Colour Television

N.T.S.C. SYSTEM, PRINCIPLES AND PRACTICE

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Preface

The object of this book is to provide a comprehensive account of the American N.T.S.C. colour television system, and the British and Continental versions of this system, from the point of view of the colour receiver engineer. Since the British reader will probably find it easier to think in terms of the monochrome system with which he is most familiar, N.T.S.C. colour television systems are explained with particular reference to the 405 line version, but wherever there are differences between the 405, 525, and 625 line systems, these differences are fully explained. A proper knowledge of the receiver demands, in turn, familiarity with the make-up of the signal and the principles of colour vision, and these aspects of N.T.S.C. have also been described in some detail.

No book on colour television can do other than draw heavily on the original work of the American National Television System Committee, as reported in references 2 and 4 of the bibliography, and here due emphasis must be placed on the major part played by the Radio Corporation of America in the development of colour television. A considerable body of opinion considers, indeed, that the present American system of colour television would more appropriately be called the R.C.A. colour television system.

The book as a whole is concerned with colour television *broadcasting*, but the same principles, cameras, display tubes, etc., are used in those industrial and closed circuit colour television applications in which only one cable is used between camera and monitor. In those few cases where the monitor is very close to the camera it may be more economic to use three connecting cables carrying the three primary signals. Monitors for this latter case present special problems in stability and are not specifically dealt with in this book.

It will be found that four chapters are of more than average length. Of these, one is concerned with transmitter coding and two deal with receiver decoding, while the fourth discusses shortcomings of N.T.S.C. systems. These subjects warrant thorough description since they include ideas and techniques quite foreign to monochrome practice.

Every attempt has been made to keep mathematics to a minimum, but an exception has been made in the appendices which are mainly concerned with the quantitive behaviour of phase locked loops. The latter are an essential feature of the N.T.S.C. system, and call for adequate treatment.

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Acknowledgements

Introduction

1.1. About this book

Colour television is a practical reality in America and Japan and a British version of the American N.T.S.C. (National Television System Committee) system has been successfully developed in this country.

This new field of modern communications makes an absorbing study. It is the culmination of large scale, co-ordinated research which has succeeded in reconciling several conflicting requirements. Colour television today is complex, but the step from black and white or monochrome television to colour is smaller than that from sound broadcasting to vision.

It is suggested that it is preferable if the reader acquires a broad outline of the subject before proceeding to a detailed examination of each part. This book has been written in a form suitable for such a study, since each chapter begins with a preview of its contents and ends with a summary of the more important points.

Colour television transmitters and receivers are both dealt with but there is a special emphasis on the latter. Familiarity with black and white television is assumed but a physical explanation of colour television is presented whenever possible rather than a mathematical description of it. In general, only elementary mathematics are used, and always with explanations in the text. Difficulties and intricacies have not been glossed over and some matters have been dealt with mathematically when this is the only satisfactory way. The reader who is not mathematically inclined can understand the conclusions reached, however, since advanced mathematics have been deliberately avoided in the descriptions of the operating principles and the circuits of transmitters and receivers.

The service engineer has been much in the authors' minds and after the basic principles have been covered in the earlier chapters, there is a considerable amount of practical information on the operation, installation and servicing of colour receivers.

Student and practical man alike will find a study of colour television rewarding, for it will become a universal form of entertainment.

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1.2. Introduction

This chapter describes in simple phrases an overall picture of the basic N.T.S.C. colour television system. This system makes use of certain limitations in the ability of the human eye and brain to register what they are actually seeing, and so the performance of the eye under various conditions is described. Finally, to prepare the ground for a more detailed discussion of the N.T.S.C. system, there is a short résumé of some relevant points in monochrome television.

1.3. Why N.T.S.C.?

Black and white television, which was a practical public service nearly a quarter of a century ago, has had a surprisingly long popularity considering its stunted representation of reality. Bereft of colour, it would have remained a laboratory curiosity had there not been very considerable difficulties in broadcasting coloured images. The extent of the growth of monochrome television has, by using so many channels in our overcrowded ether, now become a major problem in instituting a colour television service.

To tackle the problems which were holding back a satisfactory colour broadcasting service, the American Radio Industry (in 1950) placed its resources, facilities and knowledge at the disposal of its own National Television System Committee. Most of the larger American radio companies co-operated in this enterprise, and the colour television system finally evolved has been given the initials of the organizing committee, N.T.S.C. It is achieving world-wide acceptance as a practical and economic way of adding colour to our existing television services.

Simpler methods of transmitting colour have been rejected for a variety of reasons—they used too much bandwidth, gave unsatisfactory pictures, or were likely to restrict the development of future advances in the science. The N.T.S.C. system can use the same channel width as monochrome systems of similar resolution and will produce on existing monochrome receivers a good black and white version of the transmitted colour picture. Any such system is said to be *compatible*. Colour receivers for the N.T.S.C. system can also be designed to produce good black and white pictures from the existing monochrome transmissions. This is called *reverse compatibility*. A disadvantage of the N.T.S.C. system is that black and white receivers cannot be satisfactorily modified to colour reception, but such ordinary television receivers can receive the colour transmissions in monochrome.

The American N.T.S.C. system has been adapted to suit the various transmission standards used in other parts of the world.

The use of positive modulation and A.M. sound in our British 405-line system avoids some of the difficulties which arise with the original American system.

1.4. Outline of N.T.S.C. system

Fortunately, in colour television it is not necessary to transmit separate signals corresponding to every distinct colour. Just as an artist can mix a complete range of hues from three primary paints so can the television receiver give the impression of fully coloured scenes from three primary coloured lights. But the artist uses paints which appear coloured due to the reflection of light from them whereas the coloured effect in a receiver comes from transmitted light. Equal parts of primary paints can produce a black effect but a suitable mixture of three coloured light primaries, red, green and blue, produces the sensation of white.

The colour camera splits the scene it is viewing into a red, a green and a blue image, using an arrangement of mirrors and coloured light filters. At the receiver, the red, green and blue images can be recombined to form the full-coloured picture. Fig. 1.1 (a) shows a simple picture of a red tulip against a blue sky, with some white clouds. If the colours were very pure, the three component images would be as shown in Fig. 1.1 (b), (c) and (d).

Three separate camera tubes convert the three coloured images into three electrical signals, in the same way as an ordinary black and white camera tube converts the brightness of the scene into one video signal. In colour television each video signal is proportional to the amount of either red, green or blue in the picture.

1.5. Transmission of three signals

To transmit three separate signals, each in a manner similar to black and white transmission, would require three television channels, so the signals are modified before transmission in order that they can be sent in only one channel. This modification of the red, green and blue signals is referred to as *coding*. The main part of the coded signal is similar to a normal black and white television signal and consists of sync pulses and a video waveform. The video waveform is made up of a signal formed by adding a fraction of each of the red, green and blue signals together, so that it is representative of the brightness of the original scene.

The remaining part of the information, the colouring information, is transmitted with a much reduced bandwidth of between $\frac{1}{9}$ and $\frac{1}{3}$ of the monochrome bandwidth. This means that at the receiver an unsharp colouring picture is superimposed on a sharp

BLACK

RED

BLUE

GREEN

WHITE

(a) FULL COLOUR PICTURE

(b) RED COMPONENT PICTURE

(c) GREEN COMPONENT PICTURE

(d) BLUE COMPONENT PICTURE
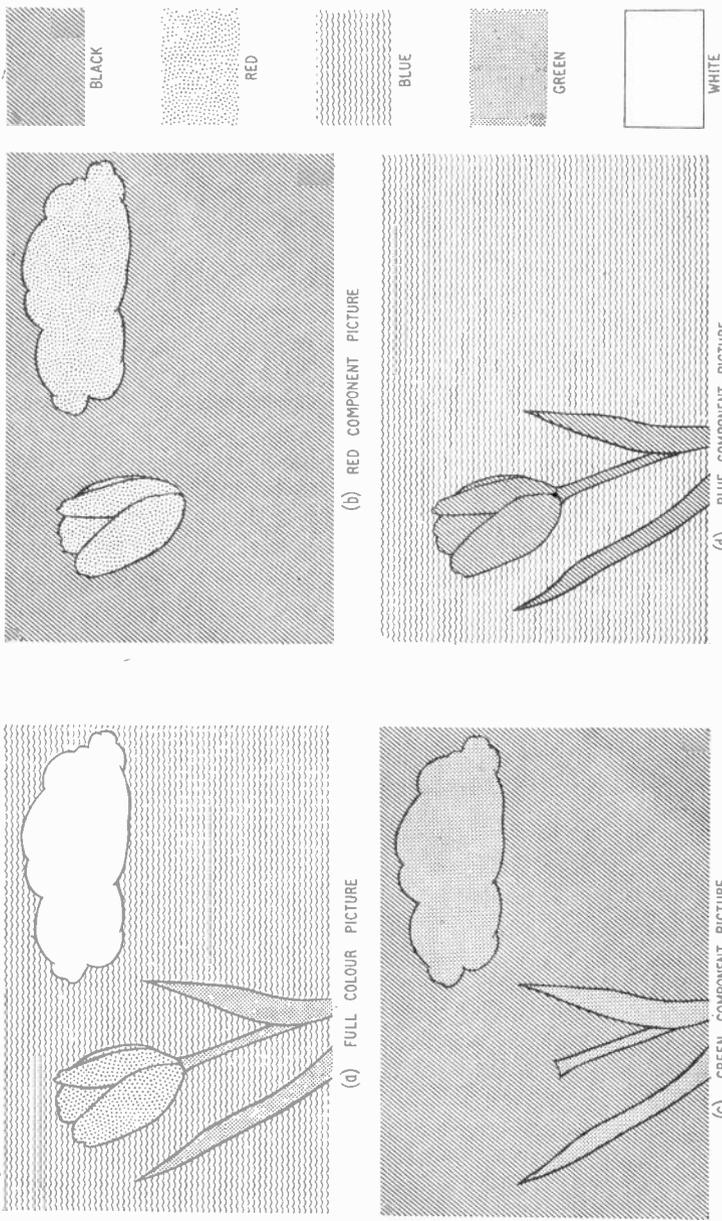
Fig. 1.1.   Analysis of a scene into its red, green and blue components

black and white brightness picture. The resulting effect is similar to the best quality photographic colour transparencies.

The electrical signal representing the scene brightness is subtracted from the signal representing the red component image so that the resultant electrical signal is proportional to the difference between the red image and a black and white version of the scene. This signal is called the *red difference signal*. A *blue difference signal* is formed in a similar manner by subtracting the brightness signal from the blue image signal. These two colour difference signals together with the brightness signal contain sufficient information to enable the receiver to reconstruct the full colour picture. Two colour difference signals are not unduly repeating the information which is already being conveyed by the brightness signal.

1.6. Colour sub-carrier

The colouring information is transmitted by modifying the colour difference signals—which are also restricted in their bandwidth—and then modulating both signals onto a carrier with a frequency,
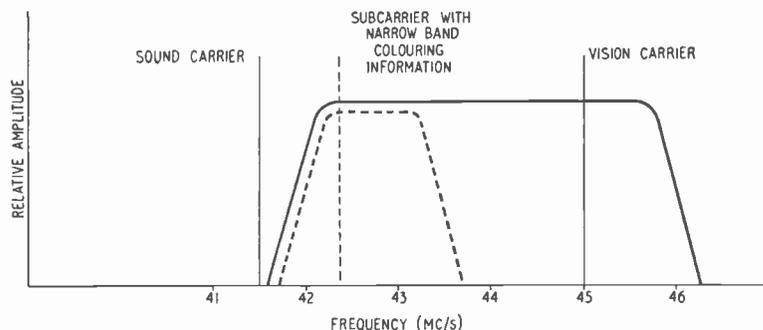
Fig. 1.2. Sub-carrier in vision band

for our British 405-line system, of about 2·7 Mc/s. This 2·7 Mc/s carrier is called the *sub-carrier*. The modulated sub-carrier is added to the brightness signal and both are modulated onto the vision carrier, and tranmsitted together (Fig. 1.2).

1.7. Reforming the colour picture at the receiver

At the colour receiver the signals are separated out again, or decoded, into their red, green and blue values, and applied to the colour display. The colour display may be a combination of three cathode-ray tubes viewed by a system of mirrors so that the three

images are superimposed on each other (Fig. 1.3). More probably the colour display will consist of a single colour cathode-ray tube which can reproduce all the three primary red, blue and green images at once. In both cases a full colour picture is seen by the viewer, including such colours as yellow, purple and brown.

To enable the colour receiver to separate the electrical signals which are carrying the colouring information, the transmitter sends
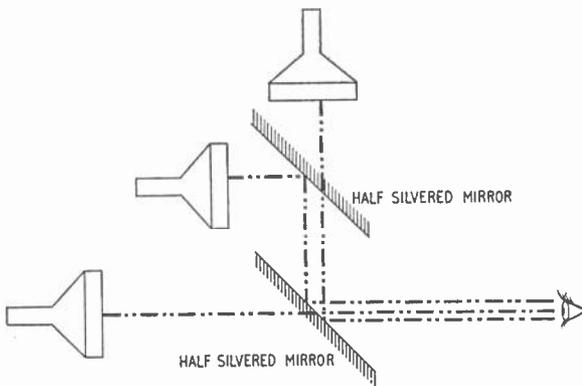
Fig. 1.3.   Three-tube colour display

a special colour synchronizing signal during the back porch of the line blanking pulse. This colour synchronizing signal is called the *colour burst*.

1.8.   Reception on ordinary black and white receivers

The presence of the sub-carrier signal in the normal video band would cause interference with black and white reception, were precautions not taken in the design of the transmitting system. These precautions are not completely successful and on good quality monochrome receivers an interference pattern of small dots can be seen on close inspection in those parts of the picture corresponding to areas which are both strongly coloured and bright.

On the British 405-line system these dots appear to move slowly upwards. Large scale tests have shown that most viewers do not find these dots annoying, and a simple modification to the receivers will eliminate most of them.

1.9.   Limitations of the human eye

The limitations of the human eye can sometimes be turned to good account. In the cinema the eye is presented with a succession

of stationary pictures and persistence of vision leads the brain to register the sensation of movement. Television carries this process of deception still further by constructing each stationary picture element by element.

The performance of the human eye varies considerably with the viewing conditions: for example, most people need a strong light in order to read fine print. For the average conditions of television viewing it is possible to draw broad conclusions about the average behaviour of eyes, although it is necessary to remember that there are variations between one person and another.

1.10. The acuity of the human eye

There is a limit to the smallness of the detail which most people can see without magnification. Thus the average limit of resolution corresponds to seeing the separate interlaced lines of a 21 in. 405-line receiver at a distance of about 10 ft, and the maximum acuity of the eye is usually taken as 1 minute of arc ($\frac{1}{60}$ of a degree). The eye can only resolve such fine detail when it is looking directly at an object.

Objects only slightly off the direct line of vision are less sharp and the colours of such objects are less clear, while at the extreme edge of the field of vision, objects are only seen as varying shades of grey.

To make up for this limitation, the eye is very sensitive to movement which occurs at the edge of the field of view and a slight change seen out of the corner of the eye attracts attention so that the eye turns to look directly at the moving object to see what it is. Hence very bright television pictures appear to flicker more when we are looking to one side of them.

Monochrome television standards assume that people will not sit too near to the receivers, and bandwidth is saved by not transmitting all the fine detail in the picture. Interlacing was adopted to save further bandwidth and the ragged edges which occur on objects rapidly moving across the picture are accepted in the same way as the line structure of the raster. Still further use is made of the limitations of the eye in colour television, to save bandwidth.

If fine detail is seen by white light, red light or green light of the same brightness, our eyes have about the same acuity, but in blue light of the same brightness there is some evidence that our eyes are not quite as acute. However, the brightness of the blue component of most pictures is much less than that of the green or red components and the sharpness of this blue component is then two or three times less critical than that of the other two images.

1.11.　Contrast and adaptation

Switching on the room lighting on a bright summer's day makes little difference to the appearance of the room. The change in brightness which can be discerned varies with the brightness of the scene and it is found experimentally that if the smallest noticeable change in a brightness B is ΔB, then

$$\frac{\Delta B}{B} = \text{constant}$$

This is the Weber-Fechner Law and holds fairly well over the range of brightness used in television.

It means that if an interfering signal is producing a brightness change ΔB in a television picture it will be more noticeable in the dark parts of the picture where B is low, than in the highlights where B is large. Typical values of the Fechner fraction $\frac{\Delta B}{B}$, which varies with the conditions under which the eye is viewing and with the colour of the light, are 0·005 for large bright areas, to 0·02 for small areas against a dark background.

Although the human eye can adapt iself to see in dark cellars and the glare of sunlit snow, it cannot accommodate such a large brightness range at one and the same time. Thus although the screen of a television receiver may look white before the set is operating, as soon as the screen fluoresces the contrast between the brightly glowing areas of the picture and the non-fluorescing areas makes the non-fluorescing areas look black instead of white.

A similar sort of adaptation occurs when the eye looks at colours which are nearly white. After looking for a few minutes at a scene with an overall blue cast, the eye corrects its impression of the scene so that the mind assumes that the scene has no overall colour cast. This is why artificial tungsten lamp lighting is acceptable by itself at night time, but looks very yellow in daylight. It is as if there were separate " automatic gain controls " in the red, green and blue responses of the eye, all operating to keep things looking as white as possible.

Similarly, when the eye is adapted to an overall blue cast a normal white colour appears to look yellowish by comparison. This colour contrast adaptation effect has been studied since Goethe's investigation (about 1800) and several attempts have been made to use it as the basis of a two-colour reproduction process in photography. Recently Land, in America, has produced very satisfying colour

pictures by photographing a scene on black and white film, once through a green filter and once through a red filter. The two black and white pictures are projected in register using white and red light respectively. With suitable scenes the eye sees a range of tones from red through pink and white to the complementary colour of greenish-blue, by the adaptation process already mentioned. Further, colours such as yellow and blue are seen by many people in areas of the picture which subtend a small angle at the eye. However, such ideas appear not to be consistently applicable to the wide range of pictures needed for colour television.

When the eye is adapted to normal levels of brightness, it is said to be adapted to *photopic vision*, that is colours are discernible. At low light levels such as moonlight, the eye changes over to *scotopic vision* and loses its ability to discern one colour from another although it can still distinguish between shades of light and dark.

1.12. Colour vision

No one as yet understands exactly the mechanism by which the eye distinguishes one colour from another. The lens of the eye focuses an image of the scene onto the back interior surface of the eye, the *retina*. There the variations in light and colour are converted into electrical impulses and the information is conveyed to the brain through an intricate interconnection of nerve paths by a system of pulse code modulation.

The retina contains two distinct types of light sensitive receptors called *rods* and *cones* after their appearance under the microscope. The rods and cones are about $\frac{1}{500}$ in. in length and very thin, about the wavelength of light in diameter, and stand on end with their small cross-section facing the light. The rods are colour-blind but are more sensitive than the cones, whereas the cones resolve detail better and appear to be responsible for colour vision.

The cones are concentrated near the centre of the retina, at the *fovea*. There are no rods here and this is the only area of the retina which will resolve fine detail; it corresponds to vision over an angle of less than $2°$. The rods are more numerous at the outer edges of the retina and are responsible for vision at low light levels, i.e., scotopic vision. In very dim light things are often better discerned by looking to one side of them so that their image falls on this more sensitive part of the retina and not on the fovea. There are about 120 million rods and 6 million cones in the human retina.

It has been popularly supposed that there are three distinct types of cones, each type sensitive to either red, green or blue light, but physiologists have so far been unable to distinguish any difference

between the cones.   At the time of writing new experiments are being made by shining light into the eye ball and studying the light which is reflected back from the retina.   There is some evidence from these experiments that there may be at least two distinct types of cones.

1.13.   Colour-blindness

A few people, $0 \cdot 003\%$ of the population, see all colours only as variations of light and dark; other people are deficient, completely or partially, in red, green or blue sensitivity, or combinations of these.

The various kinds of colour-blindness have different names. People with normal colour vision are called *trichromats*, and those (6%) who are partially down in sensitivity to one of the three primary colours, red, green or blue, are *anomalous trichromats*.

About 1% to 2% of people see only two primary hues instead of three and of these only $1\frac{1}{2}\%$ are women, for although colour-blindness is in general inherited it is essentially a male complaint.   There are various forms of this dichromatism.   About $\frac{1}{2}\%$ of the population are colour-blind people who confuse reds with greens and see both bluish green and purple light as a neutral grey colour.   This is quite certain because some people are only colour-blind in one eye and are normal in the other eye.   Another $\frac{1}{2}\%$ of the population see only blues, yellows and greys and are completely blind to hues at the extreme red end of the spectrum.   There is a rare form of colour-blindness in which the eye is relatively insensitive to the blue end of the spectrum and sees yellows as greys.

Some 5% of the population are slightly lacking in green sensitivity. A few observers are weak in red sensitivity, whilst in rare cases the blue vision is weak.

Anomalous trichromats may disagree with normal people over the correct setting of the colour balance of a colour television receiver, whereas people with other forms of colour-blindness will generally find the pictures acceptable.

1.14.   Colour perception

The colour of an object depends not only upon that object but also upon the conditions under which it is observed, such as the colour of the light shining upon it, the brightness of the light, the size of the object, the colour of other things near the object, the colours which the eye has just previously been observing and so on.

Staring fixedly at a coloured green patch for a short time causes the eye to become less sensitive to green, and transferring the gaze

to a plain white background will then produce an apparent image of a purple patch on it. These after images are complementary in colour to the original colour.

A piece of blue material laid on a larger area of red looks a different shade of blue from a piece of the same material laid on a green background. This is a form of " crosstalk " between the signals from adjacent areas of the retina.

A small piece of material looks a slightly different colour from a large piece of the same material and women are careful to match cotton to material by using the whole reel rather than a single thread. With fine threads it becomes difficult to see any colour in them until they are grouped together. The eye in fact becomes colour-blind to small objects and the objects are seen only as shades of grey.

Close inspection of a number of differently coloured cotton or silk threads attached to a piece of white card will reveal the colour of each individual thread, whereas from a suitable distance only the position of each thread is distinguishable. At intermediate distances some colours can be discerned and not others. As the apparent size of the coloured area decreases, the eye begins to lose its ability to distinguish between blue, blue-greens and greens of the same brightness, and then between yellow and grey. All colours become more pastel, but red and orange are plainly distinguished from blue-green, although not from themselves at the same brightness. Finally, as the apparent size decreases, reds and then blue-greens also become indistinguishable from grey.

This inability to see the exact colour of a small area should not be confused with the acuity of the eye in various colours, which is the ability to see fine brightness detail under one particular kind of illumination.

The eye is very easily deceived, and if a sharp brightness change is superimposed over a fuzzy colour change, a sharp transition from one bright colour to a different dark colour can be seen. The converse is also true. A slow brightness change from one area to another looks sharp if it is superimposed over a sharp colour transition. The first effect is used in colour television to save bandwidth in the colour channels. Fig. 1.4 illustrates a simple experiment to demonstrate the point.

1.15. Subjective colours

In 1826 a French monk, Benedict Prévost, observed and recorded an unusual phenomenon. A piece of board moved rapidly to interrupt a ray of light in a darkened room produced a colour

effect.   The eye sees various pastel colours if it is presented with
flashes of black and white light in certain sequences.   Benham used
the pattern of Fig. 1.5 (a) pasted onto a top; if this particular pattern
is rotated clockwise about its centre at 8 rev/s, just not fast enough
for the pattern to merge into a uniform grey colour, most people
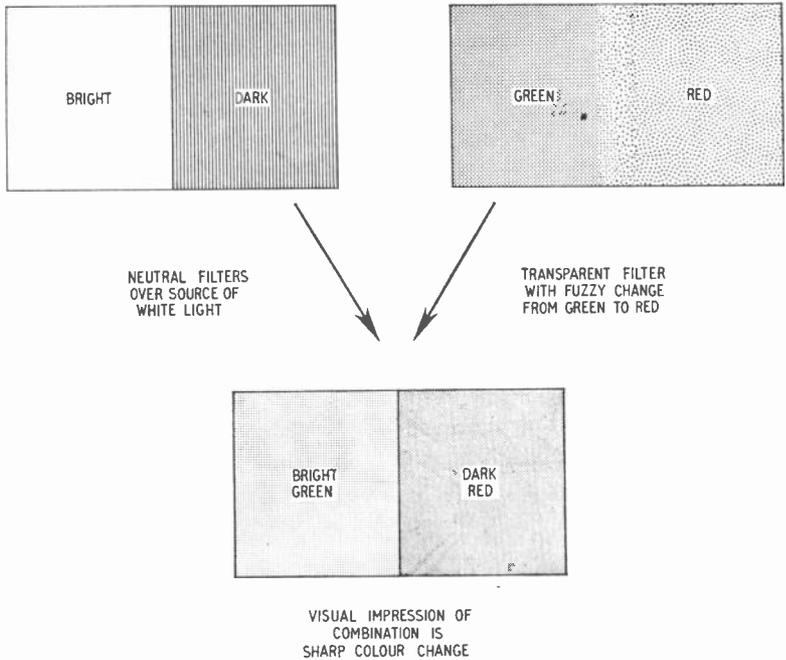
Fig. 1.4.   *Sharp brightness change superimposed on an*
unsharp colour change

see a blue colour at the outside, then green near the middle, and
finally yellow and brownish red near the centre.   Reversing the
direction of rotation causes blue to appear at the inside and red at
the outside.   Some people describe the colours as bright, others
have difficulty in discerning them at all.

The sequence of on-off waveforms needed for various colours can
be deduced from Fig. 1.5 (a), but the necessity for such a slow
repetition rate as 8 rev/s, and the paleness of the colours, makes
the phenomena unsuited to a television system.   Attempts have

been made to use the effect in televised advertisements. Two of the other patterns which have been used are shown in Fig. 1.5 (b)— Fechner's pattern—and 1.5 (c)—Rood's pattern.

The colours are generally explained on the assumption that the three supposed types of red, green and blue receptors in the eye have different build up and decay times when light is switched on and off, but there is some doubt about this explanation, for example, monochromatic light can be used to produce such effects. Similar effects occur when coloured light is shone on to Benham's rotating disc.

Rapid movement of a mixture of colours across the field of view will sometimes show up the individual component colours, whereas under practically all other conditions the eye consistently integrates the mixture into one sensation.

1.16. Some relevant monochrome revision

It is assumed that the reader of this book has a broad under-standing of monochrome television. Nevertheless, since colour television places emphasis on one or two aspects of black and white television to which little attention is normally paid, the next sections are devoted to topics which the reader may wish to revise.

1.17. Field

In British monochrome television the word "*frame*" means the $202\frac{1}{2}$ lines which constitute half a complete interlaced picture of 405
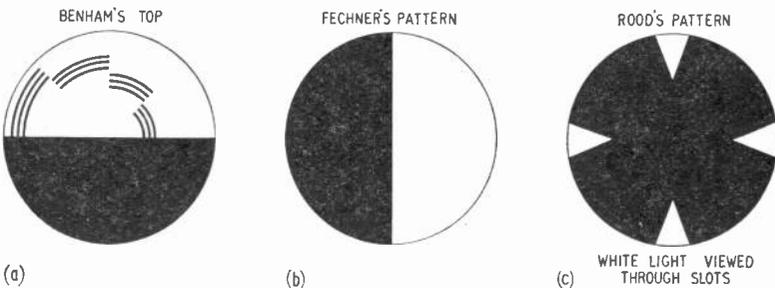
Fig. 1.5. Production of subjective colours. Each disc should be rotated at just below grey fusion frequency which is about 8 c/s

lines. In American nomenclature such a half picture of $202\frac{1}{2}$ lines would be called a field, and they use the word "frame" in the cinematograph sense of a complete picture. To avoid confusion it is becoming customary for both nations to drop the use of "frame",
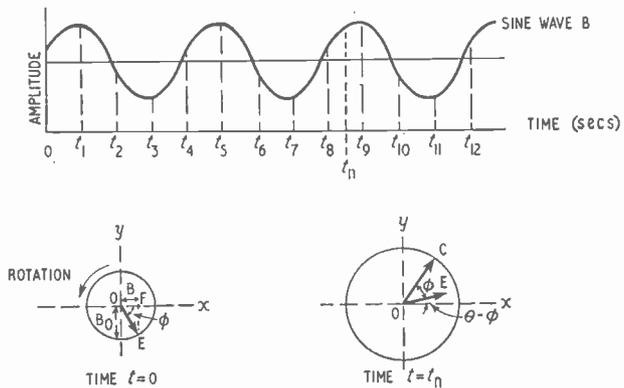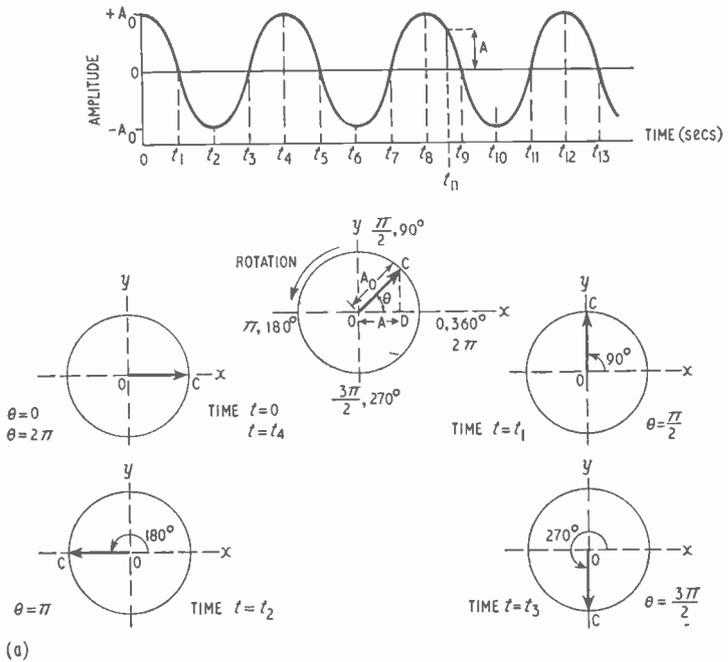
Fig. 1.6.   Phase for a sine wave oscillation

and to refer to $202\frac{1}{2}$ lines as one field, and to two interlaced fields, 405 lines, as a picture.

1.18. Phase

A sine wave such as that shown in Fig. 1.6 (a) may be considered as being generated by an anticlockwise rotating line OC, called a *phasor*, when the projection of OC on the y axis represents, at any given time, the amplitude A of the sine wave, and the projection on the x axis represents the amplitude of the cosine wave. For the triangle ODC, by definition

$$\sin \theta = \frac{DC}{OC} \qquad \cos \theta = \frac{OD}{OC} \qquad \tan \theta = \frac{DC}{OD}$$

so that $A = OC \cos \theta = A_0 \cos \theta$

If the phasor OC rotates at a frequency f cycles a second, it rotates through $360° = 2\pi$ radians, every cycle, or

$$\theta = 2\pi f t = \omega t$$

where t is the time in seconds from the phasor starting from the position shown for $t = O$. The time might have been chosen from any other position, of course.

The phase of the line OC is measured by the angle θ with respect to the x axis.   Phase is always relative, and it is necessary to specify the position from which the angle θ is being measured.   The diagrams of Fig. 1.6 (a) show the positions of OC for phases of O, $\frac{\pi}{2}$, π, $\frac{3\pi}{2}$ and 2π radians from the x axis, 2π radians repeating the original position at time $t = O$.

Fig. 1.6 (b) shows a second sine wave of the same frequency f which goes through its maximum and minimum amplitudes at a different time from that shown in Fig. 1.6 (a).   At time $t = O$ the phasor OE has a phase $-\Phi$ with respect to the x axis, so that at time t_n, when OC is at a phase angle θ, OE is at a phase angle $\theta - \Phi$ with respect to the x axis.

The sine wave B goes through its maximum value at a time $\dfrac{\Phi}{\omega}$ seconds after sine wave A has gone through its maximum, so that sine wave B may be derived from A by delaying it and, if B is smaller than A, reducing its amplitude.   Sine wave B is said to have a phase lag of Φ with respect to A, and sine wave A has a phase advance

of Φ with respect to B.   Similarly, sine wave A has a phase difference of $+\Phi$ radians with respect to B, and B has a phase of $-\Phi$ radians with respect to A.   Phase has no meaning unless it is referred to some kind of standard.   Since sine waves A and B have the same frequency they have a constant phase difference, Φ, which is related to the time delay between the two waves by

$$\Phi = \omega t_d = 2\pi f t_d$$

where f is the frequency in cycles per second, ω is the angular frequency in radians per second, and t_d is the time delay in seconds.

A curve of the phase shift suffered by a sine wave in transmission through a receiver channel may be plotted for various frequencies,
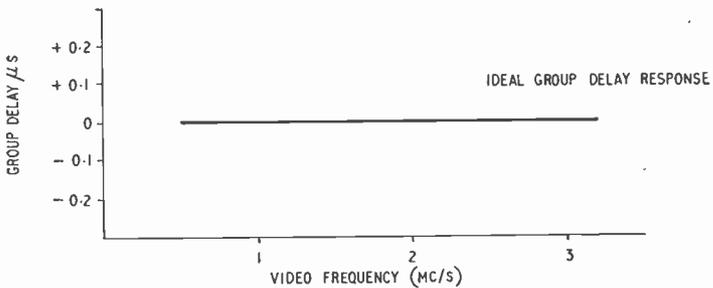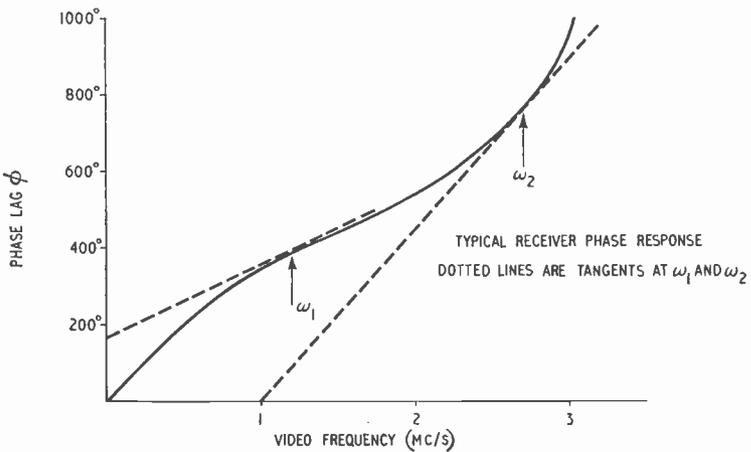
Fig. 1.7.   Phase response curves

D.C. LEVEL A_0

FUNDAMENTAL f_1

SECOND HARMONIC
$f_2 = 2f_1$

THIRD HARMONIC
$f_3 = 3f_1$
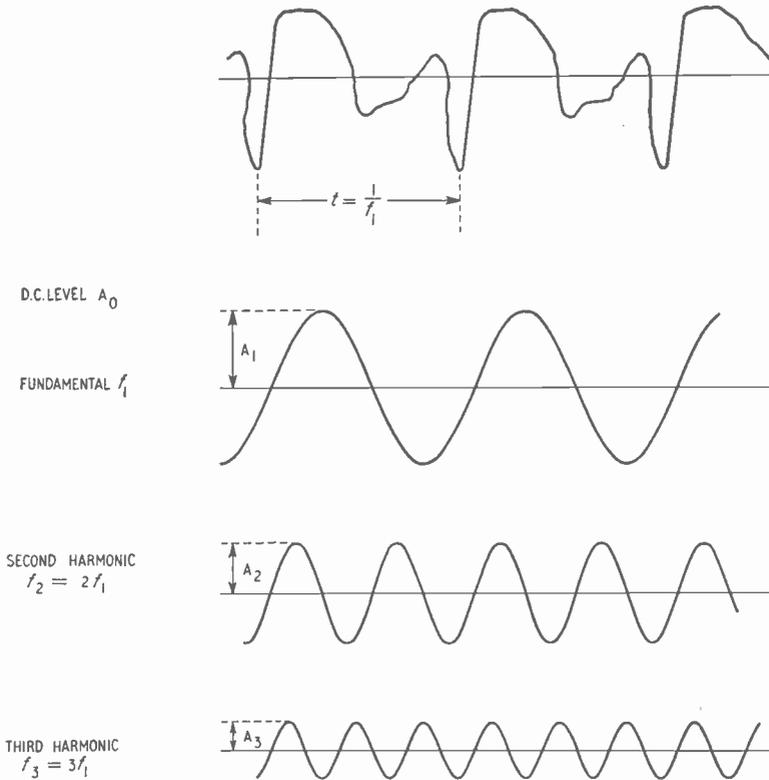
Fig. 1.8.   Fourier sine wave components of a repetitive waveform

in the same way as the amplitude response of the channel.   Some-
times the slope of the phase response curve is plotted rather than the
phase curve itself (*see* Fig. 1.7).   The slope of the curve at any fre-
quency is given, of course, by the tangent to the curve at that point.
This function is called the *group delay*, or *envelope delay* response
and it shows up variations in the linearity of the phase curve.   For a
linearly increasing phase response the group delay curve is a hori-
zontal straight line; that is to say, the group delay is a constant with
frequency, or

$$\text{Group delay} = \frac{d\Phi}{d\omega} = \text{constant}$$

The group delay at an angular frequency ω is the quantity which
matters when we are considering a carrier at frequency ω together

with the group of sidebands on either side of it. The group delay response determines the shape of the modulation waveform, the carrier envelope, and often the absolute value of the delay is immaterial.

1.19. Fourier components

Any repetitive waveform, such as in Fig. 1.8, can be considered as made up of a series of sine waves, called the *Fourier components* of the waveform. The frequencies of these sine waves are multiples of the repetition frequency of the waveform, and the amplitudes and relative phases of the sine waves depend on the shape and amplitude of the waveform. If at a particular value of time t, the instantaneous amplitudes of each of the components are summed the corresponding instantaneous amplitude of the original waveform is obtained. In general the amplitude of each component decreases as the frequency increases. Abrupt transitions in the waveform give rise to very high frequency Fourier components, and conversely, any network which will not pass the very high frequency components will not transmit an abrupt waveform change but will increase the time taken for the fastest waveform transition.

1.20. Distortionless transmission

For an electrical waveform to be transmitted without distortion, all the Fourier components must be transmitted with their amplitudes increased or decreased in the same ratio, and each component must suffer the same time delay as the others. The output waveform then has the same shape as the input waveform. Since the time delay must be constant, $\dfrac{\Phi}{\omega}$ must be constant with frequency, so that the phase shift Φ must increase linearly with frequency, as in Fig. 1.9. In this instance phase shift refers to the phase of the Fourier component after it has passed through the network compared with its phase before it entered the network.

For the type of circuits normally used in television receivers, called *minimum-phase circuits*, the amplitude response determines the phase response; attention is usually concentrated on the amplitude response since this is easier to measure and interpret. If the amplitude response is flat at all frequencies, the phase response is linear as required. Any changes in the amplitude level produce corresponding departures from linearity in the phase response but these are spread over a much wider frequency band. Any bandwidth restrictions produce phase distortion over frequencies at and

adjacent to the ends of the bands, with a corresponding distortion in the transmitted waveform.

1.21. Frequency spectra

If the transmitted picture is a stationary scene or pattern, every complete 2 field picture is the same as the other pictures, and the video waveform repeats itself every $\frac{1}{25}$ of a second. The video waveform must therefore consist of harmonics of 25 c/s, and there should be no frequencies in between these harmonics.

The amplitudes of the numerous harmonics of 25 c/s which make up the frequency spectrum of a video waveform depend upon the
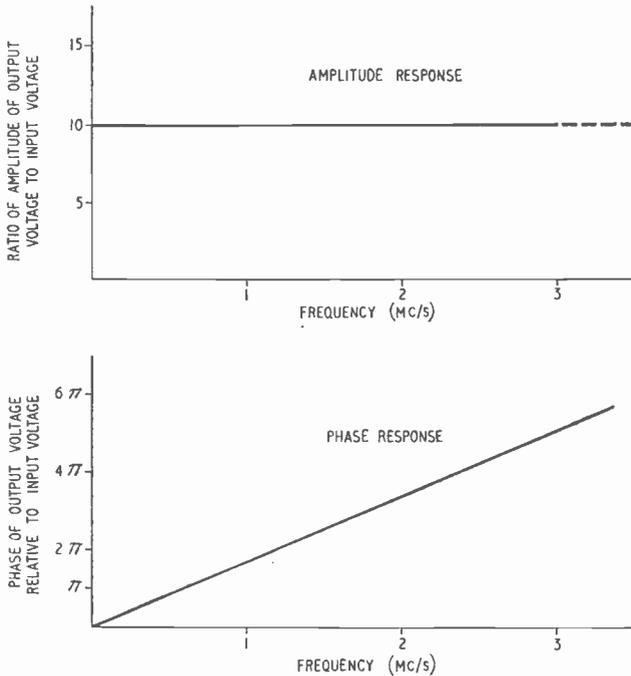
Fig. 1.9. Frequency characteristics for a distortionless transmission

particular scene being transmitted. Usually there is a marked similarity between the waveform of any particular line and that of the adjacent lines. If each line were an exact repetition of its neigh-bours, and if there were no frame blanking signals, then the complete video waveform would be composed only of harmonics of the line frequency, 10 kc/s, 20 kc/s, 30 kc/s etc., or f_L, $2f_L$, $3f_L$, etc., where $f_L = 10 \cdot 125$ kc/s. This does not happen, of course, but there is

3
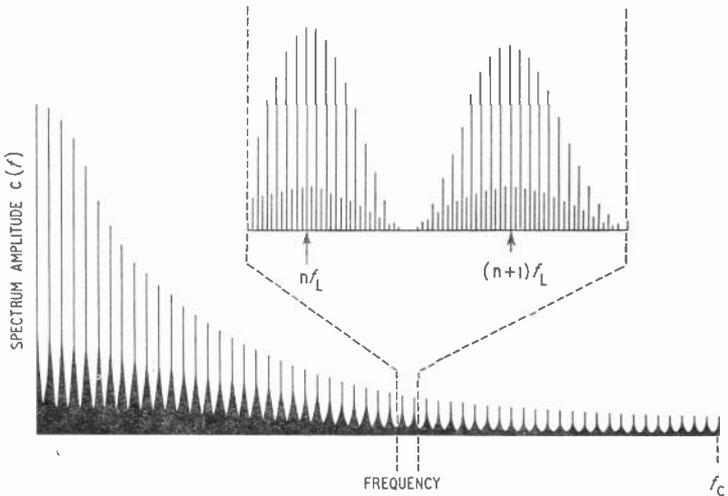
Fig. 1.10. *Frequency spectrum for a typical stationary scene (Reproduced from "Television Engineering Handbook" (Editor D. J. Fink), by D. C. Livingstone. Copyright 1957, McGraw Hill Book Company, Inc.)*

sufficient correlation between line signals for those 25 c/s harmonics which occur at and close to the 10 kc/s harmonics to be larger than those which occur half-way between these harmonics. Half-way between the 10 kc/s harmonics is described as being at the odd multiples of half the line frequency, at $\frac{1}{2}f_L$, $1\frac{1}{2}f_L$, $2\frac{1}{2}f_L$, etc., or at $(2n + 1)$ $(\frac{1}{2}f_L)$ where $n = 0, 1, 2, 3$, etc.

Measurements on actual television waveforms give frequency spectra similar to the one shown in Fig. 1.10. The only 25 c/s
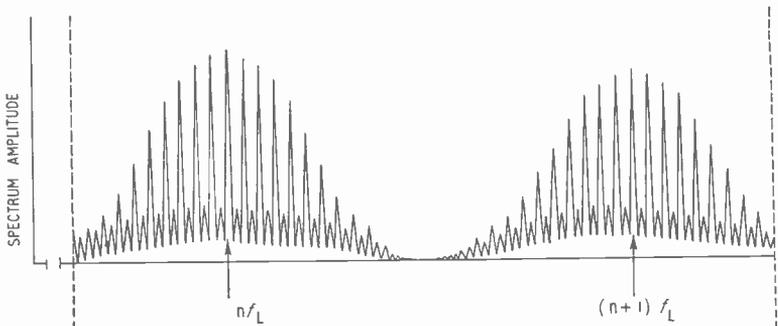
Fig. 1.11. *Frequency spectrum for a picture which is changing (Reproduced from "Television Engineering Handbook" (Editor D. J. Fink), by D. C. Livingstone. Copyright 1957, McGraw Hill Book Company, Inc.)*

harmonics which have any appreciable amplitude are those close to the 10 kc/s harmonics, and there are gaps between these 10 kc/s harmonic groups where there is very little energy.

Because the two fields of which a picture is composed are very similar in character, there is a certain amount of repetition in the waveform from one $\frac{1}{50}$ second to the next, with the result that those 25 c/s harmonics which are also 50 c/s harmonics are larger than the others.

If the transmitted picture is changing then other frequencies are introduced into the waveform, and the spectrum becomes less discrete, as in Fig. 1.11; however, the general character of the spectrum is unaltered.

1.22. Summary

The colour camera analyses the image of the scene into its red, green and blue components. The three electrical signals which arise from these primary images are then encoded into a form suitable for transmission. In the N.T.S.C. system the transmitted signals are a brightness signal similar to the one used for mono-chrome television, and a sub-carrier which is modulated with the colour information.

Monochrome receivers operate satisfactorily on the brightness signal and ignore the sub-carrier, although the interference pattern which the latter produces is noticeable on close inspection.

Colour receivers decode the transmitted signals into the original red, green and blue voltages and generate red, green and blue images which are combined into one full colour picture.

Since the eye is not a perfect recording instrument, bandwidth can be saved by not transmitting those picture variations which the eye would not be able to see from a normal viewing distance.

Colour Measurement

2.1. Introduction

A television system is not capable of transmitting colour as colour, it can only transmit electrical signals proportional to certain characteristics of colour which can be measured. Fortunately, the measurement of colour has been the subject of considerable scientific study and it has been given the name *colorimetry*. This chapter outlines the basic principles of measuring colour.

White light when split by passing it at a small angle through a prism, separates out into a band of rainbow coloured lights called a *spectrum*. By making a mixture of two or more of these primary colours all shades of colour can be reproduced. White light can be imitated by mixing primary red, green and blue, as can all shades of colours, except for a few very pure hues. Similarly, each shade can be identified by finding the red, green and blue components which when mixed together will produce the same sensation in the eye as the original colour. In the colour television receiver, three primary colours, red, green and blue are varied in strength to produce the fully coloured scene.

These ideas must be elaborated in some detail in order to appreciate how the colour television camera may analyse a colour and how the receiver may then reproduce the same colour with some precision. In particular, a very useful, if rather theoretical way of plotting colours on a diagram will be developed.

2.2. Characteristics of light

Before a colour can be measured quantitatively, the variable qualities of colour have to be defined. Various characteristics can be used but three qualities are always necessary to specify a colour exactly. Different terms are used to describe the subjective sensation caused by a particular quality of colour and the corresponding objective measurement of that quality.

2.2.1. BRIGHTNESS

Brightness is the quality of a colour which is concerned with the " power " of the colour and is best explained by an example. If

the light from a 500 watt lamp passes through a piece of red glass, then the resultant colour is brighter than the dull red colour given when the same piece of glass is placed in front of a 15 watt lamp. The sensation of brightness is a variable factor and has to be specified in order to define a colour accurately. It will be shown in this chapter that brightness has a slightly more accurate meaning if it is described as *luminance*.

2.2.2. HUE

This is the most noticeable quality of colour. The light from a 500 watt lamp passing through a red glass will give a colour which has a red hue, but when the light is passed through a green glass the resultant colour will have a green hue. Colours of different hue may or may not have the same brightness, so the quality of hue is a second independent variable. Later it will be shown that the hue of a colour depends on the *dominant wavelength* of the light.

2.2.3. SATURATION

The third quality of colour, saturation, is the variation in depth of colour of the same hue and it must not be confused with brightness which is a measure of the power of the light itself. If the light from a 500 watt lamp passes through a red glass and is allowed to fall on to a sheet of white paper, then the colour reflected from the paper is a clear bright red. But if at the same time white light is also allowed to fall on to the paper the reflected colour, although still of the same hue, i.e., red, appears to the eye as a paler shade of red and is said to be pink. The more white light there is falling on the paper, the paler or less saturated the reflected colour becomes. Corresponding to the sensation of saturation, an objective term *purity* is defined, which can be measured precisely.

2.3. The nature of light

Light consists of electromagnetic waves and has the same nature as other electromagnetic waves such as radio waves, X-rays, gamma rays and heat rays. The difference between these various radiations is one of wavelength as shown in Fig. 2.1. The eye is sensitive to radiations which have a wavelength in the range 0·000015 to 0·00003 in., i.e., the eye responds to wavelengths from 380 to 780 mμ. Frequency instead of wavelength is often used for radio waves and light waves can be similarly described

$$\lambda f = c$$

where λ is the wavelength in metres, f is the frequency in c/s and

c is a constant equal to the velocity of light, i.e., 3×10^8 m/s. So visible waves correspond to a frequency of about 10^9 Mc/s.

The eye and the brain together recognize each wavelength as a distinct hue (*see* Fig. 2.1) so that light of wavelength 700 mμ creates the sensation of red and light of wavelength 510 mμ looks green.
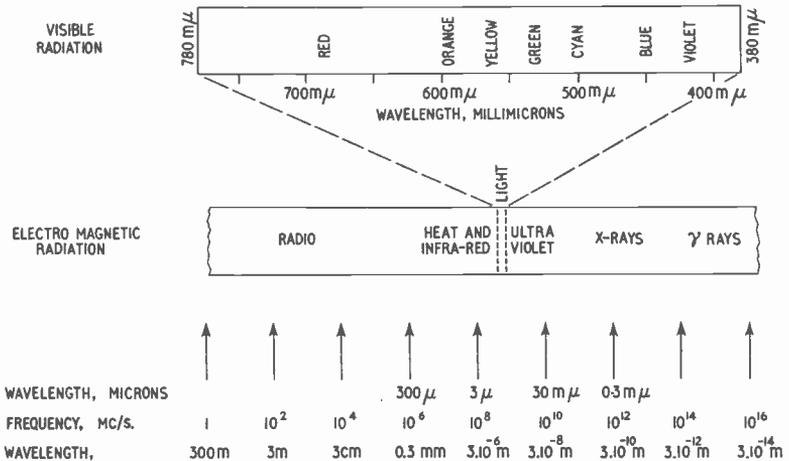
Fig. 2.1. The spectrum of electromagnetic radiation

The colours mentioned in Fig. 2.1 are called *spectrum colours* and these merge into each other so that, for example, light of wavelengths 480 mμ and 470 mμ will have a slightly different blue hue.

Light consisting of only one wavelength or of a group of wavelengths up to a bandwidth of about 5 mμ is said to be 100% saturated or is called *monochromatic*. A sodium discharge lamp emits light of this type and the energy distribution graph (Fig. 2.2) shows how all the energy is concentrated at one wavelength. The line (1) represents the energy curve of a signal lamp light and shows how the energy, though not concentrated in only one wavelength, is distributed in the red end of the spectrum.

2.4. Non-spectrum colours

Certain colours, such as black and white, are not present in the spectrum and since these colours are present in a fully coloured scene they have to be considered separately.

2.4.1. WHITE LIGHT

White light consists of a mixture of wavelengths. Exactly equal energy at every wavelength gives a white referred to as *equal energy white*. Fig. 2.2 shows how certain lights, though approximating to white light, vary in energy distribution. The examples given are the light from a north sky, sunlight, and the light from a gas-filled tungsten lamp.

The sensation of white can be produced by various simple mixtures. Thus blue and yellow lights mixed in the right proportion look white, as does an appropriate mixture of green and purple. Pairs of colours which produce white light are called *complementaries*, so that green is the complementary colour to purple, and purple is the complementary colour to green. The sensation of white can
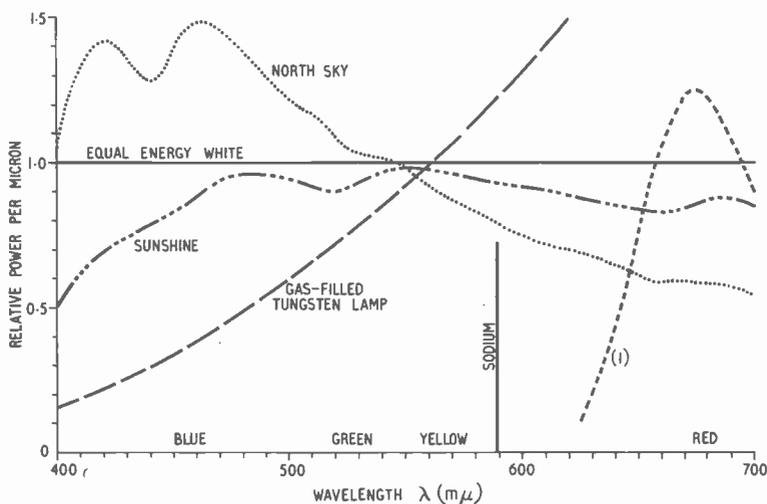
Fig. 2.2. Spectral energy distribution of lights

also be produced by mixing amounts of the red, green and blue wavelengths.

2.4.2. BLACK

Black is the absence of light of all wavelengths. But the range of brightness which the eye can register at the same time is limited (*see* Section 1.11), so that if there is a marked difference between bright and dull areas on a television screen, for example a white

LIGHT
'A'
LOOKS
YELLOW

(a)

LIGHT
'A'
LOOKS
BROWN

(b)
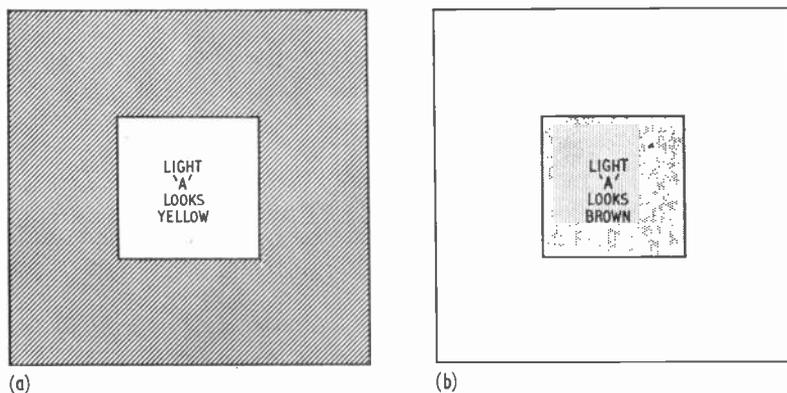
*Fig. 2.3. The sensation of brown. (a) has a black surround
and (b) has a bright surround*

patch near a dark grey area, the dull areas will appear black to the eye because of the limitation of its range.

2.4.3. PURPLE

The spectrum in Fig. 2.1 contains no mention of purple colours. Such colours are called *magenta* in colorimetry and they are produced by lights which consist of a mixture of red and blue wavelengths. These colours are not shown in the spectrum because the red and blue shades are at the extremes of the visible wavelength band and so do not merge into each other.

2.4.4. BROWN

Just as the eye is deceived into recording a dull area with a bright surrounding as black (*see* Section 2.4.2) it is also deceived into recording a dim area of yellow or orange light with a bright surrounding as brown. Brown light is therefore not a true colour but is the effect of seeing dim yellow or orange light in contrast with a brighter area (*see* Fig. 2.3).

2.5. Colours of objects

An object appears coloured by virtue of the light which is reflected from it. A white object lit by white light looks white, but lit by red light, it looks red. A red object absorbs light of all colours except red, and reflects the red light.

Paints produce the effect of colour by absorbing some wavelengths and reflecting others. This is fundamentally different from adding one coloured light to another coloured light. Paints sub-

tract colour from the incident light and we see what is left. This method of producing colour sensation is called the *subtractive process*. Adding lights together, as in colour television, is called an *additive process*. To be more precise, while photographic colour transparencies are purely subtractive, paints do produce a slight additive effect as well as a subtractive one.

2.6. Measurement of brightness

The eye is not equally sensitive to light at different wavelengths. For example, a yellow-green light of wavelength 555 mμ will appear brighter to the eye than an orange light of wavelength 600 mμ, when both lights have the same energy level.

The eye is also deceived by the conditions under which it views a light source. A light source seen in a darkened room appears to be much brighter than when it is seen in sunshine.

Although the intensity of a light could be measured simply by the energy per unit time at each wavelength, this measure of radiation would not bear a simple relation to the sensation of brightness. So a different method of measuring the brightness of a light source has been devised which takes into account the wavelength response of the eye, whilst still providing an objective measurement of luminous intensity.
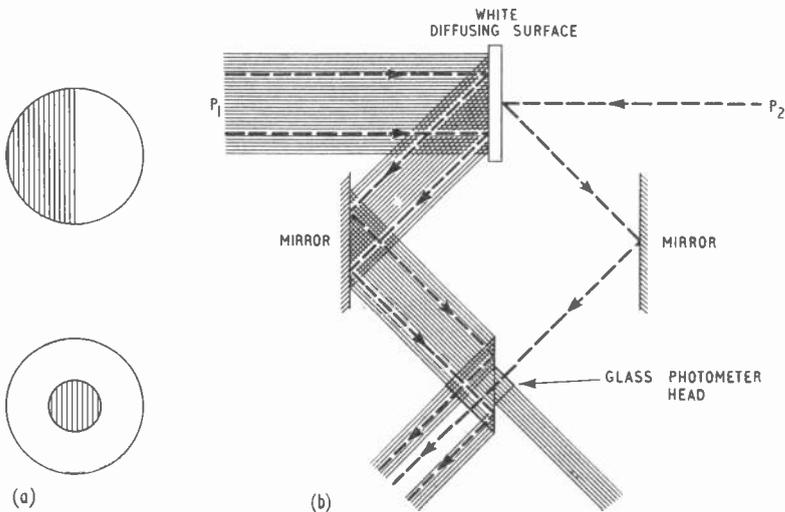
Fig. 2.4. Photometer. (a) shows two types of visual field and (b) typical construction

2.6.1. RELATIVE LUMINOSITY

If the viewing conditions were standardized, it would be possible to measure the relative brightness or *luminosity* of two colours. By using an instrument called a *photometer*, the principle of which is shown in Fig. 2.4, it is possible to view two adjacent patches of coloured light against a black background. The radiant energy of each light source can be adjusted to make the two areas appear equally bright.

If the first colour has a power P_1 watts and the second colour a power P_2 watts

$$\text{Relative luminosity of } \frac{\text{colour 1}}{\text{colour 2}} = \frac{P_2}{P_1}$$

Since light at a wavelength of 555 mμ appears brighter per watt than at any other wavelength, it is used as the reference wavelength to which all other wavelengths are compared and a curve of the relative luminosity of every wavelength is shown in Fig. 2.5. This
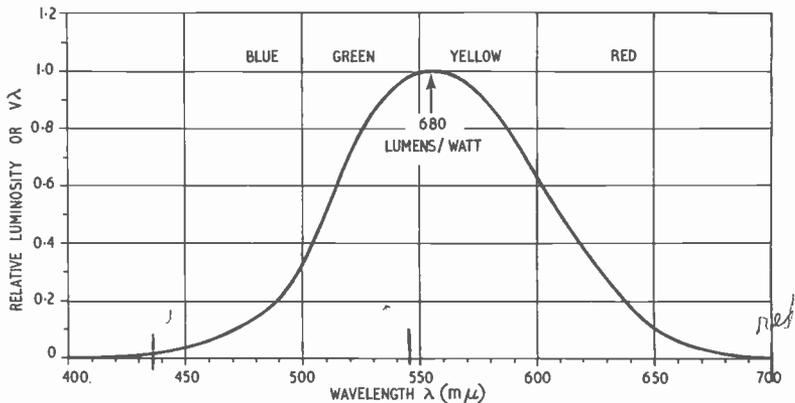
Fig. 2.5. The curve of relative luminous efficiency (Vλ) for monochromatic radiation

curve varies from one human to another, so an average curve called the *standard observer's luminosity function* has been fixed by international agreement.

This curve is valid for photopic vision (i.e., when the brightness of a scene is great enough for the eye to distinguish colour) but for scotopic vision (i.e., when the level of brightness in a scene is so low that the eye can only register different shades of black and grey)

the whole curve moves bodily down to wavelengths which are 40 mμ shorter.

2.6.2. THE LUMEN

Luminous flux is the name given to the total quantity of light emitted in unit time, when measured by its ability to produce the sensation of brightness; it is measured in terms of a unit called the *lumen.* At 555 mμ wavelength, a watt of light power corresponds to 680 lumens of luminous flux, whereas at other wavelengths a watt of light gives correspondingly less sensation and (*see* Fig. 2.5) gives 340 lumens at 510 mμ.

In order to define a unit of luminous flux, a theoretical standard source has to be defined and this has been fixed by international agreement.

This standard source is a point source which radiates evenly in every direction so that if the point were the centre of a sphere, an even amount of radiation would fall on each part of the sphere's surface. The power of the source is equal to $\frac{1}{60}$ of the total radiated power from one square centimetre of a perfectly black body at 2042°K.

A sphere can be divided into 4π unit solid angles or steradians. A solid angle is the angle subtended by a spherical surface at its centre, and is measured by the ratio of the area of the surface to the square of its radius.

The lumen is the luminous flux emitted from a standard source down a unit solid angle.

2.6.3. UNIT FOR AN EMITTING SURFACE

The objective measurement of the brightness of an emitting surface is called its photometric brightness, or *luminance.* There are three variables which must be taken into account when defining luminance and these are the type of surface, its area and the amount of light being emitted. The *foot lambert* is the unit of luminance and is defined as the luminance of a perfectly diffusing surface emitting one lumen per square foot, where a perfectly diffusing surface is one which looks equally bright from whatever angle the surface is viewed. Most direct-viewed cathode-ray tubes satisfy this requirement. The maximum luminance of a monochrome television screen is usually in the region of 50 to 80 foot lamberts.

The overall luminance of a mixture of lights of several wavelengths is the sum of the individual luminances. If the light has a continuous spectrum as in Fig. 2.2, then the total luminance is the

sum of each small group of wavelengths $\delta\lambda$. If the spectrum distribution is given in terms of energy then the energy at each wavelength must be corrected by the luminosity factor (*see* Fig. 2.5) to take into account the variations in reception of the eye. This correction can be denoted by $\bar{y}$.

The total luminance can then be written

$$L = \int B_\lambda \, \mathrm{d}\lambda$$

where B_λ is the mean luminance over the wavelength range λ to $\lambda + \delta\lambda$, and the integration need only be between the limits of 380 and 780 mμ. Then also

$$L = K \int E_\lambda \bar{y} \, \mathrm{d}\lambda$$

where E_λ is the energy at the wavelength λ and K is the maximum luminous efficiency of radiation, or 680 lumens per watt.

If this light source shines on an object which reflects differing amounts of light at different wavelengths, the total luminance of the object will be

$$L_r = K \int R_\lambda E_\lambda \bar{y} \, \mathrm{d}\lambda$$

where R_λ is the proportion of the incident light reflected at the wavelength λ.

2.7 Colorimetry

It is possible to specify any particular colour by its spectral distribution of energy, as in Fig. 2.2, and to reproduce the same colour by generating light at a great number of wavelengths throughout the spectrum to simulate the original spectral distribution. Fortunately for television engineers, a simpler method is possible.

Most colours can be matched visually by mixing together only three coloured lights. These three primary colours, or *reference stimuli* as they are called, are chosen in the red, green and blue regions of the spectrum since it is found that this enables the widest range of other colours to be matched. The instrument used to carry out such visual matching is called a *colorimeter*. The observer looks into an eyepiece and sees the colour to be matched in one half of a split circle as in the photometer (Fig. 2.4) and in the other half sees the mixture of reference stimuli, that is, a mixture of

red, green and blue lights. The amounts of these lights can be adjusted until the mixture matches the test colour.

The red light can be described as an R* light, the green as a G* light, etc., where * indicates that the particular colour is being discussed without specifying its luminance. Precision colorimeters use as reference stimuli a narrow band of spectrum wavelengths, the international standards being 700, 546 and 436 mμ. The controls which vary the amounts of red, green and blue light in the mixture are calibrated in arbitrary units. If an unknown colour C* has been matched by u units of R* plus v units of G* and w units of B*, then the result is written

$$cC^* \equiv uR^* + vG^* + wB^*$$

where $\equiv$ means, in this context, that the colours match.

C*, R*, G* and B* have no numerical significance in this equation but merely indicate the kind of light being used, whilst c indicates the amount of C* being matched.

2.7.1. GRASSMAN'S LAW

It is found experimentally that if

$$c_1C_1^* \equiv u_1R^* + v_1G^* + w_1B^*$$

and if also

$$c_2C_2^* \equiv u_2R^* + v_2G^* + w_2B^*$$

then the mixture of $c_1C_1^*$ and $c_2C_2^*$ is matched by

$$c_1C_1^* + c_2C_2^* \equiv (u_1 + u_2)R^* + (v_1 + v_2)G^* + (w_1 + w_2)B^*$$

In fact the equations representing such colour matches obey the simple algebraic rules of addition and subtraction.

The numbers c, u, v and w could be expressed in terms of watts per square foot, but this leads to large differences in the numerical values of v and w and even if luminance units such as foot lamberts are used, the mean luminance of typical primary colours may differ by a factor of twenty.

2.8. Tristimulus values

To make the numbers involved approximately equal, the units of R*, G* and B* are specified in terms of an agreed reference colour, such as the equal energy white. This agreed reference colour is

matched in the colorimeter and the scales are adjusted so that all three read unity for this match

$$wW^* \equiv 1R^* + 1G^* + 1B^*$$

$$\therefore \qquad w = 1 + 1 + 1 = 3$$

Using these arbitrary units, the match to colour C* may be

$$CC^* \equiv RR^* + GG^* + BB^*$$

where

$$C = R + G + B$$

The numerical quantities R, G and B, are called the tristimulus values of the colour CC^*. Tristimulus values are a complete specification only if the primary colours R*, G* and B*, and the reference colour W* are known, and if w, or the luminance of the W* reference light, is also given.

2.9. Trichromatic units

To overcome the variations due to observers using the same reference light at different luminances, it is usual to divorce from the equations the amount of light used, by dividing throughout by the total amount, which is $R + G + B$,

when

$$\frac{C}{R+G+B}C^* \equiv \frac{R}{R+G+B}R^* + \frac{G}{R+G+B}G^* + \frac{B}{R+G+B}B^*$$

$$1\,C^* \equiv \frac{R}{R+G+B}R^* + \frac{G}{R+G+B}G^* + \frac{B}{R+G+B}B^*$$

$$\equiv rR^* + gG^* + bB^*$$

where

$$r = \frac{R}{R+G+B} \qquad g = \frac{G}{R+G+B} \qquad b = \frac{B}{R+G+B}$$

so that

$$r + g + b = 1$$

The amount of C* in this equation is then said to be one trichromatic unit, abbreviated to one T-unit.

2.10.   Luminosity coefficients

There are two ways of expressing a match to 1 foot lambert of the reference white W*.   Using foot lamberts as units:

$$1W^* \equiv l R^* + m G^* + n B^*$$

from the V_λ curve (Fig. 2.5), it is clear that l, m and n are by no means equal.   Alternatively, trichromatic units may be used, when by definition:

$$3W^* \equiv 1R^* + 1G^* + 1B^*$$

or

$$1W^* \equiv \tfrac{1}{3}R^* + \tfrac{1}{3}G^* + \tfrac{1}{3}B^*$$

so that for this particular reference white

$$3 \text{ T-units of white} \equiv 1 \text{ foot lambert}$$
$$1 \text{ T-unit of red} \quad \equiv l \text{ foot lamberts}$$
$$1 \text{ T-unit of blue} \quad \equiv m \text{ foot lamberts}$$
$$1 \text{ T-unit of green} \equiv n \text{ foot lamberts}$$

In general, for

$$1C^* \equiv r R^* + g G^* + b B^*$$

in T-units, the luminance of 1 T-unit of C* is given by

$$rl + gm + bn \text{ foot lamberts}$$

l, m, n, are called *luminosity coefficients* since they give the relative luminance of equal T-units of the red, green and blue primaries.

2.11.   Chromaticity co-ordinates

In converting to T-units any indication of the absolute magnitude of the luminance of the colour is lost.   The remaining information must relate in some way to the hue and saturation only, and is called the *chromaticity* of the colour.

r, g and b, are known as *chromaticity co-ordinates* since they carry the chromaticity information.   As

$$r + g + b = 1$$

for any colour, any two chromaticity co-ordinates are enough to determine the chromaticity.   Fig. 2.6 shows a chromaticity

diagram on which the chromaticity of colours can be plotted, in this case in terms of r and b. When $r = 1$, then $g = b = 0$ so that $r = 1$, $b = 0$ must be the chromaticity of the red primary R^*. Similarly $r = 0$, $b = 1$ is the chromaticity of B^* and when $r = b = 0$, then $g = 1$ and the origin of the chromaticity diagram must

Fig. 2.6.   A typical chromaticity diagram
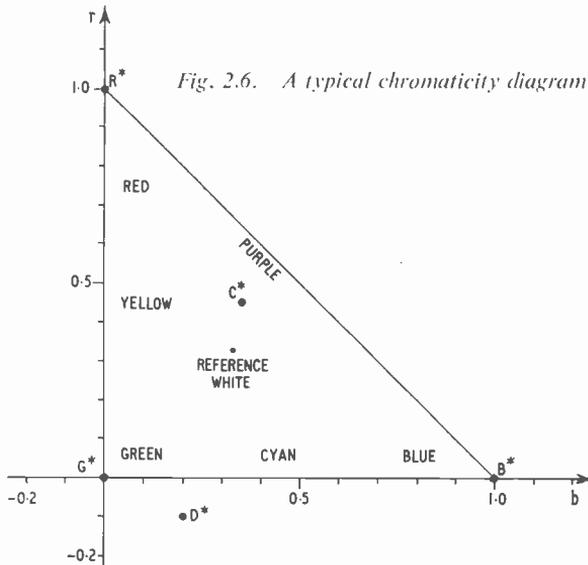
represent the green reference stimulus G^*.   A point such as C^* represents a colour with chromaticity $r = 0.45$, $b = 0.35$ and hence $g = 0.2$.   This colour C^* can be written as

$$1 \cdot C^* = 0.45R^* + 0.2G^* + 0.35B^*$$

and represents a desaturated purple colour.

Notice that since $r + b$ cannot be greater than 1, chromaticities measured in terms of R^*, G^* and B^* must lie inside the triangle R^*, G^*, B^*, providing r, g and b are positive.   But some very saturated colours cannot be matched in the colorimeter with the particular primaries chosen, notably the blue-green spectrum colours.   In such cases it is always possible to obtain a match by adding a particular amount of one of the reference stimuli to the sample colour in one half of the colorimeter field of view, and

matching this combination with a mixture of the remaining two reference stimuli in the other half.

$$DD^* + RR^* \equiv GG^* + BB^*$$

which can be written as

$$DD^* \equiv -RR^* + GG^* + BB^*$$

whence

$$r = \frac{-R}{-R + G + B}$$

The colour D* then has a chromaticity which lies outside the colour triangle, and D* cannot be reproduced in its full saturation by any real mixture of R*, G* and B*.

2.12. Maxwell colour triangle

It is not necessary to use the rectangular co-ordinates of Fig. 2.6 for plotting chromaticities and R*, G* and B* can be located at
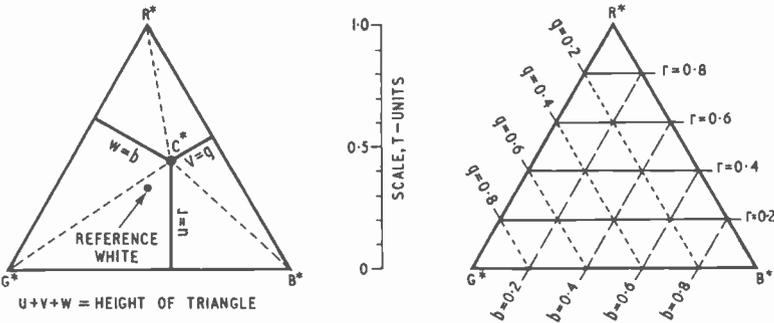
Fig. 2.7.   The Maxwell colour triangle

the vertices of an equilateral triangle, as in Fig. 2.7. The chromaticity of a colour C* is then given by the lengths of the perpendiculars to the sides of the triangle, u, v and w. Since the sum of the areas of the small triangles R*G*C*, R*B*C* and B*C*G* is always equal to the area of the main triangle R*G*B*,

$$u + v + w = \text{constant}$$

For a triangle R*G*B* of unit height, this constant is unity, and $u = r$, $v = g$ and $w = b$ are the chromaticity co-ordinates.   At the

4

reference white point, $r = g = b = \frac{1}{3}$, which is at the geometrical centroid of the triangle. Fig. 2.7 shows lines of constant values of r, g and b on such a triangle.

2.13. Chromaticity of colour mixtures

A colour formed by mixing r units of R* with g units of G* has a chromaticity which lies on the straight line joining R* to G*. The exact position of the resultant chromaticity depends on the relative amounts of R* and G*. If r and g are measured in T-units, then the chromaticity is at the point D*, Fig. 2.8,

where

$$\frac{R^*D^*}{D^*G^*} = \frac{g}{r}$$

with a total magnitude given in T-units of $r + g$. The reader may prove this for himself by calculating the chromaticity of the mixture in terms of the tristimulus values of the separate colours. This process of finding the chromaticity of a mixture is similar to finding the centre of gravity of two weights.

Similarly, mixtures of $(r + g)D^*$ with bB^* lie along the line D*B*, at a point such as C*,

where

$$\frac{D^*C^*}{C^*B^*} = \frac{b}{r + g}$$

Normally, data on the luminance of colours are required in foot lamberts or similar absolute units and not in T-units. The conversion is made by using the luminosity coefficients and knowing the luminance of the reference white used in the colorimeter.

2.14. General colour triangle

A more general chromaticity representation is when R*, G* and B* are at the vertices of any triangle as in Fig. 2.9. If the reference colour for which $r = g = b$ is given as the point Y*, which in general is not the centroid of the triangle, a simple geometrical construction will give the lines of constant r, g and b. The lines joining the vertices of the triangle to Y* meet the opposite sides of the triangle at the points S, T and U. The points T and U are points at which $r = \frac{1}{2}$, and the side G*B* of the triangle represents the line for which $r = 0$. The two straight lines TU and B*G* meet at a point O through which all the lines of constant r pass.
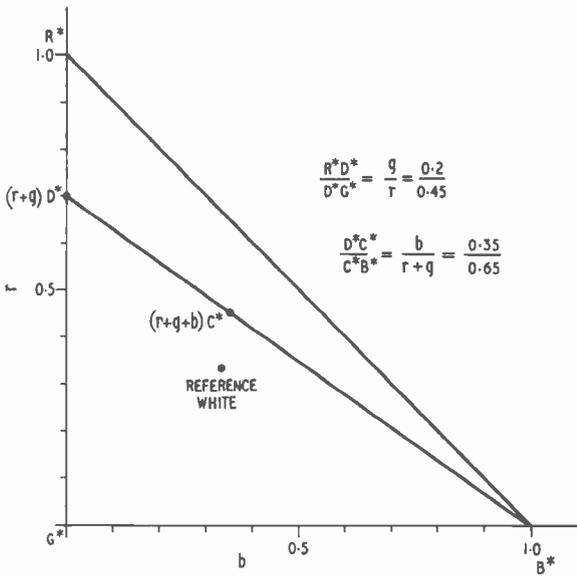
Fig. 2.8.   The chromaticity of colour mixtures
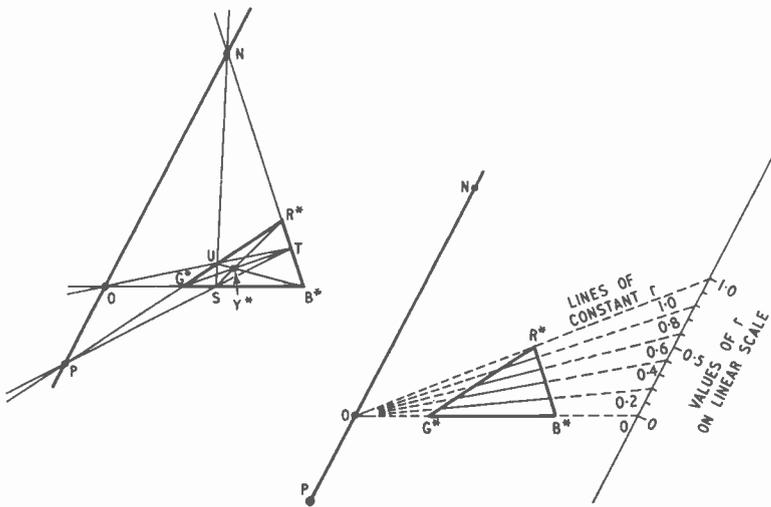
Fig. 2.9.   The general colour triangle
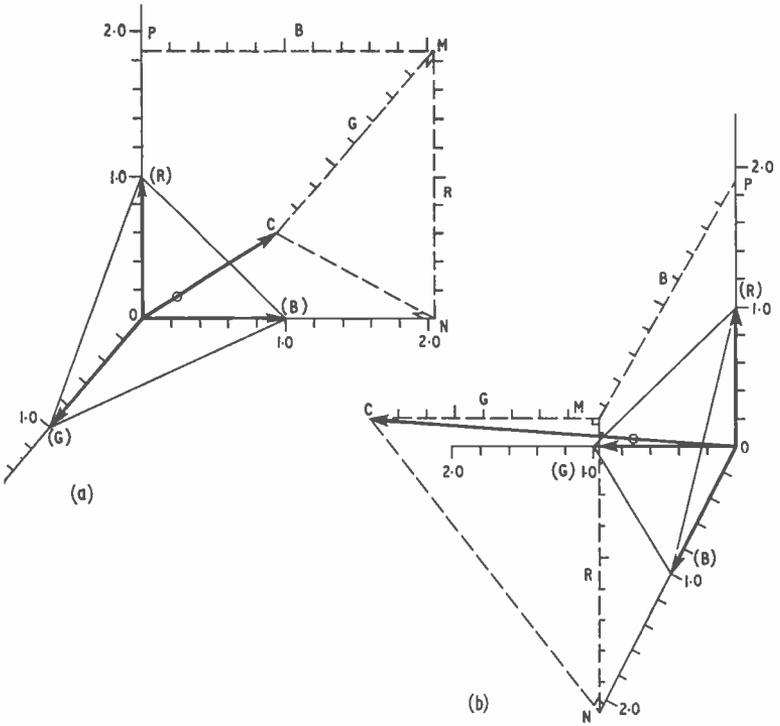
Fig. 2.10.   Two views of colour space

Similarly, *TS* meets R*G* at *P*, through which all the lines of constant *b* pass and *SU* meets B*R* at *N*, through which all the lines of constant *g* pass.   The points *O*, *P* and *N*, also lie on a straight line and any straight line parallel to *PN* has the useful property that the lines of constant *r* make linear intercepts on it, proportional to the value of *r*, and similarly for the lines of constant *g* and *b*.

2.15.   Colour space

By divorcing luminance from chromaticity, chromaticity can be plotted as a two-dimensional quantity.   It is quite practical to use three-dimensional space to plot all the three qualities of a colour, including luminance.

Thus, if three axes are taken to represent the directions of unit vectors **(R)**, **(G)** and **(B)**, which are conveniently taken to be mutually perpendicular, then the projections of a vector **OC** on to these three

axes represent the tristimulus values of the colour **OC**, as shown in Fig. 2.10.

Using vector notation

$$\mathbf{OC} = R(\mathbf{R}) + G(\mathbf{G}) + B(\mathbf{B})$$

where the direction of the vector **OC** represents the chromaticity of the colour. **OC** varies as the luminance of the colour C* increases, but is only directly proportional to luminance for any given fixed direction of **OC**, since luminance is the scalar or arithmetical sum $R + G + B$. The direction of the vector which is equidistant from the R, G, B axes represents grey colours of varying luminance.

On such a colour space diagram the Maxwell chromaticity diagram is the triangle formed by joining the ends of the unit vectors, and the intersection of the plane of this triangle with the colour vectors gives the chromaticity of that colour.
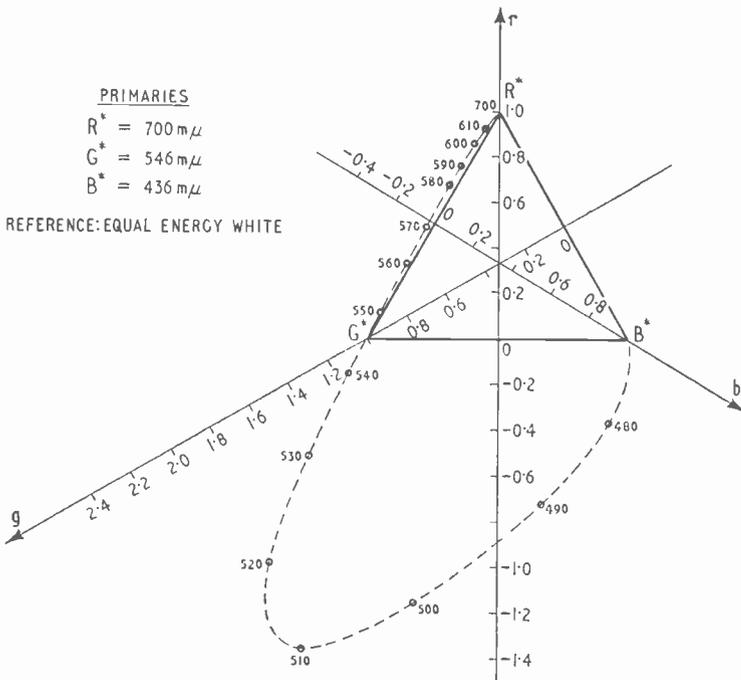
Fig. 2.11.   Chromaticities of the spectrum colours

The chromaticity diagram of Fig. 2.6 is the projection of this Maxwell triangle on the *ROB* plane.

Since the units in which the primaries are measured have different luminosity coefficients, the vectors which terminate on the Maxwell triangle do not have the same luminance, and this plane is not a plane of constant luminance.

It is generally more convenient to work with a two dimensional chromaticity diagram.

2.16. Chromaticities of the spectrum colours

The chromaticities of the spectrum colours for an equal energy white may be measured and plotted, when they are found to form a curve as in Fig. 2.11.   As in all colour matching experiments, it is necessary to define the primary lights used and also the reference or normalizing white.   For most of the spectrum wavelengths a colour match can only be obtained by transferring one of the primary colours to the spectrum colour side of the colorimeter, so that one of the chromaticity co-ordinates is negative.

At the extreme ends of the spectrum locus the wavelength scale becomes very cramped and a large wavelength change produces little or no change in the chromaticity.

2.17. Colour mixture curves

The tristimulus values obtained by matching the equal energy spectrum colours with a particular set of primaries may be plotted on a relative scale as in Fig. 2.12, for primary units chosen to match a given reference white.   Such colour mixture curves are used a great deal in colorimetry and average curves have been agreed for the standard observer and are referred to as $\bar{r}$, $\bar{g}$ and $\bar{b}$ curves.   The curves in Fig. 2.12 are not the standard international curves but a set based on the primaries used in colour television.

The amount of red primary required to match any colour whose energy distribution with wavelength $E(\lambda)$ (*see* Fig. 2.2) is given, may now be found by integrating the product of $\bar{r}$ and $E(\lambda)$ at each wavelength throughout the visible spectrum

$$R = \int_0^\infty \bar{r}E(\lambda)\, d\lambda$$

This is usually done graphically and practical limits 380 to 780 mμ are sufficient.   For instance, at every 10 mμ the values of $E(\lambda)$ and $\bar{r}$ may be taken and multiplied together and a new curve plotted with these values.   The total area under the new curves is the

required tristimulus value. Areas below the zero tristimulus axis are negative. Similarly G and B may be found. Once the colour mixture curves are known the colorimeter may be discarded and purely physical measurements of the spectral energy distribution of a colour made, from which the tristimulus values may be calculated. Such energy measurements can be made with an instrument called a *spectrophotometer*.

The colour mixture curves themselves may be calculated from the chromaticity co-ordinates of the spectrum, and the V_λ curve.
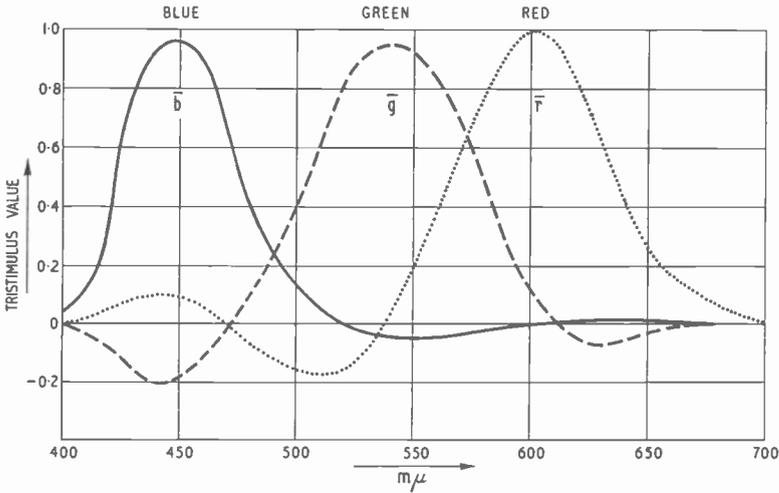
Fig. 2.12. Colour mixture curves for an equal energy spectrum, using the N.T.S.C. reference stimuli and Illuminant C as normalizing white (Based on information from "Colour Television Engineering" by Wentworth. Copyright 1955. McGraw Hill Book Company, Inc.)

The areas under each $\bar{r}$, $\bar{g}$ and $\bar{b}$ curve must be equal. These areas are

$$\int_0^\infty \bar{r}\,\mathrm{d}\lambda, \ \int_0^\infty \bar{g}\,\mathrm{d}\lambda \ \text{ and } \ \int_0^\infty \bar{b}\,\mathrm{d}\lambda$$

2.18. Transformation to different reference stimuli

If a colour C^* has been matched with one set of primaries R^*, G^* and B^* in terms of the unit amounts of R^*, G^* and B^* which match the chosen reference colour, it is possible to calculate the tristimulus values of C^* in terms of any other set of primaries, say, X^*, Y^* and Z^*. It is necessary to know the specification of X^*, Y^* and Z^* in terms of the R^*, G^* and B^* reference stimuli.

$$1(X^*) \equiv k_1R^* + k_2G^* + k_3B^*$$

$$1(Y^*) \equiv k_4R^* + k_5G^* + k_6B^*$$

$$1(Z^*) \equiv k_7R^* + k_8G^* + k_9B^*$$

The brackets round X^*, Y^* and Z^*, indicate that the units of measurement are the T-units of the $R^*G^*B^*$ system. From these equations, in which k_1 to k_9 are known constants, it is possible but tedious to calculate the values of a_1 to a_9 in these equations

$$1R^* \equiv a_1(X^*) + a_2(Y^*) + a_3(Z^*)$$

$$1G^* \equiv a_4(X^*) + a_5(Y^*) + a_6(Z^*)$$

$$1B^* \equiv a_7(X^*) + a_8(Y^*) + a_9(Z^*)$$

from which a colour $1C^*$ which is given by

$$1(C^*) \equiv rR^* + gG^* + bB^*$$

can be converted to

$$1(C^*) \equiv c_1(X^*) + c_2(Y^*) + c_3(Z^*)$$

This will give the match to 1 T-unit of C^* in the T-units of the old $R^*G^*B^*$ system. This unit trichromatic equation for C^* can be converted to the T-units of the new $X^*Y^*Z^*$ system if the new reference colour W^*_2 is given, since

$$1W^*_1 \equiv \tfrac{1}{3}R^* + \tfrac{1}{3}G^* + \tfrac{1}{3}B^* \qquad \text{in the old system}$$

and $\qquad 1W^*_2 \equiv \tfrac{1}{3}X^* + \tfrac{1}{3}Y^* + \tfrac{1}{3}Z^* \qquad$ in the new system.

Usually the reference whites are the same. So finally

$$1C^* = xX^* + yY^* + zZ^*$$

Fig. 2.13 shows a set of 3 such primaries, X^*, Y^* and Z^*, on a chromaticity diagram. The chromaticity axes shown are the g and r axes rather than the r and b axes which have been used up to now. r and b axes are usually more convenient for colour television, whilst g and r axes are normally used for colorimetry. The reason for changing the chromaticity diagram at this juncture will become clearer in the next section. The positions of X^*, Y^* and Z^*, shown in Fig. 2.13, are outside the spectrum locus which encloses the chromaticities of all real colours, and hence X^*, Y^* and Z^*

do not correspond to any physically recognizable lights. They may perhaps be thought of as super-saturated colours. They are mathematical abstractions only but there are marked advantages in referring colours to such primaries.

If the point C marks the chromaticity of a particular colour C* on the R*, G* and B* chromaticity diagram, let the line from X
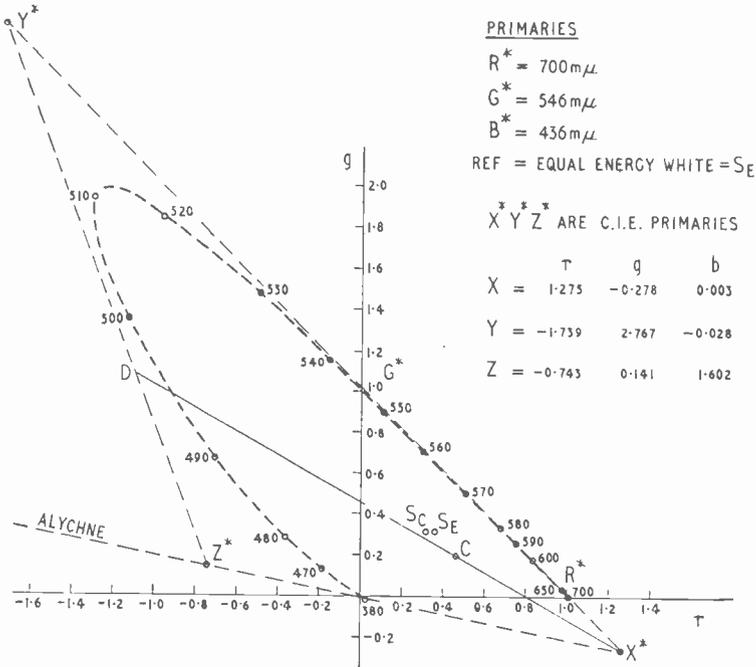
Fig. 2.13. *Transformation to new reference stimuli, showing the red-green chromaticity diagram spectrum locus for a standard observer*

through C meet YZ at D. Then C* will be matched by a mixture of X* and D* in proportions DC to CX, in T-units. Similarly D* will be matched by a mixture of Y* and Z* in the proportions ZD to DY.

2.19. C.I.E. standard reference stimuli X*, Y* and Z*

An international committee, the Commission Internationale de l'Eclairage or C.I.E., have adopted a set of non-physical primaries, X*, Y* and Z*, to simplify colorimetric calculations (*see* Fig. 2.13).

These non-physical primaries have the following chromaticities:

X^* $r = 1 \cdot 2750$ $g = -0 \cdot 2778$ $b = 0 \cdot 0028$

Y^* $r = -1 \cdot 7394$ $g = 2 \cdot 7674$ $b = -0 \cdot 0280$

Z^* $r = -0 \cdot 7429$ $g = 0 \cdot 1409$ $b = 1 \cdot 6020$

The units of X^*, Y^* and Z^* are such that a mixture of 1 unit of each matches equal energy white.

Since the chromaticities of all spectrum colours lie inside the triangle formed by the points marking the chromaticity of X^*, Y^* and Z^*, now all real colours have only positive chromaticity coefficients. Any colour C^* is matched by

$$CC^* = XX^* + YY^* + ZZ^*$$

$$1C^* = xX^* + yY^* + zZ^*$$

where

$$x = \frac{X}{X + Y + Z}, \ \ y = \frac{Y}{X + Y + Z} \ \text{and} \ z = \frac{Z}{X + Y + Z}$$

which are the C.I.E. chromaticity co-ordinates (*see* Section 2.11). It should be noted that although X^*, Y^* and Z^* are non-physical colours, it is possible to make a colorimeter which will measure the chromaticities of real colours directly in terms of the tristimulus values, X, Y and Z of the colours X^*, Y^* and Z^*. For example, to measure the amount of X^*, a spectrum colour of the same dominant wavelength as X^*, i.e., a red colour, is displayed in the known side of the colorimeter field and reference white is added to the other side. The amount of reference white added desaturates the unknown colour to the same extent that the chromaticity of the chosen spectrum colour is less than X^*.

The standard reference stimuli have been carefully chosen. In Fig. 2.13 it can be seen that the line joining Y^* and X^* runs along the spectrum locus for an appreciable range of colours at the red end of the spectrum so that all the colours along this line have zero blue coefficients. X^* and Z^* are located on the line where all the points have zero luminance, that is the line

$$lr + mg + nb = 0$$

where l, m and n, the luminosity coefficients of the $R^*G^*B^*$ primaries, are in the proportion 1 to $4 \cdot 5907$ to $0 \cdot 0601$ and this line (X^*Z^*) nearly passes through the origin.
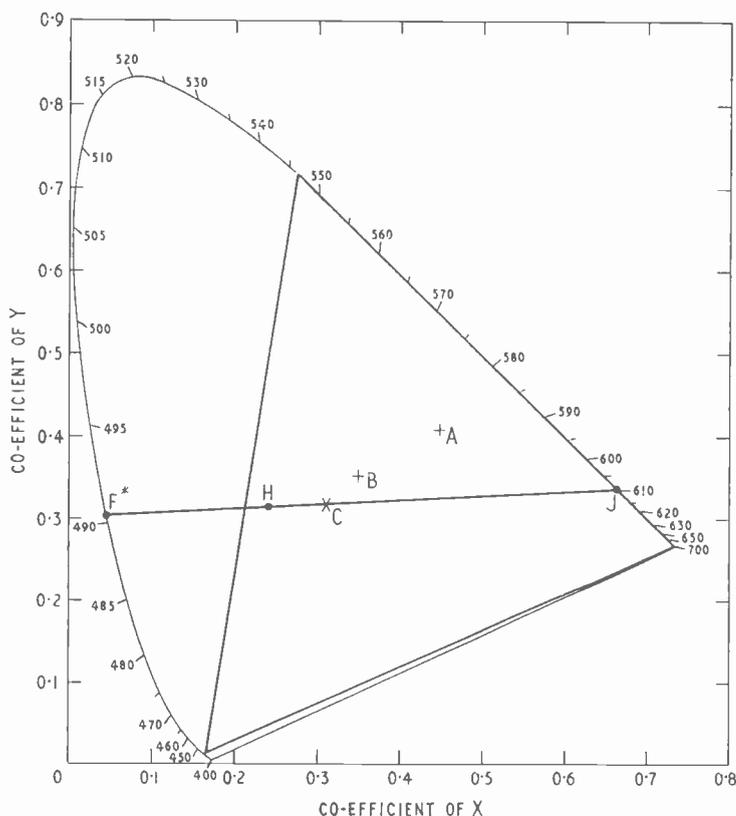
Fig. 2.14. The C.I.E. chromaticity diagram

This line is called the zero luminance line or the *alychne* and is used as the x-axis in the chromaticity diagram shown in Fig. 2.14. This is a convenient arrangement since the luminance of any colour is then equal to the number of Y* primary units needed to match the original colour. This rotation of the axes will be clear if the axes of Figs. 2.13 and 2.14 are studied.

The spectrum locus is plotted using the new axes on Fig. 2.14 and the chromaticity locations of the C.I.E. standard illuminants are also shown.

Illuminant *A* is the light from a gas-filled tungsten lamp.

Illuminant *B* is similar to sunlight.

Illuminant *C* is similar to the light from a north sky.

Equal energy white is located at the point $y = \frac{1}{3} = x$. The various groups of colours of similar chromaticities are shown in Fig. 2.15.

2.20. Light characteristics in colorimetry

The three characteristics of light which give rise to the subjective sensations of brightness, hue and saturation can all be measured objectively in colorimetry. The corresponding terms are luminance, dominant wavelength and purity. These objective terms are only used to describe the measured qualities and are not synonymous with the words used to describe the sensation which they produce in the observer.
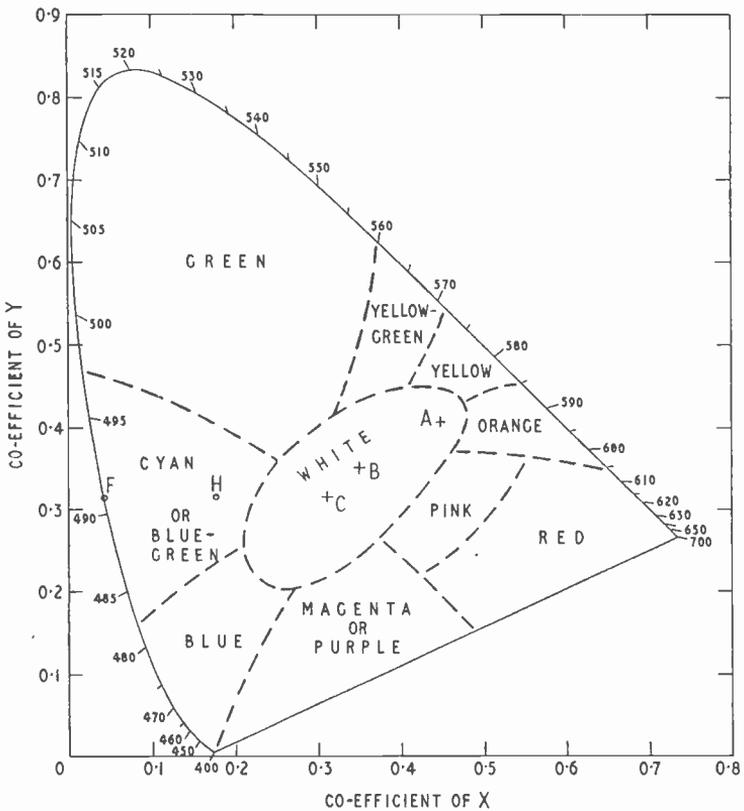
Fig. 2.15. *Approximate boundaries of colours on the C.I.E. chromaticity diagram*

2.20.1. LUMINANCE

Luminance has already been defined in Section 2.6.3. It corresponds to the subjective appreciation of brightness, but whether a given luminance looks bright or dark to the observer depends on the adaptation of his eye.

Luminance is denoted by the tristimulus value Y of the Y* primary and is usually measured in foot lamberts. In transforming tristimulus values to chromaticity co-ordinates, information about the luminance of the colour is lost.

2.20.2. DOMINANT WAVELENGTH

A straight line from a reference white such as Illuminant C in Fig. 2.14 to a spectrum colour F* passes through all the colours formed by a mixture of spectrum colour F* and the reference white. All these colours are said to have the same dominant wavelength as F*, but whether they produce the same hue sensation depends on the colour adaptation of the observer's eye. Strictly, the point marked F* in Fig. 2.14 should be called the chromaticity of the colour F*.

2.20.3. PURITY

The spectrum colour F* has 100% saturation or *purity* and the line $CF*$ represents the desaturation of the colour F* with the reference white. The purity of the colour with chromaticity represented by H will be equal to $\frac{100}{FC} HC\%$. To be precise, this is the definition of *excitation* purity. The reference white is completely desaturated or is at zero purity. The point which marks the reference white on the chromaticity diagram is referred to as the *achromatic point*.

2.20.4. DEFINITION OF COMPLEMENTARY COLOURS

If a line is drawn from a spectrum colour F* through the reference white to the other side of the spectrum locus, the point J marks the complementary colour to F*. All complementary colours can be found in this way by drawing a line through the point C.

2.21. Addition of colours in the C.I.E. system

Colours are added to form mixtures of different kinds of light in the C.I.E. system in the same way as in the RGB system, but now the calculation of luminance in absolute units is much easier.

Colours are normally specified in terms of their x and y chromaticity co-ordinates and their luminance, thus

$$\text{Colour } 1 = x_1, y_1, Y_1$$
$$\text{Colour } 2 = x_2, y_2, Y_2$$

These must be changed to tristimulus values before adding them together.

Let

$$X_1 + Y_1 + Z_1 = D_1$$

Then

$$x_1 = \frac{X_1}{D_1} \text{ and } y_1 = \frac{Y_1}{D_1}$$

and hence

$$D_1 = \frac{Y_1}{y_1}$$

and

$$X_1 = x_1 D_1$$
$$= \frac{x_1}{y_1} Y_1$$

further

$$Z_1 = z_1 D_1 = (1 - x_1 - y_1) D_1$$
$$= \frac{1 - x_1 - y_1}{y_1} Y_1$$

Similarly for X_2 and Z_2.

Then the tristimulus values of the mixture are

$$X_{1+2} = X_1 + X_2$$
$$Y_{1+2} = Y_1 + Y_2$$
$$Z_{1+2} = Z_1 + Z_2$$

and the chromaticity co-ordinates are

$$x_{1+2} = \frac{X_1 + X_2}{(X_1 + X_2) + (Y_1 + Y_2) + (Z_1 + Z_2)}$$
$$= \frac{X_1 + X_2}{D_1 + D_2}$$
$$y_{1+2} = \frac{Y_1 + Y_2}{D_1 + D_2}$$

As an example, consider the mixture of two colours given in Table 2.1.

Table 2.1
ADDITION OF COLOURS

	x	y	Luminance
Colour 1	0·2	0·6	18 foot lamberts
Colour 2	0·3	0·1	8 foot lamberts

$$D_1 = \frac{Y_1}{y_1} = \frac{18}{0·6} = 30$$

Therefore

$$X_1 = x_1 D_1 = 6$$
$$z_1 = 0·2, \text{ so } Z_1 = 6$$

also

$$D_2 = \frac{Y_2}{y_2} = \frac{8}{0·1} = 80$$

Therefore

$$X_2 = 24$$
$$z_2 = 0·6, \text{ so } Z_2 = 48$$

Therefore from these equations

$$X_{1+2} = 30, \quad Y_{1+2} = 26 \text{ and } Z_{1+2} = 54$$

so

$$D_{1+2} = 110$$

therefore

$$x_{1+2} = \frac{30}{110} = 0·27$$

$$y_{1+2} = \frac{26}{110} = 0·24$$

$$Y_{1+2} = 26 \text{ foot lamberts}$$

This chromaticity may be found graphically as in Fig. 2.16. It must lie on the line connecting $x_1 y_1$, the point P*, to $x_2 y_2$, the point Q*, at a position which divides the distance between P* and Q*
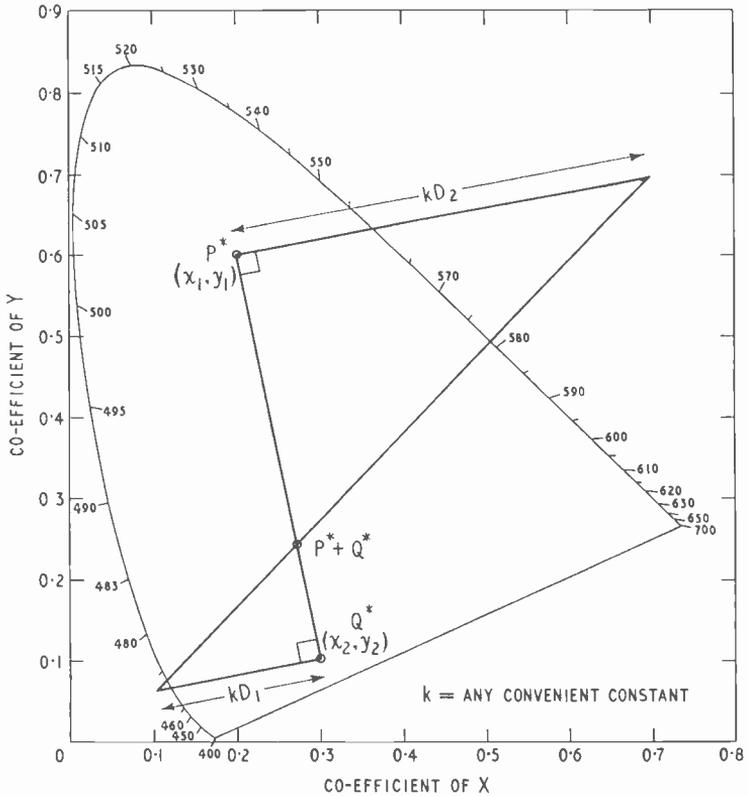
Fig. 2.16. Graphical construction for additive mixtures

inversely as the ratio of the number of T-units in colour P* to those in colour Q*. Therefore erect a line perpendicular to P*Q*, at the point P* with a length proportional to D_2 equal to kD_2. Erect also a line perpendicular to P*Q* at point Q* on the opposite side to the perpendicular at point P*, with a length proportional to D_1 equal to kD_1. The line joining the ends of these perpendiculars crosses the line P*Q* at the required chromaticity.

2.22. Chrominance

As the luminance of a colour increases, so the tristimulus values XYZ or RGB increase, but the chromaticity co-ordinates do not change. Another quantity called *chrominance* is used in colour television, and this can now be defined as the colorimetric difference

*Effect of removing each primary colour from a correctly balanced
picture; (above) red absent from picture; (centre) green absent
from picture; (below) blue absent from picture*
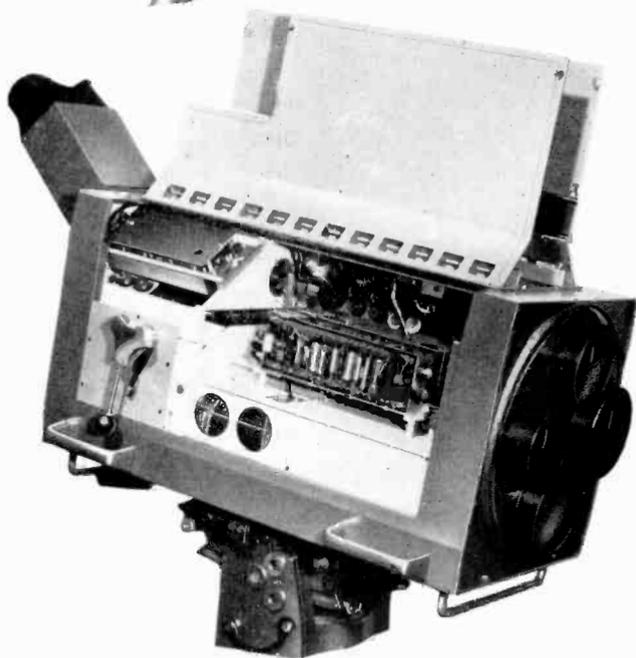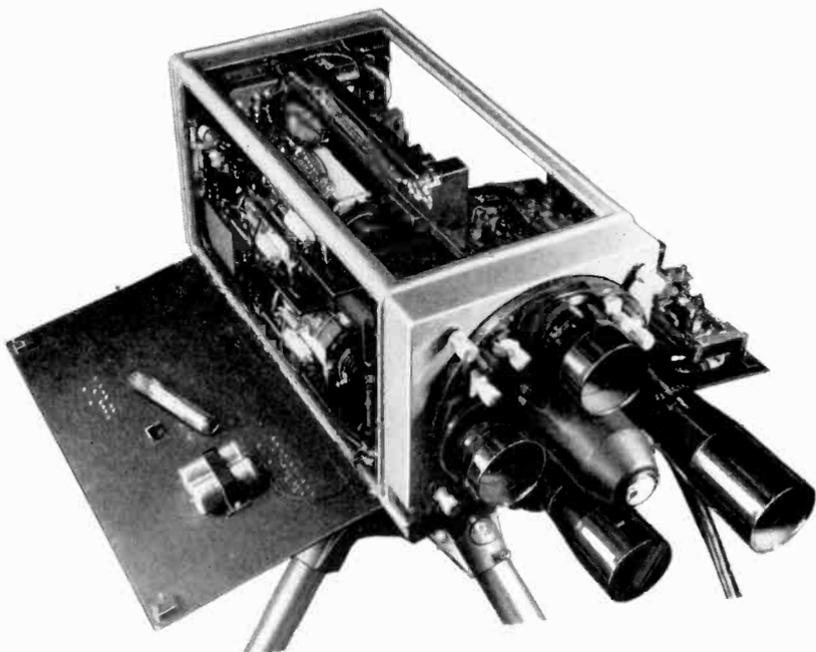
Two types of colour television camera. The top photograph shows an EMI camera, while the lower one a Marconi three-vidicon colour camera. (Courtesy EMI Ltd. and Marconi Ltd.)

between a colour and a reference white light of the same luminance as the colour. This quantity can only be illustrated in a three-dimensional colour space, such as the XYZ space of Fig. 2.17.

Since only Y controls luminance, planes parallel to the OZX plane are of constant luminance. Thus the vectors **OC** and **OW** representing a colour and a white of the same luminance must end in a plane parallel to OZX through the appropriate value of Y. The vector **WC** then represents the chrominance of the colour C.

The chrominance increases as the lengths of the vectors **OC** and **OW** increase with increase in luminance. The chrominance also increases as **OC** moves away from **OW**, that is, as **OC** represents colours of increasing saturation.

2.23. Perceptibility of changes on the chromaticity diagram

The specification and plotting of chromaticities follow linear laws, but the eye does not distinguish equally between chromaticity changes at one side of the diagram and changes at the other side. A very small change in chromaticity at the blue end of the diagram is noticeable, while the chromaticity of a green colour must move quite a long way over the diagram before the eye can notice any change in the corresponding colour.

Fig. 2.18 shows the relative susceptibility of the eye to chromaticity changes. The chromaticity change in going from a marked dot to any point on the surrounding oval produces a colour change which is about equally noticeable for each dot. The smallest
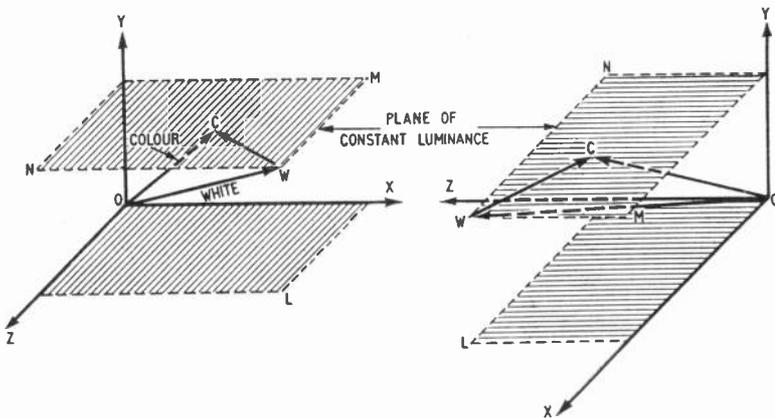
Fig. 2.17. Two views of chrominance vector in colour space

5

noticeable change depends on the viewing conditions but is approximately one hundredth of that shown.   The x, y diagram corresponds more closely to the subjective discrimination of the eye than do the x, z, and y, z diagrams, because the eye is not as sensitive to the blue component.

2.24. Summary

Any colour can be matched by amounts of three reference primaries, thus

$$CC^* \equiv RR^* + GG^* + BB^*$$
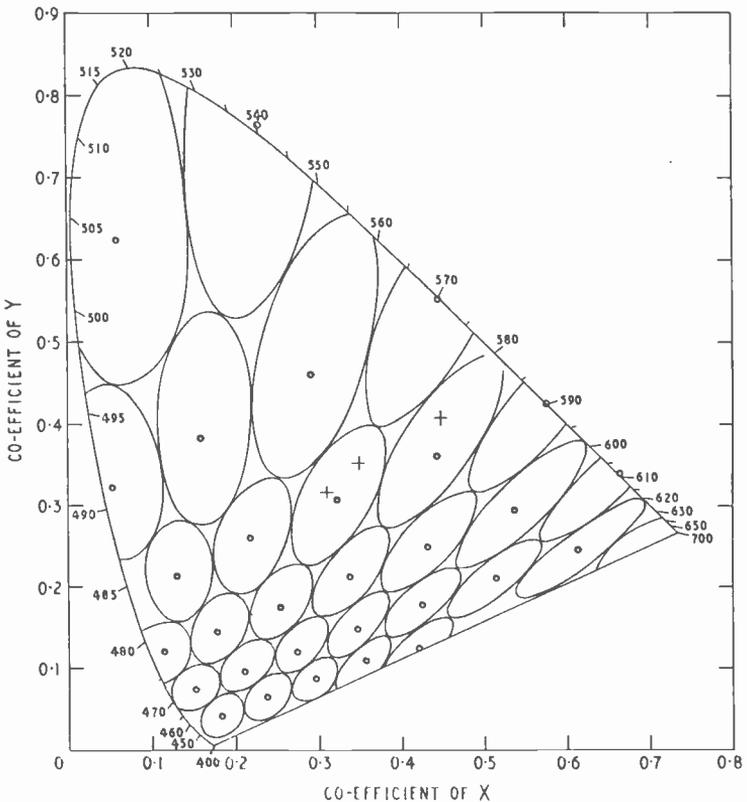
Fig. 2.18.   Approximate perceptibility of chromaticity changes.   These are Judd's equally perceptible ellipses (100 × minimum perceptible change in normal viewing)

or

$$1C^* \equiv rR^* + gG^* + bB^*$$

where

$$1 \equiv r + g + b$$

R^*, G^* and B^* are measured in T-units, so that for a given reference white

$$3W \equiv 1R^* + 1G^* + 1B^*$$

R, G and B are called tristimulus values and increase with luminance. The chromaticity co-ordinates r, g and b, are independent of luminance. Equal numbers of T-units of the primaries R^*, G^* and B^* contribute to the luminance of the mixture in the ratio of the luminosity coefficients l, m and n.

The tristimulus value of a mixture is the sum of the corresponding individual tristimulus values of the separate colours. The luminance of a mixture is the sum of the luminances of the separate components. A colour specified in terms of one set of primaries can be specified in terms of another known set of primaries without further measurement. The most convenient representation is in terms of the C.I.E. non-physical primaries X^*, Y^* and Z^*. The chromaticity of a mixture of two lights lies on the straight line joining the chromaticities of the constituent colours, at a point which divides the line in the inverse ratio of the T-units of each colour.

A quantity, chrominance, is defined which is the colorimetric difference between a colour and a reference white of the same luminance.

Colour Picture Tubes, Cameras and Film Scanners

3.1. Introduction

The simplest way to reproduce a colour picture is to reproduce separately the red, green and blue primary images and then to superimpose them so that the eye merges the three images into one full-colour picture. The fusion of the three images may be accomplished in several ways. If a separate display tube is used for each primary image and the pictures are combined optically, then at any instant all three primary images are being traced out and the reproduction is said to be made by a *simultaneous display*.

The three primary images may be presented to the eye in quick succession and the persistence of the eye's vision will merge the red, green and blue pictures into one coloured whole. This is a *sequential display*.

The eye will also fuse colours together if small areas of different hue are placed adjacent to each other. When small areas are viewed from such a distance that the eye has difficulty in resolving each individual area, the overall colour sensation is that of a normal mixture of the colours involved. The small areas in such a reproduction may be lit either sequentially or simultaneously.

All these methods and many ingenious combinations of them have been used to produce colour television displays. Only a few will be described in brief detail in this chapter, and of these the only colour tube which is made in quantity is the R.C.A. shadow mask tube. This tube will be discussed at length in Chapter 11 and the short outline given in this chapter must not be taken as a measure of the practical importance of the shadow mask tube.

Considerable inventiveness has likewise been shown in the suggestions put forward for the design of colour cameras, but here also only one type of camera is in general use. It uses three normal monochrome camera tubes, each of which produces a signal proportional to one of the primary images. For the transmission of colour film, flying spot scanners are used as in monochrome television but

three photocells are now necessary, each of them sensitive to one primary colour.

In general both transmitting and receiving equipment aims at reproducing a colorimetric match to the original scene although the viewing conditions and the absolute luminance levels are different. Departures from this principle are only made on the grounds of expediency and cost.   There is some doubt as to whether such an exact match gives the viewer the most pleasing reproduction but it is considered to be better practice to leave such subjective considerations to the artistic control of the studio producer.

3.2.   Display primaries

The choice of chromaticities for the primary lights used to produce the receiver's colour picture determines the range of reproduced chromaticities which are practical.   As large a range, or *gamut*, of chromaticities as possible is obviously desirable, although the efficiency of light output and practical manufacturing limitations restrict the choice.   The primary chromaticities which have been standardized for use in the N.T.S.C. system are discussed in the next chapter and are plotted in Fig. 3.1.

Gelatine colour absorption filters for converting white light to the primary colour lights, and phosphor powders for use in cathode-ray tubes, can all be manufactured to produce approximately the N.T.S.C. primaries.

Typical phosphors are zinc phosphate for red, zinc orthosilicate for green and a zinc sulphide activated with silver or copper for blue. The red and green phosphors usually have a medium after-glow or persistence, which falls exponentially, still giving $\frac{1}{e}$ or 37% of the original light output about 2 or 3 ms after the electron bombardment ceases.   The blue phosphor has a shorter persistence, falling to 37% in about 0·1 ms.   Objects moving across the screen may therefore leave a yellow trail behind them as the red and green phosphors continue to glow after the blue phosphor has ceased phosphorescence.

The luminous efficiency of these phosphors is low, only a few per cent of the electrical energy being converted into radiant energy, with the green phosphor having the highest efficiency.   Ten or so lumens per watt can be obtained at 20 kV, but the light output may saturate at current densities of a few $\mu A/cm^2$, which is often reached for the red and blue phosphors in projection type tubes. Because of the relatively low efficiency of the red phosphor used in
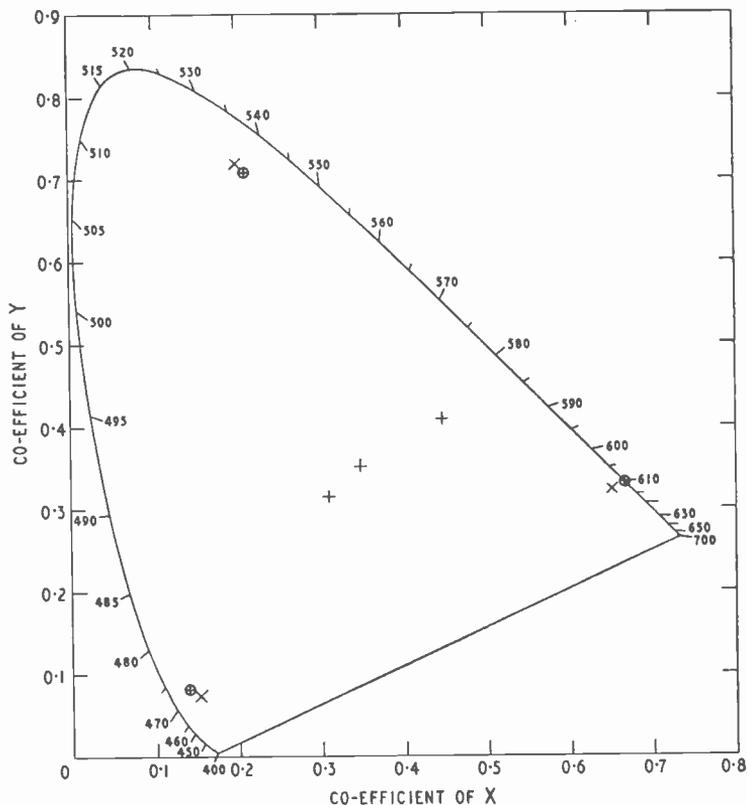
Fig. 3.1. *Typical phosphor characteristics. P22 screen phosphor chromaticities after screening are shown as* × *and N.T.S.C. reference stimuli are shown as* ⊕

Table 3.1

TYPICAL PHOSPHOR CHARACTERISTICS

Colour	Composition	C.I.E. co-ordinates				Persistence	
		Dry powder		After screening		Elapsed time msec	% brightness left
		x	y	x	y		
Red	$Zn_3(PO_4)_2$:Mn	0·674	0·326	0·650	0·322	33	12·7
						40	8·2
Green	Zn_2SiO_4:Mn	0·203	0·728	0·200	0·718	33	8·9
						40	5·1
Blue	ZnS:Ag	0·146	0·052	0·151	0·073	40	0·01

direct viewed tubes, a phosphor which glows more orange than the N.T.S.C. red is sometimes employed.

3.3. Three-tube displays

Perhaps the simplest type of colour display, in conception, consists of three projection type tubes arranged so that each is throwing a picture on to the same screen. Each tube phosphor glows in one of the primary colours and gelatine colour absorption filters may also be used to modify further the chromaticity of the screen face. Video signals proportional to either the red, green or blue tristimulus values of the scene are applied to each tube.

The input-output characteristics of the three channels must match exactly if the same ratio of red, green and blue signals is to produce a dark grey at low brightness levels and a white of the same neutral colour at high brightness levels. A neutral grey scale such as the central gradation blocks of the Test Card " C " is very useful for testing the match between the three colour channels at different input levels. Not only must the amplifier channels match but the voltage input-light output characteristics of the cathode-ray tubes must also be identical. The production of a good neutral grey scale is one of the major problems in any three-gun display. Typical errors are the dark greys looking reddish in tone, the middle greys appearing pinkish and the whites looking neutral.

Care is needed to ensure that all three images are the same size and shape, and the three identical deflection yokes are fed from the same timebase circuits. The tubes may be offset from the projection lens axes so that the images coincide, or each tube may be mounted on the inclined axis of its own lens and close to the next tube, when the resulting geometrical distortion may be corrected by modifying the scanning waveforms.

The images on three direct viewed tubes may be combined optically by mirrors so that the observer sees only one composite image, as in Fig. 3.2. This type of arrangement is sometimes called a *trino-scope* or a *jumbo*. The optical path length from the observer's eye to each tube is made the same for all three tubes so that the three images appear the same size. Such arrangements are bulky and only one observer can view directly on the axis. A rather more compact arrangement may be made by placing the red tube on the other side of the viewing axis, turning its mirror through 90° and moving it back so that if forms a cross with the blue mirror, but this arrangement gives a dark line across the centre of the picture. The mirrors used may be semi-transparent silver mirrors but these prevent a large part of the light emitted by the tubes from reaching

the eye. Very efficient mirrors called *dichroic* mirrors are normally used.

3.4. Dichroic mirrors

A dichroic mirror transmits light at all wavelengths except for a group at one end of the spectrum, which it reflects. Thus a blue reflecting dichroic mirror may transmit all wavelengths above 500mμ
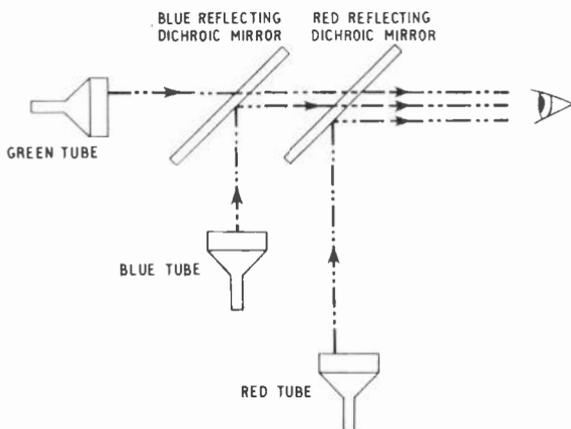
Fig. 3.2. Principle of direct-viewed three-tube display

and reflect all wavelengths below 460mμ, with a gradual changeover from transmission to reflection for the region between 460mμ and 500mμ. Similarly, a red reflecting dichroic mirror may change over to transmission at about 580mμ. The wavelengths at which the changeover occurs depend on the angle of incidence of the light, the changeover moving to longer wavelengths as the angle of incidence decreases. An increase of perhaps 30mμ may occur in changing from 45° to normal incidence.

Dichroic mirrors are sheets of glass which have been coated not with silver but with alternate layers (normally seven but may be as high as twenty) of materials with high and low refractive indices. Each layer is usually a quarter of a wavelength of light thick at the rejection wavelength. At each interface some of the incident light is reflected and the thicknesses of the layers are adjusted so that the reflected rays cancel at some wavelengths and add at other wavelengths. This is the way in which a thin film of oil on water produces coloured patches.

To correct for the change in angle of incidence which can occur across an inclined dichroic mirror with wide angle optical systems, it is possible to taper the dichroic film from one side of the mirror to the other.

Because some light is also reflected from the other side of the glass, very bright objects in the field of view may give rise to spurious images which are just noticeable.

3.5. Rotating-disc displays

The difficulties which arise with three-tube displays over registration of the three images and the matching of the three channels to ensure good grey scale reproduction, can both be avoided by the use of only one amplifier chain and one cathode-ray tube which sequentially amplify the red, green and blue signals.   This was the colour system used by Baird in 1928, a more sophisticated form of which was officially adopted in America for a short period from 1950 to 1951.

At the instant when the channel is amplifying the red signal, a red colour filter is mechanically placed in front of the white glowing cathode-ray tube.   When the signal changes to the green signal, the red filter is replaced by a green filter and so on.   The mechanical change of the coloured filters can be achieved by rotating a disc with coloured sections (*see* Fig. 3.3) in front of the cathode-ray tube. The shape of each coloured filter depends on the relative position of the tube face and the centre of the disc and on the decay time of the C.R.T. phosphor.   Drums of filters which revolve around the complete C.R.T. have also been used.

The mechanical difficulties of such arrangements for high rates of rotation limit the device to field sequential systems in which the
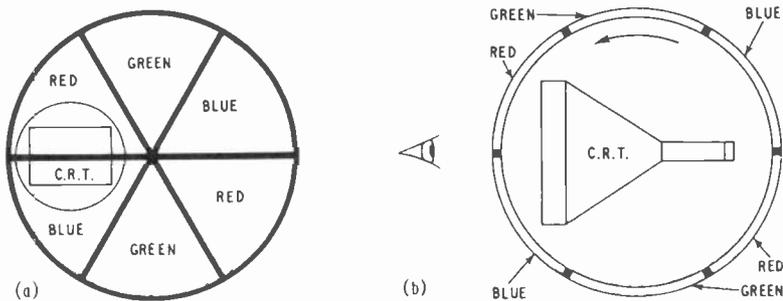
Fig. 3.3.   Rotating colour filters, (a) disc and (b) drum

colour is changed only once every field. With reasonably pure primary filters there is also a considerable loss of light (typically 5 to 1) from the white cathode-ray tube, but good colorimetric performance can be obtained.

