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Transistorization

A WELL-KNOWN FEATURE of technological progress is a feeling of diminishing satisfaction with material possessions as new models supersede them. This is particularly true of electronic equipment, which is still subject to rapid genuine development, as distinct from styling 'improvements' and other gimmicks of planned obsolescence. What was once a thoroughly satisfactory device seems to become increasingly inconvenient, sometimes to the point of drudgery—an observation which can be confirmed in the homes of many electronic engineers, who are generally not leading beneficiaries of the affluent society. How bulky and heavy that old television set now seems; what an irritating length of time the 'table-model' receiver takes to warm up when you want to hear the weather forecast; and even if the owner is strong enough to move them, there remains the tyranny of the mains lead to restrict their field of movement.

On the other hand, the new impedance bridge on p. 3 of this journal will be seen to have some obvious advantages—it is small and light, has zero warm-up time and very low power consumption, and is completely isolated from earth. Even the least sophisticated reader will recognize that these features come from using transistors, the least experienced user will appreciate the extra operating convenience, and the most accident-prone service engineer will be relieved at the freedom from lethal voltages. But they might also wonder why it is that, fifteen years after the invention of the transistor and with some six thousand types now commercially available, there are not more transistorized instruments on the market.

Why should a competitive, progressive industry like electronic instrumentation appear to stick conservatively to the thermionic valve when the transistor offers such advantages? There are good reasons for this apparent backwardness. First, it must be remembered that instruments must always be a step ahead in performance of the equipment they are measuring, but many transistors are still inferior to valves in such features as stability, noise and frequency characteristic. Furthermore, a transistor is inherently a current-operated low-impedance device while most measurements require a voltage-sensing instrument with high input impedance. Last, but not least, is the economic consideration—the extra convenience of the transistorized instrument might be outweighed by the extra cost.

It would therefore have been unwise to rush into transistorization until these drawbacks were overcome. This has been done in part by progressive improvement in transistors themselves, which are steadily increasing in performance, tightening in tolerance, and decreasing in price. But it has also called for a new design approach—it is seldom just a case of replacing valves with transistors on a one-for-one basis, since this usually results in inferior performance; on the other hand, the size and sheer convenience of transistors allow the designer freedom to include more elaborate circuitry than was practical with valves, and this can more than compensate for any initial inferiority of the transistor, although at the cost of some increased complexity. Likewise in environmental considerations, an awareness of the transistor's limitations has led to greater design precautions, backed up by extensive environmental testing in order to ensure a panclimatic performance comparable with the best valve circuits.

Gaining on the swings what you lose on the roundabouts is unlikely to happen at the fun fair but it is certainly one of the effects of using transistors instead of valves in an electronic instrument (*Photograph by courtesy of Butlin's Limited*)

Turning to particular types of instrument, the frequency counter was one of the first to benefit from transistors on account of their superior switching characteristics and suitability for use on stacked printed boards; they also have a better on/off ratio and are more tolerant of variations in supply and drive level, and drift in value of associated components. These features of transistors had already been utilized in the design of computers, whose influence can be seen in the logic circuits of the TF 1417 counter described in the December, 1961, issue of this journal. Here is a good example of transistors contributing to great savings in size, weight and heating, and improvement in stability and reliability.

A less obvious field for exploiting transistors is the measuring oscilloscope with its need for high input impedance, d.c.-coupled amplifiers and e.h.t. supply voltages. Nor can any dramatic reduction in size be expected since this is limited by the cathode ray tube and the number of front panel controls. This is the type of instrument that lends itself to hybrid circuits using both transistors and valves.

Signal sources such as oscillators and signal generators may not appear to be ideal subjects for transistorization. Problems here include the need for linearity over a wide range of signal levels and high-grade performance over a frequency band often extending beyond the scope of transistors. However, the prospect of miniaturization, fast warm-up, low hum content, freedom from microphony and general ruggedness promised by transistors cannot stay unrealized for long.

Eventually, then, we are likely to see a trend towards transistorized instruments in this journal since it is our policy to introduce them as it becomes advantageous, and technically and economically sound, to do so. They will represent a high investment of research, design know-how and pre-production testing, but promise to open up a new era of performance, versatility and convenience in electronic instrumentation. J. R. H.

A 1% Universal Bridge . . . . . . TYPE TF 2700

by E. C. CRAWFORD,
Graduate I.E.E.

TF 2700 is a small economically priced bridge of 1% basic accuracy for measuring inductance, capacitance and resistance.
This article describes the constructional features of the bridge, touching upon the details of both the electrical and mechanical design, and then gives examples of some of the more unusual experiments in impedance measurement which the layout of this bridge particularly facilitates.

NO ELECTRICAL engineering laboratory or design establishment should these days be without the facility for making precise measurements of components such as can be done with the $\frac{1}{4}$% Universal Bridge TF 1313. But many engineers and designers will prefer to have a 'personal' bridge more readily to hand for routine work, and the prerequisites for such a bridge are perhaps summed up by the words compactness, convenience, accuracy, versatility and economy.

The type of bridge circuitry used is traditional in design, using the Wheatstone bridge for resistance, the Hay or Maxwell for inductance and the capacitance-resistance bridge of series or parallel form for capacitors; Fig. 2 illustrates these basic circuits. As the oscillator and the detector amplifier are transistorized and consume very little power (about 70 mW) a battery is adequate for energizing these as well as the d.c. Wheatstone bridge. This makes for portability and also for versatility as will be shown later. It may be compared in size and appearance with the established TF 868B Bridge, both being shown together in Fig. 1.

The ratio arm components, together with the function switch are mounted on the front panel, while the printed circuit amplifier and the oscillator with its isolating transformer are mounted on the rigidly supported side frames. At the rear is the seamless plastic battery box. The type of battery supplied and recommended for replacement is of layer cell construction used throughout the world in transistor radios; it has an excellent shelf life, large capacity for its size and a great reluctance to leak and corrode its surroundings. Should, however, the completely discharged battery be inadvertently left in its box, the plastic will protect the instrument against any leaks and will itself be easy to clean out.

In a pocket at the rear of the instrument is an illustrated instruction book giving details of both the ordinary and the extraordinary uses of the bridge. This is not provided because the instrument is particularly difficult to use, but because it draws attention to facilities which might otherwise be unappreciated. The diagrams show the external connections necessary and the appropriate limiting conditions in special cases are tabulated.

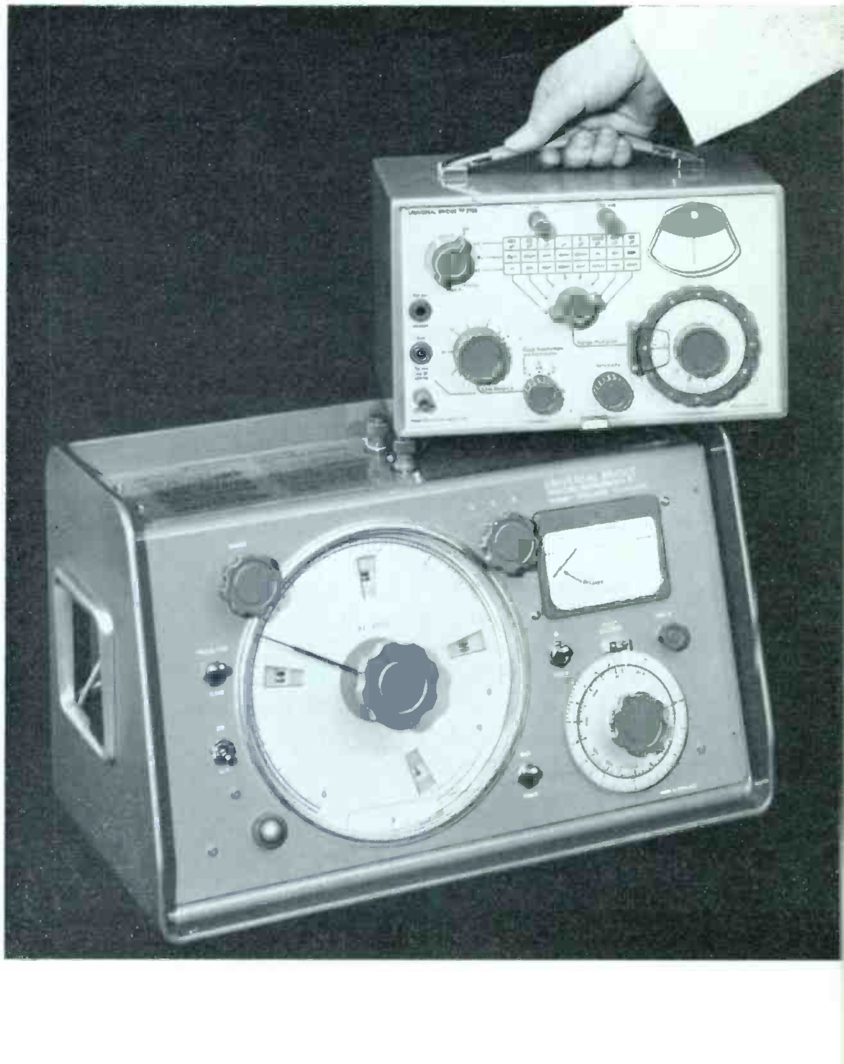
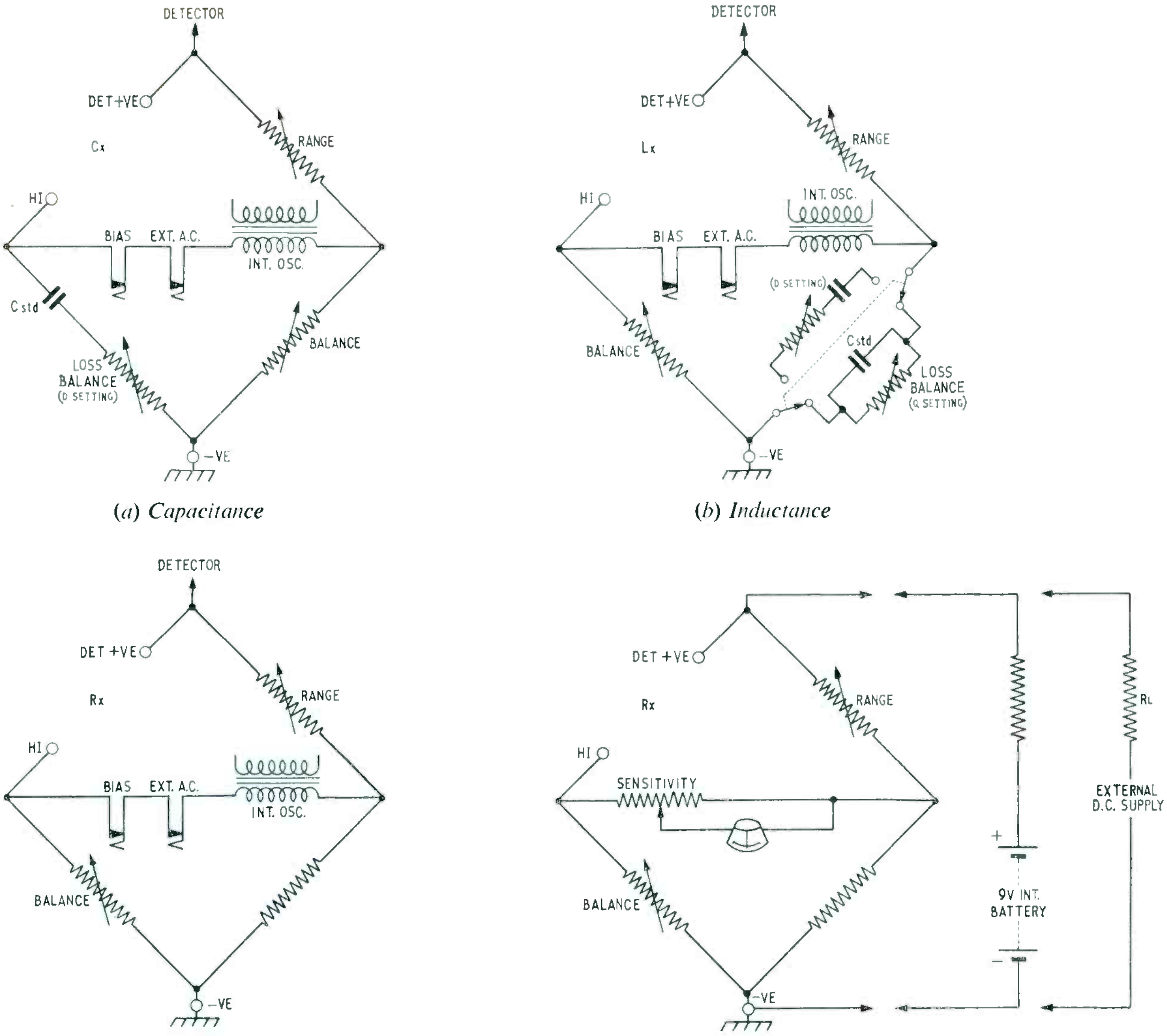


Fig. 1. TF 2700 with its predecessor TF 868B; an obvious gain—dimensionally. Closer comparison will also show a gain in performance



(*a*) *Capacitance* (*b*) *Inductance*

(*c*) *Resistance* (*A.C.*) (*d*) *Resistance* (*D.C.*)

Fig. 2. Basic bridge circuits

Balance Controls

Those readers who are familiar with the $\frac{1}{4}$% Universal Bridge TF 1313 will appreciate a display system which eliminates the guesswork of dial interpolation when it comes to reading the value indicated. The major drawback of the single dial system is its lack of resolution unless the dial is of very large size such as in the TF 868B. This has a main dial 8 inches in diameter. The total height of the front panel in the TF 2700, however, is only 6½ inches so that the possibility of a useful single dial system was automatically ruled out.

Instead of a single variable resistor a switched resistor decade is used together with an interpolating fine balance variable. The switch is controlled by the large rim-type knob which surrounds the inner dial and knob of the fine balance, and it can be seen in Fig. 3 that the cursor is arranged so that the answer appears in an in-line form as a convenient multiplier to the full scale range value indicated by the Range and Function switches.

It is of interest to compare the accuracy of this readout system with that obtained previously with the much larger 8-inch dial on the TF 868B. There are two critical points on a dial read-out system which should be studied; these are the full scale point and the point on the scale corresponding to full scale on the next lowest range—the range crossover point. With decade range switching these two points are at 100% and 10% of the full scale. The readability of intermediate values between 10 and 100% will, of course, lie between these two extremes. Below 10% it is customary to use a lower range for improved discrimination.

At full scale the TF 868B calibration extends for 0·110 inch per 1% change, corresponding to 1·5° rotation. As the TF 2700 uses a decade switch, its calibrated dial covers only 10% of the full range, so that 1% change at full scale gives 25° rotation or 0·6 inch of scale length. At 10% of full scale the logarithmic calibration of TF 868B has only 0·04 inch of scale per 1% change in

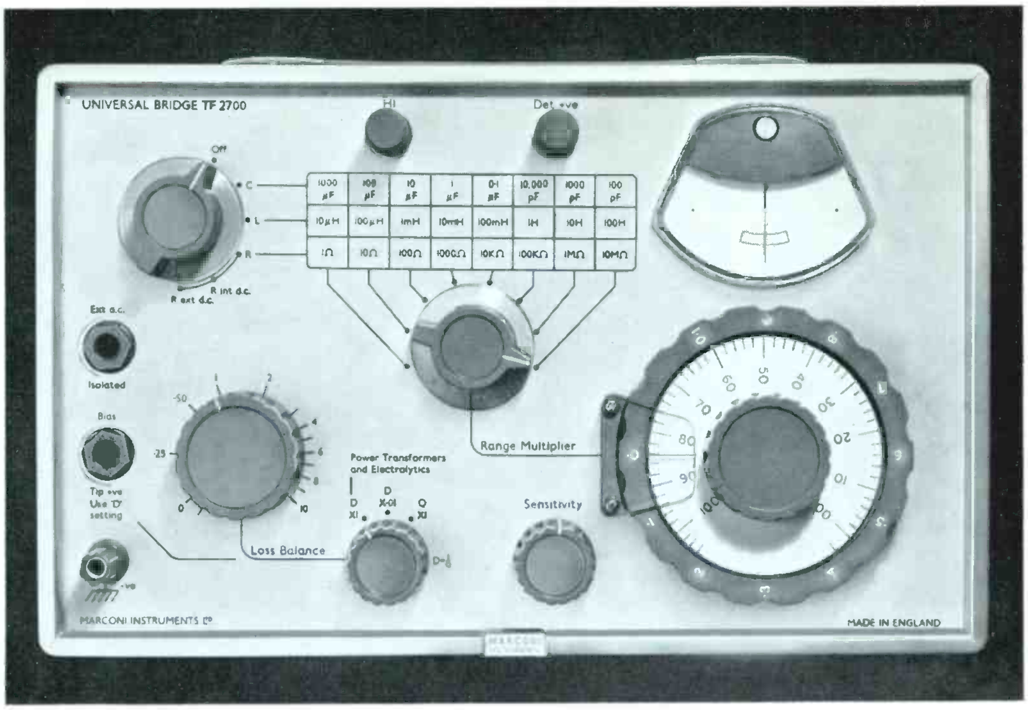


Fig. 3.
Dual concentric balance controls make high discrimination reading a feature of the Universal Bridge, TF 2700

value, realized by 0·6° of rotation; the TF 2700 decade system however is linear and 1% change is represented by 0·06 inch of scale or 2·5° of rotation. In terms of rotation or 'feel' the advantage of TF 2700 is between 4 and 16 times and, despite the much smaller dimensions, the scale length corresponding to 1% is always greater by between $1\frac{1}{2}$ to 5 times.

