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G. Marconi

CONTENTS

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Editor: H. M. DOWSETT, M.I.E.E., F.Inst.P.
Assistant Editor: L. E. Q. WALKER, A.R.C.S.

FREQUENCY MEASURING EQUIPMENT

The necessity for a more accurate method of measuring frequency than that provided by wavemeter is stressed and an alternative, depending on the generation of a standard frequency, is discussed. A description follows of the Marconi Frequency Measuring Equipment Type 482C, which operates on this principle. An essential feature of the apparatus is the production of a family of standard frequencies by means of synchronised multivibrators. Details of four methods of frequency measurement are given and the "Overall Uncertainty" (this term is defined in the article) is shown to vary from 1 *part in* 10^5 *for the least accurate to* 5 *parts in* 10^7 *for the most accurate method.*

FROM the beginning of radio transmission and reception the need for accurate measurements of frequency has always been recognised. The earliest measurements were made by wavemeter, but the considerable increase in transmitters during recent years has necessitated a more accurate method of measurement. The highest accuracy achieved by wavemeter measurements has, owing to thermal and mechanical difficulties, never exceeded about 1 in 10,000, and this is quite insufficient for checking compliance with the International regulations governing the permissible frequency variation of broadcast transmitters.

A more accurate measurement is by comparison with a standard frequency sufficiently close to enable a heterodyne beat to be obtained with the unknown frequency. The accuracy of such a measurement depends on the accuracy of the standard frequency and the device for measuring the heterodyne beat.

The Marconi Company early began the development of Frequency Measuring Equipment on these lines, and for some time apparatus has been available employing a Master Oscillator of good long-time and short-time stability, synchronised frequency dividers rich in harmonics providing a complete series of standard frequencies in decade steps, and Interpolating Oscillators which facilitate rapid measurements between the standard frequencies generated by the frequency divider. The instrument is provided with means for routine checking against Standard Time Signals.

The apparatus will be considered in detail under the following headings. (1) Description, (2) Methods of frequency measurement and (3) Accuracy of these methods.

(1) Description.

The Marconi 482C Frequency Measuring Apparatus is mounted in a 19-inch rack, approximately 6 feet high and it consists of five panels, arranged in the following order from top to bottom as shown in the photograph in Fig. 1:

(A) High Frequency Interpolating Oscillator of fundamental range 3,000 to 6,000 kcs. and 6,500 to 15,000 kcs. sec.

(B) Medium Frequency Interpolating Oscillator of fundamental range 150 to 300 and 500 to 1,000 kcs./sec.
(C) Five Multivibrators of fundamental frequencies, 1,000, 100, 10, 1 and 0.1 kcs./sec. and a synchronous clock.

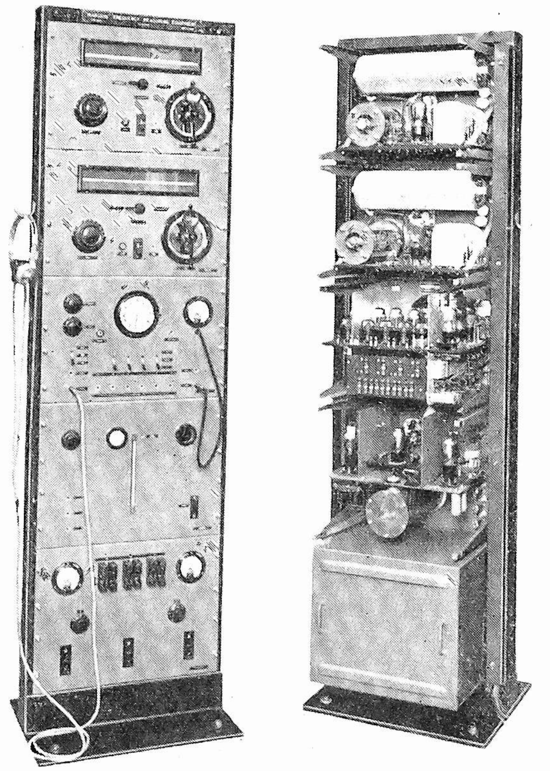


FIG. 1.

(D) Master Oscillator containing a temperature controlled crystal oscillating at a frequency of 250 kcs./sec.
(E) Power Supply.

A schematic diagram of the apparatus is given in Fig. 2.

Power Supply.

Power for operating the apparatus is obtained from the supply panel, which is suitable for all standard A.C. supply voltages and frequencies. A valve rectifier supplies 180 volts H.T., and a copper oxide rectifier 2 volts L.T. Adequate smoothing reduces the ripple voltage in the L.T. and H.T. supply line to less than .1 per cent. and .01 per cent. respectively. While the apparatus may be operated from the A.C. supply only, it is recommended that floating batteries are used for both H.T. and L.T. supply since these can maintain the apparatus in correct operation during short period supply failures, a necessary precaution if the synchronous clock is being checked against standard time signals.

Master Oscillator.

The Master Oscillator is controlled by a crystal of low temperature-coefficient housed in a constant temperature tubular oven, shown in Fig. 3. Temperature

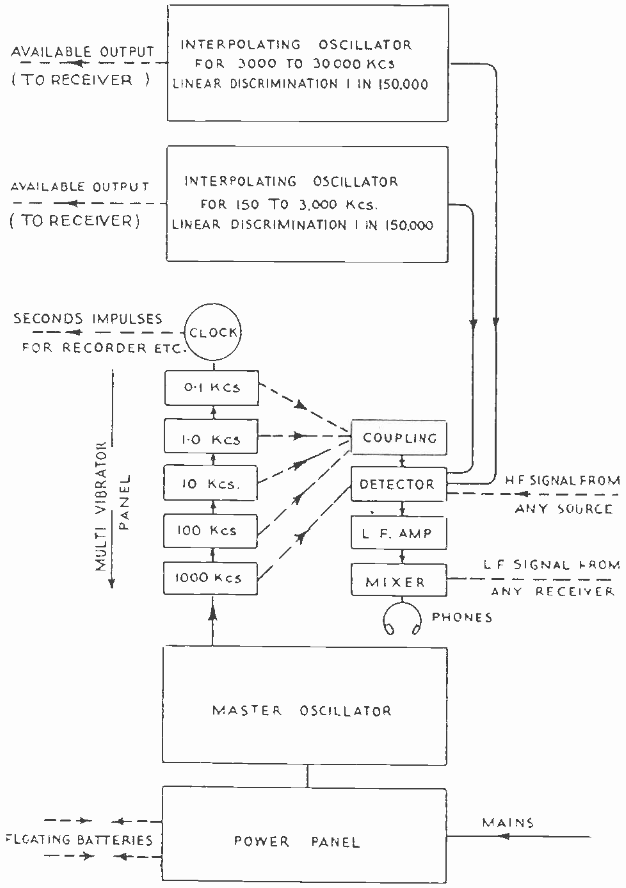


FIG. 2.

control is maintained by using the amplified voltage from an A.C. resistance bridge to operate a gas-filled relay valve connected to the heater elements. Two arms of the bridge have small and the other two arms large, temperature-resistance co-efficients. The latter arms are in close contact with the heater windings so that little temperature hunting occurs; the maximum hunt period is about 20 seconds. Valve amplification between the relay and bridge makes the effect of variation of ignition temperature of the gas-filled relay almost negligible. Extra heater windings fitted in the ends of the crystal container modify the internal temperature distribution so that variations of ambient temperature have minimum effect at the crystal. The curves in Fig. 4 show the short time (about 3 hours) taken for the Master Oscillator to reach temperature and frequency stability from rest.

One of the chief difficulties in maintaining really high stability of frequency concerns the prevention of capacitance changes to the crystal as a result of random vibration. An interesting feature of the Master Oscillator is the suspension of the crystal in a silk cradle and the use of very large air-gaps. The effect of the air-gap on frequency, is shown in Fig. 5, and the curve indicates that a crystal-electrode

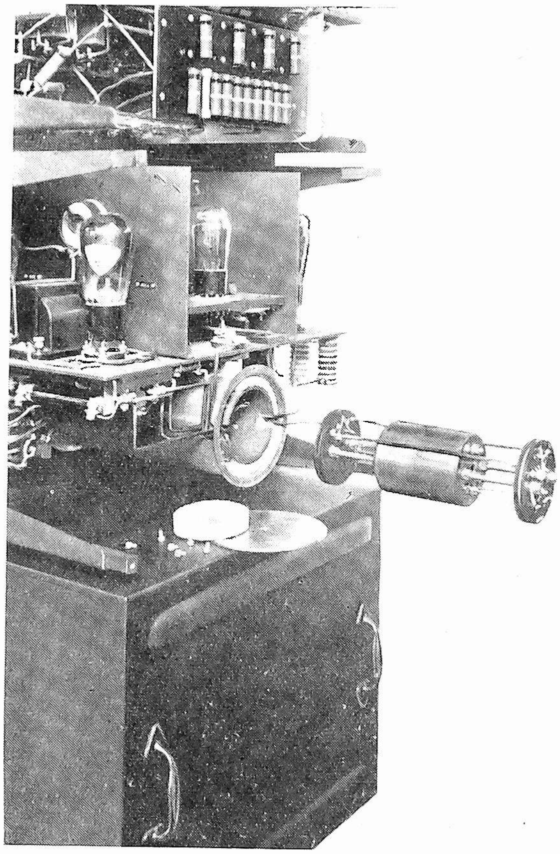


FIG. 3.

spacing midway between the second and third air-gap resonances gives a much smaller frequency change for a given (unintentional) displacement of the crystal. Each air-gap resonance is clearly shown by inflections in the curve. The gaps on either side of the crystal are made equal so that crystal displacement produces changes which tend to cancel each other. Fig. 6 gives curves of frequency variation plotted against angular displacement of the holder with respect to the crystal. A cross-coupled double-triode valve is used to maintain the crystal in oscillation and a variable capacitance shunted by a low "Q" coil in one grid circuit gives convenient fine control of frequency. This capacitance is connected to a dial giving a frequency variation of approximately 1 in 10^7 per scale division, and it can be adjusted by a key fitting into the front panel.

The output from the Master Oscillator is obtained via a screened grid separator valve, the anode circuit of which is tuned to 1,000 kcs./sec., the fourth harmonic of the oscillator. The effect of changes of the output load on the stability of the Master Oscillator is therefore negligible.

The following table indicates the stability of the Master Oscillator under varying conditions :—

CAUSE.	EFFECT ON FREQUENCY.
Sustained ambient temperature change	Less than 1 in 10^7 per degrees C.
Barometer change	Less than 2 in 10^7 per cm. Hg.
High tension change	Less than 1 in 10^7 per 1 per cent.
Low tension change	Less than 1 in 10^7 per 1 per cent.
Extreme output load change	Less than 1 in 10^7.

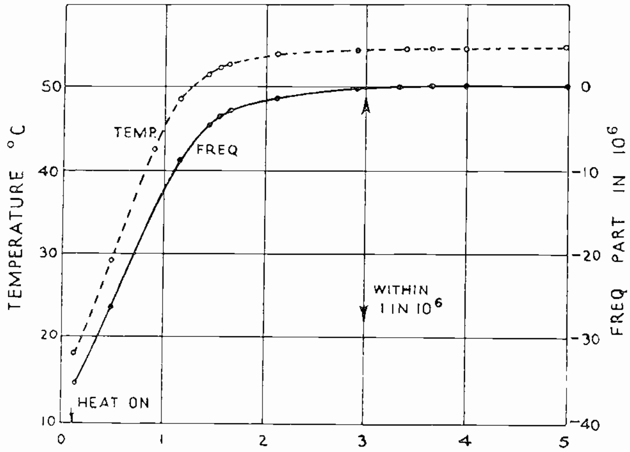


FIG. 4.

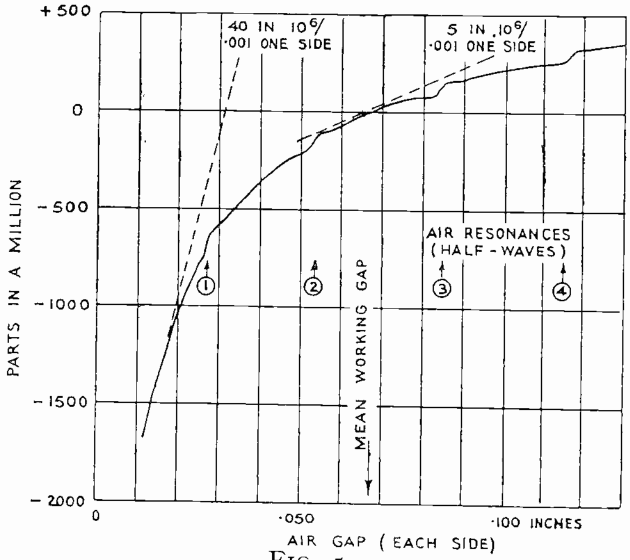


FIG. 5.

Multivibrators.

In order to obtain a family of standard frequencies from one standard frequency, the latter must be used to control an oscillator producing a wave form rich in harmonics. A multivibrator oscillator has sharp discontinuities in wave form and is thus suitable for the purpose.

It can be synchronised with a frequency close to its free working frequency, and, moreover, frequency division is possible if a series of multivibrator circuits of decreasing frequency steps are used, each synchronised by the preceding stage. Decade division is generally preferable and the Multivibrator panel has five multivibrators operating at fundamental frequencies cf 1,000, 100, 10, 1 and .1 kcs./sec. Each is synchronised by the previous stage, except the first, which is connected to the separator valve in the Master Oscillator. The last multivibrator drives a synchronous clock which is not self-starting. Second and Minute contacts on the clock enable comparison with Standard Time Signals to be made on the Marconi Double Pen Recorder Type UG7. The ambiguity of the Second contact is not greater than about .005 second. Each multivibrator—a double triode valve with grids and anodes cross-connected—is synchronised by a pulse introduced into one anode circuit from the previous stage.

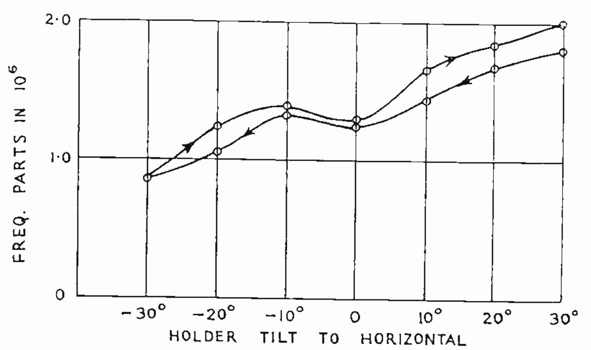


FIG. 6.

Recent improvements in the multivibrator circuits have reduced to small proportions feedback of one multivibrator circuit to the succeeding stage, have enabled synchronism to be maintained over larger variations in supply voltage (L.T. from about 1.6 to 2.3, H.T. from about 130 to 230 volts), have removed discontinuous changes in synchronism, and have reduced considerably the anode currents taken by the multivibrators thus prolonging the life of the valves.

Coupling switches connect each multivibrator (except the first which is permanently coupled) to a socket on the front panel and to a listening post consisting of a detector and two valve L.F. amplifier. Any set of multivibrator harmonics may therefore be used for heterodyne purposes.

On the panel is a switch having the following three positions :—

(1) The telephones are connected to the output of the L.F. amplifier and beats are heard between the Multivibrator and Interpolating Oscillator outputs.

(2) The telephones are connected to the output of any receiver which may have been connected to the equipment. In this case the output from the appropriate Interpolating Oscillator is normally mixed with the received signal at the aerial of the receiver.

(3) A switch position midway between (1) and (2) allows both outputs to be heard simultaneously. The volume levels of positions (1) and (2) are separately controlled.

Interpolating Oscillators.

The two Interpolating Oscillators are almost identical in design and each has two frequency ranges. The Medium Frequency Oscillator covers from 150 to 300 kcs./sec. and 500 to 1,000 kcs., and the High Frequency Oscillator from 3,000 to 6,500 kcs. and 6,500 to 15,000 kcs./sec. Frequencies not in these fundamental ranges can easily be measured by using harmonics.

Coarse frequency control is obtained by capacitance variation. A twin-ganged capacitor with slow motion drive is coupled to an illuminated scale drum calibrated in kcs./sec. and the appropriate scale is brought into view as the capacitor is turned. The range switch besides selecting the appropriate coils, controls the position of the scale drum. A Colpitts circuit is employed and the oscillator is very stable after the initial short warming-up period.

Fine frequency control is obtained by variation of inductance and is specially designed to give a linear frequency law. This variable inductance control has the added advantage of providing almost constant percentage frequency discrimination over all the frequency ranges. The fine control has a circular scale and its pointer is illuminated from behind to avoid parallax. This pointer is mechanically coupled to a second pointer which is caused to travel across the calibrated scale drum operated by the coarse frequency control.

Brass rotors acting as "short circuit" turns, advance into the coils on a screw thread cut in the spindle which is supported on one rigid and one sprung nut to eliminate backlash.

The total travel of the spindle is $7\frac{1}{2}$ turns giving a frequency change of 5 per cent., or frequency discrimination of approximately 0.7 in 10^5 per half division on the scale.

"Pulling" of the oscillator is reduced by the use of a separator screened grid valve and slow beats with the multivibrator sources are easily obtained. A socket on the panel enables the separator valve output to be connected to any desired receiver.

Having described the apparatus it is necessary now to consider the method of using it for frequency measurements.

(2) Methods of Use.

There are four convenient ways of using the apparatus, viz.:

(1) By R.F. interpolation using the Interpolating Oscillators.

(2) By heterodyning one of the Interpolating Oscillators with a strong multivibrator harmonic in such a way that the sum or difference of these two frequencies equals that of the received signal.

(3) By using a tone generator method for measuring the audible heterodyne note produced by a suitable multivibrator harmonic beating with the received signal.

(4) By counting the heterodyne beats between the received signal and a multivibrator harmonic.

