

APPLICATION NOTE

**Design of HF wideband power
transformers**

ECO6907

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1 INTRODUCTION

Transmission line power transformers can be used to perform a variety of functions, among which are phase reversal, balanced to unbalanced coupling, impedance transformation and hybrid functions. Such transformers find many applications in wide-band power amplifiers for both s.s.b. transmitters in the h.f. region and f.m. transmitters in the lower v.h.f. region.

The properties of a practical h.f. power transformer are discussed here and their effect on transformer performance is analysed. Since losses must be kept low, in practice the transformer will use a ferrite core. Further, we have limited the discussion to cores without an air-gap since these have a low stray magnetic field, a high permeability, and can cover the power range (up to 80 W) dealt with here. Data (dimensions, permeability values etc.) on all core types can be found in our Data Handbook "*Soft Ferrites*", MA01. A glance through the Handbook will show the wide range of materials, dimensions and types from which the designer may choose. It must be remembered, of course, that when cores constructed in two parts (pot-cores and cross-cores, for example) are used, the type without an air-gap must be selected.

Throughout we have aimed at giving practical solutions to the problems posed by material and design limitations. In particular, compensating techniques for extending the frequency range of a number of transformer configurations are discussed. To give an idea of some application possibilities, practical examples in several transformer configurations have been worked, using transformer cores from our range of ferrites.

2 TRANSFORMER SPECIFICATION

The transformer design considerations dealt with in this publication are:

- Maximum power level to be handled
- Frequency range
- Input and output impedance
- Allowable reflection and resistive losses.

How a transformer can meet the above considerations for a particular application is analysed in the following three sections. The first two sections deal with the influence of the core and transmission line respectively on transformer performance, and the third with mismatch compensation techniques.

3 INFLUENCE OF THE CORE ON PERFORMANCE

3.1 Primary Inductance

This inductance determines the amount of reflection at the low frequency end of the band. It can be calculated using the formula:

$$L = \mu_0 \mu_r n^2 A / l$$

in which:

L = inductance in H

$\mu_0 = 4 \pi 10^{-7}$ (rationalised M.K.S. units)

μ_r = relative permeability

A = average ferrite cross section in m²

l = average length of the lines of force in m

n = number of turns between the input connections.

In a simple example, like the phase reversing transformer, this relation holds. Other cases may require a transformation (see Section 7.1).

If degrading of performance at the high end of the band is to be avoided, the value of L must not be higher than really necessary. A good practical value is:

$$L = 4R/\omega_{\min}$$

in which:

R = midband input resistance in Ω

$\omega_{\min} = 2\pi$ times the minimum frequency in Hz.

Where requirements are severe the compensation technique described in Section 5.1 may be used.

3.2 Core Losses

The losses caused by the core material will be represented here as a resistance (R_p) in parallel with the input. This resistance depends on:

- The sort of ferrite material
- The frequency
- The quantity L/μ_r
- The maximum flux density B_{\max} .

In the small signal case ($B_{\max} \rightarrow 0$), R_p can be calculated with the aid of curves of the type shown in Fig.1⁽¹⁾. In these curves a comparison is made between different core materials based on equal core dimensions and equal number of turns. It can be seen that 4C4 and 4C6 are the best materials for frequencies above approximately 2.5 MHz. In the high v.h.f. region IZ2 ferroxplana shows interesting properties as can be seen from the same figure.

The power handling capability of a transformer is closely dependent on the behaviour of R_p as a function of B_{\max} . For the section of the B-H curve with which we are dealing, B_{\max} can be calculated using the formula:

$$B_{\max} = V_{\max}/\omega \times A \times n$$

in which:

B_{\max} = maximum flux density in T⁽²⁾

$\omega = 2\pi$ times frequency in Hz

A = ferrite cross section in m²

n = number of turns

V_{\max} = maximum value of voltage across n turns in V.

(1) The curves of Figs 1 to 7 have been drawn from measurements on single samples of the ferrite materials. Thus the average curves may differ somewhat from those shown.

(2) The letter T stands for Tesla, the unit of magnetic flux density in the SI unit system. The following relationship holds:

$$1T = 1 \text{ Wb/m}^2 = 1 \text{ V}_{\text{sec}}/\text{m}^2 = 10000 \text{ gauss.}$$

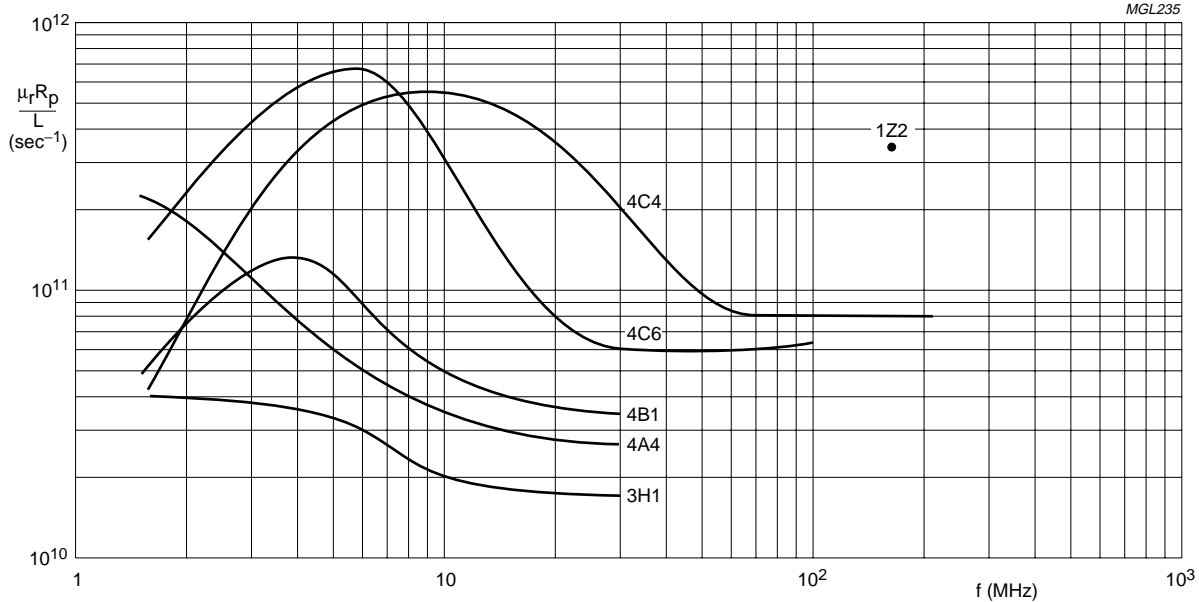


Fig.1 Curves of $\mu_r R_p / L$ plotted against frequency for our various ferrites. The curves have been plotted for small signal conditions ($B_{\max} \rightarrow 0$).

In Figs 2 to 7 the quantity $\mu_r R_p / L$ is given for different ferrite materials as a function of the product $B_{\max} \times f$ with the frequency as a parameter. The product $B_{\max} \times f$ has been chosen because, for most transformers, its value remains constant for changing frequency. From Figs 2 to 7 it can be seen that R_p decreases as B_{\max} increases, especially at lower frequencies. This forms the primary limit on the power handling capability of these transformers. If 4C4 material (Fig.5) is used in the h.f. region, the $B_{\max} \times f$ product must not be higher than approx. 2×10^4 T.Hz. Combining this with the choice of L according to the second equation in Section 3.1, we find that the power loss caused by the core material will be no more than 1%. At frequencies of 30 MHz and higher it seems that higher $B_{\max} \times f$ products, perhaps up to 10^5 T.Hz, can be used. For 1Z2 ferroplana this has already been confirmed by measurements at 165 MHz.

A very conservative choice of the B_{\max} value must also be avoided because this leads to a greater length of the transmission line and consequently more loss at the high end of the band.

