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430 MHz Yagi with open sleeve dipole feed

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Summary

The sleeve dipole is a popular antenna for situations where: the feed line has to extend from an extremity of the antenna, a ground provision isn't available, it needs to be slim and high power handling is required. A popular alternative for HF and VHF is the end-fed half-wave dipole.

Many sleeve dipole designs that go around are not well designed and introduce significant common mode current (braid current) in the feed line. This affects antenna performance. This document shows a design procedure (theoretically and practically) for sleeve dipoles with focus on acceptable common mode feed line current. It also shows that a sleeve dipole has less useful bandwidth compared to a similar size half wave dipole and that additional measures may be required to further reduce the feed line common mode current.

Based on the sleeve dipole a 4 element vertically polarized Yagi for the 70 cm amateur radio band is designed and implemented. Measurement setups for tuning the sleeve and construction tips are given.

Content

1. Introduction

1.1. General

Feeding a horizontally oriented half wave dipole is easy, just use a balun and run the cable straight down (figure 1.1 A). There will be no interaction between the cable and the antenna.

When the dipole has to be vertically mounted, at least the lower quarter wave part of the radiator will "see" the cable (Figure 1.1 B). One option is to first run the cable horizontally, and then go down. Frequently this isn't an attractive option.

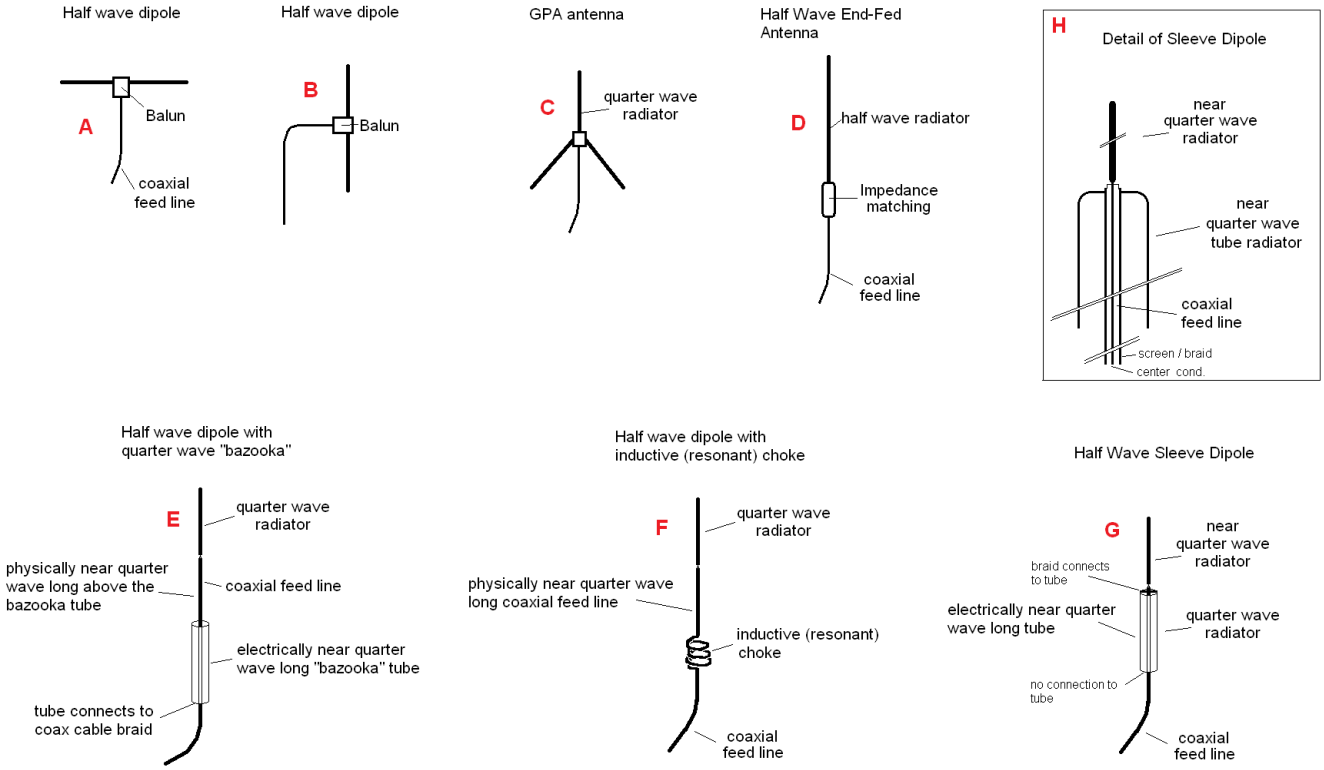


Figure 1.1. various antenna geometries for omnidirectional vertical polarization

There are several solutions for this. A simple sturdy one is to use the quarter wave ground plane antenna (figure C). The downslope of the 3 or 4 near quarter wave "radials" provides good 50 Ohms match and relatively low common mode (screen) cable current or low mast current.

When one wants a very slim structure one can divert to the half wave end-fed antenna (figure D). As the half wave end-fed impedance can be in the kOhm range, a matching circuit with relative high voltage capability is required (when using a transmitter). When using a relative thick radiator, the half wave end-fed antenna introduces significant common mode current into the cable or mast. This will affect

the radiation pattern of a half wave dipole. One of the chokes to be discussed can be used to suppress cable or mast common mode current.

Other solutions add a high impedance gap in the cable that doesn't affect the current and voltage inside the coaxial feed line.

Figure 1.1 E shows the bazooka choke (or bazooka balun). The space between the inner side of the tube and the braid of the coaxial feed line inside the tube acts as an electrically short circuited quarter wave line. This introduces high impedance across the tube edge and the feed line. The quarter wave radiator and the braid of the feed line above the bazooka tube acts both as radiator. So it is a real half wave dipole

The bazooka choke electrically disconnects itself from the dipole.

Figure 1.1 F shows the inductive and/or resonant choke. The feed line is formed into a coil just below the actual dipole. Note that also here the braid above the coil is one half of the radiator. The currents inside the feed line are not affected by the coil, but the outside braid / screen current experiences high impedance due to the coil.

1.2. The Sleeve Dipole

Figure 1.1 G shows the sleeve dipole. Figure H shows a detail of how the tube connects to the coaxial feed line. The bazooka tube is now used upside down and forms near half of the radiator.

The space between the inner side of the tube and the braid / screen must act as an electrically quarter wave shorted line. This generates high impedance isolating the feed line below from the antenna. It should be noted that the quarter wave tube should be somewhat shorter to compensate for capacitance. This will be discussed later.

As the tube has physical length somewhat below a quarter wave, the top part should be somewhat longer so that the combination of quarter wave radiator and quarter wave braid resonates (at about 70 Ohms).

In the bazooka design the tube serves as an isolator only, but in the sleeve dipole it acts as part of radiator and isolator.

Designs according to figure 1.1 D, E and G can be very thin so that they can be mounted into a thin glass fiber tube or fishing pole that acts as radome.

It should be noted that designs E, F and G have about 70 Ohms impedance. When very good 50 Ohms SWR is required, some form of matching is required.

Note:

There are many designs available on the Internet. Most of them ignore the effect of the plastic protective sheath that is in between the inner side of the tube and the braid of the feed line. This slows down the propagation speed of waves (Vf < 1). Therefore the tube should have a physical length of less then a quarter wave.

"forgetting" to compensate for Vf < 1 will result in low impedance across the gap and high common mode current in the feed line. Many designs try to correct this with lots of ferrite.

1.3. This document

This document treats the Sleeve Dipole in more depth. A yagi-uda antenna is built were the sleeve dipole (with some matching) is inside a fishing rod and the 3 yagi elements are clipped onto the fishing rod. For those that don't have confidence with matching, a yagi design is given that doesn't require matching.

The document assumes that the reader is familiar with:

- Basic knowledge of coaxial transmission lines (characteristic impedance, velocity factor (Vf), cable inductance, cable capacitance).
- currents in coaxial feed lines (inner screen current, outer screen current and center conductor current)
- common mode current (outer screen current in a coaxial feed line) and common mode impedance.
- Simple RLC circuits
- Impedance behavior of short circuited and open quarter wave line
- Concept of Characteristic impedance of (antenna) wires without return conductor.

2. Analysis of the Sleeve Dipole

2.1. Half wave dipole

2.1.1. Center fed half wave dipole

The half wave center fed dipole can be modeled based on two near quarter wave long transmission lines for modeling the reactive behavior versus frequency. The impedance at resonance can be modeled via series resistors.

Figure 2.1A shows the model. Though a ground is present in the model, in real world the two wires have a return path via capacitive coupling between them.

Thick conductors give lower characteristic impedance giving larger impedance (or SWR) bandwidth.

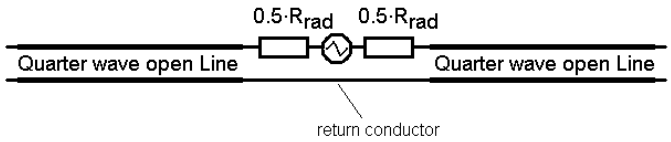
Figure 2.1B shows the end-fed model. Why do we need this? One of the two quarter wave sections is also the sleeve that should “isolate” the feed line from the dipole.

Figure 2.1C shows some additional formulas to calculate some properties.

The impedance curve of the center fed half wave dipole behaves as a series resonant circuit around the resonant frequency.

Transmission Line Model for Half-Wave monopole and Half-Wave Dipole

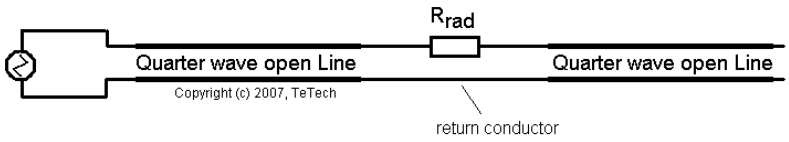
A **Model for HW dipole**



R_{rad} = 40 .. 73 Ohm

$$Z_{0\_line} \approx 138 \cdot \log\left(\frac{0.25 \cdot \lambda}{d_{wire}}\right)$$

B **Model for HW monopole**



R_{rad} = 40 .. 73 Ohm
(no ground plane)

R_{rad} = 53 .. 97 Ohm
(perfect conduct. ground plane)

The Transmission lines do not radiate. All radiation effects are in the radiation resistance.

HW monopole input impedance can be found with the Quarter Wave transmission line transformer formula ($Z_I \cdot Z_L = Z_0^2$):

C

$$Z_{Res\_HW\_monopole} = \frac{(Z_{0\_line})^2}{R_{rad}} \qquad Q = \frac{\pi}{4} \cdot \frac{Z_{0\_line}}{0.5 \cdot R_{rad}}$$

Please note: R_{rad} can change significantly because of presence of (for example) a groundplane parallel to the antenna (as with NVIS antennas). This affects both Q (so Bandwidth) and input impedance.

Figure 2.1, Half wave center-fed and end-fed properties

2.1.2. End-fed half wave dipole (monopole)

The impedance at resonant frequency for a center-fed half dipole depends only slightly on radiator diameter. The impedance for an end-fed half wave dipole depends however on the diameter.

Larger diameter gives lower $Z_{0\text{-line}}$. $(Z_{0\text{-line}})^2$ is present in the formula (figure 2.1 C). Therefore larger diameter gives lower end-fed impedance.

Below the resonant frequency the impedance becomes inductive, above it becomes capacitive. The impedance vs frequency curve behaves as a parallel resonant circuit.

An example: **Half wave dipole for 435 MHz from 6.5 mm tubing**

Lambda = 0.69 m
Expected impedance at resonance = 60 Ohms (just a guess).

Free space characteristic impedance of tubing

$Z_{0\text{-line}}$ = 200 Ohms

$Q_{HW\text{-}dip}$ = 5.2 (83 MHz impedance bandwidth)

$BW_{SWR=2}$ = 60 MHz (for 60 Ohms reference!)

End-fed impedance = 670 Ohms

Let us assume 1Arms feed current. In that case the input power is 60 W.

When feeding at an extremity (end-fed scheme), that would result in 200 Vrms at each tip of the dipole.

Imagine you want to feed this dipole using a bazooka choke, inductive choke or sleeve choke (Figure 1.1 E, F or G). You would like to have about 20 dB common mode suppression. In that case maximum 100 mA may leak out of the dipole into the feed line that goes down.

That would require an impedance of at least 200V / 0.1A = 2 kOhms. This impedance should have a small real part (resistive part) otherwise the impedance dissipates some of the RF power. When the bandwidth of the dipole is well above the useful bandwidth of the choke, one may subtract 670 Ohms.