Methods 2, 3 and 4 require a knowledge of the approximate value of the unknown frequency. The choice of any particular method must depend on the position in the frequency spectrum of the frequency to be measured and its proximity to a strong multivibrator harmonic.

Method 1. R.F. Interpolation.

The Interpolating Oscillator is suitably coupled to the aerial of a receiver which is tuned to receive the station whose frequency is required. The output of the receiver is connected to the phones on the multivibrator panel. The Interpolating Oscillator frequency is adjusted to give zero beat with the incoming signal when the three position switch connects the receiver output to the phones. The switch is then moved to the position connecting the phones to the Multivibrator Listening Post. The oscillator is calibrated at the time of use against suitable multivibrator harmonics on either side of the incoming signal frequency position. The unknown frequency can be obtained by simple interpolation since the oscillator frequency scale is linear between the two multivibrator harmonics.

Rapid measurements may be made by this method and errors due to heterodyning with a harmonic instead of the fundamental of the interpolating oscillator are most unlikely to occur. Care must be taken when a superheterodyne receiver is used, since an apparent heterodyne may be obtained when the Interpolating Oscillator frequency is equal to the receiver oscillator frequency or the sum of the receiver oscillator frequency and the intermediate frequency of the receiver. Both these conditions are readily recognisable since detuning of the receiver changes the heterodyne note. Zero beat may in certain conditions be difficult to distinguish and an improvement can be obtained by adjusting the Interpolating Oscillator to 1,000 $\sim$/sec. from the unknown frequency and beating this against the 1,000 $\sim$/sec. multivibrator with the mixing switch on the multivibrator panel in the mid-position.

The Interpolating Oscillator frequency will then be 1,000 cycles/sec. higher or lower than the signal frequency, and a check test for approximate zero beat will remove any ambiguity.

Method 2. Heterodyning by the sum or difference of Interpolating Oscillator and Multivibrator Frequencies.

The Medium Frequency Interpolating Oscillator may be used to bridge the gap between one powerful 1,000 kcs./sec. multivibrator harmonic and the signal, and a heterodyne note will be heard in a receiver when the sum or difference of these two frequencies approaches that of the received signal.

The outputs from Multivibrator and Interpolating Oscillators are both coupled to the aerial or to some suitable point in the receiver and the output of the receiver connected to the phones. The Interpolating Oscillator can be calibrated as in the first method.

This method is most suitable for measurements on weak signals at high frequencies using a selective and sensitive receiver and is more accurate than Method 1. The approximate frequency must first be known and this can be done by Method 1. Ambiguity may be caused in some cases where the incoming signal frequency is nearly half-way between two powerful multivibrator harmonics since heterodyning is obtained with sum and difference terms. Except in special cases Method 1 can remove this ambiguity.

Method 3. Tone Generator as Comparator.

In this method the Multivibrator output is coupled to the receiver input and the resulting audio frequency output is mixed with a tone generator to give zero beat. A separate tone generator is not required since the Medium Frequency Interpolating Oscillator may be used in conjunction with the multivibrator as a tone generator. This has the advantage that the Interpolating Oscillator can be calibrated at the time of use, its scale is linear and frequency discrimination is good, for example if the oscillator is used in the region of 160 kcs. a discrimination of approximately 400 divisions per kilocycle may be obtained.

Comparison of the tone generator and receiver output is made by means of the three position switch on the multivibrator panel.

The audio tone from the receiver may be measured by a stroboscope but it should be noted that harmonics in the tone may produce ambiguity.

This method is much easier than Method 2, and is very useful for measuring the drift of medium frequency stations whose frequency is within 3 kcs. sec. of that of a strong multivibrator harmonic.

Method 4. Slow Beats with a Multivibrator Harmonic.

This form of measurement is only suitable when the unknown frequency is within approximately 30 cycles/sec. of a multivibrator harmonic.

The output from the multivibrator is coupled to the aerial of the receiver and the slow beats are counted visually or aurally against a clock or by the Marconi Double Pen Recorder Type UG7. The practical range of beat is from approximately 0.1 to 5 cycles/sec. for clock counting and from .01 to 30 cycles/sec. for recorder counting. It is advisable (essential in the case of the pen recorder) to heterodyne the multivibrator frequency by one of the Interpolating Oscillators so that the beat between multivibrator and signal may be brought into the pass band of the I.F. circuits of the receiver. The setting of the oscillator does not affect the measurement since the slow beat frequency modulates the audible heterodyne note.

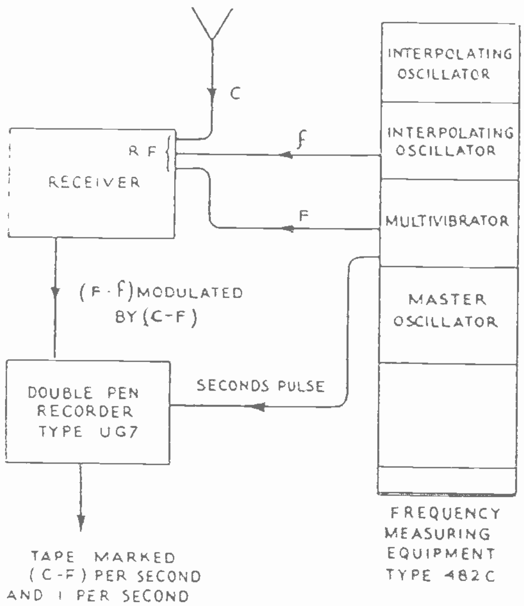


FIG. 7.

In Fig. 7 is given a schematic diagram of the apparatus set up for slow beat measurement on the double pen recorder.

This method of measurement is the most accurate since it depends only on the accuracy of the Master Oscillator frequency and is particularly useful for checking the frequency drift of medium frequency stations of good stability.

Accuracy of Frequency Measurement.

In defining the performance of precision apparatus it is essential that the meanings assigned to the terms generally used, namely, " Error," " Uncertainty " and " Accuracy " should be clearly stated. " Error " is the difference between the real or absolute value and the measured value. It cannot be completely expressed as a simple numerical value but only as less than a particular value. " Uncertainty " is the maximum error possible (though not perhaps likely) in a particular measurement process. Generally several uncertainties each requiring separate consideration are involved and their summation gives a reliable overall uncertainty. " Accuracy " is the inverse of uncertainty and represents the merit of the measurement

For any particular measurement using the Marconi 482C Frequency Measuring Equipment the overall accuracy depends upon—

(1) The amount of " Uncertainty " which must be attributed to the Master Oscillator Frequency. (This is in the nature of a " Datum Uncertainty " which is always applicable and is independent of the method of comparison.)

(2) The amount of " Uncertainty " introduced by the method used to compare the unknown frequency with the multivibrator harmonics. (This can be called the " Comparison Uncertainty ").

The results will be summarised after a separate discussion of these " Uncertainties."

Master Oscillator Uncertainty.

This " Uncertainty " has been found to be less than 1 part in 10^6 over periods of many weeks and for short periods it is not greater than ± 3 or 5 in 10^7. The graph in Fig. 8 shows the variation in frequency of two Master Oscillators checked against a third.

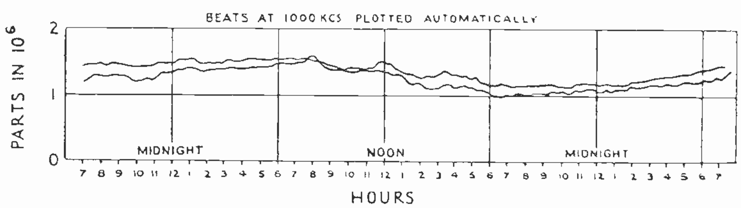


FIG. 8.

With a correctly installed equipment subjected to the usual routine checks, it is therefore possible to assume that the " Datum Uncertainty " is never greater than ± 1 in 10^6, and that, apart from abnormal changes of ambient or supply conditions, the Uncertainty is no more than ± 5 in 10^7.

Before making use of the latter assumption, proper evidence from reputable Time Signals should be available in respect of several days prior to the measurement. Official circulars, issued periodically concerning the best Time Signals available, sometimes show changes of correction as large as 0.02 seconds, which corresponds to 2.3 in 10^7 for a 24-hour run.

Comparison Uncertainties.

The Comparison Uncertainties will vary according to the method of measurement and each of the four methods will be examined separately.

Method 1. *By R.F. Interpolation.*

The Uncertainty is less than 1 in 10^5 over both ranges of Interpolation Oscillators.

Method 2. *Sum or difference of Local R.F. sources as heterodyne.*

The Comparison Uncertainty in this case is variable, depending on the frequency of the unknown signal in relation to the 1,000 kcs./sec. multivibrator harmonic. For example, if the unknown frequency is about 12,200 kcs./sec. it may be measured by using the 12th harmonic of the 1,000 kcs./sec. multivibrator, and the 150 to 300 range of the Interpolating Oscillator. The latter has an Uncertainty of 1 in 10^5 which is 2 in 200 kcs./sec., thus the Comparison Uncertainty of the measurement becomes 2 in 12,200 kcs. or 1.64 in 10^7. If the unknown frequency is about 12,900 kcs., the Comparison Uncertainty becomes 9 in 12,900 kcs. or 6.98 in 10^7. The lowest range in the Interpolating Oscillator is 150 kcs., so that the difference frequency, viz., 13,000—100 cannot be used. A frequency of 12,800 would have made this possible and the Comparison Uncertainty would be 2 in 12,800, or 1.56 in 10^7.

As the unknown frequency is decreased the Comparison Uncertainty increases and finally approaches that of the Interpolating Oscillator itself.

Method 3. Using a Tone Generator as a Comparator.

If the Interpolating Oscillator is used in conjunction with the 160 kcs. multivibrator harmonic to provide a tone generator, a discrimination of approximately 2.5 cycles/sec. per division on the Interpolating Oscillator dial is obtained, so that assuming readings may be made to $\frac{1}{2}$ division the Uncertainty is less than 1.6 cycles/sec. This Uncertainty is fixed and independent of the measured frequency so that at a frequency of 500 kcs./sec. the Uncertainty is 1.6 in 500,000, or 3.2 in 10^6, whereas at 7,000 kcs./sec. it is only 0.23 in 10^6.

Method 4. Slow Beats with a Multivibrator Harmonic.

The Uncertainty of this method is practically that of the Master Oscillator itself, which for a well run equipment is ∓ 0.5 in 10^6.

Summary of Overall Uncertainties of Measurement.

If 0.5 in 10^6 is assumed to be the Master Oscillator Uncertainty the overall Uncertainties may be summarised as follows :—

Method.	Uncertainty.
1	± 1 in 10^5.
2	± 0.5 in 10^6 $\pm$ 10 cycles/sec.
3	± 0.5 in 10^6 $\pm$ 1.6 cycles/sec.
4	± 0.5 in 10^6 $\pm$ 0.1 cycles/sec.

Conclusion.

The Marconi 482C Frequency Measuring Equipment has been described and the methods of measurement, together with the overall Uncertainties involved in the use of these methods, have been discussed. In conclusion it is worth noting that the probable Error for an average measurement is less than the Uncertainty shown.

N. Lea.
K. R. Sturley.

A NEW TYPE OF ANTI-FADING AERIAL

This article describes a new type of anti-fading aerial recently erected in Brazil.

THE Marconi Anti-Fading Aerial described herewith has been in operation for Radio Tupi at São Paulo, Brazil, since the beginning of August 1937, and it is now permissible to state that from the outset the results have been consistently

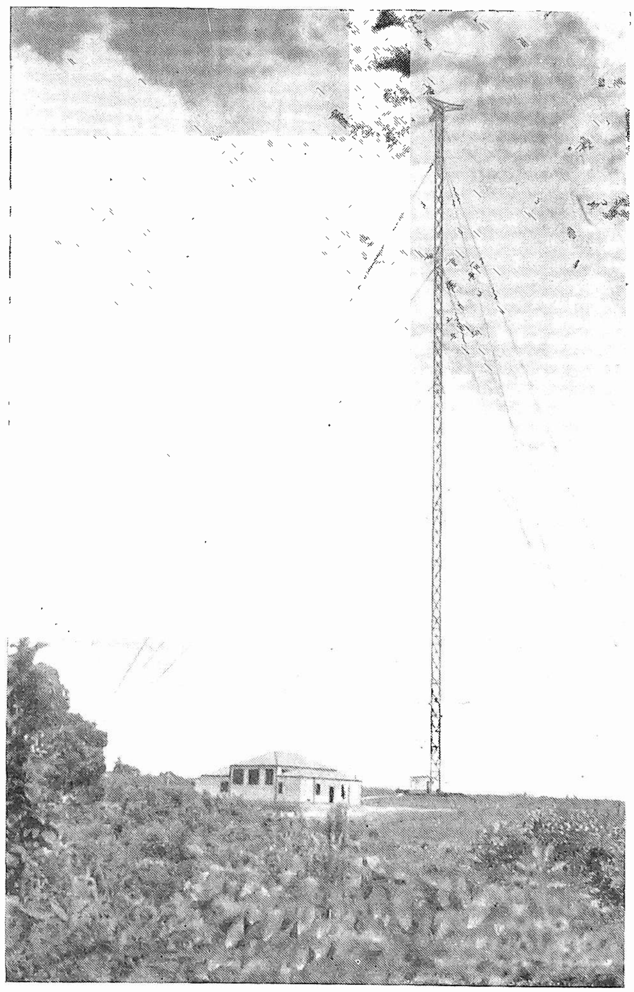


FIG. 1.

satisfactory. An inspection of the illustrations will reveal an entirely new principle in design : briefly the aerial consists of a triangular stayed mast supporting a three-wire symmetrically disposed plain vertical radiator, the three wires being suspended from a top framework formed of three 5-metre arms, thus giving the radiating surface an effective diameter of 10 metres. The object of the scheme is threefold—to eliminate the capacity effect of the stays as regards current distribution; to present an almost perfectly uniform wave impedance throughout the entire height of the radiator; and to reduce the high frequency voltages across the stay insulators.

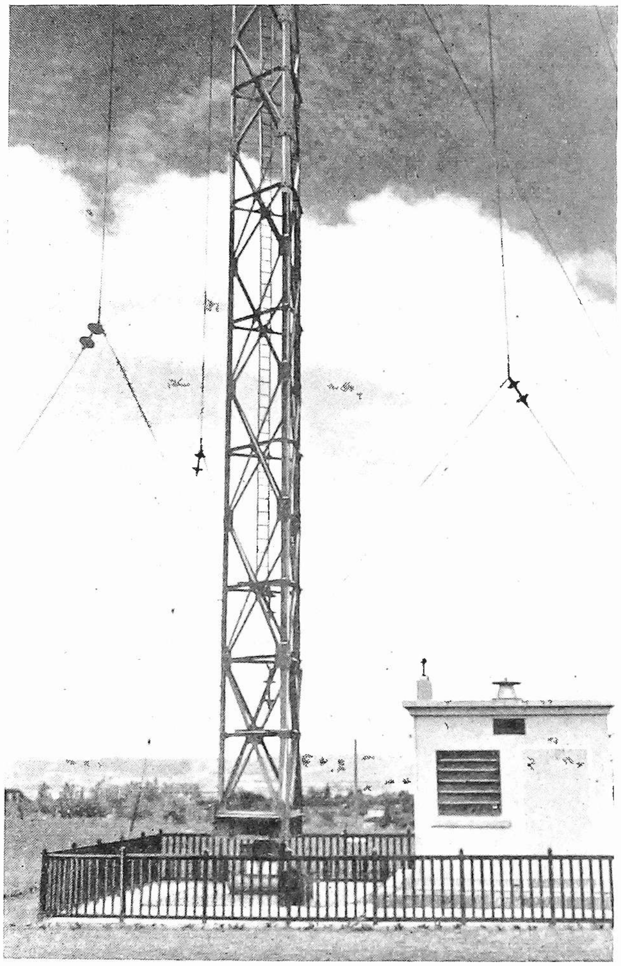


FIG. 2.

That the first object has been achieved is shown by the fact that the ratio of half the free natural wavelength, which is 309 metres, to the physical length of the radiator, which is 145 metres, is 93 per cent., whereas the ratio in the case of a stayed mast radiator of similar structure, and for which the stay end capacity effects have been carefully reduced to a minimum, is of the order of 80 per cent. Since for a given sinusoidal current distribution it is height alone that determines effective radiation, it will be appreciated that 93 per cent. is a figure of unusual merit.

From the foregoing it is obvious that the three vertically supported wires, on which the stays have no direct and very little indirect effect, must approximate closely to the ideal of a vertical uniform conductor.

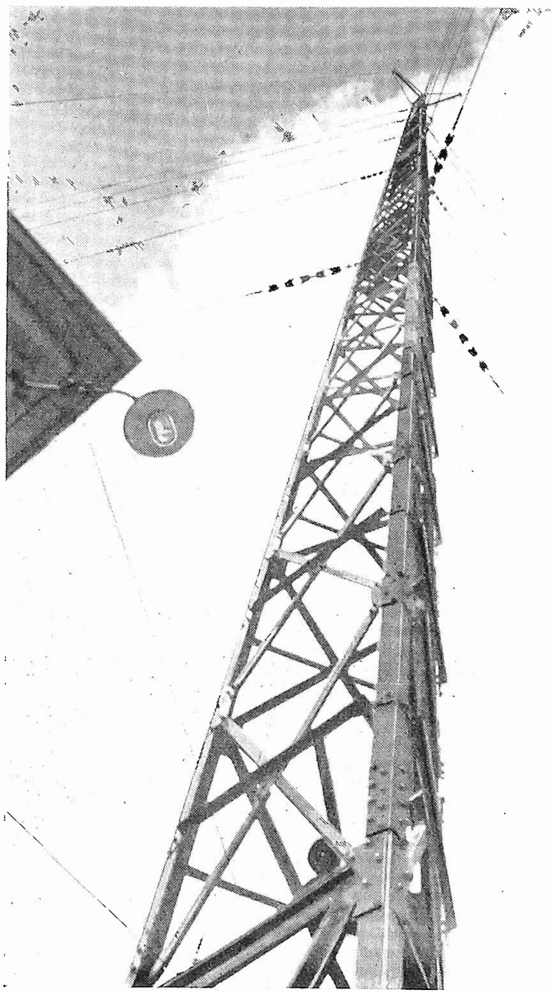


FIG. 3.

