

RF APPLICATION NOTE:

SEME LAB

PUSH-PULL CIRCUITS and WIDEBAND TRANSFORMERS



Push-Pull Transistors

Semelab plc produces a wide range of push-pull MOSFETs and this application note is intended as a guide to some circuit design principles which are particularly appropriate when using these devices. Two examples of Semelab push-pull MOSFETs are shown in figure 1.

Fundamentally, a push-pull circuit uses a pair of effectively separate transistors operating 180° out of phase with one another. If good amplitude and phase balance is maintained between the signals in each half of the device, then an RF ground will exist at the midpoint. This approach leads to several advantages over single-ended designs:

- Input and Output Impedances Doubled
- Reduced Even Order Harmonics
- Increased Output Power
- Higher Bandwidths Possible
- Reduced Effect of Common Lead Inductance



Figure 1 – Semelab MOSFETS

Perhaps the most significant disadvantages are the need for differential RF excitation and the fact that excellent symmetry is required in both the matching circuit and the device itself. Semelab resolves the latter of these issues by individually selecting each side of their push-pull MOSFETs and rigorously testing them to ensure exceptional conformity between the threshold voltage and transconductance parameters. A generic push-pull circuit configuration is shown in figure 2.

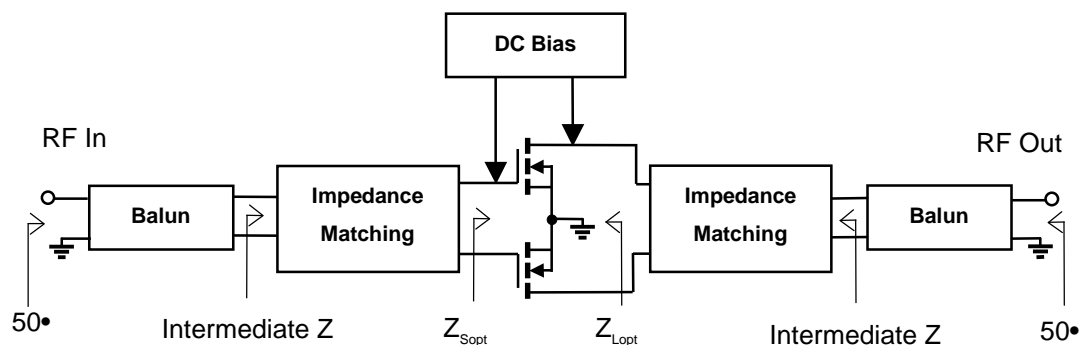


Figure 2 – Generic Push-Pull Amplifier Schematic

Balanced to Unbalanced Transformations

As has already been mentioned, push-pull transistors require splitting the RF signal at the input (along with the reciprocal operation i.e. the need to recombine the output). Circuit elements that provide this function are referred to as *baluns* (from balanced to unbalanced). The ideal balun would split the signal into halves of equal amplitude without loss, along with providing an 180° differential phase shift across all frequencies. Although “ideal” baluns are only theoretical devices, several approaches can be used to build practical baluns with excellent performance over wide bandwidths. Principle methods include:

- Conventional Coil Transformers
- Transmission Line Transformers
- Microstrip Structures (Wilkinson Divider; Line and Ring Hybrids)

The choice of method is largely dependant on frequency. For HF (up to 30MHz) magnetically coupled coil transformers are suitable. Above these frequencies, leakage inductance and parasitic capacitance degrades the performance making them a poor choice. VHF and UHF circuits (30MHz – 1GHz) most commonly employ transmission line balun structures, built from coaxial cable or twisted pairs. Microstrip structures such as Wilkinson dividers also offer very good performance, but these are limited by their physical size (in the order of $\lambda/2$) and are therefore only practical at higher frequencies. Substrates with high dielectric constants can mitigate this problem to some extent.

After the balun, each half of the matching network is usually calculated by conventional means treating each half of the transistor as if it were a single ended device*. One important point to note is the difference between balanced (i.e. drain to drain) impedance and unbalanced (drain to ground) impedance. In a well matched circuit, the drain to drain impedance should be twice that measured from drain to ground.

