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A Design Method for

Wide Band Balanced and Screened Transformers

in the Range 0·1 — 200 Mc/s

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A method for the design and construction of wide band transformers is described in which the physical characteristics of the transformer are determined so that it appears as an ideal transformer together with a symmetrical low-pass filter with the appropriate iterative impedance. The winding, which may be encapsulated, is screened and has a form which achieves a high degree of balance. Examples are given of transformers so designed having an insertion loss of about 0·5dB together with a good balance and impedance match over a wide frequency range.

IN a survey of wide band transformers currently used in aerial feeder systems at radio receiving stations a requirement was seen for a wide band transformer having an improved performance and which was at the same time cheap and easy to produce; it was this requirement which stimulated the work of which the present article is the subject. Although wide band transformers have applications in many fields, e.g. video amplifiers, measuring apparatus, it should be borne in mind throughout that the particular application to aerial feeder systems is the underlying theme.

H.F. aerials are commonly erected as far as is practicable from the receiving station and transmission lines up to a mile long are used to feed the signals to the receivers. The transmission line may be either coaxial cable, in which case a balanced to unbalanced transformer is used to match the aerial to the cable, or a combination of an open wire balanced feeder and coaxial cable, when two transformers are required, the first, a balanced to balanced, to match the aerial to the open wire feeder, and the second, a balanced to unbalanced, to match the open wire feeder to the coaxial cable. At 20Mc/s the loss of open wire feeders is about 1dB/mile and that of coaxial cable, 0·2dB/100ft. It is therefore apparent that the insertion loss of the matching transformers must be extremely low.

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List of Symbols

- L_p = Primary inductance
- C_{sp} = Electrostatic capacitance between screen and primary
- C_{cp} = Electrostatic capacitance between core and primary
- C_{wp} = Electrostatic capacitance of primary winding
- C_{ss} = Electrostatic capacitance between screen and secondary
- C_p = Effective primary capacitance
- C_s = Effective secondary capacitance
- L_L = Leakage inductance } referred to the primary
- R_w = Core losses } referred to the primary
- R_L = Load resistance
- Q = Normalized susceptive component of primary admittance for given r.
- f_1 = Lower frequency limit of the transformer
- f_2 = Upper frequency limit of the transformer
- f_c = Cut-off frequency of equivalent low-pass filter
- Z_o = Iterative impedance of equivalent low-pass filter
- Z_2 = Impedance of equivalent filter at f_2
- N = Turns per μH factor for core
- n = Turns ratio
- n_p = Primary turns
- n_s = Secondary turns
- r = Standing-wave ratio

Losses, other than transformer core losses, will be introduced if there is mismatch between aerial and feeder or between feeder and coaxial cable. The standing-wave ratio (s.w.r.) is a convenient measure of impedance match and this must therefore be as close to unity as possible. This is particularly important in the case of an aerial, e.g. the heptagon, which must be accurately terminated. An s.w.r. of 1·25 over the band is considered to be the worst mismatch which can be tolerated in such instances—this has been taken as a guiding figure in making transformer specifications.

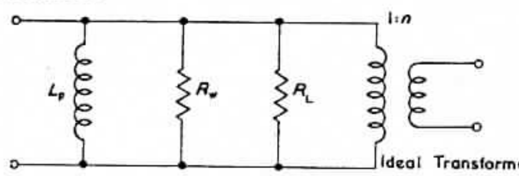


Fig. 1. Low frequency equivalent circuit of a transformer referred to the primary

Further, any unbalance in the transformers will result in a greatly increased noise pick-up and crosstalk in the feeders and hence a high degree of balance, defined for the purposes of this article as twenty times the logarithmic ratio of the voltage across the balanced winding to the out-of-balance voltage, is required. $20 \log \frac{V_1}{V_L}$

General Considerations

It is well known that a physical transformer is replaceable by an ideal transformer coupled with a network, whose form depends on whether high or low frequencies are under consideration.

At low frequencies the transformer is equivalent to the circuit of Fig. 1 where L_p is the inductance of the primary winding and R_w is a resistive shunt which represents the core losses referred to the primary. The insertion loss, and hence s.w.r., depends on the admittance of this parallel combination—the lower this admittance the better the performance of the transformer. R_w and L_p are determined by the efficiency of the core and the number of turns on the primary winding; thus for any specified transformer it is necessary that there should be sufficient turns to provide a maximum shunt admittance at the lowest frequency.

In the high frequency case when the admittance of the R_wL_p combination may be neglected the equivalent circuit referred to the primary becomes that shown in Fig. 2. The primary and secondary capacitances C_p and C_s, and the total leakage inductance L_L become the controlling factors and form a π section low-pass filter. The technique is to so arrange the winding form that C_p and n^2C_s are equal and of such a value that when combined with L_L the filter becomes a constant K π section with an iterative impedance equal to that of the primary impedance of the required transformer.

Standard filter equations may be applied so that if Z_0 is the iterative impedance and f_c the cut-off frequency.

$$Z_0 = (L_L/C)^{\frac{1}{2}} \text{ and } f_c = \frac{1}{\pi(L_L C)^{\frac{1}{2}}}$$

where $C = 2C_p = 2n^2C_s$.

The cut-off frequency is thus determined by the leakage inductance and the shunt capacitances. In practice the latter can be kept small and need not increase with the number of turns. The leakage inductance, however, increases with the number of turns, i.e. f_c decreases, and it is apparent that the upper and lower frequency limits are related and have conflicting requirements. A core must therefore be used

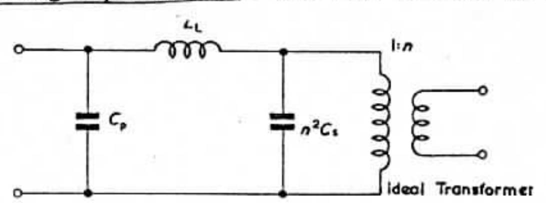


Fig. 2. High frequency equivalent circuit of a transformer referred to the primary

which will permit the maximum shunt admittance to be obtained with the fewest possible turns.

In order to produce a transformer which has a good balance ratio it can be shown that the winding must be of such a form as to have equal coupling between the primary and each half of the secondary and equal winding capacitances for each half of the secondary. Hence in the particular case of a balanced to unbalanced transformer this further requirement must be met.

