Vertical 2 x 5/8 lambda dipole for the 2-meter band

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After setting up a shortwave MagLoop on a balcony of the apartment on the second floor and a shortwave dipole at the end of a boom of the MagLoop standpipe, an antenna for VHF was still missing. The most useful would be a vertically polarized omnidirectional antenna for the 2-meter band. Since the "antenna system" already looked a bit adventurous for non-radio amateurs (my XYL), no further "stick" should be attached, but the VHF antenna should be placed on the boom as inconspicuously as possible.

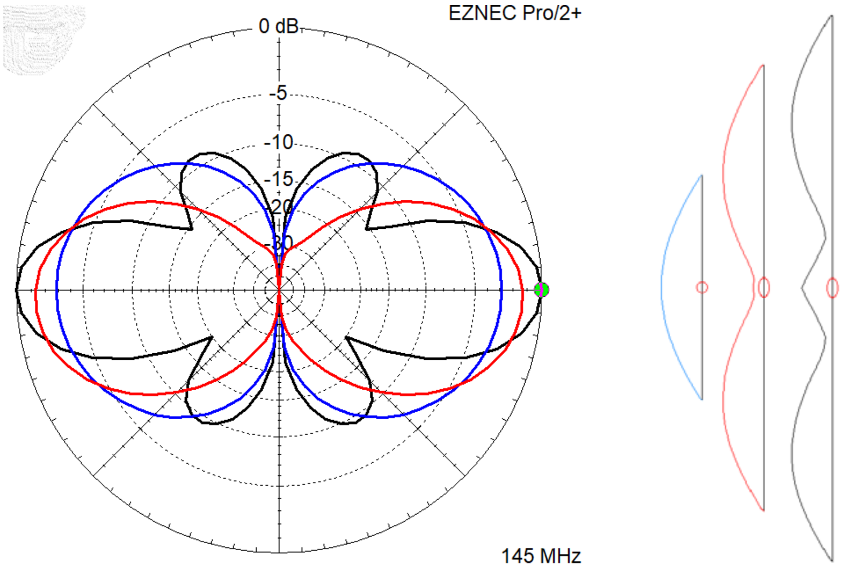
Figure 1: 2 x 5/8 λ (lambda) dipole for VHF between shortwave dipole and shortwave MagLoop

Since a detailed EZNEC /1/ model of the shortwave antennas with their connections and fastenings to the metal balcony railing was already available, /2/, some antenna alternatives could be checked in advance using simulation. It turned out that the 90 cm long boom on the MagLoop's standpipe forms a miserable "counterweight" for monopole antennas such as λ/4 or λ/2 radiators: The terminal current of the radiator must also flow on the boom and from there onto the standpipe of the MagLoop. This creates significant currents on the entire antenna system with current antinodes and current nodes, especially on the horizontal boom and the vertical standpipe, which leads to undesirable radiation in all directions, sometimes also with horizontal polarization, and as a result to a significant loss of directivity in the horizontal plane. Resonant radials, such as the quarter-lambda ground plane, would improve the situation, but would also increase the optical "signature" in an unpleasant way.

However, there is significantly less parasitic current on the antenna system if a symmetrical vertical dipole antenna is placed in front of the boom: Then the lower arm of the dipole forms the counterweight for the upper dipole arm and no current flows directly onto the boom, which is transverse to the dipole. So, primarily currents are only induced in the parts of the antenna system that are parallel to the dipole axis, i.e. into the vertical standpipe of the MagLoop and the shortwave dipole. As a test, a provisional wire dipole in a GRP pipe was tied to the boom and properly received the nearest 2-meter relays. But can there be a little more gain than the (theoretical) 2.15 dBi of a 2 x λ/4 dipole?

It's the length that counts!

With a length of around 1 m, a vertical 2 x λ/4 dipole is still easy to handle; it could easily be twice as long. In Figure 2 you can see a comparison of the radiation patterns of dipoles in free space with different lengths up to a total length of 2.5 m.



