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MARCONI INSTRUMENTATION

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MARCONI INSTRUMENTS LIMITED

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ENGLAND

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Specialization

THE TERM 'specialist' has in this post-war era acquired a disreputable connotation and, as defined by one cynic, is a person who knows more and more about less and less. A notable exception to this occurs in medical circles however, where to be a specialist is still to be revered by the layman. With the rapid expansion of the various branches of technology specialization is unavoidable, and the graduate engineer soon finds that even though he may have been equally instructed in such subjects as the theory of structures, heat engines, or even semiconductors, specialization must commence immediately he leaves his academic confines. The days when a man could be a 'scientist', so that it was possible for the laws of electromagnetic induction to be formulated by a professor of chemistry, are definitely over.

If specialization is to be deprecated then we at Marconi Instruments are more fortunate than most, as our particular field of endeavour — namely measurements — brings us into contact with the majority of the numerous aspects of electronics. A cursory glance through this and previous issues of *Instrumentation* will show how diverse are our interests, but inevitably, in common with all large organizations, specialist teams are necessary for the design of our major product lines. This not only ensures continuity with each successive design, but enables the engineer, by concentrating his effort, to learn and possibly solve some of the problems confronting the users of our instruments.

Logically this team spirit has now been extended to our Production Division in the form of product lines to concentrate the various skills required for different instruments at each stage in their growth. Assembling an oscilloscope, for example, obviously requires different techniques from those needed by a precision line oscillator; likewise the wiring of a $\frac{1}{4}\%$ bridge presents different problems from a pulse generator. Most important of all in the calibration and test stage, familiarity with all the details of a particular series of designs is essential. We are confident that with this specialization at all stages through design to production a greater consistency of performance in every detail will be obtained from our products.

The two main articles in this issue deal with the specialized technique of measuring radiated power as a precaution against dangerous dose rates. This is a relatively new hazard to health which, with modern high-power ground and airborne radar, has become sufficiently serious for regulations to be laid down for the protection of operators and public alike.

R.F. heating of body tissues has, of course, long been turned to advantage in the form of diathermy, but the possibility of adverse effects of radiation began to be suspected in the early days of radar. Although unproved scientifically for several years we now know that the thermal effect of microwave radiation can indeed cause serious bodily damage. This may take the form of a dangerous rise in body temperature or selective heating that can result in irreversible damage such as cataract of the eye. We know that the percentage of airborne radiation converted into heat generally increases with frequency, although there are unpredictable effects such as standing waves or cavity resonances to be considered; at the lower radar frequencies the heat is largely absorbed in the deeper tissues, and is therefore particularly dangerous since it is unlikely to be detected by the body's sensing elements which lie mostly at the surface.

Specialization is not confined only to modern industry but has existed since man first settled to live in communities. In this electronic age an ancient craft like thatching is still much in demand. Here is a skilled thatcher at work on an old-world cottage in Dorset, but no doubt this traditional roof will not be complete until a contemporary aerial array has been erected on the chimney stack by yet another specialist

The maximum safe continuous dosage at any frequency from 30 Mc/s to 30 Gc/s is widely accepted as 10 mW/cm². This figure assumes complete absorption by the body and that no more than half the body surface can be irradiated from one source; when averaged over the total body surface, this power in biological terms represents 100% overload of the body's natural heat-dissipating mechanism, but this is considered well within its capabilities. In engineering terms a dose of 10 mW/cm² is received about 30 ft from an S Band transmitter radiating a mean power of 10 kW from an open ended waveguide (see *Electronics Weekly*, April 18th, 1962); this shows that the extent of the hazard is not merely restricted to such imprudent behaviour as actually peering down the waveguide.

As far as is known the effect of r.f. radiation on the body is entirely thermal; one way of solving the measurement problem would therefore appear to be the observation of the temperature rise in a sample of human tissue, but this is hardly practical and, in any case, involves a material that is rather outside the field of electronic instrumentation. An early device called the Richardson Dosimeter did, in fact, use this principle, but using a blob of gelatine to simulate the tissue; although capable of accurate results, it was rather sluggish and sensitive to ambient temperature. The new Marconi Instruments Power Meter Type TF 1396A described on p. 147 is an all-electronic device using thermistors as heat-sensing elements, and provides a practical means of taking precautions against this latest hazard to health.

In contrast to this new technique of radiation measurement, one of our major specializations has always been the design of signal generators, and over the years there has been a continuous improvement in performance to match the advances made in receiver design. Through this long experience we have of necessity come to learn some of the problems associated with the design and production of receivers. In this respect we are familiar with the standard test conditions, specified by the British Post Office, for checking receiver selectivity involving the application of two simultaneous signals of slightly different frequencies to the input. Looking at this type of test from a signal generator viewpoint the input arrangements to the receiver appear to be extravagant in output level requirements. We would therefore like to draw special attention to p. 156 of this issue where we have taken the unusual step in this short article of posing a question, and we would ask receiver experts to consider the alternative arrangements which are suggested there.

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MARCONI INSTRUMENTS

R.F. Radiation Power Meter . . . *TYPE TF 1396A*

by S. G. PRITT

This instrument has been designed to meet the principal recommendations of the British Post Office relating to the measurement of intense radio frequency radiations, which have possible harmful effects upon the human body. It consists of a thermistor power meter suitable for use between 10 Mc/s and 10 Gc/s and is intended to be fed from aerials and attenuators designed for customers' specific frequency requirements. When used with such, the Meter is capable of indicating radiation intensity over the range of 1 to 20 mW per sq. cm with an error which does not exceed ±2 dB.

WITH THE ADVENT of modern high powered radar, radio and television transmitting equipment concern has been expressed in some quarters regarding the possible harmful effects of intense r.f. radiation on the human body. In consultation with the many interested bodies—the Medical Research Council being but one of them—the General Post Office issued, in 1960, a publication *Safety Precautions Relating to Intense Radio-Frequency Radiation.** This sets out a code of good practice intended to safeguard members of the public, operating and maintenance personnel on transmitter sites, and research, experimental and test personnel in establishments and factories. The publication also includes a general specification of radiation intensity measuring equipment. It is this specification which forms the basis of design of Marconi Instruments' R.F. Radiation Power Meter, Type TF 1396A.

Fig. 1. A Radiation Power Meter in operation with S Band aerial

General Design Considerations

It is recommended that the human body should never be subjected to a continuous daily exposure from an electromagnetic field of greater intensity than 10 mW per sq. cm. This level of radiation does in fact allow for a considerable safety factor but nevertheless is generally accepted as the upper limit of safe intensity, and in this instrument it corresponds to half scale deflection of the meter. High radiation levels cause damage to human tissue in the first instance by direct heating. The eyes are particularly vulnerable organs in this respect in that they possess no temperature sensitivity nor are they subject to the human refrigeration process. Methods have been devised which measure the field intensity in terms of the temperature rise in a radiation absorbent material but the time constant of such devices are normally too long when several readings have to be taken on site. A rise and decay time constant of one second for a radiation meter is accepted as being an accurate simulation of the human tissue thermal time constant.

Rather than make temperature rise measurements it is quicker and more convenient to make use of a d.c. meter and an r.f. to d.c. transducer which in the TF 1396A takes the form of a pair of thermistors. These have been chosen in preference to a crystal detector as their output is always a function of the average input power and they are mechanically stable. The d.c. output from a crystal diode on the other hand follows, to some extent, the peak r.f. level so that amplitude modulation tends to produce an output which is not proportional to r.m.s. voltage or power. The characteristics of the crystal diode are also subject to changes caused by variations of temperature and vibration. The diode is very sensitive but the

* H.M. Stationery Office, Code No. 43–182.

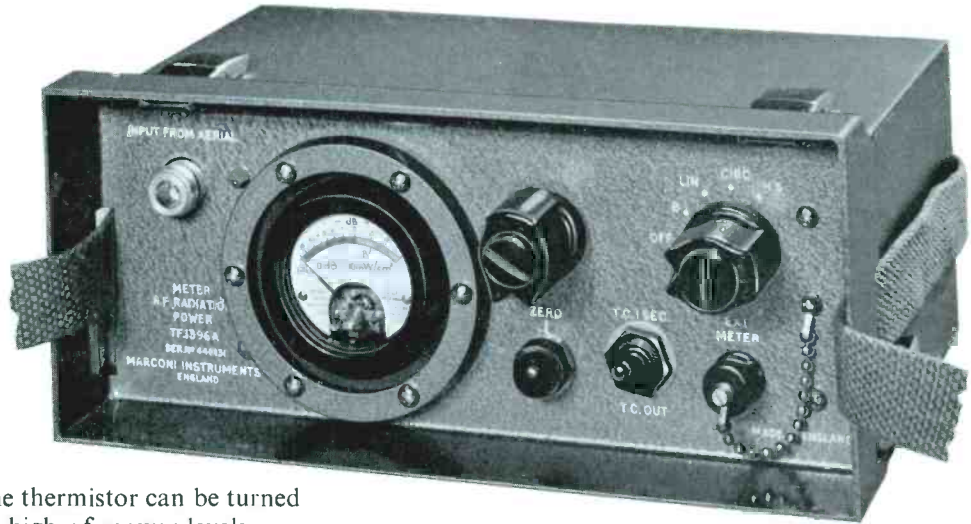


Fig. 2.
Marconi Instruments R.F. Radiation Power Meter, Type TF 1396A

comparative insensitivity of the thermistor can be turned to advantage when measuring high r.f. power levels.

In order that the thermistors should give a true indication of r.f. power it has to be arranged that they constitute a matched load across a waveguide or coaxial transmission system. Since any particular waveguide only operates over a limited frequency band, the TF 1396A transmission system of necessity had to be a coaxial line and the thermistors are therefore installed in a 50 Ω coaxial thermistor mount inside the type 'N' connector in the instruments.

The fundamental concept for the complete Monitor was one of a sectionalized design—aerial, transmission and power measurement. This arises because of the necessity of providing a calibration and test procedure which can be executed in the factory using v.s.w.r. and insertion loss measurements.

To transfer electromagnetic energy from the incident field to the Meter an aerial is required which presents a constant aperture to all frequencies in the band to which it is sensitive. Since it is not possible to design an aerial whose effective aperture, when working into a truly matched coaxial system, is small enough for the induced power to be fed direct into the thermistor mount (in the present instrument an incident field of 10 mW per sq. cm corresponds to an input of 5 mW to the thermistors) it is necessary to use an attenuator between the two. It is convenient in this application to use 'lossy' coaxial cable, of predetermined length, corresponding to the power loss required, for this attenuation element.

General Description

Fig. 1 shows the Radiation Power Meter in its operational position, slung by a strap from the shoulder of the operator, whose hands are then free to use the controls and orientate the aerial. A detachable aerial is chosen to suit the particular frequency band in which the incident r.f. field is known to be propagated.

The aerial is connected by a comparatively loss free flexible cable to the type 'N' socket on the control panel. This socket is in fact part of an aerial coupling unit comprising the attenuator and cable frequency characteristic correction filter. This coupling unit is housed within the protection of the main instrument case and is also changed to suit the particular aerial in use. To change the frequency band of operation it is only necessary to open the quick release fasteners of the case, change the coupling unit and having replaced the case reconnect the appropriate aerial.

The Power Meter, and as such it may be used in isolation, is calibrated over the range of −10 to +3 dB relative to 10 mW per sq. cm. (i.e. 1 to 20 mW per sq. cm), this is shown in Fig. 2. This calibration holds for both vertically and horizontally polarized incident fields by suitably rotating the aerial about its horizontal axis. Another position of the function selector switch doubles the sensitivity of the Monitor to permit the linearly polarized aerials to be used in circularly polarized fields, wherein their excitation is reduced by 3 dB. A third position of the switch provides an increase in sensitivity of approximately 6 dB relative to that obtained in the normal position when measuring linearly polarized fields. This additional sensitivity is useful for search tests prior to taking accurate measurements. This facility provides the opportunity to detect fields of lower intensity.

For use in the open, on transmitter sites, such an instrument as this must be independent of mains supplies, light-weight, portable and capable of withstanding rain showers. To this end it is battery operated and position 'B' of the function switch provides a Check Battery condition with normal current drain. The 'B' calibration on the meter indicates the lower useful level of battery voltage. The instrument is lightly yet robustly constructed, is provided with control panel guard fence, and the whole unit is shower proofed by suitably constructed control spindles, meter and rubber sealed removable case.

Inside the outer case there is a second clip-on cover to protect the internal wiring, components and batteries. The latter are mercury dry cells, having an indefinite shelf life and an expected minimum life of 24 hours on continuous operation.

The control knob, other than that for the function switch, is the fine ZERO setting control whilst the removable cap beneath provides protection for the preset coarse ZERO control.

The toggle switch enables the one second time constant to be cut out. This permits a fast meter response to help the operator to orientate the aerial to the maximum output position.

The jack at the right hand end of the panel is provided to enable measurements of incident field strength to be made at a location remote from the aerial by plugging in an external meter. To facilitate this the aerials are provided with a ¼ B.S.W. tapped hole so that they may be mounted on to a camera tripod.

