Wideband Loop Antenna
Matching Networks

by

Chris Trask / N7ZWY
Sonoran Radio Research
P.O. Box 25240
Tempe, AZ 85285-5240

Senior Member IEEE

Email: christrask@earthlink.net
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Introduction

Loop antennas are of interest to a wide range of users, from shortwave listeners (SWLs) and radio amateurs to designers of direction-finding receiver systems. SWLs and radio amateurs living in confined areas such as apartments or in communities having antenna restrictions find loop antennas and especially active loop antennas to be a practical solution as they can offer directional performance similar to that of a dipole antenna while taking up a considerably smaller space, and their small size makes them readily adaptable to mechanical rotation.

However, the high inductive reactance of the loop antenna impedance is detrimental to wideband performance, and remote tuning is often employed for achieving good performance and enjoying the highly desireable magnetic field response, which provides some degree of immunity from electric field noise from sources such as lightning discharges, faulty mains transformers, and flourescent lighting.

Wideband Active Receiving Loop Antenna Design Practices

Many designs for wideband active receiving loop antennas overcome the high inductive reactance of the antenna by way of two methods, the first of which is to connect the antenna terminals to a high input impedance amplifier, such as the gate terminal of an FET. Although this method overcomes the signal voltage loss due to the high inductance of the antenna impedance, it does so at the expense of the highly desireable magnetic field response characteristics of the loop antenna. With the high load resistance of the amplifier, very little current flows through the antenna and the reception is dominated by the less desireable electric field response, the consequence of which is that the loop antenna becomes more suscpetible to electric field noise.

The second method popular with many wideband active receiving loop antenna designs is to connect the antenna terminals to a low input impedance amplifier, such as the emitter terminal of a common-base bipolar transistor. To a degree, this is an improvement over the previous method as more current will flow through the antenna and the magnetic field properties are partially restored, but the antenna impedance still inhibits the antenna current and the reception becomes an undesirable blending of electric and magnetic field resonse. In addition, the signal voltage seen at the antenna terminals is greatly diminished, requiring a high degree of amplification.

The combination of the diminished signal voltage and the noise figure (NF) of the amplifier results in a substantial increase in the apparent noise temperature of the antenna, which has serious impacts on the dynamic range of the receiving system. At least one commercial loop antenna, the Wellbrook WL1030, uses this method where the antenna terminals are connected to the inverting summation junctions of a pair of high performance operational amplifiers, which are points of virtually zero impedance in the negative feedback topology.

Loop Antenna Impedance

Before we address the design of wideband matching networks, we need to gain an understanding and appreciation of the impedance of loop antennas, the nature of which is a serious drawback in the design of wideband matching networks. It is well known that the loop antenna impedance consists of a small real part R_{ant} (consisting of the radiation resistance plus bulk and induced losses) in series with a large inductance L_{ant}, which renders the loop antenna as being a high Q source (1):

$$Q_{ant} = \frac{\omega L_{ant}}{R_{ant}} \tag{1}$$

where ω is the frequency in radians per second.

There is more than enough literature available about loop antennas that the basic theory really does not need to be repeated here, and very thorough treatments are available from King (2), Kraus (3), Terman (4) and Padhi(5). Most authors provide little discussion about the impedance of the loop antenna, other than to demonstrate that the impedance is dominated by a large series inductance and is a cascade of parallel and series resonances (6). A few go further and show that the loop antenna impedance can be seen as a shorted transmission line. Terman (4) makes use of such a method, which is usable for frequencies below the first parallel resonance.

An IEEE paper published in 1984 (7), provides a very useful means for estimating the real and imaginary parts of the loop antenna impedance, the latter of which is a refinement of the method proposed by Terman, and which the authors of that paper further refine by providing scalar coefficients for use with a wide variety of geometries that are commonly used in the construction of loop antennas. In their approximation, the radiation resistance is determined by:

$$R_{ant} = a\, tan^{b}\left(\frac{k_0\, L}{2}\right) \qquad (2)$$

where L is the perimeter length of the loop antenna and the wave number k_0 is defined as:

$$k_0 = \omega\sqrt{\mu_0\, \varepsilon_0} \qquad (3)$$

where μ is the permeability of free space ($4\,\pi\, 10^{-7}$ H/m), and ε is the permittivity of free space ($8.8542\, 10^{-12}$ F/m). The coefficients *a* and *b* in Eq. 2 are dependent on the geometry and the perimeter length of the loop antenna, a list of values being provided in Table 1.

The inductive reactance of the loop antenna impedanceis determined by:

$$X_{ant} = j\, Z_0\, tan\left(\frac{k_0\, L}{2}\right) \qquad (4)$$

where Z_0 is the characteristic impedance of the equivalent parallel wire transmission line, defined as:

$$Z_0 = 276\, ln\left(\frac{4\, A}{L\, r}\right) \qquad (5)$$

where A is the enclosed area of the loop antenna and r is the radius of the antenna conductor.

A highly detailed report from the Ohio State University Electroscience Laboratory in 1968 (8) provides a thorough analytical means for estimating the real and imaginary parts of the impedance of single and multi-turn loop antennas, as well as the antenna efficiency.

