RF Transformers Part 2: The Core

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In the first part of this article (June 1995, pg. 36) the different winding topologies for wideband, ferrite-loaded transformers were introduced. This article concentrates on selecting core shapes and material types. [Note: Figure numbering is continued from part 1 of this article, ed.]

Some ferrite manufacturers sell "designer's kits" of cores with a wide variety of shapes, sizes and materials. I suspect that these are not really used as designer's kits, but as bodger's kits, where the idea is to keep on winding transformers until one works. This part of the article will encourage a better engineering approach to the subject.

Transformer Loss

Imagine a transmitter connected to a load using a length of coaxial cable. It is clear that winding the cable onto a ferrite core will cause no additional signal loss: this is because the current flowing on the inner conductor exactly cancels the current flowing in the outer conductor so there is no magnetic flux in the core.

Now imagine that the case of the load is RF 'hot'. There will now be a small out of balance current flowing through the windings. This is referred to in the text books as the magnetizing current. Because it causes a magnetic flux in the core, it will have losses associated with it.

For simplicity, a simple Guanella phase inverter wound on a Philips type RCC14/5-4C65 core will be discussed first. The core is torroidal, with an external diameter of 14.5mm and is made of 4C65 grade material. The circuit diagram of this transformer is shown in Figure 17. Physically, this transformer consists of a few turns of miniature coaxial cable (type MCX) wound on a torroidal core. At the center of the winding, the inner conductor is connected to the outer of the second half of the winding, and vice-versa.

Figure 17 also shows a lumped element equivalent circuit of the transformer. Its low frequency loss is dominated by the low reactance of L_P, the winding inductance. This is well covered in the literature. In contrast, authors have had considerable difficulty in predicting the mid-band insertion loss and high frequency performance of transformers. The mid-band loss is dominated by R_P, the parallel equivalent core loss. The high frequency loss was discussed in Part 1, but it is sometimes dominated by C, the overall winding capacitance. To predict the transformer loss, values for L_P, R_P and C must be estimated. In the following paragraphs, the values of L_P and R_P change with frequency.

Predicting anything about inductors and transformers is notoriously diffi-

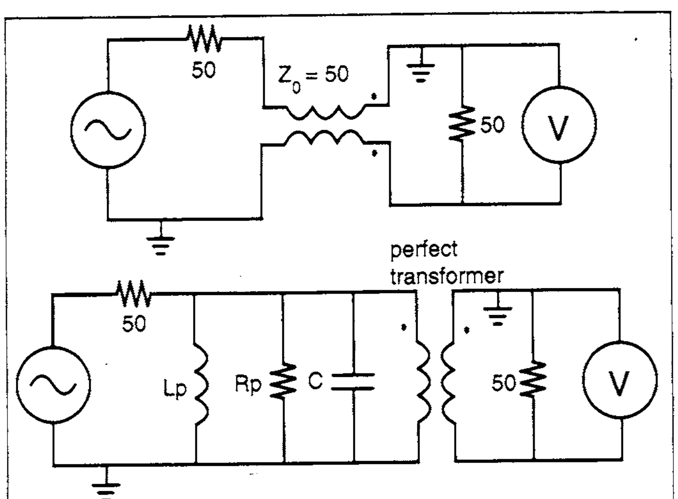

Figure 17. The Guanella inverter with a lumped element equivalent circuit.

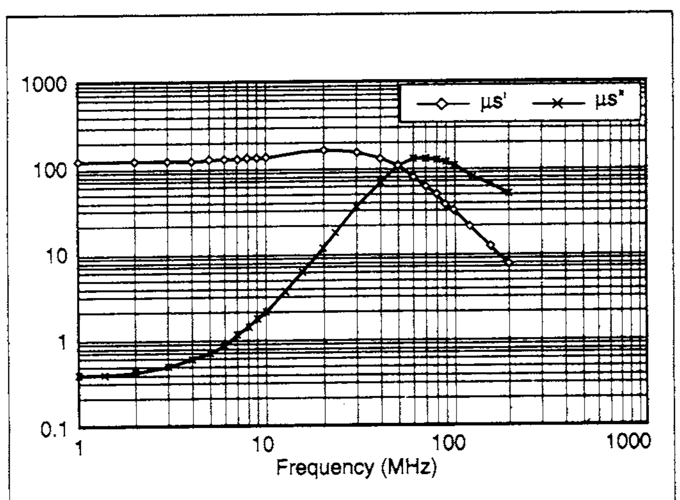

Figure 18. Complex permeability spectrum of Philips 4C65 material.

cult. To simplify the problem, make some assumptions:

a) The Q of the coaxial choke is dictated solely by the core material
b) The windings have no losses
c) There is no dielectric loss in the core
d) Small signal conditions apply at all times

Be aware that this is skating on rather thin ice, but proceed with caution.