Two constructions are widely used to meet these requirements. The first is a two winding transformer toroidally wound on a ferrite or mumetal ring core, with a copper screen separating the windings. With this method the individual elements of capacitance and leakage inductance are difficult to control and external compensating circuits are usually included. Winding capacitances tend to be high and restrict the cut-off frequency. The second is an auto-transformer with a concentric spiral winding of copper tape, again on a ring core of ferrite or mumetal. The auto-transformer has a greatly reduced leakage inductance and quite good bandwidths can be obtained as a result, but it suffers from an inherently bad balance ratio because the coupling between the primary and each half of the balanced winding are not equal. The effect depends on the number of turns and becomes worse with increasing turns ratio.

Neither of the transformers quoted above satisfies the conditions for optimum performance. A construction is required in which the leakages and winding capacitances can be accurately and independently controlled and which enables a high degree of winding balance to be attained. It will further be of advantage if the transformer is compact and can be wound on a pot type core which will itself provide adequate screening without external additions.

The present design provides a construction which complies closely with these requirements and which enables a greatly improved performance to be obtained.

Constructional Details

The transformers can be constructed using either a single layer of wire or a concentric spiral of tape for the primary winding. The latter has the disadvantage that the winding capacitance and hence the total primary capacitance increases rapidly as the number of turns is reduced. The winding capacitance of the layer winding represents such a small part of the total primary capacitance, that no such effect is present; however, the leakage inductance of this type of winding is greater than that of the spiral, and for most cases the spiral winding is preferred.

The primary is wound on a thin former built up by winding polystyrene or other suitable tape round the centre rod of a Ferroxcube type LA1 pot core assembly.

This core was chosen because of its high initial permeability, and is preferred to an 'E' type core as the winding is totally enclosed in a high permeability shell reducing stray flux to a negligible value. There is therefore no need for a screening can and the assembly may be potted in a suitable casting resin, or contained in moulded polythene as required. The main requirement of the core is that it should permit the required shunt admittance to be obtained with the minimum number of turns, the ring piece is therefore ground to reduce the gap as far as is practical resulting in a considerable increase in shunt inductance for a given number of turns. This process is controlled by testing each core with a standard winding.

For a balanced primary winding of either type a single unshorted turn of copper tape is wound over the completed primary to form an electrostatic screen, and connected to the core. The thickness of insulation between winding and core, and winding and screen, is arranged to ensure equal winding-core and winding-screen capacitances. In the case of an unbalanced layer wound primary, one end of the winding is connected to the screen, however, it should be noted that for reasons detailed later it is not possible to take a transformer designed for balanced working and convert it to unbalanced by simply earthing one side of the primary. When a spiral winding is used for an unbalanced primary, the last turn is connected to the core and acts as a screen.

The material and thickness of the winding to core and winding to screen insulation together with the width of the winding (and in the case of the spiral, the material and thickness of the inter-leaving material), are all controllable and determine the primary shunt capacitance.

A single layer wire secondary is wound over the screen or the last turn of the spiral as the case may be to provide a symmetrically placed winding, the gauge of wire is chosen so that a winding length appropriate to the required capacitance is obtained. This ensures equal coupling between the primary and each half of the secondary, also equal winding capacitance from each half of the screen. The leakage inductance can be controlled by the thickness of insulation between the screen and the secondary winding, as can the secondary capacitance. The latter is also controlled by the material used for this layer, and by the secondary winding length, i.e. the wire gauge used. For some applications, e.g. aerial transformers, a static leak is required. The secondary is then centre-tapped and the tap connected to the core. As this point is at core potential due to the balanced condition of the winding, making the connexion has no effect on the performance.

A further variation is to use a single layer wire primary and a concentric spiral tape secondary. In this case the first turn of the secondary spiral is earthed and acts as a screen, this form of winding is physically more difficult to achieve, and is only advantageous in certain circumstances where a small secondary capacitance is required. It will therefore not be discussed in detail, but the arguments relating to the other forms of winding are applicable.

The Components of the Equivalent Circuits

Primary Inductances

The amount of primary inductance L_p that is required for a given s.w.r. or insertion loss is dependent on the admittance of the combination of L_p, R_w, and R_L in Fig. 1. It has been pointed out previously that these considerations are only valid at low frequencies, and it is therefore necessary to make L_p sufficiently large to meet the requirements of s.w.r. at the lower limit, f_1, of the frequency range specified. A simplified method of calculating L_p sufficiently accurate for practical applications is as follows.

Consider the parallel combination of the transferred load R_L and the primary shunt inductance L_p, and let the s.w.r. be limited to r. The normalized admittance of R_L and L_p must fall at a point on a Smith Chart within a circle of radius r. The normalized conductive component will remain constant at 1·0 and the range of the suspective component may be found as the intersection of the unity constant conductance circle and the circle of s.w.r. equal to r. Let this value be Q, then

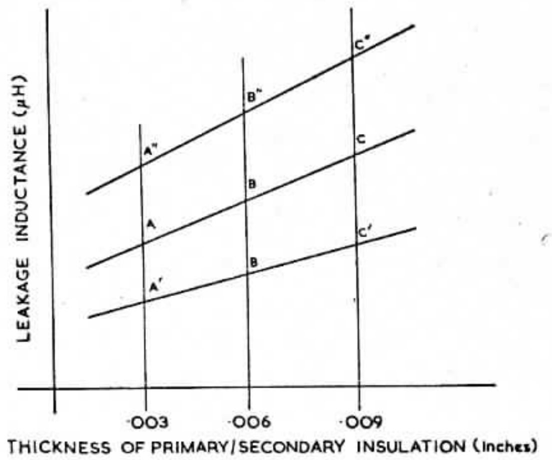


Fig. 3. Relationship between interwinding insulation and leakage inductance

$$Q = (R_L/2\pi f L_p) \text{ or } L_p = (R_L/2\pi f Q)$$

Thus if the chosen s.w.r. is 1·25, f_1 is 1Mc/s and R_L is 80Ω, from the Smith Chart $Q = 0{\cdot}225$ and hence

$$L_p = \frac{80 \times 10^6}{2\pi \times 10^6 \times 0{\cdot}225} = 56{\cdot}6\mu\text{H}$$

If N is the turns per μH factor for the core, the primary turns required for an inductance L_p are given by $n_p = N(L_p)^{\frac{1}{2}}$. The minimum attainable leakage inductance, L_L, increases with primary turns, and as low values for L_L are necessary at high frequencies it is desirable to make N small. As described previously, N is decreased by grinding the ring piece of the pot core assembly. Using LA1 material it is found convenient to make N unity, and hence $n_p = (L_p)^{\frac{1}{2}}$; thus 7 turns will result in 40μH, 8 turns in 64μH etc.

