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IN THIS ISSUE

ELECTRONIC INSTRUMENTS FOR TELECOMMUNICATIONS AND INDUSTRY

RESUME D'ARTICLES PUBLIES DANS LE PRESENT NUMERO

MESURES DE BRUIT A L'AIDE DE VOLTMETRES ELECTRONIQUES

Les facteurs qui influencent la mesure de bruit gaussien à l'aide de voltmètres électroniques ne répondant pas à la valeur efficace réelle font l'objet de la présente étude. Les caractéristiques de bruit gaussien et la relation entre le bruit et les sinusoïdes contenues dans les paramètres qui affectent les lectures d'instruments sont discutées. A partir des valeurs relatives du facteur de forme pour une sinusoïde et pour le bruit, il est possible d'obtenir des résultats quantitatifs avec des voltmètres à réponse moyenne. Des résultats analogues pourront être acquis grâce à des instruments à réponse de pointe, sous condition que l'on connaisse le régime de charge et de décharge du détecteur d'entrée. Il est aussi très important, quand on utilise des voltmètres électroniques, de connaître le facteur de pointe et la largeur de bande. Il faut, de plus, considérer le temps d'intégration lorsqu'on s'intéresse au cas de bruits de très basse fréquence ("flicker", par exemple).

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GENERATEUR DE BRUIT BLANC A 12 VOIES TYPE TM 7816

L'instrument décrit étend les techniques usuelles de charge des circuits en bruit blanc aux équipements multiplex. La possibilité de tester le système multiplex seul ou accompagné des systèmes à faisceaux hertziens à faible capacité est aussi discutée. Le bruit parasite, mesuré, peut s'exprimer en picowatts, en dBa, en dBrnC ou en rapport signal/bruit. Les influences de la distorsion d'intermodulation et de la diaphonie dans des équipements multiplex peuvent être mesurées individuellement, ou globalement.

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OSCILLOSCOPE POUR ESSAIS DE FAISCEAU RADAR TYPE TF 2200/6M1

Il est possible de mesurer la linéarité de l'équipement de faisceaux de données radar à l'aide de techniques dérivées. Le dispositif d'affichage associé doit être accompagné d'une voie suivant l'axe des X et d'une possibilité d'écrans anti-radar et de gains élevés suivant l'axe des Y. Une version spécialement modifiée de l'oscilloscope type TF 2200 est à présent disponible qui satisfait à ces exigences.

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L'ANALYSE SPECTRALE À BANDE ÉTROITE ET L'ANALYSEUR SPECTRAL OA 1094A/3

L'analyse des spectres en H.F. sur des bandes étroites est une exigence très difficile à satisfaire par un analyseur spectral. Cependant, l'appareil d'analyse de spectre H.F. type OA 1094A/3, tout en conservant la totalité des fonctions de son prédécesseur, le OA 1094A, convient pourtant parfaitement à ce genre de travail grâce à ses performances améliorées dans ce domaine. Les possibilités nouvelles de l'instrument sont détaillées en même temps qu'il est fait mention de certains phénomènes qui peuvent réduire la qualité des affichages de spectres à bande étroite.

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ZUSAMMENFASSUNG DER IN DIESER NUMMER ERSCHEINENDEN BEITRÄGE

RAUSCHMESSUNGEN MIT ELEKTRONISCHEN SPANNUNGSMESSGERÄTEN

Es werden die Faktoren untersucht, welche einen Einfluß auf die Messung von Gaußchem Rauschen mittels elektronischer Spannungsmeßgeräte haben, die nicht den wahren Effektivwert anzeigen. Die hierbei wichtigen Eigenschaften des Gaußchen Rauschens und das Verhältnis zwichen Rauschen und sinusförmigen Spannungen bei den die Anzeige beeinflussenden Parametern werden besprochen. Bei bekannten relativen Werten des Formfaktors für sinusförmige Spannungen und für Rauschen können quantitative Messungen mit einem Instrument für die Anzeige von Mittelwerten gemacht werden. Ähnliche Ergebnisse lassen sich mit Spitzenwertanzeigern erreichen, wenn die Zeitkonstanten für Ladung und Entladung im Eingangsgleichrichterkreis bekannt sind. Bei Benutzung von elektronischen Spannungsmeßgeräten muß man außerdem den Spitzenfaktor und die Bandbreite wissen. Bei Rauschen mit sehr niedriger Frequenz muß die Integrationszeitkonstante ebenfalls mit in Betracht gezogen werden.

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RAUSCHGENERATOR TYPE TM 7816 FÜR MESSUNGEN MIT WEISSEM RAUSCHEN BEI 12 KANÄLEN

In diesem Aufsatz wird ein Gerät beschrieben, mit welchem die zur Zeit angewendete Methode der Intermodulationsmessung mit weißem Rauschen auf Multiplexeinrichtungen ausgedehnt werden kann. Methoden zur Prüfung von Multiplexeinrichtungen allein oder von ganzen Mikrowellen-Richtfunkanlagen mit geringer Kanalzahl werden diskutiert. Das Fremdrauschen kann in Pikowatt, dBa, dBrnC (bezogen auf den Normalrauschpegel 1 pW mit C-Filter) oder als Rauschabstand im Kanal gemessen werden. Die Auswirkungen von Übersprechen und Intermodulationsverzerrungen in Multiplexeinrichtungen können getrennt oder zusammen gemessen werden.

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OSZILLOGRAPH TYPE TF 2200/6M1 FÜR MESSUNGEN AN RADAR-RICHTFUNKSTRECKEN

Bei Radar-Datenübertragungsanlagen kann die Linearität durch Bestimmung der Ableitungen entlang der Amplitudenkurve gemessen werden. Die hierzu benötigte Anzeigeeinrichtung muß einen wechselstrommäßig gekoppelten X-Kanal besitzen, während der Y-Kanal eine hohe Verstärkung und einen großen dynamischen Bereich haben muß. Um diese Anforderungen zu erfüllen, ist jetzt eine speziell abgeänderte Ausführung des Oszillographen TF 2200 lieferbar.

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SCHMALBANDIGE SPEKTRUMANALYSE MIT DEM SPEKTRUMANALYSATOR TYPE OA 1094A/3

Die Analyse schmalbandiger Bereiche des Hochfrequenzspektrums stellt sehr strenge Anforderungen an einen Spektrumanalysator. Der Spektrumanalysator OA 1094A/3 für den HF-Bereich ist für diesen Zweck hervorragend geeignet. Seine Arbeitsweise wurde für diese Betriebsart verbessert, während alle Funktionen des Vorgängers OA 1094A beibehalten wurden. In diesem Aufsatz werden die zusätzlichen Eigenschaften dieses neuen Gerätes beschrieben und gewisse Erscheinungen erwähnt, welche die Anzeige schmaler Spektralbereiche verschlechtern können.

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EDITORS

P. M. RATCLIFFE, A.M.I.E.E.

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For some exercises the peak value is the important parameter as any climber knows—

Mount Snowdon, North Wales

J. Allan Cash

Peak mean or r.m.s.

FROM AUDIO FREQUENCIES to microwaves measuring an a.c. voltage is simple—until r.m.s. is specified. By definition, the r.m.s. value of an alternating waveform is determined by comparison with a direct voltage; the two are equivalent when each produce the same heating effect in identical resistors. This is not just an academic definition as present standards of r.f. voltage measurement are by reference to d.c. voltages. It is also a technique used in our range of r.f. absorption power meters, where a thermocouple is used as the sensing element to ensure that the measured voltage is independent of waveform.

Unfortunately, the paradoxical situation exists that the most versatile electronic voltmeters, in respect of frequency range, sensitivity, stability, time constant and, by no means least, economy, do not measure the r.m.s. value. The majority of them sense either the mean, or peak, or peak-to-peak value of a waveform.

When commercial voltmeters were first marketed, electronics was an infant mainly concerned with communications where the sine wave predominated. Thus the convention was established, although deprecated by some, of calibrating electronic voltmeters in terms of the r.m.s. value of a sine wave. When used on other waveforms accurate results are still realisable provided the form or peak factor is known and the appropriate correction made to the indication.

Nowadays white noise is being increasingly used for dynamic testing of systems, and because of its complex nature it is often assumed that true r.m.s. voltmeters are essential to determine the voltage. For this reason the characteristics of our peak and mean reading meters have been examined, both theoretically and experimentally, and in the article of page 18 correction factors are given to convert the meter indication to the true r.m.s. value. For completeness, this information on noise measurement has also been given for our new peak-to-peak Electronic Millivoltmeter type TF 2603 which we are confident will further advance the art of voltage measurements, as will be seen when details are published in a future issue. P.M.R

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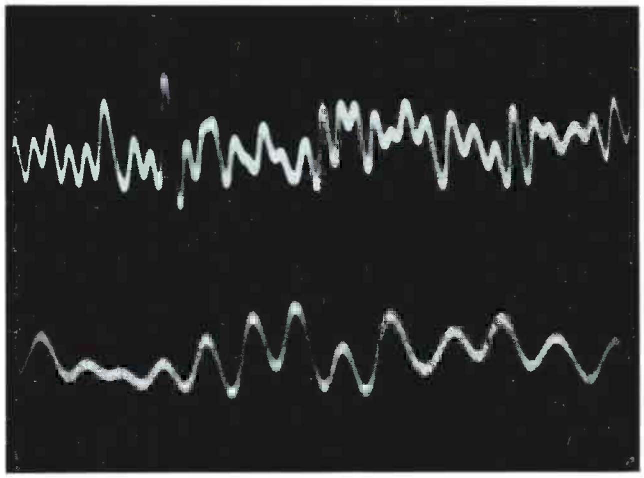


Fig. 1.

Noise measurements

WITH ELECTRONIC VOLTMETERS

by
P. BRODERICK,
B.Sc., Grad. Inst. P.

The factors affecting the measurement of Gaussian noise on electronic voltmeters which do not respond to the true r.m.s. value are examined. The relevant characteristics of Gaussian noise and the relationship of noise to sinusoids in those parameters which affect meter reading are discussed. Knowing the relative values of the form factor for a sinusoid and noise, quantitative results can be obtained with average responding voltmeters. Similar results may be obtained with peak responding meters providing the rate of charging to discharging times of the input detector is known. In addition it is important when using electronic voltmeters to know the peak factor and bandwidth; for very low frequency noise the integrating time should also be considered.

MOST VOLTMETERS are designed to measure repetitive waveforms and to indicate the r.m.s. amplitude of an equivalent sinusoid. The response of the meter is, however, not necessarily to the r.m.s. value but most often to the average value or the peak or peak-to-peak amplitude. How the meter responds dictates how to interpret the readings when noise or non-repetitive waveforms are being measured. Before considering the effect of different responses to noise it is useful to summarize some of the salient features of noise itself.

Some characteristics of noise

The sources of electrical noise are many and varied, and result in several types of power spectrum. For example, electrons in thermal equilibrium with the molecules in a conductor display a Brownian motion in the form of thermal noise, so that all conductors produce thermal noise (also called Johnson noise) to some extent. Thermionic valves produce noise of a different type, namely 'shot noise' arising from the random rate of arrival of electrons at the anode of the valve. Boundaries or contacts in composition resistors and in semiconductors give rise to fluctuation noise (also called flicker or 1/f noise); this is thought to be due to a random fluctuation in the conductivity across a boundary as charges traverse it. Noise also arises from such causes as atmospheric disturbances and multiple interference from external coherent sources.