This superior readout means that comparisons between components can be made to much better accuracy than ever before on a 1% bridge.

The switched decade is not extravagant in resistors. Only four resistors are used to give every value in hundreds up to 1000 Ω. Instead of the usual 1, 2, 4, 8 sequence we use 1, 2, 3 and 6 as this not only uses lower power resistors but is more convenient to switch in the required manner. 4 is given by 3 + 1 and 5 by 3 + 2 and so on.

In addition to adjustment to within 0·1%, the resistors are arranged to have the correct inductance to match the stray capacitance in the circuit. This assists in making measurements at high audio frequencies when too much reactance makes it difficult to obtain a good sharp balance.

Bridge Ratio Arms

The range arm is switched in decade values from 0·1 Ω to 1 MΩ. Higher values use metal film type resistors, while the two lower values are wire wound to realize the required accuracy. The 0·1 Ω resistor in particular must be adjusted *in situ* to allow for all the various series switch contact resistances; the latter being reduced by paralleling up of contacts to the practical limit where further paralleling would give negligible extra stability.

0·1 Ω is an unusually low value for a resistor in this position, but it was included to extend the capacitance range of the bridge to 1000 µF. With this extension of range automatically comes a resistance range of 1 Ω full scale and inductance of 10 µH full scale. The 1000 µF range is frequently required whereas the low resistance and inductance ranges are not necessarily so to the general user. However, there are occasions when measurements of such low values are useful.

The higher values of range resistor are susceptible to shunt capacitance and precautions are taken in the wiring and the arrangement of the switching so that this capacitance is minimized.

As this bridge is intended to be as light, compact and economical as possible, consistent with the main objective of 1% basic accuracy for L, C and R, it was decided at the outset that defined accuracy for Q or power factor would not be included in the specification. This leads to considerable simplification of the 'range resistor' arm as precise phase compensation of the individual resistors is no longer required. Reasonable limits, however, must still be maintained or else difficulties in balancing the main component may arise. With the exception of the lowest range it was decided to keep the stray inductance or capacitance within limits such that at a Q of 1 at 1 kc/s the additional L or C error would not exceed 0·25% due to the effect. The lowest range is rather exceptional as there is so much inevitable inductance in the wiring to the switch compared with the 0·1 Ω range resistor. By and large, however, the error when measuring electrolytic capacitors of hundreds of microfarads is insignificant.

The capacitance standard against which capacitors and inductors are compared is an accurate 0·1 µF polystyrene type with variable resistors arranged either in series or parallel to simulate the loss of the component

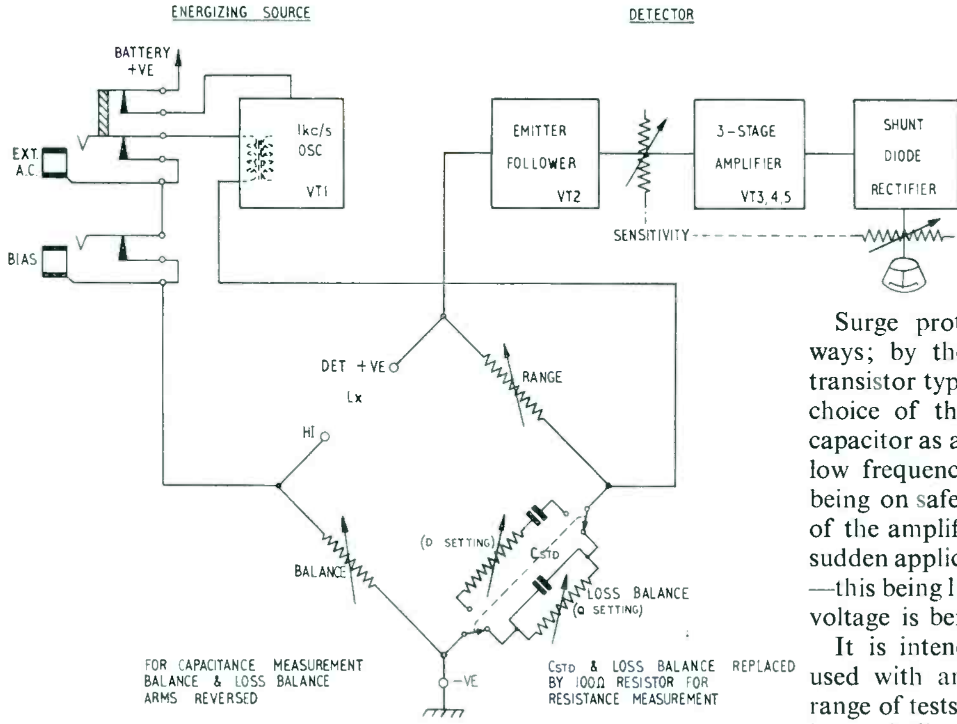


Fig. 4. Functional diagram of complete instrument

on test. As was previously mentioned a defined Q accuracy is not included in the specification, but a calibration corresponding to the average of the variable resistor used is printed on the front panel so that Q settings of different components on the same bridge may be compared against one another.

With the extended electrolytic capacitor range has also come an extended tan δ or D range (roughly the same as power factor). This is especially helpful as low frequency measurements of electrolytics may quite readily be made and the common D range of 0 to 0·1 is often not sufficient.

Detector and Oscillator

Between two opposite corners of the bridge network there is an impedance of about 1 MΩ maximum ranging down to almost zero at extreme range limits. This is necessarily either the load for the oscillator drive or the source impedance to the detector. As it is easier to provide a balanced e.m.f. to drive the bridge than a balanced high impedance detector, it follows that the detector has to contend with the range of source resistance mentioned. With mains operated valve equipment there is virtually no problem, but transistors tend to have two fundamental drawbacks when used as input stages to amplifiers. First their natural input resistance is too low, and second they are liable to accidental burn-out by surge overload. The naturally low input resistance means that the bridge sensitivity decreases as the range impedance increases—this is overcome by two expedients. First is to use an emitter follower for the input stage so that an increased input resistance results. Second is to increase the bridge e.m.f. on the two highest ranges as extra compensation. By these means approximately constant sensitivity is achieved.

Surge protection is also effected in two ways; by the choice of a robust junction transistor type for the input stage and by the choice of the value of the input coupling capacitor as a compromise between safety and low frequency response, the main emphasis being on safety. The criterion was the ability of the amplifier input stage to withstand the sudden application of a d.c. potential of 500 V—this being liable to happen when bias of that voltage is being applied to test components.

It is intended that the bridge should be used with any audio frequency for a wide range of tests and for this reason the detector is aperiodic, continuous tuning being quite impractical in this instrument. The basic amplifier has a nearly uniform response from 20 c/s to over 20 kc/s but the surge limiting input capacitor cuts the 20 c/s gain by about 20 dB. This may be readily allowed for by increasing the external a.f. drive if necessary.

The oscillator uses a single transistor with tuned collector and mutual inductance coupling to the base, operating in class C. Such a circuit is not critically dependent upon battery voltage or choice of transistor for operation and with a high Q tuned circuit the harmonic level is satisfactorily low. A low harmonic level is necessary for measuring components which are frequency conscious and when detecting the unbalance with an aperiodic amplifier. The tuned inductor of the oscillator forms the primary of the bridge drive transformer, the secondary being well isolated by an electrostatic screen to avoid other than pure mutual inductive coupling. In order to avoid frequency pulling or other results of too heavy a load on the performance of the oscillator, a resistor is used in series with the secondary as a minimum load.

On most of the ranges the oscillator e.m.f. is only 160 mV. On the two highest impedance ranges an increased voltage is tapped off the transformer to approximately equalize the sensitivity.

Maximum frequency pulling occurs when the load on the secondary is such that the series resistance and reactance are equal. This only occurs on the 100 μF range when 47 μF is on test and here the frequency change is only $\frac{1}{2}$%. On other ranges the change is much less with any load.

The current drain on the 9 V battery to both the amplifier and the oscillator is only $7\frac{1}{2}$ mA, the amplifier taking 5 and the oscillator $2\frac{1}{2}$ mA.

SOME UNUSUAL MEASUREMENTS

Frequency Characteristic of Capacitors

This measurement is carried out using an external variable frequency oscillator and an isolating transformer such as Type TM 7120 which is available as an optional accessory (see Fig. 5). By connecting these to the EXT A.C. jack with the plug supplied, the internal 1 kc/s oscillator is disconnected and the external audio frequency supply is placed in the position formerly occupied by the isolated secondary of the internal bridge drive transformer.

However, it is possible if some care in layout of the equipment is used, that this transformer may be dispensed with. This is only possible because the battery operation of TF 2700 permits its complete isolation from true earth.

The capacitor to be checked is then connected to the test terminals and the bridge controls adjusted for balance at different frequencies. Some results are shown in Fig. 6.

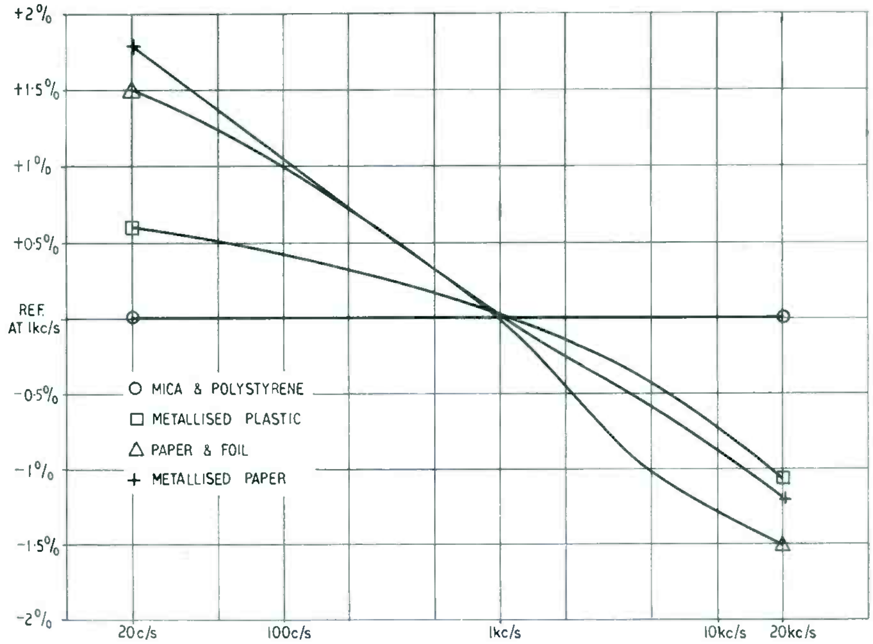


Fig. 6. Frequency characteristics of capacitors

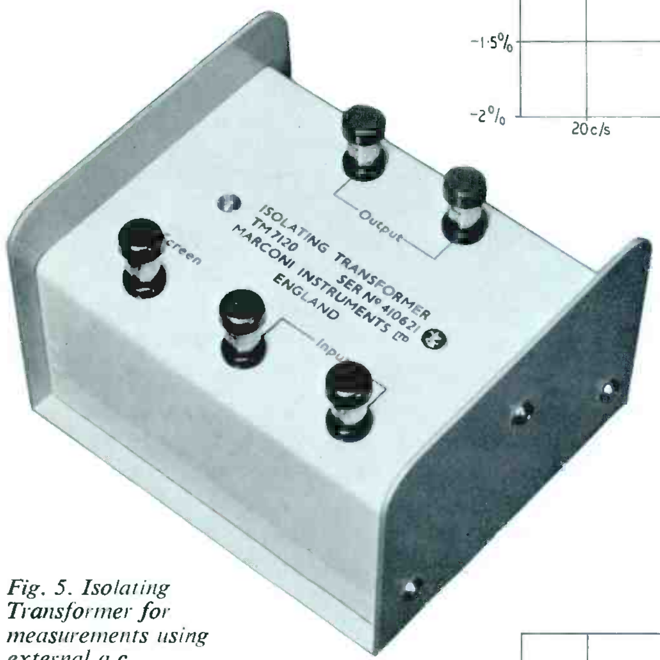
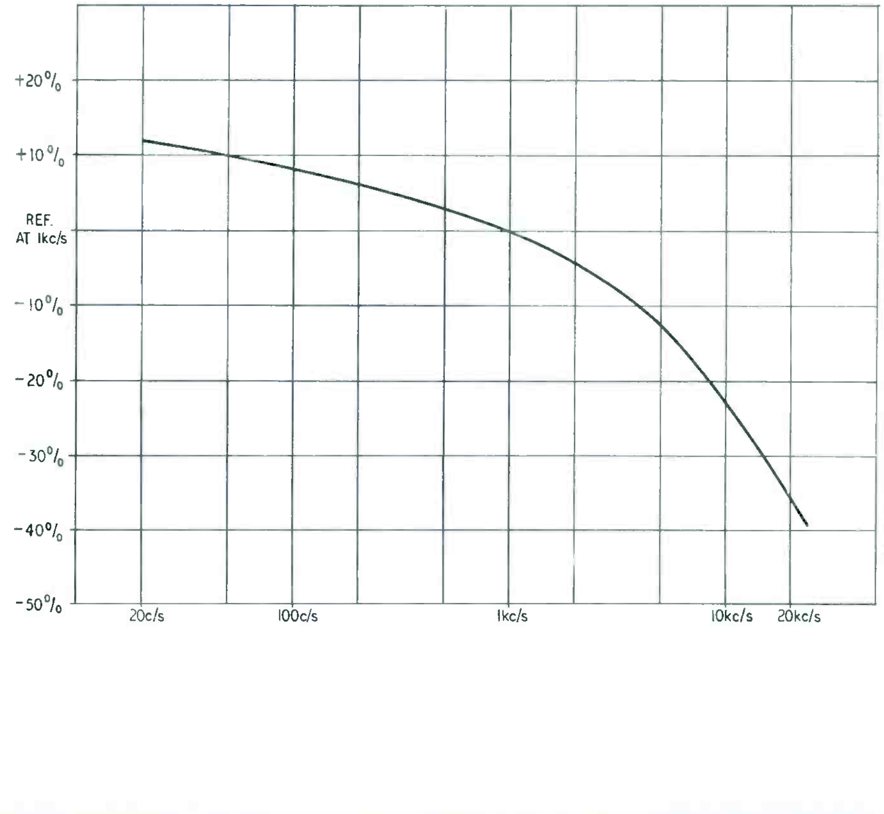


Fig. 5. Isolating Transformer for measurements using external a.c.

It is interesting to note that many ordinary capacitors alter by about –1% between 1 kc/s and 10 kc/s—it is obviously important to be aware of such characteristics when selecting close tolerance capacitors for high frequency use. The change in capacitance is associated with dielectric loss; the low loss capacitors do not show any appreciable change.

Particularly frequency conscious are some electrolytics, which are usually manufactured to have their rated value at mains supply frequency (see Fig. 7).

