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NUMBER 47: MAY 1963

Vertical Aperture Correction using
Continuously Variable Ultrasonic Delay Lines

by

D. HOWORTH, B.Sc. Tech., Grad.I.E.E.
(Research Department, BBC Engineering Division)

PART I

AN EXPERIMENTAL VERTICAL APERTURE CORRECTOR

PART II

A CONTINUOUSLY VARIABLE DELAY UNIT
FOR TELEVISION SIGNALS

BRITISH BROADCASTING CORPORATION

PRICE FIVE SHILLINGS



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FOREWORD

THIS is one of a series of Engineering Monographs published by the British Broadcasting Corporation. About six are produced every year, each dealing with a technical subject within the field of television and sound broadcasting. Each Monograph describes work that has been done by the Engineering Division of the BBC and includes, where appropriate, a survey of earlier work on the same subject. From time to time the series may include selected reprints of articles by BBC authors that have appeared in technical journals. Papers dealing with general engineering developments in broadcasting may also be included occasionally.

This series should be of interest and value to engineers engaged in the fields of broadcasting and of telecommunications generally.

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VERTICAL APERTURE CORRECTION USING CONTINUOUSLY VARIABLE ULTRASONIC DELAY LINES

SUMMARY

Part I of this monograph describes the principle of operation, construction, and calibration of a simple vertical aperture corrector which has been made possible by the use of ultrasonic delay lines. The performance of this aperture corrector and some of the problems involved in the design of an operational version are discussed. The method of calibration is based on a 'Calibration Factor' obtained by a simple and convenient pulse procedure. It is shown that the steady-state amplitude/frequency characteristic, and also the effect of the aperture corrector on signal-to-noise ratio, can be determined from this factor.

In Part II the construction and characteristics of an ultrasonic delay line using mercury as the transmission medium are discussed in some detail, and an experimental variable delay unit suitable for a 405-line television system is described. There is an account of the difficulties encountered in constructing the video delay unit and some indication is given as to the possibility of extending the use of the line to television systems requiring a higher video bandwidth. In addition to its use in the vertical aperture corrector, such a unit may have other television applications.

PART I

AN EXPERIMENTAL VERTICAL APERTURE CORRECTOR

1. Introduction

The need for aperture correction* of television signal sources is well known and horizontal aperture correction is included in most television systems. The problem of applying vertical aperture correction is more difficult and has for many years virtually been ignored. The development of ultrasonic delay lines, however, has made the construction of a vertical aperture corrector possible.

This report describes an experimental vertical aperture corrector first put forward by W. G. Gibson and A. C. Schroeder of R.C.A.¹ The principal of operation is similar to that of the horizontal aperture corrector known as a derivative equalizer. In the simple practical example to be described approximations to the first and second derivatives are generated.

2. Principle of Operation

Gouriet has shown² that if a signal is linearly distorted during transmission, the resulting signal can be corrected by adding proportions of its own successive derivatives.

$$\text{Hence } g(t) = a_0 f(t) + a_1 f'(t) + a_2 f''(t) + \dots$$

where $g(t)$ is the original signal

$f(t)$ is the distorted signal

$f'(t), f''(t), \dots$ are successive derivatives of $f(t)$

and a_0, a_1, a_2, \dots are constant coefficients.

* Aperture correction may be defined as the correction of the distortion produced by the finite size and non-uniform flux distribution of the scanning spot. The nature of the television scanning process clearly necessitates entirely different approaches to the problems of applying this correction in the horizontal and vertical directions.

This method is often used for horizontal aperture correction and can be adapted for use as a vertical aperture corrector by using difference signals to approximate to the derivatives.

The fundamental definition of the first derivative of a function is

$$f'(t) = \lim_{\tau \rightarrow 0} \frac{f(t) - f(t - \tau)}{\tau}$$

Hence an approximation to the first derivative can be obtained by subtracting from the function a delayed version of itself. The limitation of the difference method of approximating to a first derivative can be demonstrated by considering a simple sinusoid.

$$\text{Let } f(t) = \sin \omega t$$

$$\text{Then } \Delta_1(t) = \frac{f(t) - f(t - \tau)}{\tau}$$

$$= \frac{2}{\tau} \left[\sin \frac{\omega \tau}{2} \right] \left[\cos \omega \left(t - \frac{\tau}{2} \right) \right] \quad (1)$$

It can be seen that as $\tau \rightarrow 0$ the first difference signal $\Delta_1(t)$ approaches the true first derivative $\omega \cos \omega t$. For finite values of τ , however, the difference is only a close approximation to the derivative for small values of ω with respect to $2/\tau$. Fig. 1 shows the departure from the ideal case; it will be seen that the difference signal can only be considered to approximate to the derivative up to a frequency of $1/2\tau$. Successive differences can be similarly produced by introducing more delayed signals.

Aperture correction by adding difference signals is readily applicable to the production of a vertical aperture corrector. If the delay τ is made equal to the duration of

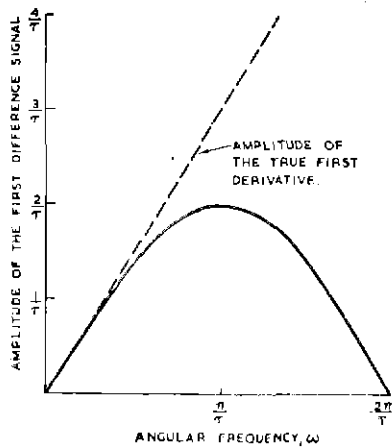


Fig. 1 — Amplitude of the first-difference signal versus angular frequency.

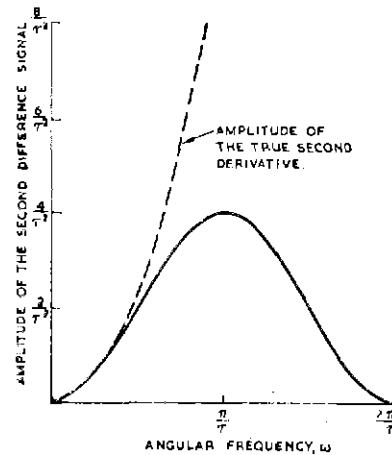


Fig. 2 — Amplitude of the second-difference signal versus angular frequency.

one television line, all true horizontal information, being the same for successive lines, will be cancelled, leaving only vertical information. Furthermore, the television system can be considered to be a sampling system in the vertical direction, each row of picture elements being sampled in turn. In this case, the maximum frequency which can be carried by the system is $1/2\tau$, and this conveniently is also the maximum frequency up to which the difference signal can be considered to be an approximate derivative.

Most uncorrected television apertures are, in the author's experience, symmetrical and can be corrected by means of even-order differences. The simple aperture corrector described in this report was designed primarily to produce a second-difference signal.

The second-difference signal can be produced by subtraction as follows:

$$\Delta_2(t) = \frac{\Delta_1(t) - \Delta_1(t - \tau)}{\tau}$$

where $\Delta_1(t)$ and $\Delta_1(t - \tau)$ are first differences and $\Delta_2(t)$ is the second difference.

Once again it is instructive to consider the case of a simple sinusoid in order to find the limitations of the second-difference signal.

Let $f(t) = \sin \omega t$

$\Delta_1(t)$ is given in equation (1)

and it can be shown that

$$\Delta_1(t - \tau) = \frac{2}{\tau} \left[\sin \frac{\omega\tau}{2} \right] \left[\cos \omega \left(t - \frac{3}{2}\tau \right) \right] \quad (2)$$

From equations (1) and (2) it follows that

$$\Delta_2(t) = -\frac{4}{\tau^2} \left[\sin^2 \frac{\omega\tau}{2} \right] \left[\sin \omega(t - \tau) \right] \quad (3)$$

As in the case of the first difference it can be seen that as $\tau \rightarrow 0$ the second difference $\Delta_2(t)$ approaches the true second derivative $-\omega^2 \sin \omega t$. For finite values of τ , however, the

difference is only a close approximation to the derivative for values of ω small with respect to $2/\tau$. Fig. 2 shows the departure from the ideal case; it will be seen that the second-difference signal can only be considered to approximate to the second derivative up to a frequency of $1/2\tau$, that is, over the same frequency range as applies for the first difference.

It can be seen from equations (1), (2), and (3) that the difference signals are all delayed with respect to the input signal; the first-difference signals are delayed by odd multiples of $\tau/2$ and the second difference is delayed by τ . It is necessary for the difference signals to be accurately timed with respect to the signal which they are to correct; therefore the second-difference signal can be used to correct the input signal delayed by τ , but the first-difference signals cannot be used unless further delays of $\tau/2$ are available. A more convenient method of producing a first-difference signal, which would be correctly timed with respect to the input signal delayed by τ , is to subtract the signal delayed by 2τ from the undelayed signal; however, this difference signal will only be effective over a frequency range of one-half that of the second-difference signal.

If the two first-difference signals are not used, a considerable circuit simplification can be obtained by not generating them directly. Consider

$$\Delta_2(t) = \frac{\Delta_1(t) - \Delta_1(t - \tau)}{\tau}$$

Substituting for $\Delta_1(t)$ and $\Delta_1(t - \tau)$ their values in terms of the original function we have

$$\Delta_2(t) = \frac{1}{\tau} \left[\frac{f(t) - f(t - \tau)}{\tau} - \frac{f(t - \tau) - f(t - 2\tau)}{\tau} \right]$$

This simplifies to

$$\Delta_2(t) = \frac{1}{\tau^2} \left\{ \left[f(t) + f(t - 2\tau) \right] - 2f(t - \tau) \right\} \quad (4)$$

Thus the second-difference signal can be generated from the appropriate delayed signals by means of two simple adding circuits.