Computer simulation routines such as EZNEC also provide a useful means for estimation the loop antenna impedance. Together with papers and reports such as those mentioned herein, they allow the designer to gain an understanding of the nature of the loop antenna impedance. They are not, however, suitable substitutes for actual measurements and

Configuration	$L/\lambda \le 0.2$		$0.2 \le L/\lambda \le 0.5$	
	a	b	a	b
Circular	1.793	3.928	1.722	3.676
Square (side driven)	1.126	3.950	1.073	3.271
Square (corner driven)	1.140	3.958	1.065	3.452
Triangular (side driven)	0.694	3.998	0.755	2.632
Triangular (corner driven)	0.688	3.995	0.667	3.280
Hexagonal	1.588	4.293	1.385	3.525

Table 1 - Coefficients to be Used with Equation 2

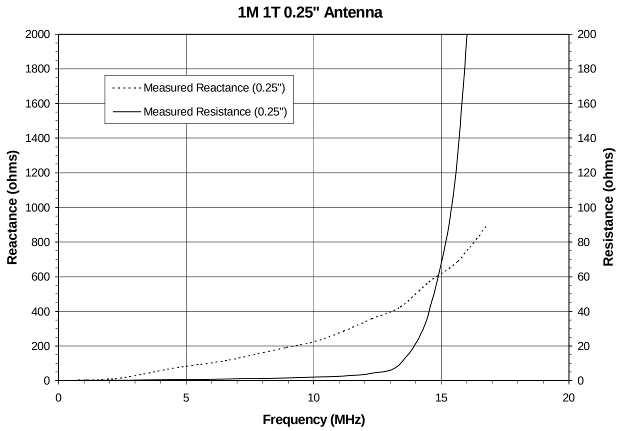


Fig. 1 - Measured Impedance of 1 meter Diameter Loop Antenna made with 0.25" Copper Tubing

the designer should always rely to measured data, especialy when designing matching networks.

Fig. 1 shows the measured terminal impedance of a 1m diameter loop made with 0.25" copper tubing. In order to ensure that the loop antenna is properly balanced,a 1:1 BalUn transformer is used to interface the loop antenna with the impedance bridge. Loop antennas that are fed unbalanced have dramatically different impedance characteristics and radiation patterns from those that are fed balanced (9).

In the process of designing matching networks for adverse impedances such as those of loop antennas, it is very useful to devise lumped element equivalent models as some analysis and optimization routines, such as PSpice, do not have provisions for including tables of measured data for interpolation. Fig. 2 illustrates two rudimentary lumped element models, the first being usable up to and slightly beyond the first parallel resonance and the second being usable to the point prior to where the impedance becomes asymptotic, or about 25% below the first parallel resonance. Far more detailed models can be devised that include subsequent resonances and anti-resonances (10), but they would serve little purpose here as the application here is focused on frequencies below the first resonance.

In general, I use the more detailed model for PSpice simulations and the simpler one for illustrations such as those to be used later herein. For the 1m diameter loop made from 0.25" copper tubing, the element values are

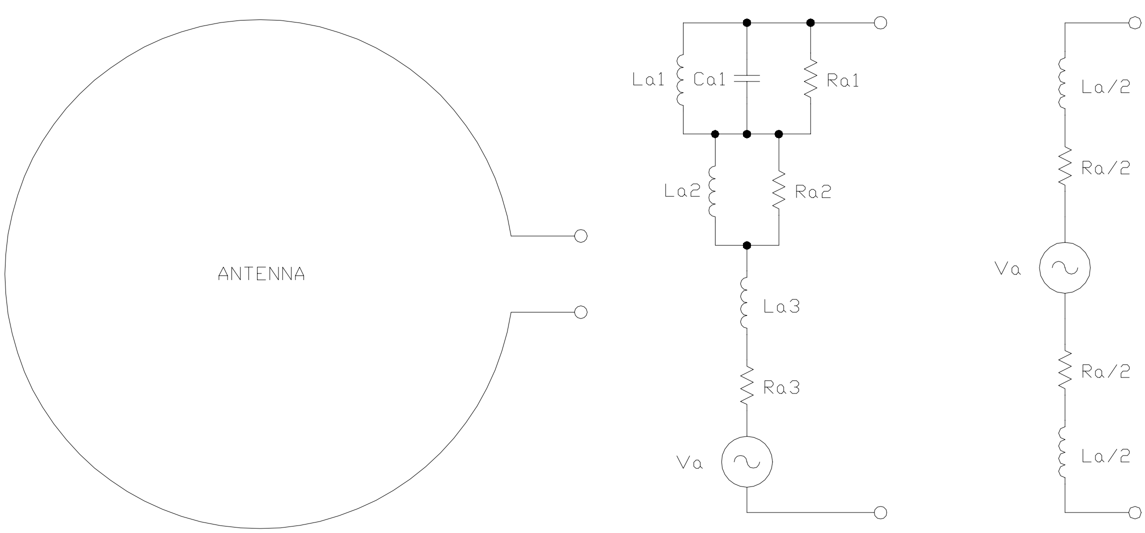


Fig. 2 - The Loop Antenna (left) Together with Detailed (centre) and Simplified (right) Lumped Element Impedance Equivalent Models.