Fig. 7. Frequency characteristic of a 10 μF 100 V electrolytic capacitor



Voltage Characteristic of Capacitors

In series with the EXT A.C. jack mentioned above is a BIAS jack for use with a polarizing supply of up to 500 V. The proportion of this voltage that appears across the capacitor on test depends only upon its leakage relative to the accompanying range resistor. For the usual capacitor this results in a negligible loss of polarizing voltage which is only 0·1 V for 1 mA/µF leakage. Useful application of this bias is not necessarily limited to electrolytic capacitors, but also to some of the high dielectric constant ceramics and allied types which are also voltage conscious.

Inductance of a Wire-wound Potentiometer

This entails some complicated description so reference will be made to Fig. 8 in which the bridge is switched to the R condition.

Two additional capacitors are required, C_2 fixed and C_1 variable and calibrated. To establish a non-inductive standard of comparison a composition resistor, of almost the same resistance value as the wire-wound potentiometer to be measured, is first balanced on the bridge, keeping C_1 near its minimum value by a convenient choice of C_2. This is done at the highest audio frequency possible to increase the reactive sensitivity.

The wire-wound potentiometer is substituted for the composition resistor, and the bridge rebalanced by C_1 disturbing the other bridge controls as little as possible. The inductance can then be found by calculation.

Two measurements were made as follows:

(a) Nominal 500 Ω 2 W wire-wound potentiometer.
Actual value, $R_x = 475\ \Omega$ measured at 100 c/s
Carbon resistor used as non-inductive standard, 472 Ω
C_2, fixed, 100 pF
C_1 with carbon resistor, 240 pF at 20 kc/s
C_1 with wire-wound resistor, 750 pF at 20 kc/s
$\Delta C_1 = 750 - 240 = 510$ pF.

C_1 is in parallel with the 100 Ω standard, R_S, in the TF 2700 and the product $\Delta C_1.100\ \Omega$ gives the value of the time constant which just balances the inductive time constant of the wire-wound potentiometer.

Time constant, $C_1 R_S = 510 \times 10^{-12} \times 10^2$.
Therefore $L/R_x = 5{\cdot}1 \times 10^{-8}$
and as $R_x = 475\ \Omega$
$L = 475 \times 5{\cdot}1 \times 10^{-8} \times 10^6\ \mu\text{H}$
$= 24{\cdot}2\ \mu\text{H}$.

(b) The second example was a nominal 50 Ω wire-wound potentiometer. C_2 was a fixed 100 pF as before and the carbon standard was a 48 Ω ¼ W.

ΔC_1 was found to be 870 pF and R_x to be 49·3 Ω
$\therefore L/R_x = 870 \times 10^{-12} \times 10^2 = 8{\cdot}7 \times 10^{-8}$
and $L = 49{\cdot}3 \times 8{\cdot}7 \times 10^{-8} \times 10^6\ \mu\text{H}$
$= 4{\cdot}28\ \mu\text{H}$.

Lower value potentiometers are usually relatively more inductive because the thicker wire used to maintain the resolution involves more turns per ohm.

It can be shown that the impedance of the first example will have increased by only 1% at 450 kc/s due to the measured inductance and this could be compensated for by a parallel capacitor of 107 pF ($C = L/R^2$).

It would have been extremely difficult if not impossible to measure this inductance by normal means because of the low Q.

At 20 kc/s the Q of the 500 Ω potentiometer is

$$2\pi fL/R = 6{\cdot}28 \times 2 \times 10^4 \times 24 \times 10^{-6} \times 500^{-1} = 0{\cdot}006$$

and this may be compared with the lowest limiting value for Q of 0·03 for the normal type of bridge to be capable of achieving a balance with any pretence to accuracy.

The method outlined may be extended to resistors of much higher relative inductance but the C_1 required may get very large. But using a decade capacitance box of 0·1 µF maximum (10×0·01 µF, 10×0·001 µF and 10×100 pF) for C_1, a comfortable overlap between this technique and the normal inductance bridge techniques may be obtained.

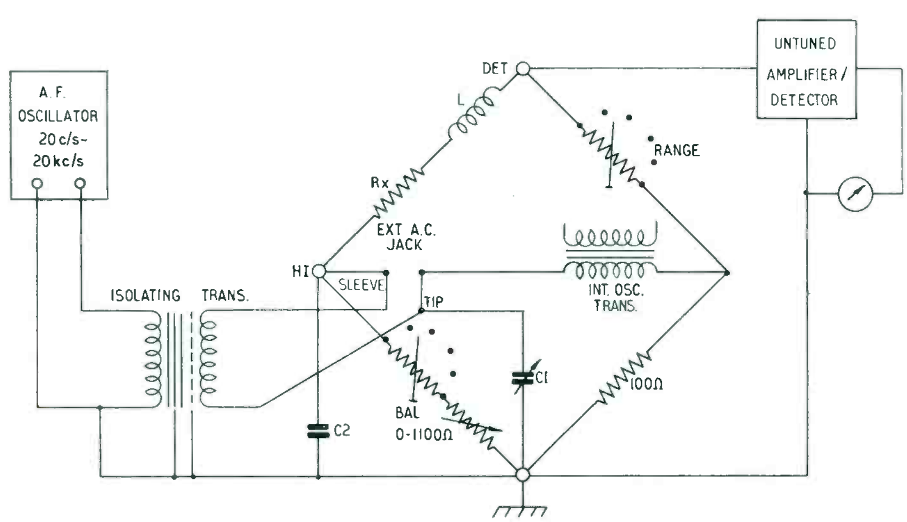


Fig. 8. Measurement of wire-wound potentiometers

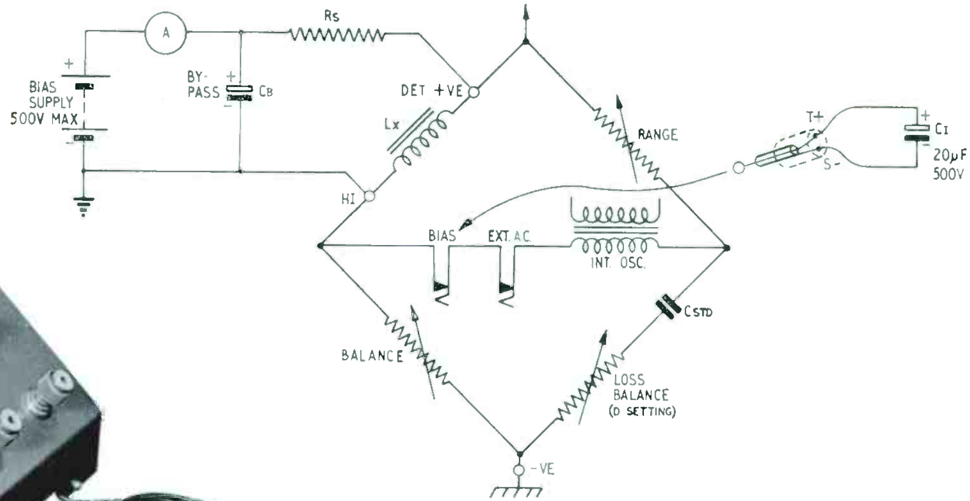


Fig. 9.
A simple method of passing d.c. current through an inductor under test

Fig. 10.
D.C. Choke Adaptor, TM 6113, enables d.c. currents up to 200 mA to be passed through an inductor under test

Variation of Inductance with superimposed d.c.

There are a number of methods of connecting the d.c. supply to the inductor but the one used in this example is shown in Fig. 9.

This method may be used provided R_S is at least 1000 Ω per henry because the extended D range up to 10 permits a wide range of parallel loss to be balanced out. Battery operation of the bridge and the battery supply for the d.c. makes the arrangement insensitive to hand capacitance if a wooden table or bench is used.

The results were as follows:

I	0 mA	2 mA	$4\frac{1}{2}$ mA	8 mA	18 mA
L	3H	1·56H	0·28H	0·14H	0·099H

The core was of mumetal laminations without a gap.

Low Power Resistance Measurements

This last example illustrates one of the advantages of a.c. resistance measurement. When a very low power resistor or, as in this example, a sensitive meter coil is to be measured, the normal d.c. bridge feeds in too much power because of the comparative insensitivity of a conventional d.c. meter as a balance indicator. With low frequency a.c. the resistance found is usually very close to the d.c. value and the advantage of an amplifier detector is available to increase sensitivity without overload.

When using the TF 2700 to measure the d.c. resistance of a universal meter it was obviously subject to severe overload on the mA and μA ranges. Using the internal 1 kc/s source the balance null was masked by the reactance, but at 20 c/s sensible answers were obtained as under:

Meter range	50 μA	250 μA	1 mA	10 mA
Resistance	5 kΩ	500 Ω	180 Ω	18 Ω

Low level a.c. may also be used to advantage when exploring the effect of temperature on sensitive elements as the self-heating is very low and the thermal e.m.f's have no effect upon the balance.

ABRIDGED SPECIFICATION

Ranges

CAPACITANCE: 0·5 pF to 1100 μF in 8 ranges of 110 pF to 1100 μF full scale.

INDUCTANCE: 0·2 μH to 110 H, in 8 ranges of 11 μH to 110 H full scale.

RESISTANCE: 10 mΩ to 11 MΩ in 8 ranges of 1·1 Ω to 11 MΩ full scale.

Q INDICATION: 0 to 10 at 1 kc/s.

D INDICATION: 0 to 0·1 or 0 to 10 at 1 kc/s.

Accuracy

BASIC ACCURACY: ±1% of reading ±0·1% of range full scale.

RESIDUALS:

Co, approx. 0·2 pF.
Lo, approx. 0·2 μH.
Ro, approx. 2 mΩ.

Bridge source

INTERNAL: 1 kc/s oscillator for a.c. measurements of C, L and R.
9 V battery for d.c. measurements of R between 1·2 Ω and 50 kΩ.

EXTERNAL: A.F. from 20 c/s to 20 kc/s may be used for C, L and R measurements.
D.C. may be used for R in place of the internal battery.

Facilities

Non-linear elements may be tested by applying a variable d.c. bias or a.f. signal.

Power supply

Internal 9 V battery, world standard type; usual current drain approximately 7 mA.

Dimensions and weight

Height	*Width*	*Depth*	*Weight*
8 in	$11\frac{1}{2}$ in	8 in	$8\frac{1}{2}$ lb
(20·5 cm)	(29 cm)	(20·5 cm)	(3·8 kg)

MARCONI INSTRUMENTS

621.396.615.12

HIGH OUTPUT VERSIONS OF *Standard Signal Generator* . . . *TYPE TF 144H*

by J. M. PARKYN

TF 144H/4 to -H/6 is a new range in the TF 144H series giving an additional high output facility to supersede the existing TF 144H to -H/2. These new versions offer a direct output of 2·75 V e.m.f. at virtually zero source impedance, which gives 100 mW in a 75 Ω load.

WHILE BELONGING to a line some 28 years old, the TF 144H Standard Signal Generator is in fact a comparatively new design having been introduced in 1960.[1, 2] However, it is already time for a number of new versions offering an additional high output facility to supersede the existing TF 144H series. Before describing the extra facility in detail a list is given below showing the original models related to their successors.

Original model	*New high output version*
TF 144H Civilian type, free standing with bench case	TF 144H/4
TF 144H/1 Civilian type with dust cover for rack mounting	TF 144H/5
TF 144H/S Services type approved with bench case (*CT 452*)	TF 144H/4S (*CT 452A*)
TF 144H/2S Services type approved with bench case and accessories (*CT 452 set*)	TF 144H/6S (*CT 452A set*)

Notwithstanding an article by the same author in the previous issue of *Instrumentation*,[3] there are certain tests which require a power from the signal generator larger than the usual figure. The original series, TF 144H to TF 144H/2S, offered a 2 V e.m.f. with a 50 Ω source impedance and a 2 V direct output suitable for only high impedance loads, e.g. a frequency counter. The new versions deliver an output of 100 mW in contrast with the 20 mW ($E^2/4R$) maximum available power from previous models.

The high output is monitored directly at the output socket and is delivered, therefore, from a virtually zero source impedance. The 100 mW feature has been introduced to facilitate testing of the Marconi's Wireless Telegraph 1 kW output linear r.f. amplifier which is now in widespread use for m.f./h.f. transmission. At the standard SET CARRIER meter deflection with the HIGH/NORMAL OUTPUT switch at HIGH the output from the new versions of TF 144H is 100 mW in 75 Ω, i.e. 2·75 V. In the NORMAL setting the instrument offers all the previous facilities of output level and impedance. A simplified diagram (Fig. 2) shows the additions within the dotted area. The added preset resistor is inside the r.f. box but accessible from outside, having removed the outer case or dust cover. It is set so that when in circuit it lowers the sensitivity of the output meter so that 2·75 instead of 2·0 V is required for SET CARRIER deflection. It could be

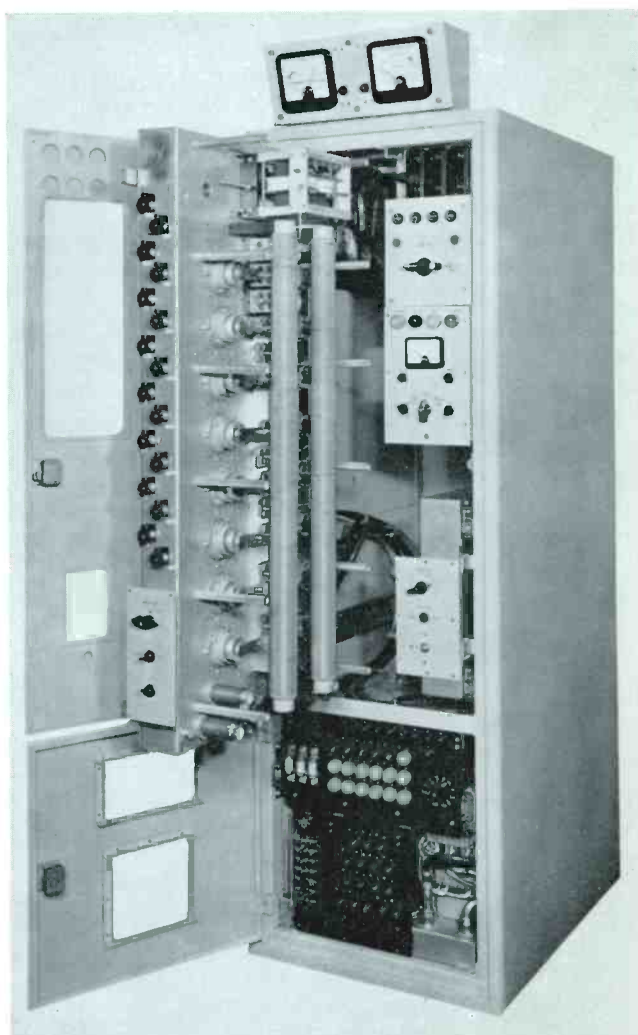


Fig. 1. The Marconi's Wireless Telegraph Co. Ltd. Linear R.F. Amplifier, NT 203, which may be tested with the new version of the TF 144H

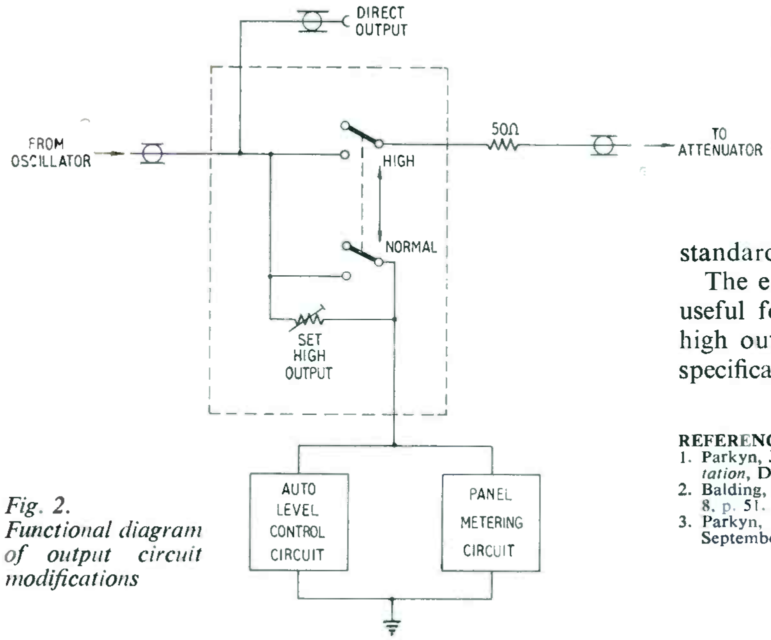


Fig. 2. Functional diagram of output circuit modifications

set for any figure between 2·0 and 3·0 V but standard instruments are supplied with the 2·75 V setting. As will be seen in the diagram the automatic level control monitor is also lowered in sensitivity by the same preset, so the operation of the switch causes very little apparent effect on the output meter since the automatic system maintains very closely the standard deflection.