This example shows that the end-fed impedance is of importance as it “dictates” the minimum required isolation impedance between the dipole and the down running feed line.

Why 20 dB common mode suppression?

The antenna current in the example is 1A.This generates a certain E-field at distance. Imagine that the 100 mA common mode current generates a proportional E-field, then that E-field is 10% of the antenna field. That could change the antenna E-field with -0.9 .... +0.8 dB.

In real world the CM current will generate less then 10% of the antenna E-field. The radiation pattern of a traveling wave component is different, and some of the common mode power is dissipated instead of radiated. So when you can reach 20 dB common mode suppression, actual deviation from the standard 2.13 dBi for a half wave dipole will be less then +/- 0.4 dB.

2.2. The sleeve dipole, basics

Figure 2.2 shows a sleeve dipole with expected E-field lines. Only the field lines above the structure are shown. The structure is normally used vertically where the top quarter wave radiator points upwards.

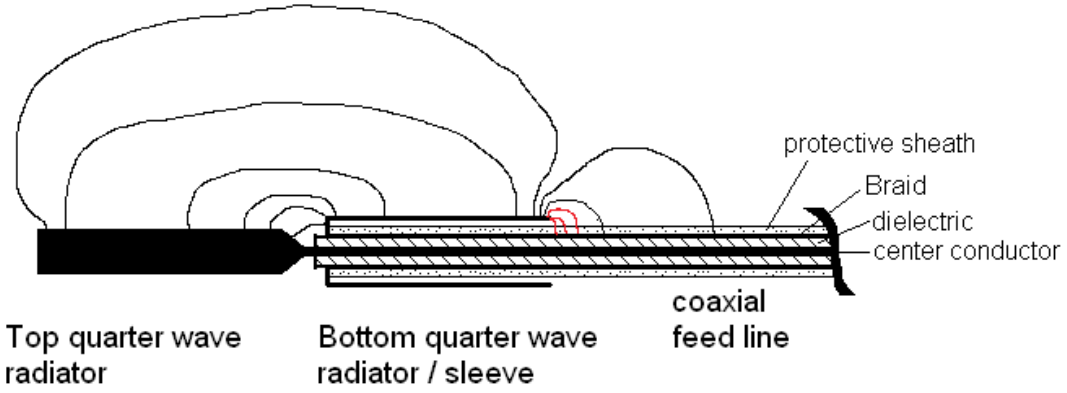


Figure 2.2, Sleeve Dipole with electric field

The gap between the right side edge of the sleeve and the braid of the feed line should show high impedance so that the dipole is isolated from the feed line. The dipole consists of the Top quarter wave radiator and the Bottom quarter wave radiator / Sleeve.

In real world the electrical length and the physical length of the sleeve will be less then a quarter wave. This is because of:

- Fringe capacitance (red field lines)
- Dielectric material in the sleeve (mostly the sheath and spacers)

2.2.1. Fringe capacitance

The small distance between the right side edge of the sleeve and the braid of the feed line is small. Therefore there will be a relative strong E-field between the edge and the feed line. These lines are shown in red. This field introduces capacitance. The capacitance is more compared to the fringe capacitance of a clean cut coaxial cable. This is because of the protrusion of the feed line (that serves as the center conductor for the sleeve).

The fringe capacitance for a clean cut approximately 5 mm cable is in the 50 fF (0.05 pF) range. The total fringe capacitance from edge to braid of feed line is in the range of 0.2 pF for a 2 mm thick braid.

Simulation shows that for a 6 mm sleeve and 2 mm feed line the fringe capacitance is about 0.25 pF. This capacitance should be counteracted by some inductive behavior of the transmission line in the sleeve. Therefore the electrical length is somewhat less then a quarter wave.

Calculation of the length reduction is very simple. You need the capacitance/m for the transmission line that is formed by the inner side of the sleeve and the outside of the braid of the coaxial feed line. To find the capacitance/m you need to know the characteristic impedance and the velocity factor. You may also calculate the capacitance directly using coaxial capacitance formula.

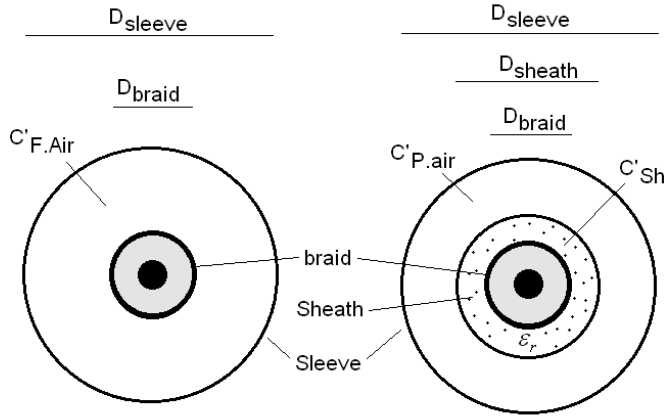
When (for example) the capacitance is 50 pF/m and the fringe capacitance is 0.25 pF, you need to remove 0.25/50 = 5 mm from the sleeve (or sleeve tube). Calculation will be discussed in the next paragraph

2.2.2. Dielectric inside the sleeve

Dielectric material reduces Vf. Therefore the physical length will be less then the required near quarter wave electrical length.

$$le_{phys} = Vf \cdot le_{electr} \quad [m]$$

When the space between the braid and inner side of the sleeve is fully filled with PE or PTFE, Vf = 0.66. When the space is partly filled, calculation is required and it is easier then you think. You don't need differential calculus. The procedure is shown in figure 2.3.



General coaxial capacitance formula

$$C' = \frac{2 \cdot \pi \cdot \varepsilon_0 \cdot \varepsilon_r}{Ln(D/d)} \quad [F/m, \quad Vs/Am]$$

Calculation of effective dielectric constant

Calculate: $C'_{F.Air}$ $C'_{P.air}$ C'_{Sh}

Calculate: $C'_{eff} = \frac{C'_{P.air} \cdot C'_{Sh}}{C'_{P.air} + C'_{Sh}}$

$$\varepsilon_{r.eff} = \frac{C'_{eff}}{C'_{F\ Air}}$$

$$Vf = \frac{1}{\sqrt{\varepsilon_{r.eff}}}$$

Figure 2.3, calculation of effective capacitance and Vf

Spacers
Sometimes spacers are required to keep the feed line centered. When you have only one thin spacer at the open end of the sleeve, there is a voltage (capacitive) effect only as the H-field is very low.

Calculate the spacer's additional capacitance ($C_{s.add}$). For air only inside the sleeve, you need to use $\varepsilon = \varepsilon_{r.s}-1$ for calculation of the additional capacitance. $\varepsilon_{r.s}$ is the

relative permittivity of the material of the spacer. The shorter the length of the spacer, the smaller the additional capacitance.

Reduce the length of the transmission line with: $C_{s.add}/C'$. C' is the capacitance/m of the sleeve transmission line (this is not the capacitance of the feed line).

When you have evenly distributed spacers (think of minimal 3), consider each spacer as a piece of transmission line with smaller Vf, and reduce the length of the sleeve with the total amount of length reduction due to Vf. When you use PE, PP or PTFE (Vf = 0.66) spacers, you can reduce the sleeve length with 0.44*(total length of spacers). So when you have 4 spacers of 3 mm, you need to reduce the length with 0.44*4*3 mm = 5.3 mm.

2.2.3. Limitation of common mode suppression

One would expect very high common mode suppression when having very high sleeve choke impedance. Even when optimally corrected for fringe capacitance and dielectric inside the sleeve.

Forget very high common mode suppression, there are limitations.
The red field lines are very close to the structure and can be treated with electrostatic theory. The field lines further away are deformed due to size no longer small to wavelength. The effect is no longer comparable with a capacitive effect, so one can't compensate this with further length reduction of the sleeve. It is more or less radiation coupling. It can best be modeled as a resistor parallel to the gap (that is across edge of sleeve and braid) when using PSPICE modeling.

In real world > 26 dB is hardly possible, so put a resistor across the sleeve / feed line transition. This limits the attenuation to a practical value when making SPICE models using transmission lines. For simulating common mode behavior of the feed line, use the Z0_line formula from figure 2.1A to calculate the common mode characteristic impedance of the braid of the coaxial cable (feed line).

Further practically feasible common mode reduction can be obtained by adding another choke. It is important that the choke doesn't create high common mode impedance at the sleeve / feed line transition. A second choke with high common mode impedance should be connected about 0.25 lambda below the sleeve. CM reduction is further discussed in paragraph 2.3 and 2.4.

Impedance bandwidth of the sleeve
The useful bandwidth of a sleeve is generally well below the useful bandwidth of the dipole itself.

Our 435 MHz example dipole requires > 1330 Ohm sleeve impedance to get 20 dB CM suppression.

When using 2.5 mm coaxial cable as feed line (RG 316) and a 6.5 mm sleeve, the sleeve's internal characteristic impedance is 60 Ohms ($Z_{0.int.sleeve}$). The formulas

below can be used to find the frequencies where the impedance drops below 1330 Ohms

$$Z_{short} = j \cdot Z_c \cdot \tan(2\pi \cdot le/\lambda) \approx \cdot \frac{4}{\pi} \cdot \frac{Z_c}{jv} \qquad v = \frac{\omega}{\omega_{0.25}} - \frac{\omega_{0.25}}{\omega}$$

For small values of v,

$v = 2\Delta\omega/\omega_{0.25} = 2\Delta f/f_{0.25}$

$f_{0.25}$ = the frequency where the sleeve is in quarter wave resonance.

For Z_{sleeve} = 1330 Ohms and $Z_{0.int.sleeve}$ = 60 Ohms, v = +/- 0.057

For $f_{0.25}$ = 435 MHz, this results in Δf = +/- 12.4 MHz

Total bandwidth = 24.8 MHz for CM suppression > 20 dB. The SWR=2 bandwidth of the center fed was about 60 MHz, So subtraction of 670 Ohms (from 2000 Ohms) is allowed.

Of course the antenna can be used outside the 24.8 MHz bandwidth, but with more deviation from the standard G = 2.14 dBi for a half wave dipole.

When operating off-resonance, the reduced impedance of the sleeve loads the end of the dipole. This will reduce the actual bandwidth of the dipole!

2.3. PSPICE modeling of sleeve dipole

2.3.1. Half wave dipole PSPICE model

PSPICE modeling is a good way to derive several properties. It is much faster then full 3D EM modeling. Of course when done, it is good to run a full 3D EM simulation to rule out mistakes...

Figure 2.4 shows the half wave dipole transmission line model.

- The quarter wave lines have 200 Ohms characteristic impedance (using formulas from figure 2.1).
- R1 and R2 model 60 ohms resonance impedance
- VAm2 is current measuring instrument to measure the center radiator current.
- Rref sets the reference impedance for the SWR measurement.

The black trace shows SWR. The useful bandwidth is significantly larger then 40 MHz, as SWR is still below 2. The pink trace shows the radiator current.

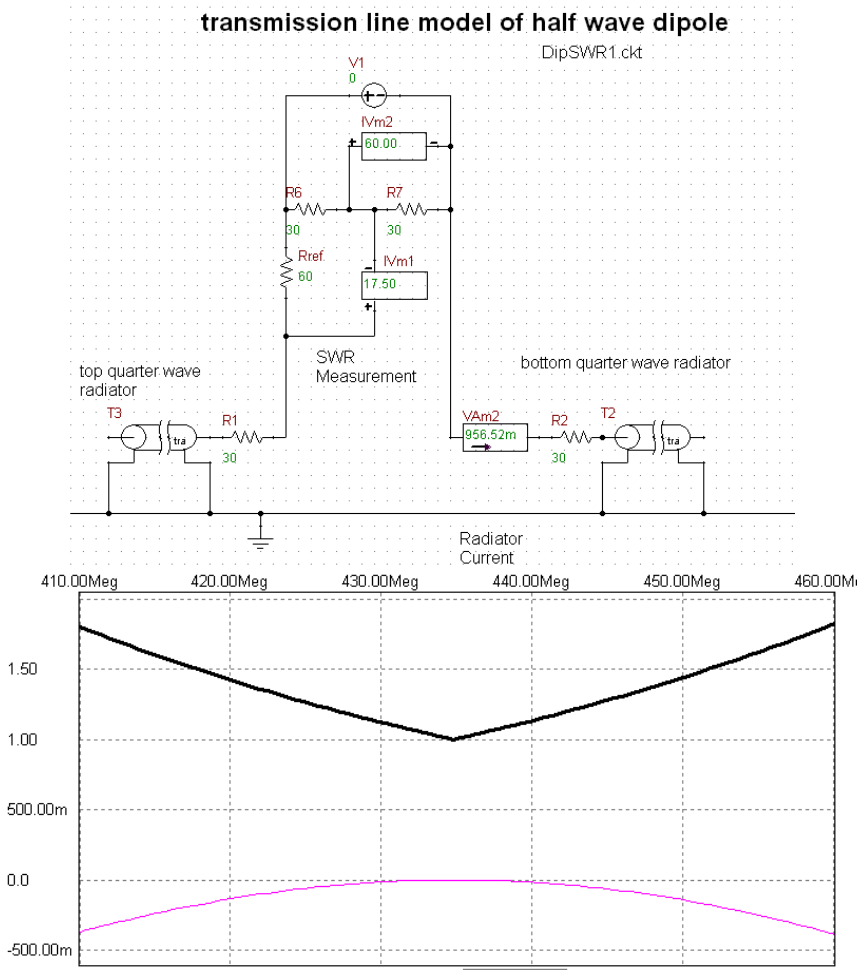


Figure 2.4, half wave dipole transmission line model with SWR curve

2.3.2. Half wave Sleeve dipole PSPICE model

Figure 2.5 shows the half wave sleeve dipole transmission line model.

