Some Thoughts on Portable and Tunable Mobile and Small Loop Antennas

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Introduction

Hiker hams often use vertical whip antennas in combination with a trailing dragged wire as a ground radial. This is described in K. Siwiak, KE4PT, “Ionospherica –How Dipoles Radiate – the Hiker’s Bent Dipole,” *QRP Quarterly*, Spring 2014, pp 27-28. But how good is that solution? This paper tries to clarify some things about these antennas, as well as mobile verticals and small loops. We will also present some definitions and measurements.

The detailed paper on loop antennas and mathematics for loops as large as 0.3 to 0.4 wavelengths in circumference can be found in K. Siwiak, KE4PT, and R. Quick, W4RQ, “Small Gap-Resonated HF Loop Antennas” *QST*, Sep., 2018.

As useful introduction to the problem of electrically short antennas is:

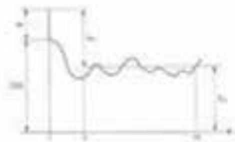
https://www.microwavejournal.com/articles/32231-tuning-electrically-short-antennas-for-field-operation

Definitions

Our definitions are based on Zinke and Brunswick, *Fundamentals of RF and Microwave Techniques and Technologies*, Springer 2021, Editors: Ulrich L. Rohde. Hans L. Hartnagel, Matthias Rudolph, Ruediger Quay, Chapter 6: Electromagnetic Radiation and Antennas, chapter 6. Note that we retain the Figure and equation numbering and notation of Zinke and Brunswick. Some of the drawing still contains some German language comments but the English translation is added.

Field strength, in physics, means the *amplitude* of a vector-valued field (in V/m for an electric field ***E*** and in A/m for a magnetic field ***H***). An electromagnetic field results in both an electric field strength and a magnetic field strength. As an application in radio frequency telecommunications the signal strength excites a receiving antenna and thereby induces a voltage at a specific frequency and polarization in order to provide an input signal to a radio receiver.

Effective height



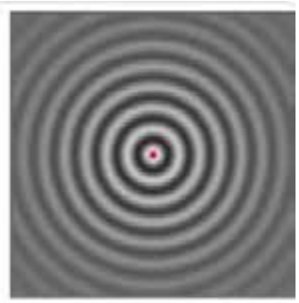
In telecommunication, the effective height, or effective length, of an antenna is the height of the antenna's center of radiation above the ground. It is defined as the ratio of the induced voltage to the incident field. Wikipedia

The reference level in V/m addresses the effective height of an antenna, not its physical height or length. So a distant transmitter radiates power, for example 1 kW, and at the place of reception generates field strength as defined above. There is a magnetic field and an electric field; they are perpendicular to each other. The whip antenna, which is short relative to a wavelength, is essentially a voltage probe, while a infinitesimal loop antenna, which is very small in diameter compared to the wavelength, is a magnetic probe.

The electric field-strength is measured in V/m and the magnetic filed strength is in A/m.

Gain is often referenced to isotropic radiation, which is really mathematical convenience.

Isotropic radiator



An isotropic radiator is a theoretical point source of electromagnetic or sound waves which radiates the same intensity of radiation in all directions. It has no preferred direction of radiation. It radiates uniformly in all directions over a sphere centred on the source. Wikipedia

Exact Value

The exact value of Impedance of free space is universally accepted and defined value is-

Impedance of free space Z0 = 119.9169832 πΩ = 376.73031346177... Ω.

Formula

The Impedance of free space can be mathematically written as –

$$Z_0 = \frac{E}{H} = \sqrt{\frac{\mu_0}{\varepsilon_0}} = \mu_0 c_0 = \frac{1}{\varepsilon_0 c_0}$$

Where,

- μ0 is the permeability of free space.
- ε0 is the permittivity of free space.
- C0 is the speed of light in free space.
- H is the magnetic field strength.
- E is the electric field strength.

Z0 some time also referred as the admittance of free space.

Hope you have understood the Impedance of free space or the characteristic impedance of free space along with value and units and the formula.

[From **https://byjus.com/physics/impedance-of-free-space/**]

The available signal, in V/m, is not easily determined as different antennas and different receivers will indicate different result. Calibration is needed and the *antenna factor* must be specified. Therefore a special test receiver, as seen pictured below, is needed.

The Rohde & Schwarz PR 100 with the test antenna.

That receiver includes a table listed by name of the various available antenna factors. The antenna shown connected to the R&S PR 100 is calibrated and covers the frequency range from 9 kHz to 20 MHz

An integrated amplifier if needed can be activated. Another more powerful receiver is the ESMD monitoring receiver, more about it below.

The peak gain of a half-wave **dipole**, neglecting electrical inefficiency, is equal to the directive **gain**, which is 1.64 (2.15 dBi). The following is a basic antenna theory over-review.

Introduction in to Electromagnetic Radiation and Antennas

Basic concepts of radiation, some mathematical treatment

We will consider the emission of waves in free space; we also assume that the ground, metallic reflectors, etc. have ideal conductivity. Also we assume initially that the diameter of the antenna wire is negligible thickness and that the dipole/antenna is resonant at the frequency of operation. For electrically short antenna, a matching circuit for the resonant condition is needed. Either a loading coil or a capacitive hat or a combination of both may be used. In this paper we first focus on the dipole, then later on the small HF loop antenna.

The antenna forms the transition between the transmission line and propagation of waves in free space [literally: "between line-conducted propagation and free wave propagation"]. One might think of the antenna as a transducer coupling the applied RF voltage and current at the transmitter or receiver, with the E and H fields in space. The following sections will examine the two elementary dipoles: the electric (or Hertzian) dipole and the magnetic dipole realized as a very small current loop (loop antenna).

While a professor in Karlsruhe, Heinrich Hertz (February 22, 1857 to January 1, 1894) demonstrated that analogous to light waves, electromagnetic waves can be refracted and reflected (1887/88) and are propagated like light waves. Hertz also observed the passage of cathode rays (electrons) through metal foil.

1.1 Field equations and radiation pattern for the Electric or Hertzian dipole

Two alternating point charges $+q$ and $-q$ have a separation $\Delta l << \lambda/4$ such that the charge current $i(t)$ can be assumed to be independent of location (**Figure {6.}1/1**). The field around this "Hertzian dipole" is rotationally symmetrical, i.e. all components at any point P are independent of the azimuth angle φ. We will assume that the component in the direction of propagation is E_r and the component perpendicular to it in the plane of the drawing is E_θ. No E field is present perpendicular to the plane of the drawing, i.e. $E_\varphi \equiv 0$. The only magnetic field component is H_φ (perpendicular to the plane of the drawing). Since $H_r \equiv 0$ and $H_\theta \equiv 0$, this is an E wave or TM wave

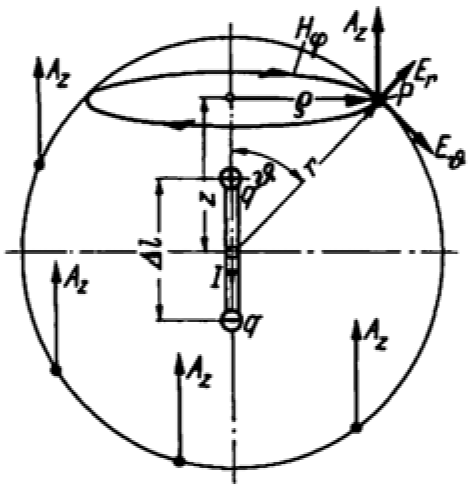


Abb. 6.1/1 Hertzscher Dipol mit Feldkomponenten E_ϑ, H_φ, E_r. A_z ist die einzige Komponente des Vektorpotentials

Figure {6.}1/1. Hertzian dipole with field components E_θ, H_φ, E_r. A_z is the only component of the vector potential.

We will now consider the magnitude of the field components at a distance r for the given current $i(t)$ and separation Δl of the charges, i.e. for H_φ , E_θ, $E_r = f(r, \theta)$. We will derive them hereafter from the vector potential. The vector potential ***A*** (see chapter 5) is introduced with

$$\boldsymbol{B} = \text{rot}\,\boldsymbol{A}. \qquad (\{6.\}1/1)$$

On the other hand, according to Maxwell's first equation (Ampère's law), we have

$$\text{rot}\,\boldsymbol{H} = \boldsymbol{J} + \frac{\partial \boldsymbol{D}}{\partial t}. \qquad (6.1/2)$$

J is the line current density and $\partial\boldsymbol{D}/\mathrm{d}t$ is the density of the displacement current. In other words, for location-independent μ we have

$$\frac{1}{\mu}\text{rot rot}\,\boldsymbol{A} = \boldsymbol{J} + \frac{\partial \boldsymbol{D}}{\partial t}. \qquad (6.1/3)$$

This equation has the solution

$$\boldsymbol{A} = \frac{\mu}{4\pi}\iiint\limits_V \frac{\boldsymbol{J}\left(t - \frac{r}{c}\right)}{r}\,\mathrm{d}V. \qquad (6.1/4)$$

Here, dV is the volume element that transports the current density $\boldsymbol{J}(t)$, r is the distance between the dipole and point, and c is the speed of light. Because the effect of $\boldsymbol{J}$ at the point is delayed or "retarded" by the time period r/c, $\boldsymbol{A}$ is known as the "retarded vector potential".

Therefore, we can calculate the vector potential for the entire space if we know the current distribution at the place of origin. For the case of the Hertzian dipole, we have

$$\mathrm{d}V = A\,\mathrm{d}z.$$

Here, A is the cross-section of the conductor between the two charges and dz is its length element. The current density in case of purely harmonic excitation is

$$J_z \mathrm{e}^{\mathrm{j}\omega\left(t-\frac{r}{c}\right)} = \frac{I}{A}\mathrm{e}^{\mathrm{j}\omega\left(t-\frac{r}{c}\right)} = \frac{I}{A}\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}r}\,\mathrm{e}^{\mathrm{j}\omega t}. \qquad (6.1/5)$$

Then, the vector potential of the dipole of length Δl has only the component

$$A_z = \frac{\mu}{4\pi}\int_{-\Delta l/2}^{+\Delta l/2} \frac{I}{A}\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}r}\frac{A\mathrm{d}z}{r} = \frac{\mu I \Delta l}{4\pi r}\,\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}r} \quad 1 \,. \qquad (6.1/6)$$

[In the above integration, we must have $\Delta l \ll r$. This allows us to move the average value of $r(z)$ in front of the integral since it is independent of z.]

According to equation ({6.}1/6), the vector potential is constant (see **Figure {6.}1/1**) on a spherical surface arranged around the Hertzian dipole with radius r. If r grows by one wavelength, the phase of A_z rotates by 2π. All of the field components can be derived from A_z. Equation ({6.1}/1) becomes

$$H_\varphi = \frac{1}{\mu}(\mathrm{rot}\,A)_\varphi. \qquad (6.1/7)$$

Since only A_z is present, it is more convenient to perform the following calculation in cylindrical coordinates than in spherical coordinates.

Given that

$$(\mathrm{rot}\,A)_\varphi = \frac{\partial A_\varrho}{\partial z} - \frac{\partial A_z}{\partial \varrho} \qquad (A_\varrho \equiv 0) \qquad (6.1/8)$$

we obtain the following from equation ({6}.1/7):

$$H_\varphi = -\frac{1}{\mu}\frac{\partial A_z}{\partial \varrho}. \tag{6.1/9}$$

Since A_z is known in terms of its dependence on the sphere's radius r, it is more convenient to write

$$H_\varphi = -\frac{1}{\mu}\frac{\partial A_z \partial r}{\partial r\, \partial \varrho}. \tag{6.1/10}$$

Now according to **Figure {6.}1/1** we have

$$r_2 = \varrho^2 + z^2 .$$

We take the partial derivative with respect to ρ and find

$$2r\frac{\partial r}{\partial \varrho} = 2\varrho + 0$$

or

$$\frac{\partial r}{\partial \varrho} = \frac{\varrho}{r} = \sin \vartheta . \tag{6.1/11}$$

We thus have

$$H_\varphi = -\frac{1}{\mu}\frac{\partial A_z}{\partial r}\sin \vartheta . \tag{6.1/12}$$

According to equation ({6.1}/6), we have

$$H_\varphi = \mathrm{j}\frac{I\Delta l}{2\lambda}\frac{\sin \vartheta}{r}\mathrm{e}^{-\mathrm{j}\frac{2\pi}{\lambda}r}\left(1 + \frac{1}{\mathrm{j}\frac{2\pi r}{\lambda}}\right) . \tag{6.1/13}$$

If $r \ll \lambda$, then H_φ is determined essentially only by the second term in the parentheses. This case is known as the close "near field". Here, H_φ is proportional to $1/r^2$. In the "far field" ($r \gg \lambda$), the first addend in the parentheses dominates. Here, H_φ only decreases proportional to $1/r$. For example, we are still in the near field at a distance of 10m for f = 50 MHz (λ = 6m), but we are already in the far field at a distance of 100 m . The field strength components of the far and near fields are mutually phase-shifted by 90° . In the far field, the energy corresponding to H_φ^2 decreases inversely proportional to the sphere's surface, i.e. proportional to $1/r^2$. The same inverse-square law also applies in optics and acoustics in the far field.

Using Maxwell's first equation, we can determine the electric field components E_θ and E_r from H_φ.

$$\text{rot}\,\mathbf{H} = \mathrm{j}\omega\mathbf{D} + \mathbf{J}\,. \tag{6.1/14}$$

Outside of the conductors we have **J** = 0 and

$$\mathbf{D} = \varepsilon_r \varepsilon_0 \mathbf{E} \qquad \text{mit} \qquad \varepsilon_r = 1.$$

{"mit" --> "where"}

In other words, we have

$$\mathbf{E} = \frac{1}{\mathrm{j}\omega\varepsilon_0}\text{rot}\,\mathbf{H} = \frac{1}{\mathrm{j}\frac{2\pi c}{\lambda}\varepsilon_0}\text{rot}\,\mathbf{H} = \frac{Z_0}{\mathrm{j}\frac{2\pi}{\lambda}}\text{rot}\,\mathbf{H}\,. \tag{6.1/15}$$

Here, $Z_0 = \sqrt{(\mu_0/\varepsilon_0} \approx 377\ \Omega$ is the wave impedance of free space, (see section 4.2). For the transverse component E_θ of the electric field, we have the following according to equation ({6.1}/15):

$$E_\vartheta = \frac{Z_0}{\mathrm{j}\frac{2\pi}{\lambda}}(\text{rot}\,\mathbf{H})_\vartheta\,. \tag{6.1/16}$$

For $H_r = 0$ (E waves!), we have

$$(\text{rot}\,\mathbf{H})_\vartheta = -\frac{1}{r}\frac{\partial(rH_\varphi\,)}{\partial r} \tag{6.1/17}$$

also

$$E_\vartheta = \mathrm{j}Z_0\frac{I\Delta l}{2\lambda}\frac{\sin\vartheta}{r}\mathrm{e}^{-\mathrm{j}\frac{2\pi r}{\lambda}}\left(1 + \frac{1}{\mathrm{j}\frac{2\pi r}{\lambda}} + \frac{1}{\left(\mathrm{j}\frac{2\pi r}{\lambda}\right)^2}\right)\,. \tag{6.1/18}$$

The first term in parentheses again belongs to the far field while the last two terms determine the near field. E_θ and H_φ represent the transverse components which are relevant for energy transport. As we can see from equations ({6.1}/13) and ({6.}1/18), they are in phase in the far field ($r >> \lambda$) so we can write

$$E_\vartheta = H_\varphi \, Z_0 \,. \tag{6.1/19}$$

Here, both *r* and *θ* are missing, i.e. this relationship holds at every point in the space as long as $r >> \lambda$. When making field strength measurements, it thus does not matter whether we measure H_φ or E_θ. Despite the spherical propagation, we encounter the same relationship between the electric and magnetic transverse field strengths as on a line. Since E_θ and H_φ are involved in the energy transport, the Poynting radiation vector (**Figure {6.}1/2a**) is

$$\mathbf{S}_r = \tfrac{1}{2} \, \mathbf{E}_\vartheta \times \mathbf{H}^*_\varphi \,. \tag{6.1/20}$$

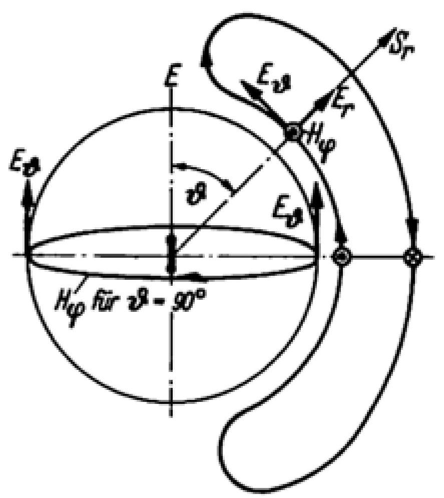


Abb. 6.1/2a. Prinzipielles Feldbild des Hertzschen Dipols im Fernfeld

Figure {6.}1/2a. Basic field pattern for the Hertzian dipole in the far field.

