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MARCONI INSTRUMENTATION

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MARCONI INSTRUMENTATION

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MARCONI INSTRUMENTS LIMITED

ST. ALBANS

ENGLAND

EDITORS

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

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A Volume of Progress

IN THE COMMUNICATIONS INDUSTRY reflections are definitely undesirable because of the various forms of distortion they generate. However, as this issue completes the two year cycle of another volume, some reflections will be deliberately introduced, it is hoped without distortion, to review electronic instrument progress during the preceding and current year—a period in which, on an average, each month produced a new design.

In anticipation of things to come, page one aptly set the scene for what proved to be the main theme of this volume—namely transistorization. As can be seen, with few exceptions, the new instruments featured in this journal have utilized the current solid state techniques. It was realised at the outset that the decision to go solid state would have repercussions throughout the Company. In the design stage circuits have become more elaborate, and to ensure panclimatic operation the oven and refrigerator are now as commonplace development tools as the signal generator and oscilloscope. Presentation of information has become more of a problem as the front panel diminishes, although this was turned to advantage and the opportunity seized to recommence with a new set of common standards.

The new design concept first appeared with the announcement of the modular audio and medium frequency instruments, oscillators and attenuators with the by-product of signal sources. Alone and in combinations these units allowed us to create a new record in introducing eight new instruments in one issue. Subsequent events have shown the utility of such a flexible system in the ease with which it has been possible to originate further instruments, either by selecting different combinations, or, as in the case of the Transmission Measuring Sets, designing an additional unit to accompany existing ones. An audio frequency version of the Transmission Measuring Set is described briefly on page 206 of this issue, and it is shown why such an instrument is required in addition to the medium frequency version. Also in the audio field were two new bridges and a redesign of the Wave Analyser and Distortion Factor Meter. To match the low distortion that is now commonplace in audio oscillators a Tunable Rejection Filter was added to our range to extend the measuring facilities of the Wave Analyser. A circuit element of special interest, which might be considered as the heart of the Wave Analyser, is described here on page 195.

With the redesign of these older instruments the familiar type numbers are fast disappearing to be replaced by the "2000" Series. Blocks of numbers have been allocated in advance to the various product lines, and eventually a meaningful pattern will emerge as new replaces old. It is hoped that this system, rather than the previous random selection, will prove helpful to our customers.

Naturally considerable effort was also concentrated on our major product line—signal generators. As was seen in the previous issue this produced the M.F./H.F. A.M. Signal Generator type TF 2002, the first all solid state generator of its kind. Considerable investigation was required in its design to determine the most advantageous method of utilizing transistors, not just to equal the performance of thermionic valves, but also to give improvements and extend the facilities. Here again the construction is of unit form, but owing to the wide range of signal levels available, it highlighted the problem of earth loops in parallel with signal paths. This was overcome by the use of a novel component which is described here on page 204, as it also offers a solution to a similar problem frequently encountered in the laboratory when measuring a high insertion loss component using a generator and receiver, both with earth connections to a mains supply.

A bridge which will not be found in the Marconi Instruments' range. The photograph shows the new road bridge spanning the Firth of Forth, linking Edinburgh with the North

The trend in frequency modulated generators is also apparent from this volume. Here the emphasis has been on the need for wider deviations and higher modulation frequencies—requirements which have been satisfied by suitable modification to three of our standard designs.

For frequency and time measurement, the popularity of the 10 Mc/s Counter/Frequency Meter type TF 1417/2 continued to grow producing several variants on the basic design. These provided the solution to specialized measurement problems, and for quick reference the main features together with the particular application were collocated in one article. The measurement range was also extended up to 510 Mc/s by the availability of a frequency converter operating on the familiar heterodyne principle. Also in parallel with this came the new Counter type TF 2401 which with the appropriate plug-ins allows direct frequency measurement in excess of 50 Mc/s, or down to 0·1 μsec in time.

Perhaps the least likely candidate for transistorization is the oscilloscope, but despite this the 6 Mc/s transistorized Double Beam Oscilloscope type TF 2202 made its appearance at the beginning of the year as a bench or rack mounted instrument. As with counters, oscilloscopes are another product where there has been a constant demand for the "special". This has been met by modified versions of the Oscilloscope type TF 2200, five of which were described in one issue.

The majority of instruments considered above can be classified as general purpose, but there has also been considerable development in more specialized fields for such diverse applications as television monitoring and air traffic control. Another field in which Marconi Instruments have for many years played a leading role is in the measurement of intermodulation distortion in multi-channel communications systems using the white noise test signal specified by international agreement. In many of these systems microwave radio links are used with as as many as 40 repeater stations in a link. As each repeater can introduce distortion which is cumulative in the system, stringent test conditions are specified to ensure that the interference in the telephony system does not become intolerable. The original White Noise Test Set type OA 1249 was capable of measurements on systems containing up to 960 channels, but nowadays 1800 channel systems are in operation, and 2700 channel systems are under active development. To meet this anticipated increase in capacity a 2700 channel test set has been developed to replace the 960 channel version, and it is described in this issue on page 187. It is a tribute to modern components and circuit techniques that a three-fold increase in capacity has been achieved with a very considerable reduction in size and weight.

To summarize, it can be said that this volume has introduced to you, our customers, the new era of solid state instrumentation. This will be continued in the succeeding years to the extent that the instrument employing thermionic devices, and no doubt there will be applications where these are still advantageous, will be the exception rather than the rule.

P.M.R.

621.317.799:621.396.44

MARCONI INSTRUMENTS

NEW DESIGN

A 2700 CHANNEL
White Noise Test Set . . . . . *TYPE OA 2090*

by
H. C. GRIBBEN

The subject of the white noise method of intermodulation testing is introduced by describing briefly the frequency division multiplex system used in cable and radio multichannel links. Reasons for the use of white noise are given and the method of making an intermodulation measurement is explained. Design requirements for such a test set are considered in relation to the recommendations of the international committees C.C.I.R. and C.C.I.T.T. The new transistorized white noise test set, capable of testing systems of capacities from 12 channels to 2700, is described from the viewpoint of the user rather than in circuit detail.

IN A MULTI-CHANNEL cable or radio link, each telephone channel occupies a frequency band of 300 c/s to 3400 c/s; and in the frequency division multiplex (F.D.M.) system a number of such channels, each channel allocated a bandwidth of 4 kc/s, are placed side by side by means of frequency transposition. Twelve channels placed side by side in this manner is called a basic primary group and occupies the band of frequencies 12 kc/s to 60 kc/s. Secondary groups or supergroups are formed by assembling five such basic groups of twelve channels and transposing them to various positions in the frequency spectrum in the manner shown in Fig. 1.

The frequency spectrum occupied by the transposed channels is referred to as the baseband and is either transmitted direct along cables or used to frequency modulate the carrier in the case of a microwave radio link. The waveform of this total multiplex signal, consisting of the addition of a large number of transposed voice frequencies, closely resembles white noise having the same band of frequencies and distribution of peak amplitudes, i.e. gaussian distribution of peaks. For this reason it is convenient, when testing a link, to simulate fully loaded conditions by applying white noise having the correct level and baseband frequency spectrum.

Good intelligibility is the criterion of performance of a multichannel link system. To secure this, noise, which causes a deterioration in intelligibility, must be kept to a minimum. Main sources of noise are:

1. Intermodulation noise due to amplitude and phase non-linearity throughout the system.
2. Thermal noise generated in amplifiers and receivers.

It is the purpose of a white noise test set to provide a simple and accurate means of comparing the noise produced by the link with the level of an output signal due to the applied white noise. The resulting ratio is called the noise power ratio (N.P.R.) of the system.

A simple understanding of the method of measurement may be obtained if we assume that, say, a 600 channel system is completely loaded with normal speech traffic except for one channel. This channel is used as a listening-post and should be completely silent, but because of non-linearity throughout the link system, intermodulation products occur which are noticeable as noise. It is immaterial which channel is used as the quiet channel although intermodulation will in general be different at different points in the base band.

These conditions may be simulated by a white noise test set. A white noise generator is used in place of the speech traffic. Its output frequency range, as shown in Fig. 2a, is limited by high- and low-pass filters according to the capacity of the system under test. A quiet channel

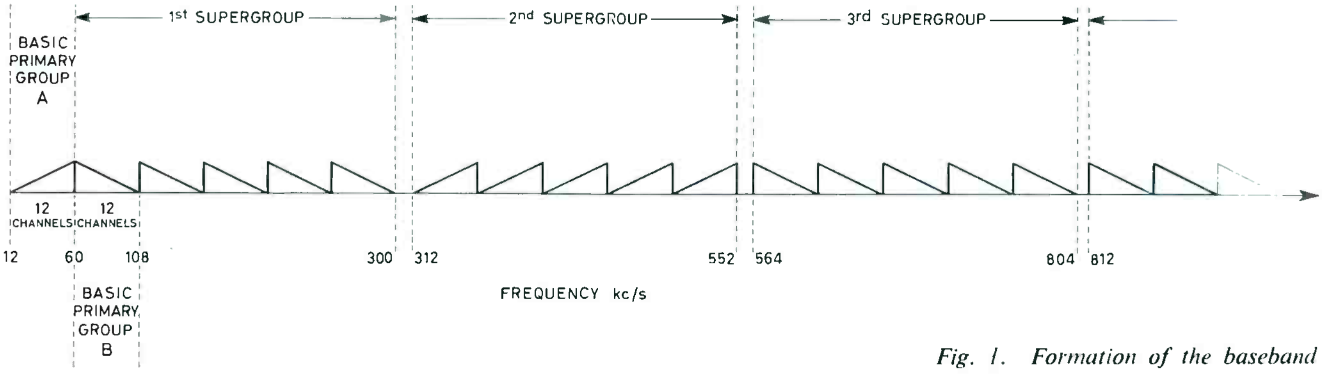
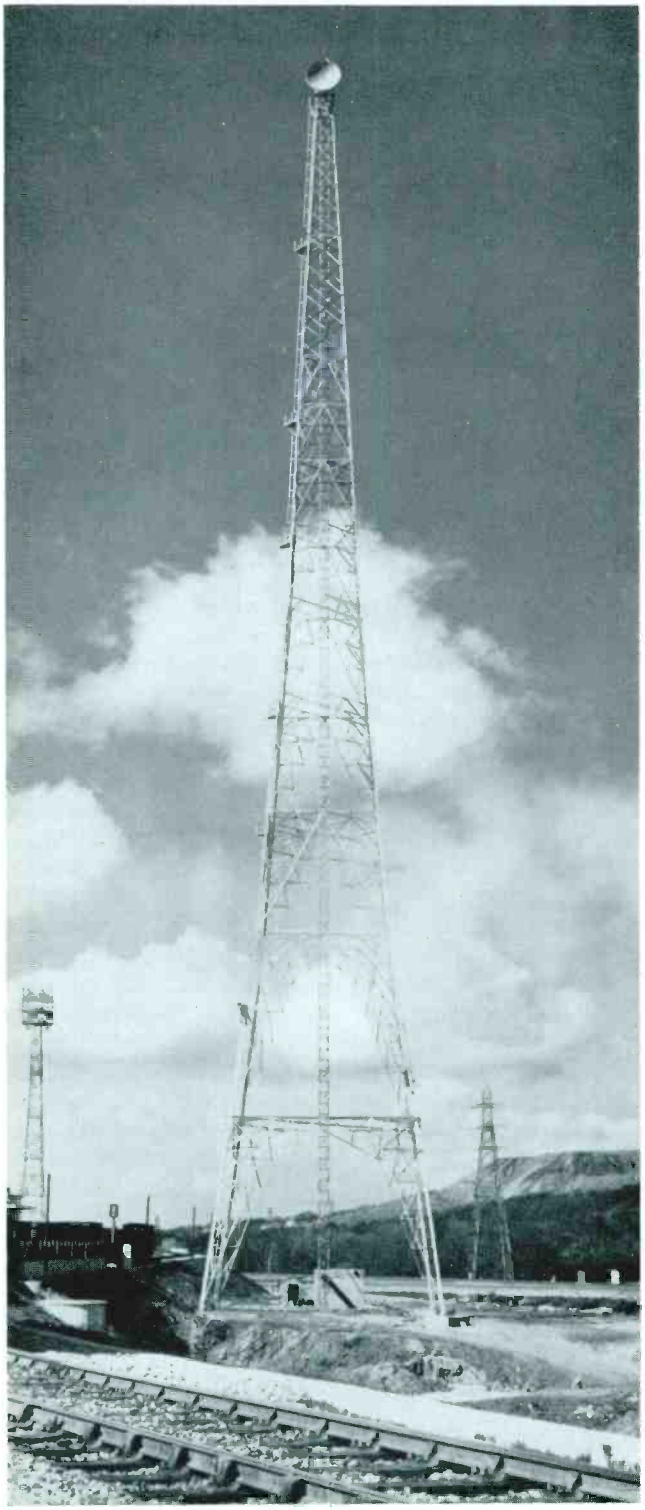


Fig. 1. Formation of the baseband



A microwave aerial tower for British Railways' 7,000 Mc/s. multichannel link between Newcastle and York. This link, which was supplied and installed by the Marconi Company Limited, has an ultimate capacity of 300 channels and is used for telephone traffic and high-speed data transmission

is produced by switchable band-stop filters, as shown in Fig. 2b. The white noise signal is fed to the baseband input at the sending end of the link, by-passing the channelling equipment.

The second part of the test set is a receiver which is switch-tuned to the frequency of the band-stop filter used to produce the quiet channel. It is connected to the baseband output at the receiving end of the link—again excluding the channelling equipment.

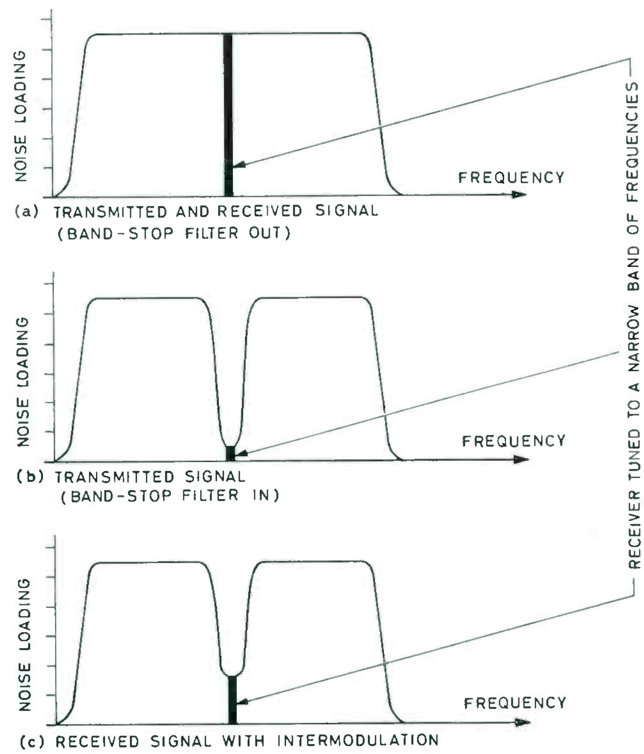


Fig. 2. Principle of operation of white noise test set

A measurement is made by setting the generator to the correct output level with the band-stop filter out and adjusting the receiver meter to the reference mark. When a band-stop filter is switched in, a narrow band of frequencies is attenuated by about 90 dB and the receiver meter deflection will fall. If there is no noise generated in the link equipment the meter deflection would be restored by reducing the receiver input attenuator by this same amount, i.e. 90 dB. Because of intermodulation and thermal noise it will be found that attenuation will have to be reduced by, perhaps, only 50dB, as shown in Fig. 2c. This change in attenuation is the noise power ratio of the system.

