ROHDE&SCHWARZ

ANTENNA BASICS

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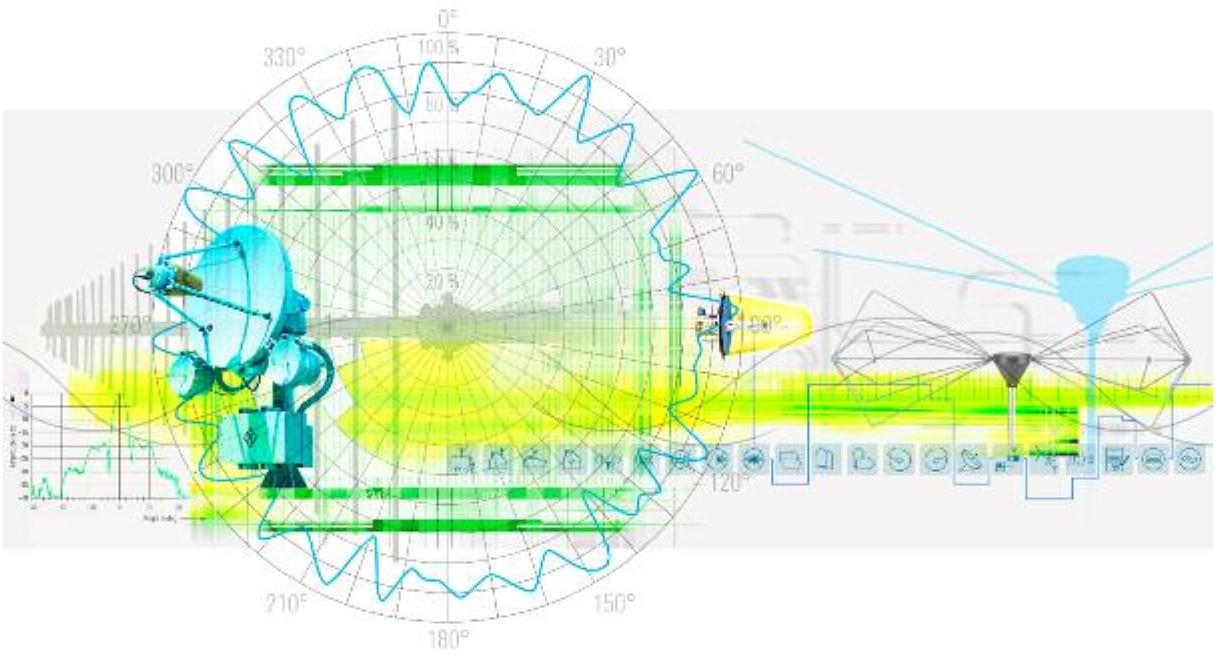


C O N T E N T S

1 Introduction

Antennas act as **transformers** between conducted waves and electromagnetic waves propagating freely in space (Fig. 1.1). Their name is borrowed from zoology, in which the Latin word *antennae* is used to describe the long, thin feelers possessed by many insects.

The oldest existing antennas, such as those used by *Heinrich Hertz* in 1888 during his first experiments to prove the existence of electromagnetic waves, were in theory and in practice not so very different from an RF generator; in fact, resonant circuits are still frequently used even today as a means of explaining the individual properties of antennas. It was not until around 1900 or even later, when transmitting and receiving stations were being built, that a clear distinction was made and antennas were classified as separate components of radio systems.

At first glance, modern antennas look very much the same as their "ancient forebears"; however, as a rule they are optimized at great expense for their intended application. Communications technology strives first and foremost **to transform one wave type into another with as little loss as possible**. This requirement is less important in the case of test antennas, which are intended to provide a **precise measurement of the field strength at the installation site** to a downstream test receiver; instead, their physical properties need to be known with high accuracy.

The following chapters can describe only a few of the many forms of antenna that are in use today. The explanation of the physical parameters by which the behavior of each antenna can be both described and evaluated is probably of wider general use.

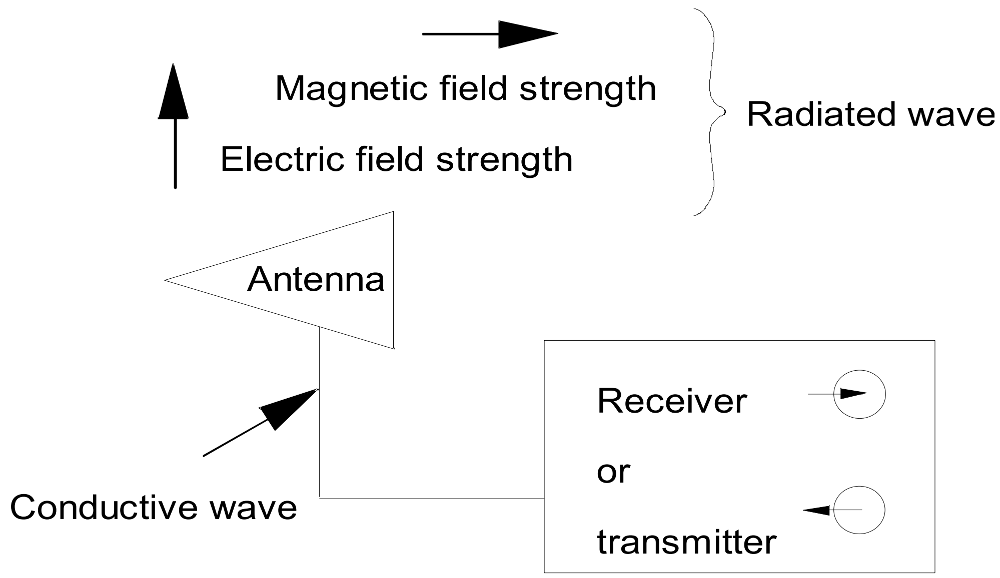


Fig. 1.1 The antenna as an electromagnetic transformer

The starting point for any theoretical investigation of wave propagation is the system of equations presented by **James Clerk Maxwell** in **1864**, which exactly describes how all field parameters behave in time and space:

James Clerk Maxwell 1884 :

$$\text{rot}\,\mathcal{H} = \nabla \times \mathcal{H} = \mathcal{G} + d\mathcal{D}/dt$$

$$\text{rot}\,\mathcal{E} = \nabla \times \mathcal{E} = -\,d\mathcal{B}/dt$$

$$\text{div}\,\mathcal{B} = \nabla \bullet \mathcal{B} = 0$$

$$\text{div}\,\mathcal{D} = \nabla \bullet \mathcal{D} = \rho$$

From these and the material equations

$$\mathcal{D} = \varepsilon\,\mathcal{E} \qquad \mathcal{G} = \kappa\,\mathcal{E} \qquad \mathcal{B} = \mu\,\mathcal{H}$$

linking the field parameters it is possible to derive the second order differential equation known for historical reasons as the **telegraph equation**,

$$\nabla \times \nabla \times \mathcal{F} + \kappa\,\mu\,d\mathcal{F}/dt + \mu\,\varepsilon\,d^2\mathcal{F}/dt^2 = 0 \quad \text{or}$$

$$\text{rot rot}\,\mathcal{F} + \kappa\,\mu\,d\mathcal{F}/dt + \mu\,\varepsilon\,d^2\mathcal{F}/dt^2 = 0$$

in which the symbolic letter $\mathcal{F}$ stands for either the electric or the magnetic field strength. Provided that conductivity is vanishingly small ($\kappa = 0$) and parameters are purely sinusoidal at a constant frequency, the **wave equation**

$$\nabla \times \nabla \times \mathcal{F} - \omega^2\,\mu\,\varepsilon\,\mathcal{F} = 0 \quad \text{or}$$

$$\text{rot rot}\,\mathcal{F} - \omega^2\,\mu\,\varepsilon\,\mathcal{F} = 0$$

can be derived from the equation above.

The simplest solution to this differential equation is known as a **plane wave in loss-free, homogeneous space**, for which the following relations are valid:

- The vectors of the electric and magnetic field strengths are perpendicular, both mutually and to the direction of propagation (direction of energy transport, see Fig. 2.1)

- Electric field strength E and magnetic field strength H are mutually connected via

  the free space **field characteristic impedance** $Z_0 = 120\,\pi\,\Omega = 377\,\Omega$

as follows: $E = Z_0 \cdot H$

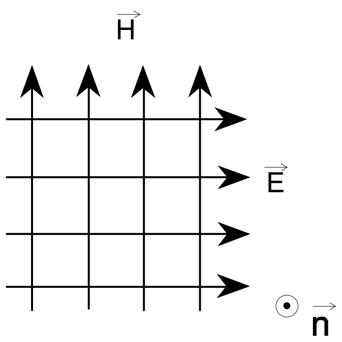


Fig. 2.1 Plane wave

Therefore in the case of plane waves in homogeneous space (and only in that case!), only one parameter needs to be measured in order to determine both field strengths. However, plane waves quite simply cannot exist: In order to possess the stated properties, they would have to originate from a source (antenna) located at an infinite distance. It can be shown, however, that when the distance from the transmitting antenna is sufficiently large (usually one or a few wavelengths with the exception of electrically large antennas) each wave can be considered a plane wave to a sufficient level of accuracy. The spatial domain in which this case holds true is known as the **far field**. Whereas useful field-strength measurements are as a rule taken in the far field of the antenna, this cannot always be guaranteed in the case of RFI field-strength measurements. When investigating radiocommunications links, however, it is almost always possible to assume far field conditions.

The direction of the electric field strength vector is called the **direction of polarization** or simply the **polarization direction**. A distinction must be made between the following types of polarization:

- linear: The magnitude of the electric field strength vector changes periodically .

- circular: The electric field strength vector is constant with respect to the magnitude and rotates about the vector that points in the direction of propagation.

- elliptical: The electric field strength vector changes in magnitude and direction; its peak position can be described by an elliptic equation.

Linear-polarized waves near the earth's surface can also be referred to by their surface area (vertical, horizontal polarization).

There is **polarization mismatch** when the polarization of the electromagnetic wave does not match the polarization of the receiving antenna. The attenuation during linear-polarized reception of circular-polarized waves (and vice versa) amounts to 3 dB, whereas the attenuation between circular polarizations rotating in opposite directions and between orthogonally linear polarizations theoretically increases beyond all limits.

The **simplest imaginable antenna** is the **isotropic radiator**; this too does not exist in practice, but makes an excellent theoretical model. An isotropic radiator, which is a dimensionless point in space, generates waves with spherical wavefronts that are radiated uniformly in all directions. When the ideally matched transmitter power P_S is applied to it, then at distance r this gives rise to the

radiation density

$$S = \frac{P_S}{4 \pi r^2}$$

(often also known as power density) (see Fig. 2.2). Since the power density can also be determined in the far field of an antenna from the product of the field strengths in accordance with the equation

$$S = E \cdot H$$

the **free-space far field strength of the isotropic radiator** is

$$E_0 = \frac{\sqrt{30\ \Omega\ P_S}}{r}$$

This in turns gives rise to the following dimensional equation which is highly usable in practice

$$E_0 = 173 \cdot \frac{\sqrt{P_s}}{r}$$

where E_0 is expressed in mV/m when the power is in kW and the distance is in km.

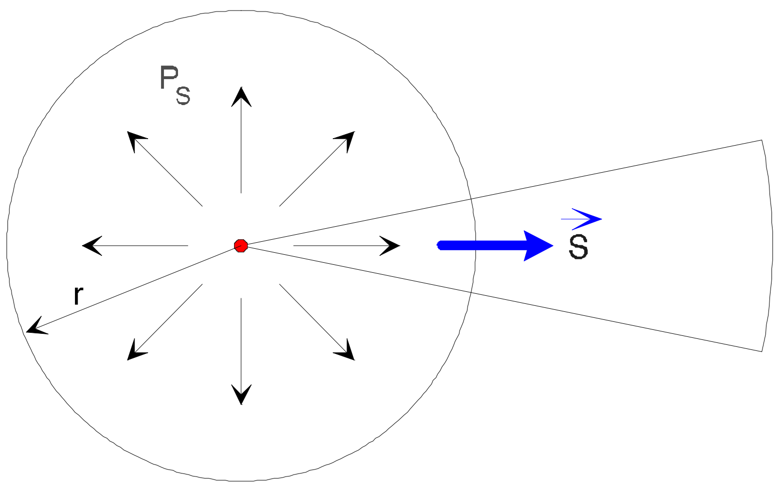


Fig. 2.2 The isotropic radiator in homogeneous space

However, the assumption of straight-line, free-space propagation that is totally independent of frequency is a crude simplification. It turns out to be more accurate but also far more complicated to calculate the receive field strength by also including the effects of the physical phenomena **diffraction, refraction and reflection**, to which all electromagnetic waves are subject.

Long and ultralong waves for example do not propagate in a precisely straight line, since they are diffracted by the earth's surface. Short waves are commonly reflected by the ionosphere and still shorter waves by obstacles. Fig. 2.3 shows an overview of the mechanisms affecting wave propagation at different frequencies. For field strength forecasts on the basis of a wide variety of parameters, relevant tables, atlases and computer programs are being used to an ever increasing extent, but to present and discuss them here would go beyond the scope and objectives of this paper.

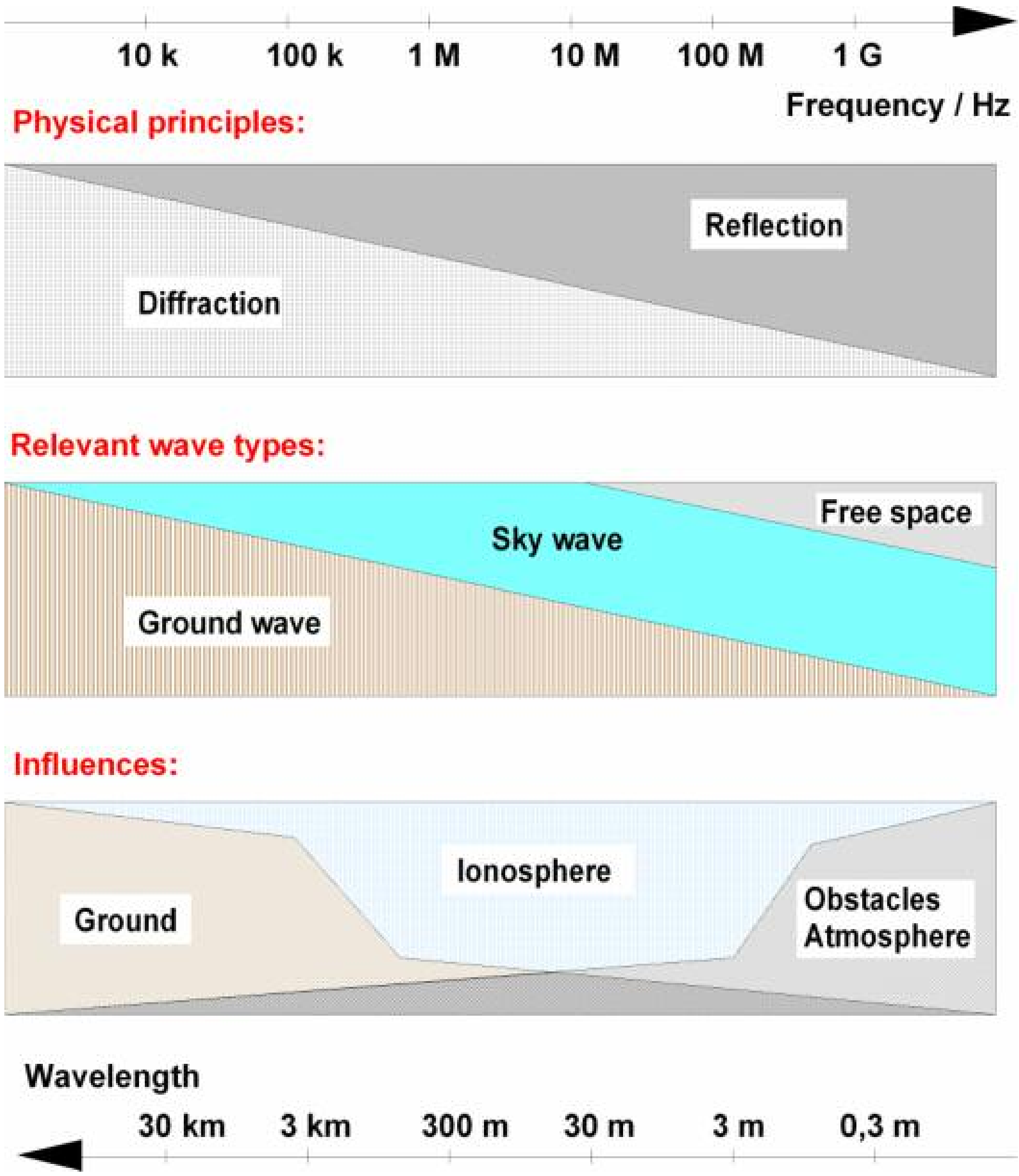


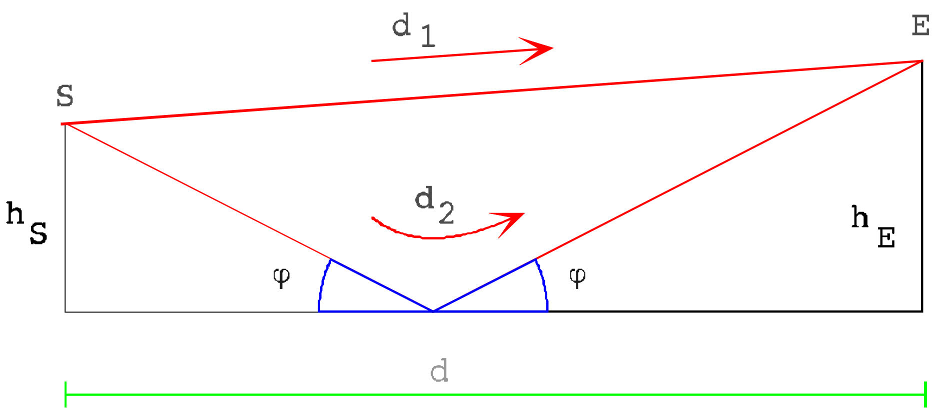
Fig. 2.3 Wave propagation at different frequencies

The analytical calculation of a propagation path in which antennas are transmitting and receiving over a highly conducting ground still has practical significance even today, since on the one hand the results that can be achieved have a high level of accuracy, and on the other, such configurations are frequently used in EMI and antenna measurements.

In accordance with Fig. 2.4 the receive field strength E can be computed from the vector superposition of the wave that is directly incident at the reception site and the wave that is reflected at the surface. However, for scalar (and therefore simpler) addition the directly incident and reflected field strength vectors must be parallel, which occurs only in the case of horizontal polarization. For this reason (and given ideal conductivity) the reflection coefficient is R = - 1, which further simplifies the equation shown in Fig. 2.4. If however parallel vectors and ideal ground conductivity cannot be assumed, vector addition must be performed, while the reflection coefficient is a complex function of path geometry, frequency and ground characteristics.

The argument of the exponential function in Fig. 2.4 is further simplified if the distance between the two antennas is very much larger than the height of the antennas (with the incident waves only just touching). The influence of the earth's curvature need only be included in the calculation when the distance between the transmitting and receiving antennas is extremely large; however, in the configurations that occur in practice the divergence coefficient can usually be set to α = 1.

It should also be mentioned that in configurations which are used for antenna calibration (reflection test site) or interference field measurements, it can only rarely be assumed that the angle of incidence and emergence φ shown in Fig. 2.4 is small; on the contrary, a very short test distance requires a more complex calculation.



$$E = E_0 \left| 1 + \alpha R e^{-j k (d_2 - d_1)} \right|$$

$$\alpha = 1$$

für ebene Erde
for plain ground

$$k (d_2 - d_1) \approx 4 \pi \frac{h_S h_E}{d \lambda}$$

$\varphi \longrightarrow 0$

Fig. 2.4 Wave propagation over an ideally conducting surface

3 Antenna Parameters

As already mentioned in the Introduction, the task of antennas is to transform one wave type into another. The direction in which this energy conversion takes place is of no importance for the operational principle or for ease of understanding. Therefore transmitting and receiving antennas do not require different investigation methods (reciprocity principle), and the physical parameters described below are valid both for transmission and reception. This is true even if parameters can be measured only during transmission or only during reception, or when their specification seems to apply to only one of the two cases. Active antennas are the only exception to this: as purely receiving antennas they are non-reciprocal. A clear distinction must be made between transmitting and receiving antennas if, for example, the maximum permissible transmitter power is to form part of the investigation. However, this does not affect the operating principle and the parameters.

3.1 Radiation Pattern

The **spatial radiation behavior** of antennas is described by their **radiation pattern** (normally in the far field). Only an isotropic radiator would exhibit the same radiation in every spatial direction, but this radiator cannot be implemented for any specified polarization and is therefore mainly suitable as a model and comparison standard. Dipoles and monopoles possess directivity: An electrically short dipole in free space has the spatial radiation pattern shown in Fig. 3.1 with nulls in the direction of the antenna's axis (also known as the antenna boresight).

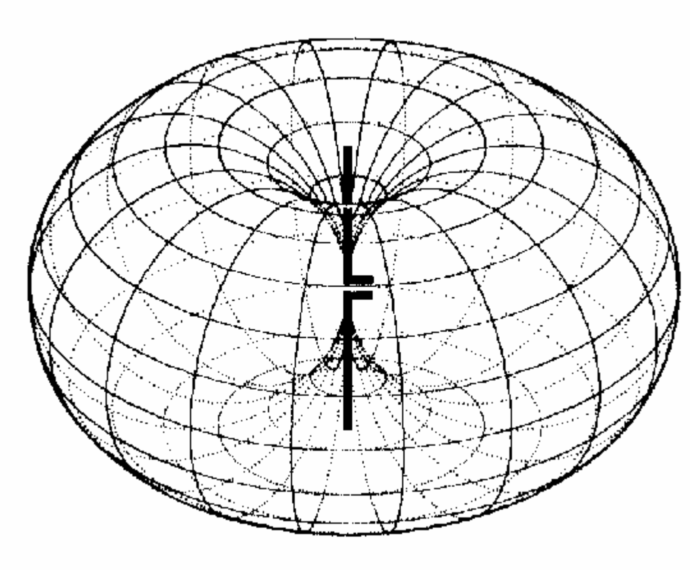


Fig. 3.1 Spatial radiation pattern of a dipole

The following are part of a complete and rigorous description of the spatial radiation distribution

1. the ϑ component of the field strength as a function of the angles ϑ and φ
2. the φ component of the field strength as a function of the angles ϑ and φ

with regard to magnitude and phase; the angles ϑ and φ describe the elevation and azimuth.

It is often sufficient to specify instead the **field strength magnitude** in the desired direction of polarization (copolarization direction), since this magnitude is usually standardized to the field strength maximum:

$$C(\vartheta,\varphi) = \frac{E(\vartheta,\varphi)}{E_{max}(\vartheta_0,\varphi_0)} = \frac{H(\vartheta,\varphi)}{H_{max}(\vartheta_0,\varphi_0)}$$

For the sake of greater simplicity, in many cases it is possible to limit oneself to two sections through the spatial radiation pattern:

- The **azimuth radiation pattern** describes the dependency of the field strength on the azimuth angle φ for the elevation angle ϑ at which the maximum of the spatial radiation distribution lies. A special case of the azimuth radiation pattern is the **horizontal radiation pattern** for ϑ = 90°.

- The **vertical radiation pattern** describes the dependency of the field strength on the elevation angle ϑ at the azimuth of the spatial maximum and of the corresponding opposite direction.

  Fig. 3.2 shows the vertical radiation pattern of a quarter-wave vertical antenna. This is important in cases such as shortwave transmission.

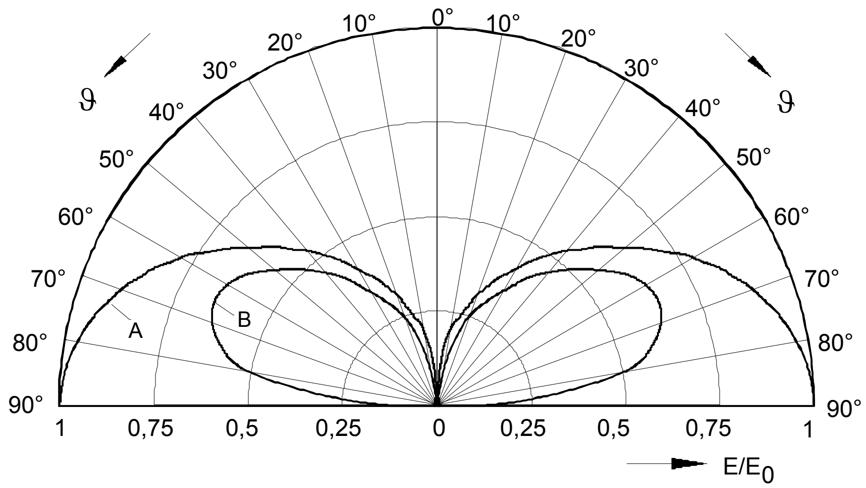


Fig. 3.2 Radiation pattern of a vertical $\lambda/4$ monopole over ideally (A) and moderately (B) conducting ground

Whereas the antennas that can be implemented in practice by no means achieve uniform radiation on all sides, the expression **directional antenna** is applied only to antennas with a **sifnificant unilateral preference** for one spatial direction.

When assessing radiation patterns it should be observed whether the solid angle represents the **field strength** distribution (as is most usually the case nowadays) or the **power** distribution. Power patterns are distinguished by narrower beam widths and smaller side lobes. Whereas a display in spherical or polar coordinates such as that in Fig. 3.2 is commonly chosen for its clarity, a Cartesian (and probably logarithmic) display usually provides a significantly more detailed and precise picture of the radiation characteristics.

In view of the reciprocity principle already mentioned, it will be sufficient here and in the following sections to investigate the transmission case only.

3.2 Directivity Factor

The **directivity factor D** is defined as the ratio between the radiation intensity F_{max} generated in the main direction of radiation and the radiation intensity F_i which a lossless isotropic radiator generates at the same **radiated power P_t** (ITG/NTG 2.1/01). Instead of the radiation intensity, the power density which is given by the Poynting vector is as follows

$$\underline{\mathbf{S}} = \underline{\mathbf{E}} \times \underline{\mathbf{H}} \qquad \text{where } \underline{\mathbf{S}} \text{ perpendicular to } \underline{\mathbf{E}} \text{ perpendicular to } \underline{\mathbf{H}} \text{ in the far field,}$$

can also be specified at the same distance r from the antennas (bold and underlined letters are used to identify vectors).

The following therefore applies:

$$D = \frac{F_{max}}{F_i} \quad \text{where} \quad F_i = \frac{P_t}{4\pi}$$

3.3 Gain

Gain G is the ratio between the radiation intensity F_{max} generated in the main direction of radiation and the radiation intensity F_{i0} which a lossless isotropic radiator generates at the same **input power P_{t0}** (ITG/NTG 2.1/01):

$$G = \frac{F_{max}}{F_{i0}} \quad \text{where} \quad F_{i0} = \frac{P_{t0}}{4\pi}$$

In contrast to the directivity factor, therefore, if **antenna efficiency η** is also included, the applicable relationship is:

$$G = \eta \cdot D$$

Furthermore the **practical gain G_{pract}** is defined in DIN 45030. This parameter can be measured, whereas the gain defined above assumes ideal matching and must be determined from the practical gain and from the magnitude of the reflection coefficient r by evaluating the equation

$$G = G_{pract} \frac{1}{1 - |r|^2}$$

Both gain and directivity factor are commonly specified in logarithmic form

$$g = 10 \lg G \text{ dB} \qquad \text{and} \qquad d = 10 \lg D \text{ dB}$$

In special applications, gain is still occasionally specified without reference to the isotropic radiator or with reference to a direction other than the main direction of radiation. In the case of divergence from the applicable recommendations and standards, the chosen reference quantity must always be mentioned.

