\mu$sec. If we assume that 99 per cent of the change is completed in this time, a single *RC* circuit could have a time-constant not exceeding 110,000 pF–$\Omega$. Such a circuit gives a response of $-$ 3 db at 1·4 Mc/s.

We can say, therefore, that from the point of view of sync separation the frequency response of the circuits before the amplitude filter should extend up to about 1·5 Mc/s. In most cases, of course, the response is maintained well above this figure for the pre-sync-separator circuits are the picture-signal channel. The necessity for this response must be borne in mind, however, in cases where there is a separate channel feeding the limiter.

In the circuits following the sync separator the response need not be as good, but as there are often several circuits involved it is wise to keep the *RC* products of each coupling of the order of 100,000

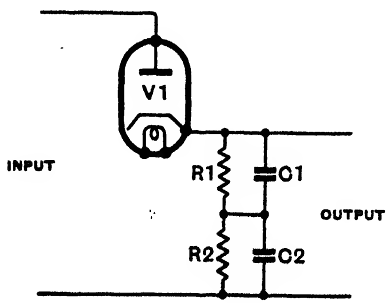



**Fig. 16.26—With a diode sync separator the effective time-constants are different for the leading and trailing edges of the pulses**

pF–Ω. With stray capacitances of the order of 20 pF. This leads to the use of coupling resistances of about 5,000 Ω.

In the case of a coupling following a diode, the conditions are rather different. Values chosen in the way just described will still be satisfactory from the point of view of frequency response, but when a diode is involved it is sometimes possible to use a higher time-constant for the coupling components without harmful effects.

When the diode circuit is of the type shown in Fig. 16.26, where $V_1$ is conductive on the sync pulses only, the bias being provided by $R_2C_2$, the effective resistance is less when the valve is conductive than when it is non-conductive. $R_1$ is the load resistance and $C_1$ represents the stray capacitances. At the beginning of a sync pulse the diode is conductive and the time-constant is $C_1R_1r_a/(R_1 + r_a)$. When the diode ceases to conduct at the end of the pulse, however, the time-constant becomes $C_1R_1$.

The a.c. resistance of the diode, $r_a$, may be as low as 1,000 Ω, and when it is, no value of $R_1$ can make the time-constant high enough to distort the leading edge of the pulse appreciably, assuming a normal value for the stray capacitance. At the end of the pulse, the rate at which it can die away is governed by $C_1R_1$, and if the product is too large trouble will be experienced. Distortion of the shape of the pulse is much less serious at its end than at its beginning, however, and it is quite reasonable to make $R_1$ several times as large as one would do if it affected the leading edge of the pulse.

A similar effect also occurs with a pentode when the grid voltage is a positive-going pulse and here also it is possible to increase the value of the coupling resistor.

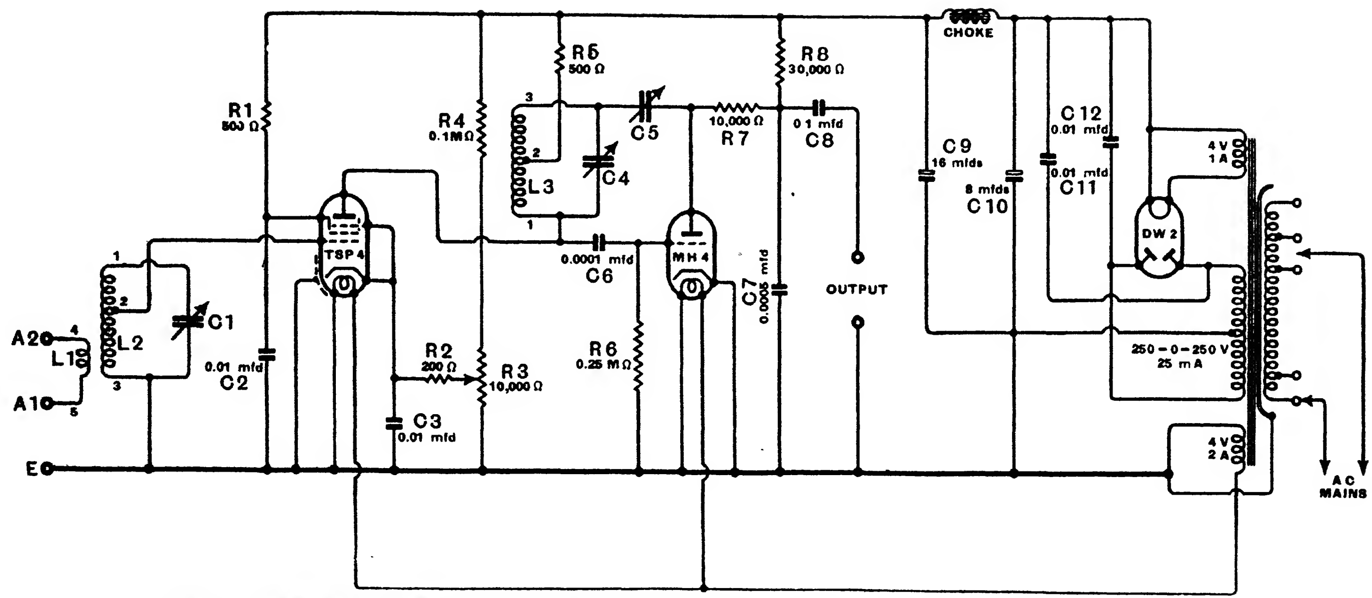



Fig. 17.1—This diagram shows the circuit of a simple adaptor for sound reception. The output is connected to the pick-up terminals of any standard broadcast set

CHAPTER 17

# Sound Reception

THE SOUND ACCOMPANIMENT to vision is radiated on a frequency 3·5 Mc/s lower than that of the vision channel. The necessary receiver can be entirely separate from, or integral with, the vision receiver. The latter course is usually adopted in commercial apparatus on the grounds of economy, but a separate receiver is often convenient in experimental equipment.

For reception over moderate distances a straight set is entirely adequate and is inexpensive. Up to some ten miles a reacting detector with a triode a.f. stage feeding into either a single triode output valve or a pair in push-pull is adequate.

At greater distances one s.f. stage should be added, and for experimental purposes a two-valve unit feeding into the pick-up terminals of any broadcast receiver is often useful.

With a good a.f. amplifier very high quality reproduction can be secured. Owing to the high frequency of the signal a large amount of reaction can be used without affecting quality; using only a moderate amount, the selectivity is high enough to avoid interference from the vision signal.

The circuit diagram of such a unit is given in Fig. 17.1 and is of a unit built by the author many years ago. It is still satisfactory for its purpose but, of course, one would now use more modern types of valve. The SP41 or EF50 would be now more suitable for the s.f. stage and a choice between them would be dictated largely by whether it is more convenient to use valves with 4-V or 6·3-V heaters.

At extreme distances the superheterodyne will probably become advisable. One s.f. stage, frequency-changer, two i.f. stages, diode detector, and one a.f. valve before the output stage will normally be adequate at any point where the vision signal can be received at useful strength.

The i.f. circuits should not be made too selective, otherwise serious difficulty from oscillator drift and microphony is likely to be experienced. Perfect oscillator stability is very difficult to achieve and the normal amount of drift will prove very troublesome unless the selectivity of the i.f. amplifier is considerably lower than is usual in a broadcast receiver.

It is, therefore, advisable to legislate for a bandwidth of the order of 50 kc/s in the sound i.f. amplifier. The circuits used should consequently be of fairly high inductance and low capacitance, so that fairly high gain per stage can be secured. There is no need to strive for very low capacitance circuits, as in the case of the vision amplifier, and it is quite satisfactory to use trimmers of some 50-pF capacitance.

When separate superheterodynes are used for sound and vision the oscillator and intermediate frequencies must be carefully chosen if interference between the two receivers is not to occur. The receivers will usually be close together and it will be very difficult to provide adequate screening for the prevention of interference.

The possibilities of mutual interference are so great when two separate superheterodynes are used in close proximity that this arrangement should be avoided whenever possible. It is usually simpler to adopt commercial practice and combine the two receivers with common s.f. and frequency-changer stages.

When only one or neither of the receivers is a superheterodyne there are few difficulties, but it will usually be necessary to employ a common aerial and feeder. One method of connecting otherwise independent sets to a common aerial is shown in Fig. 17.2. Rejector circuits $L_1C_1$ tuned to the sound signal and $L_2C_2$ tuned to the vision signal are inserted in the connections to the vision and sound receivers respectively. As they are inserted in a low impedance circuit, the rejectors should have a fairly low $L/C$ ratio, and a capacitance of the order of 200 pF will usually be about right.

Although convenient in experimental apparatus, it is not common practice to use separate receivers. Nearly always the vision and sound receivers are in some degree combined. In a straight set the first two s.f. stages and in a superheterodyne the s.f. and frequency-changer stages, and sometimes the first i.f. stage also, are common.

The intervalve couplings of these stages are given a bandwidth covering both vision and sound channels, and they amplify the signals of both channels simultaneously. There are three possible cases, depending on whether the vision channel is single- or double-sideband and, if the former, on whether the selected sidebands are those adjacent to or remote from the sound channel. With double-sideband reception of Alexandra Palace a total band of 41·5 Mc/s to 47·75 Mc/s is needed for the common stages. With single-sideband reception of the remote sidebands the same band is

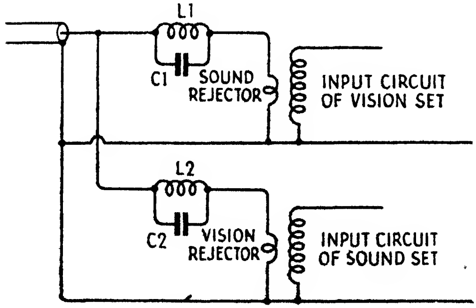



**Fig. 17.2—A method of connecting vision and sound receivers to a common aerial through rejector circuits is shown here**

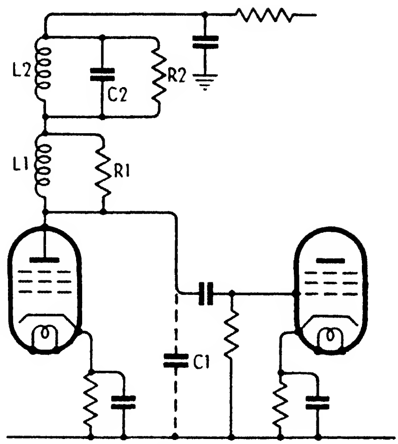



**Fig. 17.3—An i.f. coupling containing circuits tuned to both sound and vision channels**

satisfactory and usually adopted. There is, however, the alternative of providing a double band, one of 45 Mc/s to 47·75 Mc/s and another narrow one centred at 41·5 Mc/s. With single-sideband reception of the adjacent sidebands the band needed is only 41·5 Mc/s to 45 Mc/s, actually only about 0·75 Mc/s more than the vision band itself. This last case is the one which will be general in the future.

The usual practice is to provide a single wide band for all circuits which are common to both channels. There are some receivers, however, in which the double band referred to above is included.

One such arrangement is shown in Fig. 17.3; the normal coupling for the vision channel is $L_1$ tuned by the stray capacitance $C_1$ and damped by $R_1$. The additional circuit $L_2C_2$ is tuned to the sound channel and is independently damped by $R_2$.

The circuit is advantageous when the vision-channel band is considerably spaced from the sound channel, for then the normal bandwidth of the vision channel need not be extended to include the sound channel and higher gain is obtained. In addition, it is possible to secure higher gain from the sound channel than from the vision if this is desired. In general it is not, since it increases the risk of cross-modulation. Between the two channels there is a trough of low response, so that the response curve of the coupling has the form of Fig. 17.4.

It is not usual to have more than two amplifying stages common to both channels for two reasons. In the first place, such common stages are satisfactory only at low signal levels; when the signal amplitude becomes too great cross-modulation is very likely to occur. Secondly, after the signals are separated into individual channels these independent amplifiers must each have sufficient selectivity to eliminate the signals of the other channel. This is always easier to achieve when there are many couplings than when there are only a few.

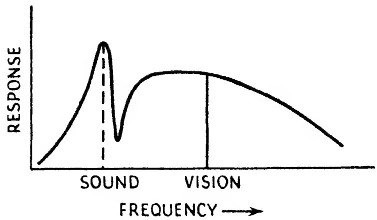



**Fig. 17.4—The general form of the response curve of the circuit of Fig. 17.3**

One method of separating the signals is shown in Fig. 17.5. This is very similar to the coupling of Fig. 17.3, but only the voltage developed across the sound circuit is applied to the sound channel. If this circuit $L_1C_1$ is tuned to the sound frequency, the maximum signal is obtained in the sound channel but it is also applied to the vision channel which is undesirable. It is often better, therefore, to tune this circuit for minimum sound signal in the vision channel. This represents quite a small amount of mistuning so that the loss of sound signal is small, but it helps greatly in rejecting the sound signal from the vision channel.

Another method sometimes used is to feed the sound channel from a cathode-circuit sound-channel rejector fitted to the vision channel. Thus, in Fig. 11.5 the sound channel can be fed from the cathode of the valve.

In nearly all these separating circuits the arrangement is one which looks as though the sound circuit would also act as a sound rejector in the vision channel. However, it is found that the tuning

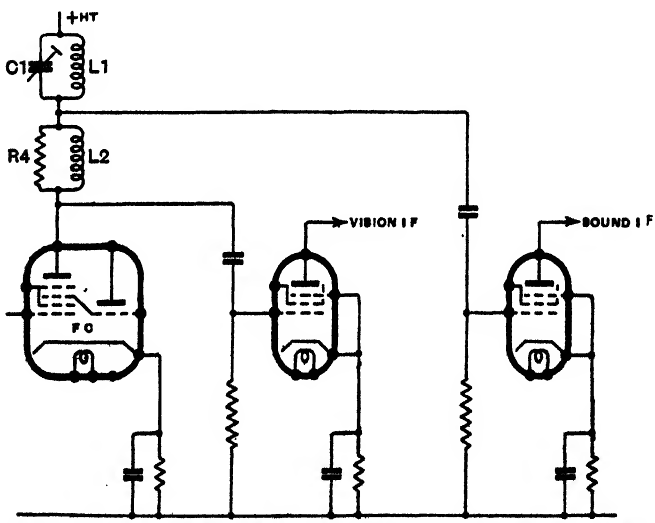
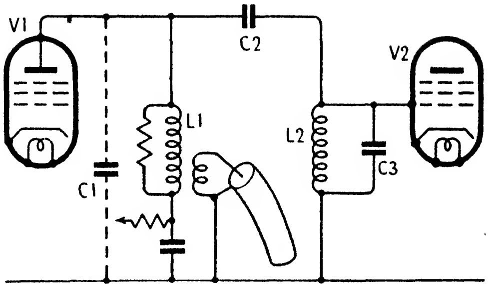



**Fig. 17.5—The frequency-changer is common to both sound and vision, and the signals are separated in its anode circuit as shown here**

**Fig. 17.6—This diagram shows a convenient way of separating the sound and vision channels**




points of the sound circuit corresponding to maximum sound signal in the sound channel and to minimum sound signal in the vision channel are not the same.

This is most easily seen from the particular arrangement of Fig. 17.6 but other circuits are similar. Here $V_1$ is the last common valve and $V_2$ is the first sound-channel valve. The v.f. coupling is $L_1$ tuned by stray capacitance; the v.f. signal is fed out in this case through a coupling coil to a feeder which carries the v.f. signal to another chassis. The arrangement of the sound-channel input circuit $L_2C_3$ coupled by $C_2$ is precisely that of the rejector circuit of Fig. 11.4.

Maximum rejection of the sound signal from the vision channel occurs when $\omega^2L_2(C_2+C_3)=1$. The maximum sound input to $V_2$ occurs when $\omega^2L_2\left(C_3+\frac{C_1C_2}{C_1+C_2}\right)=1$, ignoring the effect of $L_1$. If $C_1$ and $C_2$ are both small in relation to $C_3$ the two tuning points are very close together and the loss of signal in the sound channel if the circuit is tuned for rejection in the vision channel is very small. It is usually then best to tune the circuit for minimum sound signal in the vision channel; one may gain 6–10 db sound signal rejection for perhaps 1–3 db loss in the sound channel.

This process really means tuning the circuit so that the trough in the curve of Fig. 17.4 falls on the sound signal. This is, of course, the response curve of the coupling to the vision channel only and tuning to the trough does not mean that one is doing this in the sound channel. As far as this is concerned one is merely tuning somewhat off the peak.

Another separation circuit is shown in Fig. 17.7; $L$ is a normal tightly-coupled vision-channel coupling and $L_1$, $L_2$, $C$ form the sound-channel input circuit coupled to $L$ by $L_1$. As in the case of the circuit of Fig. 17.5, it is usually best to adjust the tuning of this circuit for minimum signal in the vision channel.

If these double resonant conditions do not exist, it is necessary to tune the circuit for maximum signal in the sound channel and to obtain the rejection in the vision channel with a separate circuit.

The sound channel after the separating stage follows normal

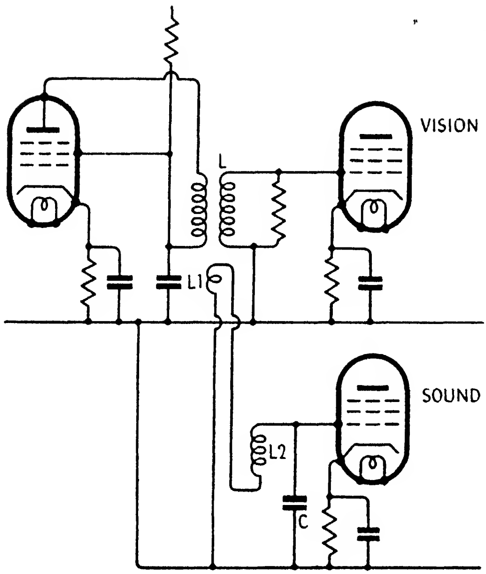



**Fig. 17.7—Another vision- and sound-signal separator is shown here**

practice. In the case of a superheterodyne a bandwidth of 50–100 kc/s should be provided in order to permit a reasonable amount of oscillator drift. In a straight set it will be difficult to obtain a smaller bandwidth. In any case, such a bandwidth is desirable if an ignition-interference suppressor is fitted.

The gain needed depends on the detector and a.f. circuits and follows the normal practice of sound receivers so that it need not be treated here. If the valves used are of the same type as in the vision channel roughly one-half the number of stages will be needed. Thus, in a straight set the author has used two common stages followed by two vision-channel stages and one sound-channel stage. In a superheterodyne he has used one s.f. stage and a frequency-changer common to both channels followed by a three-stage vision channel and a two-stage sound channel.

The gain control should normally act equally in both channels. This is easily arranged if there are two common stages by controlling these stages only. If there is only one common stage then three stages must be controlled—the common stage, the first vision-channel stage and the first sound-channel stage and the valves in the last two should be alike. This arrangement ensures that whenever the gain control is set for the proper input to the c.r. tube the sound detector also has the input for which it is designed and changes of signal strength which are corrected by the gain control do not affect the relative strengths of the output sound and vision signals.

When a superheterodyne is used with a common frequency-changer for both channels the vision-channel intermediate frequency is usually settled first. The sound-channel intermediate frequency must be 3·5 Mc/s different, but it can be higher or lower. This is the only matter left for choice and even this will usually be settled as lower because it gives the lower frequency, and so the better frequency stability, for the oscillator.

This frequency is also rather better because the selectivity of normal resonant circuits is slightly greater below resonance than above. It, therefore, helps in rejecting the sound signal from the vision channel if the sound intermediate frequency is lower than the vision. The effect is not large, however, at the frequencies usually adopted. It decreases in importance as the frequencies are raised and above some 15 Mc/s becomes negligible.

CHAPTER 18

# Interference

THE TERM INTERFERENCE COVERS anything added to the picture or sound which is not in the original transmissions. Strictly speaking, therefore, it includes noise. It is not, however, customary to call noise interference and as it arises chiefly in the first circuits and valve, it is dealt with in Chapter 12.

Interference proper falls under three main headings—that produced internally in the receiver, that caused by other signals, and that caused by external non-radio apparatus. The first of these may occur in the superheterodyne and be caused by harmonics of the intermediate frequency being fed back from the output to the input circuits; it is treated in Chapter 13. The straight set is free from this trouble but both straight set and superheterodyne can, of course, exhibit interference effects which are caused by parasitic oscillation in some stage or in associated equipment.

The effect produced might be indistinguishable from that caused by an interfering unmodulated external signal, and would probably be apparent only when the television signal is being received. It would, therefore, be rather hard to distinguish from external interference. The author has not met a case of this kind, but it is certainly a possibility and should be considered when persistent interference is encountered.

Interference of external origin and caused by unwanted radio signals is fortunately not very common, especially in the areas of higher field strength. When the interfering signals are on frequencies outside the vision-channel bandwidth the remedy is, of course, increased selectivity. This may not always be easy to obtain but is always possible.

The most serious interference occurs when it lies within the receiver bandwidth. If the interfering signal is unmodulated a pattern is produced on the picture corresponding to the beat with the vision signal. If this beat frequency is above some 2·5 Mc/s the effect is to break up the lines into dots so that instead of the raster having the normal appearance of parallel horizontal lines it is of a mesh character. Provided that it is not too intense it detracts very little from the picture quality.

When the beat is of lower frequency, but is above the line-scanning frequency, the appearance is of vertical or diagonal shadow bands which increase in width as the frequency is lowered. The pattern is rarely steady since the frequencies involved are not precisely constant. The bands become horizontal when the beat frequency falls below the line frequency.

When the interfering signal is modulated, as it usually will be, the same general patterns are obtained but they vary in accordance with the modulation.

The normal frequency allocations of stations should preclude this type of interference and, in fact, they normally do so. The allocations are such that there are no stations operating on frequencies within the receiver bandwidth which are situated in regions where they are likely to produce appreciable signal in the television service area. However, under abnormal propagation conditions remote stations may produce sufficient signal to cause trouble and there is always the possibility of interference from harmonics of stations operating on lower frequencies.

If the receiver bandwidth is 42–48 Mc/s, the second harmonic of stations in the band 21–24 Mc/s may cause trouble, as may also the third harmonic of 14–16 Mc/s, the fourth of 10·5–12 Mc/s, the fifth of 8·4–9·6 Mc/s, and so on.

There are two ways in which such harmonic interference can occur; they may be radiated by the transmitter, they may be generated in the receiver.

The first is the more usual for it is impossible completely to prevent harmonic radiation from a transmitter, although with good design it can be kept to a very low level. Interference with television reception from such radiation is unlikely to be important except towards the limit of the service area and then only when the receiver happens to be located fairly close to the transmitter.

An exception occurs when the transmitter concerned is an amateur one in the 14–16 Mc/s band and the interference may then occur even in areas of high field strength. This is because an amateur transmitter, like a television receiver, is usually in a residential district and so might well be within only a hundred yards or so of the receiver. In spite of its low power, its close proximity may lead to severe interference.

Such interference may not be the fault of the transmitter but may lie in the receiver. There may be sufficient 14–16 Mc/s signal picked up to overload the first valve and produce the harmonics at this point. If this is the case it is simple to introduce a wavetrap in the feeder—an acceptor circuit connected across the feeder is satisfactory—tuned to the fundamental of the interfering signal. If the harmonics are radiated from the transmitter, however, the only real remedy is at the transmitter, but in some cases an improvement can be effected by adopting single-sideband reception, choosing the particular side of the carrier away from the interference.

Apart from this, the only thing that can be done to reduce external interference is to use a directional aerial system with its minimum pointed to the interfering signal. Even this fails if it happens to lie in the same direction as the television transmitter.

Probably the most common source of non-radio interference is the ignition system of a car. The amount of trouble experienced depends very much on the location of the receiver. In residential areas having a high field strength the interference is rarely severe and away from main roads the traffic density is small. The amount of interference occurring under such conditions is thus quite tolerable. On a main road, with a high traffic density, however, the interference may be sufficiently continuous to be troublesome even if its intensity relative to that of the signal is not very great.

This applies also to reception in areas of lower field strength away from main roads, for as the receiver must be operated at higher gain each individual car causes interference over a greater distance and so for a longer time.

The right remedy is, of course, the use of interference suppressors on the cars themselves. This is not under the control of the television user, however, and it is thus fortunate that quite a lot can be done at the receiver to mitigate its effects.

The field of interference produced by an ignition system is local to the car and the first step is naturally to place the aerial as high and as far from the road as possible. In some cases it may pay to use a directional aerial with its minimum towards the road.

To obtain the full benefit from this it is important that the signal pickup of the set itself and of the feeder should be negligibly small. This necessitates good screening of all early receiver circuits and proper attention to the feeder; in particular, if this last is of the unscreened twin-wire type, careful attention must be given to the balancing of the input coupling transformer.

All the foregoing applies to both sound and vision, and when all has been done that can be by these methods, there remains the possibility of reducing the trouble by the use of limiters. Although

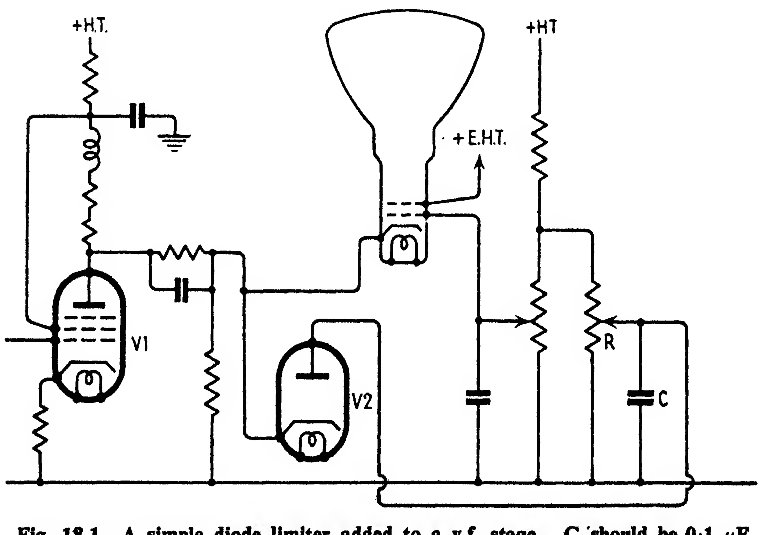



**Fig. 18.1—A simple diode limiter added to a v.f. stage. C should be 0·1 $\mu$F capacitance or more and the limiting level is controlled by R**

the technique may be basically the same, the practical form of the circuits is different in the vision channel from those adopted in the sound and it is desirable to treat them separately.

The waveform in the receiver corresponding to ignition interference consists in the v.f. circuits of pulses of large amplitude and short duration. They brighten the trace and so produce white spots on the picture. The pulse amplitude is often much larger than the signal amplitude corresponding to peak white with the result that not only is the interference very bright but, because the scanning spot is defocused, it appears as a splodge of large area.

This can be avoided by using a limiter at the c.r. tube input which is set to limit all signals to the peak-white level. No interference can then appear brighter than the white level and defocusing is avoided. Ignition interference is then restricted to white pinpoints and the usual bright splodges do not appear.

One very simple limiter is shown in Fig. 18.1 and consists merely of a diode $V_2$, a potentiometer $R$ and by-pass capacitor $C$ of 0·1 $\mu$F capacitance added to a standard v.f. stage. $R$ is adjusted so that the potential of its slider relative to earth is the same as that of the tube cathode on a peak-white signal. On all signals less than peak white the cathode of $V_2$ is positive with respect to its anode, the valve is non-conductive and does nothing. When the tube cathode falls below peak white the diode becomes conductive and ideally it ties the tube cathode to the slider of $R$ and prevents its potential from changing further.

In practice, the diode resistance is appreciable and the limiting action is not perfect. The circuit greatly reduces interference but does not always completely prevent some defocusing of the spot.

Another simple limiter* is shown in Fig. 18.2 applied to a v.f. stage handling a signal of a polarity suitable for feeding to the cathode of the c.r. tube. A low-resistance diode is used for $V_1$ and $C_1$ is 0·1 $\mu$F with $R$ 10 M$\Omega$. When $V_1$ is non-conductive $R$ is high enough for the circuit to have negligible effect. When $V_1$ is conductive, there is heavy negative feedback which reduces the interference. The diode bias is obtained automatically from the charge built up on the capacitor.

