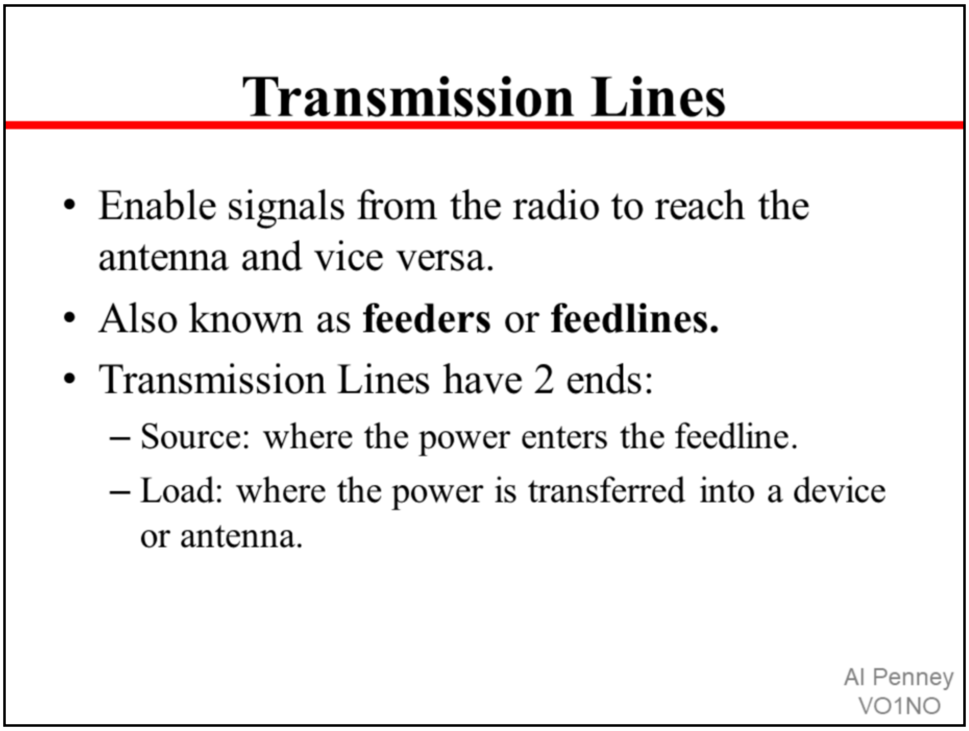

Transmission Lines
Advanced Course
VO1NO



In radio-frequency engineering, a **transmission line** is a specialized cable or other structure designed to conduct alternating current of radio frequency, that is, currents with a frequency high enough that their wave nature must be taken into account. Transmission lines are used for purposes such as connecting radio transmitters and receivers with their antennas (they are then called feed lines or feeders), distributing cable television signals, trunklines routing calls between telephone switching centres, computer network connections and high speed computer data buses.

Ordinary electrical cables suffice to carry low frequency alternating current (AC), such as mains power, which reverses direction 100 to 120 times per second, and audio signals. However, they cannot be used to carry currents in the radio frequency range, above about 30 kHz, because the energy tends to radiate off the cable as radio waves, causing power losses. Radio frequency currents also tend to reflect from discontinuities in the cable such as connectors and joints, and travel back down the cable toward the source. These reflections act as bottlenecks, preventing the signal power from reaching the destination. Transmission lines use specialized construction, and impedance matching, to carry electromagnetic signals with minimal reflections and power losses. The distinguishing feature of most transmission lines is that they have uniform cross sectional dimensions along their length, giving them a uniform *impedance*, called the characteristic impedance, to prevent reflections. Types of transmission line include parallel line (ladder line, twisted pair), coaxial cable, and planar transmission lines such as stripline and microstrip. The higher the frequency of electromagnetic waves moving through a given cable or medium, the shorter the wavelength of the waves. Transmission lines become necessary when the transmitted frequency's wavelength is sufficiently short that the length of the cable becomes a significant part of a wavelength.

At microwave frequencies and above, power losses in transmission lines become excessive, and waveguides are used instead, which function as "pipes" to confine and guide the electromagnetic waves. Some sources define waveguides as a type of transmission line; however, this article will not include them. At even higher frequencies, in the terahertz, infrared and visible ranges, waveguides in turn become lossy, and optical methods, (such as lenses and mirrors), are used to guide electromagnetic waves.

Optimum Transmission Line

- Does not radiate signal from the line itself.
- No loss of signal passing through the line.
- Constant electrical characteristics throughout its length.

- Unfortunately, there is no such thing as an ideal transmission line!

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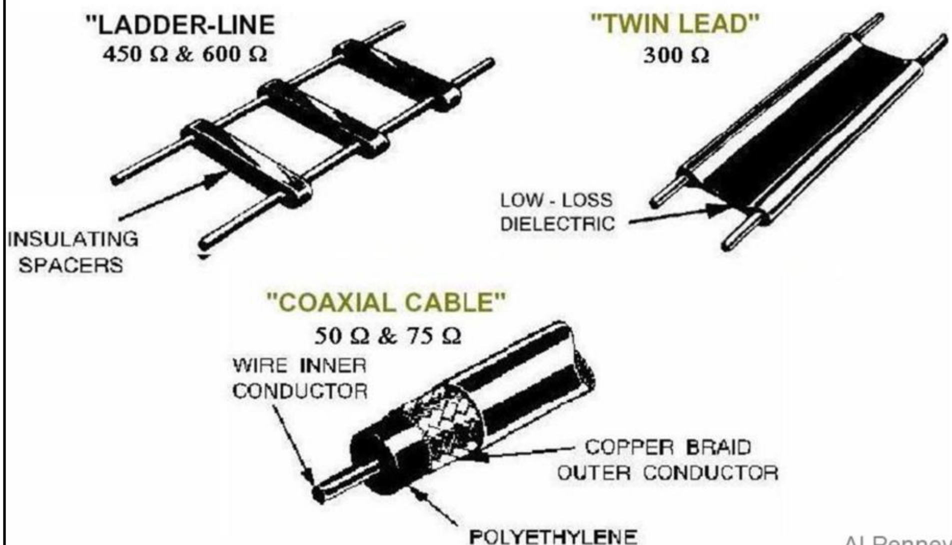
RF Transmission and Reception Feedline Types



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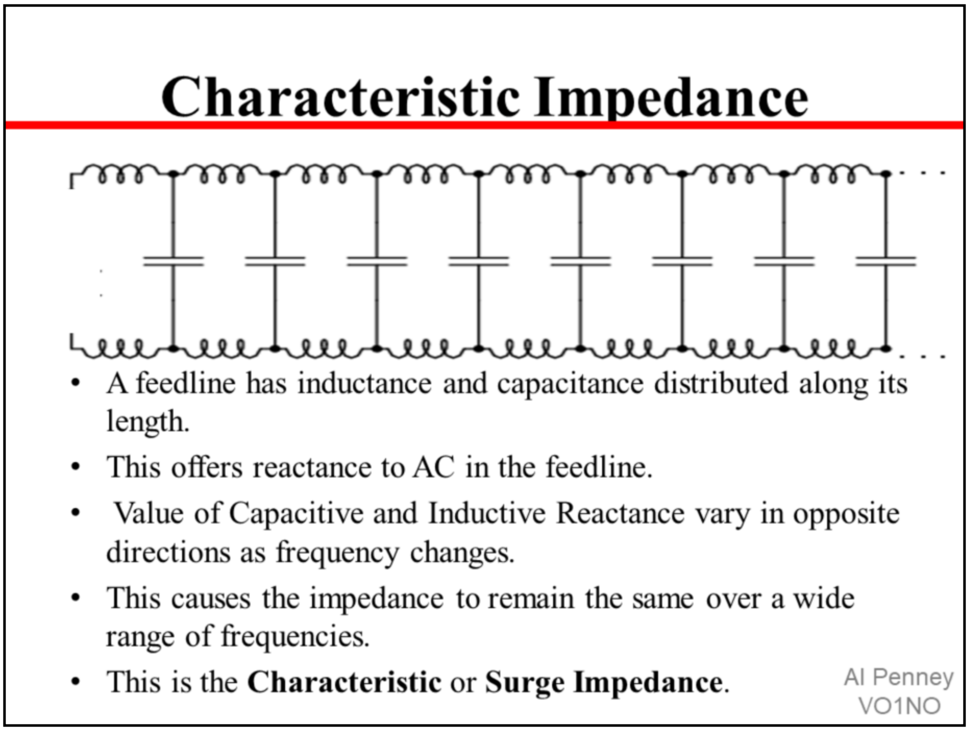

CHARACTERISTIC IMPEDANCE

If the line could be "perfect"—having no resistive losses—a question might arise: What is the amplitude of the current in a pulse applied to this line? Will a larger voltage result in a larger current, or

is the current theoretically infinite for an applied voltage, as we would expect from applying Ohm's Law to a circuit without resistance? The answer is that the current does depend directly on the voltage,

just as though resistance were present. The reason for this is that the current flowing in the line is something like the charging current that flows when a battery is connected to a capacitor. That is, the line has capacitance. However, it also has inductance. Both of these are "distributed" properties. We may think of the line as being composed of a whole series of small inductors and capacitors, connected as in **the figure**, where each coil is the inductance of an extremely small section of wire, and the capacitance is that existing between the same two sections. Each series inductor acts to limit the rate at which current can charge the following shunt capacitor, and in so doing establishes a very important property of a transmission line: its *surge impedance,* more commonly known as its *characteristic impedance*. This is abbreviated by convention as Z0.

Characteristic Impedance

- Abbreviated $\mathbf{Z_o}$
- Value depends on
  - the physical dimensions of the line; and
  - The relative positions of the conductors.
- $\mathbf{Z_o}$ = ratio of voltage to current at any given point.
- Value does not depend on length.

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Characteristic Impedance

- Characteristic Impedance is an AC effect – you cannot measure it using DC.
- Actual resistance losses are called **Copper Losses**.
- **Skin Effect** causes higher losses as frequency increases.

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Skin Effect

- Tendency of AC to become distributed within a conductor such that the current density is largest near the surface of the conductor, and decreases with greater depths in the conductor.
- The electric current flows mainly at the "skin" of the conductor, between the outer surface and a level called the **skin depth**.
- The skin effect causes the effective resistance to increase at higher frequencies where the skin depth is smaller, thus reducing the effective cross-section of the conductor.

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Skin effect is the tendency of an alternating electric current (AC) to become distributed within a conductor such that the current density is largest near the surface of the conductor, and decreases with greater depths in the conductor. The electric current flows mainly at the "skin" of the conductor, between the outer surface and a level called the **skin depth**. The skin effect causes the effective resistance of the conductor to increase at higher frequencies where the skin depth is smaller, thus reducing the effective cross-section of the conductor. The skin effect is due to opposing eddy currents induced by the changing magnetic field resulting from the alternating current. At 60 Hz in copper, the skin depth is about 8.5 mm. At high frequencies the skin depth becomes much smaller. Increased AC resistance due to the skin effect can be mitigated by using specially woven litz wire. Because the interior of a large conductor carries so little of the current, tubular conductors such as pipe can be used to save weight and cost.

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Cross-sectional area of a round conductor available for conducting DC current

"DC resistance"

Cross-sectional area of the same conductor available for conducting low-frequency AC

"AC resistance"

Cross-sectional area of the same conductor available for conducting high-frequency AC

"AC resistance"

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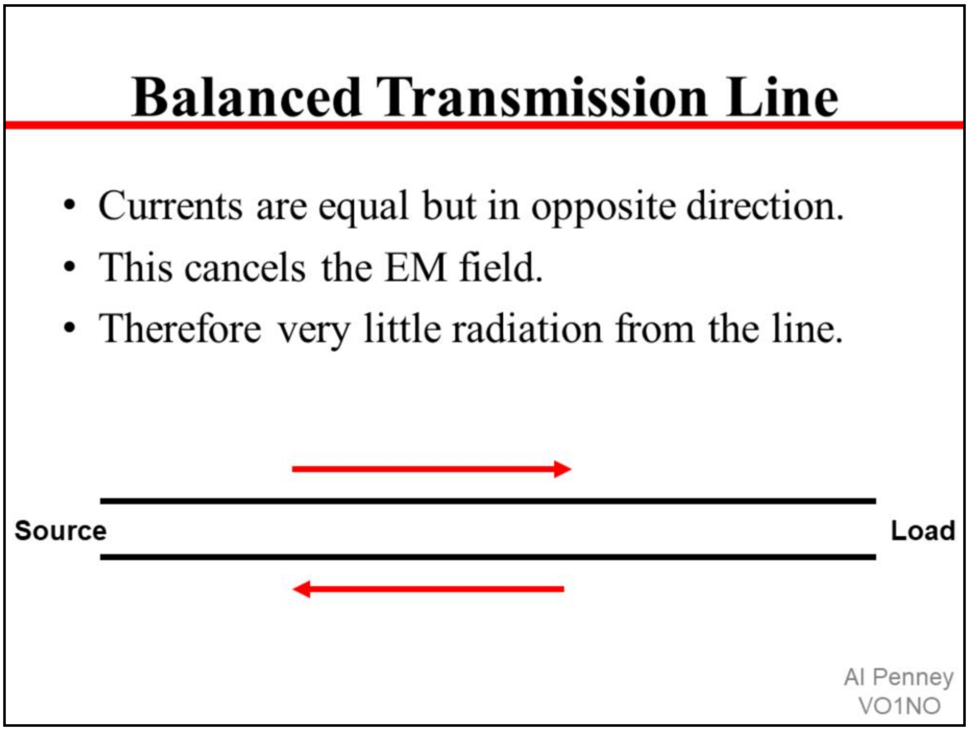


At radio frequencies, every conductor that has appreciable length compared with the wavelength in use *radiates* power¾every conductor is an antenna. Special care must be used, therefore, to minimize

radiation from the conductors used in RF transmission lines. Without such care, the power radiated by the line may be much larger than that which is lost in the resistance of conductors and dielectrics

(insulating materials). Power loss in resistance is inescapable, at least to a degree, but loss by radiation is largely avoidable.

Radiation loss from transmission lines can be prevented by using two conductors arranged and operated so the electromagnetic field from one is balanced everywhere by an equal and opposite field

from the other. In such a case, the resultant field is zero everywhere in space¾there is no radiation from the line. For example, two parallel conductors having currents I1 and I2 flowing in opposite

directions. If the current I1 at point Y on the upper conductor has the same amplitude as the current I2 at the corresponding point X on the lower conductor, the fields set up by the two currents are equal in

magnitude. Because the two currents are flowing in opposite directions, the field from I1 at Y is 180° out of phase with the field from I2 at X. However, it takes a measurable interval of time for the field from X to travel to Y. If I1 and I2 are alternating currents, the phase of the field from I1 at Y changes in such a time interval, so at the instant the field from X reaches Y, the two fields at Y are not exactly 180°

out of phase. The two fields are exactly 180° out of phase at every point in space only when the two conductors occupy the same space¾an obviouslyimpossible condition if they are to remain separate

conductors.

The best that can be done is to make the two fields cancel each other as completely as possible. This can be achieved by keeping the distance between the two conductors small enough so the

time interval during which the field from X is moving to Y is a very small part of a cycle. When this is the case, the phase difference between the two fields at any given point is so close to 180°

that cancellation is nearly complete.

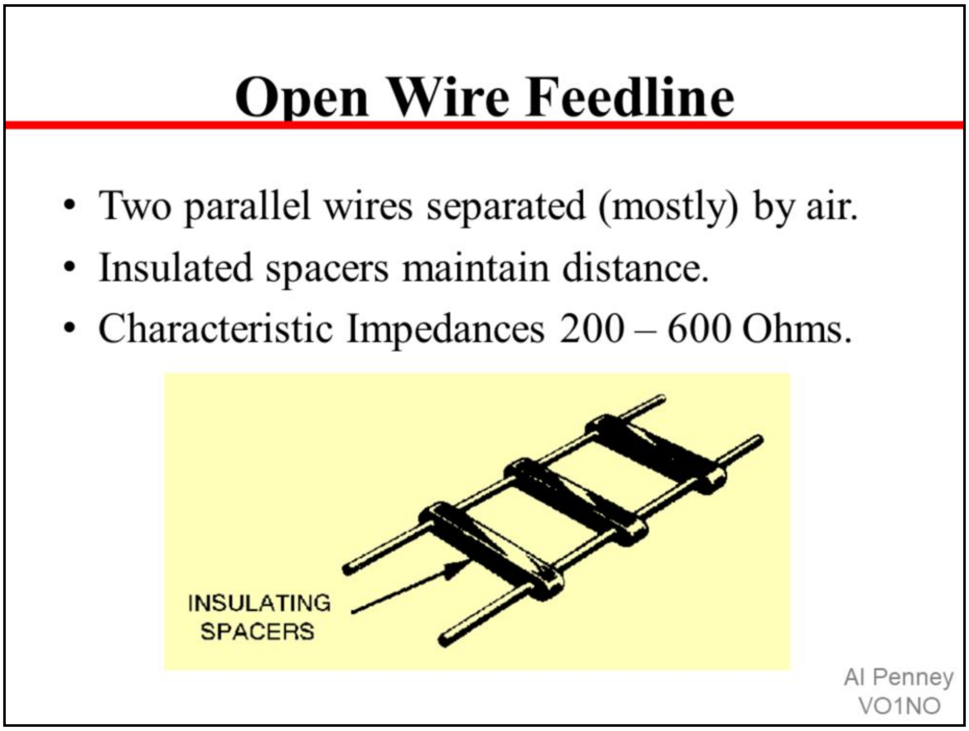
Practical values of d (the separation between the two conductors) are determined by the physical limitations of line construction. A separation that meets the condition of being “very small” at one frequency may be quite large at another. For example,

if d is 6 inches, the phase difference between the two fields at Y is only a fraction of a degree if the frequency is 3.5 MHz. This is because a distance of 6 inches is such a small fraction of a wavelength

(1 λ = 281 feet) at 3.5 MHz. But at 144 MHz, the phase difference is 26°, and at 420 MHz, it is 77°. In neither of these cases could the two fields be considered to “cancel” each other. Conductor separation

must be very small in comparison with the wavelength used; it should never exceed 1% of the wavelength, and smaller separations are desirable. Transmission lines consisting of two parallel conductors

as in Fig 1A are called *open-wire lines*, *parallel-conductor lines* or *two-wire lines*.



Parallel feeders go back to the beginnings of radio. By 1930, the "two-wire untuned feeder system" was a standard ARRL *Handbook* feature. The *Jones Radio Handbook* of 1937 provides a table of line losses showing the advantages of open-wire feeders (a 440-Ohm line in the table) over lower impedance twisted-pair feeders (p. 70). The use of 600-Ohm lines was fairly standard, using a spacing of about 6". "To reduce radiation from the feeders to a minimum, the two wires should not be more than 10 to 12 inches apart." (*The Radio Amateur's Handbook*, 7th Ed., ARRL, 1930, p. 162) Rarely did hams exceed the 6" spacing.

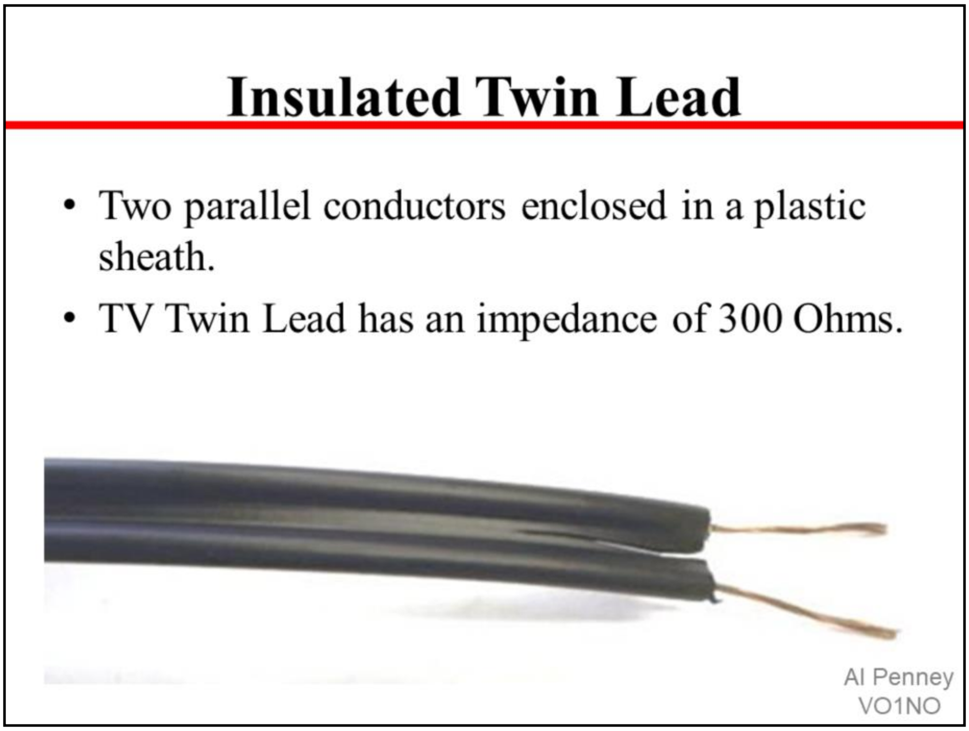
Every transmission line has a characteristic impedance, and parallel transmission lines are no exception. The characteristic impedance (Zo) of a line depends on the physical properties of the line. For a 2-wire set, we have only two properties of note (assuming the use of a very conductive material, such as copper): the diameter of the wire and the spacing between the wires.

Twin lead is a form of parallel-wire balanced transmission line. The separation between the two wires in twin-lead is small compared to the wavelength of the radio frequency (RF) signal carried on the wire.[3] The RF current in one wire is equal in magnitude and opposite in direction to the RF current in the other wire. Therefore, in the far field region far from the transmission line, the radio waves radiated by one wire are equal in magnitude but opposite in phase (180° out of phase) to the waves radiated by the other wire, so they superpose and cancel each other. The result is that almost no net radio energy is radiated by the line.

Similarly, any interfering external radio waves will induce equal, in phase RF currents, traveling in the same direction, in the two wires. Since the load at the destination end is connected *across* the wires, only *differential*, oppositely-directed currents in the wires create a current in the load. Thus the interfering currents are canceled out, so twin lead does not tend to pick up radio noise.

However, if a piece of metal is located sufficiently close to a twin-lead line, within a distance comparable to the wire spacing, it will be significantly closer to one wire than the other. As a result, the RF current induced in the metal object by one wire will be greater than the opposing current induced by the other wire, so the currents will no longer cancel. Thus nearby metal objects can cause power losses in twin lead lines, through energy dissipated as heat by induced currents. Similarly, radio noise originating in cables or metal objects located near the twin-lead line can induce unbalanced currents in the wires, coupling noise into the line.

In order to prevent power from being reflected from the load end of the line, causing high SWR and inefficiency, the load must have an impedance which matches the characteristic impedance of the line. This causes the load to appear electrically identical to a continuation of the line, preventing reflections. Similarly, to transfer power efficiently into the line, the source must also match the characteristic impedance. To connect balanced transmission line to unbalanced line like coaxial cable, a device called a balun must be used.



win-lead cable is a two-conductor flat cable used as a balanced transmission line to carry radio frequency (RF) signals. It is constructed of two stranded copper or copper-clad steel wires, held a precise distance apart by a plastic (usually polyethylene) ribbon. The uniform spacing of the wires is the key to the cable's function as a transmission line; any abrupt changes in spacing would reflect some of the signal back toward the source. The plastic also covers and insulates the wires.

Twin lead can have significantly lower signal loss than miniature flexible coaxial cable at shortwave and VHF radio frequencies; for example, type RG-58 coaxial cable loses 6.6 dB per 100 m at 30 MHz, while 300 ohm twin-lead loses only 0.55 dB. However, twin-lead is more vulnerable to interference. Proximity to metal objects will inject signals into twin-lead that would be blocked out by coaxial cable. Twin lead therefore requires careful installation around rain gutters, and standoffs from metal support masts. Twin-lead is also susceptible to significant degradation when wet or ice covered, whereas coax is less or not affected in these conditions. For these reasons, coax has largely replaced twin-lead in most uses, except where maximum signal is required.

Ladder Line

- Variant of Insulated Twin Lead.
- To reduce losses, some of the plastic is cut away.
- Characteristic Impedance of 450 Ohms.

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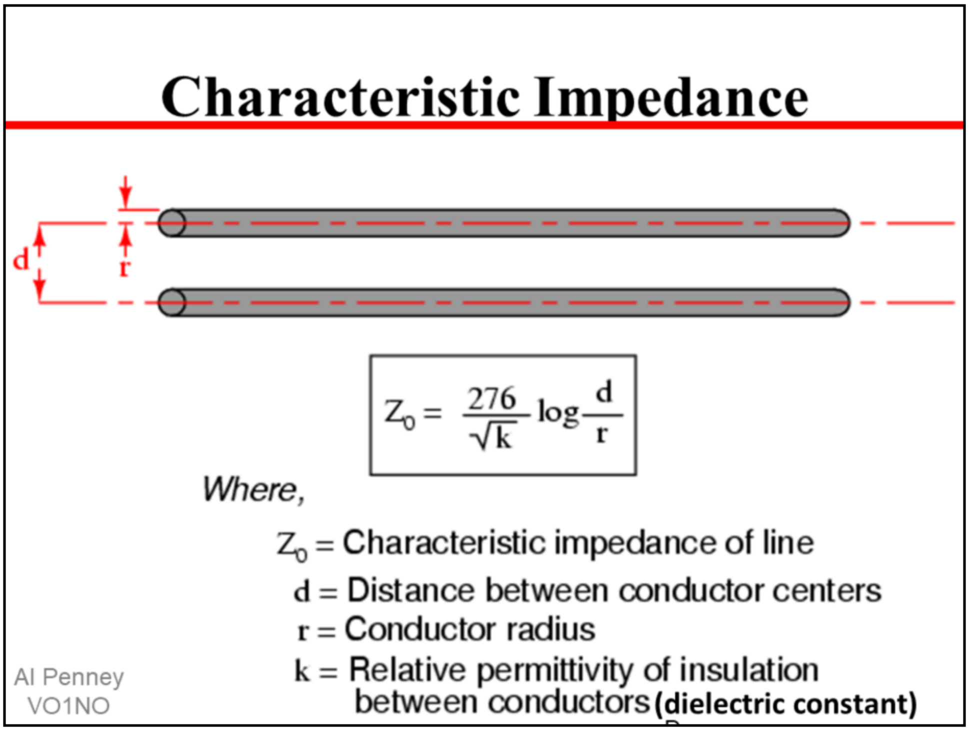
Ladder line or "window line" is a variation of twin lead which is constructed similarly, except that the polyethylene webbing between the wires which holds them apart has rectangular openings ("windows") cut in it. The line consists of two insulated wires with "rungs" of plastic holding them together every few inches, giving it the appearance of a ladder. The advantage of the "windows" is that they lighten the line, and also reduce the amount of surface on which dirt and moisture can accumulate, making ladder line less vulnerable to weather-induced changes in characteristic impedance. The most common type is 450 ohm ladder line, which has a conductor spacing of about an inch.



Identical units of measurement must be used in both terms of the fraction.

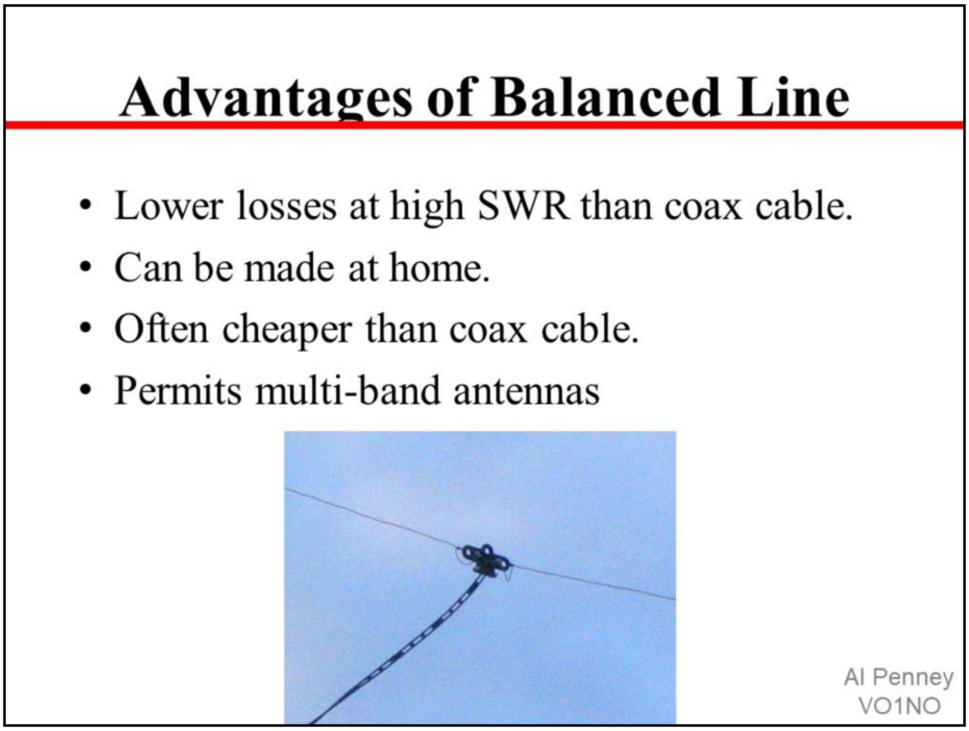
k = dielectric constant of insulation between conductors. For air it is 1.00059 (round off to 1 for practical applications).

The dielectric constant k is the relative permittivity of a dielectric material. It is an important parameter in characterizing capacitors. It is unfortunate that the same symbol k is often used for Coulomb's constant, so one must be careful of this possible confusion.

The **relative permittivity** of a material is its (absolute) permittivity expressed as a ratio relative to the vacuum permittivity.

Permittivity is a material property that affects the Coulomb force between two point charges in the material. Relative permittivity is the factor by which the electric field between the charges is decreased relative to vacuum.

Likewise, relative permittivity is the ratio of the capacitance of a capacitor using that material as a dielectric, compared with a similar capacitor that has vacuum as its dielectric. Relative permittivity is also commonly known as the **dielectric constant**, a term still used but deprecated by standards organizations in engineering as well as in chemistry.



Open-wire line has the advantage of both lower loss and lower cost compared to coax. 600-Ω open-wire line at 30 MHz has a matched loss of only 0.1 dB. If you use such open-wire

line with the same 5:1 SWR, the total loss would still be less than 0.3 dB. In fact, even if the SWR rose to 20:1, the total loss would be less than 1 dB. Typical open-wire line sells for about 1/3 the cost of good quality coax cable.

Open-wire line is enjoying a renaissance of sorts with amateurs wishing to cover multiple HF bands with a single-wire antenna. This is particularly true since the bands at 30, 17 and 12 meters

became available in the early 1980s. The 102-foot long “G5RV dipole,” fed with open-wire ladder line into an antenna tuner, has become popular as a simple all-band antenna. The simple 130-foot long flattop

dipole, fed with open-wire 450-Ω “window” ladder-line, is also very popular among all-band enthusiasts.

Despite their inherently low-loss characteristics, open-wire lines are not often employed above about 100 MHz. This is because the physical spacing between the two wires begins to become an

appreciable fraction of a wavelength, leading to undesirable radiation by the line itself. Some form of coaxial cable is almost universally used in the VHF and UHF amateur bands.

Disadvantages of Balanced Line

- Spacing must be kept constant.
- Cannot be buried, laid on ground, or run alongside a conductor.
- Impedance varies in rain and with icing.
- Safety hazard if touched while transmitting.
- Impedance higher than radio antenna terminal, so an impedance matching device is necessary.
- Spacing on VHF/UHF is an appreciable fraction of the wavelength, so line radiates.

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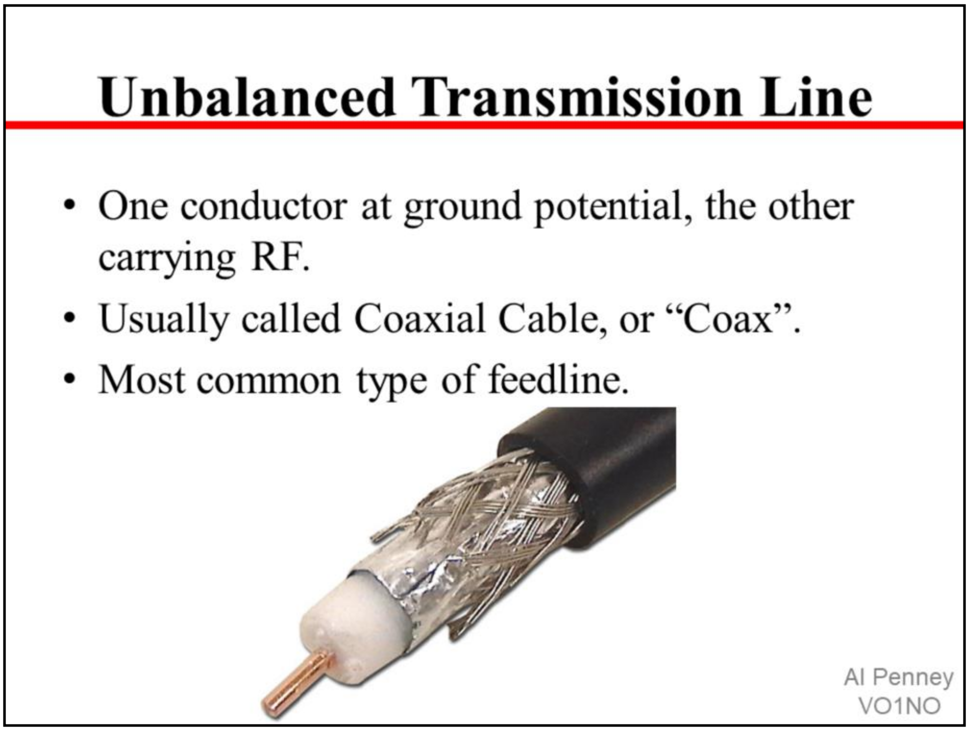


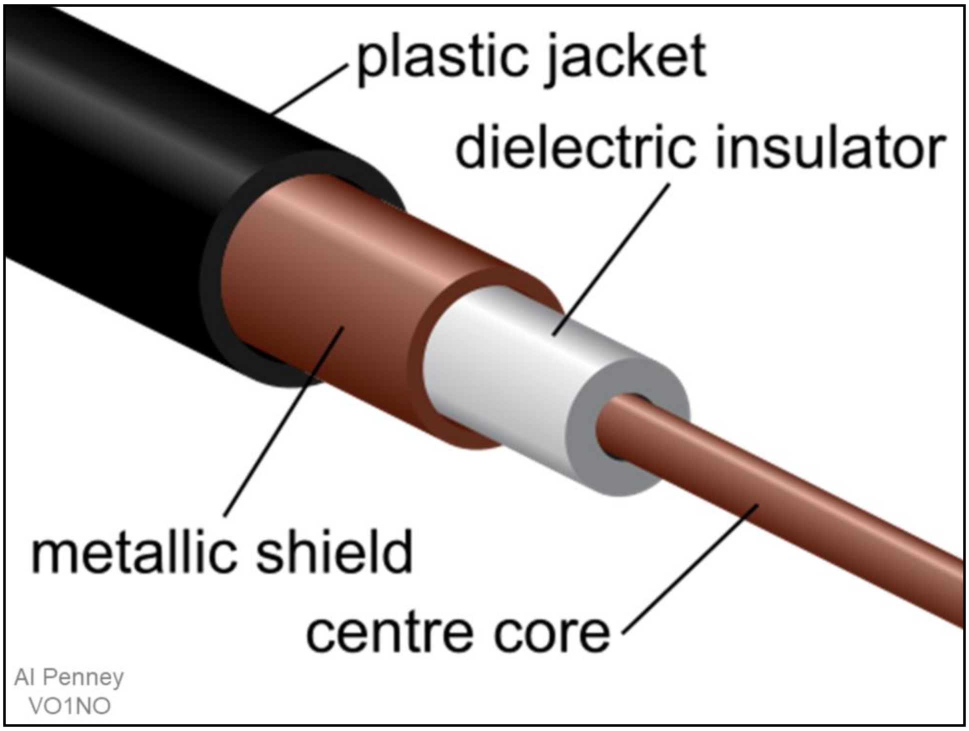
An **unbalanced line** is a transmission line, often coaxial cable, whose conductors have unequal impedances with respect to ground; as opposed to a balanced line. Microstrip and single-wire lines are also unbalanced lines.

Any line that has a different impedance of the return path may be considered an unbalanced line. However, unbalanced lines usually consist of a conductor that is considered the signal line and another conductor that is grounded, or is ground itself. The ground conductor often takes the form of a ground plane or the screen of a cable. The ground conductor may be, and often is, common to multiple independent circuits. For this reason the ground conductor may be referred to as *common*.

A coaxial line (coax) has a central signal conductor surrounded by a cylindrical shielding conductor. The shield conductor is normally grounded. The coaxial format was developed during World War II for use in radar. It was originally constructed from rigid copper pipes, but the usual form today is a flexible cable with a braided screen. The advantages of coax are a theoretically perfect electrostatic screen and highly predictable transmission parameters. The latter is a result of the fixed geometry of the format which leads to a precision not found with loose wires. Open wire systems are also affected by nearby objects altering the field pattern around the conductor. Coax does not suffer from this since the field is entirely contained within the cable due to the surrounding screen.

Coaxial lines are the norm for connections between radio transmitters and their antennae, for interconnection of electronic equipment where high frequency or above is involved, and were formerly widely used for forming local area networks before twisted pair became popular for this purpose.

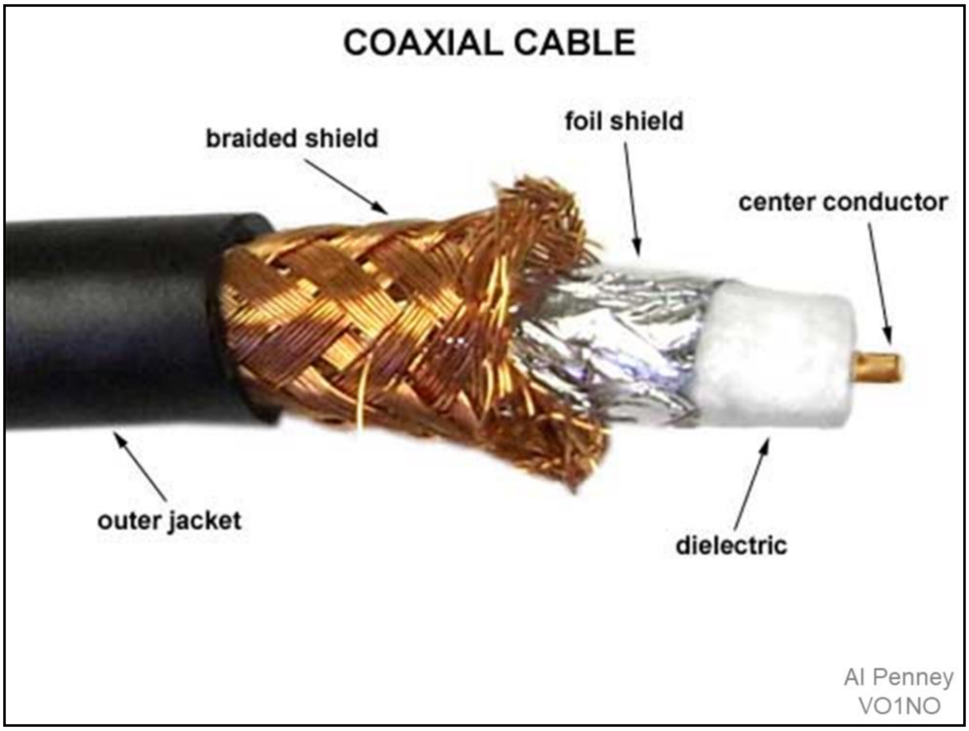
Triaxial cable (triax) is a variant of coax with a second shield conductor surrounding the first with a layer of insulation in between. As well as providing additional shielding, the outer conductors can be used for other purposes such as providing power to equipment or control signals. Triax is widely used for the connection of cameras in television studios.



Coaxial cable conducts electrical signal using an inner conductor (usually a solid copper, stranded copper or copper plated steel wire) surrounded by an insulating layer and all enclosed by a shield, typically one to four layers of woven metallic braid and metallic tape. The cable is protected by an outer insulating jacket. Normally, the shield is kept at ground potential and a signal carrying voltage is applied to the center conductor. The advantage of coaxial design is that electric and magnetic fields are restricted to the dielectric with little leakage outside the shield. Further, electric and magnetic fields outside the cable are largely kept from interfering with signals inside the cable. This property makes coaxial cable a good choice for carrying weak signals that cannot tolerate interference from the environment or for stronger electrical signals that must not be allowed to radiate or couple into adjacent structures or circuits. Larger diameter cables and cables with multiple shields have less leakage.

Common applications of coaxial cable include video and CATV distribution, RF and microwave transmission, and computer and instrumentation data connections.

The characteristic impedance of the cable is determined by the dielectric constant of the inner insulator and the radii of the inner and outer conductors. In radio frequency systems, where the cable length is comparable to the wavelength of the signals transmitted, a uniform cable characteristic impedance is important to minimize loss. The source and load impedances are chosen to match the impedance of the cable to ensure maximum power transfer and minimum standing wave ratio. Other important properties of coaxial cable include attenuation as a function of frequency, voltage handling capability, and shield quality.



Coaxial cable design choices affect physical size, frequency performance, attenuation, power handling capabilities, flexibility, strength, and cost. The inner conductor might be solid or stranded; stranded is more flexible. To get better high-frequency performance, the inner conductor may be silver-plated. Copper-plated steel wire is often used as an inner conductor for cable used in the cable TV industry.

The insulator surrounding the inner conductor may be solid plastic, a foam plastic, or air with spacers supporting the inner wire. The properties of the dielectric insulator determine some of the electrical properties of the cable. A common choice is a solid polyethylene (PE) insulator, used in lower-loss cables. Solid Teflon (PTFE) is also used as an insulator, and exclusively in plenum-rated cables. Some coaxial lines use air (or some other gas) and have spacers to keep the inner conductor from touching the shield.

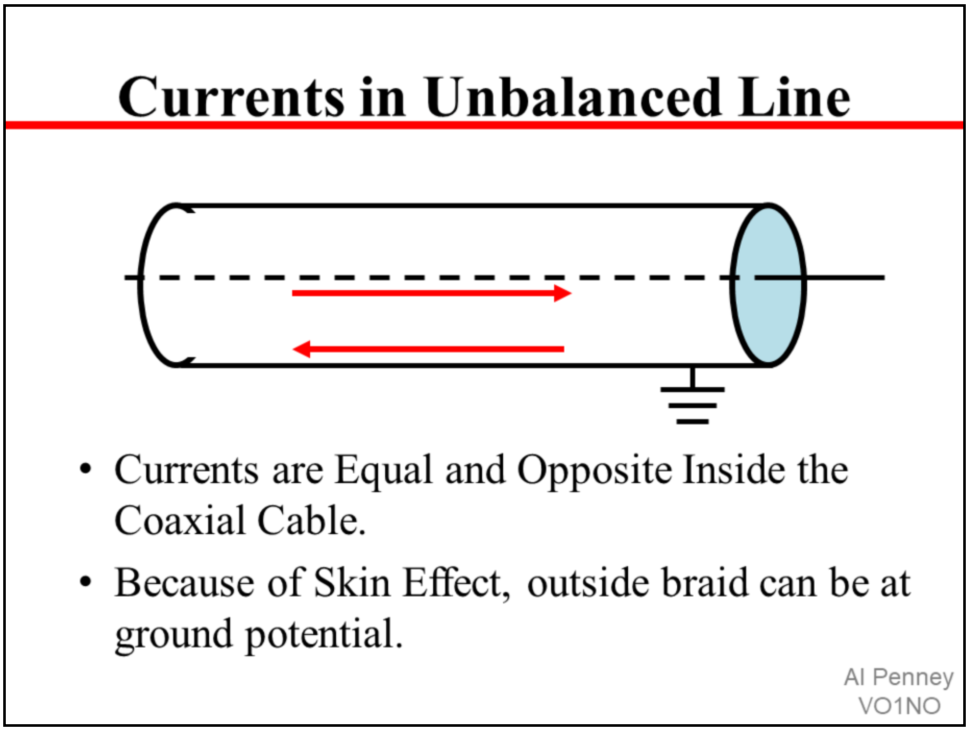
Many conventional coaxial cables use braided copper wire forming the shield. This allows the cable to be flexible, but it also means there are gaps in the shield layer, and the inner dimension of the shield varies slightly because the braid cannot be flat. Sometimes the braid is silver-plated. For better shield performance, some cables have a double-layer shield. The shield might be just two braids, but it is more common now to have a thin foil shield covered by a wire braid. Some cables may invest in more than two shield layers, such as "quad-shield", which uses four alternating layers of foil and braid. Other shield designs sacrifice flexibility for better performance; some shields are a solid metal tube. Those cables cannot be bent sharply, as the shield will kink, causing losses in the cable. When a foil shield is used a small wire conductor incorporated into the foil makes soldering the shield termination easier.

For high-power radio-frequency transmission up to about 1 GHz, coaxial cable with a solid copper outer conductor is available in sizes of 0.25 inch upward. The outer conductor is corrugated like a bellows to permit flexibility and the inner conductor is held in position by a plastic spiral to approximate an air dielectric. One brand name for such cable is *Heliax*.

Coaxial cables require an internal structure of an insulating (dielectric) material to maintain the spacing between the center conductor and shield.
The dielectric losses increase in this order: Ideal dielectric (no loss), vacuum, air, polytetrafluoroethylene (PTFE), polyethylene foam, and solid polyethylene. A low relative permittivity allows for higher-frequency usage. An inhomogeneous dielectric needs to be compensated by a non-circular conductor to avoid current hot-spots.

