

BAND I TELEVISION-TRANSMITTER DESIGN, WITH PARTICULAR REFERENCE TO THE TRANSMITTERS AT CRYSTAL PALACE

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SUMMARY

The paper describes television transmitting equipment designed to operate in two parallel chains to ensure reliability.

The vision transmitters are designed to work with considerable distortion and noise in the input signal, and the noise-rejection and clamp circuits developed for this purpose are discussed. The use of tetrodes in the output stage of the vision transmitter enables receiver-type valves to be used throughout the modulator, and permits the advantage of high-level modulation to be exploited.

The methods of phasing of both radio and vision frequencies are described.

The implications of colour television on design are discussed and the means adopted in these transmitters to assure faithful colour transmission are described.

(1) INTRODUCTION

The design of television transmitters is governed to a major extent by the types of valve available. Previous designs¹⁻³ used triodes as grid-modulated amplifiers, and the difficulties of driving and modulating these valves for the television service were overcome by expending considerable power in both the r.f. driving stages and the modulating stages. In a later design⁴ some effort was made to overcome these difficulties by adopting low-level modulation.

The relative merits of low- and high-level modulation have been discussed elsewhere,⁵ and there is little doubt that, if tetrodes are available as grid-modulated r.f. amplifiers, the choice will be, wherever possible, in favour of high-level modulation.

Air-cooled tetrodes of 10 kW anode dissipation are now available, and the Crystal Palace transmitters are examples of equipment which uses these valves and has been designed to operate on any of the three more popular television standards, i.e. the 405-line British standard, the 525-line F.C.C. standard and the 625-line C.C.I.R. standard.

By the use of tetrodes throughout, the whole of the r.f. amplifier stages for a 15–20 kW transmitter can be accommodated in one cabinet 2 ft 6 in wide, 3 ft 9 in deep and 7 ft high. The modulator, together with a large amount of waveform-correction apparatus and modulator power supplies, is housed in a similar cabinet.

The positive-modulation edition of the vision transmitter has an overall efficiency of 33% on peak white.

The sound transmitter for the BBC must operate on a.m. standards which exist little outside this country. Conventional class-B modulation of the anode of the final r.f. stage is employed; the final stage uses push-pull triodes in earthed-cathode connection, since a design of this circuit existed and was used as the penultimate stage of a sound transmitter already in use by the BBC.⁶

(1.1) Preliminary Design Considerations

To increase reliability of service from Crystal Palace, two transmitters are operated in parallel—each transmitter working into mutually independent sections of the aerial system. Co-

phasing of the two r.f. channels and of the two video-frequency channels is necessary and means for effecting this are described in Section 2.1.

Provision is made for switching both aerial sections to either transmitter and also for switching both transmitters to one aerial section.

Either transmitter removed from its aerial can be separately tested into a resistance load without interrupting service. In each transmitter, apart from care in generously rating all component parts, provision is made for switching out immediately all the more complicated arrangements which are included to cater for unusually bad input signals, and to provide the simplest possible chain of amplifiers as a basic minimum in case of fault, even at the expense of some performance. For example, the best possible stability of radiated black level is achieved by means of overall feedback operating in the transmitter clamp circuits; but in the event of fault the feedback can be removed, leaving a conventionally clamped transmitter in operation. Clamp-pulse failure is cared for by switching the transmitter to 'd.c. restoration'. The deterioration of performance experienced by this immediate switchover in both cases is measurable but not of such a magnitude as to cause annoyance at the receiver.

Where suspected units cannot be switched out of service, full metering of all independent current paths and full waveform monitoring facilities are provided.

Examination of BBC specifications and those of authorities in other countries reveals that, to conceive an economic vision transmitter substantially unchanged to meet the various standards, two basic editions are required, one for positive, and the other for negative, modulation. The difficulty of combining these into one edition is due to the greater peak power available from a given valve under negative-modulation conditions and the different power levels corresponding to 'black' or 'pedestal' in the television waveform. These two factors lead to rather different power supplies and operating conditions. Furthermore, for grid modulation in the negative sense the main vision-frequency amplifier is dealing with components of current corresponding to picture signals as it approaches the minimum of its current excursion. To preserve adequate reactive current-handling capability, this minimum is substantial and considerably in excess of the minimum adequate for the corresponding excursion into the synchronizing-pulse region in positive-modulation systems.

Colour television, while not yet commonly used, is sufficiently near for customers to demand that current equipment will deal with any colour system now known and likely to be operated within the life of the equipment. On the assumption that N.T.S.C. type signals will be used, we may assess the additional power requirements in the vision transmitter and design accordingly. These additional requirements are not easy to meet in the preliminary design stage. Detailed provisions will be discussed later, but it is clear that, while it is possible to meet the more stringent phase/frequency response required, it is not economically possible to design video-frequency and r.f.

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amplifiers with amplitude/phase characteristics adequate to deal with the overall requirement of only $1-2^\circ$ of phase shift when the signal changes from one comparatively black to a saturated colour picture of full luminance. This difficulty is overcome partially by liberal design and more completely by precorrection circuits.

In the general engineering of this type of band I transmitter the physical separation of vision-frequency circuits into units has been determined by a general consideration of the functions and specifications of particular circuit arrangements. Where it has been possible to assign a complete identity and performance

specification to a circuit arrangement, it has been designed as a unit to facilitate its integration into more general applications. This has the advantage of giving flexibility and of partial standardization at the expense of sacrificing overall gain per stage, since, in order to achieve flexibility, it has been necessary in some cases to have both input and output signals at standard level.

The design of power supplies for high-voltage power has undergone a significant development over the past few years. In turn, mercury-vapour and selenium rectifiers have enjoyed some popularity. In the present series of transmitters the designers

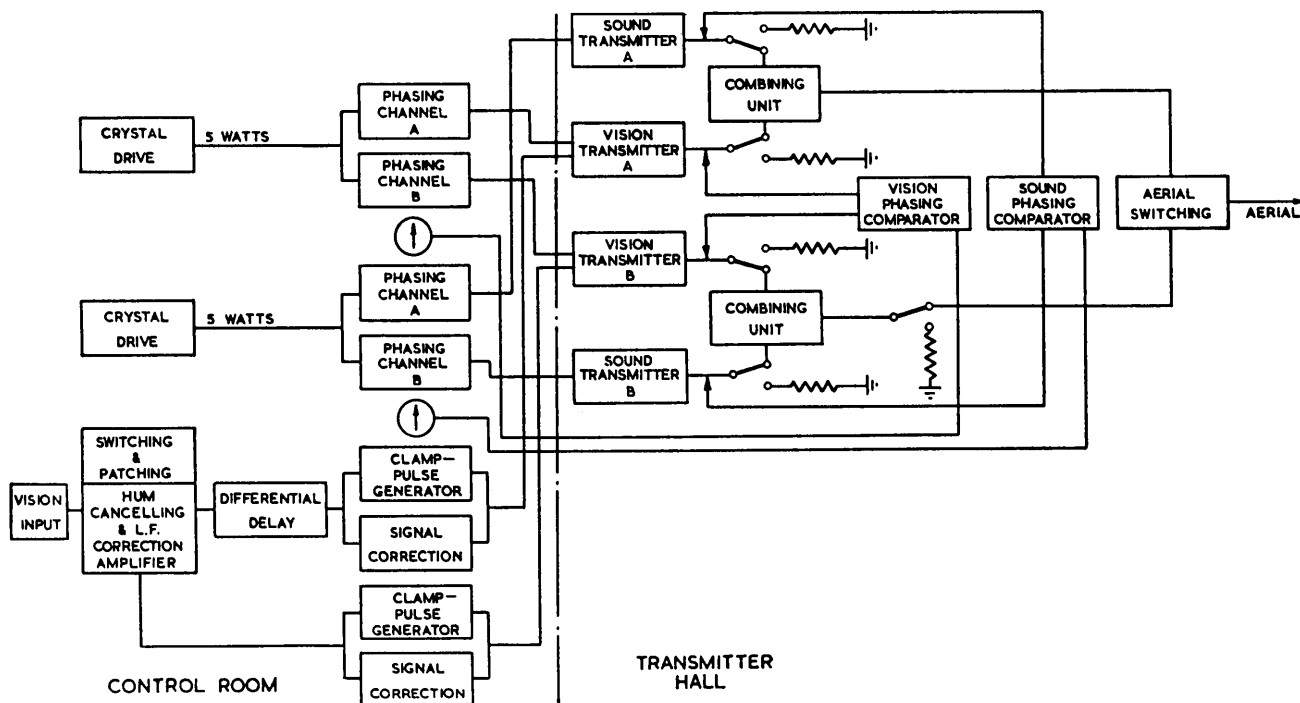


Fig. 1A.—Schematic of paralleling arrangements.