The extra output from these new models will also be useful for driving bridges and many other uses where high output is essential. For completeness an abridged specification of TF 144H series is given below.

REFERENCES

1. Parkyn, J. M. 'Standard Signal Generator Type TF 144H'. *Marconi Instrumentation*, December 1960, 7, p. 229.
2. Balding, J. 'Standard Signal Generator Type TF 144H'. *Ibid.*, September 1961, 8, p. 51.
3. Parkyn, J. M. 'Is a High Output Signal Generator Always Necessary?' *Ibid.* September 1962, 8, p. 156.

ABRIDGED SPECIFICATION

Standard Signal Generator, Types TF 144H/4, /5, /4S, /6S

RANGE: 10 kc/s to 72 Mc/s, in 12 bands.

CALIBRATION ACCURACY: ±1%.

FINE TUNING: Calibrated directly in % frequency change. Discrimination: 1 division = 0·01%. Total cover: 1%.

CRYSTAL CHECK: 400 kc/s and 2 Mc/s crystals.

Output

IMPEDANCE: 50 Ω at calibrated output; v.s.w.r. better than 1·25: 1.

CALIBRATED OUTPUT: 2 μV to 2 V e.m.f.

Low outputs down to 0·2 μV using 20 dB Pad TM 5573.

DIRECT OUTPUT: Normal: 2 V directly monitored. High: 2·75 V directly monitored.

Modulation

INTERNAL A.M.: 400 c/s and 1 kc/s.

DEPTH: 0 to 80%.

EXTERNAL A.M.: 20 c/s to 20 kc/s.

MARCONI INSTRUMENTS APPLICATION NOTE

621. 376. 33 : 621. 396. 619. 13

PROGRAMMABLE F.M. DEVIATION METER

by W. OLIVER

There is a growing demand in the U.S.A. for programmable test equipment; this indicates a trend towards automation in the electronics industry. Marconi Instruments do not at present manufacture programmed test equipment, but this article indicates a method of programming the TF 791D Carrier Deviation Meter over a limited range of measurement.

RECENTLY a requirement arose for the use of a TF 791D Deviation Meter in a rack installation for production testing of f.m. transmitters. The particular test requirements were as follows:

1. When either a 47·95 or 60 Mc/s r.f. signal was applied from a low power f.m. transmitter, the deviation meter was to have provision for remote local oscillator switching so that no manual tuning operation was needed. The local oscillator was to be crystal controlled to ensure accurate measurement of low deviations.
2. Two f.m. deviation measurements were required at each carrier frequency. These were ±15 kc/s for normal modulation and ±500 c/s for an l.f. tone modulation which would be used to unlock a tone operated squelch gate in an f.m. receiver.

A Marconi Instruments TF 791D Carrier Deviation Meter was the obvious choice of instrument for this application, because it has provision for crystal control of the local oscillator and suitable deviation ranges of ±5, 25, 75 and 125 kc/s full scale. To meet requirement 1 the following conditions had to be met:

(a) An i.f. of 325 kc/s $\pm$10 kc/s must be produced between the incoming signal and the crystal controlled local oscillator to suit the TF 791D.

(b) The local oscillator must operate above the signal frequency for the DEVIATION + and − switch on the TF 791D to read in the correct sense.

It follows that the two local oscillator frequencies should be 48·27 Mc/s $\pm$10 kc/s and 60·325 Mc/s $\pm$10 kc/s. Since 48 and 60 are both factors of 12, an obvious method of meeting the requirements of (a) is to use the 4th and 5th harmonics of a nominal 12 Mc/s crystal. Using a 12·067 Mc/s crystal the 4th harmonic is 48·268 Mc/s and the 5th harmonic 60·335 Mc/s, which are both within the requisite 10 kc/s of the wanted local oscillator frequencies.

By this means requirement 1 was met, since if either 47·95 or 60 Mc/s was applied to the instrument, an i.f. would be produced within $\pm$10 kc/s of nominal without any tuning operation of the TF 791D. The r.f. sensitivity under these conditions was 15 mV at 47·95 Mc/s and 20 mV at 60 Mc/s. It was suggested that 30 mV be applied to ensure adequate limiting under all conditions. It will be noted that the crystal frequency chosen (12·067 Mc/s) is outside the recommended crystal frequency range of the TF 791D (4 to 10 Mc/s). However, no difficulty was experienced with any of the instruments modified.

For requirement 2 deviation measurements of $\pm$500 c/s and $\pm$15 kc/s were needed. This meant that remote switching of deviation from the $\pm$25 kc/s range to the $\pm$5 kc/s range was required to provide accurate results with good discrimination.

The basic circuit of the pulse count discriminator used in the standard TF 791D is shown in Fig. 1. The deviation range is selected by the resistor value switched into circuit by SB, the deviation range switch.

Remote switching of deviation range was accomplished as shown in Fig. 2. A resistor, Rx, was switched in parallel with the $\pm$5 kc/s deviation range resistor, by a relay contact RL1/1. The resistor value was selected so that the parallel combination was equal to 388 Ω which is the value of the $\pm$25 kc/s deviation range resistor.

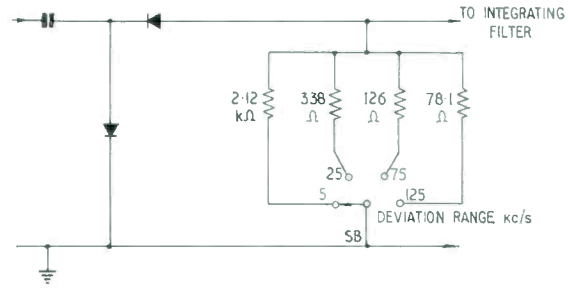


Fig. 1. Basic circuit before modification

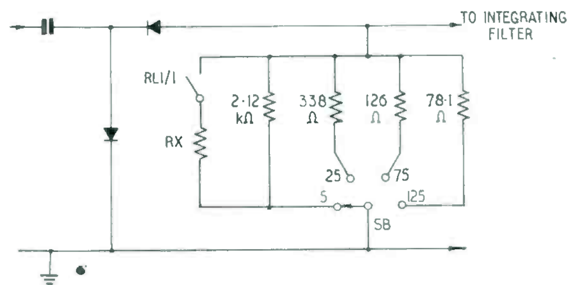


Fig. 2. Basic circuit with remote switching

The deviation meter was then re-standardized using the Bessel Zero method on all deviation ranges and SB was turned to the $\pm$5 kc/s deviation range position. With RL1 energized by an external switch this deviation range became $\pm$25 kc/s full scale.

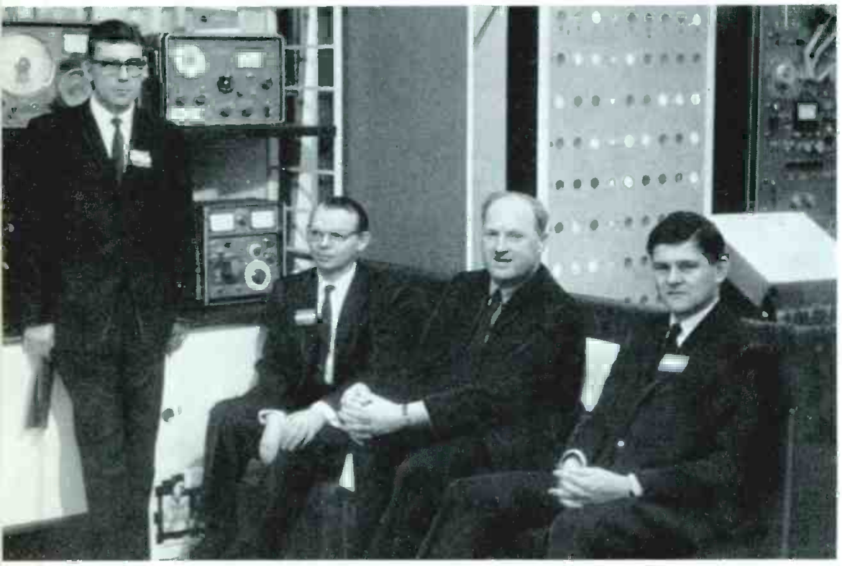
The relay used was a miniature r.f. switching type with its operating coil connections brought out to a BNC socket mounted near the fuse holders at the side of the instrument. The relay was mounted on a bracket close to SB to minimize lead lengths.

The modifications outlined provided the following features:

1. Automatic tuning to either 47·95 or 60 Mc/s input frequency.
2. Remote deviation switching from $\pm$5 kc/s to $\pm$25 kc/s full scale.
3. The instrument performs as a standard TF 791D until the relay coil is operated.

The units were rack mounted and used in a programmed test gear assembly for aligning low power f.m. transmitters.

This requirement shows a trend towards more sophisticated and therefore more expensive test gear. However, its advantages in production line assembly are great, and the increased production rate with less skilled personnel would soon recover the initial cost of the test equipment.

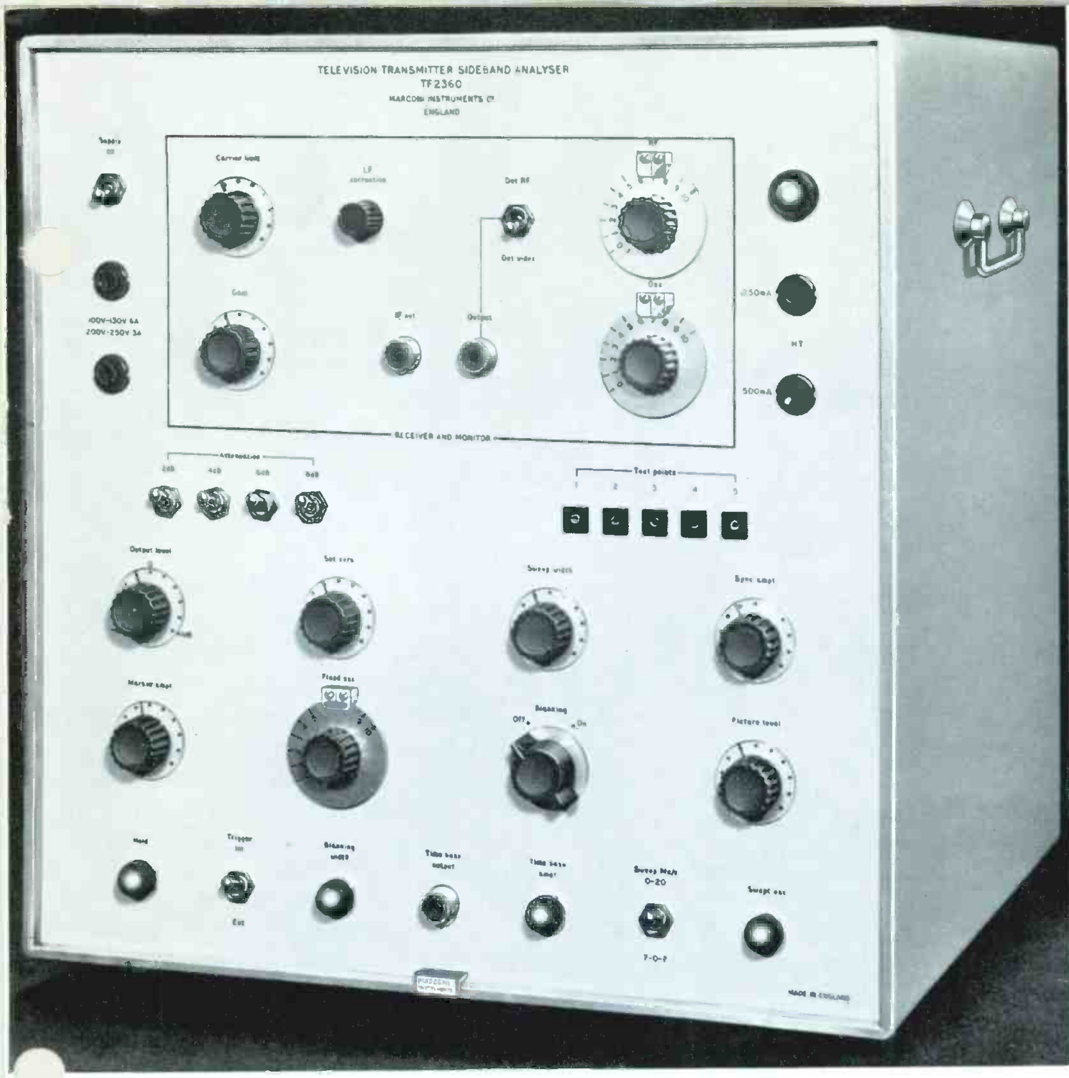


TF 791D, top centre, at the International Radio and Electronics Show with M.I. executives. Left to right: Mr. V. J. O'Herlihy, Service Manager, U.S.A.; Mr. W. Oliver, Assistant Sales Manager, U.S.A. - the author of this article; Mr. A. G. Wray, Deputy Chief Engineer, St. Albans; and Mr. K. Elkins, Assistant Sales Manager, U.S.A.

621. 317. 74: 621. 397. 611

MARCONI INSTRUMENTS

NEW DESIGN



TYPE TF 2360

TF 2360, cased as a transportable equipment. A rack mounting version is also available

Television Transmitter Sideband Analyser

by R. J. LUDBROOK,
B.Sc.(Eng.),
Graduate I.E.E.

This instrument has been developed from the OA 1241 Television Sideband Analyser which is in use in many transmitting stations in this country and abroad. Although the basic method of displaying the sideband response is similar, the new instrument makes use of improved circuit techniques and components resulting in an instrument of improved performance and considerably smaller physical size. Besides its ability to make dynamic measurements on transmitter and modulator chains on 405, 525 or 625 line systems in Bands I and III, an additional unit is available which extends the range of the instrument to cover the U.H.F. Bands IV and V. In this article the principles of operation are described, outlining the tests which can be made, together with a circuit description.

THE FREQUENCY RESPONSE of a television transmitter or modulator system is closely controlled to a strict specification and an accurate determination is important. The Television Transmitter Sideband Analyser enables the response of a transmitter to be examined visually to a high order of accuracy using sweep frequency techniques. A video sweep signal covering all the required modulation frequencies modulates the transmitter under test, and the resulting r.f. output consisting of carrier plus sidebands is detected and the outline of the response is displayed on an oscilloscope. Blanking and synchronizing signals may be applied to the instrument to provide a composite test signal; this enables the test to be carried out without disturbing the black level clamping networks and d.c. restorers in the modulator chain, which otherwise could affect the accuracy of the measurement, apart from lengthening the duration of the test.

Measurement of the response of video systems alone may also be made. A detector circuit is incorporated which will monitor the sweep output from the video equipment under test whether of video or composite form and present the detected response on an oscilloscope.

