CHAPTER 11

# DESIGN OF RADIO FREQUENCY INDUCTORS

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## SECTION 1 : INTRODUCTION

The space available for coils in a radio receiver is invariably limited, and part of the design work on each coil is the obtaining of maximum $Q$ in a minimum volume. In the case of i-f transformers the volume is clearly defined by the shield used, and the same applies to r-f coils in cans, but where shielding is not used with r-f coils the increased $Q$ obtainable from larger diameter formers is offset against the increased damping from components and magnetic materials, such as the receiver chassis, adjacent to the coil.

The resulting form factor for each individual winding requiring maximum $Q$ and minimum self-capacitance is usually such that the length of the winding is very approximately equal to its diameter. To obtain this shape for i-f, broadcast and short-wave coils the winding method is varied.

I-F transformers use universal windings, and where the $Q$ and distributed capacitance requirements are not severe a single coil wound with a large cam (perhaps $\frac{3}{8}$ in.) will be satisfactory, even without an iron core, if litz wire [see Sect. 5(i)J] is used. However higher $Q$ values are usually required than are obtainable in this way and in such cases each winding consists of two or more pies, perhaps with an iron core.

The same types of coils are satisfactory in the broadcast band, although in the case of signal frequency coils which are tuned over a ratio of more than three to one it is essential to use very narrow windings or two or more pies to reduce the distributed capacitance of the winding. Litz wire and iron cores are commonly used to give a secondary $Q$ of 100 or more.

Another method of decreasing distributed capacitance is the use of progressive universal windings in which the winding finger travels along the former in addition to moving to and fro. Such coils are rarely built up to a greater height than five wire thicknesses, and are equally useful with or without iron cores. Solenoids are also used either with iron cores or on comparatively large diameter formers if an iron core is not used.

At frequencies between 2 and 6 Mc/s (which are used in receivers providing continuous coverage between the normal short-wave and broadcast bands, or for oscillators giving second harmonic mixing for the short wave band) the progressive universal winding is particularly useful because the winding length of solenoids may be too great with the usual range of former diameters, and the number of turns is insufficient to allow them to be split into many sections.

A set of worm driving gears which spreads the appropriate number of turns over a length approximately equal to the coil diameter is required. This usually necessitates a faster worm drive than is used for broadcast coils. Litz wire with a large number of strands is useful in obtaining maximum $Q$.

For the 6 to 18 Mc/s band the solenoid is used, almost always with solid wire, and this type of coil remains useful at least up to 100 Mc/s. Increased wire diameters and spacings are used at higher frequencies to obtain form factors similar to those used at low frequencies.

## SECTION 2 : SELF-CAPACITANCE OF COILS

*(i) Effects of self-capacitance (ii) Calculation of self-capacitance of single-layer solenoids (iii) Measurement of self-capacitance.*

### (i) Effects of self-capacitance

The self-capacitance of a coil is due to the electrostatic coupling between individual turns and between the turns and earth. When the self-capacitance is between uninsulated turns in air its $Q$ may be high, but the greater the amount of dielectric in the field of the coil the greater will be the losses.

Short wave coils of enamelled wire on a solid former do not have serious dielectric losses if good quality materials are used, but at broadcast and intermediate frequencies the universal windings used have comparatively high-loss dielectrics and unless self-capacitances are kept to a low value the reduction in $Q$ may be appreciable.

In the case of coils which are to be tuned over a range of frequencies, comparatively small values of shunt capacitance can have a large effect on the possible tuning range.

In all cases the self capacitance of a coil has an apparent effect on its resistance, inductance and $Q$, and at frequencies considerably below the self resonant frequency of the coil—

$$\text{apparent inductance} = L\left(1 + \frac{C_0}{C}\right)$$

$$\text{apparent resistance} = R\left(1 + \frac{C_0}{C}\right)^2$$

$$\text{and apparent } Q = Q/\left(1 + \frac{C_0}{C}\right)$$

where $C_0$ = self-capacitance of coil
$C$ = external capacitance required to tune $L$ to resonance
$L$ = true inductance of coil
$R$ = true resistance of coil
and $Q$ = true $Q$ of coil.

### (ii) Calculation of self-capacitance of single-layer solenoids

Until recently the work of Palermo (Ref. C3) had been taken as the standard on the self-capacitance of single-layer solenoids. However Medhurst (Ref. C2) has disputed the theoretical grounds on which Palermo's work is based. As the result of a careful analysis and a large number of measurements he states that the self-capacitance, $C_0$, of single-layer solenoids with one end earthed, and without leads, is

$$C_0 = HD \ \mu\mu\text{F} \qquad (1)$$

where $D$ = diameter of coil in cm.

and $H$ depends only on the length/diameter ratio of the coil. A table of values of $H$ is given below and when used in conjunction with a capacitance correction for the "live" lead, Medhurst states that the accuracy should be within 5%.

TABLE 1

| $\left(\frac{\text{Length}}{\text{Diameter}}\right)$ | $H$ | $\left(\frac{\text{Length}}{\text{Diameter}}\right)$ | $H$ | $\left(\frac{\text{Length}}{\text{Diameter}}\right)$ | $H$ |
|---|---|---|---|---|---|
| 50 | 5.8 | 5.0 | 0.81 | 0.70 | 0.47 |
| 40 | 4.6 | 4.5 | 0.77 | 0.60 | 0.48 |
| 30 | 3.4 | 4.0 | 0.72 | 0.50 | 0.50 |
| 25 | 2.9 | 3.5 | 0.67 | 0.45 | 0.52 |
| 20 | 2.36 | 3.0 | 0.61 | 0.40 | 0.54 |
| 15 | 1.86 | 2.5 | 0.56 | 0.35 | 0.57 |
| 10 | 1.32 | 2.0 | 0.50 | 0.30 | 0.60 |
| 9.0 | 1.22 | 1.5 | 0.47 | 0.25 | 0.64 |
| 8.0 | 1.12 | 1.0 | 0.46 | 0.20 | 0.70 |
| 7.0 | 1.01 | 0.90 | 0.46 | 0.15 | 0.79 |
| 6.0 | 0.92 | 0.80 | 0.46 | 0.10 | 0.96 |

Lead capacitance can be determined separately and added to the coil self-capacitance. Fig. 11.1 (from Medhurst) can be used for this purpose.

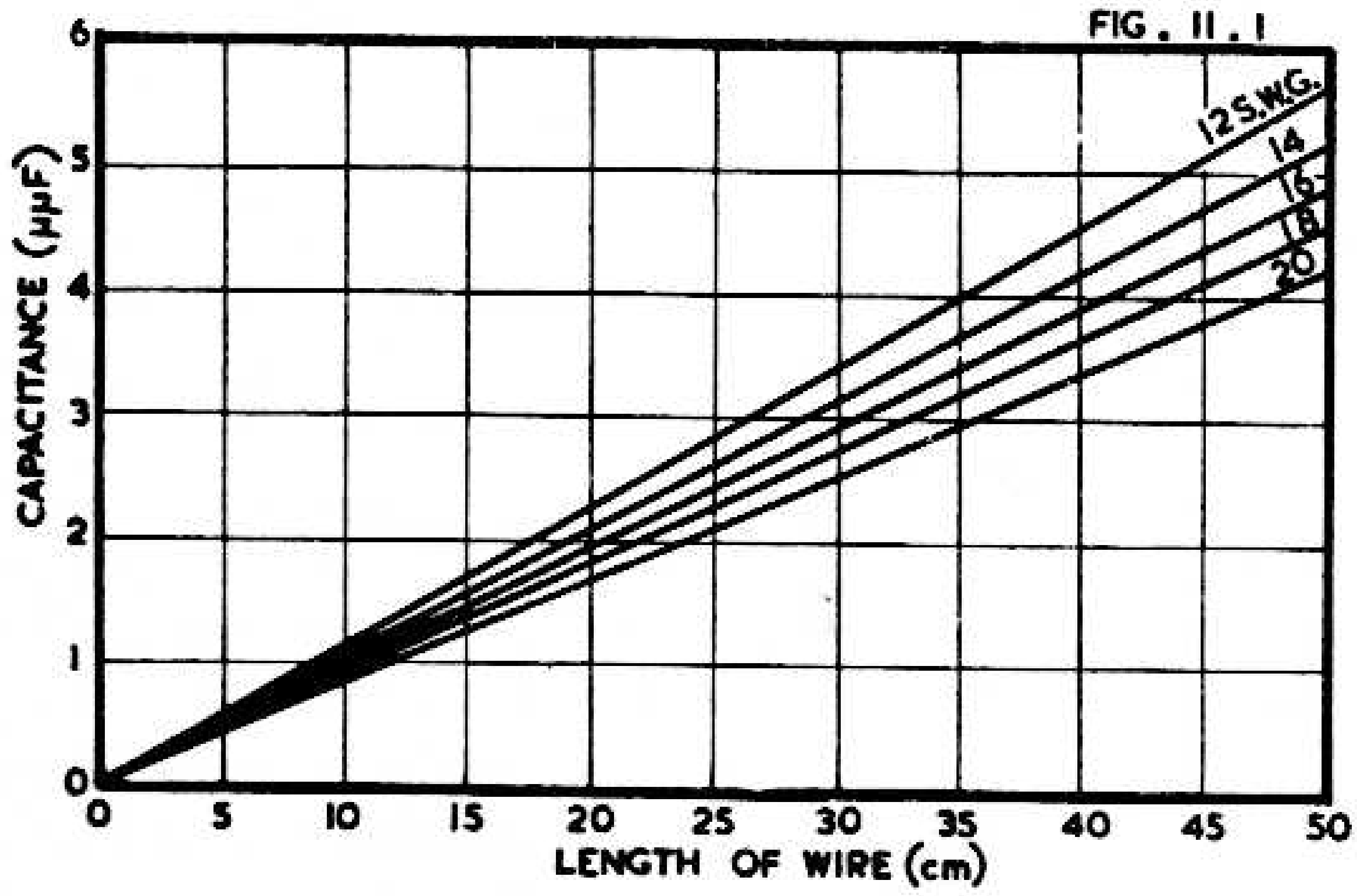


*Fig. 11.1. Variation of capacitance with wire length for vertical copper wires of various gauges (Ref. D6).*

It is interesting to compare Medhurst's formula with another by Forbes Simpson [Ref. G17 and Sect. 5(i) of this chapter]. Forbes Simpson's formula is applicable to coils with a length/diameter ratio of unity, a pitch of 1.5 times the wire diameter and with leads at each end equal in length to the diameter of the coil. His formula is

$$C_0 = D(0.47 + a)\ \mu\mu\text{F} \tag{2}$$

where $D$ = diameter of coil in cm.

and $a$ = a constant depending on the gauge of wire, lying between 0.065 for 42 S.W.G. (38 A.W.G. approx.) and 0.11 for 12 S.W.G. (10 A.W.G. approx.) where one lead of the coil is connected to chassis.

In eqn. (2) there is no equivalent of the $H$ in eqn. (1) because the length/diameter ratio is a constant. The length of "live" lead taken into account in eqn. (2) is equal to $D$, and it is possible to determine its effect from Fig. 11.1. For 12 S.W.G. wire the capacitance is 0.11 $\mu\mu$F per cm. so that eqn. (1) (with lead correction) could be written

$$C_0 = 0.46D + 0.11D = 0.57D$$

for the conditions of eqn. (2) and using 12 S.W.G. wire. Similarly, using the value of $a$ given for 12 S.W.G. wire in eqn. (2), eqn. (2) could be written

$$C_0 = 0.47D + 0.11D = 0.58D.$$

The data in Fig. 11.1 are not sufficient to compare the equations using the lower limit of $a$ given by Forbes Simpson but for 20 S.W.G. eqn. (1) becomes

$$C_0 = 0.55D$$

for the conditions of eqn. (2), while eqn. (2) for 42 S.W.G. becomes

$$C_0 = 0.535D.$$

## (iii) Measurement of self-capacitance

A graphical method of self-capacitance determination due to Howe is shown in Fig. 11.2. Values of external capacitance in $\mu\mu$F required to tune a coil to resonance are plotted against $1/f^2$ where $f$ is the resonant frequency in Mc/s. The self-capacitance is the negative intercept of the straight line with the capacitance axis.

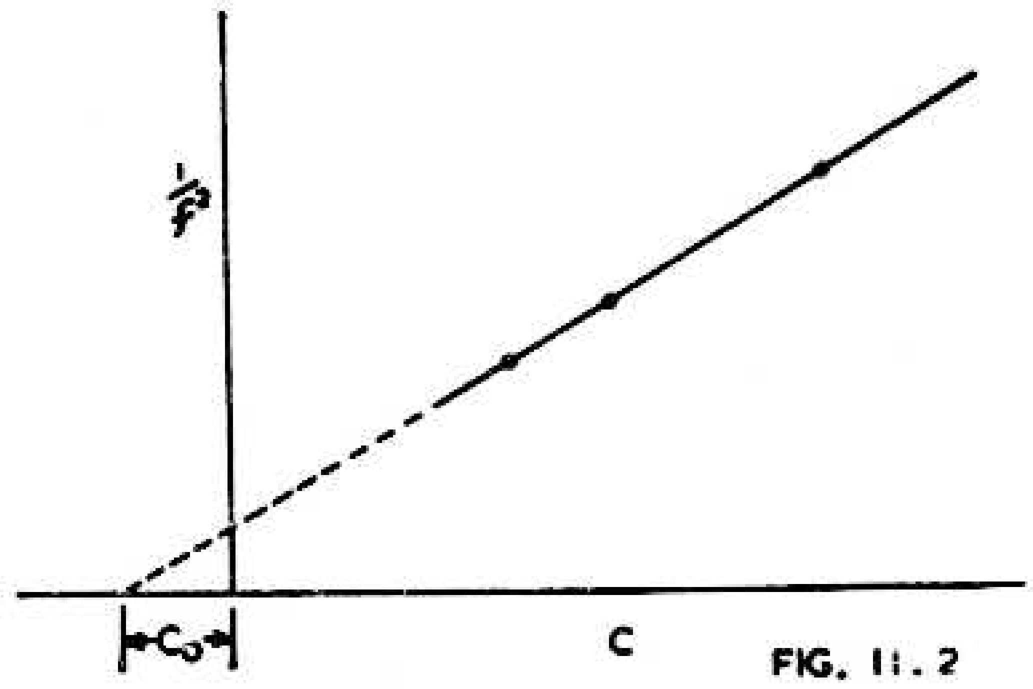


*Fig. 11.2. Construction for determination of self-capacitance of coil.*

An alternative method is to determine the external capacitances necessary to tune a coil to frequencies of $f$ and $2f$. If the two capacitances are $C_1$ and $C_2$ respectively,

$$\text{self-capacitance} = \frac{C_1 - 4C_2}{3}.$$

# SECTION 3 : INTERMEDIATE-FREQUENCY WINDINGS

*(i) Air-cored coils (ii) Iron-cored coils (iii) Expanding selectivity i-f transformers (iv) Calculation of gear ratios for universal coils (v) Miscellaneous considerations.*

## (i) Air-cored coils

Most commercial receivers have two i-f transformers with four circuits tuned to about 455 Kc/s. To obtain adequate selectivity in such a case the required $Q$ for each winding (mounted on the chassis but not connected in the circuit) is 100 or more.