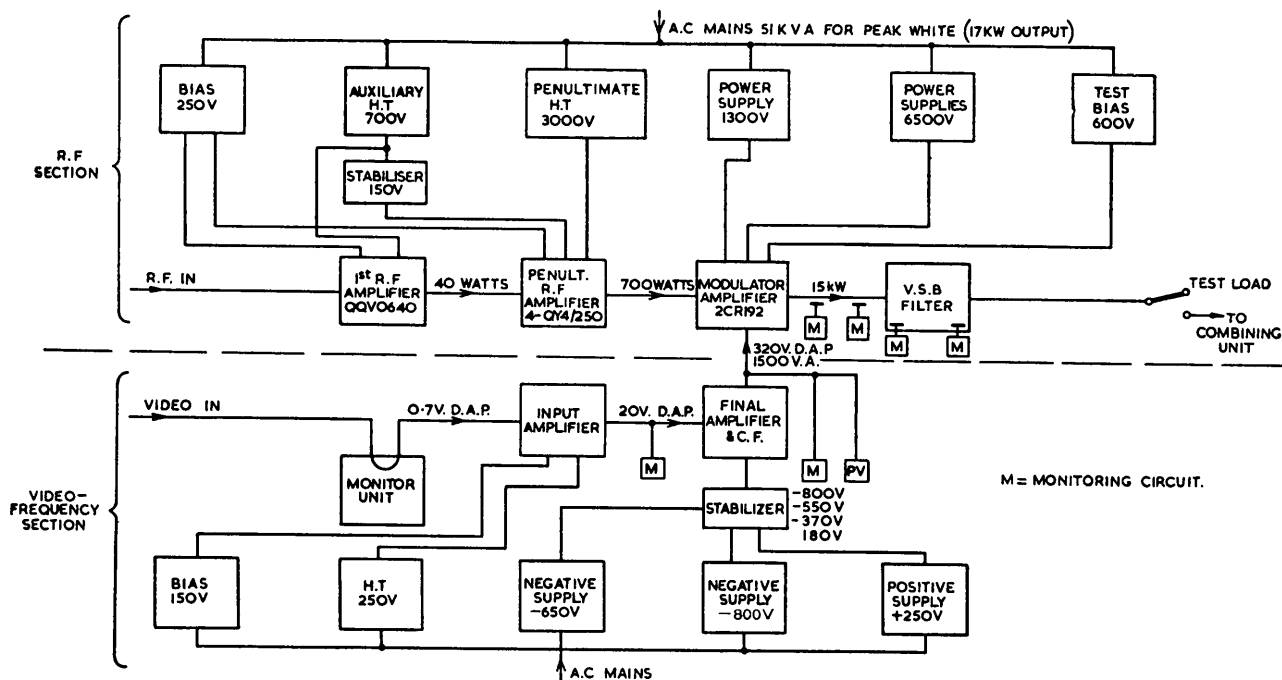


Fig. 1B.—Schematic of vision transmitter.

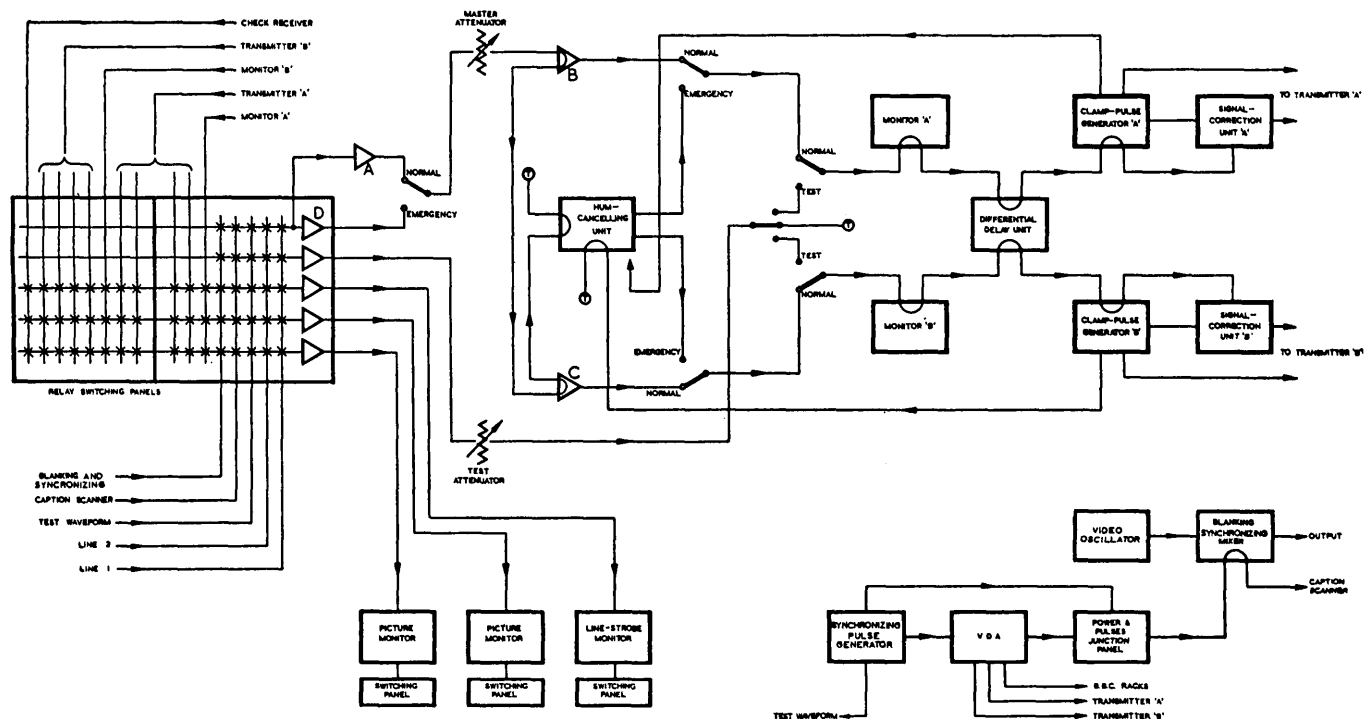


Fig. 1c.—Schematic of control room (vision transmitter).

favour rectifiers of the xenon type which have the high-efficiency characteristics of the mercury-vapour rectifier and the superior starting-up features of metal rectifiers.

Switching, control and protection equipment has undergone little basic change, although some attempt has been made to simplify this part of the equipment. It is significant in the progress of electronic engineering that the incidence of fault is now predominantly in the control and switching systems, where it is more difficult to anticipate failure by replacement.

(2) THE CONTROL ROOM

Fig. 1 includes a block schematic of the vision and sound transmitters and the signal-input switching and all test and monitoring equipment as used by the BBC. It will be seen that each transmitter receives its input vision signal from a series of units mounted in the control room.

(2.1) R.F. Drive and Phasing

Experience in paralleling sound-broadcast transmitters¹² operating in l.f., m.f. and h.f. bands has shown that, so long as a common drive is used and some method of phase correction is available, satisfactory performance is obtained. Distortion, noise, linearity, etc., appear always to be roughly a mean of the figures for the individual transmitters. Tests on a pair of Band I vision transmitters showed precisely the same trend, but it was evident that differential video phase control was also required.

The r.f. drive and phasing apparatus for both sound and vision transmitters is contained in cabinets of standard 19 in rack centres, 7 ft high. In these are mounted the vision and sound drives, the phasing amplifiers, power supplies and fault indication panels. Each unit is duplicated, and available for immediate switching. The relation of these units to the whole equipment is shown in Fig. 1A.

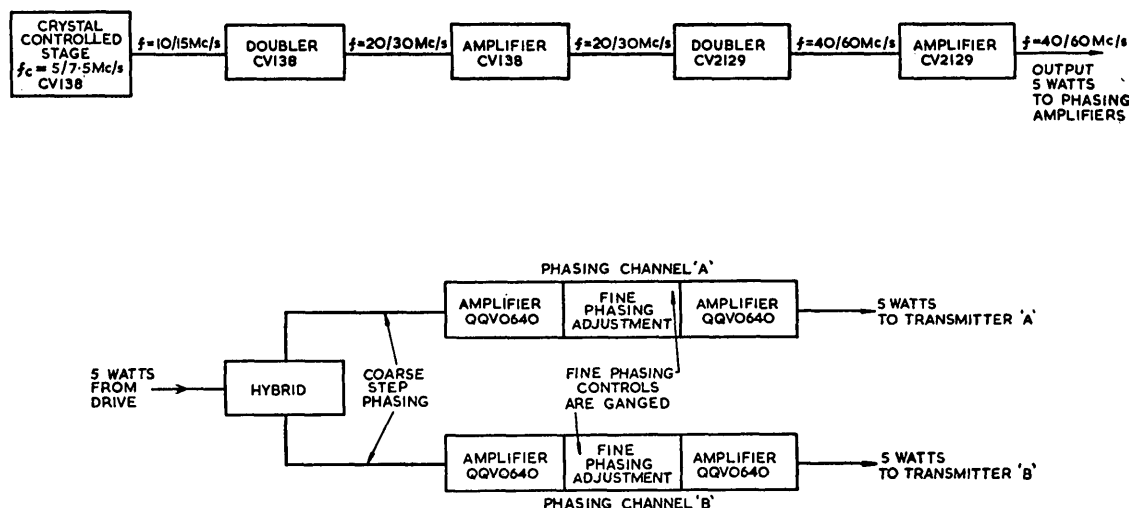


Fig. 2.—Schematic of vision and sound drives and phasing amplifier.

Both sound and vision drives are identical except for frequency (Fig. 2).

Phase indication is provided by comparing pick-up from three points in the feeder system. Three voltage vectors of equal amplitude are derived, one at 0° , one at 180° and a third at nominally 90° . The anti-phase pair are from one transmitter output feeder and the quadrature vector is from the other transmitter output feeder. The deviation from the nominal 90° of the quadrature vector is a measure of the phase difference between the output signals and is obtained by arithmetical subtraction of the resultants of vector additions of the $0^\circ + 90^\circ$ vectors and the $180^\circ + 90^\circ$ vectors. Adjustment of the fine phasing control to produce zero indicates the co-phasing of the transmitter output signals. A centre-zero meter gives the phase, in correct sense, of the output of transmitter A relative to that of transmitter B.

Phase difference can be measured up to 50° maximum, and adjustment is provided to give co-phasing with an accuracy of better than 1° . This is found to be a relatively stable adjustment requiring very occasional attention.

(2.2) Vision Input Equipment

The control room contains all the apparatus necessary to provide a hum-free distortion-corrected signal, and clamp pulses for the transmitter video-frequency circuits, so that the transmitter itself contains only straightforward amplifiers and limiters. Controls for setting the clamp and feedback levels are brought from the transmitter to the control desk. Two lines (one spare) from the studio are brought to a set of programme selector switches on the control desk. Three additional inputs are avail-

able for the caption scanner, the test waveforms and a spare. (See Fig. 1c.)

