Mr. Somerville.

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The Measurement of Random Noise in the presence of a Television Signal

by

L. E. WEAVER, B.Sc., A.M.I.E.E.
(Designs Department, BBC Engineering Division)

BRITISH BROADCASTING CORPORATION

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FOREWORD

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

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

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THE MEASUREMENT OF RANDOM NOISE IN THE PRESENCE OF A TELEVISION SIGNAL

SUMMARY

It is becoming increasingly important for authorities concerned with the generation, distribution, and radiation of television signals to have available an accurate method of measuring random noise in the presence of a signal. This monograph describes two realizations of a simple method which is based upon sampling the random noise in the known minimum-energy regions of the video spectrum. The first has been used quite extensively and has proved to give very consistent and reliable results, while the second and more sophisticated version is at the moment still in the course of development.

The advantages of the method are:

(*a*) The apparatus is simple and stable and the measurements can be repeated to within 1 dB. Also, no undue degree of skill is required on the part of the operator.

(*b*) Both the noise spectrum and the total noise power in the band of frequencies are measured at the same time.

(*c*) Measurements can be made accurately in the presence of pilot control frequencies and interference of a relatively continuous nature such as pick-up from broadcast stations and mains hum. Furthermore, the accuracy is not impaired by relatively large amplitude non-linearity such as is encountered, for example, in gamma correction amplifiers.

(*d*) One is not restricted to a particular test signal of an artificial type, but, within limits, measurements can be made during the transmission of a wide range of picture signals. Also, it is simple to measure the noise level at various amplitudes of the picture signal.

The disadvantage is that there is a minimum noise level, depending upon the type of signal present and the frequency of measurement, which can be measured to a given degree of accuracy. This limitation, however, has not been found at all restrictive in practice and, in general, an overall accuracy of approximately ± 1 dB can be expected.

1. Introduction

The measurement of the signal/random fluctuation noise ratio in a television transmission link or in the signal from a television camera has for a long time been one of the most difficult to make in a truly satisfactory manner.

The problem arises as a result of the nature of random fluctuation noise, which may be considered as a perfectly random ensemble of impulses which have amplitudes lying between zero and, in theory, infinite amplitude.

Neither the voltage nor the current at any instant has any significance by itself, since it is implicit in a truly random process that any single event can give no information about any other. On the other hand, the distribution is known statistically so that the probabilities of occurrence of impulses with given amplitudes are known. As would be expected, the impulses having infinite amplitudes also have vanishingly small probabilities of occurrence.

An r.m.s. random noise voltage or current can be defined as the mean of the sums of the squares of the instantaneous voltages or currents taken over a sufficiently long period. Ideally, the r.m.s. voltage can be measured by a suitably designed valve voltmeter, but in practice, at least over the video frequency range, the only characteristic of random fluctuation noise which can readily be measured absolutely is the long-term mean power.

This mean power is capable of being measured very accurately as a function of the heat dissipated in a resistive element as in a thermocouple or thermistor bridge, and in any instance where the random noise level is not affected by the absence of signal the r.m.s. noise voltage may be derived very simply in this manner from the mean power and the value of the resistor in which it is dissipated. Then the ratio of the peak-to-peak picture signal voltage at that point in the circuit to the r.m.s. noise voltage, expressed in decibels, gives the signal-to-noise ratio in the standard form. This assumes, of course, that no hum voltages or other spurious signals are present on the circuit during the noise voltage measurement, since these should not properly be included as part of the random fluctuation noise level.

For this reason it is common to measure the random noise voltage with an inserted high-pass filter which removes the energy below about 5 or 10 kc/s. The effect of hum voltages, which may contain relatively high harmonics of the mains frequency, is thereby eliminated with a quite negligible error. With a 3-Mc/s band of noise which has a flat spectrum only about one three-hundredth of the noise power is lost, and if the noise has a triangular distribution with frequency the error is even less.

However, it is becoming increasingly common for transmission systems to need at least the transmission of line synchronizing pulses to keep them operating normally, and in such instances where the signal-to-noise ratio has to be measured with a signal present, less straightforward methods must be used. This situation, of course, also occurs with the output signal from a television camera or other picture originating device, where at least the blanking waveforms must be present in every case.

The simplest method makes use of the fact that the decreasing probability of occurrence of the higher noise peaks, in conjunction with the non-linear characteristic of the eye, results in a given random noise level appearing to have a fairly well defined amplitude as seen on a waveform monitor. This quasi peak-to-peak amplitude can be converted to the equivalent r.m.s. value and the ratio formed with the peak-to-peak amplitude of the signal, also measured on the waveform monitor. Provided the signal has an easily identifiable form, the eye is capable of estimating its amplitude within the envelope of the superimposed random noise with a fair degree of accuracy.

The method has two important disadvantages. Firstly, it is subjective and the measurement of the quasi peak-to-peak voltage of the noise depends upon the brightness level at which the waveform monitor is operated, the ambient lighting, and the state of adaptation of the eyes of the observer, as well as upon his judgment. In a test where experienced engineers measured a given noise level on a television signal independently and under their own preferred conditions of use of the waveform monitor, a range of 5 dB in the estimate of the quasi peak-to-peak value was recorded. Secondly, the various authorities are not in agreement on the value of the conversion factor from quasi peak-to-peak to r.m.s. voltage, and values ranging from 14 to 18 dB are given. This is evidently not very satisfactory, especially in view of the fact that when the random noise level is visible on a television picture, a change in its level of 1 dB is usually perceptible.

It should, perhaps, be added that the consistency of the subjective method is found to be very much improved when no signal is present, but in that case, of course, the noise voltage can be measured directly by means of an instrument.

A more elaborate method makes use of a television picture signal containing a uniformly illuminated area, produced either by means of a camera or synthetically, and isolates a portion of this signal by a time selection process; the result is a train of pulses with the noise level of the signal superimposed upon their horizontal portions. The modulating pulses are then subtracted from the latter signal so as to leave only a succession of bursts of random noise which can be measured as a power.