3. Description of the Vertical Aperture Corrector

Fig. 3 is a block diagram of a complete vertical aperture corrector, which provides both a first-difference and a second-difference signal. The necessary delays are produced by means of two of the video delay units which are described in Part II of this monograph, and which use continuously variable ultrasonic delay lines. These lines were developed by Mullard Research Laboratories and are now available commercially from Mullard Equipment Ltd. The input signal and the signals from the video delay units are fed to a video processing unit in which they are added and subtracted according to equation (4); the second-difference signal thus obtained is added to the input signal which has been delayed by one line. The first-difference signal is generated by subtracting the output of the two-line video delay unit from the input signal. Hence this difference is only effective over a frequency range of one-half that of the second difference.

The complete circuit diagram of the video processing unit is shown in Fig. 4.

4. Performance

Before proceeding to describe the performance of the experimental aperture corrector, it must be emphasized that the principal delay used is one television line, but because of the interlace of the television system this is equivalent to two lines of the actual picture. Therefore the maximum frequency which can be corrected is equivalent to two picture elements; the vertical aperture corrector may be compared to a horizontal aperture corrector giving maximum correction at 1.5 Mc/s and zero correction at 3 Mc/s.

In view of this fact, the effectiveness of the experimental aperture corrector was surprising. In order to obtain a subjective assessment of the effect of the corrector on the

quality of a television picture, a group of observers was shown a series of moving pictures which was obtained from a high-quality flying-spot 35-mm film scanner. They were asked to assess the picture quality according to the scale:

- | | |
|----------------|----------------|
| 1. Excellent | 4. Rather poor |
| 2. Good | 5. Poor |
| 3. Fairly good | 6. Very poor |

As might be expected, the results of this test showed that the corrector is more effective on some pictures than on others. In every case all the observers voted that the quality of the corrected pictures was better than, or at least equal to, that of the uncorrected pictures. The average improvement was voted to be about one grade but in extreme cases an improvement of three grades was registered.

The effectiveness of this rather poor* experimental vertical aperture corrector serves to show how bad the quality of a television picture is in the vertical direction. If the vertical aperture corrector could be made to operate upon successive picture lines it is probable that a further improvement could be achieved. As Gibson and Schroeder pointed out,¹ such a vertical aperture corrector could be made if a good-quality field-delay were available.

The vertical aperture corrector increases the visibility of random noise and other unwanted signals having components in the vertical direction, such as moiré patterns and stationary noise patterns, for example phosphor grain. In the case of a picture source whose signal-to-noise ratio is marginal, the amount of vertical aperture correction which could be used might be limited by the increase in noise. No separate assessments of these defects have been made; however, the subjective tests indicate that they do not nullify the improvement in picture quality.

5. Method of Calibration

The unit could be calibrated by measuring steady-state amplitude/frequency characteristics but this would be tedious, especially as it would involve blanking the

* At least by horizontal aperture corrector standards.

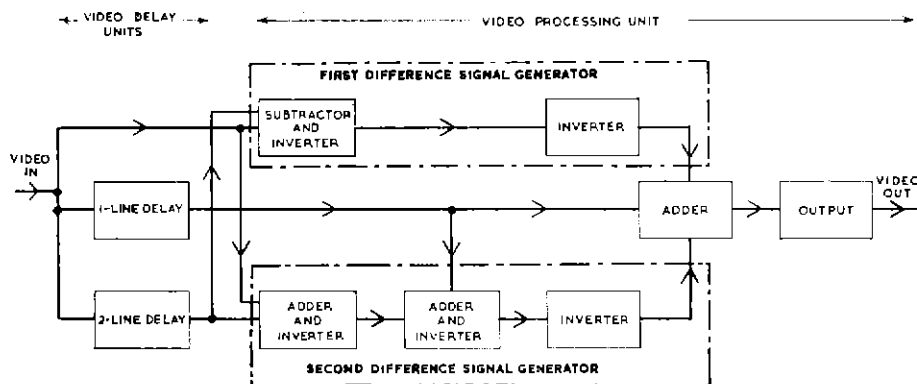


Fig. 3 — Block diagram of the experimental vertical aperture corrector.

variable-frequency test wave in order to simulate a television signal. The pulse method to be described can be carried out rapidly, uses standard equipment, and yields information from which the amplitude/frequency characteristics and the effect of the aperture corrector on signal-to-noise ratio can be calculated.

5.1 Test Procedure

A standard '2T' pulse is fed into the vertical aperture corrector and the delay lines are adjusted until the resultant pulse at the output just splits into three separate components: one positive pulse and two equal-amplitude negative pulses will result (see Fig. 5). The magnitudes of these pulses are measured by means of an oscilloscope and the ratio of negative- to positive-pulse magnitude computed. From a knowledge of this ratio, which can be conveniently termed the 'calibration factor', the steady-state amplitude/frequency characteristics and the effect of the unit on signal-to-noise ratio can be calculated.

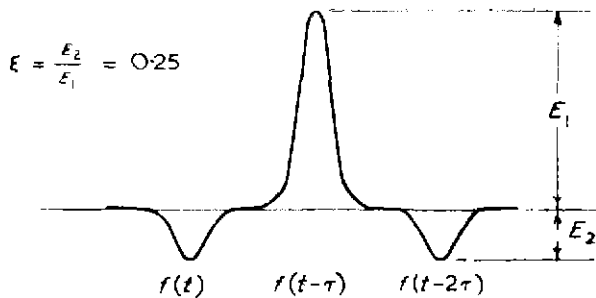


Fig. 5 — Output of the aperture corrector (adjusted for measurement of the calibration factor).

5.2 Calculation of Amplitude/Frequency Characteristics

Let the calibration factor be denoted by ξ . It is apparent from Fig. 5 that the output of the vertical aperture corrector when adjusted for normal working conditions is given by the equation

$$g(t) = f(t - \tau) - \xi[f(t) + f(t - 2\tau)] \quad (5)$$

where $f(t)$ and $g(t)$ are the input and output signals respectively and τ is the duration of one television line.

In order to obtain the steady-state amplitude/frequency characteristics of the aperture corrector let

$$f(t) = \cos \omega t$$

$$\text{hence } g(t) = \cos \omega(t - \tau) - \xi[\cos \omega t + \cos \omega(t - 2\tau)]$$

This expression simplifies to

$$g(t) = [1 - 2\xi \cos \omega\tau] \cos \omega(t - \tau) \quad (6)$$

from the modulus or amplitude of which the amplitude/frequency characteristics can be calculated for various values of ξ .

In a practical vertical aperture corrector it is preferable to design the circuits so that adjustment of the amount of

equalization does not change the magnitude of the output signal corresponding to a plain area of uniform grey. In this case the output of the vertical aperture corrector can be represented by the equation

$$g'(t) = f(t - \tau) + v[2f(t - \tau) - \{f(t) + f(t - 2\tau)\}]$$

where v is a constant depending upon the amount of equalization. The term in square brackets, which represents the correction signal, reduces to zero for a 'd.c. signal' corresponding to uniform grey.

Simplifying this equation gives

$$g'(t) = \left\{ f(t - \tau) - \frac{v}{1 + 2v} [f(t) + f(t - 2\tau)] \right\} (1 + 2v) \quad (7)$$

By comparing equations (5) and (7) we see that

$$\xi = \frac{v}{1 + 2v} \quad (8)$$

Equation (8) shows that with this arrangement the calibration factor ξ cannot exceed 0.5 and thus the corrected picture as a whole cannot change polarity.

In the experimental vertical aperture corrector the maximum value of ξ which is available is 0.25; this is found to be more than adequate for most purposes. Amplitude/frequency characteristics of the vertical aperture corrector for various values of ξ and a line duration of 100 μ s are given in Fig. 6; this diagram may be adjusted to suit any television standard by a linear transformation of the scale of the frequency axis.

5.3 Calculation of the Effect on Signal-to-noise Ratio

Let the magnitude of the input signal corresponding to white be E_1 volts and the r.m.s. voltage of the noise associated with it be e_1 . The signal-to-noise ratio of the input signal may be defined as

$$N_1 = 20 \text{ Log}_{10} \frac{E_1}{e_1}$$

Consider the effect of the vertical aperture corrector on the signal alone. The output voltage E_0 corresponding to white will be given by equation (5) in Section 5.2. Hence, remembering that a delay τ has no effect upon a d.c. signal representing large areas of uniform grey level,

$$E_0 = E_1[1 - 2\xi]$$

The calculation of the effect of the vertical aperture corrector on noise is more complicated but it may be considerably simplified, without introducing an appreciable objective error, if only the noise components above about ten times the television line-scanning frequency are considered.* In this case the noise components of the three

* Noise components below about ten times the television line-scanning frequency will appear mainly as fluctuation in brightness from line to line, which will give the picture a striated appearance. As the r.m.s. magnitude of these striations is proportional to the square root of their bandwidth, they will be at least 20 dB below the r.m.s. magnitude of the full video bandwidth noise. Hence the 'subjective error' introduced by this simplification will also be small. In practice, the action of circuits which are intended to restore the 'd.c. component' of the video signal from line to line will help to minimize the visibility of the striations.