All such field sequential receivers suffer from two defects called *colour flicker* and *colour break-up*. Since the luminosities of the red, green and blue pictures are not in general the same, a brightness

R = RED
Y = YELLOW
W = WHITE
C = CYAN
B = BLUE
G = GREEN
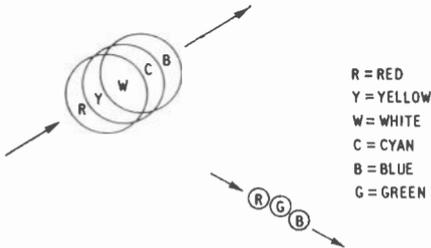
Fig. 3.4. Colour break-up of a moving object, due to displacement of primary images during sequential scanning

flicker occurs unless a complete cycle of 3 colour fields occurs in, at most, $\frac{1}{50}$ of a second. If each colour field occupies $\frac{1}{50}$ of a second, then when any area of the transmitted picture contains only one of the primary colours, e.g. a red patch, it is only lit for $\frac{1}{50}$ second in every $\frac{3}{50}$ second or just over 16 times a second.

If any object in the picture changes its position during the cycle of 3 colour fields it is seen as a series of red, green and blue images as it moves across the screen (Fig. 3.4). This is called *colour break-up* or *colour fringing*. It can also occur if the viewer moves his head quickly so that the red, green and blue images fall on different parts of the retina. However, if the colour field repetition rate is high enough to overcome the brightness flicker, then colour break-up effects are usually negligible.

Since a simultaneous transmission provides all the colour information all the time, a sequential display can be used at the receiver for either simultaneous or sequential transmission.

3.6. Lawrence type tubes

Various types of colour display tubes using a phosphor screen made up of strips of red, green and blue phosphors have been proposed (Fig. 3.5). The electron beam can be made to fall on any particular one of the three primary phosphor strips by means of a wire grid placed between the electron gun and the screen. Such tubes are called *Lawrence tubes* after the American physicist who suggested this arrangement, but particular laboratory models have

also been called " Chromatron ", " Chromaton ", focus mask tubes and post-deflection focus (P.D.F.) tubes.

The wire grid is spaced about a centimetre behind the phosphor screen and the wires are connected together in two sets. All the wires of one set are in line with the red phosphor strips and all the wires of the other set are in line with the blue phosphor strips. There are no wires in line with the green strips, of which there are twice as many as there are red or blue strips.

If there is no voltage between the two sets of grid wires, the electron beam passes through the wires and is focused on to the green-glowing phosphor strips. This focusing action is obtained by keeping the mean voltage of the grids many thousands of volts below the potential of the phosphor screen.

By placing three to five hundred volts between the two sets of grid wires, the electrons are bent towards the nearest positive wire and alight on the corresponding red or blue phosphor strip depending on the polarity of the applied voltage. In this way the grid wires control the primary colour which is being reproduced at any instant and the video signal must be switched so that the corresponding red, green or blue signal is applied to the electron-gun grid-cathode space.
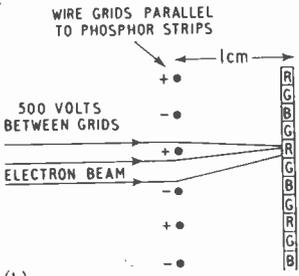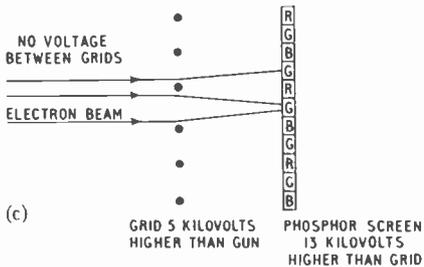
Fig. 3.5.   Operation of Lawrence tube.   (a) shows phosphor stripes viewed from the front; (b) and (c) are side views

The switching is most conveniently done at field frequency but can be done at either line or picture element frequency. The sub-carrier frequency itself can be used to synchronize the video and wire-grid switching. In such cases the tube is operated with the phosphor strips horizontal and parallel to the scanning lines. By applying a sine wave voltage between the wire grids the electron beam is swept across the phosphor strips and the intensity of the beam is controlled by a signal gated out from the video information. Since the beam traverses the green strip twice in each cycle, the phase of the gating signal must either be adjusted to miss the green strip on one crossing per cycle or the gating width for green must be half that for red and blue. Differences in efficiency between the blue, green and red phosphors can also be compensated by altering the gating width. About seventy volts of video drive are needed. Various ingenious ways of deriving the gated video signal have been proposed but at the time of writing no commercial receivers using such tubes have been produced.

The two sets of grid wires have appreciable capacity, between 1,000 and 2,000 pF, and some 20 to 30 watts of power have to be generated for the wire-grid switching, which necessitates effective shielding of the generator circuits. The two grids themselves do not radiate appreciably as they are only a fraction of a wavelength apart (10^{-5} for sub-carrier switching). Ideally, each strip and the focused electron spot should have a width of at most $\frac{1}{3}$ that of a picture element. Such close spacing of the strips has not yet been achieved and there is some doubt as to whether adequate cathode life could be obtained with the required small intense spot.

Three-gun versions of the focus mask tube have been made in the laboratory. The three electron guns are aligned so that the wire grids focus each electron stream on to only one colour phosphor, and no switching voltage is required.

3.7. Sensing tubes

Another promising type of colour tube, which has been demons-trated but is not yet commercially available, also uses a phosphor screen of red, green and blue strips. Instead of directing the electron beam on to the required colour phosphor the beam (as it traces out its normal raster) scans over each vertical strip in turn. In order that the video circuits may select the right primary colour signal for the single electron gun, it is necessary for the display tube to be able to tell the video circuits which colour phosphor the electron beam is falling upon at any particular instant. Various ways of
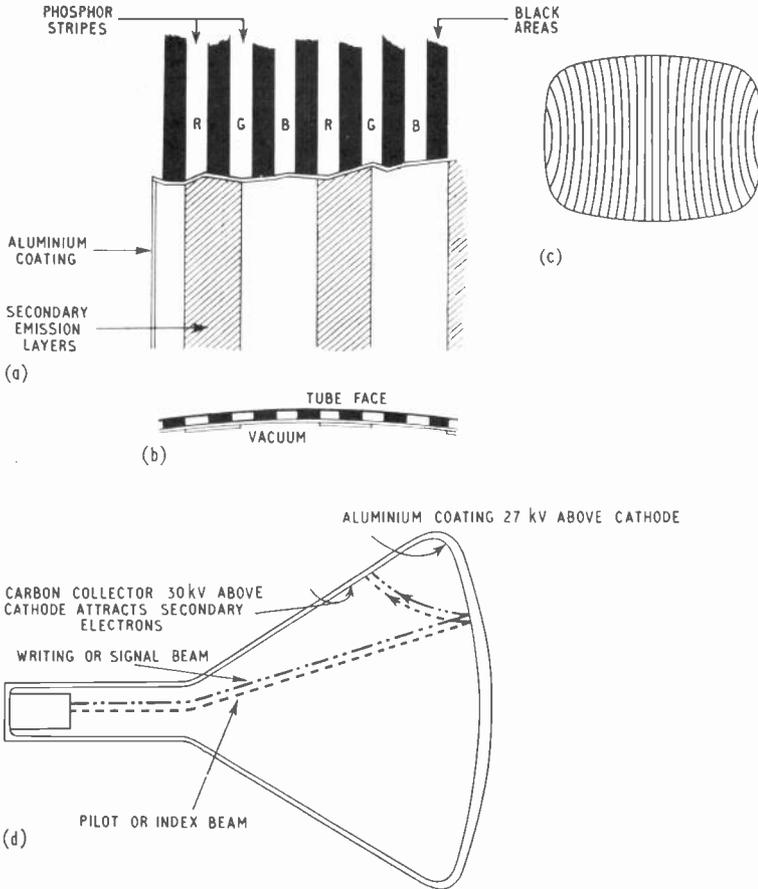
Fig. 3.6.   Philco Corporation "Apple" or sensing tube.   (a) is the back elevation and (b) the plan.   (c) shows pincushion arrangement of phosphor strips and (d) is a cross-section view of the tube

producing this sensing information have been patented but the most successful method appears to be that used by Philco Corporation in their " Apple " tube.

In the " Apple " tube the phosphor screen of vertical red, green and blue strips has dark non-luminescent areas between each phosphor strip, each dark area being as wide as a phosphor strip. This ensures that the scanning spot can illuminate any primary colour without desaturation.   The efficiencies of the phosphors are adjusted by the inclusion of non-activated material so that the

unmodulated scanning beam produces a white raster. The reference white and the grey scale balance are thus not functions of the circuitry or of the tracking between three electron guns.

Behind the normal aluminium screen backing another layer of vertical strips of high secondary emission material (magnesium oxide) is deposited, with one strip to each triad of phosphor strips, the magnesium oxide overlapping 40% of the total width of each triad (*see* Fig. 3.6) which is 0·06 in.

As the scanning beam passes over a secondary emission strip or index strip as it is called, it emits a burst of secondary electrons which are collected by a carbon coating on the inside of the tube bulb. The index strips lie behind the red phosphor strips and the bursts of secondary emission can be used to synchronize the gating of the red colour signal, and hence of the green and blue signals if the linearity of scan over this short width of a triad is reasonable. However, if an averaging process is used in the generation of the green and blue signals, then the overall linearity of the line scan must be good.

Since the writing electron beam disappears for the black parts of the picture and varies in intensity with the picture colour, a second electron beam, i.e. the index beam, is provided and scans with the writing beam. The index beam has no picture modulation and does not disappear in the black parts of the picture but remains as a just visible background of about $\frac{1}{2}$ foot lambert. The amplitude of the index beam is varied at a high frequency outside the video band so that tuned circuits can extract the index beam information from the signals produced by the video information.

The " Apple " tube has usually been operated so that it acts as its own decoder. The chrominance sub-carrier frequency is converted by a heterodyning process to the indexing frequency, which depends on the stripe geometry and line scan. The conventional chrominance signal then goes through its colour phase sequence (*see* Chapter 4) in step with the scanning of the writing beam over the phosphor strips.

The width of the triads is graded across the tube face to allow for average scanning non-linearity, and the pattern of triads is made slightly pincushion in the corners to allow for average scanning coil distortion. As the transit time of the secondary electrons from the index strip to the collector varies from the centre of the tube outwards, the phase of the index strips with respect to the phosphor strip is varied to compensate.

As with the Lawrence tube, narrow strips and very small spot sizes are necessary to achieve high definition and good saturation.

3.8. Shadow mask tubes

The most successful colour display tube is the R.C.A. shadow mask tube. This tube has been manufactured and sold for a number of years in some quantity and is capable of producing a very good 405-line colour picture. It is a three-gun tube and is basically suitable for simultaneous systems although it can be used for sequential displays. The phosphor screen of a 21 in. shadow mask tube is made up of just over a million phosphor dots, arranged in triads of red, green and blue (Fig. 3.7). About $\frac{3}{4}$ in. behind the phosphor screen
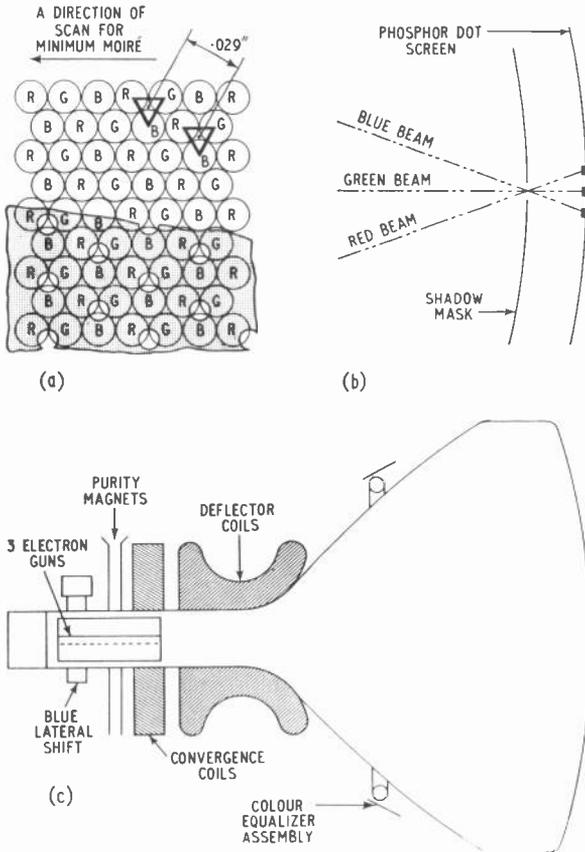
Fig. 3.7. R.C.A. 21 in. shadow mask tube. (a) shows an enlarged view of the shadow mask with the phosphor dot screen behind it and (b) the principle of the geometrical separation of three electron beams. (c) is a cross-sectional view of the tube

is a metal plate with 357,000 holes in it, each hole being aligned with
one of the triads of phosphor dots.   The three electron guns are
mounted symmetrically about the tube axis and the geometry of the
whole arrangement is such that electrons from any one gun can only
reach one colour of dots, the metal mask shadowing the dots of the
other two colours.

The focused electron beam of each gun covers an area of two or
three triads and the majority of the electrons are absorbed by the
metal plate, only some 20% reaching the phosphor screen.   Despite
this absorption of electrons, a high light brightness of 20 foot lam-
berts is usual with an E.H.T. voltage of 23 kV.

The phosphor dots are deposited on the inside of the face plate
glass, which is curved with a radius of 26 in.   The shadow mask
plate is made of copper-nickel alloy or, in later tubes, of steel about
0·006 in. thick and is curved to match the phosphor dot screen
curvature.

The mounting of the shadow mask is designed so that as the
mask heats up the expansion moves the apertures in the direction
of the beam travel.

Several auxiliary devices are needed outside the shadow mask
tube in order to obtain good colour pictures.   The three electron
beams have their centres of deflection (*see* Section 11.6) inside the
deflector coils, at a distance of approximately 15 in. from the
centre of the phosphor screen.   The electron guns are aligned so
that the undeflected spots approximately converge at the centre of
the screen.   To achieve exact convergence in the undeflected position,
or *static* convergence, the electron beams are moved by means of
external magnets or electromagnets placed round the tube neck.
The convergence magnets move the three spots radially, the flux
from the external magnets being carried around the electron
beams by pole pieces built into the gun structures (*see* Fig. 3.8).
Thus any two spots can be made to overlap.   The lateral move-
ment required to make the third spot coincident is provided by the
blue lateral shift magnet at the end of the tube, where there are
further pole pieces inside the tube neck around the blue beam.

As the three beams are deflected, the point of convergence moves
away from the phosphor screen to a position inside the tube itself.
The three electron-beam spots therefore diverge (Fig. 3.9) and must
be brought into coincidence again by varying the convergence fields
dynamically.

This *dynamic* convergence is achieved by applying parabolic
current waveforms to electromagnets on the convergence magnet
assemblies.   These waveforms are derived from the frame and line

REMOTE CONTROL
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VERTICAL ASTIGMATISM
COMPENSATOR
E/TAD 3238

BLUE FILTER
E/TAD 3246

HORIZONTAL ASTIGMATISM
COMPENSATOR   E/TAD 3239

MIRROR
E/TAD 3242

INCOMING LIGHT

IRIS HOUSING

SUPPLY LEAD
TO IRIS MOTOR
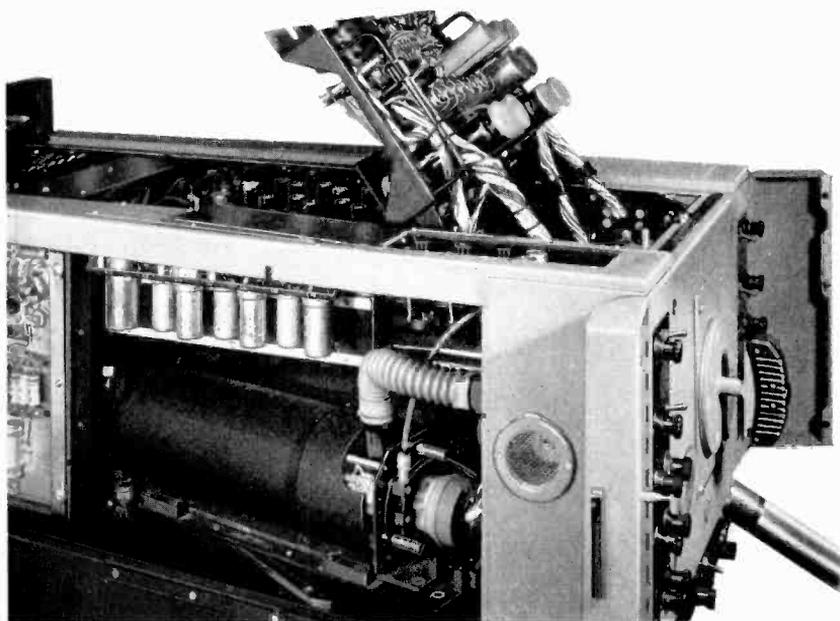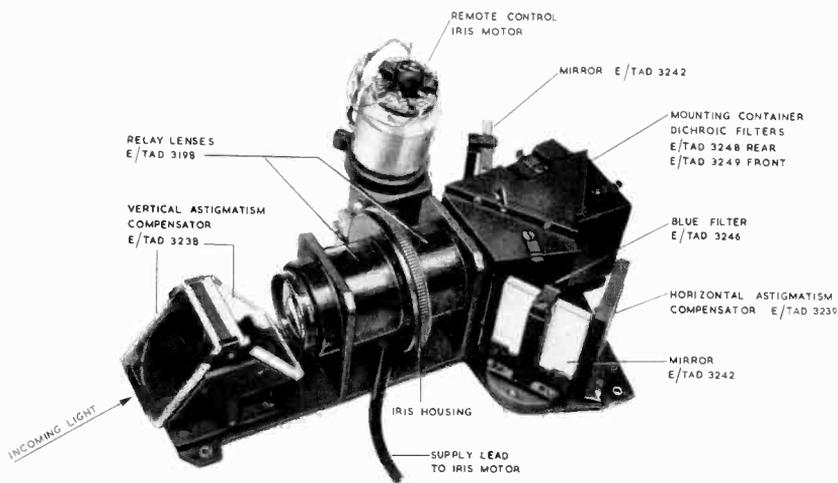
The Marconi image orthicon colour camera with the covers re-
moved. The optical system which is shown in the top photograph
is in the front part of the camera. (Courtesy Marconi Ltd.)
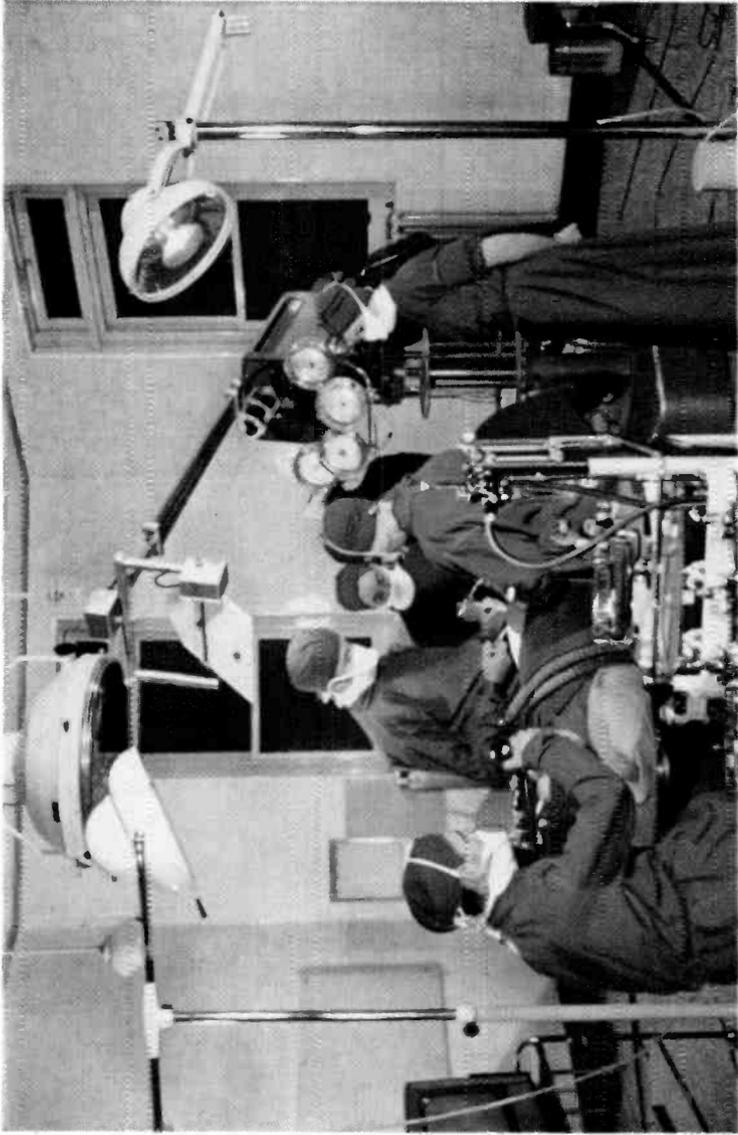
A typical scene in the operating theatre at the Hammersmith Hospital. The camera, with lights mounted on the front, photographs the actual operation via a 45° mirror vertically above the patient and is transmitted by a closed loop television circuit to a lecture room remote from the theatre. (Courtesy E.M.I. Ltd.)
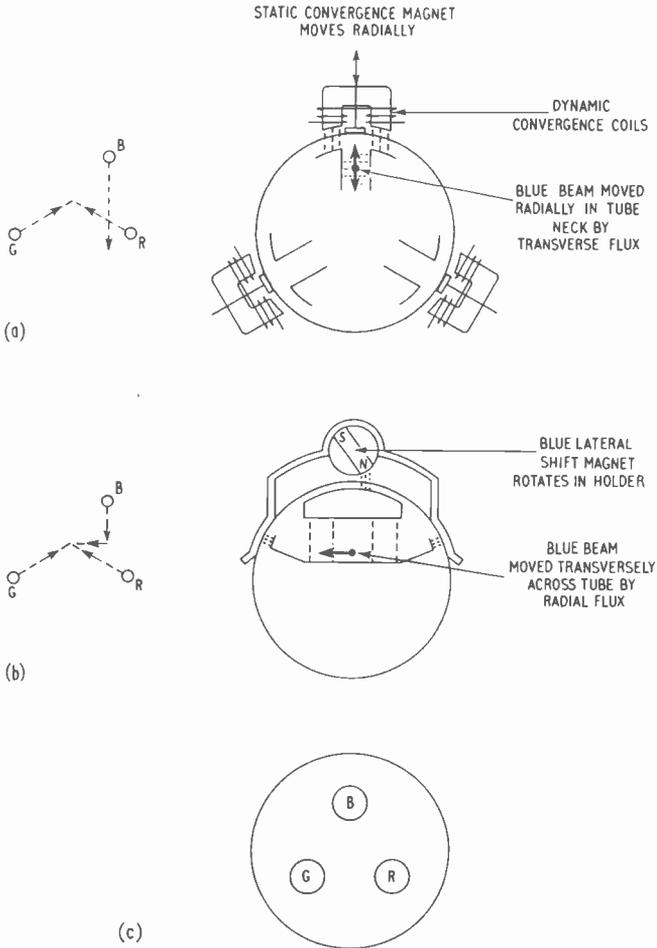
Fig. 3.8. *Control of beam convergence in a shadow mask tube.
(a) shows dynamic convergence coils, (b) the blue lateral shift
magnet rotating in holder and (c) the arrangement of three guns
viewed from the socket of the tube*

time base waveforms since the convergence correction required
depends on the position of the spot on the raster.

In order that the red electron beam shall land on only red phosphor
dots, its point of origin must be located with some precision with
respect to the shadow mask. Purity magnets are used to move all
three electron beams about in the tube neck until they appear to

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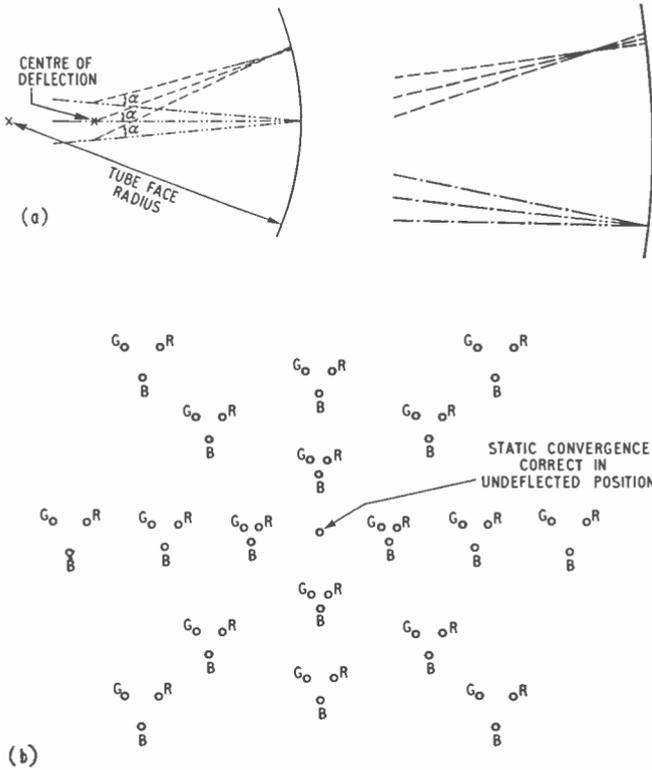
Fig. 3.9. *Convergence of three-gun tube.* (a) *shows the relative flatness of the tube face and the tilt on the guns for dynamic convergence control, and* (b) *shows the approximate separation of spots as three beams traverse the screen*

originate from the correct positions. The purity magnets may consist of two annuli magnetized transversely to the tube axis so that rotation of the two rings relative to each other alters the strength of the field, while rotation of both rings as a unit rotates the direction of the magnetic field across the tube neck.

Unwanted magnetic fields which vary across the picture area will upset the alignment of the electron beams and spoil the colour purity.

Upon installation, and occasionally after the receiver has been moved to another position, the shadow mask and associated ferrous material in the receiver must be demagnetized (or *degaussed*).

Finally, cancellation of any remaining field may be achieved by adjusting a number of small colour equalizer magnets around the bulb, although in later tubes it has been possible to dispense with such equalizer magnets.

Moiré patterns can arise through beat effects between the line scanning structure of the raster and the shadow mask pattern but these are negligible if the scanning lines are parallel to one of the lines into which the shadow mask holes fall, and reach a maximum at $\pm$ 30° from these directions.

The three electron beams are focused electrostatically, all three focus electrodes being connected together. The electron guns are of tetrode construction and the three grid characteristics can be made to match over the grey scale by using high screen potentials to keep the characteristics relatively linear, and then adjusting the screen potentials individually. Typical operating conditions are

Gun screen grid to cathode potential	200 to 500 V
Black-out grid to cathode voltage	-50 to -150 V
Focus voltage	4 to 5 kV
Final anode voltage	20 to 25 kV

The holes in the shadow mask are etched by a photo-engraving process from a master negative. Each shadow mask is then used to lay down its own phosphor screen by a photographic process. The tube face is coated with a photo-resist layer containing one of the colour phosphors, say the red. The mask is placed in its correct position and a point light source, situated at the position from which the red electron beam will eventually appear to be emanating, is used to expose the photo-resist layer. The coating is then washed in a suitable solvent when the exposed dots will have been hardened by the light so that only the unexposed parts are washed away. This process is repeated for the other two colour phosphors.

3.9. Colour cameras

Both image-orthicon and vidicon tubes have been used in colour cameras. Vidicon tubes make a much smaller and lighter camera but suffer from the usual vidicon lag, which produces smear on moving objects at low light levels. The image-orthicons must be operated below the knee of their characteristics so that all three tubes operate over a similarly linear part of their input-output curve. The input-output characteristic of vidicon tubes is never linear; however, its curvature is just about right to correct for the

input-output characteristic of display tubes so that no gamma correction (*see* Section 4.11) is necessary.

The red, green and blue camera tubes are identical in design and are mounted alongside each other in the camera housing (Fig. 3.10). The optical system is arranged so that the light from the objective lens in the turret is split into its primary red, green and blue images by dichroic mirrors and care is taken that the image on each tube is identical in shape and size with its neighbours. Colour filters are used to match exactly the combined spectral responses of the tubes and dichroic mirrors to the required colour character-istics. Neutral density filters are inserted to adjust the red and

N.D. ≡ NEUTRAL DENSITY FILTERS
C ≡ COLOUR CORRECTING FILTERS
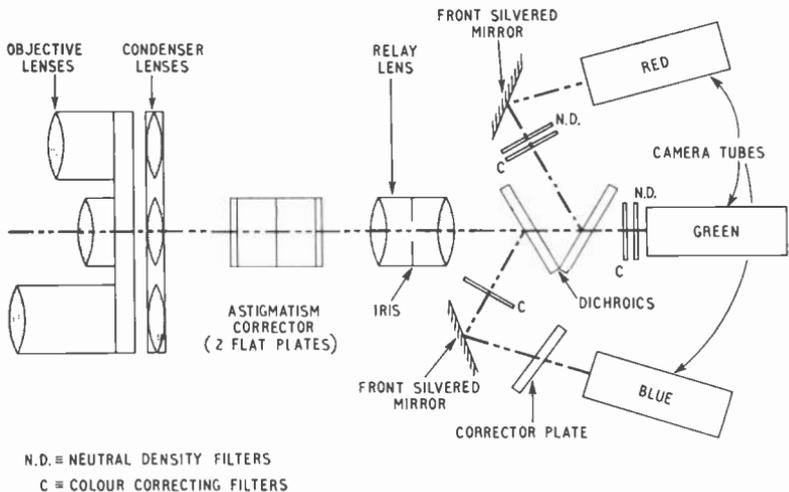
Fig. 3.10.   Schematic plan of an R.C.A. colour camera

green camera sensitivities so that all three tubes operate over similar portions of their transfer characteristics.

Since the dichroic mirrors have a finite thickness, the refraction of a ray of light through them introduces an asymmetry between the vertical and horizontal directions (Fig. 3.11) which is corrected by the insertion of two further plain sheets of glass of the same thick-ness as the dichroic mirrors, and at the same angle, but rotated about the light axis by 90°. An additional corrector is needed in the blue channel so that the blue light passes through the same thickness of glass as the red and green light. Sometimes the

Fig. 3.11.   Asymmetry produced by dichroic mirrors.   Astigmatism correcting plates are rotated through 90° from the dichroic mirrors about the light axis.   (a) shows a horizontal and (b) a vertical section through the dichroic mirror
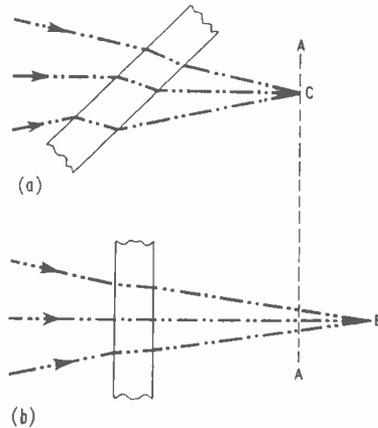
dichroic film of the mirror is itself sandwiched between glass to ensure equality between transmitted and reflected rays.

To leave room for all these mirrors and filters it is necessary to use a relay optical system.   The real image produced by the objective lens is relayed to the mosaics of the camera tube by an auxiliary lens system, with a magnification of unity.   A turret of condenser lenses is mounted in the plane of the real image produced by the objective lenses in order to collect as much light as possible.   The
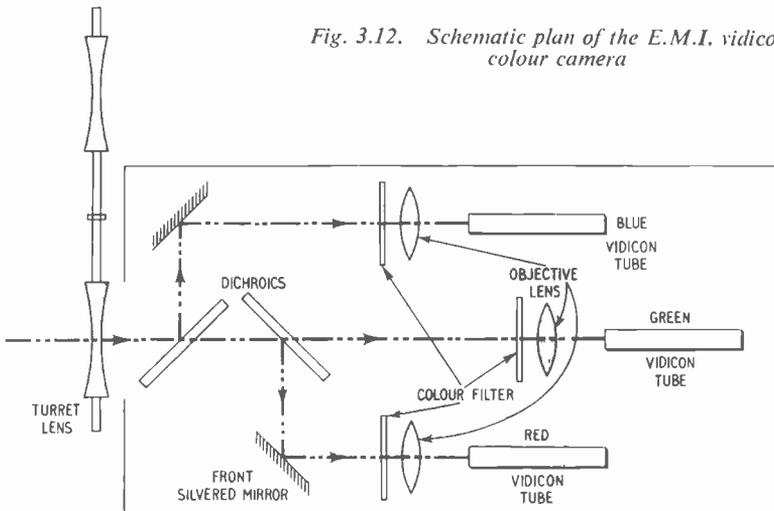
Fig. 3.12.   Schematic plan of the E.M.I. vidicon colour camera

turret of condenser lenses rotates with the objective lens turret but only the objective lens turret moves longitudinally for focusing. The iris diaphragm is mounted in the relay lens system and is remotely controlled.   The temperatures of the three image orthicons are kept within the range 85–120°F.

Studio lighting levels are commonly of the order of 400 to 1,000 foot lamberts, several times the usual monochrome level.

3.10.   Vidicon colour camera

Fig. 3.12 shows a block diagram of an E.M.I. colour camera which uses small vidicon tubes.   The optical system is novel since each vidicon camera tube has its own narrow angle (9°) objective lens.   The dichroic and reflecting mirrors are placed in front of the objective lenses where the narrow angle of field keeps the angle of incidence over the dichroic mirrors fairly constant and astigmatic errors are avoided.

The turret lenses are special colour corrected doublets which convert the angle of view to either 6°, 18° or 29°.

3.11.   Double and single tube cameras

Suggestions have been made for single tube cameras employing a coloured filter strip over the mosaic, similar to the phosphor screen structure of an " Apple " display tube.   Other suggestions have included two-tube cameras, the first tube producing a high definition monochrome picture while the second tube uses a colour grid to provide a low definition colour picture.   So far only the three-tube cameras appear to be practical.

3.12.   Camera circuits

The viewfinder on colour cameras is usually a small monochrome cathode-ray tube accepting either the green signal output or the luminance signal.

Camera amplifier circuits are fairly normal in conception but of necessity are in triplicate.   Special correcting waveforms may be added to the scanning waveforms of one or two of the camera tubes to aid the registration of the three images.   Aperture compensation, shading correction if required, blanking insertion and gamma correction must be carried out as in monochrome practice but always the three signals must be treated similarly.   Particular care must be taken that the three channels black out together.   The standard voltage output is 0·7 volt peak-to-peak across 75Ω for red, green and blue, or 1·0 volt peak-to-peak if synchronizing pulses are added.

3.13. Chromacoder

In this type of studio equipment a single camera tube is used with a rotating disc to provide a field sequential colour signal. The sequential signal is then standards converted to a simultaneous signal for transmission (Fig. 3.13). This type of pick-up has been called *simulquential.*

The field sequential signal does not need to conform to the bandwidth of the transmission standards and much of the objection to field sequential operation can thereby be overcome. The pick-up camera operates at three times the field frequency of the transmission, with three times the line scanning frequency, so that three times the video bandwidth is required in the camera circuits. The resulting
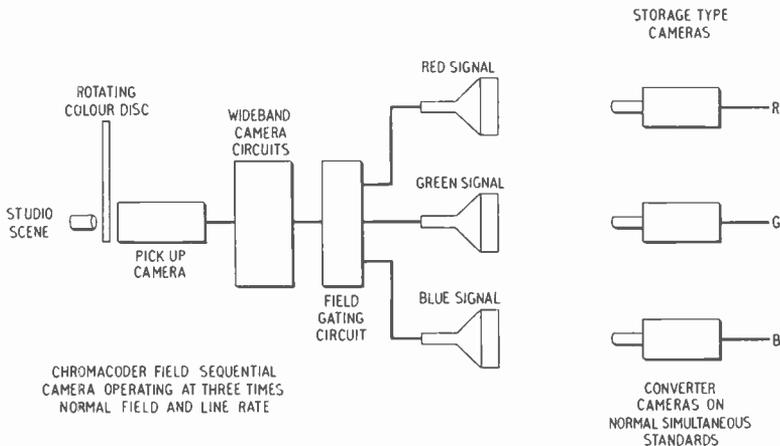
Fig. 3.13. Chromacoder arrangement of cameras and converter tubes

red, green and blue signals are gated out and displayed on separate monitor display tubes. The monitor display tubes are monochrome with a relatively long afterglow and each monitor is viewed by a separate camera operating on the normal simultaneous transmission standards. The combination of long persistence phosphors and storage in the converter camera tubes, which are typically C.P.S. Emitrons, enables the information provided in one high repetition rate chromacoder field to be read off at the slower simultaneous field rate. Thus for 405-line standards the chromacoder would provide a complete red field, scanned in $\frac{1}{150}$ of a second, once every $\frac{1}{50}$ second. This red field would then be read off by the

converter camera in $\frac{1}{50}$ second whilst the other two converter cameras were also reading off their fields.

To overcome moiré patterns the pick-up camera and the monitor tubes may be scanned vertically whilst the converter cameras scan horizontally, as normal.

3.14. Colour film scanners

These are basically similar to monochrome film scanners in the methods used to present the scanning apparatus with a stationary film frame. The film may move continuously and an optical device used to maintain a projected image of a frame stationary in the gate. The long pull-down time required may be offset by only illuminating the film frame for a fraction of a field period and using a storage type camera. The camera viewing arrangements are then similar to the beam splitting arrangements used in colour cameras.

For flying-spot type scanners which have no storage the film image must be kept stationary during the entire scanning process. A very bright image of an unmodulated cathode-ray tube raster is focused on to the film frame, the light passing through the film is collected by a condenser lens system and is then split into red, green and blue beams by an arrangement of dichroic mirrors and filters. Each beam is focused on to a photocell which transforms the light into an electrical signal. Since the phosphor of the cathode-ray tube producing the original raster has an appreciable afterglow, each of the three photocell amplifiers must have decay time correction.

3.15. Colorimetric requirements

Colour television systems are engineered, as far as possible, to reproduce a colorimetric match to the original scene. The colorimetric analysis which the camera performs must therefore be related to the primaries which the receiver is going to use. These primaries and the reference white fix the relative spectral response curves which the red, green and blue cameras must have.

The colour response of a camera channel is the net response of the optical system, the dichroic mirrors, the gelatine colour correcting filters and the camera mosaic. Typical mosaic responses are shown in Fig. 3.14 for the vidicon and image-orthicon colour camera tubes. The resultant overall responses are required to have the same shapes as the colour mixture curves $\bar{r}$, $\bar{g}$, $\bar{b}$, for the receiver primaries (*see* Section 2.23). Thus if the input to the
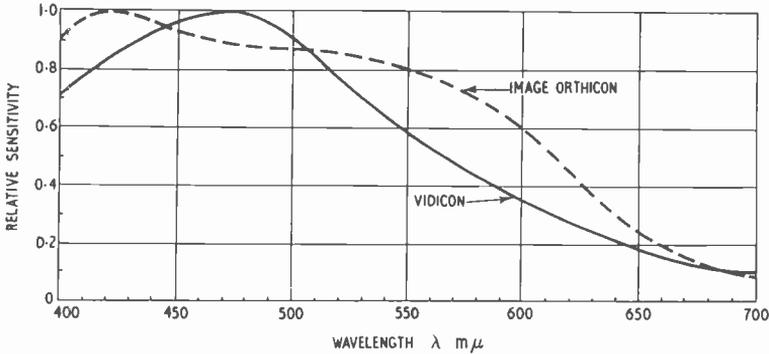
Fig. 3.14. Spectral response curves of two typical colour camera tubes. Illumination is equal energy white

camera is a spectrally pure radiation, of wavelength λ, the receiver will reproduce a match to this colour if the camera outputs are proportional to the ordinates on the colour mixture curves at the wavelength λ. For non-spectral colours the camera mosaics will effectively integrate the product of energy distribution in the light reflected into the camera and the transmission through the camera's colour filters, over the spectrum

$$\int_{380}^{780} \bar{r} \; E(\lambda) \; d\lambda = R$$

so that the camera output is proportional to the required tristimulus value.

The colour mixture curves for the N.T.S.C. primaries and reference white are given in Fig. 2.12. It will be seen that they have appreciable negative lobes. These negative lobes can be simulated by using more than three camera tubes but this is not yet a practical proposition. It has been suggested that the camera spectral responses should follow the C.I.E., $\bar{x}$, $\bar{y}$ and $\bar{z}$ colour mixture curves which have no negative lobes, and the camera outputs matrixed from X, Y and Z, to give the required RGB outputs. Such matrixing results in a worse signal-to-noise ratio.

The present practice is to match the camera response characteristics to the positive parts of the colour mixture curves, with some trimming of the sides of the responses to offset the loss of the negative lobes, as in Fig. 3.15.

The colour mixture curves of Fig. 2.12 and the camera curves of Fig. 3.15 assume that the scene lighting is equal energy white. In
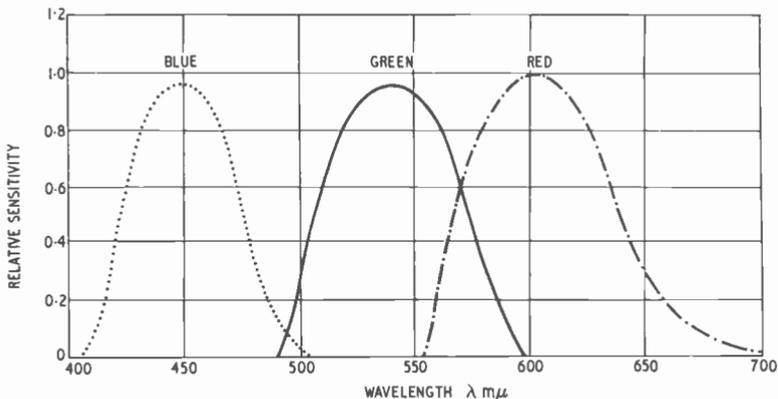
Fig. 3.15.   Practical camera taking responses

principle the filters should be modified for other types of lighting so that the combined effects of the spectral distributions of the lighting and the responses of the camera mosaics, filters, dichroics, etc., total the curves of Fig. 3.15.

3.16.   Electronic masking

Colour films have certain residual colour errors of a systematic nature which arise from imperfections in the dyes used.   These errors are equivalent to a form of crosstalk between the red, green and blue signals and can be partially corrected at the transmitter by introducing a cancelling crosstalk between the three colour signals.   The film errors have a logarithmic variation with brightness but the correction applied at the transmitter is based on linear signals and is only partially effective, although it can be applied after gamma correction has been carried out.

Colour films usually have a higher contrast range than the television system can accommodate and the extra gamma correction needed has the effect of reducing the saturation of the reproduced colours.   Correction can also be made for this desaturating effect.

The process of correction is called *electronic masking* by analogy with the photographic process of masking in which a correction transparency or mask is printed in register with the original film. The electronic masking signal takes the form of a colour difference signal (*see* Section 4.13) which is added to each primary signal.

Thus the masked green signal would be

$$E''_G = K_4 E'_G + (K_1 E'_R + K_2 E'_G + K_3 E'_B)$$

where $K_1 + K_2 + K_3 = 0$, so that the correction signal disappears for greys when $E'_R = E'_G = E'_B$.   The constants can be computed from colorimetric measurements on the film but are usually adjusted on the basis of subjective assessment of the resulting picture quality.

3.17. Summary

The colour camera or film scanner analyses the scene to be transmitted into its three primary colour components and develops an electrical signal proportional to the integrated distribution coefficient of the corresponding receiver primary.   The electrical signals undergo various correction processes before being transmitted.   These processes take account of deficiencies in the original film stock, transmitting equipment, and receiving equipment.

The most successful type of receiving tube is the three-gun shadow mask tube, despite all the external correcting adjustments which are necessary.

CHAPTER **4**

Transmitter Coding

4.1. Introduction

After the scene to be televised has been analysed by the colour
camera, the red, green and blue outputs have to be transmitted in
an agreed form. In the N.T.S.C. system this agreed form is deter-
mined partly by engineering considerations and partly by more
general requirements.

The main part of the signal is similar to a monochrome television
signal and the modulation waveform contains sync pulses combined
with a video signal representing the brightness of the scene being
televised (Fig. 4.1 (a)). This luminance signal will produce a
satisfactory black and white picture on either a monochrome or a
colour receiver and is said to be *compatible*.

To produce a colour picture on a colour receiver the transmitted
signal must contain further information. In the N.T.S.C. system
this further information represents the difference between the colour
of the picture element being transmitted and a white light of the same
luminance or photometric brightness, and this difference becomes
smaller as the colour in the scene approaches a pastel shade or a
grey tone and also becomes smaller as the luminance of the scene
decreases (Fig. 4.1 (b)). This extra information is the *chrominance*
of the scene, described in Section 2.32, and needs two quantities
or numbers to specify it.

The two chrominance parameters are added to the main lumi-
nance signal in such a way as to cause minimum interference to
monochrome reception. They are first modulated onto a carrier
whose frequency falls in the upper part of the video frequency spec-
trum, at 2·7 Mc/s for the 405-line system. The combined signal at
2·7 Mc/s, with sidebands, is then modulated onto the main vision
carrier, which may be 45 Mc/s for example, along with the luminance
signal (Fig. 4.2). The first carrier at 2·7 Mc/s is then called a
sub-carrier. To modulate two quantities onto the one sub-carrier
frequency requires special techniques at the transmitter, and the
detection of such a signal at the receiver necessitates circuits which
are unlike those normally used in monochrome television. Such a

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relatively complex process was developed only after more obvious methods of transmission had been shown to be unsatisfactory.

4.2. Early methods of colour transmission

Since there are three quantities to be transmitted, the simplest solution would be to use three separate television channels (Fig. 4.3), in the same way as the monochrome vision and sound signals

Fig. 4.1. Modulation wave-forms for monochrome and colour. The signals are:
(a) monochrome video and colour luminance;
(b) chrominance;
(c) complete colour signal, luminance plus chrominance
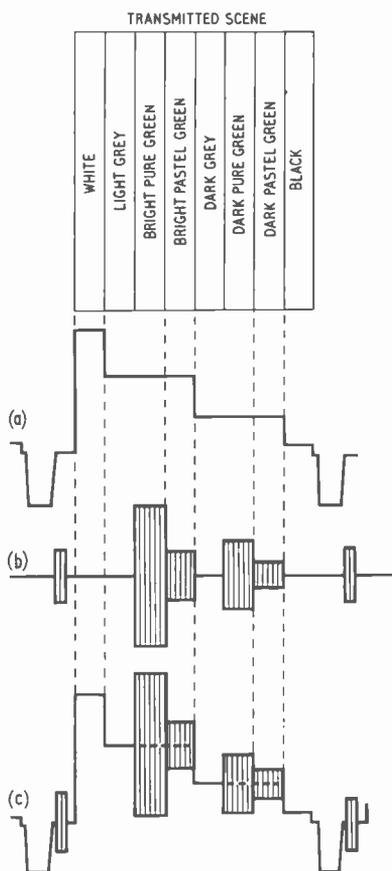
are transmitted separately. This necessitates an increase of rather more than three times in the bandwidth occupied by the vision information. The three transmitted images must be registered accurately at the receiver so that the red picture lies exactly over the green

picture and both must coincide geometrically with the blue image. Further, if the colour balance of the resultant picture is to be consistent, the three channels, including the transmission paths, must have the same gain and identical input-output linearity, for long periods.

4.3. Field sequential systems

To overcome these difficulties, many systems have been proposed using a single channel only and transmitting the red, green and blue
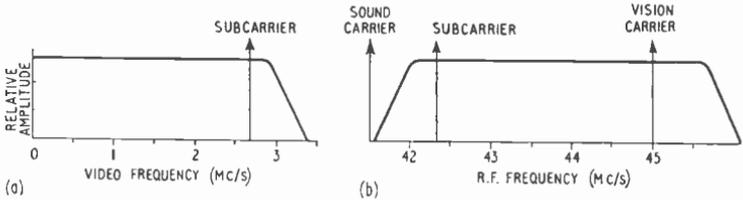
Fig. 4.2.   Addition of sub-carrier to vision signal

information sequentially along the one channel.   Baird demonstrated such a system in 1928 using two rotating discs, each with segments of red, green and blue light filters, rotating in synchronism before the camera tube and the receiver tube (*see* Section 3.5). Each primary coloured filter remained over the tube face for the period of one field.   A more advanced system, developed by the
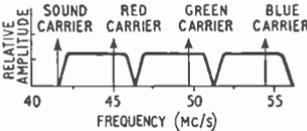
Fig. 4.3.   Three simultaneous colour channels

Columbia Broadcasting System, was adopted in the United States by the Federal Communications Commission as the official American Colour System in 1950, but was discarded in 1951.

Such field sequential systems have two major drawbacks.   If the picture being radiated consists mainly of tones of one hue, such as green, then the red and blue fields are black or nearly so, and each green field is succeeded by two black fields.   This increases the brightness flicker of the reproduced picture.   Even when the transmitted picture consists of grey tones, the luminance of successive fields varies since the system primaries have different luminosity coefficients (*see* Section 2.16).   To overcome such flicker effects

the field scanning rate must be increased by a factor of three, in which case even when only one primary colour is being transmitted, the time between active fields is the same as for monochrome television. Such an increase in field scanning rate requires an equivalent increase of three times the bandwidth, which makes the total channel width similar to that for three-channel simultaneous operation.

A second disadvantage is colour fringing, described in Section 3.5, although the increase in field rate needed to overcome brightness flicker is normally also enough to overcome colour fringing.

4.4. Line sequential systems

To avoid these defects, attempts have been made to send the three primary colours at line sequential rate, that is to say, one line in red, the next in green and so on. This reduces the large area flicker and fringing effects to line effects, but they are still annoying.

4.5. Dot sequential systems

A logical development was to send the colour information at picture element sequential rate, or one picture element in red, the next in green and so on. Each line is then made up of a sequence of red, green and blue dots. Such a dot sequential method of sending the information within the restricted bandwidth of a normal monochrome channel was found to be similar to sending the information simultaneously and led to the use of the sub-carrier technique.

4.6. Bandwidth saving in simultaneous systems

This dual character of the dot sequential system renewed interest in simultaneous systems and, particularly in America, considerable engineering effort was devoted to reducing the wide bandwidth previously required for simultaneous three-channel operation.

4.7. Non-engineering considerations

A public colour television service must provide more than an academic solution to transmitting coloured images from one laboratory to another; it must be economically suited to society. The present national production per day of monochrome receivers sells for about the price of one complete transmitter, including its site and buildings. Simplicity and cheapness of the receiver can be seen to be of paramount importance while complexity and cost at the transmitter are secondary considerations.

The ether space is already overcrowded with communication services of one kind and another and ideally the colour system

should allow colour broadcasting to take place with only a few minor modifications using the same channel frequencies and bandwidths and the same cable and radio links as are in use for monochrome. Such a colour broadcast service ought not to interfere with other radio services. This ideal system should give good black and white reproduction on the millions of present-day monochrome receivers and should allow them to be cheaply converted to colour reception. The quality of the colour picture should satisfy public demand for a few decades, reception difficulties should not be increased and studio and domestic equipment should be reliable and rugged.

Above all, the system adopted should be such as to encourage the development of new techniques and devices and to be capable of absorbing advances in the art as they are made. It is difficult to know when this criterion has been met.

Any change in a widely used public service tends to cause considerable financial expenditure and such changes should, in the public interest, be as infrequent as possible. If the advent of colour broadcasting necessitates any major upheaval in reception equipment it is clearly politic to revise the system on all scores such as resolution, aspect ratio, compatibility with foreign programme sources, etc. Such changes are only worth while if they procure major improvements in picture quality. Systems so far demonstrated represent only compromise solutions to these conflicting requirements.

The N.T.S.C. system is the most successful one yet field tested. It makes use of certain properties of the eye to reduce the amount of information which must be transmitted. The extra information which is needed to change a monochrome picture into a colour picture is transmitted inside the normal channel bandwidth in such a way that it hardly upsets the monochrome picture. Although the amount of this extra colour information is small, the resulting picture does not look blurred as the eye is deceived by the sharp luminance picture which is transmitted.

4.8. Choice of primary colours and white

In any colour television system it is necessary to standardize the primary colours, or reference stimuli, which are used in transmission. The primaries adopted in the N.T.S.C. system have been chosen to allow a wide range of colours to be reproduced and they are colours which can be achieved at the receiver with cathode-ray tube phosphors which are reasonably efficient. The position of the N.T.S.C. red, green and blue primaries on a C.I.E. diagram are shown in
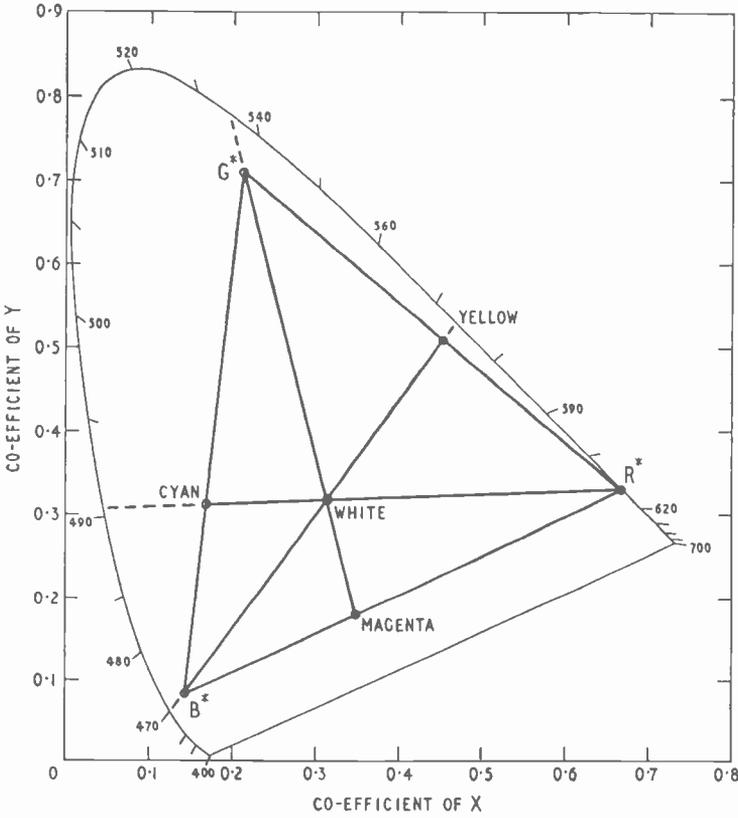
Fig. 4.4. N.T.S.C. system reference stimuli

Table 4.1

N.T.S.C. PRIMARIES AND WHITE

	x	y	z		x	y	z
Red	0·67	0·33	0·00	Cyan	0·17	0·31	0·52
Green	0·21	0·71	0·08	Magenta	0·35	0·18	0·47
Blue	0·14	0·08	0·78	Yellow	0·45	0·51	0·04

The reference white—C.I.E. Illuminant C   0·3101   0·3163   0·3736

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Fig. 4.4.   Any colour which falls inside the triangle R*G*B* can be reproduced; colours outside the triangle can be represented by the correct dominant wavelength but at lower purity (*see* Chapter 2).

The complementaries of these primaries are represented by the points where the line from one primary drawn through the white point Y meets the opposite side of the colour triangle.   These complementary colours are called cyan, magenta and yellow.

4.9.   Receiver white

The white point Y determines the relative size of the units in which R, G and B are measured, these units being such that one unit of red plus one unit of green plus one unit of blue light produces 3 units of the chosen white light Y, which is called the *normalizing illuminant*.   The single unit of red light gives rise to a signal voltage of 1 in the red channel, and similarly for green and blue.   The combination of these three lights produces 3 units of white, but the corresponding luminance signal voltage is unity, as will be seen in Section 4.15.

The particular white chosen for N.T.S.C. receivers is C.I.E. standard Illuminant C, a rather bluish white representing the light from an overcast north sky at midsummer, and not unlike the colour of many modern monochrome television tubes.   When the camera output signals are equal, no sub-carrier is transmitted and the receiver must reproduce Illuminant C.

4.10.   Studio white

It is cheaper and easier to use lighting in the studio which is more red than the equal energy white theoretically assumed.   The practice is to use whatever illuminant is available at the camera and to normalize the camera outputs so that when a nominal white card is held in front of the camera the three outputs are equal and no sub-carrier is transmitted.   On such a signal the receiver reproduces Illuminant C, and the producer monitors the transmission on this assumption.

4.11.   Gamma correction

The light output from cathode-ray tubes is not directly proportional to the modulating grid-cathode voltage, even when constant voltages such as bias and signal set-up have been taken into account.   This non-linearity distorts the tone gradation in monochrome television, and a typical graph of input voltage against light output is shown in Fig. 4.5 (a), where the peak amplitudes of both have been normalized to unity.   The law connecting effective signal voltage V_e—that is

to say, voltage above the black-out bias on the tube—with L, the light output in foot lamberts, is approximately

$$L = k(V_e)^\gamma$$

where normally $\gamma = 2\cdot2 \rightarrow 2\cdot7$

By analogy with photographic work, the exponent is usually represented by the Greek letter gamma (γ).

The transmitter is predistorted to correct for the transfer characteristics of the cathode-ray tube and this applies to both black and white and colour television. The camera output signal is passed through
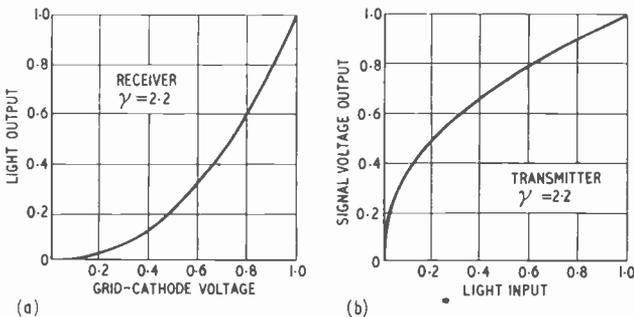
Fig. 4.5. Input-output curves for receiver and transmitter

a gamma corrector circuit with an input-output characteristic, Fig. 4.5 (b) which is the inverse of Fig. 4.5 (a).

The gamma circuit is sometimes called a *rooter* since it carries out the mathematical operation of taking the gamma root $\sqrt[\gamma]{}$, which is approximately the square root, of the incoming voltage.

In N.T.S.C. television, this transmitter distortion leads to some departure from the ideal performance of the system (*see also* Chapters 5 and 16); but it does improve the signal-to-noise ratio in the darker parts of the picture by about 12dB.

N.T.S.C. gamma correction is based on a value of $\gamma = 2\cdot2$. Each red, green and blue camera output is corrected separately, and this distortion is performed after the camera outputs have been normalized on the studio white and before any coding of the signals has taken place, that is to say, before the luminance or chrominance signals have been formed. To indicate that gamma correction has taken place the symbols representing signal voltages are marked with a prime if they refer to circuit points after the gamma correctors; thus G represents a voltage in the green camera

channel before the gamma corrector and G' a voltage which has passed through the correcting circuits.

4.12. Constant luminance principle

Everyday experience as well as scientific measurement indicates that the eye is more conscious of changes in luminance than changes in hue or saturation. Similarly, experiments in colour television have shown that noise which is altering the luminance of the picture is more noticeable than noise which is only altering the chromaticity, and indeed that observers can tolerate an increase in the amount of noise in the colour channel of some 8dB if this is not affecting the brightness of the picture.

As far as possible modern colour systems and receivers take advantage of this by making variations in colour signals produce only variations in chromaticity of the reproduced picture. The luminance of the picture is constant however the other colour signals vary and whatever interference is superimposed on them. This is called the *constant luminance principle*. If one of the three signals is a function only of the luminance of the colour then the receiver can be designed to obey the constant luminance principle.

In the N.T.S.C. system the main signal, termed Y', carries the monochrome or luminance information and is formed by adding fractions of the red, green and blue signals together so that these signals contribute to the brightness signal in proportion to their *luminosity coefficients*. For the N.T.S.C. primaries and illuminant C' these luminosity coefficients are red $= 0.299$, green $= 0.587$ and blue $= 0.114$. For most engineering purposes it is sufficiently accurate to write

$$Y' = 0.30R' + 0.59G' + 0.11B'$$

Since the quantities Y', R', G' and B' represent voltages they are often written E'_Y, E'_R, E'_G and E'_B.

The camera outputs are normalized or adjusted so that they give equal outputs on the studio white, thereby fixing the relative units in which R, G and B are measured. The maximum value of each signal on peak white is 1, so that $0.30R'$ equals 0.30 whilst peak white is being transmitted. For a grey picture which gives half the maximum R', G' and B' signals

$$Y' = 0.30(0.50) + 0.39(0.50) + 0.11(0.50)$$
$$= 0.15 + 0.295 + 0.055$$
$$= 0.50$$

In general, if $R' = G' = B'$, then

$$Y' = 0\cdot30R' + 0\cdot59R' + 0\cdot11R'$$
$$= R' = G' = B'$$

If we give R, the camera signal before gamma correction, the value of $0\cdot5$ then, for a grey picture

$$Y' = 0\cdot30(0\cdot50)^{\frac{1}{2\cdot2}} + 0\cdot59(0\cdot50)^{\frac{1}{2\cdot2}} + 0\cdot11(0\cdot50)^{\frac{1}{2\cdot2}}$$
$$= 0\cdot30(0\cdot73) + 0\cdot59(0\cdot73) + 0\cdot11(0\cdot73)$$
$$= 0\cdot73$$

which is different from the value of Y' when $R' = R^{\frac{1}{2\cdot2}}$ and $= 0\cdot50$.

The actual value in volts of any of these signals depends on the point in the system at which they are measured. Thus a peak Y' signal which may be 1 volt when it is formed from R', B' and G' may be represented by a signal of 1 mV at the receiver aerial, by a few volts at the detector, and by a hundred or so volts at the cathode-ray tube. The relative values of Y', R', G', and B', are carefully maintained, however, and it is convenient to relate them to the corresponding peak signal value so that the amplifier gains can be ignored.

For a pure red colour, $G' = B' = 0$, and R' will vary between 0 and 1 depending on the luminance of the red; the corresponding Y' signal will vary between 0 and 0·30.

4.13. Colour difference signals

The information which must be transmitted apart from Y' is first derived as two colour difference signals $(R' - Y')$ and $(B' - Y')$, obtained by subtracting the derived Y' signal from the red and blue camera output voltages.

In the three signals, Y', $(R' - Y')$ and $(B' - Y')$ there is sufficient information for the receiver to calculate R', G' and B', which are the voltages it needs to apply to the display tube, since

$$R' = (R' - Y') + Y' \qquad [4.1]$$

$$B' = (B' - Y') + Y' \qquad [4.2]$$

and since

$$Y' = 0\cdot30R' + 0\cdot59G' + 0\cdot11B' \qquad [4.3]$$
$$= 0\cdot30Y' + 0\cdot59Y' + 0\cdot11Y'$$

and hence

$$0 = 0 \cdot 30(R' - Y') + 0 \cdot 59(G' - Y') + 0 \cdot 11(B' - \overset{\backprime}{Y}') \quad [4.4]$$

that is

$$G' - Y' = -\frac{0 \cdot 30}{0 \cdot 59}(R' - Y') - \frac{0 \cdot 11}{0 \cdot 59}(B' - Y')$$

and

$$G' = Y' - \frac{0 \cdot 30}{0 \cdot 59}(R' - Y') - \frac{0 \cdot 11}{0 \cdot 59}(B' - Y') \quad [4.5]$$

$$= Y' - 0 \cdot 51(R' - Y') - 0 \cdot 19(B' - Y') \quad [4.6]$$

From both Equation 4.3 and 4.6 it is clear that

$$G' = 1 \cdot 70 Y' - 0 \cdot 51 R' - 0 \cdot 19 B' \quad [4.7]$$

Notice that with these three signals, Y', $(R' - Y')$ and $(B' - Y')$, for a peak white signal in which

$$R' = G' = B' = 1$$

then $Y' = 0 \cdot 30(1) + 0 \cdot 59(1) + 0 \cdot 11(1) = 1$

so that $B' - Y' = 1 - 1 = 0$

and $R' - Y' = 1 - 1 = 0$

and the colour difference signals disappear, as indeed they do for any amplitude of white such as

$$R' = G' = B' = \tfrac{1}{4}$$

This illustrates one advantage of using colour difference signals; for a monochrome scene only the luminance signal needs to be transmitted and for desaturated colours the colour difference signals are small, which improves the compatibility of the signal. In practice, for 90 % of the time it is found that the colour difference signals have less than one-third of their maximum values. Similarly, colour difference operation also has the advantage that changes in relative gain between the three transmitted signals do not affect the colour balance of the grey scale, since the colour difference signals are zero for greys. Before they are finally transmitted, the colour difference signals are reduced in bandwidth and the receiver has to derive a low-definition colouring picture to superimpose on

the sharp luminance picture. Colour difference signals have an advantage here in that the receiver design is simplified compared with transmission of a wide band Y', and, say, narrow band R' and B' signals. In the latter case, the derivation of a combined high definition luminance signal and a low definition colour signal for each primary colour is more involved, because, for example, if the R' signal were applied directly to the red display, the high frequency luminance components carried by the Y' signal would be absent. One way to derive the required signal would be to add to the low definition R' signal those high frequency components of Y' which fall outside the R' pass-band, but this requires careful tailoring of the frequency response shapes in the overlap region. A better method of obtaining the combined signal would be to form $(R' - Y')$ and $(B' - Y')$ signals at the receiver, bandwidth-limit them, and then apply them to the displays, together with the high definition Y' signal. Obviously, the transmission of bandwidth-limited $(R' - Y')$ and $(B' - Y')$ signals avoids this complication at the receiver. It would be possible to transmit any two difference signals such as $(R' - Y')$ and $(G' - Y')$. However, it will be found that the green difference signal for any colour is always smaller than, or at most equal to, either the red or the blue difference signal for that colour. Hence, the use of $(G' - Y')$ in transmission would require gain at the receiver for the derivation of the third difference signal, with a corresponding deterioration in noise performance.

The values of the colour difference signals are not always those which one might superficially expect for a particular colour. For a saturated red signal, in which $G' = B' = 0$, and $R' = 1$

$$Y' = 0 \cdot 30$$
$$\text{and} \quad (B' - Y') = -0 \cdot 30$$
$$\text{and} \quad (R' - Y') = 0 \cdot 70$$

Table 4.2 includes example values of the three colour signals for various typical cases; the I and Q values will be explained in Section 4.16. The reader will find it instructive to check the arithmetic in this table and to work out a few examples for himself. Notice that the blue difference signal $(B' - Y')$ does not vanish for colours which contain no blue component, but is negative. For a bright pure blue colour, the $(R' - Y')$ signal does not disappear but is $(0 - 0 \cdot 11) = -0 \cdot 11$. Note also that the desaturated red taken as an example in Table 4.1 may be considered as a grey signal of value $R' = G' = B' = \frac{1}{2}$ plus a saturated red in which

Table 4.2

TYPICAL SIGNAL VALUES

Colour	Camera Voltages			Y'	Colour Difference Voltages		Transmission Signals		Sub-carrier	
	R'	G'	B'		$(R'-Y')$	$(B'-Y')$	I'	Q'	Amplitude $\sqrt{(I')^2+(Q')^2}$	Phase $\left(\tan^{-1}\dfrac{I'}{Q'}\right)+33°$
White	1	1	1	1	0	0	0	0	0	
Grey	0·2	0·2	0·2	0·2	0	0	0	0	0	
Black	0	0	0	0	0	0	0	0	0	
Bright Pure Red	1	0	0	0·30	0·70	-0·30	0·60	0·21	0·63	103°
Less Bright Pure Red	0·5	0	0	0·15	0·35	-0·15	0·30	0·11	0·32	103°
Bright Pastel Red	1	0·5	0·5	0·65	0·35	-0·15	0·30	0·11	0·32	103°
Dark Pastel Red	0·2	0·1	0·1	0·13	0·07	-0·03	0·06	0·02	0·06	103°
Bright Pure Magenta	1	0	1	0·41	0·59	0·59	0·28	0·52	0·59	61°
Dark Pure Magenta	0·3	0	0·3	0·12	0·18	0·18	0·08	0·16	0·18	61°
Bright Pure Green	0	1	0	0·59	-0·59	-0·59	-0·28	-0·52	0·59	-119°
Bright Pure Blue	0	0	1	0·11	-0·11	0·89	-0·32	0·31	0·44	-13°
Bright Pure Cyan	0	1	1	0·70	-0·70	0·30	-0·60	-0·21	0·63	-77°
Bright Pure Yellow	1	1	0	0·89	0·11	-0·89	0·32	-0·31	0·44	167°

$R' = \frac{1}{2}$, $G' = B' = 0$, and that the colour difference signals are the same as for the example of a medium bright saturated red in which $R' = \frac{1}{2}$ and $G' = B' = 0$. Notice that a saturated red of high luminance gives rise to larger colour difference signals than the same red at a lower luminance, although the chromaticity of the colour has not altered.

4.14. Constant luminance

If there is no gamma correction this type of colour difference signal satisfies the constant luminance principle. If a noise signal voltage is added to the colour difference signals it does not affect the luminance of the reproduction but only the chromaticity. As an example, consider the case of a colour represented by

$$0 \cdot 70R^* + 0 \cdot 40G^* + 0 \cdot 50B^*$$

which gives rise to camera voltages of $R = 0 \cdot 70$, $G = 0 \cdot 40$ and $B = 0 \cdot 50$ and to transmitted signals, which for clarity in this particular case are indicated by a subscript t, of

$$Y_t = 0 \cdot 30R + 0 \cdot 59G + 0 \cdot 11B$$
$$= 0 \cdot 30(0 \cdot 70) + 0 \cdot 59(0 \cdot 40) + 0 \cdot 11(0 \cdot 50)$$
$$= 0 \cdot 50$$
$$(R - Y)_t = (0 \cdot 70 - 0 \cdot 50) = 0 \cdot 20$$
$$(B - Y)_t = (0 \cdot 50 - 0 \cdot 50) = 0$$

The addition of a noise voltage of, say, $0 \cdot 10$ to either or both colour difference signals will not affect the luminance of the received colour.

The new signals may be

$$Y_t = 0 \cdot 50, \quad (R - Y)_t = 0 \cdot 30 \quad \text{and} \quad (B - Y)_t = 0 \cdot 10$$
$$\therefore \quad R = (R - Y)_t + Y_t = 0 \cdot 80$$
$$B = (B - Y)_t + Y_t = 0 \cdot 60$$
$$G = Y_t - 0 \cdot 51(R - Y)_t - 0 \cdot 19(B - Y)_t$$
$$= 0 \cdot 50 - 0 \cdot 51(0 \cdot 30) - 0 \cdot 19(0 \cdot 10)$$
$$= 0 \cdot 33$$

These give rise to a received colour of luminance

$$0 \cdot 30R + 0 \cdot 59G + 0 \cdot 11B$$
$$= 0 \cdot 30(0 \cdot 80) + 0 \cdot 59(0 \cdot 33) + 0 \cdot 11(0 \cdot 60)$$
$$= 0 \cdot 50$$

which is the luminance of the original colour without the noise voltages added.

In general, for transmitted signals Y_t, $(R - Y)_t$ and $(B - Y)_t$, the receiver can deduce that

$$R = (R - Y)_t + Y_t$$
$$B = (B - Y)_t + Y_t$$
$$G = Y_t - 0 \cdot 51(R - Y)_t - 0 \cdot 19(B - Y)_t$$

<div align="right">from Equation 4.6</div>

and these values give rise in a linear system to a colour of luminance
$0 \cdot 30R + 0 \cdot 59G + 0 \cdot 11B$

or

$$0 \cdot 30(R - Y)_t + 0 \cdot 30Y_t + 0 \cdot 59Y_t - 0 \cdot 30(R - Y)_t - 0 \cdot 11(B - Y)_t$$
$$+ 0 \cdot 11(B - Y)_t + 0 \cdot 11Y_t$$
$$= Y_t$$

whatever variations occur in $(R - Y)_t$ and $(B - Y)_t$.

It is possible to obtain constant luminance operation without using colour difference signals. For example, if the three transmitted signals were Y_t, R_t and B_t, then for the same colour as used in the previous case, $Y_t = 0 \cdot 50$, $R_t = 0 \cdot 80$ and $B_t = 0 \cdot 60$ if a noise voltage of $0 \cdot 1$ is added to R_t and to B_t.

The receiver then works out that

$$G = 1 \cdot 70(0 \cdot 50) - 0 \cdot 51(0 \cdot 80) - 0 \cdot 19(0 \cdot 60) \qquad [4.7]$$
$$= 0 \cdot 85 - 0 \cdot 41 - 0 \cdot 11$$
$$= 0 \cdot 33$$

Then as before the reproduced luminance on a linear system would be

$$0 \cdot 30(0 \cdot 80) + 0 \cdot 59 (0 \cdot 33) + 0 \cdot 11(0 \cdot 60) = 0 \cdot 50$$

The essence of constant luminance operation is thus to transmit a monochrome signal which is accurately a function of luminance, and of luminance only.

In practice the system departs from this ideal constant luminance operation because of the need to gamma correct the signals. The subjective addition of the brightness of the three primary components is only carried out after they have been raised to the power γ, and the
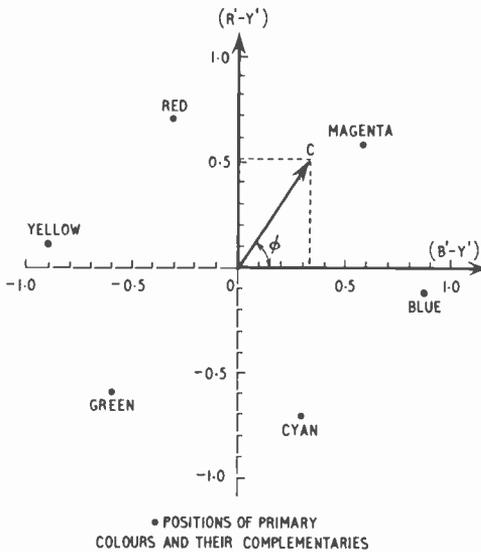
Fig. 4.6. Colour difference axes

examples quoted are therefore only approximately accurate. This departure from constant luminance operation is discussed further in the next chapter.

The chrominance signal which is finally transmitted in the N.T.S.C. system is obtained after the $(R' - Y')$ and $(B' - Y')$ signals have both been reduced in amplitude.

4.15. Reduction of colour difference signal amplitudes

Describing the chrominance information by two quantities $(R' - Y')$ and $(B' - Y')$ makes it convenient to represent the colour being transmitted by a diagram such as Fig. 4.6. The point C indicates a colour whose chrominance is measured by the projections of OC on the $(R' - Y')$ and $(B' - Y')$ axes.

$$(R' - Y') = OC \sin \phi$$

$$(B' - Y') = OC \cos \phi$$

The particular colour illustrated is a magenta colour. From Table 4.1 the chrominance signals corresponding to the other colours of the spectrum can be plotted and the positions of these primary colours are indicated in Fig. 4.6. The angle the line OC makes
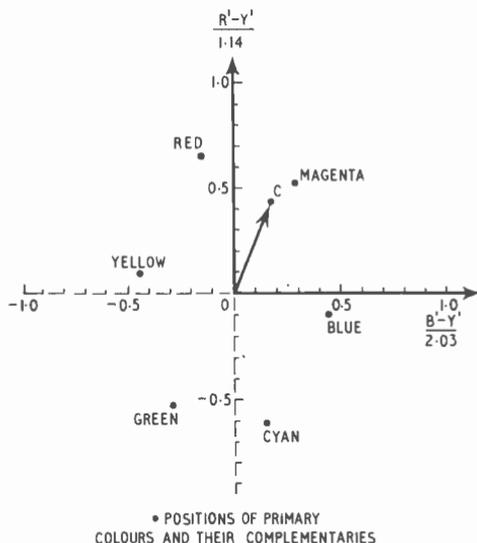
Fig. 4.7. Weighted colour difference axes

with the $(R' - Y')$ and $(B' - Y')$ axes indicates the dominant wavelength of the colour, but the length of OC increases with both luminance and the purity of the colour.

It is expedient to change the scale along the $(B' - Y')$ axis by plotting $\dfrac{1}{2 \cdot 03} (B' - Y')$ rather than $(B' - Y')$ itself, in order to reduce the amplitude of the transmitted chrominance signal. Similarly, the scale along the $(R' - Y')$ axis is changed by plotting $\dfrac{1}{1 \cdot 14} (R' - Y')$ instead of $(R' - Y')$. Such a reduction in chrominance signals avoids undesirable overload effects at the transmitter. The considerations which led to the choice of such reduction factors as $\dfrac{1}{2 \cdot 03}$ rather than a half are discussed in Section 5.4.

The diagram of Fig. 4.6 is replotted using these new scales in Fig. 4.7. Notice that the positions of the primary colours are now different although they correspond to the same values of $(R' - Y')$ and $(B' - Y')$. It is important to check the scale along any diagrams with $(R' - Y')$ and $(B' - Y')$ axes to see whether the reduction factors $\dfrac{1}{1\cdot14}$ and $\dfrac{1}{2\cdot03}$ have been used or not, since all the angles and lengths on the diagram are generally different for the two cases. This reduction in amplitude of $(R' - Y')$ and $(B' - Y')$ is called the weighting of the difference signals.

4.16. Choice of colour axes

As well as reducing the amplitude of the chrominance signals it is also necessary to reduce their bandwidth in order that they may be superimposed on the luminance signal without causing undue interference. Section 4.21 deals with the choice of bandwidths for the two chrominance signals. It is advantageous to reduce the width of one chrominance signal more than that of the other signal.

Fig. 4.8. I' and Q' axes
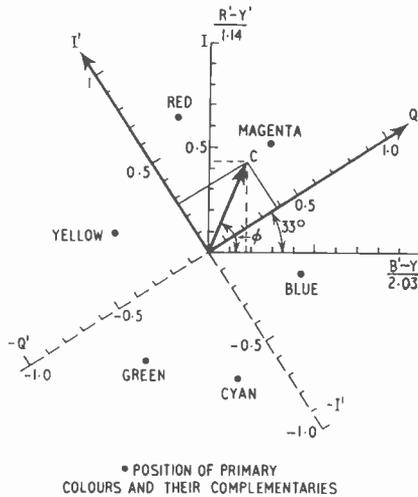
• POSITION OF PRIMARY
COLOURS AND THEIR COMPLEMENTARIES

The $(B' - Y')$ axis could be made the narrow band axis but from the data in Section 1.14 it is reasonable to make the narrow band axis lie amongst those colours which the eye is first unable to distinguish as the size of the coloured area decreases. If the eye cannot see changes in these colours in small areas, then it will not notice, from a reasonable viewing distance, that such changes are

not being transmitted.   The narrow band axis is chosen to lie among
the green-magenta colours.   The exact choice of position is not
critical and the ideal axis varies with the luminance changes associated
with the small coloured areas; it has been standardized as an axis
which makes an angle of 33° anticlockwise from the $\frac{1}{2 \cdot 03}$ $(B' - Y')$
axis, as in Fig. 4.8.   This new axis is called the Q' axis.   The
second chrominance axis is chosen at right angles to the Q' axis
and is called the I' axis.   The I' axis then lies among the orange cyan
colours for which the eye has maximum chromaticity discrimination
in medium-small areas.   This sensitivity to chromaticity changes
does not vary quickly with the position of the I' axis and the choice
of the I' axis perpendicular to the Q' axis is quite acceptable.   On
the $(R' - Y')$ and $(B' - Y')$ axes of Fig. 4.6 the I' and Q' axes
make angles of 139° and 20° with the $(B' - Y')$ axis and have scales
of $1 \cdot 5I'$ and $1 \cdot 8Q'$; the importance of checking the scale used for
$(B' - Y')$ and $(R' - Y')$ is thus underlined.

The colour C of Fig. 4.8 can be represented either by its projec-
tions on the old $\frac{1}{2 \cdot 03}$ $(B' - Y')$ and $\frac{1}{1 \cdot 14}$ $(R' - Y')$ axes, or by its
projections on the new I' and Q' axes, and there is a direct relation
between the values of I' and Q' for a colour C and the values of
$(R' - Y')$ and $(B' - Y')$ for the same colour.

The chrominance component along the new Q' axis can be con-
sidered as the sum of the projections of the $\frac{1}{1 \cdot 14}$ $(R' - Y')$ and
$\frac{1}{2 \cdot 03}$ $(B' - Y')$ components along the Q' axis.

$$\therefore \qquad Q' = \frac{(R' - Y')}{1 \cdot 14} \sin 33° + \frac{(B' - Y')}{2 \cdot 03} \cos 33° \qquad [4.8]$$

Similarly

$$I' = \frac{(R' - Y')}{1 \cdot 14} \cos 33° - \frac{(B' - Y')}{2 \cdot 03} \sin 33° \qquad [4.9]$$

The negative sign occurs because the $(B' - Y')$ component lies
along the negative part of the I' axis.

4.17. Modulation of the sub-carrier

The two I' and Q' chrominance signals are transmitted by modulat-
ing them onto a sub-carrier.   Two sine waves are generated at the
same frequency but differing in phase by 90° so that the phasors

representing these two sine waves (Fig. 4.9) could well be drawn along the $I' - Q'$ axes in Fig. 4.8.   One phasor is then amplitude modulated with the I' information, and the other phasor is amplitude modulated with the Q' signal.   This modulation is carried out in balanced modulators (*see* Chapter 6) so that if there is no I' signal,
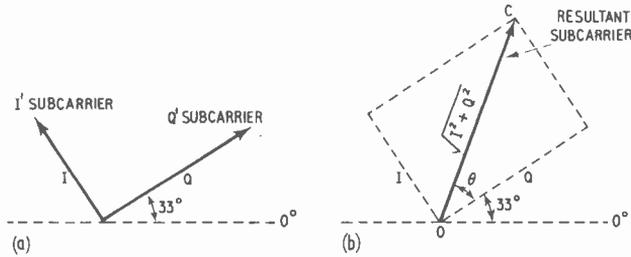
Fig. 4.9.   *Resultant of two sub-carriers*
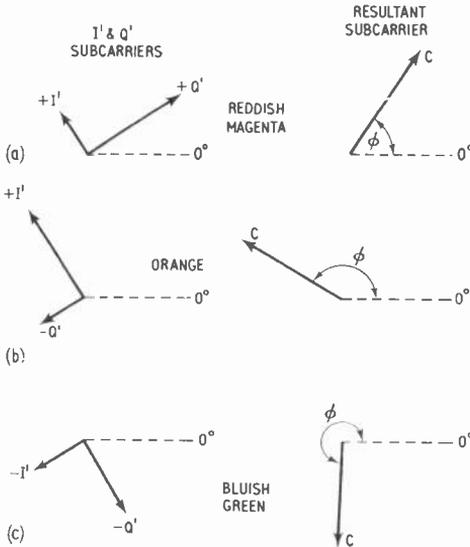
Fig. 4.10.   *Equivalent sub-carrier representations*

the sub-carrier at I' phase has zero amplitude, and if $Q' = 0$ as well then both sub-carriers disappear.   In this way the D.C. value of the I' signal is represented by the mean of the peak amplitude of the sub-carrier at I' phase and the modulation sidebands of this sub-carrier represent the variations of I' about this D.C. level. Negative values of I' correspond to a reversal in direction of the phasor or a

phase change of 180° in the I' sub-carrier.　This method of operation in which the carrier has no standing value when the modulation is zero, unlike the modulation of the sound carrier, is sometimes called *suppressed carrier working*.

4.18.　Representation by a single sub-carrier.

Since the I' and Q' sub-carriers have the same frequency, they can be represented by a single resultant phasor of the same frequency, as in Fig. 4.9.　Variations in the amplitude of the I' and Q' components make the resultant sub-carrier vary in both amplitude and phase (Fig. 4.10 gives examples of reddish magenta, orange and bluish-green), and this type of modulation can be considered in two ways.　A single sub-carrier can be imagined as conveying amplitude modulation proportional to $\sqrt{(I')^2 + (Q')^2}$ and phase modulation proportional to $\tan^{-1} \dfrac{I'}{Q'}$. Trigonometrically the resultant sub-carrier OC can be written as the sum of a Q' signal represented by

$$Q' \sin (\omega t + 33°),$$

and the I' signal represented by

$$I' \sin (\omega t + 33° + 90°) \text{ or } I' \cos (\omega t + 33°)$$

i.e. $OC = I' \cos (\omega t + 33°) + Q' \sin (\omega t + 33°)$

$$= \sqrt{(I')^2 + (Q')^2} \left[\frac{I'}{\sqrt{(I')^2 + (Q')^2}} \cos (\omega t + 33°) \right.$$
$$\left. + \frac{Q'}{\sqrt{(I')^2 + (Q')^2}} \sin (\omega t + 33°) \right]$$
$$= \sqrt{(I')^2 + (Q')^2} \left[\sin \phi \cos (\omega t + 33°) + \cos \phi \sin (\omega t + 33°) \right]$$
$$= \sqrt{(I')^2 + (Q')^2} \sin (\omega t + 33° + \phi)$$

and
$$\tan \phi = \frac{I'}{Q'}$$

Alternatively, the signal can be considered as composed of two sub-carriers each of which carries only amplitude modulation, and these two sub-carriers are 90° apart in phase.

4.19.　Zero phase reference

As the phase of the sub-carrier is important, it is necessary to standardize on a reference phase.　The phase of the positive direction of the $(B' - Y')$ axis is called zero phase, and angles are

measured anticlockwise from this phase, so that the positive I' axis is at phase $123°$ (*see* Fig. 4.8). The $(B' - Y')$ and $\frac{1}{2 \cdot 03}(B' - Y')$ axes lie along the same direction.

4.20. Video signal

The sub-carrier signal is added to the luminance signal and the mixed sync signal so that the complete video signal is a combination of all three, as is shown in Fig. 4.1. The sub-carrier signal thus rides up and down on the luminance signal. Viewed on an oscilloscope synchronized to the line scanning frequency the sub-carrier, where it is present, appears as a blur, which is represented in Fig. 4.1 by the rectangular outline of the chrominance waveform. The sub-carrier is a high video frequency and even if the colour it represents is constant from line to line, the phase of the sub-carrier relative to the line sync pulse reverses on every line (see Section
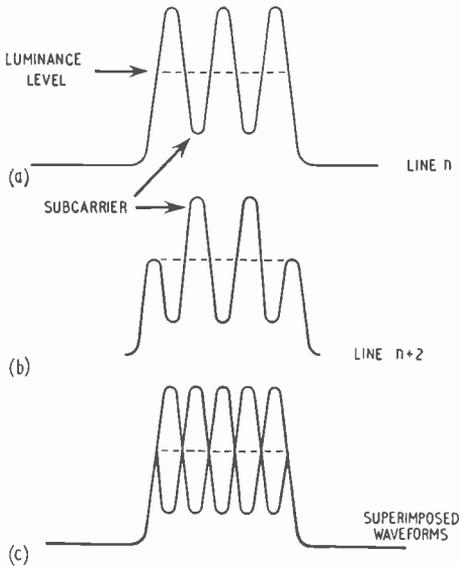
Fig. 4.11. *Video waveforms for vertical colour bar*

4.32) so that only the envelope of the waveform is clearly visible unless a high speed scan is used to pick out the chrominance, in which case the detailed picture of the waveforms is as represented in Fig. 4.11.

8

4.21. Bandwidths of the chrominance signals

To reduce the visibility of the interference caused by putting the sub-carrier in the frequency band which is normally occupied only by luminance information (*see* Fig. 4.2), the sub-carrier frequency is chosen to be a high luminance frequency. This necessarily implies that the sideband frequencies of the sub-carrier are limited on the sound carrier side. Thus for a sub-carrier frequency of 2·7 Mc/s
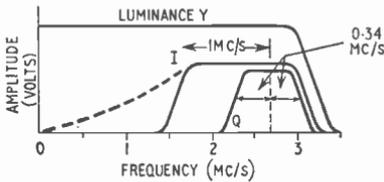
Fig. 4.12. N.T.S.C. bandwidths. Dotted line is upper limit of transmitter tolerance on I *bandwidth*

Table 4.3

N.T.S.C. BANDWIDTHS

Q bandwidth			I bandwidth		
525 and 625 lines	**405 lines**		**525 and 625 lines**	**405 lines**	
400 kc/s	300 kc/s	< −2dB	1·3 Mc/s	1 Mc/s	< −2dB
500 kc/s	340 kc/s	< −6dB	3·6 Mc/s	2·5 Mc/s	⩾ −20dB
600 kc/s	450 kc/s	⩾ −6dB			

on the 405-line system, the sub-carrier sidebands can only extend up to about 0·3 Mc/s in the high frequency direction, i.e., 3·0 to 2·7 Mc/s.

If the sub-carrier is to be modulated double sideband, this limits the rate at which the chrominance information can be changed to 0·3 Mc/s for both colour difference signals. This means that there can be no sharp vertical colour boundaries and the sharpest change from an orange colour to a cyan colour of the same luminance will spread along about $\frac{1}{4}$ in. on a horizontal line of a 21 in. receiver. On most pictures, the colour changes will occur in conjunction with luminance changes and the sharp edge of the luminance change will camouflage the unsharp chromaticity change (*see* Section 1.14). Nevertheless the wider the bandwidth which can be allotted to the chrominance signal, the sharper the final picture will appear. This

bandwidth restriction does not affect the sharpness of colour changes in a vertical direction.

On the low frequency side of the sub-carrier there is room for a much wider modulation bandwidth but this would mean using asymmetric sideband modulation as is used for the main vision carrier. Such unbalanced transmission produces an unwanted signal which has a phase at right angles to the wanted signal, that is to say, *quadrature crosstalk* occurs. This is discussed further in Chapter 8. The result is that an asymmetric Q' signal would produce a spurious signal at right angles in the I' phase.

To avoid this crosstalk the Q' signal is sent double sideband. The I' signal is sent asymmetric (*see* Fig. 4.12) but is double sideband over the whole frequency band of the Q' signals and does not therefore crosstalk into the Q' signal over this frequency band. Outside this frequency band the I' signal is asymmetric but the quadrature crosstalk components due to this part of the signal fall outside the Q' pass-band and are rejected by the Q' channel. In this way the two signals can be received without mutual interference.

The transmitter response specifications are

I' **channel**	Q' **channel**
at 1·0 Mc/s less than 2dB down	at 300 kc/s less than 2dB down
at 2·5 Mc/s at least 20dB down	at 340 kc/s less than 6dB down
	at 450 kc/s at least 6dB down

The system characteristics so far discussed are common to the American N.T.S.C. and the versions adapted (A.N.T.S.C.) to both 405 and 625 lines. The bandwidths, however, vary from one system to another. Thus, for 525 lines

I' **channel**	Q' **channel**
at 1·3 Mc/s less than 2dB down	at 400 kc/s less than 2dB down
at 3·6 Mc/s at least 20dB down	at 500 kc/s less than 6dB down
	at 600 kc/s at least 6dB down

For 625 lines the suggested chrominance bandwidths are the same as for the American system. If an 8 Mc/s channel is available for each station, as in the U.S.S.R., the optimum utilization of the extra megacycle of bandwidth is probably to increase the luminance pass-band by 0·5 megacycle and use the other 0·5 megacycle to increase the vestigial sideband of the vision carrier to 1·25 Mc/s. The latter reduces the effects of vestigial sideband distortion,

particularly on negative modulation. If the sub-carrier frequency is left unaltered, the extra luminance bandwidth can be used to increase the chrominance bandwidth and to ease the problem of providing a sharp luminance and chrominance cut-off at the sound channel frequency.

4.22. Time matching of signals

Since the bandwidths of the Y', I' and Q' signals are different, a sharp transition in the picture will give rise in general to a sharp transition in the Y' signal, a slow transition in the I' channel, and a
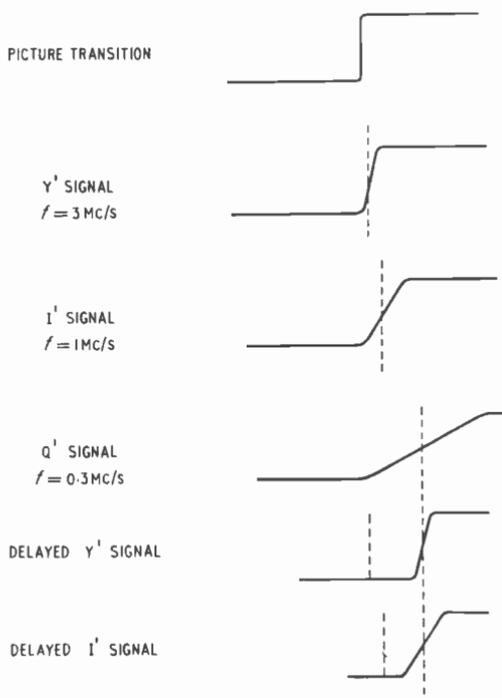
Fig. 4.13. Time matching of signals. Build-up time, 10–90%, 1/2f. Transmitter tolerance is $\pm\,0{\cdot}07\mu s$ (405 lines) and $\pm0{\cdot}05\mu s$ (525 lines)

still slower change in the Q' channel, as in Fig. 4.13. The fastest possible rate of rise, measured from 10% to 90% of the amplitude change, is given by $T = \dfrac{1}{2f}$ where f is the bandwidth of the channel concerned. The time delay of each signal depends upon the band-

width of the channel through which it is passing, and the attenuation slope at the cut-off frequency.

For these three transitions to appear to be coincident in time, the centres of the transitions should coincide. This is achieved by delaying the I' and Y' signals with respect to the Q' signal. The BBC transmitter specification for this time matching is that the signals as transmitted shall be coincident in time; but *see* Section 6.3.

4.23. The final chrominance signal

The chrominance signal as it is finally transmitted consists of two components, I' and Q', amplitude modulated onto two sub-carriers which are 90° apart in phase and can therefore be represented as in Fig. 4.8.

The amplitudes of the I' and Q' signals are calculated from the reduced values of the colour difference signals

$$\frac{(R' - Y')}{1 \cdot 14} \quad \text{and} \quad \frac{(B' - Y')}{2 \cdot 03}$$

according to the Equations 4.8 and 4.9, which are

$$Q' = \frac{(R' - Y')}{1 \cdot 14} \sin 33° + \frac{(B' - Y')}{2 \cdot 03} \cos 33° \qquad [4.8]$$

and

$$I' = \frac{(R' - Y')}{1 \cdot 14} \cos 33° - \frac{(B' - Y')}{2 \cdot 03} \sin 33° \qquad [4.9]$$

since $\cos 33° = 0 \cdot 839$ and $\sin 33° = 0 \cdot 545$

$$I' = 0 \cdot 736(R' - Y') - 0 \cdot 268(B' - Y') \qquad [4.10]$$

$$Q' = 0 \cdot 478(R' - Y') + 0 \cdot 413(B' - Y') \qquad [4.11]$$

The combined resultant of these two sub-carriers, (Section 4.18) has an amplitude $\sqrt{(I')^2 + (Q')^2}$ and makes an angle $33° + \tan^{-1}\dfrac{I'}{Q'}$ with respect to the $\dfrac{(B' - Y')}{2 \cdot 03}$ axis. The $\dfrac{(B' - Y')}{2 \cdot 03}$ axis lies along the old $(B' - Y')$ axis but the scale is different.

Table 4.1 gives sample values of $(I')^2 + (Q')^2$ and the phase angle $\tan^{-1}\dfrac{I'}{Q'} + 33° = \phi$ for various colours. The reader should work through some of these examples.

A sub-carrier resultant of specified amplitude OC and phase ϕ can be resolved into its components along any chosen pair of axes, as well as along the I', Q' or $\dfrac{(R' - Y')}{1\cdot14}$ and $\dfrac{(B' - Y')}{2\cdot03}$ axes, provided the modulation frequencies being considered fall inside the region where I' and Q' are both double sideband (*see* Chapter 8).

4.24. The combined luminance and chrominance signal

The preceding sections of this chapter have laid the basis for describing the final formulation of the N.T.S.C. video signal.

The luminance signal is first formed from the red, green and blue camera outputs, after the camera amplifier gains have been normalized on studio white and the signals have been gamma corrected

$$Y' = 0\cdot299R' + 0\cdot587G' + 0\cdot114B'$$

where the luminosity coefficients have been given to 3 decimal places.

The colour difference signals $(R' - Y')$ and $(B' - Y')$ are next formed and these are reduced in amplitude to $\dfrac{(R' - Y')}{1\cdot14}$ and $\dfrac{(B' - Y')}{2\cdot03}$ to limit the peak excursions of the transmitted signal.

The reduced colour difference signals are now converted to their I' and Q' equivalent signals by resolving them along the I' and Q' axes, that is to say, I' and Q' are formed thus, from Equations 4.10 and 4.11

$$I' = 0\cdot74(R' - Y') - 0\cdot27(B' - Y')$$
$$Q' = 0\cdot48(R' - Y') + 0\cdot41(B' - Y')$$

Up to this point the signals have been wideband but the frequency bandwidth of the Q' sub-carrier is now limited. The I' signal is also bandwidth limited but not so severely as the Q' signal, and is time delayed.

The Q' signal is next amplitude modulated on to a sub-carrier, which can be represented as

$$Q' \sin(\omega t + 33°)$$

where ω is the angular frequency of the sub-carrier and equals $2\pi f$ where $f \simeq 2\cdot7$ Mc/s for the 405-line system, t represents time in seconds, and 33° represents the phase of the sine wave with respect to the zero phase of the $(B' - Y')$ axis.

The I' signal is amplitude modulated on to a sub-carrier which has the same frequency as the Q' signal, but is shifted in phase by 90° and thus can be represented by a cosine function $I' \cos(\omega t + 33°)$.

During the final modulation on to the main vision carrier the I' sub-carrier bandwidth is made asymmetric as in Fig. 4.12, by the conventional frequency characteristic of the main modulator.

Both these signals are added to the delayed luminance signal to form the combined video signal M'

$$M' = Y' + Q' \sin (\omega t + 33°) + I' \cos (\omega t + 33°)$$

This is sometimes written more formally to emphasize that it is voltages which are being considered.

$$E'_M = E'_Y + k \{E'_Q \sin (\omega t + 33°) + E'_I \cos (\omega t + 33°)\}$$

where k is normally unity, but can vary between 1 and $0·3$ if it is desired to make the signal more compatible by decreasing the amplitude of the sub-carrier. Decreasing k decreases the signal-to-noise ratio of the chrominance signal at the receiver.

For the slow modulation changes which correspond to frequencies where I' and Q' are both double sideband, there is no need to refer to the I' and Q' axes and the signal can be written in terms of $(R' - Y')$ and $(B' - Y')$

$$E_M = E'_Y + \frac{E'_B - E'_Y}{2·03} \sin \omega t + \frac{E'_R - E'_Y}{1·14} \cos \omega t$$

To help in transforming from R', G' and B' signals to I' and Q' signals, etc., the following identities can all be derived from the equations already given.

4.25. Camera signals in terms of transmission signals

$$R' = Y' + 0·956I' + 0·621Q'$$
$$G' = Y' - 0·272I' - 0·647Q'$$
$$B' = Y' - 1·106I' + 1·703Q'$$
$$R' = Y' + 1·14S'_R$$
$$G' = Y' - 0·581S'_R - 0·394S'_B$$
$$B' = Y' + 2·03S'_B$$

where

$$S'_R = \frac{R' - Y'}{1·14}$$

and

$$S'_B = \frac{B' - Y'}{2·03}$$

These equations can be neatly written in matrix form, thus

$$
\begin{bmatrix} R' \\ G' \\ B' \end{bmatrix} = \begin{bmatrix} 1 & 0\cdot956 & 0\cdot621 \\ 1 & -0\cdot272 & -0\cdot647 \\ 1 & -1\cdot106 & 1\cdot703 \end{bmatrix} \begin{bmatrix} Y' \\ I' \\ Q' \end{bmatrix}
$$

and the circuits which perform such calculations are called matrix circuits.

4.26. Transmission signals in terms of camera signals

$$Y' = 0\cdot299R' + 0\cdot587G' + 0\cdot114B'$$
$$I' = 0\cdot596R' - 0\cdot275G' - 0\cdot322B'$$
$$Q' = 0\cdot211R' - 0\cdot523G' + 0\cdot312B'$$

4.27. Chrominance transmission signals in terms of colour difference signals

$$I' = 0\cdot736(R' - Y') - 0\cdot268(B' - Y')$$
$$Q' = 0\cdot478(R' - Y') + 0\cdot413(B' - Y')$$
$$O = 0\cdot299(R' - Y') + 0\cdot587(G' - Y') + 0\cdot114(B' - Y')$$
$$(R' - Y') = -1\cdot963(G' - Y') - 0\cdot381(B' - Y')$$
$$(G' - Y') = -0\cdot509(R' - Y') - 0\cdot194(B' - Y')$$
$$(B' - Y') = -2\cdot623(R' - Y') - 5\cdot144(G' - Y')$$

4.28. Colour difference signals in terms of I' and Q'

$$(R' - Y') = 0\cdot956I' + 0\cdot621Q'$$
$$(G' - Y') = -0\cdot272I' + 0\cdot647Q'$$
$$(B' - Y') = -1\cdot106I' + 1\cdot703Q'$$

4.29. Colour difference signals in terms of camera signals

$$(R' - Y') = 0\cdot701R' - 0\cdot587G' - 0\cdot114B'$$
$$(G' - Y') = -0\cdot299R' + 0\cdot413G' - 0\cdot114B'$$
$$(B' - Y') = -0\cdot299R' - 0\cdot587G' + 0\cdot886B'$$
$$S'_R = \frac{(R' - Y')}{1\cdot14} = 0\cdot615R' - 0\cdot515G' - 0\cdot100B'$$
$$S'_B = \frac{(B' - Y')}{2\cdot03} = -0\cdot147R' - 0\cdot289G' + 0\cdot436B'$$

4.30. N.T.S.C. tristimulus values in terms of C.I.E. values

The camera outputs, R', G', B' are sufficient to determine the C.I.E. co-ordinates of the transmitted colour and vice versa, and these relations are, (*see* Section 2.25)

$$R = 1 \cdot 91(X) \quad - 0 \cdot 532(Y) \quad - 0 \cdot 288(Z)$$
$$G = - 0 \cdot 985(X) \quad + 1 \cdot 999(Y) \quad - 0 \cdot 028(Z)$$
$$B = 0 \cdot 058(X) \quad - 0 \cdot 118(Y) \quad + 0 \cdot 898(Z)$$

for a reference white of $0 \cdot 981(X) + (Y) + 1 \cdot 182(Z)$ or unit luminance of illuminant C.

$$(X) = 0 \cdot 607R + 0 \cdot 174G + 0 \cdot 200B$$
$$(X) = 0 \cdot 299R + 0 \cdot 587G + 0 \cdot 114B$$
$$(Z) = 0 \cdot 000R + 0 \cdot 066G + 1 \cdot 116B$$

Notice that these equations refer to the camera signals before they have been gamma corrected.

In all of the above equations the numerical constants have been given to three decimal places in order that the reader, when calculating signal values, might be able to rely on the second decimal place in the final answer. If only two decimal places are used in the calculations, the answer may differ significantly from published values. For most engineering purposes an accuracy of two decimal places is sufficient.

It is only permissible to use colour difference signals such as $(R' - Y')$ at the receiver for modulation frequencies which lie within the Q bandwidth, i.e., for large areas of colour.

4.31. The colour burst signal

In order that the receiver shall be able to decode the colour difference information it must be able to recognize the phase of the sub-carrier signal and must therefore have a phase reference with which to compare it. Only then can the receiver tell I' phase from Q' phase, or, to look at it another way, recognize the amount of phase modulation on the sub-carrier.

The transmitter provides a reference phase by generating a continuous sine wave of sub-carrier frequency at a phase standardized as 180°, that is to say, it is opposite in phase to the positive direction of the $(B' - Y')$ axis. In Chapter 5 it is shown that of the four convenient phases of $\pm (R' - Y')$ and $\pm (B' - Y')$ the reference

signal has the least visibility on the receiver display for the
$-(B' - Y')$ phase.

A section of this continuous sine wave (Fig. 4.14) is transmitted
on the back porch of the line blanking period by gating out 9 cycles
of the reference signal every line (Fig. 4.15). The phase of this
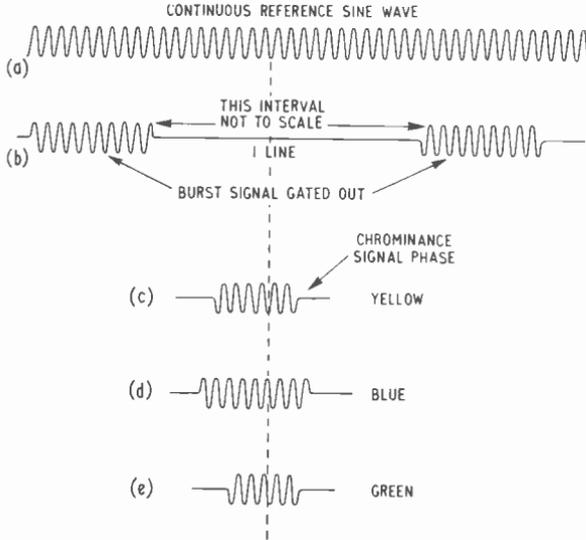
Fig. 4.14. Formation of colour burst signal

burst of reference signal is constant with respect to time but due to
the frequency relationship between sub-carrier and line frequencies
the phase of the burst with respect to the line sync pulse reverses on
successively transmitted lines, that is to say between lines 1 and 3,
or 6 and 8, etc.

The peak-to-peak amplitude of this colour burst signal is the
same as the height of the sync pulse signal, i.e. 0·3 of the peak white
transmitted signal.

4.32. Frequency relationships

The approximate choice of frequency for the sub-carrier is deter-
mined by the bandwidth required for the Q' channel. On mono-
chrome receivers the sub-carrier appears as an interfering C.W. signal
and produces a beat pattern of bright and dark dots along the scan-
ning lines, the dots falling into the familiar pattern of interference
stripes across the picture. The higher the sub-carrier frequency,

the finer and less obvious the texture of the beat pattern becomes. If the sub-carrier frequency f_{sc} is too close to the upper frequency limit of the luminance channel there is insufficient room for the upper sidebands of the Q' sub-carrier (Fig. 4.12). To further reduce the visibility of the dot pattern, the exact sub-carrier frequency in the required approximate range is chosen so that the dots interlace; this happens if the sub-carrier frequency f_{sc} is any odd multiple of half the line scanning frequency f_L

$$f_{sc} = (2n + 1)\tfrac{1}{2}f_L$$

where n is any whole number. The sub-carrier peaks, corresponding to white dots, are then displaced on successive scanning lines since there is always half a cycle of sub-carrier left over at the end of each line. On any one field, therefore, the dots are neatly staggered. On the next field the dots are staggered with respect to the dots on the next upper line of the previous field, since there are, for example, $202\tfrac{1}{2}$ lines in a field so that the first half line of the second field is really a continuation of the last half line of the first field, the 203rd. This is illustrated for a simple system with very quick flybacks in Fig. 4.16.

The succeeding 3rd field rescans the odd lines again and the sub-carrier peaks are now displaced relative to the dots on the first
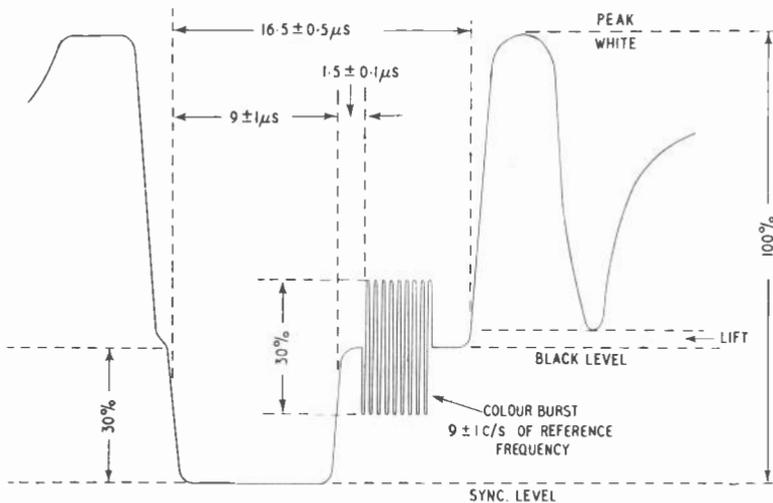
Fig. 4.15. The colour burst reference signal. The burst is not transmitted during field sync pulses
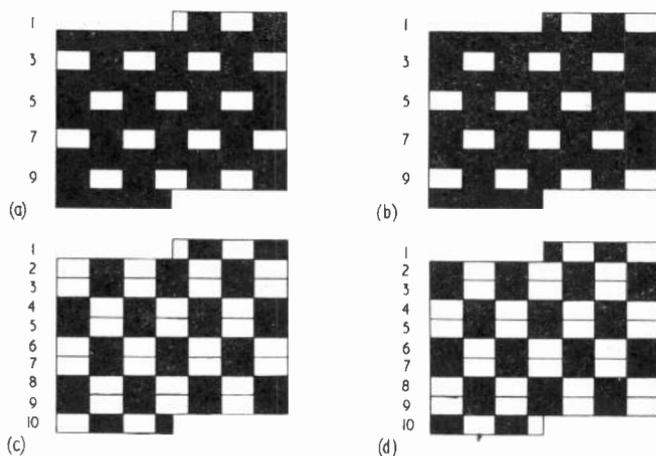
Fig. 4.16.   Dot interlacing.   (a)—picture 1, field 1; (b)—picture 2,
field 3; (c)—picture 1, fields 1 and 2; (d)—picture 2, fields 3 and 4

field.   Similarly on the fourth field the dots fall between those
illuminated on the even lines of the second field.

The net result is that over two complete pictures, i.e., $\dfrac{1}{12 \cdot 5}$ s,
the peaks of the sub-carrier on one picture fall into the troughs of the
sub-carrier on the next picture.   This behaviour arises because the
line frequency is an odd multiple of the picture frequency, f_P

$$f_L = 405 f_P$$

so that the sub-carrier frequency is an odd multiple of half the picture
frequency as well as of half the line frequency,

$$f_{sc} = \tfrac{1}{2}(2n + 1) f_L$$
$$= \tfrac{1}{2}(2n + 1) 405 f_P$$

and as 405 and $(2n + 1)$ are both odd, their product is odd.

Although the eye does not store the image very well over this
relatively long period, the net subjective effect is a marked decrease
in the annoyance value of the beat pattern, equivalent to a decrease
in sub-carrier amplitude of about 20dB.   This principle is used in
monochrome television to reduce the interference between stations
sharing the same channel.

For 405 lines a convenient odd multiple of half the line frequency, which has low value factors to make the transmitter dividing circuits simple, is $\frac{525}{2}f_L$, $(525 = 7 \times 5 \times 5 \times 3)$.

4.33. Asynchronous operation

To ease the design of the reference circuits in the receiver it is usual to unlock the field frequency from the mains and to make f_F exactly 50 c/s however the mains frequency is varying, so that f_L is then exactly

$$\frac{405}{2}f_F = 10{,}125 \text{ c/s}$$

and the sub-carrier frequency is constant at

$$\frac{525}{2}f_L = 2{\cdot}6578125 \text{ Mc/s}$$

4.34. Vision-sound carrier spacing

To reduce the visibility of any beat patterns produced between the sub-carrier and the sound carrier, this frequency interval is also made an odd multiple of half the line frequency and the vision carrier-sound carrier spacing becomes a multiple of f_L

$$f_{\text{vision}} = f_{\text{sound}} + 350 f_L$$

$$= f_{\text{sound}} + \frac{4}{3}f_{sc}$$

This means that the vision-sound carrier spacing cannot be maintained at exactly $3 \cdot 5$ Mc/s, but must now be $3 \cdot 54375$ Mc/s. In practice the receiver tuning of the sound carrier is more critical than the vision carrier and for the benefit of the monochrome receivers already in use the sound carrier frequency of the station is left unaltered and the vision carrier is increased slightly in frequency.

In the 525-line system, although F.M. sound is used and the vision carrier-sound carrier spacing cannot therefore be locked, nevertheless the visibility of the beat between the sub-carrier and the sound carrier is reduced, on average, by making the vision carrier-undeviated sound carrier spacing a multiple of the line frequency. The nominal line frequency is $15 \cdot 75$ kc/s and nearest harmonic to $4 \cdot 5$ Mc/s is the 286th at $4 \cdot 5045$ Mc/s. However, the change to $4 \cdot 5045$ Mc/s might upset the intercarrier operation of

existing monochrome receivers, so the carrier spacing is kept at exactly 4·5 Mc/s by reducing the line frequency from 15·75 kc/s to 15·73426 kc/s.

That is

$$\text{Line frequency } f_L = \frac{4\cdot5 \times 10^6}{286} = 15{,}734\cdot26 \text{ c/s}$$

$$(\text{monochrome} = 15{,}750 \text{ c/s})$$

$$\text{Field frequency } f_F = \frac{2}{525}f_L = 59\cdot94 \text{ c/s}$$

$$(\text{monochrome} = 60 \text{ c/s})$$

As far as the sub-carrier frequency is concerned, a Q' bandwidth of 0·5 Mc/s must be accommodated in the video band of 4·0 Mc/s, so that the sub-carrier frequency must be near 3·5 Mc/s. A convenient multiple is 455, so that

$$\text{sub-carrier frequency } f_{sc} = \frac{455}{2}f_L = 3{,}579{,}545 \text{ c/s}$$

where $\qquad\qquad\qquad 455 = 13 \times 7 \times 5$

Similarly, in the 625-line system the 5·5 Mc/s vision carrier-sound carrier spacing is made a multiple of the line frequency. In this case the 352nd harmonic of the nominal line frequency of 15·625 kc/s is exactly 5·5 Mc/s, so that the monochrome scanning frequencies are acceptable.

$$\text{Line frequency } f_L = \frac{5\cdot5 \times 10^6}{352} = 15{,}625 \text{ c/s}$$

$$(\text{monochrome} = 15{,}625 \text{ c/s})$$

$$\text{Field frequency } f_F = \frac{2}{625}f_L = 50 \text{ c/s}$$

$$(\text{monochrome} = 50 \text{ c/s})$$

For a Q' bandwidth of 0·6 Mc/s and a video band of 5 Mc/s, a sub-carrier near 4·4 Mc/s is indicated. A convenient multiple is 567, so that

$$\text{sub-carrier frequency } f_{sc} = \frac{567}{2}f_L = 4{,}429{,}687 \text{ c/s}$$

$$567 = 7 \times 3 \times 3 \times 3 \times 3$$

The same sub-carrier frequency will probably be used in the Russian 625-line system where the vision-sound carrier spacing is 6·5 Mc/s.

4.35. Frequency interleaving and the energy spectrum

In Chapter 1 it was pointed out that the periodic nature of the luminance signal gives a discrete line structure to the energy spectrum of the television video signal (Fig. 1.10). For similar reasons the sub-carrier signal and its sidebands also exhibit a discrete line structure and due to the odd multiple of half line frequency relationship adopted, the clumps of energy in the sub-carrier spectrum fall neatly between the clumps in the luminance signal spectrum (Fig. 4.17). The two signals can therefore coexist in the same frequency
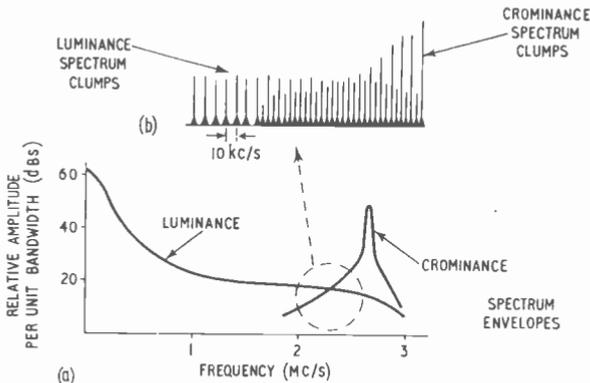
Fig. 4.17. (a) is the frequency spectrum for a typical 405-line signal. (b) shows an expansion of the frequency scale

band as separate identities. Despite this it is not practical to extract one from the other as filters with the necessary comb-teeth properties of changing from pass to reject every 5 kc/s, and ideally every 25 c/s, are not yet manufactured. However, the subjective appearance of the composite signal on a monochrome picture is improved by this interleaving technique.

4.36. Delay characteristics

The design of the colour receiver I.F. circuits is rather more difficult than is the case for monochrome, since the amplitude-frequency response must be maintained out to well past the sub-carrier frequency. At the same time, the sound rejection required

is at least as much as for monochrome reception and this implies a sharp onset of attenuation at the end of the luminance band. With the minimum-phase type of circuits used in monochrome reception, such an abrupt change in the amplitude response is reflected in a lack of linearity in the accompanying phase characteristic. In the colour receivers so far produced, including those using non-minimum phase sound rejector circuits, it has not been economic to combine
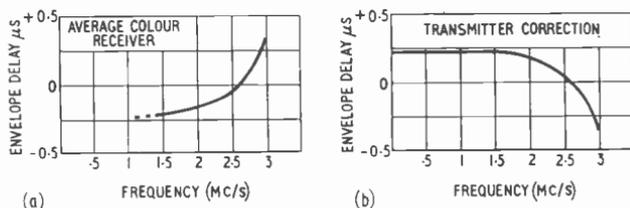
Fig. 4.18.   Envelope delay characteristics

an ideal flat amplitude response with an ideal linear phase response.

The phase responses of several types of colour receiver were measured and computed and an average delay curve drawn. The transmitter applies pre-correction to balance out this average phase error in colour receivers, and Fig. 4.18 shows these two envelope delay characteristics. The transmitter correction is only for frequencies above $1\frac{1}{2}$ Mc/s.

4.37.   Transmitter tolerances

The transmitter line and field frequencies are counted down from the sub-carrier frequency, which has a tolerance of ± 8 c/s and a maximum rate of change of $0\cdot1$ c/s². The burst is 9 cycles of sub-carrier ± 1 cycle, and its position is defined to within $\pm 0\cdot1$ μs relative to the line pulse. The tolerances on the sub-carrier signals are approximately ± 2dB in amplitude (*see* Chapter 5) and $\pm 10°$ in phase for the saturated primary colours at 75% of peak signal. The sub-carrier phase accuracy is normally better than this for flesh tones. The Y', I' and Q' signals should match each other in time to $0\cdot07$ μs. The peak white signal of the transmitter is usually reduced to 90% of its monochrome value to allow partially for the sub-carrier excursions. In other respects the colour transmission conforms to standard monochrome tolerances.

4.38.   The colour test stripe signal

For some time after the inauguration of a colour television service it is expected that the majority of the programmes will continue to

(Above) *luminance and chrominance components displayed separately;* (below) *complete lack of I' signal on left of picture and complete lack of Q' signal on right of picture*
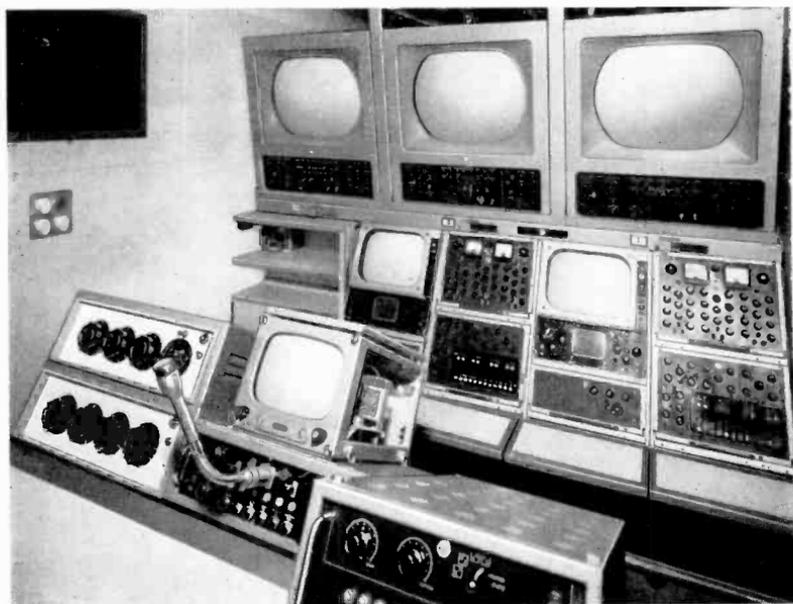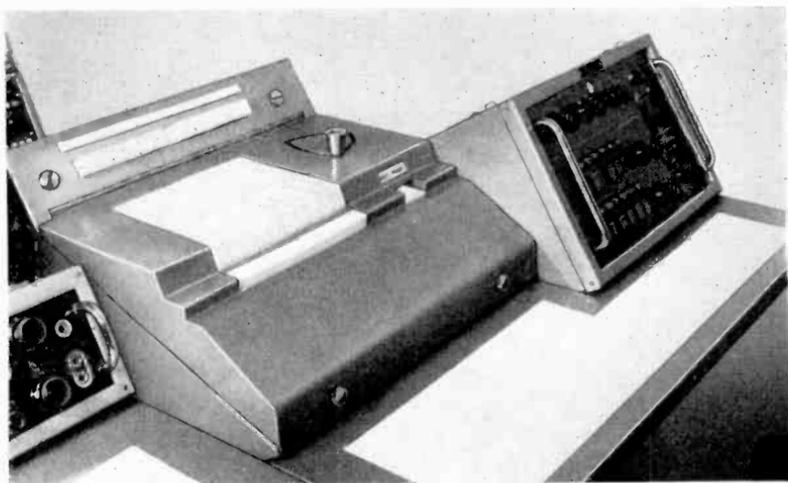
The top photograph shows the Marconi next channel direct switching vision mixer, specially designed to handle colour television signals. To the right of the mixer, is a special effects unit for producing horizontal and vertical wipes and corner inserts. The lower picture shows the camera control compartment of the Marconi two camera channel Mobile Colour Television Unit. (Courtesy Marconi Ltd.)

be in monochrome. To enable the colour performance of any particular receiver installation to be assessed whilst monochrome programmes are being radiated, it is possible for stations to transmit a special signal. The signal which has been used in America consists of a burst of sub-carrier frequency at the beginning and end of each line (Fig. 4.19). The two bursts have the same phase but their frequency does not need to be locked to the line scanning frequency.

On a monochrome receiver these bursts produce two narrow vertical stripes of beat pattern at the extreme edges of the picture which are often hidden by the picture mask.

On a colour receiver with an automatic colour killer (*see* Section 7.7) the stripes will be hardly visible since the luminance channel is designed to reject sub-carrier frequency. To make the stripes visible, the burst gate of the receiver must be delayed until it accepts the nearest stripe: this will switch on the chrominance channel and pull the reference oscillator into phase lock. The second stripe on the tube face should then appear as a green band along the edge of the monochrome picture. If the aerial or receiver has inadequate gain at sub-carrier frequency, the green stripe will not appear. If
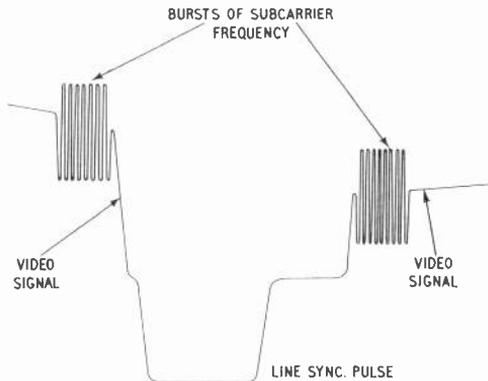
Fig. 4.19. Colour test stripe signal. Test bursts have the same amplitude and duration as colour burst, but need not be locked to line frequency

the hue control is incorrectly set, the stripe will have the wrong colour. The burst gate of the receiver may be delayed either by adjusting the line hold control or by adding capacity to the sync separator anode (*see* Section 10.4).

4.39 Summary

The colour television camera analyses the picture into three voltages R', G' and B', which represent the amounts of red, green and blue lights which the receiver must generate to produce an

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impression of the colour in the studio scene. A luminance video signal is formed by combining fractions of the primary voltages,

$$Y' = 0.30R' + 0.59G' + 0.11B'$$

The rest of the colour information is transmitted by sending bandwidth limited I' and Q' signals on a sub-carrier placed in the high frequency part of the video signal. The I' and Q' signals are derived from the colour difference signals $(R' - Y')$ and $(B' - Y')$ thus

$$I' = 0.74(R' - Y') - 0.27(B' - Y')$$
$$Q' = 0.48(R' - Y') + 0.41(B' - Y')$$

The resulting colour sub-carrier vanishes for grey tones, $I' = Q' = 0$, and its amplitude increases with both the purity and the luminance of the colour being transmitted. The phase of the sub-carrier signal with respect to the colour burst signal describes approximately the hue of the transmitted colour.

The luminance signal produces an acceptable black and white picture on normal monochrome receivers, the sub-carrier producing an interference pattern which is not obvious; the field frequency is not locked to the 50 c/s mains.

Colour Specification in N.T.S.C. Systems

5.1. Introduction

In Chapter 4 the transmitted signal has been discussed in electrical terms using as units the voltages produced by the cameras on peak white. Different colours produce different electrical signals and it is possible to plot the chromaticities of colours on a C.I.E. diagram (*see* Section 2.25), and to indicate on the same diagram the corresponding signal voltages. Such diagrams can indicate how the received picture will appear to the eye as the signal voltages change.

For convenience in measuring and recording colours the C.I.E. system does not use the red, green and blue primaries but its own set of theoretical primaries called X^*, Y^* and Z^*. Similarly, the television transmitter does not use amounts of the red, green and blue primaries to describe the colour being transmitted; instead it radiates the amounts of Y', I' and Q', and the behaviour of the system can be described in terms of these quantities.