Thermal and shot noise are characterized by having a Gaussian distribution of relative amplitudes. The spectral density is constant over all frequencies, so that the noise is referred to as 'white noise' and is the commonest type in the audio and high frequency bands. This is the noise to be considered here. Fig. 1 is an oscillogram of the output of the TF 2091 White Noise Generator at two different time base rates. By contrast, contact noise has a 1/f spectrum, i.e., the spectral density is largest at very low frequencies. Usually, at frequencies above a few hundred c/s this type of noise is relatively unimportant.

Gaussian noise

Since noise has no defined amplitude it must be described by statistical means and most text books on noise derive the probability distribution of amplitudes as:—

$$p(v)dv = \frac{1}{\sigma\sqrt{2\pi}} \exp(-v^2/2\sigma^2)\, dv \quad \ldots\ldots\ldots\ldots (1)$$

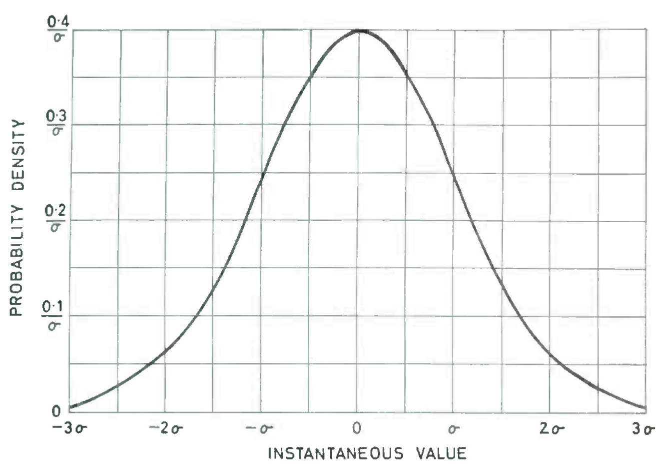


Fig. 2. Probability density function for a random variable

This is the probability that the instantaneous value of the noise will lie between v and $v+dv$, where σ is the r.m.s. value. From this equation one can plot the familiar normal distribution for a random variable—Fig. 2. The area under this curve between any two limits is the probability that the instantaneous value will lie between these limits.

Mean value and form factor for Gaussian noise

The mean absolute value, $\bar{v}$, of Gaussian noise is obtained using equation (1) directly, integrating over all values of voltage multiplied by the probability of getting that voltage. In fact, since the function is symmetrical the answer for positive values can be doubled.

$$\bar{v} = \frac{2}{\sigma\sqrt{2\pi}} \int_0^\infty v \exp(-v^2/2\sigma^2) dv$$

Therefore

$$\frac{\bar{v}}{\sigma} = \sqrt{\frac{2}{\pi}} = 0{\cdot}798 \quad \ldots\ldots\ldots\ldots (2)$$

The form factor is the reciprocal of (2), viz:

$$\text{Form factor} = \frac{\text{r.m.s.}}{\text{average}} = \sqrt{\frac{\pi}{2}} \quad \ldots\ldots\ldots\ldots (3)$$

Peak factor for Gaussian noise

To understand the action of peak responding meters for random noise inputs it is necessary to have a quantitative description of the probability of occurrence of peaks greater than the r.m.s. value. This follows from Fig. 2 but it is easier for numerical reference to plot a curve

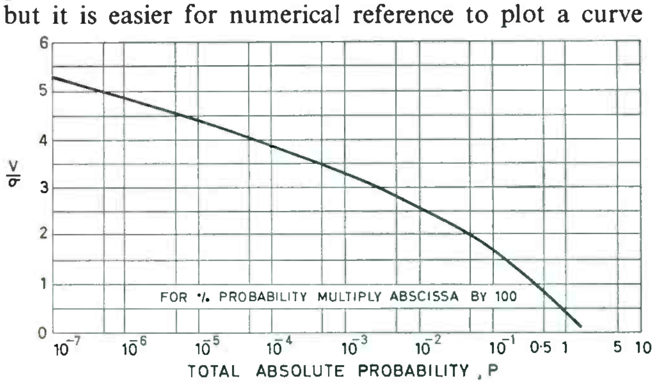


Fig. 3. Probability of peak voltages exceeding r.m.s. value

showing this probability directly. The curve must display the equation:

$$P = \int_v^\infty p(v) dv$$

$$= \frac{1}{\sigma\sqrt{2\pi}} \int_v^\infty \exp(-v^2/2\sigma^2) dv \ldots\ldots\ldots\ldots (4)$$

Equation (4) is plotted using tables[1] and is shown in Fig. 3 where both positive and negative going voltages are taken into account in showing the total absolute probability of voltages greater than the r.m.s. value occurring.

Typical values of peak factors for Gaussian noise are shown in table 1 below which is often useful if a curve is not available. From the table it can be seen that a meter with a peak factor of about 2·6 will ignore peaks which occur less than 1% of the time. On the other hand a peak factor of 4·4 ignores peaks occurring less than 0·001% of the time. Clearly less error is incurred in the latter than the former in the measurement of the true r.m.s. value.

The error involved in the direct measurement of r.m.s. noise is easily evaluated from the percentage change in average value resulting from restricting the upper limit of the integration for $\bar{v}$ to a definite multiple of the r.m.s. value, σ. Fig. 4 shows a plot of the percentage error in the indicated value for different peak factors evaluated in this way. Hence, if peak clipping occurs the meter reads low and from the curve it is possible to deduce by what percentage to increase the reading to obtain the true r.m.s. value of the noise.

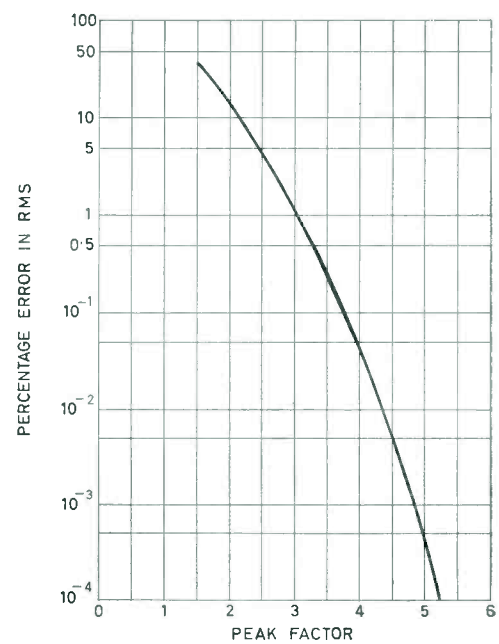


Fig. 4. Percentage error in r.m.s. indicated value for different peak factors

Table 1

Peak factors for Gaussian noise

% of time peak factor is exceeded	Peak factor $= \frac{v}{\sigma}$	$20 \log_{10} \frac{v}{\sigma}$ dB
10	1·645	4·32
1·0	2·576	8·22
0·1	3·290	10·35
0·01	3·900	11·82
0·001	4·430	12·9
0·0001	4·900	13·8
0·00001	5·300	14·5

Effect of bandwidth

The average power of noise is directly proportional to bandwidth so that the r.m.s. voltage is proportional to the square root of the bandwidth. Thus for resistor noise for example:—

$$\overline{v^2} = 4KTR\triangle F$$

where K = Boltzmann's constant,
T = absolute temperature,
R = value of resistor,
and $\triangle F$ = bandwidth.

A voltmeter should therefore have a bandwidth greater than or equal to the noise bandwidth, or noise power will be lost. If the meter bandwidth is reduced below that of the noise the final indication must be corrected by proportionality. For example, if a noise bandwidth is 100 Mc/s and the voltmeter 3 dB pass band is 10 Mc/s and shows an indication V_1 r.m.s., the true value V_2 is given by:

$$V_2 = \sqrt{10}\, V_1.$$

It is now possible with the above information to correlate the indications of different types of meter with the true r.m.s. value of a noise input.

True r.m.s. meter

A true r.m.s. responding meter can always be used to measure Gaussian noise provided the bandwidth criterion outlined in the previous section is taken into account. The meter will indicate directly the true r.m.s. value of the input noise subject to some error due to a finite peak factor.

True r.m.s. meters are designed to deal with signals with very small duty cycles, e.g., 1% or less and in this case the peak factor may be shown to be given by:

$$\text{Peak factor} = \sqrt{\frac{1}{D} - 1}$$

where D = duty ratio.

For example, if the duty cycle is 1%, $D = 1/100$ and peak factor = 10. Accordingly the amplifiers must have sufficient dynamic range to handle signals 10 times the full scale.

For Gaussian noise it is clear that a crest factor of only 4·4 results in the clipping of peaks occurring less than 0·001% of time and consequent negligible error (see Fig. 4), so that such peak factors of 6 to 10 that are generally found in r.m.s. meters are more than adequate when measuring noise. True r.m.s. meters however do not have very wide bandwidths and are not as common in a laboratory as the average or peak responding meters.

Average responding meter

The widely used type of average responding voltmeter such as the Sensitive Valve Voltmeter type TF 2600 responds to the average value of a rectified waveform and indicates the r.m.s. value of an equivalent sinusoid. When used for measuring noise the meter reading is related to the true r.m.s. value by a constant factor of 1·05 dB.

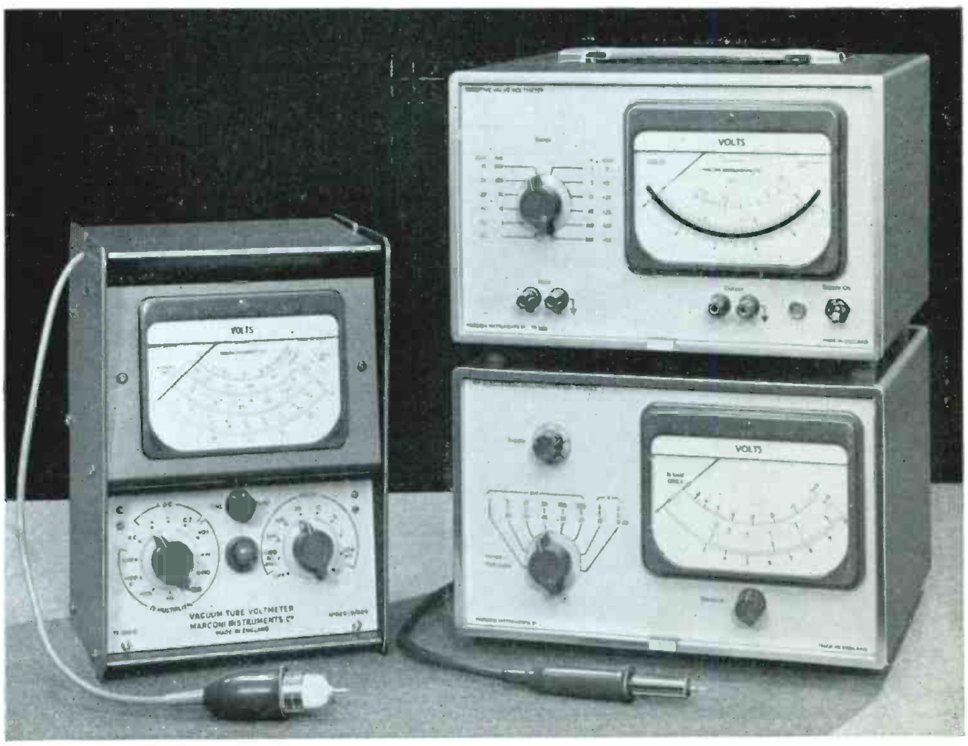
This is shown as follows:

$$\text{For a sinusoid the r.m.s. value} = \frac{\text{peak}}{\sqrt{2}}$$

$$\text{the average value} = \frac{2}{\pi}\ \text{peak}$$

$$\text{therefore the form factor} = \frac{\text{r.m.s.}}{\text{average}} = \frac{\pi}{2\sqrt{2}} = 1{\cdot}11 \quad ..(5)$$