Extension of the frequency range to cover the U.H.F. Bands IV and V involves no change in the basic instrument, but requires an additional unit which comprises a

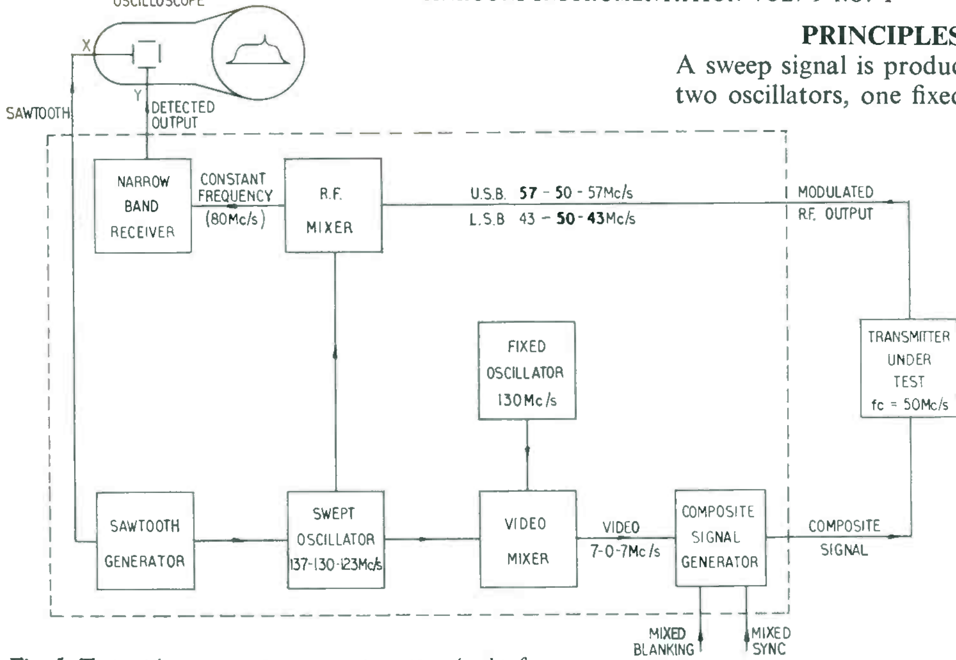


Fig. 1. Transmitter response measurement. As the frequency sweeps from 7 Mc/s to zero, the upper sideband response is displayed, and from zero to 7 Mc/s the lower

local oscillator and wideband crystal mixer, the output from which is within the range of the Sideband Analyser.

The TF 2360 Television Transmitter Sideband Analyser bears no resemblance mechanically to its predecessor, the OA 1241, and electrically the difference is very great, although the method of obtaining the video sweep and of producing the sideband display is basically similar. The composite signal is generated by a new and much improved method.

This instrument consists of a Video Sweep Generator, Receiver, Composite Signal Generator, Video and Monitor Amplifiers and Power Supplies, all contained in one unit which opens out on hinges for ease of servicing and alignment.

PRINCIPLES OF OPERATION

A sweep signal is produced by mixing the outputs from two oscillators, one fixed in frequency at 130 Mc/s and the other sweeping at either half mains frequency or half field frequency symmetrically about 130 Mc/s with a maximum excursion of at east 123 to 137 Mc/s, as shown in Fig. 1. The difference frequency produced gives a video signal which sweeps from 7 Mc/s to zero and back to 7 Mc/s at the rate of sweep, when the sweep width is maximum.

This video signal, after amplification, may be used to modulate the transmitter. Thus at any instant one of the two sidebands produced is decreasing in frequency at the same rate as the frequency of the swept oscillator; therefore if the transmitter output is mixed with the output from the swept oscillator there will be a difference frequency produced which will be constant in frequency throughout the sweep, but which will vary in amplitude according to the variation in amplitude of the appropriate sideband. The output is then fed to the input of a narrow band receiver tuned to this constant frequency, which is equal to the difference between the transmitter frequency and the swept oscillator frequency.

The detected output from the receiver consists of a d.c. voltage which varies in amplitude over the sweep as the amplitude of the sideband varies and may be displayed on an oscilloscope such as the Marconi Instruments TF 1330 or TF 2200 which may also be used for observing video signals. The sawtooth waveform which

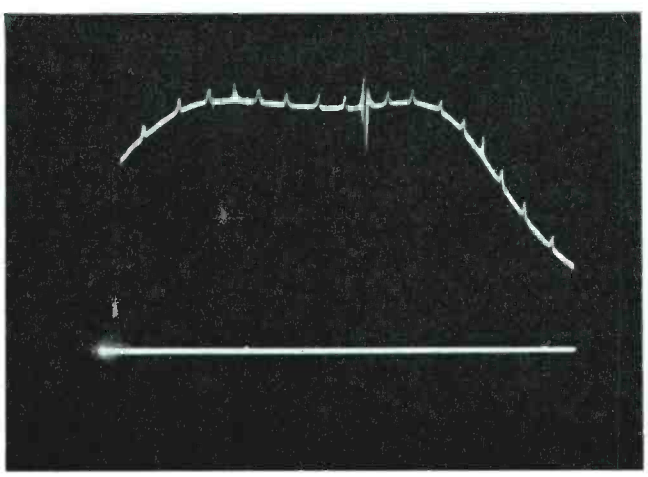


Fig. 2. Response of Band I transmitter simulator at fc = 45 Mc/s showing 1 and 5 Mc/s markers. Upper and lower sidebands are displayed on either side of the carrier with a video modulation sweep of 8–0–8 Mc/s

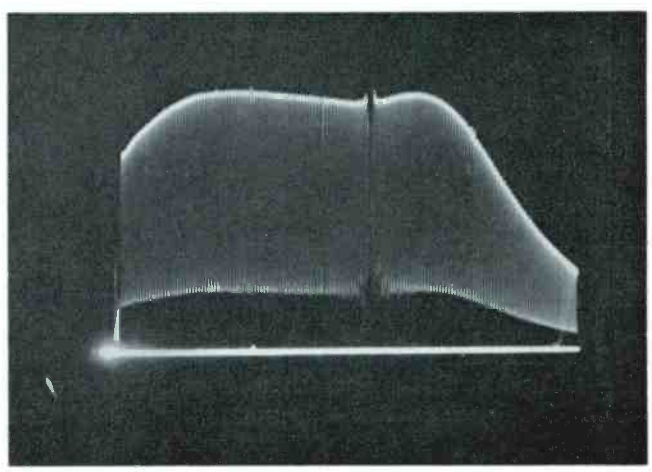


Fig. 3. Response of Band I transmitter simulator at fc = 45 Mc/s without markers. The simulator is modulated with a composite signal sweeping 8–0–8 Mc/s at the standard picture level of 1 V d.a.p. on a 405 line system

drives the swept oscillator is also available to provide the horizontal deflection for the oscilloscope. Since the sweep is symmetrical about the carrier the presentation shows the response of each sideband on either side of the carrier and thus vestigial sideband characteristics of a transmitter may be examined.

Besides the symmetrical sweep used for transmitter measurement there is also available an asymmetrical sweep which swings from below 100 kc/s to at least 20 Mc/s, as shown in Fig. 5. This may be used in video or composite signal form for measuring the response of video amplifiers or modulator systems. When switched to asymmetrical sweep the swept oscillator swings from 130 to 110 Mc/s for maximum sweep width, thus giving a difference frequency of 0 to 20 Mc/s. After application to the video system under test the output signal is fed to the video amplifier with its associated detector, the d.c. output from which is displayed on an oscilloscope as before to show the video response. Provision is made in the detector for removal of synchronizing and blanking pulses when a composite test signal is used without affecting the measurement of the video response.

When used on Bands IV and V the receiver is set to a given frequency, e.g. 80 Mc/s, so that the Sideband Analyser will accept an input frequency of, in this case, 50 Mc/s, as shown in Fig. 6. The output from the transmitter is fed to the matched input of a wideband crystal mixer, which together with a local oscillator comprises the additional unit. The local oscillator is tuned to a frequency differing from that of the transmitter by, in this case, 50 Mc/s, thus giving an i.f. output which can be accepted by the main instrument.

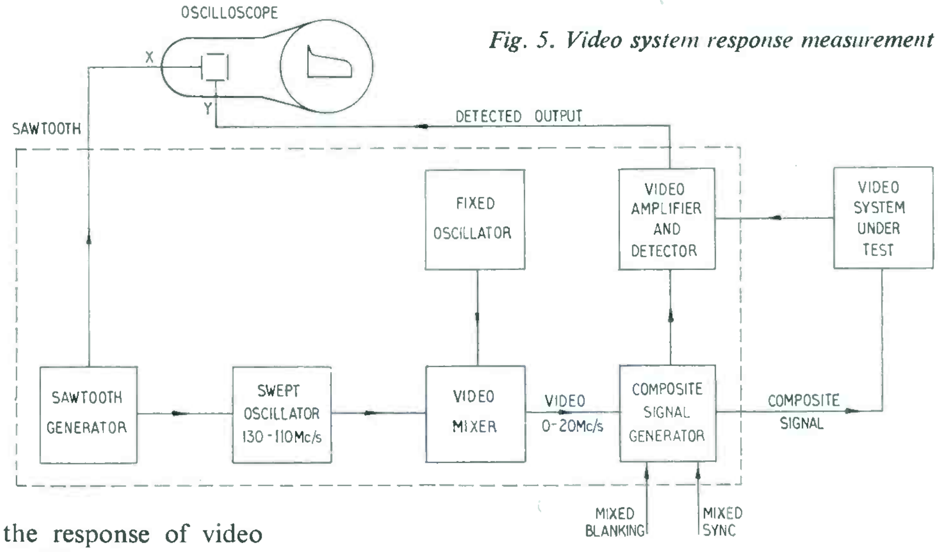


Fig. 5. Video system response measurement

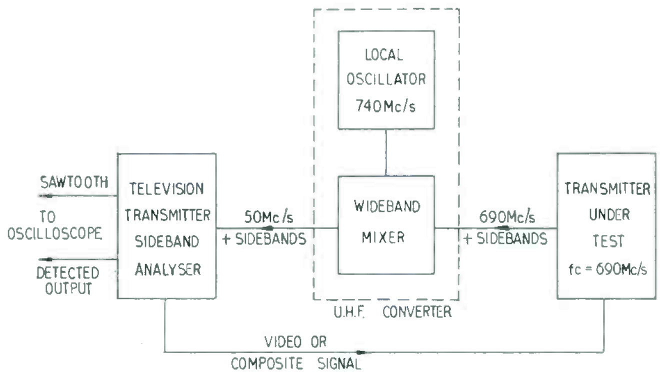


Fig. 6. Bands IV and V transmitter response measurement. The transmitter is operated in Band V at 690 Mc/s and the sideband analyser is set to accept a Band I signal of 50 Mc/s

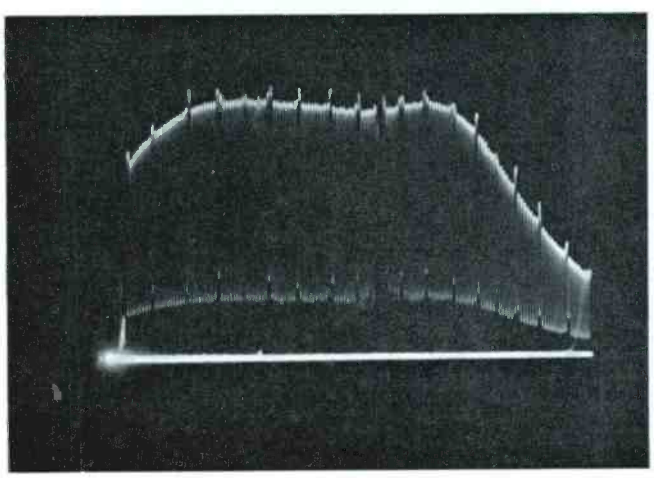


Fig. 4. As Fig. 3 but showing markers superimposed

FACILITIES AVAILABLE

The video sweep output may be of composite or non-composite form as required. Video sweep may be taken directly from the Sweep Generator output and has an amplitude variable from 0·3 to 3 V d.a.p. in 2 dB steps with a continuously variable control of ± 1 dB calibrated at 0·2 dB intervals. The output is held constant to within 0·2 dB over the sweep range by means of an a.g.c. circuit.

A composite signal may be obtained by feeding the video from the sweep generator into the composite signal generator to which is applied mixed synchronizing and blanking pulses from a standard synchronizing generator. The video component of the output signal may be varied in amplitude from 0·1 to 1 V d.a.p. Although the standard level for the composite signal is 1 V d.a.p. from sync bottom to peak white, this may be varied over a wide range to cater for different tests. The picture level can be varied up to 0·8 V d.a.p. and the sync amplitude up to 0·5 V d.a.p. so that responses can be examined at any level from peak white to full black-to-white modulation.

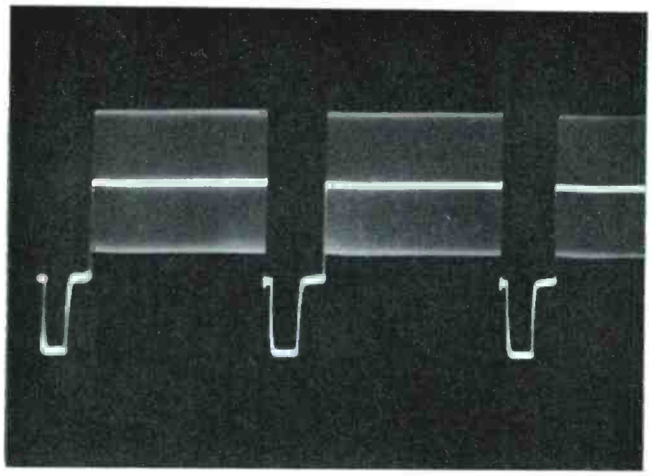


Fig. 7. Output from composite signal generator observed at line frequency on 405 line system. Standard picture level of 1 V d.a.p. with 0 to 20 Mc/s sweep

The availability of a composite signal containing blanked video and synchronizing pulses in their correct sequence and relative amplitude enables dynamic measurements to be made on transmitters or modulators at any level in the grey scale, without the need for disconnecting black-level clamping networks or d.c. restorers, thus giving a more realistic and accurate measurement.

Sweep repetition rate may be either locked internally to a 50 c/s source or externally to the field frequency. The sweep may be symmetrical or asymmetrical and varied up to at least 7 Mc/s on either side of zero or from 0 to 20 Mc/s respectively. Frequency markers are available to measure the sweep width.

For measuring the r.f. output from a transmitter a receiver is used to give a detected output for examining the response on an oscilloscope. The receiver has a coverage of 40 to 100 Mc/s and thus the transmitter frequency range, when mixed with the 130 Mc/s swept oscillator is from 30 to 90 Mc/s, covering Band I. Frequencies in the range 170 to 230 Mc/s will also give a difference frequency in the range of the receiver for Band III measurements. The receiver gain is variable and there is a carrier limiting control to limit the large carrier spike on the display to avoid distortion. The bandwidth of the i.f. amplifier in the receiver is less than 40 kc/s so that the response of a transmitter may be examined close to the carrier without interference from the carrier.

For measuring the video output from a modulator a video amplifier and detector is used to provide a d.c. output for display. As well as displaying video signals the amplifier will also accept a composite signal, an arrangement being used whereby the detector is rendered inoperative during the blanking periods to avoid distortion to the video component of the signal. A switch is provided to cater for positive or negative modulation.

The ability to measure small amplitude video signals in the presence of blanking and sync pulses is of special importance when examining the response at peak white when as small a video signal as possible is used prevent to too great an excursion into adjacent picture regions.

The output from receiver or video amplifier is fed into the monitor amplifier which presents the output in the correct polarity for display and also provides correction for the low frequency response of the oscilloscope, as well as superimposing the frequency markers upon the display. Circuit details of each section are given below with reference to Fig. 8.

VIDEO SWEEP GENERATOR

Sawtooth Generator

A large amplitude sawtooth waveform is generated from a Miller-transitron oscillator circuit, whose recurrence frequency is locked either internally to a 50 c/s source or externally to large negative-going pulses derived from the blanking pulses in the composite signal generator, and occurring at field frequency.

The output drives the following modulator stage and is also applied to a socket on the front panel via the TIME BASE amplitude control for horizontal deflection of the display oscilloscope. The amplitude of the sawtooth applied to the modulator stage may be varied by the SWEEP WIDTH control.

Modulator and Swept Oscillator

The sawtooth is amplified and inverted in the modulator stage before being fed to the magnetizing coil of a ferrite reactor. The operating point for the current in the magnetizing coil of the reactor is adjusted by the SET ZERO control which forms part of a resistance chain across the modulator. Current change in the magnetizing coil due to the sawtooth waveform causes a change in permeability of a ferrite rod upon which is wound the inductance of the tuned circuit of the swept oscillator and consequently varies its frequency of oscillation.