Design Considerations

The conditions under which noise and intermodulation tests should be made are set out as recommendations published in the handbooks of the International Committees C.C.I.R. and C.C.I.T.T., consequently the design of a white noise test set should, in general, be based on these recommendations.

TABLE 1

C.C.I.R. recommendations for baseband parameters

Maximum No. of telephone traffic channels	*Limits of band occupied by telephone channels (kc/s)*	*Frequency limits of baseband (kc/s)*	*Nominal impedance at baseband* Ω	*Relative power level per channel (dbr) at baseband interconnection points*			
				Radio equipment output	*Main repeater station* *Output*	*Main repeater station* *Input*	*Radio equipment input*
24	12–108	12–108	150 bal.	–15	–23	–36	–45
60	12–252 60–300	12–252 60–300	150 bal. 75 unbal.	–15	–23	–36	–45
120	12–552 60–552	12–552 60–552	150 bal. 75 unbal.	–15	–23	–36	–45
300	60–1300 64–1296	60–1364	75 unbal.	–18	–23	–36	–42
600	60–2540 64–2660	60–2792	75 unbal.	–20	–23 –33	–36 –33	–45
960	60–4028 316–4188	60–4287	75 unbal.	–20	–23 –33	–36 –33	–45
1800	312–8204 316–8204	300–8248	75 unbal.	–28	–33	–33	–37
2700	312–12388 316–12388	308–12435	75 unbal.	–28	–33	–33	–37

As an example, Recommendation 399 states that the power level of the signal with a uniform continuous spectrum should, for radio links with at least 240 telephone channels, be equal to $(-15 + 10 \log_{10} N)$ dBm at a point of zero relative level, N being the total number of channels in the circuit. For systems having at least 12 and not more than 240 channels the power level should be equal to $(-1 + 4 \log_{10} N)$ dBm at a point of zero relative level. This means that if the noise signal is to be applied to a point of zero relative level, the actual values of power required are those defined above. For 2700 channels this would be +19·3 dBm. If, however, the signal is to be applied to the baseband interconnection points (Table 1), as would normally be the case for maintenance tests on a radio relay system, the actual power input for a 2700 channel system conforming to C.C.I.R. recommendations would be $-37 + 19{\cdot}3 = -17{\cdot}7$ dBm. Higher power levels than this may be necessary for tests on a single amplifier, consequently, not only must the noise generator be capable of providing output levels of the order recommended by C.C.I.R. as those to be applied to points of zero relative level, but provision must also be made for accurately attenuating these higher output levels.

Useful information regarding noise levels likely to be encountered in practice can be obtained from figures given for allowable noise in a hypothetical reference circuit[1]. It is for the designer to apportion the allowable noise to the various parts of the link system; in general equal amounts are allocated to intermodulation noise and thermal noise. In the case of modem equipment only, this level could be of the order of –75 dBm per channel for each source. If a level of this order is to be measured accurately by means of the white noise method, the generator band-stop filter should attenuate the applied noise to a level at least 20 dB lower than this.

Noise and intermodulation distortion generated in the receiver itself will also affect the accuracy of measurement. If the error is not to exceed 0·1 dB, the noise generated in the receiver must be 20 dB lower than the signal to be measured. An error of 0·4 dB will result if receiver noise is 10 dB lower than the signal.

The N.P.R. corresponding to the noise level of –75 dBm and noise loading, due to the generator, of –15 dBm per channel, is 60 dB. A 60 dB attenuator in the receiver input would, therefore, be sufficient to enable the measurement to be made. However, it is necessary to make provision for checking the noise generated in the receiver itself—a back-to-back measurement—consequently the attenuator should have a range of the order of 90 dB. From this it follows that the receiver sensitivity should be –105 dBm per channel bandwidth or –111 dBm/kc/s—again with a noise loading of –15 dBm per channel.

White Noise Test Set Type OA 2090

This instrument consists of two units, a generator and a receiver. The generator unit carries the band limiting and band-stop filters and the receiver the necessary band-pass filters.

The instrument can be used for testing systems of any capacity from 12 channels, providing it is fitted with suitable filters. A total of nine filters can be fitted into the generator. By removing the bottom plate of the instrument case, filters can be changed within a matter of minutes. This same facility is provided on the receiver, which can carry a maximum of six band-pass filters. The instrument can be fitted, therefore, with filters suitable for two or three systems and any filter may be quickly changed. If a full complement of filters is not required blank panels are fitted in the vacant spaces.

A block diagram of both generator and receiver is shown in Fig. 3.

generated noise signal is not significantly impaired. This maximum output is required only for 2700 channels, consequently for lower capacity systems the overload capacity of the output stage is greater than the 10 dB figure quoted.

The output attenuator is in two sections, one having 4 steps of 10 dB, the second 11 steps of 1 dB—a total attenuation of 51 dB. Output level is calibrated, therefore, from +20 dBm to –51 dBm. Output impedance is 75 Ω, the standard for this type of equipment, although 150 Ω

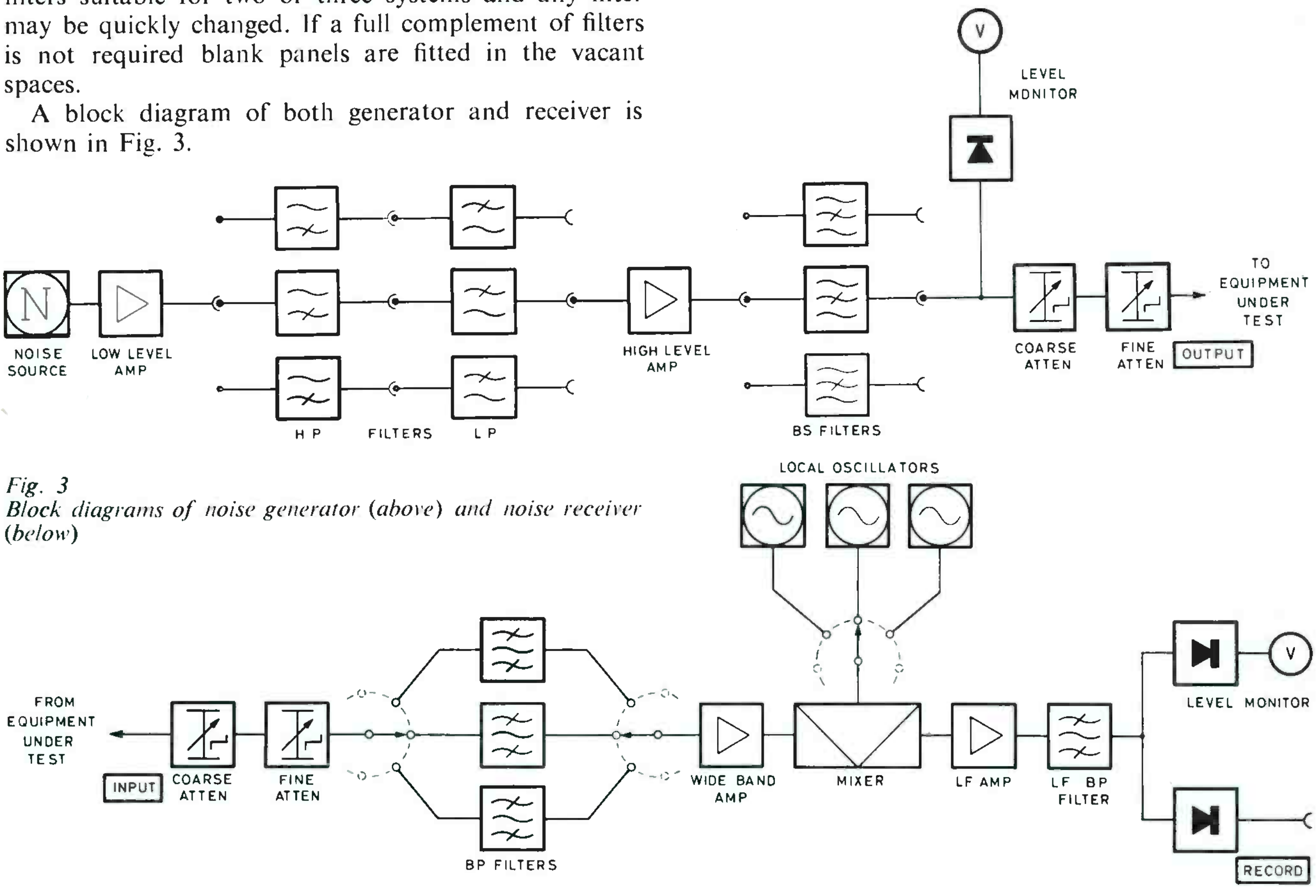


Fig. 3
Block diagrams of noise generator (above) and noise receiver (below)

Noise Generator Type TF 2091

Noise is generated over a bandwidth from below 10 kc/s to above 13 Mc/s by a biased silicon diode and passed to the first amplifier via an output control. The bandwidth of noise is now limited to the required frequency band by high-pass and low-pass filters so that the second amplifier is loaded only with the noise required and not with the total band of noise as would be the case if the filters were inserted after the second amplifier; this amplifier is followed by the selected band-stop filter, the power output monitoring circuit and the output attenuator.

The monitoring meter has 2 ranges, 0 – 10 dBm and 10 – 20 dBm, selectable by means of a front panel switch and calibrated at 1 dB intervals. Maximum output is specified as +20 dBm; at this level the output transistor will handle peak voltages more than 10 dB in excess of the r.m.s. value, therefore the gaussian nature of the

balanced outputs can be obtained by using the 75/150 Ω Transformer type TM 5955. With the attenuator set to zero attenuation, return loss is greater than 20 dB up to approximately 6 Mc/s and falls to about 15 dB at 12 Mc/s. When at least 6 dB of attenuation is switched into circuit the return loss is greater than 20 dB at all frequencies. In normal use considerably more than 6 dB of attenuation is used and the return loss is correspondingly better.

Noise Receiver Type TF 2092

In a normal superheterodyne receiver the local oscillator is tuned to a frequency several hundred kc/s higher than the received signal. If a second signal having a frequency of twice this difference frequency higher than the wanted signal appears at the receiver input, the receiver output will be affected unless the rejection of this image signal is extremely good. Where the incoming signal is a wide

White Noise Test Set type OA 2090

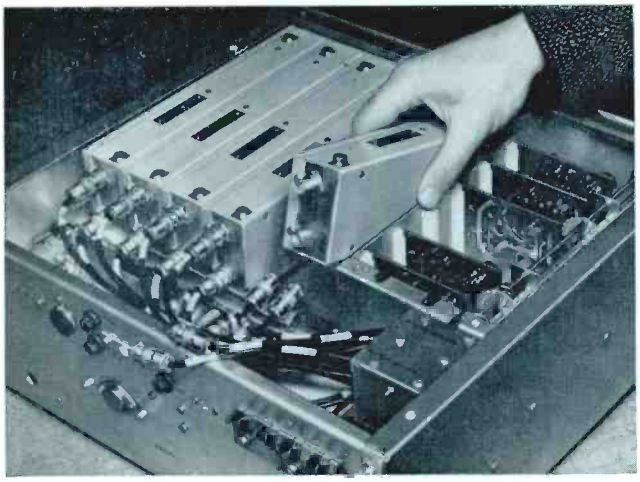
band of high level noise this effect can cause serious differences in measurement between different frequencies and different receivers.

This disadvantage of the normal superheterodyne has been overcome by adopting the method used by the British Post Office in their original white noise test set. The local oscillator is set to the centre frequency of the band-stop filter it is desired to use. The mixer is followed by a band-pass filter having a bandwidth from 500 to 1000 c/s. Consequently, two bands of frequencies spaced 500 to 1000 c/s below and above the oscillator frequency will be passed by the filter. The input band-pass filters have a bandwidth sufficiently wide to pass both signal channels; the noise bandwidth of the receiver is therefore 1 kc/s at all frequencies and possible difficulties due to insufficient image rejection are avoided. The band-pass filter following the mixer is introduced after the first stages of the high-gain I.f. amplifier so that noise generated in those stages will be reduced by the filter.

The total attenuation of the input attenuator is 111 dB in 10 steps of 10 dB and 11 steps of 1 dB. Dials fitted to the two attenuators are marked in such a manner that if they are both set to read zero for the higher level of the two N.P.R. readings, the sum of the attenuations, after switching in the band-stop filter and restoring the original meter indication, gives the noise power ratio directly.

A measurement is made by first selecting the required frequency, then setting the meter pointer to one of the reference marks on the scale by means of the OUTPUT LEVEL control, with the attenuators at "0" and the band-stop filter OUT. When the band-stop filter is switched IN the meter deflection falls and must be restored to its original setting by adjustment of the two input attenuators. The sum of the attenuator readings is the N.P.R. of the system as was explained above.

In many cases the input signal level will be such that only a very small, if any, indication will be evident on the meter when the attenuators are set to "0". If this is



The receiver accommodates up to six band-pass filters which are switch-selected at the front panel. New filters can be readily fitted as shown above

the case, it will be necessary to set the coarse attenuator to 10 or 20. This number must, of course, be subtracted from the final attenuator reading to give the N.P.R. value.

It will be noted that the receiver meter is not calibrated, but that all measurements are carried out on the attenuator by restoring the meter indication to a reference mark. This method has the advantage that meter calibration errors are avoided and measurement accuracy is not affected if inherent receiver noise occupies part of the meter scale. Maximum advantage can thus be taken of the receiver gain, which is of the order of −120 dBm/kc/s.

Level measurements may be made by first standardizing the receiver gain with the attenuators set to a fixed position, by using the noise generator as the standardizing noise source.

Filters

The design of all filters is in accordance with C.C.I.R. recommendations. Filters used in conjunction with the

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621.317.341

Measuring the Attenuation of Coaxial Cables

by
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B.Sc., Graduate I.E.E.,
A.K.C.

A method is shown whereby the attenuation of a length of coaxial cable can be measured using only a signal generator and an electronic voltmeter.

WHEN THE PARAMETERS of coaxial cables are measured, the most usually required constants are propagation coefficient, velocity ratio and characteristic impedance. There are, however, cases where the most important parameter to be measured is the attenuation per unit length. If sufficient of the line constants are known, this may be calculated, but it may be more convenient to measure the attenuation directly. The classical method of making this measurement uses a slotted line, but this is a tedious and complicated procedure, and the following analysis shows how the measurement may be carried out using only a signal generator and an electronic voltmeter—instruments that are usually readily available.