4 INFLUENCE OF THE TRANSMISSION LINE ON PERFORMANCE

4.1 Resistive Loss and Power Handling

The power loss in the transmission line depends on:

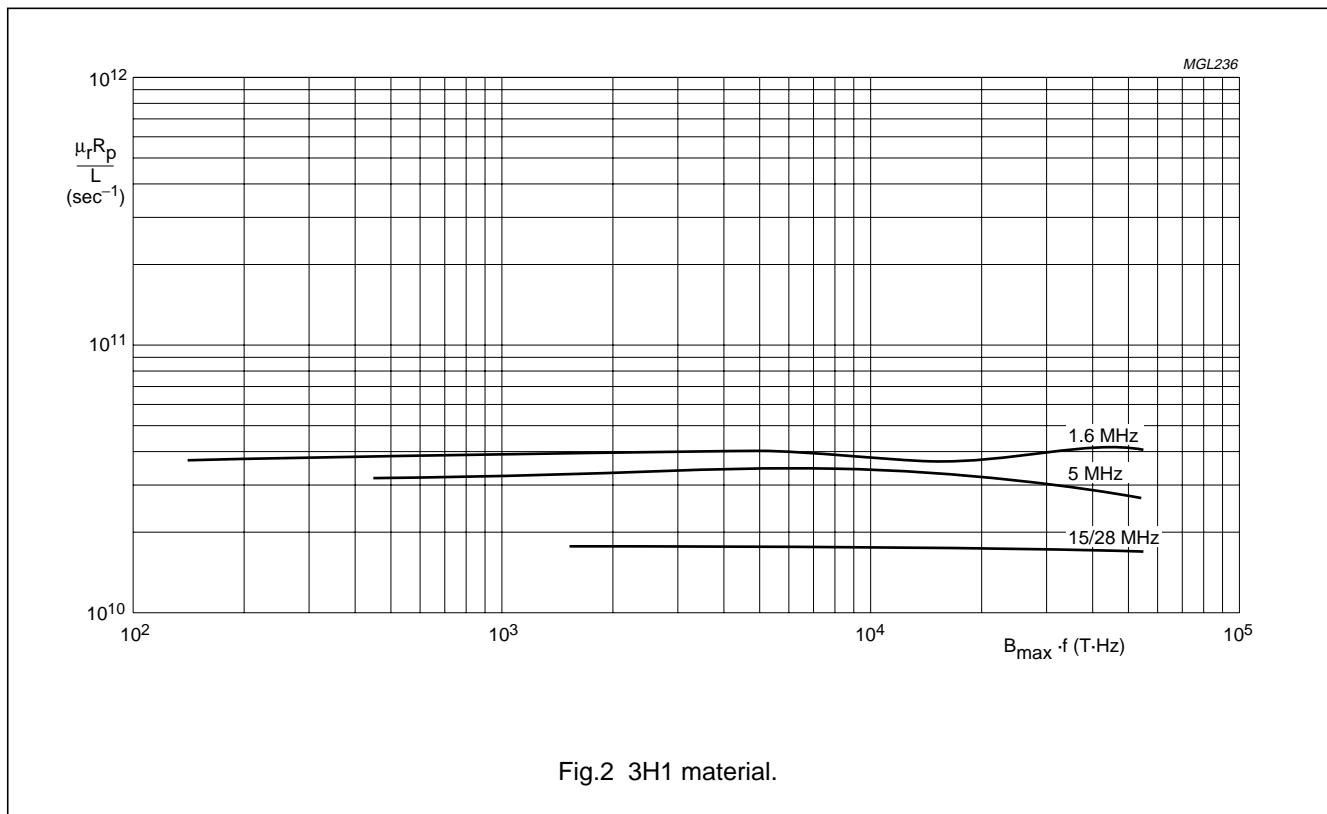
- The type of line
- The frequency
- The length.

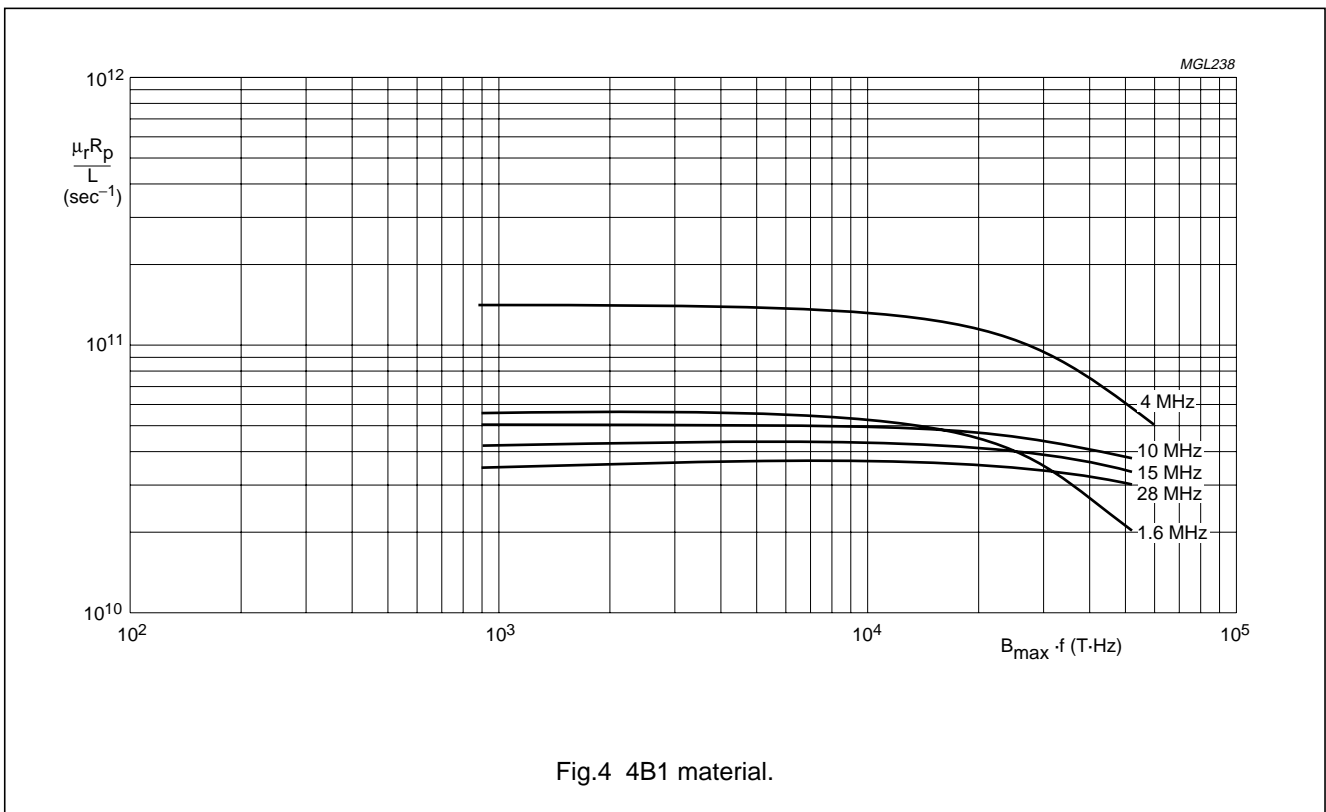
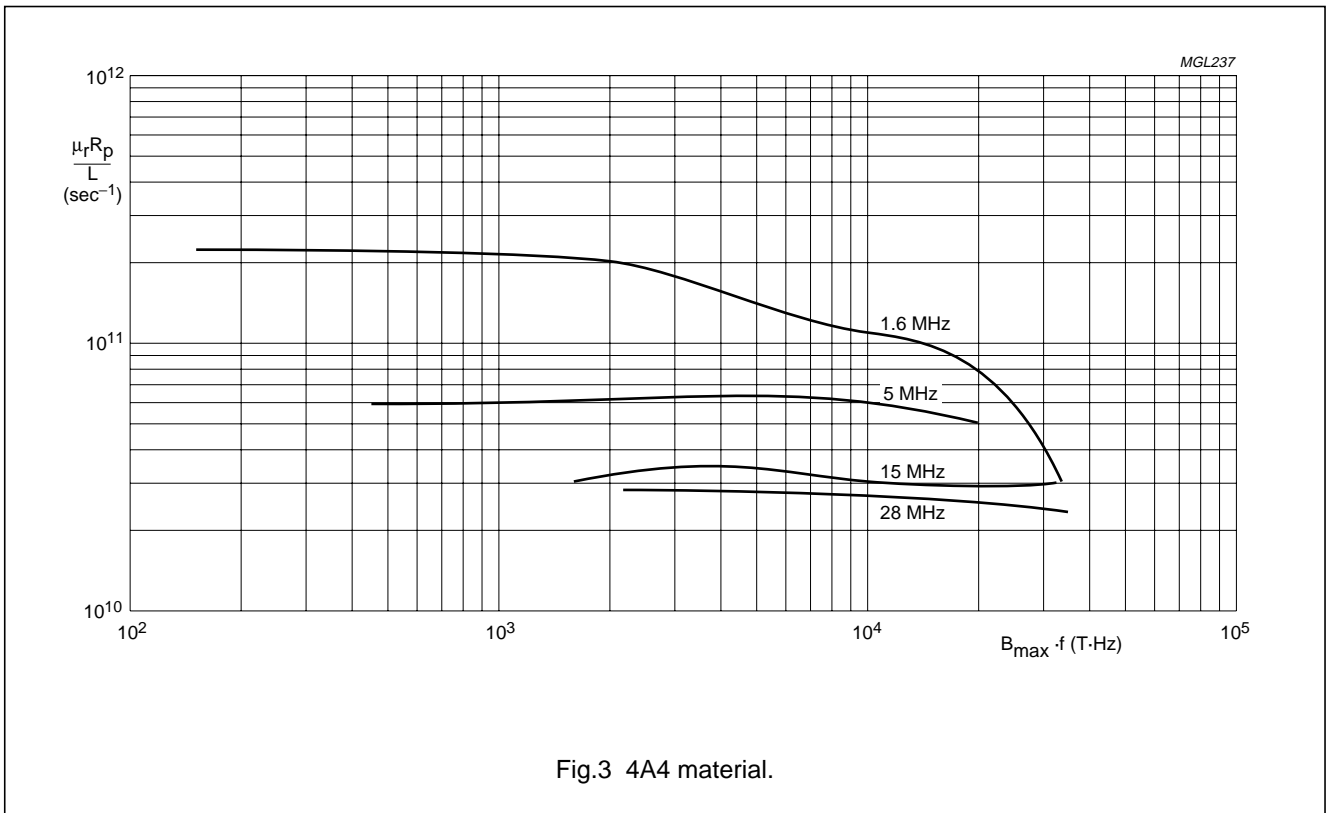
Data on power loss in some 50 Ω coaxial cables is given in Fig.8. This power loss and the allowable maximum cable temperature restrict the power handling of the cable. The maximum power which can be transmitted depends on the type of cable and the frequency; data is given in Fig.9.