- The quarter wave lines have 200 Ohms characteristic impedance (using formulas from figure 2.1).
- R1 and R2 model 60 ohms resonance impedance
- VAm2 is current measuring instrument to measure the center radiator current.
- Rref sets the reference impedance for the SWR measurement.
- T1 is the 60 Ohms transmission line that models the transmission line between the inner side of de sleeve and the feed line braid. It is connected to the edge of the lower quarter wave radiator / sleeve and the feed line's braid.
- R3 models a perfect common mode impedance of the braid of the feed line.
- VAm1 models the common mode current in the coaxial braid.

– R5 models the more or less radiation coupling between the dipole and the braid of the feed line. It avoids also that in simulation the CM suppression becomes impractically high. The capacitance (red field lines in figure 2.2) is left out.

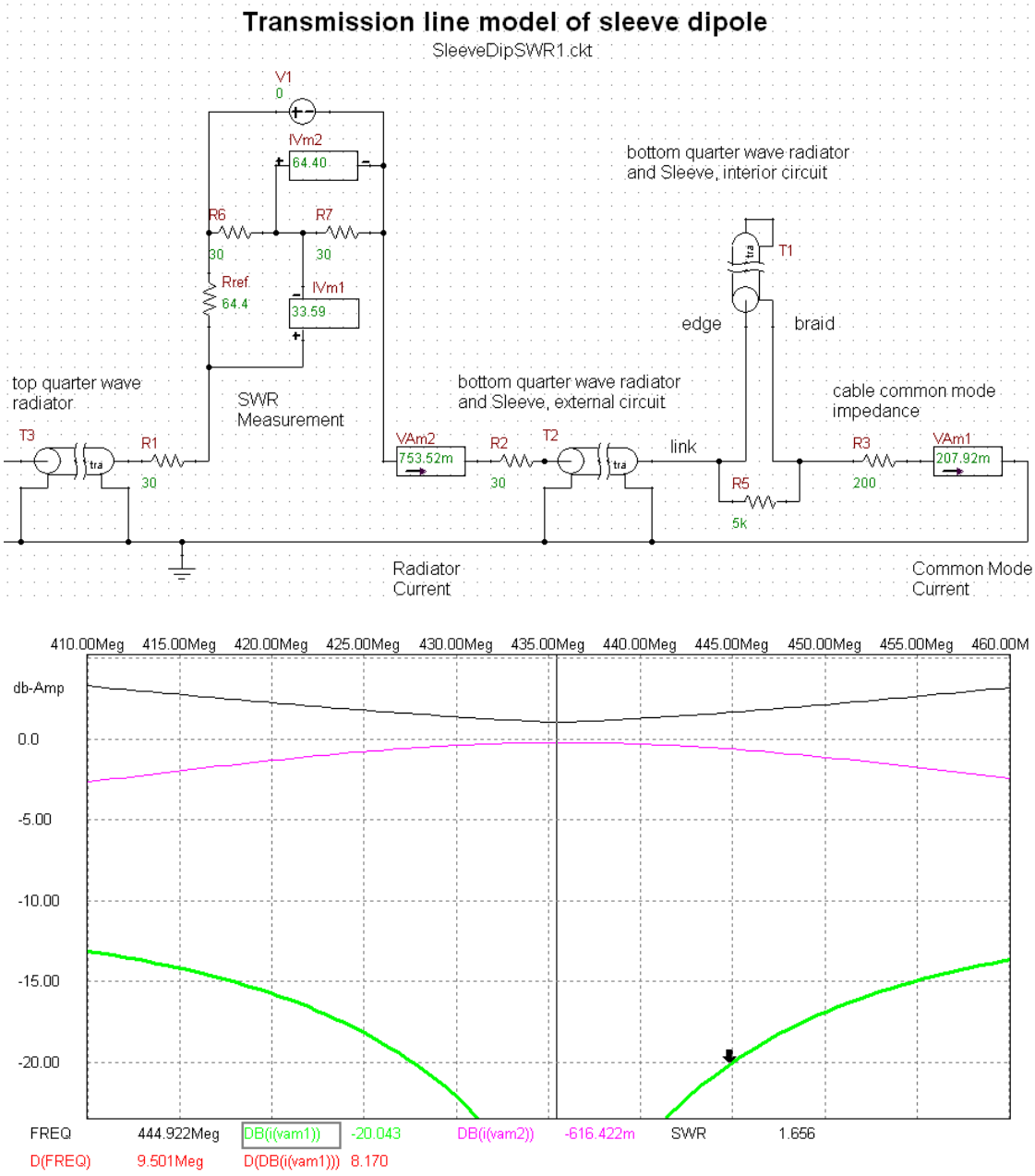


Figure 2.5, half wave sleeve dipole transmission line model with CM current

The graph has different vertical scale compared to the graph in figure 2.4.

The black trace shows SWR. The pink trace shows the radiator current. The green trace shows ratio in dB of CM current and radiator current at center frequency (1A).

The useful bandwidth (SWR=2) of the dipole is significantly reduced from about 60 MHz to 26 MHz. The useful bandwidth for CM suppression of 20 dB is 17 MHz.

Our simple hand calculation gave 24.8 MHz. This is likely because the calculation assumed that the end-fed impedance remains 670 Ohms of a span of 25 MHz, and that isn't true. It drops somewhat. We also didn't account for vector summation of impedances.

2.4. Inconvenient common mode impedance of the feed line

The PSPICE model, models the CM impedance as pure 200 Ohm real impedance. In real world the CM impedance can vary, with large imaginary part (large reactive part). This can be modeled in spice also, inclusive transmission line effects.

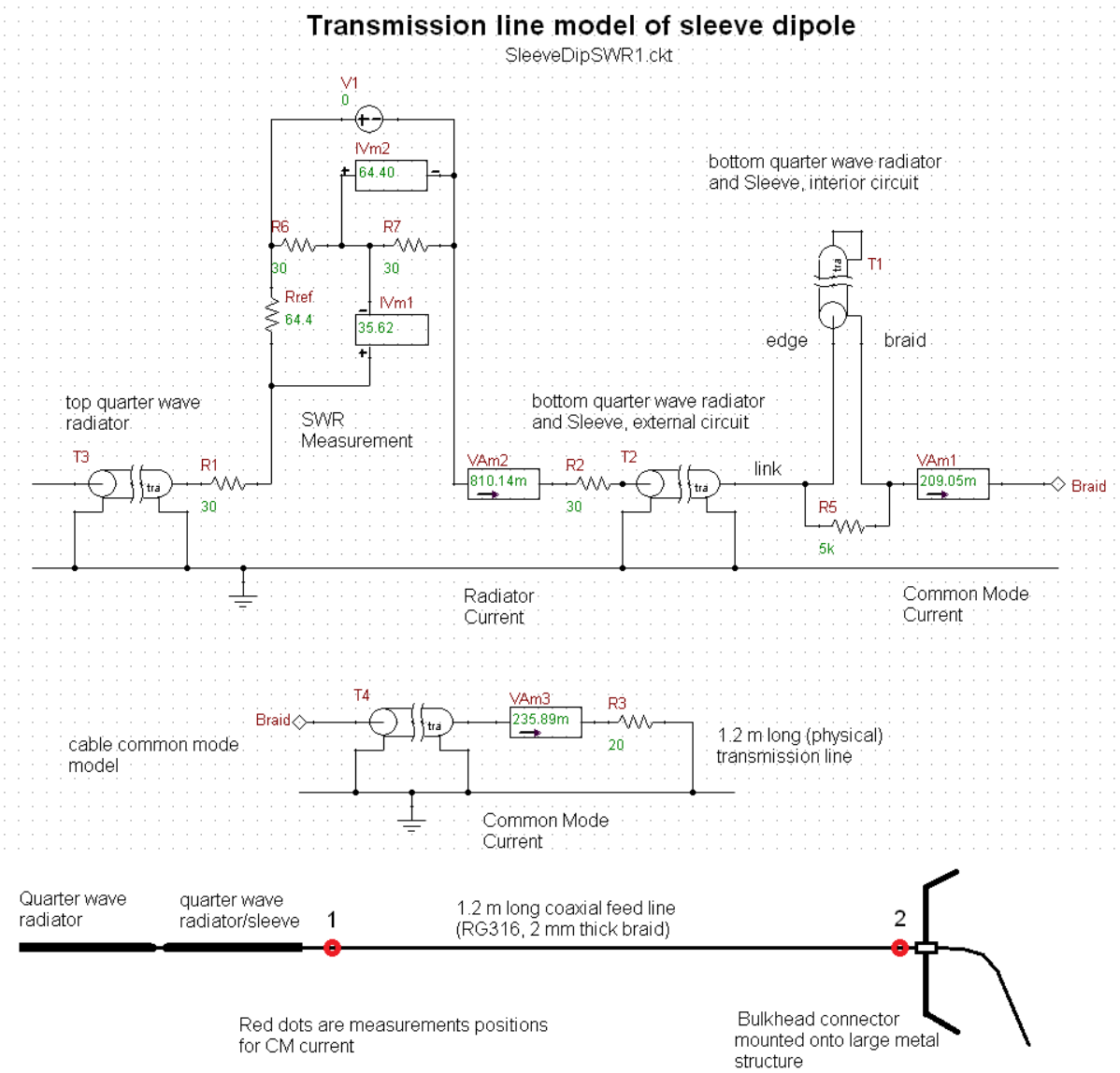
Figure 2.6 shows an example A sleeve dipole is provided with a 1.2 m long RG316 (2.55 mm) coaxial cable with connector. It is connected to a bulkhead connector that feeds the antenna signal through a metal plate. The braid of the cable is simulated using T4 (200 Ohms, 1.2 m long). The CM impedance of the metallic plate is simulated with R3 (20 Ohms). 20 Ohms is not uncommon, it can even be less.

VAm1 measures the CM current just below the sleeve, VAm3 measures the CM current just above the bulkhead connector.

SWR and CM current a few cm below the sleeve are still acceptable for the 430 to 440 MHz amateur radio band. However the CM current above the bulkhead is excessive. This will significantly affect the radiation pattern of the dipole.

The reason for this behavior is that the cable's braid provides a CM impedance that is in the range of the impedance of the sleeve to braid gap. Along the braid of the feed line that runs towards the bulkhead, that high CM impedance is converted to a lower impedance (quarter wave transformer), at the "cost" of a much higher CM current.

One can also say, the CM impedance of the braid just below the sleeve has a reasonable match to the impedance of the sleeve (that is impedance of T1 in figure 2.6). Frequently we like a good match, but here we need a bad match to get as less power possible into the braid of the coaxial feed line.



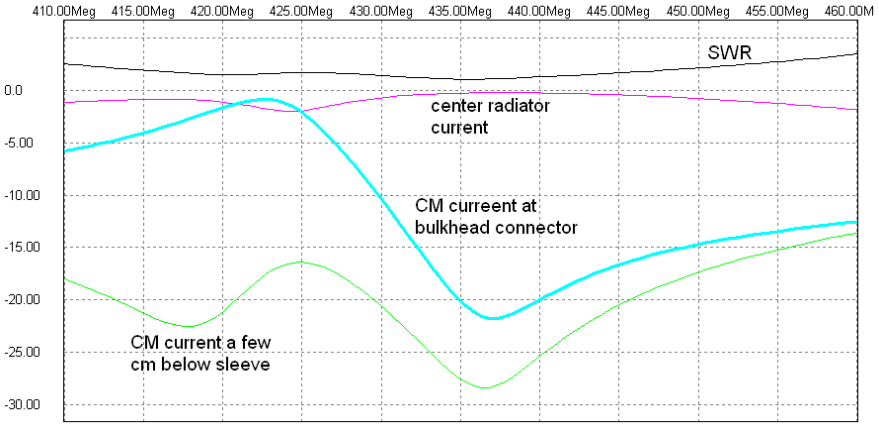


Figure 2.6, inconvenient common mode impedance

As the sleeve impedance is relatively high (to get reasonable CM suppression), it is best to provide a relative low impedance as seen from the sleeve AND/OR avoid resonance on the feed line. Having a low CM impedance of the braid just below the sleeve, gives a bad match resulting in low energy extraction from the common mode current.

