# LPDA-Antennas for Large Scale Radio Detection of Cosmic Rays at the Pierre-Auger-Observatory

von

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# 1. Introduction

Fundamental questions arise ever since humankind has been observing the sky. One of the still unanswered questions concerns the nature and the origin of ultra high energy cosmic rays (UHECRs).

Carrying energys up to several  $10^{20}$  eV, UHECRs induce a cascade consisting of millions of secondary partices if they hit the atmosphere of our earth. These extensive air showers (EAS) offer a great opportunity to detect cosmic rays - their direct observation is almost impossible due to their low flux.

The Pierre Auger Observatory is todays largest experiment to investigate UHECRs via their induced air showers. By using two complementary detection techniques, the Pierre Auger Observatory studies cosmic rays in great detail.

Beside the established detection techniques, cosmic rays can be observed by measuring radio pulses emitted during the shower evolution. Several test setups have been used to study the requirements on the hardware for radio detection at the Auger site. Within these approaches, the Black Spider LPDA, a logarithmic periodic dipole antenna (LPDA) was designed to precisely match the environmental requirements. As the test setups succeeded in the detection of EAS, a great effort to employ the radio detection technique at the Pierre Auger Observatory has recently been decided. The AERA-detector (Auger Engineering Radio Array) will cover an area of ~ 20 km<sup>2</sup> and is the first attempt to establish the radio detection on large scales in the Eev energy regime.

The used radio antennas play a key role for the successful detection of cosmic rays as they provide the sensitive component of the radio detector. We will achieve a deeper understanding of the mode of operation of LPDAs by measurements of the electrical transactions within the antenna structure and of single components of LPDAs.

The radio detection of cosmic rays makes various demands on the used antennas. We will motivate the electrical and mechanical conception of the used LPDAs as a consequence of a precise adjustment of the antenna properties to the demands of radio detection at the Auger site.

For the reconstruction of EAS from the measured radio pulses, a precise knowledge of the antenna characteristics is essential. Within this thesis, we determine the properties of the Black Spider LPDA in detail through measurements and simulations. Different test setups will be used to benchmark both, the electrical and mechanical performance of the Black Spider. A large experiment such as the AERA-project defines additional requirements on the antenna hardware. We present a LPDA design, specially developed for the AERA detector. The design will be motivated with focus on the demands of a mass production of the antennas.

# 2. Physics of Ultra High Energy Cosmic Rays

The discovery of cosmic rays was made in conjunction with investigations to a, by then, disparate phenomenon in physics. It was observed that statically charged bodies lose their charge in the course of time. This was first explained in 1896 with the discovery of radioactivity by A. H. Bequerel. In this theory active isotopes in the earth's crust decay and ionize the air which discharges the body. As further experiments showed that the radiation could not be shielded by placing lead between the earth and the body, this theory was doubted. It was finally disproved in 1912 when Victor Hess found during a balloon flight that the intensity of the ionizing radiation increases with larger heights, rather than decreases as expected for sources on the earth's surface. He concluded that at least a part of the radiation must have a extraterrestrial origin. After a series of experiments which confirmed Hess's observation he was honored with the Nobel Price in Physics in 1936 for the discovery of cosmic rays.

In 1937 Pierre Auger investigated cosmic rays in more detail and noticed coincidences between spatially separated particle detectors. He concluded that extensive particle showers are generated by high-energy primary cosmic rays that interact with air nuclei high in the atmosphere, initiating a cascade of interactions that ultimately yield a shower of secondary particles that reach ground level. His further studies showed that the energy of these cosmic ray induced air showers vary over several orders of magnitude extending up to  $10^{15}$  eV.

## 2.1 Energy Spectrum

Today the properties of cosmic rays and induced air showers are measured in detail by a series of experiments. The energy spectrum of the primary cosmic rays is shown in 2.1. It covers a range of approximately 12 orders of magnitude in energy and 32 orders of magnitude in flux. Below 10 GeV the differential flux is modulated by magnetic fields in our solar system. Above 10 GeV it can be approximated by a simple power law:

$$\frac{d^2\phi(E)}{dEd\Omega} \propto \left(\frac{E}{GeV}\right)^{-\gamma}, \ \gamma \sim 2.7 \quad . \tag{2.1}$$

On closer inspection two features are visible: named the 'knee' and the 'ankle as the spectrum reminds of a leg. At the knee at  $10^{15}$  eV only 1 particle per square meter per year is observed. Between knee and ankle the spectral index  $\gamma$  changes from



Figure 2.1: Energy spectrum of primary cosmic rays, adopted from [1].

