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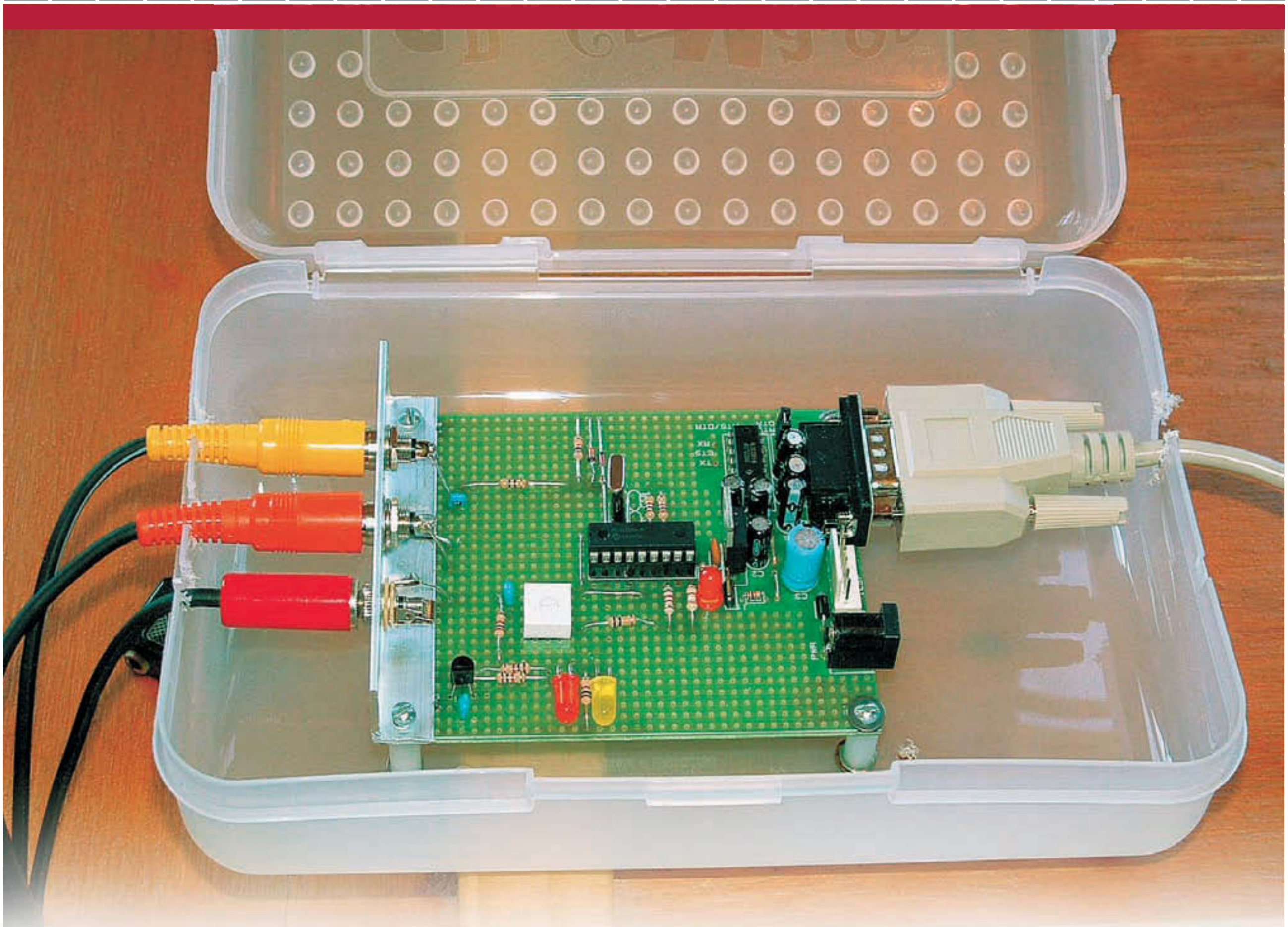

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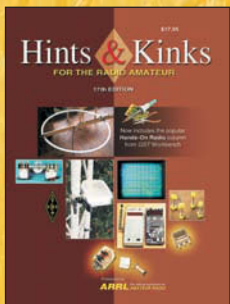

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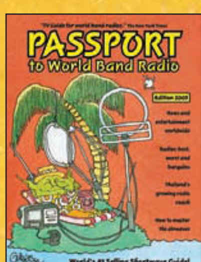

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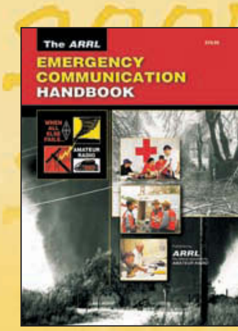

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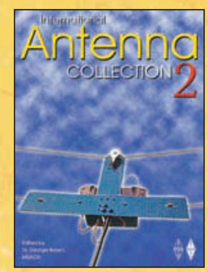

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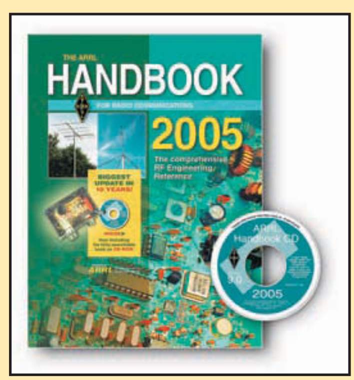

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About the Cover

A $3 PIC makes a simple TNC that can be adapted for several digital modes. The story begins on page 16.

Features

Columns

Mar/Apr 2005 *QEX* Advertising Index

Empirical Outlook

It Looked Good on Mountains of Paper

Representative government is the solution to society's need to organize and provide for itself. Our chief responsibility as citizens is to provide information to our leaders about what to do and how to do it. We judge leadership by placing our leaders' lists of accomplishments against the yardstick of our expectations.

In the 1930s, Congress passed much of its lawmaking authority to bureaucracies it created. One of those is the Federal Communications Commission (FCC), which falls under the supervision of those who manage the Commerce Department in the USA. The Commission provides ample opportunity for feedback regarding proposed legal changes, then its members vote on them. Elected officials in Congress have prerogative powers to overturn any decision. The executive branch of our government is charged with enforcing adopted decisions. It sounds like a good check-and-balance system, doesn't it? Can it be improved?

Michael K. Powell, son of erstwhile Secretary of State Colin Powell, departs as Chair of the FCC this month. Get out your yardstick and let's see whether his self-proclaimed list of accomplishments stacks up. He points to heavy fines levied against stations airing lewd material, such as those carrying the Super Bowl last year and those carrying Howard Stern's show. Ask yourself whether such televised material has decreased in general. Powell extolled the virtues of broadband-over-power-line (BPL) technology. Ask whether BPL has done more good than harm, given that much of its occupied bandwidth transports the very content he promised to fight. Powell embraced relaxed rules against owning too many stations in major markets. Ask if that has improved our lot. Ask if Powell's FCC heeded the feedback we gave, especially on those decisions affecting Amateur Radio. Consider whether his actions reflected more concern for economics than regulation. Finally, ask if the mountains of paper generated and promulgated by the FCC are worth your hard-earned dollars.

Well, it's no surprise that Mr. Powell looks good in our rear-view mirror. The question at the time of this writing is: Who comes next? Somewhat unlike Supreme Court justices, FCC candidates may be scrutinized at confirmation hearings about their proposed courses of action. After all, the FCC belongs to the same branch of government as the confirming body. Take firmly your governmental reigns and write your Congresspersons. Now is the best time to apply your leverage to the situation. If you haven't already done so, get Edward Tenner's *Why Things Bite Back: Technology and the Revenge of Unintended Consequences* (Knopf 1996, ISBN 0679747567). Read it and include in your letter some of his brilliant examples as quotations. Get with your friends and add their signatures to your letter. Send a copy to President Bush. Pray.

Our answer is yes, we can improve things. If you feel we are wrong, forget about Tenner and pick up William J. Lederer's *A Nation of Sheep* (W.W. Norton, 1961, ISBN 0393052885).

In This Issue

Chris Trask, N7ZWY, looks at designing broadband transformers, especially for power amplifiers. Chris covers parasitic effects, magnetic materials and various topographies.

Bob Ball, WB8WGA presents an inexpensive TNC that you can build for $25. Now what packet projects could you build with that?

John Champa, K8OCL, and John Stephensen, KD6OZF, describe high-speed UHF modems. Richard Kiefer, KØDK, studies end-fed verticals for hand-held transceivers. Read about the theory behind them, construction methods and results.

Al Christman, K3LC, begins a two-part study of phased vertical arrays and their ground systems. Al concentrates on four-square and certain other configurations in a thorough analysis. In "Antenna Options," LB Cebik concludes his tale of Yagis with practical construction techniques. I close the issue with a paper about distortion and noise in OFDM systems.—*Doug Smith, KF6DX;* **kf6dx@arrl.org.** □□

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.

By Chris Trask, N7ZWY

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.

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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

[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:

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

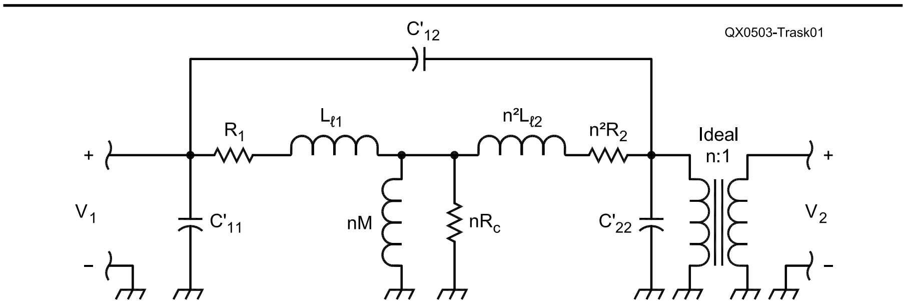

Fig 1—Complete wideband transformer model.

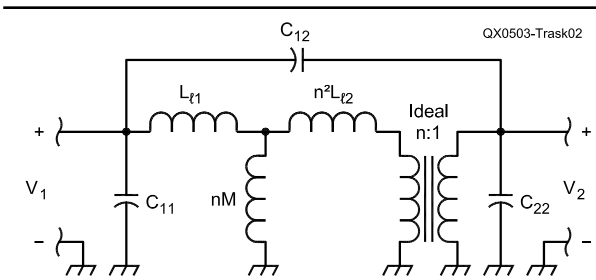

Fig 2—Lossless wideband transformer model.

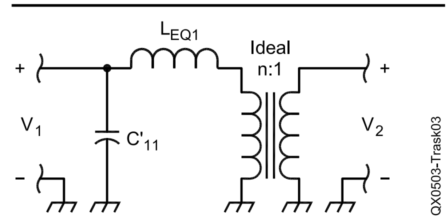

Fig 3—Equivalent circuit for determining C'_{11}.

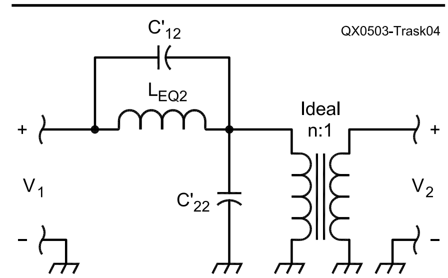

Fig 4—Equivalent circuit for determining C'_{12} and C'_{22}.

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.

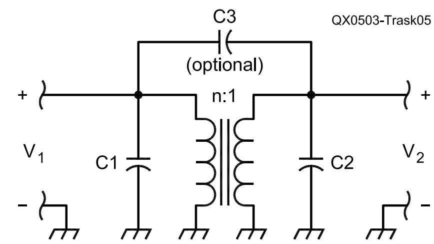

Fig 5—Matching network components.

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:

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

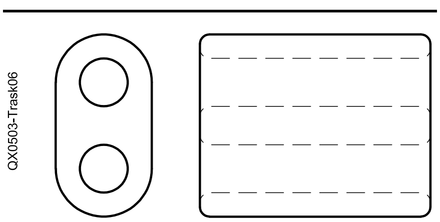

Fig 6—Binocular core.

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

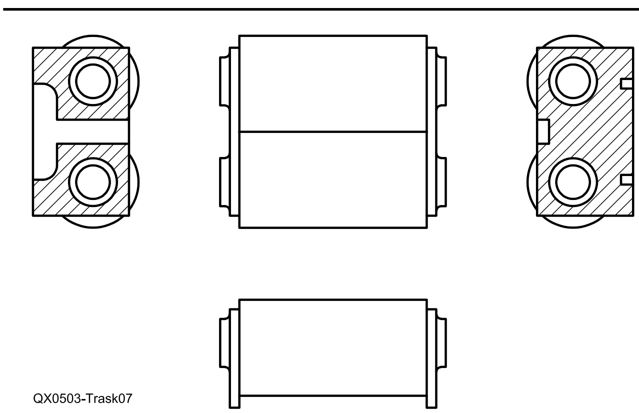

Fig 7—High-power rf transformer core.

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

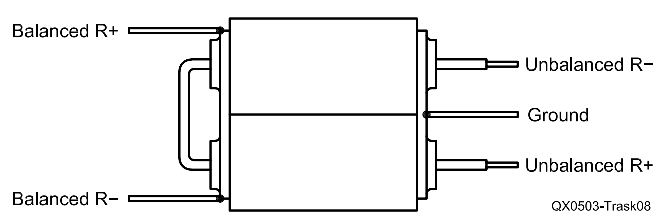

Fig 8—Conventional 1:1 balun transformer.

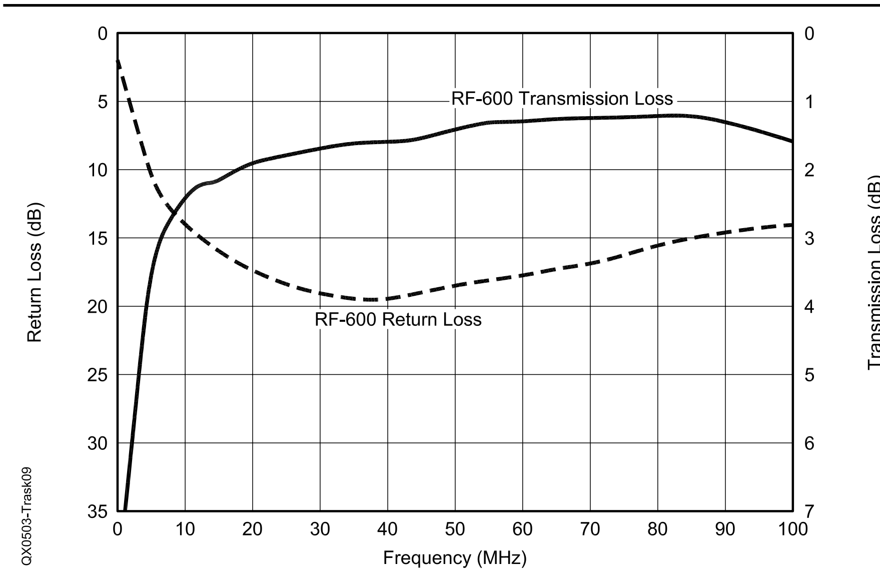

Fig 9—Conventional 1:1 balun transformer performance.

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:

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

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:

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

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:

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

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

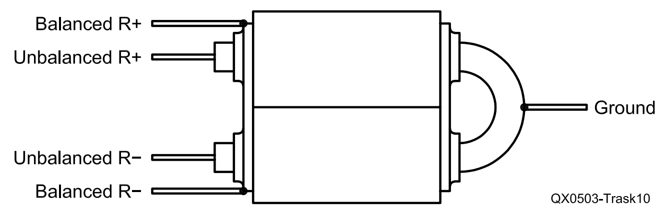

Fig 10—Transmission-line 1:1 balun transformer.

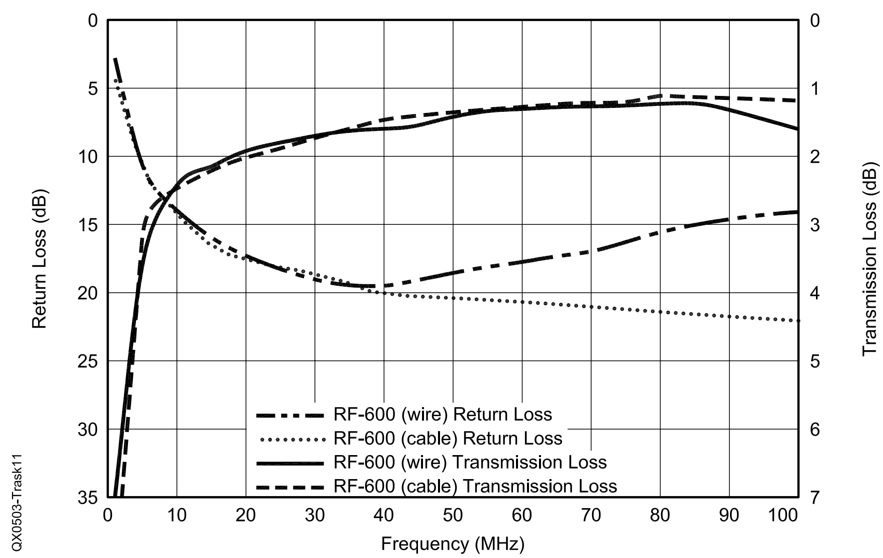

Fig 11—Conventional vs. Transmission-line 1:1 balun transformer performance.

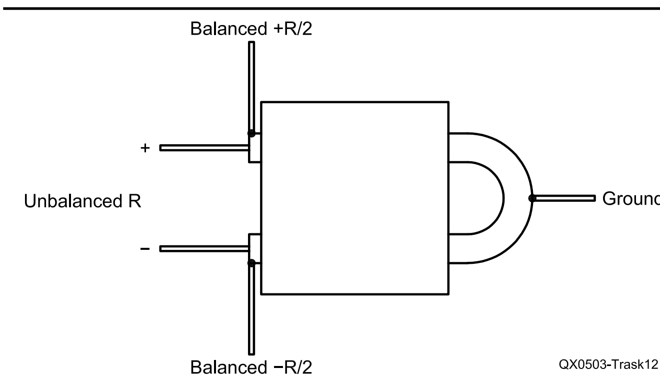

Fig 12—Coaxial transmission-line 1:1 balun.

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

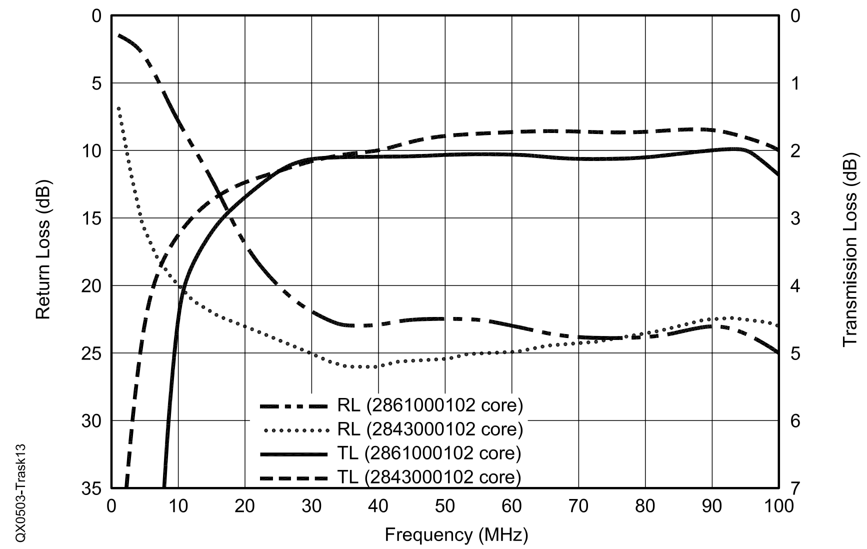

Fig 13—Coaxial transmission-line 1:1 balun performance (-0102 cores).

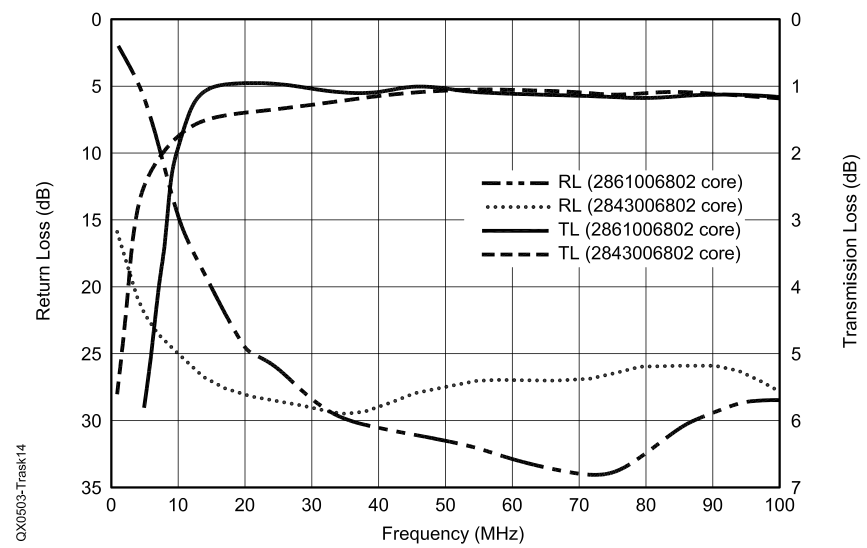

Fig 14—Coaxial transmission-line 1:1 balun performance (-6802 cores).

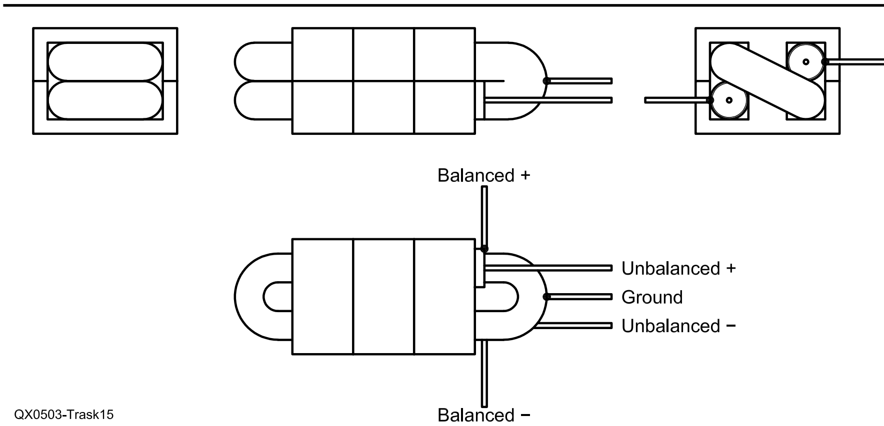

Fig 15—Extended coaxial transmission-line 1:1 balun using e-cores.

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

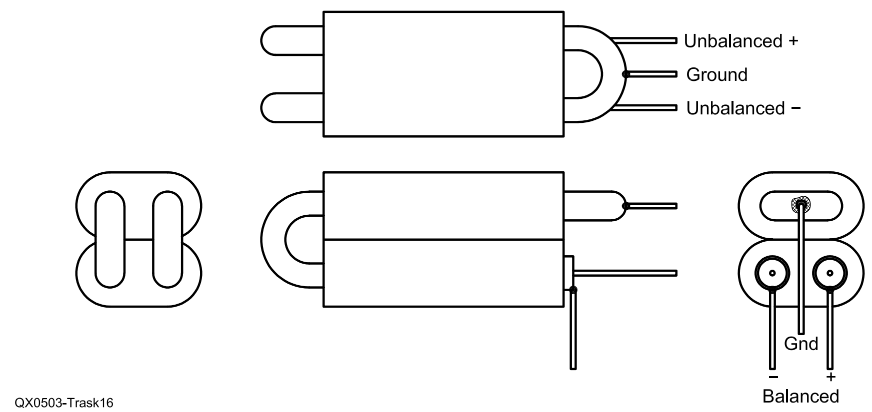

Fig 16—Extended coaxial transmission-line 1:1 baluns using binocular cores.

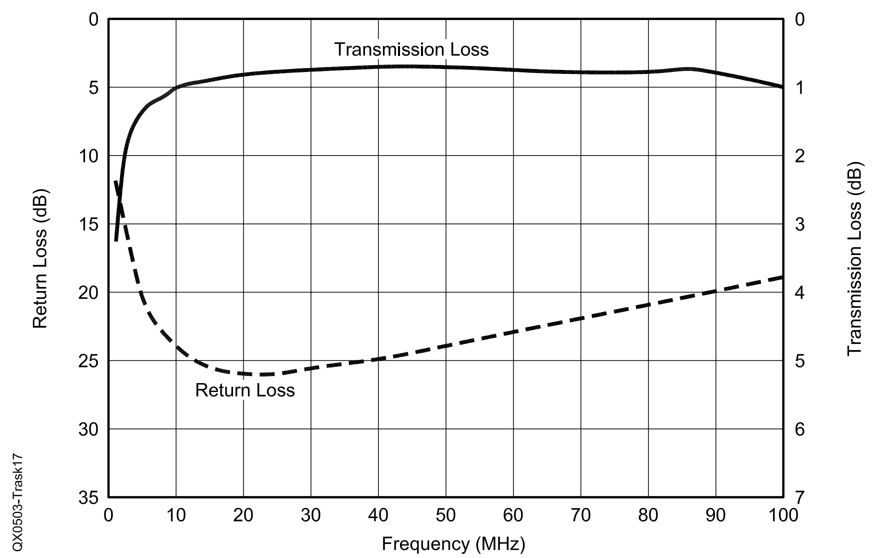

Fig 17—Extended-length coaxial transmission-line 1:1 balun performance.

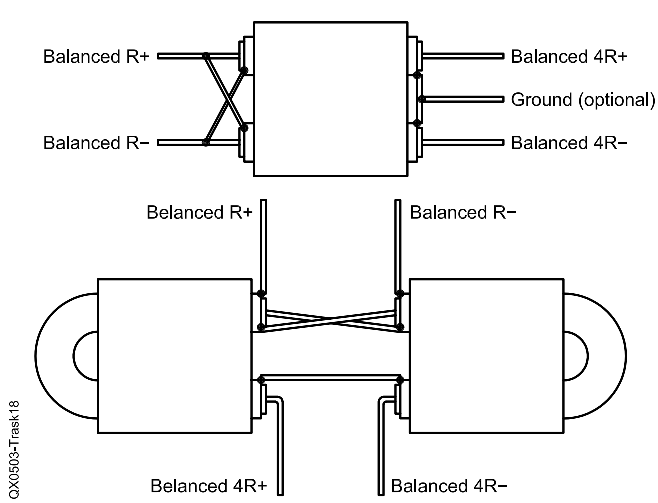

Fig 18—Coaxial transmission-line 4:1 balanced transformers.

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

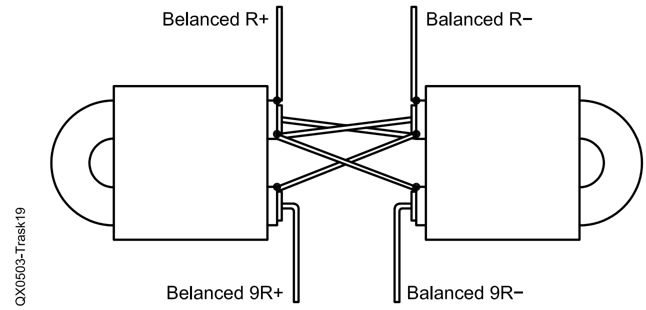

Fig 19—Coaxial transmission line 9:1 balanced transformer.

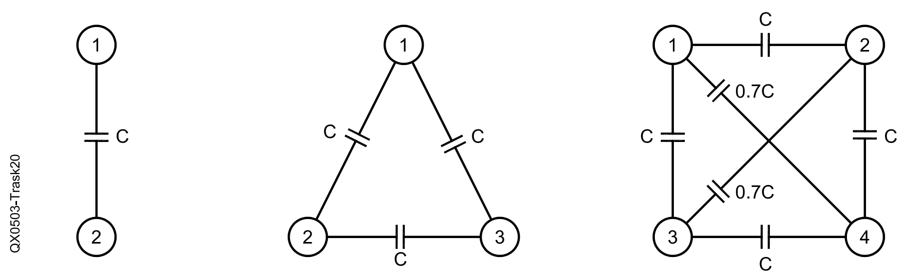

Fig 20—Twisted-wire capacitances.

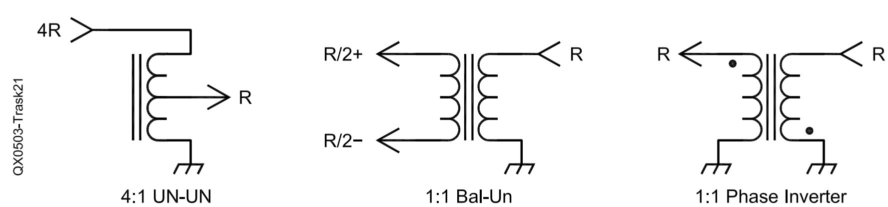

Fig 21—Two-conductor twisted-wire transformer configurations.

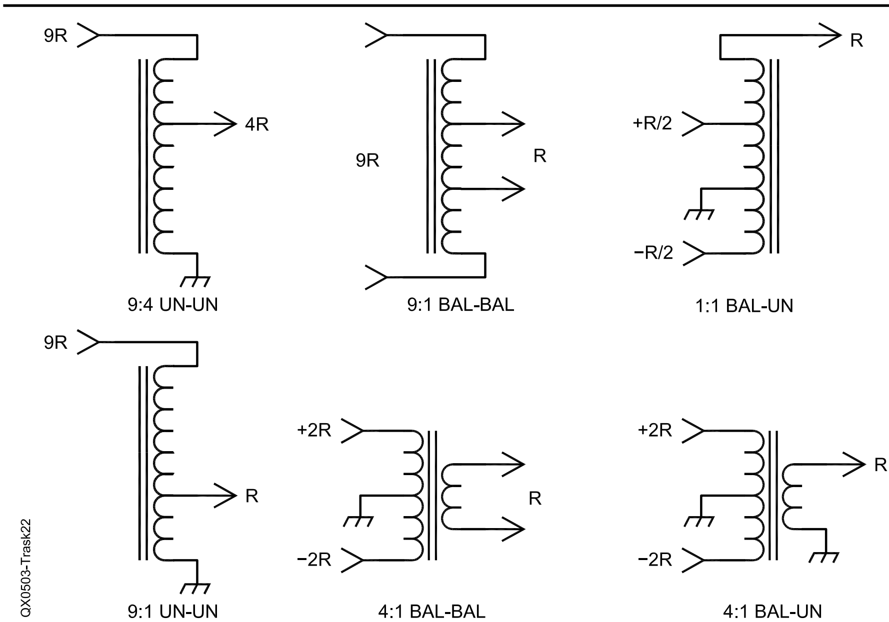

Fig 22—Three-conductor twisted-wire transformer configurations.

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

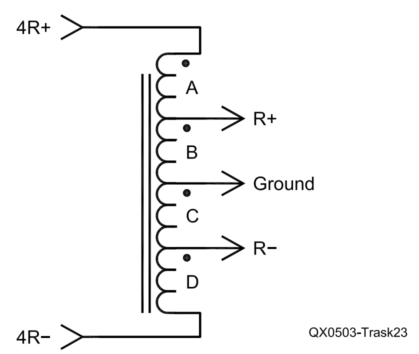

Fig 23—Balanced-balanced transformer, 4:1 transformation ratio.

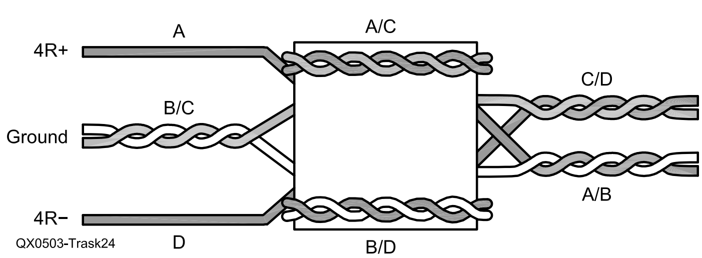

Fig 24—Balanced-balanced transformer, 4:1 transformation ratio.

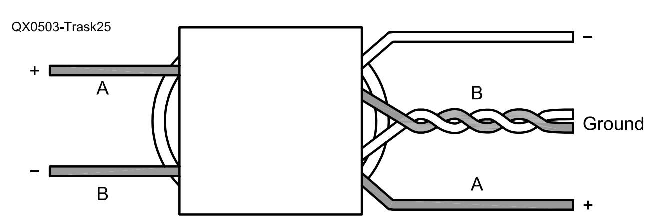

Fig 25—Balun using monofilar primary and bifilar secondary.

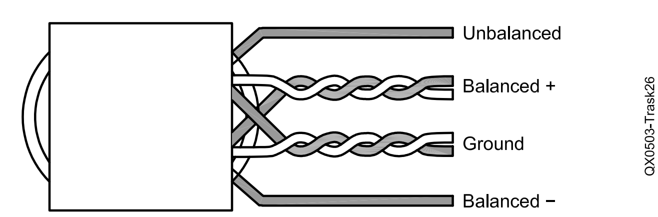

Fig 26—Trifilar wound 1:1 balun.