The meter responds to the average value and reads 1·11 times this to give the equivalent r.m.s. value. The



From the Marconi Instruments range: Vacuum Tube Voltmeter Type TF 1041C, Sensitive Valve Voltmeter Type TF 2600, Electronic Millivoltmeter Type TF 2603 same applies to Gaussian noise, thus:

$$\text{From equation (3) form factor} = \sqrt{\frac{\pi}{2}}$$

$$\text{i.e., the average value} = \frac{\text{r.m.s.}}{\sqrt{\frac{\pi}{2}}}$$

$$\text{The meter therefore indicates } 1{\cdot}11 \frac{\text{r.m.s.}}{\sqrt{\frac{\pi}{2}}} = 0{\cdot}886 \text{ r.m.s.}$$

$$\left.\begin{aligned} \text{therefore indicated reading} &= 0{\cdot}886 \times \text{ true reading} \\ &= -1{\cdot}05\,\text{dB} \end{aligned}\right\} \quad \ldots (6)$$

The criterion regarding bandwidth must of course be observed and also it is important to ensure that the voltmeter peak factor does not cause excessive clipping of large peaks. The peak factor for sinusoids = 1·414 so that a meter designed primarily for sinusoid measurements need only deal with peaks of this order. The TF 2600 amplifier however has a dynamic range which exceeds its full scale deflection by 8·8 dB. If the range selected when measuring noise is such that only the first 70% of the scale is used, peaks up to 12 dB above the r.m.s. value will not be clipped. The peak factor will therefore be equal to 4 (increasing of course for values less than 70%) and the error incurred (from Fig. 4), is less than 0·04%. Less than 1% error is obtained in the measurement of r.m.s. values for readings up to 92% of f.s.d. At full scale the error is 2%. Since the accuracy of true r.m.s. voltmeters is seldom better than 1% there seems very little to gain in using them for noise measurements.

White noise from the TF 2091 Noise Generator was used with a selected low pass bandwidth of 4·028 Mc/s and measured on several ranges of the TF 2600 Voltmeter. The input to the voltmeter was also measured on a standard true r.m.s. thermocouple meter and the two readings compared. The graph in Fig. 5 shows how closely the theoretical value of equation 6 is approached.

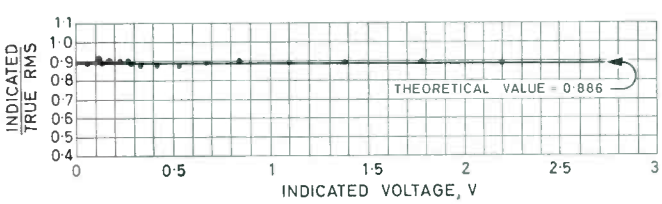


Fig. 5. Relationship between indicated value and true r.m.s. value of Gaussian noise for TF 2600 Sensitive Valve Voltmeter

Peak responding meter

The peak voltmeter relies upon the output circuit of the diode rectifier having so long a time constant compared with the lowest frequency component of the input, that a steady bias voltage Vb is maintained across the diode. Beranek[2] has shown that if the forward charging resistance, Rc, is constant and fixed by the diode plus source impedance and the reverse discharge resistance is independent of the diode and source impedances, then the bias voltage is derivable from the amplitude distribution $p(v)$ of the input voltage. The equation he obtained equates the current leaving the detector capacitor via the discharge path, Rd, to that entering it via the diode as an equilibrium condition:

$$\text{Thus:—} \quad \int_{Vb}^{\infty} \frac{v\text{-}Vb}{Rc}\, p(v)dv = \frac{Vb}{Rd}$$

where v = instantaneous input voltage

$$\text{rearranging gives:} \quad \frac{1}{Vb}\int_{Vb}^{\infty} vp(v)dv - \int_{Vb}^{\infty} p(v)dv = \frac{Rc}{Rd} \quad ..(7)$$

$p(v)$ is given by equation (1) so that the first integral is evaluated directly in terms of Vb. The second integral is obtained from tables as the area under the normal curve beyond Vb. Since the meters are calibrated to read r.m.s. with a sinusoidal input the mean level, Vb, is directly related to the r.m.s. indication. If the ratio of indicated to true r.m.s. is plotted against Rd/Rc the curve shown in Fig. 6 is obtained. The curve allows one to estimate

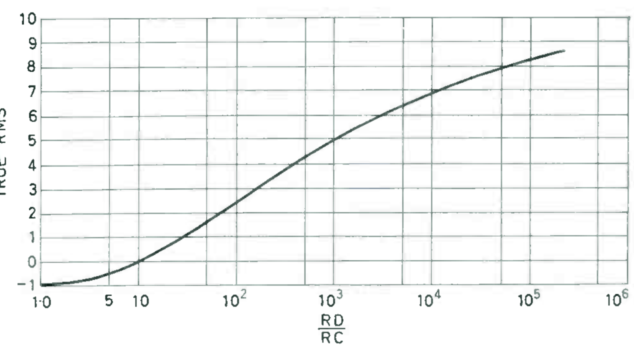


Fig. 6. Ratio of indicated to true r.m.s. for various charge-to-discharge ratios

the factor with which to multiply the peak reading meter to get the r.m.s. value. In practice, however, the curve can serve only as a guide since the charging resistance, Rc, depends upon the diode resistance which varies with input voltage. Furthermore, at very low voltages the diode has an approximate square law characteristic so that the multiplying factor will not be constant but approach unity in this region.

The TF 1041C and TF 2603 are examples of peak responding voltmeters. The first is a peak detecting meter and the second a peak-to-peak meter. Both indicate the r.m.s. value for an input sine wave. Figs. 7 and 8 show curves of indicated to r.m.s. readings versus the indicated values for both these instruments and show the correction factor to be applied to the voltmeter reading to obtain the true r.m.s. value of the noise. Both instruments show a transition from peak detecting to true r.m.s. as the voltage decreases. The TF 2603 uses a semiconductor diode so that the region at which it begins to be square law is lower (50 mV approximately) than for the TF 1041C. By the use of capacitive attenuators the region over which the TF 2603 is true r.m.s. reading can be extended to cover all ranges of the instrument. This is a useful facility since the frequency characteristic of peak voltmeters is far superior to the usual true r.m.s. voltmeter. It must be remembered again that the meter bandwidth should exceed the noise bandwidth or the appropriate correction must be applied. In the above measurements the noise bandwidth was approximately 4 Mc/s from a 75 Ω source impedance.

If low frequency noise is being measured a larger time sample of the noise may be selected, thus rendering the meter reading insensitive to the infrequent large spikes of random noise, by inserting a resistor in series with the detector. This of course adds to the charging resistor, *Rc*, and results in a smaller ratio of indicated to true r.m.s. reading so that the curves in Figs. 7 and 8 move up the appropriate amount on the vertical axis.

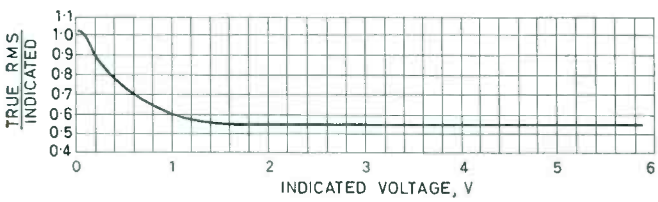


Fig. 7. Relationship between indicated and true r.m.s. values for TF 1041C Vacuum Tube Millivoltmeter with Gaussian noise

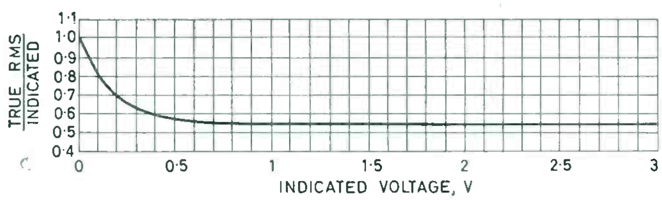


Fig. 8. Relationship between indicated and true r.m.s. values for TF 2603 Electronic Millivoltmeter with Gaussian noise

Very low frequency noise

An important factor in measuring noise of very low frequency (50 c/s and less) is the averaging or integrating time. The mean square value of a noise voltage is got by squaring the voltage and integrating over a period of time. There is of course a minimum time of observation which clearly increases the smaller the bandwidth of noise selected. If insufficient averaging time is allowed error results. It can be shown[3] that to determine within 95% certainty the mean square value, or the mean square deviation if the mean value is not zero, to within $P\%$ for a normally distributed noise having a flat spectral density over a band $\triangle f$, the integrating time, T, after square law detecting is:

$$T = \frac{4 \times 10^4}{\triangle f P^2} \text{ seconds} \quad \ldots\ldots\ldots\ldots (8)$$

$$\text{if } T \triangle f \gg 1$$

If the spectral density is not flat but frequency dependent, different integrating times are required[4]. For frequencies above 100 c/s or so the problem of integrating time vanishes for the commercial instruments mentioned.

Narrow band noise

If a narrow band of Gaussian noise is applied to an envelope detector the output waveform follows the peaks. If the input is large so that the detector is linear the spectrum of the output is no longer Gaussian but becomes Rayleigh (for low levels where the rectification is square law the output is of course still Gaussian). In this case the probability density function is different thus:—

$$p(v) = \frac{v}{\sigma^2} \exp(-v^2/2\sigma^2) \quad \ldots\ldots\ldots\ldots (9)$$

The peak factor/probability curve will consequently be different than that for Gaussian noise. Substituting (9) into (4) we can derive this distribution and Table 2 lists some typical peak factors.

Table 2

Peak factor for Rayleigh noise

% of time peak factor is exceeded	Peak factor $= \frac{v}{\sigma}$	$20 \log_{10} \frac{v}{\sigma}$ dB
1	3·04	9·6
0·1	3·62	11·4
0·01	4·29	12·6
0·001	4·8	13·6
0·0001	5·2	14·3

Conclusion

To measure Gaussian noise it is not necessary to insist on a true r.m.s. voltmeter for quantitative results. Provided bandwidth and peak factor criteria are observed the results obtained with an average responding voltmeter are a fixed ratio of the true r.m.s. value. Insufficient peak factor rating can be overcome by using less than full scale deflection or taken into account by reference to an error curve. For peak factors above 4, errors are usually insignificant anyway. Because of the random amplitude distribution it is sometimes thought that peak meters are suitable only for comparison of noise voltages. This is not the case when information is available about the meter's charge-to-discharge resistance. Better still a graph relating the correction factor for converting indicated to true r.m.s. value allows absolute measurements. Other advantages of peak detectors are the superior frequency response compared to the normal true r.m.s. voltmeter, the facility of true r.m.s. reading at low voltages directly, and the ability of the peak detector to detect asymmetry in a noise waveform by measuring separately the negative and positive parts of the noise.

