Designing Wide-band Transformers for HF and VHF Power Amplifiers

The author describes the alternatives available in the design of transformers for solid state RF amplifiers. The key parameters of different construction techniques are discussed with results shown for each.

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Introduction

In the design of RF power amplifiers, wide-band transformers play an important role in the quality of the amplifier as they are fundamental in determining the input and output impedances, gain flatness, linearity, power efficiency and other performance characteristics. The three forms of transformers that are encountered, *unbalanced-to-unbalanced* (unun), *balanced-to-balanced* (balbal), and balanced-to-unbalanced (balun), are used in various combinations to accomplish the desired goals.

Careful consideration needs to be given when making choices of the magnetic materials (if any is to be used), the conductors, and the method of construction, as the choices made weigh significantly in the overall performance of the transformer. The type and length of the conductors and the permeability of the magnetic material are the primary factors that determine the coupling, which in turn determines the transmission loss and the low frequency cutoff. The type and length of conductor used and the loss characteristics of the magnetic material also affects the coupling, and further influences the parasitic reactances that affect the high frequency performance.

Parasitics and Models

Transformers are not ideal components, and their performance is highly dependent upon the materials used and the manner in which they are constructed. The transmission losses and the low frequency cutoff are primarily dependent upon the method of construction, the choices of magnetic material and the number of turns on the windings or length of the conductors. These choices further determine the parasitic reactances that affect the high frequency performance, which include, but are not limited to, resistive losses, leakage inductance, interwinding capacitance and winding self capacitance. A complete equivalent model of a wide-band transformer is shown in Fig. 1.[1] Here, the series resistances R_1 and R_2 represent the losses associated with the conductors in the primary and secondary windings, respectively. These resistances are nonlinear, increasing with

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[1]Notes appear on page 15.

frequency because of the skin effect of the wire itself.[2] Since wide-band transformers using ferromagnetic cores have fairly short lengths of wire, the contribution of the resistive loss to the total loss is small and is generally omitted.[2] The shunt resistance R_C represents the hysteresis and eddy current losses caused by the ferromagnetic material,[3] which increases with ω^2 or even ω^3, and is significant in transformers that are operated near the ferroresonance of the core material.[2] This is a serious consideration in the design of transformers used at HF and VHF frequencies, and therefore requires that proper consideration be given to the selection of the core material.

The low frequency performance is determined by the permeability of the core material and the number of turns on the windings or length of the conductors. The mutual inductance M of Fig 1 is a result of the flux in the transformer core[4] that links the two windings. The high frequency performance is limited by the fact that not all of the flux produced in one winding links to the second winding, a deficiency known as *leakage*,[5] which in turn results in the primary and secondary leakage inductances L_{l1} and L_{l2} of Fig 1. Since the leakage flux paths are primarily in air, these leakage inductances are practically constant.[6,7]

The capacitances associated with wideband transformers are generally understood to be distributed, but it is inconvenient to model transformers by way of distributed capacitances *per se*, so lumped capacitances are used. In Fig 1, capacitor C_{11} represents the distributed primary capacitance resulting from the shunt capacitance of the primary winding. Likewise, C_{22} represents the shunt capacitance of the secondary winding. Capacitor C_{12} is referred to as the interwinding capacitance,[8] and is also a distributed capacitance. In transformers having a significant amount of wire, the inter-winding capacitance can interact with the transformer inductances and create a transmission zero. In good quality audio transformers, the inter-winding capacitance is minimized by placing a grounded copper sheet between the windings, often referred to as a Faraday Shield.

The complete model of the wide-band transformer shown in Fig 1 is well suited for rigorous designs in which the transformer is used near the limits of its performance. These and other models are well suited for use in detailed computer simulations.[9] In general practice and in analytical solutions, it is more convenient to consider the lossless wide-band transformer model shown in Fig 2. This model has been reduced to the reactive components and an ideal transformer.[10] The three model capacitances of Fig 2 are related to the model capacitances of Fig 1 in the following manner:[7]

$$C'_{11} = C_{11} + C_{12}\left(1 - \frac{1}{n}\right) \qquad \text{(Eq 1)}$$

$$C'_{12} = \frac{C_{12}}{n} \qquad \text{(Eq 2)}$$

$$C'_{22} = \frac{C_{22}}{n^2} + C_{12}\left(\frac{1}{n} - 1\right) \qquad \text{(Eq 3)}$$

Using the lossless model of Fig 2, we can devise two models that are proper subsets that can be used to measure the various reactive components. The model of Fig 3 is used to visualize the measurement of the primary shunt capacitance C'_{11} and the primary referred equivalent series inductance L_{EQ1}.[10] Likewise, the model of Fig 4 is used to visualize the measurement of the secondary shunt capacitance C'_{22}, the primary referred equivalent series inductance L_{EQ2} and the interwinding capacitance C'_{12}.[10] The turns ratio n of the ideal transformer is the actual ratio of the physical number of turns between the primary and secondary windings. The procedure for determining the values of the parasitic reactances of Fig 2 is as follows[10]:

1. With the secondary open, measure the primary winding inductance L_P at a frequency well below the high frequency cutoff of the transformer.
2. With the primary open, measure the secondary winding inductance L_S, also at a frequency well below the high frequency cutoff of the transformer.
3. With the secondary open, apply a signal at an appropriate mid-band frequency (between the low and high frequency cutoff frequencies) to the primary winding and measure the input and output voltages v_1 and v_2.
3. Calculate the coupling coefficient k using:

$$k = \frac{v_2}{v_1}\sqrt{\frac{L_P}{L_S}} \le 1 \qquad \text{(Eq 4)}$$

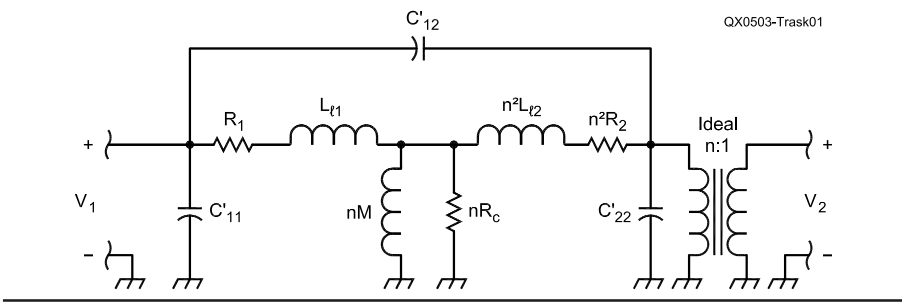

Fig 1—Complete wideband transformer model.

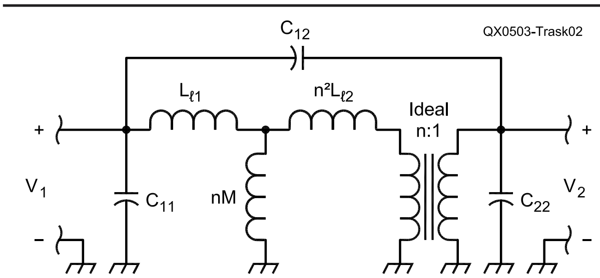

Fig 2—Lossless wideband transformer model.

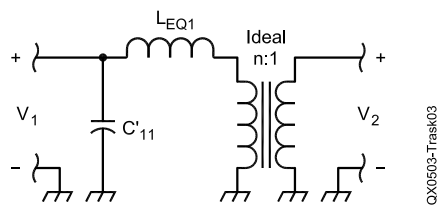

Fig 3—Equivalent circuit for determining C'_{11}.

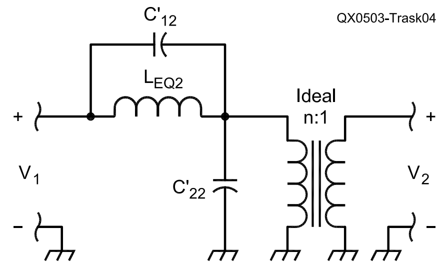

Fig 4—Equivalent circuit for determining C'_{12} and C'_{22}.

4. Calculate the mutual inductance M using:

$$M = \sqrt{L_P \; L_S \; k} \qquad \text{(Eq 5)}$$

5. Calculate the primary leakage inductance L_{l1} using:

$$L_{l1} = L_P - n\,M \qquad \text{(Eq 6)}$$

6. Calculate the secondary leakage inductance L_{l2} using:

$$L_{l2} = L_S - \frac{M}{n} \qquad \text{(Eq 7)}$$

7. Calculate the primary referred equivalent inductance L_{EQ1} using:

$$L_{EQ1} = \frac{n^2\,M\,L_{l2}}{M + n\,L_{l2}} + L_{l1} \qquad \text{(Eq 8)}$$

8. Calculate the secondary referred equivalent inductance L_{EQ2} using:

$$L_{EQ2} = \frac{n\,M\,L_{l1}}{n\,M + L_{l1}} + n^2\;L_{l2} \qquad \text{(Eq 9)}$$

9. Referring to Fig 4, connect a generator to the primary and an appropriate load across the secondary. Measure the transmission parallel resonant frequency f_{12}, which is the frequency at which the voltage across the secondary is at a minimum.