The testing equipment is set up as shown in Fig. 1, and the signal generator output frequency is adjusted to give a maximum reading, V_{max}, on the voltmeter, at a frequency as close as possible to the nominal test frequency. The signal generator frequency is then re-adjusted through the nominal test frequency, to give a minimum reading, V_{min}, on the voltmeter, again as close as possible to the nominal test frequency.

It can be shown by classical line theory (see Appendix) that, for lines that are long compared with the wavelength, the attenuation is given by

$$Al = 10 \log_{10} \frac{|V_{max}| + |V_{min}|}{|V_{max}| - |V_{min}|} \text{ dB}$$

where A = attenuation coefficient
and l = length of cable

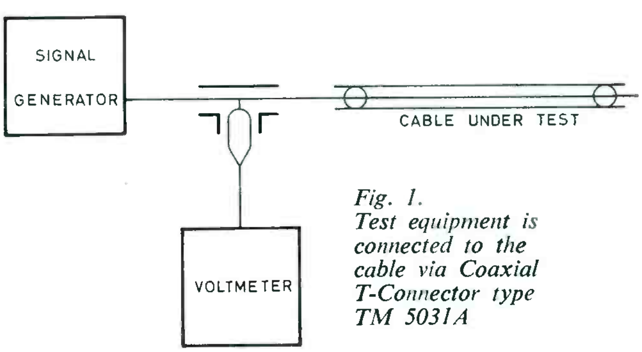


Fig. 1.
Test equipment is connected to the cable via Coaxial T-Connector type TM 5031A

Limitations of length of line and frequency

It can also be shown that the percentage change in frequency required to take the voltmeter reading from a maximum to a minimum is given by:—

$$\text{change in frequency} = \frac{v}{2lf} \times 100\%$$

generator—high-pass, low-pass and band-stop—are each fitted with a switch to enable them to be switched in or out of their particular circuit. Band-stop filters are fitted with red switches in order that they may be easily distinguished from band-limiting filters which are not used for measurement purposes and, once set for the capacity of the system under test, remain set for the series of measurements.

REFERENCE

1. C.C.I.R. Documents of the Xth Plenary Assembly, Geneva, 1963, Vol. IV, pp. 57-63.

ABRIDGED SPECIFICATION

Noise Generator, TF 2091

NOISE BAND CHARACTERISTICS: White noise is generated over the band from below 12 kc/s to above 12,388 kc/s; the noise level does not vary by more than 1 dB. When high-pass and low-pass filters are in circuit, the noise is attenuated by at least 25 dB at all frequencies lower than 20% below the high-pass cut-off frequency, and at all frequencies higher than 10% above the low-pass cut-off frequency.

STOP BAND CHARACTERISITCS: When a band-stop filter is switched in, noise is attenuated by more than 80 dB over a bandwidth of at least 3 kc/s, and by at least 3 dB at frequencies (0·02 F + 4) kc/s from the centre frequency F kc/s.

NOISE POWER OUTPUT: The reference level is adjustable up to a maximum of at least –15 dBm per 1 kc/s of bandwidth provided a total power output of +20 dBm is not exceeded. The monitor measures total power and has two ranges, 0 to +10 dBm and +10 to +20 dBm. Attenuators cover 51 dB in four 10 dB steps and eleven 1 dB steps.

OUTPUT IMPEDANCE: 75 Ω, with return loss greater than 20 dB.

Noise Receiver, TF 2092

EFFECTIVE BANDWIDTH: 1 kc/s at all input frequencies.

SENSITIVITY: Better than –115 dBm per 1 kc/s of bandwidth.

INPUT ATTENUATOR: Direct reading in noise power ratio from 0 dB to 91 dB, with additional settings of +10 dB and +20 dB, in ten 10 dB steps and eleven 1 dB steps.

INPUT IMPEDANCE: 75 Ω, with return loss greater than 20 dB.

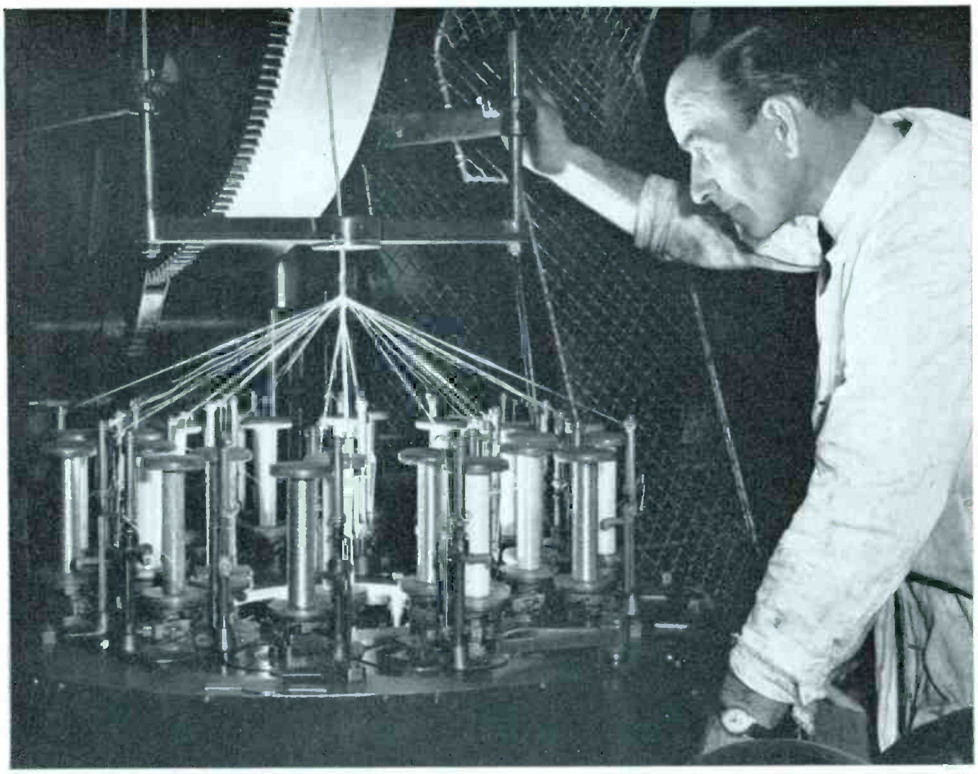
RECORDER OUTPUT: Suitable for use with 0·1mA recorders.

A reel of cable under test in a cable manufacturer's laboratory. This simple method was devised by Marconi Instruments Ltd. to speed production testing

where v = velocity of propagation
l = length of line
and f = frequency.

As the frequency is usually decided by other considerations, it would appear that the longer the line the better, as the percentage change in frequency should not be excessive. However, as the attenuation increases with the length of the line, V_{max} approaches V_{min} and the effect of a small error in either of these measurements will result in large errors in the final answer. Thus a compromise must be achieved in which the change of frequency must not be too large or the difference between V_{max} and V_{min} too small; this compromise will depend upon the variation of attenuation with frequency, and the actual attenuation of the cable involved.

Satisfactory measurements have been carried out on lengths of coaxial cable varying between 50 and 300 ft over a frequency range of 450 to 800 Mc/s, using Vacuum Tube Voltmeter type TF 1041C with Coaxial 'T' Connector type TM 5031A and U.H.F. Signal Generator type TF 1060.



In the manufacture of coaxial cables complex machines are required, such as this one which forms the outer screen

APPENDIX

The expression for attenuation can be derived[1] in the following manner.

For any line:

$V = a\,e^{-Px} + b\,e^{Px}$

where V = voltage at a point distance x from the source

P = propagation coefficient for the line $= A + jB$,

where A = attenuation coefficient

B = phase change coefficient

and a and b are constants

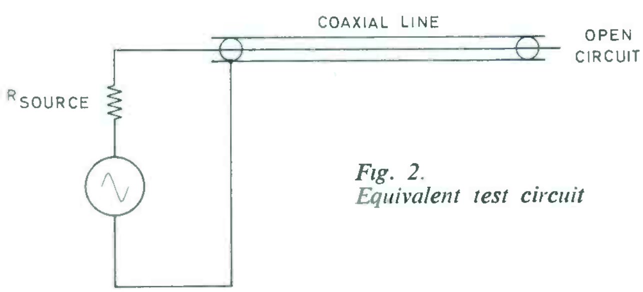


Fig. 2. Equivalent test circuit

Referring to Fig. 2,

$Vo = a + b$, since $x = 0$

where Vo = voltage at sending end of line

and $b = a\,e^{-2Pl}\,(Z_r - Z_o)/(Z_r + Z_o)$

where Z_o = characteristic impedance of the line,

Z_r = terminating impedance of the line,

and l = length of line

$\therefore\quad Vo = 2ae^{-Pl}\,(Z_r\cosh Pl + Z_o\sinh Pl)/(Z_r + Z_o)$

In the case of the open-circuited line, $Z_r = \infty$.

$$\therefore\quad Vo = 2ae^{-Pl}\cosh Pl$$
$$= 2ae^{-(A+jB)l}\cosh\,(A + jB)l$$
$$= 2ae^{-(A+jB)l}\,(\cosh Al\cosh jBl + \sinh Al\sinh jBl)$$
$$= 2ae^{-(A+jB)l}(\cosh Al\cos Bl + j\sinh Al\sin Bl)$$

When the open-circuit line is such that $Bl = 2n\pi/2$ where n is any integer then $Vo = V_{max}$

$$\therefore\quad V_{max} = 2ae^{-(Al+j2n\pi/2)}\cosh Al$$

and, if $Bl = (2n + 1)\pi/2$ then $Vo = V_{min}$

$$\therefore\quad V_{min} = 2ae^{-[Al+j(2n+1)\pi/2]}\,j\sinh Al$$

Hence, $V_{max} = 2ae^{-(Al+j2n\pi/2)}\,\tfrac{1}{2}(e^{Al} + e^{-Al})$

$$\therefore\quad |V_{max}| = 2ae^{-Al}\tfrac{1}{2}(e^{Al} + e^{-Al})$$

similarly, $V_{min} = j2ae^{-[Al+j(2n+1)\pi/2]}\,\tfrac{1}{2}(e^{Al} - e^{-Al})$

$$\therefore\quad |V_{min}| = 2ae^{-Al}\tfrac{1}{2}(e^{Al} - e^{-Al})$$

$$\therefore\quad |V_{max}| + |V_{min}| = 2ae^{-Al}\tfrac{1}{2}(e^{Al} + e^{-Al} + e^{Al} - e^{-Al})$$
$$= 2ae^{-Al}e^{Al}$$
$$= 2a$$

and $|V_{max}| - |V_{min}| = 2ae^{-Al}\tfrac{1}{2}(e^{Al} + e^{-Al} - e^{Al} + e^{-Al})$

$$= 2ae^{-Al}\,e^{-Al}$$
$$= 2ae^{-2Al}$$

$$(|V_{max}| + |V_{min}|)/(|V_{max}| - |V_{min}|) = 2a/2ae^{-2Al} = e^{2Al}$$

$$\therefore\quad Al = \tfrac{1}{2}\log_e(|V_{max}| + |V_{min}|)/(|V_{max}| - |V_{min}|)\text{ nepers}$$
$$= 10\log_{10}(|V_{max}| + |V_{min}|)/(|V_{max}| - |V_{min}|)\text{dB.}$$

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621.372.543.2

MARCONI INSTRUMENTS

A 100 kc/s Crystal Filter

by
M. TILEY,
B.Sc. Graduate I.E.E.

A description is given of the design of a highly selective crystal filter centred on 100 kc/s, and having a bandwidth of 6 c/s. An approximately maximally flat response is obtained by treating the crystals as series resonant elements in a hypothetical filter model, and then evaluating the remainder of the circuit constants by synthesizing the desired transfer function from the positions of the poles of the circuit in the complex frequency plane.

A QUARTZ CRYSTAL RESONATOR provides a useful basis for the design of tuned filters where the high selectivity required prohibits the use of conventional wound inductors. Modern low frequency crystals mounted on soldered wire supports in evacuated glass envelopes, have a Q-factor of several tens of thousands.

The electrical and mechanical properties of crystalline quartz are not the same in all directions, and hence various cuts of crystal, although resonating at the same frequency, may have widely differing properties. Of these, the most important are the temperature coefficient of frequency, and the magnitude of secondary resonances, which may be troublesome. Other considerations from a commercial standpoint are size and cost, and the ease of pulling the crystals onto frequency. The 0°X cut bar was chosen as the best compromise, and its properties will be discussed in detail.

The 0°X Cut Crystal

A quartz crystal possess three types of principal axis[1] a single optical, or Z axis, along which there is no piezo-

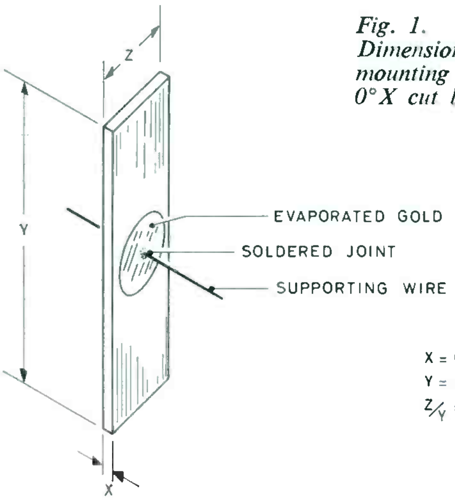


Fig. 1. Dimensions and mounting of 0°X cut bar

electric effect, three Y or mechanical axes, and three electrical or X axes. The 0°X cut bar is an on-axis cut, having its shortest dimension along an X axis, as shown in Fig. 1. The crystal resonates longitudinally along the Y axis, and the supporting wires are soldered at the centre of opposite crystal faces, putting no restraint on the vibrations. Secondary responses are thus minimized, and the equivalent circuit becomes the simple configuration of Fig. 2.

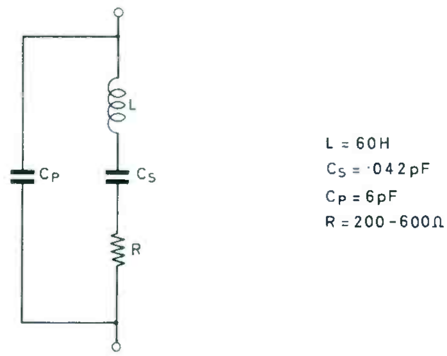


Fig. 2. Equivalent circuit of 0°X cut bar

Two other cuts of crystal are often used as resonators[2]. These are the +5°X cut and the –18·5°X cut. The +5°X cut has the lowest temperature coefficient of frequency, but difficulties with secondary frequencies are worst. The –18·5X cut is virtually free of secondary responses, but has very poor temperature coefficient. Thus the 0°X cut lying between the two, appears to be the best compromise.