Now we must still determine the radial component of the electric field strength E_r. For $H_\theta \equiv 0$, the radial component is

$$(\mathrm{rot}\, \mathbf{H})_r = \frac{1}{r \sin\vartheta} \frac{\partial(\sin\vartheta \, H_\varphi)}{\partial\vartheta} \tag{6.1/21}$$

$$E_r = \mathrm{j} Z_0 \frac{I \Delta l}{2\lambda} \frac{2\cos\vartheta}{r} \, \mathrm{e}^{-\mathrm{j}\frac{2\pi r}{\lambda}} \left(\frac{1}{\mathrm{j}\frac{2\pi r}{\lambda}} + \frac{1}{\left(\mathrm{j}\frac{2\pi r}{\lambda}\right)^2} \right) \,. \tag{6.1/22}$$

The first term in parentheses again belongs to the far field while the second term belongs to the near field. E_r is small with respect to E_θ in the far field except for the dipole axis and its surroundings. Moreover, E_r is phase-shifted by 90° with respect to E_θ and H_φ and thus does not contribute to the radiation. We can see that the Hertzian dipole produces an *E* wave (TM wave) with a structure that increasingly approaches that of the plane wave (TEM wave) in the far field as the distance grows.

Figs. {6.}1/2b and {6.}1/3 show a series of Hertzian dipole field line patterns at different time points. As we can see, the radial field strength E_r is inevitably required so that the electric field lines either end at the dipole or can close in space. The snapshots in **Figure {6.}1/3a-h** more precisely depict the process of "pinching off" (or "tying off") the field lines.

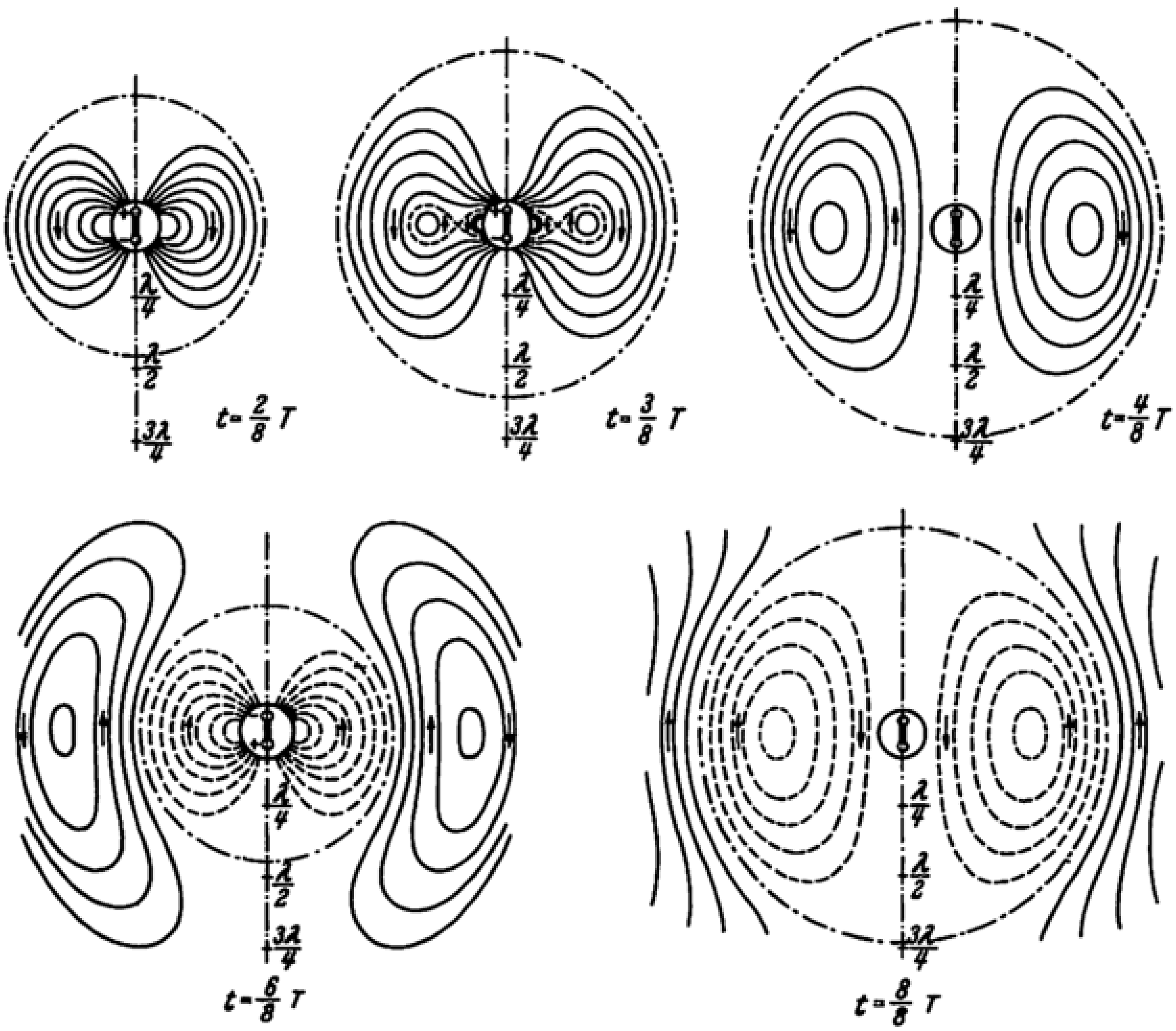


Abb. 6.1/2b. Elektrische Feldlinien des Hertzschen Dipols in Zeitabständen von 1/8*T* und 1/4*T*

Figure ({6.}1/2b). Electric field lines of the Hertzian dipole at time intervals of 1/8*T* and 1/4*T.*

The dependence of the transverse field strengths E_θ and H_φ on the angle θ for constant distance r in the far field is known as the antenna's "vertical radiation pattern" if the dipole is arranged perpendicular to the reference plane. In the far zone (or "far field"), we have

$$E_\vartheta = E_{\vartheta\max} \sin\vartheta \qquad (6.1/23)$$

and

$$H_\varphi = H_{\varphi\max} \sin\vartheta\,. \qquad (6.1/24)$$

In other words, both exhibit the same pattern. The Hertzian dipole is an omnidirectional radiator in all planes perpendicular to the dipole axis: E_θ and H_φ are independent of φ; in the planes through the dipole axis, it exhibits slight directivity (given by sin θ). For $\theta = 0$, we have $E_\theta = 0$ whereas E_r is at a maximum. The spatial far-field pattern of the Hertzian dipole has a toroidal shape. **Figure {6.}1/4** illustrates the dependency of the radiation density on θ in the far field of the Hertzian dipole.

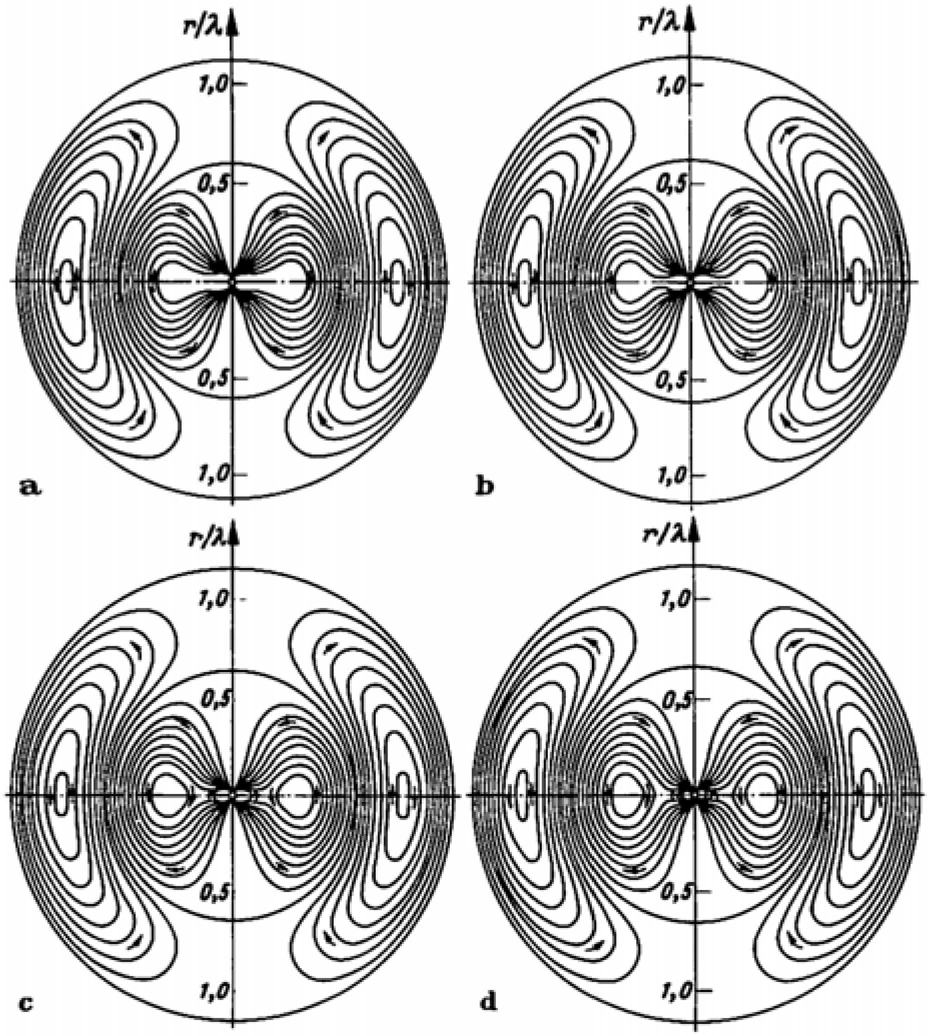
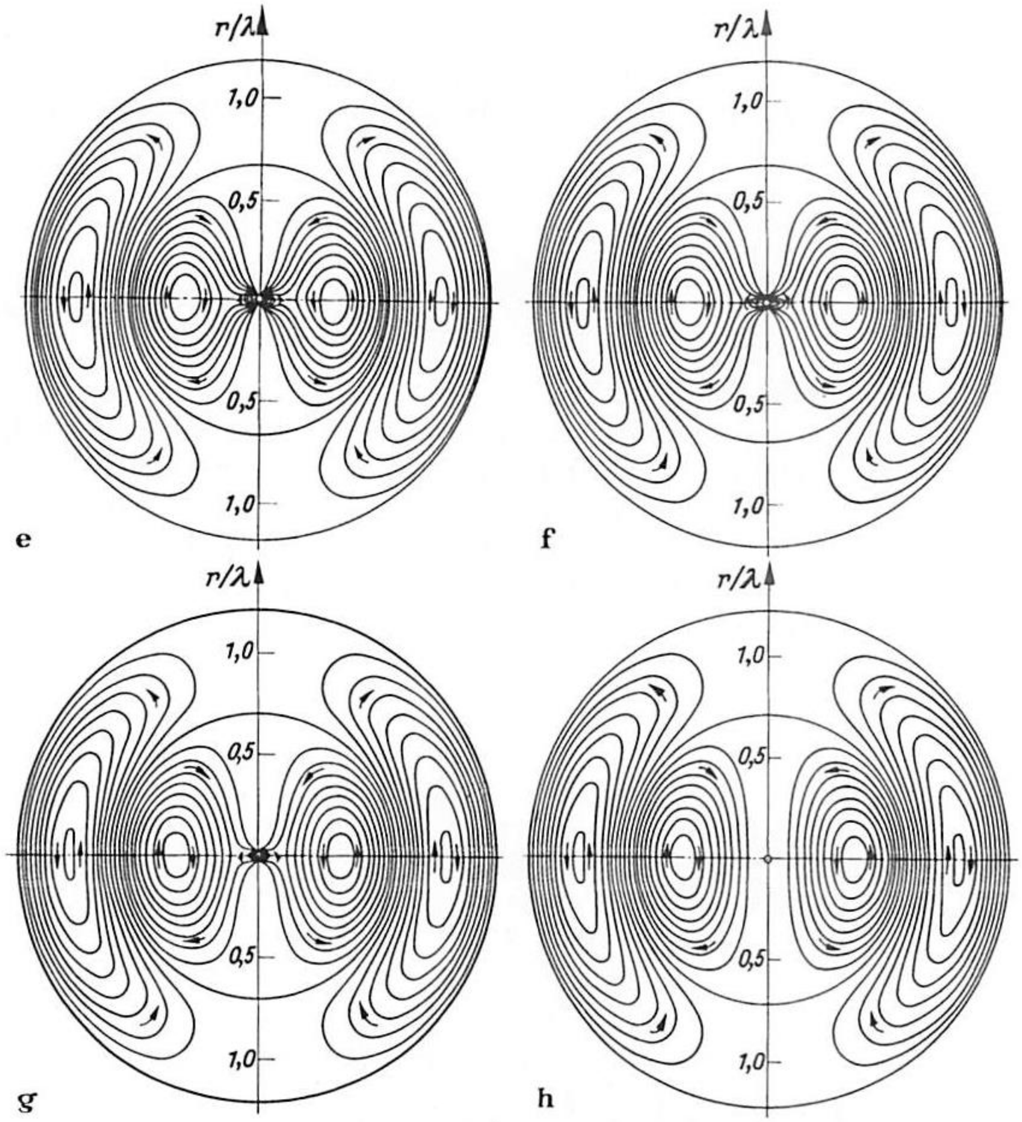


Abb. 6.1/3a–d. Elektrisches Feldbild eines Hertzschen Dipols für verschiedene Zeitpunkte, bezogen auf die Periodendauer T. **a** $t = 25/64T$; **b** $t = 26/64T$; **c** $t = 27/64T$; **d** $t = 28/64T$

Figure {6.}1/3a-d. Electric field pattern of a Hertzian dipole at different time points, referred to the period *T*. **a** $t = 25/64T$; **b** $t = 26/64T$; **c** $t = 27/64T$; **d** $t = 28/64T$



Zinke/Brunswig, Lehrbuch Hochfrequenztechn

Abb. 6.1/3 e–h. **e** $t = 29/64T$; **f** $t = 30/64T$; **g** $t = 31/64T$; **h** $t = 32/64T = T/2$

["Zinke/Brunswig, Lehrbuch Hochfrequenztechnik" --> "Zinke/Brunswig, RF Engineering textbook"]

Figure {6.}1/3 e-h. **e** t = 29/64T; **f** t = 30/64T; **g** t = 31/64T; **h** t = 32/64T = T/2

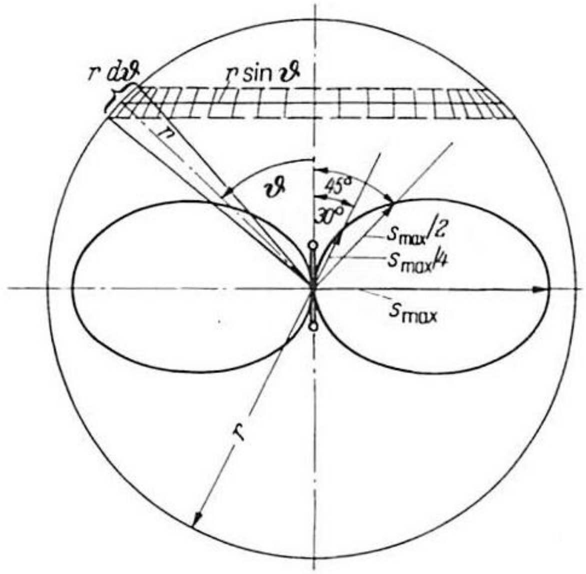


Abb. 6.1/4. Verteilung der Strahlungsdichte im Fernfeld des Hertzschen Dipols. Bezeichnungen der Kugelzone

Figure {6.}1/4. Distribution of radiation density in the far field of the Hertzian dipole. Sphere zone designations

1.2 The infinitesimal loop as an energized magnetic dipole[1] [2]

We showed in Chapter 5 that propagation processes can very often be characterized with only three field components, i.e. two components perpendicular to the propagation direction and one longitudinal component. We designate the fields as *E* fields (TM fields) or *H* fields (TE fields) depending on whether the electric field strength or the magnetic field strength is involved. The Hertzian dipole produces a field with E_θ and $H_\varphi = E_\theta/Z_0$ as the transverse components and E_r as the longitudinal component, i.e. an *E* field. There exists a dual *H* field with the transverse components H_θ and $E_\varphi = H_\theta Z_0$ and the longitudinal component H_r. A field of this sort is actually produced by a current-carrying loop[3] as shown in **Figure 6.1/5b**. By way of comparison, consider **Figure 6.1/5a** in which our Hertzian dipole has been transformed from its usual representation (**Figure 6.1/1**) into a circular-plate capacitor with plate separation Δl. (We can take this liberty because the field equations for the Hertzian dipole are valid only for distances $r >> \Delta l$, i.e. the actual near field of the technical arrangement is not present [*alternatives*: "determined", “defined”, “established”].) The electrostatic stray field of the circular-plate capacitor has

[1] The magnetic dipole is sometimes known as the “Fitzgeraldian dipoleThe Fitzgerald vector is the dual counterpart of the Hertz vector. G. Fitzgerald (1851-1901) extended Maxwell’s work on electromagnetic light theory. From 1880 on, he worked as a professor in Dublin.