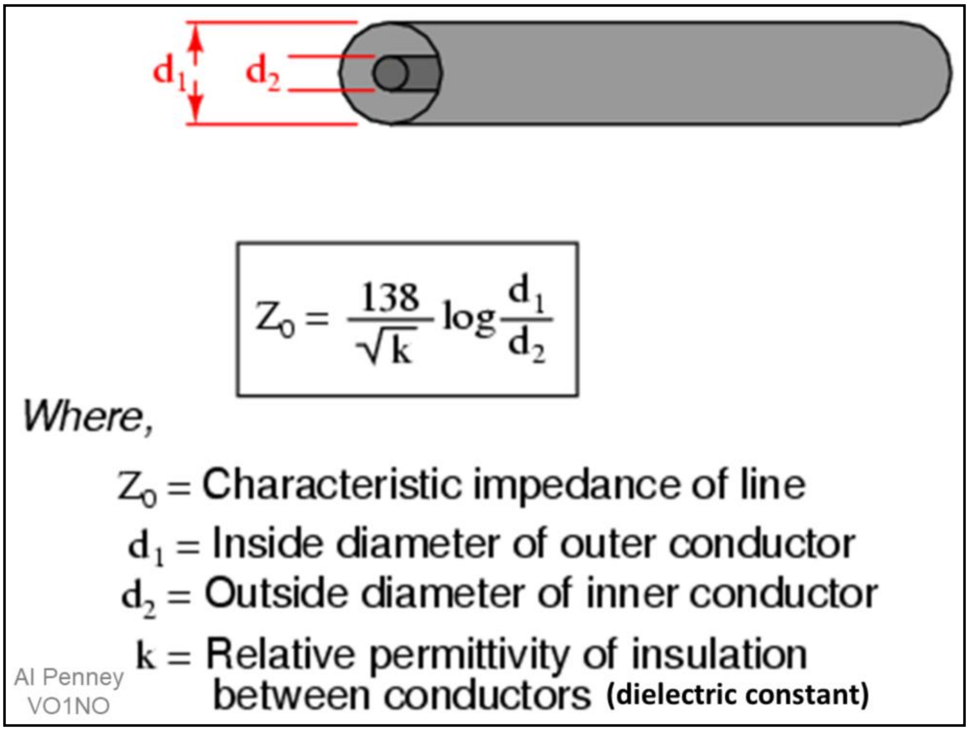
While many cables have a solid dielectric, many others have a foam dielectric that contains as much air or other gas as possible to reduce the losses by allowing the use of a larger diameter center conductor. Foam coax will have about 15% less attenuation but some types of foam dielectric can absorb moisture—especially at its many surfaces — in humid environments, significantly increasing the loss. Supports shaped like stars or spokes are even better but more expensive and very susceptible to moisture infiltration. Still more expensive were the air-spaced coaxials used for some inter-city communications in the mid-20th century. The center conductor was suspended by polyethylene discs every few centimeters. In some low-loss coaxial cables such as the RG-62 type, the inner conductor is supported by a spiral strand of polyethylene, so that an air space exists between most of the conductor and the inside of the jacket. The lower dielectric constant of air allows for a greater inner diameter at the same impedance and a greater outer diameter at the same cutoff frequency, lowering ohmic losses. Inner conductors are sometimes silver-plated to smooth the surface and reduce losses due to skin effect. A rough surface extends the current path and concentrates the current at peaks, thus increasing ohmic loss.

The insulating jacket can be made from many materials. A common choice is PVC, but some applications may require fire-resistant materials. Outdoor applications may require the jacket to resist ultraviolet light, oxidation, rodent damage, or direct burial. Flooded coaxial cables use a water blocking gel to protect the cable from water infiltration through minor cuts in the jacket. For internal chassis connections the insulating jacket may be omitted.



The center and inside of the shield carry equal and opposite direction RF currents. This ALWAYS is the case when the shield is several [skin depths] thick. We cannot force anything else to happen!

In the drawing above and below, the outside of the shield is isolated by skin effect. It behaves like a separate transmission outer conductor. Skin effect prevents any current, voltage, or field (even magnetic) from penetrating the shield when the shield is many skin depths thick. Only the breaks at the ends connect the inner and outer shield conduction layers.



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Advantages of Unbalanced Line

- Can be run alongside metal or buried.
- Same impedance as that required by our radios.
- Convenient to use.
- Weatherproof.

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Disadvantages of Unbalanced Line

- Higher losses than Balanced Feedline
- Losses increase with higher SWR.
- Water ingress a danger to coax cable.
- Kinking or bending too sharply can cause damage.
- Some connectors cause impedance “bumps”.

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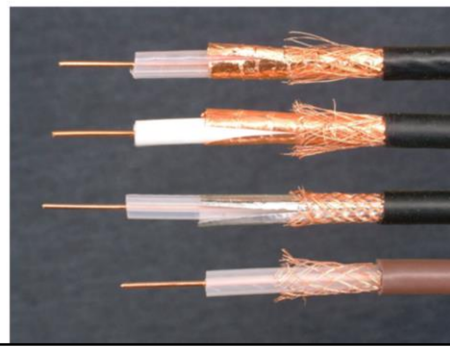
Beware Cheap Coax Cable!

- Poor quality coax cable has poor braid coverage.
- This causes signal attenuation and is susceptible to interference.



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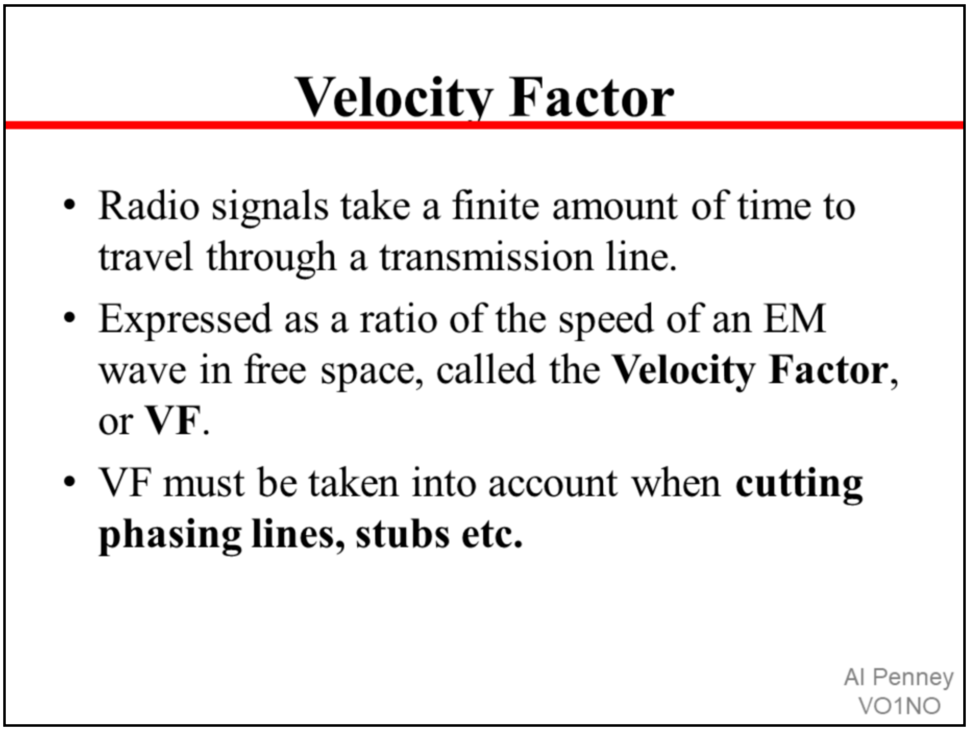
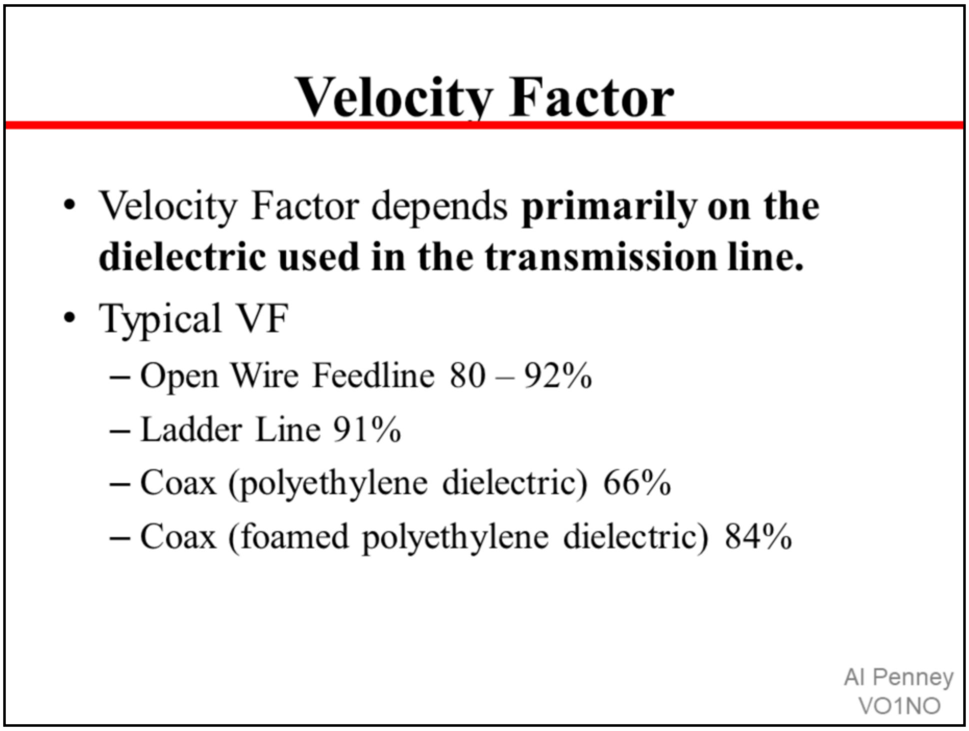


The **velocity factor** (**VF**), also called **wave propagation speed** or **velocity of propagation** (**VoP** or) of a transmission medium is the ratio of the speed at which a wavefront (of an electromagnetic signal, a radio signal, a light pulse in an optical fibre or a change of the electrical voltage on a copper wire) passes through the medium, to the speed of light in a vacuum. For optical signals, the velocity factor is the reciprocal of the refractive index.

The speed of radio signals in a vacuum, for example, is the speed of light, and so the velocity factor of a radio wave in a vacuum is unity, or 100%. In electrical cables, the velocity factor mainly depends on the insulating material.



Dielectric material codes

•FPE is foamed polyethylene

•PE is solid polyethylen

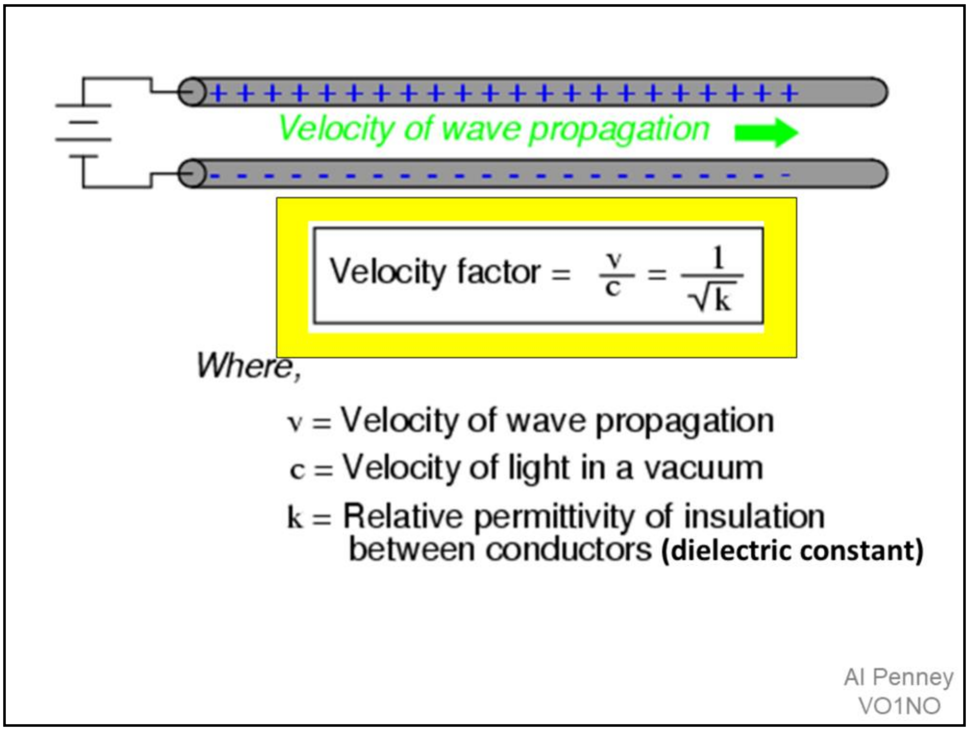
•PF is polyethylene foam

•PTFE is polytetrafluoroethylene;

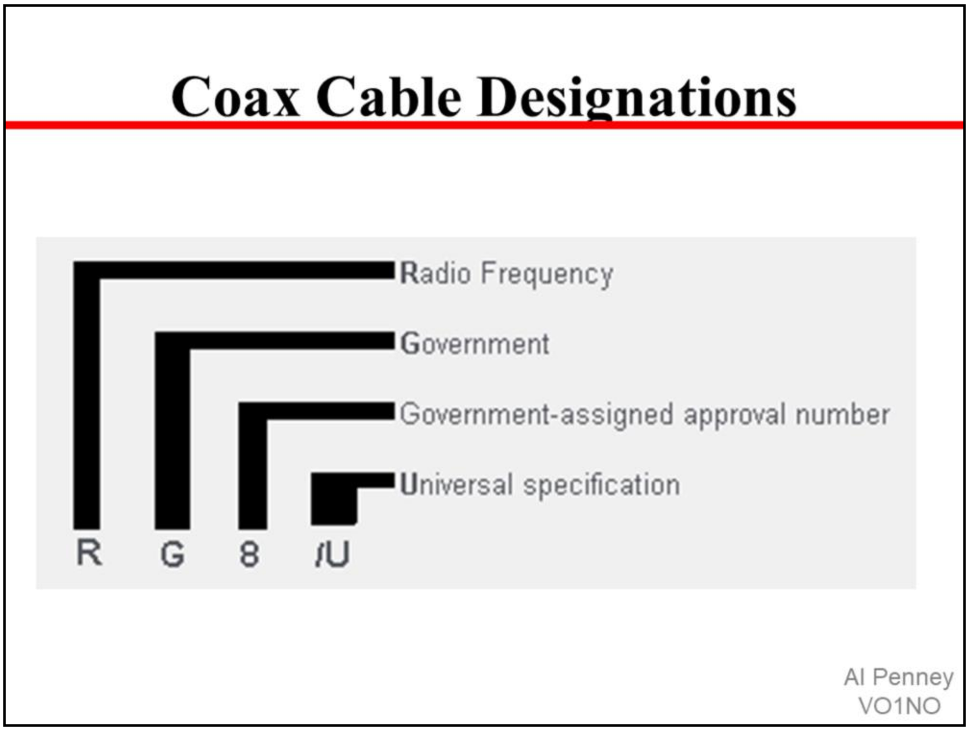
•ASP is air space polyethylene[

VF is the Velocity Factor

- •VF for solid PE is about 0.66
- •VF for foam PE is about 0.78 to 0.88
- •VF for air is about 1.00
- •VF for solid PTFE is about 0.70
- •VF for foam PTFE is about 0.84



The ratio of a transmission line's true propagation velocity and the speed of light in a vacuum is called the *velocity factor* of that line. Velocity factor is purely a factor of the insulating material's relative permittivity (otherwise known as its *dielectric constant*), defined as the ratio of a material's electric field permittivity to that of a pure vacuum. The velocity factor of any cable type—coaxial or otherwise—may be calculated quite simply by the formula in the slide:



Coaxial cables that conform to U.S. Government specifications are identified with an RG designation.
The meaning of the individual components of the designation are:

If the letters A, B or C appear before the slash (/) it indicates a specification-modification or revision. As an example, RG 8/U is superseded by RG 8A/U.

A series of standard types of coaxial cable were specified for military uses, in the form "RG-#" or "RG-#/U". They date from World War II and were listed in *MIL-HDBK-216* published in 1962. These designations are now obsolete. The RG designation stands for Radio Guide; the U designation stands for Universal. The current military standard is MIL-SPEC MIL-C-17. MIL-C-17 numbers, such as "M17/75-RG214", are given for military cables and manufacturer's catalog numbers for civilian applications. However, the RG-series designations were so common for generations that they are still used, although critical users should be aware that since the handbook is withdrawn there is no standard to guarantee the electrical and physical characteristics of a cable described as "RG-# type". The RG designators are mostly used to identify compatible connectors that fit the inner conductor, dielectric, and jacket dimensions of the old RG-series cables.

RG58/U

- 50 Ohm cable.
- Lightweight – diameter of a pencil.
- Okay for HF, but not long runs on VHF/UHF.
- Max power is a few hundred watts.

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RG-58/U is a type of coaxial cable often used for low-power signal and RF connections. The cable has a characteristic impedance of either 50 or 52 Ω. "RG" was originally a unit indicator for bulk RF cable in the U.S. military's Joint Electronics Type Designation System. There are several versions covering the differences in core material (solid or braided wire) and shield (70% to 95% coverage).

The outside diameter of RG-58 is around 0.2 inches (5 mm). RG-58 weighs around 0.025 lb/ft (37 g/m), exhibits approximately 25 pF/ft (82 pF/m) capacitance and can tolerate a maximum of 300 V potential (1800 W). Plain RG-58 cable has a solid center conductor. The RG-58A/U features a flexible 7- or 19-strand center conductor.

Most two-way radio communication systems, such as marine, CB radio, amateur, police, fire, WLAN antennas etc., are designed to work with a 50 Ω cable.

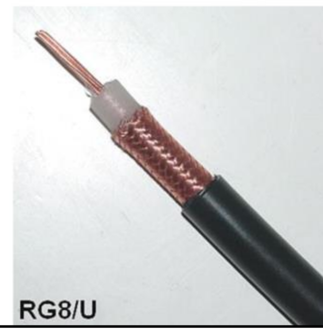
RG-58 cable is often used as a generic carrier of signals in laboratories, combined with BNC connectors that are common on test and measurement equipment such as oscilloscopes.

RG-58 in versions RG-58A/U or RG-58C/U was once widely used in "thin" Ethernet (10BASE2), for which it provides a maximum segment length of 185 meters. However, it has been almost completely replaced by twisted-pair cabling such as Cat 5, Cat 6, and similar cables in data networking applications.

RG-58 cable can be used for moderately high frequencies. Its signal attenuation depends on the frequency, e.g. from 10.8 dB per 100 m (3.3 dB per 100 feet) at 50 MHz to 70.5 dB per 100 m (21.5 dB per 100 feet) at 1 GHz.

RG8/U

- Used to be the standard 50 Ohm cable.
- Influx of cheap "RG8 Type" cables caused US Government to change to RG213 as the standard 50 Ohm cable.
- Diameter ~ 1 cm.
- Much lower loss than RG-58.
- Full legal limit on HF.



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RG213/U

- Quality 50 Ohm cable.
- Full legal limit on HF.
- RG214/U is identical but has 2 shields.

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RG 213 is a flexible coaxial cable that can be used in a number of commercial and military applications. With a 50 Ohm impendance and PVC jacket this RG 213 coax cable has low signal loss and high operation voltage for atenna cable applications. You will find RG 213 in a lot of VHF and UHF applications. Allied Wire & Cable also carries a mil-spec equivalent to RG 213 that can be found at M17/74-RG213.

RG59/U

- 75 Ohm cable meant primarily for TV.
- Suitable for low power use on HF – 400 watts.
- Also good for specialized receive antennas.
- RG11 is a higher power version.

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Low Loss Coax Cable

- Use foam dielectric or plastic spacers to reduce dielectric losses.
- Also have larger center conductors to reduce copper losses.
- Rigid shield cable is called Hardline.
- Not required on HF, but necessary for weak signal or high power applications at VHF/UHF and up.

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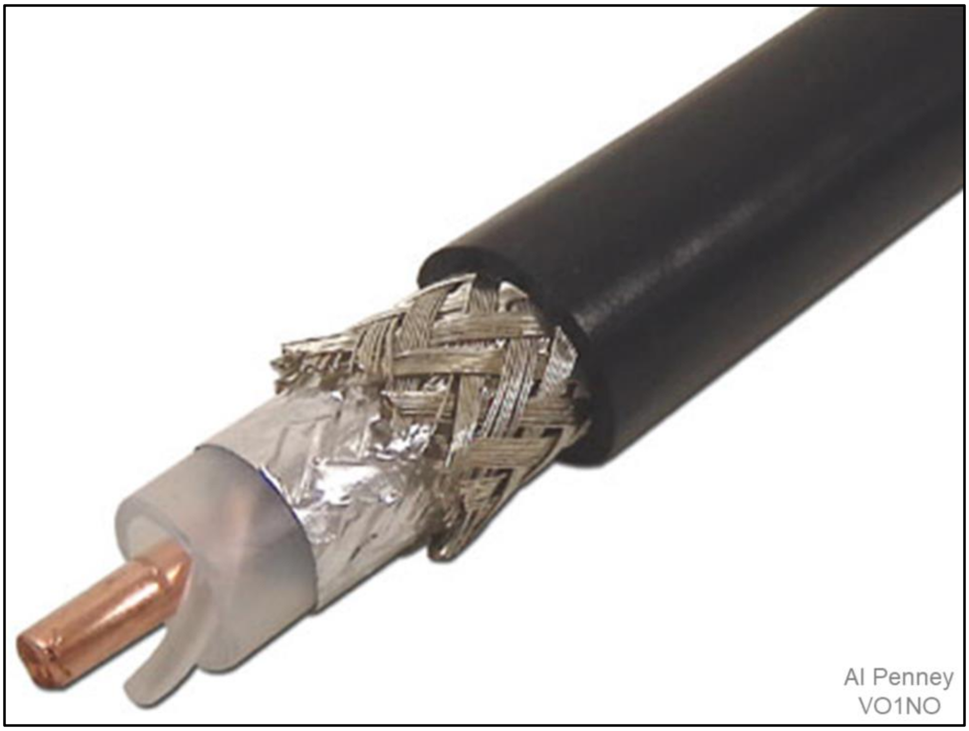
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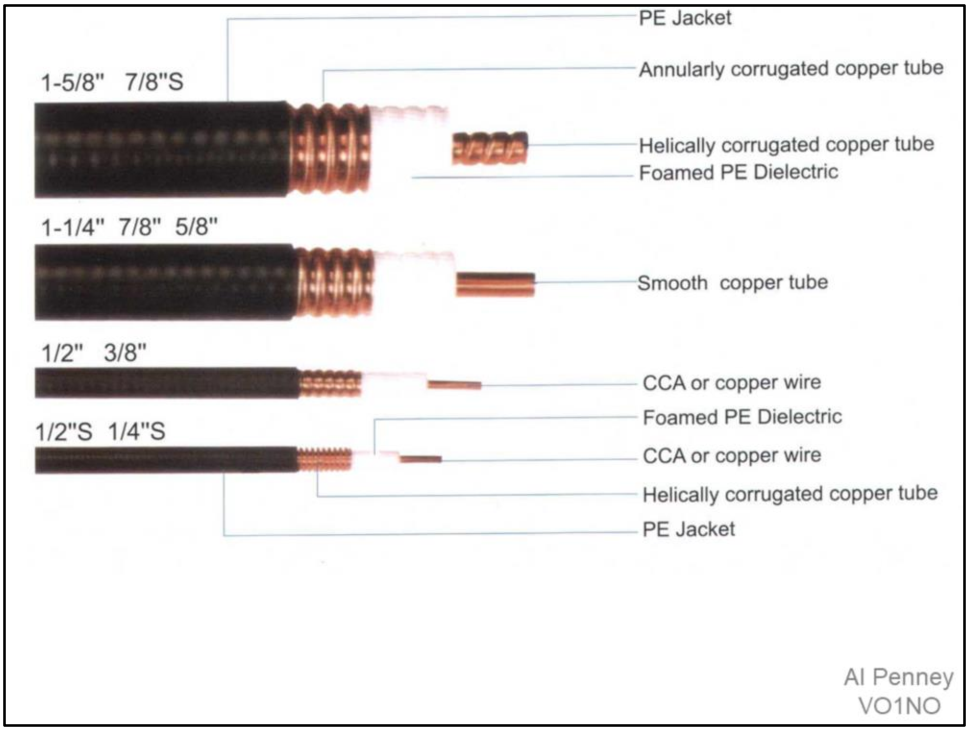
PE Jacket
1-5/8"   7/8"S
Annularly corrugated copper tube
Helically corrugated copper tube
Foamed PE Dielectric
1-1/4" 7/8" 5/8"
Smooth copper tube
1/2"   3/8"
CCA or copper wire
1/2"S 1/4"S
Foamed PE Dielectric
CCA or copper wire
Helically corrugated copper tube
PE Jacket
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Nominal Characteristics of Commonly Used Transmission Lines

RG or Type	Part Number	Nom. Z_0 Ω	VF %	Cap. pF/ft	Cent. Cond. AWG	Diel. Type	Shield Type	Jacket Matl	Jacket OD inches	Max V (RMS)	Matched Loss (dB/100) 1 MHz	10	100	1000
RG-6	Belden 1694A	75	82	16.2	#18 Solid BC	FPE	FC	P1	0.275	600	0.2	.7	1.8	5.9
RG-6	Belden 8215	75	66	20.5	#21 Solid CCS	PE	D	PE	0.332	2700	0.4	0.8	2.7	9.8
RG-8	Belden 7810A	50	86	23.0	#10 Solid BC	FPE	FC	PE	0.405	600	0.1	0.4	1.2	4.0
RG-8	TMS LMR400	50	85	23.9	#10 Solid CCA	FPE	FC	PE	0.405	600	0.1	0.4	1.3	4.1
RG-8	Belden 9913	50	84	24.6	#10 Solid BC	ASPE	FC	P1	0.405	600	0.1	0.4	1.3	4.5
RG-8	CXP1318FX	50	84	24.0	#10 Flex BC	FPE	FC	P2N	0.405	600	0.1	0.4	1.3	4.5
RG-8	Belden 9913F7	50	83	24.6	#11 Flex BC	FPE	FC	P1	0.405	600	0.2	0.6	1.5	4.8
RG-8	Belden 9914	50	82	24.8	#10 Solid BC	FPE	FC	P1	0.405	600	0.2	0.5	1.5	4.8
RG-8	TMS LMR400UF	50	85	23.9	#10 Flex BC	FPE	FC	PE	0.405	600	0.1	0.4	1.4	4.9
RG-8	DRF-BF	50	84	24.5	#9.5 Flex BC	FPE	FC	PE	0.405	600	0.1	0.5	1.6	5.2
RG-8	WM CQ106	50	84	24.5	#9.5 Flex BC	FPE	FC	P2N	0.405	600	0.2	0.6	1.8	5.3
RG-8	CXP008	50	78	26.0	#13 Flex BC	FPE	S	P1	0.405	600	0.1	0.5	1.8	7.1
RG-8	Belden 8237	52	66	29.5	#13 Flex BC	PE	S	P1	0.405	3700	0.2	0.6	1.9	7.4
RG-8X	Belden 7808A	50	86	23.5	#15 Solid BC	FPE	FC	PE	0.240	600	0.2	0.7	2.3	7.4
RG-8X	TMS LMR240	50	84	24.2	#15 Solid BC	FPE	FC	PE	0.242	300	0.2	0.8	2.5	8.0
RG-8X	WM CQ118	50	82	25.0	#16 Flex BC	FPE	FC	P2N	0.242	300	0.3	0.9	2.8	8.4
RG-8X	TMS LMR240UF	50	84	24.2	#15 Flex BC	FPE	FC	PE	0.242	300	0.2	0.8	2.8	9.6
RG-8X	Belden 9258	50	82	24.8	#16 Flex BC	FPE	S	P1	0.242	600	0.3	0.9	3.1	11.2
RG-8X	CXP08XB	50	80	25.3	#16 Flex BC	FPE	S	P1	0.242	300	0.3	0.9	3.1	14.0
RG-9	Belden 8242	51	66	30.0	#13 Flex SPC	PE	SCBC	P2N	0.420	5000	0.2	0.6	2.1	8.2
RG-11	Belden 8213	75	84	16.1	#14 Solid BC	FPE	S	PE	0.405	600	0.2	0.4	1.3	5.2
RG-11	Belden 8238	75	66	20.5	#18 Flex TC	PE	S	P1	0.405	600	0.2	0.7	2.0	7.1
RG-58	Belden 7807A	50	85	23.7	#18 Solid BC	FPE	FC	PE	0.195	300	0.3	1.0	3.0	9.7
RG-58	TMS LMR200	50	83	24.5	#17 Solid BC	FPE	FC	PE	0.195	300	0.3	1.0	3.2	10.5
RG-58	WM CQ124	52	66	28.5	#20 Solid BC	PE	S	PE	0.195	1400	0.4	1.3	4.3	14.3
RG-58	Belden 8240	52	66	28.5	#20 Solid BC	PE	S	P1	0.193	1900	0.3	1.1	3.8	14.5
RG-58A	Belden 8219	53	73	26.5	#20 Flex TC	FPE	S	P1	0.195	300	0.4	1.3	4.5	18.1
RG-58C	Belden 8262	50	66	30.8	#20 Flex TC	PE	S	P2N	0.195	1400	0.4	1.4	4.9	21.5
RG-58A	Belden 8259	50	66	30.8	#20 Flex TC	PE	S	P1	0.192	1900	0.4	1.5	5.4	22.8
RG-59	Belden 1426A	75	83	16.3	#20 Solid BC	FPE	S	P1	0.242	300	0.3	0.9	2.6	8.5
RG-59	CXP 0815	75	82	16.2	#20 Solid BC	FPE	S	P1	0.232	300	0.5	0.9	2.2	9.1
RG-59	Belden 8212	75	78	17.3	#20 Solid CCS	FPE	S	P1	0.242	300	0.6	1.0	3.0	10.9
RG-59	Belden 8241	75	66	20.4	#23 Solid CCS	PE	S	P1	0.242	1700	0.6	1.1	3.4	12.0
RG-62A	Belden 9269	93	84	13.5	#22 Solid CCS	ASPE	S	P1	0.240	750	0.3	0.9	2.7	8.7
RG-62B	Belden 8255	93	84	13.5	#24 Flex CCS	ASPE	S	P2N	0.242	750	0.3	0.9	2.9	11.0
RG-63B	Belden 9857	125	84	9.7	#22 Solid CCS	ASPE	S	P2N	0.405	750	0.2	0.5	1.5	5.8
RG-142	CXP 183242	50	69.5	29.4	#19 Solid SCCS	TFE	D	FEP	0.195	1900	0.3	1.1	3.8	12.8
RG-142B	Belden 83242	50	69.5	29.0	#19 Solid SCCS	TFE	D	TFE	0.195	1400	0.3	1.1	3.9	13.5
RG-174	Belden 7805R	50	73.5	26.2	#25 Solid BC	FPE	FC	P1	0.110	300	0.6	2.0	6.5	21.3
RG-174	Belden 8216	50	66	30.8	#26 Flex CCS	PE	S	P1	0.110	1100	1.9	3.3	8.4	34.0
RG-213	Belden 8267	50	66	30.8	#13 Flex BC	PE	S	P2N	0.405	3700	0.2	0.6	1.9	8.0
RG-213	CXP213	50	66	30.8	#13 Flex BC	PE	S	P2N	0.405	600	0.2	0.6	2.0	8.2
RG-214	Belden 8268	50	66	30.8	#13 Flex SPC	PE	D	P2N	0.425	3700	0.2	0.6	1.9	8.0
RG-216	Belden 9850	75	66	20.5	#18 Flex TC	PE	D	P2N	0.425	3700	0.2	0.7	2.0	7.1
RG-217	WM CQ217F	50	66	30.8	#10 Flex BC	PE	D	PE	0.545	7000	0.1	0.4	1.4	5.2
RG-217	M17/78-RG217	50	66	30.8	#10 Solid BC	PE	D	P2N	0.545	7000	0.1	0.4	1.4	5.2
RG-218	M17/79-RG218	50	66	29.5	#4.5 Solid BC	PE	S	P2N	0.870	11000	0.1	0.2	0.8	3.4
RG-223	Belden 9273	50	66	30.8	#19 Solid SPC	PE	D	P2N	0.212	1400	0.4	1.2	4.1	14.5
RG-303	Belden 84303	50	69.5	29.0	#18 Solid SCCS	TFE	S	TFE	0.170	1400	0.3	1.1	3.9	13.5
RG-316	CXP TJ1316	50	69.5	29.4	#26 Flex BC	TFE	S	FEP	0.098	1200	1.2	2.7	8.0	26.1
RG-316	Belden 84316	50	69.5	29.0	#26 Flex SCCS	TFE	S	FEP	0.096	900	1.2	2.7	8.3	29.0
RG-393	M17/127-RG393	50	69.5	29.4	#12 Flex SPC	TFE	D	FEP	0.390	5000	0.2	0.5	1.7	6.1
RG-400	M17/128-RG400	50	69.5	29.4	#20 Flex SPC	TFE	D	FEP	0.195	1400	0.4	1.1	3.9	13.2
LMR500	TMS LMR500UF	50	85	23.9	#7 Flex BC	FPE	FC	PE	0.500	2500	0.1	0.4	1.2	4.0
LMR500	TMS LMR500	50	85	23.9	#7 Solid CCA	FPE	FC	PE	0.500	2500	0.1	0.3	0.9	3.3
LMR600	TMS LMR600	50	86	23.4	#5.5 Solid CCA	FPE	FC	PE	0.590	4000	0.1	0.2	0.8	2.7
LMR600	TMS LMR600UF	50	86	23.4	#5.5 Flex BC	FPE	FC	PE	0.590	4000	0.1	0.2	0.8	2.7
LMR1200	TMS LMR1200	50	88	23.1	#0 Copper Tube	FPE	FC	PE	1.200	4500	0.04	0.1	0.4	1.3
Hardline														
1/2"	CATV Hardline	50	81	25.0	#5.5 BC	FPE	SM	none	0.500	2500	0.05	0.2	0.8	3.2
1/2"	CATV Hardline	75	81	16.7	#11.5 BC	FPE	SM	none	0.500	2500	0.1	0.2	0.8	3.2
7/8"	CATV Hardline	50	81	25.0	#1 BC	FPE	SM	none	0.875	4000	0.03	0.1	0.6	2.9
7/8"	CATV Hardline	75	81	16.7	#5.5 BC	FPE	SM	none	0.875	4000	0.03	0.1	0.6	2.9
LDF4-50A	Heliax –1/2"	50	88	25.9	#5 Solid BC	FPE	CC	PE	0.630	1400	0.05	0.2	0.6	2.4
LDF5-50A	Heliax –7/8"	50	88	25.9	0.355" BC	FPE	CC	PE	1.090	2100	0.03	0.10	0.4	1.3
LDF6-50A	Heliax – 1¼"	50	88	25.9	0.516" BC	FPE	CC	PE	1.550	3200	0.02	0.08	0.3	1.1
Parallel Lines														
TV Twinlead (Belden 9085)		300	80	4.5	#22 Flex CCS	PE	none	P1	0.400	**	0.1	0.3	1.4	5.9
Twinlead (Belden 8225)		300	80	4.4	#20 Flex BC	PE	none	P1	0.400	8000	0.1	0.2	1.1	4.8
Generic Window Line		405	91	2.5	#18 Solid CCS	PE	none	P1	1.000	10000	0.02	0.08	0.3	1.1
WM CQ 554		420	91	2.7	#14 Flex CCS	PE	none	P1	1.000	10000	0.02	0.08	0.3	1.1
WM CQ 552		440	91	2.5	#16 Flex CCS	PE	none	P1	1.000	10000	0.02	0.08	0.3	1.1
WM CQ 553		450	91	2.5	#18 Flex CCS	PE	none	P1	1.000	10000	0.02	0.08	0.3	1.1
WM CQ 551		450	91	2.5	#18 Solid CCS	PE	none	P1	1.000	10000	0.02	0.08	0.3	1.1
Open-Wire Line		600	92	1.1	#12 BC	none	none	none	**	12000	0.02	0.06	0.2	0.7

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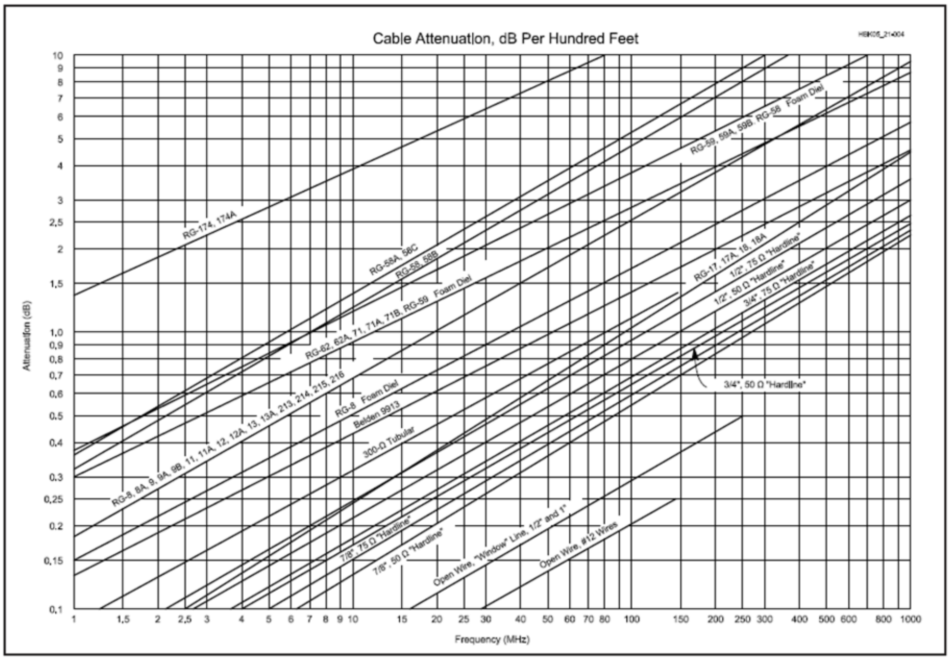
Note: Coax losses shown above are in dB for 100 feet lengths at specified frequencies. Loss is a length multiplier, so a 200 ft length would have twice the loss shown above and a 50 ft length would have half the loss.



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Matched-Line Loss for 250 ft of Three Common Coaxial Cables

Comparisons of line losses versus frequency for 250-ft lengths of three different coax cable types: small-diameter RG58A, medium-diameter RG8A, and ³/₄-inch OD 50-Ω Hardline. At VHF, the losses for the small-diameter cable are very large, while they are moderate at 3.5 MHz.

Xmsn Line	*3.5 MHz Matched-Line Loss, dB*	*3.5 MHz Loss, 6:1 SWR, dB*	*28 MHz Matched-Line Loss, dB*	*28 MHz Loss, 6:1 SWR, dB*	*146 MHz Matched-Line Loss, dB*	*146 MHz Loss 6:1 SWR, dB*
RG-58A	1.9	4.0	6.3	9.3	16.5	21.6
RG-8A	0.9	2.2	3.0	5.4	7.8	10.8
³/₄" 50-Ω Hardline	0.2	0.5	0.7	1.8	2.1	4.2

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Coax Connectors

- Enable coax cable to be connected to devices and antennas.
- Type depends on cable, use and frequency.
- Buy from reputable manufacturers.
- Look for Teflon insulation, silver coating on parts that must be soldered, and gold plating on pieces that mate.

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PL259 and SO239

- Called "UHF" connectors.
- Good on HF, somewhat useful on 2M, not any higher in frequency however.
- Connects to RG8/U and RG213/U cable.
- PL259 plugs into SO239.
- NOT weatherproof!

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Using PL259 with RG58/U

- An UG175 adaptor is needed to attach a PL259 to RG58/U.
- Get the right one – similar looking adaptors exist for RG59, but have a different diameter.

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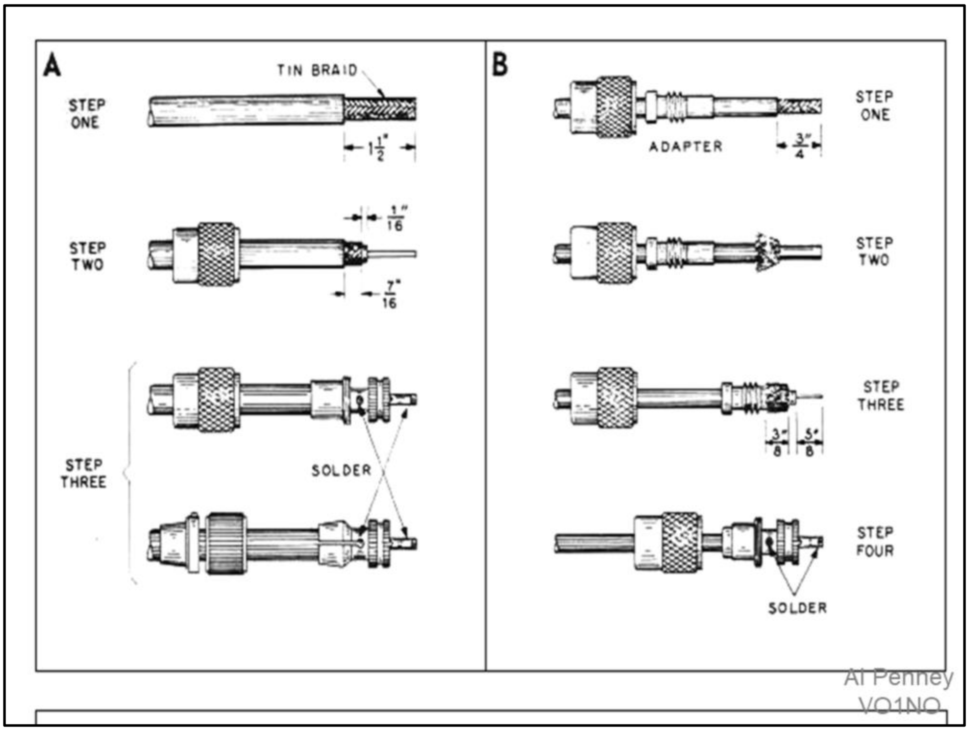

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BNC Connectors

- “Bayonet Neill Concelman”
- Quick connect/disconnect.
- Good to 2 GHz or more, but limited in power handling capability.
- Not weatherproof.

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N Connectors

- Named after Paul Neill, of Bell Labs.
- Fits RG8/U, RG213/U and other cables.
- Good to 11 GHz or higher, at full legal limit in lower bands.
- Waterproof.

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F Connectors

- Used on RG59 and RG6 for TV applications.
- Some use them for receive antennas.
- Easy to attach, but not very strong or waterproof.

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Adaptors

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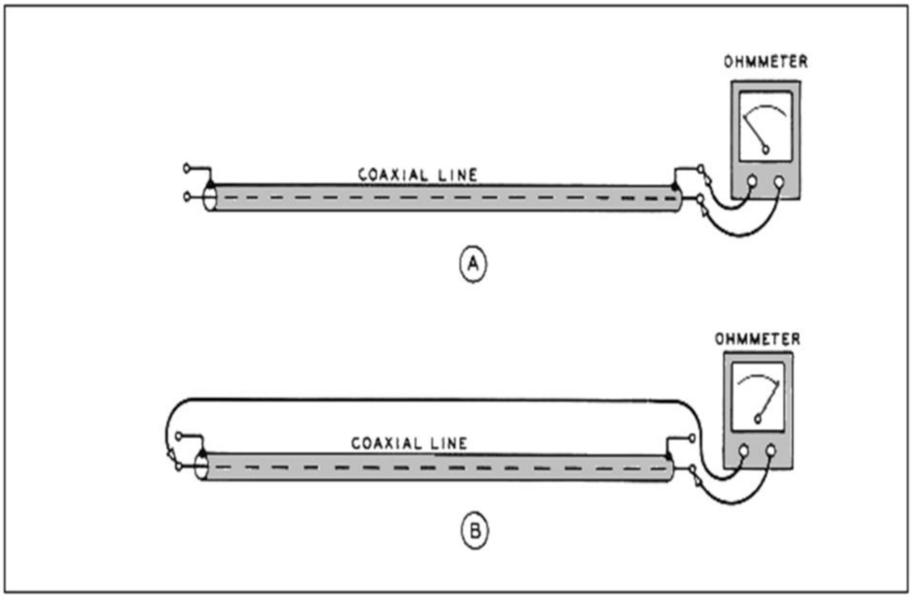

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Care of Coax Cable

- Do not kink or bend cable too sharply – the center conductor can migrate.
- Minimum bend radius is 5-10 times cable diameter.
- Protect from water ingress!
- Do not drag or step on cable.

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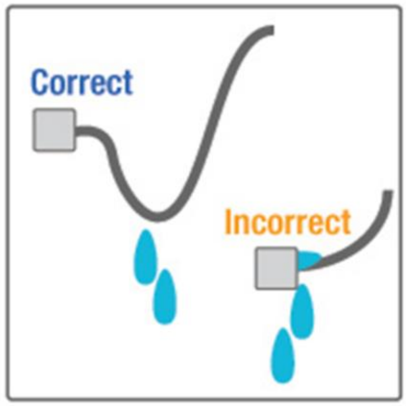
Drip Loop



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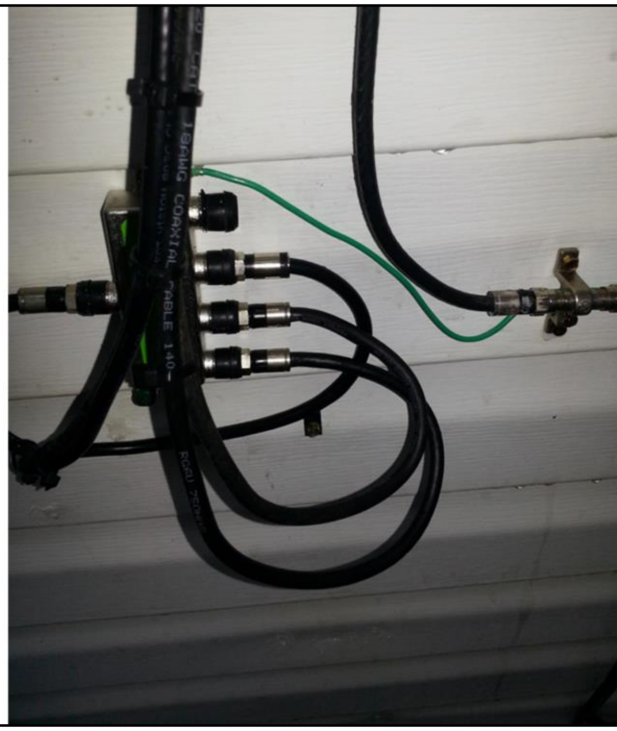
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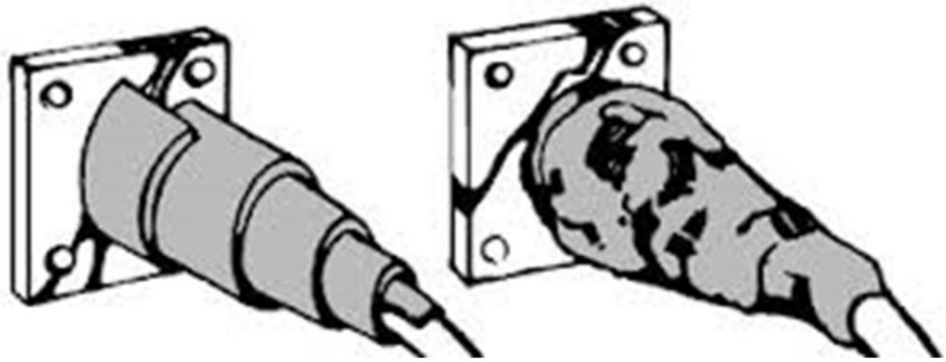

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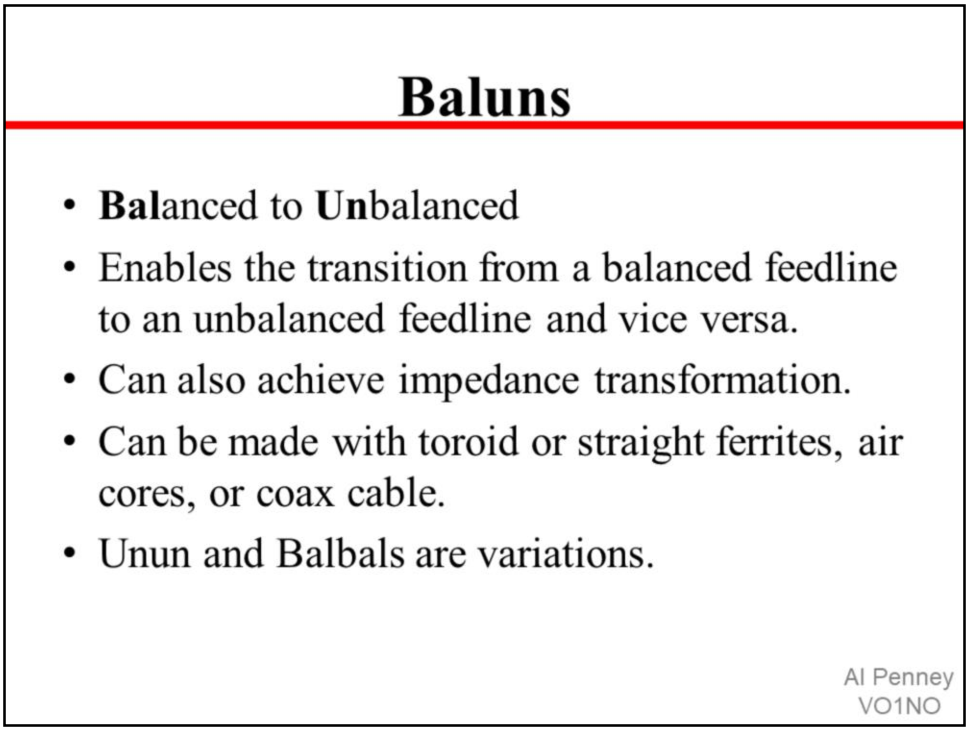


A **balun** /ˈbælʌn/ (portmanteau of "balanced to unbalanced") is an electrical device that converts between a balanced signal and an unbalanced signal. A balun can take many forms and may include devices that also transform impedances but need not do so. Transformer baluns can also be used to connect lines of differing impedance. Sometimes, in the case of transformer baluns, they use magnetic coupling but need not do so. Common-mode chokes are also used as baluns and work by eliminating, rather than ignoring, common mode signals.