Thermistor Mount

Since the thermistors measure input r.f. power the design of the thermistor mount must be such that the thermistor beads are arranged to present a non-reactive load to the coaxial transmission line. In order to attain the least reactance it is essential that the thermistors should be as small as possible and the TF 1396A uses two beads which consist of a spheroid of semi-conductor material approximately 0·010 inch diameter with two diametrically opposite connecting wires 0·001 inch diameter. Fig. 3 shows

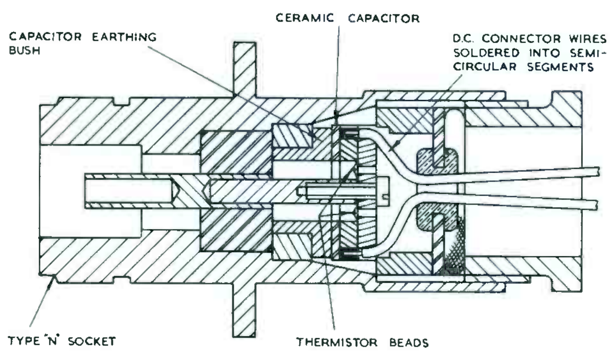


Fig. 3. Thermistor mount

how it has been arranged that these beads are presented as a load across the inner and outer conductors of the coaxial connectors with as short a path length as possible.

Our readers will be familiar with the thermistor Wheatstone Bridge technique of biasing the thermistors so that their resistances in parallel present the correct load impedance (50 Ω) to the line. The bridge is balanced in this condition, then any additional heating by incoming r.f. power produces a further lowering of thermistor resistance which in turn causes out of balance current to flow through the meter whose scale can be calibrated in r.f. power.

At room temperature the resistance of the beads is in the order of 1700 Ω and a current of approximately 14 mA is required to reduce this to 100 Ω. It is convenient in this instance to supply this as d.c., so provision is made in the mount for this current to be fed into the thermistors (which are in series to d.c.) without interfering with the r.f. characteristics. In assembly the thermistors are selected as pairs that require equal current to bring their resistances to 100 Ω; this ensures a balanced resistive load to the r.f. and contributes to a good v.s.w.r. for the mount.

In order to achieve the condition of having the thermistors in series to d.c. (as one arm of the bridge) and in parallel to r.f., it will be observed that the construction illustrated in Fig. 3 provides the necessary d.c. connections direct to the 'earthy' ends of the thermistors. These are d.c. blocked to the outer of the coaxial line by the

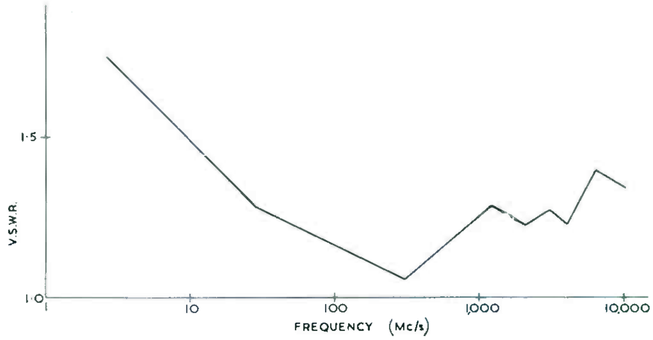


Fig. 4. Typical v.s.w.r./frequency characteristic of thermistor mount

presence of capacitors, each of which are of 5000 pF and thus ensure a low reactance earth path for r.f. currents down to 10 Mc/s. The capacitors are in fact specially made as a single high permittivity ceramic washer bonded on one face to an earthing bush and on the opposite face to two semi-circular segments which provide connections to the outer ends of the thermistors and the incoming d.c. leads. This construction also makes it unnecessary for any external connections to be made to the centre coaxial conductor which would certainly introduce an impedance mismatch. Fig. 4 shows a plot of the average v.s.w.r. over the range 10 Mc/s to 10,000 Mc/s for this design of thermistor mount on a 50 Ω transmission system.

Overload tests have shown that no damage is sustained by the thermistor when connected to a source of available power 20 times greater than the normal r.f. input (i.e. 100 mW—corresponding to an incident field of 200 mW per sq. cm). This arises to some extent as a result of the rapid decrease in resistance of the thermistors as more power is dissipated in them, with the consequent mismatch to the higher power source.

The Thermistor Bridge and its Supplies

Due to the instability of thermistors when fed from a constant voltage source it is necessary to provide a source of constant current. Of the various means of effecting this the most economic and convenient way is to make the thermistor bridge the collector resistance of a transistor. This is shown in Fig. 5 which illustrates the basic circuit diagram of the TF 1396A R.F. Radiation Power Meter. To reduce the effects of battery voltage fluctuations the base/emitter potential is stabilized by the Zener diode MR1 and the bridge current is set for null deflection on the meter (in the appropriate positions of the function selector) by coarse and fine ZERO controls RV1 and RV2.

R1, R2 and RV3 constitute the fixed arms of the Wheatstone Bridge but RV3 is made adjustable in order to optimize the resistance of the thermistors to maintain the best impedance match for the coaxial line. This is set up on test, together with RV4 and RV5 which adjust the meter sensitivity to correspond with the appropriate bridge power sensitivity.

The recommended rise and decay time constant of 1 sec is provided by the capacitor C1 in conjunction with the resistance of the meter.

On a portable instrument subject to intermittent operation, where the bridge null can be checked quickly each time the instrument is switched on, it was considered that the effect of variations of ambient temperature on the thermistor resistance could be compensated by manual setting of the fine ZERO control. In the interests of economy and battery consumption therefore no attempt has been made to include automatic ambient temperature compensation in this instrument.

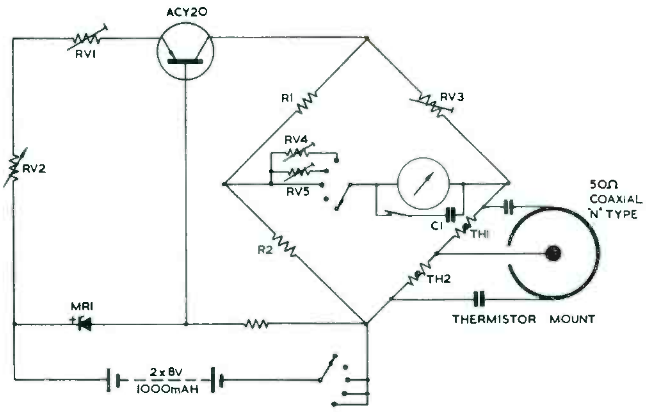


Fig. 5. Functional diagram of R.F. Radiation Power Meter, Type TF 1396A

The considerations in the design of the aerials for use with this Monitor is the subject of the following article.

ABRIDGED SPECIFICATION

Sensitivity

Meter scale calibrated from −10 to +3 dB relative to 10 mW/cm². A function selector switch enables this scale to be used for both linearly and circularly polarized radiation by suitably altering the meter sensitivity (by 3 dB). Provision is also made for a 6 dB increase in sensitivity.

Time constant

Meter readings are normally subject to a 1 second time constant. When required, a cutout switch enables the time-constant circuit to be by-passed.

Accuracy

Within −0 and +2 dB at power levels in the range 5 to 20 mW/cm².

Input impedance

50 Ω, with a v.s.w.r. better than 1·5 : 1 at frequencies between 10 Mc/s and 10,000 Mc/s.

Aerials

Detachable aerials are provided to cover frequency ranges as required.

External meter

A jack socket enables an external 25 μA meter to be connected for remote reading of high-intensity radiation. When used in this manner the aerials may be mounted on a tripod with suitable ¼ in. B.S.W. fittings.

Power supply

Two self-contained mercury batteries; each 8 V, 1,000 mAh.

621.396 677.1

Aerial Design for Power Measurement

by R. SMITH and
S. T. ROBERTS,
Graduate I.E.E.

A range of aerials for use with the TF 1396A is described. After consideration of the requirements for such aerials, a description is given of sectoral horn aerials for the S and X Band, and a discone type for the L Band. Electrical and mechanical design problems are discussed, and performance examined.

THIS ARTICLE deals with a range of aerials for use with the TF 1396A R.F. Radiation Power Meter, described on page 147. The aerials together with the TF 1396A form a complete instrument for the detection and measurement of hazardous r.f. fields.

To meet existing customer requirements X, S, and L Band aerials have been developed to cover the frequency ranges 8·73 to 10·67 Gc/s, 2·7 to 3·3 Gc/s, and 700 to 1400 Mc/s respectively.

DESIGN CONSIDERATIONS

Listed below are the qualities that an ideal aerial for radiation measurements should possess.

1. Capability of operating over a wide frequency range without tuning.
2. Constant output over the frequency range, i.e. constant effective receiving aperture, to avoid the use of correcting networks.

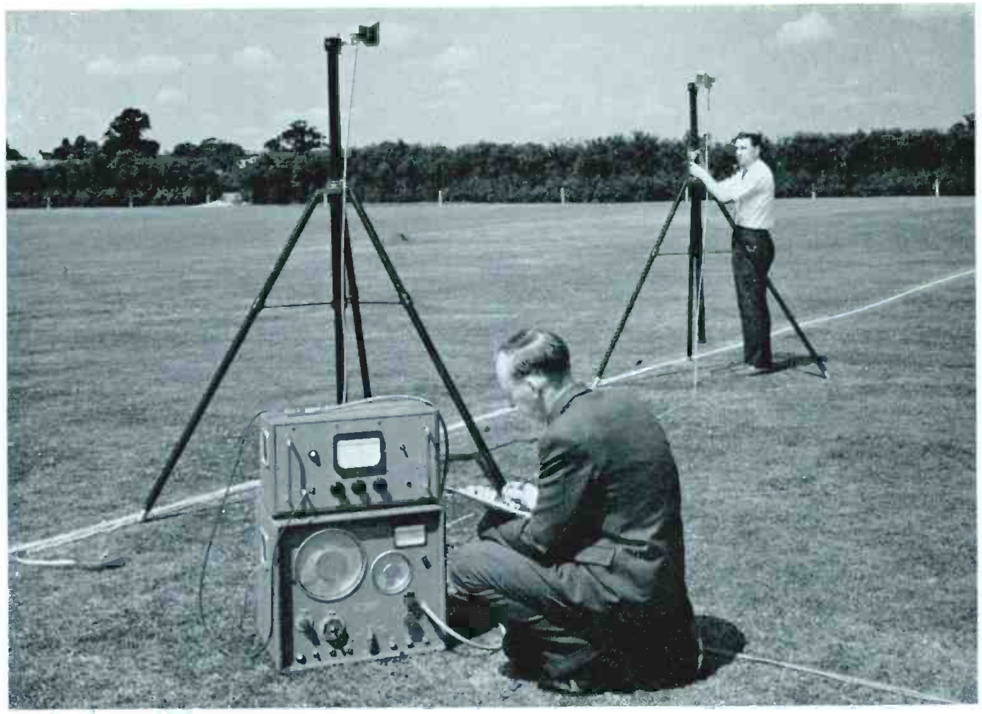


Fig. 1. S Band aerial aperture measurements in progress

3. An output small enough to feed the thermistor mount direct, and thus avoid the use of attenuating cables.
4. A constant and wide angle radiation pattern.
5. 50 Ω output impedance to provide a good match to the thermistor mount.
6. Sensitivity to all modes and directions of propagation.
7. Of robust panclimatic construction—portable and light weight.

With the above requirements in mind several types of aerial were considered—Sleeve Dipole, Equiangular Spiral, Loop, Discone, E-plane Sectoral Horn and a λ/4 Marconi. Of these the Sectoral Horn was chosen for X and S Band operation, and the Discone for L Band.

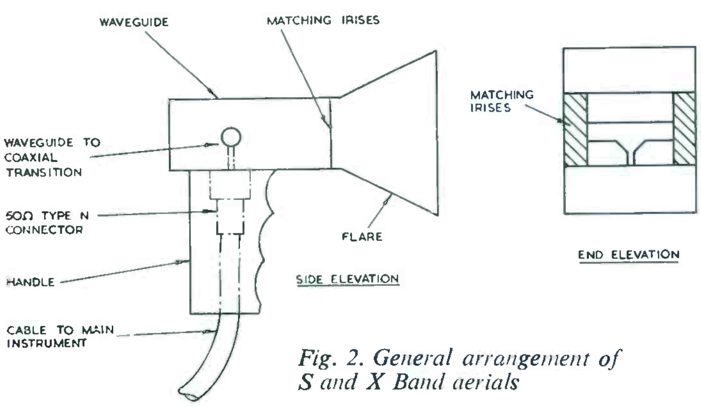


Fig. 2. General arrangement of S and X Band aerials

S AND X BAND AERIALS

Description

The X and S Band aerials are designed along similar lines, one being virtually a scaled model of the other. Each basically consists of a free space to waveguide tapered transition followed by a waveguide to coaxial transition. The two transitions are combined together by a section of waveguide to form an E-plane sectoral horn with a 50 Ω coaxial line output, as shown in Fig. 2.

A pistol grip type of handle is provided as a means of holding the aerial. When held upright the aerial is in position to receive vertically polarized waves; when held horizontal the aerial will receive horizontally polarized waves. For receiving circularly polarized waves the aerial can be held in either position but in this instance the aerial output is 3 dB down and to compensate for this the power meter sensitivity is raised 3 dB, by switching the mode selector switch from the LINEAR to CIRCULAR position.

Matching irises are included to improve the match of the flare to the waveguide section, thus giving an overall improvement in the match between free space and the 50 Ω coaxial line output. A typical curve of the v.s.w.r. looking into the X Band aerial output socket is shown in Fig. 3. The v.s.w.r. of the S Band aerial is also better than 1·3: 1 over its working frequency range.

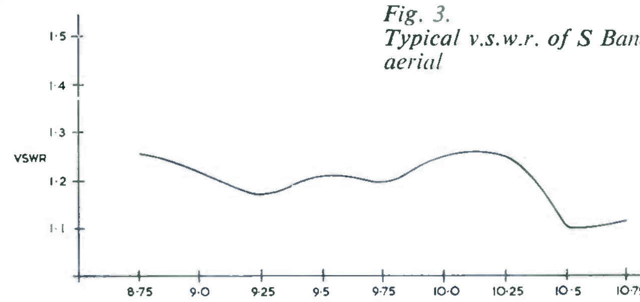


Fig. 3. Typical v.s.w.r. of S Band aerial

Waveguide to Free Space Transition

An improvement in the match between free space and the open end of a section of waveguide can be made by enlarging the waveguide in the plane of the electric field to form a flare, illustrated in Fig. 4.