3.4 Effective Aperture

The **effective aperture A_W** is a parameter that is defined especially for receiving antennas. It is a measure of the maximum received power P_r which the antenna can obtain from a plane wave of power density S:

$$P_{rmax} = S \cdot A_w$$

Although the effective aperture can definitely be thought of as a real aperture that is perpendicular to the direction of propagation of the incident wave, it is not necessarily identical to the geometric aperture A_g of the antenna. The relationship between the two apertures is described by the **aperture efficiency**

$$q = A_w / A_g$$

The effective aperture and the gain can be converted from one to the other with the aid of the equation:

$$A_w = \frac{\lambda^2}{4\pi} \cdot G$$

3.5 Effective Antenna Length

By analogy with the effective aperture, the effective antenna length (often also called the effective antenna height) is the quotient of the maximum **open-circuit voltage U_0** on the antenna terminals and the electric field strength E of the linear-polarized incident wave at optimum antenna orientation (Fig. 3.3):

$$U_0 = h_{eff} \cdot E_0 = l_{eff} \cdot E_0$$

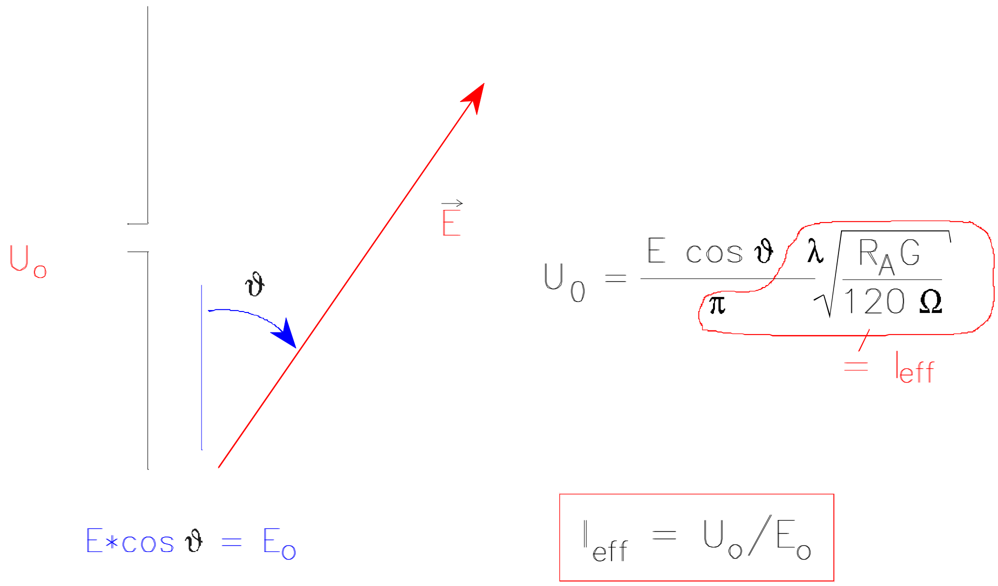


Fig. 3.3 Effective antenna length

The effective length is not identical to the geometric antenna length. In the transmission case, it can be computed from the geometric length l of the antenna and the current distribution $I(z_q)$ on the antenna by evaluating the integral

$$l_{eff} = \int_0^l \frac{I(z_q)}{I_A} \, dz_q$$

which in the example of a very "slim" half-wave dipole amounts to $l_{eff} = 0.64 \cdot l$. This can also be used to clearly explain the concept of the effective length: The effective length of an antenna is the length which a dipole homogeneously carrying the feedpoint current I_A would have to have in order to generate the same field strength in the main direction of radiation (see Fig. 3.4).

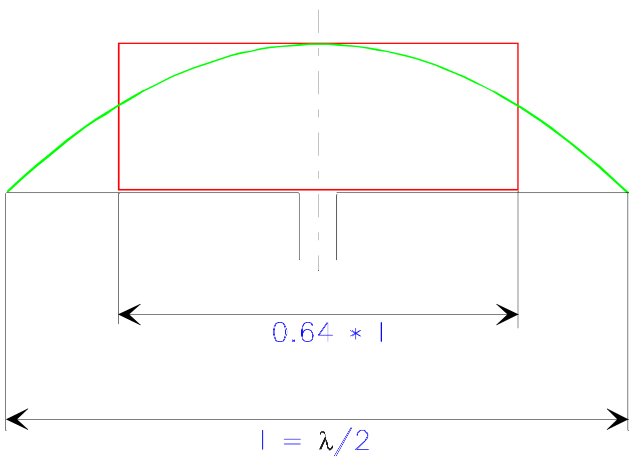


Fig. 3.4 Effective length of a half-wave dipole

In order to convert power-referenced parameters into voltage-referenced parameters the equation

$$l_{eff} = 2\sqrt{\frac{R_A A_w}{120\,\pi\,\Omega}}$$

can be used, provided the feedpoint resistance R_A of the antenna is known.

3.6 Antenna Factor

Unlike the effective length, the antenna factor K links the electric field strength E with the **voltage U_E** present on the receiver input during matching (Fig. 3.5):

$$K = \frac{\textit{Electrical Field Strength}}{\textit{Output Voltage at } 50\,\Omega}$$

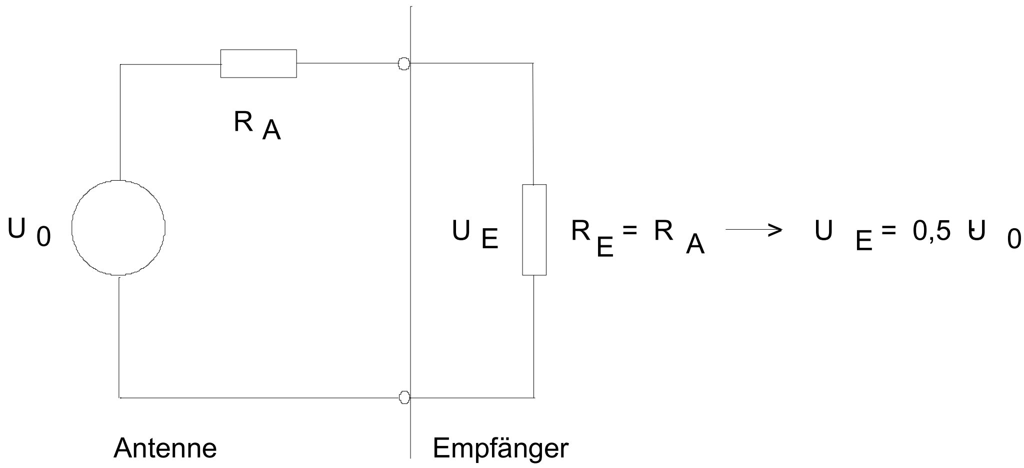


Fig. 3.5 Antenna factor

As a rule, a modern test receiver has a 50 Ω input. If $R_E = R_A = 50\,\Omega$, then $U_E = 1/2\ U_0$ and

$$K = 2 / l_{eff}$$

In order to measure a field strength it is **indispensable to know precisely either the effective antenna length or the antenna factor of the test antenna being used**.

It is often more convenient to use the antenna factor in the logarithmic form:

$$k = 20 \lg (K \cdot m)$$

In this form, the antenna factor k then only has to be added to the voltage level P_u displayed by the test receiver in order to obtain at once the **level F of the electric field strength**:

$$\frac{F}{dB(\mu V/m)} = \frac{P_u}{dB(\mu V)} + \frac{k}{dB(1/m)}$$

For the sake of completeness it should be mentioned that to obtain a precision field strength measurement the **cable loss** between the test antenna and the receiver has to be included.

Whereas throughout field strength measurement the antenna factor is commonly used as the characterizing value for the antenna, the predominant terms in general antenna engineering are gain and directivity factor. Therefore, it often proves useful to know the relationship

$$K = \frac{9{,}73}{\lambda \cdot \sqrt{G}}$$

between the antenna factor and the practical gain, which is also expressed in logarithmic form as

$$k = (- 29.8 + 20 \lg (f/MHz) - g) \ dB$$

3.7 Impedances and Resistances

The term **characteristic impedance** which is still occasionally used in connection with antennas originally came from the theory of high frequency lines. These lines can be represented both as active resistors and inductors distributed in series on the conductors and as capacitances and conductances arranged in between, which facilitates calculations. These line reactances, which can be location-dependent, are designated by a superscript vertical dash indicating the dimension "physical parameter divided by unit of length". On this basis the **characteristic impedance**

$$Z_W = \sqrt{\frac{R' + j\,\omega\,L'}{G' + j\,\omega\,C'}}$$

can be calculated. Probably because some of the calculation methods commonly used in antenna engineering are based on the characteristic impedance defined in this way, it is sometimes confused with the following impedances and resistances.

The most significant parameter for antenna operators is the complex **input impedance**

$$Z_E = R_E + j\,X_E \,,$$

being the instantaneous impedance at the antenna feedpoint (see Fig. 4.2).

Its real part is the sum of the **radiation resistance**

$$R_S = P_S / I^2$$

at the antenna feedpoint and the **loss resistance** R_L. It should be noted, however, that the radiation resistance, being the quotient of the radiated power and the square of the rms value of the antenna current, is dependent on position - as is the antenna current itself - so that in order to specify a radiation resistance it is always necessary to include a reference to the point on the antenna at which it is valid. The usual reference points are the antenna feedpoint or the current maximum. Although these two positions do in fact coincide in some antenna types, this is by no means so in all cases.

The imaginary part X_E of the input impedance disappears if the antenna is operated at **resonance**. Electrically very short linear antennas such as dipoles and vertical antennas have capacitive impedance values ($X_E < 0$) of several kΩ, whereas electrically overlong linear antennas can be recognized by their inductive imaginary part ($X_E > 0$).

The term **nominal impedance** Z_n is defined in a standard (ITG/NTG 2.1/01) and represents simply a reference value. Usually the characteristic impedance of the antenna cable is specified and the antenna impedance then has to be matched to it (as a rule Z_n = 50 Ω).

A measure of the quality of the matching is the voltage standing wave ratio s (VSWR) as the quotient of the maximum and minimum voltage on the feed line.

$$s = \frac{U_{max}}{U_{min}} = \frac{I_{max}}{I_{min}} = \frac{|U_{fwd}| + |U_{back}|}{|U_{fwd}| - |U_{back}|} = \frac{\sqrt{P_{fwd}} + \sqrt{P_{back}}}{\sqrt{P_{fwd}} - \sqrt{P_{back}}}$$

In the case of ideal matching, its value is s = 1. By converting the equation it possible to calculate from the VSWR how much of the power fed into the antenna is reflected back from the antenna to the feed line (Fig. 3.6 and table).

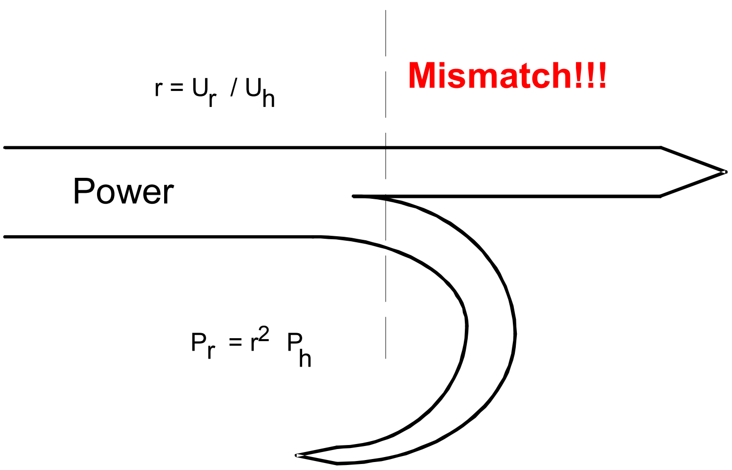


Fig. 3.6 Power reflection due to mismatching

s	1	1.5	2	3	6
Reflected power	0 %	4 %	10 %	25 %	50 %

The voltage standing wave ratio can be measured or the equation

$$s = \frac{Z_E}{Z_N}$$

can be used to calculate it from the feedpoint impedance and the nominal impedance, provided their values are real. The relationship with the magnitude of the reflection coefficient r entered in Fig. 3.6 is specified by

$$s = \frac{1 + |r|}{1 - |r|}$$

The **return loss** derived logarithmically from this

$$a_r = -20 \lg \left(\frac{U_{back}}{U_{fwd}}\right) \text{ dB} = -10 \lg \left(\frac{P_{back}}{P_{fwd}}\right) \text{ dB}$$

is also a common measure of the reflection characteristics of a cable, circuit or antenna.

There are in fact several physical parameters for describing the quality of impedance matching; these can simply be converted from one to the other as required:

Example: If a 50 Ω coaxial cable is connected to an antenna feedpoint having an input impedance of 150 Ω, this yields the voltage standing wave ratio (SWR)

s = 150 Ω / 50 Ω = 3 and thus a power reflection of 25 %.

The associated reflection coefficient (and thus the voltage ratio between the reflected and incident waves) is

r = (3-1) / (3+1) = 2 / 4 = 0.5 and the return loss therefore is

a_r = - 20 lg (0.5) = 6 dB.

Conversion table (values rounded):

SWR	r	a_r in dB	SWR	r	a_r in dB
1.002	0.001	60	1.6	0.23	13
1.004	0.002	54	1.7	0.26	12
1.006	0.003	50	1.8	0.29	11
1.008	0.004	48	1.9	0.31	10
1.01	0.005	46	2.0	0.33	9.5
1.02	0.01	40	2.5	0.43	7.4
1.04	0.02	34	3.0	0.5	6
1.1	0.05	26	4.0	0.6	4.4
1.2	0.1	20	5.0	0.67	3.5
1.3	0.13	18	7.5	0.76	2.3
1.4	0.16	15	10.0	0.82	1.7
1.5	0.2	14			

4 Basic Properties of Selected Antennas

4.1 Dipole Antennas and Rod Antennas

Dipole antennas are the most fundamental form of antenna that can be implemented. The best known example is the tuned (half-wave) dipole (Fig. 4.1).

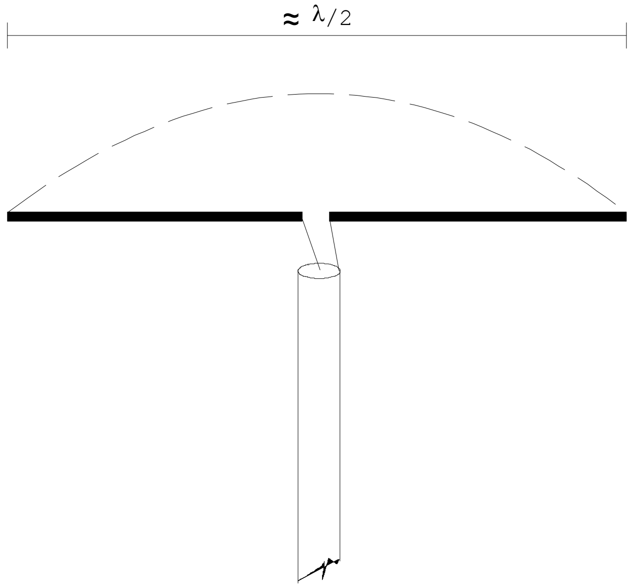


Fig. 4.1 Tuned half-wave dipole

Its length is somewhat less than half a wavelength and its input impedance at resonance is between 50 Ω and 70 Ω depending on its length/diameter ratio, so that a feed cable with a common nominal impedance can easily be connected. The current distribution on the dipole (shown by a broken line in Fig. 4.1) can be assumed to be sinusoidal in good approximation. Its radiation pattern in the E plane - being a reference plane along which the dipole axis lies - is similar to that of a monopole over ideally conducting ground (see Fig. 3.2 curve “A“ in Section 3.1); the radiation distribution in planes perpendicular to its axis (H plane) is uniform.

The name "half-wave dipole" indicates that this form of antenna can be constructed and used for one frequency only. However, experience shows that dipoles can actually be used to receive (but not transmit) broadband broadcasts. It would be possible to conclude from this that half-wave dipoles could be used at least as **test antennas**, even far from their resonance frequency. When used for broadband purposes, however, conventional dipoles experience significant problems in practice:

1. The **antenna input impedance** strongly depends on the antenna length/wavelength ratio (see Fig. 4.2), so that when operated far from the resonance frequency very significant matching problems occur. When receiving at frequencies below 30 MHz, the associated losses are acceptable to a certain extent; this is not particularly critical for measurement, even though conversion from voltage values to field strength values poses problems. For reception in the VHF range and above, however, good matching is required.

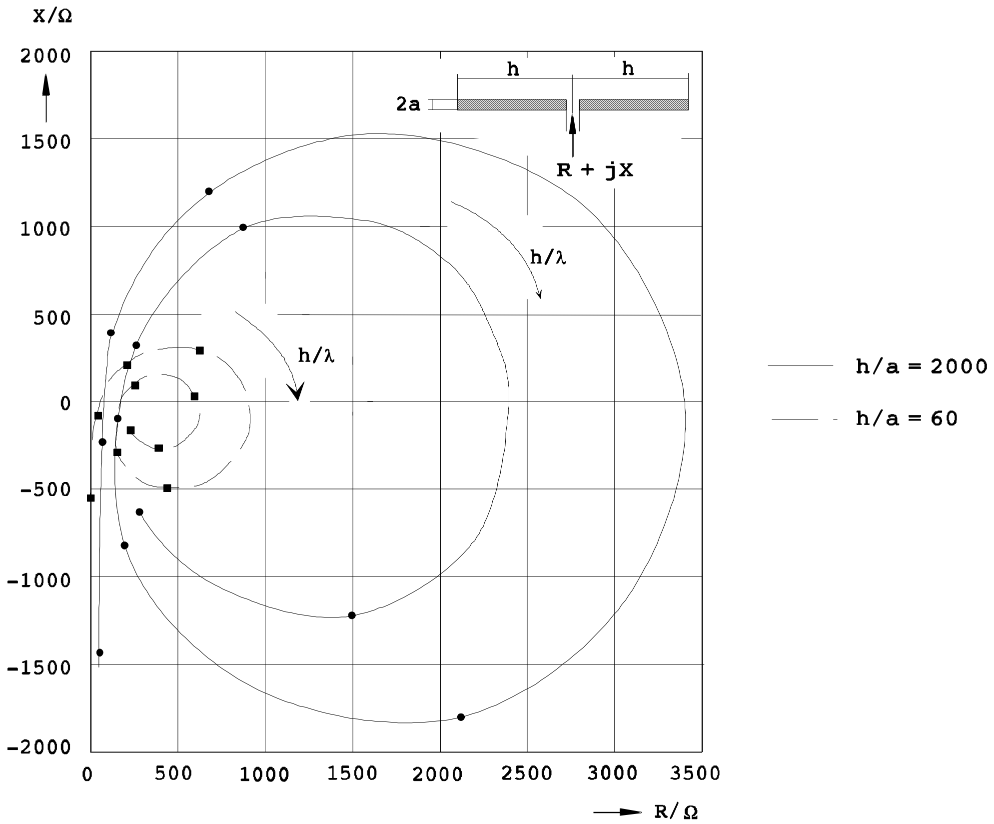


Fig. 4.2 Input impedance of "slim" and "thick" dipoles

2. Above the full-wave resonance, even the **radiation pattern** changes as a function of the antenna length/wavelength ratio to such an extent (Fig. 4.3) that it is no longer possible to clearly determine the main direction of radiation or the gain, for example.

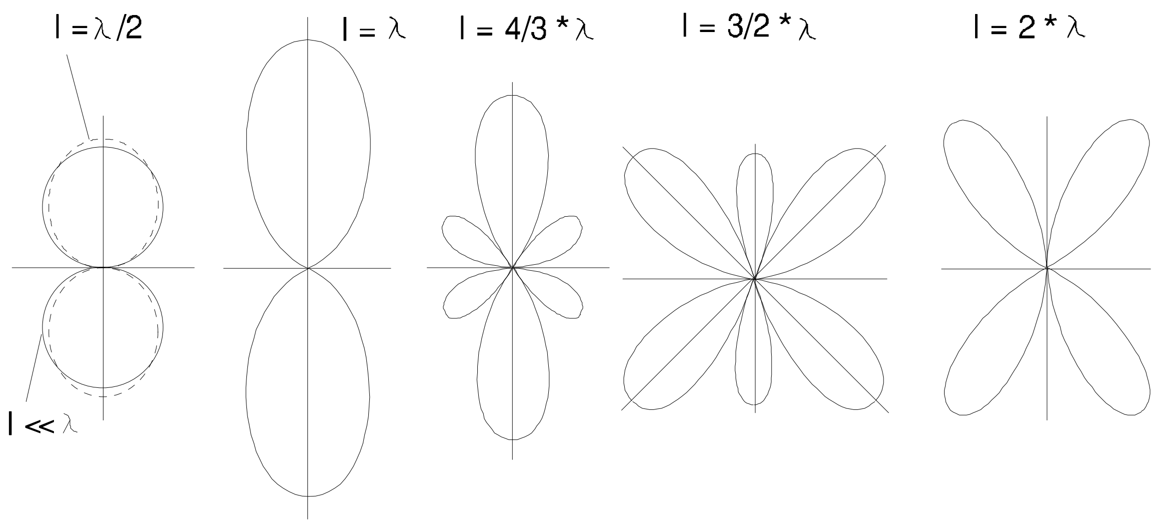


Fig. 4.3 Radiation pattern of a dipole at different frequencies

There are various ways of finding a solution:

1. The antenna rods are designed to function as **telescope rods** or to fit together, so that the antenna length can be varied to match the operating frequency. The dipole can then be operated at resonance at every frequency to which it is set.

2. The ratio of antenna length to diameter is increased. The broken-line curve in Fig. 4.2 shows by way of example the impedance characteristic of a large radius dipole. It can be seen that with decreasing length/diameter ratio the antenna characteristics depend less on the frequency. Fig. 4.4 shows some possible shapes for **"thicker" dipoles**. In order to save materials and particularly weight, such dipoles are often designed in the form of a **cage**. Another commonly used test antenna is the **biconical antenna**, which is also usually designed in the form of a cage (on the very right in Fig. 4.4).

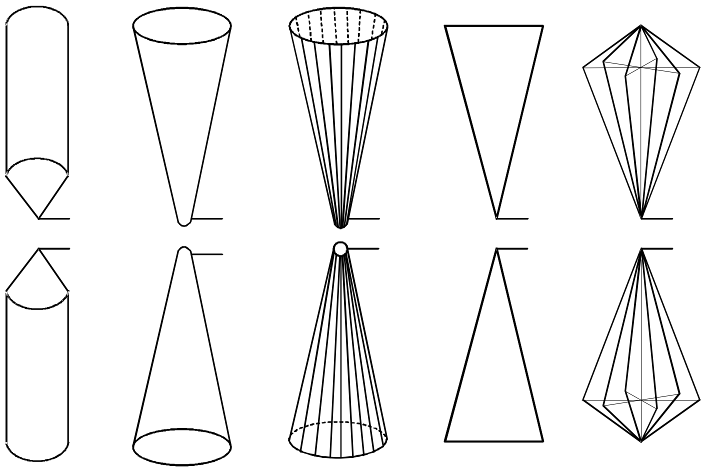


Fig. 4.4 Possible shapes of broadband dipoles

3. The **current distribution** on the antenna is altered (for instance by reactive elements, resonant circuits or ferrite rings) in such a way that at high frequencies only part of the antenna is activated (Fig. 4.5). This keeps the ratio of wavelength to antenna length almost constant even though the frequency may vary. Electrically this solution is more or less the same as the telescopic antenna described at 1. above, but with no need for the user to go to any effort or expense.

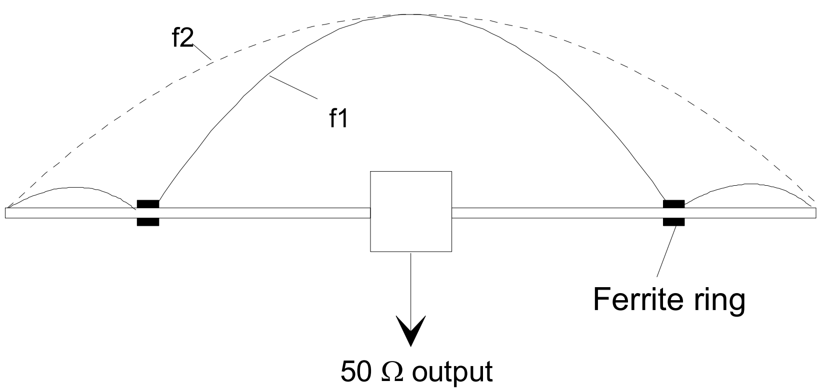


Fig. 4.5 Varying the antenna length by means of reactive elements

This principle is used in active antennas for the VHF/UHF range, for example. The considerable dipole length that can be obtained in this way produces high sensitivity even at the lower frequency limit. Fig. 4.6 shows an example in the frequency range from 200 MHz to 1000 MHz.

Fig. 4.6 Active receiving dipole with frequency-dependent dipole length

4. **Traveling waves** are generated on the antenna, for instance by a termination at the end of the radiator structure. Rhombic and Beverage antennas are based on this principle, but are now of only historical importance. A broadband version which is in use today is the terminated folded dipole (TFD) in Fig. 4.7, a loaded loop antenna whose dimensions can be kept so small that it is used in transportable systems for shortwave communications.

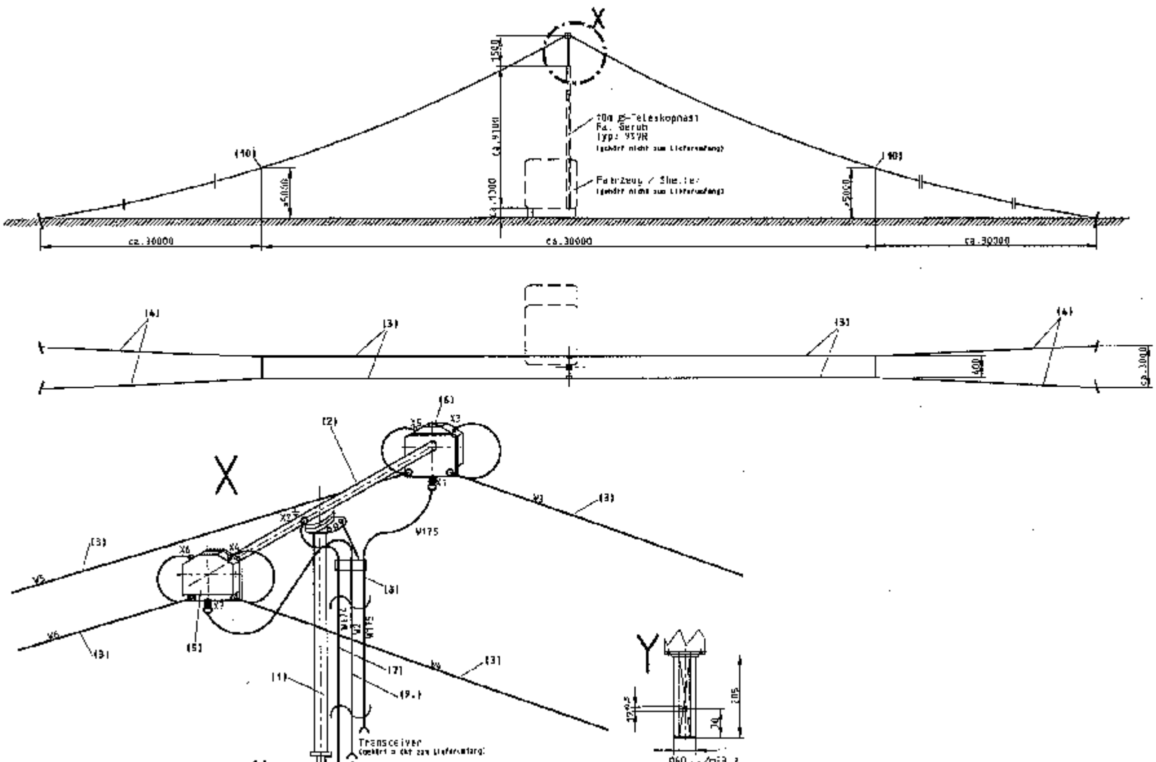


Fig. 4.7 Broadband loaded loop antenna for the shortwave range

In each of the versions described, matching the impedance to the feed cable is important, but so too is the transition between the cable, which nowadays is usually coaxial (i.e. unbalanced), and the dipoles, which are balanced (Fig. 4.8).

Unless a **balun** is used, skin currents form on the outer conductor of the coaxial cable, which

- can cause massive electromagnetic interference (EMI) when the antenna is operated as a transmitting antenna
- significantly alter the radiation pattern and other radiation characteristics
- corrupt the test results when the antenna is operated as a test antenna.