An electron-coupled Hartley oscillator with an aperiodic anode load is used as the swept oscillator. When the sweep is 7–0–7 Mc/s the current in the reactor is such that in the centre position of the SET ZERO control the swept oscillator is centred about a frequency of 130 Mc/s and swings to at least 7 Mc/s on either side for maximum sweep width. The amount of swing is determined by the SWEEP WIDTH control, while the sweep may be off-centred by the SET ZERO control.

When the sweep is 0–20 Mc/s, the operating point for the current in the reactor is changed so that the swept oscillator swings from 130 to at least 110 Mc/s for maximum sweep width.

The output from the swept oscillator is fed to the input of two wideband buffer stages using variable-mu pentodes which, as well as isolating the swept oscillator from the fixed oscillator, also control the amplitude of the video sweep output by means of the applied a.g.c. voltage.

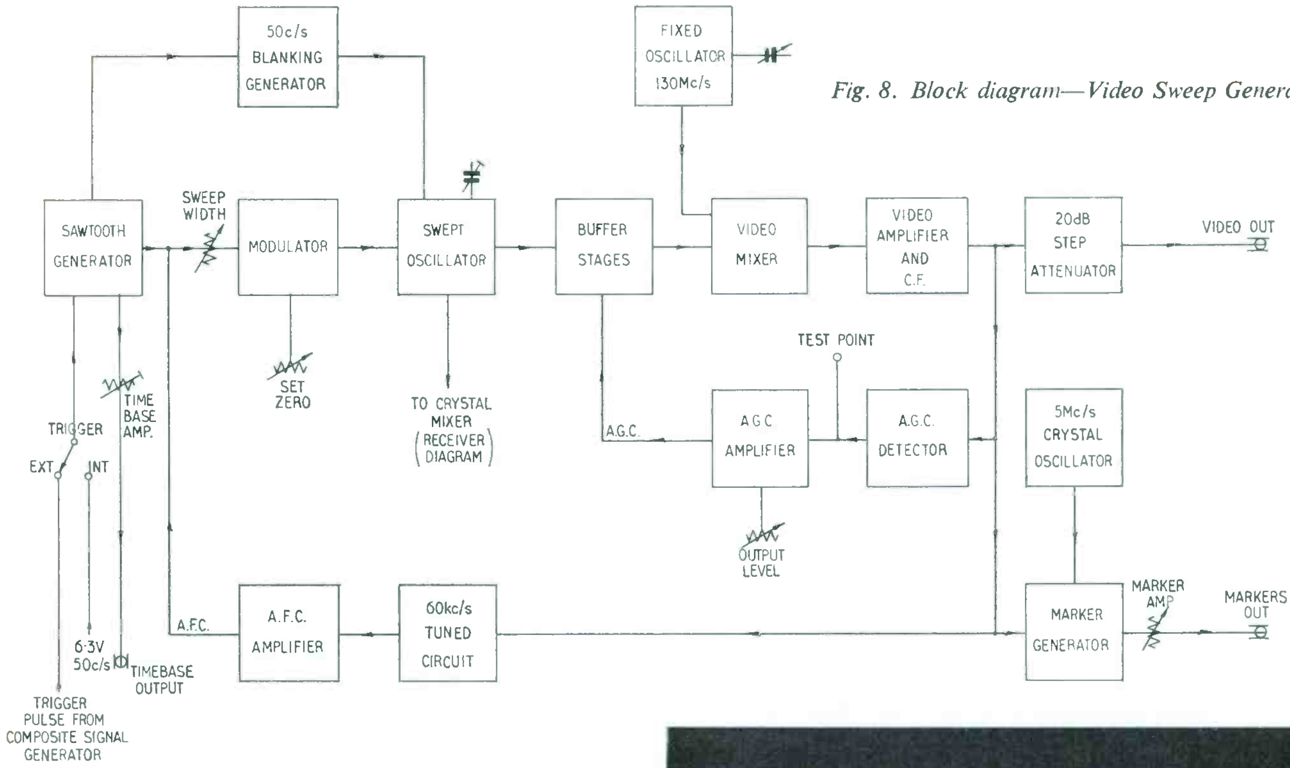


Fig. 8. Block diagram—Video Sweep Generator

50 c/s Blanking Generator

A cathode coupled monostable multivibrator is triggered from the sawtooth generator. The output from the multivibrator is connected to the grid circuit of the swept oscillator and renders it inoperative during the negative part of the multivibrator waveform. Circuit constants are such that the oscillator is allowed to function only during every alternate sawtooth rundown, and therefore the sweep repetition rate is at half mains frequency or half field frequency.

The preset BLANKING WIDTH control allows the multivibrator to be adjusted to obtain this condition.

Fixed Oscillator

This takes the form of an electron-coupled Colpitts oscillator operating at a nominal frequency of 130 Mc/s but which may be varied by 2·5 Mc/s on either side by the tuning capacitor, controlled from the front panel. Outputs from the fixed oscillator and buffer stages are applied to the input of the video mixer stage.

Video Mixer

In this stage the swept frequency signal from the buffer stages is mixed with the signal from the fixed oscillator to give a difference frequency signal which sweeps a maximum of 7 Mc/s to zero and back to 7 Mc/s or from zero to 20 Mc/s depending on the sweep selected. To ensure that the response of the video signal is dependent only on the response of the signal from the buffer stages, the amplitude of the fixed oscillator signal is considerably larger than that due to the swept signal. Inductive compensation is used to improve the frequency response.

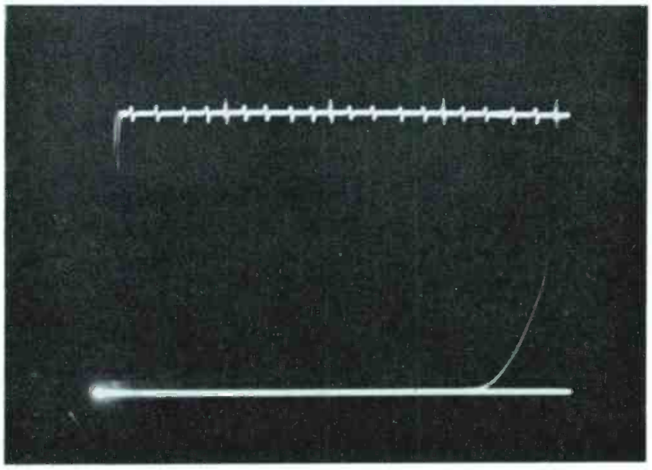


Fig. 9. Detected output from video sweep generator observed at test point, showing markers superimposed by capacitive coupling from MARKERS OUT *socket. Sweep width, 20 Mc/s*

Video Amplifier and Output Stage

A single stage video amplifier is used to obtain the required amplitude of output signal. To provide the necessary gain in a single stage a high slope frame-grid pentode having a large gain bandwidth product is employed. Optimum frequency response is obtained by shunt and series inductive compensation. A similar valve is used for the cathode follower output stage to preserve the response and provide the output signal at low impedance. The cathode follower feeds the output attenuator and also the a.g.c., a.f.c. and marker circuits. The attenuator consists of four π-section networks of 2, 4, 6 and 8 dB insertion loss at 75 Ω impedance, making a total of 20 dB which can be switched in 2 dB steps.

A.G.C. Circuit

A voltage doubler circuit is used to detect the video sweep output. The detected response is then amplified and inverted and the output clipped to remove all but the wanted a.g.c. information. After d.c. restoring to earth, the signal now consists of an inverted and amplified form of the detected response at a small d.c. level with respect to earth. This signal is applied to the variable-mu buffer stages and controls the gain inversely as the amplitude of the video output signal ensuring a flat response.

The gain of the a.g.c. amplifier may be varied by two controls, one preset for obtaining the maximum video output of 3 V d.a.p. and the other controlled from the front panel to give a variation of ± 1 dB about this level.

A.F.C. Circuit

When in the 0–20 Mc/s position the sweep does not in fact start at zero but at 60 kc/s due to the action of the a.f.c. circuit, which is inoperative in the 7–0–7 Mc/s position.

The video sweep output prior to the attenuator is fed to a series tuned circuit resonant at 60 kc/s and the voltage developed across it is applied to a voltage doubler detector. The R-C load for the detector has its time constant chosen to be large enough to maintain sufficient amplitude during the sweep to keep the a.f.c. action steady without being too large, resulting in sluggish operation. The rectified output is then fed to the a.f.c. amplifier where it is amplified and inverted and applied to the sawtooth generator causing the positive tip of the sawtooth to be restored to a d.c. level determined by the a.f.c. circuit constants, chosen so that the sweep commences at 60 kc/s.

Any change in frequency of the swept oscillator will, because of the shape of the response at 60 kc/s, produce a change in the rectified output from the a.f.c. circuit which causes the potential of the start of the sawtooth run-down to alter until the frequency is again correct. If the fixed oscillator varies in frequency, which is less likely, the a.f.c. locking circuit will still function but the swept oscillator will be corrected and thus the sweep operation will no longer be based on 130 Mc/s. This circuit functions for all settings of the sweep width control.

Marker Generator

The video feeding the a.f.c. circuit is also taken to the marker generator which operates as a self-oscillating mixer, with the frequency of oscillation crystal controlled at 5 Mc/s. Its anode load is formed by two tuned circuits in series, one resonant at 1 Mc/s, the other resonant at 2 Mc/s.

When the video sweep reaches 1 Mc/s the 1 Mc/s tuned circuit rings and produces an output at the anode, a similar action taking place in the 2 Mc/s tuned circuit when the sweep reaches 2 Mc/s. At 3 Mc/s the video mixes with the 5 Mc/s oscillator causing the 2 Mc/s circuit to ring and at 4 Mc/s the difference frequency rings the 1 Mc/s circuit. Thus there is an output at the anode at 1, 2, 3 and 4 Mc/s.

At 5 Mc/s the video beats directly with the 5 Mc/s oscillator producing a zero beat which is distinctive from the other outputs. Frequencies of 6 and 7 Mc/s mix with 5 Mc/s, again ringing the 1 Mc/s and 2 Mc/s circuits respectively. Frequencies of 8 Mc/s and above have a similar action but rely on harmonics from the oscillator.

The output is rectified and d.c. restored. Thus the final output, controlled in amplitude by the MARKER AMP. control, consists of a series of positive-going blips at 1 Mc/s intervals with every fifth marker of distinctive appearance.

RECEIVER UNIT

R.F. Mixer

The swept oscillator output and the swept transmitter output are mixed in this circuit to give a constant frequency signal which is fed to the r.f. amplifier.

A broadband resistive network is used, the transmitter being isolated sufficiently from the crystal itself so that the swept oscillator amplitude at the crystal is the larger of the two, ensuring that the output amplitude is dependent on variations due to the transmitter signal only. In addition, the attenuation introduced ensures that a good v.s.w.r. is presented to the transmitter.

Amplifiers and Detector

The output from the mixer is taken to a low-noise cascode coupled r.f. amplifier stage, the tuned output from which is loosely coupled to the tuned circuit input of the second mixer stage. There are coarse and fine tuning capacitors in each tuned circuit, to avoid the need for complicated slow motion tuning. A Hartley circuit is

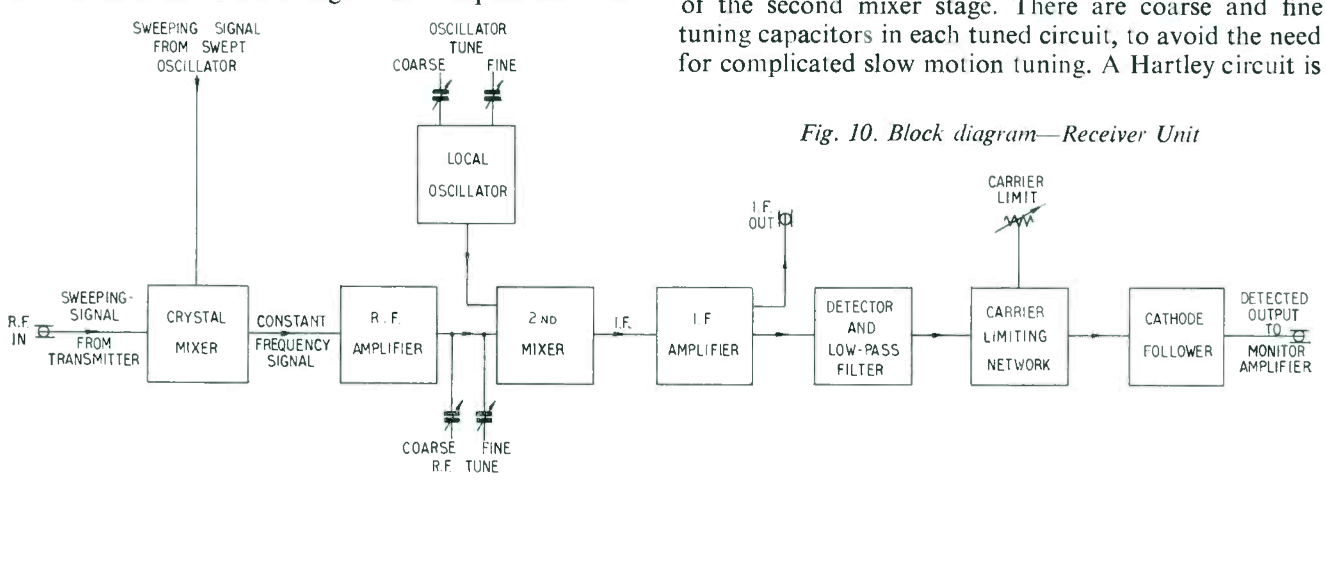


Fig. 10. Block diagram—Receiver Unit

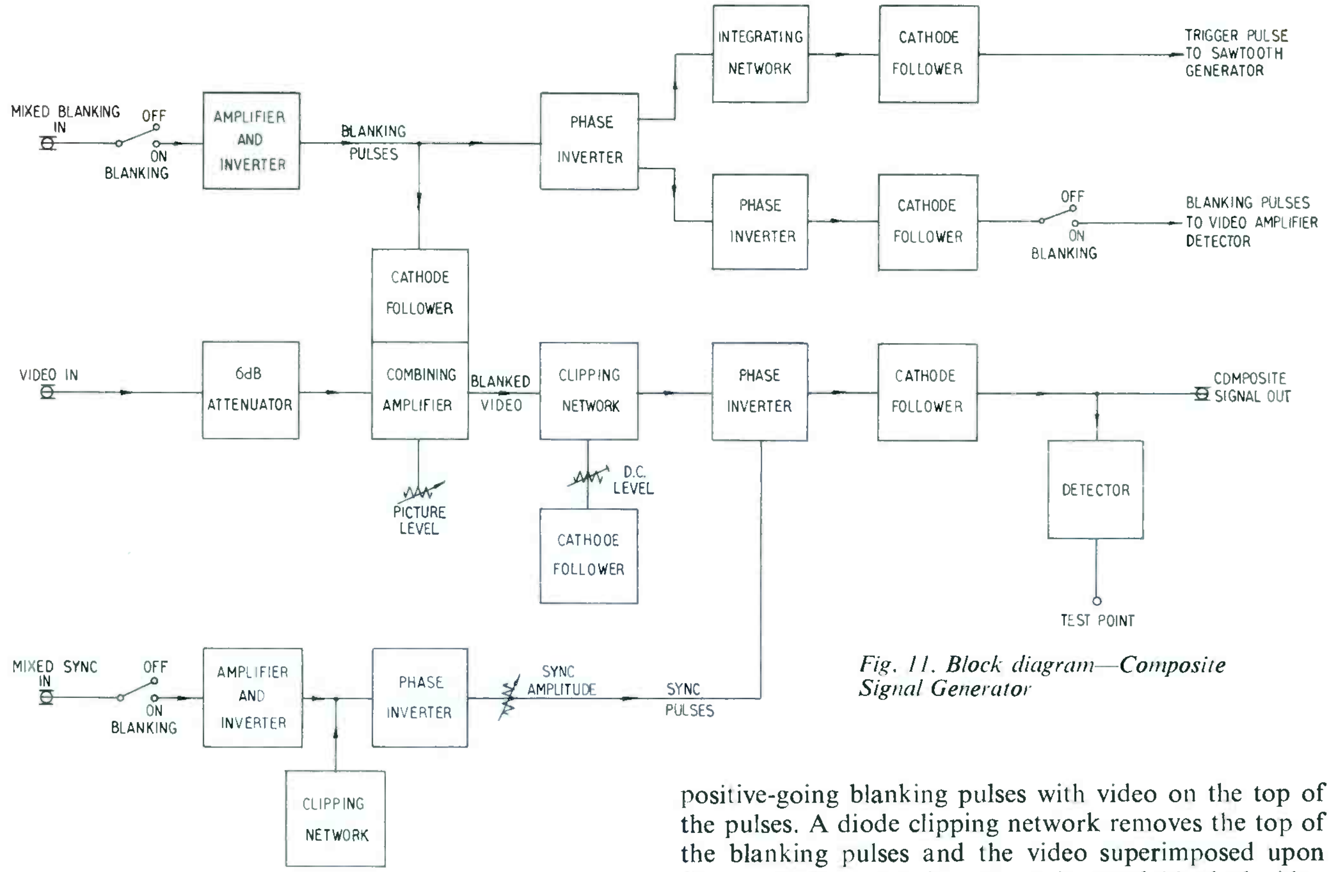


Fig. 11. Block diagram—Composite Signal Generator

used for the local oscillator which operates at 4·5 Mc/s above the signal frequency.