For best matching the physical opening D should not be less than 0·75 λ at the lowest frequency of operation. For the X Band aerial D was chosen to be λ at 8·73 Gc/s, i.e. 3·43 cm. Another consideration in the flare design is that the phase lag at the edges of the flare should be less than λ/4 with respect to that at centre; this is necessary to avoid side lobes occurring in the radiation pattern.

Therefore, from Fig. 4:

$$h = \sqrt{L^2 + \left(\frac{D}{2}\right)^2} \text{ which must be less than } \frac{\lambda}{4} + L$$

$$\therefore \quad L^2 + \frac{D^2}{4} < \frac{\lambda^2}{16} + \frac{\lambda}{2}L + L^2$$

$$\text{or } L > \frac{D^2}{2\lambda} \left(\text{neglecting } \frac{\lambda^2}{16}\right)$$

At 10·67 Gc/s (highest frequency of operation in X band) λ = 2·08 cm, and with D fixed at 3·43 cm

$$\frac{D^2\lambda}{2} = 2{\cdot}83 \text{ cm}$$

To ensure that the difference between h and L is less than λ/4, L was made equal to 3 cm.

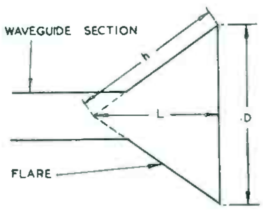


Fig. 4. Waveguide to free space transition

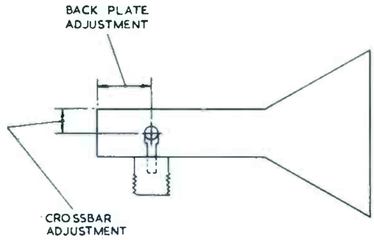


Fig. 6. Waveguide to coaxial transition

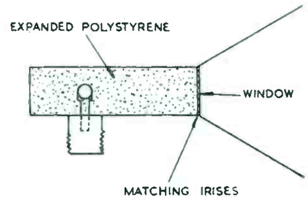


Fig. 7. Waterproofing for S Band aerial

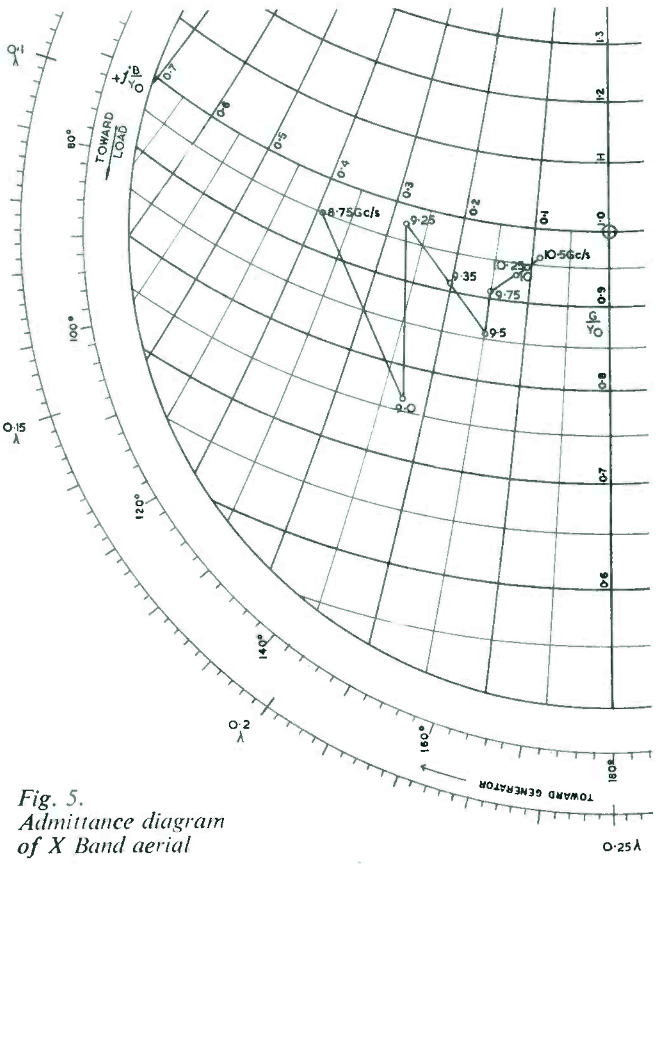


Fig. 5. Admittance diagram of X Band aerial

The v.s.w.r. looking into the X band flare was measured, and was found to rise from 1·13 at 10·5 Gc/s to 1·5 at 8·75 Gc/s. As this was not considered suitable, the admittance was measured at the throat of the flare to give the diagram of Fig. 5. From this it is seen that the flare constitutes a net shunt capacitance corresponding to +jB/Yo across the end of the waveguide section, the value of which increases as the frequency is decreased. To compensate for this shunt capacitance, inductive irises were added in the waveguide.

The irises consist of brass strips soldered in the waveguide in line with the electric field (see Fig. 2). From the admittance chart the conjugate of the capacitive admittance at 9·3 Gc/s, which is an inductive admittance of −jB/Yo = 0·25, was chosen for the irises. These were positioned 14° at 9·3 Gc/s behind the flare towards the generator. Introducing matching irises in this manner had the desired effect and improved the overall v.s.w.r. of the flare to better than 1·3: 1.

Waveguide to Coaxial Transition

A bar and post type transition was used based on the designs described in CSE Report No. 69.[1] Two laboratory prototype transitions were constructed so that the positions of the back plate and crossbar were adjustable (see Fig. 6); this was done so that the transition could be optimized for best match when the waveguide section is terminated by a flare as the design details given in the CSE report apply only when the waveguide section is terminated in a near perfect load.

Waterproofing

Waterproofing is effected by the use of dielectric windows. In the S Band aerial the window is placed across the waveguide at the throat of the flare, but for the X Band version the window is placed across the mouth of the flare. For mechanical strength the window needs to be relatively thick, but for minimum disturbance of the impedance match the window needs to be as thin as possible. Because of its very good tear-resistant qualities polystyrene sheet was chosen and provided a good dielectric window with a very low reflection coefficient. When used for the S Band aerial, however, it was not considered strong enough in view of the large area of 3 sq in. to be covered, at the throat of the flare. To overcome this problem expanded polystyrene was used as a support for the window. The expanded polystyrene which has a dielectric constant approaching unity was used to fill the waveguide section up to the point where the matching irises are situated, see Fig. 7, thus forming a solid backing for the window which is stuck in position.

The X Band waterproofing is slightly different as the cover is placed across the mouth of the flare. The need for additional support to the polystyrene sheet was not considered necessary at X Band owing to the smaller surface area of 1·2 sq in. to be covered. The effect of the waterproofing on the electrical performance of the S and X Band aerials was very small, giving rise to a slight increase in the v.s.w.r. in each case.

Aerial Characteristics

Polar diagrams of the S Band aerial at mid band are shown in Fig. 8 for both the E and H planes. The X Band diagrams are similar. In designing an aerial of this type it is also necessary to consider (*a*) how much power the aerial will deliver to the detector and (*b*) by how much the output will vary over the frequency range, for a given input. One way of obtaining this information is to measure the 'effective receiving aperture' of the aerial. Fig. 9 shows how the effective receiving aperture of the S Band aerial varies over the frequency range: 2·7 to 3·3 Gc/s.

The aperture of the aerials was measured by the air path attenuation method[2] shown in Fig. 1, using two identical aerials and calculated from the formula:

$$A = R\lambda\sqrt{\frac{P_r}{P_t}}$$

where A = aperture in sq. cm

R = distance between aerials in cm

P_r = received power

P_t = transmitted power

λ = wavelength in cm

Fig. 9 shows two curves of aperture against frequency taken at two values of R with the aerials positioned 9 ft 6 in. above the ground plane. The aperture should of course remain the same at both distances but the variations shown are due to measurement errors. One curve shows the aperture to have a maximum variation of 0·65 dB over the band while the other gives 0·85 dB. Considering the aperture to be 80 sq cm which is the mean between the curves at 2·7 Gc/s this means that in a field of 10 mW/cm² the output of the aerial will be 800 mW; this is too great to feed the thermistor mount direct which is operated at 5 mW, therefore the output of the aerial has to be attenuated by 22 dB. Attenuation of the aerial output is achieved by using a length of 50 Ω lossy cable, about 6 dB/ft; in the case of the S Band aerial 4½ ft of cable was required.

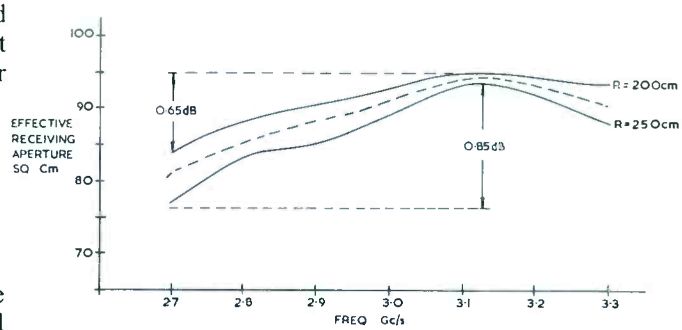


Fig. 9. S Band aerial aperture variations

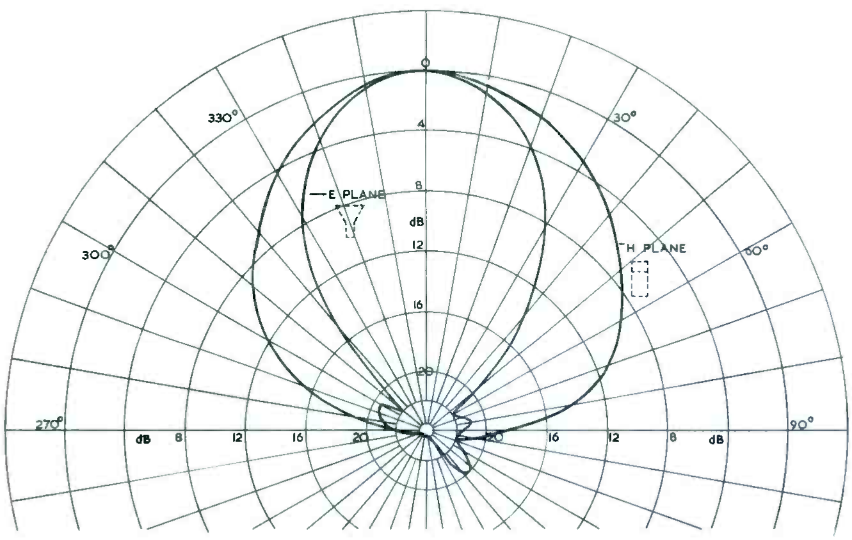


Fig. 8. Polar diagram of S Band aerial

L BAND AERIAL DESIGN

The design of this aerial presents extra problems apart from those encountered in the other two. The reasons for these extra difficulties are:

(1) The much wider bandwidth required, equal to a complete octave.

(2) The lower frequency, 700 to 1400 Mc/s, implies dimensions that would make the aerial difficult to handle.

A first approach to the solution was the obvious one of scaling a horn in direct proportion to the other aerials, but this would have resulted in a horn aperture of about two feet square which would be unmanageable.

The alternative solution was an aerial of the discone type[3] which is broadband and of reasonably constant