roughly:

Ra_1 = 5k ohms La_1 = 0.4uH
Ca_1 = 300pF

Ra_2 = 0.6 ohm La_2 = 0.05uH

Ra_3 = 1.0 ohm La_3 = 2.2uH

These values were used in the evaluation of a wide variety of passive matching networks as well as a few active solutions by way of PSpice simulations, the goal being to devise a matching network, preferrably passive, that could be coupled directly to a coaxial cable having an impedance of 50-ohms or to a subsequent amplifier stage or stages of comparable impedance.

In the overall scheme, balanced networks a prefered as they allow for the supression of common-môde interference signals such as lightning discharges, faulty mains transformers, flourescent lighting, as well as nearby high-power broadcasting stations.

Simple Matching Networks

Relatively simple matching networks, such as those used in the design of HF radio astronomy dipole arrays (11), are impractical for loop antennas due to the extreme ratio between the imaginary and real parts of the im-

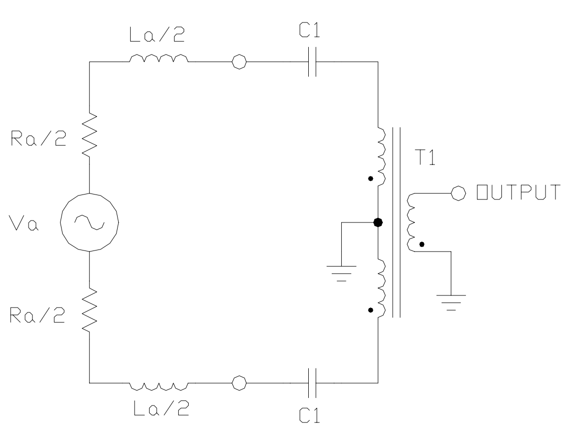


Fig. 3 - Passive Series Tuning

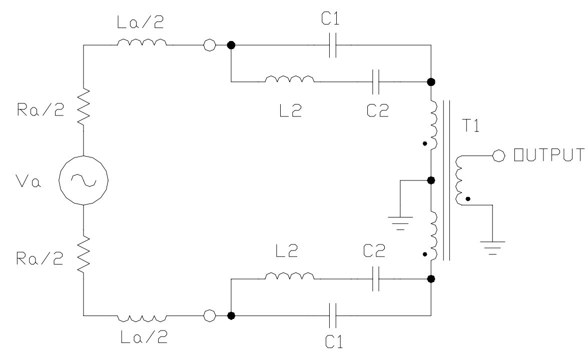


Fig. 4 - Adding Multiple Series Resonances

pedance.

At first glance, the simplified equivalent model in Fig. 2 readily suggests that adding a capacitor in series with each antenna terminal would provide a good match, which it does but only for a narrow bandwidth. This approach, illustrated in Fig. 3 provides for suberb signal-to-noise performance as the loop antenna can be matched properly to the load (1, 12). In addition, the magnetic field performance of the loop antenna can be thoroughly enjoyed, reducing the effects of noise from electric field sources, though not to the degree as would be experienced with a shielded loop antenna.

This would then suggest that adding a multiplicity of series resonances would be a suitable approach to wideband matching, such

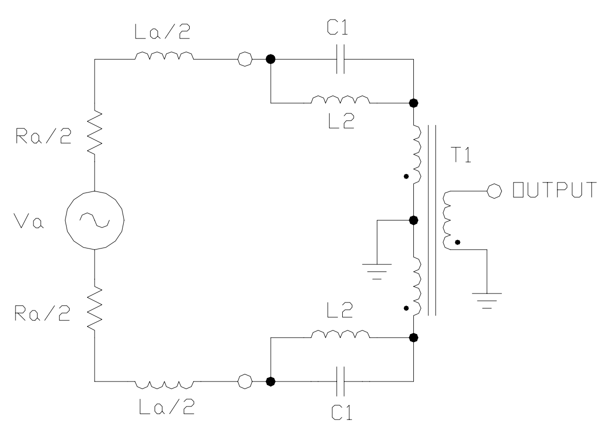


Fig. 5 - Midband Consequences of adding Multiple Series Resonances

a scheme being illustrated in Fig. 4 where the introduced arms each having a series inductance L2 and capacitance C2 provide a second series resonance with the loop antenna inductance. This, however, has serious negative consequences. Referring now to Fig. 5, the introduced series resonant arms of Fig. 4 become inductive above their resonant frequency, and the resulting inductance combined with the capacitance C1 produces a transmission zero between the two series resonances. This consequence is inescapable, however there are other topologies that can be brought to bear that will provide the desired multiplicity of series resonances without developing highly undesirable parallel resonances.

What's All This Symmetrical Lattice Stuff, Anyhow?

When devising passive circuitry, symmetrical networks are a bit easier to work with as they require fewer parameters for their characterization and the constraints on achieving a usable result are not as serious as with asymmetrical (ie - unbalanced) networks (13).

