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An Instrument for Measuring Television Signal-to-noise Ratio

by

S. M. EDWARDSON, GRAD. I.E.E.

BRITISH BROADCASTING CORPORATION

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FOREWORD

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CONTENTS

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 V.H.F./U.H.F. 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 for 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 U.H.F. 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*	NOVEMBER 1958
22.	*The Engineering Facilities of the BBC Monitoring Service*	DECEMBER 1958
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 V.H.F. F.M. 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 V.H.F. 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

AN INSTRUMENT FOR MEASURING TELEVISION SIGNAL-TO-NOISE RATIO

SUMMARY

The monograph describes an instrument which has been developed specially for the measurement of signal-to-noise ratios in television. It utilizes an input picture signal corresponding to a scene of uniform grey or white, and selects the central portion of the picture area by means of an internal gating circuit. Separation of the noise from the signal is effected by subtraction of a 'noiseless' replica of the gated picture signal and a locally generated sine wave is used in evaluating the ratio between the picture signal and r.m.s. noise levels. A range of signal to-noise-ratios from 9 dB to 51 dB can be measured, at any grey level, with an accuracy of $\pm 0 \cdot 5$ dB.

1. Introduction

1.1 *General*

The presence of random fluctuation noise in a television signal affects the picture in two ways; it causes random variations of picture brightness and it may interfere with the uniform flow of the synchronizing pulses, causing raggedness on vertical edges in the picture, or even frame slipping. Since it is now possible to design scanning circuits with an immunity to random fluctuation noise such that accurate synchronization is maintained even in the presence of noise which almost obliterates the visible picture, attention may in certain cases be restricted to the effect of the noise on picture brightness. The degree of disturbance to the viewer caused by random fluctuation noise may be determined[1, 2, 3] from a knowledge of the signal-to-noise ratio and the way in which the noise energy is distributed over both the grey scale and the video frequency spectrum. An instrument designed to measure one of the parameters of random fluctuation noise in television is described.

1.2 *Existing Methods of Measurement of Random Fluctuation Nosie*

A method of noise measurement which has been in use for many years consists of visual inspection of the waveform of the noise by means of the scanning-line display of a television waveform monitor. If the displayed waveform contains no picture-detail information, the noise will appear as moving 'grass' superimposed upon a uniform signal level and its quasi-peak-to-peak amplitude can be estimated by eye. This is possible because the probability of any noise peak exceeding a given level is found to fall to an insignificantly small value when that level is between three and four times the standard deviation (r.m.s. value). There is thus an empirical relationship between the peak amplitude estimated by the observer and the actual r.m.s. value. This ratio has been the subject of much controversy,[4] but is usually taken as 11 dB, or 17 dB when considering peak-to-peak values. If the noise waveform has a normal or Gaussian statistical frequency distribution, then the general acceptance of this empirical ratio implies that the visually determined peak value occurs (on average) only once in about 5,000 fluctuations.

A short experiment was carried out in which twenty skilled observers each made an estimate of the quasi-peak-to-peak value of the noise waveform displayed on a waveform monitor. The average value of their estimates was very close to a value 17 dB greater than the objectively measured r.m.s. value, but with a total spread of 5 dB and a standard deviation of $1 \cdot 6$ dB. Thus it may be seen that although the method is quick, it is nevertheless inaccurate.

Another method of noise measurement is similar to the one just described, but it differs in that for this measurement the oscilloscope display is arranged without any time base deflexion. This results in the noise appearing as a vertical line of varying intensity. This method gives somewhat improved accuracy but it is almost impossible to separate the noise from the television signal upon which it is superimposed, even if the television signal is merely describing a scene of uniform brightness.

Further improvements in accuracy are possible by the combination of oscilloscope displays with photographic recording, but such techniques are exceedingly slow.

It will have been noted that the above methods give no information with regard to the distribution of the noise energy throughout the video frequency spectrum and whilst for some purposes such information is essential, there are many other occasions on which the spectrum of the noise is already known, or is not of importance. An elegant means of measuring both the spectral distribution and the absolute energy of the noise has recently been developed by Weaver.[5] In this method, the video frequency spectrum is explored using a narrow-band radio receiver and the r.m.s. noise is measured at those discrete points in the spectrum where the energy contributed by the harmonics of the composite video signal is at a minimum. The overall signal-to-noise ratio is computed by a summation process from a number of such measurements taken over the video-frequency band. The method is extremely useful in those cases where a knowledge of the spectral distribution of the noise is required, but because the contribution by harmonics of the video signal cannot always be neglected when taking measurements in the middle and lower parts of the frequency spectrum, errors tend to occur when the apparatus is presented with a television signal having an overall signal-to-noise ratio in excess of 40 dB.

The instrument to be described here represents an attempt to combine rapid measurement with accuracy, but no attempt is made to measure the spectral distribution of the noise energy, only its r.m.s. value being determined. The apparatus cannot be described as simple, but internal complexity seems to have been well justified by the consequent ease of operation and its usefulness as an instrument for the routine measurement of signal-to-noise ratios.

2. The Basic System of Measurement

In order to measure the r.m.s. value (or energy) of the noise appearing in a television waveform, it is necessary first to isolate the noise from the picture signal. For this to be possible, the signal representing picture information during the time interval when the noise is to be measured must be uniform, i.e. it must represent a plain grey area of the television scene. The method adopted* is to select, by gating, only those parts of the waveform which correspond to the central portions of the active line and field periods. The waveform of the gated signal thus closely resembles the gating waveform itself, as shown in Fig. 1. The two waveforms are coincident in time and the gating waveform may therefore be regarded as a noise-free version of the gated video waveform. The noise may thus be isolated as a series of 'bursts', by subtracting the one from the other, their relative amplitudes having been carefully balanced.

The absolute r.m.s. value of this noise could be measured by an appropriately calibrated thermo-ammeter, but in fact the quantity to be measured is the *ratio* of picture signal amplitude to r.m.s. noise amplitude and an absolute r.m.s. calibration is not required. Since the picture signal amplitude is defined as the difference between blanking and white levels, whilst noise is expressed as an r.m.s. quantity, it is convenient to substitute at the input to the instrument a locally generated sine-wave signal which may be given a known relationship to both signal and noise. As illustrated in Fig. 2, the peak value of this sine wave is adjusted initially to equal the picture signal amplitude and then, by means of a calibrated attenuator, it is readjusted to provide the same r.m.s. voltmeter reading as that which was produced by the noise (both signals having passed *through the same gating circuit*). The signal-to-noise ratio is thus given by the product of the attenuation factor and the crest factor of the sine wave, or, in decibels, it is the sum of the attenuation measured in dB and 3 dB.

In the practical instrument the attenuator takes the form of a simple potentiometer and its gradations are marked in decibels 3 dB greater than the *actual* attenuation ratios to spare the user this complication. The signal-to-noise ratio can therefore be read off directly.

* Since completion of the work recorded here, it has come to the attention of the author that a Russian paper presented to the IXth Plenary Assembly of the C.C.I.R. describes another method, based upon subjective comparisons, which bears some resemblance to that described above: Annex D(XI)6, *Document* 11, Los Angeles, 1959.