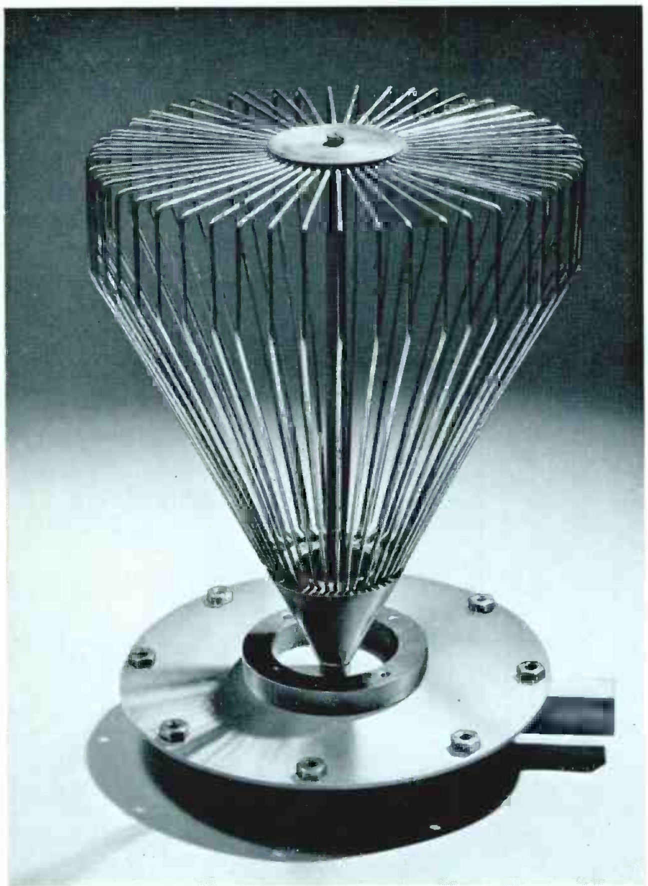


Fig. 10. L Band discone without cover

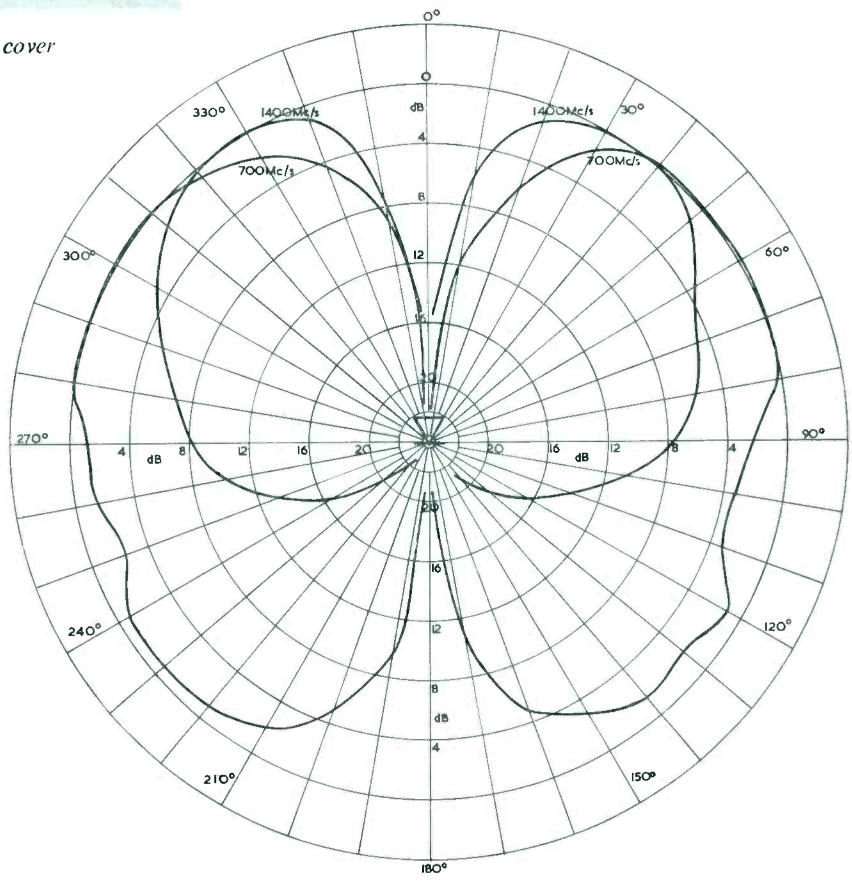
Having determined the main electrical requirements for v.s.w.r. and polar diagram performance, the method of producing a mechanically robust design had to be decided. In order to reduce the weight of the cone, a wire cage (Fig. 10) of identical dimensions to the successful solid cone was designed, with brass mounting pieces to join the separate wires together. This was found to have an identical electrical performance to the earlier solid cone. The outer cover and internal supports for the discone were all manufactured from a sheet of plastic of the polystyrene type. Adding the cover and supports was found to have a negligible effect on the electrical performance of the discone. The completed aerial was of such light and strong construction that it stood up to rough handling quite satisfactorily.

L Band Measurement Techniques

The polar diagrams of the final discone (Fig. 11) show that it had a very broad beam. It was found impractical to use the two-aerial techniques previously mentioned, in order to determine the aerial gain. The ground wave signal was always appreciable compared with the main signal, unless the aerials were brought into very close proximity, where the near field effects caused additional errors in measurement.

In order to reduce the effect of ground waves in the gain measurements, a series of high gain narrow beam Yagi arrays were designed to cover the frequency band, one of which is shown in Fig. 12. These were used as

Fig. 11. Horizontal polar diagrams of L Band aerial



polar diagram performance. Initial experiments were done on solid metal discones of different types, the angle and size of the discone being the main parameters of the design. The other factor which had a secondary effect on the v.s.w.r. was the proportions of the gap at the apex of the cone. Because of the high degree of accuracy of gain figure which the instrument had to have, all the aerials had to be designed to have a v.s.w.r. of better than 1·3 over their working bands. Adding a straight section to a conventional discone improved the v.s.w.r. relative to that which has normally been obtained without reducing its bandwidth. A figure of better than 1·25 was then easily obtainable.

standards from which the gain of the discone could be evolved. The gain was measured at three points in the working band, 700, 1000, and 1400 Mc/s. Yagi arrays have a very narrow bandwidth over which their v.s.w.r. is better than about 1·2. In the arrays used, a bandwidth of about $\pm 1\frac{1}{2}\%$ only was usable. The parameters of the arrays had to be individually adjusted until a pair of arrays was available at each of the frequencies mentioned. The gain of each Yagi was calculated by using the two-aerial method mentioned earlier. Having two Yagi arrays for each frequency it is possible to determine the gain of each Yagi and the discone using the three-aerial method.[3] By constructing the two Yagi arrays of each pair to be as nearly identical as possible, and measuring the transmission loss between each Yagi and the discone, it was found that the transmission path loss agreed within 0·1 dB, so that only one Yagi array need be used in conjunction with the discone at each frequency. Thus the number of measurements and calculations was greatly reduced without affecting the accuracy of the results. The distance between the arrays and their height above the ground were varied to check the effect of ground wave. Also, the v.s.w.r. of the aerials, together with the connecting cables, etc., was checked. The transmission loss between a Yagi and the discone was measured at the three frequencies in the band and from this the aperture and gain of the discone were determined. It should also be mentioned that it was found necessary to use the Yagi arrays as receiving aerials when measuring the polar diagrams of the discone.

It is possible to determine the gain of an aerial from its polar diagram. This has the disadvantage that any calculation cannot take into account the heat losses in the aerial which although not affecting the polar diagram shape will affect the gain and aperture figures. The heat losses could account for up to 1 dB error in the transmission path loss, but this order of error is not permissible since the total error for the instrument must not exceed + 2 dB. In order to eliminate the effect of ground wave and near field effects the discone and Yagi arrays were mounted on non-metallic supports, the aerials being about 15 ft above the ground. The range of distance between the aerials over which the transmission loss follows the formula $A = R\lambda\sqrt{\frac{P_r}{P_t}}$ was between 6 and 14 ft.

When the instrument is used in practice the aerial will only be six to eight feet above the ground so that it will be subject to a considerable multi-path transmission. It is felt that if the sum of the multi-path transmission constitutes a hazard, this hazardous condition would exist to the same extent for a person in the same position as the aerial. If the aerial is calibrated under ideal conditions using a single transmission signal of known strength, there will be no appreciable errors when the instrument is used under normal conditions. If measurements are required to be made in the absence of an operator the aerials L, S and X Band have been provided with a $\frac{1}{4}$ in. B.S.W. tapped hole, to allow mounting on a tripod.

Fig. 12 L Band Yagi test aerial

SOURCES OF ERROR

Below are listed the factors which affect the overall accuracy of the instrument.

(*a*) Due to aerial and cable

(i) Error in estimating the effective receiving aperture.
(ii) Loss due to mismatch between the aerial and its associated low loss cable.
(iii) Error due to variation in attenuation of the low loss cable over the frequency band.

(*b*) Due to components contained in the Power Meter

(i) Loss due to mismatch between aerial coupling unit and the aerial cable.
(ii) Loss due to mismatch between the aerial coupling unit and thermistor bridge.
(iii) Errors in correction filter when used.
(iv) Errors in estimating attenuation of lossy cables forming the aerial coupling unit.
(v) Errors in the power measuring accuracy of the thermistor bridge.

Taking the sum of all the possible sources of error for the worst case an overall specification of −0 to +2 dB is envisaged.

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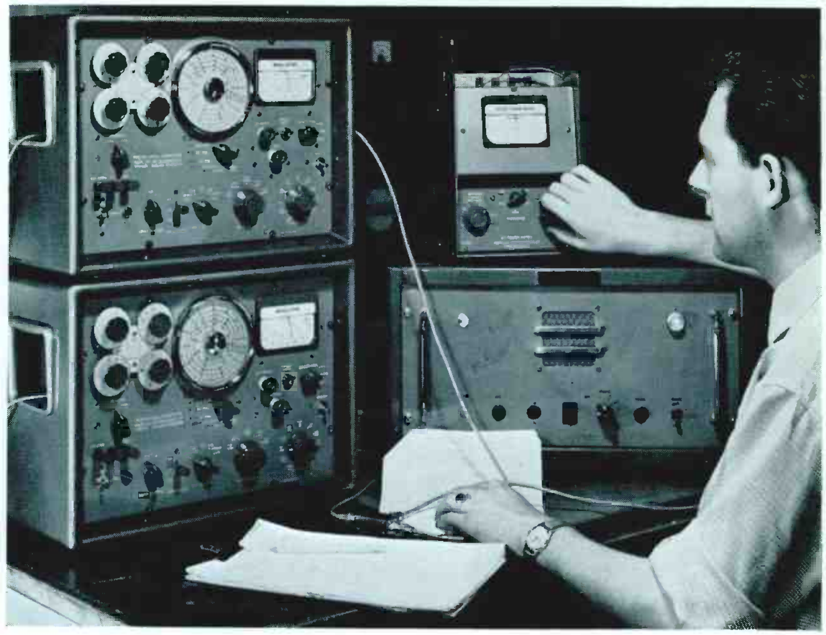
(2) C.G. Montgomery: 'Technique of Microwave Measurements'. — Radiation Laboratory Series No. 11 (McGraw-Hill 1947), p. 907.
(3) Nail, J. J.: 'Designing Discone Antennas'. *Electronics* August 1953, 26 p. 167.

621.396.615.12

Is a high output signal generator always necessary?



Two signal generator test on a receiver using F.M./A.M. Signal Generators, Type TF 995A/5

by J. M. PARKYN

Two-signal testing and loop testing of receivers normally require high outputs from signal generators. By using the techniques described, lower-output generators will often suffice.

IT IS UNUSUAL for all the required performance features of a measuring instrument to be achieved without compromises since some of the requirements will inevitably be conflicting, e.g. high output signal generators with very low leakage fields. Some compromise will *always* be necessary if an economical price is to be maintained. For this reason it is desirable to choose an instrument which is fully able to provide only the features essential for the tests envisaged. In the case of signal generator selection it is generally assumed that a large output of over, say, ¼ V is necessary for receiver testing. This is not necessarily the case as the two following examples show.

Two-signal Testing

As most receiving equipment is required to operate in congested frequency bands it is now common to augment the tests of selectivity and image response with tests involving two or more signals applied simultaneously to the aerial input to simulate more closely the receiver response under operational conditions.

To combine the output of two signal generators of equal impedance and provide a single output of the same impedance involves a loss of 3:1 if the commonly used symmetrical circuit shown in Fig. 1 is employed. As a two-signal test generally involves one strong and one weak r.f. signal separated by say the frequencies of the first i.f., a combining pad with equal loss on both arms is not the most economical system when considering a possible limitation of signal generator outputs. Fig. 2 shows an arrangement involving negligible loss in larger signal path and a 26 dB insertion loss for the signal which is required to be weaker. In practice there is less than ½ dB loss in the stronger signal path due to a very slight mismatch, but if required this can be allowed for in the calculations.

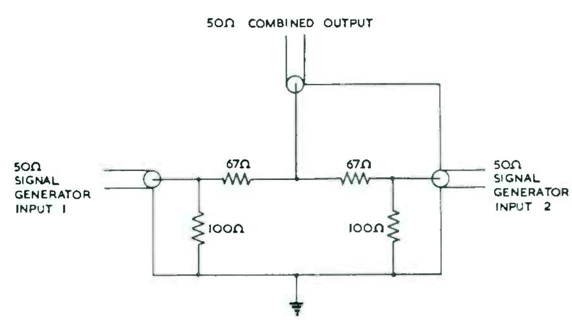


Fig. 1. 50 Ω coupling networks 10 dB loss

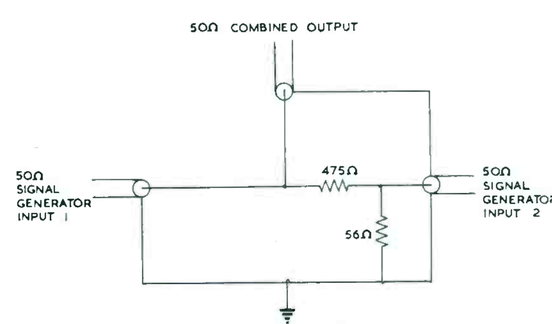


Fig. 2. 50 Ω coupling network giving zero and 26 dB loss

Loop Testing

With the advent of ferrites, which enable loop or frame aerials to be made much smaller for the same pickup, and more recently transistors, a number of moderately high performance broadcast receivers with permanently connected built-in aerials are being tested.

To test overload conditions and a.g.c. performance strong radio frequency fields up to say 0·2 V/m are required. One of the commonly used loops for generating a known field at the receiver aerial is so proportioned that the conversion of signal generator e.m.f. to volts per metre is 0·1. The relation between equivalent field intensity at the receiver aerial and transmitting loop current is given by the formula,

$$E = \frac{188{\cdot}5\, Nr^2}{D^3} \,.\, I$$

where, E = equivalent electric field intensity in volts per metre at receiver aerial

N = transmitting loop turns

r = transmitting loop radius in metres

D = distance in metres between loop centres or more accurately in the case of large radius receiver loops it is the distance between the centre of the transmitting loop and the receiver loop periphery

I = loop current in amps.