A good option (to provide relative low impedance, independent of CM impedance further down the feed line), is adding a CM impedance about a quarter wave below the sleeve. If you manage to get near real impedance of >200 Ohms a quarter wave below the sleeve, the CM impedance of the braid will never be higher then 200 Ohms. This is irrespective of the CM impedance below that impedance. It is shown in figure 2.7.

The additional impedance is modeled via Rferrite1 and T5 (200 Ohms, 170 mm long). The effect can be improved by adding another CM impedance a quarter wave below (Rferrite2).

These impedances can be generated using any choke you like, but a choke having resistive behavior has preference. The simplest solution is with suitable ferrite, not just ferrite. One may use several distributed ferrites to get a similar effect. Experimentation with measurement is definitely required to get the best results for the least amount of ferrite.

Figure 2.7 shows the simulation setup, and the results. The model could be improved by adding the frequency dependent behavior of the ferrite, but the improvement compared to just 1.2 m of cable is impressive.

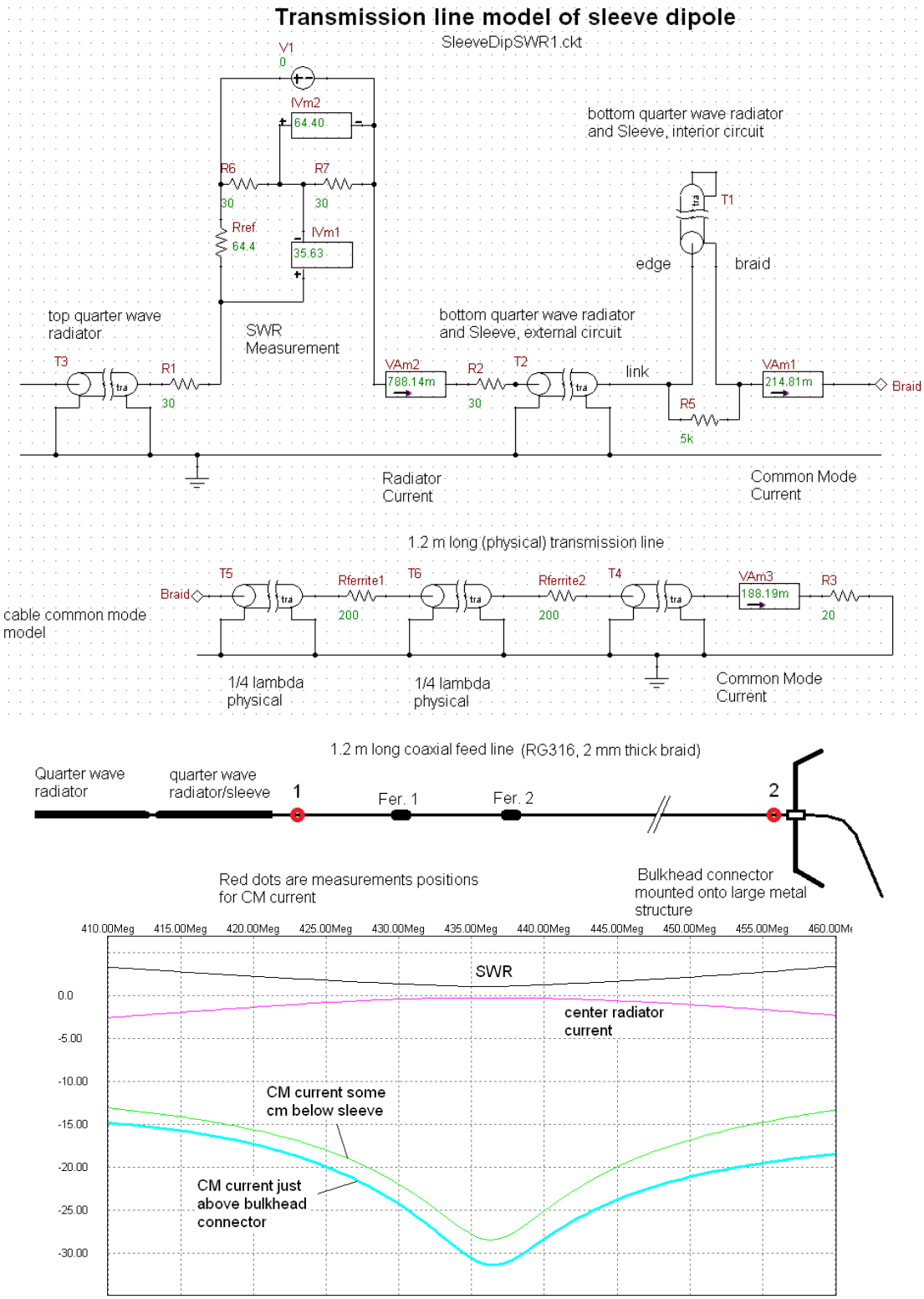


Figure 2.7 Strategy to mitigate inconvenient common mode impedance

2.5. Full EM simulation result

The structure from figure 2.7 is actually simulated in a full EM simulator. Though the sleeve has very high common mode impedance (corrected for fringing capacitance), the CM suppression is limited. This is due to the more or less radiation coupling towards the braid.

Figure 2.8 shows the EM simulation results for 1.2 m long coaxial cable.

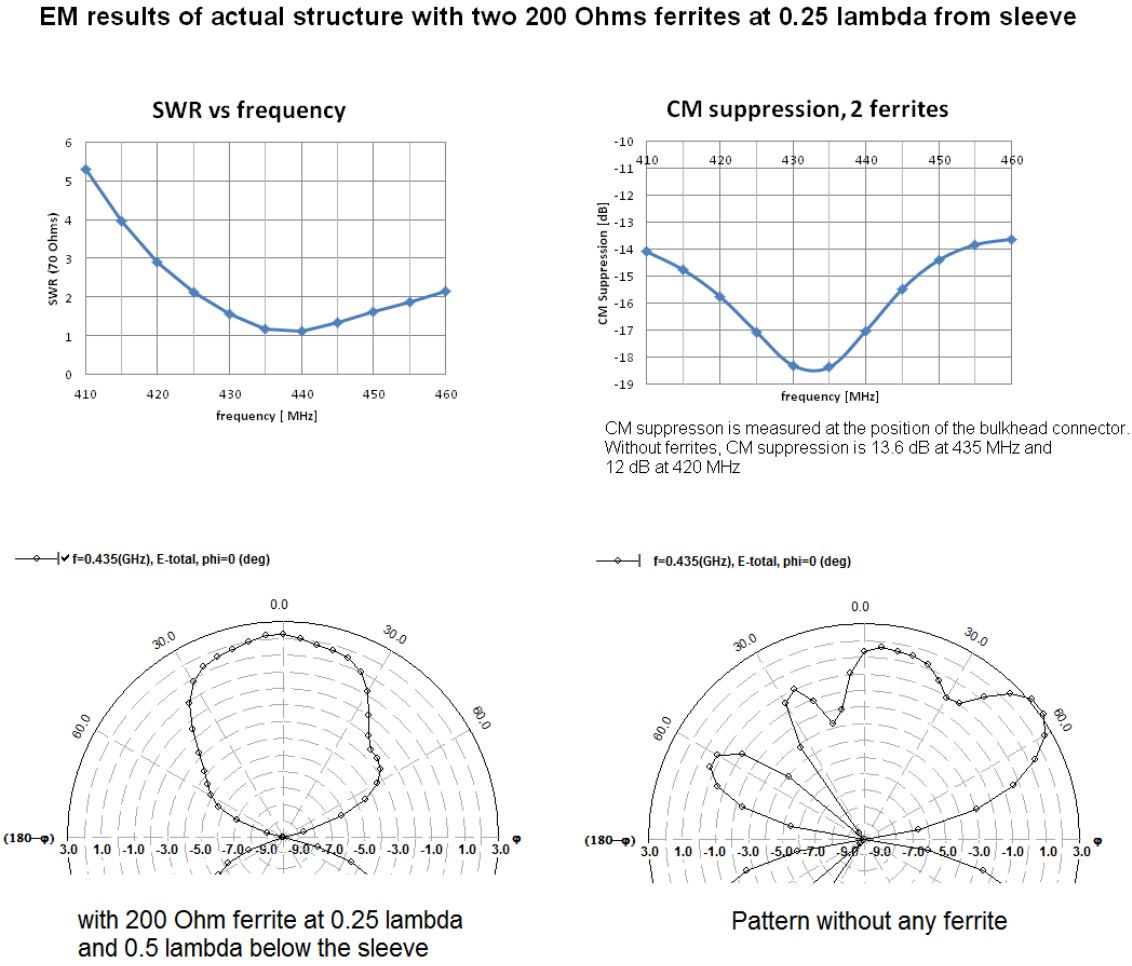


Figure 2.8 EM simulation results for sleeve dipole with 1.2 m cable.

SWR

When looking to the SWR results (70 Ohms), the resonant frequency is somewhat above 435 MHz (center of ITU region 1, 70 cm band).

Common Mode Suppression

Though the ferrites are present, >20 dB CM suppression is not met. Spice simulation gives significant "better" results. The reason is the more or less radiation coupling.

The CM current through the bulkhead connector is significantly higher then the current in the two quarter wave sections in between the ferrites and the sleeve. Even removing a quarter wave section completely (that creates a large gap) gives slight improvement only. So the CM current is induced due to “jumping” of EM fields across the quarter wave sections with ferrites.

The reason for the excessive coupling via the evanescent field is in the length of the cable between the most right ferrite and the bulkhead connector. it is near 1.25 lambda long. This gives strong resonance. Changing the length of the remaining 860 mm of cable does significantly increase the common mode suppression (in the 30 dB range). The same effect occurs when adding a third ferrite. This creates another 170 mm distance between dipole and remaining cable, and the remaining cable will be in near anti-resonance conditions.

One should be aware of undesired common mode resonance in cabling or metallic structures nearby antennas. Modeling the radiation effect in PSPICE is hard to impossible.

Radiation pattern
Despite the moderate CM suppression, the antenna radiation pattern including the cabling is very useful. The gain is 2.1 dB so very close to that of a half wave dipole in free space.

When removing the ferrites, significant standing waves are present along the 1.2 m cable. CM suppression drops to 12 dB and the radiation pattern is heavily distorted. Gain dropped to 1.4 dB. It should be noted that when varying the cable length a gain above 2.14 dBi may likely be possible. However this is an unreliable operational situation.

2.6. Conclusion

The sleeve dipole can be an option as a vertical omnidirectional antenna, but there are some things to consider.

- The relatively thin sleeve dipole has about 60…70 Ohms impedance. Impedance matching may be necessary to get optimum match and largest useful bandwidth.
- The useful bandwidth of the sleeve choke is generally below that of the dipole itself. Sufficient suppression is possible over a relative small frequency band, unless the sleeve is very thick and D_{sleeve}/D_{braid} is relatively large. Due to the sleeve, the SWR bandwidth of the dipole reduces also.
- Several designs fold the braid of the coax back onto itself. The characteristic impedance of the braid that is folded back over itself is generally in the 10 Ohms range. This is because of the small Dsleeve/Dbraid ratio and the high dielectric fill factor. Useful bandwidth will be in the 0.5% range for this arrangement.
- From theory a dipole doesn’t generate a radiation field at large distance along its major axis. However there is an evanescent field that transfers energy in the feed line, even at distance of about 0.5 lambda.

- Though common mode suppression of the sleeve dipole is reasonable, additional measures are required to mitigate standing wave / resonance effects. Without this, situations may occur where CM suppression drops below 10 dB. This is not desired for many applications as the CM current will interfere heavily with the dipole radiation. Read this as: significant pattern distortion and gain loss.
- The effect of common mode resonance and/or standing waves can be mitigated by adding chokes at quarter wave distance. Chokes with losses (that is small Im(Z)/Re(Z) ratio) have preference to avoid any common mode resonance near the dipole. Choking at distance beyond 0.5 lambda down the dipole may be required when guaranteed >20 dB CM suppression is required.