10. Calculate C'_{12} using:

$$C'_{12} = \frac{1}{L_{EQ2}\,(2\,\pi\,f_{12})^2} \qquad \text{(Eq 10)}$$

11. With the generator still connected to the primary and the secondary open, measure the input series resonant frequency f_{22}, which is the frequency at which the voltage across the primary is at a minimum.

12. Calculate C'_{22} using:

$$C'_{22} = \frac{1 - C'_{12}\,L_{EQ2}\,(2\,\pi\,f_{22})^2}{L_{EQ2}\,(2\,\pi\,f_{22})^2} \qquad \text{(Eq 11)}$$

13. Referring to Fig 3, connect a generator to the secondary and leave the primary open. Measure the output series resonant frequency f_{22}, which is the frequency at which the voltage across the secondary is at a minumum.

14. Calculate C'_{11} using:

$$C'_{11} = \frac{1 - C'_{12}\,L_{EQ1}\,(2\,\pi\,f_{11})^2}{L_{EQ1}\,(2\,\pi\,f_{11})^2} \qquad \text{(Eq 12)}$$

15. Calculate C_{12} using:

$$C_{12} = n\,C'_{12} \qquad \text{(Eq 13)}$$

16. Calculate C_{11} using:

$$C'_{11} = C'_{11} - C_{12}\left(1 - \frac{1}{n}\right) \qquad \text{(Eq 14)}$$

17. Calculate C_{22} using:

$$C_{22} = n^2\left[C'_{22} - C_{12}\left(\frac{1}{n} - 1\right)\right] \qquad \text{(Eq 15)}$$

Matching

Once the values of the parasitic reactive elements of the wide-band transformer model have been determined, it is possible to design the transformer into a matching network that not only absorbs them, but makes use of them in forming a 3-pole π network low-pass filter section.[10,11,12]

We begin by considering the fact that in a properly designed transformer the equivalent series inductance will dominate and will determine the maximum frequency for which a matching network can be realized. The input and output capacitances C_{11} and C_{22} are usually much smaller than required for realizing a π network low-pass filter section, so additional padding capacitors will be required to properly design the matching network. These three components allow us to design 3-pole Butterworth, Bessel, Gaussian, and Tchebyschev filter sections with the equivalent series inductance dictating the cutoff frequency.

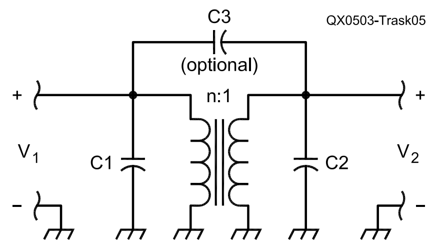

Fig 5—Matching network components.

The presence of the interwinding capacitance C_{12} suggests that a single parallel-resonant transmission zero can be included, which gives us further possibilities of inverse Tchebyschev and Elliptical (Cauer) filter sections. Since the interwinding capacitance is generally small, it will also require an additional padding capacitor to complete the design. Note, however, that adding a transmission zero to the matching net-

Table 1—Matching Section Prototype Values[13]

Filter Type	*C_{1norm}*	*C_{2norm}*	*L_{norm}*	*C_{3norm}*
Butterworth	1.000	1.000	2.000	
Bessel	1.255	0.192	0.553	
Gaussian	2.196	0.967	0.336	
Tchebyschev				
0.1 dB	1.032	1.032	1.147	
0.5 dB	1.596	1.596	1.097	
1.0 dB	2.024	2.024	0.994	
Inverse Tchebyschev				
20 dB	1.172	1.172	2.343	0.320
30 dB	1.866	1.866	3.733	0.201
40 dB	2.838	2.838	5.677	0.132
Elliptical (0.1 dB Passband Ripple)				
20 dB	0.850	0.850	0.871	0.290
25 dB	0.902	0.902	0.951	0.188
30 dB	0.941	0.941	1.012	0.125
35 dB	0.958	0.958	1.057	0.837
40 dB	0.988	0.988	1.081	0.057
Elliptical (0.5 dB Passband Ripple)				
20 dB	1.267	1.267	0.748	0.536
25 dB	1.361	1.361	0.853	0.344
30 dB	1.425	1.425	0.924	0.226
35 dB	1.479	1.479	0.976	0.152
40 dB	1.514	1.514	1.015	0.102
Elliptical (1.0 dB Passband Ripple)				
20 dB	1.570	1.570	0.613	0.805
25 dB	1.688	1.688	0.729	0.497
30 dB	1.783	1.783	0.812	0.322
35 dB	1.852	1.852	0.865	0.214
40 dB	1.910	1.910	0.905	0.154

work is not practical for transformers that go from balanced to unbalanced sources and loads (baluns) as the equivalent series inductances from the unbalanced port to the balanced ports are not identical.[10]

To begin the process of designing the wide-band transformer into a matching network, we must first decide what sort of passband performance is desired, and then select the appropriate filter prototype values from Table 1. Now, with reference to the component reference designators of Fig 5, the design process proceeds as follows:[10]

1. Calculate the maximum usable frequency ω_{max} using:

$$\omega_{max} = \frac{L_{norm}\,R_S}{L_{EQ1}} \qquad \text{(Eq 16)}$$

where R_S is the source resistance

2. Calculate the value for the input matching capacitor C_1 using:

$$C_1 = \frac{C_{1norm}}{\omega_{max}\,R_S} - C_{11} \qquad \text{(Eq 17)}$$

3. Calculate the value for the output matching capacitor C_2 using:

$$C_2 = \frac{n^2\,C_{2norm}}{\omega_{max}\,R_S} - C_{22} \qquad \text{(Eq 18)}$$

4. If required, calculate the value for the capacitor C_3 using:

$$C_3 = \frac{n\,C_{3norm}}{\omega_{max}\,R_S} - C_{12} \qquad \text{(Eq 19)}$$

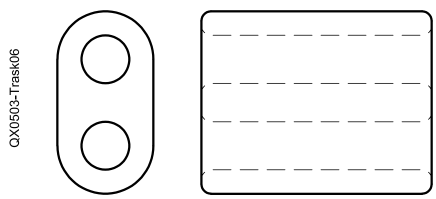

Fig 6—Binocular core.

Magnetic Materials

The first concern in the design of a wide-band transformer is the choice of the magnetic material. Both ferrite and powdered iron materials can be used, but ferrite is preferred over powdered iron as the losses are lower. Powdered iron is lossier because of the distributed air-gap nature of the material,[13] and the excessive losses not only result in decreased gain performance, but in power amplifier applications they also result in excessive heating that can damage insulating and PC board materials.

There are three essential types of ferrite materials that can be used for HF and VHF frequencies. These are listed in Table 2. The first of these is manganese-zinc (MnZn), which is generally suited for lower frequencies and low power. Fair-Rite type 77 material is an exception. It is available in the form of E-I cores which can be used for high-power transformer cores at lower HF frequencies.

The second, which will be discussed later, and undoubtedly most popular type of ferrite is nickel-zinc (NiZn). Of the three NiZn materials listed in Table 2, the Fair-Rite types 43 and 61 are by far the most widely because of their low loss, high saturation flux, and the wide variety of shapes and sizes that are available. They can be readily used for both HF and VHF applications, with the 61 material being preferred for VHF. These two ferrites will be the focus of the applications to be discussed later.

The third type of ferrite suitable for HF and VHF applications is cobalt-nickel-zinc (CoNiZn), available from Ferronics as types K and P. These ferrites are available in a limited number of shapes and sizes. Toroids made from these materials can be used to make transformer cores by stringing them along a brass tube in a frame, as will be discussed later. The one drawback to this material is that it can be permanently damaged if it is subjected to excessively high flux densities.[10]

Transformer Cores

The ferrite materials mentioned in the previous section are available in a fairly wide variety of shapes such as rods, toroids, beads (or sleeves), E-I

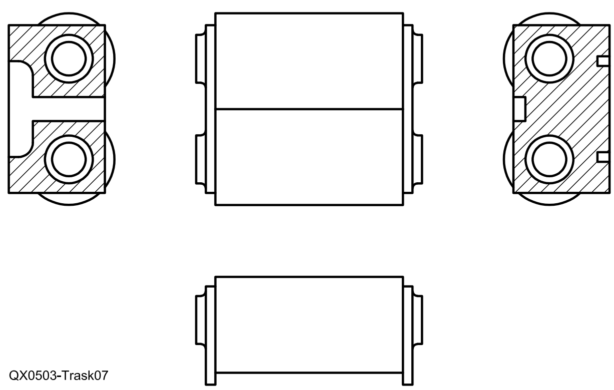

Fig 7—High-power rf transformer core.