* Optimum source and load impedances for Semelab transistors can be found at http://www.semelab.co.uk/rf/zopt_index.php

Coupled Coil Transformers

Conventional transformers operate principally by coupling signals between primary and secondary windings through a suitable magnetic material. From Faradays' law it can be shown that, for an ideal transformer, the voltage per turn is the same on each side. The relationship between impedances on the primary and secondary side can then be found as:

$$Z_{IN} = Z_{OUT} * (n_{IN}/n_{OUT})^2$$

This implies that transformers are ideally suited to impedance transformations since they can be scaled by any factor, depending only on the turns ratio. When the parasitics are negligible (i.e. at low frequencies – up to 30MHz) coupled coil transformers are often used where wideband resistance transformation is required. Various winding topologies are possible, and are chosen depending on whether DC isolation and/or balanced signals are required. Figure 3 below shows the simplest configurations. The autotransformer in 3a has a tapped continuous winding and provides a DC short between input and output. Figure 3b shows a more conventional transformer with DC isolation between primary and secondary windings. Other arrangements can allow for balanced-unbalanced operation. The centre tapped secondary transformer in 3c is a balanced signal splitter which also provides an $N^2/2$ impedance transformation.

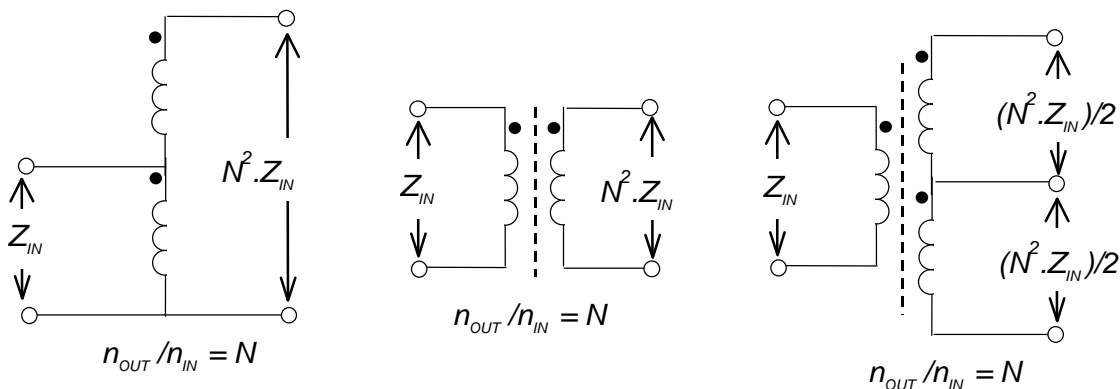


Fig 3a – Autotransformer Fig 3b – Conventional Transformer Fig 3c – Centre Tapped Secondary

Higher permeability (•) ferrite material is usually preferred for better coupling, although it should be noted that due to the increased flux density, high • cores will saturate more easily than those with lower •. Permeability can also vary as a function of current level, frequency and temperature.

Two-hole balun cores and toroids are the most commonly used ferrite structures (see Appendix C for suppliers). The choice of wire also affects performance and is essentially a trade off. Heavier wire will increase the coupling, but it will also increase the parasitic winding capacitance.

In practice, all transformers suffer from parasitic effects that cause their behaviour to deviate from the ideal. The schematic below shows an equivalent circuit for modeling the transformer.