The disadvantage of this method is that the balancing processes involved entail a great deal of careful instrumentation, and it is difficult to reduce all spurious signals to negligible amounts in a manner which is stable over reasonable periods. Moreover, when only one end of the circuit is available, the method is much less practicable since in addition to this drawback the presence of the superimposed noise makes it difficult to reconstitute the required balancing signal accurately enough from the received signal.

A third method makes use of a signal of some very simple form, for example synchronizing signals with 'lift', which is band limited as far as is practicable by means of a low-pass filter with a slow roll-off and preferably a linear phase response to minimize overshoots.

At the output of the apparatus or circuit under test, the signal is passed through a complementary high-pass filter with the object of removing all the signal components, and it is then assumed that what remains is the random noise level. This residue is measured in some suitable way, and hence the signal-to-noise ratio is calculated.

A first objection to this method is that the noise cannot be measured over the lower part of the video band. The need to preserve some of the shape of the waveform and to avoid overshoots, which may increase the depth of modulation, result in the lowest frequency of the measured random noise band being perhaps as high as 1 Mc/s. On the other hand, in some practical cases the spectrum below about 1 Mc/s is known to be sufficiently uniform for the error introduced by the omission of this region to be unimportant.

The worst feature, however, is that any amplitude non-linearity in the signal chain results in the intermodulation of signal components to give spurious signals in the range of frequencies which is assumed to contain nothing but random noise. This is particularly important in the case of the long-distance television links which normally have a low noise level and at the same time often possess a fair amount of amplitude non-linearity.

In some such instances the error in measurement may well be very serious.

A feature which is common to all three methods just described is that only the total noise power is measured, that is to say the power integrated not only with respect to time but with respect to frequency over the band of frequencies concerned. This, of course, does not give the complete noise behaviour of the apparatus or circuit. Very often the spectrum of the random noise is not uniform with frequency, as occurs, for example, in the case of the triangular noise produced by the demodulation of a frequency modulation system, and it may be of importance to know what the spectrum is. Such information has proved very valuable in the selection and grading of camera tubes, for example, for quickly checking the amount of aperture correction in a signal source, and for numbers of other purposes.

2. Principles of the Method

The method presented in this monograph depends upon a fundamental difference between random noise and the television signal.

Each single impulse of the ensemble which composes a random noise signal is an isolated phenomenon in time; the instant at which it occurs has no correlation with the time of occurrence of any of the preceding or succeeding impulses. Its spectrum must therefore be continuous, that is between any two given points in the spectrum there will be an infinite number of frequency components having infinitesimally small magnitudes.[1] The ensemble of all such impulses, that is the random noise signal, must consequently also have a spectrum which has a continuous distribution, although the amplitude may vary as a function of frequency.

A signal which consists of line synchronizing and blanking pulses and any desired amplitude of 'lift' only has complete line-to-line correlation and its spectrum is therefore discontinuous. Energy exists only at integral multiples of the line repetition frequency. The introduction of field and picture synchronizing and blanking waveforms gives rise to an effective modulation of each component of the previous line spectrum by harmonics of the field, and to a lesser extent of the picture repetition frequency, which re-

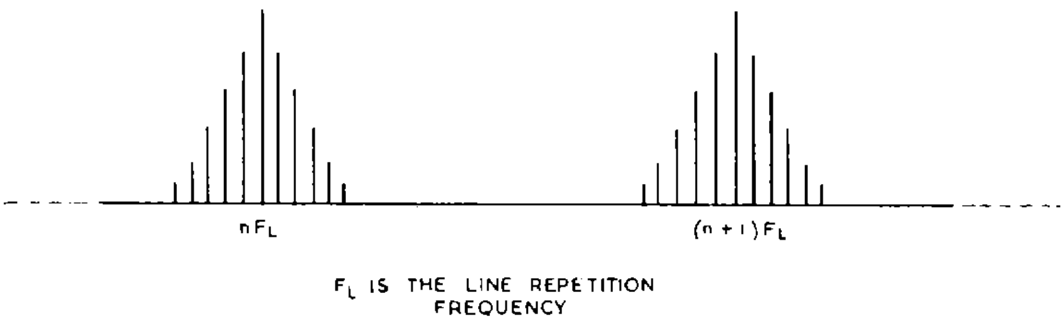


Fig. 1 — Portion of idealized video spectrum, free from noise

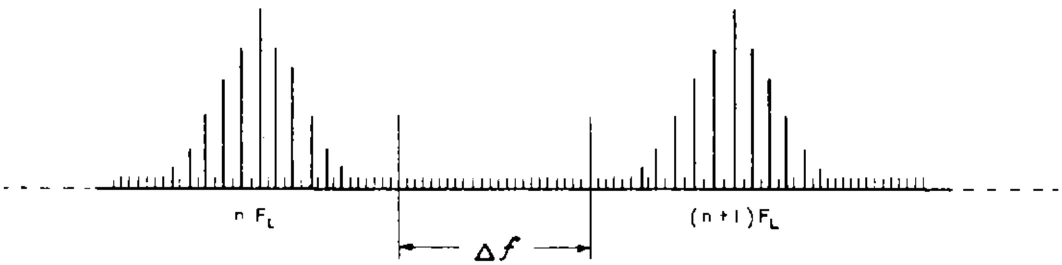


Fig. 2 — As Fig. 1, but with random noise present

sults in the type of spectrum shown in Fig. 1. Each harmonic of the line repetition frequency is surrounded by a symmetrical cluster of sidebands due to this modulation process, but, as was pointed out by Mertz and Gray in a classic paper on the subject,[2] with a picture consisting of a uniformly illuminated flat field the rate of decay of these sidebands is quite high, such that over almost all the frequency band of the signal approximately one-half of the interval between successive line harmonics contains extremely little energy.