A symmetrical network of considerable interest in network synthesis is the symmerical lattice, also known as a balanced or full lattice. Shown in Fig. 6, the symmetrical lattice consists of series impedances Za and crossarm impedances Zb. Note that the illustration of Fig. 6 can also be drawn as a bridge (14).

If the symmetrical lattice of Fig. 6 is terminated with a load resistance RL, then for a constant resistance function the series and crossarm impedances Za and Zb are related by (13, 14):

$$R_L = \sqrt{Z_a \, Z_b} \tag{6}$$

and the voltage transfer function is determined by:

$$V_{12} = V_{34}\left[\frac{1+\dfrac{Z_a}{R_L}}{1-\dfrac{Z_a}{R_L}}\right] = V_{34}\left[\frac{R_L + Z_a}{R_L - Z_a}\right] \tag{7}$$

The crossarm impedance Zb does not appear in Eq. 7 due to the symmetry and constant resistance property of the symmetrical lattice as stipulated in Eq. 6. In practice, a distinct advantage of the constant-resistance symmetrical lattice is that a single complicated symmetrical lattice can be decomposed into a cascade of simple lattices, which can then be reduced to asymmetrical networks (13).

The Darlington C Section

In the decomposition of symmetrical lattice networks to asymmetrical networks, there are many individual network topologies that can be brounght to bear. Simple networks consisting of individual inductors or capacitors are referred to as "A" sections, and simple series and parallel combinations of an inductor and ca-

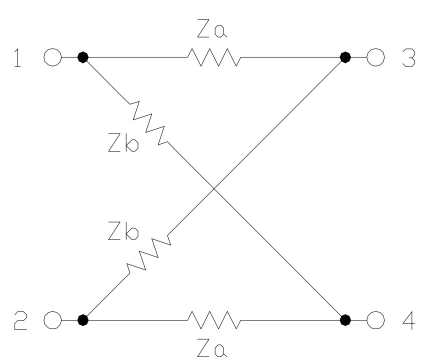


Fig. 6 -Generalized Symmetrical Lattice

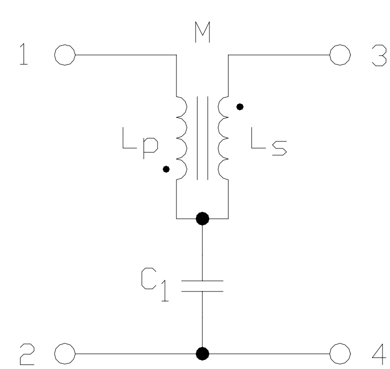


Fig. 7 - Darlington C Section in Asymmetrical Form

pacitor are referred to as "B" sections. A slightly more complicated yet very useful class of networks are known as "C" sections. Fig. 7 illustrates one of these, commonly referred to as a Darlington C Section (15) which consists of a capacitor C_1 and a two-winding transformer having a primary inductance L_p, a secondary inductance L_s, and a mutual inductance M, which is related to Lp and Ls by way of:

$$M = k\sqrt{L_p L_s} \tag{8}$$

where k is the magnetic coupling coefficient between the primary and secondary windings. For an ideal 1:1 transformer having a primary and secondary inductace L:

$$k = 1 \tag{9}$$

$$M = L_p = L_s = L \tag{10}$$

In forming a symmetrical lattice equivalent of the Darling C section, it is useful to proceed by first forming a T-network equivalent, as shown in Fig. 8. Here, the three inductances L_1, L_2, and L_3 are related to the transformer inductances L and M by way of (16):

$$L_2 = -M = -L \tag{11}$$

$$L_p = L_1 + L_2 \tag{12}$$

$$L_s = L_2 + L_3 \tag{13}$$

$$L_1 = L_p - L_2 = L_p + L = 2L \tag{14}$$

$$L_3 = L_s - L_2 = L_s + L = 2L \tag{15}$$

The asymmetrical T-network of Fig. 8 can be translated into the symmetrical lattice network of Fig. 9, where the capacitance C_2 is determined by (13):

$$C_2 = 0.5\,C_1 \tag{16}$$

and the series and crossarm inductances L_4 and L_5 are determined by:

$$\frac{L_5 + L_4}{2} = L_p = L_s = L \tag{17}$$

$$\frac{L_5 - L_4}{2} = M = -L \tag{18}$$

$$L_4 = 2L \tag{19}$$

$$L_5 = 0 \tag{20}$$

The Darlington C section is very useful in passive network transfer function realizations as it creates pole/zero pairs in the imaginary axis of the z-plane. A similar network having the transformer connected in-phase, known as a Brune C section or simply Brune section (17), creates pole/zero pairs along the real axis of the z-plane. A more complicated network known as the Darlington D section (15) creates