For an emergency, each transmitter has its own input units and clamp-pulse generator fed from different power sources. In addition, a 'normal-emergency' button on the control desk re-routes the signal from amplifiers A, B and C and passes it via amplifier D to the hum-cancelling and l.f.-correction amplifier. This unit corrects any errors in the incoming signal so that 'emergency' can also cover deterioration of the input signal. The unit gives twin outputs as required above.

All video switching is performed by means of relays, the cross-talk between lines being better than -50 dB.

(2.2.1) Distortion of Input Signals.

According to the BBC specification, input signals may have 20% superimposed hum, be differentiated by a time-constant of 2.4 millisecc ($16\mu\text{F}$ and 150 ohms) and have superimposed interference pulses up to 1.5 microsec wide equal in amplitude to the synchronizing pulses. The form of this distortion can be appreciated by reference to Fig. 3.

(2.2.2) Hum Cancellation and L.F. Correction.

The transmitter incorporates a system of hum cancellation and l.f. correction⁷ designed to remove the forms of undesirable modulation referred to above. The distortion is removed by detecting the black-level error, filtering and smoothing the

Table 1

Input frequency	Cancellation ratio
c/s	
20	80 : 1
50	300 : 1
100	60 : 1
200	25 : 1
500	6 : 1
800	3 : 1
1000	2 : 1

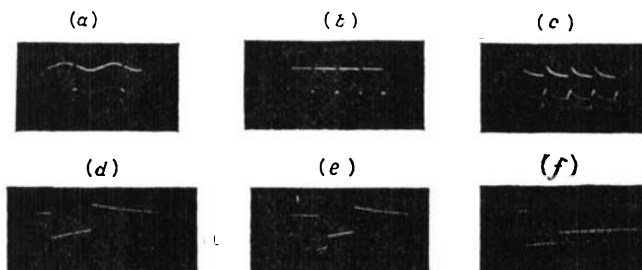


Fig. 3.—Television waveform distortion, 405-line system.

- (a) Frame distortion due to 20% 50 c/s modulation of peak-white picture.
- (b) Undistorted frame of horizontal black-bar picture.
- (c) Distortion of (b) due to CR coupling with time-constant of 2.4 millisecc.
- (d) Trace (c) expanded to show black-level distortion.
- (e) Trace (c) expanded to show frame-suppression-period distortion.
- (f) Trace (e) expanded.

resulting step waveform and feeding it back into the vision-frequency amplifier in phase opposition, so as to cancel the original distortion. The unit (Fig. 4) is necessary only when distortion due to hum is of such a level that it requires more than 30 dB attenuation of 50 c/s hum, which is the maximum that can be provided by conventional clamp circuits, or when differentiation of the waveform is excessive.

The unit operates satisfactorily with an input level from $+3$ dB

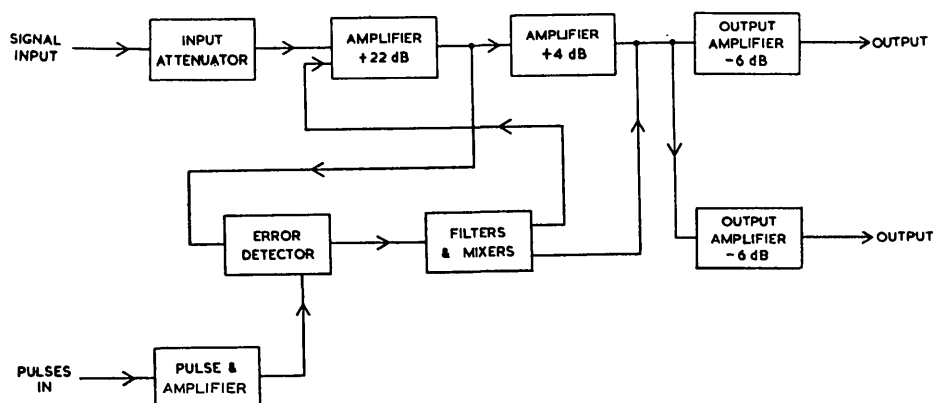


Fig. 4.—Schematic of hum-cancellation and low-frequency correction units.

to -18 dB on a standard level of 1 volt d.a.p. in 75 ohms, and gives standard output-signal level. Frequency response is flat to ± 0.2 dB up to 7 Mc/s.

As designed, the unit will cancel 50 c/s hum of up to eight times the signal amplitude with a 300 : 1 cancellation ratio; this ratio, which is optimum for 50 c/s components, varies with frequency, as shown in Table 1.

(2.2.3) Signal Correction.

The predistortion necessary to correct for the non-linearity of the r.f. grid-modulated amplifier is accomplished in the signal-correcting unit, which can provide synchronizing-pulse stretching and 'white' stretching. To maintain accurate onset of the various stretching operations, so that the overall linearity is maintained irrespective of picture content, all stages performing the correcting operations are black-level clamped, clamping signals being derived from the clamp-pulse generator unit. To obtain adequate linearity correction towards 'white', three variables are usually necessary, namely the magnitude, the onset and the law of 'stretch'.

The stretching operation is obtained in a parallel amplifier which operates in the non-linear region of its characteristics towards cut off.

The law of the correcting-signal output is adjusted by alteration of the linearizing feedback applied to this amplifier. The onset is adjusted by variation of the reference voltage to which the valve is clamped, and the magnitude of stretch is adjusted by varying the relative gain of the parallel paths.

(2.2.4) Clamp-Pulse Generation.

For normal service the clamp-pulse generator (Fig. 5) will receive input from the vision-signal source, and may therefore receive a distorted signal within the tolerances specified (see Section 2.2.1). Even under emergency conditions, when the hum-cancellation and l.f.-correction unit is in circuit, the input signal to the clamp-pulse generator may still contain noise pulses, and has therefore been designed with full noise immunity and with as much protection as possible against waveform distortion.

The clamp pulses are derived from a multi-vibrator, triggered by the lagging edge of the synchronizing pulses.

Curve A of Fig. 5 is the input waveform, showing random positive-going noise impulses. Curves B and C show the separated synchronizing pulses with the positive-going impulses removed by the delay-line method. Curve D is the waveform after differentiation; the time-constant of the leading edge (C_2 , R_2) is about 10 microsec and that of the lagging edge (C_2 , R_2 and R_3 in parallel) about 3 microsec. This circuit discriminates between synchronizing pulses and impulses of much shorter duration. Curve F shows the trigger pulses from V4, the pulse-amplitude discriminator. C_3 , R_4 (time-constant = 0.5 microsec) integrates the trigger pulses to give a self-adjusting threshold bias to V4. C_4 , R_5 have a short time-constant limiting the charge per pulse reaching C_3 , as otherwise V4 would discriminate between frame and line pulses. V5 and V6 are the pulse-forming multi-vibrator.

The pulse period is largely determined by the pulse amplitude at the anode of V5 and by C_6 , R_7 . MR3 limits this pulse amplitude, and as a result the pulse width—nominally 2 microsec—is stabilized to better than $\pm 5\%$.

The maximum delay of the clamp pulse relative to the lagging edge of the synchronizing pulse is 1.6 microsec for the delay line, plus the maximum width of a noise pulse—say 1.5 microsec—plus incidental circuit delays of 0.3 microsec. With a clamp-pulse width of 2 microsec this means that the lagging edge of the clamp pulse is 5.4 microsec after the lagging edge of the synchronizing pulse. The minimum porch width of 5.6 microsec ensures that no false clamping can exist under limiting conditions.

(2.2.5) Control Console.

Apart from normal operational switching, a considerable amount of test monitoring facilities is provided at the control console. Either of two 14 in picture monitors and a line-strobe waveform monitor can be switched to any of four inputs, four transmitter monitoring circuits or to a check receiver. One of the transmitter monitoring points can be switched locally through a monitoring amplifier to many test points in the vision-

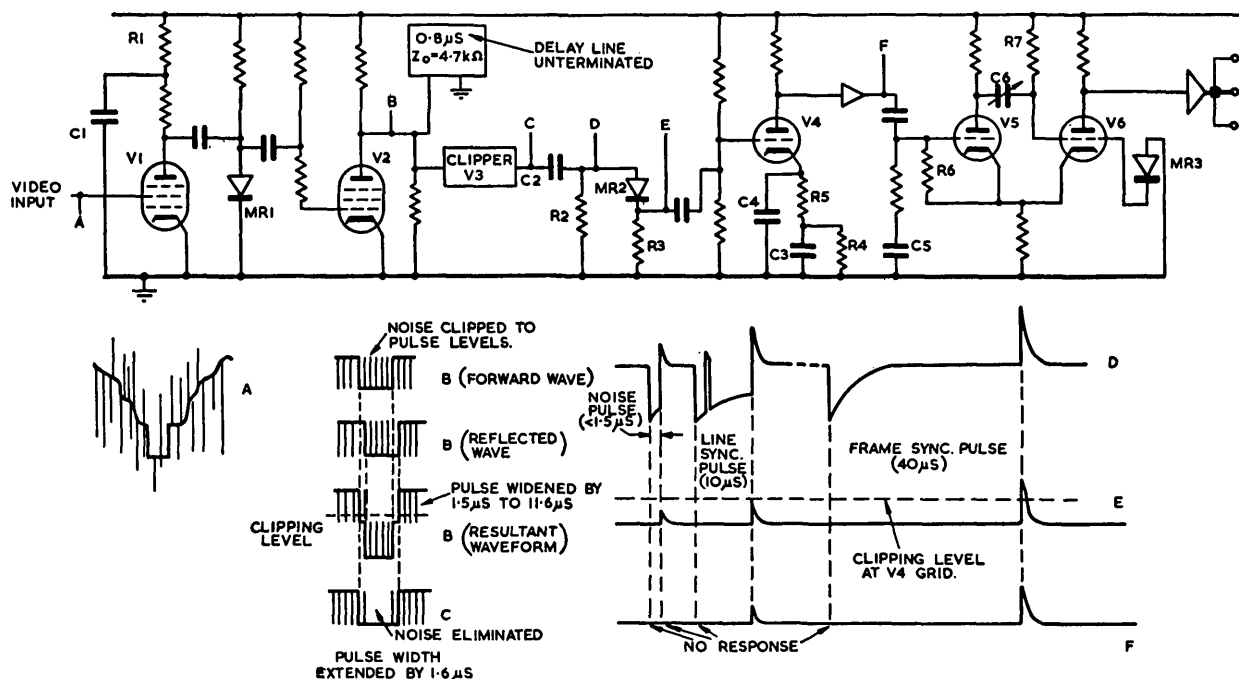


Fig. 5.—Clamp-pulse generator.