It is unfortunate that the letter Y has been adopted for both the C.I.E. fictitious green primary and for the N.T.S.C. luminance signal and its corresponding white primary. Since this practice is widespread, the letters have not been changed since the reader can easily distinguish which is meant in any particular context. Although both Y's have the same numerical value, the chromaticity of the C.I.E. Y^* is $x = 0$ and $y = 1$, while the chromaticity of the N.T.S.C. Y' is $x = 0.310$ and $y = 0.316$.

5.2. System white and primary colours

The choice of white for the colour television system theoretically affects the size of the units in which the red, green and blue components of the picture are measured. The usual choice is C.I.E. Illuminant C (*see* Chapters 2 and 4). If a warmer white such as Illuminant A is chosen, more red light is needed to produce the system white and the size of the red unit is larger. For practical

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reasons, e.g., tungsten lighting, the reference or normalizing white used at the studio is not usually Illuminant C, but the colour receiver should always be adjusted to reproduce Illuminant C when the chrominance signals are zero.

In discussing the colour performance of the N.T.S.C. system, the electrical signals will be related to the chromaticity of the colour reproduced at the receiver, on the assumption that the receiver is using the N.T.S.C. red, green and blue primaries (Fig. 4.5) and that Illuminant C is the reference white.

5.3. Y' luminance signal

The luminance signal is formed after the camera signals have been gamma corrected (*see* Section 4.14), thus

$$Y' = 0\cdot30R^{\frac{1}{\gamma}} + 0\cdot59G^{\frac{1}{\gamma}} + 0\cdot11B^{\frac{1}{\gamma}}$$

or

$$Y' = 0\cdot30R' + 0\cdot59G' + 0\cdot11B'$$

where $0\cdot30$ is the luminosity coefficient of the N.T.S.C. red primary, etc. If different red, green and blue primaries are chosen for the system, the proportions of R', G' and B' used for the luminance signal will also be different.

Because of the need for gamma correction, parts of the system are not linear; a change of $0\cdot1$ in a signal such as the R' signal produces a bigger absolute increase in the light output from the receiver tube if it is a change from $0\cdot8$ to $0\cdot9$, than if it is a change from $0\cdot1$ to $0\cdot2$.

Ideally, the luminance signal should be formed before gamma correction is carried out, while the camera signals are directly proportional to the light intensity of the scene

$$Y = 0\cdot30R + 0\cdot59G + 0\cdot11B$$

where R, G and B, without primes, indicate voltages which have not been gamma corrected. The gamma corrected luminance signal would then be

$$Y^{\frac{1}{\gamma}} = (0\cdot30R + 0\cdot59G + 0\cdot11B)^{\frac{1}{\gamma}}$$

which is not the same as the luminance signal actually used

$$Y' = 0\cdot30R^{\frac{1}{\gamma}} + 0\cdot59G^{\frac{1}{\gamma}} + 0\cdot11B^{\frac{1}{\gamma}}$$

unless $R = G = B$ when the picture is a black and white colourless scene.   For a saturated blue colour

$$Y^{\frac{1}{\gamma}} = (0 \cdot 11 \times 1)^{\frac{1}{2 \cdot 2}} = 0 \cdot 37$$

and

$$Y' = 0 \cdot 11(1)^{\frac{1}{2 \cdot 2}} = 0 \cdot 11$$

so that the Y' signal is smaller than it should be.   This result is true in general for all the saturated and near saturated colours —the luminance signal is too small.   For greys the luminance signal is correct and for desaturated colours the error is small. The luminance reproduced by the signal actually used is

$$(Y')^{\gamma} = (0 \cdot 30R' + 0 \cdot 59G' + 0 \cdot 11B')^{\gamma}$$

and for the saturated primaries and their complementaries at full brightness, i.e., $R = 1$ or 0, $G = 1$ or 0 and $B = 1$ or 0. $R = R'$, $G = G'$, $B = B'$ and $(Y')^{\gamma} = (Y)^{\gamma}$ which is not Y, of course, and is therefore incorrect.

The luminance reproduced by the theoretically correct signal is

$$(Y^{\frac{1}{\gamma}})^{\gamma} = Y$$

If the reproducing cathode-ray tube has a gamma higher than $2 \cdot 2$, as may happen for monochrome receivers, then the reproduced luminance is actually

$$(Y')^{\gamma_2} = (Y)^{\gamma_2}$$

for the saturated primaries at full amplitude, and for the second signal would be

$$(Y^{\frac{1}{\gamma_1}})^{\gamma_2} = (Y)^{\frac{\gamma_2}{\gamma_1}}$$

where $\gamma_1 = 2 \cdot 2$ and γ_2 may be $2 \cdot 7$.

Values for these signals are given in Table 5.1 for the saturated primaries and their complementaries at maximum brightness, and at three-quarters of their maximum brightness.

Since the transmitter generates in the first place the primary R', G' and B' voltages, and these are recovered at the receiver, the total luminance produced by the colour cathode-ray tube is correct.   It will be the sum of the luminances produced by $(R')^{\gamma}$

Table 5.1

LUMINANCE CARRIED BY MONOCHROME COMPATIBLE SIGNAL

Colour	Correct luminance Y	Transmitted signal Y'	Theoretically correct signal $Y^{\frac{1}{\gamma}}$	Luminance reproduced on tube with $\gamma = 2\cdot2$, i.e. colour tube		Luminance reproduced on tube with $\gamma = 2\cdot7$, i.e. monochrome tube		Correct luminance Y
				$(Y')^{2\cdot2}$	$(Y^{\frac{1}{\gamma}})^{2\cdot2}$	$(Y')^{2\cdot7}$	$(Y^{\frac{1}{\gamma}})^{2\cdot7}$	
100% Purity colours / 100% Amplitude signals								
White	1	1	1	1	1	1	1	1
Yellow	0·886	0·886	0·946	0·77	0·89	0·72	0·86	0·89
Cyan	0·701	0·701	0·851	0·46	0·70	0·38	0·65	0·70
Green	0·587	0·587	0·785	0·31	0·59	0·24	0·52	0·59
Magenta	0·413	0·413	0·670	0·14	0·41	0·09	0·34	0·41
Red	0·299	0·299	0·578	0·07	0·30	0·04	0·23	0·30
Blue	0·114	0·114	0·373	0·01	0·11	0·003	0·07	0·11
100% Purity colours / 75% Amplitude signals								
Yellow	0·664	0·777	0·830	0·57	0·66	0·51	0·61	0·66
Cyan	0·526	0·615	0·747	0·34	0·53	0·27	0·45	0·53
Green	0·440	0·515	0·689	0·23	0·44	0·17	0·37	0·44
Magenta	0·310	0·362	0·587	0·11	0·31	0·06	0·24	0·31
Red	0·224	0·262	0·507	0·05	0·22	0·03	0·16	0·22
Blue	0·086	0·100	0·327	0·01	0·09	0·002	0·05	0·09

of red light, $(G')^\gamma$ of green light and $(B')^\gamma$ of blue light, the input-output characteristic of the cathode-ray tube effectively raising the signals to the power γ, or 2·2. Since $R' = R^{\frac{1}{\gamma}}$, the total luminance is the sum of the luminance produced by R, G and B,

i.e.

$$lR + mG + nB = 0·30R + 0·59G + 0·11B$$

where l, m and n are the luminosity coefficients of R, G and B. This total luminance is proportional to the brightness of the original scene.

As the final luminance on the colour receiver tube is correct and the transmitted Y' signal is too small, the rest of the luminance must be carried by the chrominance signals, such as $(R' - Y')$ and $(B' - Y')$ or I' and Q'. The constant luminance principle is thus only approximately followed for strongly saturated colours (*see* Sections 4.12 and 4.14). Since the Y' signal is not conveying all the luminance it would perhaps be better to call it the mono-chrome signal, but common usage is to refer to it as the luminance signal.

If the transmitter radiated the true luminance signal

$$Y^{\frac{1}{\gamma}} = (0·30R + 0·59G + 0·11B)^{\frac{1}{\gamma}}$$

the receiver would need complicated circuits to derive from $Y^{\frac{1}{\gamma}}$, $(R' - Y^{\frac{1}{\gamma}})$ and $(B' - Y^{\frac{1}{\gamma}})$ the required G' value. By using Y' the receiver needs only simple linear circuits which add and subtract fractions of the transmitted signals.

5.4. Choice of burst phase

Since the chrominance signal carries some of the luminance information, the colour reference burst may produce a coloured vertical bar on the display tube if the line blanking is not adequate. The visibility of this bar is minimized by the choice of phase of the burst as $-180°$. Thus if the burst amplitude is S and it has a phase θ with respect to the $\dfrac{(B' - Y')}{2·03}$ axis, then it will have a component along this axis of $S \cos \theta$ which the receiver will interpret as a signal

$$(B' - Y') = 2·03S \cos \theta$$

Similarly, the burst will produce a red difference signal of

$$(R' - Y') = 1·14S \sin \theta$$

These two signals correspond to a green difference signal of

$$(G' - Y') = \frac{-0 \cdot 30}{0 \cdot 59}(R' - Y') - \frac{0 \cdot 11}{0 \cdot 59}(B' - Y')$$

<div align="right">see Equation 4.5</div>

$$= \frac{-0 \cdot 30}{0 \cdot 59}(1 \cdot 14\ S \sin \theta) - \frac{0 \cdot 11}{0 \cdot 59}(2 \cdot 03\ S \cos \theta)$$

For the colour burst the transmitted Y' signal is zero, so that the red, green, and blue signals to the tube are

$$R' = 1 \cdot 14 S \sin \theta$$
$$G' = \frac{-0 \cdot 30}{0 \cdot 59}(1 \cdot 14\ S \sin \theta) - \frac{0 \cdot 11}{0 \cdot 59}(2 \cdot 03\ S \cos \theta)$$
$$B' = 2 \cdot 03 S \cos \theta$$

and these produce a total luminance of

$$L_s = l(R')^\gamma + m(G')^\gamma + n(B')^\gamma$$

If R', G', or B' is negative, then it makes no contribution to the total luminance. The four most convenient phases for S are

<div align="center">Table 5.2
LUMINANCE PRODUCED BY COLOUR BURST</div>

θ	$0°$	$90°$	$180°$	$270°$
Red	0	$0 \cdot 4 S^\gamma$	0	0
Green	0	0	$0 \cdot 07 S^\gamma$	$0 \cdot 18 S^\gamma$
Blue	$0 \cdot 52 S^\gamma$	0	0	0

$\theta = 0$, $90°$, $180°$ and $270°$, and Table 5.2 gives the luminance contribution of the resulting red, green and blue signals for these four phases.

Of the four most convenient phases for the burst, $\theta = 180°$ gives the least luminance on the picture, and in this case the output is green in colour. The optimum phase angle is on the $+I'$ axis side of the $-(B' - Y')$ phase, but $180°$ is a more useful compromise.

5.5. Colour bar signal

Fig. 4.8 shows a diagrammatic picture of chrominance signals, the dots marking the ends of the phasors which represent the sub-carrier signal for each of the three primaries and their complementary colours. In considering the performance of colour

television systems and apparatus these six colours form convenient standard points to which to refer, giving rise to signals which are roughly equally spaced around the sub-carrier diagram.   One of the standard colour television test signals consists of a picture containing eight equal-width vertical bars, as in Fig. 5.1; each bar

| WHITE | YELLOW | CYAN | GREEN | MAGENTA | RED | BLUE | BLACK |
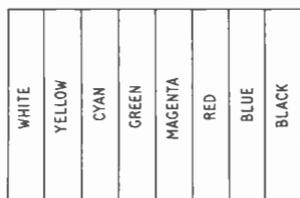
Fig. 5.1.   Colour bar test picture

is coloured with either one of the primary colours, a complementary, black or white.   The colours are usually arranged in order of luminance with the brightest colours on the viewer's left hand side. Other arrangements are sometimes used for special purposes.

The video waveform corresponding to such a saturated colour bar picture is shown in Fig. 5.2(a) and (b).   The sub-carrier peaks extend up to 1·33, or $\frac{1}{3}$ above the peak white signal, and down to −0·33, into the synchronizing region.   The weighting factors of $\frac{1}{1·14}$ for the $(R' - Y')$ signal and $\frac{1}{2·03}$ for the $(B' - Y')$ signal ensure that these excursions are not larger.   Without such weighting factors the sub-carrier would need to extend down below the zero carrier level, as shown in Fig. 5.3, which is not possible.   If the weighting factors are chosen to be small, so that the chrominance signals are small also, then the signal-to-noise ratio at the edge of the reception area is poor.   It is desirable to use the maximum amplitude of chrominance signal which can be radiated without overloading the transmitter.   Highly saturated colours at high luminance occur for only 1% of the time in average television programmes and it has been found acceptable to limit the sub-carrier excursions to within the range 1·33 to −0·33.   Between the range 1 to 1·33 the transmitter is nominally overloaded but if this is only for brief periods the effects are not serious.   In the range 0 to −0·33 the sub-carrier is extending into the synchronizing region, but here again the practical effect on receivers turns out to be small.

If the sub-carrier were modulated with the full colour difference signals $(R' - Y')$ and $(B' - Y')$, which will be written as R_D and
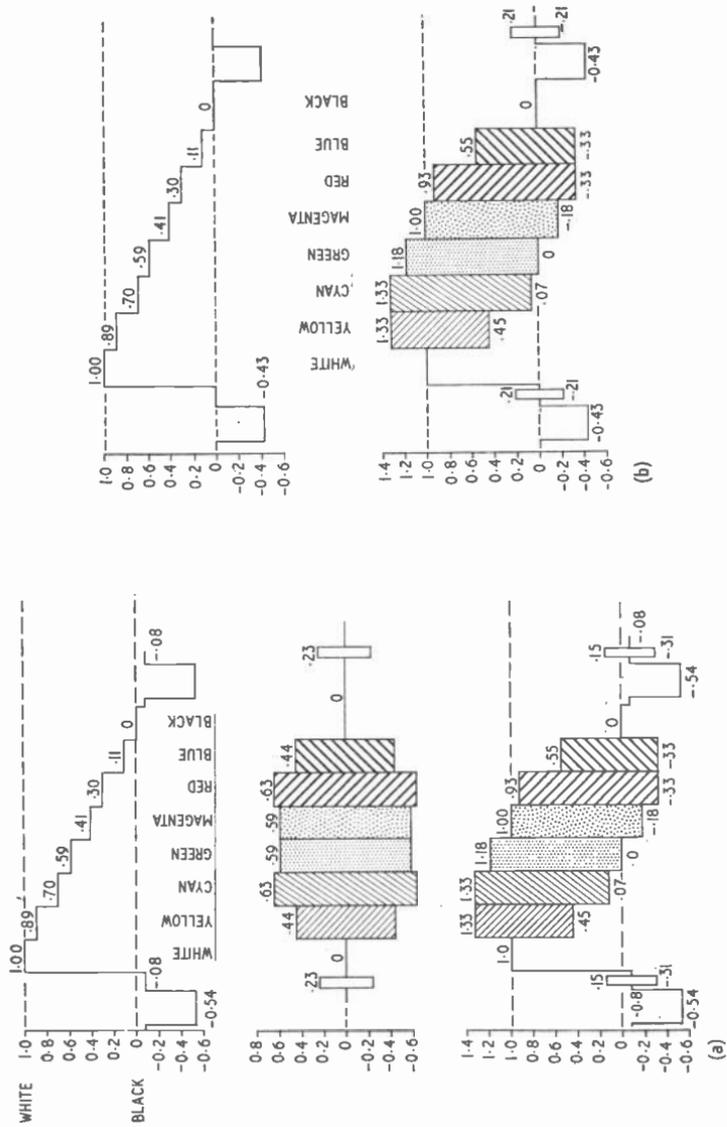
Fig. 5.2. *Saturated colour bars at full amplitude.* (a) *with set-up.* (b) *without set-up: set-up is 5% of peak white signal including sync.*

B_D, the subcarrier amplitude would be $\sqrt{R_D{}^2 + B_D{}^2}$. The peak excursions of the composite signal waveform would then be

$$Y' \pm \sqrt{R_D{}^2 + B_D{}^2}$$

and the values of this for the primary colours are shown in Fig. 5.3 and in Table 5.3.
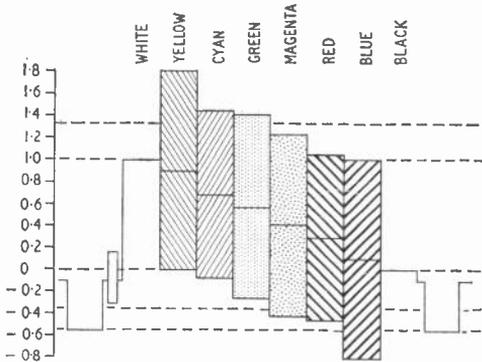
Fig. 5.3. Theoretical colour bar waveform with full colour difference signals

Fig. 5.4. Colour bar waveform with single weighting of the difference signals
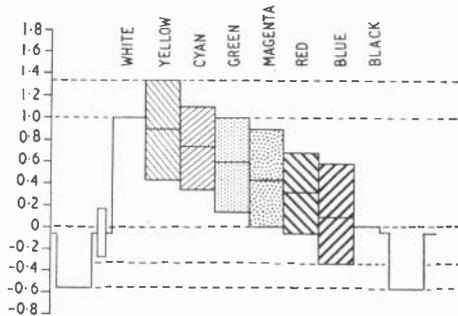
If the colour difference signals are reduced by a single weighting factor δ, thus

$$Y' \pm \delta\sqrt{R_D{}^2 + B_D{}^2}$$

δ can be adjusted so that on the maximum brightness of a saturated yellow

$$Y' + \delta\sqrt{R_D{}^2 + B_D{}^2} = 1\tfrac{1}{3}$$

and the luminance signal is

$$Y' = lR' + mG' + nB'$$

where l, m and n are the luminosity coefficients of the N.T.S.C.

Table 5.3

MAXIMUM AND MINIMUM EXCURSIONS OF THE COMPOSITE VIDEO SIGNAL

Colour	Y'	R_D	$(R_D)^2$	B_D	$(B_D)^2$	$(R_D)^2 + (B_D)^2$	$\sqrt{(R_D)^2 + (B_D)^2}$	$\delta\sqrt{(R_D)^2 + (B_D)^2}$ $\delta = 0.50$	$Y' \pm \sqrt{(R_D)^2 + (B_D)^2}$		$Y' \pm \delta\sqrt{(R_D)^2 + (B_D)^2}$	
Yellow	0·886	0·114	0·0130	−0·886	0·7850	0·7980	0·8933	0·447	1·78	−0·01	1·33	0·44
Cyan	0·701	−0·701	0·4914	0·299	0·0894	0·5808	0·7621	0·381	1·46	−0·06	1·08	0·32
Green	0·587	−0·587	0·3446	−0·587	0·3446	0·6892	0·8302	0·415	1·42	−0·24	1·00	0·17
Magenta	0·413	0·587	0·3446	0·587	0·3446	0·6892	0·8302	0·415	1·24	−0·42	0·83	0·00
Red	0·299	0·701	0·4914	−0·299	0·0894	0·5808	0·7621	0·381	1·06	−0·46	0·68	−0·08
Blue	0·114	−0·114	0·0130	0·886	0·7850	0·7980	0·8933	0·447	1·01	−0·78	0·56	−0·333

primaries.   Their values of 0·299, 0·587 and 0·114 are preferably not substituted until the end of the calculation to avoid errors in the second decimal place.   For yellow, $B' = 0$, so that

$$Y' = lR' + mG' = l + m = 1 - n$$

and

$$R_D = 1 - Y' = n$$

$$B_D = n - 1$$

$\therefore$

$$1\tfrac{1}{3} = 1 - n + \delta\sqrt{n^2 + (n - 1)^2}$$

$$= 1 - n + \delta\sqrt{n^2 + n^2 - 2n + 1}$$

$$= 1 - n + \delta\sqrt{2n^2 - 2n + 1}$$

$\therefore$

$$\delta = \frac{1\tfrac{1}{3} - 1 + n}{\sqrt{2n^2 - 2n + 1}}$$

$$= \frac{n + \tfrac{1}{3}}{\sqrt{2n^2 - 2n + 1}}$$

By substituting 0·114 for n, $\delta = 0·50$, and the values of

$$Y' \pm \delta\sqrt{R_D{}^2 + B_D{}^2}$$

for the primary colours are given in Table 5.3.

It will be seen that the total excursions of the signal are now limited to the range $1\tfrac{1}{3}$ to $-\tfrac{1}{3}$ and that these maximum values are reached by only the yellow and blue signals.   It is not necessary to limit the cyan and red signals to less than this and they also can be allowed to reach $1\tfrac{1}{3}$ and $-\tfrac{1}{3}$ if two weighting signals are used instead of one, thus

$$Y' \pm \sqrt{\alpha R_D{}^2 + \beta B_D{}^2}$$

For yellow

$$Y' = 1 - n, \quad R_D = n, \quad B_D = n - 1$$

$\therefore$

$$1\tfrac{1}{3} = 1 - n + \sqrt{\alpha^2 n^2 + \beta^2(n - 1)^2}$$

$\therefore$

$$n + \tfrac{1}{3} = \sqrt{\alpha^2 n^2 + \beta^2(n - 1)^2} \qquad [5.1]$$

For cyan

$$Y' = m + n = 1 - l, \; R_D = l - 1 \text{ and } B_D = l$$

$$\therefore \qquad 1\tfrac{1}{3} = 1 - l + \sqrt{\alpha^2(l-1)^2 + \beta^2 l^2}$$

$$\therefore \qquad l_R + \tfrac{1}{3} = \sqrt{\alpha^2(l+1)^2 + \beta^2 l^2} \qquad [5.2]$$

The same two equations are obtained from the negative excursions of the signal

$$Y' - \sqrt{\alpha R_D{}^2 + \beta B_D{}^2} = -\tfrac{1}{3}$$

for red and blue.   This arises from the symmetrical properties of the primary signals and their complementaries.

From Equation 5.1

$$(3n + 1)^2 = 9[\alpha^2 n^2 + \beta^2(n-1)^2] \qquad [5.3]$$

From Equation 5.2

$$(3l + 1)^2 = 9[\alpha^2(l-1)^2 + \beta^2 l^2] \qquad [5.4]$$

Multiply Equation 5.3 by l^2

$$l^2(3n + 1)^2 = 9[\alpha^2 l^2 n^2 + \beta^2 l^2(n-1)^2] \qquad [5.5]$$

Multiply Equation 5.4 by $(n-1)^2$

$$(n-1)^2(3l + 1)^2 = 9[\alpha^2(n-1)^2(l-1)^2 + \beta^2(n-1)^2 l^2] \qquad [5.6]$$

Subtracting Equation 5.6 from Equation 5.5

$$l^2(3n+1)^2 - (n-1)^2(3l+1)^2 = 9\alpha^2[l^2 n^2 - (n-1)^2(l-1)^2]$$

since $a^2 - b^2 = (a - b)(a + b)$, then

$$(3ln + l - 3ln + 3l - n + 1)(3ln + l + 3ln - 3l + n - 1)$$

$$= 9\alpha^2(ln - nl + l + n - 1)(ln + ln - l - n + 1)$$

$$(4l - n + 1)(6ln - 2l + n - 1) = 9\alpha^2(l + n - 1)(2ln - l - n + 1)$$

Further, since $l + m + n = 1$

$$9\alpha^2 = \frac{(4l - n + 1)(6ln - 2l + n - 1)}{-m(2ln + m)}$$

$$= \frac{(4l - n + 1)(1 - 6ln + 2l - n)}{m(2ln + m)} \qquad [5.7]$$

Substituting $l = 0.299$, $m = 0.587$ and $n = 0.114$

$$\alpha = \frac{1}{1\cdot14}$$

Similarly, multiplying Equation 5.3 by $(l - 1)^2$ and Equation 5.4 by n^2 and subtracting gives

$$9\beta^2 = \frac{(1 + 2n - 6nl - l)(1 + 4n - l)}{m(2ln + m)} \qquad [5.8]$$

and hence

$$\beta = \frac{1}{2\cdot03}$$

Using these two weighting factors of $\dfrac{1}{1\cdot14}$ and $\dfrac{1}{2\cdot03}$, both yellow and cyan signals go up to the maximum value of $1\frac{1}{3}$ and both blue and red go down to the minimum value of $-\frac{1}{3}$; in this way the best signal-to-noise ratio is obtained, at the expense of a more noticeable dot pattern on monochrome receivers.

The transmission of the colour bar signal of Fig. 5.2 provides a much more onerous condition for the transmitter than does the average television signal. Such colour bar signals can be reduced to 75% of their peak value, when they still represent 100% purity colours but at only 75% of their maximum luminance. Transmitter overload can also be reduced by transmitting the waveform corresponding to colours of full luminance and 75% saturation, but this is not such a convenient signal for receiver testing.

Fig. 5.5 shows the sub-carrier signals for the colour bar waveform at 100% and 75% of maximum amplitude, and the corresponding luminance amplitudes are shown in Table 5.4. The transmitter tolerances are $\pm10°$ in phase and 20% (approx. 2dB) in ratio of sub-carrier to luminance amplitude. Fig. 5.5 corresponds to Fig. 4.8 but the background scale has been changed from the rectangular co-ordinates which are convenient for plotting $\dfrac{(R' - Y')}{1\cdot14}$ and $\dfrac{(B' - Y')}{2\cdot03}$, to the polar co-ordinates which are convenient for measuring sub-carrier amplitude and phase; a full list of values is given in Table 5.5. The chromaticities produced by these sub-carrier signals are those given in Fig. 4.4.

The locus of the maximum excursions of the sub-carrier signal is the irregular hexagon formed by joining the ends of the sub-carrier vectors for the primary signals, as in Fig. 5.6. Our original R', G' and B' signals can be considered in colour space as being limited by

Table 5.4

LUMINANCE AMPLITUDES FOR 100% AND 75% OF MAXIMUM GAMMA CORRECTED VALUES

Primary	100%	75%
Magenta	0·41	0·31
Red	0·30	0·23
Yellow	0·89	0·67
Green	0·59	0·44
Cyan	0·70	0·53
Blue	0·11	0·08

Table 5.5

SIGNAL VALUES OF LARGE AREA PRIMARY COLOURS AT FULL LUMINANCE

Colour	R'	G'	B'	Y'	(R'−Y')	(G'−Y')	(B'−Y')	$\frac{(R'-Y')}{1\cdot14}$	$\frac{(B'-Y')}{2\cdot03}$	$\frac{(R'-Y')}{1\cdot14\,Y'}$	$\frac{(B'-Y')}{2\cdot03\,Y'}$	Q'	I'	$\sqrt{(Q')^2+(I')^2}$	θ°	φ°
White	1	1	1	1	0	0	0	0	0	0	0	0	0	0	—	—
Yellow	1	1	0	0·886	0·114	0·114	−0·886	0·100	−0·436	0·113	−0·493	−0·312	0·322	0·448	13	167
Cyan	0	1	1	0·701	−0·701	−0·299	0·299	−0·615	0·147	−0·877	0·210	−0·211	−0·596	0·632	256·5	283·5
Green	0	1	0	0·587	−0·587	0·413	−0·587	−0·515	−0·289	−0·877	−0·493	−0·523	−0·274	0·591	299·3	240·7
Magenta	1	0	1	0·413	0·587	−0·413	0·587	0·515	0·289	1·247	0·700	0·523	0·274	0·591	119·3	60·7
Red	1	0	0	0·299	0·701	−0·299	−0·299	0·615	−0·147	2·057	−0·493	0·211	0·596	0·632	76·5	103·5
Blue	0	0	1	0·114	−0·114	−0·114	0·886	−0·100	0·436	−0·877	3·829	0·312	−0·322	0·448	193	347
Black	0	0	0	0	0	0	0	0	0	—	—	0	0	0	—	—

φ is phase with respect to $\frac{(E'_R - E'_V)}{2\cdot03}$ axis.

θ is delay with respect to burst phase.
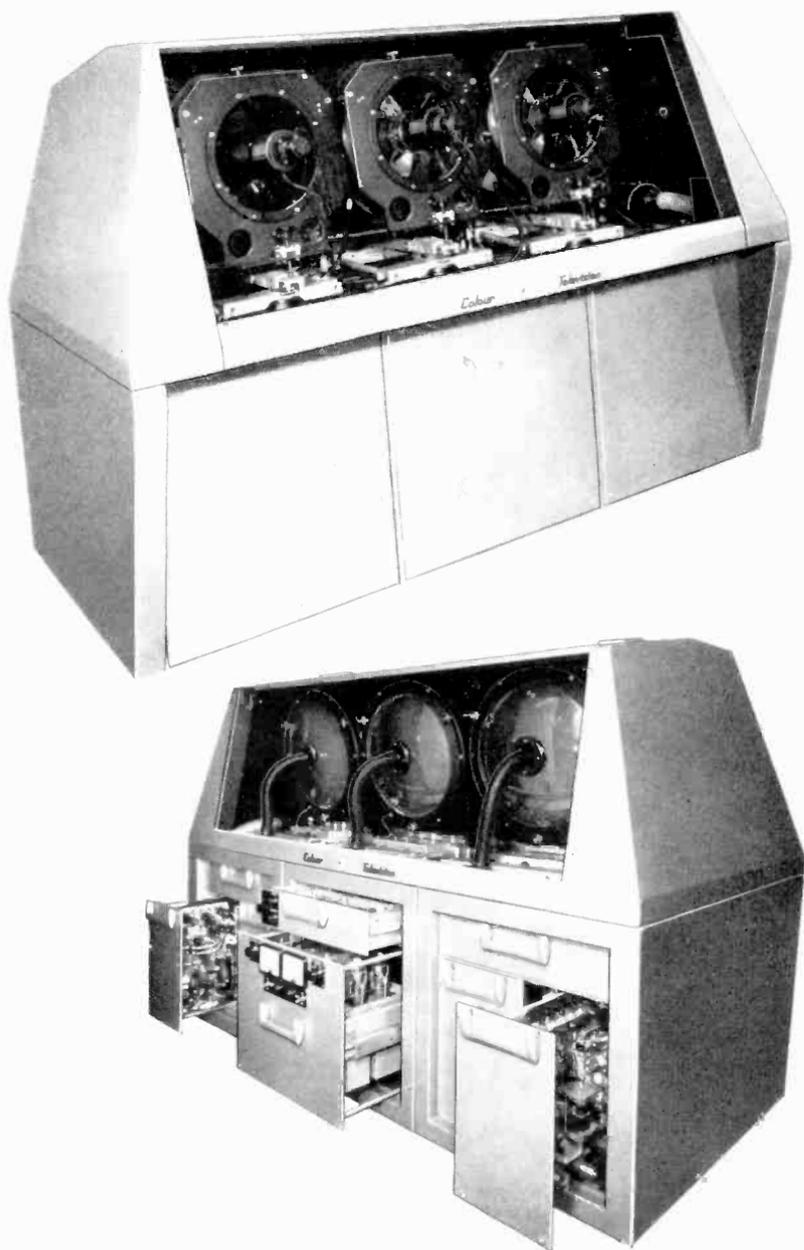
A Marconi large screen colour television projector. Picture sizes up to
12 ft × 9 ft are obtainable with a highlight brightness of 5 foot lamberts.
The three images are projected from three in-line Schmidt optical systems
with electronic correction provided to eliminate distortion due to the
lateral displacement of the two wing tubes. (Courtesy Marconi Ltd.)

A 405-line N.T.S.C. picture reproduced on a shadow mask tube

DELAYING BURST WITH RESPECT TO SIGNAL BY INSERTING CABLE IN PATH OF REFERENCE
ROTATES DEMODULATION AXES ON THIS DIAGRAM CLOCKWISE. ORANGE FLESH TONES ARE
THEN REPRODUCED TOO YELLOW—GREEN.

Primary sub-carrier signals for 100% and 75% of maximum gamma corrected values.

Transmitter tolerances are $\pm 10°$ in phase and $\pm 20\%$ (approximately 2dB) in ratio of sub-carrier to luminance amplitude.

Fig. 5.5. Sub-carrier phase and amplitude coefficients for large areas of saturated primaries

UNIT CUBE

(a)

(b)

PROJECTION ON
SUB-CARRIER PLANE
AFTER LINEAR
TRANSFORMATION

(c)

Weighting factors reduce maximum levels for R' to 0·632, G' to 0·593, B' to 0·447. Without weighting factors, Q' axis is 20·0° anticlockwise from $(B' - Y')$ axis with scale factor of 1·81Q'. I' axis is 139·1° anticlockwise from $(B' - Y')$ axis with scale factor of 1·46I'.

Fig. 5.6.   Maximum excursions of sub-carrier signals

a unit cube, and the transmitted signal excursions are derived from this cube by linear transformation so that straight lines remain as straight lines, although the angles between them, and also their lengths, alter. The hexagon in the sub-carrier plane is then the projection of the transformed cube as in Fig. 5.6.

5.6. Chrominance signals and colour triangle for unity gamma

In Chapter 4 the N.T.S.C. transmitted signals Y', I' and Q' were derived from the R', G' and B' camera signals by a series of simple calculations which are described mathematically as linear transformations. For any colour whose tristimulus values are R, G and B, it is possible to calculate both the chromaticity co-ordinates r, g and b (*see* Section 2.11), and the chrominance signals $(R' - Y')$ and $(B' - Y')$, or I' and Q'. The chromaticity co-ordinates r, g and b, are independent of the luminance of the colour. The chrominance signals increase linearly with luminance, but their values for unit luminance, $\dfrac{(R' - Y')}{Y'}$, $\dfrac{(B' - Y')}{Y'}$, $\dfrac{I'}{Y'}$ and $\dfrac{Q'}{Y'}$, are constant for a fixed chromaticity. It is thus possible to plot on a chromaticity diagram the corresponding values of chrominance signals for unit luminance.

Fig. 5.7 shows a chromaticity diagram on which the lines of constant r and constant b have been marked. The red primary R^* chromaticity is given by $b = g = 0$ and $r = 1$. Notice that a line from the blue primary, such as BD, represents a constant ratio of $\dfrac{r}{g} = \dfrac{R}{G}$, i.e. all the chromaticities along the line BD are mixtures of the blue primary with a chromaticity D which contains $\dfrac{0\cdot7}{0\cdot3}$ times as much R^* as G^*. All the lines which represent a constant value of $\dfrac{R}{G}$ pass through B^*. When the chrominance signals are zero, the colour reproduced by the receiver is the system reference white, and this is produced by one unit of R^*, one unit of G^* and one unit of B^*, so that the chromaticity of the reference white is at the centroid of the triangle. This point can be considered as the Y^* primary, since it is the chromaticity reproduced when the other two transmitted parameters are zero.

The line marking the locus of a constant value of $\dfrac{R - Y}{Y} = K$ say, can be found as follows:

since $$Y = lR + mG + nB$$
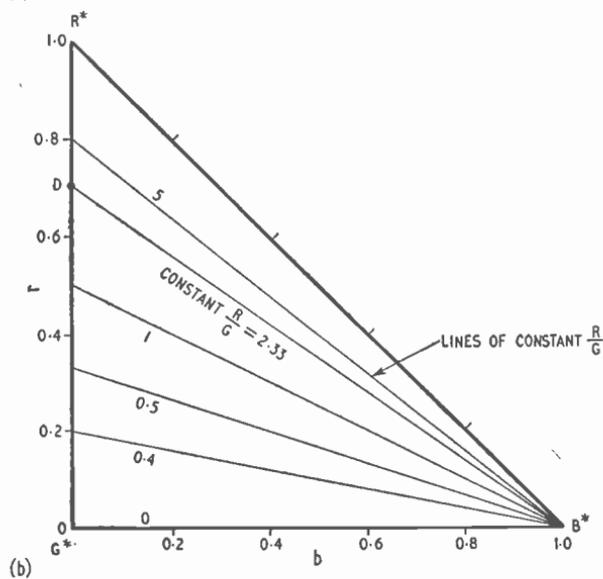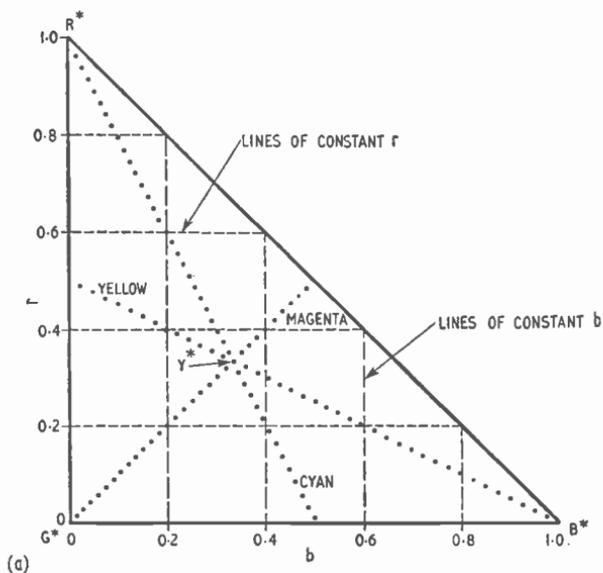
Fig. 5.7. Colour triangles: (a) with lines of constant r *and* b, *and (b) with lines of constant* R/G

where l, m and n are the luminosity coefficients of R*, G* and B*,

$$\frac{(1 - l)R - mG - nB}{lR + mG + nB} = K$$

$\therefore$ $\quad (1 - l)R - mG - nB = KlR + KmG + KnB$

i.e., $\quad [1 - l(1 + K)]R - m(1 + K)G - n(1 + K)B = 0$
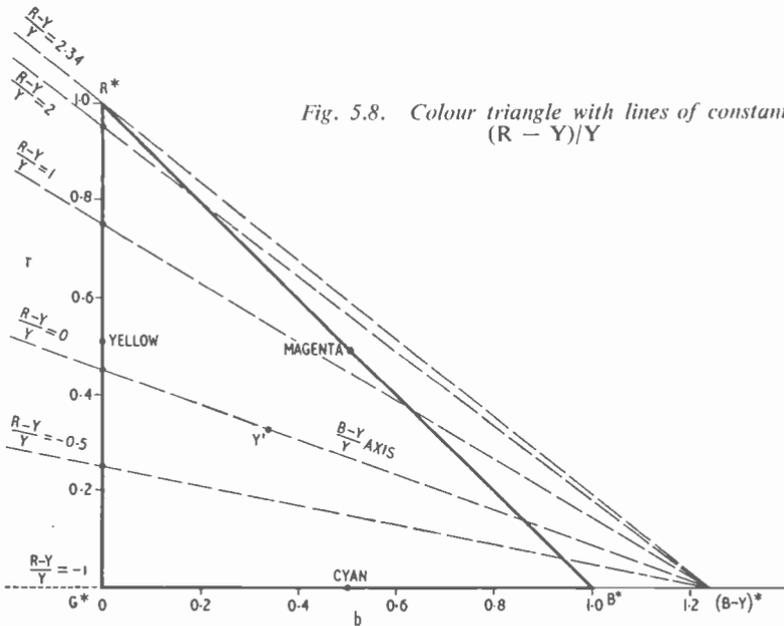
Fig. 5.8. *Colour triangle with lines of constant* $(R - Y)/Y$

dividing throughout by $R + G + B$, (assuming $R + G + B \neq 0$)

$$[1 - l(1 + K)]r - m(1 + K)g - n(1 + K)b = 0$$

$$[1 - l(1 + K)]r - m(1 + K)(1 - r - b) - n(1 + K)b = 0$$

$\therefore$ $\quad [1 - l(1 + K) + m(1 + K)]r - [n(1 + K) - m(1 + K)]b$
$$= m(1 + K)$$

This is a straight line which passes through the points

$$b = 0, \quad r = \frac{m(1 + K)}{1 - (l - m)(1 + K)} \qquad [5.9]$$

and

$$r = 0, \quad b = \frac{m}{m - n} \qquad [5.10]$$

This latter point is independent of the value of K, so that all lines of constant $\dfrac{(R-Y)}{Y}$ pass through it. It can be considered as the $(B-Y)^*$ primary and like the C.I.E. primaries does not represent a real colour. By inserting the values of l, m and n, the co-ordinates of $(B-Y)^*$ are found to be $r=0$, $b=1 \cdot 24$. This value of $b=1 \cdot 24$ places the chromaticity of $(B-Y)^*$ well outside the RGB triangle in a super-purity position which is outside the range of physically realizable colours. By inserting specific values of K in Equation 5.9, the intercepts which various lines make on the r axis are easily found (*see* Fig. 5.8). Similarly, the $(R-Y)^*$ primary can be located at $b=0$, $r=2 \cdot 04$ and the lines of constant $\dfrac{(B-Y)}{Y}$ can be found and are plotted in Fig. 5.9. From Figs. 5.8 and 5.9 the values of $\dfrac{(R-Y)}{Y}$ and $\dfrac{(B-Y)}{Y}$ can be read for any given chromaticity of the transmitted colour.

The line for $\dfrac{(R-Y)}{Y}=0$ passes through the Y* primary and can be considered as the $\dfrac{(B-Y)}{Y}$ axis; similarly $\dfrac{(B-Y)}{Y}=0$ is the line passing through the Y* primary which is the $\dfrac{(R-Y)}{Y}$ axis.

Instead of calculating the positions of $\dfrac{(R-Y)}{Y}=$ constant, the graphical construction mentioned in Chapter 2 can be used. The line joining $(B-Y)^*$ and $(R-Y)^*$ also passes through $(G-Y)^*$ and is the line of zero luminance, the *alychne* (*see* Section 2.19).

If we had known the position of this line we could have located $(R-Y)^*$ and $(B-Y)^*$ by its intersection with the sides of the colour triangle. Any line parallel to the alychne such as UV in Fig. 5.10, would have enabled the positions of the lines of $\dfrac{(R-Y)}{Y}=$ constant to be drawn. The line from $(R-Y)^*$ to Y* is the line $\dfrac{(B-Y)}{Y}=0$. The line from $(R-Y)^*$ to G* is the line $\dfrac{(B-Y)}{Y}=-1$. The intercept " e " then defines the interval for $\dfrac{(B-Y)}{Y}$ varying by steps of 1, and the scale along UV is linear.
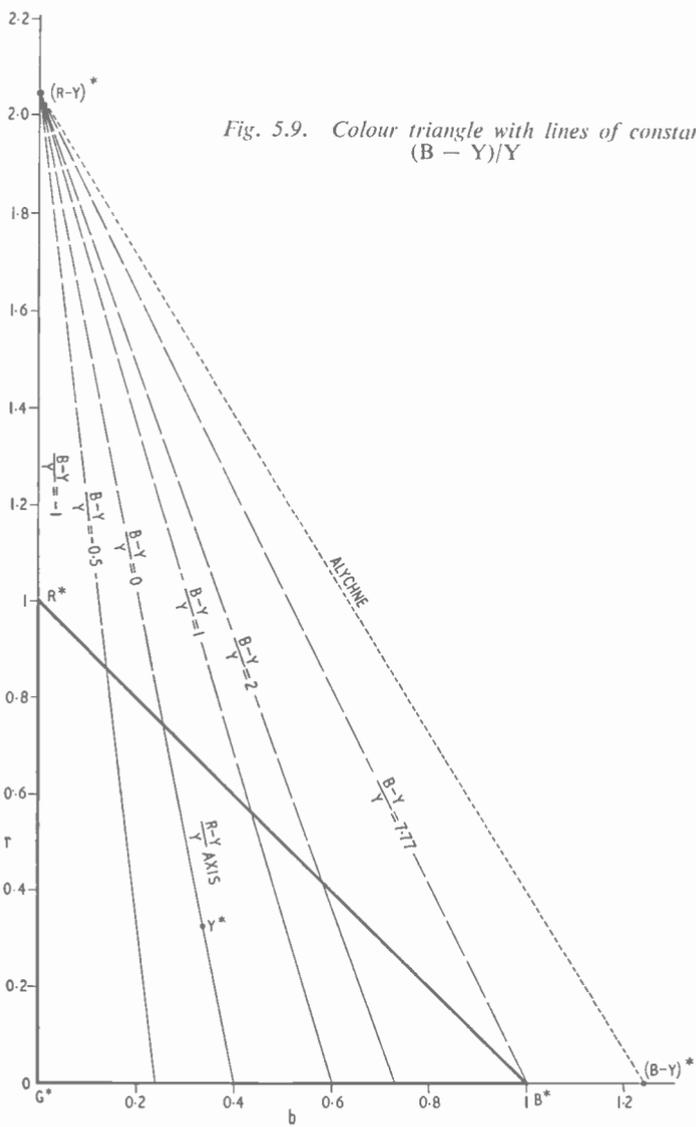
Fig. 5.9. Colour triangle with lines of constant $(B - Y)/Y$
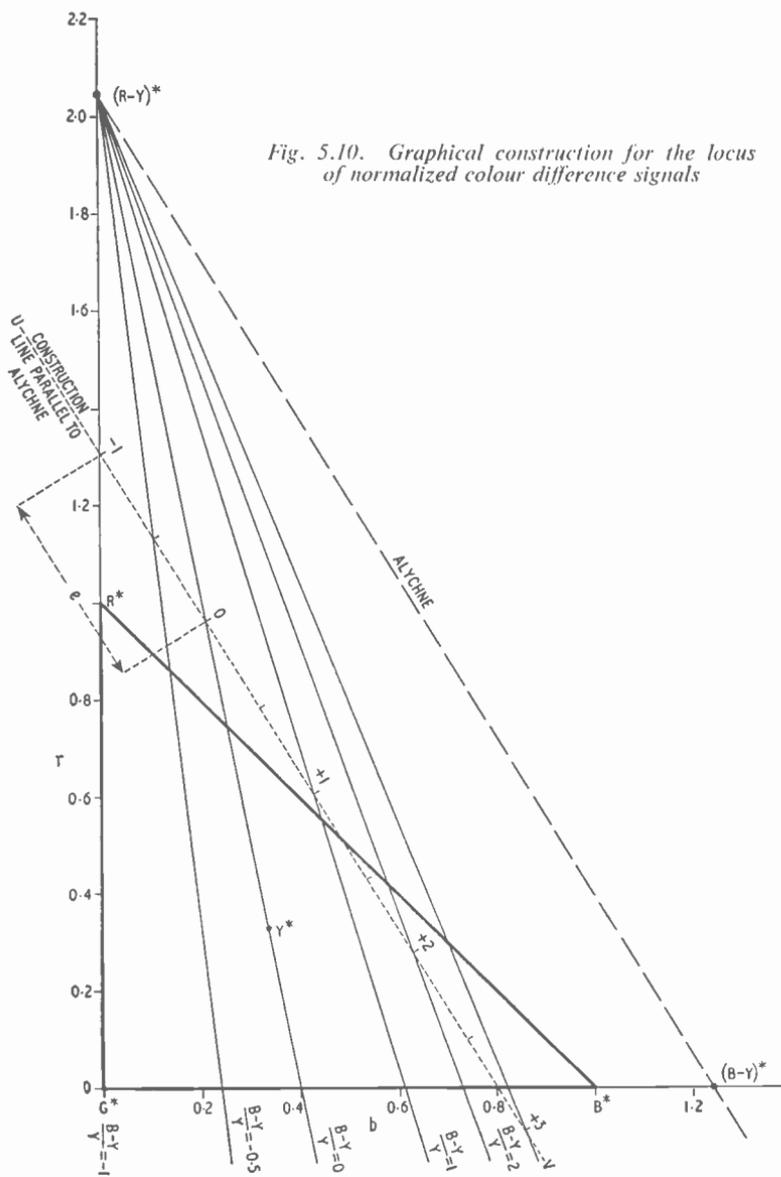
Fig. 5.10. *Graphical construction for the locus of normalized colour difference signals*
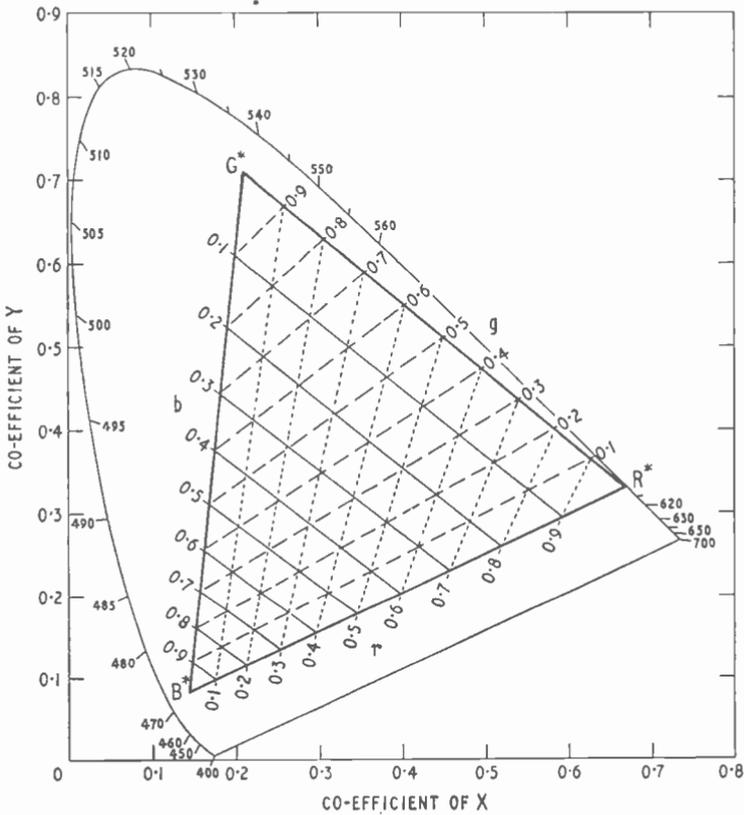
Fig. 5.11. Lines of constant r, g, b *on the C.I.E. chromaticity diagram*

The alychne could have been constructed from the equiluminous point, which is $r = 0·242$ and $b = 0·635$, for the N.T.S.C. primaries and Illuminant C.   This may be calculated from the condition that $lR = mG = nB$, and $l = 0·299$, $m = 0·587$ and $n = 0·114$.

Similar calculations can be made for I* and Q*, but it is advantageous to use the C.I.E. chromaticity diagrams.

5.7.   Chrominance signals on the C.I.E. chromaticity diagram

5.7.1.   COLOUR DIFFERENCE SIGNALS, $\gamma = 1$

Using the formulae in Section 4.30 the chromaticity of a colour $RR^* + GG^* + BB^*$ can be converted to x and y co-ordinates, and Fig. 5.11 shows lines of constant r, g and b.   This diagram
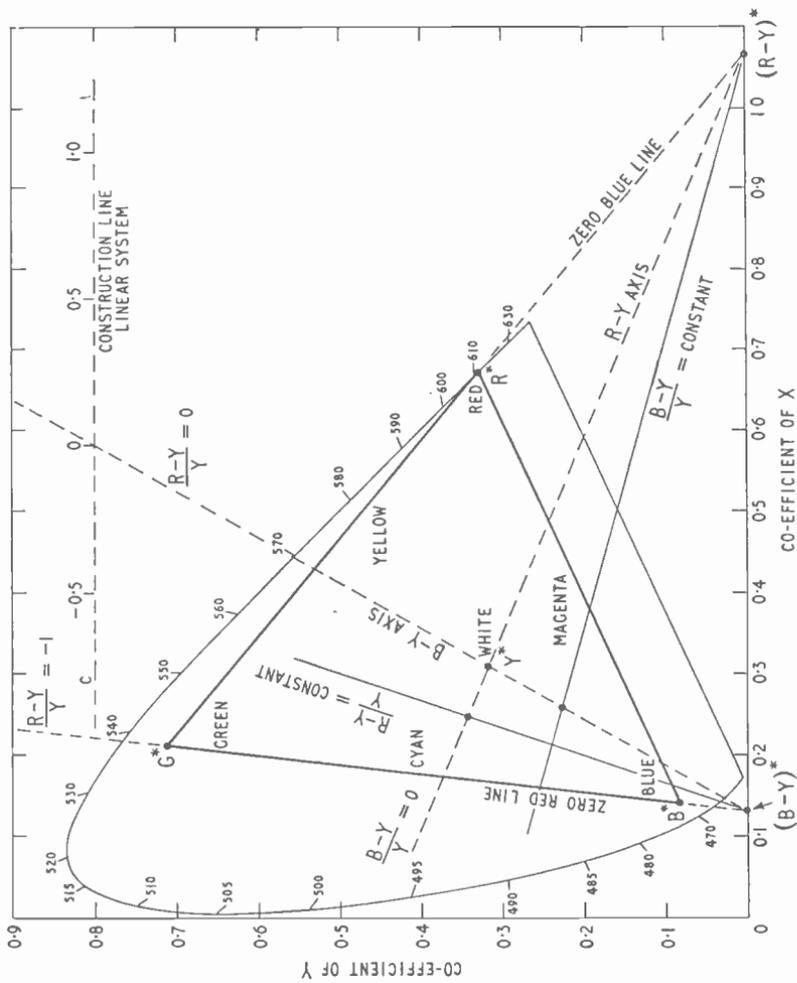
Fig. 5.12. Construction for lines of constant normalized colour difference signals on C.I.E. diagram
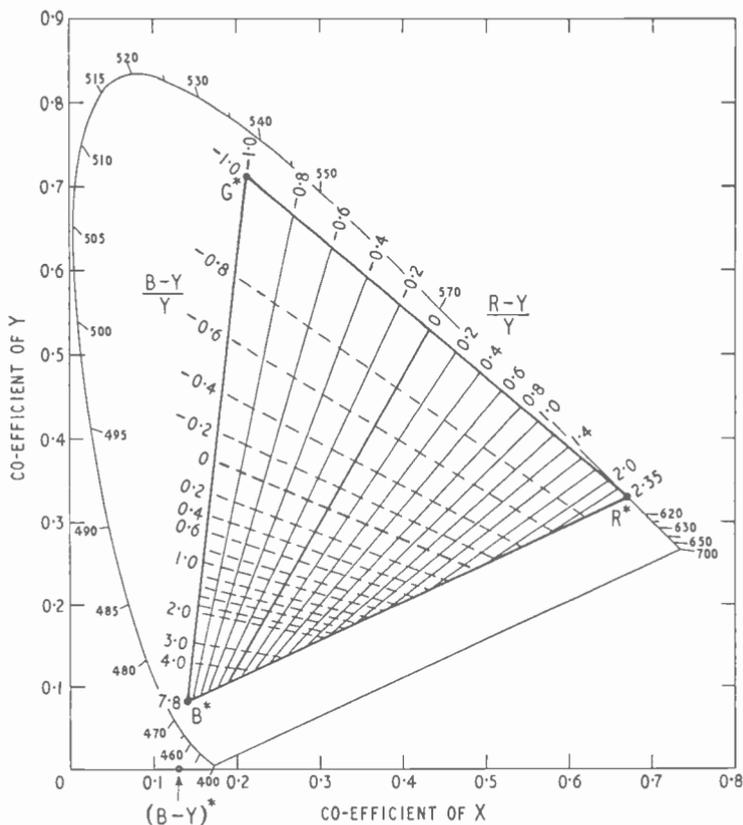
Fig. 5.13. Lines of constant normalized colour difference signals, unity gamma, on the C.I.E. chromaticity diagram

can also be constructed graphically, as described in Section 2.14. The transmission primary points Y*, $(R - Y)*$ and $(B - Y)*$ can then be found (*see* Section 5.6) and here also a graphical construction saves much tedious calculation (Fig. 5.12).   Since the chrominance primaries do not contribute to the reproduced luminance (as $\gamma = 1$) they must lie on the $Y = y = 0$ axis.   Also, as the line joining G* to R* is a line of constant $\dfrac{(B - Y)}{Y}$, it must pass through $(R - Y)*$.   The intersection of $y = 0$ and the line G*R* is therefore the $(R - Y)*$ primary and is the point $x = 1 \cdot 070$ and $y = 0$. $(B - Y)*$ is at the intersection of $y = 0$ and the line through G*B* at $x = 0 \cdot 131$ and $y = 0$.   The family of straight
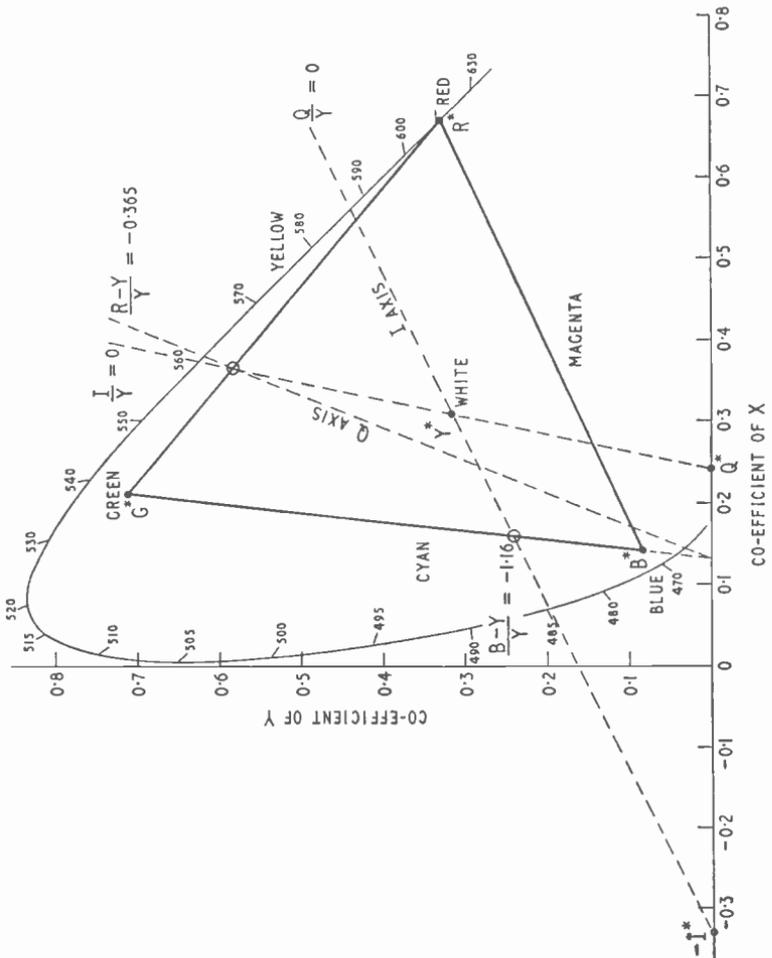
Fig. 5.14. Construction of I* and Q*, unity gamma, on the C.I.E. chromaticity diagram

lines through each chrominance primary makes linear intercepts on a line parallel to $y = 0$. The scale of these intercepts is determined from lines through the white point Y*, which give $\frac{(B - Y)}{Y} = 0 = \frac{(R - Y)}{Y}$, and from the lines through R*G* and

B*G* which give $\frac{(R - Y)}{Y} = -1 = \frac{(B - Y)}{Y}$.   Fig. 5.13 gives a plot

of such lines.   Notice that these are also plots of $\frac{R}{Y} = \frac{(R - Y)}{Y} + 1$

and $\frac{B}{Y} = \frac{(B - Y)}{Y} + 1$.   A similar construction will give $\frac{G}{Y}$.

5.7.2.   I* AND Q* SIGNALS WHEN $\gamma = 1$

The chrominance signals I and Q, normalized to unit luminance, $\frac{I}{Y}$ and $\frac{Q}{Y}$, can be similarly plotted.   The I* and Q* primaries are also zero luminance primaries and lie on $y = 0$. A simple graphical construction for the Q* primary is to consider the line through Y* and Q*, i.e. the line for which $\frac{I}{Y} = 0$ and hence

$$I = 0\cdot74(R - Y) - 0\cdot27(B - Y) = 0$$

Along this line, which is the Q axis,

$$(R - Y) = 0\cdot365(B - Y)$$

and

$$\therefore \qquad \frac{(R - Y)}{Y} = 0\cdot365 \frac{(B - Y)}{Y}$$

It must cut the side G^*R^* of the receiver primary triangle at the value for which $\frac{(B - Y)}{Y} = -1$, i.e. at $\frac{(R - Y)}{Y} = -0\cdot365$.   This point is readily located from Fig. 5.13.   The line from this point through Y* cuts the $y = 0$ axis at Q*, $x = 0\cdot245$ and $y = 0$ (Fig. 5.14).   A similar construction shows that I* lies on the negative side of the y axis, at $x = -0\cdot333$, $y = 0$.   Lines of constant $\frac{I}{Q}$ and $\frac{Q}{Y}$ can then be constructed, as in Fig. 5.15, using the lines through
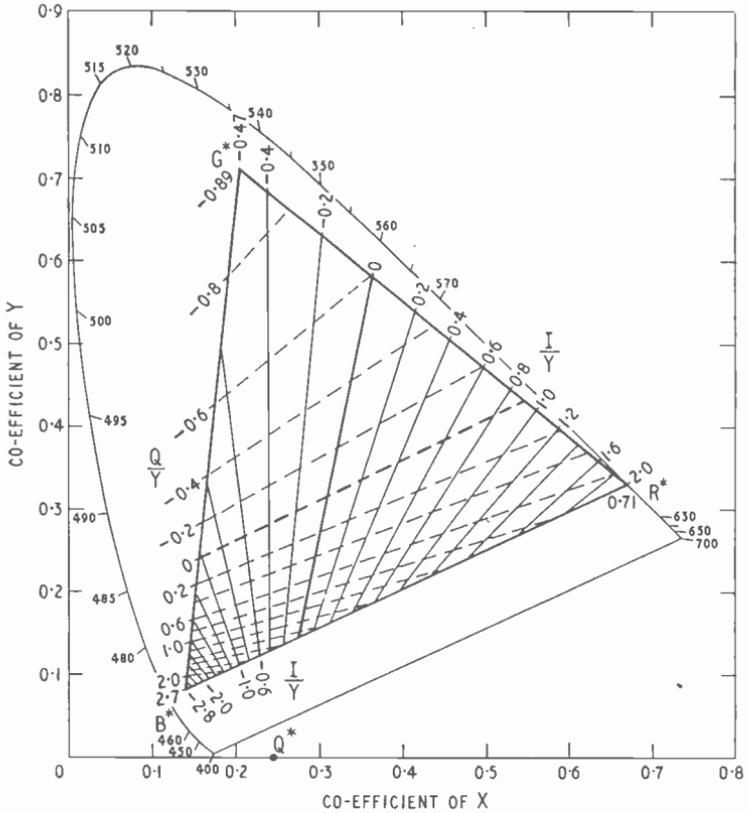
Fig. 5.15.　Lines of constant normalized chrominance transmission signals, unity gamma, on the C.I.E. chromaticity diagram

Y^* for the $\dfrac{I}{Y} = 0 = \dfrac{Q}{Y}$ lines and noting that $\dfrac{I}{Y}$ equals 2 for the red primary and $\dfrac{Q}{Y}$ equals 2.74 for the blue primary.

As $I = [0\cdot74(R - Y) - 0\cdot27(B - Y)]$ increases in value, the chromaticity point moves away from the I* primary and not towards it, and it is therefore referred to as the $-I^*$ primary.

5.7.3.　SUB-CARRIER AMPLITUDE AND PHASE WHEN $\gamma = 1$

The sub-carrier phase Φ is related to the values of $\dfrac{I}{Y}$ and $\dfrac{Q}{Y}$, thus

$$\Phi = 33° + \tan^{-1}\frac{I}{Q} = 33° + \tan^{-1}\frac{\dfrac{I}{Y}}{\dfrac{Q}{Y}}$$

$$= \tan^{-1}\frac{2\cdot03}{1\cdot14}\frac{(R-Y)}{(B-Y)} = \tan^{-1} 1\cdot78 \frac{\dfrac{(R-Y)}{Y}}{\dfrac{(B-Y)}{Y}}$$

The lines of constant sub-carrier phase are lines representing a fixed ratio of $\dfrac{I}{Y}$ to $\dfrac{Q}{Y}$ and are straight lines through the Y* primary, as shown in Fig. 5.16.   The locus of constant sub-carrier amplitude for unit luminance

$$\frac{\sqrt{I^2 + Q^2}}{Y} = \sqrt{\frac{I^2}{Y^2} + \frac{Q^2}{Y^2}}$$

is an ellipse, and curves for various normalized sub-carrier amplitude values have been plotted in Fig. 5.16.   These curves can be plotted from the values of $\dfrac{I}{Y}$ and $\dfrac{Q}{Y}$ for any given chromaticity, derived from Fig. 5.15.   For any given sub-carrier phase the variations in lumin-ance and sub-carrier amplitude make the reproduced chromaticity move along a straight line from the white point, so that a given sub-carrier phase represents a particular dominant wavelength. On the other hand, for a given value of sub-carrier amplitude per unit luminance, variations in sub-carrier phase produce mainly variations in dominant wavelength, although it is clear that these are always accompanied by small changes in purity.   The colour change resulting from any change in sub-carrier angle or amplitude can be deduced at once from this diagram if the luminance is known. When interpreting the visual appearance of the chromaticity change it must be remembered that the eye is not equally sensitive all over the diagram, and reference should be made to Fig. 2.18.

5.7.4.   CHROMATICITY MAPS WHEN $\gamma \neq 1$

So far we have assumed that the receiver cathode-ray tube has a linear relationship between voltage input and light output.   Practical colour tubes are not linear and the N.T.S.C. system predistorts the signals to correct for this, using in general a value of gamma equals 2·2.   For a colour with tristimulus values R, G and B,

the signal voltages are proportional to $R^{\frac{1}{2\cdot 2}}$, $G^{\frac{1}{2\cdot 2}}$ and $B^{\frac{1}{2\cdot 2}}$. The luminance signal is not Y, but

$$Y' = lR^{\frac{1}{2\cdot 2}} + mG^{\frac{1}{2\cdot 2}} + nB^{\frac{1}{2\cdot 2}}$$

The diagrams in Figs. 5.13, 5.15 and 5.16 must be revised for all cases where $\gamma \neq 1$. For purities up to about 30% the errors are small, but become marked for saturated colours near the complementary primaries. The new curves may be calculated from the equations of Sections 4.25 to 4.30, but a graphical approach may be more helpful to the reader.
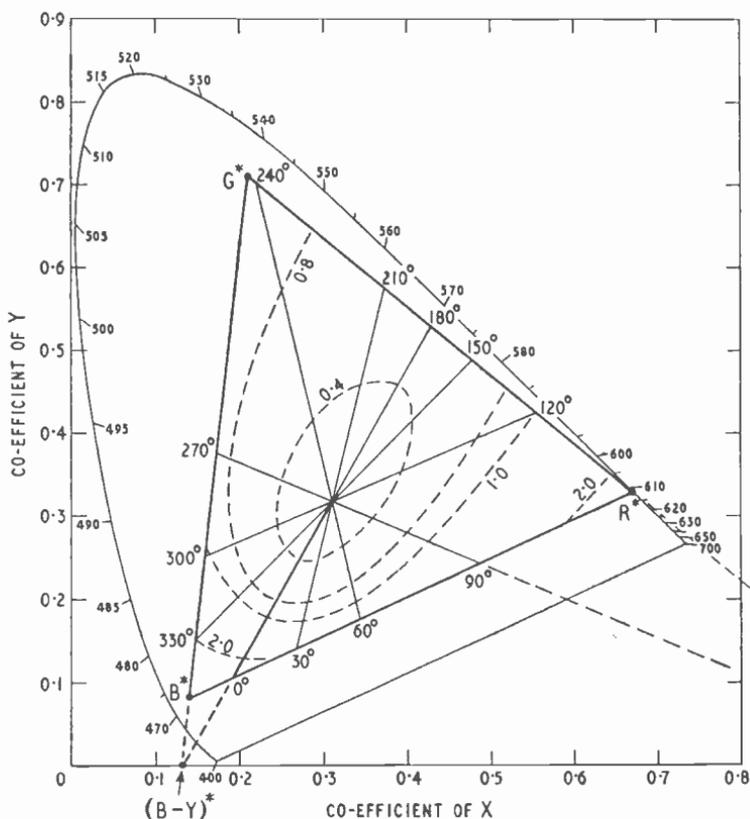
Fig. 5.16. Lines of constant sub-carrier phase and normalized amplitude, unity gamma, on the C.I.E. chromaticity diagram
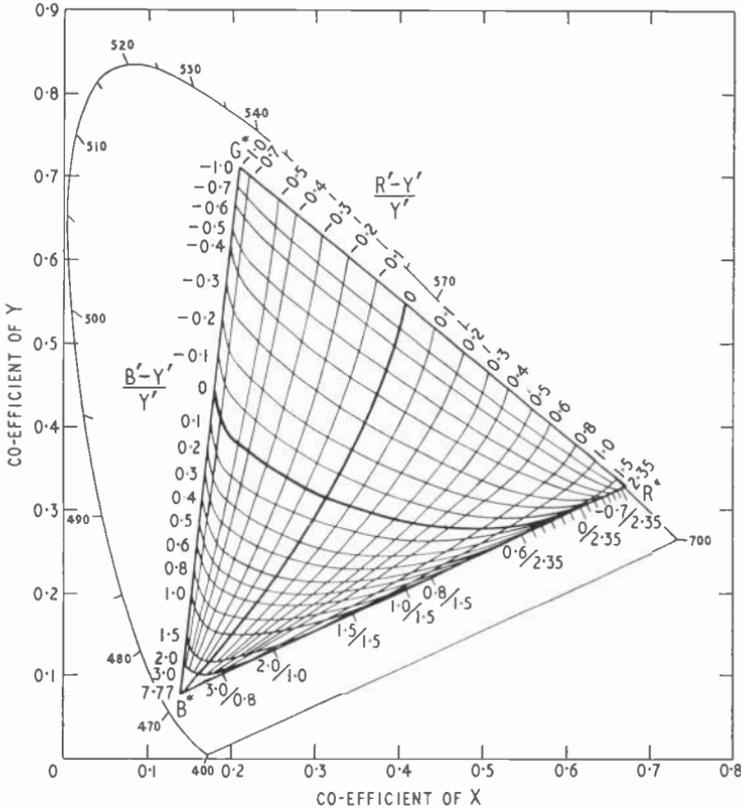
Fig. 5.17.   *Lines of constant normalized colour difference signals,* $\gamma = 2\cdot2$, *on the C.I.E. chromaticity diagram. Intersections along* B*, R* *line shown as* $\dfrac{(B' - Y')/Y' \ value}{(R' - Y')/Y' \ value}$

For the gamma corrected case

$$\frac{(R' - Y')}{Y'} = \frac{R'}{Y'} - 1$$

$$= \frac{R^{\frac{1}{2\cdot2}}}{Y'} - 1$$

Now Y' is not $Y^{\frac{1}{2\cdot2}}$ but $lR' + mG' + nB'$ so that

$$\frac{(R' - Y')}{Y'} = \frac{Y^{\frac{1}{2\cdot2}}}{Y'}\left(\frac{R}{Y}\right)^{\frac{1}{2\cdot2}} - 1$$

11

Curves of constant $\left(\dfrac{R}{Y}\right)^{\frac{1}{2.2}}$ can be drawn upon the chromaticity diagram using the construction of Fig. 5.12, but substituting for the linear scale cd on the construction line a scale of $a^{\frac{1}{2.2}}$. If the transmitted luminance signal were a true luminance signal so that $Y^{\frac{1}{2.2}} = Y'$ the lines of $\dfrac{(R' - Y')}{Y'}$ would still be linear. In fact, a plot of $\dfrac{(R' - Y')}{Y'}$ is shown in Fig. 5.17. Diagrams for the normalized
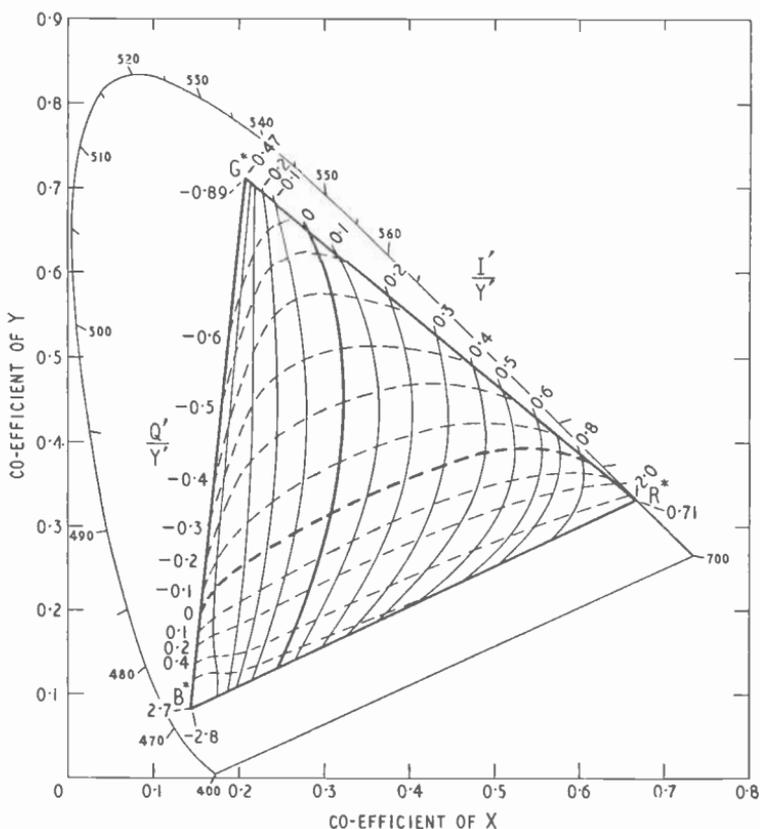
Fig. 5.18. Lines of constant normalized chrominance transmission signals, $\gamma = 2\cdot2$, on the C.I.E. chromaticity diagram. (Reproduced from "Television Engineering Handbook" (Editor D. G. Fink), by Bailey and Loughlin. Copyright 1957, McGraw Hill Book Company, Inc.)
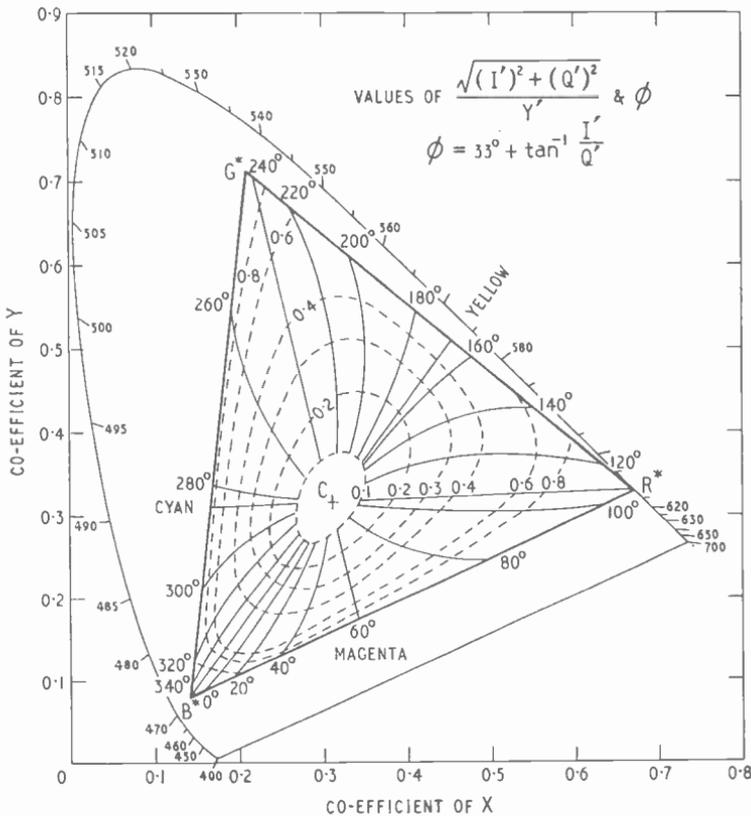
Fig. 5.19. Lines of constant sub-carrier phase and normalized amplitude,
$\gamma = 2 \cdot 2$, on the C.I.E. chromaticity diagram

chrominance signals $\dfrac{I'}{Y'}$ and $\dfrac{Q'}{Y'}$, and for the sub-carrier amplitude and

phase are given in Figs. 5.18 and 5.19.

On Fig. 5.19, the lines of constant phase angle are curved except for the special cases of the primary hues.

Fig. 5.19 can be used to determine the changes in chromaticity of the reproduced picture which occur when the sub-carrier changes, or when errors in the sub-carrier amplitude or phase are introduced into the receiver.

Although a small change in sub-carrier phase makes a dispro-portionate change in chromaticity in the green and magenta regions
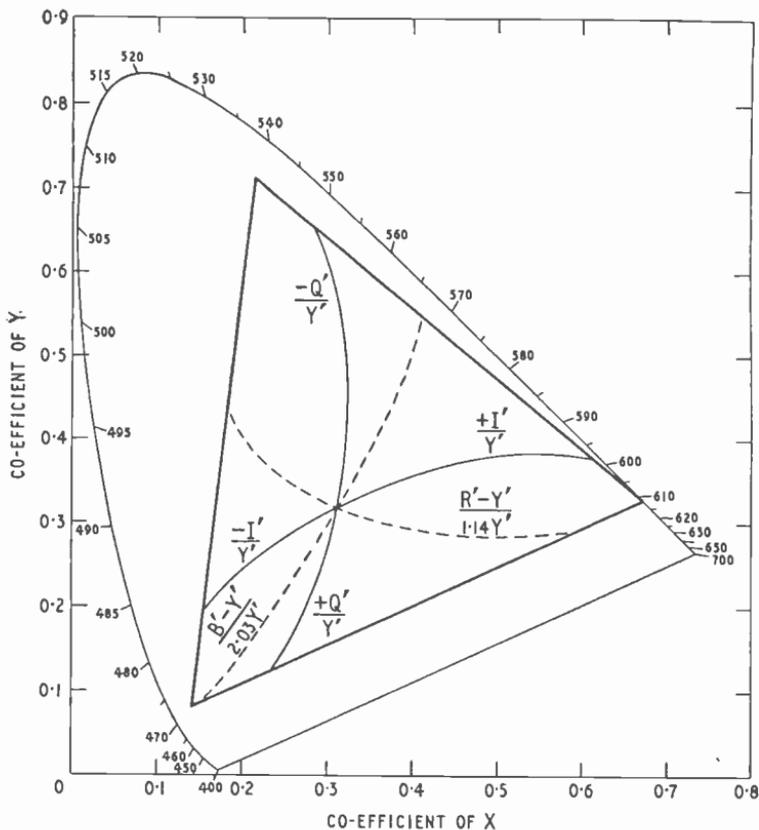
Fig. 5.20.   Chrominance signal axes on the C.I.E. chromaticity diagram

compared with a similar phase change in the blue region, the net subjective effect on the eye is fairly constant all round the diagram (*see* Fig. 2.18).   The $\frac{I'}{Y'}$ and $\frac{Q'}{Y'}$ axes, corresponding to phases of 123° and 33°, can be plotted on the chromaticity diagram and are shown in Fig. 5.20, as are the colour difference axes.

5.8.   Chromaticity diagram on normalized colour difference axes

Fig. 5.21 shows an interesting chromaticity diagram on which the axes are $\frac{(R' - Y')}{1 \cdot 14 Y'}$ and $\frac{(B' - Y')}{2 \cdot 03 Y'}$.   The *RGB* colour triangle is shown and it may be demonstrated that the $\frac{I'}{Y'}$ and $\frac{Q'}{Y'}$ axes make

their correct 123° and 33° angles with the normalized colour difference axes, just at I' and Q' axes do to the $\dfrac{(R' - Y')}{1 \cdot 14}$ and $\dfrac{(B' - Y')}{2 \cdot 03}$ axes on the sub-carrier signal diagram, in Fig. 4.8.   Lines of constant C.I.E. chromaticity co-ordinates x and y are shown for comparison. In general, plots of the chrominance signal axes on a colour triangle do not give the same phase angle relationship as these axes have on the sub-carrier phasor diagram (*see* Fig. 5.5).   An interesting exception is shown in Fig. 5.22(*a*).   The $(R - Y)$ and $(B - Y)$ axes, plotted on a simple RGB colour triangle without weighting factors and for unity gamma, are at approximately 33° to the corresponding I and Q axes.   This is fortuitous and does not, for example, hold for the triangle of Fig. 5.22(*b*).

5.9. Chromaticity transitions at colour changes

Since the Y', I' and Q' signals are not all transmitted with the full system bandwidth, the preceding sections have assumed signals which are constant or changing slowly with time.   For very abrupt changes in colour, the I' and Q' signals take time to follow the changes in chromaticity.   During a succession of very rapid changes
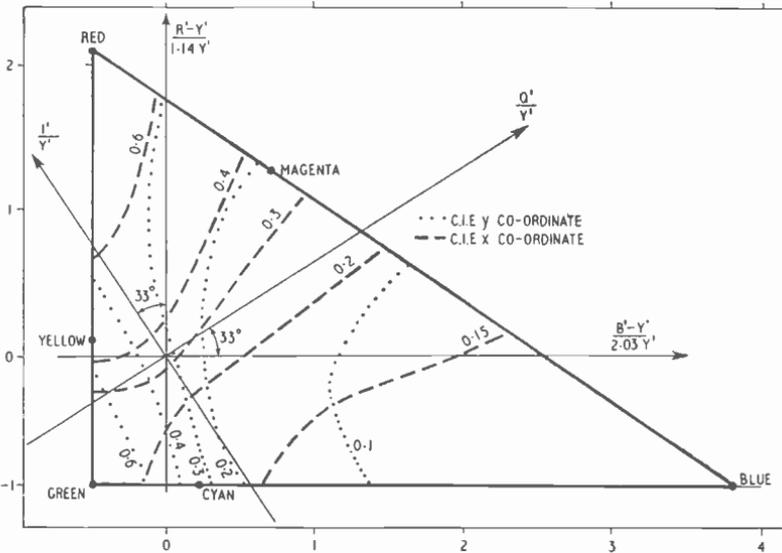
Fig. 5.21.   Chromaticity diagram in terms of $(R' - Y')/1 \cdot 14Y'$ and $(B' - Y')/2 \cdot 03Y'$
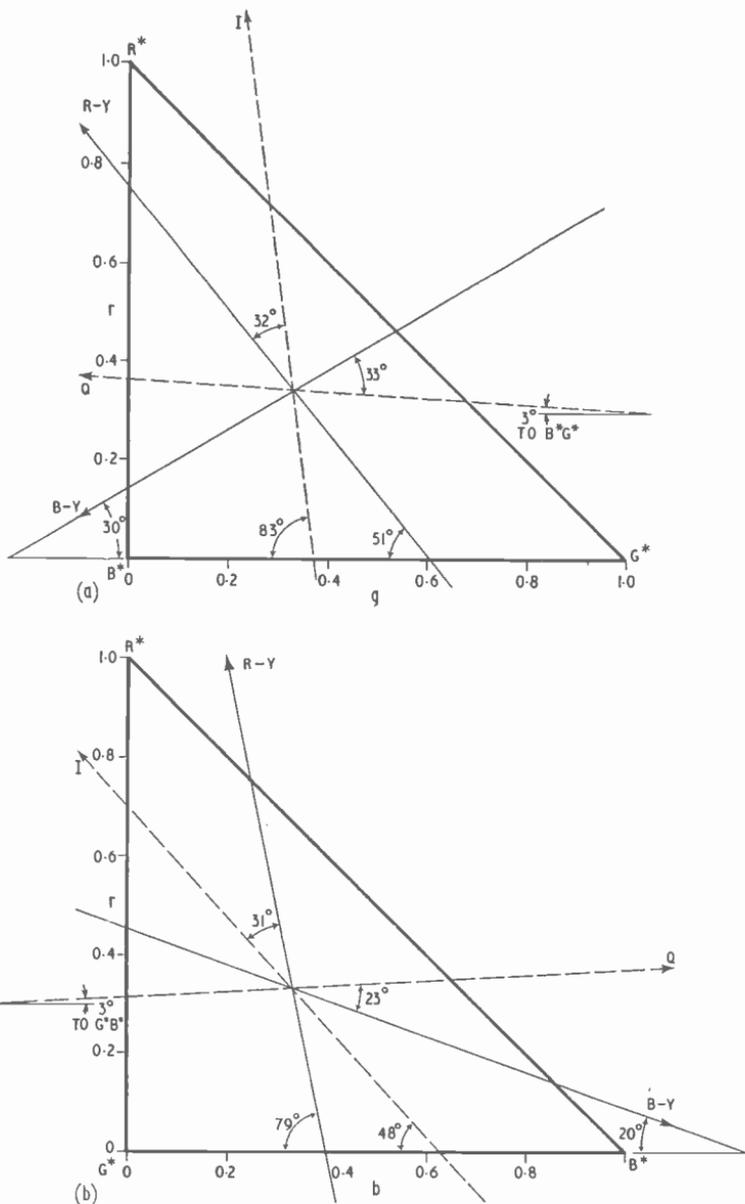
Fig. 5.22. Chrominance axes on simple colour triangle with no gamma correction; (a) with B* at origin; (b) with G* at origin

the I' and Q' signals may not be able to follow fast enough and will adopt mean values.   With frequencies above 0·34 Mc/s in the 405-line system, the Q' signal will only reproduce a mean value, and for frequencies above 1 Mc/s the I' signal also will only follow the mean value.   At such frequencies the luminance signal is the only signal which changes.   This does not mean that the fine detail in
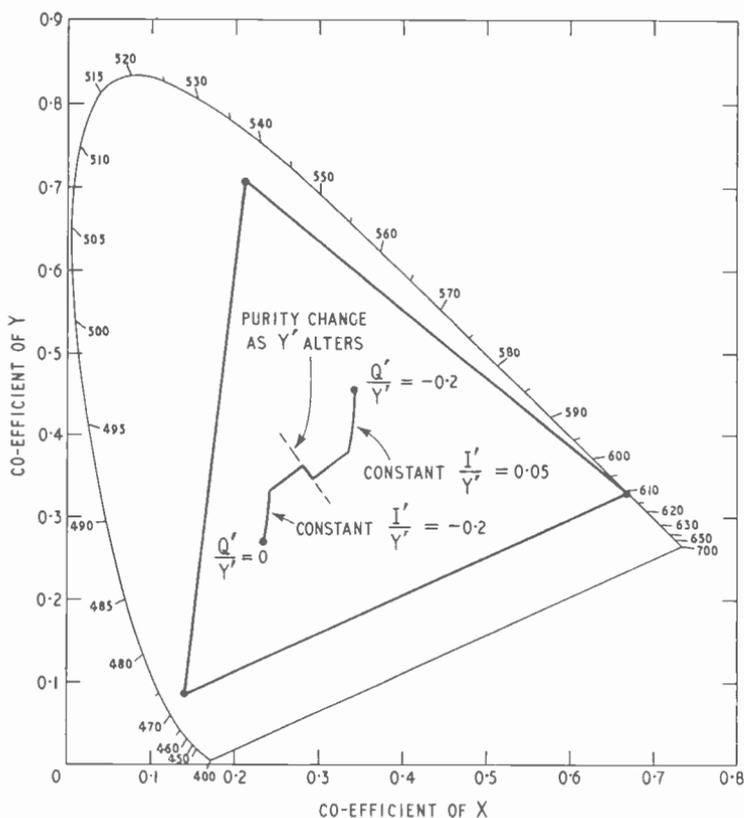
Fig. 5.23.   Reproduced chromaticity change at a colour transition

the picture can only be reproduced in black and white.   The average colour of the fine detail will be reproduced and this will change in brightness with the detail information.   Similarly, picture detail which gives rise to frequencies in the range 0·34 to 1 Mc/s will not be reproduced only in orange-cyan colours, although only the

orange-cyan component in the colour will vary. These considerations only apply along the scanning line; in the vertical direction the full system resolution is obtained in all colours.

Because of the bandwidth limitation the chromaticity reproduced by the receiver during a colour transition may not follow the chromaticity change at the transmitter, although the final reproduced chromaticity will be correct. The rates of change of the three signals Y', I' and Q' are different. The first signal to vary at a transition is the Q' signal, so that the reproduced chromaticity moves along a line of constant $\dfrac{I'}{Y'}$ until the I' channel signal begins to change (Fig. 5.23) when the chromaticity moves along a path determined by both the $\dfrac{I'}{Y'}$ and $\dfrac{Q'}{Y'}$ signals. When the luminance transition occurs the chromaticity makes an abrupt change in purity, and then continues to trace out the path determined by the $\dfrac{I'}{Y'}$ and $\dfrac{Q'}{Y'}$ waveforms until the I' waveform transition finishes, when it follows a line of constant $\dfrac{I'}{Y'}$ to its final chromaticity. The transition thus has five distinct parts during each of which the chromaticity is moving along a different path. Such transitions are further discussed in Section 16.8.

5.10. Luminance transitions at colour changes

If the constant luminance principle were obeyed exactly, the irregular and slow chromaticity transitions would be masked by the sharp luminance change produced by the Y' waveform. As the Y' signal is not the true gamma corrected luminance, the chrominance sub-carrier is carrying part of the luminance, particularly for colours of high purity.

That part of the luminance carried by the chrominance signal is bandwidth limited and can only change slowly, with the result that the reproduced luminance also shows five distinct changes through a colour transient.

As the sub-carrier goes near zero on a slow colour change, the contribution to luminance which it carries exchanges exactly with the luminance signal.   On a fast colour change, the restricted bandwidth of the chrominance channels makes the chromaticity change slow and the change in the luminance carried by the sub-carrier is correspondingly slow.   This luminance change is therefore relatively broad and does not match up with the fast luminance change carried

by the Y' signal. These effects are discussed in more detail in Chapter 16.

5.11. Summary

A graph of lines of constant sub-carrier phase and amplitude on a C.I.E. chromaticity diagram show that the sub-carrier phase determines mainly the dominant wavelength of the reproduced colour, whilst the sub-carrier amplitude for unit luminance determines mainly the reproduced purity. This is exactly true for the primary colours and nearly true for pastel colours, but becomes more incorrect as the purity of the non-primary colour increases.

During a colour transition the reproduced chromaticity can change along several incorrect paths before assuming its final correct value. The associated luminance change can also have a slow, incorrect variation superimposed upon it. These errors are usually only obvious on special test patterns, such as the colour bars, which involve changes between saturated colours.

Transmitter Coding Circuits

6.1. Introduction

The processing of the *R*, *G* and *B* signals to form a composite N.T.S.C. video signal was described in Chapter 4, and it is the purpose of this chapter to examine the circuits which can be used for performing the various operations required.

The circuits described here have been used by the authors to obtain a fully encoded N.T.S.C. signal on the 405-line standard for colour receiver test purposes, but it should be pointed out that these circuits are not the only possible arrangements which can be used. For this reason, a block diagram of each section of the apparatus will be described and then a suitable circuit will be given for it. Should the reader wish to construct a transmitter, he may apply to the construction those circuits with which he is familiar, provided they carry out the operations indicated in the block diagram.

6.2. Complete block diagram of transmitter

The operations required to form a complete R.F. signal from an *R*, *G* and *B* input signal are shown in Fig. 6.1. It is assumed that each of the three video signals is about 1 volt peak-to-peak amplitude and is line and field blanked, but without sync pulses.

Note that if the *R*, *G* and *B* input signals are obtained from a picture source, they must be individually gamma corrected before entering the unit of Fig. 6.1. On the other hand, gamma correction is not necessary if a colour bar signal (*see* Section 6.8) is used, since in this case each signal has only two possible values, namely, " on " or " off ".

The encoder unit is fed with the *R*, *G* and *B* video signals, a C.W. sub-carrier signal and mixed sync, and its output is a composite N.T.S.C. video signal. The composite video signal is then modulated on to a suitable R.F. carrier. At the same time the sound carrier must be generated and modulated with the audio signal and then both sound and vision carriers are added together. To keep the special frequency relationships which are called for in N.T.S.C. operation, the sub-carrier oscillator is used to provide both the
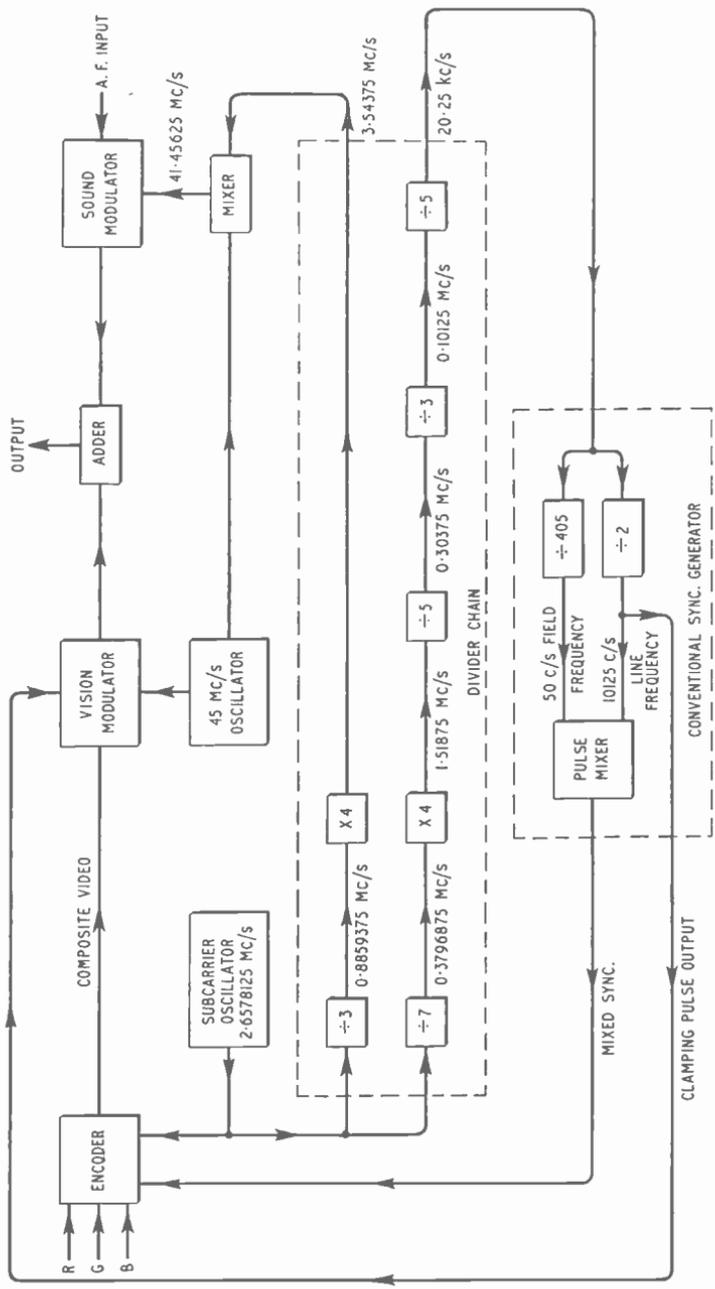
Fig. 6.1. Complete block diagram of an N.T.S.C. transmitter

mixed sync timing and the difference in frequency between the sound and vision carriers.

In order to achieve the correct frequency relationship between the sub-carrier and the scanning frequencies, the twice line frequency oscillator of a conventional sync generator must be replaced by a signal derived from the sub-carrier frequency. This signal can be obtained by dividing the sub-carrier frequency by 525 (i.e. $7 \times 5 \times 5 \times 3$) and multiplying by 4.

Again, in order to reduce the visibility of the R.F. sound carrier/R.F. sub-carrier beat, this beat frequency is made an odd multiple of half the line frequency. Since the R.F. vision carrier/R.F. sub-carrier beat is also an odd multiple of half the line frequency, it follows that the R.F. vision carrier/R.F. sound carrier beat is a whole multiple of the line frequency. The multiple which is most convenient and which gives a beat nearest to the conventional beat frequency of 3·5 Mc/s is 350. The R.F. vision/R.F. sound beat then becomes $350 \times 10{,}125$ c/s, or 3·54375 Mc/s. Since the sub-carrier frequency is $\frac{525}{2}$ times the line frequency, and the R.F. vision/R.F. sound beat is 350 times the line frequency, it follows that the sub-carrier frequency is $\frac{525}{2 \times 350}$ or $\frac{3}{4}$ of the R.F. vision/R.F. sound beat. Hence, the appropriate relationship between the vision and sound carriers can be achieved by first deriving the beat frequency of 3·54375 Mc/s by taking $\frac{4}{3}$ of the sub-carrier frequency, and then mixing this with either the R.F. vision carrier or the R.F. sound carrier. Strictly, the beat should be mixed with the conventional sound carrier of 41·5 Mc/s so that the sound traps of existing monochrome receivers are correctly tuned, but in the equipment used by the authors a crystal controlled 45 Mc/s source was already available so that for convenience the sound carrier was derived by mixing the beat with the vision carrier.

It should be noted that while a public transmitter would be required to conform to the various frequency locking arrangements described above, for most test purposes they are not essential, and an R.F. vision signal of composite video in which the sub-carrier and scanning frequencies are not locked can prove very useful.

The individual blocks of Fig. 6.1 will now be described in detail.

6.3. Encoder

The encoder accepts the R, G and B blanked video signals, a C.W. sub-carrier signal and the mixed sync signal, and provides a
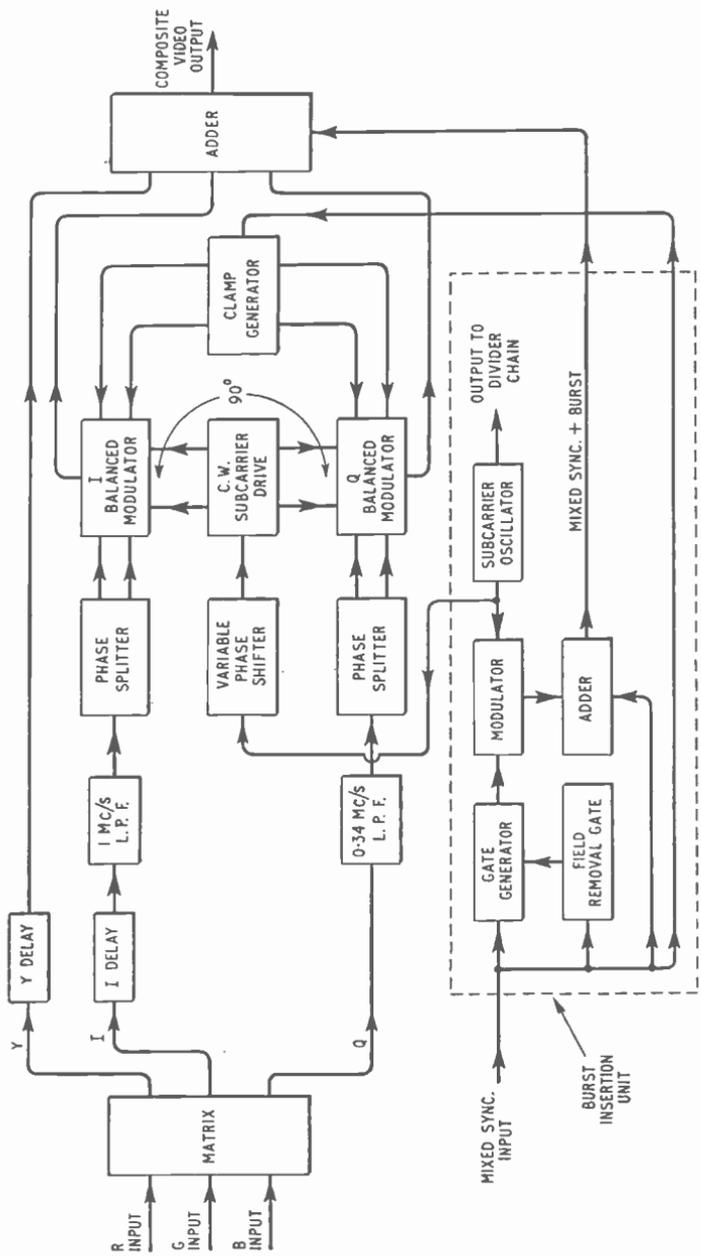
Fig. 6.2. Encoder block diagram

composite video output. A block diagram of the arrangement is shown in Fig. 6.2, while a possible circuit configuration is shown in Figs. 6.3(a) and (b). Referring to Figs. 6.2 and 6.3(a), the R, G and B video signals are first fed into a matrix network which forms the transmission parameters Y, I and Q according to the equations

$$Y = 0{\cdot}30R + 0{\cdot}59G + 0{\cdot}11B$$
$$-I = -0{\cdot}60R + 0{\cdot}28G + 0{\cdot}32B$$
$$Q = 0{\cdot}21R - 0{\cdot}52G + 0{\cdot}31B$$

The appropriate fractions of R, G and B are formed by resistor networks, while a phase inverting stage is used to obtain the negative red signal for the $-I$ parameter, and similarly for the negative green signal for the Q parameter. Note that by deriving $-I$ rather than $+I$, only one phase inversion for red is required instead of two inversions, one for green and one for blue. The $-I$ signal can be readily phase inverted in the modulator stage by appropriate connection of the sub-carrier drive transformer.

The Y signal is delayed by approximately 0·7 μs (see Section 4.22) by means of a ferrite loaded delay line which has a constant resistance termination to prevent mismatch of the delay cable by the input capacity of the following valve.

The I signal is delayed by approximately 0·5 μs also by means of a ferrite loaded delay line with a constant resistance termination, and this feeds the grid of a valve whose anode load is a low pass filter having a cut-off frequency characteristic which is about 1dB down at 1 Mc/s and 20dB down at 1·8 Mc/s. The low pass filter output is then fed to a phase splitting stage which provides a push-pull output to the I balanced modulator. The balanced modulator consists of two 6F33 valves whose anodes are strapped and whose control grids are fed push-pull by the output of the phase splitter, while the suppressor grids are fed push-pull by a C.W. sub-carrier signal of the appropriate phase. The modulator is really doubly balanced in that no output is obtained if either the I signal drive or the C.W. sub-carrier drive is removed. The I modulator output is a sub-carrier sine wave whose amplitude is proportional to the incoming I signal and whose phase is always lagging 57° on the burst phase for a positive I signal, and always leading 123° on the burst phase for a negative I signal. For a zero I signal (which must occur during the line and field blanking periods) no output is obtained.

The Q signal path is identical to that of the I signal with three exceptions. Firstly no delay is required. Secondly the low pass filter cut-off frequency characteristic is about 1dB down at 0·34
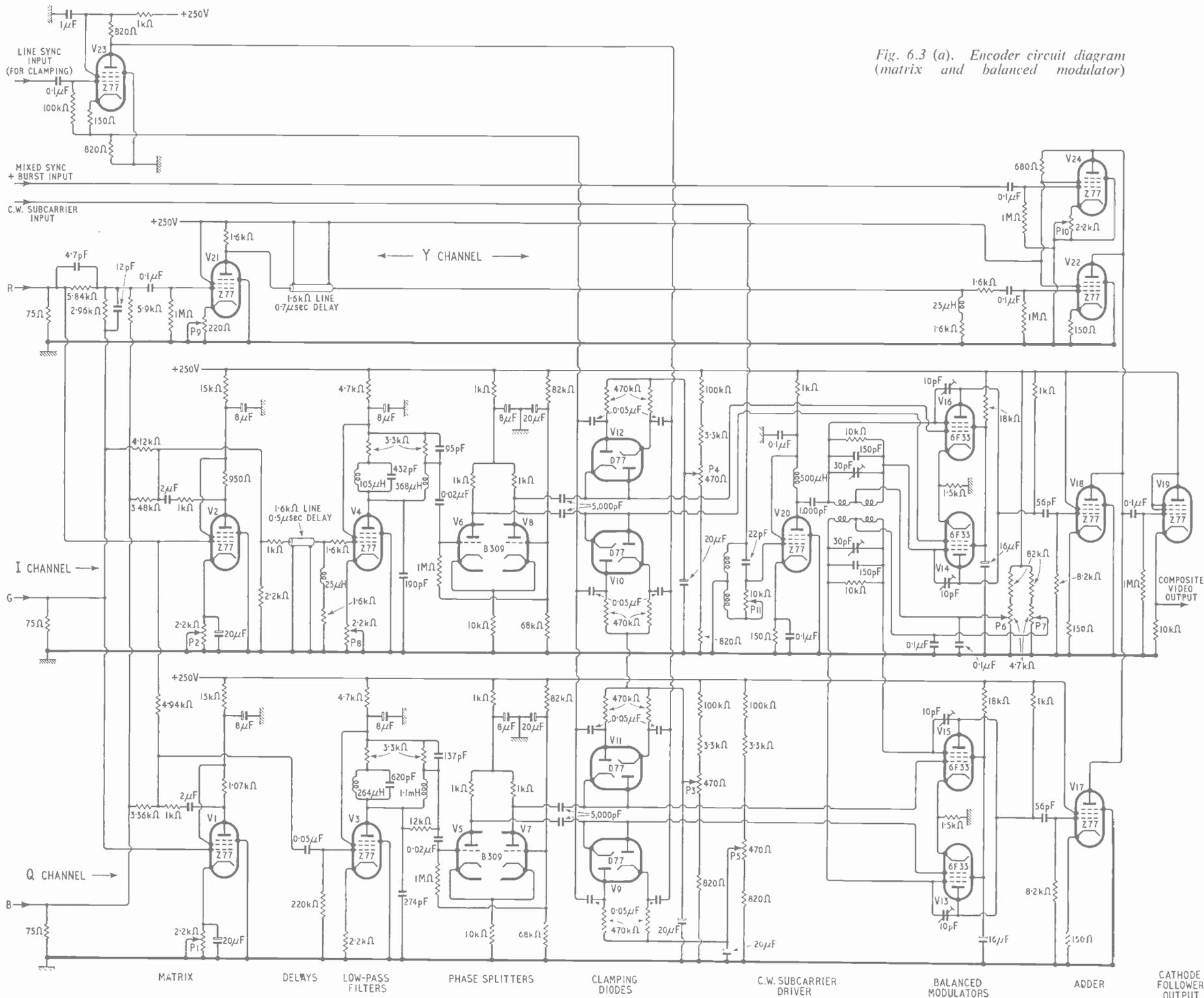
Fig. 6.3 (a). Encoder circuit diagram (matrix and balanced modulator)
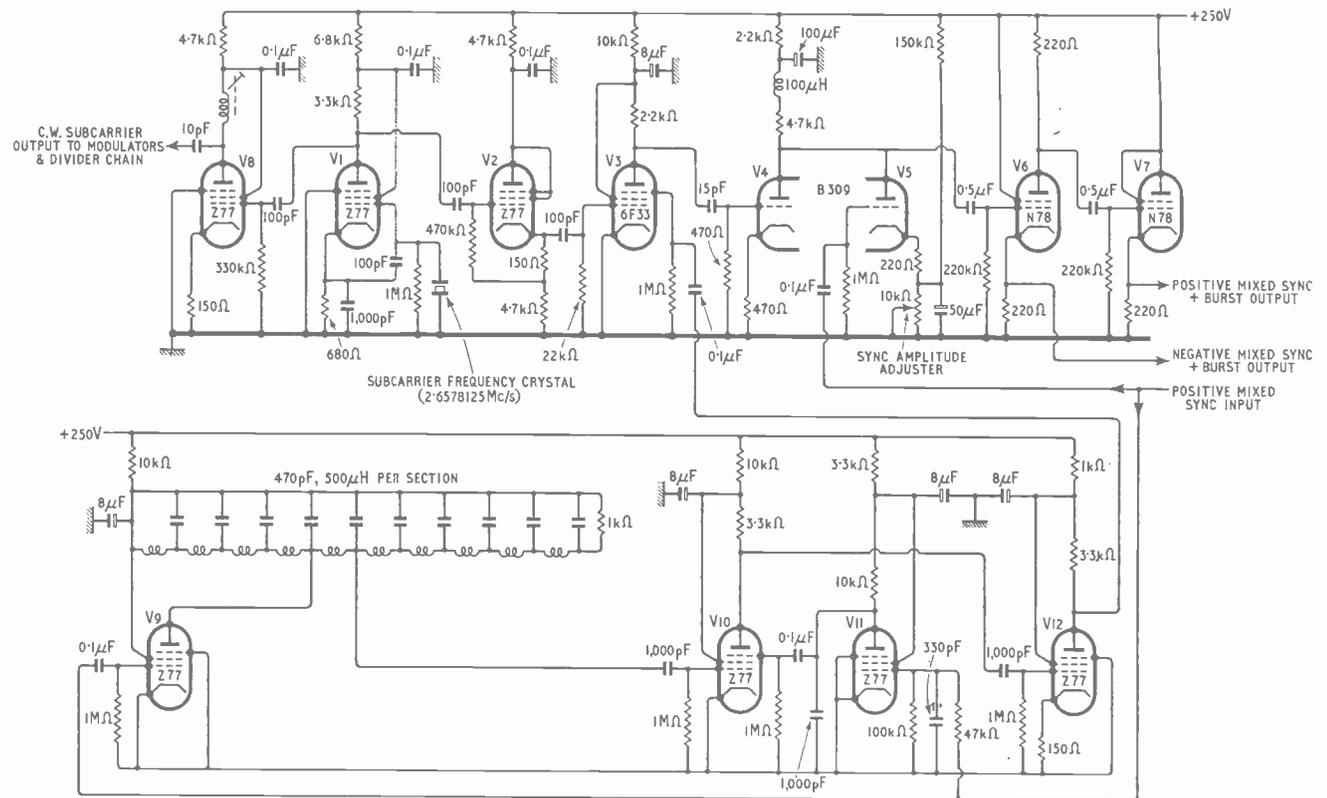
Fig. 6.3 (b). Encoder circuit diagram (burst insertion unit and sub-carrier oscillator)

Mc/s and 10dB down at 0·5 Mc/s, and thirdly the phase of the C.W. sub-carrier push-pull drive to the Q balanced modulator lags 90° on the drive to the I balanced modulator. The Q modulator output is a sub-carrier sine wave whose amplitude is proportional to the incoming Q signal, and whose phase is always lagging the burst phase by 147° for a positive Q signal, and always leading the burst phase by 33° for a negative Q signal.

It is necessary to preserve the D.C. components of the Y, I and Q signals but it is inconvenient to provide D.C. coupling throughout. Hence, A.C. coupling is employed, together with a clamping circuit. Clamping of the I and Q signals can be conveniently carried out at the grids of the I and Q modulators, while Y signal clamping is usually arranged on the composite video signal input to the R.F. modulator. If a video signal only is required, the clamping can be made on the composite video fed to the grid of the output cathode follower.

The principle of the clamping method used here is to arrange to return the point to be clamped (which is usually the grid of a valve A.C. coupled to the previous stage) to a certain D.C. potential at the start of each line scan period. To do this, line frequency pulses are fed to a clamp generator which is essentially a phase inverter providing a push-pull output of equal amplitude but oppositely phased pulses. These outputs are connected to two diodes in series, so that conduction occurs during the pulse period and the anode/cathode junction of the diodes assumes the D.C. potential to which the diode leak resistors are connected. Hence, by connecting the point to be clamped to the junction of the anode and cathode of the diodes, this point will assume the D.C. potential of the diode leak resistors during the pulse time. Note that during the period between pulses, i.e. during the active line scan, the diodes are biased off and do not affect the signal on the point where clamping is applied. The clamping pulses themselves should be long enough to ensure discharge of the coupling capacitors, yet short enough not to interfere during the active line period. As far as the balanced modulators are concerned, the clamping pulses may be as long as the line blanking period, but for clamping the composite video when sync pulses are present, they must obviously not be longer than the line sync period.

For most applications, continuous line sync pulses or mixed sync pulses provide a suitable clamp source, but in refined equipment a pulse generator which develops line frequency repetition pulses having a smaller width than line sync pulses and centred within them in time position, and with relatively poor rise and

decay times, is recommended. Then spurious differentiated pulses due to stray capacity effects are avoided in the waveform being clamped.

The purpose of the burst insertion unit, shown in the lower half of Fig. 6.2 and in circuit form in Fig. 6.3(*b*), is to add the 9 cycles of colour synchronizing burst to the mixed sync waveform. This is done by first forming a suitable gating pulse waveform which
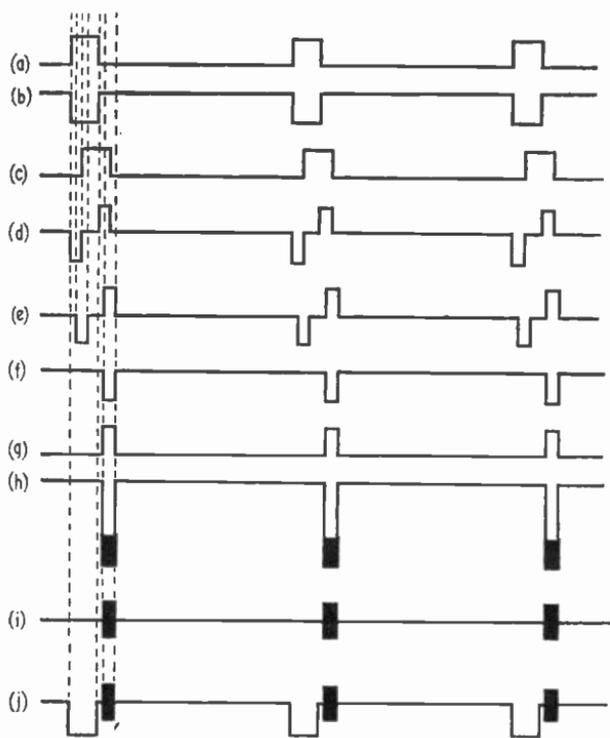
Fig. 6.4. Burst insertion waveforms. Shaded areas represent peak-to-peak sub-carrier sine wave

consists of rectangular pulses of line frequency repetition rate, having a leading edge lagging 1·5 μs on the trailing edge of the line sync waveform, and having a duration of $\frac{9}{2\cdot66}$ μs, that is, 3·4 μs. Since the burst signal should be omitted during the field sync period, a field repetition pulse which has a duration of 4 lines is applied to the gating waveform so that the line frequency pulses are removed

during the field sync period. The field removal gate can be taken from the conventional sync generator, but if this is not readily accessible it can be generated from the mixed sync waveform by means of an integrating and shaping circuit; *see* Fig. 6.3(*b*).

A convenient method of generating the burst insertion gating pulses is by way of a lumped constant delay line. The principle of operation is indicated in Fig. 6.4. Positive going line sync or mixed sync pulses are fed to the grid of a pentode whose anode is connected to a point in the delay line, which is short circuited at one end and characteristically terminated at the other.

Hence, the line sync pulses applied to the grid of the valve (Fig. 6.4(*a*)), are inverted and applied to the delay line (Fig. 6.4(*b*)). Having travelled down the delay line, the pulses are reflected and inverted at the short circuit so that they arrive back at the anode connection as delayed pulses (Fig. 6.4(*c*)). The anode waveform is therefore the sum of those in Fig. 6.4(*b*) and (*c*) and is drawn in Fig. 6.4(*d*). This waveform travels towards the characteristically terminated end of the line, being delayed in the process, and is not reflected from the characteristic termination. The output is therefore as indicated in Fig. 6.4(*e*).

This method has the advantage that there is an independent control of the pulse width (moving the short circuit towards the anode tap reduces the pulse width) and for the time position of the leading edge of the pulses; *see* Fig. 6.4(*e*). (Moving the output tap towards the characteristic termination increases the delay.)

The pulse waveform (Fig. 6.4(*e*)) can now be clipped by a grid current biased valve (a " sync separator " stage) to yield an output (Fig. 6.4(*f*)), and this is finally inverted as shown in Fig. 6.4(*g*). The field removal gate can be applied to the clipping stage to remove the line gate during the field sync period.

The burst insertion gating pulses are fed to a modulator stage which is also fed with a C.W. sub-carrier signal. In effect, this modulator is an amplifying stage which is switched on only during the gating pulse period and its output consists of the required burst signal sitting on a negative pedestal (Fig. 6.4(*h*)). Differentiation of this waveform removes the pedestal to give the burst waveform (Fig. 6.4(*i*)). The differentiation produces " pips " corresponding to the leading and trailing edges of the pedestal but if these edges are not sharp the " pips " are only small in amplitude. Finally, the mixed sync is added to give the mixed sync plus burst waveform as shown in Fig. 6.4(*j*).

The complete composite video waveform is obtained by adding together the *Y* signal, the modulated *I* and *Q* signals, and the mixed

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sync plus burst waveform.   One way of carrying out the addition
is by means of four pentode stages whose anodes are strapped, as
shown in Fig. 6.3(a).

The burst phase should lead the I phase by 57° and in order to
set this correctly a variable phase shifter is included in the C.W.
sub-carrier feed to the sub-carrier driver which operates the I and Q
balanced modulators.   The I phase should, of course, lead the Q
phase by 90°.

Several adjustments are necessary to ensure the correct operation
of the circuits shown in Fig. 6.3(a).

Thus, in the matrix circuit, the inverters V_1 and V_2 must have the
correct gain settings so that the matrix equations

$$-I = -0.60R + 0.28G + 0.32B$$

and

$$Q = 0.21R - 0.52G + 0.31B$$

are satisfied.   This may be done by connecting a video oscillator
operating at about 100 kc/s to the blue input and adjusting the
voltage output until a 1·0V reading is obtained on a valve volt-
meter connected to the grid of V_3.   Next, transfer the oscillator
output to the green input and adjust P_1 until 1·68V is obtained at
the grid of V_3.   This ensures that the green to blue ratio is 0·52
to 0·31, or 1·68, in magnitude.

Similarly, with the oscillator still connected to the green input,
transfer the valve voltmeter to the grid of V_4 and adjust the oscil-
lator output to give a reading of 1·0V.   Next, connect the oscillator
to the red input, and adjust $P2$ until 2·14V is obtained at the grid
of V_4.   This ensures that the red to green ratio is 0·6 to 0·28, or
2·14, in magnitude.   (See Fig. 6.3(a), odd numbered valves up to
17 refer to the Q channel, and even numbered valves up to 18 refer
to the I channel.)

The operating conditions of the balanced modulators may be set
up with the aid of a colour bar generator, as described in Section
6.8.   This generator provides equal peak to peak values of square
waves of R, G and B which simulate a video signal consisting of a
vertical full white stripe immediately following the line blanking,
adjacent to which is a full yellow, then cyan, green, magenta, red,
blue and black.   Thus, full white corresponds to $R = 1$, $G = 1$
and $B = 1$, while full yellow corresponds to $R = 1$ and $G = 1$
and so on.

With the colour bar output signals connected to the appropriate
R, G and B input sockets and with the mixed sync plus burst and

line clamping inputs connected, together with the C.W. sub-carrier input, connect a D.C. voltmeter between the slider of $P3$, $P6$ and $P7$ and earth in turn, and adjust each one to give a reading of 2·5V.

$P3$ sets the reference voltage to which the black level of the video signal is restored and controls one valve of each modulator pair (V_{14} and V_{15}). The second valve of each pair (V_{13} and V_{16}) may be set up as follows:—

With an oscilloscope which is capable of displaying the sub-carrier sine wave and synchronized to the line frequency, observe the waveform at the grid of V_{17}. Adjust $P5$ to balance out the sub-carrier during the line blanking period and white bar period, using the trimmers between anodes and suppressor grids of V_{13} and V_{15} for final adjustment.

Exactly the same procedure may be adopted for balancing the I modulator by connecting the oscilloscope to the grid of V_{18} and adjusting $P4$ and the trimmers connected to V_{14} and V_{16}.

The next step is to set the correct relative levels of the Y, I and Q components. With the composite video output displayed on an oscilloscope synchronized to line frequency, remove the I signal by earthing the grid of V_{18} and adjust $P9$ in the Y channel, so that the amplitude of the white bar above black level is

$$\frac{1}{2 \times 0·52} = 0·96 \text{ times}$$

as great as the peak-to-peak amplitude of sub-carrier on the green or magenta bars. This ensures that the magnitude of the Q sub-carrier signal for full green is 0·52 of the black to peak white amplitude.

Similarly, the correct I sub-carrier amplitude is set up by earthing the grid of V_{17} to remove the Q signal and adjusting $P8$ in the I channel so that the peak-to-peak amplitude of sub-carrier on the red or cyan bars is

$$2 \times 0·6 = 1·2 \text{ times}$$

as great as the amplitude of the white bar above black level.

Notice that the peak sub-carrier amplitude, not the peak-to-peak subcarrier amplitude, is equal to the I or Q signal. However, the peak-to-peak sub-carrier is more convenient to measure on an oscilloscope and this accounts for the factor of 2 in the above relationships.

The final step in setting up the correct amplitude relations consists of adjustment of the sync. amplitude relative to the burst

amplitude, and the amplitude of the burst plus sync waveform relative to Y.

Observe the mixed sync plus burst output on an oscilloscope (*see* Fig. 6.3(b)), and adjust the " sync amplitude adjuster " so that the peak-to-peak sync waveform is equal to the peak-to-peak burst waveform.   Next, observe the composite video output (Fig. 6.3(a)) and adjust $P10$ so that the sync height is 0·43 of the black to white amplitude.   This figure of 0·43 is derived as follows:—

Since 100% carrier corresponds to peak white and 30% carrier corresponds to black, the black to peak white range is 70% carrier. Hence the sync height is $\frac{30}{70}$ or 0·43 of the black to peak white range.

The above adjustments ensure the correct relative amplitudes of burst, sync, Y, I and Q signals but in addition the correct phase relation must obtain between the I and Q sub-carrier signals and the burst.   Thus, I must lead Q by 90° and the burst phase must lead the I phase by 57°.

Measurement of sub-carrier phase angles can be readily carried out by means of a calibrated phase shifter and an oscilloscope whose X amplifier can be fed with an external signal.   A suitable phase shifter is described in Section 12.5, while most oscilloscopes are fitted with an X input terminal, though unfortunately relatively few have an X amplifier with a cut-off frequency above the sub-carrier frequency of 2·7 Mc/s.   It has been found that oscilloscopes having X amplifier bandwidths of only some 300 kc/s can be used at sub-carrier frequency provided that sufficient amplitude of signal is available.

The principle of phase measurement is to apply one signal (preferably the smaller) to the Y amplifier of the oscilloscope, while the larger signal is applied via the phase shifter to the X amplifier. Since both signals are at the same (sub-carrier) frequency, the display is an ellipse.   By adjustment of the phase shifter the ellipse can be collapsed into a straight line, in which case the two signals on the oscilloscope X and Y plates are either in phase (straight line with positive slope) or in anti-phase (straight line with negative slope).

Thus, to set the phase between the I and Q sub-carrier signals, apply the composite video output, *see* Fig. 6.3(a), to the Y amplifier of an oscilloscope and a reference C.W. sub-carrier signal from the sub-carrier oscillator to the X amplifier via the calibrated phase shifter.   With a suitable signal (e.g. the green output of the colour bar generator) connected to the grid of V_4, adjust the phase shifter

PEAK TO PEAK
VALUES OF
SUBCARRIER
SINE WAVE

PHASE ANGLES
RELATIVE TO BURST

LUMINANCE
LEVELS

100% CARRIER ---1·0

BURST

30% CARRIER

0%
CARRIER

WHITE YELLOW CYAN GREEN MAGENTA RED BLUE BLACK
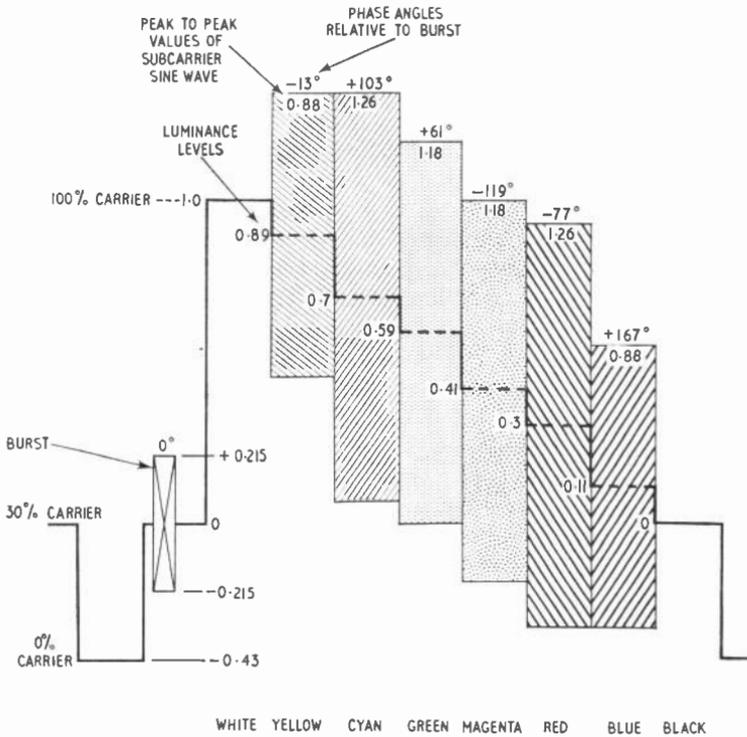
Fig. 6.5. Colour bar composite video waveform

until a straight line display is obtained.   Note the phase reading,
which gives the phase of the modulated I signal.   Switch the phase
shifter to a reading 90° lagging on this, connect the green output of
the colour bar generator to the grid of V_3 and adjust the trimming
capacitors connected to the driver transformer (Fig. 6.3(a)) so that a
straight line display is again observed on the oscilloscope.   If this
straight line has the opposite slope to that obtained previously,
reverse the connections between the driver transformer and either
the I or the Q modulator.   Re-check the phase of the modulated
I signal and repeat the whole process until the I signal leads the Q
signal by 90°.

Next, the I phase must be adjusted to lag the burst phase by 57°.
With the composite video applied to the Y amplifier of the oscil-
loscope and a reference signal applied via the calibrated phase
shifter to the X amplifier, adjust the calibrated phase shifter so that

a straight line is obtained for the burst signal. With 57° added
to the reading of the phase shifter and with the green output of the
colour bar generator connected to the grid of V_4, adjust the phase
shifter $P11$, which feeds the C.W. sub-carrier drive to the modulators
(Fig. 6.3(a)) until a straight line display is obtained for the green
bar period.

For a 100% saturated colour bar signal, the composite video
output should appear as shown in Fig. 6.5 and the phases of the
sub-carrier signals relative to the burst should be as shown.