2.7 to 3.1. The ankle at about  $10^{18}$  eV marks the energy above which cosmic rays are referred to as ultra-high-energy cosmic rays (UHECRs). Here the spectral index slightly canges to  $\gamma \sim 2.7$  again. UHECRs with energies of several  $10^{20}$  eV have been detected [2] which is remarkable as the expected event rate is below one particle per square kilometer per century.

Cosmic rays are usually defined as stable charged particles with extraterrestrial origin. Below 1 TeV the composition is well known and matches approximately the element distribution in our solar system. 90% of them are protons, 9% helium, 1% heavier nuclei and a vanishing fraction of electrons and positrons. The composition for higher energies remains uncertain. UHECRs are mostly assumed to be protons and iron, as any other nuclei would decay on their long way through cosmos.

UHECRs carry energies which are boyond the reach of todays accelerators. Thus

they provide an unique insight into physics at the highest energy scales.

However, it remains uncertain what mechanism can explain such energies. Pierre Auger suggested that it is gained by acceleration along electric fields of very large extension.

# 2.2 Origin

Contrary to particles with energies below 1 keV which come from the sun, cosmic rays with higher energies have extra solar origins. For UHECRs it is possible to constrain the number of cosmological objects considered to be sources to only a few by using theoretical models. With todays state of knowledge, an electric acceleration is only possible by pulsars which provide sufficient field gradients. The most favoured acceleration mechanisms are based on a model by Enrico Fermi [3]. Here, particles gather their energy by a statistical acceleration due to multiple scattering with turbulent magnetic fields.

These fields might occur in plasma clouds and plasma-shock-waves, emerging e.g. from super-novae. To categorize these sources, an estimate done by Hillas [4] is useful. A particle with charge  $Z \cdot e$  can be accelerated in a magnetic field B with an extension R up to an energy  $\epsilon_{cl}$ :

$$\epsilon_{cl} = Z \, e \, B \, R \sim \left(\frac{B}{10^{-4}T}\right) \left(\frac{R}{1pc}\right) 10^{20} eV. \tag{2.2}$$

It will rotate in the magnetic field until it reaches the energy  $\epsilon_{cl}$ , where it is ejected as the gyroradius would become bigger than the extension R for higher energies.

In fig. 2.2 the magnetic field strength is plotted against the extension of the accelerating region. The solid (dashed) line marks the energy of a  $10^{20}$  eV proton (iron nucleus) assuming eqn. 2.2. Only astronomical objects that lie on, or above these lines are able to accelerate cosmic rays up to  $10^{20}$  eV. Thus for the highest energies only a few source candidates have to be considered: active galaxy nuclei (AGN), radio galaxy lobes, neutron stars and white dwarfs.

Now, the question whether the cosmic rays can carry their whole energy to the observer remains.

# 2.3 Propagation

There are two dominant effects which influence the propagation of cosmic rays over large distances. Their path will be bent by the magnetic fields and, above a certain energy, they will interact with the omnipresent Cosmic Microwave Background (CMB).

Eqn.2.2 implies that cosmic rays with energies below  $10^{18}$  eV are trapped in our galaxies halo with a radius of 25 kpc, assuming a magnetic field in the order of  $\mu$ G. That is the explanation for the 'ankle' in the energy spectrum: particles with energies below  $10^{18}$  eV originating from other galaxies and subjected their to similar magnetic fields will not be able to leave their galaxy and thus never reach us. Hence cosmic rays below  $10^{18}$  eV must come from sources in our own galaxy whereas UHE-CRs have extragalactic origins. This change in the sources is visible in the spectrum.



Figure 2.2: Hillas Plot of the possible sources of cosmic rays. Diagonal lines indicate the maximum particle energy given by eqn. 2.2 [4].

However, with increasing energy the particles will be able to leave the galaxy. Intergalactic magnetic fields will also deflect charged cosmic rays. Although they are less strong compared to the galactic fields, they operate over larger distances. Thus particles will move on a chaotic path and not point back to their source. An observer will recognize an isotropic flux.