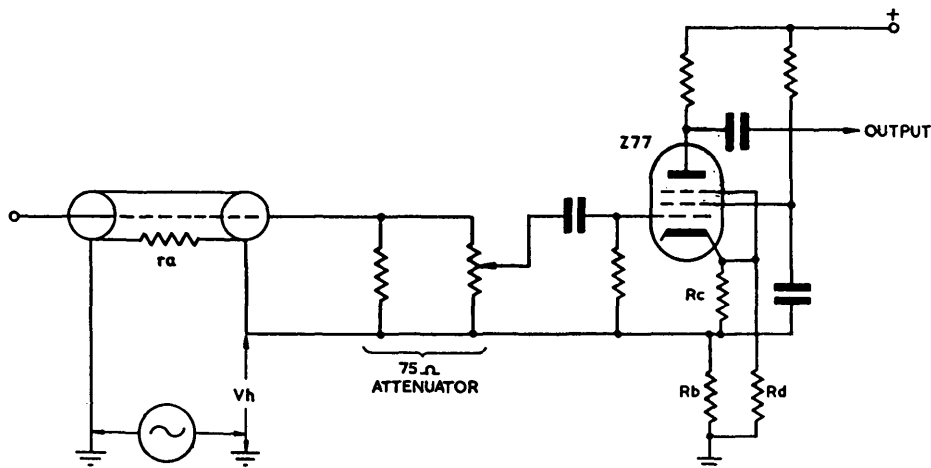


Fig. 6.—Modulation input valve circuit.

frequency chain and can likewise be used for local monitoring on a mobile monitor. Switching operations available in the event of fault to maintain service on one transmitter are provided; these are apparent in Fig. 1c.

(3) THE VISION TRANSMITTER

(3.1) The Modulator

(3.1.1) Input Stages.

Hum voltage between earthing points is liable to exist between the control room and the transmitter, and thus incoming signals will probably have superimposed hum, even if the signals have passed through the hum-correction unit in the control room.

A simple hum-cancelling arrangement has been built into the input stage of the modulator with negligible complication (see Fig. 6). By connecting the outer of the coaxial feeder to the transmitter earth through a resistance R_b , which is much larger than the series resistance of the feeder, the major part of the hum voltage due to the different earth systems will appear across it. This unwanted hum voltage therefore appears between cathode and earth, while the signal voltage with greatly attenuated hum will appear between grid and cathode. The screen and suppressor are connected to follow the cathode, so that only a very small hum voltage appears at the anode.

The resistor R_d is introduced to act in conjunction with the bias resistor R_c to provide cancellation of the residual hum in the output. Over a selection of 12 valves and without adjustment of any of the circuit component values, hum rejection exceeding 60 dB was obtained, while optimum adjustment gave more than 80 dB.

(3.1.2) Final Clamp and Black-Level Feedback.

The input signal to the d.c.-coupled final amplifier and cathode-follower is clamped to ensure a constant radiated black level. The synchronizing-pulse amplitude at the modulator output is more than sufficient to drive the transmitter to cut-off. Stable black level therefore also ensures stable synchronizing-pulse amplitude in the radiated signal.

However, some l.f. distortion, hum and long-term drift in black level may occur subsequent to clamping, and a black-level feedback system⁸ is incorporated to minimize these effects (Fig. 7A).

The transmitter output signal is rectified (waveform A). Since the transmitter is cut off during synchronizing pulses, any variation of black level corresponds to a proportional variation in synchronizing-pulse amplitude. This signal is clamped by clamp II (waveform B), the error in black level being transferred

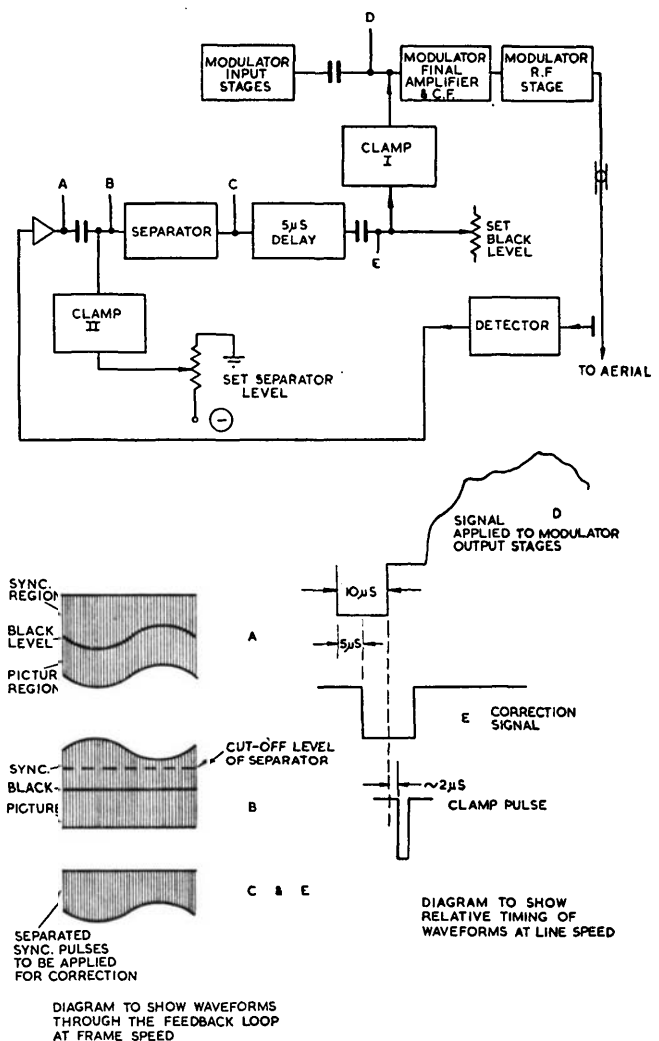


Fig. 7A.—Clamp and black-level feedback system.

to synchronizing level. The picture signals and part of the synchronizing pulses are removed by the separator (waveform C). The inverted and separated pulses are delayed by 5 microsec and used as a correction voltage for the main clamp reference, each correction pulse being coincident with a clamping pulse.

The level to which 'black' is clamped is thus jointly determined by the 'set black level' control on the desk and the correction-pulse amplitude. This technique has avoided the necessity to use a d.c.-coupled feedback loop common to earlier systems. This system will attenuate error signals at frequencies up to about one-fifth of the clamping-pulse recurrence frequency.

At the instant of clamping, a fraction of the correction-signal amplitude will be added to the signal. If this fraction is k and the loop gain is A , the criterion for stability can be shown to be

$$k \leq \frac{2}{1+A}$$

In the limiting condition black-level transients will produce a train of damped oscillations alternatively positive and negative for successive lines. For this train to decay to, say, 1% in ten lines, k must be reduced by a further 17%. In practice, k is adjusted to be 70% of the limiting value. For the case in point, $A \approx 10$ and $k = 0.13$.

To minimize the effect of the relatively large time-constant of the clamp which is determined by the above conditions, a modified type of clamp bridge has been introduced (Fig. 7B). This gives

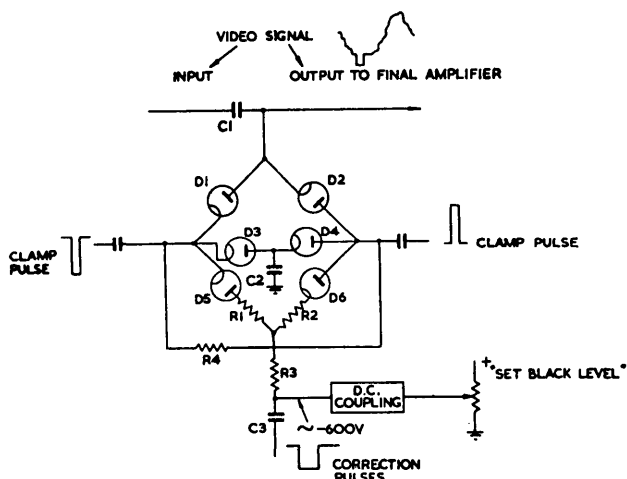


Fig. 7B.—Detail of main clamp circuit.

rapid action on errors occurring before the clamped capacitor C_1 and a slower action on feedback correction. The diodes D_1 – D_4 are part of a conventional system of short operating time-constant (0.4 microsec), the diodes being operated at high current during the clamping period to reduce the impedance as far as possible. C_2 is approximately $10C_1$. The diodes D_5 and D_6 also conduct during the clamping period, but at a much lower current. Correction is applied via these diodes at a time-constant determined by R_1 , R_2 , R_3 , C_1 and C_2 , and this is adjusted for stable operation.