It should be noted that, in the authors' opinion, the lengths of
delay line in the Y and I channels should be such that, when the
composite video is modulated onto an R.F. carrier and displayed
on an average monochrome receiver, the sub-carrier dot structure
due to a coloured portion of the picture is symmetrically placed
about that coloured portion. This may be checked, for example,
by applying the blue output of the colour bar generator to the grid
of V_{21} (Fig. 6.3(a)), and also to the grid of V_3, having first discon-
nected the V_3 grid from the V_1 anode circuit and having earthed the
grid of V_4. The resulting composite video, having been modulated
onto an R.F. carrier, should give a display on a monochrome
receiver which has equal widths of dot structure along each side
of each bar. Adjustment of the Y delay line length should be
made, if necessary, until this condition is obtained: Note that the
BBC delay specification differs from this (*see* Section 4.22).

Similarly, the I delay line may be checked by feeding the blue
bar output to V_{21} as before, and also to the input of the I delay
line with the grid of V_3 earthed. Again, adjustment of the I delay
line length should be made to give a symmetrical dot pattern.

The encoder shown in Fig. 6.3(a) can be readily converted to
equiband I and Q working. This can be done by replacing the
I delay line with a conventional lead, removing the I and Q low
pass filters in the anodes of V_4 and V_3 and replacing these by simple
resistive loads shunted by a capacitance. Thus, $1 \cdot 5k\Omega$ resistor loads
shunted by 150pF will give a 3dB response at 700 kc/s. The Y
delay line must be altered so that a symmetrical dot pattern is
observed on a monochrome receiver, as described above.

6.4. Vision modulator and carrier oscillator

The vision modulator and carrier oscillator unit accepts a com-
posite video signal and a suitable clamping waveform (such as
continuous line sync) and supplies a modulated carrier output
and also a C.W. vision carrier output from which the sound carrier
is derived, as described in the next section.

A suitable circuit arrangement is shown in Fig. 6.6. The composite video (sync negative going) is applied to the grid of the pentode V_4 and clamping is applied from the amplifier V_1 and the diodes V_2 and V_3, so that the V_4 grid video signal is instantaneously earthed during each line sync period.

The operation of the valves V_4 and V_5, which comprise the modulator stage, may be explained as follows:—

Suppose that, in the first instance, no signal is applied to the grid of V_6. Then part of the V_4 anode current will flow through
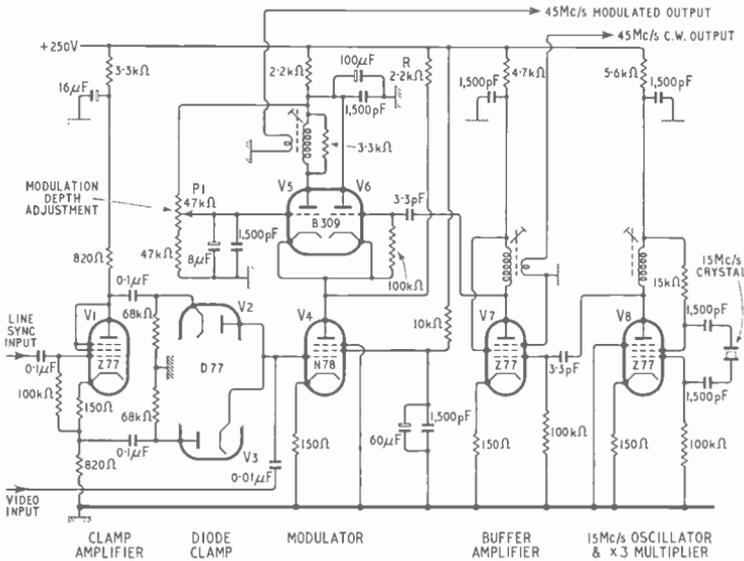
Fig. 6.6. Vision modulator and carrier oscillator

the resistor R, and the remainder will flow through V_5 and V_6. Since the grid of V_5 is held at a fixed D.C. potential, the cathodes of V_5 and V_6 will likewise be held at a fixed potential so that the current through the resistor R will be constant and independent of the V_4 grid drive. Hence, by a suitable choice of parameters, it is possible to make the current through R equal to the anode current of V_4 when the V_4 grid is earthed. Then zero V_4 grid voltage gives zero current through V_5 and V_6, the anode current of V_4 flowing wholly through the resistor R. Now when the V_4 grid is driven positive, the current through R remains the same and the increase in V_4 anode current therefore passes through V_5 and V_6. Thus,

the V_5 current bears a linear relation to the V_4 grid voltage down to the point where this is zero; that is, the operating point on the mutual characteristic of V_4 is lifted by the current bias through R.

When the carrier signal is injected into V_5 by V_6, it is modulated linearly by the V_4 signal and the modulated output is obtained from the V_5 anode load. The V_5 grid potential should be adjusted so
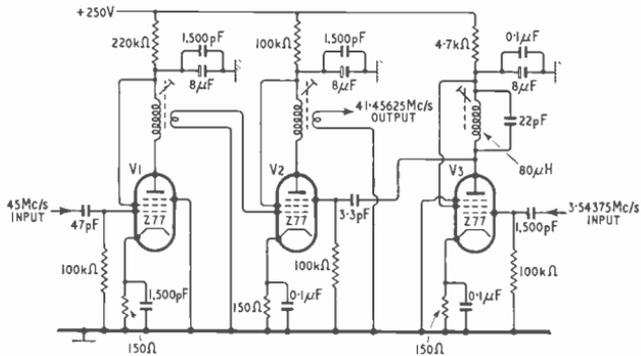
Fig. 6.7. Mixer for generating sound carrier

that the tips of the negative going sync signals correspond to zero output. This can be done by applying the modulated output to a simple crystal detector and observing the demodulated video waveform on an oscilloscope. If the oscilloscope input is now shorted out intermittently, a line will be observed corresponding to zero input and by adjusting the potentiometer $P1$ the negative tips of the line sync can be made coincident with this zero line.

The carrier frequency of 45 Mc/s is generated from a 15 Mc/s crystal oscillator which has a load tuned to 45 Mc/s. The buffer stage V_7 provides additional filtering to remove the 15 Mc/s component and it also provides a C.W. 45 Mc/s signal from which the sound carrier may be derived.

The anode loads of V_5 and V_7 consist of 45 Mc/s tuned circuits with loosely coupled secondaries. Each may be about 8 turns, with a 1 turn secondary, on a $\frac{3}{8}$ in. former with dust core tuning.

The modulated vision carrier should finally be passed through an asymmetric sideband filter and a phase correcting circuit, but for ordinary test purposes these refinements can be omitted.

6.5. Sound carrier mixer

Fig. 6.7 shows a suitable circuit for generating the sound carrier frequency from the 45 Mc/s vision carrier and the 3·54 Mc/s derived

by taking $\frac{4}{3}$ of the sub-carrier frequency. The 45 Mc/s input is fed to V_1, whose anode load is tuned to 45 Mc/s, and thence to the mixer V_2. Similarly the 3·54 Mc/s input is fed to V_3, whose anode load is tuned to 3·54 Mc/s, and thence to the suppressor grid of V_2. The V_2 anode load is tuned to 41·46 Mc/s.

The anode loads of V_1 and V_2 may be typically some 8 turns on a $\frac{3}{8}$ in. former with a dust core, the secondary windings being 1 turn.

The circuit (Fig. 6.7) can be used to derive a 45·04 Mc/s vision carrier from a 41·5 Mc/s sound carrier, by tuning the anode loads of V_1 and V_2 appropriately. This is the correct specification but as mentioned earlier it may prove more economical to derive the sound carrier since the vision carrier frequency is probably already available from a monochrome picture source.

6.6. Sound modulator

The sound modulator circuit can be similar to the vision modulator shown in Fig. 6.6, with minor modifications.

Thus (Fig. 6.6), the sound carrier is fed to the grid of V_6 and the audio input is fed via a potentiometer to the grid of V_4. The valves V_1, V_2 and V_3 must be omitted, of course, and also V_7 and
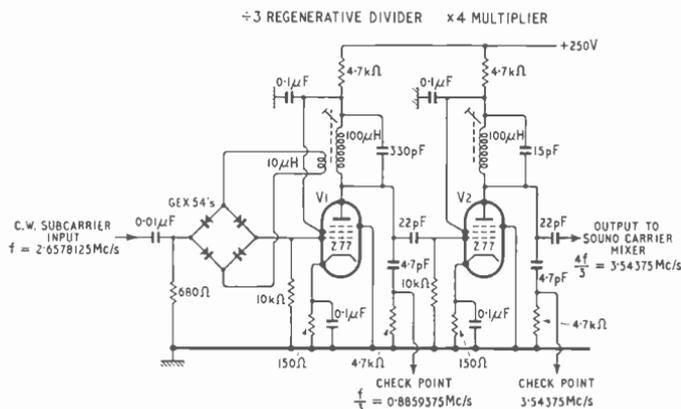
Fig. 6.8. 4/3 multiplier for deriving sound/vision beat from sub-carrier

V_8. If a sound carrier of 41·5 Mc/s and a vision carrier of 45·04 Mc/s are used, as described in Section 6.5, then V_7 and V_8 can be tuned to 41·5 Mc/s (with a crystal frequency of 13·833 Mc/s) in the sound modulator, in which case V_7 and V_8 can be omitted in the vision modulator. Since the modulating frequency is very much

lower in the sound modulator, economy of H.T. current can be effected by replacing V_4 by a Z77 and increasing R to 22kΩ. In this case an additional fixed resistor of 33kΩ must be included in the top leg of $P1$ so that the correct bias can be set for V_5.

6.7. Divider chain

A suitable circuit for deriving the sound/vision carrier beat from the sub-carrier frequency is shown in Fig. 6.8.

The C.W. sub-carrier frequency is divided by 3 in the regenerative divider V_1 and then multiplied by 4 in the multiplier V_2. The two coils forming the transformer anode load of V_1 may be wound on a $\frac{3}{8}$ in. former with a dust core, and sufficient coupling between primary and secondary is obtained if the two coils are wound adjacent to each other on the former. The primary is tuned to $\frac{1}{3}$ of the input
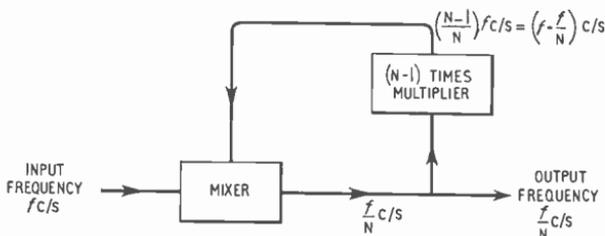
Fig. 6.9. Principle of regenerative divider

frequency. The anode load of V_2 is tuned to $\frac{4}{3}$ of the input frequency and therefore amplifies the fourth harmonic of the V_1 output.

The principle of the regenerative divider is shown in Fig. 6.9. The frequency f to be divided is fed into a mixer stage whose output load is tuned to $\dfrac{1}{n}$ of the input frequency, where n is the required division factor. Suppose that an output $\dfrac{f}{n}$ exists, then when it is multiplied by the $(n-1)$ times multiplier an output frequency $f - \dfrac{f}{n}$ results. When it is mixed with the input frequency f, an output $\dfrac{f}{n}$ is obtained. Hence, positive feedback at the frequency $\dfrac{f}{n}$ maintains continuous operation of the circuit.

In the circuit of Fig. 6.8, the multiplication and mixing processes are carried out in the diode ring in the grid of V_1. Thus, the anode
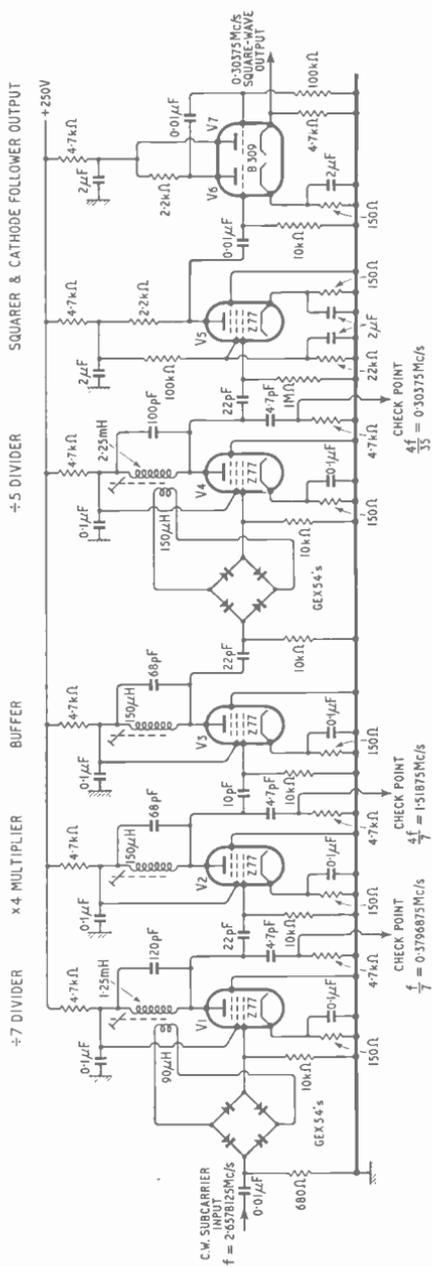
Fig. 6.10. 4/35 section of sub-carrier to twice line frequency divider chain

load of V_1 is tuned to $\frac{1}{3}$ of the input frequency and the non-linearity of the diode circuit produces a fourth harmonic of this which mixes with the input frequency to produce a $\frac{1}{3}$ frequency component which, being fed to the V_1 grid, maintains continuous operation.

It should be noted that the phase of the transformer connections of the V_1 load should be made so that no output is obtained when the input is removed. This ensures that the divider is operating regeneratively and not as a synchronized oscillator.

For setting up the correct tuning of the divider circuits, it is recommended that the input signal be displayed on one beam of a double beam oscilloscope, while the output signal from each check point in turn is displayed on the second beam. The correct frequency relationship can then easily be observed.

A suitable circuit for multiplying the sub-carrier frequency by $\frac{4}{35}$, as part of the division process to obtain a locked twice line frequency, is shown in Fig. 6.10. Here, regenerative dividers are used, the circuits being similar to those described above for the $\frac{4}{3}$ process except that the frequencies are different.

The output of the final divider of Fig. 6.10 is squared up to produce a square wave of the $\frac{4f}{35}$ frequency, so that an Eccles-Jordan type of divider may be used for the final division by 15. The regenerative type of divider tends to be unsuitable for low frequency operation because of the low frequency but high Q tuned circuits which would be required.

A suitable Eccles-Jordan circuit for carrying out the final division by 15 is shown in Fig. 6.11. The division is made in two steps of 3 and 5, respectively.

The principle of the Eccles-Jordan circuit may be described as follows:—

Considering the 3 to 1 divider as a simple example, suppose that V_1 is taking current. Then examination of the D.C. coupling from the anode of V_1 to the grids of V_2 and V_3 shows that V_2 and V_3 will be held cut off. Now suppose a positive pulse is fed to the cathode of V_1, then a positive pulse will appear at the V_1 anode and this will cause V_2 to conduct. As soon as V_2 conducts, it biases off V_1 and V_3 and equilibrium again occurs until another positive cathode pulse is applied, when V_3 will conduct. Thus, as the input pulses are applied, so each valve will conduct in turn and the output from any one valve will have a frequency equal to the input frequency divided by the number of valves in the circuit.

The output from the circuit shown in Fig. 6.11 is $\frac{4}{525}$ of the sub-carrier frequency, or 20·25 kc/s. This output can now be fed as the
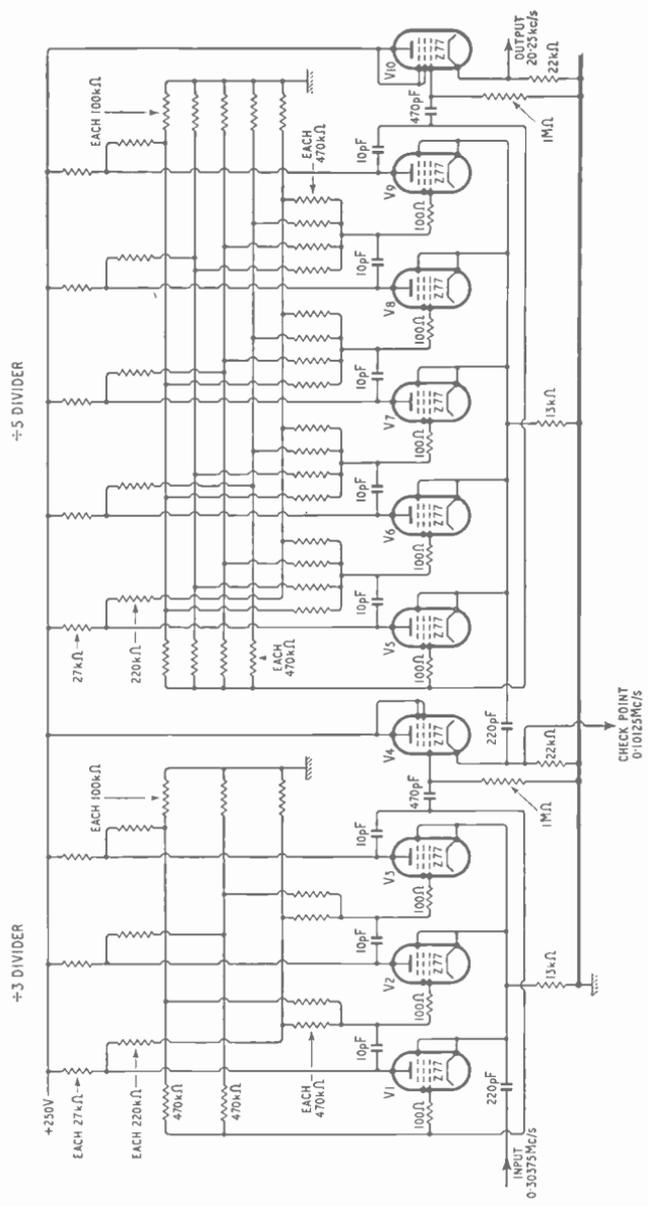
Fig. 6.11. Eccles-Jordan ÷ 15 circuit diagram. Note all screen grids are connected to H.T. line

" master oscillator " signal to a conventional mixed sync generator, as shown in Fig. 6.1.

6.8.   Colour bar generator

Suitable video sources of R, G and B inputs to the encoder may be obtained from either a colour camera or a colour flying spot scanner.   However, since both of these items are costly, it is convenient to have an inexpensive apparatus for artificially generating
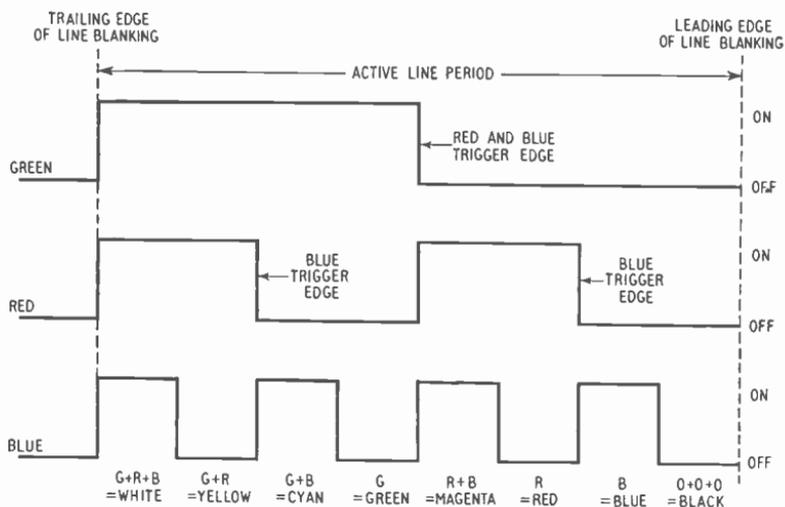
Fig. 6.12.   Colour bar generator waveforms

suitable video signals which simulate fully saturated primaries and their complementaries. Such a device is called a colour bar generator and one form of it provides an R, G and B video signal equivalent to a colour picture consisting of vertical colour stripes with full amplitude white on the left hand side, then full yellow, cyan, green, magenta, red, blue and black.   This arrangement gives the full amplitude colours in order of decreasing luminance from white on the left to black on the right.

The three waveforms required to produce this picture are shown in Fig. 6.12. All waveforms have equal peak-to-peak values and each waveform is a square wave, but the repetition frequency of the red signal is twice that of green and the frequency of the blue signal is twice that of red.   If the colour bars are to be of equal widths, the " on " and " off " periods of the B waveform should

each be $\frac{1}{8}$ of the active line period (i.e. the time of one line scan less the line blanking period), while the " on " and " off " periods of the red and green signals should be $\frac{1}{4}$ and $\frac{1}{2}$, respectively, of the active line period.

Thus, during the first $\frac{1}{8}$ of the active line period, R, G and B are each of full amplitude to give white. During the next $\frac{1}{8}$ of the active line period, R and G are still " on " but B is " off " so that yellow is obtained. Following this process through from left to right, it can be seen that white, yellow, cyan, green, magenta, red, blue and black will result.

A suitable circuit for generating the waveforms shown in Fig. 6.12 is drawn in Fig. 6.13.

Each colour channel consists of a monostable multivibrator triggered from the line blanking input, followed by a field blanking insertion stage, a phase inverter, and a cathode follower output. An amplitude adjustment is provided in each output so that the peak-to-peak values can be made equal.

The operation of the circuit may be described as follows:—

Positive going line blanking pulses are inverted by V_6 and differentiated so that positive " pips " are obtained corresponding to the trailing edges of the line blanking pulses. Now referring to the green channel, V_1 and V_2 are arranged as a monostable multivibrator in which, with no signal input, V_2 is normally conducting. The positive trailing edges of the line blanking cause V_1 to conduct and this renders V_2 non-conducting but since the V_1 anode is capacity coupled to V_2, the non-conducting periods of V_2 will last only for a time determined by the time constant of the RC circuit between the anode of V_1 and the grid of V_2, and this time can be adjusted by the potentiometer in the grid of V_2. Thus, positive going pulses are obtained at the V_2 anode whose lengths can be adjusted and whose leading edges coincide with the trailing edges of the line blanking. Note that corresponding negative going pulses are obtained at the V_2 cathode.

The V_2 output is applied to V_3 which is switched off during the field blanking period by the negative going field blanking pulses from the anode of V_7. Thus, a negative going and field blanked G waveform is obtained at the V_3 anode and this is inverted by V_4 and fed to the cathode follower output V_5.

The red channel operates in a similar manner to the green, but here the monostable multivibrator time constant is shorter. Also the negative going green waveform at the V_2 cathode is differentiated to produce positive " pips " at the times of the trailing edges of the green waveform. These trigger the red and blue multivibrators

Fig. 6.13. Colour bar generator circuit

which are also triggered by the trailing edges of the line blanking pulses.

Similarly, the blue channel multivibrator has a shorter time constant than the red. It is additionally triggered by the trailing edges of the red waveform, as well as by the trailing edges of the green and the line blanking waveforms.

The various triggering functions are illustrated in Fig. 6.12, and the amplitudes and on-off ratios of the three waveforms may be easily set up by observing each waveform in turn on an oscilloscope synchronized to the line frequency. It must be remembered that if equal bar widths are required, it is the active line period and not the whole line period, which must be divided by $\frac{1}{2}$ by the green waveform, and by $\frac{1}{4}$ and $\frac{1}{8}$ by the red and blue waveforms, respectively.

6.9. Summary

In this chapter the operations required for generating a complete R.F. signal encoded in accordance with the British adapted version of the N.T.S.C. system have been described in the form of block diagrams. In addition, actual circuits have been given for each block but, of course, any circuits which carry out the operations indicated would be suitable.

A colour bar generator has also been described, since this provides a useful and inexpensive source of R, G and B signals.

A considerable amount of the circuit complication in an N.T.S.C. transmitter is due to the requirements of frequency locking between the scanning frequencies, the sub-carrier frequency, and the R.F. vision/R.F. sound carrier beat. Again, the encoder unit which provides a composite video output from an R, G and B input is complicated by the differential bandwidth limitation of the I and Q signals. It should therefore be pointed out that an equipment which does not provide locking between the various frequencies, and which employs an equiband encoder, does prove adequate for most testing procedures, especially when a colour bar signal is available.

Thus, a simplified but extremely useful test set up would consist of an equiband encoder, a colour bar generator, and a conventional mixed sync generator all for providing a composite video signal, together with a vision modulator for providing the corresponding R.F. signal.

13

Introduction to Colour Receiver Design

7.1. Introduction

The greater part of a colour television receiver is basically similar in operation to a monochrome set. The radio frequency signal is selected, amplified and converted to the standard I.F. by a normal type turret or tuner unit. The I.F. stages are similar to monochrome practice; marked differences only appear at video frequencies.

The time bases and power supplies are like their monochrome counterparts but in general are required to deliver higher currents. For three-gun types of colour tubes convergence circuits are needed, while single-gun tubes need special gating circuits for selecting the appropriate modulation at the right time.

The greatest divergences from monochrome usage occur in the circuits for demodulating the colour difference information carried by the sub-carrier. Since the sub-carrier has two distinct kinds of modulation, at least two demodulators are required. Each demodulator must pick out the modulation which is occurring at one specific phase of the sub-carrier, say $\theta°$, and reject the modulation which the sub-carrier has at phase $\theta° \pm 90°$. Demodulators which can do this are called *synchronous demodulators* as they must be synchronized in phase with the sub-carrier.

Synchronous demodulators have to be supplied with a continuous reference sine wave signal which they can use to judge the phase of the incoming sub-carrier signal. This reference sine wave must have the same frequency and phase as the reference signal used at the transmitter when the signal is coded. The circuits which derive this receiver reference signal from the colour burst are called *reference generators*. The chrominance demodulators and the reference generators are narrow band devices and service engineers who are familiar with black and white receivers will experience little difficulty in handling these new circuits.

The type of colour cathode-ray tube which the receiver uses has a decisive effect on the exact method of signal decoding. Receivers

which use three-gun tubes or which combine optically the pictures from three separate tubes, require simultaneously, at the electron guns, voltages proportional to the three tristimulus values R, G and B.   Other types of display require the information in a different form which can always be derived from the tristimulus values although it is usually more economical to tailor the receiver to suit the display tube it is using.

Most receivers use three-gun shadow mask tubes and this is the type of receiver which will be discussed here in greater detail.

7.2. Block diagram of typical receiver

A general schematic of the receiver sections needed to perform the functions described in Section 7.1 is shown in Fig. 7.1.

The video signal at the detector consists of the normal luminance waveform with mixed sync, together with the chrominance sub-carrier, its sidebands and the colour burst signal.   It is possible to amplify these signals together but certain difficulties, to be discussed later, can be avoided if the chrominance and luminance signals are separated out at the detector.

The luminance signal is delayed with respect to the chrominance signals by a delay line which is usually a piece of coaxial cable specially made to have a low transmission velocity.   The inner of the cable is wound in a spiral and the polythene medium is loaded with ferrite dust.   This delay is needed because the chrominance signals go through narrower band circuits than the luminance signal and have different time delays.   The centres of chrominance transitions thus occur later than the centre of the corresponding luminance transition unless the luminance signal is delayed to compensate (*see* Fig. 4.13 and Section 4.22).   The luminance amplifier corresponds to the normal video amplifier, except that the final output signal required by the shadow mask tube is some 80–130V peak-to-peak video signal, depending on the screen potentials. • The chrominance sub-carrier is usually rejected by this amplifier before the signal reaches the tube.

The chrominance signal is extracted from the composite encoded video by giving the chrominance amplifier a suitable band-pass response which accepts the sub-carrier and its sidebands but rejects the unwanted lower-frequency luminance signals (*see* Fig. 7.2).   Of necessity, the higher-frequency luminance signals are also accepted by the chrominance amplifier but there is no practical way of discarding these luminance components, despite the frequency inter-leaving (*see* Section 4.35), whilst still retaining the chrominance components.   The undesired luminance components cause some
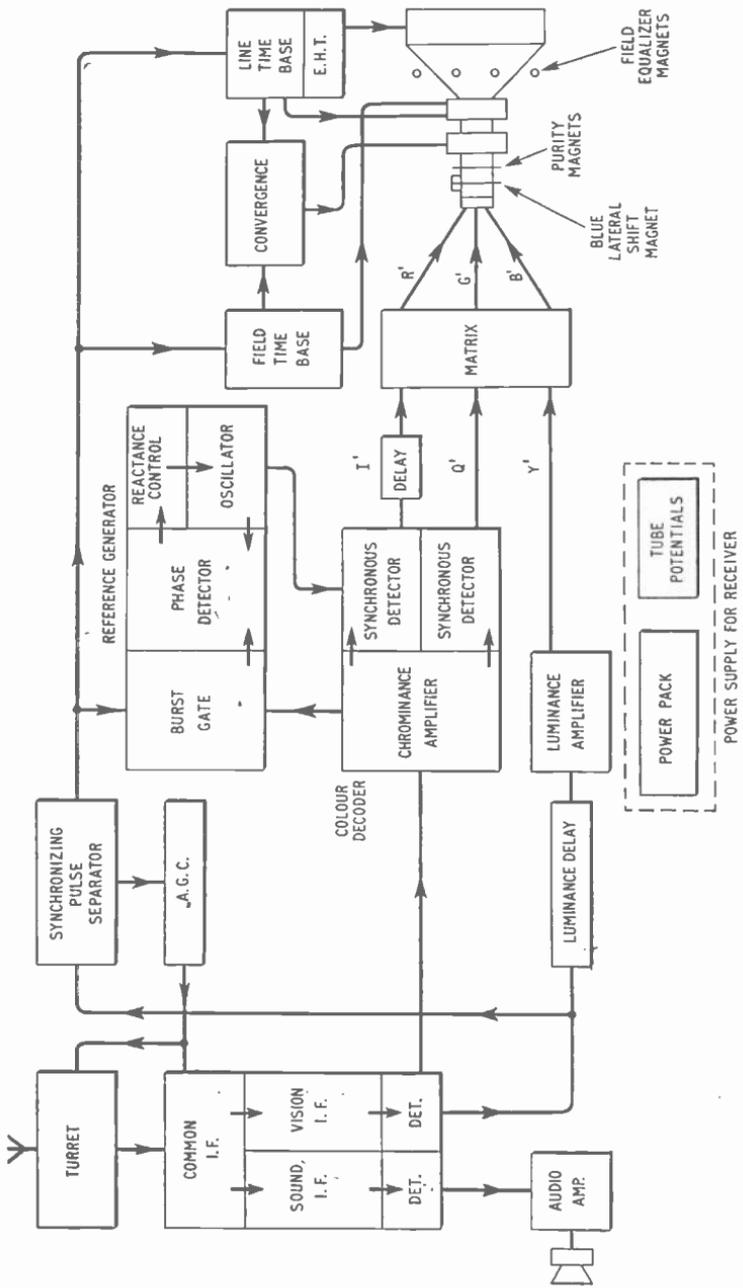
Fig. 7.1. Basic functions for a shadow mask tube receiver

spurious colour effects which are not very noticeable on colour pictures.

The complete chrominance signal is passed to each of the synchronous detectors.   At least two synchronous detectors are required and in some arrangements it is advantageous to use three.   Several types of synchronous detectors are described in Chapter 9.   One type may be thought of as a gating valve which is opened for a short
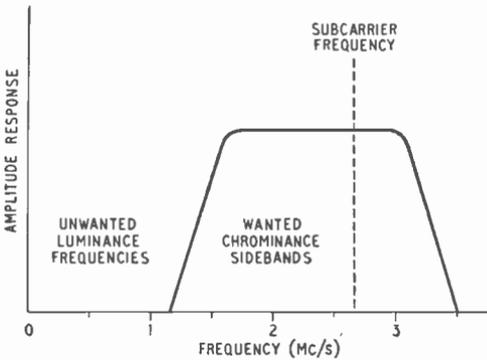
Fig. 7.2.   Chrominance band-pass response

time at the same phase on every cycle of the reference voltage and accepts the magnitude of the chrominance signal at these instants only.

The outputs of the synchronous detectors depend on the phase at which they are demodulating, but are ideally I' from one detector and Q' from the other.   The voltage level at which the synchronous demodulators operate may vary from a few volts to the hundred or so required for application to the display tube.

In general the low frequency demodulated chrominance signals are time matched if they have different pass-bands from each other and are then passed to the matrix circuits where, together with the Y' luminance signal, they are recombined to form R', G' and B'.

The matrix circuits, in effect, solve the three equations

$$R' = Y' + 0{\cdot}96I' + 0{\cdot}62Q'$$
$$G' = Y' - 0{\cdot}27I' - 0{\cdot}65Q'$$
$$B' = Y' - 1{\cdot}11I' + 1{\cdot}70Q'$$

and derive their name from the matrix notation used in algebra as a form of shorthand for writing and solving such equations.

$$
\begin{bmatrix} R' \\ G' \\ B' \end{bmatrix} = \begin{bmatrix} 1 & 0{\cdot}96 & 0{\cdot}62 \\ 1 & -0{\cdot}27 & -0{\cdot}65 \\ 1 & -1{\cdot}11 & 1{\cdot}70 \end{bmatrix} , \begin{bmatrix} Y' \\ I' \\ Q' \end{bmatrix}
$$

The matrix circuits are commonly a few resistors connected so that appropriate fractions of the three incoming signals are added together. Valve converters are normally used to get the right sign of signal for the matrix equations. The three outputs from the matrix are used to modulate the cathode-ray tube, after further amplification if necessary.

The reference generator which provides the reference frequency to the synchronous detectors may take several forms but the basic functions shown in Fig. 7.1 are representative. The stable oscillator is tuned to the sub-carrier frequency but requires to be locked very accurately in phase as well as in frequency with the transmitter's reference sine wave. The colour burst signal is gated out from the composite chrominance signal and the phase of the reference oscillator is compared with the phase of the colour burst signal. Any difference in frequency or phase between the burst and the reference oscillator produces a voltage which is used to change the frequency of the oscillator slightly until it pulls into phase lock.

The line time base provides about $1\frac{1}{2}$ A of scanning current to the deflection yoke and also generates the 22–25 kV required for the final anode of the shadow mask tube. The E.H.T. must be stabilized up to currents of 1 mA or so, since the convergence varies with the E.H.T. voltage. The electron guns of the shadow mask tube are electrostatically focused and the line time base also provides the 5kV focus voltage at negligible current. The field time base generates some 500 mA of 50 c/s or 60 c/s sawtooth current for the field deflection coils.

Voltage waveforms from both time bases are processed in the convergence circuits to provide parabolic current waveforms through the dynamic convergence coils of the shadow mask tube.

All these circuit functions will be described in more detail in the next four chapters. In the rest of this chapter we shall discuss a few general topics which arise in the design of colour receivers.

7.3. Asynchronous operation

For normal monochrome transmission in the United Kingdom the field frequency is kept in synchronism with the mains frequency. As the national grid frequency varies so the television field frequency is varied. By this means any effects on the television picture

due to mains hum, such as brightness modulation or displacement of the scanning lines, are kept stationary with respect to the picture, or at least very slow moving.

The hum effects are then much less noticeable than if they are moving up or down the picture. The most objectionable rate of movement is about 2 seconds for a complete sweep from top to bottom of the picture.

Such mains-locking enables appreciable economies in smoothing to be made in every receiver. However, in colour transmission it is desirable to keep the sub-carrier frequency constant to ease the design of the reference generator. Since

$$f_{sc} = \frac{525}{2} f_L = \frac{405}{2} \times \frac{525}{2} f_F$$

a constant sub-carrier frequency implies a constant field frequency and it is not then practical to lock to the mains frequency.

Various suggestions have been made for overcoming this difficulty, but the present practice is to operate asynchronously (*see* Section 15.2).

7.4. Crosstalk between chrominance signals

Since the I' signal is transmitted in a vestigial sideband fashion, it produces a spurious signal 90° out of phase with itself. A simple outline of the effect is given in Fig. 7.3.

An unmodulated carrier can be represented in amplitude and phase by a phasor, such as OC shown in Fig. 7.3 (a). If the carrier is modulated then its amplitude will vary so that some time later than Fig. 7.3 (a) it is represented by the phasor OC' in Fig. 7.3 (b). A convenient way of representing such a modulated carrier is to consider it to be composed of the original unmodulated carrier OC, together with its sidebands. If the modulation is a simple sine wave, the sidebands can be represented by a pair of additional phasors $C'U$ and $C'L$, the upper and lower sidebands respectively, rotating in opposite directions symmetrically around the end of the carrier OC. These sideband phasors have an amplitude dependent on the depth of modulation and rotate with an angular frequency determined by the frequency of the modulating sine wave. The resultant of the carrier and sidebands is the phasor OD shown in Fig. 7.3 (e), since the sideband phasors can be resolved into components along the carrier OC and perpendicular to OC, as in Fig. 7.3 (d). The two components perpendicular to the carrier cancel out for all positions of the sideband phasors as they rotate about C, leaving only a
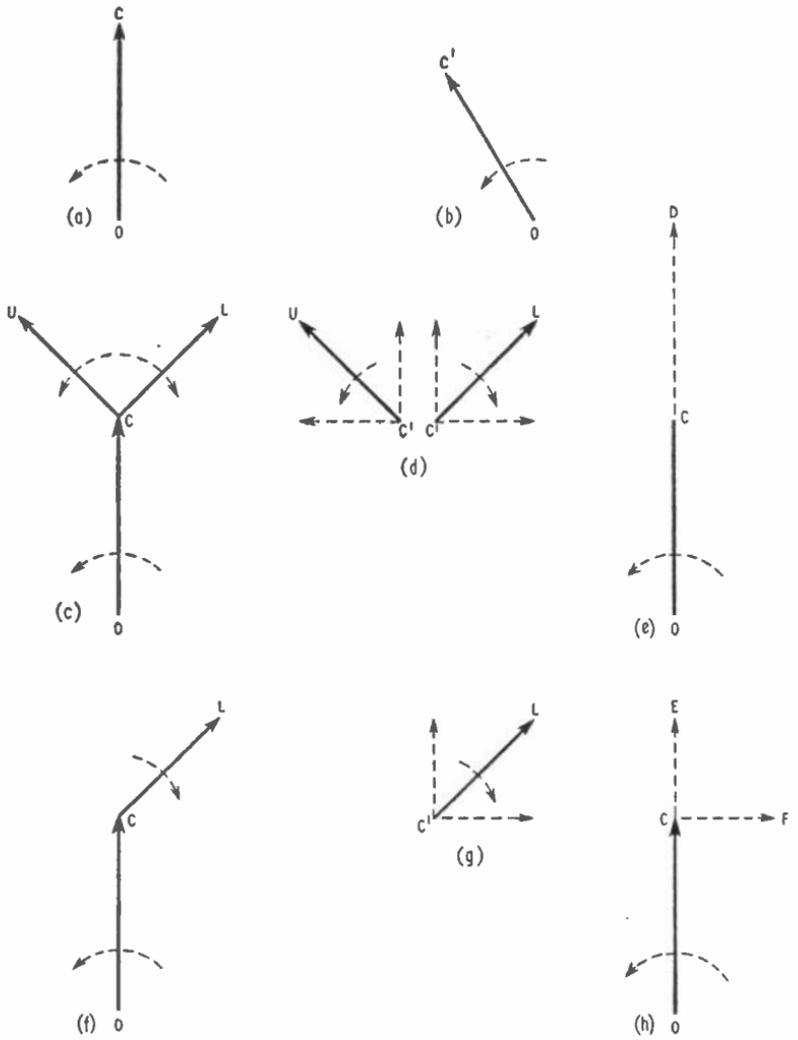
Fig. 7.3. Quadrature signal due to vestigial sideband transmission

component CD which adds to or subtracts from the main carrier phasor.

However, if one sideband is missing, as happens with the higher video frequencies in vestigial or asymmetric sideband transmissions, the two components perpendicular to the carrier no longer cancel, *see* Fig. 7.3 (g), and the resultant, *see* Fig. 7.3 (h), consists of the carrier with only half the amplitude of the original wanted modulation, CE, together with an unwanted component CF in quadrature with it.

Since there is not room between the sub-carrier frequency and the end of the video band to send the I' signal double sideband, the
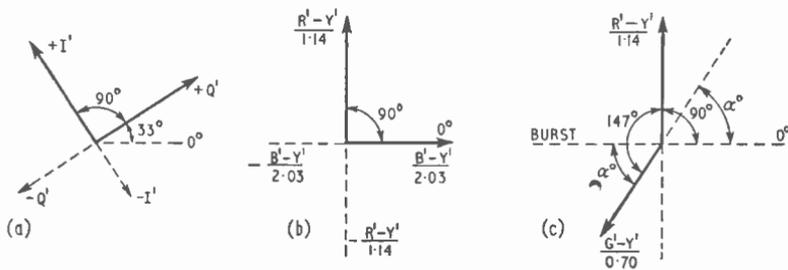
Fig. 7.4.   Demodulator axes

higher I' frequencies, from 0·34 Mc/s to 1·0 Mc/s, are sent single sideband.   These I' modulation frequencies therefore give rise to a quadrature signal, equal in amplitude to the remaining wanted modulation but 90° out of phase; i.e., in the Q' phase.

The Q' signal is double sideband throughout its modulation range of 0 to 0·34 Mc/s and does not produce a spurious quadrature signal in the I' channel.   As the I' signal is only single sideband from 0·34 Mc/s to 1 Mc/s, the spurious quadrature signal in the Q' channel falls outside the Q' acceptance bandwidth and is therefore rejected.   In the range 0 to 0·34 Mc/s the I' signal is double sideband and does not produce any crosstalk into the Q' channel.

If crosstalk from I' to Q' is to be avoided the Q' channel must not accept frequencies outside its specified bandwidth of $\pm$ 0·34 Mc/s.

7.5.   Types of chrominance channel response

If the full chrominance bandwidth of the system is to be used, one colour detector must demodulate along the I' axis (*see* Fig. 7.4), at a phase of 123°.   The chrominance frequency response for the I' channel should then be 1 Mc/s.   The sub-carrier modulation is

vestigial sideband (Fig. 7.2) and amplitude correction should be applied for the missing sidebands either before or after demodulation. Thus in Fig. 7.5 (a), if the sub-carrier chrominance response is flat, then the chrominance video response must incorporate a 6dB amplitude boost to compensate for the missing sidebands. Alternatively the response can be shaped as in Fig. 7.5 (b) so that the
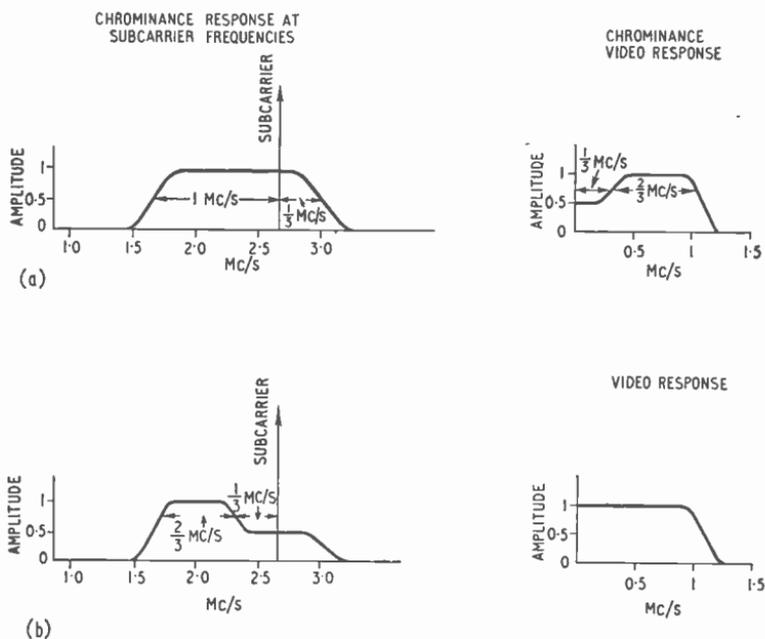
Fig. 7.5. *Amplitude correction for V′ vestigial sideband at (a) video and (b) sub-carrier*

sub-carrier is 6dB down relative to the peak of the chrominance response. In this case the following video stages can have a flat amplitude response. The important thing is the resultant overall response up to the display, and a boost in one part of the channel can be compensated for in another stage, within limits. The asymmetric sideband correction is often omitted and the sharper luminance waveform is relied upon to mask the resulting chrominance smear.

For accurate waveform reproduction the overall phase response is required to be as linear as possible. The exact shape of the phase response is not critical as far as the accurate reproduction of hue in large areas of the picture is concerned.

Receivers using the full I' bandwidth of 1 Mc/s are called *wideband chrominance receivers*. The majority of colour receivers do not use the full I' bandwidth but have both chrominance channels double sideband with the bandwidth of the Q' channel, $\frac{1}{2}$ Mc/s. Such receivers are called *narrowband chrominance receivers* or *equiband receivers*. Although narrowband receivers have relatively poor colour definition, they have certain design advantages which tend to offset the loss in chrominance resolution. One advantage is that the phase angles at which the colour detectors demodulate can be chosen without regard to any unwanted interaction between the chrominance signals and it is not essential to use I', Q' axes. If the decoding axes are not I' and Q' it is still possible to use wider bandwidths for both channels than the Q' bandwidth, but at the cost of some crosstalk between the two chrominance signals. This is called *wideband equiband operation*.

7.6. Choice of demodulation axes

In equiband receivers where both chrominance channels are only $\pm 0{\cdot}34$ Mc/s in width, there is no possibility of the I' signal crosstalking into the Q' channel due to vestigial sideband distortion. Synchronous demodulation can therefore take place along any convenient axes. An obvious choice is $(R' - Y')$ and $(B' - Y')$ axes, Fig. 7.4. These are strictly the $\dfrac{(R' - Y')}{1{\cdot}14}$ and $\dfrac{(B' - Y')}{2{\cdot}03}$ axes, and the gain factors must be remembered in designing the matrix circuits. Having obtained the $(R' - Y')$ and $(B' - Y')$ signals $(G' - Y')$ is easily obtained, since

$$Y' = 0{\cdot}30R' + 0{\cdot}59G' + 0{\cdot}11B'$$

and

$$0 = 0{\cdot}30(R' - Y') + 0{\cdot}59(G' - Y') + 0{\cdot}11(B' - Y')$$

$$\therefore \; (G' - Y') = -\frac{0{\cdot}30}{0{\cdot}59}(R' - Y') - \frac{0{\cdot}11}{0{\cdot}59}(B' - Y')$$

The simple addition of the Y' signal to the three colour difference signals $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$ then gives the three tristimulus values R', G' and B'. The display tube itself can be used to carry out this addition by modulating the three cathodes with the negative luminance signal and the three grids with the colour difference signals. It is preferable not to modulate the cathodes with the colour difference signals (*see* Section 11.8).

The $(R' - Y')$ and $(G' - Y')$ axes can also be used for demodulation and this has the advantage that the $(G' - Y')$ signal is received

at a relatively higher level than the $(B' - Y')$ signal. The angle and gain factor of the $(G' - Y')$ axis, or indeed of any other demodulation axis, is easily found.

The signal along the $(G' - Y')$ axis is the resultant of the projections on this axis of the I' and Q' signals, or, which is the same thing for narrowband signals, the projections of the $(R' - Y')$ and $(B' - Y')$ signals. Let the $(G' - Y')$ axis make an angle α with the burst phase, Fig. 7.4, and let the magnitude of the $(G' - Y')$ signal along this axis be $\dfrac{(G' - Y')}{C}$

then

$$\frac{(G' - Y')}{C} = - \frac{(R' - Y')}{1 \cdot 14} \sin \alpha - \frac{(B' - Y')}{2 \cdot 03} \cos \alpha$$

or

$$(G' - Y') = - \frac{C \sin \alpha}{1 \cdot 14} (R' - Y') - \frac{C \cos \alpha}{2 \cdot 03} (B' - Y')$$

but also

$$(G' - Y') = - \frac{0 \cdot 30}{0 \cdot 59} (R' - Y') - \frac{0 \cdot 11}{0 \cdot 59} (B' - Y')$$

$\therefore$ $$C \sin \alpha = \frac{0 \cdot 30. \quad 1 \cdot 14}{0 \cdot 59}$$

and

$$C \cos \alpha = \frac{0 \cdot 11. \quad 2 \cdot 03}{0 \cdot 59}$$

so that $C = 0 \cdot 70$ and $\alpha = 57°$.

For stability it is sometimes desirable to use three demodulators, and various axes have been suggested for such arrangements and will be discussed further in Chapter 9.

7.7. Cross-colour and parc

As well as the vestigial sidebands of I' interfering with the Q' channel, various other forms of crosstalk can occur between the components of the multiplexed colour television signal. The subcarrier itself appears in the luminance channel and causes an unwanted beat pattern on the picture (Section 15.4). The luminance signal components in the high frequency part of the video band are accepted by the chrominance channel and demodulated as if they were colour signals, when they give rise to spurious colour

effects which are called *cross-colour*. These effects are only notice-
able in the picture areas where there is fine detail corresponding to
video frequencies close to the sub-carrier frequency. The $2\frac{1}{2}$ Mc/s
block of resolution bars in Test Card C normally gives rise to such
cross-colour effects. A luminance frequency of 2·5 Mc/s is treated
by the chrominance demodulators as a chrominance frequency of

$$2·66 - 2·5 \text{ Mc/s} = 0·11 \text{ Mc/s}$$

so that a high frequency luminance signal causes low frequency
chrominance crosstalk. Cross-colour is dealt with further in
Section 16.10.

Similarly, any random noise in the high frequency part of the
luminance spectrum, which is normally relatively unimportant
because of its fine structure, is demodulated by the synchronous
detectors into low frequency noise in the chrominance channels,
where it is more obvious because of its coarser structure despite
the protection afforded by the approximately constant luminance
operation of the system. This heterodyning of random noise
frequencies produces annoying low frequency colour effects which
the Americans have called *parc*.

To remove such spurious colour effects from the monochrome
picture which the colour receiver should produce during mono-
chrome transmissions, a circuit called the *colour killer* reduces the
gain of the chrominance channel to negligible proportions during
black and white programmes. The colour killer does this auto-
matically, depending on whether or not there is a colour burst
present in the back porch of the line sync pulse.

7.8. Sound-sub-carrier beat and chrominance buzz

Beat effects between the vision and sound carriers, resulting in a
3·5 Mc/s pattern on the picture, are familiar enough in monochrome
reception. In colour reception there is the further possibility of the
chrominance sub-carrier beating with the sound carrier to produce
a black and white pattern at the difference frequency of

$$3·5 - 2·66 = 0·84 \text{ Mc/s}$$

which is a much coarser beat pattern than the vision carrier-sound
carrier beat itself.

It is also possible for enough of the 3·5 Mc/s beat frequency to
reach the synchronous detectors, where it is demodulated to produce
a coloured 840 kc/s beat pattern. Both these 840 kc/s patterns must
be guarded against in receiver design.

The vision signal may produce a buzz in the sound channel if it
is not adequately rejected, due to the low frequency picture and

field modulation carried by the vision signal. The chrominance sub-carrier is also interrupted at field frequency and is much closer to the sound carrier. Receivers which have barely adequate vision rejection in the sound channel may reproduce the familiar vision buzz on colour transmission but not on monochrome transmission, due to insufficient chrominance sub-carrier rejection in the sound channel.

7.9. Colour controls

Colour television receivers have the same customer controls as most monochrome receivers, with one or two additional knobs. Channel selection and tuning, contrast, brightness and volume (sound) are as on monochrome receivers. In addition there is usually a *colour control* knob marked " saturation ". This controls the gain of the chrominance channel and thus varies the purity of the reproduced colours, the colours becoming more pure as the gain is increased, whilst decreasing the chrominance gain eventually results in a black and white picture.

A second colour control is also incorporated in receivers but may only be accessible to the service engineer. The *hue control* knob varies the phase of the reference generator output in the receiver and thus varies the colour which a given phase of sub-carrier signal will produce. For example, rotation of the hue control may change the appearance of flesh tone from a greenish-yellow through a normal facial tone to a purplish colour.

Other controls are available for the service engineer who instals the colour receiver. For three-gun type tubes there are usually three *bias* or *background* controls, one for each electron gun, with an overriding master brightness control for the viewer. There may also be three *screen grid* controls to vary the slope of the characteristic of each gun. Occasionally there are also separate *gain* controls in each R, G and B channel. The level at which the colour killer decides that there is no burst present in the signal may be adjustable, as may the gate which is picking out the burst from the composite colour signal.

Different types of display tube also call for their own setting-up procedure, but these matters are more fully dealt with in Chapters 11 and 13, for the case of the shadow mask tube.

7.10. Power supplies and heat dissipation

The power supplies may be conventional in colour receivers, but are required to deliver roughly twice the power output of their black and white counterparts. At the same time, three-gun colour tubes

are very susceptible to unwanted stray magnetic fields.   Trans-
formers and chokes should be placed as far away from the tube as
possible, at the back of the receiver.   The fields from the trans-
formers should be kept from the tube by means of magnetic shields,
copper shorted-turns around the transformer, opposing the fields
from two transformers, and such-like techniques.

With such increased power dissipation in a 21 in. tube cabinet, the
problem of cooling is greater than in a monochrome receiver.   The
components controlling colour balance and convergence are parti-
cularly sensitive parts of the receiver and due attention must be paid
to keeping the effects of heat away from these circuits.

7.11.   Summary

As well as its normal monochrome functions of reproducing the
sound, scanning the display tube and amplifying and detecting the
luminance or Y' signal, a colour receiver has three new basic func-
tions to perform.

It must generate a continuous reference sine wave locked in
frequency and phase to the transmitted colour burst.   It must use
this reference voltage to demodulate at least two chrominance signals
from the sub-carrier signal and it must then matrix the two chromi-
nance signals with the luminance signal to produce the final red,
green and blue modulating voltages which the display tube needs.

Certain types of display tube need additional circuits for their
operation but these vary from tube to tube.

The multiplex technique used in colour transmission admits the
possibility of various forms of crosstalk between the five distinct
signals which the composite television signal contains, namely: the
luminance, the sound, the synchronizing signals and the two chromi-
nance signals.

To avoid crosstalk from I' to Q', chrominance demodulation
must be either along the I' and Q' axes or both chrominance channels
must be restricted to the narrow Q' bandwidth.

Colour Receiver Amplifiers

8.1. Introduction

The incoming R.F. signal is received and amplified by the colour receiver very much as in a monochrome receiver. The vision carrier and its colour sub-carrier are treated as one signal throughout the R.F. and I.F. stages, whilst the sound carrier is separated out in the I.F. amplifier and handled exactly as in black and white television practice. The chrominance sub-carrier signal is filtered from the vision signal either just before, at or soon after the luminance I.F. detector, and is further amplified before being synchronously detected to provide the low frequency colour-difference information. The luminance amplifier is similar to the video amplifier in conventional receivers.

8.2. R.F. tuners

Any well designed monochrome tuner unit will function satisfactorily on colour transmissions. All the extra colour information is carried in the same bandwidth as the normal monochrome transmission and no extra circuits or considerations are involved. However, whereas liberties can often be taken with the amplitude and phase response at the higher video frequencies without producing very noticeable effects on monochrome pictures, such distortions in colour reception will affect the colour sub-carrier and its chrominance signals. After the sub-carrier has been demodulated these distortions become low frequency chrominance distortions rather than high frequency luminance distortions. The lower frequency makes the chrominance waveform distortion more obvious on the reproduced picture but this is countered by the masking effect of the superimposed luminance waveform. Since the relative amplitude of the luminance carrier to the sub-carrier is conveying colour information about the purity of the transmitted colour, the amplitude response should avoid peaks around the sub-carrier frequency. Most monochrome receiver turrets are sufficiently flat and broad in their responses to be quite acceptable for colour.

The noise factor requirements of the tuner are approximately the same as for monochrome reception since the system operates

at nominally constant luminance. A good noise factor is still required as even in noisy fringe areas most observers feel that the chrominance signal adds something worthwhile to the television picture, providing they are viewing from more than four times the picture height.

8.3. I.F. stages

The design of I.F. stages for colour receivers tends to be rather more critical than for black and white receivers, although the same principles apply.

The basic requirements are that the I.F. amplifier and video stages together shall have a flat amplitude response over the pass-band of 0 to 3 Mc/s, with a linear phase response. Since the transmission is vestigial sideband this implies that either the I.F. or video stages must attenuate the low frequencies, as in monochrome receivers. The vision I.F. stages must reject the sound signal, and the sound I.F. stages must reject both the vision and chrominance signals. Protection against the signals on adjacent channels must also be provided. There must be no cross-modulation between the vision, sound and chrominance signals, and the input-output characteristics must be linear. With the types of " minimum phase " circuits which can economically be used in mass-produced receivers, compromises are necessary between these requirements.

The necessity of providing adequate rejection at sound and adjacent sound frequencies makes it difficult to keep the vision phase response linear at the edges of the pass-band. The situation at the vision carrier position is the same as in monochrome reception but it is more critical at the sub-carrier end of the band. The transmitted waveform is pre-distorted to correct for the phase response of the average receiver around the sub-carrier and this eases the receiver design.

If the overall amplitude response is flat and the overall phase response is linear, then two advantages accrue.

Firstly, there is no distortion of the Y', I' or Q' waveforms, and hence the colour transitions in the picture are not distorted. Secondly, there is no crosstalk between I' and Q' signals which could cause an I' signal to give rise to a spurious signal in the Q' channel and produce the wrong colour at the edges of coloured picture areas.

In general, phase distortion at the sub-carrier frequency will not cause incorrect hue reproduction in large areas of the picture. It is important to keep clear the distinction between phase angles in the phase-frequency characteristic of the receiver, and the phase

14

angle between the reference burst and the chrominance signal carried by the sub-carrier. It is the latter which determines the hue of the reproduced colour. The colour in large areas is not affected by phase distortion in the receiver response at 2·66 Mc/s, i.e. by the time delay between the sub-carrier and the vision carrier. For large areas of the same colour the chrominance signal is at the same frequency as the reference burst and any phase distortion in the receiver applies to both equally and does not affect the difference in phase between them, providing both burst and chrominance traverse the same paths through the receiver circuits. For example, if an all-red picture is being transmitted a colour receiver will display red no matter how distorted the phase response of the signal channel, because both the reference burst and the chrominance operate at the same point on the receiver's phase-frequency characteristic. Even when the burst and signal follow different paths, the probability is that the phase difference between them will be constant and can be taken up on the hue control.

However, phase distortion in the receiver can give rise to $I'Q'$ crosstalk. For simplicity, consider a chrominance sub-carrier E which has the phase of the I' component.

$$E = E_0 \cos (\omega t + 33°)$$
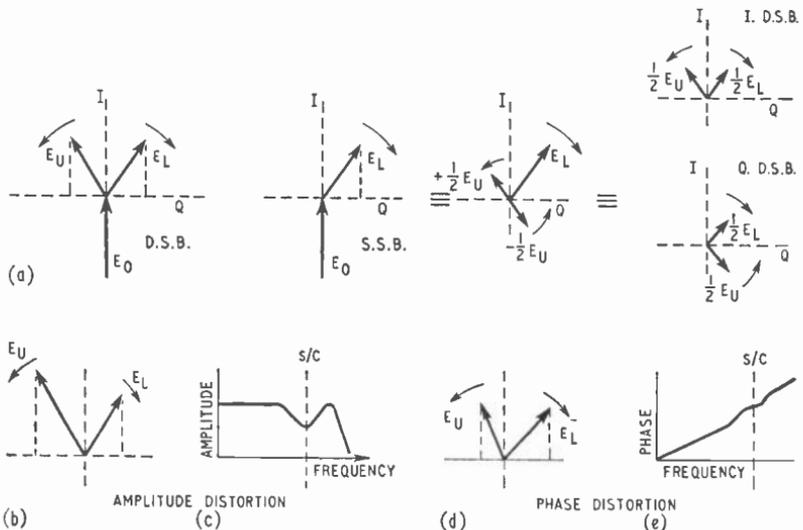
Fig. 8.1. $I'Q'$ crosstalk due to amplitude or phase distortion. The chrominance sub-carrier vector is $E = E_0 \cos (\omega t + 33°)$ with an upper sideband vector $E_0 \cos [(\omega + \omega_r)t + 33°]$ and a lower sideband vector $E_L \cos [(\omega - \omega_r)t + 33°]$

and upper and lower sidebands

$$E_U \cos \ [(\omega + \omega_r)t + 33°]$$

and

$$E_L \cos \ [(\omega - \omega_r)t + 33°]$$

representing amplitude modulation of E at a modulation frequency of ω_r as in Fig. 8.1. It is clear that both the sideband vectors will
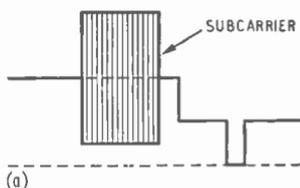
Fig. 8.2. Enhancement of the luminance amplitude due to over-amplification of the sub-carrier
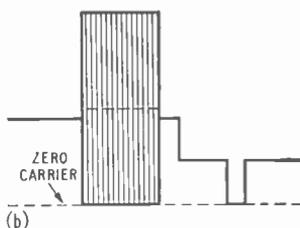
have components along the Q' axis but that these will always be equal and opposite so that there is no crosstalk into the Q' channel, over the frequency range which the Q' channel accepts and for which I' is double sideband. Any double sideband signal which does not have the sidebands equal and symmetrically disposed about the demodulation axis will give rise to crosstalk; the crosstalk waveform will differ in shape from the original waveform since all the components suffer a 90° phase change irrespective of their frequency. The sufficient condition for no crosstalk is that the receiver amplitude response characteristic should be symmetrical about the sub-carrier frequency while the phase characteristic must be skew-symmetrical. For the I' and Q' waveforms to be undistorted it is also necessary, of course, for the amplitude response to be flat and the phase response linear.

The preceding remarks apply to the overall response of the I.F. and chrominance channel up to the synchronous detectors. It is possible for the I.F. response to depart from the ideal characteristics

and to compensate for this in the chrominance amplifier.   There is
some advantage in reducing the level of sub-carrier before the lumin-
ance detector and incidentally rejecting the sound carrier more
effectively, whilst correcting the response later in the receiver.
Reducing the sub-carrier level effectively reduces the depth of
modulation and hence the single sideband distortion.   It is important
that the sub-carrier amplitude should not be emphasized in the I.F.
response.   An increase in sub-carrier level may well carry the video
waveform on the saturated low luminance colours, such as red and
blue, down to the zero carrier level as shown in Fig. 8.2, with con-
sequent rectification of the sub-carrier and enhancement of the
luminance level.

The cross-modulation requirements are similar to those for
monochrome receivers but can give rise to further intermodulation
effects such as sub-carrier buzz on the sound channel.   Gain control
is similar to black and white practice but it is important that it
shall not change the response shape and upset the carrier to sub-
carrier ratio (*see also* Section 8.6).

8.4. Sound rejection

The sound rejection in the vision I.F. must be sufficient for the
840 kc/s difference frequency beat between the sub-carrier and sound
carrier, which can be produced at the luminance detector, to be
imperceptible on the picture.   The worst case, when the 840 kc/s
beat is strongest relative to the luminance signal, is when the sound
carrier is modulated 100% and the picture is displaying a saturated
red at maximum amplitude.   Experiments indicate that a minimum
of 37 dB of sound rejection is required, which is rather less than most
monochrome receivers incorporate anyway.   This is another
instance where the colour transmission is compatible with mono-
chrome reception.   For systems using F.M. sound where the sound
carrier is not locked in frequency and hence the sound sub-carrier
beat is not an odd multiple of half the line and field frequencies,
some 20 dB more sound rejection is required.

Normal type monochrome sound rejection circuits have been
used satisfactorily in colour television receivers, as in the circuit
of Fig. 8.3.   R.C.A. have developed an interesting sound rejector
circuit called the bifilar-T trap and this has been used in American
and British receivers.   The bifilar-T circuit is illustrated in Fig. 8.4.
L_1C_1 is the trap circuit proper and determines the frequency of
rejection whilst the series bifilar elements resonate at a midband
frequency.   Theoretically the circuit is a non-minimum phase
type of circuit and the phase response is rather better than a normal

Fig. 8.3. I.F. circuits of the G.E.C. colour receiver TTIV
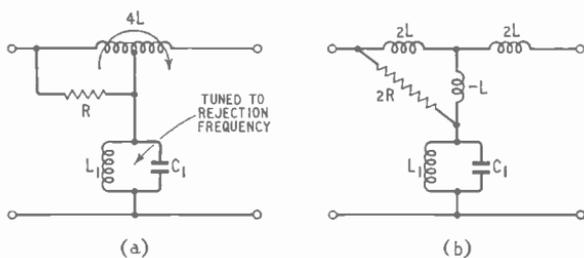
Fig. 8.4. Bifilar-T trap (a) and equivalent circuit (b)
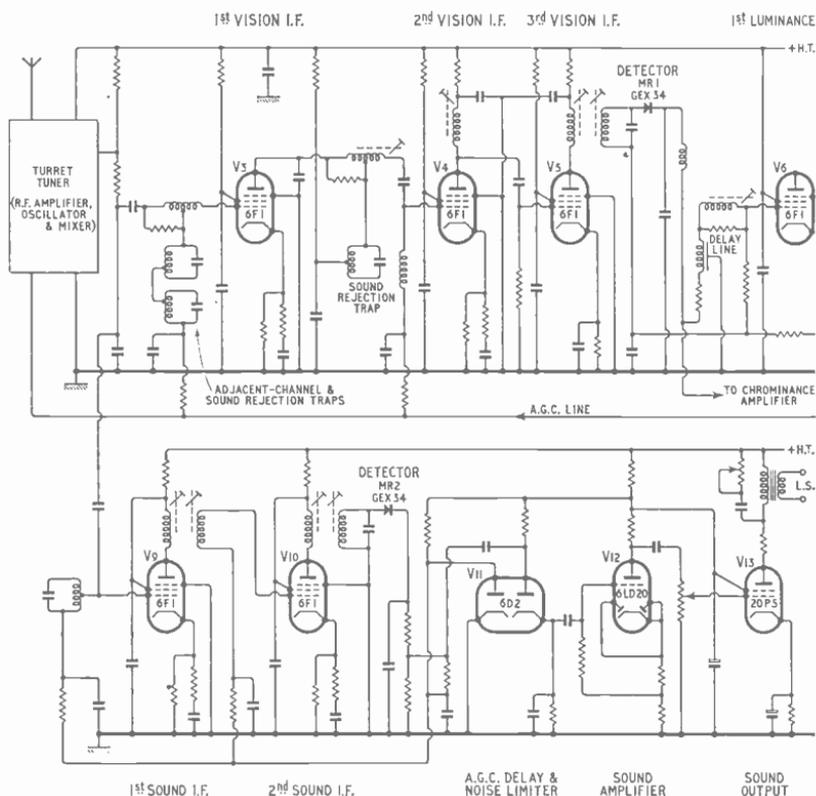
Fig. 8.5. I.F. and sound circuits of the Murphy colour receiver

type circuit, but the inevitable losses in the circuit and the fact that oscillator drift makes it difficult to keep the sound I.F. exactly at the peak rejection, tends to nullify some of its advantages. A further trap can be incorporated in the same circuit for rejection of the adjacent sound frequency. Fig. 8.5 shows how they are used in a Murphy receiver.

8.5. Vision I.F. detectors

A single diode can be used for demodulating the multiplex I.F. vision signal, as in Fig. 8.5, and the design of this stage can be quite normal providing care is taken to keep the overall video response
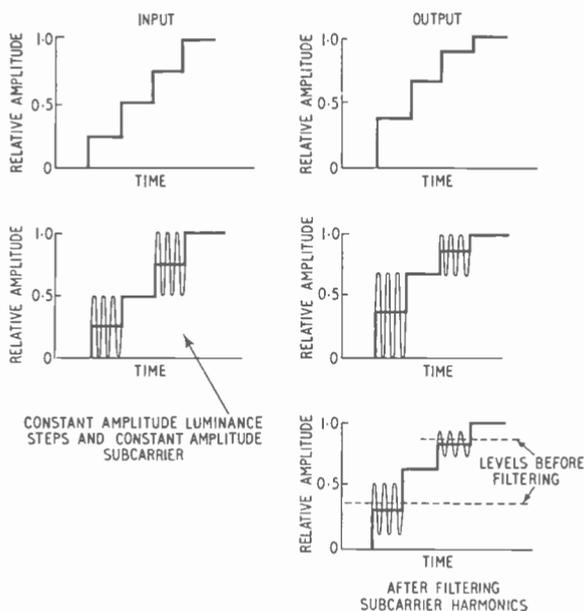
Fig. 8.6.   Differential amplitude distortion

flat to 3 Mc/s. The input impedance of diode detector circuits tends to increase over the single sideband region of the signal. This can boost the sub-carrier with respect to the carrier and may cause the sub-carrier peaks on saturated low luminance colours to extend down to the zero carrier level, and in extreme cases may cause enough distortion to enhance the luminance.

Amplification of the complete video signal as one entity raises difficulties in the design of the video amplifiers and there is some
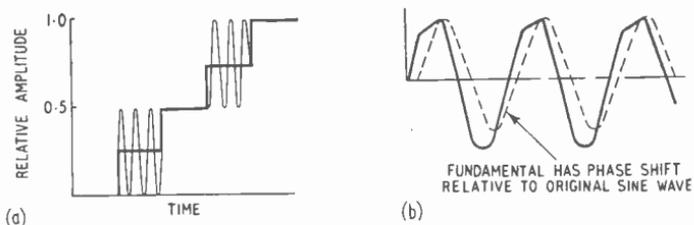
Fig. 8.7. *Differential phase distortion with diagonal clipping shown at (b)*
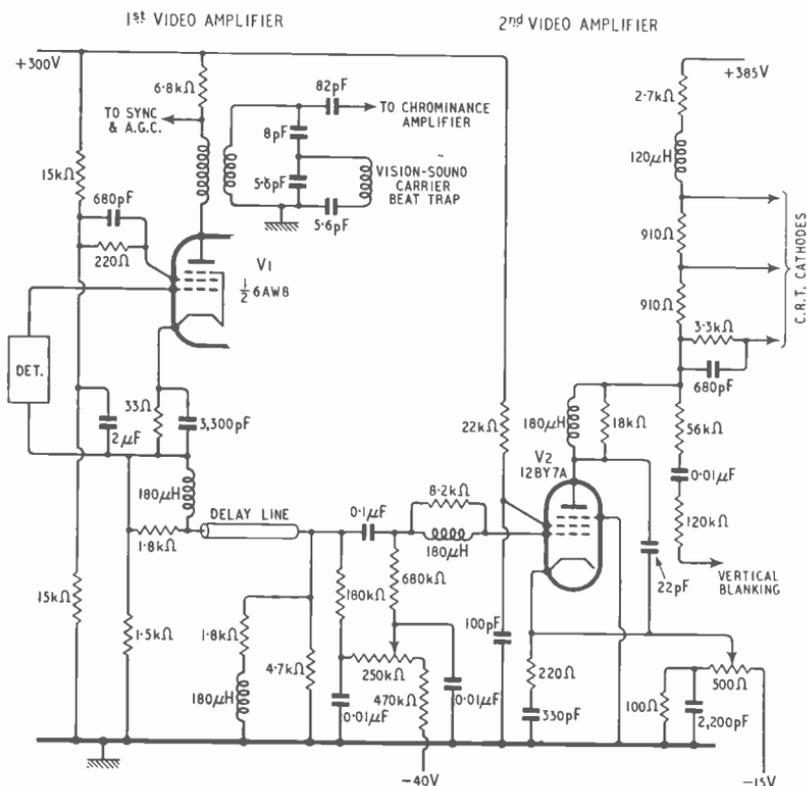
Fig. 8.8. *Video circuits of the R.C.A. colour receiver CTC 5N*

advantage in using two detectors, one accepting the full luminance bandwidth and the other the chrominance bandwidth only. The use of two detectors also overcomes difficulties due to reflections in the luminance delay cable affecting the chrominance response. An example of this kind of technique is shown in Fig. 8.3.

8.6. Luminance amplifiers

If there is only one vision detector, and the chrominance and luminance signals are amplified together in one or more video stages, then the input-output linearity of the stages becomes much more important than in black and white reception. The sub-carrier signal rides up and down on the luminance signal as the latter varies in amplitude (*see* Fig. 4.2) and it is important that neither the normalized amplitude nor the phase of the sub-carrier signal shall depend on the luminance level at which the signal passes through the video stage. If this sort of distortion does occur then, for example, the colour of actors' clothes will vary as they emerge from the shadows into the brightly lit parts of the scene and will also change with the setting of the contrast control.

When the normalized amplitude of the subcarrier varies with the luminance amplitude it is called *differential gain distortion*. Of course, for a given purity of colour the sub-carrier amplitude should change with luminance level but the ratio of sub-carrier amplitude to luminance amplitude should be constant for any particular colour however much light is falling upon it. Fig. 8.6 illustrates what happens. The sub-carrier signal is stretched non-uniformly from the positive peaks to the negative peaks. The mean level of the resultant signal is then different from the true luminance level, and since the harmonics of the sub-carrier are rejected by the video pass-band the final peak-to-peak fundamental amplitude of the sub-carrier may be different from the true value. Both these effects can change the purity of the reproduced picture.

If the phase of the sub-carrier varies with the luminance level then the effect is termed *differential phase distortion*. Such effects can occur due to diagonal clipping as in Fig. 8.7, or in the I.F. amplifiers due to variation in bias altering the amount of feedback in the circuit. Similarly, a variable cathode resistance used as a control in the luminance amplifier can cause differential phase effects due to the shunt capacitance to ground across the cathode resistor.

Fig. 8.8 illustrates the first video amplifier in an R.C.A. colour receiver. The first valve is used to split the chrominance signal
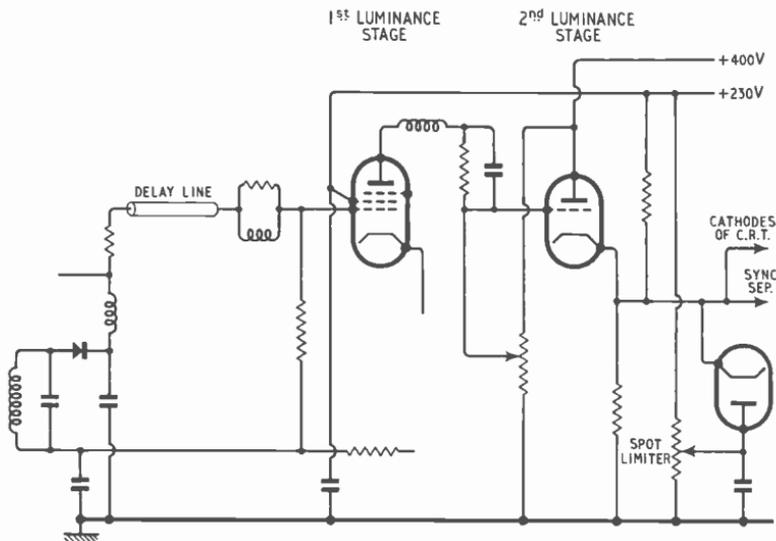
Fig. 8.9.   Video circuits of the Murphy colour receiver

from the luminance signal, and the output from the detector is applied between grid and cathode of the valve. In this way the amplification of the valve is divided between the chrominance and luminance channels. The Murphy receiver of Fig. 8.9 separates the signals immediately after the detector, whilst a G.E.C. receiver uses separate detectors, Fig. 8.10.

Once the chrominance has been separated out the luminance signal is delayed so that its waveforms will be in time coincidence with the waveforms from the narrow band chrominance circuits (*see* Section 4.25). The delay is usually accomplished by delay cable but lumped low-pass filter circuits or all-pass bridged-T phase-equalizing circuits may be used. The delay required depends upon the chrominance bandwidth and shape but is typically about 0.6 to 0.8 μs and requires about 1 ft of delay cable.

The termination of the delay cable is important if reflections and the production of spurious echoes on the picture are to be prevented. The constant resistance network termination of Fig. 8.10 is useful in this respect.

The luminance signal is carrying the 2.66 Mc/s sub-carrier which is an unwanted interfering signal in the luminance channel. Although the choice of sub-carrier frequency as an odd multiple of half the line and field frequencies helps to reduce the visibility of

the dot pattern, it can be disturbing and can produce an objectionable beat pattern with the dot structure of the shadow mask tube screen. Further, because the input-output characteristic of the display tube is curved the sub-carrier gets partially rectified by the tube, as shown in Fig. 8.11, and the resulting D.C. component of the signal decreases the saturation of the colours. On monochrome receivers this can be an advantage (*see* Chapter 15) but it is to be avoided in colour reception. The luminance amplifier therefore includes a sub-carrier rejector circuit, which is adequate to remove the dots from the large areas of the picture without unduly upsetting the luminance transient response. On Test Card C the sub-carrier rejector circuit reduces the contrast of the 2·5 Mc/s resolution bars but hardly affects the 2 and 3 Mc/s bars. Since the sidebands of the sub-carrier are similarly accepted, dots still appear in small coloured areas and at the edges of larger ones.

Even though it is not carrying chrominance, the luminance linearity is still important. Any curvature of the input-output
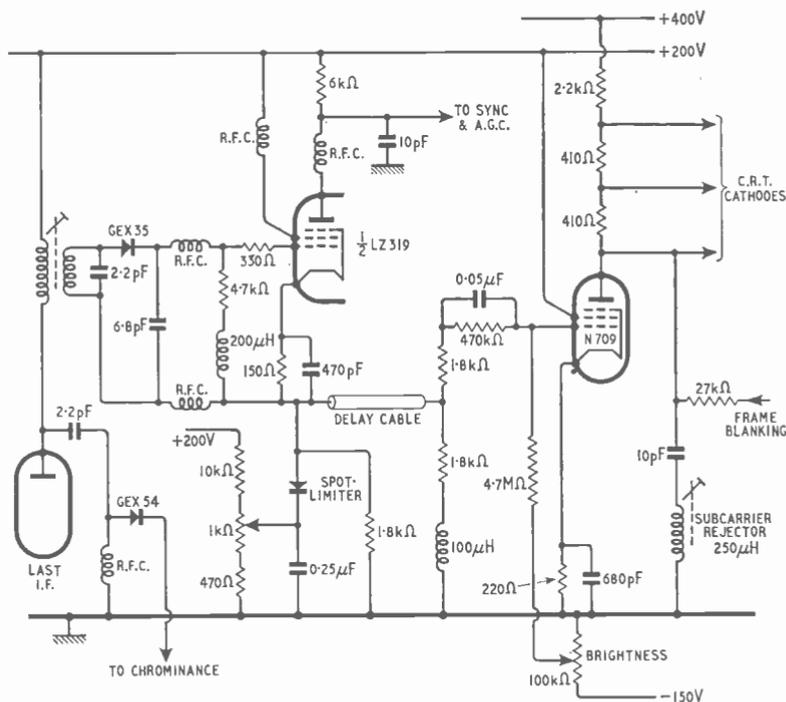
Fig. 8.10. Video circuits of the G.E.C. colour receiver TT1V

characteristic will produce the usual monochrome effects of incorrect tonal gradation and will cause variation of the saturation values.

The Y' signal after suitable wideband amplification goes to the matrix circuit. Usually the matrixing may be performed by the cathode-ray tube itself, in which case a final output voltage of about 120V peak-to-peak video may be needed.

A spot limiter may also be incorporated in the luminance channel and this will normally be sufficient because of the constant luminance principle, although it is possible to spot limit each R, G and B channel separately.

8.7. Chrominance amplifiers

Chrominance amplifiers have no counterpart in monochrome receivers. They are essentially video bandpass amplifiers accepting the sub-carrier frequency of 2·66 Mc/s with its sidebands and rejecting the lower luminance frequencies. The input to the chrominance
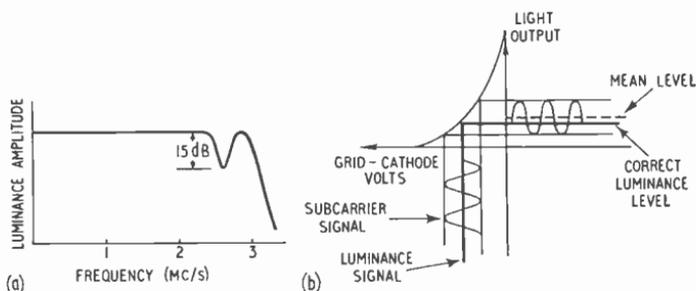
Fig. 8.11. *Rectification of the sub-carrier by the display tube.* (a) *shows the sub-carrier rejection with a notch bandwidth of 0·4 Mc/s at 3dB.* (b) *shows the display tube transfer characteristic*

amplifier may be taken either from a separate detector or from a point in the luminance channel. The later the take-off point the less gain is needed in the chrominance channel but the greater is the risk of differential phase and gain effects. A typical chrominance channel response is shown in Fig. 8.12.

From the point of view of frequency response there are two alternative designs for the chrominance channel. If full use is to be made of the I' information then the chrominance bandwidth must be wide enough to accept not only the $\pm 0·34$ Mc/s over which both I' and Q' are double sideband, but also the 0·66 Mc/s or so over which I' is single sideband. This means a total bandwidth of approximately 1·3 Mc/s. The overall response, including the R.F.
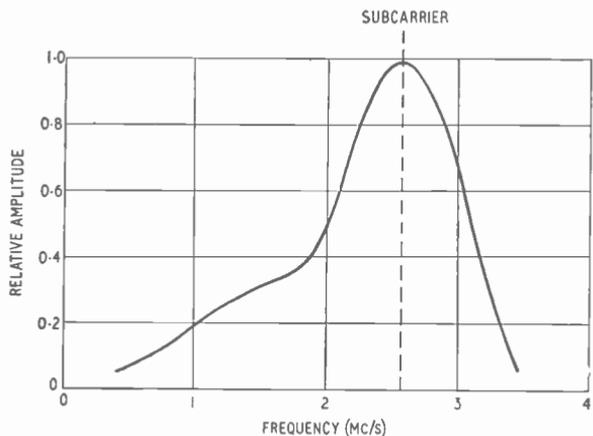
Fig. 8.12. *Typical chrominance response for an equiband receiver. The amplitude is measured between aerial and colour tube grids*
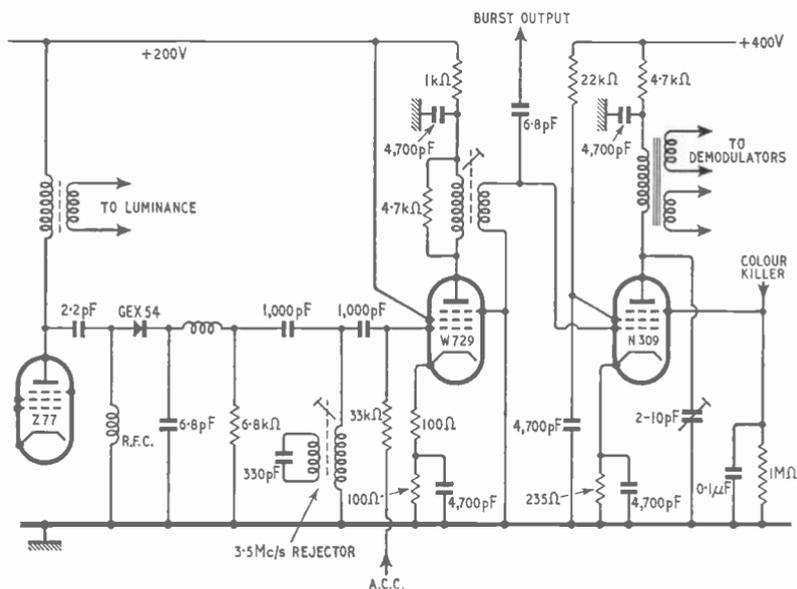
Fig. 8.13. *Chrominance amplifiers in the G.E.C. colour receiver TTIV*

and I.F. amplifiers, should be flat over the double sideband region and have a 6 dB boost over the 0·6 Mc/s region of single sideband operation in order to correct for the missing sideband. Alternatively, correction can be applied in the I' channel after synchronous detection. Some $I'-Q'$ receivers omit this amplitude correction altogether, partly because it emphasizes the cross-colour from the
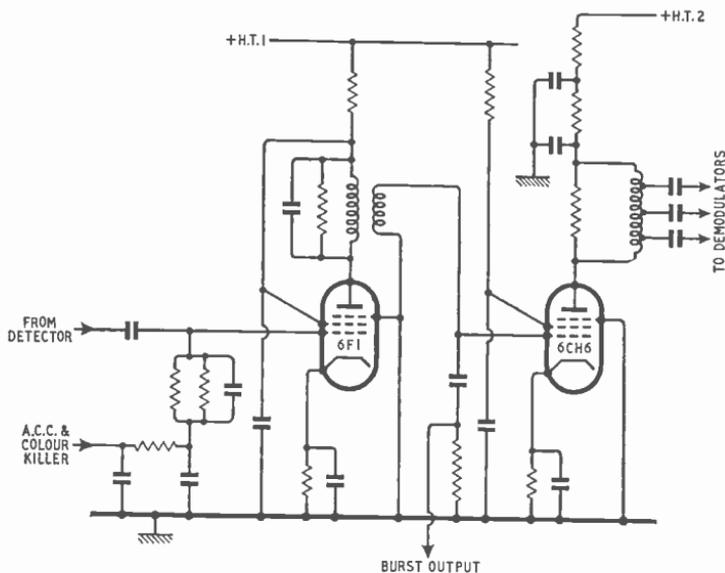
Fig. 8.14.   Chrominance amplifiers in the Murphy colour receiver

lower frequency luminance components which are accepted by the chrominance channel. The lower frequency limit of the I' channel is 1·7 Mc/s approximately, and there is a considerable amount of energy at this frequency in the average luminance signal.

For equiband operation the chrominance acceptance band can be narrower, ± 0.34 Mc/s, whilst wideband equiband receivers may use ± 0.5 Mc/s, again without boost for the missing sideband.

Figs. 8.13 and 8.14 show two British examples of chrominance amplifier design. The G.E.C. design is a wideband equiband channel while the Murphy receiver is for equiband operation. Wideband $I'Q'$ receivers should incorporate a delay in the I' channel to ensure registry of the wider band I' signal with the Q' signal.

A manual gain control is normally fitted to one or more of the chrominance amplifier stages to control the reproduced purity of

the colours and this control is usually labelled saturation.   Often, to avoid changing bias conditions, a potentiometer is used to vary the signal level at the input to the amplifier.   The contrast control operating on the gain of the I.F. stages varies both chrominance and luminance together so that the final purity of the reproduction is not changed, only the peak brightness of the picture is altered. Some form of automatic chrominance control (A.C.C.) is often fitted, to keep the chrominance gain constant at the predetermined manual setting as the receiver is switched from one channel to another, or as it fades in fringe area reception.   This A.C.C. voltage can conveniently be obtained by measuring the burst amplitude, just as the A.G.C. voltage can be obtained by measuring the sync pulse amplitude.   The derivation of the A.C.C. voltage is discussed in Chapter 10, as also is the production of the colour killing voltage.

The chrominance amplifier has to feed both or all the synchronous detectors and also the reference generator.   If the feed to the reference generator is after the point at which the colour killing is applied, then the colour killing voltage must be cancelled during the back porch so that the colour killing circuits may judge the presence or absence of the burst.

A completely automatic chrominance gain control voltage, which sets the chrominance gain to the value required for correct colour reproduction, can be obtained by comparing the peak-to-peak amplitude of the burst with the height of the sync pulse.   These two amplitudes should be the same at equivalent gain points in the receiver.   The sound carrier must be firmly rejected in the chrominance amplifier, since a sound signal which is 40 dB down at the luminance and chrominance detectors will produce sufficient amplitude of 3·5 Mc/s for it to be amplified by the chrominance channel and then beat, at the synchronous detectors, with the 2·66 Mc/s reference frequency.   The net result is the production of an 840 kc/s difference frequency which causes a coloured beat pattern on the display tube.   It is customary therefore to put a 3·5 Mc/s rejector circuit in the chrominance amplifier.

The required voltage output of the final chrominance stage depends upon the type of synchronous demodulator used.   For high level demodulation an output of some 120V may be required, whereas for low level demodulation a few volts may suffice.   Low level demodulators require further video amplifiers to raise the level of the video signals to the 120 volt level required by the shadow mask tube.   If these video stages are amplifying colour difference signals then they can have a narrow frequency response.   If the signals are matrixed to R', G' and B' before further amplification,

each video amplifier must have the full video bandwidth and the gains of the three channels must be accurately stabilized. Any variation in gain of a primary signal amplifier produces noticeable changes in hue, saturation and colour balance. Changes in gain of a colour difference amplifier produce only changes in hue and saturation and leave the grey scale balance unaltered (*see* Chapter 4). For this reason it is usual to leave the signals in colour difference form until the last stages of the receiver.

For 100V of luminance drive the colour difference amplifiers must be able to produce these maximum signal excursions

$$(R' - Y') = \pm 70\text{V}$$
$$(G' - Y') = \pm 41\text{V}$$
$$(B' - Y') = \pm 89\text{V}$$

If the display device has phosphors with different efficiencies, these figures must be suitably modified (*see* Section 11.8).

8.8. Summary

Colour tuners and I.F. strips are basically similar to good quality monochrome units and careful design is enough to prevent the vision carrier, sub-carrier and sound carrier from intermodulating unduly. Phase distortion at the sub-carrier frequency will not affect the colour reproduction of large picture areas if the chrominance and burst traverse similar paths in the receiver. For good colour transients a linear phase and flat amplitude response are needed throughout the whole video pass-band.

Inadequate sound rejection can produce both a luminance only or a coloured 840 kc/s beat pattern on the received picture. Inadequate vision rejection in the sound channel will allow the sub-carrier to produce an audible low frequency buzz.

To avoid differential gain and phase effects the video luminance and video chrominance signals are preferably amplified separately. Sub-carrier rejection is needed in the luminance channel and 3·5 Mc/s rejection in the chrominance amplifier. The Y' and I' signals must be delayed to ensure time coincidence with the Q' signal. If equi-band working is used, only the luminance signal need be delayed. Improved colour stability results if the chrominance signals are left as colour difference signals as long as possible.

A saturation control to vary the gain of the chrominance amplifier is required, and the gain of the chrominance channel must be reduced to zero, or killed, when the transmission is in monochrome.

Colour Receiver Decoding Circuits

9.1. Introduction

The purpose of the decoding circuits in a colour receiver is to derive, from a composite video input, the red, green and blue signals for driving the display tube.

Normally, the composite video input is obtained from the second detector which demodulates the I.F. signal. The usual practice is to feed the detector output into two channels, one being the luminance channel whose output is the luminance or Y' signal and the other being the chrominance channel which accepts the modulated sub-carrier signal, synchronously detects the sub-carrier, and provides an output of either I' and Q' signals, or difference signals. The combination of the Y' signal and the I' and Q', or difference signals, in a matrix circuit results finally in the required R', G' and B' signals.

Because the Y' channel of a colour receiver is almost identical with the usual video channel of a monochrome receiver, it has become common practice to refer to the processing of the modulated sub-carrier signal to yield chrominance signal outputs as the *decoding process*, since this is peculiar to colour receivers. It may therefore be stated that a colour receiver decoding circuit performs two functions: synchronous detection of the modulated sub-carrier signal and simple algebraic manipulation (if necessary) of the resulting outputs to give the difference signals $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$ and ultimately the primary signals R', G' and B'. This latter process is usually known as *matrixing*.

9.2. Synchronous detection

The modulated sub-carrier consists of two amplitude modulated carriers which have the same frequency but a 90° phase difference between them. The process of synchronous detection enables the modulation on one of the carriers to be detected without interference from the modulation on the other carrier and consequently by using two separate synchronous detectors both sets of modulation may be recovered independently.

Normally, one carrier is modulated with the I' parameter and the other with the Q' parameter, though it will be seen later that in some

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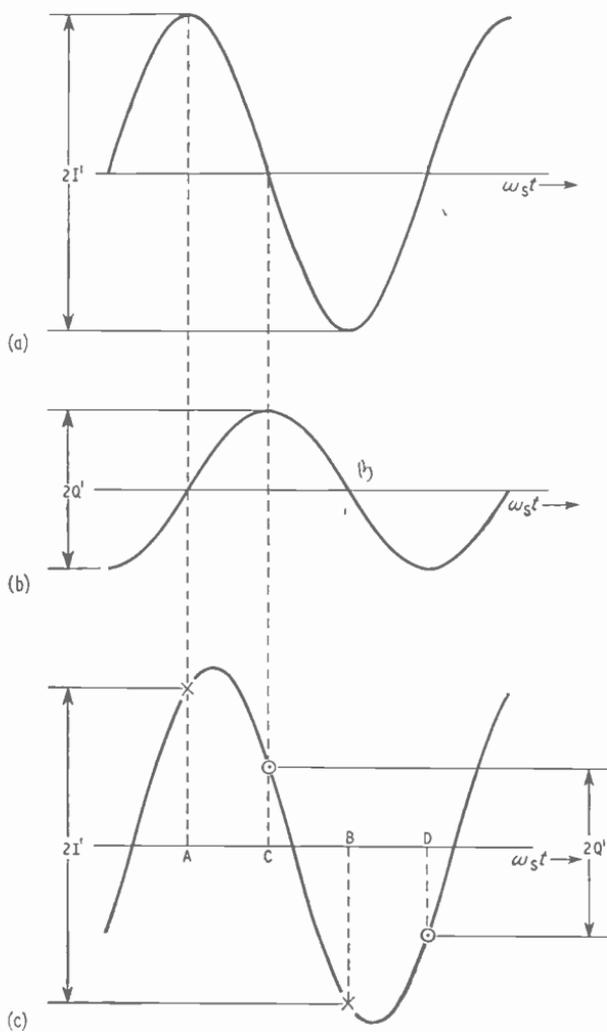
Fig. 9.1.   Principle of synchronous detection. $\omega_s = 2\pi \times$ sub-carrier
frequency.   (a) shows I' component of sub-carrier, (b) the Q' component
of sub-carrier and (c) the total sub-carrier signal

cases this arrangement is equivalent to quadrature modulation by the $(R' - Y')$ and $(B' - Y')$ signals.

As an example of the synchronous detection process, assume that the I' and Q' signals are simple D.C. values. Then the modulated sub-carrier will have two components: the first will be a sine wave of sub-carrier frequency having an amplitude proportional to the I' signal, and the second will be a similar sine wave having an amplitude proportional to the Q' signal but with a phase lagging 90° on the I' component phase. Note that, if viewed on an oscilloscope, the above modulated sub-carrier would appear as a simple sine wave because the addition of two equal frequency sine waves 90° apart in phase is itself a sine wave. Thus, the I' component of the sub-carrier is shown in Fig. 9.1 (a), the Q' component is shown in Fig. 9.1 (b) and the resulting modulated sub-carrier is shown in Fig. 9.1 (c), for a modulated sub-carrier which has 2 units of I' and 1 unit of Q'.

Note that since the I' and Q' carriers are 90° apart in phase, at the instant when either is at a positive or negative maximum, the other is at zero. Referring to Fig. 9.1 (c), the I' parameter may therefore be recovered by measuring the instantaneous value of the modulated sub-carrier at the particular instant when the I' carrier is at a maximum and the Q' carrier is zero. This time instant is labelled " A ". If the subcarrier is measured at instant " B" then a negative version of the original I' modulation will be obtained.

Similarly, Q' may be detected by measuring the modulated sub-carrier at the instant " C ", and a negative version of Q' will be obtained by measuring at the instant " D ".

While a simple case of D.C. values of I' and Q' has been taken here, it obviously follows that if I' or Q' or both have A.C. components, then these can be recovered by the same process of measuring the modulated sub-carrier at the appropriate instants in time.

An essential feature of synchronous detection is the measurement of the instantaneous amplitude of the sub-carrier signal at the appropriate time instants. This timing information is obtained from a continuous, unmodulated sine wave which is locked in frequency and phase to the originating sub-carrier oscillator at the transmitter.

This continuous sine wave is referred to as the reference signal, and the colour receiver must generate it from the colour burst or short 9 cycle sample which is transmitted during the post sync blanking period. Reference generation is discussed in detail in Chapter 10, but for the purposes of this chapter it should be assumed that a reference signal is available at the receiver.

In synchronous detector arrangements, it is usual to specify the detection axis along which detection takes place. The angular position of this axis is dependent on the phase of the reference signal used in the detector. Thus, in the example above, the I' parameter is recovered by detection along the $+ I'$ axis (which lags the burst phase by 57°) and the Q' parameter by detection along the $+ Q'$ axis (which lags the burst phase by 147°).

However, it will be seen later that in equiband working detection axes other than I' and Q' may be used, and referring to Fig. 9.2,
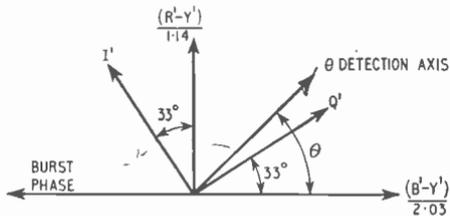
Fig. 9.2. Derivation of output of synchronous detector operating along a particular detection axis

the output from a detector operating along an axis leading the $\dfrac{(B' - Y')}{2 \cdot 03}$ axis by θ is equal to the sum of the projections of the $\dfrac{(B' - Y')}{2 \cdot 03}$ and the $\dfrac{(R' - Y')}{1 \cdot 14}$ axes on to this detection axis. Thus, the output from a detector operating along the θ axis would be

$$\frac{(B' - Y')}{2 \cdot 03} \cos \theta + \frac{(R' - Y')}{1 \cdot 14} \sin \theta$$

which is the same as

$$Q' \cos (\theta - 33°) + I' \sin (\theta - 33°)$$

provided the modulation frequencies involved are low enough to be within the Q bandwidth.

9.3. Diode detection

Several different types of synchronous detector will be described in this chapter, but by way of illustration, consider the operation of the simple double diode detector shown in Fig. 9.3.

Suppose a sub-carrier chrominance signal of amplitude S and phase θ relative to the Q' axis is fed to a two to one step-up transformer, and suppose a reference sine wave of sub-carrier frequency (derived from the colour burst) of amplitude P, and in the same phase as Q', is injected as shown. Then the amplitude of the voltage across D_1 will be equal to the vector sum of S and P, and

this voltage will appear as a positive voltage between A and C. Similarly, the vector difference between **S** and **P** will appear as a positive voltage between B and C. The voltage between A and B will then be the algebraic difference between AC and BC. This voltage between A and B can be calculated.

The voltage between A and C will equal the peak value of the vector sum of **S** and **P**. Thus, referring to Fig. 9.4 (a), if θ is the phase angle between the chrominance input signal and the reference signal, then A will be positive relative to C by an amount equal to the magnitude of the vector **LN**.

Now suppose that **P** is made very much greater than **S**, then in the triangle **LMN**, **LM** $\simeq$ **LN** and therefore

$$V_{AC} = \mathbf{LN} \simeq \mathbf{P} + \mathbf{S} \cos \theta$$

Similarly, the corresponding conditions for the voltage between B and C are shown in Fig. 9.4 (b), from which

$$V_{BC} = \mathbf{LN} \simeq \mathbf{P} - \mathbf{S} \cos \theta$$

Fig. 9.3. Double diode synchronous detector

Fig. 9.4.   Vector relations for a double diode synchronous detector
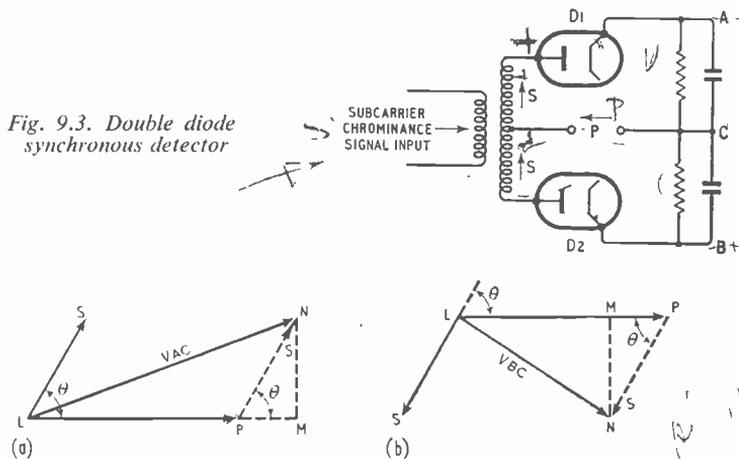
Now because the A to C and the B to C terminals are decoupled by the capacitors, the resultant voltage between A and B will be equal to the algebraic difference ($V_{AC} - V_{BC}$), that is

$$V_{AB} = \mathbf{P} + \mathbf{S} \cos \theta - \mathbf{P} + \mathbf{S} \cos \theta = 2\mathbf{S} \cos \theta$$

Now **S** cos θ is the component of **S** which has the same phase as **P**, that is, the above detector measures the amplitude of **S** in the **P**

phase direction. Hence, by making the reference **P** phase equal to the Q' phase, for example, the detector output will be equal to Q'.

Obviously, if the **P** phase is made equal to that of I' (i.e. leading the Q' phase by 90°) the detector output will be I'. Consequently, two identical double diode detectors as described above will demodulate the sub-carrier signal if both detectors are fed with the same chrominance signal but with two differently phased references, one in the Q' phase and one in the I' phase.

Note that a sign reversal of the detector output can be obtained by changing the reference phase by 180°. Thus, for a phase difference of θ between **S** and **P** the output is 2S cos θ, while for a phase difference of $(\theta + 180°)$ the output is

$$2S \cos (\theta + 180°) = - 2S \cos \theta$$

While the above results for the phase detector are approximate in that the **P** signal has been assumed to be very much greater than the **S** signal, in practice the approximation is justified since even for the severe case where **P** and **S** are equal, the maximum amplitude error is only 1dB and the maximum phase error is only 3°, for two detectors operating in quadrature. Normally **P** is made at least twice the largest **S** value. It should be noted that the diode detector circuit of Fig. 9.3 will operate satisfactorily if the sub-carrier and reference inputs are interchanged. This may be seen by rotating the vector diagram of Fig. 9.4 (b) through 180°, so that **S** rather than **P** becomes the " reference vector " between Fig. 9.4 (a) and Fig. 9.4 (b).

9.4. Matrixing

When the I' and Q' parameters have been obtained from the synchronous detectors, the next step is to derive the difference signals $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$ from I' and Q' and then Y' has merely to be added to these to give the required R', G' and B' signals. A method for deriving R', G' and B', given the outputs from an I' and a Q' detector and a Y' signal will be described. The equations to be satisfied are

$$R' = (R' - Y') + Y' = + 0 \cdot 96I' + 0 \cdot 63Q' + Y'$$
$$G' = (G' - Y') + Y' = - 0 \cdot 28I' - 0 \cdot 64Q' + Y'$$
$$B' = (B' - Y') + Y' = - 1 \cdot 11I' + 1 \cdot 72Q' + Y'$$

Notice that the total proportion of I' required is 2·07, that is, from $- 1 \cdot 11$ to $+ 0 \cdot 96$. On the other hand, for Q' the total proportion is from $- 0 \cdot 64$ to $+ 1 \cdot 72$, or 2·36. Since Q' has the greater range,

let us first see how to obtain the various proportions of Q we require, namely, $+ 0\cdot63$ for red, $- 0\cdot64$ for green and $+ 1\cdot72$ for blue.

Referring to Fig. 9.5 suppose the points A and B are the output terminals of the simple double diode detector of Fig. 9.3.   Suppose the total load resistance between A and B is L.   Call the total output
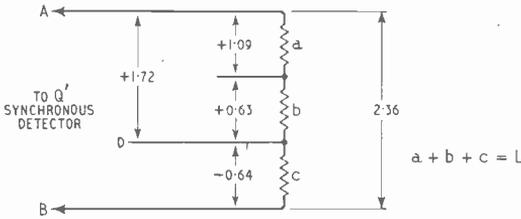
Fig. 9.5.   Q' matrix circuit
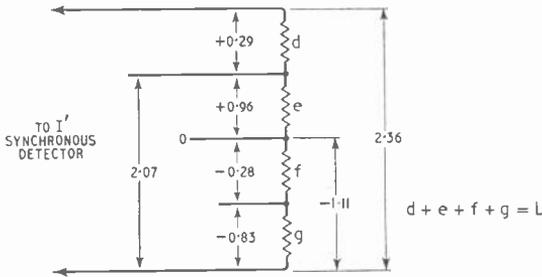
Fig. 9.6.   I' matrix circuit

$2\cdot36$.   First, find a tapping point O on the load L which divides L in the ratio of $1\cdot72$ to $0\cdot64$.   Then, for $2\cdot36Q'$ across the whole load, there is $+1\cdot72Q'$ at A and $- 0\cdot64Q'$ at B. $1\cdot72Q'$ is needed for blue and $- 0\cdot64Q'$ for green, but $+ 0\cdot63Q'$ is also required for red.   So take another tap point between O and A which will give $+ 0\cdot63Q'$.   Hence, three separate resistors a, b and c are required and the value of these can be found in terms of their sum, which is, of course, the total load L.   Thus

$$\frac{a}{L} = \frac{1\cdot09}{2\cdot36}, \quad \frac{b}{L} = \frac{0\cdot63}{2\cdot36}, \quad \frac{c}{L} = \frac{0\cdot64}{2\cdot36}$$

i.e. $a = 0\cdot462L$, $b = 0\cdot267L$, $c = 0\cdot271L$

The three required proportions of Q' are now available, and next the three values of I' will be derived.   The total Q' range was $2\cdot36$, but the total I' range is $2\cdot07$.   If the two detectors are exactly

similar, the total I' output must be taken as being 2·36 to get the correct gain relationship between Q' and I', for Q' and I' have been equally modulated (in amplitude terms) on to their carriers according to the sub-carrier equation

$$Q' \sin \omega t + I' \cos \omega t$$

Thus, in Fig. 9.6 the total I' output is 2·36, of which only 2·07 is required. A resistor d is therefore needed so that

$$\frac{d}{L} = \frac{0 \cdot 29}{2 \cdot 36}$$

For the 2·07 range of I', we can now proceed as for Q'. Thus, the tap point O will divide the 2·07 range in the ratio of 0·96 to 1·11, and the resistor values are given by

$$\frac{e}{L} = \frac{0 \cdot 96}{2 \cdot 36}, \quad \frac{f}{L} = \frac{0 \cdot 28}{2 \cdot 36}, \quad \frac{g}{L} = \frac{0 \cdot 83}{2 \cdot 36}$$

i.e. $e = 0 \cdot 406L$, $f = 0 \cdot 118L$, $g = 0 \cdot 351L$.

The various proportions of I' and Q' are now available and the final step is to add the particular I' and Q' values, and Y'.

It should be noted that, in practice, a balanced arrangement is desirable in order to minimize the effects of stray capacities on the
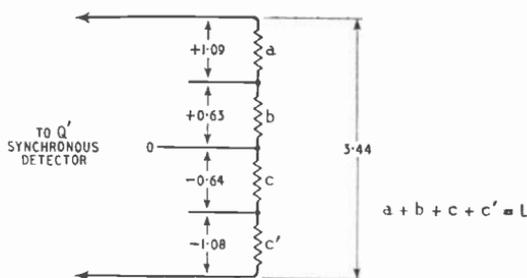
Fig. 9.7. Balanced Q' matrix circuit

various I' and Q' outputs. Then the points O of Figs. 9.5 and 9.6 would become the centre taps of the detector loads. The Q' detector load can be balanced by including an additional resistor c' in series with c, such that

$$a + b = c + c'$$

This would make $\dfrac{c'}{L} = \dfrac{1 \cdot 08}{3 \cdot 44}$, and the total load would have the proportional value 3·44 instead of 2·36. This is shown in Fig. 9.7.

The I' detector load must now be modified to have the proportional value 3·44, and this can be done by changing the resistor d so that its output becomes $(3·44 - 2·07)$, i.e. 1·37.   Next, in order to balance the I' detector load, d may be split into d_1 and d_2 (so that
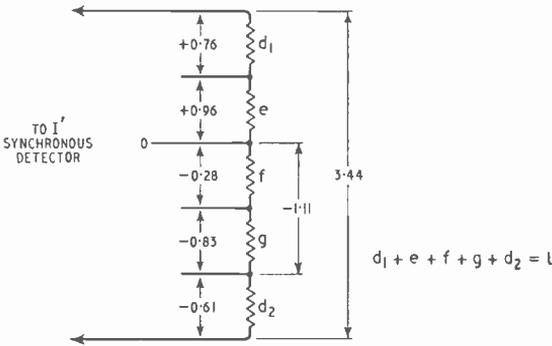
Fig. 9.8.   Balanced I' matrix circuit

$d = d_1 + d_2$), and d_2 can then be placed in series with g.   The values of d_1 and d_2 can be calculated as follows:

The output from d, must be $\dfrac{3·44}{2} - 0·96 = 0·76$

Hence, $d = \dfrac{0·76}{3·44}L = 0·221L$

Similarly, the output from d_2 must be $\dfrac{3·44}{2} - 1·11 = 0·61$

Hence, $d_2 = \dfrac{0·61}{3·44}L = 0·178L$

The balanced I' detector load is then as shown in Fig. 9.8.

The complete circuit for an $I'Q'$ diode decoder is shown in Fig. 9.9.   Note that delay must be included in the Y' channel and some delay is also necessary in the I' channel.   The I' and Q' detector outputs include low pass filters with cut-off frequencies of about 1 Mc/s and 0·34 Mc/s, respectively.

If the synchronous detectors are transformer coupled, as shown, the tap points O of the loads can be connected together and to the
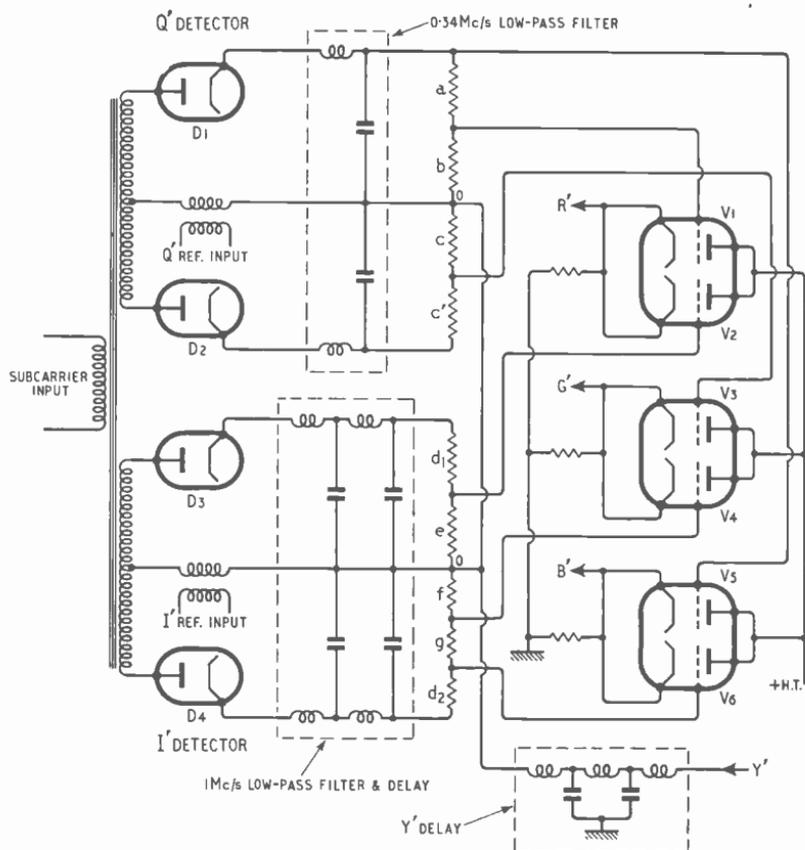
Fig. 9.9. I', Q' diode synchronous detection and matrixing

Y' signal. The appropriate voltages can now be added in cathode followers to give the required red, green and blue voltage outputs, which must be amplified individually before being applied to the display tube. The correct gain relationship between the Y' and the I', Q' parameters can be obtained by adjusting the saturation control which varies the drive applied to the synchronous detectors relative to the Y' signal.

The above matrix arrangement has been described in detail in order to illustrate the process of transformation from one set of parameters, such as I', Q' and Y', to another set, such as R', G' and B', and equal phosphor efficiencies have been assumed. In general, the matrix circuit which follows the synchronous detectors

is intimately connected with the type of detector used, and it is therefore desirable to consider the design as a complete unit.   For this reason, the detectors to be described will include the appropriate matrix circuits.

It is sometimes required to derive one of the difference signals given the other two.   This can be arranged by using a simple matrix and a phase inverting stage.   The relationship between the difference signals can be deduced from the luminance equation

$$Y' = 0.30R' + 0.59G' + 0.11B'$$

by writing

$$0.30R' + 0.59G' + 0.11B' = 1 \; Y' = (0.30 + 0.59 + 0.11) Y'$$

Hence

$$0.30R' - 0.30Y' + 0.59G' - 0.59Y' + 0.11B' - 0.11Y' = 0$$

i.e.

$$0.30(R' - Y') + 0.59(G' - Y') + 0.11(B' - Y') = 0$$

Thus to derive $(G' - Y')$, add $(R' - Y')$ and $(B' - Y')$ in the ratio of 0.30 to 0.11 so that

$$-0.59(G' - Y') = 0.30(R' - Y') + 0.11(B' - Y')$$

If this mixture of $(R' - Y')$ and $(B' - Y')$ is applied to a simple valve amplifier with a gain of $\dfrac{1}{0.59}$, the output will be $(G' - Y')$. This technique is described in Section 9.9.

9.5.   Decoding circuits in general

There are several general points in the design of synchronous detectors and matrix circuits which should be mentioned.

If difference signals and the Y' signal are applied to different electrodes of the display tube, then the Y' signal should be applied to the cathode.   This is because the difference signal circuits are relatively narrow band, and if they are included in the tube cathode, selective feedback may occur and upset the frequency response of the displayed luminance.   It should also be pointed out that a matrix circuit which accepts the Y' signal and either difference signals or $I'Q'$ signals must be capable of handling the highest Y' frequency.

Conversely, matrix circuits which do not manipulate the Y' signal need have a frequency response which is only as wide as the widest chrominance component, i.e. 1 Mc/s for $I'Q'$ working, and about 0·5 Mc/s for equiband working, in the 405-line system.

In matrix circuits, one sometimes has the choice of deriving R', G' and B' or deriving difference signals and adding Y' in the display tube. Generally speaking, the latter is preferable from the point of view of white balance stability, since for white the difference signals are zero and the arrangement is therefore independent of gain variations in the difference signal circuits.

However, in some arrangements it may be possible for the standing voltage outputs from difference signal detectors to drift differentially and thereby upset the white balance, in spite of the difference signals being zero.

Another important consideration in decoding design is the maintenance of the D.C. components of all the parameters. D.C. suppression in the luminance signal will result in the wrong value of luminance being added to the right value of difference signals, so that errors in saturation will occur. Again, D.C. suppression in the I' Q' or difference signals can produce undesirable hue errors in the colour picture. For example, suppose the whole left hand half of the picture is a saturated red, while the remaining half of the picture is grey. Lack of D.C. response in the chrominance signals would tend to give a cyan (the complementary of red) cast to the grey half of the picture. The various waveforms for this case are shown in Fig. 9.10, in which a 3dB D.C. suppression of the difference signals has been assumed. Notice that although the peak to peak values of all the signals are unchanged by the D.C. loss, the red signal becomes slightly desaturated and the grey signal becomes a desaturated cyan.

While the above type of signal would rarely occur in practice, it often happens that large areas of the picture are predominantly of one colour so that insufficient D.C. maintenance would tend to produce a complementary hue cast in the remainder of the picture.

D.C. loss in the difference signals can occur if the D.C. bleeds to the background controls (the three " brightness controls ") are of too low an impedance, but usually it is possible to maintain the D.C. to within approximately 3dB.

The Y' and chrominance signal D.C. components can be preserved either by D.C. coupling after detection (or synchronous detection) or by a D.C. clamping technique. The XZ detector described later (*see* Section 9.11), is an example of chrominance signal clamping.

In the design of synchronous detectors, it must be emphasized
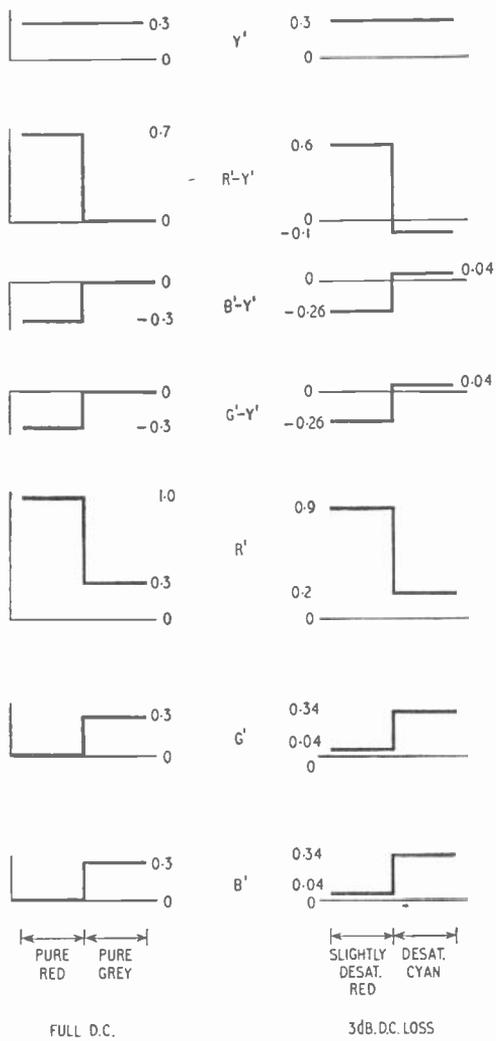
Fig. 9.10. Hue error resulting from a loss of 3dB in D.C. component of difference signals for full red to equal luminance grey transition

that when a synchronous detector is operated along one of the difference signal axes, the output will contain a factor such as $\dfrac{1}{1 \cdot 14}$ for the $(R' - Y')$ axis, and $\dfrac{1}{2 \cdot 03}$ for the $(B' - Y')$ axis. Thus, to obtain $(R' - Y')$, for example, a gain of $1 \cdot 14$ must be applied to the output of a detector operating along the $\dfrac{(R' - Y')}{1 \cdot 14}$ axis.

These factors $1 \cdot 14$ and $2 \cdot 03$ are introduced at the transmitter to limit the sub-carrier sine wave plus luminance swing so that it will not exceed, by more than $\frac{1}{3}$, the black to peak white amplitude range, as explained in Section 5.5. In decoding circuit design, it is the ratio between the factors (i.e. $\dfrac{1 \cdot 14}{2 \cdot 03}$ or $\dfrac{1}{1 \cdot 78}$) which is significant, since the absolute value of chrominance gain can be set as required by the receiver saturation control.

Another factor which is important in decoding circuits is the relative drive to the three guns of the display, assuming a three-gun tube is used. Since the red phosphor is the least efficient and the blue phosphor is the most efficient, it is recommended by the tube manufacturer that the drives be adjusted accordingly. For the R.C.A. 21AXP22A tube, for example, the R, G and B drives recommended are in the ratio 1 to $0 \cdot 8$ to $0 \cdot 6$ respectively. Thus, while an equal phosphor efficiency tube would require $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$, with a $- Y'$ signal to each cathode, the 21AXP22A tube requires $(R' - Y')$, $0 \cdot 8(G' - Y')$ and $0 \cdot 6(B' - Y')$, with cathode drives of $- Y'$, $- 0 \cdot 8 \, Y'$ and $- 0 \cdot 6 \, Y'$ respectively. These constants must be taken into account when designing the detector and matrix circuits.

9.6. Equiband working

A decoding process which recovers the I' and Q' parameters makes use of all the transmitted chrominance information; that is, (for the 405-line system) colour detail up to a bandwidth of about 340 kc/s is reproduced in three colours, and detail between 340 kc/s and 1 Mc/s is reproduced as a mixture of two colours along the I' axis, these two colours being roughly speaking an orange and a cyan together with the mean colour of the background. Above 1 Mc/s detail, the colour reproduction is in terms of Illuminant C white only. A receiver which makes use of all the transmitted information is forced to use I' and Q' detection axes, so that it may reject the crosstalking single sideband portion of its I' channel (for

I' modulation frequencies between 340 kc/s and 1 Mc/s) from its Q' channel.   Hence an $I'Q'$ receiver must include a matrix circuit arrangement to convert I', Q' and Y' to R', G' and B' for the display tube, and generally this matrix is fairly complex and therefore expensive.   In addition, delay is required in the I' channel since the I' bandwidth is greater than that of the Q' channel.

However, below 340 kc/s, there is no differential bandwidth limiting and any detection axes may be used provided the chrominance bandwidth above 340 kc/s is rejected by the receiver.   Thus, by having three colour reproduction for detail not exceeding 340 kc/s bandwidth, and for finer detail than this, reproduction in terms of Illuminant C only, together with the mean background colour, considerable simplification of the matrix circuits is possible.   In this case, the two synchronous detectors would nominally operate with equal bandwidths of 340 kc/s each, and such an arrangement is called *equiband detection*, for obvious reasons.

It might be argued that, since an equiband receiver rejects the higher frequency chrominance components, its resulting picture quality must be inferior to that of an $I'Q'$ receiver.   However, it should be borne in mind that the N.T.S.C. signal employs band-width sharing, so that $I'Q'$ operation would, on average, produce more luminance to chrominance crosstalk than equiband operation.   Thus, while an $I'Q'$ receiver would display finer colour detail, it would also display, as colour, luminance information in the video range from 1·66 Mc/s to 3 Mc/s, but the corresponding crosstalk range for narrow equiband working would only be from 2·32 Mc/s to 3 Mc/s.

A further point is that if the I' channel were to be given a 6dB single sideband boost between 340 kc/s and 1 Mc/s, the luminance crosstalk in the video range from 1·66 Mc/s to 2·32 Mc/s would be emphasized.

It would seem that further experimental work is required to determine whether $I'Q'$ or equiband operation is to be preferred, and, in fact, whether $I'Q'$ or equiband modulation should be employed at the transmitter.   However, it has been found that very good quality pictures can be obtained by using equiband detection with a slightly greater bandwidth than the Q' bandwidth, and the majority of American receivers employ this technique.   Thus, equiband operation for modulation frequencies up to about 0·5 Mc/s (in the 405-line system) gives very satisfactory results.   Strictly speaking, of course, crosstalk between the two chrominance channels occurs for modulation frequencies between 0·34 Mc/s and 0·5 Mc/s (the single sideband region), but it seems that the gain in colour

detail reproduction outweighs the inter-chrominance crosstalk over this range.

9.7. Choice of axes

With equiband operation, we are not concerned with inter-chrominance crosstalk and we can use whichever detection axes are most convenient. This leads to a number of possibilities: for example, detection along $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$ in three separate detectors, or detection along $(R' - Y')$ and $(G' - Y')$, with a simple matrix to derive $(B' - Y')$. There are also XZ detection and symmetrical detection, which are mentioned later.

The $(G' - Y')$ axis can be deduced from the equation

$$0.59(G' - Y') = -0.30(R' - Y') - 0.11(B' - Y')$$

and it is shown in Section 7.6 that detection along an axis which leads the burst by 57°, and with a gain of 0·7, will give the $(G' - Y')$ signal (*see* Fig. 9.11).

The fact that difference signals can be derived without matrixing in equiband operation reduces the total matrix operation to a simple
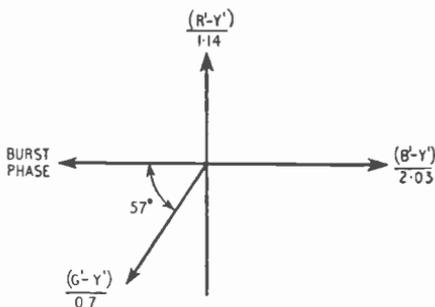
Fig. 9.11. Position of $(G' - Y')$ *detection axis*

addition of Y' to the difference signals. Even this simple matrix can be further economized by detecting at a level which is sufficiently great to enable the display tube to be driven directly from the detector output. This technique is called *high level detection.*

In this application, the three difference signals are fed to the grids of the appropriate displays, and the three display cathodes are strapped together and fed with the $-Y'$ signal. (In practice, the cathodes are normally fed with the appropriate fractions of the Y'

signal to allow for the differing phosphor efficiencies.)   The matrix operation of Y' addition is then carried out in the display tube itself.

9.8. Mixer detection

Another form of synchronous detector which is quite extensively used is a multiplicative type similar to a frequency changer.   In a
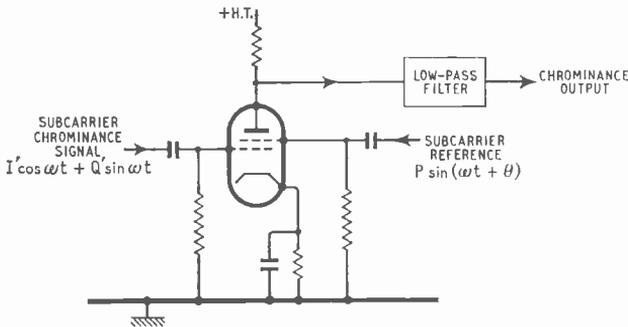
Fig. 9.12.   Mixer detection circuit

conventional mixer, an R.F. signal is fed to one grid of a valve while a local oscillator signal is fed to a second grid, and the anode current is proportional to the product of the amplitude of the two signals and has a frequency component equal to the frequency difference between the signals.

In a mixer type synchronous detector, the R.F. signal becomes the sub-carrier chrominance signal, while the local oscillator signal becomes the reference signal.   This is similar to the synchrodyne process in which the I.F. has zero carrier frequency, and yields the modulation components directly.

Referring to Fig. 9.12, suppose a reference signal $P \sin (\omega t + \theta)$ is applied to one grid of the mixer, while the sub-carrier signal

$$I' \cos \omega t + Q' \sin \omega t$$

is applied to a second grid, the phase θ being measured relative to the Q' axis.   Then the anode current i_a will be proportional to the product

$$(I' \cos \omega t + Q' \sin \omega t) P \sin (\omega t + \theta) = K i_a$$

where K is a constant depending on the valve characteristics.