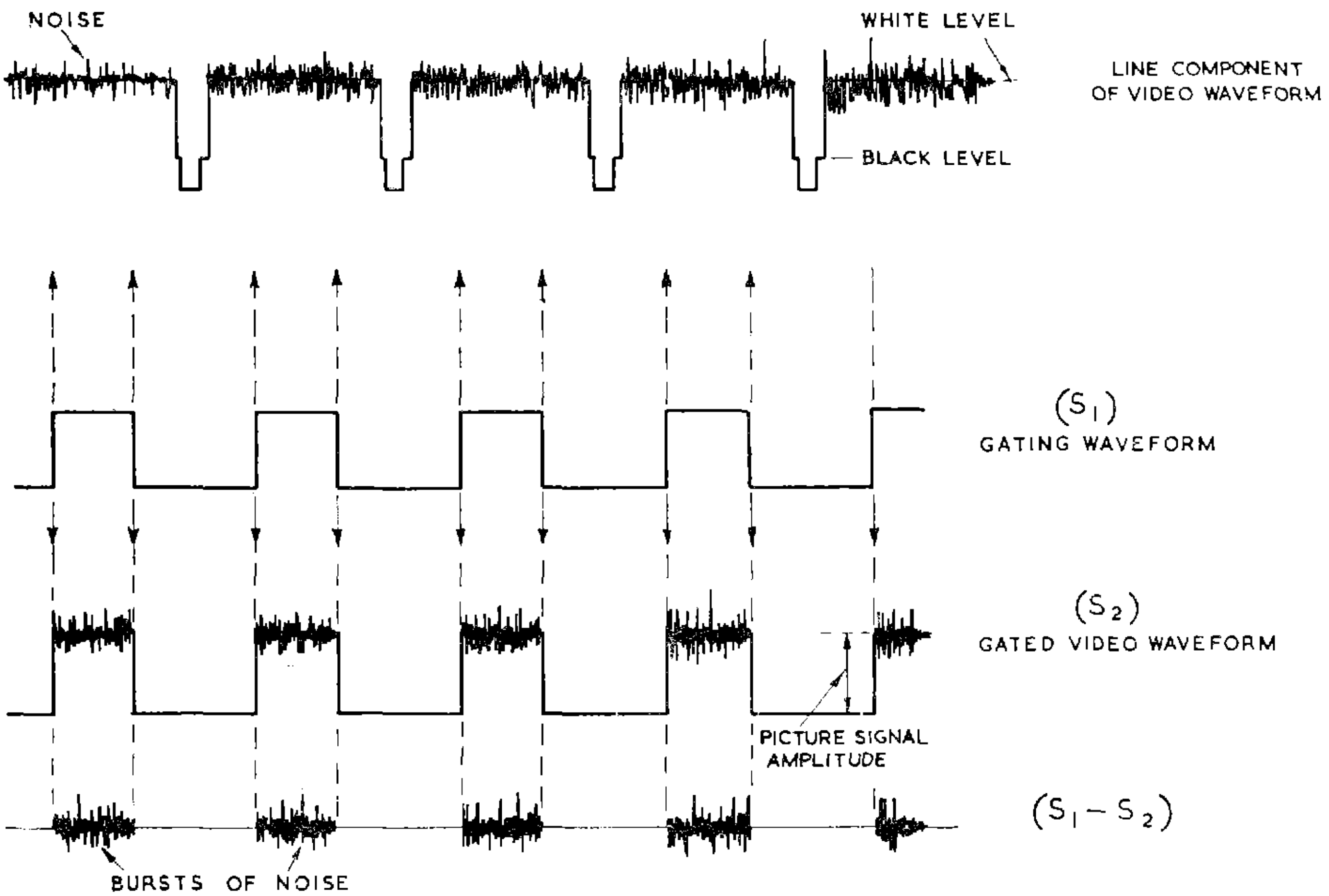

Fig. 1 — Basic subtraction process

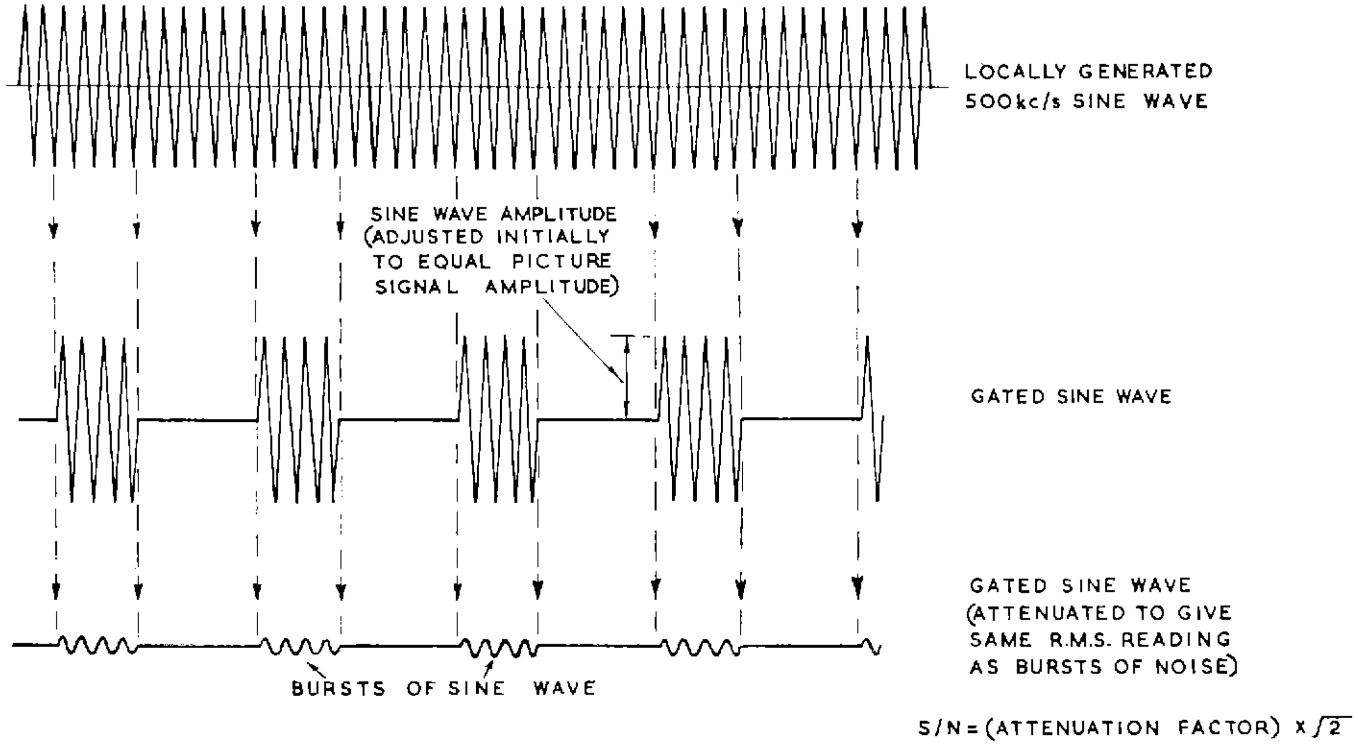

Fig. 2 — Evaluation of signal-to-noise ratio

3. Description of the Instrument

3.1 *General*

Fig. 13 is a photograph of the complete instrument. A range of signal-to-noise ratios from 9 dB to 51 dB can be measured at any video level between black and white, the result being expressed as the ratio of white picture signal to r.m.s. noise at the video level in question. The input can consist of a standard 1V d.a.p. video waveform, or separate picture and synchronizing signals, a selector switch and an internal sync-separation system being incorporated. A 3 Mc/s low-pass filter is included in the video picture signal input circuit. This is necessary because it is sometimes required to measure the signal-to-noise ratio at an early point in the television chain, where the signal has not yet been subjected to full bandwidth limitation.

The equipment is mains operated but otherwise self-contained and portable, with instructions attached to the case.

Both output meters (seen in Fig. 13) are contained in a separate box which is detached from the main body of the instrument and placed upon a level surface when in use. This is because it was found necessary to use a type of meter which must be operated in a horizontal position. In order to prevent use of the instrument with the meters still in the vertical (transport) position, the mains inlet socket is so situated as to be accessible only when the meters have been removed.

Throughout the whole of the design and development stages, reliability and ease of operation were held to be most important. Although this latter requirement caused considerable internal complications, the nature of the device is such that a fault-condition becomes immediately apparent to an intelligent user and is not likely to lead to erroneous measurements.

3.2 *Basic Operation*

Fig. 3 is a block schematic of the measuring instrument. The position of the switches depends upon which stage in the measurement process has been reached. Fig. 3 shows the basic arrangement used for the measurement of picture signal amplitude and the waveforms shown are all applicable to this condition, with the exception of the separated noise waveform (H). Details of the waveforms and their relative timings are shown in Fig. 10.

Before connecting the video signal it is necessary to adjust three 'set-zero' controls, not shown, to ensure that the balanced amplifier, phase-splitter and keyed-clamp circuits are accurately balanced in the absence of an input signal. When this operation has been correctly carried out, the outputs produced by the balanced amplifier are identical and zero deflexion is obtained on the *differentially* connected diode peak voltmeter. The output waveforms are as shown in Fig. 10 (D) and are produced by the action of the gate on the standing d.c. in the circuit. Because there is zero video input to the phase-splitter at this juncture, the negative or 'active' portions of each waveform correspond to black level. The video signal input derived from scanning a uniform scene is now amplified and subjected to keyed clamping at the input to the phase-splitter (Waveform A, Fig. 10) causing a symmetrical unbalance in the gated outputs. This can be seen at E and F in Fig. 10, the