However, simulations predict that above  $E \sim 4 \cdot 10^{19}$  eV the deflection gets so small that sources can be derived [5]. An anisotropy in the arrival directions should be observed in this case.

For even higher energies the interaction with the CMB has to be taken into account. The production of a  $\Delta^+$ -baryon becomes possible if the center of mass energy of a cosmic ray proton together with a photon of the CMB exceeds the rest mass of

the  $\Delta^+$ . This requires proton energies above  $5 \cdot 10^{19}$  eV. The interaction probability increases with larger energies. The  $\Delta^+$  will decay in two dominant channels:

$$p + \gamma_{CMB} \rightarrow \Delta^+ \rightarrow n + \pi^+ \rightarrow \Delta^+ \rightarrow p + \pi^0.$$
(2.3)

The proton will loose a fraction of its primary energy to produce the mass of the pion. If the proton energy is still sufficient to produce another  $\Delta^+$  the process will repeat.

Assuming these processes, the propagation of protons over large distances can be simulated for different initial energies. This is shown in fig. 2.3 for  $10^{20}$  eV,  $10^{21}$  eV and  $10^{22}$  eV. In all cases the proton will be decelerated below  $10^{20}$  eV after about 100 Mpc travelled distance. Therefore cosmic rays reaching the earth with energies above  $10^{20}$  eV must come from sources closer than 100 Mpc. This effect should lead to a limit to the energy spectrum. The effect is called GZK-Cutoff after Greizen, Zatsepin and Kuzmin [6, 7] who predicted it in 1963. The GZK-Cutoff has recently been observed by the Pierre Auger Observatory [8].

For UHECRs with energies above  $5 \cdot 10^{19}$  eV we can draw the following conclusion from the two propagation effects: the particles will only be slightly deflected by the magnetic fields and they must come from sources closer than 100 Mpc. There are only a few hundred possible sources within this distance, which are distributed anisotropically. Thus the observer should find anisotropic arrival directions of UHECR's.



Figure 2.3: Simulated decrease of the proton energy with traveled distance for different initial energies [9].

# 2.4 Anisotropy

The Pierre Auger Observatory was designed to investigate UHECRs in particular. The collaboration recently reported an anisotropy in the arrival directions of UHE-CRs [10].

As AGNs are one of the most promising source candidates (see. 2.2), a correlation between them and the measured events has been analyzed. To determine the significance of this correlation one needs to know the chance probability that the observed anisotropy arises from an isotropic flux. With a certain probability one will find a correlation even if the arrival directions are isotropic, especially with low statistics. If this probability is low, the measured correlation is significant, thus the anisotropy is evident. The result of this analysis depends on the following parameters:

### The threshold energy $E_{th}$ :

A low  $E_{th}$  would allow cosmic rays which have been deflected too far from their source, a high  $E_{th}$  would decrease the statistics.

The maximum angular deviation  $\phi$  from an AGN:

This cut takes the detector resolution and the fact that all UHECR's are slightly deflected into account.

The distance  $D_{max}$  up to which AGN's are considered:

This constrain is motivated by the GZK-Cutoff.

The probability to accidentally identify an isotropic distribution as an anisotropy can be minimized for the following parameters:

$$E_{th} = 5.7 \cdot 10^{19} eV, \qquad \phi = 3.2^{\circ}, \qquad D_{max} = 71 M pc.$$
 (2.4)

The result of this analysis is visualized in fig.2.4. 27 UHECR events and 442 AGNs, whereas 292 lie in the field of view of the detector, survive these cuts. 20 events were counted as correlated with AGN's. An accumulation of correlating events along the supergalactic plane around Centaurus A can be located. In average, 5.6 correlations are expected from simulations with isotropic skies for these parameters. Hence the probability to observe this result by chance can be estimated to be  $< 10^{-5}$ .

As the statistics of the experiment will strongly increase over the next years, further interesting results can be expected from the Pierre Auger Observatory.



Figure 2.4: Aitoff projection of the sky in galactic coordinates with the 442 AGN's closer than 71Mpc (red stars) and the 27 UHECR events above  $5.7 \cdot 10^{19}$ eV observed by the Pierre Auger Observatory (black circles with  $3.2^{\circ}$  angular distance). If an AGN lies within a black circle it correlates with the cosmic ray. The white area lies beyond the field of view of the detector. Blue shading indicates regions of equal exposure. The dashed line marks the supergalactic plane, the white cross Centaurus A [10].