The feedback reduces black-level deviation by 18–20 dB on a long-term basis.

(3.1.3) Final Amplifier and Cathode-Follower.

The availability of 10 kW tetrodes (CR192) suitable for grid modulation has made possible high-level modulation with a parallel arrangement of naturally cooled 25-watt tetrodes. The use of tetrodes in shunt-regulated amplifiers and cathode-followers confers several advantages, owing to their high ratio of dynamic impedance to d.c. resistance. Constant-current operation can be approached, for example, in the shunt-regulated cathode-followers without the necessity for the regulator valve to handle the bulk of the reactive current demand, as is the most

economic case when using triodes. Equal reactive-current handling by each valve of the shunt-regulated amplifier or cathode-follower is more easily obtained, and this results in greater conversion efficiency. With the screen-to-cathode voltage constant, greater anode-voltage swings are possible with a lower minimum anode voltage than is possible with comparable triodes. This has resulted in the possibility of being able to earth the h.t. positive line and in the saving of a stabilizer system. Fig. 11 shows that the peak demand of current due to modulated-stage grid current is supplied equally by both valves of the shunt-regulated pair, with only half the current demand on the shunt stabilizer system. This represents an improvement over earlier systems in which the tendency was towards making the shunt stabilizer an absorber-type circuit, resulting in a constant load equal in magnitude to the peak-loading condition.

The reduction of Miller capacitance has also confirmed advantages in the signal circuits. The price paid for these advantages is the necessity to provide stable screen supplies and negative feedback in the shunt-regulated-amplifier section, in order to recover the bandwidth lost in raising the dynamic load impedance. The d.c.-connected feedback network is a complication, but it has provided a further degree of gain stability and linearity in addition to recovering the bandwidth, and has resulted in a higher overall conversion efficiency for a given performance.

The use of a cathode-coupled pair, V_1 and V_2 , provides a useful high-impedance injection point for the feedback voltage, a high-impedance input to the amplifier and a significant degree of isolation from any residual hum on the –800-volt h.t. line.

The single pair of valves, V_2 and V_3 , cannot handle the peak reactive-current demand of the output stage and the inevitable circuit capacitances, so an additional regulator valve, V_4 , is included. On peak demands at the higher vision frequencies the a.c.-coupled regulator V_4 contributes equally with V_2 and V_3 to the load demand.

The shunt-regulated output stage V_6 and V_7 consists of five pairs of KT67 valves in parallel and can provide a peak reactive current of 0.55 amp for a mean h.t. current of 0.275 amp.

The path of the modulated r.f.-amplifier grid current is indicated by the arrows in the diagram, the only d.c. earth in the grid-current circuit being at the grid-current meter. The grid-current demand is shared equally by V_6 and V_7 , one half flowing via V_6 and V_8 and the other via the shunt stabilizer, V_7 and V_8 . Equal sharing is achieved by correct design of R_2 . The output impedance of the stage for this condition of operation is about 8 ohms.

The load on the modulator consists of $350 \mu\text{F}$ external and a further $150 \mu\text{F}$ due to the modulator output stage—a total of $500 \mu\text{F}$. The voltage excursion from black to white is about 190 volts. A reasonably shaped transient response of 0.13 microsec from black to white therefore requires reactive current of about 0.5 amp, in addition to the 250 mA transmitter-grid current.

The d.c. levels at the output terminals of the modulator are: peak white, –120 volts; black level (pedestal), –310 volts; synchronizing-pulse bottom, from –450 to –475 volts.

Peak-white limiting is provided in the final amplifier-cathode-follower coupling network by a valve driven from the output terminal, the time-constant of the limiting action being designed to give complete recovery of level during the synchronizing-pulse period and minimum deterioration of the picture during limiting.

(3.2) The R.F. Section

The connection between the penultimate r.f. amplifier and the grid-modulated amplifier calls for special consideration. The problems associated with this part of the circuit have already been commented upon and discussed.^{2,3,9} The main problem that arises is due to the flow of grid current in the grid-modulated

amplifier. In order to prevent non-linear regulation it is necessary to make the driving impedance very low compared with the peak loading due to peak grid current. Moreover, since the onset of grid current may take the form of a transient of very small rise time, the low driving impedance must either remain negligibly low over the video-frequency band or be aperiodic.

The problem has been solved in a number of ways. The most obvious and the least economical is to load the driving stage until the grid load represents only a small part of the total load. As adopted in early television transmitters,¹ this procedure also provides an aperiodic source impedance. In a post-war design of transmitter,² a hard-driven driver stage was used to produce a low impedance which the designers claimed was adequate, while in a still later design a cathode-follower driving stage was used to produce an aperiodic effective driving impedance of about 10 ohms.³

A clearer picture of both the problem and possible solutions can be obtained by drawing an analogy between this problem and a familiar video-frequency problem, namely that of providing a constant h.t. supply over the video-frequency band despite substantial changes of current—a problem usually solved by the use of constant-impedance networks or voltage stabilizers.

Suppose that the driving stage of the grid-modulated amplifier is a high-efficiency class-C amplifier with narrow bandwidth. If its output impedance could be made sufficiently low, and if a direct connection to the modulated valve could be made, this would form a reasonable practical solution, since if the output impedance of the driver restricted non-linearity to, say, 5%, the change of excitation voltage driving the grid-modulated stage over the range of modulating frequencies will be limited to 5%. In practice, this is not possible and the usual driving stage is coupled by a transformer to the output stage, in order to achieve a sufficiently low impedance; the coupled circuit or its equivalent is normally made of sufficient bandwidth. If the bandwidth of this stage is intentionally so limited that the maximum conversion efficiency is obtained, the driving-stage impedance will be aperiodic only over a limited band of frequencies; and if components of load current are demanded from it at frequencies outside its bandwidth, distortion due to the well-known ringing phenomena will occur.

When such a high-efficiency circuit exists or can be designed and it is necessary to drive a grid-modulated stage from it, we must—by analogy—first determine over what small band of frequencies the circuit may be regarded as a substantially constant resistance.

This is a direct analogy of the power supply which acts as a resistance source up to the frequency at which the smoothing components cause a significant change of impedance. In the low-pass circuit we extend the constant resistance over a larger frequency band by adding sections, and can proceed in exactly the same way in the band-pass case. In all cases, of course, $R = \sqrt{L/C}$, and in the band-pass case each tuned circuit is tuned to carrier frequency.

In practice, the application of this principle is not always simple. With modern valves of the type used in this transmitter the maximum power output is obtained at a considerable bandwidth. The physical complication of applying constant-impedance drive systems is also such that a compromise is often the better practical solution. In the Crystal Palace transmitters a small amount of damping with the simplest possible transfer connection was all that was necessary to meet the required performance. For full exploitation of valves, however, it would seem that the constant-impedance drive will in many cases offer a substantial improvement. This may be the case with Band III and Band IV transmitters, where the additional circuit elements can conveniently be made as line sections.

As is well known, the output obtainable from a given pair of push-pull valves working under class-B conditions is given by

$$P = \frac{KI^2}{16\pi C\Delta f}$$

where P = Power, watts.

I = Peak available emission current per valve, amp.

K = Circuit factor (1 for single-circuit, $\sqrt{2}$ for double-circuit and 1.5 for triple-circuit coupling).

C = Total circuit capacitance, μF .

Δf = Bandwidth between 3 dB points, Mc/s.

With this amplifier stage (which has a total circuit capacitance of $23 \mu\text{F}$ using CR192 valves) and a peak power output of, say, 18 kW, the power-capacitance product is $23 \times 18 \text{ kW} \cdot \mu\text{F}$ and is constant for a given bandwidth and emission; thus a saving of $1 \mu\text{F}$ in the circuit capacitance will increase the power output by about 800 watts. The problem of minimizing spurious circuit capacitance by careful design reconsideration becomes important.

With this transmitter the output circuit is triple tuned. For maximally flat response the input and output sections of the 'triple' are rigidly related, while the intermediate circuit can have any convenient LC ratio. The flexibility in the possible design parameters of the intermediate circuit provides an excellent aid to physical realization of the theoretical maximally-flat-response condition or any other condition chosen. For maximally-flat and Chebyshev response the circuit relations are shown in Fig. 8.

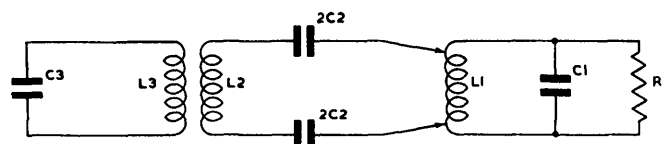


Fig. 8.—Realization of output circuit.

- (a) Maximally flat: 55 Mc/s , 3 dB, $f_c = 44 \text{ Mc/s}$
 $R_1 = 51.5 \Omega$ $C_1 = 280 \mu\text{F}$ $L_1 = 0.047 \mu\text{H}$
 $K_{12} = 0.153$ $C_2 = 17.5 \mu\text{F}$ $L_2 = 0.75 \mu\text{H}$
 $K_{23} = 0.088$ $C_3 = 23 \mu\text{F}$ $L_3 = 0.5 \mu\text{H}$
 Q-factor of primary circuit = 12.
 Q-factor of output circuit = 4.
- (b) Triple peaked: $V_p/V_s = 1.02$
 $R_1 = 51.5 \Omega$ $C_1 = 420 \mu\text{F}$ $L_1 = 0.31 \mu\text{H}$
 $K_{12} = 0.116$ $C_2 = 17.5 \mu\text{F}$ $L_2 = 0.75 \mu\text{H}$
 $K_{23} = 0.082$ $C_3 = 23 \mu\text{F}$ $L_3 = 0.57 \mu\text{H}$
 Q-factor of primary circuit = 12.
 Q-factor of output circuit = 6.