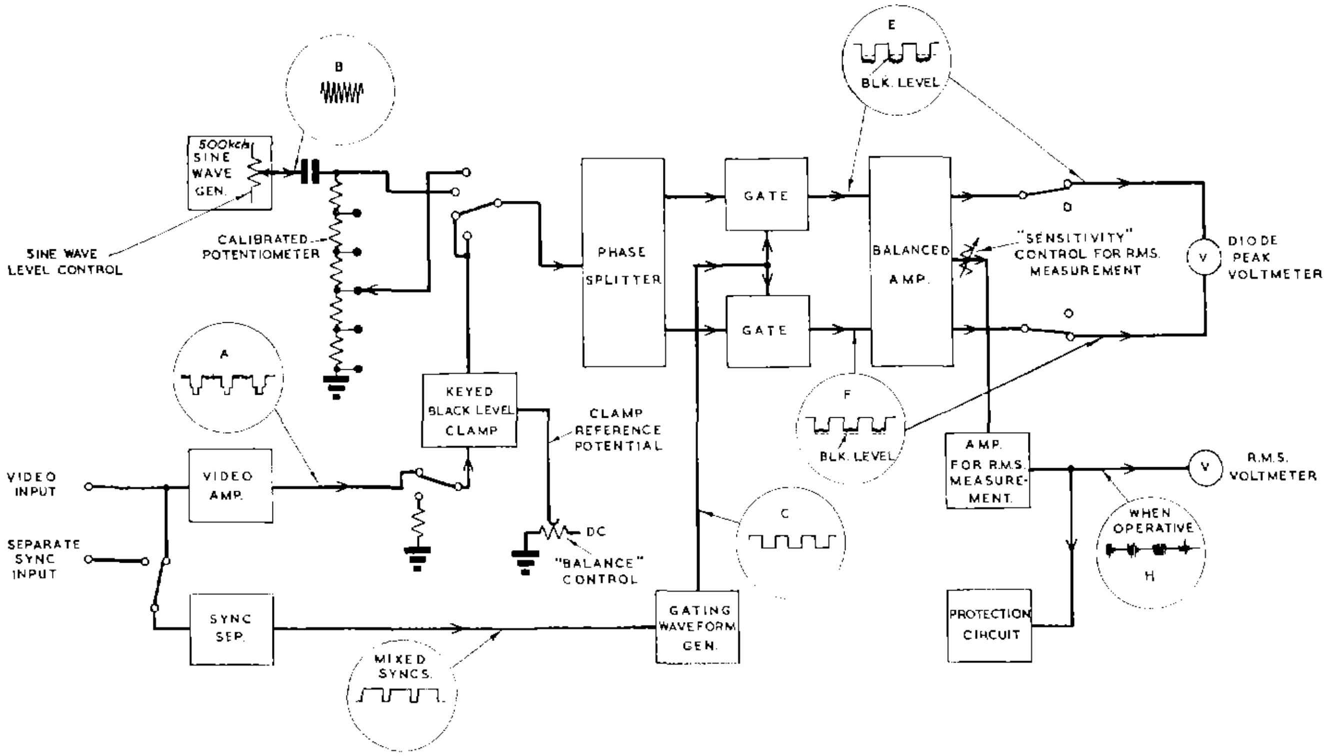

Fig. 3 — Block schematic

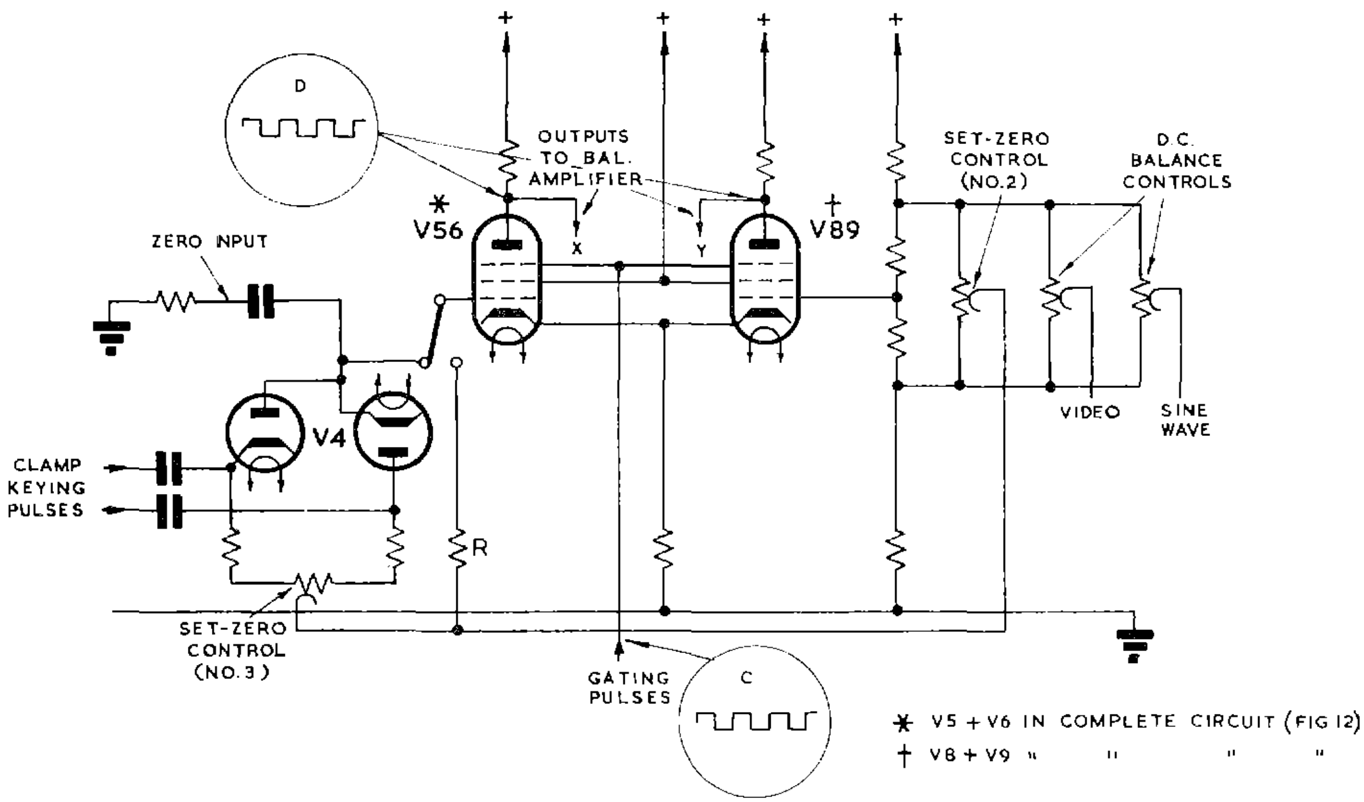

Fig. 4 — Basic phase-splitting and gating circuit. Set-zero condition

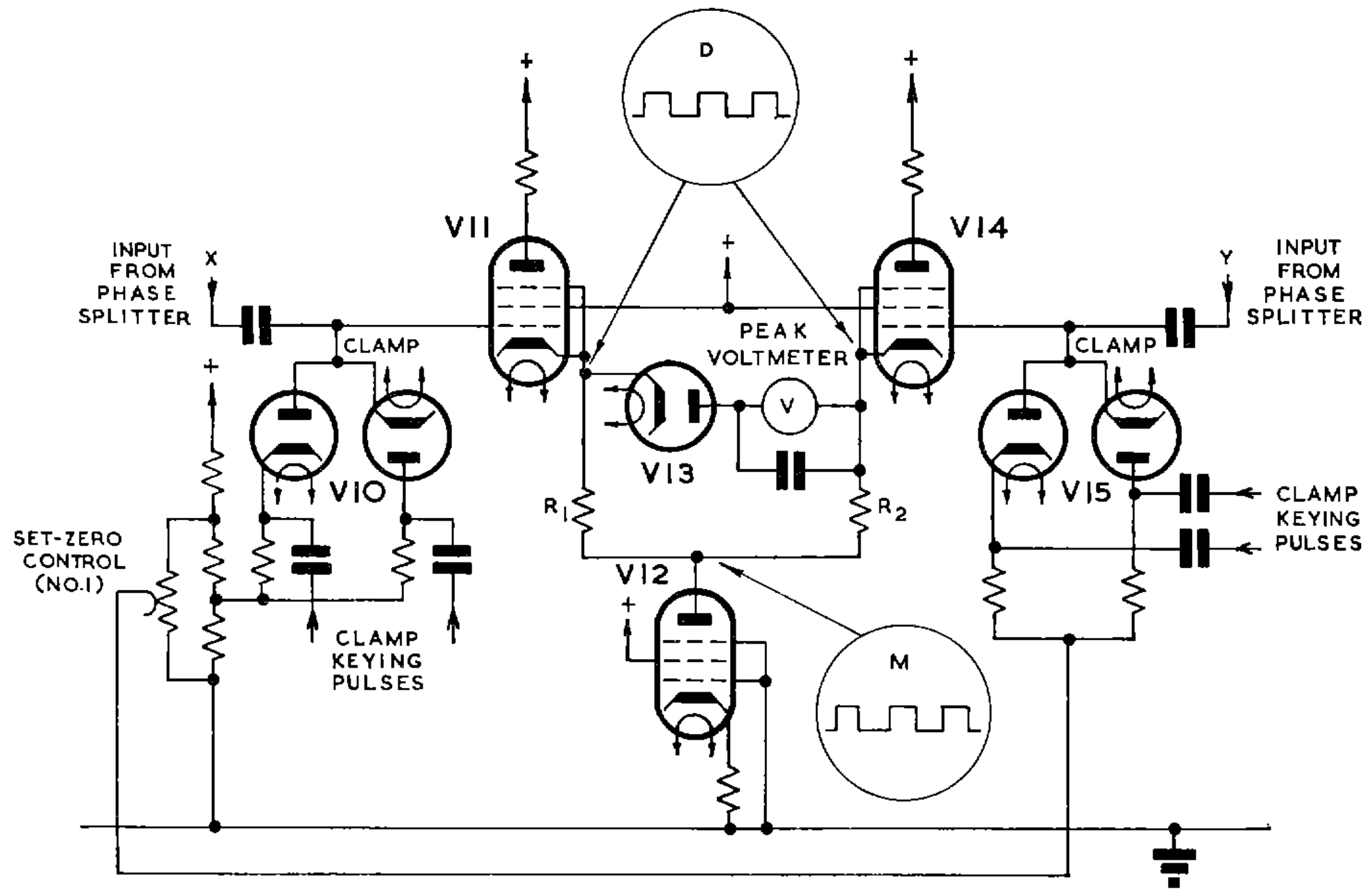

Fig. 5 — Basic balanced amplifier circuit. Set-zero condition

'active' portion of waveform E becoming more negative than black level, whilst the corresponding portion of waveform F becomes more positive. The diode peak voltmeter is connected so as to conduct on an unbalance in this direction and thus indicates the sum of the excursions of each gated video signal from black level. It is calibrated in terms of level at the input to the unit and hence measures the picture signal input amplitude, which should have the standard value of 0·7 V at white.

The next step involves the separation of the noise from the signal, using a subtraction process similar to that described earlier. The circuit connections of the balanced amplifier are now rearranged so that an amplified output is provided to drive the amplifier in the r.m.s. measurement chain, the diode peak voltmeter being disconnected. This new output is also proportional to the unbalance caused by the video signal input, but in this case the d.c. balance is restored by operation of the balance control associated with the keyed clamp at the phase-splitter input. This control merely shifts the clamp reference potential negatively so that the uniform input signal potential during the active line becomes the same as that previously held by black level. The fact that the whole input waveform is shifted by this operation is of no consequence, because only the central portions of the active line (and field) are selected by the gate. The d.c. content of the gated signal is thus accurately cancelled (by careful adjustment of balance for minimum deflexion on the r.m.s. voltmeter) and the noise waveform is isolated. A mid-scale reading of the gated r.m.s. noise voltage is obtained by adjustment of the sensitivity control at the output of the balanced amplifier. This adjustment must be left undisturbed during the remaining stages of the measurement.

We now have a measure of both the picture signal and r.m.s. noise voltages; the remaining operations are devoted to the evaluation of the ratio between them using the locally generated 500 kc/s sine wave.

The balanced amplifier is again connected to the diode peak voltmeter, but now an a.c. input to the phase splitter is obtained from the sine-wave generator, the output of which is adjusted to give the same reading on the diode peak voltmeter as was obtained from the picture signal. Waveforms J and K (Fig. 10) illustrate the form of the inputs to the voltmeter, which indicates the sum of the sine-wave amplitudes, i.e. the total 'push-pull' component.

Finally, the circuit is reverted to the r.m.s.-indicating condition and the input to the phase-splitter is obtained from the slider of the calibrated potentiometer which is adjusted until the r.m.s. voltmeter gives the same mid-scale reading as that which was produced by the noise.

The signal-to-noise ratio appears on the scale of the calibrated potentiometer.

3.3 *Gating Circuit and Balanced Amplifier*

The actions of phase-splitting and gating are performed by a single pair of amplifying stages, the outputs of which feed the balanced amplifier.

Figs. 4, 5, 6, 7, 8, and 9 show the basic circuits. V5 and V6, and again V8 and V9, are drawn as single valves, although their functions are each carried out by two valves in parallel in the equipment itself. The design called for 'suppressor-slope' pentodes in these positions, but the only available valves of this type did not have sufficient current output when used singly. For convenience, these pairs of valves are referred to in the text and the basic circuits as V56 and V89 respectively. Fig. 10 illustrates a selection of the waveforms applicable to the basic circuits. Five arrangements of this group are used to enable the necessary basic operations to be carried out, viz.:

(*a*) Peak voltmeter zero-setting.

(*b*) Indication of picture-signal amplitude (adjusted externally).

(*c*) Separation of noise and adjustment of r.m.s. noise output.

(*d*) Indication and adjustment of sine-wave amplitude.

(*e*) Adjustment of r.m.s. sine-wave output.

(*a*) *Peak voltmeter zero-setting* (Figs. 4 and 5)

In Fig. 5, the valves V11 and V14 drive the diode peak voltmeter which is connected between their respective cathodes. Pentode V12 forms a high-impedance common-cathode load, with R_1 and R_2 providing individual cathode loads for V11 and V14 respectively. This stage is thus able to handle high-amplitude 'push-push' inputs (the output appearing almost wholly at the anode of V12), whilst still accepting a 'push-pull' component to drive the differentially connected peak voltmeter.

Convenient new reference potentials are inserted in the drive waveforms to the two grids by means of the clamp circuits shown. Clamping is permissible here, even when handling a sine-wave signal, since the keying-pulses occur during the 'inactive' period of the waveform which is always in gated form at this point. The keying pulses are derived from a source which is common to the other clamping circuits in the apparatus and are timed to coincide with the video signal 'back-porch'. It is, however, erroneous to regard this particular pair of circuits as 'black-level' clamps, as will be seen. With no input signal the d.c. cathode potentials of V11 and V14 are balanced by adjusting the set-zero control (No. 1). Balance is indicated