Normalized L and R

To estimate L_P and R_P of an inductor, start with the manufacturer's graph of the complex permeability spectrum for the material to be used. Amongst others, Neosid, Philips and Siemens publish their data in this format.

Put in simple terms, the impedance of an inductor, Z_L is proportional to $+j\mu$, where μ is the permeability of the material, and is one of the factors which determines the inductance of a coil with a magnetic core. The permeability can itself be represented as a complex number. This complex permeability is given by:

$$\mu = \mu' - j\mu''. \qquad (1)$$

Substituting one formula into the other gives,

$$Z_L \propto \mu'' + j\mu'. \qquad (2)$$

So μ_s' is related to the coil's inductance and $\omega\mu_s''$ is related to the coil's resistance, where the subscript s indicates that the two components are connected in series. These can also be recalculated and expressed as parallel components μ_p' and μ_p''. For this, and subsequent calculations, the use of a computer spreadsheet takes a lot of work out of the job.

Figure 18 shows the graph of the series complex permeability spectrum for Philips 4C65 ferrite. The data have been extracted from [10], but the LF data for μ_s'' have been extrapolated below 8 MHz. This will be discussed later.

The curves for μ_s' and μ_s'' cross at f_x, which for Figure 18 is 50 MHz. At this frequency, an "inductor" wound on this core would have a Q of about 1. At a higher frequency, the impedance of the "inductor" would be almost purely resistive. However, for a low power transmission line transformer, the only requirement is that the bifilar choke should have a high impedance, never mind whether the impedance is resistive or reactive.

The manufacturer has not specified the full complex permeability above 200 MHz, however the ferrite, used in a transformer core, will work up to at least GHz. There are two reasons for this:

a) The complex permeability data are measured on a core of stated size, however the permeability becomes more dependent on the core shape and size as the frequency is raised above f_x.

b) The data on which Figure 18 is based were intended for designers of inductors. Above f_x, an inductor will have a Q of less than 1, and therefore be useless.

Figure 19 shows the same data as Figure 18, but re-calculated in terms of normalized parallel components, R_{norm} and X_{norm} (in Ωmm/turn^2). The values are given by:

$$R_{norm} = 8\pi^2 f\left(\mu_S'' + \frac{\mu_S'^2}{\mu_S''}\right)\times 10^{-4} \qquad (3)$$

$$X_{norm} = 8\pi^2 f\left(\mu_S' + \frac{\mu_S''^2}{\mu_S'}\right)\times 10^{-4} \qquad (4)$$

where f = frequency in MHz.

(All the formulae given in this article are based on those given by Snelling [2].)

Some manufacturers (e.g. Fair-Rite and Ferronics) give data for transformer core materials in graphs similar to Figure 19.

Note that $X_{norm} \propto f$. This is because the normalized parallel inductance is approximately constant. Also note that, above 20 MHz, the value of R_{norm} is approximately constant with frequency. This is important.

The shape of the graph in Figure 19 is typical of a ferrite designed for inductors with a moderate Q. For lower Q ferrites (e.g. for transformers), the bulge in the spectrum of Rnorm above 100 Ωmm/turn^2 may be less pronounced. For ferrites for RFI suppression beads, the Q of the device is of no importance, and the spectrum of R_{norm} may rise continuously, without having any maximum value.

Constanst R_{norm}?