It is necessary that the diode should conduct at frequent intervals if the charge on $C$ is to be maintained. Under conditions of severe interference $C$ can take its charge from the interference, but when there is little or none the diode must necessarily become conductive at intervals when there is only the picture signal. This results in the removal of tone values near white and in some degree it mars the picture.

The circuit is very effective in reducing interference and the effect on the picture quality is small. It is hardly noticeable when a direct comparison cannot be made, but the sudden removal of the limiter while looking at the picture reveals a distinct improvement in the highlights.

* Pye Television Model B16T Set. Review *Wireless World*, December 1946, Vol. 52, p. 403.

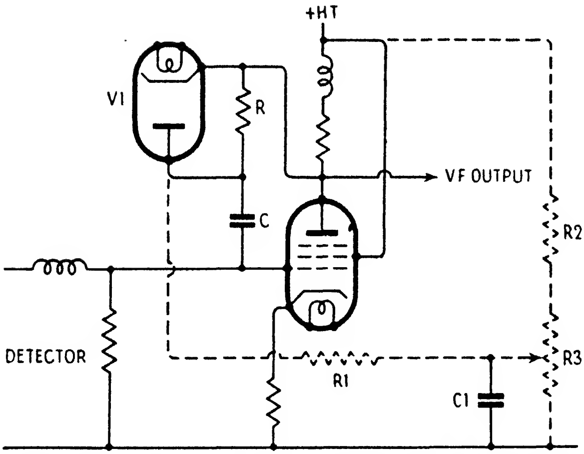



**Fig. 18.2—This limiter acts by negative feedback and bias is automatic. By removing R and adding the dotted connections a semi-fixed bias can be obtained**

The interference suppression obtainable is so good that under conditions of interference the circuit is well worth while. The effect on picture quality can be reduced at the expense of some complication and a somewhat lower efficiency of interference suppression. The resistor $R$ in Fig. 18.2 is removed and the components shown dotted are included; $R_1$ can be about 2 M$\Omega$. The diode is now biased largely with a fixed voltage derived from $R_2$ and $R_3$, and by adjusting $R_3$ it can be set so that it never comes into operation below peak white and so has no effect on picture quality. Because of this it limits rather above peak white and so permits more interference to pass. The adjustment of $R_3$ is critical and requires some care.

In the case of the sound channel a limiter can be a very effective reducer of ignition interference. In spite of this, disappointment is often expressed with its performance, especially when it is added to an existing set. It is necessary to remember that by its very nature a limiter cannot eliminate the interference, it can only reduce it. Even for it to do this satisfactorily there are certain fundamental laws which must be obeyed.

A carrier wave amplitude-modulated to a depth of 100 per cent by an audio frequency of sine waveform is sketched in Fig. 18.3, the solid line indicating the envelope of the r.f. cycles. The narrow spikes outside this envelope are intended to represent ignition noise.

This waveform is preserved through the receiver only if the bandwidth is adequate to handle the highest frequency components. The ignition interference pulses are of very short duration and so involve very high frequencies—frequencies much higher than any within the audible range.

If the bandwidth of the r.f. circuits is too small, although quite

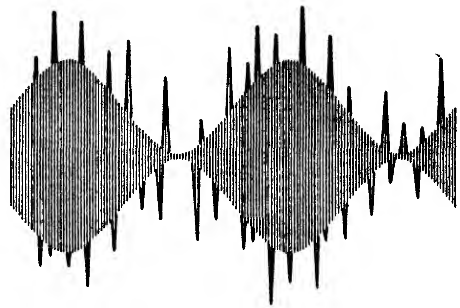


**Fig. 18.3—This sketch shows the general form of a 100 per cent modulated signal with ignition interference**

adequate for all desired modulation frequencies, it will filter out the high-frequency components of noise and the output waveform will not be like that of Fig. 18.3. The noise components will be greatly reduced in amplitude and lengthened in duration.

If a limiter is to operate successfully it is essential that the interference should be of large amplitude and short duration. If the waveform of Fig. 18.3 is preserved, the detector virtually wipes out one-half of it leaving an output like that of Fig. 18.4 (a). Passing this signal through a limiter adjusted to limit at the voltage corresponding to the peak of 100 per cent modulation brings it to the form of Fig. 18.4 (b) and it is clear that the limiter does remove quite a lot of the noise.

It should also be clear that if the pre-limiter bandwidth is inadequate the limiter will give little or no improvement, for there will be no large noise spikes for it to cut off. Instead the noise will have been converted into components of smaller amplitude and longer duration, but still of large energy content and very audible in the final result.

For success with any form of limiter it is essential to have a large bandwidth in all circuits preceding the limiter. After this point the bandwidth can be reduced, often with advantage, to that needed for the modulation only.

It is difficult to give any hard rule as to what the bandwidth

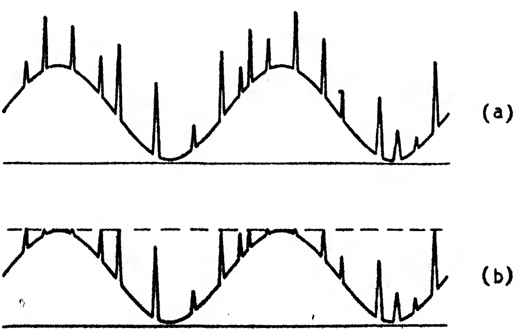



**Fig. 18.4—The a.f. signal with interference is shown after detection at (a) and after passing through a fixed-level limiter at (b)**

should be, but in general a figure of 3–5 times that needed for the modulation is necessary. Thus, if frequencies up to 10 kc/s are to be preserved in the sound channel, the pre-limiter response should be maintained up to 30–50 kc/s, which means an r.f. bandwidth of 60–100 kc/s. This is also the order of bandwidth needed to obtain freedom from oscillator drift in a superheterodyne receiver.

In addition, it is essential to fit the limiter at a point of adequate signal strength. The noise peaks must be of large amplitude compared with the pass-range of the limiter. In practice, it is inadvisable to fit a suppressor at a point at which the a.f. signal is less than about 5 V.

There are many different forms of limiter which can be adopted but they all fall into two distinct categories. There are those which have a fixed limiting level and those in which the limiting level varies with the a.f. modulation of the signal. The former are the simpler in most cases, but the latter approach more closely to the ideal.

One of the simplest of the fixed limiters is shown in Fig. 18.5 and consists simply of a short grid-base a.f. stage directly coupled to the detector. This last is conventional but the time-constant $R_1C_1$ of its load should be smaller than usual; say, $R_1 = 30$ kΩ, $C_1 =$ 30 pF or thereabouts. $R_2$ is a grid stopper to reduce r.f. applied to the a.f. stage; it should not exceed about 10 kΩ. The valve $V_2$ is given just enough grid bias by $R_3$ to avoid grid current with no signal.

An r.f. pentode is used for $V_2$ and the screen voltage is chosen so that, with the normal detector output, current cut-off occurs for a grid voltage corresponding to just over the peak of 100 per cent modulation. Thus, if the normal detector output is $-$ 2 V, on 100 per cent modulation it swings from 0 to $-$ 4 V, and the screen voltage of $V_2$ is chosen so that cut-off occurs for an input of a little more than $-$ 4 V.

The disadvantage of this system is that it demands very careful adjustment of either the detector output or of the a.f. stage screen voltage. If the limiting level is set too far from the modulation

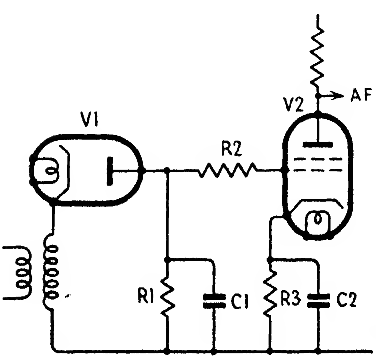
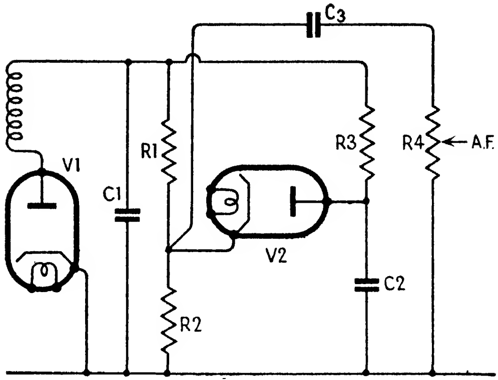



**Fig. 18.5—The simplest a.f. limiter is a short grid-base a.f. stage directly coupled to the detector**

**Fig. 18.6—A diode limiter with automatic bias**




peak, it will not be effective as a noise suppressor, while if it is set too close it may operate on the wanted signal and so cause serious distortion. Because of this methods of automatic level setting are commonly employed.

One of these—the Dickert circuit—is shown in Fig. 18.6. The detector is $V_1$ with its load resistance split into two equal parts $R_1$, $R_2$. One-half of the detector output is obtained at the junction of $R_1$ and $R_2$ and is applied through $C_3R_4$ to the a.f. amplifier. The full detector output is applied to the high time-constant circuit $R_3C_2$. The time-constant should ideally be large compared with the duration of one cycle of the lowest frequency of the modulation; common values are 0·5 MΩ and 0·5 μF.

The voltage across $C_2$ is then dependent only on the carrier level and is equal to the full detector output in the absence of modulation. If the detector output is 10 V on the carrier it varies from 0–20 V on 100 per cent modulation, but the anode of $V_2$ stays fixed at − 10 V with respect to earth. The cathode of $V_2$ is at one-half the detector output and so fluctuates between 0 and − 10 V. $V_2$ thus remains non-conductive.

On a noise pulse, however, the junction of $R_1$ and $R_2$ goes more negative than − 10 V and $V_2$ conducts and so virtually short-circuits $R_2$. The effectiveness of the suppressor depends largely upon the diode resistance in relation to $R_2$. With the low values normally permissible for $R_2$ (15–50 kΩ), the suppression is not great.

It is not uncommon, therefore, to add a second diode as shown in Fig. 18.7. In its essentials the action of the circuit is the same as before; the voltage across $C_2$ is substantially equal to the rectified carrier voltage and the voltage at the junction of $R_1$ and $R_2$ varies between this value and zero. Therefore, $V_2$ is conductive and $V_3$ is non-conductive. The purpose of $R_4$ is to complete the $V_2$ diode circuit when $V_3$ is non-conductive; it should be about 1 MΩ. On a noise pulse the anode of $V_2$ is carried rapidly negative and $V_2$ becomes non-conductive. At the same time, because this interrupts the current through $R_4$, $V_3$ becomes conductive. On a noise pulse,

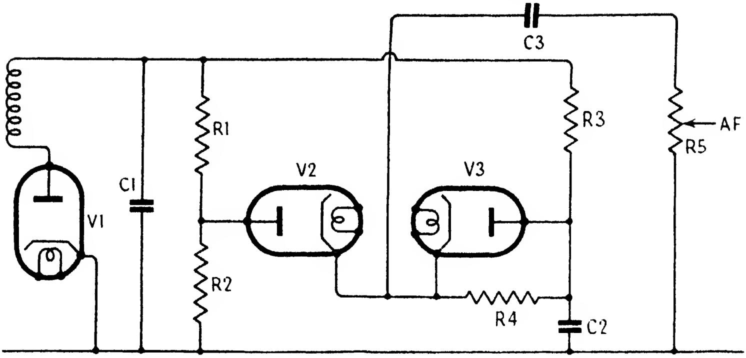



**Fig. 18.7—Double diode limiter in which $V_2$ opens the a.f. connection to the detector and $V_3$ shorts it to earth**

therefore, $V_2$ opens the a.f. connection to the detector and $V_3$ shorts it to earth through $C_2$. In fact, $V_2$ provides most of the noise suppression, but $V_3$ does help somewhat.

Rather better noise suppression is to be expected from a circuit in which the limiter bias is arranged to follow the modulation. It then comes into operation on a signal which changes more rapidly than the modulation and limits to the instantaneous level of modulation rather than to the 100 per cent level.

One such circuit is shown in Fig. 18.8; $R_1C_1$ form the detector load and the output is applied through $C_2$ to the noise limiter $V_2$. This diode is normally held conductive by the connection of $R_6$ to positive h.t. of perhaps 250 V. The diode resistance will normally be less than some 20 kΩ, so that the total resistance is this plus $R_2$ plus $R_6$ or some 3·22 MΩ. The current is thus 250/3·22 or about 80 $\mu$A and the voltage actually across the diode is of the order of 20 × 0·08 = 1·6 V.

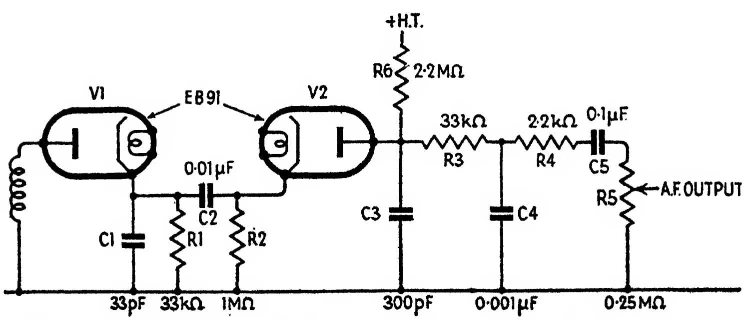



**Fig. 18.8—A diode limiter in which the limiting level follows the modulation of the signal**

If the voltage across $R_1$ does not change too quickly the voltage across $C_3$ will follow it with negligible lag and the voltage drop across the diode will hardly change. However, if the voltage across $R_1$ increases very rapidly, that across $C_3$ cannot follow it instantaneously, for $C_3$ must charge through the diode resistance. The cathode of $V_2$ is then carried positive by the noise pulse and $V_2$ is cut off leaving the voltage on $C_3$ unaffected. While $V_2$ is cut off the voltage across $C_3$ starts to rise, because it charges through $R_6$ from the h.t. supply.

Provided that the time-constant is large enough and the noise pulse of very short duration, the rise of voltage across $C_3$ can be kept small enough to have little audible effect. In this circuit as well as in most others, it is usually necessary to use a low-resistance diode of the EA50 and 6AL5 classes. The ordinary diodes such as the EB4 and 6H6 are of too high resistance for the full suppressing action to be secured.

There are many other types of limiter and it is not possible to deal with them all. The general principle, however, is to utilize the a.f. signal in some way to vary the bias on the limiter so that it comes into operation on any sudden increase of input above the instantaneous value of the wanted signal.

CHAPTER 19

# The Aerial

FOR TELEVISION PURPOSES the usual broadcast aerial is of little use for anything but short range reception. The general practice is to use a resonant aerial, because its efficiency is much higher.

Vertically-polarized waves are radiated by British transmitters and, assuming that the plane of polarization does not change during propagation, a vertical receiving aerial is needed. In practice, of course, there nearly always is some change in the plane of polarization, but under all normal conditions it is not enough to render a change from the vertical necessary in the aerial.

The simplest aerial is the vertical dipole which consists merely of a length of wire, rod or tubing one-half wavelength in length. When such an aerial, AB Fig. 19.1, is erected in free space and there is a vertically-polarized wave of wavelength $\lambda = 2AB$, standing waves occur on the aerial.

The current at any point is represented by the dotted line of Fig. 19.1, and the voltage by the dash line. At the centre of the aerial the current is a maximum and the voltage is zero, while at the ends the voltage is a maximum and the current is zero.

Consequently the impedance varies along the aerial and is a minimum at the centre and a maximum at the ends. In the practical case of an imperfect aerial the impedance is of the order of 70 Ω at the centre and 2,500 Ω at the ends.

The practical aerial, moreover, is not exactly one half-wavelength long, but is somewhat shorter. The aerial is actually a tuned circuit possessing inductance, capacitance, and resistance, but unlike an ordinary circuit in which these attributes are lumped, in the aerial they are distributed.

Now in the case of a circuit with lumped constants we are familiar with the fact that the resonant frequency is lower than the calculated value, because of the effect of stray capacitances and the inductance

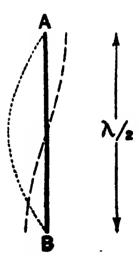



**Fig. 19.1—The current and voltage distribution in a half-wave aerial are shown here**

of connecting leads. So in the case of the aerial its actual resonant wavelength is longer than twice its physical length because of the effect of the earth, neighbouring buildings and trees and the presence of the feeder and the aerial supports. The diameter of the conductor also has an effect on the resonant length which becomes shorter as the diameter increases.

The precise length of the aerial needed for resonance in any case cannot be calculated exactly. Fortunately, however, it is not critical and it suffices to make it about 95 per cent of the ideal length. In practice, satisfactory results are secured if the following relation is used:

$$l = 1{\cdot}56\lambda \qquad 19.1$$

where $l$ is in feet and $\lambda$ in metres.

The signal is not of a single frequency, however, but occupies a band. Moreover, it is common practice to utilize the same aerial for both sound and vision channels. The band of frequencies over which the aerial is responsive must thus extend from the sound channel to the highest required frequency of the vision channel. In the case of single-sideband reception of the vision sidebands adjacent to the sound channel the aerial response must cover from the sound- to the vision-carrier frequencies—3·5 Mc/s. With single-sideband reception of the sidebands remote from the sound channel or for double-sideband reception, the aerial bandwidth needed is greater—6·25 Mc/s. In general, the aerial should resonate at the geometrical mean of the band, but since the vision channel is often more important than the sound it is not uncommon to place the resonance at the middle of the vision channel.

**A typical dipole aerial and reflector**

The mid-band frequencies and wavelengths and the aerial lengths for the various existing and proposed channels are given in Table 19.1. The figures for length should be taken as approximate only, for the diameter of the aerial rod does affect the length slightly.

TABLE 19.1

## AERIAL DIMENSIONS

| *Channel* | *Frequency (Mc/s)* | *Wavelength (m)* | *Dipole Length (ft)* | *Reflector Length (ft)* |
|---|---|---|---|---|
| 1 | 43·2 | 6·94 | 10·8 | 11·15 |
| 2 | 50 | 6 | 9·38 | 9·8 |
| 3 | 55 | 5·46 | 8·54 | 8·8 |
| 4 | 60 | 5 | 7·82 | 8·08 |
| 5 | 65 | 4·62 | 7·21 | 7·45 |

Although the aerial is a resonant circuit it is not of very high $Q$ and its bandwidth is considerable. It depends chiefly on the ratio length/diameter of the conductor and it increases as this ratio is reduced. If the ratio is too great the bandwidth will be inadequate and the definition of the picture will suffer.

Increasing the diameter does not greatly affect the other properties of the aerial. The resonant impedance is affected only to a minor degree and the length required for resonance is slightly reduced.

An exact analysis is exceedingly complex. It is fortunate, therefore, that in the case of a wideband aerial the dimensions are not critical. For television purposes the ratio $l/d$ (where $d$ is the diameter of the conductor) should never be more than 300 and it is better much smaller. This value makes the minimum diameter for Channel 1 0·036 ft = 0·43 in. In practice most commercial dipoles are made of ½-in to ¾-in tubing.

A larger diameter than this is technically desirable but is very rarely used because of the increased weight of larger tubing. For double-sideband reception of Channel 1, Alexandra Palace, a diameter of as much as 1 ft is really desirable, and the large diameter does have an appreciable effect upon the resonant length, which becomes 10·25 ft. An aerial of this diameter would be impracticable if made of tubing, but it is satisfactory to make it in the form of a cage of wires. A close approach to a cylindrical form would be given by eight wires arranged at the corners of an octagon. In practice fewer wires are satisfactory and at the Alexandra Palace transmitter only three wires in triangular formation are used. The wire spacing is some 15 in.

The centre impedance varies slightly with the diameter, but it is likely to be affected much more by the proximity of the aerial supports. It is satisfactory to take it as 72 Ω in nearly all cases.

A vertical half-wave dipole of this nature has a circular horizontal polar diagram; that is, it is equally responsive to signals in any horizontal direction. In the vertical plane the response depends on the height of the aerial above ground.

When the height at the centre is $\lambda/4$ above ground, or less, the aerial is effective for signals arriving at angles of elevation from 0–10° approximately. At greater heights up to $\lambda/2$ the efficiency at very low angles tends to decrease and at angles above 10°

to increase. Subsidiary lobes of moderate efficiency appear at about 60°.

At greater heights the low-angle efficiency tends to decrease further and more subsidiary lobes at high angles appear. Null points occur between the lobes.

Television reception is normally by direct ray and it is rare for the vertical angle to exceed some 2–3°. It would seem, therefore, that the height should not exceed $\lambda/4$. This is obviously absurd, however, for the lower end of the aerial would be at ground level.

The theoretical performance is dependent on the assumptions of perfect ground conductivity and no obstructions around the aerial. The relatively poor conductivity of the ground and the usual obstructions of buildings and trees profoundly modify the performance and in practice it is usually desirable to mount the aerial as high as possible, particularly as the field strength of the signal increases with height—especially at long range.

Nevertheless, the possibility of null points in the vertical polar diagram should be borne in mind and if poor results are secured it is a good plan to try varying the aerial height a few feet at a time.

In the usual case where the dipole is mounted on a short pole attached to a chimney stack the vertical polar diagram is more likely to be dependent on its height above roof level than that above actual ground, and low angle reception is quite good. Moreover in some cases, as will be seen later, the feeder can act in some degree as a local earth and make low-angle reception good although the aerial itself is much more than one-quarter wavelength above the actual ground level.

It is necessary to convey the energy picked up by the aerial to the receiver and a feeder is used for this. There are two basic forms of feeder, a two-wire line and a coaxial cable. Whichever is used its impedance must equal that of the aerial and receiver at the points of connection. If it does not a matching section must be inserted.

The need for matching impedances in feeders arises from two effects. The efficiency is greatest when the matching is perfect, and reflections are avoided. The latter is much the more important, for it can cause a ghost image. Thus, suppose that the aerial and feeder, on the one hand, and the feeder and receiver on the other are mismatched. If the aerial receives a pulse of r.f. energy this travels down the feeder to the receiver but because of the mismatch not all of it is absorbed by the receiver. A part of it is reflected and travels back to the aerial where, because there is another impedance mismatch, a part is again reflected and travels back to the receiver. It arrives at the receiver later than the original pulse by the time taken to travel twice the length of the feeder.

The effect can thus cause a displaced image of weaker intensity than the main one. In fact, as part of the reflected wave is again reflected, a whole series of ghost images equally spaced can be obtained. As the losses in each reflection are considerable it is not common for more than the first to be visible.

The velocity of propagation in a feeder cable depends on its design, but it is of the order of 0·65 of the free-space velocity. The transmission time per 100 ft of cable is thus about 0·16 $\mu$sec. The delay of a reflection is twice this, since the reflection travels twice along the cable, and is 0·32 $\mu$sec per 100 ft.

With a 12-in tube the picture width is 10 in and the time per active line is 82·77 $\mu$sec, so that the scanning velocity is 0·121 in per $\mu$sec. In this case, therefore, the ghost displacement is nearly 0·04 in per 100 ft of feeder cable.

A separate ghost image is unlikely to be visible as such with feeders of less than 300–500 ft for it will not be sufficiently displaced from the main image. It is unusual for feeders to exceed 100 ft in length under average domestic conditions so that ghost images from this cause are rare.

It is important to note, however, that with short feeders of 50–100 ft, although a noticeable ghost image may not be observable, such a ghost may in fact be present and have the effect of reducing definition. Although insufficiently displaced from the main image to be separately observable, it can blur the edges of the image. It is, therefore, still important to avoid reflections on the feeder even when only short lengths are used.

It should be noted that in theory it is sufficient for the feeder to be terminated properly at only one end in order to avoid a ghost image. In practice, it is rarely possible to match perfectly over a band of frequencies and it is consequently necessary to match as well as possible at each end.

At radio frequencies the impedance of a feeder is mainly a function of its inductance and capacitance per unit length. These quantities depend on the conductor sizes and spacing and upon the dielectric constant of the insulant employed. There are various formulæ for calculating the impedance, but they are only of much practical use when the dielectric is air, for it is the only dielectric having a dielectric constant accurately known, without having to measure it on each sample. Further, solid dielectrics usually necessitate a proper manufacturing process.

The only form of feeder which can readily be made is an air-spaced twin-wire line and such lines can rarely have an impedance of less

TABLE 19.2

## FEEDER CHARACTERISTICS

| *List No.** | *Impedance* ($\Omega$) | *Attenuation (db/100 ft) at* 45 *Mc/s* | 61·75 *Mc/s* | *Overall Diameter (in)* | *Type* |
|---|---|---|---|---|---|
| T.3020 | 71 | 2·8 | 3·2 | 0·23 | coaxial |
| T.3026 | 71 | 1·2 | 1·4 | 0·45 | |
| T.3098 | 80 | 2·6 | 3·0 | 0·23 | twin screened |
| T.3100 | 80 | 1·0 | 1·2 | 0·63 | |
| T.3066 | 80 | 3·2 | 3·8 | 0·1×0·16 | twin unscreened |
| T.3068 | 80 | 2·8 | 3·3 | 0·16×0·2 | |

* List number of British Insulated Callender's Cables, Ltd.

than about 300 Ω. They do not well match to an aerial and are rather clumsy.

The usual television feeders have an impedance of about 70 Ω and match well to a dipole aerial. Solid dielectric is used, usually polyethylene. Twin-wire types are obtainable with or without an external metal-braiding. The latter are sometimes of rather higher impedance—110–130 Ω—while the former are usually around 76–98 Ω. The coaxial types are nearer 70 Ω, being usually listed as 67–77 Ω, but other impedances are available; 39–50 Ω and 90–100 Ω types are made.

To give an idea of actual cable characteristics figures are quoted in Table 19.2 for some typical television feeders. These refer to a particular make, but they are typical of present practice. The cable loss decreases with size so that in long-range reception the large-diameter types are advantageous; they may also be desirable in exceptional cases where a very long length of feeder must be used.

Usually, however, little more than a 50-ft length is required and with the worst of the feeders listed the attenuation is no more than 2 db. Except at limit range, the decrease of attenuation with a better cable would not be noticeable, and most television sets use either a twin-unscreened feeder or a $\frac{1}{4}$-in, or thereabouts, coaxial cable.

A twin-wire cable is a structure balanced to earth and a dipole aerial is a similar structure. When the aerial and feeder impedances are the same it is necessary only to join the two feeder wires to the two adjacent halves of the aerial. The feeder should be led away from the aerial at right angles for at least a quarter-wavelength, Fig. 19.2 (a) but a gradual bend (b) is usually satisfactory.

At the receiver the feeder wires are connected to a coupling coil which ideally has an earthed centre-tap and is screened from the tuned circuit as shown in Fig. 19.3 (a). The simpler arrangement of (b) is often used, however. If the cable is screened then the screen is earthed at the receiver end.

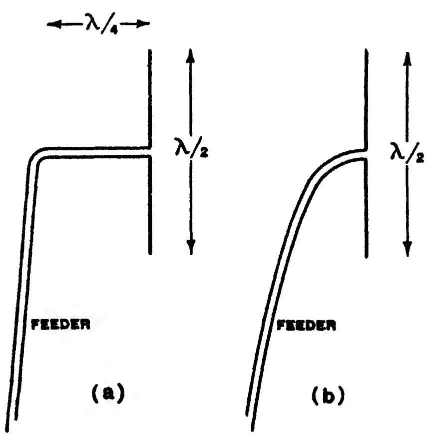



**Fig. 19.2—Two methods of bringing a feeder away from an aerial are sketched here**

With a coaxial cable matters are simpler at the receiver. A coupling coil can be used, but a simple tapping on the tuned circuit suffices as in Fig. 19.3 (c). There is, however, more difficulty at the aerial end, for a coaxial cable is an unbalanced structure. Sometimes the centre conductor is joined to one side of the aerial and the outer to the other. This is theoretically wrong, but in practice works quite well and has the merit of simplicity. The main drawback is that the outer conductor is not as " dead " as it should be. In many cases a phase-inverter is used.

In order that this may be understood it is necessary to digress somewhat and consider the properties of lines and cables. A cable has a characteristic impedance $Z_0$ as already stated and it has the property that if it is terminated by an impedance $Z_r$ at one end the impedance measured at the other end is equal to $Z_r$ provided that $Z_0 = Z_r$. Irrespective of its actual length, the input impedance equals the characteristic impedance if the remote end is matched to $Z_0$.