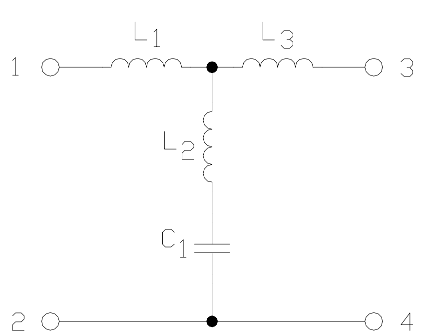


Fig. 8 - T-network Equivalent of the Darlington C Section (from ref. 13)

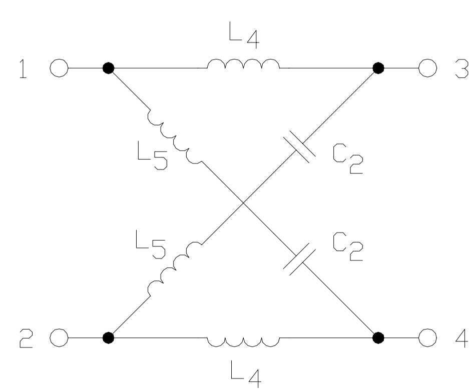


Fig. 9 - Symmetrical Latttice Equivlent of the Darlington C Section (from ref. 13)

a quatenary of complex pole/zero pairs in the z-plane. These networks and their variations all see extensive usage in the realization of passive group delay and phase shift equalizers.

Using the Darlington C section in the matching of loop antennas to resistive loads has very useful consequences, as a series of these sections can be cascaded to produce a multiplicity of series resonances without incurring the inbetween parallel resonances such as those that result from the networks shown earlier in Fig. 4 and Fig. 5.

In the realization of a loop antenna matching network using Darlington C sections, a negative consequence is that the average voltage transfer from the antenna to the load decreases as the number of matching network sections increases. In addition, it is necessary that the matching network load impedance be raised above the loop antenna radiation resistance in order to keep the matching section Q's reasonable and of practical use. In the overall design, a lower load impedance is desirable in order to retain the magnetic field performance of the loop antenna, and it becomes necessary to determine a suitable compromise between the desired overall performance and the matching network complexity.

Matching Section Evolution

In the equivalent symmetrical lattice network of Fig. 9 the inductors are not coupled,

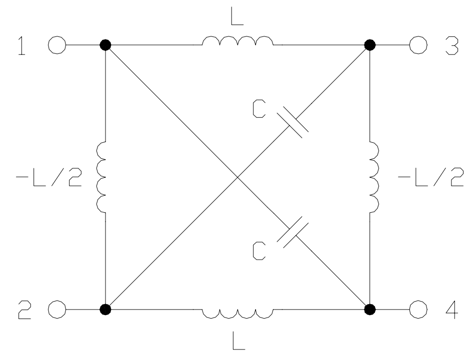


Fig. 11 - Symmetrical Latttice Equivalent of Balanced C Section

however there are distinct advantages in using a transformer in the practical realization of the network. Shown in Fig. 10, the matching network section now consists of a 1:1 transformer having equal windings of inductance L and a pair of capacitors of value C.

By virtue of the reverse polarity of the transformer windings, the equivalent model of the balanced C section of Fig. 10 now includes a pair of negative inductors in shunt across the input and output terminals, as shown in Fig. 11.

Although a negative inductance is impossible to realize, except by circuits containing active elements such as gyrators or negative impedance conveters, negative inductances such as those shown in Fig. 11 as well as those that develop in the equivalent circuits of Fig. 8 and Fig. 9 are realizied through the use of the mutual inductance of coupled inductors or trans-