Another factor affecting the frequency-temperature performance of the crystal is the length-width ratio. The frequency-temperature curve is roughly parabolic and the greater the length/width ratio the higher is the turn-over temperature. In practice it is not convenient to produce a ratio of greater than 10:1, giving a turn-over point, as shown in Fig. 3, at about 18°C which is reasonably close to average ambient conditions. A rise in temperature of 20°C above the turn-over point shifts the resonant frequency of a 100 kc/s crystal by about 1·5 c/s. In the present design this is not important, but as it is a multi-crystal filter, it is important that the behaviour of all crystals is very similar. This chiefly depends on holding a tight tolerance on the angle of cut. If two crystals differ by 10 minutes of arc, and are both set on frequency at 18°C, they will differ by 0·1 c/s at 40°C. The length/width ratio also affects temperature

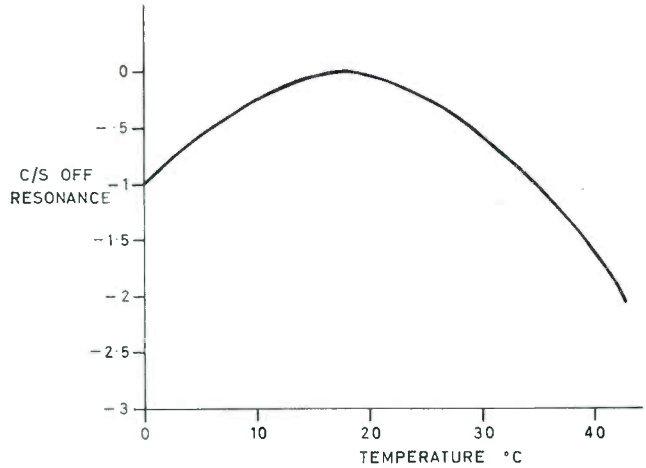


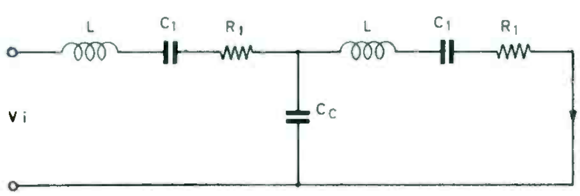
Fig. 3. Resonant frequency variations of crystal 0°X cut bar with temperature

performance, but is negligible if the ratio is held within a percent or so.

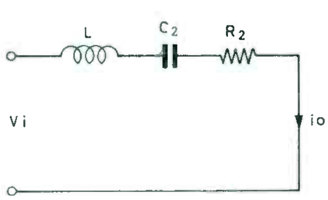
The crystal is designed to work into a capacitance of 50 pF. A small air trimmer can therefore be used to pull the crystal onto frequency. The effect of the crystal shunt capacitance is to introduce an anti-resonance at about 100 kc/s + 350 c/s. Over a narrow bandwidth around 100 kc/s, the effect is negligible, so does not invalidate the present maximally flat design, which ignores the shunt capacitance altogether. It does however considerably modify the skirt response.

Filter Model

To adequately meet the filter requirements, it is necessary to use three crystals. A useful arrangement is to split the filter circuit into two independent parts, a double tuned stage with capacitance coupling and a single tuned stage as shown in Figs. 4a and 4b respectively. It has already



(*a*) *Double crystal filter*



(*b*) *Single crystal filter*

Fig. 4. Simplified equivalent circuit of crystal filter

been mentioned that the parallel capacitance of the crystal is neglected in the design. As a circuit synthesis technique is used this makes the transfer function more manageable by reducing the number of poles and zeros.

The transfer function of the double crystal filter is

$$\left(\frac{i_o}{v_i}\right)_1 = \frac{\omega_1^2}{L}\,\frac{p}{\left(p^2+\frac{R_1}{L}p+\omega_1^2\right)\left(p^2+\frac{R_1}{L}p+\omega_1^2+\frac{2}{LC_c}\right)}$$

where $\omega_1 = \frac{1}{\sqrt{LC_1}}$

This function contains one zero at $p = o$, which need not concern us, and two pairs of conjugate poles, p_a, p_a^* and p_b, p_b^*, where

$$p_a = -\frac{R_1}{2L} + j\sqrt{\omega_1^2-\left(\frac{R_1}{2L}\right)^2}$$

$$p_b = -\frac{R_1}{2L} + j\sqrt{\omega_1^2-\left(\frac{R_1}{2L}\right)^2+\frac{2}{LC_c}}$$

These poles occur in the only quadrant of the complex frequency plane (p-plane) which corresponds to real frequencies. Their conjugates, p_a^* and p_b^* can be plotted on the p-plane but have not physical realizability.

Similarly, the transfer function of the single crystal filter is

$$\left(\frac{i_o}{v_i}\right)_2 = \frac{1}{L}\,\frac{p}{p^2+\frac{R_2}{L}p+\omega_2^2}$$

Which again has one zero at the origin, and only one pair of conjugates poles, p_c, p_c^*, where

$$p_c = -\frac{R_2}{2L} + j\sqrt{\omega_2^2-\left(\frac{R_2}{2L}\right)^2}$$

and $\omega_2 = \frac{1}{\sqrt{LC_2}}$

The overall transfer function is obtained by multiplying the two above functions together, and has therefore two zeros at the origin, and three conjugate pairs of poles.

The Maximally Flat Response

The maximally flat response for a band-pass network provides optimum flatness in the pass band with zero ripple, and also maximum rate of fall-off outside the pass band. It can be shown that to achieve maximum flatness, the poles of the transfer function must be equally spaced around a semicircle on the p-plane, with centre co-ordinates $(0, j\omega_o)$ and diameter B, where ω_o is the centre angular frequency of the filter, and B is the 3 dB bandwidth in radians. The p-plane plot for the case of a three pole network is shown in Fig 5.

The co-ordinates of the three poles positioned to give maximally flat reponse are, by simple algebra.

$$p_a = -\frac{B}{4} + j\left(\omega_o - \frac{\sqrt{3}}{4}B\right)$$

$$p_b = -\frac{B}{4} + j\left(\omega_o + \frac{\sqrt{3}}{4}B\right)$$

$$p_c = -\frac{B}{2} + j\omega_o$$

Two sets of expressions have now been derived, from which it is possible to calculate the unknown circuit constants.

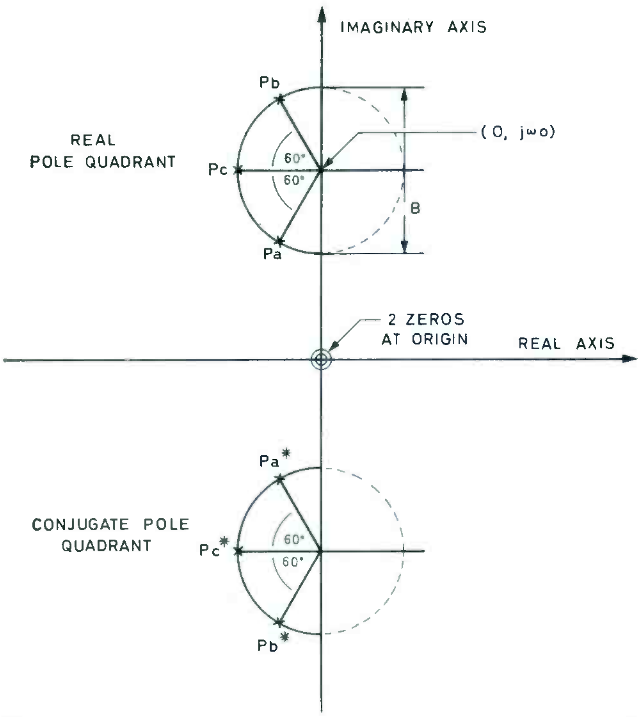


Fig. 5. Complex frequency plane representation of maximally flat response

Calculation of Circuit Constants

The values of L (60H) and ω_0 ($2\pi \times 10^5$ radians/sec) are specified by the crystal. To achieve high rejection 20 c/s off tune, the 3 dB bandwidth must be made small compared with 40 c/s, but not small enough to make tuning difficult. A compromise of 6 c/s is adopted, i.e. $B = 12\pi$ radians/sec. The remaining unknowns in the two expressions for p_a, p_b and p_c are the damping resistors R_1 R_2, the coupling capacitor Cc, and the resonant frequencies ω_1 and ω_2. These are evaluated by equating real and imaginary parts in the two sets of expressions.

Real parts: $$\frac{B}{4} = \frac{R_1}{2L} \qquad (1)$$

$$\frac{B}{2} = \frac{R_2}{2L} \qquad (2)$$

Imaginary parts:

$$\left(\omega_0 - \frac{\sqrt{3}}{4} B\right)^2 = \omega_1^2 - \left(\frac{R_1}{2L}\right)^2 \qquad (3)$$

$$\left(\omega_0 + \frac{\sqrt{3}}{4} B\right)^2 = \omega_1^2 - \left(\frac{R_1}{2L}\right)^2 + \frac{2}{LC_c} \qquad (4)$$

$$\omega_0^2 = \omega_2^2 - \left(\frac{R_2}{2L}\right)^2 \qquad (5)$$

From (1) and (2)

$$R_1 = \frac{BL}{2} = 1131\,\Omega$$

$$R_2 = BL = 2262\,\Omega$$

From (3) and (4)

$$C_c = \frac{2}{\sqrt{3}\,BL\omega_0} = 800\text{ pF}$$

$$\omega_1 \simeq \omega_0 - \frac{\sqrt{3}}{4} B = 2\pi\,(10^5 - 2{\cdot}6)\text{ radians/sec.}$$

And from (5)

$$\omega_2 \simeq \omega_0 = 2\pi 10^5\text{ radians/sec.}$$

Setting up a 100 kc/s crystal filter with the aid of a high-discrimination oscillator, oscilloscope and voltmeter

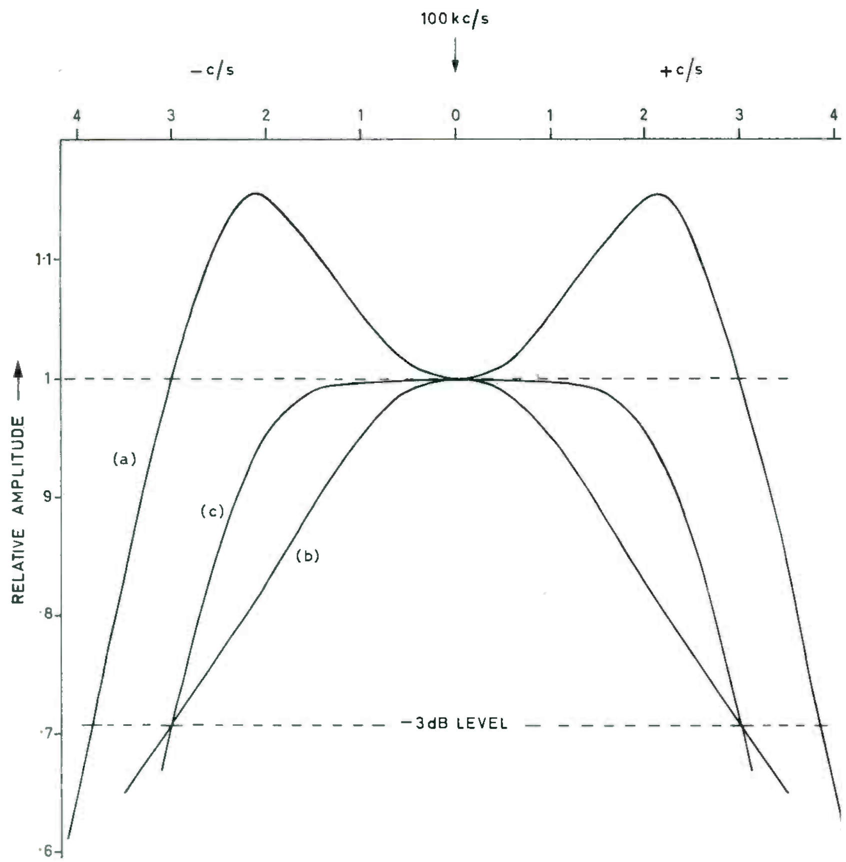


Fig. 6.

Response curves of maximally flat filter

(a) Double crystal filter
(b) Single crystal filter
(c) Overall flat response

Normalized Filter Response
The normalized frequency response of the double crystal filter can be shown to have the form

$$F_1(\triangle f) = \frac{4}{\sqrt{\left[1+\left(\frac{2\triangle f}{3}-\sqrt{3}\right)^2\right]\left[1+\left(\frac{2\triangle f}{3}+\sqrt{3}\right)^2\right]}}$$

Where $\triangle f$ is the frequency in c/s, measured with reference to *fo*, the filter centre frequency. The form of the response is shown in Fig. 6 (a), and is the response of an overcoupled double tuned stage.

The single crystal filter has the response

$$F_2(\triangle f) = \frac{1}{\sqrt{1+\frac{(\triangle f)^2}{9}}}$$

which is shown in Fig. 6(b). The overall response, which is maximally flat with a 0·1 dB bandwidth of 3 c/s is shown in Fig. 6(c).

Practical Filter Circuit
A convenient circuit arrangement is shown in Fig. 7. Emitter followers provide a suitable low impedance source for each filter section. The overall insertion loss of this circuit is typically 12 dB. If gain is required in the system, a better arrangement is to replace the emitter