# 3. Extensive Air Showers

Cosmic rays finding their way into the atmosphere of our earth will collide with an air nucleus at a certain height above ground. In this collision, secondary particles are produced which themselves carry enough energy to initiate further inelastic processes. The result is a cascade of particles propagating through the atmosphere with almost the speed of light.

These extensive air showers (EAS) offer a great opportunity to detect cosmic rays. The direct detection of UHECRs is almost impossible due to their low flux. Luckily the earth's atmosphere works as an extensive calorimeter which can be read out by special detection techniques. If the shower properties are well understood, the information carried by the primary particles, like energy, arrival direction and composition can be reconstructed by measuring their air showers.

In this chapter the basic properties of EAS are introduced and the different detection techniques are motivated. The radio detection technique is emphasized as it is of fundamental interest for the further discussion.

# 3.1 Shower Development and Simulation

Beside protons also heavier nuclei hit our atmosphere. A shower induced by a nucleus with energy E and atomic number A can be approximated by A independent proton showers with energy E/A. Therefore only proton induced showers will be discussed.

Our atmosphere has an integrated column density of  $X_{atm} \sim 1000 \ gcm^{-2}$  in vertical direction. Considering the mean path lengths in air this corresponds to  $\sim 11$  hadronic interaction lengths and  $\sim 27$  electromagnetic radiation lengths. The first interaction typically takes place at  $X_0 \sim 90 \ gcm^{-2}$  which corresponds to an average height of 20 km.

An EAS can be divided into a hadronic, an electromagnetic and a muonic cascade (see 3.1). The hadronic cascade is directly initiated by the first interaction of the cosmic ray. In this collision the cosmic ray loses about 50 % of its energy and produces a cascade of 90 % pions and 10 % kaons. The charged pions and kaons extend the hadronic cascade through further interactions. Finally they decay into muons and neutrinos:  $\pi^+ \to \mu^+ \nu_{\mu}$ ,  $\pi^- \to \mu^- \overline{\nu}_{\mu}$  and  $K^+ \to \mu^+ \nu_{\mu}$ . They form the muonic cascade. Due to their high  $\gamma$ -factor most of the muons reach ground level before they decay.

Due to their short lifetime ( $\tau \sim 10^{-16}$  s) the neutral pions decay to  $\pi^0 \to \gamma \gamma$  before they can interact further. As the  $\pi^o$  contributes a fraction of approximately 1/3 of



Figure 3.1: scheme of an proton induced air shower (left) (adapted from [11]), and sideview of a  $10^{15}$  eV proton shower simulated with CORSIKA (right) [12].

all produced pions, about 30 % of the energy of the hadronic cascade is converted into photons at each radiation length. These photons initiate the electromagnetic cascade. They produce electrons and positrons primarily through pair production and to a small extent through compton scattering. The electrons and positrons make further photons through Bremsstrahlung. In the further shower development, the electromagnetic component can amount up to 90 % of the whole particle content. The electromagnetic cascade dies out at particle energies of  $\sim 1$  MeV where electrons loose their energy only by ionization until they are stopped.

As EAS consist of millions of particles, their development can hardly be calculated analytically. Thus Monte-Carlo based showers simulations are essential for understanding shower properties. A powerful program to simulate EAS is provided by CORSIKA [13]. For a given species, energy and arrival direction of the primary particle it calculates the shower development for every shower particle by considering every possible interaction, atmospheric properties and the earth's magnetic field. In fig. 3.1 an example of a CORSIKA simulated shower is shown. On output CORSIKA delivers the position, arrival time and momentum of every particle that reaches ground level. It also gives information about the longitudinal and lateral shower profiles which are essential to reconstruct cosmic rays from measurements.

## **3.2** Shower Profiles and Methods of Observation

The shape of an EAS is almost independent of the energy and composition of the primary particle. On the other hand, the size of the shower and the atmospheric depth of the first interaction scales with the species and its primary energy. This enables measurements of cosmic ray properties via their induced air showers. Two different methods have been established. They are based on measuring either the longitudinal or the lateral shower profile. The longitudinal profile is observed in the sky by special telescopes whereas the lateral profile can be measured by an array of surface detectors.