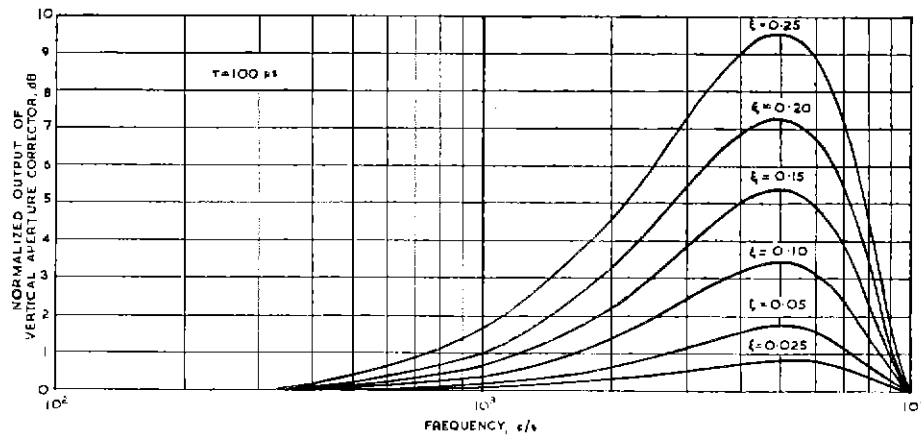


Fig. 6 — Amplitude/frequency characteristics for 405-line television system.

signals, which together form the output signal, will be virtually uncorrelated owing to the relatively long delay between them. The output noise voltage may therefore be calculated from the addition of the three noise powers.

$$\text{Hence } e_0 = e_1[1 + 2\xi^2]^{\frac{1}{2}}$$

where e_0 is the r.m.s. voltage of the output noise. The output signal-to-noise ratio is given by

$$\begin{aligned} N_0 &= 20 \text{Log}_{10} \frac{E_0}{e_0} \\ &= 20 \text{Log}_{10} \frac{E_1(1 - 2\xi)}{e_1(1 + 2\xi^2)^{\frac{1}{2}}} \end{aligned}$$

$$\text{or } N_0 = N_1 + 20 \text{Log}_{10} \frac{(1 - 2\xi)}{(1 + 2\xi^2)^{\frac{1}{2}}}$$

As the calibration factor must lie between 0 and +0.5, the final term in this expression must be negative and re-

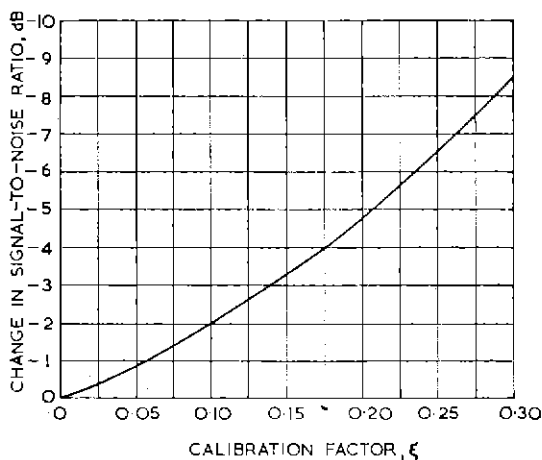


Fig. 7 — Change in signal-to-noise ratio as a function of the calibration factor.

presents the decrease in signal-to-noise ratio caused by this form of vertical aperture correction. Fig. 7 shows calculated values of the change in signal-to-noise ratio plotted against the calibration factor ξ .

6. The Design of an Operational Vertical Aperture Corrector

The following points gained from experience with the experimental vertical aperture corrector may be of use to the designer of an operational unit.

It has been stated that it is essential that the difference signals be accurately timed with respect to the signal which they are to correct. It is estimated that the tolerance on these delays should be plus or minus one-tenth of a picture element. When a crystal-controlled synchronizing pulse generator is used, the ultrasonic delay lines can be adjusted to within this tolerance and will remain within it over long periods of time. When the synchronizing pulse generator is locked to the 50 c/s mains supply, however, frequent re-adjustment of the delay lines is necessary. It would therefore be necessary, in an operational unit incorporated in a locked system, to servo-control the delay lines in order to maintain the correct timing of the difference signals.

If suitable test pulses are available in an operational television system, these can be added to the incoming signal during the field-blanking period in order to provide a means of checking the performance of the equipment continuously by the calibration method described in Section 5.

The presence of the difference signals in the output of a vertical aperture corrector can cause excursions of the video signal well below black level, and these can lead to serious synchronizing difficulties. In order to avoid this, a black-level clipper must be included in the output of the unit. Care must be taken to ensure that the blanking and synchronizing pulses which are used to feed the black-level clipper are correctly timed. If 'station pulses' which are a television line in advance of the output of the aperture corrector are used, the last line of the picture will oc-

cur during the field-suppression period and be visible if the display device does not have field fly-back suppressions. It is essential that this should not be able to occur on a transmitted picture.

7. Conclusions

This simple experimental vertical aperture corrector gives a very significant and worth-while improvement in picture quality and shows that operational vertical aperture cor-

rectors are a practical possibility. The method of calibration described has been found to be very satisfactory in practice.

A superior form of vertical aperture will become feasible when a high-quality field store becomes available. It is somewhat surprising that the vertical resolution is in general sufficiently poor for the imperfect corrector described to be so effective even on a 405-line television system; it should be even more effective on a system using more lines per field.

PART II

A CONTINUOUSLY VARIABLE DELAY UNIT FOR TELEVISION SIGNALS

8. Introduction

There are many applications in the field of television for devices capable of delaying a picture by one or more television-line periods, and ultrasonic delay lines, which were developed principally for use in computer and radar applications, have been successfully used for some of them.^{1, 3, 4} Of the delay lines at present being manufactured, those useful for television purposes have either fused quartz or mercury as the transmission medium. Whilst fused quartz is a cheaper and more tractable material than mercury, and is slightly more suitable for delay-line applications, a line having mercury as the transmission medium can have the great advantage of being made continuously variable over a range of 10:1. Such a line, made by Mullard Research Laboratories, was chosen for use in the video delay unit; it produces a delay which is continuously variable between 25 μ s and 330 μ s. For some applications, it is necessary to have a range of delay continuously variable from zero, and in order to be able to simulate this, a second mercury delay line, which has similar electrical characteristics to the variable line but a fixed delay preset to a value of 30 μ s, is included. This, however, is not required for vertical aperture correction. Fig. 8 shows a general view of the complete delay unit.

9. The Mechanism and Construction of Mercury Delay Lines

Essentially, a mercury delay line consists of a bath of mercury with an electro-mechanical transducer at each end. A suitable electrical signal applied to one of these transducers causes it to vibrate mechanically and so set up a longitudinal pressure-wave in the mercury. This is transmitted to the second transducer with a velocity of approximately 1.5 mm/ μ s, producing across it a voltage representative of the signal applied to the first transducer but delayed in time. A delay of 98.8 μ s (one television-line period for a 405-line system) can be achieved with a line length of 15 cm (6 in.).

The performance of a mercury delay line, especially with

regard to bandwidth, insertion loss, and any secondary responses which may occur, is largely dependent upon three factors. These are:

- (a) the acoustical matching of the transducers to the transmission medium;
- (b) the efficiency with which the transducer converts electrical energy into mechanical energy and vice versa (usually termed the electro-mechanical coupling factor);
- (c) the electrical impedance presented to the external circuits by the transducers.

In order to obtain a large fractional bandwidth,* it is necessary for the transducers to be acoustically matched to the mercury. It is fortunate that quartz-crystal transducers approximately match mercury and may be designed to operate at frequencies up to at least 30 Mc/s, which makes bandwidths of the order of 10 Mc/s attainable.

The transducer usually consists of an X-cut crystal driven so as to vibrate in the thickness mode (see Fig. 9). The electro-mechanical coupling of these crystals is exceedingly poor and introduces a large loss; this is the major contribution to the insertion loss of the line.† The loss due to transmission through the mercury is relatively small, being about 6 dB for a path length of 40 cm (16 in.). The total insertion loss of the line depends to some extent upon the electrical network terminating it, but is usually in excess of 50 dB at the band-centre frequency.

With such a high insertion loss it is important to try to ensure that all the signal launched into the mercury travels to the receiver along the chosen transmission path. In practice the diameter of the crystals used as transducers is 100 wavelengths of the centre frequency, giving rise to a radiation pattern of quite high directivity. The first null in the radiation pattern occurs at an angle of 40' to the normal, and the maximum of the first sidelobe is at 55' with an amplitude 18 dB less than that of the main lobe.⁵

* Fractional bandwidth = bandwidth/band-centre frequency.

† Defined as: Insertion loss = 20 log₁₀ Input Voltage/Output Voltage, dB.