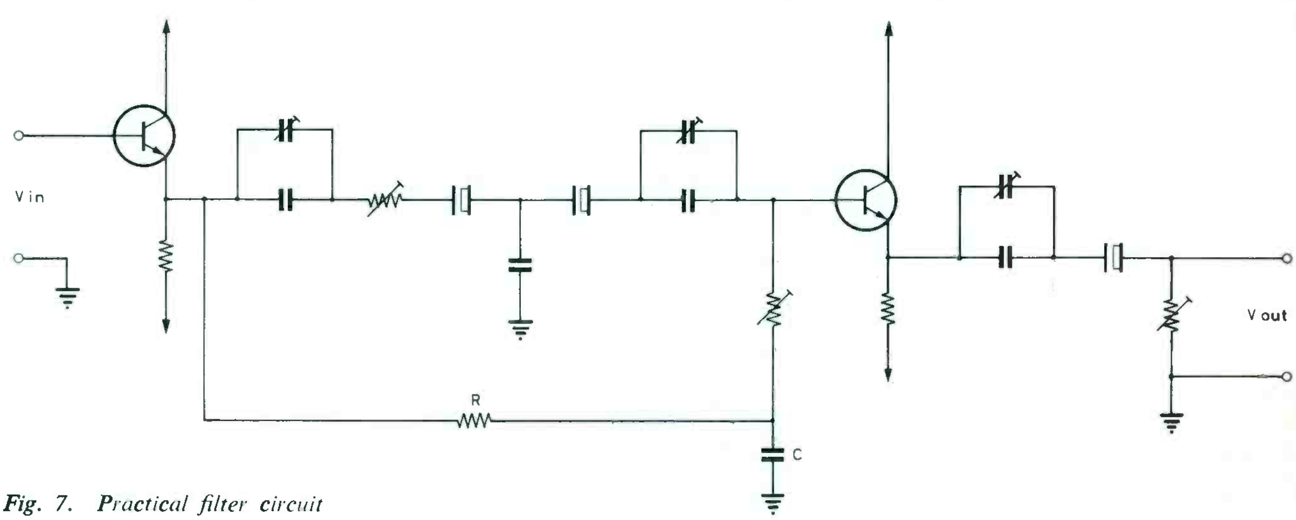
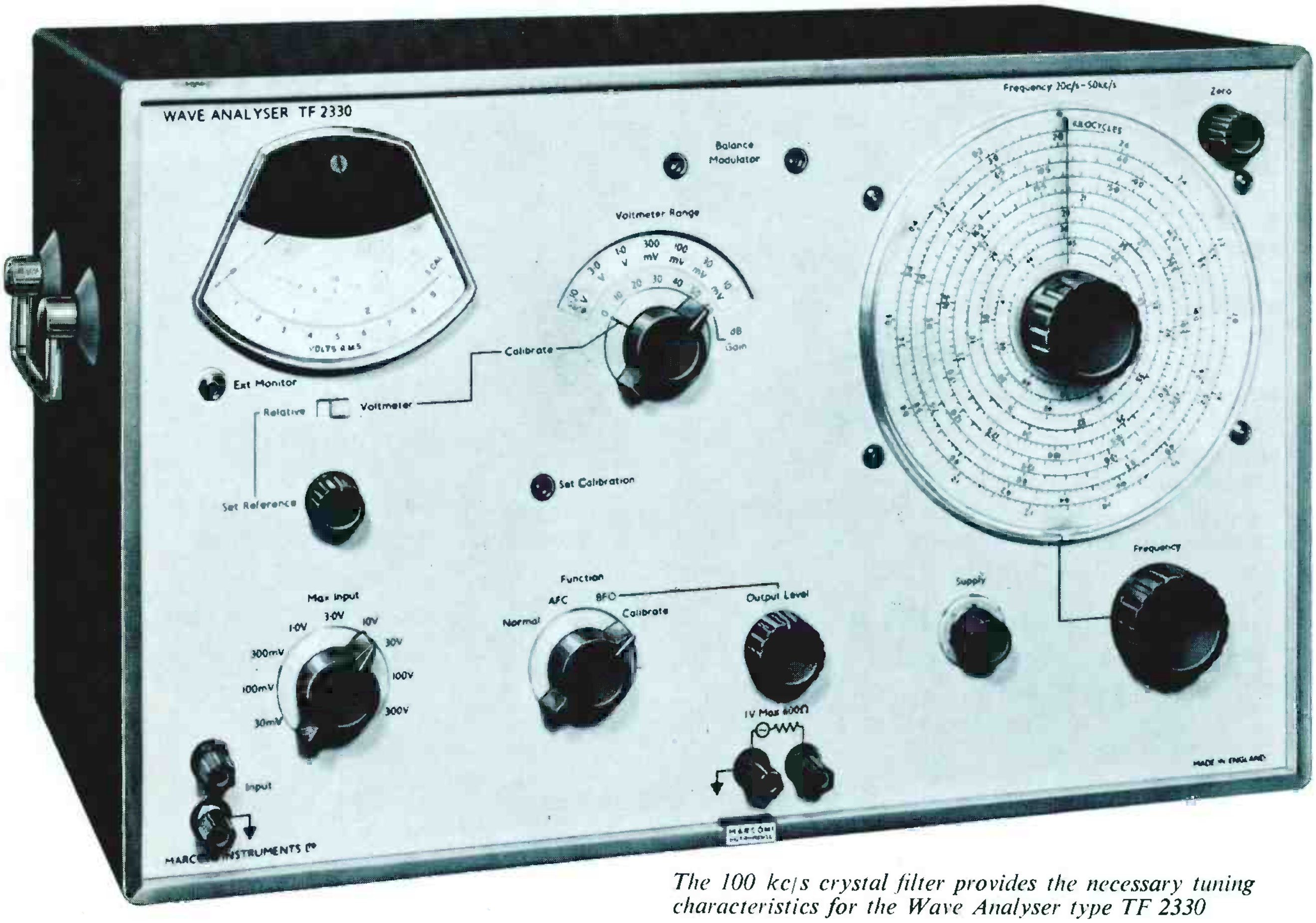


Fig. 7. Practical filter circuit



The 100 kc/s crystal filter provides the necessary tuning characteristics for the Wave Analyser type TF 2330

followers by gain stages with tuned step-down transformers in the collectors of the transistors. The circuit of Fig 8, when used to replace each emitter follow, provides a circuit with a gain typically greater than 35 dB.

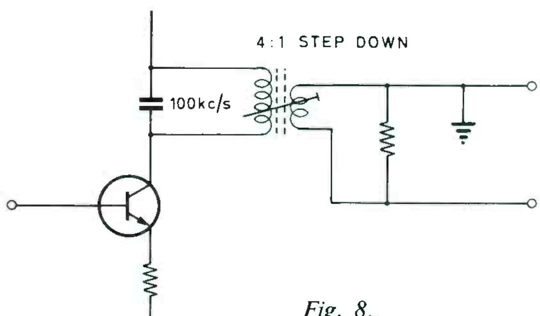


Fig. 8. Tuned amplifier for feeding filter section

Components R and C in the double crystal filter provide a small amount of voltage feedthrough with a phase angle of –90°, and has the effect of steepening the filter roll-off. The phase response of the filter is shown in Fig. 9, and it can be seen that the phase shift effectively approaches +90° at frequencies above and below resonance, and thus tends to cancel out with the fed-through voltage. The magnitude of the fed-through voltage is such as to give maximum cancellation at about ±50 c/s from 100 kc/s.

A practical circuit response is compared with the derived maximally flat response in Fig. 10, showing the

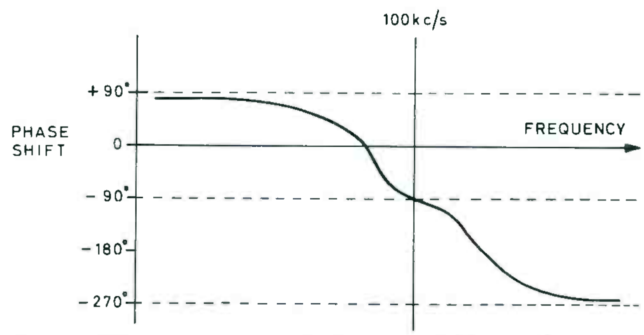


Fig. 9. Phase response of double crystal filter of Fig. 4 (a)

asymmetry produced by the crystal shunt capacitance. This capacitance also allows high frequency feed-through, which is eliminated by cascading the filter with a conventional tuned amplifier.

Setting up the filter

A useful high discrimination oscillator for setting up the filter can be produced by mixing and filtering the

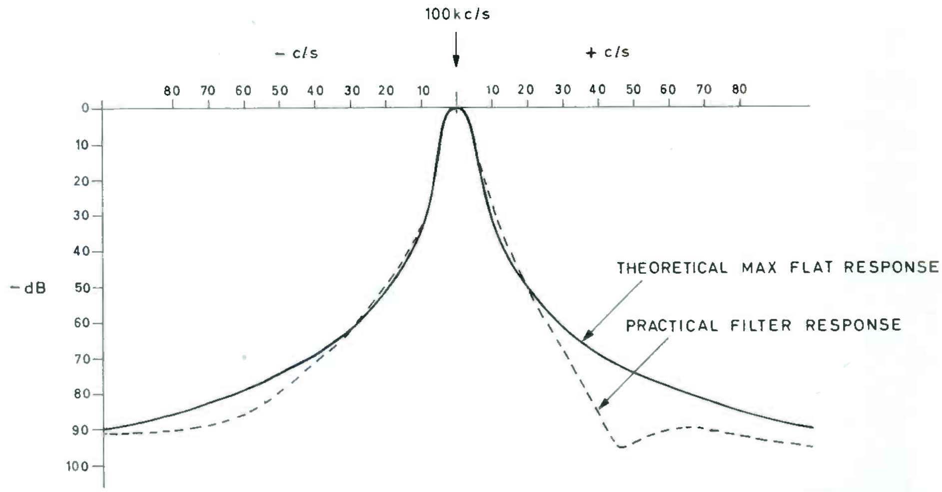


Fig. 10.
Overall filter response compared with theoretically maximally flat response

outputs from two high frequency crystal oscillators say 4·65 and 4·55 Mc/s, and pulling the frequency of one of the crystals a calibrated ±10 c/s.

The double crystal filter is first set up independently by feeding in a signal of exactly 100 kc/s, and adjusting both trimmer capacitors alternately to provide maximum outputs, with the preset potentometers set to mid-travel. The levels and frequencies of the response humps are then finally adjusted, tuning through with the high discrimination oscillator.

The single crystal filter is now independently set up for 100 kc/s centre frequency, and 6 c/s bandwidth. Finally, when the two parts have been cascaded, the trimmer capacitor of the single crystal filter is adjusted to obtain maximum flatness of response.

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621.317.335.2

UNIVERSAL IMPEDANCE BRIDGE EVALUATES COMPLEX NETWORKS

by
B. F. D. STEGER-LEWIS, B.Sc., Graduate I.E.E., A.K.C.

In many laboratories there are universal impedance bridges such as the Marconi Instruments Type TF 1313 and, whereas the use of these instruments for the measurement of component values is obvious, their use for more complex measurements is not always fully appreciated. This article deals with the measurement of a delta network of capacitors, one point of which is permanently earthed.

NOWADAYS it is not uncommon to find an underground power cable incorporating within its sheath a number of pairs of signal wires. If these wires are to be used for the transmission of data, etc., the user requires to know, among other things, the capacitance between the two wires of any pair, and this parameter must ultimately be measured after the cable has been laid.

If all the conductors within the sheath are earthed except for the pair of wires under consideration then the equivalent circuit is as shown in Fig. 1, and the expression for the total capacitance between the wires is given by:

$$C_T = C_{12} + \frac{C_{1E}\,C_{2E}}{C_{1E}+C_{E2}}$$

Where C_T = the total capacitance between the pair of wires
C_{12} = the mutual capacitance between the pair of wires
C_{1E} = the capacitance between conductor 1 and earth
and C_{2E} = the capacitance between conductor 2 and earth.

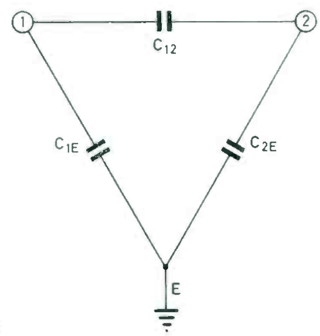
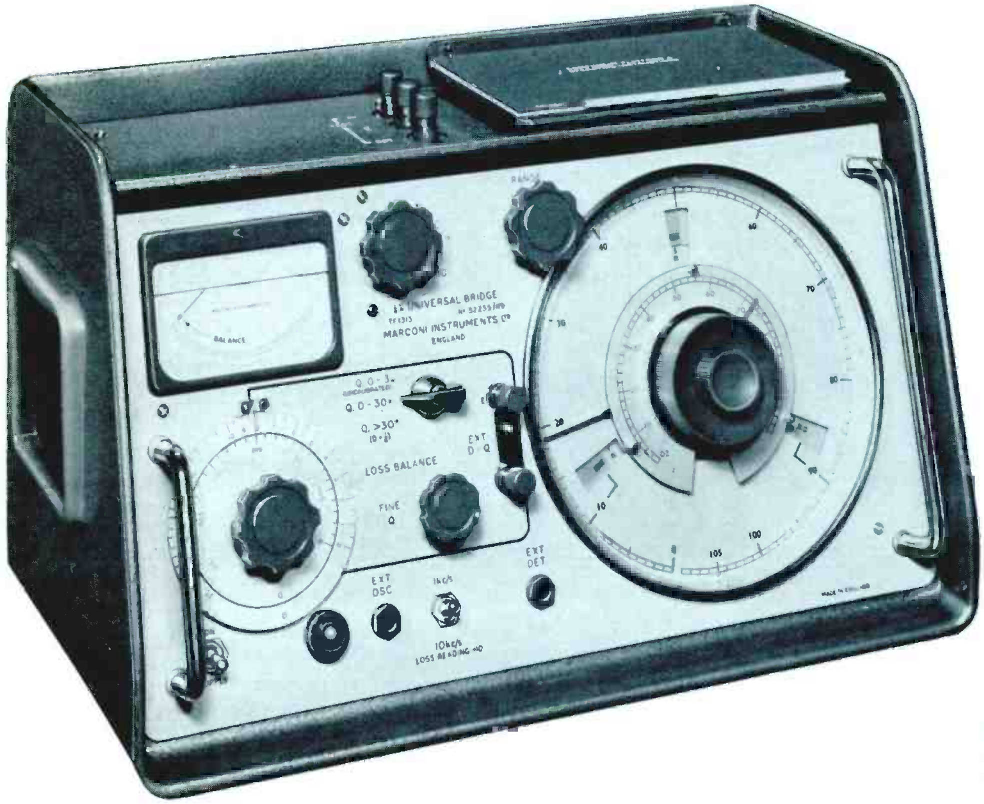


Fig. 1.
Delta network formed by capacitance between pair of wires and earth



This article deals with one of the many applications of the $\frac{1}{4}$% Universal Bridge type TF 1313

C_T may be measured directly using a universal bridge which is isolated from earth, e.g. a battery operated model such as the TF 2700 or TF 2701.

In this case the pair of wires would be connected to the bridge terminals and the capacitance measured would be C_T. If, on the other hand, the same measurement is attempted using a mains operated universal impedance bridge of the TF 1313 type, the answer is not C_T.

This does not, however, mean that such a bridge cannot be used for this measurement, and the way in which it may be used for measurements on a delta network of the type shown in Fig. 1, and more particularly for the evaluation of C_T as defined above, is the subject of this article.

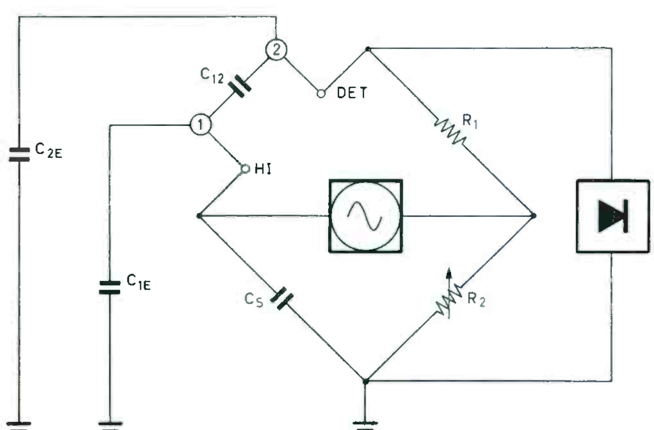


Fig. 2. Network connected to bridge terminals

First consider why it is not possible to make a direct measurement as can be done with a battery operated bridge which is isolated from earth. The basic bridge circuit is shown in Fig. 2, and it can be seen than when the HI and DET terminals of the bridge are connected to conductors 1 and 2 respectively, C_S is shunted with C_{1E}. Capacitance C_{2E} may be neglected as at balance the DET terminal is at earth potential. Thus C_{12} is measured against a standard that is no longer known, i.e. C_S in parallel with C_{1E}. If in attempting to overcome this difficulty the bridge is isolated from earth, then a more complex situation arises in which C_S is shunted by C_{1E} in series with the capacitance of the mains transformer in the bridge.