The simple conversion of 1 : 0·1 exists when $N = 3$, $r = 5$ in., D = 2 ft, and the current I is maintained by feeding the loop via a total resistance (generator + padding) of 403 Ω. This system relies upon the loop reactance being small so that the current is determined by the e.m.f. and the circuit resistance. In view of the small loop inductance this is accurate up to frequencies of several Mc/s. Other users employ a high inductance loop when the loop reactance can be used to calculate the loop current. Both these systems require high output signal generators to provide strong fields. However, by eliminating the padding resistance and calculating I from the ratio $\frac{V}{\sqrt{R^2 + X^2}}$

where R = signal generator output resistance

and X = loop reactance

a conversion approaching unity can be obtained in the long and medium wavebands.

Two signal loop tests can require very large signal generator outputs but the two economy measures outlined above can be used together with a drastic saving in signal generator output and outlay!

MARCONI INSTRUMENTS APPLICATION NOTE

621.317.382

A.C. POWER MEASUREMENT USING AN OSCILLOSCOPE

by E. J. PARKER

THE CONVENTIONAL METHOD of measuring the power dissipated in a component when a non-sinusoidal electric current is flowing, as is met in Class C amplifiers and oscillators, is to surround the component with a water-jacket, and by calorimetric measurements and calculations to find the heat and hence power dissipated. This method, although as far as accuracy is concerned is good, is very tedious and cumbersome and is sometimes impossible.

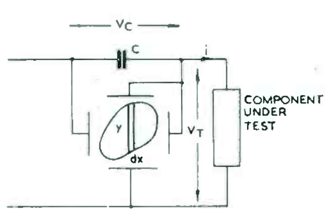


Fig. 1. Basic circuit

In the method to be discussed the voltage across the component under test is applied to one pair of oscilloscope plates and the voltage across a series capacitor is applied to the other. The area of the resultant closed figure displayed on the cathode ray tube screen is then proportional to the mean product of the two voltages, which can be shown to be proportional to the mean power dissipated over one cycle.

Let: V_T = voltage across test component, in volts

V_c = voltage across capacitor, in volts

K_1 = X-sensitivity of oscilloscope, in volts/cm

K_2 = Y-sensitivity of oscilloscope, in volts/cm

A = Area enclosed in one cycle, in cm²

C = Capacitance of series capacitor, in farads

f = Fundamental frequency, in c/s

P = Power dissipated, in watts

i = current flowing, in amperes.

$$\text{Then } A = \int_0^T y\,dx \quad \text{where } T = \frac{1}{f}$$

$$= \int_0^T \frac{1}{K_2} V_T\, d\left(\frac{1}{K_1} V_c\right)$$

$$= \frac{1}{K_1 K_2} \int_0^T V_T\, dV_c$$

$$\text{But } dV_c = \frac{idt}{C}$$

$$\therefore \quad A = \frac{1}{K_1 K_2 C} \int_0^T i V_T\, dt$$

Comparing this equation with $P = \frac{1}{T}\int_0^T ivdt$, the expression for electrical power,

$$A = \frac{1}{K_1 K_2 C} TP$$

$$\text{or } P = \frac{AK_1 K_2 C}{T} = AK_1K_2Cf$$

put $K_1 K_2 = K$

then $P = \underline{AKCf}$

In most cases, as in the anode load of an amplifier, the current flowing through the component under test will have a d.c. component. The capacitor, however, will not pass this d.c. component and so a low-pass shunt circuit has to be used in order to maintain working conditions. An inductor or resistor and inductor may be used for this. The requirements to be fulfilled are that the shunt circuit should pass negligible signal current and that the resistance be as required to maintain the correct conditions in the rest of the circuit.

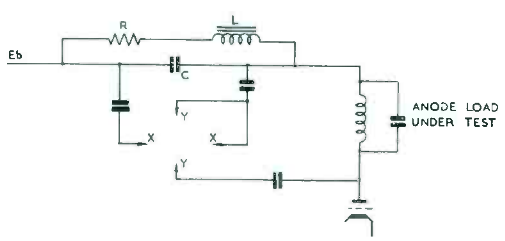


Fig. 2. Actual circuit

An actual circuit used is shown in Fig. 2. An oscilloscope having differential input facilities was used in order to measure the voltage across the anode load, as both ends of the load had signal on them.

If a balanced Y-input amplifier is not available, the voltage could be taken direct to the CRT Y-plates, using a voltage divider if required. In this case the plates should be at the correct d.c. potential, and should normally be driven balanced.

If V_T is much greater than V_c, then the Y-input need not be balanced, the oscilloscope being connected between earth and the anode, in the circuit shown. This introduces an error proportional to the difference between V_T and the vector sum of V_T and V_c.

For the circuit shown in Fig. 2 R was fixed at 1 kΩ in order to maintain the correct anode current.

The X-sensitivity was adjusted to about its mid-position.

The Y-sensitivity and the value of capacitor C were adjusted to give convenient Y- and X-deflections respectively. The relevant values measured were:—

f = 41 kc/s
k_1 = 1·5 V/cm
k_2 = 70 V/cm
C = 0·0246 μF (0·025 μF nominally)
A = 42·9 cm²

The value of inductance L was approximately 1·3 H; this passed less than 0·1 % of the signal current.

$$\therefore P = AK_1K_2Cf$$
$$= 42{\cdot}9 \times 1{\cdot}5 \times 70 \times 0{\cdot}0246 \times 10^{-6} \times 41 \times 10^3$$
$$= \underline{4{\cdot}55 \text{ W}}$$

The measurement accuracy was estimated to be as shown below:

Quantity	*Possible Error*
Frequency	0·1 %
Area	0·5 %
Capacitance	0·25 %
$K = K_1K_2$	4 %
OVERALL	5%

With more care and by calibrating the oscilloscope immediately before use at the frequency to be used, the overall accuracy could probably be improved to 2·5 %.

Having connected the circuit, the longest part of the measurement is the estimation of the area enclosed by the trace. Several methods for this were tried, each involving reproducing the outline on tracing paper. The simplest and most accurate but slowest method used transparent graph paper for the tracing and the number of squares within the area were counted. A slightly less accurate but quicker method tried was to cut out the traced shape and compare its weight with that of a known area of the same type paper. The quickest way of all was to measure the area with a polar planimeter. The accuracy of all these methods was estimated at better than 0·5 %.

This method of power measurement can be used conveniently from the audio frequency range up to several Mc/s, for powers of a few milliwatts to kilowatts with accuracies of better than 5 %, for almost any waveform.

621.317.727

MARCONI INSTRUMENTS

Suppressed Zero Voltmeter . . . *TYPE TF 1377*

by G. PETERS, B.Sc.

A description is given of a d.c. voltmeter designed to accurately measure very small changes in direct voltages up to 500 V. Variations as small as 1 mV in 100 V can be observed, and absolute measurements of voltages up to 100 V made with an accuracy of 1%. For use in voltage drift measurements, provision is made for the operation of a chart recorder.

IT IS SOMETIMES necessary to measure small changes in a relatively large direct voltage, when a normal valve-voltmeter does not provide the required scale resolution. One method which can be used to make such measurements is to back-off the major part of the voltage with a battery, and to measure only the small changes on a sensitive voltmeter. This technique may be inconvenient as the battery does not provide a continuously variable voltage, and it has the disadvantage of introducing errors due to changes of battery voltage if measurements are being taken over a long period.

The Suppressed Zero Voltmeter was designed to provide a convenient method of making measurements of this type, using the backing-off principle. By its use a discrimination of one part in 10^5 can be obtained in measurements up to 500 V. It can also be used to measure the absolute value of voltages up to 100 V with an accuracy of 1%; the maximum voltage which can be measured is 500 V.

A highly stable power supply produces a known voltage across a ten-turn helical potentiometer. An accurately

Marconi Instruments Suppressed Zero Voltmeter, Type TF 1377

For storage purposes the test and mains leads can be housed in the top of the instrument and the lid closed

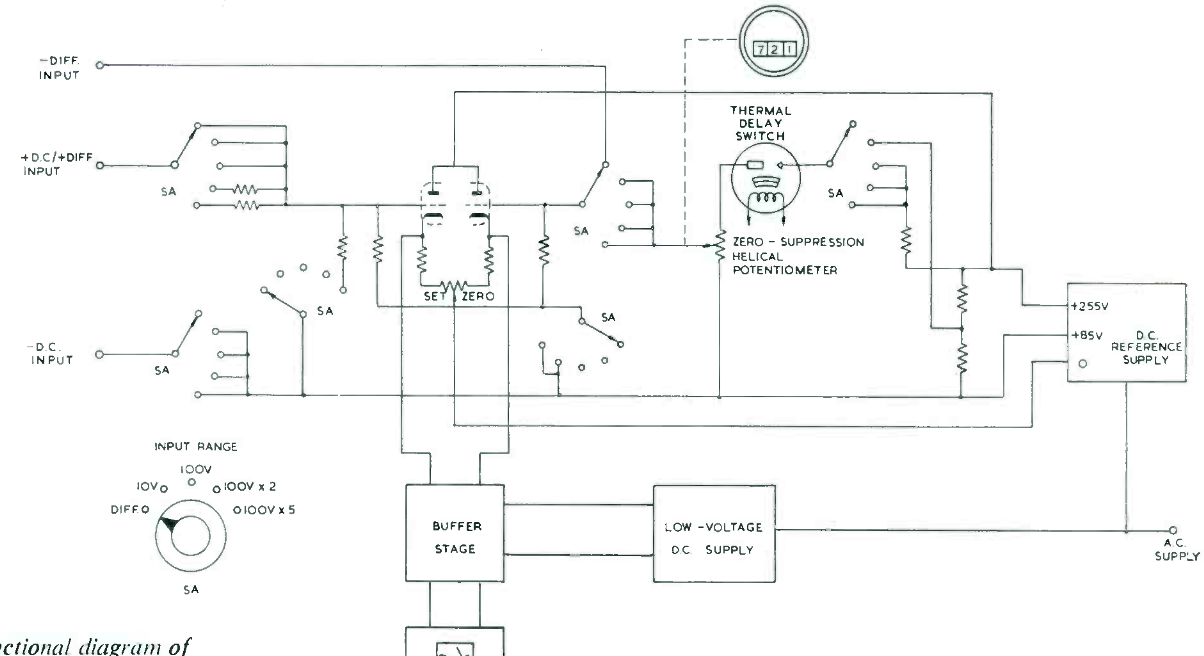


Fig. 1. Functional diagram of Suppressed Zero Voltmeter, Type TF 1377

known fraction of this voltage can be tapped off, the value being indicated on a three-digit microdial incorporated in the potentiometer control. This voltage is applied to a differential electronic voltmeter to back-off the voltage being measured. Any residual unbalance between the two voltages is indicated on a centre-zero moving-coil meter incorporated in the differential voltmeter. The meter can be switched to four sensitivity ranges between 50–0–50 V and 50–0–50 mV. The absolute value of the voltage is given by the indication on the meter added to that recorded on the dial, and any subsequent small variations in voltage can be measured on the meter. There are four ranges in the suppressed zero mode of operation, giving full scale potentiometer readings of 10, 100, 200 and 500 V. The voltmeter can also be switched for differential use, when it becomes a conventional valve voltmeter with balanced input, measuring up to 50–0–50 V full scale; in this mode the maximum sensitivity is 50–0–50 mV.

A feature of the instrument is its very high input resistance, 50 MΩ on the three lower voltage ranges and 25 MΩ on the 500 V range. The inputs are isolated from chassis so that either side of the input may be earthed. Output terminals, in parallel with the meter, are provided to connect the instrument to a chart recorder for use in long-term drift measurements. At full scale deflection the voltage available at the terminals is 25 mV.

The instrument is small and light, with low power requirements, and employs a hybrid valve and transistor circuit.

Applications

The main use of the instrument in the suppressed mode is for the measurement of small incremental changes in large voltages. One important use is in the checking and measuring of the regulation and stability of d.c. power supplies. Also, the provision of a recorder output, together with the low drift of the voltmeter, make it particularly suitable for drift measurements. In fact the Suppressed Zero Voltmeter has applications in the measurement of small changes in any physical quantity which may be expressed as a direct voltage by the aid of suitable transducers. A typical example is its use in conjunction with strain gauges.

Temperature coefficients of semi-conductor devices such as Zener diodes can be measured directly by observing the voltage changes produced by variations in temperature. In the same way temperature coefficients of resistors can be measured by passing a constant current through the resistor and observing the voltage change across it. Another possible application is for the calibration of resistance thermometers and piezo-electric transducers.

Because of its high input impedance and sensitivity, the Suppressed Zero Voltmeter can be used as a bridge detector in a similar manner to the pH Meters, TF 889/1M and TF 1093.[1] At null deflection the input impedance, which is normally greater than 50 MΩ, becomes infinitely large so the bridge circuit is not loaded by the instrument.

Circuit Summary

A block schematic diagram is shown in Fig. 1. For use in the suppressed mode the voltage under test is applied between one grid of the double triode valve and the negative end of the helical potentiometer. The voltage tapped off the helical potentiometer is applied to the other grid. In use this voltage is adjusted to give balance between the two cathodes, the digital dial being set to the nearest digit below balance. The residual unbalance voltage between the cathodes is applied, via the transistor buffer stage, used to avoid excessive loading on the valve, to the centre zero meter.

In the differential mode, the instrument is used as a simple valve voltmeter by applying the voltage to be measured directly to the grids of the two valves with the potentiometer out of operation. The voltage is read directly off the meter on its appropriate range.

D.C. Reference Supply

A diagram of the stable reference supply is given in Fig. 2.

An h.t. supply is obtained from a secondary winding on the mains transformer via a full wave rectifier and smoothing capacitor. For stabilization a two-stage neon tube circuit is used in which the second stage is supplied with constant current via a transistor.