However, since the random noise is uniformly distributed with frequency, the spectrum of such a signal with its random noise must have the appearance shown diagrammatically in Fig. 2. It is then evident that the power ΔW in the small, central frequency interval Δf must be free from components due to the signal and the ratio $\frac{\Delta W}{\Delta f}$, which may be written $\frac{dW}{df}$ with sufficient accuracy since the interval Δf is very small, represents the random noise power per cycle per second in the neighbourhood of the frequency of measurement nF. The total noise power in the complete range from zero to some upper limiting frequency F_m is then

$$W_T = \int_o^{F_m} \frac{dW}{df} \cdot df.$$

It will be shown in a later section of this monograph that such a measurement can be made conveniently and accurately, and that there is no difficulty in finding the total noise power when a single figure for the signal-to-noise ratio is required.

The only picture signal considered so far has been the uniformly illuminated field or, which is the same thing, synchronizing signals with 'lift'. With a complex still picture further sidebands of the line harmonics are produced which, in special cases, may even encroach on the intervals between neighbouring harmonics to give rise to spurious scanning components. A rapidly changing scene, as would be expected from the greatly reduced line-to-line correlation, corresponds to a highly diffuse spectrum of sidebands around each line harmonic. An alternative point of view is that used by Mertz and Gray,[2] who pointed out that move-

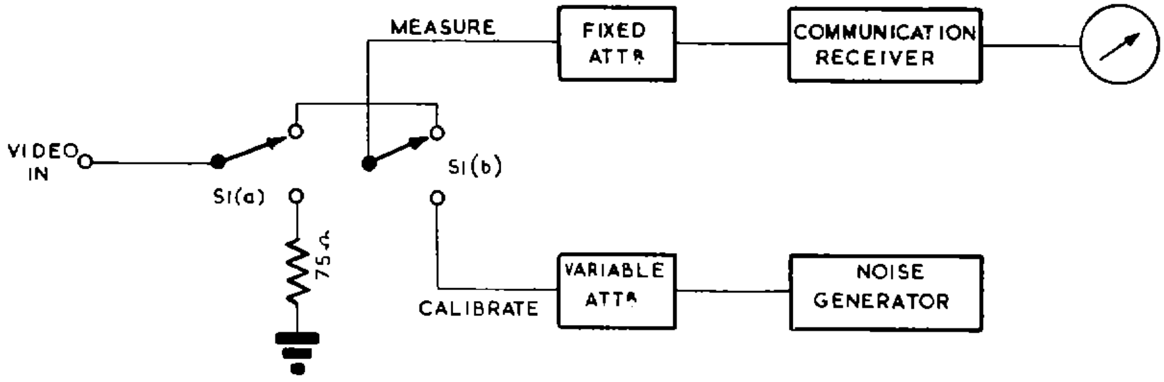


Fig. 3 — Schematic diagram of Method I

ment corresponds to a complex modulation of the picture signal components. The sidebands introduced by this modulation cause a spreading of the spectrum about each line harmonic towards, or even through, the 'empty regions'. Consequently, the accuracy of the measurement for a given random noise power will depend upon the type of picture which is being transmitted while the measurement is made, and the width of the measuring interval. However, it is quite rare in practice that the signal-to-noise ratio must necessarily be measured on a complex or moving picture.

3. Practical Description

3.1 *Method I*

Two methods of measurement have been developed. The first, which is shown schematically in Fig. 3, needs a minimum of specially designed apparatus. The ganged switches S1a and S1b, which are shown in the MEASURE position, connect the incoming video signal through a fixed 75-ohm attenuator to the input terminals of a sensitive communication receiver with a frequency range of 60 kc/s to 30 Mc/s, whose bandwidth can be varied in steps from 6,000 c/s to 100 c/s; this receiver has very good second-channel rejection.

A disadvantage of such a receiver for the present purpose is that its input circuit, even when the low-impedance 'Dipole' connection is in use, has an impedance which varies considerably over each frequency range and so does not terminate correctly the signal source presented to it.

This has been overcome by adding an input cathode-follower stage which contains an accurate 75-ohm resistor in its grid circuit; the output feeds directly into the nominally 100-ohm 'Dipole' input terminals. The source impedance of the cathode-follower is designed to be of the order of 100 ohms, so that on an average no loss in sensitivity occurs compared with the normal use of the receiver with a low-impedance aerial, and the alignment and selectivity of the receiver are unaffected.

The cathode-follower stage is contained in a compact screening box and is mounted directly on the inside of the front panel of the receiver; a coaxial socket is fitted so that the incoming signal lead can be plugged directly into its grid circuit with the 75-ohm termination. The connection to the 'Dipole' input terminals is made through a length of coaxial cable. This complete screening and careful decoupling of the supplies to the valve, which are taken from the receiver power pack, ensure that no extraneous noise or interference is picked up by the receiver input circuit.

Although a 1-volt peak-to-peak signal can be handled without distortion, sufficient sensitivity is available to make it possible to insert some attenuation before the receiver input, which completely avoids any possibility of error due to intermodulation between signal and noise components in the cathode-follower.

The audio-frequency output signal from the receiver is measured by means of a full-wave rectifier-type meter, which is designed to have a very close approximation to a square law characteristic over quite a wide range of amplitudes for frequencies up to 5 Mc/s. Careful comparison on random noise sources against a thermocouple has shown that the error is negligible for the purpose, and the instrument has the great advantage of being robust enough not to be harmed by any reasonable overload. The chief disadvantage of this type of r.m.s. reading meter for general work is that it has rather a large temperature coefficient, but in the present instance it is used as an indicator of equality of amplitudes and the absolute accuracy is of lesser importance.

In the CALIBRATE position of the switches the incoming signal is terminated in a 75-ohm resistor, and the receiver with its input attenuator is connected to an accurately calibrated 75-ohm variable attenuator in series with a white noise generator. The latter has been specially designed for this purpose, and gives a high output of random fluctuation noise, whose spectrum is uniform between approximately 100 kc/s and 5 Mc/s, into 75 ohms. The spectrum is limited by internal networks, which makes it possible to provide a calibrating meter so that the generator can be set up to give an accurate 1 mW of random noise into a 75-ohm load.

An output stage with a generous overload margin ensures that the statistical distribution of the noise peaks is modified to a negligible extent.

