# BBC
## ENGINEERING DIVISION
# MONOGRAPH

NUMBER 63: **AUGUST 1966**

# The design of transmission lines and single-stage switching circuits for a line-store standards converter

PART I

SWITCHING AND STORAGE CIRCUITS

by

E. R. ROUT, A.M.I.E.E.

and

W. K. E. GEDDES, M.A.(Cantab.), Grad.I.E.E.

(Research Department, BBC Engineering Division)

PART II

THE DESIGN OF VIDEO TRANSMISSION LINES

by

A. H. JONES, B.Sc., Grad.I.E.E.

(Research Department, BBC Engineering Division)

BRITISH BROADCASTING CORPORATION

PRICE FIVE SHILLINGS

BBC ENGINEERING MONOGRAPH

No. 63

# THE DESIGN OF TRANSMISSION LINES AND SINGLE-STAGE SWITCHING CIRCUITS FOR A LINE-STORE STANDARDS CONVERTER

## Part I

## Switching and Storage Circuits

by

E. R. Rout, A.M.I.E.E.

and

W. K. E. Geddes, M.A.(Cantab.), Grad.I.E.E.

(Research Department, BBC Engineering Division)

## Part II

## The Design of Video Transmission Lines

by

A. H. Jones, B.Sc., Grad.I.E.E.

(Research Department, BBC Engineering Division)

AUGUST 1966

BRITISH BROADCASTING CORPORATION

# PREVIOUS ISSUES IN THIS SERIES

| No. | Title | Date |
|---|---|---|
| 1. | *The Suppressed Frame System of Telerecording* | JUNE 1955 |
| 2. | *Absolute Measurements in Magnetic Recording* | SEPTEMBER 1955 |
| 3. | *The Visibility of Noise in Television* | OCTOBER 1955 |
| 4. | *The Design of a Ribbon Type Pressure-gradient Microphone for Broadcast Transmission* | DECEMBER 1955 |
| 5. | *Reproducing Equipment for Fine-groove Records* | FEBRUARY 1956 |
| 6. | *A VHF/UHF Field-strength Recording Receiver using Post-detector Selectivity* | APRIL 1956 |
| 7. | *The Design of a High Quality Commentator's Microphone Insensitive to Ambient Noise* | JUNE 1956 |
| 8. | *An Automatic Integrator for Determining the Mean Spherical Response of Loudspeakers and Microphones* | AUGUST 1956 |
| 9. | *The Application of Phase-coherent Detection and Correlation Methods to Room Acoustics* | NOVEMBER 1956 |
| 10. | *An Automatic System for Synchronizing Sound on Quarter-inch Magnetic Tape with Action on 35-mm Cinematograph Film* | JANUARY 1957 |
| 11. | *Engineering Training in the BBC* | MARCH 1957 |
| 12. | *An Improved 'Roving Eye'* | APRIL 1957 |
| 13. | *The BBC Riverside Television Studios: The Architectural Aspects* | JULY 1957 |
| 14. | *The BBC Riverside Television Studios: Some Aspects of Technical Planning and Equipment* | OCTOBER 1957 |
| 15. | *New Equipment and Methods for the Evaluation of the Performance of Lenses of Television* | DECEMBER 1957 |
| 16. | *Analysis and Measurement of Programme Levels* | MARCH 1958 |
| 17. | *The Design of a Linear Phase-shift Low-pass Filter* | APRIL 1958 |
| 18. | *The BBC Colour Television Tests: An Appraisal of Results* | MAY 1958 |
| 19. | *A UHF Television Link for Outside Broadcasts* | JUNE 1958 |
| 20. | *The BBC's Mark II Mobile Studio and Control Room for the Sound Broadcasting Service* | AUGUST 1958 |
| 21. | *Two New BBC Transparencies for Testing Television Camera Channels (Out of Print)* | NOVEMBER 1958 |
| 22. | *The Engineering Facilities of the BBC Monitoring Service* | JANUARY 1959 |
| 23. | *The Crystal Palace Band I Television Transmitting Aerial* | FEBRUARY 1959 |
| 24. | *The Measurement of Random Noise in the presence of a Television Signal* | MARCH 1959 |
| 25. | *A Quality-checking Receiver for VHF/FM Sound Broadcasting* | JUNE 1959 |
| 26. | *Transistor Amplifiers for Sound Broadcasting* | AUGUST 1959 |
| 27. | *The Equipment of the BBC Television Film Studios at Ealing* | JANUARY 1960 |
| 28. | *Programme Switching, Control, and Monitoring in Sound Broadcasting* | FEBRUARY 1960 |
| 29. | *A Summary of the Present Position of Stereophonic Broadcasting* | APRIL 1960 |
| 30. | *Film Processing and After-processing Treatment of 16-mm Films* | MAY 1960 |
| 31. | *The Power Gain of Multi-tiered VHF Transmitting Aerials* | JULY 1960 |
| 32. | *A New Survey of the BBC Experimental Colour Transmissions* | OCTOBER 1960 |
| 33. | *Sensitometric Control in Film Making* | DECEMBER 1960 |
| 34. | *A Mobile Laboratory for UHF and VHF Television Surveys* | FEBRUARY 1961 |
| 35. | *Tables of Horizontal Radiation Patterns of Dipoles Mounted on Cylinders* | FEBRUARY 1961 |
| 36. | *Some Aspects of Optical Lens Performance* | APRIL 1961 |
| 37. | *An Instrument for Measuring Television Signal-to-noise Ratio* | JUNE 1961 |
| 38. | *Operational Research on Microphone and Studio Techniques in Stereophony* | SEPTEMBER 1961 |
| 39. | *Twenty-five Years of BBC Television* | OCTOBER 1961 |
| 40. | *The Broadcasting of Music in Television* | FEBRUARY 1962 |
| 41. | *The Design of a Group of Plug-in Television Studio Amplifiers* | APRIL 1962 |
| 42. | *Apparatus for Television and Sound Relay Stations* | JULY 1962 |
| 43. | *Propagational Factors in Short-wave Broadcasting* | AUGUST 1962 |
| 44. | *A Band V Signal-frequency Unit and a Correlation Detector for a VHF/UHF Field-strength Recording Receiver* | OCTOBER 1962 |
| 45. | *Vertical Resolution and Line Broadening* | DECEMBER 1962 |
| 46. | *The Application of Transistors to Sound Broadcasting* | FEBRUARY 1963 |
| 47. | *Vertical Aperture Correction using Continuously Variable Ultrasonic Delay Lines* | MAY 1963 |
| 48. | *The Development of BBC Internal Telecommunications* | MAY 1963 |
| 49. | *Apparatus for Measurement of Non-linear Distortion as a Continuous Function of Frequency* | JULY 1963 |
| 50. | *New Methods of Lens Testing and Measurement* | SEPTEMBER 1963 |
| 51. | *Radiophonics in the BBC* | NOVEMBER 1963 |
| 52. | *Stereophony: the effect of cross-talk between left and right channels* | MARCH 1964 |
| 53. | *Aerial distribution systems for receiving stations in the l.f., m.f., and h.f. bands* | JULY 1964 |
| 54. | *An Analysis of Film Granularity in Television Reproduction* | AUGUST 1964 |
| 55. | *A Review of Television Standards Conversion* | DECEMBER 1964 |
| 56. | *Stereophony: the effect of interchannel differences in the phase/frequency and amplitude/frequency characteristics* | DECEMBER 1964 |
| 57. | *Drop-out in Video-tape Recording* | JUNE 1965 |
| 58. | *Sine-squared pulse and bar testing in colour television* | AUGUST 1965 |
| 59. | *The acoustic design and performance of a new free-field sound measurement room* | SEPTEMBER 1965 |
| 60. | *Colorimetric Analysis of Interference in Colour Television* | FEBRUARY 1966 |
| 61. | *Sporadic E Ionization and Television Interference* | FEBRUARY 1966 |
| 62. | *Automatic monitoring* | APRIL 1966 |

# THE DESIGN OF TRANSMISSION LINES AND SINGLE-STAGE SWITCHING CIRCUITS FOR A LINE-STORE STANDARDS CONVERTER


## SUMMARY

Part I of this monograph gives an account of the development of switching and storage circuits for line-store standards converters in which each storage element is associated with individual high-speed writing and reading switches, and in which a separate interpolation system is used.

Part II discusses requirements affecting the design of transmission lines suitable for the distribution of video signals within the store assembly in this form of converter. The development of two types of transmission line which meet the requirements defined is described.


## 1. Introduction to Parts I and II

The basic constituents of a line-store standards converter[1,2] are:

(*a*) A set of identical storage elements into which picture information is 'written' at a rate depending on the standard of the incoming signal and from which it is 'read' at a rate depending on the standard of the outgoing signal.

(*b*) A system of interpolation which makes allowance for the fact that each line of the converted picture generally lies between two adjacent lines of the original picture.

The storage elements are connected in sequence to input and output circuits by a number of high-speed switches. Switching may be carried out in a single stage, using individual pairs of high-speed electronic writing and reading switches associated with each of the storage elements, or alternatively by a two-stage switching arrangement in which a limited number of high-speed switches work in conjunction with a larger number of low-speed switches. For reasons which have been set out in a recent I.E.E. paper,[2] the BBC Designs and Research Departments concurrently undertook the development of operational converters using each of these two approaches. Designs Department adopted two-stage switching, and achieved interpolation by designing each storage element as a simple low-pass filter, while the Research Department project used single-stage switching, with a separate interpolation system using a number of line-period delay units.

This monograph discusses the development of some of the basic components of the Research Department converter. In the switching and storage system described in Part I the storage elements are fixed capacitors; each is mounted, together with its associated input and output switches, on a separate printed-circuit card, the arrangement being termed a 'store unit'. It is therefore necessary to make input and output video connections to each of the 546 or more store units that are required for conversion between the 405- and 625-line standards.[2] As a result, the common signal connections extend over a distance of several feet, and signal delays within the interconnection system are comparable with the time taken by a switching operation. Thus it is necessary to design the interconnection system with care in order to prevent the introduction of interfering delayed signals due to reflexions. Moreover, it is necessary to isolate the input and output connections from each other in order to limit direct crosstalk between them to an acceptable level.

Part II of the monograph outlines the requirements which influence the design of signal connections in a converter of the above type and describes the development of a transmission-line system that has proved satisfactory in practice. In this line-store converter, transmission lines were used not only for conveying signals to and from the store units but also for connection of the clock pulses which caused the store-unit switches to operate. Except where otherwise stated the development of the clock-pulse lines closely followed that of the video-signal lines.

# PART I
# SWITCHING AND STORAGE CIRCUITS

## 2. Fundamental Processes of Line-store Standards-conversion

The storage cells of a line-store converter must be associated with two switching systems; a writing system must direct each picture element of the incoming signal into its correct storage cell and a reading system must subsequently collect the stored picture elements, in the right order, at the times they are required by the outgoing standard.

One means of realizing these systems consists of combining each storage element with a writing switch and a reading switch, as shown in Fig. 1. A common connection to all the writing switches carries the input signal, and the output signal is generated in a circuit common to all the reading switches. It is convenient to give a name to the group of components that must be uniquely associated with every one of the picture elements along each line of the picture, and such a group will be referred to as a 'store

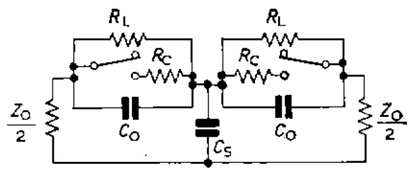


*Fig. 3 — Simplified equivalent circuit of the store-unit switches*

fundamental parameters of the store unit to be specified, it will be assumed initially that a practical switch can be represented, when closed, by a small series resistance $R_c$ and, when open, by a parallel combination of leakage resistance $R_l$ and shunt capacitance $C_o$. The basic configuration of the signal path in a complete store unit can then be depicted as in Fig. 3, which shows two switches connecting a storage capacitor $C_s$ to the input and output video transmission lines, respectively; the lines are assumed to have identical characteristic impedances $Z_o$ and to be correctly terminated at each end, thus presenting impedances of $\frac{1}{2}Z_o$ at both the input and the output of each store unit. In order to establish those values of the parameters shown in the figure that are consistent with satisfactory operation of the store unit, it is necessary to define three basic requirements of performance and make use of the results of subjective tests concerning the threshold visibility of relevant forms of picture impairment.