This $Q$ can readily be obtained with air cored coils of about 1 mH inductance provided that litz wire and comparatively large coil cans are used. Without litz it is difficult to exceed a $Q$ of 50, but even this may be sufficient when more i-f stages than usual are used.

When a single pie is used, $Q$ is to some extent dependent on coil shape but if five times the winding depth plus three times the winding width equals the external diameter, $Q$ will be close to the maximum for the wire and type of winding.

There are three methods of increasing the $Q$, and when all are used to practical commercial limits, production $Q$ figures of 150 can be maintained. The first requirement is litz wire, probably nine strands for a $Q$ of the order of 100, and twenty or more strands for a $Q$ of 150. Large coil cans are necessary, up to two inches in diameter, with formers of such a size as to make the outside coil diameter little more than half the can diameter. Lastly, self-capacitance must be reduced to a minimum, because its $Q$ is always low and in the case of a single pie winding it may amount to say 25 $\mu\mu$F, a large percentage of the total capacitance. Splitting the winding into pies reduces self-capacitance, at the same time improving the form factor, and a suitable compromise between $Q$ and economy is obtained by winding with three pies. Self-capacitance is also reduced by winding narrow pies, although a limit is set by the difficulty of winding litz wires with a cam of less than 0.1 inch, or perhaps even $\frac{1}{8}$ inch. In addition, as the cam is reduced the height of the coil increases and larger losses from damping of the coil by the can may more than offset the reduced losses in a smaller self-capacitance.

When the windings have more than one pie, the inductance is dependent on pie spacing. For this reason, and to increase the speed of winding, multi-section coils are wound with the former located by a gate, a different slot in the gate being used for each individual pie. Spacing between primary and secondary windings is kept constant by means of a double winding finger which winds primary and secondary at the same time.

A disadvantage of air cored i-f coils is that a trimmer capacitance is needed to resonate the tuned circuit. The cheapest types of trimmers usually have poor stability with respect to time, vibration, humidity and temperature, and satisfactory types may cost more than a variable iron core with provision for adjusting it.

## (ii) Iron-cored coils

The advantages of iron cores used in i-f transformer windings are that

(a) they increase the inductance of the winding, thereby giving a saving in wire cost and winding time for a given inductance,

(b) they increase the $Q$ of the winding, thereby allowing cheaper litz (or solid wire in an extreme case) to be used,

(c) they restrict the field of each winding, to an extent depending on the type of core used, and thus allow a smaller, cheaper coil can to be used without excessive damping,

(d) they provide a satisfactory method of adjusting the tuned circuit to resonance. Because of this the total tuning capacitance can be of a stable high $Q$ type (e.g. a silvered-mica capacitor), and there is no trouble from capacitance between primary and secondary trimmers giving an asymmetrical resonance curve, or from capacitance between say first i-f and second i-f trimmers leading to regeneration.

To obtain the greatest benefit from an iron core it is necessary to have all of the turns close to the core. In particular cases, especially at high frequencies, $Q$ may be decreased due to increased distributed capacitance if the coil is too close (e.g. wound on the core) but in normal i-f applications in which a former comes between coil and core the thinner the former can be made, the better will be the $Q$.

Also because of this effect it sometimes happens that by winding a coil with thinner litz (fewer strands), and so bringing the top turns closer to the iron and further from the can, an increase in $Q$ is obtained.

Another benefit from bringing all of the turns as close to the iron core as possible is that the range of inductance adjustment provided by the core is increased. If fixed capacitors with $\pm$ 10% tolerance are used to tune the i-f windings, it is necessary to provide at least $\pm$ 15% inductance adjustment from the iron core to allow for winding variations and changes in stray capacitances due to valves and wiring. It is necessary to split the i-f winding into pies to obtain a variation of this order with normal cheap i-f cores.

An additional advantage of splitting the winding is that at least half of the turns in the coil are brought much closer to the core so that $Q$ is increased above the amount to be expected from the reduction in distributed capacitance.

To obtain the previously mentioned $Q$ of 100 at an inductance of about 1 millihenry with a small iron core, a coil can somewhat larger than one inch diameter would probably be needed with a coil wound in three pies of 5 strand 41 A.W.G. litz.

Much higher $Q$'s are obtainable with special core shapes and materials, and a 455 Kc/s i-f transformer with a $Q$ of 260 in the can is mentioned in Ref. A15. However difficulty is usually experienced from instability even with valves of comparatively low slope, such as the 6SK7-GT, if i-f transformers with a $Q$ above about 150 are used.

A possibility of increasing $Q$ with iron-cored coils which does not usually arise with air-cored coils is that if the coil is wound directly on to a non-adjustable iron core or on to a small individual former, it can be mounted in different planes inside the coil can. By mounting the coil with its axis at right angles to the axis of the can it is possible to obtain an increase in $Q$ of the order of 10% (depending on the diameters of coil and can) and this method also has the advantage of making both cores adjustable from above the chassis. The chief disadvantage is that coupling between primary and secondary takes place mainly between adjacent edges of the two windings and variations in the height of the windings, due to variations in wire thickness, noticeably affect the coupling. In addition, trouble may be experienced from regeneration because with this method of mounting the field of the bottom winding extends further outside the coil can. Unless a good joint is made between can and chassis this trouble will be further aggravated.

### (iii) Expanding selectivity i-f transformers

The simplest method of expanding i-f transformer response to give peaks symmetrically spaced about the intermediate frequency is by switching a tertiary winding. Two other possibilities* of obtaining the desired result are the simultaneous switching of series and shunt coupling capacitors between the two windings, and the switching of capacitors to alter the coupling in a $\Pi$ or T network (Ref. G13). In each of these cases, however, it is necessary to switch components in high potential sections of the circuit whereas tertiary switching can be carried out in low potential circuits.

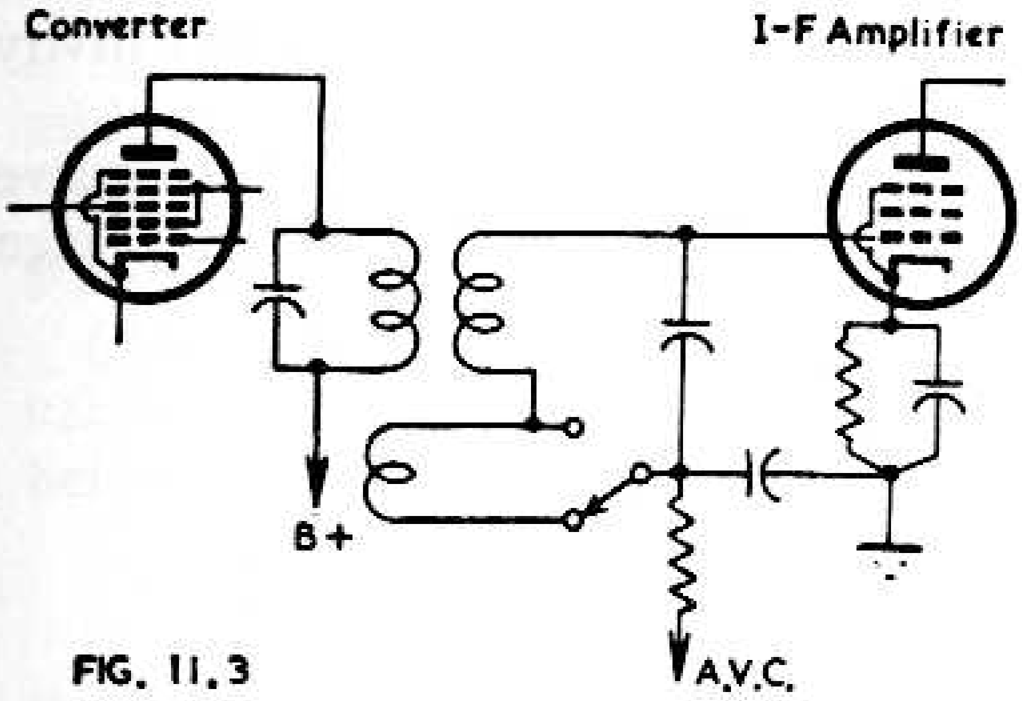


FIG. 11.3

*Fig. 11.3. Expanding selectivity with switched tertiary coil.*

Receivers have been manufactured in which movement of a complete winding altered the coupling, and whilst this eliminates switching problems, the additional mechanical problems are at least as troublesome. Moreover it is difficult to keep constant the damping effect of the coil can and the mechanical coupling of the moving coil so that the whole object of the variable coupling (i.e. symmetrical expansion of the pass-band) is liable to be lost.

Although symmetrical expansion becomes more difficult as the $Q$ of the transformer increases, it is possible with tertiary switching to maintain satisfactory symmetry in production with a $Q$ of 150 and with simple alignment procedures. In a normal receiver with a single i-f amplifier it is only necessary to expand the selectivity of the first i-f transformer because the diode damping on the secondary (and perhaps primary) of the second i-f reduces its selectivity considerably.

Fig. 11.3 gives a typical circuit. It will be seen that the inductance of the tertiary winding is switched into the tuned circuit in the expanded position. This does not lead to appreciable detuning because the inductance of the secondary is some thousands of times larger than that of the tertiary.

Suitable methods of winding the tertiary coil vary with the method of primary winding but the object is always to obtain maximum coupling with minimum tertiary inductance. When the primary is pie-wound the tertiary should be wound between two pies as close to the middle of the coil as possible. From three to five tertiary turns are required (assuming approximately critical coupling without the tertiary and three hundred or more primary turns in a 455 Kc/s i-f transformer). The type of wire is not important except for its covering, and solid wire of the finest gauge which will give no trouble with handling or cleaning is probably most suitable.

It is essential that the insulation resistance between primary and tertiary be very high. In a typical case a leakage of 100 megohms between primary and tertiary would give the control grid of the following i-f amplifier a positive voltage of five volts. Because of this the tertiary winding should be double-covered wire and the completed coil must be thoroughly dried out and then immediately impregnated in some moisture resisting compound.

If the primary consists of a single winding the tertiary is best wound in solenoid form on top of it. Similar results are obtainable for the same number of coupling turns.

To align such a transformer the switch should be turned to the " narrow " position and the receiver aligned normally for maximum gain. When the switch is turned to " broad " the output should drop, but when the signal generator is detuned, equal peaks should be found symmetrically spaced about the intermediate frequency.

*See also Chapter 9 Sect. 8 ; Chapter 26 Sect. 5.

If the peaks are not symmetrical, undesired couplings are probably responsible. All traces of regeneration must be removed and, because the stability requirements are more severe than usual, unusual effects are liable to be uncovered. For example, regeneration may occur due to coupling between a loop formed by the generator input leads and the output of the i-f amplifier (twisting generator leads will cure this) or coupling may occur between first and second i-f transformers within the steel chassis. When this happens, rotating one or both transformers will probably give cancellation but leave production receivers susceptible to the trouble. Additional spacing between transformers is advisable with each primary and secondary wired for minimum regenerative coupling.

The switch used for the tertiary winding should preferably be of the " break before make " variety. A " make before break " switch momentarily short circuits the tertiary winding during switching, giving a sudden reduction and increase in sensitivity which can be heard as a click.

Even after a satisfactory i-f selectivity curve has been obtained, the over-all curve may be too narrow if a r-f stage is used in the receiver. In such cases the r-f stage should also be expanded [Chapter 35 Sect. 5(iii)].

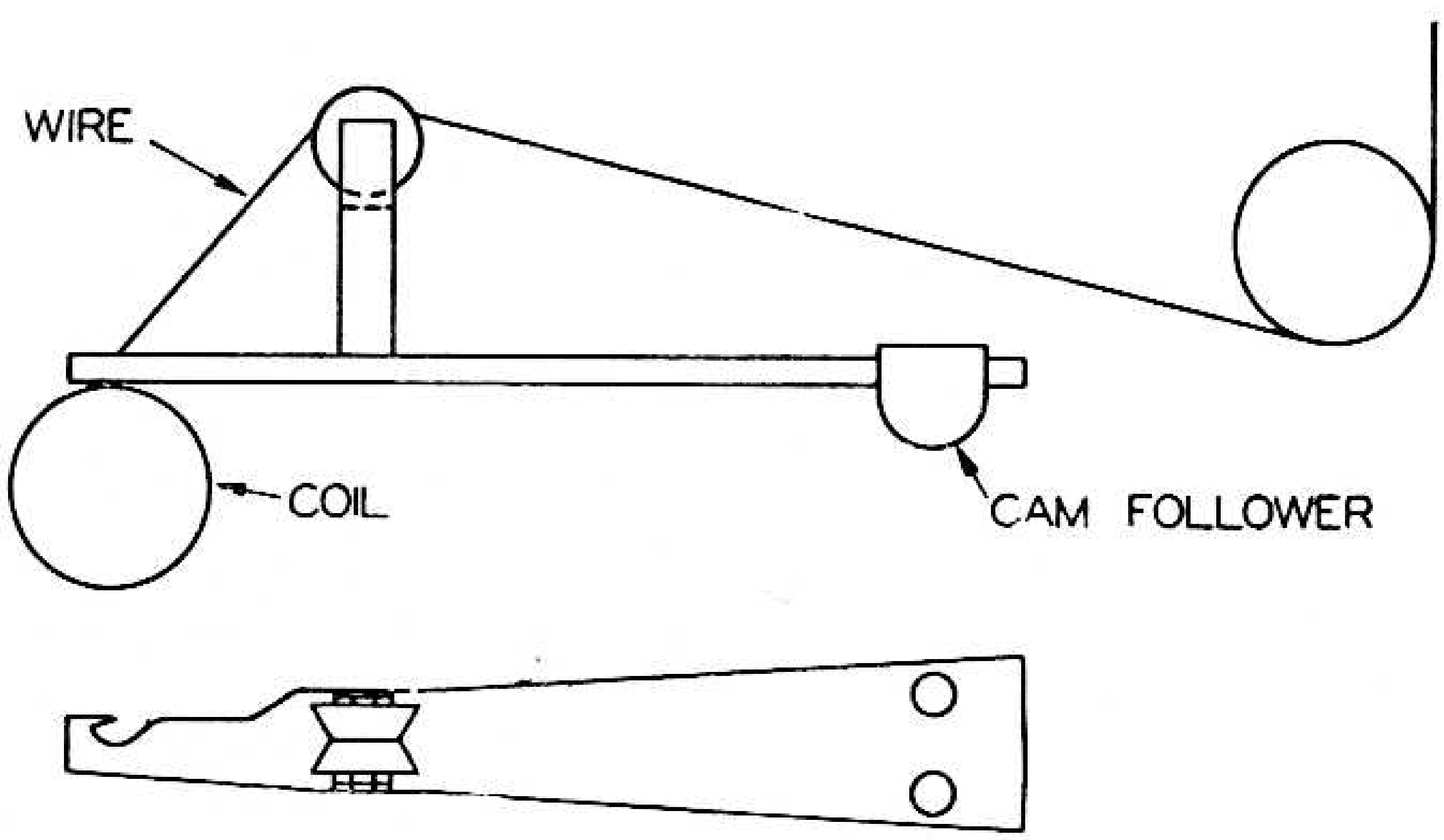


FIG. 11.4

*Fig.* 11.4. *Type of finger recommended for universal winding.*

## (iv) Calculation of gear ratios for universal coils

To obtain electrical consistency between universal coils in production runs it is essential to wind the coils with good mechanical stability. For this, machines must be adjusted correctly, but a suitable gear ratio is of equal importance.