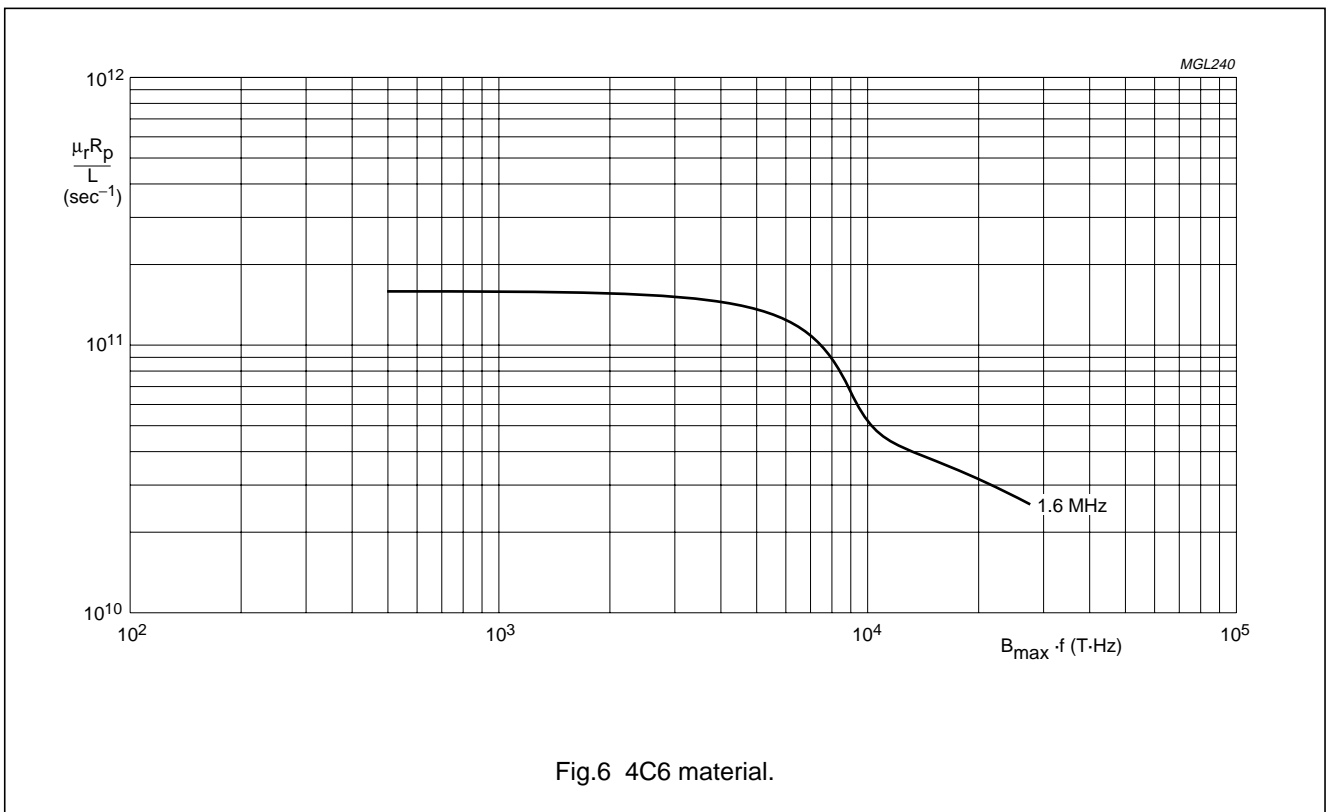
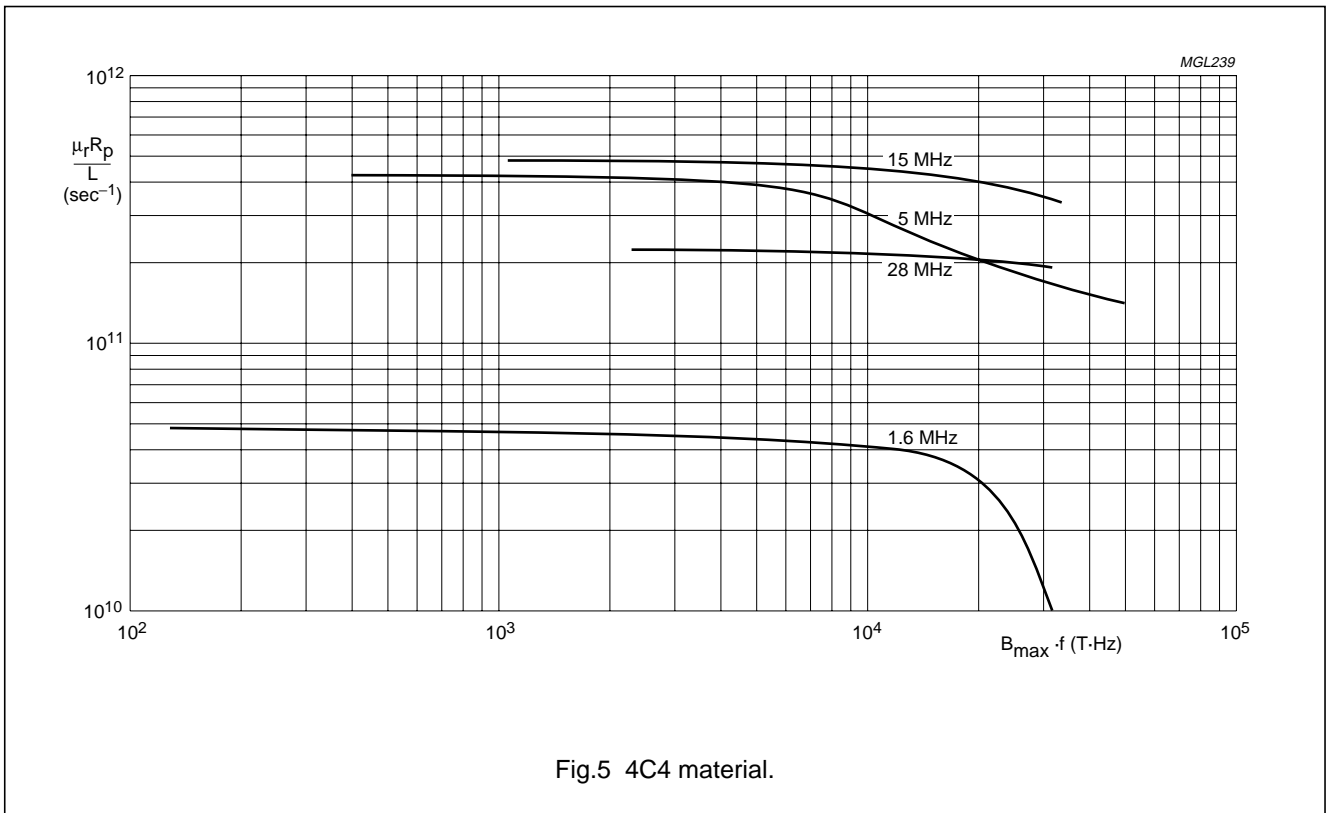
4.2 Mismatch loss