REFERENCES

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621.317.799:621.396.44.

MARCONI INSTRUMENTS

12 CHANNEL White Noise Generator . . . Type TM 7816

by
W. OLIVER, M.I.E.E.E.,
and
A. J. SPENCER, B.Sc.

An instrument is described which extends present white noise loading techniques to include multiplex equipment. Methods of testing multiplex only or complete low capacity microwave link systems are discussed. The spurious noise measured may be expressed in terms of picowatts, dBa, dBrnC or channel signal-to-noise ratio. The effects of crosstalk and intermodulation distortion in multiplex equipment may be measured separately or together.

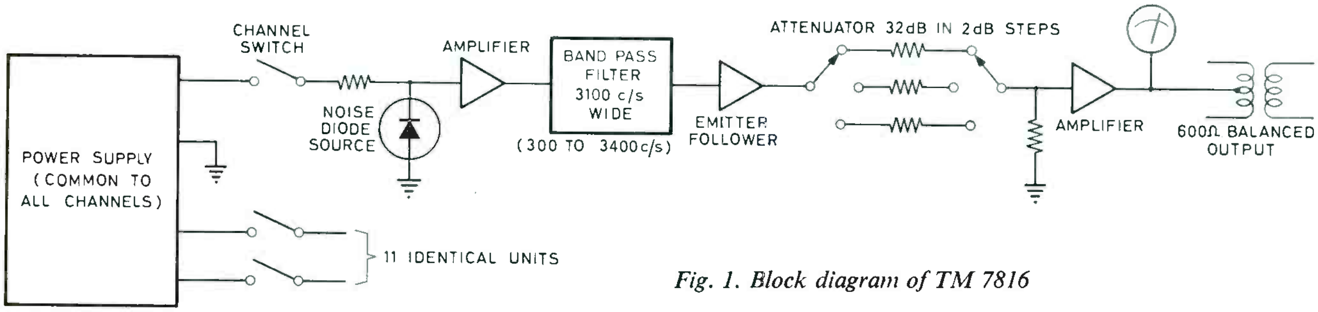


Fig. 1. Block diagram of TM 7816

NOISE GENERATOR type TM 7816 provides white noise signals for testing multi-channel telephone link equipment. To simulate twelve individual telephone circuits an independent generator with a bandwidth of 300 to 3400 c/s is used for each channel. The amplitudes of the noise signals are controlled simultaneously over a range of 32 dB in 2 dB steps using ganged attenuators. Maximum r.m.s. noise output per channel is 0 dBm into a balanced 600 Ω load. A preset gain control associated with each channel is used in conjunction with the noise level meter to equalize the amplitudes of the channels at 0 dBm. The meter is calibrated over a ± 1 dB range to provide means of monitoring the noise power output from 0 to —33 dBm per channel in 1 dB increments.

When used as a test set for small capacity links, e.g. 12 or 24 channel tropospheric scatter systems, Noise Generator type TM 7816 provides a complete test of the overall system whereas the more conventional noise loading test sets[1] measure from baseband to baseband only, leaving the multiplex equipment as untried components. For small capacity systems, the new method has the following advantages:

1. Any combination of channels can be loaded for detailed tests of intermodulation and crosstalk.
2. Complete system checking is accomplished from multiplex input to demultiplex output.
3. When a quiet channel is inserted, a known amount of loading is removed from the system, exactly simulating an unused channel.
4. Each channel can be measured separately.
5. The only auxiliary equipment needed is a sensitive voltmeter, ideally a psophometer, since the output of the measured channel is either white noise, when the channel is loaded, or spurious noise.

Applications

Basically any multi-channel system of up to 24 channels can be tested completely. Above this number, three or more 12 channel noise generators would be required, which is uneconomical, and it is suggested that group-by-group tests are made.

Specifically tropospheric scatter, i.s.b. or s.s.b. multi-channel equipments and multiplex equipment of all kinds can be tested. Since the input is at audio frequency, variations of baseband, i.f. or radiated frequency are immaterial.

A special feature of this instrument is its ability to check multiplex equipment. Both crosstalk and intermodulation can be measured, together or separately, and the effects of changes in loading determined over the entire dynamic range of the multiplex equipment.

Method of measurement

Since, in nearly every case, all tests are made through the multiplex equipment (whether or not they are connected by a transmission path or merely wired back-to-back) this type of test will be fully discussed. Although this instrument is not restricted to this test, the general principles apply even where there are deviations to suit individual circumstances.

Several of the telephone input points of the multiplex equipment are loaded with noise from the generator and the output connected back through the receiver demultiplex—see Fig. 2. A measurement is made of the resultant noise in a channel whose input is correctly terminated but which has no injected noise. Noise which appears in this quiet channel is due to idle channel noise plus contributions of crosstalk from other channels and intermodulation distortion in the system.

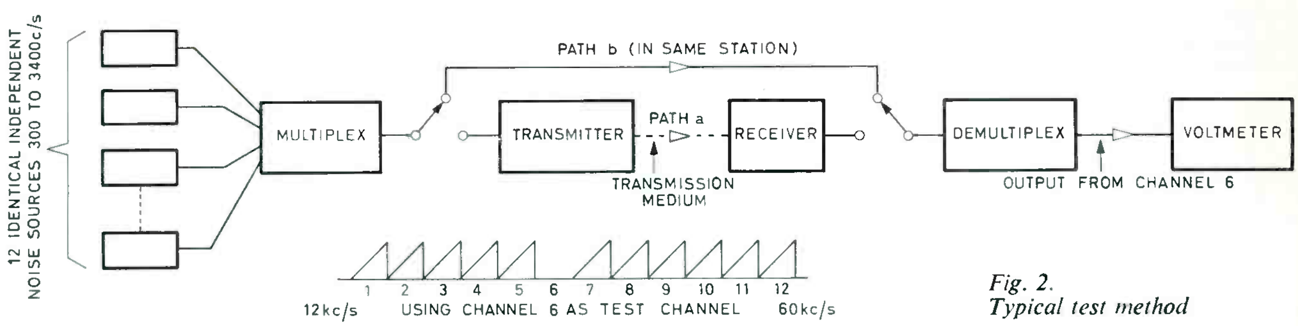


Fig. 2. Typical test method

(*a*) *Crosstalk*
Crosstalk is defined as spurious signals caused by coupling between circuits which should, in theory, be isolated. In multiplex equipment this coupling is usually the result of insufficient filtering. It can be measured by loading two adjacent channels only of a twelve channel group. Under these circumstances the total group loading will be much lower than normal and noise in the quiet channel will be mainly due to crosstalk.

(*b*) *Intermodulation*
Intermodulation distortion causes components from all channels to contribute to the noise in the idle channel. It can be separated from crosstalk by loading all channels except the quiet channel and its two adjacent channels. As discussed, crosstalk is contributed mainly by the adjacent channels and since, in a normal multiplex system, crosstalk is considerably greater than intermodulation for levels less than overload, it is necessary to avoid loading the adjacent channels to provide a guard band.
As the input level is increased the noise in the quiet channel increases fairly linearly until the overload point is reached and then a rapid increase takes place—see Fig. 3. By switching in more, or fewer, channels on the generator, measurements can be made using different activity coefficients.

Results

A typical test of a multiplex system back-to-back, with all channels except the test channel noise loaded, shows the three regions of Fig. 3:

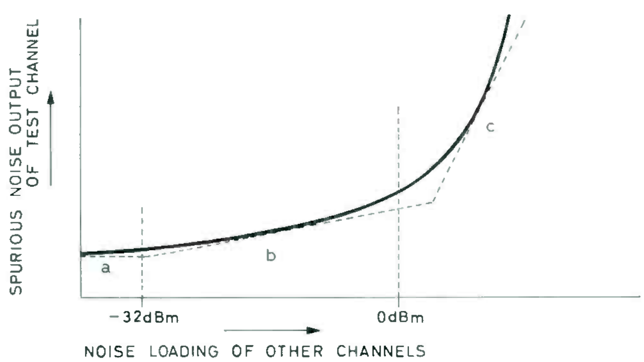


Fig. 3. Typical curve of spurious noise output with increasing loading on other channels

(*a*) Idle noise (no load), due to thermal noise and coupling from multiplex pilot tones.
(*b*) Light loading. Noise level increasing at low rate as per-channel loading is increased. Caused mainly by crosstalk from those channels immediately adjacent to the quiet channel.
(*c*) Heavy loading. Harmonic components and intermodulation distortion due to heavy per-channel loading are reflected in the steeper slope of the curve. Intermodulation distortion in the quiet channel can be produced by overloading any of the channels, unlike crosstalk which is usually traceable to the adjacent channels.

Note on channel loading levels

The measurements mentioned previously are intended to investigate the performance of the system over a wide signal level range. However, statistical studies have been made to measure the r.m.s. signal power developed by the average telephone or group of telephones[2]. From these C.C.I.R. and C.C.I.T.T. have proposed two formulas; these give the noise power required to simulate the complex speech power for a telephone circuit carrying N channels.

These formulas are used directly when testing baseband-to-baseband with white noise[1]. They are modified as shown below when measurements including multiplex are to be made. The channel noise loading powers that result simulate the equivalent channel speech power and hence provide a controlled accurate loading of the entire system.

(*a*) *Systems below 240 channels*

The C.C.I.R. adaptation of the Holbrook-Dixon[2] curves gives a loading of

$NLR = -1 + 4 \log_{10} N$ dBm0

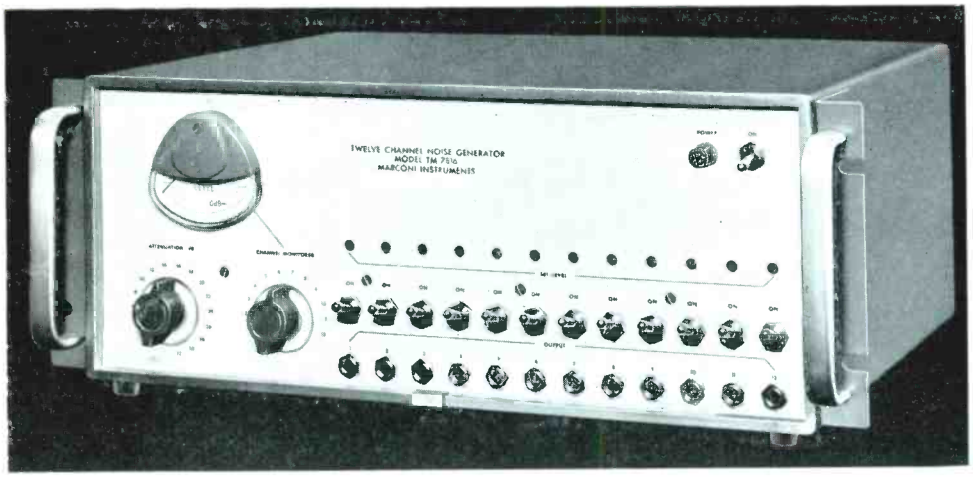
where N is the number of channels
NLR is noise loading ratio for baseband level,
and dBm0 is dB relative to test tone level.

The per-channel loading will therefore be

$-1 + 4 \log_{10} N - 10 \log_{10} N = -1 - 6 \log_{10} N$ dBm0

It can be seen that the per-channel loading will vary according to the number of channels. As examples, consider the two cases most commonly encountered when using this generator, i.e. 12 and 24 channels.