[2] Literally "supplied (or fed) magnetic dipole".

[3] Translator’s note: Literally "a loop with current flowing through it".

the same structure as the stationary magnetic near field of a circular loop with flux *Θ* [4]. Here, it does not matter whether this flux is produced in a single winding or in multiple windings with less current. We assume that the loop diameter is equal to the diameter of the circular-plate capacitor. Under this condition, the *E* lines correspond to the *H* lines of the loop even in the immediate proximity. The significant duality of the electric and magnetic dipoles is revealed by comparing the field equations that are obtained for the loop either based on decomposition into circularly arranged Hertzian dipoles or an approach along the lines of section {6.}1.1.

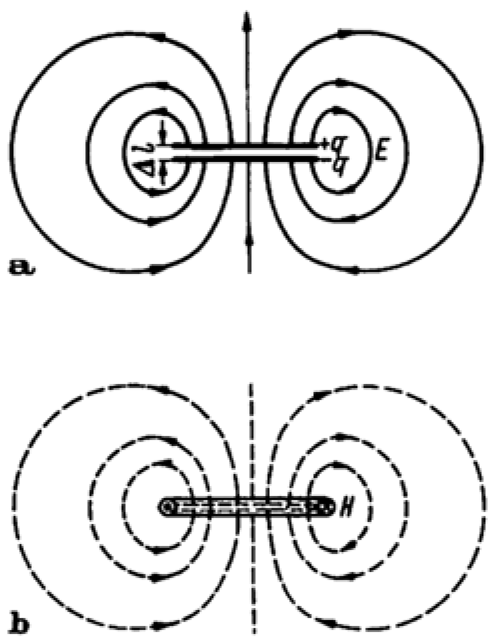


Abb. 6.1/5a, b. Nahfelder von Dipolen. a elektrisches Feld eines Kreisplattenkondensators; b magnetisches Feld einer stromdurchflossenen Windung

Figure {6.1}1/5a,b. Near fields of dipoles. **a** Electric field of a circular-plate capacitor; **b** Magnetic field of a current-carrying winding.

There is a correspondence between the field components for the Hertzian dipole

$$E_r = \mathrm{j} Z_0 \frac{\Delta l I}{2\lambda} \frac{2\cos\vartheta}{r} \times \mathrm{e}^{-\mathrm{j}\frac{2\pi r}{\lambda}} \left(\frac{1}{\mathrm{j}\frac{2\pi r}{\lambda}} + \frac{1}{\left(\mathrm{j}\frac{2\pi r}{\lambda}\right)^2} \right),$$

[4] Translator's note: "Durchflutung" is also sometimes translated as "magnetomotive force".

and

$$E_\vartheta = jZ_0 \frac{\Delta l I}{2\lambda} \frac{\sin\vartheta}{r} \times e^{-j\frac{2\pi r}{\lambda}} \left(1 + \frac{1}{j\frac{2\pi r}{\lambda}} + \frac{1}{\left(j\frac{2\pi r}{\lambda}\right)^2}\right),$$

and the field components for the magnetic dipole; respecting that $\Delta l << r$, that is, the loop is infinitesimal.

$$H_r = \frac{2\pi w A}{\lambda} \frac{I}{2\lambda} \frac{2\cos\vartheta}{r} \times e^{-j\frac{2\pi r}{\lambda}} \left(\frac{1}{j\frac{2\pi r}{\lambda}} + \frac{1}{\left(j\frac{2\pi r}{\lambda}\right)^2}\right),$$

$$H_\vartheta = \frac{2\pi w A}{\lambda} \frac{I}{2\lambda} \frac{\sin\vartheta}{r} \times e^{-j\frac{2\pi r}{\lambda}} \left(1 + \frac{1}{j\frac{2\pi r}{\lambda}} + \frac{1}{\left(j\frac{2\pi r}{\lambda}\right)^2}\right),$$

with the number of windings w and area A. In the far field, there is a correspondence between

$H_\varphi = \frac{E_\vartheta}{Z_0}$	for the Hertzian dipole	$E_\varphi = -H_\vartheta Z_0$	for the magnetic dipole,

as well as the power densities

$$S_r = \frac{1}{2} E_\vartheta H_\varphi = \frac{1}{2}\frac{E_\vartheta^2}{Z_0} = \frac{1}{2} H_\varphi^2 Z_0 \quad \text{und} \quad S_r = \frac{1}{2} E_\varphi H_\vartheta = \frac{1}{2}\frac{E_\varphi^2}{Z_0} = \frac{1}{2} H_\vartheta^2 Z_0.$$

The charge separation Δl corresponds to $2\pi wA/\lambda$.

In the loop's far field, the field components predominate that decrease proportional to $1/r$:

$$H_\vartheta = \frac{2\pi w A}{\lambda}\frac{I}{2\lambda}\frac{\sin\vartheta}{r}\,e^{-j\frac{2\pi r}{\lambda}} = H_{max}\sin\vartheta\,, \tag{6.1/25}$$

$$E_\varphi = -\frac{2\pi w A}{\lambda}Z_0\frac{I}{2\lambda}\frac{\sin\vartheta}{r}\,e^{-j\frac{2\pi r}{\lambda}} = E_{max}\sin\vartheta\,. \tag{6.1/26}$$

Thus, the radiation patterns of the loop and Hertzian dipole correspond to one another.

1.3 The Hertzian dipole and infinitesimal loop antenna as receiving antennas

We will now operate our Hertzian dipole as a receiving antenna, i.e. allow an electromagnetic wave to act upon it. As shown in **Figure {6.}1/6a**, the magnetic field strength **H** of this wave is perpendicular to the dipole axis while the propagation direction forms the angle θ with the dipole axis. We now measure the electromotive force [*alternative*: "EMF"] V_e at the dipole terminals. It is equal to the integral of the electric field strength extended over the length of the dipole:

$$U_e = \int_{-\Delta l/2}^{+\Delta l/2} \mathbf{E}\,d\mathbf{l} = -E\Delta l\sin\vartheta = U_{e,max}\sin\vartheta\,. \tag{6.1/27}$$

The Hertzian dipole’s radiation pattern is thus independent of whether it is operated as a transmitting or receiving antenna.

The same angle dependency is exhibited by the receive voltage of a loop that has a wire length that is small with respect to the wavelength. The electric field vector **E** lies in the loop plane as shown in **Figure {6.}1/6b**; propagation occurs at the angle θ with respect to the system axis. The electromotive force V_m is equal to the derivative with respect to time of the magnetic flux Φ:

$$U_m = -\frac{d\Phi}{dt} = -wA\,\sin\vartheta\,\mu_0\,j\omega H\,.$$

Here, A is the area and w is the number of windings in the loop. Based on the far-field relationship

$$\mathbf{E} = Z_0\mathbf{H} = \sqrt{\frac{\mu_0}{\varepsilon_0}}\mathbf{H} \quad \text{und} \quad \omega = \frac{2\pi c}{\lambda} = \frac{2\pi}{\lambda}\frac{1}{\sqrt{\mu_0\varepsilon_0}}$$

we find that

$$U_m = -j\frac{2\pi w A}{\lambda}E\sin\vartheta = U_{m.\,max}\sin\vartheta\,. \tag{6.1/28}$$

Likewise for the loop, the radiation patterns are thus identical for transmit and receive operation. This correspondence holds not only for the Hertzian dipole and the loop, but also for any arbitrary form since we can imagine that any antenna is composed of Hertzian dipoles. This is the subject of the reciprocity theorem for transmitting and receiving antennas (section{ 6.}1.9).

1.5 Radiation density, radiated power, radiation resistance

The vector product of the electric and magnetic vectors yields the Poynting vector **S** = **E** x **H**. Its effective value TERM-1 is a measure of the power density at the observed point while simultaneously indicating the direction of travel of the energy. For the Hertzian dipole in the far field, we have $E_\theta = Z_0 H_\varphi$ and thus

TERM-1:

$$\tilde{\mathbf{S}} = 1/2\,\mathbf{E} \times \mathbf{H}^*$$

$$S(\vartheta) = \frac{\tilde{E}_\vartheta^2}{Z_0} = \frac{\tilde{E}_{\max}^2}{Z_0} \sin^2 \vartheta = \tilde{S}_{\max} \sin^2 \vartheta . \qquad (6.1/38)$$

For example, for $\tilde{E}_{max}$ = 1 V/m, we have S_{max} = 2.65 mW/m^2. **Figure {6.}1/4** shows how S is a function of θ for the Hertzian dipole. The diagram is rotationally symmetrical about the axis of the Hertzian dipole. Integrating S over the spherical surface O that encloses the dipole (surface element dO = 2π$r$sinθrdθ), we obtain the total power radiated by the Hertzian dipole in free space

$$\begin{aligned} P_s \oiint S \mathrm{d}O &= \oiint \tilde{E}_\vartheta \tilde{H}_\varphi \,\mathrm{d}O \\ &= \frac{\tilde{E}_{\max}^2}{Z_0} \int_0^\pi \sin^2 \vartheta \; 2\pi r \sin \vartheta r \mathrm{d}\vartheta = 2\pi r^2 \frac{\tilde{E}_{\max}^2}{Z_0} \int_0^\pi \sin^3 \vartheta \,\mathrm{d}\vartheta , \\ P_s &= \frac{8}{3} \pi r^2 \frac{\tilde{E}_{\max}^2}{Z_0} = \frac{8}{3} \pi r^2 \; \tilde{H}_{\varphi\max}^2 Z_0 \end{aligned} \qquad (6.1/39)$$

or as an adapted quantity equation

$$\frac{\tilde{E}_{\max}}{\mathrm{m\,V}/m} = 212 \sqrt{\frac{P_s}{\mathrm{kW}}} \frac{1}{r/\mathrm{km}} . \qquad (6.1/40)$$

The radiated power P_s is supplied by the transmitter. We apply a suitably defined antenna current $\tilde{I}$

$$P_s = \tilde{I}^2 R_s \qquad (6.1/41)$$

and define the “radiation resistance” R_S as the equivalent ohmic resistance on which the radiated power P_s would be emitted. In the far field, we can arrange equation ({6.}1/39) in the following form with the aid of equation ({6.}1/13)

$$P_s = \frac{2}{3}\pi Z_0 \tilde{I}^2 \left(\frac{\Delta l}{\lambda}\right)^2 = 790\,\Omega\,\tilde{I}^2 \left(\frac{\Delta l}{\lambda}\right)^2 \qquad (6.1/42)$$

and thus obtain the following for the radiation resistance of the Hertzian dipole in free space:

$$R_s = 790\,\Omega \left(\frac{\Delta l}{\lambda}\right)^2. \qquad (6.1/43)$$

Δl is the length of the Hertzian dipole. As we would expect, both P_s and R_S are independent of r. The radiation resistance calculated here is referred to the feed point which carries the current $\tilde{I}$. In case of antennas with an irregular current distribution, we must take into consideration whether the radiation resistance is referred to the feed point, the base point or the location of the current maximum [*alternative*: "antinode"]. The radiation resistance R_S referred to the current maximum is naturally the smallest. It can be converted from the current maximum (current $\tilde{I}_0$) to another antenna location with separation y (current $\tilde{I}_y$):

$$\tilde{I}_y^2 R_{sy} = \tilde{I}_0^2 R_s = P_s,$$
$$R_{sy} = \left(\frac{\tilde{I}_0}{\tilde{I}_y}\right)^2 R_s. \qquad (6.1/44)$$

Given a sinusoidal current distribution

$$\tilde{I}_y = \tilde{I}_0 \cos\frac{2\pi y}{\lambda} \qquad (6.1/45)$$

we have

$$R_{sy} = \frac{R_s}{\cos^2\frac{2\pi y}{\lambda}}. \qquad (6.1/46)$$

We will now consider one half of a Hertzian dipole over earth [*alternative*: "ground"] as shown in **Figure {6.}1/9**. Part of the radiation arrives at point P via the direct path and another part is reflected by the ground. We assume the ground has ideal conductivity such that total reflection occurs. The direct and reflected radiation thus arrive at point P as if they came from the antenna and its mirror image in the ground with separation $\Delta l = 2\Delta h$.