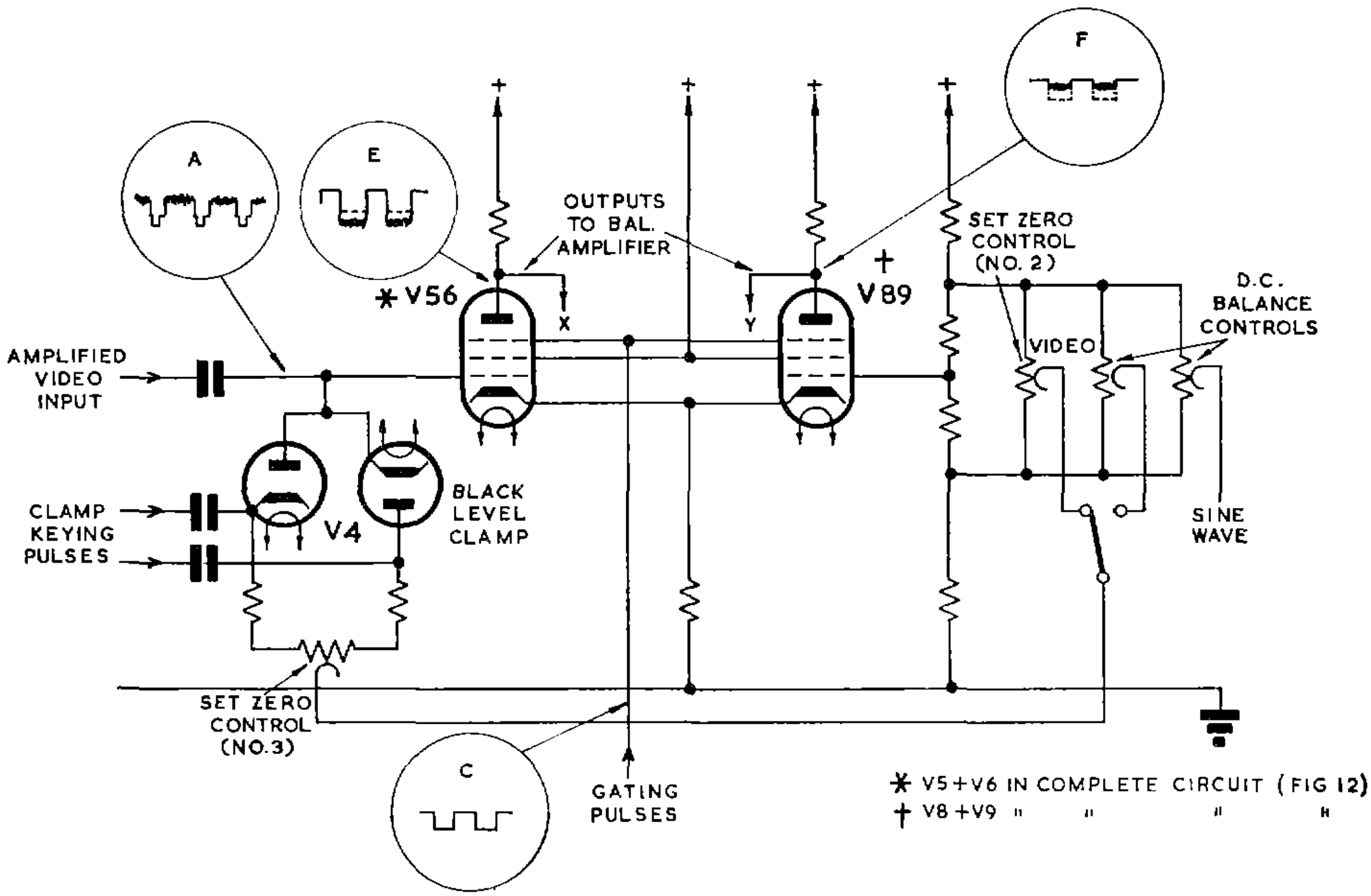

Fig. 6 — Basic phase-splitting and gating circuit. Video input condition

by the diode peak voltmeter, the electrical zero of which is slightly offset to permit a very small meter current to flow when the circuit is correctly set up.

Signals are now made available from the phase-splitting and gating stage V56 and V89 shown in Fig. 4. This circuit does not, however, receive a signal input itself at this juncture. The outputs, which are fed to V11 and V14, are shown by waveform (D) in Fig. 10 and arise from the fact that the anode currents of V56 and V89 are periodically interrupted by the gating pulses which are applied to the suppressor grids. Valves V56 and V89 form a 'long-tailed pair', V56 being the primary stage. If the anode waveforms of the two are identical, then the reading of the diode peak voltmeter will remain at zero, since V11 and V14 receive only a 'push-push' input. Any unbalance, however, will be indicated by the meter, which can be reset to zero by adjustment of the d.c. control-grid bias voltage of V56. Since, however, V56 will be required to accept both an a.c.-coupled sine wave and a clamped video signal input, each with the correct d.c. bias, it is necessary to carry out two operations. Firstly, the bias itself is adjusted by means of the set-zero control (No. 2) (a single resistance R temporarily replacing the clamp circuit at the grid of V56) and the correct insertion of this bias by the black-level clamp is next ensured by adjustment of the clamp balance (set-zero control No. 3), again for zero deflexion of the peak voltmeter.

(*b*) *Indication of Picture Signal Amplitude* (Figs. 6 and 7)

The application of the amplified and clamped video signal to the grid of V56 causes very nearly equal and opposite changes in potential to occur during the negative-going or active portions of the outputs from V56 and V89. This is shown by waveforms E and F in Fig. 10 and it will be apparent that the input to V11 and V14 now consists of combined 'push-push' and 'push-pull' signals. As described earlier the 'push-push' component appears almost wholly at the anode of V12 (waveform M), whilst the 'push-pull' signal voltage is developed between the cathodes of V11 and V14 and is indicated by the peak voltmeter. Equality of the cathode potentials of these valves during the 'inactive' period of the gated signals is maintained by the action of their respective clamping circuits, and the meter reading displayed is thus a measure of the picture-signal amplitude at the input to the unit.

(*c*) *Separation of noise and adjustment of r.m.s. noise output* (Figs. 6 and 9)

The peak voltmeter is now replaced by a variable-resistance sensitivity control (Fig. 9) and a new adjustable clamp reference potential for V56 is obtained from the video d.c. balance control (Fig. 6). An output is taken from the anode of V14, this output being produced in the following manner. As described in Section 3.2, the d.c. content of the gated video signal is cancelled by careful adjustment