As a direct measurement cannot be made it will be shown how to correct for the shunting effect due to C_{1E}, as encountered when using the bridge in the normal earthed mode. Returning to Fig. 2, at balance

$$C_{12}=\frac{R_1}{R_2}(C_S+C_{1E}) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (1)$$

But as the bridge has been calibrated with respect to C_S only, it will indicate the value of C_{12} as a lower value, A, where

$$A=\frac{R_1}{R_2}C_S$$

$\therefore \frac{R_1}{R_2}=\frac{A}{C_S}$ and substituting for $\frac{R_1}{R_2}$ in equation 1 gives

$$C_{12}=A\left(1+\frac{C_{1E}}{C_S}\right) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (2)$$

If a further capacitance of value C is added between the HI and DET terminals as shown in Fig. 3, then at balance:—

$$C_{12} + C = \frac{R_1'}{R_2'}(C_S + C_{1E}) \quad \ldots\ldots (3)$$

And again the indicated value will be some lower value, B, where

$$B = \frac{R_1'}{R_2'} C_S$$

$\therefore \frac{R_1'}{R_2'} = \frac{B}{C_S}$ and substituting for $\frac{R_1'}{R_2'}$ in equation 3 gives

$$C_{12} + C = B\left(1 + \frac{C_{1E}}{C_S}\right) \quad \ldots\ldots (4)$$

Subtracting equation 2 from equation 4 gives

$$C = (B - A)\left(1 + \frac{C_{1E}}{C_S}\right)$$

$$\therefore \; C_{1E} = \left(\frac{C}{B-A} - 1\right) C_S \quad \ldots\ldots (5)$$

and substituting for C_{1E} in equation 2 gives

$$C_{12} = A\left(1 + \frac{C}{B-A} - 1\right) = \frac{AC}{B-A} \quad \ldots\ldots (6)$$

The values A and B are known, and C may be measured using the bridge in the conventional manner as shown in Fig. 4 thus giving the value of C_{12}. By substituting this value and the known value of C_S in equation 5, C_{1E} may be found.

If the delta network connections to the HI and DET terminals are reversed as shown in Fig. 5, at balance

$$C_{12} = \frac{R_1''}{R_2''}(C_S + C_{2E}) \quad \ldots\ldots (7)$$

and the indicated value on the bridge will be D, where

$$D = \frac{R_1''}{R_2''} C_S$$

$\therefore \frac{R_1''}{R_2''} = \frac{D}{C_S}$ and substituting for $\frac{R_1''}{R_2''}$ in equation 7 gives

$$C_{12} = \frac{D}{C_S}(C_S + C_{2E})$$

$$\therefore \; C_{2E} = \left(\frac{C_{12}}{D} - 1\right) C_S$$

and substituting the expression for C_{12} obtained from equation 6 gives

$$C_{2E} = \left[\frac{AC}{(B-A)D} - 1\right] C_S \quad \ldots\ldots (8)$$

Thus having made four measurements using the configuration shown in Figs. 2, 3, 4 and 5, the indicated values being A, B, C and D respectively, it is now possible to calculate the values of C_{12}, C_{1E} and C_{2E}—the value of C_S will depend upon the bridge used, for TF 1313 it is 0.1μF. Hence the value of C_T may be calculated, using the formula

$$C_T = C_{12} + \frac{C_{1E}\,C_{2E}}{C_{1E} + C_{2E}}$$

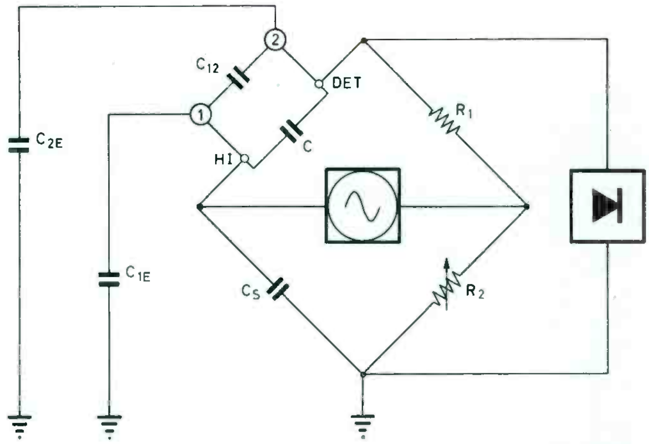


Fig. 3. Network plus auxiliary capacitor across bridge terminals

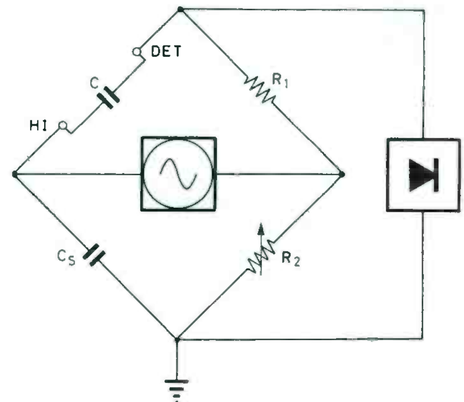


Fig. 4. Evaluating the auxiliary capacitor

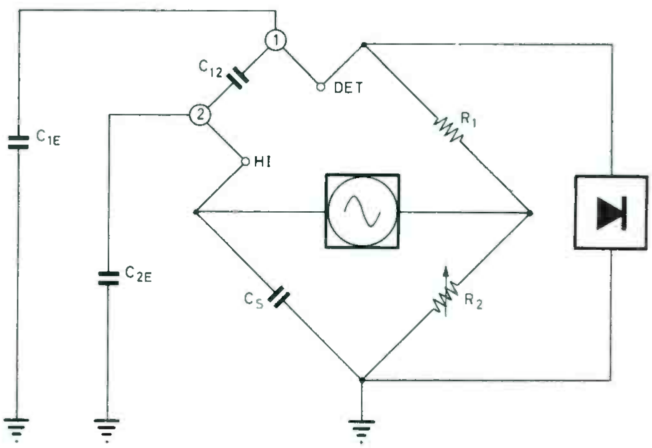


Fig. 5. Delta network reversed

A method of using a universal impedance bridge to make measurements on a complex network has been shown; this type of analysis and method is applicable to many forms of complex circuit, and can be developed to suit specific applications.

621.373.421

WIDE RANGE R-C OSCILLATOR, TF 1370A

AS THE ILLUSTRATION SHOWS, we are introducing a modernized version of the Wide Range Resistance Capacitance Oscillator type TF 1370. In addition, this new model—TF 1370A—has an extended high level output frequency response and an additional output impedance of 130 Ω.

Extending the frequency range of the amplifier has provided a high output level up to a frequency of 1 Mc/s instead of only 100 kc/s. Thus over a frequency range of 10 c/s to 1 Mc/s a signal level of 31.6 V is available, with a distortion factor of less than 1 % over the greater part of the frequency range with 2 % at the extremities.

The output impedance is basically that of a 75 Ω ladder attenuator padded to final value, hence a suitable resistor enables the additional output impedance of 130 Ω to be provided. This facility applies only to outputs up to 3.16 V.

The same basic circuit arrangements are retained for this new version using the Wien bridge system for frequency selection. Advantage was taken of the opportunity to make certain component changes to improve reliability and to ensure ease of replacement should the need arise. Housed in the new style picture frame case with steel reinforced surround, it embodies the results of air circulation studies resulting in an air flow which greatly assists cooling. The function and distribution of front panel controls including voltmeter and attenuation facilities remain unchanged.

All the original accessories, i.e., TM 6454 Attenuator Pad, TM 6222 Band-Pass Filter, and TM 6221 Unbalanced-to-Balanced Transformer are still suitable for use with this version.

MARCONI INSTRUMENTS

621.316.973

Earth Current Errors in High Frequency Measurements

by J. M. PARKYN

Errors in measurement will often occur when separate high frequency devices having a large gain or attenuation are interconnected. A typical example is the measurement of high attenuations by slideback against a standard. These errors are commonly caused by earth currents. The cause is discussed and a simple component is described which will generally effect a near perfect cure.

THERE ARE several areas of scientific measurement where possession of appropriate measuring instruments is not sufficient to ensure accuracy. Some knowledge of the many possible sources of error due to the measurement technique is also necessary. In the field of electronics the enormous power gains which are commonly used have lead to the use of very large attenuation ratios in measuring gear and considerable care is required to avoid power transfer through spurious paths. The measurement problem becomes particularly acute where these large signal ratios exist at high frequencies. Whilst no particular originality is claimed for the technique to be discussed a description in this publication for instrument users is appropriate since, although the problem arose recently in connection with the design of a signal generator[1] with internal unit construction, similar sources of error are prevalent when separate instruments are connected in cascade to form a test rig.

Spurious voltages can be developed in common earth impedances and interfere with the accuracy of measurement of small signal voltages. Also heavy stray fields can be produced with attendant errors in measurement.The trouble can become more pronounced at lower radio frequencies because in addition to the contact resistance of connectors, the current skin depth at these frequencies can approach the thickness of the coaxial cable outer. Earth return currents may then be divided between the coaxial cable and alternative earth paths, say the chassis connecting the individual units of an instrument, according to the ratio of impedances. At higher frequencies the earth return signal currents can flow entirely on the inner surface of coaxial cable outers and screening boxes. In a typical coaxial cable such as RG58C/U the skin depth becomes nearly equal to the screen thickness at 10 kc/s when about 37% of the signal current may flow on the outside of the cable. Solid copper pipes have been used as coaxial outers but this gives only a partial cure by providing a thicker wall of lower resistance.

A much more effective cure is to use a substantially greater length of standard flexible coaxial cable between the separate h.f. units with the excess length coiled to form about six turns around a ferrite core of high permeability as in Fig. 1. This solution gives basically a wide-band arrangement since the properties of the coaxial cable at high signal frequencies are quite unaffected by the coiled configuration. At lower frequencies the device acts as a unity ratio current balancing transformer to ensure exactly equal currents through the inner and outer conductors of the cable, thereby

This coaxial transformer is used in our Standards Laboratory when testing the attenuator of the new TF 2002 Signal Generator

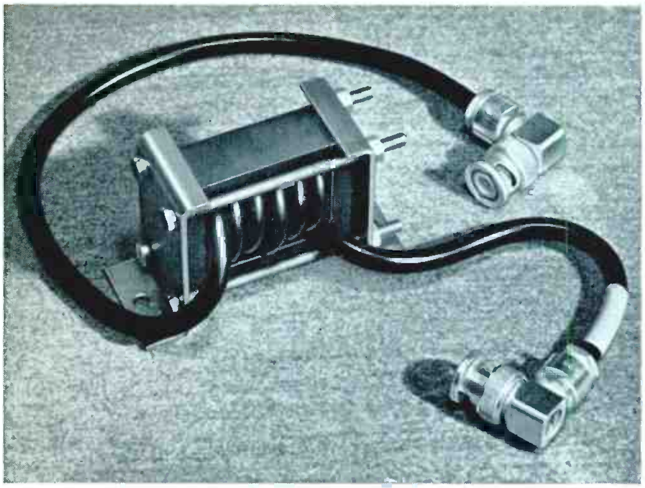
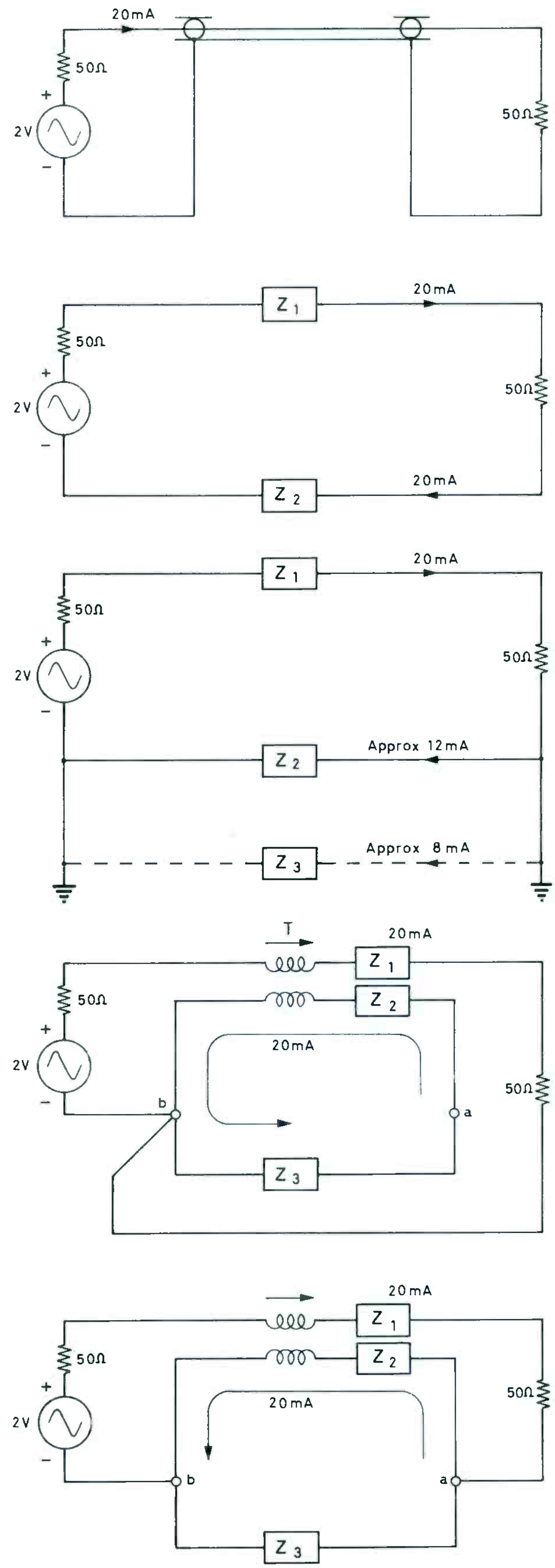


Fig. 1 (above). The coaxial transformer consists of six turns of cable on a ferrite core

Fig. 2 (right). Development of the theory of the coaxial transformer



preventing a part of the outer conductor current returning via multiple earth paths, i.e., the chassis or the earth wires in the mains supply leads of inter-connected instruments. Errors due to the spurious voltages which would be developed by earth currents are thus avoided. Because the "go" and "return" currents follow identical paths perfect field cancellation takes place. Without such a device serious stray h.f. radiation fields have been experienced, because a part of the earth return current has taken a different route from the current in the coaxial inner, to give an effective current flowing in a large loop.

There are several approaches to a detailed explanation of the working of this device. The diagrammatic treatment of a specific example which follows may be helpful. Referring to Fig. 2 (a) a 2 V e.m.f. with a 50 Ω source impedance feeds a 50 Ω load, which could be an attenuator, via a length of coaxial cable. Fig. 2 (b) represents this condition where the electrical length of the cable is short. Fig. 2 (c) shows the same system but with one side of both source and load earthed and a comparable impedance in the coaxial outer (Z_2) and earth return path (Z_3). In Fig. 2 (d) the coaxial cable has been formed into a perfect transformer of unity ratio and high inductance; this raises Z_2 to a high value causing the entire return current to flow via Z_3. For the purpose of explanation the earthy side of the load has been returned to point 'b' instead of 'a' so as to show that the transformer induces a current through Z_3 in the direction of b→a. Finally Fig. 2 (e) shows the true condition that exists with the load correctly returned to point 'a'. Under these conditions near perfect cancellation of the return current in Z_3, the earth impedance, takes place and all the return current flows via the transformer winding.