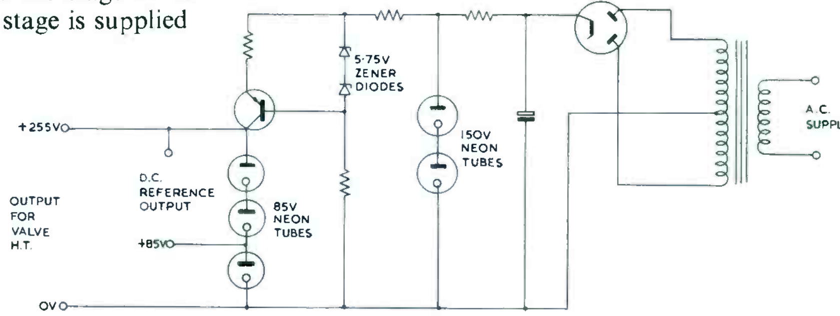


Fig. 2. D.C. reference supply

The first stage comprises two 150 V voltage-stabilizer neon tubes producing a partially-stabilized output voltage which is applied to a silicon transistor. This has a base reference voltage of 11·5 V fixed by the two 5·75 V Zener diodes, so that the emitter current is held constant. In this stage, 5·75 V diodes are used because of the low temperature coefficient of Zener diodes at this voltage. Both diodes and the transistor are mounted in brass blocks to reduce rapid variations of temperature and hence to reduce short-term drift. The constant collector current of the transistor provides the current feed for the second chain of three 85 V voltage-reference neons, chosen because of their internal resistance, which is very low compared with the output resistance of the transistor, thus ensuring a stable voltage across the neon chain.

A d.c. reference supply for the potentiometer is taken across two of the 85 V neons, the complete supply being used to provide a stable voltage for the anode of the double triode valve. Stabilization of the circuit is such that a 10% change in mains voltage produces less than 2 mV change in the meter reading.

Low-Voltage Stabilized Supply

The supply for the transistor buffer stage is provided by a low voltage stabilizer, shown in Fig. 3. A separate secondary winding on the mains transformer provides the d.c. supply via a full wave rectifier employing germanium junction rectifiers. This supply is stabilized by a series stabilizer using a germanium power transistor with a base reference voltage of 5·75 V provided by a Zener diode in series with a resistor chain across the supply rail. The output is taken from the low impedance emitter of the transistor.

This stabilizer is also used to provide the heater voltage of the voltmeter valve to avoid fluctuations in readings due to heater variations.

Potentiometer and Differential Voltmeter

This circuit is shown in detail in the block schematic diagram Fig. 1.

The d.c. reference voltage is applied across the helical potentiometer via a series resistor and a thermal delay switch. For the 100 V range, the voltage across the potentiometer is set to 99·9 V by means of a variable resistor in series with the potentiometer. Similarly on the 10 V range separate preset variable resistors set the voltage across the potentiometer to 9·99 V. The thermal delay switch is included to remove the load from the second neon chain when switching on to ensure striking of the neons.

The differential voltmeter consists of a double-triode valve with the two halves arranged as cathode followers. Their anodes are connected to the d.c. reference supply and the negative line of this supply is returned to the slider of a variable resistor in the cathode circuit. This

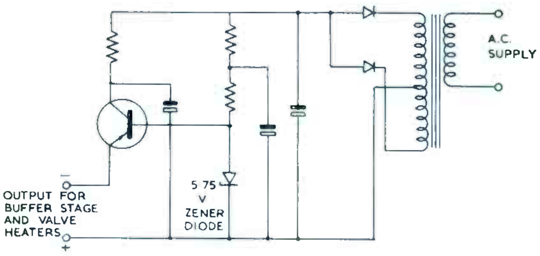


Fig. 3. Low-voltage stabilized supply

resistor is a coarse set zero control. A series variable resistor in one cathode circuit is the fine set zero control. Before use the instrument is set to zero using these controls with the three input leads connected together, the zero being indicated on the most sensitive range of the meter.

In the suppressed mode of operation, the input is applied between one grid of the valve and the negative end of the potentiometer, the input leads +DC and −DC being used. The second grid is taken to the tap on the potentiometer and is thus fed with a known voltage as indicated by the digital dial reading.

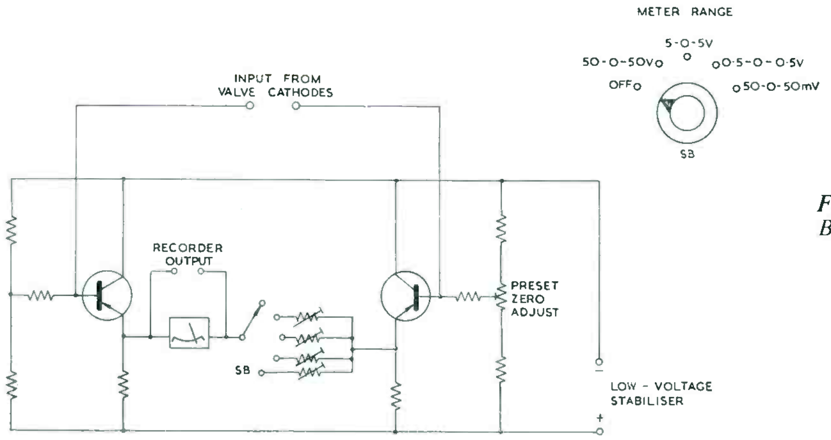


Fig. 4.
Buffer stage and meter circuit

On the 200 and 500 V ranges the positive input is taken to the first grid via potential dividers with division ratios of two and five respectively. This arrangement gives an input impedance of greater than 50 MΩ on the 200 V range and greater than 25 MΩ on the 500 V range. For these two ranges the voltage across the potentiometer is as for the 100 V range. The voltage readings on the dial and on the meter have to be multiplied by two or five as appropriate to give the true value.

In the differential mode the potentiometer is switched out of circuit and the input is applied directly between the two valve grids, with a 50 MΩ shunt resistance across them, using the +DIFF and −DIFF input leads.

The potentiometer and voltmeter circuits, as well as the reference supply, are isolated from earth.

Buffer Stage and Meter Circuit

A simplified circuit diagram is shown in Fig. 4.

In operation the difference voltage between the two cathodes of the valve is applied to the bases of two germanium transistors in the common-collector configuration. One base is taken to a potentiometer across the supply, to give a preset zero control. The output is taken from the two emitters and is applied via four switched preset variable resistors to the meter. These positions provide the four meter sensitivity ranges of the instrument, which are initially set up by means of the variable resistors.

REFERENCE

1. Murphy, B. A.: 'A pH Meter as a Bridge Detector.' *Marconi Instrumentation*, 1962. 8, p. 128.

ABRIDGED SPECIFICATION

Suppression voltage ranges

0 to 10 V d.c.
0 to 100 V d.c.
0 to 200 V d.c.
0 to 500 V d.c.

ACCURACY:

±1% of reading ±10 mV on 10 V range.
±1% of reading ± 100 mV on 100 V range.
±2% of reading ±200 mV on 200 V range.
±3% of reading ±500 mV on 500 V range.

STABILITY:

±10 mV/hr, ±3 mV short term, on 10-V range.
±30 mV/hr, ±3 mV short term, on 100 V range.

Meter ranges

50–0–50 mV.
500–0–500 mV.
5–0–5 V.
50–0–50 V.

ACCURACY:

±5% of full-scale.

Discrimination

±1 mV on 10 and 100 V ranges.
±2 mV on 200 V range.
±5 mV on 500 V range.

Input resistance

Infinite at meter null and greater than 50 MΩ off-balance on 10 and 100 V ranges.
Greater than 50 MΩ on 200 V range.
Greater than 25 MΩ on 500 V range.

Recorder output

±25 mV at full-scale deflection of meter.

AMERICAN APPOINTMENT

Mr. W. A. Buck has been appointed Manager of Marconi Instruments' organization in the U.S.A., which has its headquarters at Englewood, New Jersey. He succeeds Mr. R. J. Bailey, who is returning to the main works at St. Albans after eight years of building up the Company's very substantial business in America.

Bill Buck joined Marconi Instruments in 1954, and was posted to the U.S. Office a year later. He was Assistant U.S. Manager for a number of years, and has an excellent knowledge of the Company's business in that market. In June, 1960, he returned to Head Office at St. Albans, but since then he has maintained liaison with the U.S. Office through his work with the Export Department.

MARCONI INSTRUMENTS APPLICATION NOTE

621.317.733.3

MEASUREMENTS OF HIGH-VALUE ELECTROLYTIC CAPACITORS

by J. F. GOLDING

There is a growing need for bridges suitable for measuring high-value electrolytic capacitors. Simple modifications to the Marconi Universal Bridges can increase the capacitance range and facilitate the application of polarizing voltage.

THE ADVANCES generally associated with the introduction of transistors would not have been possible had it not been for the fact that manufacturers of the complementary circuit components have kept pace in terms of technical development. The tantalum electrolytic capacitor in particular is a noticeable example of such a component. Very high capacitance values are achieved in these capacitors, which leads to a requirement for a suitable measuring bridge.

The upper limit of the capacitance range of the TF 1313 and TF 868B bridges is of the order of 100 μF. This is adequate for most electrolytic capacitors in common use; and, as the bridge excitation voltage is quite small, it is often possible to dispense with any d.c. polarizing voltage.

Tantalum capacitors are, however, available with values up to several hundred microfarads. And for measurement of such values, either of the Marconi Universal Bridges can be modified to extend the capacitance range to 1000 μF. This is achieved by increasing the value of the internal standard capacitor by a factor of 10.

Range Extension of TF 1313

The modification is simple and need not interfere with the normal functioning of the bridge. Fig. 1 shows the electrical arrangement as applied to a TF 1313 ¼% Universal Bridge. The added circuitry is drawn in heavy lines. A 0·9 μF capacitor (C add) is switched in parallel with the existing standard capacitor which is adjusted to 0·1 μF so that the additional 0·9 μF gives a value of 1·0 μF overall.

The capacitance tolerance on high-value electrolytic capacitors is generally wide; so there is little advantage in using an expensive high stability capacitor as C add. A good quality paper-dielectric capacitor can be used providing the bridge is not subjected to large variations of ambient temperature. It should be possible to select a suitable one having the correct value from a batch of nominally 1 μF ±20% capacitors. If the selected capacitor is 0·9 μF ±1%, a bridge accuracy of the order of ±5% can be obtained allowing for temperature variations and aging. If greater bridge accuracy is required a more stable capacitor should be used.

In the TF 1313 the additional 'standard' capacitor can be secured below the top panel of the instrument as shown in Fig. 2. The switch can be of the toggle type mounted on the front panel to the right of the EXTERNAL DETECTOR socket as shown in Fig. 3.

It is shown with engraving STD (standard) for the open position and C×10 for the closed position. In the open position the bridge functions normally and the modification is ineffective. When the switch is set to C×10, balance is obtained in the normal way and the indicated capacitance is multiplied by 10.

A double-pole switch is used so that the wiring to the additional capacitor is isolated from the standard when the switch is open. This minimizes the additional stray capacitance across the standard. To preserve the accuracy

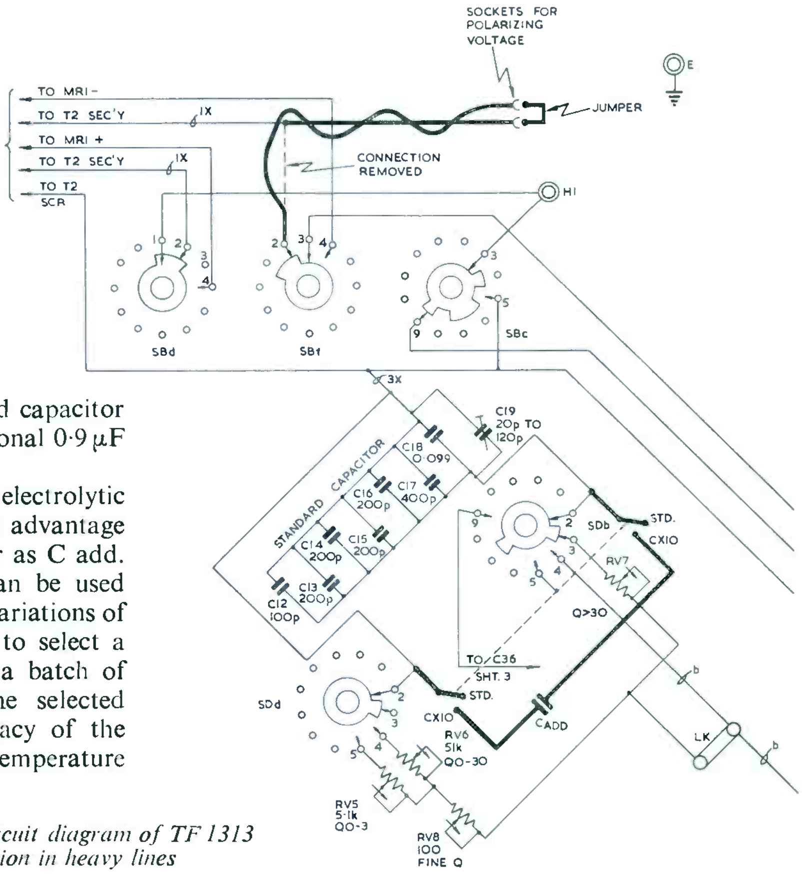


Fig. 1. Part of circuit diagram of TF 1313 showing modification in heavy lines

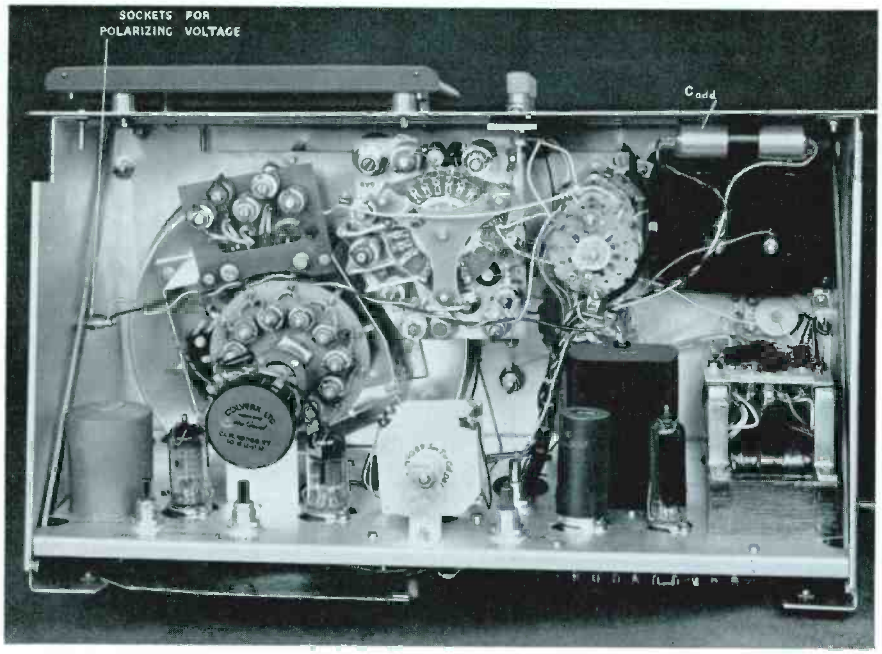


Fig. 2.
Rear view showing modifications

of the balance calibration, particularly when measuring low Q components, it is important that the coupling between this switch with its associated leads and the oscillator circuit is kept virtually zero. Providing the modification is carefully done, with this point borne in mind, the accuracy of the bridge for normal functioning will not be affected, and no further adjustment should be necessary.