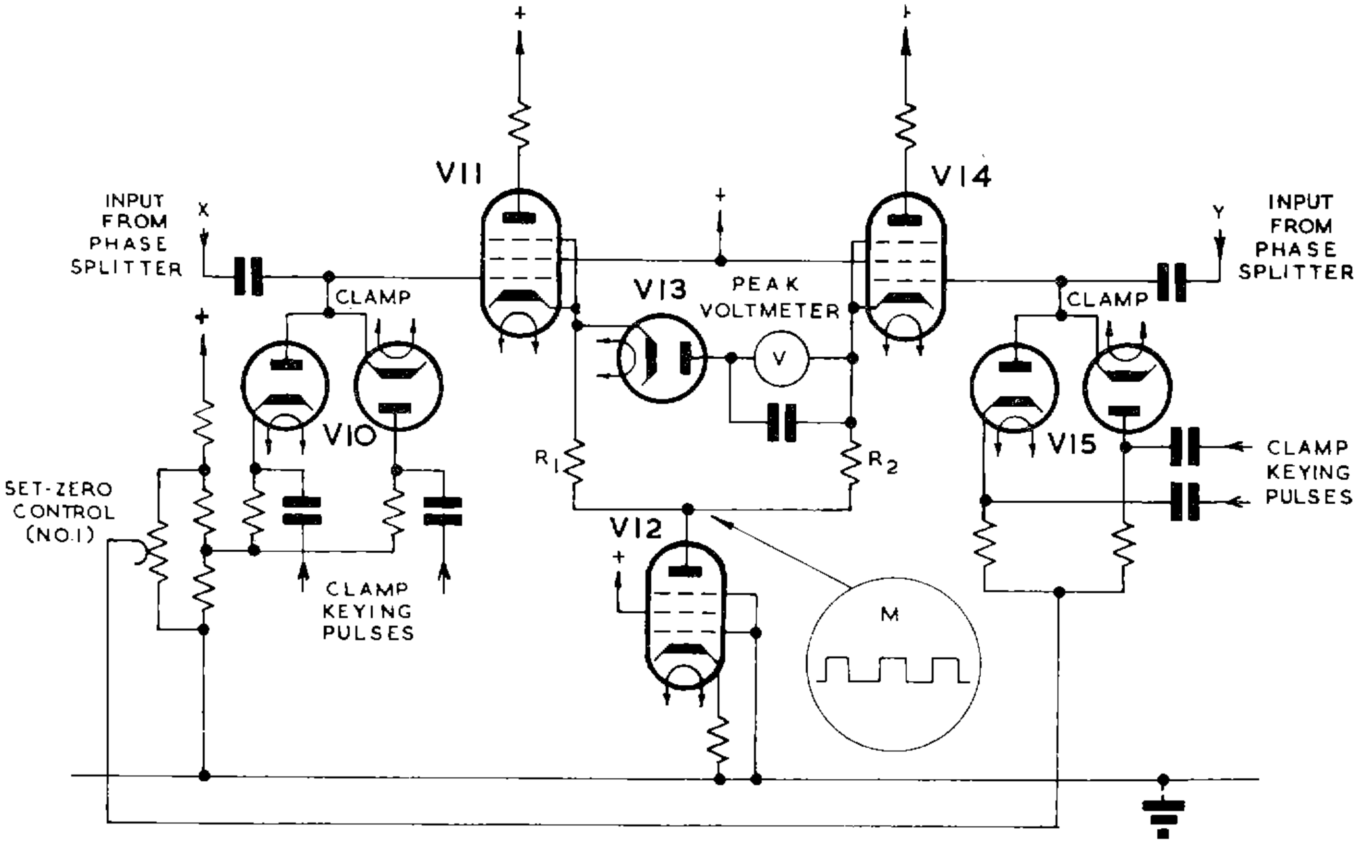

Fig. 7 — Basic balanced amplifier circuit. Peak voltage measurement condition

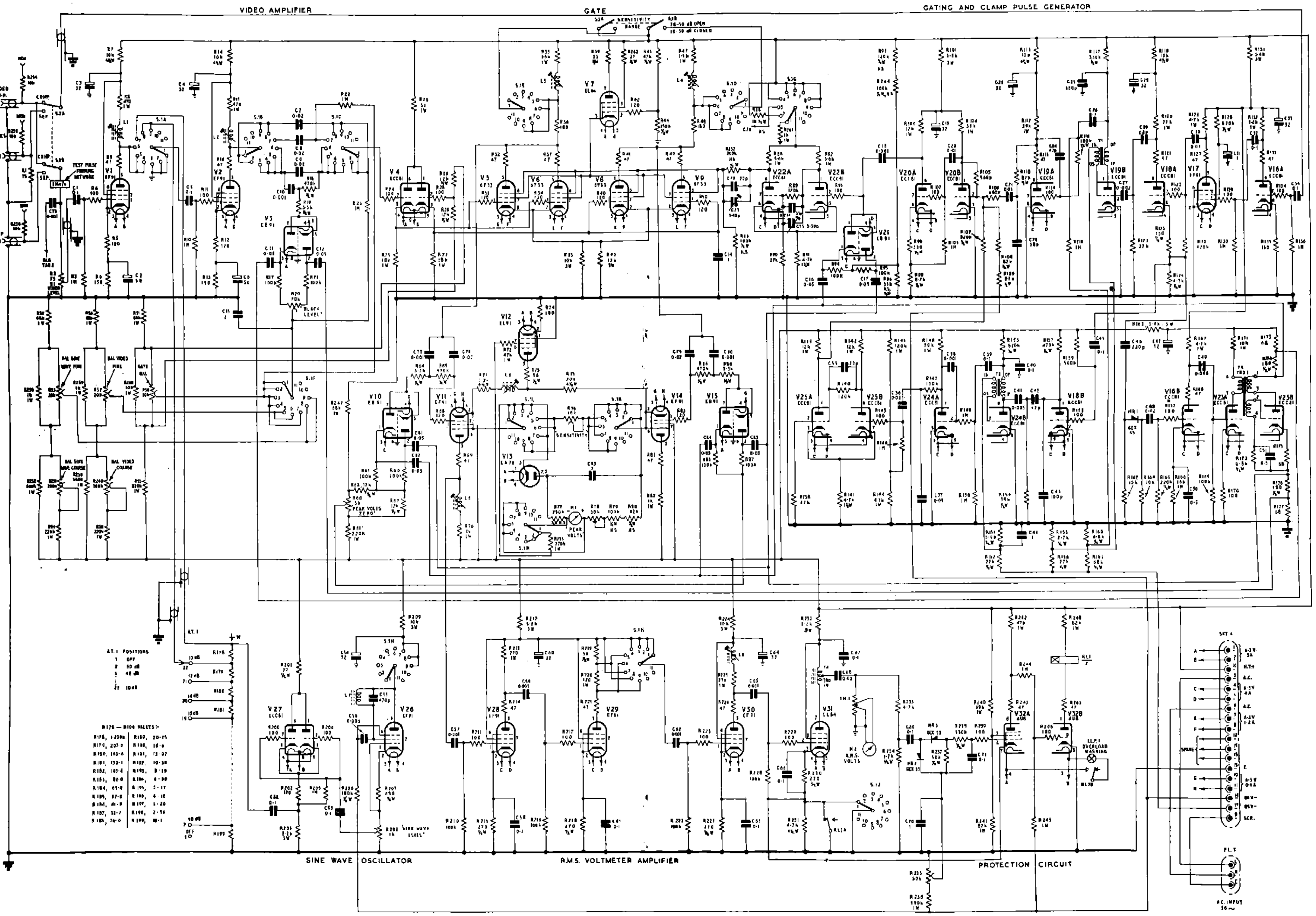

Fig. 12 — Circuit diagram

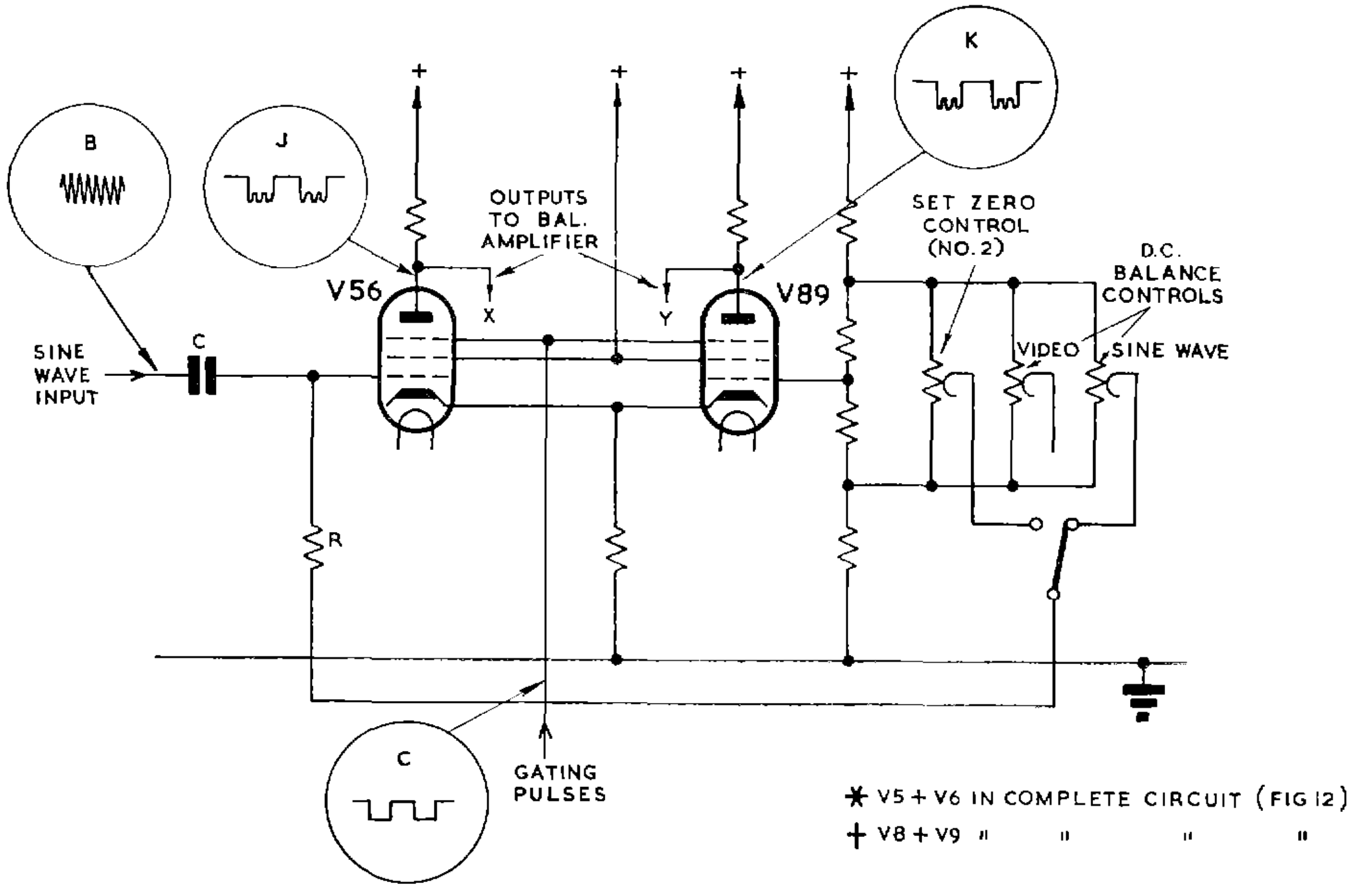

Fig. 8 — Basic phase-splitting and gating circuit. Sine-wave input condition

of the balance control, the uniform potential of the active signal being reduced to the value previously held by black level. V11 and V14 thus receive input signals from V56 and V89 similar to those shown by waveforms J and K in Fig. 10 (the noise waveform being shown as a sine wave). Because of the phase-splitting action of V56 and V89 the noise components are applied to V11 and V14 in 'push-pull' and appear in magnified form at V14 anode whilst the negative-going 'push-push' components are severely attenuated, appearing almost wholly at V12 anode as before.

The output from V14 is amplified and fed to the thermo-ammeter r.m.s. output indicator, the balance control being adjusted for minimum deflexion. A mid-scale reading of the r.m.s. noise output is obtained by means of the sensitivity control which adjusts the 'push-pull' gain of the V11 and V14 combination.

(*d*) *Indication and adjustment of sine-wave amplitude* (Figs. 7 and 8)

This condition is similar to that used for the indication of picture-signal amplitude, except that the sine-wave input is a.c.-coupled to V56 (Fig. 8) and no d.c. unbalance is caused. The peak voltmeter thus indicates the peak value of the sine wave, which is adjusted to give the same deflexion as the 'white' picture signal.

(*e*) *Indication and adjustment of sine-wave r.m.s. output* (Figs. 8 and 9)

The sine-wave output from the slider of the calibrated potentiometer is now connected to the control grid of V_1 and the potentiometer is adjusted to obtain the same mid-scale deflexion as was produced by the noise. It will be seen that, although the sine wave is a.c.-coupled, it has been necessary to provide a separate d.c. balance control for use in this condition. This is to ensure that a precise balance is maintained, having regard to the higher overall sensitivity of this circuit compared to the one used for peak voltage measurement. The adjustment of this balance control is again made for minimum deflexion on the r.m.s. output meter, followed by adjustment of the calibrated potentiometer to give the same mid-scale reading obtained when measuring the noise.

3.4 *Facility Switch*

As an aid to the simplification of operation, all the necessary circuit changes are made by a single rotary switch. The measuring process which has been described involves only five basic operations, but it was found necessary to increase this number when designing the practical instrument. Nine switch positions are provided therefore, as shown in the following table.

Basic Operation(s)	Facility Switch Position	Actual Operation(s)
Peak voltmeter zero-setting	1	Set balance of balanced amplifier
	2	Set balance of phase splitter
	3	Set clamp balance at phase splitter input
Indication of picture signal amplitude	4	Set picture signal amplitude (externally)
Separation of noise and adjustment of r.m.s. noise output	5	Balance-out video d.c. component (coarse)
	6	Balance-out video d.c. component (fine) and set gain of r.m.s. chain
Indication and adjustment of sine-wave amplitude	7	Set sine-wave amplitude (to 'white' picture value)
Adjustment of r.m.s. sine-wave output	8	Check balance (coarse)
	9	Check balance (fine) and adjust calibrated potentiometer for same r.m.s. reading as 6

Positions 5 and 8 each provide a condition of low sensitivity, intended to assist the operator in finding an approximate balance.

The signal-to-noise ratio is given by the potentiometer setting when all nine operations have been carried out.

Positions 2, 5, 6, 8, and 9 each involve an adjustment which, directly or otherwise, affects the d.c. potential at the input to the phase-splitter. Since for any particular switch position, the required potential varies very little from measurement to measurement, whereas it may vary appreciably during the process of a measurement, it has been found advantageous to provide a separate control for each switch position. In this way the operator is not compelled periodically to alter the setting of any one control and can pass rapidly through the various positions, little or no adjustment being found necessary in most cases. To repeat a measurement, for example, it is necessary only to return to Position 1 and hence to re-examine the meter readings for each position up to Position 9, checking any doubtful setting.

3.5 *Amplification of the Video and Noise Waveforms*

For two main reasons, low-gain stages of amplification are used throughout the apparatus. Firstly, low values of anode load resistance are used to reduce the necessity for frequency response correction. This is important because a portable instrument is likely to be subjected to mechani-

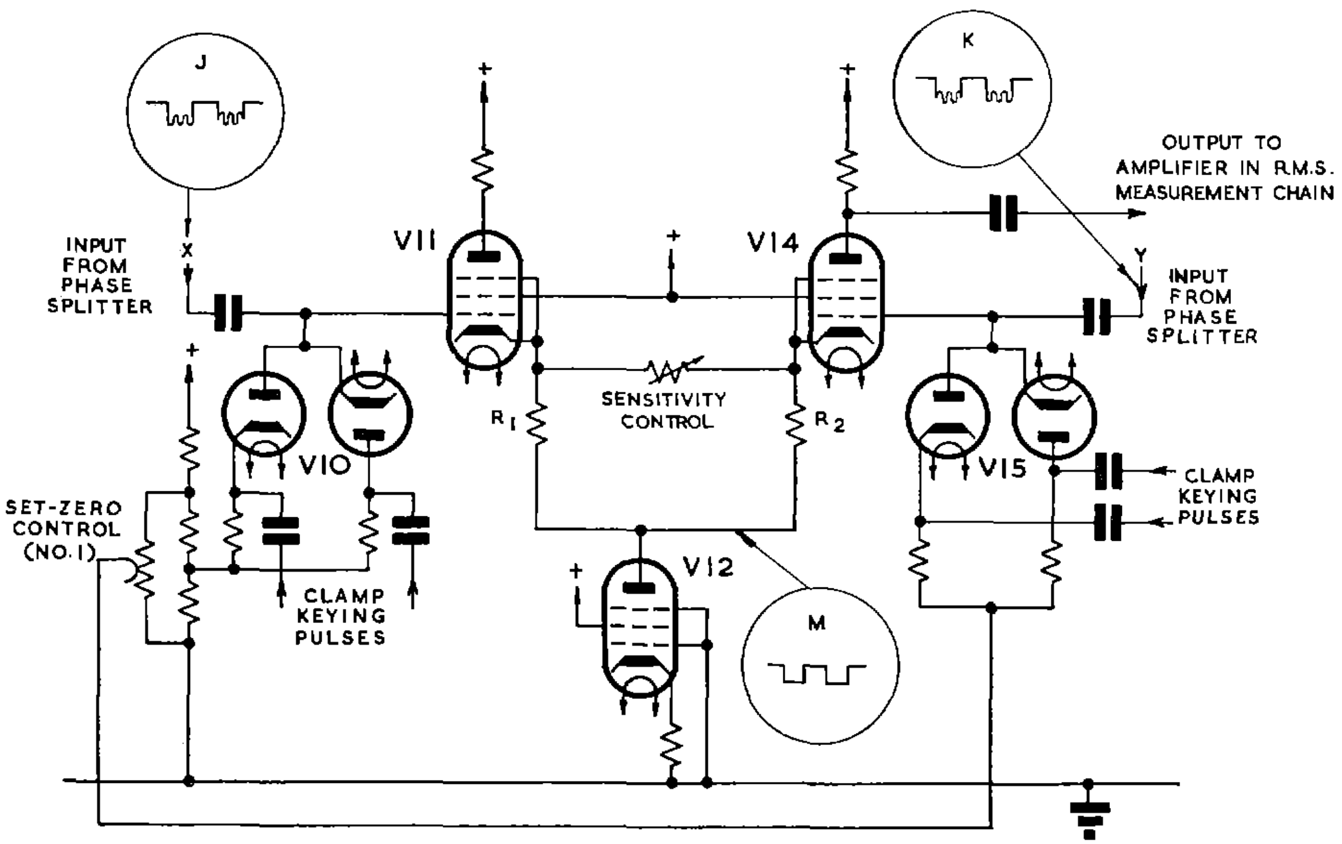

Fig. 9 — Basic balanced amplifier circuit. R.m.s. measurement condition

cal vibration which may otherwise cause variation in the frequency response of inductively compensated circuits. Secondly, cathode feedback is used in the amplifier circuits in order to reduce the dependence of gain upon valve parameters. This fact is of particular value in the case of those video amplifiers which are in use when measuring the picture-signal amplitude, since their gain defines the calibration of the peak voltmeter. Whilst such a calibration is not needed to carry out an accurate measurement of signal-to-noise ratio, it has been found to be a very useful practical facility when carrying out measurements at a number of grey levels.