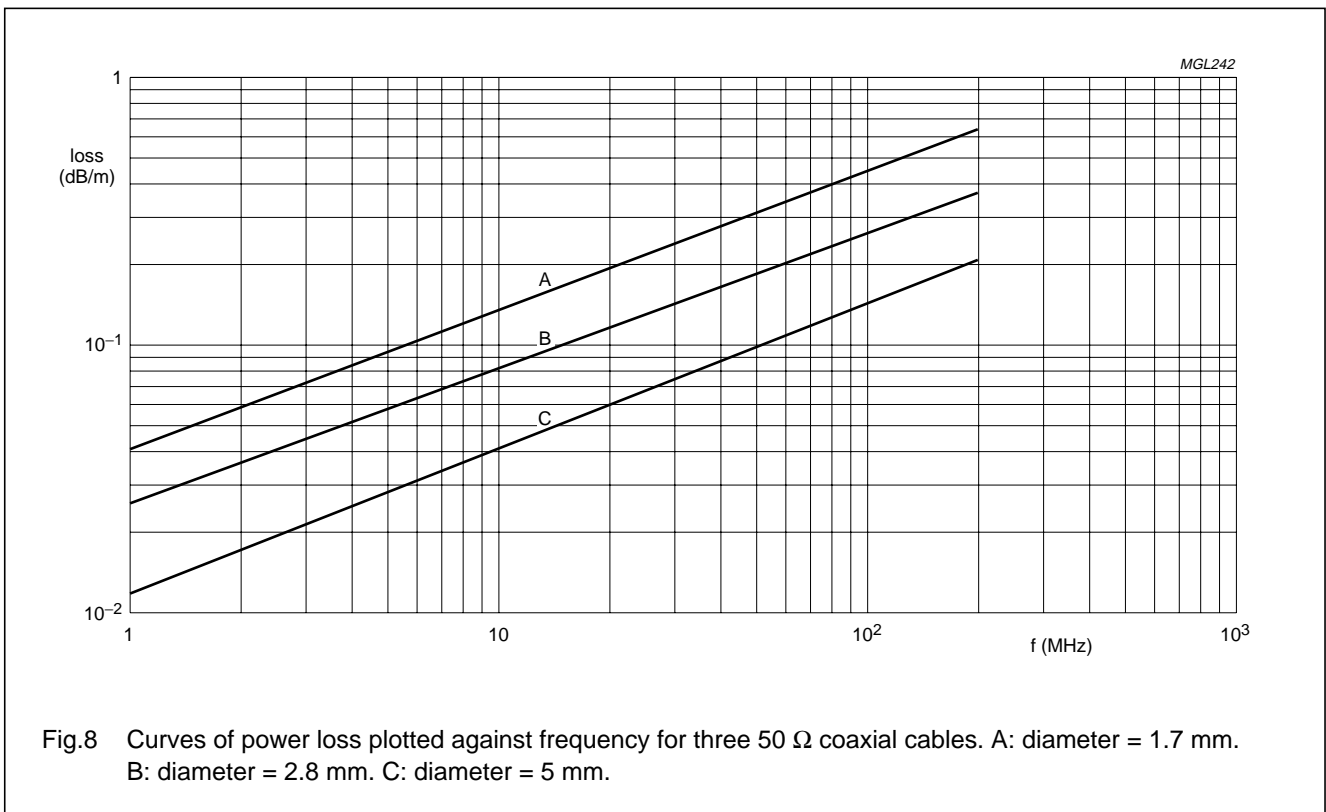
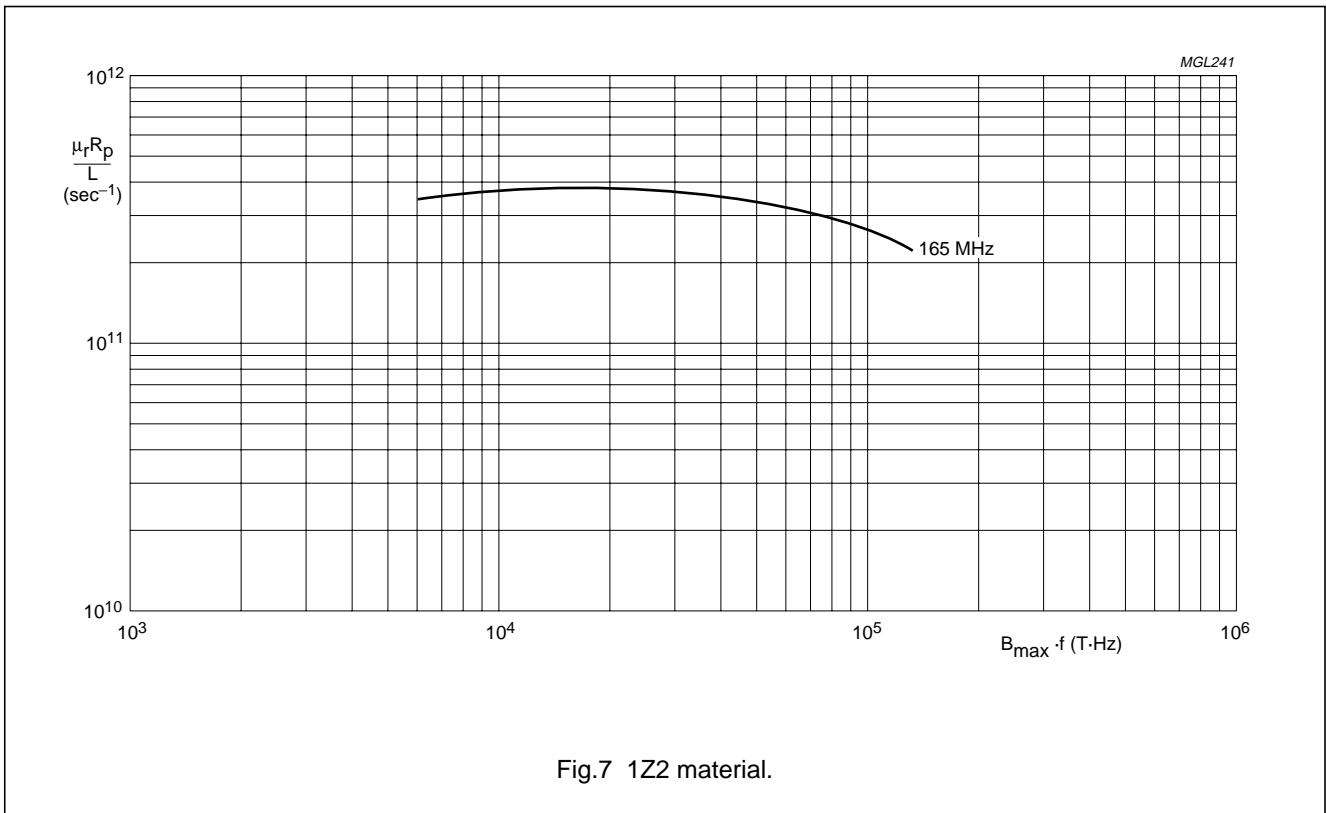
Another kind of loss caused by the transmission line can occur if the characteristic impedance of this line is not the required value. This results in a mismatch being maximum at the high frequency end of the band. The amount of mismatch depends on:

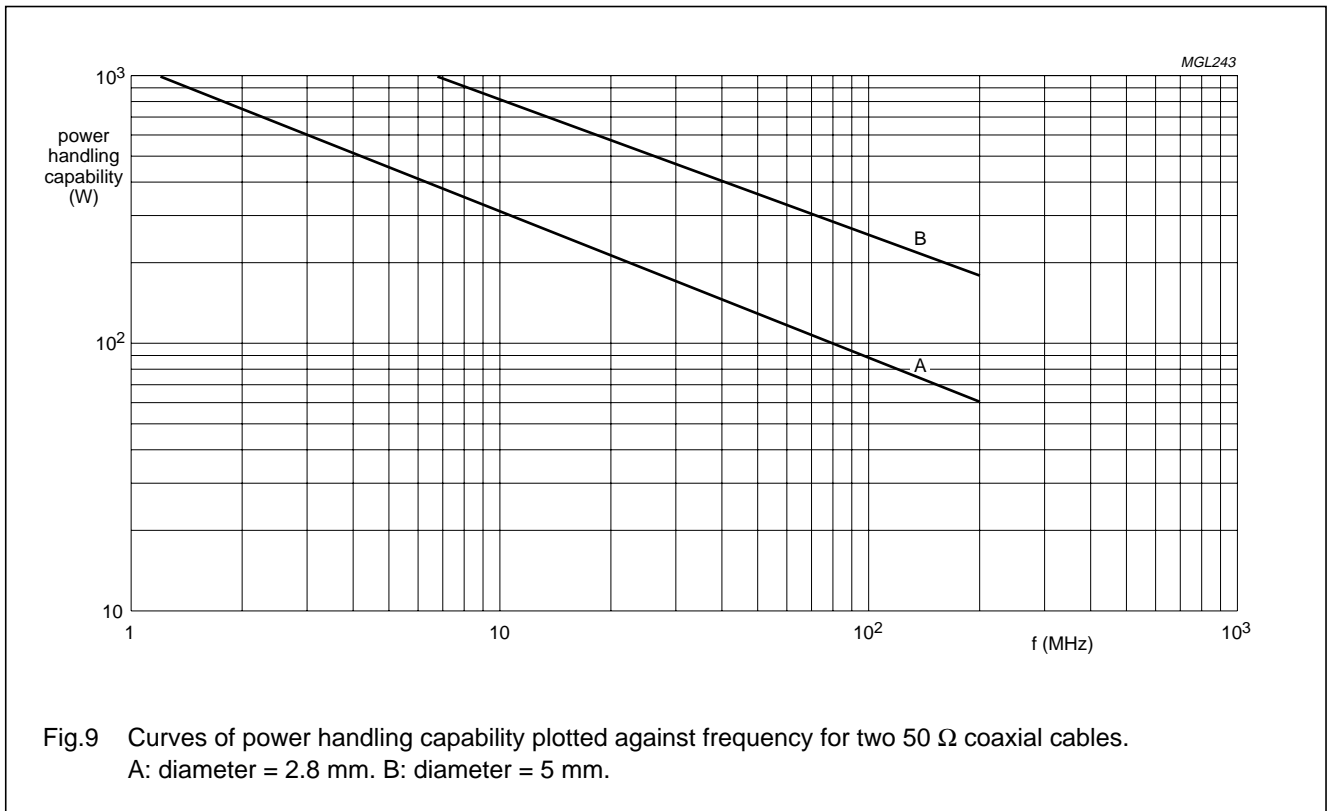
- The ratio between the length of the line and the wavelength on the line
- The ratio between the required and the actual value of the characteristic resistance.