The measurement procedure is as follows. In the MEASURE position the receiver is tuned to the frequency at which the signal-to-noise ratio is desired. The region is carefully explored by means of the fine tuning control, and no difficulty is found in identifying two consecutive line harmonics. The receiver is then tuned to the position midway between these where the output is a minimum, and the gain is adjusted to give a standard reading on the output power meter. As the tuning is varied through this central area the output indication should pass smoothly through a minimum; if by any chance subsidiary maxima are noticed, parasitic frequencies are present and another neighbouring minimum should be tried.

The switches are then thrown to the CALIBRATE position and the variable attenuator is adjusted until the same reading is obtained on the receiver output meter; the amount of this attenuation is a measure of the signal-to-noise ratio and is noted.

The receiver bandwidth is normally set at the greatest value at which the interline harmonic minimum is clearly identifiable; most often a bandwidth of $1 \cdot 2$ kc/s is used, but this may be increased or decreased according to the type of picture on which the measurement is being made.

A very simple check is available in practice. Whatever the spectrum of the superimposed random noise may be, over the very small frequency range of the receiver bandwidth it may be taken to be linear with frequency with completely negligible error.

Then if the random noise alone is being measured, a change to the next lower bandwidth of the receiver will give no change in the attenuator setting, and hence no change in the indicated signal-to-noise ratio. The receiver gain has to be increased by an amount proportional to the change in bandwidth since the reduced noise accepted by the smaller band corresponds to a smaller input voltage, but this applies equally to the MEASURE and CALIBRATE conditions.

However, if the receiver bandwidth originally included some of the diffuse sidebands from the line harmonics on either side, the reduction in the receiver bandwidth would evidently decrease the output reading in the MEASURE condition by a greater proportion than in the CALIBRATE condition, and the signal-to-noise ratio would appear to change.

Although this test is very simple and quick to perform, after some experience has been gained with the method it only becomes necessary to apply it very occasionally, say at the beginning of a series of measurements. In fact, provided the picture is of a type which does not give highly diffuse sidebands surrounding the line harmonics, and whenever possible this should be arranged, the central region between the sidebands usually appears 'flat' to an extent which most people find rather surprising on first acquaintance.

The use of a standard random noise generator for calibrating the measuring receiver in the manner described above has three important advantages.

Firstly, the bandwidth and shape of the pass band of the receiver are completely eliminated from the measurement. The signal, both during measurement and calibration, is a continuous function of frequency so that any change in the receiver pass band alters the noise power falling in that band by the same amount in the two cases. This is important, since the inevitable tracking errors in the receiver and variations in the selectivity of the radio frequency stages with frequency would otherwise give rise to serious errors.

Secondly, the signal handled by the diode detector of the receiver has the same amplitude and is of the same nature on both the MEASURE and CALIBRATE positions, so that no error is introduced by the difference in behaviour of the detector with amplitude and waveform.

Thirdly, no retuning is needed during the calibration of the receiver. This is a great practical convenience and also reduces the time taken for each measurement by a very useful amount.

The above described method has been used extensively for a variety of purposes over a period of about two years. It has been found that readings are normally repeatable to within 1 dB, and where the result can be checked by another reliable method the agreement between the two answers is always excellent.

Contrary to what might at first be supposed, no difficulty has been found in training operators in the use of the method. In fact, the manual skill demanded in tuning the receiver is hardly more than is required for tuning into a broadcast programme, and the remaining operations are very straightforward.

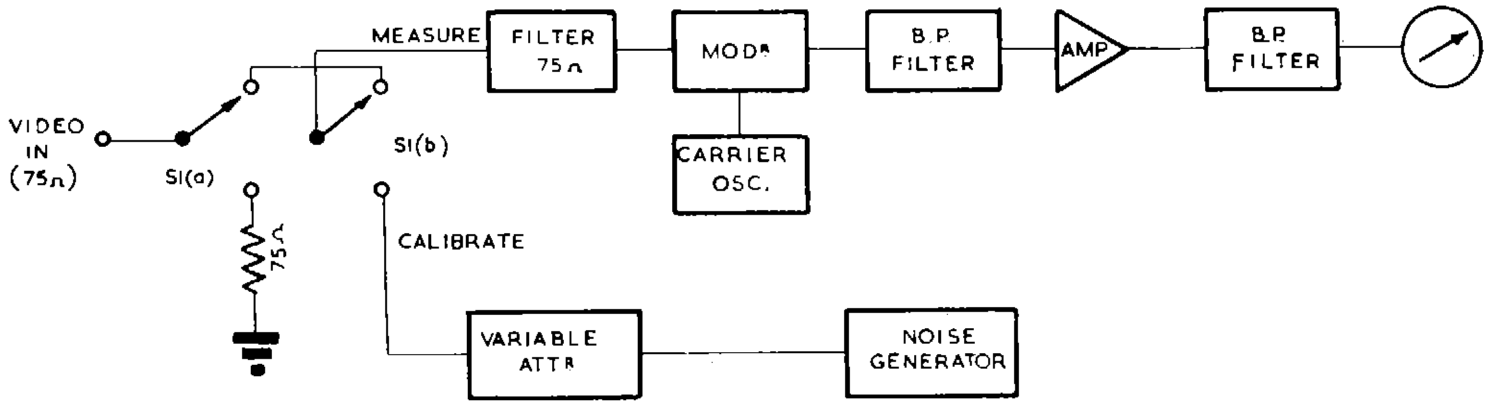
3.2 *Method II*

A more sophisticated version of this apparatus has been devised and in its prototype form has given very promising results. The schematic diagram of the method is shown in Fig. 4 (*a*), but since the instrument is still under development no circuit details will be given.