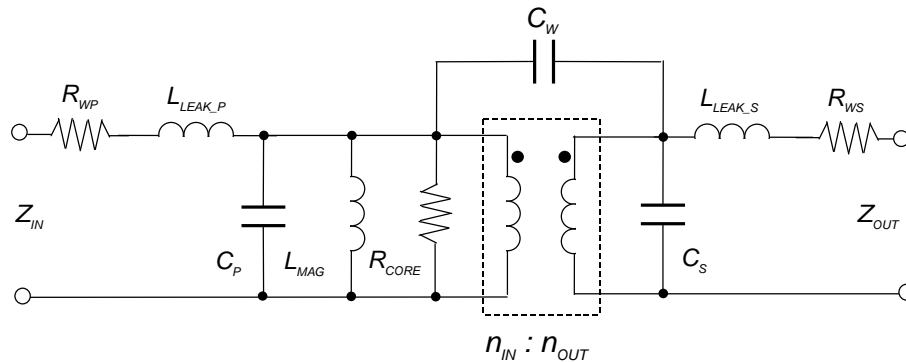


Figure 4 – Conventional Transformer Equivalent Circuit

The ideal $n_{IN}:n_{OUT}$ transformer is represented inside the dashed line. Series resistances R_{WI} and R_{WO} are the resistances of the primary and secondary windings. Although these are negligible at DC, the skin effect will increase the resistance in proportion to the frequency. R_{CORE} models the loss of energy in the transformer's core due to eddy currents and hysteresis effects. Leakage inductances represented by $L_{LEAK,P}$ and $L_{LEAK,S}$ model the magnetic flux which passes outside of the core and therefore is not coupled between the input and output ports. L_{MAG} limits the low frequency response of the transformer and it has its physical origins in the finite magnetizing inductance in the coils. At higher frequencies, the capacitive effect between windings (C_W) and between the turns of each winding (C_P and C_S) also become important.

Single winding transformers are the type most commonly used at RF. The primary winding consists of a single turn of hollow metal tubing threaded through suitable ferrite material shorted at one end via a copper strap or PCB. Insulated wire is then passed inside this tube to form the secondary. Since the primary is limited to a single turn, the impedance ratios are limited to $1:n^2$, where n is the integer number of turns in the secondary. The principle advantage of this construction method is the close proximity of the windings and ferrites, leading to minimal leakage

inductance and improved performance at high frequency. A 1:4 transformer is shown below in figure 5.

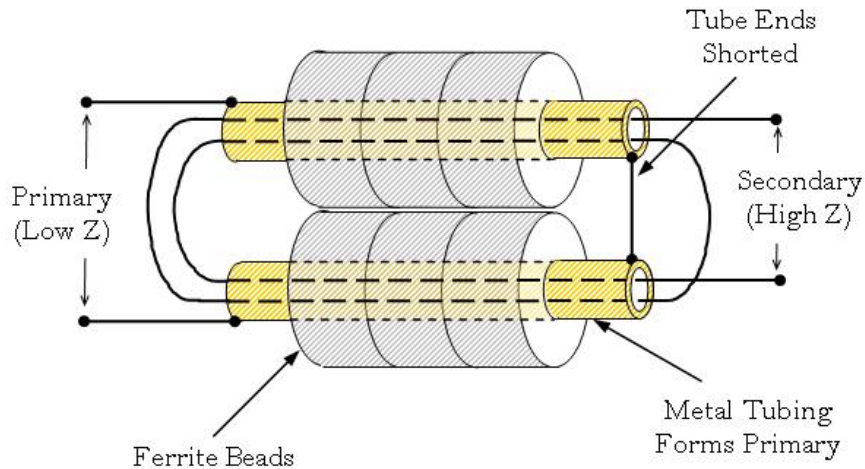


Figure 5 – Single Winding Transformer Implementation

These types of transformer can be easily constructed from coaxial cable. By using the centre conductor as the secondary, the cable can be wound to the required number of turns. The outer conductor is then removed at the open circuit end and soldered together along its remaining length. Figure 6 below shows a range of these coaxial transformers and the associated “E” and “I” magnetic cores.