Turns Ratio

The turns ratio, n, is the square root of the required impedance ratio, and the secondary number of turns n_s is given by $n \cdot n_p$.

Leakage and Lead Inductance

For a given winding form the minimum value for leakage inductance depends on the number of primary turns, but values above this minimum can be obtained by reducing the coupling factor. This is achieved by increasing the thickness of insulation between the last turn of the spiral winding and the layer winding.

In order to make practical use of these two variables in the design of a specific transformer it is advantageous to plot a number of curves as in Graph D. This is seen to consist of a family of curves of leakage inductance against thickness for various numbers of primary turns. Their use will be apparent from the examples of transformer design given later and here will only be explained the method of producing them.

Any convenient number of turns may be wound on a core and a secondary winding of the desired number of turns of suitable wire gauge is chosen as above. The leakage inductance is then measured by a short-circuit method for varying thicknesses of interwinding insulation. In Graph D an 8-turn primary winding was chosen and L_L was measured for interwinding thicknesses of ·003in. and ·006in. This is represented diagrammatically in Fig. 3. The points A, B, C . . . can then be plotted. The curve through A, B, C . . . then represents the relation leakage inductance/thickness of interwinding insulation for an 8-turn primary winding. At each point such as A, the number of primary turns per μH of leakage, N_L, may be obtained and then

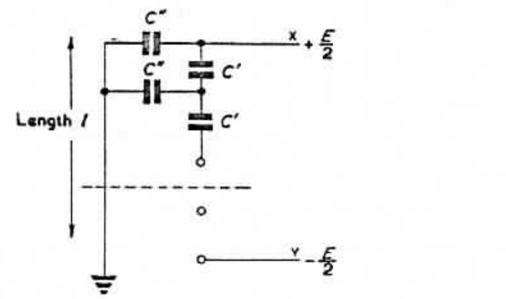


Fig. 4. Distribution of the capacitances of a single layer winding over a conducting former

using the relation $n_p = N_L(L_L)^{\frac{1}{2}}$ the leakage may be calculated for various values, of primary turns. Points A′, A″ . . . are then plotted corresponding to these values and the family of curves A′ B′ C′ . . . , A″ B″ C″ drawn. The series of curves shown in Graph A was plotted for a tape width of 0·25in. Theoretically a series should be plotted for each tape width used, but in practice, the variations of leakage with tape width has been found to be small.

When the transformer requirements, e.g. impedance or cut-off frequency, involve leakage inductances which are small it is necessary to take special care to reduce the loop inductance of the lead-out wire. This may be done by keeping the leads short and close together. A working figure for the loop inductance is 10 per cent or less of L_L. Using 24 s.w.g. wire, a lead 1in long has an inductance of about 0·02μH and referring to graph A it will be seen that this is only significant in the case of a 4-turn primary or less.

Secondary Capacitance

When a single layer coil is wound on to an earthed conducting former, the capacitive effect exhibited by the coil is modified by the balance or unbalance of the voltage applied to it. Consider the cross-section of a winding shown in Fig. 4. The capacitance seen looking into the leads XY is a function of the capacitances C' between adjacent turns and the capacitances C'' of each individual turn to the former. The capacitances C' are small and add in series and hence may be neglected. The effective capacitance across XY is thus some combination of the capacitances C''.

Suppose the winding to be of length l and let the electrostatic capacitance between the layer and the former be C. Let a balanced voltage E be applied to the coil, i.e. X and Y are at potentials $+E/2$ and $-E/2$ respectively. The voltage distribution along the coil is thus as in Fig. 5.

Let the capacitance per unit length be p, i.e. $C = pl$, and consider an element δx of l at a distance x from the centre.

The current flowing through the elemental capacitance formed by x is given by

$$\delta I = \mathrm{j}\omega p \delta x \, . \, E/2 \, . \, x/(l/2)$$

Hence the total current flowing from the positive side of the layer to the former is I, where

$$I = \mathrm{j}\omega p \, . \, E/l \, . \int_0^{l/2} x \, . \, dx$$

$$\text{or } I = (\mathrm{j}\omega C/8) \, . \, E$$

An equal current flows from the former to the negative side of the layer and thus, looking into XY, the winding appears to have a capacitive current I flowing into it.

But the current flowing through a capacitance C due to a voltage E is $i = \mathrm{j}\omega C \, . \, E$. Hence the effect of the balanced voltage distribution across the winding is to make its capacitance appear as $\frac{1}{8}$ of its electrostatic value.

Using a similar argument the effective capacitance of the winding when an unbalanced voltage is applied to it is $\frac{1}{2}$ its electrostatic value.

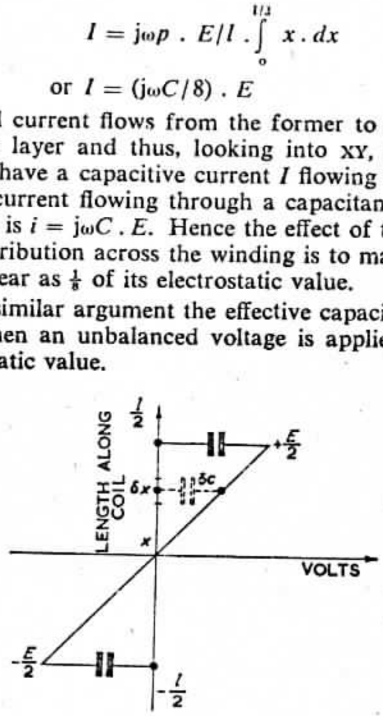


Fig. 5. Voltage distribution along the length of a balanced winding

The electrostatic capacitance between the layer and the screen may be regarded as that of a cylindrical capacitor so that

$$C_{ss} = \frac{0{\cdot}6128 \times l \times k}{\log (r_2/r_1)} \text{ pF}$$

where l is the winding length in inches, k the dielectric constant of the interleaving material and r_1, r_2 the radii of the screen and the layer winding respectively. An increment applied to both r_1 and r_2 has little effect on the result, thus small variations in the number of primary turns do not effect the capacitance.