Though not discussed here in particular, the sleeve itself can also be used as a common mode choke (bazooka style), for example for a half wave end-fed antenna.

Are these common mode issues a specialty for sleeve dipoles? No, all antennas that generate common mode current (End-fed antennas, center fed dipoles without balun, quarter wave GPA, etc) may have common mode issues. Mostly they go unnoticed as modern systems have lots of margin in the link budget. When using antennas for measurement, CM behavior is of importance but it is frequently overlooked.

3. Implementation, dipole design

3.1. Introduction

We can do almost everything in simulation now. Just let an optimizer change some measures and let it run and see the results shortly after. However there will be discrepancy between simulation and real world. Perhaps the exact geometry can't be put in simulation, simulation is not 100% accurate, or accurate material data isn't available. So experimentation and measurements are required to get a useful result.

Just an example
The useful bandwidth of the example sleeve dipole is in the 15 MHz range at 435 MHz. From the center frequency to the edge of the useful bandwidth is about 7.5 MHz, that is just 1.7%. This equals a change of 3 mm for the sleeve.

Another example
We can measure CM current in simulation very easy. There will be no interaction with the circuit. How do we measure the CM current in real world at RF? This isn't a trivial task. Common mode termination of the feed line has strong influence on CM current, and is frequency dependent. In addition reflected radiation from the dipole may reach the feed line again and introduce CM current. The cable that runs from the probe toward the spectrum analyzer or detector may affect circuit behavior. The sensor may also be sensitive for voltage.

The most demanding task when building sleeve dipoles is finding the sleeve length for minimum common mode current. Once you have it, you can just experiment with the top section of the dipole based on SWR measurement.

3.2. Determining optimum sleeve length

The first step is to correct (that is reduce) the quarter wave length because of dielectric inside the sleeve. Use the procedure from figure 2.3.

Very likely there are some spacers to keep the feed line centered. Use materials that do not absorb moisture. As spacers occupy only a part of the air inside the sleeve, you may use materials that you would normally not use for High-Q RF application. If you don't want to do the math, use PE, PP or other moderate to low loss material. Thin discs of FR4 are also an option.

When using PP, or PE discs, evenly distributed along the sleeve length, reduce the sleeve length with 33% of total spacer thickness. So when you put the spacers on top of each other and the thickness is 10 mm, you need to remove 3.3 mm from the length of the sleeve.

When having a first guess for the length of the sleeve, it is time to measure.

The measurement should mimic the actual situation as much as possible. It should not interfere with the fields near the sleeve / braid transition. Figure 3.1 shows a measurement setup that will do the job, but is troublesome to use.

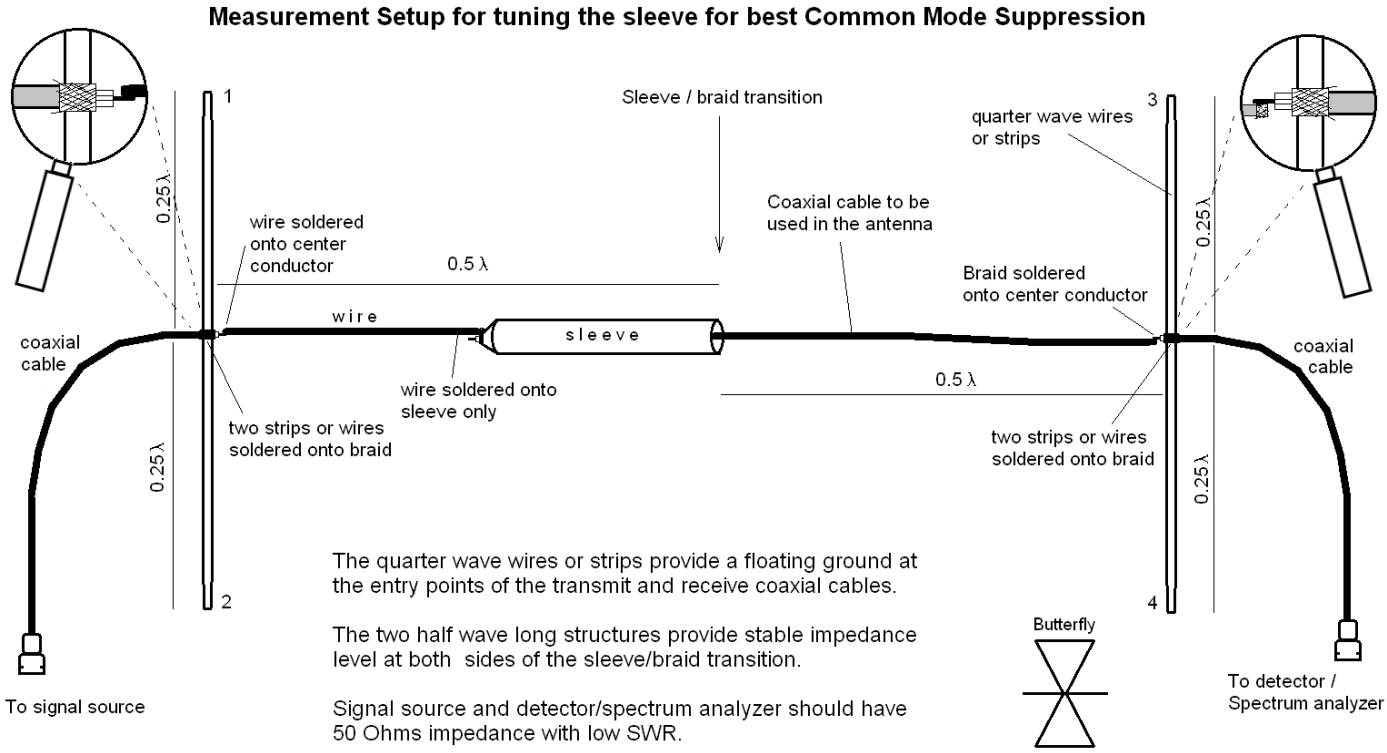


Figure 3.1 Measurement Setup for tuning the sleeve

The goal of this test setup is to provide relative low impedance on both sides of the sleeve/braid transition. This avoids undesired resonance due to the reactive impedance behavior of the sleeve/braid transition.

A scalar relative output/input ratio measurement is carried out to find the resonant frequency of the sleeve. The resonant frequency appears as a minimum (dip) in the transfer. As the sleeve is in its actual environment, no correction is required afterwards. All fringe/stray effects are taken into account during this measurement. When the dip is below the desired frequency, the sleeve is too long.

It is important that both sleeve and coaxial cable inside the sleeve represent the actual situation. Also spacers should be placed before measurement. When a radome is used it is advised to apply a piece of radome across the sleeve/braid transition as there will be some influence due to fringe field lines. It is not desired to cover the complete sleeve.

Operation of the measurement setup
RF signal is applied at the left side. The left open end of the coaxial cable is provided with a low impedance floating ground via two quarter wave strips. Then the signal travels across the wire towards the left side of the sleeve. The wire is soldered onto the sleeve. The center conductor of the coaxial feed line that runs

inside sleeve is not used. The braid of the coaxial cable that runs inside the sleeve is connected to the tapered (left side) of the braid, similar to the actual situation.

The signal travels further to the right via the sleeve. Due to the half wave overall length from RF injection point to the sleeve/braid transition, relative low impedance is maintained. The signal that crosses the sleeve/braid transition travels along the braid of the coaxial cable towards the receiving end at the right side.

It then enters the receive coaxial cable. As the length of the braid is 0.5 lambda, the sleeve/braid transition sees low impedance (50 Ohms). Two 0.25 lambda wires (3 and 4) or strips provide a good floating ground for the receive coaxial cable.

Some notes:

- When having a relative thick sleeve, the wire should also be relatively thick to avoid large impedance step at the wire/sleeve transition. A 3:1 ratio is acceptable.
- The wavelengths in the figure are electrical lengths, so you need to correct for insulation for single wire transmission line using the procedure in figure 2.3. Use $D = 0.25 \cdot \lambda$ as outer diameter (D_{sleeve} in figure 2.3), though there is no sleeve. As an example, insulation on VD 2.5 mm^2 installation wire has Vf = 0.95 at 400 MHz.
- When relative wide band measurement is carried out (say $f_{max}/f_{min} > 1.2$) it is better to change to have 4 quarter wave wires on each side, or use the butterfly. This provides low impedance over a wider frequency band. Another option is to divert to the alternative setup (paragraph 3.3).
- To increase the bandwidth of the measurement setup, one may add 150 Ohms resistors in series with the coaxial input and output (at the positions of the magnifying glasses).
- The measurement setup should be as free as possible in air. You may use polystyrene foam (best option) or single layer very dry corrugated cardboard as mechanical support. Hanging the structure (vertically) using thin rope is also possible.
- Move the complete measurement setup over about 0.1..0.25 lambda in three dimensions and check whether spread in measurement results is acceptable. Metallic structures should be at least several wavelength away from the test setup.

3.3. Determining optimum sleeve length, alternative

There are some practical issues with the setup shown in figure 3.1.

- The test setup from figure 3.1 is fully open and therefore susceptible for nearby objects. Outdoor measurement is almost mandatory
- Due to the more or less radiation coupling across the sleeve / braid transition, the dip in the transfer curve is wide and not very deep.

The two properties are observed during practical measurement. The spread in subsequent measurements with environmental changes is in the several MHz range. Simulation shows similar behavior, see figure 3.2A.

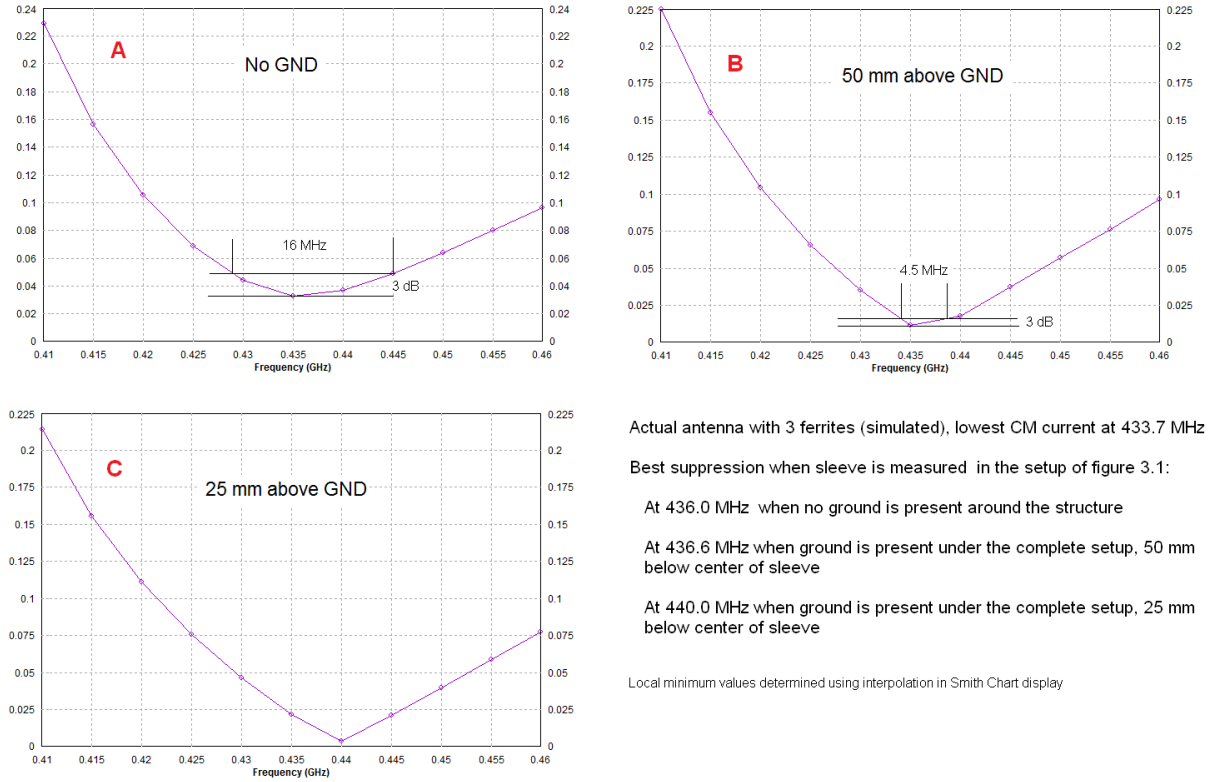


Figure 3.2 Effect on transfer function of ground below the structure

Though theoretically the setup from figure 3.1 is good, practically it is more or less troublesome. Luckily there is a way to reduce the uncertainty. It introduces some systematic error, but reduces random errors much more then the added systematic error.