Finally, absence of flash-over trouble at the stay insulators, which are of the large walnut type, leads to the inference that such troubles have been materially reduced, if not altogether eliminated.

The mast height gives a vertical distribution of 193 degrees at the working wavelength of 288.5 metres, and this height has been proved a satisfactory average as regards anti-fading effects.

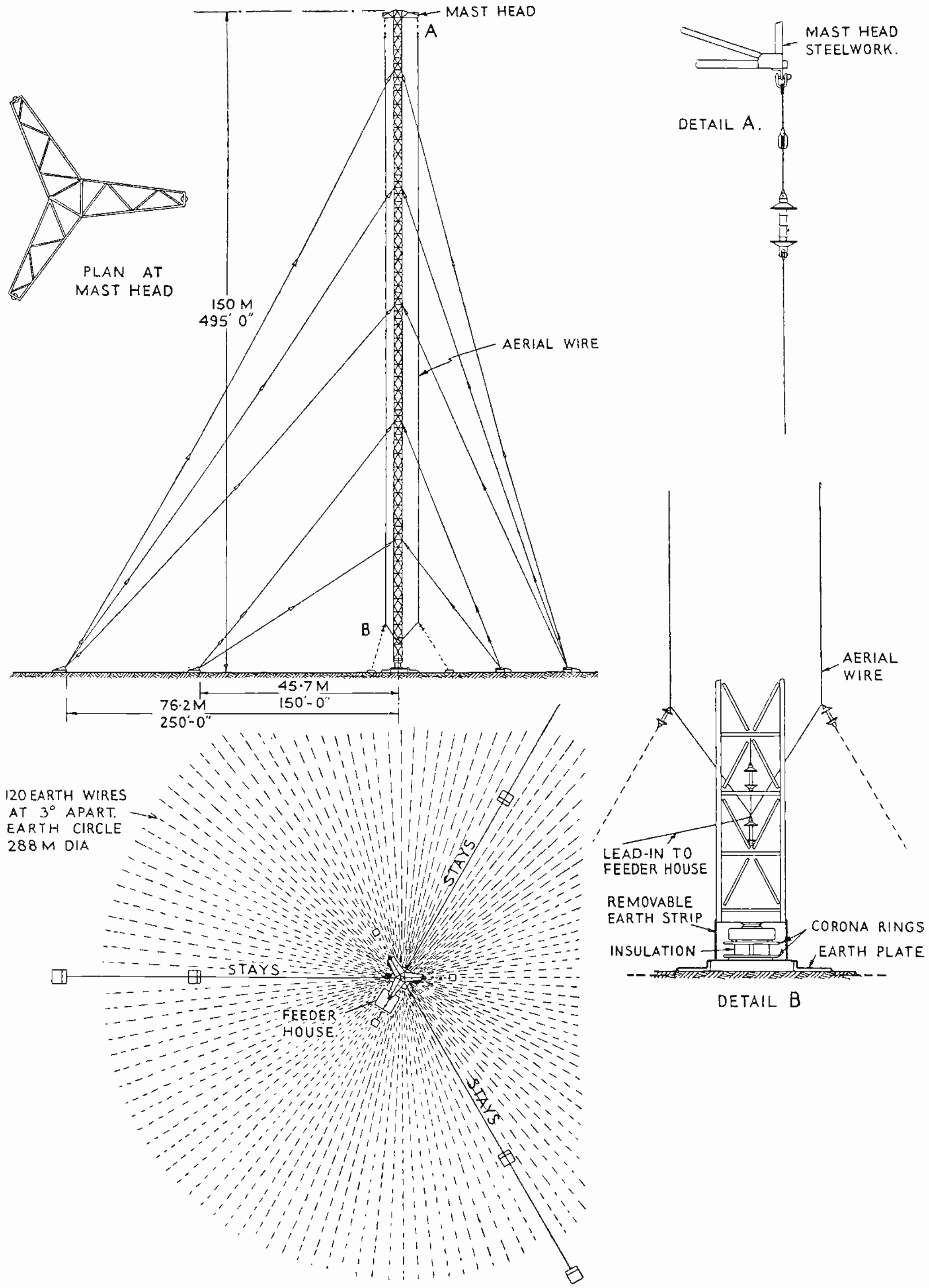

MAST HEAD
A
MAST HEAD
STEELWORK.
DETAIL A.
PLAN AT
MAST HEAD
150 M
495' 0"
AERIAL WIRE
B
45·7 M
150'-0"
76·2 M
250'-0"
AERIAL
WIRE
120 EARTH WIRES
AT 3° APART.
EARTH CIRCLE
288 M DIA
STAYS
STAYS
STAYS
FEEDER
HOUSE
LEAD-IN TO
FEEDER HOUSE
REMOVABLE
EARTH STRIP
CORONA RINGS
INSULATION
EARTH PLATE
DETAIL B

Fig. 4.

An interesting feature is that the central mast has no effective influence on the characteristics of the aerial proper. As a precautionary measure the mast was erected with a base insulator, but careful experiments under power revealed no difference in either distant or home signals whether the mast was insulated or shorted; also the terminal tuning between these two conditions only differed by a fractional amount. As will be seen from an examination of Figure 3 the mast is actually in commission with its base insulator shorted, that is, it functions as an earthed mast.

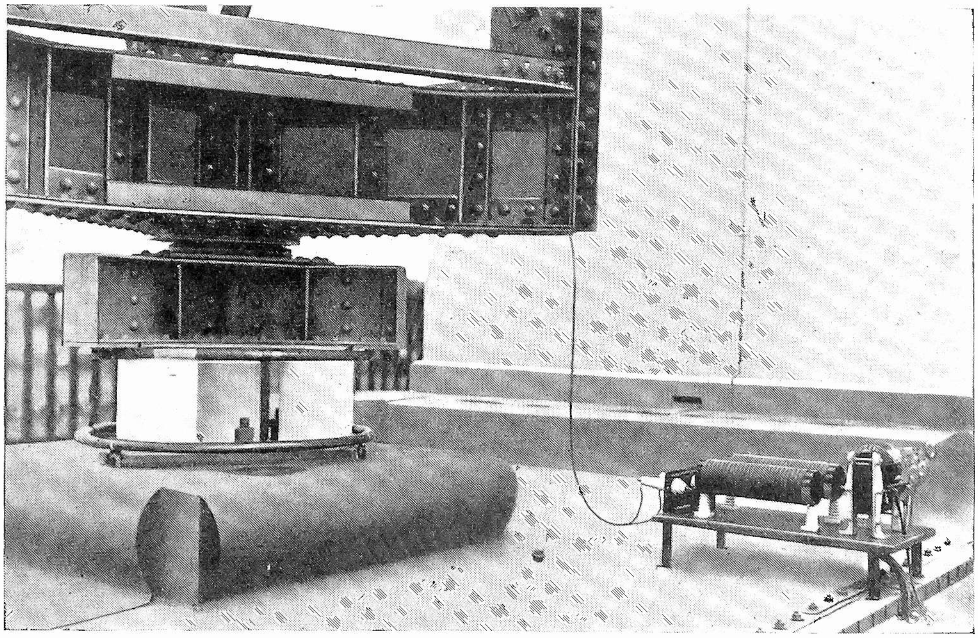


FIG. 5.

The earth system consists of 120 radials of 144 metres radius.

Power is supplied from a Marconi Series Modulated Transmitter, delivering 24 kw. to the aerial via a 600 ohm twin overhead feeder.

Figure 1 is a general view of the aerial.

Figure 2 shows in greater detail the lower portion and the adjacent feeder house.

Figure 3 gives an idea of the stay end insulation.

Figure 4 is a scale diagram of the general arrangement of mast and aerial, and Figure 5 gives details of the base insulator, one shorting bar being clearly visible.

N. WELLS.

SOME NOTES ON IRON-DUST CORED COILS AT RADIO FREQUENCIES

The introduction of iron-dust core materials has made possible the production of compact coils of high efficiency. The nature of the material is discussed and methods of manufacture outlined. The advantages and disadvantages of various shapes of core are compared, and methods of inductance adjustment noted. Consideration is given to winding desiderata and screening losses. Curves are given shewing the variation of coil magnification with frequency for various types of iron-cored coils over a wide frequency range, which enables the efficiency of most of the usual gauges of wire to be compared. The effects of the variation of temperature are noted.

THE use of ferro-magnetic powder compositions to increase the efficiency of coils used in radio-frequency apparatus is now well known and accepted practice and much work has been carried out of recent years in investigating the behaviour and characteristics of these magnetic compounds at radio-frequencies.

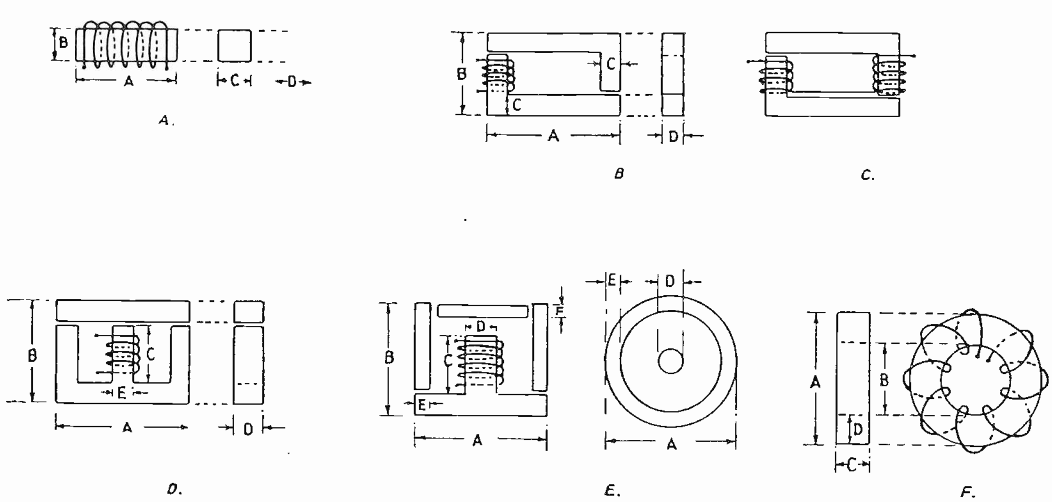
The names of Polydoroff in America and Vogt in Germany will be associated with the development of iron-dust cores in the forms which are commercially available to-day. Accounts of their work appeared in 1932 and 1933 and there was a certain amount of controversy as to the mechanism by which the layer construction, due to Vogt, served to reduce the losses inherent in a powdered iron core. In this connection the recent development of the pot type of core emphasised the fact that in the design of magnetic circuits using laminated material it is advisable to avoid allowing the path of the flux to cross the layering, as this leads to a considerable increase in the losses and in the reluctance of the circuit; these points are dealt with in greater detail in the notes on the pot type of core.

Round used iron-dust cores, made with reduced iron and paraffin wax, about 1904 and subsequently employed iron cores for inter-valve transformers at a frequency of about 3,000 kc. (100 metres) in the Type 55 Amplifiers of the Marconi Company and this was probably the first commercial application of iron cored coils to frequencies of this order. These transformers were fitted with a screw adjustment which enabled the iron core to be moved relative to the windings and gave a form of permeability tuning.

Vogt in his early experiments investigated the possibilities of iron wire cores and was satisfied that the very fine wire required to make up cores which would be effective at broadcast frequencies was not a practicable proposition and approached the problem by the alternative method of building up magnetically aligned insulated iron particles on layers of paper, the resultant material being marketed under the trade name of Ferrocart.

Polydoroff took a powdered iron and developed an insulating varnish used to insulate the iron particles and with the addition of a small percentage of binding material subjected the mixture, in heated moulds, to pressures of the order of 25 tons per square inch. The iron-dust cores which resulted were mechanically strong and capable of being machined if required, although the desired shape could normally be produced by moulding.

Powdered iron may be produced by a number of processes, mechanical such as crushing and grinding, chemical as precipitation from solution or reduction from compounds, and also by electrolytic deposition. The powdered iron required for iron-dust cores must be of high purity and consistent grain size, and although iron of high purity can be obtained by electrolysis, in which electrolytes using ferrous



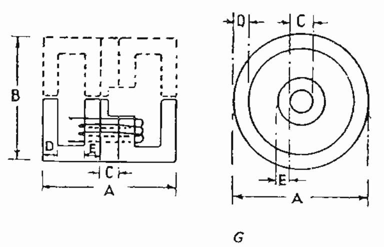


FIG. 1.

Typical approximate dimensions (in millimetres).

FIG. 1A.

A	13	22	27	13
B	6	12	8.5	—
C	6	10	8.5	—
D	—	—	—	9.5

FIG. 1B and 1C.

A	20	34	40
B	15	34	34
C	5	10	10
D	3	12	12

FIG. 1D.

A	30
B	25
C	15
D	12
E	10

FIG. 1E.

A	28	35.5	45
B	18	40	45
C	10	26	28
D	10.5	14	18
E	2	4	5
F	3	5	7

FIG. 1F.

A	57
B	38
C	12.7
D	9.5

FIG. 1G.

A	22
B	7.5
C	7.5
D	1.8
E	2.2

chloride or ferrous ammonium sulphate may be employed, iron obtained by reduction by hydrogen or from carbonyl of iron is obtainable commercially. The chemical composition of a good grade of iron powder would approximate to :—

Carbon	0.02 per cent.
Oxygen	0.10 ,,
Iron	99.88 ,,

Powdered iron, however produced, should be globular in form and of the order of 5 microns (10^{-6} metre) diameter although the optimum grain size is a function of frequency. Use is also made of other magnetic materials for the production of dust cores, nickel alloys in particular being suitable for the purpose, although the resultant permeability of such cores is but little greater than that of those constructed with powdered iron, in the case of cores intended for radio frequencies.

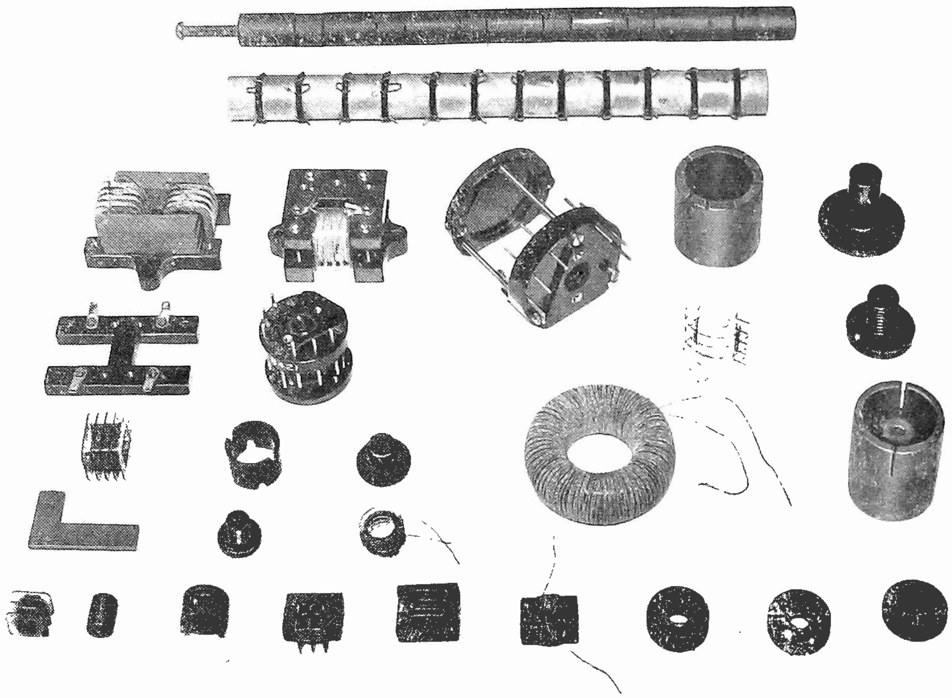


FIG. 2.

(1) *Iron dust core coil with 55 amplifier—rod contains six cores.*

(2) *Transformer used in* (1) *showing six primary and secondary windings.*

(3) *Complete coil and core refer. Fig.* 1C. (4) *Complete coil and core refer. Fig.* 1B. (5) *Clamps for pot. Fig.* 1E *(medium).* (6) *Wall of pot. Fig.* 1E *(medium).* (7) *Centre core and base. Fig* 1E *(medium).*

(8) *Clamp for cores for Fig.* 1C. (9) *Complete small pot. Fig.* 1E. (10) *Coil former used with small pot.* (11) *Adjusting screw used with medium pot. Fig.* 3.

(12) *Former for coils refer. Figs.* 1B *and* 1C. (13) *Wall of pot. Fig.* 1E *(small).* (14) *Centre core and base. Fig.* 1E *(small).* (15) *Toroidal coil and core.* (16) *Early experimental pot refer. Fig.* 5.

(17) *Laminated L section. Fig.* 1C. (18) *Adjusting screw used with small pot. Fig.* 3. (19) *Former (wound) for use with small pot. Fig.* 1E.

(20) *Core and former refer. Fig.* 1A. (21) *Core refer Fig.* 1A. (22) *Former for use with* (21). (23) *Core and former refer. Fig.* 1B. (24) *Core and former refer. Fig.* 1D. (25) *Core and coil Fig.* 1C. *Coil solid.* (26) *Core refer Fig.* 1G. (27) *Core refer Fig.* 1G. (28) *Topdisc used with experimental pot. Fig.* 5.