For 12 channels the per-channel loading will be
$-1 - 6 \log_{10} 12 = -1 - 6 \times 1{\cdot}0792 \simeq -7{\cdot}5$ dBm0



The 12 Channel White Noise Generator Type TM 7816

For 24 channels the per-channel loading will be
$-1 - 6 \log_{10} 24 = -1 - 6 \times 1{\cdot}3802 \simeq -9{\cdot}3$ dBm0

(*b*) *Systems above 240 channels*

The C.C.I.R. agreed formula is

$NLR = -15 + 10 \log_{10} N$ dBm0

which gives a per-channel loading of -15 dBm0 independent of the number of channels.

(*c*) *Data loading*

A formula (non-C.C.I.R.) based on 100% loading is used. This is

$NLR = -5 + 10 \log N$ dBm0

which gives a per-channel loading of -5 dBm0 independent of the number of channels.

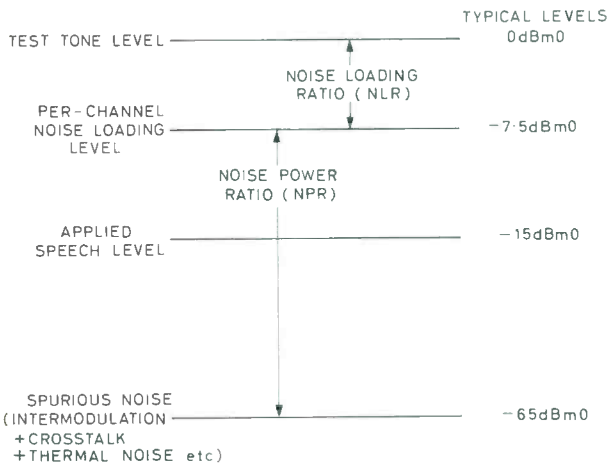


Fig. 4. Level diagram for a typical channel of a 12 channel group[3]

Signal-to-noise ratio

It can be seen from Fig. 4 that the ratio of test tone level to spurious noise is given by the formula $S/N = NPR + NLR$. Where the noise power ratio (determined using TM 7816 to measure the ratio between noise in a channel with all channels ON to noise in a quiet channel with all other channels ON) is, say, 45 dB and the per-channel loading as determined from $-1 - 6 \log N$ is, say, $-7{\cdot}5$ dBm0 (a 12 channel system C.C.I.R. loaded) then the signal-to-noise ratio equals $45 + 7{\cdot}5 = 52{\cdot}5$ dB.

However, since the speech level is usually applied 10 to 16 dB below test tone level, the actual speech signal-to-noise ratio will be that amount less.

Conversion of signal-to-noise ratio to channel noise

The channel noise can be measured directly with a psophometer or calculated as follows when the measured NPR of the system has been converted to channel signal-to-noise ratio.

1. *dBa0 noise units*

The term dBa[4] refers to F1A weighted noise measurement above a reference noise level of -85 dBm, i.e. 0dBa $= -85$ dBm (at 1 kc/s). But white noise when applied to an F1A weighting filter is attenuated by 3 dB.

0dBa $= -85 + 3 = -82$ dBm (white noise applied)

$\therefore$ to convert S/N ratio to dBa0

dBa0 $= 82 - (S/N)$

2. *dBrnc0 noise units*

The term dBrnc refers to 'C' message weighted noise measurements above a reference noise level of -90 dBm, i.e. 0 dBrnc $= -90$ dBm (at 1 kc/s). White noise applied to a 'C' message weighting filter is attenuated by 1·5 dB.

0 dBrnc $= -88{\cdot}5$ dBm (white noise applied)

To convert S/N ratio to dBrnc0

dBrnc0 $= 88{\cdot}5 - (S/N)$

3. *Picowatts of noise power*

1 pW $= 10^{-12}$ watts *or* 0 dBm $= 10^9$ pW

Picowatts of noise power are usually measured with psophometric weighting. When white noise is applied to a C.C.I.R. psophometric weighting filter it is attenuated by 2·5 dB (almost the same as F1A weighting).

Assuming 0 dBm0 = 0 dBm at the point of measurement, to convert S/N ratio to picowatts,

$$pW = \frac{10^9}{\text{antilog}_{10}\, S/N} \quad \text{unweighted}$$

$$\text{or} \quad \frac{10^9 \times 0{\cdot}56}{\text{antilog}_{10}\, S/N} \quad \text{weighted}$$

If 0 dBm0 is not equal to 0 dBm, the noise power should be corrected to suit the difference in level.

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621.317.75

MARCONI INSTRUMENTS

Oscilloscope . . . . Type TF 2200/6M1

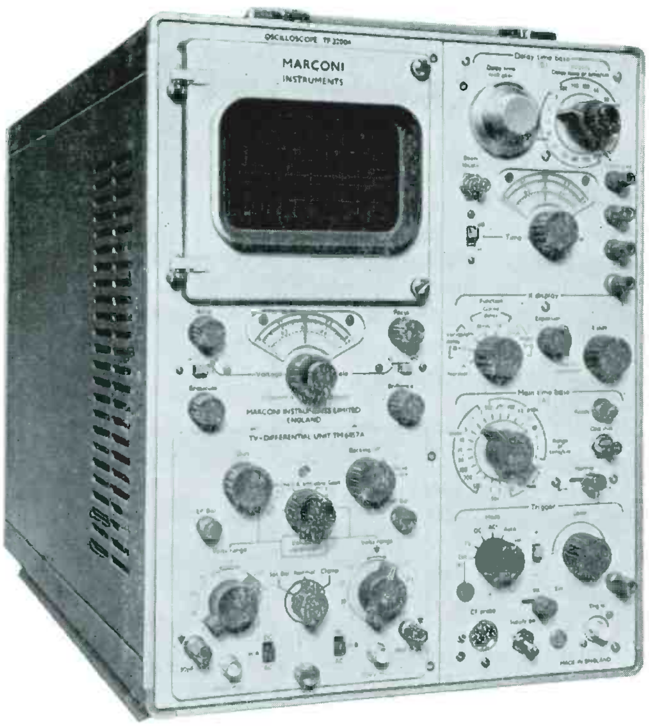
FOR RADAR LINK TESTING

by
M. W. G. HALL,
A.M.I.E.R.E.

In radar data link equipment linearity can be measured by derivative techniques. The associated display device must have a d.c. coupled X channel and high gain and windowing capability on the Y axis. A specially modified version of Oscilloscope type TF 2200 is now available to meet these requirements.

MODERN air traffic control systems rely on the integration of a chain of radar stations. To achieve this result the radar heads may be sited where topographically most suitable, and the radar information (video, trigger, aerial bearing, etc.) channelled on to the baseband of a microwave radio link for transmission to a central control centre.

On some radar link systems it is necessary to transmit angle information in analogue form with a high degree of accuracy. One method of doing this is to frequency modulate the information on the channel sub-carrier, using a high modulation index for signal to noise improvement; special techniques are employed to ensure an adequate channel linearity. With channels of this type, it is necessary to provide a convenient method of checking linearity. A derivative method is employed for this, in which a high amplitude, low frequency sweep signal



Marconi Instruments Oscilloscope Type TF 2200A

Prototype of a radar heightfinder supplied to Kuwait. It has been sand-proofed and tropicalized

(The Marconi Company Limited)

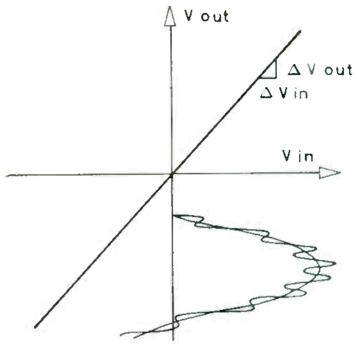
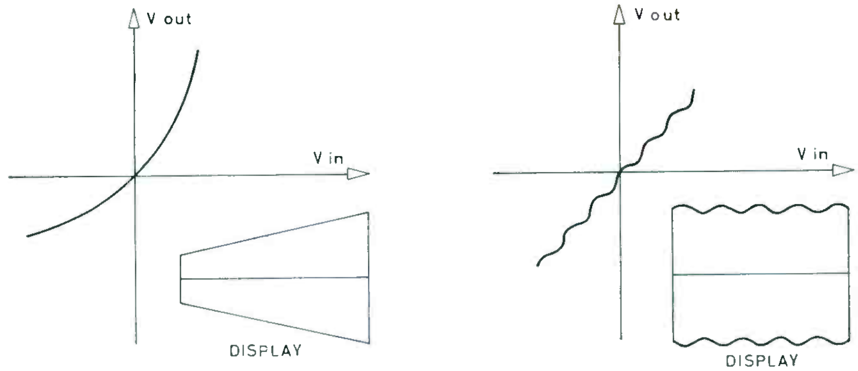


Fig. 1. Test signal applied to ideal linear transfer function



(*a*) (*b*)
Fig. 2. Effect of non-linear transfer functions

together with a low amplitude, high frequency search signal are transmitted over the channel. At the output, the two signals are separated by means of filters; the sweep signal is used to produce an X deflection on the display device and the search signal the Y deflection.

Fig. 1 shows the channel transfer function in the ideal (linear) case; Figs. 2a and 2b show examples of non-linear characteristics. The variation of the first derivative of the characteristic, $dVout/dVin$, relative to Vin is a measure of the linearity of the channel. The envelope of the display obtained by the test method described displays $\triangle$ *Vout*, which is proportional to $dVout/dVin$ if $\triangle$ *Vin* is small and constant, as a function of *Vout*. $\triangle$ *Vin* is the amplitude of the search signal.

The advantage of the method is that the displayed signal is of constant amplitude if the channel is linear, thus eliminating display linearity problems and enabling high Y gain to be employed. This means that, if sufficient display offset is available, only the top of the displayed envelope need be viewed critically. It will be seen that the display device requires to have high Y gain with exceptional windowing capability to allow accurate measurements to be made, together with a d.c. coupled X input to enable low frequency sweep rates to be employed.

Display unit

A specialized display device could have been designed for this application but it was simpler to use a specially modified version of a normal high-grade measuring oscilloscope.

In choosing the TF 2200 Oscilloscope, the following modifications were necessary.

1. The external drive facility of the X amplifier was re-arranged to be d.c. coupled and fitted with an input level control.
   To avoid congestion of the instrument front panel the additional control, labelled X INPUT LEVEL, was mounted coaxially with the existing X SHIFT control.
2. The Y shift control of the single trace plug-in unit is re-arranged so that when switched to the ×5 position of the gain control, where input sensitivity is 10mV/cm, it is still possible to window a signal of 4V peak-to-peak.
   The normal production version of the single trace plug-in unit type TM 6455 has the requisite sensitivity, namely 10mV/cm, but the maximum coverage of the Y shift control under these conditions is approximately 0·4 V. To increase this coverage the series resistance in the shift circuitry was reduced and the values of the resistances in the potentiometer chain which supplies the voltage were reduced by a factor of 10:1 which prevents the accuracy of voltage measurement via the calibrated shift control from being worse than ±5% of full scale. To make the highly sensitive shift control more manageable the normal potentiometer was replaced by a 5-turn helical control.

These simple modifications permit a general purpose measuring oscilloscope to be used in a specialized role without sacrificing any of its other facilities.

Acknowledgements

The author wishes to acknowledge the assistance of Mr. P. N. Sargeaunt of the Marconi Company Limited, Great Baddow Laboratories, in the preparation of this article.