### 3.2.1 Longitudinal Shower Profile

The longitudinal shower profile describes the development of shower properties along the shower axis, e.g., the total number of particles. To consider the changing air density, usually the integrated column density X is chosen as reference value instead of the distance the shower particles travelled through the atmosphere. Assuming that the shower disk propagates with the speed of light, the longitudinal shower profile describes also the temporal development. Thus X is a measure of the shower age. X is also referred to as slant depth.

The number of electrons  $N_e$  in the shower can be parameterized by the Geisser-Hillas function [14]:

$$N_e = N_{max} \left(\frac{X - X_0}{X_{max} - X_0}\right)^{\frac{X_{max} - X_0}{70}} exp\left(\frac{X_{max} - X}{70}\right).$$
(3.1)

Here,  $X_0$  is the atmospheric depth at the first interaction.  $X_{max}$  is the depth at the shower maximum with the maximum particle number  $N_{max}$ , both depending on the primary energy  $E_p$ :

$$X_{max} \sim X_i + 55 \cdot log \ (E_p) \ gcm^{-2}, \qquad N_{max} \propto E_p.$$
(3.2)

 $X_i$  depends on the particle species, hence the longitudinal shower profile is sensitive to the chemical composition of the cosmic rays. For UHECRs the shower maximum is at about  $X_{max} = 700 - 800 \ gcm^{-2}$ . It differs about 100  $cm^{-2}$  for the lightest and heaviest particles, protons and iron.

In general the number of particles in the shower grows from the first interaction to the maximum number  $N_{max}$ . With increasing shower age it falls off again as low energy shower particles will be absorbed or decay. For vertical showers almost the maximum number  $N_{max}$  will reach ground level as the shower maximum lies close to the ground. For inclined showers the shower maximum is higher in the atmosphere and thus only a fraction will reach ground level.

The electrons described by eqn.3.1 and other charged particles excite nitrogen atoms traversing the atmosphere. When the nitrogen deexcites a trace of fluorescence light

is left along the showers path. By tracking this trace of light with telescopes, the whole shower can be observed and the arrival direction of the cosmic ray can be derived.

The amount of light received in the telescopes is proportional to the number of charged particles at a certain shower age. By integration over the whole trace, the total visible energy  $E_{vis}$  can be observed. It is proportional to the primary particles energy  $E_p$ :

$$E_p \propto E_{vis} \propto \int N_e(X) dX$$
 (3.3)

Hence, with this technique also calorimetric measurements of the whole shower development can be done. Finally the primary particles energy can be estimated. Such a measurement of the longitudinal shower profile is shown in Fig.3.2.



Figure 3.2: Longitudinal profile of a  $4.1 \cdot 10^{19}$  eV air shower recorded with a fluorescence telescope of the Pierre Auger Observatory. The energy of the received fluorescence light along the trace is plotted against the slant depth. The red line shows the fit to the Geisser-Hillas function. The shower maximum (red dot) lies at  $809 \ gcm^{-2}$ . Note that the slant depth exceeds  $1000 \ gcm^{-2}$  because the shower is inclined (52.5° in zenith).

## 3.2.2 Lateral Shower Profile

At the primary interaction, the newly produced hadrons will carry some statistic transverse momentum with respect to the direction of the inducing cosmic ray. In further interactions the overall transverse momentum of the electromagnetic and hadronic cascade will increase. The muons will propagate in straight lines until they reach ground level. The particles of the electromagnetic cascade will spread out further with increasing shower age due to multiple scattering.



Figure 3.3: Lateral shower profile of a  $6.9 \cdot 10^{19}$  eV air shower observed with the array of surface detectors of the Pierre Auger Observatory. The particle density measured in 18 different stations is given in units of a muon that hits the detector vertically (VEM). The red line represents a fit to the lateral density function with its uncertainties (grey band).