During each closure of the appropriate switch, the storage capacitor should be charged or discharged substantially to equilibrium. This condition renders the performance of the store unit almost independent of small variations in the charging or discharging time-constant and in the duration of the switching pulse. In order to accommodate the largest variations likely to occur, the store capacitor should be charged to within 1 per cent of the voltage existing on the 'writing' video transmission line; similarly, at least 99 per cent of the charge on the storage capacitor should subsequently be transferred to the 'reading' transmission line. If a closure time of 100 ns is assumed,* these conditions impose a maximum value of 20 ns on the time constants of both the writing and reading circuits. Denoting the total resistance, $R_c + \frac{1}{2}Z_o$, by $R_t$, we therefore have:

$$C_sR_t < 2 \times 10^{-8} \text{ seconds}$$

Once charged, the storage capacitor must not be appreciably discharged by leakage during the storage period, because the magnitude of the discharge due to leakage depends on the storage time, which is not constant. The maximum tolerable rate of discharge is 1 per cent in 100 $\mu$s. Since both switches represent leakage paths, this condition imposes the restriction:

$$C_sR_l > 2 \times 10^{-2} \text{ seconds}$$

* The duration of closure must, in fact, be slightly less than 100 ns for the switches operated by 625-line clock pulses. This ensures that the disturbance imposed on the transmission line by operation of each switch has passed beyond the next tapping point (towards eventual absorption in the terminating resistance) before the next switch operates.

The stray capacitances, $C_o$, must not couple the two transmission lines sufficiently to cause perceptible contamination of the output signal by unconverted input signal. This requirement imposes the restriction:

$$100C_o < C_s$$

Thus we have three basic inequalities specifying switch performance:

$$C_sR_t < 2 \times 10^{-8} \text{ seconds}$$
$$C_sR_l > 2 \times 10^{-2} \text{ seconds}$$
$$100C_o < C_s$$

These may conveniently be rearranged and expressed in terms of appropriate practical units:

$$\left.\begin{array}{l} \text{(i) } R_t(\Omega) < \dfrac{200}{C_o\,(\text{pF})} \\ \text{(ii) } R_t(\Omega) < R_l(\text{M}\Omega) \\ \text{(iii) } \dfrac{2 \times 10^4}{R_t(\Omega)} > C_s(\text{pF}) > 100\text{F}_o\,(\text{pF}) \end{array}\right\} \text{ where } R_t = R_c + \tfrac{1}{2}Z_o$$

It can be seen from (i) and (ii) that each of the unwanted characteristics of the open switch, $C_o$ and $R_l$, imposes a maximum value on the total charging or discharging resistance $R_t$. The inequality (iii) shows that the extent to which the right-hand term of (i) exceeds the left-hand term defines a range of choice for the storage capacitance $C_s$. The quantities $R_c$ and $\frac{1}{2}Z_o$ are only involved in these expressions in terms of their sum $R_t$. However, the switch resistance $R_c$ should in fact be as small a fraction of $R_t$ as possible in order that differences between switches shall have the minimum effect on the operation of the store unit.

These inequalities enable a designer to make a quick assessment as to whether or not a semiconductor device of known characteristics merits consideration as a switching element. If, for instance, the shunt capacitance, $C_o$, of a semiconductor switch is 2·5 pF, then the storage capacitance, $C_s$, must be at least 250 pF and the value of the switch resistance, $R_c$, must be considerably less than 80 Ω, defined by (i) as a maximum value for $R_t$. If $R_t$ is assumed to be 80 Ω, $R_l$ must be at least 80 MΩ and $C_s$ must be 250 pF. If, however, $R_t$ can be made equal to (say) 20 Ω, $R_l$ need only be 20 MΩ and $C_s$ may have any value between 250 pF and 1,000 pF.

## 5. A Specimen Design of Store Unit

### 5.1 *The Switch*

So far, the requirements of a switch have been discussed solely in terms of its operating speed, its resistance and shunt capacitance in the non-conducting state, and its resistance in the conducting state. In choosing a practical circuit configuration, however, it is very desirable also that the switch be balanced; that is to say, application of the switching pulse should cause no current to flow in a short circuit joining the switch contacts. This condition demands the use of at least two semiconductor elements that must be closely matched as regards both static and

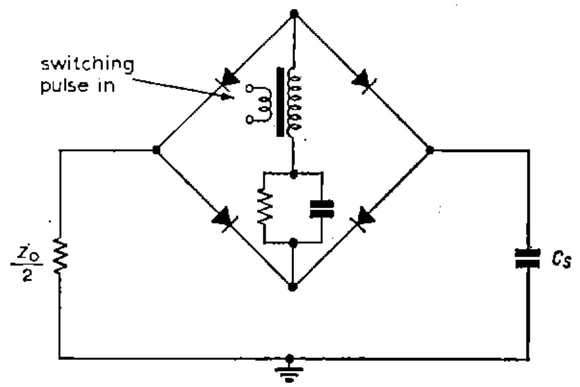


*Fig.* 4 — *Configuration of a practical switch*

transient characteristics. Use of a balanced switch avoids the need to control precisely the magnitude of the switching pulse because, provided that the pulse is large enough to operate the switch, its magnitude does not affect the output of the store unit.

Various forms of two-, three-, and four-layer semiconductor devices may be used as switches; of the possible arrangements that have been investigated, the four-diode switch shown in Fig. 4 has been found to be the cheapest circuit arrangement capable of meeting all the requirements enumerated. In such a configuration, the values of $C_o$, $R_c$, and $R_l$ for the bridge of four diodes are the same as for a single diode. For modern diodes with which the required speed of operation can be obtained, a typical value for $R_c$ is 7 Ω. As stated earlier, a transmission line can readily be designed to have an impedance of 26 Ω; the total charging or discharge resistance $R_t = R_c + \frac{1}{2}Z_o$ is, therefore, about 20 Ω. Thus the inequalities derived in the preceding section require that $C_o$ shall be less than 10 pF and $R_l$ shall be greater than 20 MΩ. This last figure precludes the use of fast germanium diodes, but the available range of fast silicon diodes includes several types which meet the above requirements. A typical diode of this type, costing about three shillings, has $R_c = 7\,\Omega$, $C_o = 2$ pF, and $R_l = 750\,\text{M}\Omega$.

When experimental switches were constructed using diodes of this type it was found that although the other conditions enumerated were easily met, an excessive degree of unbalance occurred in bridges made up at random from a batch of diodes. Lack of balance results in a constant error between the input and output signals of the store unit and is one of the factors producing striations in the converted picture. Some improvement in the ratio of video signal to added component can be made by using a larger video signal. However, this necessitates increasing the switching current with consequent aggravation of the unbalance, and it has been found in practice that little is gained by using a video level of more than 6 V, peak-to-peak.

The unbalance of a complete store unit is conveniently measured by connecting the unit between resistive terminations of $\frac{1}{2}Z_o$, operating writing and reading switches alternately, and observing the step, $\Delta V_s$, in the voltage across the storage capacitor caused by each operation of the reading switch. This step signifies a transfer of charge to the reading circuit of the same magnitude as that which would result from application of a video voltage $\Delta V_s$ to a perfectly balanced store unit.

Experiments have shown that this unbalance is primarily a high-frequency phenomenon produced by differences in the hole-storage and 'forward recovery-time' characteristic of the diodes, and despite the use of diodes that are, individually, satisfactory in these respects, it has, nevertheless, been found necessary to employ a process of selection in order to bring the unwanted added signal below the level corresponding to the threshold of perceptibility for vertical striations.

Two methods of selection have been used successfully. In the first method, selection is applied to groups of four diodes; these groups each contain diodes having similar characteristics at low frequencies, and are supplied by the manufacturers as 'matched quads'. Each quad is incorporated in a high-speed switch circuit similar to that used in the complete store unit and is classified according to the direct voltage, $\Delta V$, appearing across the storage capacitor as a result of repeated operation of the switch. Store units are then made up, the writing switch and the reading switch of each unit consisting of quads 'paired' for similar values of $\Delta V$; by connecting the two quads appropriately substantial cancellation of unbalance is achieved.

The second method of selection, which applies to single diodes, again involves determining the out-of-balance voltage, $\Delta V$, for a high-speed switch. In this method, three arms of the bridge contain diodes previously selected for identical characteristics, and the diodes from an incoming batch are, in turn, inserted in the fourth arm; each diode can thus be classified by a value of $\Delta V$. Quads are then made up, each from a group of four diodes yielding similar values of $\Delta V$. By this means diodes are matched sufficiently well for pairing of the resultant quads to be unnecessary.

When diodes chosen at random are assembled into quads, which are then used without 'pairing', the total unbalance of a store unit, $\Delta V_s$, is typically $\pm 200$ mV. This figure can be reduced to about $\pm 60$ mV by the use of unpaired quads whose individual diodes have been matched for low-frequency characteristics alone, and to about $\pm 10$ mV by either of the two selection procedures described above.

So far, the effects of circuit inductance have been disregarded, but, at the switching speeds under discussion, inductance has, in fact, a significant effect upon the charging and discharging of the storage capacitor. In a practical layout involving a transmission line and a printed-circuit board the total length of the charging or discharging path is at least two or three inches, the inductance of which is about 0·1 μH.

A simplified equivalent circuit which includes a stray inductance $L_o$ is shown in Fig. 5(*a*).

The transient voltage waveforms after closure of the switch have been calculated for four typical values of inductance and are shown in Fig. 5(*b*); it has been assumed

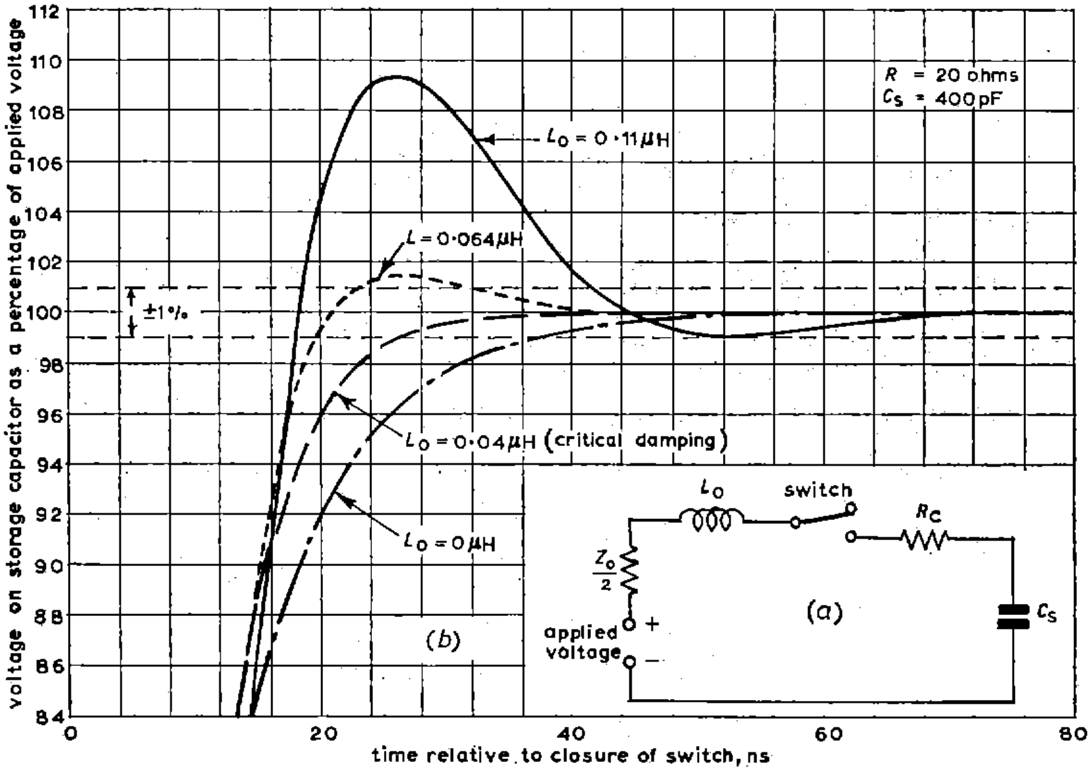


*Fig. 5 — The effect of circuit inductance on the charging process*

(*a*) Equivalent circuit including inductance $L_o$
(*b*) Voltage waveform across $C_s$ for four values of $L_o$

that the applied voltage step is positive-going and that the circuit has $C_s = 400$ pF and $R_t = R_c + \frac{1}{2}Z_o = 20\,\Omega$.

The requirement is that the voltage across $C_s$ should rise rapidly to within, say, $\pm 1$ per cent of the applied voltage and then remain within these limits. Fulfilment of this condition ensures that the stored charge is almost unaffected by small variations in the values of $L_o$ or $R_t$, or in the duration of switch closure; the charge is, in fact, related almost exclusively to the applied voltage by the value of $C_s$, which can be controlled within a very close tolerance and should remain stable with changes of temperature over a long period of time.

From Fig. 5(*b*) it can be seen that the voltage across $C_s$ remains within 1 per cent of the applied voltage after 36 ns in a circuit with no inductance, after 26 ns for $L_o = 0\cdot04\,\mu$H, and after 32 ns for $L_o = 0\cdot064\,\mu$H. For values of $L_o$ up to $0\cdot08\,\mu$H the charging of the capacitor to within 1 per cent of the applied voltage is more rapid than for $R_t$ and $C_s$ alone and the stray circuit inductance becomes a useful adjunct rather than an embarrassment. Above $0\cdot08\,\mu$H the presence of inductance extends the time for which the voltage across $C_s$ differs by more than $\pm 1$ per cent from the applied voltage, and, for $0\cdot11\,\mu$H, Fig. 5(*b*) shows this time to be 42 ns. Thus it is important to reduce the circuit path length to the point where the value of $L_o$ is of the order of $0\cdot04\,\mu$H to $0\cdot08\,\mu$H.