The bandwidth of the two-stage i.f. amplifier has been kept down to 40 kc/s so that the sideband response of a transmitter may be examined close to the carrier without interference.

Although the non-linearity in the detector has been kept to a minimum this may be completely eliminated by taking the output at the i.f. prior to the detector at the I.F. OUT socket.

Carrier Limiting Circuit

The output from the detector is applied to a cathode follower whose cathode provides the reference point for the carrier limiting circuit and which is compensated for changes in voltage which are not due to the signal. The carrier limiting circuit comprises a diode clipping network whose operating point may be set to any desired level by means of the Carrier Limit control to remove the large spike due to the carrier from the display. From the carrier limiting circuit a detected output is available, at low impedance, as one of the inputs to the Monitor Amplifier.

COMPOSITE SIGNAL GENERATOR

Mixed synchronizing and blanking signals must be supplied to this unit from a standard synchronizing generator.

First, the blanking pulses are combined with the sweep input to give a video signal interrupted by amplified positive-going blanking pulses with video on the top of the pulses. A diode clipping network removes the top of the blanking pulses and the video superimposed upon them and the result is now an inverted blanked video signal. The operating point of the clipping network is preset and the bias on the combining amplifier stage may be varied by the PICTURE LEVEL control to give the desired picture set-up.

In the inverter stage synchronizing pulses are added from the sync amplifier, after being amplified and clipped to make the amplitude independent of the input level; a SYNC AMPLITUDE control is provided.

The processed signal at the COMPOSITE SIGNAL OUT socket may be set for the required video amplitude by using the attenuator in the sweep generator and for the required picture level and sync amplitude by using the controls in the composite signal generator.

The blanking pulses from the first stage are also fed to a phase inverter from which two outputs are taken.

One output is integrated, the circuit constants being chosen so that the line blanking component is attenuated and the field blanking component appears as a large negative pulse which may be used to trigger the sawtooth generator at field frequency.

The other output provides a large positive going mixed blanking waveform to the detector in the Video Amplifier unit. A detector circuit is provided at the COMPOSITE SIGNAL OUT socket to monitor the video response.

VIDEO AND MONITOR AMPLIFIERS

When testing video systems, the output from the equipment under test is fed into the video amplifier. If a composite test signal is being used, blanking pulses from the composite signal generator are fed to the detector to render it inoperative during the blanking periods; this prevents distortion during the detection of the video component. A polarity switch enables signals with either positive or negative modulation to be displayed.

This video amplifier ensures satisfactory display of small amplitude video signals, which might be of the order of the 0·1 V d.a.p. when examining the response at peak white. The blanking of the detector is also especially important under these conditions.

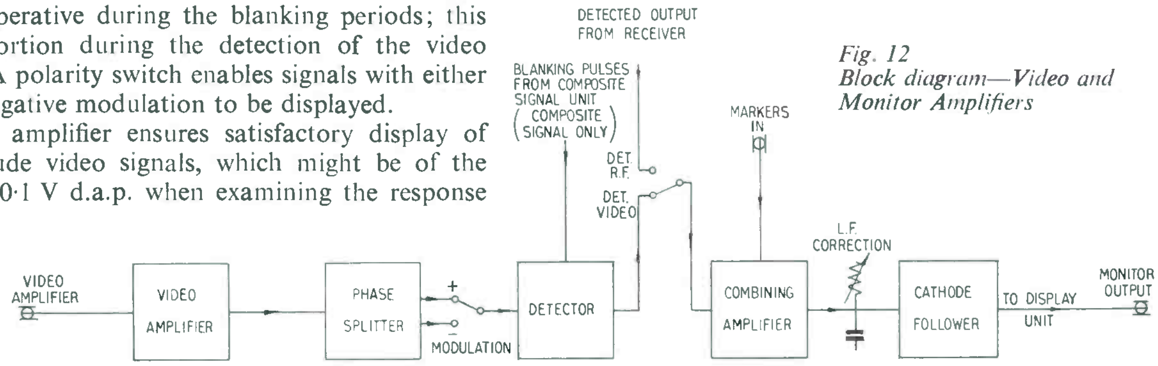


Fig. 12
Block diagram—Video and Monitor Amplifiers

The monitor amplifier is used for displaying the detected output from either the receiver or video amplifier depending on whether a transmitter or video system is being tested.

The detected signal in either case is applied to an amplifier stage which gives a positive-going output for display, and whose gain may be varied to give unity gain in the Monitor Amplifier so that the amplitude of the detected signal is preserved. Marker signals from the Sweep Generator are combined in this amplifier with the detected signal. The resulting signal is applied to the output cathode follower stage which provides a low impedance output at the MONITOR OUT socket for application to the display oscilloscope. In the grid circuit of the cathode follower a variable R-C network is used to enable the l.f. response to be adjusted to compensate for that of the oscilloscope, by means of the L.F. CORRECTION control.

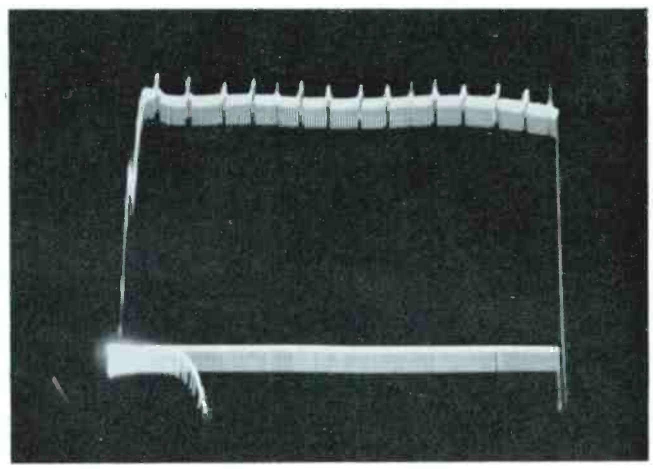


Fig. 13. Output from blanked detector in video amplifier for a composite signal input sweeping from 0 to 15 Mc/s on a 405 line system. Picture level is the standard 1 V d.a.p. and video component is 0·1 V d.a.p. corresponding to peak white

U.H.F. CONVERTER

The test signal at Band IV or V frequency is coupled to the input of a wideband diode mixer. Care is taken in the design that the impedance presented to the transmitter is matched to 50 Ω to prevent reflections. The diode current is monitored so that the diode is not overloaded and to set the relative levels of input and local oscillator for satisfactory mixing.

A variable impedance line local oscillator in the unit provides adequate power for satisfactory operation of the mixer as well as possessing good frequency stability. The frequency range may be varied from 420 to 910 Mc/s so that for an i.f. output of 50 Mc/s a transmitter range of 470 to 960 Mc/s is obtained.

ABRIDGED SPECIFICATION

Video sweep output

AMPLITUDE: 0·3 V to 3 V d.a.p.

SYMMETRICAL SWEEP: Lower limit locked to a frequency not exceeding 100 kc/s. Upper limit continuously variable up to 20 Mc/s.

ASYMMETRICAL SWEEP: Continuously variable up to 7–0–7 Mc/s.

FREQUENCY RESPONSE: Flat to within ±0·1 dB throughout the sweep.

Composite signal output

AMPLITUDES:
Video Sweep Component: 0·1 to 1 V d.a.p.
Mixed Sync Component: 0 to 0·5 V d.a.p.
Set Up: 0·1 to 0·8 V d.a.p.

SYMMETRICAL SWEEP: } As for Video
ASYMMETRICAL SWEEP: } Sweep.

FREQUENCY RESPONSE: Flat to within ±0·1 dB up to 10 Mc/s and within ±0·25 dB up to 20 Mc/s.

Video amplifier

AMPLITUDE OF SIGNAL INPUT: Standard level of 1 V d.a.p. with +ve or −ve syncs into 75Ω.

FREQUENCY RESPONSE: Flat to within ±0·1 dB up to 10 Mc/s.

R.F. response measurement

FREQUENCY RANGE: 30 to 90 Mc/s and 170 to 230 Mc/s. 470 to 960 Mc/s with additional unit TM 6936.

AMPLITUDE/FREQUENCY DISTORTION: Less than 0·25 dB.

621.317.34: 621.317.353.3

MARCONI INSTRUMENTS APPLICATION NOTE

INTERMODULATION MEASUREMENT ON LOW CAPACITY LINK SYSTEMS

by H. C. GRIBBEN

Intermodulation measurements can be made with the White Noise Test Set OA 1249 on any system having a maximum of 960 channels. On low capacity systems, however, it is necessary to apply a small correction factor to obtain the best measurement accuracy.

A table of correction factors is given for systems having from 12 to 60 channels, and the theory of measurement is explained.

MARCONI INSTRUMENTS LTD. have been manufacturing the OA 1249 series of White Noise Test Sets[1] since 1958. Originally it was considered that this method of intermodulation measurement would, in general, be confined to channel capacities of greater than 60, because C.C.I.R.[2] recommendations for white noise intermodulation testing do not apply to system capacities of less than this number of channels. Our experience is, however, that the instrument is also used extensively on low capacity systems down to 12 channels. It should be noted that measurement accuracy will be less than that obtainable on high capacity systems unless a correction is applied to take into account the amount of power extracted from the base band when a band-stop filter is switched into circuit. The value of the correction will depend upon the extent of the base-band and the characteristics of the filter.

In order to make the problem clear, the white noise method of intermodulation measurement will first be described briefly.

Cross-talk or intermodulation occurs in a multi-channel telephone system due to amplitude and phase non-linearity. The effect may be described if we consider, say, a 600 channel system in which 599 channels are occupied by speech and the 600th channel used as a listening post. Intermodulation between the 599 subscribers talking simultaneously will be apparent as noise in the 600th channel. The higher the level of the noise appearing in what should be a quiet channel, the more serious is the intermodulation.

The method used to assess this form of distortion was first described by White and Whyte in 1955.[3]

In this method, white noise at a suitable power level is applied to the input of the equipment under test and occupies the same frequency spectrum as that normally occupied by the translated speech; a fully loaded multi-channel telephone system is thus closely simulated.

If a filter having a very narrow stop-band is interposed between the white noise source and the system (Fig. 1), the conditions then existing will be equivalent to a fully loaded system except for one quiet channel.

In order to assess the amount of intermodulation present, a receiver tuned to the frequency of the quiet channel is connected to the output of the multi-channel equipment under test and the noise measured relative to the noise level present without the band-stop filter switched into the circuit. The ratio of the two levels is referred to as the noise power ratio (N.P.R.).

Returning now to the question of intermodulation measurement on low capacity systems, we are concerned with the change in power loading produced by switching 'in' the band-stop filter used to provide the quiet channel.

Consider the 70 kc/s band-stop filter characteristic shown in Fig. 2, which conforms to the recommendations of C.C.I.R. When this filter is switched into circuit, the

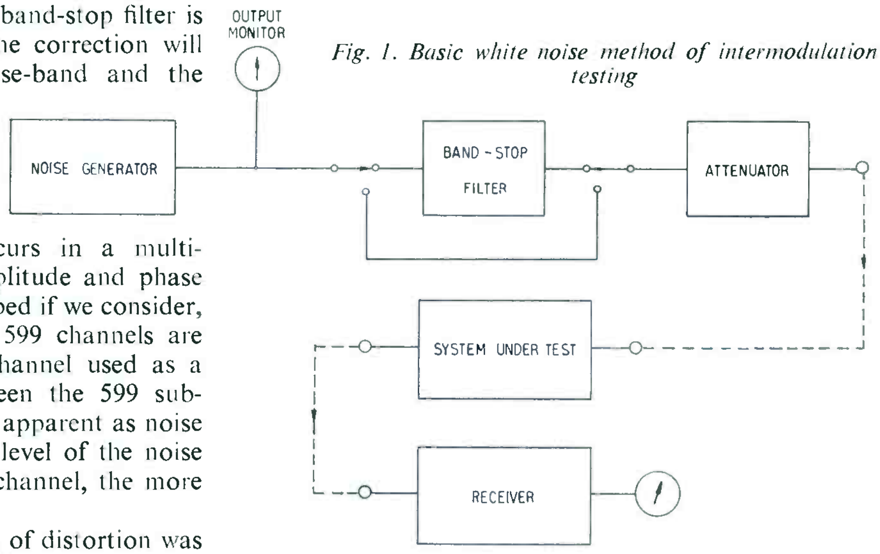


Fig. 1. Basic white noise method of intermodulation testing

noise power applied to the system under test is reduced by an amount dependent upon the shape of the filter characteristic. In this case, if all but 1% of the extracted power is taken into account, the effective bandwidth of the applied noise is reduced by 21·5 kc/s.

For a 24-channel system having a base-band from 12 to 108 kc/s, i.e. 96 kc/s bandwidth, when the filter is switched into circuit the effective bandwidth is reduced to 74·5 kc/s and the power applied to the equipment under test falls by 1·1 dB.

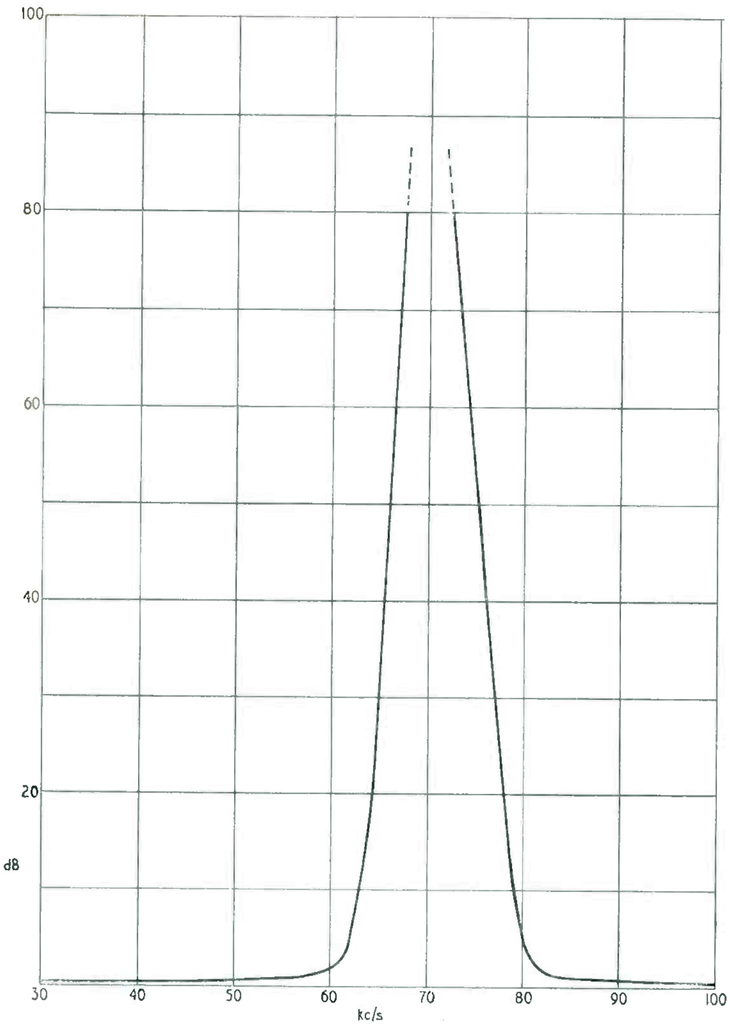


Fig. 2. Band-stop filter characteristic

If the only consideration is that the frequency deviation of the system shall be the same after switching in the filter as it was before, then the output of the White Noise Generator must be increased by this amount regardless of the fact that loading is increased only over that part of the base-band unaffected by the filter.