The method of gear ratio calculation given below has been used successfully for production coils, and gives a straight pattern on the side of each coil.

The instructions make provision for a spacing between centres of adjacent wires of 8/7 of the wire diameter. This diameter should be measured, and the required spacing depends on the type of winding finger used. The most satisfactory types, and the ones for which the spacing factor will be found suitable, are those in which the wire passes to the bottom side of the finger before being placed in position on the coil. Fig. 11.4 shows such a finger, and it will be seen that the tension on the wire pulls the finger down to the face of the coil.

Fingers in which the wire passes through a groove on the top and then takes up its own position on the surface of the coil may require more spacing than that specified. Other factors may also affect the spacing required, but a small amount of experiment will decide this. It will be noticed that some types of litz wire tend to spread when wound, and so need more spacing than would be expected from measurement.

**Symbols**

$d$ = former diameter (inches)
$c$ = cam throw (inches)
$n$ = nominal number of crossovers per turn (see Table 1)
$q$ = number of crossovers per winding cycle, i.e. before wire lies alongside preceding wire (see Table 1).
$v$ = nominal number of turns per winding cycle (see Table 1)
$R$ = gear ratio = former gear/cam gear
$w$ = modified wire diameter (inches)—see Note below
$P$ = $qc/(w + x)$ = an integer
$x$ = smallest amount necessary to make $(P + 1)/v$ an integer (inches).

Note. For fabric covered wire, $w$ = (measured diameter of covered wire) $\times$ 8/7. If the wire is enamelled only, the same formula is used but the bare wire diameter is multiplied by 8/7.

**Procedure**

(A) From $n \ngtr (2d/3c)$ determine the largest convenient value for $n$. Do not use values of $n$ less than 2 for bare enamelled wire. Obtain values of $q$ and $v$ from Table 1 below for the value of $n$ chosen.

TABLE 1

| | | | | | | | |
|---|---|---|---|---|---|---|---|
| $n$ | 4 | 2 | 1 | 2/3 | $\frac{1}{2}$ | 1/3 | $\frac{1}{4}$ |
| $q$ | 4 | 2 | 2 | 2 | 2 | 2 | 2 |
| $v$ | 1 | 1 | 2 | 3 | 4 | 6 | 8 |

(B) Determine $w$ from information given in the note above.
(C) Calculate $P$ from $P = qc/(w + x)$.
(D) Obtain $R$ from $R = \frac{1}{2}n(P + 1)/P$.

**Example 1**

Given $d = \frac{1}{2}$ in. and $c = 0.1$ in. determine the gears to wind a coil with 42 S.W.G. enamelled wire.

(A) $(2d/3c) = 1.0/0.3$. Take $n = 2$, giving $q = 2$ and $v = 1$ from the table.
(B) The diameter of bare 42 S.W.G. wire is 0.004 in., so $w$ = 0.004 57 in.
(C) $P = qc/(w + x)$ and $(P + 1)/v$ must be an integer, i.e. $P$ must be an integer since $v = 1$.
$qc/w = 200/4.57 = 43.7$.
But $P = qc/(w + x)$ = an integer.
Therefore $P = 43$.
(D) $R = \frac{1}{2}n(P + 1)/P = \frac{1}{2} \times 2(43 + 1)/43 = 44/43$.

**Example 2**

Given $d = \frac{1}{2}$ in. and $c = \frac{1}{4}$ in. determine the gears to wind a coil with 0.016 in. litz wire.

(A) $2d/3c = 2/3$. Take $n = 2/3$, giving $q = 2$ and $v = 3$.
(B) $w = 0.016$ in. $\times 8/7 = 0.0183$ in.
(C) $qc/w = 500/18.3 = 27.3$.
But $(P + 1)/v$ = an integer.
Therefore $P = 26$.
(D) $R = \frac{1}{2}n(P + 1)/P = 1/3 \times 27/26$.
To obtain suitable gears :—
$R = 2/3 \times \frac{1}{2} \times 27/26 = (28/42) \times (27/52)$.

When it is known that $n$ will be 2, as is the case with the majority of coils, the method reduces to dividing the modified wire diameter into twice the cam throw (ignoring any fractions in the answer). This gives $P$ and the required ratio is $(P + 1)/P$.

For further information on universal coil winding see Refs. I 1 to I 8.

## (v) Miscellaneous considerations

### (A) Direction of windings

Although the coupling between primary and secondary is assumed to be due to mutual inductance, the capacitive coupling is appreciable. Depending upon the direction of connection of the windings, the capacitive coupling can aid or oppose the inductive coupling. When the two types are in opposition and of the same order a slight change in one—the capacitive coupling is particularly liable to random variation—gives a much larger percentage variation in the effective coupling. In a bad case, production sensitivity variations from this cause may be quite uncontrollable.

To avoid the trouble, i-f transformers are usually connected for aiding capacitive and inductive coupling. This is done by connecting the i-f amplifier plate to the beginning of one winding and the following grid to the end of the other winding when the two coils are wound in the same direction. Other connections to give the same winding sense will give the same result.

Even with aiding couplings it is desirable to reduce capacitive coupling to a minimum to obtain a symmetrical response curve and care should be taken in the placing of tuning capacitors and with details such as keeping the grid wire of the secondary winding well away from the plate wire or the plate side of the primary winding.

### (B) Amount of coupling

An undesirable feature, from a production point of view, of i-f transformers in which the coupling is less than critical, is that receiver sensitivity becomes more dependent on i-f coil spacing. For a transformer with approximately critical coupling a spacing difference of 1/32 inch makes no appreciable difference to sensitivity in a typical case. However if the transformer were under-coupled the sensitivity change would be noticeable.

Transformers which are slightly over-coupled can be aligned for symmetrical response curves without undue trouble by detuning each winding (say with an additional capacitor of the same size as the tuning capacitor) while the other winding of the transformer is aligned for maximum sensitivity. Symmetry is of course dependent on the absence of regeneration.

### (C) Losses

Unless care is taken, the sum of a number of apparently negligible losses may result in appreciable reduction in $Q$. Many artificial coverings have greater losses than silk, but this is usually obvious when a sample coil is wound. Less obvious losses may occur in details such as the material used to seal the end of a winding or in the placing of shunt capacitors or even coil lugs close to the actual i-f winding. An assembly of straight parallel wires between top and bottom of the i-f can may cause appreciable decrease in $Q$, particularly if large blobs of solder are placed close to the windings.

To eliminate such losses it is advisable to check the $Q$ of a winding, sealed with low-loss material and baked to remove all moisture, when suspended well away from any substances which will introduce losses. The mechanical structure of the complete i-f assembly can then be added, one section at a time, and the $Q$ read at each stage, a final reading being taken when the coil is mounted in the receiver.

### (D) F-M i-f transformers

For details of F-M i-f construction and other i-f information of practical interest see Chapter 26 Sect. 4(vi).

### (E) Other ferromagnetic materials

The development of a non-metallic ferromagnetic material named **Ferroxcube** has been announced by Philips (Refs. A28, A29, A30, A31, A32, A33).

Several grades of Ferroxcube are manufactured. They have in common a high specific resistance of $10^2$ to $10^8$ ohm cm. and a high initial permeability of from 50 to 3000 depending on the type. Ferroxcube IV, which is useful to 40 Mc/s, has a permeability of 50, and Ferroxcube III with an upper frequency limit of about 0.5 Mc/s has a minimum permeability of 800.

Because of the closeness of the Curie point to room temperature (the Curie point of Ferroxcube III is 110°—160°C) some change of permeability occurs with changing temperature. For instance the permeability of Ferroxcube III can be almost halved by an increase in temperature from 20°C to 80°C. However, the permeability of Ferroxcube V is decreased less than 10% by a similar temperature change under the same conditions. Between 10° and 40°C the change in permeability averages 0.15% per 1°C for the various types of Ferroxcube.

In cases in which the magnetic circuit is normally completely enclosed, the high permeability of Ferroxcube can be used to minimize losses by means of an air gap. If the gap is such as to reduce the effective permeability to one tenth of its original value (which could still be high), losses and the effect of heat on effective permeability will also be reduced in the same proportion.

The properties of Ferroxcube allow considerable reductions to be made in the size of such items as i-f transformers or carrier-frequency filters, and it is an excellent material for magnetic screens or for permeability tuning. However its saturation point is rather low and it is not used for power transformer or output transformer cores.

Ref. A27 describes research on **ferromagnetic spinels** by the Radio Corporation of America. These spinels are ceramic-like ferromagnetic materials characterized by high permeability (up to greater than 1200), high electric resistivity (up to $10^8$ ohm cm), and low losses at radio frequencies. Wide ranges in these and other properties are obtainable by varying the component ingredients and methods of synthesis.

Ferrospinels are being used increasingly in electronic equipment operating in the frequency range of 10 to 5000 Kc/s. At power and low audio frequencies the ferrospinels are not competitive with laminated ferromagnetic materials, and at very high frequencies the losses in ferrospinels are excessive when high permeability is required. It is possible however to produce a ferrospinel, with low permeability, useful at frequencies in the order of 100 Mc/s. The application of ferrospinels as core bodies in the deflection yoke, horizontal deflection transformer and high voltage transformer for television receivers is now finding wide acceptance. The ferrospinels are especially suited to television video frequencies as their use in these components results in improved performance at lower cost and with smaller space.

In the standard broadcast receiver, the ferrospinels are expected to be used in the radio frequency circuits as " trimmers " and as permeability tuning cores. With a properly designed coil it is possible to tune a circuit, by the movement of a ferrospinel rod, from 500 to 3000 Kc/s, or to cover the standard American broadcast band (540 to 1730 Kc/s) with only three eighths of an inch movement of the rod.

By using a ferrospinel with a high electric resistivity as the core body for radio frequency inductances, the wire body may be placed on the ferrospinel without additional insulation. In fact, the conductor may be affixed by the printed circuit technique for some applications.

## SECTION 4 : MEDIUM WAVE-BAND COILS

*(i) Air-cored coils (ii) Iron-cored coils (iii) Permeability tuning (iv) Matching.*

### (i) Air-cored coils

With large diameter formers reasonable $Q$ can be obtained on the broadcast band with air-cored solenoids. However in limited spaces, coils are wound with two or more pies or by progressive universal winding when high $Q$ is required. The progressive winding has the advantage of being less susceptible to inductance variation through careless handling or winding, and from the production point of view it is desirable because it can be wound without stopping.

Although high $Q$ is not necessary for broadcast band oscillator coils, it is desirable. Low $Q$ tuned circuits need larger reaction windings which in turn give increased phase shift in the voltage fed back from the plate into the grid circuit. Consequently an even larger reaction winding is required and the result may be reactance reflected into the secondary circuit which makes it impossible to tune the required range—if the police band (U.S.A.) is to be included. In any case the reflected reactance, which is a variable throughout the tuning range, complicates the tracking problem.

A method of decreasing the coupling inductance required, which is applicable to progressive but not to ordinary universal windings, is to thread both secondary and primary wires together through the winding finger, wind the required number of primary turns, terminate the primary and proceed with the secondary winding until completed. This method is possible with progressive windings because of the gaps which occur throughout the winding, whereas with a normal universal winding there is no room for a primary between turns of the secondary.

For r-f and particularly for aerial coils the need for high $Q$ is greater. The sensitivity, signal-to-noise ratio and image ratio of a receiver are all governed by the aerial coil $Q$.

Since the amount of wire used in an aerial coil secondary is so much less than in the i-f transformers it is often possible to use more expensive litz.

Probably the most satisfactory method of coupling the aerial to the grid of the first valve (or the plate of the first valve to the grid of the second) is by means of a high impedance primary. This can take the form of a universal winding of the finest wire the machines will conveniently handle, wound at an appropriate distance from the secondary to give about 20% coupling (depending on the $Q$ of the secondary, the amount of mistracking that can be tolerated, and the types of aerial expected to be used). In the case of progressive universal secondaries it sometimes happens that the primary must be wound as closely as possible to the secondary to obtain the required degree of coupling.

The primary should be resonated by its own self-capacitance, or by an additional capacitor, below the intermediate frequency so that no normal aerial can tune the primary to the intermediate frequency and so cause instability. A typical value of primary inductance is 1 mH and a 150 $\mu\mu$F capacitor would be suitable for use with it.

To increase the aerial coil gain at the high frequency end of the broadcast band a capacitor of the order of 4 $\mu\mu$F might be used between aerial and grid of the first valve. Since the capacitor decreases image ratio at the same time as it increases sensitivity and signal-to-noise ratio, some requirement other than flat sensitivity across the broadcast band might be desirable.