REFERENCE

1. Parkyn, J. M.: 'Transistorized m.f./h.f. a.m. signal generator type TF 2002', *Marconi Instrumentation*, September 1964, 9 p. 168.

MARCONI INSTRUMENTS NEW DESIGN

621.317.743

AUDIO FREQUENCY *Transmission Measuring Set* . . TYPE TF 2332

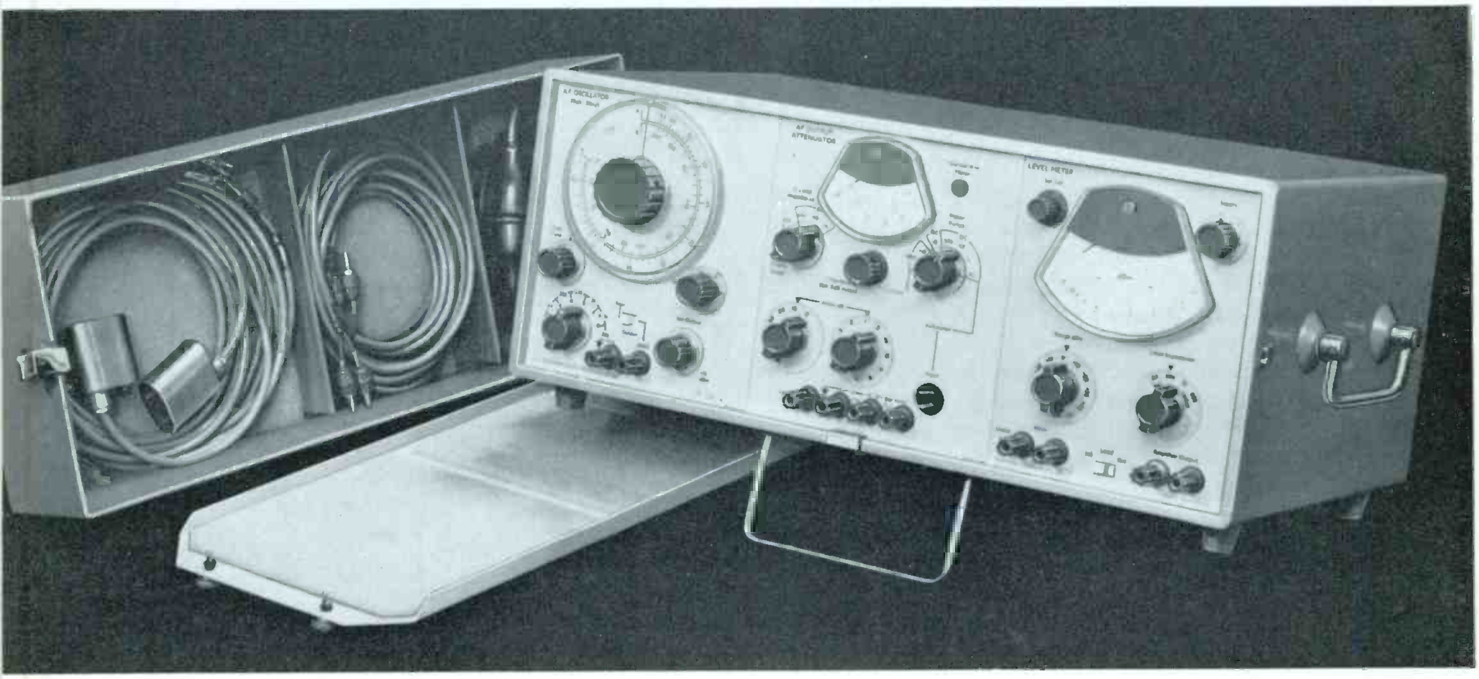
by
V. F. ARNOLD

Audio Frequency Transmission Measuring Set type TF 2332 consists of a monitored signal source and a level meter. A frequency range of 20 c/s to 20 kc/s is covered by the signal source with an output level range of +15 to –70 dBm. Impedances are 600 Ω, 150 Ω and 75 Ω switch-selected, for both signal source and level meter. Through-level measurements may be made or, alternatively, the level meter will provide the terminating load. Its range of power levels is from +25 to –70 dBm, and the frequency range is 20 c/s to 560 kc/s. A note is included on the technique of monitoring signal levels and tables are given comparing the instrument specification with C.C.I.T.T. recommendations.

IN A RECENT ISSUE of *Marconi Instrumentation*[1] a number of basic units in the new modular range of instruments were introduced. These basic units could be used either singly or in combinations to give several different instruments. One such combination formed the Transmission Measuring Set type TF 2333[2], an instrument which covers the requirements for measurements on telephone systems in the audio frequency range and for baseband measurements in multi-channel systems. The maximum output level of this test set is +3 dBm. However, for measurements on programme circuits in the frequency range 50 c/s to 20 kc/s it is necessary to make measurements at levels in the +10 dBm to +15 dBm range. In order to meet this requirement another combination of basic units has been introduced. This new combination is the Audio Frequency Transmission Measuring Set type TF 2332, which consists of an a.f. oscillator, a.f. output attenuator and level meter.

It has a maximum output level of +15 dBm at all impedances. This meets the requirements for programme circuits. The frequency discrimination is much better than that of the m.f. transmission measuring set since it covers a much smaller frequency band in the same number of ranges and also permits a better accuracy to be maintained. This feature is useful for measurements on filters and frequency responses of systems. In addition, the restricted frequency range permits a lower distortion transformer to be used and the distortion level

The audio frequency T.M.S. comprises an oscillator, a monitored attenuator and a level meter. It is supplied with splash-proof front panel lid which also provides neat stowage for a comprehensive set of measuring leads



of less than 0·1 % permits measurements to be made on low distortion systems.

Attenuator Unit

The attenuator units of both the m.f. and the a.f. transmission measuring sets have a 2 position meter range switch designated +0 and +10 dBm. Although the maximum output of the m.f. model is only +3 dBm it retains the higher range in order to economize on standard basic parts.

When the impedance switch is in any of the BAL positions, the output is derived from a transformer and may be 75, 150 or 600 Ω, according to the position of the impedance switch. The secondary winding of the transformer is floating but has a centre tap which can be earthed to give a balanced output. Alternatively, for unbalanced outputs, one side of the winding can be earthed. For 600 Ω operation only, an unbalanced output is available direct from the oscillator and not through the transformer, thus giving even lower distortion.

Monitoring System and Use of Attenuator

Because the a.f. oscillator has a 600 Ω output to enable it to be used as a self-contained unit, the monitoring system in the attenuator measures the voltage developed across the 600 Ω input impedance of the attenuator as shown in Fig. 1; this is a departure from our usual practice of monitoring source e.m.f. The voltage developed across the external load will depend on the value of attenuation switched in, on the frequency response of the transformer for balanced outputs, and the output impedance selected. Provided there is sufficient attenuation to isolate the monitor from the load, the monitor indication will be the same, irrespective of the load condition. This situation exists if 20 dB or more of attenuation is switched in. Under these conditions the monitor behaves in the same way as it would if it were monitoring the e.m.f. from a low impedance generator and the attenuator provides the 600 Ω source impedance which feeds the transformer.

In the special case where zero attenuation is used the monitor measures the voltage developed across the transformer primary when it is in circuit, or across the load directly in the 600 Ω unbalanced case. In either case, allowing for the step down of voltage for the 150 Ω and 75 Ω outputs, the monitor is effectively across the load as shown in Fig. 2. In this situation, if the oscillator output is set and thereafter not adjusted, and if the load has a reactive component and the generator frequency is varied, the monitor indication will vary with load impedance as the load sees the 600 Ω source impedance of the generator. On the other hand, if the monitor reading is kept constant, the load will appear to be fed from a zero source impedance. It is necessary, therefore, to take care where reactive loads are concerned and to use the monitor in the manner appropriate to the test being undertaken.

When high outputs are required with zero attenuation the oscillator provides the 600 Ω source impedance with an e.m.f. that is constant within ±0·3 dB over its complete frequency range.

When a reactive component is present in the load and it causes the impedance to fall below the nominal value of the load, there is clearly a limitation to the range of impedance values over which the voltage across the load can be maintained constant. It will depend on the value which it is required to develop across the load but ultimately the limit will be reached when the oscillator output is at its maximum.

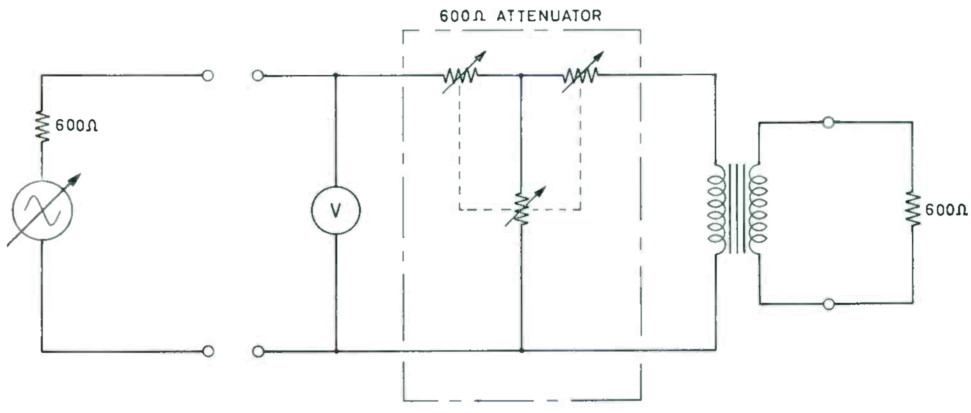


Fig. 1. The monitor indicates the oscillator voltage, which is independent of the load impedance if sufficient attenuation is switched in

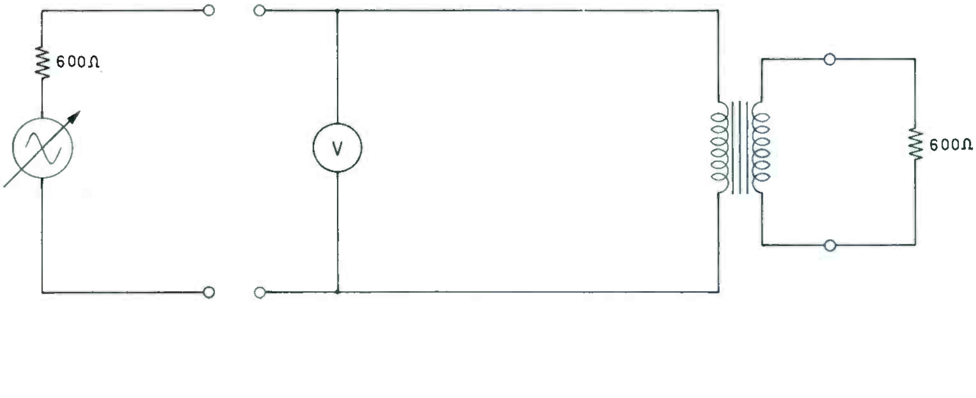


Fig. 2. In the special case of zero attenuation the voltage is monitored across the load

C.C.I.T.T. Recommendations

There are C.C.I.T.T. recommendations for telephone and programme circuits. Summaries of these recommendations for both generators and level meters are given in Table 1 together with the corresponding performance figures of the TF 2332.

REFERENCES

1. Cain, W. D., *et al*: 'Modular electronics instruments', *Marconi Instrumentation*, June, 1963, 9. p. 25.
2. Gribben, H. C.: 'Transmission measurements and the new Transmission Measuring Set type TF 2333', *Ibid*, September, 1963, 9. p. 56.

The G.P.O. International Exchange in Faraday Building, London. Transmission measuring sets play a vital part in maintaining the efficiency of international communications

(by courtesy of H.M. Postmaster-General).

TABLE 1

Comparison of C.C.I.T.T. recommendations with TF 2332 performance

	C.C.I.T.T. RECOMMENDATIONS		*TF 2332*
	Telephone circuits	*Programme circuits*	
Generators			
Frequency range	300 c/s to 3·4 kc/s	At least 50 c/s to 10 kc/s	20 c/s to 20 kc/s
Internal resistance	600 Ω	600 Ω or less than 30 Ω	600 Ω
Balance about earth	At least 52 dB	Output to be balanced about earth	At least 52 dB up to 10 kc/s with centre tap earthed
E.M.F. or level	E.M.F. at least 1·55V	+15 dBm into 600 Ω	8·5V e.m.f. or +15 dBm into 600 Ω
Frequency response with respect to 800 c/s	±0·2 dB	±0·2 dB	±0·2 dB
Harmonic distortion	Less than 2%	Less than 2%	Not greater than 0·1 %
Level Measuring Sets			
Frequency range	300 c/s to 3·4 kc/s	50 c/s to 10 kc/s	20 c/s to 20 kc/s
Input impedance	Greater than 20 kΩ or 600 Ω	Greater than 20 kΩ or 600 Ω	Greater than 30 kΩ or 600 Ω
Input balance	At least 52 dB	At least 52 dB	At least 52 dB on –10 dBm and lower ranges

ABRIDGED SPECIFICATION

Signal Source

FREQUENCY RANGE: 20 c/s to 20 kc/s in six ranges.

FREQUENCY ACCURACY: ±1%

OUTPUT: At least +15 dBm.

ATTENUATOR: 70 dB in 1 dB steps.

OUTPUT ACCURACY: ±1% of indicated dB value ±0·2 dB.

OUTPUT IMPEDANCE:
600 Ω unbalanced direct;
600 Ω, 150 Ω and 75 Ω, balanced or unbalanced.

DISTORTION: Less than 0·1% from 50 c/s to 20 kc/s.

OUTPUT METER RANGES: 0 dBm and +10 dBm (meter scaled ±6 dB on each range).

ACCURACY OF OUTPUT: Can be standardized against external standards; ±0·2 dB from 20 c/s to 20kc/s with respect to 1 kc/s.

Level Meter

FREQUENCY RANGE: 20 c/s to 560 kc/s.

LEVEL MEASUREMENT RANGES: +25 dBm to –70 dBm.

LEVEL ACCURACY: Standardized against external standard.

FREQUENCY RESPONSE: At 0 dBm relative to 1 kc/s, 50 c/s to 100 kc/s ±0·25 dB, 200 c/s to 20 kc/s ±0·1 dB.

RANGE SWITCHING ACCURACY: Up to 100 kc/s ±0·3 dB, from +20 dB to –50 dBm.

METER SCALE ACCURACY: ±0·1 dB per 1 dB increment relative to 0 dB between +5 and –5 dB.

TERMINATED INPUT RESISTANCE: Balanced, 600 Ω and 150 Ω; unbalanced, 600 Ω and 75 Ω.

BRIDGING INPUT RESISTANCE: Greater than 30 kΩ balanced; greater than 15 kΩ unbalanced.

REJECTION OF COMMON SIGNALS: At least 52 dB at 1 kc/s (on –10 dBm or lower ranges).