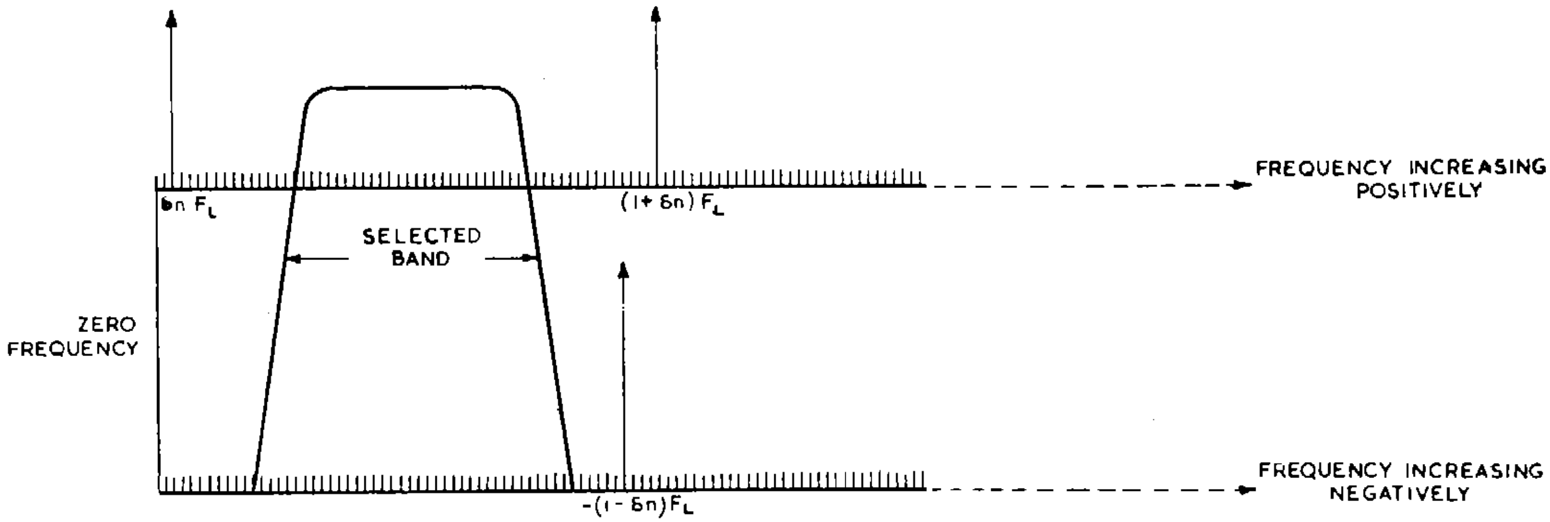
In the MEASURE position the incoming signal passes through a preselecting filter and then a modulator provided with a suitable variable-frequency carrier oscillator which covers the video range. The modulated signal passes through a narrow band-pass filter centred on approximately 5 kc/s for a 405-line input signal, through a high-gain amplifier with means for varying the gain, and then through another similar 5-kc/s narrow-band filter. Alternatively, the required selectivity may be incorporated into the amplifier. Finally, the modulated signal thus selected is measured by a meter which measures power.

In the CALIBRATE position the measuring instrument is connected through a variable 75-ohm attenuator to the standardizing noise generator, as in the previous method.

Suppose it is required to measure the power in the close neighbourhood of the nth line frequency harmonic nF_L. The carrier oscillator is set to some frequency very near this value, say $(n-\delta n)F_L$. If only the lower sideband is selected the result is as shown in Fig. 4 (*b*) where the sidebands corresponding to frequencies lower than nF_L are negative frequencies. Nevertheless, since positive and negative frequencies are not practically distinguishable they



(*a*) *Schematic diagram*



(*b*) *Selection of a narrow band of the modulated noise*

Fig. 4 — Method II

are both drawn in the direction of positive increasing frequency. The effect may be likened to a 'folding' of the spectrum about zero frequency which, because of the modulation process undergone by the signal, now represents a frequency very close to nF_L.

It is now evident that the low-frequency band-pass filters, or alternatively the selective amplifier, will pass a band of frequencies which is composed of almost exactly the centre portions of the two intervals between adjacent harmonics of the line frequency on either side of the chosen frequency. That is, they will pass the random noise falling into these intervals without interference from the signal, on the basis of the original argument, and this will be measured by the output power meter. The calibration is performed in exactly the same manner as described previously.

The oscillator frequency has purposely not been chosen to equal nF_L exactly, in order to demonstrate that a small but very useful tolerance in carrier frequency is available. This, in conjunction with the fact that each measurement can be made very expeditiously, appreciably lightens the requirement for oscillator stability. Nevertheless, it should be perfectly possible as a refinement to use an oscillator with automatic frequency control.

The object of the filter which is placed immediately in front of the modulator is to remove all signal components which might beat with harmonics of the carrier frequency to give spurious products falling into the measured band. This preselection does not need to be completely variable, but may consist of a number of suitably chosen band-pass or low-pass filters which are switched in as the oscillator frequency is changed, and other arrangements are available to suit special circumstances. A further possibility would be the use of a multi-pass-band or 'comb' filter at the input, with appropriate changes in the modulating circuitry, so that the combined noise of a number of central regions at various points in the video range could be measured simultaneously.

In a further projected model of this apparatus it is planned to attempt to simplify its use still more by removing the need for constant calibration. If the overall gain, including the frequency conversion, can be well enough stabilized, it may be possible to achieve sufficient accuracy with only an occasional recalibration against the noise generator. The instrument might then be made direct-reading, which would make such measurements even more convenient.

3.3 *Calculation of the Signal-to-noise Ratio*

Since each measurement only includes the noise power within a very small frequency range, it is evident that the total noise power, and hence the equivalent r.m.s. noise voltage, can only be determined if the total rectangular bandwidth which contains the noise is known or assumed in a given instance.

Accordingly, in a given measurement let the bandwidth of the incoming signal path be 3 Mc/s, and let the bandwidth of the calibrating noise generator be 5 Mc/s; this latter is a constant for a given generator and the figure is supplied with the instrument.

Suppose for the moment that the same output power reading is obtained from the receiver in both MEASURE and CALIBRATE positions with a setting of zero dB in the measuring attenuator. Then if the receiver bandwidth is Δf Mc/s, the power which it accepts from the calibrating source is $\Delta f/5$ milliwatt, since the generator output has been set to 1 mW in a bandwidth of 5 Mc/s.