If the electrical length of the cable is one-half wavelength, the input impedance equals the terminating impedance irrespective of the characteristic impedance.

If the electrical length of the cable is one-quarter wavelength the input impedance is $Z_0^2/Z_r$ and such a cable can be used as a transformer for impedance matching.

It is important to note that these last two cases hold exactly only at one frequency, so that over a band some deviation from the optimum condition is found.

Under steady-state conditions the voltages at the ends of a $\lambda/2$ length of cable are 180° out of phase, and at the ends of a $\lambda/4$ length they are 90° out of phase. A $\lambda/2$ length can thus be used as a phase-inverter.

This is sometimes used for matching a coaxial cable to a dipole. The arrangement is shown in Fig. 19.4. The coaxial cable is shown connected directly to the upper half of the dipole and to the lower

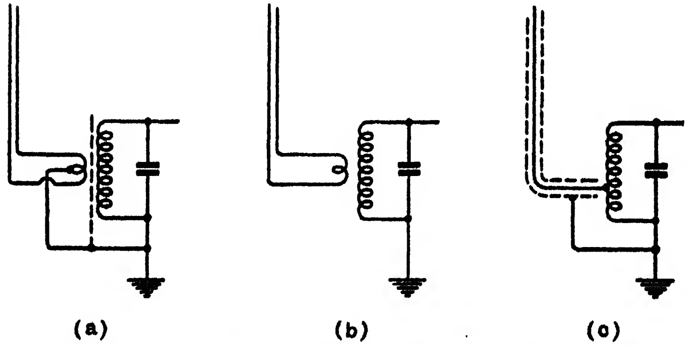
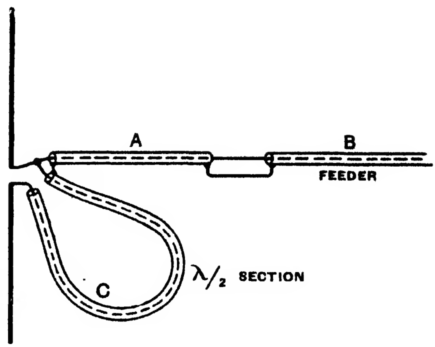



Fig. 19.3—The ideal feeder coupling to the receiver is shown at (a) and usual form at (b). The case of a coaxial cable is indicated at (c)

**Fig. 19.4—A balance-unbalance transformer can be obtained with the aid of a $\lambda/2$-section of cable**




half through a $\lambda/2$ length of feeder. The outers of the two feeders are joined.

The two halves of the aerial are thus fed in opposite phase. Each half presents an impedance of 35 $\Omega$ at the points of connection. The input impedance of the $\lambda/2$ section C is the same—35 $\Omega$—so that the total impedance presented to the feeder is the two in *parallel*, or 17·5 $\Omega$. This is too small for any normal feeder, so a $\lambda/4$ matching section A is used. To match to a 70-$\Omega$ feeder B, the impedance of section A must be 35 $\Omega$. As C is terminated by 35 $\Omega$ it is best, but not essential, for this to be of 35-$\Omega$ cable also.

This use of cable of different impedance can be avoided by the arrangement of Fig. 19.5. Here one side of the aerial is fed through a $\lambda/4$ section and the other through a $3\lambda/4$ section, so that the difference is still $\lambda/2$. If these sections are of 70-$\Omega$ impedance the input impedances of each at the junction with B are 140 $\Omega$, and the two in parallel provide 70-$\Omega$ to match B.

A word of warning should be given here. No attempt should be made to use such matching methods unless proper methods of measuring the electrical lengths of the cables are available. The

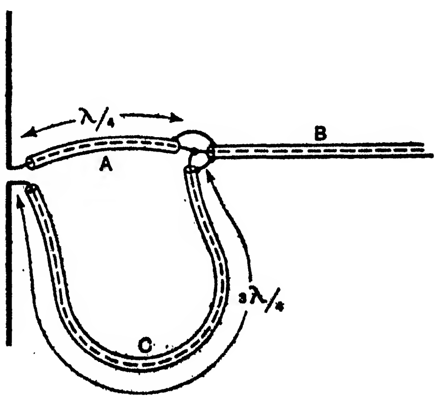



**Fig. 19.5—By using $\lambda/4$ and $3\lambda/4$ cable-sections a balance-unbalance transformer can be secured with 70-ohm cable throughout**

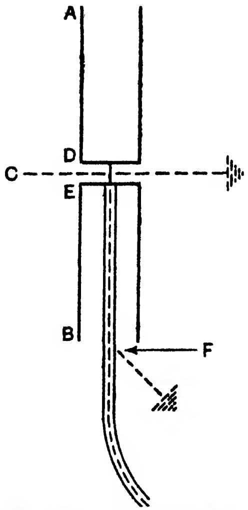



Fig. 19.6—A balance-unbalance transformer action can be secured by running the feeder inside the lower limb of the aerial

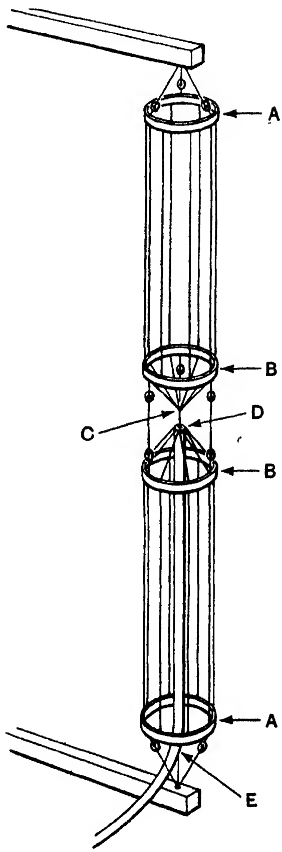



Fig. 19.7—Cage dipole of large bandwidth connected to a coaxial feeder running inside the lower half

electrical length of a solid dielectric cable is much greater than its mechanical length; it is approximately 50 per cent greater with polyethylene dielectric. It is not possible to determine the right length merely by a measurement of the physical length.

Another method of joining a coaxial cable to a dipole which is much more convenient, especially with wire-cage aerials, is merely to run the feeder up the centre of the lower limb of the aerial. This is shown in Fig. 19.6. The outer of the cable with the inside of the aerial conductor form a $\lambda/4$ resonant line of the coaxial type, short-circuited at the upper end. The impedance is, therefore, infinite at the open end, or in practice when losses are not zero, very high.

The cable is earthy up to the point of entry into the aerial and the resonant line thus provides a high impedance between earth and the lower end of the aerial, which is a high potential point. The easiest way of understanding the action is to assume the dipole to be correctly excited by a wave. Then A and B are at equal and opposite potentials with respect to the mid-point C of the system; this point is at earth potential.

There is a potential difference along the lower limb of the aerial, so that at an instant when A is positive with respect to earth, B is equally negative. At the centre D and E are also equal and opposite with respect to earth but the magnitudes of the potentials are much less since the impedance is lower. The inner conductor of the cable is joined to D and the outer to E and the cable impedance matches the aerial. The outer of the cable at E is not at earth potential for it is joined to E. It is required that it be brought to earth potential at the point F where it leaves the aerial. If it is, then the potential difference between E and C must equal that between E and F.

Writing $V_{AD}$ for the potential of A with respect to D, and so on, we have $V_{AD} + V_{DC} = V_{CE} + V_{EB}$. Also $V_{AD} = V_{EB}$ if the aerial is balanced and so $V_{DC} = V_{CE}$. The voltage across the feeder is $V_{DC} + V_{CE} = 2V_{CE}$. If F is earthy, $V_{FE} + V_{EC} = 0$ along the cable outer, hence $V_{FE} = -V_{EC} = V_{CE}$.

Regarding the inner of the aerial as the outer of a coaxial line, if F is earthy, $V_{BF} = V_{BC}$, but $V_{BC} = V_{BE} + V_{EC}$ along the aerial and $V_{BF} = V_{BE} + V_{EF}$ around the line. Therefore again $V_{EC} = V_{EF}$.

What happens is that the aerial functions normally and the feeder is also normal up to the aerial. Inside the aerial the feeder becomes more and more " alive " and there is a balanced condition at its junction.

Another effect brought about by the use of this method of feeding the aerial is that the feeder acts to some degree as a local earth at $\lambda/4$ below the centre of the aerial. It thus tends to improve the performance at low angles.

The sketch of Fig. 19.7 shows how this system can be applied to a cage dipole. Each half of the aerial proper consists of a number of wires stretched between rings A and B. The rings can be conductors or non-conductors as preferred, since they only connect points of the same potential. At the lower end of the upper limb the wires are tapered in to a point C for the connection to the centre conductor of the feeder. At the upper end of the lower limb they are tapered in similarly not to a point, but to a metal ring D. This should be of an inside diameter such that it just slips over the metal sheath of the cable and can be soldered to it for the connection to the cable outer conductor.

The aerial is most easily supported by suspending it between brackets from a wooden pole.

This method of connecting a coaxial feeder to an aerial, by running it up inside the lower member of the aerial, is not recommended with small diameter aerials because of the practical difficulty of maintaining adequate insulation between the two. It best lends itself to the large-diameter cage aerial.

It will be apparent from the foregoing that a twin-wire feeder is much easier to match to a dipole than a coaxial type and one might conclude that it is consequently the preferable type. This is not necessarily so, however, for a twin-wire feeder demands a balanced input circuit in the receiver and this is not always easy to provide.

If the receiver input is not well balanced to earth the conditions are much the same as with a coaxial cable connected directly to a dipole without any attempt at balance these. There is likely to be little difference in the performance of a screened twin-wire feeder and such a coaxial cable and the latter is usually the cheaper.

An unscreened twin-wire cable cannot be recommended for indiscriminate use, for it is likely to be satisfactory only when the receiver gain is low or when the receiver input circuit is unusually well-balanced. The absence of screening is likely, when the input circuit is not almost perfectly balanced, to increase the difficulties of maintaining stability in a straight set or avoiding i.f. harmonic interference in a superheterodyne. Receiver screening is rarely so perfect that one can willingly dispense with the extra protection afforded by a screened feeder, either twin-wire or coaxial.

In spite of the theoretical necessity for maintaining balance in the aerial and feeder circuits it is not even attempted in many practical cases. A coaxial cable, for instance, often has its inner and outer joined directly to the two halves of the dipole without any form of balance-to-unbalance transformer. The reason is that the effect on the performance is small and the arrangement is simpler.

It is, however, very unsatisfactory when the feeder passes through a region of local interference, the aerial itself being outside it. The unbalanced condition results in interference being picked-up by the feeder. In this special case it pays handsomely to adopt quite elaborate methods of achieving balance.

Reverting to the aerial itself, it is very common practice to add a reflector to the simple dipole described. This is a similar conductor to the aerial, but slightly longer, mounted behind it so that the two are in line with the transmitter. The pick-up now varies around the aerial and the horizontal polar diagram exhibits a minimum immediately behind the reflector and a maximum in front of it.

This is sketched in Fig. 19.8 and the ratio of the responses at front and back depends on the spacing of the aerial and reflector. It is a maximum when the spacing is about $0{\cdot}1\ \lambda$.

The use of a reflector increases the front pick-up as compared with that of a simple dipole. The amount of increase is not great, being usually about 4 db, but is useful at long ranges. The improvement in front-to-back ratio is often more important in reducing interference from sources immediately behind the aerial. This ratio can be as much as 10 db.

The presence of the reflector influences the centre impedance of the dipole by an amount which depends on the spacing. When this is $0{\cdot}1\ \lambda$ the impedance is greatly reduced—to around 10 $\Omega$ or so, and matching to the feeder becomes quite difficult when a considerable bandwidth is needed. For television purposes, a spacing of

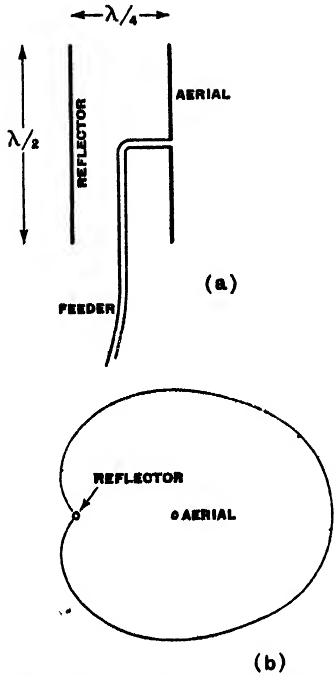



Fig. 19.8—The use of a reflector behind the aerial leads to an appreciable increase in signal strength, and makes it directional as shown by the polar diagram

$\lambda/4$ is more useful, and the centre impedance is then around 60 $\Omega$ so that little matching difficulty arises.

The proper length for a reflector spaced $\lambda/4$ from the dipole is given in Table 19.1, page 292.

If the "reflector" is mounted on the transmitter side of the aerial it will act as a director provided that it is shorter in length than the dipole and more closely spaced from it. More than one director can be used in addition to a reflector, and the gain becomes quite high. The directional properties are greatly improved.

Such aerials are usually unnecessary for anything but very long range reception. A single reflector with $\lambda/4$ spacing is very useful, however.

Within the normal service area a dipole or dipole and reflector is usually satisfactory. At great distances from the transmitter something more may be needed and a form of rhombic aerial is probably advisable. It is essentially a wideband directional aerial of quite high gain—up to some 20 db is possible and 12 db is quite usual. Its great disadvantage is that it occupies a large amount of space. In general, it is only suitable for a country district where a field is available for its erection.

The half-rhombic of Fig. 19.9 is usually most suitable, for the full rhombic, having an inverted-V section below it, demands a mast of more than twice the height. The characteristics depend chiefly upon the length $l$ of the sides and the angle $\theta$ at the apex. The sides must be several wavelengths long if the improvement over a dipole is to be worth while.

The performance depends very much on the ground conductivity for the aerial depends on ground reflection to provide virtually the other half of the rhombus. In general, therefore, it is a good plan to use it with a counterpoise consisting of several wires stretched beneath the aerial for at least its full length. The terminating resistance $R$ should be about 300 $\Omega$ and the feeder must be matched to this figure.

There is relatively little published information about this type of aerial but in one account* it is stated that the length $l$ must be an

* *Murphy News*, April 1949.

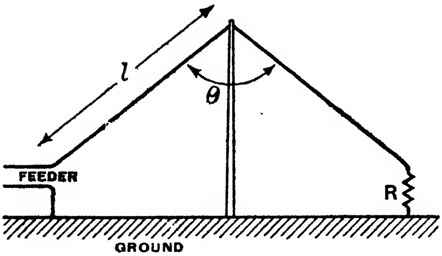



**Fig. 19.9—Vertical half-rhombic aerial for long range reception**

odd number of quarter-wavelengths greater than one while it must also be $\lambda/2$ longer than its projection on the ground. If $n$ is the number of quarter-wavelengths in $l$, then the height $H$ of the apex above ground is simply

$$H = \frac{\lambda}{2}\sqrt{n-1}$$

$$l = n\lambda/4$$

and the total length of aerial measured along the ground is $\lambda(n/2 - 1)$.

If $\lambda = 6{\cdot}94$ m and $n = 9$, the length of each wire is $15{\cdot}6$ m $= 51$ ft, the total length along the ground is $24{\cdot}3$ m $= 80$ ft and the height of the apex is $9{\cdot}8$ m $= 32$ ft.

The aerial should be terminated by a 400-Ω resistor at the end facing the transmitter. If a 70-Ω feeder is used it should be connected to the aerial through a $\lambda/4$ matching section of 170-Ω impedance.

The possibility of ghost images caused by reflections in the feeder has already been mentioned. This is not the only cause of such ghost images, however, for the signal which reaches the aerial may not have travelled to it from the transmitter by one path only; it may have travelled by several different ways. Its different paths will not all be of the same length and the components will have taken different times to reach the aerial. Ghost images, which may be negative, can be produced and the synchronizing may be upset.

The effects can be very serious, and it is not always possible completely to eliminate them. Usually, however, they can be made negligible by the correct arrangement of the aerial system. Before considering this, however, it is necessary to understand just how the ghost images are produced.

Ultra-short-waves tend to travel in straight lines, but are readily reflected by a surface large compared with the wavelength. The wavelength of the vision transmission is of the order of 6 metres, which corresponds to about 18 ft. There are consequently many objects such as buildings, hills and clouds, in the path of the signals which can cause reflection. Aeroplanes, in particular, can cause reflections which vary in intensity quite rapidly.

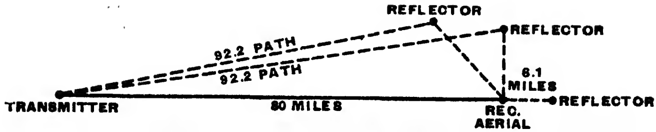



**Fig. 19.10—This sketch shows several different paths by which a signal may travel to the receiver**

Now the signal is radiated in all directions from the transmitting aerial. A part of it travels by the direct path between transmitter and receiver, but another part, which is initially radiated in a different direction, may still reach the receiver because it encounters some reflecting object.

Obviously, the reflected wave has travelled a greater distance than the direct ray and it arrives at the receiving aerial later. Consequently it produces a displaced image.*

The difference in path can readily be computed from the displacement of the ghost from the main image. The time occupied by the horizontal scanning of the visible portion of the picture is 82·77 $\mu$sec; this figure multiplied by the ratio of the displacement to the picture width gives the delay time of the signal. As the signal travels at 186,000 miles a second, or 0·186 miles per microsecond, the distance can easily be worked out. In the form of an equation,

$$D = 0{\cdot}186 \frac{d}{W} t \qquad 19.4$$

where $D$ is the distance in miles by which the path of the reflected wave exceeds that of the direct ray, $d$ is the distance between the ghost and main images, $W$ is the width of the picture and $t$ is the time in microseconds occupied by the scan of the visible portion of the picture. With $t = 82{\cdot}77$ $\mu$sec the equation reduces to

$$D = 15{\cdot}4 \frac{d}{W} \qquad 19.5$$

With a 10-in picture and a 1-in displacement of the ghost image $D$ is 1·54 miles.

A greater displacement of the ghost means that the reflecting surface is farther from the receiving aerial. Thus, 7¾-in displacement corresponds to a distance of 11·9 miles and the minimum distance of the reflector from the receiver is 5·95 miles. This occurs when it is behind the receiving aerial.

Three of the many possible points for the reflector are sketched in Fig. 19.10. It may be 5·95 miles behind the aerial, 8 or 9 miles on one side of it, or at a somewhat greater distance on one side, but nearer the transmitter. In this example, the direct path is taken as 80 miles and as the excess distance of the reflected wave is 12·2 miles, the indirect path is 92·2 miles. (In this paragraph and in Fig. 19.10 distances differ slightly from those given earlier because

* *Wireless World*, September 15, 1938, and January 19, 1939.

they were computed for a scan period of 84·5 $\mu$sec as previously used by the B.B.C. They are now 2 per cent less.)

The precise effects of reflection depend upon the relative strengths of the direct and reflected waves, their phase difference, and their time difference. This last governs the displacement of the ghost and has been dealt with.

When the reflected wave arrives at the receiver 180 degrees out of phase with the direct wave and with a considerable time delay the effect is shown in Fig. 19.11, when the reflected wave is weaker than the direct. Here (a) shows the r.f. waveform of one scanning line with modulation corresponding to a white dot on a grey background.

Many fewer cycles are shown than there are in reality, of course. White is given the amplitude of 80 arbitrary units and grey 60.

The reflected wave (b) is generally the same, but of smaller amplitude, 53 corresponding to white and 40 to grey. The phase is opposite and the time delay such that the " white " pulse comes noticeably late.

The addition of these two waves gives the composite (c) which has an amplitude corresponding to grey of only 20. The " white " pulse is reproduced at 40, but the ghost gives an amplitude of only seven. This is less than " grey " and corresponds to black. After the detector the v.f. waveform is as shown in (d). By a similar process a black square in the original can produce a white ghost.

Now if the reflected wave is in the same phase as the direct we find that the conditions are as in Fig. 19.12. The direct wave is

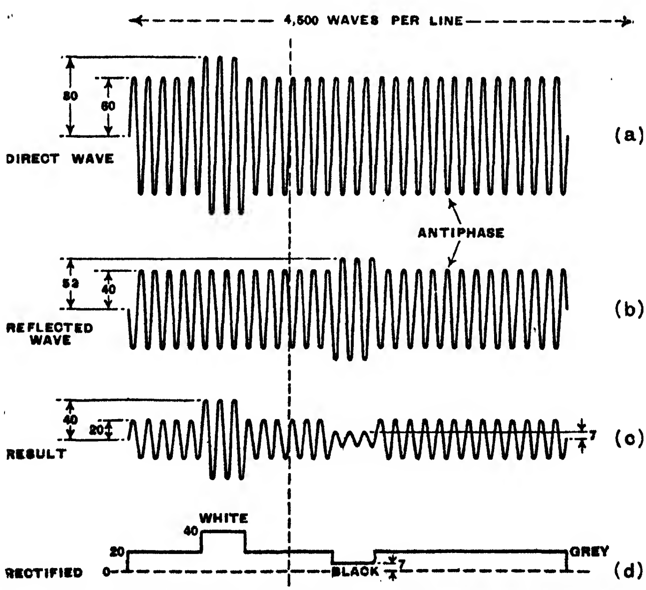



Fig. 19.11—The production of a negative ghost image is clearly shown here

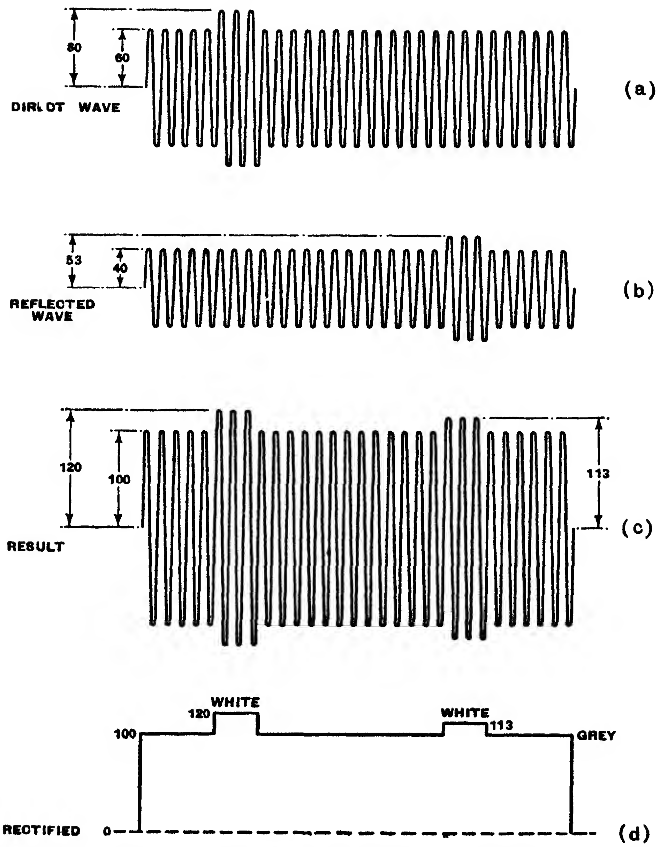



**Fig. 19.12—When the reflected wave is in the same phase as the direct the ghost image is positive**

the same as in Fig. 19.11, and the reflected wave is also the same except that its phase is now opposite and both waves are now in the same phase. The two add together as in (c) and give the v.f. signal of (d). The ghost image is now positive.

The phase obtained depends on the precise position of the aerial and moving it a distance of approximately one half-wavelength will change a ghost image from a positive to a negative.

The possibility of a ghost image is not confined to one, several are possible and each corresponds to a different path for the signal.

The elimination of ghost images is very largely a matter of aerial design. The type of aerial is not as important as its position. It is, however, very desirable that it should be as directional as possible, so that it gives little response to the reflected waves.

Where reception is carried out at a considerable distance from the transmitter and the ghost is widely spaced from the main image, indicating that the reflector is a long way from the receiver, the use of a directional aerial is probably the only way of reducing the ghost. Any alteration which can be made in the aerial position is then so small in relation to all the signal paths that it is likely to make little difference; except, of course, to change a positive ghost to a negative.

The position is rather different when the ghost is very close to the main image. The reflector is then near the receiver and an alteration in the position of the aerial may have quite a large effect.

Ghosts are not confined to long-distance reception and are commonly encountered in towns close to the transmitter, the reflector usually being a steel-framed building. In the majority of cases, however, they can be avoided by careful choice of the aerial and its position. Unfortunately, the correct cure for any given place must be determined experimentally.

Ghost images can also occur when the reflecting object is an aircraft. In this case the ghost is not stationary but moves across the picture. The general variations of signal level, however, often obscure the ghost and usually there are considerable fluctuations in the brightness of the picture while, in bad cases, the synchronizing may also be upset. Aircraft reflections produce a characteristic effect on the picture which is easily recognized once it has been seen, but which is hard to describe in words. It is chiefly produced by aircraft within a few miles of the receiver.

There is no cure for it. It is, however, claimed that some reduction in its effects is possible if the response of the v.f. amplifier is reduced at very low frequencies—say 0·2–5 c/s.

CHAPTER 20

# Special Television Circuits

THE USUAL TELEVISION circuits are dealt with in other chapters, but there are certain special arrangements which are only occasionally used but which deserve comment. Constant-resistance networks are among the more important of these.

The gain control on a television receiver is sometimes made to have only a limited range of control, and the output is adjusted to the correct level when the receiver is first installed by including an attenuator in the aerial circuit. The gain control is then used only for the adjustment of contrast and to correct for valve deterioration.

The input impedance of the receiver is made the same as the impedance of the feeder; a common value is 72 Ω. If the matching is to be correct the input and output impedances of the attenuator must be the same.

For a coaxial cable type feeder a T-network of the type shown in Fig. 20.1 (a), is the simplest, but for a two-wire feeder, a balanced attenuator of the H-type is better (b). If the input and output resistances are $R$, the values of $R_1$ and $R_2$ can readily be calculated from the following equations:

$$R_1 = R\frac{A-1}{A+1} \qquad 20.1$$

$$R_2 = R\frac{2A}{(A+1)(A-1)} \qquad 20.2$$

where $A$ is the ratio of input to output voltage.

As an example, for 20 db attenuation $A$ is 10 and when $R$ is 72 Ω, $R_1$ is 59 Ω and $R_2$ is 14·6 Ω, to slide-rule accuracy.

The resistances should, of course, be non-inductive and stray capacitances must be kept low. Because of stray capacitances, it is

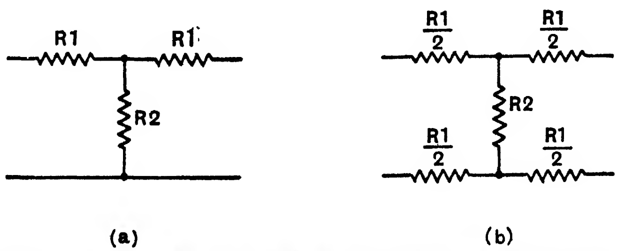



**Fig. 20.1—A T-type attenuator suitable for use with a coaxial cable is shown at (a) and an H-type attenuator for a twin-wire feeder at (b)**

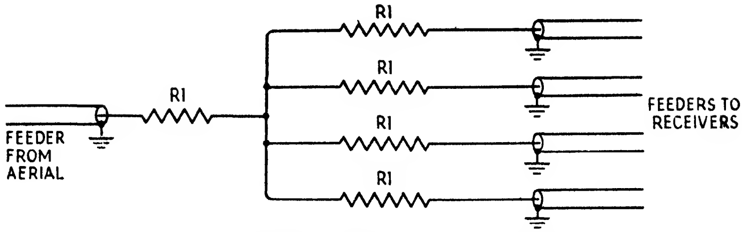



**Fig. 20.2—Circuit for feeding a number of receivers from a common aerial**

better to secure high attenuation from several low attenuation networks in series. The overall attenuation will be the sum (db) of the attenuations of each network.

It is sometimes desired to operate two or more receivers from a common aerial. This can be done by a simple resistance network which keeps all the impedances correct, but introduces some loss. The arrangement is shown in Fig. 20.2. It is required that when each cable is terminated by its characteristic impedance $Z_o$ the impedance presented by the network to each cable shall also be $Z_o$. As the network is symmetrical it is obvious that all the resistances must be alike.