Table 2—Commercial Ferrites Suitable for Power Amplifier Applications

Manufacturer	Type	Permeability (u_i)	Saturation Flux (Gauss)	Loss Factor	Usable Frequency	Resonance Frequency
Manganese-Zinc (MnZn) Ferrites						
Fair-Rite	77	2000	4900	15	1 MHz	2MHz
Nickle-Zinc (NiZn) Ferrites						
Fair-Rite	43	850	2750	85	5 MHz	10MHz
Steward	28	850	3250			
Fair-Rite	61	125	2350	32	25 MHz	50MHz
Steward	25	125	3600		15 MHz	25MHz
Cobalt-Nickle-Zinc (CoNiZn) Ferrites						
Ferronics	*K*	*125*	*3200*		*50 MHz*	*60 MHz*
Ferronics	*P*	*40*	*2150*		*80 MHz*	*100 MHz*

cores, and multi-aperture cores. Among the various multi-aperture cores available, there is one form, shown in Fig 6, that is commonly referred to as a "binocular core" as the shape suggests that of a pair of field glasses. This shape is available from numerous small sizes suitable for small-signal transformers to larger sizes suitable for power amplifiers up to 5 W. Similar cores available from Fair-Rite having a rectangular rather than an oval cross-section are available in larger sizes suitable for amplifiers of 25 W or more.

Higher-power amplifiers require cores with larger cross-sections that can accommodate the higher flux densities in the magnetic material. For these applications, it is more suitable to construct a transformer core using ferrite beads (or sleeves) supported by a frame made from brass tubing and PC board material, such as those that are available from Communications Concepts and made popular by the numerous applications notes and other publications by Norman Dye and Helge Granberg.[14,15,16,17] An illustration of a Communications Concepts RF600 core assembly is shown in Fig 7. Notice that the left-hand endplate has two separate conductors while the right-hand end plate has a single conductor. This is helpful in forming a center-tap ground connection in some applications. In an application to be described later in the design of balun transformers, it will be seen that there are times when it is advantageous to dispense with this common connection.

The transformer core of Fig 7 can be made for higher power levels by using multiple ferrite beads along the supporting tubes, such as the RF-2043 assembly offered by Communications Concepts. Such an assembly technique can also allow for the use of toroids to provide a transformer core having a larger cross section or to provide a means of using ferrite materials in the form of toroids when beads are not available as was mentioned earlier.

Conventional Wide-band Transformers

The most common method used in the design of power amplifiers for HF and VHF frequencies is shown in Fig 8. Here, a 1:1 balun is made using the transformer core previously shown in Fig 7. The balanced side of the trans-former is provided by the brass tubes that support the ferrite sleeves with the center tap being provided by the common connection foil of the right-hand endplate and the + and – terminals provided by the foil of the left-hand endplate. The unbalanced side of the transformer is provided by way of a piece of insulated wire that is passed through the tubes.

There are at least two problems with transformers constructed in this manner, the first of which is the wire for the unbalanced side of the circuit that is exposed in the left-hand end of the assembly. The field created by this exposed wire is not coupled to either the brass tubes of the balanced side of the circuit nor the ferrite material, and this results in excess leakage inductance. The second problem is that the coupling between the two sides of the circuit is not uniform as the physical placement of the wire cannot be tightly controlled. This can lead to some small amount of imbalance. Despite these problems, this form of transformer remains very popular in the design of amateur, commercial, and military HF and VHF power amplifiers.

For demonstration purposes, a 1:1 balun transformer was constructed, using a Communications Concepts RF-600 transformer core assembly, which uses a pair of Fair-Rite 2643023402 beads, made with type 43 material and having an inside diameter of 0.193 inch, an outside diameter of 0.275 inch, and a length of 0.750 inch.

The performance for this balun, shown in Fig 9, is marginal at best. The average insertion loss for HF frequencies is in the neighborhood of 2 dB, and the cutoff frequency is around 4 MHz. At higher frequencies, the insertion loss improves to 1.2 dB, but even this is of questionable value. The slowly degrading return loss is more a result of the increased losses caused by the ferrite material, as was evidenced by the fact that adding matching capacitors (see Fig 5) did little to improve the performance. The increased transmission loss above 85 MHz is due mostly to the leakage inductance caused by the exposed conductor on the left-hand end of the assembly.

Transmission-Line Transformers

The leakage inductance of the balun transformer of Fig 8, however small, is the limiting factor for higher frequency performance. To fulfill the need for wide-band transformers at higher frequencies and power, coaxial cable is often employed as the conductors. Since the coupling takes place between the inner conductor and the outer shield, there is very little opportunity for any stray inductance. This means that we can anticipate good performance at

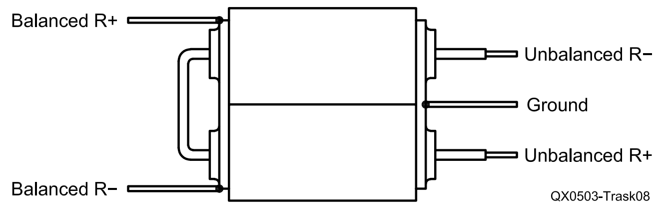

Fig 8—Conventional 1:1 balun transformer.

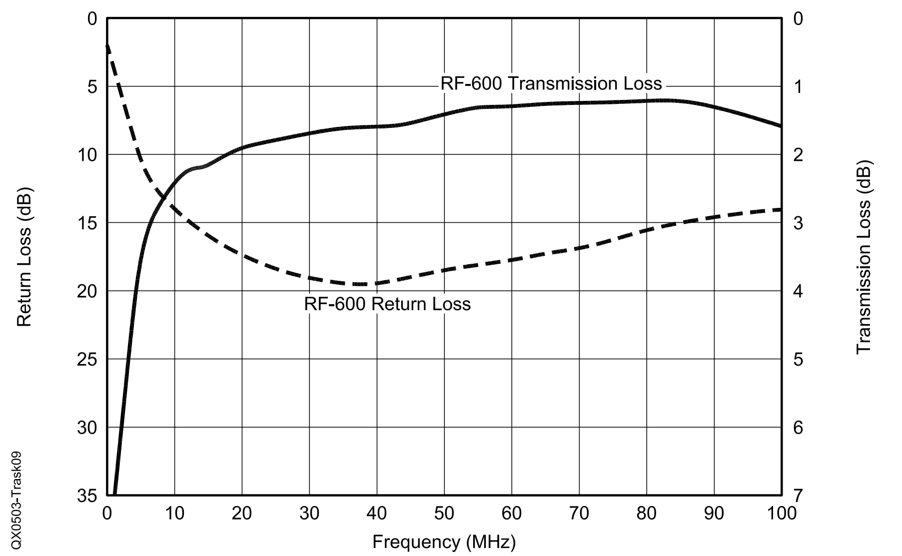

Fig 9—Conventional 1:1 balun transformer performance.

much higher frequencies, and it also means that we can usually dispense with the matching capacitors that are often used with wide-band transformers.

In the design of transmission line transformers, the cable should have a characteristic impedance that is the geometric mean of the source and load impedances:

$$Z_0 = \sqrt{Z_S \times Z_L} \quad \text{(Eq 20)}$$

In most cases, the use of coaxial cable having the exact impedance is simply not possible as coaxial cable is generally offered in a limited number of impedances, such as 50 and 75 Ω. Other impedances such as 12.5, 16.7, 25, and 100 Ω are available, but usually on a limited basis for use in military and commercial applications. Low impedances such as 6.12 Ω are difficult to achieve, although it is possible to parallel two 12.5 Ω cables, which is standard practice.[16] The insertion loss will increase as the impedance of the coaxial cable deviates from the optimum impedance of Eq. 20. For most applications, the effects of using cable having a non-ideal characteristic impedance is not great as long as the equivalent electrical length of the cable is less than λ/8. In general, the line impedance is not critical provided that some degree of performance degradation is acceptable.[16]

The equivalent electrical length of the cable is actually longer than the physical length due to the electrical properties of the insulating material between the inner and outer conductors, and the relationship is:

$$L_E \cong L_P \sqrt{\varepsilon_r} \quad \text{(Eq 21)}$$

where L_E is the equivalent electrical length, L_P is the actual physical length, and ε_r is the relative dielectric constant of the insulating material, typically 2.43 for PTFE. When the cable is inserted in a magnetic material, the equivalent electrical length is further lengthened by the magnetic properties of the material:

$$L_E \cong L_P \sqrt{\varepsilon_r \mu_i} \quad \text{(Eq 22)}$$

where μ_i is the relative permeability of the magnetic material. In general, a close approximation to the equivalent electrical length of the cable will be a combination of Eq. 21 and Eq. 22, with the former applied to the length of cable that is outside the transformer core and the latter used for that portion of the cable that is inside the transformer core.