If the noise power is assumed to be uniformly distributed over the 3 Mc/s of the signal path, then the total signal noise power is

$$\frac{\Delta f}{5} \times \frac{3}{\Delta f} = \tfrac{3}{5}\text{mW} = 0 \cdot 6 \text{ mW}$$

As would be expected, the receiver bandwidth disappears from the calculation. The equivalent r.m.s. voltage for a 75-ohm circuit is obviously $\sqrt{0 \cdot 6 \times 10^{-3} \times 75}$ volt.

The peak-to-peak signal voltage will be measured separately, probably by means of a precision waveform monitor. Let this be $0 \cdot 65$ volt peak-to-peak. Then the corresponding signal-to-noise ratio is

$$20 \log_{10} \frac{0 \cdot 65}{\sqrt{0 \cdot 6 \times 75 \times 10^{-3}}} \text{ dB} \quad \text{.........} (1)$$

However, in general a certain attenuation, say A dB, has to be inserted in the measuring attenuator in the CALIBRATE position of the switches, so that the final signal-to-noise ratio is

$$A + 20 \log_{10} \frac{0 \cdot 65}{\sqrt{0 \cdot 6 \times 75 \times 10^{-3}}} \text{ dB} \quad \text{.........} (2)$$

Obviously, for a given kind of measurement using the same apparatus the right-hand factor in the above expression is a constant, and in practice the known value in dB is simply added to the attenuator reading.

Where the spectrum of the random noise is constant with frequency, the expression (2) above gives the signal-to-noise ratio immediately. However, if this is not the case then a series of readings is obtained, each of which gives the signal-to-noise ratio which would be obtained if the random noise power at the frequency of measurement were uniformly distributed at that rate over the whole spectrum. That is, each individual value is calculated on the assumption that the spectrum is flat over the complete noise bandwidth.

In practice, a completely flat random noise frequency distribution is rarely obtained, but on the other hand over a wide field of measurements it is found that the range of variation is much less than might be supposed. Measurements made to date on transmission links and modern camera channels give distributions lying between flat and triangular with a greater tendency to approach the flat type of distribution. For example, television cable links and image orthicon camera channels rarely give more than three or four decibels change in distribution up to 3 Mc/s.

As will be shown, any type of distribution can readily be dealt with. The only advantage of the flatter spectra is that fewer readings need be taken.

The principle of the calculation consists firstly in finding the noise power equivalent to a given signal-to-noise ratio. For a 0·65-volt peak-to-peak picture signal the random noise power in a 75-ohm circuit corresponding to a signal-to-noise ratio of 20 dB is 56·3 μW, so that for 30-dB signal-to-noise ratio it is 5·63 μW and for 40 dB 0·563 μW. Any other values can very quickly be found by means of a slide-rule, or a simple calculator can be constructed.

Next, the noise power per megacycle is found. Since at the moment a 3-Mc/s video band is under consideration the noise power per megacycle corresponding to a signal-to-noise ratio of 20 dB is 18·8 μW, and proportionally for other signal-to-noise ratios. In cases where the noise bandwidth is a fixed quantity, as may often happen in television measurements, the noise power per megacycle could be found directly, and if a calculator is used the values would be indicated immediately without having to divide by the bandwidth.

Finally, the total noise power in the band is found from the measured values by a method of approximate integration, and then converted into the ratio of signal to the r.m.s. voltage corresponding to the total noise power. The whole operation is far less formidable to carry out than it seems at first glance, as the following example will demonstrate, and with a little practice it takes only a few moments.

Example:

In a certain image orthicon camera channel the measured values of signal-to-noise ratio were:

f (Mc/s)	0·5	1·0	2·0	3·0
Sig./Noise Ratio (dB)	38	38	37	35

The signal-to-noise ratio is constant below 1 Mc/s and worsens by 3 dB at 3 Mc/s, probably due to the action of the aperture corrector. The lowest reading can evidently be transferred to zero frequency in order to give a constant frequency increment of 1 Mc/s; this simplifies the approximate integration slightly but is by no means necessary. At the same time the noise power per megacycle can be written down for each reading:

f (Mc/s)	0	1	2	3
Sig./Noise Ratio (dB)	38	38	37	35
Noise Power per Mc/s (μW)	0·298	0·298	0·375	0·594

According to the familiar trapezoidal rule for approximate integration, the area under this curve is approximately equal to the frequency increment multiplied by the sum of the mean of the first and last ordinates together with all the other ordinates. That is:

Total Noise Power in 3-Mc/s band =

$$1\times[\tfrac{1}{2}(0\cdot298+0\cdot594)+0\cdot298+0\cdot375]+1\cdot119\ \mu\text{W}.$$

The signal-to-noise ratio for the complete noise band is then immediately obtained as 37 dB.

If preferred, there is an alternative method which avoids the use of the actual noise power. In this instance the noise power relative to the best signal-to-noise ratio, which is 38 dB in the example, is found by means of a slide-rule or table of logarithms, and is then divided by the bandwidth. Thus, at 2 Mc/s the signal-to-noise ratio is 1 dB worse, so that the noise power is antilog 0·1, that is 1·26 times what it is at zero frequency or 1 Mc/s.

The following table can then be drawn up:

f (Mc/s)	0	1	2	3
Noise Power relative to 38 dB	1·00	1·00	1·26	2·00
Relative Noise Power per Mc/s	0·33	0·33	0·42	0·67

The total noise power relative to 38 dB is:

$$1\times[\tfrac{1}{2}(0\cdot33+0\cdot67)+0\cdot33+0\cdot42]$$
$$=1\cdot25$$

The total noise power is therefore 10 log 1·25=1·0 dB worse than 38 dB, that is 37 dB as before.

Obviously, the calculation is simplified if the frequency intervals are chosen to be equal. If this is not so, it may be possible to split the calculation into more than one range. For instance, in the example given above the signal-to-noise ratio at zero frequency must still be assumed to be 38 dB; then the total noise power could be calculated as the sum of that between 0 and 1 Mc/s where the interval is 0·5 Mc/s and that between 1 and 3 Mc/s where the interval is 1 Mc/s.