Because of the circular cross-section of the wire, it is necessary to add a weighting factor to r_2, this will depend on the gauge of wire used, and is only effective when the interwinding insulation is less than ·003in in thickness. For design purposes it is convenient to plot a graph of C_{ss} against tape width and thickness of primary/secondary insulation. Graph C has been plotted with a weighting factor added to value of r_2 below ·003. This gives a reasonable average for wire gauges between 32 and 38 s.w.g.

PRIMARY CAPACITANCE

The capacitance of the layer winding is made up of the winding to core and the winding to screen capacitances C_{cp} and C_{sp} respectively. As in the case of the secondary, the inter-turn capacitance may be neglected and C_{cp} and C_{sp} can be considered as cylindrical capacitors. As before the static capacitance is modified by the voltage distribution with the result that for an unbalanced winding.

$$C_p = \frac{\frac{1}{2}(C_{cp} + C_{sp})}{2}$$

and for a balanced winding

$$C_p = \frac{(C_{cp} + C_{sp})}{8}$$

The capacitance of the unbalanced winding in the form of a concentric spiral is made up of the winding capacitance C_{wp} and the first-turn-to-core capacitance C_{cp}. The winding capacitance may be regarded as that of a series connexion of concentric cylinders and the expression of C_{wp} becomes

$$C_{wp} = \frac{(0{\cdot}6128)\, kw}{\log \left(\dfrac{r_o + t_1 + (n-1)\, t_2}{r_o + t_1} \right)} \text{ pF}$$

where k is the dielectric constant of the interleaving material
- w is the tape width in inches
- r_o is the sum of the radius of the core + the thickness of the former
- t_1 is the thickness of the conductor
- n is the number of turns
- t_2 is the thickness of the interleaving material.

C_{cp} may also be considered as the capacitance of a cylindrical capacitor formed between the core and the first turn of the primary winding,

$$\text{i.e. } C_{cp} = \frac{(0{\cdot}6128)\, kw}{\log (r_o/r_i)} \text{ pF}$$

where r_i is the radius of the core and other symbols have the same meaning as before. Small variations in former thickness do not significantly effect the calculation of C_{wp} as any such small increment affects both the numerator and the denominator of the expression in the logarithm. When a balanced winding is under consideration, the capacitance between the last turn and screen C_{sp}, must be included, this is again considered as a cylindrical capacitor where r_i is the radius of the last turn and r_o that of the screen. To maintain balance C_{sp} must be made equal to C_{cp} and the primary capacitance which is again modified by voltage distribution becomes

$$C_p = \frac{\frac{1}{4}(C_{cp} + C_{sp}) + C_{wp}}{8}$$

for the balanced condition, and

$$C_p = \frac{C_{cp} + C_{wp}}{2} \text{ for the unbalanced condition.}$$

Practical Design

FREQUENCY LIMITS AND CHOICE OF WINDING FORM

The lower and upper frequency limits f_1 and f_2 have conflicting requirements in that a low f_1 requires a large primary inductance which results in a large leakage inductance reducing the limit for f_2, and vice versa. Accordingly, in design either f_1 or f_2 may be taken as a starting point and the bandwidth arranged to accommodate the other limit. The relationship between f_2 and the cut-off frequency f_c of the equivalent filter is given by:

$$Z_2 = \frac{Z_o}{[1 - (f_2/f_c)^2]^{\frac{1}{2}}} \text{ where } Z_o = (Z_2/r)$$

thus for $r = 1{\cdot}25$, $f_2 = 0{\cdot}6 f_c$. Applying the standard filter equations to the equivalent circuit referred to the primary

$$L_L = \frac{0{\cdot}6 \, . \, 10^6 \, Z_o}{\pi f_2} \, \mu\text{H and } C_p = n^2 C_s = \frac{0{\cdot}3 \, . \, 10^{12}}{f_2 \pi Z_o} \text{ pF}$$

From these equations it will be seen that f_2 depends on L_L, C_p and Z_o. For example a transformer having $f_2 = 200$Mc/s, $n^2 = 4$ and $Z_o = 400\Omega$, would require $L_L = 0{\cdot}35\mu$H, and $C_p = n^2 C_s = 1{\cdot}2$pF. If Z_o is reduced to 80Ω L_L becomes $0{\cdot}07\mu$H and $C_p = n^2 C_s = 6$pF.

It was remarked earlier that a single layer wire wound primary allows a low primary capacitance, but has a rather larger leakage inductance than that obtainable with a spiral tape winding. An indication of the limits on the performance which can be achieved using the types of winding described may be obtained from Fig. 6.

The factors limiting Z_o and f_2 are mainly the leakage inductance and the primary capacitance, all of which are inter-dependent. In practice the lowest values of the leakage inductance and primary capacitance which can be obtained simultaneously are approximately $0{\cdot}1\mu$H and 5pF respectively for a spiral tape primary winding. Hence using $C_p = (0{\cdot}3 \times 10^{12})/f_2\pi Z_o$, and letting $C_p = 5$pF, curve AL defines the limits on f_2 and Z_o imposed by C_p, all points

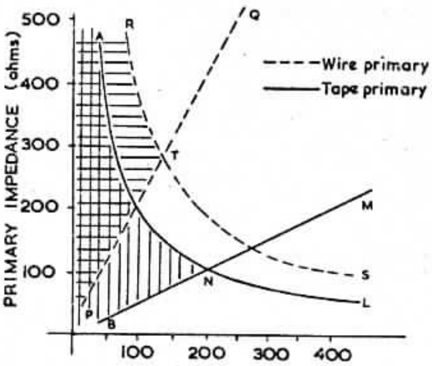


Fig. 6. Anticipated performance limits for tape and wire transformers using LA1 cores

(f_2Z_o) lying inside the concave portion of the curve being excluded. Similarly curve BM defines the limits imposed by L_L and all points (f_2Z_o) lying below BM are excluded. Thus the combinations of f_2 and Z_o which can be achieved in practice are those which lie within the region ANB of Fig. 6. The intersections of curves AL and BM correspond approximately to $f_2 = 200$Mc/s when $Z_o = 100\Omega$. Values of Z_o greater or less than this can be achieved only with a reduced upper frequency limit.