5.2 *The Switching-pulse Generator and Shift Register*

As stated in the previous section, unbalance in the four-diode switches adds an interfering signal to the output of the converter and the video signal must, therefore, be sufficiently large for this interference to be subjectively imperceptible. The required ratio has been shown[2] to be 49 dB, so that the existence of a typical degree of unbalance, equivalent to a spurious input signal of $\pm 10$ mV, necessitates a video-signal magnitude at the writing transmission line of 6 V peak-to-peak.

A further interfering signal is generated as a result of slight differences in the effective gains of individual store units; its magnitude is proportional to the difference between the potential on the input line and the potential to which the store capacitor is discharged. Since this capacitor is discharged to earth potential, the interference has been minimized by so choosing the d.c. level of the video signal on the input line that its 6-V range produces maximum excursions of $\pm 3$ V about earth potential.

To charge or discharge a 400-pF storage capacitor through 3 V* within, say, 40 ns demands a mean current of 30 mA. The value of peak current depends upon

* Although the stored potential has a total range of 6 V, it is never changed by more than 3 V at a time, because the store capacitor is discharged to zero potential, by the reading operation, between successive writing operations.

the circuit inductance and upon the forward recovery characteristics of the diodes; nevertheless, it is at least twice that of the mean current. The peak current supplied by the switching-pulse generator must exceed the peak charging or discharging current by a sufficient margin to maintain good conduction in the diodes and the design should include a margin of safety in order to allow for component tolerances and ageing effects. A peak switching current of 100 mA has therefore been taken as the design figure for the switching-pulse generator.

The use of an unnecessarily large switching current is disadvantageous because spurious signals are thereby increased and because more power is required.

As has already been stated in the discussion of the switch design, the practice of charging and discharging the store condenser to the equilibrium value and the use of a balanced switch circuit allow the switching-pulse generator to be designed to a specification permitting reasonably wide variations of pulse magnitude and duration, with a consequent saving in cost. The pulse of current through the diodes need be specified to no greater precision than is shown, in Fig. 6, by the area inside the solid lines; the duration must have a value between 60 ns and 100 ns, the maximum current should be maintained for as long as possible and the waveform must be free from significant overshoots. The dotted line in Fig. 6 represents a typical pulse shape.

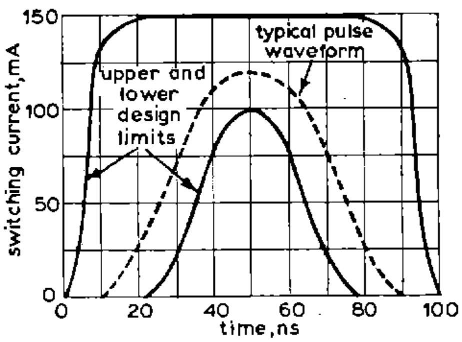


*Fig. 6 — Design tolerances for the switching pulse*

The circuit used for the switching-pulse generator must, therefore, be the simplest circuit capable of delivering a high-current pulse, of short duration and low duty-factor, with short rise-and-fall times. Two basic configurations potentially capable of meeting these requirements are the blocking oscillator and the avalanche pulse generator; circuits of both forms have been tested, using a number of types of switching transistor. It has been concluded that, while usable circuits can be evolved from either configuration, neither is wholly satisfactory. Fast transients and high currents may be obtained with a blocking oscillator, but the magnitude and shape of the pulse produced by such a circuit are very dependent upon individual semiconductors and transformers. Generators using transistors in the avalanche mode are more consistent, but are sensitive to the magnitude of the triggering pulse and, in general, require a more expensive type of transistor. It has been found possible, however, to combine the advantages of both circuit configurations by applying positive feedback to an avalanche circuit, thereby removing its sensitivity to trigger pulse magnitude. Fig. 7(*a*) shows the circuit of a complete store unit incorporating an avalanche pulse generator of this type. It will be seen that the feedback is derived by adding a tertiary winding to the transformer that is used in order to couple the unbalanced pulse generator to the balanced load presented by the switch; thus little extra cost is involved.

Fig. 7(*a*) also shows the way in which the switching-pulse generators in adjacent stores are coupled together in order to provide a 'shift register' as discussed earlier in this report. The condition that each generator shall operate only when it simultaneously receives a clock pulse and a 'hand-on' pulse from the previous store unit is met by interposing an 'and' gate between the clock-pulse input and the trigger circuit of the pulse generator. The hand-on pulse is, of course, initiated by the clock pulse immediately preceding the one with which it must coincide. Delay must, therefore, be incorporated in the coupling between adjacent store units, and this is accomplished by means of the inductors $L_d$ shown in the diagram. The inductance $L_d$ and the associated stray capacitance at the input to the avalanche pulse generator form a simple delay section.

The appearance of the circuit when embodied in a printed-board assembly is shown in Fig. 7(*b*); in order to allow the components to be seen more clearly, a small copper screen, usually mounted across the centre of the board, has been removed.

### 5.3 *Assessment of Performance*

As has already been stated, mutual capacitance coupling the input and output transmission lines causes contamination of the output signal by unconverted input signal. It has been found[2] that perceptible degradation is avoided if the capacitance is of a lower value than that corresponding to a signal-to-interference ratio of 16 dB at 3 Mc/s. It can be shown that when due allowance is made for the fact that the output signal is of lower magnitude than the input signal, the critical value of the total mutual capacity is 10 pF. Assuming 500 store units, therefore, the capacitance between the input and output of each must be less than 0·02 pF. Measurements, by means of conventional bridge techniques, have established that this condition is in fact fulfilled by the assembly shown in Fig. 7(*b*), when the copper screen mentioned above is fitted.

In order to allow the remaining aspects of performance to be rapidly assessed, a special testing rig has been constructed, and its operation is illustrated in Fig. 8(*a*). The store unit under test is subjected to a repetitive cycle of writing and reading, in which each writing operation is followed by four successive reading operations; the period of the whole cycle is equal to four line periods of the 405-line standard. The writing and reading transmission lines are simulated by terminating resistors of value $\frac{1}{2}Z_o$, and a direct voltage of magnitude $V$ and, say, negative polarity

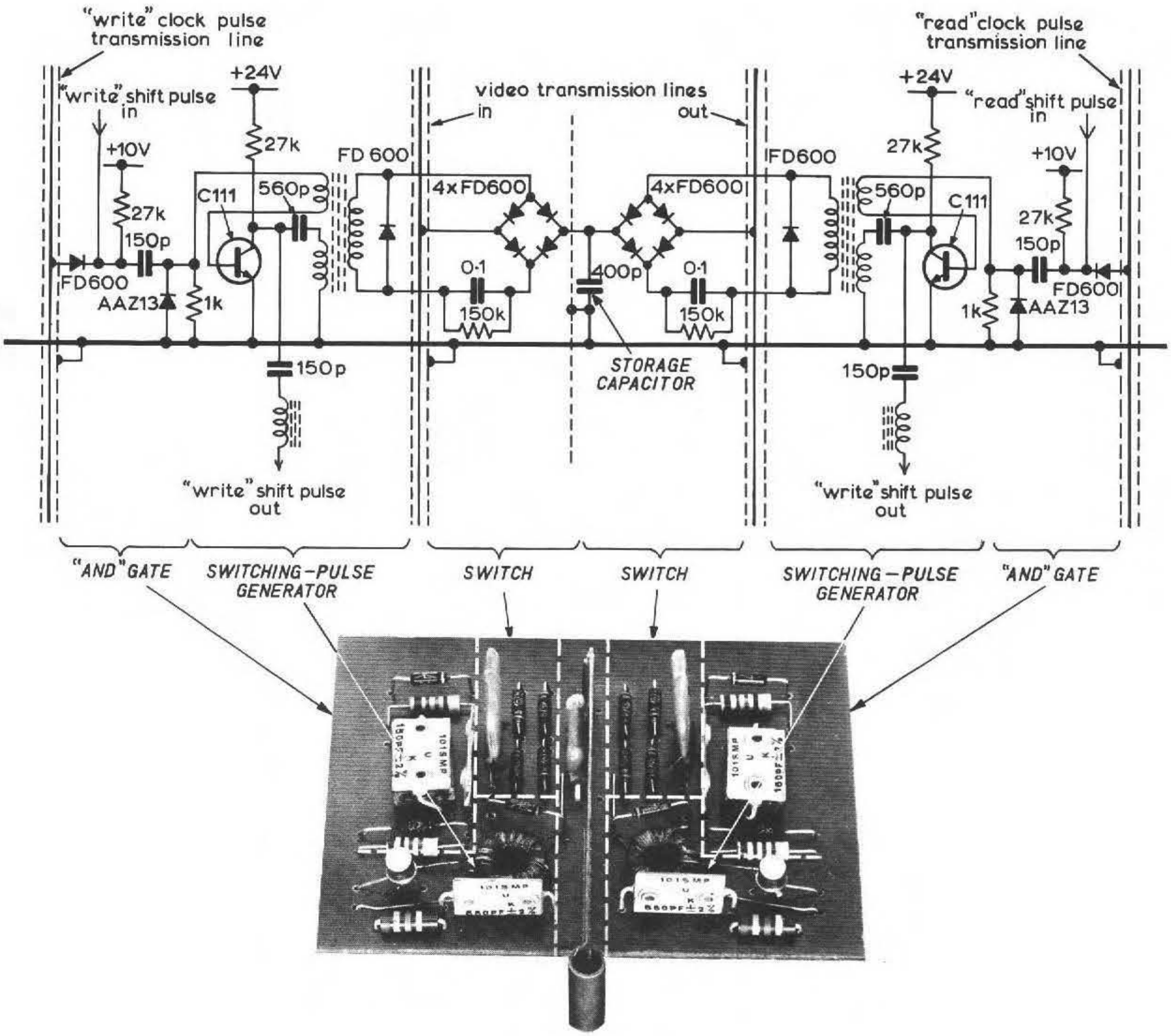


*Fig. 7 — A specimen design of store unit*

(*a*) Circuit diagram
(*b*) Store unit built on a printed-circuit board

is applied in series with the resistor on the input side of the store unit. It is arranged that the shift-register circuits are inhibited, so that all triggering pulses operate the appropriate switching circuits.

As has already been discussed, one of the requirements of a store unit is that the storage capacitor be charged or discharged almost to equilibrium by a single operation of the appropriate switching circuit. If, therefore, an ideal store unit were connected to the testing rig the first reading pulse to follow each writing pulse would result in complete discharge of the storage capacitor, whose voltage would then remain at zero until the next writing pulse caused it to fall to $-V$. The waveform observed on an oscilloscope connected to the storage capacitor would therefore be as shown at the bottom of the diagram. If, however, the reading circuits in the store unit were unable to discharge the capacitor in one operation, a further discharge would take place at the time of the second reading pulse to follow each writing pulse and a positive step would occur in the oscillogram. Furthermore, if the writing switch were of inadequate impedance in its 'off' condition, the storage capacitor would re-charge appreciably towards $-V$ during the intervals between pulses, and positive steps would also occur at the times of the third and fourth reading pulses. By reversing the connections to the writing and reading sides of the unit, and repeating the test, the charging efficacy of the writing circuits and the leakage of the 'read' switch can similarly be assessed.

Fig. 8(*b*) shows part of the storage-capacitor waveform for a typical store-unit; the deflexion sensitivity of the oscilloscope used was such that the applied voltage of $-3V$ corresponded to 300 of the small divisions on the graticule. It may therefore be seen, from the step which occurs at the time of the second reading pulse, that the residual charge left on the storage capacitor after the first reading pulse was only about 0·3 per cent of that stored by the writing process. Moreover, the absence of perceptible steps at the times of the third and fourth reading pulses shows that the effects of leakage were quite negligible.