(a) (b)

Figure 2: (a) Vertical cuts of the radiation patterns of vertical dipole antennas. Blue=2 x λ/4 (approximately 1 m), red=2 x λ/2 (approximately 2 m) and black=2 x 5/8λ (approximately 2.5 m) total length at 145 MHz. (b) The associated current distributions on the dipole antennas with source in the centre.

The classic 2 x λ/4 dipole has the widest lobe and the lowest gain, the longer versions with narrower lobes have higher gain: the 2 x λ/2 dipole around 1.6 dB and the 2 x 5/8 λ dipole even 2.8 dB more. However, with further extension the gain drops again until a null appears in the horizontal plane at a length of 2 x λ. As has long been known, e.g. Rothammel /3/, the 2 x 5/8 λ length offers the highest gain of a simple dipole, but the terminal impedance of this dipole is much higher than the approximately 70 Ω of the classic dipole, which is fed in the maximum current position, but also lower than the kΩ impedance of the 2 x λ/2 dipole, which is fed at the minimum current position, see

Figure 2b. In the simulation with EZNEC, the 2 x 5/8 λ dipole with a small conductor diameter has a terminal impedance of around 150 Ω with a capacitive reactance of around 600 Ω (corresponding to 1.8 pF) in series; the complex impedance is 150 Ω – j 600 Ω. Actually, this should not be a difficult case for a simple 50 Ω matching network!

Worth a try

The almost 3 dB improvement over the classic dipole is worth the try. Mainly only the reactive component of the antenna impedance would have to be compensated - by a series inductance of around 0.6 μH. The 150 Ω resistance component of the dipole impedance should be easily matched with a classic λ/2 balun with 200 Ω output impedance. The equivalent circuit diagram for this can be seen in Figure 3a.

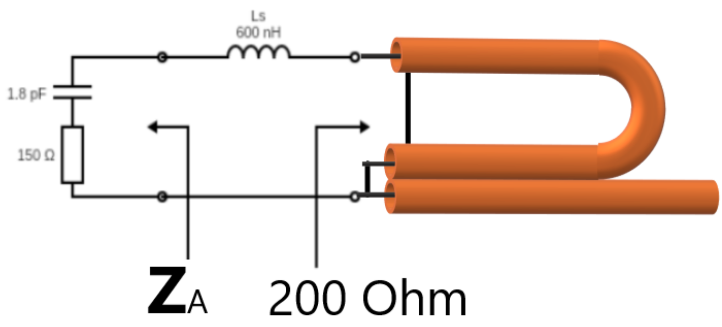


Figure 3a: Equivalent circuit diagram of the dipole impedance Z_A with matching circuit consisting of series inductor Ls and coax cable balun.

For an experimental setup, a plastic housing was used to accommodate the series inductor, the connection cable with balancing element and to fasten the dipole arms in the side walls of the housing. As the upper dipole arm, a 1.25 m long spring wire rod (chrome-plated, diameter 2.5 mm) of a shortwave mobile antenna was screwed into the upper side wall and as the lower dipole arm, a stiff copper wire with a diameter of 2 mm hung downwards from the housing wall.

Disappointing result of measuring the antenna impedance with the NanoVNA: The series inductance needs to be much smaller to compensate for the reactive component of the impedance and the measured effective resistance component is well below 150 Ω.

At first it was not clear why the theory deviated so far from the measurement in this case. To date, EZNEC simulations of “wire antennas” have mostly been close to reality – so what could have been inadequately modeled in the simulation model? The following procedure has proven useful for this question: You increase/decrease the parameters of the simulation model, i.e. dimensions (here positions, lengths and diameters of the dipole arms) and electrical variables (here the inductance of the compensation coil) and see whether you get closer to the measured impedance curve (or reflection coefficient). However, in this case there was no improvement; none of the attempted modifications led anywhere close to the measured dipole impedance curve. The solution only came from taking into account a parameter that is usually neglected in antennas, the stray capacitance of the practical fixture. This capacitance is parallel to the dipole terminals, i.e. parallel to the dipole impedance, as shown in the equivalent circuit diagram, Figure 3b:

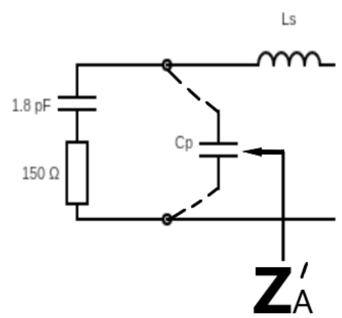


Figure 3b: Realistic dipole impedance Z_A' with stray capacitance Cp.