REFERENCE

1. Middleton, J. D., 'Oscilloscope type TF 2200', *Marconi Instrumentation*, December 1962, 8, p. 171.

MARCONI INSTRUMENTS ACQUIRE MICROWAVE COMPANY

W. H. Sanders (Electronics) Ltd., have joined the English Electric Group as a subsidiary of Marconi Instruments Ltd. Their design and manufacturing experience in microwave equipment, waveguide components and industrial control gear adds an important sector to the Group's electronic output. Although Sanders will continue at present as a self-contained unit with its existing range of products, new and advanced microwave test gear will be gradually introduced to supplement the Marconi Instruments range. Mr R. E. Burnett, Managing Director of Marconi Instruments has been appointed Chairman of W. H. Sanders.

621.317.757

MARCONI INSTRUMENTS

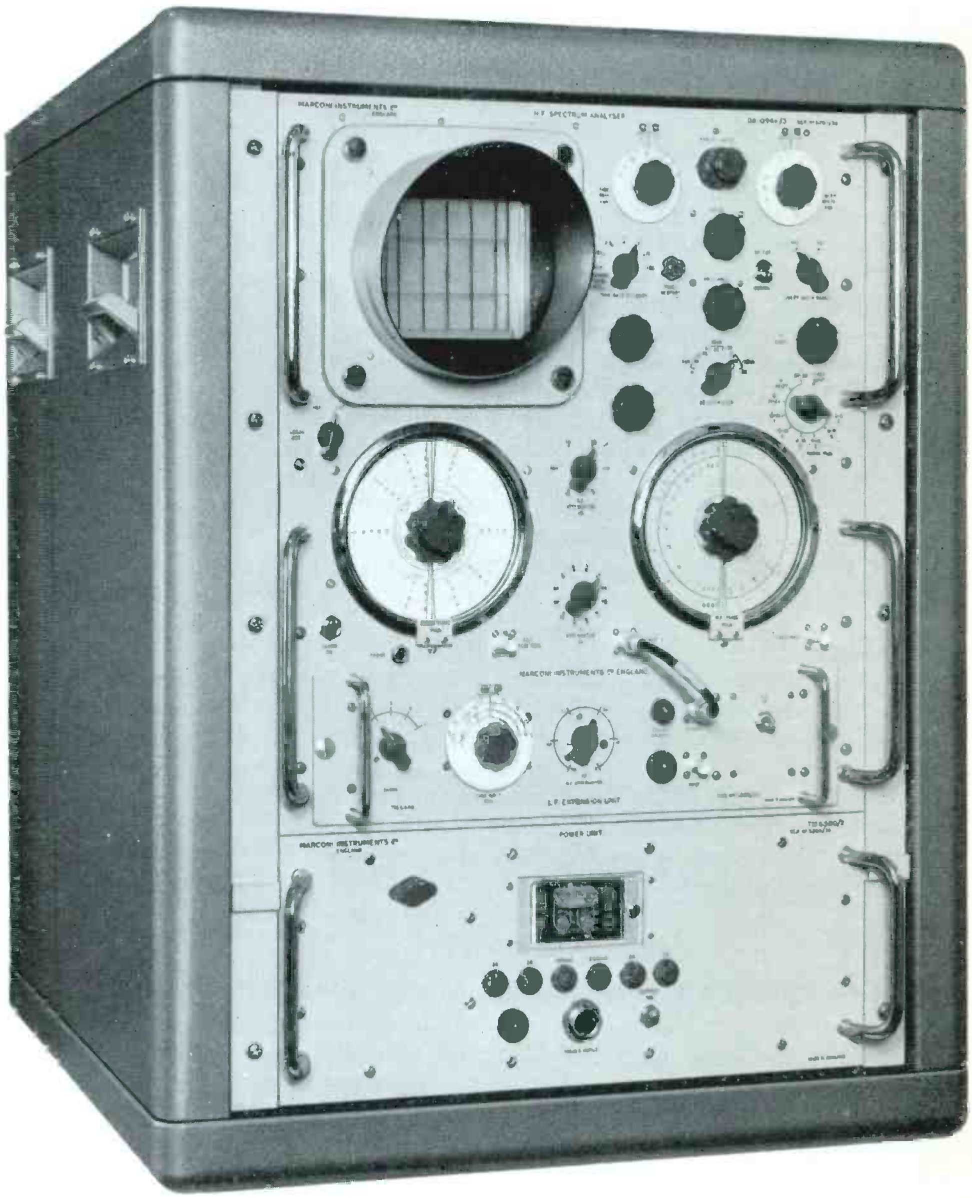
Narrow band spectrum analysis

and the

OA 1094A/3 SPECTRUM ANALYSER

by
M. J. TANT,
B.Sc., A.C.G.I.,
Graduate I.E.E.

Analysis of radio-frequency spectra over narrow bands is an exacting requirement for a spectrum analyser. Eminently suited to this purpose is the type OA 1094A/3 H.F. Spectrum Analyser which has improved performance in this mode, whilst retaining all the functions of its predecessor, the OA 1094A. The additional facilities of the new instrument are described and mention is made of certain phenomena which may degrade narrow band spectral displays.



The Marconi Instruments improved Spectrum Analyser Type OA 1094A/3

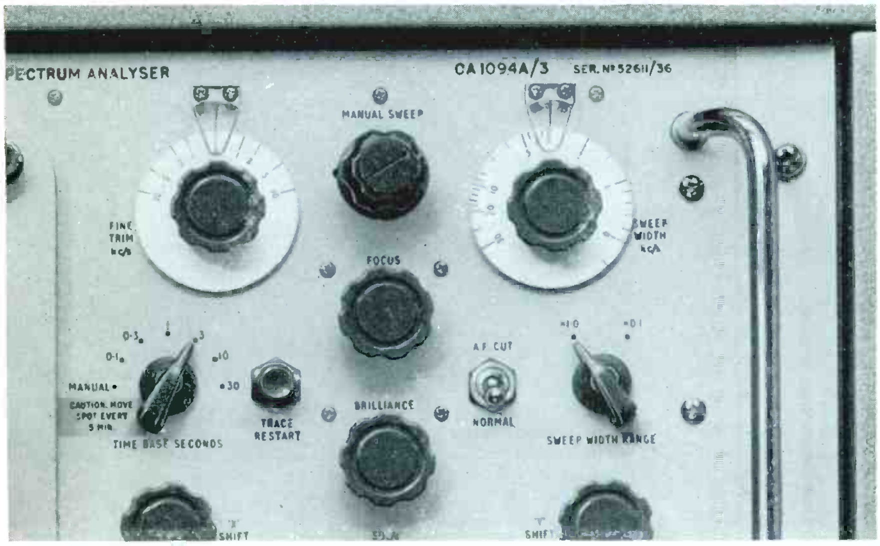


Fig. 1
Sweep controls
of the OA 1094A/3

H.F. SPECTRUM ANALYSER type OA 1094A/3 replaces the OA 1094A which has been fully described in a previous issue of *Instrumentation*[1]. The new instrument is basically similar to its predecessor, being essentially a triple superheterodyne receiver with the final stage frequency swept by the X deflection waveform. There are two facilities additional to those of the OA 1094A, namely, normal or reduced Y amplifier bandwidth and manual sweep. Relevant controls can be seen in Fig. 1, which shows the upper right hand section of the front panel. Y amplifier bandwidth may be reduced with the NORMAL/A.F. CUT switch. Manual sweep may be selected by the TIME BASE selector, the spot being set to any position across the c.r.t. face by means of the concentric coarse and fine controls. In addition to these new functions, the inherent hum sidebands have been further reduced by modifications to the power supply.

These changes are described below in more detail and the improvement in narrow bandwidth displays is shown. In conclusion, the phenomenon of ringing is described, and its importance in narrow band measurement discussed.

Mode of operation

A simplified block diagram of the spectrum analyser is shown in Fig. 2. An r.f. signal in the range 3 Mc/s to 30 Mc/s passes via the r.f. attenuator and a tunable band

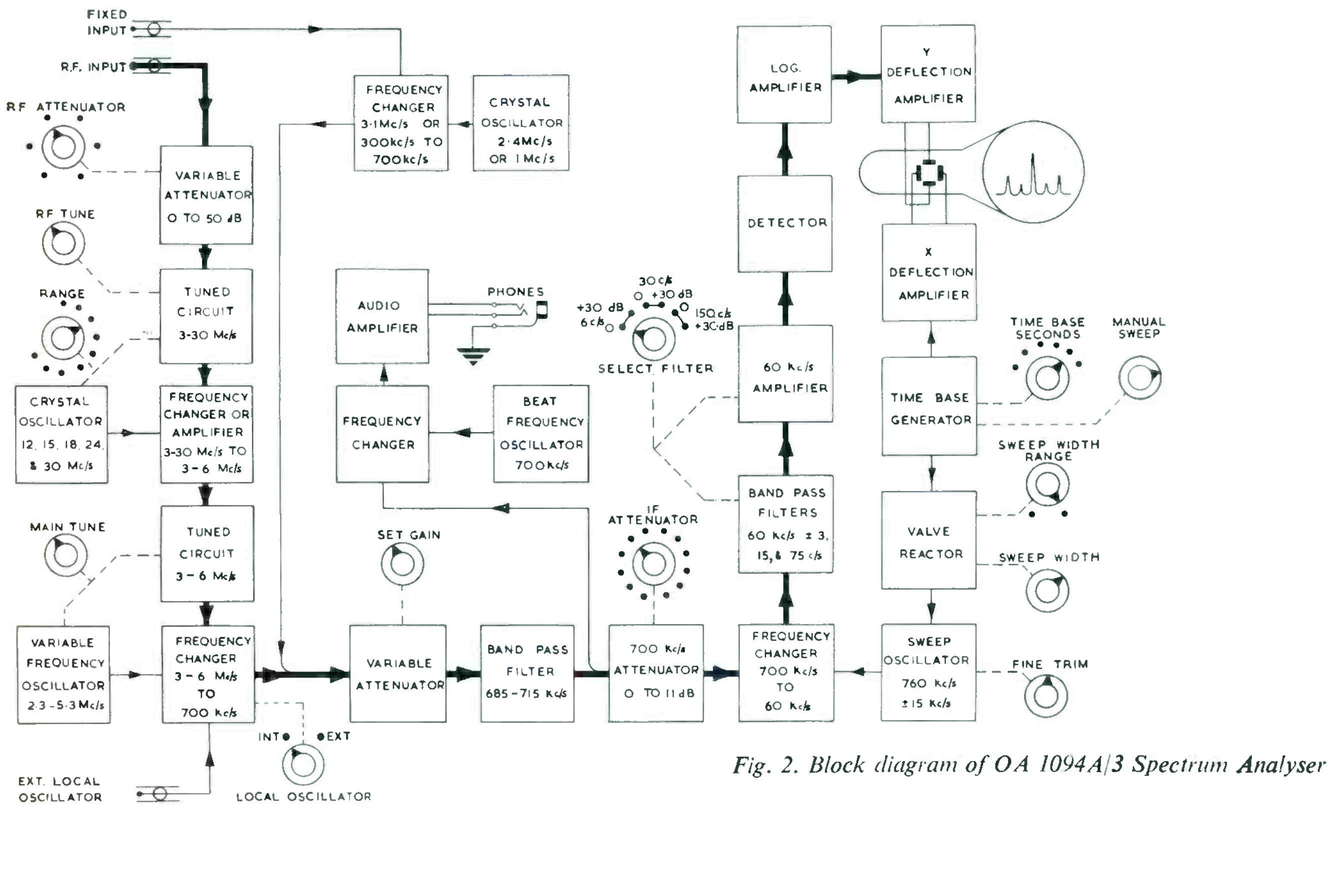


Fig. 2. Block diagram of OA 1094A/3 Spectrum Analyser

pass filter (R.F. TUNE) to the first frequency changer, where it is mixed with a crystal controlled local oscillator and translated to the range 3 Mc/s to 6 Mc/s. After filtering out the unwanted frequency components, the signal is mixed with a variable local oscillator (MAIN TUNE) in the second frequency changer, to give the second i.f. of 700 kc/s. After the SET GAIN control the 700 kc/s filter attenuates all signals outside its pass band of 700 $\pm$15 kc/s and the signal then passes, via the i.f. attenuator, to the third frequency changer.