In order to provide an indication of their performance, when incorporated in a line-store converter, eighty store units of the type shown in Figs. 7(*a*) and (*b*) were constructed using small-scale production techniques. These store units were suitably mounted and connected to video amplifiers and clock-pulse generators by means of tapped transmission lines. Fig. 9 shows photographs of two vertical strips of a 625-line Test Card 'C' converted to the

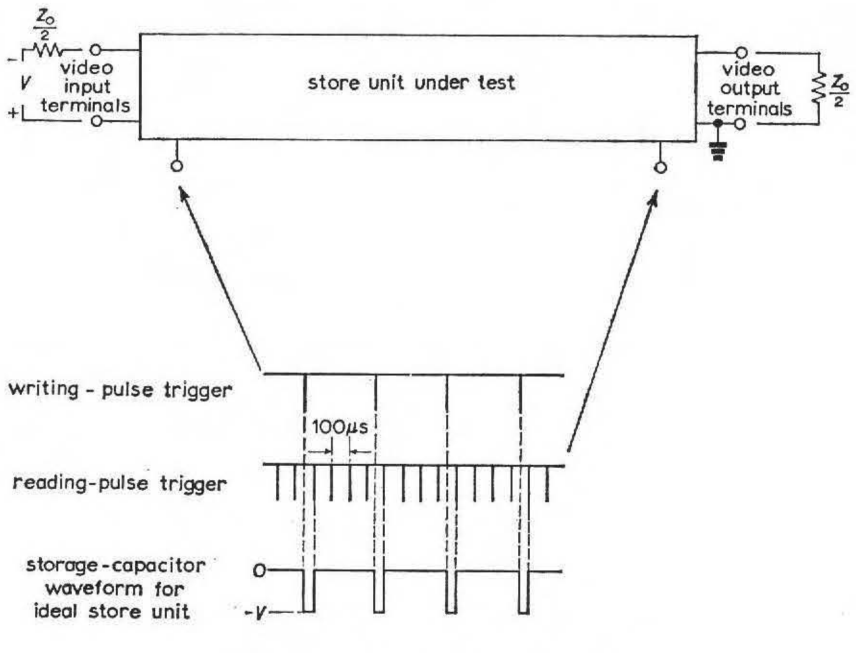


*Fig.* 8(*a*) — *Testing arrangements for store units*

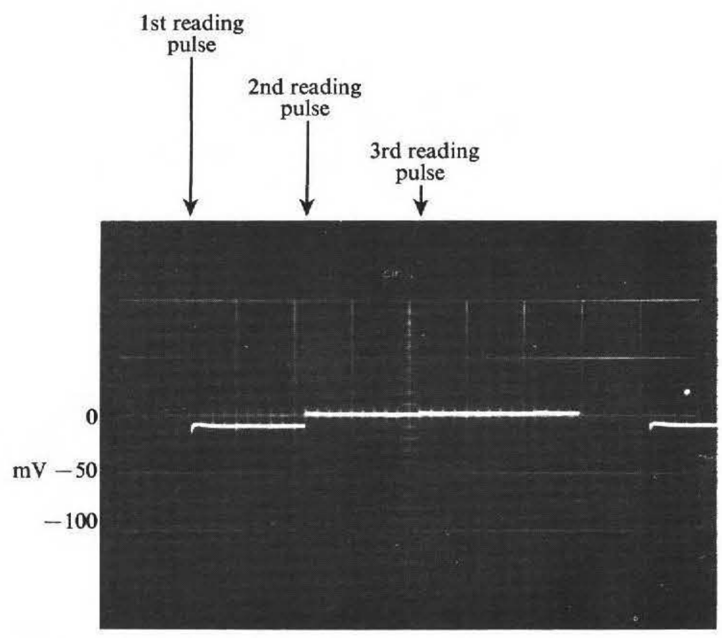


*Fig.* 8(*b*) — *Waveform on storage capacitor under test conditions*

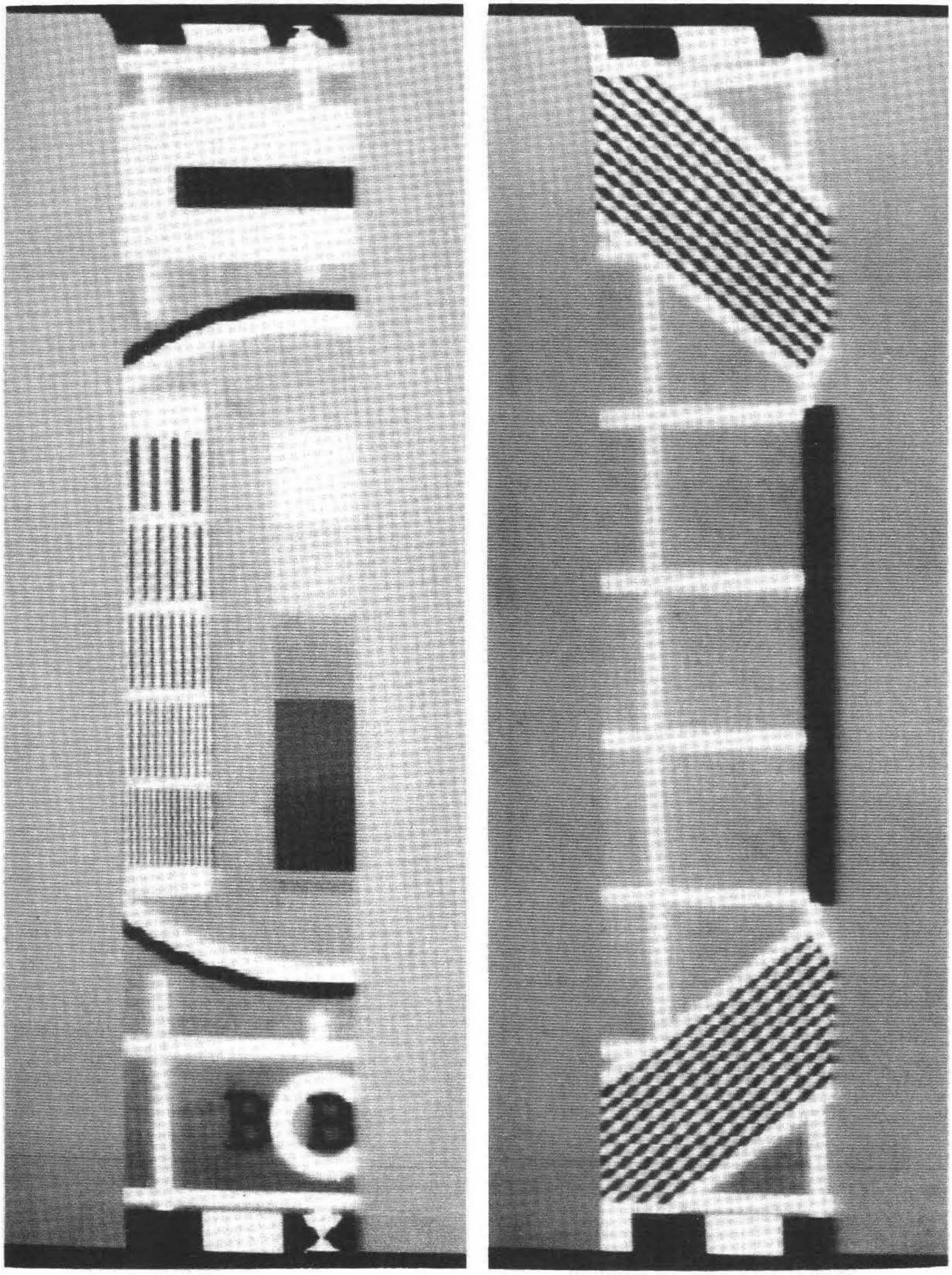

*Fig. 9 — Sections of a Test Card 'C' converted, without interpolation, from* 625 *to* 405 *lines by an experimental assembly of eighty store units*

405-line standard by this assembly; it should be noted that no interpolation was used. The photographs show that the performance of the store units was satisfactory.

## 6. Conclusions on Part I

As a result of the investigation outlined in this monograph, it has been established that satisfactory store units can be designed without recourse to elaborate circuits or unduly expensive components. Moreover, the writers do not consider that the combination of several hundred store units to form the nucleus of a practical converter would present any serious difficulties.*

* Subsequent experience with an operational converter, incorporating a full complement of store units, has shown that this assumption was correct.—Ed.

# Part II

# THE DESIGN OF VIDEO TRANSMISSION LINES

## 7. Specification of Transmission Lines

7.1 *Basic Requirements*

It will be appreciated that the route taken by a given picture-signal sample in its progress through the converter is different from that taken by other samples from the same television line. This route, comprising input connection, store unit, and output connection, depends on which store-unit input switch is closed at the time when the sample arrives at the input connection. Therefore any differences between routes cause spurious vertical striations to appear on the output picture.* All store circuits should therefore be identical in operation[2] and should be presented with the same impedance by each common signal connection. Furthermore, this impedance should be suited to the switch circuit since it will influence the store-capacitor charging action.

The interconnection system should have a bandwidth adequate for the video signals that it is required to carry (4·6 Mc/s) and it should be designed in such a way that the charging action is not disturbed by spurious signals arising from reflexions at remote points; the effects of such reflexions would vary progressively from store to store and so the interference produced would be visible as a series of broad vertical bands superimposed on the output picture.

Another cause of spurious interference patterns on the displayed output picture is 'break-through' of unconverted input signal components into the output circuits. Such break-through arises if any coupling exists between the input and output connection systems other than that provided by normal store-unit action. The maximum tolerable unwanted coupling has been established by means of subjective tests[2] and the results of these tests figure as an important specification in the design of the interconnection system.

It became apparent early in the design of the converter that the best method of satisfying the above requirements would be to use a single transmission-line for the common input connection and a second transmission-line, running parallel to the first, as the common output connection; the store units could be arranged to straddle the two lines and could be connected to them by means of tapping points. Adjacent store units would operate consecutively. A further pair of transmission lines similarly arranged would be used to distribute clock-pulse signals.

It was also decided at this time to drive the input line at one end of the group of transmission-lines and to extract the signal from the output line at the other end of the group. Thus all parts of the video signal would travel along the same total length of transmission-line irrespective of which store was being used, and so shading signals due to transmission-line attenuation would be avoided. A simplified diagram of the arrangement is shown in Fig. 10. For correct operation of the converter it is essential that the stored samples reach the output amplifier in a regular succession; it is also desirable, though less important, that the input signal be sampled at regular intervals. For any configuration of transmission-lines in which each video line runs alongside an associated clock-pulse line characterized by the same velocity of propagation, these conditions are automatically met provided only that the video signal always flows in the same direction as its associated clock pulses.

7.2 *Detailed Considerations*

7.2.1 *Impedance*

Whilst a store-unit switch is closed, the store capacitor is charged or discharged as part of a damped series-resonant circuit, half the transmission-line impedance forming about half of the damping resistance. This must be taken into account when the transmission-line impedance is chosen. Thus the impedance of the line should be sufficiently low to ensure that the charging (or discharging) current has ceased at the end of the period during which the switch is closed, yet it must be high enough as compared with the forward resistance of the diodes constituting the switch to prevent the total charging resistance from being unduly affected by the inevitable variations that occur in the values of diode resistance. (See Section 5.1.) However, it was found that only a limited range of impedances could be easily realized if lines of the two types considered prac-

* In practice, the subjective severity of these striations may be reduced by arranging that a given vertical strip of picture is handled by adjacent store units during alternate field periods.

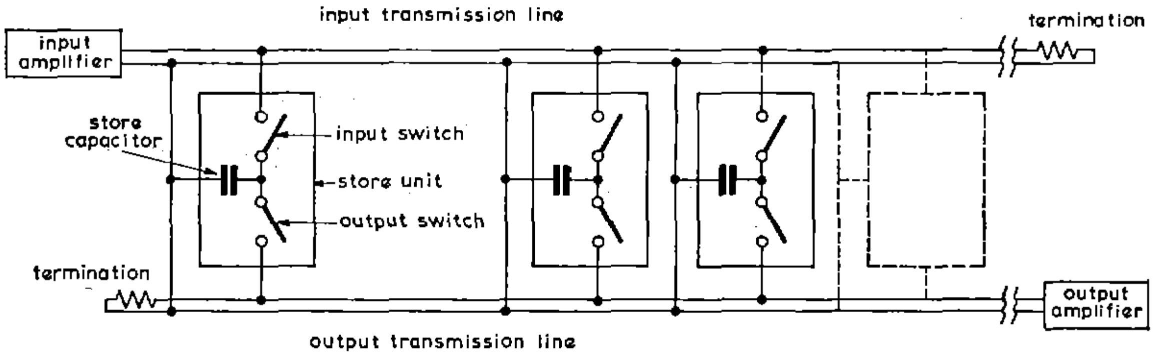


*Fig.* 10 — *Simplified block diagram of store assembly (both lines are terminated at both ends, however the above nomenclature assists the description)*

ticable were used. One type was of co-axial construction and it was found that in order to achieve uniformity with impedances of 5 Ω or less, the required degree of mechanical precision would be impracticable. The other type used flat copper strips as conductors and the width of these would be inconveniently large for impedances of less than 15 Ω. The upper limit of realizable impedance is decided by the fact that the line is loaded by the input capacitance of each inoperative store circuit. If an impedance of 55 Ω or greater were chosen, the capacitance presented at each tapping point by the associated store unit would constitute a major proportion of the capacitance determining the impedance of the line. Thus variations in store-unit capacitance would cause irregularities in impedance.

Thus it was necessary for the designs of the transmission-lines and store units to proceed together. The impedance range most convenient for transmission-line design (20 Ω to 30 Ω) also proved very suitable to the store-circuit requirements, and all the lines constructed were made to have a loaded impedance within this range.