Again in the case of a single layer wire primary winding, the practical limits of L_L and C_p are approximately $0{\cdot}4\mu$H and 3·0pF, the corresponding curves for these values are shown in Fig. 6 by PQ and RS. The greater suitability of the single layer wire primary for higher impedances will immediately be seen, the permissible values being those which lie within the region RTP. In the cross-hatched region, both types of winding are equally suitable.

Materials and Graphical Aids

For design purposes it is convenient to plot graphs of the various functions. Graph A shows the relationship between the capacitances C_{sp}, C_{cp}, and the thickness of insulation, for various tape or winding widths. Graph B is a plot of the winding capacitance C_{wp}, against turns, tape width and interleaving thickness. The relationship between C_{ss} and screen to secondary insulation for a range of winding lengths is shown in Graph C; Graph D is plotted for the purposes of the examples which follow, from measurements of L_L as described previously.

Polystyrene tape was chosen for building up the various layers of insulation because of its low dielectric constant of 2·6 and because it does not stretch or flow when wound tightly. In the examples which follow a tape thickness of ·0015in* was used as it was readily available. The copper tape used for spiral windings and screens was ·001in thick, the dimensions of the pot core limit the width to 0·375in maximum. For the wire windings enamelled copper wire was used. Table 1 gives the number of turns per 0·1in close wound for gauges between 20 and 40.

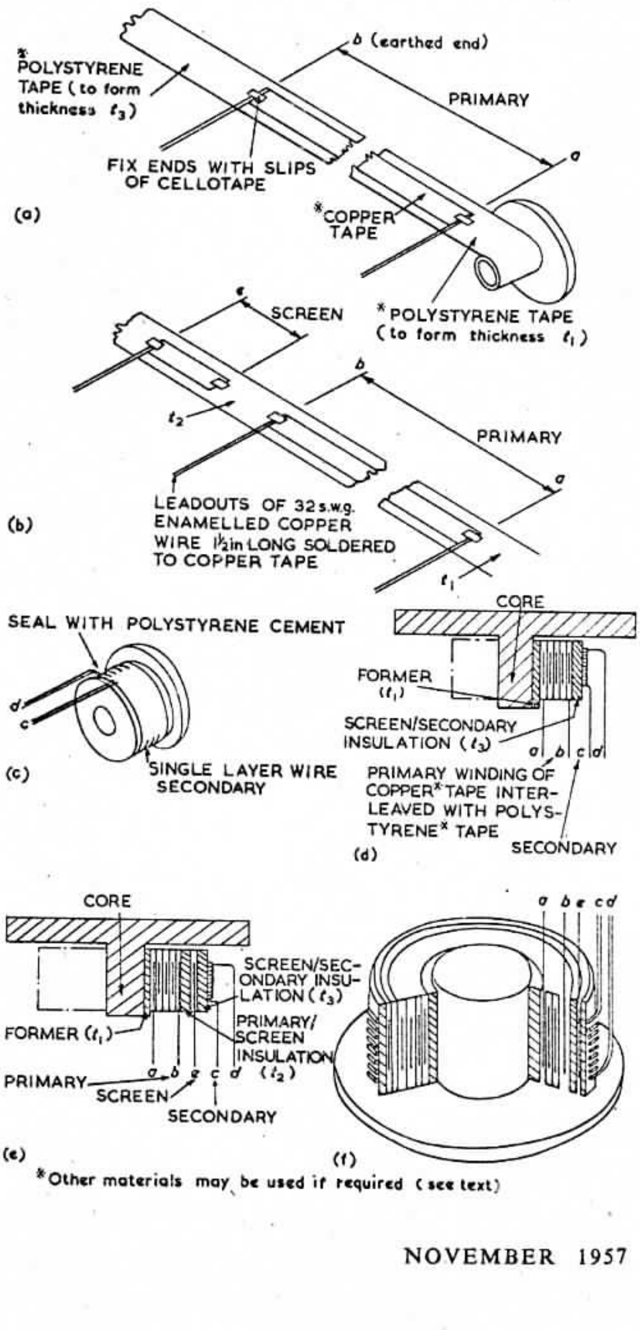
Illustrations of these constructions are given in Fig. 7

TABLE 1.

WIRE GAUGE	TURNS/0·1IN.	WIRE GAUGE	TURNS/0·1IN.
20	2·6	30	7·25
21	2·92	31	7·75
22	3·3	32	8·27
23	3·83	33	8·93
24	4·16	34	9·70
25	4·55	35	10·5
26	5·02	36	11·6
27	5·51	37	12·8
28	6·10	38	14·5
29	6·6	39	16·4
30	7·25	40	17·8

Fig. 7. Constructional details of windings

(a) Unbalanced tape primary
(b) Balanced tape primary and screen
(c) Secondary winding both types
(d) Section through balanced/unbalanced type
(e) Section through balanced/balanced type
(f) Isometric section of balanced type



*Other materials may be used if required (see text)

* = 37,5 µm

Practical Examples

Design Procedure Based on f_1 Using Spiral Tape Primary

Let the requirement be for an 80Ω unbalanced to 640Ω balanced transformer having s.w.r. not greater than 1·25 at and between 1 and 40Mc/s. The primary inductance is found to be 56·6μH and therefore, using $n_p = (L_p)^{\frac{1}{2}}$, 8 turns are required for the primary winding.