However, as a precaution it is advisable to recheck the calibration of the bridge after modifications against an external standard capacitor. Any convenient value may be used. If necessary, the trimming capacitor C19, which forms part of the internal standard, may be adjusted to compensate for any change in stray capacitance. If no suitable standard capacitors are available, a stable capacitor should be standardized by measuring its value accurately with the bridge before modification.

During the initial calibration of the bridge, the correct *in situ* capacitance of 0·1 μF is obtained by disconnecting one or more of the fixed capacitors C12 to C18. Final adjustment is made with the variable preset C19. It is important, therefore, that these capacitors remain disconnected and that the position of each component in this part of the circuit remains undisturbed. Connection to the toggle switch is best made at pins b2 and d2 of switch SD.

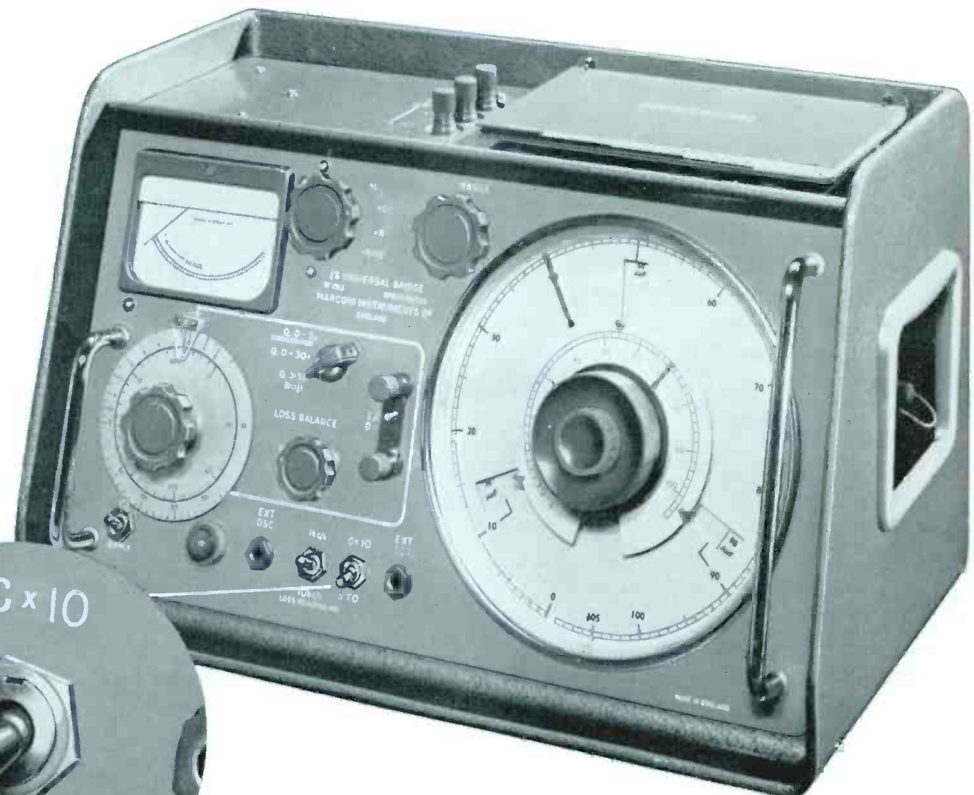


Fig. 3.
Modified ¼% Universal Bridge, Type TF 1313. The link for applying the polarizing voltage is just visible in the right-hand pocket

Range Extension of TB 868B

A similar modification can be applied to the TF 868B Universal Bridge. Connection to the internal standard capacitor is made at the terminals of the Q-tan δ switch.

The additional capacitor can be mounted on the underside of the top panel in the same way as for the TF 1313, and there is room for the STD–C×10 toggle switch adjacent to the frequency selector switch on the front panel.

This is convenient from a wiring point of view. Leads can be kept short and the changes in stray capacitance become negligible. This consideration is, of course, less important in this bridge than in the TF 1313 because the tolerance for normal operation is four times greater. Nevertheless, care should be taken to keep the leads as short as possible, and routing them too close to the front panel should be avoided.

There is no provision for adjusting the *in situ* value of the internal standard capacitor of the TF 868B.

Provision for Polarizing Voltage on TF 1313

The most satisfactory way of applying a d.c. polarizing voltage to the capacitor being tested is by connecting the d.c. source in series with the a.c. exciting voltage as shown in Fig. 4. It is, of course, important that this source shall have a low a.c. impedance; otherwise the sensitivity

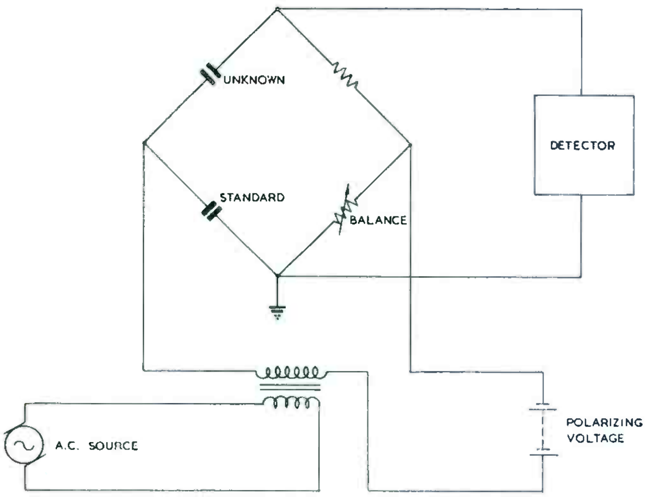


Fig. 4. Method of applying polarizing voltage

of the bridge will be reduced. The direct current passing through the electrolytic capacitor resulting from this polarizing voltage is normally small so that there is little risk of saturating the core of the bridge input transformer or of damaging the circuit. However, it is advisable to provide a d.c. limiting resistor for other reasons that will be dealt with later.

Fig. 1 shows the circuit of the bridge with the modified wiring in heavy lines. The lead from the secondary winding of transformer T2, which normally goes to contact f2 of switch SB, is disconnected from the switch and taken to an insulated tag. This tag is fixed to one of the switch support pillars, the end nut which forms part of the switch assembly being utilized to hold it in position. A twisted pair of leads are taken from the tag and contact SB f2 to a pair of banana sockets mounted into the side panel inside the right-hand handle recess. For normal use, these sockets are connected together by means of an external link. For application of the polarizing voltage, the link is removed and the d.c. source is connected to the sockets.

When the loss-balance selector switch, SD, is in the $Q > 30$ position, the loss-balance resistor is in series with the standard capacitor, so that there is no d.c. path through the lower two arms of the bridge. When the switch is set to the $Q = 0$–30 or $Q = 0$–3 positions, however, the loss-balance resistor is in parallel with the standard capacitor. Thus, when the polarizing voltage is applied, direct current flows in the loss-balance resistor, the calibrated variable arm of the bridge, and the secondary winding of the transformer T2. It is, indeed, easily possible for sufficient current to flow to damage the close-tolerance resistors in the circuit at applied voltages much below the desired polarizing voltage. As electrolytic capacitors are usually of fairly low Q, this might appear to be a serious limitation. However, the difficulty can easily be overcome by either of two methods.

The TF 1313 is fitted with a pair of terminals marked EXT D–Q. These are normally linked together by means of a shorting bar, which is effectively in series with the loss-balance resistor at all settings of the selector switch SD.

It is, thus, a simple matter to break the d.c. path for the lower Q settings of the switch by taking off the shorting bar and connecting a blocking capacitor between the terminals. It is, however, important that the reactance of this capacitor is small compared with the resistance of the loss-balance resistor at its balance setting. This is given by:

$$R = \frac{Q}{2\pi f C}$$

where R is the phase balance resistance in ohms
f is the bridge frequency in c/s
C is standard capacitance in farads

As the actual value of R is not known, the effect of the d.c. blocking capacitor on the accuracy of the bridge can be assessed from its shunting effect on the internal standard capacitor. The added shunt capacitance due to the blocking capacitor is given approximately by:

$$C_{\text{shunt}} = \frac{C^2_{\text{standard}}}{C_b\, Q^2}$$

where C_{standard} is the external standard capacitance
C_b is the value of the blocking capacitor

$$\therefore \quad \%\ \text{error} = \frac{C_{\text{standard}} \times 100.}{C_b\, Q^2}$$

$$\therefore \quad C_b = \frac{C_{\text{standard}} \times 100}{Q^2 \times \%\ \text{error}}$$

Supposing the permissible error is 5% and the lowest Q to be expected is 2, the value of blocking capacitor becomes

$$\frac{C_{\text{standard}} \times 100}{4 \times 5} = C_{\text{standard}} \times 5$$

So to give a measuring accuracy of 5% when the 1 μF standard is being used, the required value of blocking capacitor would be 5 μF, or if greater accuracy is required the capacitor may be increased in value.

An alternative method is to use the bridge with the selector switch set to $Q > 30$ and add resistance between the EXTERNAL D–Q terminals to achieve phase balance. A variable resistor would, of course, be used and its maximum value in ohms is given by the expression

$$R_{\text{max}} = \frac{1}{2\pi f C Q_{\text{min}}}$$

where f is the bridge frequency in c/s

C is the standard capacitance in farads

Q_{min} is the lowest Q to be measured

With the bridge operating at 1 kc/s using the normal 0·1 μF standard capacitor and the lowest Q to be measured being 2, the required value of the series variable resistor would be 796 Ω.

With the 1 μF standard capacitor the value could be reduced to 79·6 Ω. These values apply when the internal potentiometer is set to zero, so as an alternative, lower values of external resistance could be used in conjunction with the internal resistance which has a maximum value of 5 Ω.

It should be mentioned that, with these low values of Q, the two alternative methods will give different capacitance values corresponding to equivalent shunt capacitance and equivalent series capacitance respectively. This aspect is covered fully in the instruction book of the bridge and does not warrant detailed treatment in this article.

The D.C. Source

The d.c. polarizing source must be isolated from earth. A suitable battery is the most convenient form since the current drain should be negligible.

Steps for preventing a deliberate d.c. path through bridge arms have been described in some detail. There is the danger, however, that the capacitor being tested may break down and offer a low resistance d.c. path through the upper arms of the bridge. To prevent damage from this cause, it is advisable to apply the voltage via a series resistance of 1000 Ω /V.

In order to keep the a.c. impedance of the source as low as possible, an electrolytic capacitor of suitable working voltage and as high a value as convenient should be connected between the polarizing voltage input sockets.

Provision for Polarizing Voltage on the TF 868B

The TF 868B is not provided with external D–Q terminals. Therefore, although the circuit is similar to that of the TF 1313, the modification for injection of polarizing voltage involves the addition of these terminals. A rather elaborate modification is thus necessary if the bridge is to be suitable for low Q capacitors. If the use of the bridge is restricted to the higher Q range, however, the same modification can be applied as for the TF 1313.

PRESIDENT OF S.I.M.A.

In July 1962, Mr. R. E. Burnett, Managing Director of Marconi Instruments Ltd., commenced a strenuous year as President of the Scientific Instruments Manufacturers' Association of Great Britain. Membership of the Association (S.I.M.A. for short) consists of nearly two hundred British companies, whose products range from telescopes and microscopes to computers and electronic test equipment and includes every kind of scientific instrument.

S.I.M.A. was formed in 1937 to enable companies to act together in matters of common interest and committees are operating to deal with all the problems of trading, standards, education and relations that affect members. One of the current problems that will actively concern Mr. Burnett is the possible advent of the Common Market. Consultations on this and similar problems have taken place with Government Departments so that they may know the views of the industry on the possible impact of trade treaties with other countries. Trade Associations of the six European Common Market countries have already held preliminary discussions with S.I.M.A. on the possibility of mutual co-operation.