A large ground plane is added below the structure.

- It reduces the radiation from the structure, hence influence from nearby objects reduces significantly
- It modifies the field around the sleeve / braid transition, so it will introduce a systematic error
- It reduces the radiation coupling between the structure left and right of the sleeve / braid transition. The dip drops further and its width reduces.

The systematic error
When the structure from figure 3.1 is 50 mm above the ground plane, the influence is +600 kHz compared to no ground at all. The width of the dip in the transfer curve reduces significantly making measurement easier. When going down to 25 mm above the ground plane, the systematic error rises rapidly (+4 MHz).

The sleeve used in these simulations is placed into an antenna in simulation with 3 ferrites. Then the best CM suppression is at 433.7 MHz.

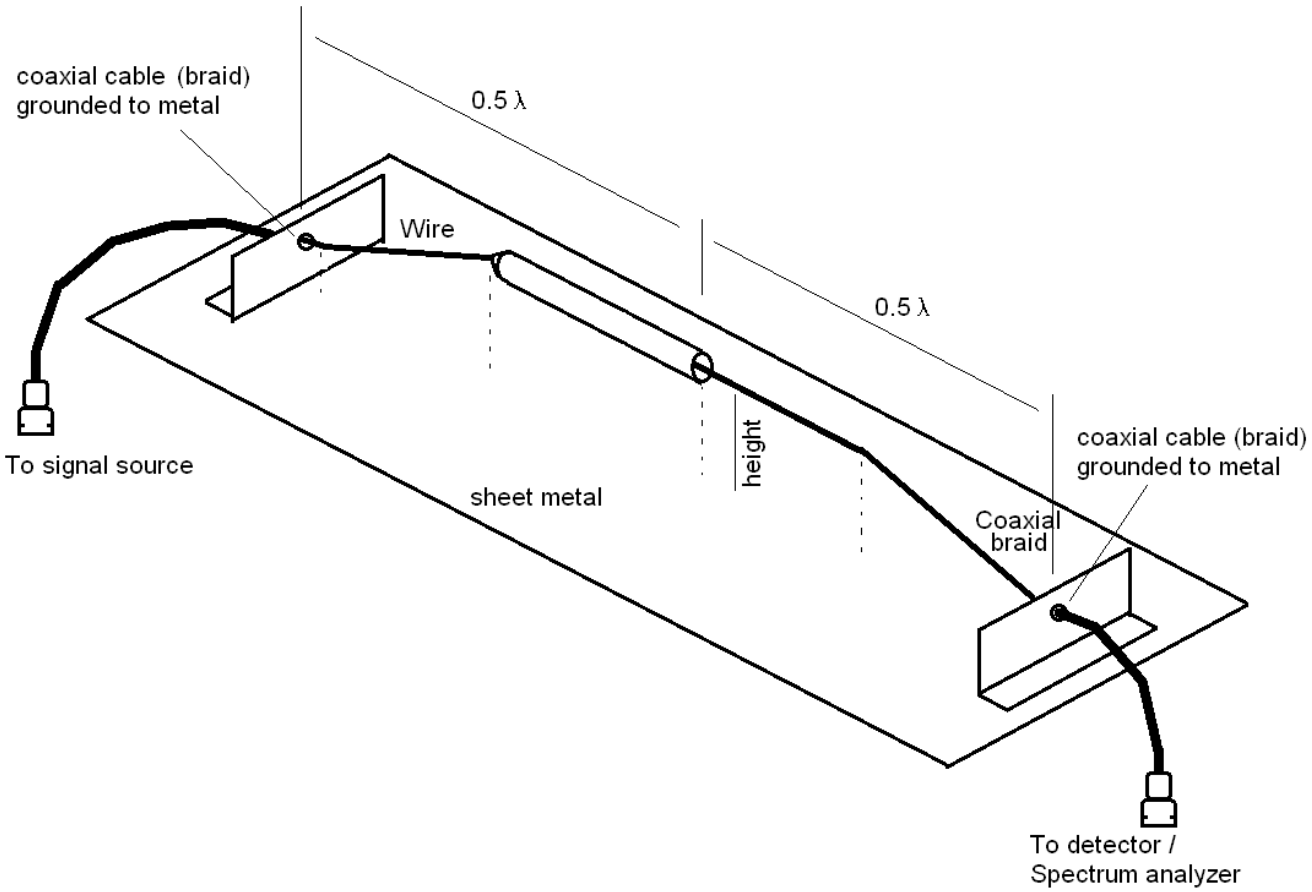


Figure 3.3 Modified Measurement Setup for Sleeve tuning

Removal of the quarter wave wires
As the structure is now relative close to a large ground plane, on may feed the structure referenced to that ground plane. The wire that runs from the left side to the sleeve may slope. Some part of the coaxial braid that runs to the right (spectrum analyzer) may slope also.

A practical setup is shown in figure 3.3. One may use bulkhead connectors at the vertical metal plates for easy removal of cables. The height should be varied to find a good compromise between easy detection of the dip in the transfer curve and systematic error due to modification of the field near the sleeve / braid transition. For 400 MHz, 50 mm can be used.

The L-shaped profile supporting the coaxial cables for signal input and spectrum analyzer may be clamped to the ground plane with magnets. Mostly the capacitive coupling trough a coating layer is sufficient, so removal of coating is mostly not required.

The pictures in figure 3.4 show the measurement setup with ground plane (still with the quarter wave floating ground wires/strips) with the 435 MHz sleeve.

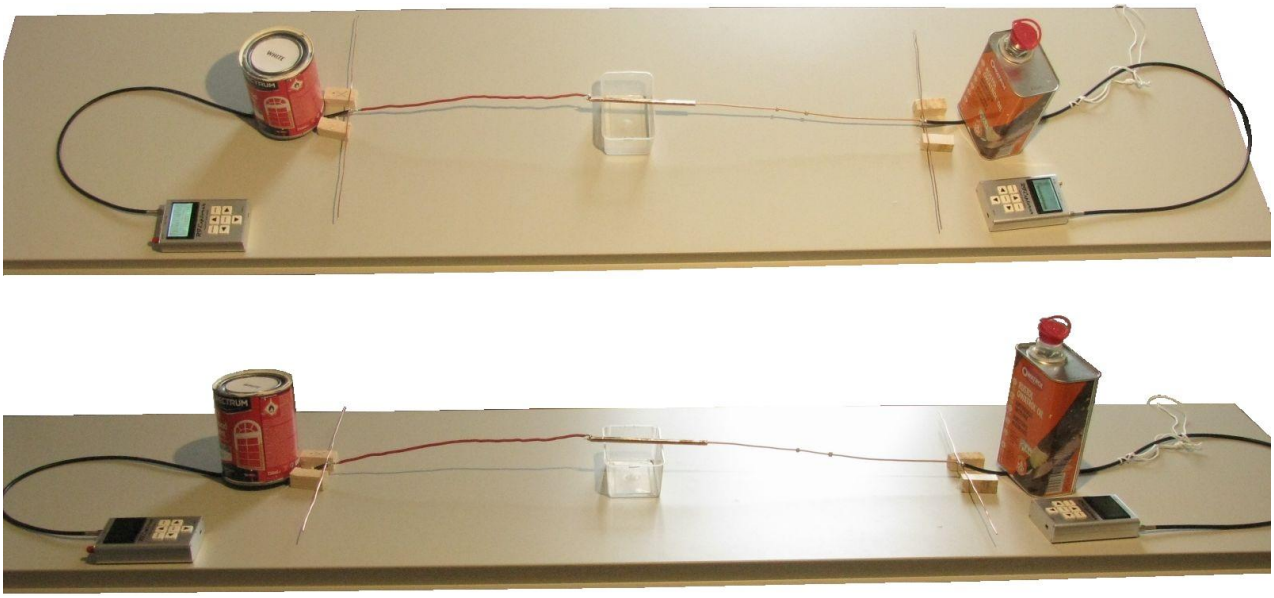


Figure 3.4 Photos of setup with ground and sloping wires

At the resonance dip (440 MHz) the detector/spectrum analyzer reads about -70 dBm with -40 dBm input. When adding the fishing rod material at the sleeve/braid transition, the resonance dip moved to 436 MHz. When using a modern (thin wall) fishing rod, the shift will be less. It is therefore good to have at least the sleeve/braid transition covered with the radome material over about 4*(sleeve diameter), both left and right of the transition.

The internals of the sleeve are shown in figure 3.5.
Sleeve thickness is 6.5 mm, made from copper foil. Length (inclusive tapered part) is 149 mm using RG316 2.5 mm thick coaxial feed line. Maximum attenuation is at 440 MHz, but when the sleeve is inside the fishing rod, this drops to 435 MHz. This is

because the material from the rod is somewhat in the fringe field between the edge of the sleeve and the braid of the feed line that goes through the sleeve.

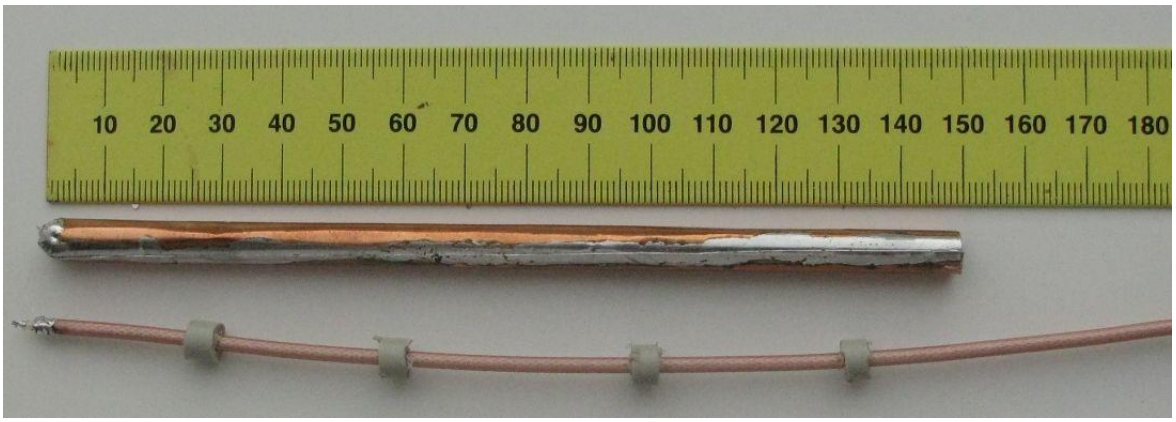


Figure 3.5 Photo of Sleeve construction

3.4. Determining sleeve length using CM current

Paragraph 3.2 and 3.3 describe measurement setups to tune the sleeve for best performance. It is possible to tune the sleeve in an actual antenna + cable situation (but not recommended).

The sleeve was tuned in a “live” situation prior to putting it in the setups describes in paragraph 3.2 and 3.3. The measurement setups were developed for a reason: Tuning in an actual antenna setting is really troublesome.

One would think: Why using such difficult measurement setups?

1. Just add some current probe around the cable and do a frequency sweep using a (scalar) VNA, or a separate signal source with detector.
2. Just grab the cable and move your hand up and down along the cable and watch variation of SWR or S11.

Figure 4.4 is used to explain why just simple measuring the braid current with a probe and a frequency sweep has severe shortcomings.