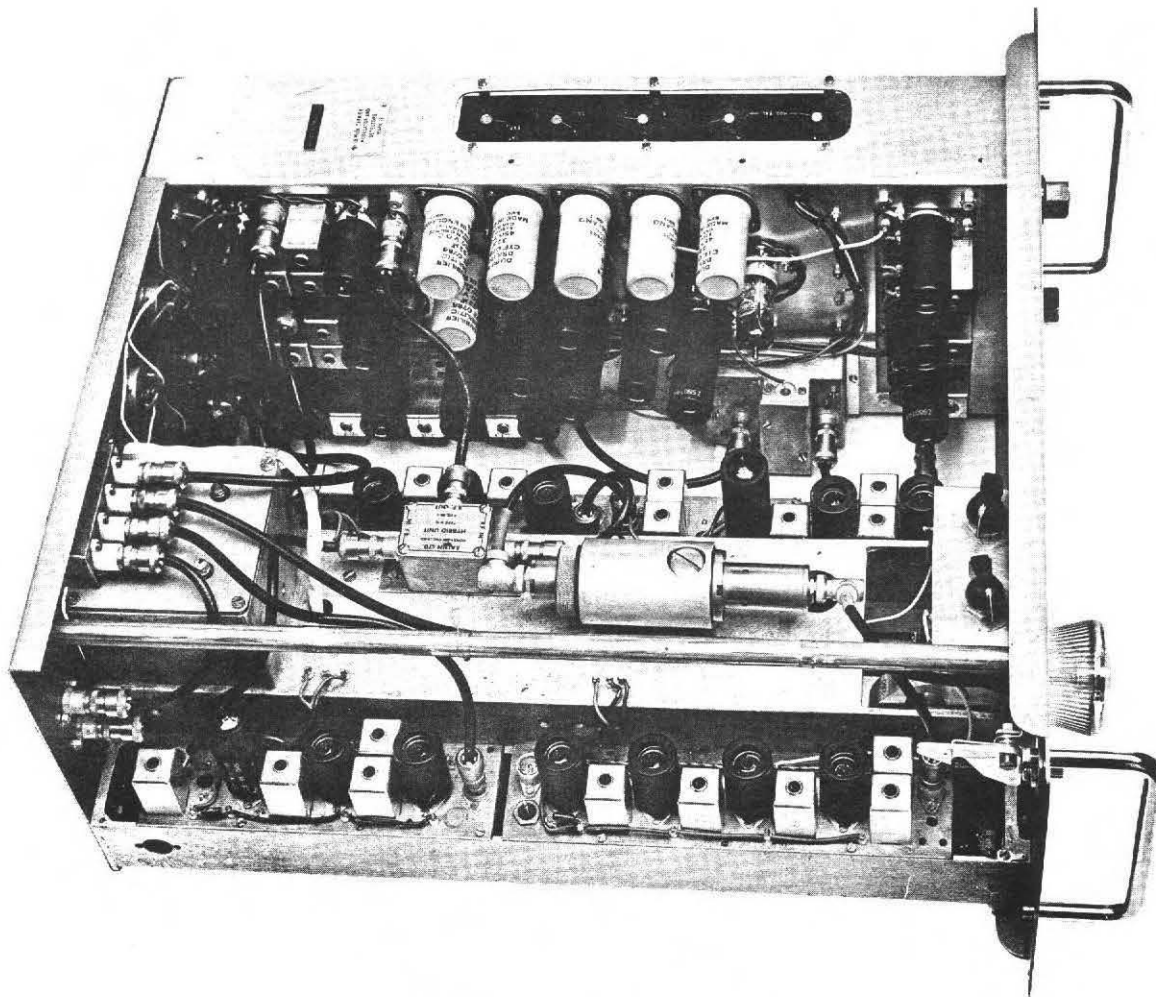


Fig. 8 — The video delay unit.

The signal radiated through the sidelobes strikes the walls of the container, usually a steel tank, and, because of the roughness of the walls, the wave becomes scattered; only a minute proportion finds its way to the receiving transducer to produce an unwanted signal. A far more troublesome form of secondary response is caused by the poor electro-mechanical coupling of each transducer; the mechanical energy not transformed into electrical energy by the receiving transducer must be re-radiated and, although a large proportion of this can be absorbed by an acoustical termination on the side of the crystal remote from the mercury, some must be re-radiated back along its original path. This energy will eventually reach the sending transducer where, after a second reflection, it will traverse the delaying medium for the third time to produce an unwanted 'third-time-round' response. This effect can be largely eliminated by tilting the receiving transducer so that the 'third-time-round' signal is incident at the angle of the first null in the radiation pattern. The angle of incidence of the main signal will then be equal to one-third of the 'null' angle, which will increase the insertion loss by

only 1.8 dB.⁵ By the use of this artifice, and careful design of the acoustical terminations of the line, the amplitude of the spurious responses can be restricted to at least 40 dB below that of the wanted signal.

The form of the attenuation/frequency characteristic of the line depends upon many factors. The centre frequency is largely determined by the transducers, the wavelength in quartz being approximately twice the crystal thickness. The bandwidth depends upon the centre frequency and is also affected by the load presented to each crystal by the transmission medium and the acoustical termination. In order to achieve maximum bandwidth, the acoustical termination can take the form of a mercury-filled cavity having a shape designed to trap and absorb the signal entering it. A simpler and more common method of termination is to provide each crystal with a lead backing. The form of the attenuation/frequency characteristic is also influenced by the orientation of the crystals. Owing to the fact that the radiation pattern of a crystal transducer varies with frequency, a change in the angle of incidence affects the attenuation/frequency characteristic of the line. When the

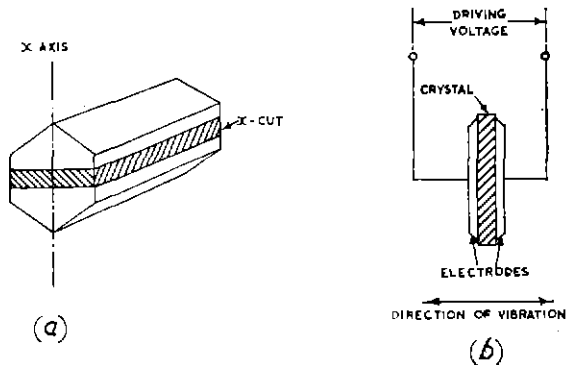


Fig. 9 — (a) Diagram of a section of a quartz crystal showing an X-cut.
 (b) An X-cut crystal operating in the thickness mode.

receiver transducer is adjusted so as to suppress the 'third-time-round' signal, the insertion loss at high frequencies is increased. Finally, as the attenuation of the acoustic wave is proportional to the square of the frequency, the transmission of the signal through the mercury has a very definite effect on the characteristic; this factor places an upper limit on the frequency which can be used with a given length of line.

Another principal factor determining the performance of the line is the electrical impedance presented to the external circuits by the transducers. Equivalent circuits for crystal transducers have been developed and described in the literature.^{5,6,7} In practice, the impedance seen across the terminals of a transducer reduces to a capacitance in parallel with a resistance; the latter, however, is of such a high value that for all practical purposes it may be ignored.

The two forms of line incorporated in the delay unit are described below.

9.1 Adjustable Line Type 4014

This is a simple type of line consisting of a mild-steel tank with a lead-backed quartz-crystal transducer suitably mounted at each end. The transmitting transducer may be moved axially over a short distance, thus providing a delay adjustment. The receiving transducer can be tilted to reduce the spurious responses. The attenuation/frequency characteristic of this line is shown in Fig. 10.

9.2 Continuously Variable Line Type 4016

The continuously variable line has a delay range of approximately 11:1. This is obtained by varying the path length of the acoustical wave as follows: both transducers are mounted at the same end of a rectangular mild steel tank. The wave radiated from the transmitting transducer enters a movable corner reflector, from which the signal is returned to the receiving transducer. The corner reflector is mounted on a steel lead-screw having a pitch such that one complete turn changes the delay by $10 \mu\text{s}$. The attenuation/frequency characteristic of this line changes with de-

lay and is shown in Fig. 11 for the maximum and minimum delay settings.

10. Description of the Video Delay Unit

10.1 General Description

The video delay unit shown in Fig. 8 consists of two separate delaying channels both driven from the same signal source. One channel uses the adjustable line to produce an output delayed by $30 \mu\text{s}$; the other channel uses the continuously variable line to produce an output whose delay may be adjusted between $25 \mu\text{s}$ and $330 \mu\text{s}$. Thus the second output can be delayed from $-5 \mu\text{s}$ to $+300 \mu\text{s}$ with respect to the first. The electrical characteristics of the lines are sufficiently similar for the design of the two channels to be identical, and the bandwidth available will accommodate, using a 15 Mc/s carrier, the amplitude-modulated

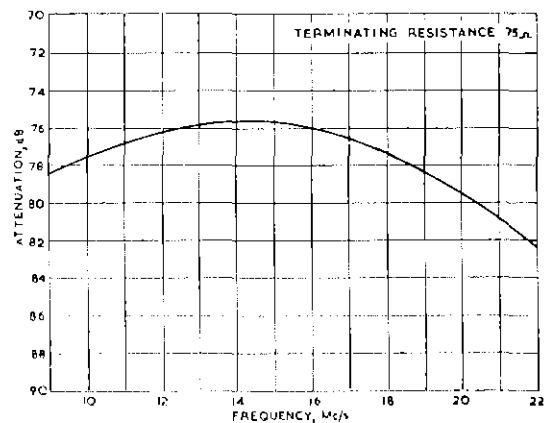


Fig. 10 — Attenuation/frequency characteristic of the adjustable mercury delay line type MS4014.