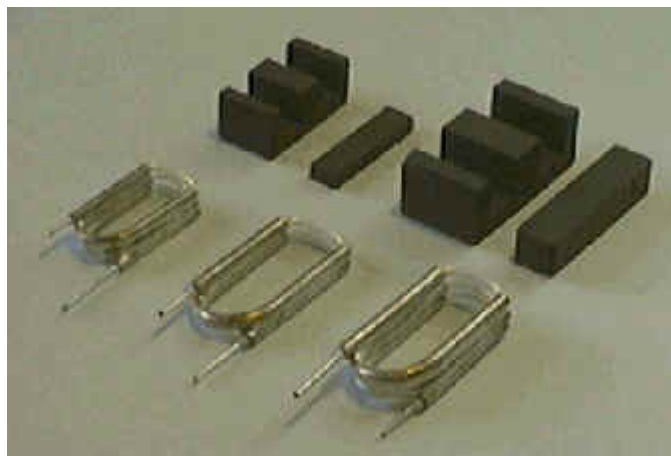


Figure 6 – Practical Single Winding RF Transformers

Generally, conventional transformers will not perform as well as transmission line transformers in terms of bandwidth, losses and power handling. They do, however, offer a wider variety of possible impedance ratios and DC isolation from primary to secondary windings.

Transmission Line Transformers

These types of transformers rely on a large variation in impedance being presented to the differential (balanced) mode and common (unbalanced) mode currents on the transmission line. This is readily achieved at high frequencies, and the low frequency performance can be extended using a suitable ferrite material around the transmission line[†]. It is important to note that unlike conventional transformers, the ferrite is not the medium being used to transfer RF power between ports. For this reason much larger bandwidths and greater efficiencies are achievable using the transmission line transformer. Degraded performance at high frequency is caused by the length of the line (max $\lambda/8$), by ground plane effects, and by non-optimum characteristic impedance of the chosen transmission line.

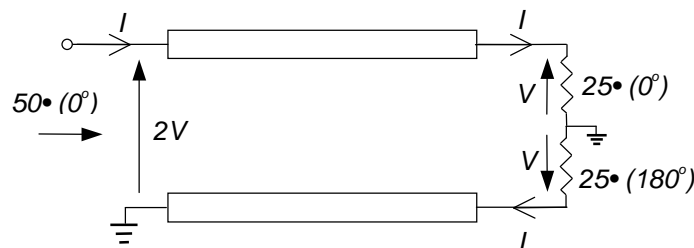


Figure 7 – Guanella Transmission Line Balun

Figure 7 shows the basic “building block” of transmission line transformers – the 1:1 unbalanced-to-balanced transformer first introduced by Guanella in 1944. Note that this provides no impedance transformation, but each side of the balanced load has an impedance half that seen at the input. This is effectively an RF transmission line equivalent of the centre-tapped transformer used at lower frequencies.

The choice of transmission line largely depends on the characteristic impedance as dictated by the choice of balun or transformer. Coaxial cables of fixed values are freely available, and these

[†] Differential mode currents produce no magnetic field external to the transmission line and therefore remain unaffected by the addition of ferrites.

can be combined to make transmission lines of varying impedances. As with most UHF/VHF transmission line transformers, Teflon insulated coaxial cable is preferred – see appendix C for a list of suppliers. The power handling capability is generally limited by the maximum allowable temperature, which itself is a function of dielectric material and cable diameter. The table in appendix B provides a rough guide. Twisted pairs are useful since a limitless range of characteristic impedances can be realised. Below are some guidelines for designing twisted pairs of specific Z_o .

- Increasing the number of twists per unit length decreases impedance
- Increasing the wire diameter will reduce the impedance
- Very low impedances are achieved by using multiple pairs in parallel
- 50 Ω wire can be made using AWG-22 enameled wire with 2.5 twists per cm

The impedance will also be affected by the wire separation, the dielectric material between them as well as the proximity of any ground plane.

The simplest type of transmission line transformer is the $\lambda/4$ line. Since the impedance match is dependant on the presence of a quarter wavelength standing wave, these transformers are inherently narrowband. They also require a transmission line of characteristic impedance which is the geometric mean of the input and output impedances to be matched i.e.:

$$Z_o = \sqrt{Z_{IN} * Z_{OUT}}$$

This can limit the practicality of this technique since many types of transmission line are only available in a limited range of impedances. However, cables can be combined to achieve Z_o variations. For example, two paralleled 25 Ω cables will produce a transmission line with a characteristic impedance of 12.5 Ω .