In VHF antennas with impedances around 50 Ω, stray capacitances hardly play a role, but in this case the shift in antenna impedance can actually be attributed to a small parallel capacitance of only about 1 pF. Figure 4 shows how the parallel capacitance Cp shifts the antenna impedance so that both the resistance and reactance components are reduced: Z_A with R_A and X_A becomes Z_A' with R_A' and X_A'; the larger Cp, the further the down shift.

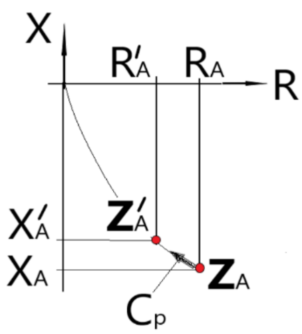


Figure 4: Impedance transformation through parallel capacitance Cp. The horizontal axis shows the resistance, the vertical axis shows the reactance. Illustration not to scale.

The stray capacitance of the implemented structure could perhaps be reduced, but it is better to specifically increase the parallel capacitance: With around 0.7 pF of additional capacitance, the resulting impedance was shifted to 50 Ω - effective resistance and a coil (series inductance) with fewer turns had to be used to match the antenna to the 50 Ω of the coaxial feed line, see Figure 5.

This success can be understood with the help of the EZNEC simulation. Figure 6a first shows the reflection coefficient versus frequency plot for a model of a 2 x 5/8 λ dipole with two series inductors of 0.35 µH each as load elements, which compensate the

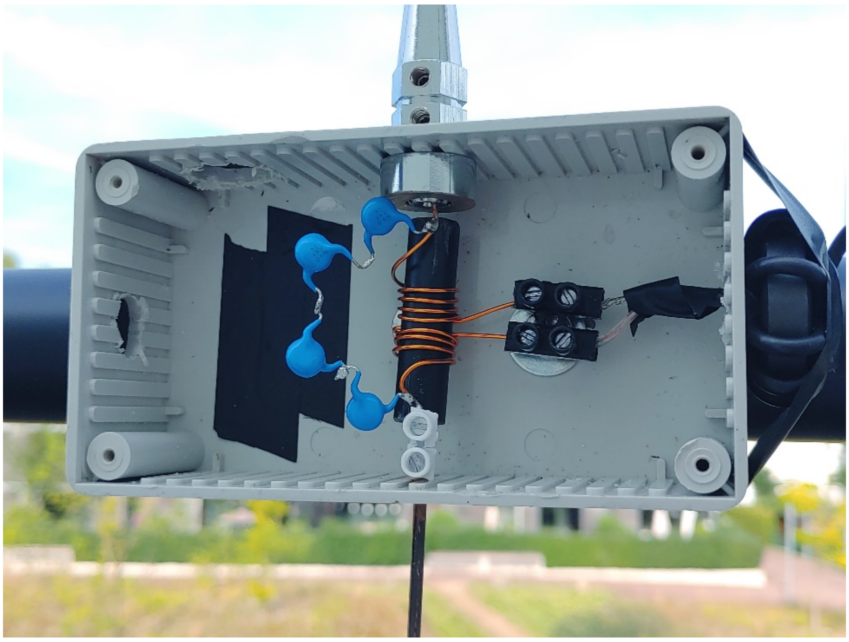
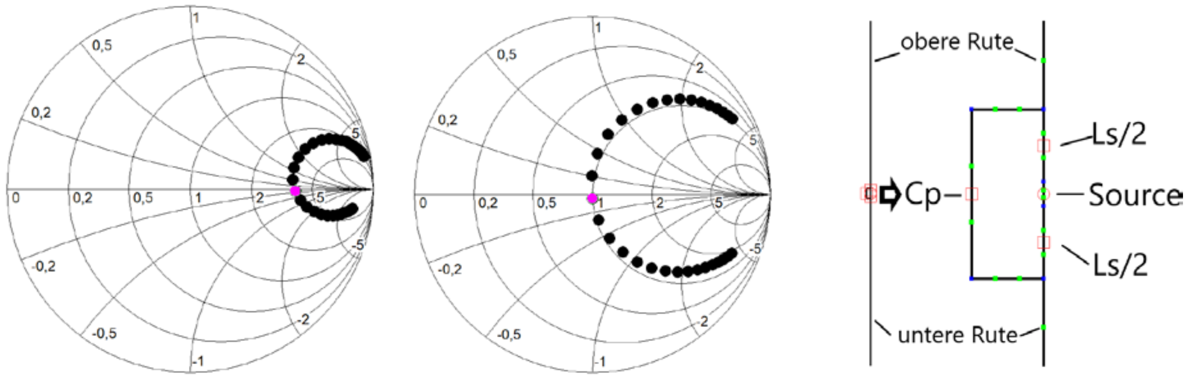