Similar considerations apply to the coupling of the r-f valve except that primary inductances are usually larger (say 4 mH) and their tuning capacitors may be increased in size until the gain is reduced to the required figure.

The direction of connection for high impedance coils and oscillator coils is the same as that already detailed for i-f transformers; the grid is connected to the end of the secondary further from the primary, to reduce capacitance coupling, and then, considering the secondary and primary to be one tapped coil wound in the one direction, the aerial (or plate) is connected to the other extreme of the winding.

### (ii) Iron-cored coils

The use of suitable iron cores greatly simplifies the problem of obtaining high $Q$ windings. In fact, for any reasonable $Q$ requirement it is only a matter of obtaining a suitable core.

Whilst for a given $Q$ and coil size a coil wound on an iron slug will be less affected by its surroundings than an air-cored coil, owing to the concentration of the flux in the core, high $Q$ secondaries (e.g. 250) will be found to require very careful placement if serious $Q$ losses are to be avoided, particularly at the high frequency end of the broadcast band. Even an open circuited primary at the appropriate distance for correct coupling may noticeably decrease the $Q$.

The primary should be wound on the opposite side of the secondary from the iron core, which will not pass completely through the secondary because of the necessity for leaving room for adjustment. Under these conditions movement of the core to compensate for differences between individual receivers will not seriously affect the coupling between primary and secondary.

To avoid losses due to adjacent components with high $Q$ coils, iron cores can be used which completely enclose the winding. These cores give higher $Q$'s than the more usual slugs with a similar winding and reduce the external field of the coil to such an extent that in extreme cases it may be difficult to obtain sufficient coupling between the secondary and an external high impedance winding.

An advantage of using an iron core in the oscillator coil is that, apart from giving increased $Q$, it can greatly increase the coupling between primary and secondary, thereby decreasing the primary reactance for a given amount of coupling.

## (iii) Permeability tuning

The increasing use of permeability tuning is an indication of the extent to which the design and production difficulties associated with this type of tuning are being overcome.

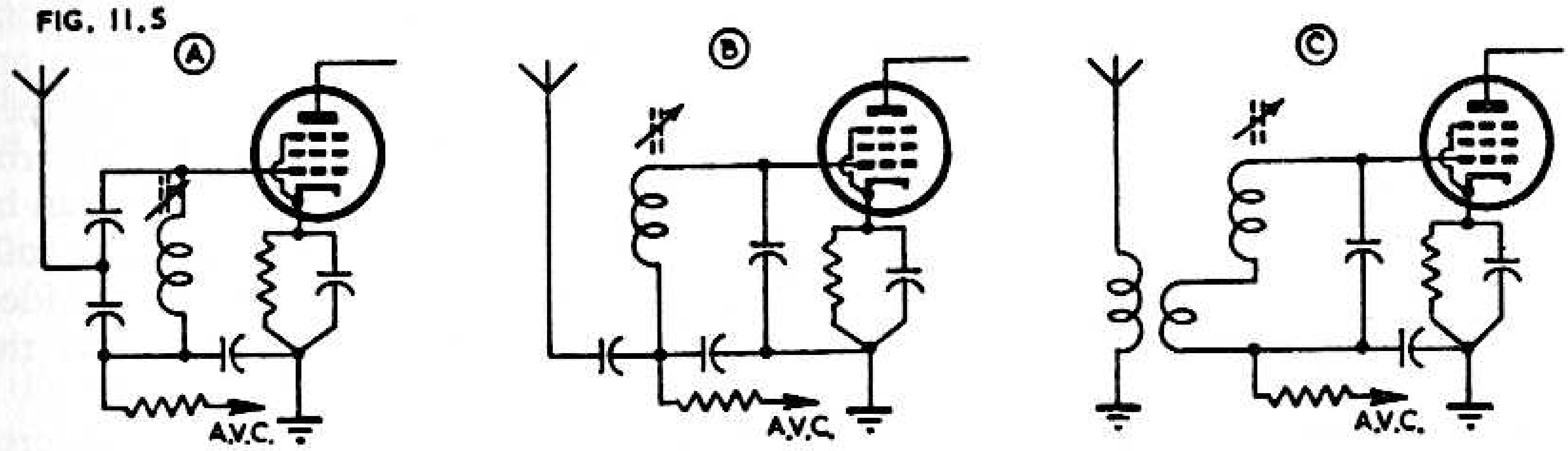


*Fig.* 11.5. *Aerial coupling circuits used with permeability tuning.*

No gang condenser is needed, so that even with the added complication in the coil assembly a cheap broadcast band tuning unit is possible. Gang microphony is automatically eliminated, and a value of tuning capacitor can be chosen which will make the effects of, for instance, capacitance variations due to valve replacement negligible over the whole tuning range. This capacitor can be a low-loss type and it may be given a suitable temperature coefficient if desired. The space saving owing to the elimination of the tuning capacitor is appreciable, even taking into account the additional space required for the iron cores and their driving mechanism.

By a suitable choice of coil and core constants the oscillator grid current can be maintained at or very close to the desired figure at all points in the broadcast band. Aerial coil gain can also be kept constant.

One of the most serious troubles in the design of a permeability tuning unit is in obtaining satisfactory coupling between the aerial and the permeability-tuned aerial coil. If a normal primary is used the coupling variations with changing core position, as the coil is tuned over the broadcast band, are excessive. Similarly a tapped inductor can not be used because the core alters the coupling between the two sections of the winding too much. Fig. 11.5A, B and C show three methods which have been used.

The circuit of Fig. 11.5A couples the aerial straight into the tuned circuit and gives good aerial coil gain when correctly aligned. However image rejection is poor and the detuning effect when different aerials are connected to the receiver is considerable.

Improved image rejection is obtained from the circuit of Fig. 11.5B but sensitivity is lost because much of the voltage developed across the tuned circuit is not applied between grid and cathode of the valve.

Fig. 11.5C gives a more satisfactory solution (Ref. A24). There is no inductive coupling between primary and permeability tuned secondary and the coupling between primary and coupling coil is unaffected by the core position. A constant gain of about three times over the band can be obtained from a coil with average effective $Q$ of 100, with satisfactory image ratios across the band and reasonable freedom from detuning effects with different aerials.

The maintenance of correct calibrations in production receivers is a problem which concerns the designer of the iron dust core, but cores have been produced with permeability consistent enough for station names for the whole broadcast band to be marked on the dial and used.

Iron cores are available with sufficient variation in permeability for the complete broadcast band to be tuned. However if the police band is required and space for mounting the tuning unit is restricted (resulting in appreciable damping on the coil) the permeability requirements for the core are severe.

The Colpitts oscillator circuit is suitable for use in permeability tuning units, and padding can be obtained by using a suitably shaped core for the oscillator coil. Alternatively the oscillator coil can be wound in two series-connected sections, one with fixed inductance (or pre-tuned for alignment purposes) and the other with inductance varied by the moving core. Many receivers have been built with complete tracking i.e. a series inductor corresponding to the usual (parallel) trimmer, and a parallel inductor corresponding to the usual (series) padder.

The rate of change of inductance with core movement does not give a linear relationship between tuning control and the frequency tuned unless special precautions are taken. Linearity, if required, can be obtained in two ways. The drive between tuning spindle and cores can include some type of worm or cam arranged to vary the speed of the drive in suitable parts of the tuning range, or the coils themselves can be progressive universal windings to give the same effect. Progressive universal coils lend themselves very well to this form of winding because the progression is provided by a screw thread and this can be cut with the pitch varied over the length of the winding to give the required tuning rate.

An important drawback to permeability tuning is the difficulty involved in providing coverage of a number of wavebands without complications which outweigh the saving of a tuning condenser.

One simple method of adding short wave ranges to a broadcast band permeability tuner is to connect the broadcast coil in parallel with a short wave coil which is to be tuned. This gives satisfactory tuning but restricted bands. It is suitable for a band-spreading circuit but when every international short wave broadcasting band is to be covered, a large number of ranges is required. By designing a special short wave permeability tuner it is possible to cover the 6 to 18 Mc/s band in two ranges (Ref. A24).

For more restricted ranges such as the F-M broadcasting band, permeability tuning is satisfactory provided that a suitable type of iron is available. Cores have been made which can be used in tuned circuits at 150 Mc/s without introducing undue loss.

References A11, A16, A20, A22, A23, A24, A34.

## (iv) Matching

Methods of coil matching depend on the type of coil and to a certain extent upon the subsequent alignment procedure.

Air-cored solenoids can be wound with a spacing towards the end of the winding equivalent to one turn. The inductance is then increased by sliding turns across the gap from the small end section to the main body of the coil, or decreased by sliding turns in the opposite direction.

Air-cored pie-wound coils can be matched by pushing pies together to increase inductance, but the method has disadvantages. Firstly, the coils are liable to be damaged, even to the extent of collapsing completely, and secondly coils which are rigid enough to withstand the pushing without damage may have enough elasticity to return towards their original positions over a period of time, thus destroying the matching. If coils matched in this way are to be impregnated it is necessary to recheck (and readjust in many cases) after impregnation, and recheck again after flash dipping, if a reasonable degree of matching accuracy is to be maintained.

Another method which has been used with air-cored coils is to wind each coil with slightly too much inductance, bake, impregnate and flash-dip the coil and then seal it in an aluminium can. A groove is then run in the can above the appropriate coil and the depth of the groove is increased until the inductance of the coil inside the can is reduced to that of a standard. All production inductance variations can be absorbed in the matching process with insignificant $Q$ loss.

When iron cores are used, matching problems are much simplified. Even when the core is not threaded it can be pushed through a winding for inductance adjustment. With threaded cores the adjustment is so simple that in many cases the coils are not matched. Given correct calibrations and close tolerance padders in a complete receiver, the oscillator coil trimming and padding can be adjusted to suit the calibration and the aerial coil trimming and padding to suit the oscillator with results which may even be (from the user's point of view) superior to those obtained from more complicated matching and alignment methods.

When matching individual windings on formers containing two or more coils it is possible for errors to occur due to variations in windings other than the one being matched. For example an aerial coil secondary may be matched to the same inductance as a standard secondary but may be coupled to a primary with an inductance different from that used on the standard coil. Under operating conditions in a receiver the two effective secondary inductances would be different.

This difficulty can be overcome by suitable connection of associated coils during matching. Normally the best procedure is to short-circuit and earth coils which are not being matched, but on occasion it may be better to earth the cold end only.

## SECTION 5 : SHORT-WAVE COILS

(*i*) *Design* (*ii*) *Miscellaneous features.*

### (i) Design*

Much has been written on the subject of designing coils suitable for use on short waves. A number of references will be found at the end of this chapter. The work by Pollack, Harris and Siemens, and Barden and Grimes is very complete from the practical design viewpoint. The papers by Butterworth, Palermo and Grover, and Terman are basically theoretical. Austin has provided an excellent summary and practical interpretation of Butterworth's four papers. Medhurst's paper gives the results of measurements which in some cases disagree with Butterworth's theoretical values. For coils whose turns are widely spaced the measurements of high-frequency resistance are in good agreement but for closely-spaced coils the measured values are very considerably below those of Butterworth. Theoretical reasons are given for these differences.

**(A) Pollack** (Ref. G4) summarizes the procedure for the optimum design of coils for frequencies from 4 to 25 Mc/s as follows :—

1. Coil diameter and length of winding : Make as large as is consistent with the shield being used. The shield diameter should be twice the coil diameter, and the ends of the coil should not come within one diameter of the ends of the shield.

2. A bakelite coil form with a shallow groove for the wire, and enamelled wire may be used with little loss in $Q$. The groove should not be any deeper than is necessary to give the requisite rigidity. The use of special coil form constructions and special materials does not appear to be justified (except for the reduction of frequency drift due to temperature changes).

3. Number of turns : Calculate from

$$N = \sqrt{L(102S + 45)/D}$$

where $S$ = ratio of length to diameter of coil,
$D$ = diameter of coil in centimetres,
and $L$ = inductance in microhenrys.

4. Wire size : Calculate from

$$d_0 = b/N\sqrt{2} = \text{optimum diameter in centimetres.}$$

where $b$ = winding length in centimetres.

*This is a revision and expansion of the chapter on this subject in the previous edition by L. G. Dobbie.

That is, the optimum wire diameter is $1/\sqrt{2}$ times the winding pitch, measured from centre to centre of adjacent turns.

(B) **Barden and Grimes** (Ref. G16) recommend for coils working near 15 Mc/s. that No. 14 or No. 16 A.W.G. enamelled wire on a form not less than one inch diameter at a winding pitch equal to twice the wire diameter is desirable. The screen diameter should be not less than twice the coil diameter. A comparison of coils of equal inductance on 0.5 in. and 1 in. forms in screens double the coil diameter indicates that the value of $Q$ is twice as great for the larger diameter coil. No. 24 A.W.G. wire was used for the small diameter coil.

(C) **Harris and Siemens** (Ref. G21) quote the following conclusions :—

(1) $Q$ increases with coil diameter.

(2) $Q$ increases with coil length, rapidly when the ratio of length to diameter is small, and very slowly when the length is equal to or greater than the diameter.

(3) Optimum ratio of wire diameter to pitch is approximately 0.6 for any coil shape. Variation of $Q$ with wire diameter is small in the vicinity of the optimum ratio ; hence, selection of the nearest standard gauge is satisfactory for practical purposes.

(D) **The shape of a coil necessary for minimum copper loss** (from Butterworth's paper) is stated by **Austin** (Ref. D7) as follows :—

(1) Single layer solenoids : Winding length equal to one-third of the diameter.

(2) Single layer discs (pancake) : Winding depth equal to one-quarter of the external diameter.

(E) Butterworth's paper (Ref. D1) deals with the copper loss resistance only, and insulation losses must be taken into account separately. **Insulation losses** are minimized by winding coils on low loss forms, using a form or shape factor which provides the smallest possible self-capacitance with the lowest losses. Thus air is the best separating medium for the individual turns, and the form should provide only the very minimum of mechanical support. Multilayer windings in one pie have high self-capacitance due to proximity of the high and low potential ends of the winding. The same inductance obtained by several pies close together in series greatly reduces the self-capacitance and associated insulation losses. Heavy coatings of poor quality wax of high dielectric constant may introduce considerable losses.