Obviously a small migration of values from the maximally-flat to the peak response condition is permissible, if necessary, without significant sacrifice of transient response.

The problem of handling both vision-frequency and radio-frequency currents in grid-modulated amplifiers entails careful design of all the current paths. To preserve the effective bandwidth of the modulated circuits the supply impedances must be aperiodic, and all connections, including earth returns, must be included in this consideration of aperiodicity. The screen-grid supply is stabilized at its source, and the cathode impedance is kept to a minimum to minimize r.f. degeneration.

On the modulation input circuit of the grid-modulated stage further precautions are necessary.

Assume a perfect modulator output impedance of R ohms from which is derived a perfect vision-frequency signal; it is then necessary to apply this signal between the grid and cathode of the modulated stage.

The capacitance of the grid-to-cathode circuit shunted by the non-linear-resistance component due to grid current represents the unavoidable load on the modulator. The connections between the modulator and modulated stage, and the chokes or

similar impedances used to enable the vision-frequency signal to operate in parallel on the modulated-stage grids without disturbing the push-pull r.f. driving conditions, present a series impedance which is predominantly inductive. Series inductance causes the voltage across the capacitance to rise with increasing frequency and increases the apparent capacitance loading on the source, thereby demanding more reactive current. Generally

speaking, the use of $\lambda/4$ lines as chokes tends to help in some cases, but whatever the form of the connection, either the capacitance loading and the voltage across the capacitance load increases rapidly or, if the characteristic impedance of the lines is low, the rise is less rapid but the capacitance load on the modulator is heavier over a greater frequency band.

Fig. 9 illustrate these effects and shows that a satisfactory

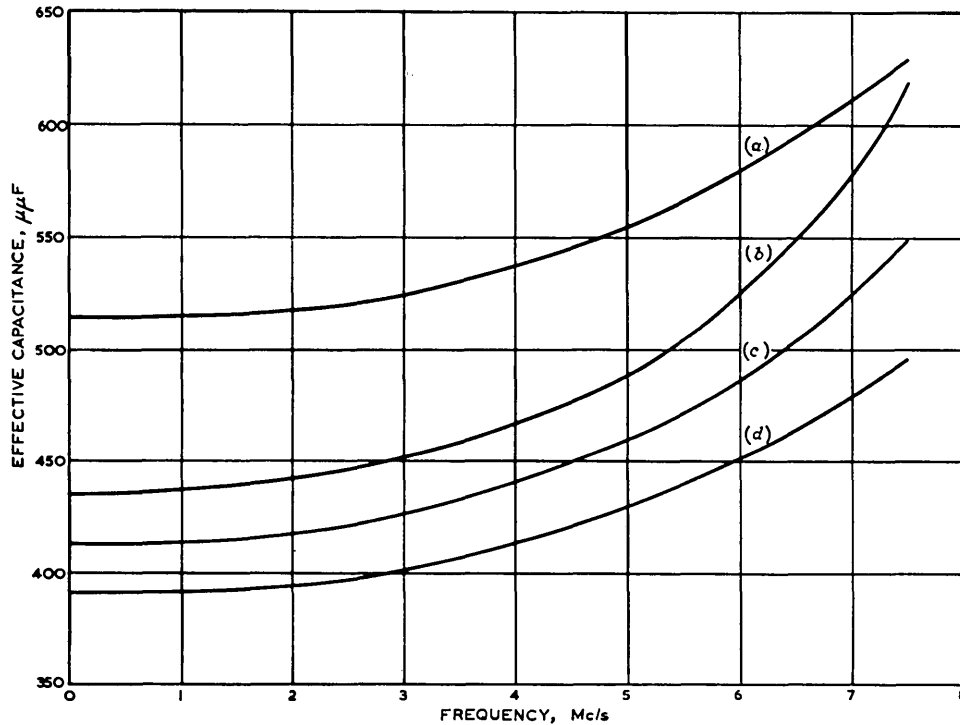


Fig. 9A.—Effect of using lines as impedances.

(a) $\lambda/4$; $Z_0 = 30 \Omega$; $f_r = 16 \text{ Mc/s}$.

(b) $\lambda/4$; $Z_0 = 60 \Omega$; $f_r = 13 \text{ Mc/s}$.

(c) $3\lambda/16$; $Z_0 = 60 \Omega$; $f_r = 14 \text{ Mc/s}$.

(d) $\lambda/8$; $Z_0 = 60 \Omega$; $f_r = 17 \text{ Mc/s}$.

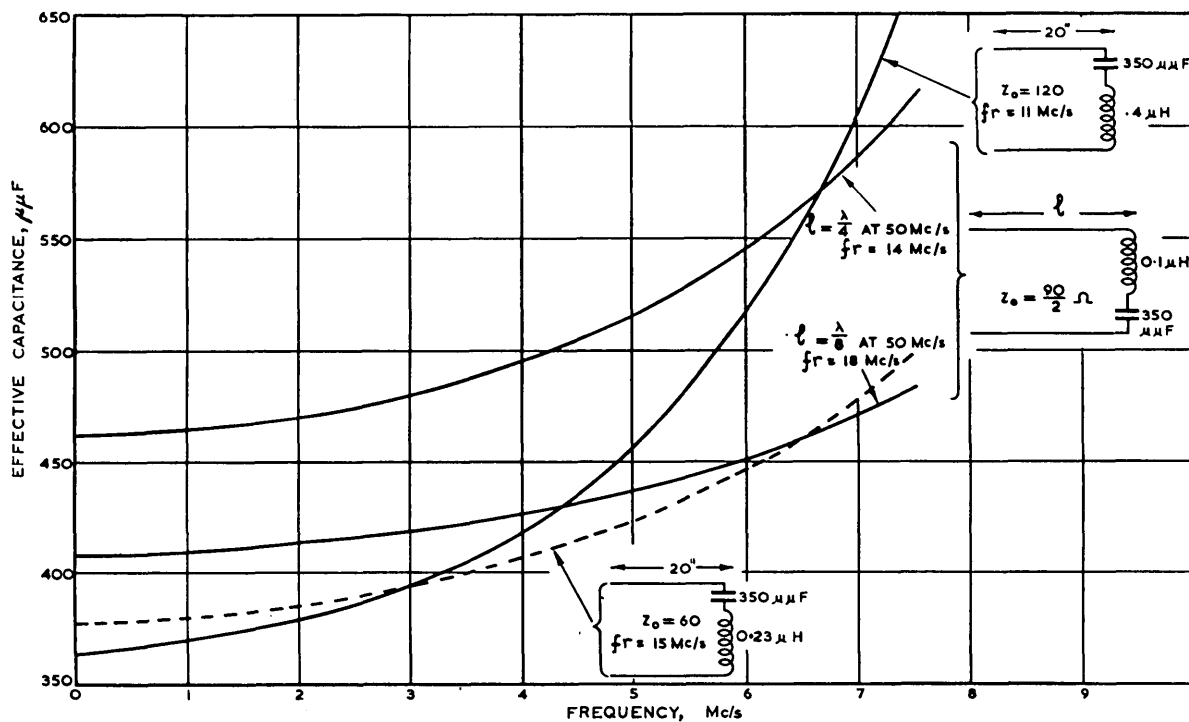


Fig. 9B.—Relative performance plotted against inductive loading.

compromise is possible at the expense of some modulator volt-ampere output.

(3.2.1) Description of the Vision Transmitter.

The vision transmitter (Fig. 1b) uses push-pull circuits with lumped circuit elements, which reduces costs, enhances reliability and gives a general appearance which engenders confidence in the minds of the operational staff.

Input from the external driving source is at 5 watts into 75 ohms and at radiated frequency.

The first stage is a push-pull-connected double-tetrode of type QQV0640, the tuned-grid circuit being inductively coupled to the feeder and the tuned-anode circuit capacitively coupled to the grids of the penultimate stage. H.T. and screen supplies to this stage are non-stabilized and are from a common supply unit. The filament heating is by alternating current.

The penultimate amplifier uses four type QY4-250 valves in push-pull-connected parallel pairs. This stage is situated immediately over the first stage, to enable short input connections to be used. These valves are self-neutralized at about 60-65 Mc/s; cross-over-capacitance neutralizing is necessary below this frequency, and parallel-capacitance neutralizing above it. The use of valves in parallel at very high frequencies is not considered ideal technically, but bearing in mind the economics of the problem, this solution proved satisfactory.

The penultimate stage uses an anode inductor comprising two vertical rods with a short-circuiting strap to give a preset inductance.

The grid of the final amplifier stage is coupled via a capacitive potentiometer to the penultimate-stage anode circuit. Damping is provided across this circuit, the overall damping impedance being about 600 ohms. Two peak rectifiers are provided to monitor the peak modulation-voltage excursion, i.e. in the positive and negative sense. This is a valuable aid in setting up the bias and driving voltages on the final amplifier. The anode circuit is triple tuned and consists of a primary inductance, of the clamped-up bar type, across the anodes of the pair of type CR192 tetrodes in push-pull, and is tuned by a screw-driven capacitor plate. A secondary balanced tuned circuit is inductively coupled to the primary inductance and directly tapped across the output circuit. This circuit converts from balanced to unbalanced 51.5-ohm output.

The filaments, which are d.c. heated, are fed via $\lambda/4$ lengths of Pyrotenax cable. The screen supply is built out to 100 ohms and the anode supply to 150 ohms by the conventional constant-impedance-network technique.

A switch in the grid circuit enables the modulation to be removed and replaced by a controllable bias supply, to permit the final stage to be set up to simulate any required modulation level.

In the power-conversion stages no hot-cathode mercury-vapour valves were permitted. In the intermediate power-level stages xenon valves are used, while the final h.t. supply is from selenium rectifiers mounted on trucks and housed in the air stream.