The current direction in the mirror image generally follows from **Figure {6.}/10**. We assume that the current in the radiator is $\mathbf{I} = I_v e_v + I_h e_h$ and in the mirror image it is $\mathbf{I'} = I_v' e_v + I_h' e_h$. Moreover, we assume that $\mathbf{E} = E_v e_v + E_h e_h$ is the electric field strength before the ground reflection and $\mathbf{E'} = E_v' e_v + E_h' e_h$ is that after the ground reflection. The horizontal component of the field strength must disappear at the earth's surface with its good conductivity:

$$E_h + E_h' = 0 .$$

In other words, we must have $I_h = -I_h'$. However, the vertical component is common to both waves:

$$E_v = E_v'$$

In other words, we also have

$$I_v = I_v'$$

In order to calculate the radiated power P_s' and the radiation resistance R_S' of our one half of the dipole with respect to ground, we now only need to integrate over the upper half sphere, i.e. only within the limits $\theta = 0$ to $\theta = \pi/2$. We thus obtain

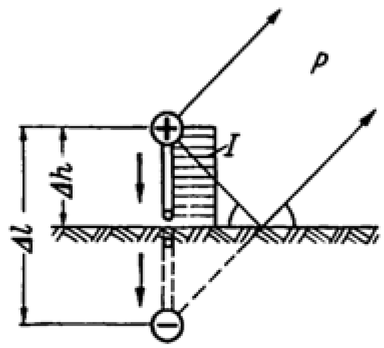
$$P_s' = \frac{1}{3}\pi Z_0 \tilde{I}^2 \left(\frac{2\Delta h}{\lambda}\right)^2 = 1580\,\Omega \tilde{I}^2 \left(\frac{\Delta h}{\lambda}\right)^2 \qquad (6.1/47)$$

and

$$R_s' = 1580 \left(\frac{\Delta h}{\lambda}\right)^2 \Omega . \qquad (6.1/48)$$

The radiation pattern is a half torus [*alternative*: "toroid"]. Given that $P_s' = P_s/2$, we obtain the field strength over level ground without losses from equation ({6.}/40):

$$\frac{\tilde{E}'_{max}}{\text{mV/m}} = 300 \sqrt{\frac{P_s'}{\text{kW}}} \frac{1}{r/\text{km}} . \qquad (6.1/49)$$



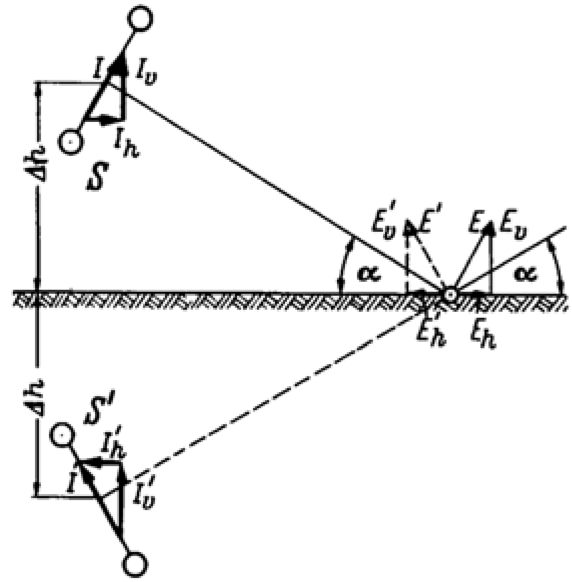


Abb. 6.1/9. Hertzscher Dipol über einer leitenden Ebene

Abb. 6.1/10. Bestimmung der Stromrichtung im Spiegelbild *S'* des Leiters *S*

Figure {6.}1/9. Hertzian dipole over a conductive plane

Figure {6.}1/10. Determination of the current direction in the mirror image *S'* of conductor *S*

In the transmitting case, the radiated power P_s is less than the electric power P_{s0} fed to the antenna. The difference $P_v = P_{s0} - P_s$ represents the thermal dissipation loss in the antenna. The antenna efficiency in the transmitting case is

$$\eta_s = P_s/P_{s0} = P_s/(P_v + P_s)\ .$$

Analogously, in the receiving case the power P_e absorbed by the antenna from the radiation field is greater than the power P_{e0} that can be transferred by the antenna to the load in the case of power matching [*alternative*: "impedance matching"]. We have $P_v = P_e - P_{e0}$ and

$$\eta_e = P_{e0}/P_e = P_{e0}/(P_v + P_{e0})\ .$$

The current distribution on the antenna in the transmitting case differs from the receiving case because the near fields differ.

The “antenna gain” *G* is defined as the product of the directivity and the antenna efficiency:

$G = D\eta$.

The receive antenna is assumed to be matched to the load impedance. Any mismatches are taken into account in the “practical antenna gain of a receive antenna” G_p. If an optimally oriented and lossless Hertzian dipole or half-wave dipole is used as the reference radiator, we obtain $G_{Hz} = G/1.5$ and $G_D = G/1.64$, respectively. The gain is typically specified in dB:

$$\frac{g}{\mathrm{dB}} = 10 \lg G \,.$$

In actual practice, the gain of an antenna is determined by comparing it with a $\lambda/2$ dipole or based on the transmission efficiency of a path. In principle, it can be determined computationally by integrating the radiation pattern; however, considerable errors can arise in the case of antennas with high gain if the side lobes[5] cannot be taken into account down to very small values.

For antennas with linear dimensions on the order of magnitude of the wavelength or greater, we use the “effective area” A to characterize the directivity properties. For receiving antennas, A is also known as the “absorption area”. In the receiving case,

$$A = P_{\mathrm{e\,max}}/S \qquad (6.1/64)$$

is an area[6] perpendicular to the propagation direction through which the maximum radiation receive power of the antenna $P_{e,\max}$ would pass if exposed to an undisturbed plane wave with radiation density S. If the reciprocity theorem holds, the effective area in the transmitting case is equal to the absorption area in the receiving case.

If we connect an assumed lossless Hertzian dipole to a receiver with input resistance R_e, then the power absorbed from the field is

$$P_{\mathrm{e}} = \frac{R_{\mathrm{e}} \Delta l^2}{(R_{\mathrm{e}} + R_{\mathrm{s}})^2} \tilde{E}^2 \qquad (6.1/65)$$

and for $R_e = R_S$ (power matching) we have

$$P_{\mathrm{e\,max}} = \frac{\Delta l^2}{4R_{\mathrm{s}}} \tilde{E}^2 \,. \qquad (6.1/66)$$

[5] Translator’s note: Literally "auxiliary maxima".

[6] Translator’s note: Would "surface" be more logical?

For the very short dipole (see section 6.2), we have

$$R_S = 80\pi^2 \left(\frac{\Delta l}{\lambda}\right)^2 \Omega$$

such that

$$P_{e\,max} = \frac{\tilde{E}^2}{Z_0}\frac{3\lambda^2}{8\pi} = SA_{Hz}\,. \qquad (6.1/67a)$$

In conjunction with equation (6.1/38), the effective antenna area of the Hertzian dipole is thus

$$A_{Hz} = \frac{3}{8\pi}\lambda^2\,. \qquad (6.1/67b)$$

The ratio of the effective area A to the aperture area A_g [*alternative*: "physical aperture"] of an antenna

$$q = A/A_g \qquad (6.1/67c)$$

is known as the "aperture efficiency". A_g usually represents the geometrical area of the radiant opening of an aperture antenna (see section 6.4). Assume we have two antennas 1 and 2 with mutual separation r. In the transmitting case, we assume power P_{s0} is supplied to the antenna, and in the receiving case the antenna delivers power P_{et} to a load from the transmit field in case of optimal orientation towards the transmitting antenna. Both antennas are assumed to be lossless. Assume the transmitting antenna has directivity D_s and the receiving antenna has effective area A_E. Moreover, we assume the propagation medium between the two antennas is lossless and isotropic and exhibits transmission symmetry. For free wave propagation and under far-field conditions, we then have the following for the transmission direction from 1 to 2:

$$P_{et}^{(2)} = \frac{P_{s0}^{(1)}}{4\pi r^2} D_s^{(1)} A_E^{(2)}\,.$$

If we swap the roles of the transmitting and receiving antennas, i.e. reverse the transmission direction, the following must hold:

$$P_{\mathrm{et}}^{(1)} = \frac{P_{\mathrm{s0}}^{(2)}}{4\pi r^2} D_{\mathrm{s}}^{(2)} A_{\mathrm{E}}^{(1)} .$$

Setting TERM-1 equal to TERM-2, due to the assumed transmission symmetry TERM-3 must then also equal TERM-4 such that

TERM-1: $P_{\mathrm{s0}}^{(1)}$ TERM-2: $P_{\mathrm{s0}}^{(2)}$ TERM-3: $P_{\mathrm{et}}^{(2)}$ TERM-4: $P_{\mathrm{et}}^{(1)}$

$$D_{\mathrm{s}}^{(1)} A_{\mathrm{E}}^{(2)} = D_{\mathrm{s}}^{(2)} A_{\mathrm{E}}^{(1)} . \qquad (6.1/67\mathrm{d})$$

Now let us assume that antenna (1) is a monopole and antenna (2) is a Hertzian dipole. From equation ({6.}/67d) in conjunction with equation ({6.}/67b), it then follows for the effective area A_K of the imaginary monopole that

$$1 \cdot \frac{3}{8\pi} \lambda^2 = \frac{3}{2} A_{\mathrm{K}} ,$$

$$A_{\mathrm{K}} = \frac{\lambda^2}{4\pi} . \qquad (6.1/67\mathrm{e})$$

Now that we know the effective area of the monopole, we can obtain the important relationship for any arbitrary antenna between its effective area and its directivity:

$$A = \frac{\lambda^2}{4\pi} \cdot D . \qquad (6.1/67\mathrm{f})$$

Taking into account the antenna losses, the effective area A is replaced by the effective area[7] $A_W = \eta A$:

$$A_{\mathrm{W}} = \frac{\lambda^2}{4\pi} \eta D = \frac{\lambda^2}{4\pi} G . \qquad (6.1/67\mathrm{g})$$

[7] Translator's note: The German text literally says "effective area" for A_W.

We thus obtain the fundamental equation for power transmission on a radio link which is known as the Friis transmission equation:

$$P_{et} = P_{s0}\frac{A_{Ws}A_{WE}}{\lambda^2 r^2} = P_{s0}G_s G_E\left(\frac{\lambda}{4\pi r}\right)^2 . \qquad (6.1/68)$$

The expression

$$a = 10 \lg \frac{P_{s0}}{P_{e\theta}} \qquad (6.1/69)$$

is known as the "path attenuation". We can rearrange it according to equation (6.1/68) to obtain

$$a = 20 \lg \frac{4\pi r}{\lambda} - 10 \lg G_s G_E$$

Here,

$$a_0 = 20 \lg \frac{4\pi r}{\lambda} \qquad (6.1/70)$$

is the "fundamental transmission attenuation"[8] of the radio link. *a* and a_0 are specified in dB.

In practical usage, we often split the fundamental transmission attenuation into individual summands and directly take the logarithm for fixed numerical values that arise. For example, we might obtain

$$a_0/\mathrm{dB} = 81{,}98 + 20 \lg \frac{R}{\mathrm{km}} - 20 \lg \frac{\lambda}{\mathrm{m}}$$
$$= 92{,}44 + 20 \lg \frac{R}{\mathrm{km}} + 20 \lg \frac{f}{\mathrm{GHz}} . \qquad (6.1/71)$$

On the other hand, the attenuation of a normal coaxial cable as shown in **Figure {6.}2.1/5** can be expressed with sufficiency accuracy in conjunction with equation ({6.}2.1/22) using the relationship

$$a_{max}/\mathrm{dB} \approx 2{,}35\sqrt{\frac{f}{\mathrm{MHz}}} \cdot \frac{R}{\mathrm{km}} . \qquad (6.1/72)$$

[8] Translator's note: One textbook I consulted called this "path loss".

By way of illustration, the two attenuation laws expressed in equations ({6.}.1/71) and ({6.}.1/72) are plotted in **Figure {6.}1/14** for an operating frequency f = 100 MHz. Among other things, diagrams of this sort are a useful tool when deciding between a cable link and a radio link. Of course, we can easily reduce the attenuation of a radio link by using transmitting and receiving antennas with suitable gain. If we say that G_g is the geometric mean of the two gain values, we obtain the attenuation of the radio link in **Figure {6}.1/14** by subtracting the value 20log G_g dB from a_0.

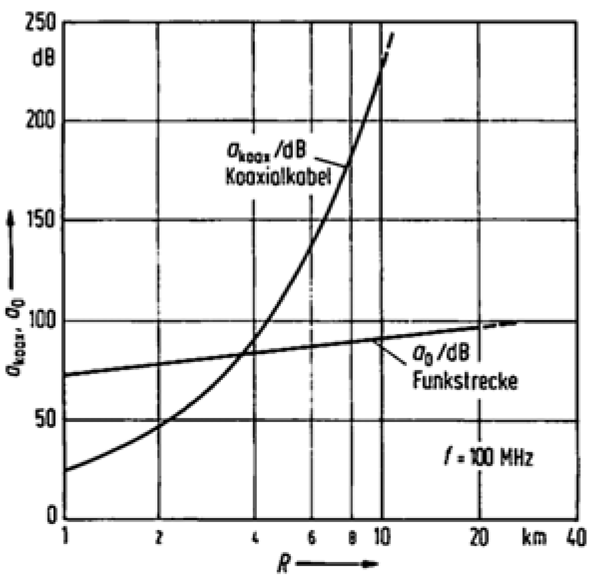


Abb. 6.1/14. Grundübertragungsdämpfung a_0 einer Funkstrecke und Dämpfung a_{koax} eines Normalkoaxialkabels als Funktion des Abstandes R. Betriebsfrequenz f = 100 MHz

Fig {6.}1/14. Fundamental transmission attenuation a_0 of a radio link and attenuation a_{coax} of a normal coaxial cable as a function of distance R. Operating frequency f = 100 MHz

["Koaxialkabel" --> "Coaxial cable", "Funkstrecke" --> "Radio link"]

6.2.2.3 Effective height of electrically short antennas.

The open-circuit voltage V_0 of an antenna is proportional to the antenna field strength E where the antenna is located:

$$U_0 = h_{eff} E \,. \qquad (6.2/11)$$

The proportionality factor h_{eff} has the dimension of length and is known as the "effective height". If the current in the antenna is location-independent (Hertzian dipole), then h_{eff} corresponds to the antenna's geometrical length l. Otherwise, the effective height is less due to the non-uniform current distribution. In this general case, we obtain h_{eff} by converting the current area into a rectangle with the same area having the maximum current I_0 as its base as shown in **Figure {6.}.2/4**. Its height is then equal to h_{eff}. Computationally, we have

$$I_0 h_{eff} = \int_0^h I_y \, dy \,, \qquad h_{eff} = \int_0^h \frac{I_y}{I_0} \, dy \,. \tag{6.2/12}$$

The definition of the effective height is closely related to that of the effective area A (see section 6.1.7):

$$A = \frac{h_{eff}^2}{4} \frac{Z_0}{R_s} \,, \qquad h_{eff} = 2\sqrt{A \frac{R_s}{Z_0}} \,. \tag{6.2/13}$$

6.2.3 λ/4 and λ/2 antennas over ground

Now that we have examined the properties of electrically short antennas where $l \leq \lambda/8$, we will turn our attention to longer antennas. Like for the transmission line, we will first consider the case in which $l = \lambda/4$. For the vertical radiation pattern of the $\lambda/4$ antenna over ground, we obtain the following from equation ({6.}.2/4):

$$F\left(\vartheta; \frac{l}{\lambda} = \frac{1}{4}\right) = \frac{\cos\left(\frac{\pi}{2}\cos\vartheta\right)}{\sin\vartheta} \,. \tag{6.2/19}$$

We can determine the radiated power and the radiation resistance of the $\lambda/4$ antenna over ground based on the power radiated in the half sphere:

$$P_s = \tilde{I}_0^2 R_s = \int_{\vartheta=0}^{\vartheta=\pi/2} \frac{\bar{E}_\vartheta^2}{Z_0} \cdot 2\pi r^2 \sin\vartheta \, d\vartheta = \tilde{I}_0^2 \cdot 60\,\Omega \int_0^{\pi/2} F^2\left(\vartheta, \frac{l}{\lambda}\right) \sin\vartheta \, d\vartheta \,,$$

$$P_s = \tilde{I}_0^2 \cdot 60\,\Omega \int_0^{\pi/2} \frac{\cos^2\left(\frac{\pi}{2} cos\,\vartheta\right)}{\sin^2\vartheta} \sin\vartheta \, d\vartheta \,,$$

$$R_s = 60\,\Omega \frac{C + \ln 2\pi - \mathrm{Ci}(2\pi)}{4} = 60\,\Omega \cdot 0{,}61 = 36{,}6\,\Omega \,; \tag{6.2/20}$$

Here, $C = 0.5772$ is Euler's constant and $C_i(x)$ is the cosine integral:

$$\mathrm{Ci}(x) = \int_{\infty}^{x} \frac{\cos u}{u}\,\mathrm{d}u, \quad \text{z. B.} \quad \mathrm{Ci}(2\pi) = -0{,}0225; \quad \mathrm{Ci}(4\pi) = -0{,}0061$$

We will now perform the analogous calculations for the $\lambda/2$ antenna. We can obtain its vertical pattern from equation ({6.}.2/4):

$$F\left(\vartheta; \frac{l}{\lambda} = \frac{1}{2}\right) = \frac{\cos(\pi\cos\vartheta) + 1}{\sin\vartheta} = \frac{2\cos^2\left(\frac{\pi}{2}\cos\vartheta\right)}{\sin\vartheta}. \tag{6.2/21}$$

We obtain the radiation resistance using the same approach as above:

$$P_s = \tilde{I}_0^2 R_s = \tilde{I}_0^2 \cdot 60\,\Omega \int_0^{\pi/2} \frac{[\cos(\pi\cos\vartheta) + 1]^2}{\sin^2\vartheta} \sin\vartheta\,\mathrm{d}\vartheta\,,$$

$$R_s = 60\,\Omega\,\tfrac{3}{4}(C + \ln 2\pi - \tfrac{4}{3}\mathrm{Ci}(2\pi) + \tfrac{1}{3}\mathrm{Ci}(4\pi) - \tfrac{1}{3}\ln 2)$$

$$= 60\,\Omega \cdot 1{,}66\,, \tag{6.2/22}$$

$$R_s = 99{,}5\,\Omega \approx 100\,\Omega\,. \tag{6.2/23}$$

Note that this radiation resistance is referred to the current maximum while the feed point (base) is located at the voltage maximum. The base resistance R_F differs similarly from R_S like the input and termination impedance of a $\lambda/4$ line:

$$R_F = \frac{Z_a^2}{R_s}\,. \tag{6.2/24}$$

$Z_a = \sqrt{L_A/C_A}$ is the line characteristic impedance of the antenna with respect to ground:

$$Z_a \approx 60\,\Omega\left(\ln\frac{2l}{d} - 0{,}6\right). \tag{6.2/25}$$

$2l/d$ is the antenna's slenderness ratio (d = diameter of the antenna conductor).