The apparent permeability (that is as measured by means of a toroidal coil) of powdered iron cores having reasonable losses, varies up to about 20, and 12 may be taken as an average commercial figure. The effective permeability, and by effective permeability is meant the ratio of the inductance of a coil with its iron-dust core to the inductance without the core, does not exceed 3.5 with most designs. A standard commercial broadcast receiver coil tested at 1,000 kc. shewed an inductance with air core of 61 microhenries, with an iron-dust core 178 microhenries, the core having an effective permeability of 2.9. An iron-dust core with a Specific Gravity of 3.8 shewed an apparent permeability of 12, while in another case a Specific Gravity of 5 shewed a permeability of 18. The use of extra high permeability alloys offers no advantages for iron-dust cores since their reluctance is dependent mainly on the gaps between the magnetic particles and it is therefore the losses of the magnetic material rather than its permeability which determine its effectiveness.

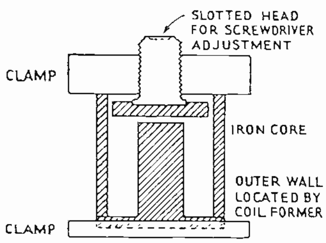


FIG. 3.

The manufacturing process of moulding lends itself to the easy production of various shapes and sizes of powdered iron-dust cores and it has been found that the performance is affected very considerably by the shape and size adopted. The

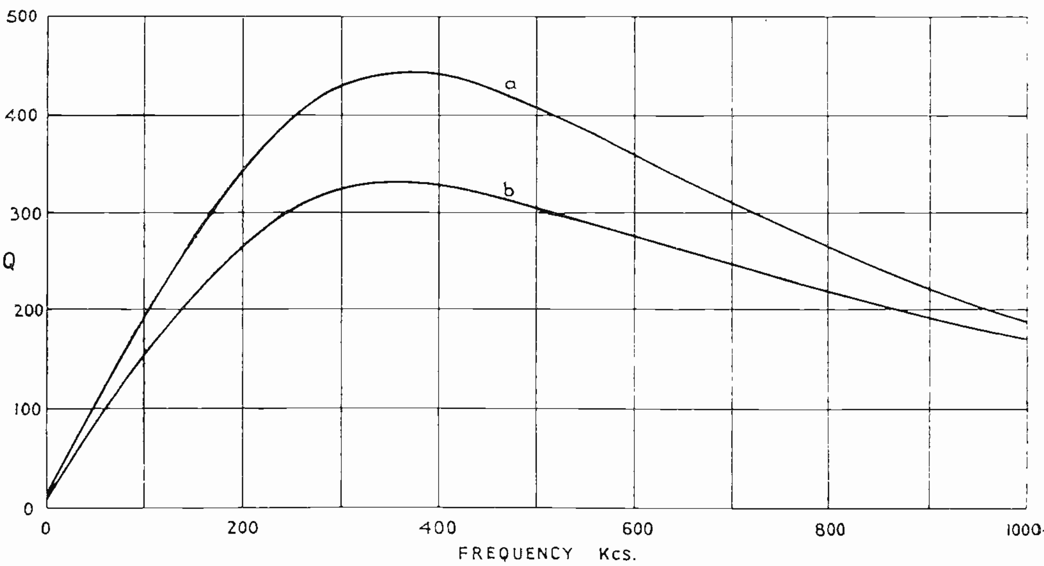
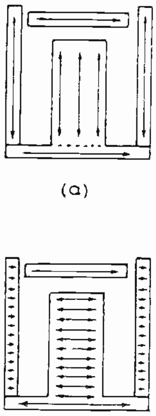


FIG. 4. 60 *Turns* 27/44 *on Ferrocart pot core* 197×10^{-6} *Hys* (*a*). *Unscreened.*
146×10^{-6} *Hys* (*b*)

simplest form of core is the open type (Fig. 1A) consisting of a straight bar core of approximately the same length as the coil in which it is inserted. This type, while capable of quite a good performance, is characterised by a comparatively large external field. Another form of core is rectangular in shape, built of L-sections, forming a closed core, which may carry a single winding on one limb (Fig. 1B) or alternatively an astatic pair of coils (Fig. 1C). A third shape is built up of E and I sections with the coil wound on the centre limb (Fig. 1D). The leakage field due to an astatic pair of coils wound on L-sections would be appreciably less than that of the other closed core types and would be of a similar order to that of the pot type of core shewn in Fig. 1E, in which the coil is wound on a centre core and entirely enclosed by an outer tube and end discs which complete the magnetic circuit. Finally there is the toroidal type of core (Fig. 1F) with an evenly distributed winding which gives a coil with the minimum of external field, but is expensive to produce owing to the difficulties of winding.



(a)

(b)

FIG. 5.

A simple method of producing an inductance variation of the order of ±5 per cent. is desirable in order that slight variations either in winding or in the permeability of the material may be compensated and coils matched to a predetermined value ; also in the case of litz wound coils adjustment by altering the number of turns is inconvenient owing to the difficulty of cleaning the stranded wire, and for this reason inductance adjustment through the medium of the iron core is of particular advantage where high Q coils are concerned. The possibility of coil matching *in situ* after the apparatus has been assembled is also a very desirable feature. For the usual requirements the coil is wound to approximately the value required and a final adjustment is obtained by altering the reluctance of the magnetic circuit, the method of variation depending on the form of core employed. In the case of an open core, adjustment is affected by a movement of the relative positions of the coil and core, maximum inductance being obtained with the core central ; with a

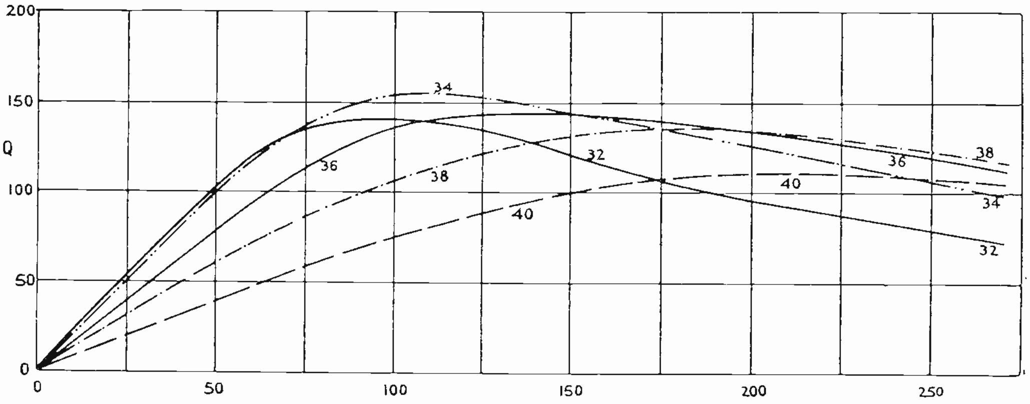


FIG. 6. *Q/f for solid wire (O.W.S.) on standard Ferrocart core I.C.* 2-4 *slot formers.* 400 *turns total.* 4.2 *millihenries. Unscreened.*

laminated rectangular core built up of L-sections, adjustment over a range of ±5 per cent. may be obtained by sliding the interlaced sections so as to alter the length of the magnetic path. For a non-laminated core consisting of either a pair of L-sections or an E and an I, the final setting is obtained by varying the air gap between the two sections before sealing. In the pot type of core a variation of inductance is obtained by an alteration in the air gap between the centre core and one disc, this alteration being effected by means of a screw mechanism in which the disc is attached to a screw

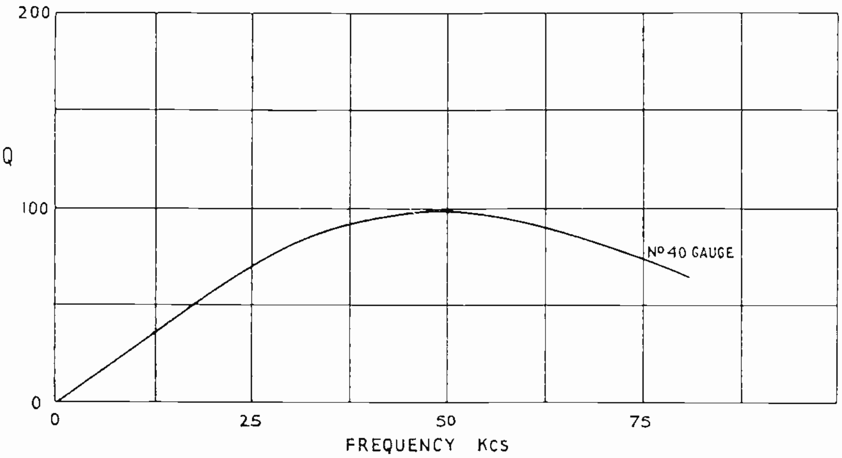


FIG. 7. *Q/f for solid wire (O.W.S.) on standard Ferrocart core I.C. Astatic.* 2-4 *slot formers.* 2,400 *turns total.* 115 *millihenries approx. Unscreened.*

operated through an insulated holder as shewn in Fig. 3. In an alternative type the pot is moulded in two identical sections, the centre core being hollow and cut away as shewn in Fig. 1G; adjustment in this case is obtained by a rotation of one

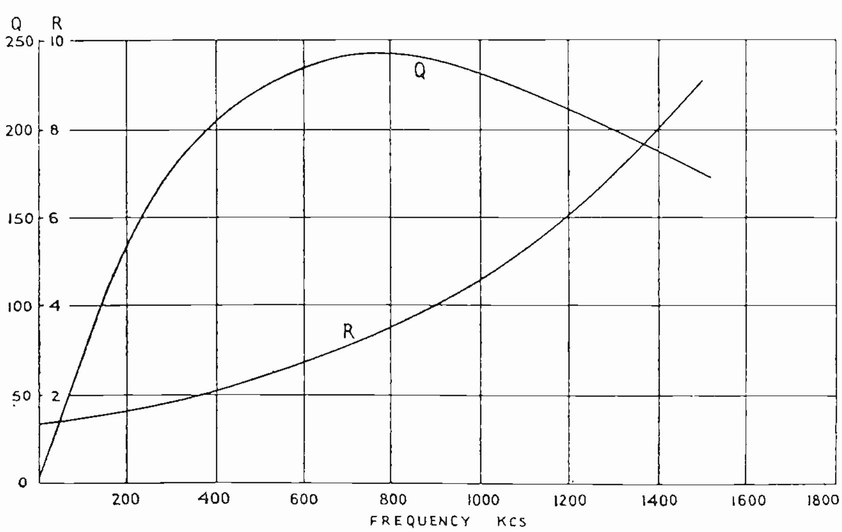


FIG. 8. *Q/f and R/f for* 60 *turns* 30/48 *Litz on Ferrocart pot core (small). Inductance set to* 170×10^{-6} *Hys. Unscreened.*

section relative to the other and by this means the effective air gap is varied and a fine adjustment of about 15 per cent. obtained. In cases where a cement is used either for holding the coils in position or for sealing a core adjustment it is advisable that a cement be employed in which the solvent does not affect the core material; cements containing amyl-acetate or acetone should be avoided.

The curves shewn in Fig. 4 illustrate the effect produced by allowing the magnetic flux to pass across the layering of Ferrocart material. Curve A relates to a pot built up as shewn in Fig. 5A so that the flux follows the direction of the layering

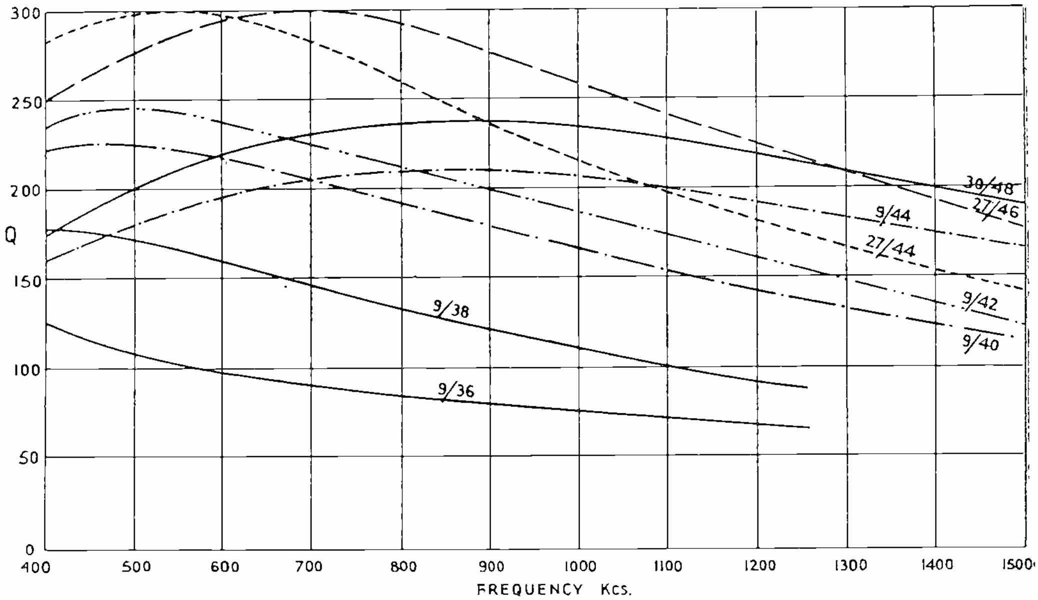


FIG. 9. *Q/f for Litz on standard Ferrocart core I.C.* 2-4 *slot formers.* 80 *turns total.* 160 *microhenries approx. Astatic. Unscreened.*

(as indicated by the arrows) throughout the magnetic circuit, whereas Curve B relates to a pot built up as shewn in Fig. 4B; and in this case it will be seen that the flux passes along the layering in the discs forming the top and bottom of the pot, but in a direction normal to the layering in the centre core and in the outer tube. With identical windings the second construction shews a drop in inductance of some 25 per cent. and a reduction in efficiency of the same order.

The generally accepted measure of the relative efficiency of a coil at a given frequency is the ratio of reactance to resistance $\frac{wL}{R}$ commonly denoted by the symbol Q; R being the equivalent series resistance of the coil at the frequency concerned and not the resistance as measured with direct current. To obtain the maximum advantage from the use of iron-dust cores careful attention must be given to the winding design, this including the design of the bobbins or formers used. It should be noted that, whereas air cored coils are usually either of single layer construction or wound so as to minimise self-capacity (e.g. by honeycomb or basket winding), iron-dust cored coils lend themselves to random pile winding and it is therefore desirable to subdivide the winding into several sections by means of slotted formers, with the object of reducing the self-capacity of the coil as far as possible. Important factors in the design of any coil are the gauge and stranding of the wire to be employed; the presence of the iron core causes a modification of the flux distribution, and for this reason the considerations governing the choice of gauge and stranding are rather different in the case of iron-dust cored coils from those obtaining with air cored coils, and the optimum wire cannot safely be predicted by the use of Butterworth's classic formulæ. It would appear that in the

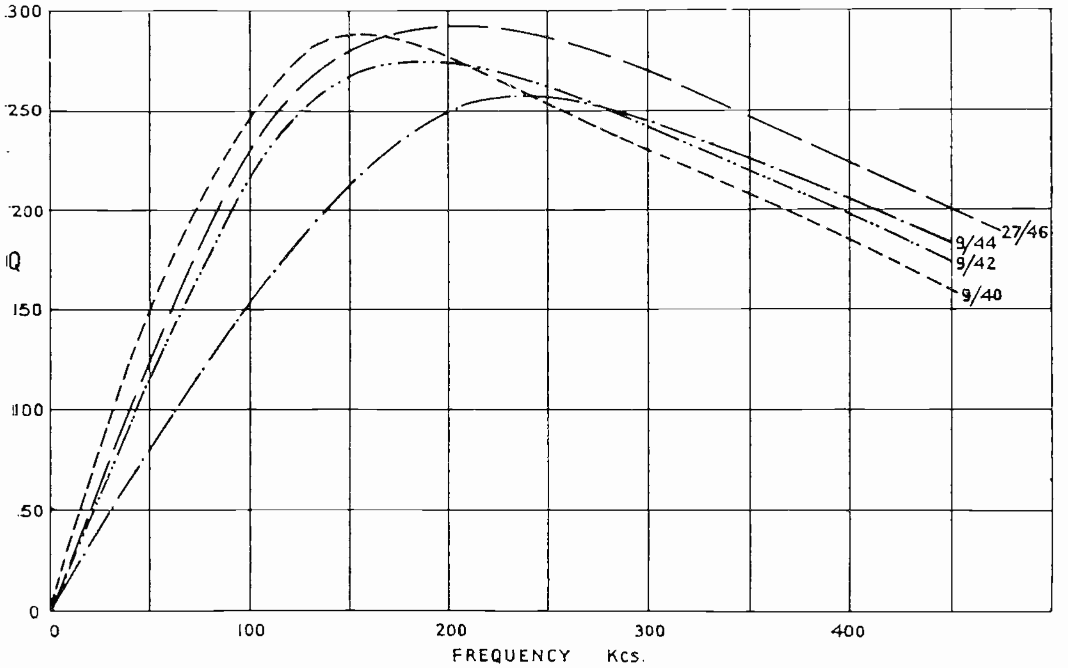


FIG. 10. *Q/f for Litz on standard Ferrocart core I.C.* 2-4 *slot formers.* 200 *turns total.* 1,100 *microhenries approx. Astatic. Unscreened.*

case of an iron cored coil a wire employing more strands of finer gauge than would be utilised for a corresponding air cored coil is necessary to obtain optimum results.

Owing to the concentration of the electro-static field on account of the reduction in size made possible by the use of iron cores, the material used in the construction of the bobbins or formers becomes of considerable importance and must be chosen with due regard for its permittivity and power factor at radio frequencies, since the dielectric losses are proportional to the product of these two quantities, while a high permittivity is further undesirable in that it tends to increase the self-capacity of the coil.