Many manufacturers publish complex permeability data for their ferrites over a limited frequency range. Some do not publish them at all. In either case, a crude, but useful rule of thumb is that near, and above f_x, the value of R_{norm} is approximately constant at about 60 Ωmm/turn^2. This does not apply for frequencies much

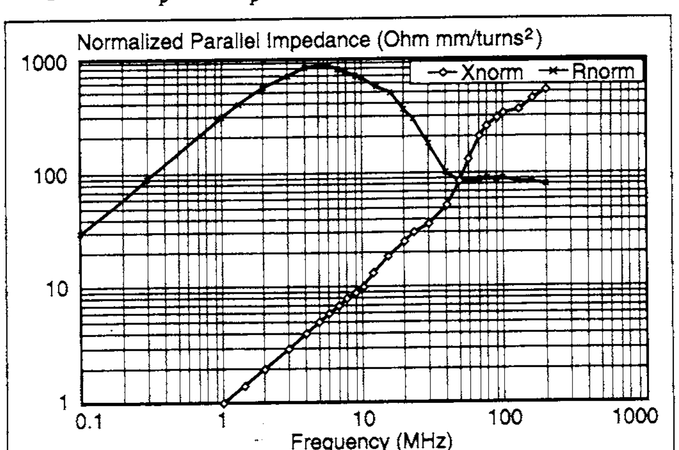

Figure 19. Normalized resistance and reactance for Philips 4C65 material.

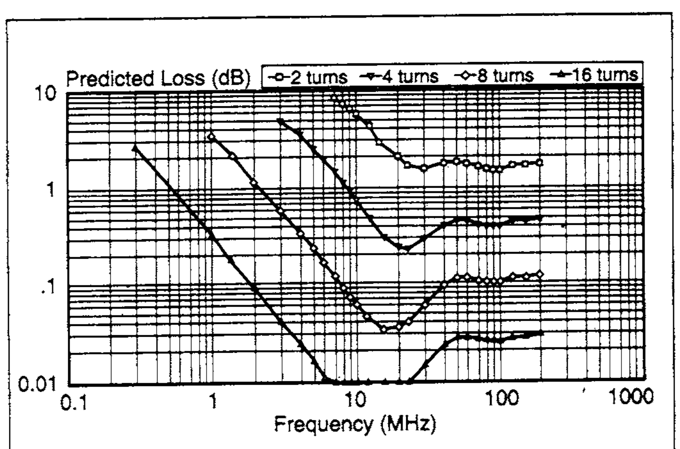

Figure 20. Predicted loss for different numbers of turns on RCC 14/5-4C65 toroid.

less than f_x.

For narrow-band use, (one decade or less), more careful selection of the material type based on the complex permeability data can result in designs with much higher values of R_{norm}. This gives a higher resistance per turn, but will only be necessary for very high power or high impedance transformers.

The complex permeability data indicate that ferrites with a high Cobalt content [11] have $R_{norm} > 60$, but these cannot be used for power applications. I have not tested them.

It can now be seen why radio amateurs have experimented with such a wide range of different materials for their transformers, and with so little definite conclusion: the main factor affecting transformer loss is Rp, and this is approximately independent of the core material.

Conventional wisdom says that a ferrite material with a small value of μ_i should be selected if the transformer loss is to be small. However, this means that the number of turns must be increased in order to keep the same value of primary inductance. This leads to a satisfactory result, because $Rp \propto N^2$. However, the same number of turns with a higher μ_i ferrite would have given a similar reduction in loss, but would give superior low frequency performance.

Finding L_P and R_P

To de-normalize R_{norm} and X_{norm}, use

$$R_P = \frac{N^2 R_{norm}}{C_1}\,\Omega \tag{5}$$

$$X_P = \frac{N^2 X_{norm}}{C_1}\,\Omega \tag{6}$$

where N = number of turns
C1= core factor in mm^{-1} (2.84 for RCC14/5 core in example). (Note that there is another core factor: C2. It is not used in this article)

The inductance is given by

$$L_P = X_P/\omega. \tag{7}$$

It should be roughly equal to the inductance calculated from the manufacturer's published value of A_L.

Some manufacturers use alternative symbols for C1, for example C1, $\Sigma(1/A)$ or l_e/A_e. Some manufacturers do not give values for C1 at all, however it can be calculated:

$$C_1 = \frac{0.4\pi\mu_i}{A_L}\,mm^{-1} \tag{8}$$

where μ_i = initial permeability of the ferrite, and A_L = inductance factor in $nH/turn^2$.

Predicting the Guanella Inverter's LF and Mid-Band Loss

To calculate A_i (in dB), the insertion loss of the Guanella inverter, use:

$$A_i = 10\log\left[\left(1+\frac{Z_0}{2R_P}\right)^2 + \left(\frac{Z_0}{2X_P}\right)^2\right] \tag{9}$$

where Z_0 = line characteristic impedance.