From transmission line theory the input impedance is given by:

$$Z_{in} = R_{in} \times \frac{1 + jr \tan \beta l}{1 + j \frac{1}{r} \tan \beta l}$$

in which:

R_{in} = midband input resistance

r = ratio between the actual and the required characteristic resistance

$\beta = 2\pi/\lambda$

λ = wavelength on the line (for 50 Ω coaxial cables approx. 67-70% of wavelength in free space)

l = length of the line.

If the deviation of Z_{in} from the required value is unacceptably large it is in many cases possible to make use of the compensation technique described in Section 5.2.

5 COMPENSATION TECHNIQUES

5.1 Compensation at Low Frequencies

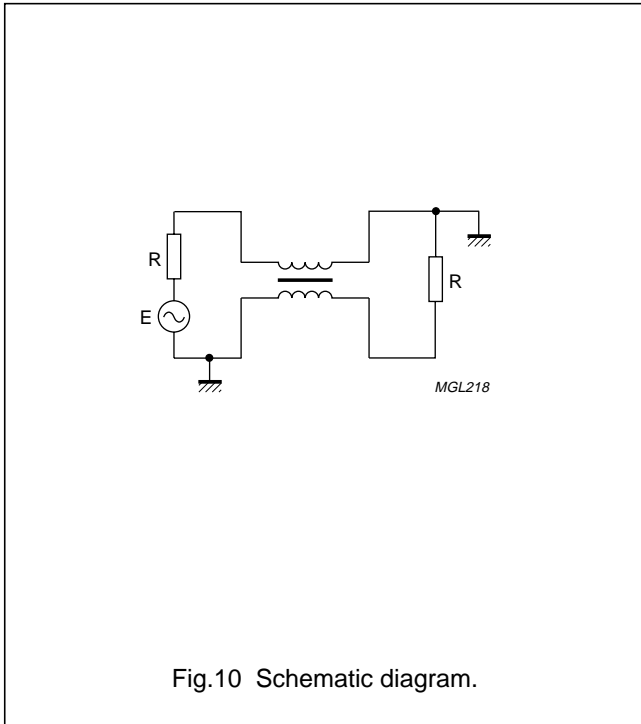


Fig.10 Schematic diagram.

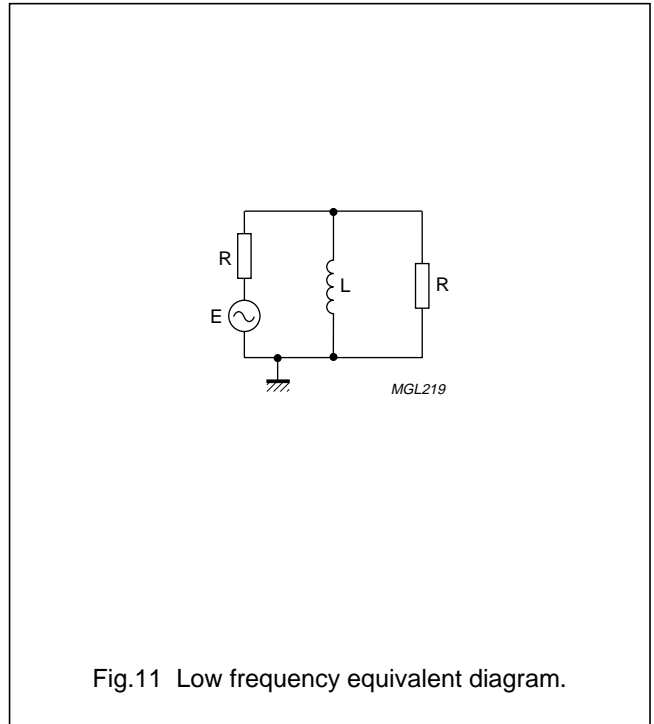


Fig.11 Low frequency equivalent diagram.

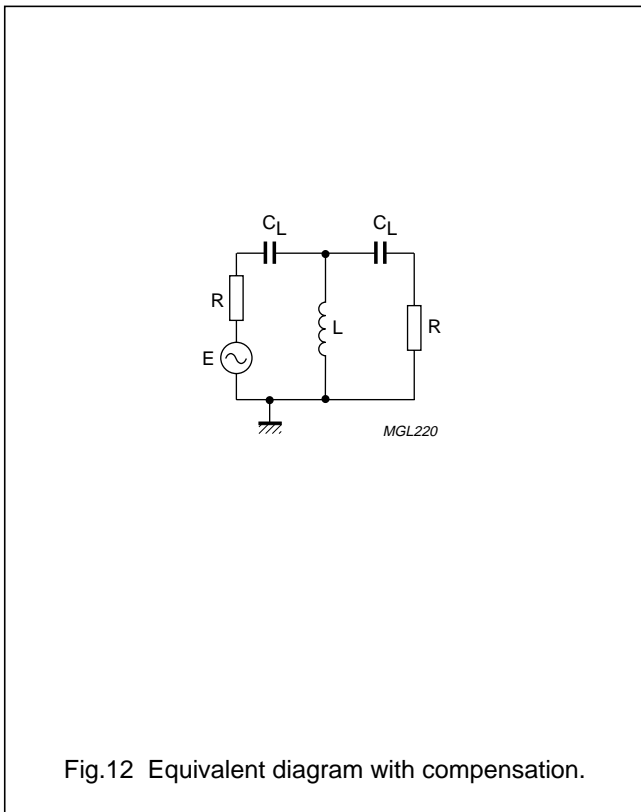


Fig.12 Equivalent diagram with compensation.

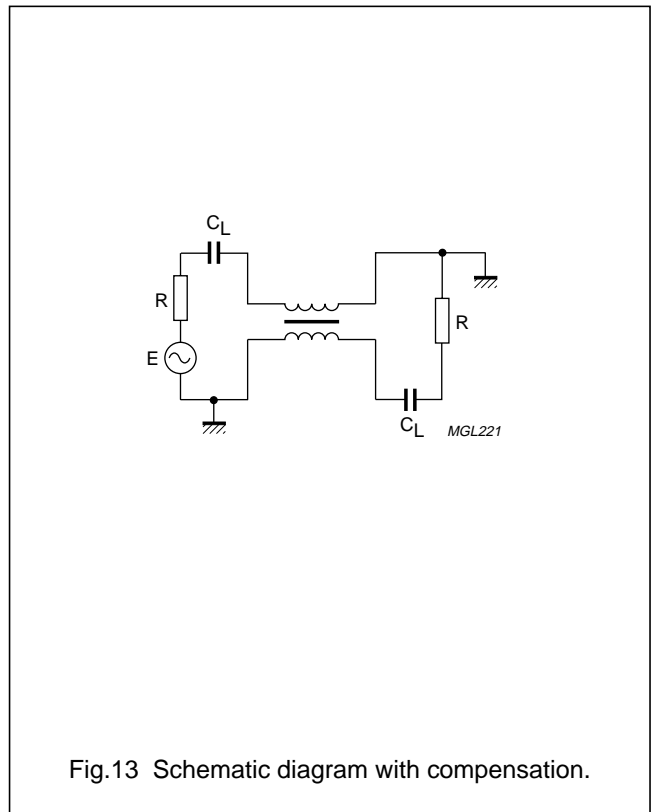


Fig.13 Schematic diagram with compensation.