Next, a thickness of interwinding insulation must be found which permits values of L_L and n^2C_s satisfying the standard filter equations:

$$L_L = (Z_o \cdot 10^6/\pi f_c)\mu\text{H} \text{ and } n^2C_s = (10^{12}/2\pi f_c Z_o)\text{pF}$$

Eliminating f_c between these two equations gives

$$n^2C_s = (10^6 L_L/2Z_o^2)$$

For a balanced secondary:

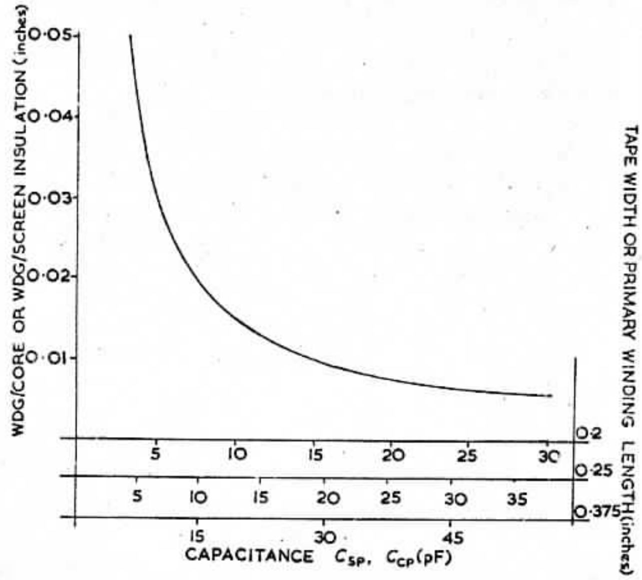
$$C_{ss} = (10^6 L_L/2Z_o^2) \cdot (8/n^2)\text{pF} \quad \text{........ (1)}$$

Values of L_L for various interwinding insulation are now substituted in equation (1), and the values of C_{ss} obtained replotted on Graph C. The point of intersection of the two curves gives the required thickness. A different result will be obtained for each winding length scale, and any convenient set of conditions meeting the requirement of f_2 = 40Mc/s may be chosen. f_2 is determined by applying the following equation to each result.

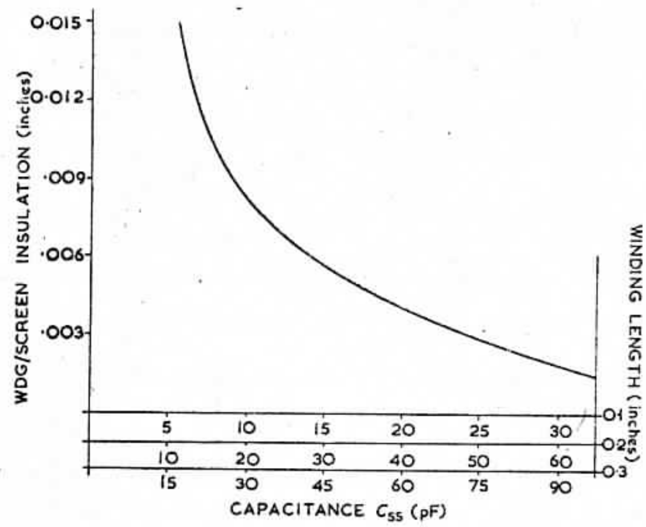
$$f_2 = (0{\cdot}6Z_o/\pi L_L)\text{Mc/s} \quad \text{.............. (2)}$$

These results are shown in Table 2

A winding length of 0·2in is chosen for convenience, and a primary tape width of 0·25in will be required to accommodate it. The winding capacitance C_{wp} of 8 turns of 0·25in tape interleaved with ·0015in insulation is shown by Graph B to be 18pF. In order to make $C_p = \frac{1}{2}(C_{cp} + C_{wp})$ equal to n^2C_s = 22pF, a former is required such that C_{cp} = 26pF. Using Graph A this is found to be ·007. The number of secondary turns is found by applying n_s =

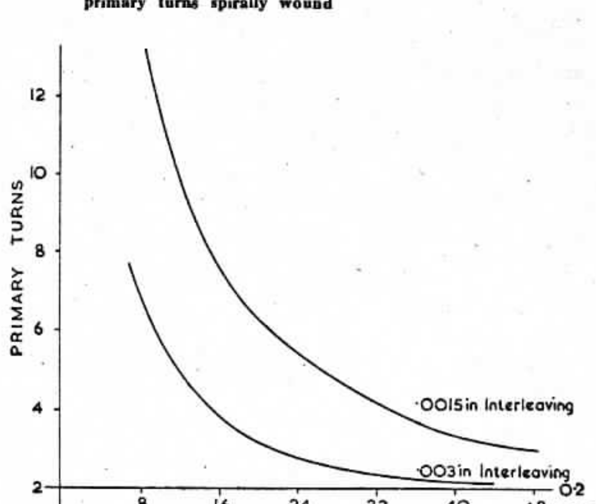


Graph A. Capacitance primary to core or screen (C_{CP}, C_{SP}) for various tape or winding widths and thicknesses of primary to core insulation



Graph C. Capacitance secondary to screen (C_{SS}) for various winding lengths and thickness of secondary to screen insulation

Graph B. Primary winding capacitance (C_{WP}) for various tape widths and primary turns spirally wound



Graph D. Relationship between leakage inductance and interwinding insulation for various primary turns

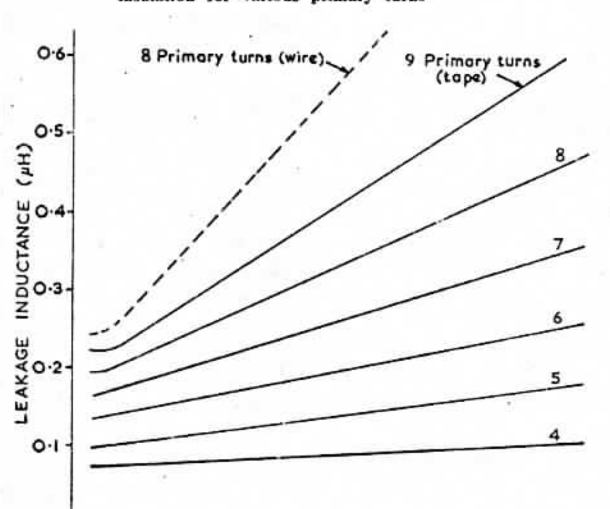


TABLE 2.