That leads to a lateral shower structure with a shower core of about 100m diameter surrounded by a disk of a few kilometers extension. 90% of the shower energy and the majority of the hadrons is concentrated in the shower core. The disk is formed by low energy electrons, photons and muons. For the disk, the particle density for a lateral distance r to the shower core can be parameterized by the Nishimura-Kamata-Greizen (NKG) function [15]:

$$\rho(r) \propto \left(\frac{r}{r_M}\right)^{s-2} \left(1 + \frac{r}{r_M}\right)^{s-4.5}.$$
(3.4)

The Moliere radius  $r_M$  in air describes the electromagnetic interaction in matter. It depends on the air density  $\rho(h)$ . The parameter s indicates the steepness of the lateral distribution. It increases with the shower age X:

$$s(X) = \frac{3X}{X + 2X_{max}}.$$
(3.5)

Hence the lateral density distribution gets flatter with increasing slant depth. The particle density at ground level can be measured by particle detectors at different distances to the shower core. Fig.3.3 shows a measurement by the surface detector array of the Pierre Auger Observatory. These measurements can be fitted by eqn.3.4. Out of this fit the shower age and the total number of particles that reach ground level can be determined. With knowledge of the longitudinal shower development (see eqn.3.1) the maximum particle number  $N_{max}$  can be calculated if the inclination of the shower is known. The primary energy of the cosmic ray is than defined as it

is proportional to  $N_{max}$ .

The inclination of the shower can be observed as follows: in first approximation the shower reaches the ground as a flat disc of a few kilometers diameter. The arrival time can be monitored for different distances to the shower core by an array of particle detectors. The shower axis can than be calculated from the timing differences.

# 3.3 Radio Signals from Extensive Air Showers

Shower particles from EAS form a disk while propagating through the atmosphere. On closer inspection the disk gets a longitudinal extension of several meters. Within this disk, processes take place which yield the emission of a radio pulse. Different theoretical approaches for the emission mechanism have been formulated.

## 3.3.1 Origin of Radio Signals from EAS

The idea that radio signals are produced in EAS was first introduced by Askaryan in 1962 [16]. In his model, Cherencov radiation is the origin of the radio emission. It is produced by shower particles when their velocity exceeds the speed of light in air. The single emissions will add up to a coherent signal if the emitting region is smaller than the radiated wavelengths. For the shower disk with a few meters extension, this is the case for wavelengths in the radio regime. As the radiation from negative and positive charges would cancel, a charge excess is needed in this approach. Such a charge excess can evolve due to the annihilation of positrons with electrons from air molecules.

Todays measurements give a strong evidence that the amplitude of the emitted radio pulse depends on the geomagnetic field, which cannot be explained within this model. Thus a geomagnetic origin of the radio signals might be promising.

This was first suggested by Kahn and Lerche. In their theory, electrons and positrons in the shower disk are separated by Lorenz force [17]. The evolving transverse current will emit dipole radiation. Like in Askaryans model it will yield a coherent radio signal. Scholten, Werner and Rusydi recently published a 'Macroscopic Description of Coherent Geo-Magnetic Radiation from Cosmic Ray Air Showers' which is based on the same idea of transverse currents [18].

A different approach for a geomagnetic radio emission was presented by Heino Falke and Peter Gorham in 2002 [19]. In this straight-forward model, the emission is coherent synchrotron radiation from charged particles rotating in the earth's magnetic field. The advantage of this model is that it is based on the well understood synchrotron theory. Within this theory the frequency spectra of synchrotron pulses emitted by highly relativistic electron-positron pairs is described. Their emission is relativistically boosted into a small cone along the shower axis. The emitted power  $P_s$  by a synchrotron pulse from a single particle with mass m and charge q is given by [20]:

$$P_s = \frac{2q^2c}{3r^2}\beta^4\gamma^4. \tag{3.6}$$

Here, r is the gyroradius,  $\beta = v_{\perp}/c$  the particles velocity perpendicular to the rotation axis in units of the speed of light and  $\gamma \sim E/m$  the Lorenz factor. The

charge is forced on the circular trajectory by the Lorenz force in the earths magnetic field B:

$$r = \frac{\gamma m v_{\perp} c}{q B}.$$
(3.7)

The total radiated power is than given by:

$$P_s = \frac{2q^4 v_{\perp}^2 B^2}{3m^4 c^5} E^2.$$
(3.8)

From this thought we can learn that the radiated power of a synchrotron emission scales with  $m^{-4}$ . Therefore electrons and positrons emit the strongest pulses. As they also give the major fraction of the particle contentin the shower, the emission of other charged particles is negligible. Furthermore the radiated power depends quadratically on the particles energy. To consider a coherent emission we can think of the shower being one particle with mass  $m = N \cdot m_e$ , charge  $q = N \cdot e$  and energy  $N \cdot E$ . The total emitted power P will than be gained by a factor  $N^2$ :

$$P = N^2 P_s. aga{3.9}$$

Thus it is characteristic for coherent emission that the radiated power increases quadratically with the particle number in the shower and not linear as predicted for incoherent emission.