Store-circuit loading affects both the impedance of the line and its propagation velocity. As is well known, the impedance of an unloaded line is given by:

$$Z_o = \sqrt{L/C}$$

and the velocity of propagation:

$$v_o = 1/\sqrt{LC}^*$$

where

$L$ = inductance per unit length
$C$ = capacitance per unit length

If the line is then loaded by a distributed capacitance of $C'$ per unit length, the impedance and the velocity are both reduced by a factor $\sqrt{1 + C'/C}$. The magnitudes involved may conveniently be illustrated by reference to the design of a practical line which is discussed later in this monograph. The effective loading by the store units and their associated connecting sockets was estimated as contributing about 200 pF per metre. In order to realize a loaded impedance of about 26 Ω it was necessary to design a line having an unloaded impedance approximating to 38 Ω and distributed capacitance of some 170 pF per metre ($\epsilon_r$ assumed equal to 4). The velocity of propagation along the line was reduced, by the loading, from about 0·50 ft/ns (0·15 m/ns) to about 0·34 ft/ns (0·10 m/ns).

* For an air-spaced line this velocity is approximately equal to $c$, the velocity of light in free space, and in general it is given by $c/\sqrt{\mu_r \epsilon_r}$ where $\mu_r$ is the relative permeability and $\epsilon_r$ the relative permittivity of the medium surrounding the conductors.

7.2.2 *Uniformity and Terminations*

When a store-unit input switch closes, the charging current drawn by the circuit produces at the switch a voltage pulse having a duration of approximately 30 ns and a magnitude proportional to the video-signal voltage existing on the input transmission line. Two voltage waves of this form therefore travel along the line, one in each direction from the switch. Discontinuities in transmission-line impedance and errors of termination cause the video-signal and either or both of the charging pulses to be reflected. The time at which reflexions return to the region of the store circuit considered depends on the distance between the store circuit and the discontinuity.

The output transmission-line carries voltage pulses due to the discharge of storage capacitors which occurs when successive store-circuit output switches close. These pulses are of similar duration to those produced at the input switches when the capacitors are charged, the train of pulses arriving at the output amplifier being amplitude modulated by the required video-signal. A similar train of pulses is propagated in the opposite direction, and it is intended that these be absorbed in the terminating resistor. However, discontinuities in the line or errors in termination result in reflexions of these unwanted pulses, which then appear as a spurious output signal.

The effect of reflexions in the input and output transmission-lines is further discussed in Section 10.

Reflexions within the transmission-lines carrying the clock-pulses cause a mistiming of the action of those stores at which the clock pulses and their reflexions overlap. This is relatively unimportant to the writing process, where the effect is a negligible horizontal shift of information in some parts of the output picture. However, such

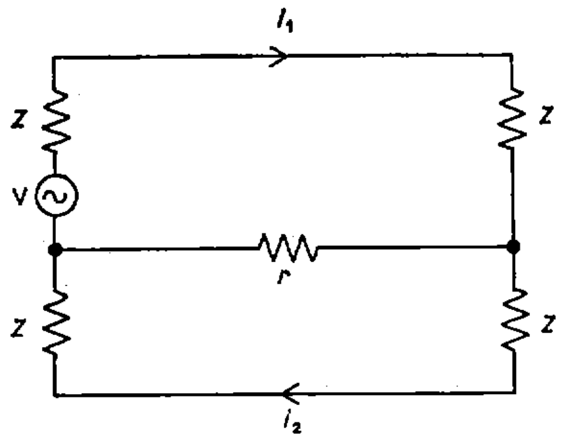


*Fig. 11 — Equivalent circuit showing common earth impedance*

disturbances to the reading process affect the regular arrival of pulses at the output amplifier, and this can cause a significant error in the output signal derived by filtering.

Approximate calculations have indicated, however, that a discontinuity in the reading clock-pulse line would not produce a visible pattern unless the modulus of the resultant reflexion coefficient were greater than 1/5. Reasonable care in design should enable such discontinuities to be avoided.

7.2.3 *Unwanted Coupling between Input and Output Lines*

Break-through of signal from the input transmission-line to the output line can arise in two ways, which are considered in the following sections.

7.2.3.1 *Coupling Due to Shared-earth Conductor*

The store circuits used in the converter considered are connected by means of unbalanced transmission-lines. The earth connection is therefore common to both input and output lines; if its resistance is significant, current in one line can produce a voltage across this resistance and so interfere with the signal voltage in the other line.

Fig. 11 shows an equivalent circuit in which two identical transmission lines share a common earth return of resistance $r$. Each line is terminated by an impedance $Z$, and the input line, driven from a matched source, carries a current $i_1$. A current $i_2$ is induced into the input line because of the shared resistance $r$.

It can readily be deduced from this circuit that if $r = xZ$, $(x << 1)$, an interfering voltage approximately equal to $x/2$ times the input-line voltage is developed in series with the output-line voltage. The magnitude of the video-signal component in the train of pulses carried by the output transmission-line may be shown to be about 1/30 of that of the video signal carried by the input line. It follows that for the interfering signal to be attenuated by at least 40 dB relative to the wanted output signal,[2] $x/2$ should be less than 1/3,000, i.e. $x < 1/1{,}500$. This means that for $Z = 25\,\Omega$, $r < 1/60\,\Omega$.

It is evident from Fig. 11 that the polarity of the coupled signal would be inverted if the direction of propagation of signal along the input line were reversed. If, therefore, the total length of the group of transmission-lines used in the store assembly is broken up into an even number of sections, this form of coupling can be substantially eliminated by arranging that in half of the sections the input and output signals flow in the same direction, while in the other half they flow in opposite directions.

Fig. 12 shows how this condition may be achieved by suitably interconnecting the transmission-lines of four adjacent sections of store assembly. Although the input and output transmission-lines (each with its associated clock-pulse line) link the store units in different sequences, correct operation of the converter is maintained by arranging that 'writing' and 'reading' hand-on pulses link the store units in the same sequence. This method of interconnection violates the condition, mentioned in Section 7.1, that all paths through the converter should involve the same total length of transmission-line. The individual sections of line are, however, comparatively short, and the resultant differences in path length are not great enough to cause perceptible shading in the converted pictures.

7.2.3.2 *Coupling Due to Inter-line Capacitance*

In the store assembly of the converter, the input and output transmission-lines run parallel to one another for a distance of the order of 50 ft (15 m). Therefore, unless they are adequately screened, their distributed mutual capacitance and mutual inductance give rise to coupling between the lines.

If the relative directions of flow of input and output signals are reversed in alternate sections, as envisaged above, the net mutual inductance is small and the coupling

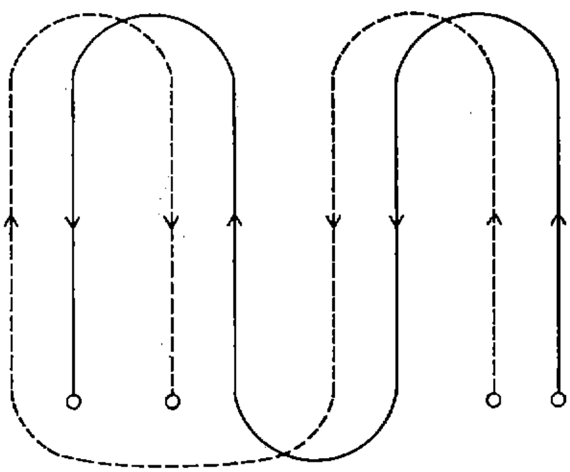

*Fig. 12 — Transmission lines interconnected in such a way as to eliminate coupling through common earth connection*

——— Input transmission lines

– – – – Output transmission lines

is predominantly capacitative. If this arrangement is not used (i.e. the lines are arranged as shown in Fig. 10) the respective contributions of mutual capacitance and mutual inductance to the coupled signal present at the output amplifier are in opposition and would, under certain ideal conditions, cancel completely;[6,7] in practice, however, the effect of capacitance is likely to predominate because of the additional capacitative introduced coupling by connection of the store boards. Therefore, in an attempt to obtain an early estimate of the amount of inter-line screening necessary, it was decided to make the assumption that no mutual inductance would exist and hence that the inter-line capacitance would be wholly responsible for the production of interfering signals.

Considering the distributed inter-line capacitance as composed of a large number of elements, the contribution that a given element makes to the interference at the output amplifier is independent of the location of the element, since the total distance travelled by each contribution is constant. Therefore, a single lumped capacitance of magnitude equal to the total distributed capacitance may be regarded as replacing the distributed capacitance in so far as the effects of interference are concerned. The equivalent circuit is shown in Fig. 13.

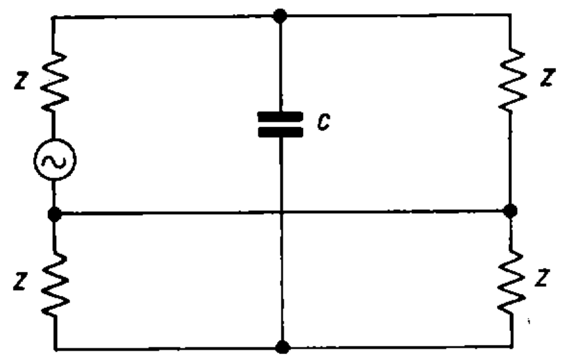


*Fig.* 13 — *Equivalent circuit showing inter-line capacitance*

Since $C$ is small, the signal is effectively coupled to the output through a differentiating network. As mentioned in Section 5.3, it has been found that under these circumstances a 3 Mc/s interfering signal must be attenuated by at least 16 dB relative to the wanted output if it is to be imperceptible. The wanted signal is, as mentioned, reduced by a factor of 1/30 in its passage through the storage system; and it may be deduced that, for $Z = 25\,\Omega$, the total inter-line capacitance must be less than 22 pF to meet the requirement.

### 7.2.4 *Mechanical Considerations*

Since the transmission lines are connected to every store unit, their design has a substantial influence on the configuration of the machine. Each store unit spans four transmission-lines and it is therefore desirable that the lines run as closely together as the coupling requirements will allow, so that their spacing does not limit the compactness of the store units.

The minimum realizable spacing between neighbouring store units depends on the way in which they are mounted and on the degree to which stray fields associated with one store are able to influence the working of adjacent stores. The spacing of stores, of course, affects the degree to which the transmission line is loaded and hence the impedance that the unloaded line should be designed to have.

Transmission-line construction is also affected by the method of connection of store units to tapping points. If soldered or other semi-permanent connections are made, ease of servicing requires either that each store unit should be very readily accessible or that it should be easily possible to remove an appropriate length of line, together with the attached stores. On the other hand, if store units are connected by means of plugs and sockets the transmission-lines may be made a fixture, provided that store units are readily accessible.

Both of the above methods were tried in the development of the converter, and very different mechanical constructions resulted.

## 8. Development of Practical Transmission-lines

Several forms of transmission line were constructed during the development of the converter and two types will now be discussed. Both were of rigid construction, forming part of the framework upon which the store units were mounted.

### 8.1 *Co-axial Transmission-lines*

Whilst the converter was in an early stage of development it was decided to design both the store-unit layout and also the transmission-line system so as to minimize as far as is practically possible the crosstalk between the input and output circuits.

Fig. 14 shows the arrangement used. The transmission lines were of square co-axial cross-section of approximately ½ in. (12·7 mm) outer dimension, their construction being as shown in Fig. 15. The inner conductors were located by means of 8BA brass studs and insulating spacers with the studs projecting on each side through insulated mounting boards to become the taps to which circuit connections were made. The store-unit printed-circuit boards were drilled to allow the taps to pass through them and soldered connections were then made to them through pigtails of copper braid. Earth connections were provided at each tapping point by means of pegs returned to brass plates sandwiched between the transmission-lines and the mounting boards. The additional pegs shown in Fig. 14 supplied power and 'hand-on' connections.