FROM GRAPH D		EQUATION (1)	FROM GRAPH C		FROM GRAPH D		EQUATION (2)
P/S INSULATION	L_L (μH)	C_{ss} (pF)	WINDING LENGTH	P/S INSULATION	C_{ss} (pF)	L_L (μH)	f_2 (Mc/s)
0·006	0·27	21	0·1	·0033 (2T)	21·5	0·22	70
0·018	0·45	35	0·2	·0072 (5T)	22·5	0·29	53
			0·3	·0117 (8T)	28·0	0·36	42.5

TABLE 3.

f (Mc/s)	INS. LOSS (dB)	BALANCE (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
1·0	0·5	50	0·96+j0·188	1·22	1·01 +j0·186	1·2
5·0	,,	,,	0·96+j0·042	1·06	1·025+j0·013	1·12
10·0	,,	,,	0·96+j0·021	1·05	1·03 −j0·015	1·12
20·0	,,	45	0·96+j0	1·04	1·05 −j0·05	1·05
40·0	,,	41	0·90+j0	1·1	1·00 −j0·04	1·09
60·0	1·0	38	0·77+j0	1·3	0·985−j0·015	1·16

$n \cdot n_p$, to be 23 turns. In order to cover a winding length of 0·2in, 36 s.w.g. enamelled wire is required.

Thus the complete winding specification is:

Core:— Ferroxcube type LA1. (Ring piece ground so that $n_p = (L)^{\frac{1}{2}}$).

Former:— 5 turns ·0015in × 0·375in polystyrene tape.

Primary:— 8 turns of ·001in × 0·25in copper tape, interleaved with ·0015in × 0·375in polystyrene tape.

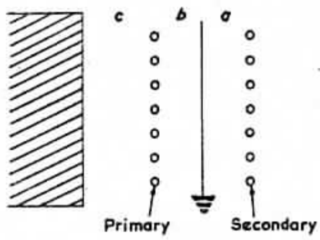


Fig. 8. Screen position using single layer wire primary

Interwinding insulation:— 5 turns of ·0015in × 0·375in polystyrene tape.

Secondary:— 23 turns of 36 gauge enamelled wire.

A transformer was constructed to this specification and measurements made of s.w.r. insertion loss and balance ratio, the results are shown in Table 3.

Design Procedure Based on f_1. Using Single Layer Wire Primary

Let the requirement be as before. Proceeding as above one finds that $n_p = 8$ and a secondary winding of 0·2in with three turns of insulation between primary and secondary results in $L_L = 0·36$ and $C_p = n^2C_s = 30$pF.

The three turns of insulation must be distributed over (a) and (b) in Fig. 8. As C_{sp} can be large without disadvantage, let (b) = one turn and (a) = two turns, then to keep C_{ss} to 30pF with the reduced screen to secondary separation, reduce the winding length proportionally to 0·15in. With (b) = one turn and a 0·25in winding length, C_{sp} = 60pF, to make $[(C_{cp}+C_{sp})/4] = 30$, C_{cp} must also be 60pF and therefore (c) must also be one turn. In order that the 8-turn primary covers 0·25in, 22 s.w.g. wire is required.

The complete winding specification is:

Core:— Ferroxcube type LA1 (ring piece ground so that $n_p = (L_p)^{\frac{1}{2}}$.

Former:— 1 turn ·0015in × 0·375in polystyrene tape.

Primary:— 8 turns 22 gauge enamelled wire.

TABLE 4.

f (Mc/s)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
1	0·96 +j0·16	1·18	1·02 +j0·156	1·17
2	0·96 +j0·08	1·08	1·02 +j0·07	1·08
5	0·96 +j0·025	1·04	1·01 +j0·018	1·04
10	0·96 +j0·025	1·04	1·01 +j0·015	1·03
20	1·11 +j0	1·11	1·06 +j0·02	1·06
30	0·84 +j0·06	1·19	1·12 +j0·03	1·13
40	0·84 +j0·12	1·25	1·142−j0·08	1·17
50	0.77 +j0·204	1·42	1·11 −j0·013	1·12
60	0·705+j0·302	1·63	0·99 −j0·018	1·2

Primary/screen insulation:— 1 turn ·0015 × 0·375in polystyrene tape.

Screen:—1 unshorted turn ·001 × 0·375in copper tape.

Screen/secondary insulation:— 2 turns ·0015 × 0·375in polystyrene tape.

Secondary:— 23 turns of 40 s.w.g. enamelled wire.

A transformer was constructed to this specification and its s.w.r. measured, the results are shown in Table 4.

Design Procedure Based on f_2 Using Spiral Tape Primary

Let the requirement be for an 80Ω unbalanced to 640Ω balanced transformer, having an s.w.r. not greater than 1·25 at and between 10 and 100Mc/s.

Using standard filter equations:

$$L_L = \frac{Z_o \cdot 0·6 \cdot 10^6}{\pi f_2} = 0·15\mu\text{H and } n^2C_s = C_p = \frac{0·3 \times 10^{12}}{\pi f_2 Z_o} = 12\text{pF}$$

∴ as $n^2 = 8$, $C_{ss} = 12$pF.

The equation $C_p = \frac{1}{2}(C_{wp} + C_{cp}) = 12$pF is satisfied if C_{wp} and C_{cp} are each made 12pF, from Graph B, $C_{wp} = 12$pF with 5 turns of 0·2in tape interleaved with ·003in material, and from Graph A C_{cp} is 12pF with 0·2in tape if a 0·012in former is used. Now if $n_p = 5$ and L_L is required to be 0·15μH, reference to Graph D shows that a thickness of 0·012in primary/secondary insulation is required. To make C_{ss} 12pF with this primary/secondary insulation, the winding length required is 0·17in (Graph C), and 14 turns of 32 gauge wire are needed.