Compensation will be illustrated by means of the phase reversing transformer. The schematic diagram is given in Fig.10 and the equivalent diagram for the low frequency end of the band is shown in Fig.11. For compensation we add two equal capacitors C_L such that a high-pass T-filter section is formed (Fig.12). According to filter theory:

$$C_L = 2L/R^2$$

The original diagram of Fig.10 is now transformed to the new one of Fig.13. If L is dimensioned according to eq.(2) the input impedance without compensation at the lowest frequency is $R/(+j4R)$. With compensation the input impedance is $(.999R/+j264R)$, illustrating that the mismatch has been reduced to a negligible level.

For some types of transformer, the capacitor at the output must have a different value from that at the input. With a $1 : n^2$ impedance transformer, for instance, it must be n^2 times smaller than the capacitor at the input.

Sometimes two or more transformers must be connected in cascade. In such a case low frequency compensation is possible if a high-pass π -filter section is used (Fig.14). When the parallel inductance of the transformers at the interconnecting point are both approximately equal to L , the capacitance of C_L must be $L/2R^2$.

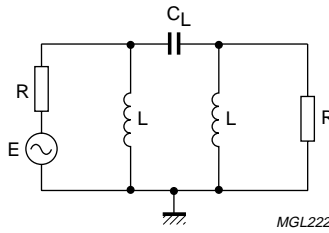


Fig.14 Equivalent diagram of cascaded transformers with compensation.

5.2 Compensation at High Frequencies

This is only necessary when the characteristic resistance of the transmission line differs from the required value. A situation often met with in practice is that in which the required characteristic resistance is lower than that of the available line. Taking the simple case of the phase reversing transformer (Fig.15) with a required characteristic resistance equal to R , we find that compensation for an actual value equal to $r \times R$ can be made as follows. In parallel with the load resistance, we connect a capacitor of such a value that at the highest frequency the real part of the input admittance becomes $1/R$. The resulting imaginary part of the input admittance is tuned out by means of a capacitor in parallel with the input. Both capacitors turn out to have the same value; given by

$$C_H = \frac{1}{\omega_{max} r R \tan \beta l} \{ 1 - \sqrt{1 - (r^2 - 1) \tan^2 \beta l} \}$$

in which $\omega_{max} = 2\pi$ times the maximum frequency. The schematic diagram is shown in Fig.15.

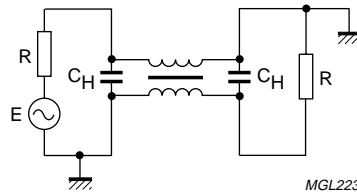


Fig.15 Phase reversing transformer with high frequency compensation.

The result of this compensation is an exact match at the maximum frequency. There will be however, a slight mismatch at lower frequencies which is many times smaller than that at the maximum frequency without compensation. A combination of high and low frequency compensation is of course possible.

6 TRANSFORMER CONFIGURATION

Because of the variety of the existing configurations it is hardly possible to give a complete survey. Therefore a restriction will be made to some principally different types.

6.1 Phase Reversing Transformer

This type has already been discussed in the previous section.

6.2 Balanced to Unbalanced Transformer

The schematic diagram is shown in Fig.16. This type can be considered as a modification of the phase reversing transformer. The primary inductance in this case is 4 times the inductance of the winding between the points A and B because of the voltage division. If low frequency compensation is used, a capacitor equal to $2C_L$ must be placed in series with each input connection (see Section 5.1) to preserve symmetry, and one capacitor equal to C_L in series with the output (point B).

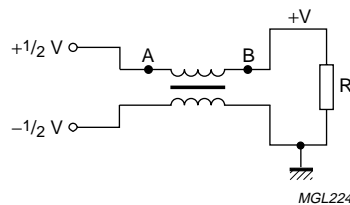


Fig.16 Balanced to unbalanced transformer.

6.3 Symmetrical⁽¹⁾ 1 : 4 Impedance Transformer

The schematic diagram is given in Fig.17. Two cables each having a characteristic resistance of $2R$, can be wound on a common core. The direction of the windings follows from the voltage division. Low frequency compensation can be made with a capacitor equal to $2C_L$ (see Section 5.1) in each of the input leads and a capacitor equal to $1/2C_L$ in each of the output leads.

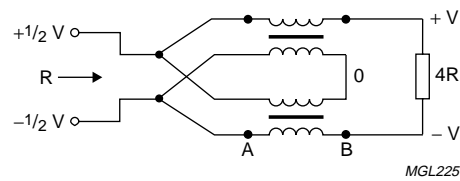


Fig.17 Symmetrical 1 : 4 impedance transformer.

6.4 Asymmetrical 1 : 4 Impedance Transformer

If, in the 1 : 4 impedance transformer of Section 6.3, the points A and B are connected to earth it is no longer possible to wind the two lines on a common core. In fact the lower line may be wound on an 'air core' because there is no voltage difference between the points A and B. The logical next step is to omit the lower line completely (with a consequent slight phase difference between the lines). Then we get the transformer shown in Fig.18. The characteristic resistance of the transmission line has again an optimum value of $2R$. But even if this value is chosen the input impedance is not constant as a function of frequency. From theory (see Ref.5) it is:

$$Z_{in} = R \times \frac{2\cos\beta l + jr\sin\beta l}{1 + \cos\beta l + j\frac{1}{r}\sin\beta l}$$

in which r = ratio between the actual characteristic resistance and $2R$.

(1) Terminology in normal use is employed to describe the various transformer configurations. However the terms 'symmetrical', 'balanced' and 'push-pull' are used here synonymously to mean 'antiphase port signals having equal amplitudes with respect to ground', while likewise 'asymmetrical', 'unbalanced' and 'single-ended' are synonymous with 'one port terminal grounded'. Strictly defined, the terms 'asymmetrical, and 'unbalanced' also apply to unequal port terminal values with respect to ground.

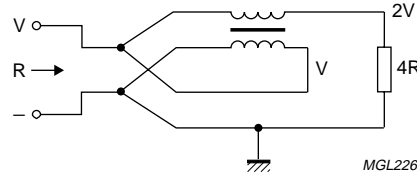


Fig.18 Asymmetrical 1 : 4 impedance transformer.

If $r > 1$ high frequency compensation is sometimes possible. A capacitor C_1 given by:

$$C_1 = \frac{1 + \cos\beta l - \sqrt{(1 + \cos\beta l)^2 - r^2 \sin^2 \beta l}}{\omega_{\max} r R \sin \beta l}$$

must then be connected across the input and a capacitor C_2 given by: $C_2 = \frac{2 + \cos\beta l - \sqrt{(1 + \cos\beta l)^2 - r^2 \sin^2 \beta l}}{4\omega_{\max} r R \sin \beta l}$

must be connected across the output.

6.5 Symmetrical 9 : 1 Impedance Transformer

The schematic diagram is given in Fig.19. The two transmission lines have an optimum characteristic resistance of $3R$. They can again be wound on a common core, and the third line omitted as in the previous case. The input impedance is

given by: $Z_{in} = 9R \times \frac{4 + 5\cos\beta l + j6r\sin\beta l}{9\cos\beta l + j\frac{6}{r}\sin\beta l}$

in which r is the ratio between the actual characteristic resistance and $3R$.

If $r > 1$ high frequency compensation is sometimes possible. A capacitor C_1 given by:

$$C_1 = \frac{4 + 5\cos\beta l - \sqrt{9(3\cos^2\beta l + 4\cos\beta l + 2) - 36r^2 \sin^2 \beta l}}{6\omega_{\max} r R \sin \beta l}$$

must than be connected across the low impedance side, and a capacitor C_2 , given by:

$$C_2 = \frac{9\cos\beta l - \sqrt{9(3\cos^2\beta l + 4\cos\beta l + 2) - 36r^2 \sin^2 \beta l}}{54\omega_{\max} r R \sin \beta l}$$

must be connected across the high impedance side.

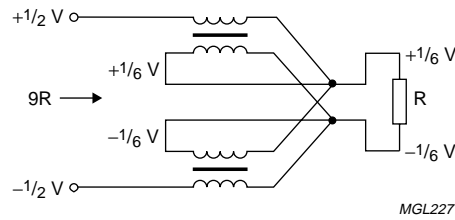


Fig.19 Symmetrical 9:1 impedance transformer.

6.6 Asymmetrical 1 : 9 Impedance Transformer

In this transformer (see Fig.20), two transmission lines are required, each having an optimum characteristic resistance of $3R$. Although these lines can be wound on a common core, the upper line must have twice the number of turns of the lower line, because of the voltage division. The third line has again been omitted.

High frequency compensation follows the same principle as that used in previous sections.

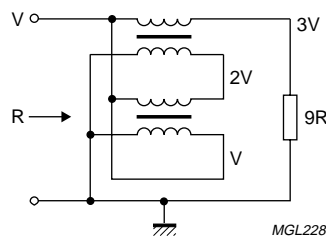


Fig.20 Asymmetrical 1 : 9 impedance transformer.