The control circuits are completely sequenced and interlocked, so that the transmitter can be brought on power by pushbuttons either at the transmitter or from the control desk.

Complete overload protection is provided, and indicator lamps on the front of the transmitter indicate location of fault.

Built into the front of the transmitter is the r.f. envelope monitor, giving a cathode-ray-oscilloscope display which enables modulation levels to be set up correctly. Radio frequency picked up from the feeder is directly displayed without the use of intervening valves or circuits, except a double-tuned circuit of 8 Mc/s bandwidth at the 3 dB points. The time-base permits display at either half line speed or half frame speed. The

monitor has its built-in power supplies and is a self-contained removable unit.

(3.2.2) Summary of Operating Conditions.

Operating conditions for the penultimate stage are as follow:

Valves	Four type QY4-250.
Power output (total)	700 watts.
H.T. voltage	2.95 kV.
Anode current	115 mA per valve
Power input	340 watts per valve.
Dissipation	165 watts per valve.
Screen voltage	150 volts.
Screen current	14 mA.
Screen dissipation	2.4 watts.

Operating conditions for the final stage are as follow:

Valves	Two type CR192.
Total circuit capacitance	23 μ F.
Power output	17 kW per pair.
Anode voltage	5.8 kV.
Anode current	2.85 amp per valve.
Anode input	16.5 kW per valve.
Anode dissipation	8 kW per valve.
Screen voltage	1.3 kV.
Screen current	180 mA per valve.
Screen dissipation	234 watts.
Anode efficiency	51.5%.
Valve efficiency	50%.
Grid current	250 mA per pair.

(3.2.3) Adjustment of Triple-Tuned Circuit.

The circuit diagram is shown in Fig. 8.

The coupling between primary and intermediate circuits is inductive and is adjusted by moving L2 relative to L3. The coupling between intermediate and output circuits is by direct tap on to the output circuit, and is adjusted by varying L2. This was found more convenient than moving the taps on the output circuit.

The output load, R1, is unbalanced, but the arrangement of the output circuit is such as to provide an unbalanced-to-balanced transformation. L1 consists of two feeder sections connected to an earth plate, C1 being connected between the free ends. The load presented to the intermediate circuit is balanced. Setting-up may be performed either by the poles and zeros method⁴ or by setting the individual parameters to the design values.¹¹ The second method was chosen and is described below.

For the triple-tuned circuit with a maximally-flat response the specified parameters are¹⁰

$$K_{23} = \frac{\Delta f}{f\sqrt{2}}$$

$$K_{12} = \sqrt{\frac{3}{2}} \frac{\Delta f}{f}$$

$$Q_1 = \frac{f}{2\Delta f}$$

where f is the centre frequency and Δf is the bandwidth between 3 dB points.

In addition, each circuit must be resonated at the centre frequency under the following conditions:

- Primary tuned with intermediate open-circuited.
- Intermediate tuned with both primary and output circuits short-circuited.
- Output tuned with intermediate open-circuited. For $f = 44$ Mc/s and $\Delta f = 5.5$ Mc/s, $K_{23} = 0.088$, $K_{12} = 0.153$, $Q_1 = 4.0$ and $C_1 = 280 \mu$ F.

In a double-tuned circuit with infinite Q-factor the fractional bandwidth between primary response peaks is equal to the coefficient of coupling. For finite Q-factors the error can be deter-

mined, but for values normally found in transmitter circuits the error can be neglected, which simplifies the measurement. Instead of using a variable-frequency oscillator and a separate detector, it has been found that a grid-dip oscillator can be used both to tune each circuit to the centre frequency and to determine the frequencies of the two peaks. In effect, the grid-dip oscillator acts both as the variable-frequency source and the detector.

The circuits are taken in pairs, i.e. primary and intermediate, then intermediate and output, and the coupling factors set as described below.

The primary circuit is correctly tuned when tuning the intermediate has no effect on the output at centre frequency. With the primary circuit thus tuned the output circuit is in tune when maximum output is obtained. With both primary and output circuits tuned, the intermediate circuit is tuned to give a symmetrical response. Residual errors are eliminated by adjustments as indicated by the theoretical error curves.³

(3.2.4) Summary of Overall Performance of Transmitter Vision-Frequency Channel.

Incoming Signal.

Nominal amplitude	1 volt d.a.p.
Nominal picture synchronizing-pulse ratio	70 : 30
Picture-signal amplitude tolerance	± 3 dB.
Synchronizing-pulse amplitude tolerance	$\pm 6-10$ dB.
Noise-pulse width	1.5 microsec (max).
Noise-pulse amplitude	± 1 volt d.a.p.
Distortion	The signal may be differentiated at a time-constant not less than 2.4 millise.
Hum	The incoming 50 c/s hum may rise to a value equal in amplitude to the signal.

Frequency Response.

Signal at low level	
Over the vision-input circuits and the modulator	0.5 dB at 5 Mc/s.
Overall	0.3 dB at 3 Mc/s.
Signal at full amplitude	
Overall	0.3 dB at 3 Mc/s.

Rise Time.

Test input signal rise-time, 0.1 microsec, sine-squared transition positive and negative going, full or part amplitude.

Output at the modulator	0.14 microsec.
Overall	0.15 microsec.

Black-Level Stability.

Better than 2% over an indefinite period.

Radiated Hum.

Expressed as a ratio of d.a.p. hum voltage to d.a.p. vision-signal voltage.

At 50 c/s	Better than 50 dB (where signal contains maximum hum).
At 300 c/s	Better than 52 dB.

(4) THE SOUND TRANSMITTER

Designed for a radiated carrier power of $4\frac{1}{2}$ kW, the sound transmitter is 15 ft long, 3 ft 9 in deep and 7 ft high.

The r.f. unit consists of three stages: input from the driving source is at 5 watts at radiated frequency; a double-tetrode (type QQV0640) in push-pull, supplies 40 watts to drive the penultimate amplifier, which uses two valves, also (type QY4-250) in push-pull, and supplies 700 watts to the modulated amplifier. Neutralizing is achieved by a variable capacitor connected between the screens of the pair of valves. The final stage, to which anode modulation is applied, uses a pair of type BR191 valves in push-pull, earthed-cathode connected with conventional bridge neutralizing. The 51.5-ohm output feeder is inductively coupled to this stage by a simple double-tuned coupled circuit.

Overload protection is by cathode-current relays and comparators using both rectified r.f. and modulation voltages from pick-up probes in the outgoing feeder.

The modulator is push-pull connected throughout and employs two stages of cathode-follower prior to the main modulator. Two stages of internal feedback are employed, and the arrangement of the stages and the feedback loops are shown in Fig. 10. Limiting diodes are also incorporated, but are not shown on the diagram.

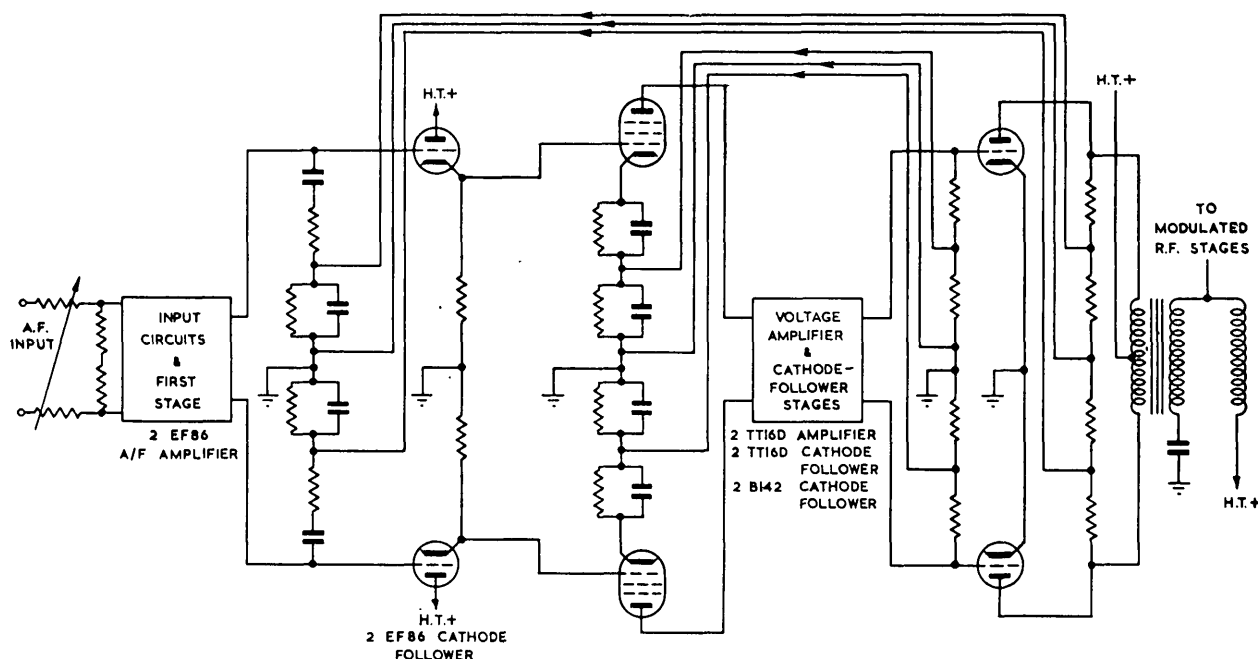


Fig. 10.—Sound modulator, showing feedback arrangements.

A built-in monitor, with cathode-ray-oscillograph indication, enables both a.f. waveforms and r.f. envelope waveforms to be displayed; this facilitates the correct setting-up of all circuits. The sound transmitter gives distortion of less than 1% from 30–40% modulation, less than 1.5% up to 70% modulation and less than 3% up to 90% modulation over the frequency band 30–10000 c/s. Input power is 14 kW at carrier frequency and 18.5 kW at 90% modulation.