In the most extreme case where all intervals are unequal, the approximate noise power for each interval would have to be calculated, and the sum of the noise powers for all intervals formed. However, this state of affairs would be unlikely to arise if one knew beforehand that the mean signal-to-noise ratio would be required, and in any case, except for detailed investigations, the total number of readings is normally very small.

As a matter of interest, the curve of noise power per megacycle in this instance was plotted in much greater detail and the area found by means of a planimeter. The resulting signal-to-noise ratio differed from the answer obtained above by only 0·1 dB. Since in any case the result of such calculations should be rounded off to the nearest decibel, because the measuring attenuator has only 1-dB steps, it is obvious that the error is quite unimportant.

It should be noticed that the above example is typical of many encountered in practice in that the noise spectrum is effectively defined by only three points, as it was in fact already known that the lowest part of the range would have a reasonably flat distribution. It is rarely that more than four or five readings need be taken and in some cases only two will suffice.

4. Limitations of the Method

Even in the case of the output of a generator of line synchronizing pulses only, where a true line spectrum should be obtained, there is always a small but measurable amount of signal of a random nature in the intervals between the line harmonics. This arises from the random noise in the generator and possible phase modulation due to time instability of the edges, as well as to the lock to the mains frequency and intermodulation with hum components. As the complexity of the signal increases, so does the amount of unwanted random components in these intervals, which puts a definite limit on the measurable signal-to-noise ratio in each case.

Naturally, under any given circumstances one should try to measure on the simplest type of signal possible, and most often some degree, at least, of choice can be exercised. A uniformly illuminated field or 'lift' is preferable to a complex picture, and the smallest amplitude suitable for the purpose should be employed. Sometimes it is possible to remove the field synchronizing and blanking components of the waveform without changing the random noise level, or even to remove the synchronizing waveform completely.

Fortunately, no great ratio is required between the circuit noise and the spurious noise since the total noise is measured as a power. If, for example, during a measurement of 1 mW of random noise 0·25 mW of spurious noise is present, the total measured is 1·25 mW which gives an error of almost exactly 1 dB, so that only a little less gives an error smaller than the accuracy of measurement and can therefore be ignored.

The spurious noise existing in typical waveform generators locked to the mains frequency was measured as described above under various conditions. Then, taking 1 dB to be the maximum tolerable error due to the presence of this spurious noise, the limiting measurable signal-to-noise ratio could be derived. With line synchronizing pulses only, which is the most likely condition for a transmission system, the figure was approximately 70 dB over the whole video band. With composite synchronizing signals the measurable signal-to-noise ratio varies with frequency in the manner shown in Fig. 5, from 46 dB at 500 kc/s to 75 dB at 4 Mc/s. The addition of full 'lift' lessens the measurable signal-to-noise ratio by about 3 dB at the lower frequencies, increasing to a reduction of about 10 dB at 4 Mc/s. This limitation has not been found at all serious in practice during the many measurements made on camera channels, although in this instance it is advantageous to remove the synchronizing signals from the output waveform of the channel.

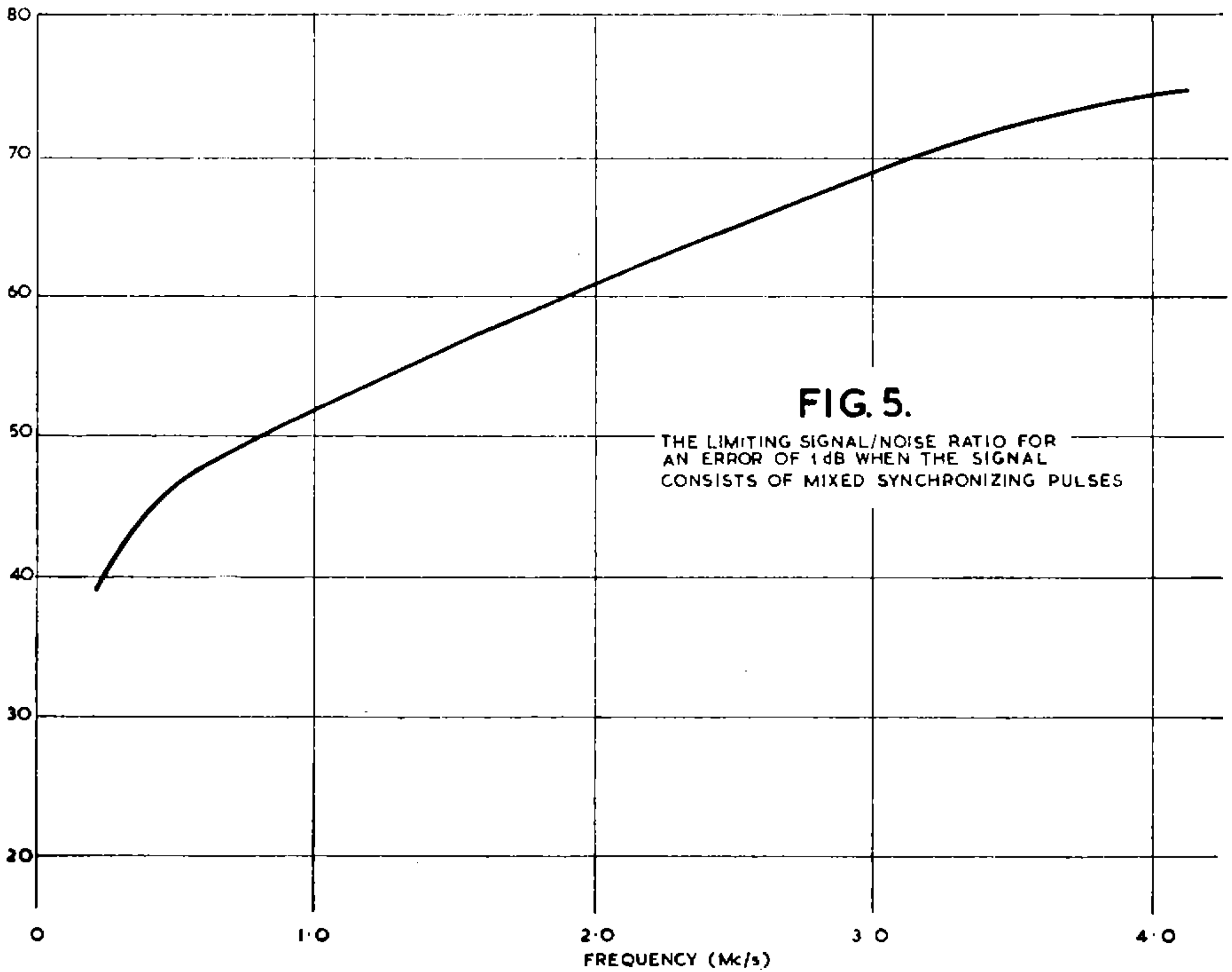


FIG. 5.