Figure 20 shows the final prediction for the loss of Guanella inverters with various numbers of turns on the RCC14/5-4C65 core. Figure 21 shows measurements of the real thing. The measurements were made on four different RCC14/5-4C65 cores, and were compensated for the loss of the coaxial cable used in the windings. The graphs become rather noisy below 0.1dB due to measuring equipment limitations.

The measured values fit the predictions quite well. Residual inaccuracies are due to the assumptions made at the start of the calculations, and because the published A_L has a ±25% tolerance. This tolerance is due to variability of both the material performance and of the core's physical dimensions. The tolerance is even greater at HF because μ_p' and μ_p'' vary with differing core size and shape.

Earlier, I stated that I had extrapolated the manufacturer's complex permeability data below 8 MHz. At low frequency, the transformer loss is dominated by X_P, the reactance of L_P. This is calculated from μ_p' which is constant at low frequencies. It can be extrapolated without fear of error. The value of R_P, which is calculated from μ_p'' has only a very small effect on the transformer loss. I have extrapolated the spectrum of μ_p'' below 8 MHz using a curve with a shape that is typical of ferrites.

LF and Mid-Band Loss of Other Guanella Transformers

The insertion loss of the complex Guanella transformer (i.e. composed of sub-transformers, and not just an inverter) can be calculated by finding the values of R_P and L_P across each sub-transformer. The differential voltage is then calculated across each of

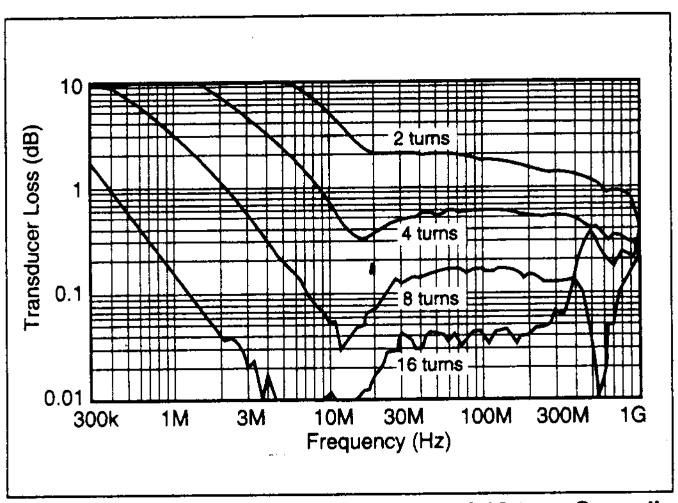

Figure 21. Measured loss for 2, 4, 8, and 16 turn Guanella inverters on RCC 14/5-4C65 toroid.

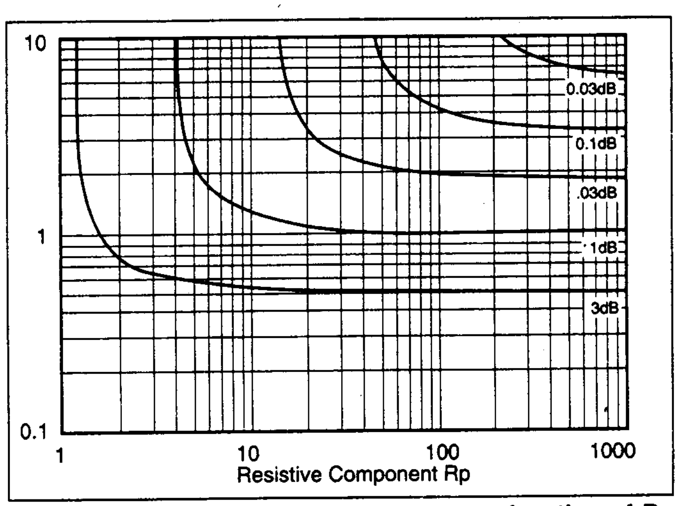

Figure 22. Transformer insertion loss as a function of Rp and Xp where Z_0 = 1 Ω

these components, and so the insertion loss can be found. This is another of the tasks that is best done with a computer simulation. Where L_P and R_P are not constant, S parameters must be used for each transformer component, and simulated using one of the newer sons of SPICE 2G.6.

It is possible to calculate the Guanella transformer's loss using matrices [12], but it is still best to use a computer for this.