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

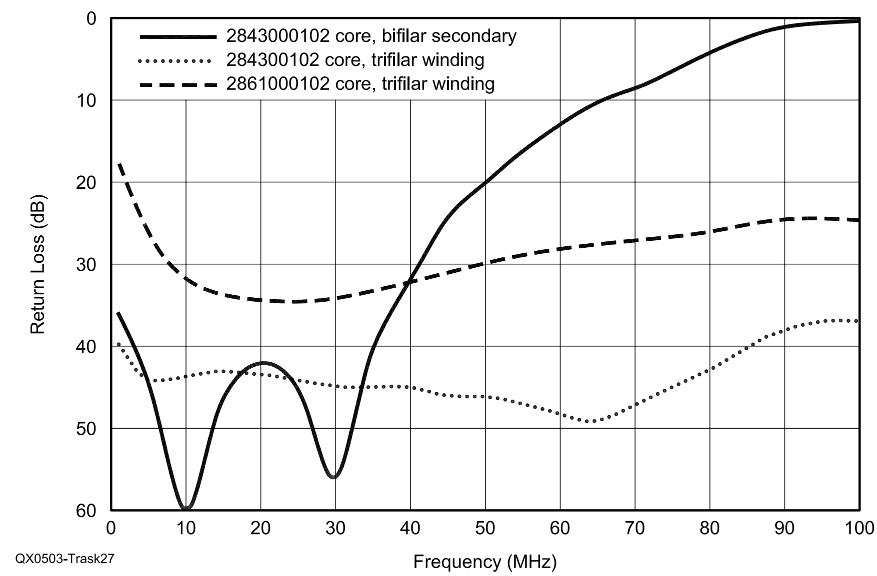

Fig 27—Twisted-wire balun return loss.

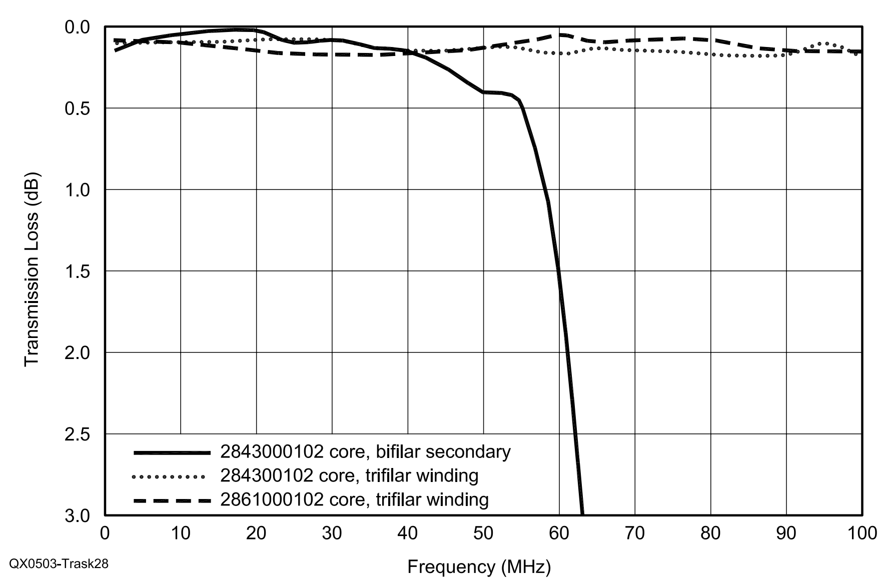

Fig 28—Twisted-wire balun transmission loss.

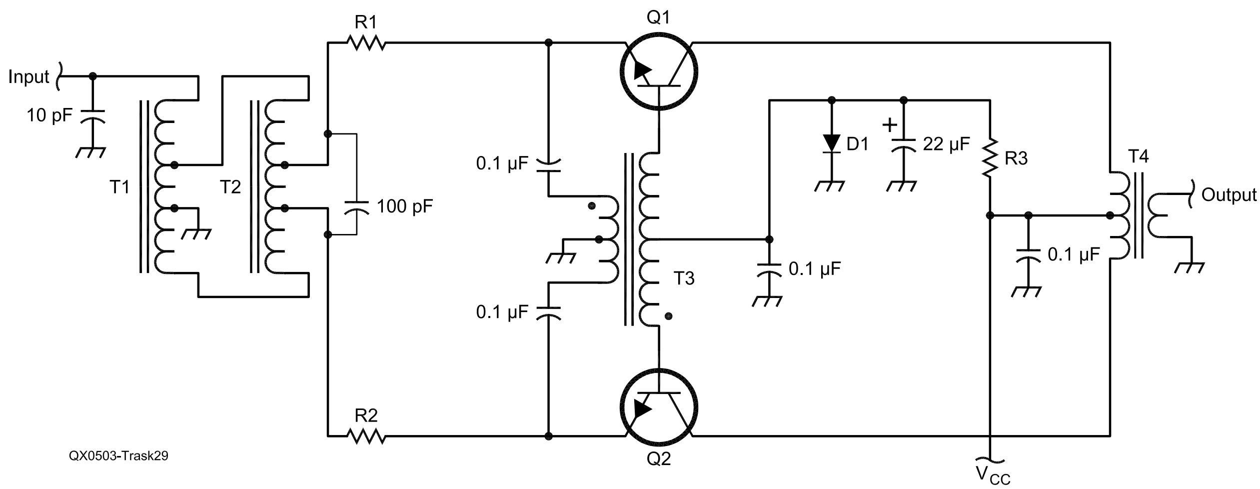

Fig 29—Power amplifier (1 W) using twisted-wire transformers.

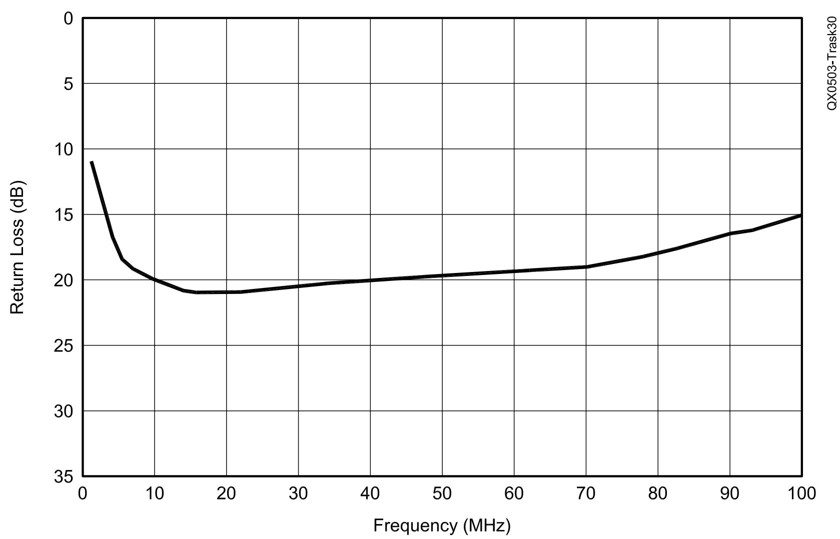

Fig 30—Return loss of 9:1 input balun.

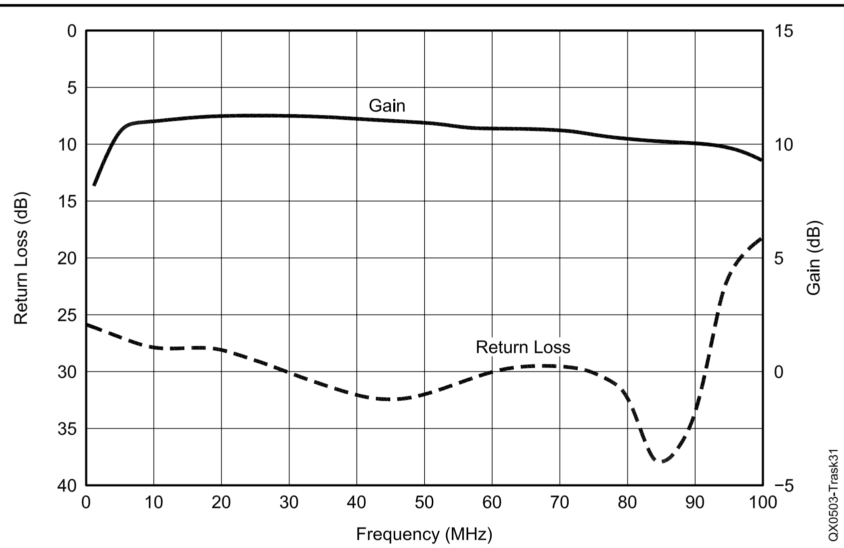

Fig 31—Gain performance of 1 W amplifier.

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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An Inexpensive Terminal Node Controller for Packet Radio

An inexpensive processor chip forms the basis for a simple-to-build TNC.

By Bob Ball, WB8WGA

Most hams who have tuned around on the 2-meter band at one time or another have heard the infamous *blurrp* of packets being sent back and forth. Sometimes it is an informal QSO, sometimes an Amateur Packet Position beacon, or maybe someone sending info on a hot DX station. Packet radio has found a wide number of applications since its introduction in the 1980s.

Did you ever want to have an inexpensive way to monitor local packet activity? Or perhaps set up your own digipeater to get a communications link to a particular point? Or possibly, actually do some programming on your own simple Terminal Node Controller (TNC)?

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If so, this simple project might be the project you have wanted. The unit can easily be built for under $25. After it is completed, the software supplied will:

- monitor local packet activity
- act as a complete digipeater
- send packet beacons at user defined intervals containing your defined text
- allow you to communicate in a round-table fashion using "converse" mode
- send APRS NMEA position reports when connected to a GPS receiver.

Note that this unit does *not* provide all functionality of the complete commercially available TNC units. It is a simplified version that captures enough functionality to get you on the air and have some fun with packet radio. If you want to expand your programming knowledge and explore AX.25 a bit, the unit described will allow you to modify the software to support just about any application you might have been considering for packet radio. This could include, for example, anything from temperature monitoring, GPS interfaces or remote control applications. The list of possibilities goes on and on. All source code for the unit is provided as a starting point and programming is possible without buying an expensive programmer. We will provide additional information on programming the processor later in this article.

I got started on this project after looking at the functionality of some of the new microprocessor chips. They just keep getting more powerful while the cost keeps dropping. After checking out a few different ones, I became attracted to the *MicroChip* series, in particular, one designated the

PIC16F88. This 18-pin dual-inline chip sells for about $3 in single units. It provides several analog to digital (A/D) channels, includes a UART to talk to a terminal, a programmable Flash program memory and a programmable EEPROM memory to store call signs and options. It can execute a program instruction every 200 nsec. Amazing! It is certainly not the Intel 8080 I used back in the early 70s! After looking at the unit I began to wonder if it could form the basis of a simple TNC.

One of the things I discovered is that many of the existing PIC processor based packet TNCs use the MX614P modem chip. Modem chips are great building blocks but do add to power consumption and cost. In addition, they can be difficult to obtain. Most of the existing TNC project designs use processor chips that have been replaced by more powerful and less expensive units.

This article will describe how to use the enhanced internal capability of these new processors to code and decode the digital signals. While the project described here is designed for packet radio, the project board and modemless techniques described should work for just about any digital mode (RTTY, Pactor, etc). So, once you have built this unit for packet use, with a little work and ingenuity, it can be reprogrammed for enhanced packet radio or other amateur digital modes.

As with many ham projects, I was able to build on the published work of many others. In particular, Mike Berg, NØQBH,[1] has done some work on modem-less receiver design using an external comparator. A number of PIC programmers[2] have documented resistor ladder network designs for older technology PIC processors that gen-

[1]Notes appear on page 25.

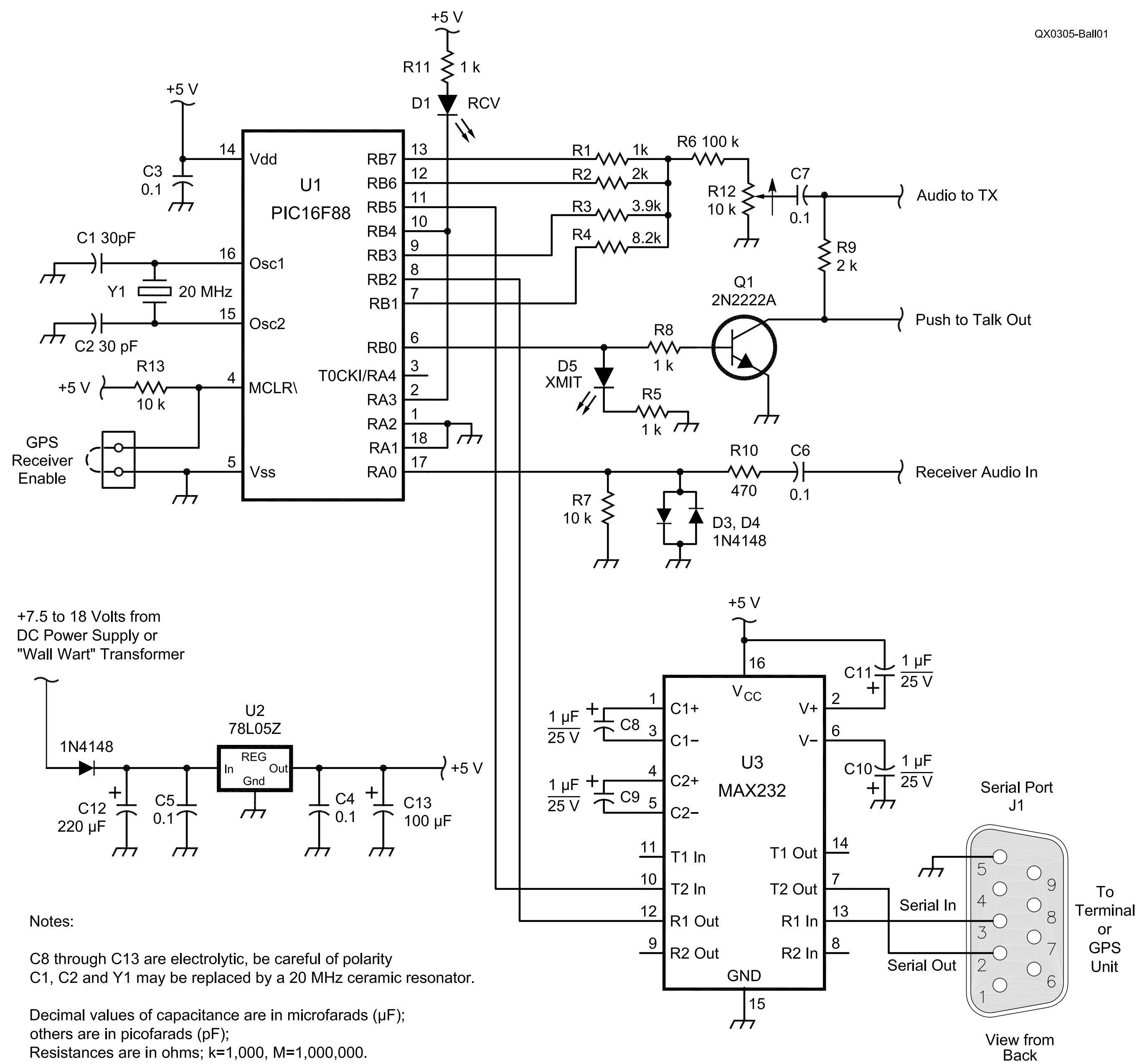

Figure 1—Schematic of the inexpensive TNC.

erate a clean sine wave needed to transmit Audio Frequency Shift Keying (AFSK) tones for transmit. I have combined this existing receive and transmit work, added a command interface which is patterned after the Tucson Amateur Packet Radio (TAPR)[3] terminal node controllers, and made use of the newest PIC technology to produce a simple TNC. It provides the following features:

- A command interface is provided that allows configurable options to be set and stored in EE ram and restored on subsequent resets.
- A monitor feature is included to allow all packets, no packets, or just packets addressed to my station to be displayed on the terminal attached to the serial port.
- Digipeating of packets (up to 255 bytes/packet) is supported.
- Alias capability allows the unit to function as RELAY type digipeater.
- Beacon capability with user set parameters is supported.
- Converse mode operation to transmit text typed from the serial port supporting "round table" chat type operation is provided.
- Support is provided to transmit APRS position beacons when a GPS unit is attached to the serial port.

Circuit Description

A schematic of the TNC is shown in Figure 1. It contains two integrated circuits, one for level conversion of the RS232 signals and the microprocessor. The microprocessor is the 18-pin DIP version of the PIC16F88. The popular MAX232 chip does the RS232 conversion.

Connections to the radio consist of audio input from the speaker jack, audio to the microphone circuit, and a connection to the push to talk circuit.

The serial port is used to connect to a standard terminal program or to a GPS receiver that outputs NMEA sentences. The terminal program is used to display received packets, send text while in the CHAT mode, and do some initial setup of the unit. Jumper J4 is used to specify what is connected on the serial port.

Note that after initial setup is complete and the station parameters are set in EERAM, the unit will run as a super simple remote digipeater without the MAX232 and serial port connections. With this configuration, it becomes a one-chip TNC.

Power to the unit can be anything from 7 to 18 V. Power from the 12 V radio can be used, or an inexpensive *wall wart* will do the job.

Receiving Packets

Packet radio, in common with many digital transmission techniques, uses *Audio Frequency Shift Keying* (AFSK) to send the stream of *ones* and *zeros*. It uses two frequencies, 1200 and 2200 Hz to represent a change in the value of the bit stream. For receiving purposes, this TNC uses a comparator to find zero crossings of the AFSK sine wave. The TNC calculates the interval between zero crossings to determine the frequency present. The software then converts the changes in tones to a data stream. To imagine this, think of a sine wave at either 1200 or 2200 Hz. If we clip the sine wave with a couple of diodes (D1 and D2 in the schematic), we get a square wave. The circuit is shown in Fig 2.

If the resulting square wave is fed to a zero-crossing-detector, a pulse is generated at every crossing. If the processor internally times the interval between these pulses, it can determine the frequency of the sine wave. The F88 processor has timers and an internal comparator circuit so all that decoding work can be done inside the chip. The software inside the PIC can also do some digital filtering and throw out frequencies outside of the passband of the audio. The resulting modem-less design is extremely sensitive and rivals the performance of the commercial modem chips like the MX614. My units routinely decode packets that are less than 1 S unit on my 2-meter radio from stations over 75 miles away.

It should be noted that the above technique could be used for sampling the frequency of any sine wave. It should work for any of the digital modes that use AFSK or for any applications that need to detect an audio tone of a certain frequency.

Transmitting Packets

To transmit, the software needs to generate either a 1200 or 2200 Hz sine wave. The frequency is changed to indicate the transmission of a digital one or zero.

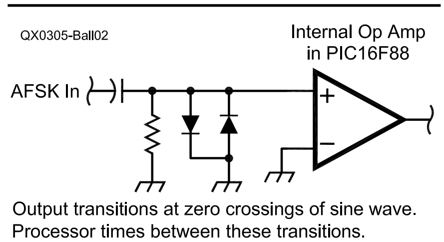

Figure 2—Determining the frequency of a sine wave using a zero crossing detector.

A sine wave is no more than a varying voltage level over time. If the processor can change the voltage level in the right amount and time frame, a sine wave with low distortion can be produced. The TNC must only generate a single tone at a time, although production of multiple simultaneous tones (i.e. DTMF) is possible. MicroChip has documented this technique in one of its application notes if you are interested in more details.

This generation can be done in software by controlling a resistor ladder network that has the correct values to generate a sine wave. If the voltage is changed often enough (say 32 times per audio cycle) at the right value, a sine wave is produced. Figure 3 shows the voltages that are generated at the wiper of R12 when the Ports RB7, RB6, RB5 and RB1 are taken through a binary sequence by the program at the correct frequency. Increasing the number of samples decreases the distortion of the sine wave but increases the real-time requirement. A sine wave adequate for this application can be generated using 32 points.

Software

Approximately 3000 words of program instruction reside on the PIC chip to do the job of taking the received packet, storing it, checking it for validity, sending the info to the serial port, and re-transmitting the packet as necessary. If you are not interested in studying or modifying the provided software, a copy of the working assembled software is provided that can be downloaded into the processor (see next section).

If you are interested in learning about PICs or how packets are formed, etc, you might want to modify the code to meet your needs. While the complete software structure will not be described here, a copy of the commented source code is available[4] and can be used as a starting point for further experimentation. The software is organized by functional processes (ie receive packet, digipeat, command processor, etc) and scheduled in a round robin manner to make modification easy. The terminal interface is implemented as a table driven function, so that additional input commands can be easily added.

The software in this unit uses published snippets of code from many hams who have shared their work on the Web. References and credits are given to all the authors in the source software. Please remember to maintain these credits if you modify the software and pass it on to others.

Construction

Construction and layout of this unit is not critical. One option is to build on a standard perforated board. The unit shown in Figure 4[5] was built on standard RadioShack prototype board with two aluminum angle brackets for supports. As with all CMOS integrated circuits, use caution when inserting the ICs to avoid electro static damage. A good wrist-grounding strap is very useful.

This unit runs at a relatively fast clock speed so bypassing of the 5 V power is important. The 0.1 µF capaci-

Spreadsheet to Calculate Sine Wave Value for TNC Transmitter							
R1	1000						
R2	2000						
R3	3900						
R4	8200						
R6	100000						
R12	10000	This is with wiper at top of pot					
					Ladder	**Time**	**To**
	MSB			**LSB**	**Out**		**Transmitter**
Value	**V1**	**V2**	**V3**	**V4**	**Eo**	**t**	**Vo**
7	0	5	5	5	2.33	1	0.21
6	0	5	5	0	2.00	2	0.18
5		5		5	1.65	3	0.15
4		5			1.32	4	0.12
3			5	5	1.00	5	0.09
2			5		0.68	6	0.06
1				5	0.32	7	0.03
1				5	0.32	8	0.03
1				5	0.32	9	0.03
1				5	0.32	10	0.03
2			5		0.68	11	0.06
3			5	5	1.00	12	0.09
4		5			1.32	13	0.12
5		5		5	1.65	14	0.15
6		5	5		2.00	15	0.18
7		5	5	5	2.33	16	0.21
9	5		0	5	2.97	17	0.27
10	5		5	0	3.33	18	0.30
11	5	0	5	5	3.65	19	0.33
12	5	5		0	3.97	20	0.36
13	5	5	0	5	4.29	21	0.39
14	5	5	5	0	4.65	22	0.42
15	5	5	5	5	4.97	23	0.45
15	5	5	5	5	4.97	24	0.45
15	5	5	5	5	4.97	25	0.45
15	5	5	5	5	4.97	26	0.45
14	5	5	5	0	4.65	27	0.42
13	5	5	0	5	4.29	28	0.39
12	5	5	0	0	3.97	29	0.36
10	5		5		3.33	30	0.30
9	5	0	0	5	2.97	31	0.27
	Note when Vo is Plotted over time, a sine wave is produced						

Figure 3—Sine wave generation on a PIC.

tors on each integrated circuit should be located right at the device with leads of one inch or less. I can't tell you how many times I have spent time debugging a digital circuit problem only to find out I had noise on the 5 V line. Bypass capacitors are cheap and small, use them freely.

If you make your own board for this project, it is always good to do some checkout before you install the integrated circuits. First apply 12 V and check for +5 on pin 14 of the PIC16F88 and pin 16 of the MAX232. If this looks good, install the chips using the wrist strap. Don't forget to program the microprocessor chip (see later section) before installation.

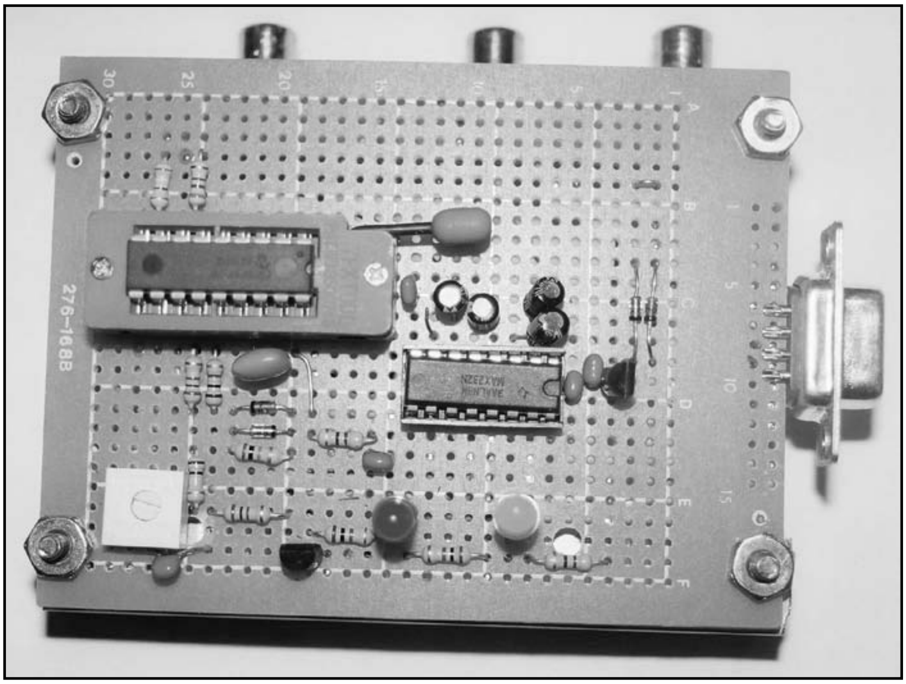

Figure 4—Breadboard style unit.

After building the prototype shown in Figure 4, I discovered another construction option, the popular Olimex prototyping board. This is the fastest way to construct the TNC.

Many projects need the basic processor, crystal, power circuit and serial connection. Olimex has provided a built and tested circuit board with this much of the circuit complete. It also includes an area to build the rest of the circuit. In the case of the TNC, approximately 1/2 of the components are already on the board.

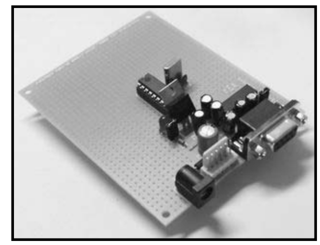

Figure 5—Olimex prototype board.

The purchased board from Olimex contains the processor chip, all serial

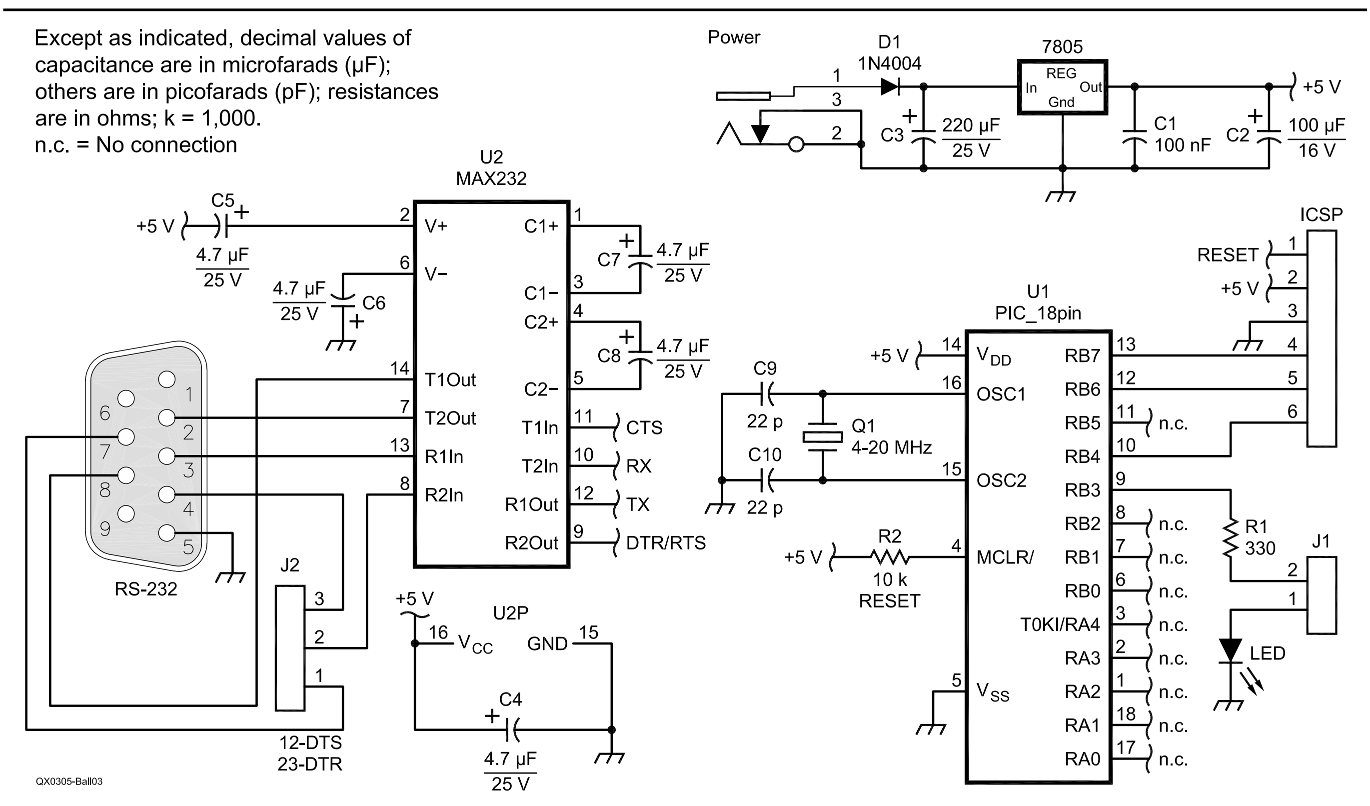

Figure 6—Schematic of the Olimex board.

Detailed Instructions for the Olimex Board

When you receive your Olimex board, it is a good idea to check it out before you add any wiring. I recommend installing the Boot Loader program which is supplied with a small checkout program that will flash the leads on the board.

Initial Checkout

1. Remove the supplied PIC16F88 and program it with the supplied Bloader Program. This should be the only time you will need an external programmer for this project. REMEMBER TO DISABLE THE MASTER CLEAR WHEN THE CHIP IS PROGRAMMED.

2. Reinstall the processor chip. With no serial port connected, power up the unit. The LEDs should alternately flash. If they do not, stop here and get a replacement board from your supplier.

Now go ahead and add the TNC wiring:

Serial connections

1. Remove J1
2. If GPS operation is desired, install wiring for GPS Receive Enable. Do not install jumper until set up is complete (see text).
3. Lead Labeled RX to Processor Pin 11
4. Lead Labeled TX to Processor Pin 8

TNC Receive Circuitry

1. Ground Pins 1 and 18 of Processor
2. Connect Processor Pins 2 and 10 to D1 Cathode
3. Connect R11 to 5V
4. Connect R10,D2, D3, R7 to Pin 17
5. Ground other lead of D2,D3, and R7
6. Connect R10 other end to C6

TX Receive Circuitry

1. Connect one end of R1,R2,R3, and R4 to Processor Pins 13,12,9, & 7 respectively.
2. Connect other end of R1,R2,R3, and R4 to R6
3. Connect other end of R6 to R12
4. Ground other end of R12
5. Connect one end of C7 to R12
6. Connect other end of C7 to output and R8
7. Connect Pin 6 of Processor to D5 Anode and other end of R8
8. Connect other end of D5 cathode to R5
9. Connect other end of D5 to ground
10. Connect R8 other end to Q1 Base
11. Connect Q1 Emitter to Ground
12. Connect Q1 Collector to R9 and PTT out
13 Connect other end of R9 to C7

Wiring is Complete. Now install the TNC software

1. Remove Power from the TNC
2. Connect a serial cable from the serial port to your PC or terminal. Start up the terminal program.
3. Remove J4
4. Install the PC portion of the loader program called Screamer
5. Start up Screamer. Hit the program button and supply the file name for the TNC software. Screamer should say "Waiting for Broadcast"
6. Power up the TNC, Screamer should program the chip and indicate successful completion
7. Exit Screamer program on PC and Start Terminal Program
8. The command interpreter should be running on your TNC. Set your options, connect the radio, and your on the air.

Figure 7—Step-by-step wiring instructions for the Olimex board.

channel components and all power components on a 100×80-mm board. With the addition of a few components, the unit is complete. Connectors for the radio interface can be based on personal preference. I use RCA audio jacks for the audio in and out and a standard phone plug for the push-to-talk connection. This allows use of the readily available audio/video cables for the audio in, transmit audio and PTT. You may want to select these connectors to meet your need.

The Olimex board does not raise the cost of the project significantly. A constructed prototyping board for the PICF88 is available for $12.95 plus shipping directly from Olimex. Other sources of the board[6] provide the board equipped with the PIC16F88 for $18.95 plus shipping. A picture of the Olimex board as it comes out of the shipping box is shown in Figure 5. The schematic of the board is shown in Figure 6. Step-by-step wiring instructions to add the parts to the Olimex board are shown in Figure 7.

A picture of the completed TNC using the Olimex board is shown in Figure 8. Note the case, which is based on a 71 cent plastic school supply container I purchased at my local WalMart store. It does a nice job of keeping the unit clean and dry.

Programming the Unit

Code Assembly

The hex file output for the TNC is provided with the software and may be used directly. If you do not wish to reassemble the software, skip this section.

If, however, you have decided you would like to experiment with the assembly code for this program, the provided source will need to be reassembled. This can be done with the free assembler and simulator provided by MicroChip. First, download and install the free MP Lab development system available from the MicroChip Web site on your PC. Using the instructions provided, assemble the provided TNC assembly file. **In the project options of MPLAB window, set the output format for the hex file to INHX8M.**

Programming the Flash Memory

Now that your code is assembled, there are a couple of options for programming the PIC16F88 chip. If you own a PIC programmer, or have a friend who has one, one method is to just download the provided hex files and you're done. If you choose to do this, the next section on the boot loader can be skipped.

Programming with the Boot Loader

If you intend to experiment with the software or install software updates, the *boot loader* is a nice way to program your unit. Once it is installed, it is much faster and convenient than using an external programming unit. With the boot loader, the chip can be reprogrammed without removing it from the circuit board using the same serial connection that is used during normal operation.

To do this, MicroChip boot loader programmers use the PIC16F88 feature that allows it to write its own instruction words. With this feature, a user can do a one-time program of a small program (called a boot loader) in high flash memory and then use this program to download the user software over the serial link. This means that the user using a regular PC with a serial link to the TNC can load software updates. Several free boot loader programs are available on the Web. I have included a site,[7] Spark Fun Electronics, that provides a freeware version that works well with the TNC. I have also included a copy of the boot loader (with permission) for the PIC16F88 from the Spark Fun Electronics site with the source code.