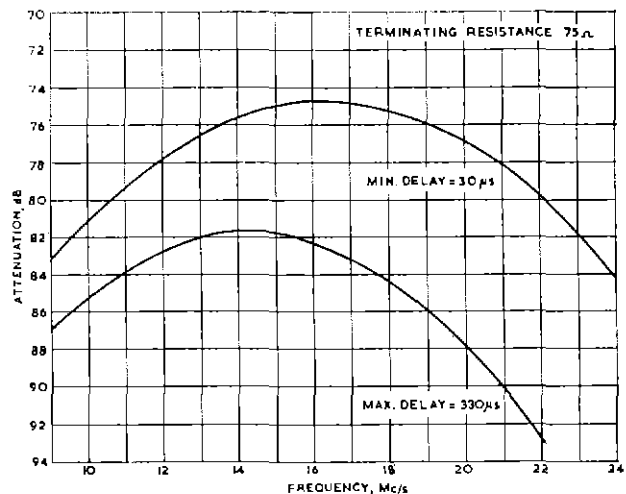


Fig. 11 — Attenuation/frequency characteristics of the continuously variable mercury delay line type MS4016.

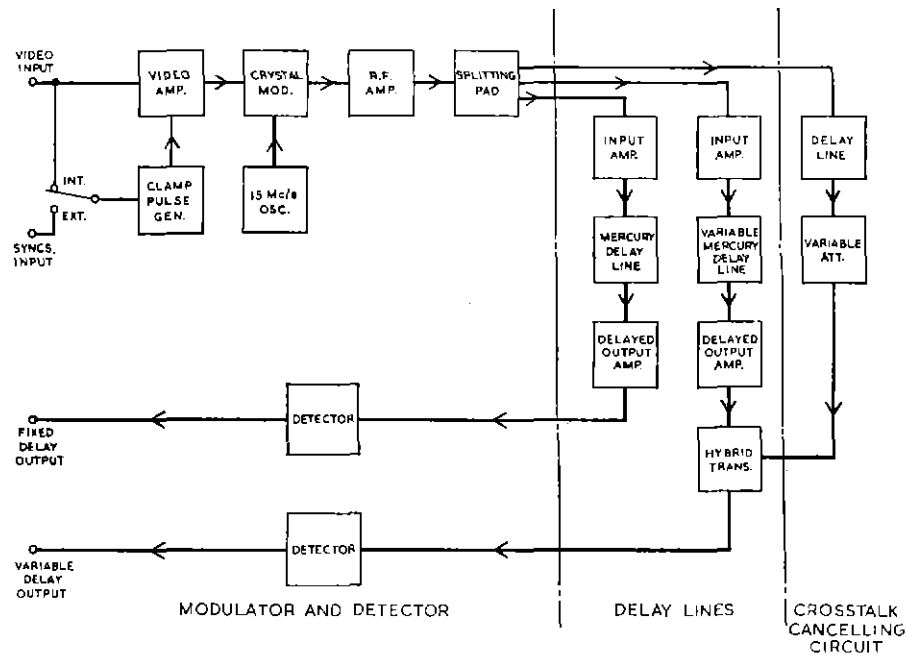


Fig. 12 — Block diagram of the video delay unit.

double-sideband transmission of a 405-line television picture.

The block diagram of the complete unit is shown in Fig. 12; the subdivisions indicate the three sub-units of which it is constructed. The video signal is modulated on a locally generated carrier and the resulting signal is amplified and split into three separate outputs. Two of these outputs are used to drive the principal delaying channels, each of which consists of an input amplifier, a mercury delay line, and an output amplifier. The third output is delayed, attenuated, and added back to the output of the continuously variable channel, in order to cancel out some direct crosstalk which occurs between the transducers. The amplitude of this crosstalk, which occurs inside the line as a result of its construction, is approximately 40 dB below that of the wanted signal; the use of the cancelling circuit reduces the crosstalk by a further 20 dB. Crosstalk is also liable to occur outside the line because of coupling between the networks connecting the transducers to the amplifiers. This has been reduced to negligible proportions by thick copper screens and the careful choice of earth connections. The delayed signals from the output amplifiers are detected by means of germanium diodes and fed through cathode followers to the output terminals.

10.2 Generation of the Modulated Signal

Fig. 13 is a diagram of the circuits used to produce the modulated signal. A 'clamped' video amplifier of conventional design is used to drive a crystal modulator. The 15 Mc/s carrier is generated by a crystal-controlled oscillator and fed to the modulator through a buffer amplifier. In order to facilitate adjustment of the modulator, the

carrier can be 100 per cent modulated by a 2.5 kc/s square wave locked to the television line-synchronizing pulses. The signal from the modulator is raised in level by an amplifier similar in design to that described in Section 10.4.

10.3 Input Amplifier

The circuit diagram of the input amplifier is shown in Fig. 14; it consists of a buffer amplifier, which permits the input line to be correctly terminated, followed by two further stages incorporating negative feedback. In order to obtain a high signal-to-noise ratio the input transducer must be driven with as large a signal as possible. For this reason, the output valve must handle large current swings, and negative feedback is used in order to reduce the non-linear distortion which would otherwise result.

The network coupling the anode of the output valve to the transducer fulfils two purposes: it transforms the signal current from the output valve into a suitable voltage across the transducer, and at the same time is used to correct the attenuation/frequency characteristic of the mercury delay line.

The design of this network was complicated by the fact that the characteristic of the variable line changes in centre frequency and bandwidth as the delay is varied (see Fig. 11). It was decided to equalize the line for the mid-setting of delay; the resulting overall attenuation/frequency characteristics at maximum and minimum delay are shown in Fig. 15. These characteristics have an approximately constant slope of 0.3 dB per Mc/s within the passband, but this will have negligible effect on a 405-line television waveform if a double-side-band amplitude modulator is used together with an envelope detector. The attenuation/

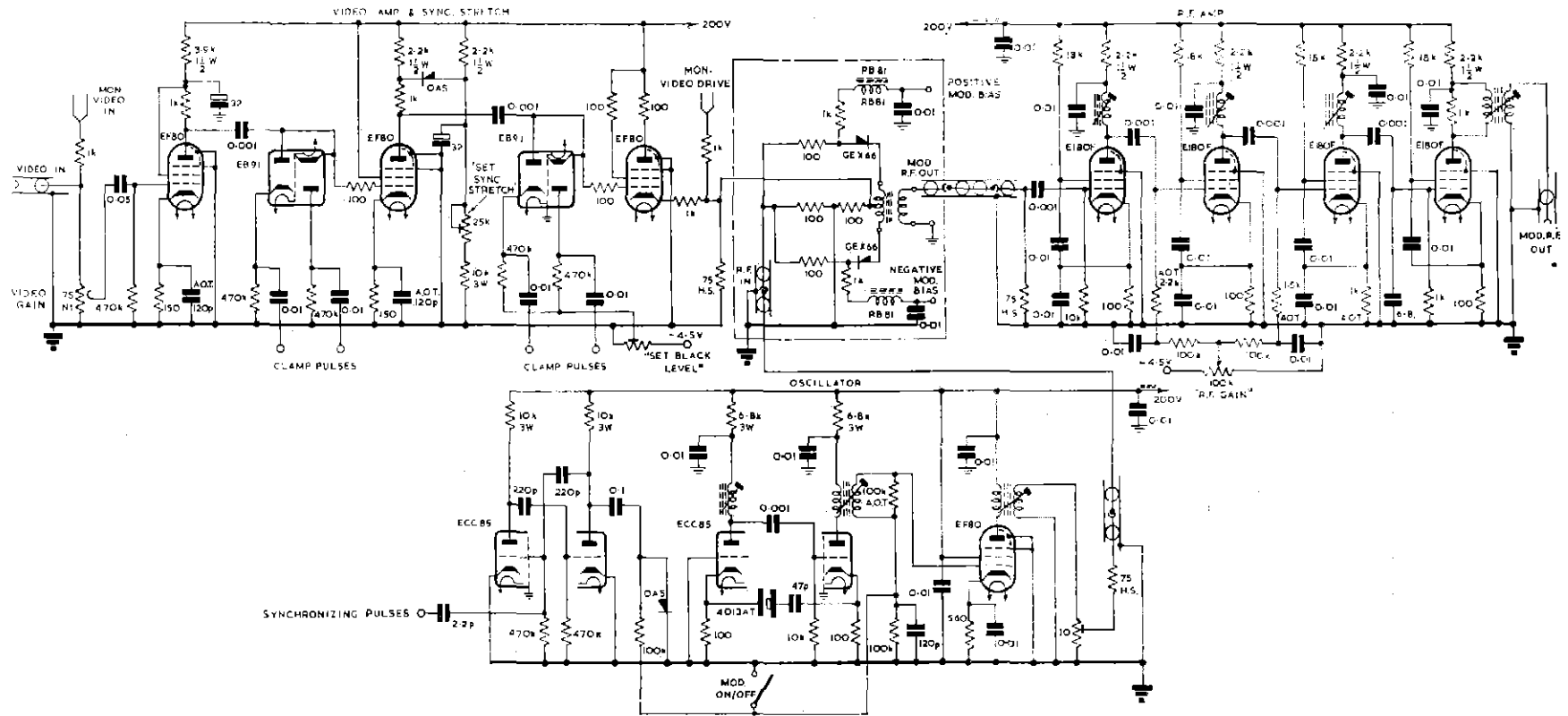


Fig. 13 — Circuit diagram of the video amplifier, oscillator, modulator, and r.f. amplifier in the Modulator and Detector Sub-unit.