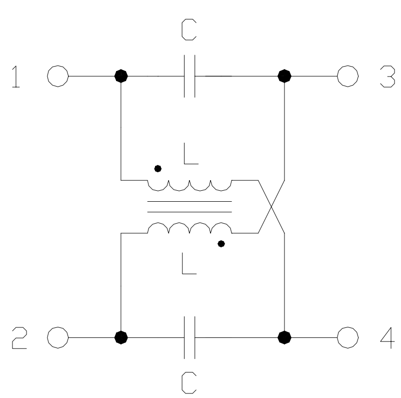


Fig. 10 - Balanced C Section

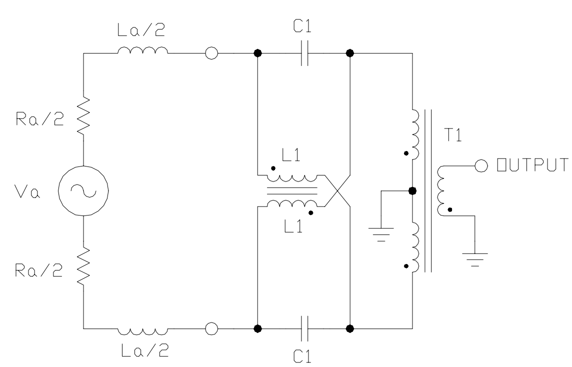


Fig. 12 - Loop Antenna Matching Network Using a Single Balanced C Section

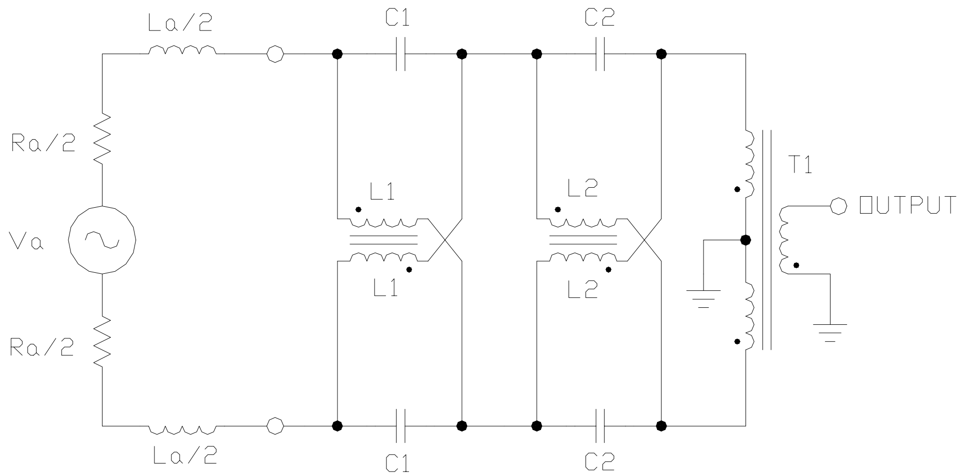


Fig. 13 - Loop Antenna Matching Network Using Cascaded Balanced C Sections

formers, as has been done here (14).

The inclusion of negative inductors in the loop antenna matching network offers a substantial opportunity to reduce the inductive component of the loop antenna impedance while at the same time decreasing the negative impact of multiple matching sections on the voltage transfer characteristics of the matching network. At the same time, the much desired multplicity of series resonances offered by the cascading of Darlington C sections is preserved.

Matching Network Design

Rather than develop the mathematics to provide a closed form solution for matching arbitrary loads using balanced C sections, a few matching networks were devised by way of PSpice modeling. The first of these networks is shown in Fig. 12, where a single balanced C section is used to match a loop antenna model to a resistive load. It was found while performing an exhaustive number of simulations that a 200-ohm load was best suited for passband ripple and high frequency cutoff.

A second network having a cascade of two balanced C sections, shown in Fig. 13, was also modeled exhaustively, and again demonstrated that a 200-ohm load was to be preferred for passband ripple and high frequency cutoff. Although this load resistance is high compared to the loop antenna radiation resist-