- Download and install the PC portion of the programmer (called *Screamer*) It is fast!
- Download the F88 version of the bootloader hex file (called Bloader -F88-20Mhz.hex) from the web site.
- Using a PIC programmer, complete the one time program of the F88 chip using the hex *Bloader* code. **When setting up the programmer for this step be sure to disable the Master Clear pin on A5 (MCLR on Pin) in the Config word. This allows PORT A5 to be used as an input lead for the GPS function.**
- Apply power to the TNC. The LEDs should alternatively flash as an indicator that the boot loader program was successfully installed.
- Connect the serial cable between the PC and the TNC. Start the *Screamer* programmer program on the PC. Select the TNC hex file (Modemless TNC V1) and click on **PROGRAM**.
- When the unit indicates *Waiting for Pic Broadcast*, cycle the power (allow 3 or 4 seconds for the capacitors to discharge) on the TNC. *Screamer* will then program the Chip in about 10 seconds.
- Close the *Screamer* program and reopen the terminal program on the PC. The command interpreter should respond to your commands. More info on using *Bloader* is provided in the *readme* file section in the bootloader code.

Your unit is now programmed. The boot loader should remain there for the life of the chip. To reprogram the unit in the future, repeat steps five through seven.

EE Program Memory

Once the program is loaded and

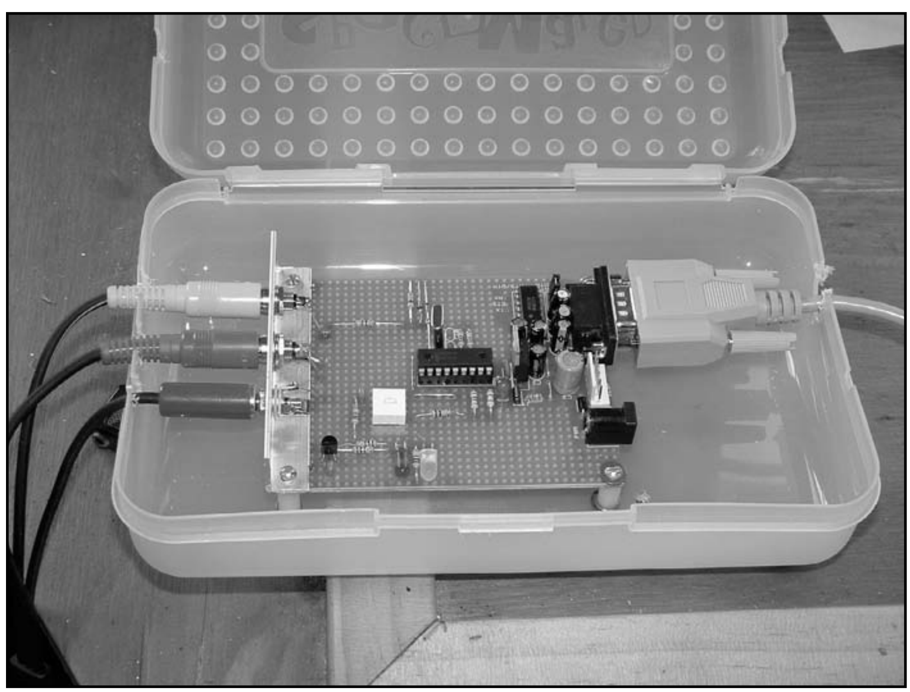

Figure 8—Completed TNC built on the Olimex board.

operational, the unit must be customized for the station using it. Information on the station call sign, beacon info, etc, must be programmed and retained. Like most TNCs, this is done from a terminal. The EE RAM of the PIC is used for this and the information is obtained over the serial port. Then a command (PERM) is executed to write the info into the EE RAM. Once the *perm* command has been executed, the data will last for many years and will withstand program changes and power transitions.

Setting the Options

Station options are set via an ascii terminal on the serial port. I use *Windows Hyperterminal*. It comes with the standard *Windows* software. Other terminal emulators will work as well. Set the terminal up at 9600 baud, 8 bits, no parity, Back Space (BS) for the rubout character, and echo. If the *echo* option is not available on your terminal program, the TNC does provide a command option to allow local echo.

After these options are set, connect a standard serial cable between your terminal and the TNC. After powering up the TNC the following message should be shown.

WB8WGA Modemless TNC V1.08
Type "Help" for Information
cmd:

If this doesn't appear, start checking the wiring around the microprocessor chip and serial connection for errors.

Once you have that working, it is time to receive some packets. Connect an audio cable from your 2-meter speaker audio to the TNC. Open the squelch and set the audio level to the point at which the LED lights. Now close the squelch and the LED should go out. If this step doesn't work, check out the wiring on the incoming audio lead and the LED wiring.

Once this is operational, you should be seeing packets being monitored on your screen.

Before we transmit with your new TNC, it is necessary to set your station parameters and save them in EERAM. The FCC seriously frowns on packets that aren't properly identified. While in command mode, the TNC will accept either upper or lower case characters.

First, you will need to set your call sign using the *mycall* command as follows:

cmd: mycall WB8WGA
OK
cmd:

Note that all callsigns have a 0 to 15 sub id, which is defaulted to zero.

The other parameter that must be set is destination call sign. This is used when the unit transmits beacons, APRS reports, or enters chat mode. This information is set via the *unproto* command. The format for this command is *unproto callsign1 v callsign2,* where callsign1 is the final destination and callsign2(optional) is the digipeater that should be used to get the packet to the destination. A max of 3 digipeaters may be used. An example is shown below:

cmd: unproto n0qbh-15 v wb8wga-10 v relay v wide
OK
cmd:

Transmitting Packets

Now it's time to send out some *blurrps*. Hook up the transmitter connections, audio and push-to-talk. Once connected to your radio, it is necessary to set the transmitter deviation. This is best accomplished with a calibrated deviation meter.

The *calibrate* command is provided to help with this step by providing a solid 1200 or 2200 Hz tone for the adjustment. It keys the transmitter and alters between the two tones each time the space bar is pressed. The opera-

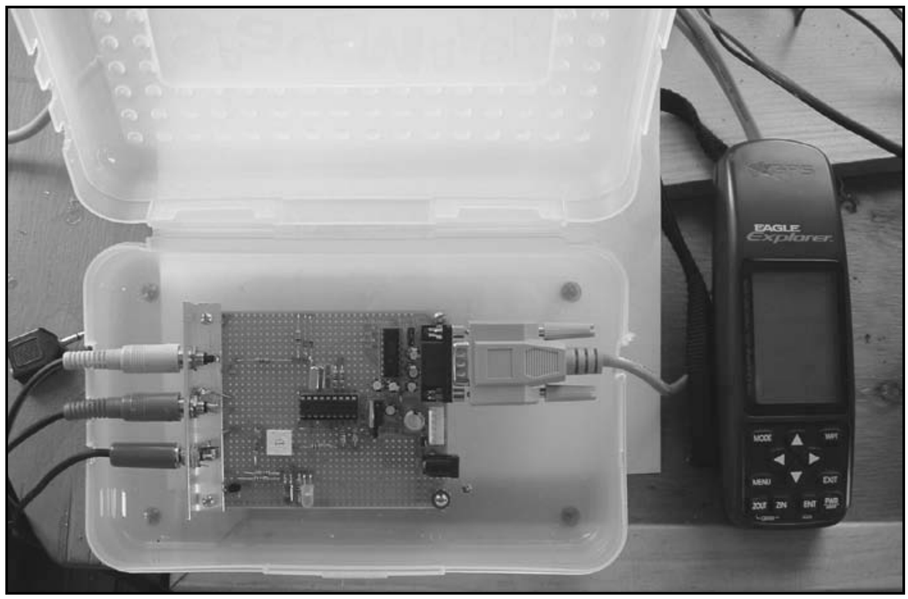

Figure 9—GPS unit connected to the TNC.

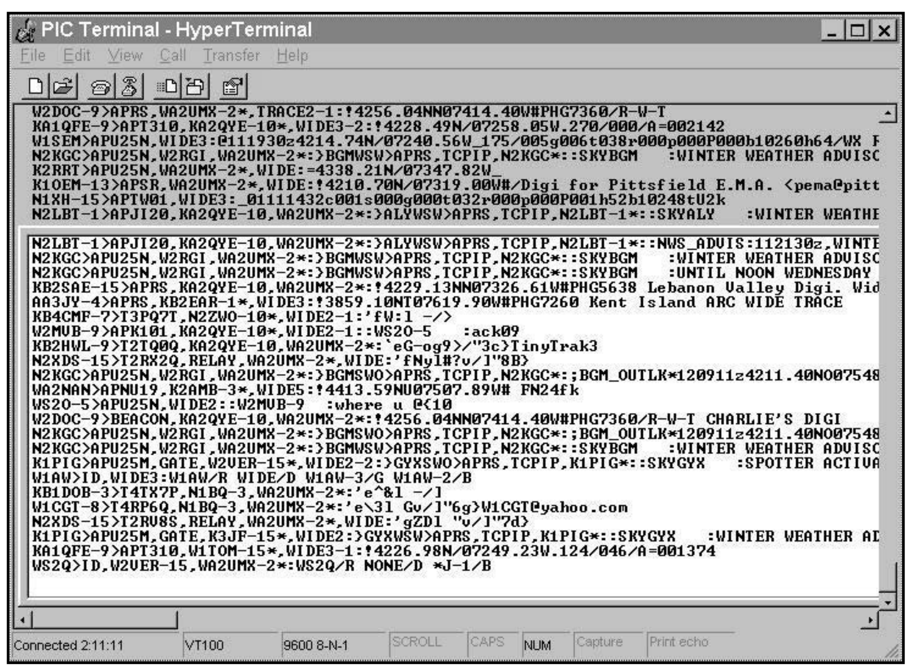

Figure 10—Packet output screen using *Hyperterminal* with monitor enabled.

tion is as follows:

cmd:CALIBRATE

Calibrate Mode. Space Bar to Toggle, Control C to Exit

<ctl-c>

OK

cmd:

You may also need to adjust the number of flags at the beginning of the packet to allow your transmitter to stabilize. This is done using the *txdelay n* command. Each increase in the number *n* adds one more flag (approximately 6 ms). The command syntax is:

cmd: TXDELAY 60

OK

cmd:

Your TNC should now be ready to go on the air. Tune your radio to a local packet frequency.

Beacon Operation

If you want your unit to send out beacons, you must enter some beacon text. This is done by entering:

cmd: btext *This is Bob in Jefferson, NH*

OK

cmd:

Then, the beacon interval can be set. Beacons can be set between 1 and 59 minutes. As an example, a beacon every half hour would be:

cmd: beacon every 30

OK

cmd:

The beacon may be turned off by entering:

cmd: beacon every 0

OK

cmd:

Monitor Operation

Three options are provided for the monitoring of packets. If you do not wish to use the serial interface of the unit (ie a dedicated digipeater), monitoring can be turned off. This can be done by the following command:

cmd: monitor off

OK

cmd:

Monitoring of all packets, independent of their call sign addresses would be selected by:

cmd: monitor all

OK

cmd:

Selective monitoring of packets addressed to *mycall* or *alias* can be accomplished by setting the monitor as follows:

cmd: monitor me

OK

cmd:

A screen capture of actual packet traffic reception as captured by *Hyperterminal* is shown in Figure 10. In this case, the *monitor* option was set to *all*.

If you don't leave a terminal connected to your TNC for monitoring (ie, a standalone remote digipeater application), it is a good idea to turn the monitor off when the EE ram configuration is complete. This will reduce power and real time consumption in the processor. The monitor feature is automatically shut off if the unit is in GPS receiver mode (Jumper J4 installed).

Alias Operation

Alias operation is optional and is used in the receive process. It allows your unit to respond to packets that might not be addressed directly to your call sign. The alias can be any 7-character string, with optional SSID. As an example, setting my alias to *relay* is done as follows:

cmd: myalias relay

OK

cmd:

Digipeating

Digipeating may be turned on and off by the following command:

cmd: digi on

OK

or

cmd: digi off

OK

If digipeating is on, any packet matching *mycall* or *myalias* will be *digipeated* on to the next station specified in the header.

Saving Your Station Parameters

After you have inputted all your parameters, you can display them all to make sure everything is correct. This is done with the *disp* command and should look as follows:

cmd: disp

TXDELAY 60

ECHO on

GPS $GPRMC

TRace off

MONitor on

DIGIpeater on

BEACON on EVERY 20

UNPROTO aprs v relay v wide v wide

MYCALL wb8wga-0

MYALIAS unit1

BTEXT Bob in Jefferson, NH

cmd:

Don't worry about the GPS or TRACE commands for now, they will be described in a later section.

If everything is correct, the parameters can be saved in EE memory with the permanent command as follows:

cmd: perm

OK

cmd:

Trace Function

For software testing or experimentation, a trace command is provided. It gives a hexadecimal display of a packet being formatted for transmit but the actual transmit is not completed. It also provides a hexadecimal display of received packets. Its syntax is:

cmd: TR xmit

OK

cmd:

or

cmd: TR rcv

OK

cmd:

Converse Mode

The TNC will allow the transmission of unnumbered information (UI) frames of anything that is typed from the serial port. This allows for a *chat* mode in which users can converse in a roundtable type operation. This is done by entering converse mode as follows:

cmd: converse

Entering Converse Mode, Hit Control C to Exit

Hello World

<ctl C>

cmd:

While in converse mode, other stations on the frequency that send frames will have their frames displayed as well, allowing for a roundtable type operation.

APRS Operation

Automatic Position Reporting System (APRS) is a fun and useful aspect of packet and Amateur Radio. The APRS system provides a good means to transmit position information, weather information, and messages. This TNC software supports some of the common APRS functions used by many hams today on the 2-meter band (typically 144.39 MHz) so it can get you on APRS very inexpensively. Several good books[8] are available from the league on APRS to get you going.

If you are using your TNC as part of a home QTH setup, you can get your newly built TNC registered in the APRS system, see your reports on the Internet, and at the same time provide a digipeater for local hams. To do this, you simply need to set the *unproto* and beacon text. Also *digi* must be turned on and *myalias* turned on if you wish to be a digipeater. As an example, I use the following command sequence to put one of my units as a digipeater from my home QTH:

cmd: myalias RELAY

OK

cmd:btext !4424.17N/07126.40W# /R Bob in Jefferson,NH

OK
cmd: unproto APRS v RELAY v WIDE v WIDE
OK
cmd:digi on
OK
cmd:

Note that you need to substitute your fixed position longitude and latitude (mine is 44.24.17 N and 71.126.40 W). The # indicator at the end indicates you serve as a digipeater.

Sending National Marine Electronics Association (NMEA) Sentences from a GPS

Another way to send APRS packets is by using your GPS unit to provide the position information. The TNC beacon feature is used to transmit location reports from a GPS unit using the serial port. The GPS unit is connected to the serial port instead of a terminal and NMEA-0183 sentences are transmitted each beacon interval instead of beacon text (btext). A picture of the TNC connected to an EAGLE TNC via the serial port is shown in Figure 9. Note to activate this feature. The GPS receiver enable function must be set **at initialization time**, telling the software to interpret input on the serial port as NMEA sentences rather than TNC commands. To return to regular mode, simply recycle the TNC power.

NMEA Setup for Your GPS

There are many types of ascii sentence structures available from GPS units. The TNC software supports $GPGGA, $GPGLL and $GPRMC sequences. These are the most common but check your GPS manual to ensure your data is compatible with one of the protocols. When the beacon timeout in the TNC has occurred, it will wait and send the next valid NMEA sentence to come from the GPS unit. Thus, your most recent location is transmitted. This sentence type must be set from the command line as follows:

cmd: GPS $GPGGA
OK
cmd:
or
cmd: GPS $GPRMC
OK
cmd:
or
cmd:GPS $GPGLL
OK
cmd:

Also remember that *mycall, unproto, and beacon* parameters must be set before the beacon can be transmitted. A common *unproto* for APRS is:

cmd: unproto APRS v RELAY v WIDE v WIDE
OK
cmd:

Check with other local APRS users to get the best route for your area.

The *beacon every n* should be set to a reasonable value to capture your position changes while not overburdening the APRS system with reports. I use 15 minute positioning reports on my system.

These options can be made permanent by using the *perm* command. After this is completed, the GPS jumper (J4) is inserted and power recycled. The unit will then start sending your position reports.

Note that the unit will continue to *digipeat packets* while in the GPS mode, assuming *digi* is on.

Future Steps

I hope this project will get you interested in packet radio operation and microprocessor usage. The hardware unit is just a starting point and is flexible enough that it can be programmed to do many useful things in your shack. Use of the Olimex experimenters' boards make it easy to get a fast start on a new project. The modemless receive and transmit techniques are good starting points for other digital modes. Use of the boot loader concept provides a lot more flexibility for changes to your ham projects. After you gain an understanding of AX25, what you put in the packets from your TNC is only limited by your imagination. Hope to hear your *blurrrp* on the air soon.

Bob Ball, WB8WGA, is a retired electrical engineer who lives in the White Mountains of northern New Hampshire. His working career spanned 30 years in the Columbus, Ohio area as a technical manager at Bell Laboratories. Bob holds a Bachelor of Science Degree in Electrical Engineering and a Master of Science Degree in Computer Science, both from the Ohio State University. He holds an Extra Class license and has been continuously licensed as an amateur operator for 40 years, first licensed at WA3ANV in 1964. Bob enjoys many facets of Amateur Radio but can most often be found on the bands running 40 and 20 meter CW.

Notes

[1]Mike Berg's work on modemless packet decoders for AX.25 can be viewed at his web site, **www.ringolake.com**.

[2]Many references on using a ladder network for generating sine waves are on the Web. MicroChip (**www.microchip.com**) has an application note and schematics, AN655, giving more information on the technique. Byon Garrabrandt, N6BG, uses this technique in his popular original Tiny Trak unit implemented on the PIC16F84. The schematic can be viewed at Byon's web site at **www.byonics.com/tinytrak/tinytrak.zip**.

[3]Many excellent resources are available at the TAPR Web site (**www.tapr.org**). Published source code contributions have been used in this project as well as information from the on-line TNC2 manual. Good papers exist on their site for understanding the AX.25 protocol. I used the AX.25 protocol specification and an excellent paper by John Hansen, W2FS, on how to implement the protocol. The command sequences are similar to the TAPR TNC1 and TNC2 Units.

[4]Hex and source files for this project are available at the League Web site (**www.arrl.org/qexfiles**).

[5]Photos for this article were provided with permission by Imre Szauter.

[6]Olimex boards with an installed PIC16F88 are available at Spark Fun Electronics (**www.sparkfun.com**) for $18.95 USD. Pictures of the board, software, and schematic were obtained (with permission) from the Web site.

[7]A boot loader for the PIC16F88 is available at no cost at **www.sparkfun.com**. It consists of 2 software parts, the small portion that is programmed on the chip (call Bloader) and the PC portion (call Screamer). For convenience, a copy of the Bloader and Screamer is also provided at the League Web site.

[8]APRS—"Moving Hams on Radio and the Internet," by Stan Horzepa, WA1LOU, provides a good overview of the history and working of APRS. □□

28 kbps to 9 Mbps UHF Modems for Amateur Radio Stations

Following on the heels of their HSMM article in the Nov/Dec 2004 issue, the authors present a protocol suite for UHF modem use.

By John Champa, K8OCL; and John B. Stephensen, KD6OZH

Preface

High Speed Multimedia (HSMM) radio within Amateur Radio currently consists primarily of the deployment by hams of inexpensive, commercially available off the shelf equipment used for radio-based local area networks (RLAN). This gear is typically one of the IEEE 802.11 standard's radios that can achieve speeds as high as 54 Mbps. The frequencies used are in the upper end of the 2400 MHz amateur band, sometimes on 902 MHz and on rare occasions the 5000 MHz band.

Although some linked HSMM radio nodes now cover an entire community, what is needed for the HSMM network to continue to rapidly grow are other methods of achieving greater range than 2.4 GHz propagation normally allows.

Early last year the ARRL HSMM Working Group decided to form a number of Radio Metropolitan Area Networks (RMAN) Project Teams to investigate such methods. Two of these teams have made substantial progress: The RMAN-VPN Team (using the Internet to connect HSMM nodes) and the RMAN-UHF Team (using lower Amateur bands such as the 440 MHz band for such linking).

What follows is a set of proposed protocols for the HSMM Orthogonal Frequency Division Multiplexing (OFDM) Modem that will allow Radio Amateurs to have all-mode voice, text, data, and video high-speed digital communications on the VHF, UHF and SHF bands. We hope to begin alpha testing of the OFDM modem this year in at least four locations—Racine, San Antonio, Tampa and Detroit. We plan to use an ATV channel in the 440 MHz band operating in a digital image mode we call Amateur Digital Video (ADV).

The Modem Physical Layer

Version: draft 5
Date: 2004-5-6
Author: John B. Stephensen, KD6OZH

1. Introduction

This document defines a set of physical layer protocols in the Open System Interconnection model for point-to-point and point-to-multi-point operation between stations operating in the Amateur Radio service. These stations may be fixed, land mobile or maritime mobile and use either directional or omnidirectional antennas in the UHF bands. This document describes the format and behavior of the protocol.

Six modems with different data rates are defined to fit the various regulatory requirements and band plans from 219 to 2450 MHz. The narrowest bandwidth was chosen to fit the current FCC regulations governing data transmission on the 219-220 and 420-450 MHz bands. The highest bandwidth was chosen to fit the largest channels allocated to data transmission in the ARRL band plan.

2. Physical Media

The physical medium interconnecting users is the electromagnetic spectrum. Only the UHF and higher-frequency amateur bands have sufficient space to allow high-speed digital links. This family of modems is designed for use in urban areas under line-of-sight (LOS) and non-line-of-sight (NLOS) conditions. Modems may use both directional and omnidirectional antennas.

Radio propagation in an urban area is characterized by strong multi-path propagation. Propagation measurements indicate that multi-path delays range from 1 to 20 µs for LOS and NLOS paths in an urban environment. The signaling rate on the radio link is limited, as the symbol period must be much longer than the maximum multi-path delay. Since we want to communicate at a data rate much higher than the symbol rate, multiple carriers must be used. The carrier spacing must be chosen to prevent mutual interference and ensure that the data on each carrier is orthogonal. The modems described here use 4.8-kBaud symbol rates with 6 kHz carrier spacing. This provides a guard band of 41.7 µs between adjacent symbols. Most intersymbol interference exists within the guard band and can be ignored.

2491 Itsell Rd
Howell, MI 48843-6458
k8ocl@arrl.net

3064 E Brown Ave
Fresno, CA 93703-1229
kd6ozh@verizon.net

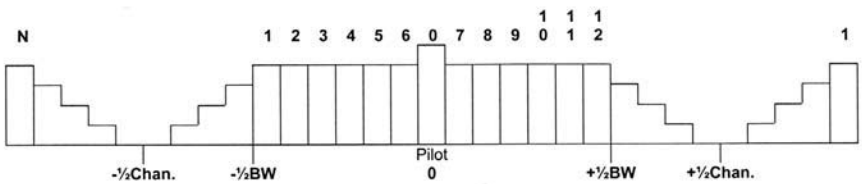

Fig 1—13-carrier transmission channel format (not to scale).

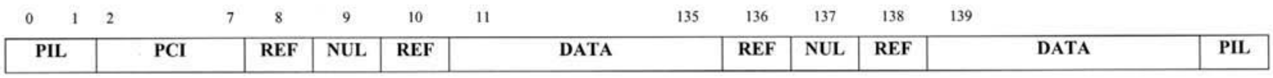

Fig 2—PHY PDU format (2 data blocks).

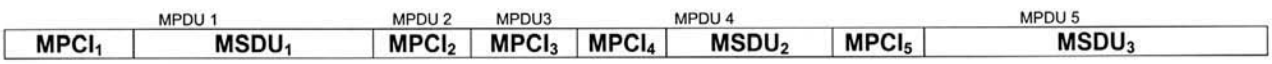

Fig 3—PHY-SDU with multiple MPDUs.

Table 1—Required minimum SNR for different modulation types (AWGN channel)

Symbol	*States per Symbol*						
Modulation	*2*	*4*	*8*	*16*	*32*	*64*	*256*
ASK	10	17	24	30	36	-	-
PSK	10	13	18	24	30	36	47
DPSK	11	15	21	27	33	39	51
QAM	-	13	-	20	-	26	32
DQAM	-	15	-	23	-	29	35

Table 2—Required frequency accuracy

Carriers	*96*	*288*
D8PSK	100.0 PPM	15.0 PPM
D256QAM	15.0 PPM	2.5 PPM

Table 3—Modem data rates for 4.8 kbaud and 6 kHz channel spacing

*Analog BPF**	*166 kHz @ ±1 dB*		*600 kHz @ ±1 dB*		*1.75 MHz @ -1 dB*	
FFT Sample Rate	96 ksps	192 ksps	384 ksps	768 ksps	1.536 Msps	3.072 Msps
No. Carriers	13	25	49	97	145	289
Signal Bandwidth	84 kHz	156 kHz	300 kHz	588 kHz	0.876 MHz	1.740 MHz
Channel Spacing	125 kHz	250 kHz	500 kHz	1000 kHz	1.500 MHz	3.000 MHz
DBPSK RC=1/2 Data Rate	*28.8 k*	*57.6 k*	*115.2 k*	*230.4 k*	*0.3456 M*	*0.6912 M*
DQPSK RC=2/3 Data Rate	*76.8 k*	*153.6 k*	*307.2 k*	*614.4 k*	*0.9216 M*	*1.8432 M*
D8PSK RC=2/3 Data Rate	*115.2 k*	*230.4 k*	*460.8 k*	*921.6 k*	*1.3824 M*	*2.7648 M*
D16QAM RC=5/6 Data Rate	192.0 k	384.0 k	768.0 k	1536.0 k	2.3040 M	4.6080 M
D64QAM RC=5/6 Data Rate	288.0 k	576.0 k	1152.0 k	2304.0 k	3.4560 M	6.9120 M
D256QAM RC=5/6 Data Rate	384.0 k	768.0 k	1536.0 k	3072.0 k	4.6080 M	9.2160 M

*Filter bandwidth is recommendation only.

3. Symbol Rate and Carrier Placement

A pilot carrier, used for timing information, is placed in the center of the carrier group. Half of the N data carriers are placed on each side of the pilot carrier and enumerated 1 through N from the lowest frequency to the highest frequency. Fig 1 shows 13 carriers with the main lobes of the carriers occupying the bandwidth, BW. The group delay must be flat over this bandwidth to minimize FFT sampling errors. Extending beyond that limit on either side are the minor lobes of the carriers. The channel spacing must be chosen so that the minor lobes of each channel's carriers are at an acceptably low level by the first carrier of the next channel.

Table 3 summarizes the various numbers of carriers and data rates for the modems. Carrier frequencies are accurate to ±100 PPM. The symbol rate shown includes a gap that is filled at the transmitter with a copy of the last ¼ of each tone. This provides a continuous waveform for the receiver FFT window even though there may be jitter. The receiver will normally sample the last part of the symbol cell to avoid inter-symbol interference that may exist in the first part of each symbol.

A number of options exist for modulating the data carriers including amplitude modulation (ASK), phase modulation (PSK) and a combination of the two (QAM). Each requires a different signal to noise ratio (SNR) to achieve a specific data rate. The SNR values summarized in Table 1 are those required for a 10^{-5} symbol error rate. Since each transmission consists of less than 11,520 symbols the block error rate can be expected to be less than 12% at these levels.

Table 2 shows the required frequency accuracy for different numbers of carriers and modulation types.

Fig 4—Data MPDU.

Table 4—PHY protocol control information (PCI) coding

PHY-PCI 012345	*Data Carrier Modulation*	*Coding Rate*	*Bits per 12 Carriers*
111111	DBPSK	1/2	6
010101	DQPSK	2/3	16
101010	D8PSK	2/3	24
111000	D16QAM	5/6	40
001110	D64QAM	5/6	60
100011	D256QAM	5/6	80

Table 5—Carrier amplitude as a function of quantity

Carriers	*1*	*13*	*25*	*49*	*97*	*145*	*289*
Amplitude (dBc):	0	-20	-23	-27	-30	-32	-36

Table 6—Puncture codes

Coding Rate	*Puncture Code 0*	*Puncture Code 1*	*Bits per 12 Carriers*
1/2	*1*	*1*	*6*
2/3	*10*	*11*	*8*
5/6	*10101*	*11010*	*10*

Six modulation types are defined for each of six modem bandwidths. Differential phase shift keying (DPSK) is used for the lowest three data rates to allow mobile operation in addition to fixed operation. As the station moves, the absolute phase varies as the strength and delay of multi-path rays vary so a fixed phase reference cannot be used. The highest three data rates are for fixed stations only. DQAM is used to compensate for phase rotation due to inaccuracies in carrier spacing.

All implementations of this modem must support a choice of DBPSK, DQPSK and D8PSK modulation. If QAM is supported, then a choice of DBPSK, DQPSK, D8PSK, D16QAM, D64QAM and D256QAM will be supported. The *italicized* data rates in Table 3 are those that must be supported for minimal compliance with this standard.

4. PHY-PDU Format

There are four special OFDM symbols used in the PHY-PDU. The null (NUL) symbol contains an unmodulated pilot carrier at the amplitude defined in Table 4 and no data carriers for one symbol period. The pilot (PIL) symbol is an unmodulated pilot carrier at maximum power (0 dBc) for one symbol period. The pilot carrier is used for frequency acquisition and Doppler shift correction.

The reference (REF) symbol allows the receiver to determine the starting phase and amplitude. The absolute phase of each carrier is set according to the formula:

$$\theta = 3.6315\, k^2$$

where k is the carrier index by frequency. The crest factor is less than 5 dB so the reference symbol shall be transmitted at 4 dB above the power levels in Table 4 to improve amplitude and phase estimation. The PCI symbols are ASK modulated REF symbols with a one at +4 dB above normal power levels and a zero at 2 dB below normal power levels using the patterns defined in Table 4. The use of REF symbols combats selective fading. The format of the PHY-PDU is shown in Fig 2.

The PHY-PDU begins with 2 PIL and 6 PCI symbols. The high amplitude single carrier allows the receiver to acquire carrier frequency lock more easily. The PCI symbols then specify the data carrier modulation as follows:

The PHY-PCI symbols have a minimum Hamming distance of 3 so 1 bit error correction is possible.

This is followed by the 3-symbol sequence REF-NUL-REF that is designed to allow the receiver to establish time synchronization under adverse conditions. Up to 125 symbols containing data may then be transmitted. If more data is to be transmitted, it is broken up into 125-symbol data blocks with the REF-NUL-REF sequence inserted in between blocks of data. This allows the transmitter and receiver clocks to resynchronize. Clock frequencies must be accurate to ±100 PPM. The last data block may be shorter than 125 symbols. The PHY-PDU ends with a PIL symbol.

Data is transmitted in the 12 to 288 outer carriers with PSK or QAM-modulated symbols. Five types of coding are used depending on the signal to noise ratio (SNR) on the link. 12, 24, 36, 48, 72 or 96 bits may be encoded in each 12-carrier group with rate 1/2, 2/3 or 5/6 forward error correction (FEC) coding. This results in data transfer rates of 1/2, 4/3, 2, 10/3, 5 or 20/3 times the number of data carriers times the symbol rate. If the number of data bits to be transmitted is less than the number of data carriers, zero bits will be added as padding. This will only be done on the last symbol in a frame.

User data bytes are serialized by placing the least significant bit into the bit stream first and the most significant bit into the bit stream last. FEC is provided by a rate ½ block convolutional code (BCC) with a constraint length of 7. The generator polynomials are $g_0 = 1011011_2$ and $g_1 = 1111001_2$. The code rate is then modified by ANDing with puncture code shown Table 6.

Encoded bits are mapped onto each OFDM symbol as follows. The first bit maps onto bit A of the lowest frequency carrier and the next onto bit A of the next higher frequency carrier until all carriers are covered. If higher-order modulation is being used, the next set of bits maps onto bit B of each carrier from lowest to highest frequency. If needed, the same mapping continues from lowest to highest frequency carrier for bits C, D, E, F, G and H. This mapping ensures that errors occurring on one carrier are spread out over the bit stream.

4.1 BPSK—One Bit per Carrier

Under very low SNR conditions, one bit is mapped on to each data carrier and all are transmitted in parallel in each symbol period as defined in Table 7. A zero bit is transmitted with normal carrier phase and a 1 bit is transmitted with inverted phase.

4.2 QPSK—Two Bits per Carrier

Under low SNR conditions, the data rate can be doubled with two data bits mapped onto each carrier as shown in Table 8. Dibit values are in Gray-code sequence so that a 90° phase error affects only one bit.

4.2 D8PSK—Three Bits per Carrier

Under moderate SNR conditions, the data rate can be tripled with three data bits mapped onto each carrier as shown in Table 9. Tribit values are in Gray-code sequence so that a 45° phase error affects only one bit.

4.4 16QAM—Four Bits per Carrier

When the SNR is higher, transmitting 4-bits per carrier quadruples the data rate. The bits are spread over the carriers in 4-bit groups using 16QAM modulation with a rectangular signal constellation as shown in Table 10. Each nibble is split into two with the least significant dibit modulating the in-phase carrier amplitude and the most significant dibit modulating the quadrature carrier amplitude.

4.5 64QAM—Six Bits per Carrier

When the SNR is very high, transmitting 6-bits per carrier results in a rate six times the base the data. The bits are spread over the carriers in 6-bit groups using 64QAM modulation with a rectangular signal constellation as shown in Table 11. Each hexbit is split into two tribits with the least significant tribit modulating the in-phase carrier amplitude and the most significant tribit modulating the quadrature carrier amplitude.