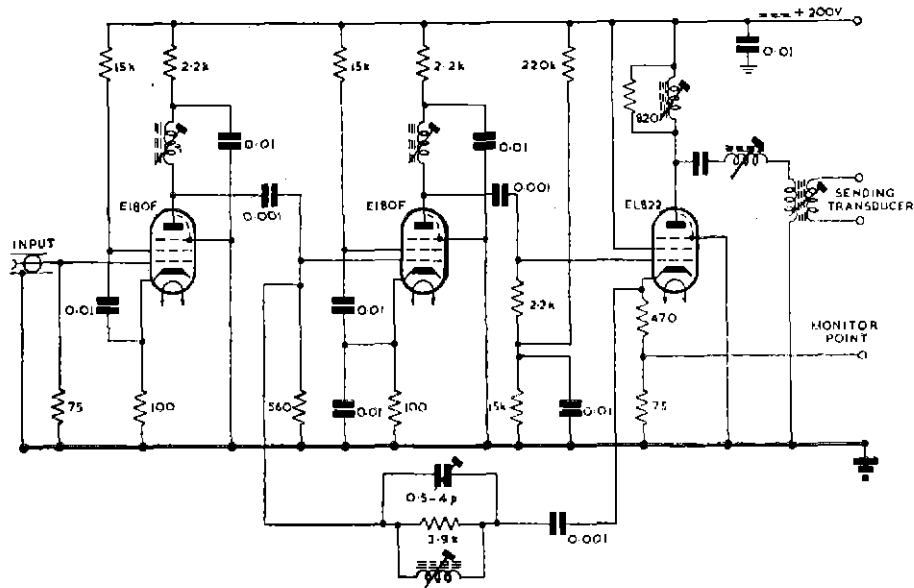


Fig. 14 — Circuit diagram of the input amplifier.

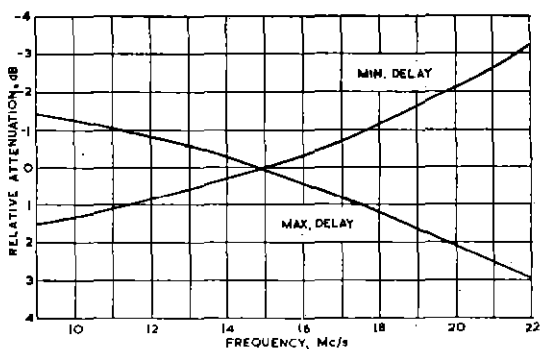


Fig. 15 — Attenuation/frequency characteristics of the continuously variable line when equalized for the mid-setting of delay.

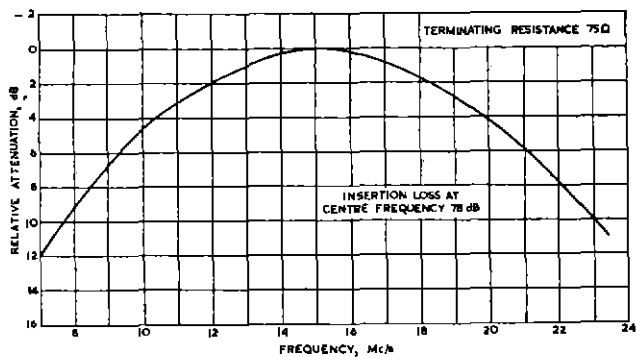


Fig. 16 — Attenuation/frequency characteristic of the continuously variable line at mid-setting of delay.

frequency characteristic of the line at the mid-setting of delay is shown in Fig. 16. In order to design an equalizer for the line, it is convenient to transform the bandpass response into an equivalent low-pass response and find a simple expression to represent this. A suitable expression is given by

$$|A(f)| = [1 + 0.94(f/10)^2 + 1.36(f/10)^4]^{\frac{1}{2}}$$

where $A(f)$ is the complex attenuation of the line and f is the equivalent low-pass frequency (Mc/s).

The simplest form of low-pass circuit which can be used as a coupling network in these circumstances is a π -section as shown in Fig. 17. The attenuation of this network is of the form

$$|B(f)| = [1 + b_1(f/10)^2 + b_2(f/10)^4 + b_3(f/10)^6]^{\frac{1}{2}}$$

By multiplying together $|A(f)|$ and $|B(f)|$ and reducing the appropriate coefficients to zero, a fourth-order maximally flat function is obtained. This process results in a set of equations relating the coefficients of $|B(f)|$ with

* For the purposes of calculation, this expression is most conveniently stated in terms of megacycles divided by ten ($f/10$).

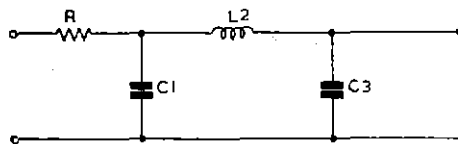


Fig. 17 — Circuit diagram of the low-pass equalizing section.

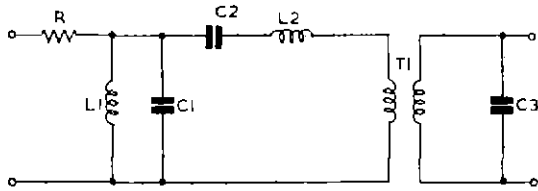


Fig. 18—Circuit diagram of the band-pass equalizing section.

those of $|A(f)|$. Thus b_1 , b_2 , and b_3 may be determined, hence we obtain the expression

$$|B(f)| = [1 + 0.94(f/10)^2 - 0.48(f/10)^4 + 1.73(f/10)^6]^{1/2}$$

By comparing this expression with the transfer function of the low-pass network, equations may be derived relating the network parameters to b_1 , b_2 , and b_3 . These equations may be used to determine the circuit component values in terms of the capacitance presented by the output valve.

The low-pass network can now be transformed into the bandpass network of Fig. 18. The transformer T1 is necessary in order to transform the transducer capacitance to a value suitable for inclusion in the coupling network; the transformer need not be perfect, for any leakage inductance can be compensated by adjustment of the series inductance L2. The characteristic of the bandpass network is shown in Fig. 19 and the combined characteristic of the coupling network and the line (at the mid-setting of delay) is shown in Fig. 20.

10.4 Output Amplifier

The circuit of the output amplifier is shown in Fig. 21; it consists of a stagger-tuned triple followed by an output stage. The amplifier has a maximum gain of approximately 60 dB which can be reduced by adjusting the negative bias applied to the grids of the second and third valves. The

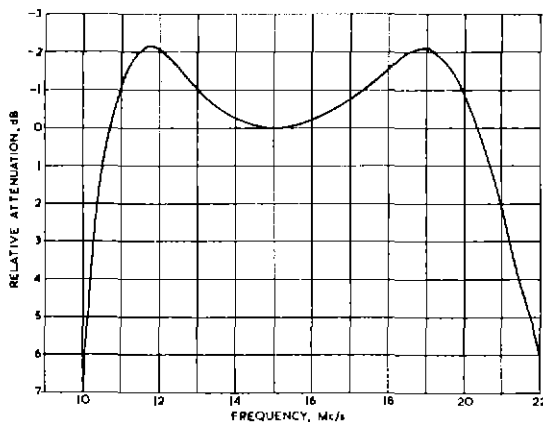


Fig. 19—Attenuation/frequency characteristic of the band-pass equalizing section.

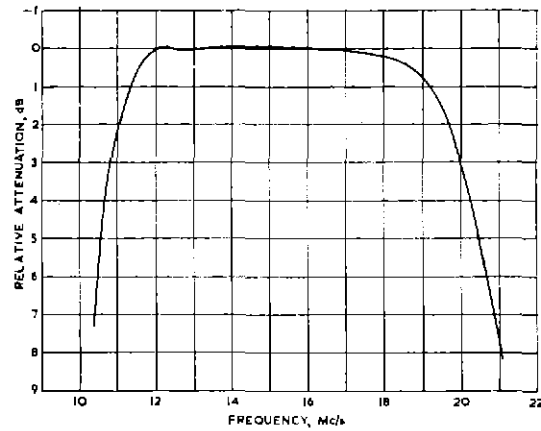


Fig. 20—Attenuation/frequency characteristic of the equalized continuously variable line for the mid-setting of delay.

output stage is designed to have unity gain and an output impedance of 75 Ω within the passband.

The signal-to-noise ratio of the video delay unit is largely determined by the noise contributed by the first valve of this amplifier; however, the amplitude of the input signal is sufficient to allow a pentode to be used.

10.5 Detector and Video Amplifier

The circuit of the detector and video amplifier is shown in Fig. 22. It consists of an amplifying stage coupled, by means of a bandpass filter, to a diode detector; the output of the detector is fed through a low-pass filter to a simple cathode-follower output stage. As the lowest sideband frequency of the modulated signal is only two octaves removed from the highest video frequency it is desirable that the low-pass filter have a rate of cut-off of at least 24 dB per octave; furthermore, the filter must not introduce any significant group-delay distortion within the video passband. These requirements are achieved by using a filter with a cut-off frequency of twice the highest video frequency and maximum attenuation at the carrier frequency; its design is described in the Appendix.