The use of overall feedback was considered when designing the video and noise amplifiers, but the idea was discarded when it was found that such circuits would be insufficiently flexible to cater for all the configurations required.

The necessity to measure the video peak voltage, after gating, created a need for the transmission of the full d.c. component to the peak voltmeter. The use of a keyed-clamp circuit enabled this requirement to be fulfilled, but because of the nature of such circuits, special precautions were necessary to nullify the effect of noise voltages which may be present during the blanking interval. In the normal course of events when measurements are made at the output of a television camera or telecine channel, such noise will not appear. If, however, the measurement is being made at a point earlier in the chain or at the output of a receiver, then noise will exist at all points in the waveform. In these circumstances, ordinary clamping circuits will cause random variation of the d.c. component, depending upon the instantaneous blanking level near the time that the clamping operation ceases. To reduce this effect an integrating or flywheel arrangement[6] was used in the input circuit of the phase-splitter, providing for the clamp to operate on the mean blanking level, i.e. that which would obtain in the absence of noise. Since this is the only part of the unit where such trouble can arise, conventionally designed clamping circuits are used elsewhere.

The linear amplification of the noise waveform presented special problems, since a well-defined peak value could not be assigned to it when calculating the signal-handling capacity of the amplifiers. In theory any assigned level, however high, will be attained sooner or later by a noise peak. Nevertheless the probability of this level being reached in a finite time diminishes very rapidly indeed as the level is raised. For example, in order to have an even chance of observing a noise peak equal to five times the r.m.s. value, it would be necessary to observe noise in a 3 Mc/s band for a fraction of a second. To observe a peak ten times the r.m.s. value, millions of years would be required. It is shown in the Appendix that if noise is amplified linearly for amplitudes of either polarity up to three times the r.m.s. value, thereafter limiting the amplitudes to this value, this limitation reduces the r.m.s. value of the output by much less than 1 per cent. To achieve this accuracy, therefore, an amplifier handling a continuous noise voltage must be capable of accepting a range of input levels which is six times the input r.m.s. value. In the instrument being described, a further allowance has had to be made for the interrupted nature of the noise waveform. Because only approximately one-quarter of the picture area is selected by the gating circuit, then the r.m.s. value of the noise waveform is reduced to about one-half of its equivalent continuous value. The amplifiers are therefore designed to accept a range of input signal levels which is at least twelve times the r.m.s. value.

3.6 *Generation of the Gating Waveforms*

The gating circuit selects the central portion of each television line and field, approximately one quarter of the active picture area being selected in this way. Suitable gating waveforms are generated within the unit. Fig. 11 illustrates a number of the waveforms associated with the generation of the gating signal and Fig. 12 is a circuit diagram of the complete instrument. With reference to Fig. 12, Pentode V17 is a conventional sync-separating stage and its control grid is fed with an amplified negative-going video signal from the anode of V16A. V18A is used to amplify and distribute the separated pulses, the output from the anode being used to trigger the blocking oscillator line sawtooth generator V19B, whilst two outputs are taken from the cathode. One output is used to trigger the clamp keying-pulse generator V16B and the other output feeds the circuit, controlled by V18B, which separates field from line sync. pulses. Considering the action of V18B, high-amplitude, negative-going sync. pulses are applied to the grid via C45, the bias resistance R159 being returned to the positive h.t. line to ensure that, by virtue of the high aiming potential, grid current flows during the interval between sync. pulses. Thus the valve normally is fully conducting, but becomes cut off for the duration of each sync. pulse. Resistor R157 and capacitor C43 form a charging circuit, with the capacitance heavily shunted by V18B when conducting. During the non-conducting period, however, C43 is allowed to accumulate a charge and it will be seen that, during the field synchronizing period, the maximum value of this charge will be nearly four times that which is permitted during the line pulse. At the end of each pulse C43 is discharged by V18B, and the resulting waveform is differentiated by C42 and R154 and used to trigger the field-frequency blocking oscillator V24B, which generates a sawtooth output across C40 (Fig. 11(*a*)). Discrimination against the lower amplitude line-pulse component is ensured by the provision of adequate external negative grid bias (via R152 and R151). V24A is a Miller integrating stage which receives a nearly linear sawtooth input from C40 and generates a parabolic waveform to feed to the cathode-coupled multi-vibrator V25 (Fig. 11(*b*)). The grid-bias for V25B and the amplitude of the input parabolic waveform are chosen so as to permit conduction only during the central portion of the active field period, R147 and C37 being included to provide the slight amount of phase shift required to centralize the waveform relative to the active field period. The output from V25 consists of a rectangular waveform at field frequency and is fed to the feedback-type adding stage V20B. To this stage is also connected the output from the Miller