As well as the impedance matching function, the $\lambda/4$ line can be used as a balun in the same way as above – by grounding one conductor on the unbalanced side and connecting each conductor to one half of a balanced load on the other side. The diagram below shows an implementation of this using coaxial cable.

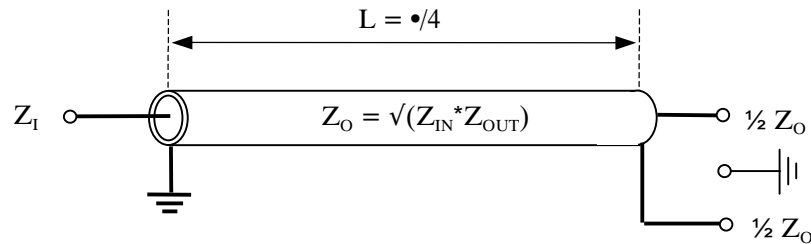


Figure 8 – Coaxial Quarter-Wavelength Balun

As well as the narrowband $\lambda/4$ transformer, a single transmission line can be used as an unbalanced to unbalanced (unun) transformer. This is capable of providing wideband 1:4 impedance transformation ratios by connecting it in the so called 'bootstrap' configuration as shown in figure 9. Here, the two conductors which constitute the transmission line are used as the primary and secondary windings in a similar way to the conventional autotransformer.

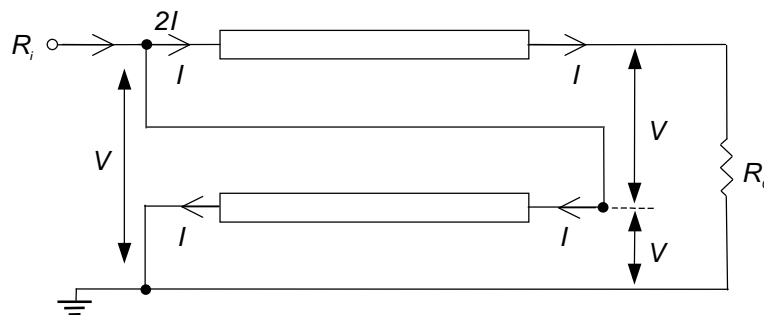


Figure 9 – Ruthroff 1:4 Unun

If a voltage V is present across the input, the same voltage will be impressed across the lower conductor of the transmission line. Charge conservation demands that the same voltage must be present across the upper conductor also. The voltage across R_o is therefore the sum of these voltages - $2V$. If the current I is to flow through the load, it must also flow through the upper conductor. But, again since both conductors have the same voltage across them, the currents through each must be identical. Therefore since $R_o = 2V/I$ and $R_i = V/2I$ then:

$$R_i = R_o/4$$

Calculations show that maximum power transfer occurs for this transformer with the optimum transmission line impedance $Z_o = 2 \cdot R_i$. For best performance the transmission line should be kept as short as possible. Figure 10 below shows two practical implementations of the 1:4 unun using

coaxial cable and wire-wrapped toroid configurations. Obviously, these devices are bilateral and simply reversing the ports will provide 4:1 step-down transformations.

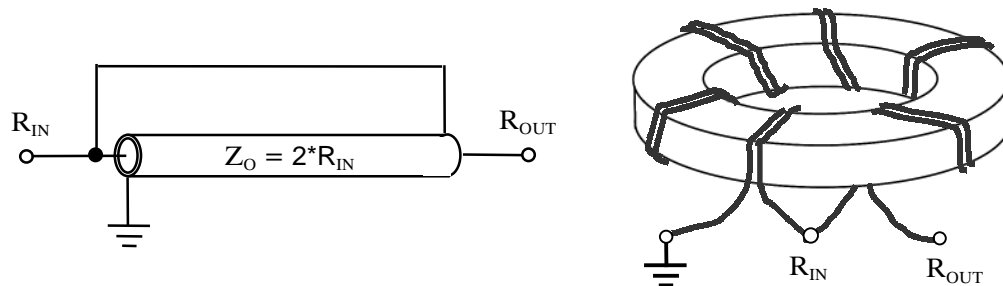


Figure 10 – 1:4 Unun Coaxial and Toroidal Implementations

This technique can be extended to other transformation ratios. By adding extra transmission lines, other $1:n^2$ transformers can easily be achieved. Figure 11 shows a 1:9 transmission line schematic alongside an implementation using coaxial cable. 1:16 transformations can be realised by adding a third conductor pair and connecting them in the same way as the second.