The Insertion Loss Formula

It is hard to visualize the implications of equation 6, so Figure 22 shows a graph of the loss for various values of R_P and X_P, where $Z_0 = 1\Omega$. This shows that the transformer's insertion loss is much more dependent on the value of R_P than on the reactance due to the winding inductance or the capacitance. As an example, for an insertion loss of 0.3 dB, the reactance should be about four-times the circuit impedance, whereas the resistance should be about 20-times the circuit impedance. An intuitive explanation for this is that reactance is non-dissipative, and serves only to increase the loss by increasing the VSWR. Resistance both increases the VSWR and dissipates power.

Auto-Transformer Loss

Figure 23 shows the graph of the insertion loss of a 50 to 75 Ω transformer wound on a Siemens B62152-A8-X30 double aperture core with a turns ratio of 9:11. The core is made of N30 grade material. Unfortunately, the manufacturer does not give complex permeability data for N30 grade for frequencies above 4 MHz. So, using R_{norm} = 60 Ωmm/turn^2, N=11, and the manufacturer's figure of C_1 = 1.78 mm^{-1}, equation 3 gives the result R_P = 4 kΩ. Because the Q of the winding will be very low, the term X_P due to the parallel inductance can be ignored, and equation 9 can be simplified to:

$$A_i = 20\log\left(1+\frac{Z_N}{2R_P}\right)\text{dB} \qquad (10)$$

where Z_N = Impedance connected across the N turns (75 Ω in this case).

This gives an estimated loss value of 0.08 dB, which is close to the value seen in Figure 23. The insertion loss rises at high frequencies towards a maximum at 1.4 GHz. This is difficult to model using SPICE, as a full model of this transformer uses 55 transmission lines. The core is small: it has two holes of 0.9mm diameter separated by 1.5mm. It is a good test of hand-eye coordination threading 11 turns through it.

HF Performance: Core Material

From my experiments, transformers with Manganese-Zinc ferrite cores give higher values of overall winding capacitance C (as defined in Figure 17) than transformers with cores of other materials. This is probably because Manganese-Zinc ferrite has such a high dielectric constant, and such low resistivity, that the core acts as a conductor. As a result, HF performance can be dominated by C. For this reason, these cores should have a plastic coating. This interposes a layer with a relatively small ε_r between the windings and the core, and thus reduces the winding capacitance. Some early transformer designs used un-coated cores, and called for an extra insulating layer to be wrapped round the core.

If C has been determined, the calculated capacitive reactance may be substituted for values of X_P in equation 9.....

For Nickel-Zinc or dust-iron cored transformers, it is less likely that the HF performance of a transformer will be determined by C. There are other sources of loss which include radiation loss and the loss due to the impedance of the spurious transmission line between the line and ground.

Finding C: The Impedance Spectra of Inductors

Figure 24 shows a graph of the impedance of a coil consisting of a ferrite toroid with various numbers of bifilar turns. The values of impedance were calculated from the values of series insertion loss measured in a 50 Ω system.

At all frequencies, doubling the number of turns gives an impedance that is increased by a factor of four. This is as expected of an inductor: if N turns give an inductance L, then 2N turns give an inductance of 4L. This is composed of the self inductance of 2 lots of N turns, which is 2L, plus their mutual inductance which is also 2L. This assumes that the coupling coefficient is 1. From this it may be seen that, at low frequencies where capacitances can be ignored, a core with 16 turns will have an impedance equal to 16^2 = 256 single turn cores in series (e.g. 256 beads threaded on a wire). For this reason, it is better to use multiple turns round a single core, providing the end-to-end capacitance does not degrade the high frequency performance too much.

Figure 25 shows the impedance achieved for various values of bifilar choke inductance and capacitance. It can be concluded from this that the 24 turn choke in Figure 24 had a capacitance of about 1 pF. This caused the sharp roll-off of impedance from 20 to 200 MHz. Above 200 MHz, resonance of the wire lengths caused the erratic behavior up to 1 GHz.

Core Shape: Guanella Transformers

To take full advantage of the Guanella transformer's good high frequency performance, the inter-turn capacitance between the transmission-line windings must be minimized, so the use of beads/sleeves /tubes for a '1 turn winding' is to be preferred over toroids with multi-turn windings.