Figure 5: Experimental setup of a 2 x 5/8 λ dipole. Above is the spring wire rod in a 3/8 inch screw connection, below the copper wire 1.25 m long. In the middle is the symmetrically divided coil (series inductance) with additional parallel capacitance made of series-connected 3 pF ceramic capacitors. On the right is the connection of a 50 Ω coax cable with a toroidal core balun outside the plastic housing.

reactance of the dipole impedance at 145 MHz. The effective resistance at 145 MHz is approximately 180 Ω. Figure 6b shows the reflection coefficient plot that results when a parallel capacitance Cp of 1.5 pF is added to the model. The corresponding model is shown in Figure 6c. In this case, the effective resistance drops to 50 Ω and at the same time the compensation of the reactance only requires a smaller series inductance Ls of 2 x 0.19 µH.



(a) (b) (c)

Figure 6: Reflection coefficient (a) without parallel capacitance, with Ls = 2 x 0.35 µH, (b) with parallel capacitance of 1.5 pF and with Ls = 2 x 0.195 µH. In (c) the EZNEC model of the dipole antenna: On the left is the overall view of the model with a total length of 2.5 m, on the right is the enlarged view of the center section with the two series inductors Ls/2, the source and the parallel capacitance Cp.

Version 2

This insight gave impetus to building an improved version of the antenna. The aim was a mechanically robust design in a slightly larger housing that also leaves space for a preamplifier (e.g. the VV145VOX from SHF-Elektronik). Secondly, the lower dipole arm should also be realized by a spring steel rod. However, this turned out to be a big problem because suitable spring steel rods, such as those used in commercial mobile antennas for the shortwave bands, are not available on the market or are not available in the required length. Instead of highly flexible spring steel, only "spring-hard" wire should be sufficient for a thin rod in a stationary antenna with little mechanical stress from wind. This wire should be made of stainless steel because of its corrosion resistance. Thin stainless steel tubes and wires with a length of 1m are readily available from model kit shops and can be joined: To form a 1.25 m long dipole arm, a 2 mm wire was inserted a few cm deep into an approximately 30 cm long tube with an outer diameter of 3 mm and an inner diameter of 2 mm and was fixed with a small screw socket (from a luster terminal). These dipole arms were inserted through the housing wall into luster terminals that are screwed to the bottom of the housing. The two rods are mechanically attached and can easily be connected to the outer ends of the coil (series inductance) in the luster terminals. Serious attenuation losses due to the lower conductivity of stainless steel compared to copper are not to be expected - in EZNEC a difference in antenna gain of around 0.2 dB is calculated. Connecting the inserted wire to the accepting tube section using a clamp as seen in Figure 7 is just as unproblematic, because Figure 2b shows: The connection point is just at the current minimum, as can be seen from the current distribution in the Figure, where the impedance is in the kΩ range. Accordingly, resistances of the order of 1 to 10 Ω caused by poor contacting can hardly lead to noticeable losses. Apart from that, even if the conductive contacting fails completely, the still existing capacitance between the two closely nested conductors would ensure the current flow. With an estimated capacity of 100 pF per connection, the simulation produces practically identical results as with a perfect connection. This means that you could even use glue instead of a clamp for fixation and a soldered connection is completely unnecessary.