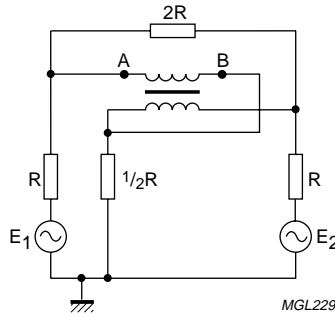


Fig.21 Single-ended hybrid.

6.7 Single-ended Hybrid

This circuit (see Fig.21) permits the combination of two signals in a common load ($R/2$) in such a way that the signal sources do not influence each other. When the two signals have different frequencies the power is equally divided between the resistors $R/2$ and $2R$. When the two signals have the same frequency, phase and amplitude, all the power is delivered to $R/2$. The optimum value of the characteristic impedance of the transmission line is equal to R . At the low frequency end of the band the isolation between the sources depends on the inductance of the winding. When the isolation is expressed as a power ratio S the inductance between the points A and B must be: $L = \frac{R}{8\omega_{\min}} \sqrt{S-1}$

Even when the optimum characteristic impedance is chosen, high frequency compensation has been found to have a marked influence on the isolation. Practically it has been found that a small capacitor must be connected in parallel with $R/2$.

To illustrate the degree of isolation that may be expected in a practical design, an example is given of a hybrid transformer designed for the frequency band of 1.6 to 28 Mhz. The requirement was to combine two signals of different frequency and equal power (3 W in 100 Ω) in a single 50 Ω load with a minimum isolation of 40 dB between the sources.

From the equation $L = \frac{R}{8\omega_{\min}} \sqrt{S-1}$, it follows that L must be 125 μH minimum. Because of this high inductance it was decided to choose 3H1 material. An available $23 \times 14 \times 7$ mm toroid belonged to 'group 5' which meant that the minimum μ_r was 2680⁽¹⁾.

The required number of turns was 9. The optimum value of the characteristic resistance is 100 Ω , but for reasons of convenience 150 Ω miniature twin lead was used. The calculated value of B_{\max} was 87 gauss at 1.6 MHz. High frequency compensation was achieved by means of a 33 pF capacitor in parallel with the 50 Ω load resistance. The isolation between the sources measured as a function of frequency is given in Table 1 below, illustrating that the minimum isolation requirement of 40 dB has been fulfilled over the desired frequency band.

(1) For the convenience of the user the toroids of ferroxcube 3H1 are delivered sorted into groups of approximately equal μ -value. The μ -value is indicated by the colour of the circumference of the toroids, see "Data Handbook System". Groups are not separately available.

Table 1

f (MHz)	ISOLATION (dB)
0.5	33
1	41
1.6	42.5
5	40.5
15	41
28	43
40	32

6.8 Push-pull Hybrid

Two different versions will be discussed. In the first one the impedance level is stepped-up by a factor of two and in the second the impedance level is stepped-down by a factor of two.

6.8.1 IMPEDANCE STEP-UP TYPE

The two transmission lines (see Fig.22) have an optimum characteristic resistance equal to R and they can be wound on a common core.

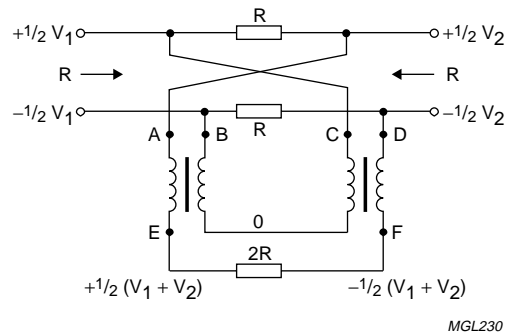


Fig.22 Push-pull hybrid with impedance step-up.

A hybrid can always be used in two ways. The first way has been mentioned in Section 6.7. The second way is the inverse of the first i.e. the power from a single source is equally divided between two loads which are isolated from each other. The latter arrangement allows a more convenient calculation of the lower frequency limit. To do this we must know the effective inductance in parallel with $2R$. The voltage across this resistor is 4 times the voltage between the points D and F and therefore the effective inductance in parallel with $2R$ is 16 times the inductance between the points D and F. For low frequency compensation the effective inductance in parallel with $2R$ can be considered as two equal inductances in series – one between points E and O and one between points F and O – each having a value of 8 times the inductance between points D and F. The hybrid can then be considered as two balanced to unbalanced transformers on a common core and the compensation can be made as described in Section 6.2.

6.8.2 IMPEDANCE STEP-DOWN TYPE

This hybrid (see Fig.23) is identical to the previous one except that the transmission lines on one side have been connected in parallel instead of in series. When this hybrid is used for the combination of two signals which have the same frequency, phase and amplitude, the ferrite core will not be magnetized. However when V_2 becomes zero, the voltage across each winding will be $V_1/4$ and the voltage across $R/2$ will be $V_1/2$. This means that the effective inductance in parallel with $R/2$ becomes 4 times the inductance of a single winding. A similar approach must be used in finding the winding V/turn when calculating B_{max} .

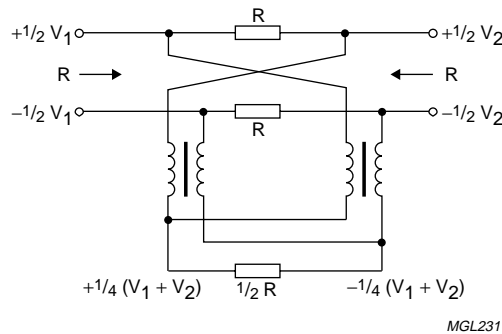


Fig.23 Push-pull hybrid with impedance step-down.

7 PRACTICAL EXAMPLES

It is possible to combine some of the functions, mentioned in Chapter 6, in one transformer. In the two examples given simple calculations show the method used in designing for a particular application.

7.1 12.5 Ω Balanced to 50 Ω Unbalanced Transformer

In this case, the transformers described in Sections 6.2 and 6.3 have been combined. The schematic diagram is shown in Fig.24. A transformer of this type is required for a wideband s.s.b. transmitter in the frequency range of 1.6 to 28 MHz. It must be able to handle a peak envelope power of 80 W. The input impedance of 12.5 Ω balanced and the output impedance is 50 Ω unbalanced. The total amount of resistive and reflection losses is required to be below 5% (power).

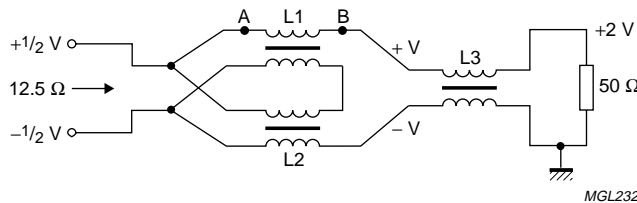


Fig.24 12.5 Ω balanced to 50 Ω unbalanced transformer.

The transformer has been wound on a single 4C4 toroid of $36 \times 23 \times 15$ mm. Windings L1 and L2 must have a characteristic resistance of 25Ω ; they consist of two 50Ω coaxial cables of 2.8 mm diameter in parallel. Winding L3 must have a characteristic resistance of 50Ω ; a single 50Ω coaxial cable of 2.8 mm diameter has been used. The winding direction follows from the voltage division. From this it is also clear that windings L1 and L2 must have an equal number of turns and that winding L3 must have twice the number of winding L1.

We can calculate the lower frequency limit from the effective inductance in parallel with the 50Ω load resistance. We allow a reactance of $+j 200 \Omega$ at 1.6 Mhz, corresponding to an inductance of $20 \mu\text{H}$, in parallel with the 50Ω load (see Section 3.1). As the voltage between points A and B is $\frac{1}{4}$ of the output voltage, the inductance between these points will be $20/16 = 1.25 \mu\text{H}$. The corresponding number of turns is:

$$n = \sqrt{\frac{L \times 1}{\mu_0 \mu_r A}}$$

in which $1/A = 9.42 \text{ cm}^{-2}$ and μ_r is taken to be 100. This gives:

$$n = 3.06$$

We take $n = 3.5$ turns, making the inductance in parallel with the 50Ω load resistance equal to:

$$20 \left(\frac{3.5}{3.06} \right)^2 = 26.2 \mu\text{H minimum}$$

The measured value was actually $39 \mu\text{H}$, because the μ_r of the core used was greater than 100.