While the use of litzendraht is necessary for the highest Q values which are obtainable with iron cored coils at broadcast frequencies, solid wire is suitable for the production of coils of reasonable losses and having the advantage of small overall dimensions; and by the careful choice of the gauge of wire the Q can be made as high as is usually necessary, particularly for the lower frequencies, say below 300 kc.

Solid wire iron cored coils can be regarded as intermediate between iron cored litz. wound coils and air cored coils both as regards expense and performance, and while little superior to well designed air cored coils, their chief advantage lies in compactness and restricted magnetic field with consequent low screening losses, which enable small clearances to be employed between the coil and its screening can, a factor which becomes of considerable importance in multi-band assemblies involving

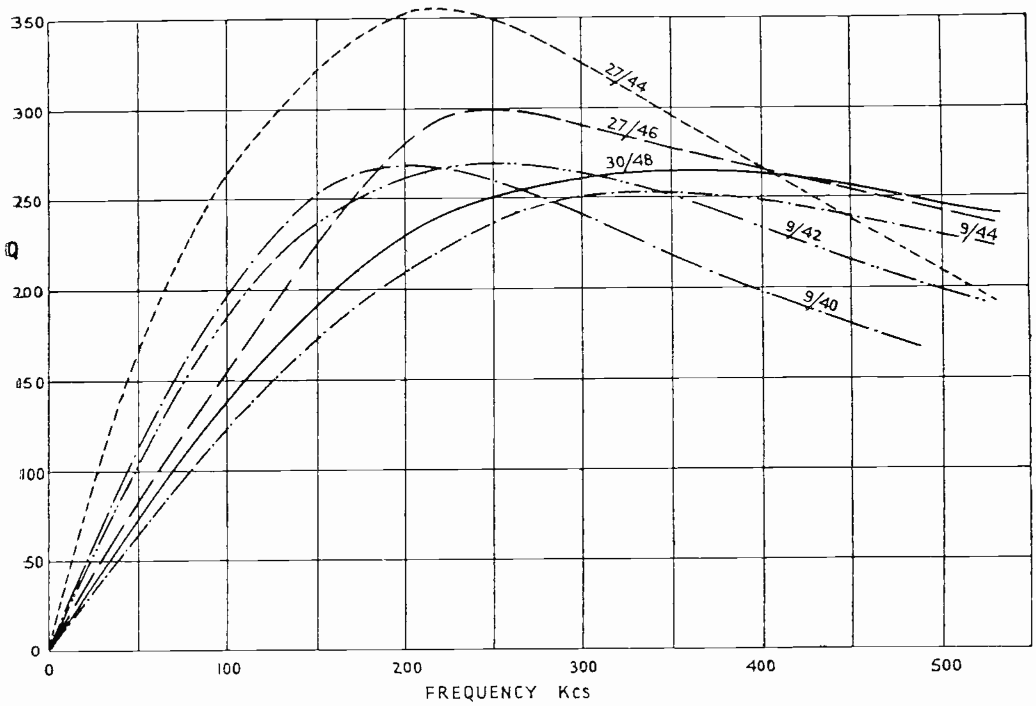


FIG. 11. *Q/f for Litz on standard Ferrocart core I.C. 2-4 slot formers. 280 turns total. 2,400 microhenries approx. Astatic.*

a large number of coils. Small coils having, inherently, restricted fields it is possible to use these coils without screening cans on occasion without recourse to the wide spacing which would be needed with air cored coils.

Reference to Figs. 6 and 7 shews that in the case of an astatic coil wound on laminated interlaced L-sections, No. 36 gauge copper wire maintains a consistently high Q value over a frequency band of 400—1,600 kc. when used in a coil of some 4 milli-henries and when these cores are used No. 36 copper wire is found to be the best general purpose gauge. Using the same core and type of winding an inductance of over 100 milli-henries can be wound with 40 gauge copper wire and a Q of 100 at 50 kc. obtained. It will be noted that while the general shape of the curve Q/f is similar in every case the peak value tends to occur at a lower frequency the higher the inductance value of the coil and the larger the gauge of wire with which it is wound.

The general relationship between the high frequency resistance R and the Q value is clearly shewn in Fig. 8. Since $Q = \frac{2 \pi f L}{R}$ and L is constant, Q will be proportional to f if R is constant, which is nearly true at the low frequencies; as the frequency rises the resistance increases more and more rapidly and the Q curve consequently bends over, passing through a maximum (which occurs at a frequency

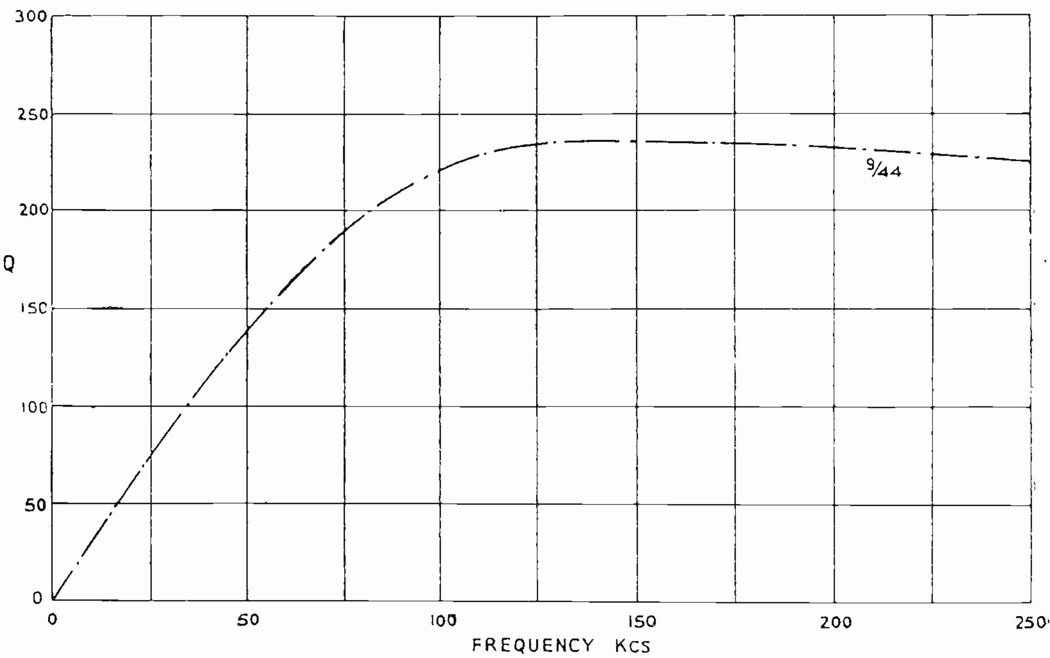


FIG. 12. *Q/f for Litz standard Ferrocart coil I.C. 2-4 slot formers. 560 turns total.* 6,800 *microhenries approx. Unscreened. Astatic.*

depending partly on the inductance and partly on the wire gauge as already noted) and actually falling as the frequency is further increased.

Figs. 9-12 shew a series of curves for coils of similar construction to those to which Figs. 5 and 6 relate, but wound with stranded wire of various gauges as detailed on the curves. It will be noted that the Q values are of a much higher order, approaching double those of the solid wire coils.

A point of interest which is brought out in several of the curves is that an increase in the number of strands, keeping the gauge of the strands constant, does not necessarily result in an increased Q value, but may cause an appreciable decrease at the higher frequencies as is shewn in Figs. 9 and 10 where 9/44 can be compared with 27/44. A remarkable similarity in the general shape of the curves for 9/44 and 30/48 wire is shewn and the slightly higher Q values given by 30/48 wire do not always justify its general use in preference to the cheaper and more robust 9/44 gauge wire. A feature of both these wires is the constancy of the Q value over the major portion of the frequency band. The very high Q values obtainable with

27/44 wire at the lower frequencies favour its use for special purposes at those frequencies, but it is hardly suitable for use over a wide frequency range on account of the rapid variation of Q with frequency.

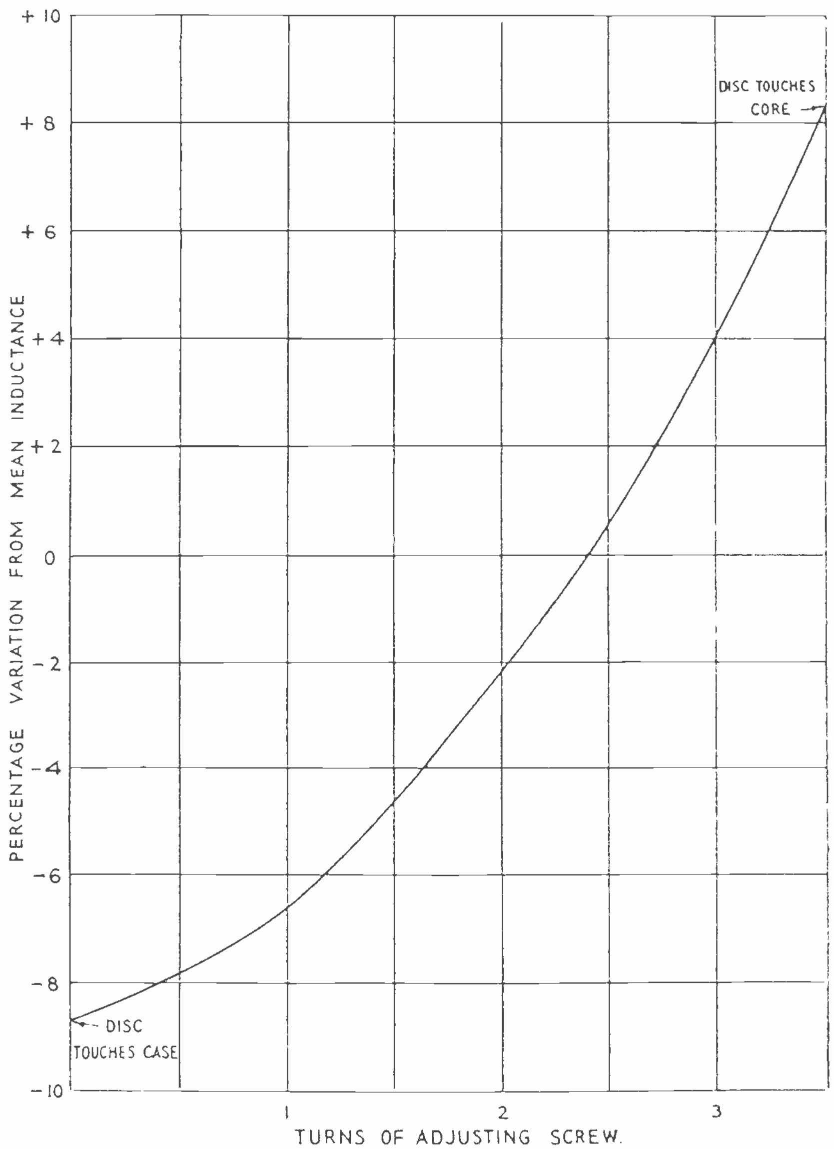


FIG. 13. *Inductance adjustment of "pot" type Ferromould core.*

Fig. 13 shews the variation of inductance of a coil wound on a small core of the pot type shewn in Fig. 1E; a fine adjustment is obtainable since about 3 complete turns of the screw are required to cover the range of variation available.

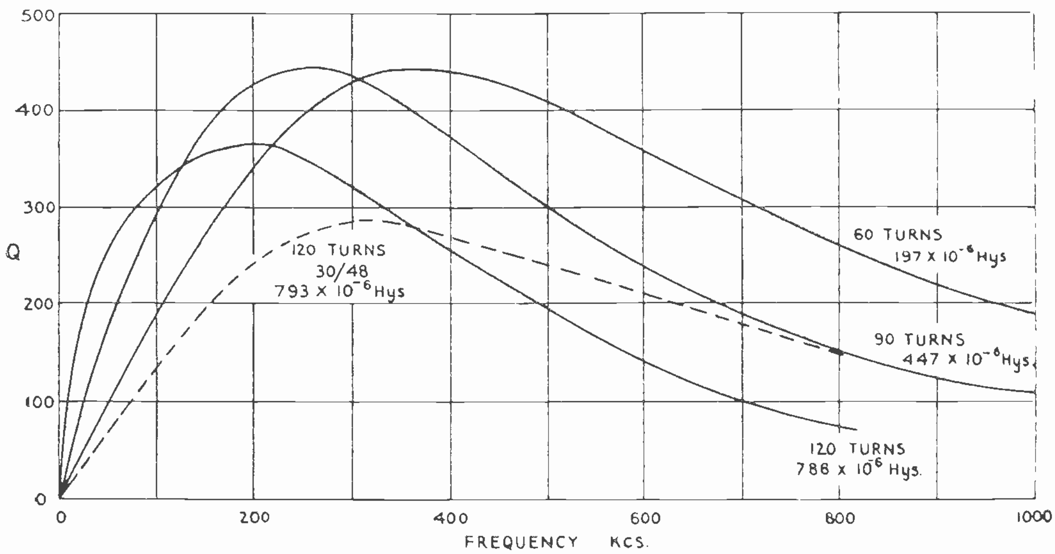


FIG. 14. *Q/f for 27/44 and 30/48 Litz on Ferrocart pot (medium).*

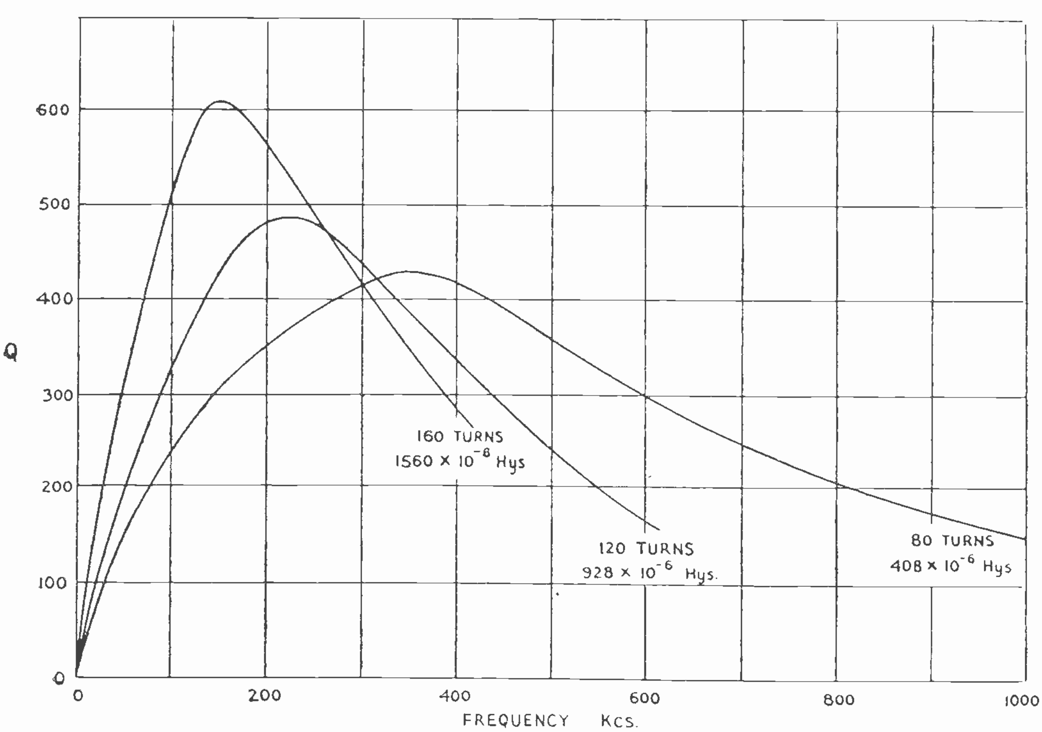


FIG. 15. *Q/f for 27/44 and 30/48 Litz on Ferrocart pot (large).*

The use of the pot type of core assists in the production of coils with very high Q values, particularly at the lower frequencies, but there is a general tendency for the Q value to vary more rapidly with frequency than is the case with coils wound on cores of the types previously considered, as will be apparent by reference to Figs. 14 and 15 which shews that 27/44 gauge wire gives a very high performance.

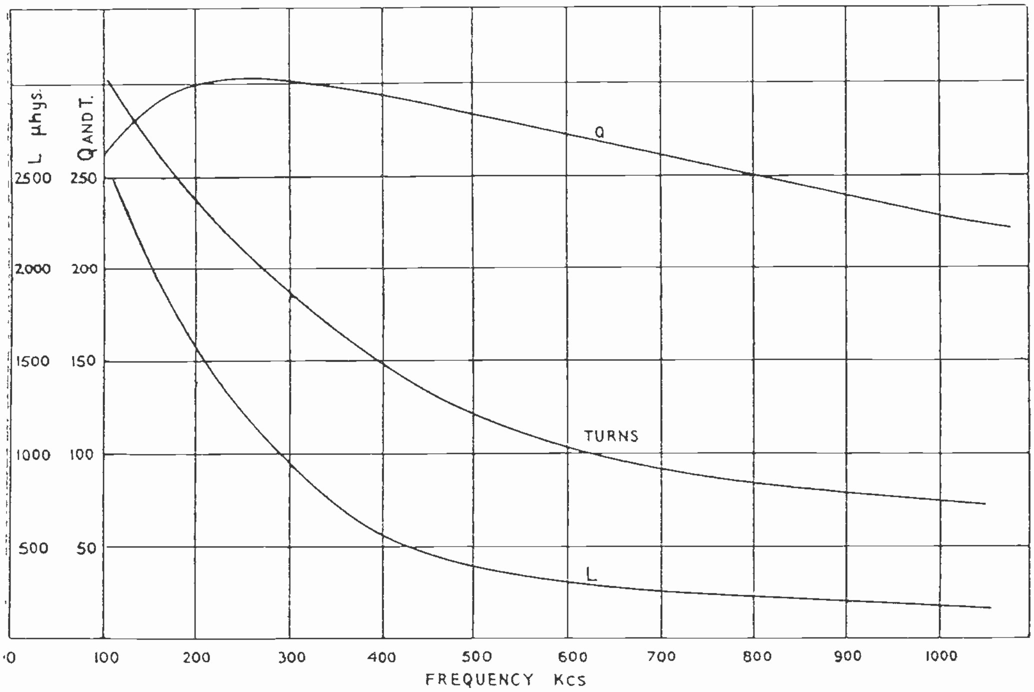


FIG. 16. *Curves for 27/46 Litz wire. Laminated L-section core I.C. Astatic windings. Optimum Q.*

Fig. 16 shews details of coils wound with 27/46 wire on laminated, interlaced L-section core of the type shewn in Fig. 1c. For any given frequency a curve can be obtained relating Q and L; this curve will shew a peak value of Q, being the highest obtainable Q at that frequency; similar curves at other frequencies can be utilised to obtain these optimum Q values and their corresponding inductance values. These abstracted values for astatic coils are shewn in Fig. 16 and it will be seen that over the frequency range covered by the measurements the optimum values of inductance approximate to those normally employed, the highest possible Q values with this type of core and wire being obtainable at approximately 250 kc.