Higher-power conversion rectifiers are of the xenon type. The transmitter is completely interlocked for sequence starting by pushbuttons at the transmitter and at the control desk.

Overload protection and indication lamps to show the position of faults are provided.

(5) OPERATIONAL RELIABILITY OF THE VISION SYSTEM

(5.1) Breakdown

The principle underlying the design has been to provide parallel paths wherever possible, so that in the event of breakdown the faulty path can be shut down and operation allowed to continue with the other. Where a common path is essential in the early part of the input equipment the 'normal-emergency' relays bring into use an alternative path complete with its own supplies. In addition, patching jacks are provided to permit the by-passing of relays which might develop faults.

The principle has also been followed of concentrating as much of the more complex equipment as possible in the control room, where it is accessible without breaking interlocks. In the black-level feedback system, where this principle cannot be followed, provision has been made to open the feedback loop and operate on simple d.c. restoration or simple clamping.

In the event of breakdown the operator's first duty is to press the 'emergency' button. If normal service is restored the fault must lie in the input equipment; if not, then in all probability one of the transmitters is at fault, and a quick check with the monitoring facilities provided will reveal which one. The operator can then shut it down from the console and continue normal programme at reduced level with the other. He can switch the

faulty transmitter to test load and apply test signals by way of the independent test channel provided; normal fault location procedure can then be carried out. Input-equipment faults can likewise be located without interrupting the programme.

(5.2) Clamping Stability

The clamp generator will accept pulse noise up to 1.5 microsec width and considerable hum and l.f. distortion. However, if pulse noise of greater width or abnormal distortion beyond the specified limits should be present on the incoming signal, mistiming of clamp pulses might result. To maintain service, all clamping must be eliminated. On the modulator this will be done by switching the clamp to d.c. restoration and in the control room by patching-out the signal-correction system. This will result in reduced power and some attenuation of synchronizing pulses, owing to the absence of stretching. Nevertheless, if the picture is important and viewable, it can still be radiated sufficiently well to continue the service.

(6) THE INFLUENCE OF COLOUR TELEVISION ON TRANSMITTER DESIGN

The transmission of colour information in addition to the conventional luminance information imposes additional demands on the transmitting system. If, for example, we consider the American N.T.S.C system, the additional demands are due to

- The presence of the colour synchronizing burst in the line-blanking period.
- The necessity to maintain the phase of colour signals within a few degrees over the complete amplitude range.

The presence of the burst in that part of the waveform normally used for clamping makes conventional 'hard' clamping unsuitable, owing to the distortion caused to the transmitted burst. Softening the clamp action by series impedance can alleviate this situation and is acceptable, provided that there are not many such clamps involved; the impedance discontinuity becoming

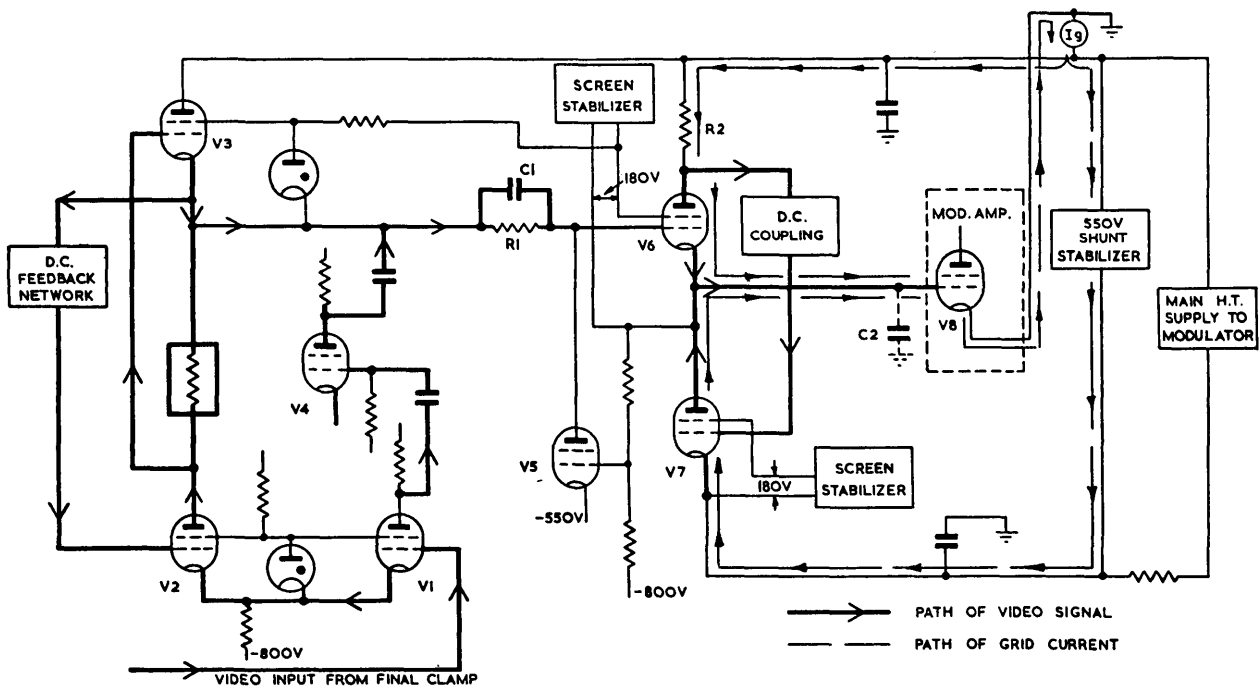


Fig. 11.—Vision-modulator output amplifier and cathode-follower arrangement.

significant as the number of clamps is increased above two or three. Alternative d.c. correcting means have been devised and will probably gradually displace conventional clamps where there is the probability of colour transmission.

The accurate phase/amplitude relation required by the colour-transmission system demands a degree of amplitude linearity far greater than that required for monochrome television and over a greater range of power demand if 100% saturated colours are to be transmitted. In the present series of transmitters care has been taken to ensure an adequate power-loading capability under true class-A conditions of the modulator. Any residual errors in the phase/amplitude relation can then be corrected by a pre-distortion method shortly to be published.

For colour transmission the transmitters will have the main clamp softened, but not to any great degree, since, with this requirement in mind, the driving-source impedance of the clamped stage has been made very low so that the disturbance to the impedance during clamping is almost negligible.

To achieve signal correction, i.e. synchronizing-pulse stretching and linearity correction, a unit alternative to that described earlier has been developed and is now available. This uses a system of d.c. correction without impedance discontinuity in the signal circuits.

(7) ACKNOWLEDGMENTS

Quite naturally, the work involved in the development and design of television transmitters is considerable, and their final design is due to a large number of people. The authors gratefully acknowledge the work done by teams under the leadership of Messrs. J. Sutton, J. E. Nixon, G. Partington, S. U. Nolan, H. Haywood and D. F. Bowers.

Thanks are also due to Mr. J. E. Nixon for assistance in preparation of the paper.

The authors desire to thank the Engineer-in-Chief of Marconi's Wireless Telegraph Co., Ltd., for permission to publish the paper.

DISCUSSION BEFORE THE RADIO AND TELECOMMUNICATION SECTION, 4TH APRIL, 1956

Captain C. F. Booth: This is indeed an historic occasion in respect of television in the United Kingdom; first, we have with us members of C.C.I.R. Study Group 11 from many countries, and secondly, the three papers presented this evening mark the passing of Alexandra Palace, which has given London its television service since 1936. It is indeed a tribute to the architects of the 405-line system, to the transmitter and receiver designers and to the British Broadcasting Corporation, that picture quality has improved almost continuously. Since the service was re-opened in 1946, great progress has been made; then there were a few thousand receivers, now there are upwards of six million, and the Corporation is providing a service covering over 93% of the population. Additionally, during the past few months, the Independent Television Authority has been set up and is providing a rapidly expanding service.

It was a bold step to transfer a well-founded service for the metropolitan area from Alexandra Palace to Crystal Palace. The reasons are explained in the paper by Messrs. McLean, Thomas and Rowden; ultimately a service will be available to upwards of one million additional people. The transfer has presented difficulties, and I should like the authors to indicate them and the precautionary steps taken.

I note from the paper by Messrs. Cooper and Morcom that the parallel technique has been solved in respect of Band I transmitters. It is an excellent achievement which should not be belittled, but I do wonder why the authors did not finish the job by fitting automatic phase control. Have they also solved the problem of paralleling Band III transmitters?

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Messrs. McLean, Thomas and Rowden state that they gave preliminary consideration to the interleaving of Band I and Band III aerials; I should like to know whether they have continued this work, and, if so, the results thereof.

I was recently privileged to inspect—with people from some 20 countries—the new Crystal Palace station; it is certainly a sound engineering job, and without doubt it will make a contribution to the television service comparable with that of its predecessor.

One final point: reference is made in the papers to the very desirable aim of achieving a design suitable for colour. I merely stress—as one paper says—that an agreement that the vision transmitter should be designed with the requirements of a colour signal of the N.T.S.C. type in mind does not imply that an N.T.S.C. system or, indeed, any colour system would be used. In fact, as is pointed out in the papers, a decision regarding colour standards is some way off.

Mr. C. E. Strong: I am not too sure about the advisability of parallel operation of transmitters in a station like this. No doubt it is effective with high-power broadcasting transmitters, but with rather modest-power transmitters including relatively complex input circuits, it might be another matter.

Parallel operation has, of course, the advantage that you might get a fault and not have even a momentary interruption of the programme. Against this, however, there are two transmitters to be similarly lined up at the same time for equal performance. The number of components in service is approaching double; there is extra complexity, perhaps, in phasing of the video and