4.6 256QAM—Eight Bits per Carrier

When the SNR is extremely high, transmitting 8-bits per carrier results in a rate 8 times the base the data. The bits are spread over the carriers in 8-bit groups using 64QAM modulation with a rectangular signal constellation as shown in Table 12. Each byte is split into two nibbles with the least significant tribit modulating the in-phase carrier amplitude and the most significant tribit modulating the quadrature carrier amplitude.

5. PHY Service Interface

The physical layer service is defined to be compatible with IEEE 802.11-1999 section 12. The modem user accesses the physical layer service through a physical service access point PHY-SAP. This section describes the physical layer service offered to the user in terms of events, called service primitives, that cross the PHY-SAP.

5.1 Data Transmission

There are six service primitives associated with data transmission.

- The user issues **PHY-TXSTART. request** to start transmission of a data frame. The parameters are the number of data carriers and modulation type.
- The provider issues **PHY-TXSTART. confirm** when it is ready to receive user data bytes for transmission.
- The user issues **PHY-DATA.request** to transmit one byte of data. The parameter is one byte of user data and a maximum of 8,640 bytes may be transmitted in one PHY-SDU. This request is only valid between PHY-TXSTART.confirm and PHY-TXEND.request primitives.
- The provider issues a **PHY-DATA. confirm** when it is ready to receive another data byte.

Table 7—DBPSK encoding

Bit A	*Phase Shift*
0	0°
1	180°

Table 8—DQPSK encoding

Dibit B A	*Carrier Phase Shift*
0 0	0°
0 1	90°
1 1	180°
1 0	270°

Table 9—D8PSK encoding

Tribit C B A	*Carrier Phase Shift*
0 0 0	0°
0 0 1	45°
0 1 1	90°
0 1 0	135°
1 1 0	180°
1 1 1	225°
1 0 1	270°
1 0 0	315°

Table 10—16QAM encoding

Lower Dibit B A	*I Amp.*	*Upper Dibit D C*	*Q Amp.*
0 0	-0.70	0 0	-0.70
0 1	-0.23	0 1	-0.23
1 1	+0.23	1 1	+0.23
1 0	+0.70	1 0	+0.70

Table 11—64-QAM encoding

Lower Tribit C B A	*I Amp.*	*Upper Tribit F E D*	*Q Amp.*
000	-0.7	000	-0.7
001	-0.5	001	-0.5
011	-0.3	011	-0.3
010	-0.1	010	-0.1
110	+0.1	110	+0.1
111	+0.3	111	+0.3
101	+0.5	101	+0.5
100	+0.7	100	+0.7

Table 12—256-QAM encoding

Lower Nibble D C B A	*I Amp.*	*Upper Nibble G H F E*	*Q Amp.*
0000	-0.707	0000	-0.707
0001	-0.613	0001	-0.613
0011	-0.518	0011	-0.518
0010	-0.424	0010	-0.424
0110	-0.330	0110	-0.330
0111	-0.236	0111	-0.236
0101	-0.141	0101	-0.141
0100	-0.047	0100	-0.047
1100	+0.047	1100	+0.047
1101	+0.141	1101	+0.141
1111	+0.236	1111	+0.236
1110	+0.330	1110	+0.330
1010	+0.424	1010	+0.424
1011	+0.518	1011	+0.518
1001	+0.613	1001	+0.613
1000	+0.707	1000	+0.707

- The user issues **PHY-TXEND. request** to complete transmission of a frame.
- The provider issues **PHY-TXEND. confirm** when frame transmission is complete.

5.3 Data Reception

There are three service primitives associated with data reception.

- The provider issues **PHY-RXSTART. indication** to signal the start of a new data frame. The single parameter is RSSI which is the pilot carrier amplitude in dBnV.
- The provider issues **PHY-DATA. indication** to transfer one byte of user data. The user must accept data at the rate it appears at the PHY-SAP.
- The provider issues **PHY-RXEND. indication** to indicate the end of a data frame. The single parameter is RXERROR, which has one of the following values:

NoError. This value is used to indicate that no error occurred during PHY-SDU reception.

FormatViolation. This value is used to indicate that the format of the received PHY-SDU was in error. This condition is detected by a FEC error.

CarrierLost. This value is used to indicate that during the reception of the incoming PHY-SDU the carrier was lost and no further processing could be accomplished. This condition is detected by the absence of the final all zeros symbol.

UnsupportedRate. This value is used to indicate that a nonsupported number of carriers or modulation type was detected.

5.3 Clear Channel Assessment

There are three service primitives associated with clear channel assessment (CCA). These are used to hold off transmission when the channel is in use by another station.

- The user issues **PHY-CCARESET. request** when it wishes to reset the CCA logic.
- The provider issues **PHY-CCARESET.confirm** when the CCA logic is reset.
- The provider issues **PHY-CCA. indication** to indicate the presence or absence of RF energy in the currently selected channel. The single parameter is STATE, which has the value BUSY or IDLE.

6. PHY Protocol

This section describes the actions taken by this physical layer entity in response to stimulus from the physical media and physical service access point.

6.1 PHY-TXSTART.request

When a PHY-TXSTART.request is received, the receiver is disabled if this is a half-duplex link. The transmitter is then enabled and the center carrier is transmitted. The initial zero symbol is then transmitted and a PHY-TXSTART.confirm is issued to the user.

6.2 PHY-DATA.request

When a PHY-DATA.request is received, the data byte is saved and a PHY-DATA.confirm is returned to the user. Data bytes are accumulated until enough are present to transmit one symbol.

6.3 PHY-TXEND.request

When a PHY-TXEND.request is received, any required padding bytes are generated and the last symbol is transmitted. The pilot and data carriers are disabled. If this is a half-duplex link, the receiver is enabled. A PHY-TXEND.confirm is then issued.

6.4 Pilot Carrier Detect

If a pilot carrier is detected with an amplitude exceeding RSSI_CCA the receiver will attempt to synchronize to the carrier frequency and then to the symbol rate detected in the pilot carrier modulation. If carrier and symbol synchronization are achieved the receiver waits for the first symbol.

6.5 First Symbol

When data carriers appear, the first symbol is immediately decoded and interpreted as the value zero. A PHY-RXSTART.indication is issued with the received pilot carrier amplitude plus the maximum and current modulation type and number of carriers specified in the PCI and the receiver begins decoding data symbols using the rules specified in the PCI.

6.6 All Other Symbols

As the receiver decodes data symbols, PHY-DATA.indications are issued to the user with the value of the bytes encoded by the symbol. One symbol results in 1 to 240 PHY-DATA.indications.

6.7 Data Carrier Loss

When the data carriers disappear, a PHY-RXEND.indication (NoError) is passed to the user and the receiver waits for another symbol.

6.8 Pilot Carrier Loss

If the pilot carrier disappears during a period when there are no data carriers, a PHY-CCA.indication (IDLE) is passed to the user and the receiver waits for a pilot carrier with no data carriers. If the pilot carrier disappears during a period when data carriers are present, a PHY-RXEND.indication (CarrierLost) is issued and the receiver waits for a pilot carrier with no data carriers.

6.9 Noise Level Increase

If the receiver detects an increase in RF energy within the channel that is not the pilot carrier, but exceeds RSSI_CCA for more than one symbol period, it sets CCA_STATE to BUSY and issues a PHY-CCA.indication (BUSY).

6.10 Noise Level Decrease

If CCA_STATE is BUSY and receiver detects an decrease in RF energy to a level less than RSSI_CCA for a period exceeding 16 symbol periods, it sets CCA_STATE to IDLE and issues a PHY-CCA.indication (IDLE).

7. PHY Management Interface

The management interface provides a means for the user to configure the modem and to collect performance information. The service primitives are defined in this section. Management operations apply to all local PHY-SAPs.

7.1 Configuration Management

The service primitive **PHY-CONFIGURE.request** transfers configuration data to the modem. There are three parameters, RSSI_CCA, TM and TC. The values and semantics of TC and TM are defined in section 5 of this document. RSSI_CCA is the CCA logic threshold in dBnV. TC and TM will retain a value of zero if the specified option is not implemented.

7.2 Performance Management

There are two service primitives that are used to request modem status information.

- The user issues a **PHY-STATUS. request** to request the current value of counters held within the modem.
- The provider issues a **PHY-STATUS.confirm** to return the current value of all management objects. The management objects are 32-bit unsigned binary values and are not modified when read. The following values are returned:

CCA_STATE – clear channel assessment logic state (BUSY or IDLE).

RSSI_CCA – RF energy level above which channel is declared BUSY.

RSSI_IDLE – average amplitude of RF energy in channel during IDLE states.

RSSI_BUSY – amplitude of RF energy in channel when last PHY-CCA.indication (BUSY) was issued.

RSSI_DATA – signal level of pilot carrier when last PHY-RXSTART. indication issued in dBnV (dB above 1 nanovolt).

PHS_ERR – maximum difference in phase between expected and actual value for last symbol received ($2^{32}=2\pi$).

AMP_ERR - maximum difference in amplitude between expected and actual value for last symbol received ($2^{32}=\pm1$).

PCI_LAST – value of last received PCI field (bits 31-3 are zero).

The following objects are counters:

TIME – increments every symbol period.

CARRIER_DETECT – increments when a PHY-CCA.indication with a value of BUSY is issued.

PCI_ILLEGAL – increments when an illegal value is detected in PHY-PCI.

FEC_DETECT – increments when a transmission error is corrected by FEC.

FEC_ERROR – increments when a transmission error cannot be corrected.

FRAMES_RECEIVED – increments when a PHY-RXEND.indication is issued.

BYTES_RECEIVED – increments when a PHY-DATA.indication is issued.

FRAMES_TRANSMITTED – increments when a PHY-TXEND.confirm is issued.

BYTES_TRANSMITTED – increments when a PHY-DATA.confirm is issued.

All management object values (including counters) are reset to a value of zero when power is applied to the modem.

8. Recommended Operating Frequencies

Operating frequencies must be selected to fit with existing ARRL band plans and FCC regulations. There are no restrictions on occupied bandwidth for data transmission above 902 MHz. In the 219-220 and 420-450 MHz bands, the maximum occupied bandwidth for data transmission is 100 kHz. However, in the 420-450 MHz band, bandwidths of up to 6 MHz may be used for amateur television, including digital amateur television (DATV). The following frequencies are recommended for operation of the OFDM modems specified in this document.

Table 13—Recommended frequencies

Band	*Sub-Band*	*No. Chan.*	*Channel Spacing*
125 cm	219 - 220 MHz	8	125 kHz
70 cm	420 - 423 MHz	12*	250 kHz
	423 - 426 MHz	1**	3.00 MHz
33 cm	903 - 906 MHz	1	3.00 MHz
	915 - 918 MHz	1	3.00 MHz
23 cm	1,248 - 1,252 MHz	4	1.00 MHz
	1,288 - 1,294 MHz	2	3.00 MHz
	1,297 - 1,300 MHz	1	3.00 MHz
13 cm	2,300 - 2,303 MHz	1	3.00 MHz
	2,396 - 2,399 MHz	1	3.00 MHz

*13 carriers for data transmission or 25 carriers for DATV

**Point-to-point DATV or experimental license for data transmission

UHF Modem MAC Sublayer for Amateur Radio Stations

Version: draft 5
Date: 2004-5-6
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1. Introduction

This document defines a medium access control (MAC) sublayer entity in the Open System Interconnection model for point-to-point and point-to-multi-point operation between fixed and mobile stations operating in the amateur radio service. This document defines the protocol implemented by this MAC entity.

The Amateur Radio service has the following unique requirements for a MAC service and protocol:

The radio links can cover a wide area. Transmitter and receiver antennas mounted at a 90 foot height above average terrain (HAAT) can provide can provide communication over a 50 mile path. A station located on a mountaintop at 1500 feet HAAT has a radio horizon of 100 miles.

The ARS requires efficient multi-cast operation. A net can have over 100 stations participating that must all be able to receive transmissions from the net control station and from each other.

Radio communication is subject to fading that results in bursts of errors. Efficient operation requires that the error rate be minimized on each communication link.

The coverage area of Amateur Radio stations results in propagation times that approach one millisecond. Carrier-sense multiple access (CSMA) techniques are not sufficient to control medium access so some form of polling must be used. The radio channel characteristics dictate error correction but traditional ARQ techniques do not work for multicasting. Consequently, the MAC entity must incorporate forward error correction (FEC) to provide a reliable multicast service. Since all stations are not in range of each other, the net control (primary) station must have the capability of forwarding MAC protocol data units (MPDUs) to all other (secondary) stations to achieve full connectivity. FEC required long data units to work effectively but many applications will transmit short data blocks. The MAC protocol must support concatenation of short MPDUs into a longer physical service data unit (PHY-SDU).

2. MAC Service Interface

The MAC service defined in this document is designed for a network of stations that all operate on the same frequency and at the same symbol (baud) rate. The available data rates on each station may vary due to differing sets of capabilities, but all neighboring stations must have a common baud rate and compatible modulation.

The primary purpose of the MAC entity is to transfer blocks of user data called MAC service data units or MSDUs. A MSDU consists of 1 to 1,536 bytes of user data that is sent from a source address to a destination address. Addresses are six bytes in length and are formed from the ARS call signs of individual stations or the name of a multicast group. Three service primitives are used to transfer user data:

The user issues an MA-UNITDATA.request when it wishes to transmit a MAC service data unit (MSDU). The parameters are the destination address, source address, length and 1 to 1,536 bytes of user data.

The provider issues an MA-UNITDATA-STATUS.indication when the user data plus MAC protocol control information (PCI) is fully processed and the MAC entity is available for further data transmission. The single parameter is transmission status, which may have the value TRANSMITTED or LOCAL_ERROR.

The provider issues an MA-UNITDATA.indication when a complete MSDU has been received. The parameters are the destination address, source address, length and 1 to

1,536 bytes of user data.

3. MPDU Format

The MAC entity will concatenate multiple MAC protocol data units (MPDUs) for transmission in one PHY-SDU whenever possible. Each MPDU consists of MAC protocol control information (MPCI) and, optionally, a MAC service data unit (MSDU). Fig 3 shows an example with five MPDUs where three contain MSDUs. The maximum PHY-SDU length is 8,640 bytes.

3.1 Data MPDU

A Data MPDU transports a complete MSDU. It consists of 21 bytes of MPCI containing the address and type fields followed by a variable-length user-data field as shown in Fig 4. The MPCI fields are defined in Table 14. DA, SA and L are obtained from the MAC service user while IA is generated by the MAC entity. IA is the next destination address while DA is the ultimate destination address.

3.2 Token MPDU

A Token MPDU contains the address of the next secondary station to transmit as shown in Fig 5 and Table 15.

T3.3 RSSI MPDU

A RSSI MPDU reports the received signal strength indication (RSSI) for one or more transmitting stations at a particular receiving station as shown in Fig 6 and Table 16. The TA and RSSI fields are repeated N times. The C and M fields indicate the transmitter capabilities at the reporting station.

3.4 MAC Address Format for Amateur Radio Stations

A modem implementation conformant to this standard shall use the locally administered ANSI/IEEE 802 48-bit address format and addresses shall be formatted as shown in Fig 7. Source addresses must be individual addresses.

The X bit shall be 0 for individual addresses and 1 for group addresses. Each address shall consist of exactly seven characters whose values are in fields C0 through C7. Each shall be encoded in 6-bit ASCII as shown in Table 17. Allowable characters are the Latin letters A through Z (case insensitive), the decimal numbers 0 through 9 and the space character.

Multicast addresses are group addresses that start with a letter or number and must consist of seven characters.

Individual addresses shall start with a letter or number and fields C0 through C5 shall be the amateur radio service (ARS) call sign assigned to the control operator. Call signs shorter than 6 characters shall be padded at the right end with spaces. The last character is an extension field. If only one modem is under control of the licensed operator or trustee, C6 shall be a space character. If more than one modem is under the control of the licensed operator or trustee, C6 shall be a non-space character.

4. Block Error Correction Code

The radio communications channel is subject to fading and impulse noise that may introduce errors in bursts. The error correction provided in the physical layer may be overwhelmed and bytes containing errors may be delivered to the MAC sublayer. A MAC-level block error correction code generates additional error correction information and distributes it over many symbols to allow correction of burst errors and increase the number of MAC-PDUs delivered to the user. This section describes the block error correction code.

A Reed-Solomon code with a symbol width of one byte, a block length of 255 bytes, a maximum data field

Table 14—Data MPDU fields

Field	*Bytes*	*Semantics*
IA	6	Intermediate MAC Address.
DA	6	Destination MAC Address.
SA	6	Source MAC Address.
L	2	802.3 Length field.
MSDU	1-1536	User data.

Table 15—Token MPDU fields

Field	*Bytes*	*Semantics*
PA	6	Primary station MAC address.
N	1	0 if token sent to PA, 1 if token sent to SA.
SA	6	Secondary station MAC Address if N = 1.

Table 16— RSSI MPDU fields

Field	*Bytes*	*Semantics*
RA	6	Reporting station MAC address.
C	1	Maximum number of carriers divided by 12.
M	1	Maximum number of bits per 12 carrier group (6-80).
N	1	Number of RSSI reports (0-255).
TA_N	6	Transmitter MAC address.
SNR_N	1	SNR of TA_N pilot carrier at RA in dB.

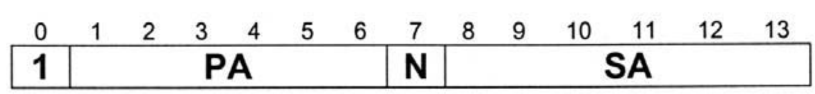

Fig 5—Token MPDU.

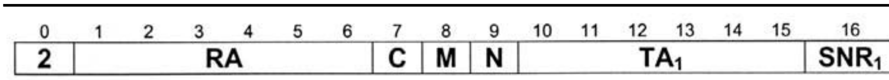

Fig 6—RSSI MPDU with one signal report.

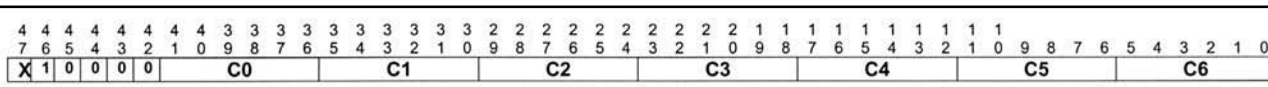

Fig 7—Amateur Radio subnetwork address format.

width of 239 bytes and a Galois field polynomial of 100011101_2 is used. This code will correct errors in as many as 8 symbols per block with an overhead of only 3.1%. When 239 data bytes are available for transmission, an encoded block of 255 bytes is generated. If the end of the PHY-SDU is reached and the number of data bytes to be transmitted is less than 239, a shortened code block is generated.

5. MPDU Forwarding

When the MAC entity receives a Data MPDU it must decide whether to deliver it locally, forward it to an adjacent station or discard it. The decision is made using the information maintained in the Neighbor Table in each station. The forwarding procedure depends on the type of destination address (individual or multicast) and the type of station (primary or secondary) involved.

Normal Forwarding: If the destination address is not a multicast address, the entries in the RSSI Table for all destination addresses are examined. If the destination addresses is the local station address, the user data is delivered to the local user at the MAC SAP. If the destination address is in the Neighbor Table, the MPDU is forwarded by setting the IA field to the destination address and transmitting the modified MPDU when this station has the permission to transmit a token. If the destination address is not in the Neighbor Table, the MPDU is forwarded by setting the IA field to the primary station address and transmitting the modified MPDU when this station has the permission to transmit a token. If this is the primary station, the MPDU is discarded.

Multicast Forwarding: If the destination address is a multicast address and this is a secondary station, the MPDU is forwarded by setting the IA field to the primary station address and transmitting the MPDU when this station has the permission to transmit a token. If the destination address is a multicast address and this is the primary station, the MPDU is forwarded by setting the IA field to the destination address and transmitting the modified MPDU when this station has the permission to transmit a token.

Each MAC entity makes decisions on the modulation method and number of carriers to use when transmitting to adjacent stations. It uses SNR data from the Neighbor Table entries for the adjacent stations that are to receive the PHY-SDU to select the modulation type and number of carriers to be used. The lowest SNR value is used to make the decision. The modulation type and number of carriers are selected based on the required SNR defined the physical layer standard. The MAC entity then checks the maximum allowed by the adjacent stations that are to receive this PHY-SDU. If the number of carriers selected is not supported by any one of these adjacent stations, the number of carriers is set to the maximum supported by all these stations and the modulation type is re-evaluated. The modulation type is then examined for each of these adjacent stations and if it is not supported, the number of bits per carrier is set to the minimum supported by all of these stations. The PHY-SDU is transmitted when the station receives permission in a Token MPDU.

6. Media Access Control

Access control is achieved via a token-passing mechanism. One station is the primary station that periodically transmits a token MPDU to selected secondary stations. The token MPDU is usually contained in a PHY-SDU that includes data transmissions to secondary stations. This token confers the right to transmit to the specified secondary station as shown in Fig 8.

The primary station transmits the token periodically to poll the secondary stations whose addresses are stored in its RSSI Table. After transmitting the token, the primary station monitors the channel status via PHY-CCA indications. If the medium is idle for more than two maximum length PHY-PDU times, the token is assumed to be lost and the primary station polls the next secondary station in the Neighbor Table.

Alternate primary stations (which may otherwise be secondary stations) may be configured. Each alternate primary station resets the ALT_PRI timer when it is passed the token. If the ALT_PRI timer ever expires, that station becomes the primary. The timer values for the first, second and third primaries are 0, 220 (1,048,576), and 221 (2,097,152) symbol periods.

7. RSSI Distribution

The Neighbor Table in the MAC MIB must be constantly refreshed with information on the communications paths between stations. The primary station polls all secondary stations by periodically transmitting an RSSI MPDU that contains the secondary station's pilot carrier amplitude at the primary station. Each secondary station broadcasts the received pilot carrier SNR by taking its idle time RSSI and the pilot carrier RSSI and calculating the SNR of the pilot carriers of all adjacent (i.e. heard) stations. The SNR information is then broadcast in an RSSI MPDU. Secondary stations that do not respond to this poll 4 times in a row are deleted

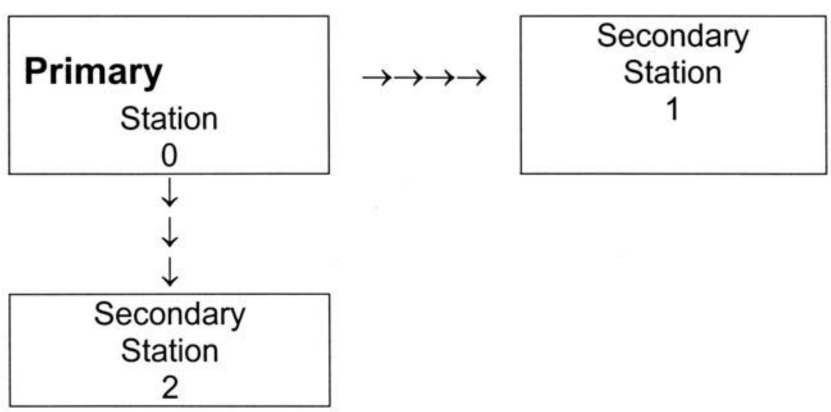

Fig 8—Token passing.

MAC Address	SNR	C	M
3ABC78	15	6	3
1465A7	23	12	6
A7779B	11	3	3

Fig 9—Neighbor Table.

from the primary station's Neighbor Table. The polling interval is every 216 (65,536) symbol periods. Fig 9 is an example of a Neighbor Table.

The Neighbor Table also contains the transmitting capabilities of each neighboring station. C is an unsigned number from 0 to 255 giving the maximum number of carriers that the station can transmit divided by 4. M is an unsigned number from 0 to 15 giving the maximum number of bits per carrier that the station can transmit.

8. MAC Protocol

This section describes the actions taken by this MAC sublayer entity in response to stimulus from the underlying physical service access point (PHY-SAP) and the user connected to the MAC service access point (MSAP).

8.1 PHY-CCA.indication

When a PHY-CCA.indication with a value of BUSY is received transmission of MPDUs is inhibited. When a PHY-CCA.indication with a value of IDLE is received, the background noise level is stored.

8.2 PHY-RXSTART.indication

When a PHY-RXSTART.indication is received, the MAC receiver entity prepares to receive and decode bytes and stores the pilot carrier RSSI.

8.3 PHY-DATA.indication

When a PHY-DATA.indication is received the data byte is transferred to the FEC decoder logic. The output is monitored for incoming MPDUs. If there are uncorrectable errors in an MPDU, the received MPDU is discarded. Otherwise, Data MPDUs are forwarded according to the process described in section 5. Each correctly decoded MPDU with the address of the local station results in a MAC-DATA.indication. RSSI MPDUs are used to update the Neighbor Table and, if received at a secondary station, cause an RSSI MPDU to be queued for transmission to the primary station.

8.4 PHY-RXEND.indication

When a PHY-RXEND.indication is received, the MAC entity discards any partial MPDU and checks for a Token MPDU. The primary station address is set by the Token MPDU. If the token specifies the local station as the secondary station with permission to transmit, CCA is checked and transmission of any accumulated MPDUs begins. A token MPDU is always sent to the primary station as part of the PHY-SDU.

8.5 MAC-DATA.request

When a MAC-UNITDATA.request is received, the MPDU is formatted and stored. The total PHY-SDU size, in bytes, is then calculated and this information is stored until the station receives permission to transmit.

8.6 Receipt of Token MMPDU

If the token MMPDU contains the local station address the MAC transmitter entity checks the channel status, waits for a value of IDLE, and issues a PHY-TXSTART.request with the appropriate carrier and modulation parameters.

8.7 PHY-TXSTART.confirm

When a PHY-TXSTART.confirm is received, the MAC entity issues the first PHY-DATA.request.

8.8 PHY-DATA.confirm

When a PHY-DATA.confirm is received, the MAC entity checks for more bytes to transmit. If so, the MAC entity issues a PHY-DATA.request. A MAC-UNITDATA-STATUS.indication is issued as transmission of each Data MMPDU completes. When all are transmitted a PHY-TXEND.request is issued.

8.9 PHY-TXEND.confirm

When a PHY-TXEND.confirm is received the receiver is enabled.

USB Interface for UHF RF Modem

Version: draft 2
Date: 2004-5-6
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1. Introduction

This document defines the user interface for an RF modem using the UHF bands in the Amateur Radio service. The interface is provided via the Universal Serial Bus (USB).

2. MAC Service and Interface

To provide compatibility with the existing LLC sublayer and network layer implementations, the service shall be as defined in ANSI/IEEE 802.2-1998 section 2.3. The user interface shall be as defined for the data class interface in the Ethernet Networking Control Model in "Universal Serial Bus Class Definitions for Communication Devices", version 1.1, 1999-1-19. This section provides a summary of the information in these documents.

The primary purpose of the modem is to transfer blocks of user data called MAC service data units or MSDUs. A MSDU consists of 1 to 1,536 bytes of user data that is sent from a source address to a destination address. Addresses are six bytes in length and their values may be a multicast address (a group of stations) or an individual address (exactly one station). Three service primitives are used to transfer user data:

The user issues a MAC-UNITDATA.request when it wishes to transmit data. The parameters are source address, destination address, user data, priority and service class.

The provider issues a MAC-UNITDATA.indication when a complete SNSDU has been received. The parameters are the source address, destination address, user data, reception status, priority and service class.

These service primitives are mapped into MAC interface data units (MIDUs) that are carried by one or more USB bulk data transfers. An OUT transfer is a request and an IN transfer is an indication. The service parameters are formatted as defined in ANSI/IEEE 802.3-2002 for the destination address, source address and length fields of a MAC protocol data unit (MPDU). A complete MIDU consists of zero or more bulk data transfers of maximal length followed by one bulk data transfer with a length less than the maximum. The bulk transfer with a length less than maximum (including zero) is the end delimiter for the MIDU.

3. Management Service and Interface

The layer and system management interface shall be as defined for the communication class interface in the Ethernet Networking Control Model in "Universal Serial Bus Class Definitions for Communication Devices",

Table 17—ARS 802-style address character set and encoding.

	Bits 5-4			
Bits 3-0	*00*	*01*	*10*	*11*
0000	(space)	0		P
0001		1	A	Q
0010		2	B	R
0011		3	C	S
0100		4	D	T
0101		5	E	U
0110		6	F	V
0111		7	G	W
1000		8	H	X
1001		9	I	Y
1010			J	Z
1011			K	
1100			L	
1101			M	
1110			N	
1111			O	

Destination Address	Source Address	Length	User Data	User Data	User Data

Fig 10—MIDU format.

Table 18—Ethernet statistics

Offset	*Field Name*	*Description*
D0	XMIT_OK	MPDUs transmitted.
D1	RCV_OK	MPDUs received.
D2	XMIT_ERROR	MPDUs not transmitted.
D3	RCV_ERROR	Total received MPDUs discarded.
D4	RCV_NO_BUFFER	MPDUs discarded due to buffer overflow.
D5	DIRECTED_BYTES_XMIT	MSDU data bytes transmitted to individual address.
D6	DIRECTED_FRAMES_XMIT	MSDUs transmitted to individual address.
D7	MULTICAST_BYTES_XMIT	MSDU data bytes transmitted to multicast address.
D8	MULTICAST_FRAMES_XMIT	MSDUs transmitted to multicast address.
D9	BROADCAST_BYTES_XMIT	MSDU data bytes transmitted to broadcast address.
D10	BROADCAST_FRAMES_XMIT	MSDUs transmitted to broadcast address.
D11	DIRECTED_BYTES_RCV	MSDU data bytes received from individual address.
D12	DIRECTED_FRAMES_RCV	MSDUs received from individual address.
D13	MULTICAST_BYTES_RCV	MSDU data bytes received from multicast address.
D14	MULTICAST_FRAMES_RCV	MSDUs received from multicast address.
D15	BROADCAST_BYTES_RCV	MSDU data bytes received from broadcast address.
D16	BROADCAST_FRAMES_RCV	MSDUs received from broadcast address.
D17	RCV_CRC_ERROR	MPDUs received with FEC error.
D18	TRANSMIT_QUEUE_LENGTH	Number of MPDUs waiting for transmission.
D19	RCV_ERROR_ALIGNMENT	Partial MPDUs received at end of PHY-SDU.
D20	XMIT_ONE_COLLISION	Token MPDUs received (secondary) or transmitted (primary).
D21	XMIT_MORE_COLLISIONS	0
D22	XMIT_DEFERRED	Number of times transmission delayed by CCA busy.
D23	XMIT_MAX_COLLISIONS	0
D24	RCV_OVERRUN	Received MSDUs discarded due to lack of buffer.
D25	XMIT_UNDERRUN	0
D26	XMIT_HEARTBEAT_FAILURE	0
D27	XMIT_TIMES_CRS_LOST	0
D28	XMIT_LATE_COLLISIONS	0
D29	undefined	Number of Neighbor Table entries.
D30	undefined	RSSI MPDUs transmitted.
D31	undefined	RSSI MPDUs received.

version 1.1, 1999-1-19. The following functions shall be implemented:

- SetEthernetMulticastFilters – a minimum of sixteen 48-bit addresses of any format shall be supported.
- SetEthernetPacketFilter–PACKET_TYPE_ALL_MULTICAST and PACKET_TYPE_ PROMISCUOUS required.
- GetEthernetStatistic – all selector codes shall be supported. Some counters may be mapped to non-802.3 information.
- The SetEthernetMulticastFilters request shall be used to set the RF modem frequency and register the 48-bit individual address and any group addresses used to identify the station.

The first entry shall be the modem frequency in kHz represented as a binary coded decimal integer. There is no hardwired MAC address and the modem will not transmit until an individual MAC address is configured.

The primary station shall be configured by including an entry with a value of $FFFFFFFFFFFF_{16}$. The alternate primary station is $FFFFFFFFFFFF_{16}$ and $FFFFFFFFFFFF_{16}$.

The statistics shown in Table 18 must be supported.

For More Information

If you have any questions about the OFDM modem or other topics regarding high-speed digital or multimedia operation, the HSMM Working Group can help you get started in this exciting part of amateur radio. You can subscribe to the ARRL IEEE 802.11b Mail List at Texas A & M University. To subscribe to this public list, go to the URL: **listserv.tamu.edu/archives/arrl-80211b.html** and select *Join or leave the list,* or see **www.arrl.org/hsmm/**.

John Champa, K8OCL, is an Extra Class Radio Amateur licensee. After education at Ohio State University, he received a commission in the US Army Military Police Corps. Most of John's civilian working career has been in the fields of safety and telecommunication engineering. He has filed four patent applications and is the inventor of the Digital Video Switch for Videoconferencing. He is also author of the text book "Videoconferencing Skills", a contributor to the 82nd Edition of The ARRL Handbook, and has had technical articles in QST, QEX *and* CQ VHF *Magazines. He is certified as a Wireless Telecommunications Engineer (Master Level) by the National Association of Radio and Telecommunications Engineers (NARTE). John is currently the Chairman of the ARRL High Speed Multimedia Networks Working Group.*

John Stephensen, KD6OZH, became interested in radio at age 11, when his grandfather bought him a crystal radio kit. During the 1960s, he built several HF receivers using vacuum tubes and other parts procured from discarded black-and-white TV sets. After attending the University of California, he and two friends founded PolyMorphic Systems, a supplier of personal computer kits, and later manufactured computers, in 1975. In 1985, he was cofounder of Retix, a networking hardware and software supplier. John received his Amateur Radio license in 1991 and has been active on bands from 7 MHz to 24 GHz, with interests including HF and microwave DXing and contesting. He has also been active on packet, satellites and on the HF bands using several digital modes. John has always designed and built his own Amateur Radio gear, some of which has been described in QEX *articles in recent years. His latest projects have centered on high-speed digital communication using DSP.* □□

High-Efficiency Antennas for Hand-Held Radios

End-fed half-wave antennas provide better efficiency for hand-held radios than quarter-wave radiators or "rubber ducks."