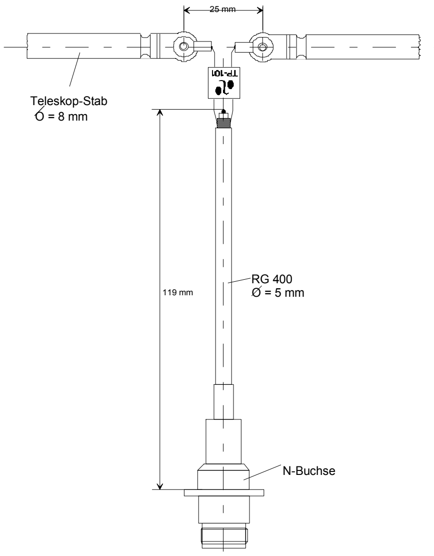


Fig. 4.8 Dipole feedpoint with balun and connecting cable

The operating principle of **rod antennas** (or **monopoles**) is based on the fact that the current distribution on an antenna structure that is only a quarter wavelength long is identical to that on a half-wave dipole (see Fig. 4.1) if the antenna element "missing" from the dipole is replaced by a highly conducting surface. As a result of this reflection principle (Fig. 4.9), vertical quarter-wave antennas on conducting ground have basically the same radiation pattern as half-wave dipole antennas. There is of course no radiation into the shadowed half of the space. The directivity factor compared to a dipole in free space thus increases by 3 dB to 5.1 dB. The input impedance is half that of a dipole, exhibiting values between 30 Ω and 40 Ω.

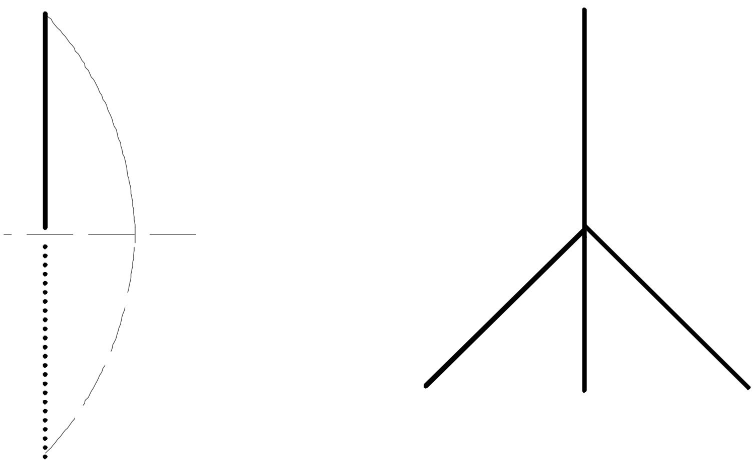


Fig. 4.9 Reflection principle and ground-plane antenna

Monopoles are to be found in the same number of different configurations as dipoles. In the lower frequency ranges they allow vertically polarized waves to be radiated. In the long and ultralong wavelength ranges they are the easiest form of antenna to implement. Broadband shortwave monopoles often come in the form of cage antennas (see Fig. 4.4; 4.10).

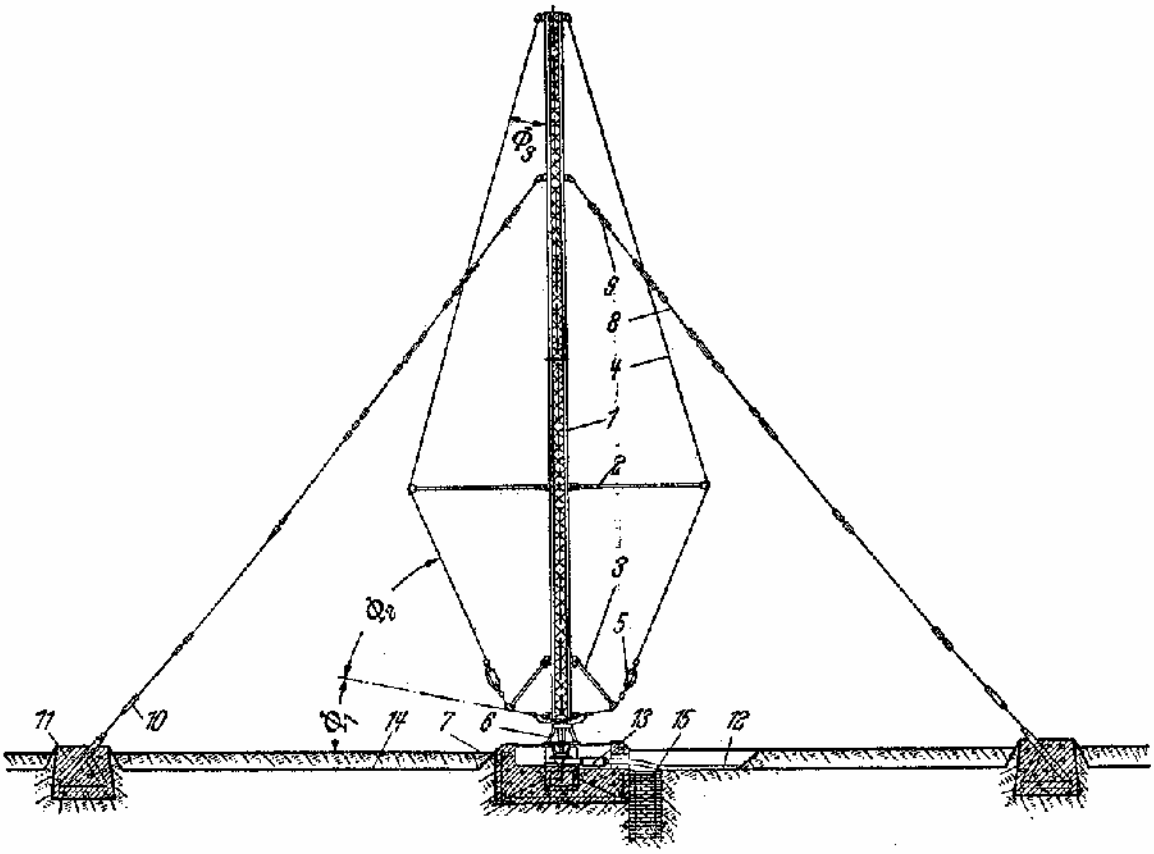


Abb. 19.4. Praktische Ausführung einer Exponentialantenne in Reusenform (Rohde & Schwarz, München).

Fig. 4.10 Shortwave broadband antenna (cage antenna)

The conducting surface on which the monopole is erected plays an important part in enabling the reflection principle to take effect. Even on reasonably conducting ground (such as a field of wet grass) and particularly on poorly conducting ground (dry sand) it is usual and helpful to put out a ground net of wires.

Monopoles are used as vertically polarized omnidirectional antennas (**ground-plane antennas**) even in the VHF/UHF range. The conducting ground can be very easily simulated from wires or rods (known as radials). If these radials are not mounted at a right angle as in the case of installation on actual ground, but instead are attached at an angle of 135° to the quarter-wave dipole (Fig. 4.9), the value of the feedpoint impedance increases to between 50 Ω and 60 Ω, making more sophisticated measures for matching the nominal impedance of commonly used coaxial cable superfluous. Since monopoles are asymmetrical antennas which do not need to have their symmetry converted, they can in many cases be connected directly to their feed cable.

Rod antennas on handheld radio units and mobile telephones use as their electrical ground plane the instrument chassis and the person operating the unit, who is capacitively coupled to it. By contrast, in the case of vehicle antennas it is possible to assume well-defined ground ratios, provided the vehicle has a metal roof.

4.2 Loop Antennas

In the case of loop antennas, the dipole consists of a usually circular or rectangular current loop which acts like a coil and functions in response to the **magnetic component of the electromagnetic field**. A distinction is generally made between loops with a circumference of about a wavelength (tuned loops, cubical quad) and those with a very small circumference relative to the wavelength. The latter are used in preference both as DF antennas and in field-strength measurement for low frequencies (below 30 MHz). In view of their small dimensions the current in the entire loop can be regarded as locally constant. Consequently they generate the same sort of field as a Hertzian dipole, i.e. a very short electrical antenna on which the current distribution is a function of position to an equally small extent. The equations for calculating the field and radiation patterns of a Hertzian dipole can therefore be used after the components of the electric and magnetic field strengths have been transposed. The radiation pattern is the same as for a Hertzian dipole (Fig. 4.11), which is why small loop antennas are also known as magnetic dipoles.

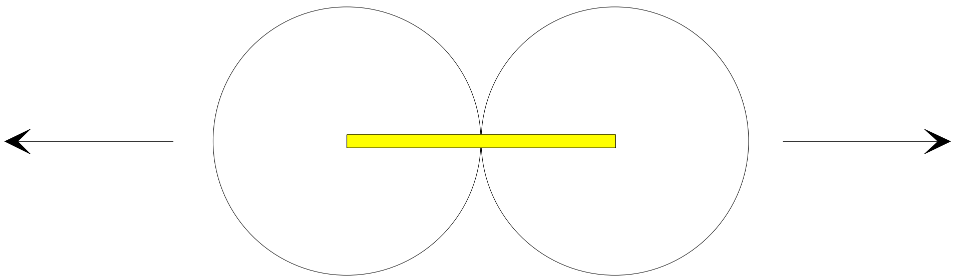


Fig. 4.11 Radiation pattern of a small, vertically installed loop antenna (seen from above) showing the main direction of radiation

The effective height of a loop antenna with loop diameter D can be calculated by means of the equation

$$h_{eff} = n \cdot \frac{\pi^2 D^2}{2 \cdot \lambda}$$

and can be increased by adding more than one winding n. For a loop diameter of 26 cm with n = 5 windings and f = 3.5 MHz, the effective height is just 19 mm. In addition, the radiation resistance is very low:

$$R_s = 20\ \Omega\ \pi^6\ n^2\ (D/\lambda)^4$$

The equation indicates that increasing the loop diameter D has a stronger effect than increasing the number of windings n.

A practical version of a single-winding loop for test purposes is shown in Fig. 4.12. As a protection against electric field effects, the actual winding is additionally surrounded by a tubular metal shield that must have a slot at one point in order not to short-circuit the magnetic field.

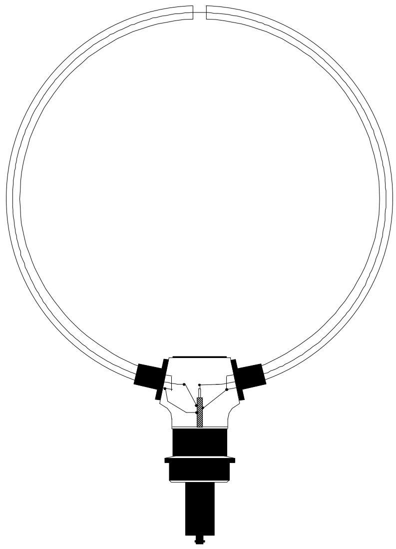


Fig. 4.12 Loop antenna

In view of their low radiation resistance, loop antennas are now used **almost always for reception purposes only (DF and test antennas)**. They have become less significant as transmitting antennas for frequencies below 30 MHz. An exception to this is the **tuned transmitting loop**, which can be equipped with a remotely controlled capacitor to make a resonant circuit (a receiving loop can be provided with additional selectivity and sensitivity in the same way). However, such loops are extremely narrowband systems and therefore have to be retuned whenever the frequency is changed (even in the same operating band). Nonetheless they are sometimes the only practicable option for transmission when space is restricted.

4.3 Active Antennas

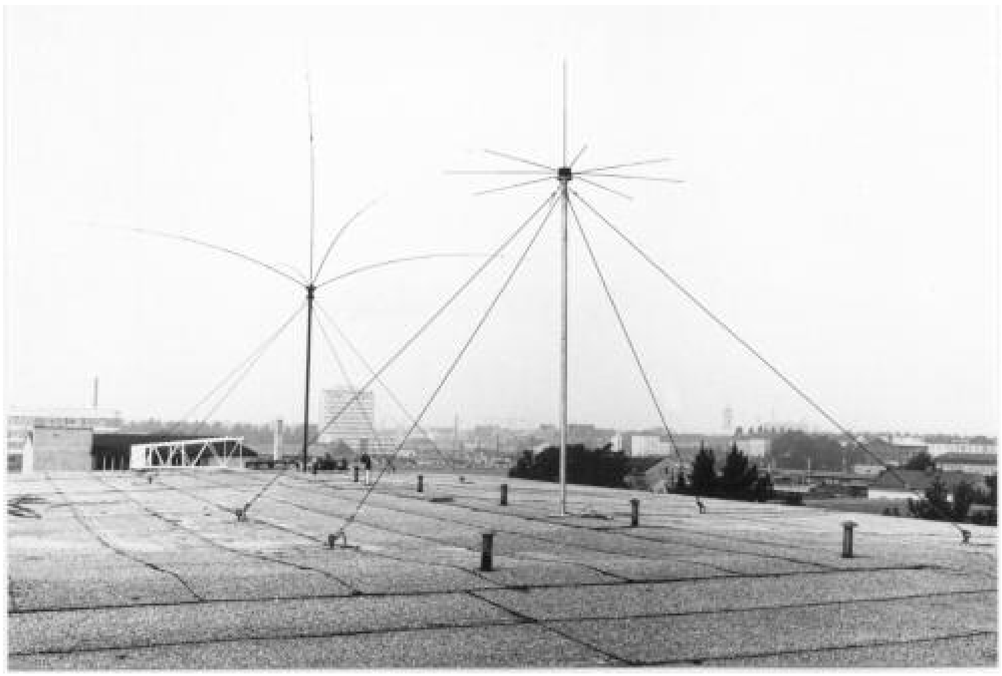


Fig. 4.13 Size comparison between passive (left) and active RF receiving antennas

Active antennas represent another possible way of implementing compact broadband antennas. They are based on the idea that drastically **shortening the dipole length** of an antenna will result in a corresponding reduction in output voltages for both the useful signal and the noise. As a consequence, the **signal-to-noise ratio,** which is the sole indicator of reception quality, **stays constant** within fairly wide limits (see Fig. 4.14)

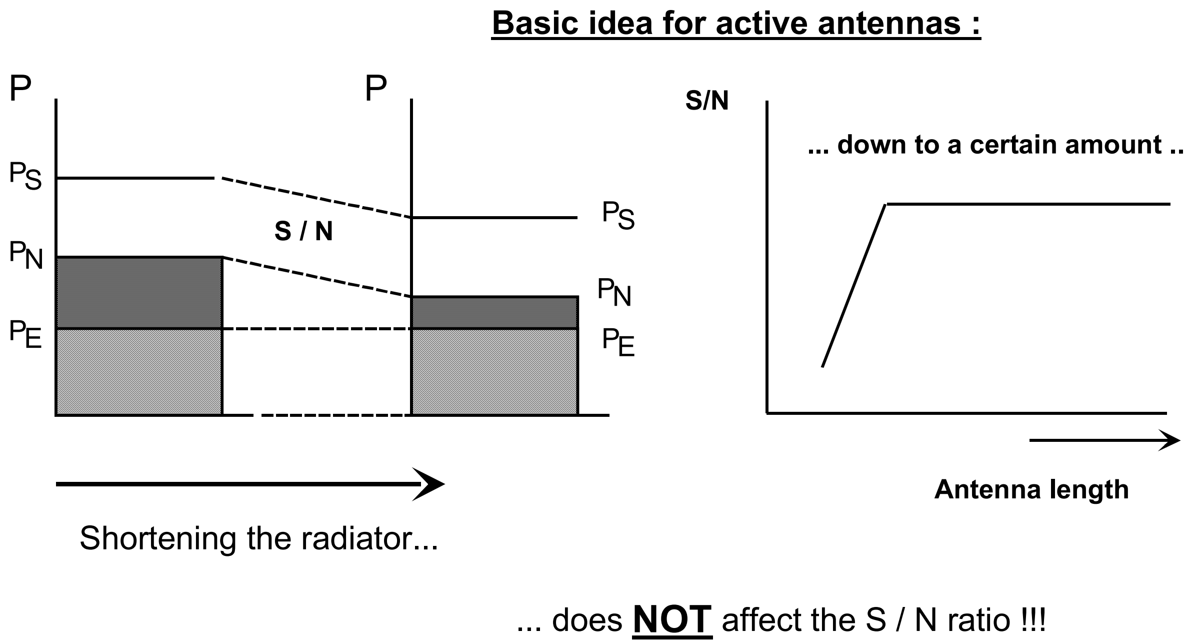


Fig. 4.14 Signal-to-noise ratio when the dipole length is shortened

The extreme change in impedance associated with shortening (see Fig. 4.2) is compensated in rod and dipole antennas by feeding the voltage on the terminals of the antenna directly to a very high-impedance active component (usually a field effect transistor) which acts as an impedance transformer and also commonly amplifies at the same time. Active antennas are therefore by definition antennas in which an **active element is attached directly to the dipoles** (Fig. 4.15), and are not to be confused with systems in which the output signal of a passive antenna is looped through a preamplifier, for example.

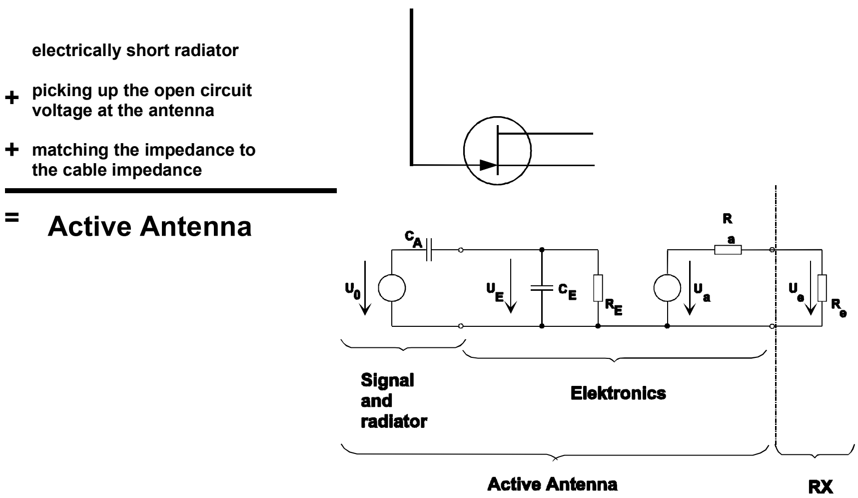


Fig. 4.15 Principle of an active antenna

The advantage of using active antennas is the fact that usually their radiation patterns are no longer a function of frequency because they have electrically short dipoles (see Fig. 4.3 left, continuous line). By carefully tuning the electronics to the antenna geometry and by using other measures, it is also possible to ensure that the antenna factor is also largely independent of frequency (see Fig. 4.16), so that field strength measurements can be carried out very easily.

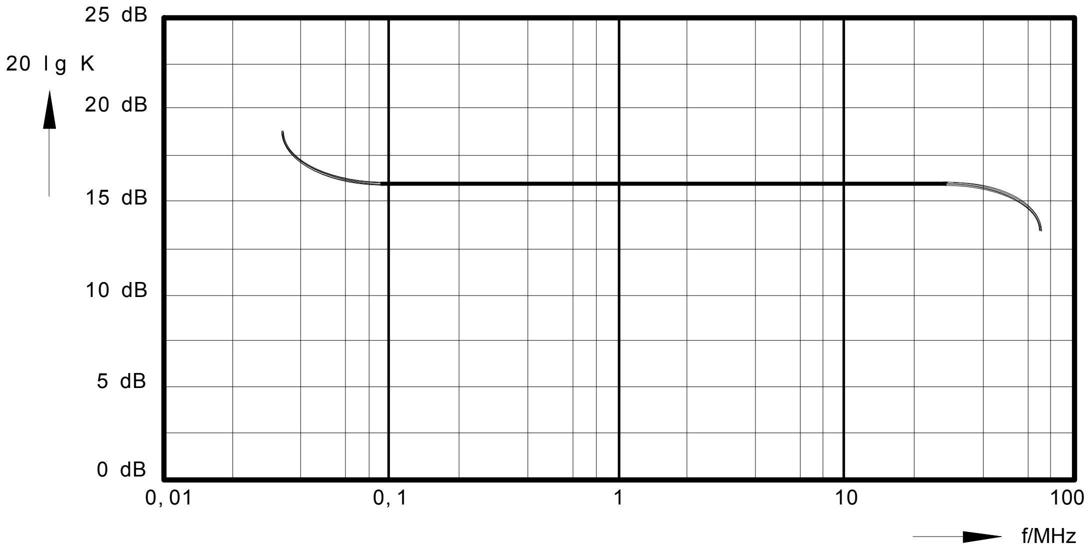


Fig. 4.16 Frequency response of the antenna factor of an active antenna

Possible versions of active antennas include all the previously mentioned radiator types – rod, dipole and loop antennas.

Active antennas are mainly but not exclusively used for low frequencies (up to about 200 MHz) at which the noise floor in the atmosphere is very high. Fig. 4.15 shows on the right a combination of active antennas in which the rod antenna covers the wide frequency range from 10 kHz to 80 MHz . The horizontal dipoles are dimensioned for 600 kHz to 40 MHz.

Due to the extremely broadband characteristics, active antennas are being increasingly used even in the higher frequency ranges (see also Fig. 4.6 and directional antennas).

Fig. 4.17 Active turnstile antenna for 20 MHz to 500 MHz

Fig. 4.17 shows an active turnstile antenna for omnidirectional reception of horizontally polarized waves in the wide frequency range from 20 MHz to 500 MHz. Unlike with active antennas for the long-, medium- and shortwave ranges, it is no longer possible to speak exclusively of electrically short dipoles. At higher frequencies, moreover, electronic noise is no longer negligible compared to the lower ambient noise. Therefore careful noise matching is required instead of high-impedance source voltage tapping at the antenna terminals.

In view of the compact dimensions of active antennas, the question of the preferred **direction of polarization** can almost always be made dependent on the problem that needs to be solved. Test specifications very often define the polarization direction. It turns out to be an advantage that a single active dipole is often enough to cover the entire frequency band in which the test is to be performed, and with such a compact and handy configuration even an obligatory change of orientation can be carried out in almost no time.

Active antennas for radiomonitoring are oriented in the same direction as the expected signals. This can be done very easily for the VHF/UHF range, because the polarization of the signals is mostly predictable. Shortwaves, for example, undergo a certain amount of reflection in the ionosphere and then exhibit elliptical polarization directions which are impossible to determine. If horizontally polarized dipoles were used, the direction of incidence would also be known, so the normal practice is to use an active monopole (Fig. 4.18). A considerably more effective, though more complex alternative is to combine two crossed, horizontally polarized active dipoles, the signals from which are added by means of a 90° coupler to obtain a horizontally omnidirectional radiation pattern (**turnstile antenna**). A turnstile antenna with a monopole (Fig. 4.19) and remote-controlled switching will be able to handle almost any imaginable reception scenario.

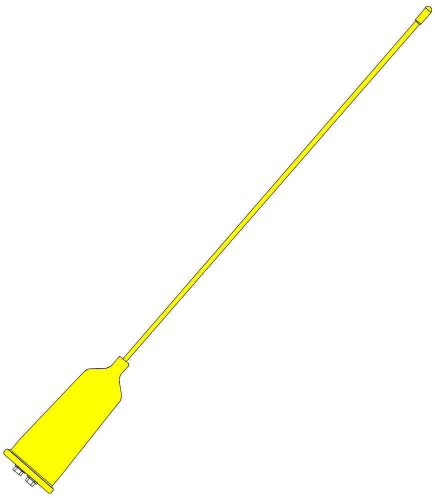
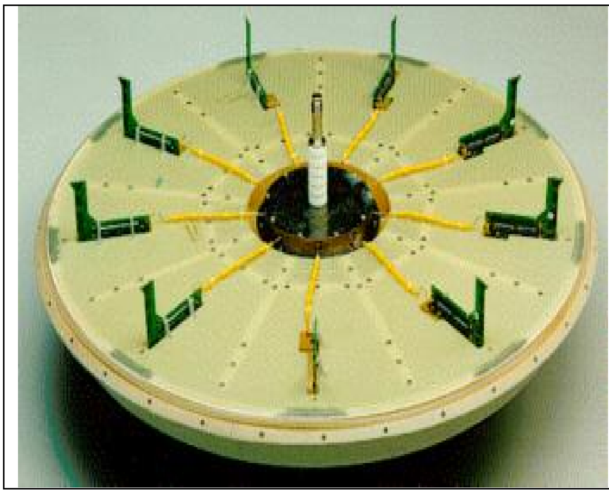


Fig. 4.18 Active monopole

Fig. 4.19 Active turnstile antenna with monopole

Active antennas are also used in direction finding (Fig. 4.20) due to the characteristics mentioned above, such as low coupling with adjacent antenna elements, compactness, as well as broadband characteristics.



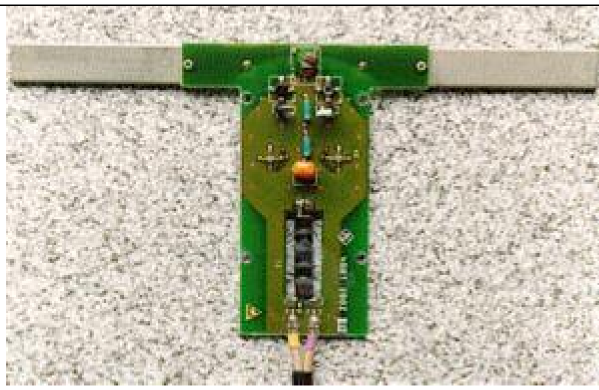


Fig. 4.20 Interferometer DF antenna (20 MHz to 1300 MHz), left, and active antenna element (right)

Here is a summary of the characteristics and rules for using active antennas:

Active antennas ...

... are smaller than comparable passive antennas

... are more broadband than comparable passive antennas

... cannot transmit

... have little coupling with their environment

... are more susceptible to interference if incorrectly installed

... are highly suitable for use as broadband test antennas

... have a frequency-independent radiation pattern

... must have sufficient large-signal immunity

... need to be very carefully balanced

... must not be installed in the midst of interference

... may be installed closely adjacent to one another

... are actually not so bad as many believe.

4.4 Directional Antennas

It has already been mentioned that ideally omnidirectional antennas cannot be produced in reality. Nonetheless only antennas that focus their radiated power in a particular spatial direction can properly be called directional antennas. At an equivalent transmitter power, they significantly improve the signal-to-noise ratio, but must be focused on the distant station so that in many cases a **rotator** has to be used. Parameters used to measure the ability to suppress signals from directions other than the main direction of radiation include not only those defined in Chapter 3, i.e. gain and directivity factor, but also the **front-to-back ratio**:

$$v = 20 \lg \frac{\text{Far field strength in main direction of radiation}}{\text{Far field strength in backward direction}} \text{ dB}$$

However, since this measure provides no information about the number and size of the radiated **side lobes,** and the gain and directivity factor provide only very little, the **radiation pattern** must always be known in order to fully assess the radiation behavior of directional antennas (Fig. 4.21).

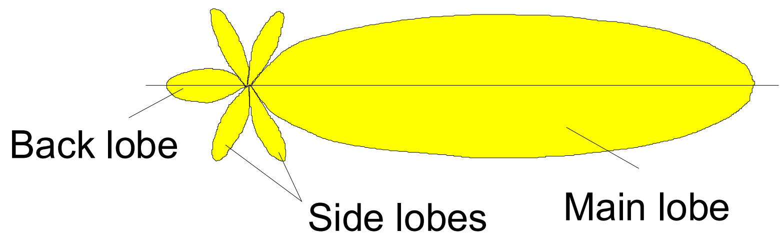


Fig. 4.21 Radiation pattern of a directional antenna (main lobe, side lobe and back lobe)

Two omnidirectional antennas that are set up within a predefined distance of one another and fed with different phases, as shown in Fig. 4.22, constitute a directional antenna in its simplest form. In the example shown, a distance of λ/4 and a phase shift of 90° have been chosen, resulting in a cardioid-shaped radiation pattern when **the far field strengths generated by the two individual radiators are added** (Fig. 4.23). Even though this configuration cannot be said to produce strongly focused radiation, the sharply defined null provides effective **suppression of interfering signals**.