The 1:1 balun transformer of Fig 8 is now modified by replacing the insulated wire conductor with an appropriate length of 0.141 inch OD 50 Ω semi-flex coax cable, with a solder-filled braid outer conductor, as shown in Fig 10. Here, the cable is bent into a U shape and passed through the holes of the transformer core. The center tap for the balanced side of the transformer is provided by soldering a wire to the outer conductor at the very center of the curve. Because of the displacement of the center tap from the endplate, the common connection provided by the copper foil on the right-hand endplate (see Fig 7) must be broken. Notice that this design places the terminals for both the balanced and unbalanced sides of the transformer on the same end of the core.

Semi-rigid coax is also available with the same 0.141 inch OD, but it is difficult to use when small radii are required. The solid outer conductor often splits or collapses if the bending radius is too small. Semi-flex will bend to smaller radii, but will still split when an excessively small radius is attempted. A mandrel, such as you would use when bending copper

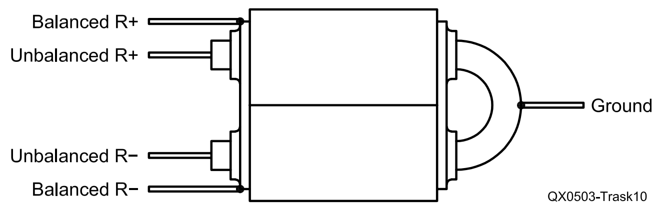

Fig 10—Transmission-line 1:1 balun transformer.

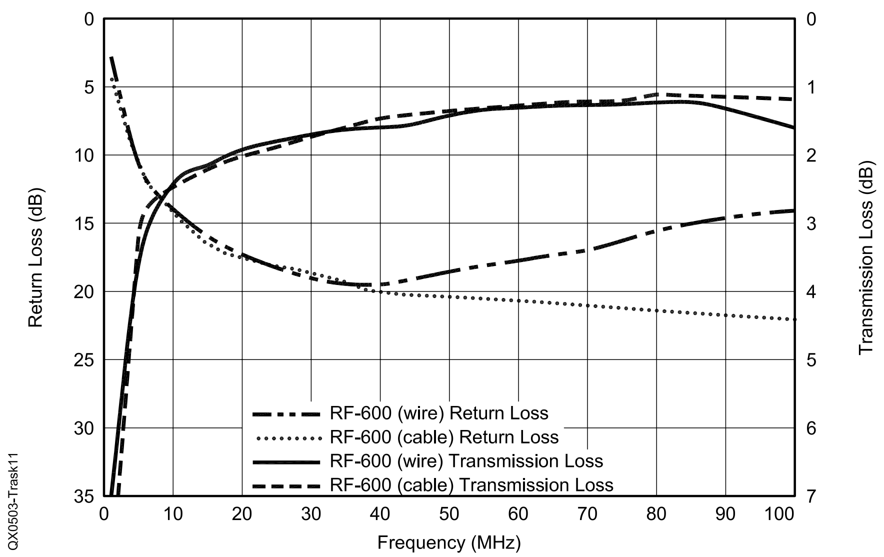

Fig 11—Conventional vs. Transmission-line 1:1 balun transformer performance.

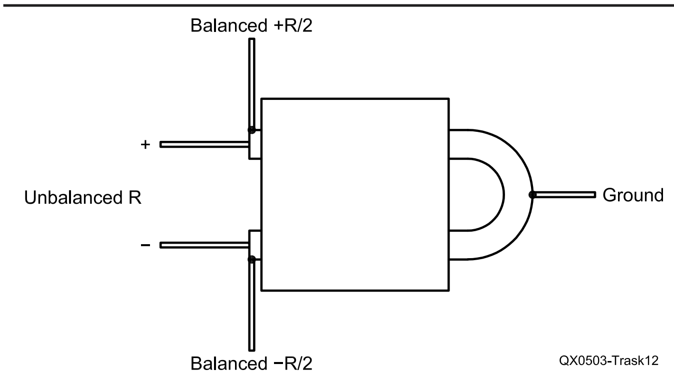

Fig 12—Coaxial transmission-line 1:1 balun.

tubing, should be used at all times when bending these cables to the small radii required. A great deal of care must be exercised, which is best done by first bending the cable to a larger radius and then slowly decreasing the radius until it is sufficiently reduced so as to pass through the two holes of the core with little effort. This method reduces the risk of splitting the outer conductor by way of distributing the mechanical stresses over a longer length of the cable. For transformer cores having larger hole diameters, larger coaxial cables such as RG-58 and RG-59 can be used, provided the outer vinyl jacket is removed.

Fig 11 shows that the use of coaxial cable has done little to improve the low frequency characteristics of the 1:1 balun transformer, however the high frequency characteristics show significant improvement, especially with regard to the return loss. With the better coupling between the two circuits, the losses induced by the ferrite material have been reduced and a better match has been attained. Also, the lack of any appreciable increase in the transmission loss above 85 MHz indicates that the leakage inductance has been reduced, as was expected by using the coax cable instead of wire for the conductor.

Transmission-Line Baluns

Replacing the wire with coaxial cable in the 1:1 balun transformer of Fig 7 and Fig 10 helps the high frequency transmission loss and return loss performance to some degree. It does not, however, improve the low frequency performance nor the transmission loss. This is due to the fact that the coupling coefficient of the transmission line transformer is highly dependent upon the length of cable used.

Let's take a broader look at the use of transmission line in the design of a wide-band transformer. In this case, we'll use a pair of 1:1 baluns as shown in Fig 12. We will use a length of 50-Ω semi-flex cable as was used in the previous example, but this time requiring a tighter radius. The core for the first of these transformers is a Fair-Rite 2843000102 binocular core, and for the second a Fair-Rite 2861000102 binocular core is used to demonstrate the differences in the performance of the two ferrite materials. The performance of these baluns is shown in Fig 13. It is immediately obvious that there is room for improvement. First, the transmission loss is 1.8 dB for the transformer using the type 43 material

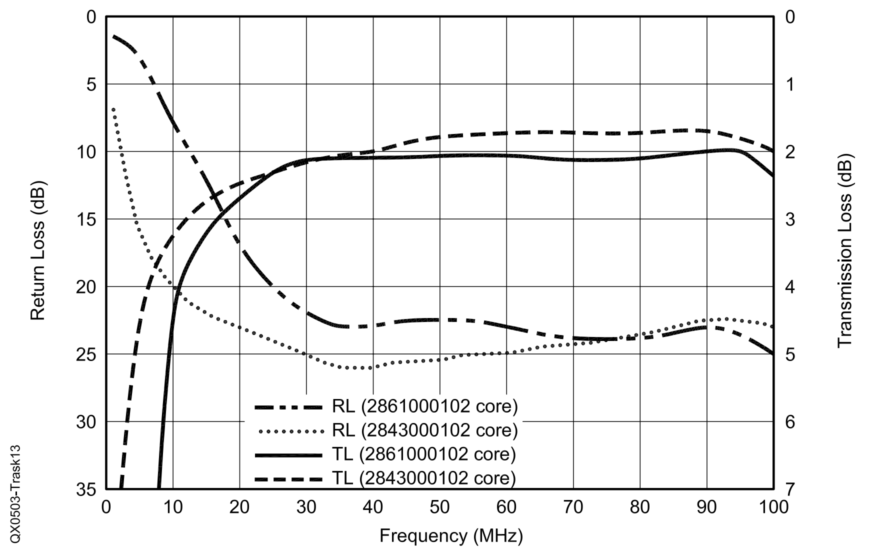

Fig 13—Coaxial transmission-line 1:1 balun performance (-0102 cores).

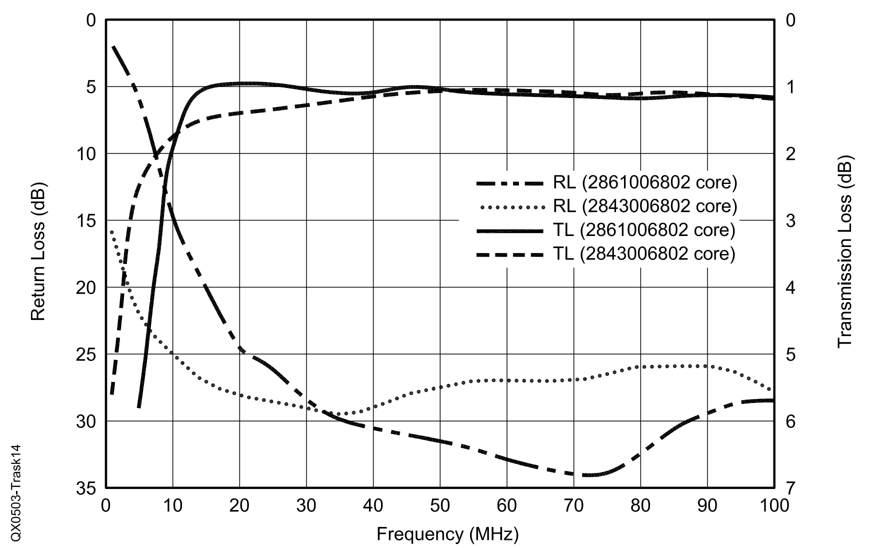

Fig 14—Coaxial transmission-line 1:1 balun performance (-6802 cores).