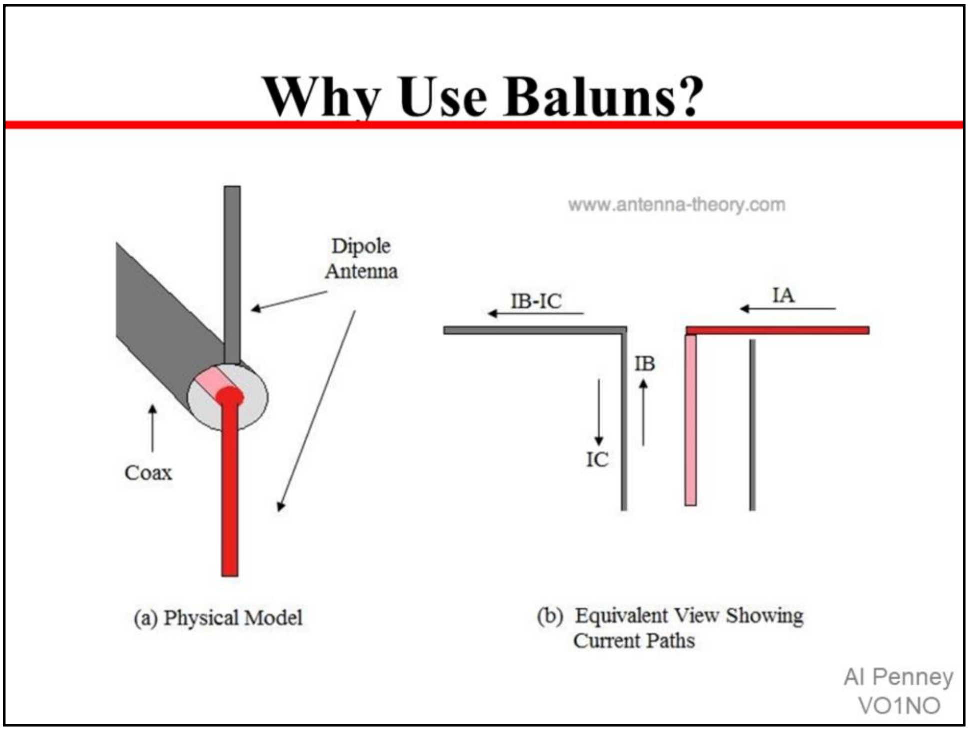
There are two variations of this device - they are:

•the **unun**, which transfers signal from one unbalanced line to another.

•the **balbal**, which transfers signal from one balanced line to another.

A *balun* is a transformer that matches a *bal*anced load (such as a horizontal dipole or Yagi antenna) to an *un*balanced resistive source (such as a transmitter output or a coaxial

feedline).

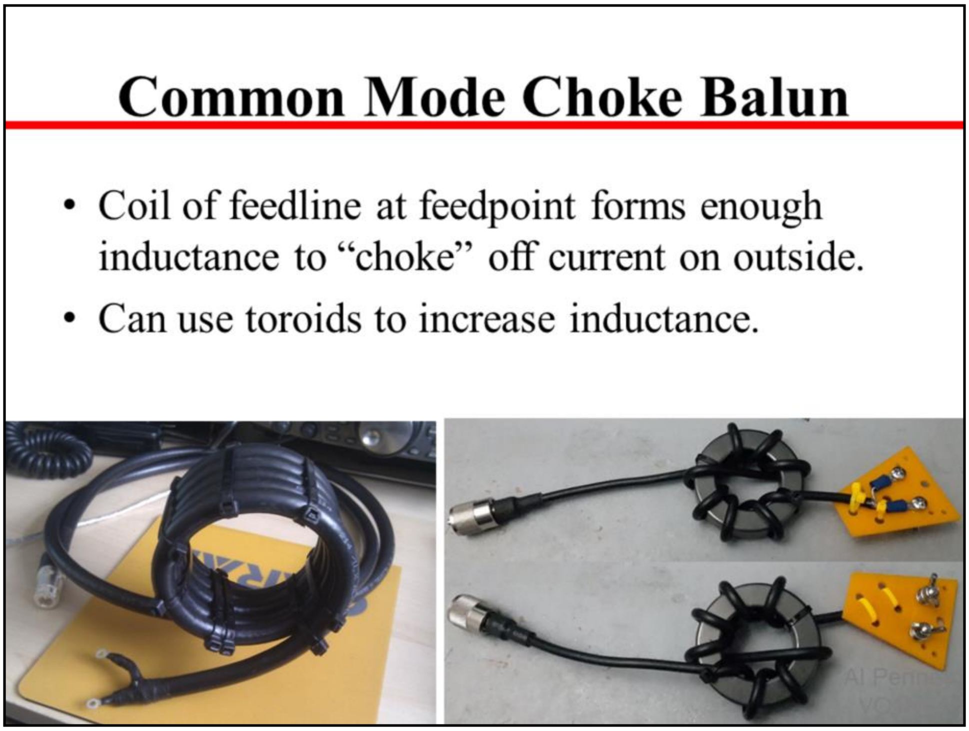


A **Balun** is used to "balance" unbalanced systems - i.e. those where power flows from an unbalanced line to a balanced line (hence, balun derives from *bal*ance to *un*balanced). As an example, consider a coaxial cable connected to a half-wave dipole antenna shown above.

In the Figure, a coaxial cable is connected to a dipole antenna. For a dipole antenna to operate properly, the currents on both arms of the dipole should be equal in magnitude. When a coaxial cable is connected directly to a dipole antenna however, the currents will not neccessarily be equal. To see this, note that the current along a transmission line should be of equal magnitude on the inner and outer conductors, as is typically the case. Observe what happens when the coax is connected to the dipole. The current on the center conductor (the red/pink center core of the coax, labeled *IA*) has no where else to go, so must flow along the dipole arm that is connected to it. However, the current that travels along the inner side of the outer conductor (*IB*) has two options: it can travel down the dipole antenna, or down the reverse (outer) side of the outer conductor of the coaxial cable (labeled *IC* in Figure 1).

Ideally, the current *IC* should be zero. In that case, the current along the dipole arm connected to the outer conductor of the coax will be equal to the current on the other dipole arm - a desirable antenna characteristic. Because the dipole wants equal or balanced currents along its arms, it is the balanced section. The coaxial cable does not necesarily give this however - some of the current may travel down the outside of the outer coax, leading to unbalanced operation - this is the unbalanced section.

The solution to this problem, however you come up with it, is a balun. A balun forces an unbalanced transmission line to properly feed a balanced component. In Figure 1, this would be done by forcing *IC* to be zero somehow - this is often called choking the current or a current choke.



Placing a “common-mode choke” whose reactance is + *j*1000 οημσ at the antenna’s feed point removes virtually all trace current on the outside of the braid. This is always true for the simple case where the feedline was dressed symmetrically, directly down under the feed point. Certain slanted-feedline lengths required additional common-mode chokes, placed at λ/4 intervals down the transmission line from the feed point.

The simplest method to create a common-mode choke balun with coaxial cable is to wind up some of it into a coil at the feed point of the antenna. The normal transmission-line currents inside the coax are

unaffected by the coiled configuration, but commonmode currents trying to flow on the outside of the coax braid are “choked off” by the reactance of the coil. This coax-coil choke could also be referred to

as an “air-wound” choke, since no ferrite-core material is used to help boost the common-mode reactance at low frequencies.

A coax choke can be made like a flat coil— that is, like a coil of rope whose adjacent turns are carefully placed side-by-side to reduce inter-turn distributed capacity, rather than in a

“scramble-wound” fashion. Sometimes a coil form made of PVC is used to keep things orderly.

This type of choke shows a broad resonance due to its inductance and distributed capacity that can easily cover three amateur bands. See **Fig 30**. Some geometries are reasonably effective

over the entire HF range. If particular problems are encountered on a single band, a coil that is resonant at that band may be added. The coils shown in **Table 3** were designed to have a high

impedance at the indicated frequencies, as measured with an impedance meter. Many other geometries can also be effective. This construction technique is not effective with twin-lead because of excessive coupling between adjacent turns.

This choke-type of balan is sometimes referred to as a “current balun” since it has the hybrid properties of a tightly coupled transmission-line transformer (with a 1:1 transformation ratio) and a coil. The transmission- line transformer forces the current at the output terminals to be equal, and the coil portion chokes off common-mode currents.

Ferrite Bead Choke

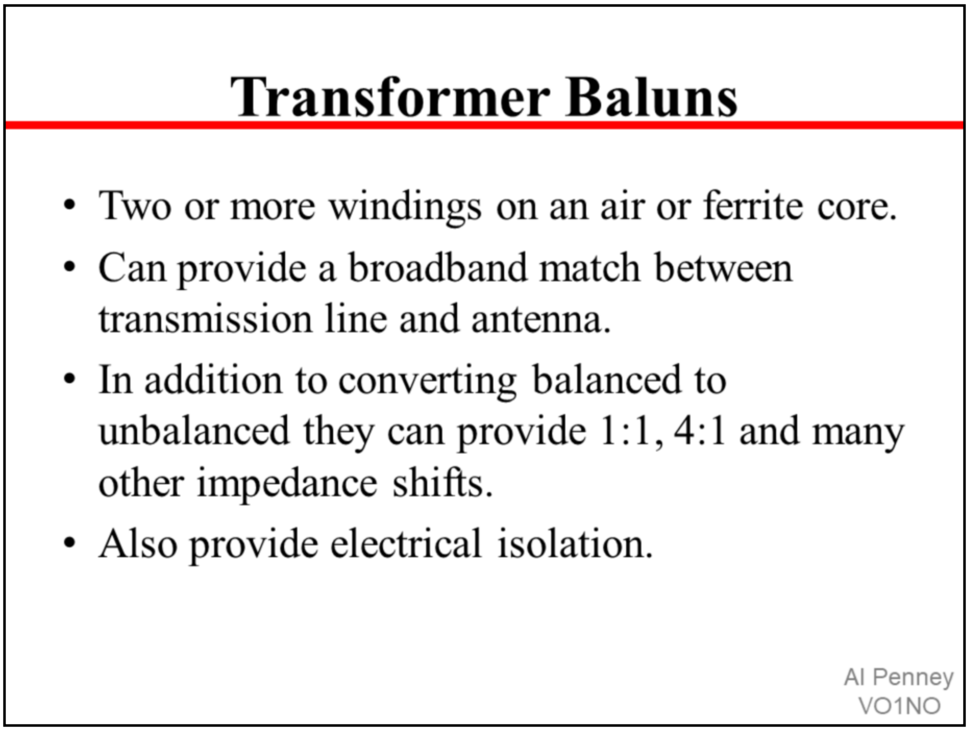
Choke Baluns

Table 3
Effective Choke (Current) Baluns

	Single Band (very effective)			***Multiple Band***
Freq (MHz)	*RG-213, RG-8*	*RG-58*	*Freq (MHz)*	*RG-8, 58, 59, 8X, 213*
3.5	22 ft, 8 turns	20 ft, 6-8 turns	3.5-30	10 ft, 7 turns
7	22 ft, 10 turns	15 ft, 6 turns	3.5-10	18 ft, 9-10 turns
10	12 ft, 10 turns	10 ft, 7 turns	14-30	8 ft, 6-7 turns
14	10 ft, 4 turns	8 ft, 8 turns		
21	8 ft, 6-8 turns	6 ft, 8 turns		
28	6 ft, 6-8 turns	4 ft, 6-8 turns		

Wind the indicated length of coaxial feed line into a coil (like a coil of rope) and secure with electrical tape. The balun is most effective when the coil is near the antenna. Lengths are not highly critical.

Al Penney
VO1NO



Toroid Impedance Matching Transformers

The toroidal transformer is capable of providing a broadband match between antenna and transmission line, or between transmission line and transmitter or receiver. Many other matching methods are frequency sensitive and must be readjusted whenever the operating frequency is changed by even a small percentage. Although

this problem is of no great concern to fixed frequency radio stations, it is of critical importance to stations that operate on a variety of frequencies or widely separated bands of frequencies.

In classical transformers, there are two electrically separate windings of wire coils around the transformer's core. The advantage of transformer-type over other types of balun is that the electrically separate windings for input and output allow these baluns to connect circuits whose ground-level voltages are subject to ground loops or are otherwise electrically incompatible; for that reason they are often called *isolation transformers*.

This type is sometimes called a *voltage balun*. The *primary* winding receives the input signal, and the *secondary* winding puts out the converted signal. The core that they are wound on may either be empty (air core) or, equivalently, a magnetically neutral material like a porcelain support, or it may be a material which is good magnetic conductor like ferrite in modern high-frequency (HF) baluns, or soft iron as in the early days of telegraphy.

The electrical signal in the primary coil is converted into a magnetic field in the transformer's core. When the electrical current through the primary reverses, it causes the established magnetic field to collapse. The collapsing magnetic field then induces an electric field in the secondary winding.

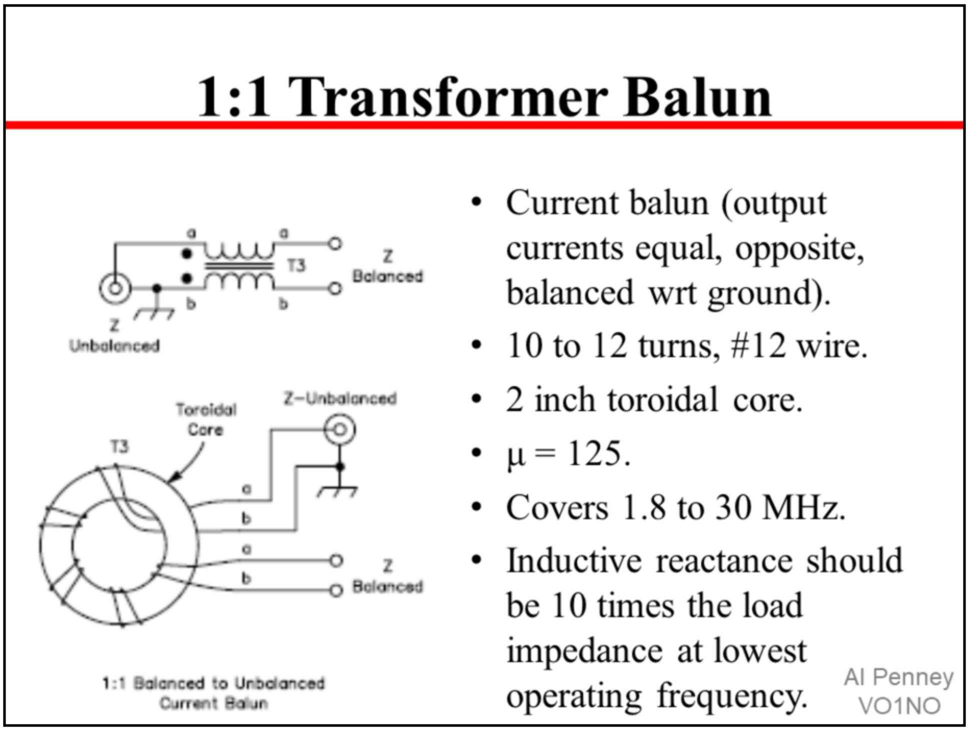
The ratio of loops in each winding and the efficiency of the coils' magnetic coupling determines the ratio of electrical potential (voltage) to electrical current and the total power of the output. For idealized transformers, although the ratio of voltage to current will change in exact proportion to the square of the winding ratio, the power (measured in watts) remains identical. In real transformers, some energy is lost inside to heating of the metallic core of the transformer, and lost outside to the surrounding environment because of imperfect magnetic coupling between the two coils.

Turns Ratio

- $Z_1 / Z_2 = (N_1 / N_2)^2$
- Where
  - Z_1 is input impedance
  - Z_2 is output impedance
  - N_1 is number of turns on transformer primary
  - N_2 is number of turns on transformer secondary
- Rearranging formula gives

N_1 / N_2 = Square root (Z_1 / Z_2)

Actual number of turns depends on core material.



A Current balun is one whose output currents are equal and opposite (balanced with respect to ground). This 1:1 balun excepted, they are usually more difficult and expensive to build than voltage baluns

Typically, 10 to 12 turns of #12 wires wound on 2.0-inch toroidal cores with μ = 125 will cover the whole range from 1.8 to 30 MHz.

Ferrite Core Inductors

The word *ferrite* refers to any of several examples of a class of materials that behave similarly to powdered iron compounds and are used in radio equipment as the cores for

inductors and transformers. Although the materials originally employed were of powdered iron (and indeed the name *ferr*ite still implies iron), many modern materials are

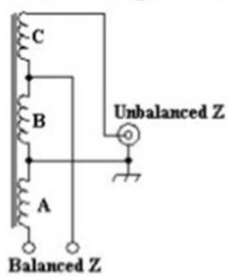
composed of other compounds. According to literature from Amidon Associates, ferrites with a *relative permeability,* or *μr,* of 800 to 5000 are generally of the manganese–zinc

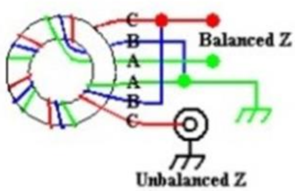
type of material, and cores with relative permeabilities of 20 to 800 are of nickel–zinc. The latter are useful in the 0.5–A� to 100–A�MHz range.


Al Penney
VO1NO

1:1 Transformer Balun

1:1 Voltage Balun



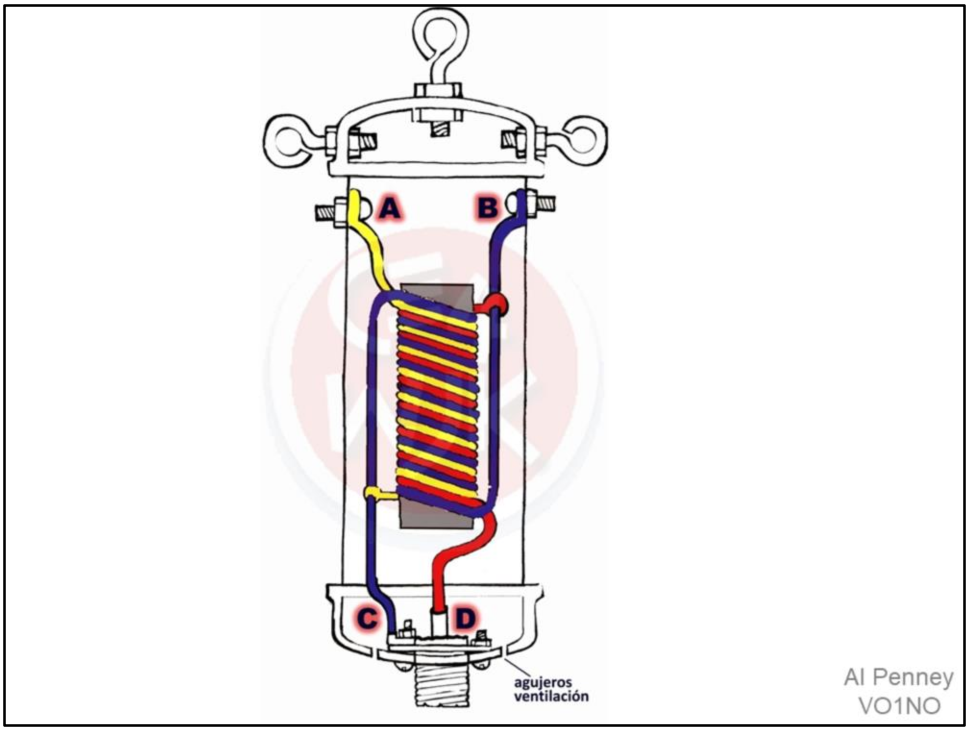


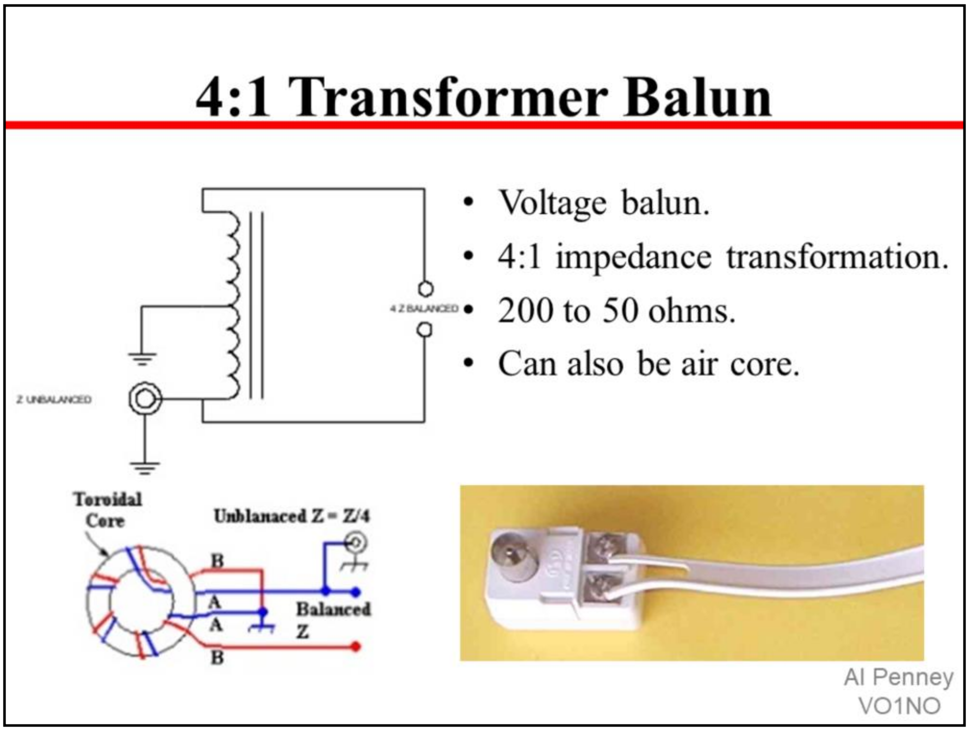
- Voltage Balun (output voltages equal, opposite, balanced wrt ground).
- Three windings connected in series.
- Can be air, toroidal or ferrite bar core.

Al Penney
VO1NO

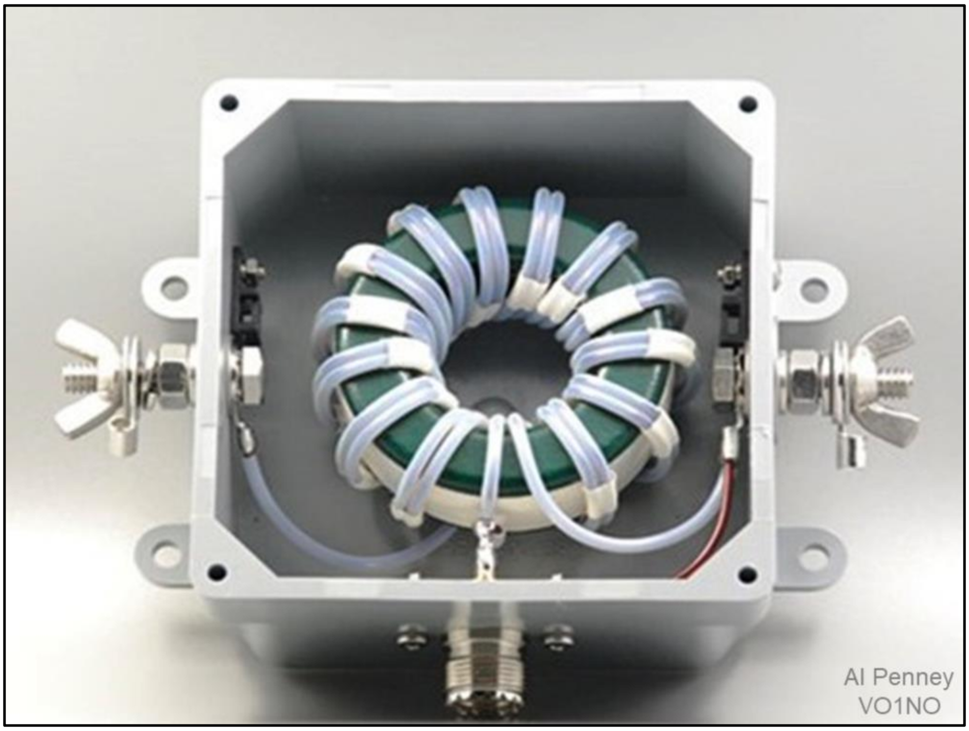
A Voltage balun is one whose output voltages are equal and opposite (balanced with respect to ground).


Power Antenna
1:1 Current / Choke
5KW: 160 to 10
number: BL
Al Penney
VO1NO


A
B
C
D
agujeros
ventilación
Al Penney
VO1NO



A Voltage balun is one whose output voltages are equal and opposite (balanced with respect to ground).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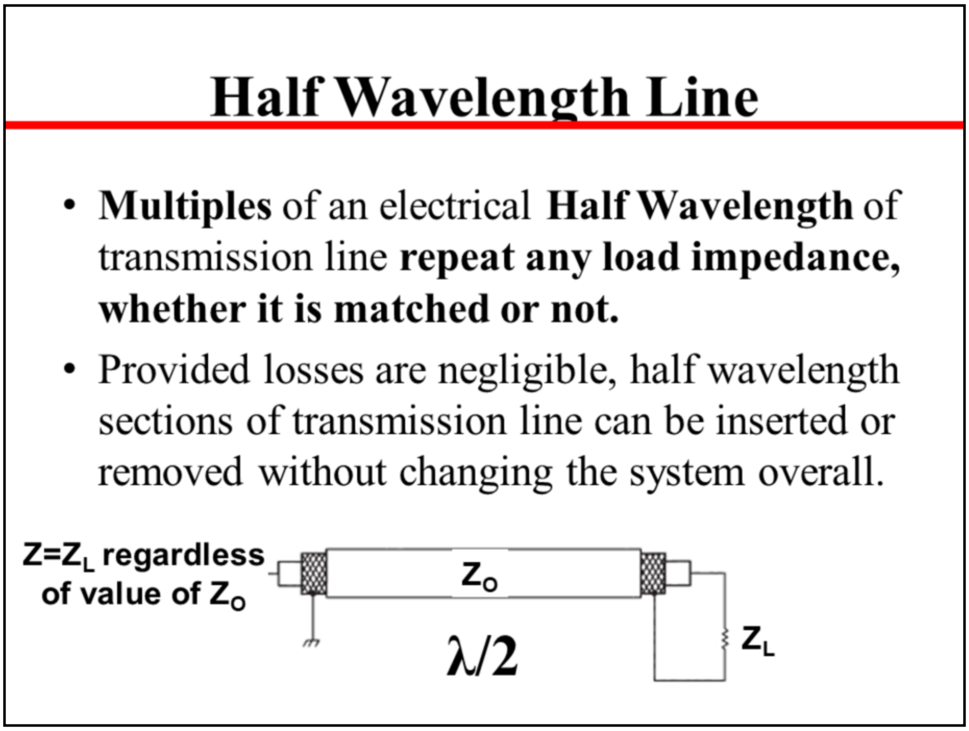
λ/2 and λ/4 Transmission Lines

- **Half wavelength:**
  - **Open** at far end presents **very high impedance** to signal source.
  - **Shorted** at far ends presents **very low impedance** to signal source.
- **Quarter Wavelength:**
  - **Open** at far ends presents **very low impedance** to signal source.
  - **Shorted** at far ends presents **very high impedance** to signal source.



Half-Wavelength Section of Line

A half–wavelength section of transmission line is “magical” because it repeats *any* load impedance, whether matched or not. Thus, for a single frequency or band of frequencies

a transmission line that is any integer multiple of λ/2 wavelength long can be used to remotely measure and monitor the feedpoint impedance of an antenna. The two caveats are:

•))> An accurate knowledge of the line’s velocity factor is needed before cutting the line to length.

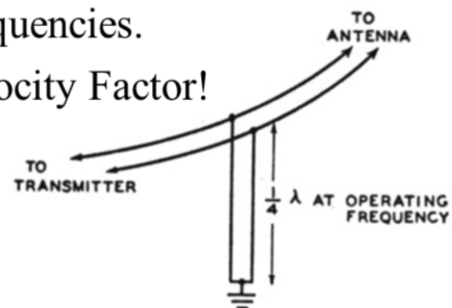
•))> A given length of line is λ/2 wavelength only over a very narrow range of frequencies.

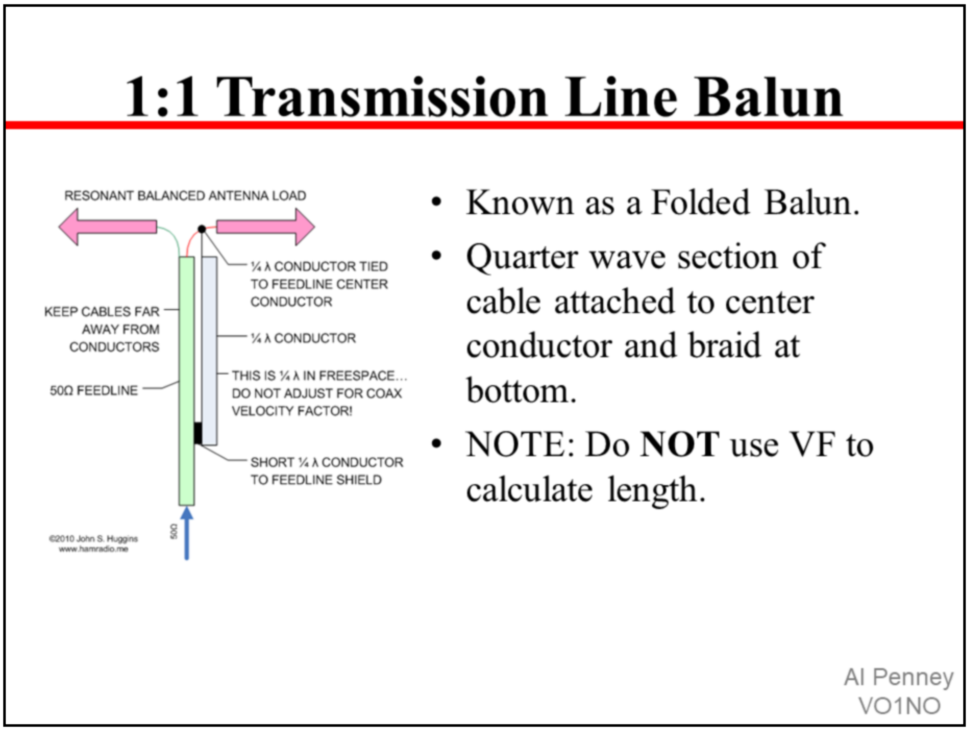
The Half-Wavelength Line

When the line length is an even multiple of 90° (that is, a multiple of λ/2 wavelength), the input resistance is equal to the load resistance, regardless of the line Z0. As a matter of fact, a line an exact multiple of λ/2 wavelength in length (disregarding line losses) simply repeats, at its input or sending end, whatever impedance exists at its output or receiving end. It does not matter whether the impedance at the receiving end is resistive, reactive, or a combination of both. Sections of line having such length can be added or removed without changing any of the operating conditions, at least when the losses in the line itself are negligible.

Quarter Wavelength Stub

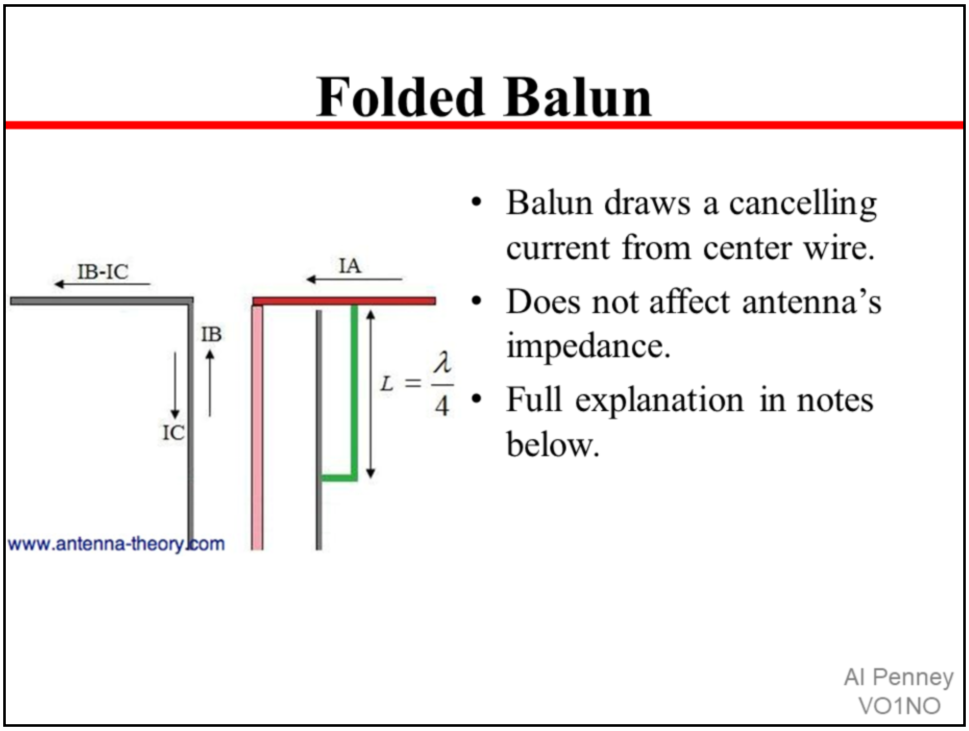
- Quarter wavelength stub with a shorted end looks like an infinite impedance to a signal.
- Can be used as a notch filter.
- Can also be used as an insulator, especially at microwave range of frequencies.
- Remember to apply Velocity Factor!





Folded Balun

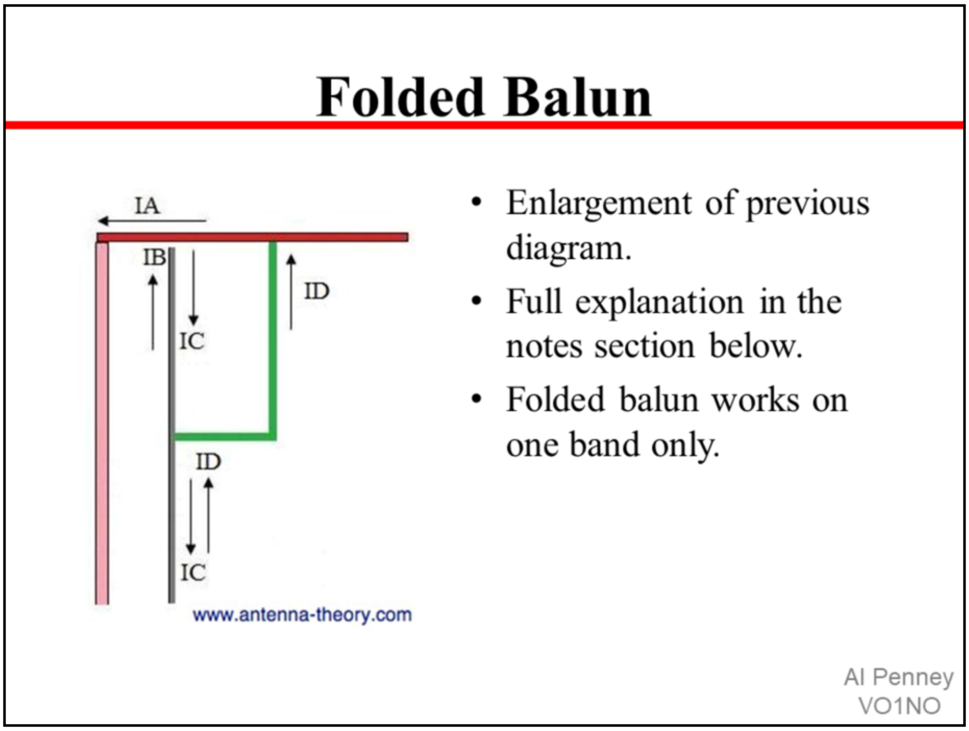
aka *Pawsey Stub* and *1/4 Wave Coaxial Balun*
Figure shows the idea behind the Pawsey stub which is known in Electrical Engineering circles as a variant of the Folded Balun. While the Gray conductor in the Figure only needs to be a wire of similar size to the coaxial cable feedline, it is often made from a scrap piece of the same cable. Each end of the outer shield of the stub is connected to the feed system. A common thought of many is since this is coaxial cable, we need velocity factor adjustments. Since the electric and magnetic fields (of the stub system) are in air, I think velocity factor does not apply.



The principle of operation of a **Folded Balun** is presented on this page (this balun is sometimes called the quarter-wavelength balun). This balun draws a cancelling current from the central arm of the coax that cancels any current that travels down the outer sleeve of the coaxial cable - thus eliminating the unbalanced condition.This balun is tricky to understand. Personally, I had to stare at it, then go to sleep, wake up, think about it some more, then it made sense. I will attempt to describe the operation here.

Note that the current on inner conductor of the coaxial cable contributes to the current that flows on the outer conductor. This could be cancelled by tying the inner conductor to the outside of the coaxial cable (because the current on the inner conductor is 180 degrees out of phase), which would cancel any net flow of current down the outer side of the coaxial cable. However, this would short-circuit any antenna connected to the coaxial cable, and thus we would have no radiation. The goal however, is to find a way to tie the current from the center cable of the coax to the outer shield such that the outer current is cancelled.

The folded balun is shown in Figure 1. The green line (which will be made of a cylindrical tube of the same dimensions and material as the main coax) is connected from the red dipole arm (that is connected to the center conductor) to the outside of the coaxial cable, a distance *L* from the end. If *L* is chosen to be a quarter-wavelength at the frequency of operation, then just as is the case for the bazooka balun, the impedance seen by this arm is infinite (due to the properties of transmission line theory and the quarter wavelength line). Hence, the first thing we need to note is that adding this balun does not affect the impedance of the dipole antenna - this means that if we can get the correct voltage to the antenna, it will radiate properly.



Now, lets try to figure out why this will achieve balanced operation. Enlarging Figure 1 and showing all of the currents on the coaxial cable and balun (and getting rid of the other half of the dipole antenna because it does not need to be illustrated here), we have Figure 2. I've introduced another current, ID, which travels along the balun.

We are going to have to understand all of the currents shown in Figure 2. The current IA is what travels down the center of the coaxial cable and onto the red arm of the dipole. The current IB is what flows on the inside of the outer shield of the coaxial cable and feeds the other dipole arm (not shown). The coaxial cable is a lossless transmission line - this means that the currents IA and IB are equal in magnitude but 180 degrees out of phase. That is, IA = - IB.

The current IC is what travels down the outside of the coaxial cable - this is the current that we would like to choke off. How large is IC? To determine this, we note that when the current IB travels to the end of the coaxial cable, it can either travel down the grey dipole arm (of Figure 1) or down the outside of the coaxial cable. The current IC will depend on what the impedance is looking down the outside of the coaxial cable relative to the impedance of the dipole arm. Lets call the impedance of the path on the outside of the coaxial cable as *Zg*.

The current ID must travel in the opposite direction to that of the current IC, because it is fed by the current IA, whereas IC is fed by IB. Here is the important observation. What is the magnitude of ID? The answer is - it must be equal to IC. Why? The answer lies in the following two observations:

First, note that the voltage on the inner conductor (pink line) is out of phase with the voltage on the outer conductor (grey line) - but they are of equal magnitude. Hence, if the attachment point of the balun to the red dipole arm (green to red connection in Figure 2), is made close to the center conductor, then the voltage at the balun connection point and the voltage at the connection between the outer conductor and the grey dipole arm are equal in magnitude but out of phase.

Second, at the point of attachment of the balun (the green-red connection in Figure 2), the impedance viewed down towards the outside of the coaxial cable must be equal to *Zg*. This is because the balun is made of the same shape and material as the main coaxial cable. Consequently, the impedance viewed down the balun and onto the main coax must be the same as that for the current IC.

Hence, if the voltages are equal in magnitude and out of phase, then the currents will cancel in the region below the balun. Consequently, balanced operation is restored.

This concept will probably take a few reads to make sense. Draw yourself some diagrams and work through it, and you will understand what is going on.

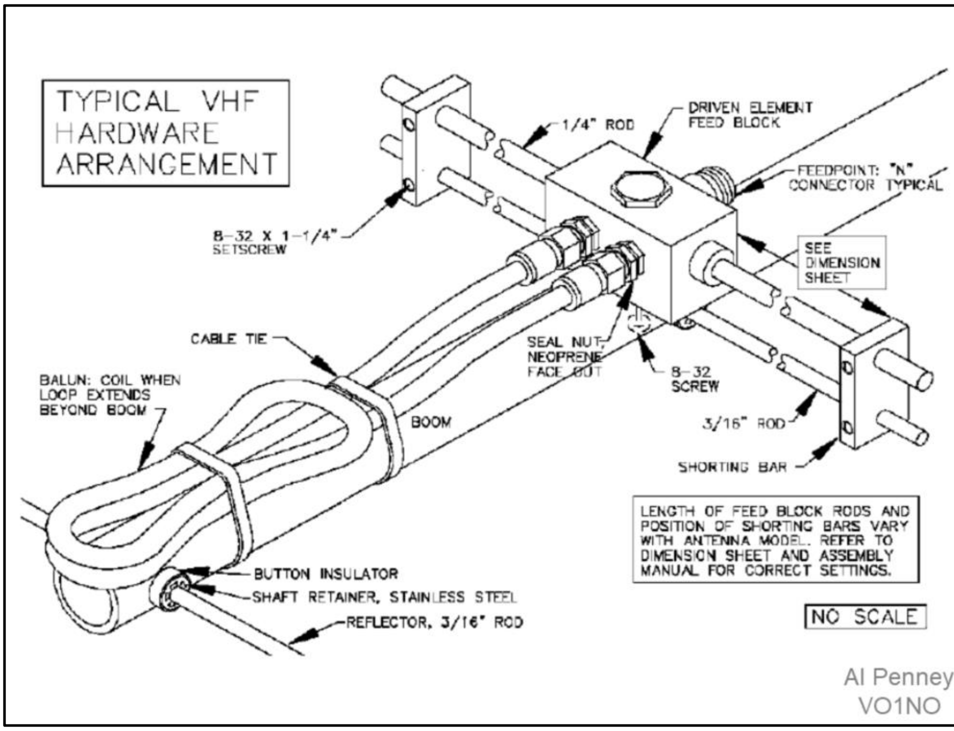
Note that because this balun requires a quarter-wavelength section to work properly, it is inherently narrowband. Away from the design frequency, the balun will no longer be a quarter-wave long and the properties described here will degrade.

Al Penney
VO1NO

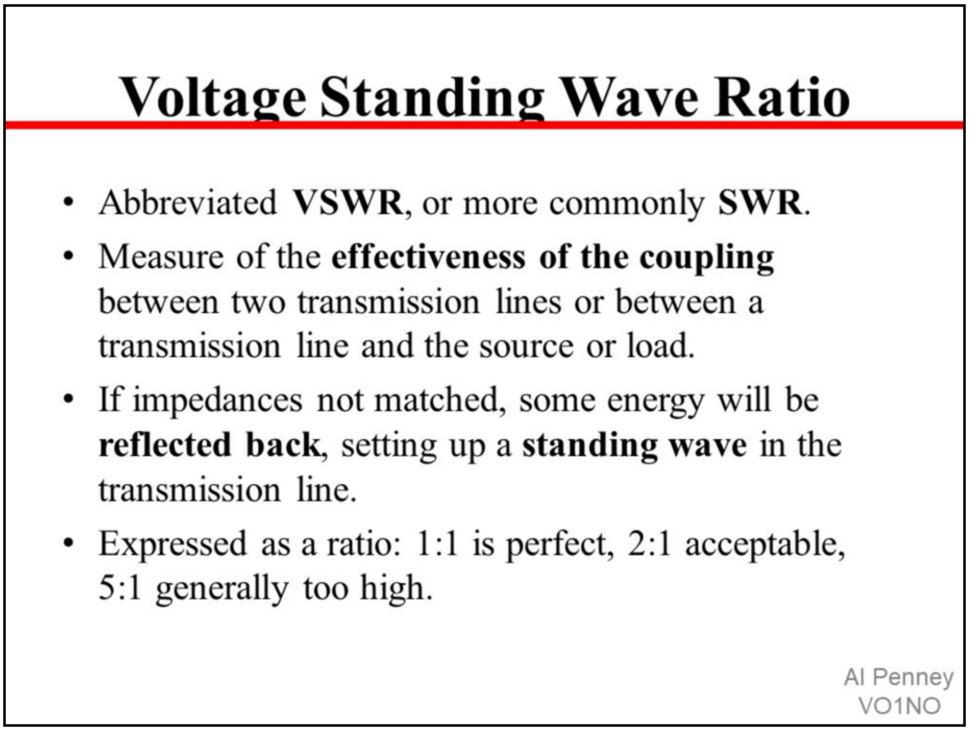
4:1 Transmission Line Balun

- Voltage Balun.
- Uses half wavelength section of coaxial cable (factor in the VF).
- Suitable for one band only.
- Often used on VHF/UHF Yagis.
- 75 ohm cable gives 300 ohms.

Al Penney
VO1NO



TYPICAL VHF HARDWARE ARRANGEMENT
DRIVEN ELEMENT FEED BLOCK
1/4" ROD
FEEDPOINT: "N" CONNECTOR TYPICAL
8-32 X 1-1/4" SETSCREW
SEE DIMENSION SHEET
CABLE TIE
SEAL NUT, NEOPRENE FACE OUT
8-32 SCREW
BALUN: COIL WHEN LOOP EXTENDS BEYOND BOOM
BOOM
3/16" ROD
SHORTING BAR
LENGTH OF FEED BLOCK RODS AND POSITION OF SHORTING BARS VARY WITH ANTENNA MODEL. REFER TO DIMENSION SHEET AND ASSEMBLY MANUAL FOR CORRECT SETTINGS.
BUTTON INSULATOR
SHAFT RETAINER, STAINLESS STEEL
REFLECTOR, 3/16" ROD
NO SCALE
Al Penney
VO1NO



In radio engineering and telecommunications, **standing wave ratio** (**SWR**) is a measure of impedance matching of loads to the characteristic impedance of a transmission line or waveguide. Impedance mismatches result in standing waves along the transmission line, and SWR is defined as the ratio of the partial standing wave's amplitude at an antinode (maximum) to the amplitude at a node (minimum) along the line.