Starting with this single sources, a gradual integration over the whole shower is performed by Falck and Gorham. By integrating over the longitudinal thickness of the shower disk, coherence effects occur. For the integration over the lateral coordinate, the particle density as given by the NKG function (eqn.3.4) is used. Finally an integration over the whole longitudinal shower evolution is performed. The result is a prediction of the following properties of the emitted radio signal.

#### **Energy Dependence:**

In the described model, the resulting radio pulse is basically a coherent superposition of single synchrotron pulses from electron-positron pairs. The number of electrons and positrons in the shower is proportional to the energy of the primary cosmic ray, see chapter 3.2.1. Hence the resulting electric field amplitude is also proportional to the primary energy. In the previous approach we found that the total emitted power of the radio pulse scales quadratically with the particle number. That is equivalent as the power carried by an electric field is proportional to the square of the electric field strength.

#### Lateral Dependence:

The lateral dependence of the pulse height is governed by the lateral and longitudinal shower structure. The pulse height is predicted to decrease exponentially with the distance to the shower core.

#### **Pulse Spectrum:**

As the starting point for this model were the frequency spectra of single synchrotron pulses, the frequency content of the resulting radio pulse can be deduced. A cutoff in the spectra above  $\sim 100$  MHz can be denoted. This results from the loss of coherence when the wavelength becomes shorter than the longitudinal extension of the shower disk. Furthermore the spectra should change with different distances to the shower core. A shift towards the lower frequencies with increasing distance is expected.

#### Azimuthal Dependance:

The amplitude of the emitted synchrotron radiation depends on the angle between the magnetic field vector and the velocity of the particle. Assuming an inclined magnetic field this would yield an anisotropy of the pulse height for different arrival directions of cosmic rays.

Although these analytic calculations take a detailed shower structure into account, they are based on approximations, e.g. far field approximations for the electric fields. They describe basic shower properties but are not sufficient for precise quantitative predictions. The precision can be increased by Monte Carlo simulations based on the geosynchrotron emission model.

### 3.3.2 Simulation of Radio Signals from EAS

Based on the geosynchrotron approach, Tim Huege developed within his PhD thesis [21] a powerful Monte Carlo simulation named REAS (Radio Emission from extensive Air Showers). REAS calculates the synchrotron pulse emitted by each electron-positron pair in the shower. The shower particles are generated with a spatial distribution based on the longitudinal and lateral shower profiles as discussed in chapter 3.2 (see eqn.3.1 and 3.4). The resulting superimposed radio signal is given for any position on the ground. The big advantage is that no approximations are necessary and parameters of the showers can be tuned easily.

Together with Falcke, Tim Huege investigated different shower properties in detail [22]. Two interesting results of their work, the pulse spectrum and the dependence on the primary energy will be discussed below.

#### **Pulse Spectrum:**

The pulse spectrum as received on ground for a vertical shower is simulated in fig.3.4 for different distances to the shower core. The spectrum shows a steep decline to towards higher frequencies. The described loss of coherence results in a chaotic signal for the high frequencies. In the coherent region the pulse amplitude E can be fitted to an exponential law:

$$|\vec{E}| \propto exp\left(-\frac{\nu - 10MHz}{\nu_0}\right). \tag{3.10}$$

The parameter  $\nu_0$  varies from 47 MHz at 20 m distance to 5 MHz at 500 m. With growing distance to the shower axis, the coherence will be lost at lower frequencies. Hence, for large distances the spectrum is dominated by the lower frequencies. For 500m distance, only frequencies below 40 MHz can be observed. This behavior has a



Figure 3.4: Frequency spectra of the radio pulse from a vertical  $10^{17}$  eV proton shower simulated for different distances to the shower core. From top to bottom: 20 m, 140 m, 260 m, 380 m and 500 m. The colored lines indicate exponential fits to the coherent region [22].

strong influence on the conception of a radio detector as it can only be designed for a limited frequency range. It should be sensitive to the lower frequencies to detect even most distant showers. A lower limit for the radio detection is given by human made radio signals and galactic noise at frequencies below 30 MHz.