In our discussion so far, we assumed a sinusoidal current distribution on the antenna. This is acceptable only for slender antennas for which l/d is large. In reality, the current distribution is not perfectly sinusoidal because the antenna behaves like a highly attenuated line. The attenuation [*alternative*:

"loss"] is primarily a consequence of the radiated power and can be taken into account by assuming that the radiation resistance is distributed over the entire antenna length. See for further information.

Figure {6.}.2/7 shows the radiation resistance of a vertical antenna over ground as a function of l/λ. – Since they are thin with respect to their length, all of the antennas discussed hitherto are narrowband. As shown in **Figure {6.}.2/8**, the plot of the resistance of an antenna of this sort between the points of current resonance ($l \approx \lambda/4$ and $3/4\lambda$) and voltage resonance ($l \approx \lambda/2$) has a very wide opening. The equivalent circuit of the antenna at the voltage resonance point is a trap circuit with the input resistance

$$R_F = \frac{Z_a^2}{R_{s(\lambda/2)}} \approx \frac{Z_a^2}{100\,\Omega}, \qquad (6.2/26)$$

Here, Z_a is the antenna's characteristic impedance and R_S is its radiation resistance. The circle becomes tighter as we decrease Z_a^2/R_S. If we increase the ratio d/l (curve B), then Z_a decreases while R_S remains practically unchanged. However, the antenna attenuation increases along with the bandwidth. For broadband antennas, Z_a^2/R_S should be as close as possible to 60 Ω. Antennas of this sort operate in a frequency range of 3:1 or more; they are implemented in the form of cylinders, cones, rotational ellipsoids (see **Figure {6.}2/9a**) and cup antennas (see **Figure {6.}2/9b**).

Since the reactance of a $\lambda/2$ dipole corresponds to that of a parallel resonant circuit, we can compensate the frequency response of the input impedance in a narrow frequency range by connecting a series resonant circuit to the input (e.g. an open $\lambda/4$ line) and thus compress the response (curve *C* in **Figure {6.}2/8)**.

Treating the antennas like lines, we ascertain that the quantities per unit length L' and C' are location-dependent. This is because thick antennas exhibit voltage resonance at lengths less than $\lambda/2$. For details on the resulting length of thick $\lambda/2$ dipoles.

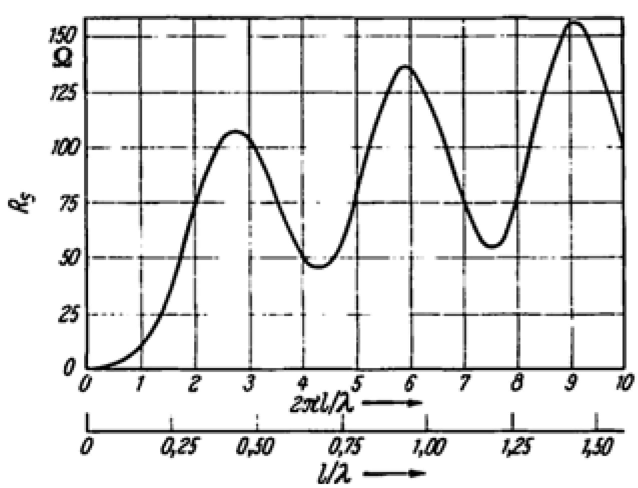


Abb. 6.2/7. Verlauf des Strahlungswiderstandes einer dünnen Antenne über Erde

Figure {6.}2/7. Plot of the radiation resistance of a thin antenna over ground

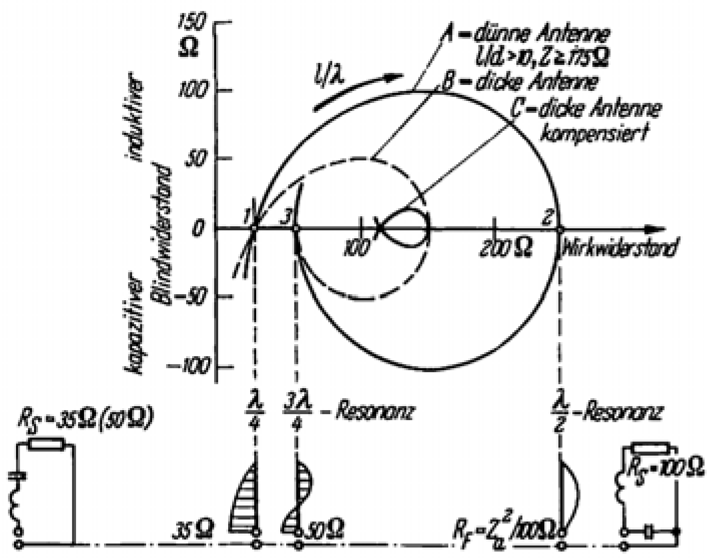


Abb. 6.2/8. Ortskurve des Eingangswiderstandes von Antennen verschiedenen Schlankheitsgrades über Erde

Figure 2{6.}/8. Plot of the input resistance of antennas with various slenderness ratios over ground

["A = dünne Antenne …" --> "A = Thin antenna …", "B = dicke Antenne" --> "B = Thick antenna", "*C* = dicke Antenne kompensiert" --> "*C* = Thick antenna compensated", "kapazitiver / induktiver Blindwiderstand" --> "Capacitive / Inductive reactance", "Wirkwiderstand" --> "Effective resistance", "…-Resonanz" --> "… resonance"]

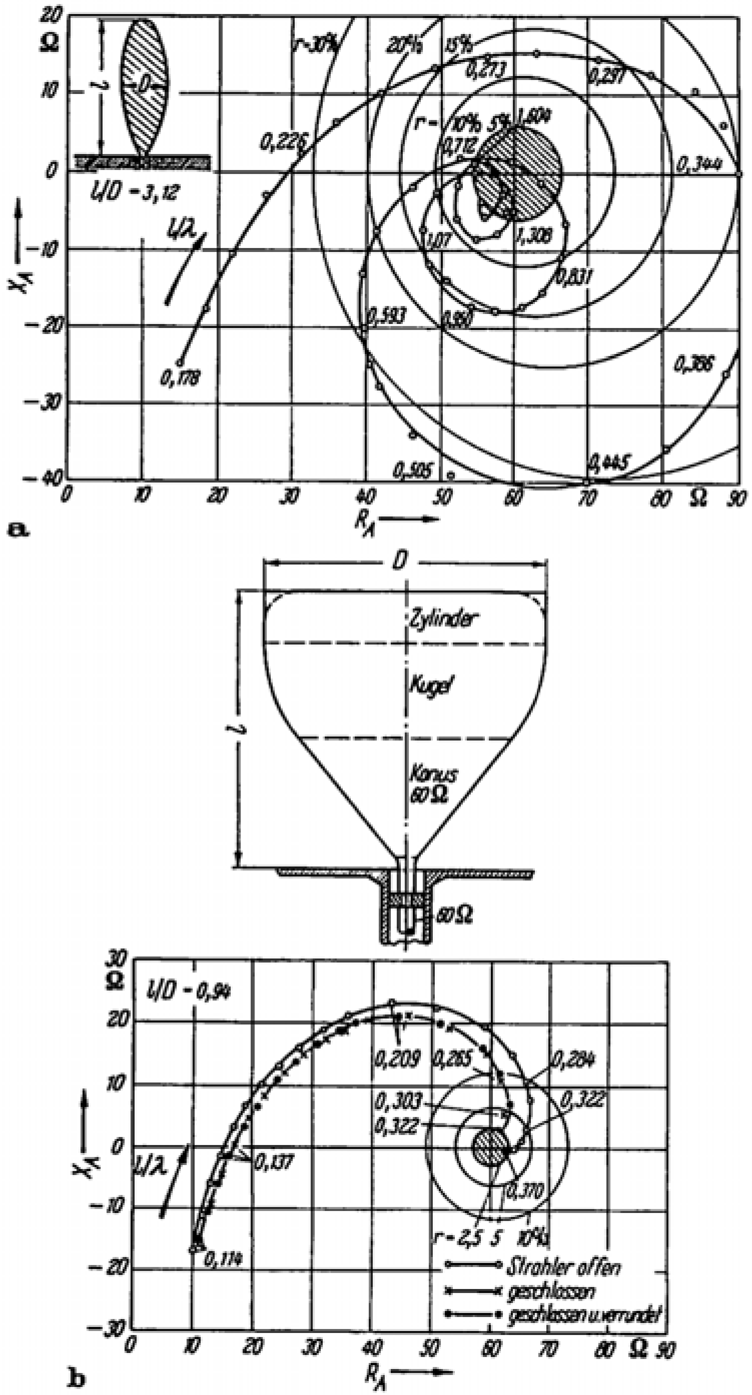


Abb. 6.2/9a, b. Ortskurven von Breitbandantennen. a Ellipsoidstrahler; b Kelchstrahler

Figure {6.}2/9a,b. Curves for broadband antennas. **a** Ellipsoidal antenna; **b** Cup antenna

["Zylinder" --> "Cylinder", "Kugel" --> "Sphere", "Konus" --> "Cone", "Strahler offen" --> "Antenna open", "geschlossen" --> "Closed", "geschlossen u. verrundet" --> "Closed & rounded"]

2.6 Loop Antennas; Circular Antennas

Loop antennas are realized in a square or round shape with one or more windings. The infinitesimal loop equation field components are valid for a *loop dimension* very much smaller than a wavelength *and* very much smaller than the distance to the field point, that is, the loop is infinitesimally small.

According to equation ({6.}1/28), the receive voltage for a loop antenna is

$$U_m = -jE\frac{2\pi wA}{\lambda}\sin\vartheta \,. \tag{6.2/30}$$

If the loop circuit is tuned using a variable capacitor, then the terminal voltage in case of resonance is

$$U_{max} = \frac{\omega L}{R} U_m \,, \tag{6.2/31}$$

where *R* (the effective series resistance) essentially contains the loop resistance at the operating frequency. Since according to section 6.2.2.3 the induced electromotive force is $V_m = h_{eff} E$, we have the following for the loop:

$$h_{eff} = \frac{2\pi wA}{\lambda} \,. \tag{6.2/32}$$

The effective height is extremely small; for a loop with an area of 1 m^2 and 10 windings, it is only 12 cm for λ = 500 m. Nevertheless, the loop antenna provides usable terminal voltages because the factor $\omega L/R$ in equation ({6.}2/31) has values of 50 to 100.

The number of windings *w* cannot be arbitrarily large since the loop's total wire length must be less than one sixth of the shortest receive wavelength λ_{min}. For a circular loop with diameter *D*, at best we obtain

$$\pi D w \approx \frac{\lambda_{min}}{6} \,. \tag{6.2/33}$$

We thus have

$$h_{eff} \approx \frac{\lambda_{min}^2}{70 w \lambda} \approx \frac{D}{4}\frac{\lambda_{min}}{\lambda} \tag{6.2/34}$$

or for $\lambda = \lambda_{min}$

$$h_{eff\,max} \approx \frac{\lambda_{min}}{70 w} \approx \frac{D}{4} \,. \tag{6.2/35}$$

For a large effective height, we must increase the diameter at the expense of the number of windings. If the dimensions of the loop are on the order of the wavelength, more complicated radiation patterns are obtained compared to the loops discussed above.

Loops with dimensions $D \ll \lambda$ are used for direction-finding. The main disruption that occurs is the "antenna effect" associated with the loop since overall the loop's metallic material forms a short open antenna with respect to the wavelength. The voltage for this antenna leads the loop's direction-dependent induced voltage by 90° and thus "clouds" the nulls **(Figure {6.}2/17)**. This antenna effect can be reduced by using a shielded loop or one that is balanced with respect to ground. However, the disruptive loop voltage is typically compensated with a voltage of equal magnitude from an auxiliary antenna ("clarification")[9].

One significant disadvantage of the loop direction-finder is its susceptibility to the "night effect". In other words, if the wave arrives after reflection on the ionosphere at an elevation angle ψ and with a polarization plane rotated by the angle γ, additional voltages will be induced by the horizontal field component in the loop's horizontal conductor sections. For an azimuth angle φ, the total voltage is then

$$U_\mathrm{m} = E \cdot 2\pi w \frac{A}{\lambda} (\sin\varphi \cos\gamma + \cos\varphi \sin\varphi \sin\gamma) \ . \qquad (6.2/36)$$

Depending on the magnitude of ψ and γ, arbitrary patterns will arise resulting in direction-finding errors.

The pattern's two minima are offset by 180°, thereby causing ambiguity in direction-finding. In order to disambiguate the direction, the loop pattern is thus overlaid with an omnidirectional pattern to obtain a cardioid with a broad null (**Figure {6.}2/18**). **Figure {6.}2/19** shows a circuit diagram for a direction-finding receiver (Telefunken) in which a single auxiliary antenna can be selected for clarification or disambiguation.

Instead of rotating the loop during direction-finding, two loops fixed at an angle of 90° can also be used. They feed two coils which are spatially offset by 90° in whose field a "search coil" is arranged in a rotatable manner. The pattern for this so-called "goniometer direction-finder" is also a double-circle pattern.

Although the loop characteristic in the plane of the loop corresponds to an omnidirectional pattern, the simple loop is not suitable for use as an omnidirectional transmitting antenna. Since a good omnidirectional pattern requires highly uniform and in-phase linear current density, a loop of this sort would have to be small with respect to wavelength. However, the radiation resistance is then too low. This difficulty can be circumvented by increasing the number of feed points ("circular antennas"). As examples of such antennas, **Figure {6.}/20a** shows a dipole square and **Figure {6.}2/20b** the "Alford loop". Due to the irregular current distribution, the patterns are not perfectly circular. In terms of their operation, circular antennas can be likened to arrangements with in-phase excitation in a circular

[9] Translator's note: Literally "unclouding".

configuration in a plane. **Figure {6.}2/21** has two examples. Circular antennas have roughly the same gain as a $\lambda/2$ dipole. In order to increase the gain in the horizontal plane, several of these antennas can be arranged on top of one another. If very high gain is required, omnidirectional radiators made of unit fields can be used as described in section 6.2.11.

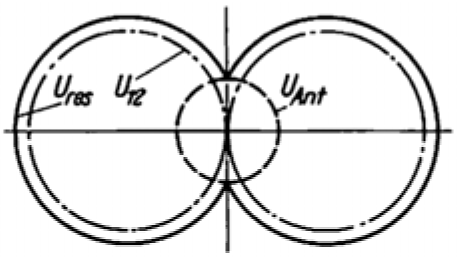


Abb. 6.2/17. Gestörte Charakteristik eines Rahmens infolge des Antenneneffektes

Figure 2/17. Distorted characteristic for a loop due to the antenna effect.