(F) In the section of **Medhurst's paper** (Ref. D6) dealing with h.f. resistance he states that so long as the wire diameter is less than one half of the distance between wire centres, Butterworth's values are applicable. An increasing error occurs as the ratio of wire diameter to distance between centres increases, the values being 190% too high when the ratio is 0.9.

A graph is presented giving variation in optimum wire spacing with variation in the ratio of coil length to coil diameter (Fig. 11.6A).

It is shown that a good approximation to the high frequency $Q$ of coils of the type measured is given by the simple expression

$$Q = 0.15R\psi\sqrt{f}$$

where $R$ = mean radius of coil (cm),

$f$ = frequency (c/s)

and $\psi$ depends on the length/diameter and spacing ratios.

A comprehensive table of $\psi$ values is presented.

(G) **M. V. Callender** (Ref. D9) points out that Medhurst's formula for $Q$ can be approximated within a few per cent by the equation

$$Q = \sqrt{f}/(6.9/R + 5.4/l)$$

where $R$ = radius of coil (cm)

and $l$ = length of coil (cm).

An even simpler expression

$$Q = 0.15\sqrt{f}/(1/R + 1/l)$$

follows the data to a few per cent provided $l > R$.

The range of conditions under which this formula applies is the same as that to which Medhurst's data refer : in particular—

(a) The ratio of wire diameter to wire spacing must approximate to the optimum shown in Fig. 11.6A.

(b) The formulae apply only for very high frequencies because of skin effect considerations. However the following table, giving the thinnest wire for which the formula applies within ± 10%, shows that most practical solenoids will be covered—

| | | | |
|---|---|---|---|
| $f$ (Mc/s) | 1.0 | 4.0 | 16 |
| wire | 22 S.W.G. | 28 S.W.G. | 37 S.W.G. |

(c) The formulae do not hold for coils of very few turns (or extremely short coils).

(d) Dielectric loss is not allowed for. This is unlikely to be serious except where the coil has a rather poor dielectric (bakelite or worse) and is used in a circuit having a low parallel tuning capacitance.

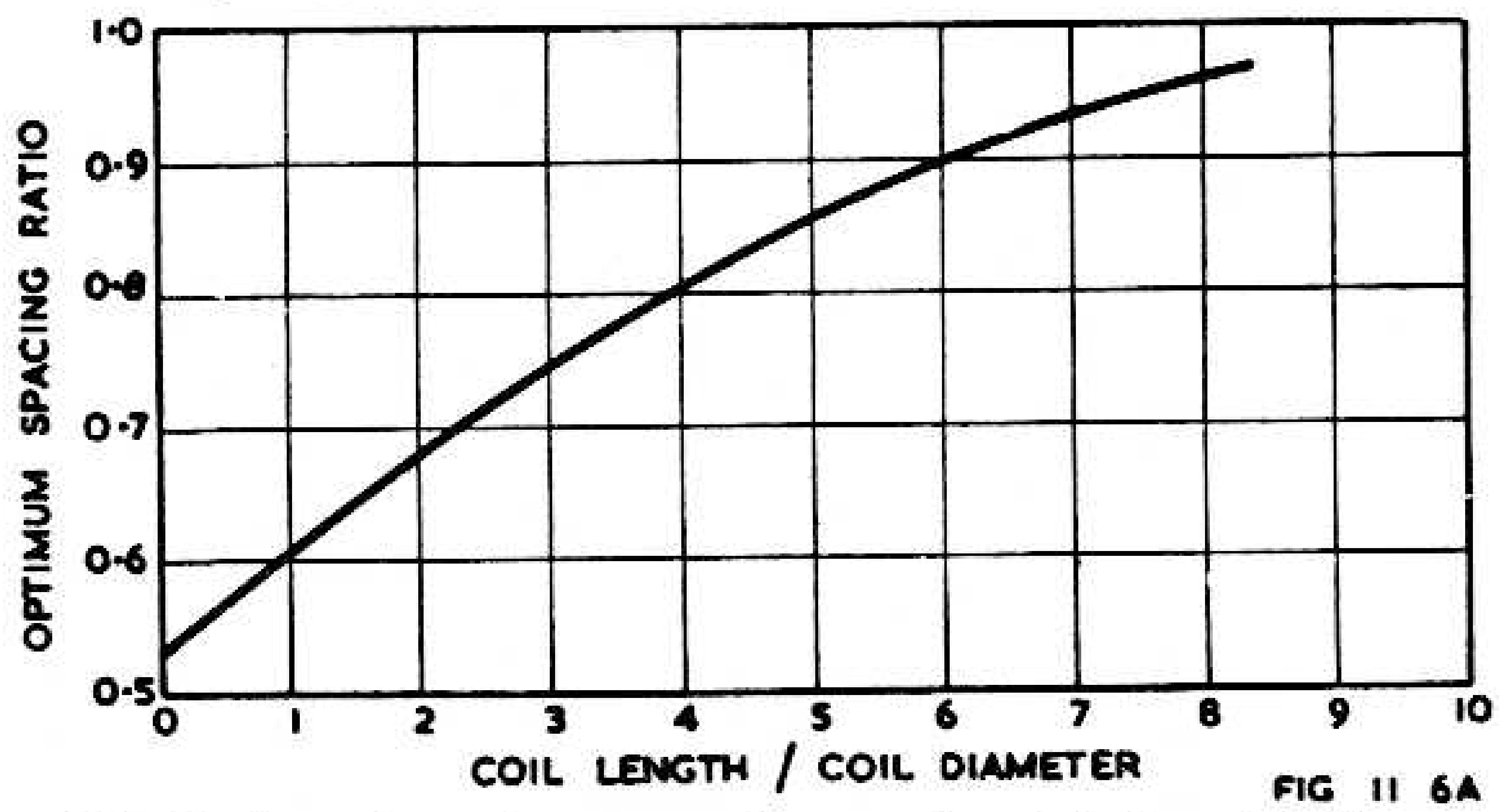


*Fig.* 11.6A. *Variation of optimum spacing ratio with length/diameter* (*Ref.* D6).

(H) **Meyerson** (Refs. G3 and G11) gives the results of measurements on a large number of coils tuned over a range between 25 and 60 Mc/s. The following information is obtained from his work :—

(a) $Q$ at any frequency within the band, and the frequency for maximum $Q$, both increase with an increase in wire size for a given coil diameter. 10 A.W.G. wire (the largest used) was nearly 10% better over the range than 12 A.W.G. wire on a one inch diameter former.

(b) Maximum $Q$ increases, and the frequency for maximum $Q$ decreases, with an increase in coil diameter for a given wire size, number of turns and number of turns per inch.

(c) The coil diameter required for highest $Q$ throughout the tuning range decreases with increasing frequency for a given wire size, number of turns and number of turns per inch.

(d) Maximum $Q$ is obtained with a spacing between adjacent wires which is slightly greater than the bare wire diameter.

(e) No variation in $Q$ was detected between coils wound with bare wire, enamelled wire or silver plated wire.

On the other hand, cotton covering decreased $Q$ by as much as 5% at 50 Mc/s and annealing of the copper wire increased $Q$ by less than one per cent.

(I) In Refs. G10 and G12 **Meyerson** gives details of work with frequency ranges between 60 and 120 Mc/s on the $Q$ and frequency stability of inductors made of wire, tubing and strip in various shapes e.g. disc, hairpin, folded lines, etc.

(J) Other points taken from various references are given below. **Dielectric losses** present in the self capacitance of the coil are reduced by altering the shape to separate the high potential end from all low potential parts of the circuit. These losses become relatively more important the higher the frequency. Thus, the shape of a solenoid for minimum total losses may need to be increased beyond what would otherwise be the optimum length.

The high frequency alternating field of a coil produces **eddy currents** in the metal of the wire, which are superimposed upon the desired flow of current. The first effect is for the current to concentrate at the outside surface of the conductor, leaving the interior relatively idle. In a coil where there are numbers of adjacent turns carrying current, each has a further influence upon its neighbour.

In turns near the centre of a solenoid the current concentrates on the surface of each turn where it is in contact with the form, i.e. at the minimum diameter. In turns at either end of a solenoid the maximum current density occurs near the minimum diameter of the conductor, but is displaced away from the centre of the coil.

Thus most of the conductor is going to waste. Multi-strand or litz (litzendraht) wires have been developed to meet this difficulty. Several strands (5, 7, 9, 15 being common) are woven together, each being of small cross section and completely insulated by enamel and silk covering from its neighbours. Owing to the weaving of the strands, each wire carries a nearly similar share of the total current, which is now forced to flow through a larger effective cross section of copper. The former tendency towards concentration at one side of a solid conductor is decreased and the copper losses are correspondingly reduced.

Litz wire is most effective at frequencies between 0.3 and 3 Mc/s. Outside of this range comparable results are usually possible with round wire of solid section, because at low frequencies " skin effect" steadily disappears while at high frequencies it is large even in the fine strands forming the litz wire, and is augmented by the use of strands having increased diameter.

Screens placed around coils of all types at radio frequencies should be of non-magnetic good-conducting material to introduce the least losses. In other words, the $Q$ of the screen considered as a single turn coil should be as high as possible. In addition, the coupling to the coil inside it should be low to minimize the screen losses reflected into the tuned circuit. For this reason the screen diameter should, if possible, be at least double the outside diameter of the coil. A ratio smaller than 1.6 to 1 causes a large increase in losses due to the presence of the screen.

The design of coils for use with iron core materials depends mainly upon the type of core material and the shape of the magnetic circuit proposed. Nearly closed core systems are sometimes used with high permeability low loss material. More commonly, however, the core is in the form of a small cylindrical plug which may be moved by screw action along the axis of the coil and fills the space within the inside diameter of the form. The main function of the core in the latter case may be only to provide a means of tuning the circuit rather than of improving its $Q$. When improvement in $Q$ is possible with a suitable material, the maximum benefit is obtained by ensuring that the largest possible percentage of the total magnetic flux links with the core over as much of its path as possible ; the ultimate limit in this direction is of course the closed core.

An excellent series of charts for the design of single-layer solenoids for a required inductance and $Q$ is presented by A. I. Forbes Simpson in Ref. G17. These charts and the instructions for their use are reprinted by permission of " Electronic Engineering " and the author (Figs. 11.7 to 11.13 inclusive).

The use of the charts is simple and gives direct answers. As so many unascertainable factors govern the final inductance of a coil in position, no attempt has been made to achieve an accuracy better than 1 per cent. The inductances indicated by the charts assume that the leads to the coil are of the same wire as the coil and perpendicular to it, and are each a coil diameter long.

Coils of this form have a low self-capacitance which is largely independent of all save the coil diameter and to a lesser extent the wire gauge.

The capacitance may be expressed as

$$C = D(0.47 + a)\ \mu\mu\mathrm{F}$$

where $D$ is coil former diameter in cms. and $a$ is a constant depending on the gauge of wire, lying between 0.065 for 42 S.W.G. and 0.11 for 12 S.W.G. wire, where one lead of the coil is connected to chassis.

**Use of charts**

The charts shown may be used as follows :

If the desired $Q$ is known, then reference to Fig. 11.8 will suggest a suitable diameter of former.

If, however, as great a $Q$ as possible is required, then the largest possible former should be used unless the coil is to be sited in a can, in which case the diameter of the former should be less than one half of the internal diameter of the can and spaced from the end by the same amount or more.

When the diameter of the former is determined, reference should be made to the appropriate chart, and as shown in Fig. 11.7 the inductance intercept $A$ corresponds to $C$ which gives the wire gauge ($D$) which gives the number of turns per inch for that gauge, and $E$ the number of turns required.