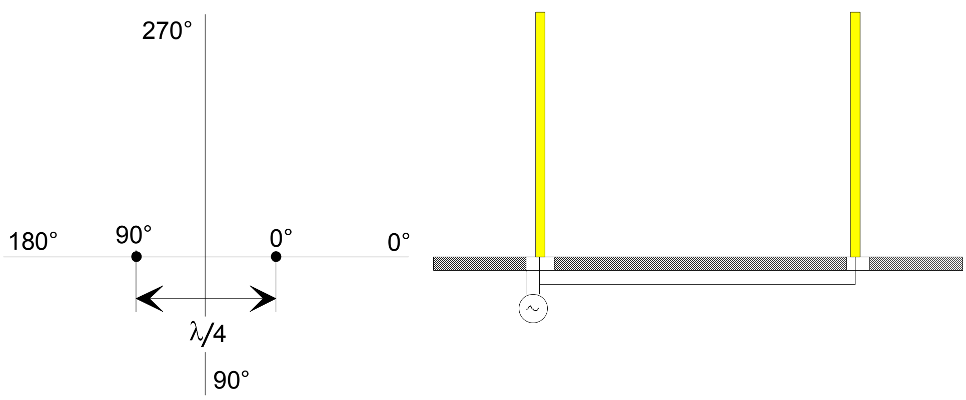
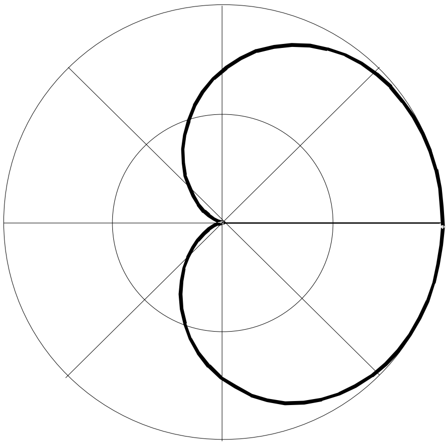


Fig. 4.22 Principle of a simple directional antenna



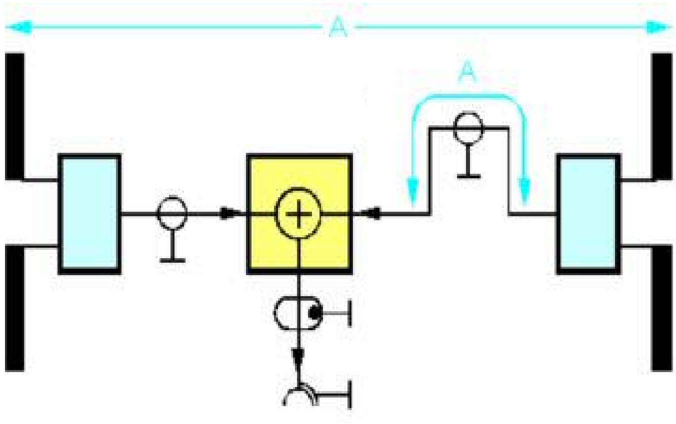


Fig. 4.23 Cardioid pattern of the directional antenna in Fig. 4.22

Fig. 4.24 Principle of a broadband VHF directional antenna with cardioid pattern

The principle of a broadband directional antenna with a cardioid pattern for the 20 MHz to 87 MHz range is shown in Fig. 4.24. Since they are designed as active antennas, the individual elements are only 1 m long. A distance of 1 m between the individual radiators is the optimum for the specified frequency range.

The active directional antenna with even smaller dimensions, shown in Fig. 4.25, produces a broadband cardioid pattern. This antenna is designed for use with portable receivers, and covers the 20 MHz to 500 MHz frequency range by means of two exchangeable modules designed as active loop antennas (a further module in the form of a log-periodic antenna extends the frequency range to 3 GHz. This type of antenna will be described in more detail in a later section).

Fig. 4.25 Portable active loop antenna for the VHF range

By combining two or more dipoles and carefully selecting phase shifts and distances between radiators, it is possible to use **diagram superposition** to generate radiation patterns whose focusing is limited by the available space. In contrast to the cable-fed dipole shown in Fig. 4.22, **Yagi-Uda antennas** use the principle of **radiation coupling**, in which only one dipole is directly fed by the cable and its radiation feeds the other antenna elements. These directional antennas are mainly used for receiving TV and VHF broadcasts, and typically have between 4 and 30 elements. An average number of elements produces gains of around 10 dB; a high number produces significantly more.

In the long-, medium- and shortwave ranges, the method of using several dipoles to generate directional patterns very quickly runs into space limitations. All the same, long- and medium-wave broadcast transmitters quite frequently use a dipole configuration of two or three monopoles similar to that in Fig. 4.22, in order either to cover a specific area or to avoid interference with their own or external stations (Fig. 4.26).

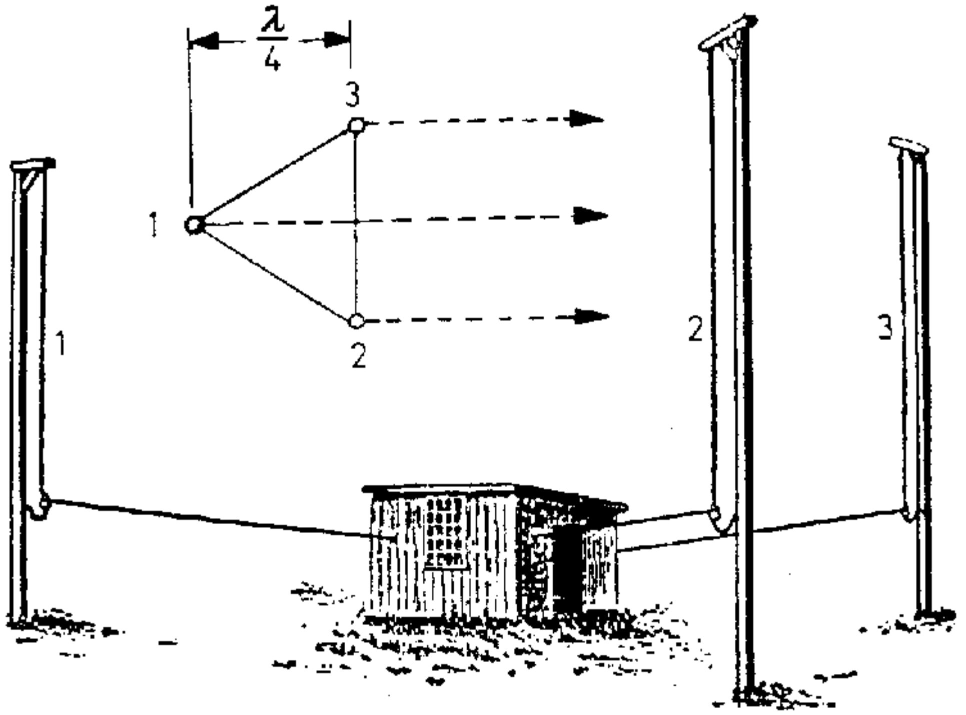


Fig. 4.26 The first directional antenna array (Karl Ferdinand Braun, 1905)

In the shortwave range it is quite usual (and even necessary) to use configurations of horizontally and vertically **stacked dipoles** in order to make sure the transmitter power is acceptable enough to be received reliably in the target coverage area. A forerunner of these **antenna arrays** was the pine-tree antenna (Fig. 4.27), first used by Grossdeutsche Rundfunk (German broadcast organization) in 1932 for foreign coverage, and taking its name from the shape of its feed elements. Its successors are used today in scarcely modified form by many international broadcast stations under the name **curtain antenna**. Unlike Yagi antennas, curtain antennas use **cable feed**. Varying the phase shifts between the individual dipoles can swing the radiation pattern. In some transmitters the curtain antennas are additionally mounted on revolving platforms so that they can transmit in any direction.

The ability to change the main direction of radiation of a sharply focusing directional antenna by purely electronic means is used to an increasing extent even in the case of antenna arrays for very high frequencies (such as for satellite radio services). Such **planar antennas** mostly take the form of a dipole curtain which, unlike a curtain antenna, is constructed facing a conducting surface. This makes it possible to etch the dipoles as conductor paths on a copper-coated board (**microstrip antenna**). This method is a cost-efficient and highly precise way of manufacturing even larger array antennas for the microwave range.

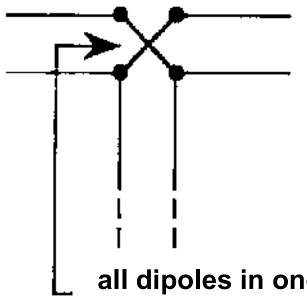
The wide range of gain values, the ability to adapt the radiation patterns to special tasks and the maturity of the technology are such great advantages of array antennas that the question arises why yet more forms of directional antennas are actually required. The answer is the maximum obtainable frequency range of one octave. In the case of antenna arrays, this range is a considerable limitation on all applications that must cover wide frequency ranges.

TANNENBAUM–ANTENNE:

Fir tree antenna,

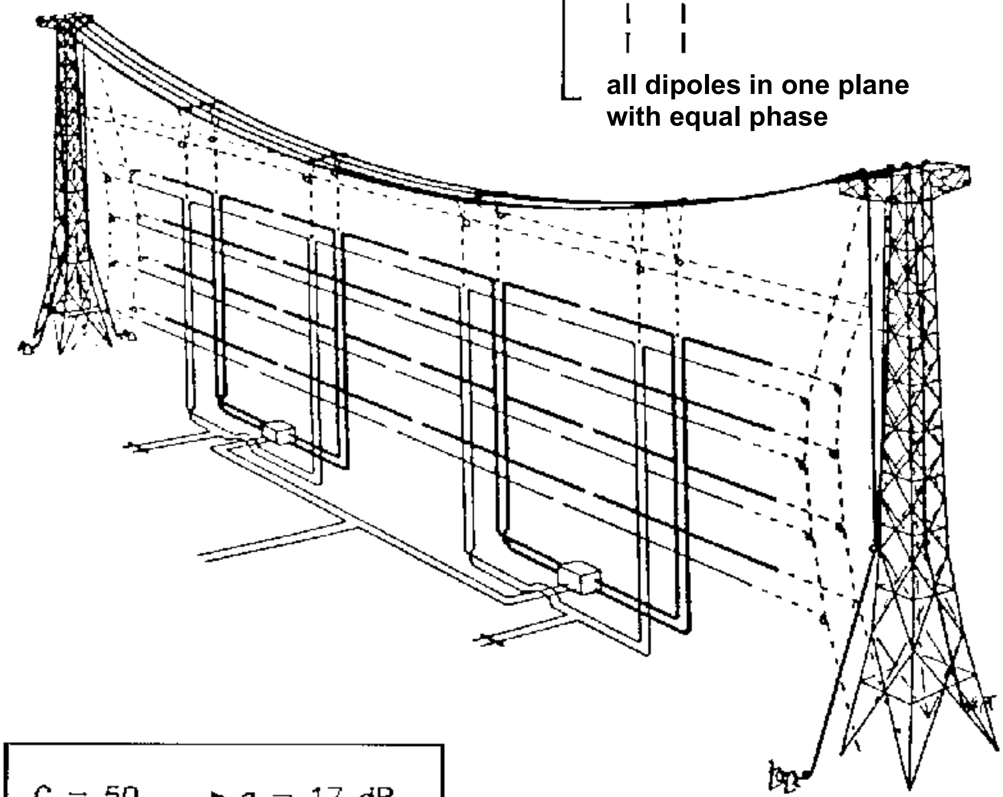
Böhme and Moser, Telefunken
German patent 1928

operational Jan 22, 1932
Königs Wusterhausen near Berlin



"similar to the branches of a fir tree"

all dipoles in one plane with equal phase



$G = 50 \longrightarrow g = 17$ dB

<u>Full wave dipoles:</u>

horizontal distance	**λ**	
vertical distance	**λ/2**	
reflector curtain	**λ/4**	**switchable, hence change of main beam possible**

Nowadays: "Curtain antenna"

Fig. 4.27 The curtain antenna as a predecessor of modern shortwave-directional antennas

Log-periodic antennas are the newest special form of directional antenna and provide a solution: In principle, they can cover any frequency range. In practice, their frequency ranges are limited:

- In the low-frequency range only by space requirements and mechanical outlay.
- In the high-frequency range by manufacturing accuracy.

The most common version is the log-periodic dipole antenna (LPDA). Beam shaping is performed by several active elements. The LPDA consists of a number of parallel dipoles gradually increasing in length and spacing (Fig. 4.28) so that the angle α formed by connecting the dipole peaks to the longitudinal axis of the antenna is constant, and so that not only the distances between radiators s_n but also their lengths l_n obey the law

$$\frac{l_{n+1}}{l_n} = \frac{s_{n+1}}{s_n} = \text{const.}$$

If such a configuration is fed from the front (starting with the shortest dipole), the electromagnetic wave first of all passes through all dipoles that are clearly shorter than half the wavelength. However, the dipoles that are just about half a wavelength long are brought to resonance and form the active zone which radiates the electromagnetic wave (in the direction of the shorter dipoles). The conducted wave therefore does not reach the longer elements.

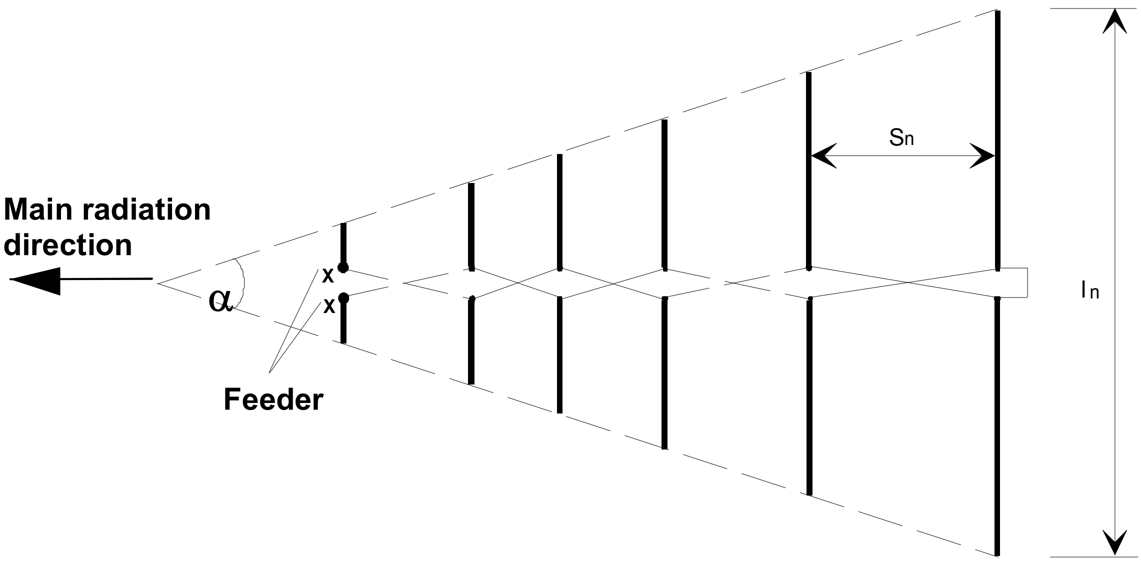


Fig. 4.28 Log-periodic dipole antenna

Due to the fact that only some of the dipoles contribute to shaping the radiation pattern, the directivity (and therefore the gain) that can be achieved by LPDAs is clearly less than is the case for example with Yagi antennas of comparable geometry and dimensions. However, the advantage of the LPDA is that in theory its bandwidth can be increased as desired.

In the shortwave range, rotatable LPDAs designed for 5 MHz to 30 MHz are used both in broadcast transmitters and in commercial radio services. If space is limited, such as on roofs, suitable shaping of the dipoles will enable LPDAs to be reduced in size compared to a configuration of conventional dipoles (Fig. 4.29).

If the entire radiotelephony and shortwave range from 1.5 MHz to 30 MHz needs to be covered, rotatable log-periodic antennas would be unacceptably large. Therefore, they are not designed to rotate in practice, but instead are suspended between a number of masts at a suitable height above the ground. Depending on their orientation for horizontal polarization they can then cover an azimuth range of about 70° within their half-power beamwidth. For vertically polarized log-periodic antennas, the half-power beamwidth amounts to around 120°.

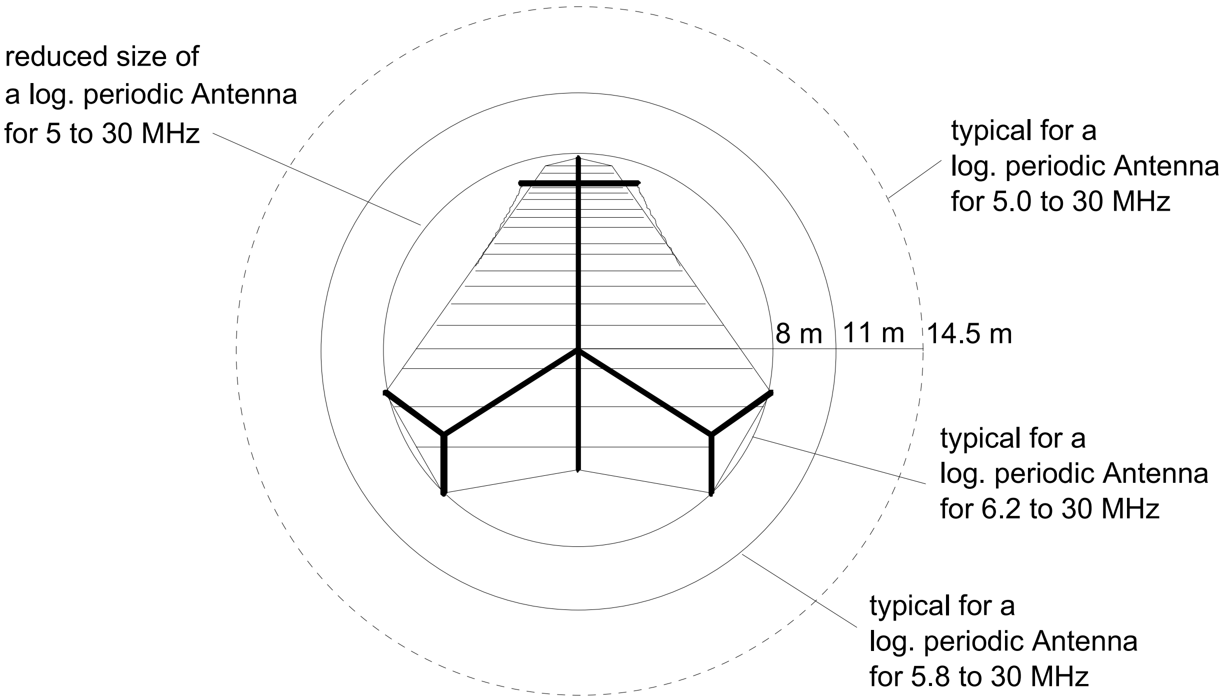


Fig. 4.29 Possible size reductions in shortwave LPDAs

Where wider azimuth ranges are needed, starshaped configurations or parts thereof, as shown in Fig. 4.30, are of advantage: Complete all-round coverage for horizontal polarization is possible using six antennas. In the case of vertical polarization, three antennas are enough due to the larger half-power beamwidth. These starshaped configurations ensure a high degree of decoupling between individual antennas and are not only space-saving but also offer cost advantages due to the multiple use of the masts.

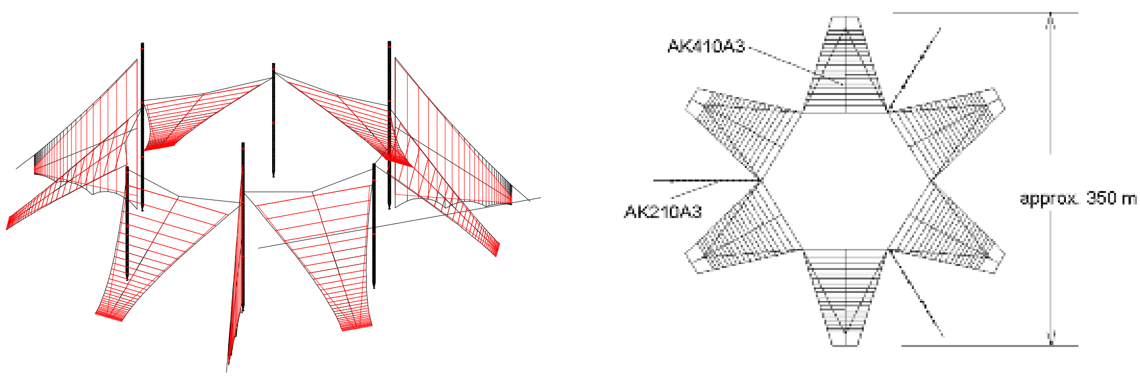


Fig. 4.30 Starshaped configuration of RF LP antennas

LPDAs for higher frequencies can be designed for still broader bandwidths and can cover for example the frequency range from 1 GHz to 26.5 GHz (Fig. 4.31 left) or 1 GHz to 18 GHz (Fig. 4.31 right), which qualifies them not only as communications antennas, but also as easy-to-use test antennas having the further advantage that the radiation pattern and gain do not change significantly versus frequency.

Fig. 4.31 Log-periodic dipole antennas for the microwave range

In radiomonitoring, it is quite common not to know the polarization of an incident signal of interest. In the shortwave range, the configuration of separate horizontal and vertical LP antennas shown in Fig. 4.30 provides the solution for this problem. For higher frequencies LPDAs can also be designed with two crossed dipole arrays as an alternative. The antennas shown in Fig. 4.31 right and Fig. 4.32 can thus be used to receive vertically and horizontally polarized signals simultaneously. Appropriate feed design or the addition of further networks permits reception of circular-polarized waves.

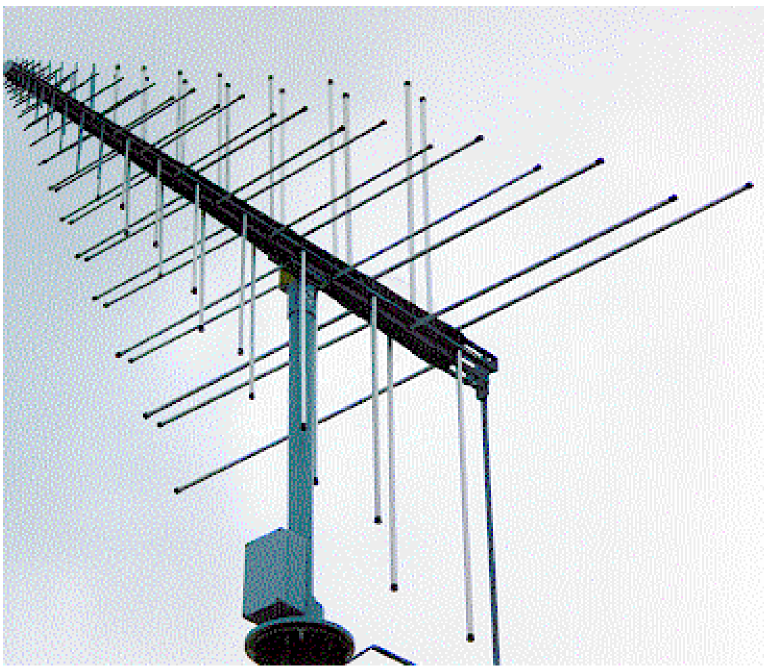


Fig. 4.32 Crossed log-periodic antenna for 80 MHz to 1300 MHz

Particularly in the higher frequency ranges, radiated transmission energy can also be focused by reflectors using geometrically optical or acoustic principles. The best known antenna of this type is the **parabolic reflector**, in which a feed such as a dipole or directional antenna is placed at the focal point of a conducting paraboloid of revolution which focuses the radiation because of its geometric shape (Fig. 4.33). The reflection theory described in a previous section makes it clear that this physical procedure is completely identical to beam shaping by means of a number of individual radiators. In some cases, provided the reflector design is appropriately modified, the feed is even placed outside the beam path in order to avoid shadowing.

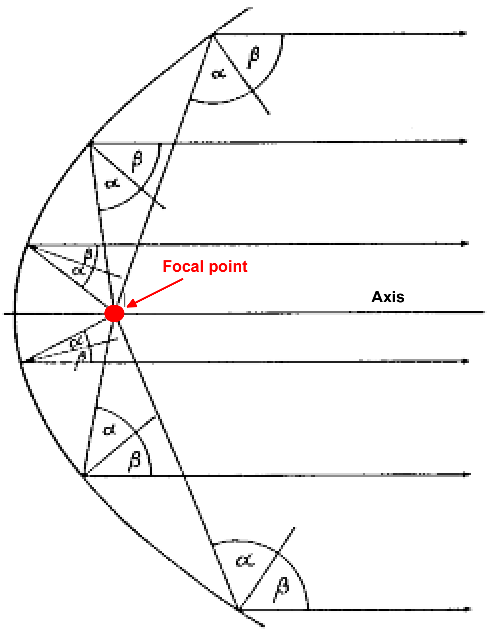


Fig. 4.33 Beam shaping by means of a paraboloid of revolution

By using parabolic reflectors of an appropriate size and suitable feeds, very **high gain values** (50 dB and above) can be obtained. The following table shows the gain values obtained over a specific frequency range for a certain number of reflector diameters. The associated half-power beamwidths of the radiation patterns are also specified.

Antenna system	**Reflector diameter**	**Frequency range**	**Gain**	**Half-power beamwidth (HPBW)**
R&S AC090	90 cm	1 GHz to 18 GHz	15 dB to 40 dB	19° to 1.3°
R&S AC120	120 cm	1 GHz to 18 GHz	19 dB to 42 dB	16° to 0.9°
R&S AC180	180 cm	1 GHz to 18 GHz	23 dB to 45 dB	12° to 0.7°
R&S AC300	300 cm	1 GHz to 18 GHz	26 dB to 50 dB	6° to 0.35°

Reflectors with larger diameters naturally increase total system costs. This is partly due to the cost of the reflector, but mainly attributable to the rotator. The rotator has to be designed not only for greater weight and higher wind loading, but also, because of the smaller half-power beamwidth, for higher setting accuracy.

The question therefore arises whether the increased financial outlay offers any advantage.

The higher gain values are intermediate quantities only, since they do not allow direct assessment of whether a monitoring task can be performed successfully. It is particularly helpful to investigate the signal field strengths at which adequate reception is obtained when using reflectors of different diameters. Fig. 4.34 shows the field strengths that can still be received with the reflector diameter as the parameter.

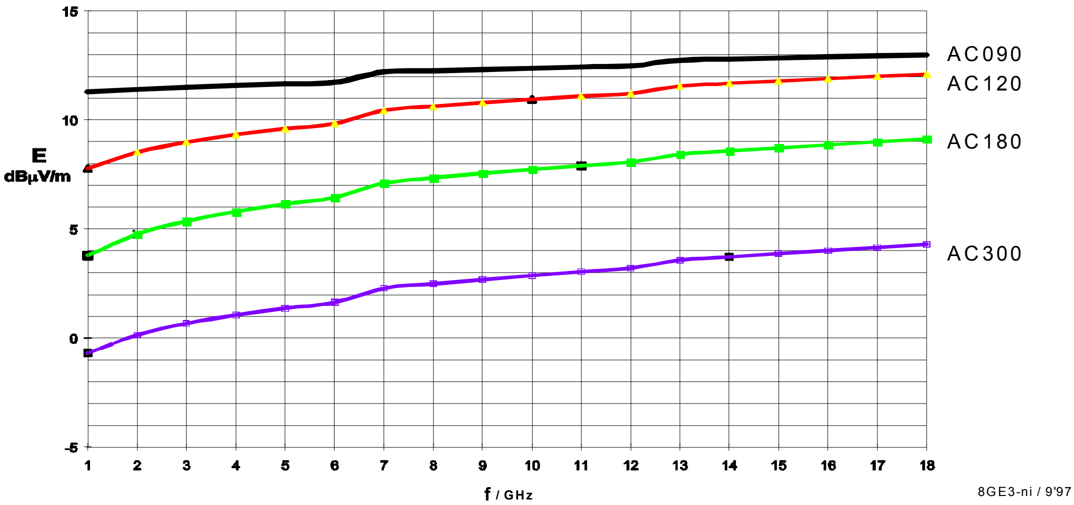


Fig. 4.34 Minimum receive field strengths for different reflector sizes (bandwidth 1 MHz; feed: crossed LP antenna from Fig. 4.30 right)

The 3 m R&S AC300 antenna makes it possible to receive signals which, depending on frequency, are 9 dB to 12 dB weaker - i.e. about an order of magnitude in terms of power - than the signals that can still be received using the 90 cm R&S AC090 antenna.

A natural restriction on increasing the gain is not due to antenna engineering, but rather to the fact that antennas with even higher directivity inevitably have ever smaller half-power beamwidths in their radiation patterns. They can then only be reliably aligned and held on their distant station at very great expense. Consequently any vibration such as that caused by the wind is capable of interrupting a microwave link. Therefore, the objective of optimization is not even higher gain, but the **broadest possible bandwidth**. LPDAs of the type shown in Fig. 4.31 are suitable as the feed for this purpose.

A complete microwave receiving antenna system with a 1.8 m reflector is shown in Fig. 4.35. Two smaller, optional reflector antennas for the ranges from 18 GHz to 26.5 GHz and 26.5 GHz to 40 GHz are added in order to extend the 1 GHz to 18 GHz frequency range of the main antenna.

Horn antennas (Fig. 4.36) are used mainly in the microwave range, not only as independent directional antennas (e.g. **test antennas** and **gain standards**, but also as feeds in reflector antennas. Horn antennas are normally fed by a waveguide. When they receive the conducted wave, they transform it into a sky wave via the flared cross-section of the horn. In the plane of the aperture, this generates an electromagnetic field which is more or less in phase. Since their dimensions usually have to be a multiple of the operating wavelength, the use of horn antennas is restricted to the **highest frequency ranges** for reasons of practicality. Like reflector antennas, horn antennas can also take a wide variety of forms. Depending on the shape of their aperture (radiation area) horn antennas can be described as rectangular, sectoral, pyramidal and conical. Further distinctions are based on the type of feed.