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Now, using the trigonometric identities

$$\sin A \cos B = \tfrac{1}{2}[\sin(A + B) + \sin(A - B)]$$

and

$$\sin A \sin B = \tfrac{1}{2}[\cos(A - B) - \cos(A + B)]$$

then

$$\mathrm{K}i_a = \frac{P}{2}\left[I' \sin(2\omega t + \theta) + I' \sin\theta + Q' \cos\theta - Q' \cos(2\omega t + \theta)\right]$$

The double frequency terms will be rejected by the low pass filter so that

$$\mathrm{K}i_a = \frac{P}{2}\left[I' \sin\theta + Q' \cos\theta\right]$$

Hence, if the reference phase θ is made $90°$

$$\mathrm{K}i_a = \frac{P}{2}I'$$

while if $\theta = 0°$

$$\mathrm{K}i_a = \frac{P}{2}Q'$$

It follows that the detection axis of the detector is the same as the phase of the reference voltage, so that if an $(R' - Y')$ output is required, for example, then the anode current will be proportional to the $(R' - Y')$ component if the reference phase is made equal to the burst phase minus $90°$.

For $I'Q'$ operation, two detectors operating $90°$ apart are required, while for equiband operation three detectors can be used along the $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$ axes. However, there are several alternative axes which may be used and the particular choice will depend on the type of matrix arrangement which follows the detector. Suitable matrix circuits are discussed under Sections 9.10 and 9.11.

In mixer detectors it should be noted that the output is proportional to the amplitude of the reference signal, so that any change in this during the line scan, for example, would produce a differential detector gain which would show as a saturation change from left to right of the picture.

While the use of a valve with two control grids has been discussed for synchronous detection, it is clear that a triode could be used if one signal (usually the sub-carrier chrominance signal) is applied to

the grid and the second signal (usually the reference) is applied to the cathode.   This technique is described in Section 9.10 on symmetrical detection.

9.9.   High level detection

It is possible to amplify the sub-carrier chrominance signal so that sufficient chrominance output is obtained after detection to drive the display tube.   This method of high level detection can be used
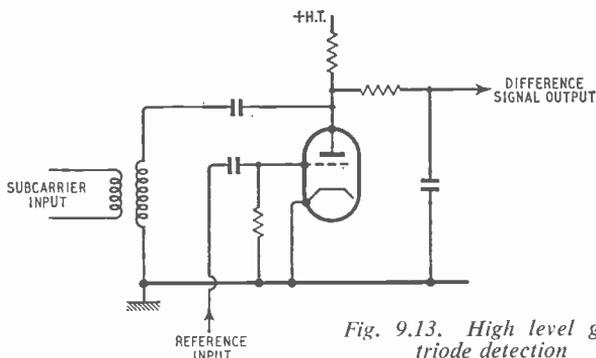
Fig. 9.13.   High level gated triode detection

economically only for equiband working, since considerable loss occurs in $I'Q'$ matrix circuits.

A popular form of high level detector is the gated triode, which is, in effect, a grid controlled rectifier.   The circuit arrangement is shown in Fig. 9.13.

A sufficiently large amplitude of sub-carrier chrominance signal is applied between the anode and cathode of the triode and a fixed phase of reference signal is applied to the triode grid.   The grid time constant is made sufficiently long to bias the triode back by grid current to class C operation, but the grid resistor is kept low so that the triode conducts heavily during the positive peaks of the reference voltage.   At the instants of conduction the anode voltage is low, usually about $+$ 25V, and is virtually independent of the H.T. line voltage, the anode load resistor, and the applied signal.

Now suppose a coloured area having a positive D.C. component of $(R' - Y')$ is being transmitted.   Then the $(R' - Y')$ component of the sub-carrier chrominance waveform will be a sine wave in the $(R' - Y')$ phase whose amplitude is equal to $\dfrac{1}{1 \cdot 14}$ of the D.C. component, and this sine wave, together with any $(B' - Y')$ sine wave

which may also be present, will be applied between the anode and cathode of the triode. If the phase of the reference sine wave on the triode grid is in the $-(R' - Y')$ phase, i.e. 90° ahead of the burst phase, then the triode anode will be clamped to $+ 25V$ at those instants when the $(R' - Y')$ sine wave is at a negative maximum.
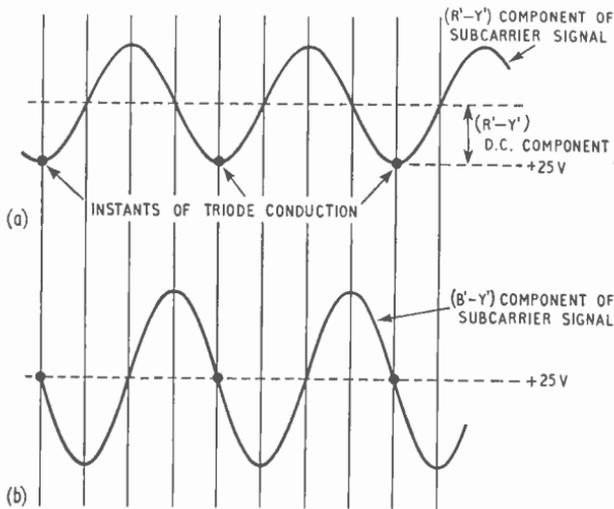
Fig. 9.14. Triode anode voltage waveforms for positive $(R' - Y')$ *and* $(B' - Y')$ *components*

This is shown in Fig. 9·14 (a) where the dots represent the conduction times of the triode.

The anode waveform is therefore a sine wave whose negative peaks are standing on a D.C. level of $+ 25V$. The *RC* filter connected to the triode anode removes the sub-carrier sine wave and leaves a positive D.C. value proportional to the original $(R' - Y')$ signal together with the $+ 25V$ clamping potential.

It is important to notice that any $(B' - Y')$ D.C. components which are present will appear as a sine wave in the anode waveform, but this waveform will be 90° out of phase with the $(R' - Y')$ sine wave so that at the instants of conduction the $(B' - Y')$ component will be zero, as shown in Fig. 9.14(b). Hence, no $(B' - Y')$ component will appear in the $(R' - Y')$ output.

If the chrominance signal contained a negative $(R' - Y')$ component, the detection process would be similar except that the conduction times would occur at the positive peaks of the

$- (R' - Y')$ sine wave reference so that a negative D.C. voltage measured from the $+ 25V$ clamping level would be obtained.   This case is shown in Fig. 9.15.   Notice that the triode anode can go appreciably negative.   For example, if a full saturated cyan is being transmitted, the normalized value of $(R' - Y')$ is $- 0.7$, so that if the display tube requires a drive of 100V, the red difference signal would be $- 70V$.

The detection process which has been described for positive and negative D.C. components applies equally well for any waveform which may be transmitted.   Thus, if the $(R' - Y')$ waveform is of the form shown in Fig. 9.16(a), the $(R' - Y')$ component of the sub-carrier chrominance signal at the triode anode will be as shown in Fig. 9.16(b) which, after filtering, yields the original waveform of Fig. 9.16(a).

It can be seen that the gated triode detector has a detection axis which is the negative of the reference phase applied to the grid.   In
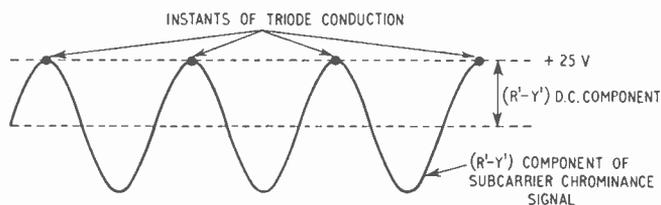
Fig. 9.15. (above): Triode anode voltage waveform for negative (R' − Y') *component.*

Fig. 9.16 (below): (a) transmitted (R' − Y') *waveform; (b)* (R' − Y') *component of triode anode voltage waveform for signal shown in (a)*
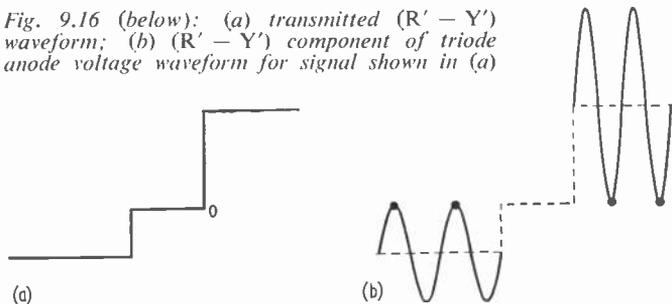
the case we have considered, detection along the $(R' - Y')$ axis is achieved when the reference phase is $- (R' - Y')$.   It follows that if we have three separate detectors fed with phases $- (R' - Y')$, $- (G' - Y')$ and $- (B' - Y')$, then the required red, green and
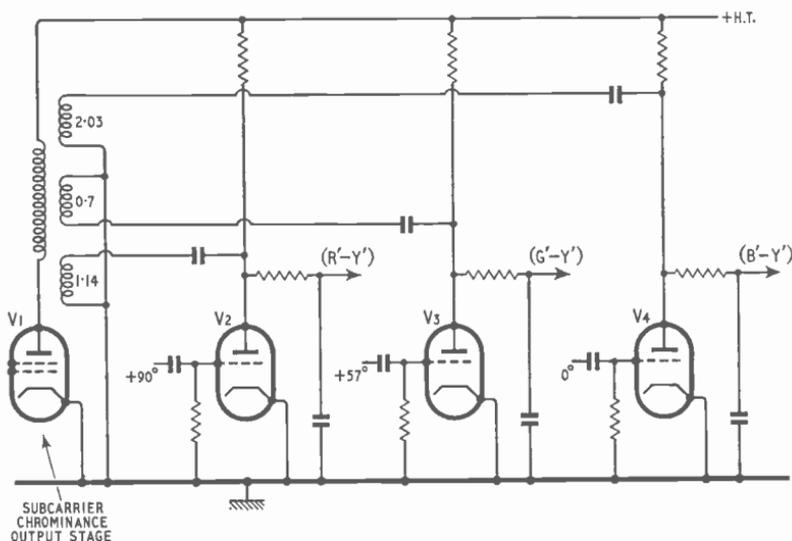
Fig. 9.17. Complete 3 detector high level decoder. Phases of reference signals are measured from the burst phase

blue difference signals will be obtained. The complete circuit for such an arrangement is shown in Fig. 9.17 in which the necessary gains of 1·14, 0·7 and 2·03 for the red, green and blue difference signals, respectively, are derived from suitable windings on the output transformer. The reference phases required are $-(R' - Y')$, $-(G' - Y')$ and $-(B' - Y')$, or $+90°$, $-123°$ and $0°$ measured from the burst phase. The inconvenient $-123°$ phase for the
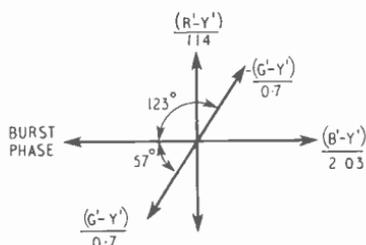
Fig. 9.18. $(G' - Y')$ and $-(G'-Y')$ detection axes

green difference detector can be obviated by reversing the green difference winding and detecting along an axis $+57°$ from burst, as shown in Fig. 9.18.

The circuit shown in Fig. 9.17 provides a very stable arrangement, since the triodes are used as on-off " switches " and the various

proportions of the difference signals are obtained from transformer windings. However, there may be some difficulty in achieving a sufficient drive for the $(B' - Y')$ detector since the gain required in this case is 2·03. The drive requirements can be eased by using $(R' - Y')$ and $(G' - Y')$ detectors and deriving the $(B' - Y')$ signal by means of a simple matrix.
Since

$$0·30(R' - Y') + 0·59(G' - Y') + 0·11(B' - Y') = 0$$

it follows that

$$-(B' - Y') = 2·72(R' - Y') + 5·37(G' - Y')$$

Hence, if we add $(R' - Y')$ and $(G' - Y')$ in the ratio of 0·30 to 0·59, the sum will be proportional to $-(B' - Y')$. If this is
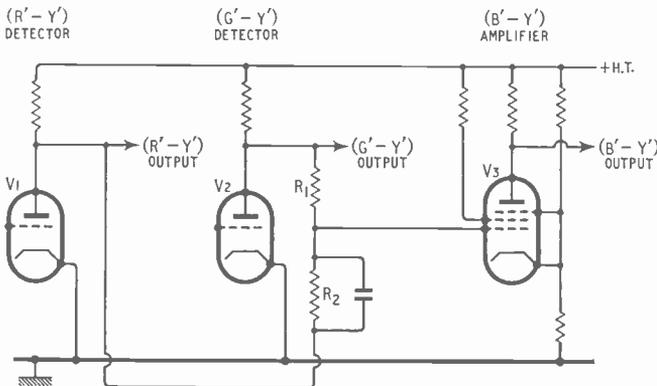
Fig. 9.19.   (B' — Y') *matrix and amplifier*

applied to a simple amplifier, the required phase inversion will be obtained, and the gain of the amplifier can be adjusted so that the correct ratio between the difference signals is achieved.

Fig. 9.19 shows a circuit of a $(B' - Y')$ amplifier which supplies the $(B' - Y')$ output given $(R' - Y')$ and $(G' - Y')$. The matrix equation above is satisfied by adjusting the values of R_1 and R_2 to obtain the correct proportions of $(R' - Y')$ and $(G' - Y')$, and the resulting $-(B' - Y')$ signal is inverted by the amplifier to give the required $(B' - Y')$ output.

In calculating R_1 and R_2, the impedance looking back into the anodes of the detectors must be taken into account (usually this is about 5kΩ), and it is preferable to measure the $(B' - Y')$ output

and adjust R_1 and R_2 until the correct $(B' - Y')$ detection phase is obtained, while disregarding, for the moment, the amplitude of the $(B' - Y')$ output relative to $(R' - Y')$ and $(G' - Y')$. Experimentally, this adjustment for phase can be made by feeding a constant amplitude but variable phase sub-carrier sine wave into the chrominance channel, and noting the phase required to give maximum $(B' - Y')$ output. This phase should be lagging $90°$ on the phase reading which makes the $(R' - Y')$ output a maximum, assuming that the relative $(R' - Y')$ and $(G' - Y')$ phases have been set up correctly.

The correct relative amplitude of $(B' - Y')$ can be obtained by adjustment of the anode load of the $(B' - Y')$ amplifier. The cathode load should be fairly high to give plenty of negative feedback, and for a $Z77$, typical anode and cathode loads would be $25k\Omega$ and $3\cdot3k\Omega$ respectively, for a 400V H.T. line. Notice that the $(B' - Y')$ amplifier is D.C. coupled.

Some improvement in $(B' - Y')$ transient response can be effected by means of a small capacitor across R_2, the larger matrix resistor.

It should be noted that it is usually necessary to arrange for different drive ratios of the difference signals to allow for the display tube phosphor efficiencies, and in the circuit of Fig. 9.17, for example, the $(G' - Y')$ output required may in fact be $0\cdot8(G' - Y')$ rather than $(G' - Y')$. If a matrix is used to derive $(B' - Y')$, it must obviously be designed to include such drive considerations.

A typical high level detection circuit using $(R' - Y')$ and $(G' - Y')$ detectors, with a $(B' - Y')$ matrix amplifier, is shown in Fig. 9.20. This circuit is used in the G.E.C. TTIV series of receivers, and is designed for the 405-line system for drive ratios of 1 to $0\cdot875$ to $0\cdot75$ for R to G to B.

In this circuit, small R.F. chokes are included in the anodes of the $(R' - Y')$ and $(G' - Y')$ detectors to prevent parasitic oscillation during the heavy positive grid drive period, and sub-carrier filtering is by way of $1\cdot5mH$ chokes which resonate at $2\cdot7$ Mc/s. The series $3\cdot3k\Omega$ resistors prevent radiation of sub-carrier harmonics which may otherwise be picked up by the I.F. or R.F. stages of the receiver and be displayed as a coloured herringbone pattern. It must be remembered that the detector anodes have sub-carrier signals of the order of 100V peak applied to them.

The standing D.C. bias for the background controls is injected from potentiometer arrangements via $270k\Omega$ resistors, and the D.C. input resistance to the difference signal outputs is increased by the $150k\Omega$ resistors which are heavily decoupled so as to maintain the
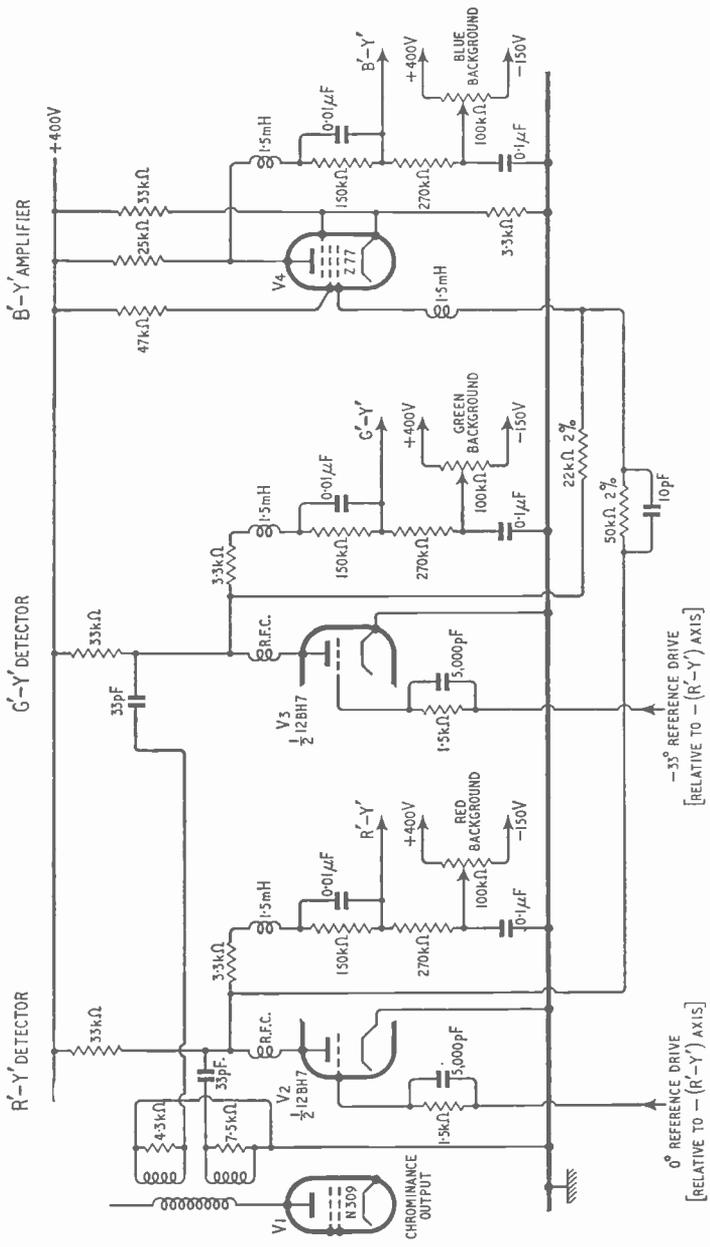
Fig. 9.20. High level detector for 405-line system

frequency response of the outputs. Some difference signal D.C. component loss occurs through the background controls, this D.C. loss being $\left(\dfrac{0\cdot27}{0\cdot27 + 0\cdot15}\right)$ or about 4dB maximum.

Correct clamping action of the detector triodes is assisted by driving the grids with a considerable amplitude (at least 50V peak) of reference signal, and from a low impedance source of 500 Ω or less.

It should be noted that in high level detectors the reference signal must not have a " kink " or slight discontinuity during the burst period, otherwise the clamping action is disturbed and the detector anodes take several microseconds to recover their correct bottoming potential. This effect shows as a coloured shading on the extreme left of the picture, and it must not be confused with the green shading due to the burst which is evident if there is inadequate line flyback suppression of the display.

Another effect to be borne in mind is that if the reference signal is removed from the detectors, their anode potentials will rise, but if a $(B' - Y')$ amplifier is used, its output will not rise proportionately because the matrix circuit is now incorrect since the anode impedances are greater. The white balance is therefore upset. If three-
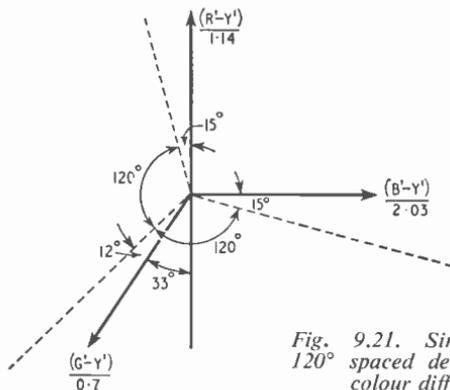
Fig. 9.21. Similarity between 120° spaced detection axes and colour difference axes

difference detectors are used, the main effect of removing the reference signals will be to produce a brightness change. Hence, it follows that reference drive must be applied to the detectors even when the receiver is operating from a monochrome signal, if bias adjustments between monochrome and colour signal operation are to be avoided.

9.10. Symmetrical detection

In symmetrical detection the principle is to take advantage of the fact that on a vector representation of the three colour difference axes $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$, three detection axes may be drawn, as shown in Fig. 9.21, which are 120° apart in phase and
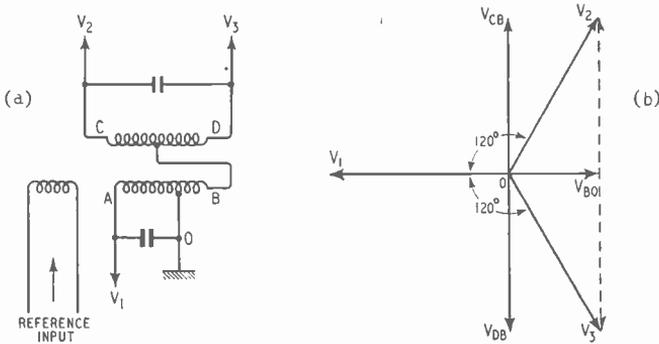
Fig. 9.22.   120° reference drive circuit

each of which does not differ by more than 15° from the phase of a colour difference axis.

Hence, three detectors operating along axes spaced 120° apart may be arranged to give outputs which are very close to the required difference signals, and a simple resistive matrix network may be included to adjust the detector outputs so that the difference signals are actually obtained.

The reference drive for the detectors may be obtained by means of the transformer arrangement shown in Fig. 9.22(a).   The secondary windings AB and CD are loosely coupled and tuned so that the voltage between B and O is 90° out of phase with the voltages between C and B, and D and B, as shown in Fig. 9.22(b).   By connecting B to the centre tap between C and D, the equal magnitude and 120° phased outputs V_1, V_2 and V_3 are obtained provided that

(i) $V_{BO} = 0.5V_1$

(ii) $V_{CB} = V_{DB}$

(iii) $V_{CB} = 0.866V_1$

A suitable symmetrical detection circuit for the American system, designed by the Hazeltine Research Corporation, is shown in Fig. 9.23.   The modulated sub-carrier signal from the chrominance
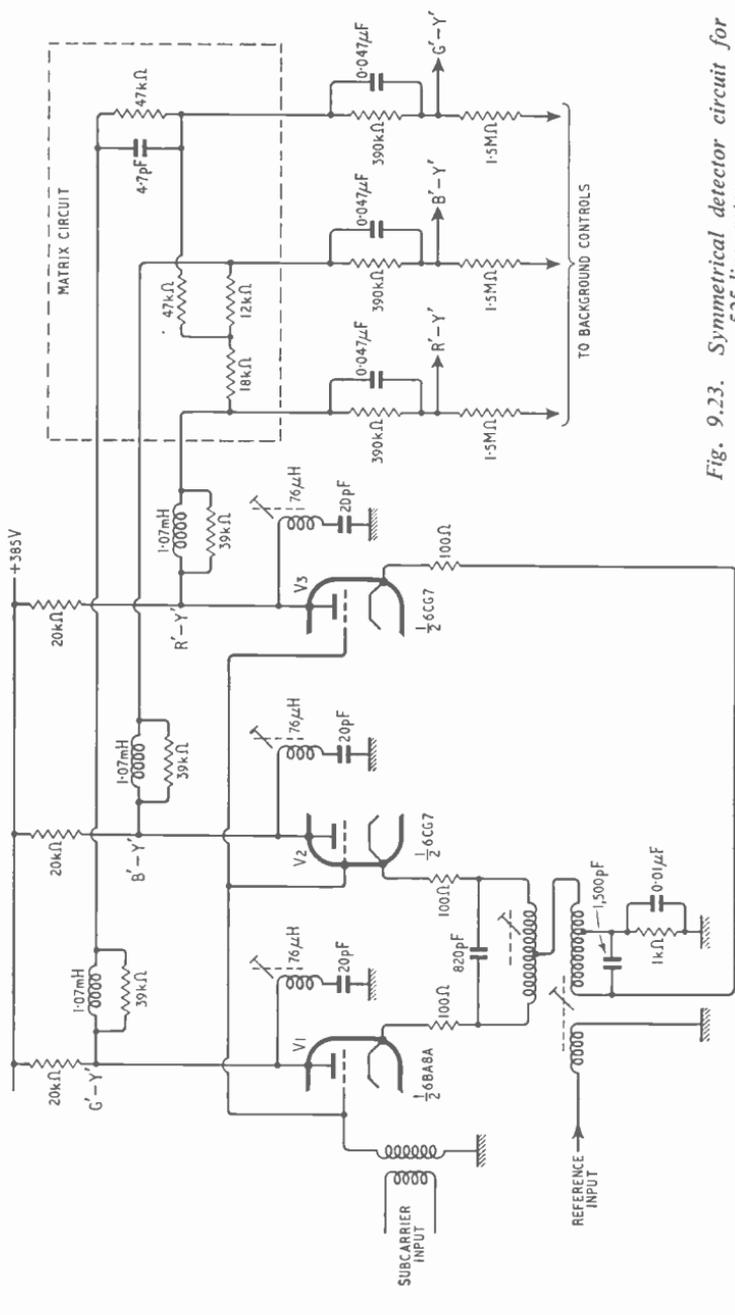
Fig. 9.23. Symmetrical detector circuit for 525-line system

amplifier is fed to each grid of the three triode detectors, while the cathodes of the triodes are returned to earth through the 120° phase shifting transformer.   The 100 Ω cathode resistors are included to stabilize against differences in conductance of the triodes.

Each detector operates along the appropriate axis as shown in Fig. 9.21, and the matrix circuit slightly changes the effective detection axes so that the correct difference signal outputs are obtained. The 4·7pF capacitor across the 47k Ω resistor in the green difference matrix input helps to maintain the green channel frequency response.

Sub-carrier frequency components appearing at the triode anodes are attenuated by series tuned traps, and the series inductance in each output forms a low pass filter in conjunction with stray capacitance and capacitative loading by the traps on one side, and stray capacitance and capacitative loading by the display tube on the other side.

The standing D.C. bias for the background adjustment is applied through 1·5MΩ to each output, and the D.C. input resistance is increased by means of 390kΩ resistors so that the D.C. bias input voltage required to be applied to the 1·5MΩ resistors is not excessive. The A.C. response is maintained by the $0·047\mu$F capacitors which shunt the 390kΩ resistors, but some difference signal D.C. component loss occurs across the 390kΩ resistors.   However, even if the background control resistance were very small, the D.C. component attenuation is only $\left(\dfrac{1·5}{1·5 + 0·39}\right)$ or about 2dB.

Since the reference drive transformer is arranged to give 120° phased outputs, the sub-carrier frequency currents due to any two of the detectors inject an equal but opposite component into the third detector, that is, the symmetry of the reference drive arrangement prevents crosstalk between the detectors.   In practice a small amount of crosstalk is to be expected since the symmetry will not be perfect.

Two advantages of the circuit shown in Fig. 9.23 are, firstly the effective coupling between the detectors is small and secondly the symmetry of the anode circuits of the detectors means that any drift in H.T. line voltage will produce equal changes in each of the difference signal outputs, so that a substantially " brightness only " change will occur in the picture.

9.11.   XZ detection

An interesting and effective decoding method known as XZ detection has been introduced by R.C.A.   The novelty of the circuit lies not so much in the detectors themselves, which are of the normal,

simple low level mixer type as described in Section 9.8, but rather in the choice of detection axes and the matrix arrangement which follows the detectors.   The circuit is intended for equiband detection of the modulated sub-carrier.

In **XZ** detection, two synchronous detectors are fed with the modulated sub-carrier signal and each detector is also fed with a particular phase of reference signal so that one detection axis is
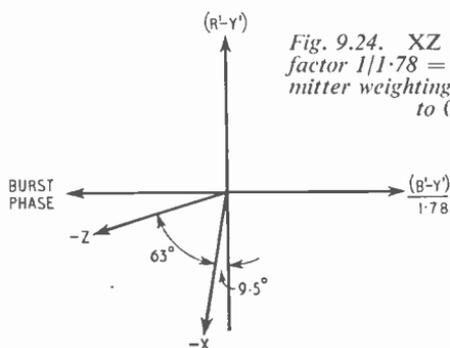
Fig. 9.24. XZ detection axes. The factor $1/1·78 = 1·14/2·03$ is the transmitter weighting of $(B' - Y')$ relative to $(R' - Y')$

$- \mathbf{X}$ and the other is $- \mathbf{Z}$, as shown in Fig. 9.24.   Notice that the output of each detector will consist of a certain proportion of the $(R' - Y')$ and $(B' - Y')$ signals, the proportions being different for the two detectors (e.g. the $- \mathbf{Z}$ detector output will contain relatively more $- (B' - Y')$ signal than will the $- \mathbf{X}$ detector output).   The two detector outputs are fed to a three triode matrix with a common cathode load, as shown in Fig. 9.25.   The triode
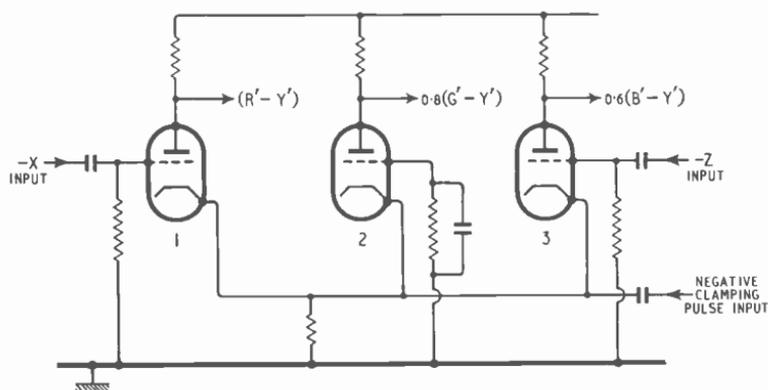
Fig. 9.25.   Basic XZ matrix circuit

arrangement is perfectly symmetrical, the anode loads being the same for each triode as well as the grid circuit time constants.

The grid of the centre triode is returned to earth, so that its cathode voltage and therefore the output from its anode will be proportional to the sum of the signals fed to the triodes 1 and 3.

Now it will be remembered that the green difference signal may be obtained by adding $-(R' - Y')$ and $-(B' - Y')$ in accordance with the equation

$$0 \cdot 59(G' - Y') = -0 \cdot 30(R' - Y') - 0 \cdot 11(B' - Y')$$

so that if the sum of the $-X$ and $-Z$ signals is made to contain
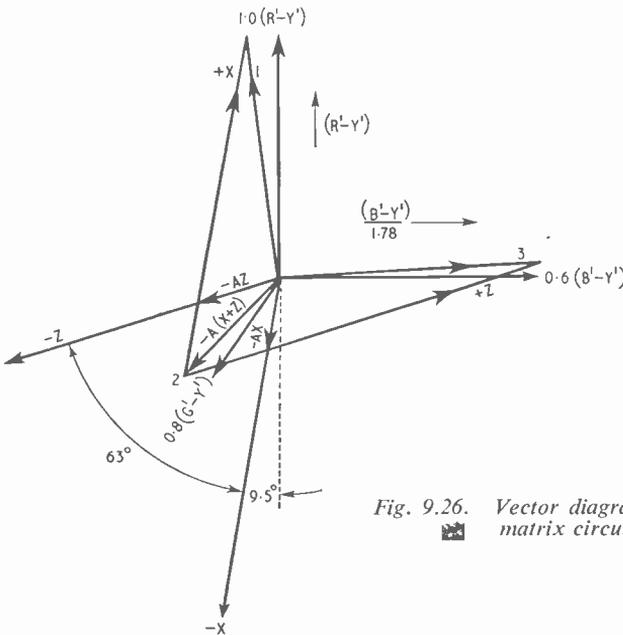
Fig. 9.26. Vector diagram of the XZ *matrix circuit*

$-(R' - Y')$ and $-(B' - Y')$ in the ratio of $0 \cdot 30$ to $0 \cdot 11$, then the output of triode 2 will be proportional to $(G' - Y')$.

Referring to the vector diagram of Fig. 9.26, the vector sum of $-X$ and $-Z$ gives the vector 2 after multiplication by a suitable factor A, which is chosen to give a green difference drive of $0 \cdot 8$ in this case. The vector 2 is proportional to the output of triode 2, and of course it should ideally be coincident with the $0 \cdot 8(G' - Y')$ vector shown. However, the red and blue difference signals are also required and in order to obtain these as well some error in each

difference output must be tolerated since there are not enough variables to satisfy all the required conditions simultaneously, unless the symmetry of the circuit is destroyed.

The grid drive to triode 1 consists of the $-X$ signal minus the cathode voltage $-A(X + Z)$, i.e. the grid drive is $-X + A(X + Z)$, so that the triode 1 output is proportional to the negative of this (because of the phase inversion from grid to anode), which is $X - A(X + Z)$, and is given by the vector 1. Again, ideally this vector should be coincident with the $(R' - Y')$ axis.

Similarly, the grid drive to triode 3 consists of the $-Z$ signal minus the cathode voltage $-A(X + Z)$, so that the triode 3 output is proportional to $Z - A(X + Z)$ and is given by the vector 3.

The X and Z vectors of Fig. 9.26 are chosen so that the outputs of the triodes 1, 2 and 3 are proportional to $(R' - Y')$, $0.8(G' - Y')$ and $0.6(B' - Y')$, the constants 1 to 0·8 to 0·6 being the phosphor efficiency drive ratios. Since the exact pure outputs cannot be obtained, X and Z have been chosen to give least crosstalk in the difference signals. Thus, output 1 consists of $(R' - Y')$ and a small amount of $-(B' - Y')$, output 2 consists of $0.8(G' - Y')$ and a small amount of $-(B' - Y')$, while output 3 consists of 0·6 $(B' - Y')$ and a small amount of $(R' - Y')$.
More exactly

$$\text{Output 1 is } 1.0 [(R' - Y') - 0.07(B' - Y')]$$

$$\text{Output 2 is } 0.8 [(G' - Y') - 0.07(B' - Y')]$$

$$\text{Output 3 is } 0.6 [(B' - Y') + 0.07(R' - Y')]$$

where the 7% crosstalk terms are the result of choosing X and Z to make the crosstalk a minimum.

Calculation of the X and Z axes is given in the next section, where it is shown that crosstalk can be avoided only when the red to blue drive ratio is 1 to 0·367. However, for a red to blue drive of 1 to 0·6, the crosstalk is only 7%, and since in practice the vectors 1, 2 and 3 may be rotated by means of the hue control until correct flesh tone is displayed, the small crosstalk effect is not significant. If desired, a resistance connecting outputs 2 and 3 may be fitted to " pull together " vectors 2 and 3 thereby reducing the crosstalk. Alternatively, crosstalk can be removed by applying a suitable fraction of the triode 3 input to the grid of triode 2, instead of earthing the latter. This technique is detailed at the end of Section 9.12.

It will be noticed that the outputs from the synchronous detectors are A.C. coupled to the triode matrix circuit. This is necessary since

the anodes of the detectors have a D.C. voltage near that of the H.T. line, but obviously some means of D.C. restoration is required. The D.C. components of the chrominance signal are recovered by a clamping technique in which a large negative pulse occurring during the line blanking period is applied to the common cathode connection of the three triodes. This pulse makes the three triodes conduct during the line blanking period so that they take grid current which charges the grid coupling capacitors. The bias on the three triodes is therefore clamped to the same voltage at the end of each line scan so that any D.C. components present in the chrominance signal will be applied to the triode grids during the line scan. The triode anodes are D.C. coupled to the appropriate electrodes of the display tube. In order to ensure correct clamping action, the sub-carrier signal fed to the synchronous detectors should be blanked during the line blanking period to remove the burst and any spurious information. This can be done, for example, by applying a suitable positive pulse to the cathode of the sub-carrier chrominance output stage, so that this becomes inoperative during the clamping period.

This clamping technique has two other advantages besides D.C. restoration. Thus, since each triode is biased by its own grid current during the flyback time, any falling off in emission in a triode will produce less grid current and hence less bias on that triode. Consequently, the gain of any triode tends to remain constant as ageing progresses. Since a large negative pulse is applied to the triode cathodes, the triode anodes are driven very negative during the flyback time so that all three displays are blacked out during retrace, thereby providing line flyback blanking.

The XZ detection circuit used by R.C.A. for the CTC 5N series of receivers for the 525-line system is shown in Fig. 9.27.

The X and Z synchronous detectors are of the mixer type and are fed with the sub-carrier signal and the appropriately phased reference signals. The cathode earth point is via a potentiometer so that the two detector gains may be equalized, and both anode circuits are identical. Some inductance correction is included to maintain the frequency response of the detector outputs to 1·3 Mc/s for 3dB down, so that " wideband equiband " operation is obtained.

Series chokes in the detector outputs attenuate sub-carrier components fed to the matrix and the matrix triode grids are connected to equal time constant circuits. The negative clamping pulse for the triode cathodes is obtained from an amplifier which is fed with a suitable pulse from the line time base output transformer. Since the clamping pulse causes a negative bias to be applied to the triode

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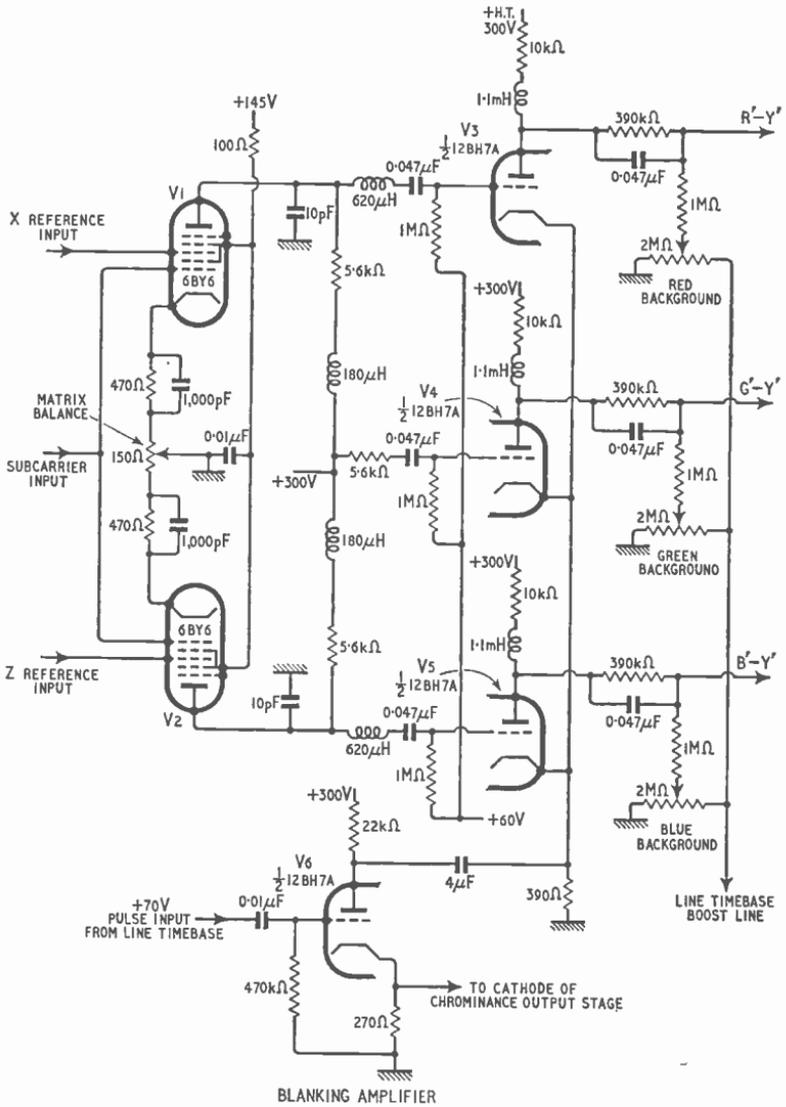
Fig. 9.27. XZ detection and matrix circuit for 525-line system

grids by grid current flow, a suitable positive bias (in this case, $+ 60V$) must be applied to the grid leaks of the triodes so that the correct grid to cathode bias is obtained.   The cathode of the blanking amplifier is connected to the cathode of the sub-carrier chrominance output stage to remove the burst during the clamping period.

Note that although the grid of the green difference triode is " earthy " to A.C. signals, it is connected to the H.T. line through an R.C. circuit of exactly the same value as in the red and blue difference triodes.   Hence, H.T. line variations will not produce
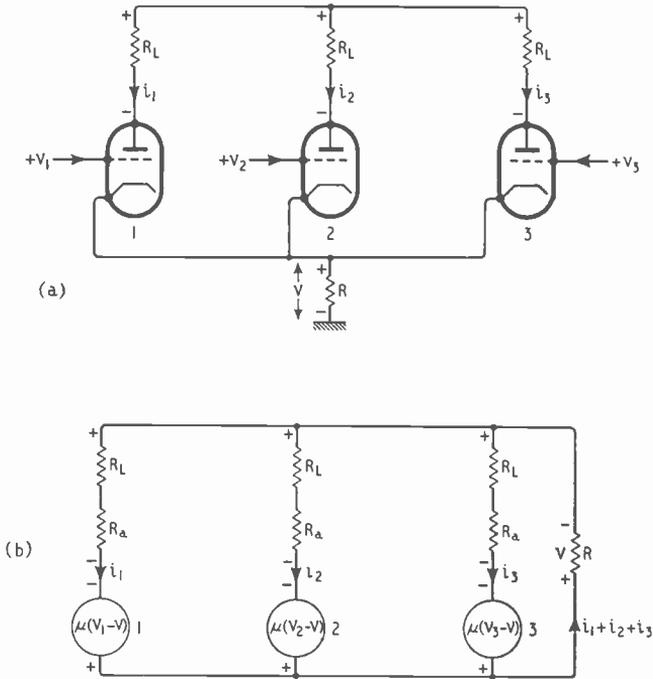
Fig. 9.28.   Three-triode matrix (a) and its equivalent circuit (b)

differential changes between the potentials of the triode anodes and consequently white balance stability will tend to be independent of such variations, which will therefore chiefly affect brightness only.

The three triode anode circuits are identical, and variable D.C. bias for background control is applied through a high impedance circuit to prevent loss of D.C. chrominance components.   The high

value of D.C. supply required as a consequence of this is obtained from the line time base boost line.

9.12.   Calculation of the X and Z axes

In order to calculate the **XZ** axes mentioned in the previous section, the operation of the three triode matrix circuit must first be investigated.   This is shown in Fig. 9.28(a), in which the anode loads R_L are equal, and the common cathode load is R.

Assuming that the three triodes are exactly similar, the equivalent circuit is shown in Fig. 9.28(b) where μ and R_a are the amplification factor and anode resistance of each triode.   Note that although we are interested in the case where v_2 is zero, the symmetry of the circuit and the calculation is preserved if v_2 is not put equal to zero until the final steps of the working.

In the circuit of Fig. 9.28(a), it is required to find i_1, i_2 and i_3 in terms of v_1, v_2 and v_3 and the circuit constants.

Thus the voltage of the equivalent generator of triode 1 is $\mu(v_1 - v)$, and this must equal the voltage drop across R, R_L and R_a.

Hence
$$\mu(v_1 - v) = v + (R_L + R_a)i_1$$

or
$$(R_L + R_a)i_1 = \mu v_1 - (\mu + 1)v \qquad [9.1]$$

Similarly
$$(R_L + R_a)i_2 = \mu v_2 - (\mu + 1)v \qquad [9.2]$$

and
$$(R_L + R_a)i_3 = \mu v_3 - (\mu + 1)v \qquad [9.3]$$

These expressions give the anode currents in terms of the cathode voltage v.   To find v, note that

$$i_1 + i_2 + i_3 = \frac{v}{R}$$

Hence, adding Equations 9.1, 9.2 and 9.3

$$(R_L + R_a)\frac{v}{R} = \mu(v_1 + v_2 + v_3) - 3(\mu + 1)v$$

from which

$$v = \frac{\mu R(v_1 + v_2 + v_3)}{R_L + R_a + 3R(\mu + 1)}$$

Substituting in Equation 9.1

$$i_1 = \frac{\mu}{R_L + R_a}\left[v_1 - \frac{R(\mu + 1)\,(v_1 + v_2 + v_3)}{R_L + R_a + 3R(\mu + 1)} \right]$$

For convenience put

$$A = \frac{R(\mu + 1)}{R_L + R_a + 3R(\mu + 1)}$$

then

$$i_1 = \frac{\mu}{R_L + R_a}\left[v_1 - A\left(v_1 + v_2 + v_3\right)\right]$$

and similarly

$$i_2 = \frac{\mu}{R_L + R_a}\left[v_2 - A(v_1 + v_2 + v_3)\right]$$

and

$$i_3 = \frac{\mu}{R_L + R_a}\left[v_3 - A(v_1 + v_2 + v_3)\right]$$

In this case, the grid of triode 2 is earthed so that $v_2 = 0$. The output voltages are as follows:—

$$\text{triode 1 gives} - \frac{\mu R_L}{R_L + R_a}\left[v_1 - A(v_1 + v_3)\right]$$

$$\text{triode 2 gives} - \frac{\mu R_L}{R_L + R_a}\left[- A(v_1 + v_3)\right]$$

$$\text{triode 3 gives} - \frac{\mu R_L}{R_L + R_a}\left[v_3 - A(v_1 + v_3)\right]$$

Now it is required to find v_1 and v_3, each of which contains both red and blue difference signals, so that $r(R' - Y')$, $g(G' - Y')$ and $b(B' - Y')$ are obtained from the anodes of triodes 1, 2 and 3, respectively, where r, g and b are the phosphor efficiency drive factors. (Typically, $r = 1 \cdot 0$, $g = 0 \cdot 8$ and $b = 0 \cdot 6$.)
Hence

$$- v_1 + A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L}\, r R_D \qquad [9.4]$$

$$A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L}\, g G_D \qquad [9.5]$$

and

$$- v_3 + A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L}\, b B_D \qquad [9.6]$$

where $R_D = (R' - Y')$, $G_D = (G' - Y')$ and $B_D = (B' - Y')$

Now

$$0 \cdot 59 G_D = -0 \cdot 30 R_D - 0 \cdot 11 B_D$$

so that Equation 9.5 may be written as

$$-A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L} g(p R_D + q B_D) \qquad [9.7]$$

where $p = \dfrac{0 \cdot 30}{0 \cdot 59} = 0 \cdot 508$ and $q = \dfrac{0 \cdot 11}{0 \cdot 59} = 0 \cdot 186$.

Equation 9.7 indicates that $v_1 + v_3$ must contain R_D and B_D in the ratio of p to q, that is, 1 to $0 \cdot 367$.   But, adding Equations 9.4 and 9.6

$$- (v_1 + v_3) + 2A(v_1 + v_3) = -(1 - 2A)(v_1 + v_3)$$

$$= \frac{(R_L + R_a)}{\mu R_L}(r R_D + b B_D) \qquad [9.8]$$

which indicates that $v_1 + v_3$ must contain R_D and B_D in the ratio of the drives r to b.   The required outputs cannot therefore be obtained

unless $\dfrac{r}{b} = \dfrac{1}{0 \cdot 367}$.

However, suppose the red to blue drive ratio is required to be 1 to $0 \cdot 6$.   This can be arranged provided that some crosstalk is tolerated. Thus, Equation 9.8 would have an R_D to B_D ratio of 1 to $0 \cdot 6$, while Equation 9.7 would have an R_D to B_D ratio of 1 to $0 \cdot 367$.   To make these two ratios equal (since Equations 9.7 and 9.8 must be satisfied simultaneously), increase the ratio in Equation 9.8 by increasing the R_D content and reducing the B_D content by allowing a crosstalk term βR_D in the B_D output of triode 3, and a crosstalk term $-\rho B_D$ in the R_D output of triode 1.

Then Equation 9.8 becomes

$$(1 - 2A)(v_1 + v_3) = -\frac{(R_L + R_a)}{\mu R_L}\Big[r(R_D - \rho B_D) + b(B_D + \beta R_D)\Big]$$

$$[9.9]$$

that is, $v_1 + v_3$ has an R_D to B_D ratio of $\dfrac{r + b\beta}{b - r\rho}$.   In addition, let us decrease the R_D to B_D ratio in Equation 9.7.   This can be done either by reducing the R_D content or by increasing the B_D content. Since p is greater than q, and since least crosstalk is required, it is better to increase the B_D content.   (For example, $p = 0 \cdot 508$ and

$q = 0.186$. If R_D were reduced by, say, 0.1, the R_D to B_D ratio would be changed from $\dfrac{0.508}{0.186}$ or 2.72, to $\dfrac{0.408}{0.186}$ or 2.2. · But if B_D were increased by 0.1, the R_D to B_D ratio would be changed from 2.72 to $\dfrac{0.508}{0.286}$ or 1.78. That is, a bigger change can be made by increasing B_D rather than by reducing R_D.) Therefore, allowing a crosstalk term $\gamma\, B_D$ in the triode 2 output, Equation 9.7 becomes

$$- A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L}\, g\left[\, 0.508R_D + (0.186 + \gamma)B_D\right] \quad [9.10]$$

which has an R_D to B_D ratio of $\dfrac{0.508}{0.186 + \gamma}$

The R_D to B_D ratios of Equations 9.9 and 9.10 must be equal, so that

$$\frac{r + b\beta}{b - r\rho} = \frac{0.508}{0.186 + \gamma}$$

If the required drive ratios are $r = 1$, $b = 0.6$, then

$$\frac{1 + 0.6\beta}{0.6 - \rho} = \frac{0.508}{0.186 + \gamma}$$

which gives the relation between the crosstalk terms. Obviously, the crosstalk terms are required to be as small as possible, and this will occur when $\rho = \gamma = \beta = x$, say.
Then

$$\frac{1 + 0.6x}{0.6 - x} = \frac{0.508}{0.186 + x}$$

or

$$0.6x^2 + 1.62x - 0.119 = 0$$

whence $x = -2.77$ or 0.07.

Taking the smaller of these two values, it can be seen that the required drive ratios can be obtained provided that the outputs of the triodes are accepted as

$$R_D - 0.07B_D$$

instead of R_D for triode 1

$$- 0.8(0.508R_D + 0.186B_D + 0.07B_D) = 0.8(G_D - 0.07B_D)$$

instead of $0.8G_D$ for triode 2

$$0.6(B_D + 0.07R_D)$$

instead of $0.6B_D$ for triode 3

Equations 9.4, 9.7 and 9.6 therefore become

$$-v_1 + A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L}(R_D - 0\cdot07B_D) \qquad [9.11]$$

$$-A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L} 0\cdot8\ (0\cdot508R_D + 0\cdot256B_D) \qquad [9.12]$$

$$-v_3 + A(v_1 + v_3) = \frac{(R_L + R_a)}{\mu R_L} 0\cdot6(B_D + 0\cdot07R_D) \qquad [9.13]$$

Substituting Equation 9.12 in Equation 9.11

$$v_1 = -\frac{(R_L + R_a)}{\mu R_L}(1\cdot406R_D + 0\cdot135B_D) \qquad [9.14]$$

Substituting Equation 9.12 in Equation 9.13

$$v_3 = -\frac{(R_L + R_a)}{\mu R_L}(0\cdot448R_D + 0\cdot805B_D) \qquad [9.15]$$

Now it is required to obtain v_1 and v_3 by synchronous detection of the sub-carrier. Suppose v_1 is derived by detecting with a gain G_1 along an axis θ_1 from the $-(R' - Y')$ axis, while v_3 is derived
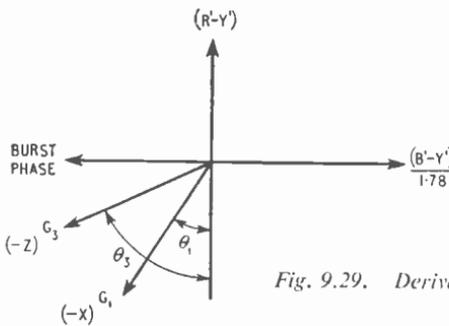
Fig. 9.29. *Derivation of the XZ detection axes*

by detecting with a gain G_3 along an axis θ_3 from the $-(R' - Y')$ axis, as shown in Fig. 9.29. Both detection axes are required to be in the third quadrant, since negative amounts of $(R' - Y')$ and $(B' - Y')$ are required in each case.

The output of detector 1 is

$$-G_1\cos\theta_1 R_D - G_1\sin\theta_1\frac{B_D}{1\cdot78} = -\frac{(R_L + R_a)}{\mu R_L}(1\cdot406R_D + 0\cdot135B_D)$$

$$\therefore \qquad \frac{\sin \theta_1}{1\cdot 78 \cos \theta_1} = \frac{\tan \theta_1}{1\cdot 78} = \frac{0\cdot 135}{1\cdot 406}$$

$$\therefore \qquad \tan \theta_1 = 0\cdot 175, \text{ i.e. } \theta_1 = 9\cdot 5^\circ$$

and

$$G_1 = \frac{1\cdot 406(R_L + R_a)}{\mu R_L \cos \theta_1} = 1\cdot 43 \frac{(R_L + R_a)}{\mu R_L}$$

The output of detector 3 is

$$-G_3 \cos \theta_3 R_D - G_3 \sin \theta_3 \frac{B_D}{1\cdot 78} = - \frac{(R_L + R_a)}{\mu R_L} (0\cdot 448 R_D + 0\cdot 805 B_D)$$

$$\therefore \qquad \frac{\sin \theta_3}{1\cdot 78 \cos \theta_3} = \frac{\tan \theta_3}{1\cdot 78} = \frac{0\cdot 805}{0\cdot 448}$$

$$\therefore \qquad \tan \theta_3 = 3\cdot 2, \text{ i.e. } \theta_3 = 72\cdot 5^\circ$$

and

$$G_3 = \frac{0\cdot 448(R_L + R_a)}{\mu R_L \cos \theta_3} = 1\cdot 49 \frac{(R_L + R_a)}{\mu R_L}$$

Note that the ratio of the detector gains required is $1\cdot43$ to $1\cdot49$, i.e. $0\cdot96$.

The detection axes θ_1 and θ_3 are defined as the $-X$ and $-Z$ axes, respectively.

The constant

$$A = \frac{R(\mu + 1)}{R_L + R_a + 3R(\mu + 1)}$$

defines the green drive according to Equation 9.12. It may be calculated by adding Equations 9.14 and 9.15 and substituting for $v_1 + v_3$ in Equation 9.12. Thus, Equations 9.14 and 9.15 give

$$v_1 + v_3 = - \frac{(R_L + R_a)}{\mu R_L} (1\cdot 854 R_D + 0\cdot 94 B_D)$$

Substituting in Equation 9.12

$$\therefore \qquad (1\cdot 854 R_D + 0\cdot 94 B_D)A = 0\cdot 4064 R_D + 0\cdot 2048 B_D$$

whence

$$A = \frac{0\cdot 4064 R_D + 0\cdot 2048 B_D}{1\cdot 854 R_D + 0\cdot 94 B_D} = 0\cdot 22$$

For a given valve type and anode load, the required value of A can be obtained by a suitable choice of the cathode resistor R.

It should be noted that the values of θ_1 and θ_3 of $9 \cdot 5°$ and $72 \cdot 5°$, respectively, have been calculated above on the assumption that equal percentage crosstalk applies in the three difference signal outputs. An alternative approach is to arrange for the error angles to be equal, so that in Fig. 9.26, the angles between vector 1 and $(R' - Y')$, between vector 2 and $0 \cdot 8(G' - Y')$, and between vector 3 and $0 \cdot 6(B' - Y')$, are all equal. Use of the equation

$$\frac{1 + 0 \cdot 6\beta}{0 \cdot 6 - \rho} = \frac{0 \cdot 508}{0 \cdot 186 + \gamma}$$

enables the equal angular error to be calculated, its value being approximately $7°$. In this case, $\theta_1 = 9°$ and $\theta_3 = 69°$, approximately. These values are very close to the equal percentage crosstalk detection axes of $9 \cdot 5°$ and $72 \cdot 5°$.

It is interesting to note that crosstalk can be avoided completely if a suitable signal is applied to the grid of triode 2. Furthermore, it is possible to arrange for this v_2 signal to be a fraction of the v_3 signal.

Thus, allowing for the case when $v_2 \neq 0$, and assuming drive ratios of $1 \cdot 0$ to $0 \cdot 8$ to $0 \cdot 6$ for r to g to b, Equations 9.4, 9.5 and 9.6, after adding the three equations so that $v_1 + v_2 + v_3$ can be eliminated, become:—

$$\frac{\mu R_L}{R_L + R_a} (1 - 3A)v_1 = (1 - 2 \cdot 405A)(-R_D) + 0 \cdot 45A(-B_D) \quad [9.16]$$

$$\frac{\mu R_L}{R_L + R_a} (1 - 3A)v_2 = (1 \cdot 81A - 0 \cdot 405)(-R_D) + (0 \cdot 9A - 0 \cdot 15)$$
$$(-B_D) \quad [9.17]$$

$$\frac{\mu R_L}{R_L + R_a} (1 - 3A)v_3 = 0 \cdot 595A(-R_D) + (0 \cdot 6 - 1 \cdot 35A)(-B_D)$$
$$[9.18]$$

Now, by a suitable choice of the constant A, it is possible to make the ratio between R_D and B_D in Equation 9.17 the same as that in Equation 9.18.

Solving the equation $\dfrac{1 \cdot 81A - 0 \cdot 405}{0 \cdot 9A - 0 \cdot 15} = \dfrac{0 \cdot 595A}{0 \cdot 6 - 1 \cdot 35A}$

it will be found that $A = 0 \cdot 245$. The root $A = \frac{1}{3}$ is trivial.

Putting $C = \dfrac{\mu R_L}{R_L + R_a}\,(1 - 3A)$ for simplicity, the above equations become:—

$$Cv_1 = 0\cdot41(- R_D) + 0\cdot11(- B_D)$$
$$Cv_2 = 0\cdot04(- R_D) + 0\cdot07(- B_D)$$
$$Cv_3^{\,\prime} = 0\cdot146(- R_D) + 0\cdot27(- B_D)$$

in which it will be seen that

$$0\cdot27Cv_3 = 0\cdot04(- R_D) + 0\cdot07(- B_D) = Cv_2$$

Hence, by applying a fraction $0\cdot27$ of the signal applied to the grid of triode 3, to the grid of triode 2, the required triode outputs of R_D, $0\cdot8G_D$ and $0\cdot6B_D$ are obtained exactly.

For a complete decoder, therefore, two synchronous detectors are required to obtain v_1 and v_3, and a simple resistive matrix to derive v_2 from v_3.

The required detection axes for v_1 and v_3 will be found to be $\theta_1 = 25\cdot5°$ and $\theta_3 = 73°$, as measured lagging on the $-(R' - Y')$ phase.   The relative gains required are $\dfrac{G_1}{G_3} = 0\cdot9$.

9.13. Chrominance signal matrixing

It has been pointed out earlier in this chapter that if $I'Q'$ working is adopted, the matrix operations after detection of the I' and Q' signals are fairly complex.   However, a decoding method has been suggested by the Hazeltine Research Corporation which may simplify the matrix circuits so that difference signal outputs are obtained directly after detection while maintaining $I'Q'$ operation.

In the usual method of sub-carrier demodulation a single sub-carrier signal is compared in phase with two separate reference signals in quadrature.   However, the reverse process may be employed in which two separate sub-carrier signals in quadrature are compared in phase with a single reference signal.   In this case, the two sub-carrier signals may be treated as " I " and " Q " signals so that matrixing can be carried out before the synchronous detection process.

Since the sub-carrier is modulated according to the picture content and therefore has sideband frequencies, the quadrature sub-carrier signals cannot be obtained from simple LC or RC circuits, but if

the modulated sub-carrier is applied to a delay line which has a delay time equal to one quarter of a cycle of the unmodulated sub-carrier frequency, the output will consist of an undistorted but quadrature phase shifted carrier version of the input.

The principle of chrominance signal matrixing is shown in Fig. 9.30.

The modulated sub-carrier signal is fed to one band-pass filter which accepts the sidebands of the I' signal (i.e. $(2 \cdot 66 - 1)$ or $1 \cdot 66$ Mc/s, to 3 Mc/s for the 405-line system) and the same modulated
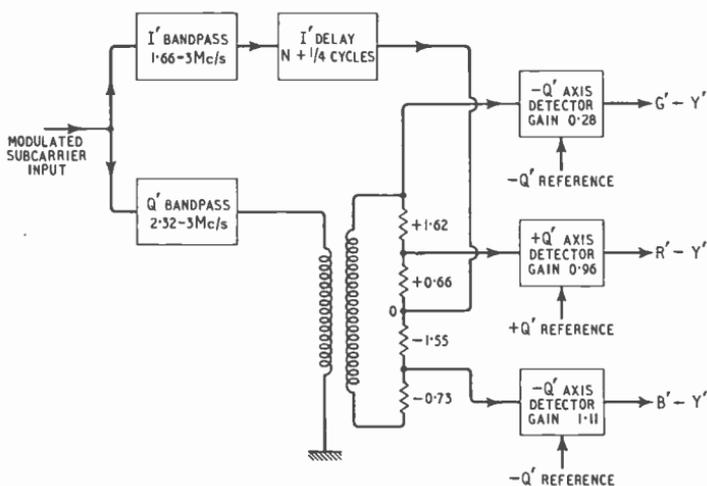
Fig. 9.30. *Pre-detection chrominance signal matrix arrangement for the 405-line system*

sub-carrier signal is also fed to a band-pass filter which accepts the sidebands of the Q' signal (i.e. $(2 \cdot 66 - 0 \cdot 34)$ or $2 \cdot 32$ Mc/s to 3 Mc/s for the 405-line system).

Now the output of the I' bandpass filter is passed through a delay line whose electrical length is equal to an exact number N of sub-carrier cycles plus a quarter cycle, the total delay being equal to that required for the time coincidence of the I' and Q' signals. Note that if this delay line output is fed to a Q' phase detector, the output will be the I' modulation.

Now the red difference signal in terms of I' and Q' is given by

$$(R' - Y') = 0 \cdot 96 I' + 0 \cdot 63 Q'$$

or

$$1 \cdot 04(R' - Y') = I' + 0 \cdot 66 Q'$$

and the blue difference signal is given by

$$(B' - Y') = -1\cdot11I' + 1\cdot72Q'$$

or

$$-0\cdot9(B' - Y') = I' - 1\cdot55Q'$$

Hence, by adding $0\cdot66$ of the output of the Q' band-pass filter to the delayed output of the I' band-pass filter and detecting the result by a Q' phase detector with a gain of 1 to $1\cdot04 = 0\cdot96$, this detector output will be equal to $(R' - Y')$.

Similarly, by adding $-1\cdot55$ of the output of the Q' band-pass filter to the delayed output of the I' band-pass filter, and detecting the result by a $-Q'$ phase detector with a gain of 1 to $0\cdot9 = 1\cdot11$, this detector output will be equal to $(B' - Y')$.

The $(G' - Y')$ signal may be obtained from the equation

$$(G' - Y') = -0\cdot28I' - 0\cdot64Q'$$

or

$$-3\cdot58(G' - Y') = I' + 2\cdot28Q'$$

Thus, the delayed output of the I' band-pass filter is added to $2\cdot28$ of the Q' band-pass filter output and the result is detected by a $-Q'$ phase detector having a gain of 1 to $3\cdot58 = 0\cdot28$, to give $(G' - Y')$.

In Fig. 9.30, the appropriate proportions of the Q' band-pass filter output required to be added to the I' band-pass filter output are shown, the additional quantity $0\cdot73$ being included to balance the Q' channel load.

It will be appreciated that the I' channel delay line must be cut sufficiently accurately to give the odd quarter cycle delay for the quadrature phase shift. One way of doing this in practice is to use two sections of line (both having the same characteristic impedance, of course), one section having less delay per unit length so that it may be used as a " fine control " of delay.

9.14. Summary

In this chapter we have seen that chrominance information may be recovered from the modulated sub-carrier signal by synchronous detection along the appropriate axes. This entails a " sampling " of the sub-carrier signal by a reference sine wave locked in frequency and phase to the transmitter's sub-carrier oscillator, the relative phase of the reference determining the detection axis.

After synchronous detection, some form of matrix circuit is required so that R', G' and B' signals can ultimately be fed to the display tube. If full use is made of the transmitted information,

synchronous detection along the I' and Q' axes is required, and a matrix circuit is necessary to convert the I' and Q' signals to the difference signals $(R' - Y')$, $(G' - Y')$ and $(B' - Y')$. The final matrix operation of adding Y' to the difference signals is usually carried out in the tube itself by feeding the difference signals to the appropriate grids of the tube, and a $- Y'$ signal to the cathodes.

An alternative decoding procedure is to detect synchronously along two or three detection axes, the bandwidths of the detected signals being made equal and rather larger than the Q' bandwidth. This " wideband equiband " technique results in some crosstalk between the higher frequency chrominance signals, but the benefit of the increased definition compared with equiband Q' bandwidth operation outweighs the undesired effects. Equiband working allows a free choice of detection axes so that the difference signals can be obtained directly from the synchronous detectors without a matrix operation, or alternatively the axes may be chosen in such a manner that a simple matrix circuit suffices to supply the required difference signals, as in symmetrical and XZ detection.

Several synchronous detector and matrix circuits have been described, and while these have been labelled as suitable for either 405-line or 525-line N.T.S.C. systems, the basic circuits are suitable for any system provided the frequency limiting components of the circuit are designed for the appropriate bandwidths.

Colour Receiver
Reference Frequency Generators

10.1. Introduction

Synchronous detection of the modulated sub-carrier signal has been described in Chapter 9, and there it was pointed out that, in order to demodulate a quadrature modulated signal, a C.W. source of sub-carrier frequency must be available which is locked in frequency and phase to the unmodulated sub-carrier at the transmitter. To enable colour receivers to generate this reference, a short sample of unmodulated sub-carrier is transmitted during the line blanking period, and this is usually referred to as the colour burst synchronizing signal. The device in the colour receiver which generates a C.W. sub-carrier signal from the burst signal is called a reference frequency generator.

The television engineer who is not yet familiar with reference frequency generators is often understandably alarmed by the function that they have to perform. Since vital hue information is carried by the phase of the modulated sub-carrier signal, it obviously follows that any spurious phase changes in the C.W. reference signal will give rise to spurious hue changes in the reproduced picture. Further, the burst signal has a very low duty ratio since only some $3 \cdot 5 \mu s$ of the transmitter's sub-carrier is transmitted only once during each $100 \mu s$ of line scan (in the 405-line system), and this would appear to suggest that any spurious signals due to noise or interference would have ample opportunity to produce hue errors in the picture.

These doubts about the practicability of generating a C.W. sub-carrier source of sufficient stability from such a short synchronizing burst have been voiced by critics of the N.T.S.C. system, and it is only fair to add that the authors themselves were originally among these critics. However, the reader may rest assured that, provided reasonable attention is paid to layout and design, the generation of a C.W. reference from the burst signal is perfectly practicable even under extremely adverse signal-to-noise ratio

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conditions. In fact, satisfactory synchronization can be obtained even when the signal-to-noise ratio is below usable level for the monochrome signal.

The fundamental principle which permits the generation of a sufficiently pure reference from the burst signal is integration. The longer the integration time, the better will be the purity of the reference output signal, and theoretically this process can be continued indefinitely. However, one practical limitation is the stability of the sub-carrier source at the transmitter, since any integrating device at the receiver must not integrate so effectively that changes in the transmitter's sub-carrier frequency cannot be accommodated.

This particular limitation is not usually of any consequence because extreme frequency stability is normally achieved by the transmitting authorities. A more realistic practical limitation is economy in the colour receiver circuit.

For those readers who are not familiar with the mechanism of integration processes, an analogy may be drawn by way of a simple experiment in photography. Suppose that a synchronized sine wave is displayed on an oscilloscope and suppose sufficient output from a random noise generator is added so that the displayed sine wave appears to be " lost " in the noise background. If a long time exposure photograph, with a corresponding adjustment of aperture, is now taken of the display, it will be found that the sine wave is quite easily visible. The longer the exposure (i.e. the longer the integration time) the more visible the sine wave will become. In effect, the random noise integrates up to give a uniform background, while the sine wave, being non-random, integrates up to give its original waveform. Note that, if, during the exposure, the frequency of the sine wave were varied, the result would be spoilt. This is the equivalent of having a longer integration time than is warranted by the stability of the transmitter's sub-carrier frequency.

The integration performed by a reference generator may be expressed in terms of a frequency response characteristic. Thus, if a signal is integrated over a time T_m, then any noise fluctuations present which have a frequency greater than $\frac{1}{T_m}$ will not contribute significantly to the output since their mean value will be near zero. Hence, only fluctuations with frequencies less than $\frac{1}{T_m}$ will be significant. Therefore integration over a time T_m is equivalent to passing

Effect of phase error of reference generator on reproduced hue;
(above) *phase delay;* (below) *phase advance*

The front view of a colour monitor with the cabinet removed. The lower photograph shows the controls at the bottom of the set

the signal and noise through a filter having a pass bandwidth of $\frac{1}{T_m}$. Since the required output from a reference generator is of sub-carrier frequency, the integration can be carried out by a filter having a pass bandwidth of $\frac{1}{T_m}$ with a centre frequency equal to the sub-carrier frequency.
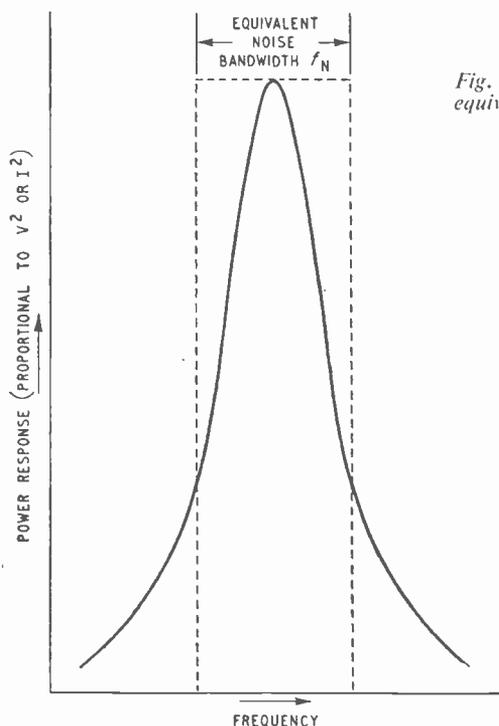
Fig. 10.1.   Definition of
equivalent noise bandwidth
of a filter

The bandwidth $\frac{1}{T_m}$ is called the *noise bandwidth*, f_N, and therefore

$$\frac{1}{T_m} = f_N$$

The equivalent noise bandwidth of a filter may be found by plotting the square of the voltage output against frequency, and then drawing a rectangular pass band characteristic having the same peak value and enclosing the same area with the frequency axis, as the physical response. The equivalent noise bandwidth

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is then the width of the rectangular response, as shown in Fig. 10.1. The square of the voltage output is used in order to obtain equivalence on a power basis.

The reference frequency generator is basically an integrator. It accepts a noisy sine wave signal and provides an output which is that same sine wave but with much reduced noise content. It is therefore equivalent to a narrow band-pass filter tuned to the subcarrier frequency.

10.2. Colour burst signal

The colour burst synchronizing signal for the 405-line British adapted N.T.S.C. system is shown in Fig. 10.2.

From the point of view of the reference frequency generator, the significant factors are:—

The peak-to-peak amplitude of the burst is equal to that of the line sync waveform.

The burst commences about $1\cdot5\mu s$ after the back porch of the line sync pulse, and lasts for not more than $3\cdot75\mu s$.

The burst is omitted during the 4 lines of field sync pulses.

The frequency of the burst is $2\cdot6578125$ Mc/s, ±8 c/s, with a maximum rate of change of $0\cdot1$ c/s^2.
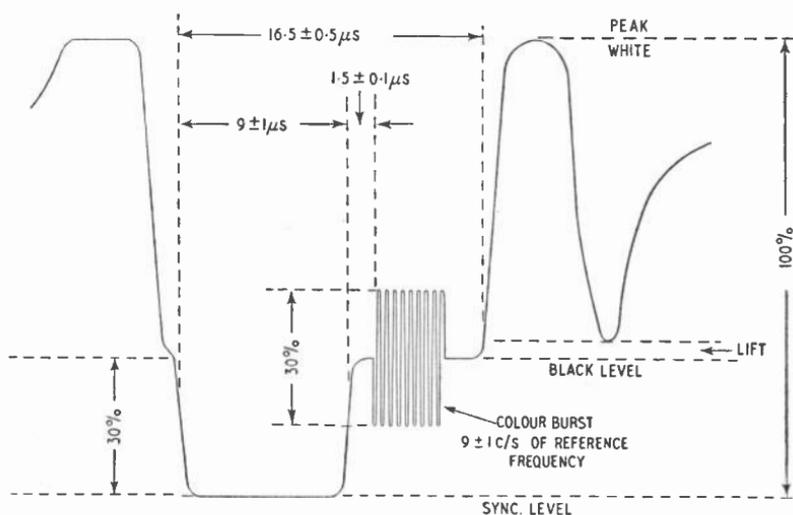

Fig. 10.2.   Colour burst synchronizing signal

It is important to realize that the colour bursts are samples of a continuous sine wave of sub-carrier frequency. That is, at the transmitter a continuous sine wave of sub-carrier is taken and applied to a gating circuit which selects the appropriate number of cycles once every line, except during the field sync period. Note that there is no phase specified for the start of the burst. This may vary as the phase of the divided down line frequency varies relative to the sub-carrier frequency. However, no matter where the time position of the line sync pulses may be relative to the burst, and no matter what the starting phase of the burst may be, it is always possible to draw a continuous sine wave starting with any one burst, which will pass through, and be exactly coincident with, all the other bursts.

Since the sub-carrier frequency is locked to an odd multiple of half the line frequency, the starting phase of the burst will change by 180° every line. Thus, an oscilloscope locked to line frequency will display the burst (or any other sub-carrier component) as two sine waves 180° apart, i.e. interlaced. On the other hand, if the oscilloscope is triggered once every other line, only a single sine wave will be displayed.

It must be emphasized that the 180° phase change in the start of the burst, and which occurs every line, is due to the fact that the time interval between the gates contains an integral number plus a $\frac{1}{2}$ cycle of sub-carrier, as a result of the frequency relationship between the sub-carrier and line frequencies.

A Fourier analysis of the burst waveform is given in Appendix 1, where it is shown that it is equivalent to a sine wave of sub-carrier frequency together with sidebands separated from it by intervals of the line frequency. The two nearest sidebands have practically the same amplitude as the sub-carrier component.

10.3. Specification of reference generator performance

Since the phase of the modulated sub-carrier relative to the burst is a function of the hue of the colour being transmitted, it is obvious that any phase changes between the output of the reference generator and the burst will produce hue errors in the reproduced colour. It therefore follows that the largest phase locking error between the reference output and the burst should ideally be such that the corresponding hue error is just not noticeable.

Neglecting the effects of noise for the moment, subjective tests have shown that while some people are capable of observing hue errors corresponding to only about $\frac{1}{2}°$ of phase error, most people

will accept the range of hue errors corresponding to phase errors of approximately $\pm 5°$. However, the authors are of the opinion that reference generators should be designed to have a phase error not greater than $\pm 2\frac{1}{2}°$ (although to date the transmitter tolerance may be as large as $\pm 10°$), to provide a reasonable safety factor. This type of phase error is called the *static phase error*. It is the maximum possible value of the phase error between the reference output and the burst.

When noise is present with the burst signal, the reference output phase will be modulated by it to give a *dynamic phase error*, as it is called. It is usual to specify random noise, since this will inevitably be present, and it is additionally the most difficult type to reject.

In a severe case, random noise in the burst signal applied to a correctly designed reference generator will show as a low frequency colour flutter on the reproduced picture. It is customary to specify the noise performance of a reference generator in terms of the r.m.s. value of the phase fluctuations obtained with a specified signal to random noise ratio. Thus, a recommended noise performance may be quoted as a $5°$ r.m.s. dynamic phase error for a unity signal-to-noise ratio of peak to peak burst to r.m.s. noise.

This figure of $5°$ r.m.s. dynamic error for a unity signal-to-noise ratio was originally suggested as a reasonable criterion, and experience has shown that it is, in fact, realistic. With such a noisy signal, flutter effects due to phase errors in the reference output are subjectively unimportant compared with the other effects present, such as time base synchronization and brightness fluctuations. For a given dynamic phase error and for a given signal-to-noise ratio, it is possible to calculate the integration time T_m and therefore the equivalent noise bandwidth

$$f_N = \frac{1}{T_m}$$

which is required. For the British system, this works out to be 140 c/s. This is discussed in Appendix 2.

A further condition which should be satisfied by a reference generator is that it should have a reasonable stabilization period, so that when switching from one channel to another only a short time is required for synchronization. A stabilization period of 1 second may be regarded as satisfactory, while one of about 3 seconds is tolerable and one of 10 seconds is poor.

A final desirable feature of a reference generator is that it should provide a reasonably constant output amplitude for all levels of

signal input, and for all possible settings of the hue control.   However, the constancy of output amplitude required will depend on the type of synchronous detector used in the decoder.

10.4.   Burst gating

The function of the reference generator is to provide a continuous sine wave output locked to the burst and to the burst only. Since, in the chrominance signal, other sub-carrier components are present which have various phase values depending on the hues of the transmitted colours, it is necessary to separate the burst from these components.   The selection of the burst to the exclusion of all other signals can be carried out by means of a suitable gating waveform which, in effect, switches the sub-carrier signal into the reference generator only at those instants when the burst is present.

Thus, a typical composite video signal is shown in Fig. 10.3(a) while the chrominance component is shown in Fig. 10.3(b).   If a gating waveform such as Fig. 10.3(d) can be generated, this can be
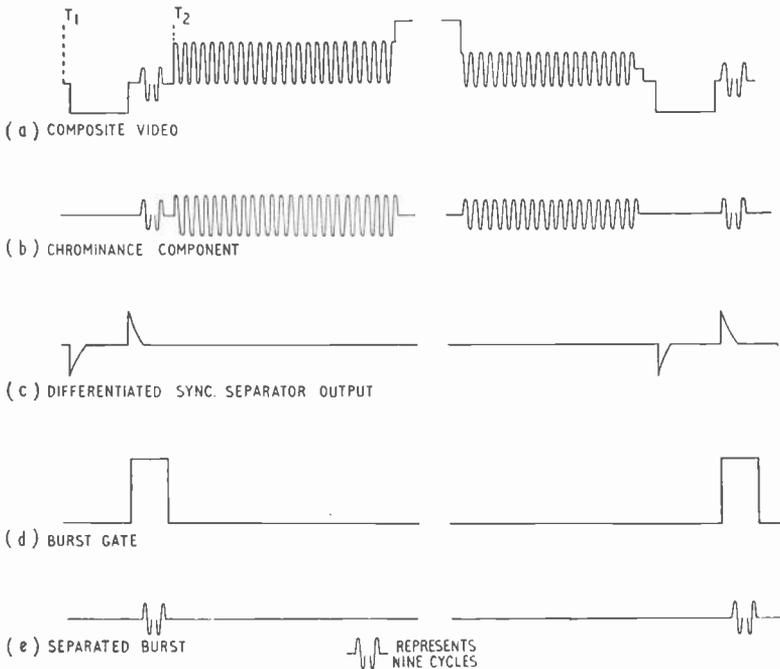

(a) COMPOSITE VIDEO

(b) CHROMINANCE COMPONENT

(c) DIFFERENTIATED SYNC. SEPARATOR OUTPUT

(d) BURST GATE

(e) SEPARATED BURST   REPRESENTS NINE CYCLES

Fig. 10.3.   Burst gating waveforms

made to switch an amplifier fed with the waveform of Fig. 10.3(a) or Fig. 10.3(b) to give the burst only output of Fig. 10.3(e).

In order to exclude unwanted sub-carrier components, the burst gate should open not earlier than the start of the pre-sync porch T_1 and should close not later than the end of the post-sync porch T_2. However, the longer the gate is open, the greater will be the noise allowed through, and therefore the ideal gate should open immediately before the burst and close immediately after it. If the burst gate width exceeds the burst width by a factor of r to 1,

Fig. 10.4.   Gate generator circuit

and the required noise bandwidth of the reference generator is f_N, then the generator itself must be designed to have a noise bandwidth of $\dfrac{f_N}{\sqrt{r}}$. Nevertheless, it is advisable to have a slight excess of gate width to allow for any relative time shift between the gate and the burst. It is essential to ensure that the gate closes before the start of the active line, otherwise disturbing hue errors will result which will vary with the hue of colours in the extreme left of the picture.

A very satisfactory method for generating the burst gating wave-form is to develop a pulse of the appropriate width by means of a monostable multivibrator which is triggered by the trailing edge of the line sync waveform. A convenient source for the latter is the anode of the conventional sync separator, where negative going mixed sync is available.

Differentiation of the sync separator output results in positive going " pips " which are coincident in time with the trailing edges of the line sync waveform, as shown in Fig. 10.3(c). Application of these " pips " to the monostable multivibrator of Fig. 10.4 results in the output shown in Fig. 10.3(d) where the length of the gating pulses can be adjusted by the time constant in the grid circuit of V_2.

The operation of the circuit of Fig. 10.4 is as follows:—

With no signal input, V_2 is biased positively relative to V_1 so that V_2 is conducting and holds V_1 non-conducting by way of cathode feedback. When a positive " pip " is applied to the grid of V_1, V_1 conducts and a negative pulse from its anode renders V_2 non-conducting. However, since the V_1 anode is not D.C. coupled to the V_2 grid, the conduction period of V_1 lasts only for a time determined by the time constant in the grid circuit of V_2. When the V_2 grid has reached a sufficiently positive value, V_2 conducts again and renders V_1 non-conducting. The circuit then remains stable until the next positive " pip " triggers V_1.

The use of this type of gate generator ensures that the gate opens before the start of the burst, in fact, about $1 \cdot 5 \mu s$ before. However, some delay can be incorporated by reducing the frequency response to the grid of V_1 by means of a series resistor, and this usually leads to about $1 \mu s$ between the opening of the gate and the start of the burst.

It is recommended that the time constant in the V_2 grid be adjusted to give a gate width of about $5 \mu s$ so that the longest burst (10 cycles) can be accommodated. For the most unfavourable condition of an 8 cycle burst, this gives an excess burst gate to burst period ratio of $1 \cdot 66$ to 1, so that the reference generator should be designed for a noise bandwidth of $0 \cdot 775 f_N$ to achieve an effective overall noise bandwidth of f_N.

It will be appreciated that the gate generator of Fig. 10.4 will produce two gates per line during the four lines of field sync pulses, where the burst signal is absent. This gives an excess burst gate width of $210 \cdot 5$ to $202 \cdot 5$, or $1 \cdot 04$, which increases the noise bandwidth by a factor of $1 \cdot 02$, which is not significant. However, if the reference generator is at the extreme end of its pull-in range, the lack of burst during the field sync period can lead to hue errors at the top of the picture.

An alternative method for generating the burst gating waveform is to derive it from the receiver's line time base. Thus, a suitable winding on the line time base output transformer can be arranged to supply a pulse of $+100V$ or so during the flyback time. If such

a pulse is applied to an integrating circuit of about $3\cdot3k\Omega$ and 100pF, the resulting pulse waveform can be used as a burst gate.

While derivation of the gate from the line time base is an economical technique, it suffers from certain disadvantages.   For example, the gate width is normally relatively long to allow for variations in the synchronization time of the time base, and of course the locking of the time base can affect the performance of the reference generator.

The amplitude of the gating waveform required will depend on the level of the chrominance signal existing where the gate is applied. The amplitude must be large enough to exclude the largest chrominance excursions (full saturated red and cyan) from the reference generator.   Thus, if gating is carried out in the final stage of the burst amplifier, the burst input amplitude may be about 5V peak so that the largest chrominance signal would be about 13V peak. The gate amplitude should therefore exceed the cut-off bias by at least 13V, leading to a value of about 20V peak.   A minimum gate amplitude of 30V or so usually allows for a satisfactory safety margin.

The gate waveform should ideally have a flat top or " on " period, and its leading and trailing edges should not have excessively fast rise and decay times otherwise spurious rings will result which may simulate the burst signal.   Note that the gating waveform applied to the gated valve will produce a gate waveform in the anode circuit, and by differentiation " pips " corresponding to the leading and trailing edges of the gate will appear with the burst waveform.

The luminance signal is another source of spurious signals which can simulate the burst.   Thus, the leading and trailing edges of the line sync waveform, and the leading edge of the line blanking when there is a white on the extreme left of the picture, all consist of step waveforms which occur close to the burst signal.   Now, if a receiver has a relatively narrow I.F. pass-band (with corresponding compensation in the chrominance amplifier response) then the sync and blanking edges can produce spurious rings having a frequency close to that of the sub-carrier component of the burst.   These spurious rings are often called sync or video *widgets*.   Note that in a monochrome transmission in which the field frequency is mains locked, a mains frequency change of only about $0\cdot1$ c/s can produce a widget having a frequency exactly equal to the sub-carrier.   (For example, the 263rd harmonic of the line frequency for a field frequency of $49\cdot9$ c/s in the 405-line system is only about 260 c/s from the sub-carrier frequency.)

Colour " unkilling " and spurious locking due to widgets can be avoided by having a wide I.F. pass-band and an accurately positioned and correct width burst gate.   (*See* Appendix 1.)

10.5.   Burst amplification

Most reference generators require a burst input amplitude of the order of 50V peak, and this amount of burst is not normally available in the receiver's chrominance channel.   The usual practice is therefore to provide an amplifier which accepts the chrominance signal from a suitable point, amplifies and gates it, and then feeds its output to the reference generator.

The amount of gain required from a burst amplifier will naturally depend on the burst take off point in the chrominance channel, but it is unlikely to exceed 50dB, in which case the burst input amplitude would be about 150 mV peak.

It should be noted that there are conflicting requirements as far as the position of the burst take off point is concerned.   On the one hand, the take off point should be from a point as close as possible to the synchronous detector input to minimize phase shift variations between the reference signal and the chrominance sub-carrier.   On the other hand, the burst should be taken as remote as possible from the synchronous detector to prevent the reference signal (which is necessarily present, usually with large amplitude, at the synchronous detector) being present with the burst, so deteriorating the pull-in performance of the reference generator.   In the authors' experience this latter requirement is by far the more significant, and it is good practice to take the burst from at least one stage before the synchronous detectors.

In order to reduce the noise content of the burst signal which is finally applied to the reference generator, the gating operation should be carried out in the last stage of the burst amplifier.   This eliminates any noise which may otherwise be introduced by post gating stages in the periods between bursts.   Further, since the final burst amplifier stage may often be a heavy current valve, H.T. current drain is conserved since the valve is switched " on " for only about 5 % of the time.

It is important to realize that although the reference generator itself is a very narrow pass-band filter, the burst amplifier should have an appreciable bandwidth up to the gated stage.   Thus, it is good practice to design for a bandwidth equal to about 40 times the line scan frequency (0.4 Mc/s in the British system) and centred about the sub-carrier frequency.   The reason for the relatively large

bandwidth is to ensure that the majority of the burst information does actually reach the reference generator. Thus, the burst waveform can be analysed into Fourier components, and the power content of any number of harmonics in the vicinity of the sub-carrier frequency can be calculated and compared with the total power in the burst waveform. It will be found that the power content of the sub-carrier component and the nearest 20 harmonics on each side of it is approximately 85% of the total
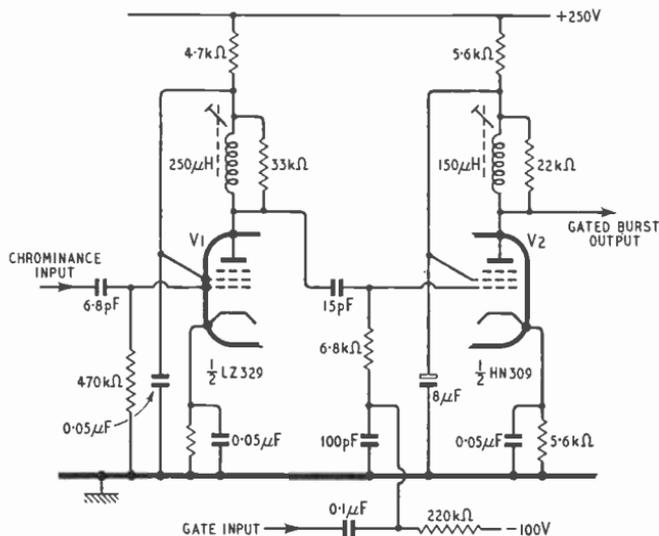

Fig. 10.5. Gated burst amplifier circuit

power in the burst waveform. Hence, for a burst amplifier bandwidth of 0·4 Mc/s (0·6 Mc/s in the American system) prior to the application of the burst gate, the burst waveform is practically undistorted so that most of the burst is included during the gating interval. (Fourier analysis of the burst waveform is described in Appendix 1.)

In the case of reference generators which are literally passive narrowband filters, there is obviously no point in maintaining the bandwidth of any post gated stage; but for A.P.C. loops, or any generator employing a phase detector, the bandwidth should be maintained even after gating to ensure that the maximum signal is available for phase control.

A typical two stage burst amplifier is shown in Fig. 10.5. The circuit is quite conventional, the anode loads being simple tuned

circuits tuned to the sub-carrier frequency, with suitable damping to give the required bandwidth of 0·4 Mc/s. Note, however, that because of the small duty ratio of the burst, the input impedance of the phase detector which is fed from the burst amplifier output is relatively low. The burst gate is applied to the lower end of a 6·8$k\Omega$ resistor connected to the grid of V_2, and some sub-carrier decoupling (100pF) is included to prevent sub-carrier components entering the gate generator.

If the gate is generated from the back edge of the line sync, as described in Section 10.4, it is advisable to apply a negative bias to the V_2 grid, as shown. The object of this bias is to cut off V_2 if mixed sync signals are absent, otherwise the amplifier would be " open " all the time and may accept spurious signals or reference output which would not have been significant during the short gating periods.

Note that an LZ329 valve and an HN309 valve can be used for both gate generation and burst amplification, to give a compact unit.

10.6.   Classification of reference generators

The preceding sections have described the performance required of a reference generator, and also the principles of amplifying and selecting the burst signal to the exclusion of other sub-carrier components.   The reference generator itself is therefore presented with the burst only waveform of Fig. 10.3(e), from which it must generate a continuous sine wave locked to the burst sine wave, that is, it must " fill up the gaps " between bursts.

Broadly speaking, reference generators fall into two classes:—

(i) *Passive integrators* which select the sub-carrier frequency component of the burst waveform by means of a narrow pass-band filter.   Note that it is not sufficient for the filter to merely exclude the sidebands on either side of the sub-carrier frequency component. Its pass-band must be narrow enough to give an equivalent noise bandwidth of about 140 c/s (in the British system) so that dynamic phase errors due to noise are subjectively acceptable.

(ii) *Dynamic integrators* which perform the filtering operation by means of a low pass filter to which is fed a beatnote signal formed by heterodyning an oscillator output with the burst signal.   The conventional dynamic integrator takes the form of an oscillator whose output is frequency and phase controlled by a phase detector which compares the oscillator and burst signals.   This arrangement is usually called an *A.P.C. loop*, that is, an automatic phase control

loop. Like the passive integrator, the dynamic integrator should have a frequency characteristic equivalent to a noise bandwidth of 140 c/s centred about the sub-carrier frequency.

Normally, the reference generator performs other functions in addition to the primary one of reference frequency generation.
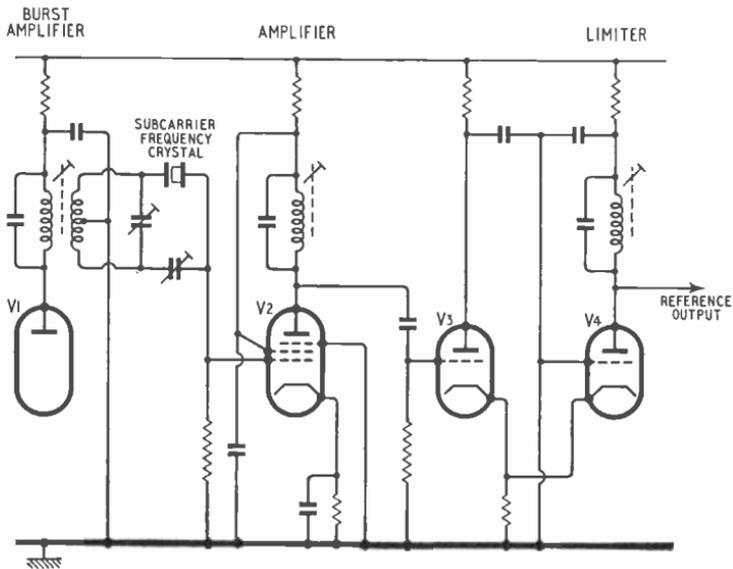

Fig. 10.6.   Circuit for a passive integrator

These include colour killing and A.C.C. or automatic chrominance control.

For reasons which will be discussed later, reference generators of the A.P.C. type are almost always used in preference to passive integrators.   However, each type will be described in the following sections.

10.7. Passive integrators

A simple passive integrator circuit is shown in Fig. 10.6.

The gated burst signal is fed to a bridge circuit which includes a quartz crystal nominally tuned to the sub-carrier frequency.   The object of the balanced arrangement is to neutralize the effect of the mounting capacitance which is unavoidably present across the crystal, thereby preventing feed through of the burst sidebands.

The crystal rings at its resonant frequency in response to the sub-carrier component of the burst waveform, and the resulting voltage is amplified.   Since the crystal voltage has an exponential decay during the period between bursts, the amplified voltage is passed to a limiter which provides a constant amplitude of sine wave output.

The crystal Q which is needed in this circuit is obviously a function of the noise bandwidth required, and it can be shown that the equivalent noise bandwidth of a simple tuned circuit is $\frac{\pi}{2}$ of its 3dB bandwidth.   Hence, if the resonant frequency is f_o c/s, the 3dB bandwidth is $\frac{f_o}{Q}$ c/s, and the noise bandwidth is

$$f_N = \frac{\pi}{2} \frac{f_o}{Q} \text{ c/s}$$

Hence

$$Q = \frac{\pi}{2} \frac{f_o}{f_N}$$

For satisfactory noise performance, f_N should be about 140 c/s for the British system, so that the Q required is

$$Q = \frac{\pi}{2} \times \frac{2 \cdot 66 \times 10^6}{140} \simeq 30{,}000$$

if it is assumed that the crystal behaves as a simple tuned circuit.

This high value of Q accounts for the use of a crystal rather than an LC circuit, and of course, crystals can be obtained which have the required Q.

There are severe limitations to this type of integrator if it is designed for the above noise bandwidth.   The difficulty is that, since the circuit Q is so high, a small change in either the crystal tuning or the transmitted sub-carrier frequency can cause very serious static phase errors.   The static phase error has two components.   One is produced by the phase shift which is caused by slightly detuning a high Q circuit and the other arises from the slightly " out of step " effect of the crystal sine wave, which accumulates during the line period before the next burst arrives.   These two sources of static phase error may be calculated as follows:

The equivalent circuit of the crystal itself is shown in Fig. 10.7(a), where L, C and r are the series components of the crystal, and S is the mounting capacitance.   Now a parallel resonance of

the crystal will occur when the applied frequency is slightly higher than the series resonant frequency, so that the series L, C components are then equivalent to an inductance l which resonates with the capacitance S, as shown in Fig. 10.7(b).

The impedance of the circuit of Fig. 10.7(b) is given by

$$Z = \frac{r + j\omega l}{1 - \omega^2 lS + j\omega Sr}$$

But if $\frac{\omega_0}{2\pi}$ is the parallel resonant frequency

$$lS = \frac{1}{\omega_0{}^2}, \qquad l = \frac{Qr}{\omega_0}, \qquad S = \frac{1}{\omega_0 Qr}$$

Substituting

$$Z = \frac{r + j\frac{\omega}{\omega_0} Qr}{1 - \frac{\omega^2}{\omega_0{}^2} + j\frac{\omega}{\omega_0 Q}}$$

and it is the phase of Z which is of interest, since variation of the phase of the voltage across the circuit relative to the burst current fed to it will determine the phase error. Hence, the phase θ_1 of Z,

(a)   (b)
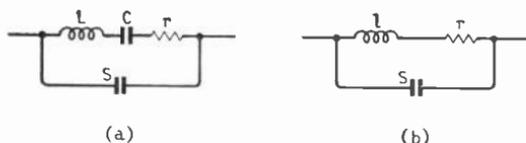
Fig. 10.7.   Equivalent circuit of a quartz crystal

which is the phase of the voltage across the circuit relative to the total current through it, is the same as the phase of the vector

$$\left[\left(1 - \frac{\omega^2}{\omega_0{}^2}\right) - j\frac{\omega}{\omega_0 Q}\right]\left(1 + j\frac{\omega}{\omega_0} Q\right) = 1 + j\frac{\omega}{\omega_0}\left[Q\left(1 - \frac{\omega^2}{\omega_0{}^2}\right) - \frac{1}{Q}\right]$$

$$\simeq 1 + j\frac{\omega}{\omega_0} Q\left(1 - \frac{\omega^2}{\omega_0{}^2}\right)$$

since $Q \gg 1$

Therefore, if f_s is the applied sub-carrier frequency, and f_0 is the parallel resonant frequency, the above vector becomes

$$1 + j\frac{f_s}{f_0}Q\left(1 - \frac{f_s{}^2}{f_0{}^2}\right)$$

Now suppose there is a tuning error Δf so that $f_o = f_s + \Delta f$, then

$$\frac{f_s}{f_o} = \frac{f_s}{f_s + \Delta f} = \frac{1}{1 + \dfrac{\Delta f}{f_s}} \simeq 1 - \frac{\Delta f}{f_s}$$

and

$$\frac{f_s{}^2}{f_o{}^2} \simeq 1 - \frac{2\Delta f}{f_s}$$

The above vector then becomes

$$1 + jQ\left(1 - \frac{\Delta f}{f_s}\right)2\frac{\Delta f}{f_s} \simeq 1 + j2\,\frac{\Delta f}{f_s}\,Q$$

so that

$$\tan \theta_1 = 2Q\frac{\Delta f}{f_s}$$

or

$$\theta_1 = \tan^{-1} 2Q\frac{\Delta f}{f_s}$$

This error θ_1 is the first of the phase shifts mentioned above, and it may be caused either by a tuning error of the crystal or by a change in the transmitter's sub-carrier frequency.

The " out of step " phase error may be calculated as follows:—

If the sub-carrier frequency is $\frac{M}{2}$ times the line frequency ($M = 525$ in the British system) and if there are 8 cycles of burst per line, then between bursts there are $\left(\frac{M}{2} - 8\right)$ cycles of sub-carrier.   Each cycle lasts for a time of $\frac{1}{f_s}$ seconds, so that the time between bursts is $\frac{(M - 16)}{2f_s}$ seconds.

Now if there is a crystal tuning error such that

$$f_o = f_s + \Delta f$$

the number of " crystal " cycles in the time

$\frac{(M - 16)}{2f_s}$ seconds is $\left(\frac{M - 16}{2f_s}\right)(f_s + \Delta f)$

However, the number of " genuine " sub-carrier cycles during this time is $\dfrac{(M-16)}{2}$, so that the total error at the start of a burst is

$$\left(\frac{M-16}{2f_s}\right)(f_s + \Delta f) - \left(\frac{M-16}{2}\right) = \frac{\Delta f}{2f_s}(M - 16) \text{ cycles}$$

Since each cycle is a phase of 2π radians, this phase error is

$$\theta_2 = 2\pi\frac{\Delta f}{2f_s}(M - 16) \text{ radians}$$

This expression may be simplified since $M - 16 \simeq M$ and $f_s = \dfrac{M}{2}f_L$ where f_L is the line frequency.

Thus approximately

$$\theta_2 = \frac{2\pi\Delta f M}{f_L M} = 2\pi\frac{\Delta f}{f_L} \text{ radians}$$

The total phase error of the passive integrator is therefore

$$\theta = \theta_1 + \theta_2 = \left(\tan^{-1} 2Q\frac{\Delta f}{f_s}\right) + 2\pi\frac{\Delta f}{f_L}$$

Since $2Q\dfrac{\Delta f}{f_s}$ will normally be small, this expression may be further simplified to

$$\theta \simeq 2Q\frac{\Delta f}{f_s} + 2\pi\frac{\Delta f}{f_L} \text{ radians}$$

or

$$\theta \simeq \frac{360}{\pi}\Delta f\left(\frac{Q}{f_s} + \frac{\pi}{f_L}\right) \text{ degrees}$$

Now if θ is not to exceed $2\cdot5°$, then Δf must not exceed about 2 c/s for a Q of 30,000, i.e. a noise bandwidth of 140 c/s.

Since the transmitter frequency tolerance is ± 8 c/s, the required static phase accuracy cannot be guaranteed for a noise bandwidth of 140 c/s.

Of course, if the Q is reduced, a larger value of Δf can be tolerated. Thus, if

$$\Delta f = 20 \text{ c/s}$$
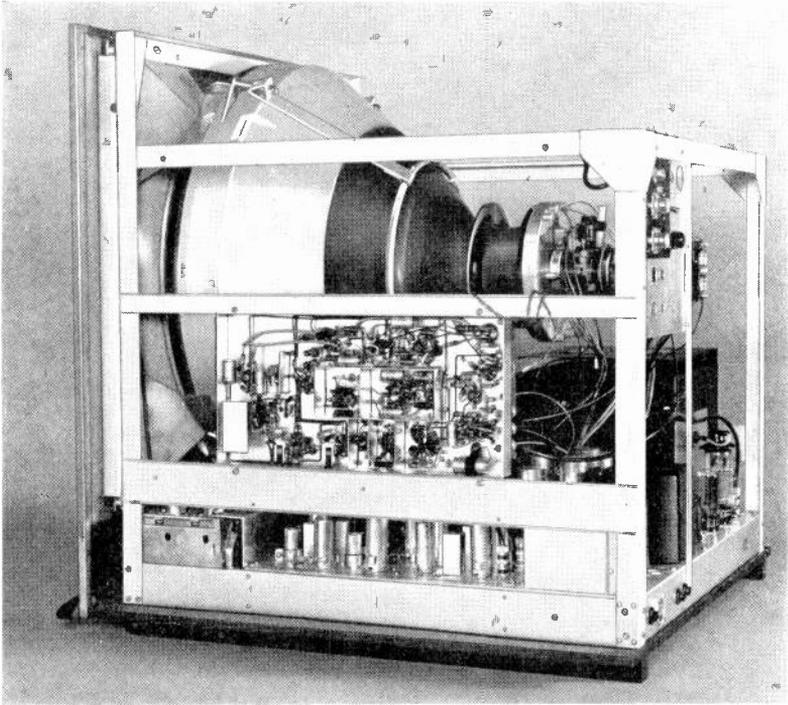
The inside of a colour monitor showing the underside of the I.F. deck and sync. separator. The two lower units are the tuner and reference oscillator

A typical complete 21-inch shadow mask receiver of a type which could go into production.   This type of receiver will accept either N.T.S.C. colour transmissions or monochrome signals

the value of Q for a static phase shift of $2 \cdot 5°$ is about 2,100, which gives a noise bandwidth of

$$\frac{\pi}{2} \times \frac{2 \cdot 66 \times 10^6}{2,100} = 2 \text{ kc/s}$$

This corresponds to a dynamic phase error of about $17°$ r.m.s. for a peak-to-peak burst to r.m.s. noise ratio of unity, whereas the desirable dynamic error is only $5°$ r.m.s.

The conflicting requirements for small dynamic and small static phase errors cannot therefore be simultaneously satisfied. A compromise may be taken, of course, or alternatively an automatic device may be included in the circuit to correct the static phase error, as shown in Fig. 10.8.

In the circuit of Fig. 10.8, the integrator is designed to have the required noise bandwidth regardless of static phase error. The integrator output is then compared in phase with the burst
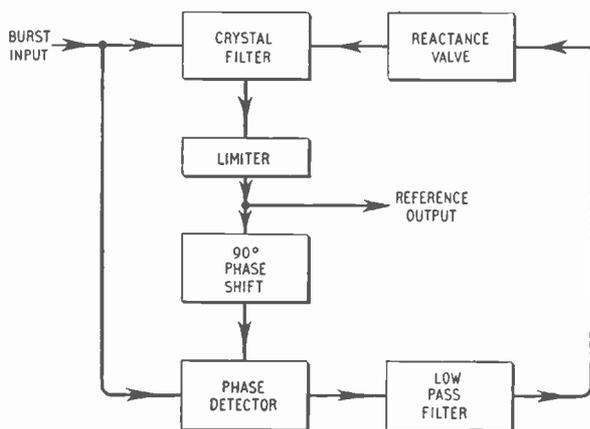
Fig. 10.8.   Passive integrator with static phase correction

signal in a phase detector whose output corrects the tuning of the integrator by means of a reactance valve, thereby reducing the static phase error. A phase shift of $90°$ must be included in one of the sub-carrier feeds to the phase detector, since phase detectors usually give a zero or equilibrium output when the two signals being compared are $90°$ apart in phase. (Phase detectors and reactance valves are discussed in Appendices 5 and 6). By including sufficient gain in the feedback loop, the static phase error may be

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reduced to the required figure, and provided the frequency pass band of the loop is small compared with the noise bandwidth, the noise performance of the integrator is not impaired.

It should be noted that in crystal filters or ringing circuits, the output voltage decays exponentially between bursts by a factor $e^{-rt/2l}$ where t is the time measured from the end of a burst. Since $\dfrac{r}{l} = \dfrac{\omega_0}{Q}$, the decay factor may be written as $e^{-\omega_0 t/2Q}$, where $\omega_0 = 2\pi \times 2 \cdot 66 \times 10^6$ radians/sec, and the maximum value of t is approximately 100×10^{-6} sec. The decay factor is therefore $e^{-266\pi/Q}$, and this has a value of $0 \cdot 9$ for a Q factor of 8,350. Hence a 10% drop in amplitude occurs during the line period for a Q of 8,350. For a Q of 30,000, the drop in amplitude during the line period is only about 3%. However, since some synchronous detectors are sensitive to any amplitude change of the reference signal, a limiter after the crystal is advisable to prevent colour changes across the picture.

Passive integrators have not proved popular in practice because of the requirement for high Q and accurately cut crystals. To
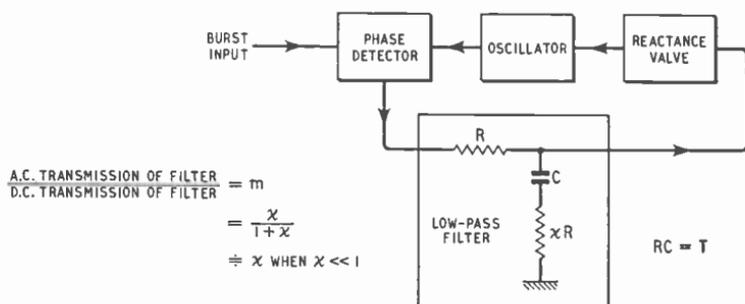
Fig. 10.9. A.P.C. loop block diagram

overcome the static phase error, a phase loop is necessary (unless a compromise is taken between static and dynamic phase errors) and the circuit then becomes strictly comparable with an A.P.C. loop from the point of view of the number of components.

10.8. Dynamic integrators or A.P.C. loops

A block diagram of a simple A.P.C. loop is shown in Fig. 10.9. The output of a stable oscillator nominally tuned to the sub-carrier

frequency is fed to a phase detector where its phase is compared with the burst waveform. The resulting beatnote output of the phase detector is then passed through a low pass filter to a reactance valve which controls the oscillator frequency.

The design of A.P.C. loops must cater for two conditions; the loop must first pull-in to frequency and phase lock, and having locked, the characteristics of the loop must be such that variations of the incoming burst phase (due to noise or interference) do not cause large variations of the oscillator output phase. Since the formulae which describe the in-sync and pull-in performance of A.P.C. loops require more advanced mathematics for their derivation, formulae will merely be quoted in the following sections. However, a full account of A.P.C. loops is given in Appendices 3 and 4 and phase detectors and reactance valves are described in detail in Appendices 5 and 6.

10.9. In-sync performance

Let us first consider the in-sync performance of the loop. Two conditions must be satisfied. First, the static phase error must not exceed a certain value, such as $\pm 2\frac{1}{2}°$. Second, the noise performance must be such that the dynamic phase error does not exceed about 5° r.m.s. for a peak-to-peak burst to r.m.s. noise ratio of unity. This corresponds to a noise bandwidth of about 140 c/s for the British system.

Before discussing the static and dynamic phase errors, it is necessary to define the various parameters which determine the loop performance. These are as follows:—

The sensitivity of the reactance valve, which is usually called β and is measured in c/s/V. Thus, one volt applied to the reactance valve grid produces a frequency change of the oscillator of β c/s. In reference generators the reactance valve is usually arranged to " look like " a capacitor. That is, a positive D.C. voltage increase on the grid reduces the frequency.

The sensitivity of the phase detector, which is usually called μ and is measured in terms of volts/radian. The meaning of μ may be illustrated by feeding a sine wave voltage of one frequency to one side of the phase detector, and a sine wave voltage of a different frequency to the other side. The phase detector output under these conditions is a sine wave having a frequency equal to the frequency difference between the signals being compared (i.e. the beatnote frequency) and a peak amplitude of μ volts. If the beatnote frequency is ω radians/sec, the phase detector output is $\mu \sin \omega t$, and if this is plotted against ωt this sine wave has a slope of

μ volts/radian at the origin. (The slope of a sine wave at the origin is numerically equal to the peak value of the sine wave.)

The usual type of phase detector used in A.P.C. loops gives an output which is proportional to the cosine of the phase difference between the two signals being compared. The detector output is therefore zero for a 90° phase difference, and it is usual to consider phase errors in terms of the angle ϕ by which the phase difference differs from 90°. That is, if there is a phase error ϕ, the detector output is $\mu \cos(90 - \phi)$, i.e. $\mu \sin \phi$.

The final parameter to be considered is the transfer characteristic of the loop filter. This is the ratio output voltage/input voltage of the filter at any frequency ω radians/second, and may be called $F(\omega)$. For the filter shown in Fig. 10.9

$$F(\omega) = \frac{1 + j\omega xT}{1 + j\omega(1 + x)T}$$

where $T = RC$

This reduces to $\dfrac{x}{(1 + x)}$ as $\omega \to \infty$, when it is called the A.C./D.C. gain ratio, m.

10.10. Static phase error

If the oscillator has a tuning error of Δf c/s, then a voltage $\dfrac{\Delta f}{\beta}$ must be applied to the reactance valve to correct it. This voltage must come from the phase detector, that is

$$\mu \sin \Delta\phi = \frac{\Delta f}{\beta}$$

where $\Delta\phi$ is the static phase error.

Hence

$$\sin \Delta\phi = \frac{\Delta f}{\mu\beta} = \frac{\Delta f}{f_c}$$

The quantity f_c is the product $\mu\beta$, and is usually called the D.C. loop gain. It represents the maximum hold-in range of the loop. The loop cannot generate the required correction voltage to tune the oscillator if the oscillator error is greater than f_c.

Since $\Delta\phi$ is usually required to be not more than $\pm 2\frac{1}{2}°$, a reasonable approximation for $\Delta\phi$ is

$$\Delta\phi = \frac{\Delta f}{f_c} \text{ radians} = \frac{180}{\pi} \frac{\Delta f°}{f_c} = \frac{180}{\pi} \frac{\Delta\omega°}{\omega_c}$$

where $\Delta\omega = 2\pi\Delta f$ and $\omega_c = 2\pi f_c$.

10.11. Dynamic phase error

The dynamic phase error may be investigated by assuming a noise-free input signal whose phase is changed sinusoidally at a particular frequency and noting the amplitude of the corresponding output phase change.   By repeating this process at different frequencies, a plot of output phase change divided by input phase change may be made against frequency.   This ratio is called the *phase transfer ratio*, usually written $Q(\omega)$, and it is analogous to the voltage transfer ratio of a filter.   In fact, it is shown in Appendix

Voltage transfer characteristic $\dfrac{V_2}{V_1}$

$$= \frac{1/j\omega C + xR}{1/\omega_c C + j\omega(1 + x) R/\omega_c + 1/j\omega C + xR}$$

$$= \frac{1 + j\omega xT}{1 - (1 + x)\omega^2 T/\omega_c + j\omega(xT + 1/\omega_c)}$$
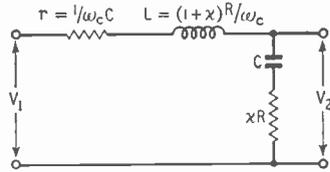
Fig. 10.10.   Equivalent circuit for the phase transfer ratio of the A.P.C. loop shown in Fig. 10.9

3 that, for the loop filter of Fig. 10.9, the phase transfer ratio of the loop is the same as the voltage transfer ratio of the circuit of Fig. 10.10.   The noise bandwidth of the loop may be found by squaring the amplitude of the $Q(\omega)$ curve, and finding the equivalent rectangular bandwidth which encloses the same area.

In Appendix 3 it is shown that

$$\frac{f_N}{2} = f_{NN} = \frac{\omega_c \left[1 + \dfrac{x^2 T \omega_c}{(1 + x)} \right]}{4 (1 + x T \omega_c)} = \frac{\omega_c (1 + m x T \omega_c)}{4 (1 + x T \omega_c)}$$

where f_{NN} is the noise semi-bandwidth.

Note that the loop does not distinguish between noise fluctuations which are above or below the sub-carrier frequency, so that the area under the $Q(\omega)$ squared curve leads to the noise semi-bandwidth.

For the British system, $f_N = 140$ c/s so that $f_{NN} = 70$ c/s, for 5° r.m.s. dynamic error and unity peak-to-peak burst to r.m.s. noise ratio.

It may be wondered why the resistor xR is included in the loop filter.   It is included to enable the static phase error and the noise semi-bandwidth to be chosen independently.

Thus, if xR is zero (i.e. short circuit)

$$f_{NN} = \frac{\omega_c}{4}$$

But the static phase error is

$$\Delta\phi \simeq \frac{\Delta\omega}{\omega_c}$$

so that

$$f_{NN} = \frac{1}{4} \times \frac{\Delta\omega}{\Delta\phi}$$

or

$$f_{NN}\Delta\phi = \frac{\Delta\omega}{4}$$

Thus, for a given tuning error $\Delta\omega$, the $\Delta\phi$ and f_{NN} cannot be independently chosen.
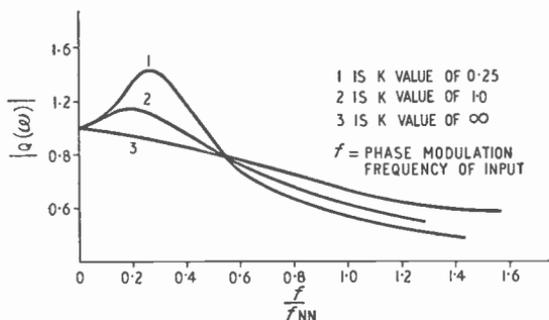

Fig. 10.11. Phase transfer characteristics for various K values

However, by including xR, ω_c can first be chosen to give the required figure for $\Delta\phi$, and then two variables, that is, x and T, are available for making f_{NN} the required figure. Further, since both x and T are available, one of these may be used to control another characteristic of the loop. Notice that the equivalent circuit of the loop, shown in Fig. 10.10, has L, C, R elements so that ringing could occur under certain conditions. It is shown in Appendix 3 that this ringing is a function of a damping factor K defined by

$$K = \frac{(1 + xT\omega_c)^2}{4(1 + x)T\omega_c}$$

$K > 1$ corresponds to overdamping (i.e. no ringing), $K < 1$ corresponds to underdamping, and $K = 1$ corresponds to critical damping. In fact, K controls the shape of the $Q(\omega)$ curve, as shown in Fig. 10.11. To avoid ringing on impulse interference, K should be 1 or more.

Notice that if $xR = 0$

$$K = \frac{1}{4T\omega_c}$$

and as shown earlier

$$f_{NN} = \frac{\omega_c}{4}$$

Thus, for given tuning and static phase errors, f_{NN} is fixed, and changing the time constant T merely changes the value of K, i.e. the shape of the $Q(\omega)$ curve.

To summarize the in-sync performance, the formulae are

Static phase error

$$\sin \Delta\phi = \frac{\Delta f}{f_c} = \frac{\Delta\omega}{\omega_c}$$

where

$$\omega_c = 2\pi\mu\beta$$

Dynamic phase error, i.e. noise bandwidth

$$f_N = 2f_{NN} = \frac{\omega_c\left[1 + \dfrac{x^2T\omega_c}{1 + x}\right]}{2(1 + xT\omega_c)} \simeq \frac{1 + x^2T\omega_c}{2xT}$$

since for the range of values which are of interest, $x << 1$ and $xT\omega_c >> 1$.

Damping factor K (preferably 1 or more)

$$K = \frac{(1 + xT\omega_c)^2}{4(1 + x)T\omega_c} \simeq \frac{x^2T\omega^2}{4}$$

It is important to notice that the in-sync performance is not merely a function of the low pass filter constants x and T. The in-sync performance is also a function of the loop gain ω_c.

10.12. Pull-in performance

The pull-in performance of the loop should be such that frequency and phase lock occur reasonably quickly after the burst signal is

applied.   If there is a small tuning error (strictly $\Delta f < mf_c$) the
loop will pull in to frequency lock immediately, and there will
be a small phase stabilization time which is usually only of the
order of hundredths of a second.

If there is a large tuning error

$$\Delta f > f_c \sqrt{2m - m^2}$$

the loop will never pull in to frequency lock.

If there is an intermediate tuning error

$$mf_c < \Delta f < f_c \sqrt{2m - m^2}$$

the loop will pull in to frequency lock, though if Δf is close to
$f_c \sqrt{2m - m^2}$ the time taken may be of the order of minutes.
When the loop has pulled in to frequency lock, rapid phase stabiliza-
tion occurs as for the case $\Delta f < mf_c$.   (*See* Appendix 4.)

The mechanism of frequency lock for the above case may be
qualitatively explained as follows:—

Suppose for the moment that the reactance valve is disconnected.
The phase detector output will then be sinusoidal, with a frequency
equal to the tuning error.   If the reactance valve is now connected,
it will try to increase the oscillator frequency during one half cycle
of beatnote, and reduce it during the other half cycle.

The beatnote waveform will therefore no longer be sinusoidal,
one half cycle corresponding to a low frequency beatnote and the
other corresponding to a high frequency beatnote.   The beatnote
waveform will have a shape as shown in Fig. 10.12.   This waveform
has unequal areas under the positive and negative half cycles, so
that it has a D.C. component.   This is stored in the filter condenser
so that it reduces the mean tuning error.   As time progresses, so
the D.C. component builds up as a voltage across the condenser,
and the mean tuning error is gradually reduced.   Eventually the
tuning error becomes less than mf_c, the loop then pulls in to fre-
quency lock immediately, and a phase transient follows.

It is interesting to note that the D.C. component will always act
in such a direction as to correct the frequency error.   The " larger "
half cycle of beatnote necessarily corresponds to a shift of oscillator
frequency towards the burst frequency.

Another point to notice is that, quite apart from the loop filter,
the loop itself has a frequency characteristic.   For example, if the
condenser C were short circuited, there would be no frequency
dependent components in the loop, yet the beatnote waveform of
Fig. 10.12 (which in this case would be a steady beatnote, so that the
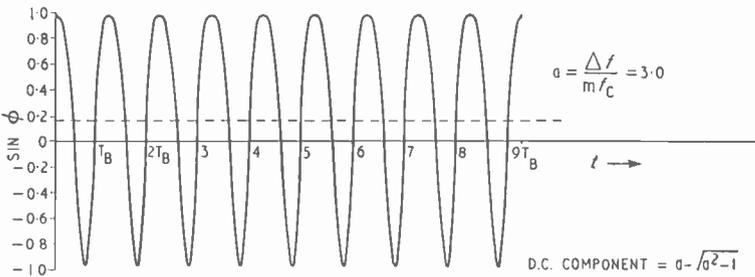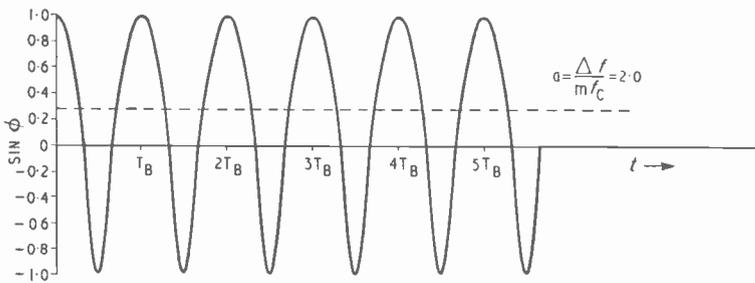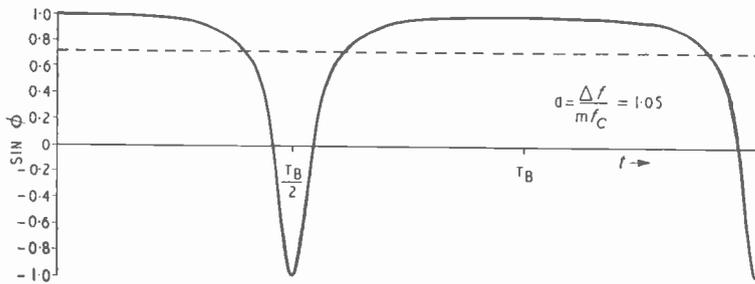loop would not pull in) would become more sinusoidal if the tuning

Fig. 10.12.   Detector beatnote output for constant loop gain (mf$_c$ = constant)
and various tuning errors.   T$_B$ is the beatnote period.   D.C. components are
shown dotted

error were increased.   The effective loop gain in this simplified
loop is mf_c c/s, which means that the loop is capable of shifting
1 radian of phase in a time $\dfrac{1}{2\pi mf_c}$ sec. Now if the tuning error is large,
the beatnote frequency is high, and the time of a half cycle of beat-
note is small.   If the time $\dfrac{1}{2\pi mf_c}$ sec is larger than the beatnote

period, the loop cannot change the phase much from one half cycle of beatnote to the next so that the beatnote tends to become sinusoidal.

Referring again to the loop filter of Fig. 10.9, the maximum pull-in range is

$$\Delta f_{\max} = f_c \sqrt{2m - m^2}$$

This may be written as

$$\Delta \omega_{\max} \simeq \omega_c \sqrt{2x}$$

for small values of m, since then $m \simeq x$.

Now the noise semi-bandwidth has been derived earlier as

$$f_{NN} \simeq \frac{1 + x^2 T \omega_c}{4xT}$$

or

$$\omega_c \simeq \frac{4f_{NN}}{x} - \frac{1}{x^2 T} = \frac{1}{\sqrt{x}}\left[\frac{4f_{NN}}{\sqrt{x}} - \frac{1}{T\sqrt{x^3}}\right]$$

The maximum pull-in range may therefore be expressed as

$$\Delta \omega_{\max} \simeq \omega_c \sqrt{2x} \simeq \sqrt{2}\left[\frac{4f_{NN}}{\sqrt{x}} - \frac{1}{T\sqrt{x^3}}\right]$$

For a constant value of f_{NN}, the pull-in range can therefore be indefinitely increased by reducing x and increasing T appropriately. This will increase the quantity

$$\frac{4f_{NN}}{\sqrt{x}} - \frac{1}{T\sqrt{x^3}}$$

and hence ω_c must be appropriately increased, since ω_c is $\dfrac{1}{\sqrt{x}}$ of the above expression. However, apart from economic considerations, the maximum pull-in range cannot exceed $\dfrac{f_L}{2}$ where f_L is the line scan frequency. Remember that the sidebands adjacent to the sub-carrier component of the burst are separated from it by f_L, and have a strictly comparable amplitude (Appendix 1). For tuning errors greater than $\dfrac{f_L}{2}$, an ambiguous beatnote between the oscillator and the nearest sideband can occur, and the oscillator may even

'lock to the sideband. This would give an incorrect reference frequency, of course, and this effect is called *side-lock*.

Quite apart from the above considerations, the pull in time would be excessive. For example, it is shown in Appendix 4 that for an optimum K value the pull in time is approximately given by

$$T_F \simeq \frac{4 \cdot 2 (\Delta f)^2}{f_{NN}{}^3}$$

If $\Delta f = 5$ kc/s, and $f_{NN} = 70$ c/s, then T_F would be just over 5 min.

This example illustrates the limitations of the straightforward A.P.C. loop shown in Fig. 10.9. Thus, for large tuning errors, pull-in time will be excessive unless f_{NN} is increased, that is, unless the noise performance is deteriorated. For the required small value of f_{NN}, therefore, reasonable pull in times can be achieved only if the tuning error Δf is kept small. This immediately suggests the use of a crystal controlled oscillator.

In fact, there are two solutions to the problem of good noise performance consistent with short pull-in time. The use of a crystal oscillator is one solution; the use of a two mode loop, or *quadricorrelator*, is the other solution.

10.13.   Crystal oscillator A.P.C. loops

A circuit diagram of a crystal controlled oscillator A.P.C. loop is shown in Fig. 10.13.

The oscillator is a modified Pierce crystal oscillator with the screen grid of the pentode acting as the " anode " of the oscillator triode. The output is taken from the actual anode of the pentode, and is therefore electron coupled to the oscillator, thereby providing some isolation. The tuned circuit in the anode is tuned for maximum output.