The layout of the store assembly was such that individual store units were not readily accessible. The experimental store assembly, containing eighty-four store units in all, was therefore divided into two sub-assemblies, each containing forty-two store units. The store units visible in Fig. 14 were some of the twenty-one units that lay above

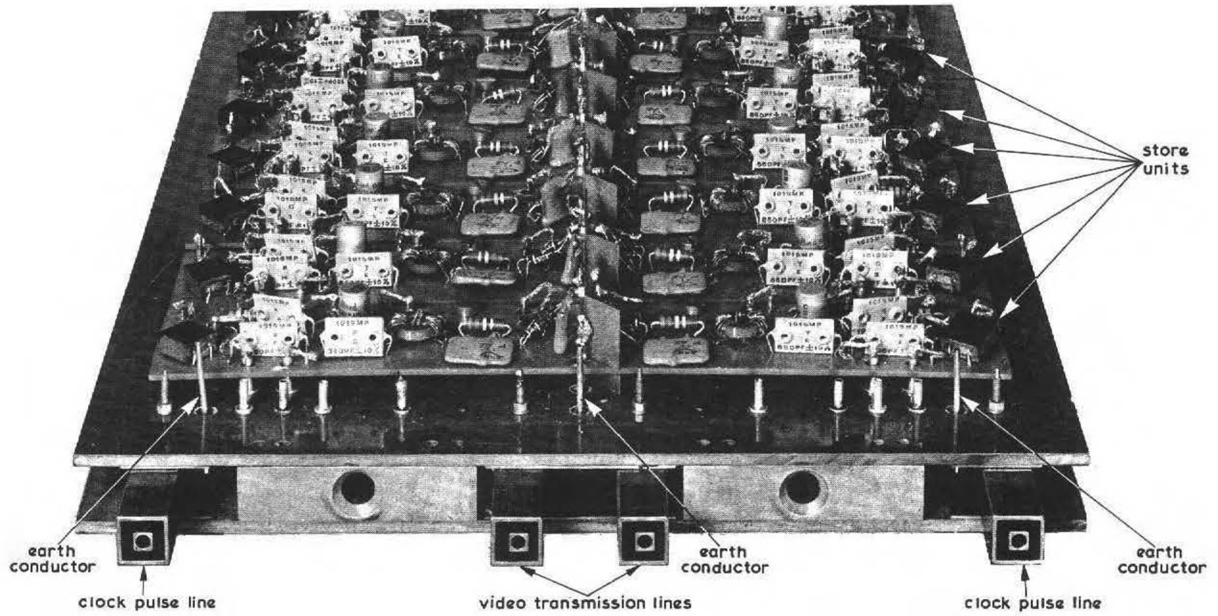


*Fig.* 14 — *Store assembly layout using co-axial transmission-lines*

the plane of the transmission lines; a further twenty-one units, not visible in the Figure, lay below this plane. The sub-assemblies were located by means of the grooves formed by the clock-pulse transmission-lines and the edges of the mounted boards. The ends of the transmission-line sections were designed to plug into U linkages of similar cross-section.

In order to determine the cross-sectional dimensions of the line necessary to produce the required impedance, a resistive analogue technique was used. This technique is possible because the electrostatic flux which exists between two conductors of different potential in a dielectric medium takes up the same distribution as the current flux between similar disposed conductors in a resistive medium.

The total electrostatic flux within a dielectric medium defines the charge $q$, on the condutcors, whereas the total current flux between conductors in a resistive medium defines the current $I$. The similarity of the two flux distributions means that similarly altering the configuration of both pairs of conductors (at constant voltage difference, $V$) changes $q$ and $I$ by the same factor. Thus, the capacitance $C$ between two conductors in a dielectric medium is inversely related to the resistance $R$ between corresponding conductors in a resistive medium. For a given dielectric the capacitance of a transmission-line is also inversely related to its characteristic impedance $Z_o$ and the value of $Z_o$ thus varies directly with the resistance of the resistive analogue.

Moreover, since electrostatic flux within a uniform loss-less transmission line is everywhere at right angles to the conductors, a two-dimensional resistive analogue describing a transverse cross-section may be used. The constant of proportionality relating the resistive analogue to the cross-section of an actual line may be determined by constructing the analogue of a line whose impedance may be easily calculated from a knowledge of its dimensions.

In this way, the impedance of a square co-axial line was determined. The analogue was constructed using a paper which had been uniformly impregnated with carbon so as to make its resistance equal to 2 kΩ per square.* Conductors were drawn on the paper using a silver paint of high conductivity.

A circular co-axial cross-section was first constructed as a reference by means of which the analogue was 'cali-

* i.e. the resistance between opposite sides of a square of any size is 2 kΩ.

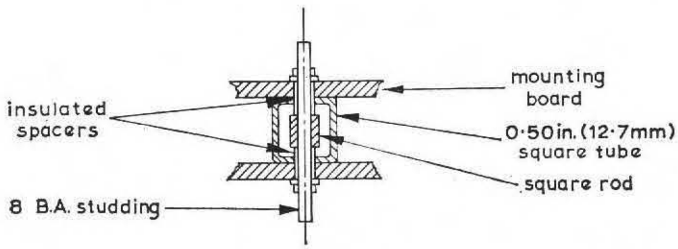


*Fig.* 15 — *Construction of co-axial transmission-lines*

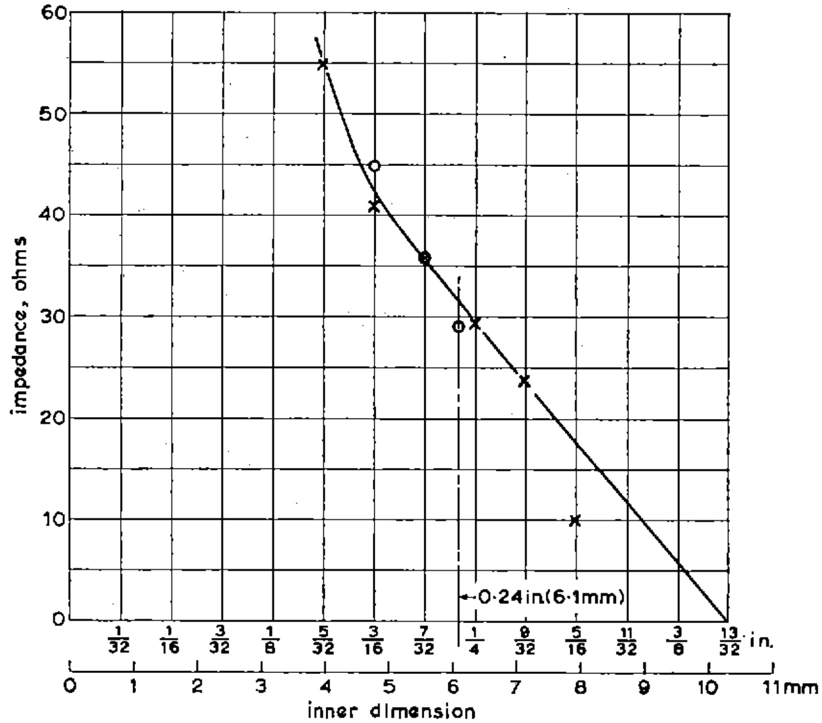


*Fig.* 16 — *Square co-axial transmission-line impedances as predicted by resistive analogue and as deduced from measurements of inductance and capacitance per unit length*

brated'. Thereafter square co-axial cross-sections were drawn. A fixed outer was used, first with a very small inner, and then with successively larger inners, measurements of resistance being made at each step. These measurements gave data from which the graph shown in Fig. 16 was constructed.

Three experimental sections of line were made, each using, as the outer, a square tube of $\frac{1}{2}$ in. (12·7 mm) outer dimension and 18 SWG (1·22 mm) thickness, with square inners of 0·24 in. (6·1 mm), $\frac{3}{16}$ in. (4·8 mm), and $\frac{7}{32}$ in. (5·6 mm). Their impedances were computed from measurements of inductance and capacitance per unit length and were found to be lower than those predicted by Fig. 16. This was largely due to the additional capacitance introduced by the studs and spacers. Half of the studs and spacers were therefore removed, and the line capacitance was again measured. By this means an allowance could be made for tapping-point capacitance and when this was done the measured impedances shown in Fig. 16 resulted. Agreement between the predicted and measured impedances is good, considering the limited accuracy of the resistive analogue and the tolerances achieved in transmission-line construction.

For the experimental store assembly (shown in Fig. 14) transmission lines having $\frac{3}{16}$ in. (4·8 mm) inners were used which, in the absence of taps, would have resulted in a line impedance of about 45 Ω. This impedance was lowered to about 35 Ω by the taps and spacers and then to about 26 Ω by the stores and earthing plates.

The assembly constructed in this way housed a total of eighty-four store units. Its performance was entirely satisfactory. However, at this time a reappraisal of the mechanical design of the store assembly was made and it was considered desirable to reduce the size of the store units and to attempt the development of a simpler and more compact configuration of store-units and transmission-lines. These considerations led to the system which will now be described.

### 8.2 *Strip Transmission-lines*

Strip transmission-lines manufactured from double-sided printed circuit board have been in use for a number of years as a means of transmitting signals at microwave frequencies.[8,9] They have the advantages of light weight, compactness, and simplicity, and they can be made with great precision. It was therefore thought that they might offer a very convenient solution to the problem in hand.

At the same time it was decided to attempt a reduction in the size of the store-unit circuit boards and to make them easily and separately detachable by the use of printed circuit edge connectors, the sockets being made permanent fixtures connected to the transmission-line and power-supply systems.

These measures all tended to increase the unwanted stray capacitance coupling unconverted signals to the output, and it was therefore necessary to ensure that the limit calculated in Section 7.2.3 was not exceeded.

The twenty-four-way sockets were first examined. It was found that if all pins of one socket, other than those making the video connections to the two transmission lines, were left unconnected, the total inter-line capacitance introduced by connection of 600 such sockets would be about 144 pF. This could be reduced to about 42 pF if pins between the video connections were earthed and would be no greater than 2·4 pF if, in addition, each socket were shrouded by an earthed screen. Suitable shrouds were therefore provided.

Fig. 17 shows the way in which the transmission lines were constructed. The 'live' conductors were formed by removing the copper coating from the upper surface of the printed circuit board so as to leave narrow strips. The copper coating which covered the reverse side of the board formed the common earth connection. Additional strips were left on the copper surface of the board to provide earth and power connections.

The transmission lines were designed to have a loaded impedance of 25 Ω. The spacing between adjacent store units was to be $\frac{7}{8}$ in. (22·2 mm) and it was estimated that each store unit, with its connector, would load the line with a capacitance of 5 pF. The relative permittivity of the base material used for the lines was taken to be 4.

From these parameters it may be deduced by means of the formulae given in Section 7.2.1 that the line must have an unloaded impedance of 36·5 Ω.

Published curves[9] show that for a base material of relative permittivity 4 and thickness 0·06 in. (1·57 mm) the strip should have a width of 0·21 in. (5·34 mm) for an unloaded impedance of 37·5 Ω. Practical tests on an experimental board indicated that a width of 0·19 in. (4·83 mm) was necessary; this small difference probably occurred because the relative permittivity was not exactly equal to 4. A diagram showing the dimensions finally used is given as Fig. 18.

Tests on an experimental layout having the above dimensions indicated that the total capacitance coupling the input and output lines as constructed would be of the order of 3·8 pF but that this would be approximately halved when the earthed socket-shrouds were in position; this reduction of the capacitance between the lines themselves was fortuitous, the shrouds having been installed in order to reduce capacitance between the contacts of the connectors.

The coupling capacitance may be estimated by use of the formula deduced in the Appendix. This formula neglects the fact that the fields produced by the lines are somewhat confined by the high permittivity of the printed circuit board and consequently leads to the pessimistic estimate of 5·1 pF.

Fig. 17 also shows that the store-unit sockets were attached to the transmission-line system by bending the pins of the socket to make contact with the strips and then soldering to form the connection. The four central pins were soldered to an earthing strip which was connected to the copper coating on the reverse side of the board at