The input impedance at any cable connector with that cable disconnected is

$$R_{IN} = R_1 + (R_1 + Z_o)/n$$

where $n$ is the number of receivers. Now we must have $R_{IN} = Z_o$, therefore,

$$R_1 = Z_o \frac{n-1}{n+1} \qquad 20.3$$

Since the aerial feeder is matched to the network it delivers to it the normal current. As the branches are all identical the current divides equally between them and each receiver is fed with $1/n$ of the current.

The input to each receiver is, therefore, less than the aerial feeder output by 20 log $n$ db. The network loss is one-half of this, for if the aerial feeder power output were divided equally and entirely among the receivers, the receiver input would be less than the feeder output by 10 log $n$ db. This loss is the price which must be paid for having all feeders properly matched.

In practice, $n$ will usually be 2, so the insertion loss is 3 db for each receiver and each set has an input 6 db below what it would have if it were the only set and were connected directly to the aerial feeder. For a 72-Ω feeder, the resistances must each be 24 Ω. With three receivers the loss is 4·77 db and the receiver input is 9·54 db below the feeder output; the resistances must be 36 Ω.

The system is very satisfactory in areas of moderately high field strength, but is clearly to be avoided when signals are weak.

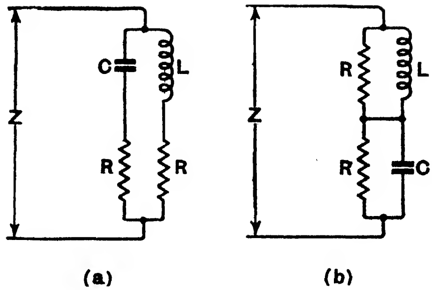



**Fig. 20.3—If the component values are properly related, these networks have a constant resistance R**

Another circuit which can be of great importance is shown in Fig. 20.3, in two forms. If $L/C = R^2$, then $Z = R$ at all frequencies, and the network behaves as a resistance of value $R$. The importance of this is that it is sometimes possible to convert an unavoidable and unwanted capacitance $C$ into a resistance $R$ by adding two resistances $R$ and an inductance $L$.

One application of the circuit is to the smoothing system of the h.t. supply. The ordinary arrangement has the disadvantage that its impedance viewed from the receiver rises greatly at very low frequencies. As can be seen from Fig. 20.4, the impedance $Z$ viewed from the receiver end is at high frequencies virtually only the reactance of $C_1$. The reactance of the choke $L_1$ is too high to have much effect.

As the frequency is lowered the one rises and the other falls and eventually there is resonance with the choke $L_1$ and the capacitors $C_2$ and $C_1$. The impedance is then very high and resistive. At still lower frequencies, it is inductive.

As the frequency is lowered, series resonance between $L_1$ and $C_2$ occurs and the impedance is again resistive and low in value. For lower frequencies still, the impedance is mainly capacitive and rises until at zero frequency it is the total d.c. resistance of the circuit.

These impedance changes are reflected by variations in the gain of even a single stage of v.f. amplification, and where many stages are involved, the effects are even more serious, for feedback comes into play. In a receiver with one v.f. stage the impedance variations are not very important.

They can, however, be overcome by using the arrangement of Fig. 20.5, which is an adaptation of the circuit of Fig. 20.3 (a). The filter impedance $Z$ is equal to $R$ and the impedance of one section consequently forms the resistance in series with $L$. In the case of

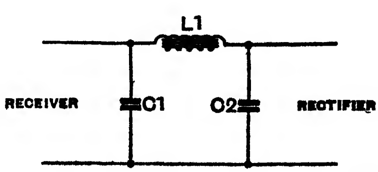



**Fig. 20.4—A typical section of an h.t. smoothing circuit**

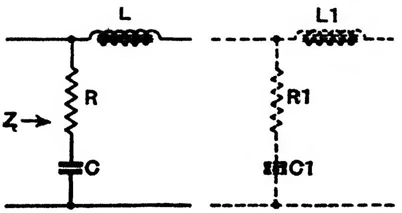



**Fig. 20.5—This diagram shows a constant-resistance smoothing circuit**

the first section the resistance of the rectifier and mains transformer is effective and settles the value of $R$.

Then $L/C = R^2$ and both $L$ and $C$ should be as large as possible in order to obtain good smoothing. In general, the choke $L$ will have resistance which should be added to the output resistance on the right-hand side to obtain the value of $R$.

Thus, suppose the rectifier precedes $L_1$ in Fig. 20.5, and the resistance is 1,000 Ω and $L_1$ has a resistance of 500 Ω. Then $R_1$ is 1,500 Ω and $L/C = 2{\cdot}25 \times 10^6$. With a similar choke resistance in the next stage $R$ becomes 2,000 Ω and $L/C = 4 \times 10^6$. The resistance gets progressively higher as more sections are added; for these values $C_1$ should be 8·9 $\mu$F and $C$ should be 5 $\mu$F with chokes of 20 H.

The constant-resistance feature is a great advantage because it prevents variations of gain in the receiver at low frequencies and it also makes feedback effects relatively simple and easily calculable. The main disadvantage of the circuit is that the smoothing per stage is much less than with a normal filter, because of the resistance $R$ in series with the capacitor $C$.

In the case of a receiver, therefore, it is probably cheaper to use a stabilized h.t. supply, for the voltage and current are only moderate. The neon stabilizer gives a substantially constant and very low impedance down to the lowest frequencies. It may need shunting with a capacitor to be effective at high frequencies.

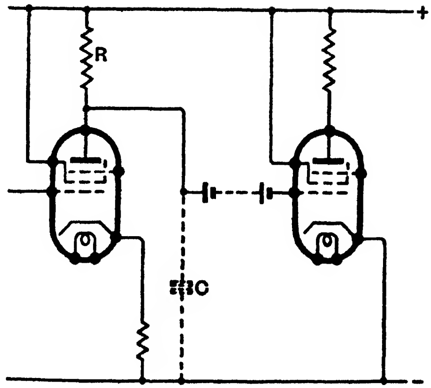
Conditions in a transmitter are rather different, for voltage and current are too high for neon stabilizers to be practicable. The constant-resistance network, therefore, is used for the h.t. smoothing in the Alexandra Palace transmitter* and the poor smoothing efficiency is overcome by using a 500-c/s supply instead of the usual 50 c/s.

One case in which the circuit can be used to get rid of a stray capacitance occurs when direct coupling is used between two stages. A battery must be employed in the coupling to maintain the correct grid potential on the second valve and has a stray capacitance $C$ to earth which shunts the coupling resistance $R$ as shown in Fig. 20.6.

The effect of this capacitance can be avoided by arranging the circuit as in Fig. 20.7. Here the two resistances $R$ with $L$ and $C$ form the network of Fig. 20.3 (b). The two coils have the same

* "The Marconi-E.M.I. Television System, Part II. The Vision Input Equipment," by C. O. Browne, B.Sc., *The Journal of the Institution of Electrical Engineers*, December 1938.

Fig. 20.6—With direct coupling the bias battery has a capacitance C to earth




inductance and are very tightly coupled; there is, however, a certain leakage inductance $L_1$ which appears in series with the secondary. Also, the battery has an internal resistance $R_1$.

$C_1$ and $R_1$ are included, therefore, to offset the effects of the battery resistance and $L_1$, which they do by forming the network of Fig. 20.3 (a). When $L/C = R^2$ and $L_1/C_1 = R_1{}^2$, the effect of the capacitance $C$ is removed and the effective coupling consists of $R$ only. Of course, the input and output capacitances of the valve remain.

We now come to valve circuits and of these the cathode-follower is undoubtedly the most important. The circuit is shown in Fig. 20.8, and it will be seen that the input is applied between grid and

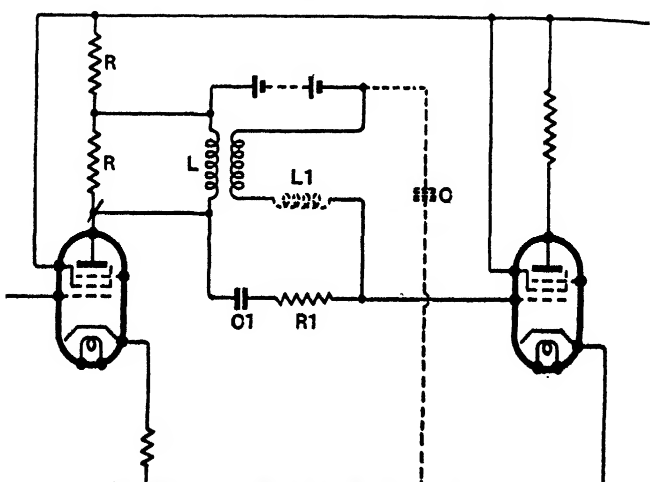



Fig. 20.7—The effect of the battery capacitance to earth can be converted into that of a resistance R by means of the network shown here

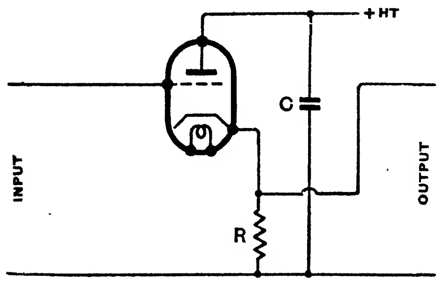



Fig. 20.8—The basic cathode-follower circuit

earth while the output is taken between cathode and earth, across the resistance *R*. The anode is maintained at earth potential to alternating currents by the large capacitor.

The output is in the same phase as the input and is slightly less than the input in magnitude. There is thus no voltage amplification but slight attenuation. The chief value of the stage lies in its possession of a very low output impedance—of the order of a few hundred ohms only. It can consequently be used to match a low impedance circuit and so can be employed as a form of impedance transformer. A cathode-follower can be designed to have an output impedance to match a low-impedance cable.

Such a cable can be terminated at the other end by a resistance across which an ordinary amplifier is connected. It is, however, possible to employ a modification of the cathode-follower in which the input is applied between cathode and earth and the output is taken bewteen anode and earth.

The circuit is shown in Fig. 20.9. It should be pointed out that here, as in Fig. 20.8, only the general arrangements are shown. Detailed modification is often made to obtain the correct grid bias.

The cathode-follower is also often used in modified form with a coupling resistance in the anode circuit as well as in the cathode. When these resistances are of the same value, the output at the

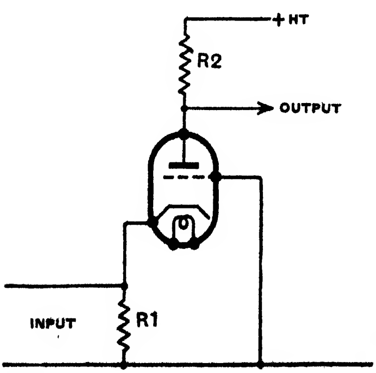



Fig. 20.9—The cathode-input circuit shown here is sometimes adopted

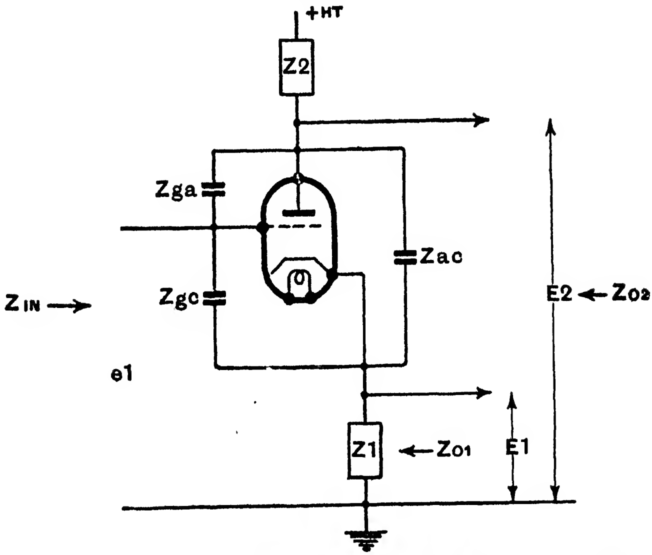



**Fig. 20.10—This diagram shows the general arrangement of a cathode-follower having anode as well as cathode circuit impedances and appreciable interelectrode capacitances**

anode is numerically equal to that at the cathode but of opposite phase. Such an arrangement is widely used as a phase-splitter for feeding push-pull a.f. amplifiers.

It also has very useful applications in television receivers, and the circuit merits detailed consideration. The general circuit, therefore, is shown in Fig. 20.10, and it is a straightforward but laborious process to derive equations for the input and output impedances and for the amplification. Unfortunately, the resulting equations are too complex to be of much general use. The complexity is, of course, due to the inclusion of the interelectrode capacitances and when these are negligible the equations are quite simple.

We shall, therefore, only deal with the more important cases here, first with the general case neglecting valve capacitances. We have:

$$\left.\begin{aligned} A_1 &= \frac{E_1}{e_1} = \frac{\mu Z_1}{r_a + Z_2 + Z_1(1+\mu)} \\ A_2 &= \frac{E_2}{e_1} = \frac{-\mu Z_2}{r_a + Z_2 + Z_1(1+\mu)} \\ Z_{01} &= \frac{Z_1(r_a + Z_2)}{r_a + Z_2 + Z_1(1+\mu)} \\ Z_{02} &= \frac{Z_2\{r_a + Z_1(1+\mu)\}}{r_a + Z_2 + Z_1(1+\mu)} \end{aligned}\right\} \qquad 20.4$$

$$Z_{in} = \infty$$

The true cathode-follower has $Z_2 = 0$. $Z_{ga}$ then comes across the input terminals and $Z_{ac}$ is in shunt with $Z_1$. In the expressions which follow $Z_{ac}$ is assumed to be included in $Z_1$ and $Z_{ga}$ is not included in $Z_{in}$. This is done for simplicity in the formulæ, for $Z_{ac}$ and $Z_{ga}$ can readily be taken into account separately when required. We have:

$$\left.\begin{aligned} A_1 &= \frac{Z(g_m + 1/Z_{gc})}{1 + Z(g_m + 1/Z_{gc})} \\ Z_{in} &= Z + Z_{gc}(1 + g_m Z) \\ Z_{01} &= \frac{Z}{1 + g_m Z} \end{aligned}\right\} \qquad 20.5$$

where $Z = Z_1 r_a/(Z_1 + r_a)$

The third case is when $Z_1 = 0$. This corresponds to the case of an ordinary amplifier. $Z_{gc}$ then comes across the input and is disregarded below, while $Z_{ac}$ is assumed to be included in $Z_2$. Then

$$\left.\begin{aligned} A_2 &= \frac{Z/Z_{ga} + g_m Z}{Z/Z_{ga} + 1} \\ Z_{in} &= \frac{Z - Z_{ga}}{1 + g_m Z} \end{aligned}\right\} \qquad 20.6$$

where $Z = Z_2 r_a/(Z_2 + r_a)$

A case of particular interest arises with the cathode-follower when $Z_{gc}$ is the reactance of a capacitance $c_{gc}$ and $Z_1$ consists of a resistance $R_1$ and capacitance $C_1$ in parallel. The input impedance $Z_{in}$ can be expressed in the form of a resistance $R_{in}$ and capacitance $C_{in}$ in parallel where

$$R_{in} = R_1 \frac{(1 + C_1/c_{gc})^2 + (1 + R_1/r_a + g_m R_1)^2/\omega^2 c_{gc}^2 R_1^2}{1 + R_1/r_a - g_m R_1 C_1/c_{gc}}$$

$$20.7$$

$$C_{in} = c_{gc} \frac{1 + c_{gc}/C_1 + (1 + R_1/r_a)(1 + R_1/r_a + g_m R_1)/\omega^2 C_1^2 R_1^2}{(1 + c_{gc}/C_1)^2 + (1 + R_1/r_a + g_m R_1)^2/\omega^2 C_1^2 R_1^2}$$

The point of particular importance is that $R_{in}$ is negative when $g_m R_1 C_1/c_{gc} > 1 + R_1/r_a$ and this is a normal condition. Thus if $R_1 = r_a = 10{,}000\ \Omega$ and $g_m = 2$ mA/V, while $c_{gc} = 5$ pF and $C_1 = 10$ pF, then at 2 Mc/s $R_{in} = -35{,}000\ \Omega$.

The input capacitance, which naturally varies with frequency, is quite small, being in the case considered only 0·455 pF at very low frequencies and rising to 0·48 pF at 2 Mc/s. The input capacitance of the stage as a whole is thus substantially the grid-anode capacitance only.

One case when both cathode and anode coupling impedances are used is capable of a reasonably simple solution. This is when $Z_1 = Z_2$ and $Z_{gc} = Z_{ga}$; this is the push-pull phase-splitting circuit widely used in a.f. amplifiers. It has certain television applications (Fig. 21.1).

Under these conditions:

$$\left.\begin{aligned} Z_{in} &= \frac{Z_1+Z_{gc}}{2} \\ A_1 &= \frac{g_m + Y_{gc} + \dfrac{2}{Z_1 + Z_{gc}}}{g_m + Y_{gc} + 2Y' + Y_1} \\ A_2 &= \frac{g_m + Y_{gc} + \dfrac{2}{Z_1 + Z_{ge}}(1 - g_m Z_1 Z_{gc} Y')}{g_m + Y_{gc} + 2Y' + Y_1} \end{aligned}\right\} \quad 20.8$$

where $Y_{gc} = 1/Z_{gc}$; $Y' = 1/Z' = j\omega c_{ac} + 1/r_a$; $Y_1 = 1/Z_1$;

It will be noted that $A_1$ is not equal to $A_2$. It will be found, however, that in most practical cases the difference is small.

A modification of these cathode-impedance circuits is shown in Fig. 20.11. The input is applied to the cathode and the output is taken from the anode. The grid-cathode capacitance appears across $Z_1$ and is included in it, while the grid-anode capacitance appears across $Z_2$ and is included in it. The input impedance is low and given by

$$\left.\begin{aligned} Z_{in} &= \frac{Z_2 + r_a + Z_2 r_a/Z_{ga}}{1 + \mu + r_a/Z_{ga}} \\ A &= \frac{e_2}{e_1} = \frac{Z_2(1 + \mu + r_a/Z_{ga})}{Z_2 + r_a + Z_2 r_a/Z_{ga}} = \frac{Z_2}{Z_{in}} \end{aligned}\right\} \quad 20.9$$

The output is in the same phase as the input, just as with the cathode-follower. The stage gain, however, is not of the order of unity, but actually very slightly greater than the gain of an ordinary amplifier.

This fact together with the absence of a phase-reversal makes the stage particularly useful. Unfortunately, the input impedance is

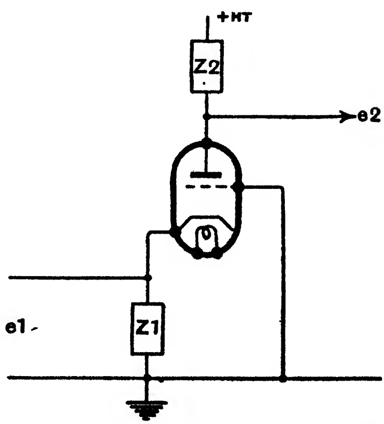



**Fig. 20.11—This diagram shows the generalized circuit of a cathode-input stage**

too low for it to be of wide application. It can, however, be used for terminating a low impedance line. The cathode-follower can be regarded as a step-down impedance transformer and is used at the start of a line. The cathode-input circuit can be regarded as its complement—as a step-up impedance transformer—and used at the end of the line.

The fact that the cathode-follower can feed a low-impedance circuit requires some qualification, for this stage must not, in general, be used to feed a circuit of high capacitance. When the cathode resistance $R$ of Fig. 20.8 is shunted by capacitance, there is a serious risk of the valve having its anode current cut off by a negative-going input pulse of an amplitude which the stage would otherwise easily handle.

Suppose that with no capacitance across $R$ the valve will handle an input of 50 V and that cut-off occurs when the grid is 10 V negative with respect to the cathode. Then take the initial condition as being with the grid and cathode each 50 V above earth and the grid-cathode potential as zero. A negative-going input pulse of 60 V amplitude would then bring the grid to $-$ 10 V with respect to earth and the cathode to earth. The grid-cathode voltage would be $-$ 10 and the valve would be just at the cut-off point.

When there is capacitance across $R$ conditions are very different. Initially the capacitance is charged to 50 V, but it cannot change its charge instantaneously so that when the negative-going pulse occurs, the cathode remains at + 50 V. The grid is at $-$ 10 V, so that the grid is 60 V negative with respect to cathode and the valve is cut off. Any input exceeding the normal grid base of the valve results in anode current cut-off, so that in this case the input must be restricted to 10 V only.

The valve remains cut off until the charge on the capacitance has leaked away through $R$ and this occupies a time dependent only on their values. The normally low output impedance of the valve is ineffective because it is cut off.

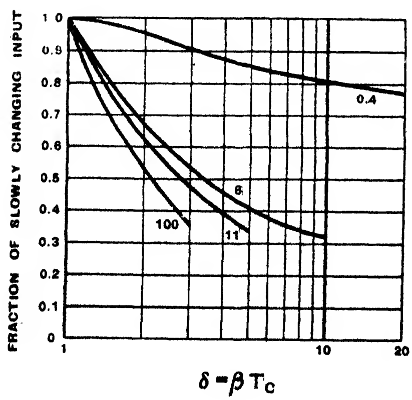



**Fig. 20.12—These curves show how the cut-off factor is related to the time-constants of the cathode circuit and input voltage, defined by $\beta T_c$, and the valve characteristics defined by B**

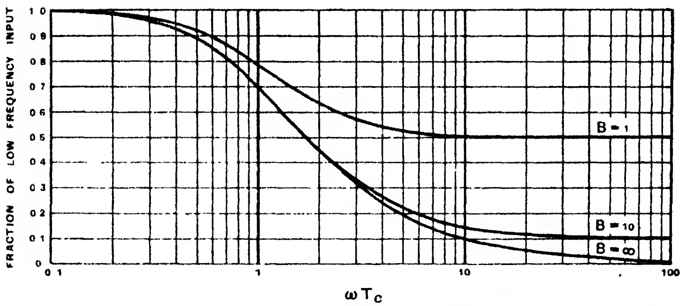



**Fig. 20.13—The relation between the cut-off factor and $\omega T_C$ is shown here for sine waves for several values of B**

Looking at the matter in terms of frequency response, the valve is cut off because the capacitance forms a virtual short-circuit to $R$ at high input frequencies and so prevents the normal negative feedback from occurring. It is this negative feedback which enables the stage to handle a large input under normal conditions.

If the cathode-follower is used feeding a circuit of appreciable capacitance it is necessary to restrict the input voltage to a value not greatly in excess of that which the valve will handle without feedback. Alternatively, the cathode-circuit impedance must be kept substantially constant up to the highest frequency included in the input. If it can be done, the latter is the better.

When a cathode-follower is used, therefore, the cathode circuit must be treated as carefully as the anode circuit of an ordinary amplifier. The same remedies for the unwanted effects of stray capacitance are applicable—in particular the use of a correcting inductance in series with the cathode resistor.

Viewing the matter from the pulse standpoint, the requirement for the avoidance of cut-off is that the time-constant of the cathode circuit be small enough to permit the cathode voltage to change at least as rapidly as the grid voltage.

The performance of the phase-splitter (Fig. 20.10) is almost identical with that of the cathode-follower, there being only a small numerical difference.

The approximate calculation of the cathode resistance is easily carried out* with the aid of the curves of Figs 20.12 and 20.13. The former relates a quantity $\beta T_c$ to the fraction of the maximum slowly changing input which can be applied without causing cut-off. There are several curves for different values of $B$. The circuit is that of Fig. 20.14, and $B = (2 + \mu)R_c/r_a$ when $R_1 = R_c$ and $C_1 = C_c$; for the true cathode-follower $R_1 = 0$ and then $B = (1 + \mu)R_c/r_a$ .

The curves are based on the assumption that the input is a pulse with an exponential leading edge; that is, a perfect pulse pre-distorted

* "Cathode-Follower Internal Impedance," by Harold Goldberg. *Proc. Inst. Radio Engrs.* November 1945.
"Cathode-Follower Dangers," by W. T. Cocking, *Wireless World*, March 1946.

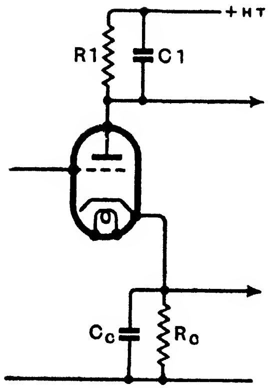



**Fig. 20.14—The push-pull phase-splitter is a modified cathode-follower. The determination of Rc is the same for both if $R_c = R_1$ and $C_c = C_1$**

by its passage through a single *RC* circuit of time-constant $1/\beta$. It is often desirable to relate this to the frequency response of the preceding circuit and it is convenient to do this by taking $\beta = \omega/\sqrt{3}$ where $\omega$ is 6·28 times the frequency for which the response is $-$ 3 db. This assumes that the preceding circuits have the same effect on the waveform as a single *RC* circuit, which is not correct. There is, however, no general accurate solution possible and this assumption does enable an approximation to the correct value to be found.

As an example of the use of the curves consider a cathode-follower with $\mu = 10$, $r_a = 3\ \text{k}\Omega$, $C_c = 30$ pF and the preceding circuits having a response of $-$ 3 db at 2·5 Mc/s. What is the maximum value of $R_c$ (a) to avoid cut-off at full input and (b) for cut-off at inputs exceeding 0·9 of the full?

We have for (a) $\beta = \omega/\sqrt{3}$ and from Fig. 20.12 we see that for full output $\beta = 1/T_c = 1/C_cR_c$. Therefore $R_c = \sqrt{3}/\omega C_c$ and is 3·67 kΩ.

In the case (b) we require to know $B$ but cannot calculate it until $R_c$ is known. However, in the region of 0·9 for the cut-off factor the results are only slightly dependent on $B$. We may expect $R_c$ to be a little above the value in (a) and so can estimate $B$ on the basi s of $R_c$ being, say, 5 kΩ. This gives $B = 18{\cdot}35$ and from Fig. 20.12, $\beta T_c = 1{\cdot}16$. Therefore, $R_c = 4{\cdot}25\ \text{k}\Omega$.

The curves of Fig. 20.13 give similar information for sine-wave inputs. Here $\omega = 6{\cdot}28$ times the input frequency.

Although it has not yet found its way into commercial receivers, the latest development, "spot-wobble," deserves mention. It is a method of reducing the visibility of the line structure and of interline flicker. It reduces the vertical definition somewhat, but leaves the horizontal unaffected. In effect, the scanning spot is elongated in the vertical direction by deflecting it very rapidly a very small amount. This is done by feeding a special deflector coil, additional to the ordinary deflector coils, with current at a frequency of about

10 Mc/s. The power needed is small and within the capabilities of a single-valve oscillator.

Spot-wobble appears to be definitely advantageous when a large, bright and very well focused picture is obtained. It is of doubtful advantage with a picture under 7-in in height and is certainly of no value when the focus is poor.

CHAPTER 21

# The Complete Receiver

ALL THE INDIVIDUAL SECTIONS of a television receiver have been described in the previous chapters in considerable detail. It is now necessary to give some indication of how they fit together to form a complete television receiver. This is probably best done by giving a number of complete circuit diagrams of both old and new apparatus.

There are, of course, many forms which a complete set can take. It will be obvious from the earlier chapters of this book that in each section of the equipment there are many alternative ways of obtaining the required action. Some are intrinsically better than others, but it is necessary for each section to fit in with and work with other sections, so that it is rarely possible to choose the form of any section solely on its own merits. It must be selected also to fit in with other sections.

Even then the possible arrangements become quite large and it is impracticable to illustrate more than a few. Only sets with electromagnetic deflection will be described for the alternative electric deflection is now virtually obsolete for television. Its use is now confined to the experimenter and, even then, to small tubes of 6-in diameter and under. For the benefit of those interested in early television sets however, it may be useful to point out that full constructional details of a set embodying a 12-in electrostatic tube have been published.*

An early example† of a receiver including an electromagnetic tube is illustrated in Figs. 21.1–3. The receiver circuit is shown in Fig. 21.1 and it will be seen that it is a superheterodyne having one s.f. stage $V_1$, a triode-hexode frequency-changer $V_2$, three i.f. stages $V_3$, $V_4$ and $V_5$ at a frequency of 13 Mc/s, a detector $V_6$ and one v.f. stage $V_7$. Following this there is a phase-splitter $V_9$ with d.c. restoration by $V_8$ on its input, and a sync separator $V_{10}$.