The reactance valve anode is capacity coupled to the crystal, and its tuned circuit is adjusted so that the beat between the reference signal and the sub-carrier is nominally zero when the reactance valve grid is earthed. This condition is easily observed on a colour picture, when all the colours will be seen to pass through the spectrum range at the beat frequency.

Some cathode D.C. feedback is included in the reactance valve for stabilizing the valve characteristics, and heavy decoupling is advisable to prevent hum modulation of the oscillator phase, particularly for a series heater chain. It should be noted that for crystal oscillators the sensitivity of the reactance valve is relatively
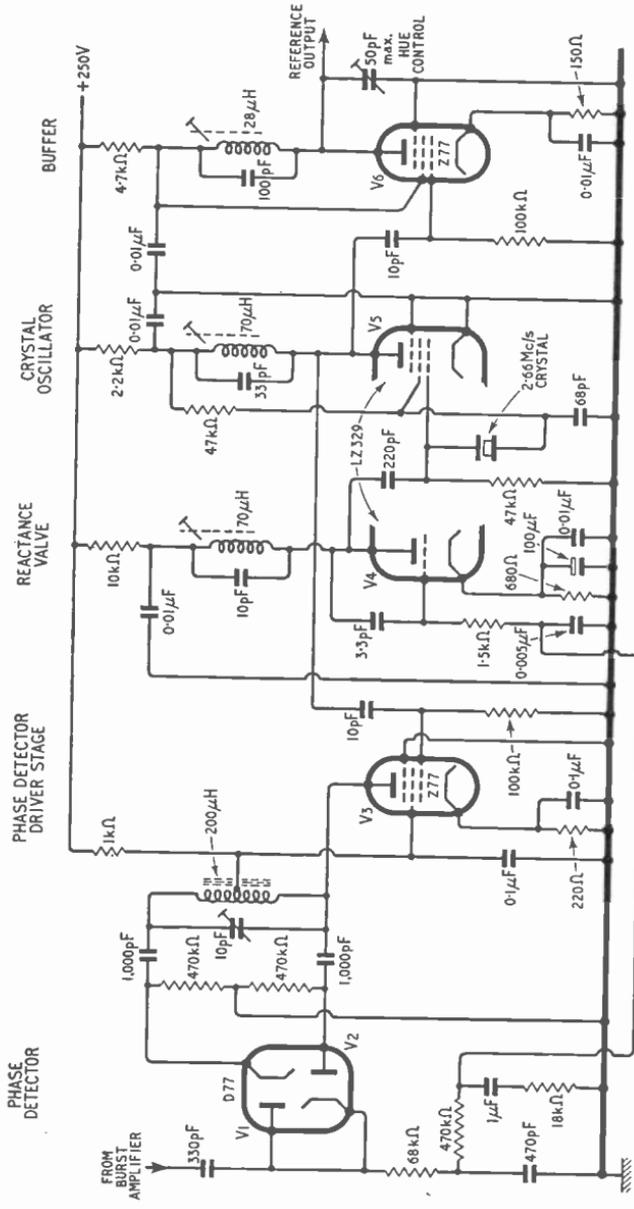
Fig. 10.13. Circuit of a 2·66Mc/s crystal oscillator A.P.C. loop

low because a large capacity change is required to shift the crystal frequency.   The sensitivity is also a function of the crystal frequency accuracy.

For example, if a large capacity is required to be added to achieve the correct frequency, the reactance valve itself must be able to provide a relatively large capacity change.

The oscillator drive to the phase detector is obtained from an amplifying stage whose anode circuit includes one half of a centre tapped auto-transformer.   This transformer should have very tight coupling between its two windings so that its push-pull output is symmetrical about earth.   A convenient arrangement is a Ferrox-cube " E " and " I " core of B2 material, since the high permeability ensures tight coupling and also reduces the self capacitance of the winding.

A total of about 10 turns of wire (preferably bifilar wound) is usually required to tune the transformer to 2·66Mc/s with an added trimmer of 10pF.

It should be noted that an accurate balance is desirable in the phase detector, otherwise a spurious D.C. unbalance voltage will be fed to the reactance valve which, in turn, will detune the oscillator.

Accuracy of balance is dependent on the symmetry of the two transformer windings and the equality of the two leak resistors of (in this case) 470kΩ.

Balance accuracy may be checked by measuring the D.C. phase detector output by means of a high impedance voltmeter connected to the junction of the 68kΩ resistor and 470pF capacitor.   (Alternatively, shift of oscillator frequency may be measured).   The voltage should be measured as zero, firstly with the burst signal absent and the oscillator signal present, and secondly with the burst signal present and the oscillator signal absent.   The first of these measurements checks the equality of the resistors and transformer windings, and the second measurement checks that the impedances of the two halves of the circuit have equal phase angles. In this connection the trimmer should be adjusted for accurate balance rather than maximum output, though normally these two settings are close together.   As a general rule, the unbalance voltage should not exceed about 1 % of the smallest voltage applied to the phase detector, and it is advisable to use matched pairs for the leak resistors.

The hue control shown makes use of the phase shift between current and voltage which changes as a tuned circuit is tuned through resonance.   Considering a simple circuit of L, C and R

all in parallel, a phase shift of 45° occurs when C is changed by an amount

$$\delta C = \frac{1}{\omega_o R}$$

from its resonant value, and the magnitude of the impedance is then 3dB below its resonant value. Thus, if $R = 3k\Omega$

$$\delta C = 19 \cdot 7 pF$$

for a 45° phase change at 2·7 Mc/s. If a 50pF trimmer is used, and the circuit is tuned by adjustment of the inductance when the trimmer is set half way, then a phase shift of more than $\pm 45°$ can be made by adjustment of the trimmer. The use of a relatively large fixed capacity in the tuned circuit enables lead capacity to be accommodated. Screened lead is necessary and its length will be dependent on the physical position of the hue control relative to the reference oscillator.

The value of the load resistance R will depend on the type of synchronous detectors used in the demodulators. $3k\Omega$ has been chosen in this case as a fairly representative value. Coupling between the hue control and the oscillator itself should be sufficiently small to prevent change of oscillator tuning with hue control setting. This is one reason for the buffer output stage, which also isolates the load from the oscillator. However, with a carefully laid out circuit the buffer stage can be omitted provided the output is sufficiently large and of low enough impedance.

Note that the hue control may be included in the burst signal side rather than the oscillator side.

The loop design is a function of the static and dynamic phase errors which can be tolerated, the maximum tuning error to be expected, and the pull-in time which can be allowed. Typical values for a crystal controlled oscillator A.P.C. loop are as follows:—

Burst drive to phase detector   80V peak

Reference drive to phase detector   35V peak

$\mu = 35$ volts/radian

$\beta = 150$ c/s/V

$f_c = 5 \cdot 25$ kc/s/radian

Maximum oscillator drift ± 200 c/s

Static phase error $\pm 2 \cdot 2°$

Noise semi-bandwidth $f_{NN} = 100$ c/s

Shunt time constant of filter $xT = 18 \times 10^{-3}$sec

A.C./D.C. gain ratio of filter $m \simeq x = 10 \times 10^{-3}$

Pull-in time for maximum tuning error 0·25 sec

A very important practical consideration in A.P.C. circuits is the amount of stray feedback from the oscillator side to the burst side of the phase detector. If feedback is present, the loop will try to lock to its own oscillator; that is, the phase detector will tend to change the oscillator frequency so that a 90° phase difference occurs across the phase detector. The feedback therefore shows as
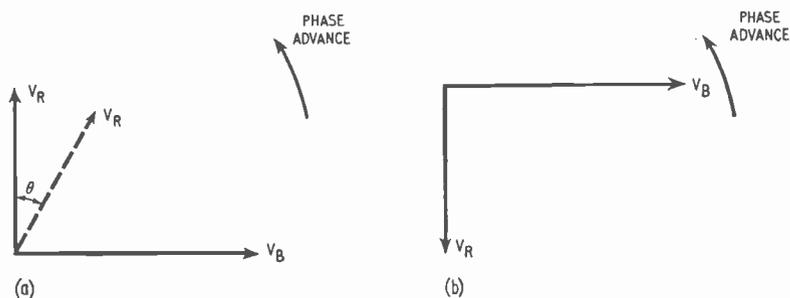
Fig. 10.14. Phase relation between reference and burst signals for the circuit shown in Fig. 10.13. V_R is the oscillator drive to top diode. (a) shows unstable and (b) stable positions

an unbalance which is a function of hue control setting, and it can be checked by measuring the oscillator frequency as the hue control is rotated through its entire range with no burst applied. In case the hue control affects the frequency directly, this can be checked by earthing the grid of the reactance valve while the oscillator frequency is measured for various hue control settings.

The feedback between oscillator and burst input can be reduced by fitting a screen directly across the diodes. Note, however, that this will only reduce internal feedback. External feedback can occur if the burst take off point is contaminated with reference signal.

It has already been stated that when an A.P.C. loop is locked, a 90° phase difference exists between the signals applied to the phase detector (apart from the small static phase error). Whether this is a 90° lag or lead will depend on the circuit configuration, but as an example, consider the circuit shown in Fig. 10.13. Suppose the burst drive is represented by V_B while the reference drive to the top diode is V_R. First, draw V_B as a reference vector as shown in Fig. 10.14(a). Next, assume that V_R leads V_B by 90°.

If this is the correct phase relationship, then if V_R is displaced in phase, the control action should restore it.   Suppose V_R is displaced clockwise by an angle θ, then the control action should try to advance the phase of V_R to restore it to its initial position.   However, because V_R is now more nearly in phase with V_B, the bottom diode will provide the larger output voltage, and since the output is taken from the cathode of the bottom diode the output voltage will be positive.   This will try to reduce the oscillator frequency, that is, retard the phase of V_R.   The control action therefore acts in the same direction as the initial displacement θ, and the stable position of V_R is consequently as shown in Fig. 10.14(b).

If the above procedure is applied to Fig. 10.14(b), it will be found that the control action acts oppositely to the initial displacement.

This method of determining the sign of the 90° phase relationship is useful for checking that the correct reference phase can be achieved within the range of the hue control.   It is also useful for determining the sign of the D.C. output voltage in the case of two mode loops, as discussed in the next section.   Note that a 180° phase shift can be obtained by connecting the anode of the driver stage to the other end of the push-pull transformer.

10.14.   Two mode loops or quadricorrelators

It has already been mentioned that simple A.P.C. loops have conflicting design requirements in that good pull-in performance cannot be achieved unless in-sync noise performance is sacrificed. The crystal oscillator loop overcomes this problem by its inherent stability which allows a fast pull-in for a relatively weak control.

However, if a two mode loop is used, one mode can be designed to give good pull-in performance regardless of in-sync conditions, while the other mode can be designed for good in-sync performance regardless of pull-in performance.

In this case the pull-in performance may be designed to cater for relatively large tuning errors, and a crystal oscillator is not required. It is estimated that an oscillator stability of about ± 2 kc/s is feasible at 2·7 Mc/s for a carefully designed LC oscillator, so that a pull-in range of ± 3 kc/s should allow a reasonable factor of safety.   A stability of $\pm \dfrac{f_L}{2}$, where f_L is the line scan frequency, is the maximum allowable drift if sidelock is to be avoided.

The change over from one mode to the other can be carried out by an automatic switch which recognizes the in-sync condition of the loop, and acts accordingly.   Since this " switch " usually operates

from a quadrature relation between two signals, the term " quadri-correlator " is often used to describe two mode loops.

Fig. 10.15 shows a block diagram of a typical quadricorrelator. It consists of a conventional A.P.C. loop which is designed for the required in-sync noise performance regardless of pull-in considerations, and a second or auxiliary phase detector to which
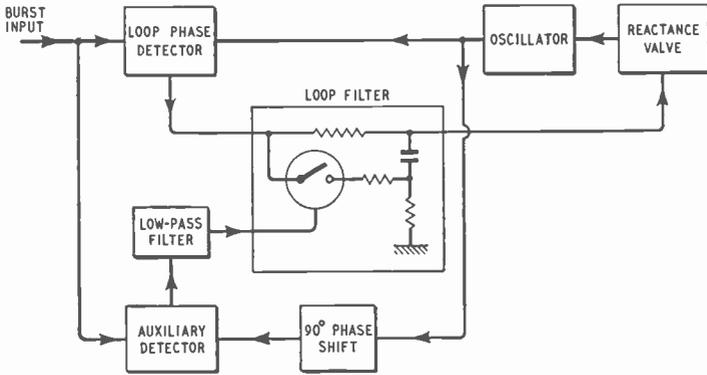

Fig. 10.15.   Quadricorrelator block diagram

is fed the burst signal and a quadrature signal from the oscillator. The object of the auxiliary detector is to produce a D.C. output voltage only when the loop has synchronized, and a zero output for all other conditions.   After passing through a low pass filter, the auxiliary detector output operates a " switch " which is closed when the output is zero, i.e. when the loop is unlocked, and open when there is a D.C. output from the auxiliary detector, i.e. when the loop is locked.   When the switch is closed, the transmission is increased through the filter and the design is arranged for good pull-in performance.   As soon as the loop has locked, the switch is opened and the loop filter is then suitable for good noise performance.

It will be remembered that during pull-in, the beatnote output of a loop phase detector is typically as shown in Fig. 10.12.   The beatnote output from the auxiliary detector is a quadrature version of this, and while the conventional beatnote of Fig. 10.12 is proportional to $\sin \phi$, the auxiliary detector output is proportional to $\cos \phi$.   This is readily deduced from $\sin \phi$, and is shown in Fig. 10.16.   Notice that this waveform is symmetrical about the time axis and therefore has a zero D.C. component.   If the auxiliary

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detector output is taken through a low pass filter to remove the A.C. components, the output will be zero whenever a beatnote exists, that is, when the loop is not locked.

When the loop has locked, however, ϕ is zero (apart from the very small static phase error) and the auxiliary detector output is then a D.C. voltage equal to

$$\mu_A \cos \phi \simeq \mu_A$$

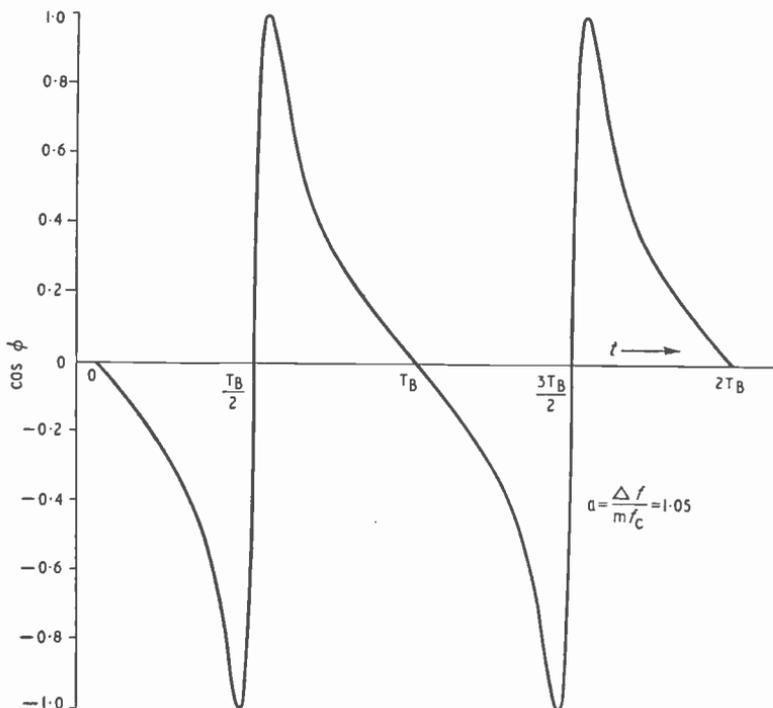where μ_A is the sensitivity of the auxiliary detector. This D.C. voltage can be positive or negative, as desired, the sign being

Fig. 10.16.   Quadrature detector beatnote

dependent on whether a 90° lag or lead is included. Normally a negative D.C. output is required so that a valve can be biased off to change the loop filter characteristic.

A typical quadricorrelator circuit, designed by the Hazeltine Research Corporation, is shown in Fig. 10.17. This design is

Fig. 10.17. 3·58 Mc/s quadricorrelator circuit

based on the American system sub-carrier frequency of 3·58 Mc/s, but of course it can be modified for 2·66 Mc/s operation by changing the various tuned circuits appropriately.

The conventional A.P.C. loop consists of a detector V_1, reactance valve V_2 and oscillator V_3. A buffer stage V_4 feeds a quadrature network which supplies two quadrature phased reference outputs for the receiver synchronous detectors.

The auxiliary detector V_5 is fed with the same burst signal as the loop detector V_1, but its reference feed is 90° out of phase with the V_1 reference. The auxiliary detector output is taken through a simple low pass filter of 100kΩ and 0·22 μF, so that any A.C. components are heavily attenuated. As explained earlier, this auxiliary detector output is zero when the loop is not locked, and (in this case) a negative D.C. voltage when the loop is locked.

Now the conventional loop filter is designed so that the noise bandwidth is satisfactorily small for in-sync performance, but a cathode follower V_6 is arranged so that it can increase the A.C. transmission through this filter and thereby increase the noise bandwidth for the unlocked condition.

The cathode follower grid resistor is returned to the auxiliary detector output, so that in the unlocked condition of the loop, the cathode follower operates and causes rapid pull-in. When pull-in has occurred, the auxiliary detector output is a negative D.C. voltage, and this cuts off the cathode follower so that the original filter circuit is restored.

The essential additions to the circuit of a simple A.P.C. loop therefore consist of an auxiliary phase detector and a cathode follower by-pass. However, because the pull-in range is independent of the in-sync design, very much greater oscillator drift is permissible so that the oscillator need not be crystal controlled.

This type of quadricorrelator is often called the variable ratio quadricorrelator because the A.C./D.C. gain ratio m (and therefore mf_c) is increased for pull-in. Instead, it is possible to increase the phase detector sensitivity μ while keeping m constant, so that $f_c(=\mu\beta)$ and therefore mf_c, which controls pull-in, is increased. This is the action of the variable gain quadricorrelator, and in this case the auxiliary detector bias is used to reduce the gain of one of the burst amplifier stages (and therefore the burst drive to the phase detector) when the loop has locked.

It is perfectly practicable to employ the variable gain and variable ratio techniques simultaneously, but in the authors' opinion the loop gain f_c should be high even for the in-sync condition to keep the static phase error small.

Typical values for a variable ratio quadricorrelator circuit are as follows:—

Burst drive to both detectors   80V peak

Reference drive to both detectors   35V peak

$\mu = 35$ volts/radian

$\beta = 2$ kc/s/V

$f_c = 70$ kc/s/radian

Maximum oscillator drift ± 3 kc/s

Static phase error $\pm 2\cdot 5°$

In-sync noise semi-bandwidth $f_{NN} = 100$ c/s

In-sync shunt time constant of filter 6×10^{-3} sec

In-sync A.C./D.C. gain ratio of filter $0\cdot 5 \times 10^{-3}$

Out of sync A.C./D.C. gain ratio of filter 15×10^{-3}

Pull-in time for maximum tuning error $1\cdot 5$ sec

It should be noted that the increase in the value of the A.C./D.C. gain ratio for the not-in-sync condition, produced by the cathode follower by-pass, is equivalent to an increase in the value of the shunt resistor of the filter, and not to a decrease in the value of the series resistor.   This is shown in Appendix 7.

10.15. Colour killing

If a black and white picture is displayed on a colour receiver whose chrominance circuits are operative, spurious colour effects will occur due to noise and luminance components which fall within the chrominance pass band.   If a burst is present, the effects of the latter are subjectively reduced because of the frequency lock between the receiver reference generator and the line scan, but in the absence of a burst it is very desirable to switch off the chrominance circuits.   This is particularly true if A.C.C. (*see* Section 10.16) is fitted, in which case the chrominance gain is a maximum when there is no burst.

Since the presence of the burst is a reliable indication of a colour transmission, it is reasonable to place the responsibility on the reference generator for automatically switching the chrominance circuits.   Such a switching action is appropriately referred to as *colour killing*.

In order to switch off the chrominance circuits, a negative bias is generally convenient.   Hence, the colour killer must provide a negative D.C. voltage when the burst is absent and a zero voltage

when it is present.   However, reference generators normally give the converse of this; for example, the auxiliary detector output of a quadricorrelator is zero for not-in-sync (i.e. no burst) and a negative D.C. for in-sync.   However, a negative voltage which appears only when the reference generator has synchronized or when there is a burst, can be utilized for operating a valve which will give the required bias for killer switching.

A suitable circuit is shown in Fig. 10.18.   A triode is used as a grid controlled rectifier; that is, when its grid is earthed, the triode anode and cathode operate as a rectifying diode, so that a negative
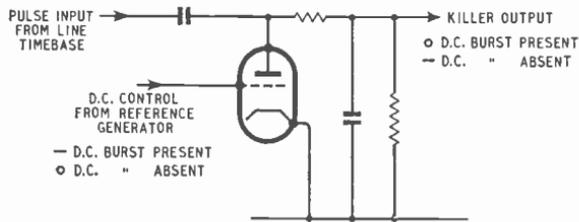
Fig. 10.18.   Colour killer circuit

D.C. output is obtained if a waveform is applied to the anode. This waveform may be a 50 c/s sine wave (e.g. the mains supply) or preferably a pulse waveform from the receiver line time base. The higher frequency of the line waveform allows smaller (and therefore quicker action) time constants to be used in the smoothing circuit.   Positive going pulses are desirable since a negative D.C. output is required.

Of course, when a negative bias is applied to the triode grid, the rectification action is prevented and the killer output becomes zero.

Sometimes the triode grid standing bias is made adjustable so that the killer output becomes zero only when the D.C. control from the reference generator has reached a sufficiently large negative value; a spurious negative D.C. due to noise, for example, can thereby be " backed off ".   Such a bias control is usually called the *killer threshold adjust*.

In the case of a simple A.P.C. loop, such as Fig. 10.13, the D.C. control for the colour killer may be obtained from the anode of the lower of the two phase detector diodes.   The D.C. voltage at this point is increased negatively when a burst is present.   When there is no burst the D.C. voltage here is equal to the peak value of the reference drive.   This method of deriving the killer control bias is

unreliable for poor signal-to-noise ratio signals, because it depends on simple amplitude detection of the burst; any noise present can also produce a D.C. voltage. Thus, a noisy monochrome signal may be interpreted as a colour signal.

In the case of two mode loops, however, the D.C. output of the auxiliary detector is a very reliable indication of the presence of a burst. The fundamental reason for this is that the auxiliary detector, like the loop detector, operates as a synchronous, and not as an amplitude, detector. Thus, when the loop is not synchronized, noise present with the incoming burst will produce fluctuations above and below zero, which, when integrated by the low pass filter, will produce a zero D.C. output. When lock has occurred, noise fluctuations will tend to reduce the D.C. output somewhat, since fluctuations in the phase will always act in a positive direction; the output is proportional to $\cos \phi$, where ϕ is the phase error, and as ϕ varies about zero the value of $\cos \phi$ will vary from unity to less than unity. However, the D.C. output can always be made large enough to avoid ambiguity. Note that the use of the auxiliary detector for colour killer bias means that the chrominance channel is made operative only when the loop has locked. Spurious colour is therefore avoided during pull-in.

It must be remembered that the killer bias must not be applied to the chrominance channel prior to the burst take-off point, otherwise the reference generator will never receive a burst signal. However, should this situation arise, it may be overcome by adding the burst gate pulse to the killer bias so that, for monochrome transmissions, the killer action applies only to the active part of the line scan. In general, this technique is not recommended since spurious colour will be displayed on the left of the picture unless carefully adjusted line blanking is employed.

10.16. Automatic chrominance control

It is possible to arrange an automatic gain control for the chrominance channel of a receiver so that the chrominance output remains practically constant as the chrominance input level changes. Such an arrangement removes the probable need for a change in saturation adjustment from one station to another, and it also tends to compensate for any differential fading between the chrominance and luminance signals. To distinguish this type of gain control from the conventional A.G.C. which may be fitted to the receiver, it is usually called A.C.C., or automatic chrominance control.

As in the colour killer, the burst is the only reliable measure of the chrominance signal as it appears during colour transmissions

only and its amplitude is independent of picture content. In fact, the negative D.C. voltage which is used to bias off the colour killer valve whenever the burst is present may be used as an A.C.C. voltage by feeding it back to control the gain of a chrominance stage prior to the burst take-off point, provided that this negative voltage increases with burst amplitude.

Thus, if the negative bias appearing at the anode of the lower diode of the simple loop of Fig. 10.13 is fed back to the chrominance stage which feeds the burst into the loop, then the burst input will tend to remain constant and, of course, so will the chrominance signals corresponding to a given picture. Manual saturation control may then be applied after the burst take-off point.

The above method of A.C.C., like the killing action, will be unreliable under poor signal-to-noise conditions because the D.C. control is obtained from amplitude detection. This would mean that when switching from a " clean " to a " noisy " transmission, the saturation would decrease in the latter case, because the noise would simulate a larger burst.

It follows that reliable A.C.C. is best accomplished by synchronous detection, and the D.C. output of the auxiliary detector of a two mode loop may be utilized for this purpose, as well as for colour killer control, provided that the auxiliary detector output is arranged to be substantially dependent on the burst amplitude rather than the reference amplitude. This can be done by making the burst feed to the auxiliary detector smaller than the reference feed.

10.17. Reference generator circuits

The circuits which have so far been given for passive and dynamic integrators have deliberately not included colour killers or A.C.C. for the sake of simplicity. However, more complete circuits will now be given, and these will additionally illustrate various modifications which may be encountered in reference generator designs.

10.18. Passive/dynamic integrator

Fig. 10.19 shows a circuit designed by R.C.A. for their CTC-5N series of receivers. It is a hybrid design in that the output of a passive integrator is used to injection lock an oscillator whose frequency is controlled by the crystal of the passive integrator.

The gated burst is fed to a transformer whose secondary is connected to the oscillator grid via a quartz crystal. The mounting capacitance of the crystal is neutralized by a $1 \cdot 6pF$ capacitor connected to the appropriate anti-phase end of the primary winding,
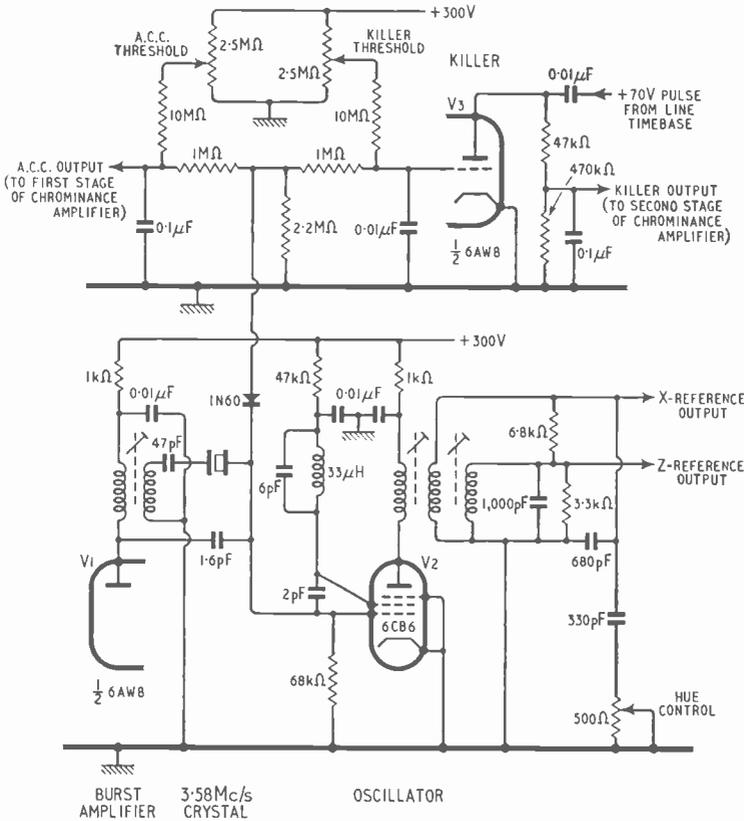
Fig. 10.19.   Circuit of a 3·58Mc/s passive/dynamic integrator

and the circuit so far described is similar to a simple passive integrator, the narrow bandwidth (quoted as 400 c/s) of the crystal determining the noise performance. Apart from any noise components within this pass-band, the signal fed to the oscillator grid is substantially the subcarrier component of the burst.

The electron coupled oscillator is arranged so that the crystal is included in its grid circuit, hence the oscillator frequency is crystal controlled. Consequently, when a burst is present, the sub-carrier component of it synchronizes the oscillator. The hue control operates by changing the effective capacitative load on the output tuned circuit.

When the burst is absent, the oscillator develops a constant negative bias at its grid. When a burst is present, however, the

sub-carrier component of it produces a larger bias voltage by rectification at the oscillator grid. In addition, a crystal diode voltage doubles this bias by conducting on the negative peaks of the C.W. Hence, by backing off the standing oscillator bias by means of threshold controls, the D.C. voltage output may be used for colour killing and A.C.C.

Note that although the killer bias is developed by simple amplitude detection, its noise performance is adequate because of the narrow crystal pass-band.

It should be noted that the noise performance of this circuit is entirely dependent on the crystal parameters. A very narrow noise
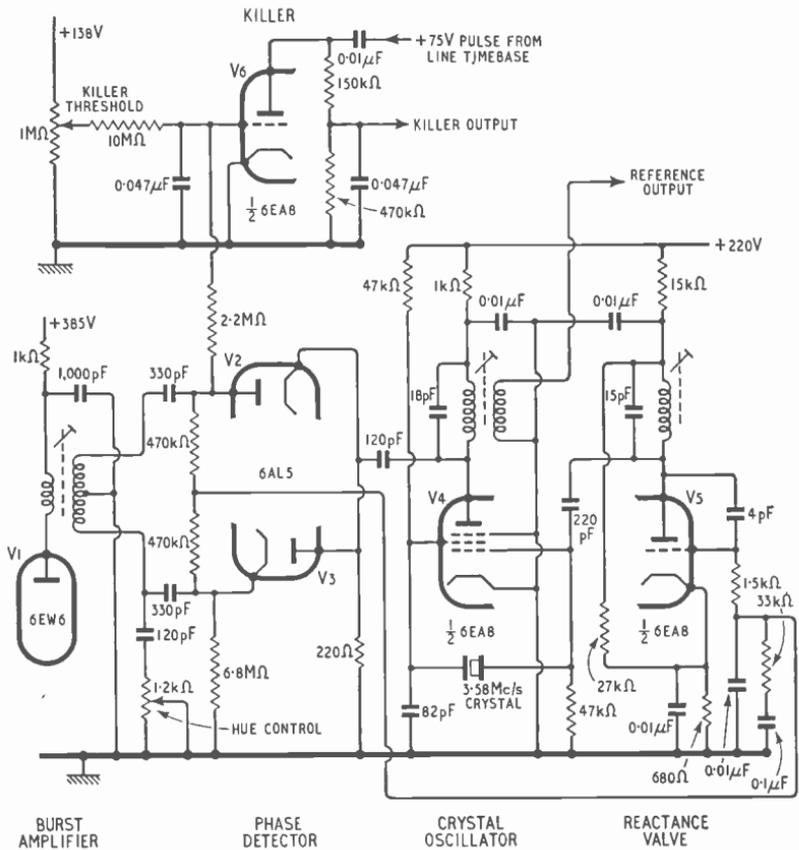
Fig. 10.20. Circuit of a simple A.P.C. loop and killer

bandwidth can be achieved only by the use of high Q crystals which would lead to large static phase errors.

10.19. Simple A.P.C. loop

A straightforward and effective crystal oscillator A.P.C. loop, used by R.C.A. in their CTC9 series of receivers, is shown in Fig. 10.20.

The general principles of this type of circuit have already been discussed, but note that here the burst signal is fed push-pull to the detectors, and the detector output is taken from the junction of the diode leak resistors. The hue control varies the phase of the burst signal rather than the reference signal.

The bias for the killer valve is obtained from the negative voltage appearing at the anode of the top diode, and since a negative voltage due to the reference appears at this point when the burst signal is absent, a positive backing off voltage is applied to the killer grid from a killer threshold control. This control is also used to back off any spurious D.C. voltage caused by noise which may be present with the monochrome signal.

The positive bias action of the killer threshold control tends to lift the anode of the top diode positive. This would normally upset the D.C. balance of the phase detector, but a compensating negative D.C. component is obtained by bleeding the cathode of the lower diode to earth through $6 \cdot 8 \text{M}\Omega$.

10.20. Quadricorrelator circuit

A quadricorrelator circuit used in the G.E.C. TTIV series of receivers is shown in Fig. 10.21.

This circuit is designed for series run heater operation, and certain precautions are advisable to prevent hum modulation of the reference phase. Thus an earthed cathode Colpitts oscillator is used, one end of the heater of the oscillator and reactance valve (LZ329) combination being earthed. Next in the chain are the filter by-pass and loop and auxiliary detector heaters. The cathode of the filter by-pass is tapped down to reduce hum effects, since any hum on this cathode will not be reduced when the valve is cut off for in-sync operation.

The electron coupled oscillator has large swamping capacitors in its tuned circuit to aid stability, and a negative temperature coefficient capacitor provides compensation for warm-up drift. The H.T. supply to the screen of the oscillator is obtained from a potentiometer to reduce the dependence of oscillator frequency on H.T. line variations. Coarse oscillator frequency adjustment

Fig. 10.21. 2·66Mc/s quadricorrelator circuit

is by way of a brass slug tuned coil, and a fine oscillator frequency control can be made with a 25pF trimmer.

The loop detector is fed via an *RC* phase shifting circuit of 4·7pF and 4·7kΩ so that the reference phase at the loop detector leads the phase of the transformer push-pull output by about 70°. The magnitude of the auxiliary detector output is therefore proportional to cos 20° (or 0·94) instead of cos 0° (or 1) which would obtain for a quadrature phase. The non-quadrature relation therefore only slightly affects the magnitude of the auxiliary detector output.

The *RC* circuit of 4·7pF and 4·7kΩ also reduces the amplitude of the reference fed to the loop detector so that the detector output is chiefly dependent on the reference signal amplitude. One of the coupling capacitors is adjustable so that an accurate balance may be obtained for the loop detector.

The reference feed to the auxiliary detector is large so that the output is chiefly dependent on the amplitude of the burst signal. This detector is arranged to give a detected burst pulse output rather than a D.C. output so that the detected burst pulse may be amplified by an " audio " amplifier. The anode waveform of this amplifier therefore consists of positive going pulses whose amplitude is proportional to the burst signal. These pulses are D.C. restored one way by the diode *D*1 and the reverse way by the diode *D*2, and the outputs of the two D.C. restorers are connected in series. The result of this arrangement is that, when not in sync, the diode circuits give equal but opposite D.C. outputs, that is, zero D.C. output. When in sync, however, one diode circuit gives a larger D.C. output than the other, and a net negative D.C. output is obtained. Because of the appreciable gain of the " audio " amplifier, a large D.C. output indicative of lock is readily obtained, and reliable killer action results even for very small burst inputs; in fact, provided the burst signal is large enough to achieve lock, the killer becomes completely " unkilled ". This arrangement avoids the occurrence of a " semi-killed " state in which, although the loop is locked, the auxiliary detector output is not sufficient to switch off the killer completely and so a reduction in saturation is observed.

The amplified burst output mentioned above may also be used for A.C.C. purposes, in which case very strong A.C.C. action is obtained because of the high gain of the system.

10.21. Alternative possibilities for reference generation

Before leaving the subject of reference generators, brief mention should be made of two techniques which may well prove useful.

The first of these is applicable to colour receivers which feed difference signals to the display tube. Thus, since the burst phase is $-(B' - Y')$, a correctly adjusted receiver should give a zero output in its $(R' - Y')$ output corresponding to the synchronously detected burst. Therefore, if a gating circuit is arranged to " look at " the detected $(R' - Y')$ output during the burst period only, and if this gated output is used to A.P.C. an oscillator from which the reference signal is derived, then the oscillator reference will automatically adjust its phase to give a zero output for the detected burst in the $(R' - Y')$ phase. This, of course, corresponds to the correct hue setting, so that no manual hue control is required. Provided the gain of the A.P.C. loop is adequate, the responsibility for correct hue setting then rests entirely on the accuracy of the transmitted burst phase.

Further, if a similar gating system is applied to the receiver's detected $(B' - Y')$ output, then a negative going detected burst pulse should be obtained when the receiver's reference signal is correctly synchronized. This signal may therefore be used (as the equivalent of an auxiliary detector output) for quadricorrelator switching of the A.P.C. loop filter, and also for colour killing and A.C.C.

Preliminary tests carried out by the authors have indicated that the above technique is feasible. It has the advantage of economy in that the whole of the receiver's chrominance amplifier is used for burst amplification, and of course no hue control should be required.

While on the subject of automatic control, it should be pointed out that automatic setting of the saturation is practicable. Thus, since the peak-to-peak burst amplitude is equal to the sync pulse height when the chrominance gain bears the correct ratio to the luminance gain, a circuit which compares the detected burst and the sync pulse height may be arranged to automatically adjust the saturation accordingly.

The second technique (which as far as the authors are aware has not yet been investigated), takes advantage of the fact that in the case of N.T.S.C. systems which use A.M. sound, the sound/vision carrier beat is simply related to the sub-carrier frequency. Thus, for the British adapted N.T.S.C. system, for example, the sub-carrier frequency is $\frac{3}{4}$ of the sound/vision carrier beat. Consequently, if a strong carrier beat can be derived which is sufficiently free of sound and vision modulation, the reference may be easily generated from it by means of a simple $\frac{3}{4}$ regenerative divider.

The possible disadvantage of this technique is the need to remove the modulation, but a fundamental advantage (particularly for

positive modulation systems in which the vision carrier during the active line period is never less than 30%) is the advantageous duty ratio compared with that of the conventional burst.   Note that this technique is not applicable to F.M. sound N.T.S.C. systems.

10.22. Summary

In this chapter we have seen that the receiver's reference sine wave which has to be locked in frequency and phase to the transmitter's sub-carrier sine wave can be derived from the transmitted burst signal.   Certain specifications as to the accuracy of the static and dynamic phase of the reference sub-carrier must be met to reduce overall hue errors and hue fluctuations due to noise.

The transmitted burst signal must be gated out of the sub-carrier waveform to prevent interference from the picture content, and the resulting gated burst waveform is then suitable for reference frequency generation.

The reference frequency generator itself is the equivalent of a very narrow (of the order of 100 c/s) band-pass filter tuned to the sub-carrier frequency, and it may be passive or dynamic in form. Most generators used at present are dynamic integrators in the shape of A.P.C. loops, and they are either crystal oscillator controlled or two mode quadricorrelators.   Either extreme oscillator stability or two mode action is necessary so that fast pull-in or stabilization time may be achieved simultaneously with good noise performance.

Besides providing a reference signal, most reference generators also supply a D.C. bias voltage which can be used for switching off the receiver's chrominance channel during monochrome transmissions (colour killing) and for automatic gain control of the chrominance channel (A.C.C.).

CHAPTER **11**

Operation of the Shadow Mask Tube

11.1. Introduction

Most of the colour receiver circuits so far described can be used
with any type of display device, although it is usually more econo-
mical to operate single-gun tubes with their own special types of
decoding circuits. This chapter is concerned with those circuits
which are peculiar to the shadow mask tube. The basic principles
of construction and operation of this three-gun display tube have
been outlined in Section 3.8. Here we shall be concerned with the
details of the time base, convergence and bias circuits for a typical
shadow mask tube, or aperture mask tube as it is sometimes called.

11.2. Electrode voltages

Each of the three electron guns of the shadow mask tube is of
tetrode construction and electrostatically focused. Separate con-
nexions are brought out for the cathode, modulator grid and screen
electrode of each gun. The 6·3V, 0·6A heaters are connected in
parallel and the focus electrodes are all common to the one terminal.
The phosphor screen, mask and final anodes are operated at 20 to
25kV with a maximum total beam current of 800 to 1,500μA. The
focus voltage is typically 4,000V. Both focus and e.h.t. voltage can
be obtained from the line time base. Since purity and convergence
adjustments depend upon the e.h.t. voltage, whilst varying beam
current load can affect the line scan and boost voltage, it is necessary
to regulate the e.h.t. supply.

There is considerable choice of screen-cathode voltage, 400 to 600V
being usual, although higher screen voltages can give improved grey
scale tracking. The grid-cathode bias may be 50 to 150V, although
the actual D.C. voltage on the cathode with respect to earth will
depend upon the type of signal drive arrangement.

Typical circuits for providing bias controls are shown in Fig. 11.1.

11.3. Deflection yoke

The three guns of the shadow mask tube necessitate a larger
diameter neck than is usual with monochrome tubes. The R.C.A.
21CYP22 tube has a neck diameter of 2 in., which is to be compared
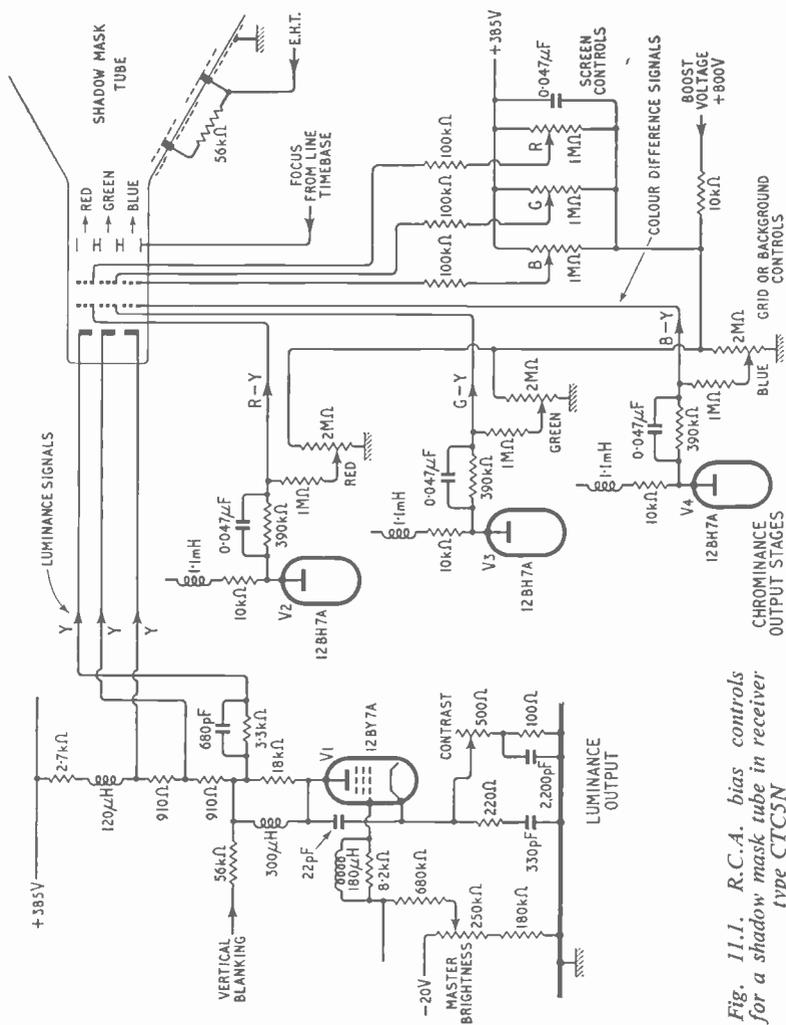
Fig. 11.1. R.C.A. bias controls for a shadow mask tube in receiver type CTC5N

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with the $1\frac{1}{3}$ in. for a monochrome 70° tube. The deflection yoke has, therefore, to fill a larger volume than normal with magnetic flux. At the same time, if all three beams are to be treated alike, field aberrations must be avoided over a larger volume than is tolerated with black and white tubes, and graded windings are essential. To further improve the uniformity of the field, the deflection coil window area is usually made even larger than the size fixed by the tube. Unwanted field effects due to the end turns of the yoke must be minimized by bending them sharply away from the tube neck. The angular separation of the three beams is only two degrees and external magnetic and electrostatic fields must be kept to a minimum. To prevent any stray field from the deflection coils reaching backwards into the convergence region, the deflection
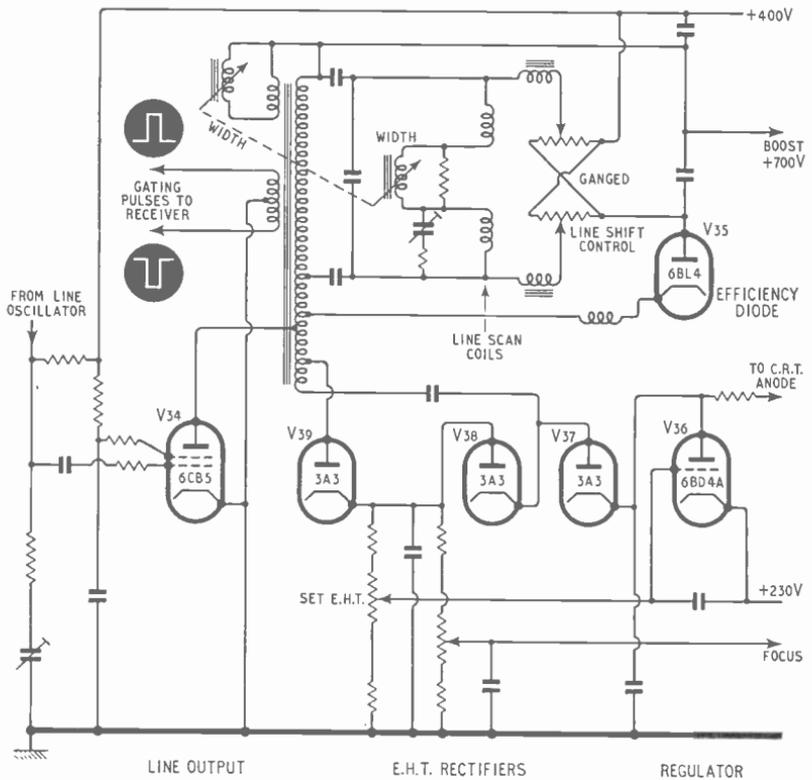
Fig. 11.2. Line time base of a Murphy colour receiver

yoke is screened from the convergence coils by means of a copper shield and a ring of high permeability magnetic material.

The equilaterally spaced guns are not symmetrical with respect to the scanning fields and the defocusing effects which occur with uniform deflecting fields produce convergence errors which are different for the three guns.   Since the blue image has the least luminance and produces the least distressing effects when it is in error, the guns are arranged to ease the convergence of the red and green spots by making the red and green convergence errors similar.   Capacitive unbalance effects may be different for the three guns and ringing on the line time base may be more marked than on monochrome.

Circuit characteristics of a typical R.C.A. deflection yoke are:

Horizontal deflection coils 11·8mH, 7Ω and a deflection angle of 70° for a peak-to-peak current of 1·7A.
Vertical deflection coils 115mH, 55Ω and a deflection angle of 55° for a peak-to-peak current of 0·5A.

11.4.   Line time base

The basic design of the line time base can be conventional. Either directly synchronized or flywheel oscillators can be used and the output stage can utilize an energy recovery diode circuit. Sample designs of line time bases are given in Figs. 11.2 and 11.3

The e.h.t. supply of the Murphy receiver is obtained from the voltage doubler circuit of V_{37} and V_{39}, the third rectifier V_{38} maintains the average positive potential on the anode of V_{37} at a value slightly below that of V_{39}.   This arrangement is more efficient than using a high resistance in place of V_{38}.   The e.h.t. voltage is stabilized by the shunt regulating action of V_{36}, whose grid voltage varies as the beam current of the cathode-ray tube alters the potential on the cathode of V_{39}.   The focus voltage is also obtained from the cathode of the first rectifier V_{39}.   The heaters of two of the rectifiers are fed from a separate transformer as there is not room on the core of the output transformer for more than one winding.   In this particular receiver the line time base also provides pulses for gating out the colour burst and for the colour killer circuits.

In the G.E.C. receiver the e.h.t. is produced by a single rectifier and the focus voltage by another rectifier.   The shunt regulator triode is controlled by a fraction of the boost voltage since this varies with the beam current loading on the output stage.

It is not practical to obtain picture shift by the simple means often adopted in monochrome receivers, such as a magnetic shuffle plate.   The shift field must be uniform over all three beams and

Fig. 11.3. Line time base of a G.E.C. colour receiver

must not interact with the convergence fields.   Vertical and hori-
zontal shifts of about $\pm$ 5 % of the picture size are normally obtained
by passing D.C. through the field and line deflection coils.   In
Fig. 11.2 two ganged potentiometers are cross-connected to allow
a picture shift either side of the central position.   The D.C. used
for picture shift is kept out of the line transformer by isolating con-
densers, and the shift circuits are prevented from short circuiting the
deflector coils by isolating inductances.

The design of the output transformer follows the general principles
developed for good quality monochrome receivers, but it is pre-
ferable to keep the flux density as low as is practicable.   The anode
circuit to deflector coil turns ratio may be 1·6 to 1 and the secondary
inductance may be of the order of 75mH.   The e.h.t. overwinding
needs careful design and insulation, of course, and may be encased
in a silicone rubber moulding.

All the e.h.t. points and connectors are housed in small anti-
corona shields and pressings so that they will not corona under
humid and dusty conditions.

Any high voltage stream of electrons will generate X-rays when
the electrons collide with metallic electrodes and similar structures.
The use of X-ray absorbing lead glass for valve and cathode-ray
tube envelopes minimizes the emission of X-rays outside the receiver
and all commercial designs are quite safe in this respect, giving con-
siderably less than the maximum permissible dosage even in the
immediate vicinity of the cabinet.

However, the receivers should not be operated for any length of
time without the protective glass over the tube face and service
engineers should not spend long periods close to unshielded line
time base rectifiers and shunt stabilizers when the chassis is out of
the cabinet.

The final anode coating on the inside walls of the R.C.A. 21CYP22
tube is split into two sections which are connected together through a
resistance of 50kΩ.   This is to safeguard the cathodes, should a
momentary internal arc occur.   The e.h.t. supply must always be
connected to the outer of the two coatings.

The e.h.t. smoothing is obtained by the capacity between the
final anode coating on the tube bulb and the graphite coating on the
outside.   A typical e.h.t. regulation is 5 %.   An interesting method,
due to R.C.A., of stabilizing the e.h.t., is to inject positive pulses
into the grid of the line output valve so that it conducts during the
flyback periods and absorbs some of the e.h.t. power.   The regulat-
ing pulse amplitude is controlled by the boost line and effectively
maintains the e.h.t. voltage constant.   However, this arrangement

Fig. 11.4. *Field time base of a G.E.C. colour receiver*

calls for increased dissipation in the output valve and the total time base current varies with the picture content.

Line time base interference, both radiated and mains borne, needs special preventive measures. The mains leads should be filtered and both the time base and the deflector coils should be as completely screened as is practicable.

11.5.   Field time base

The vertical deflection circuits are conventional except that increased output is called for, compared with 70° monochrome circuits. A representative circuit is shown in Fig. 11.4. The output transformer turns ratio is typically 7 to 1, with an output valve anode current of about 40mA.

11.6.   Purity

The alignment of the three electron beams in a shadow mask tube is readily carried out by adjusting the purity magnets and deflection yoke so that the electrons from any particular gun hit only one colour of phosphor dots (*see* Section 3.8). The geometry of the holes in the mask and the phosphor screen dots is such that the electrons must appear to come from one of three points, called the *colour centres*, if each beam is to produce a uniform picture in one of the primary colours. The process of deflection is such that any deflected beam appears to have passed through a point about half way along its undeflected path in the deflection yoke and such a point is called the deflection centre for that beam (*see* Section 11.7).

The purity magnets produce a transverse field of up to 10 gauss in the tube neck and are used to move the three beams bodily about in the neck. The deflection yoke must also be moved axially along the neck until the plane of the deflection centres coincides with the plane of the colour centres, when the purity magnets can be used to make the deflection centres coincide precisely with the colour centres. This coincidence can be judged by the colour purity of the screen with two of the guns blacked out. It is usual to leave the red field on, since this has the highest beam current, until the purity is as good as possible, finally checking for a compromise position on all three fields in turn.

The purity magnets correct both for tolerances in manufacture which may cause the initial paths of the electron beams to be in error so that they do not pass through the colour centres in the deflection plane, and for positional errors caused by stray magnetic fields, including the earth's field. Since the purity magnets act on all three beams they can only correct for uniform errors. The

effect of the earth's magnetic field depends upon both the strength of the field and the orientation of the tube in that field and purity may need to be corrected each time the receiver is moved to a new position. However, these effects are normally small and the beams are masked by the aperture mask to a size smaller than the phosphor dot separation so that there is a reasonable tolerance on purity.

Non-uniform effects may be produced after the beams have been deflected since they change direction relative to the earth's magnetic
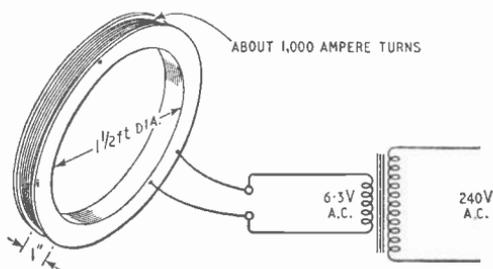
Fig. 11.5. *Degaussing coil*

field as they are scanned across the picture. Such effects may also be caused by large metallic masses near the tube and by inaccuracies in deflection yoke manufacture. They can be largely cancelled out by an empirical adjustment of the field equalizing magnets, which need only produce movements of the beam of $\pm$ 5 thousandths of an inch at the edges of the picture.

If these adjustments do not result in satisfactory purity the tube and its mounting should be demagnetized. This is readily carried out with a degaussing coil and should always be done when installing a new tube. A suitable degaussing coil (of 18 in. diameter), Fig. 11.5, consists of 28 turns of copper wire (14 S.W.G.) carrying an alternating current of 30A, which may conveniently be derived from a mains transformer. Alternatively, about 500 to 3,000 ampere-turns can be produced by using higher impedance coils operating from higher alternating voltage sources.

The degaussing coil, with the alternating current flowing through it, should be slowly moved around the front and sides of the tube and then gradually withdrawn several feet away from the receiver before switching off. In this way the alternating magnetic field is gently reduced to zero, which gives the most thorough demagnetizing effect. The purity adjustments should be checked again after the static convergence has been carried out.
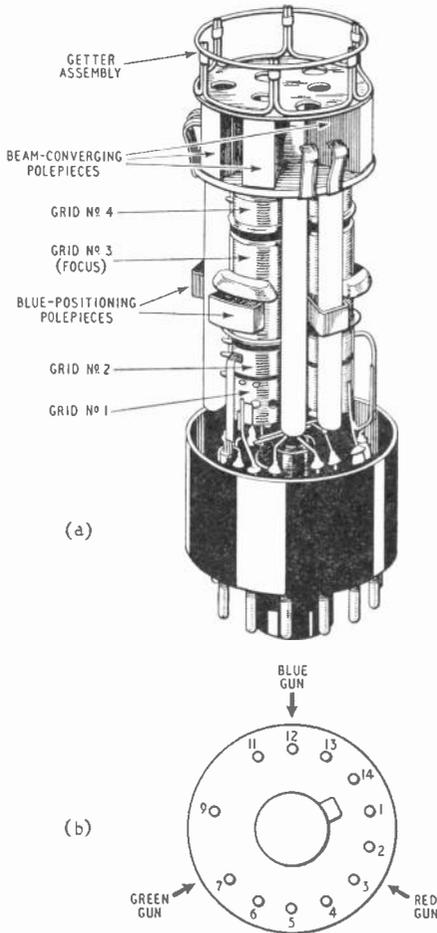
Fig. 11.6. (a) electron gun and (b) base connections of an R.C.A. shadow mask tube. Pin connections are:

1	Heater	7	Green screen G2
2	Red modulator G1	9	Focus grid G3
3	Red screen G2	11	Blue screen G2
4	Red cathode	12	Blue modulator G1
5	Green cathode	13	Blue cathode
6	Green modulator G1	14	Heater

11.7. Convergence

The convergence yoke must be so placed on the tube neck that the poles sit immediately over the internal pole pieces of the electron guns and the blue lateral shift magnet must sit over its corresponding pole pieces (*see* Fig. 11.6), which may be clearly seen through the glass neck.

The static convergence (*see* Section 3.8) is adjusted until the three beams precisely overlap at the centre of the picture, using if necessary the blue lateral shift magnet.   The magnetic field strengths necessary to achieve static convergence may be up to 7 gauss and can be produced either by permanent magnets or by passing D.C. through the vertical convergence coils.

Converging the three electron beams to produce coincident spots at the screen centre in the undeflected position is not, in itself, sufficient to ensure that the three spots remain coincident as the beams are scanned over the raster.   The electron guns are separated by about 1 in. and are tilted a little over 1° towards the tube axis so that the beams tend naturally to converge roughly at the screen centre.   The three electron beams therefore have slightly different paths through the magnetic deflecting fields and this results in the three spots separating out as they scan across the picture area.   The three beams may be thought of as parts of a very thick electron beam which suffers deflection defocusing as it is swept across the raster.   This lack of dynamic convergence has to be corrected by applying A.C. magnetic fields to the convergence pole pieces so that the angle between the beams is reduced as the deflection increases. To understand how the scan distortions vary between the three electron beams and why the convergence yoke currents have their particular waveshape, briefly consider the mechanism of magnetic deflection.   Non-uniformity in the deflecting fields and their interaction with stray magnetic fields may produce scan distortions as in monochrome television but only the case of three electron beams traversing a uniform deflecting field will be considered.

Any one electron beam experiences a force proportional to the current in the deflecting coils, and this force acts in a plane perpendicular to the direction of the magnetic field and at right angles to the direction of motion of the electrons.   The electrons therefore move along a circular path whilst they are in the deflecting field, as in Fig. 11.7(a).

When they leave the deflecting field they fly off in a straight line again, tangentially to their circular path.   The electrons then appear to have come from a point C inside the deflecting field half way along their deflected path, and for smallish angles of deflection this

(a)

RADIUS OA
PERPENDICULAR
TO INCIDENT BEAM

CIRCULAR PATH

DEFLECTED RAY PERPENDICULAR
TO RADIUS OB

ELECTRON BEAM

UNDEFLECTED RAY

DEFLECTION CENTRE

LIMITS OF
MAGNETIC FIELD
WHICH IS PERPENDICULAR TO PAPER

(b)

LINE SCAN
DEFLECTING FIELD

CONSTANT VERTICAL DEFLECTION

FLAT SCREEN
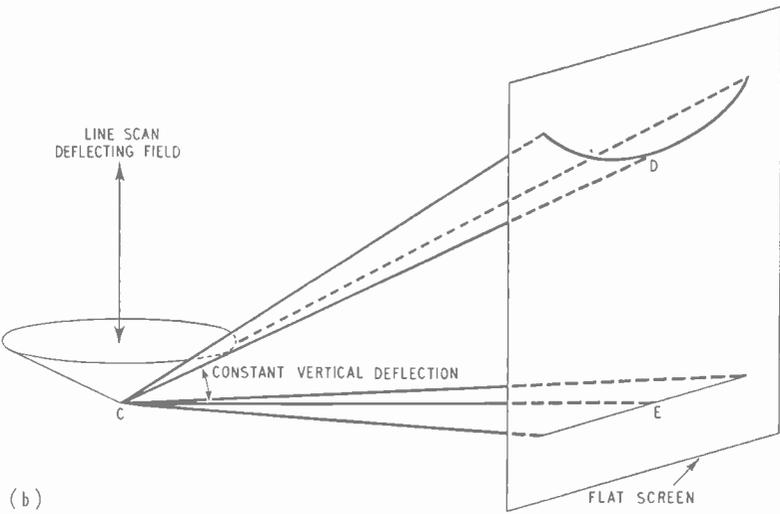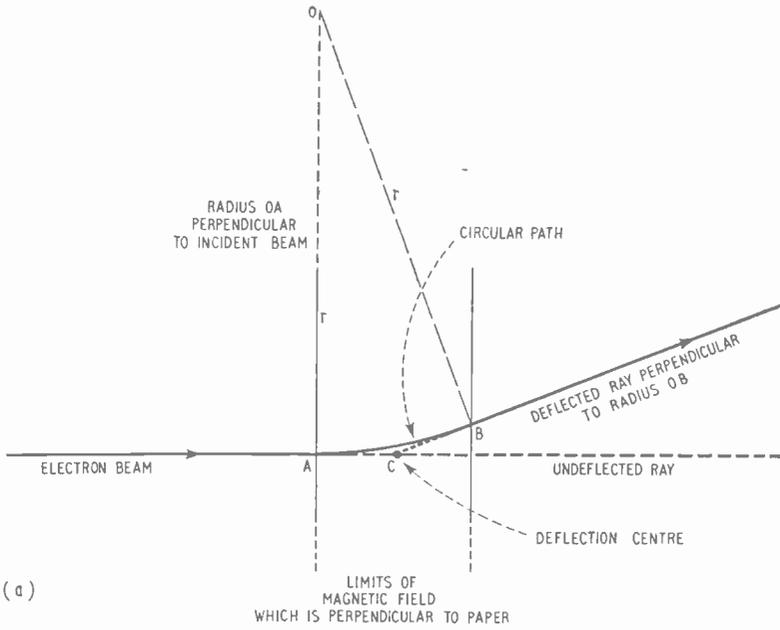
Fig. 11.7.   Magnetic deflection

point is a fixed position whatever the angle of deflection. This fixed point is called the *deflection centre*.

Consider a beam CD which is given a fixed vertical deflection, as in Fig. 11.7(b), and is then scanned horizontally. The beam traces out part of the surface of a cone whose apex lies at the deflection centre and whose axis is parallel to the magnetic field producing the horizontal deflection. If the beam strikes a flat screen, the line traced out on the screen is only straight for the special instance of a ray such as CE which is perpendicular to the magnetic deflecting field. In all the other cases the flat screen intersects the suface of the cone in a curve. This conic section is a hyperbola and the amount of curvature increases as the vertical deflection increases. Similar considerations apply to all four sides of the raster and give rise to the well-known pincushion distortion. If the screen is a sphere with its centre at the deflection centre then the pincushion distortion disappears and a truly rectangular raster is obtained. Practical tubes have spherical screens whose centre of curvature lies some way behind the deflection centre and the resulting raster produced by a uniform deflecting field shows some pincushion distortion, but less than with a flat screen.

For the case of the three-gun colour tube, each red, green and blue raster shows such pincushion distortion. The blue horizontal scan only traces a straight line for the special case when the scan passes through the horizontal projection on the screen face of the blue deflection centre, i.e. when the electron beam is perpendicular to the vertical magnetic field which is producing the horizontal scan. Similarly, vertical lines on the blue raster are only straight in the one case for which a vertical line passes through this same projection of the blue deflection centre.

Since the three guns have different deflection centres, the green and red guns are tracing out curved lines when the blue gun is tracing out a straight line, and so on, as in Fig. 11.8. Thus, although the three beams may produce one spot in the undeflected position, the paths traced out by each gun show a curvature which depends on the position of the spot with respect to the projection on the screen of the appropriate deflection centre.

For example, when the blue beam has such a fixed vertical deflection that it strikes the point X, Fig. 11.8, the horizontal scan then makes the blue spot trace out the solid curve shown. However, when the green gun is firing at the point X, or X', for which it needs a rather greater vertical deflection than the blue gun did, the horizontal scan traces out a green line which has rather greater curvature than the blue, and whose lowest point is offset from X to X' and is
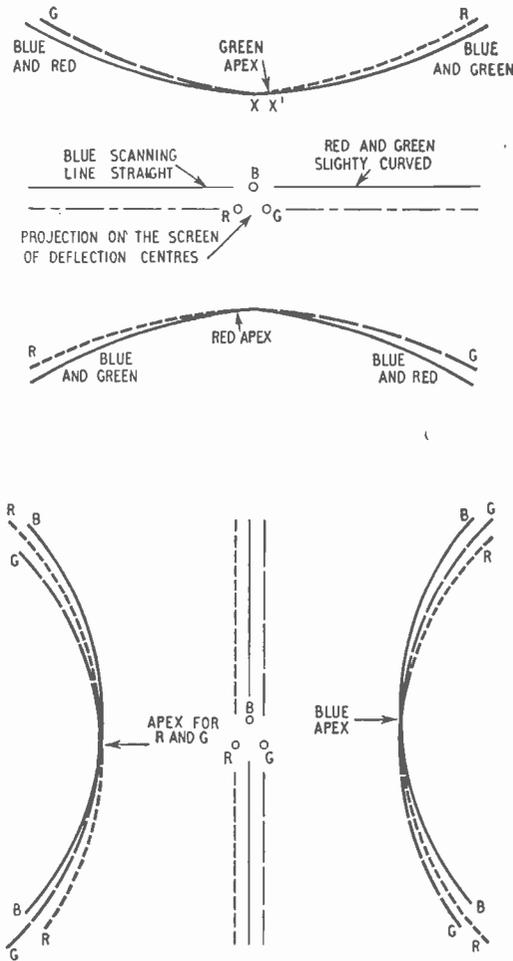
Fig. 11.8. Scan shape variations due to offset of deflection centres. The actual displacements are of the order of several millimetres

vertically above the green deflection centre.   Similarly, the red trace through the point X has the same curvature as the green trace but is displaced to the left of the blue trace.   For the case of a 70° shadow mask tube with a flat screen and the deflection centres indicated, the sideways shift of the red and green traces with respect to the less curved blue trace results in the red trace separating out on one side and the green trace on the other.   For rather greater separation of
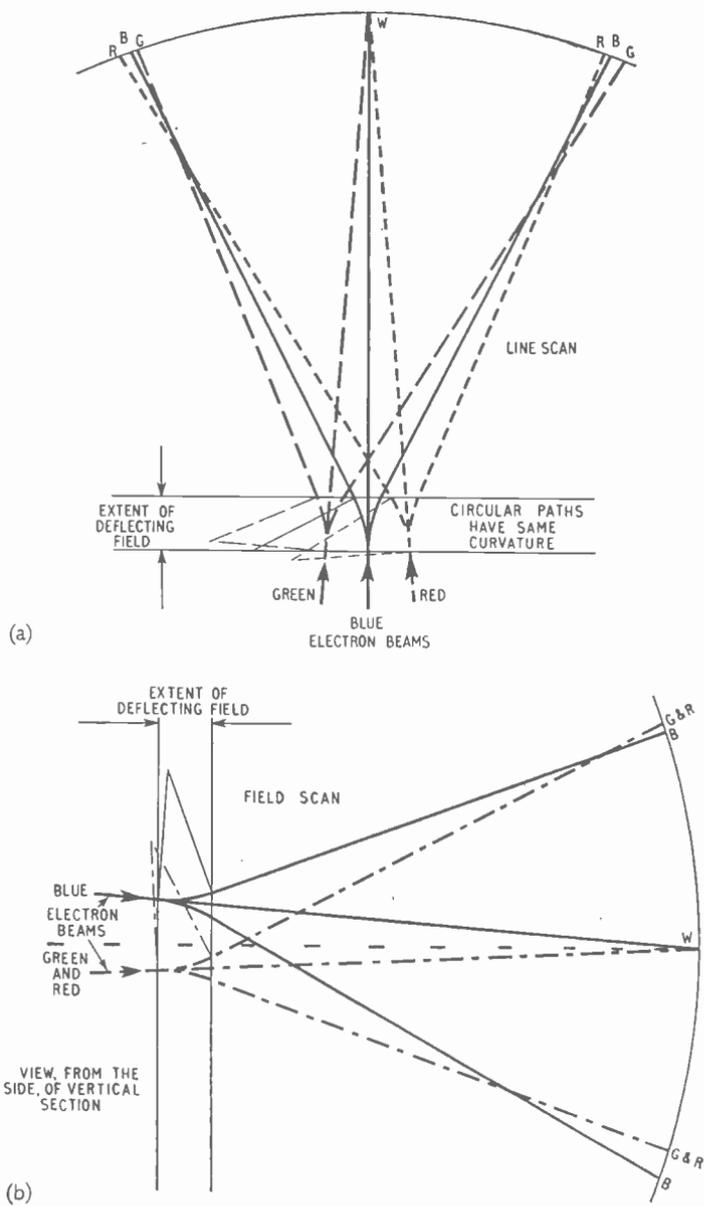
Fig. 11.9. Beam displacements due to tilt of guns (a) during horizontal scanning; (b) during vertical scanning

the deflection centres the green trace on the right hand side may fall below the blue line and the red trace on the left hand side may also fall below the blue trace.

For vertical lines traced out with constant horizontal deflection, the blue lines cross over the other lines.   The red and green traces touch at their common apex, but on the right hand side of the screen the red trace has the greatest curvature whilst on the left hand side the green trace is more curved than either the blue or red.

A further effect arises due to the tilt of the electron beams.   As the beams scan horizontally, they do not stay vertically above each other as they trace out their curved horizontal paths, the red scan leading the blue at any instant while the green spot lags behind. This effect is illustrated in Fig. 11.9 (a).   The blue beam is symmetrical viewed from above and is in line with the tube axis.   The tilt on the red and green guns causes the red and green beams to converge on the blue beam before the screen is reached and then to diverge again so that the spots separate out on the screen face.   Notice that the blue spot is not midway between the red and green spots.   Any two of the beams intersect in a point which traces out a paraboloid inside the tube bulb as the raster is scanned.

Similarly in the vertical direction, illustrated in Fig. 11.9(b), a timing error also occurs.   The red and green beams are equally offset and tilted and suffer the same vertical displacement.   The blue gun is tilted by a larger amount and the blue spot always appears below the red and green spots, the displacement being greater at the bottom of the scan than at the top.

The three beams trace out their respective rasters as in Fig. 11.8, but the time at which they pass any particular point depends upon the alignment of the electron guns.

Fig. 11.10 illustrates the combined displacement of the beams on a raster, when the red, green and blue spots converge at the centre of the screen.   The four triangles of dots give the relative positions of the three beams at instants when the scan is at the centres of the raster edges.   By applying appropriate currents to the convergence yoke coils the spots can be moved radially so that they are more nearly coincident.   The required current waveforms have approximately the shape of flat parabolas and are also sketched in Fig. 11.10.

For vertical deflection the red and green vertical convergence waveforms are the same, but the blue convergence current waveform is of opposite sign.   Both are asymmetrical with respect to the scan centre, or tilted.   In the horizontal case the red and green convergence currents are different, but both are still opposite in sign to the blue waveform.   Theoretically, perfect correction is not
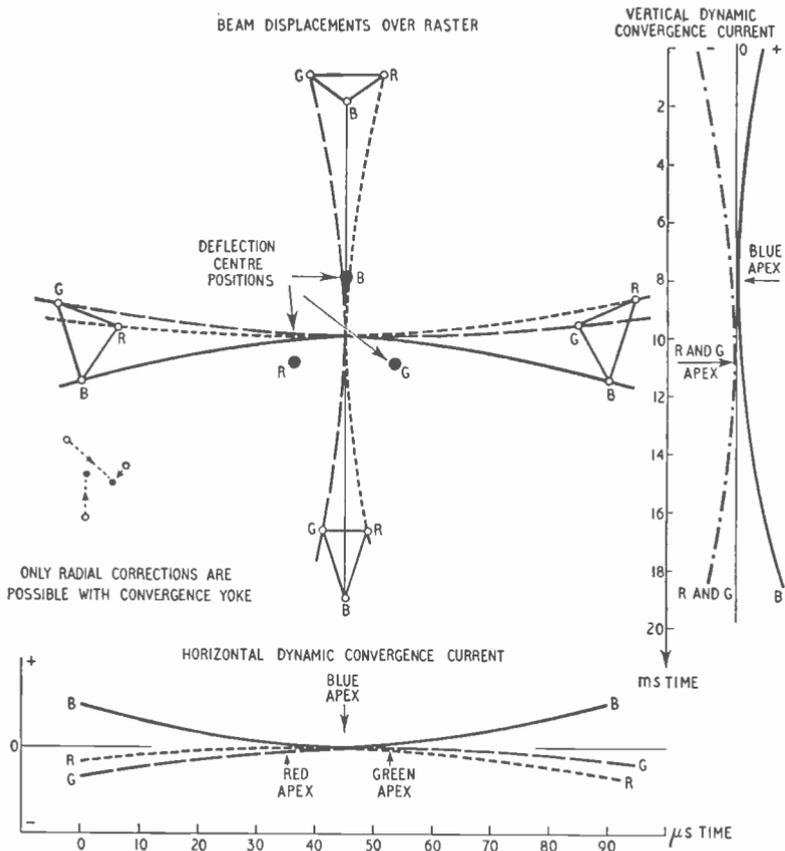
Fig. 11.10.   Total convergence errors and required corrections

possible in the horizontal direction because the colour dot triangles near the ends of the line scan are not isosceles. The dynamic convergence coils and their associated pole pieces can only move the dots along radial lines and these lines do not meet at one point. This is analogous to the problem of obtaining static convergence with the blue lateral shift magnet. To overcome this defect, some receivers use a dynamic blue lateral shift coil wound around the blue lateral shift magnet, as in Fig. 11.11.

In practice the time base sweeps are not precisely linear with time and the convergence waveforms which are derived from them must correct for the actual displacements of the spots. Similarly, convergence adjustments should only be made after the time bases have

been set up.   Normally parabolic waveforms are derived from the time bases and a tilt or sawtooth waveform added to produce the required asymmetry (*see* Fig. 11.12).   A few receivers have used sine wave convergence waveforms for the horizontal correction but this is only just tolerable.   The circuits used for convergence vary considerably, from those using several valves to those using none, and depend upon the type of convergence yoke used.   The basic principles of the wave shaping circuits used are familiar enough in general electronics and are summarized in Fig. 11.13.   Integration of a sawtooth waveform produces a parabolic waveform and repeated parabolas may be conveniently derived from various points in the time bases.   Early yokes used the same pair of bifilar windings for both vertical and horizontal convergence, and isolating inductors
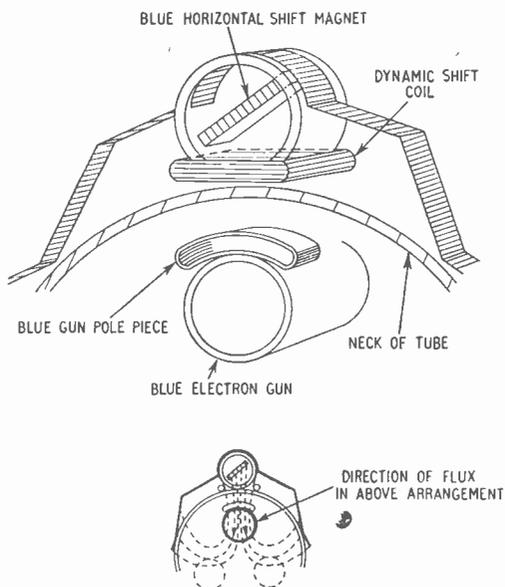
Fig. 11.11.   Blue dynamic lateral shift

were then needed to keep the line waveforms out of the frame circuits (Fig. 11.14).   Wave shaping circuits in the anode circuit of the field output valve derived both parabolic and sawtooth waveforms and the latter could be reversed in sign.   Although only one set of coils and controls is shown in Fig. 11.14 there were, of course, three sets.   For the horizontal convergence, voltage pulses from a
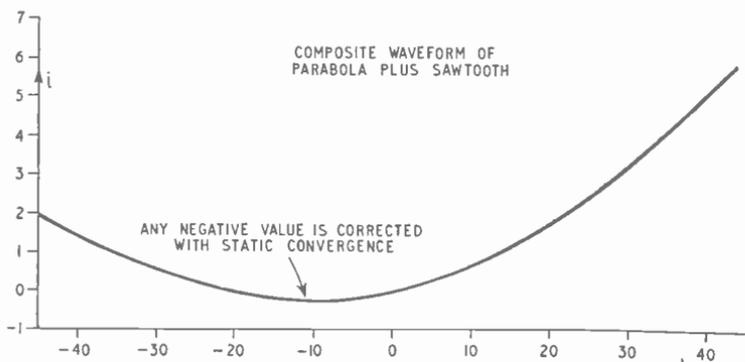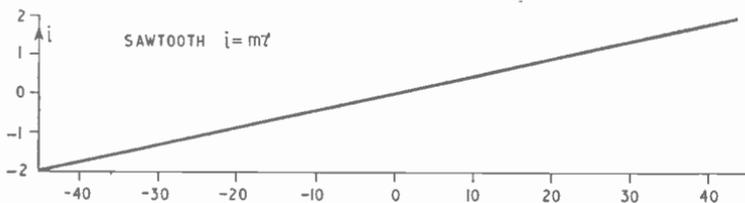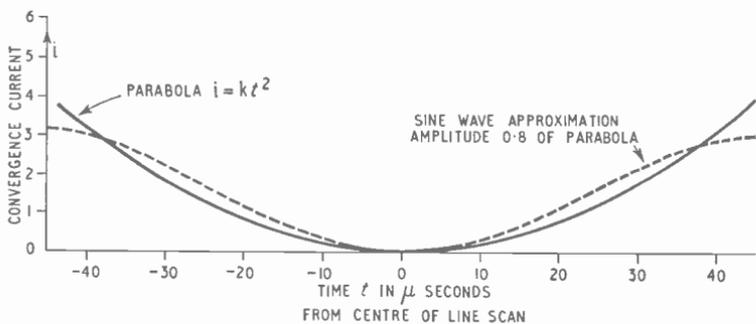
Fig. 11.12. Convergence waveform tilt by adding sawtooth to parabola

INTEGRATION OF CONSTANT CURRENT
$i_1 = k_1$
GIVES A SAWTOOTH WAVEFORM
$\int i_1 dt = k_1 t$

CIRCUIT ANALOGY

INTEGRATION OF SAWTOOTH
$i_2 = k_1 t$
GIVES A PARABOLIC WAVEFORM
$\int i_2 dt = \frac{1}{2} k_1 t^2$

$L\frac{di}{dt} = v$
$Li = \int v\, dt$

SAWTOOTH VOLTAGE

L

PARABOLIC CURRENT

$v = cq$
$= c\int i\, dt$

SAWTOOTH CURRENT

C

PARABOLIC VOLTAGE

q IS CHARGE ON C
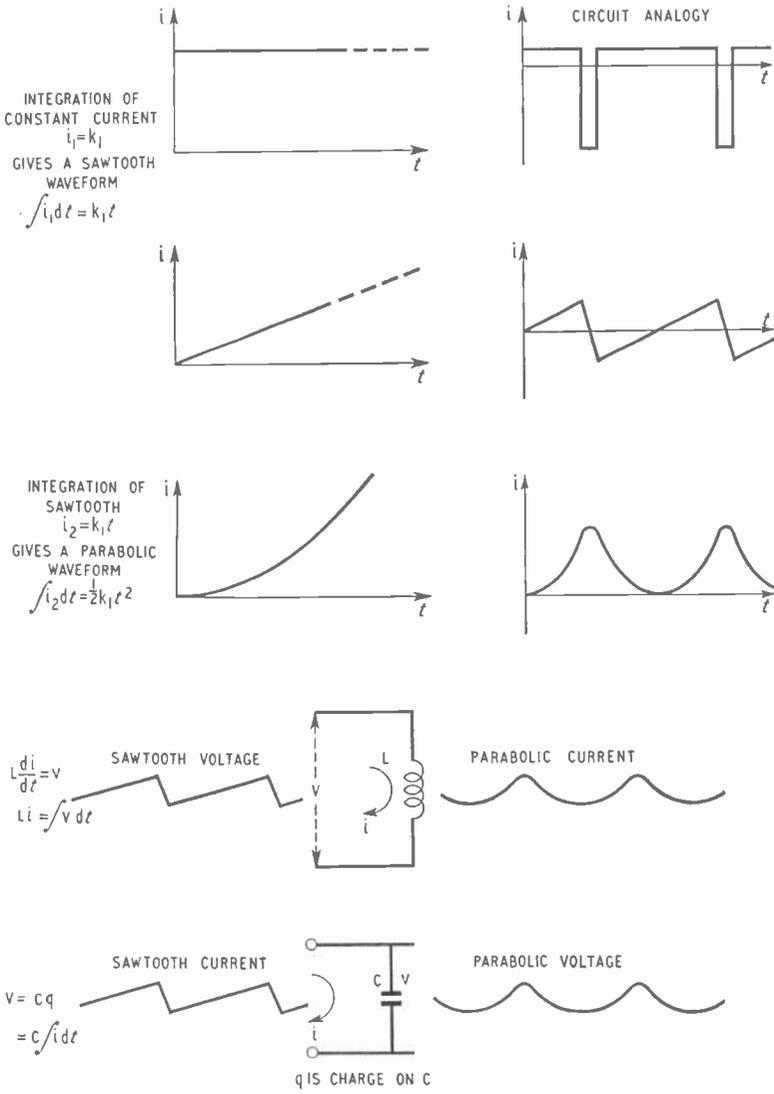
Fig. 11.13.   Derivations of parabolic waveforms
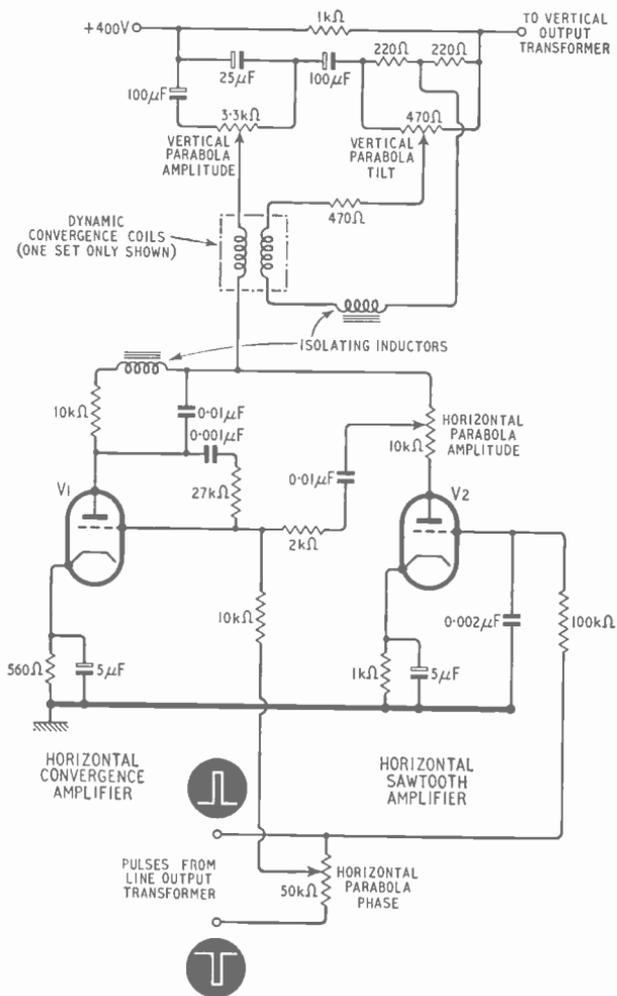
Fig. 11.14. Early R.C.A. convergence circuits

special winding on the line time base output transformer were converted to sawtooth waveforms, amplified, and then further amplified and integrated to produce parabolas.

Later convergence yokes have separate windings for the horizontal and vertical convergence, as shown in Fig. 11.15, and this obviates the need for isolating inductors.   The $\frac{L}{R}$ ratio of these coils is high so that they provide a degree of current integration directly.   A simplified version of the convergence circuits used in one receiver with this type of convergence yoke is shown in Fig. 11.16.   Positive going flyback voltage pulses from the line time base transformer (*see* Fig. 11.3) are integrated at the point A by L_1, C_1 and R_1. The resulting sawtooth voltage is applied to the convergence coil L_L where it produces a parabolic current waveform.   At the line scanning frequency the impedance of L_2 is of the same order as that of R_1, whilst at harmonics of the line frequency the impedance of L_2 becomes relatively larger.   The result is to shift the phase of the fundamental of the parabola relative to its harmonics and produce a tilt on the waveform.

A sawtooth voltage derived from the cathode circuit of the frame output valve (Fig. 11.4) produces in the convergence coils L_F a parabolic current waveform.

Small pulses of either positive or negative polarity from the tilt windings on the frame output transformer (Fig. 11.4) are also applied to the convergence coil L_F to produce an adjustable sawtooth tilt to the vertical parabola.

D.C., conveniently obtained from the line output valve, is passed through the centre tap of the tilt winding to the vertical convergence coil to provide static convergence.   The static convergence may be adjusted either side of zero by R_2.

The complete convergence circuit is shown in Fig. 11.17 and should be read in conjunction with Figs. 11.3 and 11.4.   The blue lateral shift sawtooth convergence current waveform is provided by pulse voltages from the line time base which produce a sawtooth magnetic field acting vertically on the blue lateral pole piece.   The permanent magnet in the blue lateral shift assembly is used to bias the dynamic field about the required mean position.

To ease the setting up of the convergence circuits, it is possible to D.C. re-insert on the apices of the parabolic waveforms so that the static convergence does not need constant readjustment as the dynamic fields are being adjusted.   Fig. 11.18 shows an R.C.A. circuit, in simplified form, which does this.   The essentially similar red and green horizontal convergence adjustments are common.
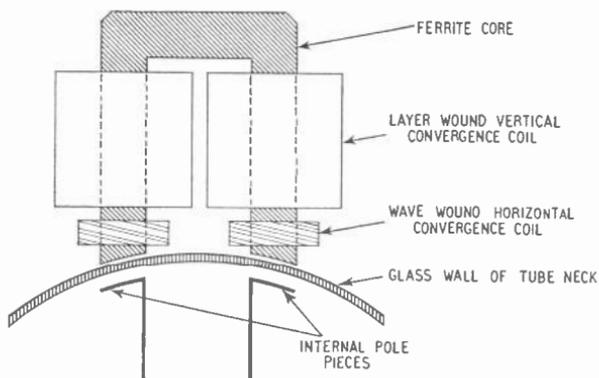
Fig. 11.15.   Dual winding convergence magnet
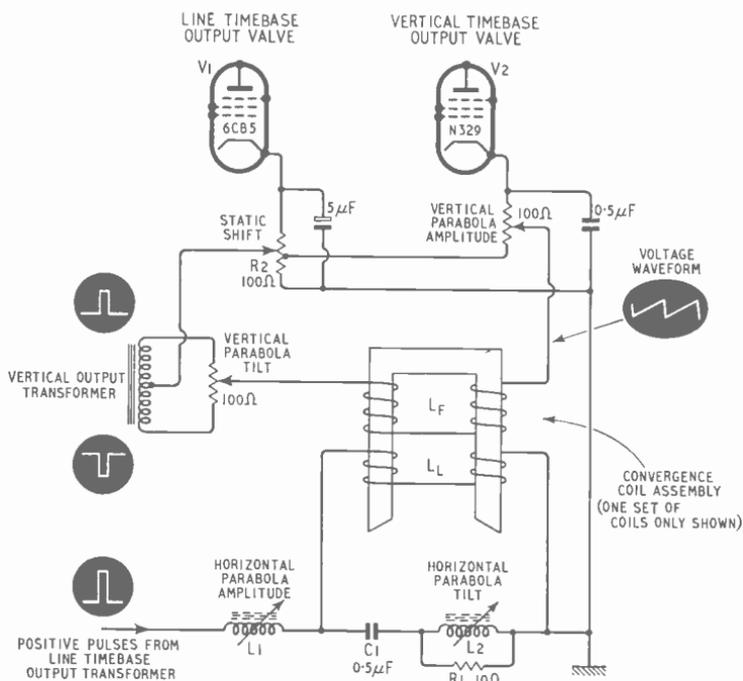
Fig. 11.16.   Simplified passive convergence circuit
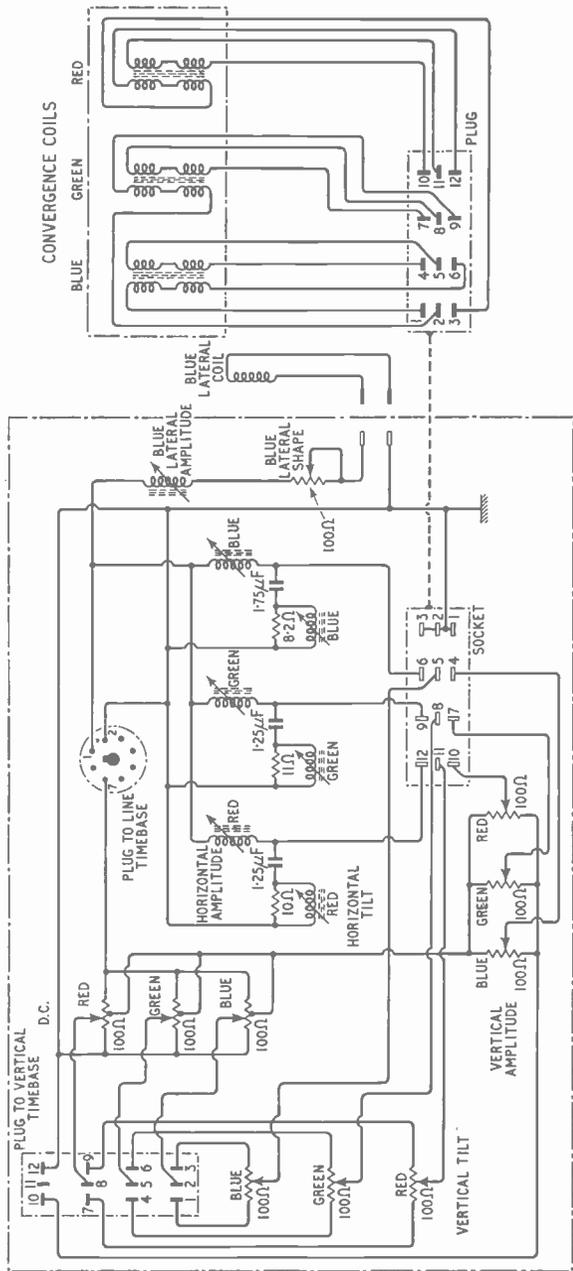
Fig. 11.17. Complete passive convergence circuit

11.8. Grey scale tracking

A typical curve of modulator grid-cathode voltage against beam current or screen brightness for a shadow mask tube gun is given in Fig. 11.19. The red, green and blue screen grid potentials are used to adjust the gun g_m characteristics so that they have the same shape. The modulator grid brightness controls or sit or background controls as they are variously termed, are then adjusted so that the three characteristics overlap, so to speak, and all three guns black out together.

The drive waveforms to each grid-cathode space are proportioned to the ratios required by the differences in phosphor efficiencies. Typical ratios are R to G to $B = 1$ to 0.8 to 0.6. The modulator grid input capacitance is 7pF for each gun and the cathode input
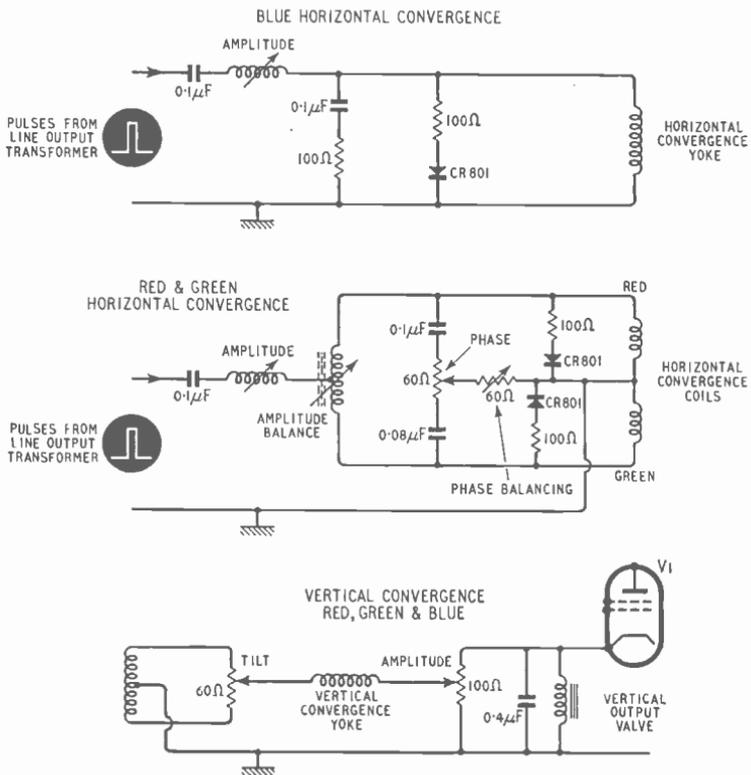
Fig. 11.18. Simplified R.C.A. convergence circuits using D.C. restoration
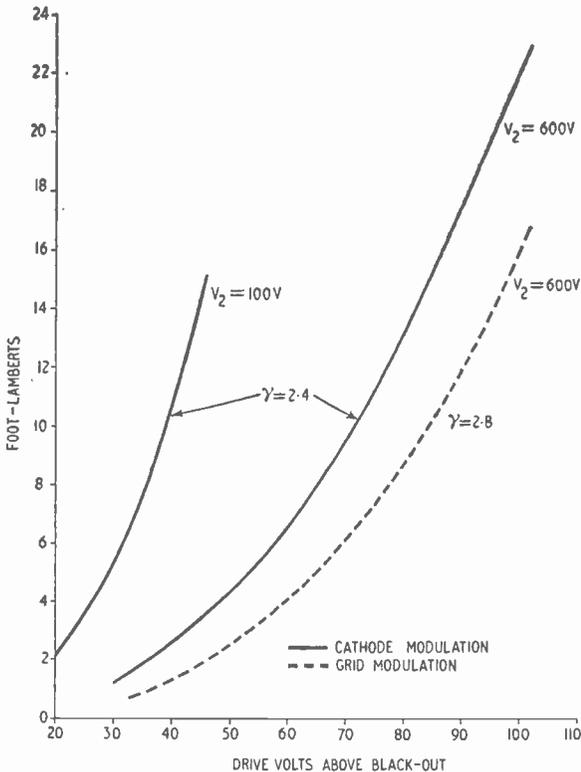
Fig. 11.19. Typical input-output curve for shadow mask tube. Final anode volts are 25 kV and the tube white is cold blue white. V_2 = *screen volts above cathode for zero drive*

capacitance is 16pF for the three guns combined. When the tube itself is used to matrix the luminance and colour difference signals to the tristimulus values, it is normal to apply the luminance signal to all three cathodes and the colour difference voltages to the grids. In this way the relatively large average luminance signal does not produce varying currents in the three colour difference circuits, which may have different impedances. Also, the colour difference circuits are narrow band and would provide frequency selective feedback if used in the cathode circuit.

Adjustment of the grey scale tracking is the most critical of the setting up procedures with the shadow mask tube. Particular care must be taken that the dark greys are neutral in colour, with the chromaticity of Illuminant *C*, since errors here produce a more

subjectively objectionable effect than colour casts in the white high-lights. Although setting up can be carried out on a picture, a gradation step wedge such as the centre blocks of Test Card C is preferable. The eye is very sensitive to changes in white balance over the brightness scale and it is difficult to measure the changes in gun g_m which produce noticeable visual effects. A detailed account of grey scale adjustment is given in Chapter 13.

11.9. Summary

The practical circuit details inherent in shadow mask tube opera-tion have been discussed. The absorption of electrons by the mask necessitates the use of a high e.h.t. voltage to obtain adequate brightness. This, together with the larger tube neck, calls for increased scanning power. The time bases must be reasonably linear if good convergence is to be obtained and slight errors in convergence make a noticeable difference to picture definition before the colour fringing effects become objectionable as such. It must be remem-bered that magnetic fields affect the colour purity and although a permanent magnet waved near the tube face produces the most interesting colour patterns, some of these patterns will probably stay there when the magnet is removed.

Grey scale tracking is very important and the grey colour chosen must be close to the reference colour, Illuminant C, if the colour reproduction is to be correct. In particular, green casts have objectionable effects on the complexion of the actors in a scene and such incorrect flesh tones must not be corrected by means of the hue control.

The reader is reminded to be careful about the danger of X-rays.

Colour Receiver Test Equipment and Performance Measurements

12.1. Introduction

Many of the instruments required for measuring the performance of colour receivers will already be possessed by organizations which are equipped for adequate performance testing of monochrome receivers. The additional apparatus required for colour work is chiefly concerned with accurate determination of frequencies and phases in the region of the sub-carrier (e.g. 2·7 Mc/s in the 405-line system).

In this chapter a list is given of the apparatus required for colour work, from the point of view of the repair shop and the performance testing laboratory. The use of the equipment is then discussed.

It is anticipated that all the apparatus mentioned in what follows either is or will be commercially available, but in case this is not so, two essential items, namely, a crystal controlled sub-carrier frequency oscillator and a calibrated phase shifter, will be described in detail so that the engineer may construct them for himself.

It is, of course, assumed that a radiated Test Card C or D and a colour signal will be available.

12.2. Servicing equipment required

From the point of view of the service engineer, the following apparatus will be necessary. The list is not necessarily arranged in order of importance.

1. General purpose high impedance (20,000 ohms/V) meter for voltage, current, and resistance measurement.
2. Valve voltmeter.
3. " Battery box ", or variable source of D.C. voltage up to about 10V.
4. Video oscillator covering the frequency range from about 50 c/s to 5 Mc/s, output at least 1V.

5. Signal generator covering the R.F. and I.F. bands, output about 100mV, and preferably capable of 100% modulation by an audio frequency square wave.

6. A swept oscillator or "wobbulator" for the R.F. and I.F. bands, with a frequency sweep adequately covering the video band.

7. An oscilloscope with a Y amplifier response at least up to the highest video frequency, and with an X amplifier input terminal. Ideally, the X amplifier should have as good a frequency response as the Y amplifier, though an X amplifier bandwidth up to about 0·5 Mc/s may suffice. Time calibration "pips" of 1μs are very useful for checking burst gate widths.

8. An accurate wavemeter (of the BC221 type, for example) for measuring sub-carrier frequencies.

9. A (monochrome) television waveform generator which can provide
   (a) About 5 different luminance levels or "greys" for checking colour balance tracking.
   (b) A thin white grid on a black background, consisting of about 16 equally spaced vertical lines and 12 equally spaced horizontal lines. This type of signal is essential for convergence adjustment.

10. A crystal controlled sub-carrier frequency oscillator having an output of at least 1V.

11. A calibrated phase shifter giving 360° maximum phase shift at sub-carrier frequency.

12. A degaussing coil for demagnetizing shadow mask tubes. This is easily constructed, as described in Section 11.6. Constructional details for items 10, and 11, which are easily made, are given below in Sections 12.4 and 12.5.

12.3. Performance measurement equipment required

For thorough investigation of colour receiver performance, the following items should be added to the above list.

1. A group delay equipment, or some means of measuring phase characteristics.

2. A travelling microscope for time base linearity measurements.

3. A colorimeter, such as the Harrison colorimeter, for purity and colour balance measurements.

4. An Illuminant *C* reference for setting the correct white chromaticity.

5. A local source of colour signal, such as a colour bar signal, available at video and R.F. and with the correct frequency lock between the synchronizing and sub-carrier signals. C.W. sub-carrier should also be available, and a sound carrier with the correctly locked frequency spacing from the sub-carrier. (This is discussed in Chapter 6.)

6. For design work, a colour picture source such as an *R*, *G* and *B* flying spot scanner is highly desirable.

7. A vectorscope for measuring the amplitude and phase of the modulation carried by the quadrature modulated sub-carrier. This instrument is chiefly of use for checking the accuracy of transmitter encoders.

A block diagram of one form of vectorscope is shown in Fig. 12.1. The composite video waveform to be checked is passed through a band-pass filter which accepts the modulated sub-carrier, and this is then synchronously detected in two detectors operating along the
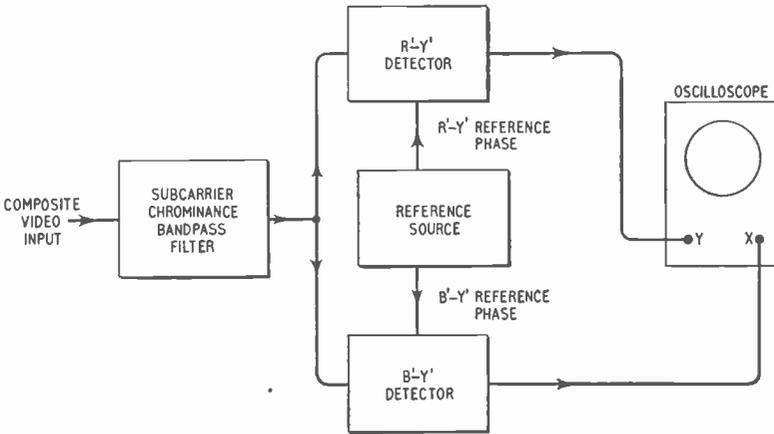
Fig. 12.1.   Vectorscope block diagram

$(R' - Y')$ and $(B' - Y')$ axes. The reference source for the detectors may be obtained either from the transmitter's C.W. sub-carrier oscillator, or from a reference generator locked to the incoming burst.

If the two detectors have equal gains, then the oscilloscope display will show the appropriate amplitude and phase of the original

chrominance signal as plotted on the $(R' - Y')$ and $(B' - Y')/$ 1·78 diagram.

█Normally the vectorscope is used for displaying the vectors corresponding to the colour bar waveform, and this gives a quick and accurate check of the encoder if a scale calibrated with the
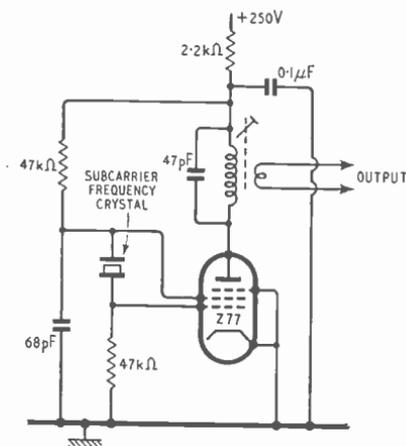
Fig. 12.2. Sub-carrier frequency crystal oscillator

correct amplitudes and phases of the three primary and their complementary colours relative to the burst is placed over the display.

12.4. Crystal controlled sub-carrier oscillator

A suitable circuit for a crystal controlled oscillator is shown in Fig. 12.2. The oscillator has an electron coupled output from the anode of the pentode, the oscillation occurring between the control and screen grids. The circuit shown gives about 5mA peak of anode current for a 2·65781 Mc/s crystal, suitable for the 405-line N.T.S.C. system. Normally, the oscillator is required to feed a load of about 70 ohms, so with a ten to one step down transformer the anode load is about 7kΩ, which results in an output of about 3·5V peak, or approximately 2·5V r.m.s. This oscillator is of most use if the crystal can be made to oscillate within a few cycles of the actual sub-carrier frequency, since it can then be used for testing reference generators as well as for phase measurements.

12.5. Calibrated phase shifter

A calibrated phase shifter can easily be made from suitable lengths of co-axial cable of the type often used for television aerial downleads. The principle of such a phase shifter relies on the time

delay which occurs when a signal is passed down a characteristically terminated cable. If a sine wave voltage is applied to a cable of such a length that a delay equal to the period of one cycle of the sine wave occurs, then the cable output sine wave will lag the input by one cycle, or 360°. Clearly, fractions of 360° may be obtained by using shorter lengths of cable, and a convenient arrangement consists of various cable lengths which can be switched in or out between input and output.

A diagram showing one possible switching method is shown in Fig. 12.3(a), in which the total length of cable used is equal to one cycle of delay, or 360°. By appropriate choice of switches, any phase delay from 0° to 360°, in 1° steps, may be obtained. However, a rather more convenient arrangement which saves some mental arithmetic is shown in Fig. 12.3(b), in which the total length of cable required is equivalent to 365°. The extra 5° of cable is certainly well worth while.

The total physical length of cable required for a 360° phase shift will depend on the effective transmission velocity in the cable used, and the actual frequency at which the phase shift is required. It is important to remember that the cable phase shifter will read phase correctly for only one value of frequency, though a correction factor can be employed for other frequencies. For example, for a frequency equal to twice that for which the device is designed, the phase readings must be multiplied by two. Also, the phase reading will be correct only if the cable is characteristically terminated. It is shown in Appendix 8 that a 1% mismatch produces a maximum phase error of about $\frac{1}{3}$°.

The transmission velocity of most types of co-axial cable is about $0.6 \times 3 \times 10^8$ m/s, that is, 180 m/μs. For a sub-carrier frequency of 2·66 Mc/s, the delay required for one cycle is $\frac{1}{2 \cdot 66}\mu$s, so that the length of cable required for a 360° shift is $\frac{180}{2 \cdot 66}$ m, or about 75 yd.

However, for any given cable it is advisable either to consult the manufacturer concerning the transmission velocity, or alternatively a rough check on the delay per yard may be made by measuring the width of the rectangular waveform (on a time calibrated oscilloscope) produced when a square wave input is applied to a known long length of short circuited cable. This width will equal twice the delay of the line, as shown in Fig. 12.4.

Calibration of the phase shifter can be carried out by first accurately determining the physical length of line required to give a 360° phase
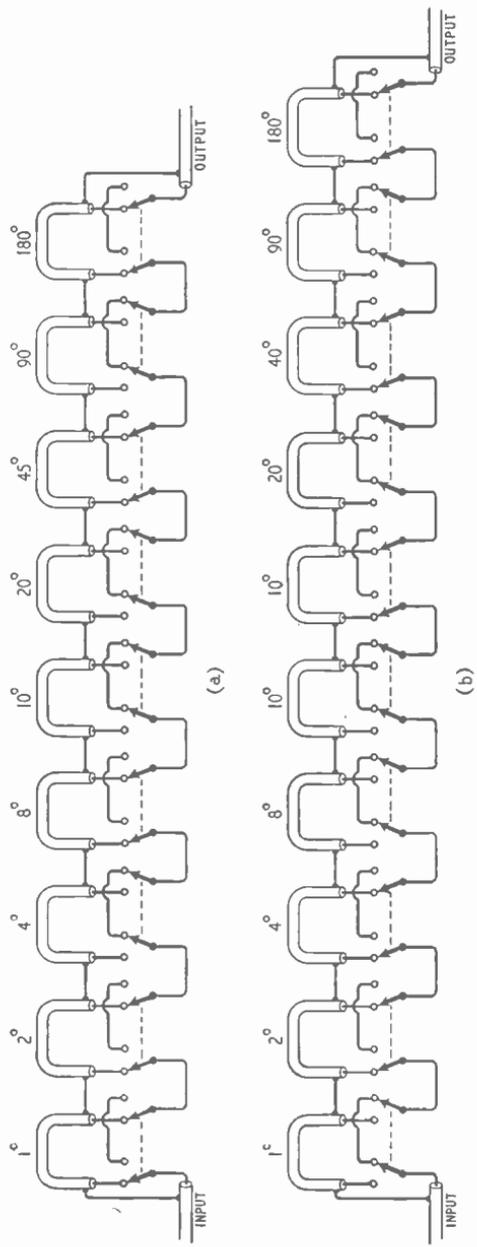
Fig. 12.3. Delay cable phase shifter (a) shows a 360° total delay and (b) a 365° total delay. Both show a 236° delay.

shift, and then cutting the line in proportionate measured physical lengths; for example, for 1°, a fraction ($\frac{1}{360}$) of the 360° length is required.   This makes the reasonable assumption that the cable has uniform characteristics.

The physical length of cable required for a 360° phase shift is best determined by measuring, at the appropriate frequency, the reactive component of the input impedance of a short circuited length of cable which is known to be near to (obviously preferably in excess
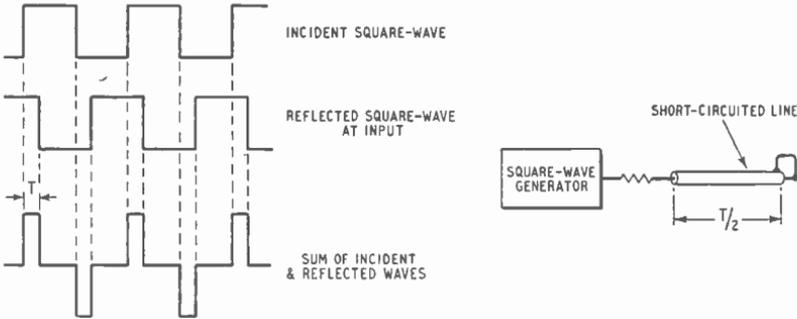
Fig. 12.4.   Approximate measurement of the delay of a coaxial cable

of) the length which corresponds to one wavelength at the required frequency.   The correct length is obtained when the input impedance is purely resistive.

(The input impedance of a short circuited line having a characteristic impedance Z_0 is

$$\frac{Z_0\left(\tanh \alpha\, l + \text{j} \tan \dfrac{2\pi}{\lambda}\, l\right)}{1 + \text{j} \tanh \alpha\, l \tan \dfrac{2\pi}{\lambda}\, l}$$

where α is the attenuation constant, λ is the wavelength of the signal in the line, and l is the length of line.   If $l = \lambda$, the reactive terms become zero.)

However, if an impedance bridge is not available, an alternative method of measurement consists of first arranging for a straight line Lissajous figure to be displayed on an oscilloscope using a voltage source of crystal controlled sub-carrier frequency.   Some phase shift may have to be introduced (for example, by way of a suitable R.C. circuit) to allow for phase shift in the oscilloscope circuits.

Having arranged for a straight line Lissajous figure to be displayed from a source of sub-carrier frequency, the estimated 360° length of

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correctly terminated cable is introduced into the X or Y path to the oscilloscope, and the length is adjusted until a straight line Lissajous is again obtained.

12.6. Luminance channel performance

Satisfactory operation of the luminance channel of a colour receiver requires adequate amplitude versus frequency response and a sufficient output voltage to drive the display tube.   In addition, the luminance delay must be such that luminance transients are centred as accurately as possible within the corresponding chrominance transients.

The receiver frequency response up to the conventional diode detector may be measured in the same way as for conventional monochrome receivers, such as by feeding an R.F. signal into the aerial socket and measuring the diode current for various frequencies covering the R.F. band.   Alternatively, of course, a wobbulator may be used.

Again, measurement of the luminance video stages may be carried out by using the techniques of monochrome video stage measurement but it is recommended that a wobbulator method be employed so that any reflections caused by the luminance delay line may be easily observed, together with the luminance notch introduced by the sub-carrier frequency rejector which is required to remove large area sub-carrier dot structure.

Direct measurement of the luminance video channel is difficult in practice because it requires a video signal to be introduced in the diode detector circuit, and the appropriate generator source impedance corresponding to the detector impedance must be included. This impedance is not easily determined and it is therefore better to derive the characteristic of the luminance video stages by taking an overall characteristic from aerial input to luminance output, and then to subtract from this the response of the R.F. and I.F. stages up to the detector.

The overall response may be conveniently measured by first investigating the D.C. and low frequency response, and then the high video frequency response.   In these measurements, the A.G.C. action must be rendered inoperative by connecting a suitable negative bias to the A.G.C. line.

To determine the D.C. and low frequency response, a vision carrier frequency signal is fed into the aerial socket, and the luminance output is displayed on a D.C. coupled oscilloscope.   The D.C. change produced by switching the signal on and off is then noted, and compared with the peak-to-peak value of the displayed output

when the same R.F. signal is modulated 100% by an audio frequency square wave, for example. The ratio of these two measurements gives the D.C. to A.C. gain ratio of the receiver. Note that if a D.C. coupled oscilloscope is not available, the D.C. change may be measured by means of a high impedance voltmeter, and of course in this case the oscilloscope trace of the modulation waveform must be calibrated. It is advisable to arrange for the D.C. output change to be about half that produced by a typical picture to ensure that overload is avoided.

The high video frequency response may be compared with that at low frequencies by feeding two signals into the aerial socket and measuring the luminance output. Thus, a vision carrier frequency is added in a resistive pad to another carrier whose frequency can be varied over the R.F. pass-band, and the corresponding beatnote luminance output may be measured by means of a valve voltmeter or oscilloscope, either of which must be capable of registering the highest video frequency. An alternative and more satisfactory arrangement is to replace the variable frequency generator by a wobbulator, when an oscilloscope display of the luminance output will have an envelope corresponding to the overall response, again provided that the oscilloscope frequency response is adequate.

The maximum voltage output of the luminance amplifier may be determined by feeding a video signal into the detector circuit and measuring the corresponding output on an oscilloscope or valve voltmeter. While this method will give no reliable indication of luminance video frequency response (because the impedance of the feed will almost certainly not be correct), the maximum output performance will not be affected. The measurement should be carried out for a D.C. video input and also for an input of the highest video frequency, for it is possible for a video amplifier to have good high frequency response for low amplitude signals but not for large amplitude signals. This effect can occur if a cathode compensated video valve does not have a large enough current swing available to charge the capacitance by-pass.

The delay introduced by the luminance delay line can be measured by feeding a square wave or sharp pulse into the detector circuit, and displaying simultaneously on an oscilloscope the original signal and the luminance output. A single-beam oscilloscope may be used if the two signals to be compared are added together in a suitable high impedance capacitance pad.

However, the actual value of the time delay required will depend on the bandwidth of the chrominance channel, and the best way to check the timing is to use a colour signal from a picture source which

consists of a vertical strip, the top third of which is red, the middle third green and the bottom third blue, all on a black background. The width of the strip should be such that the chrominance signals produced correspond to the highest chrominance frequency which the receiver is capable of resolving. By looking at the picture display of such a signal, the relative positions of the luminance and chrominance signals can be easily checked to see if the delay is too long or too short, and whether it is the same for all three colours.

12.7. Chrominance channel performance

As in the luminance channel, the amplitude versus frequency characteristic and the maximum available output voltage are of interest in the chrominance channel of the receiver, that is, the section of the receiver which accepts the modulated sub-carrier signal from the composite video and ultimately provides difference signals either to a matrix circuit or to the display tube.

Additionally, the effective gain and detection axes of the synchronous detectors and the A.C.C. performance, need to be investigated.

It should be noted that the measurements required of a chrominance channel will depend on the type of circuits employed, and the bandwidths of the various amplifiers and synchronous detectors will be different for $I'Q'$ and equiband working. However, it is very probable that the majority of receivers will use the display tube as the adding matrix for the colour difference and luminance signals, and in the measurement techniques to be described it will be assumed that the difference signal outputs are available at the display tube. In addition, it will be assumed that equiband working is employed, though if this is not the case the various modifications to the measurements should be fairly obvious: for example, in $I'Q'$ working the bandwidths of the I' and Q' detector outputs must be individually checked, as must the following I' bandwidth channel and Q' bandwidth channel. Again, I' channel delay must be such that I' and Q' transients occur symmetrically.

The amplitude versus frequency characteristic may ·be conveniently determined by a similar method to that used for the luminance channel. A vision carrier signal and a wobbulator output are added in a resistive pad and the result is fed to the aerial input socket. An oscilloscope display of each of the colour difference outputs in turn will then give the individual overall chrominance responses.

Note that a zero beat will be observed between the sweeping video frequency and the reference generator frequency. This frequency

mixing occurs at the synchronous detectors, and the zero beat corresponds to the D.C. difference signal position on the trace. The bandwidth is easily determined by feeding a video oscillator output into the oscilloscope as well as a difference signal output, when a second zero beat will be observed corresponding to the video oscillator frequency.

The maximum output of the chrominance channel is best determined by feeding a colour signal of colour bars into the aerial socket, and observing each colour difference output on an oscilloscope. By increasing the saturation control, the point at which the waveform
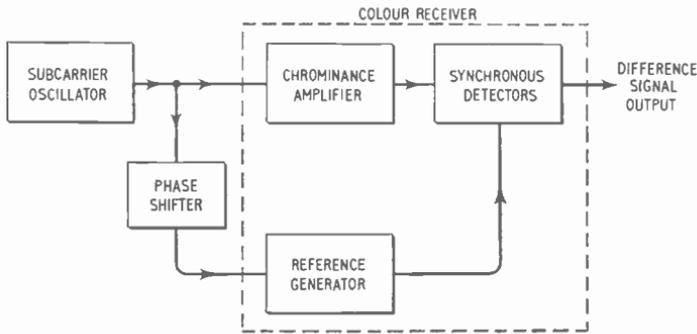
Fig. 12.5. Test set up for the determination of synchronous detector gains and detection axes

limits indicates maximum drive without overload. Note that the colour difference signals have positive and negative excursions depending on the particular colour, and the overload may occur in either direction. For example, for full pure red, the relative difference signal amplitudes are,

$$(R' - Y') = 0 \cdot 7, (G' - Y') = -0 \cdot 3$$

and

$$(B' - Y') = -0 \cdot 3$$

while for full pure cyan,

$$(R' - Y') = -0 \cdot 7, (G' - Y') = 0 \cdot 3 \text{ and } (B' - Y') = 0 \cdot 3$$

A suitable arrangement for measuring synchronous detector performance is shown in Fig. 12.5.

A crystal controlled sub-carrier oscillator is connected to a suitable point in the chrominance channel, and also to the receiver's reference

generator burst input via a calibrated phase shifter. It may be necessary to feed a monochrome signal into the receiver to generate a burst gate, otherwise the reference generator may not lock to the sub-carrier oscillator input.

By connecting a valve voltmeter to each difference signal output in turn and by adjusting the phase shifter, the phase reading corresponding to the maximum and minimum output from a difference signal channel may be determined, and hence the corresponding detection angles may be derived. For example, if maximum $(B' - Y')$ output occurs at a phase shifter reading of 100° and a maximum $(R' - Y')$ output occurs at a reading of 10°, then the red difference axis leads the blue difference axis by 90°. Measurement of minimum as well as maximum values checks the symmetry of detection for negative and positive difference signal components.

The relative amplitudes of the maximum (for example) values of the difference signal outputs will give the relative gains of the detection axes. In this connection, it must be remembered that certain factors are introduced by the transmitter, and also by the receiver to allow for the unequal phosphor efficiencies.

For example, the transmitter factors are $\dfrac{1}{1 \cdot 14}$, $\dfrac{1}{2 \cdot 03}$ and $\dfrac{1}{0 \cdot 7}$ for the

red, blue and green difference signals respectively, so that a receiver which uses equal phosphor efficiency drives will give maximum difference signal outputs (using the above method of measurement) in the ratio of 1·14, 2·03 and 0·7, or 1 to 1·78 to 0·615 for red to blue to green respectively. If the receiver has a phosphor efficiency drive ratio of 1 to 0·6 to 0·8 for red to blue to green, respectively, then the above method of measurement will give maximum difference signal values in the ratio 1 to 1·07 to 0·49.

A quick check on the synchronous detector performance may be made by feeding a colour bar signal into the receiver and observing the display in each of the primary colours in turn. If the colour bar signal has the usual form of vertical colour stripes of white, yellow, cyan, green, magenta, red and blue, then the red content should have the display shown in Fig. 12.6(a), the blue as in Fig. 12.6(b) and the green as in Fig. 12.6(c). These displays may be observed either by switching off two guns at a time (e.g. to observe the red display, switch off blue and green, etc.), or by observing the complete display through suitable red, blue and green filters. The " on " periods of any one waveform should be equal, and all the " off " periods should be equally " black ". This test is quite severe and the correct displays will be obtained only if the saturation
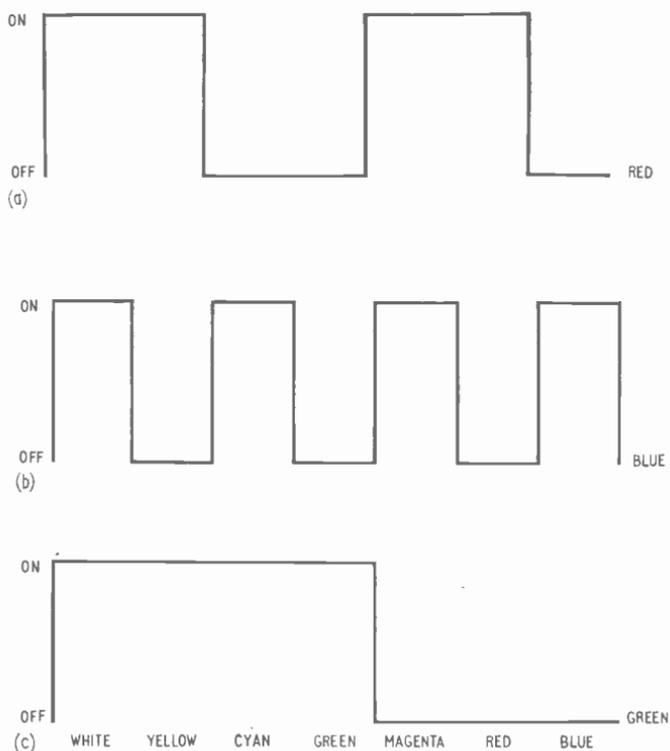
Fig. 12.6. Individual gun displays for the correct reception of colour bar signal

and hue control settings, and the relative detection axes and gains, are all correct.

The automatic chrominance control performance is best determined by the use of a local source of colour signal, such as a colour bar signal. If the receiver is adjusted to receive such a signal normally, and one of the difference signal outputs is displayed on an oscilloscope, the amount of chrominance plus burst attenuation required at the encoder to reduce the difference signal output by 6dB may be determined. This attenuation figure is then the A.C.C. figure of merit.

The amount of chrominance gain in hand may be determined by the above method by turning the saturation control to maximum

and measuring the amount of attenuation which can be introduced at the encoder before the difference signal output drops below its normal level.

12.8. Reference generator performance

Measurements which are indicative of the performance of colour receiver reference generators include frequency drift and pull-in performance, maximum static phase error and range of hue control, dynamic phase error due to noise and phase shifts due to inaccuracy of burst gating. A measurement which gives an indication of the amount of stray feedback from reference generator output to burst input is also advisable.

Apart from its primary function of reference frequency generation for the synchronous detectors, the reference generator also provides auxiliary services such as colour killing and A.C.C. The measurement of A.C.C. performance has been described in the previous section, but killer performance will be discussed below.

Frequency measurements on reference generators are best carried out with a precision wavemeter, and this must be as loosely coupled as possible to the reference output to prevent " pulling " of the wavemeter, and to ensure that no signal from the wavemeter interferes with the generator.

Frequency drift versus time after switch on and frequency shift due to mains voltage changes are easily measured. Pull-in performance is best measured by noting the generator frequency when no signal is applied to the receiver, and measuring the time for pull-in immediately after applying a colour signal. The time to pull-in from various detunings can be determined by deliberately altering the tuning of the generator, and before each measurement some time ($\simeq \frac{1}{2}$ min) should be allowed to permit the generator to become stable at the frequency to which it is tuned, since long time constant circuits may be present in the generator circuitry.

It should be noted that some idea of pull-in performance can be obtained by feeding a stable but variable frequency into the burst input, and a monochrome signal into the receiver aerial socket to operate the burst gate, if necessary. However, this type of measurement is of doubtful accuracy for two reasons. Firstly, if the line scan is not correctly locked to the burst, spurious frequencies closer than a spacing equal to half the line scan frequency may be produced next to the sub-carrier component, and this will impair pull-in performance (*see* Appendix 1). Secondly, since some receivers may have an excess of burst gate width, the effective burst duty ratio is greater than it should be and this will give an optimistic pull-in

performance. It is therefore advisable to measure pull-in performance by using a correct burst waveform which is locked to the line scan in accordance with the N.T.S.C. specification.

Maximum static phase error may be determined by means of the set up shown in Fig. 12.7. A stable but variable frequency generator is set to the sub-carrier frequency, and its output is fed into the reference generator at the burst input. It may be necessary to additionally feed a monochrome signal to the receiver in order to operate the burst gate. The reference output is fed to the Y-plates of an
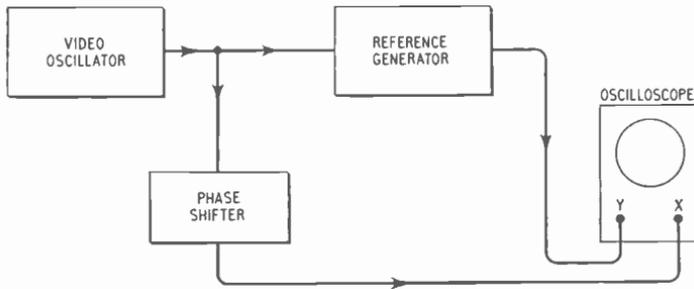
Fig. 12.7. Reference generator phase measurement set up

oscilloscope while a second output from the signal generator is fed to the X-plates via a calibrated phase shifter. Adjustment of the latter should enable a straight line Lissajous figure to be obtained when the signal generator is set to the sub-carrier frequency. Next, change the signal generator frequency until it differs from the sub-carrier frequency by an amount equal to the limit of the reference generator pull-in range (which has already been determined) and measure the phase change required to restore the Lissajous figure to a straight line. The measurement should be carried out for both positive and negative tuning errors, and the signal generator frequency is best measured by a wavemeter since small frequency changes are involved.

The above measurement will tend to give an optimistic result if there is an excess of burst gate width. An accurate measurement can be made if a correct burst waveform and a continuous sub-carrier sine wave are both available from the transmitter, in which case the burst waveform can be used for synchronizing the reference generator, while the continuous sine wave may be used for the oscilloscope X-plates. Detuning of the reference generator may be carried out by adjustment of its frequency trimmer. It is possible to use the burst waveform for the X-plate deflection instead of a

continuous sine wave, but because of the low duty ratio of the burst a rather " faint " Lissajous results.

The set up shown in Fig. 12.7 may be used for measuring the phase shift range of the hue control by noting the phase reading required to obtain a straight line Lissajous for one extreme end setting of the hue control, and subtracting the phase reading required to restore the straight line Lissajous at the other extreme end of the hue control.

An estimate of the dynamic phase error may be made by operating the receiver from a colour bar signal, and feeding the composite video to the Y-plates of an oscilloscope and the reference output to the X-plates via a calibrated phase shifter. Each colour bar will be displayed as an ellipse, and as the signal to noise ratio is reduced (by attenuating the signal input to the receiver, for example) so each ellipse will show a rotational ". blur " due to noise.   Measurement of the phase shift required to produce comparable peak to peak phase changes for a " clean " signal can be made by adjustment of the phase shifter.   Better accuracy can be obtained by this method if a " clean " composite video signal is always available for the oscilloscope display, but of course this is possible only if a colour bar generator is available.