The SWR is usually thought of in terms of the maximum and minimum AC voltages along the transmission line, thus called the **voltage standing wave ratio** or **VSWR** (sometimes pronounced "vizwar". For example, the VSWR value 1.2:1 denotes an AC voltage due to standing waves along the transmission line reaching a peak value 1.2 times that of the minimum AC voltage along that line. The SWR can as well be defined as the ratio of the maximum amplitude to minimum amplitude of the transmission line's currents, electric field strength, or the magnetic field strength. Neglecting transmission line loss, these ratios are identical.

Calculating SWR

- SWR meter compares forward voltage V_f and reverse voltage V_r
- SWR = $V_f + V_r / V_f - V_r$

- Can also be calculated by comparing impedances of the load Z_O and the line Z_L
- **SWR = Z_O / Z_L**

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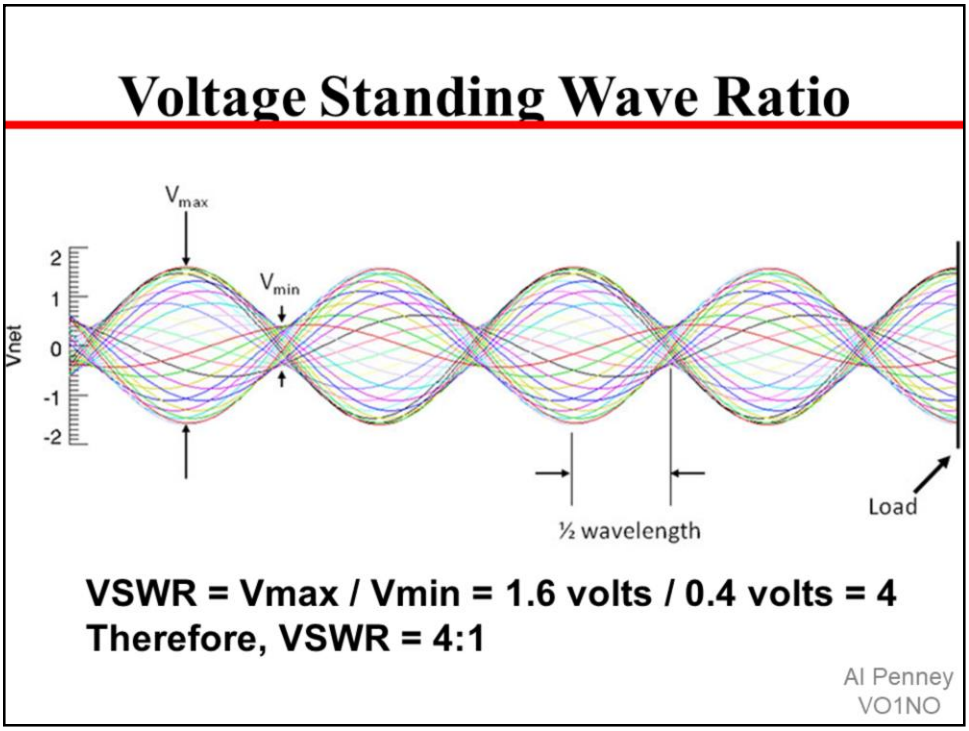
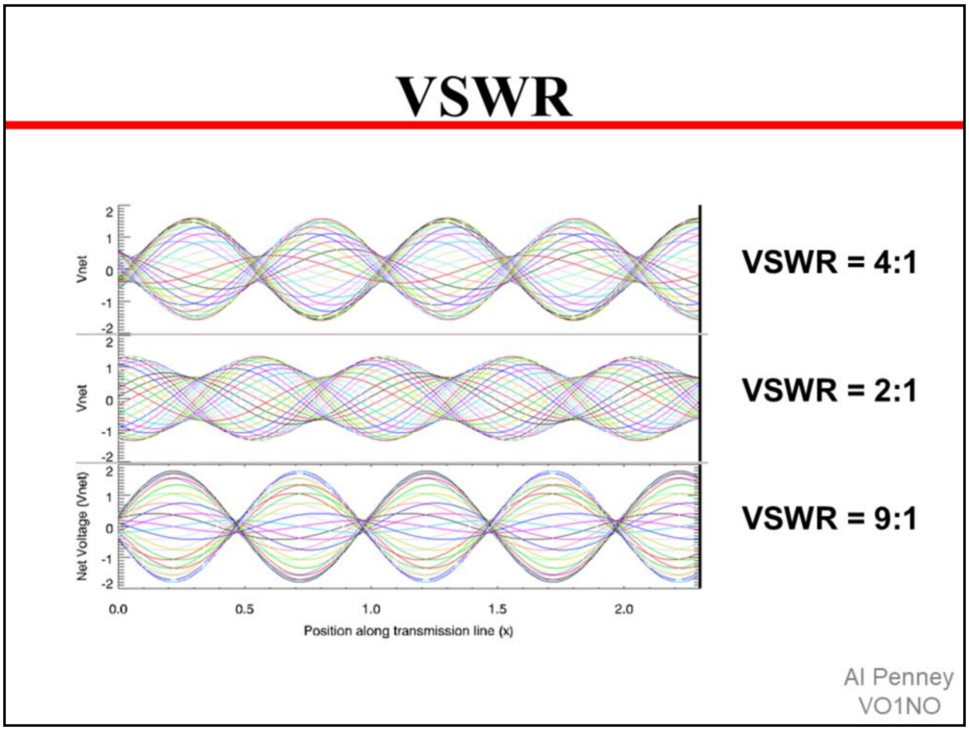
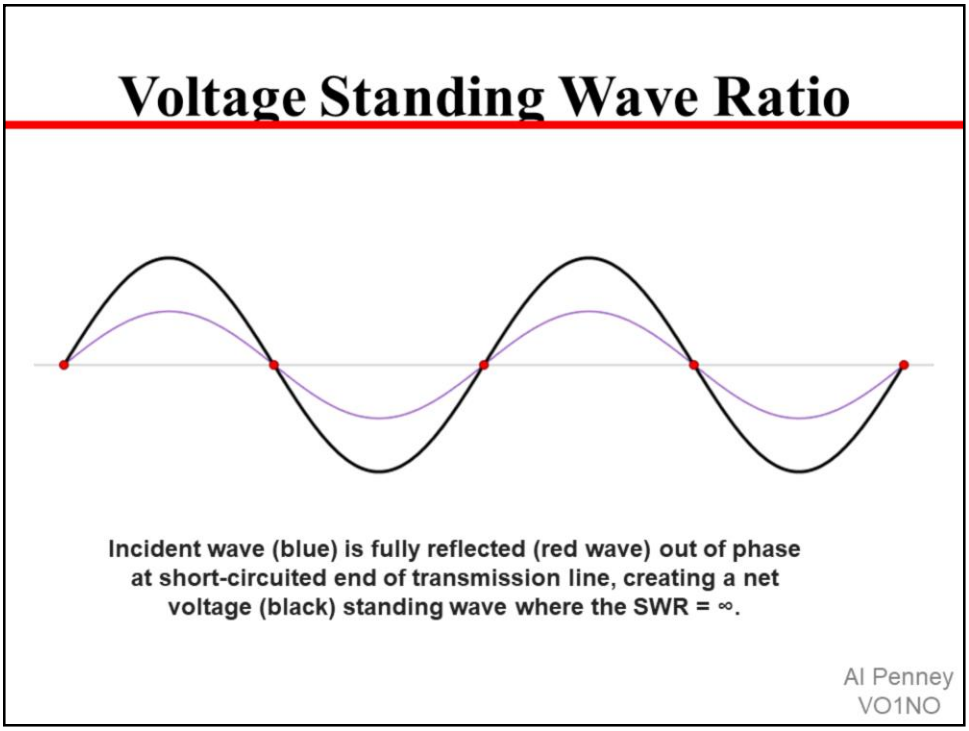


Standing waves on a transmission line, with the resulting voltages (VNET) shown in different colors during one complete cycle of the applied voltage. The generated wave with an amplitude of 1 travels from left to right and is partially reflected by the load at right. SWR = V_{MAX} / V_{MIN} = 1.6 / 0.4 = 4.0. *(Graphic created by Wikipedia contributor, Interferometrist and is used under the Creative Commons Attribution-Share Alike 4.0 International license.)*



Standing waves on transmission line, net voltage shown in different colors during one period of oscillation. Incoming wave from left (amplitude = 1) is partially reflected with (top to bottom) Γ = 0.6, −0.333, and 0.8 ∠60°. Resulting SWR = 4, 2, 9.



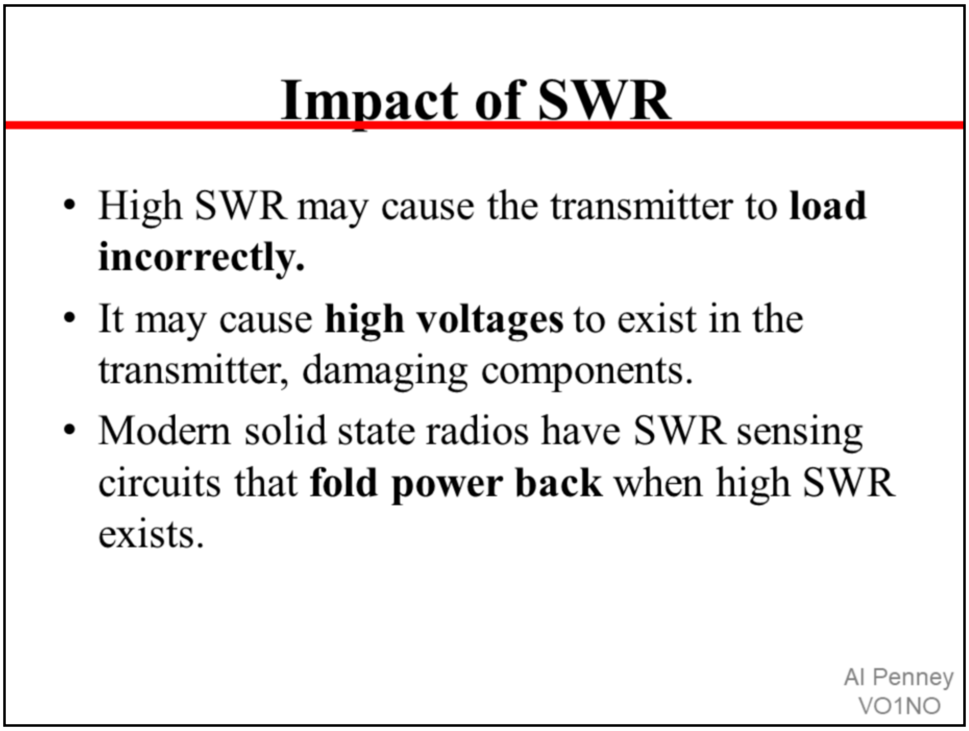
Impedance matching

SWR is used as a measure of impedance matching of a load to the characteristic impedance of a transmission line carrying radio frequency (RF) signals. This especially applies to transmission lines connecting radio transmitters and receivers with their antennas, as well as similar uses of RF cables such as cable television connections to TV receivers and distribution amplifiers. Impedance matching is achieved when the source impedance is the complex conjugate of the load impedance. The easiest way of achieving this, and the way that minimizes losses along the transmission line, is for the imaginary part of the complex impedance of both the source and load to be zero, that is, pure resistances, equal to the characteristic impedance of the transmission line. When there is a mismatch between the load impedance and the transmission line, part of the forward wave sent toward the load is reflected back along the transmission line towards the source. The source then sees a different impedance than it expects which can lead to lesser (or in some cases, more) power being supplied by it, the result being very sensitive to the electrical length of the transmission line.

Such a mismatch is usually undesired and results in standing waves along the transmission line which magnifies transmission line losses (significant at higher frequencies and for longer cables). The SWR is a measure of the depth of those standing waves and is, therefore, a measure of the matching of the load to the transmission line. A matched load would result in an SWR of 1:1 implying no reflected wave. An infinite SWR represents complete reflection by a load unable to absorb electrical power, with all the incident power reflected back towards the source.

It should be understood that the match of a load to the transmission line is different from the match of a *source* to the transmission line or the match of a source to the load *seen through* the transmission line. For instance, if there is a perfect match between the load impedance Z_{load} and the source impedance $Z_{source}=Z^{*}_{load}$, that perfect match will remain if the source and load are connected through a transmission line with an electrical length of one half wavelength (or a multiple of one half wavelengths) using a transmission line of *any* characteristic impedance Z_0. However the SWR will generally not be 1:1, depending only on Z_{load} and Z_0. With a different length of transmission line, the source will see a different impedance than Z_{load} which may or may not be a good match to the source. Sometimes this is deliberate, as when a quarter-wave matching section is used to improve the match between an otherwise mismatched source and load.

However typical RF sources such as transmitters and signal generators are designed to look into a purely resistive load impedance such as 50Ω or 75Ω, corresponding to common transmission lines' characteristic impedances. In those cases, matching the load to the transmission line, $Z_{load}=Z_0$, *always* ensures that the source will see the same load impedance as if the transmission line weren't there. This is identical to a 1:1 SWR. This condition ($Z_{load}=Z_0$) also means that the load seen by the source is independent of the transmission line's electrical length. Since the electrical length of a physical segment of transmission line depends on the signal frequency, violation of this condition means that the impedance seen by the source through the transmission line becomes a function of frequency (especially if the line is long), even if Z_{load} is frequency-independent. So in practice, a good SWR (near 1:1) implies a transmitter's output seeing the exact impedance it expects for optimum and safe operation.

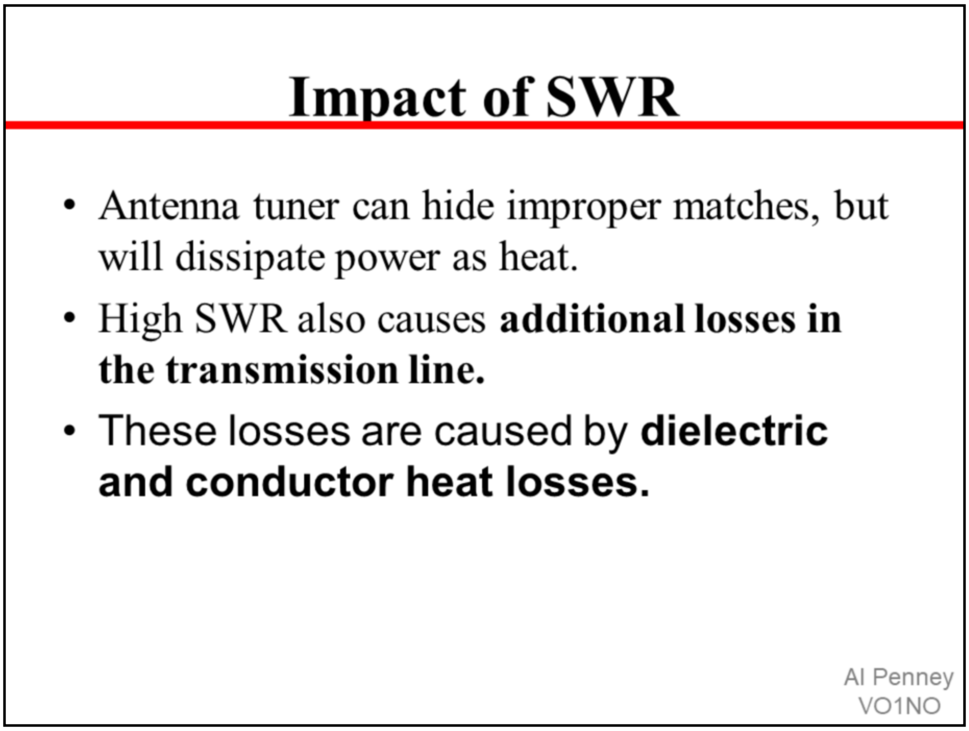


Practical implications of SWR

The most common case for measuring and examining SWR is when installing and tuning transmitting antennas. When a transmitter is connected to an antenna by a feed line, the driving point impedance of the antenna must match the characteristic impedance of the feed line in order for the transmitter to see the impedance it was designed for (the impedance of the feed line, usually 50 or 75 ohms).

The impedance of a particular antenna design can vary due to a number of factors that cannot always be clearly identified. This includes the transmitter frequency (as compared to the antenna's design or resonant frequency), the antenna's height above and quality of the ground, proximity to large metal structures, and variations in the exact size of the conductors used to construct the antenna.

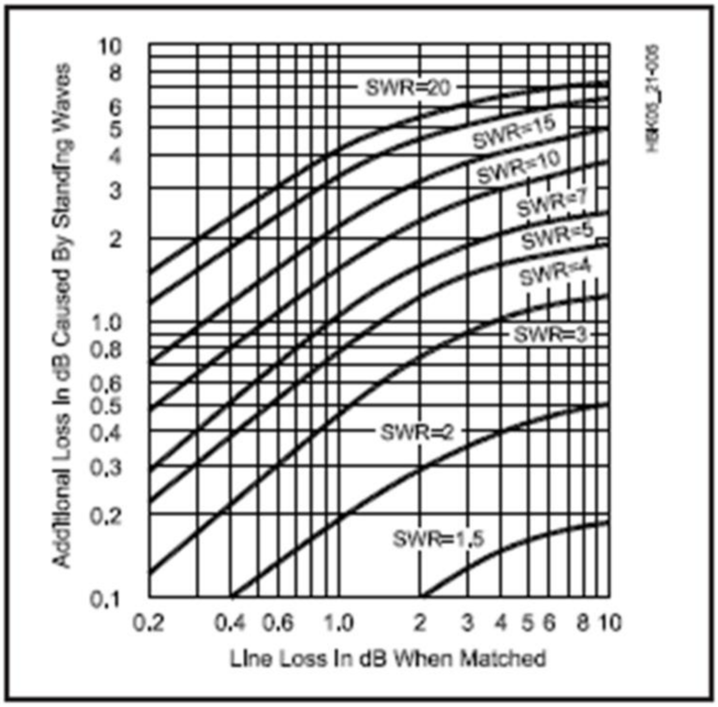
When an antenna and feed line do not have matching impedances, the transmitter sees an unexpected impedance, where it might not be able to produce its full power, and can even damage the transmitter in some cases. The reflected power in the transmission line increases the average current and therefore losses in the transmission line compared to power actually delivered to the load. It is the interaction of these reflected waves with forward waves which causes standing wave patterns, with the negative repercussions we have noted.



Matching the impedance of the antenna to the impedance of the feed line can sometimes be accomplished through adjusting the antenna itself, but otherwise is possible using an antenna tuner, an impedance matching device. Installing the tuner between the feed line and the antenna allows for the feed line to see a load close to its characteristic impedance, while sending most of the transmitter's power (a small amount may be dissipated within the tuner) to be radiated by the antenna despite its otherwise unacceptable feed point impedance. Installing a tuner in between the transmitter and the feed line can also transform the impedance seen at the transmitter end of the feed line to one preferred by the transmitter. However, in the latter case, the feed line still has a high SWR present, with the resulting increased feed line losses unmitigated.

The magnitude of those losses are dependent on the type of transmission line, and its length. They always increase with frequency. For example, a certain antenna used well away from its resonant frequency may have an SWR of 6:1. For a frequency of 3.5 MHz, with that antenna fed through 75 meters of RG-8A coax, the loss due to standing waves would be 2.2 dB. However the same 6:1 mismatch through 75 meters of RG-8A coax would incur 10.8 dB of loss at 146 MHz. Thus, a better match of the antenna to the feed line, that is, a lower SWR, becomes increasingly important with increasing frequency, even if the transmitter is able to accommodate the impedance seen (or an antenna tuner is used between the transmitter and feed line).

Certain types of transmissions can suffer other negative effects from reflected waves on a transmission line. Analog TV can experience "ghosts" from delayed signals bouncing back and forth on a long line. FM stereo can also be affected and digital signals can experience delayed pulses leading to bit errors. Whenever the delay times for a signal going back down and then again up the line are comparable to the modulation time constants, effects occur. For this reason, these types of transmissions require a low SWR on the feedline, even if SWR induced loss might be acceptable and matching is done at the transmitter.



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ADDITIONAL POWER LOSS DUE TO SWR

The power lost in a given line is least when the line is terminated in a resistance equal to its characteristic impedance, and as stated previously, that is called the *matched-line loss*. There is however an *additional loss* that increases with an increase in the SWR. This is because the effective values of both current and voltage become greater on lines with standing waves. The increase in effective current raises the ohmic losses (I2R) in the conductors, and the increase in effective voltage increases the losses in the dielectric (E2/R).

The increased loss caused by an SWR greater than 1:1 may or may not be serious. If the SWR at the load is not greater than 2:1, the additional loss caused by the standing waves, as compared with the loss when the line is perfectly matched, does not amount to more than about 1/2 dB, even on very long lines. One-half dB is an undetectable change in signal strength. Therefore, it can be said that, from a practical standpoint in the HF bands, an SWR of 2:1 or less is every bit as good as a perfect match, so far as additional losses due to SWR are concerned.

However, above 30 MHz, in the VHF and especially the UHF range, where low receiver noise figures are essential for effective weak-signal work, matched-line losses for commonly available types of coax can

be relatively high. This means that even a slight mismatch may become a concern regarding overall transmission line losses. At UHF one-half dB of additional loss may be considered intolerable!

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The SWR meter is placed between the transmitter and the feedline to the antenna

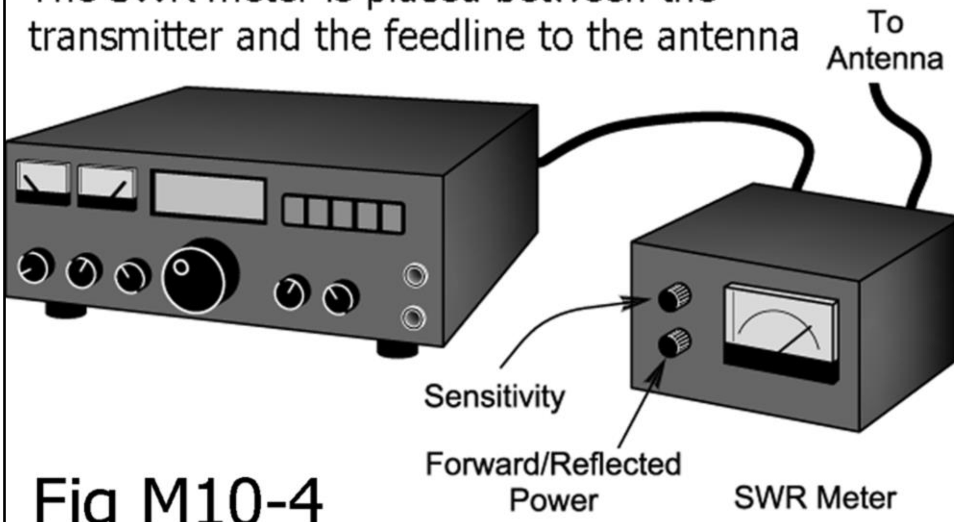

Fig M10-4

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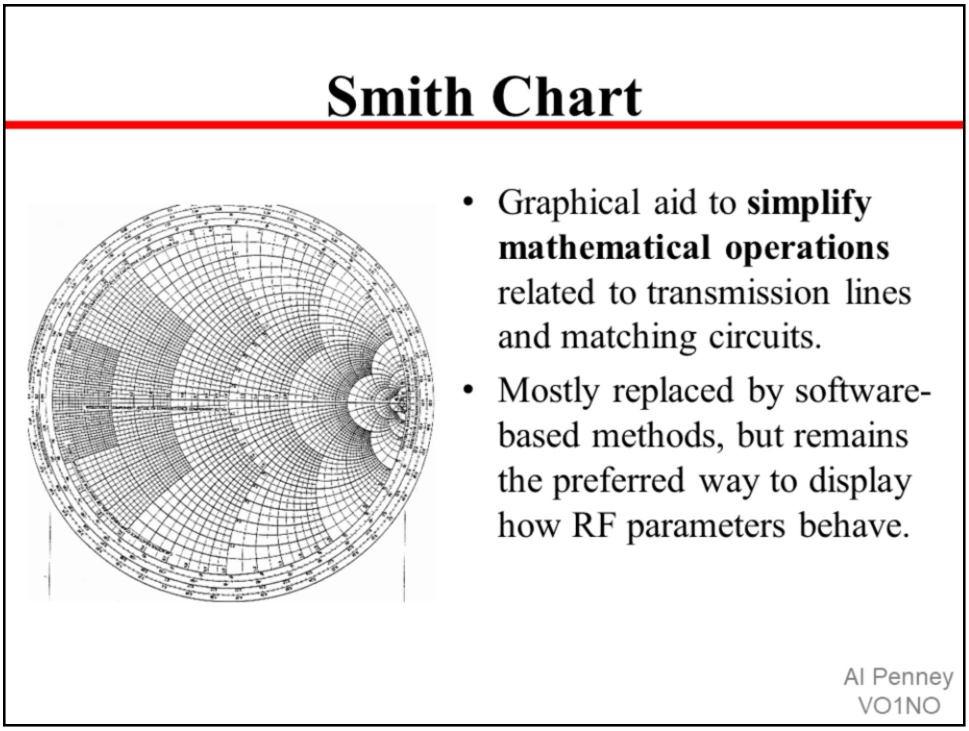
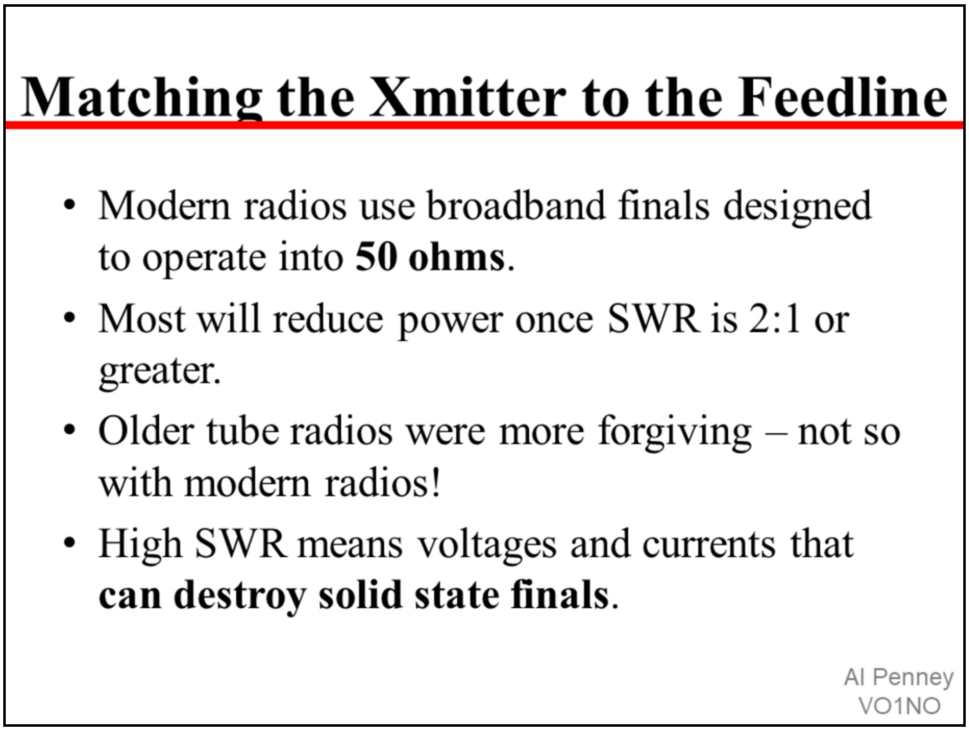


The **Smith Chart** is a fantastic tool for visualizing the impedance of a transmission line and antenna system as a function of frequency. Smith Charts can be used to increase understanding of transmission lines and how they behave from an impedance viewpoint. Smith Charts are also extremely helpful for **impedance matching**, as we will see. The Smith Chart is used to display an actual (physical) antenna's impedance when measured on a Vector Network Analyzer (VNA).

Smith Charts were originally developed around 1940 by Phillip Smith as a useful tool for making the equations involved in transmission lines easier to manipulate. See, for instance, the input impedance equation for a load attached to a transmission line of length *L* and characteristic impedance ZO. With modern computers, the Smith Chart is no longer used to the simplify the calculation of transmission line equatons; however, their value in visualizing the impedance of an antenna or a transmission line has not decreased.

The **Smith chart**, invented by Phillip H. Smith (1905–1987), is a graphical aid or nomogram designed for electrical and electronics engineers specializing in radio frequency (RF) engineering to assist in solving problems with transmission lines and matching circuits. The Smith chart can be used to simultaneously display multiple parameters including impedances, admittances, reflection coefficients, scattering parameters, noise figure circles, constant gain contours and regions for unconditional stability, including mechanical vibrations analysis. The Smith chart is most frequently used at or within the unity radius region. However, the remainder is still mathematically relevant, being used, for example, in oscillator design and stability analysis.[6]

While the use of paper Smith charts for solving the complex mathematics involved in matching problems has been largely replaced by software based methods, the Smith charts display is still the preferred method of displaying how RF parameters behave at one or more frequencies, an alternative to using tabular information. Thus most RF circuit analysis software includes a Smith chart option for the display of results and all but the simplest impedance measuring instruments can display measured results on a Smith chart display.

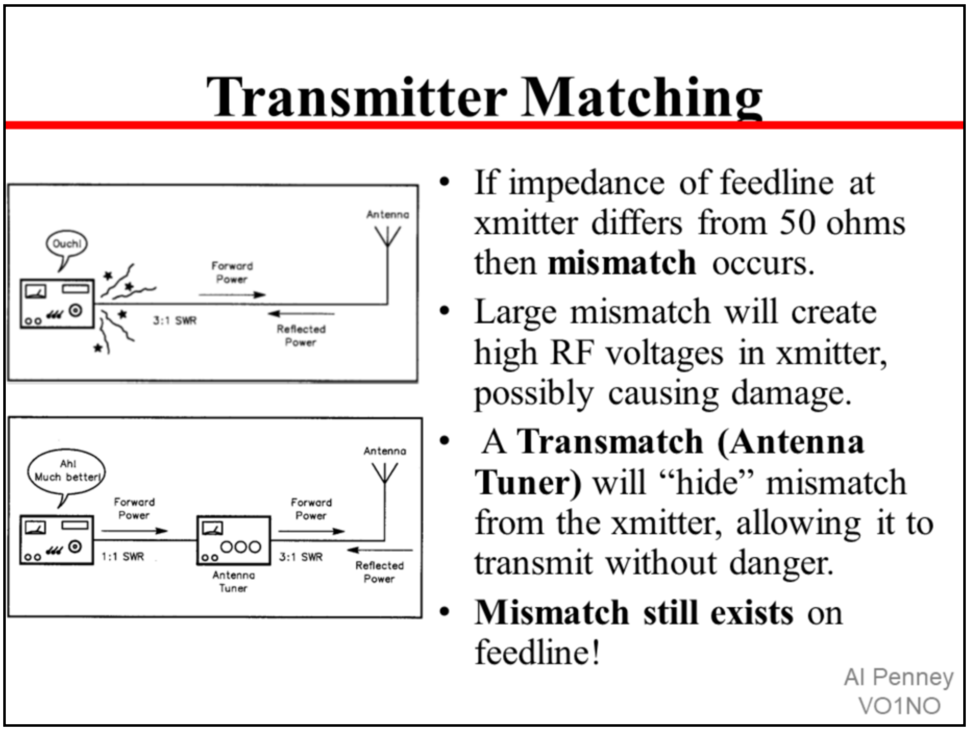


Modern amateur transceivers use broadband, untuned solid-state final amplifiers, designed to operate into 50 οημσ. Such a transmitter is able to deliver its rated output power—at its rated level of distortion—only when it is operated into the load for which it was designed. An SSB transmitter that is "splattering" is often being driven hard into the wrong load impedance.

Further, modern radios often employ protection circuitry to reduce output power automatically if the SWR rises to more than about 2:1. Protective circuits are needed because solid-state devices can

almost instantly destroy themselves trying to deliver power into the wrong load impedance. Modern solid-state transceivers often include built-in antenna tuners (often at extra cost) to match impedances

when the SWR isn't 1:1. Older vacuum-tube amplifiers were a lot more forgiving than solid-state devices—they could survive momentary overloads without being instantly destroyed. The pi-networks used to tune and load old-fashioned vacuum-tube amplifiers were able to match a fairly wide range of impedances.



MATCHING THE LINE TO THE TRANSMITTER

The impedance at the input of a transmission line is uniquely determined by a number of factors: **the frequency, the characteristic impedance Z0 of the line, the physical length,**

velocity factor and the matched-line loss of the line, plus the impedance of the load (the antenna) at the output end of the line. If the impedance at the input of the transmission line connected to the transmitter differs appreciably from the load resistance into which the transmitter output circuit is designed to operate, an impedance-matching network must be inserted between the transmitter and the line input terminals.

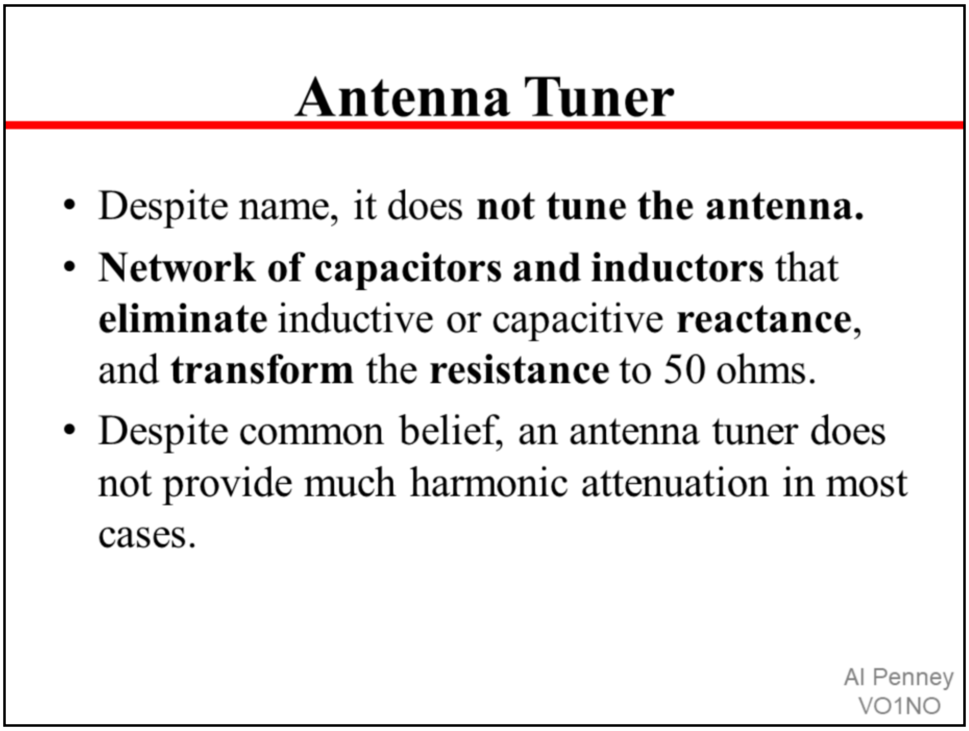
In older ARRL publications, such an impedance-matching network was often called a *Transmatch*. This is a coined word, referring to a "Transmitter Matching" network. Nowadays, radio amateurs commonly

call such a device an *antenna tuner*.

The function of an antenna tuner is to transform the impedance at the input end of the transmission line—whatever it may be—to the 50 οημσ needed to keep the transmitter loaded properly. An antenna

tuner does *not* alter the SWR on the transmission line going to the antenna. It only ensures that the transmitter sees the 50-οημ load for which it was designed.

http://www.arrl.org/files/file/Technology/tis/info/pdf/9401070.pdf



What an "antenna tuner" actually tunes

Despite its name, an antenna "tuner" does not actually tune the antenna. It matches the complex impedance of the transmitter to that of the input end of the feedline. The input impedance of the transmission line will be different than the characteristic impedance of the feedline if the impedance of the antenna on the other end of the line does not match the line's characteristic impedance. The consequence of the mismatch is standing waves on the feedline that alter the line's impedance at every point along the line.

If both the tuner and the feedline were lossless, tuning at the transmitter end would indeed produce a perfect match at every point in the transmitter-feedline-antenna system. However, in practical systems lossy feedlines limit the ability of the antenna tuner to change the antenna's resonant frequency. If the loss of power is low in the line carrying the transmitter's signal to the antenna, a tuner at the transmitter end can produce a worthwhile degree of matching and tuning for the antenna and feedline network as a whole. But with lossy, low-impedance feedlines like the commonly used 50 Ohm coaxial cable, maximum power transfer only occurs if matching is done at the antenna in conjunction with a matched transmitter and feedline, producing a match at both ends of the line.

In any case, regardless of its placement, an ATU does not alter the gain, efficiency, or directivity of the antenna, nor does it change the internal complex impedance of the antenna itself.

Efficiency and SWR

If there is still a high standing wave ratio (SWR) in the feedline beyond the ATU, any loss in that part of the feedline is typically increased by the transmitted waves reflecting back and forth between the tuner and the antenna, causing resistive losses in the wires and possibly the insulation of the transmission line. Even with a matching unit at both ends of the feedline – the near ATU matching the transmitter to the feedline and the remote ATU matching the feedline to the antenna – losses in the circuitry of the two ATUs will slightly reduce power delivered to the antenna.

Hence, operating an antenna far from its design frequency and compensating with an ATU between the transmitter and the feedline is not as efficient as using a resonant antenna with a matched-impedance feedline, nor as efficient as a matched feedline from the transmitter to a remote antenna tuner attached directly to the antenna.

Harmonic Attenuation in an Antenna Tuner

One potentially desirable characteristic of an antenna tuner is the degree of extra harmonic attenuation it can provide. While this is desirable in

theory, it is not always achieved in practice. For example, if an antenna tuner is used with a single, fixed-length antenna on multiple bands, the impedances presented to the tuner at the fundamental

frequency and at the harmonics will often be radically different, thus attenuating harmonics. There are some situations in Amateur Radio where the impedance at the second harmonic is essentially the same as that for the fundamental however, providing little attenuation to harmonics. This often involves "trapped" antenna systems or wideband log-periodic designs. For example, a system used by many amateurs is a "tribander" Yagi that works on 20, 15 and 10 meters. The second harmonic of a 20-meter transmitter feeding such a tribander can be objectionably strong for nearby amateurs operating on 10 meters. This is despite the approximately 60 dB of attenuation of the second harmonic provided by the low-pass filters built into modern solid state transceivers. A linear amplifier can exacerbate the problem, since its second harmonic may be suppressed only about 46 dB by the typical pi-network output circuit used in most amplifiers. Even in a trapped antenna system, most amateur antenna tuners will not attenuate the 10-meter harmonic much at all, especially if the tuner uses a high-pass T-network. This is the most common network used commercially because of the wide range of impedances it will match. Some T-network designs have attempted to improve the harmonic attenuation using parallel inductors and capacitors instead of a single inductor for the center part of the T.

Unfortunately, this often leads to more loss and more critical tuning at the fundamental, while providing little, if any, additional harmonic suppression in actual installations.

Harmonics and Pi-Network Tuners

In a trapped antenna system, if a different network is used for an antenna tuner (such as a lowpass Pi network), there will be additional attenuation of harmonics, perhaps as much as 30 dB for a

loaded Q of 3. The exact degree of harmonic attenuation, however, is often limited due to the stray inductance and capacity present in most tuners at harmonic frequencies. Further, the matching range

for a Pi-network tuner is fairly limited because of the range of input and output capacitance needed for widely varying loads.

Harmonics and Stubs

Far more reliable suppression of harmonics can be achieved using shorted quarter-wave transmission- line stubs at the transmitter output. A typical 20-meter $\lambda/4$ wavelength shorted stub (which looks like an open circuit at 20 meters, but a short circuit at 10 meters) will provide about 25 dB of attenuation to the second harmonic. It will handle full legal amateur power too. In short, an antenna tuner that is capable of matching a wide range of impedances should not be relied on to give additional harmonic suppression.

Antenna Tuner Circuits

- Common Antenna Tuner circuits include:
  - **L Network;**
  - **Pi Network;** and
  - **T Network.**
- Other types include:
  - **Inductive Coupling Networks**; and
  - **Series Networks**.
- Each has advantages and limitations.

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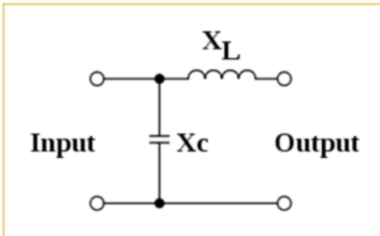
L Network



- So named because of shape.
- Uses variable capacitor and variable inductor.
- Four different combinations possible, though only two are commonly used.

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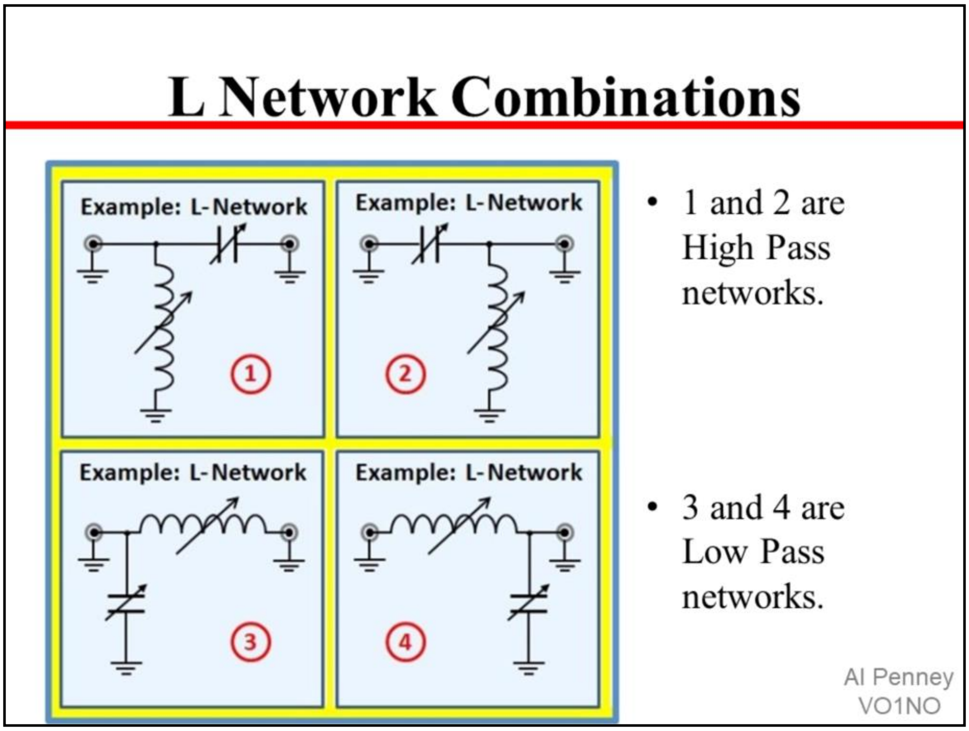


Shown above are the 4 "possible" configurations of an L-Network.

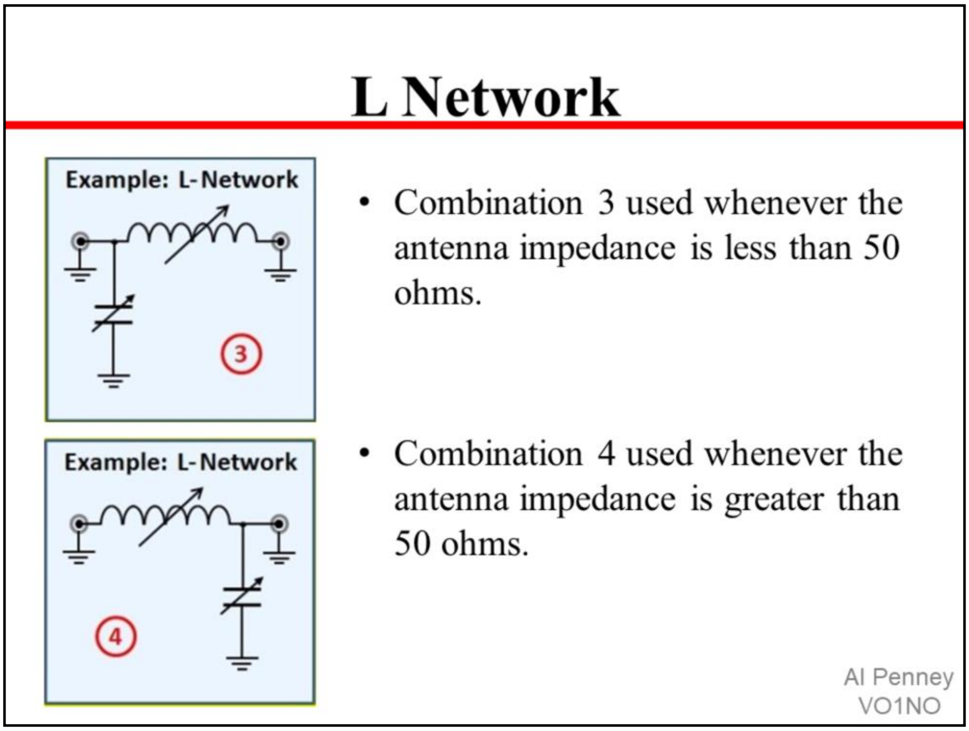
Versions 1 and 2 are high-pass matching networks.

Versions 3 and 4 are low-pass matching networks.

In practice, usually only versions 3 and 4 are used for Antenna Matchboxes.

•**VERSION 3** is used whenever the antenna has **LOW impedance** (less than 50 Ohms)

•**VERSION 4** is used whenever the antenna has **HIGH impedance** (greater than 50 Ohms)



- **VERSION 3** is used whenever the antenna has **LOW impedance** (less than 50 Ohms)
- **VERSION 4** is used whenever the antenna has **HIGH impedance** (greater than 50 Ohms)

L Network Pros and Cons

- Pros:
  - Easy to tune to optimum setting;
  - Only one setting possible per band; and
  - Less loss than other types of matching networks.
- Cons:
  - Limited impedance matching range; and
  - Requires robust components;

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Advantages of L-Network:

•Easy to tune to the optimum setting

•Only one setting (per antenna, per band) will yield the optimum SWR

•When set for minimum SWR, it always corresponds to maximum efficiency

•Fewer components than with a T-Filter (less knobs to tune)

•LESS LOSS THAN WITH T-FILTER !!! (BIGGEST ADVANTAGE)

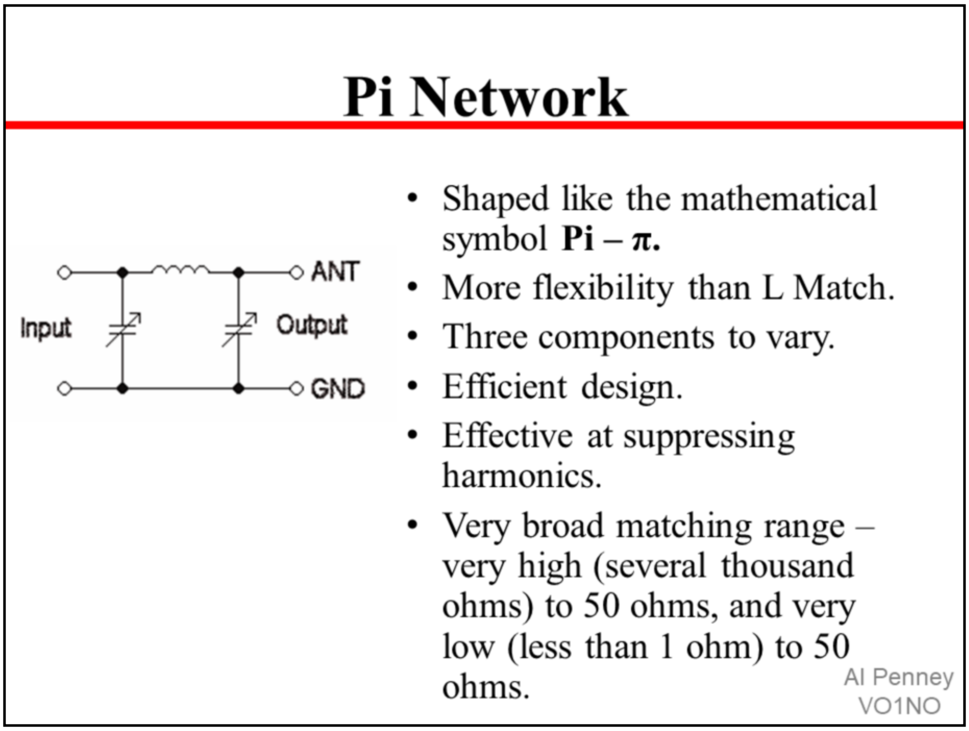
Disadvantages of L-Network:

•To cover the same impedance matching range as the T-Filter, it requires some pretty extreme component values, AND. . .