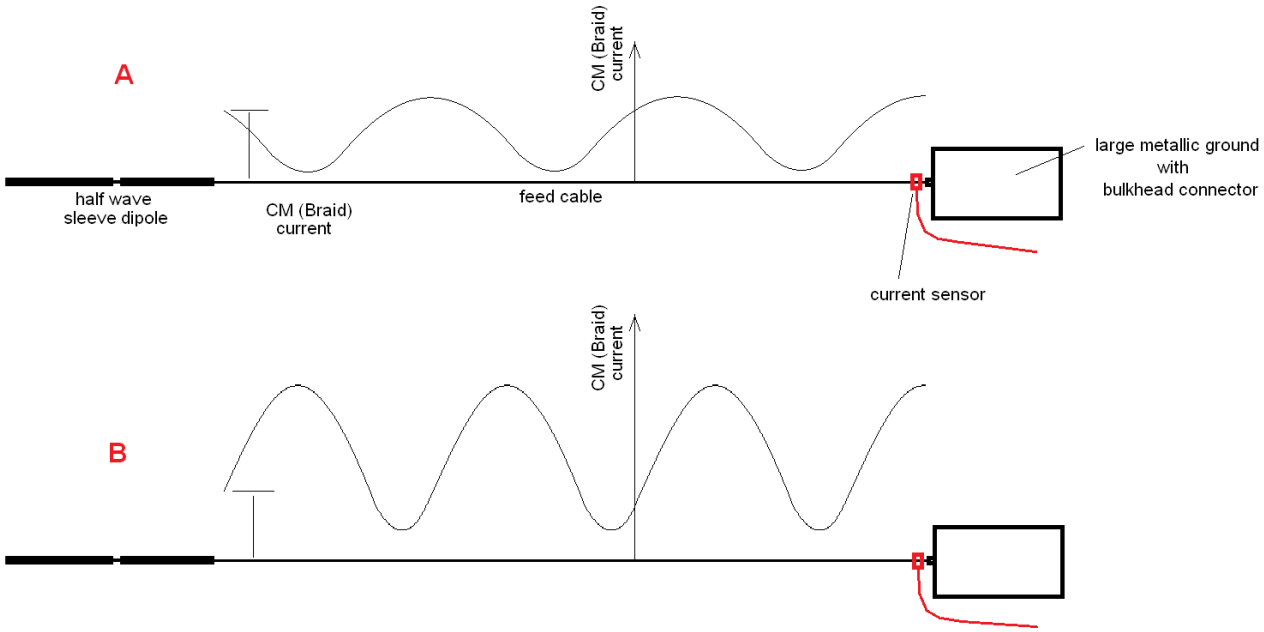


Figure 4.4. Current amplitude along the braid of a feed line

Figure 4.4A shows a sleeve dipole with a feed line The antenna is fed via the bulkhead connector. The bulkhead connector is mounted on a large metal part. Note that in real world the setup is used vertically. The feed line continues to the signal source via the bulkhead connector, but is not shown in the picture.

A current probe/sensor is present around the cable just above (here left of) the bulkhead connector. The oscillating line shows the Common Mode current amplitude along the line. Due to the low impedance behavior of the large metallic ground a strong standing wave pattern is present along the braid of the feed line. A current maximum is present at the bulkhead connector because of the low CM impedance. When moving the current sensor, the standing wave pattern can actually be measured.

In figure 4.4A, the CM current amplitude at the current sensor is a bit higher compared to the CM current just below (right of) the sleeve.

Now we raise the test frequency. This reduces the distance between CM current maximums as wavelength reduces. Figure 4.4.B shows the new situation. Though the CM current just below (right of) the sleeve is the same as in the A figure, the CM current detected by the sensor is almost twice as high! Therefore changing frequency changes the ratio between CM current just below the Sleeve and the measured CM current.

There are more drawbacks.

- When changing frequency, the common mode impedance as seen from the sleeve into the feed line changes. As we know, this can lead to an undesired good match between sleeve impedance and CM impedance.
- When using a thin upper quarter wave radiator, you need to check SWR to avoid too much change of net input power.

- Due to some form of radiation coupling between the sleeve and the feed line, the “dip” is relatively wide (see paragraph 3.3, figure3.2) and not very deep. Nearby metallic or dielectric objects (masonry, concrete, etc) will increase uncertainty.

Can we make this setup better? Yes!

1. By providing matched Common Mode impedance at the bulkhead so that there is a predominant travelling wave pattern along the braid. You need to check the current along the line to verify that there is a dominant traveling wave.
2. By keeping the frequency constant and change the sleeve length to find the lowest common mode current. Make sure to have a current maximum at the sleeve – cable transition to avoid undesired resonance. You also need to extend the upper quarter wave radiator when reducing the length of the sleeve.

Moving your hand along the cable and watching SWR.
It is not a joke. At UHF frequencies cable length can be several wavelengths When you grab the cable and move your hand up and down, you change significantly the CM impedance as seen from the sleeve into the braid. Due to the coupling between the sleeve and the braid, this will change SWR (or S11).

Best method is to change the sleeve length instead of frequency and notice the trend of SWR/S11 change during moving your hand over > 1 lambda.

It is definitely not recommended, but when no other measurement equipment is available, it may result in a working antenna.

3.5. Further reduction of common mode current

The free space common mode impedance of 2 mm braid is in the 250 Ohms range at 450 MHz. The CM impedance just below the sleeve depends however on the (metallic) structures near or connected to the cable. To avoid unexpected behavior it is advised to add additional CM filtering.

One may add a sleeve upside down with the open side 0.25 lambda below the sleeve of the dipole. One can also use suitable ferrite. The impedance should be above the CM impedance of the coaxial cable.

For the 435 MHz design, 5 fair-rite material 43, 12.7 mm long, 6.35 mm thick ferrite sleeves are used (order number 2643250402). They provide 140 Ohms impedance. They are not the best for this application, but they are remaining stock from an 868 MHz antenna design, and they fit in the fishing rod.

The lowest impedance at 435 MHz is 150 mm above the top ferrite. This is measured by measuring the standing wave pattern along the braid using the setup of figure 3.4 (but without the sleeve). So the first ferrite is 150 mm below the edge of the sleeve.

Below the 5 sleeves the RG316 is connected to H155 5 mm foam dielectric cable because of the significantly lower loss compared to RG316.

A picture of the complete dipole with ferrite is shown in figure 3.6.

3.6. Making a dipole

When you don't add a matching circuit, you can expect close to 70 Ohms, so this will give some mismatch. For the top section the thickest round tube is used that just fit inside the fishing rod. The tube is made out of 0.05mm copper foil. This increases bandwidth so the antenna can be used over the 430 to 440 MHz band without further matching with $VSWR_{50\_Ohms} < 2$.
The photo below shows (from left to right)

- H155 cable (Foam dielectric)
- 5, 6.35 mm Fair-rite mat 43 ferrite sleeves (order number 2643250402)
- RG316 2.5 mm cable with spacer
- Sleeve / lower half of radiator, 150 mm above ferrite
- Top half of radiator forming a half wave dipole.

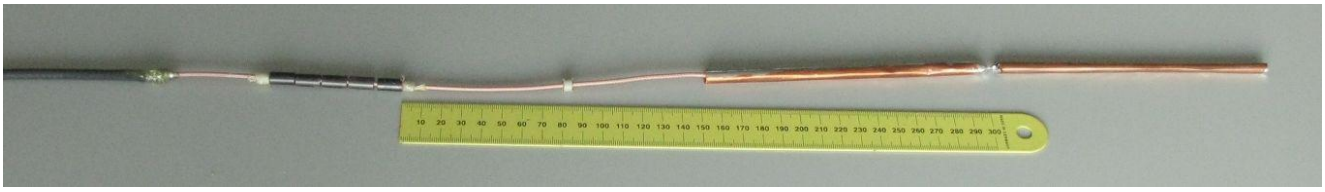


Figure 3.6 Photo of complete dipole with ferrite

The 5 ferrite sleeves are fixed using two, 5 turn barrel knots. Depending on wire thickness, you may use a standard whipping knot. These knots give sufficient friction, without compressing the cable. They are impregnated with solvent glue (contact adhesive), or other adhesive that sticks to the sheathing material. The glue should also be on the cable just below and above the knots This will avoid any movement of the 5 ferrites.

You may notice that the upper half of the radiator is not as thick as the sleeve. The upper half radiator has a thickness of about 5 mm. 6 mm doesn't fit into the fishing pole.

The dipole has about 70 Ohms impedance at resonance. This can be matched to 50 Ohms. See the matching paragraph in chapter 4 for (some) guidance.

4. Implementing a 4 element Yagi antenna

This chapter “converts” the vertical dipole into a vertical Yagi as shown below

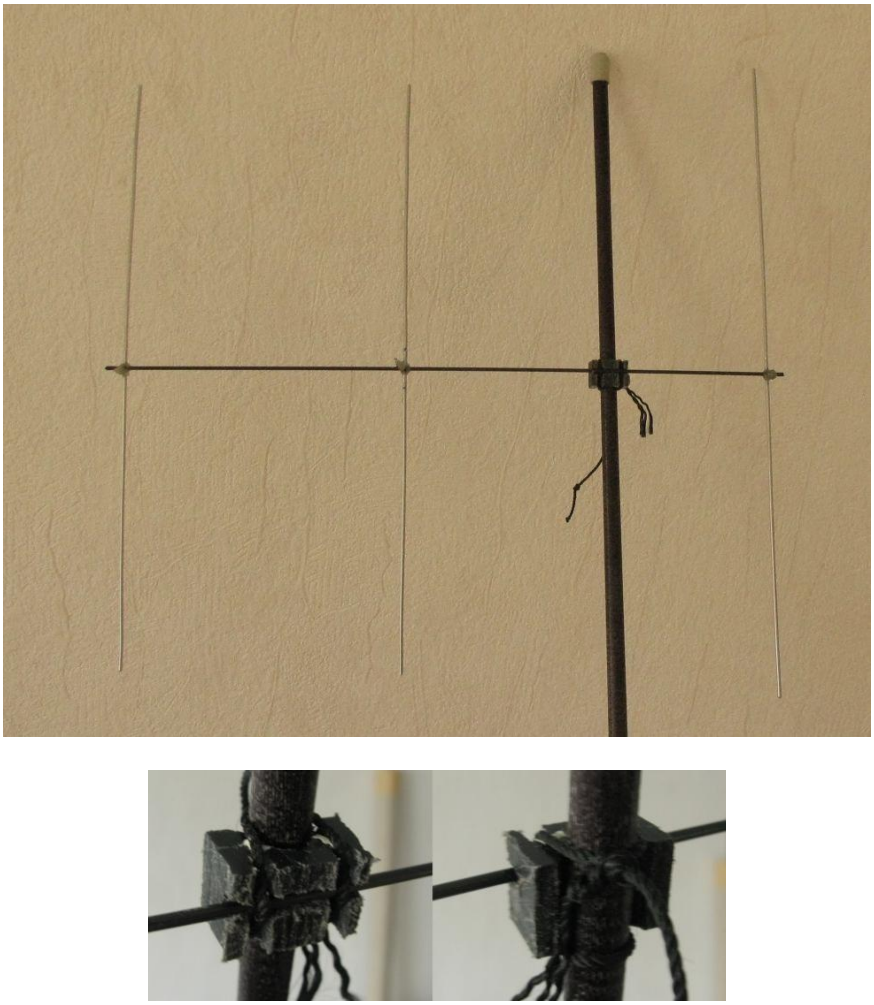


Figure 4.1 Photo of Yagi mounted onto the sleeve dipole (inside fishing pole)

Some construction tips

- The boom (350 mm) is made of 2 mm carbon fiber. You may also use 3 mm glass fiber. This will give similar stiffness.
- The elements are made of 2 mm hard drawn aluminum wire. You may also use stainless steel, or brass, but this will increase the weight.
- The elements are mounted onto the boom using square lashings and are impregnated afterwards with thick epoxy resin/glue. A paint layer is added for UV-resistance.

- The antenna is mounted onto the boom with a cross adapter with a simple lashing using a friction knot. For permanent fixing you may use a square lashing with (epoxy) impregnation, so refresh your Scouting pioneering skills.

Do not make the square lashings very tight as that my damage the materials. The rigidity comes from the epoxy impregnation. You may also use several layers of paint instead, but curing time for good rigidity takes longer.

When using epoxy, you need some paint layers for UV-protection. Most epoxy resins have poor UV-resistance.

When using a conductive boom (such as carbon), the boom should be exactly at the feed point of the dipole, otherwise RF current is induced in the boom. When the sleeve is significantly short compared to a quarter wave length, a conductive boom has to be placed just above the feed point. Best option is to base the boom placement on change of SWR. When doing this, do not mount the yagi elements.

4.1. Yagi design

Warning:

Once the sleeve has been tuned for best CM suppression, do not change the length of the sleeve anymore. For tuning or matching you only change the length of the top quarter wave radiator.

The Yagi design is a 4 elements DK7ZB 28 Ohms Yagi, providing 7.2 dBd (9.5 dBi). There is one difference; the elements are out of 2 mm aluminum (except for the radiator as that is the Sleeve dipole). Therefore the element lengths have to be corrected. After some EM simulation:

element	Distance [mm]	Length [mm]
Reflector	80	344
Radiator	0	tbd
Director 1	110	311
Director 2	140	303

Distances are in between the elements (center-center) overall boom length = 330 mm. Add 10 mm excess length for mounting, so start with a boom of 350 mm long.

The boom is a 2 mm carbon fiber rod. You may use 3 mm glass fiber instead.

Note:

For people that want to design an end-fed radiator instead of a sleeve dipole, one needs to “tune” the length of the radiator based on Common Mode feed line current (see paragraph 3.4). End-fed impedance at lowest common mode current (real part of parallel equivalent circuit) is almost twice as high as value based on the formula in the annex.