This fast-growing and important industry, for which Mr. Burnett will now be the voice, produced equipment to the value of £190 million last year, an increase of 16 per cent above 1960.

Summaries of Articles appearing in this issue

RESUME D'ARTICLES PUBLIES DANS LE PRESENT NUMERO

WATTMETRE HF A RADIATION TF 1396A

Cet instrument a été conçu pour répondre aux principales recommendations des autorités postales britanniques concernant la mesure des radiations intenses de fréquence radioélectrique qui peuvent avoir des effets dangereux sur le corps humain. Il est composé d'un wattmètre à thermistor utilisable pour la mesure des variations de 10 MHz à 10 GHz et il est destiné à fonctionner grâce aux antennes et aux atténuateurs conçus pour satisfaire les exigences des clients concernant la fréquence spécifique. Utilisé de cette façon, le wattmètre est capable d'indiquer l'intensité de radiation d'une gamme de 1 à 20 mW par cm² avec une marge d'erreur qui n'excède pas +2 dB.

Page 147

ANTENNES POUR WATTMETRE HF A RADIATION TF 1396A

L'auteur y décrit une gamme d'antennes à utiliser avec le TF 1396A. Après avoir examiné l'utilité de telles antennes, il donne une description de cornets pour les bandes S et X et d'une antenne disque-cone pour la bande L. Une discussion concernant les problèmes de construction électrique et mécanique est suivie d'un examen du rendement de ces antennes.

Page 150

Y A-T-IL UNE NECESSITE POUR UN GENERATEUR A HAUTE PUISSANCE DE SORTIE?

Les essais à deux signaux et les mesures en boucle des récepteurs exigent normalement de hautes puissances de sortie. En utilisant les techniques décrites, l'emploi des générateurs à basse sortie sera souvent suffisant.

Page 156

MESURE DE PUISSANCE C.A. UTILISANT UN OSCILLOSCOPE

La puissance dissipée dans une composante par un courant non-sinusoïdal est mesurée par une méthode calorimétrique peu commode à utiliser. Dans la methode plus pratique de l'oscilloscope, la tension qui passe par l'instrument a l'essai, est appliquée à une paire des plaques de déviation et la tension qui passe par un condensateur en serie est appliquée à l'autre paire de plaques. La puissance moyenne peut être calculée de la surface du boucle fermé resultant qui apparaît sur l'oscilloscope.

Page 157

VOLTMETRE A FAUX ZERO TF 1377

Ce voltmètre a été conçu pour la mesure précise des variations très faibles en courant continu jusqu'à 500 V. Des variations aussi faibles que 1 mV y peuvent être observées et des mesures absolues des tensions peuvent être effectuées jusqu'à 100 V avec une précision de 1 %.

Pour l'emploi dans les mesures des dérivations de tension il a été prévu un abaque d'enregistrement.

Page 159

PONTS UNIVERSELS TF 1313 ET TF 868B MESURES DE CONDENSATEURS ELECTROLYTIQUES A HAUTE VALEUR

Il y a une demande croissante pour des ponts permettant la mesure des condensateurs électrolytiques à haute valeur. De simples modifications apportées aux ponts universels Marconi peuvent augmenter la gamme de capacitance et faciliter l'application d'une tension de polarisation.

Page 163

ZUSAMMENFASSUNG DER IN DIESER NUMMER ERSCHEINENDEN BEITRÄGE

HOCHFREQUENZ-STRAHLUNGSLEISTUNGSMESSER TYP TF 1396A

Dieses Gerät wurde entsprechend den wichtigsten Empfehlungen der britischen Postverwaltung zur Messung starker Hochfrequenz-Strahlungsfelder gebaut, die unter Umständen nachteilige Effekte auf den menschlichen Körper haben können. Das Gerät besteht aus einem Thermistor-Leistungsmesser, der sich für den Frequenzbereich von 10 MHz bis 10 GHz eignet und der über Dämpfungsglieder von Antennen gespeist wird, welche für die vom Käufer gewünschten Frequenzen gebaut sind. In dieser Kombination kann das Messgerät für die Strahlungsdichteanzeige im Bereich von 1 bis 20 mW/cm² benutzt werden, wobei der Fehler den Wert von +2 dB nicht übersteigt.

Seite 147

ANTENNEN FÜR DAS HOCHFREQUENZ-STRAHLUNGS-MESSGERÄT TF 1396A

Eine Antennenserie für das Gerät TF 1396A wird beschrieben. Auf eine Betrachtung der an solche Antennen gestellten Anforderungen folgt eine Beschreibung von Sektor-Hornantennen für das S-Band und das X-Band, sowie einer Doppelkonusantenne für das L-Band. Die elektrischen und mechanischen Grundlagen der Konstruktion werden besprochen und die Leistungsfähigkeit untersucht.

Seite 150

IST EINE HOHE AUSGANGSLEISTUNG BEI MESS-SENDERN WIRKLICH NOTWENDIG?

Bei Doppelsignal-Messungen und Schleifenmessungen an Empfängern werden gewöhnlich hohe Ausgangsleistungen von Mess-Sendern verlangt. Mess-Sender mit geringerer Ausgangsleistung genügen jedoch, wenn die beschriebenen Methoden benutzt werden.

Seite 156

WECHSELSTROM-LEISTUNGSMESSUNGEN MIT EINEM KATHODENSTRAHLOSZILLOGRAPH

Die bei nicht sinusförmigem Strom von einem Schaltungsteil aufgenommene Leistung wird normalerweise mit einem umständlichen kalorimetrischen Verfahren gemessen. Bei der bequemeren Oszillographenmethode wird die an dem Schaltungsteil liegende Spannung an ein Ablenkplattenpaar und die Spannung an einem in Serie liegenden Kondensator an das andere Paar gelegt. Die mittlere Leistung wird aus der Fläche berechnet, die auf dem Bildschirm von der geschlossenen Kurve eingeschlossen wird.

Seite 157

DER SPANNUNGSMESSER TF 1377 MIT UNTERDRÜCKTEM NULLPUNKT

Ein Gleichspannungsmessinstrument zur Messung kleiner Gleichspannungsänderungen bis zu 500 V wird beschrieben. Schwankungen bis herunter auf 1 mV bei 100 V können abgelesen werden und absolute Spannungsmessungen bis 100 V können mit einer Genauigkeit von 1 % durchgeführt werden. Bei Spannungsschwankungsmessungen kann ein Registrierinstrument angeschlossen werden.

Seite 159

DIE UNIVERSAL-MESSBRÜCKEN TF 1313 UND TF 868B MESSUNGEN AN ELEKTROLYTKONDENSATOREN HOHER KAPAZITÄT

Es besteht ein steigender Bedarf an Messbrücken, die sich zur Messung von Elektrolytkondensatoren hoher Kapazität eignen. Durch einfache Änderungen an den Marconi Universal-Messbrücken lässt sich der Kapazitätsbereich erweitern und das Anlegen einer Polarisationsspannung leichter durchführen.

Seite 163

SOMMARIO DEGLI ARTICOLI PUBBLICATI IN QUESTO NUMERO

MISURATORE DELLA POTENZA DI RADIAZIONE R.F. TF 1396 A

Questo strumento è stato progettato per soddisfare le raccomandazioni principali della Poste e Telecomunicazioni Britanniche concernenti la misura di intense radiazioni di radio frequenza, le quali possono avere effetti dannosi sul corpo umano. Consiste di un misuratore di potenza a termistor, adatto per frequenze comprese tra 10 MHz e 10 GHz, ed è previsto per essere alimentato de antenne e attenuatori progettati a seconda delle frequenze specificate dai clienti. Quando venga usato con tali antenne e attenuatori, lo strumento è capace di indicare valori dell'intensità di radiazione compresi nell'intervallo da 1 a 20 mW/cm^2 con un errore che non supera i 2 dB. Pagina 147

ANTENNE PER IL MISURATORE DI POTENZA RF, TF 1396A

Viene descritta una serie di antenne da adoperarsi con il TF 1396A. Le caratteristiche che si richiedono in tali antenne vengono considerate, e antenne a settore di tromba per le bande S e X ed una del tipo discocono per la banda L vengono descritte. Problemi della progettazione elettrice e meccanica vengono discussi e le caratteristiche operative vengono esaminate. Pagina 150

E' SEMPRE NECESSARIO UN GENERATORE DI SEGNALI DI ELEVATA POTENZA D'USCITA?

Prove su ricevitori secondo lo schema a due segnali o in circuito chiuso richiedono di regola elevate potenze d'uscita dei generatori di segnali. Qualore si faccia uso dei metodi qui descritti, generatori di potenza più bassa risultano adeguati. Pagina 156

LA MISURA DELLA POTENZA IN CORRENTE ALTERNATA A MEZZO DI UN OSCILLOSCOPIO

La potenza in un componente da una corrente non sinusoidale viene di solito misurata a mezzo del metodo calorimetrico, che è poco practico. Il metodo ben più pratico, qui descritto, consiste nell'applicare la tensione che si trova ai capi del componente in prova a una delle coppie di placche di un oscilloscopio, e la tensione ai capi di un condensatore in serie all'altra coppia. La potenza media può calcolarsi a partire dall'area della curva chiusa tracciata dall'oscilloscopio. Pagina 157

VOLTMETRO A ZERO SOPPRESSO TF 1377

Viene descritto un voltmetro c.c. progettato per misure di precisione di piccolissime variazioni di tensioni continue fino a 500 V. Possono osservarsi variazioni fino a 1 mV in 100 V, e possono farsi misure assolute di tensione fino a 100 V con una precisione dell' 1 per cento. Per misure continue di fluttuazioni di tensione è prevista la connessione a uno strumento registratore.

Pagina 159

PONTI UNIVERSALI TF 1313 e TF 868B MISURE DI CONDENSATORI ELETTROLITICI DI ALTO VALORE

Ponti adatti per misurare condensatori elettrolitici aventi valori alti di capacità vengono richiesti da sempre più parti. Semplici modifiche ai ponti universali Marconi possono allargare l'intervallo di capacità e facilitano l'applicazione di una tensione di polarizzazione. Pagina 163

RESUMENES DE ARTICULOS QUE APARECEN EN ESTE NUMERO

MEDIDOR DE RADIACIONES EN RF TF 1396A

Este instrumento ha sido desarrollado con el fin de satisfacer las recomendaciones de las autoridades británicas, con relación a las medidas de radiaciones de mayor intensidad que puedan dañar el cuerpo humano.

Consiste de un medidor de potencia térmica adecuado para el uso en la gama de frecuencias desde 10 Mc/s hasta 10 Gc/s, y es la intención de que se alimente con antenas y atenuadores que sean desarrollado particularmente para satisfacer los requerimientos de los clientes.

Usado así el medidor es capaz de indicar radiaciones con intensidades desde 1 hasta 20 mW por cm^2 con un error que no es más de 2 dB. Página 147

ANTENAS PARA USAR CON EL MEDIDOR DE POTENCIA RADIADA EN R.F. TF 1396A

Se describen varias antenas para usar con el medidor TF 1396A. Después de considerar los requerimientos de estas antenas se da una descripción de antenas de bocina para las bandas S y X y una antena de tipo disco-cono para la banda L. Se discuten los problemas eléctricos y mecánicos del diseño y se examina el rendimiento de estas antenas. Página 150

¿ ES NECESARIO QUE SIEMPRE SE USE UN GENERADOR DE SEÑALES CON MAYOR SALIDA DE POTENCIA ?

Las pruebas de receptores con dos señales y pruebas con bobina de marco, mayormente necesitan que la salida de los generadores sea de mayor potencia.

Con el uso de técnicas que aquí se describen, generadores con salida de menor potencia, pueden resultar muchas veces suficiente.

Página 156

MEDIDAS DE POTENCIA EN C.A. CON UN OSCILOSCOPIO

La potencia disipada en un componente por una corriente sinusoidal es mayormente medida por un método que es algo penoso.

En este sistema, mucho más conveniente, con un osciloscopio el voltaje através del componente es aplicado a un par de placas del osciloscopio y al otro par se le aplica el voltaje através de un condensador en serie.

La potencia media se calcula midiendo el área del cuadro resultante que se muestra en el osciloscopio.

Página 157

VOLTIMETRO DE CERO SUPRIMIDO TF 1377

Se describe un voltímetro de c.c. diseñado para medir cambios muy pequeños en tensiones directos hasta 500 voltios. Variaciones muy pequeñas como de 1 mV en 100 Voltios se pueden observar y medidas exactas de tensiones hasta 100 voltios se pueden hacer con una exactidud de 1%. Para medir las desviaciones de voltaje hay provisión para usar el instrumento con un registrador de carta.

Página 159

PUENTES UNIVERSALES TF 1313 y TF 868B MEDICIONES DE CONDENSADORES ELECTROLITICOS DE MAYOR VALORES

Existe la necesidad cada vez más para puentes que puedan medir condensadores electrolíticos de mayor valores. Con unas modificaciones sencillas a las Puentes Universales Marconi se puede aumentar el margen de capacidad y facilitar la aplicación de la tensión de polarización.

Página 163

SIGNAL GENERATORS

NOISE GENERATORS

OSCILLATORS

OSCILLOSCOPES

PULSE GENERATORS

MODULATION METERS

DISTORTION ANALYSERS

SPECTRUM ANALYSERS

FREQUENCY COUNTERS AND METERS

POWER METERS

VOLTMETERS

IMPEDANCE BRIDGES, DIELECTRIC AND Q METERS

MOISTURE METERS

pH METERS

IMAGE AMPLIFIERS FOR INDUSTRIAL AND MEDICAL X-RAY EQUIPMENT

ATOMIC POWER INSTRUMENTATION

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