THE LIMITING SIGNAL/NOISE RATIO FOR AN ERROR OF 1 dB WHEN THE SIGNAL CONSISTS OF MIXED SYNCHRONIZING PULSES

5. Acknowledgment

Finally, the author would like to acknowledge his indebtedness to his colleague Mr J. E. Holder, in particular for his invaluable assistance with the work on the design of the calibrating noise source and for the suggestion of the principle of Measurement Method II.

6. References

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SUMMARIES OF SOME RECENT BBC PATENT APPLICATIONS

PAT. APP. NO. 29302/57

IMPROVED TELEPRINTER TRANSLATOR

Inventors: H. C. CHANDLER and F. L. COOMBS

The statement of invention reads:

According to the present invention there is provided a relay circuit arrangement comprising two transistors of opposite conductivity types, means for applying input voltages to the bases of the transistors, an output circuit connected in series between the two emitters and a point of fixed potential, and means for connecting the two collectors to terminals of a current source which are of opposite polarity relatively to the point of fixed potential. The transistors may be of the junction type, one being P–N–P and the other N–P–N. Alternatively, point contact transistors of P and N type may be used.

The invention is a transistor device which enables a teleprinter to be worked directly from a line or from potentials derived from a tone without the need of a telegraph relay.

PAT. APP. NO. 30741/57

DIRECTIONAL LOUDSPEAKER

Inventor: D. E. L. SHORTER

The statement of invention reads:

According to the present invention there is provided a loudspeaker comprising two vibratory diaphragms arranged spaced apart in the direction of their common axis of vibration, means for generating from acoustic signals to be reproduced derived signals in which the amplitudes at lower frequencies are increased relatively to the amplitudes at higher frequencies, and means for applying the derived signals to drive the diaphragms in phase opposition.

The invention is a highly directional loudspeaker of small size suitable for a loudspeaker telephone communication system.

PAT. APP. NO. 32061/57

A 90° PHASE-SHIFT NETWORK SUITABLE FOR GENERATING A VESTIGIAL-SIDEBAND TELEVISION SIGNAL

Inventors: G. G. GOURIET, G. F. NEWELL, and W. PROCTOR WILSON

The statement of invention reads:

According to the present invention there is provided an electrical phase-shifting network comprising a delay device terminated at one end and provided with a plurality of output terminals so connected thereto that in response to the application of an input signal to an input terminal of the device an output signal appears at each output terminal after a predetermined time delay τ, which is different for different output signals, means for adjusting, varying or setting the amplitudes of the output signals in such a manner that relatively to the amplitude of a predetermined one of the output signals appearing at a predetermined one of said output terminals and for which the delay τ has a predetermined value τ_a, the amplitudes of the other output signals diminish with increasing values of the modulus of τ', where $\tau'=\tau_a-\tau$, and have a sign dependent upon the sign of τ', and means for adding together said other output signals so adjusted to produce a combined output signal.

The network described in the invention causes the rotation by 90° of every Fourier component of a television signal and provides controllable attenuation of the low frequencies without introducing group-delay distortion.

PAT. APP. NO. 12011/58

IMPROVED MICROPHONE WINDSHIELD

Inventors: H. D. HARWOOD and R. J. PACKER

The statement of invention reads:

According to the present invention, there is provided a microphone having around it a windshield through which the flow resistance per unit area is greater in the path of wind from a direction which produces relatively high wind noise than in the path of wind from a direction which produces a relatively low wind noise.

The invention describes a microphone windshield that can be made of smaller size, for a given acoustic performance, than known windshields.

PAT APP. NO. 24116/58

IMPROVED ACKERMANN STEERING MECHANISM

Inventor: L. E. H. O'NEILL

The statement of invention reads:

According to the present invention a steering mechanism comprises a steering column rotatably mounted in a gear carrier plate and having a first toothed gear fixedly mounted thereon, the gear carrier plate being movably supported upon a frame and having rotatably mounted thereon second and third toothed gears driven directly or indirectly by the first gear, first and second wheel shafts rotatably mounted in the frame and carrying road wheels and levers provided with driving pins working in slots in the second and third gears respectively, and one or more eccentrics engaging between the gear carrier plate and the frame, rotatable by the steering column, and serving to displace the gear carrier plate relatively to the frame in such a manner that the axes of the second and third gear wheels are substantially coincident with the axes of the first and second wheel shafts respectively when the road wheels are set for straight course and depart increasingly from coincidence as the position of minimum turning circle is approached.

The invention has for its principal object to provide an improved steering mechanism with which a given degree of accuracy can be obtained in a relatively simple manner over a wider range of turning circle radii than in known Ackermann mechanisms.

PAT. APP. NO. 27878/58

MOVABLE-SPINDLE FILM MAGAZINE

Inventor: N. F. CHAPMAN

The statement of invention reads:

According to the present invention there is provided spooling means comprising, within a magazine, a support carrying a take-off and a take-up spindle, together with means for automatically causing the said support to move relatively to the magazine during the film run in a direction to bring the take-off spindle closer to the wall of the magazine and the take-up spindle further from the wall of the magazine.

The invention provides a magazine which, for a given length of film or tape contained therein, is of considerably smaller size than conventional magazines.

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