By Richard Kiefer, KØDK

I started evaluating and building antennas for hand-held radios because I needed an efficient, lightweight antenna for backpack and mountaineering activities in remote locations of the western US. My typical backcountry outing involves carrying a variety of equipment, including a hand-held radio and antenna 10-20 miles into mountainous terrain. The campsite destinations and operating locations are typically 30-100 miles from the nearest VHF or UHF repeater. Under these conditions, reliable communications over long distances at low power with the most lightweight equipment is the goal. In developing the optimum hand-held radio antenna for such excursions I have considered, modeled and constructed many types such as rubber ducks, Yagis, vertical coaxial collinear arrays, dipoles and end-fed half-waves. After all the experimentation, I have concluded that the best solution, when you consider portability, ease of use and efficiency, is an end-fed half-wave antenna mounted in the normal manner on top of the radio. This style of antenna is lightweight, easily stored in a pack and self-supporting, which is important above the timberline, where there are no trees or bushes. The total weight of my Icom IC-Q7A with two AA lithium batteries and the $^1/_2$ wave antenna is 6.6 oz.

A half-wave antenna, whether end or center fed, can provide high transmission and reception efficiency and is much better than a shortened rubber duck. For convenience, the half-wave antenna is best fed from the end with a matching network, rather than in the center. The radiating element can be either a fixed length of stainless steel wire or a telescoping element. The homebrew antennas described here all use SMA connectors, which are common on most contemporary hand-held radios. As far as I know, no commercially available half-wave antennas have SMA connectors. Because of the high-Q coil used in the matching network, the efficiency of these homebrew antennas approximates that of a center-fed vertically polarized dipole. In this article, I describe the construction and performance of two homebrew end-fed half-wave antennas for 2 meters and 70 cm. I also compare the performance of these antennas with some commercial half-wave, quarter-wave and "rubber duck" antennas.

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Fig 1—Author with hand-held and homebrew 2-m half-wave antenna.

Fig 2—A hand-held with a KØDK 2-meter half-wave telescoping antenna.

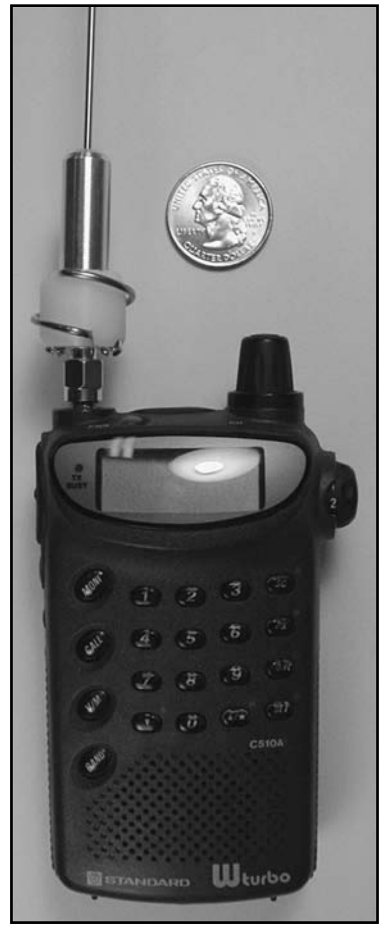

Fig 3—A hand-held with a KØDK 70 cm half-wave whip antenna.

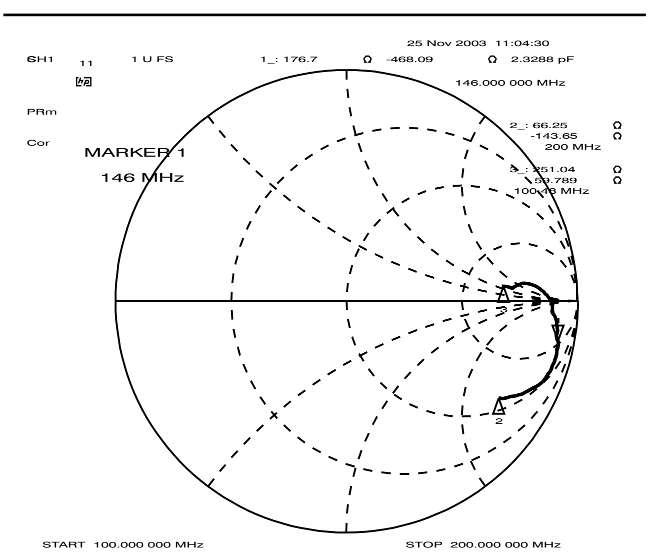

Fig 4—Impedance at the end of an unmatched 2-meter half-wave antenna element over the frequency range from 100 to 200 MHz.

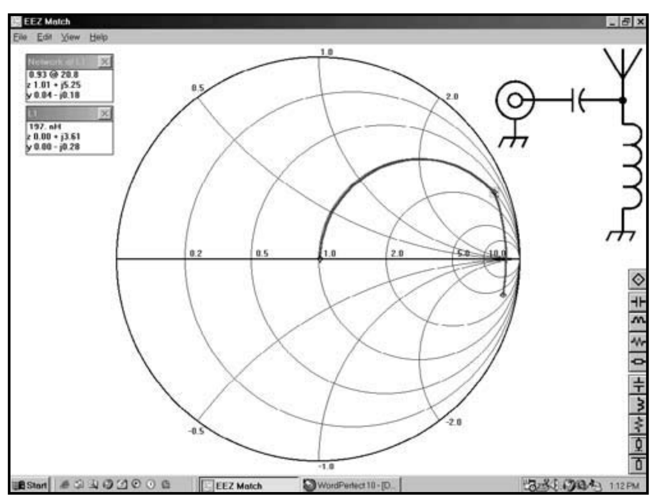

Fig 5—A Smith Chart plot of the impedance match necessary for the 2-meter half-wave antenna element. The 177–j468 Ω of the unmatched element is matched to 50+j0 with a parallel inductor of 197 nH and a series capacitor of 4.2 pF. The match is shown from the antenna end of the circuit.

Homebrew Half-Wave Antennas for Hand-Held Radios

The half-wave antennas described here use SMA connectors and a matching network as shown in Figs 2 and 3. One version of the antenna is designed for the 2-m band and the other for the 70-cm band. They both use the same construction technique, materials and matching network circuit topology. One particular advantage to a half-wave antenna with the matching network at its base is that the current maximum is high above your head in the middle of the radiating element. By keeping the current away from your head, the absorptive radiation losses and detuning are reduced. Both are important factors for hand-held operation. In addition, the body does not act as part of the antenna system (with current flowing through the capacity to your hand) as with a quarter-wave antenna or rubber duck. With a high-powered hand-held radio, it may also be safer to have the current maximum high above your head rather than near your eyes and brain.

Impedance Match for the end-fed half-wave antennas

If you measure the drive-point impedance at the end of a half-wave radiator mounted on a hand-held (see Fig 4), it is high, but finite, requiring a step-up network to match 50 Ω. Both antennas described here are fed at the end with a matching network consisting of a series capacitor and a parallel inductor. This topology, shown in Fig 5, works for any end-fed half-wave antenna independent of frequency. This is true because the impedance seen at the end of the half-wave radiator is always somewhere on the righthand side of the Smith Chart

For example, the measured drive-point impedance of a 36-inch-long stainless-steel wire at 146 MHz is 177-j468 Ω as shown by the Smith Chart. The impedances at 100 and 200 MHz are also given. This impedance is about the same whether the radiator is a stainless-steel wire or a telescoping element of approximately the same length.

The impedance-matching network transforms the 177–j468 Ω at the end of the radiating element to match the output of the hand-held radio's power amplifier. Thus, the hand-held is able to efficiently drive the antenna, to produce the same radiation pattern and efficiency as a center-fed dipole. A little power is lost in the matching network itself, heating the coil. This is because the Q of the coil is much lower than the Q of the Teflon-insulated capacitor.

The impedance match is accomplished as shown in Fig 5 by first rotating counterclockwise around the Smith Chart.[1] With a parallel inductor to ground of 197 nH to land on the constant-R circle that passes through the center.[2] This is the 50 Ω constant-R circle. Then the remaining inductive reactance is canceled with a series capacitor of 4.2 pF to "zero in" on the center of the Smith Chart at 50+j0. It also happens to be very convenient that the shunt component nearest the antenna element is a parallel inductor because it acts as a dc short to ground for ESD protection of the hand-held antenna connection. This same matching-circuit topology is used for both the 2-meter and 70-cm versions of the half-wave antennas.

Accurate Hand-Held Radio Impedance Measurements

To achieve proper impedance matching of hand-held antennas you must be able to accurately measure the drive point Z under a realistic condition. This condition must include the electrical effects of a person holding the radio and its antenna at a normal height above ground of about 5 feet. The drive-point Z is made up of the radiation and loss impedances. They are determined, in part, by the dimensions of the hand-held radio itself, its proximity to the

[1]Notes appear on page 45.

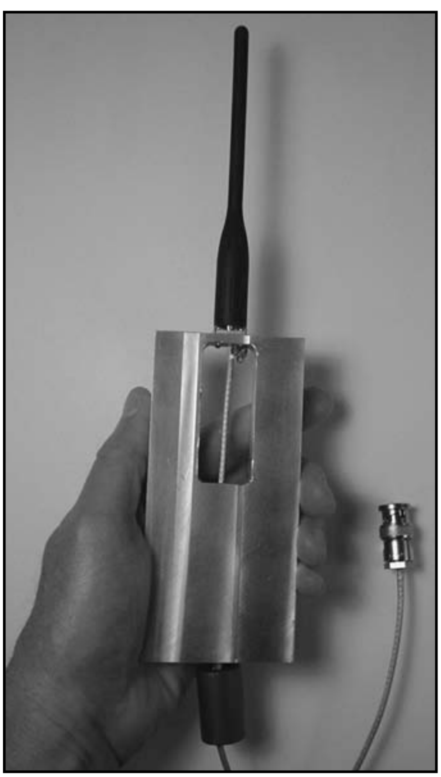

Fig 6—A dual-band rubber duck mounted to the hand-held simulator for measurements.

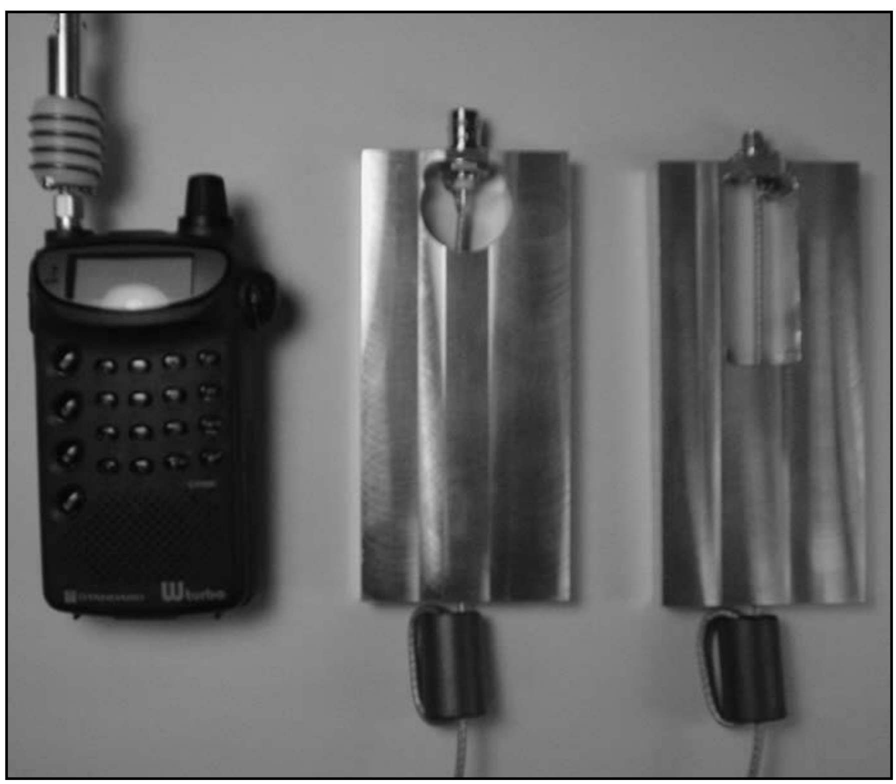

Fig 7—hand-held simulators used to measure the impedances and SWRs of hand-held antennas with either a BNC or SMA connectors.

human body and the electrical connection produced by the capacitance and conductance of the hand that holds the radio. All of these factors, along with the length of the antenna element, determine the impedance measured at the base of the antenna. This drive-point impedance must be as close to the radio-manufacturer's specification as possible. The best drive-point impedance is usually, but not always, $50+j0\ \Omega$. For example, I know of at least one power amplifier chip, the RF2172 from RF Micro Devices, that requires a load of $20-j1.6$ to achieve its maximum power output. In such cases, the optimum load impedance can only be determined by power-amplifier load-pull measurements, a subject beyond this discussion.

So to make accurate and representative hand-held antenna-impedance measurements, I use two hand-held simulators made from 0.5×5×2.5-inch aluminum blocks shown in Figs 6 and 7. These blocks approximate the size of a modern hand-held radio and provide the proper coupling to the human body when held in the hand. Each block has a coax connector mounted on one end with a 3 foot length of RG-318 Teflon coax attached. The coax is decoupled from the hand-held simulator by a large #43 material ferrite bead. This makes the impedance measurements and radiation pattern largely independent of the position of the coax. One hand-held simulator has a SMA connector mounted and the other a BNC. With one simulator or the other, I can measure a wide variety of commercial and homebrew antennas over a wide frequency range.

The instrument used to measure antenna impedance is a Hewlett-Packard HP-8753D network analyzer, which can determine the complex impedance of any antenna right at the connector on the simulator. The measurement reference plane is moved to the base of the antenna by calibrating the analyzer for S11 at the connector by the standard method using a 50-Ω reference load, an open and a short. Then all impedance measurements are made while the antenna and simulator are held in the hand as shown in the Fig 6.

For example, the impedance measurement for the homebrew 2-meter telescoping end-fed half-wave antenna is shown in Fig 8. The impedance is $59.5+j0.6\ \Omega$ for a SWR of 1.2:1 at 146 MHz. That's not too bad for having started at $177-j468\ \Omega$. At the band edges, the antenna impedance is $54.6-j18$ at 144 MHz and $64.3+j9.7$ at 148 MHz. The SWR rises to 1.4 at 144 MHz and 1.35 at 148 MHz. The SWR does not rise to 2:1 until ±5 MHz of the center frequency.

Antenna Construction Details

The mechanical construction of the 2-meter and 70-cm versions of the half-wave antennas is essentially the same. The electrical circuit topology is also the same. Only a few of the dimensions, the number of coil turns and the radiating element-lengths, are different. In addition, the 2-meter version may use a radiating element that is made of either a length of stainless steel spring wire or a telescoping whip. The mounting details are slightly different for each radiator. Each antenna then consists of the main parts in Table 1, with dimensions given for the 2-meter version

The radiating element is approximately ½-wave in length, on either 2 meters or 70 cm but the length need not be exact. The impedance at the end of the element will be somewhere to the far right of the Smith Chart if you are fairly close to a half-wave length. Also, the upper end of the telescoping element or the spring wire element should be capped with a bumper so it doesn't cause an injury. The telescoping antenna element is bolted into a hole in the mounting stud as shown in Figs 9 and 10. Then this entire antenna element assembly screws into the top of the coil form to clamp down the upper end of the coil wire. The upper end of the coil wire is set into a groove machined into the upper surface of the coil form. The depth of the cut is 0.010 inches less than the diameter of the wire, and the clamping force flattens the wire a little. This makes a solid electrical contact to between the coil and the radiating section of the antenna.

Fig 9—Matching network for 2-meter end-fed half-wave, telescoping element version.

In a similar manner, the SMA connector assembly fits into a machined groove on the bottom surface of

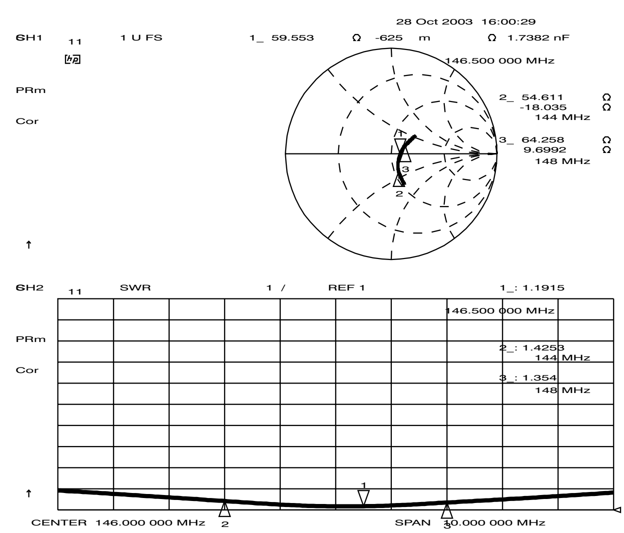

Fig 8—Drive-point impedance of the 2-meter version of the KØDK end-fed half-wave telescoping hand-held antenna.

Table 1
Main Parts List for the 2-meter Antenna

Radiating element —RadioShack 35 inch telescoping replacement antenna or a 36 inch length of stainless steel spring wire 0.062 inches in diameter.

Coil form—0.70-inch diameter Delrin rod, cut with a shallow exterior left hand thread at 6 turns/inch. The interior thread is 24 turns/inch (full length).

Coil—4½ turns of #12 AWG bare copper wire.

Antenna-element mounting post—0.40-inch-diameter aluminum rod, threaded on one end with 24 turns/inch for ¾-inch.

Capacitor—made from a piece of Teflon insulated wire about 2 inches long.

SMA male connector—mounted to one end of the coil form with two #3-48 machine screws.

the coil form to mechanically clamp the lower end of the coil. The depth of this cut is also 0.010 inches less than the diameter of the wire, for a tight press contact. See Figs 9 through 14. The SMA connector is held in place with two screws that thread into two holes in the coil form. The force produced on the upper and lower ends of the coil is sufficient to create a solid gas-tight, corrosion-resistant connection.

The series capacitor of the matching network is formed by a Teflon-insulated wire inserted into a hole drilled in the center of the threaded antenna element mounting post. The gimmick wire fits into a hole inside the threaded form, then the mounting post screws into the form. The capacitance is formed between the wire and the inside surface of the hole in the element mounting piece. The capacitor dielectric is the Teflon insulation of the short piece of wire as shown in the figures. The capacitor value is trimmed by cutting off the end in small increments, while observing the drive-point impedance of the antenna on the network analyzer using the hand-held simulator described above.

If the stainless-steel wire element is used instead of the telescoping element, the mounting stud is slightly different, as shown in Figs 12 and 13. The stainless steel wire is held in place with a setscrew as shown in the figures. The mounting stud then screws into the top of the coil form in the same manner. The capacitor is formed as before.

Notice that the coil provides a dc-ground connection between the antenna element and the case of the radio. This provides ESD protection for the radio front-end circuits in case you create a static discharge to the antenna. It also makes the antenna a lightning rod if you are operating outdoors. I have felt an electric-current tingling sensation while holding a hand-held on a mountaintop using this type of antenna. The air was highly charged with thunderstorm activity in the area, and I quickly descended to a lower, more protected location.

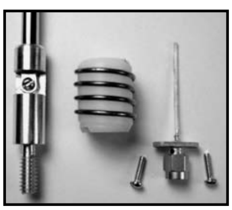

Fig 10—Disassembled matching network parts; radiator mount, coil, wire capacitor attached to SMA connector.

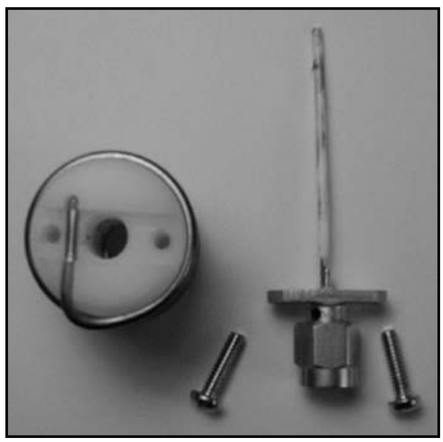

Fig 11—Bottom view of the 2 meter coil with the SMA connector and capacitor wire.

Fig 12—Top view of the 2 meter matching coil with the telescoping element.

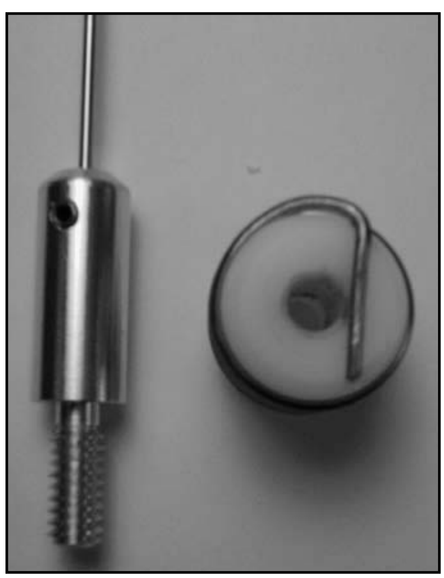

Fig 13—Top view of the 2 meter matching coil with the wire element.

Antenna Performance Comparisons

The performance proof of the homebrew end-fed half-wave antennas is in the measurement of gain and SWR. The relative gains are determined by a comparison with several different commercial antennas in outdoor measurements at an open-air test site (OATS) that is free of obstructions. The method of gain measurement for both the 2-meter and 70-cm antennas is as follows.

A battery-powered radio held in the hand, first at arm's length and then near the head, is used as a transmitting device for all antennas of the comparative tests. The power output of the Standard model C510A radio is 300 mW, which produces enough received signal strength in a nearby receiving antenna to be well above the noise floor of the receiving device. The hand-held is also equipped with with a 6 dB attenuator at its output. The attenuator is used to help present a consistent drive-point impedance, or high return loss, for the hand-held power amplifier. This insures that the power amplifier generates constant power for all antennas. The receiving antenna is a simple vertically polarized center-fed dipole. For the 2-meter measurements, the dipole is located 14 feet above the ground at a distance of 55 feet. For the 70-cm measurements, the dipole is located 5.5 feet above the ground at a distance of 40 feet.

A signal is transmitted by the hand-held first while held at arm's length and then held just a few inches from the head, in the normal talking position. These two received signal-strength measurements indicate how proximity to the user's head effects the radiation efficiency of the antenna. The receive signal strength was measured on a Hewlett Packard HP-8560A spectrum

Fig 14—2 meter matching network with the wire element.

analyzer set to 5-dB per division. This scale factor provides about 0.1-dB measurement resolution, which is far better than the measurement accuracy or repeatability for these tests.

For consistent readings, it is important to make the RF-level measurements insensitive to small variations in the position of the transmitting hand-held. This is best achieved when the ground reflection sums with the direct wave. So, the receiving antenna is positioned in the far field at a height and distance for a field strength maximum as calculated by

$$h1 = \left(\frac{\lambda}{4}\right)\left(\frac{s}{h2}\right) \qquad \text{(Eq 1)}$$

Where

$h1$ = height of the first maximum of the receiving antenna

$h2$ = height of the hand-held's transmitting antenna at its center

λ = wavelength transmitted, 2 m or 70 cm

s = the horizontal distance between the two antennas

For example, the 2-meter measurements are made at $h1$=14 feet, $h2$= 6.5 feet and s=55 feet. The 70-cm measurements are made at $h1$ = 5.5 feet, $h2$ = 4 feet and s = 40 feet. See the reference by John Kraus[3] for more details about choosing the best height for the antennas.

Results of the 2-Meter Gain Measurements

Here is how the performance of the homebrew end-fed half-wave antennas compares with a variety of commercial antennas. All measurements are made with the antennas vertically polarized. The gain of the homebrew antennas is as good or better than all of the commercial half-wave antennas and much better than any of the tested rubber ducks (Fig 15).

Notice in Table 2 that the greatest received signal strengths are created by the five end-fed half-wave antennas. The measured signals from all of the half-wave antennas are within 0.7 dB of each other. As a group, the end-fed half-wave antennas are an average of 1.4 dB stronger than the EMCO center-fed reference dipole. The EMCO reference dipole is less efficient possibly due to losses in its coaxial feedline and internal broadband balun. Notice also that the transmit signal strength of the end-fed half-wave antennas is not reduced more than a decibel or so when the hand-held is positioned in the normal talking position, near the head, rather than at arms length. The half-wave antennas are more efficient when near the body because the current maximum is at the center of the antenna and high above the head. There is also high current in the matching network coil at the base of the antenna, but its magnetic field seems to be confined enough so as not to interact much with the head and hand of the user.

In contrast, all of the rubber-duck antennas are much less efficient than any of the half-wave antennas. For example, the average of the transmit signal strengths of the rubber duck antennas is 5.9 dB weaker than the average of the half-wave antennas. The worst rubber duck is 7.9 dB weaker than the best half-wave antenna. This is very significant when you consider that a 6 dB increase in signal strength is the equivalent of multiplying your transmit power by a factor of four. (This loss applies on both transmit and receive.) Generally, the on-air tests of all of the antennas described in this article behave as the measurements would lead you to expect. In other words, the lower-gain antennas with high SWR usually are weaker into distant repeaters.

Fig 15—The hand-held antennas used for measurement comparisons to the homebrew half-wave antennas for 2 meters and 70 cm. In order from the left—Larson telescoping half-wave, AEA telescoping half-wave, Pryme quarter wave whip, Standard dual band for C510A, Larsen 2M, Larsen 70cm, Icom dual band for IC-Q7A, Standard dual band for C528A.

Table 2
2-Meter Antennas: Gain and SWR
(Note: See text "Accurate Hand-Held Radio Impedance Measurements".)

	At arm's length		*Near head*	
Antenna F_0=146.5 MHz (Vertical polarization)	*RF Level (Δ) in dBm*	*Impedance (SWR)*	*RF Level (Δ) in dBm*	*Impedance*
EMCO center-fed reference dipole tuned for 146.5 MHz [–19.5 (+0.0)]				
KØDK end–fed half–wave, whip	–18.7 (+0.8)	60–*j*0.6 (1.2)	–19.8 (–0.3)	73–*j*6
KØDK end–fed half–wave, stiff whip	–18.2 (+1.3)	55–*j*3.5 (1.1)	–19.3 (+0.2)	62–*j*10
KØDK end–fed half–wave, telescope	–17.7 (+1.8)	55–*j*8.9 (1.2)	–19.0 (+0.5)	58–*j*18
KØDK quarter wave, tape	–20.5 (–1.0)	93–*j*21 (2.0)	–22.5 (–3.0)	91–*j*17
Larsen end–fed half–wave, telescope	–17.3 (+2.2)	32–*j*28 (2.2)	–18.3 (+1.2)	43–*j*37
AEA end–fed half–wave, telescope	–18.3 (+1.2)	67–*j*0.6 (1.4)	–19.2 (+0.3)	78–*j*10
Pryme quarter wave, whip	–22.8 (–3.3)	100–*j*62 (2.9)	–25.0 (–5.5)	94–*j*57
Larsen 2 m rubber duck	–23.2 (–3.7)	56–*j*69 (3.4)	–25.6 (–6.1)	64–*j*68
ICOM 2 m/70-cm rubber duck for IC–Q7A	–24.2 (–4.7)	60–*j*26 (1.7)	–26.8 (–7.3)	64–*j*27
Standard 2 m/70-cm rubber duck for the C510A	–25.2 (–5.7)	63–*j*28 (1.7)	–28.0 (–8.5)	73–*j*26
Standard 2 m/70-cm rubber duck for the C528A	–23.3 (–3.8)	51–*j*53 (2.7)	–26.0 (–6.5)	55–*j*52

The results of the 2 Meter Impedance Measurements

All of the following antenna impedance measurements are made using the hand-held simulators shown above held at arms length. The drive-point impedances of all the antennas is measured using the HP8753D network analyzer in the manner described above. Read the network analyzer displays as follows - The upper part of the figure is a Smith Chart showing the impedance measured at the connector of the antenna plotted over a span of 10 MHz centered at 146 MHz. Three markers show the Z measured at 146.5, 144 and 148 MHz. Both the real and complex part of Z are shown as calculated from the S-Parameter S11. The lower part of the figure is a plot of SWR over the same frequency range as calculated from the measured impedance in relation to reference impedance of 50 W. The SWR markers are located at the same three frequencies. The bottom line of the graph indicates a SWR of 1:1 and the top line a SWR of 11:1.

From the Smith Chart in Fig 16 you can see that the SWR of the homebrew telescoping half-wave is pretty low across the band. This allows the hand-held power amplifier to drive the maximum power into the antenna for all frequencies. For some of the other measured antennas where the SWR is high the power amplifier of a given hand-held is often not able to develop enough voltage to drive the maximum power into the antenna. For example, the Pryme whip and the Larsen rubber duck exhibit a particularly high SWR making it unlikely that a hand-held will deliver its rated output power into the antenna. This factor contributes to the overall "inefficiency" of the antenna even though the power not transmitted into the ether is not lost as heat.

A low SWR for a hand-held antenna is also important when receiving a signal too because of the mismatch loss. For example, the received signal strength mismatch loss is about 1.25 dB for a SWR of 3:1. This is not a very great loss but can be important when signals are weak in a remote location or when you are inside a building or vehicle where the path loss to the repeater is very high.

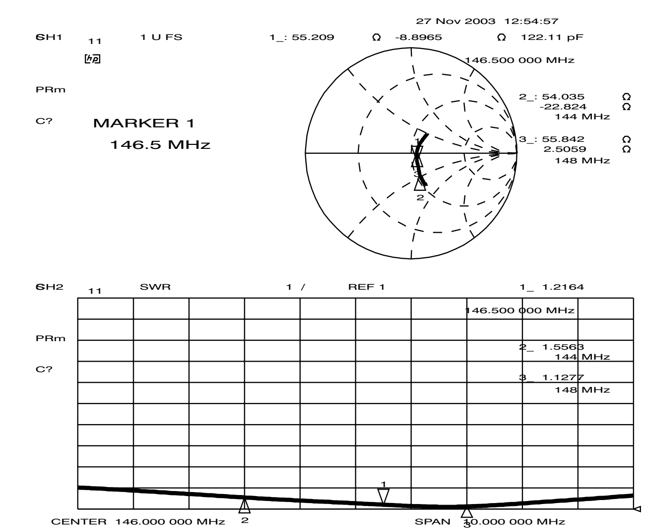

Fig 16—Impedance and SWR of the KØDK end-fed half-wave, telescope version.

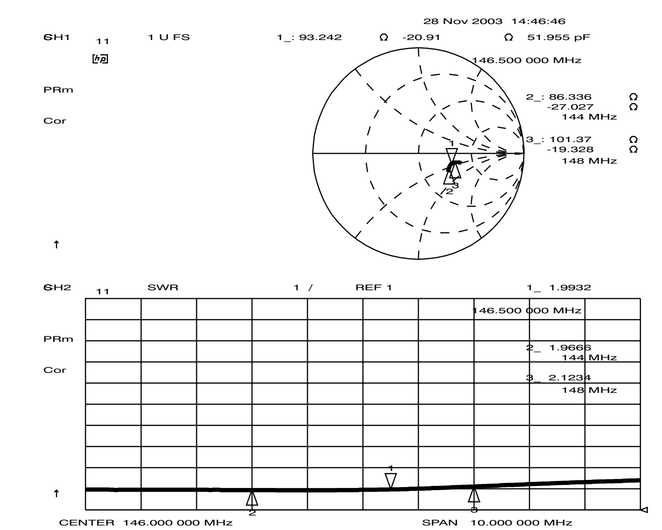

Fig 17—Impedance and SWR of the KØDK quarter wave, tape measure material.

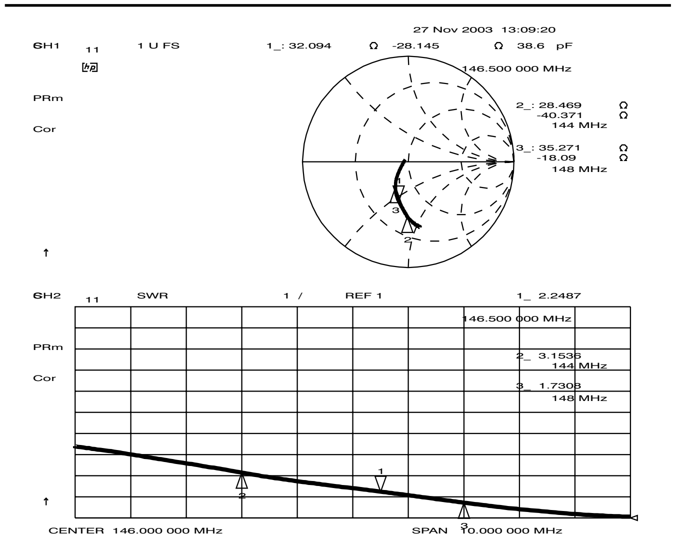

Fig 18—Impedance and SWR of the Larsen end-fed half-wave, telescope.

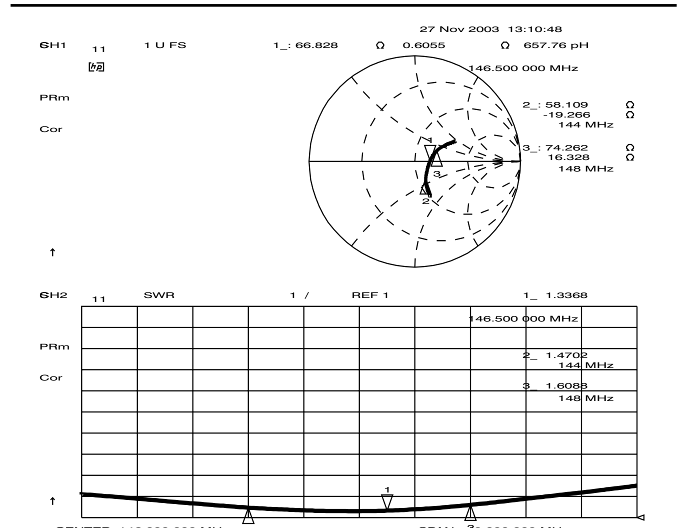

Fig 19—Impedance and SWR of the AEA end-fed half-wave, telescope.