The bottom frequency limit f_1 may be checked by applying the equation:

$$f_1 = \frac{Z_o}{2\pi \times 0·225 \times L_p} \text{ Mc/s. In this case } f_1 = 6·3\text{Mc/s.}$$

TABLE 5.

f (Mc/s)	INS. LOSS (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.	BAL (dB)
1	1·0	1·02+j0·34	1·39	0·985+j0·392	1·48	>50
2	0·75	1·02+j0·16	1·18	1·0 +j0·181	1·2	>50
5	0·5	1·04+j0·12	1·13	1·015+j·056	1·054	>50
10	0·5	1·04+j0·08	1·08	1·071+j.015	1·03	>50
20	0·5	1·02+j0·041	1·03	1·07 +j·02	1·07	48
30	0·5	0·99+j0·03	1·02	1·1 +j·03	1·1	42
40	0·5	0·96+j0·08	1·09	1·16 +j0	1·16	41
50	0·5	0·9 +j0·05	1·12	1·16 +j·025	1·17	40
60	0·75	0·9 +j0	1·1	1·18 +j·121	1·22	38
70	1·0	0·9 +j0	1·1	1·14 +j·141	1·2	36
80	0·75	0·83+j0·08	1·21	1·08 +j·161	1·14	32
90	1·0	0·83+j0	1·2	1·03 +j·136	1·142	27
100	1·5	0·87+j0·1	1·18	0·975+j·101	1·12	26

TABLE 6.

f (Mc/s)	INS. LOSS (dB)	NORMALIZED ADMITTANCE (640Ω)	S.W.R.	NORMALIZED ADMITTANCE (80Ω)	S.W.R.
0·5	0·5	1·04 −j0·402	1·52	0·92 −j0·382	1·36
1·0	0·25	1·06 −j0·127	1·15	0·92 −j0·183	1·23
10·0	0·25	1·06 +j0·06	1·09	1·0 +j0·05	1·054
20·0	0·25	1·05 +j0·12	1·14	1·1 +j0·06	1·12
30·0	0·5	1·03 +j0·17	1·18	1·25 −j0·07	1·15
40·0	0·75	1·03 +j0·32	1·37	1·24 −j0·32	1·42
50·0	1·0	0·994+j0·3	1·38	1·025−j0·49	1·61
60·0	1·5	0·96 +j0·42	1·52	0·77 −j0·49	1·83

The complete winding specification is:

Core:— Ferroxcube type LA1 (ring piece ground so that $n_p = (L_p)^{\frac{1}{2}}$.

Former:— 8 turns ·0015 × 0·375in polystyrene tape.

Primary:— 5 turns 0·2in × ·001in copper tape, interleaved with ·003in × 0·375in polystyrene tape.

Primary/secondary insulation:— 8 turns ·0015in × 0·375in polystyrene tape.

Secondary:— 14 turns 32 gauge enamelled wire.

Such a transformer was wound, and on measurement was found to have an s.w.r. of 1·42 at 100Mc/s. This was caused by too large a capacitive component on the secondary side and the transformer was rewound using 37 gauge wire with the results given in Table 5.

Encapsulation

A convenient finish for this type of transformer particular when required for outdoor use with aerial systems, is encapsulation. Typical examples are shown in Fig. 9 where Araldite casting resin has been used. Small quantity production of the transformer detailed in example (1) has been undertaken and consistent results obtained after encapsulation. The measurements shown in Table 6 were taken using one of these production transformers after encapsulation. The balance was not affected.

Production Cost

Complete transformers encapsulated as shown in Fig. 9, and suitable for use as wide band matching units for aerial systems, have been produced for under £3. The cost of the single plastic mould used was 5s., a multiple mould would be required for quantity production.

Conclusion

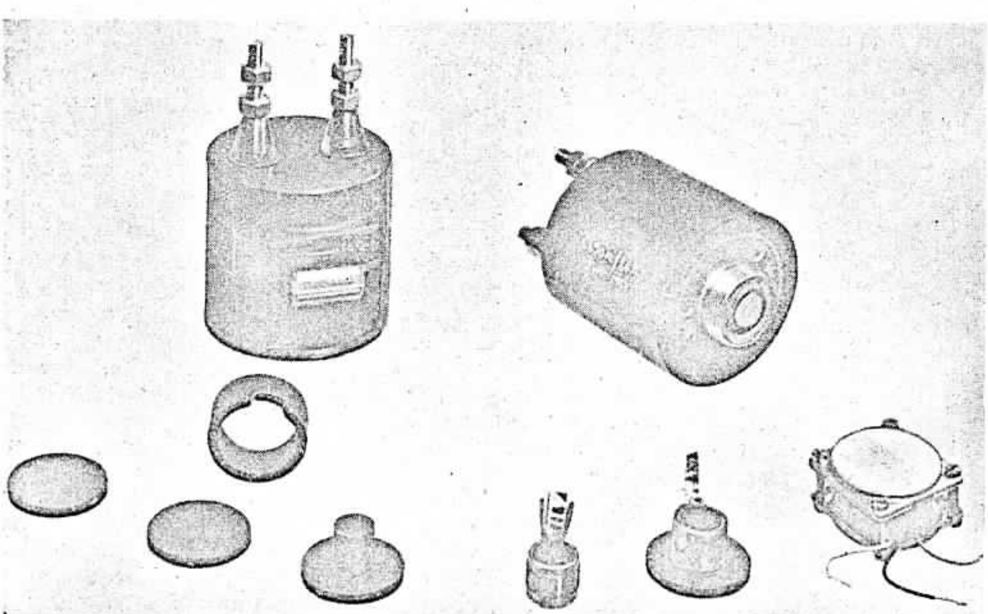
The article demonstrates that a wide band transformer which complies closely with a specified performance may be produced at low cost using this design method.

Some of the practical limitations imposed by frequency and impedance are indicated, but it should be remembered that these correspond to a required impedance match of s.w.r. = 1·25. If this restriction is relaxed f_2 will be increased by a factor determined by the equation given.

For aerial matching purposes the frequency range of a transformer is normally specified between 2dB points, using this criterion the frequency range of the transformer described in example (1) is 0·15 to 80Mc/s and that of example (3) 0·3 to 170Mc/s.

It is recognized that the performances obtained so far are by no means optimum. Further work will be devoted to extending the range of the transformers, one possible approach being to arrange the transformer elements to conform to an M of MM' configuration so that a greater portion of the pass-band of the equivalent filter may be utilized; another is to reduce all physical dimensions. By these means it seems likely that the upper frequency limit for small impedances and ratios may be extended to at least 500Mc/s.

Fig. 9. Component parts and completed transformer, encapsulated in Araldite, for use in aerial feeder systems



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