Phase shifts due to burst gating inaccuracies are normally only significant in receivers which employ line time base gating.   Using the above arrangement of a colour bar Lissajous display, the phase

RED		
GREEN	CYAN	MAGENTA
BLUE		

Fig. 12.8.   Burst gate accuracy test picture

shifter can be adjusted until a chosen ellipse becomes a straight line. The phase shift caused by variation of the line time base hold control can then be measured.   The most critical condition for burst gating accuracy obtains when the colour bar on the extreme left of the picture is blue, since this has a phase almost opposite to that of the burst.

A very sensitive subjective test of burst gating accuracy can be made if a signal corresponding to the colour picture shown in Fig. 12.8 is available.   If the burst gate is too long, for example, the cyan and magenta bars will show hue changes which correspond to

the red, green and blue sections.   Note, however, that similar hue changes can occur due to insufficient D.C. gain in the chrominance channel, but this signal provides a severe overall test.

A final measurement of reference generator performance which is recommended concerns the effect of the hue control setting on the free running generator frequency.   With the receiver operating on a monochrome signal, and with the killer disabled, the peak-to-peak change of generator frequency should be measured as the hue control is slowly rotated through its range.   There are two possible causes for this frequency change.

Firstly, direct coupling between the hue control circuit and the oscillator tank circuit, and secondly, stray feedback from the generator output back to the burst input.   Disabling the killer may tend to emphasize the latter effect, if stray coupling is by way of the chrominance channel, for example.

12.9.   Colour killer performance

Two features of colour killer action are of importance: one is the reliability of the interpretation of a noisy monochrome signal as a monochrome signal and the other is sufficient sensitivity to recognize a burst for the smallest colour signal for which the chrominance channel will operate satisfactorily.

While operating the receiver from a monochrome signal, this may be attenuated and the receiver gain correspondingly increased until the killer " unkills ".   Knowing the input signal level and the receiver noise factor, the signal to noise ratio at which the killer fails may be estimated.

It should be noted that there are two general methods for deriving a killer bias.   One involves amplitude detection of the burst, in which case noise can be confused with burst signal for relatively high signal-to-noise ratios.

The other involves synchronous detection of the burst (this type is normally found only in two mode A.P.C. loops), and is reliable under extremely adverse signal-to-noise conditions.   In fact, it is not likely that the killer can be made to fail for the latter case unless a considerable noise voltage is added to the receiver input at maximum receiver gain.

The sensitivity of the killer circuit may be checked by operating the receiver from a colour signal and reducing the signal level until the receiver has no gain in hand.   Having adjusted the saturation control for a satisfactory colour picture, the killer bias should be shorted out.   If the saturation then increases, the killer detector is not producing sufficient output to remove the killer bias.

12.10. Performance of time bases

Measurement of colour receiver time base linearity does not involve techniques unfamiliar in monochrome receivers. A white grid on a black background signal is convenient for linearity measurement by some form of travelling microscope, and in order to simplify the procedure it is convenient to have only one gun (e.g. the green) in operation. It should be noted that time base linearity is particularly important in a colour receiver since the convergence circuits will not operate correctly unless a certain standard of linearity is achieved.

12.11. E.h.t. supply performance

, Adequate e.h.t. voltage, current, and regulation are essential for good picture quality, and it is recommended that a regulation curve of e.h.t. voltage versus current be plotted. An electrostatic meter may be used for voltage measurement, while a 1mA full scale meter will suffice for current measurement of the cathodes of the tube.

12.12. Convergence accuracy

For checking the convergence of the three images, a convenient signal is a white grid on a black background. The maximum misalignment of any two colours along a given grid line, both horizontal and vertical, should be measured. The measurements should be carried out for two areas of the picture, the first is an inner area enclosed by a circle having a radius equal to half the radius of the tube face, and the second is an outer area extending from the inner zone to within about an inch of the edge of the tube. The extreme edges of the picture may normally be ignored.

The measurements above should be carried out about two minutes after switching on, and repeated after about one hour to check the convergence stability.

12.13. Purity and colour balance performance

Lack of purity on the tube face is easily noticeable if a plain raster is displayed. Areas of impurity can be measured by a colorimeter after the central area of the tube has been set to the colour of Illuminant C, and a plot of the spurious colours can be made on a C.I.E. diagram. Drift of overall colour balance, due to temperature effects or mains voltage variations, may be determined by measuring the change of chromaticity of the central area of the tube face.

12.14. Summary

This chapter lists the apparatus considered necessary for servicing and for measuring the performance of colour receivers. As far as

servicing is concerned, very little additional apparatus is required to be added to existing monochrome receiver test gear, and details are given of a crystal oscillator and calibrated phase shifter.

Various techniques are described for the measurement of phase shifts at sub-carrier frequency, and for checking demodulator and reference generator performance. In addition, measurements which should be carried out on the remaining colour receiver circuits are discussed.

Receiver Installation

13.1. Introduction

The installation of a colour receiver is sometimes regarded with some trepidation, even by the most competent monochrome television engineer. But provided that he understands the basic principles involved, an engineer who is competent enough to install a monochrome receiver satisfactorily is perfectly capable of installing a colour receiver. The basic principles are outlined below.

The major portion of the apparatus in a colour receiver is concerned with the provision of a black and white picture. In fact, most of the circuits used in this part of the receiver are identical with monochrome practice, the essential difference being the display tube itself. The shadow mask tube may be regarded as the equivalent of three tubes, one red, one green and one blue, " wrapped up " as one tube. In order to achieve a satisfactory black and white picture, it should be realized that two conditions must be satisfied.

Firstly, each gun must excite only one colour. The red gun must provide a pure red raster over the whole scanned area, and similarly the green and blue guns must provide pure green and pure blue rasters. This requirement is referred to as *purity*.

Secondly, the relative brightness of the three rasters must be such that a white raster is subjectively observed. Furthermore this white must not change colour from the lowlights to the highlights. This adjustment is called *grey scale tracking*.

With similar video signals applied to all three guns, a black and white picture will result if the above conditions are satisfied, but the presence of a picture entails a further requirement, namely that the three pictures be superimposed one on the other, or registered. This registration of the images is called *convergence*, and it has two general adjustments. *Static convergence* which may be looked upon as picture shift of each complete raster relative to the other two and *dynamic convergence* which varies across each raster and increases with distance from the centre of the tube.

The remainder of the adjustments for a black and white picture are exactly the same as for a monochrome receiver, i.e. adjustment

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of time bases, brightness and contrast, spot limiting, focus, and oscillator tuning.

It cannot be emphasized too strongly that one of the most important features of colour receiver installation is the attainment of a good black and white picture. A good colour picture cannot be obtained from a receiver which exhibits a poor monochrome picture.

In case the reader is alarmed at the requirements for purity, grey scale tracking, and convergence, it should be noted that it is only the purity which will usually need attention, since the grey scale and convergence adjustments will have been set by the manufacturer. The purity can be affected by stray magnetic fields, however, so that the optimum adjustment at the final location of the receiver may be different from that at the factory, where a field free room may be used. However, provided the receiver has not been subjected to severe mechanical shock, the purity adjustment will consist merely of degaussing and edge purity correction, as described below. In view of the fact that the purity can be influenced by stray magnetic fields, it is recommended that the receiver should be installed as remotely as possible from large masses of iron or steel, such as radiators or girders.

The remainder of the circuitry in a colour receiver is the chrominance circuit which decodes the colouring information contained in the transmitted signal, and applies the appropriate signals to the three guns so that the otherwise black and white picture is displayed in colour. There are essentially three components in a complete chrominance circuit.

The first is a chrominance amplifier which accepts the sub-carrier signal (2·7 Mc/s in the 405-line system) and its sidebands. This amplifier usually consists of one or two stages with simple tuned circuit loads.

The second is a synchronous detector which demodulates the amplified sub-carrier signal, and ultimately provides three difference signal outputs which are fed to the appropriate grids of the tube, while the ordinary monochrome signal is fed to the cathodes of the tube.

The third is a reference frequency generator which is synchronized by the transmitted colour burst signal, and whose output is used by the synchronous detector as a phase reference.

There are usually two controls associated with the chrominance channel. The saturation control which alters the output of the chrominance amplifier and therefore the " strength " of the colour (with this control turned right down a monochrome or nearly monochrome picture will result, while a " too coloured " picture

will be obtained if this control is turned right up), and the hue control which determines the hue of the displayed colours. Turned one way, flesh tones will be reproduced too green and turned the other way, they will be too magenta.

Before considering installation procedure in detail, it must be remembered that different manufacturers will have different designs and therefore different installation instructions. Naturally, the manufacturer's instructions must be strictly observed, but the following are likely to be typical. Furthermore, it is assumed that no faults are present as fault conditions are considered in Chapter 14.

13.2. Aerial requirements

A special aerial is not required for a colour receiver. An aerial installation which provides a good quality signal as displayed on a monochrome receiver should be perfectly satisfactory for operating a colour receiver.

There are, however, one or two points to bear in mind. Remember that while a video frequency in the neighbourhood of 2·5 Mc/s in the 405-line system corresponds to fine detail in a monochrome picture, and may not be regarded by some as being essential for an acceptable monochrome picture, the sub-carrier in a colour signal is about 2·7 Mc/s and this carries low frequency colour information which is obviously essential to a colour picture. To take an extreme example, if an aerial installation were to provide a flat video frequency response up to 2·5 Mc/s with a sudden cut-off above this, then 2·5 Mc/s bars of Test Card C would be visible but the signal would be quite unsuitable for colour reception.

Again, if a close echo or " ghost " is present, this would have a relatively small effect on a monochrome receiver but it could have a disastrous effect on a colour receiver if the phase of the spurious signal were such as to reduce the sub-carrier signal appreciably. However, such a situation is rather unlikely, and in the authors' experience, if a ghost is present which can be tolerated on a monochrome receiver, then for 90% of cases it can also be tolerated on a colour receiver. While on the subject of ghosts, it may be advisable to point out that although the sub-carrier phase can be altered by a ghost signal, the hue control can always be adjusted so that the large area hues in a colour picture are correct. Hue errors are therefore usually confined to vertical edges of objects, the width of the incorrect hue being equal to the echo displacement.

As a general rule, therefore, an aerial installation which can provide up to 3 Mc/s definition on a monochrome receiver should be perfectly satisfactory for colour reception. Normally, this
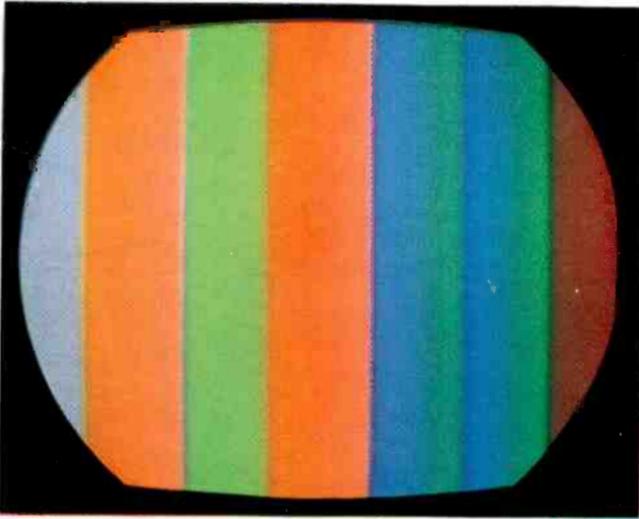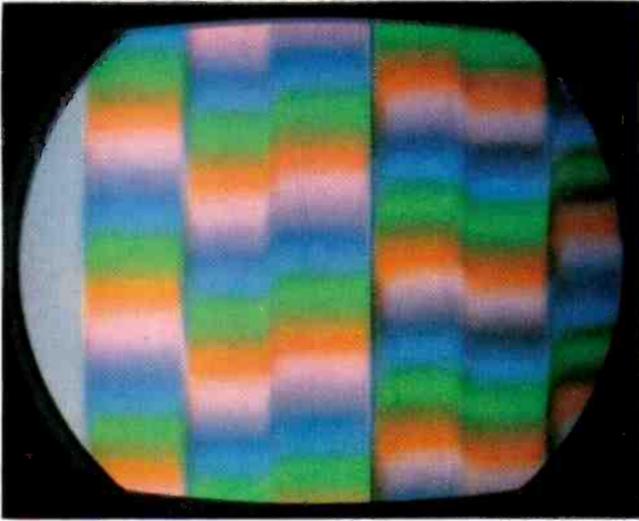
Colour beats on a picture produced by (above) *colour bar signal with reference generator out of lock;* (below) *colour bar signal with reference generator in a side-lock*
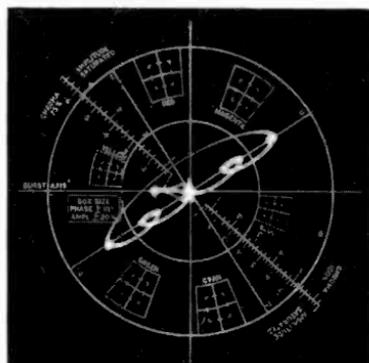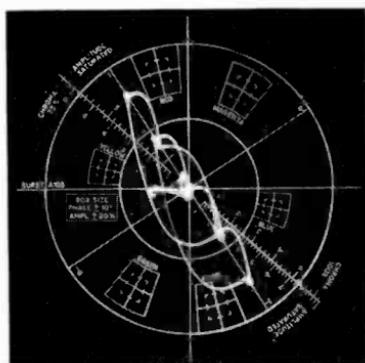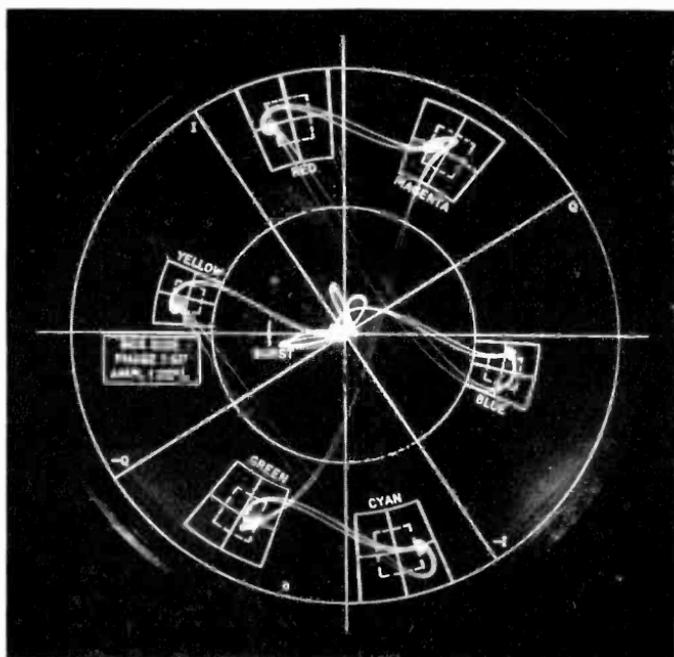
Typical displays from the Marconi Vectorscope are for checking the performance of the encoding equipment. The top photograph shows a colour bar display showing vectors representing the six primary and complementary colours lying within the appropriate boxes. The photograph on the left shows the same display with the Q component removed. The photograph on the right shows the same display with the 1 component removed. (Courtesy of Marconi Ltd.)

entails the correct choice of array for any particular location, careful positioning and orientation of the array, and preservation of the impedance matching between aerial and receiver. However, the conscientious engineer will recognize the foregoing as routine procedure for a monochrome receiver aerial installation.

Any ghost images which may be present, if acceptable on a monochrome receiver, should also be acceptable on a colour receiver.

An interesting echo effect is sometimes observed in the case of the 405-line system when a monochrome to colour (or vice versa) change is made at the transmitter. Since the vision carrier has a slightly different frequency in the two cases, it is possible for a positive echo in one case to change to a negative echo in the other for a fixed aerial position. This effect can be observed on a monochrome receiver, of course.

13.3. Monochrome adjustments

The first task in colour receiver installation is to obtain a satisfactory black and white picture.

A large number of the receiver circuits required for monochrome operation are identical in form with those of standard monochrome receivers, the essential difference being the display tube and its associated apparatus.

Apart from the usual time base, contrast, tuning, focus and brightness adjustments, the additional adjustments which may be required (but in any case must be checked) are concerned with raster purity, convergence and grey scale tracking.

13.4. Purity

With no signal applied to the receiver, adjust the brightness control so that the raster is visible and use the focus control to obtain clearly defined scanning lines.

The raster should be approximately white in colour, but the important point to notice is whether the colour is uniform all over the raster. If it is not, the tube must be demagnetized by means of a degaussing coil (see Section 11.6). Move the degaussing coil slowly around the sides and front of the receiver, and while maintaining a circular movement of the coil, gradually move back from the tube face to a distance of five or six feet before switching off. The purity should now be satisfactory, and a sensitive check can be made by switching off the appropriate guns to obtain an all-red raster. There may be some residual impurity near the edge of the raster, in which case the nearest edge purity magnet should be adjusted (see Figs. 13.1 and 13.2).

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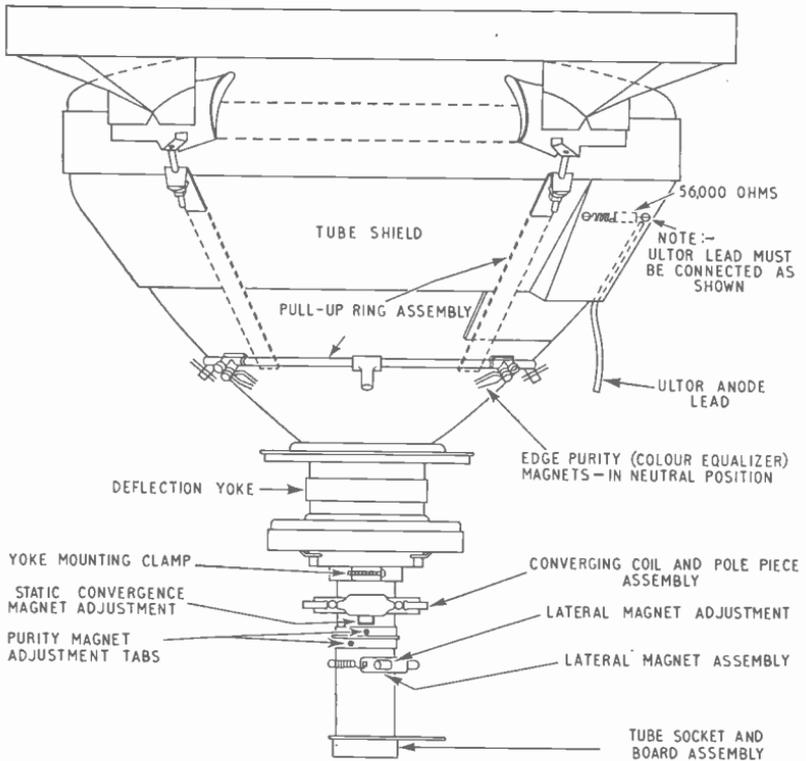
Fig. 13.1.　Tube adjustments and components

If the degaussing process and the edge purity magnet adjustments do not result in good purity, the purity magnets on the neck of the tube and the deflection yoke assembly position must be investigated.

Set the purity magnets on the tube neck to neutral by moving the tabs together, and also set the edge purity magnets to neutral so that the two magnets in each position are in line one above the other. Move the deflection yoke as far back as possible and switch off the green and blue guns. If these guns cannot be extinguished by turning down the screen and background controls, connect 100kΩ resistors between the grids of the green and blue guns and ground.

Now rotate the purity magnet around the tube neck and at the same time adjust the tabs relative to one another until there is a uniform red area at the centre of the tube. Move the deflection yoke forward until a position is found which gives the best overall

red raster consistent with absence of neck shadow. Check the purity of the green and blue guns independently; always adjust so that purity errors show only at the edges of the screen. Correct any edge impurity by means of the edge purity magnets.

13.5. Time bases

The time base adjustments are similar to the conventional monochrome procedure, i.e., height, width and linearity controls should be adjusted as required, while observing a grid or Test Card circle. However, it should be noted that time base adjustment can affect convergence and hence the time bases should be set up before convergence is attempted. Time base adjustment can be conveniently carried out by observing one display only, e.g., the green.

13.6. Convergence

Before attempting to set up the convergence, ensure that both time bases are correctly adjusted. A generator which provides a
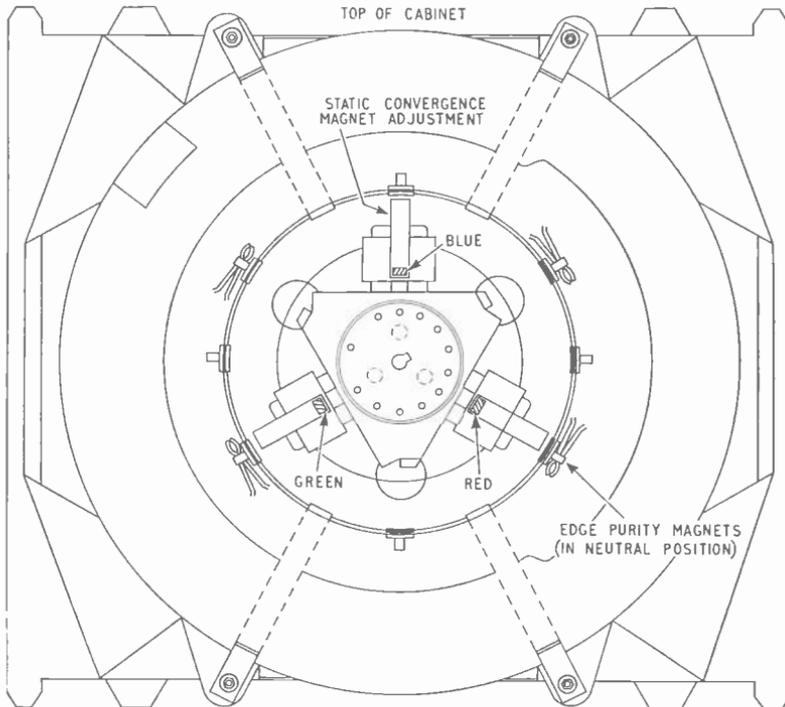
Fig. 13.2. Edge purity and centre convergence magnets

cross-hatch of fairly fine white lines is preferable for dynamic convergence adjustment.

A preliminary static convergence adjustment should first be made by applying the cross-hatch signal to the display and adjusting the bias on the guns so that all three colours are operative. After adjusting the focus, and with all the dynamic convergence controls set to mid-range, adjust the red, green and blue centre convergence magnets (Fig. 13.2) and the lateral magnet (Fig. 13.3) so that the three images are registered at the centre of the screen. The direction of movement of the images is shown in Fig. 13.4(a) for the magnets,
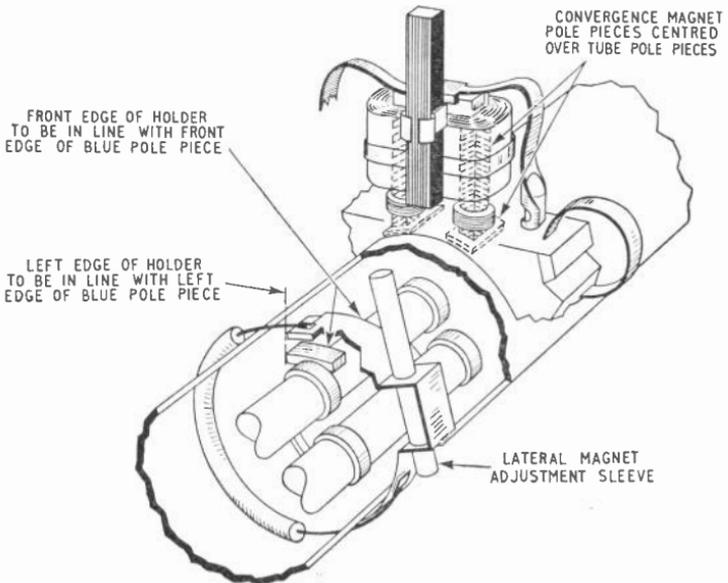
Fig. 13.3.   Location of convergence and lateral beam magnets
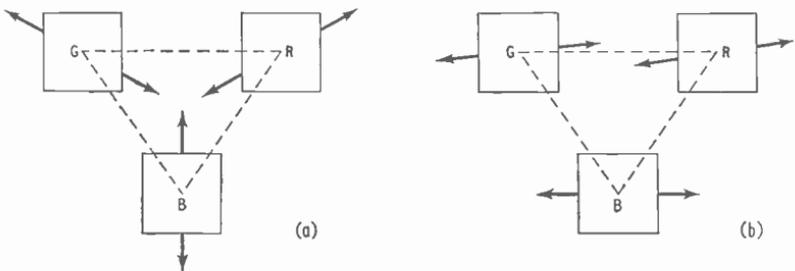
Fig. 13.4.   Image movement pattern

while the lateral magnet produces a movement (Fig. 13.4(b)) in which the red and green movement is opposite to the blue.

Next, check the purity, and re-adjust if necessary. For vertical convergence adjustment, apply the cross-hatch pattern, and observe the vertical centre bar of the pattern. Adjust the red and green tilt controls to produce equal displacement of the red and green lines at the top and bottom of the line, as shown in Fig. 13.5(a). Next, adjust the red and green amplitudes to produce straight, parallel, red and green vertical lines. Now gradually reduce the red and green amplitudes to converge the red, green and blue lines adjusting the red and green tilt controls as necessary, to keep the lines parallel. It should be possible either to make the three images overlap, or show a slight displacement of red at one side and green at the other with all lines parallel from top to bottom. In the latter case, re-adjust the static convergence magnets so that the three images are superimposed to give a vertical white line, Fig. 13.5(b).

For setting up the blue vertical convergence, examine the centres of the extreme top and bottom horizontal bars of the cross-hatch. Adjust the blue vertical tilt and amplitude controls to produce an equal downward displacement of the blue line from the top and bottom bars, as shown in Fig. 13.5(c). Reduce the blue vertical amplitude to converge all horizontal bars at the centre, from top to bottom, adjusting the blue vertical tilt as necessary, to give the result shown in Fig. 13.5(d).

The horizontal convergence adjustment will depend on the type of circuits employed, but as an illustration a typical procedure for a recent design will be given. In this, six controls are provided, three right hand, $B1$, $RG1$ and $RG2$ and three left hand, $B2$, $RG3$ and $RG4$.

With the cross-hatch pattern applied, adjust $B1$ to make the blue line at the right hand centre a straight horizontal line, as shown in Fig. 13.6(a). Now adjust $B2$ for a straight horizontal blue line at the left hand centre, as shown in Fig. 13.6(b).

Adjust $RG1$ so that the vertical lines at the right hand side converge, Fig. 13.6(c).

Adjust $RG2$ to make the horizontal red and green lines converge at the right hand side as indicated in Fig. 13.6(a). Re-adjust $B1$ to make the blue line at the right centre fall on the converged red and green lines, and re-adjust $RG1$ for convergence of the vertical lines on the right hand side.

Adjust $RG3$ to make the vertical lines on the left hand side converge, as in Fig. 13.6(d).
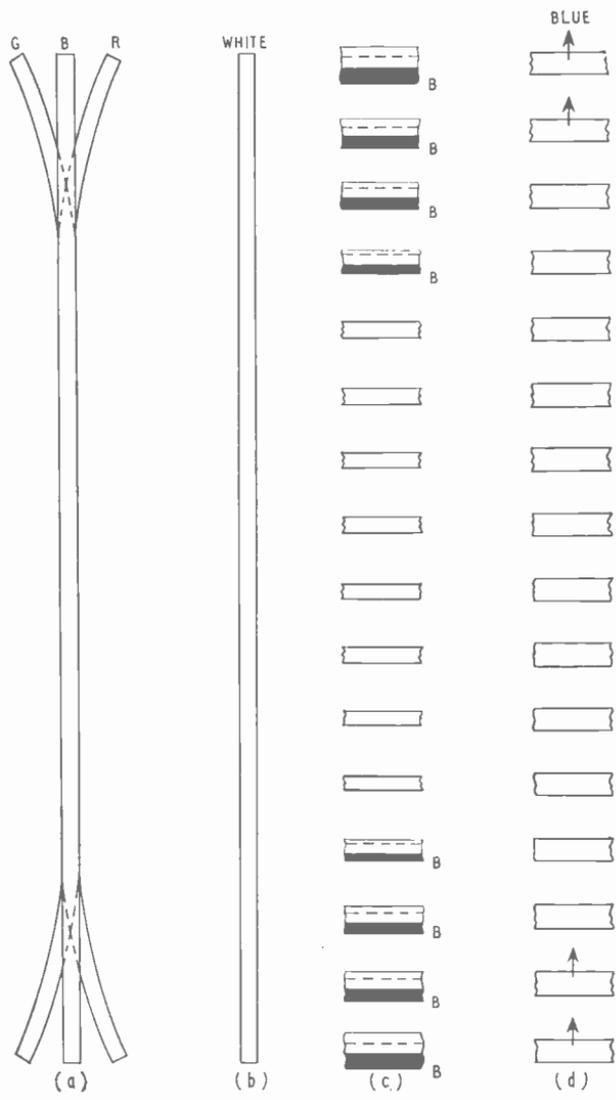
Fig. 13.5.   Vertical convergence patterns

Adjust *RG*4 to make the horizontal red and green lines converge on the left hand side, Fig. 13.6(b).   Repeat adjustment of *RG*3 to compensate for any interaction.

Re-adjust *B*2 to make the blue line on the left hand centre fall on the converged red and green lines.

The cross-hatch pattern should now be converged on all parts of the screen, as in Fig. 13.6(b).

13.7.   Grey scale adjustment

The purpose of the grey scale adjustment is to provide a constant white chromaticity at all drive levels, and a convenient signal is one containing about five steps from black to peak white.   Six controls have to be adjusted, three of which are control grid or
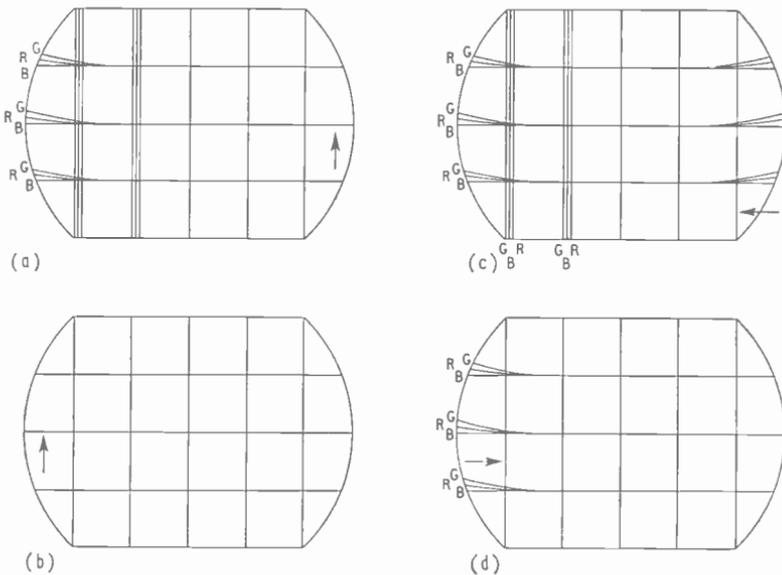
Fig. 13.6.   Horizontal convergence patterns

background controls used for setting the highlight colour and three screen controls for setting the lowlight colour.   In order to achieve as bright a picture as possible, at least one of the screen controls should be at a maximum after the setting up is completed.

With the three screen controls at maximum and the background controls at minimum, set the contrast and brightness controls to give a satisfactorily contrasted picture.   The first step is to decide

which of the three screen controls should remain at maximum, and this can be done by adjusting the three background controls to give white in the highlights, and noting the lowlight colour.

If the lowlight colour is cyan, the red screen should remain at maximum. Similarly, for a magenta lowlight, the green screen should remain at maximum, and for a yellow lowlight the blue screen should remain at maximum. Thus, the complementary lowlight colour is the screen which should remain at maximum.

If the lowlight colour is red, green or blue, then gradually turn down the screen of the predominant colour. If the lowlights become grey, leave the remaining two screens at maximum and adjust the background of the reduced colour to give white highlights. Check the lowlight colour again, if necessary adjusting the screen which was turned down. In this case the tracking is completed.

If the lowlights become cyan, magenta or yellow rather than grey, then this indicates that the red, green, or blue screens should remain at maximum.

Having determined which screen should remain at maximum, do not adjust either the screen or background controls for this colour. Adjust the remaining two screen controls for grey lowlights, and the remaining two background controls for white highlights, re-adjusting the two screens for lowlights if necessary.

13.8. Colour adjustments

The chrominance channel adjustments are chiefly concerned with the saturation and hue controls, but first it is advisable to check the operation of the reference frequency generator and its associated circuits.

13.9. Reference frequency generator

The reference frequency generator has at least two and sometimes three functions in a colour receiver. It provides a constant frequency reference which is locked to the colour burst. It recognizes the presence of the colour burst and switches the output chrominance stages on or off depending on whether the burst is present or absent, respectively (colour killing, as it is called) and in some cases it provides automatic chrominance control, or A.C.C.

The free running frequency of the reference generator can be checked by observing a colour picture, removing the burst feed to the generator (by grounding it, for example) and also the killer bias which ordinarily disables the chrominance output. The colour content of the picture will then appear as horizontal colour beats

if the saturation control is turned up sufficiently. As the frequency control of the generator is adjusted, so the horizontal beats will reduce in number until finally, when the correct setting has been reached, the colour beat becomes almost zero and drifts through zero beat. Application of the colour burst should now bring the generator into lock.

In some receivers, the burst gate which selects the colour burst from the chrominance signal may be operated from the line time base. In this case, adjustment of the line hold control will affect the operation of the reference generator, and the manufacturer's instructions should be closely followed on this point.

The colour killer may be affected by noise entering the receiver with the signal, so that a noisy monochrome signal may switch on the chrominance output stages. In order to prevent this, a killer threshold adjustment is sometimes provided, and this is usually set so that with a normal level of monochrome signal applied, the cross-colour just disappears.

13.10. Saturation control

The saturation control is usually a customer operated control, since the " colouring " of the picture is to some extent a matter of taste. Also, newcomers to colour television tend to require rather more than the correct saturation but after some hours of viewing a lower and more nearly correct saturation is preferred by most people.

However, while the acceptable range of the saturation control is quite large, there is strictly only one position for which the colour receiver will reproduce the transmitted colours most accurately and it is really the duty of the installation engineer to set the control at this position.

Before attempting to adjust the saturation, first turn the control right down and, if necessary, short out the burst input to the reference generator so that a monochrome picture is displayed. Check that the purity, grey scale, contrast and brightness are acceptable.

Now slowly turn up the saturation while observing flesh tone and, if possible, any obviously saturated colours. Eventually a point will be reached where the saturation is too high. Now reduce the saturation until the coloured content seems reasonable. For the present, ignore any incorrect hues in the picture, for example too green flesh tones.

13.11. Hue control

Having set the saturation control as above, adjust the hue control so that flesh tones are neither too green nor too magenta. It

is recommended that only flesh tones should be observed while adjusting the hue control, since this is a colour known by everyone, and occurs most frequently. Further, it seems to be the most sensitive colour in a subjective sense.

In view of the results of subjective tests carried out by the authors on the relationship between hue control setting and viewer preference, it would seem that while correct flesh tone lies between slightly too green and slightly too magenta, it is preferable to err on the side of magenta rather than green. After setting the hue control, re-adjust the saturation control, if necessary.

13.12. Effect of other adjustments on saturation and hue

It should be pointed out that the degree of saturation of a colour television picture can be appreciably altered by adjustment of the master brightness control. Turning the brightness down will tend to increase saturation; turning brightness up will decrease saturation. Note that this effect is not caused by an interaction of circuits. It is simply that a particular brightness level requires a particular chrominance level.

Again, if the white balance is set too green, for example, and the hue control is set to give acceptable flesh tone, then what should be a pure red will be reproduced with a magenta cast.

The above points help to illustrate the importance of achieving correct black and white reproduction before adjustment of the chrominance channel is attempted.

13.13. Summary

In this chapter we have seen that the first task in receiver installation is to obtain a satisfactory black and white picture. This necessitates a good aerial, a pure raster (i.e. a uniform colour over the whole raster), a satisfactory grey scale and accurate registration or convergence so that the three pictures are superimposed. Apart from the adjustments one would make to a monochrome receiver, it is likely that the only additional requirement will be degaussing the tube, since the grey scale and convergence adjustments will probably not need attention, though they should be checked.

Having achieved a satisfactory black and white picture, it is advisable next to check the free running frequency of the reference generator. Finally, the saturation control should be set to give a picture which is judged to be sufficiently but not excessively coloured, while the hue control should be set to give correct flesh tone. The engineer may rest assured that the adjustment of the saturation and hue controls is very simple, and calls for little practice.

Colour Receiver Fault Finding

14.1. Introduction

Colour television receivers contain more components and circuits than monochrome receivers but are no more difficult to service. In general the colour display tube provides more information on the location of any fault condition and this extra information offsets the increase in the number of stages. Experience, a systematic approach and adequate test instruments are all valuable aids to the colour service engineer, as in all servicing. For some years to come there will be considerable design differences between the various makes of receiver and the service man must have a clear picture in his mind of the block schematic diagram of the particular set he is repairing. Whenever possible he should work with the manufacturer's service handbook beside him. There are basic differences between receivers using *RGB* amplifiers and those using luminance and colour difference channels and these differences affect the appearance of symptoms. The picture on the cathode-ray tube will quickly localize the seat of the trouble however, and the usual monochrome techniques of measuring voltages, currents and waveforms will soon identify the specific fault if the general principles of operation of colour reception are borne in mind.

The previous chapter gives an outline of the basic arrangement of colour receivers and describes various adjustments which can be made to the specifically colour parts of the circuits. The reader is recommended to study Chapter 13 and to familiarize himself with the colour test techniques given in Chapter 12.

Throughout this chapter the bandwidths, etc., quoted refer to the 405-line system.

14.2. Safety precautions

Most of these will be second nature to the service man from his experience with monochrome receivers. The fumes from faulty selenium rectifiers should not be inhaled and the rectifiers should not be brought into contact with the bare skin. The power supplies will usually have a lower impedance than is normal with monochrome receivers and shocks can be rather more serious. This is

particularly true of the e.h.t. supply which will be over 20kV with a stabilized output of a milliamp or so. Colour tubes are heavier and more awkward than black and white ones and may have more energy stored in their screen capacity, so the e.h.t. terminals should be discharged before handling. The risk of implosion is the same as with black and white tubes but the replacement cost is higher.

Out of its cabinet, the chassis may generate some X-rays when it is operating. The most likely sources are the front face of the tube when the safety glass has been removed, the e.h.t. rectifier and the e.h.t. stabilizer valve. Portable meters which indicate the strength of such fields are available on the market, or small photographic film strips in suitable light-opaque wrapping may be left near the suspected source or worn by the engineer. Subsequent development of the film will indicate, from a knowledge of the exposure time, the seriousness or otherwise of the amount of radiation. In general, no hazard to health is likely to arise but the dangers should be appreciated and a balanced view adopted.

14.3. Initial examination with no signal input

Before making any detailed examination of the receiver a cursory and rapid inspection of the general receiver functions should be made, as in normal service procedure. Indeed, so much of the servicing of colour television receivers is similar to monochrome work that most of the procedure will be given in brief note form, suitable for quick reference at the work bench. The work bench itself should be sited well clear of anything likely to disturb the magnetic field, such as large iron or steel objects, transformers, etc.

(a) Inspect interior for signs of fire or damage.
(b) Check mains taps.
(c) Switch on and scrutinize for signs of smoke, voltage breakdown or corona.
(d) Raster should be uniformly white.

    (i) If no raster or insufficient raster—treat as monochrome and check e.h.t. and other tube electrode voltages and time bases.

    (ii) If raster uniformly coloured, check that all three guns are operating and adjust screen and grid voltages for white screen. Check gun cathode potentials.
Yellow raster indicates insufficient blue electron beam.
Magenta raster indicates insufficient green electron beam.
Cyan raster indicates insufficient red electron beam.

(iii) If raster has coloured patches, follow installation procedure for purity adjustments, field edge equalizer magnet adjustments and if necessary degauss.

(e) All three beams should go through optimum focus at same setting of focus control. If one beam is significantly out of focus at optimum setting, only cure is to replace tube.

(f) Turn up contrast and check that luminance and sound channels have sufficient gain to produce noise—treat as monochrome.

14.4. Examination with monochrome signal input

Monochrome test patterns are an essential part of colour television testing. Linearity and convergence patterns, resolution wedges, Test Card C or D, etc., are all useful (*see* Chapter 12),
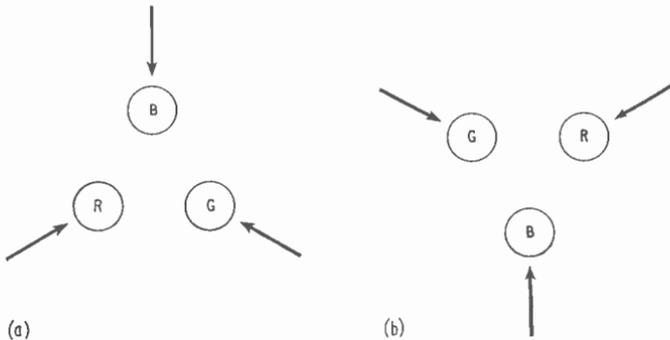
Fig. 14.1. Incorrect static convergence. The limit spot positions are shown with the convergence field (a) too large, (b) too small

as are normal monochrome picture transmissions, but check that aerial has adequate response over $2\frac{1}{2}$ Mc/s to 3 M/cs region.

(a) Check luminance gain—as for monochrome.

(b) Re-check purity and focus.

(c) Check raster size and linearity, as monochrome. If convergence is incorrect it may be convenient to adjust linearity with only one gun on (green is the brightest). Unless linearity is good it may not be possible to converge rasters accurately.

(d) Check static convergence (*see* receiver installation procedure). If faulty, check magnets, including blue lateral shift magnet and/or D.C. current through convergence yoke windings.

If blue spot cannot be moved below green and red spots, convergence field or current is too high. If blue spot cannot be moved above other two spots, convergence field or current is too low (Fig. 14.1). Try reversing magnets in their holders. As last resort, change tube.

(e) Check dynamic convergence (*see* receiver installation).
   (i) If one beam cannot be controlled, check convergence yoke winding.
   (ii) If centre but not the sides can be converged, check horizontal convergence amplitude control and current through it, with oscilloscope. Check input waveform from line time base.
   (iii) If raster comes into convergence on one side of screen before the other, check horizontal convergence tilt or phasing control.
   (iv) If raster will converge at sides but not at top and bottom, check vertical convergence controls, waveforms, and yoke windings. Check input from field time base.
   (v) Look out for misleading effects. Poor frequency response in one colour channel may cause fringe of colour on one or both sides of convergence spot. Similarly, poor focus of one beam or low emission in one gun, will produce colour halo around convergence spot. Inspect primary pictures one at a time. Sudden jumps in convergence positions may be due to faulty e.h.t. stabilization.

(f) Check grey scale tracking (*see* installation procedure, Section 13.7).
   (i) Adjust tube grid (background) controls so that highlights are neutral and a reasonable match to Illuminant C. If dark greys are then coloured (say red), reduce screen volts of appropriate gun (i.e. red) or increase screen volts of other two guns if possible, and re-adjust grid volts.
   (ii) If *RGB* amplifiers used, check relative *RGB* gains and maximum outputs.
   (iii) If neutral grey scale not possible, check range of grid and screen volts.
   (iv) Check tube cathode volts and check video output stages, including colour difference outputs if used. Remember stages may be D.C. coupled. Check gains of *RGB* amplifiers.

(v) Check matrix resistance values, unless these only control colour difference signals.

(vi) Check spot limiters.

(g) Check frequency response and R.F.-I.F. alignment — as monochrome.

(i) If one block of resolution bars, say 2 Mc/s, is missing from Test Card C, sub-carrier rejector is off tune.

(ii) If marked echoes at intervals of about $1\mu s$—check termination of luminance delay cable.

(iii) Check that the luminance response is fairly flat to past the sub-carrier frequency when the sub-carrier rejector circuit is disconnected (or shorted as appropriate).

(iv) Incorrect convergence will markedly affect picture resolution.

(h) Check colour killer operation.

(i) Adjust saturation to maximum. If coloured patches appear on picture where there is fine monochrome detail, then colour killer is not operating.

(ii) Check killer threshold adjustment, if any.

(iii) Check bias on chrominance stages.

(iv) Check D.C. output from auxiliary phase detector in reference generator or its equivalent.

(j) Check chrominance gain. This is more conveniently done on colour transmission. However, if killer voltage is shorted or removed, and saturation turned to maximum, cross-colour effects should show on monochrome transmission if chrominance channel is amplifying.

14.5. Examination with colour signal input

Do not attempt to check the colour performance until the monochrome performance is correct. If only a colour signal is available, operate the colour killer or turn down the saturation or bias-off the chrominance channel, so that the receiver reproduces a monochrome picture. Good colour reproduction is not possible if the monochrome grey scale tracking is faulty, or if the picture white is not Illuminant C.

Be sure that the aerial and any distribution system in use has adequate response, without peaks, in the vicinity of 2·7 Mc/s.

The colour bar waveform is a convenient signal for fault tracing as the service engineer will soon recognize the video and chrominance waveforms to be expected on this signal (Fig. 5.2).

   (a) If monochrome picture but no colour.

      (i) Check fine tuning.

     (ii) Short out or disable colour killer. If coloured beats, reference drive exists but oscillator is not locked. If not even coloured beats, there is no reference drive.

     (iii) Check saturation control voltage and bias on chrominance stages.

     (iv) Check gain of chrominance amplifier.

     (v) Check that chrominance is being fed to reference generator.

     (vi) Check burst gate is properly timed and burst is reaching phase detector.

     (vii) Check D.C. bias on reference oscillator as test that oscillator is oscillating.

     (viii) Check oscillator is approximately on tune.

     (ix) Check sufficient amplitude of reference is reaching synchronous detectors.

     (x) Check waveforms at inputs and outputs of synchronous detectors.

     (xi) Check waveforms through matrix circuits.

   (b) No colour lock. With killer disabled, horizontal bands of colour moving vertically across the picture show that the reference oscillator is slightly off tune. The nearer it is on tune, the broader and fewer these bands become.

     (i) Check burst input to loop detector.

     (ii) Check burst gating.

     (iii) Check bias at grid of reactance valve. If bias is large, earth grid of reactance valve. If colour bars then roll slowly near in-tune position, loop discriminator is faulty. If colour remains badly out of sync, then oscillator or reactance tube is faulty.

   (c) Long pull-in time. With killer disabled, if horizontal bands of colour do pull into sync but take a minute or so to do this, check these three points.

     (i) Tuning of oscillator.

     (ii) If quadricorrelator circuit, check auxiliary discriminator output.

     (iii) Operation of variable time constant in feed to reactance valve.

   (d) Incorrect colour. The colour bar signal is a great aid in diagnosing faulty colour response.
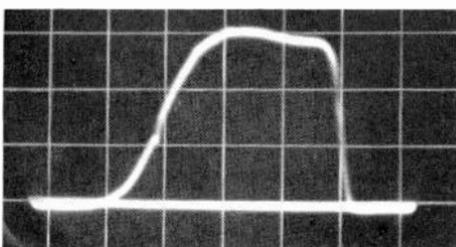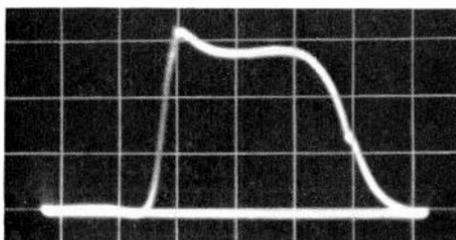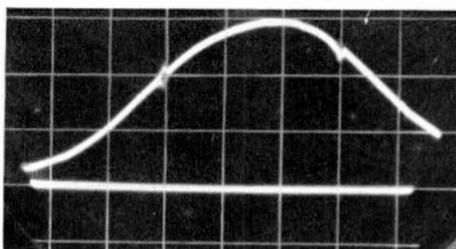
*Typical frequency characteristics of a colour
receiver measured at the luminance detector
with the sweep input applied to different points.
In the top photograph the signal is applied to the
third I.F. stage grid and shows the sub-carrier
marker on the left. In the centre and lower
photographs the input is applied to the mixer
grid and aerial input respectively. (Frequency
sweep is from right to left.)*

Receiver image misregistration; (above) effect caused by excessive delay in receiver luminance channel; (below) effect on black-and-white picture reproduced on a shadow mask tube with static convergence grossly out of adjustment

(i) Note absence of any colours. This will usually localize a fault in a particular synchronous detector or its subsequent amplifiers and matrix, if any.

Orange and cyan colours only—insufficient $Q'_{\bullet}$ signal (insufficient $(B' - Y')$ is similar).

Green and magenta colours only—insufficient I' signal (or $(R' - Y')$.

When the I' signal fails there is usually very little colour left in a normal outdoor scene.

If demodulator or colour output valve dies, the D.C. coupling to the display may give picture a colour cast similar to the colour axis of the decoder, e.g. failure of $(R' - Y')$ demodulator can give overall surplus of red in picture.

If reference oscillator pulls into side lock the wrong colours will be obtained and these will give the appearance of lack of drive.

(ii) Check setting of hue control. If all colours are present but occupy wrong positions, check operation of hue control, reference feed, phase shift circuits, and matrix (*see* Chapter 12 test methods).

Critical test is to view colour bar output from one gun only (black out other guns or view through sharp cut optical colour filter).

Adjustment of hue, contrast and saturation controls should make all visible bars the same brightness. If this is not possible, check demodulation angles of synchronous detectors, and matrix. Check all three guns.

(iii) If only pale and desaturated colours are present— check fine tuner, chrominance gain, A.C.C. voltage from reference generator, R.F.-I.F. alignment, side lock of reference oscillator.

(iv) Coloured hum bars. Check heater-cathode leakage in appropriate colour channel, reactance valve and oscillator; check also for hum on grid of reactance valve.

(e) Faulty colour transitions.

(i) Colours smear. Check fine tuner and chrominance bandwidth. Check sufficient reference drive to synchronous detectors.

(ii) Colour displaced from luminance outline of object. Check delay circuits. Colour too far to viewer's right is insufficient luminance (and I') delay.

(iii) Incorrect colour edges. Insufficient luminance high frequencies. Too wide Q' channel response.

Too wide chrominance bandwidths in equiband receivers. Incorrect chrominance channel alignment.

Feedback from luminance sub-carrier rejector circuit to chrominance channel.

Note that the colour bar generator is not, in general, a realistic test of colour transitions—especially for the green-magenta transition where all of the three basic rectangular waveforms are changing, the green signal is being switched off and the red and blue signals are being switched on. Note also, from Chapter 16, that the N.T.S.C. system has certain deficiencies in its transition performance, although these are relatively minor on the average programme scene.

(f) Coloured beats.

(i) Check reference oscillator is in-sync.

(ii) Check 3·5 Mc/s rejector in chrominance channel is on tune.

(iii) Check sound I.F. rejection.

(iv) Check input level not high enough to cause cross-modulation.

(v) Carrier interference at R.F. or I.F. near to sub-carrier frequency.

Note that such interference may come from harmonics of reference drive or chrominance.

For example :

The 13th harmonic of the sub-carrier is 34·5515625 Mc/s.

The vision carrier I.F. is normally 34·65 Mc/s.

The 14th harmonic of the sub-carrier is 37·209375 Mc/s.

The standard sub-carrier I.F. is 37·31 Mc/s.

(g) Monochrome beats on colour transmission.

(i) Check sound rejector circuits in luminance channel.

(ii) Consider possibility of c.w. interference.

(h) Picture on sound on colour transmission only.

Insufficient rejection in sound channel at sub-carrier frequency.

(j) Colour appears and disappears.

(i) Check killer threshold setting.

(ii) Check A.C.C. circuit.

(iii) Check level of burst signal to reference generator.

  (iv) Check timing of burst gating.
  (v) Check intermittent oscillation of reference oscillator, or frequency jumping.
(k) Colour varies with picture content.
  (i) Check grey scale tracking.
  (ii) Check burst gate timing.
  (iii) Check for differential phase effects in common luminance—chrominance stages.
  (iv) Check A.C.C. is operating and that chrominance signal is not limiting before saturation control.
  (v) Check D.C. level maintenance.
(l) Colour accuracy varies over picture area.
  (i) Check purity.
  (ii) Hum in phase discriminator, reactance valve or reference oscillator.
  (iii) Hum in chrominance channel.
  (iv) Check constancy of reference output from reference generator.

14.6. Miscellaneous

14.6.1. NOISE

In fringe area reception, random noise may appear in both the luminance and colour channels. It will generally be more noticeable in the white parts of the picture where it can produce coloured flecks. In areas of saturated primary colour, such as green, it will change the colour from one green to another green and the only obvious fluctuation will be in the luminance noise (*see* Section 4.14). However, aside from this effect, faulty reference oscillator sync can give rise to hue fluctuations, particularly in the subtractive primary colours yellow, cyan and magenta, due to noise. This may be due to insufficient integration in the reference generator, or the quadricorrelator may not have changed over to the in-lock position.

14.6.2. APPEARANCE OF BURST

A vertical bar of green on the left hand side of the picture will probably be the colour sync burst appearing on the horizontal flyback due to insufficient line blanking.

14.6.3. LINE TIME BASE RADIATION

The increased energy in the line time base makes it even more important to keep the line time base shielding and mains filtering intact.

14.6.4. REFERENCE OSCILLATOR RADIATION

This is a new source of television receiver interference with other services. Screening and decoupling of reference generator and demodulators must be above reproach.

It is also important to prevent the reference signal from spraying back into the burst signal path and eventually reaching the input to the reference generator; this can cause the reference oscillator to go well off frequency (*see* Section 10.13). When replacing components, wiring or screening, these points must be borne in mind.

14.7. Summary

In servicing the colour television picture much information can be obtained from the picture itself. Always work for a good monochrome picture before trying to analyse an obscure colour fault. Become familiar with the picture appearance when either one of the primary colours or one of the chrominance signals is missing. Once the approximate location of the fault has been established, the normal monochrome techniques of voltage, current and waveform measurements will pin-point the faulty component.

Monochrome Reception
of N.T.S.C. Signals

15.1. Introduction

The N.T.S.C. system is designed in such a way that the colour signal on an unmodified monochrome receiver will be displayed as a black and white picture of the transmitted colour scene. This compatibility of the N.T.S.C. signal is an extremely important feature because it allows colour to be added to an existing monochrome system, without modification of existing channel allocations, and also the need to duplicate the transmission for colour and monochrome receivers is removed.

In a perfectly compatible system, the operation of a monochrome receiver by a colour signal should be indistinguishable from its operation by a monochrome signal, but while the N.T.S.C. system does approach this ideal, the monochrome display of an N.T.S.C. signal is not identical with that of a conventional monochrome signal.

The N.T.S.C. signal has been described in Chapter 4, and it will be remembered that it consists of a monochrome signal to which a sub-carrier signal carrying the chrominance information has been added. Now consider the effects of an N.T.S.C. signal as displayed by a monochrome receiver.

15.2. Asynchronous working

Apart from the sub-carrier, the N.T.S.C. monochrome signal differs from its conventional counterpart in that the scanning frequencies are derived by division from the sub-carrier frequency, and consequently the line and field scanning frequencies are crystal controlled and therefore independent of the mains supply frequency. Thus, the N.T.S.C. signal entails asynchronous working of all receivers, so that sufficient smoothing must be included to prevent undulations on the picture and brightness hum bar modulation.

The reader may wonder why it is necessary to crystal control the sub-carrier frequency, and hence the scanning frequencies, at the transmitter. Why not have a nominal sub-carrier frequency of

2·66 Mc/s, divide down to the scanning frequencies, and compare the field frequency with the mains supply frequency and thereby control the sub-carrier frequency to give mains synchronous working?

If this scheme were adopted, colour receivers would not necessarily be able to generate a sub-carrier reference frequency which is essential in order to demodulate the sub-carrier signal. For example, if the supply frequency changed by 2 % (from 50 c/s to 49 c/s), the sub-carrier frequency would also change by 2 % and 2 % of 2·66 Mc/s is about 53 kc/s. Now it is shown in Section 10.12 that the gated nature of the colour burst synchronizing signal puts a limit, for the 405-line system, of $\pm$ 5 kc/s on the frequency difference between the transmitted sub-carrier frequency and the free running frequency of the receiver's reference frequency generator. Hence, conventional synchronous working would be possible only if the mains supply frequency were held to within about 0·2 %, and for economical reasons in the colour receiver, preferably much less than this. A change of only 0·02 % in sub-carrier frequency (about 530 c/s) would be difficult to accommodate in the colour receiver, and would severely limit the choice of reference generator.

However, one method of synchronous working has been suggested which could be adopted in colour transmissions. The sub-carrier frequency is crystal controlled in the usual way so that colour receiver reference frequency generation is unaffected, but a mains synchronous, or very nearly mains synchronous, field frequency is derived by changing one of the dividing factors in the system. Thus, instead of varying the master oscillator frequency in sympathy with mains frequency variations, a division factor is changed appropriately. In this case, the line frequency could be crystal controlled by dividing down from the sub-carrier frequency, and the field factor 405 could be changed appropriately, but always to an odd number in order to preserve interlace. Alternatively, the sub-carrier frequency to twice line frequency factor ($\frac{525}{4}$ for the 405-line system) could be changed (again to an odd number of quarters to preserve the dot interlace), and in this case the number of lines would not alter, since the line and field frequencies would change together.

Although maintaining mains synchronous operation may involve complicated circuit arrangements, since the transmitter and not the receiver is involved, the complication is probably justified.

15.3. Carrier frequencies

In the 525- and 625-line N.T.S.C. systems the vision and sound carriers are the same as in their corresponding conventional

monochrome systems.   This is necessary in order to maintain the same frequency difference between the vision and sound carriers so that intercarrier sound operation is not affected on existing monochrome receivers.

In the British 405-line N.T.S.C. system, however, the difference between the vision and sound carriers is changed from the conventional 3·5 Mc/s to approximately 3·54 Mc/s (strictly, 3·54375 Mc/s). This change is a result of locking the frequency difference between the sub-carrier and the sound carrier to an odd multiple of half the line frequency so that any beat between the sound carrier and the sub-carrier has minimum subjective visibility.

The usual practice is to keep the same sound carrier frequency as in the conventional monochrome case, and to increase the vision carrier frequency by about 40 kc/s.   By this means the sound carrier traps in existing receivers are still correctly tuned, while the small increase in vision carrier frequency causes a very slight drop in carrier response at the nominal 6dB down point.

Thus, the 405-line N.T.S.C. vision carrier is very slightly higher than its conventional value, and this change has a negligible effect on monochrome receivers.

15.4.   Sub-carrier dots

The most significant difference between a conventional monochrome signal and an N.T.S.C. type of signal is the presence of the sub-carrier dots in those areas of the picture where the colour is not the Illuminant C reference white.   These dots produce two direct effects on a monochrome receiver: firstly, they appear as an interference effect, and secondly, they produce an increase in luminance due to rectification by the non-linear light output versus voltage input characteristic of the cathode-ray tube.   Both these effects increase with the purity times luminance product of the transmitted colour, since it is this product which determines the amplitude of the sub-carrier signal.

As far as the interference effect of the dot structure is concerned, a BBC report on compatibility tests on the 405-line system states that 2·7 % of a total of 624 engineer observers found the dot pattern " somewhat objectionable ", while 1·1 % found it " definitely objectionable ", on a motion picture test.   Furthermore, the picture size influences the results to the extent that more than one grade of subjective difference exists between pictures under 9 in. high compared with pictures over 12 in. high.   Thus, if an observer were to grade the dot structure as " somewhat objectionable " on a small picture, he would grade it as rather more than " definitely

objectionable " on a large picture. The subjective grading used in these tests was as follows:

Criterion	Grade
Imperceptible	1
Just perceptible	2
Definitely perceptible but not disturbing	3
Somewhat objectionable	4
Definitely objectionable	5
Unusable	6

It is interesting to note that the results obtained for static slides showed that 2·2% of observers found the dot pattern " somewhat objectionable " while 0·6% found it " definitely objectionable ". Since the colorimetric purity of scenes in the motion picture was not higher than that of the slides, it is suggested that the more unfavourable results for the moving pictures may have been caused by strobing effects which upset the dot integrating effect, and also there is the possibility that observers may have based their comments for the moving pictures on those scenes in which the dots were most conspicuous.

The dot structure on a monochrome display can be reduced either by rolling off the video response before the sub-carrier frequency, or by including a notch filter tuned to the sub-carrier frequency. The former method naturally severely restricts the picture definition, and is therefore not recommended, but it has been found that a notch filter with a rejection of about 12dB at the sub-carrier frequency and with a 3dB bandwidth of about 400 kc/s considerably reduces the large area dot structure in the case of the 405-line system, and produces only a slight degradation of definition. Thus, the $2\frac{1}{2}$ Mc/s bars of Test Card C have reduced contrast, but the 3 Mc/s bars and edge definition are scarcely affected.

It should be noted that only the large area dot structure can be reduced since the dot structure along edges corresponds to sidebands of the sub-carrier frequency which fall outside the greatest rejection range of the notch filter. Complete rejection of all dot structure can be achieved only by limiting the video response to about 1·7 Mc/s in the case of the 405-line system, with unreasonable consequences to definition.

In view of the results obtained from compatibility tests, the BBC recommend that a notch filter is essential in the case of the larger

displays, if these are not to become an obstacle to the introduction of a 405-line N.T.S.C. system.

15.5. Sub-carrier dot rectification

Besides appearing as an interfering pattern in coloured areas, the subcarrier also produces a contribution to the luminance in such areas owing to the non-linear law of the monochrome display tube.

Thus, consider a coloured area having a monochrome signal value $Y' = 0.30R^{\frac{1}{\gamma}} + 0.59G^{\frac{1}{\gamma}} + 0.11B^{\frac{1}{\gamma}}$ and a sub-carrier signal
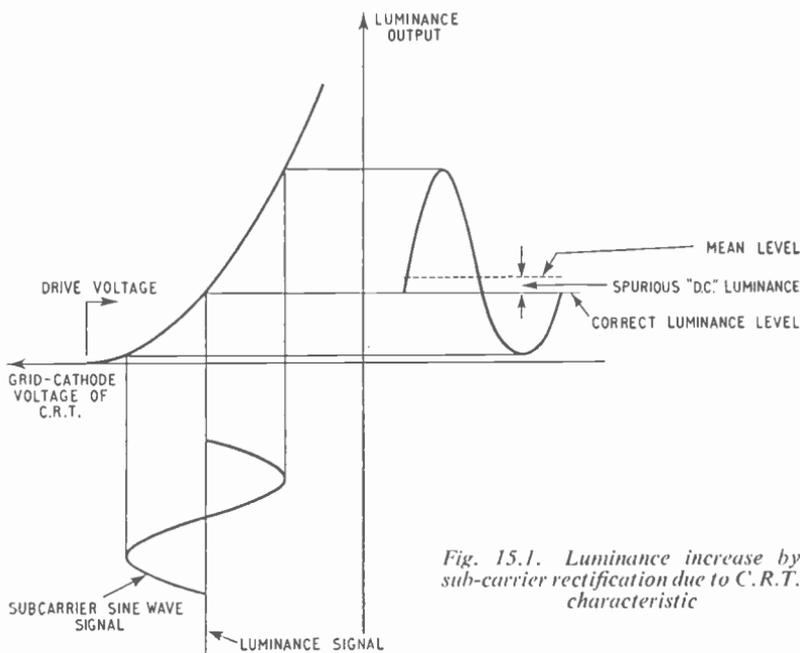
Fig. 15.1. Luminance increase by sub-carrier rectification due to C.R.T. characteristic

$S \sin \omega t$, displayed on a monochrome tube whose luminance output L is related to the signal drive V according to the equation

$$L = KV^{\gamma}$$

where K is a constant.

For a square law display, the luminance output for the colour signal will then be given by

$$L = K (Y' + S \sin \omega t)^2$$
$$= K\left[(Y')^2 + \frac{S^2}{2} + 2Y'S \sin \omega t - \frac{S^2}{2} \cos 2 \omega t\right]$$

The mean value of the displayed luminance is then

$$L = K \left[(Y')^2 + \frac{S^2}{2} \right]$$

Hence, rectification of the sub-carrier signal produces a contribution to the displayed luminance which is proportional to the square of the sub-carrier amplitude in the case of a square law display. Fig. 15.1 shows diagrammatically how the sub-carrier sine wave voltage produces a distorted luminance output waveform which has a spurious " D.C." component.

In practice, most monochrome tubes have a power law of about 2·7, so that a closer approximation to the display luminance would be

$$L = KV^3 = K \, (Y' + S \sin \omega t)^3$$

Using the identities

$$2 \sin^2 \theta = 1 - \cos 2\theta$$

and

$$2 \sin^3 \theta = \sin \theta - \sin \theta \cos 2\theta$$
$$= \sin \theta - \tfrac{1}{2} [\sin (\theta - 2\theta) + \sin (\theta + 2\theta)]$$

it follows that the mean luminance is

$$(Y')^3 + \frac{3 Y'}{2} S^2$$

It is interesting to evaluate the display luminance for the full primaries and their complementaries for the case of square and cube law displays. Table 15.1 illustrates the results.

It can be seen that the luminance contribution of the sub-carrier makes the displayed luminance approach the correct Y value for

Table 15.1

RELATIVE LUMINANCE VALUES OF SATURATED COLOURS

Colour	S	$(Y')^2$	$(Y')^2 + \dfrac{S^2}{2}$	Y	$(Y')^3 + \dfrac{3 Y'}{2} S^2$	$(Y')^3$
Red	0·63	0·090	0·288	0·30	0·205	0·027
Green	0·59	0·348	0·522	0·59	0·513	0·205
Blue	0·44	0·012	0·109	0·11	0·033	0·001
Cyan	0·63	0·490	0·688	0·70	0·760	0·343
Magenta	0·59	0·168	0·342	0·41	0·283	0·069
Yellow	0·44	0·792	0·889	0·89	0·965	0·705

both the square law and the cube law displays, and in fact, in the square law case, the displayed luminance is nearly correct for all six colours.

Notice that if the sub-carrier luminance contribution were absent, the displaced luminance would be proportional to the figures in the $(Y')^2$ or $(Y')^3$ columns, depending on the display law. These show considerable variations, and the effect on the luminance display is to produce a non-panchromatic characteristic. Thus, a red of luminance 0·3 would be displayed as a luminance of 0·09 or 0·027 (depending on the display law), while a green of luminance 0·3 would be displayed as a luminance of 0·188 or 0·082 for square and cube law displays, respectively, and for $\gamma = 2\cdot2$.

However, the BBC compatibility tests indicated that viewers show no marked preference for or against the presence of the sub-carrier signal, as far as its effect on tonal gradation is concerned.

15.6. Transient effects

The above discussion on the luminance contribution of the sub-carrier applies to large picture areas into which the relative frequency characteristics of the luminance and sub-carrier signals do not enter. But in the case of a sudden change from one colour to another, or even from one luminance level to another without a chromaticity change, it is obvious that the transient response will be affected by the restricted bandwidth of the sub-carrier signal. Since, for all colours other than Illuminant C, the sub-carrier carries some of the luminance information, then clearly this luminance component is restricted in bandwidth to the same extent as the chrominance signals, and hence the sub-carrier luminance contribution must have a relatively poor rise time. All luminance transients (except grey ones) are therefore built up of a fast Y' transient and a slow sub-carrier luminance transient.

In addition, it is possible for a monochrome receiver to display a luminance transient when the original transition is between two different colours of equal luminance. This effect can occur when the chromaticity passes near the white point during the transition, for then the sub-carrier luminance contribution is necessarily smaller than it was at the start of the transient, or will be at the end of the transient. Thus, the sub-carrier luminance contribution has a minimum value or a " notch " when the transition passes nearest to the white point, and because of the bandwidth limitation of the sub-carrier this notch is broadened out.

The initial and final values of a luminance transient, as displayed on a monochrome receiver, will depend on the actual colours between

which the transition occurs, and during the transient the luminance will also be dependent on the chromaticity path taken by the originating transition. As an illustration of the processes involved, consider a transition between a full pure red and a pure green of
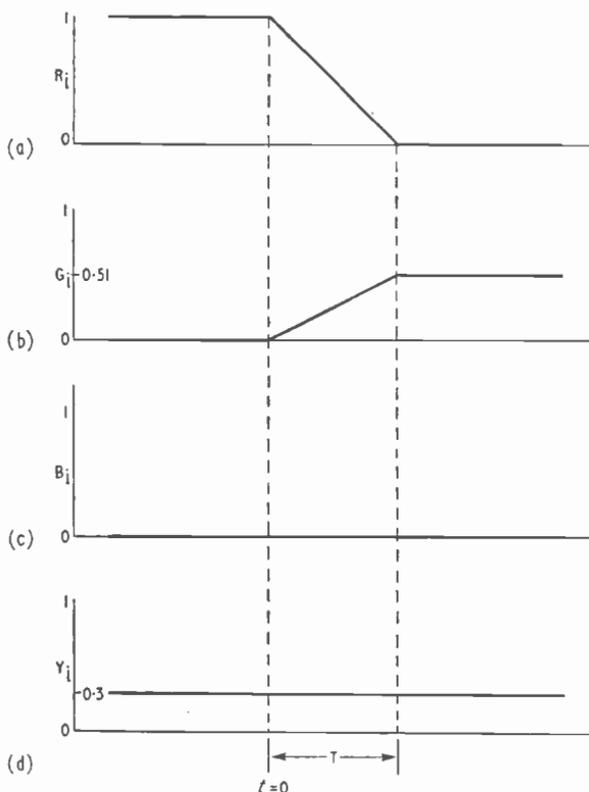
Fig. 15.2. Instantaneous tristimulus values and luminance of transition from full red to pure green of equal luminance before encoding

equal luminance. For simplicity, assume a gamma of 2 and a square law display so that the displayed luminance is proportional to

$$(Y')^2 + \frac{S^2}{2}$$

as derived in Section 15.5.

Let us further assume that the colour transition being televised leads to instantaneous tristimulus values which vary linearly with

time, so that if the time t is measured from the start of the transition which is complete after a time T

$$R_t = \left(1 - \frac{t}{T}\right) \qquad G_t = \frac{0 \cdot 30}{0 \cdot 59} \frac{t}{T} \qquad \text{and } B_t = 0$$

are the instantaneous numerical tristimulus values.

These values are plotted in Fig. 15.2 and notice that the true luminance Y_t remains constant throughout the transition.

Now the displayed luminance will have two components, one due to $(Y')^2$ and one due to $\frac{S^2}{2}$.

The luminance produced by the Y' signal will be

$$
\begin{aligned}
(Y')^2 &= [0 \cdot 30\ R_t{}^{\frac{1}{2}} + 0 \cdot 59\ G_t{}^{\frac{1}{2}} + 0 \cdot 11\ B_t{}^{\frac{1}{2}}]^2 \\
&= \left[0 \cdot 30\ (1-x)^{\frac{1}{2}} + 0 \cdot 59\ \left(\frac{0 \cdot 30}{0 \cdot 59} x\right)^{\frac{1}{2}}\right]^2 \\
&= 0 \cdot 09 + 0 \cdot 087x + 0 \cdot 6\ \sqrt{0 \cdot 177x\ (1 - x)}
\end{aligned}
$$

where $x = \dfrac{t}{T}$.

This luminance contribution is plotted in Fig. 15.3 (a) for values of x between 0 and 1, i.e. during the transient.

Since the sub-carrier amplitude S is given by

$$S^2 = \frac{(R' - Y')^2}{1 \cdot 14^2} + \frac{(B' - Y')^2}{2 \cdot 03^2}$$

and since

$$(R' - Y') = 0 \cdot 7\ (1 - x)^{\frac{1}{2}} - (0 \cdot 177x)^{\frac{1}{2}}$$

and

$$(B' - Y') = -0 \cdot 3\ (1 - x)^{\frac{1}{2}} - (0 \cdot 177x)^{\frac{1}{2}}$$

it follows that

$$
\frac{S^2}{2} = \frac{0 \cdot 49 - 0 \cdot 313x - 1 \cdot 4\ \sqrt{0 \cdot 177x\ (1 - x)}}{1 \cdot 14^2 . 2} \\
+ \frac{0 \cdot 09 + 0 \cdot 087x + 0 \cdot 6\ \sqrt{0 \cdot 177x\ (1 - x)}}{2 \cdot 03^2 . 2}
$$

which gives the instantaneous luminance contribution of the sub-carrier. This is plotted in Fig. 15.3(b) assuming that the transient
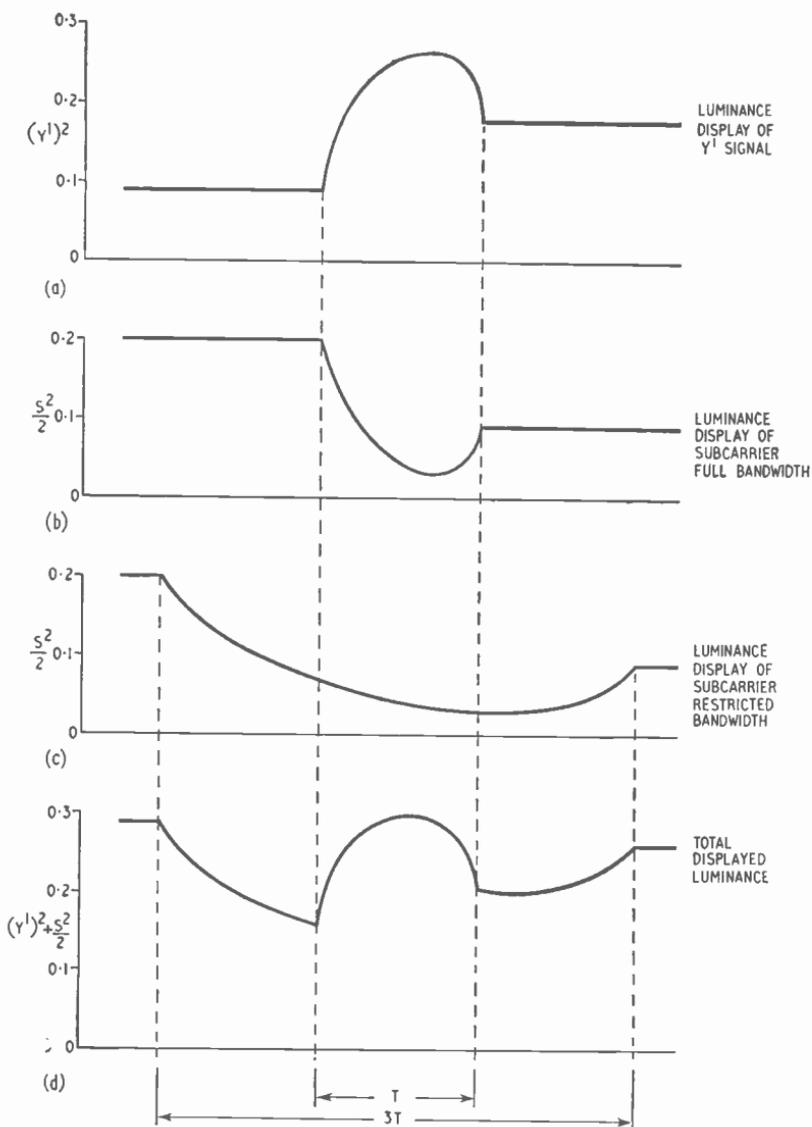
Fig. 15.3. Monochrome display of the transition from full red to pure green of equal luminance for a square law display. $\gamma = 2$

is complete after a time T.   However, since the sub-carrier signal has a restricted bandwidth, the curve has been re-plotted on the assumption that the transient takes three times as long, Fig. 15.3(c).

The total displayed luminance, which is the sum of the $(Y')^2$ and $\dfrac{S^2}{2}$ values, is shown in Fig. 15.3(d).

There are several points of interest in Fig. 15.3.   Notice that, although the correct luminance is constant, not only does $(Y')^2$ start and finish at different values but it also exhibits an overshoot. Hence, even if the sub-carrier were not transmitted, an entirely spurious luminance transient would be displayed.

Again, the graph of $(Y')^2 + \dfrac{S^2}{2}$ starts and finishes at different values, but these are closer to the correct luminance level of $0\cdot3$ than are the initial and final values of $(Y')^2$.   However, a slow and entirely spurious luminance transient is displayed.

Note that, in Fig. 15.3, it has been assumed that Y' is delayed at the transmitter so that the centre of the Y' transient coincides in time with the centre of the slower $\dfrac{S^2}{2}$ transient.   This makes the transient more symmetrical and therefore subjectively more acceptable.

Unlike the large area sub-carrier luminance contribution, the transient sub-carrier luminance effects cannot be removed by a sub-carrier rejector filter.   This is because the sidebands of the sub-carrier are significant during fast transients, and they would fall outside the rejection range of a notch filter.   Sub-carrier luminance transient effects can be removed only by limiting the video response of the monochrome receiver to exclude all possible sidebands of the sub-carrier, and this would lead to a video response of only about $1\cdot7$ Mc/s instead of the usual 3 Mc/s for the 405-line system.

15.7.   Burst visibility

Another effect associated with an N.T.S.C. type signal arises from the presence of the colour burst synchronizing signal, which is transmitted so that colour receivers may generate a continuous sine wave locked to the transmitter sub-carrier frequency for synchronously detecting the quadrature modulated sub-carrier.   This burst occurs during the post line synchronizing period, which usually forms part of the line flyback period of a monochrome receiver, and consequently the sub-carrier dots due to the burst are " stretched out " by the relatively fast flyback scan.   The effect of this is to produce vertical lines down the left hand side of the picture as shown

in Fig. 15.4. In general, these burst lines are visible only on those pictures which contain a black area on the left-hand side, and their visibility is a function of the D.C. gain of the receiver, the line flyback time, and the setting of the black level control. Complete invisibility of the burst lines can be achieved by applying a line blanking pulse to the C.R.T. or video stage, in a similar manner

Fig. 15.4. *Display of colour burst on a monochrome receiver. The effect has been emphasized for clarity*

to the field blanking which is often employed to remove the line scan during the field flyback.

Burst visibility on monochrome receivers does not appear to be a significant effect, for the BBC compatibility tests showed that only $1 \cdot 2\%$ of the general public, and $1 \cdot 8\%$ of engineers in the tests, found burst visibility " very annoying ".

15.8. Synchronizing effects

The sub-carrier signal occupies an appreciable part of the sync pulse range of the video waveform, particularly for full saturated red and blue, and for the colour burst. However, the colour burst occurs some time after line synchronization has taken place, and its

position does not change with picture content.   One would have expected to find greatest sensitivity to line synchronizing trouble for full saturated red or blue on the extreme right of the picture, where the sub-carrier signal would be only about 1μs from the leading edge of the line synchronizing pulse, and, of course, for greatest sensitivity to field synchronizing trouble, saturated red or blue at the bottom of the picture would appear to be the severest test.

The BBC compatibility tests indicate that synchronizing difficulties were mentioned by about 5% of observers, but the severity of the effects is difficult to assess.   Previous tests have shown that colour transmission causes little trouble with receiver synchronizing circuits. In the authors' experience, synchronizing trouble has been found only when the transmitted signal is not correctly balanced.   Thus, if sub-carrier is present on the synchronizing waveform (which, of course, is not in accordance with N.T.S.C. standards), then the receiver line time base synchronization is " pulled " so that sub-carrier dots on the picture are arranged in vertical lines, that is, the dot interlace is destroyed.

15.9. Sound buzz

It will be appreciated that the sub-carrier which is present in a colour signal is necessarily blanked at the line and field frequencies, in the same way that a conventional monochrome signal is blanked. Therefore, the sub-carrier and its sidebands are modulated with line and field frequency components.   The field frequency component can, under certain circumstances, cause a buzz in the sound channel of a receiver.   There are two mechanisms by which this buzz may enter the audio channel; either by cross-modulation in the R.F.-I.F. channel, or the sub-carrier sidebands in the upper video region may be accepted directly by the sound channel.

Adequate sound channel selectivity and freedom from cross-modulation should ensure satisfactory sound reception.   The BBC compatibility tests showed that $4\cdot4\%$ of the general public and $1\cdot8\%$ of engineer observers regarded sound buzz as being either moderately or very annoying.   The difference between the general public and engineer observer results may be due to misalignment of the general public receivers; for example, an incorrectly tuned sound I.F. could produce sound buzz.

15.10. Summary

An N.T.S.C. colour signal consists of a monochrome signal to which a sub-carrier signal, carrying the colour information, has been added.   When this signal is received on a monochrome set, the black

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and white picture produced differs in certain respects from a picture produced by a monochrome transmission.

The monochrome component of the colour signal differs from a black and white signal in two ways. Firstly, the line and field frequencies are constant and so the field frequency of the colour signal is not mains synchronous. Secondly, except for white and shades of grey, it is not a true measure of luminance. This difference occurs because the luminance signal of a colour transmission is formed by adding together the appropriate proportions of R, G and B signals, which have been individually gamma corrected, whereas the luminance signal of a black and white transmission is gamma corrected as a whole. The effect of this difference on reception is to cause spurious luminance transients.

The sub-carrier appears as an interfering signal which causes a dot pattern on those areas of the scene which are coloured when televised, but since the sub-carrier frequency is locked to the scanning frequencies, this pattern has minimum visibility.

The action of the sub-carrier on the non-linear light output versus voltage drive characteristic of the C.R.T. produces an increase of luminance. This increase tends to correct the luminance distortion which is caused by the monochrome component signal but this correction is only effective over large picture areas, because the sub-carrier has a limited bandwidth.

All colour transitions, except those between shades of grey, will be displayed as a fast transient superimposed on a slower transient which is approximately complementary in shape.

Minor effects caused by the monochrome reception of a colour signal can include visibility of the burst, scan synchronization effects and sound channel buzz.

Shortcomings of N.T.S.C. Systems

16.1. Introduction

An N.T.S.C. colour picture suffers from certain defects of a second order, and it is the purpose of this chapter to point out how they arise. It must be emphasized that these defects are not serious enough to qualify as objections to the N.T.S.C. system as such, and in fact many of them would occur in other types of system to a more marked degree. Most of the defects in the N.T.S.C. system can be removed by allowing additional complication at the receiver, though modification of the transmitter only can be effective in some cases. It should be noted that the term " defect " used in this section refers to a departure, however slight, from a theoretically perfect picture. The reader can be assured that, from the authors' personal viewing of N.T.S.C. pictures, the colour picture which suffers from all the defects mentioned below would nevertheless be described as " very good ", and would have a colour quality rather better than that obtained with good colour photographs.

The N.T.S.C. system discussed below is that which satisfies the specification given in Chapter 4. That is, three primary signals, R, G and B are each gamma corrected individually, then encoded in terms of full bandwidth Y' and differentially bandwidth limited I' and Q' signals having bandwidths of approximately one third and one ninth, respectively, of the Y' signal.

Criticism of such a system can cover many features. For example, only three primaries are used, with the result that some colours, particularly in the blue-green region, cannot be reproduced with full saturation. Again, the " detail colour blindness " quality of human vision is relied upon to produce the subjective effect of full definition in each primary colour. These defects are of a very small order, of course, and are small compared with the fact that the picture is necessarily quantized in a line structure.

Defects of a more noticeable order are concerned with constant luminance failure and distortion of transitions between colours,

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coupled with non-uniform noise performance. All these are a result of the type of gamma correction used which, while producing these defects, is nevertheless economical from the point of view of the receiver design.

Again, bandsharing of the video spectrum by the luminance and chrominance information, and single sideband transmission of the upper part of the I' spectrum, both contribute to spurious effects which are further complicated by the presence of noise.

Further defects arise from the use of asymmetric sideband transmission of the main vision carrier, or single sideband distortion, as it is often called. This particular effect is also a function of the sign of modulation used, while the use of F.M. or A.M. of the sound carrier can also produce different subjectively noticeable beat effects, particularly with regard to the relatively low frequency beat between the sound carrier and the R.F. chrominance sub-carrier. The relative performances of a negative modulation, F.M. sound N.T.S.C. signal and a positive modulation, A.M. sound N.T.S.C. signal are briefly discussed below, but in the authors' opinion the latter is unquestionably the better system as far as picture quality is concerned.

Let us consider some of these defects in more detail.

16.2. Constant luminance failure

One consequence of individual gamma correction of the R, G and B signals before encoding is that the luminance signal Y' then does not carry all the luminance information. The balance is therefore carried by the chrominance parameters I' and Q' (or $(R' - Y')$ and $(B' - Y')$) so that any noise which may be present in the chrominance channel can affect the displayed luminance. Hence a degradation of subjective noise performance is produced.

Let us consider the case of a large area colour, so that the effects of differential bandwidth limiting do not enter. Then the receiver display will have the signals Y', $(R' - Y')$, $(G' - Y')$, and $(B' - Y')$ applied to it, and if the display has a γ law then the displayed colour will have the following tristimulus values:—

Red
$$(Y' + R' - Y')^\gamma = (R')^\gamma = R$$

Green
$$(Y' + G' - Y')^\gamma = (G')^\gamma = G$$

Blue
$$(Y' + B' - Y')^\gamma = (B')^\gamma = B$$

Hence, the displayed colour will have the correct chromaticity and luminance. However, let us investigate how much of this luminance is carried by Y', and how much is carried by the difference signals. That is, let us find the displayed luminance in terms of the Y' and colour difference signals.

The displayed luminance is

$$0 \cdot 30R + 0 \cdot 59G + 0 \cdot 11B = Y$$

the correct value. Hence

$$Y = 0 \cdot 30[Y' + (R' - Y')]^\gamma + 0 \cdot 59[Y' + (G' - Y')]^\gamma$$
$$+ 0 \cdot 11[Y' + (B' - Y')]^\gamma$$

For simplicity, assume that $\gamma = 2$ (strictly $\gamma = 2 \cdot 2$), so that

$$Y = 0 \cdot 30[(Y')^2 + (R' - Y')^2 + 2Y'(R' - Y')]$$
$$+ 0 \cdot 59[(Y')^2 + (G' - Y')^2 + 2Y'(G' - Y')]$$
$$+ 0 \cdot 11[(Y')^2 + (B' - Y')^2 + 2Y'(B' - Y')]$$

Collecting terms

$$Y = (Y')^2 + 0 \cdot 30(R' - Y')^2 + 0 \cdot 59(G' - Y')^2 + 0 \cdot 11(B' - Y')^2$$
$$+ 2Y'[0 \cdot 30(R' - Y') + 0 \cdot 59(G' - Y') + 0 \cdot 11(B' - Y')]$$

But

$$Y' = 0 \cdot 30R' + 0 \cdot 59G' + 0 \cdot 11B' = (0 \cdot 30 + 0 \cdot 59 + 0 \cdot 11)Y'$$

$$\therefore \quad 0 \cdot 30(R' - Y') + 0 \cdot 59(G' - Y') + 0 \cdot 11(B' - Y') = 0$$

and hence

$$Y = (Y')^2 + 0 \cdot 30(R' - Y')^2 + 0 \cdot 59(G' - Y')^2 + 0 \cdot 11(B' - Y')^2$$

$$= (Y')^2 + Y_s{}^2$$

It can therefore be seen that, although the displayed luminance has the correct value of Y, some of this luminance that is, $Y_s{}^2$, is carried by the difference signals. The fundamental reason for this is that

$$Y' = 0 \cdot 30R^{\frac{1}{\gamma}} + 0 \cdot 59G^{\frac{1}{\gamma}} + 0 \cdot 11B^{\frac{1}{\gamma}}$$

instead of

$$Y' = (0 \cdot 30R + 0 \cdot 59G + 0 \cdot 11B)^{\frac{1}{\gamma}} = Y^{\frac{1}{\gamma}}$$

This point has been further discussed in Section 5.3. Notice that there is one exception to this failure of constant luminance,

i.e. if the transmitted colour is Illuminant C, then the difference signals are zero and $Y = (Y')^\gamma$. This arises because, if $R = G = B$ (which is so for Illuminant C), then

$$Y' = 0 \cdot 30 R^{\frac{1}{\gamma}} + 0 \cdot 59 G^{\frac{1}{\gamma}} + 0 \cdot 11 B^{\frac{1}{\gamma}} = (0 \cdot 30 + 0 \cdot 59 + 0 \cdot 11) R^{\frac{1}{\gamma}} = R^{\frac{1}{\gamma}}$$

and also

$$Y^{\frac{1}{\gamma}} = (0 \cdot 30 R + 0 \cdot 59 G + 0 \cdot 11 B)^{\frac{1}{\gamma}} = R^{\frac{1}{\gamma}}$$

The fraction of the true luminance carried by the Y' signal is given by $\dfrac{(Y')^2}{Y}$ for a γ of 2. This fraction is sometimes called the *constant luminance index*, and its ideal value is unity.
Therefore

$$\frac{(Y')^2}{Y} = \frac{(0 \cdot 30 R^{\frac{1}{2}} + 0 \cdot 59 G^{\frac{1}{2}} + 0 \cdot 11 B^{\frac{1}{2}})^2}{0 \cdot 30 R + 0 \cdot 59 G + 0 \cdot 11 B}$$

and if numerator and denominator are divided by $R + G + B$, then

$$\frac{(Y')^2}{Y} = \frac{(0 \cdot 30 r^{\frac{1}{2}} + 0 \cdot 59 g^{\frac{1}{2}} + 0 \cdot 11 b^{\frac{1}{2}})^2}{0 \cdot 30 r + 0 \cdot 59 g + 0 \cdot 11 b}$$

This equation shows how constant luminance failure varies with chromaticity; thus, for Illuminant C, $r = g = b$ and $\dfrac{(Y')^2}{Y} = 1$, but for pure blue

$$\frac{(Y')^2}{Y} = \frac{(0 \cdot 11 b^{\frac{1}{2}})^2}{0 \cdot 11 b} = 0 \cdot 11$$

so that for pure blue the Y' signal carries only 11% of the true luminance. On the other hand, for pure green the Y' signal carries 59% of the luminance.

It follows that the greater the purity of a colour, the greater is the failure of constant luminance, and the nearer the dominant wavelength is to the lowest luminance primary, that is blue, the greater is the failure of constant luminance. Thus, constant luminance failure is zero for Illuminant C, and a maximum for pure blue. It would therefore appear that the amount of luminance carried by the chrominance parameters is a function of both the purity and the dominant wavelength of the colour.

16.3. The elliptical sub-carrier

The present N.T.S.C. specification leads to an elliptical relation between sub-carrier amplitude per unit Y' and sub-carrier phase,

for a given constant luminance index and a gamma of two.   This may be shown from the equation

$$\frac{Y}{(Y')^2} = 1 + \frac{0\cdot30(R' - Y')^2 + 0\cdot59(G' - Y')^2 + 0\cdot11(B' - Y')^2}{(Y')^2}$$

which was derived in the previous section, by substituting

$$0\cdot59(G' - Y') = -0\cdot30(R' - Y') - 0\cdot11(B' - Y')$$

$$(R' - Y') = 1\cdot14S \sin \theta$$

and

$$(B' - Y') = 2\cdot03S \cos \theta$$

where S and θ are the sub-carrier amplitude and phase.

Hence

$$\frac{Y}{(Y')^2} = 1 + \frac{S^2}{(Y')^2}\left(0\cdot30 \times 1\cdot14^2\left(1 + \frac{0\cdot30}{0\cdot59}\right)\sin^2 \theta\right.$$

$$+ 0\cdot11 \times 2\cdot03^2\left(1 + \frac{0\cdot11}{0\cdot59}\right)\cos^2 \theta$$

$$\left. + \frac{2 \times 0\cdot3 \times 0\cdot11 \times 1\cdot14 \times 2\cdot03}{0\cdot59} \sin \theta \cos \theta\right)$$

$$= 1 + \frac{S^2}{(Y')^2}(0\cdot59 \sin^2 \theta + 0\cdot54 \cos^2 \theta + 2 \times 0\cdot13 \sin \theta \cos \theta)$$

$$\therefore 0\cdot59 \sin^2 \theta + 0\cdot54 \cos^2 \theta + 0\cdot13 \times 2 \sin \theta \cos \theta = \frac{1}{\left(\dfrac{\dfrac{S}{Y'}}{\sqrt{\dfrac{1}{K} - 1}}\right)^2}$$

where $K = \dfrac{(Y')^2}{Y}$ is the constant luminance index.

Now the polar equation to an ellipse of semi-axes a and b, the former of which is rotated through an angle μ relative to $\theta = 0$, is given by

$$\frac{1}{a^2} \cos^2(\theta - \mu) + \frac{1}{b^2} \sin^2(\theta - \mu) = \frac{1}{r^2}$$

or else it may be given by

$$\left(\frac{1}{a^2}\sin^2\mu + \frac{1}{b^2}\cos^2\mu\right)\sin^2\theta + \left(\frac{1}{a^2}\cos^2\mu + \frac{1}{b^2}\sin^2\mu\right)\cos^2\theta$$

$$+ \left(\frac{1}{a^2} - \frac{1}{b^2}\right)\sin\mu\cos\mu \times 2\sin\theta\cos\theta = \frac{1}{r^2}$$

Hence, the equation between $\dfrac{\dfrac{S}{Y'}}{\sqrt{\dfrac{1}{K}-1}}$ and θ is an ellipse whose

constants μ, a and b, can be determined from the equations

$$\frac{1}{a^2}\sin^2\mu + \frac{1}{b^2}\cos^2\mu = 0\cdot59$$

$$\frac{1}{a^2}\cos^2\mu + \frac{1}{b^2}\sin^2\mu = 0\cdot54$$

and

$$\left(\frac{1}{a^2} - \frac{1}{b^2}\right)\sin\mu\cos\mu = 0\cdot13$$

Hence

$$\mu = 50\cdot5°, \;\; a = 1\cdot2 \;\; \text{and} \;\; b = 1\cdot5$$

The equation between θ and $\left(\dfrac{S}{Y'}\right)$ may therefore be written as

$$\frac{1}{1\cdot2^2}\cos^2(\theta - \mu) + \frac{1}{1\cdot5^2}\sin^2(\theta - \mu) = \frac{1}{\left[\dfrac{1}{\sqrt{\dfrac{1}{K}-1}} \dfrac{S}{Y'}\right]^2}$$

or

$$\frac{1}{\left(\dfrac{S}{Y'}\right)^2} = \frac{K}{(1-K)}\left[\frac{1}{1\cdot2^2}\cos^2(\theta - \mu) + \frac{1}{1\cdot5^2}\sin^2(\theta - \mu)\right]$$

$$= \frac{1}{\left(1\cdot2\sqrt{\dfrac{1-K}{K}}\right)^2}\cos^2(\theta - \mu) + \frac{1}{\left(1\cdot5\sqrt{\dfrac{1-K}{K}}\right)^2}\sin^2(\theta - \mu)$$

For each value of the constant luminance index therefore, the sub-carrier amplitude per unit Y' describes an ellipse as θ varies, and the smaller the value of K, the larger are the semi-axes of the ellipse, as Fig. 16.1 shows.   This illustrates that constant luminance failure becomes more severe as the chromaticity approaches the perimeter of the primary triangle.   Note that the largest ellipse
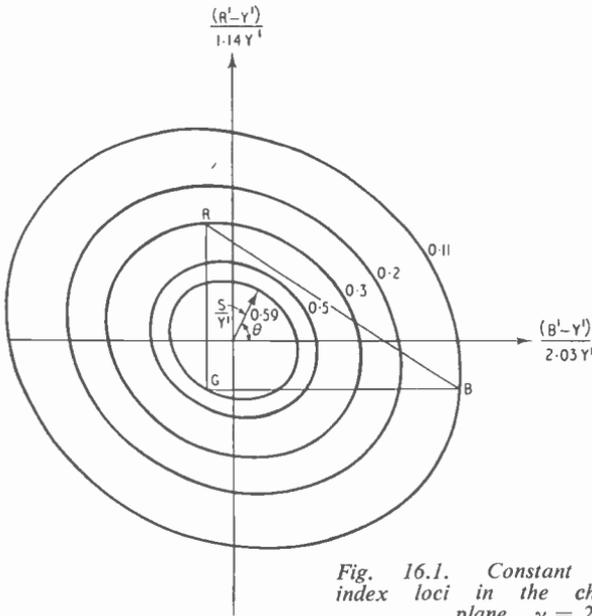
Fig. 16.1.   Constant   luminance index   loci   in   the   chrominance plane. $\gamma = 2$

which can be drawn to include a physical chromaticity passes through the blue primary, showing that constant luminance failure is most severe for the pure blue chromaticity.

16.4.   The circular sub-carrier

If the sub-carrier amplitude per unit Y' were circularly related to the sub-carrier phase for a given value of the constant luminance index, then the latter would be independent of the sub-carrier phase. That is, all the luminance carried by the sub-carrier would be a function only of the sub-carrier amplitude per unit Y''.   In this case, it would be possible for the transmitter to derive a correction signal from the sub-carrier amplitude, and add this to the Y' signal. The modified Y' signal would then carry all the luminance, so constant luminance operation would be achieved, and in addition,

monochrome receivers would operate panchromatically, provided they were fitted with a sub-carrier rejector. Colour receivers would require some additional circuitry for deriving the conventional Y' signal from the corrected signal, of course, otherwise they would display too much luminance. Remember that, as far as a colour receiver is concerned, correct large area luminance is achieved when the conventional Y' signal is transmitted.

A circular sub-carrier can be achieved by a modification of the sub-carrier make-up at the transmitter. Suppose instead of modulating one sub-carrier with $\dfrac{(R'-Y')}{1 \cdot 14}$ and the quadrature sub-carrier with $\dfrac{(B'-Y')}{2 \cdot 03}$, we modulate one sub-carrier with

$$[(R' - Y') + \lambda(B' - Y')]C$$

and the quadrature sub-carrier with $(B' - Y')D$, where λ, C, and D are suitable constants.

Then, if S and θ are the sub-carrier amplitude and phase

$$[(R' - Y') + \lambda(B' - Y')]C = S \sin \theta$$

and

$$(B' - Y')D = S \cos \theta$$

Now if $\gamma = 2$, the luminance contribution of the sub-carrier is
$$Y_s{}^2 = 0 \cdot 30(R' - Y')^2 + 0 \cdot 59(G' - Y')^2 + 0 \cdot 11(B' - Y')^2$$

$$= \left(0 \cdot 30 + \frac{0 \cdot 30^2}{0 \cdot 59}\right)(R' - Y')^2 + \left(0 \cdot 11 + \frac{0 \cdot 11^2}{0 \cdot 59}\right)(B' - Y')^2$$
$$+ \frac{0 \cdot 066}{0 \cdot 59}(R' - Y')(B' - Y')$$

But

$$(B' - Y') = \frac{S \cos \theta}{D}$$

and

$$(R' - Y') = \frac{S \sin \theta - C\lambda(B' - Y')}{C}$$
$$= \frac{S \sin \theta - \dfrac{C\lambda S}{D} \cos \theta}{C}$$
$$= S\left(\frac{\sin \theta}{C} - \frac{\lambda}{D} \cos \theta\right)$$

Substituting in the equation for $Y_s{}^2$, it follows that

$$Y_s{}^2 = 0\cdot30\left(1 + \frac{0\cdot30}{0\cdot59}\right)S^2\left(\frac{\sin\theta}{C} - \frac{\lambda}{D}\cos\theta\right)^2$$

$$+ 0\cdot11\left(1 + \frac{0\cdot11}{0\cdot59}\right)\frac{S^2}{D^2}\cos^2\theta$$

$$+ \frac{0\cdot066}{0\cdot59} \times \frac{S^2}{D}\left(\frac{\sin\theta}{C} - \frac{\lambda}{D}\cos\theta\right)\cos\theta$$

$$\therefore \quad \frac{0\cdot59\,Y_s{}^2}{S^2} = \frac{1}{D^2}(0\cdot267\lambda^2 + 0\cdot077 - 0\cdot066\lambda)\cos^2\theta$$

$$+ \frac{0\cdot267}{C^2}\sin^2\theta + \frac{1}{DC}(0\cdot066 - 0\cdot534\lambda)\sin\theta\cos\theta$$

If this is to be independent of θ, then

$$\frac{(0\cdot267\lambda^2 + 0\cdot077 - 0\cdot066\lambda)}{D^2} = \frac{0\cdot267}{C^2}$$

and

$$0\cdot066 - 0\cdot534\lambda = 0$$

in which case

$$\frac{0\cdot59\,Y_s{}^2}{S^2} = \frac{0\cdot267}{C^2}$$

which is independent of θ.

Solving the two equations, $\lambda = 0\cdot124$ and $C = 1\cdot91\,D$

The sub-carrier make-up for circularity is therefore

$$1\cdot91\,D[(R' - Y') + 0\cdot124(B' - Y')]$$

modulated on to one sub-carrier, instead of the conventional $\dfrac{(R' - Y')}{1\cdot14}$, and $D(B' - Y')$ modulated on to the quadrature sub-carrier, instead of the conventional $\dfrac{(B' - Y')}{2\cdot03}$.

The circular sub-carrier amplitude is given by

$$D\sqrt{1\cdot91^2[(R' - Y') + 0\cdot124(B' - Y')]^2 + (B' - Y')^2}$$

where the constant D may be chosen to limit the sub-carrier excursions in the video range.　The greatest excursion must occur for one of the full purity complementaries, and evaluating for these three colours, we have the results shown in Table 16.1.

Table 16.1

CHOICE OF CIRCULAR SUB-CARRIER WEIGHTING FACTOR

Colour	Y'	$D\sqrt{1\cdot91^2[(R'-Y')+\cdot124(B'-Y')]^2+(B'-Y')^2}$	Sum
Cyan	0·70	$1\cdot3D$	$0\cdot70+1\cdot3D$
Magenta	0·41	$1\cdot4D$	$0\cdot41+1\cdot4D$
Yellow	0·89	$0\cdot89D$	$0\cdot89+0\cdot89D$

If the greatest sub-carrier excursions are limited to $1\frac{1}{3}$ times the 0 to 1 video range, as in the conventional case, the possible values of D are given by the equations

$$0\cdot70 + 1\cdot3D \ = 1\cdot33, \quad \text{or } D = 0\cdot485$$

$$0\cdot41 + 1\cdot4D \ = 1\cdot33, \quad \text{or } D = 0\cdot655$$

$$0\cdot89 + 0\cdot89D = 1\cdot33, \quad \text{or } D = 0\cdot495$$

The smallest value of D must be chosen, i.e. 0·485, so that the maximum excursion of 1·33 is not exceeded for any colour. (For example, if $D = 0\cdot655$, then $0\cdot70 + 1\cdot3D = 1\cdot55$, which exceeds

Table 16.2

MAXIMUM SIGNAL EXCURSIONS FOR SATURATED COLOURS USING CIRCULAR SUB-CARRIER SIGNAL

Colour	Y'	$Y' + S$	$Y' - S$	S
Red	0·30	0·93 (0·93)	−0·33 (−0·33)	0·63 (0·63)
Green	0·59	1·27 (1·18)	−0·09 (0·00)	0·68 (0·59)
Blue	0·11	0·54 (0·54)	−0·32 (−0·33)	0·43 (0·44)
Cyan	0·70	1·33 (1·33)	0·07 (0·07)	0·63 (0·63)
Magenta	0·41	1·09 (1·00)	−0·27 (−0·18)	0·68 (0·59)
Yellow	0·89	1·32 (1·33)	0·46 (0·45)	0·43 (0·44)

1·33.)　The excursions for all six full purity colours can now be calculated, assuming $D = 0\cdot485$ (see Table 16.2).

The conventional N.T.S.C. figures are given in brackets, and it can be seen that apart from blue and yellow, the sub-carrier amplitude is greater than or equal to the conventional values.　Hence, slightly improved signal-to-noise ratios should be obtained except

for blue and yellow, but it can be seen that the differences are quite small.

For decoding the circular sub-carrier at the receiver, the synchronous detection axes must be altered slightly compared with those required for the conventional elliptical sub-carrier.

The transmission axes are shown in Fig. 16.2. The output of a synchronous detector operating along the ϕ axis shown will give an output

$$0.485[1.91(R' - Y') + 0.124 \times 1.91(B' - Y')] \cos \phi$$
$$- 0.485(B' - Y') \sin \phi$$

$$= 0.485[1.91 \cos \phi(R' - Y') + (0.124 \times 1.91 \cos \phi$$
$$- \sin \phi)(B' - Y')]$$

If this is to be a pure $(R' - Y')$ signal, then

$$\sin \phi = 0.124 \times 1.91 \cos \phi$$

whence

$$\phi = 13.3°$$

The output will then be

$$0.485 \times 1.91 \cos \phi(R' - Y') = 0.9(R' - Y')$$

Hence, if two synchronous detectors are used to obtain the red and blue difference signals, they must operate with a phase difference of
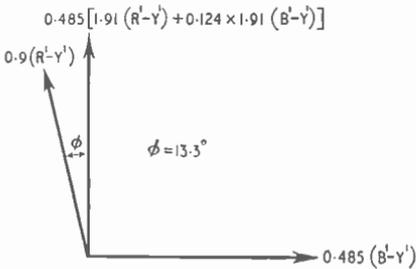
Fig. 16.2. Transmission axes for a circular sub-carrier

103.3° instead of the 90° separation required for a conventional elliptical sub-carrier.

16.5. Effects of non-linear encoding of chrominance parameters on displayed chromaticity

The conventional gamma correction leads to constant luminance failure, and hence a subjective degradation of performance results when noise is present. This effect is brought about by the luminance

carrying property of the chrominance signals, and it would seem that these signals tend to change luminance rather than chrominance whenever severe constant luminance failure is occurring. This immediately suggests that the extent of chrominance control by the chrominance parameters will vary so that for chromaticities where constant luminance failure is severe, the control is weak, and the chrominance signals will not have much effect on the chrominance of the display. Since there is no large area colour distortion, the control of the chrominance signals over the displayed chrominance must be correspondingly greater for chromaticities where there is only mild constant luminance failure. Thus for low purity colours, a given change in a chrominance signal will produce a relatively large change in displayed chrominance but only a small change in displayed luminance.

For high purity colours a given change in a chrominance signal will produce a relatively small change in displayed chrominance but a large change in displayed luminance.

The effects of the luminance controlled by the chrominance signals have been considered, so now it remains to examine the displayed chrominance controlled by these signals; and more particularly, the relationship between the sub-carrier amplitude and the displayed purity, and between the sub-carrier phase and the displayed dominant wavelength will be investigated, so that the effects of noise fluctuations on the received chromaticity can be estimated.

16.6. Sub-carrier amplitude control of displayed purity

To examine the control of sub-carrier amplitude over the purity of the displayed colour, consider a colour having a total tristimulus value D and a chromaticity

$$r = a + p, \quad g = a, \quad b = a$$

Since $r + g + b = 1$, it follows that

$$3a + p = 1$$

so that

$$a = \frac{(1 - p)}{3}$$

The chromaticity is therefore

$$r = a + p = \frac{(1 + 2p)}{3}, \qquad b = a = \frac{(1 - p)}{3}$$

Referring to Fig. 16.3, this chromaticity has a dominant wavelength equal to that of the red primary, and as p is varied the purity will vary but the dominant wavelength will remain constant. If $p = 0$, the purity is zero and the chromaticity is Illuminant C. If $p = 1$ the purity is 100% and the chromaticity is pure red. Now let us find the relationship between the sub-carrier amplitude S and the quantity p, and determine how variation of one will affect the
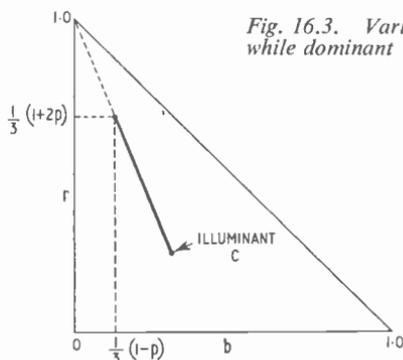
Fig. 16.3.   Variation of p changes purity, while dominant λ remains constant at red

other.   Notice that the dominant wavelength will remain constant at red so that the sub-carrier phase does not enter as a variable. (This is explained in the next section.)

It has been shown earlier that the sub-carrier amplitude is given by

$$S = \sqrt{\alpha^2(R' - Y')^2 + \beta^2(B' - Y')^2}$$

where

$$\alpha = \frac{1}{1\cdot14}, \qquad \beta = \frac{1}{2\cdot03}$$

Since $r = \dfrac{R}{D}$, or $R = rD$, where $D = R + G + B$ (*see* Chapter 2) it

follows by substitution that

$$\frac{S^2}{D^\gamma} = \alpha^2\left(r^{\frac{1}{\gamma}} - 0\cdot30r^{\frac{1}{\gamma}} - 0\cdot59g^{\frac{1}{\gamma}} - 0\cdot11b^{\frac{1}{\gamma}}\right)^2$$

$$+ \beta^2\left(b^{\frac{1}{\gamma}} - 0\cdot30r^{\frac{1}{\gamma}} - 0\cdot59g^{\frac{1}{\gamma}} - 0\cdot11b^{\frac{1}{\gamma}}\right)^2$$

Substituting the values of r, g and b in terms of p we have

$$\frac{3^{\frac{2}{\gamma}}S^2}{D^{\frac{2}{\gamma}}} = \alpha^2\Big[0.70(1 + 2p)^{\frac{1}{\gamma}} - 0.59(1 - p)^{\frac{1}{\gamma}} - 0.11(1 - p)^{\frac{1}{\gamma}}\Big]^2$$

$$+ \beta^2\Big[0.89(1 - p)^{\frac{1}{\gamma}} - 0.30(1 + 2p)^{\frac{1}{\gamma}} - 0.59(1 - p)^{\frac{1}{\gamma}}\Big]^2$$

$$= 0.70^2\alpha^2\Big[(1 + 2p)^{\frac{1}{\gamma}} - (1 - p)^{\frac{1}{\gamma}}\Big]^2 + 0.30^2\beta^2\Big[(1 + 2p)^{\frac{1}{\gamma}} - (1 - p)^{\frac{1}{\gamma}}\Big]^2$$

i.e.

$$\left(\frac{3^{\frac{1}{\gamma}}S}{D^{\frac{1}{\gamma}}}\right)^2 = \Big[(1 + 2p)^{\frac{1}{\gamma}} - (1 - p)^{\frac{1}{\gamma}}\Big]^2(0.70^2\alpha^2 + 0.30^2\beta^2)$$

or

$$S = K\frac{D^{\frac{1}{\gamma}}}{3^{\frac{1}{\gamma}}}\Big[(1 + 2p)^{\frac{1}{\gamma}} - (1 - p)^{\frac{1}{\gamma}}\Big]$$

where

$$K = \sqrt{0.70^2\alpha^2 + 0.30^2\beta^2} = 0.63$$

Now find how S changes when a change is made in p, assuming that D remains constant all the time.

Differentiating

$$\frac{\mathrm{d}S}{\mathrm{d}p} = \frac{K}{\gamma}\left(\frac{D}{3}\right)^{\frac{1}{\gamma}}\Big[2(1 + 2p)^{\frac{1}{\gamma}-1} + (1 - p)^{\frac{1}{\gamma}-1}\Big]$$

If $p = 0$

$$\left|\frac{\mathrm{d}S}{\mathrm{d}p}\right|_{p=0} = \frac{K}{\gamma}D^{\frac{1}{\gamma}}3^{1-\frac{1}{\gamma}}$$

If $p \to 1$

$$\left|\frac{\mathrm{d}S}{\mathrm{d}p}\right|_{p\to1} \to \infty$$

(since

$$(1 - p)^{\frac{1}{\gamma}-1} = \frac{1}{(1 - p)^{1-\frac{1}{\gamma}}} \to \infty \text{ if } \gamma > 1 \text{ and } p \to 1)$$

Hence, for small values of p, S increases slowly as p increases. For large values of p, S increases rapidly as p increases.

A plot of S against p is shown in Fig. 16.4, and it can be seen that equal increments in p, produced by changing the purity of the transmitted colour, produce small changes in S for low values of p, and large changes in S for high values of p. At the receiver, a change in the received value of S has a large effect on p for low
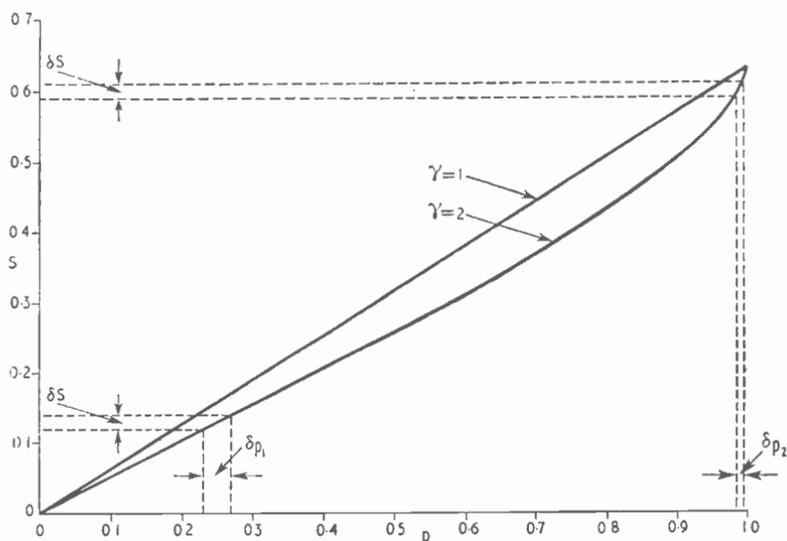
Fig. 16.4.   Relation between sub-carrier amplitude and p for cases where $D = 1, \gamma = 2$; and $D = 1, \gamma = 1$

values of S, and a small effect on p for high values of S. Consequently, noise will change p more markedly at low values of p, and less markedly at high values of p.

In the linear case, $\gamma = 1$ and $S = K Dp$ and $\dfrac{dS}{dp} = KD$.

Hence, noise fluctuations will produce equal effects on p, independent of p. That is, noise performance will be uniform for $\gamma = 1$.

16.7.   Sub-carrier phase control of displayed dominant wavelength

Let us now examine the relationship between the phase of the sub-carrier and the dominant wavelength of a colour, and hence determine the relative control of one over the other.

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If θ is the phase of the sub-carrier relative to the blue difference axis, then

$$\tan \theta = \frac{\alpha(R' - Y')}{\beta(B' - Y')}$$

where $\alpha = \dfrac{1}{1\cdot 14}$ and $\beta = \dfrac{1}{2\cdot 03}$

Substituting $0\cdot 30R' + 0\cdot 59G' + 0\cdot 11B'$ for Y', it follows that

$$\tan \theta = \frac{\alpha}{\beta}\left(\frac{0\cdot 70R^{\frac{1}{\gamma}} - 0\cdot 59G^{\frac{1}{\gamma}} - 0\cdot 11B^{\frac{1}{\gamma}}}{0\cdot 89B^{\frac{1}{\gamma}} - 0\cdot 59G^{\frac{1}{\gamma}} - 0\cdot 30R^{\frac{1}{\gamma}}}\right)$$

Dividing numerator and denominator by $(R + G + B)^{\frac{1}{\gamma}}$

$$\frac{\beta}{\alpha} \tan \theta = \frac{0\cdot 70r^{\frac{1}{\gamma}} - 0\cdot 59g^{\frac{1}{\gamma}} - 0\cdot 11b^{\frac{1}{\gamma}}}{0\cdot 89b^{\frac{1}{\gamma}} - 0\cdot 59g^{\frac{1}{\gamma}} - 0\cdot 30r^{\frac{1}{\gamma}}}$$

$$= \frac{0\cdot 70(r^{\frac{1}{\gamma}} - g^{\frac{1}{\gamma}}) - 0\cdot 11(b^{\frac{1}{\gamma}} - g^{\frac{1}{\gamma}})}{0\cdot 89(b^{\frac{1}{\gamma}} - g^{\frac{1}{\gamma}}) - 0\cdot 30(r^{\frac{1}{\gamma}} - g^{\frac{1}{\gamma}})}$$

Hence, for a given value of θ, r can be plotted against b, remembering that $r + g + b = 1$. By repeating these plots for various values of θ, a family of curves is obtained, as shown in Fig. 16.5. Notice that these curves converge towards the primaries and diverge away from the complementaries, so that a change in the dominant wavelength of a chromaticity produces a relatively large change in the sub-carrier phase angle in the region of a high purity primary dominant wavelength and a relatively small change in phase angle in the region of a high purity complementary. In terms of receiver noise performance, therefore, the displayed dominant wavelength of high purity complementaries will be more affected by phase fluctuations than will the dominant wavelength of high purity primaries.

For low purity chromaticities, the relationship between dominant wavelength and sub-carrier phase angle is similar to the linear or $\gamma = 1$ case.

For consider a chromaticity $r = \frac{1}{3} + \delta_1$, $g = \frac{1}{3} + \delta_2$, $b = \frac{1}{3} + \delta_3$, where the δ quantities are small compared with $\frac{1}{3}$.

Then $r^{\frac{1}{\gamma}} = (\tfrac{1}{3} + \delta_1)^{\frac{1}{\gamma}} = (\tfrac{1}{3})^{\frac{1}{\gamma}}(1 + 3\delta_1)^{\frac{1}{\gamma}} = (\tfrac{1}{3})^{\frac{1}{\gamma}}\left(1 + \dfrac{3\delta_1}{\gamma}\right)$, by the Binomial expansion.

Similarly, $g^{\frac{1}{\gamma}} = (\tfrac{1}{3})^{\frac{1}{\gamma}}\left(1 + \dfrac{3\delta_2}{\gamma}\right)$ and $b^{\frac{1}{\gamma}} = (\tfrac{1}{3})^{\frac{1}{\gamma}}\left(1 + \dfrac{3\delta_3}{\gamma}\right)$

Substituting in the expansion for $\dfrac{\beta}{\alpha}\tan\theta$ given earlier

$$
\begin{aligned}
\frac{\beta}{\alpha}\tan\theta &= \frac{0{\cdot}70\left(1 + \dfrac{3\delta_1}{\gamma} - 1 - \dfrac{3\delta_2}{\gamma}\right) - 0{\cdot}11\left(1 + \dfrac{3\delta_3}{\gamma} - 1 - \dfrac{3\delta_2}{\gamma}\right)}{0{\cdot}89\left(1 + \dfrac{3\delta_3}{\gamma} - 1 - \dfrac{3\delta_2}{\gamma}\right) - 0{\cdot}30\left(1 + \dfrac{3\delta_1}{\gamma} - 1 - \dfrac{3\delta_2}{\gamma}\right)} \\[2mm]
&= \frac{0{\cdot}70(\delta_1 - \delta_2) - 0{\cdot}11(\delta_3 - \delta_2)}{0{\cdot}89(\delta_3 - \delta_2) - 0{\cdot}30(\delta_1 - \delta_2)} \\[2mm]
&= \frac{0{\cdot}70(r - g) - 0{\cdot}11(b - g)}{0{\cdot}89(b - g) - 0{\cdot}30(r - g)}
\end{aligned}
$$

which is the same as the linear case.

Notice that in the linear case, a linear relationship exists between $\tan\theta$ and the chromaticity co-ordinates, so that a particular value of θ will completely specify the dominant wavelength. In the case where $\gamma \neq 1$, however, this is true only for the three particular cases $r = g$, $r = b$, and $g = b$. These three equations give straight line loci on the chromaticity diagram which pass through the white point and the appropriate primary and complementary points. The dominant wavelengths of all chromaticities lying on these loci are therefore completely specified by the sub-carrier phase angle, but for all other chromaticities the phase angle is not a sufficient specification of the dominant wavelength, and another parameter, such as the sub-carrier amplitude for a given luminance, is required as well. This leads to the conclusion that, unless the displayed dominant wavelength is that of a primary or complementary, the displayed dominant wavelength will be a function of both the saturation and hue control settings of the receiver.

It should be pointed out that Fig. 16.5 shows the connection between chromaticity and sub-carrier phase angle, and if the subjective effects of phase variations are required, the non-linear property of the conventional chromaticity diagram in terms of equally perceptible chromaticity differences must be taken into account, as described in Section 2.23.

16.8.   Effects of gamma correction on transient response

The preceding sections have been concerned with the effects of gamma correction on the display of large area colours for which the limited bandwidth of the chrominance signals is not involved. However, for relatively small area colours, or for transitions between different colours, it is to be expected that certain disturbances will occur as a result of constant luminance failure.

Basically, distortion of transients arises because the luminance is carried partly by the high definition Y' signal and partly by the relatively low definition chrominance signals. Hence, if there is any sudden change from one colour to another which also involves a change in the luminance contribution of the chrominance signals, the resulting displayed colour change will have a fast luminance transient controlled by the Y' signal superimposed upon a relatively slow luminance transient controlled by the bandwidth limited chrominance signals. In addition, the instantaneous chromaticity
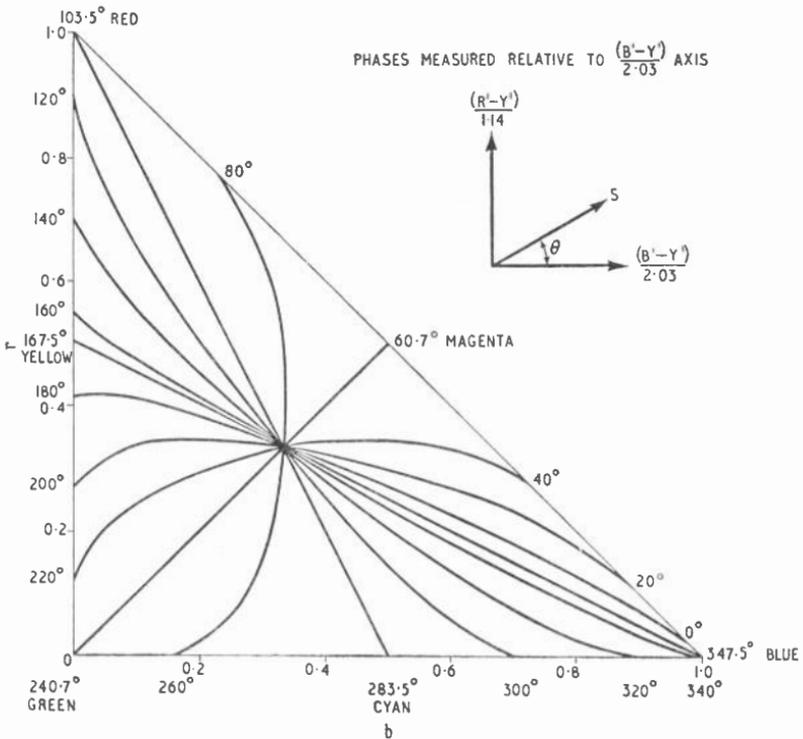
Fig. 16.5.   *Chromaticity loci for constant sub-carrier phases,* $\gamma = 2$

of the originating transient will not be faithfully reproduced, but since the human eye is more sensitive to luminance changes these will provide the greater subjective effects.

The exact form of transient distortion will depend on the actual colours before and after the transient, and also during the transient. In Section 16.2, it was shown that, for the case $\gamma = 2$, the displayed luminance is given by

$$Y = (Y')^2 + Y_s{}^2$$

Hence

$$\mathrm{d}\,Y = 2\,Y'\mathrm{d}\,Y' + 2\,Y_s\,\mathrm{d}\,Y_s$$

so that for a luminance change $\mathrm{d}\,Y$ in the original picture, this will be correctly displayed if Y' and Y_s are present at the display tube. But if $\mathrm{d}\,Y$ is a fast transient, only Y' will actually reach the receiver as a full bandwidth signal, and the ratio

$$\frac{\text{luminance displayed}}{\text{luminance in original picture}} = \frac{\mathrm{d}(Y')^2}{\mathrm{d}\,Y} = \frac{Y'\mathrm{d}\,Y'}{Y'\mathrm{d}\,Y' + Y_s\mathrm{d}\,Y_s}$$

gives a measure of the distortion. The ideal value of this ratio is unity, in which case there would be no distortion, but this can occur only for the cases $Y_s = 0$ or $\mathrm{d}\,Y_s = 0$. Thus, there is no transient distortion when the transient is between shades of grey, i.e., when $Y_s = 0$, or when it is between colours which always lie on a contour of constant Y_s throughout the transition, i.e., when $\mathrm{d}\,Y_s = 0$.

In all other cases some distortion will occur, and this will be a function of the initial and final colours and also the intermediate colours, of the originating transient.

In order to illustrate the mechanism of luminance transient distortion, consider three separate examples. In each case, assume that the colour transient being televised is accomplished linearly with time, and that the Y' bandwidth is sufficient to accommodate the Y transient.

Let us assume that the restricted bandwidth of the sub-carrier signals increases the duration of the transient threefold, and in addition let us assume that there is no differential bandwidth restriction of the chrominance signals.