The beaded line is useful where the minimum frequency is above 30 MHz. Choose a thin transmission line, and beads that are a tight fit over it. This will maximize each bead's impedance. Many short beads are better than a few long ones with the same overall length, because the gaps between the beads reduce the end to end capacitance of the transformer.

The toroid is useful where good LF performance is required. It should have the smallest possible hole through the middle, consistent with the turns being reasonably separated. This will give the highest value of C_1, and thus the best low frequency performance.

The twin-hole core usually gives an inter-turn transmission-line with too low an impedance for use on Guanella transformers.

Core Shape: Ruthroff Transformers

Toroids are usually the best shape for N-filar Ruthroff transformers. As shown in Part 1, the best high frequency performance is to be had by using the shortest possible total wire length; so use the core that gives the shortest winding length for 1 turn (l_w, in mm), and use as few turns as possible. The best low frequency performance is achieved using the greatest

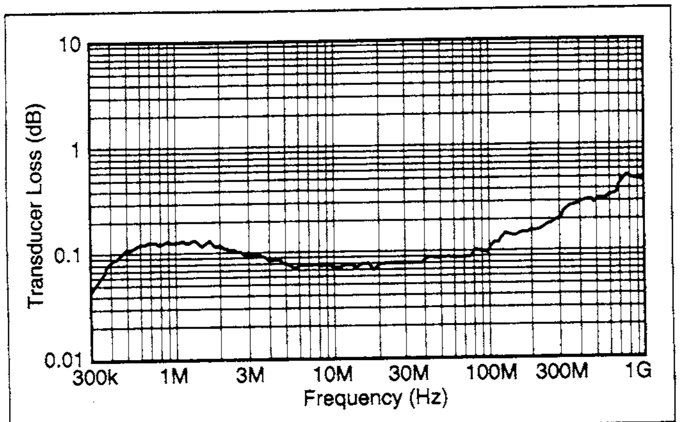

Figure 23. Loss of 50:75 Ω transformer on A8X30 core.

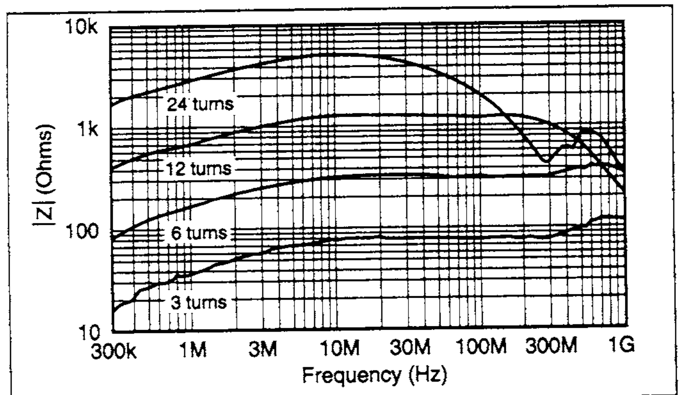

Figure 24. Impedance vs. frequency for choke wound on K37X38 toroid.

number of turns and a core with the lowest core factor. There is thus a trade-off between the winding length, the core factor, and the potential bandwidth of the transformer. This is related to the form factor (F), which is calculated from:

$$F = I_w C_1 \tag{11}$$

The core with the smallest value of F should be selected.

Core Shape: Auto-Transformers

For auto-transformers, the same considerations as for the Ruthroff transformers apply, with the exception that the high frequency performance is sacrificed for winding convenience. For low power transformers, use twin-hole cores. Large twin-hole cores are hard to come by, so for high power designs, use two tube cores side by side. The tube cores can be made from a stack of suitable torroids.

Windings

The wire used should be as thin as possible. For some microwave directional couplers, coaxial cable the thickness of a human hair is used. On Nickel-Zinc or dust-iron cores, try to leave a gap of at least half a wire width between adjacent turns where the winding passes through the center of the toroid. The windings should extend over 330°, leaving a gap of 30° between the winding ends. These measures all reduce the capacitance. There is a limit to this: if the wires are made too small, the required transmission line characteristic may be difficult to achieve, and the line's loss may be increased due to the decreasing skin depth with frequency.

Core Heating

For low power transformers, the only effect of R_P is on the insertion loss, which is an inconvenience which can usually be removed with a little extra amplification. (The exception to this being low-noise designs, which must have low-loss transformers.) However, R_P also sets a limit to the transformer's power handling capacity due to core heating. To handle high powers R_P must be quite large. For example, take a 400 W 200:50 Ω HF balun. The core dissipation is to be 1 W maximum. The required bifilar choke impedance can be calculated to be 50 × 400/1 = 20 kΩ.