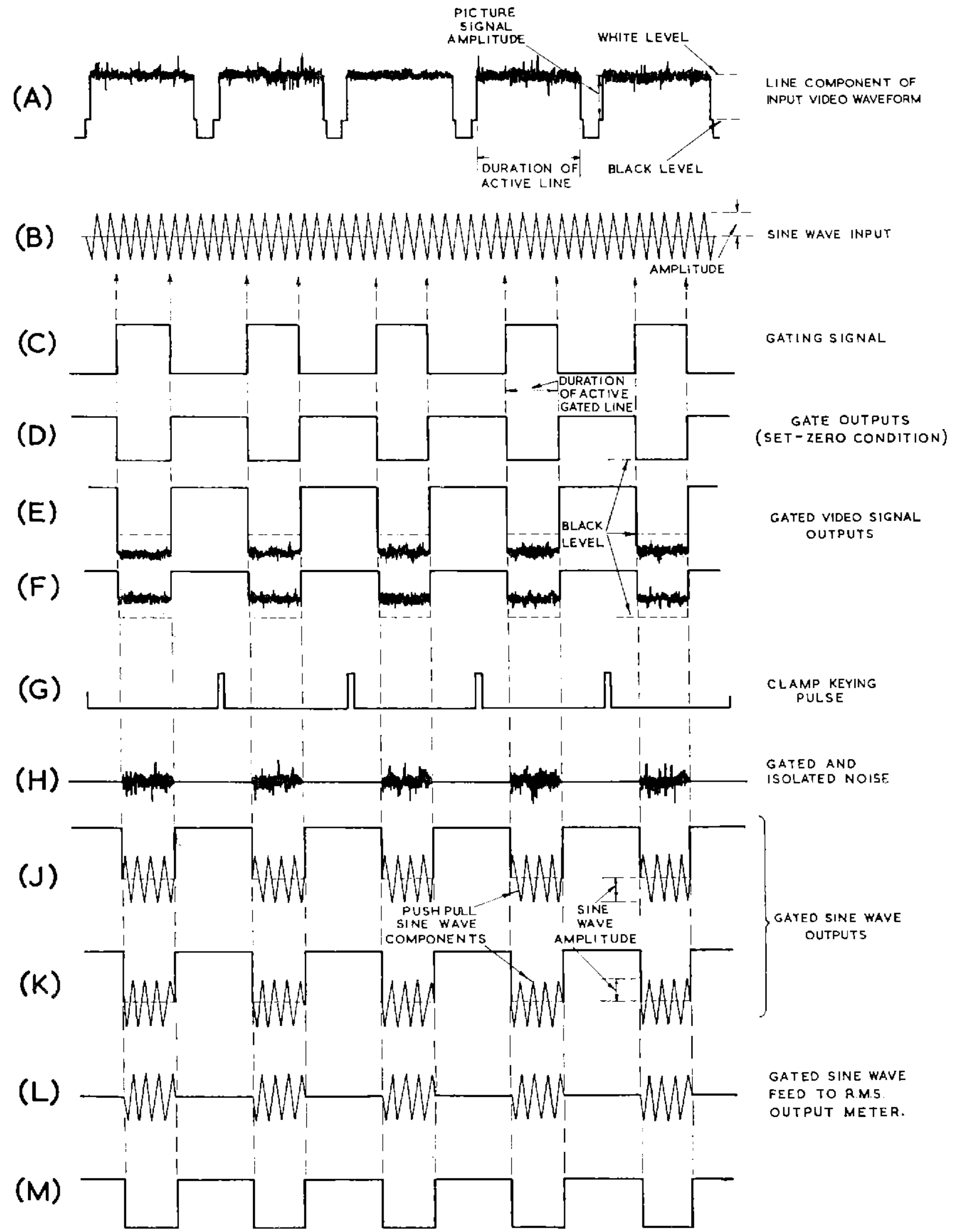

Fig. 10 — Typical waveforms

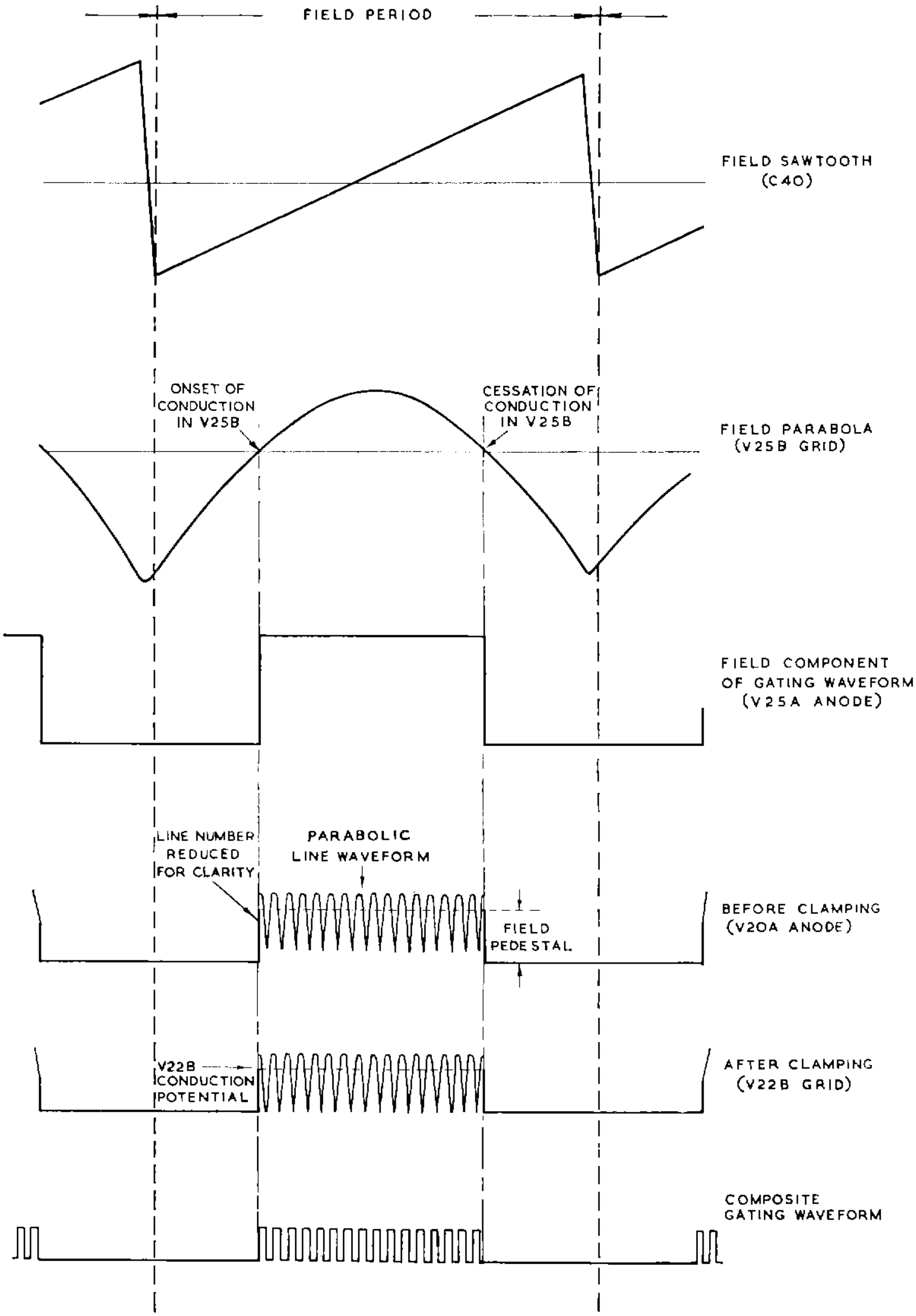

Fig. 11 — Generation of the gating waveform

integrator V19A, which generates a line-frequency parabolic waveform in conjunction with V19B, the circuit operation being similar to the field-frequency case just described. The adding stage V20B is designed to be cut off by the negative-going or 'inactive' portion of the field waveform resulting in an output waveform from V20A as shown in Fig. 11(*d*). This is applied to the cathode-coupled multivibrator V22 and causes conduction of V22B during the central portions of the selected lines. The output from V22A (Fig. 11(*f*)) constitutes the composite gating waveform which is connected to the suppressor grids of the gating valves. In order to avoid errors in the width of the line gating pulses which may be caused by variations in the height of the field 'pedestal', the grid input waveform of V22B is subjected to line-by-line clamping as shown in Fig. 11(*d*) and (*e*).

3.7 *R.M.S. Output Indication*

An industrial-type 4 mA vacuum thermocouple is used to drive the horizontally mounted output microammeter. This meter is calibrated with a decibel scale in order to allow interpolation between the discrete 2 dB steps of the calibrated potentiometer.

The thermocouple and microammeter combination were chosen in consultation with Ernest Turner, Electrical Instruments Ltd, in order to obtain optimum stability of meter-scale law against both a change in thermocouple and changes in ambient temperature. Since it is necessary to drive the output indicator from a valve amplifier and bearing in mind the frequency response and the peak signal-handling capacity required when amplifying a noise voltage, it will be appreciated that the thermocouple heater current and impedance are defined within fairly close

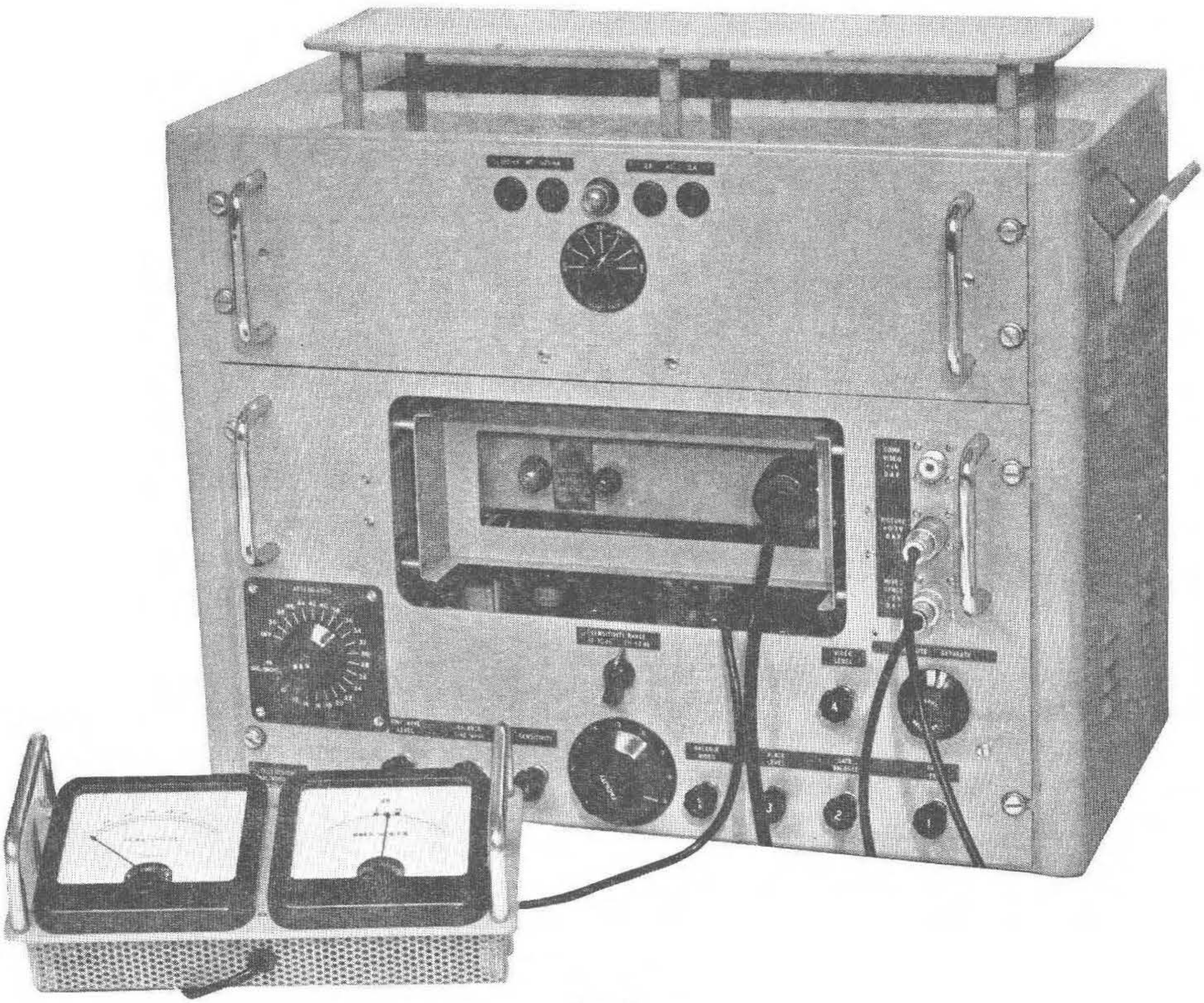

Fig. 13

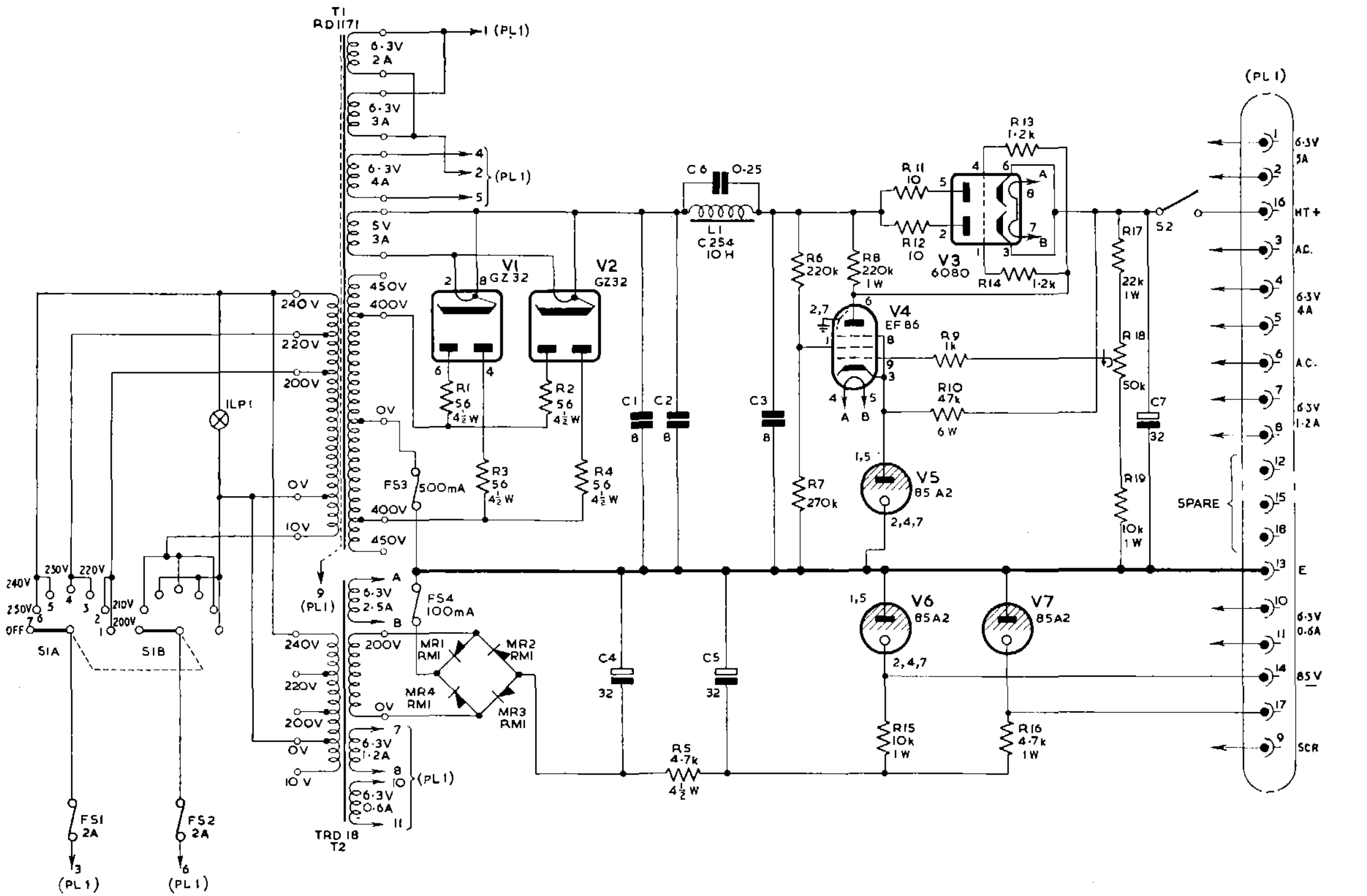

Fig. 14 — Power supply unit

limits. To ease this problem an output transformer was used which provided a current gain of 6 dB with an overall frequency response which was uniform from 60 kc/s to 3·5 Mc/s. The advantage of rejecting frequencies below 60 kc/s is that the effects of any unwanted components (such as 'tilts' and 'bends') which may occur within the gating interval are greatly reduced.