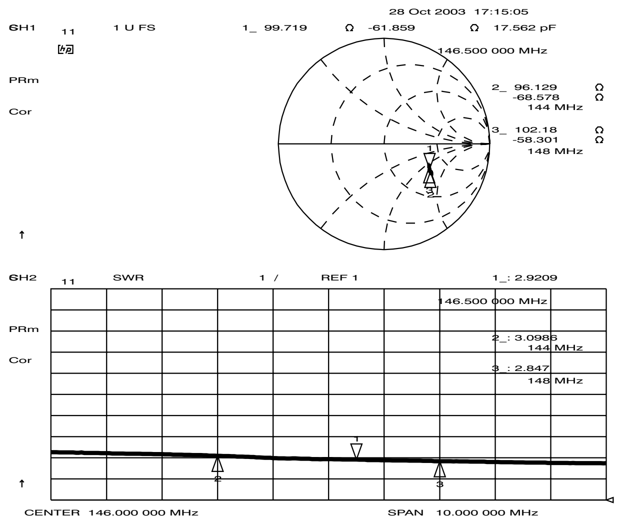

Fig 20—Impedance and SWR of the Pryme 2M/70cm quarter wave whip.

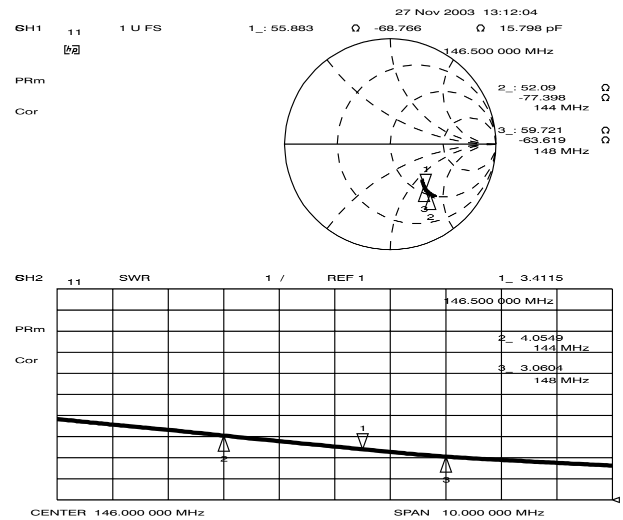

Fig 21—Impedance and SWR of the Larsen 2M rubber duck.

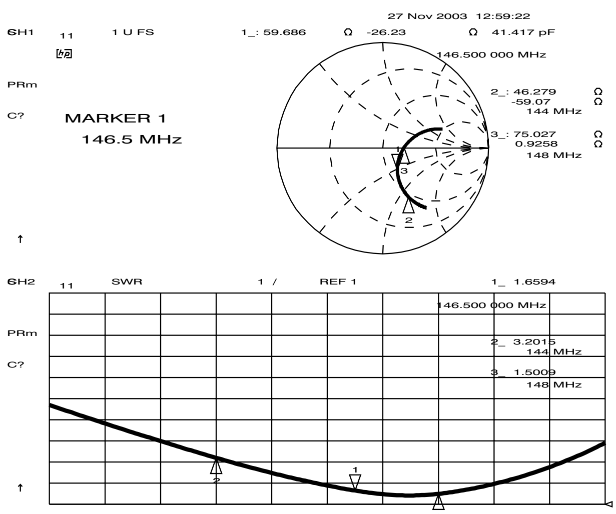

Fig 22—Impedance and SWR of the Icom 2M/70cm rubber duck for IC-Q7A hand-held.

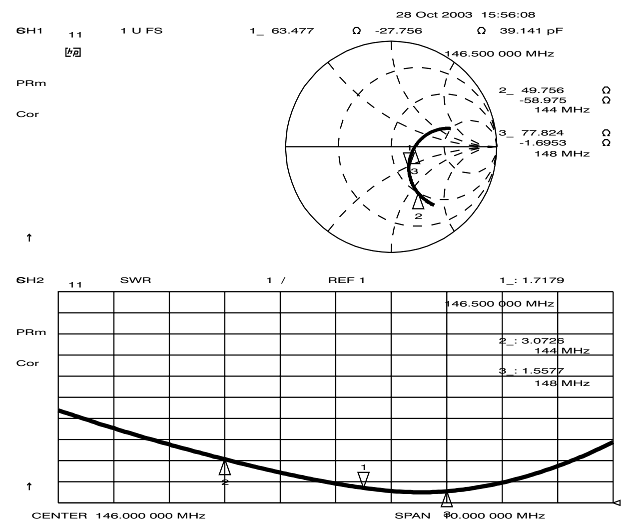

Fig 23—Impedance and SWR of the Standard 2M/70cm rubber duck for the C510A hand-held.

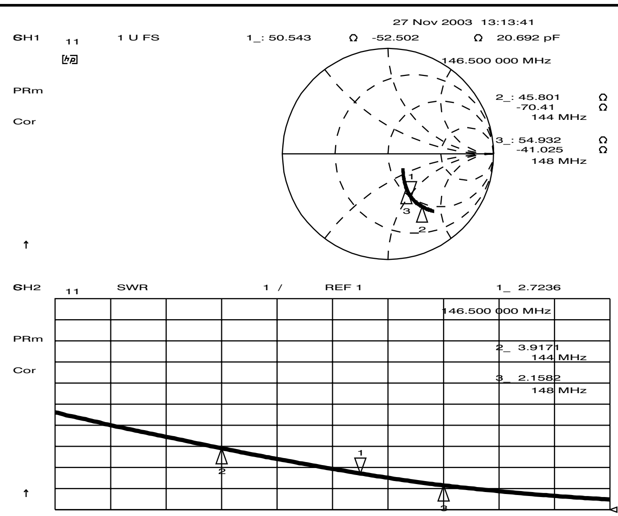

Fig 24—Impedance and SWR of the Standard 2M/70 cm rubber duck for C528A hand-held.

Results of the 70-cm Gain Measurements

Again, the homebrew half-wave antenna shows more gain than the rubber ducks. Refer to Table 3. In some cases, the advantage is slight, probably because the rubber ducks are much closer to a half-wave length on 70 cm than they are on 2 meters. For example, one half wavelength at 448 MHz is 33.5 cm and most of the rubber ducks used for this experiment are about 15-20 cm long. The transmit signal strengths of the rubber ducks are only an average of 4 dB worse than the

Table 3
70-cm antennas: Gain and SWR
(Note: See text "Accurate Hand-Held Radio Impedance Measurements".)

Antenna *F_0=146.5 MHz* *(Vertical polarization)*	*At arm's length* *RF Level (Δ) In dBm*	*Impedance (SWR)*	*Near head* *RF Level (Δ) in dBm*	*Impedance*
EMCO center-fed reference dipole tuned for 448 MHz [–32.5 (+0.0)]				
KØDK end-fed half-wave, whip	–34.3 (–1.8)	42+*j*10 (1.3)	–32.3 (+0.2)	46+*j*9.5
Pryme quarter-wave, whip	–40.0 (–7.5)	122+*j*19 (2.4)	–35.5 (–3.0)	137+*j*18
Larsen 70-cm rubber duck	–36.5 (–4.0)	56–*j*67 (3.4)	–39.8 (–7.3)	68–*j*58
ICOM 2 m/70cm rubber duck for IC–Q7A	–35.3 (–2.8)	38–*j*50 (3.0)	–36.0 (–3.5)	39–*j*41
Standard 2 m/70-cm rubber duck for the C510A	–34.8 (–2.3)	59–*j*48 (2.4)	–36.7 (–4.2)	56–*j*32
Standard 2 m/70-cm rubber duck for the C528A	–36.2 (–3.7)	66–*j*91 (4.4)	–38.6 (–6.1)	76–*j*75

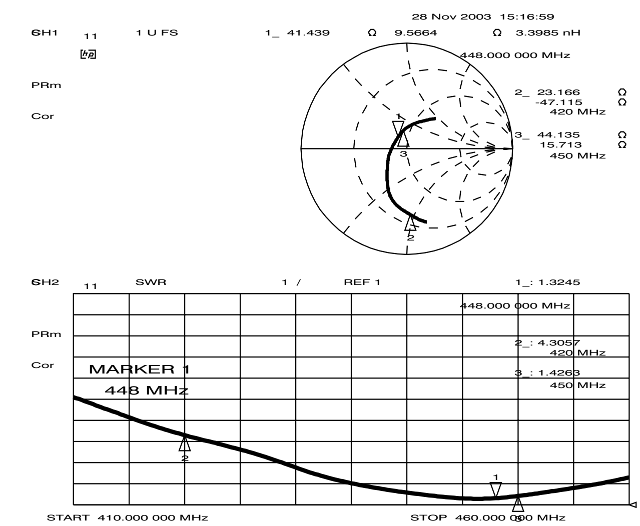

Fig 25—Impedance and SWR of the KØDK 70-cm end-fed half-wave, wire version.

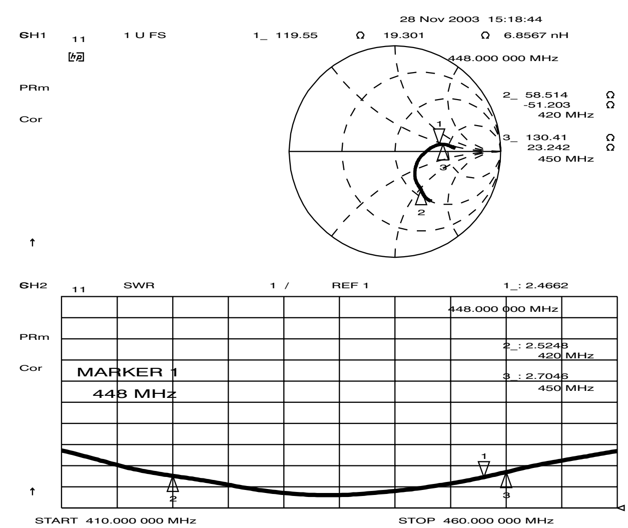

Fig 26—Impedance and SWR of the Pryme dual band whip at 70-cm.

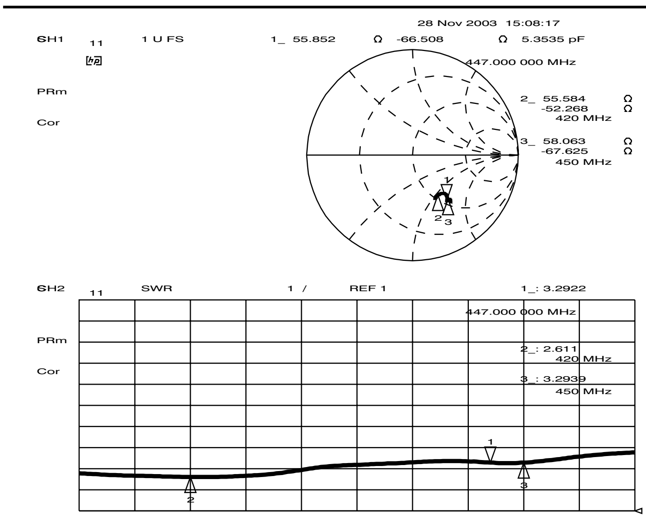

Fig 27—Impedance and SWR of the Larsen 70-cm rubber duck.

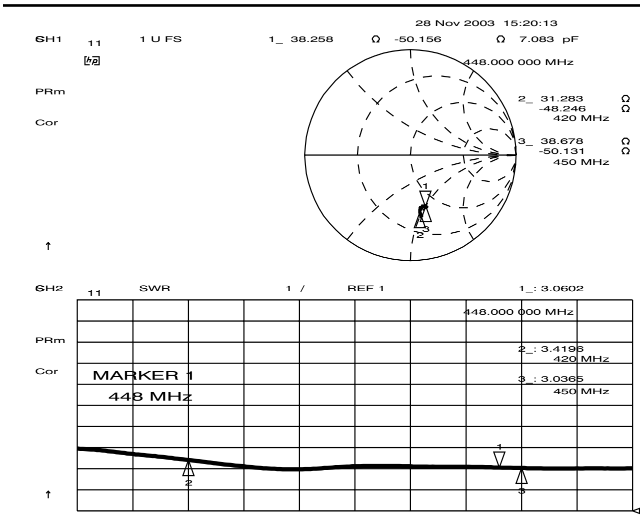

Fig 28—Impedance and SWR of the ICOM dual-band for IC-Q7A at 70-cm.

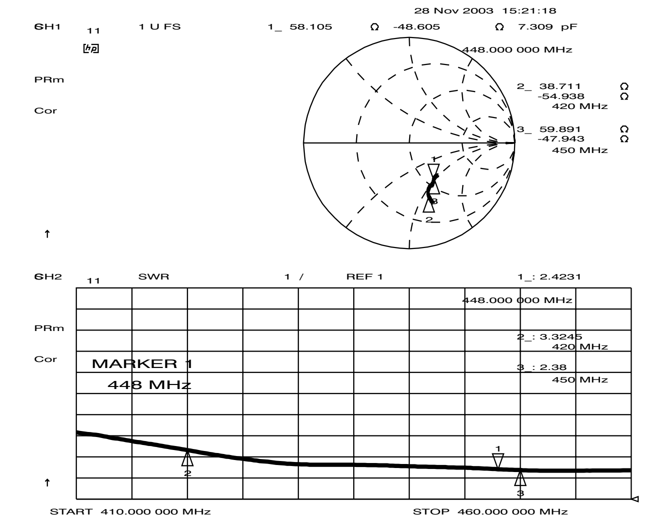

Fig 29—Impedance and SWR of the Standard C510A dual band duck at 70 cm.

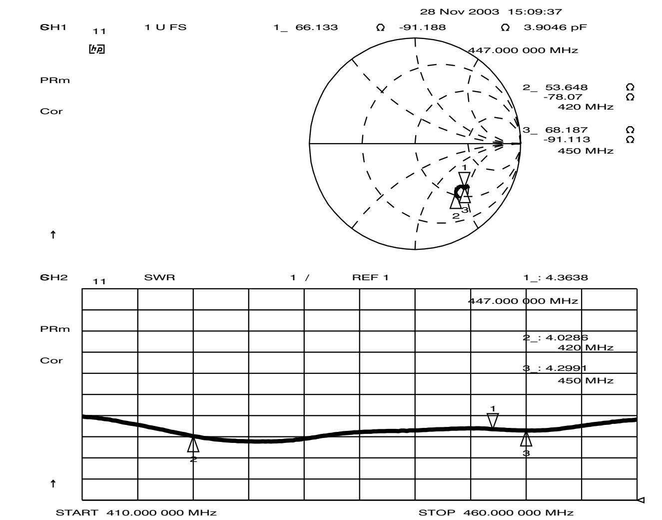

Fig 30—Impedance and SWR of the Standard C528A dual-band at 70 cm.

homebrew half-wave. This is not as large a difference as with the 2-meter antennas. Moving the rubber ducks closer to the head also reduces their gain as on 2 meters. The two whip antennas increased in gain when brought closer to the head of the user for unknown reasons. It is possible that the body acts as a reflector? It is also likely that the margin of measurement error for all of the data below is somewhat greater at 448 MHz than at 146.5 MHz because of the much shorter wavelength. Much smaller dimensional and positional differences also make a larger change in the readings.

*Richard Kiefer has been a licensed ham since 1959. His early interest in Amateur Radio led to Bachelor's and Master's degrees in Electrical Engineering with an academic emphasis in the subjects of radio-frequency and analog circuit design. He has worked as an electronics engineer since 1970, designing a variety of analog and RF circuits, antennas and microprocessor systems. He has also written software and firmware in several languages to control various circuits and hardware. Richard has worked for several electronic product-development companies, including IBM, Hewlett-Packard, Martin-Marietta and Armco Autometrics. For the past 24 years he has been the principal of Kiefer Electronic Development (*__www.KED-Wireless.com__*), an electronic product-development consultancy specializing in radio-frequency product design. He also holds five US patents. He currently enjoys working SSB DX on the 10-40 meter HF bands with a Yagi stack on a 100 foot rotating tower. He also works DX mobile through his EchoLink connected VHF repeater located at Boulder, Colorado. Richard's most recent interest is the study of efficient radio-spectrum management, including the Amateur Radio allocations, in relation to the recommendations of the FCC Spectrum Policy Task Force.*

Notes

[1] *P. Smith, Electronic Applications of the Smith Chart* (McGraw-Hill Book Company)
[2] C. Bowick, *RF Circuit Design*, (Howard W. Sams and Company) chapter 4, p 66.
[3] J. D. Kraus, *Antennas* (2nd edition, McGraw-Hill Book Company) section 18-3b, pp 811-813 and appendix E pp 870-871. ☐☐

A Study of Phased Vertical Arrays

The author describes a number of possible multi-element vertical antenna arrays, outlining their features and the differences in expected performance.

By Al Christman, K3LC

Introduction

The 4-square phased-vertical array is widely used by DXers and contesters for operation on the low bands from 40 through 160 meters. There are a number of different permutations of this array that are useful to hams, including variations in element spacing and current phasing. Several larger arrays are available, using from six to nine elements, and these also deserve consideration. This article reviews the performance of the typical 4-square array in terms of forward gain, front-to-back ratio, front-to-rear ratio, and azimuthal beamwidth. A variety of modified 4-squares are also included, along with the 6-hex and 9-circle array designs.

Element Description

A frequency of 3.65 MHz in the 80-meter band was selected for the computer analysis, which was performed using EZNEC-4.[1] A full-size quarter-wave vertical element composed of # 12 AWG copper wire was placed over average soil with a conductivity of 5 millisiemens per meter and a dielectric constant of 13. The segment lengths for the vertical monopole were tapered in accordance with the most conservative NEC guidelines, with the shortest (feed-point) segment having a length of 6 inches. I selected # 16 AWG copper wire for the radials, since this wire is widely available at reasonable cost. The inner segment of each buried radial slopes downward into the soil from the base of the vertical element, to a depth of 3 inches, with the major portion of the radial extending horizontally at this same depth.

I decided to utilize 5000 feet of wire in the radial ground-screen for each element, and conducted a series of tests to determine the optimum number and length for these buried wires. Simulation on the computer indicated that I could get the same gain using anywhere from 46 to 52 radials, whose corresponding lengths spanned the range from 108.7 to 96.2 feet. Because I planned to use arrays containing either 4, 6, or 8 symmetrically-placed elements (the 9-circle array positions one element in the center), I decided to utilize 48 radials per element, with each one having a maximum length of about 104.2 feet.

Modelling Considerations

For all of the arrays, I configured the radials in a fashion similar to what would be found at an AM-broadcast antenna site. Fig 1 illustrates this technique, and shows the computer-generated model for the conventional 4-square. Copper wires (# 12 AWG) serve as "bus-bars" to which radials from adjoining elements are soldered where they intersect. This methodology prevents any overlap of one element's ground-screen with another, and reduces the total amount of wire needed for the radial system. For consistency, I oriented each array so that its main beam would fire at an azimuth of 45°, common practice here in the north-eastern part of the US.

Classical 4-Square

The first array to be examined was the popular 4-square, which has now been in use for several decades, since it was popularized in the well-known 1976 *QST* article.[2] In this configuration, the four elements are placed at the corners of a square whose side-length is 0.25 λ. Normally the feed system is set up so the array fires through the corners of the square. This can be accomplished by driving all four elements with equal-amplitude currents. The usual current phase-angles are 0° for the rear element, –90° for the two side elements, and –180° for the front element. EZNEC predicts a peak forward gain of 6.16 dBi at a take-off angle (TOA) of 23°, with a front-to-back ratio (FBR) of 16.58 dB in the elevation plane. In the azimuthal plane the FBR is 22.71 dB, while the half-power beamwidth (HPBW) is 100.6°.

Fig 2 shows the elevation-plane radiation pattern for the array, and also illustrates the pattern that results when firing through the sides of the square. To do this, all four elements are driven with equal-amplitude currents, as before, but now the two rear elements are driven at a phase-angle of 0°, while the two front elements are both driven at –90°. In this case the maximum forward gain is 5.18 dBi at 25° TOA, and the FBR is 12.87 dB. In the azimuthal plane the FBR is 21.97 dB, while the HPBW is 130.8 degrees.

[1]Notes appear on page 51.

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Fig 3 displays both azimuthal-plane radiation patterns together on the same plot, showing how the secondary feed configuration helps to fill in those regions where the primary beams are at their half-power points.

4-Square with W8JI Feed

As Tom, W8JI, points out,[3] the phase-angles of the currents that are used to drive the typical 4-square array are far from optimal, and should be much larger. W8JI suggests that the two middle elements be driven with currents whose phase angle is –120° with respect to the rear element, while the current in the front element should have a phase-angle of –240°. This strategy yields a peak forward gain of 6.61 dBi at a TOA of 22°, with a front-to-back ratio of 25.23 dB in the elevation plane. In the azimuthal plane, the FBR is 31.50 dB and the HPBW is 80.2°. The radiation patterns are shown in Fig 4 and 5. Notice that almost half a decibel of extra gain has been obtained from the same physical array as before, simply by changing the phase-angles of the base currents. The new pattern is much sharper, as evidenced by a significant reduction in the azimuthal beam-width, and the FBR is very high, although there are some fairly-large side lobes.

Since the HPBW has been reduced from 100.6 to 80.2°, the four traditional directions of fire (through the corners) no longer provide good coverage of all sectors of the compass. Thus, it is advantageous to design the feed system of this array so it can fire through the sides of the square as well as through the corners. To do this, we could simply repeat the strategy used earlier with the classic 4-square, feeding the two rear elements at 0° and the two front elements at –90°, which would produce the same values of gain, FBR, etc, as were quoted above. However, by driving the two front elements at –110°, we can increase the maximum forward gain to 5.57 dBi at 24° TOA, while the FBR improves to 16.51 dB. In the azimuthal plane, the FBR is now 18.36 dB, and the HPBW is 116.6°. These radiation patterns are also displayed in Fig 4 and 5.

Expanded 4-Square

Another way to get more gain from

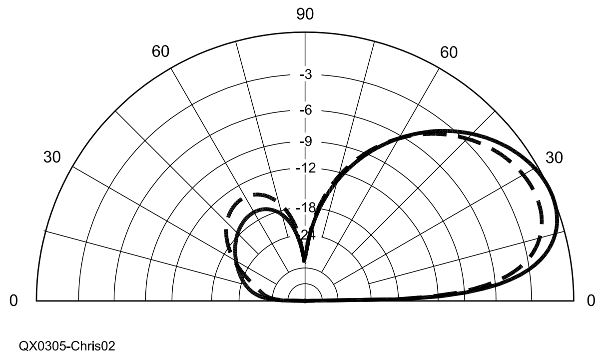

Fig 2—Elevation-plane radiation patterns for the conventional 4-square array. Solid trace: firing through the corner; gain = 6.16 dBi at 23° take-off angle, front-to-back ratio = 16.58 dB. Dashed trace: firing through the side; gain = 5.18 dBi at 25° take-off angle, front-to-back ratio = 12.87 dB.

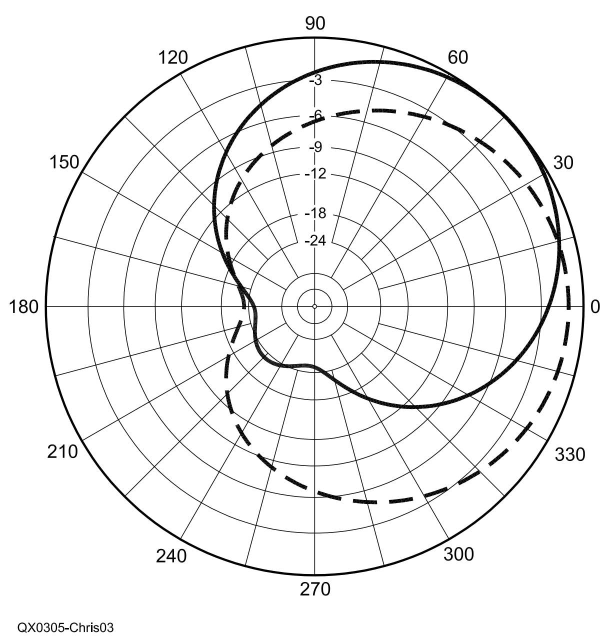

Fig 3—Azimuthal-plane radiation patterns for the conventional 4-square array. Solid trace: firing through the corner; front-to-back ratio = 22.71 dB, half-power beamwidth = 100.6°. Dashed trace: firing through the side; front-to-back ratio = 21.97 dB, half-power beamwidth = 130.8°.

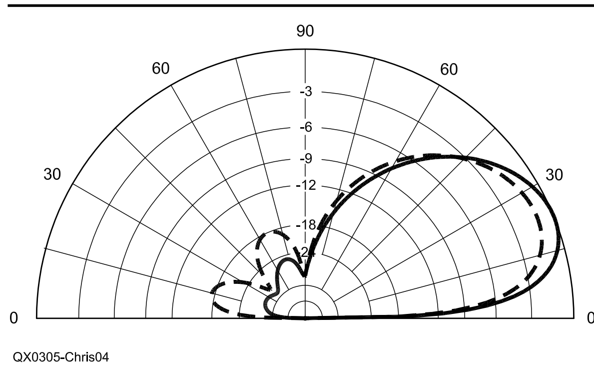

Fig 4—Elevation-plane radiation patterns for the 4-square array, with W8JI-style feed. Solid trace: firing through the corner; gain = 6.61 dBi at 22° take-off angle, front-to-back ratio = 25.23 dB. Dashed trace: firing through the side; gain = 5.57 dBi at 24° take-off angle, front-to-back ratio = 16.51 dB.

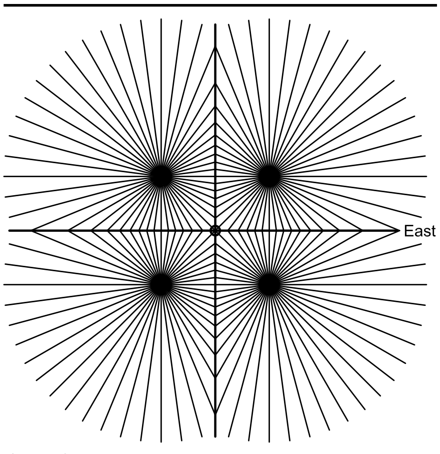

Fig 1—A bird's-eye view of a traditional 4-square phased-array, showing details of the broadcast-style radial ground-screen. Each element has 48 radials, whose maximum length is 104.2 feet.

a 4-square is to make it larger by spacing the elements farther apart. If we want to use the same phase-angles as in the traditional array (0, –90, –90, and –180°), then a diagonal spacing of 0.5 λ is suitable,[3] which means the side-length of the square expands from 0.25 λ to about 0.354 λ. The peak forward gain rises to 6.84 dBi at a TOA of 24°, and the FBR is 20.98 dB in the elevation plane. In the azimuthal plane, the FBR is 35.26 dB (although there are large side-lobes) and the HPBW is 87.0°. Fig 6 and 7 illustrate the radiation patterns.

As was true with the W8JI-style feed, the azimuthal-plane beam-width of this antenna is less than 90°, so we need to make the array fire through the sides of the square as well. We can use the traditional phase-shifts of 0 and –90°, which yield a peak forward gain of 6.11 dBi at 27° TOA, and a FBR

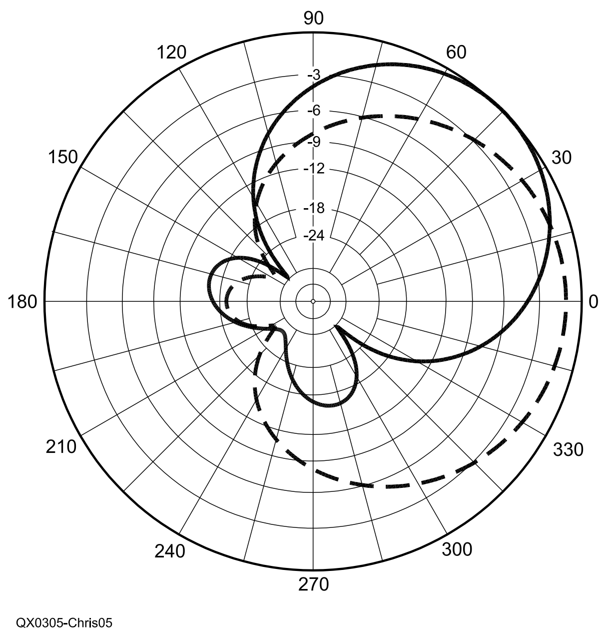

Fig 5—Azimuthal-plane radiation patterns for the 4-square array, with W8JI-style feed. Solid trace: firing through the corner; front-to-back ratio = 31.50 dB, half-power beamwidth = 80.2°. Dashed trace: firing through the side; front-to-back ratio = 18.36 dB, half-power beamwidth = 116.6°.

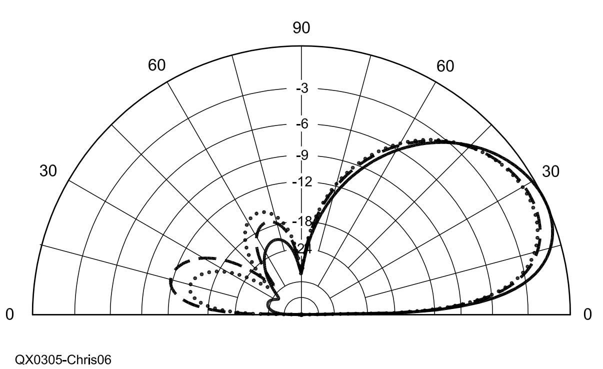

Fig 6—Elevation-plane radiation patterns for the expanded 4-square array. Solid trace: firing through the corner; gain = 6.84 dBi at 24° take-off angle, front-to-back ratio = 20.98 dB. Dashed trace: firing through the side, current phase-angles of 0 and –90°; gain = 6.11 dBi at 27° take-off angle, front-to-back ratio = 11.01 dB. Dotted trace: firing through the side, current phase-angles of 0 and –80°; gain = 5.98 dBi at 27° take-off angle, front-to-back ratio = 13.71 dB.

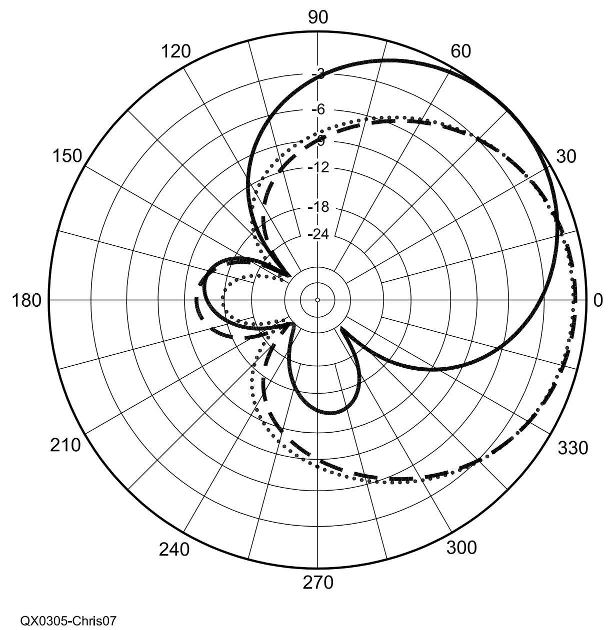

Fig 7—Azimuthal-plane radiation patterns for the expanded 4-square array. Solid trace: firing through the corner; front-to-back ratio = 35.26 dB, half-power beamwidth = 87.0°. Dashed trace: firing through the side, current phase-angles of 0 and –90°; front-to-back ratio = 12.94 dB, half-power beamwidth = 105.4°. Dotted trace: firing through the side, current phase-angles of 0 and –80°; front-to-back ratio = 16.98 dB, half-power beamwidth = 111.0°.

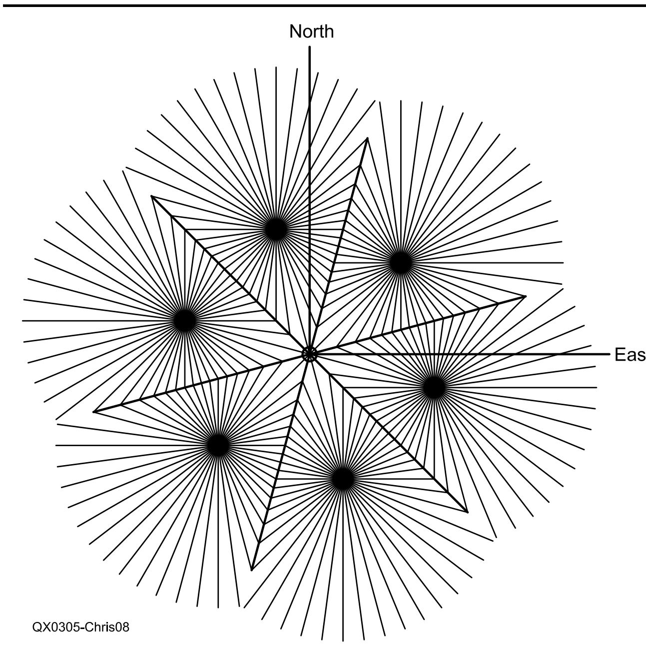

Fig 8—A bird's-eye view of the K3YA 6-element hexagonal array, showing details of the broadcast-style radial ground-screen. Each element has 48 radials, whose maximum length is 104.2 feet.

of just 11.01 dB in the elevation plane. There are two minor lobes in the rear portion of the radiation pattern, but the lower one is quite large. In the azimuthal plane, the FBR is 12.94 dB and the HPBW is 105.4°.