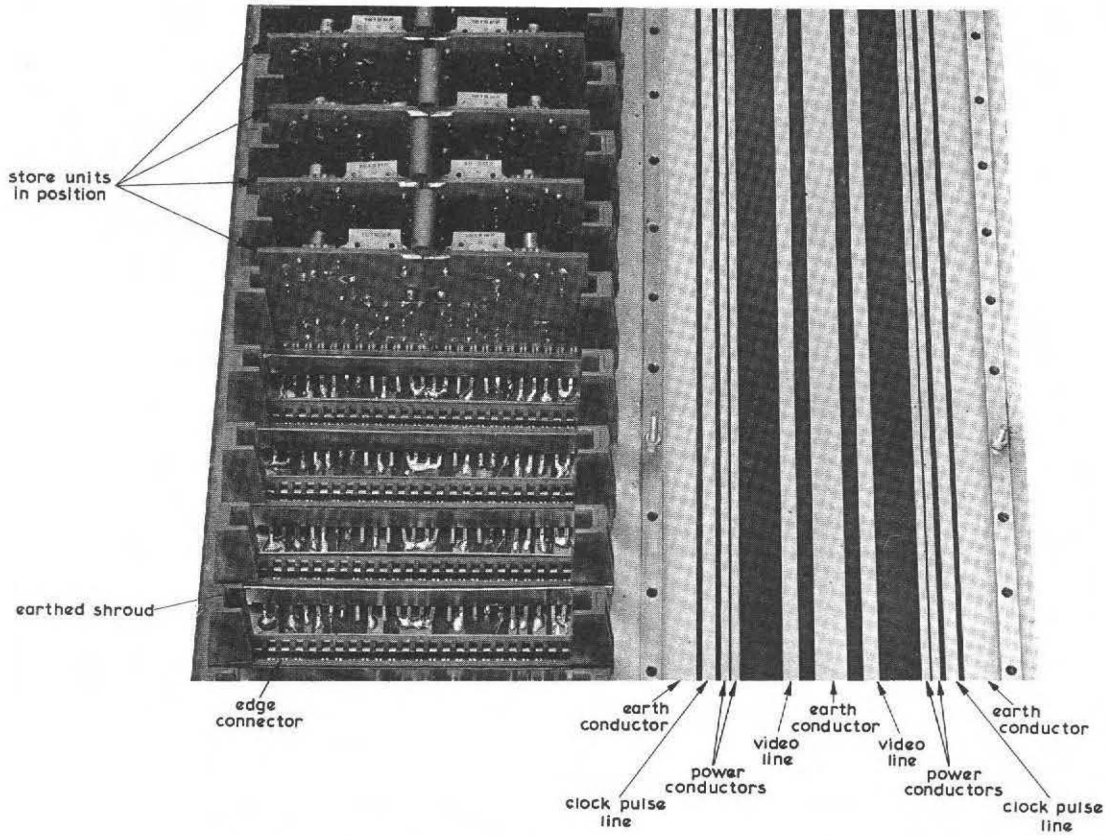


*Fig. 17 — Store assembly layout using strip transmission-lines*

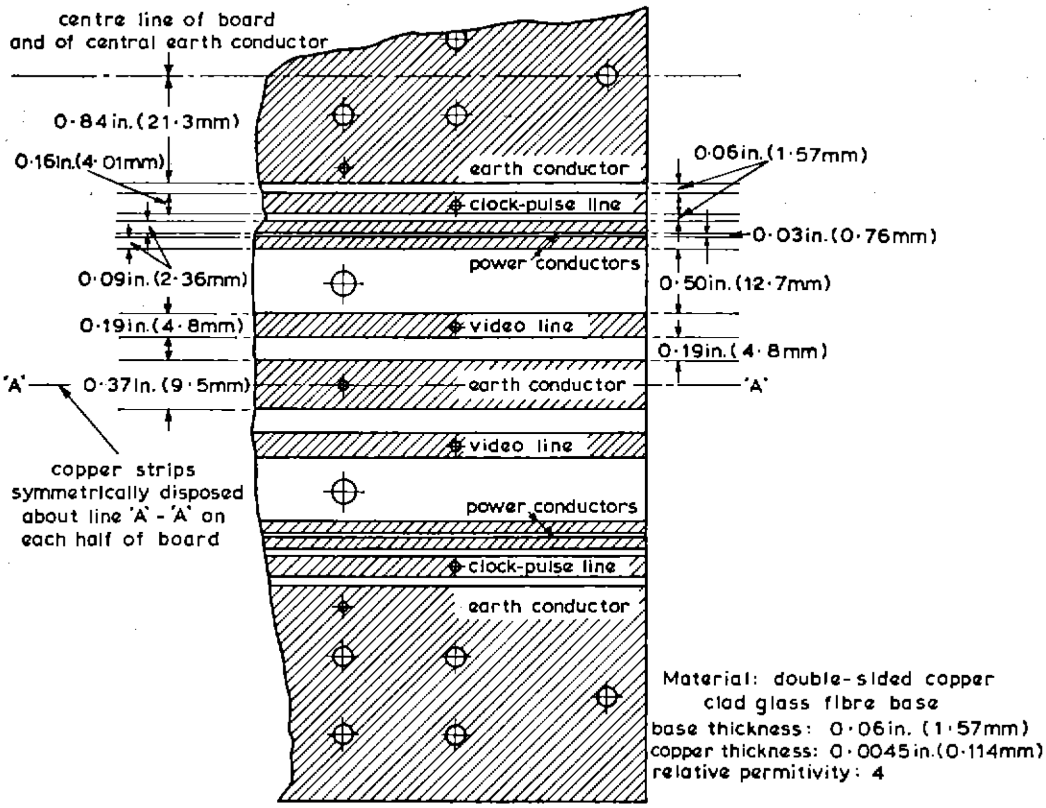


*Fig.* 18 — *Construction of strip transmission-lines*

each store position by means of a wire passing through a hole in the board. The shrouds were earthed at both ends by means of the fixing screws.

The prototype converter was constructed in this way, the store assembly being divided into two halves, one mounted on each side of the machine. A photograph of the machine showing half of the store assembly is reproduced as Fig. 19. The complete assembly comprised sixteen sections of multiple-strip line, as illustrated in Fig. 17, arranged in pairs in which one section was joined 'end-on' to a second section. Each pair of sections was associated with seventy-two store units and the pairs of sections were connected together by cables of the appropriate impedance. The total length of multiple-strip line was 45 ft (13·72 m). The connections were made in the way discussed in Section 7.2.3.1, in order to produce cancellation of 'break-through' signals developed across the finite impedance of the common earth connection. The clock-pulse signals were arranged to travel in the same direction as the associated video signals in order to ensure that correct timing was maintained.

The performance of the prototype store assembly was generally satisfactory; however, some broad stripes were just visible in plain areas of the converted picture, indicating transmission-line irregularities. The main cause of these is thought to have been slight mis-match between the pairs of sections of line and the 25 Ω co-axial cables joining them.

## 9. Measurements made on Prototype Machine

Transmission-line measurements were made on the prototype machine at the stage when one half of the store assembly was complete and working as part of the standards converter. The second half of the assembly was installed but was not connected to the first half, nor were its store units inserted.

The capacitance of each of the video lines in the second half of the assembly was first measured and found to be 1,845 pF. Next the velocity in each line in the second half was determined by injecting a pulse of short rise-time and measuring the time taken for it to return from the open-circuited end. This gave a velocity of 0·41 ft/ns (0·12 m/ns). (This should be compared with 0·50 ft/ns (0·15 m/ns) in the line before loading by socket capacitance and 1 ft/ns (0·30 m/ns) which is the velocity in an unloaded air-spaced line.) It was therefore deduced that, at each tapping point, the loading produced by the socket alone was 1·93 pF. The impedance of the lines in this condition was 31 Ω.

A similar pulse was then added to the video waveform during the line-blanking period. By this means the velocity of signals within the first ('working') half of the store assembly was measured and found to be 0·32 ft/ns (0·098 m/ns). Since capacitative loading reduces velocity and characteristic impedance by the same factor, it was deduced that the impedance of the fully loaded line was 24 Ω. This corresponds to a total loading of about 5·7 pF

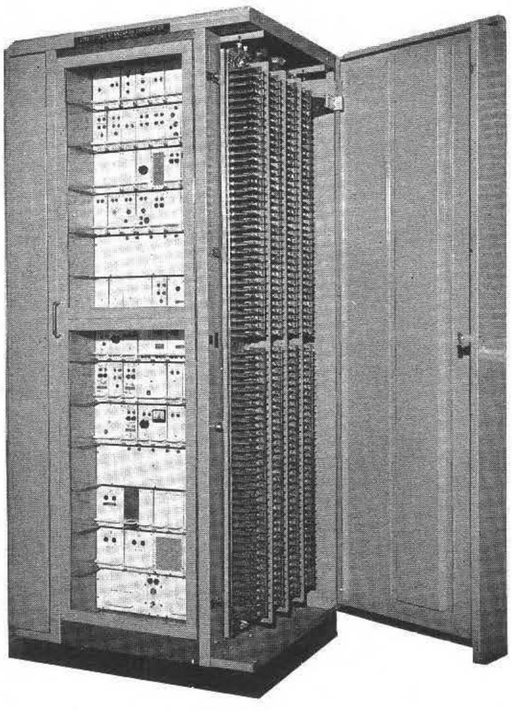

*Fig. 19 — Prototype machine showing half of store assembly*

at each tapping point, as compared with the value of 5 pF estimated previously.

## 10. Reflexions within the Transmission Lines

The effects of mis-termination of the transmission lines, and of mis-match between adjoining sections of the lines, will now be considered qualitatively.*

### 10.1 *Reflexions in the Input Line*

One effect of a mis-termination at the 'terminated' end of the input line is to produce an echo whose delay is a function of horizontal position in the converted picture. The total time taken to traverse the line is about 140 ns and the direct and reflected signals are therefore separated by a maximum interval of about 280 ns (i.e. about three picture elements). Virtual imperceptibility of such echoes requires the level of the reflected signal to be more than 32 dB below that of the direct signal corresponding to a maximum permissible mis-termination of $\pm 5$ per cent. No echoes were, in fact, perceptible in pictures derived from the complete converter. There is, however, a further consequence of the reflexion of video signals in the input line.

When a store-unit input-switch closes, current is taken from the transmission line as the store capacitor is charged, and the video voltage is correspondingly reduced. This current is shown diagrammatically in Fig. 20(*a*). It will be seen that although charging is virtually complete after 30 ns the switch is kept closed for 80 ns in order to render the stored charge independent of the precise value of the store unit's charging time-constant; the total period between clock pulses is, for the 625-line standard, 91 ns.

Whilst the store unit is taking current, a voltage pulse is produced at its input. This 'charging pulse' may typically be of the form shown in Fig. 20(*b*). Its duration is approximately 30 ns, and its peak magnitude about a third of the undisturbed video-signal level.

A pulse of this type is propagated in either direction along the transmission line. If the line is mis-terminated at either end or contains discontinuities, reflected charging pulses may be returned to the region in which they were generated at such times as to interfere with the signal being stored by the circuit that produced them, or by the next or even the next-but-one store to operate. The phenomenon will be discussed for the case where a single reflexion is produced by mis-termination at the non-driven end of the transmission line. The effect of a reflected charging pulse is then to reduce the magnitude of the reflected video-signal encountered at a given store unit for about 30 ns in each 91 ns period. Because of its short charging time-constant, the signal stored by a store unit is determined only by those signals reaching it during the latter part of the 80 ns period for which its input switch is closed. Thus a store unit is affected by a reflected charging pulse only if the unit's position along the line is such that it is subject to the pulse during this critical period. Fig. 21 shows the factor by which the reflected wave is effectively reduced, as a function of the position of the store unit. The right-hand peak corresponds to store units that are affected by the reflexion of their own charging pulses; the central and left-hand peaks correspond to store units affected by the reflexions of pulses from the previous and next-to-previous store units respectively. Thus, in plain areas of the picture, the change of video level associated with reflexion from a mis-termination is modulated by the reduc-

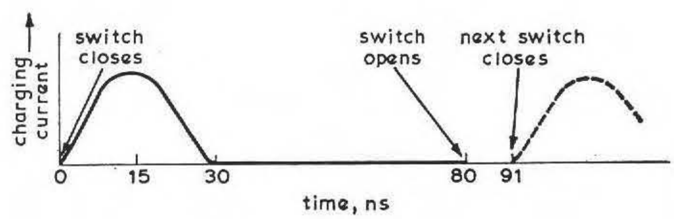


*Fig. 20(a) — Store unit charging current*

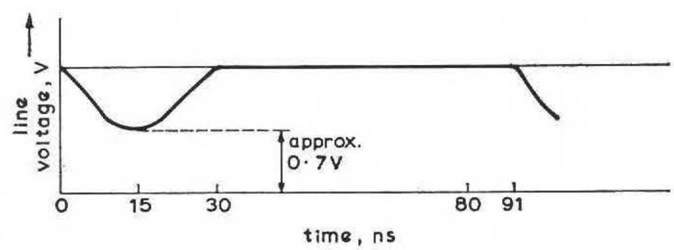


*Fig. 20(b) — Voltage pulse produced at input to store*

* This section includes an outline of some considerations by G. D. Monteath.

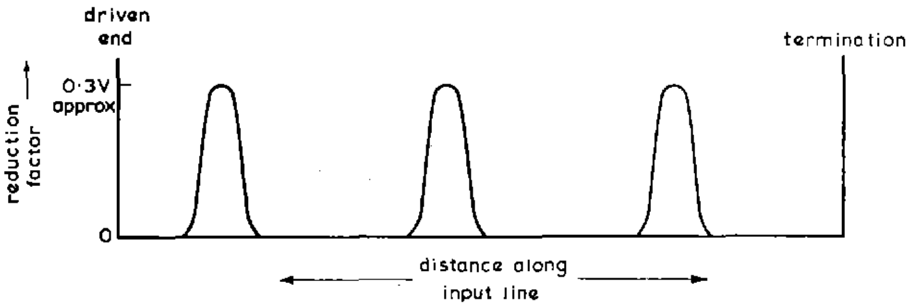


*Fig. 21 — Reduction of reflected video-signals by reflected charging pulses, measured at the times when store switches operate*

tion factor shown in Fig. 21, and non-uniformity of video level is produced in the form of broad vertical bands.

Where a single discontinuity occurs abruptly somewhere within the line, effects similar to those described apply between the discontinuity and the driven end. Between the discontinuity and the non-driven end of the line, banding still occurs, but now results from reflexion of the charging pulses that are propagated towards the driven end.

The spectrum of the charging pulses has been estimated as extending to 30 Mc/s, and it is therefore important to achieve low values of series inductance and shunt capacitance in the termination of the line, in order to maintain satisfactory matching up to this frequency.