•It also requires re-configuration of the inductor and capacitor to all 4 circuits shown above.

•In practical applications, achieving the extreme component values needed requires switching the capacitor back and forth between the input and output side of the network, switching multiple capacitors in parallel with the main tuning capacitor, and swapping the positions of the inductor and capacitor.

- **Without all of the above, its impedance matching range is limited, and not as broad as that of the T-Filter matchbox.**
- **There are very few commercial (manual) L-Network tuners on the market and all of them are relatively expensive. Here are a couple:**
- **TEN-TEC Model 238/238A/238B/238C** ("C" is the current model)
- **Viking Nye Matchbox** (only available on the used market)



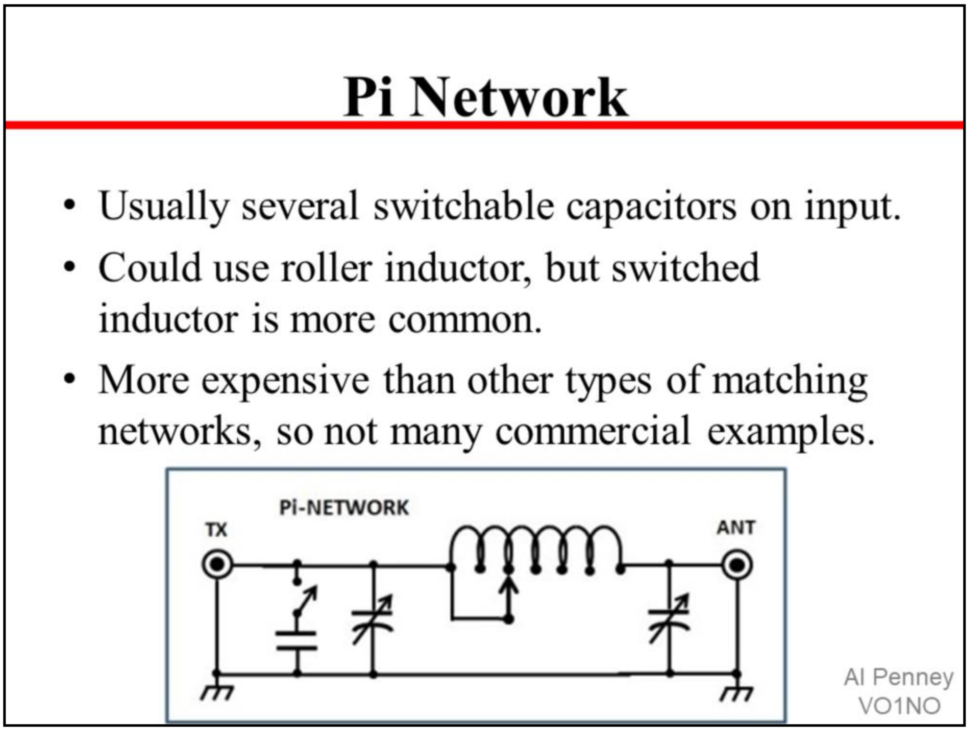
THE PI-NETWORK

The impedances at the feed point of an antenna used on multiple HF bands varies over a very wide range, particularly if thin wire is used. The transmission line feeding the antenna transforms

the wide range of impedances at the antenna's feed point to another wide range of impedances at the transmission line's input. This often mandates the use of a more flexible antenna tuner than an Lnetwork.

The Pi-network offers more flexibility than the L-network, since there are three variables instead of two. The only limitation on the circuit values that may be used is that the reactance

of the series arm, the inductor L in the figure, must not be greater than the square root of the product of the two values of resistive impedance to be matched.



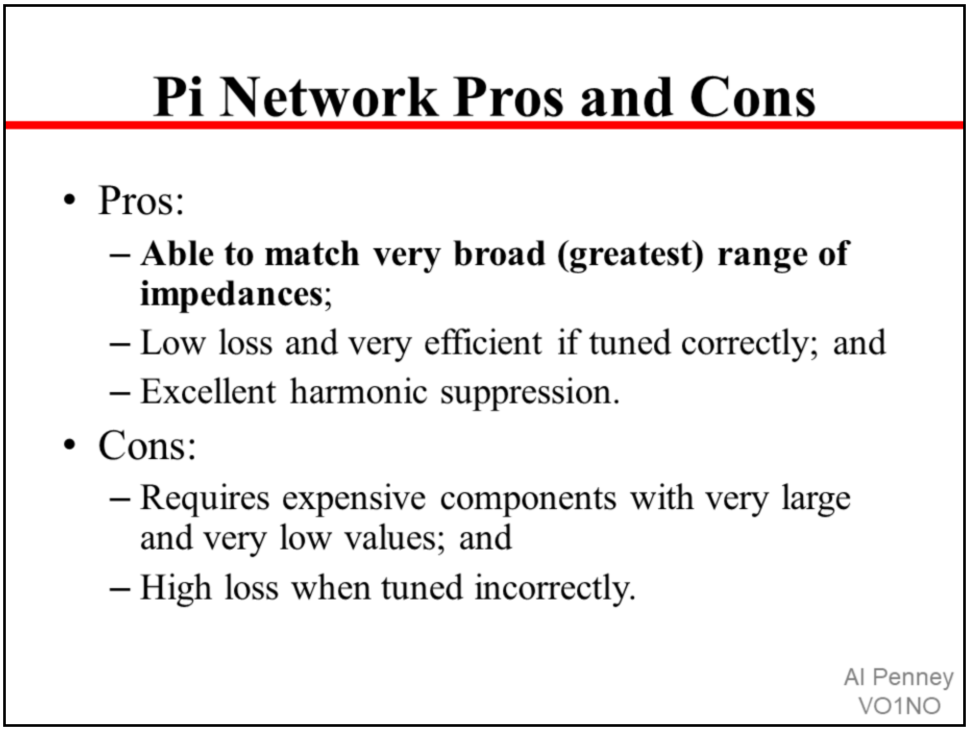
The Pi-Network Matchbox, like the Link-Coupled Matchbox, is more of a device from the past, rather than the present. This is in no way due to its performance, but rather to the cost of the components required to implement a Pi-Network Matchbox capable of matching the required impedances on all HF bands.

The circuit above shows an extra capacitor on the input side. In reality there would probably be several capacitors, which are switchable on demand, depending on the antenna's impedance. We do this rather than using a larger variable capacitor because a larger variable capacitor would have too high of a minimum capacitance to be used on 10m.

Although you could build this matchbox with a roller inductor, you usually only see it with switched inductance. The only Pi-Network matchbox still on the market today is a Remote Matchbox and uses switched inductance.

The Pi-Network matchbox has very good efficiency and also has the added advantage of surpressing harmonics. Although this is old technology, it is still a great concept. It's only problem is, it is expensive to implement in a way that it has a broad impedance matching range over the entire hf spectrum.

The Pi-network may be used to match a low impedance to a rather high one, such as 50 to several thousand ohms. Conversely, it may be used to match 50 οημ to a quite low value, such as 1 οημ or less. For antenna-tuner applications, C1 and C2 may be independently variable. L may be a roller inductor or a coil with switchable taps. Alternatively, a lead fitted with a suitable clip may be used to short out turns of a fixed inductor. In this way, a match may be obtained through trial. It will be possible to match two values of impedances with several different settings of L, C1 and C2. This results because the Q of the network is being changed. If a match is maintained with other adjustments, the Q of the circuit rises with increased capacitance at C1. Of course, the load usually has a reactive component along with resistance. The effect of these reactive components can be compensated for by changing one of the reactive elements in the matching network. For example, if some reactance was shunted across R2, the setting of C2 could be changed to compensate, whether that shunt reactance be inductive or capacitive. As with the L network, the effects of real-world unloaded Q for each component must be taken into account in the Pi-network to evaluate real-world losses.



ADVANTAGES OF THE Pi-NETWORK:

•**Able to match a very broad range of impedances**

•**Able to match the entire hf spectrum (assuming sufficient switchable components)**

•**Low loss / very efficient (if tuned correctly)**

•**Has excellent harmonic surpression**

DISADVANTAGES OF THE Pi-NETWORK:

•**High loss when tuned wrong**

•**Requires components with very large and very low values**

•**Components are VERY EXPENSIVE. Much more so than with other matchbox types**

•**Only one commercial source of Pi-Network Matchboxes (HamWare* in Germany)**

•**HamWare AT615-U (1.5 kW)**

•**This is a manual remote matchbox with separate matching unit for outside and control unit for the desktop**

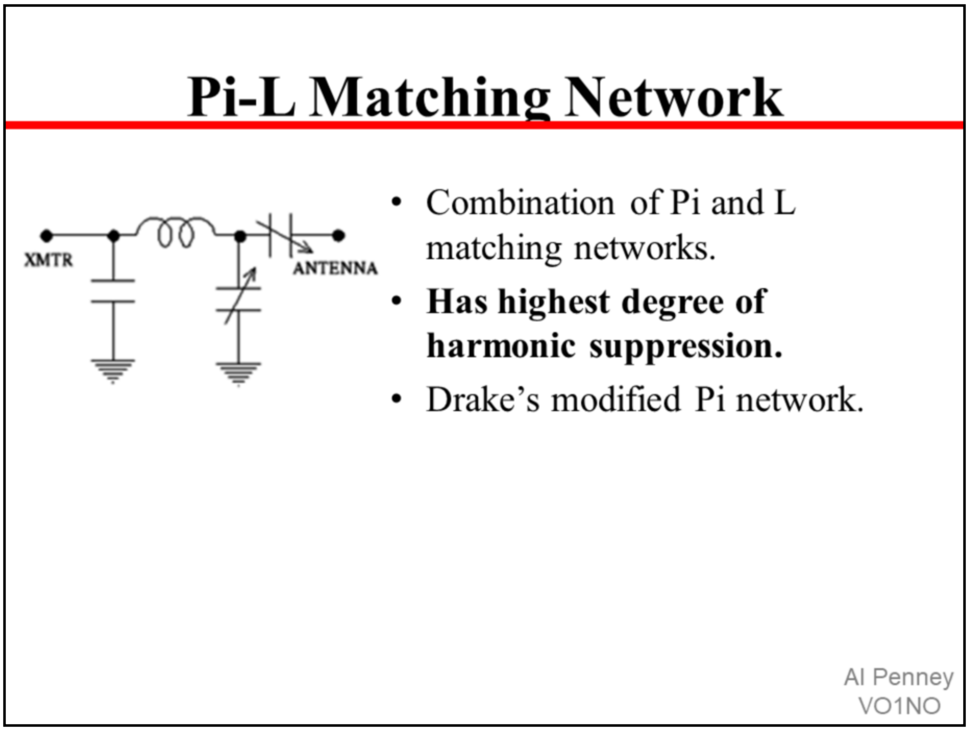
•**HamWare also make two similar Symmetrical Pi-NETWORK**

matchboxes. More on this under Symmetrical Matchboxes.

COMMENTS:

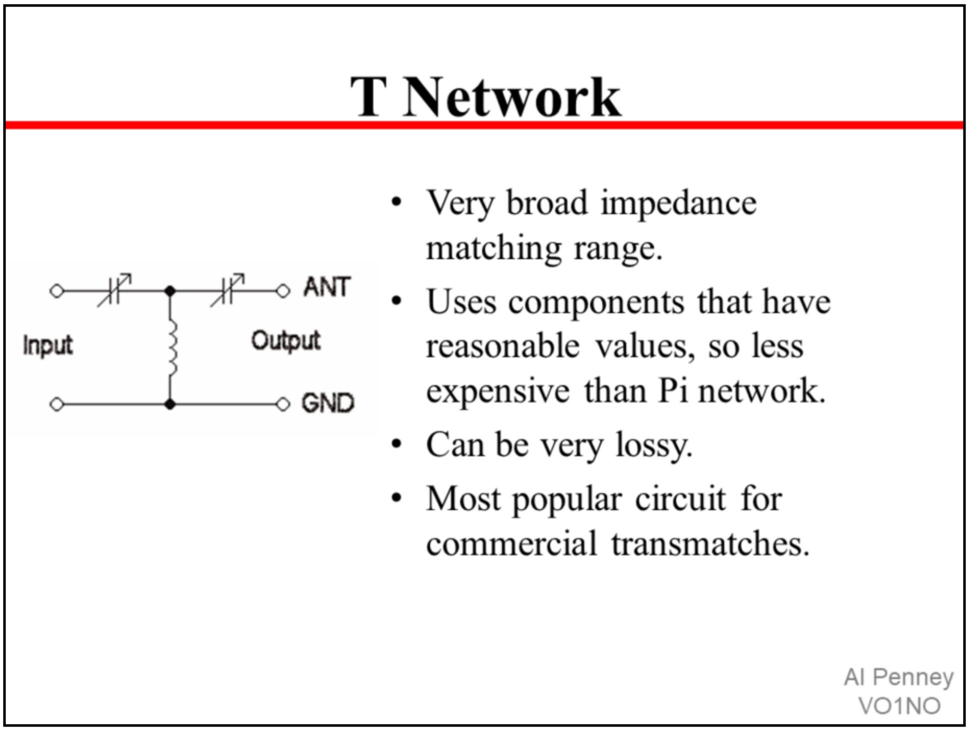
The Pi-Network Matchbox is a great technology, with very low internal loss and it also offers good harmonic surpression. Unfortunately the cost of components today has delegated this technology to the history books.

In the past, companies such as R.L. DRAKE marketed Pi-Network matchboxes. In the meantime these have become collectors' items and are difficult to find.



Drake's modified π-network

A modified version of the π-network is more practical as it uses a fixed input capacitor, which can be several thousand picofarads, allowing the two variable capacitors to be smaller. A band switch selects the input capacitor and inductor. This circuit was used in tuners covering 1.8 to 30 MHz made by the R. L. Drake Company.



In the past 30 years, the **T-Network** has established itself as the most popular matchbox design found in most ham shacks. It is a simple design that easily covers multiple bands, and it can be built using common, inexpensive components.

The T-Network can easily match a broad range of impedances, making it the ideal matchbox for matching "whatever comes your way". Unfortunately it is not always the most effecient matchbox.

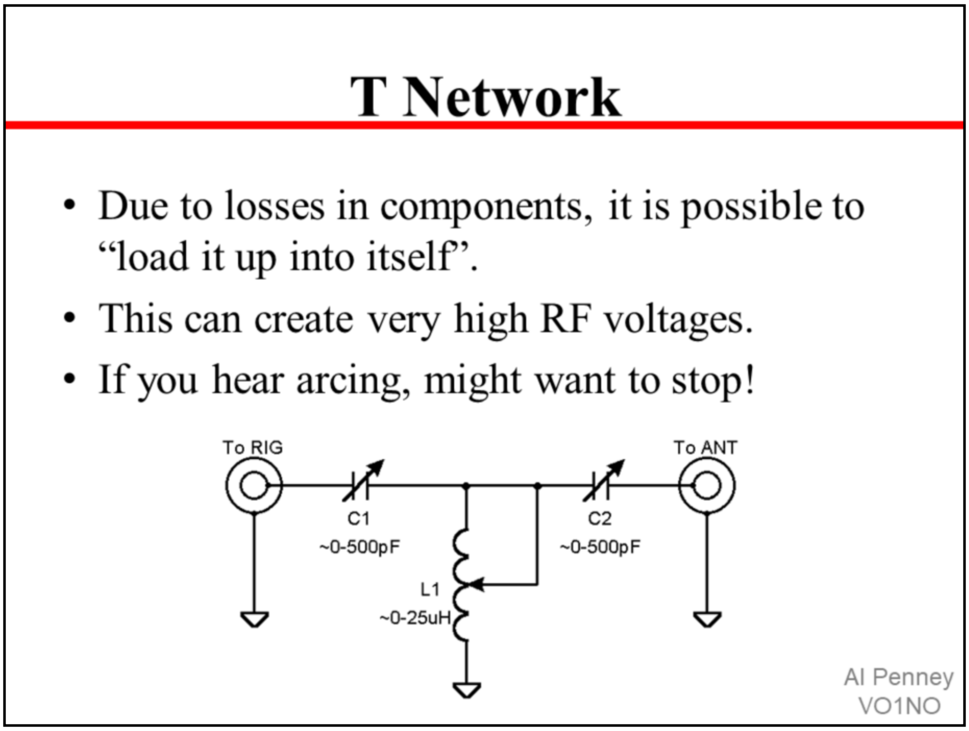
The BIGGEST PROBLEM with the T-Network is its internal loss - power lost inside the matchbox. Through improper tuning by the operator, the losses can be even higher - sometimes 50% or more.

THE T-NETWORK

Both the Pi-network and the L-network often require unwieldy values of capacitance—that is, *large* capacitances are often required at the lower frequencies—to make the desired transformation to 50 οημσ. Often, the range of capacitance from minimum to maximum must be quite wide when the impedance at the output of the network varies radically with frequency, as is common for multiband, single-wire antennas. The T-network shown is capable of matching a wide range of load impedances and uses practical values for the components. However, as in almost everything in radio, there is a price to be

paid for this flexibility. The T-network can be very lossy compared to other network types. This is particularly true at the lower frequencies, whenever the load resistance is low. Loss can be severe if the

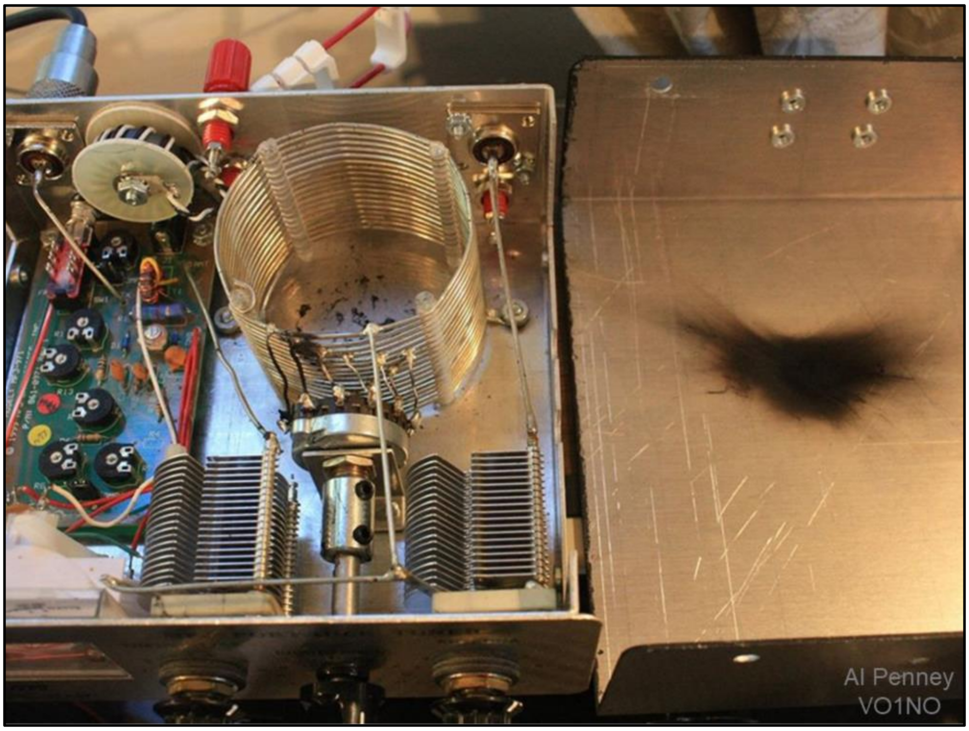
maximum capacitance of the output capacitor is low.



Due to the losses in the components in a T-network, it is quite possible to "load it up into itself," causing real damage inside. For example, see the example in the ARRL Antenna Book where a T-network is loaded up into a short circuit at 1.8 MHz. The component values look quite reasonable, but unfortunately *all* the power is dissipated in the network itself. The current through the output capacitor at 1500 W input to the antenna tuner would be 35 A, creating a peak voltage of more than 8700 V across the output capacitor. Either of the two capacitors will probably arc over before the power loss is

sufficient to destroy the coil. However, the loud arcing might frighten the operator pretty badly!

The point you should remember is that the T network is indeed very flexible in terms of matching to a wide variety of loads. However, it must be used judiciously, lest it burn itself up. Even if it doesn't fry itself, it can waste that precious RF power you'd rather put into your antenna.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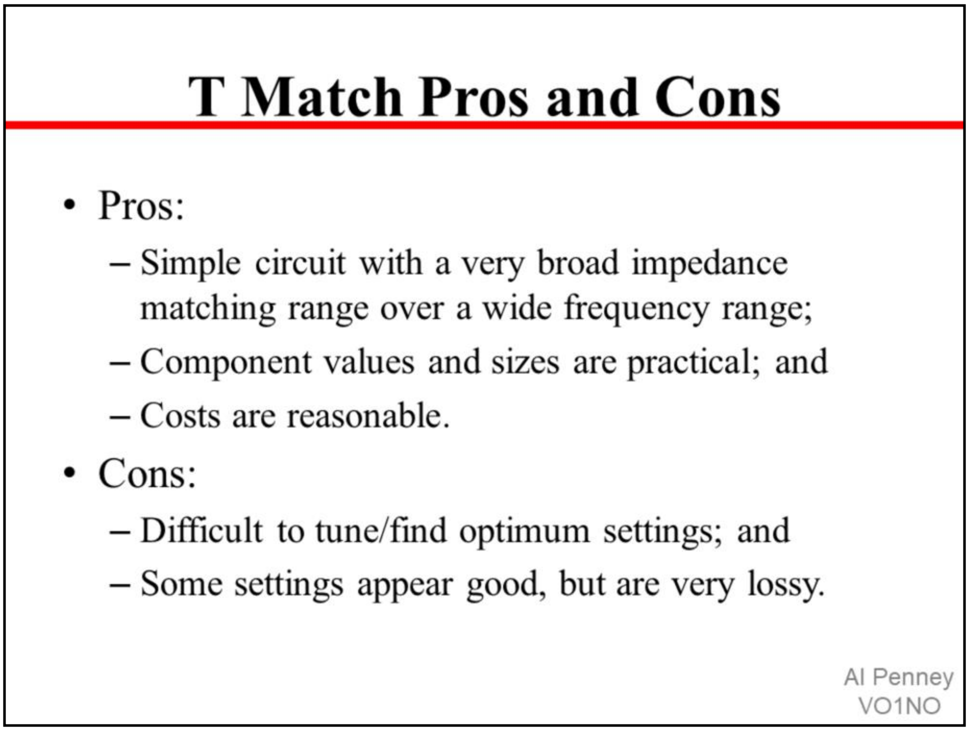

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The BIGGEST PROBLEM with the T-Network is its internal loss - power lost inside the matchbox. Through improper tuning by the operator, the losses can be even higher - sometimes 50% or more.

Advantages of T-Network Matchboxes:

•Simple circuit covers a broad frequency range

•Able to easily match a very broad impedance range

•Component values and physical sizes are practical

•Component costs are reasonable

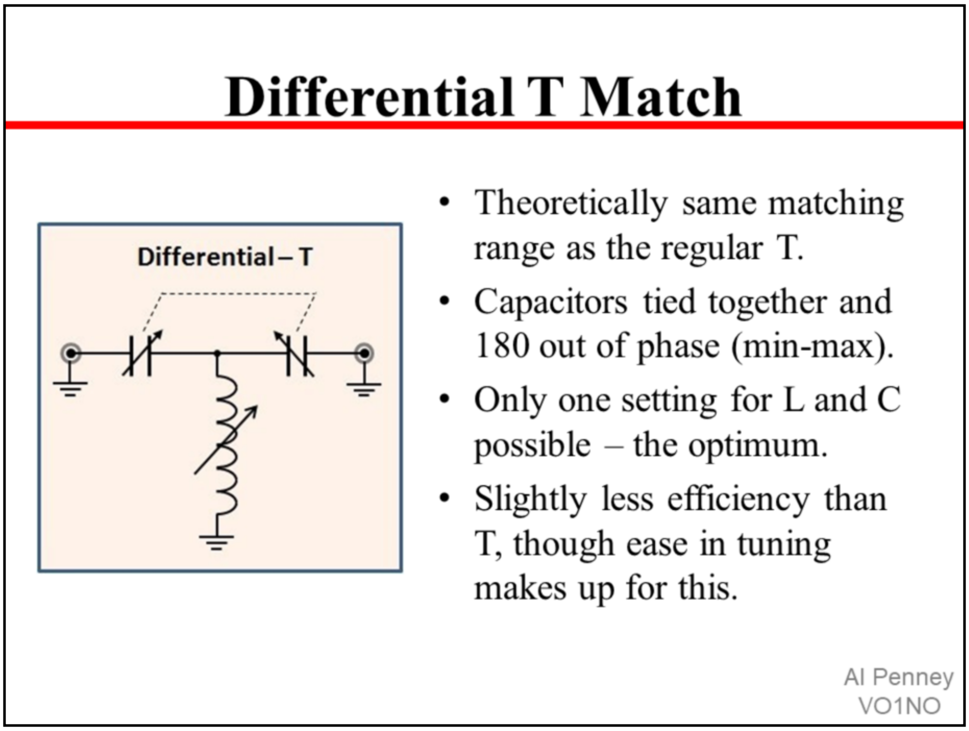
Disadvantage of T-Network Matchboxes:

- **Difficult to tune / difficult to find the best setting**
- **Multiple settings "appear" to be optimum (show low SWR), but . . .**
- **. . . when tuned to these wrong settings, these matchboxes heat up, have unnecessary loss, and can even burn up.**
- **VERY OFTEN USED AT LESS THAN OPTIMUM EFFICIENCY SETTING !!!**
- **The operator needs to engage his brain when tuning!**

COMMENTS:

This is probably the best matchbox technology for most hams who are just beginning and do not know yet what they will ultimately need. Selecting this matchbox technology is a low-risk choice, BUT IF YOU DON'T TUNE IT RIGHT, YOU MAY LOSE UP TO 50% OF YOU POWER INSIDE THE MATCHBOX. It is possible to get better solutions for some antenna applications, but at least this matchbox should always find a good match to the antenna. **GOOD STARTER MATCHBOX!**

IF YOU CHOOSE THIS MATCHBOX, YOU <u>MUST</u> LEARN HOW TO TUNE IT PROPERLY!



The " Differential-T " . . . (a good compromise)

- Theoretically it has the same broad matching impedance range as the normal T-Filter matchbox.
- It only finds one setting of L and C which provide the best SWR. This is also the setting with the highest efficiency. This eliminates the possibility of tuning to the wrong position. (nothing can go wrong)
- In theory, slightly more loss than a properly tuned classical T-Matchbox.
- Although it theoretically has a slightly lower efficiency than the standard T-Matchbox, in practical situations it is better because you can never tune it to the wrong settings, so you always have it tuned to its optimum setting.
- BUT, the physical limits of the size of the enclosure prohibits using variable capacitors with as many pF as the standard T-Filter uses; as a result, the impedance range that you can match with the models that are on the market is not as great as the range of a standard T-Filter matchbox.

At first glance, it looks just like the classical T-Filter matchbox.

A closer look reveals the differences:

1. The input and output capacitors are tied together and tuned

together.

2. The input and output capacitors are 180 degrees out of phase; when the input capacitor is at maximum capacitance, the output capacitor is at minimum, and vice versa.

These simple changes result in the circuitry finding only one single point of best SWR, and this point always coincides with maximum effeciency. This eliminates the main problem with the original T-Filter matchbox: user error.

In the interest of keeping costs low, instead of using two separate capacitors ganged together, typical commercial units use a single "differential capacitor". This capacitor is a lot longer than a normal variable capacitor, and mandates using a much deeper cabinet. To fit it all into a manageable size, the maximum value of each half of the capacitor is not as high as the value typical used for each capacitor in the original T-Filter. As a result, the Differential-T matchboxes currently on the market do not cover as broad of an impedance matching range as the original T-Filter matchbox.

Differential T Pros and Cons

- Pros:
  - Easier to tune than regular T Match; and
  - At minimum SWR it is at the optimum settings.
- Cons:
  - Actual tuning range not as broad as regular T Match;
  - Differential capacitors are long, requiring a deep cabinet that is too large for the desk; and
  - Usually more expensive because of capacitor.

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ADVANTAGES OF THE DIFFERENTIAL-T MATCHBOX:

•Easier to tune to minimum SWR than the classical T-Matchbox

•At minimum SWR, the Differential-T is always operating at its most efficient tuning position.

DISADVANTAGES OF THE DIFFERENTIAL-T MATCHBOX:

•The impedance matching range of the Differential-T Matchbox is not as broad as that of a classical T-Matchbox.

•Their over-sized (long) differential variable capacitor mandates a very deep enclosure, which is often difficult to place on your desktop.

•There are very few Differential-T matchboxes on the market. As a result, these tuners are usually more expensive than a classical T-Matchbox. Examples:

•MFJ-986 (rated 1.5 kW)

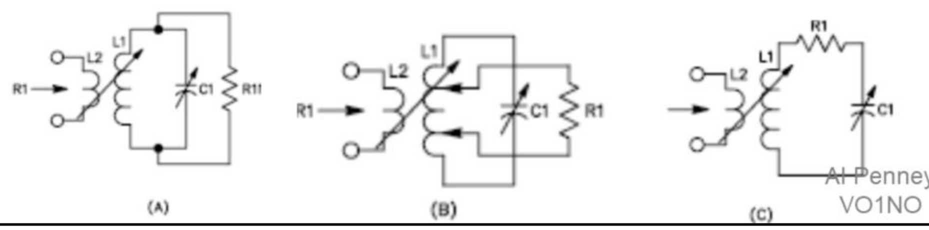
•Palstar AT-500 (rated 500w)

- **Palstar AT2KD (rated 2 kW)**

BOTTOM LINE: THE DIFFERENTIAL-T IS A GOOD CHOICE WHEN YOU ARE LOOKING FOR A HIGH POWER MATCHBOX WITH A TUNING RANGE SIMILAR TO THE RANGE OF THE BUILT-IN TUNERS FOUND IN TRANSCEIVERS.

Inductive Coupling Networks

- Also known as Link Coupling.
- Specialized form of matching network.
- Avoids direct connection between transmitter and antenna – prevents front end overload near other powerful transmitters.
- Usually used with balanced line.



Circuit arrangements for inductively coupled impedance-matching circuit. A and B use a parallel-tuned coupling tank; B is equivalent to A when the taps are at the ends of

L1. The series-tuned circuit at C is useful for very low values of load resistance, R1.

Link coupling offers many advantages over other types of systems where a direct connection between the transmitter and antenna is required, using a balanced type of transmission line. This is particularly

true at 3.5 MHz, where commercial broadcast stations often induce sufficient voltage to cause either rectification or front-end overload. Transceivers and receivers that show this tendency can usually

be cured by using only magnetic coupling between the transceiver and antenna system. There is no direct connection, and better isolation results, along with the inherent band-pass characteristics

of magnetically coupled tuned circuits. Although link coupling can be used with either single-ended or balanced antenna systems, its most common application is with balanced feed.

Link-Coupling is one of the oldest technologies used for matching an antenna to the transmitter. There are many different configurations of link-coupling, from very simple, to very elaborate circuits using differential capacitors. The efficiency is very good but covering multiple bands can be a challange, especially for the bandswitch.

This is probably the main reason that Link-Coupling is not found commercially available.

Series Matching

- Inductors, capacitors and load all **connected in series.**
- Impedances and reactances involved require high-voltage capacitors.
- Very high circulating currents require large wires and inductor core to avoid saturation.

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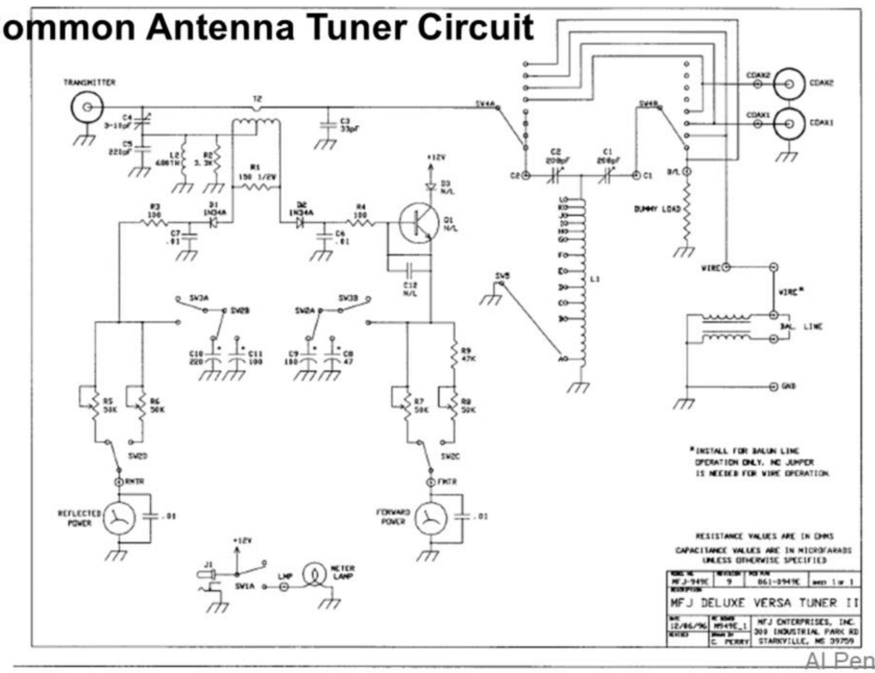
Common Antenna Tuner Circuit



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Matching Feedline to the Antenna

- There are several ways to **match the feedline** to the **feed point** of the antenna, including:
  - Attaching coax or twinlead **directly** to the dipole element;
  - The **Balun Match;**
  - The **Delta Match;**
  - The **T-Match;**
  - The **Gamma Match;**
  - The **Omega Match;**
  - The **Hairpin Match;**
  - The **Quarter Wavelength Transmission Line Match;**
  - The **Series Section Transmission Line Match;** and
  - The **Stub Match.**

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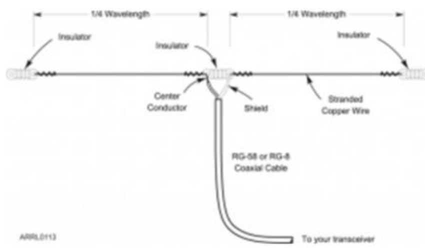
Connect Coax Directly to Antenna



- Pros
  - Simple;
  - Easy; and
  - Cheap.
- Cons
  - May be impedance mismatch;
  - Can get RF on the outside of the shield; and
  - Pattern of a gain antenna can be distorted.

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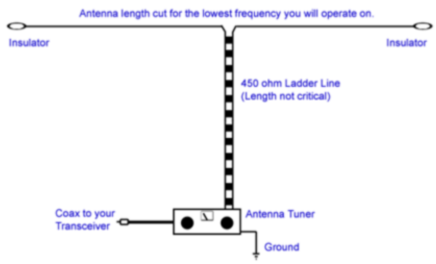
Connect Twinlead Directly to Antenna



- Pros:
  - Multiband;
  - Low loss transmission line.
- Cons:
  - Requires transmatch;
  - Twinlead requires careful routing.
  - Antenna pattern changes with frequency.
  - May get RF in shack.

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Using a Balun

- Different types and impedance ratios available to match a variety of antennas.
- 1:1 and 4:1 ferrite baluns common on HF.
- Coax baluns more common on VHF/UHF yagis, especially using a 4:1 coax balun to feed a 200 ohm T Match.

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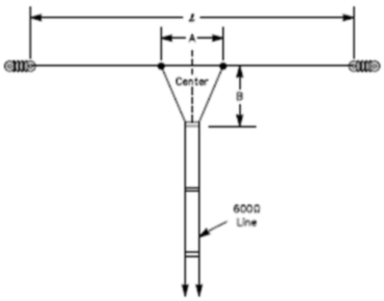
Delta Match



- Impedance at center of a dipole is too low for open twinlead.
- Impedance increases towards the ends of the dipole.
- Between the center and ends is a point where the dipole impedance equals the feedline impedance.

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A resonant antenna has properties similar to those of a tuned circuit. The impedance presented between any two points symmetrically placed with respect to the center of a λ/2 wavelength

antenna will depend on the distance between the points. The greater the separation, the higher the value of impedance, up to the limiting value that exists between the open ends of the antenna. This is also

suggested in the figure. The impedance ZA between terminals 1 and 2 is lower than the impedance ZB between terminals 3 and 4. Both impedances, however, are purely resistive if the

antenna is resonant.

This principle is used in the *delta matching system* shown in **the figure**. The center impedance of a λ/2 wavelength dipole is too low to be matched directly by any practical type of air-insulated parallel-conductor line. However, it is possible to find, between two points, a value of impedance that can be matched to such a line when a "fanned" section or delta is used to couple the line and antenna. The

antenna length l is that required for resonance. The ends of the delta or "Y" should be attached at points equidistant from the center of the antenna. When so connected, the terminating impedance for

the line will be resistive. Obviously, this technique is useful only when the Z0 of the chosen transmission line is higher than the feed-point impedance of the antenna. Based on experimental data for the case of a typical $\lambda/2$ wavelength antenna coupled to a 600-Ω line, the total distance, A, between the ends of the delta should be 0.120 ωαϖελενγτη for frequencies below 30 MHz, and 0.115 ωαϖελενγτη for frequencies above 30 MHz. The length of the delta, distance B, should be 0.150 ωαϖελενγτη. These values are based on a wavelength in air, and on the assumption that the center impedance of the antenna is approximately 70 οημσ. The dimensions will require modifications if the actual impedance is very much different.

The delta match can be used for matching the driven element of a directive array to a transmission line, but if the impedance of the element is low—as is frequently the case—the proper dimensions for

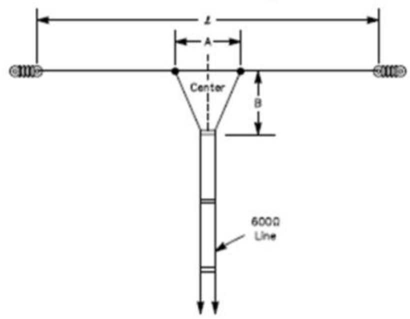
A and B must be found by experimentation.

The delta match is somewhat awkward to adjust when the proper dimensions are unknown, because both the length and width of the delta must be varied. An additional disadvantage is that there is

always some radiation from the delta. This is because the conductor spacing does not meet the requirement for negligible radiation: The spacing should be very small in comparison with the wavelength.

Delta Match

- Assuming 70 ohm center impedance:
  - A = 0.12 λ below 30 MHz and 0.115 λ above
  - B = 0.15 λ for all frequencies
- Note that impedance in a yagi is lower than 70 ohms, so these dimensions will change.



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Delta Match Pros and Cons

- Pros:
  - Easy way to match open line to dipole;
  - Adjustable to ensure proper match: and
  - Can also connect coax to dipole without need to have insulator in the center (using 4:1 balun).
- Cons:
  - Matching section radiates; and
  - Cannot remove reactive impedance elements so a stub may be required.

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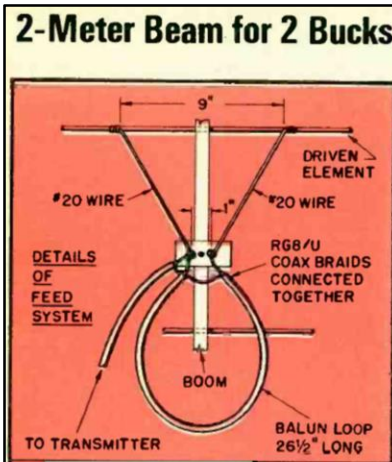
Delta match

The delta match for of Yagi matching is one of the more straightforward solutions. It involves fanning out the ends of the balanced feeder to join the continuous radiating antenna driven element at a point to provide the required match.

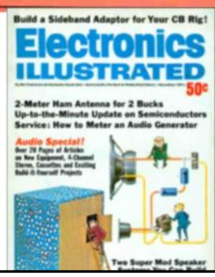
Both the side length and point of connection need to be adjusted to optimise the match.

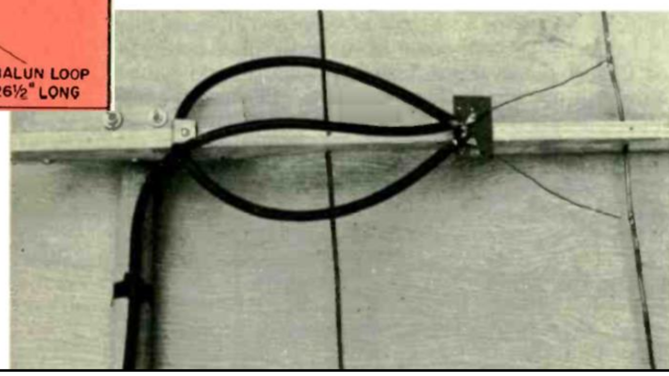
One of the drawbacks for using the Delta match for providing Yagi impedance matching is that it is unable to provide any removal of reactive impedance elements. As a result a stub may be used.

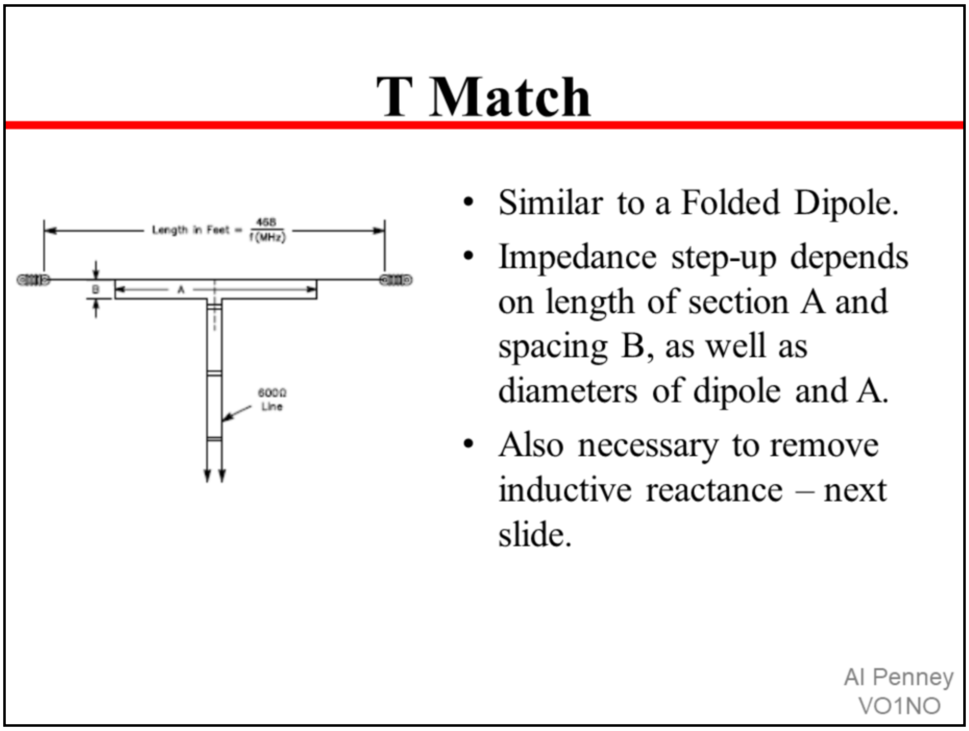
My First 2M Antenna!



Electronics Illustrated Magazine
November 1971







The T Match

The current flowing at the input terminals of the T match consists of the normal antenna current divided between the radiator and the T conductors in a way that depends on their relative diameters and

the spacing between them, with a superimposed transmission-line current flowing in each half of the T and its associated section of the antenna. See **Figure**. Each such T conductor and the associated antenna conductor can be looked upon as a section of transmission line shorted at the end. Because it is shorter than λ/4 wl it has inductive reactance. As a consequence, if the antenna itself is exactly resonant at the operating frequency, the input impedance of the T will show inductive reactance as well as resistance. The reactance must be tuned out if a good match to the transmission line is to be obtained. This can be done either by shortening the antenna to obtain a value of capacitive reactance that will reflect through the matching system to cancel the inductive reactance at the input terminals, or by inserting a capacitance of the proper value in series at the input terminals as shown in next slide. Theoretical analyses have shown that the part of the impedance step-up arising from the spacing

and ratio of conductor diameters is approximately the same as given for a folded dipole. The actual impedance ratio is, however, considerably modified by the length A of the matching section

(Fig 9). The trends can be stated as follows:

1) The input impedance increases as the distance A is made larger, but not indefinitely. In general there is a distance A that will give a maximum value of input impedance, after which further

increase in A will cause the impedance to decrease.

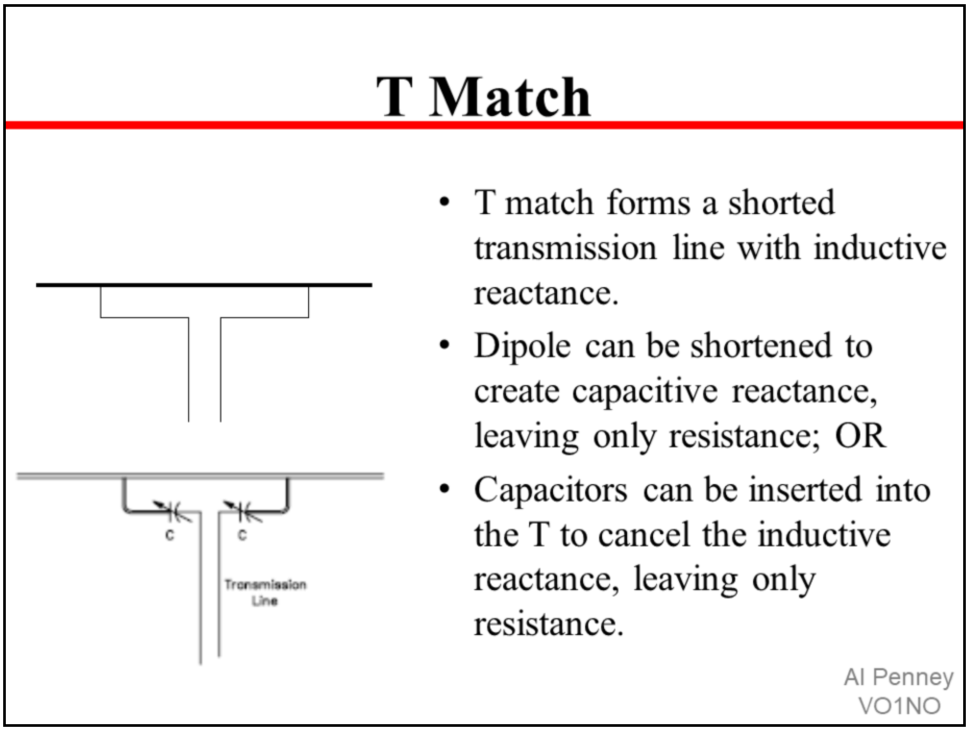
2) The distance A at which the input impedance reaches a maximum is smaller as the ratio of diameters d2/d1 is made larger, and becomes smaller as the spacing between the conductors

is increased.