The effect of screening on iron-dust cored coils is dependent on the form of the core used, as well as the closeness of the screen, those types of core with the largest stray fields being most affected by the presence of the screening can, both inductance and Q values being reduced. In the case of pot type cored coils and those wound astatically on L-sections, where the external field is comparatively small, the reduction is not very serious even with a close fitting screen, being of the order of 5 per cent. in inductance and 10 per cent. in Q; for example the following figures will serve to indicate the order of the effect :—

Core. Open type similar to Fig. 1A.			27 mm. × 8.5 mm. × 8.5 mm.
	Micro-henries.		Coil 25.4 × 25.4 × 16 mm.
No screen—L.	262	Q. 160	
Large screen	235	147.6	Screen 90 × 70 mm. diameter.
Small screen	210	135.8	Screen 50 × 40 mm. diameter.
Core. Closed type similar to Fig. 1C.			34 × 28 × 4.5 mm.
	Micro-henries.		Coil 25.4 × 25.4 × 10 mm.
No screen—L.	120	Q. 156.9	
Large screen	117	153	Size as above.
Small screen	115	140	Size as above.
Core. Pot type similar to Fig. 1E.			18 × 21 mm. diameter.
	Micro-henries.		
No screen—L.	290	Q. 273	
Large screen	283	253	63.5 × 63.5 × 44 mm.
Small screen	276	235	42 × 42 × 54 mm.

The material of which the screen is constructed affects the screening loss to a marked degree, and while non-magnetic metals such as copper and aluminium are suitable, it is found that the presence of iron close to the coil introduces serious losses unless a non-magnetic screen is interposed. The effect of allowing an earth contact with the iron-dust core is shewn by the following figures which relate to an astatic pair of coils of 48 turns each wound with 27/44 Litz wire on a core similar to that shewn in Fig. 1C, the measurements being taken at a frequency of 1,000 kc.

Normal—L. (micro-henries)	222	Q. 155
Core earthed with a single wire contact	222	90

Coils utilising powdered iron cores shew little variation of their electrical properties with variation of temperature. The early Ferrocart shewed a slight rise in permeability for the first temperature cycle, after which the permeability variation became cyclic. In the case of another material no such ageing effect was present and the temperature coefficient of permeability was negative. To indicate the order of the effect to be expected the following figures may be quoted, these including the variation due to the coil former :—

°C. At 200 kc.	L. (micro-henries).
17.8	1,960
60	1,940
18.9	1,960

and serve to shew that the inductance change over the temperature range likely to be encountered in practice may be taken as negligible for most purposes. It should, however, be noted that the use of rubber and low melting-point wax for adjusting and locking purposes is not to be recommended for tropical use.

With reference to level, Polydoroff has stated that with the material he originally developed there was no change in permeability with a variation of flux density of

.01 to 10 Gauss, and tests of Ferrocart over a similar range shew a similar constancy of permeability. It is not be to expected that the maximum figure would be approached in ordinary receiver practice.

Receiver design has been influenced by the introduction of iron-dust cored coils and the combination of compactness and high efficiency found in the modern receiver would hardly be possible without the use of this type of coil. Multi-band switch assemblies have been made commercially possible, although it is not advisable to wind coils for more than one range on the same core, and the use of iron cored coils for the higher frequencies, above 2 megacycles, may only be justified on the score of compactness and ease of adjustment rather than on efficiency. It should be emphasised that the components associated with a coil in a circuit, such as the valve holder and tuning condenser, should be chosen with due regard to the overall efficiency required; for example, it would usually be uneconomic to employ an ultra-low loss condenser for tuning purposes in conjunction with a compact medium efficiency coil; but where a high Q tuned circuit is required it is essential to employ, with the high efficiency coils, components in which all sources of loss are minimised.

C. AUSTIN.
A. L. OLIVER.

MARCONI-T.C.M. HIGH FREQUENCY CABLES

Since the introduction of the concentric feeder by the Marconi Company in 1926 for use in their short wave beam system, the use of this type of transmission line has become widespread, not only in short wave radio installations, but more latterly in connection with wired carrier-current transmission for multi-channel telegraph and telephone working.

In this article, prepared by the Company's Research Department, in collaboration with Dr. E. W. Smith, of Messrs. Telegraph Construction & Maintenance Co., Ltd., details are given of a continuous feeder, with a lead outer conductor and a minimum of supporting dielectric between inner and outer conductors. An outline is then presented of some methods of measurement of H.F. cables and of determining separately the losses due to conductors and to dielectric. Finally some notes are given on the installation of this type of cable.

DURING the past few years considerable interest has been shown in cables in which air is used as the greater part of the dielectric, and great ingenuity has been exercised in their design.

Dry air is an ideal dielectric where high breakdown strength is not of paramount importance, and in the design of cables having very small dielectric loss its use is inevitable. The best form of solid insulation known gives transmission losses at least 50 per cent. higher.

The recent rapid development of the air dielectric cable is due to the extension of the frequency range of cable transmission beyond carrier frequencies of 100 kilocycles or so up to that required for television, i.e., at least one megacycle. In providing for this extension by the means to be described, it became evident that a properly designed cable need not be limited in frequency range to even a megacycle, but could be used, in lengths of several miles, for frequencies as high as 50 megacycles.

At the same time, air-dielectric types of cable were developed for auxiliary wiring, so that in all cases where high frequency energy is confined to wired circuits, a special air-dielectric wire is now available to take the place of the less efficient solidly insulated conductor.

It is of interest first to classify the three chief types of high frequency cable according to the means adopted for keeping the conductors separated in the surrounding air. These are

(1) By insulating discs disposed laterally on the conductors, spaced at regular intervals along the length of the cable.

(2) By open helices of insulating material wound round the conductor or conductors.

(3) By insulating tubes in which the conductor is centred either by regularly spaced lateral constrictions or longitudinal ribs in the tube, or by kinks in the conductor itself.

In general, the first type has the advantage of requiring the least amount of solid dielectric.

Description of the Cable.

The cable about to be described is manufactured by the Telegraph Construction & Maintenance Co. Ltd., of Greenwich, S.E.10, in association with Marconi's Wireless Telegraph Co. Ltd.

It is a coaxial cable of the first type, and has a solid copper wire as centre conductor, spaced within a concentric tube of lead as outer conductor.

The spacing discs are moulded of the synthetic resin known as trolitul. This material has a dielectric constant of about 2.5 and a power factor of the order of 4×10^{-4}, these figures being by far the lowest now obtainable with this class of material. Trolitul also possesses the important advantage of being non-hygroscopic, and extended observations have shown that its electrical and mechanical properties have no tendency to change with time.

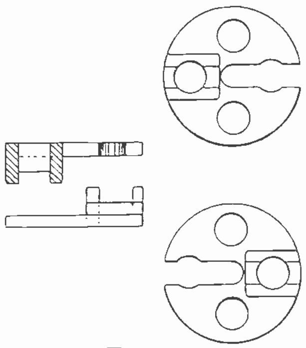


FIG. 1.

The discs are shown in Fig. 1.

They are designed for assembling in pairs, each having a slot to pass the conductor, and a raised rib of the same dimensions as the slot, along the opposite radius thereto. In the assembly the pairs are cemented together and to the conductor, so that the rib of one disc fits into, and closes, the slot of the other. A light and rigid spacing unit, consisting of the minimum amount of a high quality dielectric, is thus formed.

The discs are assembled on the conductor with a spacing between pairs of normally 3 inches. Continuous lengths up to 500 metres are readily accommodated on suitable drums, and after the assembly is complete, the outer conductor, which consists of a lead tube some 80 mils in radial thickness, is directly extruded over the discs from a lead press. The melting point of the trolitul discs is sufficiently high to prevent softening during the short time that they are under the plastic lead, and special blocks of reinforced discs are assembled on the conductor at predetermined and marked distances to enable the lead press to be stopped and re-charged as required. There is thus no practical limitation to the length of cable which can be sheathed in one operation, although it has been found convenient in manufacture and in subsequent handling to limit the length to about 500 metres.

While for a given diameter of cable the minimum attenuation is obtainable with a copper outer conductor, there are several advantages in using lead, particularly on account of ease of manufacture. If copper were used, a composite conductor, e.g., a number of tapes longitudinally applied, would be essential to preserve flexibility, and a waterproof covering would have to be added. A study of the economics of construction has shown that the present design, in which the lead sheath combines the functions of return conductor and waterproof covering, costs less for a given attenuation than a cable in which both conductors are of copper. It is important to note that the lead acts as a very effective screen against high frequency interference from external sources.

A section of cable after the lead sheathing process, with the sheath cut away to show the centre conductor and spacing system, is shown in Fig. 2.

This "core" is then normally provided with a double steel tape armouring, which with jute servings and compound forms a protected cable which can be buried directly in trenches.

The illustrations above are of the cable known as A.S.26—this being the most commonly used size. Another smaller size, A.S.33, is manufactured in a similar manner. The two cables are designed to have the same impedance, so that they may be interconnected without reflection losses. The larger size, A.S.26, armoured and laid in trenches, is normally used as the main cable, while the smaller size, A.S.33, is normally used unarmoured for internal distribution from the main cable.

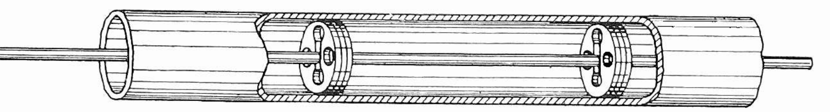


FIG. 2.

Dimensions and Electrical Data.

The table below gives details of the size and attenuation of the above cables :—

	A.S.26	A.S.33
Diameter of inner conductor.. ..	0.127 in.	0.064 in.
Inner diam. of outer ,, ...	0.75 in.	0.375 in.
Outer ,, ,, ,, ,, ...	0.91 in.	0.52 in.
Approx. characteristic impedance, ohms	100	100
Max. Attenuation, db. per mile at 2 Mc.	4.5	9.5
,, ,, ,, ,, ,, 10 ,,	10	23
,, ,, ,, ,, ,, 20 ,,	14	33
,, ,, ,, ,, ,, 40 ,,	20	47

These figures for attenuation are slightly higher than those obtained on measurements of a sample length, as described later in this article, and allow a small margin for manufacturing tolerance.

It has been stated that the ideal of an all-air dielectric has been closely approached, and this will be seen to be the case on consideration of the capacity and power factor obtained.

The average capacity of a large number of manufactured lengths of A.S.26 cable is 0.0540 μF. per mile (0.0335 μF. per km.). The highest capacity obtained on separate lengths was 0.0543, and the lowest 0.0538. Corresponding figures for A.S.33 cable are average 0.0541, highest 0.0544, lowest 0.0539. The average figure 0.0540 for A.S.26 cable corresponds to a dielectric constant of 1.07. The measured power factor of the cable has never exceeded 5×10^{-4}, and in some cases values of less than half this figure have been obtained. The insulation resistance averages 10,000 megohm-miles.

Although both sizes of the cable were primarily designed as receiving feeders, they are capable of being used for low power transmission. For example, the larger, A.S.26, has been found capable of transmitting a current of 8 amps at 21 megacycles.

Testing of H.F. Cables.

If maximum transmission efficiency is to be realised in any communication system utilising cables, two important features of the cable should be known, viz., its characteristic impedance and its attenuation, over the frequency range required.

Whilst on a test length of cable these characteristics can be determined at high frequencies by certain of the methods normally used in connection with low frequency cables, others, not possible at low frequencies, are particularly suited to radio-frequency measurements.

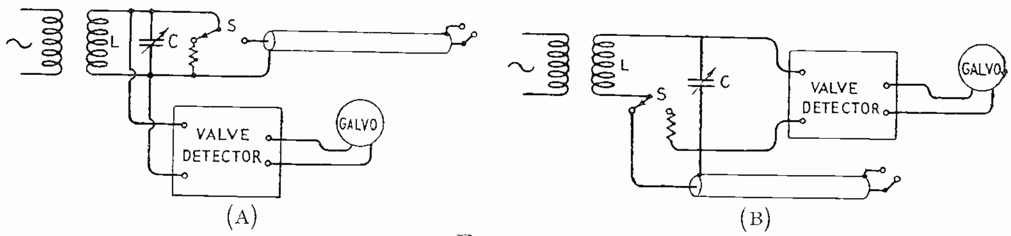


FIG. 3.

Some of the methods employed in the measurement of Marconi-T.C.M. H.F. cables will be considered here.

(I) RESONANCE METHOD.

This method depends upon the well-known property of a transmission line, namely, that when open-circuited or short circuited at the far end, it becomes resonant at frequencies corresponding to which the line length is an integral number of quarter waves. At these frequencies it will appear as a purely resistive load to the generator.

For any length of cable :—

Let : Z_o = Characteristic Impedance of Cable (ohms).
α = Attenuation in Nepers.
β = Phase shift in radians.
Z_s = Input impedance when the far end is short circuited (ohms).
Z_f = Input impedance when the far end is open circuited (ohms).
R = Total Conductor Resistance (ohms).
L = Inductance (Henrys).
C = Capacity (Farads).
G = Leakance (ohms).

The input resistance of a cable shorted at the far end is :—

$$Z_s = Z_o \tanh [\alpha + j\beta]$$

At all frequencies for which $\beta = n\frac{\pi}{2}$, (where n is any integer) this becomes

$$Z_s = Z_o \tanh \left[\alpha + jn\frac{\pi}{2}\right]$$

$$= \begin{cases} Z_o \coth \alpha \ldots\ldots\ldots \text{for odd values of } n & \text{(ia)} \\ Z_o \tanh \alpha \ldots\ldots\ldots \text{,, even ,, ,, } n & \text{(ib)} \end{cases}$$

When the cable is open circuited at the far end the input resistance is given by :—

$$Z_f = Z_o \coth [\alpha + j\beta]$$

and when $\beta = n\frac{\pi}{2}$,

$$Z_f = Z_o \coth\left[\alpha + jn\frac{\pi}{2}\right]$$

$$= \begin{cases} Z_o \tanh \alpha \text{ for odd values of } n & \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \text{(iia)} \\ Z_o \coth \alpha \text{ ,, even ,, ,, } n & \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \text{(iib)} \end{cases}$$

Dividing (iia) by (ia):

$$\frac{Z_f}{Z_s} = \frac{Z_o \tanh \alpha}{Z_o \coth \alpha} = \tanh^2 \alpha.$$

$$\therefore \qquad \alpha = \tanh^{-1}\sqrt{\frac{Z_f}{Z_s}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \text{(iii)}$$

for odd values of n.

Dividing (ib) by (iib):

$$\frac{Z_s}{Z_f} = \frac{Z_o \tanh \alpha}{Z_o \coth \alpha} = \tanh^2 \alpha$$

giving: $$\alpha = \tanh^{-1}\sqrt{\frac{Z_s}{Z_f}} \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad \text{(iv)}$$

for even values of n.

Multiplying (ia) by (iia) or (ib) by (iib) gives in each case

$$Z_s \,.\, Z_f = [Z_c \tanh \alpha]\,[Z_o \coth \alpha], \text{ so that } Z_o = \sqrt{Z_s \,.\, Z_f} \quad \ldots \quad \ldots \quad \text{(v)}$$

in each case.

The attenuation is thus given at odd quarter-wave frequencies by expression (iii) and at even quarter-wave frequencies by (iv). It is to be understood, of course, that in (iii) and in (iv), Z_f and Z_s are corresponding values for the same value of n. The purely resistive nature of Z_s and Z_f at integral quarter-wave frequencies is shown in expression (i) and (ii). Also, since coth α and tanh α are reciprocal, Z_s and Z_f assume alternately high and low values with respect to Z_o, for successive values of n. For odd values of n, Z_s will be high and Z_f correspondingly low whilst for even values of n their relative magnitudes are reversed.

This resonant behaviour of cables at certain frequencies offers at once a simple means of determining Z_f and Z_s at each of these frequencies, by a resistance substitution method, and reference to Fig. 3 will make clear the underlying principle.

An inductance L and variable condenser C form a tuned circuit which is connected to the input of a sensitive valve-detector and loosely coupled to a variable frequency oscillator. The cable is connected in parallel or in series with this tuned circuit, depending respectively upon whether the input impedance of the cable is high or low.

It has been shown that this input impedance can be high, or correspondingly low with respect to Z_o, so that it is in either case comparable to the dynamic resistance of the tuned circuit to which it is connected. By means of a switch S the cable can be replaced in the circuit by a variable known resistance.

The procedure in making measurements is first to find the lowest (1st quarter wave) frequency at which the addition of the cable to the circuit causes no detuning of the circuit. Under this condition the cable is introducing no reactance into the circuit. This frequency in megacycles will be approximately equal to $246/l$ where l is the length of the cable in feet.