11. Performance

The degradation of a video signal by the delay unit may be judged from the reproductions of Test Card 'C' shown in Figs. 23, 24, and 25. Fig. 23 shows the input signal, whilst Figs. 24 and 25 show the signal delayed by 150 μ s and 300 μ s respectively. The residual crosstalk discussed in Section 10.1 can be observed in Fig. 24; it is more pronounced in Fig. 25 owing to the increased insertion loss of the line. The ratio of peak-to-peak video signal (excluding synchronizing pulses) to r.m.s. noise is dependent upon the delay setting. It varies from 38 dB at maximum delay to 44 dB at minimum delay.

Although the experimental unit described in this monograph was designed for a 405-line television system, the

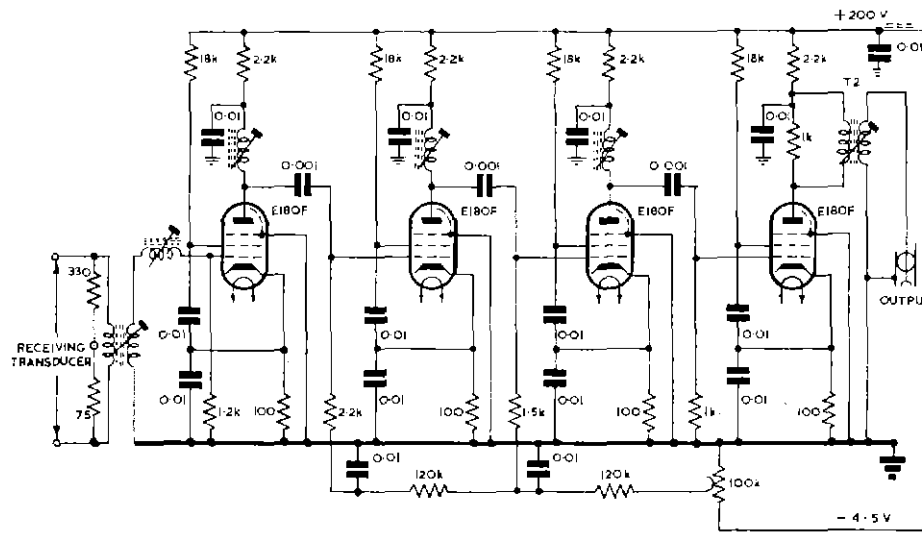


Fig. 21 — Circuit diagram of the output amplifier.

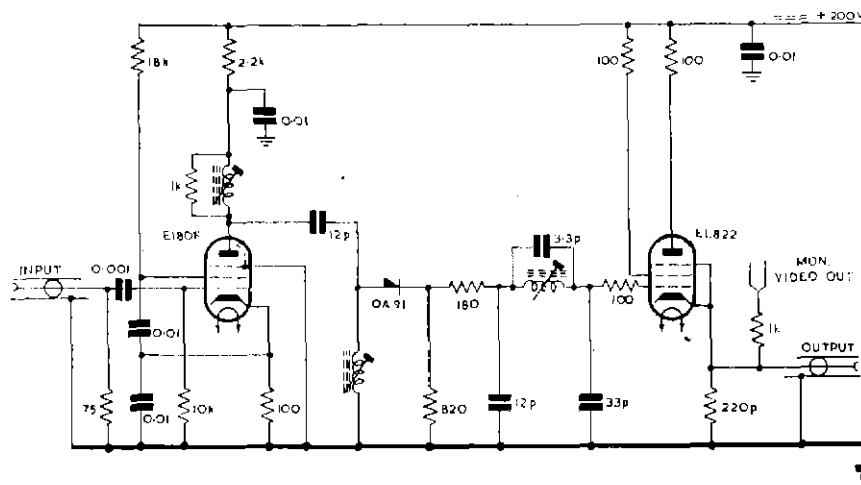


Fig. 22 — Circuit diagram of the detector and video amplifier.

equalized delay line has sufficient bandwidth for a 625-line system if a single-sideband system of modulation is used.* In this case, the change of slope of the attenuation/frequency characteristic with delay could no longer be tolerated, and it would become necessary to introduce a variable equalizer (preferably ganged to the delay control). When using the unit with a television system locked to mains frequency, it is necessary to adjust the delay at frequent intervals in order to compensate for changes in line-scan period. An improved version of the unit would incorporate a servo-mechanism, controlled by the television line-timing pulses, moving one of the transducers so as to

* It may be possible to extend the use of the line to an 819-line system by means of a more complex equalizing network.

maintain a delay of one line-scan period or a constant proportion thereof.

12. Conclusions

The mercury delay line is eminently suitable for the vertical aperture corrector described in this monograph, and could also be used in other television applications requiring delays up to a few line-scan periods, such as bandwidth-compression experiments or synchronous standards conversion. If the line were to be included as an integral part of a television system which was locked to the mains frequency, the problem of varying the delay of the line according to the mains frequency would first have to be solved.

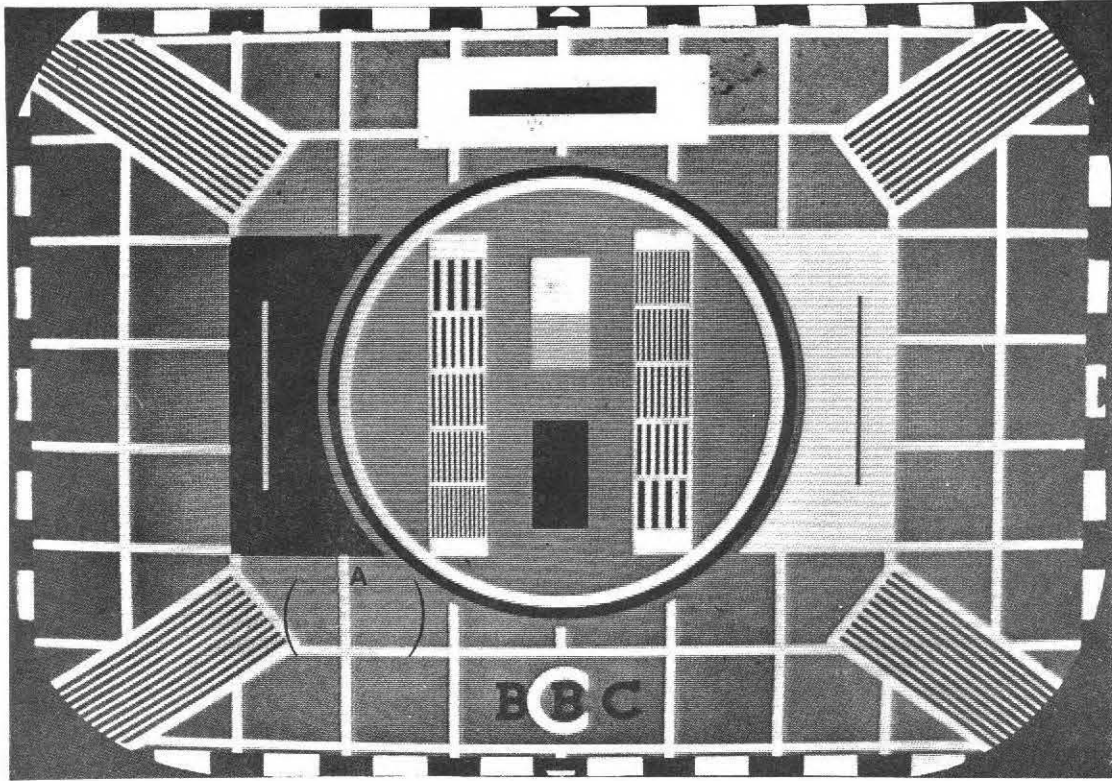


Fig. 23 — Picture input to the video delay unit.
 A — An example of moiré patterning due to printing process.

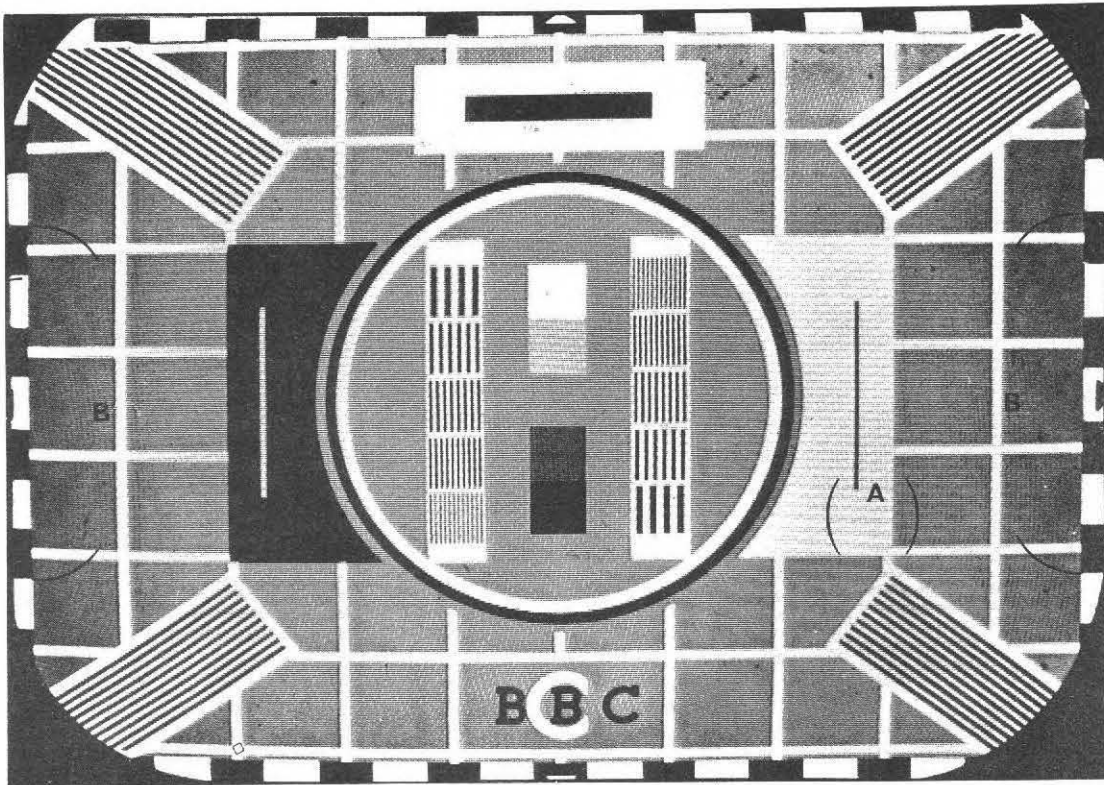


Fig. 24 — Picture output from the video delay unit after a delay of 150 μ s.
 A — An example of moiré patterning due to printing process. B — An example of crosstalk.