The sawtooth time base waveform applied to the X plates of the c.r.t. via the X amplifier is also used to drive a variable reactor which sweeps the third local oscillator to a maximum of $\pm$15 kc/s about its centre frequency of 760 kc/s. At various points on the time base sweep, components in the spectrum of the input signal (now in the range 685 kc/s to 715 kc/s) will combine with the swept local oscillator to give a third i.f. of 60 kc/s. This 60 kc/s signal passes through one of the three alternative band pass crystal filters, is detected, amplified logarithmically, and applied to the Y plates. The input signal frequency spectrum is thus displayed on the c.r.t. with the X axis linear in frequency and the Y axis calibrated in decibel amplitude ratio.

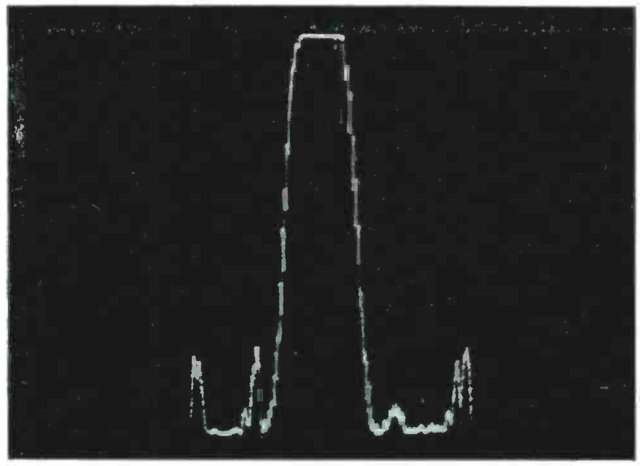
A.F. cut facility

The 60 kc/s filter has switched 3 dB bandwidths of 150 c/s, 30 c/s or 6 c/s and the slopes of the filters are of the order of $\frac{12}{B}$ dB per c/s over the 60 dB measurement range of the instrument, where B is the 3 dB static bandwidth of the filter. For narrow band displays where components are only tens of c/s apart, the 6 c/s filter is required to allow adequate resolution of adjacent frequency terms.

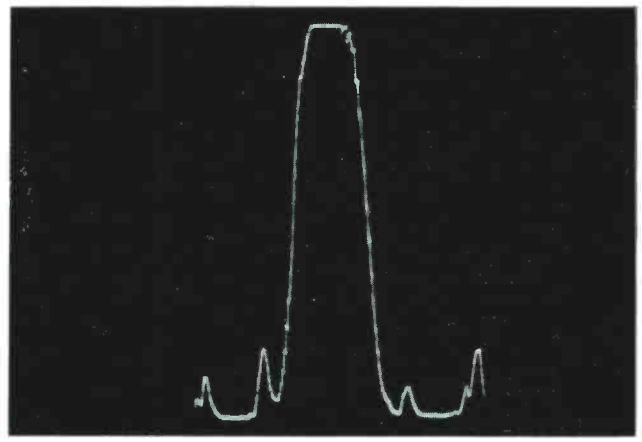
The 6 c/s filter has an average slope of about 2 dB per c/s and the analyser has two 'non crystal locked' local oscillators which are naturally susceptible to a certain amount of random f.m. As an example, consider random f.m. of only $\pm$5 parts in 10^7. This results in random frequency shift of $\pm$1·5 c/s in the second frequency changer (when set to 3 Mc/s) and $\pm$0·4 c/s in the swept oscillator. Thus the displayed signal is subject to random shifts of $\pm$2 c/s and the trace may fluctuate up to $\pm$4 dB on the skirts of the filter. The same calculation for the 30 c/s and 150 c/s filter would give $\pm$0·8 dB and $\pm$0·16 dB fluctuation respectively, so the worst degradation occurs with the 6 c/s filter which is necessary for narrow bandwidth displays.

Fig. 3 (*a*) shows the 50 c/s and 100 c/s hum sidebands of an s.s.b. transmitter with single tone modulation and with carrier suppressed. Due to random modulation of the second local oscillator, there is an amplitude uncertainty of perhaps 2 dB. The transmitter frequency of 6·3 Mc/s is equivalent to 5 Mc/s for the second local oscillator, so this type of display can result from frequency shifts of only 2 parts in 10^7.

It is, however, possible to integrate these fluctuations, without degrading the required display, by reducing the bandwidth of the Y amplifier which is the purpose of the



(*a*) *Normal*

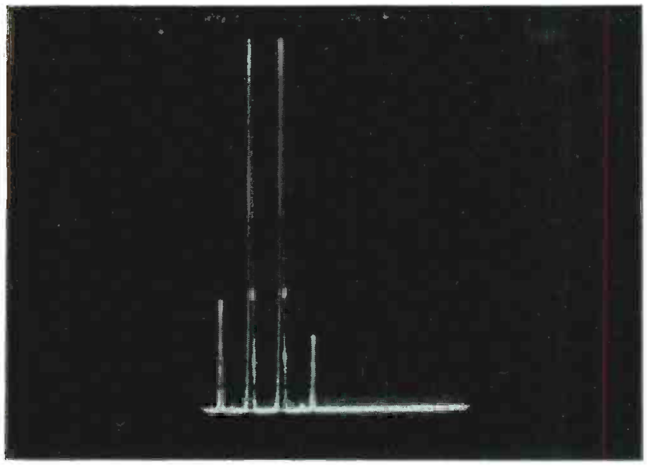


(*b*) *A.F. cut*

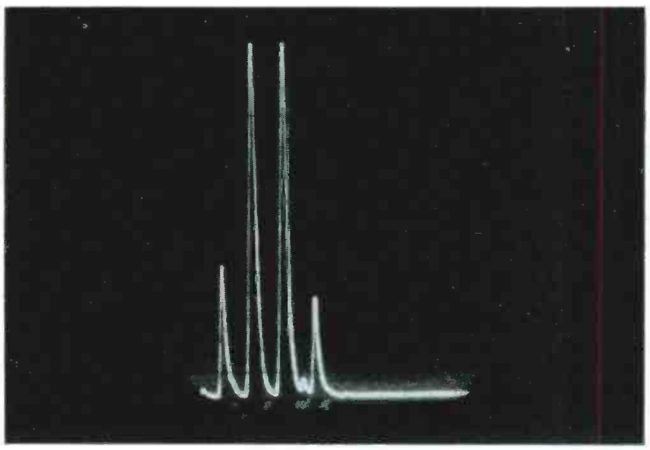
Fig. 3. Spectrum of s.s.b. signal with hum sidebands at $\pm$50 c/s and $\pm$100 c/s

A.F. CUT position of the NORMAL/A.F. CUT switch. Fig. 3 (*b*) shows the display of Fig. 3 (*a*) but with the a.f. cut facility used. The improvement is sufficient to allow accurate measurement of the hum sidebands and the 2 dB uncertainty of Fig. 3 (*a*) has been largely removed.

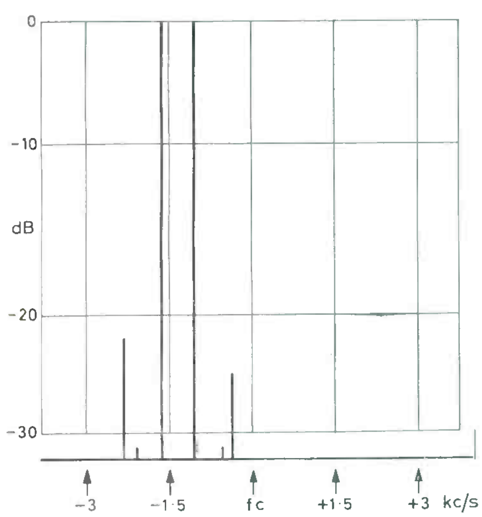
When wider sweep widths are required the 30 c/s or 150 c/s filters are used. The amplitude fluctuations are reduced 5 times and 25 times respectively and the A.F. CUT facility is not required. With these wider sweeps, the required display has a wider bandwidth and the NORMAL position should be used. Reducing the Y amplifier bandwidth will attenuate the higher frequency components and degrade the display. Fig. 4 shows the output of an s.s.b. transmitter, two tone modulated. The carrier is suppressed and coincident with the centre vertical graticule. Loss of detail is clearly shown in Fig. 4 (*b*), the a.f. cut condition, where the increased time constant tends to mask detail.



(a) Normal

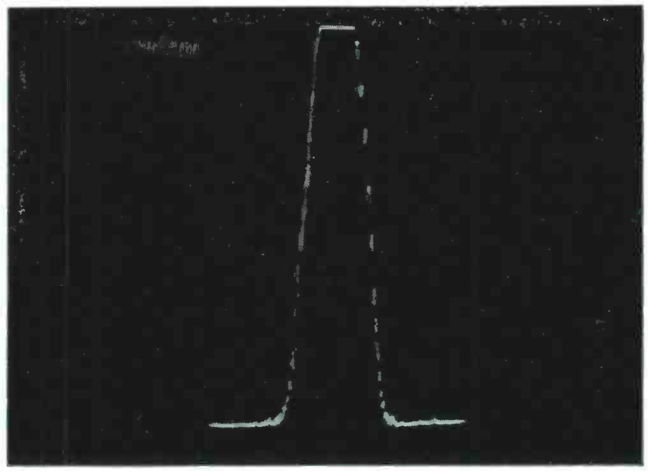


(b) A.F. cut

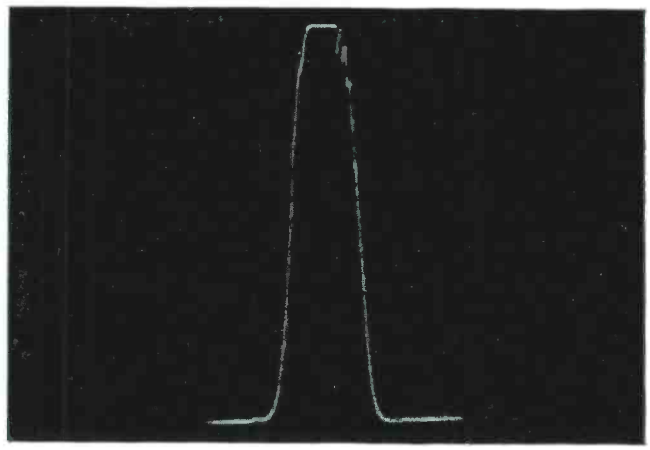


(c) Idealized

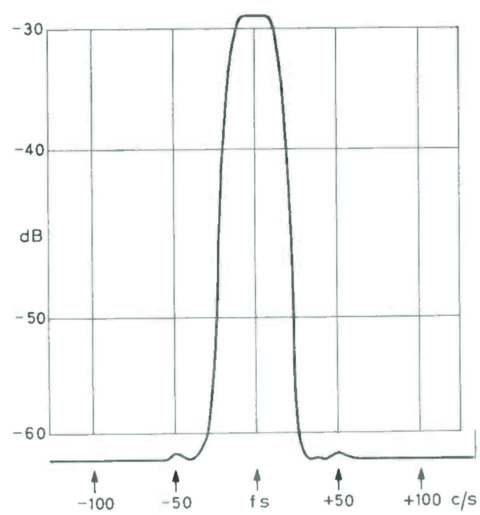
Fig. 4. Spectrum of two-tone s.s.b. signal with intermodulation components.