The addition of the cable will, however, introduce resistance into the circuit and thereby decrease the voltage across C. This will be shown by a change of galvanometer reading. The value of this resistance can be found by substituting one of known value, by the switch S, to give the same galvanometer reading as with the cable in circuit.

The comparison resistances can be of the "$\frac{1}{2}$-watt" metallized type, provided their H.F. resistance is known. At frequencies greater than about 20 megacycles per second, resistances of 10,000 ohms or higher may show considerable discrepancy between their effective high frequency value and their D.C. value. By paralleling and by graphical interpolation, the resistance of the cable can be closely matched.

Having determined Z_f and Z_s as above, at the first quarter wave frequency, further measurement is made at each harmonic frequency, where the cable will exhibit alternate quarter- and half-wave resonance effects. Substitution in the appropriate formula (iii) or (iv) of values of Z_f and Z_s for each frequency will then give α, and Z_o can be found from formula (v).

The resonant conditions of the feeder (as defined in terms of multiple quarter waves), with the far end open and shorted, should, in a perfectly uniform cable, occur at precisely the same frequency. Any irregularity in the cable, such as uneven spacing between inner and outer, will result in open and short circuit resonance occurring at different frequencies since they introduce additional points of reflection in the cable, and the disparity between the two frequencies is a measure of the extent of the irregularity. However, unless such irregularity is of a serious nature this disparity will be very small, and within the limits of good manufacturing tolerance, should be less than about 0.3 per cent.

It has been mentioned in an earlier paragraph that the first resonant frequency in Mcs. is given approximately by $f_{mc} = \frac{246}{l}$ where l is in feet which assumes a transmission velocity equal to that of light. The actual resonant frequency will, however, be found to be lower than this, since the presence of solid dielectric in the cable increases its capacity and therefore increases the phase-shift per unit length. The "velocity ratio" is then given directly as $\frac{f_o}{f_{mc}}$ where f_o is the observed resonant frequency. The velocity ratio of the Marconi-T.C.M. cable is found to be .97 for A.S.26 and .96 for A.S.33.

(2) Voltage Ratio Method.

In an open circuited cable, the ratio of the voltages V_1 at the input and V_2 at the output is, by general line theory:

$$\frac{V_1}{V_2} = \cosh\,[\alpha + j\beta]$$

so that if the length of the cable is known, the angle β at the frequency of measurement can be calculated, applying the necessary correction for velocity ratio, and α can then be found.

Fig. 4 illustrates this method. Diode voltmeters are used, the static diode current being balanced out by a local battery and potentiometer P.

At the odd quarter wave resonant frequencies of the cable,

$$\beta = \left[2n + 1\right]\frac{\pi}{2} \qquad \text{and } \frac{V_1}{V_2} = \cosh\left[\alpha + j\,(2n+1)\frac{\pi}{2}\right] = \pm j \sinh \alpha$$

so that the modulus $\left|\frac{V_1}{V_2}\right| = \sinh \alpha$ .. .. (vi)

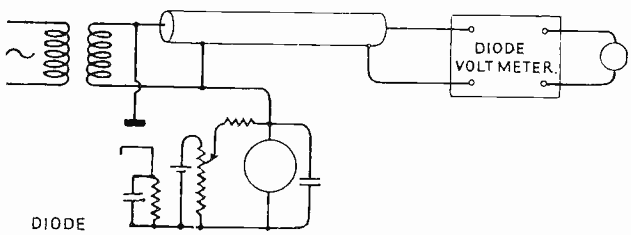


FIG. 4.

Owing to the resonant voltage build-up at the far end of the cable, $\frac{V_1}{V_2}$ will be small, and if it is less than 0.1, then α is given very closely by the ratio $\frac{V_1}{V_2}$.

If the input current to the cable is also measured in the resonant condition, the characteristic impedance can also be found directly. For multiplying (iii) and (v) gives

$$\alpha Z_0 = Z_f = \frac{V_1}{I_1} \text{ where } I_1 \text{ is the input current}$$

so that $$Z_0 = \frac{V_1}{I_1} \cdot \frac{1}{\alpha}$$

and when α is small enough to be replaced by $\frac{V_1}{V_2}$, then :

$$Z_0 = \frac{V_2}{I_1}$$

(3) CURRENT RATIO METHOD.

So far the measuring technique has stressed the application of resonance effects, although the voltage ratio method is not wholly limited thereto. In considering methods involving current measurement, it would appear feasible also to apply similar principles, for example, to measure the input and output current of a shorted cable. Whilst theoretically this is possible, experience has shown that considerable inaccuracy may result in practice, due to the insertion of an instrument at the supposedly short-circuited end. It is essential at high frequencies that the short circuit be as effective as possible.

If the cable is correctly terminated, however, by its characteristic impedance, the ratio of input to output current gives at once a reliable attenuation factor, which moreover can be obtained at any frequency. It is important that the termination, which includes the resistance of the instrument, be correct, otherwise the slight reflections due to mistermination will cause the measured attenuation to alternate the frequency about the correct value. If the termination is correct and I_1 and I_2 are respectively the input and output currents, then

$$\frac{I_1}{I_2} = e^{\alpha + j\beta} = e^{\alpha}\underline{\lfloor \beta}$$

so that the attenuation α, in nepers is given by :

$$\alpha = \log_e \left| \frac{I_1}{I_2} \right|.$$

Separation of Losses.

The methods outlined above admit of a direct measurement of the total cable attenuation and were found to be in good agreement.

However, with the object of comparing conductor and dielectric losses, separate measurements of each were carried out on the A.S.26 type cable. These were intended to supplement, rather than to check, previous results, and since they necessarily involved some degree of assumption in analysis and in the methods adopted, exact agreement with more simple measurements was not expected. Nevertheless, summation of the losses showed a disparity of only about 10 per cent. at 10 megacycles.

Here again the basis of measurement was that the length of cable should form part of a tuned circuit, but in this case resonance effects were deliberately avoided, and the test specimen was made short enough to be well below resonance whilst at the same time showing a measurable effect on the tuned circuit.

A length of 10 feet was chosen as effecting this compromise, and with this length measurements were carried out at frequencies up to 12 megacycles per second.

Conductor Resistance.

The test length, with inner and outer conductors joined at one end by a substantial copper plug, is placed in series with a shunted thermocouple in a tuned circuit of low L/C ratio. This circuit is excited through a neutralised 25 watt amplifier. The inductance and all connections are made up of thick copper strip to keep the circuit resistance to a minimum. The current reading is noted. The test piece is then replaced by a " standard " piece of copper cable of similar dimensions, the H.F. resistance of which can be calculated from Russell's formula :

$$R = 41.2 \times 10^{-4} \left[\frac{1}{r_1} + \frac{1}{r_2} \right] \sqrt{f} \text{ ohms per km.}$$

where r_1 = Inner radius of outer conductor in cms.
r_2 = Outer ,, ,, inner ,, ,, ,,

The object in using a standard length is to keep constant the inductance of the circuit. Resistances of known value are then connected across the end of this standard and a curve is plotted connecting currents and resistances. From this curve and the known current reading of the test piece, its resistance can be found.

Dielectric Loss.

This can be found from measurements of the dynamic impedance of a tuned circuit with and without the test length. Since the power factor is very low, the tuned circuit, if it is to be affected to a measurable extent by the cable, must itself have a high dynamic impedance. This can be achieved by the use of a high grade condenser and of copper tube inductance of large diameter.

To ensure minimum disturbance of current distribution when the test piece is added, a standard length should be used in this case also. This standard must have minimum of supporting dielectric. The impedance of the circuit is first measured by means of a valve voltmeter of high input resistance with the standard across it, and then again with the test piece substituted for the standard. In this case, since the cable is measured as a condenser, the far end of the test length, and of the standard, is of course left open.

Since the addition of the test piece is equivalent to adding a parallel resistance to the circuit, the value of this resistance can then be found.

The dynamic impedance can be readily obtained by the well-known detuning method. The circuit is first tuned to resonance. The capacity change required to reduce the voltage across the tuned circuit to $\frac{1}{\sqrt{2}}$ of its resonance value on each side of resonance is then found. Let this change be denoted by $\triangle C$, then :

$$\text{Dynamic Impedance} = \frac{2}{\omega \triangle C}$$

This capacity change will be found in practice to be very small, of the order of one micro-microfarad, so that an accurately calibrated vernier condenser is necessary in parallel with the main tuning condenser.

If R_s is the value of cable shunt resistance found as above, and C the capacity of the cable, the power factor is given at once as $\frac{1}{\omega C R_s}$ at the frequency $\frac{\omega}{2\pi}$.

The power factor was measured as above at various frequencies up to 8 megacycles, and the results showed it to be independent of frequency, the mean value being 0.04 per cent.

Summation of Losses.

It can be shown that at high frequencies the attenuation of the cable can be very closely represented by the expression :

$$\alpha = \frac{R}{2Z_o} + \frac{GZ_o}{2} \quad \text{(nepers)}$$

where R is the total conductor loss, G is the leakance—equal to $\frac{1}{R_s}$—and Z_o the characteristic impedance. From values of R and G therefore as found above, α can be obtained by summation of the losses.

It was found that the attenuation figures derived in this way were less than those due to other methods. The proportion of total loss attributable to dielectric was, however, found to vary from about 5 per cent. at 3.5 megacycles to 8 per cent. at 10 megacycles.

These figures amply justify the claim made earlier in this article, that the ideal of an air spaced cable has been closely approached.

Fig. 5 shows the measured attenuation, in decibels per mile, against frequency, as obtained on test lengths of A.S.26 and A.S.33, from one megacycle to 100 megacycles per second. For attenuations above 15 megacycles per second, where due to experimental errors measurements are necessarily less accurate, the curves are shown as broken lines. Nevertheless, the results obtained by the various methods

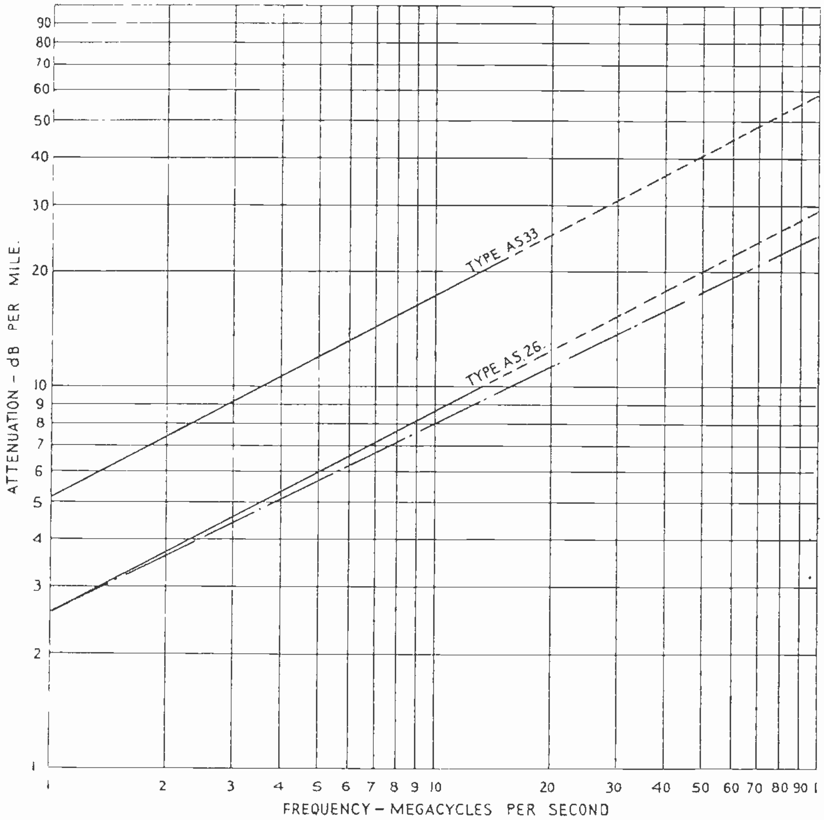


FIG. 5.

show that the curves do represent with fair accuracy the attenuation at the higher frequencies. The chain line curve is derived by deducting from the total attenuation the proportions as found above for dielectric loss, and shows therefore the loss due to conductors only. It will be seen that the slope of this curve very closely approaches the value of 0.5 which theoretical considerations would predict since the high frequency resistance of the conductors increases as the square root of frequency. The difference between the square root curve and the actual loss curve gives therefore the loss due to leakage.

It will be clear that in considering the merits of a high frequency cable, the slope of its attenuation-frequency characteristic affords a ready means of assessing the extent of the dielectric loss. Since this increases directly as the frequency, its contribution to total loss becomes greater as the frequency is increased.

Installation of Cables.

These cables can be installed without difficulty provided due care is exercised. It must be remembered that much of the mechanical strength possessed by ordinary solid dielectric cables has been necessarily sacrificed to obtain the desired high frequency characteristics, and that, unless protected, they do not present much more resistance to damage than ordinary lead tube. With the protection of steel tape armour, however, they will withstand all the handling necessary for outside installation, and may be buried directly in the ground.

The larger cable, A.S.26 steel tape armoured, may be bent to radii of 2 feet, while the smaller, A.S.33, may be bent to radii of 9 inches when used unarmoured for internal distribution wiring.

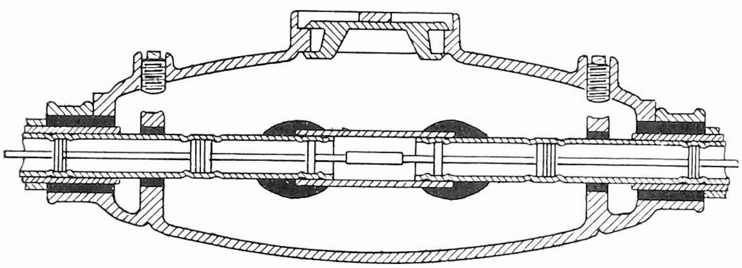


Fig. 6.

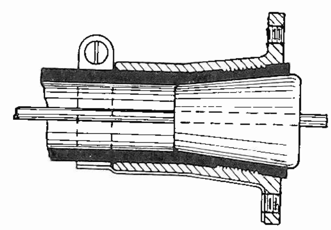


Fig. 7.

The cables are thoroughly desiccated in the works, and further drying is unnecessary if cutting, jointing and terminating are carried out with due precaution against the admission of moisture.

In jointing, the centre conductors are scarfed, soldered, bound with fine wire, and resoldered, while the outers are connected by a lead sleeve with wiped joints. For this operation, the trolitul discs at the joint are reinforced by ceramic jointing washers, so that when the former soften under the heat applied in making the wiped joints, the latter maintain the inner conductor in its proper position. For outside installation, the joints are enclosed in compound-filled iron joint boxes, the armouring being terminated in end clamps in these boxes. A typical arrangement is shown in Fig. 6.

For terminating the cables, a special gland has been designed in which both conductors are securely held. This is illustrated in Fig. 7.

It will be seen that while the gland entry fits the outer conductor, the bore of the gland increases towards the flange. The cable end, after insertion into the gland, is expanded to conform accurately with the taper bore, and is secured in this position by a taper plug drilled to take the centre conductor. This plug is smeared with hot compound and then tapped into the cable end, thereby providing an effective seal.

These cables are at present chiefly used in connecting aerial systems to radio-receiving stations, and upwards of five miles have already been installed for this purpose in various countries, whilst a further two miles is in course of manufacture for other projected installations.

At the same time they are clearly well suited to long distance multi-channel carrier working, or for television transmission.

E. W. Smith.
R. F. O'Neill.

THE EFFECT OF THE ELEVEN YEAR SUNSPOT CYCLE ON SHORT WAVE COMMUNICATION

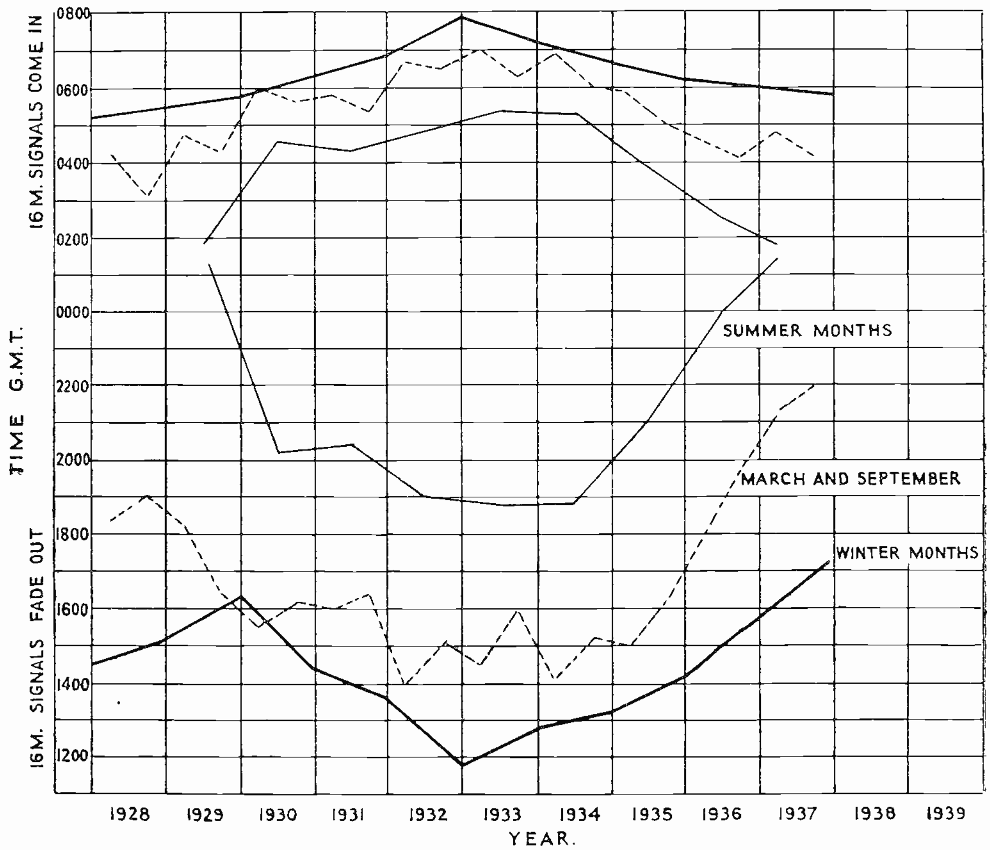
The following gives, very briefly, some of the changes which have been noticed in short wave communication with the change in the eleven year sunspot cycle.