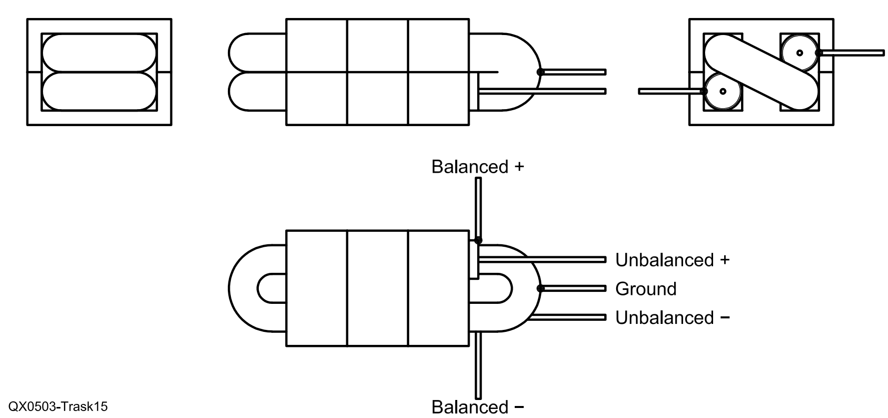

Fig 15—Extended coaxial transmission-line 1:1 balun using e-cores.

and 1.5 dB for the type 61 material. The cutoff frequency is 2.5 MHz for the type 43 material and 11 MHz for the type 61 material.

Another pair of transformers were constructed, this time using Fair-Rite 2843006802 and 2861006802 binocular cores, approximately twice as long as the previous -0102 cores. As shown in Fig 14, this increase in the length of transmission line improves the transmission loss to about 1.1 dB for both materials. As expected by virtue of the longer line length, the cutoff frequencies are significantly lower, less than 1 MHz for the type 43 material and 4 MHz for the type 61 material.

Clearly, the longer length of coaxial cable has distinct advantages in terms of insertion loss and cutoff frequency. It would therefore appear obvious that increasing the length of the cable and ferrite balun core further would result in additional performance improvement. However, in the design of power amplifiers we often encounter a limitation in terms of the amount of physical board space that is available for the various components.

A solution to increasing the length of the cable without sacrificing valuable board space is to form the cable into a series of two or more loops and embed it into an E-core, a two-turn version of which is shown in Fig 15.[16,17] Here, six pieces of ferrite E-core have been cemented together to form a single piece of ferrite. The method of construction is to first cement two sets of three pieces of core material together to form the upper and lower halves of the to be completed core. Next, the cable is formed to the shape necessary to fit within the channels of the core. Finally, the cable is placed inside the channels of one core half and the second half is cemented in place.

The construction itself is fairly straightforward, but implementing it onto a circuit board reveals a couple of problems, specifically the length of the leads for the two balanced ports are unequal and the coax loop on the right-hand side interferes with the unbalanced and balanced positive terminals. At lower frequencies the inequality of the lead lengths will not present sufficient imbalance in lead inductance to create any problems, but with increasing frequency the transformer will become unbalanced and compensation will be required to offset the excessive lead inductance, which will be difficult to bring into balance. An additional constraint in the use of this approach is that the required E-cores are only available in the Fair-Rite type 77 material, which is not well suited

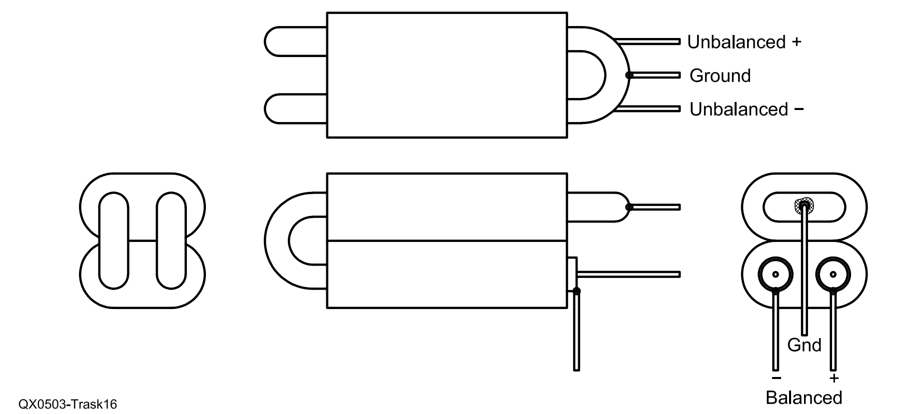

Fig 16—Extended coaxial transmission-line 1:1 baluns using binocular cores.

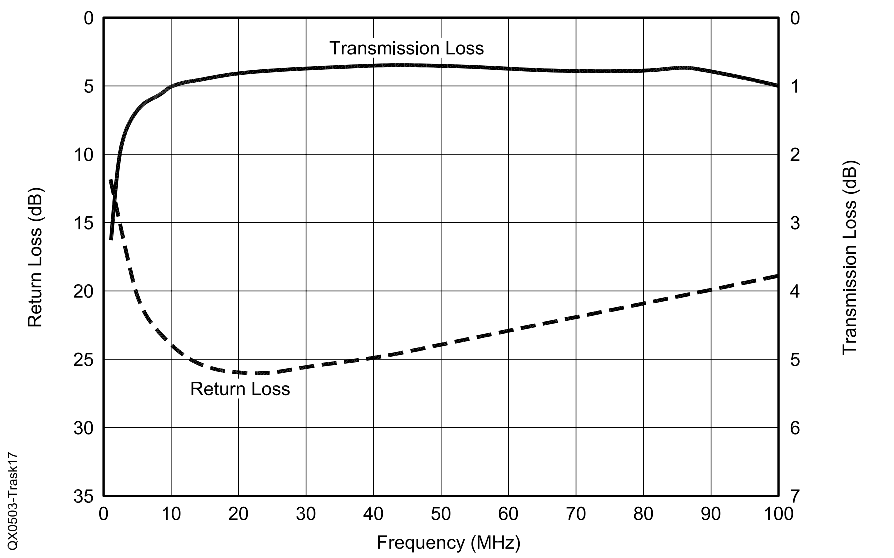

Fig 17—Extended-length coaxial transmission-line 1:1 balun performance.

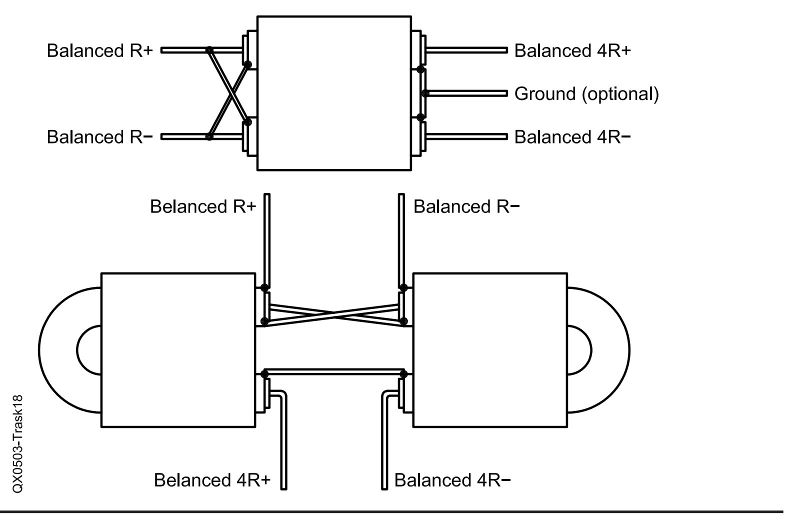

Fig 18—Coaxial transmission-line 4:1 balanced transformers.

above lower HF frequencies. Even with these shortcomings, the wide-band transformer approach of Fig 15 is well worth consideration for applications at HF frequencies.

An alternative approach is shown in Fig 16. A pair of ferrite binocular cores have been used in place of the E-cores of Fig 15. Here, both ends of the cable have equal lead lengths, and there is no mechanical interference to be dealt with. The construction presents no more difficulty than before. The cable is first formed into a U shape, then passed through the holes of the first, or upper binocular core. The free ends of the cable are then bent back over the first core, and subsequently passed through the holes of the second or lower core. The two cores may then be cemented together to make the assembly whole. A pair of endplates similar to those shown for the left-hand end of the transformer-core assembly of Fig 7 may be used to hold the two cores together and to ease mounting the transformer on the amplifier PC board.