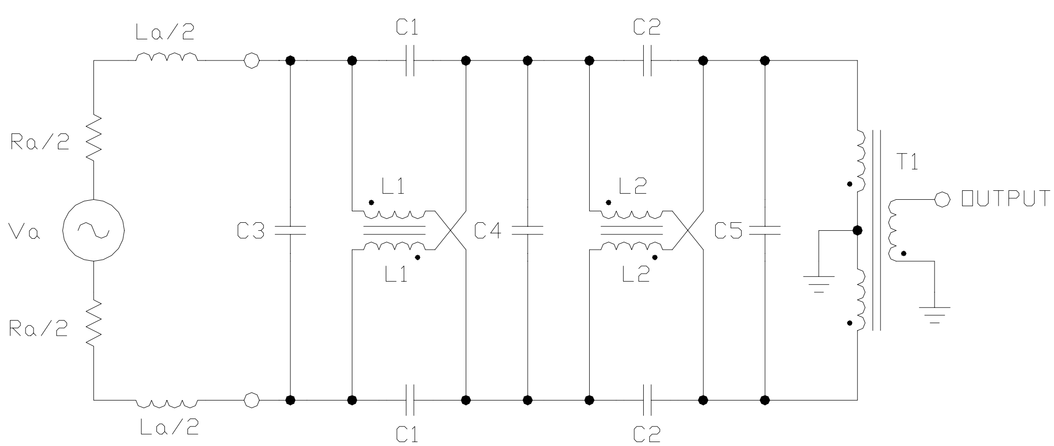


Fig. 14 - Adding Shunt Capacitors

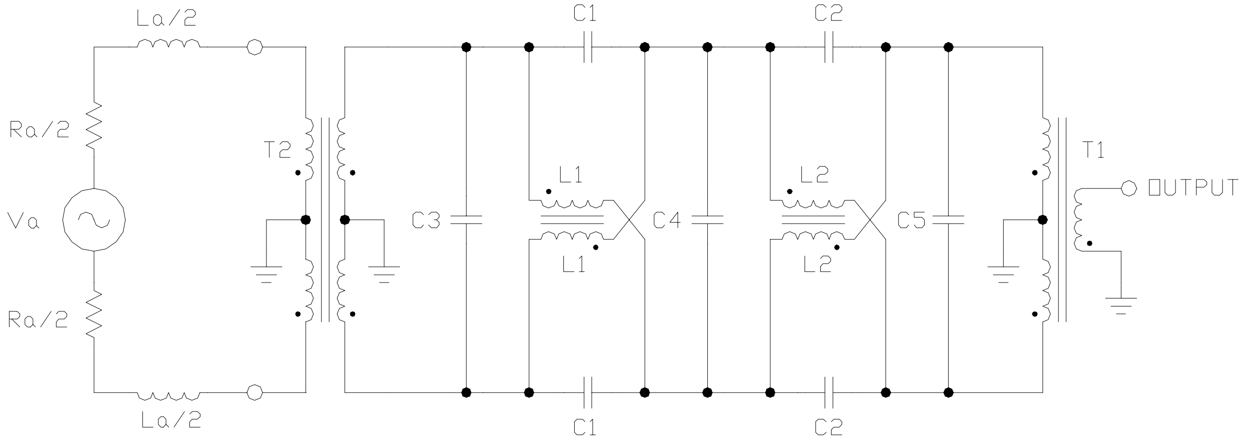


Fig. 15 - Adding an Input Transformer

ance, it is very convenient for matching directly to a 50-ohm load as the output transformer T1 becomes either a 2CT:1 wideband transformer or a 4:1 Guanella transmission line transformer (TLT), both of which are very easily constructed.

As shown in Fig. 14, shunt capacitors were added to the matching network to add additional poles to the response. This further improves the passband ripple and allows for increasing the ripple bandwidth to within 5% of the parallel resonance of the loop antenna.

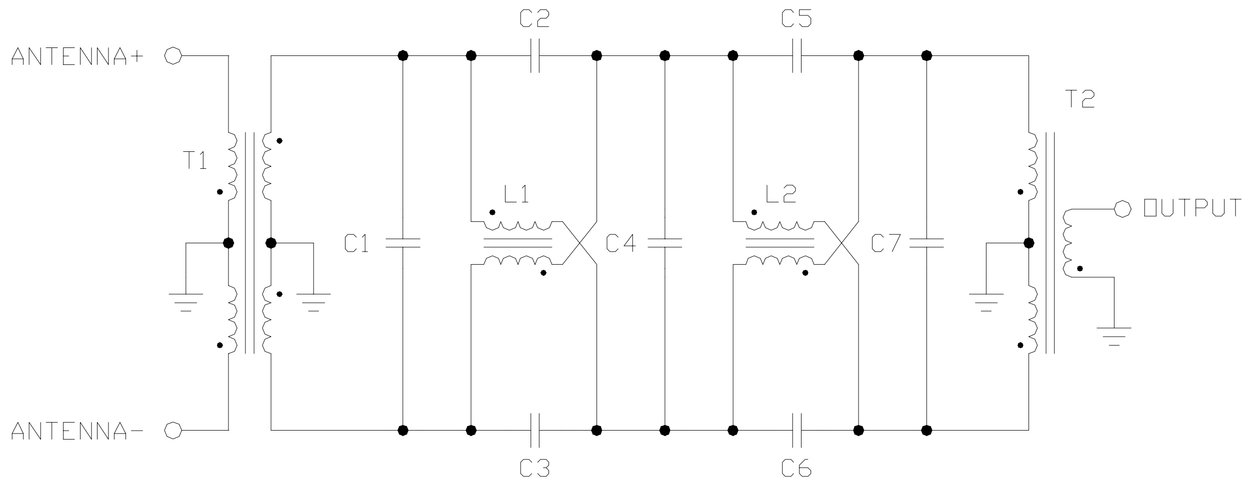
At the end of the design process, an input coupling transformer was added between the loop antenna and the matching network. This addition fulfills a number of purposes, the first being to maintain the current and voltage balance of the loop antenna itself. Secondly, it provides the first defense in reducing noise caused by common-môde signals that would result from noise nources such as lightning discharges, faulty mains transformers, and flourescent lighting. In addition to the electrical aspects, the mechanical construction of the input transformer provides a means of mounting the loop antenna to the matching network assembly.

Prototype Construction and Testing

The final schematic and parts list for the matching network of Fig. 15 is shown in Fig. 16. A prototype of this circuit was constructed on a 1.2”x4.0” piece of Ivan board (0.80” squares, 0.10” apart on 1/16” G-10 epoxy fiberglass, similar to FR-4), as shown in the photograph of Fig. 17. Silvered mica capacitors were used throughout as they are most suitable for HF and VHF circuitry that requires high Q capacitors for tuned circuits.

The input coupling transformer T1 was made with a Communication Concepts RF800-0 Balun Transformer core assembly. By using #6 AWG copper wire for the antenna with a single layer of heat-shrink PVC tubing as an insulator, a transmission line transformer having a characteristic impedance of 7 ohms is formed, which is well suited for the task at hand (18, 19, 20).