Fig. 4.35 Rotatable microwave antenna system for 1 GHz to 40 GHz

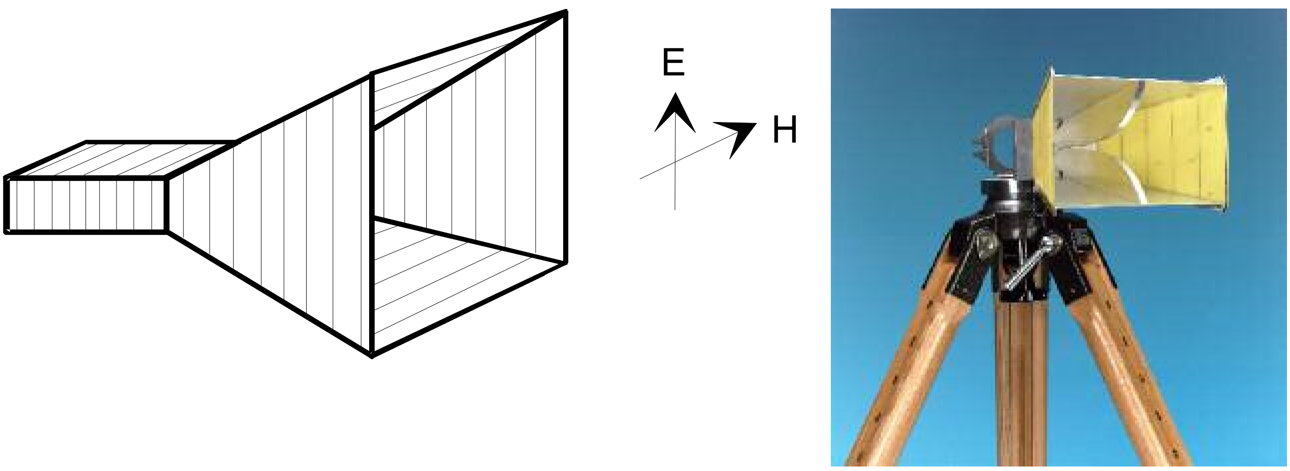


Fig. 4.36 Horn antenna with rectangular aperture

Fig. 4.37 Rotatable receiving antenna system for 10 GHz to 3 GHz
combining a large number of the antennas described here in a small amount of space

4.5 Pattern Shaping and Beam Steering

Section 4.4 already described that and how separate dipoles can be arranged with defined spacing and fed with predetermined phases in order to generate radiation patterns and thus configure directional antennas. The radiation pattern of such an antenna is of course invariable over time. In many applications, however, there is an advantage in being able to steer the radiation pattern and for instance swing the main lobe. This can be done by using one of the methods described below.

4.5.1 Electronic Beam Steering

If the outputs of the elements of an array antenna ahead of summation are provided with electronically variable attenuators and phase shifters (Fig. 4.38), highly flexible, nearly inertia-free pattern shaping is possible.

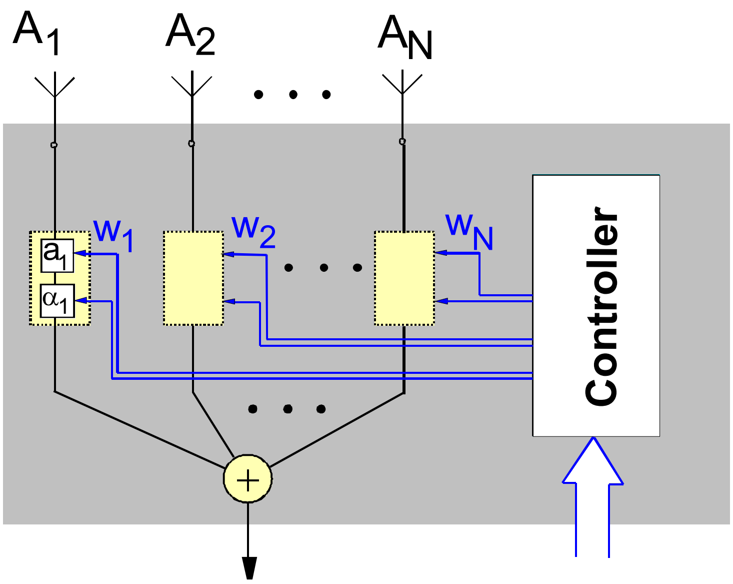


Fig. 4.38 Block diagram of a configuration for electronic pattern steering

If only the main lobe of the pattern is to swing in a particular direction, amplitude weighting with a_i is normally not required. The **phase-controlled antenna array** obtained in this way is known as a **phased array.**

Normally – i.e. if the range of swing is wide, or if there are stringent requirements with regard to the shape of the pattern during swinging, or there is simultaneous null steering - amplitude weighting is necessary. Amplitude and phase weighting can be combined in the complex weighting

$$w_i = a_i e^{j\alpha_i}$$

If an antenna array consists of n antennas, it is possible to satisfy n-1 of the requirements to be met by the radiation pattern that will be generated.

4.5.2 Digital Beam Steering

If the output signals of the antenna elements are not weighted in analog form as shown, but entirely by computation after digitizing the signals (Fig. 4.39), this is known as digital or even synthetic pattern shaping. With the appearance of increasingly powerful components for digital signal processing, this method of pattern shaping is also becoming increasingly important in radiomonitoring and radiolocation.

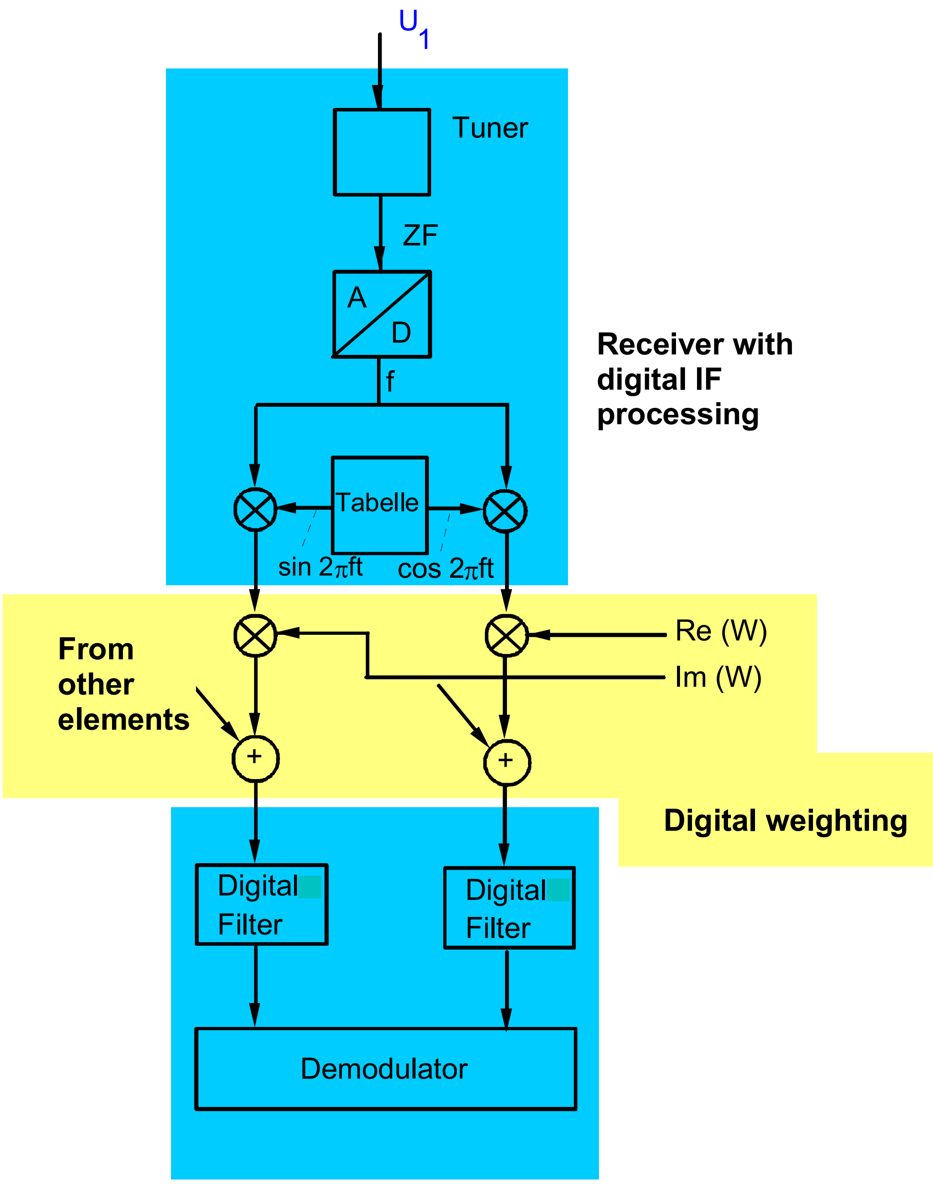


Fig. 4.39 Digital complex element weighting by signal processing in the baseband

4.5.3 Adaptive Antennas

Antennas are said to be adaptive when they comply with the following definition:

Definition:

Adaptive antennas

shape their **radiation pattern automatically**
in accordance with specified **criteria**.

The **objective** is to **maximize** the **signal-to-noise ratio**

The precondition for setting the weighting automatically is to expand the antenna setup in Fig. 4.38 by a control loop and to introduce boundary conditions that contain information on differences between useful and interfering signals (Fig. 4.40).

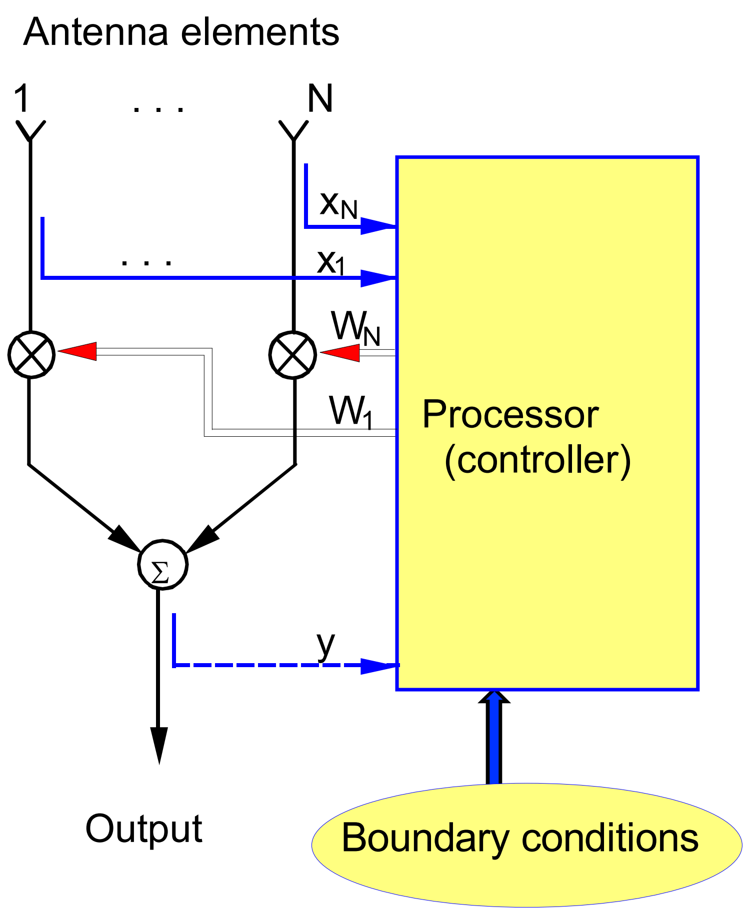


Fig. 4.40 Block diagram of an adaptive antenna

Fig. 4.41 shows how the radiation pattern could look after an antenna system with three elements, for example, has been adapted to a particular interference scenario.

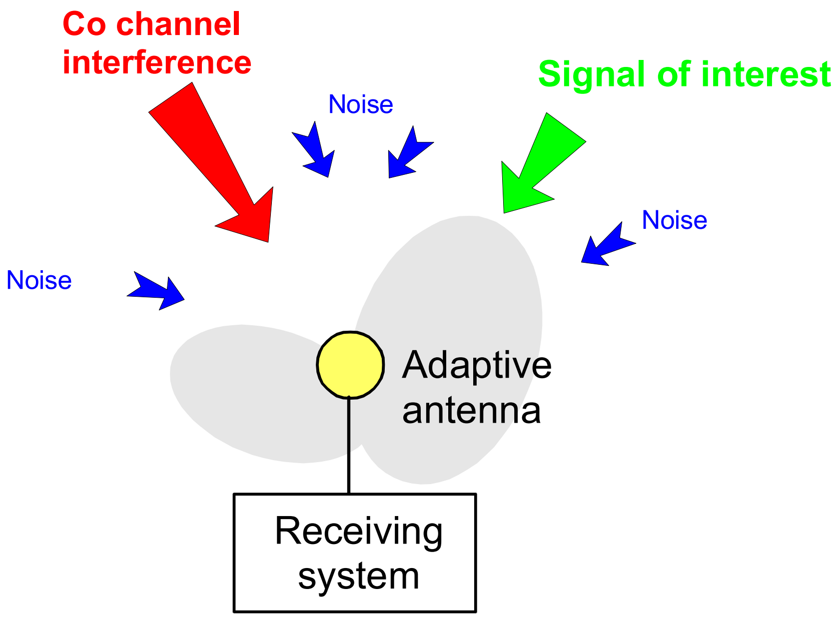


Fig. 4.41 Example showing the response of an adaptive antenna

As already mentioned in Section 4.5.1, the flexibility of pattern shaping mainly depends on the number of antenna elements N:

N elements yield a degree of freedom

K = N - 1

e.g. number of elements N = 3
makes it possible to independently set

2 nulls
or
1 null + 1 maximum

5 Test Antennas and their Application

5.1 Introduction to Field Strength Measurement

Anyone wishing to measure electric and magnetic field strengths, which is becoming an increasingly important activity in view of the ever-rising number of radio services and the fast-growing significance of EMI suppression, will have to deal in detail with measuring instruments. There is also an occasional tendency to overlook the fact that field strength measurement cannot be reliably performed without thorough knowledge of the test antenna.

The expression **useful field-strength** is generally used in connection with radio emissions and stands in contrast to the term **unwanted radiation**, which refers to the interference generated by items such as household appliances and motor vehicle ignition systems. The juxtaposition of all kinds of communications services using electromagnetic waves to transmit information requires rules to ensure that the various services do not cause mutual interference. However, radio waves do not stop at national boundaries, which means that their use also results in international problems. **Radiomonitoring** was therefore put on an international footing in the International Telecommunications Convention (Atlantic City 1947) since when there has been constant worldwide improvement.

Coverage measurements such as those performed on mobile radio or broadcasting networks also fall within the scope of useful field-strength measurement. Network operators are regularly required to generate a signal with adequate field strength in a specified area, and to verify this using measurements.

EMI measurements, however, has as its objective the detection and quantitative assessment of unintentionally emitted radiation. Basically every electrical or electronic device (vacuum cleaners and computers alike) is capable of emitting electromagnetic waves. The legislator therefore requires that electromagnetic fields generated in this way shall not exceed certain maximum values, and in most cases manufacturers or importers must produce evidence that this requirement is satisfied.

In view of the strong increase in wanted and unwanted emissions, it is all the more important to check devices and systems (including even entire aircraft) for their **immunity to external fields**. Field strength measurements are necessary in order to obtain quantitatively usable and comparable evidence.

The instruments needed to measure field strengths can be divided into four or, more accurately, five groups:

- **Antennas**, which transform the electromagnetic wave into a conducted wave (or vice versa) and whose electrical characteristics must be accurately known. A single antenna or an antenna array will be used, depending on the application.

- **Signal distribution**, which can be implemented by a simple RF cable. When there are several switchable antennas, signal distribution is sometimes very complex. It may include switchable relay matrixes, diode switches and electronic signal multipliers.

- **Receivers**, which cover the frequency range of interest. They are equipped with different demodulators depending on the intended purpose, and must comply with applicable requirements regarding selectivity and bandwidth. These requirements are often defined in relevant standards. The receiver <u>must</u> have a calibrated test instrument for the voltage at its antenna input and <u>should</u> have a computer interface so that both measurement and analysis of the measured data can be performed automatically.

- **Control units**(s), which automate measurement and result processing, and by so doing can significantly reduce personnel costs. Industry-compatible, advanced-technology computers or PCs are ideal for this purpose. Some test receivers are already equipped with the control software and interfaces necessary for handling data or can be upgraded with the appropriate options.

Combining the modules mentioned with a well-thought-out installation of system modules that provide **protection from lightning, electromagnetic interference** (EMI) and the like will make little or no difference to the convenience, speed and reliability of measurement.

Equipping and organizing correctly for field strength measurement can consequently incur very considerable outlay. For this reason, manufacturers at one time described their highly expensive instruments as modest field strength indicators in order to avoid the word measurement in this context.

5.2 Antenna Calibration

Section 3.6 makes clear that knowledge of one of the following parameters

- effective aperture
- gain
- effective length / effective height
- antenna factor

is indispensable when measuring field strength. A purely computer-based definition of the parameters for the antenna type being used (or such information taken from a catalog, data sheet or manual) is normally not enough to ensure the accuracy demanded for many measurement tasks. Test antennas are therefore calibrated by the manufacturer, by a specially accredited institute or by the user. In this way, each antenna is provided with an **individual report** which not only shows the antenna factor or gain (often as a function of frequency) but on occasions must also include a section on error investigations.

5.2.1 Calibration Methods

National and international standards institutions and telecommunications authorities have defined a series of methods for calibrating test antennas. The purpose of these is

- to work toward **standardization** of the terms and expressions used and to ensure reproducibility of the results obtained during calibration and measurement

- to take into account the **physical characteristics** of the frequency ranges and forms of antenna under investigation, and where possible to compensate any systematic errors that such characteristics may cause.

If the electric field strength is precisely known at the point where the antenna being calibrated is located, the antenna factor for this antenna can be obtained by evaluating the defining equation for the antenna factor (**standard field method**).

For a clearer understanding it should be remembered that the majority of terms defined in the context of radiation and wave propagation assume fully **idealized environmental conditions** (free space, matching and quite often far field). Therefore, a distinction first needs to be made between **two classes of test methods**:

- **Free space methods** require a test path that very closely simulates free space propagation conditions.

- In **reflection methods** the increase in field strength caused by reflection from a highly conducting surface is utilized for calibration.

Various methods are also used for the **evaluation of the measurement**:

Use of the **comparison method** assumes that a **reference antenna** is available so that its data can be compared with the antenna being calibrated.

In the **reciprocity method**, the required antenna data is determined from the **transmission loss** between the transmitting and receiving antennas. A further distinction needs to be made between the **two-antenna method** and the **three-antenna method**: The two-antenna method is based on two antennas that are as identical as possible, whereas the three-antenna method (for a correspondingly higher measurement outlay) is not affected by the same restriction.

The test methods just mentioned are summarized in the table below.

<u>Summary of commonly used calibration methods:</u>

	Free space method	Reflection method
	Standard field method	
Comparison method	Comparison method DIN 45003 Standard receiving antenna method NBS Report 5539	Reference antenna method ANSI C63.5
Reciprocity method (two identical antennas)	Reciprocity method to DIN 45003 ARP 958	Standard site method to ANSI C63.5
Reciprocity method (three antennas)	Reciprocity method to DIN 45003 and R&S company standard ARP 958 (3 antennas)	Standard site method to ANSI C63.5 (3 antennas)

5.2.2 Measurement Methods

Standard field method

The standard field method is certainly the most natural test method, because it can be derived directly from the defining equation of the antenna factor (see 3.6). Standard fields can be generated in otherwise free space as well as in shielded or unshielded spaces [DIN 45300, NBS Techn. Note 1309, IEEE Std 291-1991]. The antenna factor is determined from the measured antenna output voltage following an elementary calculation, **without taking into account the measurement errors inherent in the test instruments used for calibration**.

A decisive influence on the accuracy of this test method is the precision with which a field can be computed in advance at a specific point in space. This can be done successfully, for example, in the **magnetic near field** of a two-wire line or small loop antenna at low frequencies (below 30 MHz) for which the error limits are in the region of +/- 0.2 dB when working with great care. Fig. 5.1 shows an example of a setup for near field calibration of a loop antenna in the frequency range from 9 kHz to 30 MHz. The measurement distance is typically D = 0.5 m - 1.5 m and loop diameters are less than 2 r = 1 m.

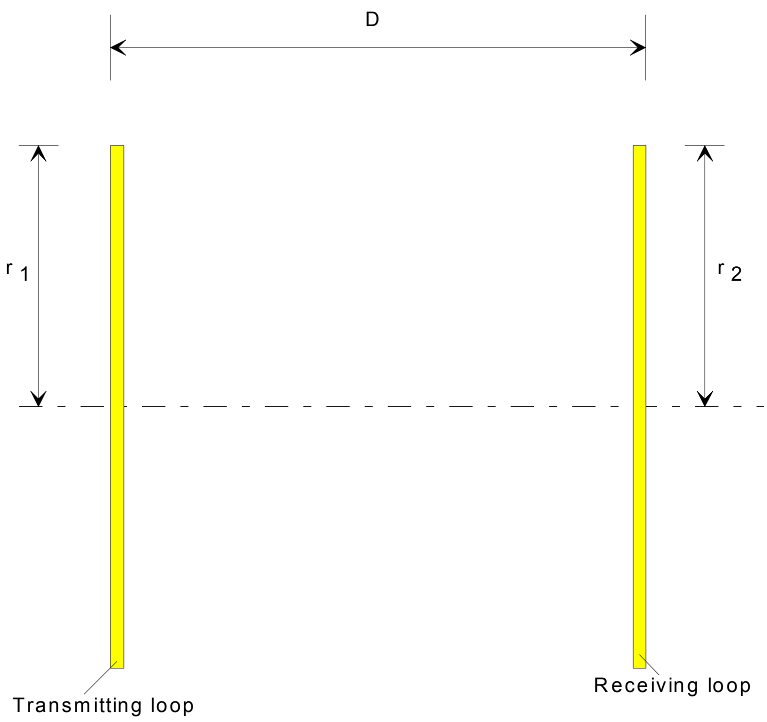


Fig. 5.1 Test setup for near field calibration of a loop antenna

With good approximation, the average value of the magnetic field strength on the surface extending from the coaxially attached receiving loop is

$$|H_{av}| = \frac{I_0 \cdot F_1}{2 \cdot \pi \cdot R^3} \sqrt{1 + k^2 R^2}$$

where I_0 current in transmitting loop

$F_1 = \pi \cdot r_1^2$ area of transmitting loop

$k = 2\pi/\lambda$

$R^2 = D^2 + r_1^2 + r_2^2$,

provided the condition

$$16 \cdot r_1 \cdot r_2 < R^2$$

is fulfilled. The quotient of the **magnetic field strength calculated** in accordance with the formula listed above and the **measured antenna output voltage** is the desired antenna factor. It should be noted that in the method described it is not - as defined in Section 3.6 - the electric field strength, but rather the magnetic field strength that is referenced to the output voltage. Since this does not agree with the strict definition of the antenna factor, the electric field strength derived from multiplying the above formula by the field characteristic impedance Z_0 is sometimes specified in the technical literature and in data sheets instead of the formula for the magnetic near field. To prevent any misunderstanding it must be made clear that the resulting expression serves only as a defining equation and has nothing to do with the electric near field strength. Most sources take this into account, using designations such as pseudo-E field or equivalent electric far field.

With today's state of the art, standard field measurement methods are used only in cases where no higher accuracy is going to be achieved by using other measurement methods. Such cases include the calibration of antennas in the frequency range below 30 MHz, whereas for higher frequencies the methods described below deliver much more accurate results. However, these methods are partly or totally unsuitable for low frequencies.

Free space method

At first it would seem obvious to use free space calibration methods because the antenna parameters that need to be determined are themselves free space parameters. This would require a **completely homogeneous total space** in which transmitting and receiving antennas would be a virtually **infinite distance** apart.

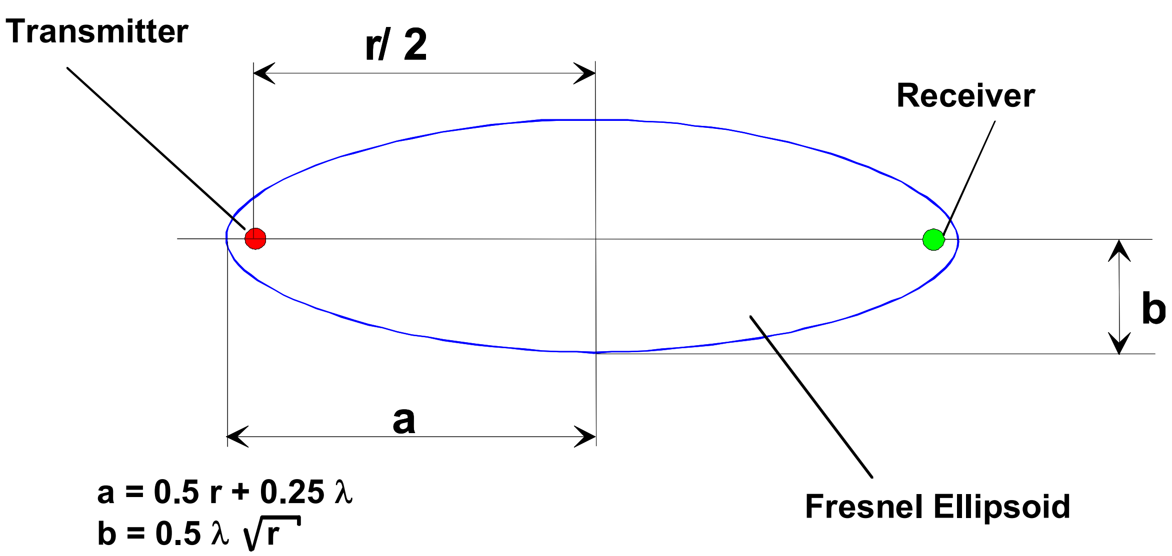


Fig. 5.2 Free space conditions

In practice the situation is limited to a propagation path having an **unobstructed Fresnel ellipsoid** (see Fig. 5.2). The extent to which such a configuration matches ideal free space conditions depends first and foremost on the frequency and the antenna radiation pattern:

- On the one hand, the dimensions of the ellipsoid to be kept unobstructed increase in proportion to the frequency,
- on the other, the antennas used in practice at higher frequencies also have larger dimensions.
- On the one hand, environmental influences that could affect the free space conditions become negligible as antenna directivity increases,
- on the other, high directivity complicates accurate antenna alignment and is more sensitive to the actual antenna being calibrated.

In practice, applying this method is mainly recommended for the **frequency range above 100 MHz**, and at the same time the use of high-directivity antennas (parabolic reflectors) is not advisable. With linear-polarizing antennas, **vertical polarization** is preferable to horizontal polarization in order to reduce ground effects.