Reducing the phase-angle of the driving-point currents for the two front elements from –90 to –80° improves the performance by equalizing the size of the two rear lobes. Now the maximum forward gain falls a bit, to 5.98 dBi at a TOA of 27°, but the FBR rises to 13.71 dB in the elevation plane. In the azimuthal plane, the FBR increases to 16.98 dB, with a HPBW of 111.0°. All of these additional patterns are included in Fig 6 and 7.

K3YA 6-Hex Array

In the early part of 2004, I had the opportunity to visit Joe Johnson, K3RR, at his new QTH near Gettysburg, PA. While there, Joe told me of his plans to construct a 6-element array of short verticals that would be used exclusively for receiving purposes on 80 meters. As K3RR described the antenna design, which he credits to K3YA, I realized that it would make an excellent transmit array as well, and Joe agreed. Fig 8 is a computer rendition of the antenna with its ground screen, when viewed from directly overhead.

The design is very clever, and consists of six elements which are equally spaced to form a hexagonal shape. The spacing between adjacent elements is equal to the distance from any element to the center of the array. Initially, I performed some computer analysis of the array over a very simple MiniNEC-style ground in order to optimize the element spacing, and found that Joe's suggested value of 70 feet worked well at an operating frequency of 3.65 MHz. If the inter-element spacing is made larger, the forward gain will rise, but the azimuthal-plane beamwidth decreases.

Only four of the elements are active at any given time, while the other two are open-circuited at their bases. To fire northeast (see Fig 8), both the northeast and southwest elements are idle, while the other four are all driven with equal-amplitude base currents. The two elements in the northwest quadrant form an end-fire (cardioid) pair, as do the two elements in the southeast quadrant. Together, these two 2-element sub-arrays are then fed as a broadside array that beams to the northeast. The two rear elements have a current phase-angle of 0°, while the two front elements are fed at –105°. A phase-angle of –90° can also be used to drive the two front elements, but the gain and FBR in the elevation plane are not as good. Fig 9 and 10 display the radiation patterns. The peak forward gain is 7.53 dBi at a TOA of 24°, and the FBR is 16.69 dB in the elevation plane. In the azimuthal plane, the FBR is 19.24 dB, with a half-power beamwidth of 71.4°.

In the K3YA array, we can see that the element spacing of the two end-fire pairs is greater than what would typically be used, while the broadside spacing is somewhat smaller than normal. However, despite the compromises in spacing, this 6-element configuration provides the low-band operator with the equivalent of three separate four-element reversible arrays, using a system which requires only 6 elements.

While discussing phased vertical arrays with John, WØUN, he pointed me to the Web-site of Greg, W8WWV.[4] Greg has done a lot of work with 6-element hexagonal arrays, and has constructed a multi-band antenna at his QTH using an inter-element spacing of just 40 feet. The K3YA-style 6-hex fires through the corners, but W8WWV's array fires through the sides, and all six elements are driven at all times. In essence, Greg's antenna functions as two 3-element arrays which are parallel to one another, except that the two middle elements have been moved outward so they are farther apart than the front and rear pairs (which produces the hexagonal shape).

I chose to keep the hexagon's element spacing fixed at 70 feet, as before, and then adapted Greg's current phase-angles to achieve a half-power beamwidth which was similar to that of the K3YA version. In W8WWV's array, the two center elements receive twice as much current as the front and rear elements, so I did the same with mine. In the final version here, the two center

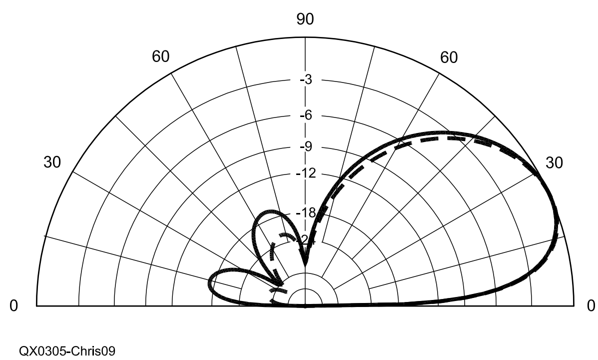

Fig 9—Elevation-plane radiation pattern for the two 6-element hexagonal arrays. Solid trace—K3YA version: gain = 7.53 dBi at 24° take-off angle, front-to-back ratio = 16.69 dB. Dashed trace—W8WWV version: gain = 7.54 dBi at 23° take-off angle, front-to-back ratio = 21.94 dB.

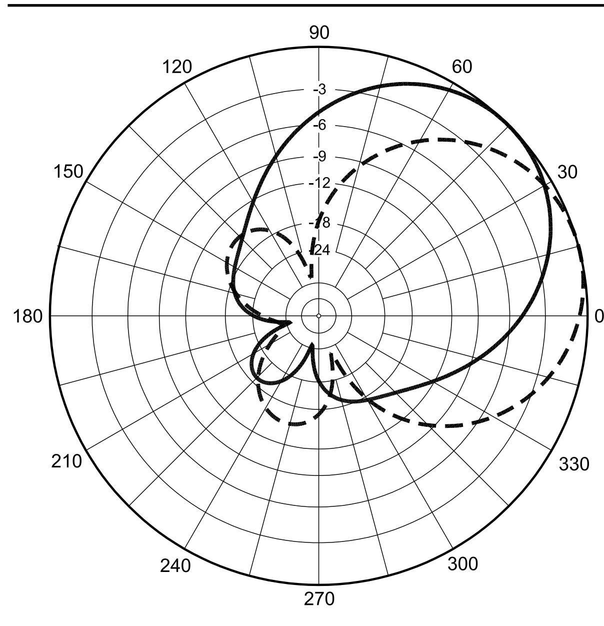

Fig 10—Azimuthal-plane radiation pattern for the two 6-element hexagonal arrays. Solid trace—K3YA version: front-to-back ratio = 19.24 dB, half-power beamwidth = 71.4°. Dashed trace—W8WWV version: front-to-back ratio = 33.67 dB, half-power beamwidth = 71.2°.

elements have the reference current phase-angle of 0°, while the two front and two rear elements are fed at –100° and +100° respectively. Figures 9 and 10 display the new radiation patterns, in conjunction with those of the K3YA array. The W8WWV-style antenna has a peak forward gain of 7.54 dBi at a TOA of 23°, and the FBR is 21.94 dB in the elevation plane. In the azimuthal plane, the FBR is 33.67 dB (although there are significant side-lobes), and the HPBW is 71.2°. Comparing the two versions of the 6-hex, we can see that both yield similar gains, to within 0.01 dBi, although the W8WWV design has a slightly-lower take-off angle. The azimuthal-plane beam-widths are also extremely close, which was my intention. The two arrays differ mainly in their ability to reject unwanted signals arriving from the back, with the W8WWV antenna being the clear winner. Theoretically, it would be possible to combine both versions together into a single antenna (using a lot of networks and switching circuitry) to create a 6-element hexagonal array with 12 directions of fire.

WØUN 9-Circle Array

The final antenna to be investigated is the 9-circle array of John Brosnahan, WØUN.[5] This antenna is formidable in design and construction, and the feed network is rather complex, but it pays big dividends, both in terms of gain and front-to-back ratio. One element is installed at the center of a circle whose radius is 0.39 λ at the frequency of interest, while the other 8 elements are placed at equal intervals around this circle. At 3.65 MHz, the radius of the circle is 105.1 feet. There are 8 directions of fire, directly outward through each of the 8 outer elements, and all 9 elements are active at all times.

When beaming to the northeast, the extreme front and rear (NE and SW) elements are fed with equal-amplitude (1.0 A relative) currents. The phase-angle of the current into the NE element is –180°, while that of the SW element is +180°. The center element and the two elements located directly to its northwest and southeast are all fed with currents whose phase-angle is 0°; the central element has a relative current amplitude of 3.0, while the two out-riggers have unity (1.0) current amplitude. The two intermediate front/side elements both have base currents of 1.64 at a phase-angle of –90°. (The suggested current amplitude of 1.66 was adjusted slightly to provide better performance.) As already stated, the rear-most element is driven with a

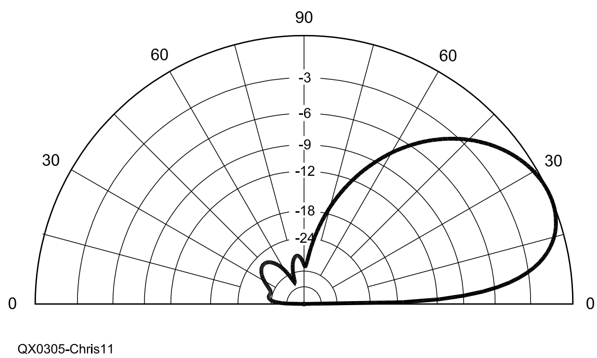

Fig 11—Elevation-plane radiation pattern for the WØUN 9-element circular array. Gain = 8.75 dBi at 23° take-off angle, front-to-back ratio = 26.72 dB.

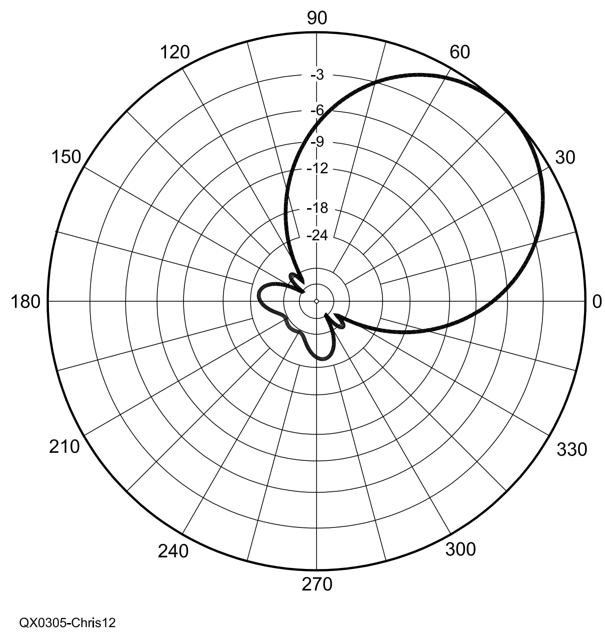

Fig 12—Azimuthal-plane radiation pattern for the WØUN 9-element circular array. Front-to-back ratio = 33.81 dB, half-power beamwidth = 60.4°.

Table 1—Performance Data for the Phased Vertical Arrays discussed herein. Elevation-plane data includes the gain and take-off angle (TOA) of the main lobe, the front-to-back ratio (FBR) and the front-to-rear ratio (FRR). Azimuthal-plane data includes the front-to-back ratio (FBR) and front-to-rear ratio (FRR), along with the half-power beamwidth (HPBW). The three 4-square arrays all have two lines of data; the first line is for the main beam when firing through the diagonal of the square, and the second line is when firing through the side of the square.

	Elevation-Plane Data			*Azimuthal-Plane Data*		
	Gain & TOA	*FBR*	*FRR*	*FBR*	*FRR*	*HPBW*
Array	*(dBi & deg)*	*(dB)*	*(dB)*	*(dB)*	*(dB)*	*(deg)*
9-Circle	8.75 @ 23	26.72	32.10	33.81	32.51	60.4
6-Hex (W8WWV)	7.54 @ 23	21.94	30.79	33.67	20.84	71.2
6-Hex (K3YA)	7.53 @ 24	16.69	21.63	19.24	22.17	71.4
Expanded	6.84 @ 24	20.98	30.21	35.26	21.73	87.0
4-Square	5.98 @ 27	13.71	18.93	16.98	16.65	111.0
4-Square	6.61 @ 22	25.23	30.58	31.50	22.23	80.2
W8JI-feed	5.57 @ 24	16.51	21.11	18.36	17.34	116.6
Regular	6.16 @ 23	16.58	21.66	22.71	21.22	100.6
4-Square	5.18 @ 25	12.87	20.13	21.97	12.96	130.8

current of 1.0 at an angle of +180°, while the two intermediate back/side elements both have base currents of 1.64 at a phase-angle of +90°.

The maximum forward gain of this array is 8.75 dBi at a TOA of 23°, and the FBR in the elevation plane is 26.72 dB. In the azimuthal plane, the FBR is 33.81 dB, and the HPBW is 60.4°. Plots of the radiation patterns are included as Figures 11 and 12.

Conclusions

This article has summarized the capabilities of a variety of phased-vertical arrays that are of interest to the low-band operator. Table 1 lists the important performance parameters of each array, for reference and comparison. In this table, I have listed the "FRR" (front-to-rear ratio) in addition to the typical front-to-back ratio (FBR). For each array, the customary "front- to-back ratio" data has already been given in the text and in the accom-panying figures, but these statistics fail to account for the presence or absence of multiple secondary lobes, which can be large. To perform the FRR calculations, I averaged the gain at five-degree increments throughout the rear portion of each radiation pattern, and then subtracted this value from the peak forward gain in order to obtain the FRR value.

Each of these arrays was also simulated using a ground-screen consisting of 50 100-foot radials per element. In this scenario, none of the radials were truncated in length, although many had to be buried at different depths in order to avoid unwanted "collisions" which led to software crashes. In every case, this alternative ground-screen configuration yielded slightly-higher values of forward gain, ranging from about 0.1 to 0.3 dB. Since all of the radials in these ground screens are full-length, considerably more wire is required to construct them.

Notes

[1]EZNEC is available from Roy Lewallen, W7EL, PO Box 6658, Beaverton OR 97007.

[2]D. Atchley, W1CF, H. Stinehelfer, ex W2ZRS, and J. White, PhD, "360-degree-Steerable Vertical Phased Arrays," *QST*, April 1976, pp 27-30.

[3]T. Rauch, W8JI, **www.w8ji.com/tx_four_square.htm**.

[4]G. Ordy, W8WWV, **www.seed-solutions.com/gregordy/Amateur%20Radio/Experimentation/HexArray/8040Array.htm.**

[5]J. Devoldere, ON4UN, *ON4UN's Low-Band DXing, Third Edition*, ARRL, 1999, pp 11-68—11-71.

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Antenna Options: A Yagi Case Study Part 3—Construction Options

By L. B. Cebik, W4RNL

In the first episode of this "Tale of 3 Yagis," we explored the design options for a three-element 2-meter Yagi intended for field use and restricted to a 30-inch boom or smaller. Our options included high-gain, high F/B and wide-bandwidth versions of the antenna. Each option provided design dimensions for round-tubing elements ranging from 1/8 up to 1/2-inch in diameter. In the second episode, we examined some of the element materials other than round tubing that we may use for the Yagis. As well, we looked at the process of correlating these materials to the dimensions in the first part of this exercise.

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In this final portion of the exercise, we shall explore some of the construction options involved in building the small 3-element Yagi—whatever the selected design and element material. Our perspective will not be commercial construction, but rather what we can accomplish within a typical home shop.

Part 3: Building 3-Element Yagis for Different Uses

A commercial antenna designer might begin with a set of operating or use specifications and then select materials and construction methods that will achieve those goals. However, the average home antenna builder often begins from a different position. He or she has some materials, some shop abilities (and limitations) and some uses for the final product. The next step is usually reaching a physical design that combines these starting points into one antenna. Therefore, let's examine a few building options for both permanent (long-term) installations and for portable (field) antennas.

Long-term Construction Techniques

Utility antennas on 2 meters can make good use of 1/2" to 1" nominal PVC pipe as the boom material. The elements are light enough that you can place a T fitting off center (to avoid the driver position) and still have a stable mounting. Indeed, you may even extend the boom rearward so that the boom-to-mast assembly is behind the reflector. This system is especially suitable for vertically oriented Yagis to suppress interactions between the mast and the array. The non-conductive boom also means that you may use the dimensions in Part 1 as a direct guide to

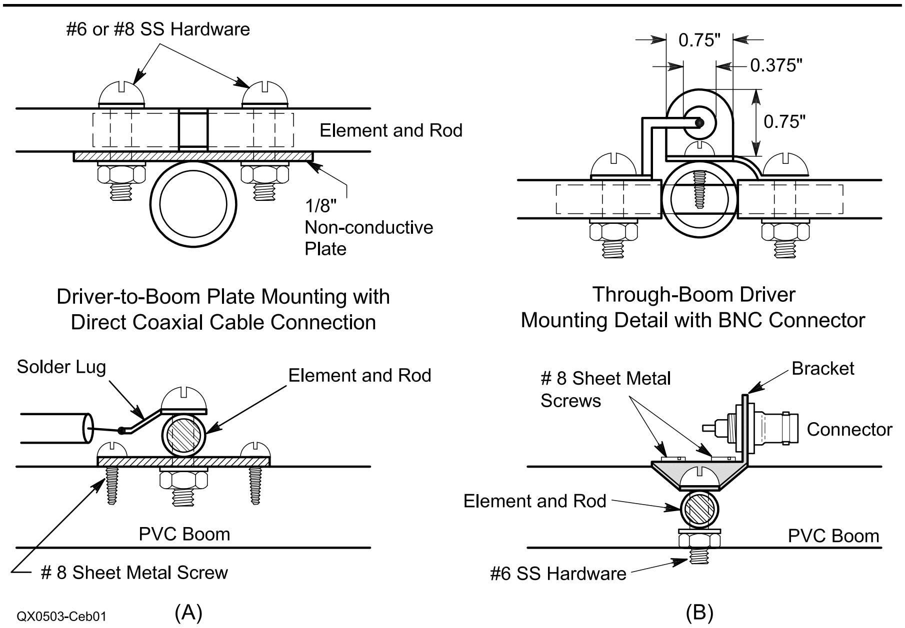

Fig 1—Some alternatives for driver mounting assemblies.

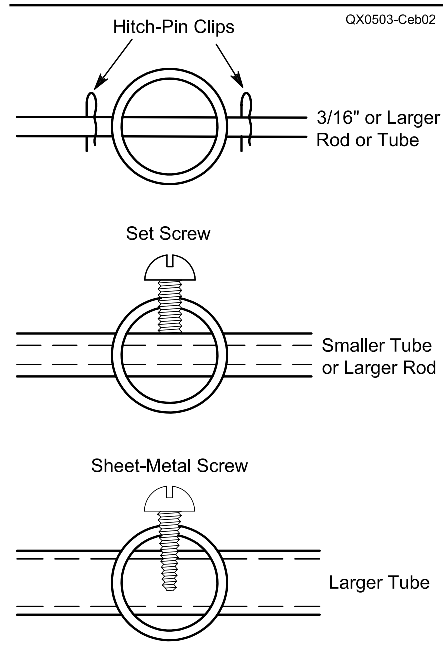

Fig 2—Some alternatives for parasitic element mounting.

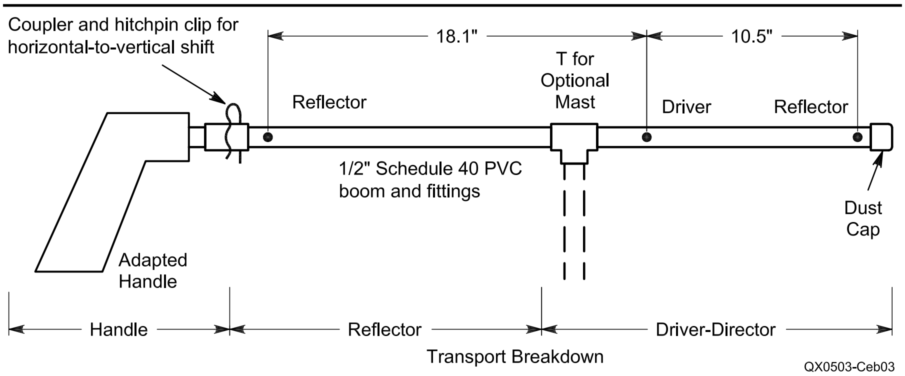

Fig 3—Boom sketch of the demonstration field antenna that uses 3 different alternative element materials.

Fig 4—The demonstration Yagi broken into pieces for storage and transport.

construction without adjustment for the use of a metallic boom.

Not all white PVC found in the US is UV resistant. If the white PVC in your area is UV susceptible, then use the gray electrical-conduit version; it tends to be more uniformly UV resistant. Other potential boom materials include fiberglass and other resin-based non-conductive tubular materials.

For long-term installations, I recommend round elements, either aluminum tubes or rods. All of the designs that appear in Part 1 use split driver elements to minimize the number of mechanical connections on the driver assembly. The low-impedance drivers are resonant in the high-gain version for use with a 35-37 Ω $\lambda/4$matching section. In the maximum F/B version, they are designed for about 25 Ω of capacitive reactance, for use with a beta or hairpin match. The very wide-bandwidth design requires direct connection to a 50-Ω coaxial cable. In each case, the design places the antenna connections at the split-driver terminals.

Fig 1 gives us several alternatives for assembling the driver to the boom. Although the graphic shows alternatives A and B, we actually have 4 major combinations, plus any number of adaptations you may create based on available materials. Alternative A uses a small plate (Plexiglas, polycarbonate or acrylic) with the drive tube anchored to it. A non-conductive insert (fiberglass, CPVC, or similar) aligns the two halves of the driver and strengthens the tube against crushing when the bolts are

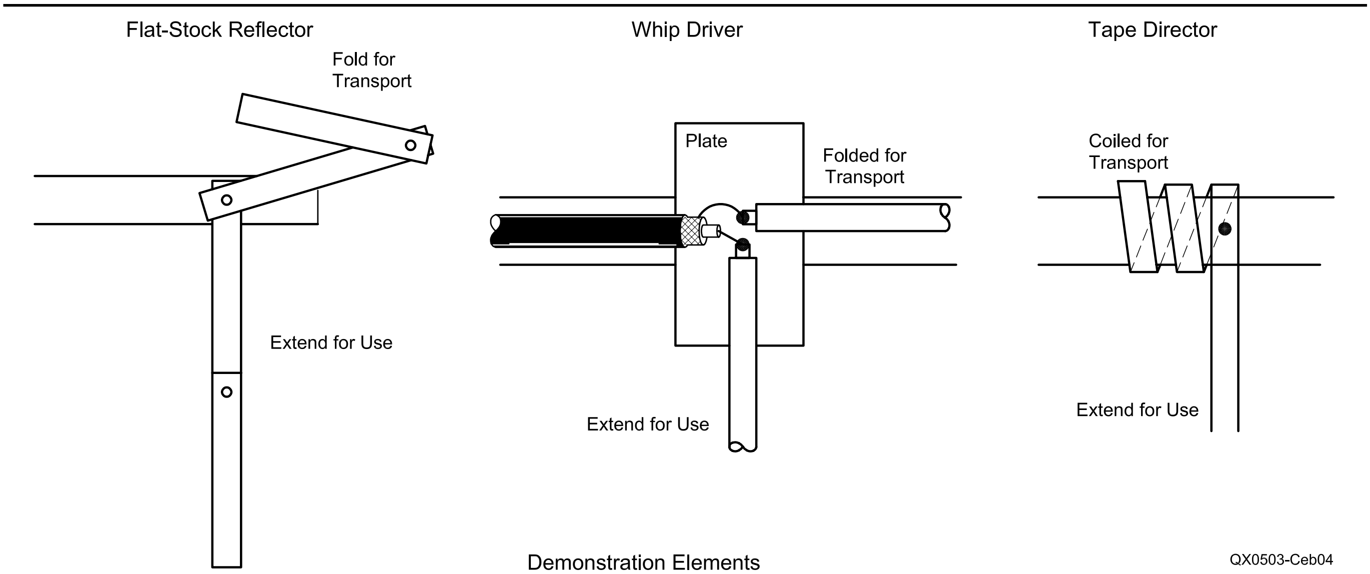

Fig 5—Elements used in the demonstration field antenna, shown extended for use and folded or coiled for transport.

tightened. Even if the parasitic elements pass through the boom, the very slight misalignment of elements relative to their ideal plane will create no operational difficulties.

Alternative A shows a direct connection to coaxial cable without the use of a connector. This technique allows you to use a cable length of your choosing to dress the lead to the boom or to connect the feedline directly to the antenna. The open end of the coax—with its implicit ring connectors for attachment to the driver terminals—requires sealing. Plasti-Dip and similar products have proven reliable in this service, and they are less bulky than coax sealant and tape. Indeed, over the years, I have come to prefer this system.

If you prefer to use a connector, alternative B shows a simple mounting bracket that not only holds the connector, but also extends to one side of the element. Light L stock (1-inch wide by 1/16-inch-thick) can support the BNC connector shown in the sketch or a standard UHF connector. Do all drilling before trimming the L stock to final size for easier handling while machining the required holes. The sketch also shows the bracket with rounded upper edges. It's easiest to arrive at this shape using a disk sander. (Do not use a grinder designed for steel.) Clearly, you may adapt the connector bracket for use with the plate assembly in alternative A. Likewise, you may use direct coaxial-cable connections to each side of the driver that passes through the boom.

The through-boom driver shown in the sketch uses 1/2-inch tubing with a 3/8-inch fiberglass or similar rod or tube that actually passes through the boom. There is a limit for minimum effective

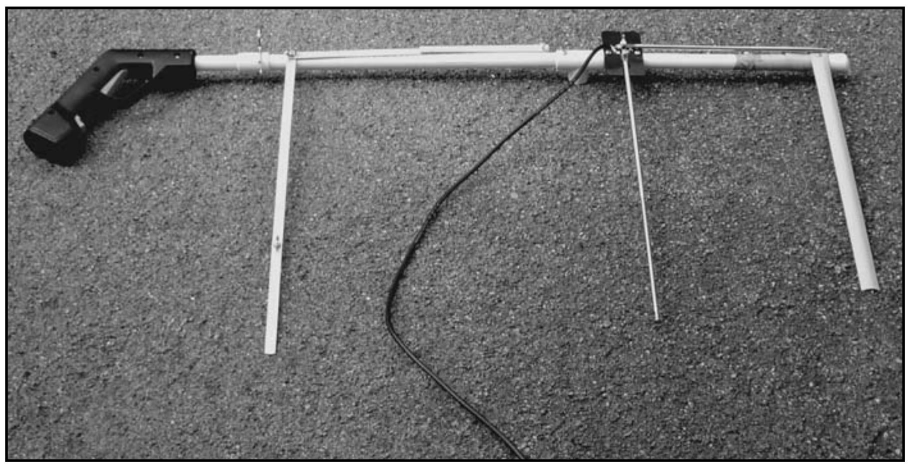

Fig 6—A working three element Yagi for 2-meters demonstrating the use of a variety of element materials and a special handle for field exercises.

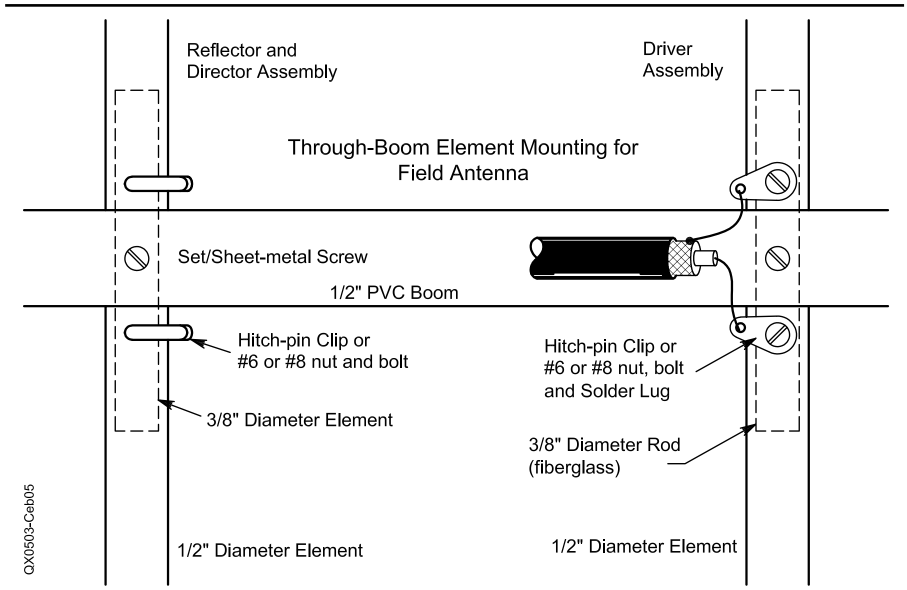

Fig 7—A method of using tubular elements for a field antenna for disassembly into a compact transport package.

insulating rod size to support the driver, since a hardware hole will pass through both the element and the rod. You may use this larger driver with any of the dimension sets in Part 1, using thinner material for parasitic element diameters. The "fat" driver will require a reduction in length to bring the beam to its proper SWR curve, but it will not otherwise affect performance.

There is no good reason why the parasitic elements for a simple 3-element Yagi should not pass though the PVC boom. There are three (or more) different suitable systems for holding the elements in place. See Fig 2 . The top option works well with rods and tubes at least $^3/_{16}$" in diameter. It uses minimum-size hitch-pin clips on either side of the boom to secure the element. Some builders use C- or E-clips, relying upon their spring action to hold the element in place against the boom. All such hardware should be non-corrosive, however. The middle sketch shows the use of a setscrew. If you under-size the hole, most stainless-steel bolts (#6 to #10, depending on the element and boom diameters) will self-tap the material for a firm seating. You may install a nut below the setscrew head and tighten it to the boom after securing the element. For larger tube sizes, you may use the last option, a sheet-metal screw that penetrates both the PVC boom and a hole at the center of the element.

Both the hitch pin and the setscrew mounts benefit from a small bit of filing. The hitch-pin clips require holes through the round rods or tubes. A small jeweler's file can create a flat spot no greater than about $^3/_{32}$-inch diameter without weakening the element. The flat spot eases the drilling if you do not have a drill press. A similar flat spot at the element center gives a setscrew a good surface for the bite necessary to secure the element.

The techniques suggested here have resulted in numerous solid Yagis with up to six elements and with boom lengths up to five feet or more. (As boom length increases, rear mounting becomes less suitable.) All of the designs in Part 1 use booms shorter than 30 inches and the elements are suitable for the hardware-store materials noted along the way. *Do not* use hardware-store materials for the elements themselves. High-grade 6063 or 6061 aluminum tubing or rod is best for long-term installations.

Construction of Field Antennas

Antennas for the field call for some special techniques. You may take an antenna built according to the preceding suggestions into the field; however, that antenna has a permanent size, about 40×30 inches. Hence, it is a bit ungainly for transport in an auto trunk or other confined space. Hallmarks of a good field antenna are that it stores compactly for transport and is ready for use with minimal field assembly. The ideal situation is one that requires no tools to transform the transport package into a working antenna.

You can achieve these goals in many ways. There are as many ways to successfully build a good field antenna as there are alternative materials for antenna elements. To demonstrate what is possible (in local talks for clubs and other functions), I created a hybrid Yagi using separate techniques for each of the three elements. The design uses the very wide-band design as its basis, although there is no reason not to use any of the other versions. Since you will likely use a single material best suited to your operating goals, your own field antenna will pick the design that is most apt to those goals.

My hybrid begins with a length of $^1/_2$ inch nominal PVC pipe. The actual outer diameter of this pipe is a little over $^7/_8$ inch. I placed a T fitting just behind the driver position. The fitting is aligned for horizontal mounting of the antenna on a boom, using PVC screw fittings to increase the boom diameter until it matches the mast on which the antenna will sit. For many field uses, you need not cement the fitting in place. Press-fitting the T will provide a secure and durable connection between the boom sections for most field operations. You may use a sheet-metal screw to secure the unglued side of the T , but that requires a screwdriver. The forward end of the boom has a cap to keep out bugs and debris.

Fig 3 shows the general arrangement of the boom, along with the special rear section. At the rear end of the boom, there is an in-line coupler. It attaches to the forward boom sections with a large hitch-pin clip. I drilled the end of the boom with two holes at 90° angles. Hence, I can change the orientation of the antenna from horizontal to vertical and back again simply by removing the hitch-pin clip, twisting the boom, and re-installing the clip. You can use the same system with a rear boom-to-mast attachment system rather than the funny handle shown in the sketch.

Most fox-hunting antennas that I have seen use rubber hand-grips in line with the antenna. These grips are most suitable to point Yagis at satellites, but are not ergonomically suitable for aiming the antenna straight ahead. Therefore, I took the pistol grip handle from a defunct electric weed cutter and replaced the steel tube with a short length of PVC. Since the in-line coupler bears the lever-force of the entire antenna ahead of it, I cemented the coupler to the handle-end pipe. Fig 6 reveals that I left the trigger in place, since it is smooth, while the bare opening without it has sharper, less-comfortable edges.

The photo also shows the demonstration elements. The driver uses collapsible whips taken from TV "rabbit ears." The reflector uses flat stock, and the director uses a length of steel measuring tape. Both Figs 5 and 6 show the elements, each with one half ready for use and the other half stored for transport.

The $^1/_2$-inch-wide flat-stock reflector would not store well if we used a full half-element length on each side of the boom. Instead, I used #8 bolts and wing nuts for the outer section so that it could fold back on itself and fit entirely behind the driven element during storage. For the reflector, I used a #10 wing nut and bolt that passes entirely through the boom to secure the reflector in place for both use and transport. Thin ($^1/_{16}$ inch) stock seemed a bit flimsy initially but has held up well during use.

The driver whips retract for storage and extend for use. Setting them requires a tape measure or other measuring strip to get the correct length each side of center. (I wrote the measurement on the boom.) #8 hardware secures the position of the driver. I ground shallow grooves with a rotary tool in the small Plexiglas plate so that the driver stays in either the use or storage position once I tighten the wing nuts. The feed point bolts also hold the length of coax that I have devoted to the antenna.