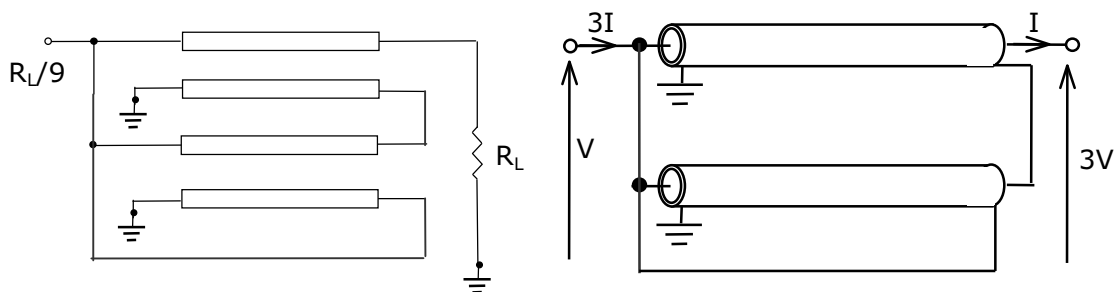


Figure 11 – 1:9 Ruthroff Unun

One of the prime factors limiting high frequency performance is the phase error caused by the arbitrary length of the transformers' interconnections. If these connections were made using a transmission line of the same length, velocity factor and impedance as the transformer line itself, then the phase error would be eliminated. An additional advantage is that the physical shape of the transformer is also no longer restricted by the need to bring connecting points close together. Devices of this type are referred to as equal delay transformers. Figure 12 shows how the principle is applied to the 1:4 Ruthroff Unun seen before in figure 9.

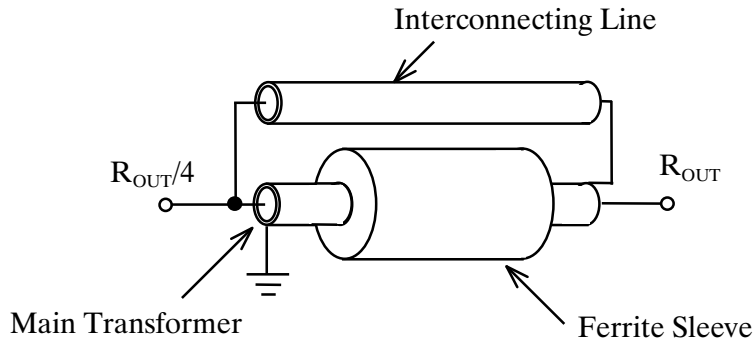


Figure 12 – 1:4 Equal Delay Ruthroff Unun

Note the absence of any magnetic material on the interconnecting line

In push-pull circuits, impedance transformations between a balanced source and balanced load are often required. By combining a number of basic building blocks in a range of parallel/series combinations balanced to balanced impedance ratios of 1:n² can be achieved. The diagrams below show 1:4 and 1:9 Guanella transmission line transformers based on this principle. Like the quarter-wavelength transformer, the optimum transmission line impedance is the geometric mean of input and output; however, small deviations from this can be permitted if some bandwidth degradation is acceptable.