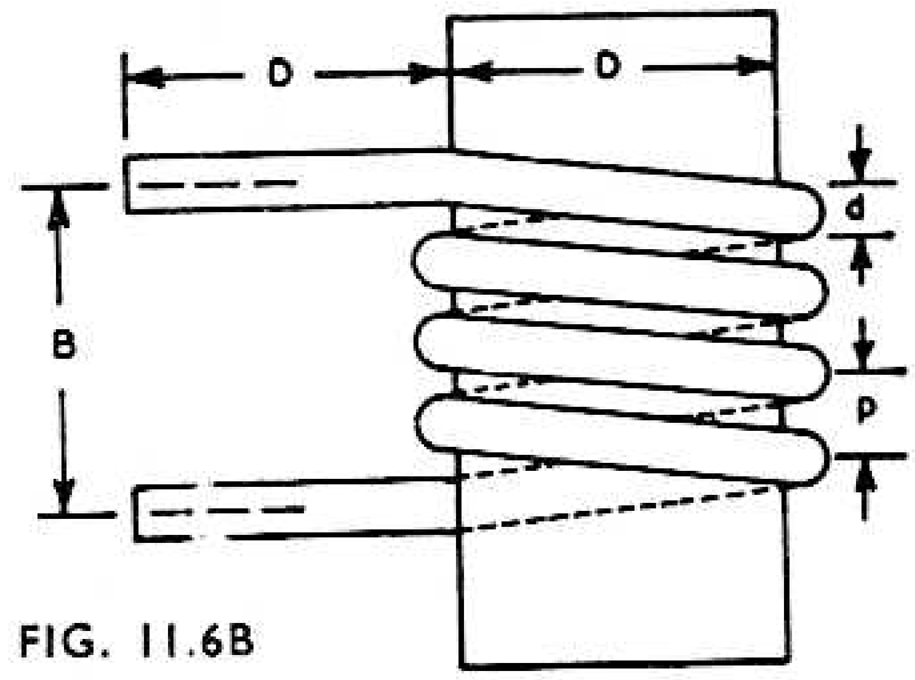


FIG. 11.6B

*Fig.* 11.6B. *Dimensions of coil referred to in text.*

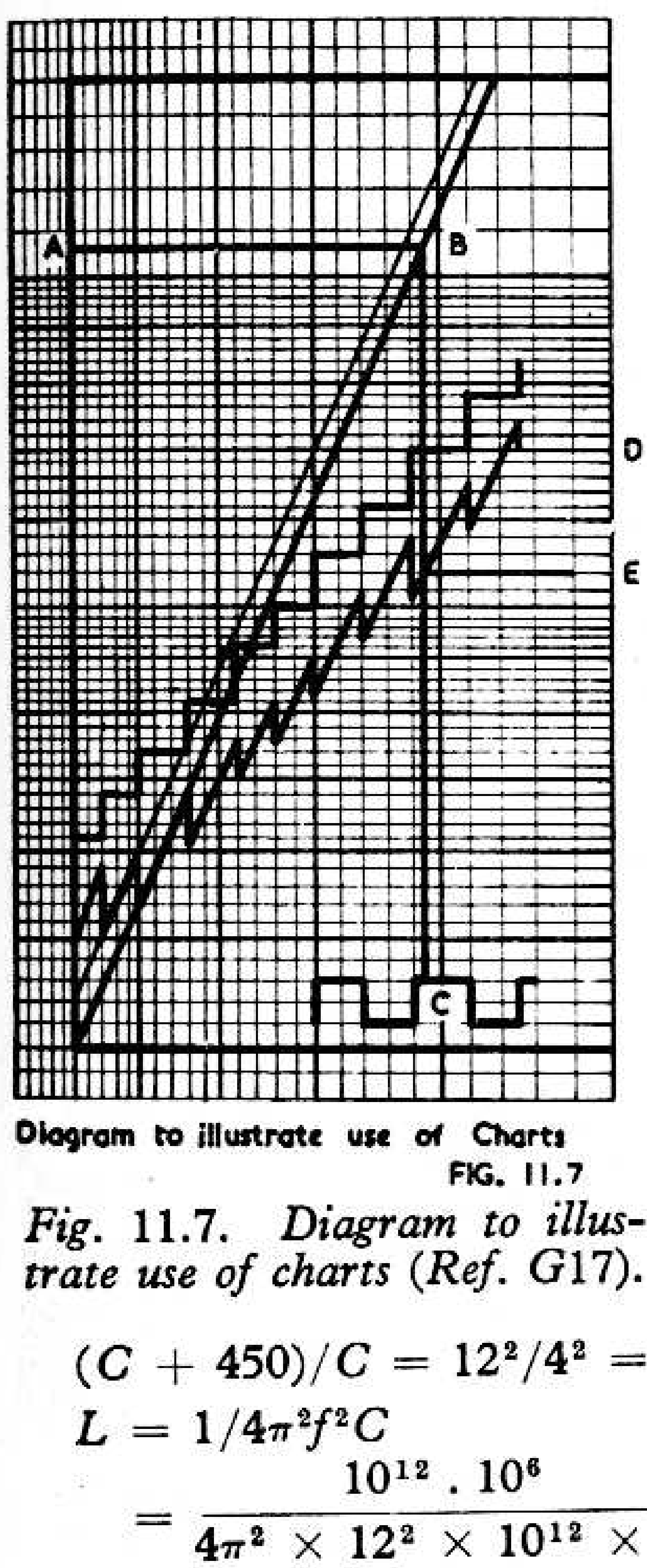


Diagram to illustrate use of Charts
FIG. 11.7

*Fig.* 11.7. *Diagram to illustrate use of charts* (*Ref.* G17).

It will be seen that " even " gauges of wire have been used. If it is desired to use an " odd " gauge, a little simple geometry will soon give the result.

The $Q$ in air of a coil wound to this information, when tuned by capacitances between 45 and 500 $\mu\mu$F may be read directly from Fig. 11.8.

The $Q$ values shown in the chart have been corrected for coil capacitance and are somewhat higher than those indicated by the usual $Q$ meter. The accuracy of these curves is rather less than that of the inductance charts, but various values have been checked by several methods which have given substantial agreement.

**Example**

As an example* of the use of the curves, suppose we have a variable condenser of 450 $\mu\mu$F swing and desire to find a coil with which it will tune from 12 Mc/s to 4 Mc/s. If the total capacitance at 12 Mc/s. be $C$ $\mu\mu$F then

$$(C + 450)/C = 12^2/4^2 = 9 \text{ and } C = 56.25\ \mu\mu\text{F}$$

$$L = 1/4\pi^2 f^2 C$$

$$= \frac{10^{12} \,.\, 10^6}{4\pi^2 \times 12^2 \times 10^{12} \times 56.25}\ \mu\text{H} = 3.15\ \mu\text{H}.$$

Inspecting Fig. 11.8 we see that the $Q$ at 4 Mc/s. will lie between 116 for $\frac{1}{2}$ in. and 168 for a $\frac{7}{8}$ in. former, while that at 12Mc/s. will lie between 182 for a $\frac{1}{2}$ in. and 265 for $\frac{7}{8}$ in. former.

If we choose a $\frac{5}{8}$ in. former we find that the line corresponding to this in Fig. 11.10 cuts the $L$ line at a wire diameter of 0.025 in. This line intersects T.P.I. at 23.8 and the " No. of turns " is 18.5 with a wire gauge of 22 S.W.G.

Referring to Fig. 11.8—we find for $L = 3.15\ \mu$H :

| | | | | | |
|---|---|---|---|---|---|
| $C$ $\mu\mu$F | 45 | 100 | 200 | 400 | 500 |
| $Q$ | 212 | 195 | 175 | 150 | 142 |

$Z_o$ is readily obtained by writing $Z_o = Q\sqrt{L/C}$.

*This example is presented in a different form from that in the original article.

If, however, we had used the ½ in. former we would find that 18 turns of 26 S.W.G. at 37 T.P.I. would give

| $C$ $\mu\mu$F | 45 | 100 | 200 | 400 | 500 |
|---|---|---|---|---|---|
| $Q$ | 181 | 165 | 147 | 123 | 117 |

## (ii) Miscellaneous features

### (A) Matching

Short wave coils are frequently matched by altering the position of the last few turns on the former. This sometimes leads to damage to the coil, particularly if there is an interwound primary.

An alternative method, which is applicable to coils terminated by passing the wire diametrically through the former, is to push sideways this terminating wire in the

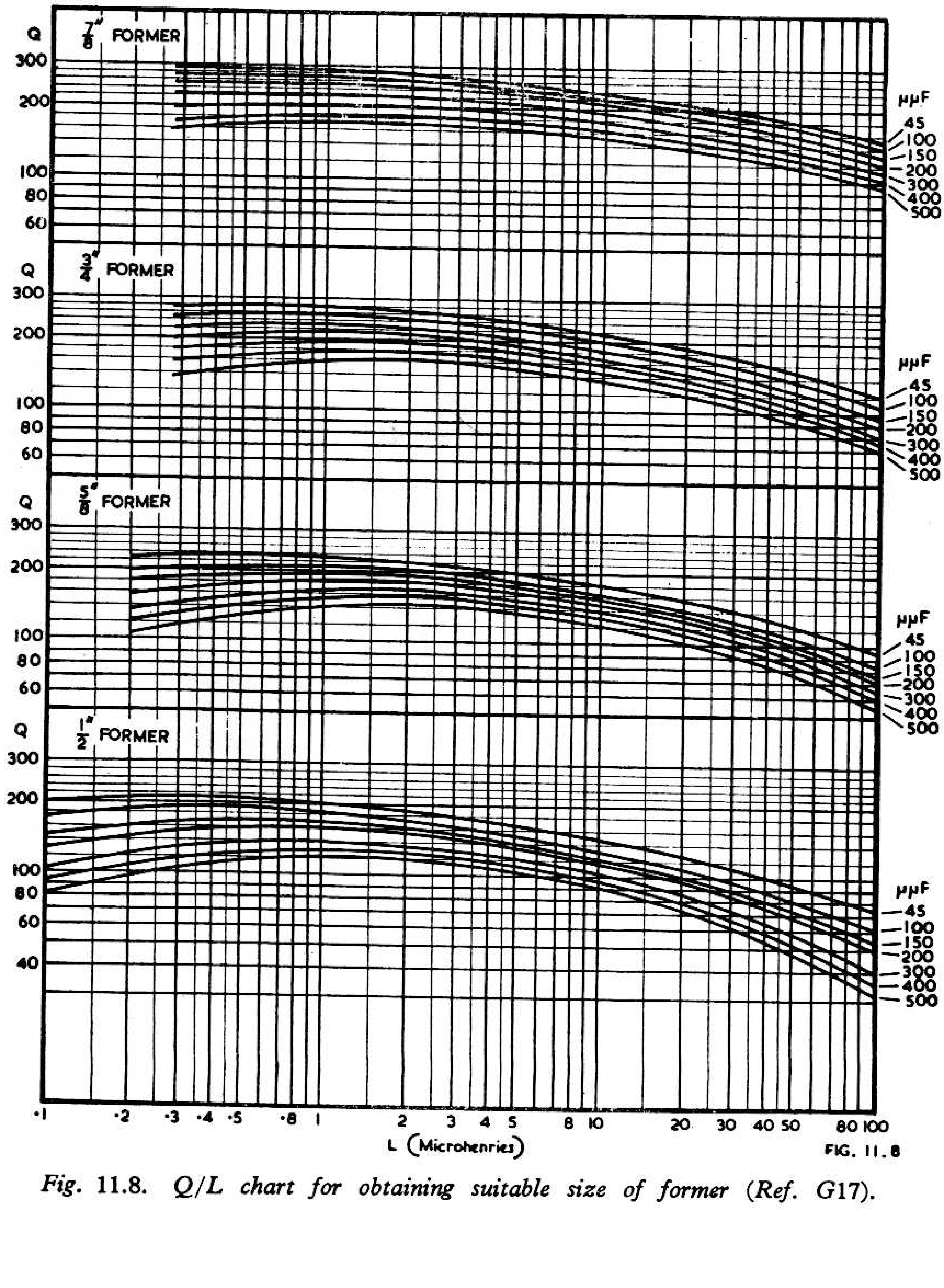


*Fig.* 11.8. *Q/L chart for obtaining suitable size of former (Ref.* G17).

middle of the coil. In this way the last half turn of the coil can be increased almost to a complete turn or decreased to a small fraction of a turn. Both the first and last half turns on the coil can be treated in this way, thereby providing a possible inductance adjustment equivalent to approximately one complete turn of the winding. When matching short wave coils it is essential that associated primaries, and any other coils on the same former, be connected in a similar manner to their ultimate connection in the receiver, i.e. earthed ends of coils should be earthed, and hot ends either earthed or left open depending on which is more suitable.

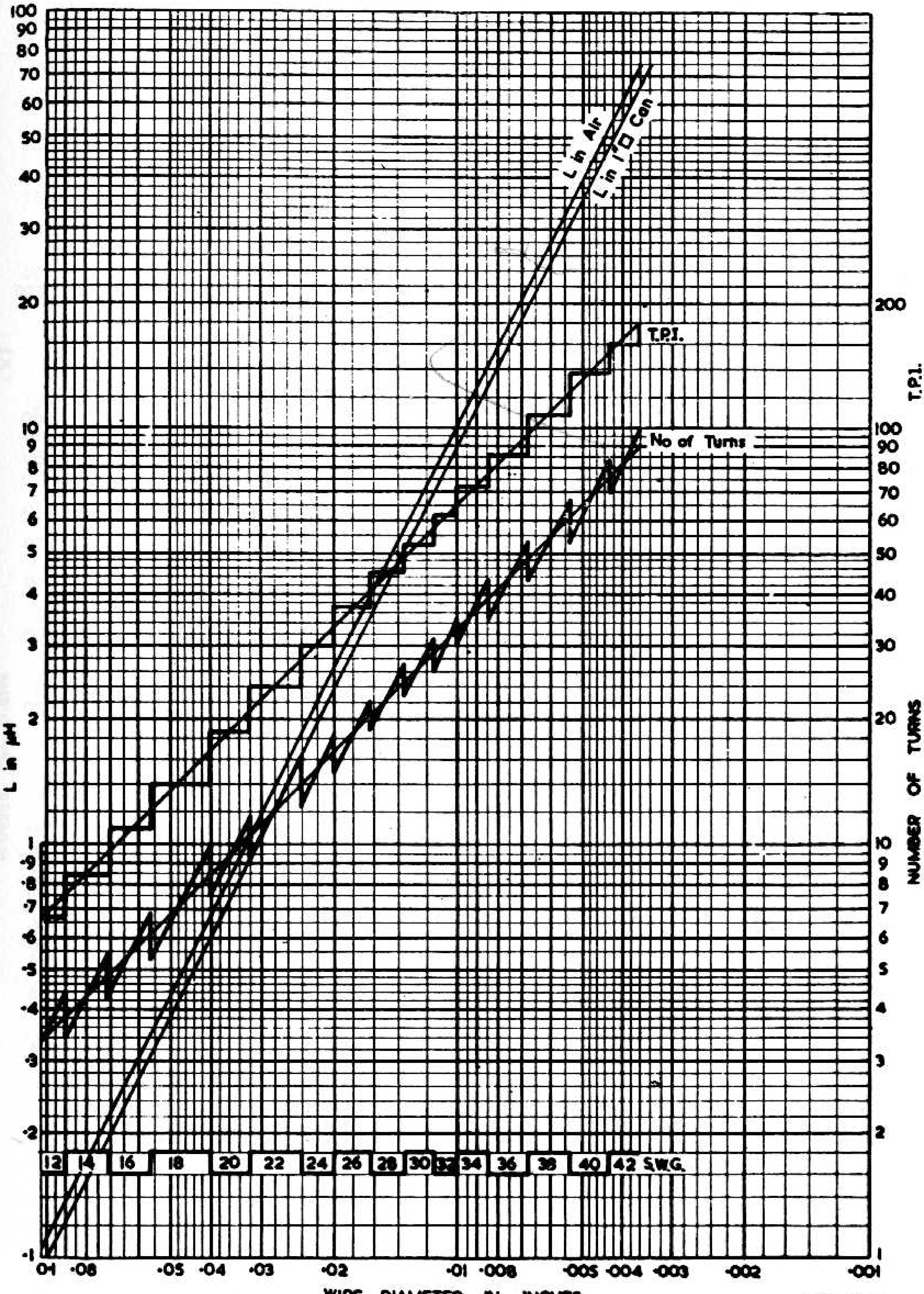


FIG. 11.9

*Fig.* 11.9. *Optimum* Q, ½ *in. former* (*Ref.* G17).

Iron cores are widely used for adjusting the inductance of short wave coils. Because at higher frequencies increased losses in cores offset the advantage of increased permeability, the cores in some cases do not increase the $Q$ of a coil, but provide a convenient method of inductance adjustment.

In shortwave oscillator coils an iron core also has the advantage of increasing the coupling between primary and secondary for a given winding. Thus increased grid current may be obtained although $Q$ is not increased.