The director uses a length of steel measuring tape. A single sheet-metal screw fastens the element to the boom, although I placed a few thin washers between the tape and the boom to maintain the tape curvature. The tape does not require re-positioning for storage. Instead, wrap the tape around the boom and secure it with a piece of duct tape or equivalent. In fact, you can use sections of cardboard tubing from a roll of paper towels to slide over the coiled elements. In either case, guard your face when opening the element; it will spring to position very rapidly.

You can store the entire antenna in a three-foot-long storage unit that is only about four inches wide (plus the handle). Alternatively, you can remove the handle and break the boom at the T for more compact storage. Fig 4 shows the pieces in full-storage mode.

Does the hybrid field antenna work? Since I selected materials for the director and reflector that are very close equivalents to $^1/_2$-inch-diameter round elements. I used the spacing for the very wide-band design for those elements. Then, I simply adjusted the driver length to give the 50-Ω impedance curve for that design. I used the two fattest sections of the whip, and the resulting length was not much longer than the value shown in the Part-1 tables, about 19.25 inches on each side of center.

For field use, especially if you plan to use the antenna in a hand-held activity like fox hunting, you will need to determine the correct driver length for a normal use position well in advance of going into the field. You may also discover that for different orientations and heights above ground, the required driver lengths may differ.

If you prefer the design security of using tubular elements, Fig 7 shows one method of achieving a compact storage package and a full-size array of the type that you choose. You may construct the boom in the same manner as for the alternative element materials, using a three-piece break-down for transport. However, the element positions will have stubs protruding about 1.5 inches on each side of the boom. The reflector and director stubs will be $^3/_8$-inch-diameter aluminum tubes, while the driver stub will be a length of $^3/_8$-inch-diameter fiberglass or other non-conductive rod or tube. Secure each stub through the boom with a sheet-metal screw as a permanent mounting.

Although you may believe in thinner rods for field use, $^1/_2$-inch 6063-T832 tubing weighs very little more than $^3/_{16}$" solid rod. Table 1 provides comparative weights of rods and tubes used in common amateur antenna construction. The material is drawn from the Web site maintained by Texas Towers. It applies to aluminum tubing with a wall thickness of 0.058." Alternative materials with thicker or thinner walls will, of course, change the weight per foot. For the project at hand, the elements are between 3.0 and 3.3 feet long so totaling the element weight is easy math.

In most cases, the boom will outweigh the sum of the elements and their hardware by a good margin. Using larger materials adds little to the antenna weight, but allows fastening with nuts and bolts. The half-elements use $^1/_2$-inch-diameter tubes, none of which is longer than about 20 inches. You may attach the reflector and director outer element halves to the stubs with #6 or #8 nuts and bolts or with hitch-pin clips. (Do not exceed #8 hardware, or the hole may weaken the stub.) The driver uses nuts and bolts, plus solder lugs, to attach the element halves and form connection points for the coax and any matching device (such as a hairpin).

Table 1—Aluminum 6061-T6 Rod and 6063-T832 Tube Weights

Type	*Outside Diameter*	*Weight/12"*
Rod	0.125"	0.015 lbs
Rod	0.1875	0.032
Rod	0.25	0.058
Tube	0.375	0.044
Tube	0.50	0.095
Tube	0.625	0.104
Tube	0.75	0.127
Tube	0.875	0.150
Tube	1.00	0.202

The advantages of using aluminum tubes as elements for the field antenna include general strength. Overgrown fox-hunting field sites that can snag the elements may still test the antenna's sturdiness. However, use of the antenna at an emergency or Field-Day site for FM or similar applications is unlikely to encounter such tests. The disadvantages include the need for small hardware to assemble the antenna. If you opt for this type of field antenna, be sure that the transport package includes both extra hardware to replace pieces lost in the grass and tools for assembly. A dedicated screwdriver and nut-driver are essential.

In the end, the decisions concerning the methods of construction will rest upon your intended uses, the availability of materials, and your own assessment of your construction skills. However, somewhere in this collection of ideas—and other ideas that you develop—will be a Yagi that you can build yourself.

Conclusion

The hybrid demonstration antenna is simply a potpourri of ideas that you can adapt to both field and long-term antennas for 2 meters. In fact, we have surveyed a wide variety of factors that go into a home-brew utility Yagi for this band.

1. We examined three different Yagi designs: a high-gain version, a maximum front-to-back ratio model, and a very wide-band unit.
2. We also saw 2 ways of matching a 25-Ω driver impedance to a 50-Ω coax line using a $^1/_4$-wavelength matching section with a resonant driver and a potential hairpin match with a driver that is capacitively reactive.
3. We explored a variety of alternative materials that builders of field antennas use instead of rods and tubes, and we measured them at 146 MHz to find their nearest round equivalents.
4. Finally, we explored various ideas for constructing both long-term and field antennas using common materials from hardware outlets.

Now, you have no excuse for not building your own 3-element 2-meter Yagi, whatever your operating goals. In fact, I would expect you to have some building ideas that yield an antenna better than any of the samples that you have seen in these notes. Those ideas increase the number of options we have. The greater the number of options, the closer that we can match our antenna to the job for which we need it. □□

Distortion and Noise in OFDM Systems

A brief look at performance issues in medium- and high-speed systems.

By Doug Smith, KF6DX

Preface

In this essay, orthogonal frequency-division multiplexed (OFDM) signals are defined as those comprising a plurality of subcarriers, evenly spaced in frequency and angle-modulated, amplitude-modulated, or both. Here, orthogonal means mutually exclusive rather than at right angles. The discussion pertains to all Amateur Radio digital voice systems known to me and to many other high-speed digital schemes—some being contemplated and some already deployed—throughout the communications world.

My goals are: 1) to explain certain deleterious effects caused by amplitude and phase distortion in radio transceivers and propagation media, 2) to mention a few methods for combating those effects, and 3) to encourage further intercourse on the topic. The concepts presented are not new, but they delineate some of the challenges still faced by equipment designers. Your comments are most welcome.

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OFDM Basics

This section is a simplified explanation of how and why OFDM works. Readers familiar with OFDM may skip to the next section.

To get information from one point to another via radio, we have but three physical properties of radio waves to exploit: amplitude, phase and frequency. Really, phase and frequency are related in that FM and PM are both forms of angle modulation.

A high-speed digital modulation scheme using a single modulated subcarrier tends to produce a large occupied bandwidth. Multiple subcarriers may be employed to reduce bandwidth while maintaining high throughput capacity. Such a scheme also mitigates the effects of some of the distortions outlined below. The symbol (baud) rate of each subcarrier is chosen with respect to anticipated channel impulse responses.

In a typical 3-kHz bandwidth system, dozens of individual subcarriers may be used. Each is modulated digitally, usually using some form of phase-shift keying (PSK) or quadrature amplitude modulation (QAM); each carries a fraction of the total throughput burden. For instance, a system having 36 subcarriers, each modulated with quadrature PSK (QPSK) at 50 baud, produces a raw throughput rate of (36)(2)(50)=3600 bps. The factor of two indicates that at each symbol time, two bits of information are conveyed by QPSK. The actual rate of useful throughput is usually reduced by the needs for synchronization and error-detection and -correction. Those requirements use up some of the total capacity.

In such a system, the center frequencies of each subcarrier are spaced evenly. That is, the frequency difference between sub-channels is constant. The designer would use the entire available bandwidth, perhaps selecting a 75-Hz separation. Sub-channel centers in Hz might then be 300, 375, 450, 525, 600 and so on. See Fig 1.

Limitations of OFDM Systems

Peak-To-Average Ratio

One issue in OFDM systems has to do with peak-to-average ratio, usually called crest factor. It turns out that combining multiple subcarriers as described may result in a signal producing occasional peak values far above its average or RMS value. That is a disadvantage because it limits the total energy that may be transmitted in peak-limited transmitters. Every transmitter is, after all, peak-limited in some way.

Several methods have been used to reduce crest factor. Those include clipping—but sophisticated coding techniques are now employed that code the data to prevent large crest factors. Such techniques are generally referred to as *coded* OFDM (COFDM). They do their jobs by exploiting redundancy among sub-channels.[1] COFDM may also refer to coding to employ error-detection and -correction systems. The latter type of coding naturally reduces channel capacity in the absence of errors.

[1]Notes appear on page 59.

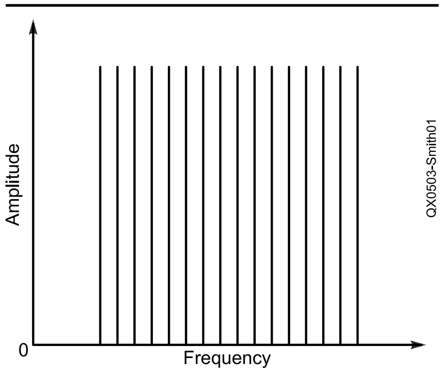

Fig 1—A typical 16-tone OFDM spectrum, unmodulated.

Intermodulation Distortion

In both transmitters and receivers, intermodulation distortion (IMD) forms a distinct limitation for OFDM systems. Because of cube-law nonlinearities, any two adjacent subcarriers produce IMD products by their very presence that fall near the center of adjacent sub-channels—right where they do the most damage. Two or more nonadjacent subcarriers may or may not produce IMD that also falls in another sub-channel.

QAM involves both amplitude and phase modulation. Each QAM subcarrier therefore produces IMD on its own because its amplitude is continually changing.

IMD shows up as noise in demodulated data from any particular sub-channel, since the data in one sub-channel usually do not correlate with those in another sub-channel. Such noise shows up as a "noise cloud" around the points in a constellation display, often used to show coherence of received signals, and in signal-to-noise ratio (SNR) calculations performed on those signals. Fig 2 shows a so-called constellation display. Each cloud represents one subchannel.

Using OFDM signals, typical Amateur Radio HF transceivers employing class-AB power amplifiers (PAs) produce maximum SNRs in the range 20-25 dB because of IMD effects. That, in turn, limits the bit error rates (BER) of OFDM systems. Automatic level control (ALC) may also produce deleterious effects.

The solution involves ensuring the highest degree of linearity possible in radio transceivers. That is most often achieved at power levels well under the maximum rating of transmitters. Class-A PAs may be at a distinct advantage here for AM or SSB. Digital signal processing (DSP) has been used to adaptively correct for linearity errors.

Pre-distortion of exciter waveforms is a useful method for combating IMD; however, protagonists have found that bandwidths of up to five times the corrected bandwidth must be used for pre-distortion. While the technique has shown success in reducing low-order IMD, it often increases high-order IMD. Post-correction filtering is therefore often required. IMD in receivers may similarly be corrected. Pre-distortion requires knowledge of both amplitude and phase responses.

Phase and Group-Delay Distortion

Where digital angle modulation is involved, the relative phases of signals

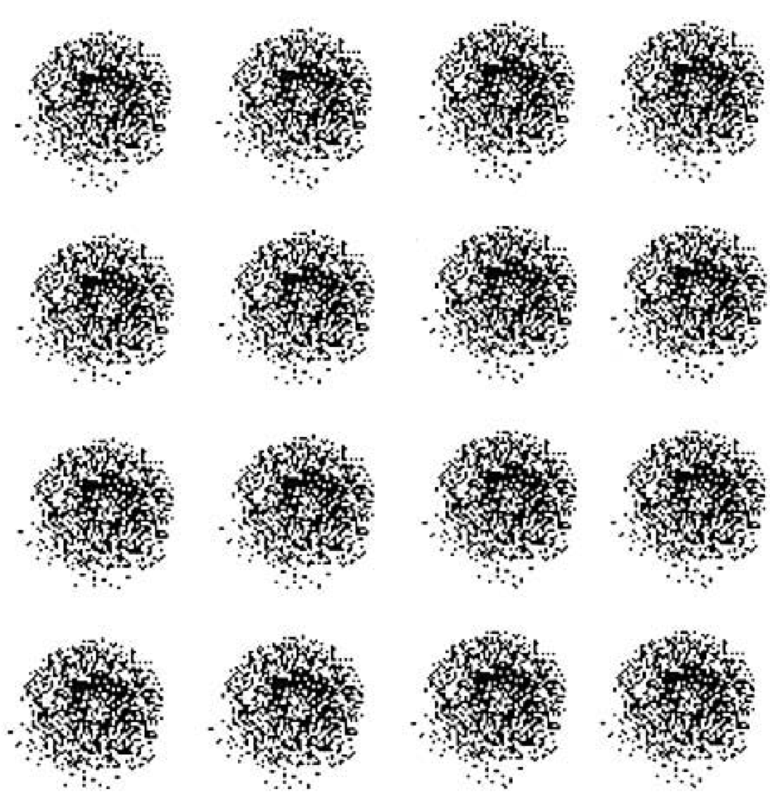

Fig 2—A 16-tone constellation display, showing somewhat noisy or distorted conditions.

are critical. When phase shift is not directly proportional to frequency, group delay is not constant. That may mean high frequencies propagate from transmitter to receiver faster than low frequencies, or vice versa. Data in one sub-channel may no longer correlate with those in another and some of the bits get out of time alignment, obviously degrading performance.

Adaptive DSP methods can deal with group-delay distortion quite nicely by equalizing the channel to achieve relatively flat group-delay characteristics.[2] Adaptive equalizers may use a training sequence or may adjust themselves "on the fly" to do their jobs. Such subsystems have been employed in telephone modems for many years and they have found their way into high-speed radio systems.

Those that continually adjust themselves are to be preferred over those using training sequences because data loss is minimized. Group delays that are continuously changing require that approach. Such is the situation on an HF radio path—quite unlike the typical telephone line, which does not normally change much over short time frames. Short-term phase instability, though, remains an issue.

Noise and Interference

Phase Noise

Noise is the perennial enemy of the communicator. It is evidently present at all frequencies and it confuses digital demodulators according to well-known models. One source of noise over which digital communications systems designers have some control is called phase noise.

Phase noise is the unwanted phase modulation of frequency-control elements in a transceiver by thermal noise.[3] Thermal noise is associated with the random (Brownian) motion of atomic and subatomic particles caused by heat. Many modern frequency-synthesis techniques have made great strides in reducing the phase noise of frequency sources, but such noise is still in play for the vast majority of systems.

In the case of closely spaced subcarriers in an OFDM system, each subcarrier mixes with a local oscillator (LO) in a transmitter to produce RF. In so doing, it carries with it the phase noise of the LO. That noise resides adjacent to the subcarrier's RF position in the transmitted wave. Some of it overlaps adjacent and nearby sub-channels, degrading their maximum SNR. LO phase noise in a receiver worsens the problem because uncorrelated noise powers add.

When subcarriers are spaced as closely as 75 Hz, the effect can be quite significant. 50- or 60-Hz power-line frequencies may be present. Careful design and measurement are necessary to mitigate the effects of phase noise.

Interference

The presence of uncorrelated signals obviously degrades the performance of high-speed digital communications systems. Interference is often beyond the control of the operator. Determination of the interference immunity of any particular modulation format is a complex session in engineering statistics. Any theory about performance must be supported by empirical evidence.

Amateur Radio digital voice systems are currently allowed in the U.S. only on the phone bands. Interference thus is limited mainly to that from other analog phone or image signals. It remains to be seen what level can be tolerated with present systems, both for those suffering the interference and those who interfere. Error detection and correction is central to the issue.

Error Detection and Correction in Digital Systems

In the face of the above-mentioned obstacles, error detection and correction is strictly necessary for high-speed digital communications over radio. An OFDM signal transmitted with a fairly low SNR without error correction cannot stand much degradation on its way to a receiver and its demodulator before being hopelessly corrupted.

Many advanced algorithms have been developed that produce robust performance with minimal overhead. Overhead refers to the additional data that must be sent to achieve error detection and correction. The most rugged and efficient algorithms increase transmitted data by about 50%.

Acknowledgements

Many thanks to the following individuals for their helpful comments and suggestions on this article: Cédric Demeure, Thales Communications; Ulrich Rohde, Synergy Microwave; Charles Brain and George Bednikoff, ARRL Digital Voice Working Group.

Notes

[1]G. A. Davidson, "Digital Audio Coding: Dolby AC-3," in *The Digital Signal Processing Handbook*, V. K. Madisetti and D. B. Williams, Eds. 1998, CRC Press LLC, Boca Raton, FL.

[2]D. Smith, KF6DX, "Deconvolution in Communications Systems," *QEX/Communications Quarterly*, Sep/Oct 2001, ARRL, Newington, CT; also appears in *20th ARRL and TAPR Digital Communications Conference Proceedings*, Sept 2001, ARRL.

[3]D. Stockton, GM4ZNX, "AC/RF Sources (Oscillators and Synthesizers)," in *The ARRL Handbook for Radio Amateurs*, 79th Ed.; D. G. Reed, Ed., 2001, ARRL. □□

Letters to the Editor

A Software-Defined Radio for Digital Communications (Nov/Dec 2004)

John,

First, let me compliment you on your excellent article in *QEX*. I want to also thank you for your other contributions to the Amateur Radio art. I look forward to your future contributions.

I do want to bring to your attention an erroneous assumption. In your reference to my previous *QEX* articles, you state, "Unfortunately, the poor opposite-sideband suppression, which rarely exceeds 50 dB, wastes dynamic range." I can understand this comment based on a historical analog perspective. With DSP we have the ability, for free, to correct for phase and amplitude imbalances in software. Theoretically, this could be done to perfection in the frequency domain.

We are currently using time-domain correction (because it was easy), which nulls the opposite sideband into the noise floor on a single frequency. In practice, it gives better than 70 dB of suppression across the passband. We have even automatically nulled the image.

We have incorporated a pulse generator into the hardware that we plan to use soon for automatic frequency-domain equalization using the actual measured impulse response of the receiver. This should give us exceptional image rejection with our 10-Hz bin size in the frequency domain.

Thanks again and keep up the great work.—*73, Gerald Youngblood, AC5OG, FlexRadio Systems, 8900 Marybank Dr, Austin, TX, 78750;* **gerald@flex-radio.com**

Gerald,

70 dB of suppression is a significant improvement for direct-conversion receivers for the 160- to 6-m amateur bands, and I hope that this improvement can be extended. I was designing for a different frequency range, however, and I have not yet seen more than 50 dB of image rejection quoted for a UHF direct-conversion receiver. The technology required to support both narrow- and wide-band modes, including satellite communication, on the 160-m to 13-cm bands differs from that used at lower frequencies. The second half of the article "HSMM Radio Equipment" in the same issue of *QEX* describes the UHF OFDM modem that is the initial application.

73, John Stephensen, KD6OZH, 153 Gretna Green Way, Los Angeles, CA, 90049; **kd6ozh@amsat.org**

Hi John,

Thanks for your response. The trick is that you don't much care how bad the analog section is. You can correct for any phase and amplitude imbalance between I and Q in the DSP code. In fact, our analog circuit has only around 40 dB of image suppression. We get the extra 30 dB or so in software. It takes only a couple of lines of code in the time domain. It is more complicated in the frequency domain, but you can approach near perfect rejection when you know the impulse response of the receiver.—*73, Gerald*

Hi Gerald,

The process sounds similar to compensating for distortion introduced by propagation effects in the radio channel for digital communication. The limitations are how often you can measure these effects versus how fast they change and the overall dynamic range of the ADC. For narrow bandwidths, sigma-delta ADCs with over 120 dB of dynamic range are available, and it will be interesting to see how far the image rejection can be pushed toward that limit.

It will also be interesting to see whether the final impediment for high-performance DC receivers at VHF and UHF can be removed in the future. The three main issues with DC receivers have been microphonics, limited image rejection and 1/f noise. Microphonics are remedied by an RF amplifier with very low reverse gain, and image rejection is now being improved. 1/f noise generated in the mixer remains a problem, as it creates a band of frequencies near 0 Hz that is unusable. At low frequencies, where atmospheric noise is high, the problem is minimized. As the frequency increases, increasing RF gain compensates to some extent, but increases signal levels at the mixer and lowers the dynamic range of the receiver.

At low radio frequencies, transistors or diodes can be used that confine 1/f noise to frequencies below 100-1000 Hz, depending on the device used, but very low-1/f-noise devices tend to create mixers with low IM intercepts. As the LO frequency increases, devices with higher transition frequencies must be used in the mixer to increase switching speed and limit IM products. The 1/f region expands to 10 MHz in microwave devices. I haven't seen much improvement in 1/f noise in the past 20 years, so it remains a problem for narrow-bandwidth high-frequency DC receivers. Superhets bypass the problem with an IF above the 1/f frequency.—*73, John, KD6OZH*

John,

From my experience, the issue is more phase and amplitude balance versus frequency, not linearity of the ADC over the dynamic range. At a specific frequency, you can null the image into the noise floor of the receiver with no problem in the time domain. We just need to implement frequency-domain correction to get rid of the changes by frequency. That we will do soon.

The SDR-1000 is really a single-conversion receiver, where the first IF is at dc, and the second IF is at 11 kHz. 1/f noise is not an issue because we stay away from anything below 1 kHz. Agreed, the SDR-1000 is a direct-conversion superhet.—*73, Gerald*

John,

I am really looking forward to getting your widget. We will be using it to build a C-band-to-C-band transponder prototype. It does appear to be at the top of its class.

Your note contains one inaccuracy. Gerald's receiver is not direct-conversion. Including the software I have written it is a dual-conversion superhet. With a high end 24-bit sound card, it has over 100 dB of dynamic range, the MDS is less than –140 dBm, and the noise figure, with the additional preamplifier modification board, is 3 dB through 60 MHz. It has a BDR greater than 130 dB (Gerald publishes greater than 90 dB, but he did that test with a 16-bit sound card). The secret to the quadrature-sampling detector (QSD) is that it is the "convolution" of a sample-and-hold mixer and roofing filter all in one. It is pretty neat. The only problem we have with the hardware is solvable in software. We do have spurs that are caused by the DDS chip. There is not a single thing we can do about them—except avoid them.

We are doing avoidance of the AM/PM spurs that most people know about. This is because—in software—the second IF frequency is variable. The DDS steps between discrete frequencies and the remaining tuning is done in software. The discrete frequencies are those that minimize DDS spurs. I believe they can be completely avoided in the passband by using more of the range inside the roofing filter.

All spurs could be avoided by having a tracking PLL with the reference driven by the DDS. That comes at a price of increased phase noise, so this must be carefully designed.

The front end has too much gain for low bands on HF but even so, it has anIP3 of +29 dBm with a high-end sound card. Gerald has provided an attenuator (10 dB) and the IP3 is then a whopping +39 dBm from 160 through 6 meters.

Comparing your new SDR to Gerald's is comparing apples to oranges. They are ideally aimed at different audiences. Your offering can be used as a narrow-band HF transceiver, but it will never have the dynamic range, BDR and so on of the SDR-1000. It is impossible to have enough filter/downsampler stages to get sufficient processing gain to overcome the naturally huge dynamic range and linearity of the QSD. On the other hand, to get this performance, we cannot hope to use the QSD on wide-bandwidth signals; in that arena. I think your tool will be the best receiver we can afford to do the job, and possibly the best tool in existence. With your widget, I can scan an entire Amateur Radio band on HF and classify, detect, demodulate, and display information on every strong signal in the band. That would be the end of my usage of it on an HF band. To prosecute the detected signal, I would tune the SDR-1000 to it to get the IC-781 or Ten-Tec Orion receiver performance.

On wide-bandwidth signals such as C-C Rider for AMSAT Eagle, we have *one* choice to prototype it and experiment: *your SDR*. I very much look forward to having it.—*73, Robert McGwier, N4HY, 64 Brooktree Rd, East Windsor, NJ, 08520-2438;* **rwmcgwier@comcast.net**.

Robert,

I used "direct-conversion" as it has been used in *QST* and *QEX,* when image rejection is performed in software. In commercial circles, people have been using the term "low-IF" where digital signal processing is used instead of an RF bandpass filter to eliminate the image frequency. "Single-conversion low-IF transceiver" may be a description that best describes Gerald's hardware. The method of avoiding DDS spurs is innovative and reduces the cost of the radio by eliminating the PLL and multiple VCOs that would otherwise be required. In amateur circles, the dynamic range is only exceeded by more complex high-end Superhets, such as the W7AAZ/W4ZCB/G3SBI Triad receiver.

You are correct that comparing the DCP-1 to the SDR-1000 is comparing apples to oranges. The SDR-1000 is the high-speed equivalent of a sound card with an on-board DSP. It can process RF directly and filter it with a digital down-converter instantiated in the FPGA, similar to the KD7O design. However, I am using it as an IF signal processor and I pointed out in my article that a superhet receiver architecture with a mixer and filter ahead of the ADC provides more dynamic range than direct digitization of RF. The SDR-1000 does the same for a sound card. *—73, John, KD6OZH*

Doug,

I just started playing with Xilinx' new version of ISE with speed files for the production hardware. The difference is amazing. The carry logic doubles in speed and even the multipliers are faster.

Change the text below Fig 22. 147.456 MIPS is now 258.048 MIPS. This means 1 billion filter taps per second and 64 million radix-2 butterflies per second. Also, the caption of Fig 22 is incorrect. Only 36 k of RAM is used.*—73, John, KD6OZH*

RF Power Amplifier Output Impedance Revisited (Jan/Feb 2005)

Dear Sir:

I found Robert Craiglow's article extremely interesting. Unfortunately, an error in Appendix C has muddied the results. After correcting this error, the article demonstrates the following:

1. An RF power amplifier (PA) can be represented by a Thevenin equivalent source with constant parameters. (In real life, a good approximation.)

2. As a corollary, viewed from the output terminals, a PA is a linear device.

3. As a corollary, the PA is conjugately matched with its load when the power output-load relation is at a maximum.

4. Features of a PA may prevent loading to a maximum in power output—a "wall" may be encountered. In that case, a conjugate match is not obtained.

5. The load-variation method of measuring the source resistance of a PA is exactly equivalent to the signal-injection method as can be demonstrated by the compensation theorem.

There is a simple demonstration that, on a *differential basis*, any source having a maximum power output at some load impedance is conjugately matched to that load, regardless of the linearity of the source. This demonstration is based on obtaining the value of dI/dR_L by differentiating the power output with respect to R_L and comparing that value with the derivative of current from a Thevenin equivalent source.

The error in Appendix C occurs in Eq C3. The fundamental value of i_p, i_{p1}, is given by the integral:

$$i_{p1} = \frac{1}{\pi}\int_{-\pi}^{\pi}(\Delta E\cos\phi)\cos\phi\, d\phi \quad \text{(Eq 1)}$$

This integral may be separated into three integrals,

$$\frac{\Delta E_p}{\pi}\int_{-\pi}^{\frac{\theta}{2}}\left(\cos\phi - \cos\frac{\theta}{2}\right)\cos\phi\, d\phi +$$

$$\frac{1}{\pi}\int_{-\frac{\theta}{2}}^{\frac{\theta}{2}}\left(\cos\phi - \cos\frac{\theta}{2}\right)\cos\phi\, d\phi +$$

$$\frac{1}{\pi}\int_{-\frac{\theta}{2}}^{\pi}\left(\cos\phi - \cos\frac{\theta}{2}\right)\cos\phi\, d\phi \quad \text{(Eq 2)}$$

The functionality of the integrand results from the fact that the tube does not conduct until the signal voltage reaches $\Delta E_p\cos(\theta/2)$. Notice that the grid is being driven by $E_c+E_g\cos\phi$. The effective test-signal voltage driving the tube is thus $\Delta E_p[\cos\phi - \cos(\theta/2)]$. See Mr. Craiglow's Eq A3.

The first and last of the three integrals contribute nothing to the current as the tube is cut off. Therefore:

$$\Delta I_{p1} = \frac{\Delta E_p}{\pi r_p}\int_{-\frac{\theta}{2}}^{\frac{\theta}{2}}\left(\cos\phi - \cos\frac{\theta}{2}\right)\cos\phi\, d\phi$$

$$= \Delta E_p \frac{\theta - \sin\theta}{2\pi r_p} \quad \text{(Eq 3)}$$

It follows that:

$$R_s = \frac{2\pi r_p}{\theta - \sin\theta} \quad \text{(Eq 4)}$$

which agrees with all other calculations of the source impedance.—*Albert*

E. Weller, Jr, 1325 Cambridge Blvd, Columbus, OH, 43212; **aeweller@att.net.**

Doug,

Bob Craiglow's sincere and well-intentioned effort to help resolve the conjugate-match argument missed its objective because of some poor choices for analysis. The conjugate-match debate has been limited to linear RF power amplifiers since they alone are used for SSB.

A theoretically ideal linear class-B amplifier is a good choice because it allows analysis with simple equations. The conduction angle must be 180° for linear operation. Therefore, all discussion regarding other conduction angles serves no useful purpose. The only plate-current waveform is a half sine wave. It is that shape for any signal amplitude in the linear region.

In the theoretical class-B case, the plate resistance R_p is the same over the entire conduction area. R_s is then always twice the value of R_p because the tube only conducts half the time. More precisely, it is the ratio of the amplitude of the half-sine-wave plate current pulse to the peak amplitude of its fundamental component.

An equation in Terman's *Radio Engineering Handbook* shows that the relationship between the tube's μ and its plate resistance is: $g_m R_p = \mu$. It turns out that triodes with a medium μ of roughly 40 have a low enough R_p so that a certain load resistance R_L will provide a conjugate match. Such tubes are better adapted to class-C operation—for example, the 812A, which has a μ of 29.

On the other hand, the type 811A tube, a similarly sized tube designed for linear class-AB modulator use, has a high μ of 160. This characteristic makes it an excellent tube for linear RF power amplifiers also. The high μ causes the plate resistance to be approximately five times as high as that of the 812A tube.

To represent such a high-μ tube in Bob's Fig 2, the grid-voltage lines would have only one-fifth as much slope. This would have provided a much more convincing argument to demonstrate that a conjugate match does not establish correct tube operation.

The efficiency for theoretical class-B operation is: $\eta = (\pi/4)(e_p/E_B)$, where e_p is the RF peak plate voltage, and E_B is dc plate voltage. The RF plate voltage must remain to the right of the diode line since the tube is nonlinear on the left side. Notice that R_p and R_s are not part of this efficiency equation, therefore neither directly determines efficiency or plate dissipation, as Bob pointed out.

Introducing the 3/2-power tube characteristic contributes nothing toward resolving the debate, either. This characteristic is inherently nonlinear. In addition, constant-current characteristic curves must be used with a straight load line. Manufacturers have found ways to modify tube geometry to approach ideal class-AB operation.

Regarding my measurement method of using a small test signal 10 kHz away from the amplified signal, Bob reached an incorrect conclusion. The voltage amplitude of the small test signal is only on the order of 2% of the amplified signal's amplitude, thus the principle of superposition can still be used.

In the last part of his conclusion, Bob clearly doesn't understand how the low R_s of the class-C amplifier in an AM broadcast transmitter is used to broaden transmitter-antenna bandwidth. It is not by loading. The class-C tube acts substantially as a voltage generator. The electrical length of the circuitry and coax that couple the tube to the antenna is adjusted to provide an odd multiple of 90° phase delay. This creates an impedance inversion; therefore, the antenna is effectively driven by a current source, which produces a flatter frequency response. That is explained in the *National Association of Broadcasting's Engineering Handbook*.

In the Jul/Aug 2001 *QEX*, I showed how to compute R_s from a set of constant-current tube curves. Unfortunately, I no longer have access to lab facilities where I could take measurements to show equal results, thereby validating both methods.—*73, Warren Bruene, W5OLY, 7805 Chattington Dr, Dallas, TX 75248-2492*

Doug:

Here are some errors I found in my article. I'm still working on a reply to the letter from Mr. Weller.

In the second line after Eq 3, "$\theta = \omega t$" should be "$\phi = \omega t$." In this article, ϕ and θ have very different meanings. In the sentence after Eq A4, the word "for" should be moved from just before the word "and" to just after the words "Solving Eq A4" as "Solving Eq A4 for . . ."

In the second sentence of the second paragraph of Appendix D, change "meaning R_X" to "meaning *of* R_X." At the end of column 2, paragraph 2, sentence 1, on p 35, add endnote 2 as "measurement applies to RF PAs.[2]"

Change the location of endnote numbers "[1, 2]" in the Conclusions from after "Warren's[1, 2]" to before Warren's as "none exists.[1, 2] Warren's . . ."

Eq 15 should read

$$Zs = \Delta Ep / \Delta Ip_1$$

In the line between Eq C4 and Eq C5, change "Ip_1" to "$\mathrm{I}p_1$" (Roman type). Eq C1 should start

"$\Delta ip(\phi) =$"

—73, Bob Craiglow, 2204 Glass Rd NE, Cedar Rapids, IA, 52402; **craiglowrl@juno.com**

An Effective 40- and 80-Meter SSB/CW Receiver (Jan/Feb 2005)

Gentlemen,

First, thanks for the excellent job you folks did when publishing my article in *QEX*. I am truly flattered that my receiver made the front cover of the premier radio amateur experimenter's journal. I now have bragging rights at the ham club. There were a couple of minor errors, though, so I'll provide some feedback.

My correct e-mail address is **dlyndon@direcpc.com** not **director.com** as listed on p 3. I am also listed as **AK4AA@arrl.org** and it should be **AK4AA@arrl.net.**

In Fig 3, the ground connection is missing from the bus at the bottom of the second filter section. On p 6, column 3, line 16, insert "is about 6 dB" after the word "mixer."

Thanks again for your kind attention.—Best regards and 73, *Dave Lyndon, AK4AA, 85 Woody Farm Rd, Hot Spring, NC 28743;* **dlyndon@direcpc.com** □□

In the next issue of QEX/Communications Quarterly

Frank Witt, AI1H, explains methods for measuring cable loss. He compares those methods, giving mathematical relationships and discussing accuracy and resolution—way to go!

Daniele Moretti, IØFGR, and Silvano Ricci, IØLVA, tell us about their experiences on the 6-mm band. Don't miss this interesting story about millimeter waves. □□