A single example of the 1:1 balun transformer of Fig 16 was constructed, using a pair of the longer Fair-Rite 2843006802 binocular cores. The test results shown in Fig 17 indicate that further lengthening of the coaxial cable continues to improve the performance. A comparison of the three balun examples using Fair-Rite type 43 ferrite material is listed in Table 3, where the transmission loss is as an average over what would be considered the usable frequency range.

Even with the lower transmission loss of the balun transformer of Fig 16, this performance of the transmission-line balun transformer remains significant. When used as an output transformer in a power amplifier, this excess loss degrades the power efficiency, which should be taken into consideration in the overall design.

Other Transmission-Line Transformers

There are many possible impedance ratios that can be realized using transmission-line transformers. Fig 18 shows two methods for making balanced transformers having an impedance ratio of 4:1.[15,16,17,18] The first of these makes use of a single binocular core, and it should be

Table 3

Configuration	*Cutoff Frequency*	*Insertion Loss*
Double 2843006802 Core	1.1MHz	0.8dB
Single 2843006802 Core	3.9MHz	1.1dB
Single 2843000102 Core	11MHz	1.8dB

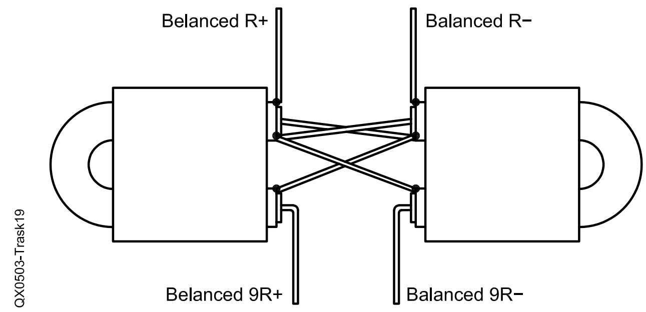

Fig 19—Coaxial transmission line 9:1 balanced transformer.

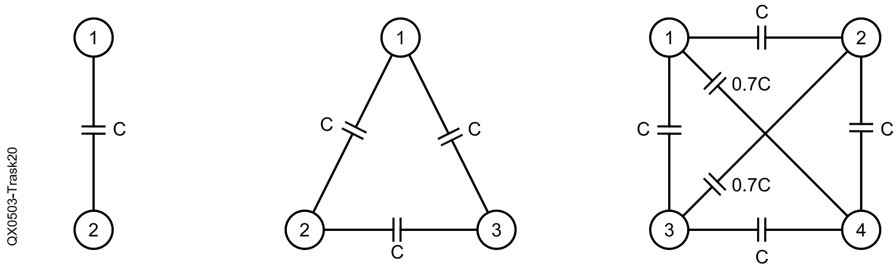

Fig 20—Twisted-wire capacitances.

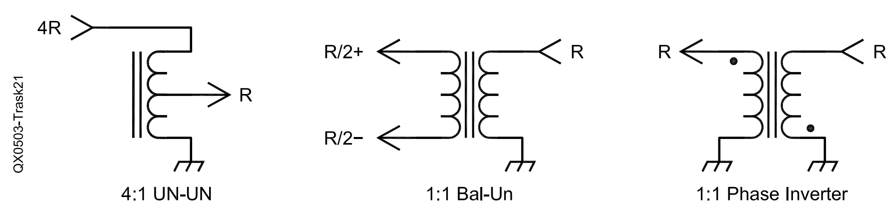

Fig 21—Two-conductor twisted-wire transformer configurations.

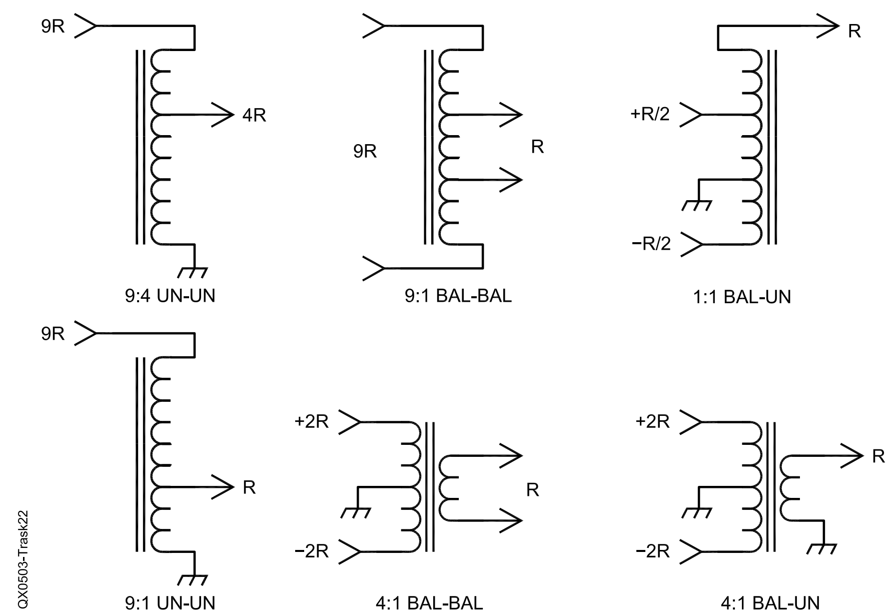

Fig 22—Three-conductor twisted-wire transformer configurations.

obvious from the examples shown for making balun transformers that the core should be as long as possible. The second method makes use of the same bent cable design used for making the earlier baluns of Fig 12,[18] and the extended length design of Fig 16 can also be used here to conserve board space, decrease the cutoff frequency, and decrease the transmission loss.

Fig 19 shows a method by which a balanced transformer having an impedance ratio of 9:1 can be realized, also using the bent cable designs of Fig 12 and Fig 16.[16, 17,18]

Numerous additional transformers can be realized by way of a variety of combinations that will result in integer impedance ratios.[19]

Twisted Wire

The method of using coaxial cable has been widely used in the design of wide-band transformers for high power and high frequencies, but the limitations imposed by low frequency performance and excessive transmission loss are a direct consequence of shortened line lengths. These are often imposed on the designer due to limited physical space. We could make transformers at lower frequencies and with lower transmission losses by extending the approach of Fig 16 by using more cores, but this method would have its own limitations and can become fairly expensive to produce. An alternative to the use of coaxial cable for the conductors is twisted wire. It is by far more flexible than cable, and with proper attention to a few details it can yield transformers having exceptional performance in terms of bandwidth and transmission losses.

It is first necessary to understand that you cannot twist together any number of wires when making a transformer, and Fig 20 illustrates why not. In the first case, a pair of wires twisted together (bifilar) have a fairly uniform distributed interconductor capacitance. Both conductors will have a uniform and equal distributed inductance due to the uniform distributed mutual inductance between the two conductors, a characteristic that allows twisted wire to be seen as a form of transmission line.[20]

In the second case, three wires twisted together (trifilar) exhibit the same properties. The distributed capacitance is the same between all three conductors, as is the distributed inductance. This implies that the characteristic impedance and propagation constant for all three wires are the same.[21]

In the third case, four wires twisted together (quadrifilar) show dissimilar properties. While the interconductor capacitance between adjacent conductors is the same, the capacitance between diagonally opposite conductors is reduced by $1/\sqrt{2}$, and the mutual inductance between these conductors is also reduced. Now, the characteristic impedances and propagation constants for the conductors are not identical, and this results in unequal coupling as well as phasing problems. This is more noticeable as the amount of wire is increased beyond $\lambda/8$. Because of this, it is best to use no more than three twisted wires in the design of wide-band transformers.[21]

Another aspect of twisted wire that needs consideration is the insulation. The polyurethane nylon insulation used on magnet wire offers the best in terms of dielectric properties and losses, but is easily abraded when passed through the holes of a ferrite binocular core. Teflon and polyvinyl chloride (PVC) insulated wire is more durable, but the insulation tends to have higher dielectric losses. In all, magnet wire is the better choice provided you exercise adequate care when assembling the transformer.

Twisted-Wire Transformers

The schematics of Fig 21 illustrate a few of the transformers that can be realized using a pair of twisted wires. The 4:1 autotransformer and the 1:1 unbalanced-to-unbalanced (unun) phase inverter are fairly obvious, whereas the 1:1 balun requires that a ground reference be supplied elsewhere in the circuit. This configuration is convenient when used with an additional transformer such as the 4:1 balbal that will be discussed later in this section.