The set does not include any provision for sound-channel reception since it was intended that an entirely separate receiver should be used for this. Because of this absence of a sound channel the receiver is very flexible in its vision operation and can be adjusted for either single- or double-sideband operation merely by altering the oscillator frequency. In normal operation the set was used chiefly in the single-sideband condition to select the sideband remote from the sound channel because only in this way could freedom from sound-

* *Wireless World*, July 2, 9, 16, 23 and 30, 1937.

† *Wireless World*, June 29, July 6, 13 and 20, 1939.

channel interference be readily secured, owing to the absence of sound rejectors.

The oscillator operates at 58 Mc/s—above the vision carrier. This frequency was chosen because at the time the set was designed (1939) there were virtually no signals at the second-channel frequency of 71 Mc/s. This is now hardly a valid reason and one would now usually choose the alternative frequency of 32 Mc/s in order to obtain a more stable oscillator performance. In this set, however, on account of the absence of sound-channel circuits—in particular, rejectors—quite a large amount of oscillator drift can be tolerated and the frequency of 58 Mc/s proved quite satisfactory.

The s.f. and first i.f. stages have adjustable grid bias for gain control and the network $R_{13}$, $R_{14}$, $R_{15}$ is arranged to vary both $g_1$ and $g_3$ bias in a fixed ratio in order to prevent serious changes of the input capacitance or resistance of the valves. All intervalve couplings are single circuits stagger-tuned and a bandwidth of 3 Mc/s can be obtained by their proper adjustment.

The output of the v.f. stage $V_7$ is positive-going on the picture, and negative-going on the sync pulses. The tube used had the cathode internally joined to the heater and so the signal had to be fed to its grid. It was considered unwise to join the grid of the tube directly to the anode of $V_7$ and so a capacitor coupling with d.c. restoration was necessary.

The output of $V_7$ could have been fed directly to a sync separator but, unless this was a diode type or had two stages, its output would have been in the wrong phase to suit the saw-tooth generators. As will be evident from the chapter on sync separation, the most satisfactory form of limiter is a tetrode or pentode which limits the picture signal by anode current cut-off. It requires a negative-going picture input and delivers negative-going sync pulses at its output. This is correct for the saw-tooth generators in this set, but the signal is in the opposite phase to that needed by the tube.

In order to reconcile these conflicting requirements the phase-splitter $V_9$ was included. The d.c. component of the signal is restored on the grid by $V_8$ and the tube is fed from the cathode by direct connection. An output of opposite phase is taken from the anode to the sync separator $V_{10}$, d.c. restoration being again effected, this time by the grid-cathode path of $V_{10}$.

The time bases used in this set are saw-tooth current generators. The line-scan oscillator is $V_{12}$ of Fig. 21.2. The winding $L_{11}$ of $T_1$ acts as an auto-transformer to feed the deflector coil $L_{13}$; a " damping diode " $V_{13}$ is used. The sync pulses are fed to the grid of the generator through a diode $V_{11}$.

The frame circuit is similar, but the coils $L_{14}$, $L_{15}$, $L_{16}$, and $L_{17}$ of the generator $V_{15}$ are wound on the deflector yoke itself. The circuit is essentially the same as that of the line and the sync pulses are fed through the diode $V_{14}$ after being integrated by $R_{34}$ and $C_{27}$. In order to keep the picture central it is necessary for the mean flux in the core of the deflection yoke to be zero and this is achieved with

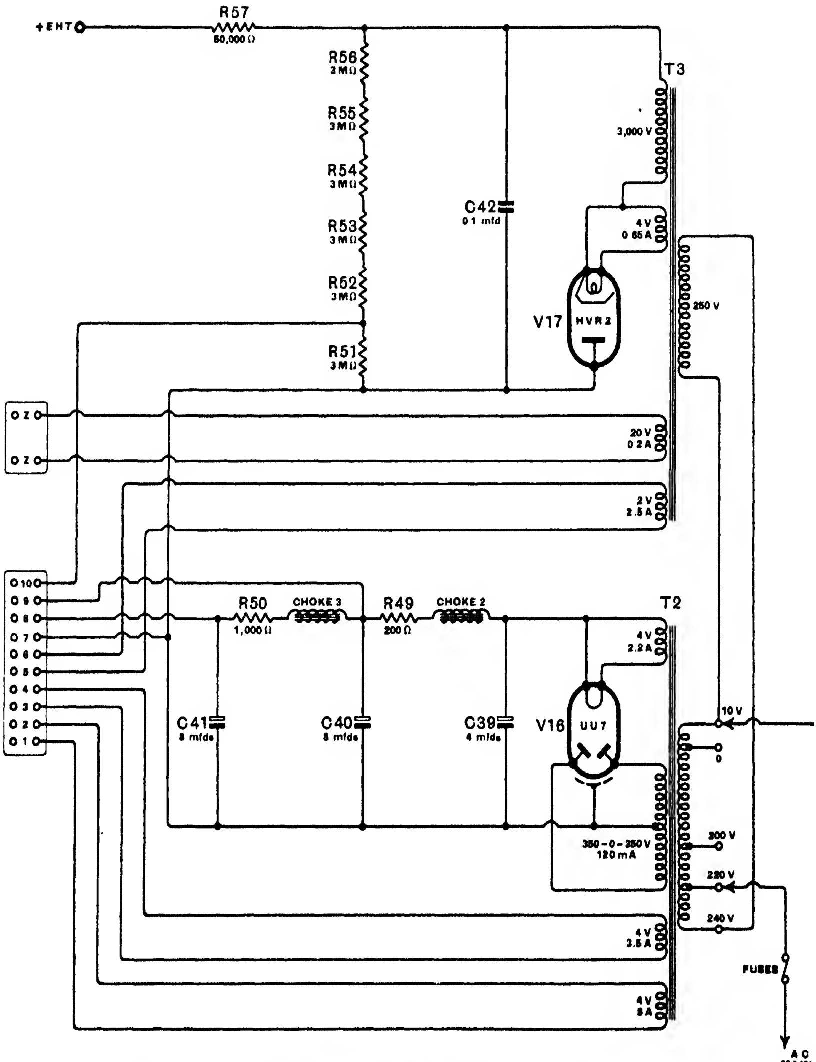



**Fig. 21.3—This diagram shows the arrangement of the power unit for the receiver of Figs. 21.1 and 21.2**

the aid of $L_{16}$ and $L_{17}$. Current is fed through them to buck the difference between the mean grid and anode currents in $L_{14}$ and $L_{15}$. Changing currents are kept out of $L_{16}$ and $L_{17}$ by $C_{35}$ in conjunction with the choke.

The focus coil is connected in series with the time-base h.t. supply and the total current through it is adjustable by $R_{47}$. The power unit is shown in Fig. 21.3 and is self-explanatory.

In its mechanical form this receiver was built in two main units—receiver and power pack. The former comprised two sub-units bolted together. One was a steel chassis holding everything shown in Fig. 21.2, including the tube, and the other was an aluminium channel for the receiver proper of Fig. 21.1. The valves were mounted in line, except for the detector, v.f. stage and sync separator which were on a side member. The canned coils were mounted between the valves and served as screens between them, while the oscillator components were in a can alongside the valve. No under-chassis screening was used. A top view of the set is shown in Fig. 21.4.

Details of another more modern receiver, again for electromagnetic deflection, are shown in Figs. 21.5–13. This set was designed primarily for the constructor at a time when components and materials were in very short supply and consequently full constructional details were given of all special television components.* The design was also carried out to be as non-critical as possible in order to permit quite wide variations in some materials and this accounts for the fact that in some places more valves are used than would be normal in commercial practice.

A unit construction was adopted and two alternative receiver units were described. One was a straight set and the other a superheterodyne.†

* *Wireless World*, January, February, March, May, July, August, September, October, November and December 1947. Reprinted under the title "Television Receiver Construction." See also April 1949.

† *Wireless World*, February and March 1949. Reprinted under the title "Superheterodyne Television Unit."

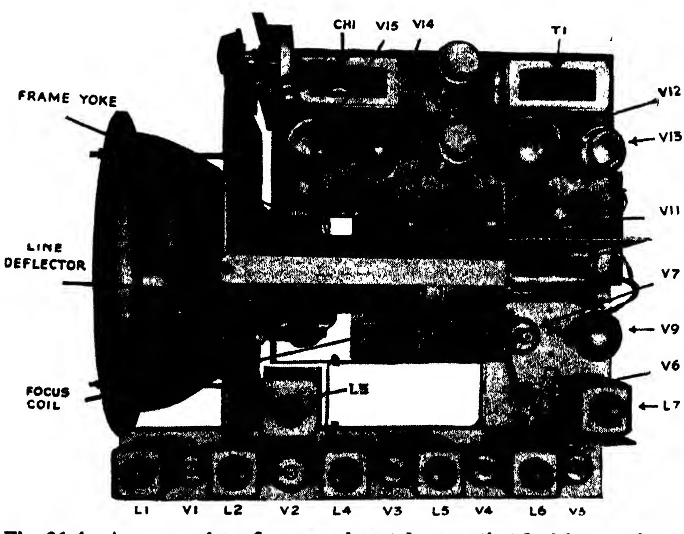



**Fig. 21.4—An upper view of an experimental magnetic television receiver**

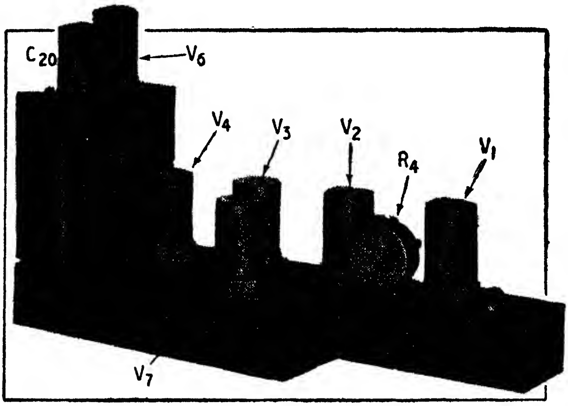



**Fig. 21.6—A general view of the receiver unit is shown here. The vision and sound channels are in two strips bolted together and the v.f. components are included in a rear compartment.**

The first unit—the straight set—has the circuit shown in Fig. 21.5. It is designed for double-sideband operation on Channel 1, Alexandra Palace, and gives sound and vision outputs. There are four s.f. stages in the vision chain of which $V_1$ and $V_2$ operate on both channels and $V_3$ and $V_4$ on vision only. Single-circuit stagger-tuned couplings are used to provide the 6-Mc/s bandwidth. The input resistance and capacitance of the valves are stabilized by 33-Ω cathode resistors shunted by 50-pF capacitors—$R_2$ and $C_2$ being an example.

A sound-channel rejector $T_8$ is included and the sound pick-out circuit $T_6$ is tuned to act as a further rejector. One stage $V_7$ is provided in the sound channel and feeds a diode detector $V_8$. No a.f. amplification is included since the set is intended for use with any good a.f. amplifier or can even feed into the pick-up terminals of a standard broadcast receiver.

On the vision side a diode detector is used with a compensated load $R_{19}$, $L_3$ and one v.f. stage. This has a compensated load $R_{21}$, $L_4$, $L_5$ and feeds the signal to the cathode of the c.r. tube through the network $R_{23}$, $R_{24}$, $C_{21}$. In conjunction with $R_{20}$, $R_{21}$ and $C_{20}$ this forms a circuit which has a substantially flat response down to zero but which limits the potential of the output terminal. It is included to prevent the heater-cathode rating of the tube—100 V—from being exceeded.

The r.f. coils are all slug-tuned with 0 B.A. brass slugs in formers of $\frac{3}{8}$-in outside diameter. The vision channel coils are double-wound for stability. The chassis is a copper channel with screens across the valveholders for the single-ended valves. The purely sound-channel parts are in a similar channel mounted alongside and bolted to the vision chassis. The v.f. components are in a separate compartment above chassis and designed to bring the output terminal close to the tube base. A view of the receiver unit is given in Fig. 21.6.

The alternative superheterodyne unit was designed for higher sensitivity and to be more easily adaptable for Channel 4, Birmingham. The circuit is shown in Fig. 21.7. The s.f. stage $V_1$ and the frequency-changer $V_2$ and $V_3$ are common to both sound and vision channels. There are three i.f. stages $V_8$, $V_9$ and $V_{10}$ operating single-sideband with a bandwidth of 3 Mc/s—10–13 Mc/s, the vision carrier being at 13 Mc/s. In the sound channel there are two i.f. stages $V_4$ and $V_5$ at 9·5 Mc/s, the bandwidth being about 100 kc/s.

Stagger-tuned circuits are used in the vision channel. The s.f. couplings $T_1$ and $T_2$ are double wound for stability, but in the i.f. amplifier single windings $L_8$ and $L_{10}$ are sufficient at the lower frequency. $T_5$ is double wound, however, this time because the detector requires a low-resistance circuit. Between $V_2$ and $V_8$ a link-coupled pair of staggered circuits is used. This is done largely because, for mechanical convenience, the two valves are on different chassis and the connecting cable is about a foot long and has a capacitance of about 22 pF. To prevent this from having any appreciable effect a considerable step down from the tuned circuits is needed at each end. With the type of coil used it is not possible to obtain sufficient coupling for the bandwidth needed without also staggering the resonance frequencies of $T_3$ and $T_4$.

The vision channel is provided with the sound-channel rejectors $L_7$, $L_9$ and $L_{11}$ which together give 40 db attenuation of this signal. The coils used are of the same form as those in the straight set but for all sound channel coils, including the rejectors, polystyrene formers and copper slugs are used in order to reduce losses.

The detector and v.f., $V_{11}$ and $V_{12}$, circuits of the vision channel are substantially the same as in the straight set, but a simple noise limiter $V_{13}$ is added.

In the sound channel the detector $V_6$ is followed by a noise limiter $V_7$ and again no a.f. amplification is included. Both $V_6$ and $V_7$ are germanium crystal-valves, but thermionic diodes can be used.

The gain control $R_3$ operates to vary the bias on $V_1$, $V_4$ and $V_8$, thus keeping the relative gain of the sound and vision channels constant.

Whichever receiver is used, its output is taken to the tube cathode directly and through a 0·1 $\mu$F capacitor to the frame time-base unit. The circuit of this is shown in Fig. 21.8 and d.c. restoration is effected at the grid of $V_1$ since the sync pulses are positive-going. This valve is the main sync separator and the output at its anode is taken to the line time-base unit and also through the critical time-constant $C_2R_5$ to $V_2$. This is a further limiter which acts in conjunction with $C_2R_5$ as an inverse frame-pulse separator, the valve conducting only on the back edges of the frame pulses.

The saw-tooth generator is a blocking oscillator $V_3$. It is synchronized by $V_2$, the anodes of the two valves being joined together. When $V_2$ conducts its current flows through $L_1$ and produces a negative pulse on the anode of $V_3$. A positive pulse is also produced on the grid by the transformer action and this triggers the valve.

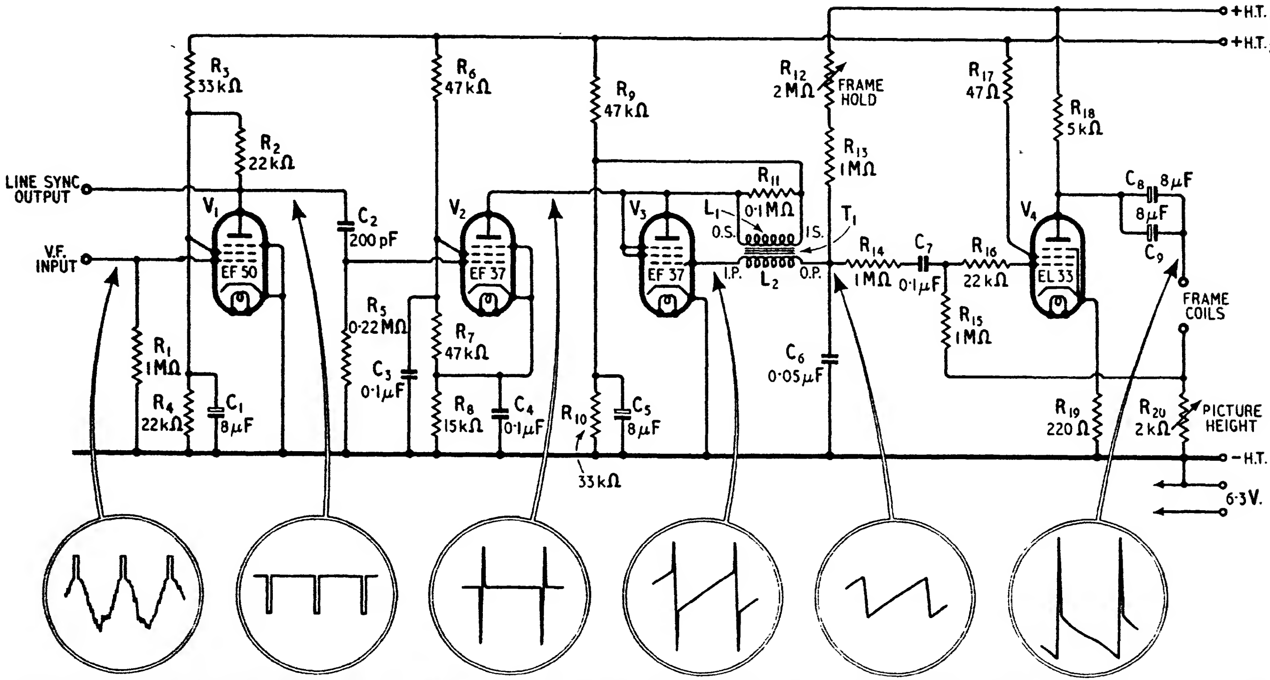
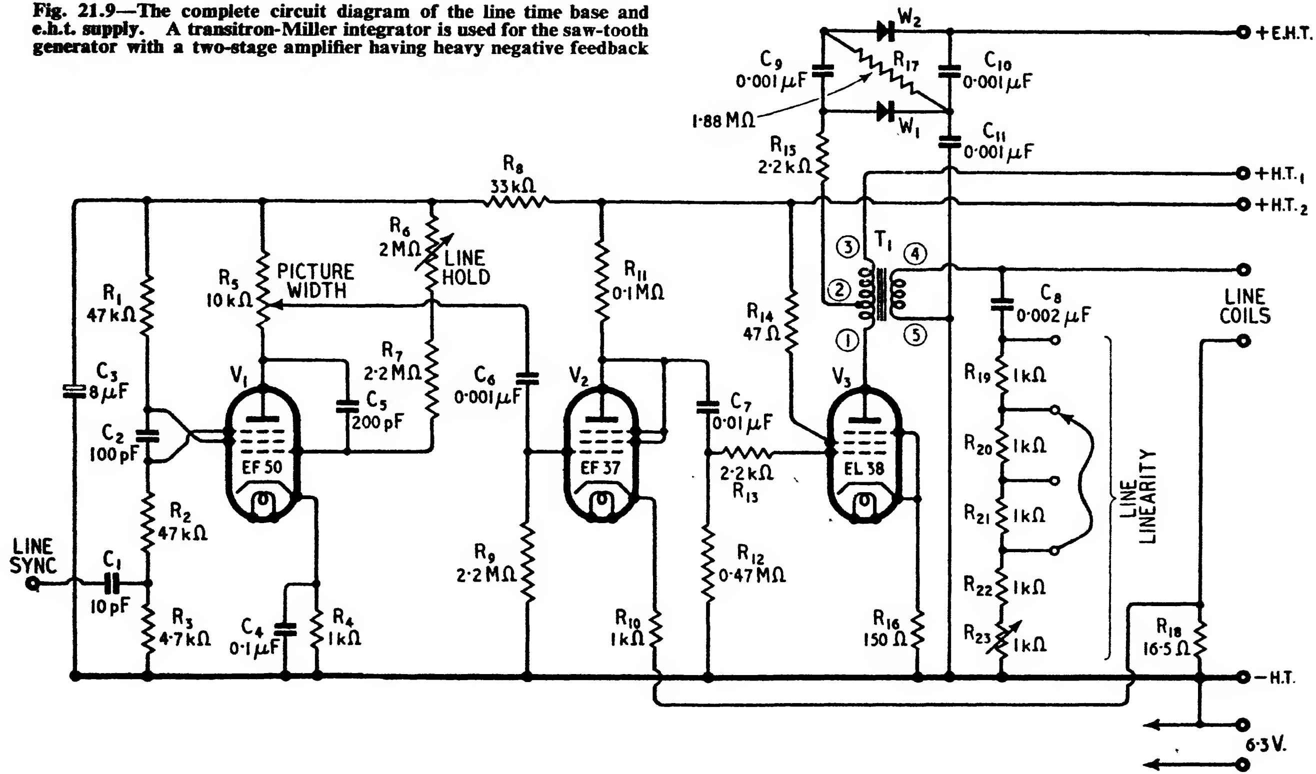



**Fig. 21.8—The circuit diagram of the sync separator and frame time base is shown here. $V_1$ is the main sync separator and $V_2$ the frame pulse separator ; $V_3$ is the frame blocking-oscillator saw-tooth generator and $V_4$ the output valve feeding the deflector coils**

**Fig. 21.9—The complete circuit diagram of the line time base and e.h.t. supply. A transitron-Miller integrator is used for the saw-tooth generator with a two-stage amplifier having heavy negative feedback**



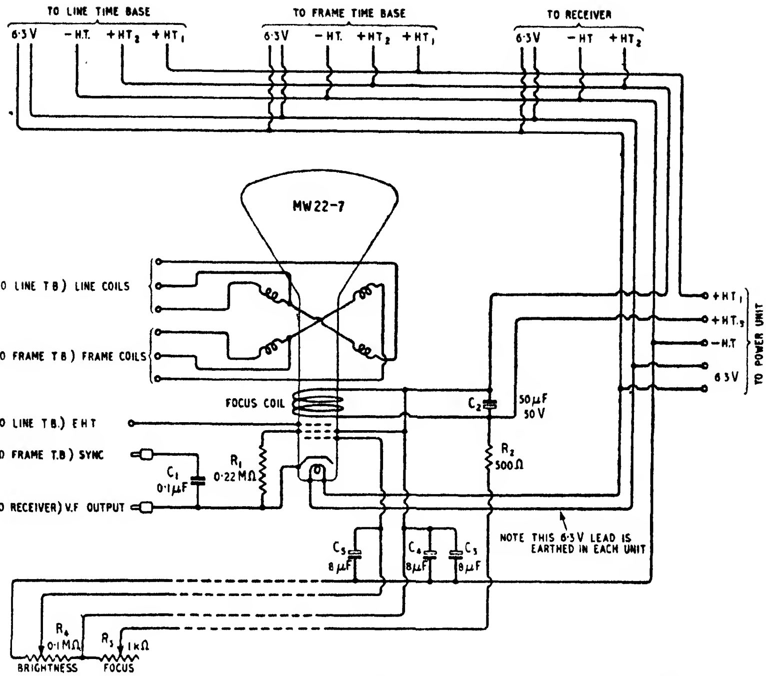



**Fig. 21.10—The connections on the main frame are shown in this diagram. They comprise the inter-unit connections and the leads to the c.r. tube, deflector coils and focus coil, and a few associated components**

The saw-tooth voltage across $C_6$ is applied to the pentode $V_4$ which feeds high-inductance deflector coils through an RC coupling $R_{18}$, $C_8$, $C_9$. The variable resistor $R_{20}$ acts as a picture height control and is in series with the deflector coil. It operates by providing negative feedback in conjunction with $R_{14}$ and $R_{15}$. Linearity is obtained by the curvature of the valve characteristics which is arranged to offset the inverse curvatures of the other circuit elements. Negative feedback also helps in producing linearity.

The line time-base unit consists of a Miller-integrator transitron saw-tooth generator $V_1$, Fig. 21.9, to which the line sync pulses are applied through a differentiating circuit $C_1R_3$. The linear negative-going saw-tooth at the anode is applied to the output valve $V_3$ through a phase-reversing stage $V_2$. Negative feedback is applied to the cathode of $V_2$, the feedback voltage being developed across $R_{18}$ in series with the low-inductance deflector coils. The feedback is considerable and an exceedingly linear scan is obtained whenever $V_3$ is conductive. Immediately after the fly-back $V_3$ is still cut off, and the linearity here is controlled by the damping circuit $C_8R_{19-23}$.

The high-voltage supply for the tube is taken from the pulse developed on fly-back, a voltage-doubler using metal rectifiers $W_1$ and $W_2$ being employed. A supply at 5 kV is obtained.

The interconnections of the various units together with the focus coil and control are shown in Fig. 21.10 and the power unit in Fig. 21.11.

The three basic units, receiver, line time base and frame time base are hinged to the main framework and have flexible connections so that they can be opened out for access to their parts. Fig. 21.12 shows two of the units so opened and Fig. 21.13 shows the top one of the two units of the superheterodyne opened up. For access to the lower unit the two can be lifted together.

An example of modern commercial practice is shown in Fig. 21.14, which is the circuit of the Pye B18T receiver. It is a transformerless set suitable for operation on either a.c. or d.c. mains of 230 V or more. It is a straight set with four s.f. stages on vision. The first two stages are common to vision and sound and have independent gain controls biasing both control and suppressor grids to keep the input capacitance and resistance constant. One control is labelled " Sensitivity " and the other is called " Contrast."

The intervalve couplings are a mixture of single-circuit and coupled-pair types, dust-iron cores being used for trimming. Cathode-circuit sound-channel interference rejectors are included in $V_3$ and $V_4$ and the former is used also as the sound pick-out circuit to feed the two-stage r.f. amplifier $V_{13}$ and $V_{14}$.

The vision-channel detector $V_5$ is one half of a duo-diode, the other half of which serves as a noise limiter $V_6$. The v.f. stage $V_7$ feeds the signal to the cathode of the c.r. tube and also to the sync separator $V_8$. The output of this valve is capacitance coupled to

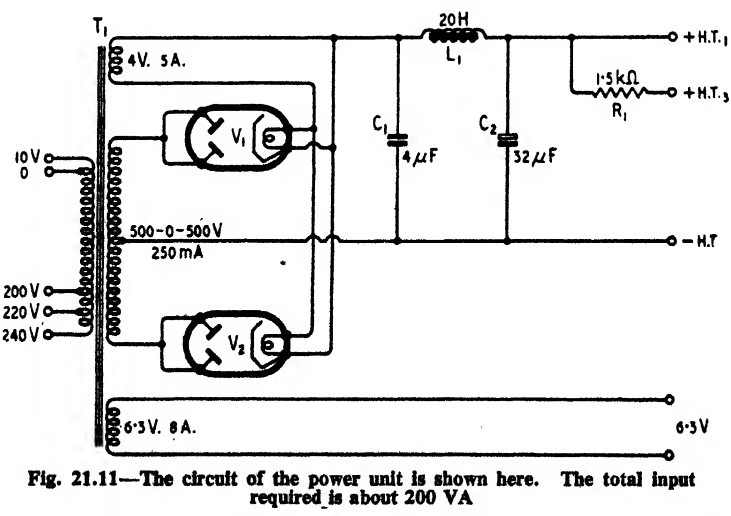



**Fig. 21.11—The circuit of the power unit is shown here. The total input required is about 200 VA**

the line blocking oscillator $V_{18}$ and directly coupled to the frame pulse separator $V_9$ and $V_{10}$.

The frame time base has two valves $V_{11}$ and $V_{12}$, a blocking oscillator and a pentode amplifier respectively. The linearity circuit is a little unusual. The charging capacitor is split into two in series, $C_1$ and $C_2$, the latter being coupled to the cathode circuit of $V_{12}$ by $R_1$. The voltage developed across the cathode RC network by the cathode current of the valve is of parabolic form and is fed back through $R_1$. This, in conjunction with valve curvature, effects linearization of the trace.

The line saw-tooth generator is also a blocking oscillator and feeds the pentode $V_{19}$. The e.h.t. supply is taken from the line fly-back, a valve rectifier $V_{20}$ being used. No reservoir capacitance is shown, since the c.r. tube used has an external metal coating to its bulb and the capacitance between this and the internal graphite coating of the anode serves for it. A damping diode $V_{21}$ is used and provides h.t. boost as well as giving current economy. The boosting voltage is negative with respect to the h.t. supply and so is applied to the cathode of $V_{19}$; the line marked " D " is the most negative point of the set.