3) The maximum impedance values occur in the region where A is 40% to 60% of the antenna length in the average case.

4) Higher values of input impedance can be realized when the antenna is shortened to cancel the inductive reactance of the matching section. The T match has become popular for transforming

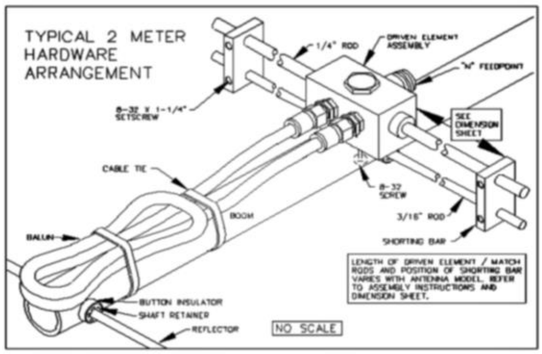
the balanced feed-point impedance of a VHF or UHF Yagi up to 200 οημ. From that impedance a 4:1 balun is used to transform down to the unbalanced 50 οημ level for the coax cable feeding the Yagi.



Each such T conductor and the associated antenna conductor can be looked upon as a section of transmission line shorted at the end. Because it is shorter than $\lambda/4$ wavelength it has inductive reactance. As a consequence, if the antenna itself is exactly resonant at the operating frequency, the input impedance of the T will show inductive reactance as well as resistance. The reactance must be tuned out if a good match to the transmission line is to be obtained. This can be done either by shortening the antenna to obtain a value of capacitive reactance that will reflect through the matching system to cancel the inductive reactance at the input terminals, or by inserting a capacitance of the proper value in series at the input terminals as shown in **the figure**

T Match

- T Match often designed for 200 Ohm impedance and fed with 50 ohm coax through a 4:1 Coax Balun.



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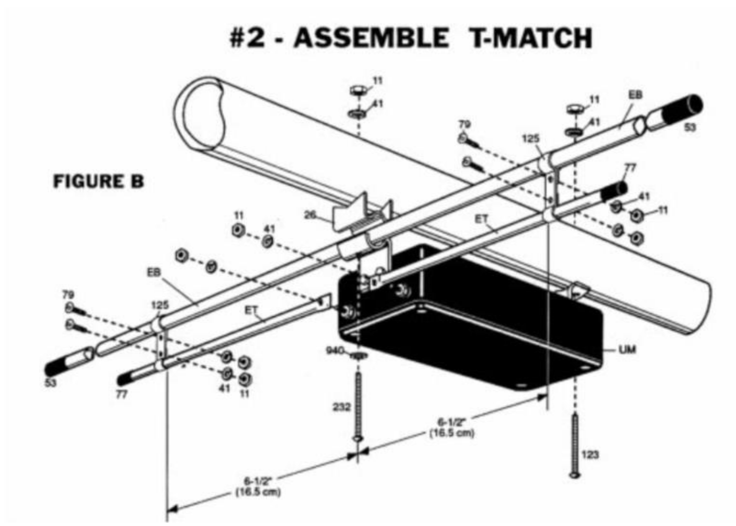
T Match Cushcraft 17B2 Yagi



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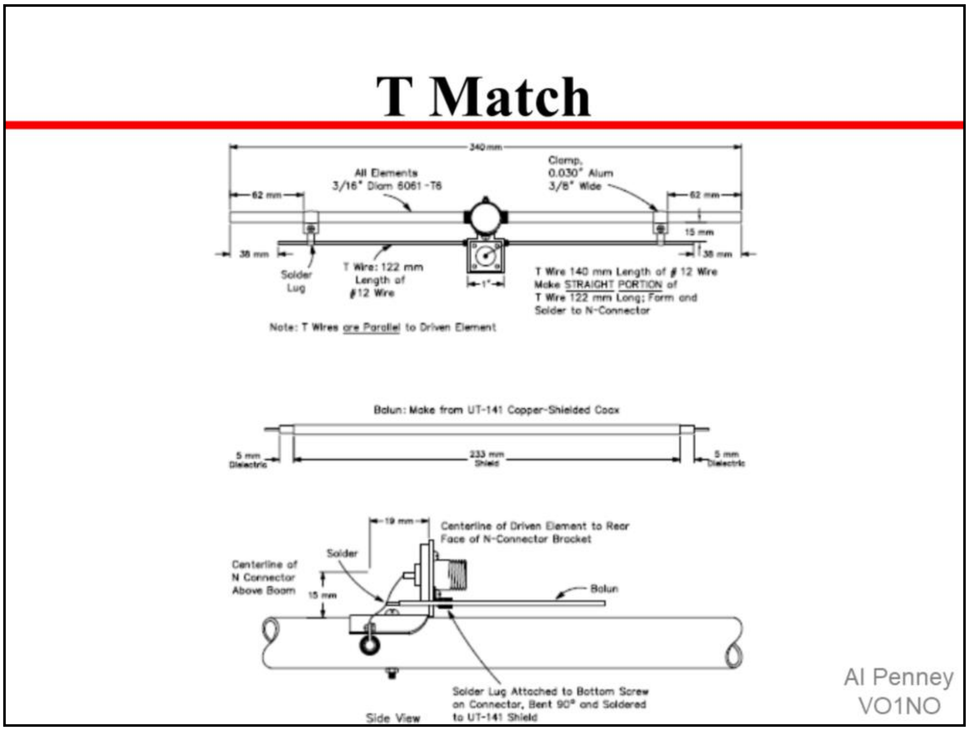
4:1 coax balun can be hidden inside a box mounted at the feedpoint. This makes for a neat installation that is protected against the elements.



The T-Match provides an impedance matching function, as well as providing a proper de coupling, balun function. The use of a half wave balun keeps the RF energy going to the driven element and nowhere else! The standard T-Match originally had series capacitors going to each side of the driven element. The improved "modified" T-Match does away with the capacitors. In its place, we adjust the length of the driven element to add some capacitive or inductive reactance to cancel out any reactance at the feed point. The antenna feed point consists of the following components:

1. Driven element
2. T-Match sliding bars
3. T-Match wires or rods, (This wire is much smaller in diameter than the driven element.)
4. Half wave balun made from semi rigid Teflon coax cable

T Match Pros and Cons

- Pros:
  - When used with VHF/UHF yagis, it provides a convenient way to match the driven element to a 4:1 coax balun.
  - Balanced, so yagi pattern is not distorted.
- Cons:
  - Single band.

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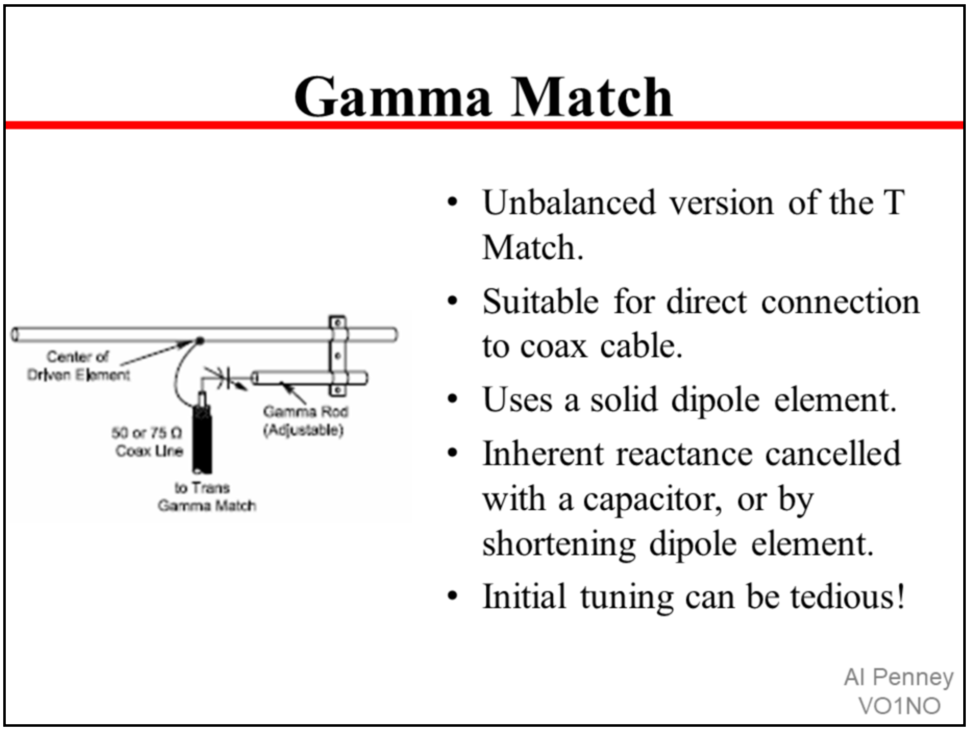


The gamma-match arrangement shown in Fig 10B is an unbalanced version of the T, suitable for use directly with coaxial lines. Except for the matching section being connected between the center and

one side of the antenna, the remarks above about the behavior of the T apply equally well. The inherent reactance of the matching section can be canceled either by shortening the antenna appropriately or by using the resonant length and installing a capacitor C, as shown in Fig 10B.

For a number of years the gamma match has been widely used for matching coaxial cable to all metal parasitic beams. Because it is well suited to “plumber’s delight” construction, where all the metal

parts are electrically and mechanically connected, it has become quite popular for amateur arrays.

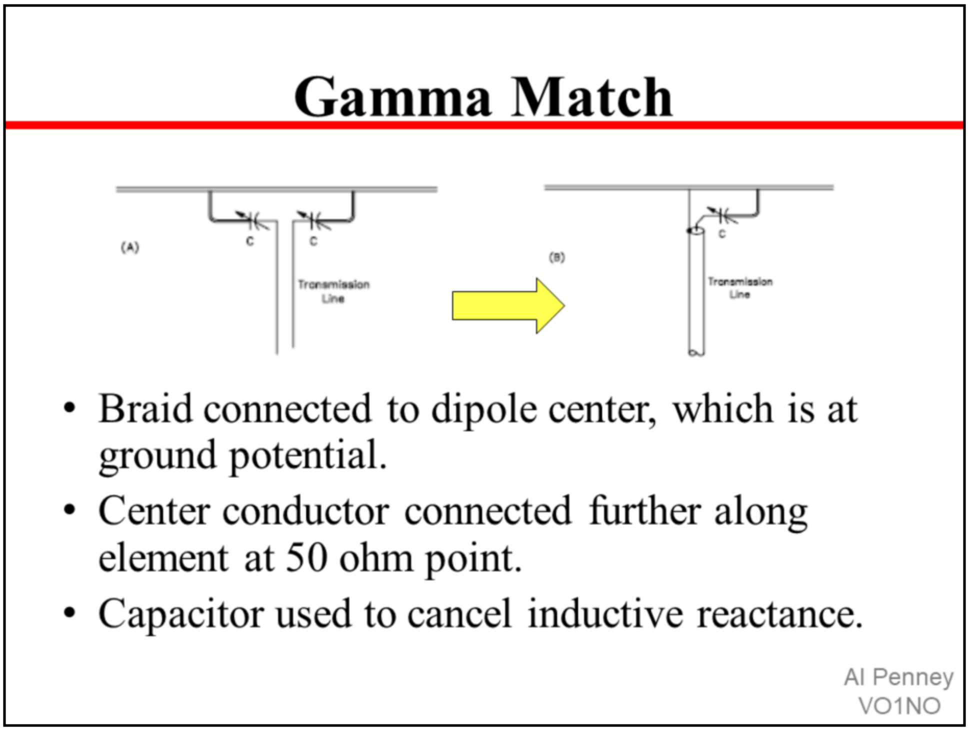
Because of the many variable factors—driven-element length, gamma rod length, rod diameter, spacing between rod and driven element, and value of series capacitors—a number of combinations

will provide the desired match. The task of finding a proper combination can be a tedious one, as the settings are interrelated. A few “rules of thumb” have evolved that provide a starting point for the

various factors. For matching a multielement array made of aluminum tubing to 50-οηµ line, the length of the rod should be 0.04 to 0.05 ωαϖελενγτη, its diameter 1/2 to

1/3 that of the driven element, and its spacing

(center-to-center from the driven element), approximately 0.007 ωαϖελενγτη. The capacitance value should be approximately 7 pF per meter of wavelength. This translates to about 140 pF for 20-meter operation. The exact gamma dimensions and value for the capacitor will depend on the radiation resistance of the driven element, and whether or not it is resonant. These starting-point dimensions are for an array having a feed-point impedance of about 25 οημ, with the driven element shortened approximately 3% from resonance.



Gamma match

The gamma match is often used for providing Yagi impedance matching. It is relatively simple to implement.

As seen in the diagram, the outer of the coax feeder is connected to the centre of the driven element of the Yagi antenna where the voltage is zero. As a result of the fact that the voltage is zero, the driven element may also be connected directly to a metal boom at this point without any loss of performance.

The inner conductor of the coax is then taken to a point further out on the driven element - it is taken to a tap point to provide the correct match. Any inductance is tuned out using the series capacitor.

When adjusting the RF antenna design, both the variable capacitor and the point at which the arm contacts the driven element are adjusted. Once a value has been ascertained for the variable capacitor, its value can be measured and a fixed component inserted if required.

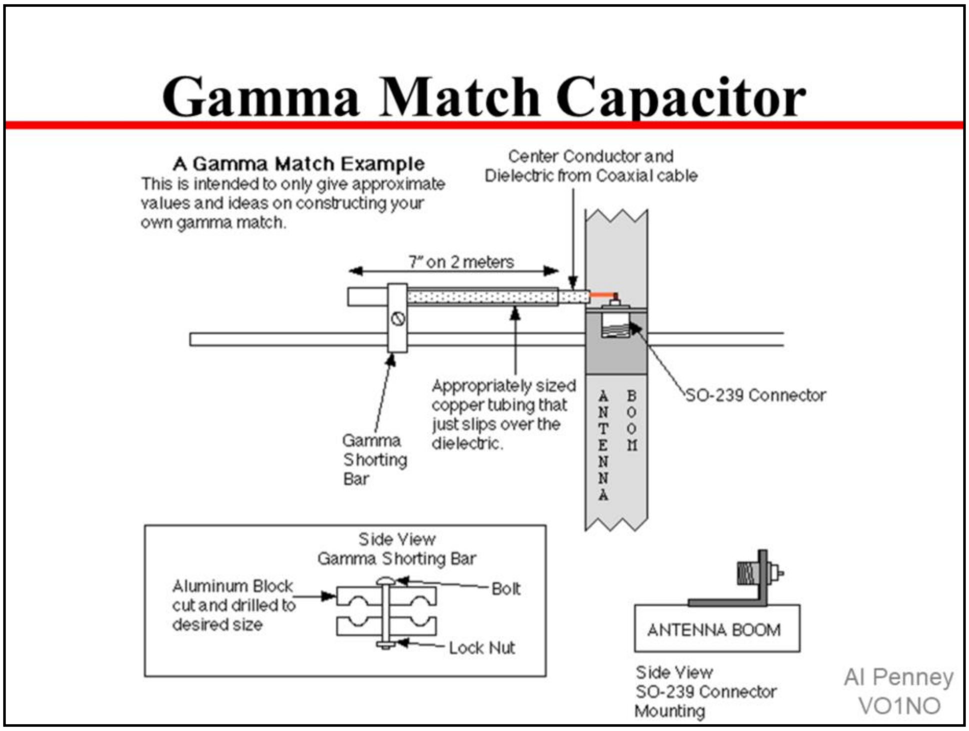
As **Fig. 26** shows, we make our feedline connection to a rod or double rod that parallels the element for a distance, at which point we connect a shorting strap to the rod and the element. The gamma match permits a direct conversion of the antenna feedpoint impedance to 50 Ohms. In the gamma match, the coax braid is connected directly to the electrical center of the element.

The Tee double rod system is most effective in raising the impedance to a much higher value than 50 Ohms. 200 Ohms is a convenient value, since a 4:1 balun then provides a fine 50-Ohm match, while performing the task of attenuating common mode currents at the same time.

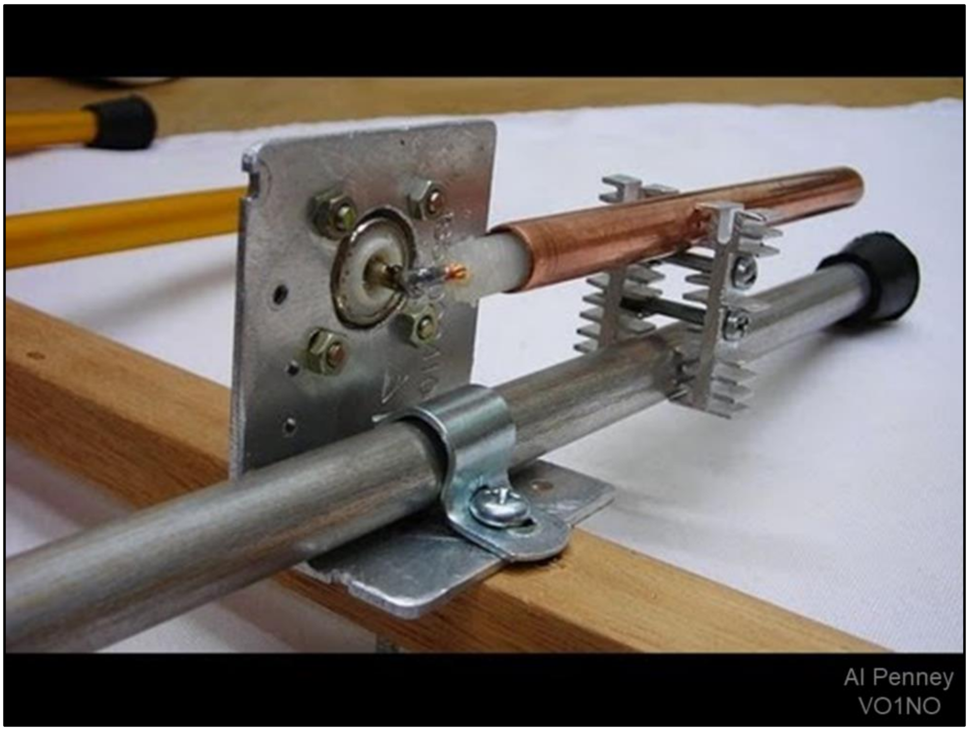
In VHF beams, many builders have noted and measured some pattern distortion with the gamma match. Hence, the Tee has almost become a standard for beams having other than a 50-Ohm feedpoint impedance. HF beam makers have more confidence in the gamma match, whose parts are a relatively smaller proportion of the driver structure than they are at VHF and UHF. However, the Tee is inherently balanced.

The dimensions of both matching systems depend upon the relative sizes of and the spacing between the element and the rod material. Calculations are best done with software specifically designed for the task. YO (Yagi Optimizer), by K6STI, for example, has such auxiliary software, and independent utility calculation programs are available as freeware from many sources.

Note that the sketches in **Fig. 26** did not include a series capacitor at the feedpoint and rod junction, a standard feature in many basic system sketches. In most instances, variable capacitors do not withstand well the rigors of weather. Moreover, the reactance of the system can be set (or nulled, depending upon one's perspective) by adjusting the length of the driven element.



Instead of a discrete fixed or variable capacitor, most manufacturers use a polyethylene dielectric in a metal tube that serves as the gamma arm, with the center conductor attached to the center conductor in the polyethylene. This gives stable performance and is easily waterproofed.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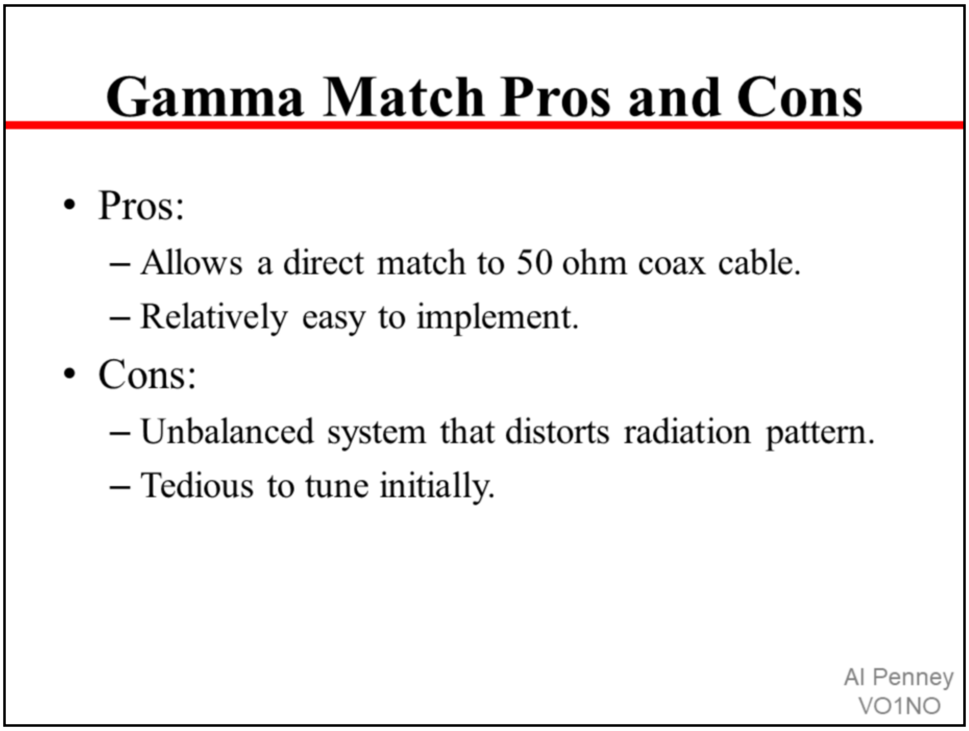

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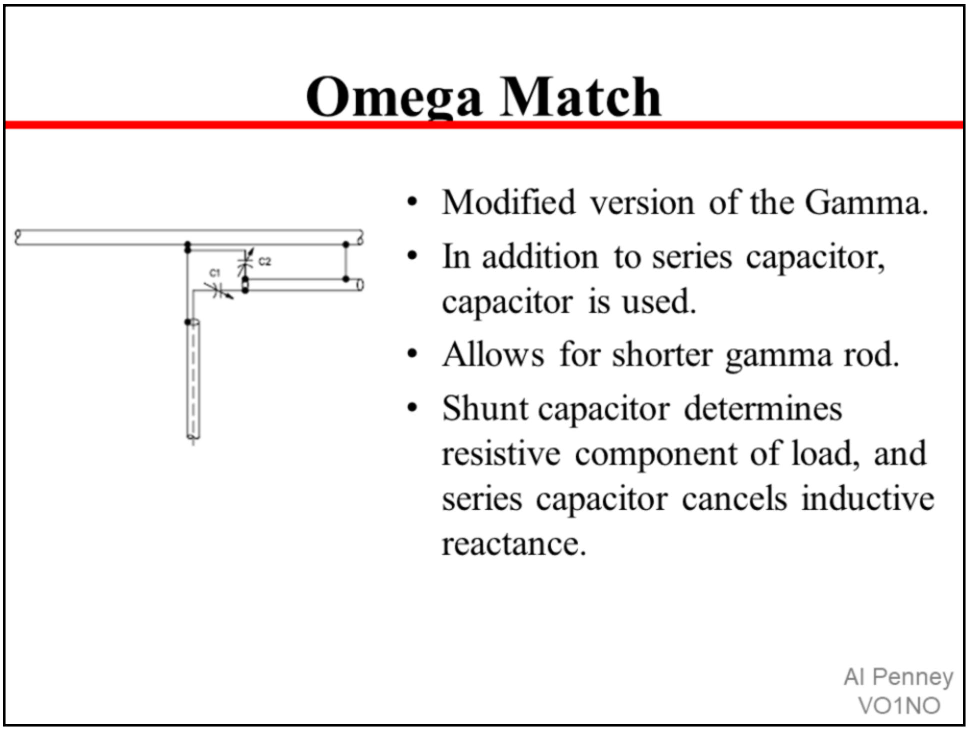
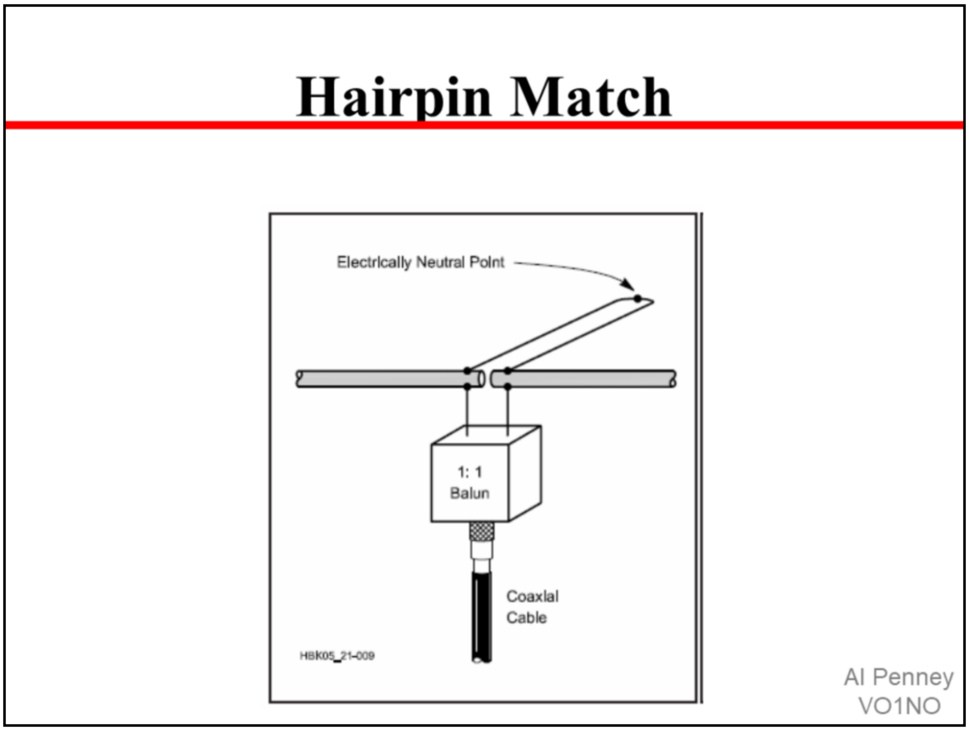


The Omega Match

The omega match is a slightly modified form of the gamma match. In addition to the series capacitor, a shunt capacitor is used to aid in canceling a portion of the inductive reactance introduced by the

gamma section. This is shown in **Fig 12**. C1 is the usual series capacitor. The addition of C2 makes it possible to use a shorter gamma rod, or makes it easier to obtain the desired match when the driven

element is resonant. During adjustment, C2 will serve primarily to determine the resistive component of the load as seen by the coax line, and C1 serves to cancel any reactance.



The usual form of the *hairpin match* is shown in **Fig 13**. Basically, the hairpin is a form of an Lmatching network. Because it is somewhat easier to adjust for the desired terminating impedance than

the gamma match, it is preferred by many amateurs. Its disadvantages, compared with the gamma, are that it must be fed with a balanced line (a balun may be used with a coax feeder, as shown in Fig 13— see section later in this chapter about baluns), and the driven element must be split at the center. This latter requirement complicates the mechanical mounting arrangement for the element, by ruling out “plumber’s delight” construction.

As indicated in Fig 13, the center point of the hairpin is electrically neutral. As such, it may be grounded or connected to the remainder of the antenna structure. The hairpin itself is usually secured

by attaching this neutral point to the boom of the antenna array. The *beta match* is electrically identical to the hairpin match, the difference being in the mechanical construction of the matching section. With

the beta match, the conductors of the matching section straddle the boom, one conductor being located on either side, and the electrically neutral point consists of a sliding or adjustable shorting clamp placed around the boom and the two matching-section conductors.

The capacitive portion of the L-network circuit is produced by slightly shortening the antenna driven element. For a given frequency the impedance of a shortened $\lambda/2$ wavelength element appears as the antenna resistance and a capacitance in series. The inductive portion of the resonant circuit is a “hairpin” of heavy wire or small tubing which is connected across the driven element center terminals.

Hairpin Match

- Usually used in monoband Yagis where radiation resistance is less than 50 ohms.
- Shortening driven element gives it capacitive reactance.
- Careful selection of the hairpin can cancel reactance and increase radiation resistance to 50 ohms for a perfect match.
- Hairpin is sometimes replaced with a coil inductor.
- Also used in short vertical antennas.
- Beta Match is similar to Hairpin Match.

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http://ehpes.com/n6mw/HairpinArticleFix.pdf

Multi-element higher frequency Yagis are sometimes hairpin matched to provide a balanced 200 ohm antenna impedance that is further transformed with a 4:1 λ/2 coaxial "step up balun." This provides a final 50 ohm unbalanced antenna impedance that is a good match to a 50 ohm (unbalanced) coaxial cable

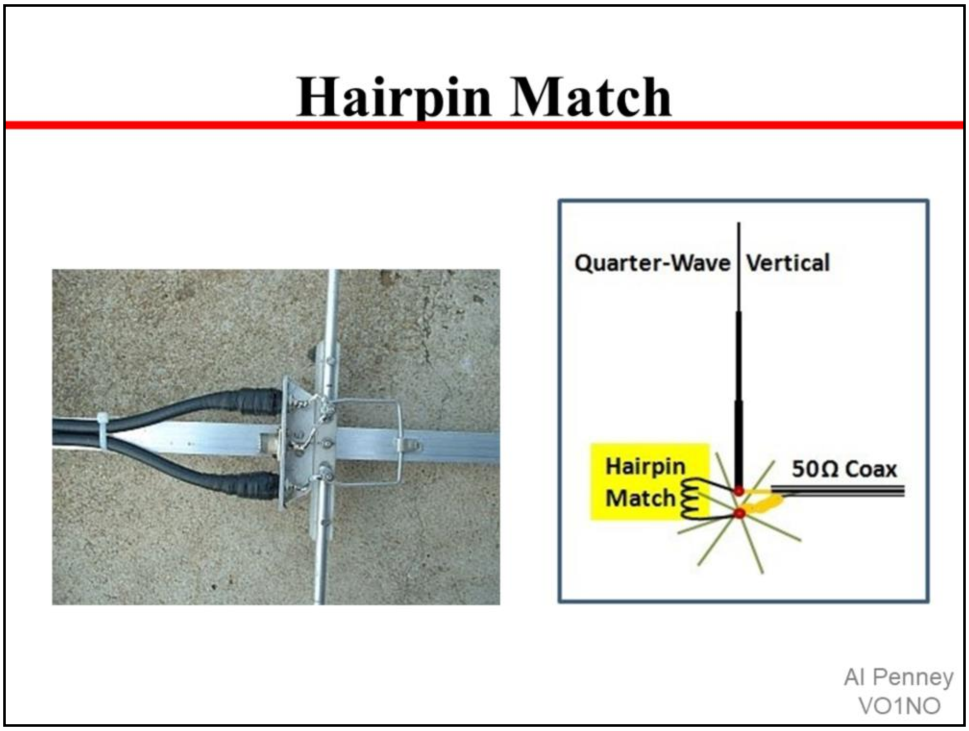
"Hairpin" matching is nothing more than adding an inductance directly across the feed point of the antenna. The inductance may be a simple wire coil or an extended U-shaped wire or rods. This U-shape is of course the reason for the name hairpin and many applications of this matching style use that form of inductance. Sometimes this method is called "shunt" matching, but beware because other forms of matching also use the word shunt although some are rather different.

How Hairpin Impedance Transformation Works

If an antenna has an unmatched complex impedance of Za (Ra+jXa), adding an inductance across the antenna input (sometimes loosely referred to as a "shunt") gives an equivalent circuit as shown in the figure, where Ro+jXo is the resulting transformed complex impedance output. The object is then to select the added inductance value to cause Ro to be close to the feed line characteristic impedance (often 50 ohms) with the net output reactance Xo near zero.

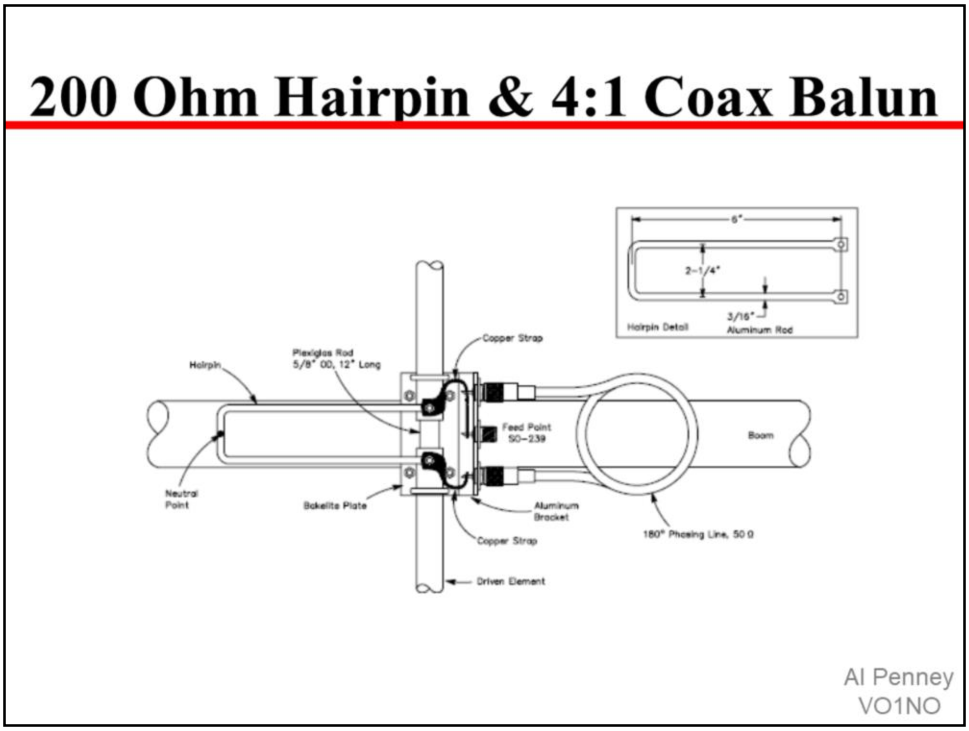
By adding just one inductance of our choice, you want to both get Ro to be near the feed line impedance and make Xo be near zero. You might wonder if it is possible to do both with just one additional component. The answer is generally no. However, if Za has one of a range of appropriate values, it will work. The equation that describes this parallel circuit is not difficult, although some may be put off by the need to deal with complex numbers. This is expanded on in the sidebar. The upshot is if Ra is less than Ro, there is always a capacitive Xa (negative) and a hairpin inductance XL that will produce a perfect match. or Za Ro +j Xo Hairpin (XL) At first blush you might think if I just short out the antenna input with a bit of wire, won't much of the current coming up the feed line just go through the added wire and not the antenna - and that won't be so good. It turns out that the magnitude of the current in the hairpin is generally comparable with that in Za, and it can be larger than the feed line current. This is because the hairpin and Za currents are far from being in phase. But it turns out that this delicate balance makes the hairpin matched antenna appear as an impedance of Ro, even though there are (mostly) non-radiating currents in the hairpin .

The first requirement for use of the hairpin is that the unmatched antenna resistive part, Ra, must be significantly less than the desired feed line impedance, Ro. Second, the unmatched antenna reactive part, Xa, must be capacitive (negative) and also must be near the required value calculated in the sidebar. Xa can often be adjusted by modest changes in the length of the driven element or by addition of loading. Once the appropriate Xa is available, there is a hairpin inductive reactance, XL, that can provide a perfect match



Multi-element higher frequency Yagis are sometimes hairpin matched to provide a balanced 200 ohm antenna impedance that is further transformed with a 4:1 λ/2 coaxial “step up balun.” This provides a final 50 ohm unbalanced antenna impedance that is a good match to a 50 ohm (unbalanced) coaxial cable.

Short verticals sometimes use a hairpin to match their low impedance to 50 ohm coax cable.



Refer to **Figure** for driven element and hairpin match details. A bracket made from a piece of aluminum is used to mount the three SO-239 connectors

to the driven element plate. A 4:1 transmission line balun connects the two element halves, transformingthe 200-οημ resistance at the hairpin match

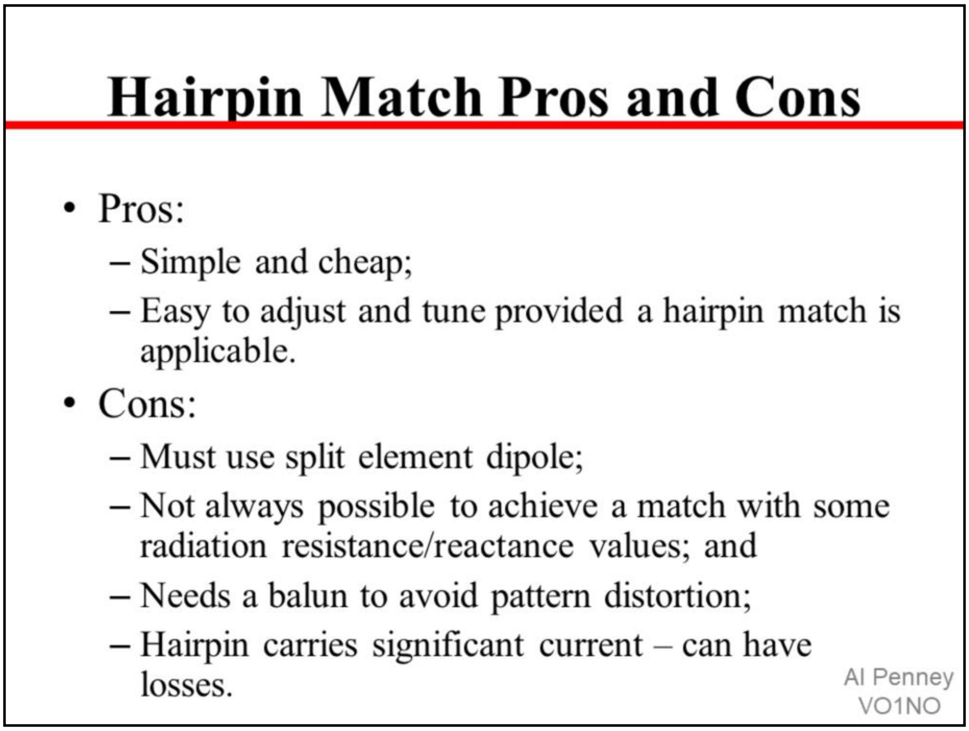
to 50 οημ at the center connector. Note that the electrical length of the balun is λ/2 wl, but the physical length will be shorter due to the velocity factor of

the particular coaxial cable used. The hairpin is connected directly across the element halves. The exact center of the hairpin is electrically neutral and

should be fastened to the boom. This has the advantage of placing the driven element at dc ground potential. The hairpin match requires no adjustment

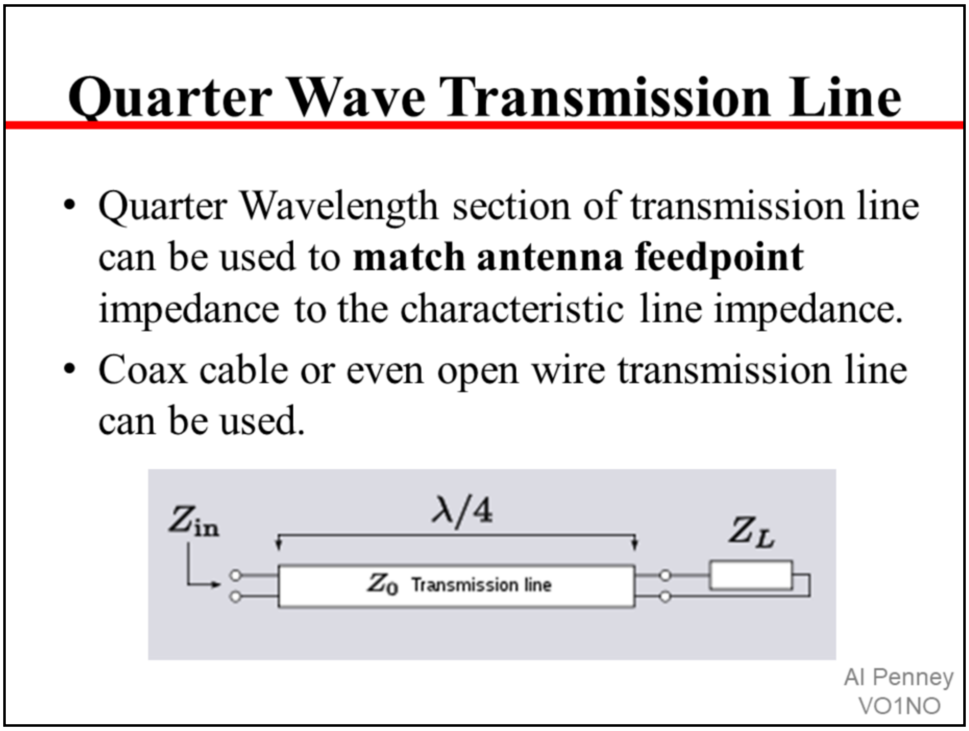
as such. However, you may have to change the length of the driven element slightly to obtain the best match in your preferred portion of the band. Changing the

driven-element length will not adversely affect antenna performance. *Do not adjust the lengths or spacings of the other elements—they are optimized already*.



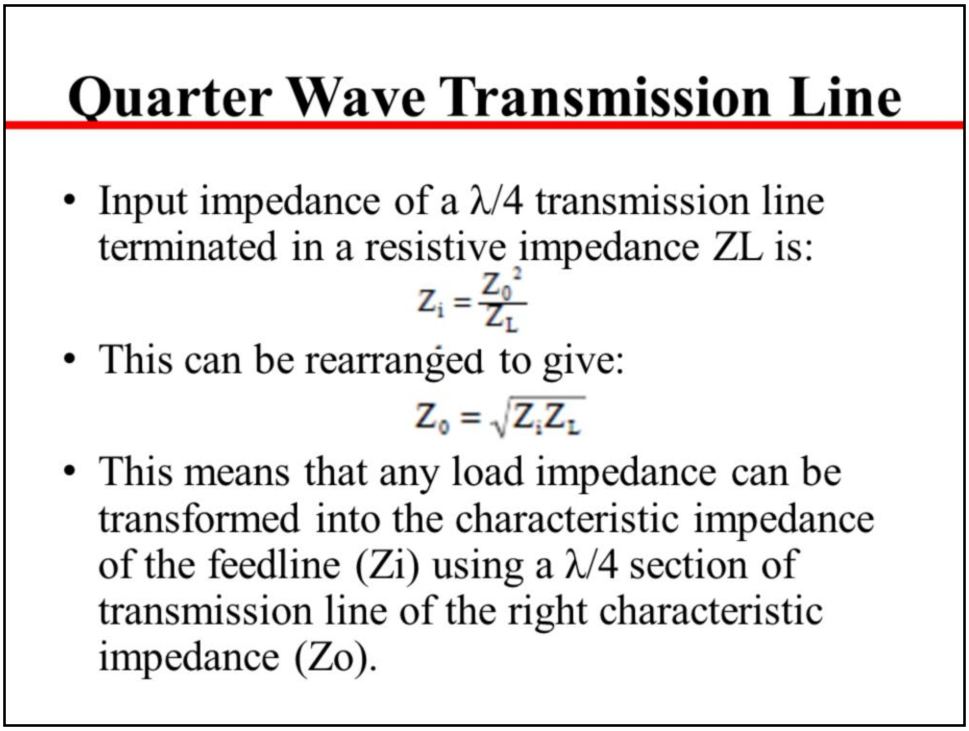
Drawbacks of Hairpin Match

The most glaring weakness of hairpin match, for Yagis, is the mechanical complication of the need to split the driven element into two insulated halves. The good news is that it is then easy to measure the unmatched impedance with a modest antenna analyzer. While the Ra range for matching is pretty wide, the required Xa range is a bit less forgiving, with variations of about 25% from the optimal value leading to best case SWR above 1.5:1. Adjustment of the driven element to get the optimal Xa may be difficult under those conditions where the length is largely fixed. Furthermore, the value of Ra also depends (but usually to a lesser degree) on the driven element length so it all can be a bit of a moving target. The hairpin match will leave you with a balanced antenna. With a coaxial cable feed line, which is unbalanced, conventional wisdom is that a common mode choke or another style 1:1 balun will probably needed to limit potential distortion of the pattern from common mode currents. There will be significant current flowing in the hairpin inductance so the size of the wire and the quality of the connections need to be considered to assure there are no significant ohmic losses. Some antenna analyzers do not provide the sign of Xa. This can be resolved by looking at the behavior of Xa with frequency as compared to the model, if your model is reasonably close to reality. It is also possible to add a short length of coax and find the transformed Za change compared to predictions of a code like TLW for Ra and +/-Xa. In another regard, this change in impedance at the end of different length feed lines can be confusing when measuring poorly matched antennas remotely since the impedance at the antenna can be quite different from that at the far end of the feed line, unless you happen to have an exact multiple of a half wave of cable. However, the SWR does not depend significantly on cable length so looking for an SWR minimum is still a useful strategy.



A **quarter-wave impedance transformer**, often written as **λ/4 impedance transformer**, is a transmission line or waveguide used in electrical engineering of length one-quarter wavelength (λ), terminated with some known impedance. It presents at its input the dual of the impedance with which it is terminated.

It is a similar concept to a stub; but, whereas a stub is terminated in a short (or open) circuit and the length is chosen so as to produce the required impedance, the λ/4 transformer is the other way around; it is a pre-determined length and the termination is designed to produce the required impedance.



The impedance-transforming properties of a λ/4 wavelength transmission line can be used to good advantage for matching the feed-point impedance of an antenna to the characteristic impedance of the line. As described in Chapter 24 of the ARRL Antenna Book, the input impedance of a λ/4 wavelength line terminated in a resistive impedance ZL is

Zi = ZO^2 / ZL

where

Zi = the impedance at the input end of the line

Z0 = the characteristic impedance of the line

ZL = the impedance at the load end of the line

Rearranging this equation gives

Zo = Square Root (Zi x ZL)

This means that any value of load impedance ZL can be transformed into any desired value of impedance Zi at the input terminals of a λ/4 line, provided the line can be constructed to have a characteristic

impedance Z0 equal to the square root of the product of the other two impedances.

The factor that limits the range of impedances that can be matched by this method is the range of values for Z0 that

is physically realizable. The latter range is approximately 50 to 600 Ω. Practically any type of line can be used for the matching section, including both air-insulated and solid-dielectric lines.

The λ/4 wavelength transformer may be adjusted to resonance before being connected to the antenna by short-circuiting one end and coupling that end inductively to a dip meter. The length of the short-circuiting conductor lowers the frequency slightly, but this can be compensated for by adding half the length of the shorting bar to each conductor after resonating, measuring the shorting-bar length between the centers of the conductors.