(a) Normal



(b) A.F. cut



(c) Idealized

Fig. 5. Spectrum of 5 Mc/s oscillator signal

Manual sweep

As previously mentioned, the effect of small random frequency shifts in the local oscillators is to degrade the display when the 6 c/s filter is used, because of its high slope as expressed in dB per c/s. However, the top of the filter has zero slope, and any frequency fluctuations will have far less effect on this portion than on the skirts. The manual sweep facility allows the time base to be stopped and the spot manually located on the top of the desired frequency component. Relative amplitude measurements may then be made with any inaccuracies due to random frequency shifts much reduced.

With the TIME BASE selector set to MANUAL, a potentiometer network is substituted for the time base generator and this provides both the X shift on the c.r.t. and frequency shift to the third local oscillator via the reactor. Two concentric potentiometers are included in the network; the coarse control allows shift across the entire width of the tube face, and the fine control gives approximately $\pm 5\%$ coverage of the coarse control.

Hum sidebands

Fig. 3 shows the 50 c/s and 100c/s hum sidebands of an s.s.b. transmitter, but any measurements made on their relative amplitudes are not very meaningful unless the hum sidebands of the spectrum analyser are negligible in comparison. For accurate hum sideband measurement over the full amplitude range of the instrument, the inherent sidebands should be better than 60 dB down.

The predominant cause of inherent hum sidebands in the OA 1094A was mechanical vibration caused by the smoothing choke in the power supply. This choke, together with the two mains transformers is now mounted on a sub-chassis with anti-vibration mountings and adequately screened.

The overall effect upon hum is shown in Fig. 5, for the spectrum of a 5 Mc/s transistorized oscillator. In Fig. 5 (*a*) the hum sidebands cannot be distinguished from the noise, but in the A.F. CUT mode, they can just be detected. From the idealized diagram it can be seen that the inherent hum sidebands are much better than 60 dB down, so allowing the full amplitude range of the instrument to be used to within 50 c/s separation of adjacent terms.

Ringing

It is a well known phenomenon that a single tone swept sinusoidally in frequency (frequency modulation) will exhibit a spectrum consisting of frequency terms spaced at intervals equal to the modulating frequency. In the spectrum analyser the signal is frequency swept past a crystal filter by a sawtooth modulating signal. The detected output of the filter is then displayed. This is a frequency modulation process so it is to be expected that the displayed dynamic filter characteristic will show distortion due to the application of an f.m. spectrum instead of the single tone which is assumed for the static characteristic. This distortion may be expected to worsen as the filter bandwidth is decreased, the sweep width increased or the time base speed increased.

For the single tone swept past a Gaussian filter of 3 dB bandwidth B, a sweep width of W c/s and a time base of t secs, the following relationships apply:—

$$V_D = V_S \left[1 + 0{\cdot}195 \left(\frac{W}{tB^2}\right)^2\right]^{-\frac{1}{4}} \qquad (1)$$

$$R = B\left[1 + 0{\cdot}195 \left(\frac{W}{tB^2}\right)^2\right]^{\frac{1}{2}} \qquad (2)$$

where V_D is the dynamic peak amplitude of the displayed term,
V_S is the static peak amplitude,
R is the resolution (dynamic 3 dB bandwidth) of the displayed term,
and B the static 3 dB bandwidth.

Thus we see that as the static bandwidth is decreased, the time base speed increased or the sweep width increased, the trace will show distortion in the form of reduced peak amplitude and broadening of the 3 dB bandwidth. In fact the trace also exhibits damped oscillation of the trailing edge and displacement to the right. These distortion phenomena are known as ringing[2].

The crystal filters in this spectrum analyser approximate to Gaussian form and distortion may be seen in precisely the manner dictated by equations (1) and (2).

The static 60 dB bandwidth of the 6 c/s filter is approximately 60 c/s, but at a value of $\frac{W}{tB^2}$ of 2, the 60 dB bandwidth has broadened to approximately 100 c/s. In this condition 50 c/s hum sidebands will be just inside the filter characteristic and their measurement will not be possible. The particular oscillograms shown here were all taken at $\frac{W}{tB^2} = 0{\cdot}8$. A value of $\frac{W}{tB^2}$ of less than 1·0 is generally suitable for narrow band measurements. The additional facilities of this spectrum analyser can only be used to their fullest advantage if attention is paid to selection of the time base and sweep width as mentioned above.

REFERENCES

1. Deichen, J. H., 'H.F. Spectrum Analyser type OA 1094A', *Marconi Instrumentation*, December 1961, **8**, p. 88.
2. Slater, F., 'Ringing in spectrum analysers', *Ibid.*, March 1958, **6**, p. 149.

SOMMARIO DEGLI ARTICOLI PUBBLICATI IN QUESTO NUMERO

MISURE DI RUMORE CON VOLTMETRI ELETTRONICI

Si esaminano i fattori che influenzano la misura di rumore Gaussiano mediante voltmetri elettronici che non rispondono ai vero valore efficace. Vengono discusse le attinenti caratteristiche del rumore Gaussiano e la relazione fra rumore e sinusoidi riguardo ai parametri che influenzano le indicazioni dello strumento. Conoscendo i rispettivi valori del fattore di forma per una sinusoide e per il rumore, si possono ottenere risultati quantitativi con voltmetri rispondenti al valore medio. Simili risultati possono essere ottenuti con strumenti che rispondono al valore di cresta purchè sia noto il rapporto tra i tempi di carica e di scarica del rivelatore di entrata. Nell'uso di voltmetri elettronici è inoltre importante conoscere il fattore di cresta e la larghezza di banda; per rumore di frequenze molto basse occorre anche considerare il tempo di integrazione.

Pagina 18

GENERATORE DI RUMORE BIANCO SU 12 CANALI TIPO TM 7816

Viene descritto uno strumento che estende i presenti metodi di caricamento con rumore bianco in modo da renderli applicabili ad apparecchiature multiplex. Si discutono procedimenti di prova del solo multiplex o di impianti completi di ponti radio di bassa capacità a microonde. Il rumore spurio misurato può essere espresso in picowatt, dBa, dBrnC o rapporto segnale/disturbo di canale. Gli effetti di diafonia e di distorsione di intermodulazione in apparati multiplex possono essere misurati separatamente o nel complesso.

Pagina 23

OSCILLOSCOPIO PER MISURE IN IMPIANTI DI COLLEGAMENTO PER RADAR TIPO TF 2200/6M1

In apparati di collegamento per la trasmissione di dati radar la linearità può essere misurata con metodi a derivata. I relativi dispositivi di presentazione visiva debbono avere un canale X ad accoppiamento diretto ed un elevato guadagno sull'asse Y con capacità di scelta della porzione della forma d'onda in senso verticale. E' ora disponibile una versione dell'oscilloscopio tipo TF 2200 appositamente modificata per fare fronte a queste esigenze.

Pagina 26

ANALISI SPETTRALE A BANDA STRETTA E L'ANALIZZATORE DI SPETTRO TIPO OA 1094A/3

L'analisi di spettri a radio frequenza su bande strette è un requisito gravoso per un analizzatore di spettro. Eminentemente adatto a tal fine è l'analizzatore di spettro per alta frequenza tipo OA 1094A/3 le cui prestazioni in tale modo di funzionamento sono state migliorate, pur conservando tutte le funzioni del predecessore, il tipo OA 1094A. Vengono descritte le aggiunte apportate al nuovo strumento e si fà menzione di alcuni fenomeni che possono deteriorare la presentazione visiva di spettri a banda stretta.

Pagina 28

RESUMENES DE ARTICULOS QUE APARECEN EN ESTE NUMERO

MEDIDAS DEL RUIDO CON VOLTIMETROS ELECTRONICOS

Se examinan los factores que afectan la medida del ruido gaussiano con voltímetros electrónicos, que no responden al valor medio verdadero. Se estudian las características esenciales del ruido gaussiano y la relación entre el ruido y sinusoides, en aquellos parámetros que afectan a la lectura de los medidores. Conociendo los valores relativos del factor de forma de una sinusoide y un ruido, se pueden obtener resultados cuantitativos con voltímetros de una respuesta media. También se consiguen resultados parecidos con voltímetros con respuesta de pico, siempre que se conozca la relación entre los tiempos de carga y descarga del detector de entrada. Cuando se utilizan voltímetros electrónicos es, además, importante conocer el factor de pico y la anchura de banda; en caso de ruido de muy baja frecuencia, debe también considerarse el tiempo de integración.

Página 18

GENERADOR DE RUIDO BLANCO CON 12 CANALES TM 7816

Se describe un equipo que amplía las técnicas de carga de ruido blanco existentes, para incluir los equipos multiplex. Se estudian los métodos de probar los equipos multiplex sólo, o los sistemas completos de enlace por microondas de baja capacidad; el ruido espúreo medido puede expresarse en pW, dBa, dBrnC, o como relación señal/ruido en el canal. Los efectos de diafonía y de distorsión de intermodulación en los equipos multiplex, pueden medirse por separado o en conjunto.

Página 23

OSCILOSCOPIO PARA PRUEBA DE ENLACES DE RADAR TF 2200/6M1

En los equipos de enlace para transmisión de datos por radar, puede medirse la linealidad por medio de técnicas de derivación. El dispositivo asociado de presentación debe tener un canal X acoplado a c.c., alta ganancia y capacidad de pantalla en el eje Y. Existe una versión especial modificada del osciloscopio tipo TF 2200, que cumple estos requisitos.

Página 26

EL ANALISIS DEL ESPECTRO EN BANDA ESTRECHA Y EL ANALIZADOR DE ESPECTRO OA 1094A/3

El análisis del espectro en radiofrecuencia, en bandas estrechas, es una condición esencial de los analizadores de espectro; cumple especialmente esta condición el analizador de espectro en a.f., OA 1094A/3, que trabaja satisfactoriamente para estas medidas, conservando al mismo tiempo todas las posibilidades de su predecesor, el OA 1094A. Se describen las nuevas aplicaciones de esta moderna versión, haciendo notar ciertos fenómenos que pueden degradar las presentaciones espectrales en banda estrecha.

Página 28

SIGNAL GENERATORS

SWEEP GENERATORS

FREQUENCY COUNTERS
AND METERS

OSCILLOSCOPES

NOISE GENERATORS

OSCILLATORS

PULSE GENERATORS

MODULATION METERS

DISTORTION ANALYSERS

SPECTRUM ANALYSERS

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