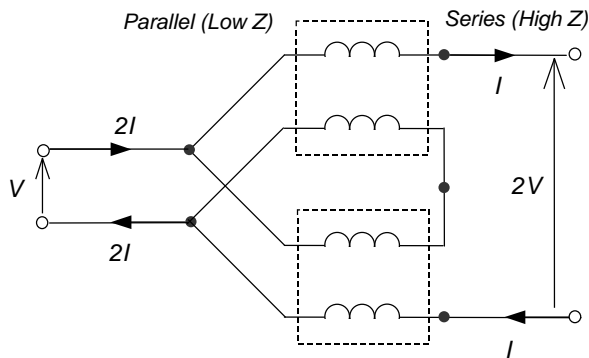


Fig 13a – 1:4 Balanced-Balanced Transformer

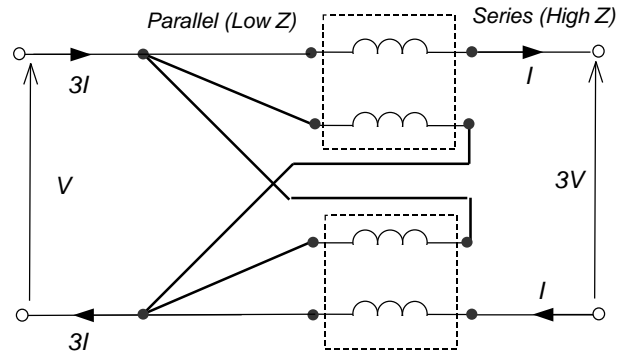


Figure 13b – 1:9 Balanced-Balanced Transformer

These transformers are often constructed of coaxial cable, and it is good practice to form the cable into a suitable shape to keep the interconnects as short as possible. The limitation of squared integer impedance ratios can be avoided to some extent by combining the conductors in

more complex arrangements, but the benefits of doing this must be weighed against the practicality of the design and the possible loss of bandwidth.

Appendix A – AWG Chart

AWG	Diameter (mm)	Diameter (mil)	Area (mm ²)	•/km (@20°C)
0	8.25	325	53.49	0.322
2	6.54	258	33.62	0.513
4	5.19	204	21.15	0.815
6	4.12	162	13.30	1.30
8	3.26	129	8.367	2.10
10	2.59	102	5.261	3.30
12	2.05	80.7	3.310	5.25
14	1.63	64.2	2.081	8.34
16	1.29	50.8	1.309	13.2
18	1.02	40.2	0.823	20.9
20	0.81	31.9	0.518	33.3
22	0.64	25.2	0.326	53.0
24	0.51	20.1	0.205	84.2
26	0.40	15.7	0.129	134
28	0.32	12.6	0.081	213
30	0.25	9.84	0.051	339
32	0.20	7.87	0.032	539
34	0.16	6.30	0.020	856
36	0.13	5.12	0.013	1361
38	0.10	3.94	0.008	2164
40	0.08	3.15	0.005	3340

Appendix B – Approximate Coaxial Cable Power Handling

Outer Diameter	Cable Type	1MHz	10MHz	100MHz	500MHz
1.7mm	Flexible	1kW	300W	90W	-
2.8mm	Flexible	1kW	800W	250W	-
1.1mm	Semi-Rigid	-	-	68W	32W
2.2mm	Semi-Rigid	-	-	330W	140W
6.4mm	Semi-Rigid	-	-	1.2kW	515W

Appendix C

C1 - Coaxial Cable Suppliers

Micro-Coax Pottstown, PA (+1-800-223-2629)
<http://www.micro-coax.com>

Gigalink Rotterdam, The Netherlands (+31-10-789-11-22)
<http://www.gigalink-mce.com>

NaF Technology Hawsung-City, Kyunggi-Do, Korea (+82-31-352-6671)
<http://www.naftech.co.kr>

Storm Products Santa Monica, CA (+1-310-319-5388)
<http://www.stormproducts.com>

C2 - Ferrite Suppliers

Eastern Components W. Conshohocken, PA (+1-610-825-8610)
<http://www.eastern-components.com>

Fair-Rite Wallkill, NY (+1-845-895-2055)
<http://www.fair-rite.com>

Magnetics Pittsburgh, PA (+1-412-696-1333)
<http://www.mag-inc.com>

Magnetics Group Bethlehem, PA (+1-610-867-7600)
<http://www.magneticsgroup.com>

C3 – Balun/Transformer Suppliers

XFMRs Camby, IN (+1-317-834-1066)

<http://www.xfmrs.com>

Coilcraft Europe Cumbernauld, Scotland (+44-123-673-0627)

<http://www.xfmrs.com>

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