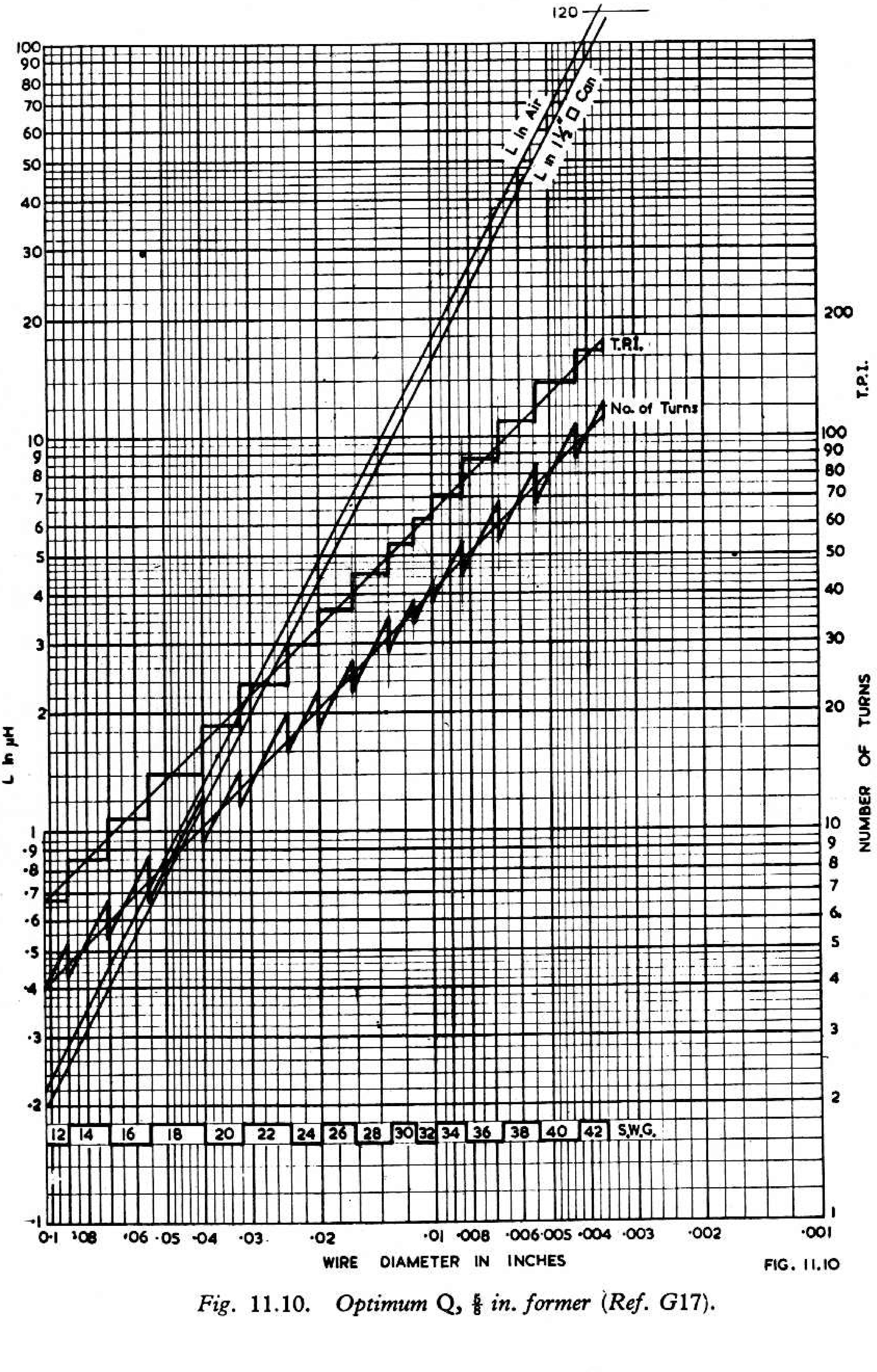


*Fig.* 11.10. *Optimum* Q, ⅝ *in. former* (*Ref.* G17).

At frequencies of the order of 100 Mc/s the difficulties of manufacturing suitable iron cores are greater although they are made and used (Ref. A22) but it is comparatively simple to wind air cored coils with a *Q* of say 300. Because of this, inductance adjustment can also be carried out by means of copper slugs which are adjusted in the field of the coil in the same way as are iron cores at lower frequencies. The slugs are sometimes silver plated to minimize losses, but on the other hand brass slugs are also used, apparently without undue losses.

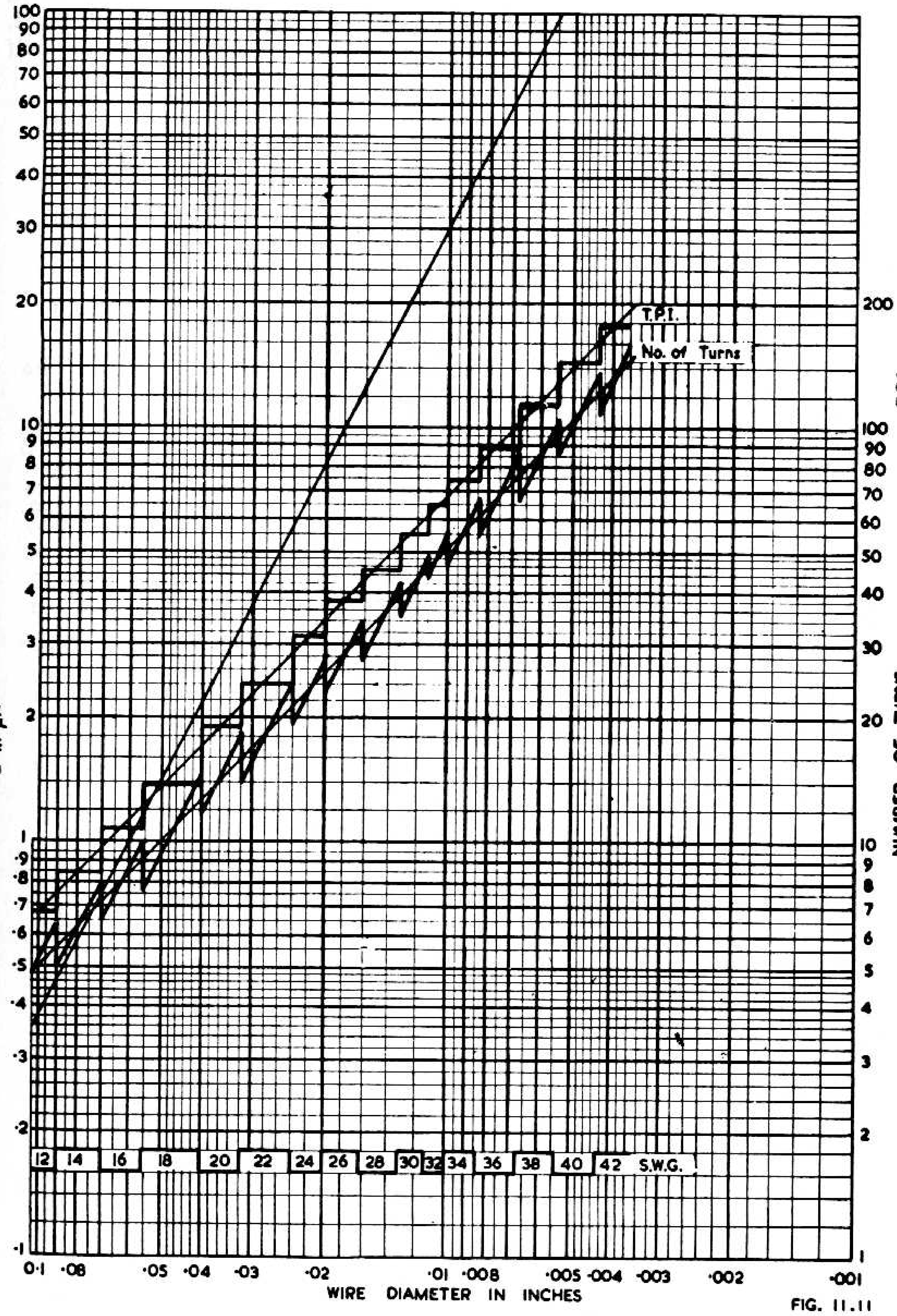


*Fig.* 11.11. *Optimum* Q, ¾ *in. former* (*Ref.* G17).

**(B) Self-capacitance**

To reduce capacitances in a coil with an interwound primary winding without noticeably decreasing the coupling, it is possible—if the spacing factor of the secondary winding allows it—to wind the primary against the wire on the low potential side of each space between secondary wires. A definite advantage is gained from this method of winding and at the same time the cause of appreciable random deviations in coil capacitance is removed.

It is usually advisable to start an interwound primary winding just outside a tuned secondary winding for a desired coupling with minimum secondary capacitance. Best results will probably be obtained with the primary started between one half

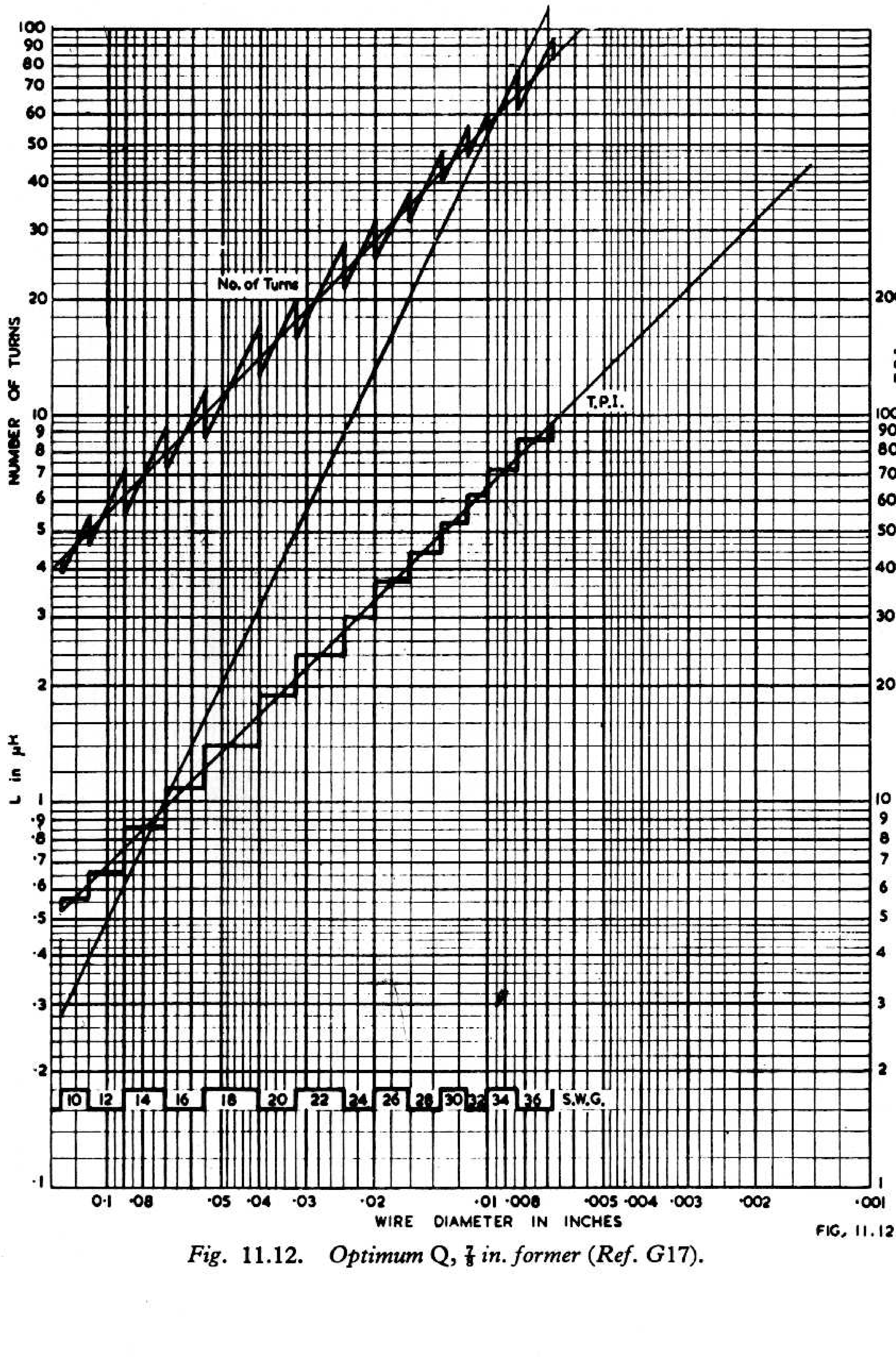


*Fig.* 11.12. *Optimum* Q, $\frac{7}{8}$ *in. former* (*Ref.* G17).