The addition of a third wire increases the number of practical wide-band transformer configurations, as shown in the schematics of Fig 22. The 9:4 and 9:1 ununs and the 4:1 balun are all fairly obvious. Both the 4:1 and 9:1 balbals require a ground reference elsewhere in the circuit, just as with the bifilar 1:1 balun shown earlier. The 1:1

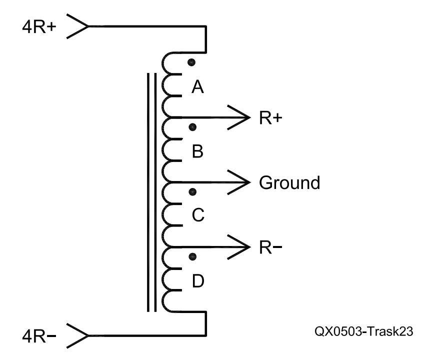

Fig 23—Balanced-balanced transformer, 4:1 transformation ratio.

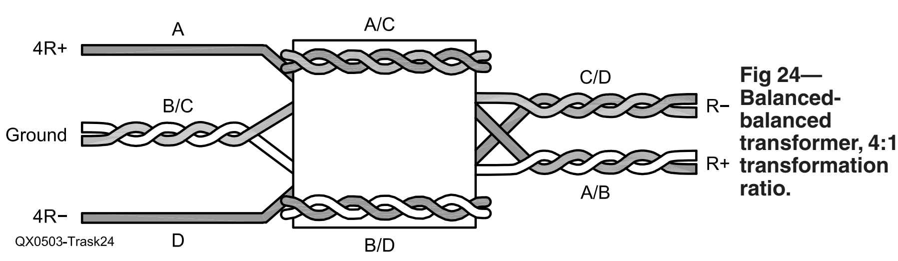

Fig 24—Balanced-balanced transformer, 4:1 transformation ratio.

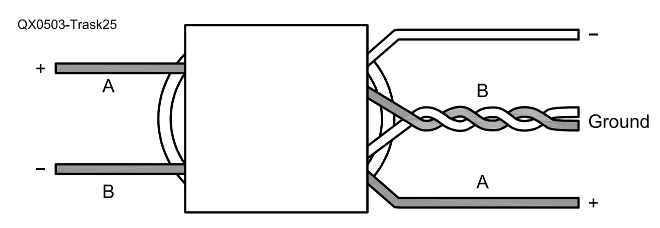

Fig 25—Balun using monofilar primary and bifilar secondary.

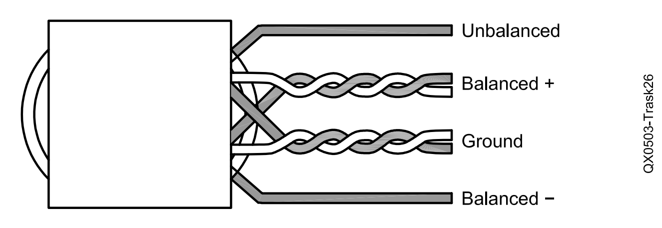

Fig 26—Trifilar wound 1:1 balun.

balun offers the best possible performance for this essential type of transformer as will be shown later, and it provides the center tap that is needed when using the 4:1 and 9:1 trifilar balbals.

An additional wide-band transformer that finds wide usage is the 4:1 balbal shown schematically in Fig 23. It would appear at first that this would be a suitable application for quadrifilar twisted wire, but due to the unequal coupling considerations discussed earlier, an alternative approach is used. Shown in Fig 24, the transformer of Fig 23 is realized by winding a pair of bifilar conductors around the outside of the balun transformer core. Winding the wires along the outside results in an increased amount of leakage inductance, which is minimized by using the twisted wire. The wires are twisted together as A/C and B/D pairs, as shown in the designations used in both Fig 23 and Fig 24. This transformer is particularly convenient as it provides a center tap that is needed when using the bifilar 1:1 balun and the 4:1 and 9:1 trifilar balbals.

Twisted-Wire Baluns

Perhaps the best way to illustrate the distinct advantages of twisted wire is to examine the performance of a couple of variations of the 1:1 balun. The number of configurations for this transformer are too numerous to mention, and we are going to narrow it down to just two forms. The first of these is shown in Fig 25, and consists of a monofilar (single wire) primary and a bifilar secondary. This form of construction is very convenient as it places the unbalanced terminals on one end of the core and the balanced terminals on the opposite. The example to be used here consists of two turns of #26 bifilar and four turns of #26 monofilar wound on a Fair-Rite 2843000102 binocular core. A pair of 47 pF mica capacitors were used for the matching capacitors C_1 and C_2 (ref. Fig 5).

The second form to be evaluated is the trifilar balun shown in Fig 26, which had earlier been shown schematically in Fig 22. Two of these were evaluated, each having three turns of #26 trifilar wire, the first on a Fair-Rite 2843000102 core (as used in the monofilar/bifilar balun) and the second on a 2861000102 core so as to again illustrate the the difference in performance tetween the two ferrite materials. The leakage inductances for both of the trifilar baluns were negligible, so no matching capacitors were required.

As shown in Fig 27, all three transformers show good return loss (>15 dB) down to 1 MHz. At higher frequencies, the monofilar/bifilar balun of Fig 25 degrades beyond 55 MHz while the trifilar baluns of Fig 26 continue to be usable beyond 100 MHz. This performance is repeated in the transmission losses shown in Fig 28, where all three transformers exhibit less than 0.25 dB loss for lower frequencies. The monofilar/bifilar balun of Fig 25 degrades rapidly beyond 55 MHz while the trifilar baluns of Fig 26 show little sign of degradation up to 100 MHz.

Twisted-Wire Transformer Power Amplifier Design

Let's now take a look at a 1 W power amplifier using twisted-wire wide-band transformers. The schematic of Fig 17 is a push-pull augmented class-AB amplifier.[22,23,24] The transistors Q_1 and Q_2 are 2N4427s, and diode D_1 is a 1N4001. The emitter resistors R_1 and R_2 are 6.0 Ω, each comprised of two 12 Ω 1206 size SMT resistors in parallel. This

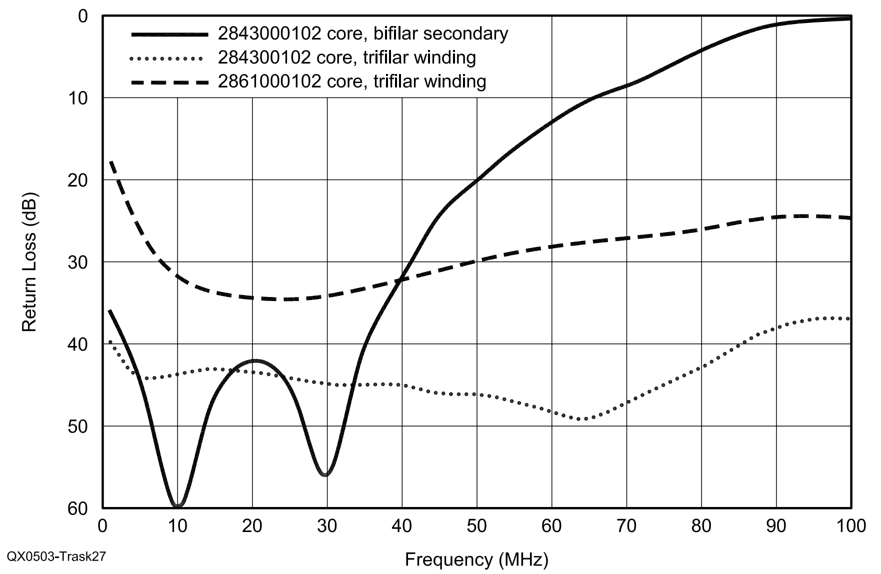

Fig 27—Twisted-wire balun return loss.

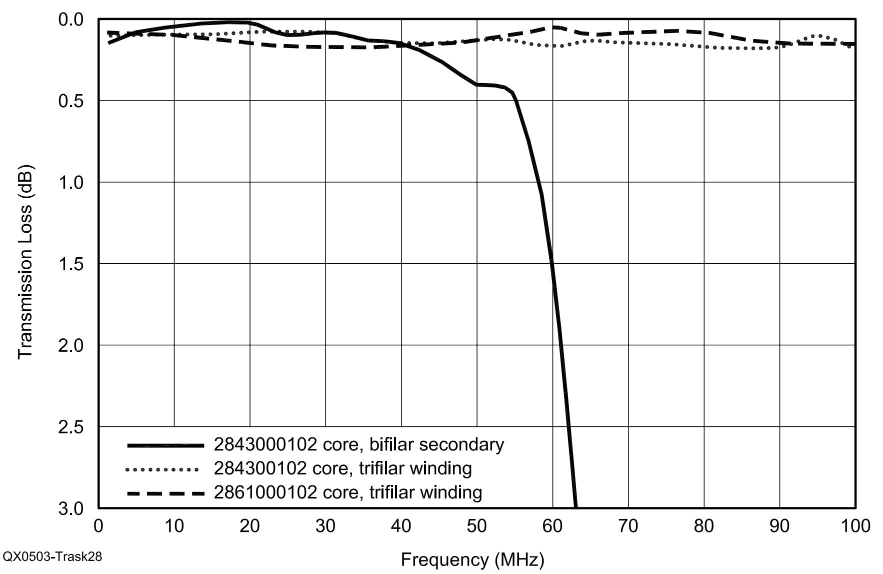

Fig 28—Twisted-wire balun transmission loss.