Quarter-Wave Matching Section

A quarter-wavelength line exhibits a transforming property, converting low-resistance loads to higher ones, and capacitive reactances to inductive. The best way to understand and use this feature is with the *Smith chart* or asimilar tool. One important application of this property is to transform a load impedance to a higher (or lower) main transmission line impedance. For instance, suppose an antenna for 10 MHz has a feedpoint impedance of 100 ∧ resistive. It's not a very good match for 52-∧ coaxial cable, but a λ/4 wavelength section of 75-∧ RG-59 or RG-11 coaxial cable for television systems will transform the antenna feedpoint impedance from 100 to 52 ∧.

Impedances for the λ/4 wavelength transformer are calculated (for resistive loads) as follows:

ZL x Zi = Zo^2

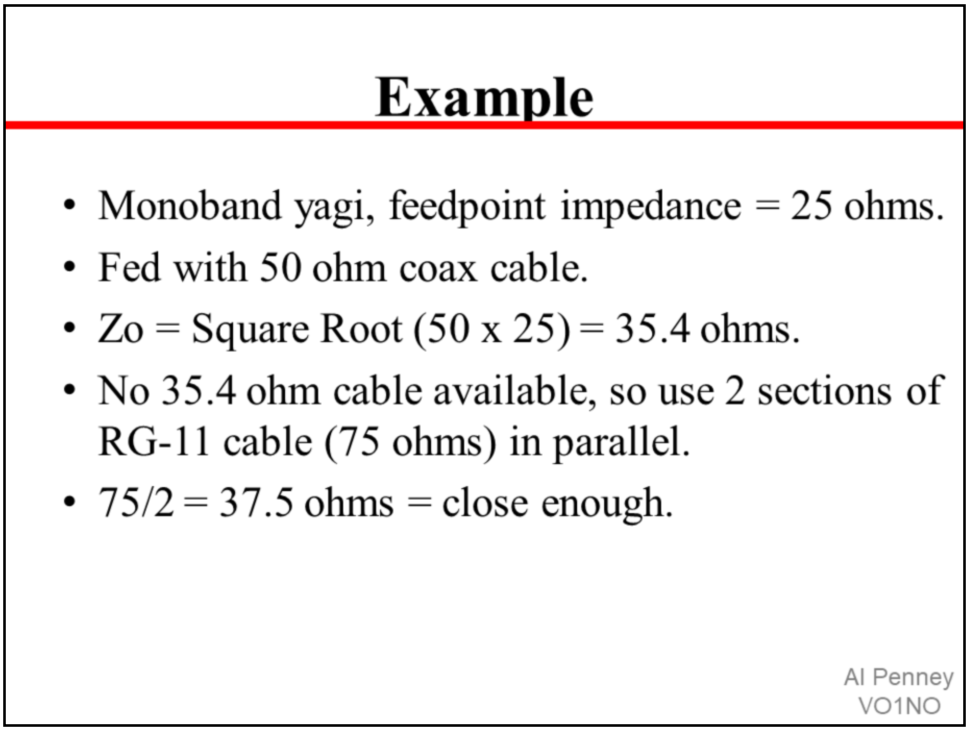
where *ZL* = load or antenna feedpoint impedance

*Z*o = characteristic impedance of λ/4 wavelength section

Zi = characteristic impedance of system transmission line

The λ/4 transmission line transformer is sometimes called a *Q-section matching network.*

The λ/4 wavelength transformer may be adjusted to resonance before being connected to the antenna by short-circuiting one end and coupling that end inductively to a dip meter. The length of the short-circuiting conductor lowers the frequency slightly, but this can be compensated for by adding half the length of the shorting bar to each conductor after resonating, measuring the shorting-bar length between the centers of the conductors.



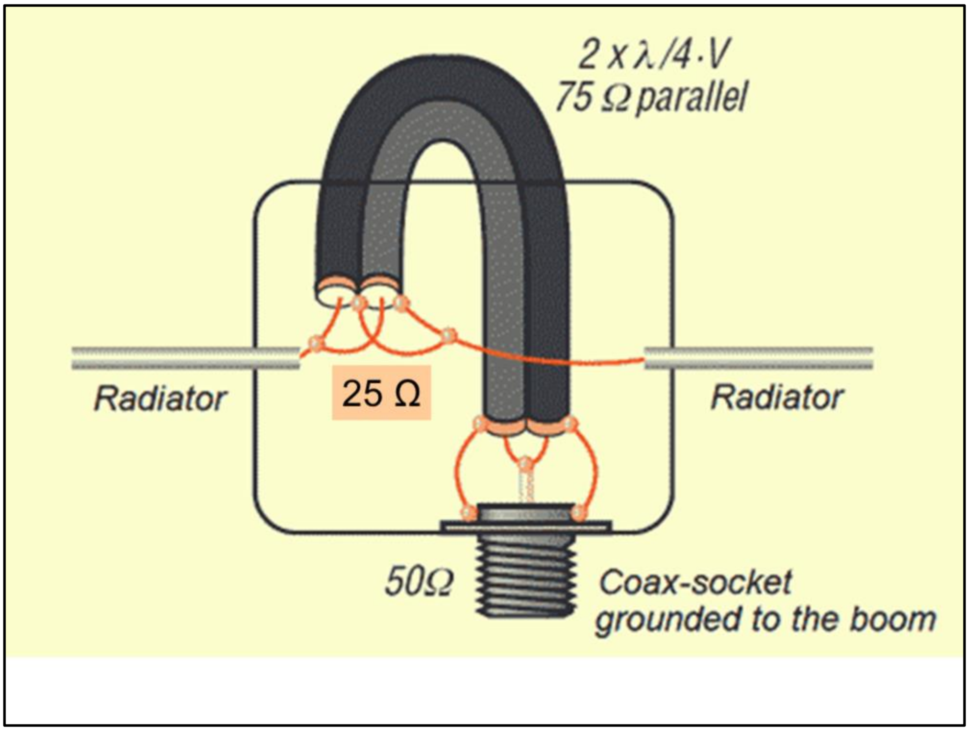
Yagi Driven Elements

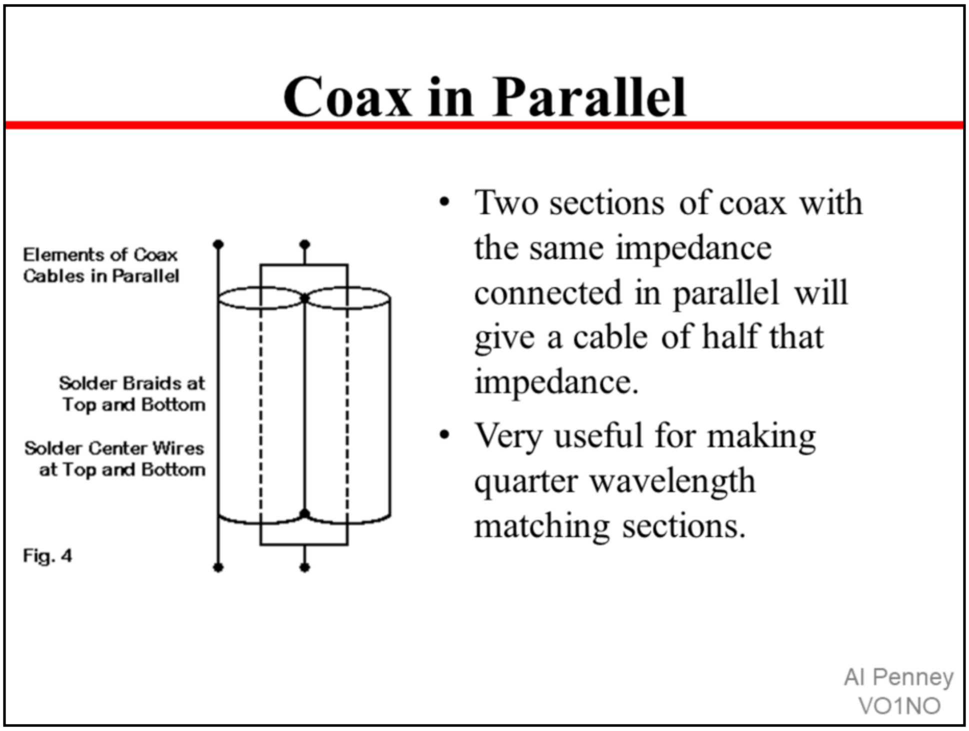
Another application for the λ/4 “linear transformer” is in matching the low antenna impedance encountered in close-spaced, monoband Yagi arrays to a 50-Ω transmission line. The impedances at the

antenna feed point for typical Yagis range from about 8 to 30 Ω. Let’s assume that the feed-point impedance is 25 Ω. A matching section having

Zo = Square Root (50 x 25) = 35.4 ohms

is needed. Since there is no commercially available cable with a Z0 of 35.4 οημσ, a pair of λ/4 wavelength long 75-οημ RG-11 coax cables connected in parallel will have a net Z0 of 75/2 = 37.5 οημ, close enough for practical purposes.


2 x λ/4·V
75 Ω parallel
Radiator
25 Ω
Radiator
50Ω
Coax-socket
grounded to the boom



Parallel-Connected Coaxial Cables

The use of parallel-connected coax cables of the same length normally involves cases where we need a cable of 1/2 the Zo of a single line of the same cable. Obviously, this opens up the door to

having cables of some odd but useful characteristic impedances. Two 50s in parallel give us a 25-Ohm cable--a value not commercially available. Two 70s give us 35 Ohms, a value very useful

in 1/4-wavelength matching sections. Two 93s give us 46.5 Ohms, again useful in matching sections. Two 125s give us 62.5 Ohms, once more useful in matching sections or as a main line.

Fig. 4 provides data on how to hook up cables to achieve a parallel connection. The two center wires are soldered together at both ends. As well, the two braids are soldered together at both

ends. If you are using RG-58 (50-Ohms) or RG-59 (70 Ohms), the two cables will--with proper preparation of the ends--fit inside the barrel of a standard coax male connector. You can seal the

spaces with a non-reactive, non- conductive sealant--after everything is properly soldered. However, consider using direct splices between the parallel coax section and any main cable to

which you connect it. If you work in the VHF/UHF region, especially with small cables, you may be able to avoid the introduction of small reactances by using a carefully constructed direct cable

splice. You can always strengthen the junction by adding some surface stiffening between the linked sections. Heat shrink tubing with a little epoxy inside (but not on the junctions of wires) can

add strength to spliced thin lines.

Paralleled cables will provide roughly twice the power handling capacity of a single cable in the HF region. The surface area of the center wires is doubled. However, be sure to calculate your power

limit for the new impedance and the current peaks that will be present at this impedance. One of the main uses of parallel-connected coax cables is to form a 1/4- wavelength matching

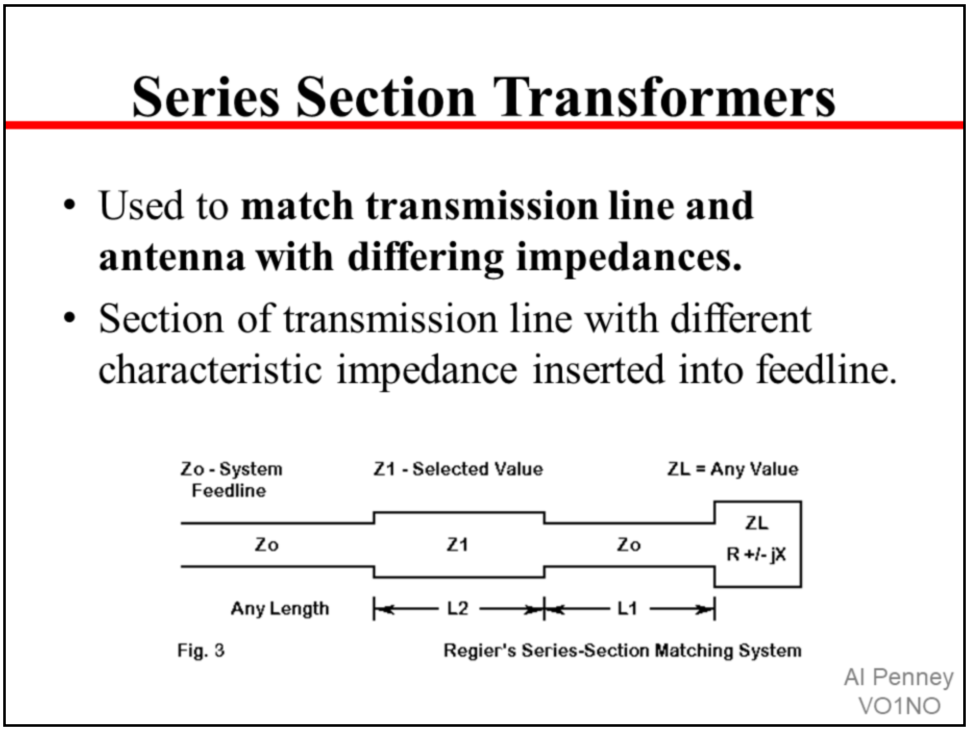
section with an impedance that is correct for the desired match. The ideal matching section impedance is the square root of the product of the input and output impedances. Here are a few

examples of common matching section applications.

Antenna Impedance	**Line**	**Zo**	**Match Section Zo**
2-element Quad	**115**	**50**	**76 (use 70 Ohm cable)**
3-element Yagi	**25**	**50**	**35 (use RG-83 or paralleled 70-Ohm cables)**
2-element Yagi	**40**	**50**	**45 (use paralleled RG-62 sections)**
Dipole	**72**	**50**	**60 (use paralleled RG-63 sections)**

$\lambda/4$ Matching Section Pros and Cons

- Pros:
  - Many different impedance lines available;
  - Sections are short so low loss; and
- Cons:
  - Frequency specific – one band only.



When the load impedance of an antenna does not match the characteristic impedance of the transmission line feeding the antenna, we often wish to effect a match. Similarly, where we wish to shift from one kind of transmission line to another having a different characteristic impedance, we must also effect a match. The are numerous means of obtaining matching conditions including (with deference to the limitations of each) an L-C network, a standard transformer, and a transmission-line transformer (balun). Often overlooked in the search for a matching system is the "series-section" transformer or series matching system.

A series matching system consists of several elements: a load impedance (Zl), a characteristic impedance of a line to be match to the load (Zo), and up to two lengths of transmission line constituting the impedance transformer to effect a match between the load impedance and the line impedance. The transmission line(s) used to effect the match can be specified in terms of their characteristic impedances, which we may call Z1 and Z2, where Z1 is always the section closest to the load. Figure 1 sketches the typical series matching situation.

The advantages of using a series matching system are many. First, for almost any set of conditions, one can find commercially made transmission lines to serve as the elements of the system. Second, losses are generally low because the required lengths for Z1 and Z2 are short. Third, connections between the line lengths can be made with standard commercial connectors. We can weatherproof these connectors by standard means. The result is that we can avoid using fixed and variable lumped components (coils and capacitors) to create the impedance transformation, components which are often more difficult to weatherproof.

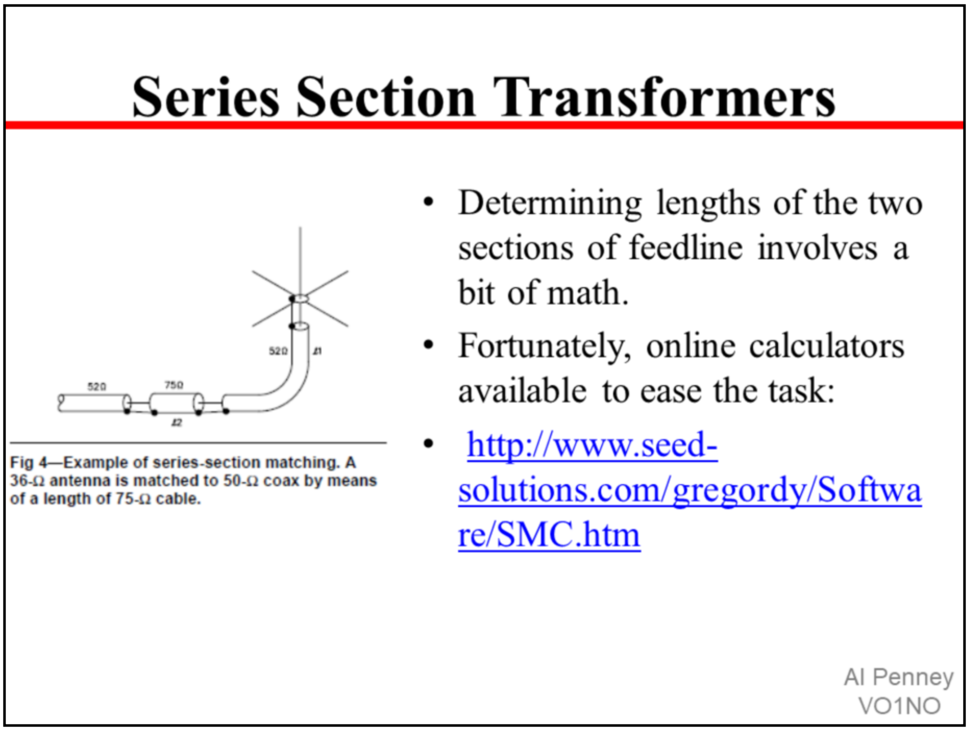
The key limitation to all series section matching systems is that they are frequency specific. Since all are composed of lengths of transmission line that will be specified in electrical degrees, the physical length of the lines will vary with frequency. In most cases, the effective operating bandwidth of these systems will be quite sufficient to cover any of the ham bands (at least above 80 meters). However, they are not broad-banded systems in the sense that a well-designed impedance-transforming balun or unun is broad-banded.

Series-Section Transformers

The *series-section transformer* has advantages over either stub tuning or the λ/4 transformer. Illustrated in **Fig 3**, the series-section transformer bears considerable resemblance to the λ/4 transformer.

(Actually, the λ/4 transformer is a special case of the series-section transformer.) The important differences are (1) that the matching section need not be located exactly at the load, (2) the matching section

may be less than a quarter wavelength long, and (3) there is great freedom in the choice of the characteristic impedance of the matching section. In fact, the matching section can have *any* characteristic impedance that is not too close to that of the main line. Because of this freedom, it is almost always possible to find a length of commercially available line that will be suitable as a matching section. As an example, consider a 75-Ω line, a 300-Ω matching section, and a pure-resistance load. It can be shown that a series-section transformer of 300-Ω line may be used to match *any* resistance between 5 Ω and 1200 Ω to the main line. Frank Regier, OD5CG, described series-section transformers in Jul 1978 *QST*. This information is based on that article. The design of a series-section transformer consists of determining the length L2 of the series or matching section and the distance L1 from the load to the point where the section should be inserted into the main line. Three quantities must be known. These are the characteristic impedances of the main line and of the matching section, both assumed purely resistive, and the complex-load impedance. Either of two design methods may be used. One is a graphic method using the Smith Chart, and the other is algebraic. You can take your choice. (Of course the algebraic method may be adapted to obtaining a computer solution.)

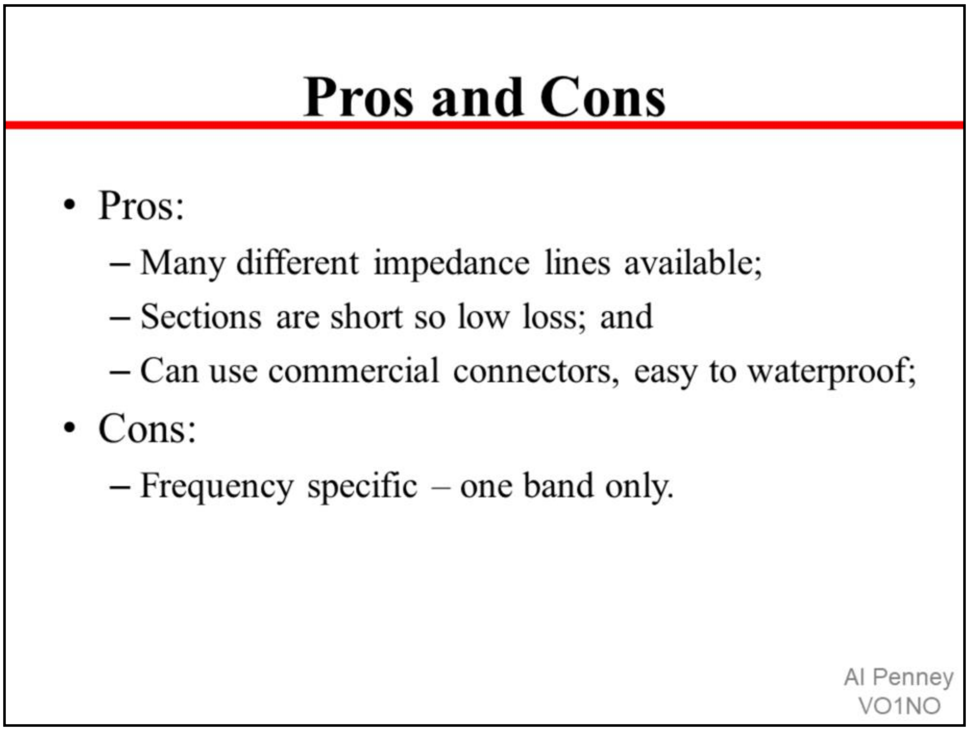


SMC computes the lengths of two different types (impedances) of transmission lines. When these two lines are connected in series (in the right order!), they perform an impedance match between a 50 Ohm transmission line, and a complex impedance, usually the impedance at an antenna. The goal is to reduce the SWR to 1.0.

In the typical scenario, the antenna impedance is close to, but not exactly, 50 Ohms of resistance (50 + j 0 Ohms). The SWR will, therefore, be greater than 1.0. It is desired to lower the SWR as much as possible. Available resources include 50 Ohm transmission line, and transmission line of another impedance. Often this is 75 Ohm line, such as RG-11. It could be a parallel combination of 50 Ohm line, creating a 25 Ohm line. It can be any line or parallel combination of lines that has an impedance other than 50 Ohms. An *antenna analyzer* or other measurement device is used to determine the impedance at the antenna. The frequency, antenna impedance, and the important characteristics (impedance and velocity factor) of the two lines are supplied to SMC. It then computes two lengths, a length of 50 Ohm line, and a length of the other line. The 50 Ohm line is connected directly to the load, usually the antenna. The other impedance line is connected to the opposite end of the 50 Ohm line. Their combination results in the transformation of the load impedance into 50 Ohms at the input of the second line section. From that point, standard 50 Ohm transmission line runs the remainder of the distance back to the radio, with an SWR of 1.0.

This technique will not work on all load impedances. There is a finite range of impedance values which can be matched by this technique. The range is related to the impedance values of the two transmission lines. The range is increased when the impedance between the two different transmission line impedances increases. In practice, the typical impedance values of verticals, dipoles, and even loops, can be matched back to 50 Ohms by using nothing more than 50 and 75 Ohm line. If you are running low power, this could be, for example, RG-8X and RG-59. If you are running high power, this could be RG-8 and RG-11. Certainly use transmission line which can handle your power level.

This approach works because any transmission line which is not terminated in its characteristic impedance will act as an impedance transformer. The impedance at the input will be a function of the impedance values (line and load) and the length of the line. By cleverly picking specific lengths of line, and lines of different impedance, it may be possible to arrive at a desired value, such as 50 + j 0 Ohms, a match to the rest of the feedline and radio.

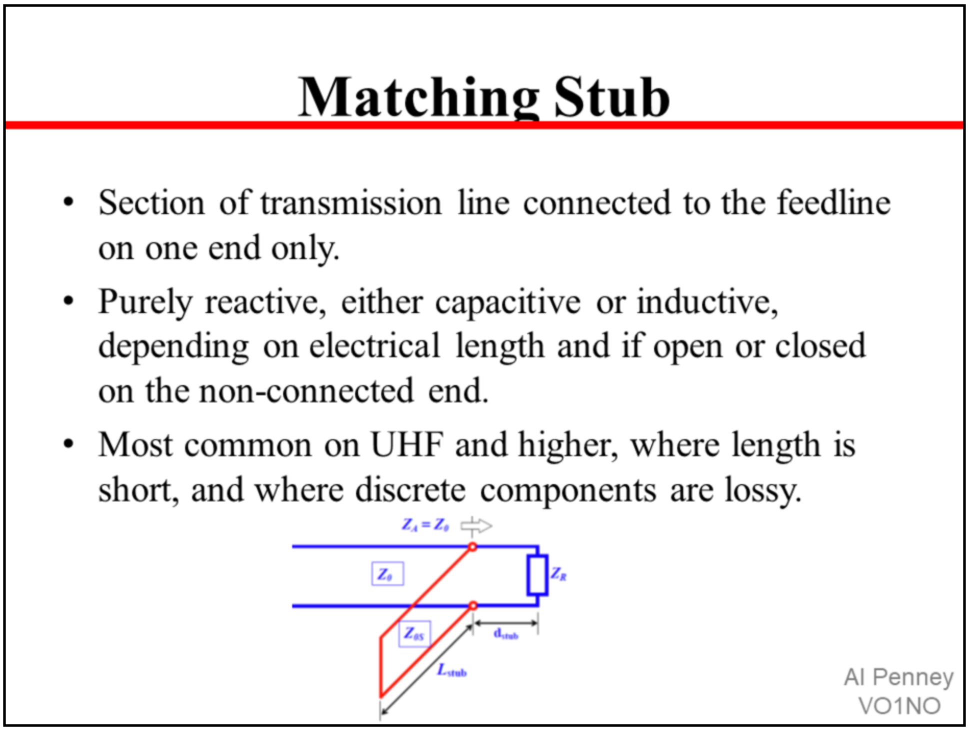


One can find commercially made transmission lines to serve as the elements of the system. Second, losses are generally low because the required lengths for Z1 and Z2 are short. Third, connections between the line lengths can be made with standard commercial connectors. We can weatherproof these connectors by standard means. The result is that we can avoid using fixed and variable lumped components (coils and capacitors) to create the impedance transformation, components which are often more difficult to weatherproof.

The key limitation to all series section matching systems is that they are frequency specific. Since all are composed of lengths of transmission line that will be specified in electrical degrees, the physical length of the lines will vary with frequency. In most cases, the effective operating bandwidth of these systems will be quite sufficient to cover any of the ham bands (at least above 80 meters). However, they are not broad-banded systems in the sense that a well-designed impedance-transforming balun or unun is broad-banded

Non-50/75 Ohm Coaxial Transmission Lines

Impedance	Types
12.5 Ω	RG192, RG193, RG194
25 Ω	RG73, RG191, RG230, RG328 M17/124
35/37 Ω	RG83, RG100, RG264
93/95 Ω	RG7, RG22, RG43, RG57, RG62, RG71, RG111, RG130, RG131, RG133, RG180, RG294, RG317 M17/15, M17/30, M17/56, M17/90, M17/95, M17/100, M17/137, M17/139, M17/177, M17/178, M17/182, M17/185, M17/195,
100 Ω	RG285, RG287, RG383
125 Ω	RG23, RG24, RG63, RG79, RG89, RG160 M17/16, M17/31, M17/218
140 Ω	RG102
150 Ω	RG72, RG125
185 Ω	RG114 M17/47, M17/208
190 Ω	RG146



Matching Stubs

As explained in Chapter 24, a mismatch-terminated transmission line less than λ/4 long has an input impedance that is both resistive and reactive. The equivalent circuit of the line input impedance

at any one frequency can be formed either of resistance and reactance in series, or resistance and reactance in parallel. Depending on the line length, the series resistance component, RS, can have any

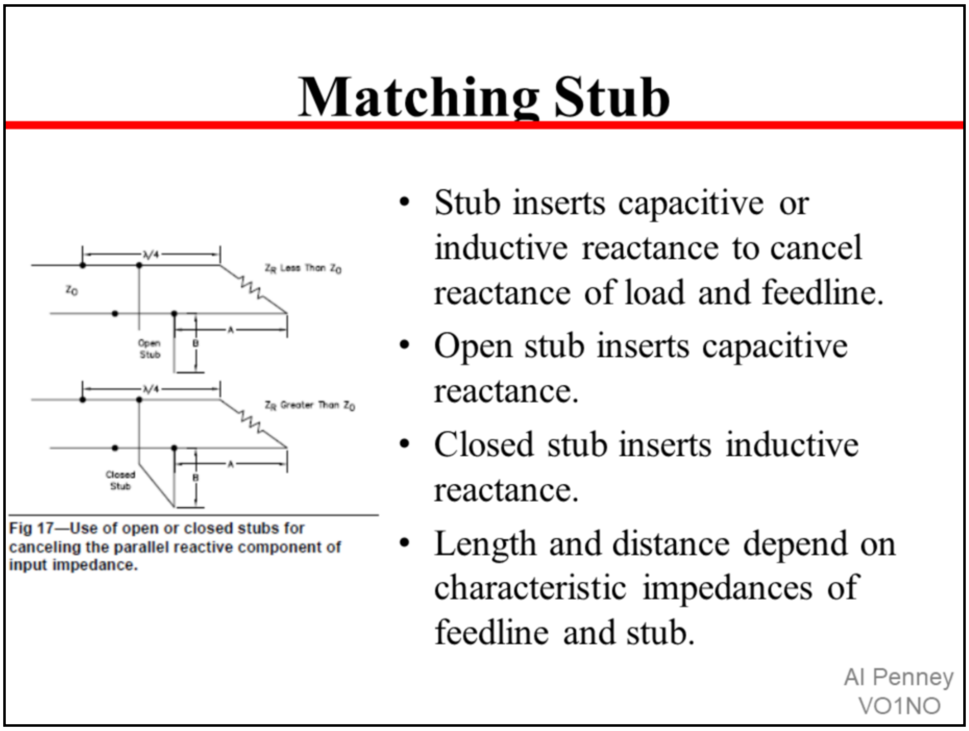
value between the terminating resistance ZR (when the line has zero length) and Z0 2/ZR (when the line is exactly λ/4 wavelength long). The same thing is true of RP, the parallel-resistance component.

In microwave and radio-frequency engineering, a **stub** or **resonant stub** is a length of transmission line or waveguide that is connected at one end only. The free end of the stub is either left open-circuit or (always in the case of waveguides) short-circuited. Neglecting transmission line losses, the input impedance of the stub is purely reactive; either capacitive or inductive, depending on the electrical length of the stub, and on whether it is open or short circuit. Stubs may thus function as capacitors, inductors and resonant circuits at radio frequencies.

Stubs work by means of standing waves of radio waves along their length.

Their reactive properties are determined by their physical length in relation to the wavelength of the radio waves. Therefore, stubs are most commonly used in UHF or microwave circuits in which the wavelengths are short enough that the stub is conveniently small. They are often used to replace discrete capacitors and inductors, because at UHF and microwave frequencies lumped components perform poorly due to parasitic reactance. Stubs are commonly used in antenna impedance matching circuits, frequency selective filters, and resonant circuits for UHF electronic oscillators and RF amplifiers.

Stubs can be constructed with any type of transmission line: parallel conductor line (where they are called Lecher lines), coaxial cable, stripline, waveguide, and dielectric waveguide. Stub circuits can be designed using a Smith chart, a graphical tool which can determine what length line to use to obtain a desired reactance.



RS and RP do not have the same values at the same line length, however, other than zero and λ/4 wavelength. With either equivalent there is some line length that will give a value of RS or RP equal to the characteristic impedance of the line. However, there will be reactance along with the resistance. But if provision is made for canceling or "tuning out" this reactive part of the input impedance, only the resistance will remain. Since this resistance is equal to the Z0 of the transmission line, the section from the reactance-cancellation point back to the generator will be properly matched.

Tuning out the reactance in the equivalent series circuit requires that a reactance of the same value as XS (but of opposite kind) be inserted in series with the line. Tuning out the reactance in the equivalent

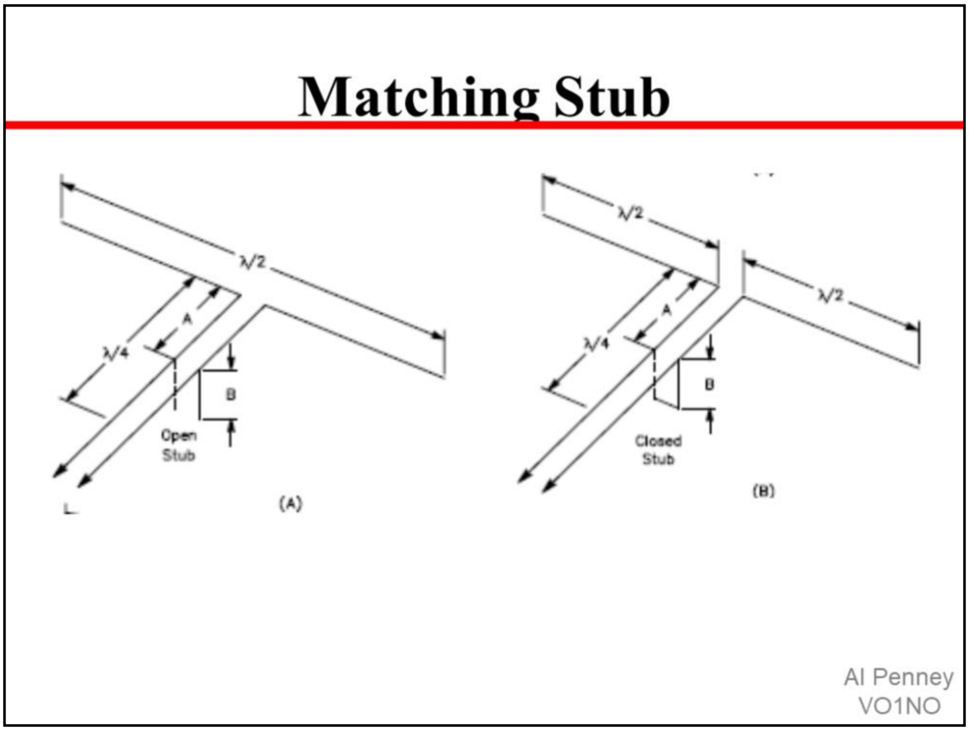
parallel circuit requires that a reactance of the same value as XP (but of opposite kind) be connected across the line. In practice it is more convenient to use the parallel-equivalent circuit. The transmission

line is simply connected to the load (which of course is usually a resonant antenna) and then a reactance of the proper value is connected across the line at the proper distance from the load. From this point

back to the transmitter there are no standing waves on the line.

A convenient type of reactance to use is a section of transmission line less than λ/4 wavelength long, terminated with either an open circuit or a short circuit, depending on whether capacitive reactance or inductive reactance is called for. Reactances formed from sections of transmission line are called *matching stubs*, and are designated as *open* or *closed* depending on whether the free end is open or short circuited. The two types of matching stubs are shown in the sketches in **Fig 17**.

The distance from the load to the stub (dimension A in Fig 17) and the length of the stub, B, depend on the characteristic impedances of the line and stub and on the ratio of ZR to Z0. Since the ratio of ZR to Z0 is also the standing-wave ratio in the absence of matching (and with a resonant antenna), the dimensions are a function of the SWR. If the line and stub have the same Z0, dimensions A and B are dependent on the SWR only. Consequently, if the SWR can be measured before the stub is installed, the stub can be properly located and its length determined even though the actual value of load impedance is not known.



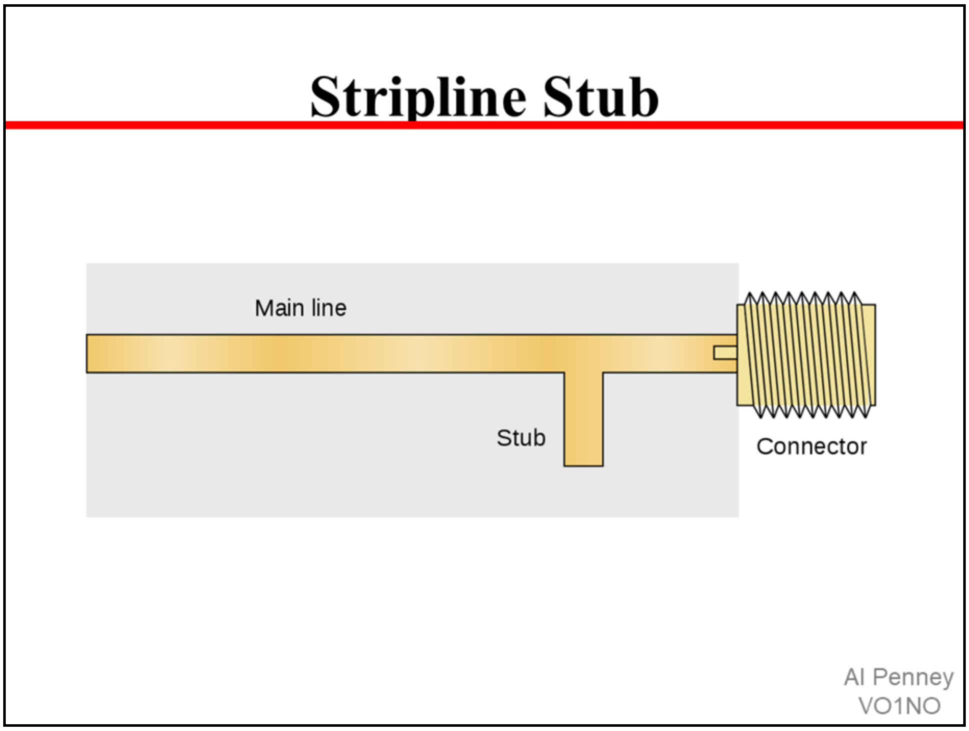
Typical applications of matching stubs are shown above, where open-wire line is being used. From inspection of these drawings it will be recognized that when an antenna is fed at a current loop, as in Fig 18A, ZR is less than Z0 (in the average case) and therefore an open stub is called for, installed within the first $\lambda/4$ wavelength of line measured from the antenna. Voltage feed, as at B, corresponds to ZR greater than Z0 and therefore requires a closed stub.

Stub matching

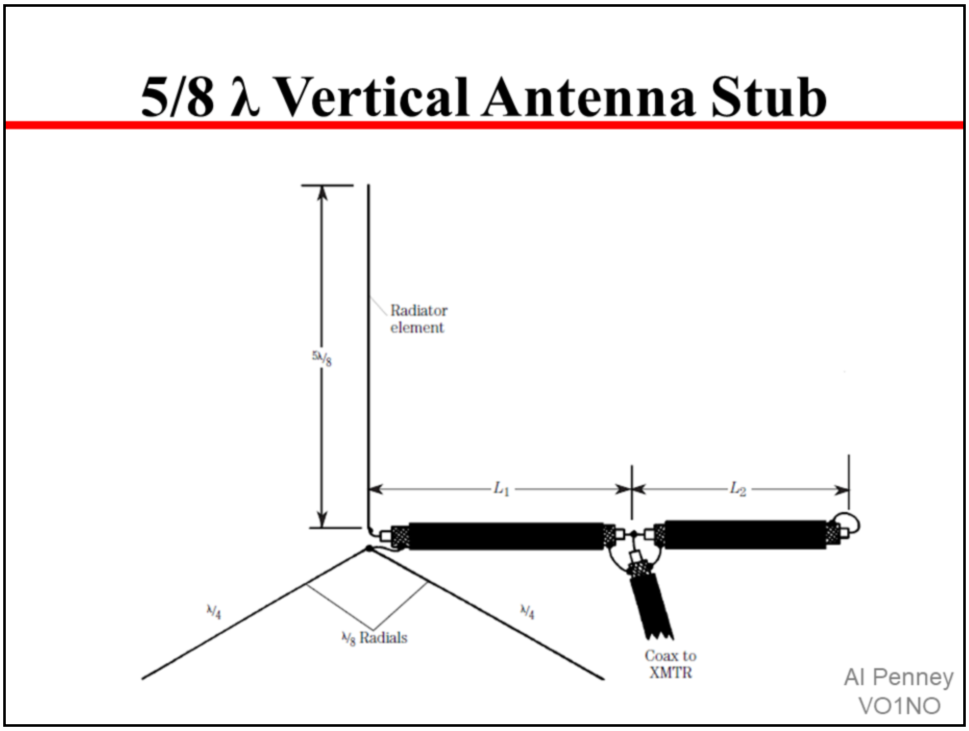
In a stripline circuit, a stub may be placed just before an output connector to compensate for small mismatches due to the device's output load or the connector itself.

Stubs can be used to match a load impedance to the transmission line characteristic impedance. The stub is positioned a distance from the load. This distance is chosen so that at that point the resistive part of the load impedance is made equal to the resistive part of the characteristic impedance by impedance transformer action of the length of the main line. The length of the stub is chosen so that it exactly cancels the reactive part of the presented impedance. That is, the stub is made capacitive or inductive according to whether the main line is presenting an inductive or capacitive impedance respectively. This is not the same as the actual impedance of the load since the reactive part of the load impedance will be subject to impedance transformer action as well as the resistive part. Matching stubs can be made adjustable so that matching can be corrected on test.

A single stub will only achieve a perfect match at one specific frequency. For wideband matching several stubs may be used spaced along the main transmission line. The resulting structure is filter-like and filter design techniques are applied. For instance, the matching network may be designed as a Chebyshev filter but is optimised for impedance matching instead of passband transmission. The resulting transmission function of the network has a passband ripple like the Chebyshev filter, but the ripples never reach 0 dB insertion loss at any point in the passband, as they would do for the standard filter.



In a stripline circuit, a stub may be placed just before an output connector to compensate for small mismatches due to the device's output load or the connector itself.



This figure shows a single stub tuner on a 5/8λ vertical (Carr 2001). (Straw 2007, p26.12-15) explains the technique.

As the radiating element of a vertical monopole increases beyond λ/4 wl, both the resistance and the reactance seen at the series feedpoint increase, as well. Thus, for any significant increase in the length of the vertical radiator, direct matching to commonly available transmission lines will result in increased SWR on the transmission line, so some form of matching network or ATU should be employed. Again, an alternative is to directly ground the base of the radiator and use some form of shunt-feed network to accomplish the proper match.

The base impedance of a 5/8-wavelength antenna is about 1600 ∧—certainly not an acceptable match for anything but a custom-designed transmission line. So some form of impedance matching is needed.

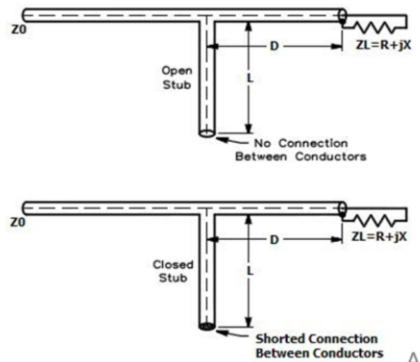
For a single-band 5/8 ωλ antenna, one option is to use a form of *stub matching* employing unbalanced coaxial cable, such as shown above. The lengths are given by

L1 (meters) = 38 / F (MHz) = L1 (feet) = 122 / F (MHz)

L2 (meters) = 9 / F (MHz) = L2 (feet) = 30 / F (MHz)

Stub Calculator

- Online stub calculator at: http://www.arcticpeak.com/antennapages/single_stub_match.html



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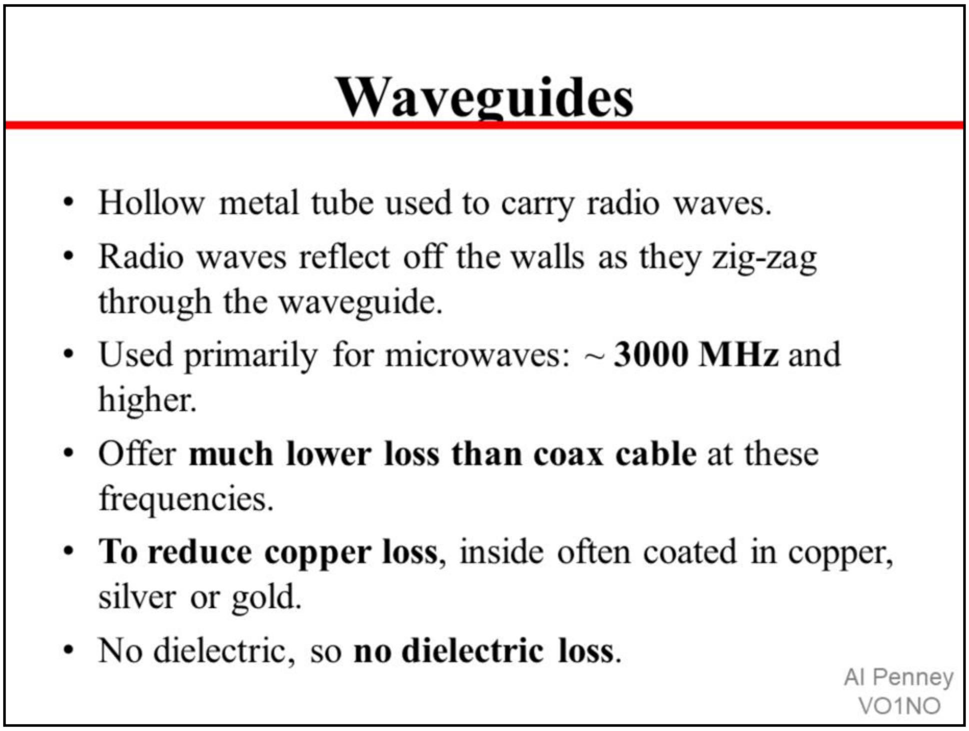
Matching Stub Pros and Cons

- Pros:
  - Multiple bands possible if several stubs used;
  - Can replace lossy discrete components at UHF and higher frequencies.
- Cons:
  - Stubs can be unwieldy on HF;
  - Single band only (though multiple stubs possible);



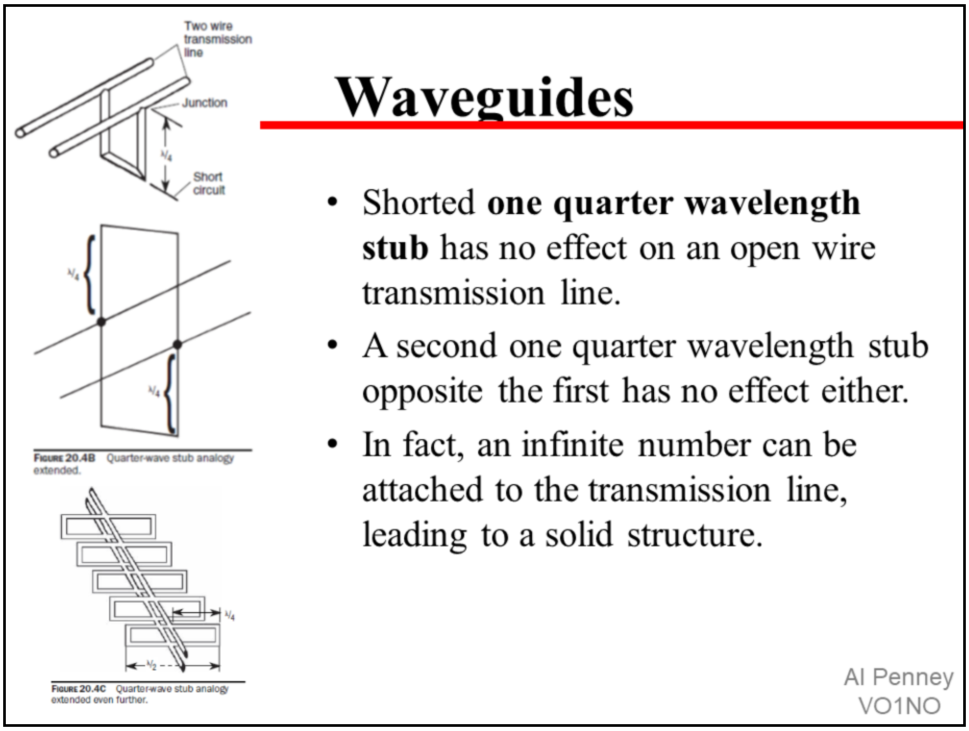
A **waveguide** is a hollow metal pipe used to carry radio waves. This type of waveguide is used as a transmission line mostly at microwave frequencies, for such purposes as connecting microwave transmitters and receivers to their antennas, in equipment such as microwave ovens, radar sets, satellite communications, and microwave radio links.