Any deviation from the requirement for an infinite distance between transmitting and receiving antennas inevitably causes a phase error, as can be seen from Fig. 5.3; this error is due to the fact that a wavefront that is not fully plane cannot reach every point on a spatially extended antenna simultaneously.

$$r > \frac{\pi (d_1 + d_2)^2}{4\ \lambda\ \Delta\varphi}$$

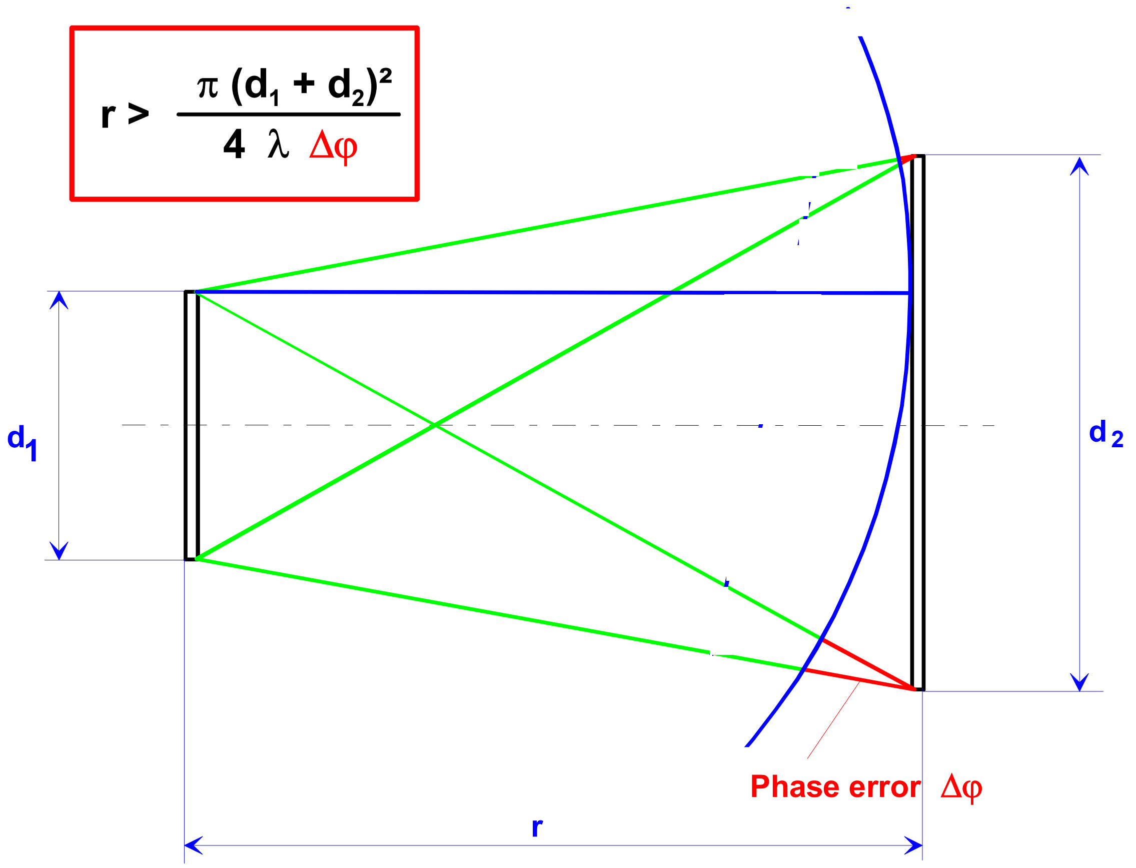


Fig. 5.3 Definition of the far field

In order to determine the measurement distance at which approximately far field conditions apply, particularly in the case of antenna structures that are large in relation to the wavelength, the equation in Fig. 5.3 can be used after a value has been specified for the maximum permissible phase error. This value is commonly defined as $\Delta\phi$ = 11.25°, but this is by no means mandatory. Far field conditions are definitely breached if the measurement distance is less than one wavelength.

When the free space method is used, the comparatively **high degree of accuracy** resulting from the fact that the implementation of the far field conditions and the accuracy of the mutual alignment of the antennas are the only system-related sources of error appears to be an **advantage**. However, the same set of circumstances proves to be a **disadvantage at low frequencies**.

Fig. 5.4 shows a typical free-space calibration setup used, for example, for **calibrating log-periodic dipole antennas** at frequencies between 50 MHz and several GHz. When analyzing the measurement results, it should be observed that the **position of the phase center is frequency-dependent in LPDAs** (see Chapter 4) and the distance R shown in Fig. 5.4 is not identical to the value to be used in the equation for calculating the receive field strength.

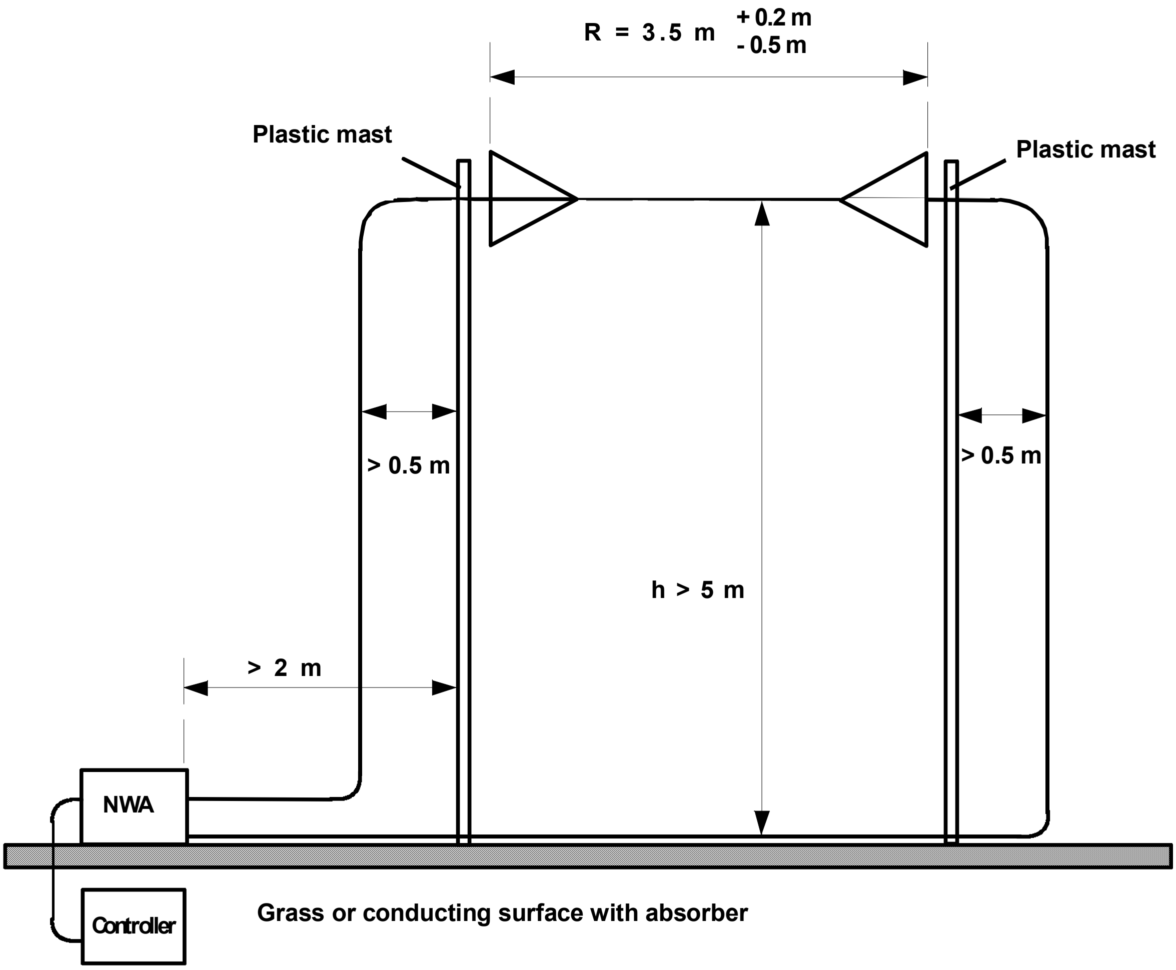


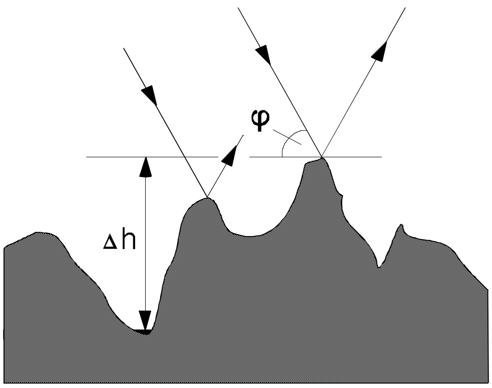
Fig. 5.4 Test setup for free space calibration

Reflection method

If there is no way of reliably guaranteeing that free-space propagation conditions will be maintained, these conditions are deliberately breached by **introducing a highly conducting surface**. Since the field strength that can theoretically be received at the reception site when using this type of configuration is fairly easy to calculate, conclusions about the antenna data are possible on the basis of the measured transmission loss.

When calibrating in line with the reflection method it is very important to ensure that

- the field strength maximum is actually used by making an appropriate **height adjustment** to the antenna being calibrated
- the reflecting surface is either **very highly conducting** or $|R_h| = 1$ is not valid (see Fig. 2.4 and associated text)
- the reflecting surface is **smooth enough**; this can generally be assumed to be so if the Rayleigh criterion is fulfilled (see Fig. 5.5)
- the reflecting surface is **large enough** and reflections of currents induced in the surface at its boundary are reliably avoided or suppressed by suitable measures (slots, notches, etc)
- the test path and its surroundings are **free from other reflectors and obstacles**.



$$\Delta h < \frac{\lambda}{16 \sin \varphi}$$

Fig. 5.5 Definition of smooth surfaces

The **advantage** that the described method can also be used **at low frequencies** and **in a relatively small space** is confronted with the **disadvantages** of a **great many sources of error** (reflection behavior, resonance suppression, antenna directivity) and **increased outlay** compared with the free space measurement.

It seems advisable in this case to use **horizontal polarization**, which

- makes it simpler to calculate the field concerned
- ensures reduced coupling between antennas and cables, as well as a much more strongly attenuated ground wave
- makes it possible to ignore scattering at the antenna cable.

It is an important requirement (and a further argument in favor of preferring horizontal polarization over vertical polarization) that the antennas to be calibrated by this method have **omnidirectional behavior in the vertical**, as is the case for instance with horizontally aligned dipoles. Otherwise, the field calculation becomes very complicated, since in that case the directly incident wave and the reflected wave

- are radiated with different rms power and
- receive different weighting on reception.

If this cannot be avoided in some cases, it is frequently possible to subsequently take into account the vertical radiation pattern using appropriate correction values. However, this complicates the measurement considerably and also is a further potential source of error.

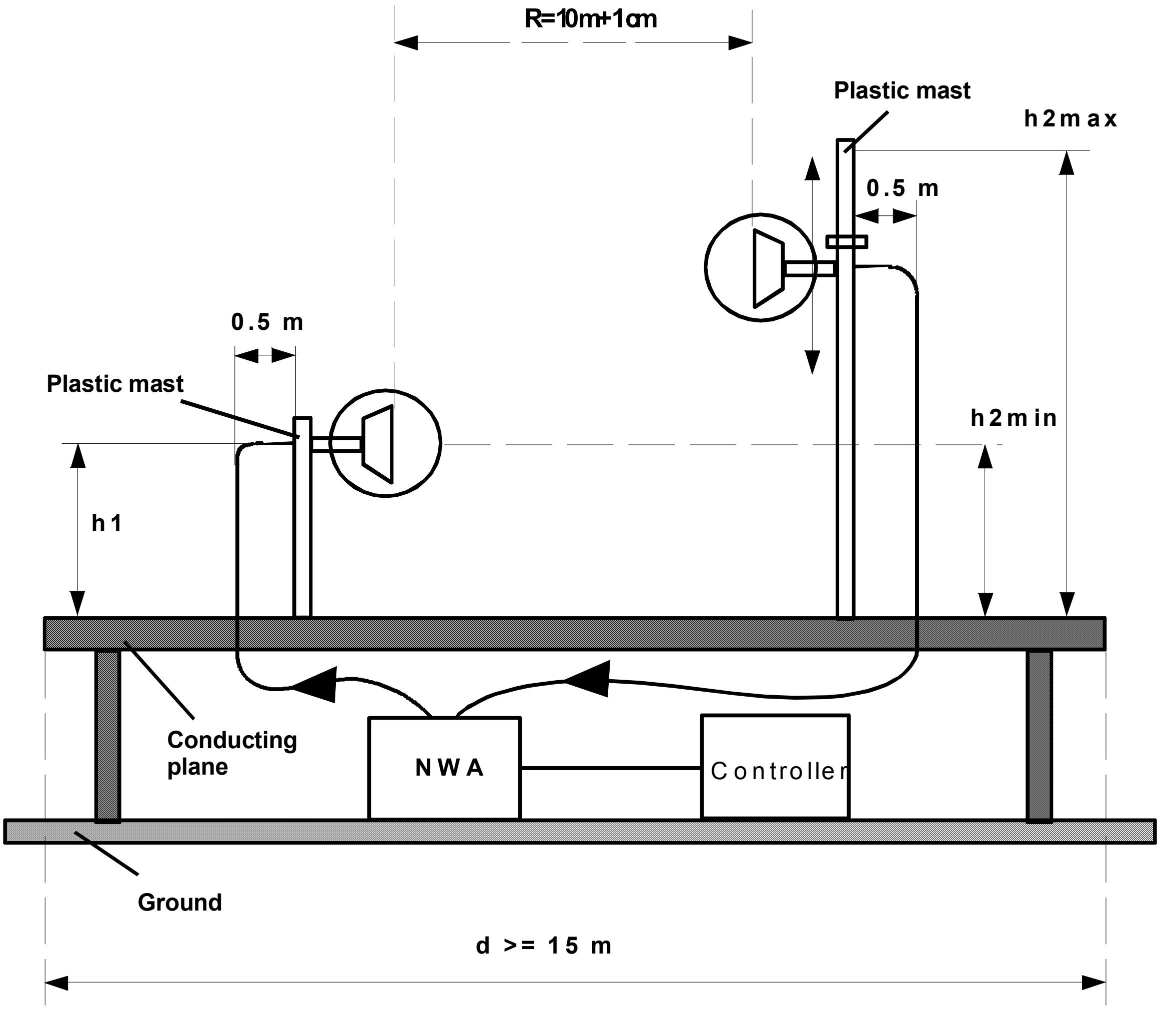


Fig. 5.6   Test setup for reflection path calibration

The geometry for reflection path calibration of biconical dipoles in the 20 MHz to 300 MHz frequency range is shown in Fig. 5.6, which also shows the minimum length of the conducting surface for this method. The lateral dimension can be considerably smaller (a few meters). The maximum adjustable height of the receiving antenna primarily depends on the available mast, but should be at least 4 m. It is recommended that the chosen minimum height and the transmitting antenna height are adequate (h > 1.5 meters). This will ensure that coupling of the antennas with the conducting surface is prevented as far as possible. However, some standards specify correction values for antennas that are set lower.

5.2.3 Evaluation Methods

Whatever the chosen test setup, the next question is the method to be used to analyze and evaluate the measurement data. There are basically two different methods available for this purpose. The method about to be described, the comparison method, is gradually losing ground in the area of precision measurement.

Comparison method

Antenna comparison methods can be used for measuring both the practical gain and the quantitative radiation pattern (in free space measurement), provided a **calibrated reference antenna** is available. Both antennas are introduced one after the other into a constant electromagnetic field. When doing so it is particularly important to ensure that

- the **field strength** at the reception site **does not alter** when the antennas are changed
- both antennas are placed at **precisely the same spot** and with the **same alignment**.

A comparison of the output voltages U_e and U_x for both antennas will then yield the required **antenna factor**

$$K_x = K_e \frac{U_e}{U_x}$$

where the index e relates to the reference antenna and the index x to the antenna requiring calibration. It is not necessary to know the magnitude of the electric field strength. However, its relative value must be **continuously monitored** in order to ensure compliance with the requirements mentioned above.

The **accuracy of the available reference antennas** is a limiting factor in the overall precision of the method. For instance the NBS (US National Bureau of Standards) specifies a tolerance of +/- 1 dB for reference antennas, and naturally this Fig. has to be added to other measurement tolerances which arise in any case. Therefore measurements in which the total error limit must not exceed +/- 1 dB cannot be performed by using the comparison method.

Reciprocity methods

Unlike the comparison method, the following test methods do not need a reference antenna. Antenna factors or gain are instead obtained from a **comparison between measured and calculated field parameters**.

Thus in the case of a test path complying with **free space conditions**,
the relationship

$$P_e = P_s \frac{\lambda^2}{16\,\pi^2\,r^2} G_s\,G_e$$

applies to the ratio of received to radiated power P_e/P_s at wavelength λ and antenna spacing r when the transmitting and receiving antennas have gains of G_s and G_e respectively. If the **power ratio**

$$P_e/P_s = |S_{21}| \dashrightarrow 10 \lg |S_{21}| = s_{21}$$

is then measured by means of a **scattering parameter measurement** with the aid of a network analyzer, the sum of the logarithmic antenna gains is obtained from the measured parameter and the **basic free-space transmission factor**

$$L_0 = 10 \lg \frac{P_s}{P_e}$$

in the form of the expression

$$g_e + g_s = s_{21} + L_0 .$$

This can be used not only to determine the **geometric mean of the gains of both antennas**, but also

- if the gain of one of the two antennas is known or
- if the antennas are identical,

to then determine their **individual gains** g. From this information the required **antenna factor**

$$k = - 29.8 \text{ dB} + 20 \lg (f/\text{MHz}) - g$$

can be calculated. It should however be pointed out again that all the gain values mentioned here are **practical gains**. They are likely to include losses due to matching and may need to be converted into ITG standard gain. It goes without saying that **before actually starting the measurement, the cable loss** must be determined and taken into account when evaluating the results.

The **reflection test path** can be calibrated in the same way. In this case, care must be taken to ensure that the field strength present at the reception site can assume the maximum value E_{max} defined by the equation shown in Fig. 2.4, and that by appropriate height variation the receiving antenna is also positioned at **field strength maximum**. The measured ratio U_S/U_e between the voltages indicated on the transmitter output and the receiver input is the **field attenuation**

$$A = \frac{\sqrt{30\ \Omega\ P_s\ G_s}\ 79{,}6\ 10^6\ K_s\ K_e}{2\ f\ E_{max}}\ \text{m/s}$$

and from its logarithmic form a = 20 lg A together with the **field strength level**

$$F_{max} = 20 \lg \frac{E_{max}}{1\ \mu\text{V/m}}$$

the **sum of the antenna factors**

$$k_s + k_e = 20 \lg f/\text{MHz} + \text{Fmax} + a - 48.9 \text{ dB}$$

can be calculated.

It is obvious that the **accuracy** of the method described depends on the **calculated field parameters**. Free-space or reflection-path conditions must be adhered to with care, as must the far field condition.

It is quite permissible, and also recommended in the case of reflection test paths, to investigate a test path for **increasing reliability** with regard to factors that conflict with these requirements and if necessary to compensate for these factors by means of correction values, in other words **to calibrate even the test site itself from time to time**. For this purpose, the field attenuation A specified above can be used to derive the **normalized site attenuation** (NSA)

$$A_0 = \frac{A}{K_s\, K_e} \quad \text{or} \quad a_0 = 20 \lg A_0$$

which can then be stored for subsequent comparison. The NSA may be thought of as a single, antenna-independent parameter describing the behavior of the test site.

The fact that the method described is able to yield **purely and simply the geometric mean** of two gains can be avoided by applying the **three-antenna method**. In this method, just about **any three antennas** are cyclically interchanged and measured against one another. This method, which in this example relates to the free space test path, produces a set of equations with three unknown gains g:

Measurement:	Produces equation:
1 - 2 --> s_{21}'	$g_1 + g_2 = s_{21}' + L_0$
2 - 3 --> s_{21}''	$g_2 + g_3 = s_{21}'' + L_0$
3 - 1 --> s_{21}'''	$g_3 + g_1 = s_{21}''' + L_0$

Provided that the basic transmission factor L_0 during the measurements is constant (and therefore all measurements are performed in the exactly identical configuration), the set of equations can be solved and produces the **individual gain values**

$$g_1 = 1/2 * (s_{21}' - s_{21}'' + s_{21}''' + L_0)$$

$$g_2 = 1/2 * (s_{21}'' - s_{21}''' + s_{21}' + L_0)$$

$$g_3 = 1/2 * (s_{21}''' - s_{21}' + s_{21}'' + L_0).$$

Therefore, the three-antenna method involves a **higher measurement outlay** in comparison with the reciprocity method, but allows **simultaneous calibration of three different antennas**. Moreover **no further requirements** such as similarity of type, calibration in advance, etc, must be fulfilled.

The **calibration of active antennas** is a special case. These cannot be used to transmit and therefore cannot be cyclically interchanged in the way envisaged in the three-antenna method. A common solution is to interchange the antennas once anticyclically and then to take any further possible resulting errors into account. As an alternative it would of course be possible to consider the comparison method.

However, active antennas (particularly rod antennas, for which the methods described here cannot be used at all or only with difficulty) are frequently calibrated by **measuring the transfer function of their active component**. For this purpose (see Fig. 5.7) the passive antenna part is simulated by an equivalent circuit of the same frequency response (a capacitor is often enough) and a signal from a generator connected at the correct impedance is then applied to it. Due to the high input impedance this turns out to be the open-circuit voltage $U_0 = U_G/2$ which would generate an electromagnetic wave at the dipole feedpoint. If the voltage U_E at input impedance $R_E = 50\ \Omega$ is then measured at the antenna output (e.g. using a test receiver), the antenna factor can be determined from this voltage ratio and the computed effective length:

$$K = \frac{U_0}{U_E} \cdot \frac{1}{l_{eff}}$$

Particularly in the case of short passive dipoles, the effective length can be determined from the total mechanical length:

$$l_{eff} = l_{tot}/2$$

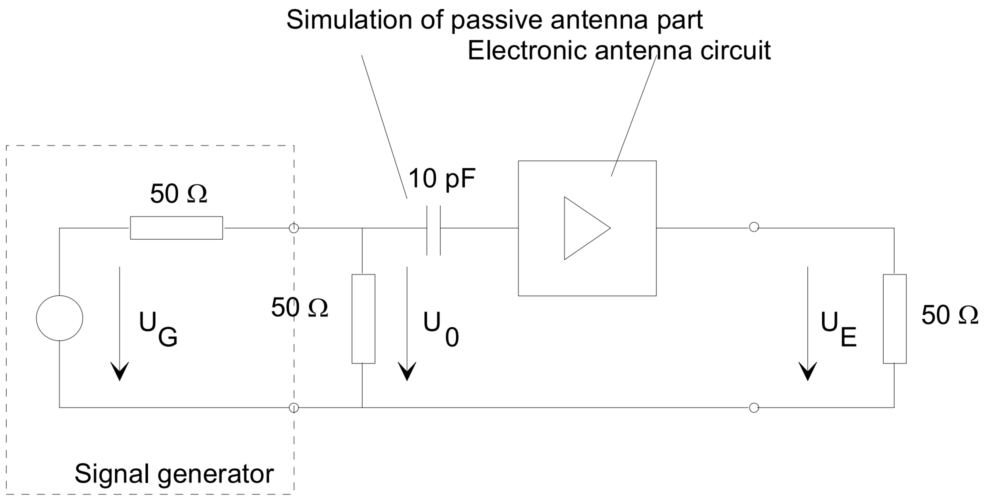
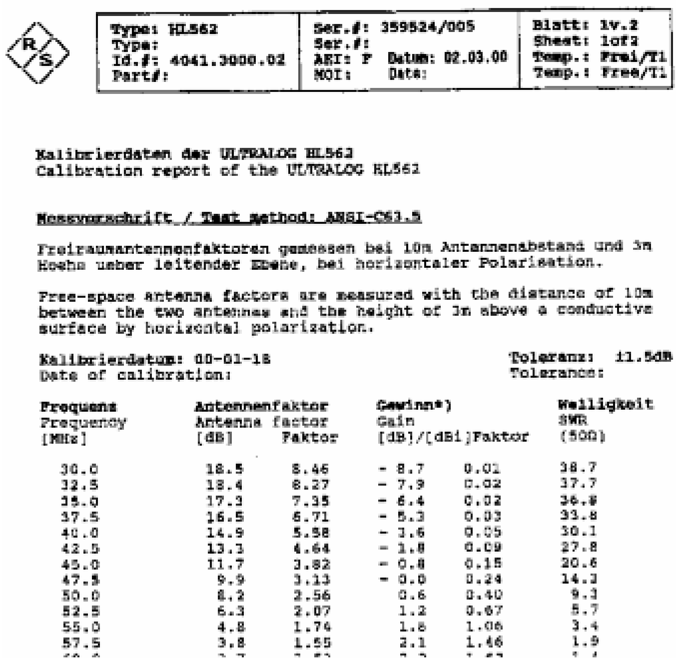


Fig. 5.7 Test setup for calibrating active antennas

In conclusion, Fig. 5.8 shows an example of a calibration report (specifying the date and serial number as well as the calibration guideline used). Such a report should accompany every antenna described by its supplier as a **test antenna**:



Type: HL562 Type: Id.#: 4041.3000.02 Part#:	Ser.#: 359524/005 Ser.#: ARI: P Datum: 02.03.00 MOI: Date:	Blatt: 1v.2 Sheet: 1of2 Temp.: Frei/T1 Temp.: Free/T1

Kalibrierdaten der ULTRALOG HL562
Calibration report of the ULTRALOG HL562

Messvorschrift / Test method: ANSI-C63.5

Freiraumantennenfaktoren gemessen bei 10m Antennenabstand und 3m Hoehe ueber leitender Ebene, bei horizontaler Polarisation.

Free-space antenna factors are measured with the distance of 10m between the two antennas and the height of 3m above a conductive surface by horizontal polarization.

Kalibrierdatum: 00-01-18 Toleranz: ±1.5dB
Date of calibration: Tolerance:

Frequenz Frequency [MHz]	Antennenfaktor Antenna factor [dB]	 Faktor	Gewinn*) Gain [dB]/[dBi]	 Faktor	Welligkeit SWR (50Ω)
30.0	18.5	8.46	- 8.7	0.01	38.7
32.5	18.4	8.27	- 7.9	0.02	37.7
35.0	17.3	7.35	- 6.4	0.02	36.8
37.5	16.5	6.71	- 5.3	0.03	33.8
40.0	14.9	5.58	- 3.6	0.05	30.1
42.5	13.3	4.64	- 1.8	0.09	27.8
45.0	11.7	3.82	- 0.8	0.15	20.6
47.5	9.9	3.13	- 0.0	0.24	14.3
50.0	8.2	2.56	0.6	0.40	9.3
52.5	6.3	2.07	1.2	0.67	5.7
55.0	4.8	1.74	1.8	1.06	3.4
57.5	3.8	1.55	2.1	1.46	1.9

Fig. 5.8 Example of a calibration report

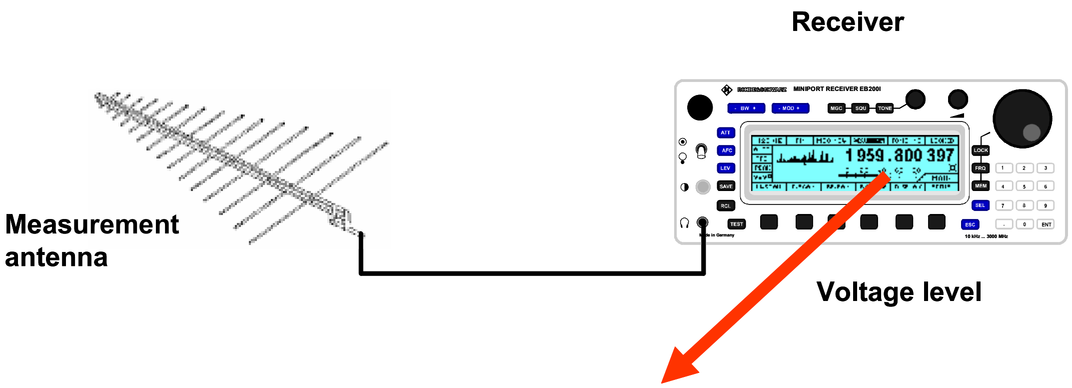
5.3 Field Strength Measurement

A field strength measurement (see 5.1) always requires (at least)

1 **test receiver** and
1 **test antenna** (as well as a connecting cable between the antenna and the receiver)

(Fig. 5.9). Computer-aided measurement and analysis has simplified the practical work and is therefore increasingly accepted. This should, however, not lead to the mistaken conclusion that it also increases the precision of the measurement. The degree of precision depends solely on the quality of the receiver and the antenna as well as on the care exercised in performing the measurement.

The **test receiver** must obviously be suitable for the required frequency range and should be calibrated regularly.



$$F\ [\mathrm{dB}(1\ \mu V/m)] = P_u\ [\mathrm{dB}(1\ \mu V)] + k\ [\mathrm{dB}(1/m)]$$

Field strength level **Antenna factor**

Fig. 5.9 Test setup for field strength measurements

Any type of antenna that is suitable for the frequency range is suitable as a **test antenna** (and therefore as a transformer between sky waves and conducted waves). However, whereas communications technology first and foremost strives to transform one wave type into another with as little loss as possible, test antennas have the task of providing a downstream test receiver with an **accurate measurement of the field strength at the installation site**. In fact, a receiver is nothing but a frequency-selective voltmeter and measures nothing but voltages (irrespective of what might be displayed). It is clear therefore that every antenna can also be used as a test antenna provided that its characteristics are known with sufficient accuracy, and in particular the **antenna factor for calculating the field strength from the voltage value indicated by the receiver**. This antenna factor can be specified generally for each antenna type; its value for a tuned and matched half-wave dipole is $K = 2\ \pi / \lambda$, for example. If the computed antenna factor is not accurate enough for field strength measurements (which is regularly the case with RFI field strength measurement), it is measured individually for each test antenna (antenna calibration). In contrast, with antennas used for communications purposes and for useful field-strength measurements it is permissible to specify the antenna factor purely on the basis of type. The main test antennas are half-wave and broadband dipoles (biconical antennas), active dipoles and monopoles, log-periodic antennas or loop, spiral and horn antennas.