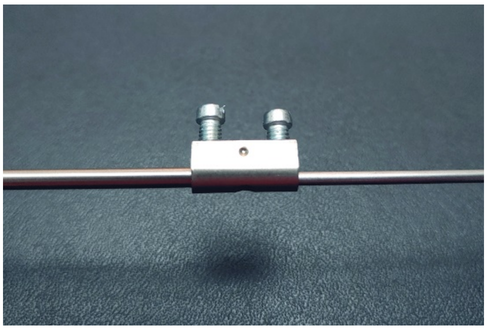


Figure 7: The screw socket of a luster terminal is used to fix and contact the wire rod inserted into the tube (on the left in the picture).

Most likely, the loss resistance of the coil (series inductance) could cause a noticeable drop in antenna gain. However, the series resistance of only 0.6 Ω calculated in OptiCoil /4 / only leads to a loss of around 0.1 dB in the EZNEC simulation. Perhaps a little less optimistically, overall losses (excluding losses due to currents in the neighboring antennas and other metallic conductors and in the ground) can be estimated at around 0.5 dB.

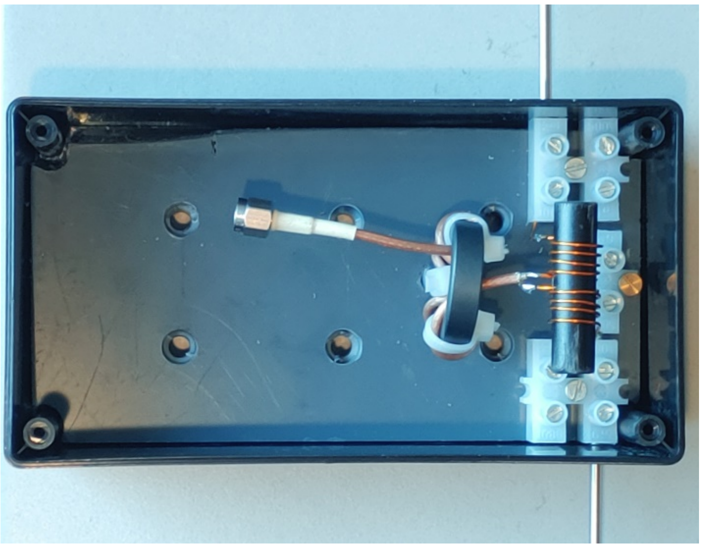


Figure 8: Antenna housing 152 mm x 82 mm with a PVC plate to stiffen the housing base and to attach the luster terminals. Dipole arms at the top and bottom fixed in the luster terminals. The coil on a 10 mm diameter GRP tube in the middle above the luster terminals. Connection to the inner coil ends via coaxial cable with toroidal ferrite balun.

To mount the dipole antenna on the boom tube, a suitable aluminum plate with two grounding clamps, see Figure 9, was attached vertically to the boom and the antenna housing was screwed onto the aluminum plate through the holes in the housing base.