Here is an overview of the effective height of different antennas:

Antenna	Radiation resistance ohms	Source
Center-fed half-wave dipole	73.1[18]	Kraus 1988:227,Balanis 2005:216
Short dipole of length $\lambda/50 < L < \lambda/10$	$20\pi^2\left(\frac{L}{\lambda}\right)^2$	Kraus 1988:216, Balanis 2005:165,215
Base-fed quarter-wave monopole over perfectly conducting ground	36.5	Balanis 2005:217, Stutzman & Thiele 2012:80
Short monopole of length $L \ll \lambda/4$ over perfectly conducting ground	$40\pi^2\left(\frac{L}{\lambda}\right)^2$	Stutzman & Thiele 2012:78-80
Resonant loop antenna, 1 λ circumference	~100	Weston 2017:15, Schmitt 2002:236
Small loop of area A with N turns (circumference $\ll \lambda/3$)	$320\pi^4\left(\frac{NA}{\lambda^2}\right)^2$	Kraus 1988:251, Balanis 2005:238
Small loop of area A with N turns on a ferrite core of effective relative permeability μ_{eff}	$320\pi^4\left(\frac{\mu_{\text{eff}}NA}{\lambda^2}\right)^2$	Kraus 1988:259, Milligan 2005:260

The small loop radiation resistance for the very small loop (second to last entry in the Table) can also be written more conveniently in terms of C_λ, the loop circumference in wavelengths,

$$R_{RAD} = 197.3C_\lambda^4$$

where we consider an N = 1 turn loop. The equation assumes a constant current around the loop. A more accurate expression for the current (valid for C_λ, < 0.4) in the gap-fed loop is,

$$I(\phi) = \left\{1 - 2C_\lambda{}^2 \cos(\phi)\right\}$$

where ϕ is the angle around the loop circumference, and the feed gap is at ϕ = 0. Furthermore we can predict the far field null depth using this two term Fourier expansion for the current, and also obtain the wave impedance at the center of the loop. It is easy to show that for this current approximation the *exact* expression for the null depth is

$$N_{dB} = -20\log(2C_\lambda)$$

and the *exact* wave impedance at the loop center is,

$$Z_W = \frac{E_{center}}{H_{center}} = -j\eta_0 C_\lambda$$

where η_0 = 376.73 Ω, the free space intrinsic impedance. Thus rational criteria can be chosen for deciding when a small loop can qualify as a "magnetic field probe" (see K. Siwiak, KE4PT, "Small HF Loop Antennas are not Magnetic Loops," Mar. 2020, *QEX* (Technical Notes), p. 24). If, for example, we require that the magnetic field probe have a far field null depth of at least 20 dB, than C_λ must be less than 0.05, which will result in a near-field wave impedance of 5% of the free space intrinsic impedance, and a maximum current deviation of 0.5% from a constant value.

Practical Antennas

The dipole is a very useful antenna, as the current radiates perpendicularly to the wire, some energy gets radiated in the air useful for NVIS operation (near vertical incident skywave). NVIS operation can occur when the dipole antenna is close to the ground and horizontally polarized), and certain ionospheric conditions are met. The image below shows a pair of horizontal dipoles That have built-in tuners.

That is, each side has two dipole which puts the antenna in the category of two thick dipoles or fan dipoles. In the background for DX operation is the StepperIR antenna and the vertical antenna is "hidden" in the flagpole.

Below is the theoretical impedance curve of a dipole in free space. As the antenna wire is fairly thick, the resonant impedance is at maximum 2000 Ω reactive and resistive. In practice the wire is more like 2 mm or 0.1 inch, which results in much higher impedances, up to 20,000 Ω.

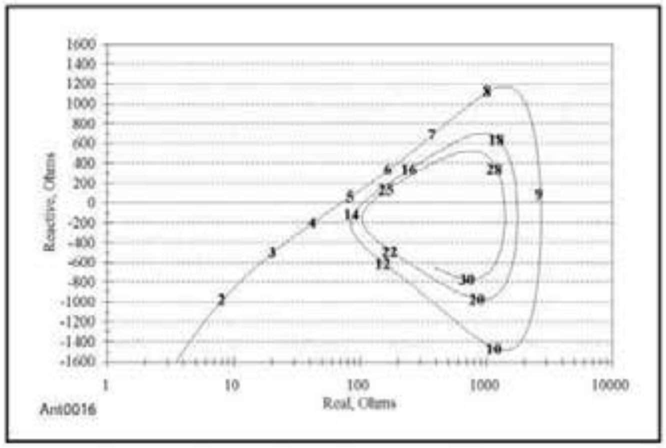
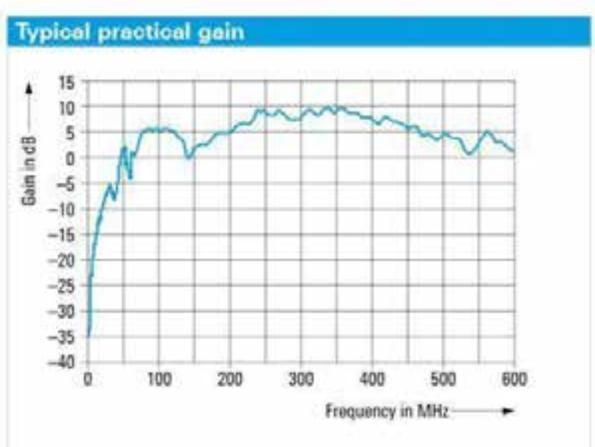


Figure 2.11 — Feed point impedance versus frequency for a theoretical 100-foot long dipole in free space, fed in the center and made of thick 1.0-inch diameter wire. Once again, the excursion in both reactance and resistance over the frequency range is less with this thick wire dipole than with thinner ones.

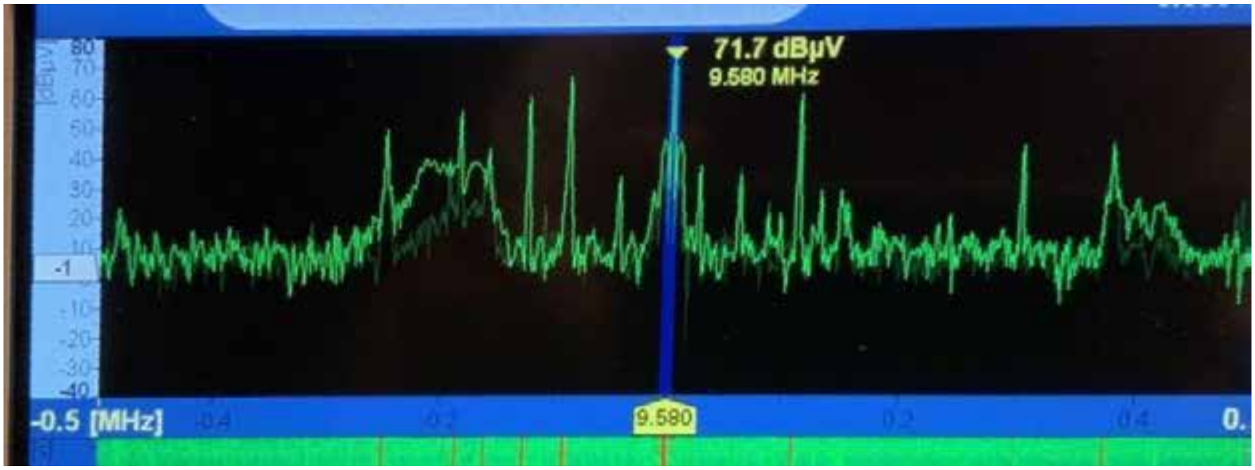
[Source: ARRL Antenna Book.]

As we now move to the vertical antenna, the following picture below shows a vertical 23 feet tall antenna and next to it a combination active receiving antenna (actually two antennas in one housing) from 9 kHz to about 600 MHz .The R&S®HE055 active omnidirectional receiving antenna has an extremely wide frequency range from 1.5 MHz to 600 MHz.



This receive only antenna has a protection circuit so when transmitting it does not get destroyed.

Vertical rod or whip antennas have their place, as seen above, this vertical is driven by an antenna tuner, it works very well from 7 MHz to 50 MHz. For amateur bands 1.8 to 4 MHz it's a bit short, but given the environmental noise it works sufficiently well in receive. In transmit the efficiency is about 2.5 %. This antenna will also be used to compare with the small HF loop antenna operating above 7 MHz.



The picture above shows the frequency occupancy of the 9.68 MHz broadcast band (±500 kHz).The evening noise level in a 2.4 kHz bandwidth is around 10 dBμV

A shorter VHF/UHF antenna version while using a Manpack radios is convenient but lossy as the housing of the radio is too small compared to a wavelength. That applies also to all hand-held radios.

At home the Manpack on the balcony is fun .

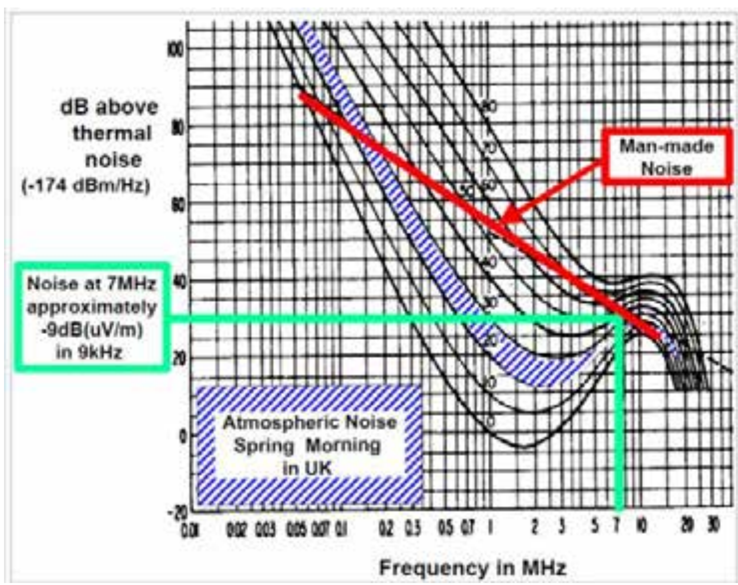
And on the sailboat it is easily placed.

Some Real Life Tests

Let us remember that the receiver with an antenna that collects voltage needs to translate the field strength into a signal (typically into a 50 Ω termination). The field strength in V/m needs to be multiplied with the effective antenna height. The source impedance is 377 Ω, as we learned .This is practically an open circuit voltage, which collapses in value at 50 Ω termination significantly dependent on the antenna length. In the case of the active antenna the 1 m tall antenna rod is terminated into an FET with much higher input impedance, several 1000 Ω,

This system has a power gain of about 8.7 dB.

From a practical point the absolute input voltages is less relevant. What is important is the signal/noise ratio. The 1 m tall active antenna at 7 MHz is definitely to short, and is not resonant. A quarter wave long antennas would be 10.4 m long and the 1 m vertical collects only 10% of the available signal but also only a tenth of the environmental noise. Most CCIR ambient noise publications are outdated because recent modern RF noise sources like air conditions, refrigerators and other appliances that generate a lot of electric noise. The wideband measurement shown above clearly indicate this.



Most use resonant dipoles, so the number in dBμV/m is useless, for 40 m one must multiply this number by 10 or add 10 dB to the power reading. This now has also to be related to the receiver bandwidth. Away from the house there are still enough high tension wires another electrical noisemakers that things don't change too much.

It turned out that the ham station normally do not transmit long enough for good measurements so we went to the 7.335 MHz broadcast frequency where we found Radio Cuba with fairly constant signal, and little fading . Other relevant tests were done at slightly below 10 MHz.

The active antenna next to the chimney is connected to an R&S monitoring receiver, type ESMD, calibrated in dBμV and an input impedance of 50 Ω. The noise level was as expected to be about 10 dBμV/2.4 kHz bandwidth. For the broadcast station at 7.335 the reading was 50 dBμV=300 μV (into 50 Ω), the 9.58 MHz station was 71.7 dBμV. If we use the definition of S 9 =50 uV= 34dBμV, then the signal absolute is 60 dBμV=1mV x 3.9 ~ 4 mV. 71.7 dBμV -34dBμV ~ 38 dB above S 9. Most modern receiver can handle this easily.

For most of the tests, the Rohde & Schwarz monitoring receiver were used, see picture above.

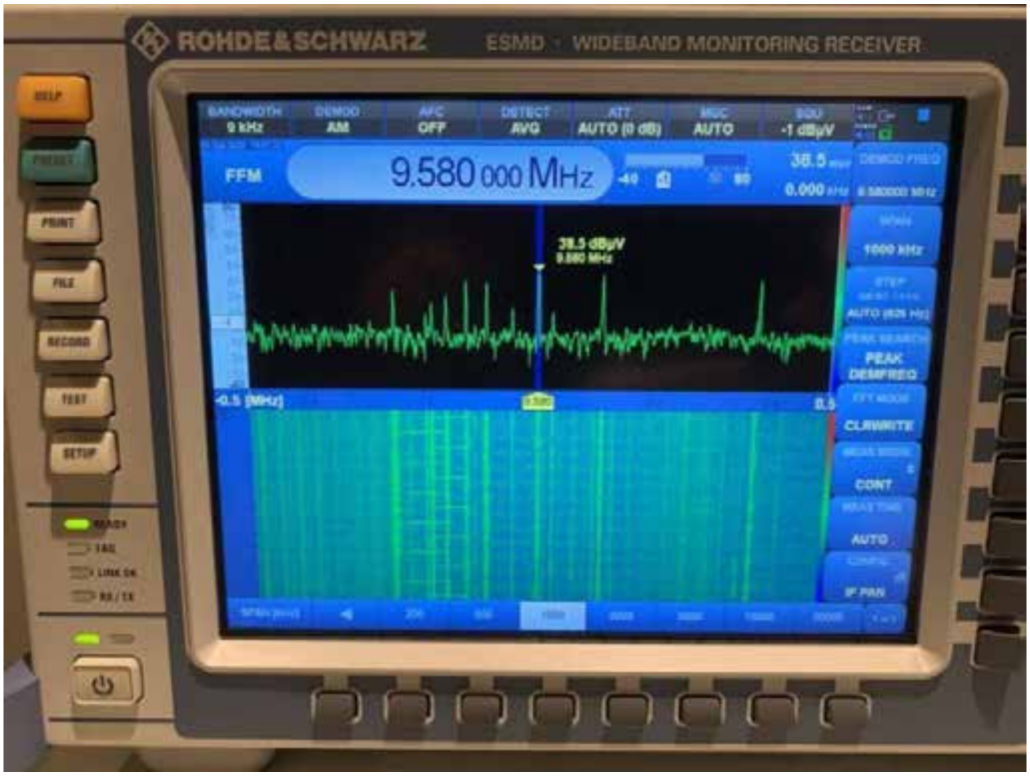
The R&S®ESMD features a wide frequency range (8 kHz to 40 GHz), outstanding receive characteristics, 80 MHz real-time bandwidth (base unit: 20 MHz) and a wealth of functions. Thanks to its sophisticated reselection stages, the receiver can be directly connected to a wideband monitoring antenna. This is an operating scenario that requires high large-signal immunity and high sensitivity, particularly in the presence of many strong signals.

The marine whip antenna showed 71.7 dBμV ~ 4 mV in to 50 Ω.

Important: This is a non-resonant but electrically matched antenna so the familiar current distribution does not apply!

The 1 m rod (Definition μV/m) as a reference active antenna to the ESMD test receiver indicates a treading of 38.5 dBμV.

The on screen dynamic range of the receiver covers -40 dBμV to 80 dBμV. This range can be shifted, as an example from -20 dBμV to 100 dBμV (100 mV).



Usinng the 1m tall active antenna there are quit a number of strong signals. Using the radio amateur definition of S9 =50uV=14dBuV this signal with 38 dBuV is very strong for such s short antenna.

Important information: The free space impedance is 377 Ω and the termination (receiver input) is 50 Ω so the down transformation is (377 +50)/50= 427/50= 8.54 or in dB 20log(8.54) =18.6 dB. The active antenna impedance transformation compensates this so the electric field is roughly 40 dBµV/m=100 µV/m

In comparison the 8 m non resonant antenna supplied 4 mV/50 Ω. The unmatched 1 m whip antenna would reduce EMF by practically 18.6 dB or reduce the voltage to 400 µV/50 Ω.