In the prototype converter, banding attributable to reflexions in the input line was just perceptible to trained observers, but was not sufficiently severe to cause significant degradation of the overall picture quality.

### 10.2 *Reflexions in the Output Line*

The wanted signal on the output line consists of a train of discharge pulses produced at the times of closure of store-circuit output switches; the pulses are of similar shape and duration to the charging current pulses shown in Fig. 20(*a*), and the train is amplitude modulated by the required video-signal.

If a discontinuity occurs within the line, all discharge pulses originating at store locations between the discontinuity and the termination (see Fig. 10) are proportionately modified in amplitude as they pass the discontinuity. Pulses originating near to the discontinuity but on the other side, and travelling towards the output amplifier, are equally modified by the superimposition of reflected pulses of negligible delay. As the point of origin moves away from the discontinuity towards the output amplifier, a delay accumulates between the required discharge pulse and the reflected pulse. The effect of a discontinuity in the output line is thus to produce an echo whose delay is a function of horizontal position; but, as stated in connection with the input line, no such echoes were in fact perceptible in the converted picture, indicating that a satisfactory degree of transmission-line uniformity had been achieved.

## 11. Conclusions on Part II

The design of a transmission-line system for a line-store converter is largely determined by the parameters and layout of the store circuits which it feeds, together with the requirement of low inter-line capacitance. In turn, the method chosen plays a deciding influence in the design of the store assembly. Two suitable types of transmission-line have been developed. Of these the strip-line system is preferable, on grounds of cheapness, reliability, and ease of construction.

## 12. Acknowledgements


The authors wish to acknowledge their indebtedness to Messrs V. G. Devereux, F. A. Bellis, and G. C. Wilkinson, who were responsible for much of the work upon which this monograph is based.


## 13. References


1. British Patent Specifications, Nos. 928,730 and 966,318.
2. Rainger, P., and Rout, E. R., **Television Standards Conversion using a Line-store,** Proc. I.E.E., 113, No. 9, pp. 1437–56, Sept. 1966.
3. Rainger, P., **A New System of Standards Conversion,** I.E.E. Television Conference Digest, pp. 170–2, May–June 1962.
4. Lord, A. V., and Rout, E. R., **An Outline of Synchronous Standards Conversion using a Delay-line Interpolator,** I.E.E. Television Conference Digest, pp. 167–70, May–June 1962.
5. Rout, E. R., British Patent Application No. 26838/63.
6. Oliver, B. M., **Directional Electromagnetic Couplers,** Proc. Inst. Radio Engrs., 42, No. 11, pp. 1686–92, 1954.
7. Monteath, G. D., **Coupled Transmission Lines as Symmetrical Directional Couplers,** Proc. I.E.E., 102B, No. 3, pp. 383–92, 1955.
8. Grieg, D. D., and Engelmann, H. F., **Microstrip—a New Transmission Technique for the Kilomegacycle Range,** Proc. Inst. Radio Engrs., 40, No. 12, pp. 1644–50, 1952.
9. Dukes, J. M. C., **The Application of Printed Circuit Techniques to the Design of Microwave Components,** Proc. Inst. Radio Engrs., 105B, No. 20, pp. 155–72, March 1958.
10. Dukes, J. M. C., **An Investigation into some Fundamental Properties of Strip Transmission Lines with the Aid of an Electrolytic Tank,** Proc. I.E.E., 103B, No. 9, pp. 319–33, 1956.

# APPENDIX*

## MUTUAL CAPACITANCE BETWEEN STRIP TRANSMISSION-LINES

Fig. 22 represents a cross-section through a pair of strip transmission-lines A and B each of which have spacing $d$ between the strip and the ground plane and are separated by a distance $D$. Line A carries a charge $+q$ per unit length and it is assumed that the lines are air-spaced.

So far as electrostatic fields are concerned, the ground plane may be replaced by images of the two strip conductors in the manner indicated in Fig. 23 provided that the line PQ which replaces line A also carries a charge $q$ per unit length. The strip lines A and B may thus be replaced by the balanced transmission-lines PQ and RS.

If $D$ is sufficiently large compared with the other dimensions, the charges may be regarded as located at the centres of the strips P and Q. Then the electric force $E$ at the centre of the space between conductors R and S is the resultant of the two forces $E'$ produced by $+q$ and $-q$ separately.

i.e. $$E = 2E' \operatorname{Sin} \theta$$

where $\theta$ is the angle between the forces $E'$ and the neutral plane.

$$= 2 \frac{q}{2\pi \varepsilon_o \sqrt{D^2 + d^2}} \cdot \frac{d}{\sqrt{D^2 + d^2}}$$

$$\simeq \frac{qd}{\pi \varepsilon_o D^2}$$

since $D \gg d$.

Returning now to Fig. 22, the same force $E$ will be present within line B due to the charge on line A, and may be regarded as constant within line B.

Therefore the potential induced in line B is

$$Ed = qd^2/\pi \varepsilon_o D^2$$

Now if the lines have capacitance per unit length equal to $C_o$, and the voltage applied to a line A is $V$, then

$$q = C_o V$$

Moreover the voltage induced into line B will be

$$VC_m/C_o$$

where $C_m$ is the mutual capacitance.

Therefore:

$$VC_m/C_o = C_o V d^2/\pi \varepsilon_o D^2$$

$$C_m = C_o^2 d^2/\pi \varepsilon_o D^2$$

In the absence of fringing fields, $C_o$ would be $\varepsilon_o w/d$ per unit length, where $w$ is the width of the strip. The existence of fringing may be allowed for by taking $C_o = \varepsilon_o w'/d$, $w'$ being termed the 'effective width'.

Fig. 24 is a graph which may be used to obtain $w'/d$ from $w/d$. $C_m$ may therefore be expressed as

$$\varepsilon_o w'^2/\pi D^2$$

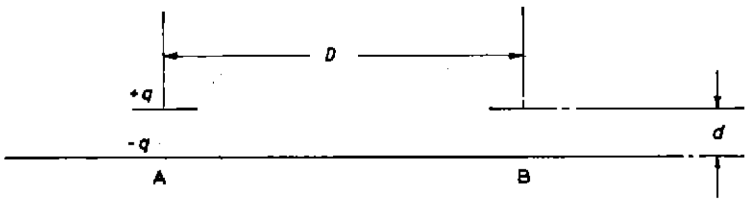


*Fig. 22 — Cross-section of two strip transmission-lines immersed in a uniform dielectric*

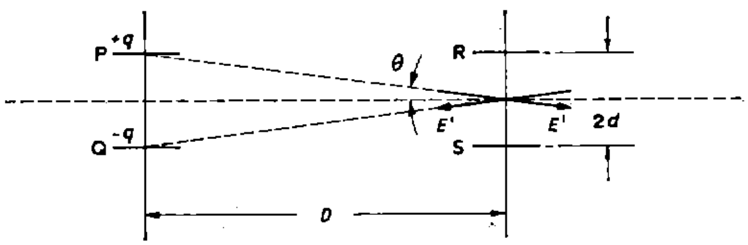


*Fig. 23 — Cross-section of a pair of transmission lines having a field pattern similar to that of the arrangement shown in Fig. 22*

* This Appendix is based on work contributed by J. B. Izatt.

Using the dimensions of the lines constructed for the prototype machine, we obtain $w/d=3$, so that $w'/d=5\cdot3$. So, since $w=\frac{3}{16}$ in. (4·8 mm), $w'=5\frac{3}{16}$ in.

$D=\frac{15}{16}$ in. (23·8 mm), and $\varepsilon_o=8\cdot854\ 10^{-12}$

Therefore, $C_m=0\cdot35$ pF/m

$=5\cdot1$ pF per 48 ft (13·72 m).

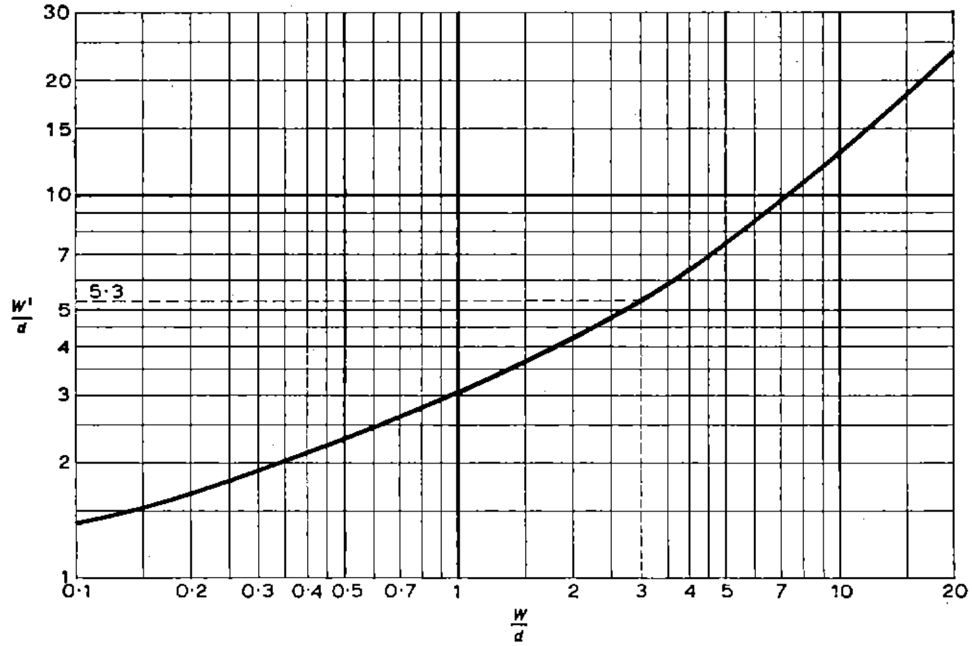


*Fig. 24 — Graph for obtaining $w'/d$ from $w/d$*

This graph was derived from information given in Fig. 8 of reference 10.

## BBC ENGINEERING TRAINING MANUALS

The following manuals by members of the Engineering Division of the BBC have been prepared primarily for the Corporation's operating and maintenance staff. They have been made available to a wider public so that the specialized knowledge and experience contained in them may be open to all interested in the engineering side of sound and television broadcasting.

**Sound and Television Broadcasting: General Principles**—K. R. Sturley, Ph.D., B.Sc., M.I.E.E. 45s. net, by post 46s. 5d. 378 pp.

*This manual explains the basic principles of sound and television broadcast engineering and operations.*

**Microphones**—A. E. Robertson, B.Sc.(Eng.), A.M.I.E.E. 75s. net, by post 76s. 3d. 357 pp.

*Explains the principles of microphones currently used in sound and television broadcasting.*

**Television Engineering: Principles and Practice**—S. W. Amos, B.Sc.(Hons.), A.M.I.E.E., and D. C. Birkinshaw, M.B.E., M.A., M.I.E.E.

*Vol. I: Fundamentals, camera tubes, television optics, electron optics.* 45s. net in U.K. only, by post 46s. 3d. 301 pp.

*Vol. II: Video-frequency amplification.* 45s. net, by post 46s. 3d. 270 pp.

*Vol. III: Waveform generation.* 45s. net, by post 46s. 3d. 223 pp.

*Vol. IV: General circuit techniques.* 35s. net, by post 36s. 3d. 277 pp.

These manuals are published by ILIFFE BOOKS LTD, DORSET HOUSE, STAMFORD STREET, LONDON, S.E.1, for the British Broadcasting Corporation, and are available from the publishers or from BBC PUBLICATIONS, 35 MARYLEBONE HIGH STREET, LONDON, W.1.

## ENGINEERING TRAINING SUPPLEMENTS

| | | s. | d. |
|---|---|---|---|
| No. 1 | Some Transmitter Problems | 2 | 6 |
| No. 3 | Harmonic Distortion and Negative Feedback in Audio-frequency Amplifiers | 7 | 0 |
| No. 4 | Some Fundamental Problems in Radio Engineering | 4 | 0 |
| No. 5 | An Introduction to Wideband Aerials | 3 | 0 |
| No. 6 | Programme Meters | 3 | 0 |
| No. 7 | Basic Principles of Television Lighting | 5 | 6 |
| No. 9 | Frequency Modulation | 6 | 6 |
| No. 10 | Long-wave and Medium-wave Propagation | 4 | 0 |
| No. 11 | Lighting for Television Outside Broadcasts | 6 | 0 |
| No. 12 | Transistors | 3 | 0 |
| No. 14 | Colorimetry | 4 | 6 |

These Engineering Training Supplements are available from BBC TECHNICAL PUBLICATIONS SECTION, HAREWOOD HOUSE, HANOVER SQUARE, LONDON, W.1.

*Published by the British Broadcasting Corporation, 35 Marylebone High Street, London, W.1. Printed in England by The Broadwater Press Ltd, Welwyn Garden City, Herts.*
*No. 6790*