The maximum flux density B_{max} can be calculated when the maximum voltage across the windings is known. The peak value of the voltage across the 50Ω load resistance for an 80 W power level is:

$$V_p = \sqrt{2 \times 80 \times 50} = 89.5\text{V}$$

and between points A and B: $89.5/4 = 22.35 \text{ V}$. The ferrite cross section for this core is $0.976 \times 10^{-4} \text{ m}^2$. So at 1.6 MHz the maximum flux density is:

$$B_{\text{max}} = \frac{V_{\text{max}}}{\omega A n} = \frac{22.35}{2 \times \pi \times 1.6 \times 10^6 \times 0.976 \times 10^{-4} \times 3.5} = 65.4 \times 10^{-4} \text{ T or } 65.4 \text{ gauss.}$$

This corresponds to a $B_{\text{max}} \times f$ product of $1.05 \times 10^4 \text{ T} \cdot \text{Hz}$. From Fig.5 it can be seen that up to this level the variation in core losses is only very small. The minimum value of the parallel loss resistance referred to the 50Ω level (see Fig.5) is:

$$R_p = \frac{L}{\mu_r} \times 41 \times 10^9 = \frac{26.2 \times 10^{-6}}{100} \times 41 \times 10^9 = 10.7 \text{ k}\Omega$$

So the core losses are: $50/10700 \times 100\% = 0.47\%$ maximum.

The cable can handle a power of 170 W at 28 MHz and it has a loss of 0.135 dB/m at this frequency. The length of the cable (the sum of L_1 and L_3) is about 60 cm, giving a maximum cable loss of approximately 1.9%.

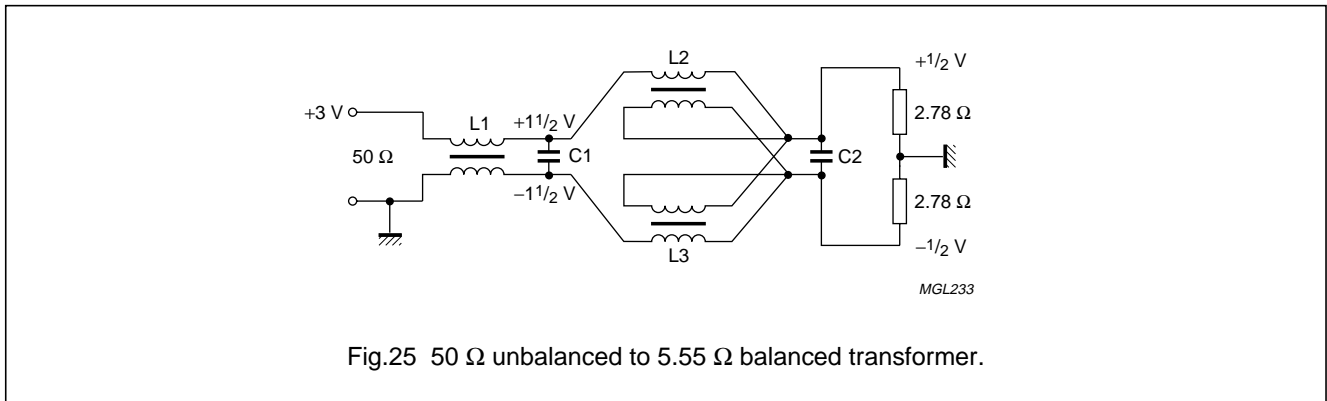
When the transformer was driven from an unbalanced 50Ω source and loaded with two resistors of 6.25Ω at the balanced side (connecting point of the resistors earthed) the power loss and asymmetry were so small that they could not be measured. The input impedance at the unbalanced side as a function of frequency is given in Table 2.

Table 2

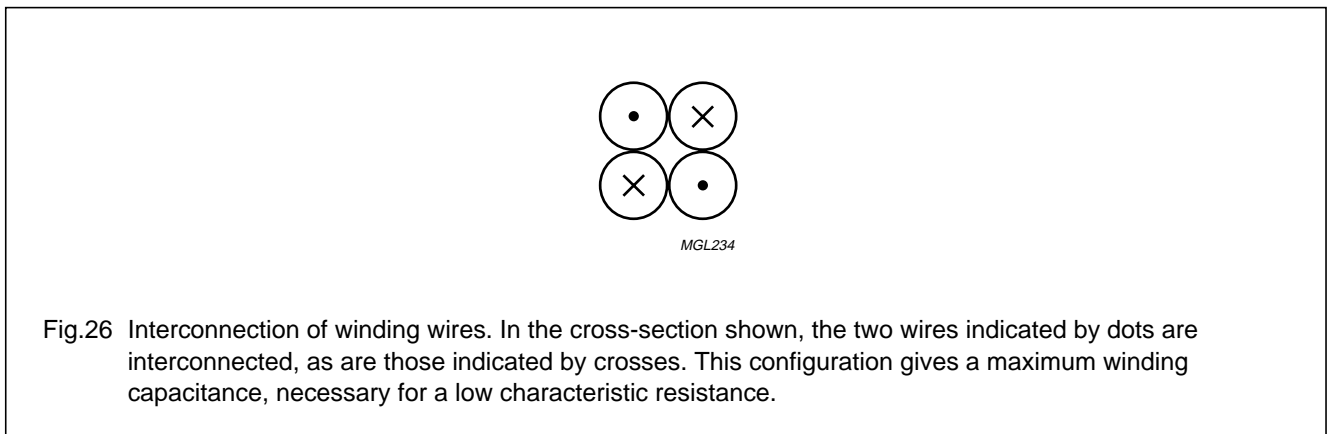
f (MHz)	R _p (Ω)	X _p (Ω)
1.6	49.3	+j 395
5	49.6	+j 1250
15	50.2	+j 21200
28	50.2	-j 1060

7.2 50 Ω Unbalanced to 5.55 Ω Balanced Transformer

In this case the transformers described in Section 6.2 and 6.5 have been combined. The schematic diagram is given in Fig.25. A transformer of this design is again required for a wideband s.s.b. transmitter in the frequency range of 1.6 to 28 MHz. The required peak envelope power handling is 5 W, with an input impedance of 50 Ω unbalanced and an output impedance of 5.55 Ω balanced.



The transformer can be wound on a single 4C4 toroid of 23 × 14 × 7 mm. Winding L₁ must have a characteristic resistance of 50 Ω; this has been approached by a twisted pair of enamelled copper wires of 0.45 mm diameter (R_c = 52 Ω). Windings L₂ and L₃ must have a characteristic resistance of 16²/₃ Ω. For this purpose four enamelled copper wires of 0.22 mm diameter have been twisted and diagonally interconnected as shown in Fig.26; the characteristic resistance was found to be 30 Ω, making a high-frequency correction necessary. The required compensation, determined experimentally gave values for C₁ of 22 pF and for C₂ of 270 pF.



From the voltage division it follows that winding L_1 must have $1\frac{1}{2}$ times the number of turns of windings L_2 and L_3 . The required number of turns can be calculated in the same way as in the previous section. To get a parallel input inductance of 20 μH minimum, 9 turns are required for winding L_1 . The measured value was approx. 30 μH .

Calculation of B_{max} according to the method of Section 7.1 gives a value of approx. 40 gauss at 1.6 MHz. This is so low that a smaller core could be chosen. The reason for not doing this is that the smaller core has not sufficient room for the windings. In Table 3 the input impedance is given as a function of frequency.

Table 3

	WITHOUT H.F. COMPENSATION		WITH H.F. COMPENSATION	
	R_p (Ω)	X_p (Ω)	R_p (Ω)	X_p (Ω)
1.6	49	+j 300	48	+j 370
5	50	+j 360	49	+j 840
15	56	+j 220	50	+j 1400
28	71	+j 213	51	-j 11000

If there is no high frequency compensation, the variation in impedance is rather large. This is caused by:

- The characteristic resistance of the transmission lines L_2 and L_3 being 30 Ω instead of $16\frac{2}{3}$ Ω .
- The electrical length of the windings L_2 and L_3 . Measurements have shown that the wave-length on these lines is only 41% of the wavelength in free space – the decrease in wave-length being caused by the twisting of the wires, the dielectric constant of the wire insulation and the influence of the μ_r of the ferrite core.
- The inductance of the load resistors.

The power loss measured at 28 MHz was 0.34 dB. At lower frequencies it was too small to be measured accurately. Voltage asymmetry between both terminals was less than 4%.

8 REFERENCES

1. W. Hilberg;
Einige grundsätzliche Betrachtungen zu Breitband-Übertragern; N.T.Z. 1966, Heft 9, p. 527 - 538.
2. A. Muser, J. Zen, R. Armbruster;
Transformateurs d'impulsions rapides a lignes de transmissions; L'onde électrique No. 446, Mai 1964, p. 520 - 529.
3. H. Nielinger;
Optimale Dimensionierung von Breitbandanpassungsnetzwerken; N.T.Z. 1968, Heft 2, p. 88 - 91.
4. O. Pitzalis, Jr., T.P.M. Couse;
Practical Design Information for Broadband Transmission Line Transformers; Proc. IEEE (Letters) April 1968, p. 738 - 739.
5. C.L. Ruthroff;
Some Broad-Band Transformers; Proc. IRE August 1959, p. 1337 - 1342.