A Mini-Circuits T4-6T may be used for the output transformer T2, however a much better performing transformer is easily made with four turns of #34 trifilar wire through a Fair-Rite 2843002402 binocular core, as is shown in Fig. 17 (20, 21, 22). Details for the construction of this transformer are shown in Fig. 18. Basically, one wire from the trifilar twist is separated out to the right for the secondary winding. One wire each of the remaining two wires at opposite holes are brought out for the ends of the primary winding, and the remaining two wires



Parts List

C1, C7 - 36pF Silvered Mica
C2, C3 - 100pF Silvered Mica
C4 - 27pF Silvered Mica
C5, C6 - 120pF Silvered Mica

L1 - 3.3uH (see text)
L2 - 2.7uH (see text)

T1 - 2CT:2CT 7-ohm Transmission Line Transformer (Communication Concepts RF800-0, see text)

T2 - 2CT:1 consisting of four turns of #34 trifilar wire on Fair-Rite 2843002402 core, or Mini-Circuits T4-6T (see text)

Fig. 16 - Schematic and Parts List

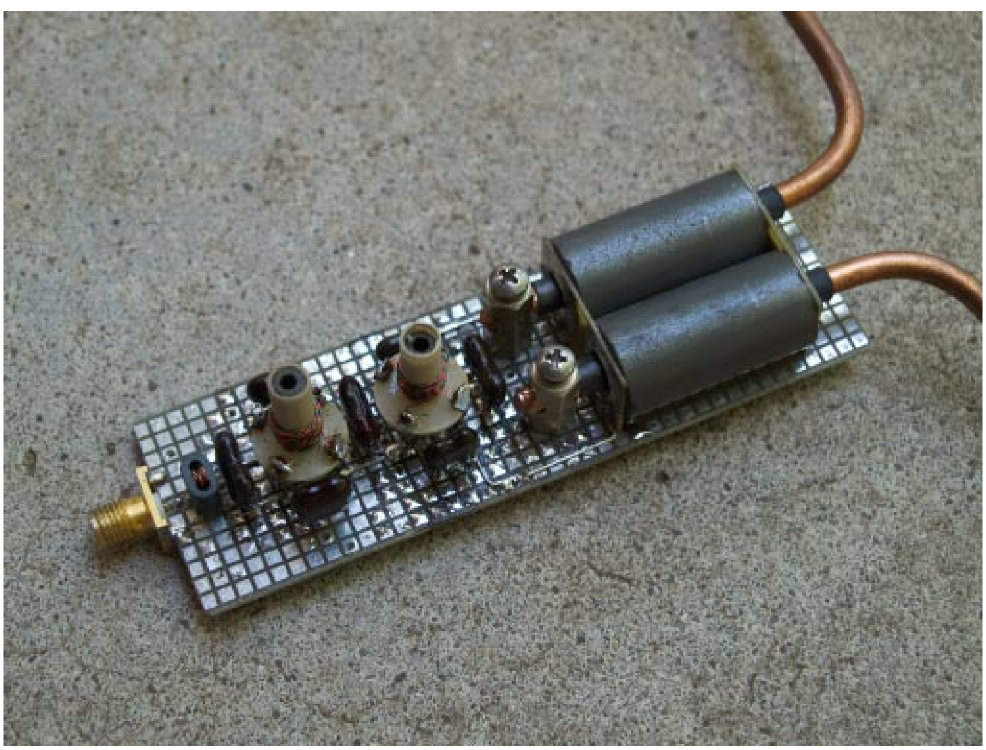


Fig. 17 - Prototype with Variable Inductors

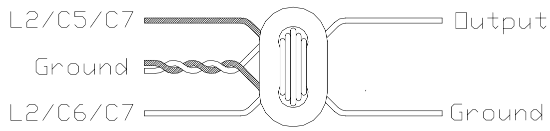


Fig. 18 - Transformer T2 Construction Details

are twisted together to form the grounded cetre tap.

Inductors L1 and L2 are made with 20 turns of #36 bifilar wire on adjustable coil forms (manufacturere and part number unknown). With Micrometals TH36-0202 threaded cores used for the adjustable slugs, these inductors had an adjustment range from 2.15uH to 4.37uH, well beyond the circuit requirements.

Initial testing was conducted by coupling a signal generator to a 1 meter diameter loop of #6 AWG wire by way of six turns of #20 insulated wire at the centre, directly opposite the matching network, thus simulating the voltage source shown in the various model illustrations herein. Reults showed that the rsponse was very close to that pridicted in the PSpice simulations, thus showing that the use of the balance C section as a matching network element is a valid approach.

On-air testing showed that the sensitivity of the 1 meter diameter loop antenna with the wideband matching network is comparable to that of a 1 meter long whip, meaning that the a high gain low noise amplifier of at least 20dB of gain is needed to make the antenna fully usable. Although this may seem a bit disappointing, it should be pointed out that the directional characteristics of the loop antenna are a distinct advantage over the omnidirectional characteristics of the short vertical whip antenna.

Closing Remarks

The wideband loop antenna matching networks described herein are the results of a considerable amount of PSpice simulation and prototype evaluations. There was no doubt from the start that the results would be far less than those that are enjoyed by series tuning such as that shown in Fig. 3. The end results show that there may be some slight advantage above active loop antennas that employ no matching at all. If anything, this investigation into wideband matching networks for loop antennas has helped define the limits as to what can be realized. For instance, a usable antenna could be realized in which a narrower band of frequencies is matched to a much lower impedance.

Hopefully this paper will provoke comment and discussion with the goal of improving the techical aspects of wideband loop antenna matching. Sometimes new ideas can bring out related or even unrelated ideas that otherwise would not happen without work such as this being made available.

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