A half-wave rectifier is used in the h.t. supply circuit and on d.c. supplies arrangements are made to cut it out of circuit and also to use part only of the smoothing choke. This is done to reduce the voltage drop as much as possible. All heaters are series connected and there is a tapped voltage-dropping resistor and a thermistor-type regulator.

In the sound channel a diode detector is used followed by a diode noise limiter $V_{16}$ and a pentode output stage $V_{17}$. The first of the

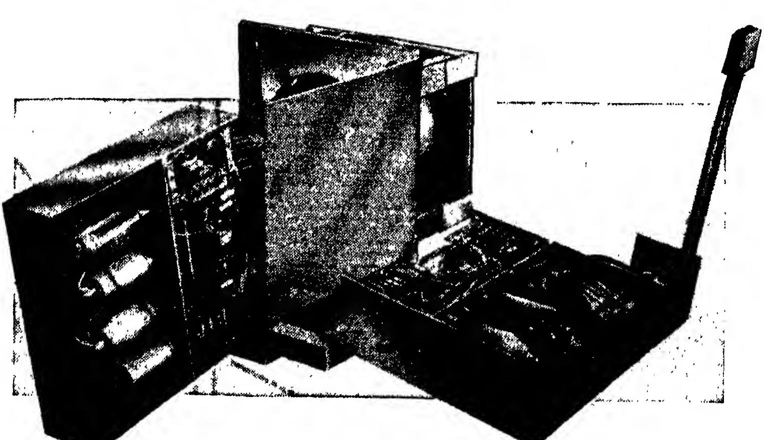


**Fig. 21.12—A rear view of the assembly showing the frame and line time-base units opened out for access to the interior. The only connection that need be broken is the plug connector to the " Sync Input " socket on the frame unit**

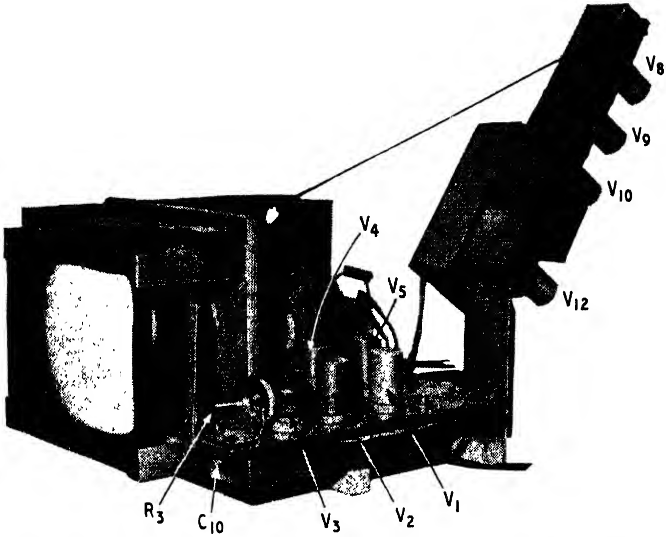



**Fig. 21.13—In this photograph the upper deck of the superheterodyne receiver is shown hinged up so that access to the underside can be obtained**

sound-channel s.f. stages $V_{13}$ has a.g.c. applied to it, mainly so that changes in the setting of the contrast control do not affect the sound volume.

A permanent magnet is used for focusing. Its effect is varied by a mechanical adjustment on the magnet itself and the focus control is thus a pre-set one.

CHAPTER 22

# Faults and their Remedies

As IN THE CASE of ordinary sound equipment, defects may occur in television apparatus and it is only more difficult to trace their source because they are still much less familiar. It is proposed, therefore, to describe some of the more common symptoms and the faults responsible.

If the picture is upside down it means merely that the two connections to the frame deflecting coil (electromagnetic tube) or plates (electrostatic tube) need reversing. Similarly, if the picture is reversed from left to right, the two leads to the line deflecting coils or plates must be reversed.

Somewhat more difficult to deal with is the case when the raster ceases to be rectangular. In estimating the rectilinearity of the raster due allowance must be made for the curvature of the end of the c.r. tube.

With electric deflection trapezium distortion may occur. The raster then tends to the form sketched in Fig. 22.1. This distortion indicates unbalanced time-base amplifiers.

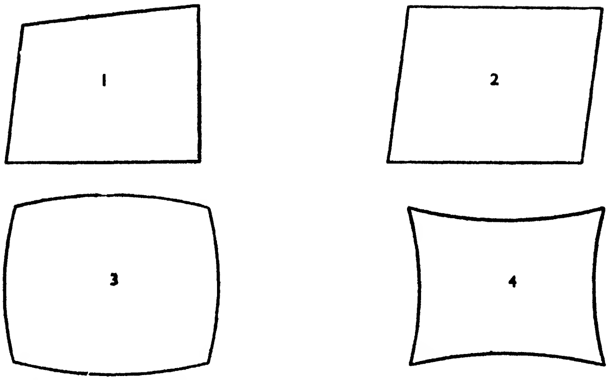
In general, it is only met with in as severe a form as shown when asymmetrical deflection is used. With the usual push-pull deflection, trapezium distortion is rarely severe and when it is detectable it is a sign that the amplifiers are badly out of balance.

With electromagnetic deflection it is possible to obtain a raster of the form shown exaggerated in Fig. 22.2. This indicates that the planes of the line and frame deflecting coils are not at right angles. Unless the whole assembly is a sealed unit, it is a simple matter to rotate one pair of coils with respect to the other until a truly rectangular raster is secured.

Another form of distortion which occasionally occurs is barrel or pin-cushion distortion, shown at Figs. 22.3 and 4 respectively. These are caused by non-uniformity of the deflecting fields. The remedy is more difficult and lies in redesigning the deflecting coils (see Chapter 6).

Although a rectangular raster may be obtained it is not necessarily perfect. It is common to find that the focusing is uneven. A sharp focus is usually obtained over the centre of the raster, but it deteriorates towards the edges and may become very poor indeed.

With electromagnetic deflection the trouble usually lies in the coil design and usually occurs to an appreciable degree only when the design has been carried out for maximum economy in deflecting power. Beyond a redesign of the coils there is little that can be done.




**Figs. 22.1–4—Trapezium distortion is shown at 1 and trouble due to a twisted deflecting system at 2. Barrel and pin-cushion distortion are shown at 3 and 4**

With an electrostatic tube the trouble is to some degree inherent. It can often be greatly reduced, however, by applying a suitable bias voltage to the deflecting plates as described in Chapter 3.

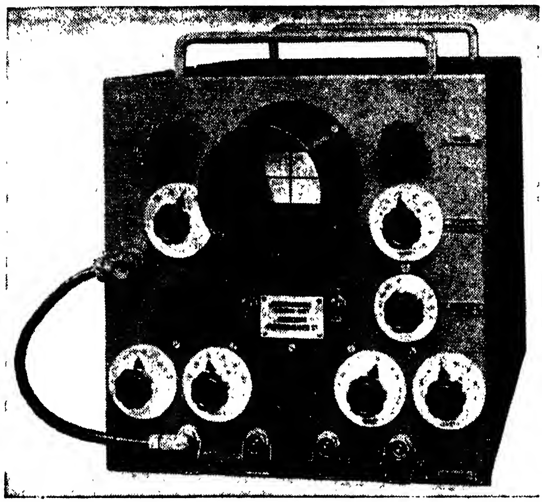
Uneven spacing of the lines is a sign of a non-linear frame deflecting voltage or current. Generally, it is evident as a closing up of the lines at the bottom of the picture. In order to remedy it the cause of the non-linearity must be found, and a good oscilloscope is a great help in this since the waveform can readily be checked at the various stages.

The requirements for good waveform have been discussed in detail in the chapters dealing with time-base generators and amplifiers, and it is consequently unnecessary to go into the matter here. It may be remarked, however, that non-linearity is more likely to occur as the result of the ageing of a valve in a magnetic than in an electric time base. In the former, the valves are usually run harder and there is less factor of safety for valve deterioration.

Non-linearity is particularly objectionable in the case of the line scan, where it generally shows up as a cramping of the right-hand side of the picture. Quite a small amount can be very noticeable on a rapid panning shot, because an object moving across the picture changes its size as it enters the non-linear portion.

Ripple from the mains supply causes very different effects according to its point of entry. In the vision receiver, or in the tube electrode supplies, it causes dark bands across the picture. In the time bases it causes a wavy edge to the raster.

As full-wave rectification is generally used for the vision receiver h.t. supply the main hum frequency is 100 c/s, and if the smoothing is insufficient two dark bands will appear across the picture. When the picture is locked they will be stationary, but are more easily seen with an unsynchronized picture and a blank raster, as they then move vertically across it.



**Marconi Instruments Electrostatic Viewing Unit**

As half-wave rectification is used in most cases for the tube h.t. supply, inadequate smoothing here generally leads to only a single dark band.

Ripple on the line time base makes the vertical edges of the picture wavy. Usually both edges are equally affected, but occasionally only one edge may have the wave on it. This is when the hum is reaching only one output valve of a push-pull time-base amplifier for electric deflection.

The stray field from a mains transformer can deflect the beam of the c.r. tube directly if it is sufficiently close. Serious deformation of the raster then occurs and can be remedied by moving the transformer farther away, using a transformer of smaller external field, or fitting a mu-metal shield to the tube. This last is expensive.

It should not be forgotten, too, that the leakage field from a loudspeaker will seriously distort the raster if the speaker is too near the tube. In a compact assembly, it is inadvisable to use a speaker with a magnet of high flux-density.

Perhaps the commonest fault and at the same time the most difficult to cure is faulty interlacing. In a bad case, the lines of successive frames are nearly superimposed so that the scanning lines appear very broad and with noticeable gaps between them. Owing to the apparent width of the lines, one may at first think that the focus is at fault, but examination will show that the picture detail along the lines is good.

If no *critical* setting of the frame-hold control will cause interlacing, the trouble is almost certainly caused by pulses from the line time base reaching the frame time-base generator. Almost complete freedom from interaction is necessary if perfect interlacing is to be secured.

meant a gradual tailing off of all objects on their right-hand sides, instead of the normal sharp boundary. In a severe case, the flare may be visible for an inch or so on the right of the object.

The author met with this in experimental equipment. The output of the h.t. unit was too great for the receiver and a dropping resistance was consequently inserted in series with it, but a by-pass capacitor was forgotten. The dropping resistance was thus common to anode and screen circuits of the v.f. stage.

At low frequencies the v.f. stage functioned largely as a triode with the h.t. dropping resistance as a coupling resistance. At higher frequencies stray capacitances came into play and also the various i.f. decoupling circuits exercised an effect, with the result that the gain fell off and the valve reverted gradually to its normal mode of operation as a pentode. The connection of an 8-$\mu$F capacitor on the receiver side of the h.t. supply completely cured the trouble. The capacitor should, of course, have been included in the first place.

Inadequate low-frequency response gives just the opposite effect on the picture. All outlines are sharply reproduced, but the main body of the tone value disappears. Thus, a black object will be reproduced with little difference of tone from the background, but its outlines will be clear. The effect is known as plastic and it is as though objects were in relief but of uniform shade. It can occur with a normal superheterodyne if the oscillator is incorrectly adjusted so that the intermediate frequency is too far removed from the i.f. pass-band.

More common is an effect often known as "black after white." Immediately to the right of a sharp transition from black to a lighter background there appears a white line following the contour of the black object. If the object is white, then the following line is black.

In severe cases, there are several such lines alternating in black and white and eventually shading off into the background. Usually, however, only a single line is noticeable with a second one detectable on close examination. The effect is very similar to that produced by reflections.

In a well-designed receiver the effect is nearly always due to misadjustment of the tuned circuits. It occurs chiefly when one circuit is inadequately damped and is mistuned to one side of resonance. In the case of an amplifier, therefore, in which circuits are overcoupled or mistuned to give a double-humped curve and another fairly sharply-tuned circuit is arranged to fill up the trough in the overall response curve, quite a small amount of mistuning will produce the effect.

It is common, too, to find that it can be produced at will by the adjustment of the oscillator frequency. With a good i.f. amplifier it does not occur when the oscillator is set to give an intermediate frequency equal to the mid-band frequency. As the oscillator is mistuned, however, so that the intermediate frequency comes to one side of the pass-band, the effect often appears over a small range of frequencies. It can also be produced by inadequate damping of a correction circuit in a v.f. stage.

The effect is not always undesirable. If present in the correct degree, which is small, it can give a very useful sharpening of the picture. If the bandwidth is rather narrow, the definition will not be very good; the finer points of the detail will be missing and the edges of objects will be somewhat blurred.

By adjusting the receiver so that there is a small amount of " black after white " the edges of objects appear much sharper and although there is no more real detail in the picture, the effect is greatly improved. It should be noted, however, that if the effect is obtained by mistuning the oscillator so that the intermediate frequency moves from the centre of the pass-band there will also be an increase of detail. Under this condition single-sideband reception is approached and the effective bandwidth is increased.

In general, only a very moderate amount of " black after white " should be permitted. It should not be noticeable at the normal viewing distance and should only be visible on close inspection of the picture.

CHAPTER 23

# Television Servicing

BY THE TERM SERVICING is meant the location and repair by amateur or professional of a defect in equipment which has previously been functioning correctly. It is an obvious extension of the meaning to include the tracing of faults in new and experimental apparatus.

There is this difference, however; in the former case it is known that the apparatus is of sound design and that the search is consequently for a fault which has developed after use, whereas in the latter case there are in addition the possibilities of faulty design, wrong connections, wrong values of components and unsatisfactory layout. The difficulties are consequently greater.

Television servicing does not differ essentially from ordinary wireless servicing, the chief requirements being a good knowledge of " how it works," and plenty of common sense. Elaborate test apparatus is unnecessary, but if there is a lot of work to be done it may be valuable in saving time.

The essential test apparatus comprises a multi-range a.c. and d.c. voltmeter and milliameter. On the a.c. range it should be possible to check accurately heater voltages from 2 V to 6·3 V and currents of about 1·5–3 A, since some c.r. tubes have current- and not voltage-rated heaters. Incidentally, the meter impedance on the current ranges should be known, so that allowance can be made for the voltage drop caused by its insertion.

These are the important a.c. ranges. Voltage ranges of 200–1,500 V are also desirable, but not essential, for checking the h.t. secondaries of mains transformers and the mains voltage itself.

The direct-current ranges should be wider and it should be possible to check anything from 1 V to some 1,200–1,500 V and current up to at least 150 mA and preferably 250 mA. The usual multi-range instrument has for its lowest range a full-scale reading of 3–5 mA. This is on the high side for checking the tube anode current, or the grid current of an oscillator, and for these purposes a separate meter with a full-scale reading of 1 mA is very useful. It is not an essential item, however, for these particular tests are not often needed.

If the multi-range meter includes ohmmeter ranges, so much the better. An ohmmeter is very handy for continuity testing and for checking resistances. Again it is not essential, for one of the voltmeter ranges can be used instead with an external battery.

If it can be afforded, a good electrostatic voltmeter is very handy for checking the high-voltage supply. An ordinary voltmeter is useless, since it takes far too much current. If an electrostatic voltmeter is not available, however, it is possible to check the high-voltage supply indirectly.

This is done by checking the resistance values in the bleeder across the output of the high-voltage unit, needless to say with the unit switched off. The earthy end is then opened and a milliameter is inserted in series with the bleeder. The product of the current in milliamperes and the bleeder resistance in megohms gives the voltage across the bleeder in kilovolts. Thus, 0·2 mA through 20 MΩ corresponds to 4 kV.

A modulated test oscillator for adjusting the s.f. and i.f. circuits is highly desirable, if not essential. It should have a fairly accurate frequency calibration and its output should be controllable over a wide range. It is particularly important that it should be capable of being detuned by accurately known amounts since many sets require the use of several slightly differing frequencies during alignment.

If much television servicing is to be done, a cathode-ray oscilloscope is almost indispensable. It should include a linear time base operating up to 5,000 c/s and down to 12·5 c/s and it should include an amplifier for the signal.

It is rather important that the amplifier should be first class and have a frequency response maintained up to at least 100,000 c/s. Many excellent oscilloscopes for ordinary work fail for television in this respect. A poor response in the oscilloscope amplifier seriously distorts the trace of the line sync pulses and even the line saw-tooth waveform.

If the oscilloscope is to be of any use for indicating the linearity of a frame-scan waveform it is essential that the amplifier should have a substantially perfect response down to 50 c/s. This means that the — 3-db response must come below 1 c/s.

If the amplifier is not up to the necessary standard, the oscilloscope will have to be used without it in many cases. It must be remembered, however, that it is quite easy to make up an external amplifier and this should be done when an otherwise good oscilloscope is available. The amplifier should be designed on the lines of a v.f. stage.

The time base in the oscilloscope should preferably be fed for synchronizing purposes from the output of the amplifier. If it is fed directly from the work signal, that is the receiver under test, the oscilloscope time base usually upsets the receiver synchronizing.

The oscilloscope is most valuable for checking the operation of the sync separating circuits and also for investigating the time bases. With electric deflection checking is easy, but with electromagnetic deflection it will usually be necessary to insert a resistance in series with the deflecting coils.

Here it is a saw-tooth current that is needed, but the oscilloscope requires a voltage input. If a resistance of 10–20 Ω is included in series with the deflecting coil and the oscilloscope connected across it a good trace will usually be obtained and the effect of the resistance on the time base will be small.

Some oscilloscopes are provided with deflecting coils so that currents can be investigated directly. In the author's experience, however, these coils are unsuitable for checking television time bases, for their introduction in series with the deflecting coils upsets the operation of the time base.

The oscilloscope is much more useful in television servicing than in ordinary work on broadcast receivers. In development work on time bases and sync separators it is indispensable.

Owing to the large number of valves used in television apparatus a good valve tester is highly desirable. Complete reliance cannot be placed upon it, however, in the case of valves used in certain hard-valve time-base oscillator circuits.

Valve characteristics of which the valve tester takes little or no account then play an important part, and it is necessary to rely upon a substitution test. With such circuits one cannot predict with any certainty whether or not a different valve of similar normal characteristics will function satisfactorily.

The grid current characteristic plays an important part in many time-base oscillators and this is not included among the published valve data in most cases. Two valves of very similar normal characteristics may have very different grid current characteristics.

Testing the cathode-ray tube offers a difficult problem unless a substitute is available. The anode current can be checked by including a meter in series with the anode lead, *taking great care over the insulation.* With a good modern magnetic tube at 5,000 V and a " white " raster, a current of the order of 100–150 $\mu$A is to be expected. It will, of course, vary with different tube designs and will be generally less with electrostatic tubes than with electromagnetic.

An abnormally high current accompanied by a dull picture is a probable indication that the tube is soft. A very low current, on the other hand, would tend to show a failure of the cathode emission. Normal current, but poor brilliancy, would indicate a failure of the fluorescent screen.

It is, however, necessary to know what the normal current should be before these tests can be anything but the roughest. This is the difficulty and in most cases at present the tube must be suspected only when everything else has failed.

Normal service procedure is naturally followed and differs very little from that adopted for ordinary apparatus. The symptoms usually enable the fault to be placed as in one section of the apparatus and then general testing in that section enables it to be located. Thus, a faulty raster or cramping of part of the picture indicates a defect in the time bases, whereas poor synchronizing is more probably caused by the sync separator or its associated circuits.

In this connection, it should be noted that poor synchronizing is sometimes caused by an incipient breakdown of insulation. One case which the author investigated was found to be due to the line output transformer. The chief symptom was very poor frame synchronizing, but on listening carefully small sounds of sparking could be heard from the line transformer.

On every line fly-back the e.m.f. across the transformer primary broke down the insulation and a small spark occurred. Eventually, of course, the transformer broke down completely, but towards the end of each frame the sparking led to a pulse being developed on the frame oscillator which triggered it.

In connection with television servicing it is necessary to emphasize the importance of care, because high voltages are used and bad shocks can be obtained. The high-voltage equipment is probably less dangerous to life than the secondary of an ordinary mains transformer. It is current that kills and the maximum current output of the high-voltage equipment is severely limited.

A current of some 15–25 mA is sufficient to kill and as the body resistance can be only a few thousand ohms, quite a low voltage will suffice to produce a fatal shock. An h.t. battery of only 100 V and with low internal resistance is capable of doing so under favourable conditions.

When the skin is dry the contact resistance is fairly high. On lightly touching a high-voltage point the current is consequently fairly small and the involuntary muscular contraction brings the hand away at once.

While this muscular contraction is a great safeguard in most cases, it can be a great danger in others. As long as one only touches a high-voltage point, the hand jerks away, but if one grabs it the muscular contraction makes it impossible to let go.

The most important safety rule is to keep one hand in the pocket while working and to see that the floor of the room is *dry*. Apart from concrete, most floor materials are insulators, but none are if damp.

Even if one does get a shock the current is then unlikely to pass with any serious intensity near a vital part. A high voltage applied across one hand, while unpleasant, is much less dangerous than if it is connected between both hands.

Television apparatus is actually no more dangerous than an ordinary receiver or amplifier, provided that the high-voltage equipment is properly protected. The average broadcast set has over 1,000 V peak across the transformer secondary.

## ABBREVIATIONS

| | | |
|---|---|---|
| a.c. | = alternating current. | |
| a.f. | = audio-frequency | = Any frequency within the audible range of some 20–20,000 c/s. |
| a.g.c. | = automatic gain control. | |
| (a.v.c. | = automatic volume control is often less correctly used for a.g.c.) | |
| b.f. | = beat-frequency | = The sum or difference of two frequencies which are so mixed that beats are produced. |
| c.r. | = cathode-ray. | |
| d.c. | = direct current. | |
| h.f. | = high-frequency | = A relative term; at one time used for r.f. |
| i.f. | = intermediate-frequency | = The frequency at which amplification is chiefly carried out in the superheterodyne ; the beat frequency. |
| l.f. | = low-frequency | = A relative term, at one time used for a.f.; except when used comparatively, now confined to frequencies of 50 c/s or less. |
| p | = peak | = $\sqrt{2}$ times the r.m.s. value of a sine wave alternating current or voltage. |
| p–p | = peak-to-peak | = $2\sqrt{2}$ times the r.m.s. value of a sine wave; the difference between the extremes of voltage or current excursion of a non-sinusoidal waveform. |
| r.f. | = radio-frequency | = Any frequency higher than about 30 kc/s; often used for s.f. |

| | | |
|---|---|---|
| r.m.s. | = root-mean-square | = The effective value of alternating current. |
| s.f. | = signal-frequency | = The frequency of the incoming signal. |
| sync | = synchronizing. | |
| t.b. | = time base | = (In television) apparatus for generating a saw-tooth voltage or current waveform. |
| v.f. | = vision-frequency (or video-frequency) | = Frequencies produced by the scanning process and corresponding to the picture detail; the modulation frequencies of the vision signal. |
| z.f. | = zero-frequency | = When applied to voltage or current, an unchanging voltage or direct current; rarely used. |

## UNITS AND THEIR ABBREVIATIONS

*Capacitance* — Symbol $C$

F = farad.
$\mu$F (or mfd.) = micro-farad = $10^{-6}$ F.
$\mu\mu$F (or mmfd.) = micro-micro-farad = $10^{-12}$ F.
pF = picofarad = $\mu\mu$F.

*Frequency* — Symbol $f$

c/s = cycles per second.
kc/s = kilocycles per second = $10^3$ c/s.
Mc/s = megacycles per second = $10^6$ c/s.

*Inductance* — Symbol $L$

H = henry.
mH = millihenry = $10^{-3}$ H.
$\mu$H = microhenry = $10^{-6}$ H.

*Resistance* — Symbol $R$

$\Omega$ = ohm
k$\Omega$ = kilohm = $10^3$ $\Omega$
M$\Omega$ = megohm = $10^6$ $\Omega$

*Time* — Symbol $t$

sec = second
msec = millisecond = $10^{-3}$ sec
$\mu$sec = microsecond = $10^{-6}$ sec

## DEFINITIONS

*Flyback* = The rapid return of a scanning spot after the completion of a vertical or horizontal scan.

| | |
|---|---|
| *Frame frequency* | = The number of vertical scans per second. |
| *Interlacing* | = A method of scanning whereby (in the B.B.C. transmissions) alternate lines are scanned in the first, third, fifth, etc., frames and the intervening lines in the second, fourth, sixth, etc., frames. |
| *Line frequency* | = The number of scanning lines traced per second. |
| *Picture frequency* | = The number of complete pictures per second (one-half the frame frequency with the present 2 : 1 interlace). |
| *Raster* | = The illuminated rectangle built up on the screen of the c.r. tube by the scanning process. |
| *Saw-tooth waveform* | = A current or voltage waveform such that when current or voltage is plotted against time the resulting curve resembles the tooth of a saw; for television the waveform is usually linear and a saw-tooth wave can then be defined as a current or voltage which grows to a certain maximum linearly with time and then decreases to its minimum in a very much smaller amount of time. |
| *Scanning* | = The process of exploring a picture point by point in an orderly method, and hence, the similar process of building up the picture at the receiver. |
| *Screen (of c.r. tube)* | = Fluorescent screen; the end of the c.r. tube coated with fluorescent material. |
| *Video frequency* | = Vision frequency; see v.f. |

NOTE.—In the U.S.A. the term " field " is used instead of " frame " and " frame " is used instead of " picture." References to frame frequency in American literature are thus to picture frequency in British parlance.

## APPENDIX 2

# INTEGRATORS

THE NORMAL TELEVISION usage of an integrator is dealt with in Chapter 16. It should be realized, however, that an integrator is nothing but a circuit having a frequency response falling at high frequencies. The ordinary *RC* intervalve coupling has in the coupling resistance and the shunt stray capacitance the form of an integrator, just as in the coupling capacitance and grid leak it has the form of a differentiator.

The basic circuit is shown in Fig. A.2.1 and it will be seen that like the differentiator it consists of capacitance and resistance in series. The output, however, is taken across the capacitance instead of the resistance. The current follows the same law and at the instant of applying a step wave it is $E/R$ when $C$ is initially uncharged. The output is then zero and as time progresses it rises exponentially and eventually equals the input.

At any instant after the application of a step wave

$$E_o = e_{in} (1 - \epsilon^{-t/T}) \qquad \text{A.2.1}$$

where $T = CR$.

If the capacitance has an initial voltage $e_1$ then this is modified to

$$E_o = e_1 + (e_{in} - e_1) (1 - \epsilon^{-t/T}) \qquad \text{A.2.2}$$

With a pulse input, the initial response can be calculated from these equations up to the time of the end of the pulse. Then the voltage decays in accordance with the relation

$$E_o = e_2 \, \epsilon^{-t/T} \qquad \text{A.2.3}$$

where $e_2$ is the voltage across $C$ at the end of the pulse.

The output from a train of pulses is calculable by the successive application of Equs. (A.2.2) and (3) for each change of input voltage.

The general form of the output voltage (b) for a pulse input (a) is shown in Fig. A.2.2 and the result with a train of pulses corresponding to the television sync signals is illustrated in Fig. 16.19.

The purpose of the integrator is to produce an output having an amplitude which is dependent on, and ideally proportional to, the duration of a pulse, so that long and short pulses can be separated

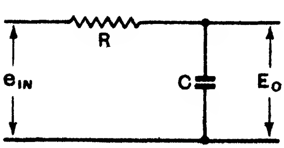
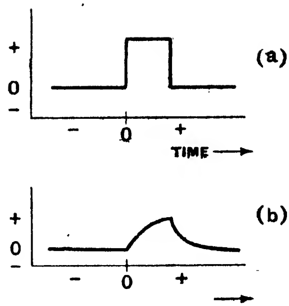



**Fig. A.2.1—Basic integrator circuit**

**Fig. A.2.2—With an input pulse (a) the output has the form (b)**




on an amplitude basis. For this result to be achieved it is necessary that the time-constant be such that $E_o$ is less than $e_{in}$ at the end of the longest pulse which is applied. It is usually desirable for $E_o$ to be considerably less than $e_{in}$ at this time.

The exact values are by no means critical and it is usually satisfactory to make $T$ about five times the duration of the longest pulse.

The symbol $\epsilon$ used above and elsewhere in this book has the numerical value 2·718. Tables of powers of $\epsilon$ are commonly used for evaluating expressions containing them.