When you don't feel confident with matching the dipole to work properly in a 28 Ohms design, use the 50 Ohms design (6 dBd) instead:

element	Distance [mm]	Length [mm]
Reflector	83	347
Radiator	0	tbd
Director 1	32	320
Director 2	137	297

Distances are in between the elements (c-c) overall boom length = 252 mm. Add 10 mm excess length for mounting, so start with a boom of 272 mm long.

Tuning is now done by changing the upper radiator only. Though the gain is somewhat less, the radiation pattern is good to suppress co-channel effect when using Single Frequency Network repeaters (for example the Dutch PI2NOS repeater network).

4.2. Matching the 28 Ohms yagi design

As the radiator for this DK7ZB design has 28 Ohms resonant impedance, matching is required. Due to the space, the number of options is limited.

Formulas and basic matching circuit are shown in figure 4.2.

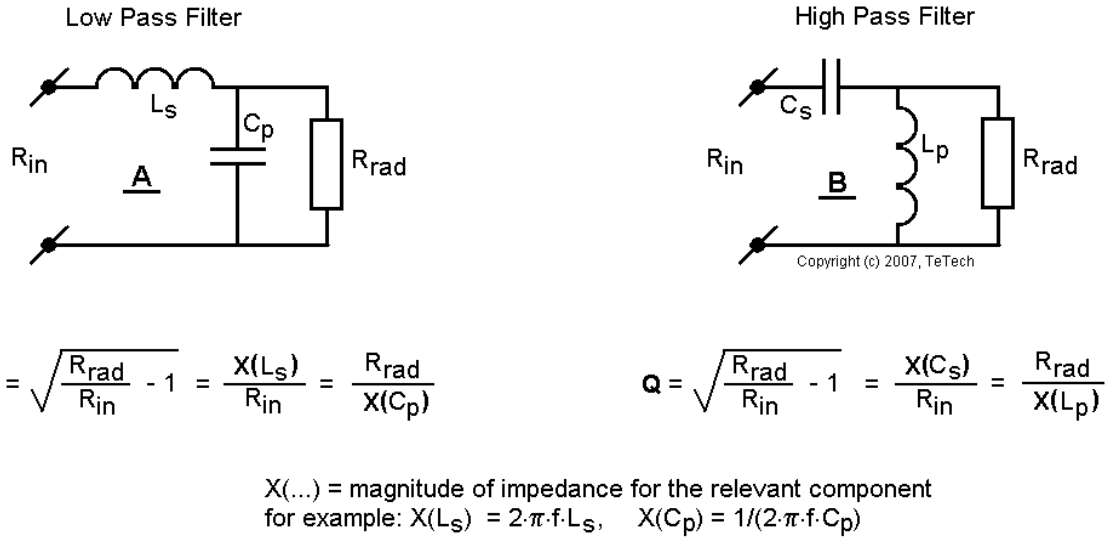


Figure 4.2 Circuit and formulas for matching

Figure A shows the low pass version, Figure B shows the high pass version.

You can only transform to a higher impedance, so Rrad > Rin. When you want to step down, you need to use the circuits in reverse. The Numerator of the fraction must always be larger then the denominator. That means $R_{rad}/R_{in} > 1$.

4.2.1. Using the dipole as omni-directional radiator

Rrad = 70 Ohms, Rin = 50 Ohms.

Q = rt(70/50 – 1) = 0.63.

Reactance of parallel component: 70/0.63 = 110 Ohms (3.3 pF)
Reactance of series component = 50*0.63 = 32 Ohms (11 nH)

Experimentation is necessary to build the matching circuit.

One option is to use a series inductor between the center conductor of the coaxial cable and the upper quarter wave radiator and an external capacitor. The external capacitor is copper foil or tight fitting tube that goes onto the outside of the fishing rod. It should overlap partly the top and bottom (sleeve) section of the half wave radiator.

As mentioned, this requires experimentation. It will not be discussed further as focus is onto the sleeve dipole fed yagi.

4.2.2. Matching the Yagi

When you change the top resonator length to get real impedance (resonance), it will show 28 Ohms to the coaxial feed line (SWR just under 2, after correction for feed cable loss). We need 50 Ohms. Therefore the matching circuits (figure 4.2) have to be used the other way around. I prefer an inductor parallel to the dipole (high pass version B). This reduces coupling to nearby 2 m and HF antennas.

Rrad = 50 Ohms (cable side), Rin = 28 Ohms (antenna side)

Q = rt(50/28 – 1) = 0.89.

Reactance of parallel component: 50/0.89 = 56 Ohms (6.5 pF or 20 nH)
Reactance of series component = 28*0.89 = 25 Ohms (9 nH or 15 pF)

The series component (that is in series with the antenna) is the easy one. It can be generated by making the upper radiator shorter (gives series capacitance) or by making it longer (gives series inductance).

The parallel component is the difficult one, as space inside the fishing pole is very limited.

As the radiators are tapered, there is some space to make a 20 nH inductor. A single turn of 0.5 mm thick wire with 6 mm diameter gives about 9 nH.

Some experimentation shows that two turns of 1 mm wire around a 5 mm drill bit gives about 20 nH. See the picture below. When space is very limited in your fishing rod, you may go down to a wire diameter of 0.5..0.6 mm.

It is wise to make the taper angle somewhat smaller compared to my tubes. It will give you more space to add the inductor. Pictures of the matching coil are shown in figure 4.3.

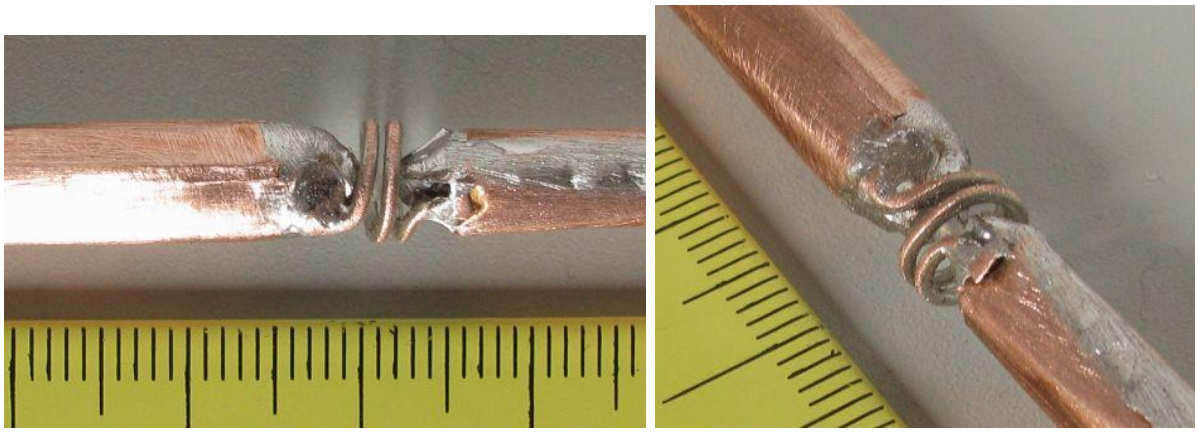


Figure 4.3 Matching coil to transform from 28 to 50 Ohms

The 1 mm thick wire was fully tinned. Some material had to be removed to have a good fit inside the fishing pole. Therefore the bare copper is visible now.

As mentioned, when you don't feel happy doing the matching, you may divert to the 50 Ohms Yagi design. This only requires changing the length of the top quarter wave radiator.

4.3. Using fishing rods and glass fiber masts

Fishing rods and specialized fiberglass poles for antennas (such as Spiderbeam, Sotabeams, DX-Commander, etc), are strong lightweight products, but are brittle. They can't handle point or line pressure well.

You may visit https://tetech.nl/divers/Slim_2m_EndFed.pdf
Chapter 6 gives tips on how to use fiberglass rods and how to (temporary) mount them onto a pipe/mast using a piece of (ply)wood and some rope. Picture 4.5 and 4.6 (in the referenced document) show a procedure on how to make a hairpin spring strain relieve for positioning the antenna inside the rod.

Of course you can use a metal or carbon mast/pole, as long as the top section that contains the antenna + ferrite choke is non-conducting.

You have reached the end of the document!

Hopefully this document will encourage people to build good working sleeve dipoles and or sleeve dipole fed yagi antennas. Instead of a sleeve dipole, a half-wave end-fed can also be used, but is more difficult to build. At upper VHF and UHF sleeve dipoles become attractive as a rigid sleeve is an option. It is rugged and has good power handling capability.

The annex shows an overview of formulas that are used throughout the document.

Response time to comment and questions depend on available time. You can expect days to some weeks depending on professional activities.

Wim Telkamp, PA3DJS

5. Annex

This annex shows several formulas that are useful when designing your own antenna.

Characteristic impedance (Z_{coax}) for coaxial cable. Use same unit for D and d. ε_r is relative dielectric constant of the cable insulation

$$Z_{coax} \approx 138 \cdot \sqrt{\varepsilon_r} \log\left(\frac{D}{d}\right) \quad [Ohm]$$

Characteristic impedance for wires without return conductor (that is the common mode impedance or just CM impedance)

$$Z_{0\_line} \approx 138 \cdot \log\left(\frac{0.25 \cdot \lambda}{d_{wire}}\right) \quad [Ohm]$$

End-fed half wave dipole / monopole input impedance at resonance (based on 60 Ohms center fed impedance). When slightly off-resonance, Z_{Res_HWmono} is the resistive part (Rp) in a parallel equivalent representation of total impedance.

$$Z_{Res\_HWmono} \approx 310 \cdot \log^2\left(\frac{0.25 \cdot \lambda}{d}\right) \quad [Ohm]$$

Value is valid for a dipole only. When used as a radiator in a 28 Ohms yagi, Z_{Res_HWmono} is about factor 1.8 higher at minimum common mode current.

Quality factor of the equivalent series RLC circuit for a center fed half wave dipole. Quality factor of the equivalent parallel RLC circuit for an end fed half wave dipole/monopole. The Q-factor is based on the -3 dB impedance bandwidth. The SWR=2 bandwidth is about 0.71·(impedance bandwidth)
Use same units for wavelength (λ) and wire diameter (d).

$$Q_{HWdip} = Q_{HWmono} \approx 3.6 \cdot \log\left(\frac{0.25 \cdot \lambda}{d}\right)$$

Value is for dipole only. When used in a Yagi, Q-factor will depend on yagi design and impedance behavior no longer matches that of an RLC parallel resonant circuit.

Coaxial capacitor formula. Le is length and has to be in meters. Use same unit for D and d

$$C = \frac{2 \cdot \pi \cdot \varepsilon_0 \cdot \varepsilon_r \cdot le}{Ln(D/d)} \qquad [F,\ As/V]$$

Coaxial capacitance in F/m and inductance in H/m

$$C' = \frac{3.34n \cdot \sqrt{\varepsilon_r}}{Z_0} \quad [F/m,\quad As/Vm] \qquad L' = 3.34n \cdot \sqrt{\varepsilon_r} \cdot Z_0 \quad [H/m,\quad Vs/Am]$$

Impedance (Zshort) of a short-circuited quarter wave line (lossless)
It shows LC parallel circuit resonance near the quarter wave frequency. $\omega_{0.25}$ is the radian frequency where the transmission line shows high impedance resonance, the so-called quarter wave resonance frequency. ω_0 is the LC resonant frequency.

"v" is from German "Verstimmung" or Dutch "verstemming" and is the misalignment factor, amount of detuning or detuning factor. When there is no misalignment (or detuning), v = 0. Large detuning gives large v. A negative tangent means capacitive behavior, positive tangent means inductive behavior.

$$Z_{short} = j \cdot Z_c \cdot \tan(2\pi \cdot le/\lambda) \approx \cdot \frac{4}{\pi} \cdot \frac{Z_c}{jv} \qquad v = \frac{\omega}{\omega_{0.25}} - \frac{\omega_{0.25}}{\omega}$$

Equivalent LC parallel circuit $\omega_0 = \omega_{0.25}$

$$Z_p = \frac{\omega_0 \cdot L_p}{jv} \qquad L_p = \frac{4}{\pi} \cdot \frac{Z_c}{\omega_0} \qquad v = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} \approx \frac{2 \cdot \Delta\omega}{\omega_0}$$

Capacitor values: use $X_L = X_C$ at the resonant frequency. The formulas above are useful for modeling a short-circuited quarter wave line via an LC parallel resonant circuit. A resistor can be added in parallel to model the loss of the transmission line.