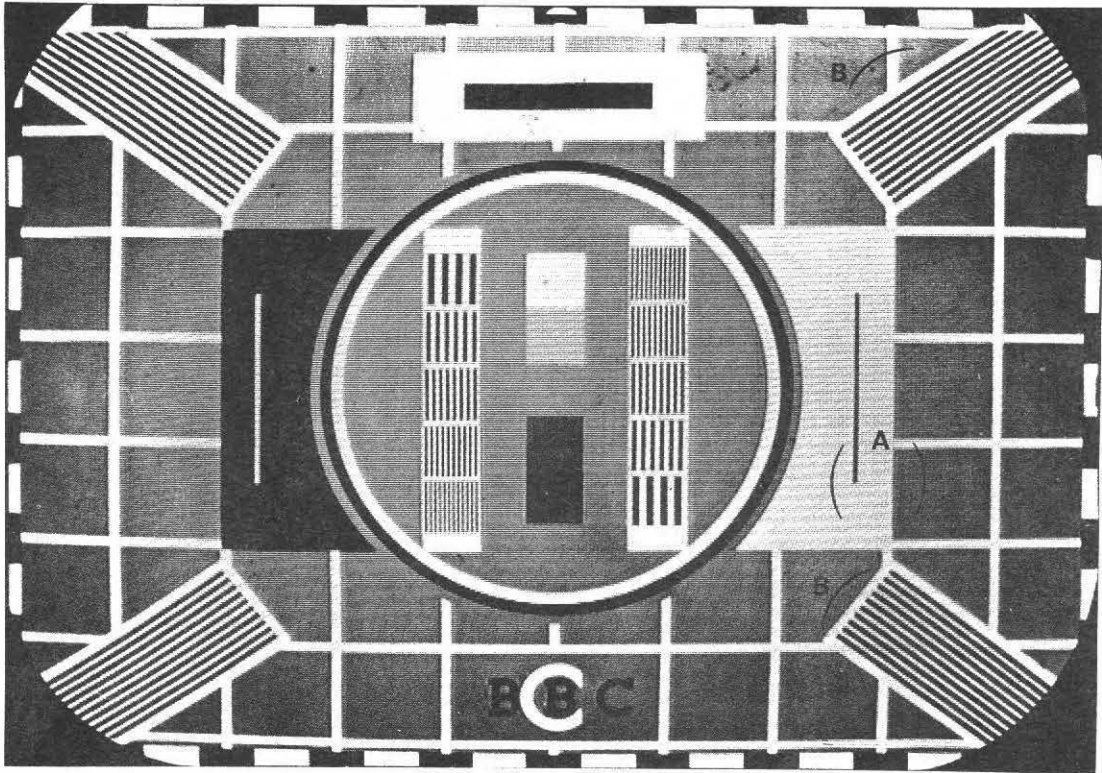


Fig. 25 — Picture output from the video delay unit after a delay of 300 μ s.
 A — An example of moiré patterning due to printing process.
 B — An example of crosstalk.

13. Acknowledgment

The author would like to acknowledge the assistance of Mr K. J. Wright in the design and construction of the signal processing unit of the vertical aperture corrector.

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APPENDIX

DESIGN OF A DETECTOR FILTER

The circuit of the filter section used in the detector is shown in Fig. 26. The attenuation function of such a network has a pole-zero configuration as shown in Fig. 27 and is of the form

$$A(p) = \frac{1 + a_1 p + a_2 p^2 + a_3 p^3}{1 + b_2 p^2}$$

The condition for such a function to have a maximally flat frequency response may be found by multiplying the function by $A(-p)$ and equating, as far as possible, the coefficients of equal powers of p in the numerator and denominator. Hence

$$A(p) \cdot A(-p) = \frac{1 + (2a_2 - a_1^2)p^2 + (a_2^2 - 2a_1 a_3)p^4 - a_3^2 p^6}{1 + 2b_2 p^2 + b_2^2 p^4}$$

and the conditions for a third-order maximally flat response are

$$2a_2 - a_1^2 = 2b_2 \tag{9}$$

$$a_2^2 - 2a_1 a_3 = b_2^2 \tag{10}$$

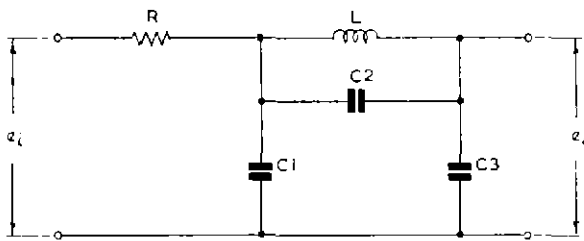


Fig. 26 — Circuit diagram of the detector filter.

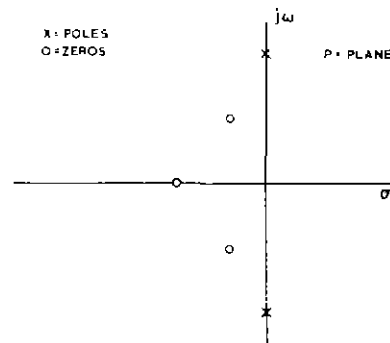


Fig. 27 — Poles and zeros of the attenuation function of the detector filter.

In order to specify the performance of the network, the coefficients (a_1, a_2 , etc.) may be stated in terms of a cut-off frequency (ω_0) and a parameter 'm' where

$$m = \left[1 - \left(\frac{\omega_0}{\omega_\infty} \right)^2 \right]^{\frac{1}{2}}$$

and ω_∞ is the frequency at which there is maximum attenuation. It is apparent from Fig. 27 that the frequency of maximum attenuation is the point at which the poles occur on the imaginary axis, hence

$$b_2 = \frac{1}{\omega_\infty^2} \tag{11}$$

The value of the cut-off frequency may be defined as

$$\omega_0^2 = \frac{a_1}{2a_3} \tag{12}$$

This definition of ω_0 was chosen in order to make the normalized transfer function of the network identical to that of a similar conventional 'm-derived' filter section having the same values of m and ω_0 .

Solving equations (9), (10), (11), and (12), for the required coefficients we obtain the expression

$$A(p) = \frac{1 + \frac{2m}{\omega_0} p + \frac{1 + m^2}{\omega_0^2} p^2 + \frac{m}{\omega_0^3} p^3}{1 + \frac{1 - m^2}{\omega_0^2} p^2} \tag{13}$$

An analysis of the circuit of Fig. 26 leads to the expression

$$A(p) = \frac{1 + R(C_1 + C_3)p + L(C_2 + C_3)p^2 + RL(C_1C_2 + C_2C_3 + C_3C_1)p^3}{1 + LC_2p^2}$$

The values of the circuit elements may now be determined by equating the coefficients of this expression to those of expression (13), and solving the equations so set up.

The solutions of these equations are tabulated in Table 1; the values of the circuit components for the conventional 'm-derived' filter are included for comparison.

TABLE 1

<i>Circuit Component</i>	<i>Detector Filter</i>	<i>'m-derived' Filter</i>
R	R	R
C_1	$\frac{1}{\omega_0 R} \frac{2m^2 - 1}{2m}$	$\frac{m}{\omega_0 R}$
C_2	Chosen to resonate with L at f_∞	
C_3	$\frac{1}{\omega_0 R} \frac{2m^2 + 1}{2m}$	$\frac{m}{\omega_0 R}$
L	$\frac{R}{\omega_0} \frac{4m^3}{2m^2 + 1}$	$2m \frac{R}{\omega_0}$
Terminating Resistor	∞	R

**CORRECTION TO ENGINEERING MONOGRAPH NO. 46
(THE APPLICATION OF TRANSISTORS TO SOUND BROADCASTING)**

The last two lines of the caption to Fig. 3 (page 9) should be amended to read:

Transistor (a) $i_{co} = 9.5 \mu\text{A}$, $h_{fe} = 103$

Transistor (b) $i_{co} = 18 \mu\text{A}$, $h_{fe} = 88$

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