EXAMPLE 1

Saturated full blue to a grey of $Y = 0\cdot04$. The relevant waveforms are shown in Fig. 16.6. Since the original transient is taken
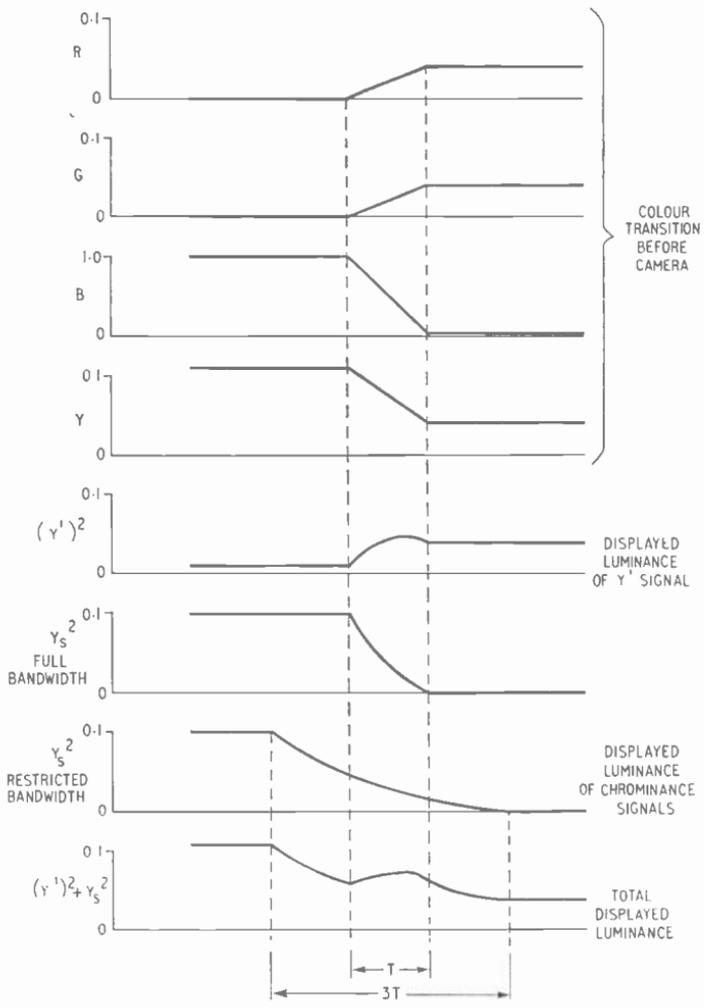
Fig. 16.6. Example 1. Blue to grey transition

as a linear function of time, the tristimulus values at any instant are given by

$$R_i = 0\cdot04\frac{t}{T}, \qquad G_i = 0\cdot04\frac{t}{T}$$

and

$$B_i = 0\cdot96\left(1 - \frac{t}{T}\right) + 0\cdot04$$

where t is measured from the start of the transient, which is completed after a time T. Hence the instantaneous luminance is

$$Y_i = 0\cdot30R_i + 0\cdot59G_i + 0\cdot11B_i$$

and similarly

$$Y'_i = 0\cdot30R_i^{\frac{1}{2}} + 0\cdot59G_i^{\frac{1}{2}} + 0\cdot11B_i^{\frac{1}{2}}$$

Then

$$Y_{si}{}^2 = Y_i - (Y'_i)^2$$

Notice that the luminance carried by Y' increases during the transient, while Y decreases. Also the initial and final thirds of the total displayed luminance period are entirely dependent on the $Y_s{}^2$ signal as far as frequency response is concerned.

EXAMPLE 2

Saturated full blue to a grey of $Y = 0\cdot11$. In this case, illustrated in Fig. 16.7, the instantaneous tristimulus values are

$$R_i = 0\cdot11\frac{t}{T}, \qquad G_i = 0\cdot11\frac{t}{T} \text{ and } B\hat{\imath} = 0\cdot89\left(1 - \frac{t}{T}\right) + 0\cdot11$$

Although there is no change in Y throughout the transient, there is a considerable disturbance in $(Y')^2$ and $Y_s{}^2$, and since the latter is bandwidth limited, these two disturbances are not complementary in shape so that an entirely spurious luminance transient is displayed.

EXAMPLE 3

Pure red of $R = 0\cdot403$ to a greenish blue $R = 0$, $G = 0\cdot018$ and $B = 1$. The instantaneous values are

$$R_i = 0\cdot403\left(1 - \frac{t}{T}\right), \qquad G_i = 0\cdot018\frac{t}{T} \text{ and } B_i = \frac{t}{T}$$

and the waveforms are shown in Fig. 16.8.
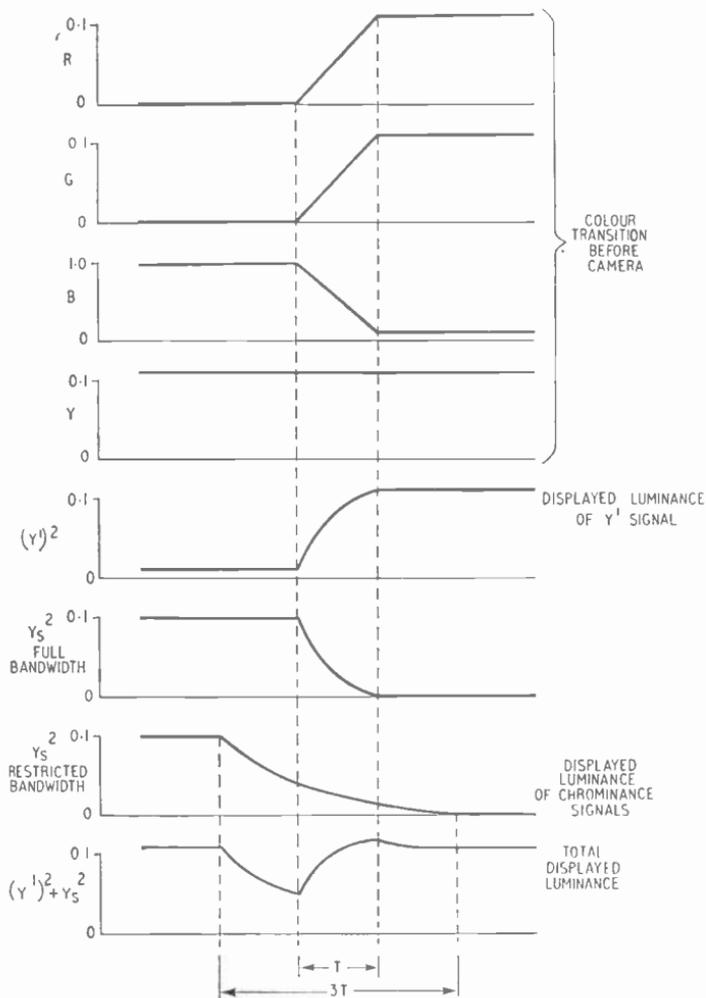
Fig. 16.7.　Example 2.　Blue to grey transition at constant luminance
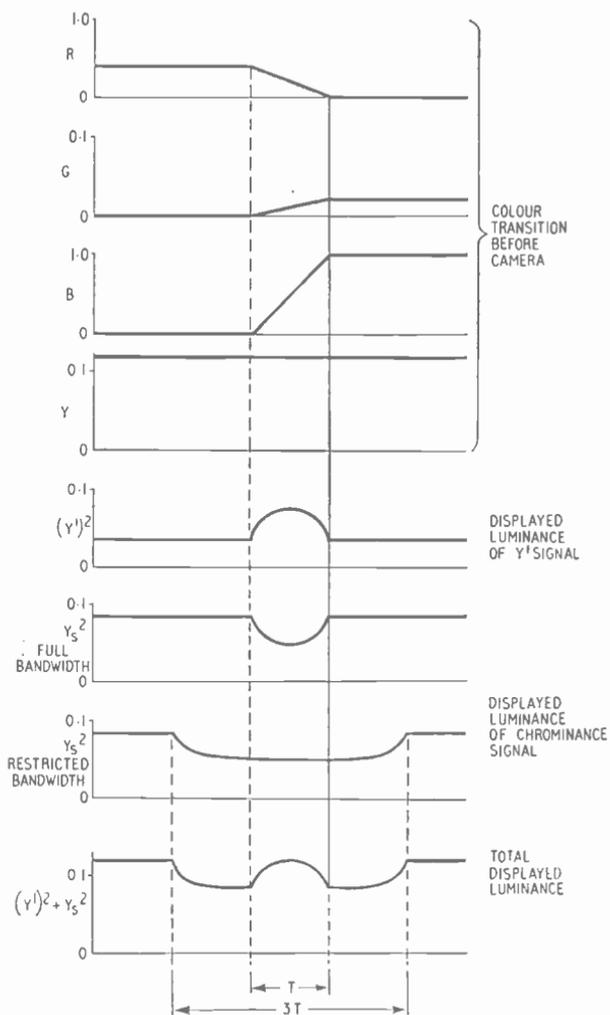
Fig. 16.8. Example 3. Transition between colours having equal Y, $(Y')^2$ *and* Y_s^2 *values*

In this case, in spite of the fact that Y, $(Y')^2$ and $Y_s{}^2$ each have the same initial and final values, and although Y is constant throughout the transient, nevertheless an entirely spurious luminance transient is displayed. Notice that a drop in luminance occurs on each side of the centre of the transient owing to the falling luminance contribution of the slow Y_s signal. Notice also that the displayed luminance is correct at the centre of the transient because it is here that the values of the restricted bandwidth and full bandwidth $Y_s{}^2$ signals are equal.

It should be pointed out that in all the above examples the sharp transitions of the displayed luminance will be rounded off as a

Fig. 16.9. Contours of constant $Y_s{}^2$ *in chrominance plane. Equal time intervals are indicated on the* I' *and* Q' *axes*
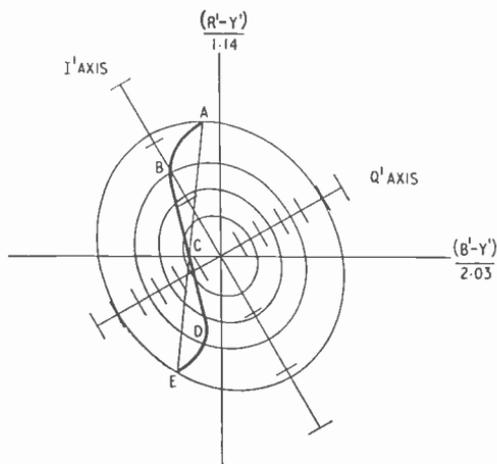
consequence of the finite luminance channel bandwidth, but this practical point has been omitted from the figures for the sake of simplicity.

It is interesting to examine the above type of transient in more general terms. In Section 16.2 it was shown that

$$Y_s{}^2 = 0\cdot30(R' - Y')^2 + 0\cdot59(G' - Y')^2 + 0\cdot11(B' - Y')^2$$

For constant values of $Y_s{}^2$, $\dfrac{(R' - Y')}{1\cdot14}$ may be plotted against $\dfrac{(B' - Y')}{2\cdot03}$ in the chrominance plane. The contours of constant $Y_s{}^2$ are ellipses, as shown in Fig. 16.9. (Note that these ellipses

differ from the loci of Fig. 16.1 in that the latter are plotted for constant values of $\dfrac{(Y')^2}{Y}$, and since $(Y')^2 = Y - Y_s{}^2$, it follows that the ellipses of Fig. 16.1 correspond to constant values of the ratio $Y_s{}^2$ to Y.)

The larger ellipses of Fig. 16.9 correspond to the larger values of $Y_s{}^2$, and it will be seen that a transient between two colours such as A and E, each of which has the same value of $Y_s{}^2$, will have a straight line path AE if the red and blue difference axes have equal bandwidths. For a wideband I', narrow band Q' receiver, the transient will follow the path $ABCDE$. In either case, the sub-carrier luminance contribution at the centre C of the transient is smaller than at A or E, so that a sub-carrier luminance transient occurs although the transient is between colours of equal sub-carrier luminance. This luminance transient is broadened by the bandwidth limitation of the chrominance signals so that a darkening occurs at the transient which is not cancelled by the corresponding opposite polarity luminance contribution of the higher definition Y' signal. The dip in luminance contribution of the sub-carrier during the transient is often called a *luminance notch*.

Notice that the transient passes near the white point so that desaturation occurs at the centre of the transient. If the transient could be made to follow the elliptical path between A and E, the luminance notch would be removed completely and the saturation would be higher. This corresponds to the case $dY_s = 0$.

16.9. The I' Q' fallacy

Another consequence of constant luminance failure which is of interest is concerned with the underlying principles of wideband I', narrow band Q' operation. In a conventional $I'Q'$ system, crosstalk between I' and Q' or vice versa does not occur over the double sideband frequency spectrum common to I' and Q'. Within the single sideband region of the high frequency I' components, Q' is absent because it has been bandwidth limited at the transmitter, and therefore Q' cannot crosstalk to I'. However I' can crosstalk to Q' over this region, but such crosstalk can be eliminated by including a Q' pass filter at the receiver.

Now the need for a filter to remove I' crosstalk from the Q' channel may be questioned; since high frequency components of Q' produce colours on the Q' axis, these colours should not be visible owing to the effective tritanopia of the central fovea of the human eye. Put another way, since the high frequency Q' components should not be visible, neither should the crosstalk of I' to Q' be

visible. The fact that a Q' filter is required in practice suggests that the spurious luminance components of I' to Q' crosstalk are subjectively more noticeable than the spurious colour components of crosstalk, in which case it follows that Q' bandwidth limitation at the transmitter removes luminance information which would be visible, although the corresponding chrominance components would not be visible.

It will be shown that the rejection of high frequency Q' components by the transmitter produces a " blind " luminance axis so that for colours lying on this axis the single sideband or high frequency components of I' do not affect luminance, and since the high frequency Q' components are absent, the axis is luminance sensitive only to the low frequency Q' components.

Referring to the equation for the displayed luminance in terms of Y' and the chrominance signals

$$Y = (Y')^2 + 0 \cdot 30(R' - Y')^2 + 0 \cdot 59(G' - Y')^2 + 0 \cdot 11(B' - Y')^2$$

if $\gamma = 2$

Substituting

$$(R' - Y') = 0 \cdot 96I' + 0 \cdot 63Q'$$

$$(G' - Y') = -0 \cdot 28I' - 0 \cdot 64Q'$$

$$(B' - Y') = -1 \cdot 11I' + 1 \cdot 72Q'$$

it follows that

$$Y = (Y')^2 + 0 \cdot 46(I')^2 + 0 \cdot 69(Q')^2 + 0 \cdot 15I'Q'$$

If Y' and Q' are held constant, the control of Y by I' is given by

$$\frac{\mathrm{d}Y}{\mathrm{d}I'} = 0 \cdot 92I' + 0 \cdot 15Q'$$

Similarly, if Y' and I' are held constant, the control of Y by Q' is given by

$$\frac{\mathrm{d}Y}{\mathrm{d}Q'} = 1 \cdot 38Q' + 0 \cdot 15I'$$

The axes $\frac{\mathrm{d}Y}{\mathrm{d}I'} = 0$ and $\frac{\mathrm{d}Y}{\mathrm{d}Q'} = 0$ are plotted on the chrominance plane diagram in Fig. 16.10. The equations of these axes are

$$I' = -\frac{0 \cdot 15}{0 \cdot 92}Q', \quad \text{and} \quad I' = -\frac{1 \cdot 38}{0 \cdot 15}Q' \text{ respectively.}$$
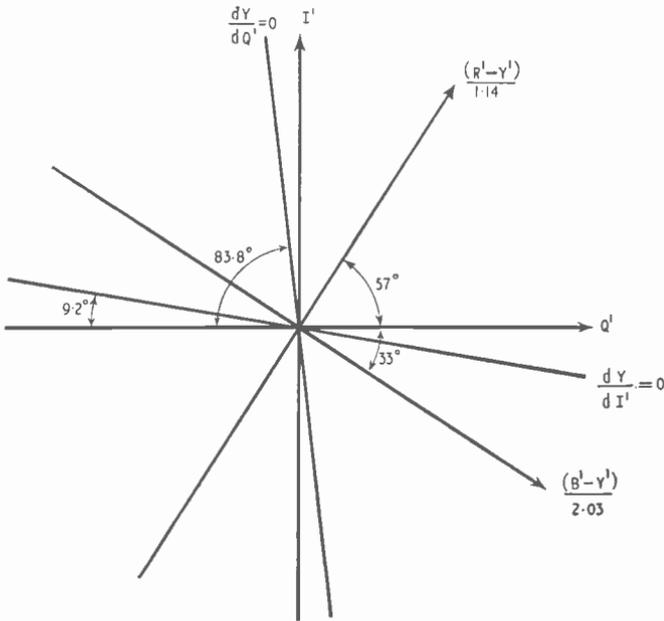
Fig. 16.10. Luminance control by I' *and* Q'

For colours lying on the axis $\dfrac{dY}{dI'} = 0$, the I' signal has no control over luminance at any frequency. Such control would normally come from the Q' signal and since Q' is bandwidth limited, there is no luminance control by any high frequency chrominance component along this axis. Thus, luminance transient response is limited for colours on the $\dfrac{dY}{dI'} = 0$ axis, and for colours near this axis the single sideband components of I' will have little effect on luminance.

16.10. Effects due to multiplexing luminance and chrominance signals

An essential feature of the N.T.S.C. system is the sharing of part of the video band by the luminance and by the sub-carrier and its sidebands. While a colour television system using out of band chrominance could be envisaged, this would necessitate an appreciably wider bandwidth than that required for the corresponding monochrome system and one of the great advantages of N.T.S.C. operation, i.e., the ability to add colour to an existing monochrome

system, would be lost. Additionally, the provision of increased R.F. and I.F. bandwidth in colour receivers, and the possibility of frequency dependent propagation effects, may present economic difficulties. However, multiplexing the luminance and chrominance signals produces certain spurious effects, as follows:—

16.10.1. CROSS-COLOUR

The luminance channel accepts the sub-carrier and its sidebands as a luminance signal, giving a dot structure interference which itself produces undesirable luminance effects. Similarly, the chrominance channel accepts those components of luminance which lie in the region of the sub-carrier, and interprets them as chrominance information. This effect is usually called *cross-colour*.

The N.T.S.C. system is designed so that the above spurious effects have reduced visibility in a subjective sense. Thus, the sub-carrier signal is zero for white, and is small for the statistically frequent pastel colours. Also, locking the sub-carrier frequency to an odd multiple of half the line scan frequency produces an integrating effect which reduces dot visibility. Both these properties of N.T.S.C. systems reduce the spurious effects of chrominance to luminance crosstalk. Again, spurious colour effects produced by luminance to chrominance crosstalk tend to integrate out because of the frequency relationship between the sub-carrier and scanning frequencies, so that a spurious colour on one picture is added to its complementary on the succeeding picture.

It is interesting to note that, for a monochrome transmission (i.e., one in which the colour burst is absent), the receiver reference frequency generator is necessarily unlocked so that the optimum conditions for reducing spurious colour effects do not apply. In other words, cross-colour effects are more serious when the burst is absent, and these spurii are emphasized subjectively because the picture is nominally monochrome. Furthermore, if the receiver is fitted with A.C.C., the chrominance gain will be set to maximum. The need for some colour killing mechanism which switches off the receiver chrominance channel during " burstless " reception is therefore obvious.

16.10.2 PARC NOISE

In the same way that luminance components in the frequency region of the sub-carrier are interpreted as chrominance information by the receiver, giving rise to cross-colour, so any noise components in the appropriate frequency band of the luminance signal are

similarly displayed as " coloured noise " or *parc noise* as it is some-
times called.   The noise in the luminance signal which produces
parc noise can originate either in the picture source at the trans-
mitter or in the transmission path to the receiver, or both.   Notice
that cross-colour and parc noise, while originating in the higher
frequency video band (somewhere near 2·7 Mc/s in the British
N.T.S.C. system) are nevertheless displayed as low frequency
chrominance information because of the heterodyning action of the
reference frequency generator.   This tends to increase their
visibility.

16.11.   Transmitter notch filter

Cross-colour and parc noise can be reduced by including a
rejection filter in the transmitter luminance channel which attenuates
luminance components in the region of the sub-carrier frequency.
The exact characteristics of such a filter are best determined by
experiment, as too much attenuation will spoil luminance defini-
tion while too little attenuation will be ineffective.   An attenua-
tion of 6dB at sub-carrier frequency, with a 3dB bandwidth of
± 200 kc/s, is likely to be near optimum.   This type of luminance
notch filter must not be confused with the luminance notch filter
in a colour receiver.   The latter, it will be remembered, is included
to remove the large area dot structure of chrominance to luminance
crosstalk.

16.12.   Effects of gamma characteristic on cross-colour

Although cross-colour visibility is reduced by locking the sub-
carrier frequency to an odd multiple of half the line scan frequency,
in point of fact the display tube has an approximately square law
relation between light output and voltage input, and this produces
relatively high visibility terms which would be absent for a linear
display.   The generation of these higher visibility components is
probably best illustrated by an example.

Suppose a " monochrome " sine wave of amplitude M, and
oscillating about a mean luminance level M, is transmitted.   Then
the luminance signal is

$$Y' = M(1 + \sin \omega_1 t)$$

where $\dfrac{\omega_1}{2\pi}$ is the frequency.   Suppose that a burst is transmitted,
so that the receiver chrominance channel is operative, and the
receiver reference frequency generator is correctly locked.   Let us

further suppose that no chrominance information is transmitted. If $\frac{\omega_2}{2\pi}$ is the sub-carrier frequency and if $\frac{\omega_1}{2\pi}$ falls in the acceptance band of the chrominance channel, the colour receiver will derive the chrominance difference signals as follows:—

$$(R' - Y') = 1 \cdot 14M \sin(\omega_1 - \omega_2)t$$

$$(B' - Y') = 2 \cdot 03M \cos(\omega_1 - \omega_2)t$$

$$(G' - Y') = -M[0 \cdot 58 \sin(\omega_1 - \omega_2)t + 0 \cdot 38 \cos(\omega_1 - \omega_2)t]$$

The $(R' - Y')$ output is obtained from the $(R' - Y')$ synchronous detector by heterodyning the incoming $M \sin \omega_1 t$ signal with a reference signal $A \sin\left(\omega_2 t + \frac{\pi}{2}\right)$, while the $(B' - Y')$ output results from heterodyning $M \sin \omega_1 t$ with a reference $A \sin \omega_2 t$. The higher frequency terms are eliminated by the receiver chrominance filters, and the gains $1 \cdot 14$ and $2 \cdot 03$ are applied by the receiver to compensate for the reduced chrominance signals in a colour transmission.

The luminance signal is

$$Y' = M(1 + \sin \omega_1 t)$$

but if a luminance notch filter is included in the receiver (as it must be), then if this has an attenuation of $\frac{1}{m}$ to 1 at the frequency $\frac{\omega_1}{2\pi}$, the luminance signal actually reaching the display will be $M(1 + m \sin \omega_1 t)$

For a square law display, the displayed colour will be

$$R = [M(1 + m \sin \omega_1 t) + 1 \cdot 14M \sin(\omega_1 - \omega_2)t]^2$$

$$G = [M(1 + m \sin \omega_1 t) - M(0 \cdot 58 \sin(\omega_1 - \omega_2)t + 0 \cdot 38 \cos(\omega_1 - \omega_2)t)]^2$$

$$B = [M(1 + m \sin \omega_1 t) + 2 \cdot 03M \cos(\omega_1 - \omega_2)t]^2$$

In these expressions, in addition to the terms $M^2(1 + m \sin \omega_1 t)^2$ which would be obtained if the chrominance channel were inoperative, terms having pulsatances of $(\omega_1 - \omega_2)$, $2(\omega_1 - \omega_2)$, $(2\omega_1 - \omega_2)$ and ω_2 are obtained.

If we assume that ω_1 is a multiple N of the line scan frequency L, so that the signal $M(1 + m \sin \omega_1 t)$ appears as vertical stripes,

and if we take $\omega_2 = (2P + 1)L/2$, where P is an integer, in accordance with the N.T.S.C. specification, then

$$\omega_1 - \omega_2 = NL - (2P + 1)\frac{L}{2} = (2N - 2P - 1)\frac{L}{2} = \text{Odd} \times \frac{L}{2}$$

$$2(\omega_1 - \omega_2) = (2N - 2P - 1)L = \text{Odd} \times L$$

$$2\omega_1 - \omega_2 = 2NL - (2P + 1)\frac{L}{2} = (4N - 2P - 1)\frac{L}{2} = \text{Odd} \times \frac{L}{2}$$

Hence terms involving $(\omega_1 - \omega_2)$, $(2\omega_1 - \omega_2)$ and ω_2 have low visibility, while those involving $2(\omega_1 - \omega_2)$ have relatively high visibility.

(Note that if the display were linear, the chrominance terms would all be functions of $(\omega_1 - \omega_2)$, which has low visibility.)

If we assume that the low visibility terms are, in fact, not visible, then the displayed output will be

$$R = M^2[(1 + m \sin \omega_1 t)^2 + 0 \cdot 65(1 - \cos 2(\omega_1 - \omega_2)t)]$$

$$G = M^2[(1 + m \sin \omega_1 t)^2 + 0 \cdot 24 - 0 \cdot 10 \cos 2(\omega_1 - \omega_2)t \\ + 0 \cdot 22 \sin 2(\omega_1 - \omega_2)t]$$

$$B = M^2[(1 + m \sin \omega_1 t)^2 + 2 \cdot 06(1 + \cos 2(\omega_1 - \omega_2)t)]$$

Notice that the factor m reduces some of the low visibility terms, but does not affect the high visibility $2(\omega_1 - \omega_2)$ terms.

Since

$$(1 + m \sin \omega_1 t)^2 = 1 + 2m \sin \omega_1 t + \frac{m^2}{2}(1 - \cos 2\omega_1 t)$$

and since ω_1 is well above I cut off, we may write $1 + \dfrac{m^2}{2}$ for $(1 + m \sin \omega_1 t)^2$ if we are considering chromaticity rather than luminance.

The chromaticity co-ordinates of the displayed colour are therefore

$$r = \frac{1}{D}\left[1 \cdot 65 + \frac{m^2}{2} - 0 \cdot 65 \cos 2(\omega_1 - \omega_2)t \right]$$

$$g = \frac{1}{D}\left[1 \cdot 24 + \frac{m^2}{2} - 0 \cdot 10 \cos 2(\omega_1 - \omega_2)t + 0 \cdot 22 \sin 2(\omega_1 - \omega_2)t \right]$$

28

$$b = \frac{1}{D}\left[3 \cdot 06 + \frac{m^2}{2} + 2 \cdot 06 \cos 2(\omega_1 - \omega_2)t\right]$$

where

$$D = 5 \cdot 95 + \frac{3}{2}m^2 + 0 \cdot 22 \sin 2(\omega_1 - \omega_2)t + 1 \cdot 31 \cos 2(\omega_1 - \omega_2)t$$

It is interesting to evaluate the chromaticity co-ordinates over a complete cycle of $2(\omega_1 - \omega_2)t$ for the extreme case of $m = 0$. The result is shown in Fig. 16.11.

Since $m = 0$, the transmitted A.C. component of luminance, $m \sin \omega_1 t$, is not reproduced, and apart from the D.C. component
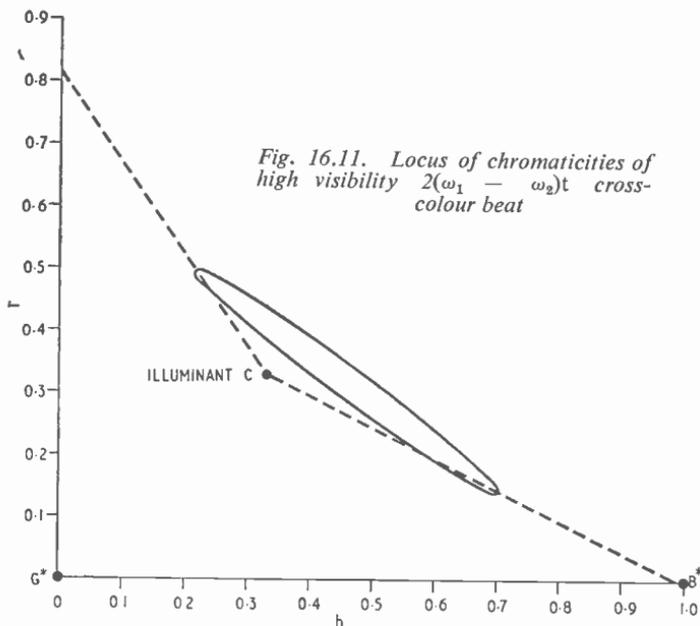
Fig. 16.11. *Locus of chromaticities of high visibility* $2(\omega_1 - \omega_2)t$ *cross-colour beat*

M^2, the display is entirely composed of spurious terms. Thus, vertical coloured bars having a frequency of $\dfrac{(\omega_1 - \omega_2)}{\pi}$ are obtained, and the chromaticities vary between an orange of $r = 0 \cdot 81$ and $g = 0 \cdot 19$ with a purity of 34%, and a near blue of $b = 0 \cdot 98$ and $g = 0 \cdot 02$ with a purity of 57%, as shown in Fig. 16.11.

It will be appreciated that if a genuine chrominance signal is present in addition to luminance components of the type $M \sin \omega_1 t$,

then further spurious terms will be generated.   However, the sub-jective effect of such spurii is likely to be reduced by the presence of the genuine colour information in the picture.

16.13.   Spurious colour in white areas

If a white area is present in an otherwise coloured picture, any noise accepted by the chrominance channel will produce coloured flecks in the white area.   For reasonable signal conditions, e.g. greater than $100\mu V$ in Band I, such " coloured " noise is sub-jectively acceptable, but for fringe area reception it may reach

Fig. 16.12.   Sideband and carrier vectors for double and single sideband
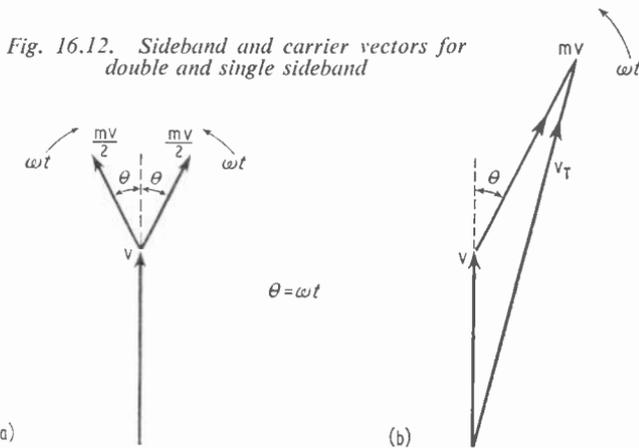
annoyance level.   Hence, a subjective improvement might be obtained in fringe areas if a device is incorporated in the chromin-ance channel which rejects all chrominance information which is less than a certain pre-set level.   Then white, and all pastel or low luminance colours, would be reproduced as a white chromaticity without " coloured " noise.

16.14.   Single sideband distortion

If an amplitude modulated R.F. signal is applied to a conven-tional amplitude sensitive detector, the detected output waveform differs from the original modulating signal if one sideband is omitted.   Since all television receivers now operate single sideband over a substantial part of the higher frequency video spectrum, the video waveform appearing at the conventional second detector of a receiver is distorted.   This distortion increases with modulation

depth, and is often called *single sideband* (or sometimes *quadrature*) *distortion*. It is of some importance in N.T.S.C. colour transmissions, since the sub-carrier and its sidebands fall entirely within the single sideband region.

In negative modulation systems the effective modulation depth is greater for the black to white range than it is for positive modulation systems and single sideband distortion is therefore more serious in the former case.

The mechanism of the distortion may be illustrated by considering the simple case of a sine wave modulated carrier which is applied to a conventional detector after the removal of one sideband. Fig. 16.12(a) shows the double sideband case with two equal vectors rotating in opposite directions at the modulating frequency, the carrier amplitude being V and the modulation depth index m, i.e. $100m\%$ modulation. The detector measures the magnitude of the vector sum of the three vectors, which can be seen to lie always in phase with the carrier.

In the single sideband case of Fig. 16.12(b), if the sideband is boosted 6dB relative to the carrier, the vector sum is equal to the double sideband case only at the two instants when the sideband is in phase or in anti-phase with the carrier. At all other instants, the detector output, which is proportional to the magnitude of the vector V_T, will not be the same as the double sideband case, and distortion will result.

Notice that in the single sideband case V_T has a component in quadrature with the carrier (hence the term quadrature distortion) which is absent in the double sideband case because here the quadrature components of the two sidebands are equal and opposite, and therefore cancel.

The magnitude of V_T is given by

$$V_T{}^2 = V^2 + m^2V^2 + 2mV^2 \cos \theta$$

by the trigonometric cosine rule.

Hence

$$V_T = V(1 + m^2)^{\frac{1}{2}}\left(1 + \frac{2m}{1 + m^2} \cos \theta\right)^{\frac{1}{2}}$$

If $m \ll 1$, then by the Binomial expansion

$$V_T \simeq V(1 + m^2)^{\frac{1}{2}}\left(1 + \frac{m}{1 + m^2} \cos \theta\right) \simeq V(1 + m \cos \theta)$$

which is the same as the double sideband case in which

$$V_T = V + \frac{mV}{2}\cos\theta + \frac{mV}{2}\cos\theta$$

and is undistorted.

But suppose the modulation is 100%, i.e. $m = 1$, then

$$V_T = V2^{\frac{1}{2}}\left(1 + \frac{2}{2}\cos\theta\right)^{\frac{1}{2}} = V2^{\frac{1}{2}}\left(2\cos^2\frac{\theta}{2}\right)^{\frac{1}{2}}$$

since

$$1 + \cos\theta = 2\cos^2\frac{\theta}{2}$$

Hence

$$V_T = 2V\cos\frac{\theta}{2}$$

(in which it must be remembered that V_T is a magnitude so that the magnitude of $2V\cos\frac{\theta}{2}$ must be taken for any value of θ) while the undistorted value is $V_T = V + V\cos\theta$. These two cases are shown in Fig. 16.13(a) and Fig. 16.13(b), respectively.

The original sine wave modulating signal is distorted to the same shape as that of a full wave rectified sine wave. It has a D.C. component of $\frac{4V}{\pi}$ instead of the correct value V and a fundamental component amplitude of $\frac{8V}{3\pi}$ instead of the correct value V.

If one considers the case of a video waveform in which a sub-carrier sine wave is oscillating about a luminance level, the sine wave mentioned above may be taken as an example of the sub-carrier sine wave. Hence, for a 100% modulation case, the sub-carrier amplitude is reduced since only the fundamental of the sub-carrier can be accommodated in the video frequency range and in addition a spurious D.C. component appears which affects the luminance level. In positive modulation systems the D.C. component increases the luminance level, while in negative modulation systems the luminance level is decreased by the spurious D.C. component, since in this case an increase in D.C. level at the detector causes a decrease in displayed luminance.

It has been shown that the distortion is small for $m \ll 1$ and a maximum for $m = 1$. For intermediate values of m the distortion itself will have an intermediate value which can be calculated in

terms of elliptic integrals, and these are convenient as they are extensively tabulated.

The distortion of a sine wave modulating signal (such as the sub-carrier signal) can be investigated by making a Fourier analysis of the waveform. The waveform has a shape which is somewhere between Fig. 16.13(a) and Fig. 16.13(b), depending on the value
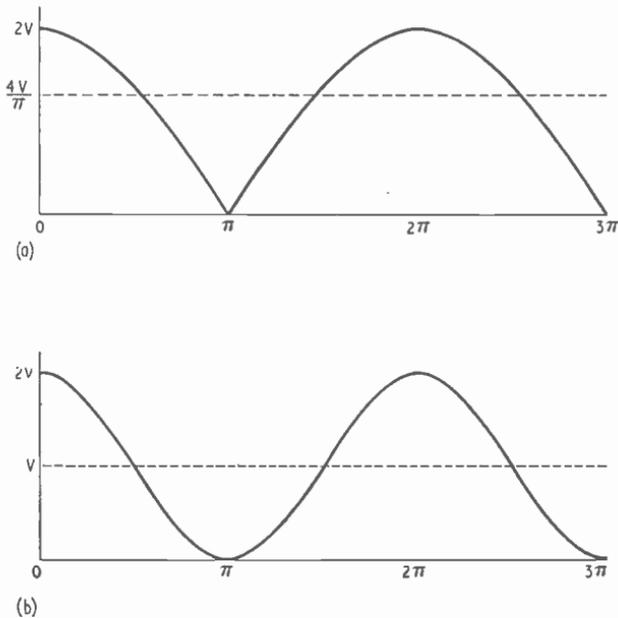
Fig. 16.13. Detector output for 100% modulation for (a) single sideband and (b) double sideband

of *m*. It is symmetrical about $\theta = \pi$ and can therefore be expressed as a cosine series, and it is fortunate that this symmetry applies for all values of *m*, otherwise a phase shift (and therefore a hue shift) would be produced which would vary with modulation depth.

The detector output V_T is given by

$$V_T = V(1 + m^2)^{\frac{1}{2}}\left(1 + \frac{2m}{1 + m^2}\cos\theta\right)^{\frac{1}{2}}$$

which may be expressed as a Fourier series

$$V_T = a_0 + a_1 \cos\theta + a_2 \cos 2\theta + \ldots.$$

If we are considering the case of distortion of the sub-carrier, then

$$\theta = 2\pi f_s t$$

where f_s is the sub-carrier frequency, and we are concerned only with the D.C. term a_0 and the fundamental amplitude a_1, since the higher harmonics fall outside the video band and will be rejected by the video output stage.

Then

$$a_0 = \frac{2}{2\pi} \int_0^\pi V_T \, d\theta \ \text{ and } \ a_1 = \frac{2}{\pi} \int_0^\pi V_T \cos\theta \, d\theta$$

Substituting for V_T in terms of θ, it can be shown that

$$a_0 = \frac{2V}{\pi}(1+m) \int_0^{\pi/2} \left[1 - \left(\frac{2\sqrt{m}}{1+m}\right)^2 \sin^2\phi \right]^{\frac{1}{2}} d\phi = \frac{2V}{\pi}(1+m)E$$

and

$$a_1 = \frac{2V}{3\pi m}(1+m)(1+m^2)\left[\int_0^{\pi/2}\left[1 - \left(\frac{2\sqrt{m}}{1+m}\right)^2 \sin^2\phi \right]^{\frac{1}{2}} d\phi \right.$$

$$\left. - \frac{(1-m)^2}{1+m^2} \int_0^{\pi/2}\left[1 - \left(\frac{2\sqrt{m}}{1+m}\right)^2 \sin^2\phi \right]^{-\frac{1}{2}} d\phi \right]$$

$$= \frac{2V}{3\pi m}(1+m)(1+m^2)\left[E - \frac{(1-m)^2}{1+m^2}K \right]$$

where

$$K = \int_0^{\pi/2}\left[1 - \left(\frac{2\sqrt{m}}{1+m}\right)^2 \sin^2\phi \right]^{-\frac{1}{2}} d\phi$$

and

$$E = \int_0^{\pi/2}\left[1 - \left(\frac{2\sqrt{m}}{1+m}\right)^2 \sin^2\phi \right]^{\frac{1}{2}} d\phi$$

are complete elliptic integrals of the first and second kind respectively, and may be looked up in tables for given values of $\left(\dfrac{2\sqrt{m}}{1+m}\right)^2$

For any given colour, the value of m can be calculated by taking the ratio of the sub-carrier amplitude to the mean carrier level

$$\text{MODULATION DEPTH INDEX} = m = \frac{A}{B}$$

(a)   (b)
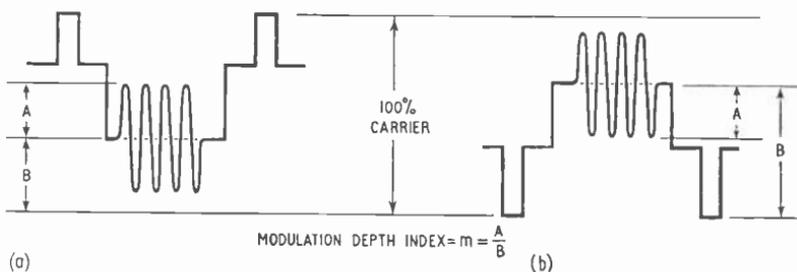
Fig. 16.14.   Modulation depths for negative and positive modulation systems

about which the sub-carrier is oscillating.   Fig. 16.14(a) shows the
negative modulation case, and Fig. 16.14(b) the positive modulation
case.   Notice that the greatest modulation depth is always smaller
in the positive modulation case because the synchronizing pulses
lift the level of carrier about which the sub-carrier signal oscillates.

By using the formulae quoted above, it is possible to calculate
the relative signal voltages applied to the display tube when a
colour bar signal is transmitted.   For the case of the transmission
of the three full primaries and their full complementaries, Fig.
16.15 shows the signals at the red, green and blue displays for the
positive and negative modulation systems.   The undistorted signals
are either 0 or 1 for each colour.   In the negative modulation case
it was assumed that peak white was at $12 \cdot 5\%$ carrier, and that cyan
and yellow were clipped at 0% carrier.   Thus for cyan and yellow
the distortion introduced by clipping has been included in addition
to the single sideband distortion.

It can be seen that, in the negative modulation case, single side-
band distortion tends to reduce luminance and therefore increase
saturation.   However a further effect is produced by the clipping
of the sub-carrier for yellow and cyan which tends to reduce satura-
tion, and in the case of yellow this reduction in saturation pre-
dominates over the increase in saturation caused by single sideband
distortion.   For cyan, the single sideband distortion effect is the
greater so that the saturation is preserved in this case.   If fully
saturated colours are transmitted, as in Fig. 16.15, the saturation
is unaffected by single sideband distortion since the saturation
cannot be further increased, i.e., the negative going signals applied
to the display are no more effective than zero signals.   For non-
fully saturated colours, however, the negative going signals will
tend to increase the saturation of the displayed colours.

In the positive modulation case, no clipping occurs provided
the transmitter can accept satisfactorily the additional drive require-
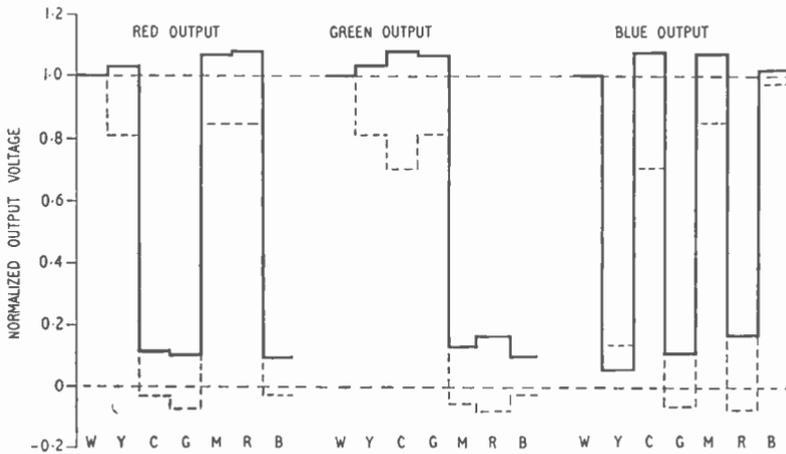
Fig. 16.15. Voltage output single sideband distortion for colour bars. Con. tinuous lines show positive modulation and dotted lines negative modulation. Undistorted levels are either 0 or 1

ment of the sub-carrier peaks. Single sideband distortion increases the luminance and reduces the saturation, and is greatest for red since this gives the largest m value.   Notice that the greatest single sideband distortion effect in the negative modulation case occurs for cyan, the complementary of red.

Some idea of the subjective effects of single sideband distortion can be obtained from Fig. 16.16, which shows the relative light outputs in the case of negative and positive modulation systems, assuming a square law relation between light output and voltage input for the display device.   That is, Fig. 16.16 is the " square " of Fig. 16.15.   The squaring process further reduces the luminance in the negative modulation case, and increases the saturation in the positive modulation case, and it would appear that the subjective effects of single sideband distortion are reduced when positive modulation is used.   However, it must be pointed out that a full luminance colour bar signal is the most severe test of single sideband distortion effects, although bright saturated yellows are quite likely to occur in practice.

16.15.   Effects of method of modulation of vision carrier and sound carrier

The techniques employed in the N.T.S.C. colour system, while originally designed for the American 525-line standard which uses negative modulation and F.M. sound, have been readily adapted

to positive modulation, A.M. sound systems, such as the present British 405-line system. It is instructive to consider some of the more important differences between these two methods of vision and sound modulation. It will be appreciated that differences in line number or picture frequency are relatively minor from a theoretical viewpoint, though there may be practical economic considerations in the provision of wide video bandwidths in receivers. Apart from this detail, the 525-line and 625-line systems may be regarded as substantially similar, but this similarity does not include the 405-line British system because of its different vision and sound modulation processes.

Before comparing the colour performance of negative and positive modulation systems, it may be advisable to point out the most important of the very significant differences between the corresponding monochrome systems, since these are often not generally appreciated. These differences may be listed as follows, in which negative modulation, F.M. sound systems and positive modulation, A.M. sound systems are designated N and P, respectively:—

(a) Line and field synchronization are more susceptible to noise or interference in N systems. Thus, flywheel line time bases are required in $'N$ system receivers for B and I signal
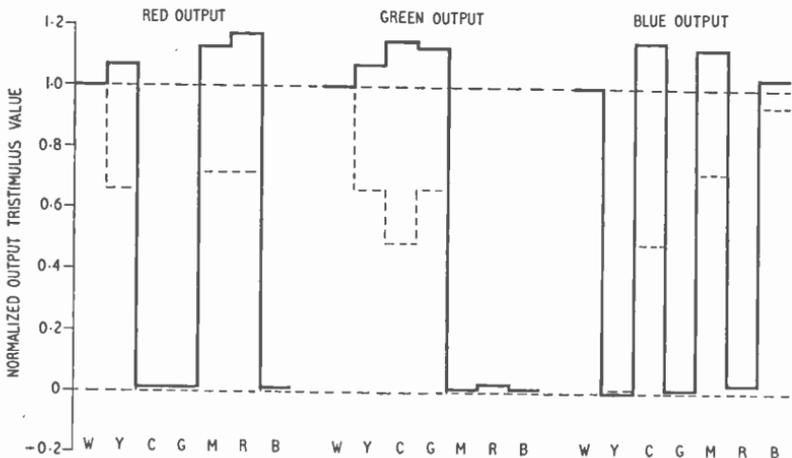
Fig. 16.16. *Light output single sideband distortion for colour bars. Continuous lines show positive modulation and dotted lines negative modulation. Undistorted levels are either 0 or 1*

levels below about $1mV$, whereas for P system receivers the corresponding signal level is about $200\mu V$.

Again, field synchronization in N systems can be upset by interference so that a field " bounce " or even collapse of the field scan is produced. This effect is almost unknown in P systems.

(b) N systems give black display of ignition interference, P systems give white. Note that grey is probably the optimum.

(c) A.G.C. is apparently easier with N systems, since each line sync pulse is a sample of 100% carrier. However, gating is normally essential to prevent A.G.C. action by interference, and A.G.C. in N systems is then easier only on the grounds of greater inherent gain.

(d) N systems tend to flatten picture whites as modulation drives the carrier towards zero. P systems can have a large peak white excursion for short periods while the black level remains constant, but on N systems this can be done only by changing the black level correspondingly.

(e) Single sideband distortion affects mainly picture content in N systems, and mainly synchronizing waveform in P systems, but the synchronizing properties are not impaired.

(f) D.C. maintenance is more difficult in N systems since absence of the carrier drives the display above peak white if simple D.C. coupling is used.

(g) Inter-carrier sound can be used in N systems.

Before discussing this point it may be as well to describe the inter-carrier sound technique. In inter-carrier sound operation, receiver sound rejection is normally not greater than about 20dB so that an appreciable amplitude of the beat between sound and vision carriers appears at the vision detector. Since the sound is frequency modulated, the audio information does not produce L.F. interference on the picture, and the sound carrier-vision carrier beat which does appear is a high video frequency outside the normal video band, and therefore not normally objectionable. In any case, this beat can be removed from the video by means of a suitable tuned rejector, if desired. The beat between sound and vision carriers can be passed through a limiter and discriminator to yield the audio information.

The advantages normally claimed for inter-carrier sound are economic ones, in the sense that vision and sound I.F.'s are amplified together, and that performance is relatively independent of local

oscillator tuning. However, it must be remembered that economically the cost of a limiter and discriminator in N systems must be weighed against the cost of the I.F. sound amplifier and an A.M. detector in P systems.

As far as oscillator drift is concerned, the inter-carrier technique does have some advantage, but this is not as great as one might suppose. The reader may believe that since the sound-vision carrier beat is not subjectively worrying on the picture, and since the beat amplitude will increase rather than decrease for reasonable detunings, then local oscillator stability is unimportant. But while it is true that the audio output will be relatively independent of tuning, it must be remembered that picture quality will not be, for L.F. or H.F. video boost will result depending on the direction of the tuning error. Again, if the detuning is sufficient to bring the sound carrier to the edge of the sound rejection characteristic, demodulation will result and produce audio breakthrough on the picture. Strictly speaking, therefore, inter-carrier operation allows a larger oscillator drift to the extent that the bandwidth over which a 20dB rejection can be maintained is greater than that over which a 40dB rejection can be maintained, 40dB being the approximate figure for P systems. Also, a stable discriminator is easier to design for frequencies in the region of the sound-vision carrier beat (about 5 Mc/s) rather than for sound I.F. frequencies (about 38 Mc/s).

Turning now to the colour performance of N and P systems, it seems that the former suffers from these disadvantages:

(a) Full saturated yellow and cyan cannot be transmitted without clipping the sub-carrier, unless the sub-carrier amplitude is reduced relative to the vision carrier.

(b) Single sideband distortion is more significant (*see* Section 16.14).

(c) D.C. maintenance is essential in colour receivers to prevent colour distortion, and this is more difficult in N system receivers (*see* Section 16.15(f) (Monochrome effects)).

(d) Accurate burst gating is more difficult in N systems because of synchronizing difficulties.

(e) Unlike the sound carrier-vision carrier beat, the beat between sound carrier and sub-carrier is a relatively low frequency and therefore has high visibility. Considerable sound rejection (about 60dB) is required in N system receivers to remove this. In P system receivers, the monochrome figure of 40dB rejection is adequate, since the sound carrier-sub-carrier beat can be locked to an odd multiple of half the line

scan frequency, and this gives the subjective effect of approxi-mately 20dB rejection compared with the unlocked case. Although it is possible in principle to lock the unmodulated sound-vision carrier beat in an F.M. sound system, in practice this is not very successful.

Note that the greater sound rejection required in N systems reduces the permissible oscillator drift compared with P systems, even if inter-carrier sound is used (*see* Section 16.15(g) (Mono-chrome effects)).

It would appear that while some schools of thought may regard the difference between N and P monochrome systems as being marginal, or even in favour of N systems, there is overwhelming evidence in favour of positive modulation, A.M. sound systems for N.T.S.C. colour use.

16.16. Summary

In this chapter we have seen that individual gamma correction of the red, green and blue video signals before encoding at the transmitter, leads to constant luminance failure and hence a degradation of noise performance. Gamma correction also entails a non-linear relation between the transmission parameters and the chromaticity of the colour being televised, so that the effects of noise on the reproduced chromaticity are greater for pastel colours than for saturated colours, and greater for complementary than for primary hues. While there is no large area colour distortion, con-stant luminance failure coupled with the bandwidth limitation of the chrominance signals (and hence of the luminance carried by the latter) causes distortion of transitions between colours and in some cases produces luminance transients which are entirely spurious.

Bandsharing or multiplexing of the luminance and sub-carrier signals leads to crosstalk between them which, when applied to a non-linear display, produces spurious effects.

Single sideband distortion at the vision detector reduces the sub-carrier amplitude and generates a spurious D.C. component which tends to increase saturation in negative modulation systems and decrease saturation in positive modulation systems. Because the effective modulation depth is greater in the former case, negative modulation systems suffer the greater distortion.

The sign of vision modulation and the type of sound modulation used in an N.T.S.C. system have an appreciable effect on the per-formance of colour receivers, and there are clear indications that positive modulation and A.M. sound are to be preferred.

Analysis of burst waveform

A.1.1. Calculation of Fourier coefficients

The burst waveform for the shortest burst of 8 cycles is shown in Fig. A.1.   It has been drawn over a period of 2 lines, during which the sub-carrier completes 525 cycles in the British system.   The fundamental frequency is half the line frequency since the waveform is repetitive after two line periods.

The waveform has been drawn symmetrically about π so that only cosine terms will be present in the Fourier series.

The equation of the burst waveform during the burst is $S \cos Nx$,

where $N = 525$, and $x = \dfrac{\omega_L t}{2}$

where $\omega_L = 2\pi$ times the line frequency.

If a_n is the amplitude of the nth harmonic, then

$$\pi a_n = \int_0^\alpha S \cos Nx \cos nx \, dx + \int_{\pi-\alpha}^{\pi+\alpha} S \cos Nx \cos nx \, dx$$
$$+ \int_{2\pi-\alpha}^{2\pi} S \cos Nx \cos nx \, dx$$

Since

$$\cos Nx \cos nx = \frac{1}{2}[\cos(N + n)x + \cos(N - n)x]$$

then

$$\frac{2\pi a_n}{S} = \left[\frac{\sin(N + n)x}{N + n} + \frac{\sin(N - n)x}{N - n}\right]_0^\alpha$$
$$+ \left[\frac{\sin(N + n)x}{N + n} + \frac{\sin(N - n)x}{N - n}\right]_{\pi-\alpha}^{\pi+\alpha}$$
$$+ \left[\frac{\sin(N + n)x}{N + n} + \frac{\sin(N - n)x}{N - n}\right]_{2\pi-\alpha}^{2\pi}$$
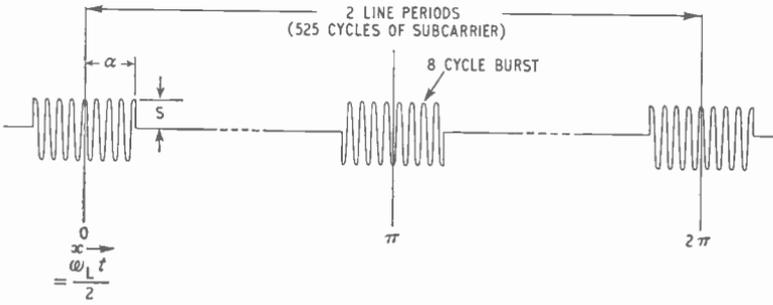
Fig. A.1.   8 cycle burst waveform over two line periods

Hence

$$\frac{2\pi a_n}{S} = 0 \qquad \text{for } n \text{ even}$$

and

$$\frac{2\pi a_n}{S} = \frac{4\sin(N+n)\alpha}{N+n} + \frac{4\sin(N-n)\alpha}{N-n} \qquad \text{for } n \text{ odd}$$

The harmonics of interest are those close to the sub-carrier frequency, i.e. the values of n close to N.

Hence, approximately

$$\frac{\pi a_n}{S} = \frac{2\sin(N-n)\alpha}{N-n}$$

Putting $m = (N-n)$, (m is always even)

$$\frac{\pi a_n}{2S} = \frac{\sin m\alpha}{m}$$

where $m = 0$ gives the sub-carrier frequency component, $m = 2$ gives the nearest harmonic lower than the sub-carrier frequency, and $m = -2$ gives the nearest harmonic higher than the sub-carrier frequency.

Note that the sub-carrier component has an amplitude a_N which is given by

$$\frac{\pi a_N}{2S} = \underset{m\to 0}{\text{Limit}} \ \frac{\sin m\alpha}{m} = \alpha$$

where α is the fraction of the twice line cycle occupied by half the burst, i.e.

$$\alpha = \frac{4}{525} \times 2\pi = \frac{8\pi}{525}$$

Hence

$$a_N = \frac{2S\alpha}{\pi} = \frac{16S}{525} = 0{\cdot}0305S$$

The burst waveform may therefore be represented by the series

$0{\cdot}0305S \cos \omega_s t$

$$+ \frac{2S}{\pi}\left[\frac{\sin 2\alpha}{2}\cos(\omega_s - \omega_L)t + \frac{\sin 4\alpha}{4}\cos(\omega_s - 2\omega_L)t + \dots \right]$$

$$+ \frac{2S}{\pi}\left[\frac{\sin 2\alpha}{2}\cos(\omega_s + \omega_L)t + \frac{\sin 4\alpha}{4}\cos(\omega_s + 2\omega_L)t + \dots \right]$$

where $\alpha = \frac{8\pi}{525}$, $\omega_s = 2\pi \times$ sub-carrier frequency

and $\omega_L = 2\pi \times$ line frequency.

Note that the amplitude of the nearest harmonic to the sub-carrier component, and on each side of it, is $\frac{\sin 2\alpha}{2\alpha}$ of the amplitude of the sub-carrier component. This ratio is very nearly unity, since

$$\frac{\sin 2\alpha}{2\alpha} \simeq 1 - \frac{2}{3}\alpha^2 = 1 - 0{\cdot}00153$$

Care must be exercised in reference generator designs to avoid locking to these adjacent sidebands. This is called *side-lock*.

The above series clearly illustrates the frequency interleaving principle of the N.T.S.C. system. Thus, the fundamental component of the burst waveform has a frequency of $\frac{\omega_L}{4\pi}$ (the 262nd harmonic counting downwards from the sub-carrier component) which is half the line frequency, and all the harmonics are spaced from this at intervals of the line frequency. These harmonics therefore fall half way between the luminance harmonics, whose fundamental frequency is equal to the line frequency (neglecting field frequency components) and whose harmonic spacing is again equal to the line frequency.

A.1.2. Burst amplifier bandwidth

The mean square value of the waveform is given by the series

$$\frac{1}{2}(0{\cdot}0305S)^2 + \frac{1}{2}\cdot\frac{4S^2}{\pi^2}\left[\frac{\sin^2 2\alpha}{2^2} + \frac{\sin^2 4\alpha}{4^2} + \frac{\sin^2 6\alpha}{6^2} + \cdots\right]$$

$$+ \frac{1}{2}\cdot\frac{4S^2}{\pi^2}\left[\frac{\sin^2 2\alpha}{2^2} + \frac{\sin^2 4\alpha}{4^2} + \frac{\sin^2 6\alpha}{6^2} + \cdots\right]$$

$$= \frac{1}{2}(0{\cdot}0305S)^2 + \frac{4S^2}{\pi^2}\left[\frac{\sin^2 2\alpha}{2^2} + \frac{\sin^2 4\alpha}{4^2} + \frac{\sin^2 6\alpha}{6^2} + \cdots\right]$$

for any number of harmonics.

The total mean square value can be calculated as follows:

The area under the square of the burst waveform for values of x from $x = 0$ to $x = \alpha$ is given by

$$\int_0^\alpha S^2 \cos^2 Nx \, dx = \frac{S^2}{2}\int_0^\alpha (1 + \cos 2Nx)\, dx = \frac{S^2}{2}\left[x + \frac{\sin 2Nx}{2N}\right]_0^\alpha$$

$$= \frac{S^2}{2}\alpha$$

The total area under the waveform over a two line period is therefore $2S^2\alpha$, so that the mean square value is

$$\frac{2S^2\alpha}{2\pi} = \frac{2S^2 \times 8\pi}{2\pi \times 525} = \frac{8S^2}{525} = 0{\cdot}0152S^2$$

This is the mean square value for an infinite number of harmonics. If the sub-carrier component and 20 harmonics on each side are taken, their mean square value is 85% of this total. The efficiency is shown in Fig. A.2.

The above waveform analysis has assumed that the burst is present once every line, but strictly speaking there is also a field frequency component because the burst is omitted during the field sync pulses. However, since only 4 lines are affected in $202\frac{1}{2}$, the field frequency harmonics are very small.

It should be pointed out that, as far as a colour receiver is concerned, the above analysis of the burst waveform assumes that the gating circuit which separates the burst does not itself introduce a spurious signal having a line frequency repetition rate. If the gating circuit does produce a ring, or allows through a sync widget, then such a spurious signal will have a fundamental frequency equal to the line frequency and it will therefore have harmonics which are

integers of the line frequency. These harmonics will be spaced by
$f_L/2$ from the burst waveform sidebands and consequently, if large
enough, may limit the pull-in range of a reference generator to
$\pm f_L/4$ instead of $\pm f_L/2$ which applies in the case of a pure burst
waveform. Such spurious signals can be reduced by ensuring that
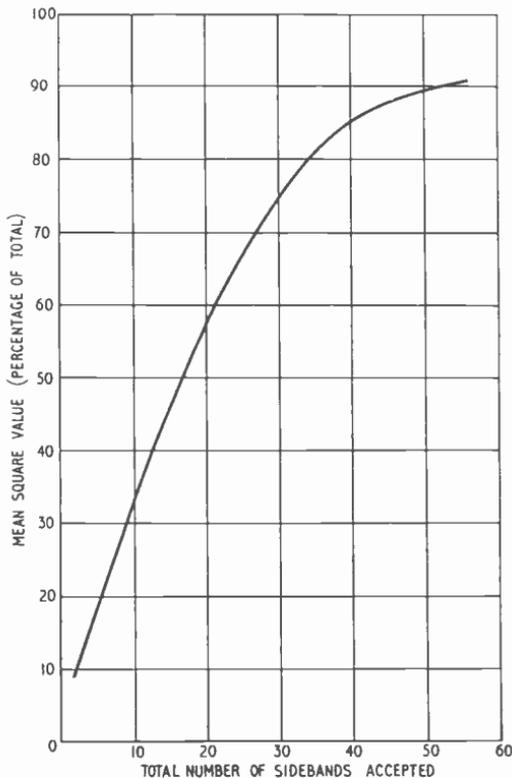
*Fig. A.2.   Graph of the efficiency of recovery of burst
energy against the bandwidth of the burst channel*

the gating waveform does not have very fast rise and fall times,
and by positioning the gating waveform accurately to exclude
widgets.

If a colour signal is used in which the line frequency is not correctly
locked to the sub-carrier frequency (or alternatively is not held to
within about 0.01% of its nominal value) spurious line harmonics
may very appreciably impair the pull-in performance of a reference
generator and may cause side-lock to a frequency very close to the

correct sub-carrier component.   This is in no way a function of the burst waveform itself which may be regarded as amplitude modulation of the sub-carrier by a rectangular waveform of line frequency repetition rate which has an " on " period equal to the burst duration.   This modulating waveform has harmonics which are multiples of the line frequency, so that after modulation the frequencies produced will be equal to the sub-carrier frequency plus and minus multiples of the line frequency.   Thus, regardless of the numerical relationship between the sub-carrier and line frequencies, the nearest sidebands to the sub-carrier component will be spaced from it by the line frequency.   On the other hand, since a slight change in the fundamental frequency of a spurious line frequency component can produce an appreciable change in the absolute frequency of a high order harmonic, then if the gating circuit introduces spurious line frequency components into the burst waveform, pull-in performance can be reduced.   For example, if the line frequency were 20 c/s high at 10,145 c/s (about 0.2% high) then the 262nd harmonic would be 2,657,990 c/s, which is only 177·5 c/s higher than the sub-carrier component of 2,657,812·5 c/s.   Such a close spurious signal could seriously impair pull-in performance, and it is therefore necessary to use a correctly locked line and sub-carrier colour signal for a fair test of reference generator pull-in performance.   This would not apply to reference generators which do not introduce spurious line frequency signals into the gated burst signal.

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Relation between
dynamic phase error, S/N ratio and
noise bandwidth

From the point of view of reference generator design, it is useful to express the r.m.s. phase error for a given signal-to-noise ratio in terms of equivalent noise bandwidth.

Thus, if the burst amplitude is S, the signal during the burst is $S \cos \omega_s t$, where ω_s is 2π times the sub-carrier frequency. The presence of random noise will cause fluctuations in this signal, and the phase errors which result are of particular interest as far as the reference generator is concerned. If the instantaneous noise which is present with the signal after integration by the reference generator is $N(t)$, this may be considered as the sum of two noise components in quadrature such that

$$N(t) = a(t) \cos \omega_s t + b(t) \sin \omega_s t$$

If a and b are typical instantaneous amplitudes of the quadrature components, the total noise power will be given by

$$\overline{N^2} = \overline{(a \cos \omega_s t + b \sin \omega_s t)^2}$$

which represents the mean square value of the cosine and sine components, the amplitudes of which have a random variation with time.

The mean square value of cosine or sine waveforms is half the square of the amplitude, so that

$$\overline{N^2} = \frac{\overline{a^2}}{2} + \frac{\overline{b^2}}{2}$$

where $\overline{a^2}$ is the mean of the squares of all the a components and $\overline{b^2}$ is the mean of the squares of all the b components.

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But by symmetry,

$$\overline{a^2} = \overline{b^2}$$

so that

$$\overline{N^2} = \overline{a^2} = \overline{b^2}$$

That is, the r.m.s. value of each of the quadrature components is equal to the r.m.s. noise N, or

$$a_{RMS} = b_{RMS} = N_{RMS}$$

Now the instantaneous amplitudes of signal and noise may be represented by the vector diagram of Fig. A.3 in which the instantaneous phase error ϕ is given by

$$\tan \phi = \frac{b}{S + a} \qquad \cdot$$

which reduces to $\phi = \dfrac{b}{S}$ if the noise is not excessive. Hence, approximately

$$\phi_{RMS} = \frac{b_{RMS}}{S} = \frac{N}{S}$$

The r.m.s. noise N above is that which is present with the signal after integration by the reference generator. That is, before integration, the noise has a value N_w in the video band f_w (3 Mc/s in

Fig. A.3.   *Derivation of phase error due to noise components*
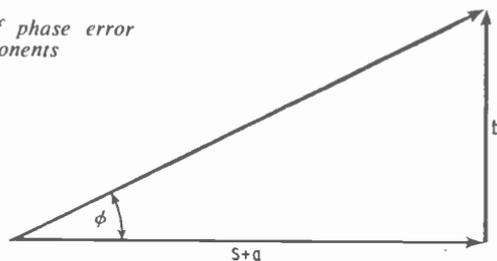
a=INSTANTANEOUS NOISE COMPONENT IN PHASE WITH SIGNAL S
b=INSTANTANEOUS NOISE COMPONENT IN QUADRATURE WITH SIGNAL S

the British system), so that the noise power is N_w^2 for a bandwidth f_w, or $\dfrac{N_w^2}{f_w}$ per unit bandwidth. Consequently, if the reference generator has an integration time T_m corresponding to a noise

bandwidth f_N, the noise power after integration will be $\dfrac{N_w{}^2}{w} f_N$, and this is equal to N^2.

Hence

$$\phi_{RMS} = \frac{N}{S} = \frac{N_w}{S} \sqrt{\frac{f_N}{f_w}}$$

However, since the signal is present for only a fraction d of the time, where d is the duty ratio of the burst, the effective integration time is dT_m, i.e. the effective noise bandwidth is $\dfrac{f_N}{d}$. Allowing for the burst duty ratio, the r.m.s. phase error becomes

$$\phi_{RMS} = \frac{N_w}{S} \sqrt{\frac{f_N}{df_w}} \text{ radians}$$

$$= \frac{180}{\pi} \frac{N_w}{S} \sqrt{\frac{f_N}{df_w}} \text{ degrees}$$

In the American N.T.S.C. system, $f_w = 4 \cdot 3$ Mc/s and $d = \dfrac{16}{455}$ so that

$$\phi_{RMS} = 0 \cdot 148 \sqrt{f_N} \frac{N_w}{S}$$

In the British N.T.S.C. system, $f_w = 3$ Mc/s and $d = \dfrac{16}{525}$ so that

$$\phi_{RMS} = 0 \cdot 189 \sqrt{f_N} \frac{N_w}{S}$$

Note that the adverse tolerance of an 8 cycle burst is chosen. The duty ratio of the burst is derived by noting that during 2 consecutive line periods there are 455 or 525 cycles of sub-carrier (depending on the system) and 16 cycles of burst.

For a dynamic phase error of 5° r.m.s. and a signal-to-noise ratio of $\dfrac{S}{N_w} = \dfrac{1}{2}$ (i.e. peak-to-peak burst equal to r.m.s. noise), the required noise bandwidth is

$$f_N = 285 \text{ c/s for the American system}$$
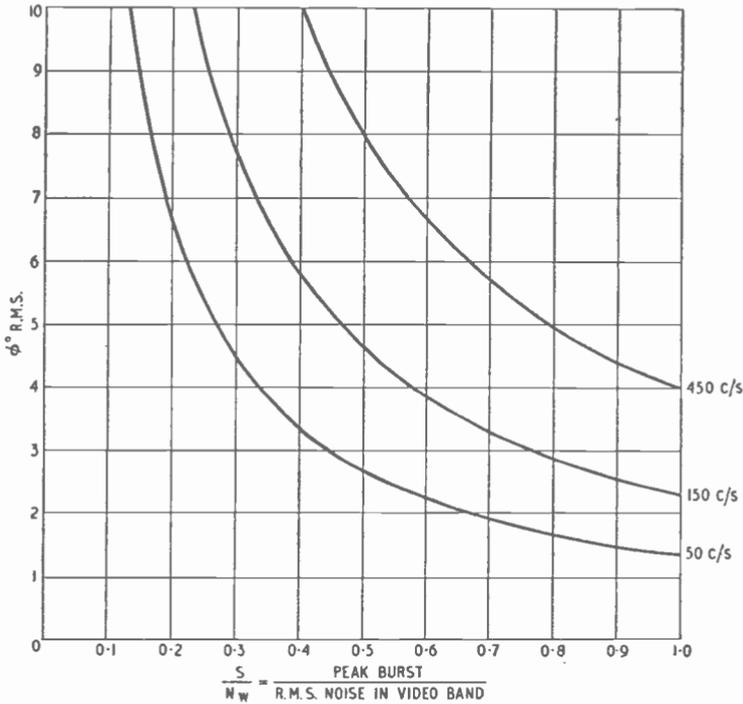
Fig. A.4. R.m.s. phase error as a function of signal-to-noise ratio for noise bandwidths of 50 c/s, 150 c/s and 450 c/s (British system)

and

$$f_N = 175 \text{ c/s for the British system}$$

If excess burst gating width is to be allowed, these figures must be reduced. (*See* Section 10.4.) Thus, for an excess burst gate of 1·5 to 1, the noise band-widths become

$$f_N = 232 \text{ c/s for the American system}$$

and

$$f_N = 143 \text{ c/s for the British system}$$

In Fig. A.4, the r.m.s. phase error (in degrees) has been plotted against signal-to-noise ratio for three different noise bandwidths.

For any given noise bandwidth, the r.m.s. phase error is inversely proportional to the signal-to-noise ratio, so that a design for a

particular S to N_w ratio automatically ensures less dynamic error for higher S to N_w ratios.

For a given signal-to-noise ratio, the dynamic phase error is proportional to the square root of the noise bandwidth. Thus, reducing the noise bandwidth 9 times (from 450 c/s to 50 c/s, for example) reduces the dynamic phase error by a factor of 3.

In-sync performance of an
A.P.C. loop

Let us assume that the loop is synchronized in frequency and phase, and it is required to determine the static and dynamic phase errors in terms of the loop parameters.

The loop parameters include the sensitivity of the reactance valve, that is, the frequency shift of the oscillator in cycles per second produced by one volt applied to the reactance valve, usually called β c/s/V; the sensitivity of the phase detector, that is, the voltage output of the phase detector for a phase difference of one radian between the two signals being compared, usually called μ volts/radian; and finally, the output voltage versus input voltage characteristic of the loop low pass filter, which we shall call $F(\omega)$, to indicate that this voltage ratio is a function of the frequency $\frac{\omega}{2\pi}$ of the voltages involved.

As far as the in-sync performance is concerned, small values of static or dynamic phase errors are of interest in which case the phase detector output is virtually linear with phase angle. Strictly, the usual type of phase detector gives an output which is proportional to the sine of the angle by which the phase difference between the signals being compared differs from 90°.

Thus, if one signal is $\sin(\omega_s t + \theta)$ and the other is

$$\sin(\omega_s t + \phi + 90°)$$

the phase detector output is $\mu \sin(\theta - \phi)$. However, if $(\theta - \phi)$ is small, then

$$\mu \sin(\theta - \phi) \simeq \mu(\theta - \phi)$$

A.3.1. Static phase error

If the oscillator has an initial tuning error of Δf c/s, then the reactance valve requires an input voltage of $\frac{\Delta f}{\beta}$ to correct this error. Since this is a static error, the correction voltage is D.C. and the low

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pass filter transmission is therefore 100%, so that the phase detector output voltage must be $\dfrac{\Delta f}{\beta}$. Hence, if the phase difference between output and input signals is $\Delta\phi_0$, which is the static phase error, then

$$\mu \sin \Delta\phi_0 = \frac{\Delta f}{\beta} \simeq \mu \, \Delta\phi_0$$

that is

$$\Delta\phi_0 = \frac{\Delta f}{\mu\beta} = \frac{\Delta f}{f_c}$$

where $f_c = \mu\beta$ is usually called the D.C. loop gain. (Note the paradox that although f_c is the D.C. loop gain, it is measured in cycles per second per radian since $\mu =$ volts/radian and $\beta =$ cycles/second/volt.)

Thus, the static phase error is directly proportional to the initial detuning error, and inversely proportional to the loop gain. For a given detuning error, therefore, the static phase error may be reduced as desired by making $\mu\beta$ large enough.

As far as dynamic phase error is concerned, the variations of output phase due to noise fluctuations of the input phase are of interest.

A simple approach to this problem is to assume a noise free input phase which is given a sinusoidal variation with time, and to calculate the magnitude of the corresponding output phase at various frequencies. This is analogous to finding the voltage transfer characteristic of a filter, except that here phase angle changes with frequency are of interest instead of voltage amplitude changes. Having calculated the phase transfer or phase following characteristic of the loop, the equivalent noise bandwidth and hence the dynamic phase error may be determined.

A.3.2. Phase transfer ratio

Assume that the loop is frequency and phase locked and that the input phase is changed by an amount $\Delta\theta$ at a time t. If the corresponding output phase change is $\Delta\phi$, the phase detector output change will be $\mu(\Delta\theta - \Delta\phi)$. The voltage output from the loop filter will therefore be $\mu(\Delta\theta - \Delta\phi)F(\omega)$, where $F(\omega)$ is the voltage transfer ratio of the filter. This voltage will produce a frequency shift of $\mu(\Delta\theta - \Delta\phi)F(\omega)\beta$ c/s by way of the reactance valve, i.e. a shift of $f_c(\Delta\theta - \Delta\phi)F(\omega)$ c/s or $\omega_c(\Delta\theta - \Delta\phi)F(\omega)$ radians/second, where

$\omega_c = 2\pi f_c$. This frequency shift, however, must be equal to the rate of change of output phase, since the loop is frequency locked.

Hence

$$\omega_c(\Delta\theta - \Delta\phi)F(\omega) = \frac{d\Delta\phi}{dt}$$

$$= p\Delta\phi$$

where p represents the operator $\dfrac{d}{dt}$

Therefore

$$\Delta\theta = \Delta\phi + \frac{p\Delta\phi}{\omega_c F(\omega)}$$

or

$$\frac{\Delta\phi}{\Delta\theta} = \frac{F(\omega)}{F(\omega) + p/\omega_c}$$

If the input phase variation is sinusoidal, then $p = j\omega$ and

$$\frac{\Delta\phi}{\Delta\theta} = \frac{F(\omega)}{F(\omega) + j\omega/\omega_c} = \frac{\text{amplitude of output phase change}}{\text{amplitude of input phase change}}$$

This ratio is often called $Q(\omega)$, to indicate that it is, like $F(\omega)$, a function of the frequency $\dfrac{\omega}{2\pi}$ at which the phase is varied.

By using the above relation, it is possible to calculate the phase following ratio $Q(\omega)$ for any given A.P.C. loop. For example, if the low pass filter of the loop is as shown in Fig. 10.9, then

$$\frac{\text{Output voltage of filter}}{\text{Input voltage of filter}} = F(\omega) = \frac{1 + j\omega xT}{1 + j\omega(1 + x)T}$$

where

$$T = RC$$

Then

$$Q(\omega) = \frac{F(\omega)}{F(\omega) + j\omega/\omega_c} = \frac{1 + j\omega xT}{1 - \omega^2(1 + x)T/\omega_c + j\omega(xT + 1/\omega_c)}$$

It must be remembered that $Q(\omega)$ is a ratio of output to input phases and not voltages. However, since voltage transfer characteristics are usually more familiar, the expression for $Q(\omega)$ above is the

same as the voltage transfer characteristic of the circuit of Fig. 10.10, which has a resistance of $\dfrac{1}{\omega_c C}$ in series with an inductance of $(1 + x)\dfrac{R}{\omega_c}$ in the top arm.

It is very important to notice that the expression for $Q(\omega)$ is not only a function of the filter characteristic $F(\omega)$. The loop constant $\omega_c = 2\pi f_c$ is also involved. While the low pass filter will modify the output voltage of the phase detector, the effect on the output phase will clearly depend on the magnitude of this voltage (i.e. the phase detector sensitivity) and its frequency shifting capabilities (i.e. the reactance valve sensitivity).

A.3.3. Dynamic phase error

A plot of $|Q(\omega)|$ against frequency is shown in Fig. 10.11 for the filter configuration of Fig. 10.9.

As far as phase variations are concerned, the A.P.C. loop behaves like a narrow band filter tuned to the sub-carrier frequency, and it is possible to find the equivalent noise bandwidth by squaring the $|Q(\omega)|$ curve, and finding the equivalent rectangular bandwidth of the same height and which encloses the same area, as described in Section 10.1.

Notice, however, that the filter shape of $Q(\omega)$ responds equally well to frequencies above and below the sub-carrier frequency, so that the $|Q(\omega)|^2$ curve will correspond to the noise semi-bandwidth, usually called f_{NN}, such that

$$f_{NN} = \frac{f_N}{2}$$

Hence

$$f_N = 2f_{NN} = 2 \int_0^\infty |Q(\omega)|^2 \, df$$

where the integral is the area under the squared $Q(\omega)$ curve.

$$\therefore \qquad\qquad f_{NN} = \int_0^\infty |Q(\omega)|^2 \, d\left(\frac{\omega}{2\pi}\right)$$

$$= \frac{1}{2\pi} \int_0^\infty |Q(\omega)|^2 d\omega$$

Substituting for $|Q(\omega)|^2$ according to

$$|Q(\omega)^2| = \frac{1 + \omega^2 x^2 T^2}{\left[1 - \frac{\omega^2}{\omega_c}(1 + x)T\right]^2 + \omega^2\left(xT + \frac{1}{\omega_c}\right)^2}$$

it will be found that

$$\frac{f_N}{2} = f_{NN} = \frac{\omega_c\left[1 + \frac{x^2 T\omega_c}{(1 + x)}\right]}{4(1 + xT\omega_c)}$$

Notice that, for a given value of ω_c, two variables are available for satisfying this relation. Hence, apart from the noise bandwidth, some control may be exercised over the shape of the $Q(\omega)$ curve to prevent resonant ringing due to impulse interference. To investigate this, consider the operational form of $Q(\omega)$ which is

$$Q(p) = \frac{1 + pxT}{1 + \frac{p^2}{\omega_c}(1 + x)T + p\left(xT + \frac{1}{\omega_c}\right)}$$

This is the general expression for $Q(\omega)$, in which the quantities are not necessarily sinusoidal. The previous expression in which $p = j\omega$, applies only to sinusoidal variations. Hence, if $\Delta\phi$ and $\Delta\theta$ are the output and input phases, respectively, then

$$\frac{\Delta\phi}{\Delta\theta} = Q(p) \qquad \text{or} \qquad \frac{1}{Q(p)}\Delta\phi = \Delta\theta$$

The complementary function of this differential equation is given by

$$\frac{1}{Q(p)} = 0$$

that is

$$1 + \frac{p^2}{\omega_c}(1 + x)T + p\left(xT + \frac{1}{\omega_c}\right) = 0$$

or

$$(1 + x)Tp^2 + (1 + xT\omega_c)p + \omega_c = 0$$

Solving this equation for p, the quantity under the square root sign is

$$(1 + xT\omega_c)^2 - 4(1 + x)T\omega_c$$

If a damping factor K is defined by $K = \dfrac{(1 + xT\omega_c)^2}{4(1 + x)T\omega_c}$, then

if $K > 1$, the quantity under the square root sign is positive and over damping occurs ;

if $K < 1$, the quantity under the square root sign is negative so that oscillatory ringing occurs ;

if $K = 1$, the damping is critical and near optimum conditions are obtained in practice.

Hence, having chosen ω_c so that the static phase error is within the required limit for a given detuning error, x and T may be chosen so that the required noise bandwidth is obtained. Further, since x and T may be chosen independently, some control is available over the shape of the $Q(\omega)$ curve.

For the range of values which are of interest, $x \ll 1$ and $xT\omega_c \gg 1$ so that

$$f_{NN} \simeq \frac{\omega_c(1 + x^2T\omega_c)}{4xT\omega_c}$$

$$= \frac{1 + x^2T\omega_c}{4xT}$$

$$= \frac{1}{4xT} + \frac{x\omega_c}{4}$$

and

$$K \simeq \frac{(xT\omega_c)^2}{4T\omega_c} = \frac{x^2T\omega_c}{4}$$

so that

$$f_{NN} \simeq \frac{1 + 4K}{4xT}$$

or

$$xT \simeq \frac{(K + \frac{1}{4})}{f_{NN}}$$

and

$$x^2T \simeq \frac{4K}{\omega_c}$$

Hence

$$\frac{x^2T}{xT} = x \simeq \frac{4K}{\omega_c} \frac{f_{NN}}{(K + \frac{1}{4})}$$

and

$$T = \frac{1}{x} xT \simeq \frac{\omega_c(K + \frac{1}{4})}{4Kf_{NN}} \frac{(K + \frac{1}{4})}{f_{NN}}$$

$$= \frac{\omega_c(K + \frac{1}{4})^2}{4Kf_{NN}^2}$$

Hence, x and T may be calculated for a given noise bandwidth and for given values of ω_c (which determines the static phase shift for a given detuning) and K (which determines the shape of the response curve).

Note that since

$$f_{NN} = \frac{1}{4xT} + \frac{x\omega_c}{4}$$

a value of x can be found which makes f_{NN} a minimum. This value $x_{\min}$ is given by

$$x_{\min} = \frac{1}{\sqrt{T\omega_c}}$$

in which case the damping factor for least noise bandwidth is

$$K_{\min} = \frac{x_{\min}^2 T\omega_c}{4} = \frac{1}{4}$$

and the minimum value of f_{NN} is

$$f_{NN\min} = \frac{2}{4x_{\min}T} = \frac{1}{2}\sqrt{\frac{\omega_c}{T}}$$

However, to avoid ringing on impulse type interference, it is advisable to use a K value nearer to 1.

Pull-in performance of an A.P.C. loop

In this section, formulae relating to the pull-in performance of the loop shown in Fig. 10.9 will be derived.

The problem is to determine the time taken for the loop oscillator to lock in frequency and phase to the burst signal, when the loop oscillator has an initial tuning error of $\Delta\omega$ rads/sec.

A.4.1. The loop parameters

The phase detector sensitivity, μ volts/radian.

The reactance valve sensitivity, β c/s/V.

The D.C. loop gain, $f_c = \mu\beta$.

The operational form of the voltage transfer ratio of the loop filter, F(p), which is equal to the ratio of the output voltage to the input voltage of the filter.

A.4.2. The loop differential equation

The loop performance will be determined by a differential equation which may be derived as follows:

Suppose the instantaneous voltage output of the loop filter is v volts, then this will produce a frequency shift of βv c/s in the oscillator frequency. If the initial oscillator error was Δf c/s, it is now reduced to $(\Delta f - \beta v)$ c/s or $(\Delta\omega - 2\pi\beta v)$ rads/sec by the action of the reactance valve.

If the instantaneous phase error between the input and output signals is ϕ rads, the instantaneous frequency error is $\dfrac{d\phi}{dt}$ rads/sec. Therefore

$$\frac{d\phi}{dt} = \Delta\omega - 2\pi\beta v$$

But the phase detector output is $\mu \sin \phi$, and $\dfrac{v}{\mu \sin \phi} = F(p)$

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Eliminating v from these two equations

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} = \Delta\omega - 2\pi\beta v = \Delta\omega - 2\pi\mu\beta F(p) \sin\phi = \Delta\omega - \omega_c F(p) \sin\phi$$

or

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + \omega_c F(p) \sin\phi = \Delta\omega$$

This is the required equation which is the relation between ϕ and t. However, for the loop filter of Fig. 10.9 the operational form of $F(\omega)$ in which $p = \dfrac{\mathrm{d}}{\mathrm{d}t} = j\omega$ is

$$F(p) = \frac{1 + pxT}{1 + p(1 + x)T}$$

and if this is substituted in the differential equation, the resulting equation cannot be solved except by numerical methods.

A.4.3. The simplified loop

An approximate solution can be found by making use of the results for a simple loop filter. Thus, if the loop filter is merely a resistive network as shown in Fig. A.5, the voltage transfer ratio is

Fig. A.5.   Simple resistive loop filter
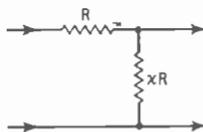
VOLTAGE TRANSFER RATIO $= \dfrac{xR}{R+xR} = \dfrac{x}{1+x} = m$

simply given by $F(p) = m$. The equation for this simplified loop is therefore

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + m\omega_c \sin\phi = \Delta\omega$$

or

$$\mathrm{d}t = \frac{\mathrm{d}\phi}{\Delta\omega - m\omega_c \sin\phi}$$

Hence

$$m\omega_c t = \int \frac{\mathrm{d}\phi}{\dfrac{\Delta\omega}{m\omega_c} - \sin\phi} + \text{constant}$$

This may be evaluated, and there are two solutions depending on whether $\dfrac{\Delta\omega}{m\omega_c}$ is less or greater than one. If $\dfrac{\Delta\omega}{m\omega_c} < 1$, and if the phase at $t = 0$ is ϕ_0, then the solution is given by

$$\frac{\tan\dfrac{\phi}{2} - \tan\dfrac{\phi_\infty}{2}}{\tan\dfrac{\phi}{2} - \cot\dfrac{\phi_\infty}{2}} = \left[\frac{\tan\dfrac{\phi_0}{2} - \tan\dfrac{\phi_\infty}{2}}{\tan\dfrac{\phi_0}{2} - \cot\dfrac{\phi_\infty}{2}}\right] e^{-m\omega_c t \cos\phi_\infty}$$

where ϕ_∞ is the value of ϕ after an infinite time, and is given by $\sin\phi_\infty = \dfrac{\Delta\omega}{m\omega_c}$. That is, ϕ_∞ is the ultimate static phase shift.

The above expression is rather similar to that for the voltage decay curve of a discharging condenser

$$v = V e^{-t/CR}$$

and the " time constant " is

$$\frac{1}{m\omega_c \cos\phi_\infty} = \frac{1}{m\omega_c\sqrt{1 - \left(\dfrac{\Delta\omega}{m\omega_c}\right)^2}}$$

$$= \frac{1}{\sqrt{(m\omega_c)^2 - (\Delta\omega)^2}}$$

However, the value of ϕ at any time t will be a function of the initial phase ϕ_0, since the " V " is a function of ϕ_0. As an example of phase stabilization time, suppose that $\phi_\infty = 2°$ and $\phi_0 = 90°$.

Then, if t_3 is the time taken for the phase error to reach $3°$ (i.e. within $1°$ of the ultimate phase), it will be found that

$$m\omega_c t_3 = 4\cdot75$$

A typical practical value for $m\omega_c$ would be about 200 radians/second/radian so that

$$t_3 = \frac{4\cdot75}{200} \text{ sec}$$

or about 24 msec.

The important point to notice about the above solution is that ϕ approaches ϕ_∞ without changing by more than 2π. In other words

there is no frequency slip, only a phase transient. The design of
A.P.C. loops is usually more concerned with the case where fre-
quency slip does occur, since the time for frequency pull-in is very
much greater than that required for the phase to approach reason-
ably closely to ϕ_∞.

The " frequency slip " case occurs when $\dfrac{\Delta\omega}{m\omega_c} > 1$, for which the

solution is

$$m\omega_c t = \frac{2}{\sqrt{a^2 - 1}}\left[\tan^{-1}\left(\frac{a\tan\dfrac{\phi}{2} - 1}{\sqrt{a^2 - 1}}\right) - \tan^{-1}\left(\frac{a\tan\dfrac{\phi_0}{2} - 1}{\sqrt{a^2 - 1}}\right)\right]$$

where

$$a = \frac{\Delta\omega}{m\omega_c} = \frac{\Delta f}{mf_c}$$

The meaning of this equation can be clarified by writing

$$y = \frac{a\tan\dfrac{\phi}{2} - 1}{\sqrt{a^2 - 1}} \qquad Y = \frac{a\tan\dfrac{\phi_0}{2} - 1}{\sqrt{a^2 - 1}}$$

and

$$m\omega_c\sqrt{a^2 - 1} = \frac{2\pi}{T_B}$$

Then, the equation becomes

$$\frac{\pi}{T_B}\,t = \tan^{-1}y - \tan^{-1}Y$$

$$\therefore \qquad \tan\frac{\pi t}{T_B} = \frac{y - Y}{1 + yY}$$

or

$$y = \frac{Y + \tan \pi t/T_B}{1 - Y\tan \pi t/T_B}$$

that is   .

$$y = \tan\left(\pi t/T_B + \Phi\right)$$

where $\tan\Phi = Y$

Now since

$$y = \frac{a \tan \dfrac{\phi}{2} - 1}{\sqrt{a^2 - 1}}$$

then

$$\tan \frac{\phi}{2} = \frac{1 + y\sqrt{a^2 - 1}}{a}$$

But

$$\sin \phi = \frac{2 \tan \phi/2}{1 + \tan^2 \phi/2}$$

$$= \frac{2a[1 + y \sqrt{a^2 - 1}]}{a^2 + (1 + y \sqrt{a^2 - 1})^2}$$

$$\therefore \qquad \sin \phi = \frac{2a \left[1 + \sqrt{a^2 - 1} \tan \left(\dfrac{\pi t}{T_B} + \Phi\right)\right]}{a^2 + \left[1 + \sqrt{a^2 - 1} \tan \left(\dfrac{\pi t}{T_B} + \Phi\right)\right]^2}$$

The quantity $\sin \phi$ is proportional to the phase detector output, which is $\mu \sin \phi$. Hence, the above expression gives the waveform of the detector output, and notice that while $\sin \phi$ has the same value for ϕ as for $(\phi + 2\pi)$, the right hand side of the equation has the same value for t as for $(t + T_B)$. That is, $\sin \phi$ is cyclic with a period

$$T_B = \frac{2\pi}{m\omega_c \sqrt{a^2 - 1}}$$

Since $a = \dfrac{\Delta f}{mf_c}$

$$T_B = \frac{1}{mf_c\sqrt{\left(\dfrac{\Delta f}{mf_c}\right)^2 - 1}} = \frac{1}{\sqrt{\Delta f^2 - (mf_c)^2}}$$

The waveform of $\sin \phi$ is shown in Fig. 10.12 for three different values of $\dfrac{\Delta f}{mf_c}$. Notice that this beatnote waveform is not sinusoidal, but tends to become sinusoidal as $\dfrac{\Delta f}{mf_c}$ increases. Notice also that the peak positive and negative excursions are equal, and the waveform

becomes inverted if the sign of " a " is negative, i.e. if the oscillator error Δf is negative.

To summarize the performance of the simplified loop, if $\dfrac{\Delta f}{mf_c} < 1$ the loop pulls in without slipping a cycle, and the phase error ultimately reaches a value equal to the static phase error. The phase detector generates a control voltage proportional to $\dfrac{\Delta f}{mf_c}$, i.e. proportional to the tuning error, so that no frequency slip occurs.

If $\dfrac{\Delta f}{mf_c} > 1$, this loop does not pull in, and a steady beatnote output results from the phase detector. This waveform has a D.C. component, however, which reduces the mean tuning error, but for large tuning errors the waveform tends to become sinusoidal so that the D.C. component is then small. For various values of Δf (greater than mf_c) it would be possible to measure the D.C. component in terms of its frequency shifting capabilities, and obtain a curve of D.C. control against Δf. This D.C. control may be calculated by finding the D.C. component of the beatnote waveform. Thus, the differential equation of the simple loop is

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + m\omega_c \sin \phi = \Delta\omega$$

and if this is integrated over a cycle of beatnote

$$\int_t^{t+T_B} \frac{\mathrm{d}\phi}{\mathrm{d}t}\,\mathrm{d}t + \int_t^{t+T_B} m\omega_c \sin \phi \,\mathrm{d}t = \int_t^{t+T_B} \Delta\omega\mathrm{d}t$$

i.e.

$$2\pi + \int_t^{t+T_B} m\omega_c \sin \phi \,\mathrm{d}t = \Delta\omega T_B$$

The integral in this equation is the area under a cycle of the phase detector filter output, and the D.C. component of the output is therefore this integral divided by T_B. Calling this D.C. component $m\omega_c \overline{\sin \phi}$

$$m\omega_c \overline{\sin \phi} = \Delta\omega - \frac{2\pi}{T_B}$$

$$= \Delta\omega - 2\pi\sqrt{\Delta f^2 - (mf_c)^2}$$

or

$$mf_c \overline{\sin \phi} = \Delta f - \sqrt{\Delta f^2 - (mf_c)^2}$$

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Hence, for a positive tuning error (oscillator frequency too high) a positive D.C. component is generated which tends to reduce the oscillator frequency. If the tuning error is negative, the opposite sign of the square root applies and

$$mf_c \overline{\sin \phi} = - [\Delta f - \sqrt{\Delta f^2 - (mf_c)^2}]$$

that is, the D.C. component is negative. In this case the beatnote waveform of Fig. 10.12 is inverted.

A plot of $mf_c \overline{\sin \phi}$, which is the D.C. component of the waveform applied to the reactance valve (measured in terms of frequency) is shown in Fig. A.6. Notice that the control is strong up to $\Delta f = mf_c$, but for values of $\Delta f > mf_c$ the control becomes progressively weaker as the beatnote waveform becomes more like a
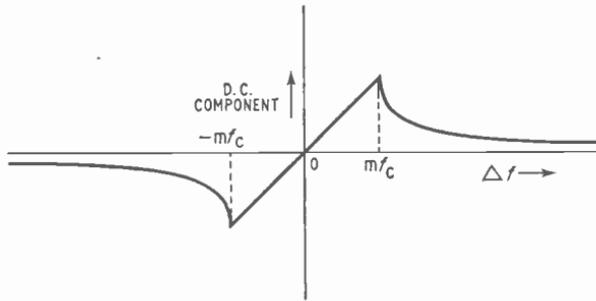
Fig. A.6.   D.C. component of beatnote waveform as a function of tuning error

sine wave. Notice also that although the transfer characteristic of the simplified loop filter is not frequency selective, the loop behaves as though there is a falling frequency response in that the control voltage reduces as the beatnote frequency increases. The reason for this behaviour is the loop time constant $\dfrac{1}{mf_c}$. Thus, if mf_c is very large, $\dfrac{1}{mf_c}$ is very small and can change the phase appreciably during the short time of one cycle of beatnote. On the other hand, if mf_c is very small very little phase change can occur during a beatnote cycle.

The results obtained above for the simplified loop can be applied to the loop which uses the filter shown in Fig. 10.9, in which a condenser is included in series with the shunt resistor. As in the

simplified loop, if the tuning error is small so that $\dfrac{\Delta f}{mf_c} < 1$, the loop does not slip a cycle and there is merely a phase transient. If $\dfrac{\Delta f}{mf_c} > 1$, however, a beatnote output from the phase detector results, and the D.C. component of it is stored in the condenser. The resulting mean beatnote frequency is thereby reduced, and a larger D.C. component is therefore obtained, which again reduces the mean beatnote frequency still further. This process continues (provided that the tuning error, though greater than mf_c, is not excessive. The limits for the tuning error will be derived later) until the tuning error has been reduced to mf_c. The loop has then frequency locked, and a phase transient occurs as in the simplified loop when $\Delta f < mf_c$.

A.4.4. Quantitative analysis of the loop

The differential equation of the loop, quoted earlier, is

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + \omega_c F(p) \sin\phi = \Delta\omega$$

For the loop filter of interest (Fig. 10.9),

$$F(p) = \frac{1 + pxT}{1 + p(1 + x)T}$$

Putting $x = \dfrac{m}{1 - m}$ $\left(\text{i.e. } m = \dfrac{x}{1 + x}\right)$

$$F(p) = \frac{1 - m + pmT}{1 - m + pT}$$

which may be written as

$$F(p) = m + \frac{1 - 2m + m^2}{1 - m + pT}$$

Substituting this in the equation above, it follows that

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + m\omega_c \sin\phi + \left[\frac{1 - 2m + m^2}{1 - m + pT}\right]\omega_c \sin\phi = \Delta\omega$$

or

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + m\omega_c \sin\phi = \Delta\omega - \left[\frac{1 - 2m + m^2}{1 - m + pT}\right]\omega_c \sin\phi$$

This equation is the same as that for the simplified loop except that here the right hand side of the equation is as quoted above, while for the simplified loop it is merely $\Delta\omega$.

The quantity

$$\Delta\omega - \left[\frac{1 - 2m + m^2}{1 - m + \mathrm{p}T}\right] \omega_c \sin\phi$$

therefore represents an instantaneous tuning error which reduces as pull-in progresses.

Writing

$$\omega = \Delta\omega - \left[\frac{1 - 2m + m^2}{1 - m + \mathrm{p}T}\right] \omega_c \sin\phi$$

the differential equation becomes

$$\frac{\mathrm{d}\phi}{\mathrm{d}t} + m\omega_c \sin\phi = \omega$$

Using the results obtained for the simplified loop, the mean value of $m\omega_c \sin\phi$ over a cycle of beatnote is

$$m\omega_c \overline{\sin\phi} = \overline{\omega} - \sqrt{\overline{\omega^2} - (m\omega_c)^2}$$

where $\overline{\omega}$ is the mean tuning error. In the simplified loop this quantity was, of course, constant. Here, however, it increases as $\overline{\omega}$ reduces.

From the equation

$$\omega = \Delta\omega - \left[\frac{1 - 2m + m^2}{1 - m + \mathrm{p}T}\right] \omega_c \sin\phi$$

the mean value of ω is given by

$$\overline{\omega} = \Delta\omega - \overline{\left[\frac{1 - 2m + m^2}{1 - m + \mathrm{p}T}\right] \omega_c \sin\phi}$$

The quantity on the extreme right hand side is the mean value of the frequency shift produced by the condenser voltage, and since the time constant T of the loop filter is normally very much greater than the beatnote period, it would seem to be legitimate to assume

that this voltage is a result of applying the D.C. component of the phase detector output $\omega_c \overline{\sin \phi}$ to the loop filter.

That is

$$\overline{\left[\frac{1 - 2m + m^2}{1 - m + pT}\right] \omega_c \sin \phi} \simeq \left[\frac{1 - 2m + m^2}{1 - m + pT}\right] \omega_c \overline{\sin \phi}$$

This may be shown as follows:

The operational impedance of the loop filter is

$$Z_p = (1 + x)R + \frac{1}{pC}$$

For an applied voltage $\omega_c \overline{\sin \phi}$, the instantaneous current is

$$\frac{\omega_c \overline{\sin \phi}}{Z_p}$$

and the voltage across the condenser is

$$\frac{\omega_c \overline{\sin \phi}}{Z_p} \cdot \frac{1}{pC} = \frac{\omega_c \overline{\sin \phi}}{1 + p(1 + x)T}$$

This voltage is applied to the reactance valve through a potentiometer of xR and R, so that a voltage

$$\frac{R}{(1 + x)R} \cdot \frac{\omega_c \overline{\sin \phi}}{1 + p(1 + x)T}$$

is applied to the reactance valve from the condenser.

Putting $1 + x = \dfrac{1}{1 - m}$, the voltage applied to the reactance valve from the condenser is

$$\frac{(1 - m) \omega_c \overline{\sin \phi}}{1 + \dfrac{pT}{(1 - m)}} = \frac{(1 - 2m + m^2) \omega_c \overline{\sin \phi}}{1 - m + pT}$$

From the two equations

$$\omega_c \overline{\sin \phi} = \frac{1}{m}\left[\bar{\omega} - \sqrt{\bar{\omega}^2 - (m\omega_c)^2}\right]$$

$$\omega_c \overline{\sin \phi} = \left[\frac{1 - m + pT}{1 - 2m + m^2}\right] (\Delta\omega - \overline{\omega})$$

that is,

$$\omega_c \overline{\sin \phi} = \frac{\Delta\omega}{(1 - m)} - \frac{\overline{\omega}}{(1 - m)} - \frac{T}{(1 - m)^2} \frac{d\overline{\omega}}{dt}$$

After eliminating $\omega_c \overline{\sin\phi}$ and putting $\rho = \dfrac{\overline{\omega}}{m\omega_c}$ and $\rho_0 = \dfrac{\Delta\omega}{m\omega_c}$,

it follows that

$$\frac{(1 - m)}{mT} dt = \frac{d\rho}{m\rho_0 - \rho + (1 - m)\sqrt{\rho^2 - 1}}$$

If T_F is the time required to pull in from $\overline{\omega} = \Delta\omega$ to $\overline{\omega} = m\omega_c$ (that is, T_F is the frequency pull-in time) the corresponding values of ρ are $\rho = \rho_0$ to $\rho = 1$, and therefore

$$\frac{(1 - m)T_F}{mT} = \int_{\rho_0}^{1} \frac{d\rho}{m\rho_0 - \rho + (1 - m)\sqrt{\rho^2 - 1}}$$

A.4.5. Pull-in range

Before attempting to carry out this integration, it is possible to calculate the maximum pull-in range. Thus, the above integral is the area enclosed between the curve of $\left[\dfrac{1}{m\rho_0 - \rho + (1-m)\sqrt{\rho^2-1}}\right]$ plotted against ρ, and the ρ axis, between the limits $\rho = \rho_0$ and $\rho = 1$. Now if a real value of ρ can occur between ρ_0 and 1 which makes $\left[m\rho_0 - \rho + (1 - m)\sqrt{\rho^2 - 1}\right]$ zero, the reciprocal will be infinite and so will be the area under the curve. This will correspond to an infinite value of T_F.

Now if

$$m\rho_0 - \rho + (1 - m)\sqrt{\rho^2 - 1} = 0$$

then

$$m(2 - m)\rho^2 - 2m\rho_0\rho + (1 - m)^2 + m^2\rho_0^2 = 0$$

and therefore

$$\rho = \frac{2m\rho_0 \pm \sqrt{4m^2\rho_0^2 - 4m(2 - m)[(1 - m)^2 + m^2\rho_0^2]}}{2m(2 - m)}$$

This cannot have a real value if

$$4m^2\rho_0{}^2 < 4m(2 - m)\,[(1 - m)^2 + m^2\rho_0{}^2]$$

that is

$$m\rho_0{}^2 < 2 - m$$

or

$$\rho_0 < \sqrt{\frac{2}{m} - 1}$$

The pull-in time T_F will therefore be finite provided that

$$\frac{\Delta f_{\max}}{mf_c} < \sqrt{\frac{2}{m} - 1}$$

or

$$\Delta f_{\max} < f_c\,\sqrt{2m - m^2}$$

where $\Delta f_{\max}$ is the maximum tuning error.

A.4.6. Pull-in time

The above integral can be evaluated, but the solution is cumbersome. However, we are normally interested in loops in which $m \ll 1$, and making this approximation, we have

$$(1 - m)\frac{T_F}{mT} = \int_{\rho_0}^{1} \frac{\mathrm{d}\rho}{\sqrt{\rho^2 - 1} - \rho}$$

$$= \int_{\rho_0}^{1} \frac{\sqrt{\rho^2 - 1} + \rho}{(\sqrt{\rho^2 - 1})^2 - \rho^2}\,\mathrm{d}\rho$$

that is

$$(1 - m)\frac{T_F}{mT} = \int_{1}^{\rho_0} (\sqrt{\rho^2 - 1} + \rho)\,\mathrm{d}\rho$$

$$= \tfrac{1}{2}\,[\rho_0\sqrt{\rho_0{}^2 - 1} + \rho_0{}^2 - 1 - \cosh^{-1}\rho_0]$$

Now provided ρ_0 is not near 1, the quantity on the right hand side is approximately equal to $\rho_0{}^2$.

For example, if $\rho_0 = 5$ then

$$(1 - m)\frac{T_F}{mT} = 23{\cdot}1 \simeq 5^2)$$

Therefore

$$(1 - m)\frac{T_F}{mT} \simeq \rho_0{}^2 = \left(\frac{\Delta f}{mf_c}\right)^2$$

This expression does not include the restriction that T_F becomes infinite when $\rho_0^2 \geq \dfrac{2 - m}{m}$, but by writing

$$\frac{(1 - m)T_F}{mT} = \frac{\rho_0^2}{1 - \dfrac{m}{(2 - m)} \rho_0^2}$$

this restriction is included.   Since $\dfrac{1 - m}{m} = \dfrac{1}{x}$, it follows that

$$T_F = \frac{xT\rho_0^2}{1 - \dfrac{m}{(2 - m)} \rho_0^2}$$

For the range of values which are of interest

$$m \simeq x \ll 1$$

so that

$$T_F \simeq xT\rho_0^2 = xT\left(\frac{\Delta f}{xf_c}\right)^2$$

$$= \frac{T}{x}\left(\frac{\Delta f}{f_c}\right)^2$$

Using the expressions

$$x \simeq \frac{4Kf_{NN}}{\omega_c(K + \frac{1}{4})} \quad \text{and} \quad T \simeq \frac{\omega_c(K + \frac{1}{4})^2}{4Kf_{NN}^2}$$

which were derived at the end of Appendix 3 the pull-in time may be expressed as

$$T_F \simeq \frac{\omega_c^2(K + \frac{1}{4})^3}{4^2K^2f_{NN}^3}\left(\frac{\Delta f}{f_c}\right)^2$$

$$= \left(\frac{\pi}{2}\right)^2 \frac{(K + \frac{1}{4})^3}{K^2}\frac{(\Delta f)^2}{f_{NN}^3}$$

For given values of Δf and f_{NN}, K may be chosen to make T_F as small as possible.   The minimum value of $\dfrac{(K + \frac{1}{4})^3}{K^2}$ occurs when $K = \frac{1}{2}$, and then

$$T_F \simeq \left(\frac{\pi}{2}\right)^2 \frac{27}{64} 4 \frac{(\Delta f)^2}{f_N N^3}$$

$$\simeq 4 \cdot 2 \frac{(\Delta f)^2}{f_N N^3}$$

If $K = 1$

$$T_F \simeq 4 \cdot 8 \frac{(\Delta f)^2}{f_N N^3}$$

and if $K = \frac{1}{4}$

$$T_F \simeq 4 \cdot 93 \frac{(\Delta f)^2}{f_N N^3}$$

It is therefore preferable for K to be greater than $\frac{1}{2}$ rather than less than $\frac{1}{2}$, and since the shape of the $Q(\omega)$ response is optimum near $K = 1$, it would appear that a K of 1 gives (very nearly) best in-sync and best pull-in performance.

Phase detector

A typical phase detector circuit which is often used in reference oscillators is shown in Fig. A.7(a).

One signal to be phase compared is fed push-pull to two diodes, while the second signal is fed to the anode and cathode junction of the diodes. The output is taken either from point X (through a
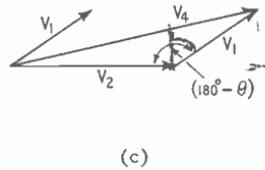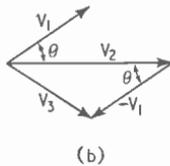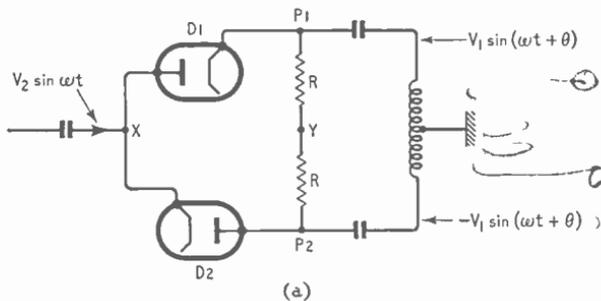
Fig. A.7. Basic phase detector circuit

suitable resistor to prevent capacitance loading of the preceding circuit) while Y is earthed, or from point Y while X is earthed through a suitable resistor.

A.5.1. Sensitivity of the phase detector

If V_1 is the peak value of half the total push-pull drive, and V_2 is the peak value of the second signal, then the peak value of the voltage across diode D_1 is the vector difference between V_2 and V_1. If this difference is V_3, as shown in Fig. A.7(b), then

$$V_3{}^2 = V_1{}^2 + V_2{}^2 - 2V_1V_2 \cos\theta = V_1{}^2 \left[1 + \left(\frac{V_2}{V_1}\right)^2\right]$$

$$\times \left[1 - \frac{\dfrac{2V_2}{V_1}}{1 + \left(\dfrac{V_2}{V_1}\right)^2} \cos\theta\right]$$

Now it is possible to make $V_1 \gg V_2$, so that $\left(\dfrac{V_2}{V_1}\right)^2$ is small compared with 1. Under these conditions, V_3 is approximately given by

$$V_3 \simeq V_1 \left[1 + \left(\frac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}} \left[1 - \frac{\dfrac{V_2}{V_1}}{1 + \left(\dfrac{V_2}{V_1}\right)^2} \cos\theta\right]$$

$$= V_1 \left[1 + \left(\frac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}} - \frac{V_2 \cos\theta}{\left[1 + \left(\dfrac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}}}$$

Since V_3 is the peak value of the voltage across diode $D1$ the point P_1 will have a positive D.C. voltage relative to X of V_3, where V_3 is given by the above expression.

Similarly, point P_2 will have a negative D.C. voltage relative to X which will be equal to the vector sum of the peak values of V_2 and V_1, as shown in Fig. A.7(c). Calling this sum V_4,

$$V_4{}^2 = V_1{}^2 + V_2{}^2 - 2V_1V_2 \cos(180° - \theta) = V_1{}^2 + V_2{}^2 + 2V_1V_2 \cos\theta$$

and therefore

$$V_4 \simeq V_1 \left[1 + \left(\frac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}} + \frac{V_2 \cos\theta}{\left[1 + \left(\dfrac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}}}$$

where V_4 is a negative voltage relative to X.

Now the total D.C. voltage across the resistors R is $V_3 + V_4$ and the D.C. current through them is $\dfrac{V_3 + V_4}{2R}$. The voltage drop across the top resistor is therefore

$$\frac{V_3 + V_4}{2R} R = \frac{V_3 + V_4}{2}$$

The point Y is therefore positive relative to X by a voltage of

$$V_3 - \left(\frac{V_3 + V_4}{2}\right) = \frac{V_3 - V_4}{2}$$

$$\simeq - \frac{V_2 \cos \theta}{\left[1 + \left(\frac{V_2}{V_1}\right)^2\right]^{\frac{1}{2}}}$$

i.e. the point Y is negative to X by approximately $V_2 \cos \theta$. Hence, if X is earthed through a resistor, the output at Y will be $- V_2 \cos \theta$. On the other hand, if Y is earthed and the output is taken from X through a resistor, the output will be $+ V_2 \cos \theta$.

Note that the output is proportional to the cosine of the phase angle between the signals being compared, and to the amplitude of the smaller signal, which is numerically equal to the sensitivity of the phase detector. In the above example, therefore, if V_2 had been made greater than V_1, so that $\frac{V_1}{V_2}$ was less than 1, the output would have been $V_1 \cos \theta$. In reference generators it is consequently advisable to make the burst signal applied to the detector larger than the oscillator signal. The phase detector output sensitivity is then proportional to the oscillator drive, which is generally " cleaner " than the incoming burst signal.

A.5.2. Effect of gated nature of burst

There is another reason for making the burst signal larger than the oscillator reference signal. Thus, it has so far been assumed that the reference and burst signals are both continuous sine waves. However, the burst signal is actually a short sample of sine wave and this fact changes the characteristics of the phase detector somewhat.

Thus, referring to Fig. A.8(a), this shows a large push-pull signal and a small " single ended " signal. The waveform v_{p1} is the instantaneous voltage at the point $P1$ of Fig. A.7(a), v_{p2} is the instantaneous voltage at P_2 and v_x is the instantaneous voltage at X, assuming Y to be earthed, and $\theta = 180°$.

If the larger signal is the burst, then after the burst has occurred the bias on the diodes will be at $+ V_1$ and $- V_1$ volts, D.C. These bias voltages are sufficiently large to prevent diode conduction during the period between bursts, that is, the phase detector operates as a peak detector. On the other hand, if the small " single ended " signal is the burst signal, then any D.C. voltage developed during the burst will be discharged by the conduction of
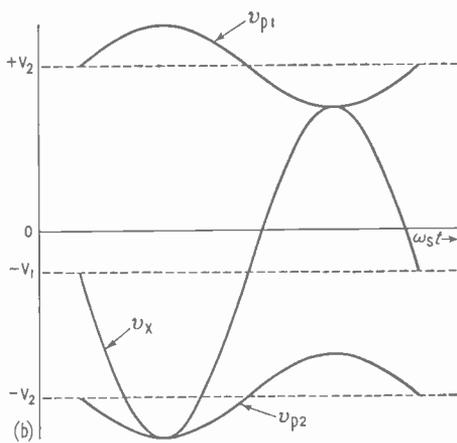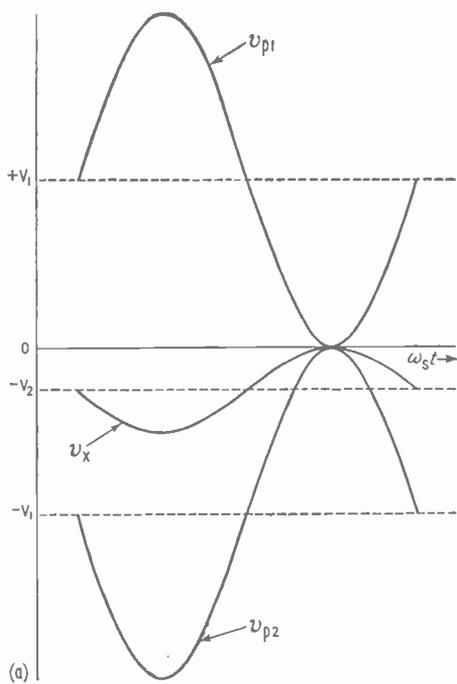
Fig. A.8. Phase detector drive waveforms for
(a) large push-pull signal, (b) small push-pull
signal

the appropriate diode on the peaks of the reference signal, after the burst has occurred.

A similar situation obtains if the " single ended " signal is larger than the push-pull signal, as shown in Fig. A.8(b). A little thought will show that provided the peak burst signal is greater than twice the peak reference signal, the detector will tend to operate as a peak detector in either case. Thus, a greater efficiency results by making the peak burst signal exceed twice the reference signal, whichever signal is made the push-pull drive.

A.5.3. Justification for approximations

The reader may query the approximations made under the assumption that $V_1 \gg V_2$ (or vice versa). In practice, " very much greater than " may, in this instance, be taken to mean " two to one " or more. In fact, if x is put for $\dfrac{V_2}{V_1}$, in the example of Section A.5.1 the approximate detector output is

$$- xV_1 \cos \theta$$

while the exact output is given by

$$\frac{V_3 - V_4}{2} = \frac{V_1 \sqrt{1 + x^2 - 2x \cos \theta} - V_1 \sqrt{1 + x^2 + 2x \cos \theta}}{2}$$

The ratio

$$\frac{\text{Approximate output}}{\text{Exact output}}$$

is a maximum when $\theta = 90°$ or $270°$, and it then has a value of $\sqrt{1 + x^2}$.

If $V_1 = 2V_2$, $x = 0.5$ and the above ratio is $\sqrt{1.25}$, or 1.12. Thus, for a two to one amplitude ratio, the approximate output is only 12% high even under the worst conditions.

It is interesting to note that while $\theta = 90°$ or $270°$ gives the maximum value of the above ratio, the approximate and exact outputs are in fact both equal (i.e. zero) when θ is exactly $90°$ or $270°$. However, $\theta = 90°$ or $270°$ gives the limiting value of the maximum ratio.

Reactance valve

A typical reactance valve circuit is shown in Fig. A.9, in which only those components relating to the A.C. voltages and currents are shown.

The purpose of the reactance valve is to vary the tuning of the tank circuit by means of a D.C. voltage applied to the grid of the triode. If C and R are both small, their series impedance is substantially capacitative and the voltage across R therefore leads the
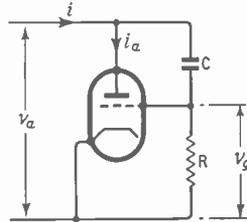
Fig. A.9. Reactance valve circuit

applied voltage by approximately 90°, that is v_g leads v_a by 90°. The anode current i_a, which is $g_m v_g$, therefore leads v_a by 90°. The impedance $\dfrac{v_a}{i_a}$ is therefore capacitative, the value of which is a function of the g_m of the triode. The g_m is itself a function of the D.C. bias on the triode, and hence varying the D.C. bias will alter the tuning of the tank circuit.

A.6.1. Value of the impedance $\frac{v_a}{i}$

The current through the C, R circuit is

$$\frac{v_a}{R + \dfrac{1}{j\omega C}} = \frac{j\omega C v_a}{1 + j\omega CR}$$

Hence

$$v_g = \frac{j\omega C v_a R}{1 + j\omega CR}$$

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and since $i_a = g_m v_g$ (provided that the triode R_a is large) it follows that

$$i_a = \frac{j\omega CR g_m v_a}{1 + j\omega CR}$$

Hence

$$i = i_a + \frac{j\omega C v_a}{1 + j\omega CR}$$

$$= \frac{j\omega C v_a}{1 + j\omega CR}[g_m R + 1]$$

and therefore

$$\frac{v_a}{i} = \frac{1 + j\omega CR}{(g_m R + 1)j\omega C}$$

$$= \frac{1}{(g_m R + 1)}\left[R + \frac{1}{j\omega C}\right]$$

The triode and its CR circuit therefore behaves like a resistance $\frac{R}{g_m R + 1}$ in series with a capacitance of $C(g_m R + 1)$, that is a resistance

$$\frac{R}{(g_m R + 1)}\left(1 + \frac{1}{\omega^2 C^2 R^2}\right)$$

in parallel with a capacitance

$$\frac{(g_m R + 1)C}{1 + \omega^2 C^2 R^2}$$

The resistance damps the tank circuit without affecting the tuning, while the capacitance has a component $\frac{g_m RC}{1 + \omega^2 C^2 R^2}$ which is proportional to g_m.

For values of interest in reference generators ωCR is small compared with 1 so that the effective shunt capacitance due to g_m may be taken as $g_m RC$.

A.6.2. Sensitivity of reactance valve

The effect of the reactance valve on the tuning of the tank circuit may be calculated:

If the tank circuit is tuned to a frequency ω radians/second, with an inductance L_T and a capacitance C_T, then

$$\omega^2 = \frac{1}{L_T C_T}$$

We wish to know how ω changes if a small change is made in C_T.

Hence differentiating with respect to C_T

$$2\omega \frac{d\omega}{dC_T} = -\frac{1}{L_T C_T^2} = -\frac{\omega^2}{C_T}$$

Therefore

$$dC_T = -2C_T \frac{d\omega}{\omega} = -2C_T \frac{df}{f}$$

where $f = \dfrac{\omega}{2\pi}$

(The negative sign indicates the direction of tuning, i.e. increasing dC_T produces a negative effect on frequency, thereby reducing it.)

Now suppose for example, the tuning needs to be altered by 2 kc/s at 2·7 Mc/s, the tank circuit capacitance being 500pF.

Then

$$dC_T = -2 \times 500 \times 10^{-12} \times \frac{2 \times 10^3}{2 \cdot 7 \times 10^6} = -0 \cdot 74 \text{pF}$$

That is, we require a tank circuit capacitance change of 0·74pF.

If the reactance valve changes its g_m by 0·1mA/V for a bias change of 1 V, then numerically

$$0 \cdot 1 \times 10^{-3} RC = 0 \cdot 74 \times 10^{-12}$$

i.e. $$RC = 7 \cdot 4 \times 10^{-9}$$

If R is made 2·2kΩ, then $C = 3 \cdot 35$pF. These values will therefore give a reactance valve sensitivity of 2 kc/s/V.

It is usual to include a decoupled cathode feedback resistor in the reactance valve to stabilize the valve characteristics against ageing. The effective g_m of the valve with a feedback resistor R_F is

$$\frac{g_m}{1 + g_m R_F}$$

and it is this value which must be used in calculating the sensitivity.

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For example, a typical valve characteristic would give a $g_m = 2\cdot7$mA/V for a bias of -1 V, and a $g_m = 1\cdot6$mA/V for a bias of -2 V. If a decoupled 1kΩ cathode feedback resistor is used, the effective g_m is

$$\frac{2\cdot7}{1 + 2\cdot7} = 0\cdot73\text{mA/V}$$

at -1 V, and

$$\frac{1\cdot6}{1 + 1\cdot6} = 0\cdot61\text{mA/V}$$

at -2 V. The change in g_m for a 1 V bias change is then

$$0\cdot73 - 0\cdot61 = 0\cdot12\text{mA/V}$$

It is important to remember that the above type of reactance valve reduces the tank circuit frequency as the bias voltage is made more positive. If C and R were interchanged (a suitable blocking condenser being included in series with R) the reactance valve would behave as a variable inductance, and making the bias voltage more positive would reduce the inductance and therefore increase the frequency. This inductance type of reactance valve is not normally used in reference generators, since the oscillator has a high C to L ratio for stability reasons, and is therefore very sensitive to changes in inductance.

Quadricorrelator equivalent filter

The standard A.P.C. loop filter is shown in Fig. A.10(a). This
has a voltage transfer characteristic

$$\frac{V_2}{V_1} = \frac{1 + j\omega CxR}{1 + j\omega C(R + xR)}$$

When used in a variable ratio quadricorrelator circuit, the trans-
mission through this filter is increased by applying the voltage at
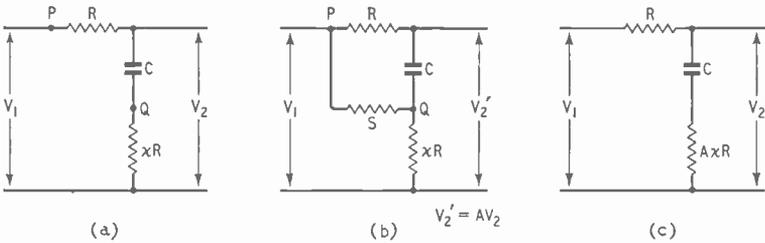
Fig. A.10. Equivalent circuit of a quadricorrelator filter

P to Q via a cathode follower. Hence, no phase shift is introduced
and the equivalent filter circuit becomes that shown in Fig. A.10(b),
where S is a resistance whose value is such that

$$V_2' = AV_2$$

i.e., the filter transmission is increased by a factor A.
From Fig. A.10(b)

$$\frac{V_2'}{V_1} = \frac{1 + \dfrac{j\omega CxR(S + R)}{S + xR}}{1 + \dfrac{j\omega C[RS + xR(S + R)]}{S + xR}}$$

Comparing the two ratios $\dfrac{V_2}{V_1}$ and $\dfrac{V_2'}{V_1}$, note that

$$xR \text{ is now } \quad \frac{\left(1 + \dfrac{R}{S}\right)xR}{1 + \dfrac{xR}{S}}$$

and

$$R + xR \text{ is now } \frac{RS + xR(S + R)}{S + xR}$$

i.e.

$$R \text{ is now } \frac{1}{1 + \dfrac{xR}{S}} R$$

Normally, x is less than $0\cdot01$ and the increase in transmission through the filter is at least 10 to 1. That is, $A = 10$ (at least), so that $\dfrac{R}{S} \simeq 10$ also, i.e. $\dfrac{R}{S} \simeq A$

Thus

$$xR \text{ is now } \frac{(1 + A)xR}{1 + xA} \simeq AxR$$

and

$$R \text{ is now } \frac{1}{1 + 0\cdot1} R$$

Hence, for those values of x and A which are of interest, the equivalent circuit of the increased transmission filter is as shown in Fig. A.10(c).

Effect of resistive mismatch on the value of the phase shift along a characteristically terminated cable

In the co-axial cable phase shifter described in Section 12.5, the need to accurately terminate the cable to prevent errors in phase reading is mentioned. The error produced by a given mismatch may be calculated:

For a line of length l, propagation constant $P(= \alpha + j\beta)$, characteristic impedance Z_O, termination load Z_L, the ratio of input to output voltage for a sine wave signal is given by

$$\frac{V_I}{V_O} = \cosh Pl + \frac{Z_O}{Z_L} \sinh Pl$$

If the line is assumed to be lossless, $\alpha = 0$ and therefore $P = j\beta$. The characteristic impedance Z_O becomes purely resistive, i.e. $Z_O = R_O$.

Then

$$\frac{V_I}{V_O} = \cos \beta l + j\frac{R_O}{Z_L} \sin \beta l$$

If θ is the phase by which V_O lags V_I, then

$$\tan \theta = \frac{R_O}{Z_L} \tan \beta l$$

Now suppose the load impedance Z_L is purely resistive, and equal to R_L, and suppose that

$$\frac{R_O}{R_L} = 1 + \epsilon$$

that is, ϵ is the fractional mismatch. Then if the change in phase

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due to the mismatch is δ such that the actual phase shift θ is given by

$$\theta = \theta_O + \delta$$

where $\theta_O = \beta l$ is the phase shift for correct termination, then

$$\tan(\theta_O + \delta) = (1 + \epsilon) \tan \theta_O$$

Expanding the left hand side,

$$\frac{\tan \theta_O + \tan \delta}{1 - \tan \theta_O \tan \delta} = (1 + \epsilon) \tan \theta_O$$

If δ is small, then $\tan \delta \simeq \delta$ so that

$$\frac{\tan \theta_O + \delta}{1 - \delta \tan \theta_O} \simeq (1 + \epsilon) \tan \theta_O$$

whence

$$\delta \simeq \frac{\epsilon \tan \theta_O}{\sec^2 \theta_O}$$

$$= \frac{\epsilon}{2} \sin 2\theta_O$$

$$= \frac{\epsilon}{2} \sin 2\beta l$$

The error angle δ is therefore a function of the mismatch and the length of line. It is a maximum when $\sin 2\beta l = \pm 1$, i.e. when

$$2\beta l = \frac{\pi}{2} \quad \text{or} \quad \frac{3\pi}{2}$$

The maximum value of the error δ is therefore given by

$$\delta = \frac{\epsilon}{2} \text{ radians} \quad \text{or} \quad \frac{90\epsilon^\circ}{\pi}$$

and it occurs for phase shifts of 45° and 135°, i.e. when $\beta l = \frac{\pi}{4}$ or $\frac{3\pi}{4}$.

Hence, for a 10% mismatch, for example, $\epsilon = 0 \cdot 1$ and $\delta \simeq 3°$. It is therefore recommended that the mismatch be limited to not more than 1%, in which case the maximum error is only about one third of a degree.

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