Applying the same argument to a 60-channel system having a base-bandwidth of 240 kc/s, the same 70 kc/s filter will cause a reduction in power of approximately 0·45 dB when it is switched into circuit. A higher frequency filter having a proportionately wider bandwidth will cause a greater reduction in power when switched into the same base-band.

For channel capacities greater than 60 channels the effect is small, whatever the filter frequency, and need not be taken into consideration.

The table shows the corrections for various filters and channel capacities. The power indicated on the White Noise Generator output meter must be increased by the value of the correction after the filter is switched in.

A change in meter reading occurs when a band-stop filter is switched into circuit due to an accompanying impedance change, and does not accurately indicate the change in power output of the White Noise Generator. The necessary correction must, therefore, be added to whatever the meter indicates, by increasing the generator noise output control.

REFERENCES

1. Roper, G., 'Intermodulation Testing Using the White Noise Test Set Type OA 1249A'. *Marconi Instrumentation*, March 1959, 7, p. 17.
2. Comité Consultatif International des Radio-communications. Los Angeles, 1959.
3. White, R. W., and Whyte, J. S., 'Equipment for Measurement of Inter-channel Crosstalk and Noise on Broadband Multi-channel Telephone Systems'. *P.O.E.E.J.*, October 1955.

Power Loss due to the Introduction of Band-stop Filters

The following table gives the loss of power relative to the total power in the noise band. The ratio is quoted in dB's for 12, 24, 36, 48 and 60 channel systems.

		Loss in dB's					
Filter	*Centre Frequency kc/s*	*12-Channel System 12–60 kc/s*	*24-Channel System 12–108 kc/s*	*36-Channel System 12–156 kc/s*	*48-Channel System 12–204 kc/s*	*60-Channel System 12–252 kc/s*	*60-Channel System 60–300 kc/s*
TM 6295	14	0·8	0·4	0·25	0·2	0·15	—
6304	27	1·1	0·5	0·3	0·25	0·2	—
6296	34	1·7	0·8	0·5	0·35	0·3	—
5841	40	2·1	0·9	0·6	0·45	0·35	—
6329	50	1·7	0·8	0·5	0·35	0·3	—
6297	56	1·6	0·85	0·55	0·4	0·3	—
5842	70	—	1·1	0·7	0·5	0·4	0·4
5843	105	—	1·2	1·15	0·8	0·65	0·65
6299	135	—	—	0·55	0·4	0·3	0·3
6298	152	—	—	0·6	0·45	0·35	0·35
5844	185	—	—	—	0·55	0·45	0·45
6301	245	—	—	—	—	0·45	0·45
6330	270	—	—	—	—	—	0·55
5845	290	—	—	—	—	—	0·5

Summaries of Articles appearing in this issue

RESUME D'ARTICLES PUBLIES DANS LE PRESENT NUMERO

PONT UNIVERSEL 1% TF 2700

L'appareil TF 2700 est un pont universel permettant de mesurer inductance, capacité et résistance à moins de 1% d'erreur.

Cet article décrit la structure de ce pont, fournissant certains détails sur ses caractéristiques électriques et mécaniques, puis indique quelques-unes des expériences les plus originales en mesures d'impédance que la disposition de ce pont facilite.

Page 3

VERSIONS HAUTE PUISSANCE DU GENERATEUR STANDARD TF 144H

Une nouvelle gamme d'appareils vient de sortir dans la série TF 144H. Il s'agit des modèles TF 144H/4 à TF 144H/6 qui, par leur puissance supérieure, sont appelés à remplacer les modèles TF 144H à TF 144H/2. Ces nouveaux modèles ont une puissance directe de sortie de 2,75 V (f.e.m.) à un niveau pratiquement nul d'impédance initiale, ce qui donne 100 mW pour une résistance de charge de 75 ohms.

Page 10

MESUREUR D'EXCURSION DE FREQUENCE A PROGRAMME

On constate aux Etats-Unis un accroissement de la demande d'équipement d'essai pouvant être programmé, ce qui indiquerait une tendance vers l'automatisation dans l'industrie de l'électronique. La Sté Marconi Instruments ne fabrique pas à présent de matériel d'essai à programmes, mais cet article décrit une méthode permettant de programmer le mesureur d'excursion de fréquence TF 791D pour une gamme limitée de mesures.

Page 11

ANALYSEUR DE BANDE LATERALE D'EMETTEUR DE TELEVISION TF 2360

Cet instrument a été élaboré à partir de l'analyseur de bande latérale de télévision OA 1241 qui est en service dans de nombreuses stations émettrices de télévision, tant à l'étranger qu'en Grande-Bretagne. Bien que la méthode de présentation de la réponse de bande latérale reste pareille, le nouvel instrument offre une grande réduction d'encombrement alliée à une amélioration de rendement grâce aux circuits et composants perfectionnés dont il est doté. Outre son aptitude à effectuer des mesures dynamiques sur les chaînes de modulateurs et d'émetteurs à 405, 525 ou 625 lignes dans les bandes I et III, il peut étendre son action aux bandes UHF, IV et V, grâce à l'adjonction d'une unité supplémentaire, à présent disponible. Dans cet article sont expliqués les principes de fonctionnement; on y trouvera également un aperçu des essais réalisables et une description des circuits.

Page 13

MESURE D'INTERMODULATION DANS LES SYSTEMES DE LIAISON DE FAIBLE ENVERGURE

L'appareil d'essais a bruit blanc OA 1249 permet de se livrer aux mesures d'intermodulation dans tous les systèmes comprenant au maximum 960 canaux. Pour les systèmes de faible envergure, il faut toutefois appliquer un facteur de correction pour obtenir le maximum de précision.

La table de facteurs de correction s'applique aux systèmes de 12 à 60 canaux et est accompagnée d'une explication théorique du principe de ces mesures.

Page 21

ZUSAMMENFASSUNG DER IN DIESER NUMMER ERSCHEINENDEN BEITRÄGE

1% UNIVERSAL—RLC—MESSBRÜCKE TF 2700

Das Gerät TF 2700 ist eine preisgünstige Messbrücke zur Messung von Induktivitäten, Kapazitäten und Widerständen mit einer Genauigkeit von 1%. In dem Aufsatz werden die konstruktiven Eigenheiten der Messbrücke beschrieben, wobei Einzelheiten über die elektrische und mechanische Konstruktion erwähnt werden. Hierauf folgen Beispiele von einigen weniger üblichen Versuchen bei Impedanzmessungen, welche sich wegen der Brückenanordnung besonders leicht durchführen lassen.

Seite 3

NEUE AUSFÜHRUNGEN DES MESS-SENDERS TF 144H MIT HÖHERER AUSGANGSLEISTUNG

Die neuen Ausführungen TF 144H/4 bis /6 der Typenserie TF 144H haben einen zusätzlichen Ausgang für eine höhere Ausgangsleistung und ersetzen die bisherigen Ausführungen TF 144H bis H/2. Sie besitzen einen direkten Ausgang mit einer Leerlaufspannung von 2,75 V bei einem Innenwiderstand, der praktisch gleich Null ist, wobei 100 mW an einen Abschlusswiderstand von 75 Ohm abgegeben werden können.

Seite 10

PROGRAMMIERBARER FREQUENZHUBMESSER

In den USA besteht eine steigende Nachfrage nach programmierbaren Prüfgeräten. Dies ist ein Anzeichen für die Entwicklung in Richtung auf Automatisierung in der Elektronik-Industrie. Die Firma Marconi Instruments stellt zur Zeit keine programmgesteuerten Prüfgeräte her. In diesem Aufsatz wird jedoch eine Methode zur Programmierung des Frequenzhubmessers TF 791D über einen begrenzten Messbereich angegeben.

Seite 11

FS-SENDER SEITENBANDANALYSATOR TF 2360

Dieses Gerät wurde aus dem FS-Sender Seitenbandanalysator OA 1241 entwickelt, welcher in vielen Sendestellen in Grossbritannien und der übrigen Welt benutzt wird. Obwohl die Grundmethode der Anzeige der Seitenbandkurve in ähnlicher Weise beibehalten worden ist, werden in dem Gerät eine verbesserte Schaltungstechnik und bessere Einzelteile verwendet, wodurch sich bei diesem Gerät eine bessere Arbeitsweise und kleinere Abmessungen ergeben haben. Es besteht nicht nur die Möglichkeit zur Durchführung von dynamischen Messungen an Sender-und Modulatorschaltungen für 405, 525 und 625 Zeilen in den FS-Bändern I und III, sondern es sind ausserdem weitere Einheiten lieferbar, mit denen der Bereich des Gerätes auf die UHF-Bänder IV und V erweitert werden kann. In diesem Aufsatz werden die Arbeitsweise und die Schaltung beschrieben und Prüfungen angegeben, welche sich durchführen lassen.

Seite 13

KREUZMODULATIONSMESSUNGEN AN NACHRICHTENVERBINDUNGEN MIT GERINGER KANALZAHL

Mit dem Prüfgerät Typ OA 1249 können mit Hilfe von 'weissem Rauschen' Kreuzmodulationsmessungen in jeder Anlage für maximal 960 Kanäle durchgeführt werden. Bei Anlagen mit geringerer Kanalzahl muss zur Erzielung einer optimalen Genauigkeit ein kleiner Korrekturfaktor angewendet werden.

Eine Tabelle mit Korrekturfaktoren für Anlagen mit 12 bis 60 Kanälen ist angegeben und die Theorie der Messung wird erklärt.

Seite 21

SOMMARIO DEGLI ARTICOLI PUBBLICATI IN QUESTO NUMERO

PONTE UNIVERSALE ALL'1% TF 2700

Il TF 2700 è un ponte di piccole dimensioni e di basso costo, adatto per misurzaioni d'induttanza, capacitanza e resistenza, con una precisione dell'1%.

Questo articolo descrive le caratteristiche di costruzione di questo ponte, soffermandosi brevemente su dettagli riguardanti sia la parte elettrica che la parte meccanica e quotando esempi inerenti ad alcuni tra i più insoliti esperimenti circa le misurazioni dell' impedenza, le quali sono rese enormemente più facili dall'ordinamento schematico di questo ponte.

Pagina 3

NUOVE VERSIONI AD ALTO RENDIMENTO DEI GENERATORI DI SEGNALI CAMPIONI TF 144H

Il TF 144H/4 fino ad -H/6 rappresenta un nuovo raggio d'azione per la serie di generatori di segnali TF 144H, la cui facilità nell' ottenere un'aggiuntivo alto rendimento, rende l'attuale TF 144H-H/2 superato. Il diretto rendimento di queste nuove versioni è di 2,75 V f.e.m. con un'origine d'impedenza praticamente zero, il quale sviluppa 100 mW ad un carico di 75 Ω.

Pagina 10

MISURATORE DI DEVIAZIONI F.M. PROGRAMMABILE

Negli Stati Uniti d'America esiste al momento una forte richiesta per dispositivi di collaudo; questo indica una tendenza verso l'automatizzazione dell'industria elettronica. Marconi Instruments, non produce ancora alcun dispositivo di collaudo programmato, ma quest'articolo vi indica un metodo per la programmazione del Misuratore di Deviazioni FM TF 791D attraverso una limitata gamma di misurazione.

Pagina 11

ANALIZZATORE DI BANDA LATERALE PER TRASMETTITORI TV TF 2360

Quest'apparato è una versione perfezionata dell'Analizzatore di Banda Laterale per Transmettitori TV OA 1241, il quale è tutt'ora in uso presso varie stazioni trasmittenti in questo paese ed all'estero. Sebbene il metodo principale dell'esposizione della potenza sonora delle bande laterali sia simile, questo nuovo strumento s'avvantaggia di una perfezionata tecnica per circuiti e componenti, e come risultato si ha uno strumento di miglior prestazione e di dimensioni considerabilmente più piccole.

A parte l'abilità nell'ottenere misurazioni dinamiche su trasmettitori e su catene modulatrici dei sistemi a 405, 525 e 625 linee nelle Bande I e III, un'altro dispositivo è ottenibile per estendere il raggio d'azione dello strumento e coprire in tal modo le Bande U.H.F. IV e V.

In questo articolo vengono spiegati i principii di funzionamento sottolineando le prove che possono essere fatte, ed una descrizione del circuito.

Pagina 13

MISURAZIONE DELLE INTERMODULAZIONI DEI SISTEMI DI COLLEGAMENTO A BASSA CAPACITA'

La misurazione delle intermodulazioni può essere fatta su qualsiasi sistema con un massimo di 960 canali, usando il Complesso di Misura con Rumore Bianco OA 1249. Ad ogni modo, su sistemi a bassa capacità è necessario applicare un determinato coefficiente correttivo onde ottenerne la più accurata misurazione possibile.

E'fornita una tabella dei coefficienti di correzione per quei sistemi da 12 a 60 canali, con spiegazione delle teorie di misurazioni.

Pagina 21

RESUMENES DE ARTICULOS QUE APARECEN EN ESTE NUMERO

PUENTE UNIVERSAL 1% TF 2700

El puente universal TF 2700 es un instrumento económico y pequeño con una exactitud del 1% para mediciones de inductancia, capacitancia y resistencia.

Este artículo describe varios particulares del puente, tocando sobre los detalles del diseño eléctrico y mecánico y pues da varios ejemplos de experimentos extraordinarios en mediciones de impedancia que este puente puede particularmente facilitar.

Página 3

VERSIONES CON SALIDA MAYOR DEL OSCILADOR TF 144H

En la serie TF 144H, los tipos TF 144H/4 hasta -H/6 dan la facilidad adicional de salida mayor y reemplazan los existentes TF 144H hasta -H/2.

Estas nuevas versiones ofrecen una salida directa de 2,75 voltios con una impedancia de casi cero dando una salida de 100 mW en 75 ohmios.

Página 10

DESVIOMETRO EN ONDA PORTADORA QUE SE PUEDE PONER EN PROGRAMA

En los E.U. se va pidiendo cada vez más, equipos de comprobación que se puedan poner en programa. Esto indica una tendencia a la automatización en la industria electrónica. Hasta la fecha Marconi Instruments no fabrica estos equipos, pero este artículo indica un modo de poner en programa el desviómetro en onda portadora TF 791D sobre un margen limitado de mediciones.

Página 11

ANALIZADOR DE BANDAS LATERALES EN TRANSMISORES DE TELEVISION TF 2360

Este instrumento ha sido desarrollado del analizador de banda lateral para televisión OA 1241 que se usa en muchas estaciones emisoras de televisión en este país y en el extranjero. Aunque el método fundamental de demostrar la respuesta de la banda lateral es parecido, este nuevo instrumento hace uso de técnicas en circuitos y componentes mejorados que resultan en un instrumento mejorado y considerablemente màs pequeño.

Además de su habilidad en hacer mediciones dinámicas en emisoras y moduladoras en sistemas de 405, 525 y 625 lineas de las Bandas I y III, existe una unidad adicional que extiende el margen del instrumento para cubrir las Bandas UHF IV y V.

Este artículo describe el modo de operación y las pruebas que se pueden hacer y una descripción del circuito.

Página 13

MEDIDAS DE INTERMODULACION EN SISTEMAS DE ENLACE CON BAJA CAPACIDAD

Medidas de intermodulación se pueden hacer con el equipo de prueba por ruido blanco OA 1249 en cualquier sistema que tenga una capacidad máxima de 960 canales.

En sistemas de baja capacidad, sin embargo, es necesario que se aplique un pequeño factor de corrección para obtener las medidas de mayor exactitud. Se da una lista de factores de corrección para sistemas que tengan 12 hasta 60 canales y se explica la teoría de mediciones.

Página 21

SIGNAL GENERATORS

NOISE GENERATORS

OSCILLATORS

OSCILLOSCOPES

PULSE GENERATORS

MODULATION METERS

DISTORTION ANALYSERS

SPECTRUM ANALYSERS

FREQUENCY COUNTERS AND METERS

POWER METERS

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