For the sake of completeness it should be mentioned that for a precision field strength measurement the **cable loss** between the test antenna and the receiver has to be included.

As a rule, the antenna factor of a test antenna is supplied by the manufacturer in the form of an individual table for each unit. It can often be entered in the test receiver (or the analysis software), and the receive field strength can then be read straight off and analyzed without the need for tiresome recalculations.

The frequently observed practice of simply measuring the input voltage or input power of a receiver without any knowledge of the antenna data and cable data can at least provide evidence as to whether or not any connection can be established at all. However, such a measurement is not a field strength measurement, but rather a **coverage measurement**.

Especially in mobile radio, the problem often is that coverage and field strength have to be measured in a large area. Measurement therefore needs to be performed in a **moving vehicle** (Fig. 5.10), and this poses two problems:

1. The **antenna factor of** the antenna used for measurement must be determined.
2. The **position of the vehicle** must be defined and documented as accurately as possible.

Determining the antenna factor of an antenna attached to the metal roof of a vehicle is by no means trivial. Monopoles (rod antennas) l/4 in length are normally used for this purpose. These monopoles typically have a gain of around 5 dB and are omnidirectional in the horizontal plane when positioned on an infinite, ideally conducting surface. In practice however, radiation patterns of vehicle antennas exhibit **uncircularities** amounting to between a few and many dB, and this needs to be taken into account when evaluating measurements. An occasional solution is therefore to assume that the average gain of such a configuration is approximately 2 dB. However it is also permissible to place the whole vehicle on a turntable in an anechoic chamber and take accurate measurements (i.e. calibrate). Such measurements also yield the aforementioned average gain, for example.

Fig. 5.10 Automated field strength measurement using a test vehicle

Three principal methods can be used for **determining and documenting the vehicle's position** during a measurement:

- The distance traveled is measured by means of a pulse generator attached to the vehicle
- Position is continuously determined with the aid of a travel pilot installed in the vehicle. This involves calculating the geographic coordinates from the vehicle's speed and its movements around its spatial axes
- Position is continuously determined with the aid of a global positioning system (GPS) satellite navigation system

and subsequently transferred by computer onto a map (Fig. 5.11).

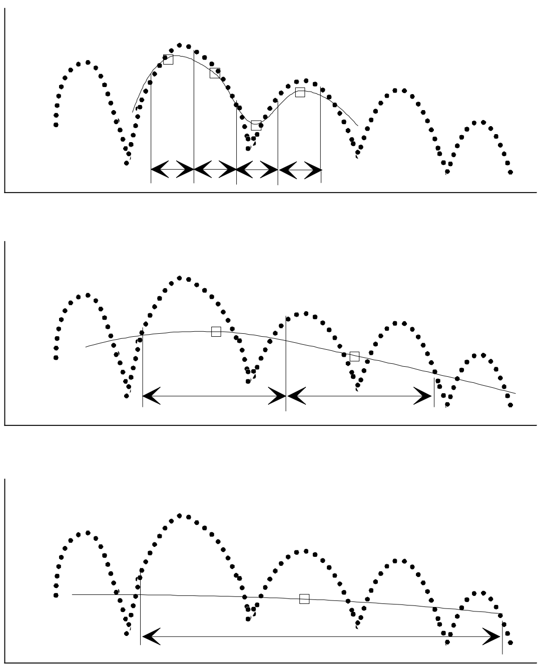
Fig. 5.11 Digitized map of Augsburg, Germany, showing mobile radio base stations and test results (by courtesy of E-Plus)

It is quite normal and useful to introduce the results of field strength measurements in the form of attenuation factors for particular regions, locations, landscape features etc into the wave propagation model used for planning, and thus add self-measured values to land use factors (clutter factors, see Chapter 3) taken from tables, if special regional features make this advisable. This method is often known as calibrating the model, and must not be confused with the method described above for determining the antenna factor.

With computer-aided measurement data acquisition and storage, the data files created during propagation or coverage measurements can quickly reach a colossal size. Furthermore, the purpose of a test drive is seldom to investigate the complicated and largely unreproducible fine structure of the local field strength distribution, which as a rule is purely statistical in nature. On the other hand it does make sense to reduce the measured values by **averaging** so that

- repeating a measurement will produce similar results (in statistical terms: scattering stays as low as possible)
- significant information such as environmentally-influenced field strength distribution is, however, not eliminated by averaging.

The choice of a suitable averaging interval is therefore particularly important, as Fig. 5.12 makes clear.



Averaging length too short:
Mean value scattering too widely

Averaging length correct:
Long term fading structure
still recognizable

Averaging length too long:
Long term fading structure
no longer recognizable

Fig. 5.12 Averaging field strength values

Various reliable mathematical methods are available for averaging and for determining useful measurement intervals, but they are not discussed here. Instead the technical literature should be consulted for more detailed information.

6 Measurements on Antennas

It would be possible but by no means a trivial task to **measure the complete range** of antenna data presented in Chapter 3, which in itself describes only a section of the entire spectrum of conceivable antenna characteristics. To determine, for instance, a radiation pattern requires not only a rotator but also a signal generator and test receiver, or better still a complete test setup in order to ensure the necessary synchronism between transmitter and receiver as well as convenient acquisition of measurement data. Gain and antenna factor measurements are regulated by a multitude of national and international standards which must be strictly applied, especially when the calibration of test antennas is concerned. Research institutes, universities, antenna manufacturers and similar organizations must therefore invest **considerable resources** if they wish to correctly measure the specifications of an antenna.

Yet often it is by no means necessary to check every characteristic of an antenna. In particular when the data of an antenna type or individual antenna has already been measured and judged to be satisfactory, a **functional check** is often all that is needed to make sure that the antenna is operating as planned.

One possible way of carrying out a functional check would be to connect a transmitting antenna to the associated transmitter and use a receiver to check whether the transmission can be received. Obviously such a method can be both costly and inaccurate. It is better to check the antenna's **optimum matching**. This makes it possible to obtain information about whether the total energy produced by the transmitter is being fed to the antenna and then radiated by the antenna rather than being reflected back to the transmitter. With suitable test methods it will also be possible to localize any mismatching should the need arise. If more advanced measures are needed, the operator of the antenna will normally contact the antenna supplier, whose test facilities are usually more comprehensive and advanced.

6.1 Standing Wave Ratio and Parameters Used

The quickest and easiest way to check the matching of an antenna and its ability to function is to **determine its input impedance or its voltage standing wave ratio**. If the input impedance deviates from the specified value (in practically all cases 50 Ω), the voltage standing wave ratio (also called the standing wave ratio) increases to values greater than 1.0 and power is reflected back toward the transmitter (see Section 3.7). SWR values of less than 1.0 are physically impossible, as is shown in Section 3.7 (perpetuum mobile), and the value $s \equiv 1.0$ represents only the theoretical case of ideal matching. Commonly available antennas have SWRs between $s = 1.2$ and $s = 1.5$ depending on requirements.

Determining the input impedance of an antenna makes it possible to calculate the standing wave ratio, whereas measuring the SWR provides only limited information about the impedance ratio. Nevertheless in practice this limitation is accepted, especially as in most cases the effort involved is much lower. As explained in Section 3.7, it is enough to measure just one of the three parameters voltage standing wave ratio, reflection coefficient or return loss, since the other missing values can be calculated from it (see also the conversion table).

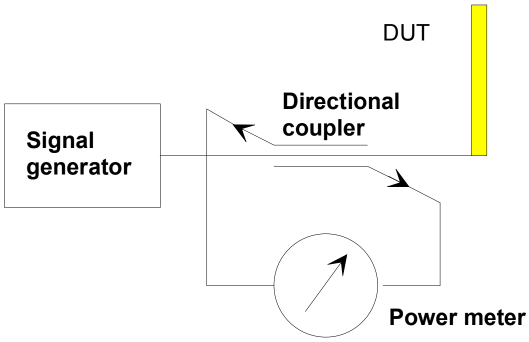


Fig. 6.1 Measuring the standing wave ratio

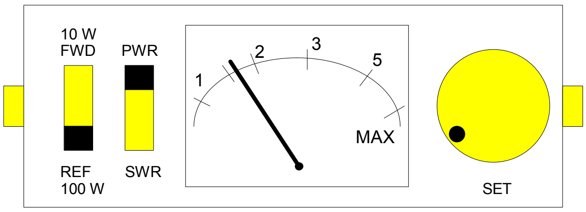


Fig. 6.2 Control panel of a simple SWR bridge

The **basic method for measuring a standing wave ratio** is indicated by the defining equation specified in Section 3.7: The quantities that have to be determined are the power of the wave fed to the antenna (device under test or DUT) and the power of the wave reflected from it. Fig. 6.1 shows a typical test setup in which a signal generator feeds a signal into the transmitting antenna. Directional couplers tap part of the power being routed in both directions and feed it to a suitable measuring instrument. The voltage standing wave ratio can then be determined from a comparison of the two power values.

It is obviously **important** that the **antenna is kept completely clear during SWR measurement** and that there are **no obstacles of any kind in the beam direction**; otherwise waves reflected from objects in the vicinity and radiated back into the antenna can fundamentally corrupt the result of the test. If necessary the measurement must be performed in an **anechoic chamber**.

The **measuring instruments that are permissible** for a measurement of this kind are distinguished by

the physical parameter they measure (voltage or power)
their user-friendliness
how they display the results
their measurement accuracy.

In addition, the **signal generator** must provide **constant frequency and level values**.

In most case, **voltage** is measured rather than **power**. Whatever the case, the results display can be calibrated either in units of power or in s values. Otherwise it is necessary (and does not involve special effort) to calculate the SWR from the voltage ratio.

User-friendliness can first and foremost be measured by whether it is necessary to switch over or even change the instrument in order to measure the forward and reflected wave. If the requirements for measurement accuracy are not particularly high, measuring instruments such as the one in Fig. 6.2 are available at a very low price. It can measure the power (switch position PWR) in two ranges (10 W and 100 W) and - with the aid of a built-in directional coupler - the standing wave ratio (switch position SWR). With the switch in the FWD position, the rotary knob on the right is used to adjust the pointer on the instrument to the MAX mark. By switching over to REF it is possible to read off the standing wave ratio straight from the scale. Power meters that fulfill stricter requirements for precision are much more expensive and as a rule allow a wider choice of display formats (power, voltage, SWR, etc). Cross-pointer instruments which cover three scales with the aid of two pointers are also quite common. The pointers have a scale each for the power of the forward wave and the reflected wave, and their point of intersection on a third scale indicates the SWR. Tedious switching over or even changing instruments is thus avoided, and all parameters of interest are displayed simultaneously.

Although the previously discussed test methods are entirely sufficient for many practical applications with regard to precision and user-friendliness, they do not provide two properties that are virtually indispensable in professional practice:

- A roundabout process with notepad and pen is just about the only way the test results can be processed any further.

- Displaying the result as a frequency response is a cumbersome manual task (for example subsequent plotting of a curve).

If the intention is to optimize an antenna, the effect of the measures taken can be observed at a particular frequency, but not over the entire useful frequency band.

Modern **sweep display units** and **network analyzers** avoid this disadvantage. They **cover a user-definable frequency range** and display the **result graphically** on a screen. If necessary, the result can be saved, plotted or processed with the aid of a PC. It is possible to measure (or rather display) power and/or voltage levels, input impedances in various formats (Cartesian, Smith chart), voltage standing wave ratio, reflection coefficient, return loss and further parameters. It is also often possible to perform measurements by means of reflectometry (see Section 6.2).

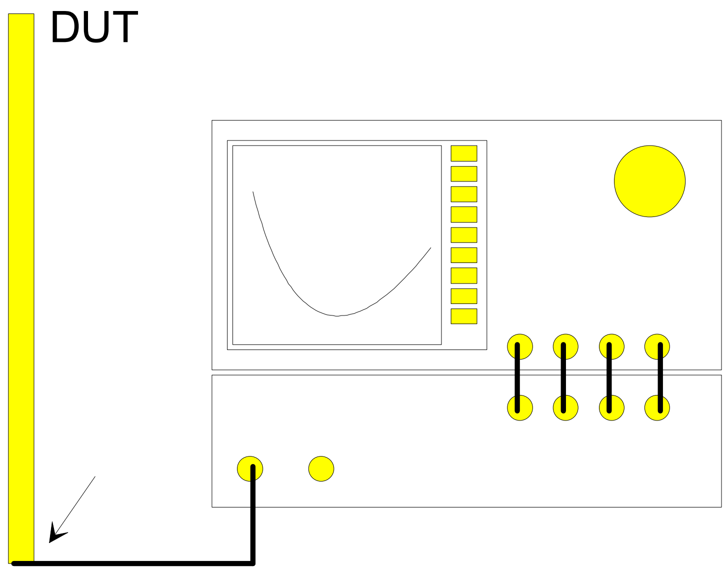


Fig. 6.3 Test setup with network analyzer (diagram)

Operating a modern network analyzer is by no means a trivial exercise. Since the procedures are not the same for every manufacturer, it would be impossible to give an all-embracing introduction here. Most manufacturers provide a detailed **manual** complete with applications and examples. **Studying** this manual is of paramount importance. Instruments of this kind are usually programmable and in some cases can be remote-controlled via PC or will operate automatically. However, the ease of operation and the convenient evaluation of measured data will **not** necessarily ensure precise measurement. As with any kind of measuring instrument, the achievable **measurement accuracy** is shown in the documentation supplied by the manufacturer and does **not** depend on the number of decimal places in the output.

Before starting a measurement the network analyzer (NWA) must be **calibrated**, which goes hand-in-hand with the decision on **what exactly** is going to be measured:

* Calibrating on the **input connectors** (see Fig. 6.3) produces measurement results for the **complete setup consisting of coaxial cable and antenna**. The cable loss is therefore included in the measurement result (the measured standing wave ratio is lower than that of the antenna alone); the antenna impedance cannot be determined in this way at all.

* The result produced by calibrating on the **antenna feedpoint** (see arrow in Fig. 6.3) represents the specifications of the **antenna**.

The data determined with the aid of the NWA is **valid for the surface** (or point) in which the **calibration** was performed. As a rule, the calibration procedure itself consists in providing the NWA **consecutively** with an **open circuit** (cable removed), a **short-circuit** and a **50 Ω match** (connector accessory). The NWA then uses the three impedances Z = ∞ , Z = 0 and Z = 50 Ω to calculate all the values it needs. The calibration must be **repeated** if a **different frequency range** is chosen. Further details on how to operate the instrument and navigate the menus should be looked up in the manual supplied by the manufacturer. At this point, it should also be borne in mind that the antenna being measured must be able to radiate in an **unobstructed space**.

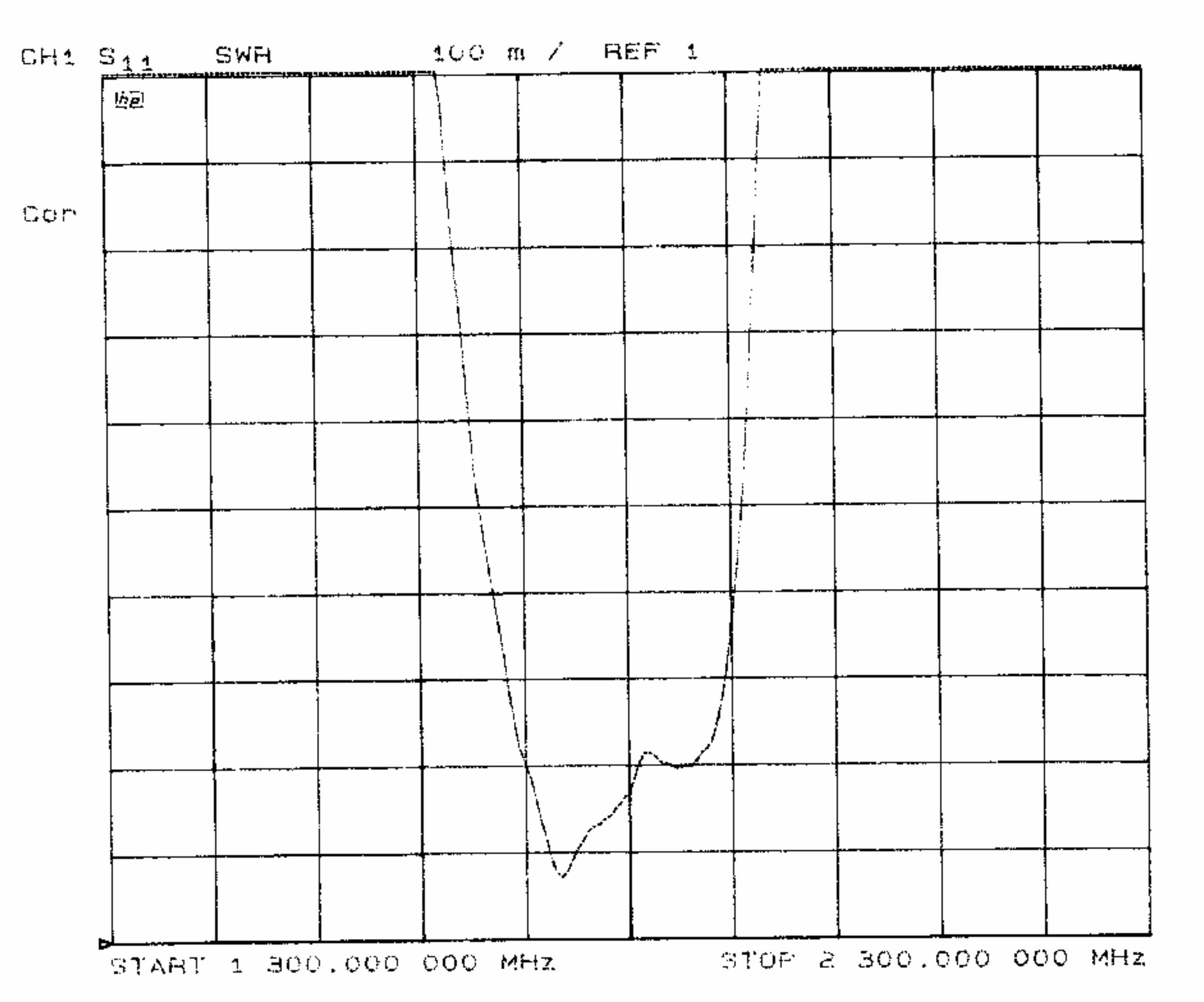


Fig. 6.4 SWR of a base station antenna in the frequency range from 1.3 GHz to 2.3 GHz

Fig. 6.4 shows the **standing wave ratio** (**SWR**) of a BTS antenna in the 1.3 GHz to 2.3 GHz frequency range as obtained with the aid of an NWA. The horizontal scale is divided into 100 MHz / division and the vertical scale is 0.1 / division. The horizontal baseline (marked by ∇ on the left-hand edge of the screen in the NWA used) represents the value s = 1. As a rule, these parameters can all be set as required; some NWAs provide a self-scaling facility (AUTOSCALE or similar). The very wide frequency range initially selected allows an **overview measurement for determining the resonance range** (in this case 1.7 GHz to 1.9 GHz), but not **finer analysis** (resolution not fine enough). For this purpose, the **frequency range** has to be suitably reduced **to the useful band** (and the NWA must then be recalibrated - see above). Fig. 3.5 shows that the SWR exceeds the value 1.2 slightly in only a very narrow range around 1.8 GHz, and therefore the antenna certainly fulfills the requirement for s <1.3.

Many NWAs permit the use and positioning of **markers** (∇ in Fig. 6.5) in order to determine values of special interest even more accurately. In the example shown below, marker 1 has been set to the SWR maximum.

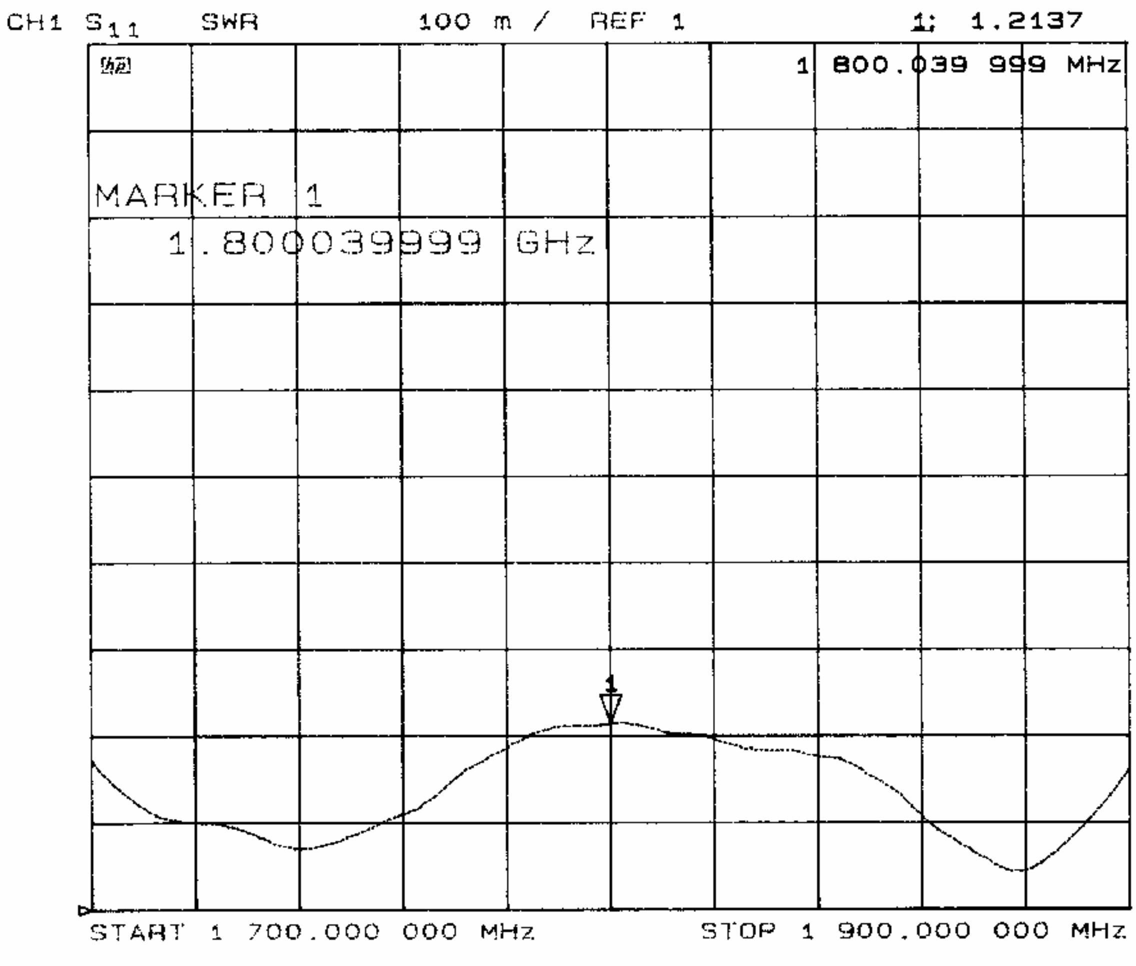


Fig. 6.5 SWR of the same base station antenna (BTS antenna) in the operating frequency band

For a still more accurate investigation, the vertical scaling can be changed as appropriate in order to obtain a finer resolution of the SWR. However, this step has been omitted in the present case.

Instead the associated **reflection coefficient** for the same frequency range is displayed in Fig. 6.6. The baseline now represents the value r = 0, and for the sake of clarity the value 0.01 / division has been chosen for the vertical scaling. The position of the marker is unchanged relative to the previous measurement and thus has the value r = 0.0898 ≈ 0.09 in the SWR maximum (see also Section 3.7 and the table on page 17).

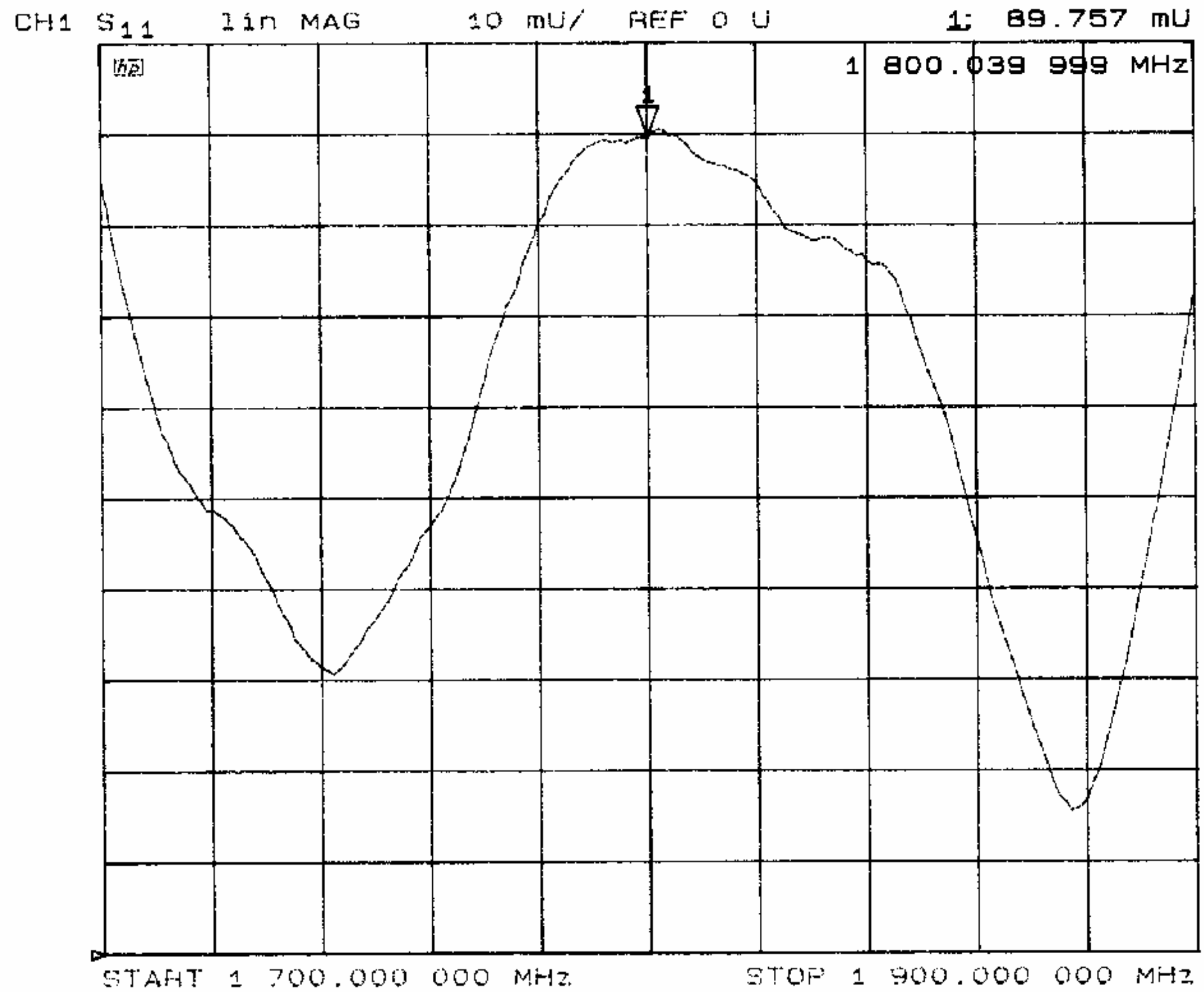


Fig. 6.6 Reflection coefficient of the same BTS antenna

The **display format** to select (standing wave ratio or reflection coefficient) is entirely **up to the wishes of the operator**. At this point, it is worth repeating that the NWA always measures the **same physical quantity** (voltage) and merely **converts** differently depending on the selected display format.

This also applies to the curve for the **return loss** (also measured on the same antenna) shown in Fig. 6.7. Its **negative value** is displayed by the instrument used for the measurement, i.e. the voltage ratio reciprocal to the defining equation indicated in Section 3.7. However, with a little practice it is possible to visualize the curve upside down. Using a different measuring instrument can make this unnecessary.

As a look at the conversion table shows, the value indicated at the marker also corresponds to the SWR maximum.