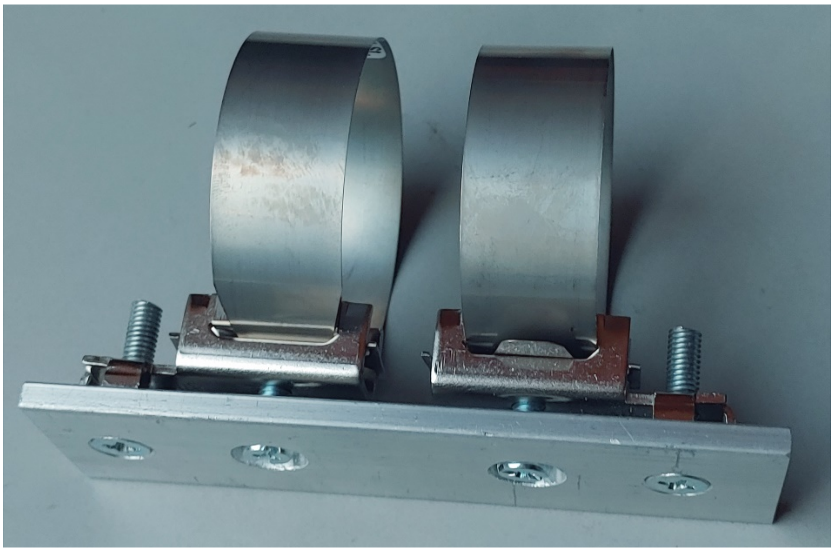


Figure 9: Mounting plate for the antenna housing with two grounding clamps for attaching to the boom tube.

Matching and results

To tune the antenna, three parameters must fit together, the dipole length, the parallel capacitance and the compensation inductance. As shown, the resulting reactance of the dipole impedance can be compensated for by more or fewer turns on the two-part coil (series inductance). First, however, a coil was installed that realized the compensation inductance expected from the simulation. In the position of the antenna on the support tube, as shown in Figure 1, the wire rods were inserted into the openings of the 3 mm tubes and clamped in order to achieve the total length of the dipole arms of around 1.25 m (the total length may well be a few cm more or less). In this condition, a NanoVNA was used at the end of a connecting cable to measure the reflection coefficient. Here a Smith Chart representation is absolutely necessary in order to distinguish between resistive and reactive components of the impedance. Secondly, the calibration must be carried out in such a way that the measured reflection coefficient refers exactly to the end of the coax cable where the two inner coil ends are connected, because the antenna impedance can only be measured there. In the following tuning process, the effective resistance of the antenna impedance was matched by adjusting the stray capacitance: As can be seen in Figure 8, the two dipole arms in the housing lie side by side on the PVC housing base, separated by the width of the luster terminals. The stray capacitance between the two ends can be increased by increasing the overlap of the tube ends until the measured impedance has an effective resistance of approximately 50 Ω. To do this, it was necessary to push the tube of the lower dipole arm into the middle luster terminal, while the tube of the upper dipolar arm did not protrude from the upper luster terminal. Only with this setting the measured antenna impedance was used to determine the required series inductance for compensation of the reactance component of the antenna impedance; from this, the number of coil turns was redetermined accordingly and the inductance fine-tuned by moving the turns on the coil body.

A result of measuring the impedance of the tuned antenna is shown in Figure 10. In the diagram of the reflection coefficient two dips can be seen. The measured center frequency of 144.4 MHz in the 2-meter band can be easily increased by further inserting one of the wire rods into the corresponding tube, thus shortening the dipole, but the matching bandwidth of 8 MHz (at VSWR = 2.6) is completely sufficient for operation over the entire 2-meter band. In addition, the total length of the dipole arms of 2.5 m together with the series inductance also results in a resonance in the 6-meter band, with 6 MHz also giving sufficient match bandwidth for operation over the entire 6-meter band (although in DL up to 51.0 MHz vertical polarization is not permitted).