Sanity check: The efficiency of an electrically short antenna follows the formula

$$R_S = 80\pi^2 \left(\frac{\Delta l}{\lambda}\right)^2 \Omega$$

So the ratio is ~ 800 x $(1/30)^2$ / 800 x $(8/30)^2$ = $(1/30)^2/(8/30)^2$= 1/64 = 10log (64) or -18 dB; *QED*!

The Japanese company Diamond has recently introduced an improved set of mobile antennas (HF##CL), see photograph below.

They are mechanically very stable and seem to have a high Q.

Relative to an SWR of 2.6 dB, 3dB bandwidth the loaded Q is 6.975 MHz/110 kHz = 63.

This can be derived from the measurement seen below.

This measurement is actually tricky as the R&S network analyzer model FSH4 needs to have a proper counterpoise or set of radials. A somewhat simpler test is to check if the resonant frequency varies if the analyzer is touched. In this case two resonant cables where used like a set of radials.

The measurements are actually quite accurate and at resonant frequency show the expected deviation in frequency out of the box and an SWR of about 1.7. This is to be expected for this type of antenna and as most radio transmitters reduce power above an SWR of 1.5 what needs to be correctly called a line flattener is needed.

The pictures below shows part of the setup. The measurement was taken at night were we have strong signals in the 40 m Band. We did not have a mobile antenna for 10 MHz

The peak signal was a French language station at 7.375 MHz peaking 48.5 dBµV while the same station at the 8 m marine whip antenna was 75 dBµV, see below

The whip antenna does not have the luxury of an active amplifier but an impedance transformation

Z_0= SQRT (L/C)= SQRT (40 µH/14 pF)=1700 Ω, R(Resonance) = 1700 x Q = 1700x63 = 106 kΩ. The antenna will be extremely sensitive to detuning by any close-by objects.

Again, for an electrically short element:

$$R_S = 80\pi^2 \left(\frac{\Delta l}{\lambda}\right)^2 \Omega$$

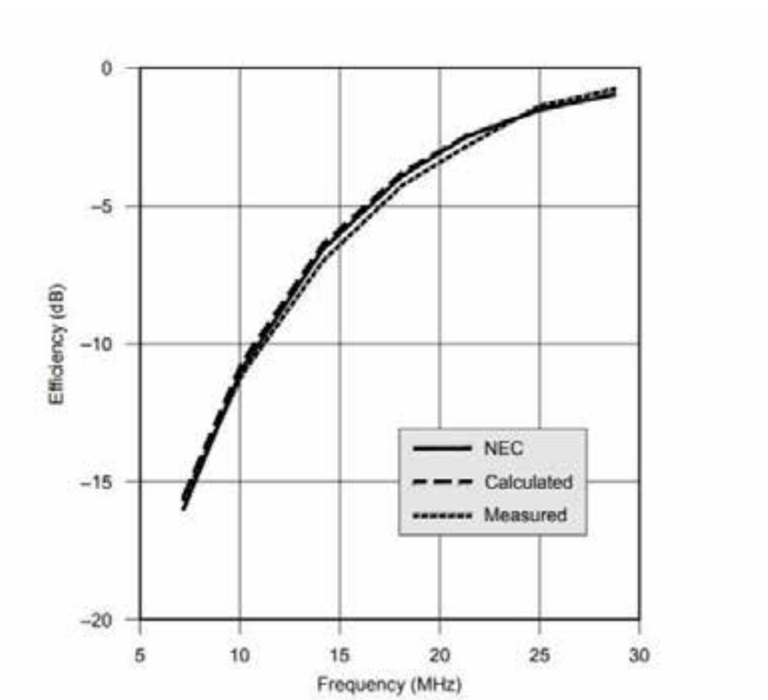
R_S ~ 800 $(2.5/40)^2$ = 4 Ω. For the SWR= 1.7 or 50x1.7 = 85,the loss is 85/4 = 21.2 or 26 dB. The measured difference was 25 dB relative to the 8 m whip . The error of 1.5 dB can be explained by fading during the measurement.

Then the Manpack, using the build in tuner was matched with a mobile antenna. This resulted in the same 1 dBμV noise level. A short wave receiver typically has a noise figure of 10 dB without a pre-amp or 10 dB S/N at 0.3 μV, or -10 dBμV. At this 1 dBμV level a 10 dBμV signal is needed. And yes, 28 MHz or even 50 MHz is much quieter,

One more important point; we all look at the absolute input voltage, I prefer dBμV at 2400 Hz IF bandwidth over dBm and unless an SDR with accurate S meter is used, we really don't know what is going on. Don't we love receivers that state S-0 (by correct definition should be 65 nV) at 2 μV already? And at Q 5?

Considering Small HF Loop Antenna Performance

The small HF loop, an AlexLOOP, pictured above has efficiency shown in the figure below.



Small HF Loop efficiency. [Source: K. Siwiak, KE4PT, and R. Quick, W4RQ, "Small Gap-Resonated HF Loop Antennas" *QST*, Sep., 2018.]

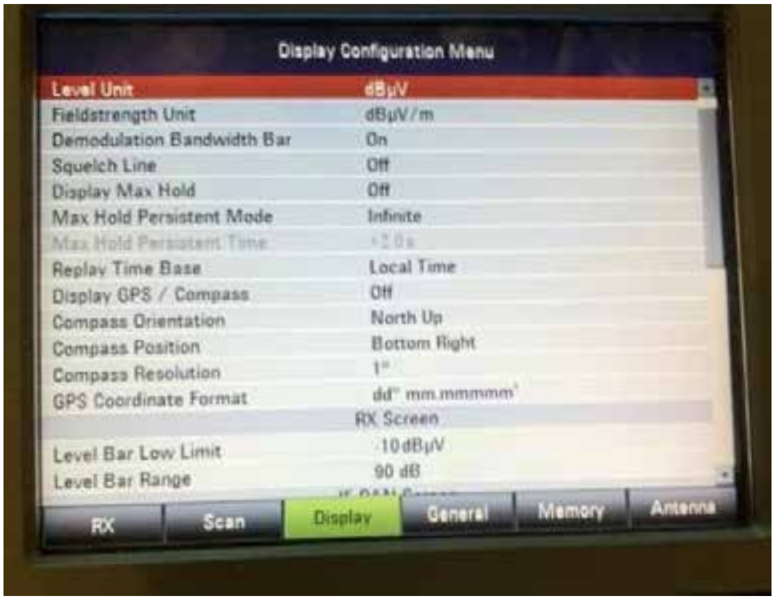
Kai Siwiak, KE4PT, shown using a small HF loop in a portable station at a public park.

The Mobile Station

As the mobile station used is the R&S®MR300xH/U Advanced Multiband Tactical Radios, as they cover 1.5 MHz to 512 MHz and operate from CW to FM in all needed modulations. In addition, being an SDR, the input level indicator was optimized to be in dBμV from internal noise (CW =-30 dBμV, SSB/2.4 kHz = -19d BμV, with +110 dBμV max) to an extremely loud signal, S-9 = 35 dBμV, or 84 dB above S-9.

The absolute accuracy is ±1 dB. Here is a view of the front panel.

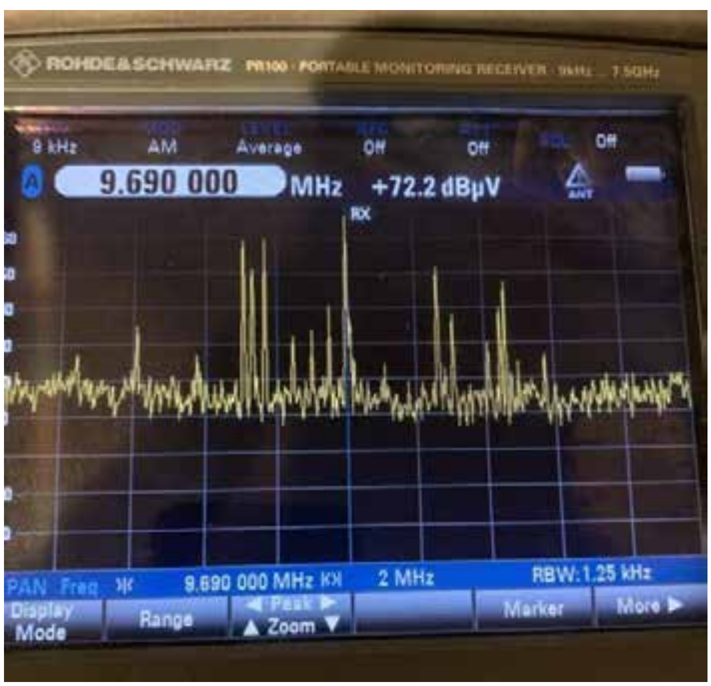
The yellow wire is a “counterpoise” of 5 m length. The displays shows 25 dBμV; C14 refers to the channel 14, which happens to be in the 20 m band , tunable in 1 Hz steps, and +J3E refers to upper side band .



Setting of the PR100 receiver

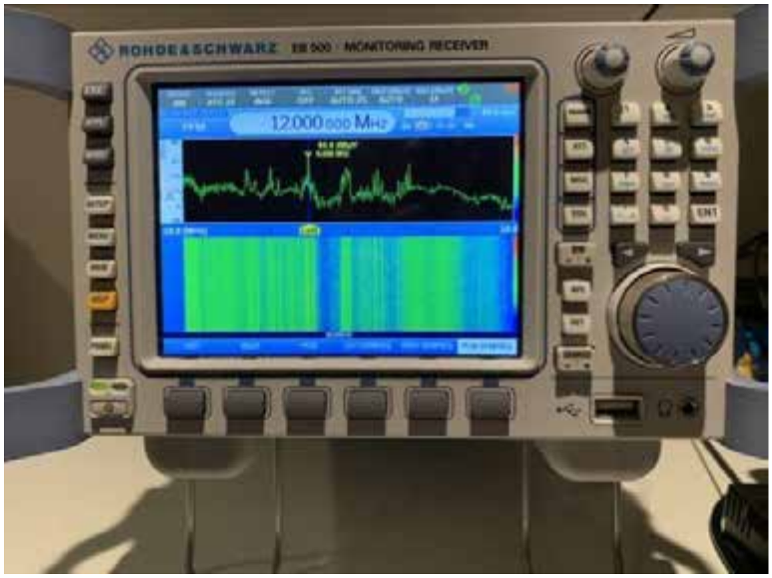
When the loop antenna is used, the wide band noise is multiplied with its narrow band width and we see the noise peak, resonant condition

Tuning the loop antenna to the 30 m broadcast range the wanted signal at 52 dBuV shows the peak while the high Q suppresses the adjacent signals.



This picture shows the same signals from the 23 feet whip antennna

Frequency occupancy in the evening at about 10 MHz with a tuned 23 feet rod antenna

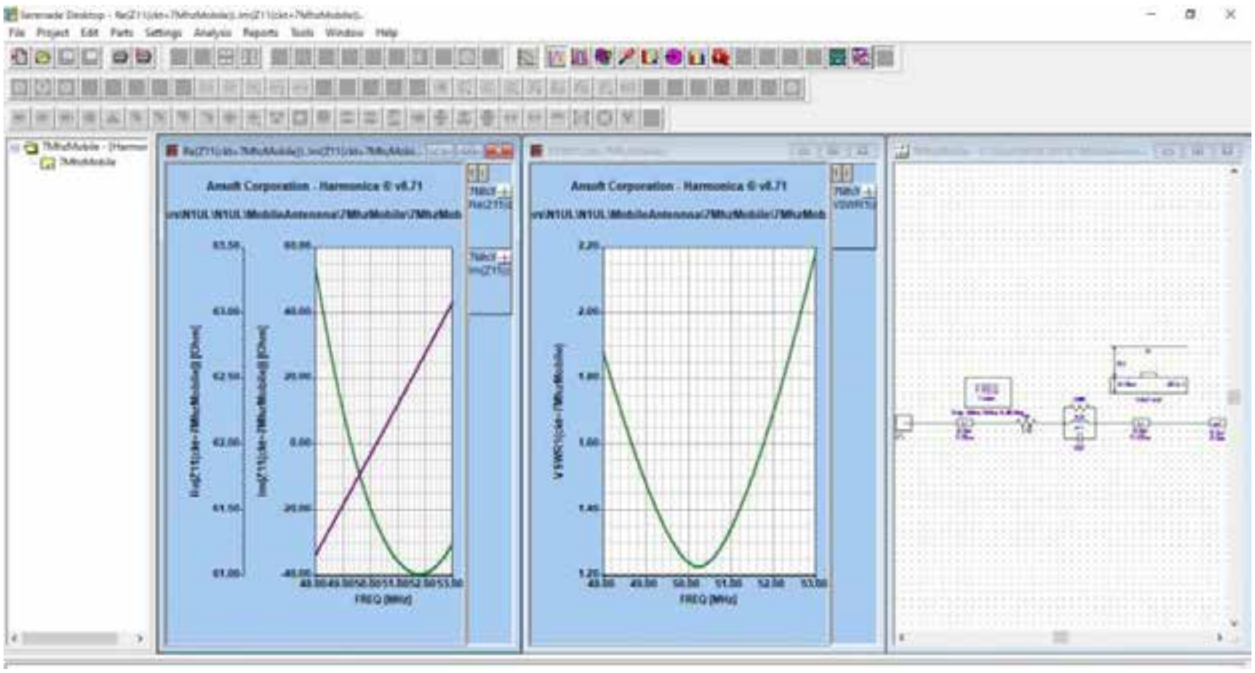


Rohde & Schwarz EB 500 high performance SDR receiver to monitor the frequency occupancy set from 2 MHz to 22 MHz using an active antenna, at 3:41 PM there were still huge signals present. The calibration accuracy is 0.1 dB

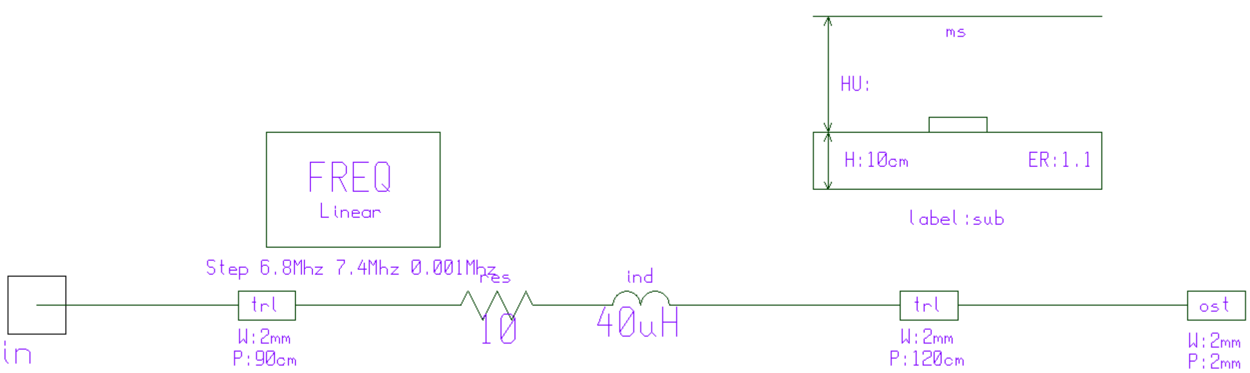
Whip Antenna Simulations

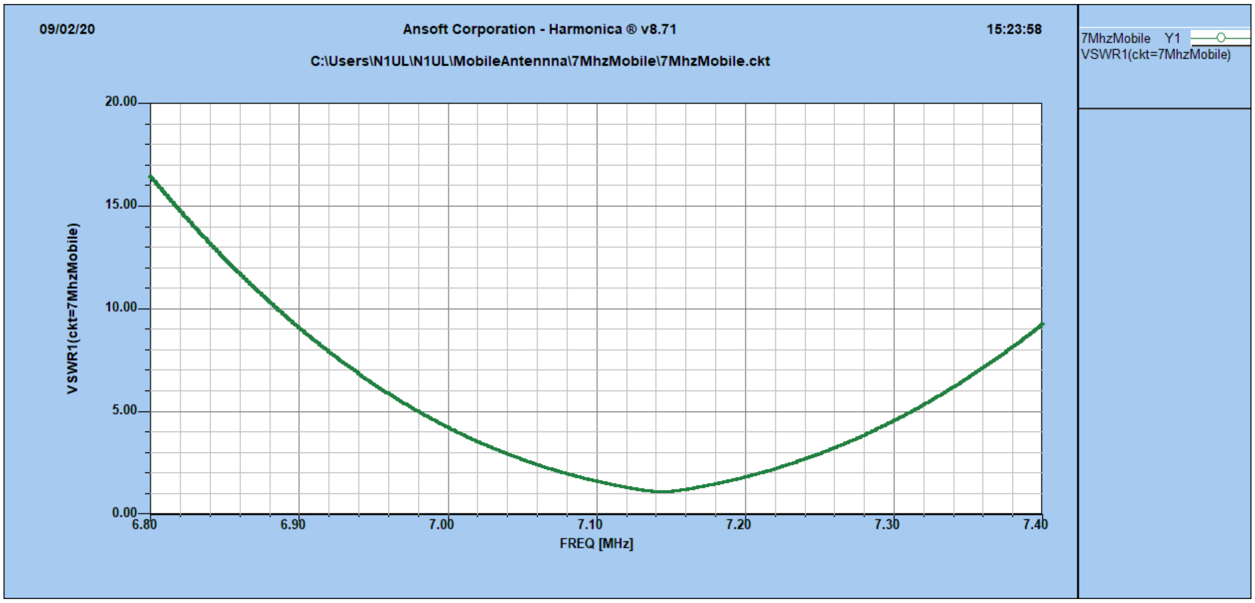
The majority of antennas are simulated with the NEC software but as we are less interested in radiation pattern but in matching and correct SWR and operating Q; a field simulator using tanh functions for resonant elements adds an additional level of accuracy to it. Multi element arrangement simulations are possible if the multiple coupled line element model is used. Up to 8 parallel elements with arbitrary spacing and diameter are possible. It is based on the Spectral Domain Method with enhanced accuracy and speed.