In this article R_P has been calculated as a small-signal value derived from the complex permeability data. At high power there will be additional hysteresis loss, which must be calculated from the manufacturer's data. This reduces the actual value of R_P, however the small-signal value provides a useful practical guide to the expected core heating.

Intermodulation Products

It has been stated that transmission line transformers do not suffer from intermodulation products because the magnetic fluxes of the two windings cancel. This is not true. In general, for each bifilar winding there is a difference between the average voltages at the input and output. This voltage is connected across the bifilar choke, and causes a small out of balance current to flow. This is true even where the transformer is being used solely as a delay element: the voltage difference is provided by the phase shift of the RF.

This magnetization current will exercise the ferrite's hysteresis loop. Unfortunately, the only simple way to

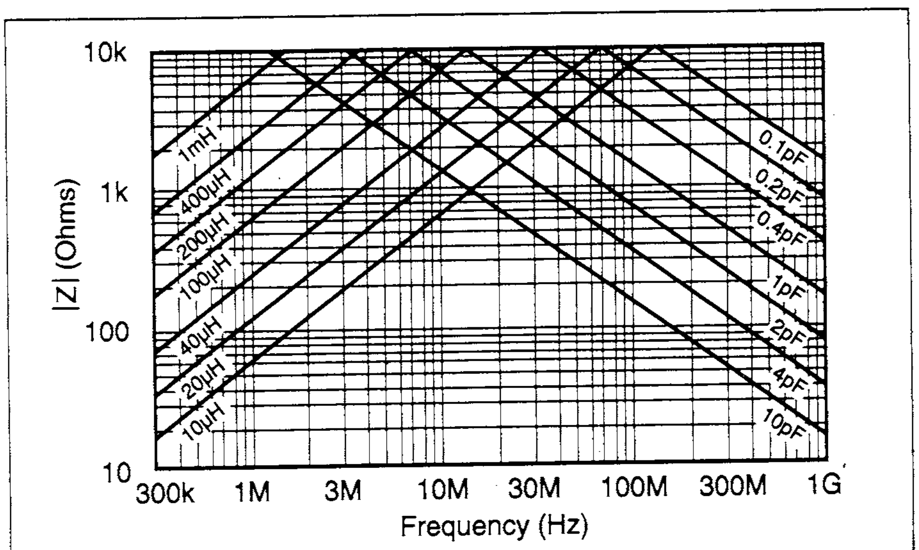

Figure 25. Impedance levels for various values of bifilar choke inductance and capacitance.

predict the resultant intermodulation products is by using CAD calculations. I have not tried this.

The Non-Working Transformer

In Part 1 of this article, I made an unsupported statement regarding Figure 3. I said that the transformer consisting of a few turns on a ferrite ring would not work at RF. The time has come to both qualify and justify this statement.

The transformer in Figure 3 is obviously a pure Faraday type, and is clearly the wrong shape, as defined above. To work efficiently, it must operate on the LF part of the permeability characteristic where the resistive losses are low. However, more turns would be needed to achieve sufficient primary inductance to make the transformer work. The only way that the transformer in Figure 3 could work was if the input and output impedances were small, say less than 1 Ω. I have measured a transformer like the one in Figure 3 in a 50:12.5 Ω circuit. The center of its pass-band was at 5 MHz, where its insertion loss was 2 dB. Its bandwidth was very small.

Conclusion

It has been argued that data for core materials that are given for conventional transformers and inductors are of no use when trying to predict the performance of transmission-line transformers. This is not true. To select a grade of ferrite for a transformer, the manufacturer's complex permeability or impedance spectra should be consulted. These, after mathematical manipulation, can give a good estimate of transformer loss.

At the start of this article, I pointed to the apparent disagreement between Snelling's and Sevick's view of transmission line transformers. I hope that it is now obvious that there is no disagreement because:

a) Following Snelling's advice on 'conventional' transformer design leads to the construction of transmission-line transformers

b) Sevick did not claim the auto-transformer as a transmission-line device.

I hope that this article has encouraged RF designers to build their own transformers. There is much still to be discovered about them. They are also one of the last components that the circuit designer can design for himself [13]. *RF*

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