The error in the measured signal-to-noise ratio due to the loss of information below 60 kc/s is treated as being negligibly small.

To prevent damage to the meter or thermocouple if accidentally overloaded, a protection circuit has been incorporated. A valve-controlled relay is arranged to remove the driving voltage when it exceeds the full-scale value, the detecting circuit being designed to be approximately equisensitive to both noise and sine-wave signals. Because the thermocouple and meter are normally operated well within their maximum ratings, the time delay in relay operation is not serious, particularly as the manufacturer's specification permits a 50 per cent overload for 'short' periods. Operation of the protection circuit is indicated by the lighting of a lamp, and the circuit is so interlocked as to require the operator to return to Position No. 1 on the Facility Switch and recommence the measurement.

3.8 *Accuracy*

It will be noted that the peak voltmeter is of the simple diode type and will give accurate readings only when provided with the correct type of input; it cannot, for example, accurately measure the peak value of a signal with a very low duty factor, such as a narrow pulse. This characteristic of the circuit helps to improve accuracy when measuring the value of picture signal plus noise, in that the noise peaks are approximately averaged, giving a reading of peak voltage corresponding to a noiseless video signal. The effect is difficult to analyse quantitatively and therefore a number of oscilloscope comparisons have been made at different added noise levels, to ascertain the maximum error likely to be incurred. The maximum discrepancy between peak measurements of television signal and sine wave is of the order of 3 per cent. This is by far the greatest error and the tendency is for the measured signal-to-noise ratios to be too high by an amount varying from 0 to $+0{\cdot}25$ dB. It is not considered worth while to introduce a correction for this relatively small amount, particularly as the next largest source of error is always biased in the opposite direction. The latter effect is due to the inability of the instrument and its user to balance out accurately the video component, leaving noise as the only contributor to the r.m.s. output. Again, it is difficult to be precise about the magnitudes involved, since they are very small and variable, but it is felt that an overall accuracy figure of $\pm 0{\cdot}5$ dB is a safe limit to claim for the device as a whole.

3.9 *Power Supplies*

Fig. 14 shows the circuit of the power supply unit which is located on a separate chassis and panel, housed within the same cabinet as the measuring instrument.

A conventional series regulator is included in the 330 V h.t. supply and neon stabilizers are used to stabilize the two -85 V low-current lines used for biasing purposes.

4. Conclusions

The instrument described in this monograph has fully justified the effort made in its design and construction: it has, in particular, demonstrated its usefulness in routine checking of the performance of camera tubes and telecine machines. A number of lessons have been learned, however, and should the opportunity arise a second model would incorporate a number of practical changes. In particular, suitably disposed electro-mechanical switches would replace the existing Facility Switch unit, the inclusion of which imposed a severe limitation to freedom of layout.

The author wishes to acknowledge the important contribution made by Mr I. G. Gibbs in the design and development of this instrument.

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FUTURE DEVELOPMENTS

Another method of measuring signal to noise has been investigated. Broad details were published in the E.B.U. Review.* The new method does not involve balancing and promises to be a robust technique for field measurements.

* Weaver, L. E., **Some New Video Measurements Techniques and Apparatus,** E.B.U. Review, Part A, Technical, No. 64, December 1960.

APPENDIX

THE AMPLIFICATION OF A RANDOM NOISE VOLTAGE

General

Any device which is used to indicate the r.m.s. value of a random noise voltage must be able to handle peak voltage excursions of either polarity in excess of this voltage if serious error is to be avoided.

The magnitude of the error in an r.m.s. measurement caused by peak voltage limitation has been determined. Close agreement has been found between theory and practice.

*Theoretical Procedure**

Assuming that the noise amplitude has a normal probability density distribution we can calculate its r.m.s. value and the change in r.m.s. value caused by a known amount of peak-amplitude limitation.

If u=noise voltage

$$\phi(u)=\frac{\lambda}{\sqrt{\pi}}e^{-\lambda^2u^2}=\text{probability density}$$

The standard deviation σ of the unlimited distribution is given by

$$\sigma^2=2\int_0^\infty u^2\phi(u)\,du$$

This is a well-known expression giving

$$\sigma^2=\frac{1}{2\lambda^2}$$

In the case of a distribution limited to a maximum value $\pm U$, all values of $|u|$ up to $|U|$ will be included by

$$2\int_0^U u^2\phi(u)\,du$$

whereas excursions beyond $\pm U$ will have the value $\pm U$, but with a probability of occurrence still denoted by $\phi(u)$.

Thus the standard deviation of the limited distribution:

* Mathematical Analysis by Dr R. D. A. Maurice.

$$S^2=2\int_0^U u^2\phi(u)\,du+2\int_U^\infty U^2\phi(u)\,du$$

The ratio $\frac{S}{\sigma}$ gives the ratio of the r.m.s. value of the limited distribution to that of the unlimited distribution for various values of U (U being expressed in terms of σ).

Fig. 15 shows $\frac{S}{\sigma}$ plotted against various values of $\frac{U}{\sigma}$ together with a similar curve for a sine wave.

Experimental Procedure

A multiplier type photo-electric cell was used as a noise source, adjustment of output being obtained by controlling the brightness of a lamp adjacent to the photo cathode. A two-stage valve amplifier provided the drive to a 1 600 ohm, 1·25 mA vacuum thermocouple, the amplifier being designed to handle peak amplitudes having some thirty to forty times the r.m.s. value. A diode circuit connected across the heater of the thermocouple provided a means for adjustable symmetrical peak limitation. Care was taken to ensure that the clipping action was very close to the ideal. To this end, the change in r.m.s. value of a sine wave was measured when subjected to peak clipping, close agreement with the calculated figures being recorded. A further precaution was taken by using a noise spectrum limited to frequencies below about 100 kc/s.

A number of curves were plotted to investigate the effect of bandwidth changes on the amplitude distribution, but, as can be seen, no marked differences are to be observed and the overall agreement with the calculated curve appears to be very good.

Conclusion

A waveform consisting of continuous random noise can be amplitude limited to a value which is plus and minus three times the r.m.s. value of unlimited random noise with a change in r.m.s. value of much less than 1 per cent.

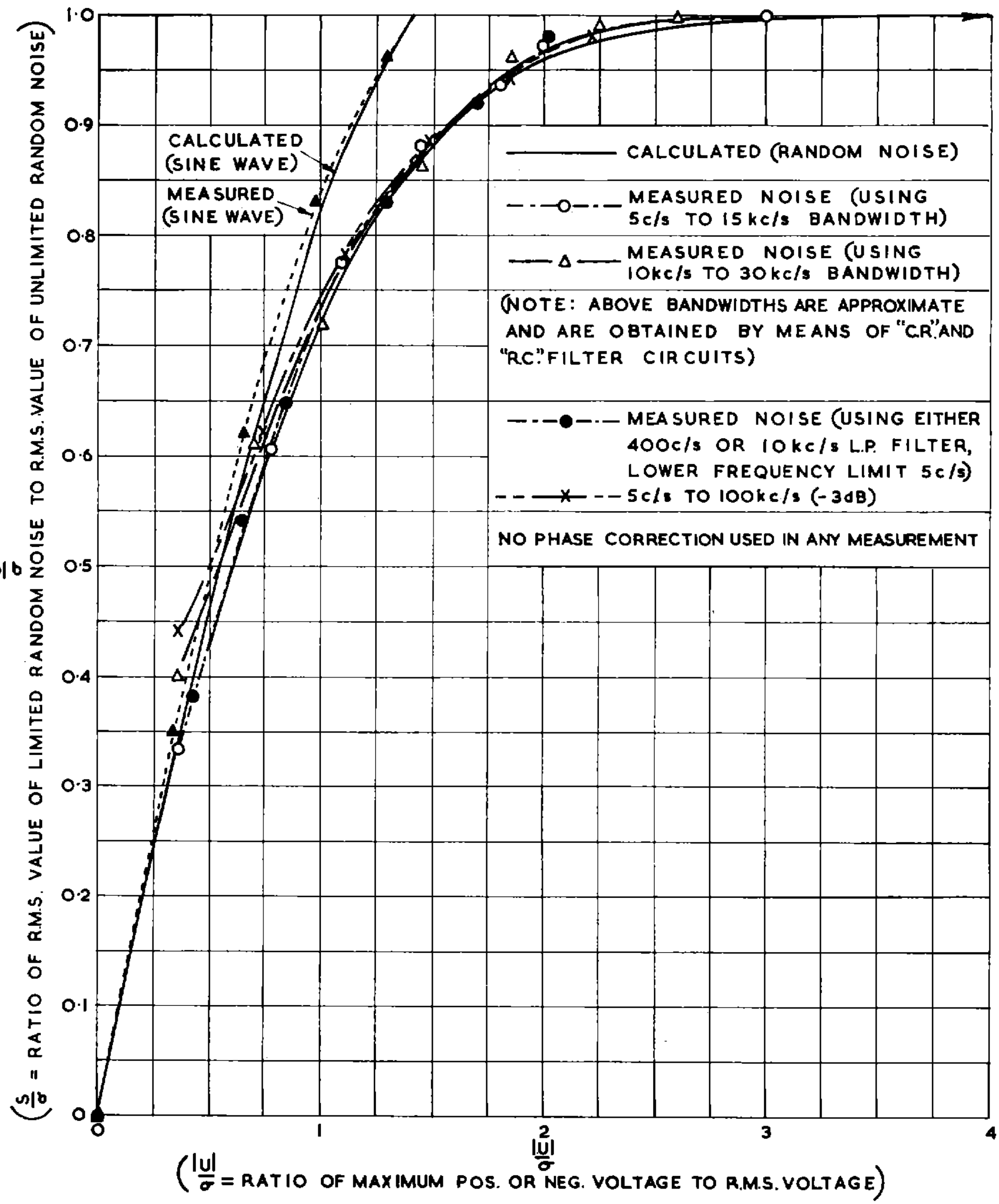
CALCULATED (SINE WAVE)
MEASURED (SINE WAVE)
CALCULATED (RANDOM NOISE)
MEASURED NOISE (USING 5c/s TO 15kc/s BANDWIDTH)
MEASURED NOISE (USING 10kc/s TO 30kc/s BANDWIDTH)
(NOTE: ABOVE BANDWIDTHS ARE APPROXIMATE AND ARE OBTAINED BY MEANS OF "C.R." AND "R.C." FILTER CIRCUITS)
MEASURED NOISE (USING EITHER 400c/s OR 10kc/s L.P. FILTER, LOWER FREQUENCY LIMIT 5c/s)
5c/s TO 100kc/s (-3dB)
NO PHASE CORRECTION USED IN ANY MEASUREMENT
$\left(\frac{S}{\sigma}\right.$ = RATIO OF R.M.S. VALUE OF LIMITED RANDOM NOISE TO R.M.S. VALUE OF UNLIMITED RANDOM NOISE$\left.\right)$
$\frac{S}{\sigma}$
$\left(\frac{|U|}{\sigma}\right.$ = RATIO OF MAXIMUM POS. OR NEG. VOLTAGE TO R.M.S. VOLTAGE$\left.\right)$
$\frac{|U|}{\sigma}$
0
0·1
0·2
0·3
0·4
0·5
0·6
0·7
0·8
0·9
1·0
1
2
3
4

Fig. 15

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