The microwave simulator, see below,

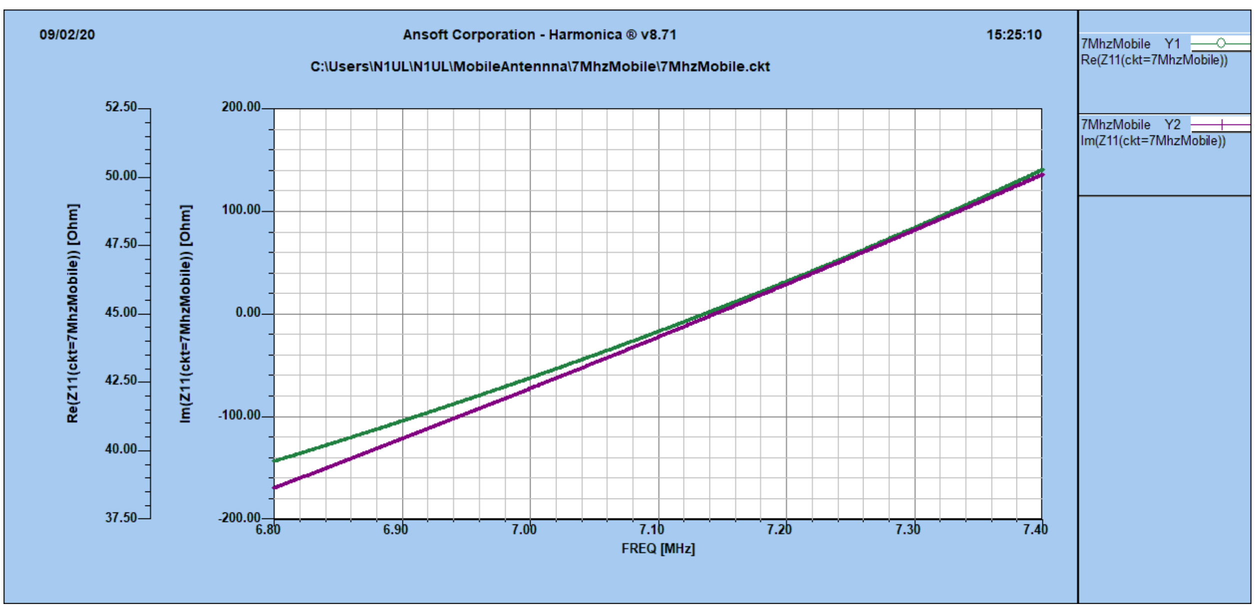


Using the Serenade Microwave /RF simulator to look at the resulting resonance as a function of the loading coil and the steel rod parts

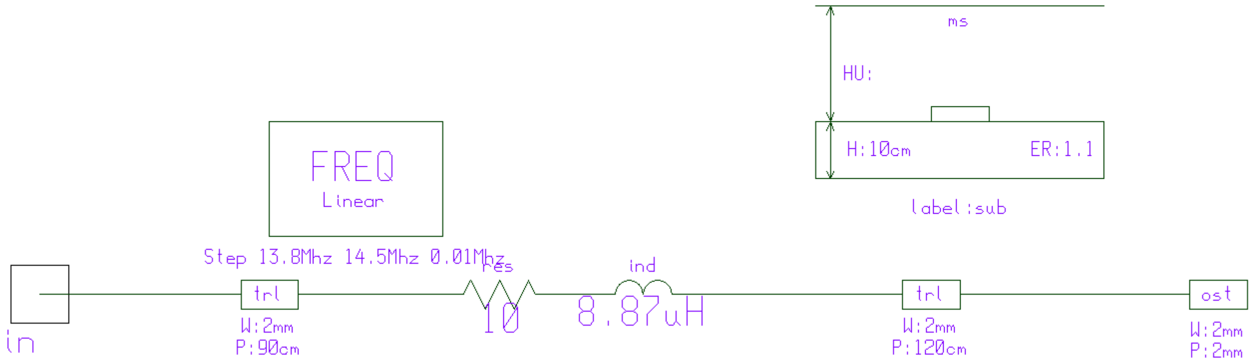




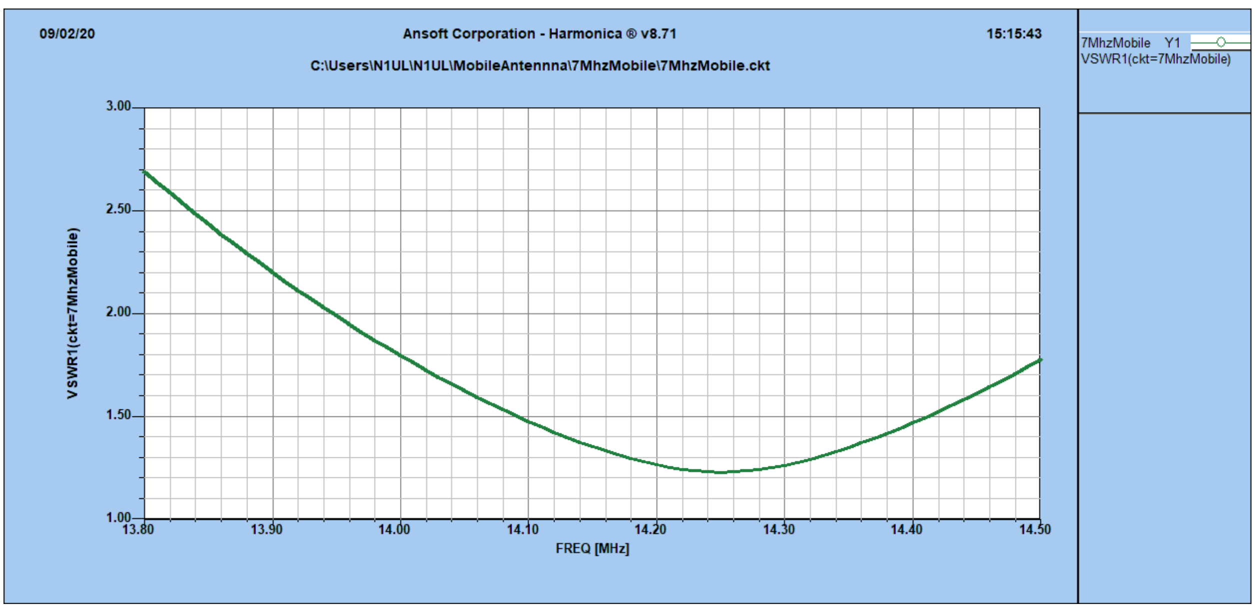
40 m mobile rod antenna SWR



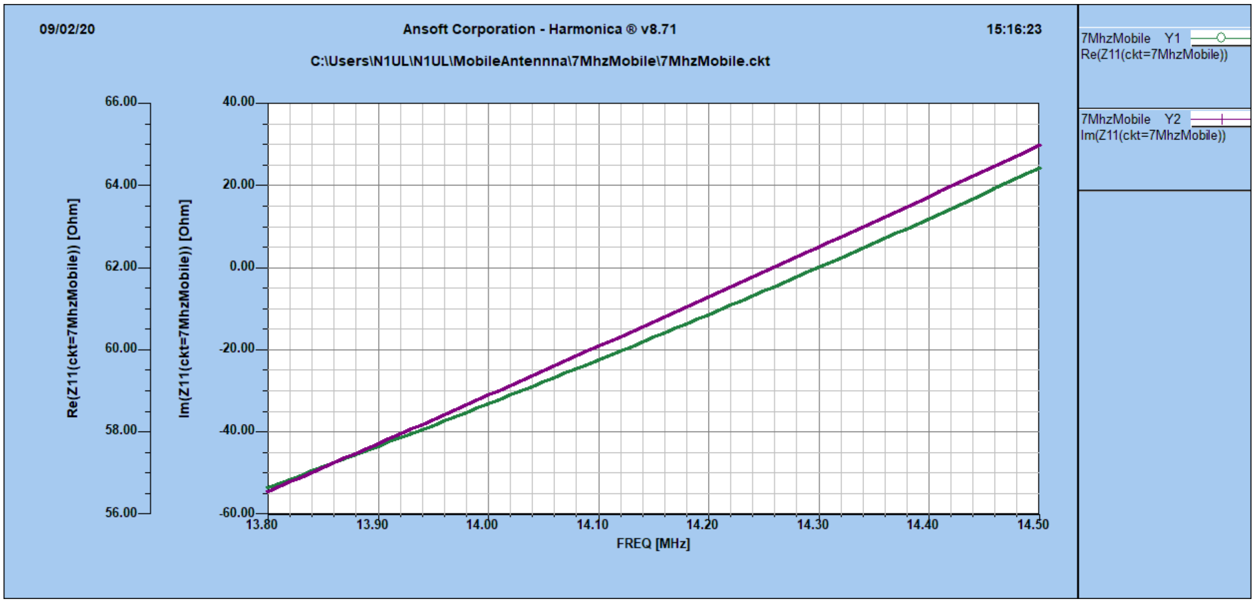
40 m mobile antenna input impedance



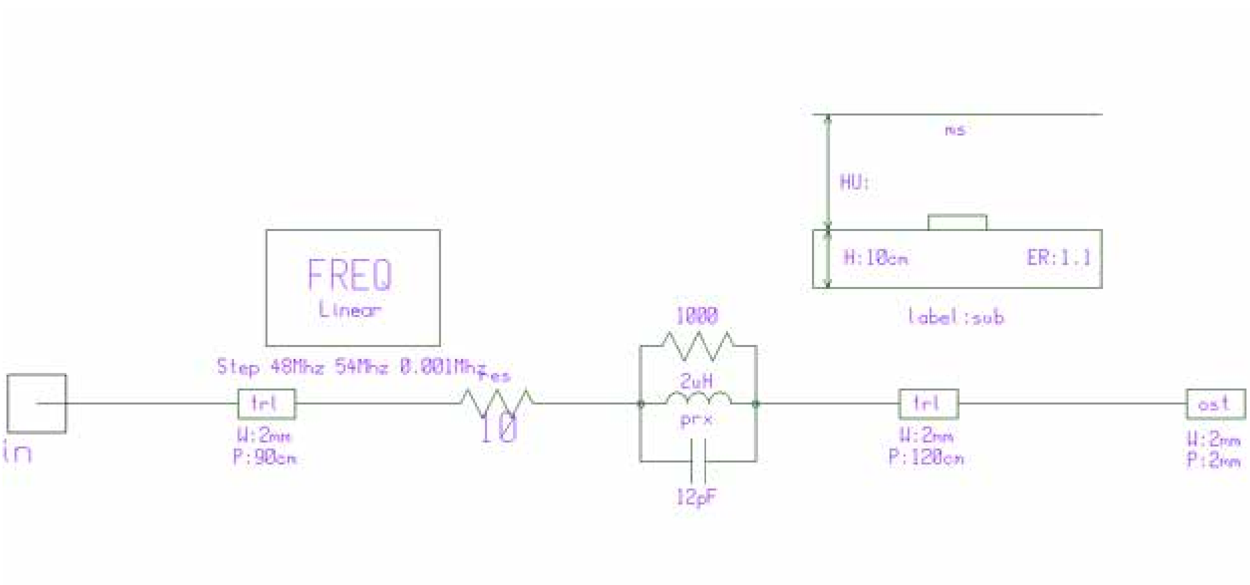
Equivalent circuit parameters of the 20m mobile antenna



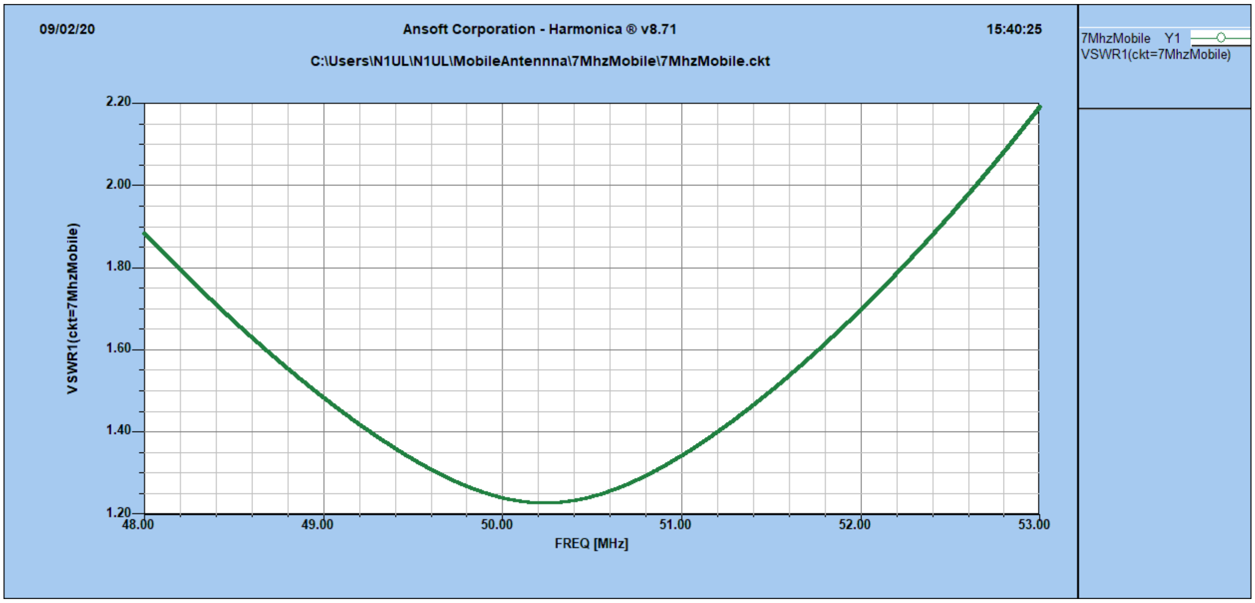
VSWR simulation of the 20 m mobile antenna



Input impedance of the 20 m mobile antenna



Equivalent modeling parameters for the 6 m mobile antenna



VSWR simulation of the 6 m antenna

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