## DIFFERENTIATORS

THE MAIN USE OF differentiating circuits in television receivers lies in sync separation and it is treated in Chapter 16. As the name implies the output voltage is ideally proportional to the time rate of change of the input. In practice, the circuits normally used are far from this ideal.

The basic differentiating circuit is shown in Fig. A.3.1 and has the form of an *RC* intervalve coupling. It differs from such a coupling only in the value assigned to the product $CR = T$.

When the input voltage is a step function of the form shown in Fig. A.3.2, and the capacitance is uncharged, the initial current at the moment of the step ($t = 0$) is limited only by the circuit resistance. This is $i = e_{in}/R$. The output voltage, however, is $E_o = iR$, hence $E_o = e_{in}$. As the capacitance is uncharged, the full input voltage appears as the output.

However, as time goes on the capacitance charges and a voltage appears across it. The output voltage is then the input less this capacitance voltage, and the initial output voltage is not maintained. The output at any instant is given by

$$E_o = e_{in}\,\epsilon^{-t/T} \qquad \text{A.3.1}$$

and has the form sketched in Fig. A.3.3.

When the input is a pulse it can be regarded as a positive step followed by a negative, or vice versa, the interval between the steps equalling the pulse duration. The capacitance may or may not be charged when this second step is applied, depending on the relation between the time-constant $T$ and the pulse duration.

In all cases the full input voltage change occurs at the output at the instant of the change of input. The voltage effective on the subsequent decay, however, is the maximum voltage on the output and not the change. Thus, suppose there is a 10-V input pulse. The output jumps to + 10 V and then decays exponentially in accordance with Equ. (A.3.1) until the end of the pulse is reached. Let the output at this instant be 6 V.

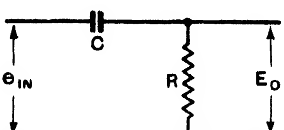



**Fig. A.3.1—Basic differentiator circuit**

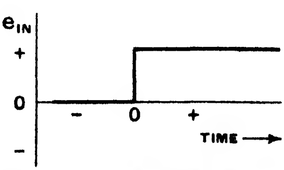
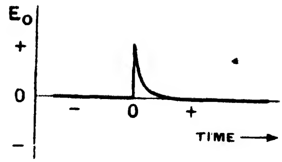



**Fig. A.3.2—Positive-going step wave**

**Fig. A.3.3—Output voltage from a differentiator with a positive-going step-wave input**




The change of input is in the negative direction and is 10 V. The output, therefore, changes to 6 – 10 = – 4 V instantly. This then starts to decay exponentially and the proper value to insert in Equ. (A.3.1) for $e_{in}$ is – 4 V in order to calculate the output.

When a pulse (a), Fig. A.3.4, is applied the output wave takes the form sketched in (b), (c) and (d) according to whether $T$ is larger, equal to, or less than the time of duration of the pulse. The last condition is the correct one for a differentiator, but the others are also very important.

Case (b) arises when a coupling must pass the wave with as little distortion as possible. If a pulse has a duration $t_1$, and a drop in output of 2 per cent at its end is the maximum permissible, then in Equ. (A.3.1), $E_o/e_{in} = 0{\cdot}98\ \epsilon^{-t_1/T} \approx 1 - t_1/T$. Consequently $T = 50\ t_1$.

This is a useful rule to remember. *For not more than 2 per cent distortion in the coupling, the time-constant must not be less than 50 times the pulse duration.* It applies also to other periodic waves having a definite duration, such as saw-tooth waves.

As an example, in a frame time base the saw-tooth duration is 18·6 msec, so that the time-constant of an intervalve coupling must be 930 msec = 0·93 sec = 0·93 $\mu$F–M$\Omega$. For $R$ = 2 M$\Omega$, $C$ = 0·5 $\mu$F or more.

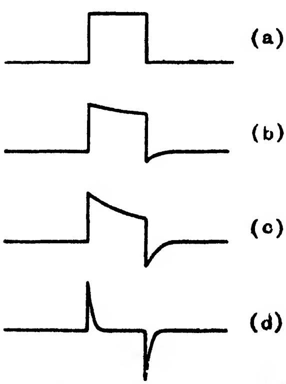



**Fig. A.3.4—An ideal pulse is shown at (a) and the output wave at (b), (c) and (d) for a time-constant much larger than, equal to, and smaller than the pulse duration**

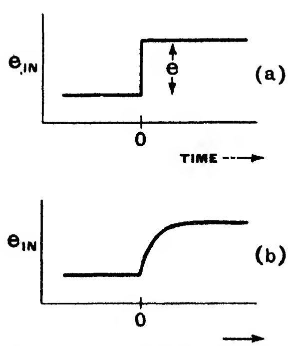



**Fig. A.3.5—An ideal pulse has a leading edge like (a) but in practice the edge is distorted to a form more like (b)**

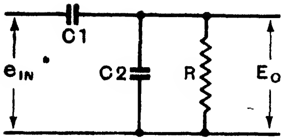



Fig. A.3.6—The capacitance $C_2$ represents the input capacitance of the following valve

On the other hand for the line scan the duration is 82·77 $\mu$sec so $T = 4{,}140\ \mu\text{sec} = 4{,}140$ pF–MΩ, and with $R = 2$ MΩ, $C$ need be only 2,070 pF.

The second case (c) of equal time-constant and pulse duration is less commonly used and is explained in Chapter 16 for one application.

The third case (d) represents the true differentiating action. The usual requirement here is that at the end of the pulse the voltage

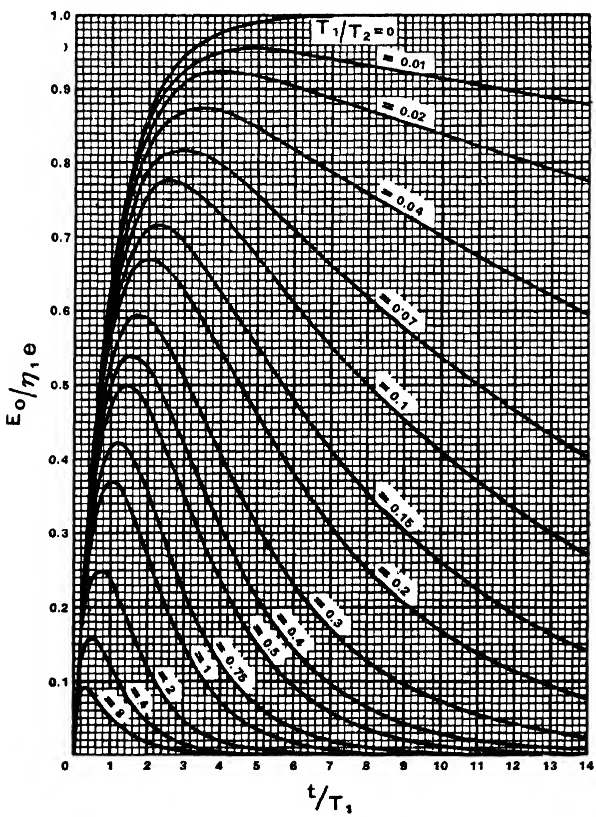



Fig. A.3.7—This family of curves shows the output wave as a function of time for a series of values of $T_1/T_2$. The input wave is the outer curve for $T_1/T_2=0$

should have fallen to, say, 2 per cent of its maximum. Therefore $\epsilon^{-t_1/T} = 0{\cdot}02$, or $t_1/T = 4$ (approx), giving $T \doteq t_1/4$.

For a differentiator, therefore, the rule is that *the time-constant should not exceed one-quarter of the pulse duration.*

It has been assumed in the foregoing that the input pulse is perfect. In practice it has been distorted somewhat by its passage through the early circuits and in any case a perfect pulse cannot be generated. The edges of the pulse are not truly vertical and a finite time is involved in the change from one input voltage to another. Perfect and practical pulses have leading edges of the forms sketched in Fig. A.3.5 at (a) and (b) respectively. In the latter the shape of the edge is exaggerated for clarity.

The main effect of this is to reduce the output voltage level and it is important because of this. In many cases a specific output voltage is needed from a differentiator and it is then necessary to take the time of rise of the pulse into account.

It is not possible to cover all the forms which the edges of the pulse can assume but it is sufficient for most purposes to take the shape as exponential. In what follows it is assumed that the edges of the pulse have the shape which would be taken by an ideal pulse which has been passed through a single *RC* circuit of time-constant $T_1$. It is convenient, but not very accurate, to relate this to the frequency response of the preceding equipment by taking

$$T_1 = 1/2\pi f \qquad \text{A.3.2}$$

where $f$ is the frequency for which the response is $-$ 3 db.

$$\text{Then } E_o = e_{in}\,\eta_1\,\frac{\epsilon^{-t/T_1} - \epsilon^{-t/T_2}}{T_1/T_2 - 1} \qquad \text{A.3.3}$$

provided $T_1 \neq T_2$,

where $T_2 = R(C_1 + C_2)$

$\eta_1 = C_1/(C_1 + C_2)$

The circuit is shown in Fig. A.3.6 and differs only from that of Fig. A.3.1 by the inclusion of $C_2$ to represent the input capacitance of the following valve. The performance* is represented by the curves of Fig. A.3.7 while Figs. A.3.8 and 9 enable circuit values to be selected.

The usual television differentiator has $C_1 = 50$ pF and $R = 50$ kΩ and $C_2$ may well be 10 pF, so that $T_2 = 3$ μsec. For a frequency response of 2·5 Mc/s at $-$ 3 db, $T_1 = 0{\cdot}0636$ μsec and $T_1/T_2 = 0{\cdot}0212$. For these values of capacitance $\eta_1 = 0{\cdot}83$.

From Fig. A.3.8, $\eta_2$, the ratio of the maximum output to the input, is 0·92 and $t_1/T_2 = 3{\cdot}9$. The peak output is thus $E_o = \eta_1\eta_2 e_{in} = 0{\cdot}77\ e_{in}$ and this peak occurs at $t_1 = 3{\cdot}9 \times 0{\cdot}0636 = 0{\cdot}25$ μsec after the onset of the pulse. The overall efficiency of a typical differentiator is thus about 75 per cent.

Occasionally inductance is used instead of capacitance in a differentiator and the output is then taken across the inductance.

* "Square-Wave Differentiating Circuit Analysis," by G. P. Ohman, *Electronics*, August 1945.

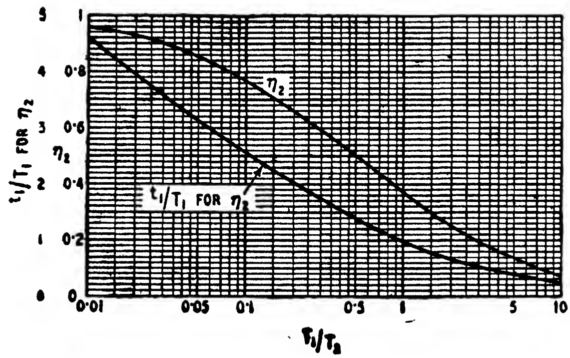



Fig. A.3.8—These curves show the relation between $T_1/T_2$ and $t_1$, the interval after the onset of the pulse at which the maximum efficiency $\eta_2$ is secured

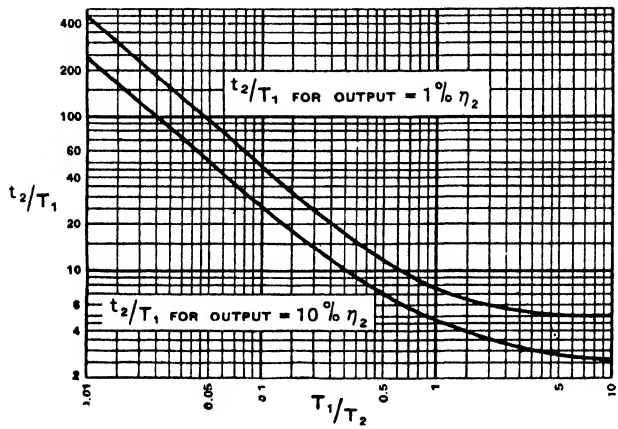



Fig. A.3.9—The time $t_2$ at which the output has fallen to 1 per cent or 10 per cent of the maximum is shown here

This has the advantage that a double-wound coil can be used and the output taken from the secondary; as a result an output of the same or the opposite polarity to the input can be secured merely by reversing the connections to the secondary. In addition, the output can be obtained without any point on the secondary being earthed and this sometimes simplifies the feed to certain time bases.

The circuit is shown in Fig. A.3.10 and to a first approximation Equ. (A.3.1) can be used if $T = L_p/R$ and $E_o$ is multiplied by $n$ = ratio secondary/primary turns. According to the rule given earlier, for differentiating line sync pulses $T$ should be about 2·5 $\mu$sec; therefore, $L_p/R = 2{\cdot}5$ in mH and k$\Omega$. With $R = 10$ k$\Omega$, $L_p$ = 25 mH.

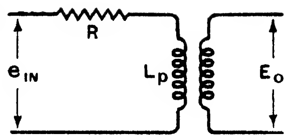



**Fig. A.3.10—Differentiator with inductance instead of capacitance**

The simple circuit of Fig. A.3.10, however, is not really representative of the operation for it is impossible to construct a transformer without self-capacitance and this can greatly modify the performance. For the circuit to be non-oscillatory it is necessary to have $C \leqslant L/4R^2$. In the example this means $C \leqslant 2{\cdot}5 \times 10^{-2}/4 \times 10^8 = 6{\cdot}25 \times 10^{-11}$ F $= 62{\cdot}5$ pF. This is possible and the circuit can be used, but it needs much more careful design than the usual capacitance system.

# D.C. RESTORATION

A SERIES OF RECURRENT pulses of duration $t_1$ and interval $t_2$ is applied to an *RC* coupling which removes the d.c. component. The input waveform is like (a) of Fig. A.4.1 and the output has the general form (b). The tops and bottoms of the pulses slope by an amount depending on the time-constant of the coupling and the pulse train has shifted as a whole with reference to the zero level by an amount dependent on $t_1$ and $t_2$. It has shifted because the coupling does not pass the d.c. component of the input signal.

Restoration of this component can be effected by connecting a diode in shunt with the resistance as shown in Fig. A.4.2. The diode conducts only when its anode is positive with respect to its cathode—ignoring the initial electron velocity which makes the cut-off point around $-$ 1 V.

Referring to Fig. A.4.1, the diode is conductive during the periods $t_1$ and non-conductive during the periods $t_2$. When it is conductive it has a resistance $r_a$ which for simplicity will be taken as very small compared with $R$.

During the positive-going pulses, $C$ charges through $r_a$, the effective voltage being $V = E - v_2$. The current $i_1 = \dfrac{V}{r_a}\epsilon^{-t/T_1}$ at any time after the start of the pulse. The total charge gained by $C$ during one pulse of duration $t_1$ is $q_1 = \dfrac{V}{r_a}\displaystyle\int_0^{t_1} \epsilon^{-t/T_1}dt = VC\,(1 - \epsilon^{-t_1/T_1})$ where $T_1 = Cr_a$.

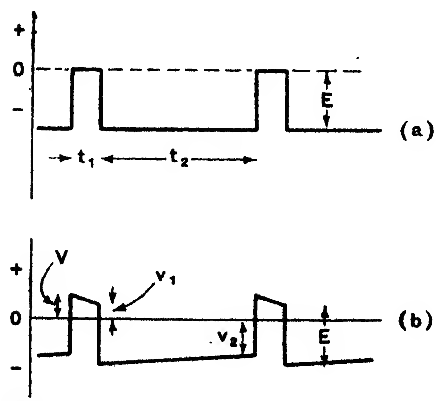



**Fig. A.4.1—A typical series of pulses is shown at (a) and in greatly exaggerated form after d.c. restoration at (b)**

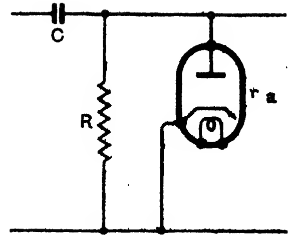



**Fig. A.4.2—Typical d.c. restoring circuit**

During the pulse intervals $t_2$ the diode is non-conductive and the time-constant is $T_2 = CR$. The voltage effective at the start of an interval is $v_1 - E$ and the current is $i_2 = \frac{v_1 - E}{R} \epsilon^{-t/T_2}$. The total charge gained by $C$ during the interval $t_2$ is

$$q_2 = \frac{v_1 - E}{R} \int_0^{t_2} \epsilon^{-t/T_2} dt = (v_1 - E)(1 - \epsilon^{-t_2/T_2})$$

Now $v_1$ is the voltage existing at the end of the pulse $t_1$ and is consequently $V \epsilon^{-t_1/T_1}$. Therefore,

$$q_2 = C\,(V\epsilon^{-t_1/T_1} - E)(1 - \epsilon^{-t_2/T_2}).$$

As $v < E$, $q_2$ is negative, which means that $C$ loses charge during the interval $t_2$.

In the equilibrium condition the charge gained during $t_1$ must equal the charge lost during $t_2$, hence,

$$q_1 = -q_2 = VC\,(1 - \epsilon^{-t_1/T_1}) = C\,(E - V\epsilon^{-t_1/T_1})\,(1 - \epsilon^{-t_2/T_2})$$

$$\text{or } \frac{E}{V} = \frac{1 - \epsilon^{-t_1/T_1}\,\epsilon^{-t_2/T_2}}{1 - \epsilon^{-t_2/T_2}}$$

If $t_1 \ll T_1$ and $t_2 \ll T_2$, then

$$\frac{E}{V} \approx \frac{t_1/T_1 + t_2/T_2}{t_2/T_2} = 1 + \frac{t_1}{t_2} \cdot \frac{R}{r_a}$$

On the television signal $t_1 = 10$ $\mu$sec and $t_2 = 90$ $\mu$sec, approximately, except during the frame pulses. Table A.4.1 shows the value of $E/V$ as a function of $R/r_a$, and in columns (3) and (4) the

**D.C. RESTORER PERFORMANCE**

**TABLE A.4.1**

| (1) $\frac{R}{r_a}$ | (2) $\frac{E}{V}$ | (3) *V Black Lines (volts)* | (4) *V White Lines (volts)* | (5) *Black Level (from zero) on Black Lines (volts)* | (6) *Black Level (from zero) on White Lines (volts)* |
|---|---|---|---|---|---|
| 10 | 2·1 | 3·75 | 12·5 | −5·25 | — |
| 20 | 3·2 | 2·82 | 9·4 | −6·18 | — |
| 50 | 6·5 | 1·38 | 4·6 | −7·62 | −4·4 |
| 100 | 12·1 | 0·745 | 2·48 | −8·26 | −6·52 |
| 200 | 23·2 | 0·388 | 1·29 | −8·61 | −7·71 |
| 500 | 56·5 | 0·16 | 0·54 | −8·84 | −8·47 |
| 1,000 | 112·1 • | 0·08 | 0·27 | −8·92 | −8·73 |

value of $V$ when $E$ is 9 V and 30 V respectively, corresponding to black and white lines for a 30-V maximum signal. Ideally, $V$ would be zero for both conditions. Columns (5) and (6) give the amplitude of sync pulse between zero and black level and these are the important figures because they represent the signal level upon which a sync separator must function.

When $R/r_a = 1{,}000$, instead of the black level being at $-9$ V at all times it is $-8{\cdot}92$ V on black lines and $-8{\cdot}73$ V on white. This is a negligible difference and represents the usual case of d.c. restoration when $R = 1$ M$\Omega$ and $r_a = 1$ k$\Omega$. The use of low values of $R/r_a$ is definitely bad, and a ratio of 100 : 1 is probably the smallest tolerable.

A normal value for $C$ is $0{\cdot}1$ $\mu$F. This makes $Cr_a = 100$ $\mu$F–$\Omega$ and $CR = 100{,}000$ $\mu$F–$\Omega$. These figures are both large compared with 10 $\mu$sec and 90 $\mu$sec respectively, so that the conditions for the validity of the approximation used in the analysis are met.

Electrically the main requirement is that $CR$ should be large enough in relation to $t_2$ for the drop of voltage during $t_2$ to be very small. If it is not the picture will tend to shade off darker to the right. This means that $v_2$ should be nearly equal to $E - v_1$.

$$\text{Now } v_2 = (E - v_1)\,\epsilon^{-t_1/T_1}$$

$$\text{and when } t_2 \ll T_2$$

$$\frac{v_2}{E - v_1} \approx 1 - \frac{t_1}{T_2}$$

For a 2 per cent loss during a line, $v_2/(E - v_1) = 0{\cdot}98$ and $t_2/T_2 = 0{\cdot}02$, making $T_2 = 4{,}500$ $\mu$F – $\Omega$. With $R = 1$ M$\Omega$, $C$ can be as low as $0{\cdot}0045$ $\mu$F. The usual value of $0{\cdot}1$ $\mu$F thus offers a large factor of safety.

## ELECTROMAGNETIC DEFLECTION

### Transformer Coupling

The current in the deflector coil during the scan period $\tau_1$ is assumed to have the ideal form

$$i_L = I_L\,(t/\tau_1 - 1/2) \qquad \text{A.5.1}$$

$0 \leqslant t \leqslant \tau_1$.

It is also assumed that the equivalent circuit of Fig. A.5.1 applies with sufficient accuracy for practical purposes. Then

$$e_L = I_L\left[\frac{L_L}{\tau_1} + R_L\left(\frac{t}{\tau_1} - \frac{1}{2}\right)\right] \qquad \text{A.5.2}$$

$$e_S = I_L\left[\frac{L_L + L_S(1-k^2)}{\tau_1} + (R_L + r_S)\left(\frac{t}{\tau_1} - \frac{1}{2}\right)\right] \qquad \text{A.5.3}$$

$$i_S = c - \frac{I_L}{2} + I_L\frac{t}{\tau_1}\left[\frac{L_L + L_S}{k^2 L_S} - \frac{\tau_1(R_L + r_S)}{2k^2 L_S}\right] + I_L\frac{t^2}{\tau_1^2}\cdot\frac{\tau_1(R_L + r_S)}{2k^2 L_S} \qquad \text{A.5.4}$$

where $c$ is a constant of integration the value of which it is not necessary to determine. The second and third terms represent linear and square-law currents. The latter must be supplied to the circuit by the valve if a linear coil current is to be obtained.

As a practical alternative the circuit values may be chosen so that its magnitude is negligibly small. If the square-law current is to be a fraction $a$ of the linear current, the necessary condition is

$$L_S = \frac{\tau_1(R_L + r_S)(1 + 1/a)}{2} - L_L \qquad \text{A.5.5.}$$

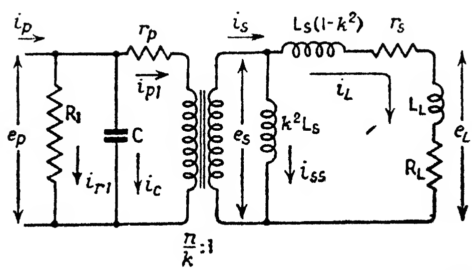



**Fig. A.5.1—Equivalent circuit of transformer-coupled stage feeding a deflector coil**

As this leads to an inconveniently high value of $L_S$, this relation is not often adopted and the necessary square-law current is supplied to the circuit.

### Line Scan

On the line scan it is usually permissible to ignore the effect of the series resistances $r_p$, $r_s$ and $R_L$ without serious error. The effect of $R$ and $C$ is also very small except on fly-back.

Using capital letters for peak-to-peak values, the deflector-coil volt-amperes are

$$E_L I_L = I_L^2 L_L/\tau_1 \qquad \text{A.5.6}$$

and the primary volt-amperes are

$$E_p I_p = I_L^2 \frac{L_L + L_S}{k^2 L_S} \cdot \frac{L_L + L_S(1 - k^2)}{\tau_1} \qquad \text{A.5.7}$$

The transformer efficiency* is, therefore,

$$\eta_T = \frac{E_L I_L}{E_p I_p} = \frac{k^2 L_L L_S}{(L_L + L_S)\{L_L + L_S(1 - k^2)\}} \qquad \text{A.5.8}$$

This has an optimum value

$$\eta_{T\,(\mathrm{opt})} = \frac{k^2}{(1 + \sqrt{1 - k^2})^2} \qquad \text{A.5.9}$$

when $\quad L_S = L_L/\sqrt{1 - k^2} \qquad$ A.5.10

With this relation

$$E_p = I_L \frac{L_L}{\tau_1} \cdot \frac{n}{\sqrt{\eta_{T(\mathrm{opt})}}} \qquad \text{A.5.11}$$

$$I_p = I_L/n\sqrt{\eta_{T(\mathrm{opt})}} \qquad \text{A.5.12}$$

### Line Fly-back

On fly-back, it is simplest to use the equivalent circuit of Fig. A.5.2 and when equation (A.5.10) is met, $L = n^2 L_L$. If the peak primary current at the start of fly-back is $i_o$ and $C$ is charged so that terminal (2) is $E_p$ volts below terminal (1) the instantaneous voltage and current are:—

$$e = [-E_p \cos \omega t + (i_o/\omega C + E_p/2\omega CR) \sin \omega t]\, \epsilon^{-\alpha t} \qquad \text{A.5.13}$$

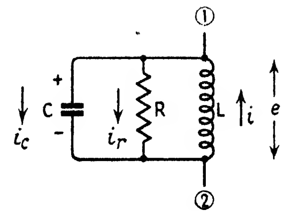
$$i = [i_o \cos \omega t + (i_o/2\omega CR + E_p/\omega L) \sin \omega t]\, \epsilon^{-\alpha t} \qquad \text{A.5.14}$$

where $\alpha = 1/2CR$; $\omega = \sqrt{1 - L/4CR^2}\,/\,\sqrt{LC}$

Exact solutions of these equations are difficult. However, most interest is attached to them in the regions of $\omega t = \pi/2$ for (A.5.13), and $\omega t = \pi$ for (A.5.14) and the cosine and sine terms respectively are then very small.

* This is field efficiency and no idea of power dissipation in the transformer is involved. By it is meant the ratio of the energy stored in the magnetic field of the deflector coil to the sum of the energy stored in the fields of the deflector coil and the transformer.

**Fig. A.5.2—Simplified circuit valid on line fly-back**




It is sufficiently accurate to write

$$e = \frac{i_o}{\omega C} \sin \omega t \, \epsilon^{-\alpha t} \qquad \text{A.5.15}$$

$$i = i_o \cos \omega t \, \epsilon^{-\alpha t} \qquad \text{A.5.16}$$

The first negative current maximum occurs when $\omega t = \pi$ and then

$$i = -i_1 = -i_o \, \epsilon^{-\pi\alpha/\omega}$$

If $x$ = fractional overshoot $= i_1/i_o = \epsilon^{-\pi\alpha/\omega}$

$$I_p = i_o (1 + x) \qquad \text{A.5.17}$$

(In reality, $I_p$ is slightly less than this, for the proper point to start the following scan is not at the negative maximum, but slightly later where the rate of change of current equals the scan rate of change of current. The point is difficult to calculate and it is convenient and not very inaccurate to take the negative current maximum instead.)