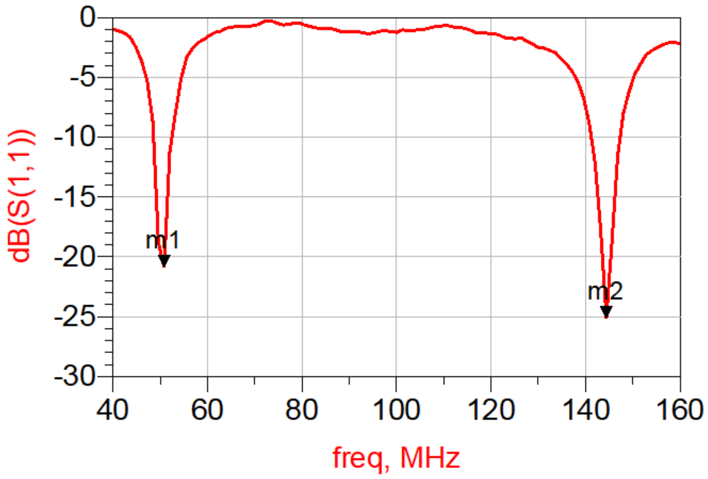


Figure 10: Measurement of the reflection coefficient (S11) of the dipole antenna. Marker m1 at 50.8 MHz, marker m2 at 144.4 MHz.

This second resonant frequency is no coincidence: Just as in this case a second resonant frequency of the antenna occurs at about a third of the upper frequency, the antenna could also be designed as a 2 x 5/8 λ dipole for the 70-centimeter band (total length approx. 87 cm) – this would realize a second resonant frequency as a λ/2 dipole in the 2-meter band.

Three antennas close together – does that work well?

As mentioned at the beginning, the VHF dipole also induces currents on the adjacent two antennas and their support tubes and the metallic balcony railing - the reaction of these currents on the VHF antenna leads to a slight shift in its impedance (compared to the impedance in free space) and to distortions of their radiation pattern. The impedance shift is already compensated for by the tuning described. The pattern distortion must be accepted, but the simulation does not suggest any deep drops, so that there will hardly be any noticeable impairment in operation. Conversely, the MagLoop and the HF dipole also induce currents in the VHF dipole when transmitting (3.5 MHz - 30 MHz). However, the VHF dipole is so small relative to the wavelengths at shortwave that the simulation does not show any significant deterioration in the impedance and radiation characteristics of the two HF antennas, and this also is not seen in real operation.

What could be more of a problem is that even with only a small amount of mutual coupling from one antenna to the other, if one antenna has a high transmission power, a power level can be induced on the other antenna that could be dangerous for the receiver input connected to it. The NanoVNA was therefore used to measure the mutual coupling (the S21 - scattering parameter) between the antennas. Between the HF dipole and the VHF dipole S21 was found around -25 dB in the VHF and around HF range; the coupling to the MagLoop was at least 5 dB lower in both cases. So, if the dipole is used to transmit at 100 W in the HF range, around 320 mW of input power must be expected at the VHF transceiver, a value that is well above a limit of around 100 mW that experience has shown is still considered safe, for an Rx Input stage without strong out-of-band filtering. Conversely, there is no risk if the VHF Transceiver transmits a maximum of 20 W, since a maximum of only 63 mW of power arrives at the HF transceiver – in this case, there is no need for any reduction in coupling. In order to reduce the cross-coupling of HF power into the VHF transceiver, a high pass or a bandpass could be inserted into the housing of the VHF dipole, which would allow the VHF signal to pass through and block or at least lower the HF signal. A more simple solution is to use a stub line: A coaxial cable stub is connected parallel to the outgoing cable and has a length of a quarter wavelength at 145 MHz and a short circuit at the end. The cable exhibits an open circuit as input impedance in the 2-meter band and thus allows the VHF signal to pass through unhindered. At 30 MHz, however, the stub already represents a low inductive reactance of around 16 Ω, which creates a mismatch at the HF frequency and reduces the passage of the HF signal by around 5 dB; the lower the frequency, the higher the suppression through the stub line. This would mean that the coupled HF power on the VHF transceiver would already be reduced below the critical threshold over the entire HF range. Nevertheless, a slight blocking effect is noticeable on the VHF transceiver when the shortwave station is transmitting. A side effect of the stub is that the dipole is short-circuited for dc and set to ground potential, meaning that electrostatic charges that potentially are dangerous for the VHF transceiver are already grounded in the antenna.

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