The electromagnetic waves in a (metal-pipe) waveguide may be imagined as travelling down the guide in a zig-zag path, being repeatedly reflected between opposite walls of the guide. For the particular case of **rectangular waveguide**, it is possible to base an exact analysis on this view. Propagation in a dielectric waveguide may be viewed in the same way, with the waves confined to the dielectric by total internal reflection at its surface.

A **closed waveguide** is an electromagnetic waveguide (a) that is tubular, usually with a circular or rectangular cross section, (b) that has electrically conducting walls, (c) that may be hollow or filled with a dielectric material, (d) that can support a large number of discrete propagating modes, though only a few may be practical, (e) in which each discrete mode defines the propagation constant for that mode, (f) in which the field at any point is describable in terms of the supported modes, (g) in which there is no radiation field, and (h) in which discontinuities and bends may cause mode conversion but not radiation.

The dimensions of a hollow metallic waveguide determine which wavelengths it can support, and in which modes. Typically the waveguide is operated so that only a single mode is present. The lowest order mode possible is generally selected. Frequencies below the guide's cutoff frequency will not propagate. It is possible to operate waveguides at higher order modes, or with multiple modes present, but this is usually impractical.

Waveguides are almost exclusively made of metal and mostly rigid structures. There are certain types of "corrugated" waveguides that have the ability to flex and bend but only used where essential since they degrade propagation properties. Due to propagation of energy in mostly air or space within the waveguide, it is one of the lowest loss transmission line types and highly preferred for high frequency applications where most other types of transmission structures introduce large losses. Due to the skin effect at high frequencies, electric current along the walls penetrates typically only a few micrometers into the metal of the inner surface. Since this is where most of the resistive loss occurs, it is important that the conductivity of interior surface be kept as high as possible. For this reason, most waveguide interior surfaces are plated with copper, silver, or gold.



Consider a quarter-wavelength shorted stub. A short circuit is transformed to an open circuit by a quarter-wavelength of any transmission line. Thus, the "looking-in" impedance of such a stub is infinite, so when it is connected in parallel across a transmission line the stub has no electrical effect on passing waves for which the stub length is $\lambda/4$. In other words, at its resonant frequency,

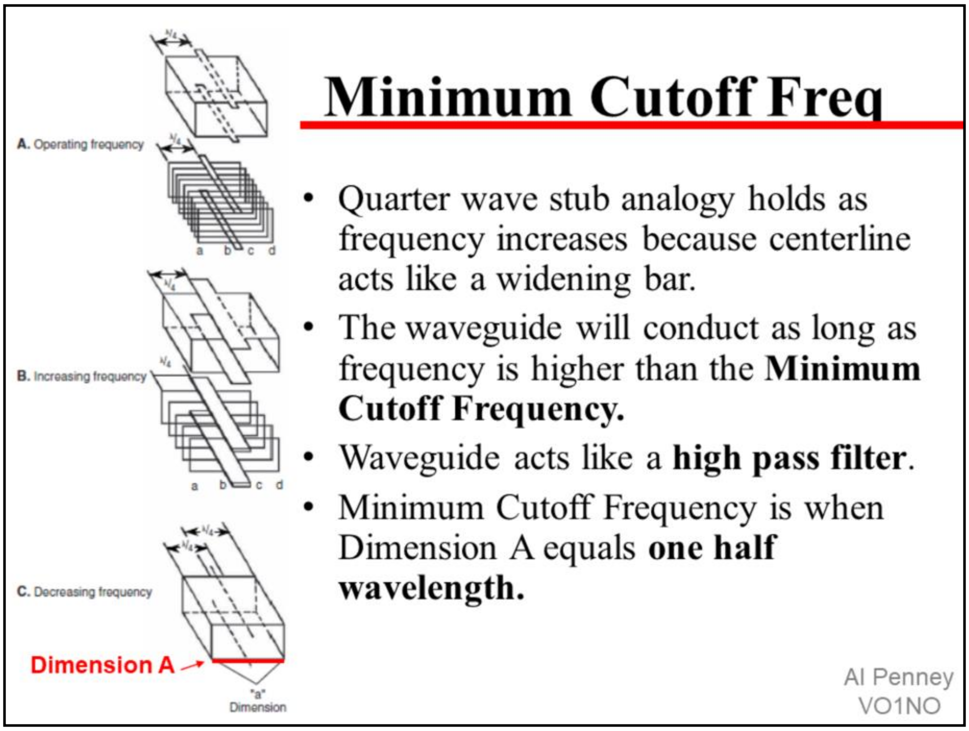
the stub is a *metallic insulator* and can be used to physically support the transmission line.

Similarly, we can connect a second $\lambda/4$ stub in parallel with the first across the same points on the transmission line without loading down the line. This arrangement

effectively forms a half-wavelength pair. The impedance is still infinite, so no harm is done. But if we can do this at one point along the transmission line, we can do it at

many (Fig. 20.4C). Ultimately, if we do it everywhere, we have totally surrounded the original parallel-wire transmission line with a metal skin! The waveguide is analogous to an infinite

number of center-fed "half-wave pairs" of A�quarter-wave shorted stubs connected across the line. The result is the continuous metal pipe structure of the common rectangular waveguide.



On first glance, an explanation based on λ/4 wavelength shorted stubs would seem to be valid only at one frequency. It turns out, however, that the analogy also holds up at other frequencies so long as the frequency is higher than a certain *minimum cutoff frequency.* The waveguide thus acts like a high-pass filter. Waveguides also have an upper frequency limit; between the upper and lower limits, waveguides support a bandwidth of 30 to 40 percent of the cutoff frequency. Between segments the centerline (which represents the conductors in the parallel line analogy) of the waveguide becomes a “shorting bar” that widens as the operating frequency gets higher and the λ/4 wavelength stub on each side gets shorter or narrows as the operating frequency becomes lower and the λ/4 wavelength stub lengthens.

At the cutoff frequency, the fraction of the *a* dimension that acts as though it is a transmission line conductor is a minimum; in other words, the back-to-back λ/4 stubs virtually touch at the midpoint

of *a*. At this frequency, $a = \lambda/2$ wavelength. Above that frequency, the effective or apparent width of the conductor increases and ½ wavelength < *a*.

Below the cutoff frequency, the chamber ceases to function as a waveguide; instead, it acts like a conventional parallel transmission line with a pure reactance connected across the two conductors. Thus,

the (low-frequency) cutoff frequency is defined as the frequency at which the *a* dimension is less than $\lambda/2$ wavelength.

Waveguide Propagation Modes

- Signals in regular transmission line are Transverse Electromagnetic (TEM) – E and H fields orthogonal to direction of travel and each other.
- TEM waves will not propagate in a waveguide due to **boundary conditions**.
- E field must be orthogonal to surface.
- H field must not be orthogonal to surface.

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Propagation Modes in Waveguides

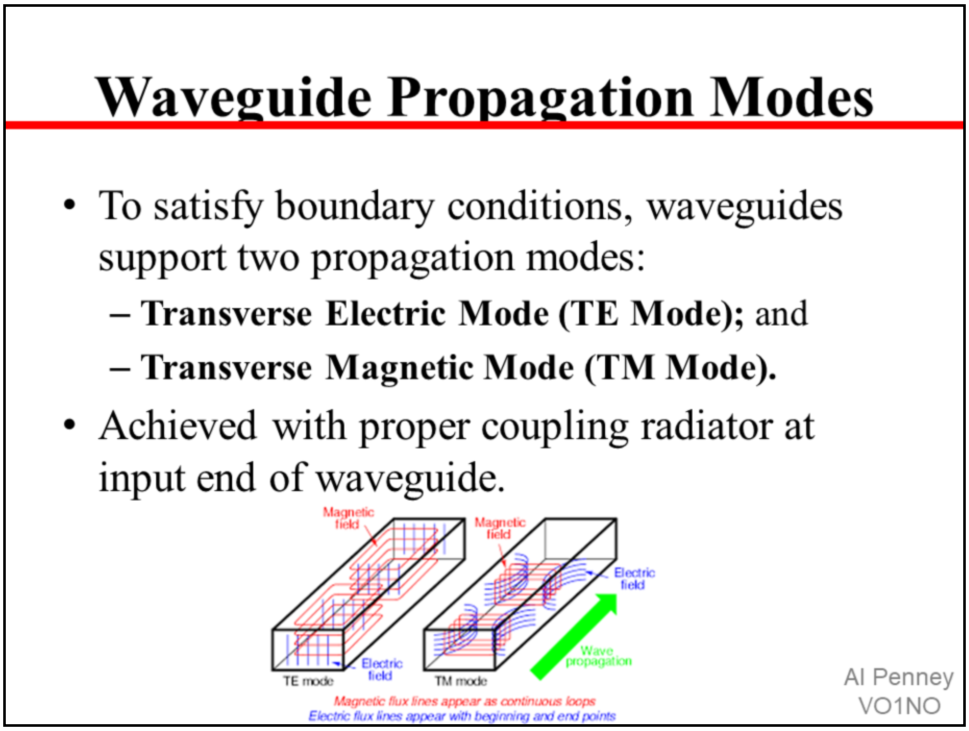
Whether in a conventional transmission line or a microwave waveguide, the signal propagates as an electromagnetic wave, not as a longitudinal current. The transmission line supports a *transverse electromagnetic* (TEM) field. The word *transverse* indicates that the wave is propagating at right angles to both the electric and the magnetic fields. In addition to the word *transverse* the E- and H-fields are said to be "normal" or "orthogonal" to the direction of travel—three different ways of saying the same thing: right-angledness.

Boundary Conditions

In contrast, a TEM wave will not propagate in a waveguide because different constraints, called *boundary conditions,* apply. Although the wave in the waveguide propagates through the air (or inert gas dielectric) in a manner similar to free-space propagation, the phenomenon is bounded by the walls of the waveguide—clearly a far different situation than for a TEM wave in free space!
The boundary conditions for waveguides are these:

•)>> The electric field must be orthogonal to the conductor in order to exist at the surface of that conductor.

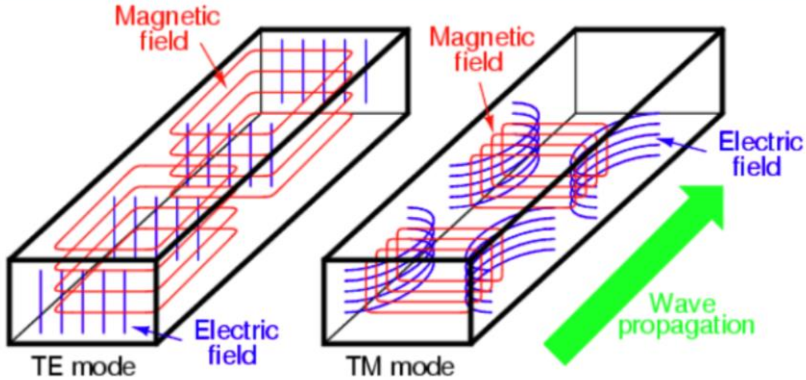
•)>> The magnetic field must not be orthogonal to the surface of the waveguide.



In order to satisfy these boundary conditions the waveguide supports two types of propagation modes: *transverse electric mode* (TE mode) and *transverse magnetic mode* (TM mode). The TEM mode for radio waves propagating through free space violates the boundary conditions because the magnetic field is not parallel to the surface and so does not occur in waveguides. The transverse electric field requirement means that the E–field must be perpendicular to the conductor wall of the waveguide. This requirement is met by use of a proper coupling scheme at the input end of the waveguide. A vertically polarized coupling radiator will provide the necessary transverse field.

One boundary condition requires that the magnetic (H) field must not be orthogonal to the conductor surface. An H–field that is at right angles to the E–field (which *is* orthogonal to the conductor surface) meets this requirement (see Fig. 20.6). The planes formed by the magnetic field are parallel to both the direction of propagation and the wide surface dimension of the waveguide. As the wave propagates away from the input radiator, it resolves into two components that are not along the axis of propagation and are not orthogonal to the walls. The component along the waveguide axis violates the boundary conditions, so it is rapidly attenuated. For the sake of simplicity, only one component is shown in Fig. 20.7. Three cases are shown in Fig. 20.7A, B, and C, respectively: high, medium, and low frequency

Comparison of fields in TE and TM modes



Magnetic flux lines appear as continuous loops
Electric flux lines appear with beginning and end points

www.opamp-electronics.com/tutorials/waveguides_2_14_08.htm

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Waveguide Propagation Modes

- Two points to remember:
  - **In the transverse electric mode, a component of the magnetic field is in the direction of propagation**; and
  - **In the transverse magnetic mode, a component of the electric field is in the direction of propagation**.
- Waveguides are too big a subject to fully cover here – check the ARRL UHF/Microwave Experimenter's Manual for more information.

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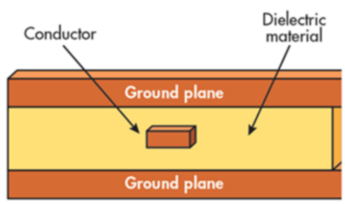
Stripline

- Transmission line consisting of a flat conductor inside a printed circuit board with a ground plane both above and below.



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Stripline

Stripline is a flat conductor with a ground plane both above and below the conductor. The variant of stripline where the space between the two ground planes is completely filled with a dielectric material is sometimes known as *triplate*. Stripline can be manufactured by etching the transmission line pattern on to a printed circuit board. The bottom of this board is left completely covered in copper and forms the bottom ground plane. A second board is clamped on top of the first. This second board has no pattern on the bottom and plain copper on the top to form the top ground plane. A sheet of copper foil may be wrapped around the two boards to electrically bond the two ground planes firmly together. On the other hand, stripline for high power applications such as radar will more likely be made as solid metal strips with periodic dielectric supports, essentially air dielectric.

Stripline is a transverse electromagnetic (TEM) transmission line medium invented by Robert M. Barrett of the Air Force Cambridge Research Centre in the 1950s. Stripline is the earliest form of planar transmission line.

A stripline circuit uses a flat strip of metal which is sandwiched between two parallel ground planes. The insulating material of the substrate forms a dielectric. The width of the strip, the thickness of the substrate and the relative permittivity of the substrate determine the characteristic impedance of the strip which is a transmission line. As shown in the diagram, the central conductor need not be equally spaced between the ground planes. In the general case, the dielectric material may be different above and below the central conductor.

To prevent the propagation of unwanted modes, the two ground planes must be shorted together. This is commonly achieved by a row of vias running parallel to the strip on each side.

Like coaxial cable, stripline is non-dispersive, and has no cutoff frequency. Good isolation between adjacent traces can be achieved more easily than with microstrip. Stripline provides for enhanced noise immunity against the propagation of radiated RF emissions, at the expense of slower propagation speeds when compared to microstrip lines. The effective permittivity of striplines equals the relative permittivity of the dielectric substrate because of wave propagation only in the substrate. Hence striplines have higher effective permittivity in comparison to microstrip lines, which in turn reduces wave propagation speed

Stripline

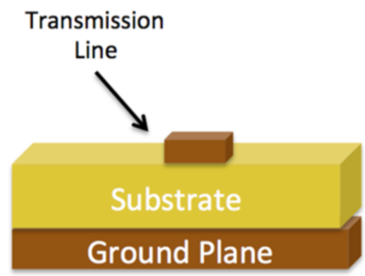
•Greater isolation of transmission lines

•Supports more densely populated designs (traces are smaller, large number of internal layers possible)

•Requires stricter manufacturing tolerances

Microstripline

- Transmission line similar to stripline, but open above the conductor.
- No dielectric or ground plane above conductor.
- **Poorer shielding than coax cable**.



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Microstrip

Microstrip is similar to stripline but is open above the conductor. There is no dielectric or ground plane above the transmission line, there is only dielectric and a ground plane below the line. Microstrip is a popular format, especially in domestic products, because microstrip components can be made using the established manufacturing techniques of printed circuit boards. Designers are thus able to mix discrete component circuits with microstrip components. Furthermore, since the board has to be made anyway, the microstrip components have no additional manufacturing cost. For applications where performance is more important than cost a ceramic substrate might be used instead of a printed circuit. Microstrip has another small advantage over stripline; the line widths are wider in microstrip for the same impedance and thus manufacturing tolerances and minimum width are less critical on high-impedance lines. A drawback of microstrip is that the mode of transmission is not entirely transverse. Strictly speaking, standard transmission line analysis does not apply because other modes are present, but it can be a usable approximation.

Microstrip

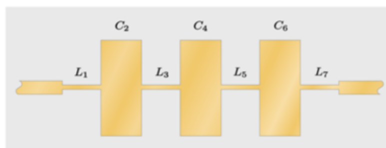
•Dielectric losses are less (when using identical materials)

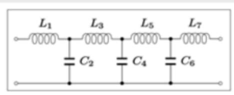
•Cheaper and easier to manufacture

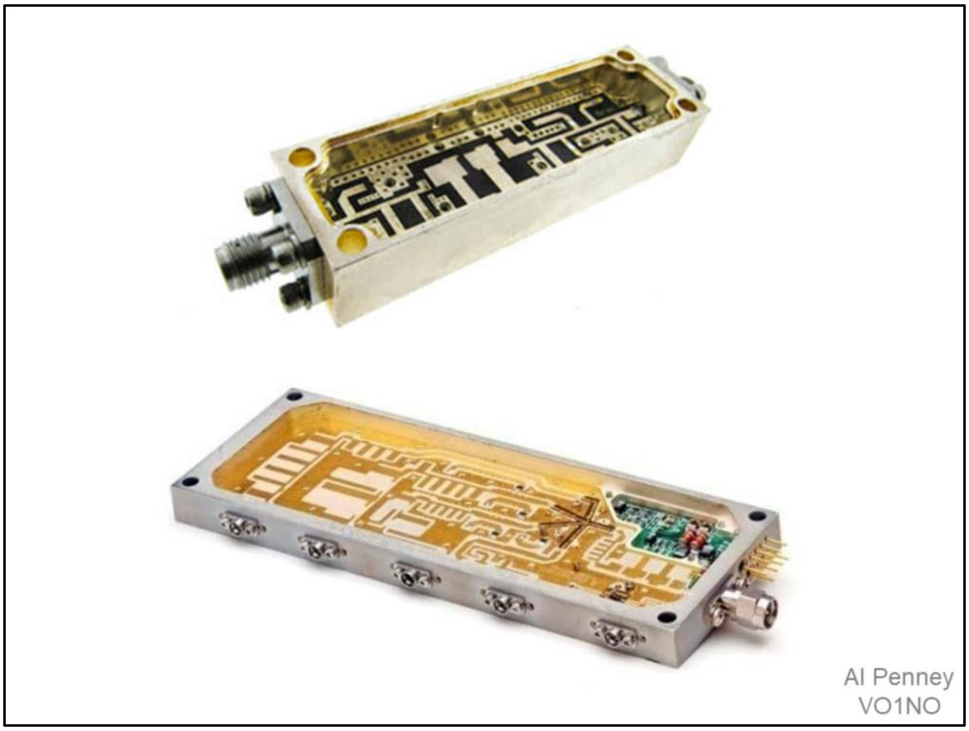
•Location of traces on top and bottom layers leads to easier debugging

Microstripline Components

- A variety of discrete components can be built by varying the shape of the copper trace, including inductors, capacitors, stubs, transformers, filters etc.
- Consult reference documents for more info.



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Questions?

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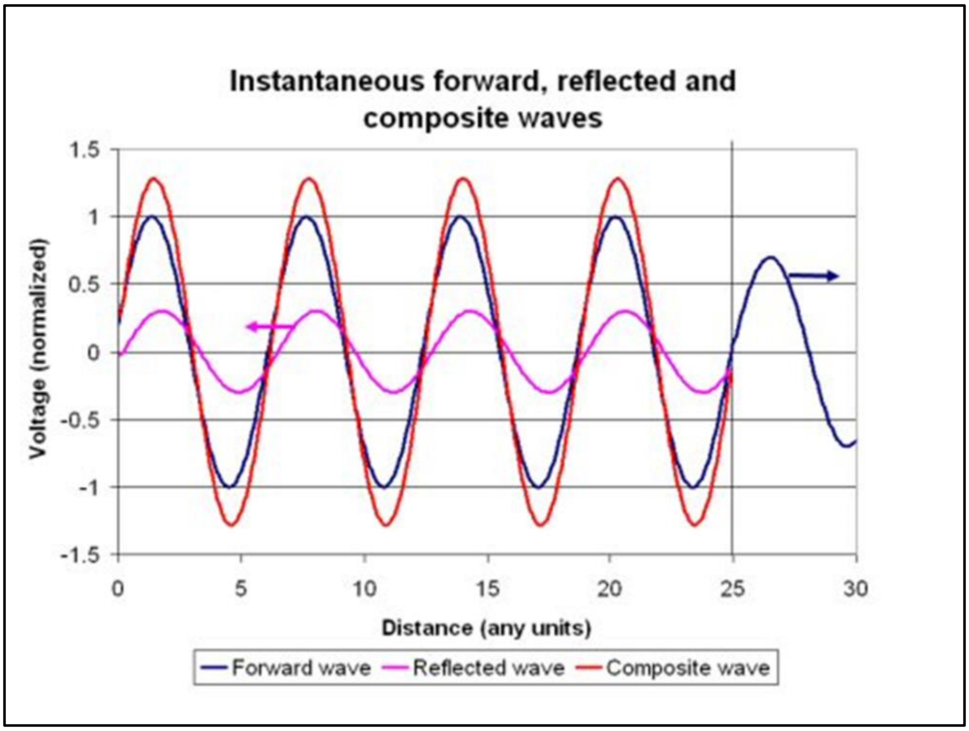
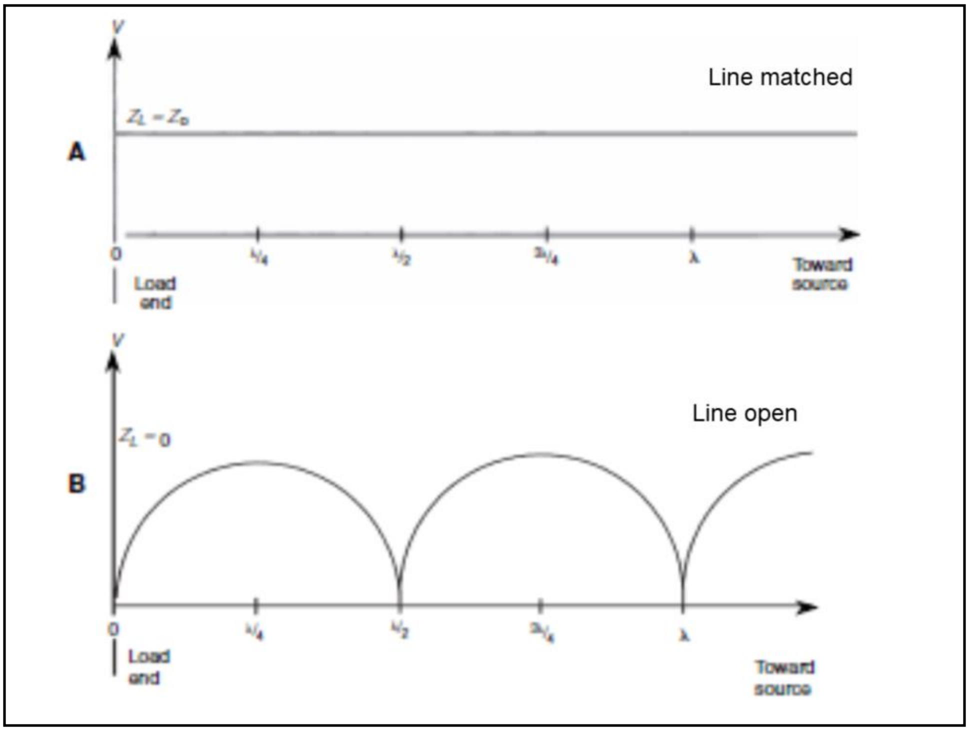


Instantaneous forward, reflected and composite waves
1.5
1
0.5
0
-0.5
-1
-1.5
Voltage (normalized)
0
5
10
15
20
25
30
Distance (any units)
Forward wave
Reflected wave
Composite wave



Steady-State Response of the Transmission Line

When a CW RF signal is applied to a transmission line, the excitation is sinusoidal (Fig. 4.9), so investigation of the steady–state ac response of the line is useful. By *steady-state*

we mean a sine–wave excitation of constant amplitude, phase, and frequency. Our initial examination of the reflected wave that results when the load impedance

does not match $Z0$ of the transmission line used a step function, but the same reflection occurs when the transmission line is excited with a pure sinusoid at a single frequency.

Here, too, when a transmission line is not matched to its load, some of the energy is absorbed by the load and some is reflected back up the line toward the source. The interference of incident

(or "forward") and reflected (or "reverse") waves creates *standing waves* on the transmission line.

If the voltage or current is measured along the line, it will vary, depending on the load, according to Fig. 4.10. In the absence of special instrumentation, such as a directional

coupler, the voltage or current seen at any point will be the vector sum of the forward and reflected waves.

Figure 4.10A is a voltage-versus-length curve for a matched line—i.e., where $Z_L = Z_0$. The line is said to be "flat" because both voltage and current are constant all along the line. But now consider Fig. 4.10B and C. Figure 4.10B shows the voltage distribution over the length of the line when the load end of the line is *shorted* (i.e., $Z_L = 0$). Of course, at the load end the voltage is zero,which results from the boundary condition of a short circuit. As noted earlier, the same conditions are repeated every half-wavelength along the line back toward the generator.

When the line is unterminated or open-circuited (i.e., Z_L = □), the pattern in Fig. 4.10C results. Note that it is the same shape as Fig. 4.10B for the shorted line, but the phase is shifted 90 degrees, or a quarter-wavelength, along the line. In both cases, the reflection is 100 percent, but the phase relationships in the reflected wave are exactly opposite each other.

Of course, if Z_L is not equal to Z_0 but is neither zero nor infinite, yet another set of conditions will prevail on the transmission line, as depicted in Fig. 4.10D. In this case, the nodes exhibit some finite voltage, V_{MIN}, instead of zero.

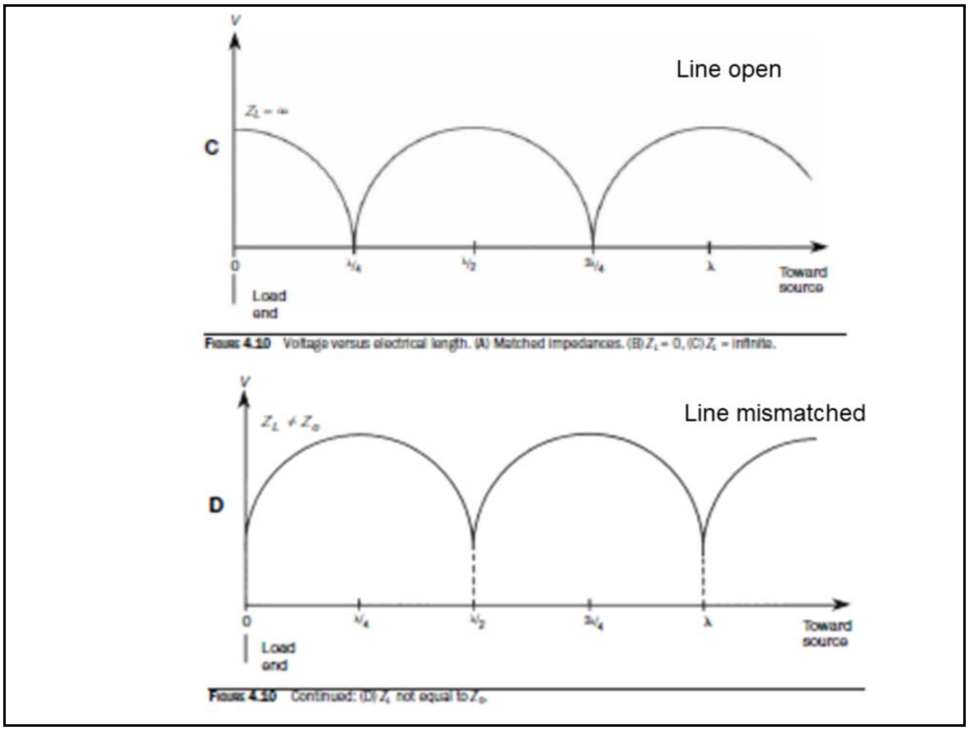


Figure 4.10 Voltage versus electrical length. (A) Matched impedances. (B) $Z_L = 0$, (C) Z_L = infinite.

Figure 4.10 Continued: (D) Z_L not equal to Z_o.

When the line is unterminated or open-circuited (i.e., Z_L = □), the pattern in Fig. 4.10C results. Note that it is the same shape as Fig. 4.10B for the shorted line, but the phase is shifted 90 degrees, or a quarter-wavelength, along the line. In both cases, the reflection is 100 percent, but the phase relationships in the reflected wave are exactly opposite each other.

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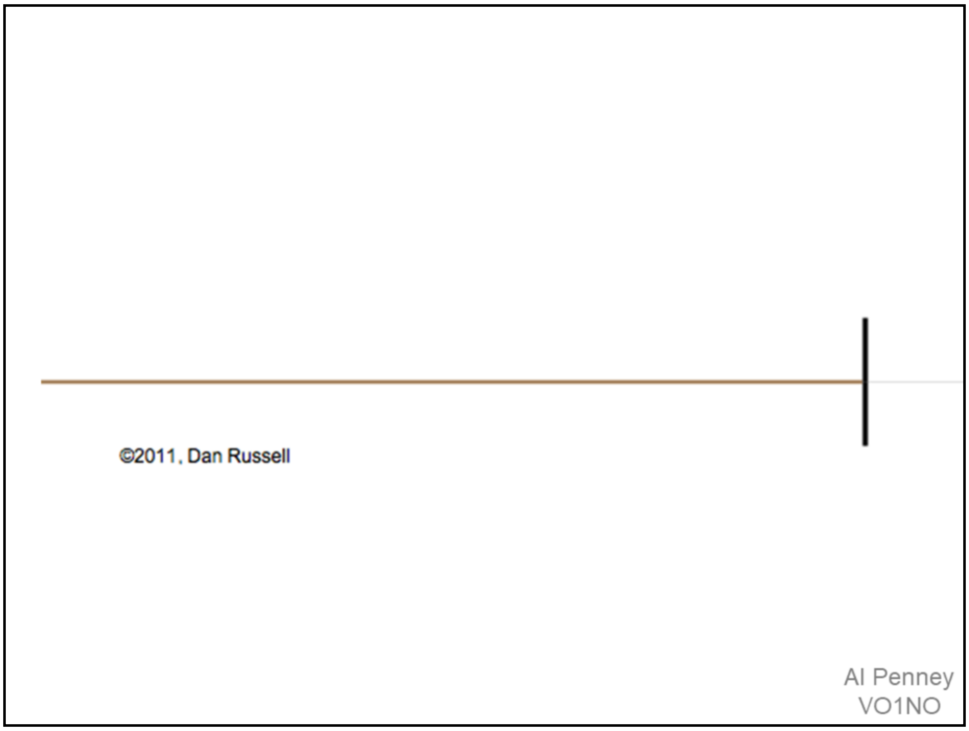
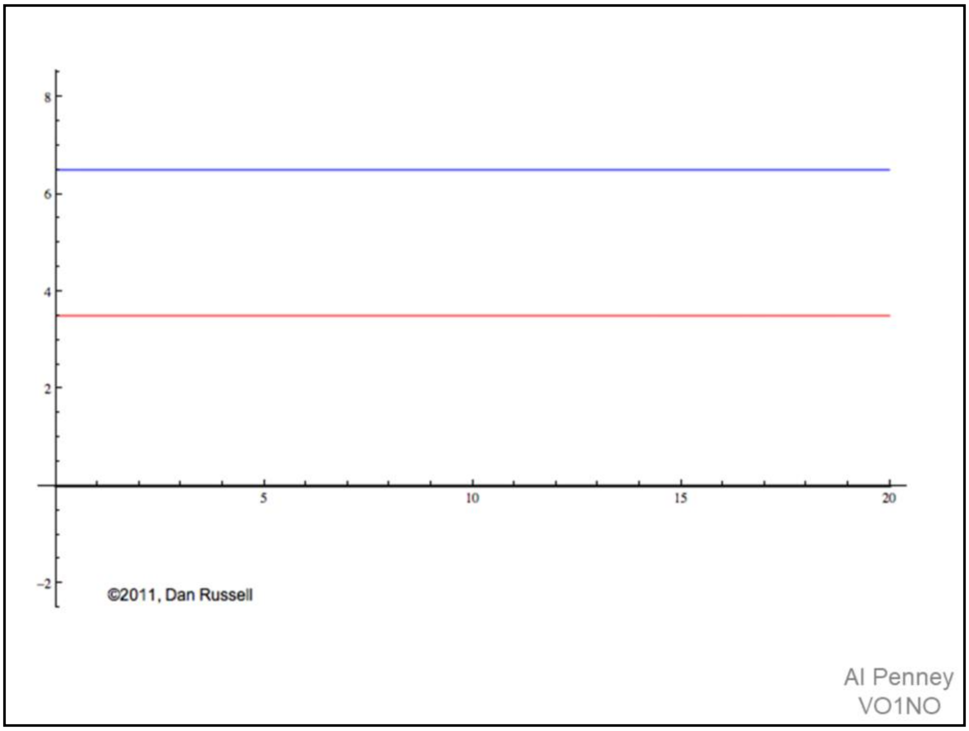
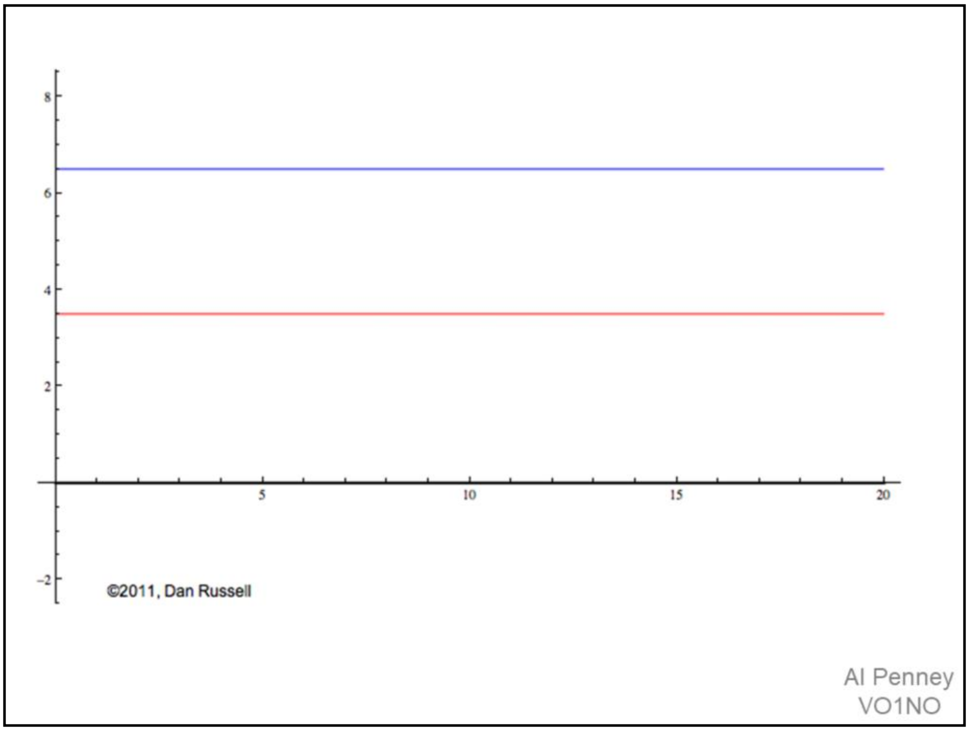


The Incident Wave Train
A wave train is just a long sinusoidal wave that may have a beginning (or end) but which is long enough that once the front of the wave has passed by, the wave just looks like an infinitely long sinusoidal wave function. The animation at left shows such a sinusoidal wave train traveling toward the right. I've shown the wave passing through a "boundary" without any change in the amplitude or phase of the wave. This is what we would expect to see for a wave encountering a second medium that has exactly the same characteristic mechanical impedance as the first medium. The characteristic mechanical impedance is the square root of the product of the elastic modulus and the mass density of the medium. Even though the materials might be different, as long as the characteristic impedances are the same, the wave train will pass directly from the first medium into the second medium without any change.

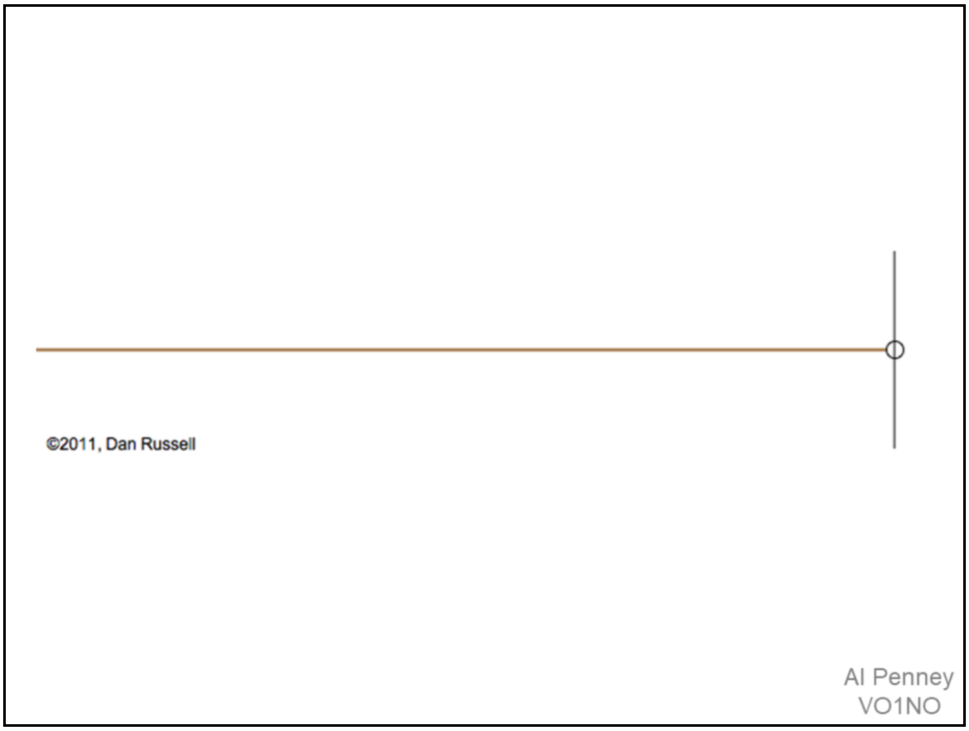


Standing Waves (superposition of two waves traveling in opposite directions)
A traveling wave represents a disturbance that moves from one location to another location at a speed which depends on the elastic and inertia properties of the medium. The animation at left shows two traveling waves: the blue wave is traveling to the right and the red wave is traveling to the left. Both waves have the same amplitude, the same frequency, and the same wavelength.
The black signal in the animation represents the superposition of these two oppositely directed traveling waves. As these waves pass through each other and add together, they create a standing wave - a pattern which neither moves left or right, but simply oscillates up and down as a function of time. The amplitude of this standing wave is twice that of the individual waves when the two waves are in phase so that peaks and valleys line up. The amplitude of the standing wave is zero when the two waves are completely opposite phase so that they peaks of one wave line up with the valleys of the other wave; the two wave amplitudes cancel each other out.

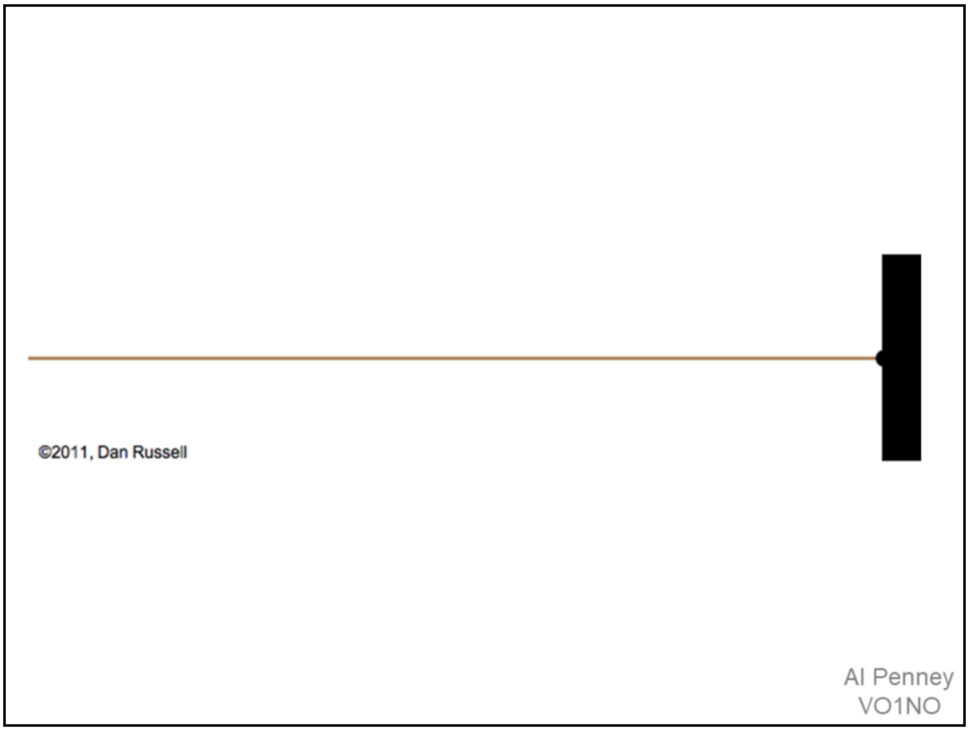


This animation is almost the same as the one above, except that the red left-going wave is now upside down. This changes the initial shape of the combined sum wave as the two waves begin to pass through each other, but after a short time, the standing wave looks exactly the same. The bottom line is that whenever two waves with the same amplitude and the same frequency are traveling in opposite directions in the same medium, the result is a standing wave.

A traveling wave carries energy and momentum with it as it travels; for a traveling wave the kinetic and potential energy both travel with the wave and have the same value at a specific position at a specific time. In a traveling wave, the potential and kinetic energy do not trade back and forth like they do for oscillation. However, a standing wave (which results from two waves traveling in opposite directions) does not involve the propagation of energy. The energy in a standing wave is trapped between the nodes, and simply trades back and forth between kinetic and potential forms.



Standing Waves Due To Reflections from Hard and Soft Boundaries
One way to get two waves traveling in opposite directions is to have a single wave train reflect from a boundary. The animation at left shows a wave train moving toward the right, incident upon a free boundary. A free boundary means that there is no force to limit the displacement. Mathematically, this results in the string having zero slope at the boundary. As was the case for a wave pulse reflection from a free end, the reflected wave has the same upright orientation as the incident wave, but is now traveling in the opposite direction.
As the incident wave train is reflected from the free end, a standing wave is quickly formed as the right-going incident wave is superposed with the left-going, upright reflected wave. The standing wave neither moves right or left, but simply oscillates up and down. Points on the string that do not move (where the amplitude is zero) are called **nodes**. Locations where the standing wave pattern has a maximum amplitude are called anti-nodes.



This animation shows how a wave train reflects from a hard, rigid boundary to form a standing wave. Just as we saw for a wave pulse reflection from a fixed end, the reflected wave is inverted (upside down) compared to the incident wave, and is now traveling in the opposite direction.

As the reflected wave superposes with the incident wave, a standing wave is quickly formed. There is no net transfer of energy The fixed end of the string is a node.

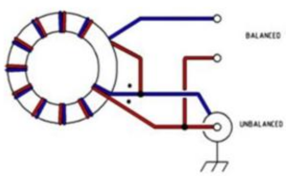
Now let's see what happens when an incident wave train encounters a boundary with an impedance that is somewhere between hard and soft. Now, only part of the incident wave is reflected; the rest is either absorbed by or transmitted into the second medium with the different impedance. The orientation of the reflected wave (whether upright or inverted) depends on whether the new material has a larger or smaller impedance. For this particular animation, the reflected wave has half of the amplitude of the incident wave, and is inverted.

The resulting superposition looks kind of like a standing wave, but it doesn't stay still. It appears to lurch forward toward the right. This should make sense since only a part of the energy carried by the original wave is reflected from the boundary; the rest of the incident energy is either absorbed by the impedance termination or transmitted into the second medium. As a result there is more energy moving to the right than there is being reflected to the left. There is a net energy flow to the right.

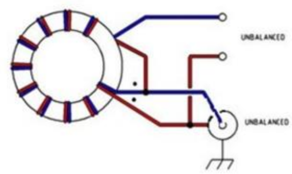
If you watch the above animation very carefully you will notice that the locations where the wave pattern reaches a maximum value are fixed in space. Likewise, the locations where the wave pattern has a minimum value are also fixed locations. So, while there is a net transfer of energy toward the right, there is an aspect of this wave pattern that does not change with time or space.

When is a Balun not a Balun?

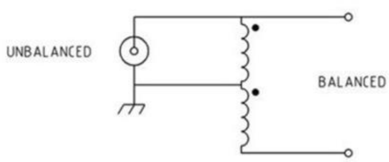
When it’s a UNUN

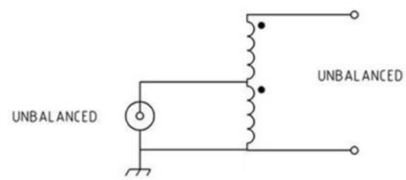


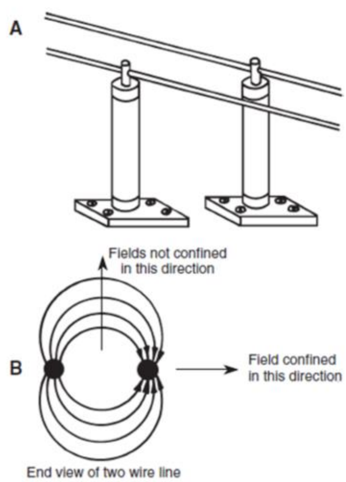
4:1 Balun



4:1 Unun







Many directly-fed Yagi antenna designs for HF and 50 MHz that are optimized for the best gain and front-to-back performance have feed-point impedances of approximately 20 to 25 Ohms, at resonance. Obviously, this will not allow a direct connection to a 50-ohm feedline without a serious mismatch (2:1 or more). There are various ways to match the driven element to the feed-line successfully; Gamma Match, T-Match, and the Hairpin (aka Beta Match) are favorites. The Gamma match is an outdated, unbalanced system that typically distorts the antenna radiation pattern. The T-match is basically two Gamma Match systems on either side of the boom, which may correct the imbalance, but is a mechanical nightmare and is difficult to tune correctly.

In a superior-balanced Yagi feed system, the driven element must be insulated from the boom and it must be split at the boom so that the driven element halves are also insulated from each other, like a dipole. A balanced Yagi can make use of a balun for balanced distribution of RF currents and a proper, predictable radiation pattern.