The varying conditions during the past few years are illustrated by the times at which the 16 metre wavelength used on the Indian Beam circuit to London ceased to be of use in the afternoon or night and commenced to be of use in the morning.

These times are dependent on electron limitation in the F2 layer and therefore give some indication of the changes which have occurred in the electron density of this layer.

THE Marconi short wave beam services have now been in operation during practically a complete sunspot cycle. The last sunspot maximum was in 1928-1929 and we are now approaching the next maximum.

Short wave wireless communications have undergone a marked change during this period, and routes in the higher latitudes have been effected to a greater extent



than routes in the central latitudes, probably due to effects connected with the magnetic poles.

During the sunspot minimum years, for a given time of the day or night, it was necessary to use considerably longer wavelengths than had been used during the last sunspot maximum years on the northern latitude routes such as London to Montreal. On the more southern routes such as London to India and London to Capetown, the original day and night wavelengths were used throughout the cycle, but the times at which the change from the day to the night wave was made varied considerably. Measurements of critical frequency for the E and F layers during sunspot maximum years have shown that reflections can be obtained at higher frequencies for a given time of day or night than for the same times during the sunspot minimum years, but that this change is much greater for the F2 layer (average 300 kilometres height) than for the F1 (200 kilometre) layer or the E (100 kilometre) layer. From records it appears that the change of conditions throughout this cycle has not been quite regular. A rather sudden change towards the sunspot minimum values occurred during the year 1930, after which the change was slower through the minimum years with an abrupt change towards the maximum conditions towards the end of 1935. The change in conditions during 1930 was probably connected with the fact that magnetic storms were abnormally severe during that year. The diagram from data kindly supplied by Cables and Wireless shows the changes which have taken place in the London to India circuit.

The curves are for the 16-metre day wave, and they show the times at which the signals on this wavelength ceased to be of use in the afternoon or night, and the times at which signals became useful again in the morning. Times between these are covered by the night wave of 34 metres.

During the month of June in the sunspot maximum years, it has been possible to use the 16-metre wave throughout the 24 hours on this route.

As an example of the change in conditions during the cycle on the more Northern routes, wavelengths of the order of 70 metres were used between London and Montreal during the later part of the winter nights in the sunspot minimum years, whereas wavelengths of the order of 30 metres can be used during the sunspot maximum years at this time.

K. W. TREMELLEN.

PATENT ABSTRACTS

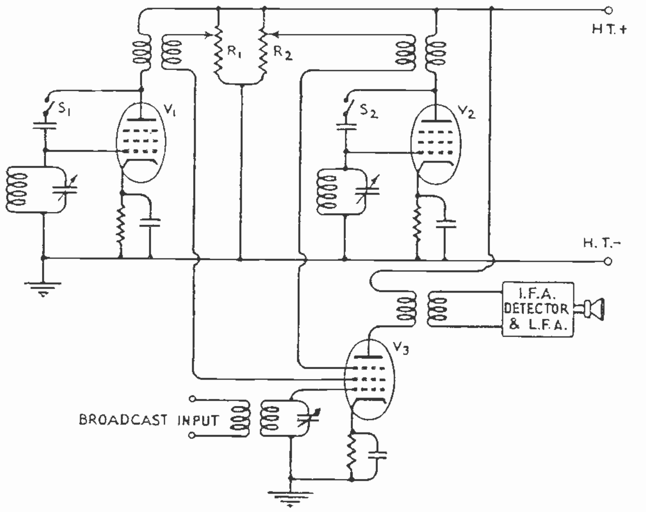
Under this heading abstracts are given of a selection from the most recent inventions originating with the Marconi Co. These abstracts stress the practical application of the devices described.

COMBINED BROADCAST AND TELEVISION RECEPTION.

Application date, September 18th, 1936. **No. 481,893.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and A. A. Linsell.

A method is here disclosed of using the television section of a combined television and all-wave receiver to produce the heterodyne oscillation for the frequency changer of the all-wave section. The heterodyne is the difference frequency derived by detection of two ultra short wave oscillations obtained from the television section. The short wave oscillations may be obtained from the oscillator or oscillators of a superheterodyne television receiver, or from two H.F. stages of a straight R.F. television receiver back coupled to produce oscillation. Such a circuit is shown in the figure. Valves V_1 and V_2 are two R.F. amplifier valves in the television section and, when closed, switches S_1 and S_2 provide capacitive feedback between the anode and grid circuits.



PATENT No. 481,893.

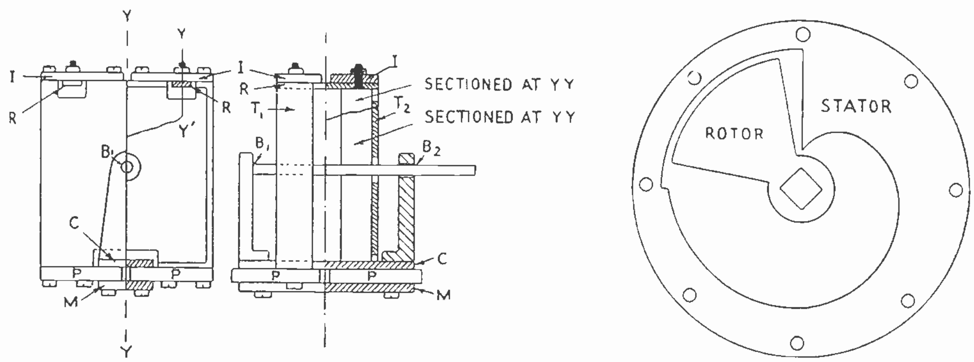
Oscillation occurs and the outputs are connected to two electrodes of the multigrid frequency changer of the all-wave section. Suitable bias is applied to these electrodes by means of potentiometers R_1 and R_2. The difference frequency is produced in the frequency changer and this combines with the input signal to form the intermediate frequency of the all-wave section. The difference frequency may, of course, be obtained by detection in a separate valve before application to the frequency changer of the all-wave section. For television reception switches S_1 and S_2 are open and the two stages act normally as R.F. amplifiers.

VARIABLE CONDENSER CONSTRUCTION.

Application date, July 21st, 1936. **No. 478,637.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and T. D. Parkin.

Multisection variable condensers in which temperature effects are reduced to a minimum are the subject of this specification. The important features are the division of the insulation securing the top and bottom of the stator plates into four sections and the use of a rotor plate of circular sector shape. The four insulating sections are held together by strips of the same metal as the rest of the condenser so that distortion due to the different coefficient of expansion of the insulation is much reduced.



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The drawing on the left indicates the construction of the condenser. The two rotor shaft bearings B_1 and B_2 are fixed to two metal strips C and M between which are sandwiched the divided stator insulating plates P. The stator housings T_1 and T_2 secured to the insulating plates P are cut away to allow clearance for the strip C. A similar construction holds the housings at the top, the insulation being split into four sections I as for the bottom plates P, and secured by two brass strips R. If the same metal, e.g., brass is used for the metallic parts of the condenser the whole structure expands or contracts at the same rate in all directions, and there is no distortion due to the insulating material.

An improved type of rotor and stator plate is also described. The rotor is part of a sector of a circle and the stator is a shaped circular disc as shown on the right. The correct capacity law curve is obtained by shaping the stator plate. This has the advantage of using rotor plates of uniform radius so that the tendency to warp with temperature change is almost entirely eliminated. The circular construction of the stator plate also tends to reduce warping. The angle of the rotor vane sector may be varied to suit any particular requirements and mechanical balance of the rotor may be obtained by staggering adjacent rotor and stator sections.

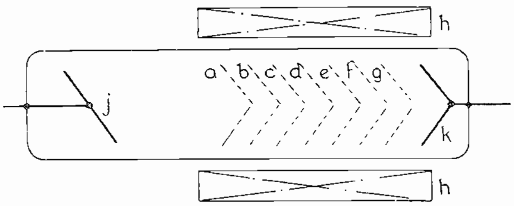
ELECTRON MULTIPLIER SYSTEMS.

Application date, October 21st, 1936. **No. 483,575.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and G. B. Banks.

This specification discloses an improved type of electron multiplier of the type wherein a primary electron stream is multiplied by secondary emission from a series of grids, these grids being situated between the cathode from which the primary electrons are emitted and the anode where the multiplied secondary electron stream is collected.

Obviously to obtain the greatest amount of secondary emission from any grid it is essential that as many as possible of the electrons approaching the grid shall strike it to produce secondary emission and shall not simply pass through it. In order to secure this the grids may be so arranged that their electrostatic fields are distorted so as to impart a "swirling" action to the electrons; for example they



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may be canted to the lines of force produced by the magnetic field applied to the multiplier.

Alternatively, as shown in the figure, the grids *a*—*g* may be cone shaped and arranged so as to produce an electrostatic field having a component normal to the magnetic field produced by the solenoid *h*. The electrons produced by the photo-emissive cathode *j* and collected by the anode *k* will then perform rotary movements and will produce the maximum amount of secondary emission.

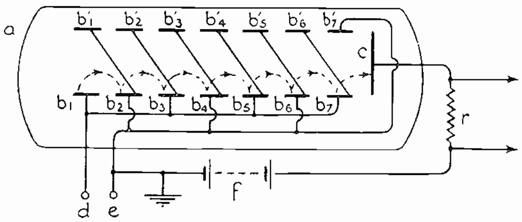
In order to minimise external connections the various voltage applied to the electrodes may be obtained from a potentiometer included in the envelope of the multiplier, only the end leads of this being brought out through the envelope.

ELECTRON MULTIPLIER SYSTEMS.

Application date, June 25th, 1936. **No. 478,813.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and G. B. Banks.

The majority of electron multipliers suffer from two serious disadvantages. The first is that a high operating potential is required and the second that the form of electrode structure necessitates a relatively large number of external connections being taken through the glass wall which encloses the electrode.



PATENT NO. 478,813

Both of these drawbacks are avoided in the method of electron multiplication disclosed in the present patent.

In the figure *a* is an evacuated envelope containing a primary cathode b_1 from which electrons are emitted either thermionically or photo-electrically, a series b_2—b_7 of secondary emitter electrodes, a series of field electrodes b'_1—b'_7 and an output anode c. b'_1 is connected to b_2, b'_2 to b_3 etc. ; b_1, b_3, b_5 and b_7 are connected to each other and, through the envelope, to d ; b_2, b_4 and b_6 are connected to each other and through the envelope to e. The anode c is connected, through the envelope, through a resistance r to the positive of a battery f, the negative of which is connected to e, which is earthed. A high frequency voltage is applied between d and e, and a magnetic field is applied across the space between the two planes occupied by electrodes b_1—b_7 and b'_1—b'_7.

When light falls on b_1 electrons are emitted and as, due to the high frequency voltage, b_2 becomes positive these electrons travel to b_2 and cause secondary emission from b_2. The frequency of the voltage applied between d and e is arranged so that, just as the original electrons reach b_2, b_2 reverses polarity, b_3 and b'_2 become positive and the secondary electrons proceed to b_3. The process is repeated until the output anode A is reached.

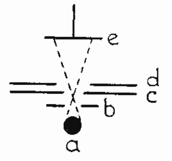
It will be appreciated that in place of the usual high voltage supply a relatively small H.F. voltage is needed in this construction and that moreover the number of external leads has been reduced to three.

BEAM VALVE SYSTEMS.

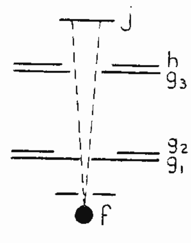
Application date, October 26th, 1936. **No. 483,827.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and G. F. Brett.

This invention relates to improved forms of electron beam valves, its principal object being to reduce the potentials which must be applied to the control grid in order to control the electron beam current (i.e., to increase the mutual conductance of the valve) and to make the beam less convergent. The normal electron beam valve, shown in Fig. A, comprises a rectilinear cathode behind a slotted control electrode b. An accelerator anode c follows, in front of which is an apertured suppressor electrode d and a collector electrode e. In this type of valve a somewhat divergent beam (as shown by the dotted lines) is obtained.



(A)



(B)

PATENT No. 483,827.

In the proposed construction therefore (Fig. B) a rectilinear cathode f lies immediately behind slotted first and second accelerator anodes g_1 and g_2. At a relatively great distance in front of g_2 is a third slotted accelerator anode g_3 in front of which is the suppressor electrode and the collector electrode j. The slotted aperture on g_2 is considerably wider than in g_1, and the aperture in h wider than in g_3. In use a relatively small positive potential is applied to g_1 and considerably higher potentials to g_2 and g_3. By this construction the divergence of the electron beam is diminished and the mutual conductance of the valve increased.

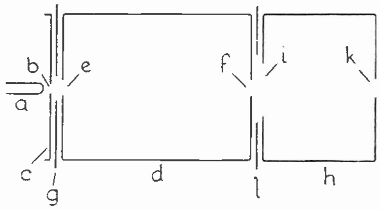
CATHODE RAY TUBE SYSTEMS.

Application date, August 27th, 1936. **No. 480,711.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and N. Levin.

This invention discusses means whereby cathode ray tubes can be operated on relatively low voltages. This is achieved by providing improved electrode systems as shown in the figure.

The emission point of a cathode a is placed close behind an aperture b in a plate c which serves as a first anode. Co-axial with this is a second cylindrical anode d having apertured ends e and f. d is insulated from c by a mica disc g. A third anode h mounted co-axially with c and d also has apertured ends i and k, and is insulated from d by means of a mica disc 1.



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In use a may be maintained at zero potential, c may vary between zero and +100 volts, d may be maintained at +170 volts and h at +500 volts. The apertures b and e form an electron lens and f and i a second electron lens.

An essential feature of the electrode system is that in each electron lens the aperture which is farther away from the cathode is of substantially larger diameter than the other aperture of the lens. Thus e is larger than b and i is larger than k.

The apertures are so dimensioned that the combination be causes the beam to be only slightly divergent or convergent and the combination fi focuses the beam. The final spot produced is bright, both on account of the fact that the beam can be of relatively high current, and also that very sharp focusing may be obtained.

A modulating grid electrode can be interposed between a and b if the tube is to be used for television purposes.

TIME BASE SYSTEMS.

Application date, July 21st, 1936. **No. 483,545.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and D. J. Fewings.

This invention relates to an improved method whereby saw-toothed waves can be produced for use in time bases for television purposes.

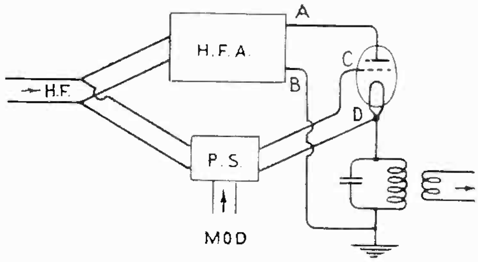
In Fig. A a is an oscillator designed to give an approximately sinusoidal output, this being taken through condensers b and c to two deflection plates e and f of an elementary cathode ray tube d. The electron stream from the cathode of this tube is arranged to be of ribbon formation and, in its undeflected position, would pass through a slit aperture in the plate g, behind which is a collector anode h. Now when the beam is deflected by the oscillator a, it will normally be collected by the plate g, except once every half cycle of the oscillator a, when it will pass undeflected through g and be collected by h. Thus once every half cycle there will be a sharp fall of current in the resistance j and consequent rise of potential of g. This rise of potential E which maintains the system in the carrier condition. When speech input ceases, the voltage across the condenser decays relatively slowly, but when it reaches a predetermined value the relay armature returns to the rest contact, the potential developed from E is suddenly removed from the grid of the valve and the carrier is cut off, but the carrier is maintained at its full value until then.

PHASE MODULATION SYSTEMS.

Application date, July 21st, 1936. **No. 478,708.**

Patent issued to Marconi Wireless Telegraph Co., Ltd., and W. T. Ditcham.

This invention discloses a simple high efficiency system of modulation whereby an amplitude modulated output is obtained by a phase control method, the modulation potentials being arranged to control the phase relationship of the H.F. potentials on the grid and anode of the modulating valve.



PATENT NO. 478,708.

One method of carrying out this system of modulation is shown in the figure, where H.F. energy from a constant frequency driving source passes through an H.F. amplifier (HFA) of the conventional type. Across the tank circuit of the amplifier is connected a modulator valve in series with the output tuned circuit as shown. The initial H.F. driving source is also coupled to the grid of this modulating valve through a phase shifting (PS) network as shown, the modulation being connected to the phase shifting network in such a way that changes of modulation potential control the phase relationship of voltage across the grid and cathode points CD of the modulation valve relative to the potentials developed from the tank circuit of the amplifier.

It is clear that when the potential on the grid of the modulating valve is made highly negative the modulating valve acts as an open circuit, whereas in the condition of maximum positive potential on the grid, the valve is very nearly short-circuited. Since there is an H.F. alternating potential across the modulator output circuit which varies in amplitude with conductivity of the modulator, the energy delivered to the output circuit will be a function of the phase control of grid swing relative to that of the output H.F. voltage of the amplifier: This means that a change of phase of grid potential will operate to change the amount of power being supplied to the output circuit.

A point of interest to note is that since there is an alternating H.F. potential across both the modulator valve and the H.F. circuit, only one half cycle of this potential is effective and the circuit as shown is of the half wave type, but clearly a pair of modulators can be incorporated in order to provide a full-wave arrangement.