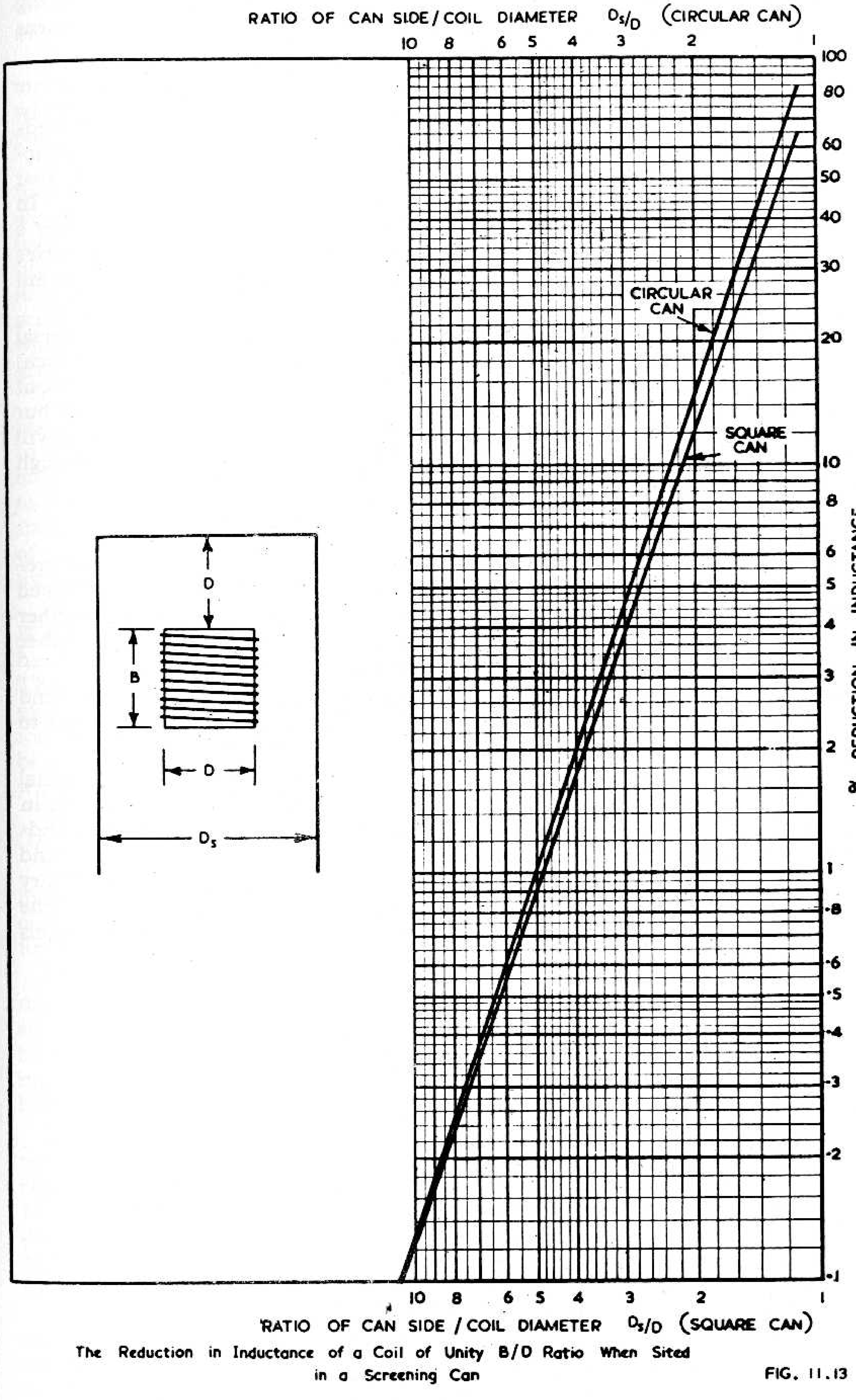


*Fig.* 11.13. *Effect of circular and square screening cans on inductance (Ref. G17).*

and one and one half turns outside the cold end of the secondary, close spacing being used for parts of the primary outside the secondary and the external primary turns being wound as close to the last secondary turn as possible

**(C) Aerial primary windings**

On the 6 to 18 Mc/s short wave bands both interwound and external high impedance primaries are used. External primaries have the advantage of reducing the capacitive coupling between the aerial and the grid of the first valve with the result that oscillator re-radiation from superheterodyne receivers is noticeably reduced. In addition, impulse interference is apparently decreased. The gain obtainable over a band of frequencies is at least as good as that with low impedance interwound primaries and the tracking troubles experienced when a particular aerial resonates an interwound primary within the band are not so likely to occur.

A satisfactory method of winding for an external primary is to use a universal winding (perhaps 36 or 38 A.W.G. S.S.C.) with a small cam (1/10 inch or 1/16 inch) and space the primary as closely as possible to the cold end of the secondary. About 25 turns on a 3/8 inch former would be suitable for the range from 6 to 18 Mc/s but final details must be determined in a receiver. Additional capacitance coupling will sometimes provide increased gain at the high frequency end of the band, although with the loss of some of the advantages outlined above.

**(D) Direction of windings**

For interwound oscillator coils the method of connection is the same as that previously described—considering primary and secondary as one continuous tapped winding in the one direction, the grid is connected to the end of the secondary further from the primary (for a tuned grid oscillator) and the plate is connected to the other extreme, i.e. the end of the primary further from the grid connection.

Interwound aerial and r-f coils are connected in the opposite sense, i.e. grid to end of secondary further from primary, and plate (or aerial) to end of primary closer to grid.

External primaries are connected in the same sense as oscillator coils i.e. aerial and grid at extremes of the two windings when considered as one tapped winding in the same direction. Care is needed with short wave coil connections because secondary and external primary are wound on different types of machines and may be wound in opposite directions. This necessitates reversing the connections to the primary winding. Examination of a coil will show that if the primary is moved from one end of the secondary to the other it will also be necessary to reverse the connections to the primary.

## SECTION 6 : RADIO-FREQUENCY CHOKES

(*i*) *Pie-wound chokes* (*ii*) *Other types.*

### (i) Pie-wound chokes

Radio frequency chokes are of two main types, those used above their main self-resonant frequency and those used below it. In either case the choke is normally required to have, over a range of frequencies, an impedance higher than some minimum value and a high reactance which does not have sudden changes in its value. The need for high impedance is obvious, and a constant value of reactance is necessary if the choke is in parallel with a ganged tuned circuit.

At frequencies below its natural resonance the impedance of a choke is due to its inductive reactance, modified by the self-capacitance as shown in Sect. 2(i) of this chapter, and so increases with frequency until self-resonance is reached. The choke may have one or more resonances, series as well as parallel, as explained below, but above some high frequency its impedance is approximately equal to the reactance of its self-capacitance, i.e. the impedance is inversely proportional to frequency and eventually reaches values low enough to render the choke useless.

Thus to obtain high impedance over a wide band of frequencies a choke requires maximum inductance and minimum self-capacitance. For lower frequencies universal windings are used, usually split into two or more spaced pies with each pie as thin as possible for minimum self-capacitance.

To keep the impedance high over a band of frequencies it is necessary to avoid any tuning of sections of the choke by other sections. A typical case of such tuning would occur in a r-f-c consisting of two pies of different sizes. The impedance-frequency curve of such a choke would show two maxima at the self-resonant frequencies of the two sections, but between these peaks there would be a serious drop in impedance at the frequency at which the inductive reactance of one pie became series resonant with the capacitive reactance of the other. At such a frequency there would also be a sudden change in the effective reactance of the choke which would give tracking errors.

Series resonance of the type described occurs whenever pies with different self-resonant frequencies are connected together. To eliminate them it is not sufficient to wind several pies with the same numbers of turns because the effects of mutual inductance in different sections of the coil still give different self-resonant frequencies to adjacent pies. To avoid this, Miller (Ref. H6) recommends winding progressively smaller pies towards the centre of a multi-pie r-f-c to make the self resonant frequency of each pie the same. He gives figures which show a considerable increase in uniformity of impedance over a range of frequencies when this method is adopted.

In addition to the need for individual pies to be resonant at the same frequency, Wheeler (Ref. H5) gives the requirement (for two-pie chokes) that the pies should be connected so that mutual inductance opposes the self-inductance of the pies. This almost entirely removes minor resonances which occur at approximately harmonically related frequencies, and gives a very smooth curve of apparent capacitance for the r-f-c. The spacing between pies and the number of turns on the pies should be experimentally adjusted for best results with the choke located in the position in which it is to be used.

## (ii) Other types

At high frequencies (above 10 Mc/s) self resonant chokes can be wound to give a high impedance over a reasonably wide band of frequencies. The winding length of such chokes for a given former diameter and wire size can be obtained from Ref. H1 for frequencies between 10 and 100 Mc/s.

In other cases it is necessary to design chokes for maximum impedance, i.e. maximum inductance and minimum self-capacitance with a minimum of resistance. Such an application occurs in the battery input circuit of a vibrator unit. Solenoids have been used for this purpose but higher impedance can be obtained with pyramid winding (say eight turns on the first layer, seven on the second and finishing with a top layer of one turn).

Even better results are obtainable by using an iron core as a former, although the winding must be arranged in such a way that the iron core does not increase the self-capacitance unduly.

Another application requiring a minimum-resistance r-f-c is a mains filter. Universal windings of, say, 26 A.W.G. wire are used in some cases, but where better performance is required a wooden former of perhaps 2 inches diameter is used, and slots in the former are filled with jumble windings of heavy wire.

## SECTION 7 : TROPIC PROOFING

(*i*) *General considerations* (*ii*) *Baking* (*iii*) *Impregnation* (*iv*) *Flash dipping* (*v*) *Materials.*

### (i) General considerations

The use of a suitable wax or varnish is not the only way in which the moisture resisting properties of a coil can be improved. The higher the impedance of a circuit, the greater the effect of a certain resistance in parallel with it. Thus a leakage between coil terminals (or other appropriate parts of the circuit) due to moisture will have a greater effect on a high $Q$ circuit than on a low $Q$ one. Depending upon the importance attached to maintenance of performance in humid conditions, it may be desirable to use low impedance (low $Q$ and low $L$) circuits even at the expense of an additional stage in a receiver.

Another aspect of the question is that the moisture absorption of a coil is governed by its surface area. High $Q$ coils usually have a larger surface (due to the use of pie-winding for instance) and thus are more liable to deteriorate rapidly under humid conditions.

This has been demonstrated in a test in which two sets of coils were used. The first set consisted of 1 mH coils wound in a single pie with a 3/8 in. cam, giving an initial $Q$ of 75, the second set of 1 mH coils were wound with three pies and a 1/10 in. cam giving a $Q$ of 145. Both sets of coils were baked, impregnated and flash dipped and then exposed to conditions of high humidity. At the end of one week the single pie windings had an average $Q$ of 65, and the three pie windings an average $Q$ of about 60. As the test continued the improved performance of the single pie windings became even more marked. In addition the variations between individual coils were much less in the single pie windings, which may have been due to the more variable nature of the flash-dip on the three-pie windings.

No coating or impregnating treatment for coils prevents moisture penetration if the exposure to high humidity is prolonged. On the other hand, some treatments will delay the ingress of moisture much longer than others—for instance the $Q$ of a coil with a single flash-dip will fall much more rapidly than that of a coil which has been vacuum impregnated and flash-dipped twice. The corollary should not be overlooked ; after prolonged exposure to humidity the coil with a single flash dip will return to its original performance when the humidity is decreased more rapidly than will the other. Conditions of use determine requirements, but it cannot be taken for granted that the treatment which gives the slowest deterioration in $Q$ is the best for all types of operation.

Apart from maintaining $Q$, moisture proofing is also required to maintain insulation resistance between different windings or between windings and earth. This is particularly important in the case of interwound coils (e.g. an oscillator coil) in which one winding is earthed and the other is of fine wire with a d.c. potential applied to it. Leakage between the two windings in the presence of moisture can produce electrolysis and a consequent open-circuit in the fine wire.

### (ii) Baking

Irrespective of the moisture proofing treatment which is to be applied to a coil, the first requirement is to remove all moisture. This is done most conveniently by baking at a temperature above the boiling point of water. Materials used in coil winding are liable to be damaged at temperatures greatly in excess of 100°C and a satisfactory oven temperature is from 105° to 110°C.

The time required depends on the components being treated, and it is essential for all parts of a coil to be raised to oven temperature and maintained at that temperature at least for a short time. Baking for a quarter of an hour is a minimum for simple coils, and half an hour is a more satisfactory time.

One detail of oven design which must not be overlooked is that provision must be made for removing the water vapour as it is expelled from the coils. A current of fresh air passing through the oven is satisfactory for this purpose.

Whether the baking is followed by impregnation or flash-dipping only, the next process should begin before the coils have cooled.

### (iii) Impregnation

Although vacuum impregnation gives the best penetration of the impregnant and best removal of moisture vapour and gas, a " soaking " treatment is often used and gives satisfactory results for commercial requirements.

A typical soaking treatment calls for the coils to be immersed in wax which is maintained at 105° to 110°C for a period of from one quarter to one half hour, depending on the type of winding, e.g. a thick universal winding needs more time than a spaced solenoid. In any case the coil should be left in the wax until no more bubbles are given off.

The temperature of the wax-tank needs accurate adjustment as apart from damage to coils which may result if it is too high, the characteristics of the wax may be altered by excessive heating.

Soaking in varnish (air-drying or oven-drying) is also possible but it is necessary to keep a close check on the viscosity.

### (iv) Flash-dipping

Flash-dipping is commonly practised without impregnation and properly carried out it can be a satisfactory commercial treatment. The flash-dip should be of the same type of material as normally used for impregnation.

A wax-tank for flash-dipping can conveniently be maintained at a lower temperature than one for soaking, and a suitable temperature is one at which a skin of wax just develops on the surface of the tank on exposure to air. By keeping the top of the tank covered when not in use the skin will not be troublesome and it can readily be cleared away when necessary.

With the wax temperature low, a thicker coating is obtained on the coil when dipped and the tendency for " blow-holes " to develop is minimized. This is partly due to the thicker coating and partly because the coil itself is heated less by the wax and so has less tendency to expel air through the coating of wax as it is setting.

Once a hole has developed it is desirable to seal it with a hot instrument. To attempt to seal it with a second flash-dip usually results in failure as air is expelled through the hole in the first coat as the coil is warmed up and this punctures the second coating in the same place.

If flash-dipping alone is used for moisture proofing a minimum treatment might be, first flash-dip, seal all blow-holes with wax, then second flash dip.

Similar troubles are experienced with holes when varnish sealing is used.

### (v) Materials

The first requirement for an impregnating compound is good moisture resistance. It is also important that its losses be low at the frequencies concerned, and it is quite possible to obtain impregnants which can be used without noticeably affecting the $Q$ of the impregnated winding.

Because it is usually desirable to have minimum self-capacitance in a coil, the dielectric constant should be low, and the temperature coefficient should also be known. Waxes can be obtained with dielectric constants having a negative temperature coefficient and this is useful in offsetting the temperature coefficient of the coil itself, which is positive in all usual cases.

Apart from the resistance of a moisture proofing compound to moisture absorption, its surface properties are important. A compound which does not wet easily is desirable and tests on moisture proofing should be designed to separate the effects of surface leakage from those of absorption.

Where flash-dipping only is carried out, some of the properties mentioned above are not of such importance but they cannot be ignored completely.

Other points which must be considered when choosing a moisture-proofing compound are that it must be capable of maintaining its properties over the range of temperatures to which it is liable to be subjected. Melting of waxes is the trouble most likely to be encountered, but crazing (and the consequent admission of moisture) may be experienced at low temperatures.

Waxes and varnishes do not themselves support **fungus growth**, but they are liable to collect coatings of dust which may do so. Suitable precautions to avoid this may be desirable. More important is the avoidance in coils as far as possible of materials which are subject to fungus attack. Some types of flexible tubing fall in this category and should not be used, and the same applies to some cheap coil former materials. The fabric covering of wires also may support fungi and should be completely coated with the wax or varnish to avoid this occurrence.

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