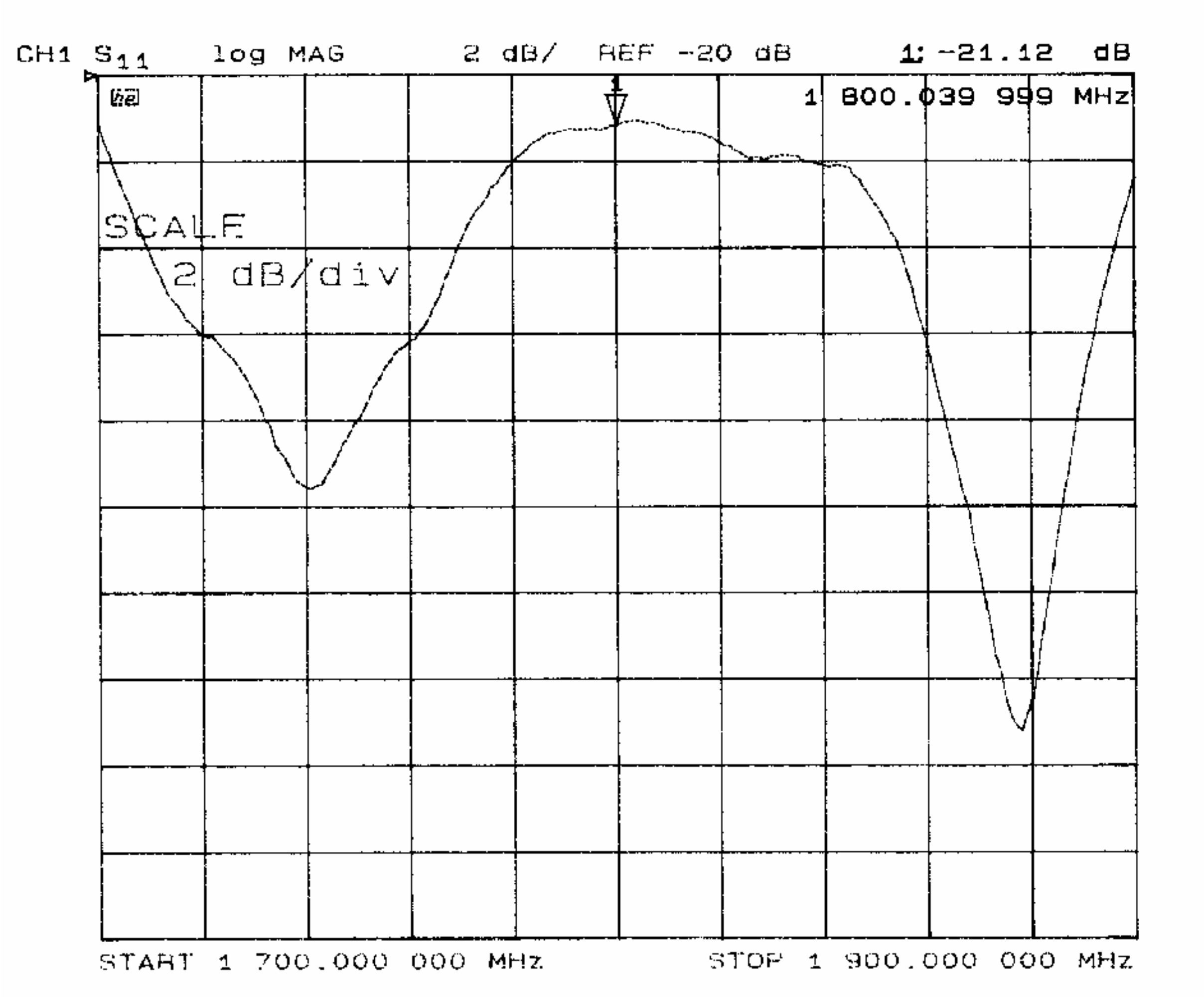


Fig. 6.7 Return loss of the same BTS antenna

Fig. 6.8 shows a Smith chart for the **complex antenna input impedance**. This curve has not been measured but calculated by the NWA from the measured complex voltages. Also indicated are the components producing the impedance shown by the marker, which in this case were a (very small) ohmic resistor and a capacitor. In many cases, this information makes the chart easier to understand; in the present case, for instance, it makes clear that at SWR maximum the antenna is being operated below its resonance frequency, but this is no substitute for accurately measuring the capacitance or inductance.

Although the display as a **Smith chart** often seems unclear at first glance (if necessary, it can be replaced, for example, by the display of the real and imaginary parts of the impedance), this example quickly shows that all impedances in the operating frequency range are close to the nominal value of 50 Ω.

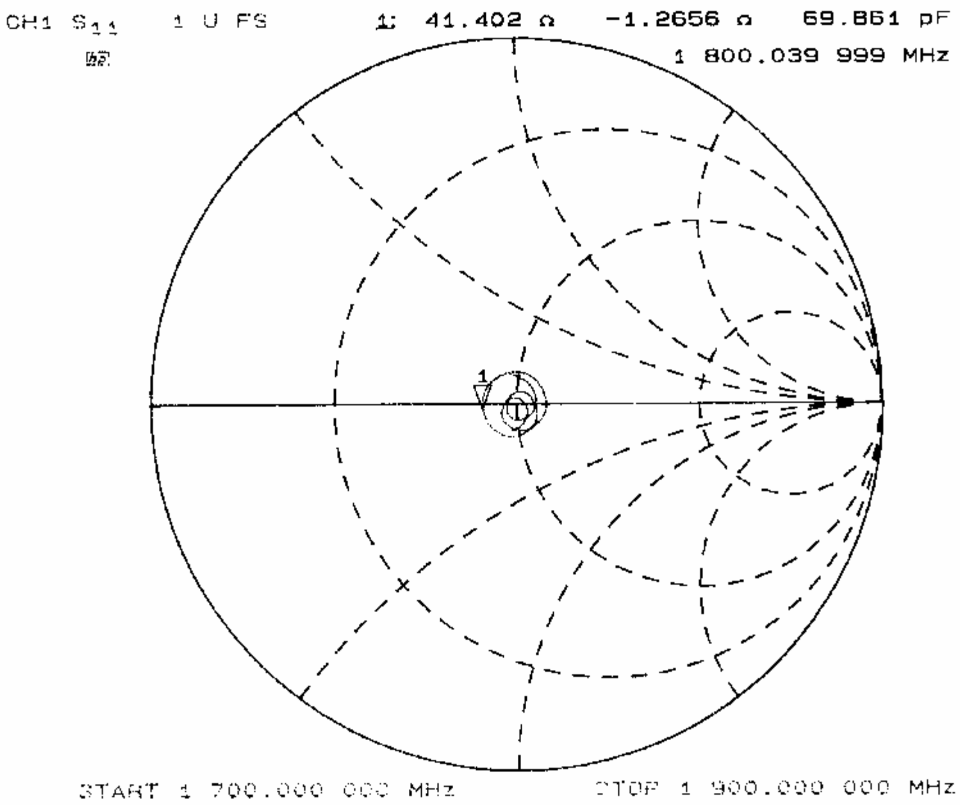


Fig. 6.8 Complex input impedance of the same BTS antenna

In conclusion, it should be pointed out again that antenna measurements (particularly measurements of patterns and SWR) must be performed **in a reflection-free environment**, i.e. at an open-field test site or in an anechoic chamber (Fig. 6.9).

Fig. 6.9 Antenna measurements at an open-field test site (left) and in an anechoic chamber (right)

6.2 Pulse Reflectometry (TDR)

Pulse reflectometry (also known as **time domain reflectometry** or **TDR**) is a method of **localizing mismatches and discontinuities**. The principle is shown in Fig. 6.10.

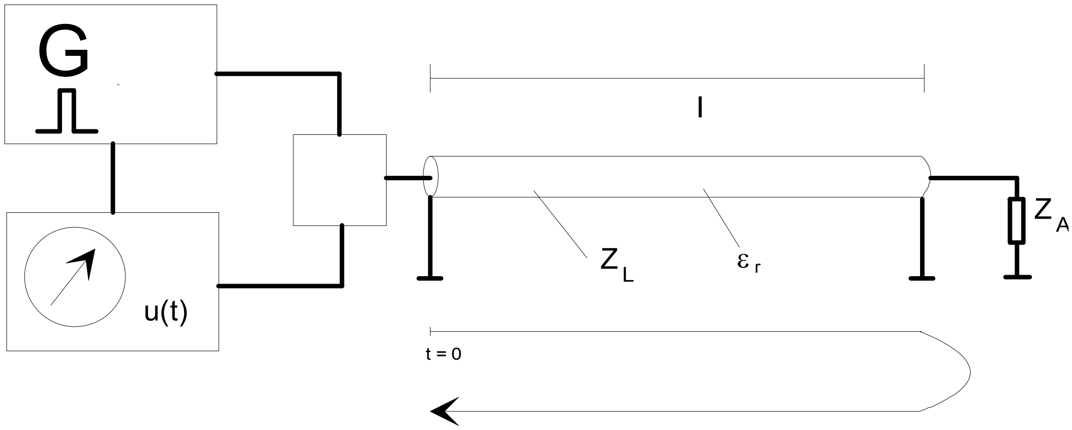


Fig. 6.10 Pulse reflectometer (diagram)

A pulse generator transmits a **steep-slope voltage pulse** to the setup under test and at the same time triggers a display unit (oscilloscope). Every discontinuity (characteristic impedance jump, interruption, short-circuit) causes a reflected pulse which can be observed on the oscilloscope. The **delay difference between the transmitted and reflected pulses** indicates the **location of the discontinuity**. The entire setup thus resembles a closed, one-dimensional wire **radar system** suitable first and foremost for checking and investigating antenna cables and RF cable distribution systems (even if built-in or below ground).

The **locating accuracy** (resolution in the direction of observation) depends decisively on the **slope steepness** of the transmitted pulse and the **bandwidth** of the oscilloscope. Pulse rise times should always be <850 ps. With rise times of ≈ 200 ps and using a modern oscilloscope with a bandwidth of 1 GHz and a cutoff frequency of 2 GHz, reflection points with a resolution of 1 cm or 2 cm can be localized reliably. Conventional oscilloscopes cannot meet these requirements. The tendency is to use sampling oscilloscopes, in which the signal to be displayed is sampled pointwise and displayed on the screen.

The **pulse duration** must be dimensioned in such a way that the reflected pulse occurs on the oscilloscope either clearly before or clearly after the trailing edge of the transmitted pulse, i.e. it depends on the length of the line system under test. Fig. 6.11 shows how the shape of the transmitted pulse affects that of the reflected signal.

The **maximum location distance** depends on the following quantities:

Cable loss
Power of the pulse generator
Sensitivity of the receiver

and can amount to several kilometers.

Pulse shape of the transmitted signal: Shape of the received signal:

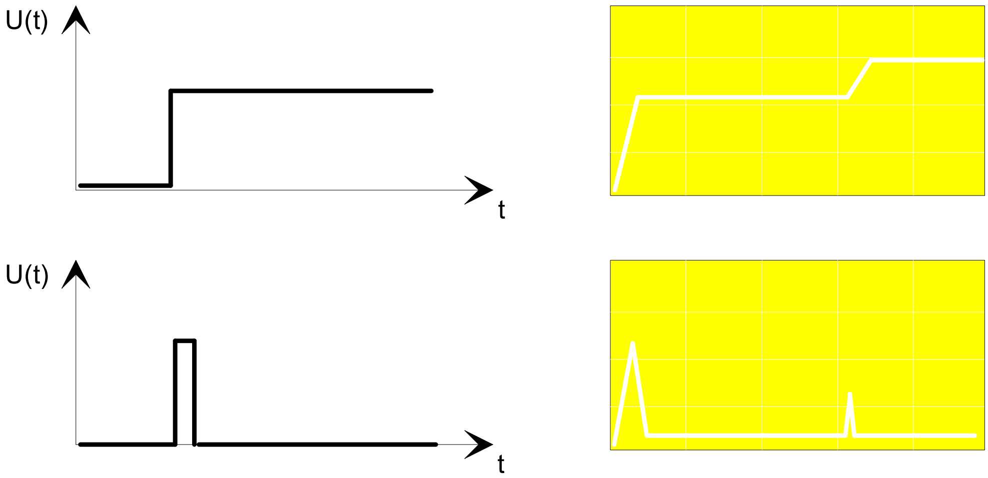


Fig. 6.11 Signal received from a discontinuity as a function of the transmitted pulse shape

The **propagation velocity** of a pulse on a line depends on the dielectric constant ε_r . If a pulse is to travel a line length l, the time required for this is

$$t = \frac{l}{c_0 / \sqrt{\varepsilon_r}} \qquad \text{where the velocity of light } c_0 \approx 3 \cdot 10^8 \text{ m/s.}$$

When analyzing a measurement, however, it should be noted that a pulse caused by a reflection at distance l has to travel along the cable **twice** (out and back), so that it is only after time

$$t = 2 \cdot \frac{l}{c_0 / \sqrt{\varepsilon_r}}$$

that it reaches the receiver. It therefore goes without saying that the **longest possible time**, which is a function of the total length of the cable under test, **must** be **determined before starting a measurement**, so that the appropriate scaling and settings can be selected on the display unit and, if necessary, the pulse shape can be adapted in line with the criteria discussed above. For example, in a 2 m cable having the dielectric constant ε_r = 2.5 this amounts to

$$t = 2 \cdot \frac{2\text{ m}}{3\text{ m / s}} \cdot 10^{-8} \cdot \sqrt{2{,}5} = 2.11\ .\ 10^{-8}\text{ s} = 21.1\text{ ns} \ .$$

If the dielectric characteristics of a cable are not known before starting a measurement, a cable of length l (accurately measured) should be connected as a DUT, left unterminated or provided with a short-circuit, and then the dielectric constant should be calculated from the measured delay t by transposing the formula shown above:

$$\varepsilon_r = \left[\frac{c_0 \cdot t}{2 \cdot l}\right]^2$$

Pulse reflectometry can also be performed using certain modern **network analyzers**. They do not operate in the same way as the pulse reflectometer shown in Fig. 6.10, but yield the same result. Without going into detail, it should be mentioned that the timing characteristic h(t) can be derived from the measured frequency response H(jω) of a DUT with the aid of the mathematical operation known as inverse **Fourier transform**:

$$H(j\omega) = \int_{0}^{\infty} h(t)\, \exp(-j\,\omega\, t)\, dt$$

As with the sampling oscilloscope, information is displayed on the screen. Fig. 6.12 was produced by this type of instrument, for example. It shows the impulse response of two coaxial cables, each approx. 1 m in length, which are connected in the correct order and with the correct characteristic impedance to an RF coupler and a 50 Ω resistor.

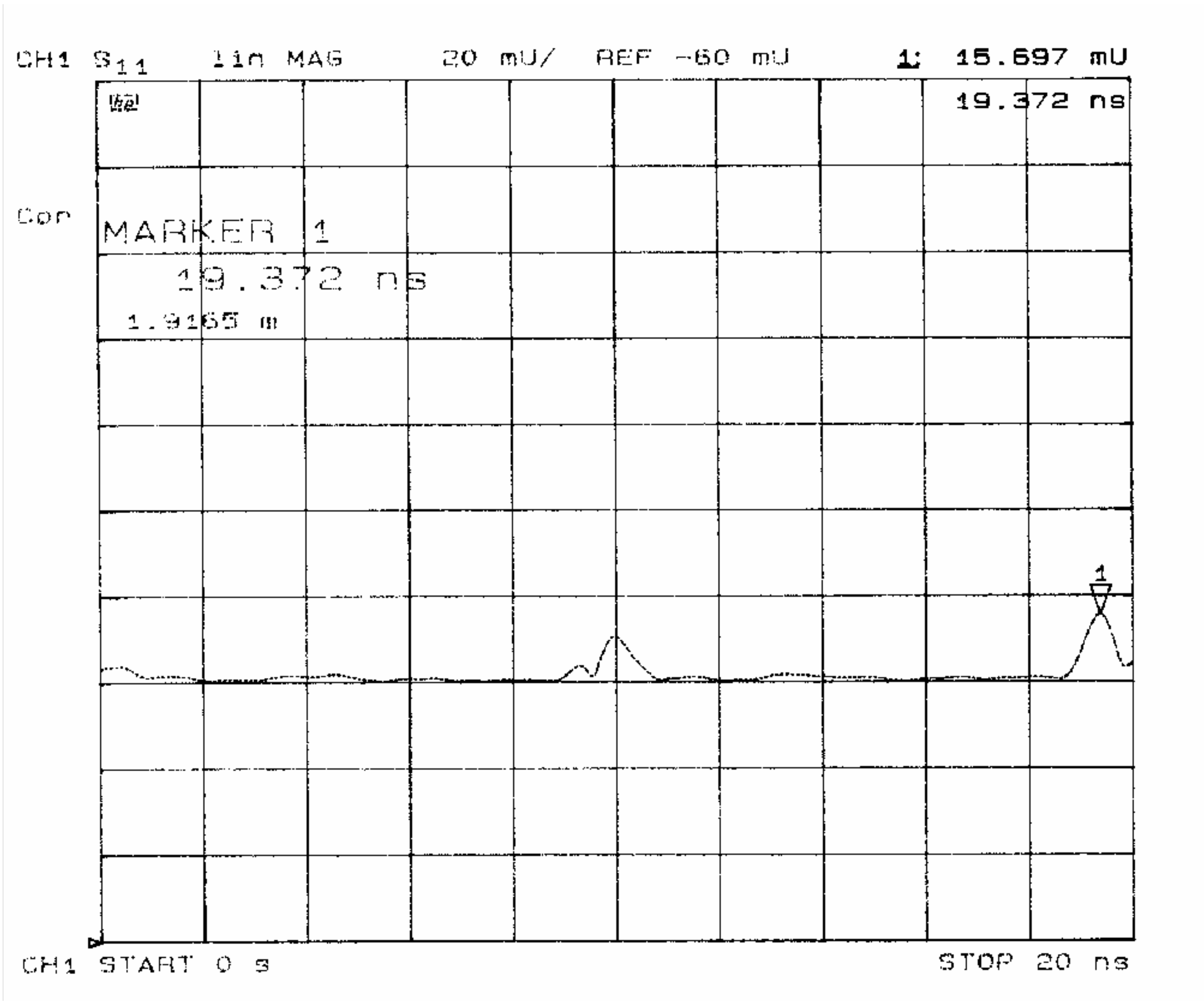


Fig. 6.12 Impulse response of two interconnected and terminated coaxial cables

As can be seen from the marker, the horizontal time axis can also be calibrated in units of length. In the example shown here, 20 ns correspond to cable length l = 2 m.
It has already been taken into account that the signal travels this path twice.

The first thing to be recognized is the coupling between the two cables at exactly 1 m (t = 10 ns). This causes the reflection coefficient r = 0.01 (scale division 0.02 / division), which (see table on page 13) corresponds to the SWR value s = 1.02. The match at a distance of 1.92 m has the reflection coefficient r = 0.016 and the SWR s ≈ 1.03.

Fig. 6.13 shows the impulse response of the BTS antenna from the previous examples, which is connected to a 1 m coaxial cable. After the antenna connector (= marker position), whose reflection coefficient r = 0.01 gives no cause for complaint, there is a 20 cm piece of cable that causes no reflections worth mentioning. The subsequent peaks indicate the complicated "inner life" of the antenna, i.e. splitters, baluns and dipoles.

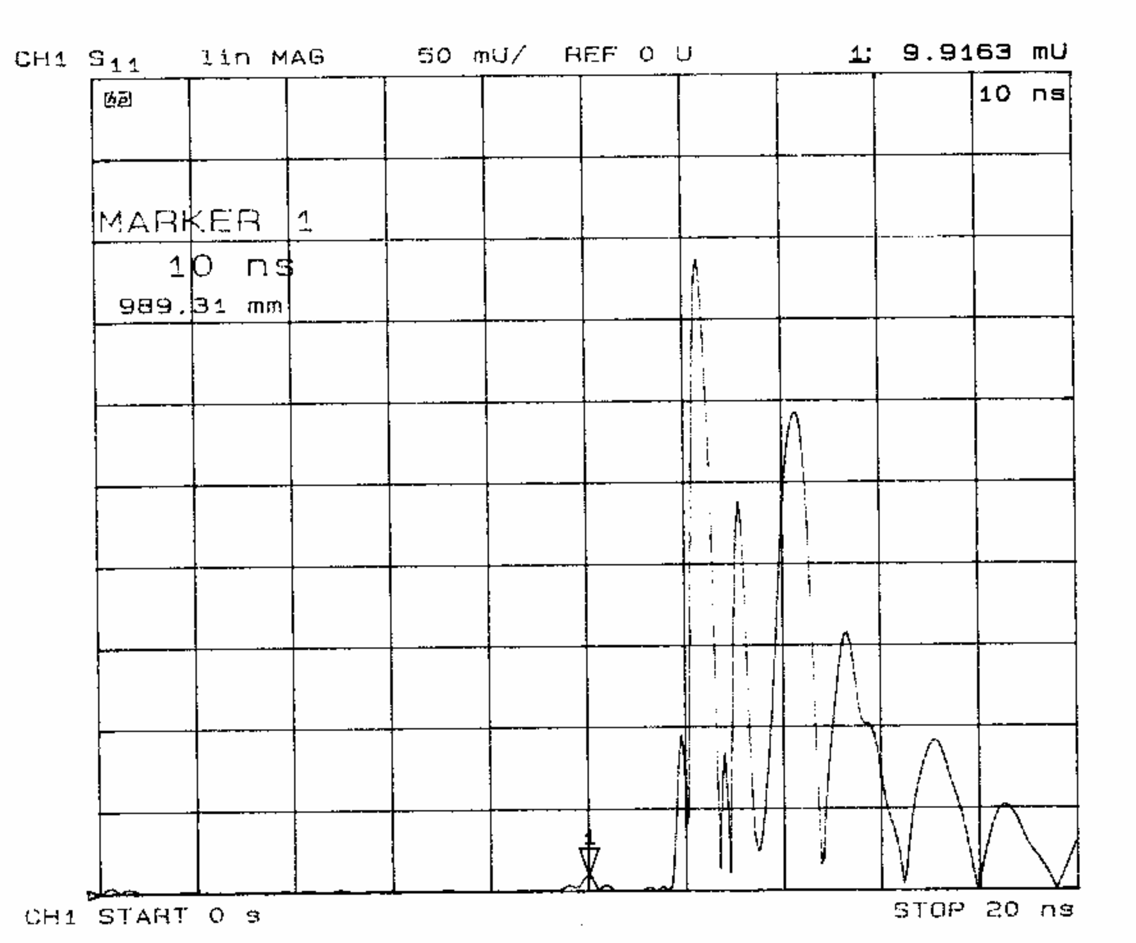
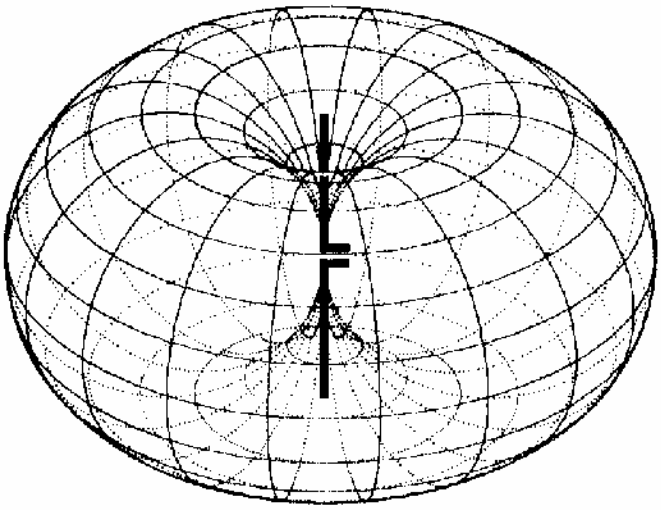


Fig. 6.13 Impulse response of a BTS antenna

It must be pointed out that the described TDR methods permit **no quantitative conclusions to be made about the reflection behavior of frequency-dependent components**. This is understandable considering that a steep pulse (just like the NWA) produces a **very wide frequency spectrum**, but that an antenna, for example, can only be matched in a particular frequency band. TDR methods can generally be used for **localizing discontinuities** and for measuring **reflection characteristics, but only in the case of frequency-independent components** (cables, connectors, etc).

7 Overview of the Main Antenna Properties

Antenna type	Directivity factor	Directivity factor in dB	Effective length	Radiation resistance
Isotropic radiator	1	0	-	-
Electrically short vertical antenna on conducting surface	3	4.7	h/2	$40 \left(\frac{\pi h}{\lambda}\right)^2 \Omega$
$\lambda/4$ antenna on conducting surface	3.3	5.1	0.16λ	40Ω
5 $\lambda/8$ antenna on conducting surface	6.6	8.2		
Electrically short dipole	1.5	1.8	l/2	$20 \left(\frac{\pi l}{\lambda}\right)^2 \Omega$
Half-wave dipole	1.64	2.1	$\lambda/\pi = 0.32 \lambda$	73Ω
Turnstile antenna	0.82	- 0.86		
Full-wave dipole	2.4	3.8	$>> \lambda$	200Ω
Electrically small loop antenna with n windings	1.5	1.8	$n \cdot \frac{\pi^2 D^2}{2 \cdot \lambda}$	$20 \Omega \pi^6 n^2 (D/\lambda)^4$
Full-wave loop (ring, circumference = 1 λ)	2.23	3.5		
Yagi-Uda antenna (6 elements)	typ. 10	typ. 10		
Log-periodic antenna in free space	typ. 4 to 5	typ. 6 to 7		
RF log-periodic antenna over conducting surface	typ. 4 to 20	typ. 6 to 13		



8 References and Recommended Further Reading

Author:	Title:	Publisher:
Balanis	Antenna Theory	John Wiley & Sons
Beckmann	Die Ausbreitung der elektromagnetischen Wellen *(The Propagation of Electromagnetic Waves)*	Becker & Erler
Braun	Planung und Berechnung von Kurzwellenverbindungen *(The Planning and Calculation of Shortwave Links)*	Siemens
Chatterjee	Antenna Theory and Practice	John Wiley & Sons
Collin	Antennas and Radiowave Propagation	McGraw-Hill
Davies	Ionospheric Radio	Peregrinus
Dombrowski	Antennen *(Antennas)*	Porta München Technik Berlin
Grosskopf	Wellenausbreitung (2 volumes) *(Wave Propagation)*	Hochschultaschenbücher
Heilmann	Antennen (3 volumes) *(Antennas)*	Hochschultaschenbücher
Hille/Krischke	Antennenlexikon *(Antenna Lexicon)*	vth
Hock/Pauli	Antennentechnik *(Antenna Engineering)*	Expert
Jasik	Antennas	McGraw-Hill
Kraus	Antennas	McGraw-Hill
Lee	Mobile Communications Design Fundamentals	Howard W. Sams
Lee	Mobile Cellular Telecommunication Systems	McGraw-Hill
Maslin	HF Communications	Pitman
Meinke/Gundlach	Taschenbuch der Hochfrequenztechnik *(Manual of High Frequency Engineering)*	Springer
Reithofer	Praxis der Mikrowellenantennen *(Microwave Antenna Practice)*	UKW-Berichte
Rothammel	Antennenbuch *(The Antenna Book)*	Franckh-Kosmos
Spindler	Das große Antennenbuch *(The Big Book of Antennas)*	Franzis
Stirner	Antennen (3 volumes) *(Antennas)*	Hüthig
Wiesner	Fernschreiben und Datenübertragung auf Kurzwellen *(Shortwave Telephony and Data Transmission)*	Siemens
Zuhrt	Elektromagnetische Strahlungsfelder *(Electromagnetic Radiation Fields)*	Springer

Selection of standards on the subject of antenna calibration:

DIN 45003, Meßverfahren für Ton- und Fernsehrundfunk-Empfangsanlagen im Frequenzbereich von 30 bis 1000 MHz *(Test Method for Sound and TV Broadcast Receiving Systems in the Frequency Range from 30 MHz to 1000 MHz)*

NBS Report 5539, Methods for Accurate Measurement of Antenna Gain, National Bureau of Standards, November 1957, Boulder (CO), USA

ANSI C-63.5, American National Standard for Calibration of Antennas

ARP 958, Aerospace Recommended Practice, Broadband Electromagnetic Interference Measurement Antennas; Standard Calibration Requirements and Methods, Society of Automotive Engineers Inc., January 1968, New York (NY), USA

IEEE Std 291-1991, IEEE Standard Methods for Measuring Electromagnetic Field Strength of Sinusoidal Continuous Waves, 30 Hz to 30 GHz

IEEE Std 149-1979, Standard Test Procedures for Antennas

CISPR/A(CO)48, Test Sites for Measurement of Radio Interference Field Strength