The voltage is a maximum when $\omega t = \pi/2$ and then

$$e = E_M = \frac{i_o}{\omega C} \epsilon^{-\pi\alpha/2\omega} \qquad \text{A.5.18}$$

From (11), (12), (17) and $L = n^2 L_L$ this can be put in the form

$$\frac{E_M}{E_p} = \frac{\tau_1}{\omega LC (1 + x)} \epsilon^{-\pi\alpha/2\omega} \qquad \text{A.5.19}$$

Since the end of the fly-back of duration $\tau_2$ is taken as occurring when $\omega t = \pi$, $\omega = \pi/\tau_2$. From this and the alternative value of $\omega$ given after equation (A.5.14)

$$\frac{E_M}{E_p} = \pi \frac{\tau_1}{\tau_2} \cdot \frac{\epsilon^{-\pi\alpha/2\omega}}{(1 + x)(1 - L/4CR^2)} \qquad \text{A.5.20}$$

From the definition of $x$

$$\frac{\alpha}{\omega} = \frac{\log_e 1/x}{\pi} = \frac{y}{\pi} = \sqrt{\frac{L/4CR^2}{1 - L/4CR^2}}$$

whence

$$\frac{E_M}{E_p} = \pi \frac{\tau_1}{\tau_2} \cdot \frac{\sqrt{x}}{1 + x}\left(1 + \frac{y^2}{\pi^2}\right) \qquad \text{A.5.21}$$

$$L = \frac{\tau_2^2}{\pi^2 C} \cdot \frac{1}{1 + y^2/\pi^2} \qquad \text{A.5.22}$$

$$R = \tau_2/2yC \qquad \text{A.5.23}$$

## Design Formulæ

The following formulæ are derived from the foregoing by inserting $\tau_1 = 82{\cdot}77$ $\mu$sec and $\tau_2 = 16$ $\mu$sec and reducing to the most convenient practical units. Three amounts of overswing (10 per cent, 30 per cent and 100 per cent) have been allowed for. The last is impossible but represents approximately the limiting case of a very low-loss circuit.

| $x$ = | 0·1 | 0·3 | 1·0 | |
|---|---|---|---|---|
| $L$ = <br> $R$ = <br> $E_M$ = <br> $i_o$ = | $16{,}900/C$ <br> $3{,}470/C$ <br> $7{\cdot}17E_p$ <br> $0{\cdot}91I_p$ | $22{,}700/C$ <br> $6{,}650/C$ <br> $7{\cdot}86E_p$ <br> $0{\cdot}77I_p$ | $26{,}000/C$ <br> $8{,}000/C$ <br> $8{\cdot}1E_p$ <br> $0{\cdot}5I_p$ | mH, pF <br> kΩ, pF <br> V <br> mA |

$$n = \sqrt{L/L_L}$$

$$L_S = L_L/\sqrt{1 - k^2}$$

$$L_p = n^2 L_S$$

$$\eta_{T(opt)} = \frac{k^2}{(1 + \sqrt{1 - k^2})^2}$$

$$E_p = 0{\cdot}0121\ I_L\ L_L\ n/\sqrt{\eta_{T(opt)}} \qquad \text{[V, mA, mH]}$$

$$I_p = I_L/n\sqrt{\eta_{T(opt)}}$$

*Example*

Let $L_L = 8{\cdot}9$ mH, $I_L = 600$ mA, $k = 0{\cdot}98$, $x = 0{\cdot}1$, $C = 100$ pF, $R_L = 15\ \Omega$.

Then $L = 169$ mH, $R = 34{\cdot}7$ kΩ, $n = 4{\cdot}35$, $L_S = 52$ mH, $L_p = 0{\cdot}98$ H, $\eta_{T(opt)} = 0{\cdot}66$, $E_p = 346$ V, $E_M = 2{\cdot}73$ kV, $i_o = 170$ mA.

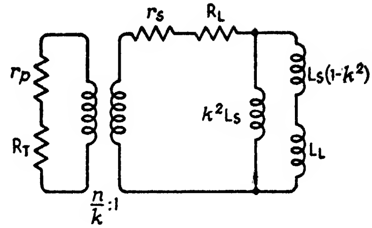
NOTE.—$E_M$ is the peak voltage with respect to positive h.t. The peak voltage with respect to earth is $E_M$ plus the h.t. voltage and the latter must be at least 70 V plus the cathode bias voltage plus $E_p$. In the above example, the h.t. supply must be at least 420 V and the peak voltage to earth is at least 3·15 kV. In estimating the h.t. voltage needed $E_M$ should be increased by some 10 per cent to allow for the voltage drops across the series resistance of components.

## Frame Scan

Equations (A.5.1–5) apply to the frame scan as well as to the line, but subsequent approximate equations are not necessarily valid. With the large value of $\tau_1$ (18·6 msec) appropriate to the frame scan the back e.m.fs across the deflector coil and leakage inductances are usually negligible compared with the voltage drops across the resistive elements.

Although equation (A.5.5) is valid, it is not in a useful form for the determination of the transformer secondary inductance $L_S$ because it requires $r_S$ to be known and this cannot be known until

**Fig. A.5.3—Simplified equivalent circuit for frame scan**




$L_S$ has been found. However, for a given form of construction the ratio $r_S/L_S$ is substantially independent of $L_S$ and equation (A.5.5) is usefully re-written in the form

$$L_S = L_L \frac{\frac{\tau_1}{2} \cdot \frac{R_L}{L_L}\left(1 + \frac{1}{a}\right) - 1}{1 - \frac{\tau_1}{2} \cdot \frac{r_S}{L_S}\left(1 + \frac{1}{a}\right)} \qquad \text{A.5.24}$$

Referring to Fig. A.5.1 the effect of $C$ is negligible on both scan and fly-back, $R$ represents transformer core losses combined with any resistance connected across the primary, and apart from their effect on linearity dealt with above, the inductances all have negligible effect.

The valve then works into a substantially resistive load

$$R_a = \frac{RR_p}{R + R_p} \qquad \text{A.5.25}$$

which should be chosen for the optimum performance of the valve. The effective resistance viewed from the transformer primary is

$$R_p = n^2 \left[R_L + r_S\left(1 + \frac{L_S}{r_S} \cdot \frac{r_p}{L_p}\right)\right] \qquad \text{A.5.26}$$

since $k^2 \approx 1$.

## Frame Fly-back

On fly-back the circuit is approximately equivalent to Fig. A.5.3—approximate only because $k^2L_S$ is transferred from the left to the right of $r_S + R_L$ for simplicity.

The total resistance effective in the primary circuit is

$$R_e = R_T + R_p \qquad \text{A.5.27}$$

and the inductance is

$$L_e = \frac{n^2 L_S \{L_L + L_S(1 - k^2)\}}{L_L + L_S} \qquad \text{A.5.28}$$

The current decays exponentially and the change of current is 99 per cent complete when the exponent is 4·5; that is,

$$\tau_2 R_e/L_e = 4{\cdot}5 \qquad \text{A.5.29}$$

Now $R_T$ is the parallel value of $R$ with the a.c. resistance of the valve, so

$$R_T = \frac{r_a R}{r_a + R} \qquad \text{A.5.30}$$

These relations are sufficient to formulate the design equations.

## Design Formulæ

The following formulæ are derived from the foregoing by inserting $\tau_1 = 18{\cdot}62$ msec and $\tau_2 = 1{\cdot}38$ msec and reducing to the most convenient practical units.

$$L_S = L_L \frac{0{\cdot}00931\,(1 + 1/a)\,R_L/L_L - 1}{1 - 0{\cdot}00931\,(1 + 1/a)\,r_S/L_S} \qquad [\text{H}, \Omega]$$

$$r_S = L_S\,(r_S/L_S) \qquad [\text{H}, \Omega]$$

$$\frac{R_p}{n^2} = R_L + r_S\left(1 + \frac{L_S}{r_S} \cdot \frac{r_p}{L_p}\right)$$

$$\frac{L_e}{n^2} = \frac{L_L + L_S\,(1 - k^2)}{1 + L_L/L_S}$$

$$\frac{R_T}{n^2} = 3{,}260\,\frac{L_e}{n^2} - \frac{R_p}{n^2} \qquad [\text{H}, \Omega]$$

$$n = \sqrt{\left[\frac{r_a R_a}{r_a + R_a} \middle/ \frac{(R_p/n^2)(R_T/n^2)}{(R_p/n^2) + (R_T/n^2)}\right]}$$

$$R_p = n^2\,(R_p/n^2) \qquad L_p = n^2 L_S$$

$$R = \frac{R_p R_a}{R_p - R_a} \qquad r_p = L_p\,(r_p/L_p)$$

$$I_p = \frac{I_L}{n}\left(1 + \frac{L_L}{L_S}\right)$$

$$i_o = I_p + i_1 = \text{peak current in transformer primary}$$

where $i_1$ is the minimum permissible anode current.

$$i_{ao} = \left(I_p + i_1\right)\left(1 + \frac{r_p}{R}\right) + n\,I_L\,\frac{R_L + r_S}{2R}$$

$$= \text{peak anode current of the valve.}$$

$$E_S = I_L\,\frac{R_L + r_S}{2} \qquad [\text{V, A}, \Omega]$$

$$E_p = n\,I_L\,\frac{R_L + r_S}{2} + i_o\,r_p \qquad [\text{V, A}, \Omega]$$

*Example*

Let $L_L = 8{\cdot}9$ mH, $R_L = 15\ \Omega$, $k = 0{\cdot}99$, $a = 0{\cdot}2$, $r_S/L_S = r_p/L_p = 10$, $I_L = 600$ mA, $i_1 = 10$ mA, $r_a = 50$ kΩ, $R_a = 8$ kΩ.

Then $L_S = 1{\cdot}88$ H, $r_S \doteq 18{\cdot}8\ \Omega$, $R_p/n^2 = 52{\cdot}6\ \Omega$, $L_e/n^2 = 0{\cdot}0463$ H, $R_T/n^2 = 98{\cdot}4\ \Omega$, $n = 14{\cdot}14$, $R_p = 10{\cdot}5$ kΩ, $R = 33{\cdot}6$ kΩ,

$I_p = 42{\cdot}6$ mA, $i_o = 52{\cdot}6$ mA, $i_{ao} = 58{\cdot}6$ mA, $E_p = 341{\cdot}5$ V, $L_p = 376$ H, $r_p = 3{\cdot}76$ kΩ.

If $r_S/L_S = r_p/L_p = 5$, all other initial values remaining unchanged, the figures become:—

$L_S = 1{\cdot}15$ H, $r_S = 5{\cdot}75$ Ω, $R_p/n^2 = 26{\cdot}5$ Ω, $L_e/n^2 = 0{\cdot}0316$ H, $R_T/n^2 = 76{\cdot}5$ Ω, $n = 18{\cdot}7$, $R_p = 9{\cdot}26$ kΩ, $R = 59$ kΩ, $I_p = 32{\cdot}3$ mA, $i_o = 42{\cdot}3$ mA, $i_{ao} = 45{\cdot}6$ mA, $E_p = 202$ V, $L_p = 403$ H, $r_p = 2{\cdot}02$ kΩ.

## Resistance-Capacitance Coupling

### Frame Scan

The equivalent circuit is shown in Fig. A.5.4 and as before the coil current is assumed to have the ideal form

$$i_L = I_L\,(t/\tau_1 - 1/2) \qquad \text{A.5.1}$$

on the scan. As $\tau_1$ is large, 18·6 msec, the rate of change of current is small and the back e.m.f. across $L_L$ is small compared with the voltage drop across $R_L$. Neglecting it, therefore,

$$e_p = c_1 - I_L\frac{R_L}{2} + I_L\frac{t}{\tau_1}\left\{R_L - \frac{\tau_1}{2C}\right\} + I_L\frac{t^2}{\tau_1^2}\cdot\frac{\tau_1}{2C} \qquad \text{A.5.31}$$

where $c_1$ is a constant of integration.

Now $i_r = e_p/R$, therefore,

$$i_T = \frac{c_1 - I_L\dfrac{R}{2}\left(1 + \dfrac{R_L}{R}\right)}{R} + I_L\frac{t}{\tau_1}\left(1 + \frac{R_L}{R} - \frac{\tau_1}{2\,CR}\right) +$$

$$I_L\frac{t^2}{\tau_1^2}\cdot\frac{\tau_1}{2\,CR} \qquad \text{A.5.32}$$

At $t = 0$, the valve supplies a certain current $i_1$, the minimum permissible anode current and $i_L = -I_L/2$, this current being supplied by the discharge of $C$. The current in $R$ is thus $i_1 + I_L/2 = i_{TO}$ and the initial value of $e_p$ is $e_{po} = R\,(i_1 + I_L/2)$.

From the above equations

$$e_{po} = c_1 - I_L\frac{R_L}{2} = i_1\,R + I_L\,\frac{R}{2} \qquad \text{A.5.33}$$

$$\therefore\ c_1 = i_1\,R + I_L\frac{R}{2}\left(1 + \frac{R_L}{R}\right) \qquad \text{A.5.34}$$

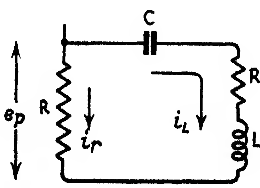



**Fig. A.5.4—Equivalent circuit for a resistance-capacitance coupled stage**

Therefore, the maximum value of $e_p$, when $t = \tau_1$, is

$$E_p = i_1 R + I_L (R_L + R/2) \qquad \text{A.5.35}$$

and

$$I_p = I_L \left(1 + \frac{R_L}{R}\right) \qquad \text{A.5.36}$$

The ratio of the square-law to the linear components of $i_T$ is

$$a = \frac{\tau_1/2CR}{1 + R_L/R - \tau_1/2CR} \qquad \text{A.5.37}$$

whence

$$C = \frac{\tau_1}{2(R + R_L)} \cdot (1 + 1/a) \qquad \text{A.5.38}$$

### Frame Fly-back

On fly-back, $C$ can be ignored and the current decreases exponentially with an exponent $-t(R + R_L)/L_L$. The change of current is 99 per cent complete when the exponent is 4·5, hence as a minimum

$$R + R_L = 4{\cdot}5\, L_L/\tau_2 \qquad \text{A.5.39}$$

It has been assumed above that the a.c. resistance of the valve is very large compared with $R$. If it is not, the value used for $R$ should be that of the actual coupling resistance and valve resistance in parallel. Then $I_T$ will not be the anode current, but the current of the constant-current generator equivalent to the valve.

### Design Formulæ

For $\tau_1 = 18{\cdot}62$ msec, $\tau_2 = 1{\cdot}38$ msec.

$$R = 3{\cdot}26\, L_L - R_L \qquad [\text{k}\Omega, \text{H}]$$

(This is a minimum permissible value.)

$$C = \frac{9{\cdot}31}{R + R_L}(1 + 1/a) \qquad [\mu\text{F}, \text{k}\Omega]$$

$$I_p = I_L (1 + R_L/R)$$

$$E_p = i_1 R + I_L (R_L + R/2) \qquad [\text{V}, \text{mA}, \text{k}\Omega]$$

*Example*

Let $L_L = 1$ H, $R_L = 1{\cdot}7$ kΩ, $I_L = 40$ mA, $a = 0{\cdot}2$, $i_{am} = 10$ mA.

Then $R = 1{\cdot}56$ kΩ, $C_c = 17{\cdot}2$ μF, $I_p = 83{\cdot}5$ mA, $E_p = 115$ V. In practice it would usually be better to increase $R$ above its minimum value and so reduce $I_p$, since a higher value of $E_p$ is usually permissible.

If $a$ is made small, so that negligible square-law current need be supplied, $C$ increases greatly. Thus if $a$ is made 0·02, $C$ becomes 146 μF.

# APPENDIX 6

## RADIO-FREQUENCY AMPLIFIERS

### Stagger-Tuned Single-Circuit couplings with optimum stagger

All stages have the basic circuit of Fig. A.6.1 with inductance $L$, capacitance $C$, and resistance $R$, while the valve has a mutual conductance $g_m$. The components in different stages are distinguished by numerical subscripts. The order of stages is arbitrary and the actual order is dictated by other considerations.

SINGLE-STAGE AMPLIFIER

$$\left.\begin{aligned} nCR &= 159\sqrt{(S_n^2 - 1)} \\ &\text{or can be read from curve A, Fig. A.6.3} \\ A &= g_m R \\ L &= 25{,}400/f_r^2 C \end{aligned}\right\} \quad \text{A.6.1}$$

The two first equations also give the gain per stage of an optimum stagger amplifier of any number of stages, but the value of $R$ obtained is not necessarily that to be used in any stage.

TWO-STAGE AMPLIFIER

$$\left.\begin{aligned} C &= \sqrt{C_1 C_2} \\ nCR &= 225\sqrt[4]{(S_n^2 - 1)} \\ &\text{or can be read from Curve B, Fig. A.6.3} \\ f_r &= f_m\sqrt{(1 - n^2/4f_m^2)} \\ L &= 25{,}400/f_r^2 C \\ y &= 159/f_r CR \\ a^2 &= \frac{2 + y\sqrt{4 - y^2}}{2 - y^2} \\ R_1 &= RaC/C_1 \qquad R_2 = RC/aC_2 \\ L_1 &= La^2C/C_1 \qquad L_2 = LC/a^2C_2 \\ f_{r1} &= f_r/a \qquad f_{r2} = f_r a \\ A &= \frac{g_{m1}\, g_{m2}\, R^2}{2}. \end{aligned}\right\} \quad \text{A.6.2}$$

### THREE-STAGE AMPLIFIER

$$\left.\begin{aligned}
C &= \sqrt{C_2 C_3} \\
R &= \frac{318\sqrt[6]{S_n^2 - 1}}{nC} \\
f_r &= f_m\sqrt{(1 - n^2/4f_m^2)} \\
y &= \frac{n}{2f_r\sqrt[6]{(S_n^2 - 1)}} \\
a^2 &= \frac{1 + 1{\cdot}73y + 0{\cdot}5\,y^2}{1 - y^2} \\
L &= 25{,}400/f_r^2 C \\
R_1 = CR/2C_1 \qquad & R_2 = CRa/C_2 \qquad R_3 = CR/aC_3 \\
L_1 = LC/C_1 \qquad & L_2 = La^2C/C_2 \qquad L_3 = LC/a^2C_3 \\
f_{r1} = f_r \qquad & f_{r2} = f_r/a \qquad f_{r3} = f_r a \\
A &= g_{m1}g_{m2}g_{m3}\left(\frac{159}{nc}\right)^3\sqrt{(S_n^2 - 1)}
\end{aligned}\right\} \quad \text{A.6.3}$$

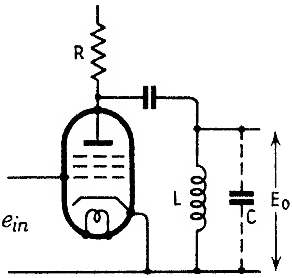
### FOUR-STAGE AMPLIFIER

$$\left.\begin{aligned}
C &= \sqrt{C_1 C_2} \\
R &= \frac{416\sqrt[8]{(S_n^2 - 1)}}{nC} \\
f_r &= f_m\sqrt{(1 - n^2/4f_r^2)} \\
y &= \frac{n}{2f_r\sqrt[8]{(S_n^2 - 1)}} \\
a_1^2 &= \frac{1 + 1{\cdot}207\,y^2 + 1{\cdot}414\,y\sqrt{1 - y^2/4}}{1 - 1{\cdot}707y^2} \\
a_2^2 &= \frac{1 - 1{\cdot}207y^2 + y\sqrt{1 - 2{\cdot}9\,y^2}}{1 - 1{\cdot}707y^2} \\
L &= 25{,}400/f_r^2 C \\
L_1 = LCa_1^2/C_1 \qquad & R_1 = RC\,a_1/C_1 \qquad f_{r1} = f_r/a_1 \\
L_2 = LC/a_1^2C_2 \qquad & R_2 = RC/a_1C_2 \qquad f_{r2} = f_r a_1 \\
L_3 = LCa_2^2/C_3 \qquad & R_3 = RC\,a_2/2{\cdot}414\,C_3 \qquad f_{r3} = f_r/a_2 \\
L_4 = LC/a_2^2C_4 \qquad & R_4 = RC/2{\cdot}414\,a_2C_4 \qquad f_{r4} = f_r a_2 \\
A_T &= g_{m1}g_{m2}g_{m3}g_{m4}\left(\frac{159}{nC}\right)^4\sqrt{(S_n^2 - 1)}
\end{aligned}\right\} \quad \text{A.6.4}$$

### BAND-PASS COUPLINGS

The circuit has the form of Fig. A.6.2 and the first step is to determine $nCR$ from curve C, Fig. A.6.3, then

Fig. A.6.1—Basic circuit of a single-circuit coupling




$$R_1 = (nCR)/nC_1 \qquad R_2 = (nCR)/nC_2$$

$$A = g_m \frac{\sqrt{R_1 R_2}}{2}$$

$$L_1 = 25{,}400/f_r^2 C_1 \qquad L_2 = 25{,}400/f_r^2 C_2$$

$$k = 1/\sqrt{[1 + 3{\cdot}94 \times 10^{-5}(f_r CR)^2]}$$

$$M = k\sqrt{L_1 L_2}$$

*Symbols*

$f_m$ = mid-band frequency
$f_r$ = resonant frequency
$n$ = bandwidth $= f_2 - f_1$
$f_1, f_2$ = limits of bandwidth
$S_n$ = ratio of $\dfrac{\text{response at } f_1}{\text{response at } f_1 \text{ and } f_2}$
$A$ = amplification
$M$ = mutual inductance
$k$ = coupling coefficient

*Units*

In all these equations, $R$ is in k$\Omega$, $L$ in $\mu$H, $C$ in pF, frequency in Mc/s, $g_m$ in mA/V.

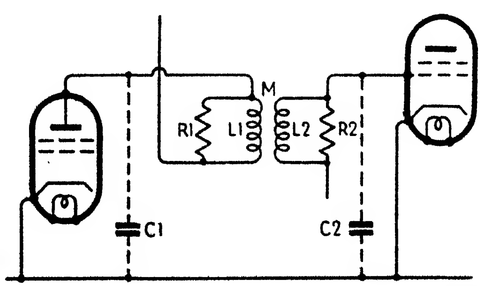



Fig. A.6.2—Two circuit coupling of the band-pass type

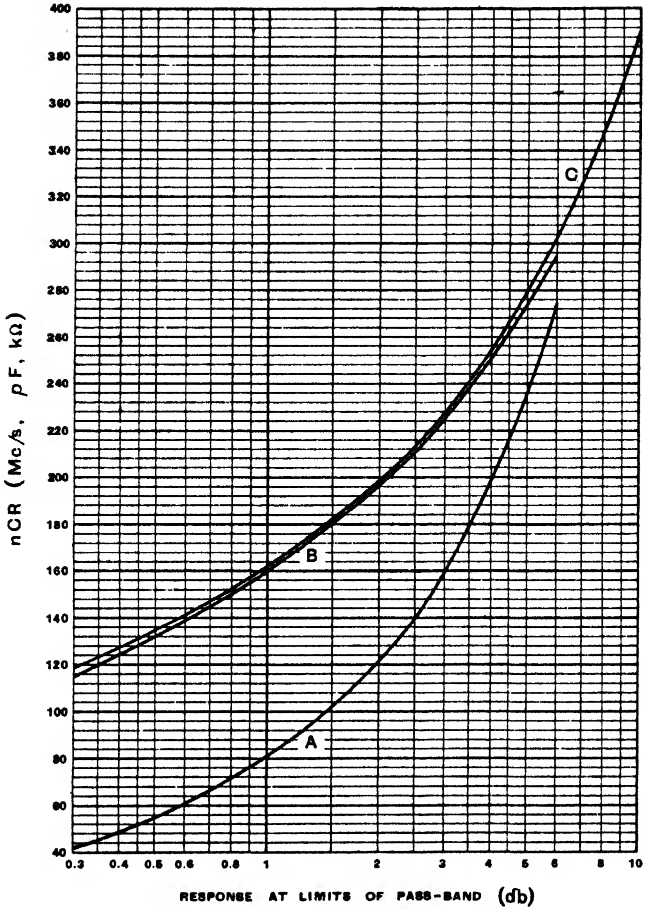



**Fig. A.6.3—Design curves for r.f. couplings are shown here. Curve A refers to single circuits tuned all to the same frequency, curve B to stagger tuning and curve C to band-pass tuning. The bandwidth is n (Mc/s), the total capacitance C (pF), and the resistance R (kΩ)**

## REJECTOR CIRCUITS

Referring to Fig. A.7.1, when $R_1 \ll \omega L_1$

$$Z = R_S + jX_S = \frac{1}{j\omega C_1} + \frac{j\omega L_1(1 - \omega^2 L_1 C_2) + \omega^2 L_1 C_2 R_1}{(1 - \omega^2 L_1 C_2)^2 + \omega^2 C_2{}^2 R_1{}^2} \quad \text{A.7.1}$$

When $R_1$ is small, then very nearly,

$$X_S = 0 \text{ when } \omega^2 L_1 C_2(1 + C_1/C_2) = 1$$

$$X_S = \infty \text{ when } \omega^2 L_1 C_2 = 1$$

The condition $X_S = 0$ corresponds to the rejection frequency $f_\infty$ of the circuit, while the condition $X_S = \infty$ corresponds to the frequency of the adjacent edge of the vision-channel pass-band, $f_\infty + \Delta f$. Therefore,

$$(1 + \Delta f/f_\infty)^2 = 1 + C_1/C_2$$

and $$\frac{C_1}{C_2} = \frac{2\Delta f}{f_\infty}\left(1 + \frac{\Delta f}{2f_\infty}\right) \approx \frac{2\Delta f}{f_\infty} \quad \text{A.7.2}$$

At $f_\infty$, $X_S = 0$ and

$$Z = R_S = \frac{\omega_\infty{}^2 L_1 C_2 R_1}{(1 - \omega_\infty{}^2 L_1 C_2)^2 + \omega_\infty{}^2 C_2{}^2 R_1{}^2}$$

$$= \frac{R_1}{\dfrac{(C_1/C_2)^2}{1 + C_1/C_2} + \dfrac{C_2 R_1{}^2}{L_1}}$$

If $Q = \omega_\infty L_1/R_1$

$$R_S = \frac{1}{\omega_\infty C_1 Q \dfrac{C_1}{C_2}\{1 + 1/(QC_1/C_2)^2\}} \quad \text{A.7.3}$$

$$= \frac{1}{4\pi \Delta f C_1 Q\left(1 + \dfrac{\Delta f}{2f_\infty}\right)\left\{1 + 1\Big/\left(Q\dfrac{2\Delta f}{f_\infty}\left[1 + \dfrac{\Delta f}{2f_\infty}\right]\right)^2\right\}} \quad \text{A.7.4}$$

and when $\Delta f \ll 2f_\infty$ this reduces to

$$R_S \approx \frac{1}{4\pi \Delta f C_1 Q\{1 + (f_\infty/2\Delta f Q)^2\}} \quad \text{A.7.5}$$

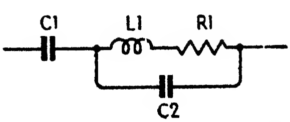



Fig. A.7.1

At $f_\infty + \Delta f$, $X_S = \infty$ and

$$Z = \frac{1}{j\omega C_1} + \frac{L_1}{C_2 R_1} = \frac{1}{j\omega C_1} + R_D \qquad \text{A.7.6}$$

Now if $Q$ is the same at this frequency as at $f_\infty$

$$R_D = \frac{L_1}{C_2 R_1} = \frac{\frac{C_1}{C_2} Q}{\omega_\infty C_1} \qquad \text{A.7.7}$$

If the circuit across which the rejector is connected has shunt resistance $R$ and capacitance $C$ the attenuation at $f_\infty$ introduced by the rejector is approximately

$$20 \log R/R_S \quad \text{db} \qquad \text{A.7.8}$$

and it is obviously desirable to have

$$R_D \gg R$$
$$C_1 \ll C$$

if the effect of the rejector in the vision-channel pass-band is to be small.

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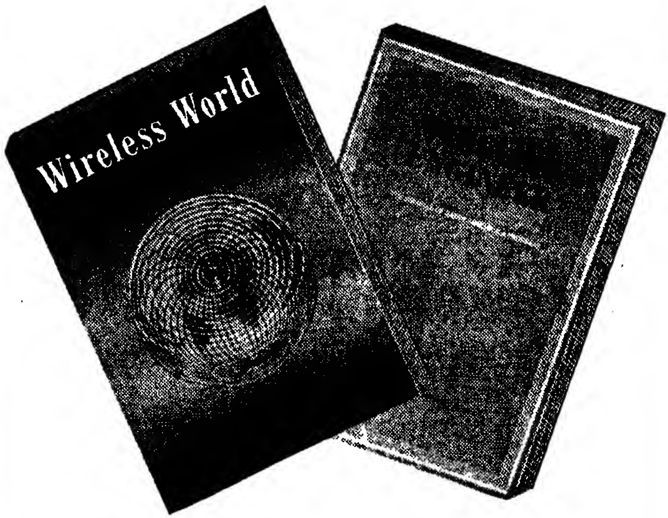
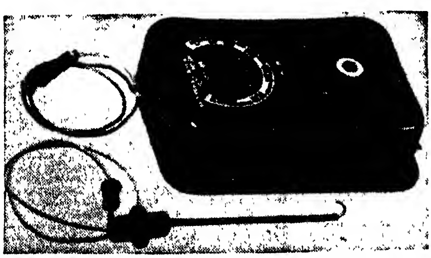

Wireless World



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