Summaries of Articles appearing in this issue

RESUME D'ARTICLES PUBLIES DANS LE PRESENT NUMERO

APPAREIL DE CONTRÔLE À BRUIT BLANC. TYPE OA 2090, POUR 2700 VOIES

On introduit la notion de contrôle de l'intermodulation à l'aide d'une source de bruit blanc, en décrivant d'abord succintement le système multiplex à répartition en fréquence couramment employé pour liaisons à voies multiples, soit par câble soit par faisceau hertzien. On explique pourquoi et comment le bruit blanc est employé pour mesurer l'intermodulation. Dans le cadre des recommendations des comités internationaux CCIR et CCITT, on considère les conditions à remplir par un tel appareil de mesure. On décrit alors, du point de vue de l'usager plutôt que de celui du montage, le nouvel appareil transistorisé de contrôle par bruit blanc, qui est capable de contrôler les systèmes avec un nombre de voies s'étendant de 12 à 2700.

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MESURE DE L'AFFAIBLISSEMENT D'UN CABLE COAXIAL

On propose une méthode pour mesurer l'affaiblissement d'un tronçon de câble coaxial à l'aide seulement d'un générateur de signaux et d'un voltmètre électronique.

Page 192

UN FILTRE PIÉZO-ELECTRIQUE POUR 100 kHz

On décrit la réalisation d'un filtre piézo-électrique extrêmement sélectif, avec une bande passante de 6 Hz centrée sur 100 kHz. La caractéristique affaiblissement-fréquence est approximativement du type dit "de Butterworth", et elle est obtenue comme suit: les cristaux piezoélectriques sont représentés comme éléments série d'un montage hypothétique équivalent au filtre, et les autres éléments de ce montage sont alors calculés en faisant la synthèse de la fonction de transfert désirée, à partir de la position des pôles dans la représentation de ce montage dans le domaine des fréquences complexes.

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MESURE DES IMPÉDANCES DANS LES CIRCUITS COMPLIQUÉS AVEC UN PONT UNIVERSEL

Dans la pulpart des laboratoires on trouve un pont universel de mesure d'impédances, tel que le modèle TF 1313 de la société Marconi Instruments. L'emploi de ces appareils pour mesurer les composants va de soi, mais on ne se rend pas toujours compte de la possibilité de faire des mesures dans les cas plus compliqués. Cet article explique comment on peut mesurer un montage de trois capacitiés en triangle, même lorsque l'un des sommets est inévitablement raccordé à la terre.

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ERREURS DANS LE MESURES EN HAUTES FRÉQUENCES, DUES AUX COURANTS DE TERRE

Lorsque plusieurs appareils en haute fréquence ayant soit un gain soit un affaiblissement importants sont branchés l'un sur l'autre, des erreurs de mesure interviennent souvent. Un éxemple typique est la mesure d'affaiblissements importants par comparaison directe avec un atténuateur standard. Ces erreurs sont généralement dues à des courants de terre. On explique la cause de ce phénomène et on décrit un composant très simple qui parvient généralement à éliminer ces erreurs presque complètement.

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DÉCIBELMÈTRE B.F. MODÈLE TF 2332

Le décibelmètre B.F. modèle TF 2332 se compose d'un oscillateur avec contrôle de la puissance de sortie et d'un indicateur de niveau. La plage de fréquences de l'oscillateur s'étend de 20 Hz à 20 kHz, et sa gamme de puissances de +15 à –70 dBm. L'impédance de l'oscillateur et de l'indicateur de niveau peut être, au choix, 600, 150 ou 75 Ohm, en montage équilibré ou non. L'indicateur de niveau peut être branché pour mesures "en passant" mais il peut aussi lui-même faire office de charge. Il est prévu pour la plage de fréquences de 20 Hz à 560 kHz et il est calibré de +25 à –70 dBm. On explique la technique de la mesure de niveaux de transmission, et on donne des tables comparant les caractéristiques de l'appareil décrit avec les recommendations du CCITT.

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ZUSAMMENFASSUNG DER IN DIESER NUMMER ERSCHEINENDEN BEITRÄGE

EIN 2700-KANAL INTERMODULATIONSMESSGERÄT TYPE OA 2090

Einer Behandlung der Intermodulationsmessung mit weißem Rauschen geht eine kurze Beschreibung der bei Koaxial—und Richtfunkverbindungen verwendeten Frequenzmultiplexmethode voraus. Die Gründe für die Benutzung von weißem Rauschen werden angegeben und die bei der Durchführung einer Intermodulationsmessung angewendete Methode wird erklärt. Die an ein solches Meßgerät zu stellenden Anforderungen werden in einem Vergleich mit den CCIR- und CCITT-Empfehlungen behandelt. Das neue transistorisierte Intermodulationsmeßgerät, welches Anlagen für 12 bis 2700 Kanäle prüfen kann, wird mehr vom Standpunkt des Benutzers als vom schaltungstechnischen Standpunkt aus beschrieben.

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DIE MESSUNG DER DÄMPFUNG VON KOAXIALKABELN

Hier wird beschrieben, wie die Dämpfung eines Koaxialkabels unter Verwendung von nur einem Meßsender und einem elektronischen Voltmeter gemessen werden kann.

Seite 192

EIN 100-kHz-QUARZFILTER

In diesem Aufsatz wird ein stark selektives Kristallfilter mit einer Bandbreite von 6 Hz bei einer Mittenfrequenz von 100 kHz beschrieben. Eine ungefähr maximal geebnete Durchlaßkurve wird dadurch erhalten, daß die Kristalle als Serienresonanzkreise in einem hypothetischen Filtermodell behandelt werden und dann die übrigen Schaltungskonstanten durch eine Synthese der gewünschten Übertragungsfunktion aus der Lage der Pole für die Schaltung in der komplexen Frequenzebene bestimmt werden.

Seite 195

EINE UNIVERSAL-MESSBRÜCKE ZUR AUSMESSUNG KOMPLIZIERTER SCHALTUNGEN

Viele Laboratorien besitzen eine Universal-Meßbrücke wie die Ausführung Type TF 1313 der Firma Marconi Instruments. Obwohl die Benutzung eines solchen Gerätes zur Messung von Werten von einzelnen Schaltungselementen selbstverständlich ist, dürfte dessen Anwendung zur Durchführung komplizierterer Messungen nicht überall bekannt sein. Dieser Aufsatz befaßt sich mit der Messung einer Dreieckschaltung aus Kondensatoren, wobei ein Punkt immer geerdet ist.

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FEHLER DURCH ERDSTRÖME BEI HOCHFREQUENZ-MESSUNGEN

Häufig entstehen Meßfehler, wenn einzelne mit hoher Verstärkung oder Dämpfung arbeitende Hochfrequenzgeräte untereinander verbunden werden. Ein typisches Beispiel ist die Messung hoher Dämpfungswerte mittels Kompensation durch ein Dämpfungsnormal. Dabei treten gewöhnlich Fehler durch Erdströme auf. Die Ursache dafür wird besprochen und eine einfache Schaltung beschrieben, mit der sich im allgemeinen eine fast vollkommene Abhilfe schaffen läßt.

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TONFREQUENZ-DÄMPFUNGMESSGERÄT TYPE TF 2332

Das Tonfrequenz-Dämpfungsmeßgerät Type TF 2332 besteht aus einer mit einem Kontrollinstrument versehenen Signalquelle und einem Pegelmesser. Die Signalquelle hat einen Frequenzbereich von 20 Hz bis 20 kHz bei Ausgangspegeln von +15 bis –70 dBm. Sowohl die Signalquelle als auch der Pegelmesser besitzen symmetrische oder unsymmetrische Anschlüsse mit 600, 150 und 75 Ohm. Die Pegelmessungen können als Durchgangsmessungen durchgeführt werden. Andererseits kann der Pegelmesser auch als Abschluß verwendet werden. Sein Pegelmeßbereich erstreckt sich von +25 bis –70 dBm und der Frequenzbereich von 20 Hz bis 560 kHz. Die Methode zur Messug von Signalpegeln wird kurz behandelt und Tabellen für einen Vergleich der Gerätedaten mit den CCITT-Empfehlungen sind beigefügt.

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SOMMARIO DEGLI ARTICOLI PUBBLICATI IN QUESTO NUMERO

UN COMPLESSO DI MISURA CON RUMORE BIANCO PER 2700 CANALI, IL TIPO OA 2090

Si introduce l'argomento del metodo a rumore bianco per la misura dell'intermodulazione tramite una breve descrizione del sistema di multiplex a divisione di frequenza impiegato con linee in cavo e ponti radio a più canali. Si adducono le ragioni per l'uso del rumore bianco, e si spiega come venga effettuata una misura di intermodulazione. I requisiti da soddisfare in sede di progetto di un tale complesso di misura sono considerati in relazione alle raccomandazioni dei comitati internazionali C.C.I.R. e C.C.I.T.T. Il nuovo complesso di misura con rumore bianco, transistorizzato, idoneo per prove su impianti con capacità da 12 a 2700 canali, viene descritto piuttosto dal punto di vista di chi intenda servirsene che nelle particolarità circuitali. Pagina 187

MISURA DELL'ATTENUAZIONE DI CAVI COASSIALI

Si descrive un metodo che consente di misurare l'attenuazione di un tratto di cavo coassiale usando soltanto un generatore di segnali ed un voltmetro elettronico. Pagina 192

UN FILTRO PIEZOELETTRICO DA 100 kHz

Si descrive la progettazione di un filtro piezoelettrico a selettività elevata, con frequenza centrale di 100 kHz e banda passante di 6 Hz. Una curva di risposta all'incirca massimamente piatta viene ottenuta trattando i cristalli piezoelettrici come elementi di risonanza in serie entro una ipotetica struttura filtrante, e calcolando poi tutte le altre costanti circuitali col sintetizzare la funzione di transmissione desiderata a partire dalle posizioni dei poli del circuito nel piano della frequenza complessa. Pagina 195

STUDIO DI RETI COMPLESSE CON UN PONTE UNIVERSALE DI IMPEDENZE

Molti laboratori dispongono di ponti universali di impedenze quali il tipo TF 1313 della Marconi Instruments, e, mentre l'impiego di questi strumenti per la misura dei valori di parti staccate è ovvio, non ne viene sempre apprezzato in pieno l'uso per misure più complesse. Questo articolo tratta della misura di una rete a delta costituita da condensatori, un punto della quale è permanentemente collegato a terra. Pagina 200

ERRORI INTRODOTTI DA CORRENTI DI TERRA IN MISURE AD ALTA FREQUENZA

Quando apparati ad alta frequenza dotati di un grado elevato di guadagno o di attenuazione sono collegati insieme, essendo fisicamente separati, si verificano spesso errori di misura. Ne è un esempio tipico la misura di forti attenuazioni mediante compensazione rispetto ad un campione. Questi errori risultano comunemente da correnti di terra. Se ne discute la cagione, e si descrive un semplice componente che in generale rappresenta un rimedio quasi perfetto. Pagina 204

COMPLESSO PER MISURE DI TRASMISSIONE AD AUDIOFREQUENZA TIPO TF 2332

Il complesso per misure di trasmissione ad audiofrequenza tipo TF 2332 si compone di una sorgente di segnali controllata e di un misuratore di livello. La sorgente di segnali copre un campo di frequenze da 20 Hz a 20 kHz con un livello di uscita variabile da +15 a −70 dBm. Le impedenze, sia della sorgente di segnali che del misuratore di livello, sono di 600, 150 e 75 Ω bilanciati o sbilanciati. Si possono effettuare misure di segnali di passaggio, od alternativamente il misuratore di livello può fornire il carico di chiusura. I livelli di potenza ammissibili vanno da +25 a −70 dBm, entro un campo di frequenze da 20 Hz a 560 kHz. L'articolo include una nota sulla tecnica dell'osservazione di livelli di segnali, come pure tabelle in cui i dati funzionali dello strumento sono posti in confronto con le raccomandazoni del CCITT.

Pagina 206

RESUMENES DE ARTICULOS QUE APARECEN EN ESTE NUMERO

EQUIPO DE PRUEBA DE RUIDO BLANCO, EN 2.700 CANALES, TIPO OA 2090

Se da a conocer la cuestión del método de ruido blanco para pruebas de intermodulación, con una breve descripción del sistema múltiplex de división de frecuencia, empleado en radioenlaces multicanales y por cable. Se explican las razones del empleo del ruido blanco, así como el procedimiento para medidas de intermodulación. Se considera también la necesidad de dicho equipo de prueba en relación con las recomendaciones de los Comités Internacionales C.C.I.R. y C.C.I.T.T. Se describe—desde el punto de vista de los usuarios más que respecto a detalles de circuitos—el nuevo equipo de prueba de ruido blanco, transistorizado, que puede efectuar pruebas de sistemas de 12 a 2.700 canales.

Página 187

MEDIDA DE ATENUACION EN CABLES COAXIALES

Se explica un procedimiento para medir la atenuación en cables coaxiales, utilizando sólo un generador de señal y un voltímetro.

Página 192

FILTRO A CRISTAL DE 100 kHz

Se describe un filtro a cristal de alta selectividad, centrado en 100 kHz, y con una anchura de banda de 6 Hz. Se obtiene una respuesta plana máxima, aproximada, tratando los cristales como elementos resonantes en un hipotético modelo de filtro; después se valora el resto de las constantes del circuito, sintonizando la función de transferencia deseada en la posición de los polos del circuito, en el plano complejo de frecuencia.

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EL PUENTE DE IMPEDANCIA UNIVERSAL VALORA REDES COMPLEJAS

En muchos laboratorios existen puentes de impedancia, universales, tales como los Marconi Instruments tipo TF 1313 y, aunque es conocido el uso de estos aparatos para medir valores de componentes, no se reconoce siempre su utilidad para medidas más complejas. Este artículo trata de la medida de una red delta, de condensadores, uno de cuyos puntos está siempre conectado a tierra.

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ERRORES EN MEDIDAS DE A.F. DEBIDOS A LAS CORRIENTES TELURICAS

Se producen, frecuentemente, errores al efectuar estas medidas, cuando se conectan instrumentos de A.F. con atenuación o ganancia grandes. Un ejemplo típico es la medida de atenuaciones grandes por comparación con las normas. Estos errores los producen, corrientemente, las corrientes telúricas. Se estudian sus causas, y se describe un sencillo componente, cuyo empleo es, generalmente, un remedio eficaz.

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MEDIDOR DE TRANSMISION EN A.F. TIPO TF 2332

Este equipo consta de una fuente de señal controlada y un medidor de nivel. El margen de frecuencia, de 20 Hz a 20 kHz, está cubierto por la fuente de señal, con un nivel de salida de +15 a −70 dB. Las impedancias son 600, 150 y 75 ohms. equilibradas o no, tanto en la fuente de señal como en el medidor de nivel. Pueden efectuarse medidas a través del medidor de nivel, o, alternativamente, el medidor de nivel puede simular carga de terminal. El margen de medida de niveles de potencia es de +25 a −70 dBm, para frecuencias de 20 Hz a 560 kHz. Se incluye una nota sobre la técnica de niveles de señal controlados, así como unas tablas comparativas de las características del instrumento con las recomendaciones del C.C.I.T.T.

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