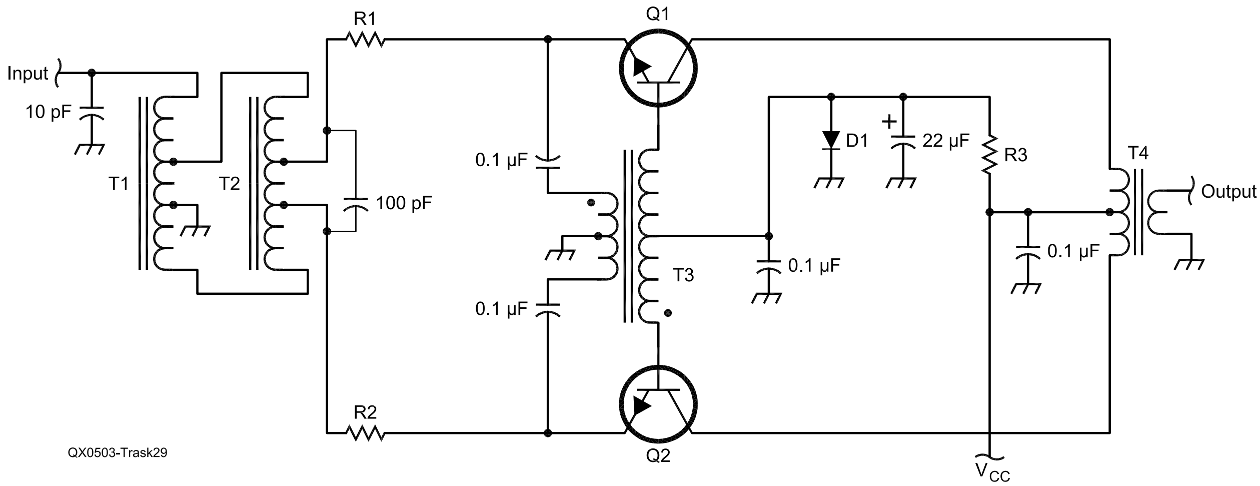

Fig 29—Power amplifier (1 W) using twisted-wire transformers.

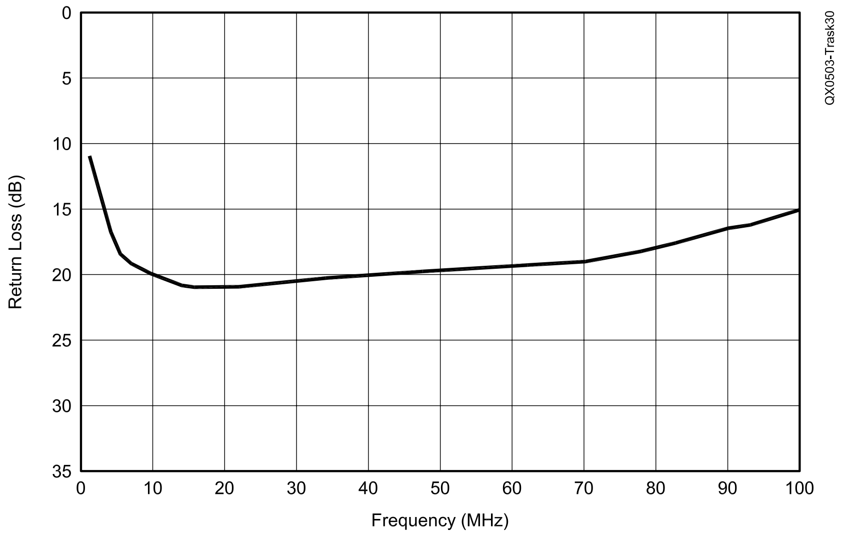

Fig 30—Return loss of 9:1 input balun.

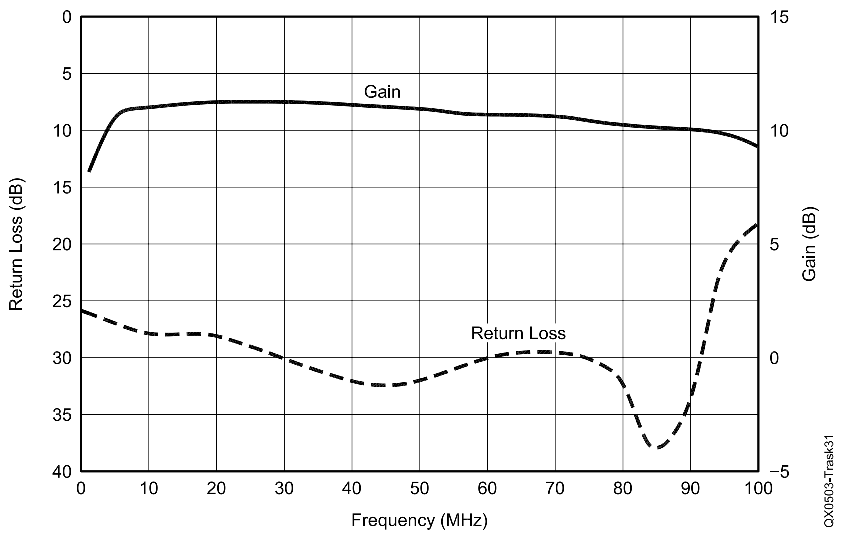

Fig 31—Gain performance of 1 W amplifier.

assumes a 0.25 Ω emitter input resistance for the two transistors. The supply voltage is 12 V, and resistor R_3 is adjusted so that each transistor draws a quiescent bias current of 10 mA.

T_1 is a 1:1 balun transformer made with three turns of #32 trifilar wire wound on a Fair-Rite 2843002402 binocular core as shown in Fig 26. The 4:1 balbal transformer T_2 is made as shown in Fig 24, with three turns of #32 bifilar wire on each side of a Fair-Rite 2843002402 binocular core. The augmentation transformer T_3 is also made on a Fair-Rite 2843002402 binocular core with #32 bifilar wire, with one turn on the primary side and two turns on the secondary side. The 4:1 output balun transformer T_4 is made of three turns of #26 trifilar wire on a Fair-Rite 2843000102 binocular core, as shown in Fig 14.

A few words should be said here to explain the theory of linearity augmentation, a new method for improving the linearity of common-base amplifiers. In the schematic of Fig 29, the signal voltages at the emitters of transistors Q_1 and Q_2 are inverted, amplified, and coupled to the bases of transistors Q_1 and Q_2. These emitter signal voltages are a result of the finite nonlinear emitter resistances of the two transistors, and are the principal source of nonlinear distortion in common-base amplifiers. By inverting these voltages and coupling them to the transistor bases, the signal voltages at the emitters are reduced by as much as 85% for a transformer turns ratio of 2:1. As the turns ratio is increased, the input current to the transistor emitters becomes increasingly dependent upon the fixed resistors R_1 and R_2, thereby improving the linearity.

In the push-pull amplifier, the action of the augmentation transformer further causes the transistors to turn on and off faster during their respective negative and positive cycles, thereby improving the collector efficiency to 95% or more.

The input return loss for the transformers T_1 and T_2 is shown in Fig 30. These tests were made with the transformers terminated with a single 12 Ω load resistor to make the measurements easier and to show the performance of the transformers themselves. The matching capacitors were 10 pF at the input and 68 pF at the 12 Ω load resistor.

The gain and return loss of the amplifier is shown in Fig 31. By applying augmentation, the emitter input resistances of the transistors are reduced to about 0.25 Ω, which with the 6.0 Ω emitter input resistors provides an almost perfect match out to 95 MHz after increasing the value of the second matching capacitor to 100 pF to accommodate the inductive input reactance of the two transistors. With a current gain of 2.0 at the input and an additional current gain of 2.0 at the output, the power gain should be in the order of 12 dB minus 0.75 dB for the three transformers, or 11.25 dB. The power gain is at the expected 11.25 dB for most of the HF bands, and has a low cutoff frequency of around 1 MHz which could be improved with more turns on the three transformers. The gain drops slightly with increased frequency due to stray capacitance at the transistor collectors.

Acknowledgements

I would like to thank Fair-Rite Products Corporation for having been generous in supplying some of the ferrite cores that were used in the preparation of this article. The participation of manufacturers is always welcome when making extensive evaluations such as has been made here, and especially when small quantities of parts used in such evaluations are not readily available through distributors.

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