#### **Energy Dependance:**

Fig.3.5 shows the increase of the pulse amplitude with the primary energy for various distances from the shower axis. As reference value the 10 MHz component of the electric field is chosen. The signal should be coherent for all reviewed distances at 10 MHz (see fig.3.4). The dependence of the electric field strength  $|\vec{E}|$  on the primary energy  $E_p$  can be described by a power law:

$$|\vec{E}(r, E_p)| = E_0 \cdot \left(\frac{E_p}{10^{17} eV}\right)^{\kappa(r)},$$
(3.11)

with  $\kappa(r)$  close to one.  $E_0$  changes from 10.18  $\mu V m^{-1} M H z^{-1}$  at 20 m distance to  $E_0 = 0.265 \ \mu V m^{-1} M H z^{-1}$  at 500 m. This approximately linear behavior is expected for coherent emission as discussed in the previous chapter.

Furthermore, fig.3.4 and 3.5 give information about the lateral dependence. A steep decline of the electric field strength with growing distance to the shower axis can be observed for every frequency and primary energy respectively. A new version of REAS, REAS2, is coupled to CORSIKA (see chapter 3.1). CORSIKA gives a more realistic information of the particle content in the shower than the parameterizations used in REAS1. However, comparisons show that there is no dramatic change in the



Figure 3.5: Simulation of the 10 MHz component of the electric field for vertical showers from  $10^{15}$  eV to  $10^{19}$  eV at five different distances. From top to bottom 20 m, 100 m, 180 m, 300 m and 500 m. The lines are fits to the simulated data [22].

results from both versions [23].

Most of the air shower properties predicted by theoretical approaches and simulations have been verified through measurements which will be discussed now.

## 3.3.3 Measurements of Radio Signals from EAS

Inspired by Askaryans idea J.V. Jelly and co-workers first detected a radio pulse from an EAS in 1964. Jelly used an array of 72 antennas which were triggered by Geiger counters on the ground. Other research groups experimented with different setups and proved coincidences between radio signals and the inducing air showers. Due to insufficient theoretical models, technical problems and the great success of other detection methods, the interest in the radio detection technique ceased about ten years after its discovery.

Within todays giant cosmic ray experiments and Falcke and Gorhams promising geosyncrotron approach, in 2002 the topic was reanimated. Experimental investigations resumed in 2004 with the two large experiments LOPES and CODALEMA. LOPES (LOFAR Prototype Station), located within the KASCADE cosmic ray detector at the Forschungszentrum Karlsruhe, was actually planned to develop hardware for the LOFAR (Low Frequency Array) telescope. It uses an array of inverted V-dipole antennas externally triggered by KASCADE.

CODALEMA (Cosmic ray Detection Array with Logarithmic ElectroMagnetic Antennas) first used the antenna hardware of the Nancay Radio Observatory but is now supplied with fat dipole antennas. These are triggered by four stations of particle detectors on the ground. Beside these externally triggered experiments a great effort for a large self-triggered antenna array within the Pierre Auger Observatory has recently been decided [24]. Different antenna types and designs of autonomous self-triggered readout stations have been tested at the site during the last years. The location provides great possibilities to compare the results with other detection techniques. The Auger Radio effort is the first attempt to establish the radio detection technique on large scales.

In general the reconstruction of cosmic ray properties via the radio detection technique works similar to the method using surface particle detectors described in chapter 3.2.2. The radio signals evolving with the particle disk form a footprint when they reach the ground. Antennas placed within this footprint will receive electric field strengths up to several  $\mu V m^{-1} M H z^{-1}$ . The unit  $\mu V m^{-1} M H z^{-1}$  is used as the total amplitude observed by an antenna depends on the bandwidth of the antenna. Thus, dividing the amplizude by the bandwidth leads to a quantity comparable to other experiments. Comparisons between the measured pulse heights in the different antennas allow a reconstruction of the shower core position and the primary particle energy if the lateral distribution is known. The timing differences between the received signals define the shower axis.

From measurements of air showers in coincidence with the Haverah Park experiment H.R. Allan and collaborators were able to investigate the influence of different parameters to the radio signal. The results can be summarized in the following parameterization for the electric field amplitude  $\epsilon_{\nu}$  per unit bandwidth:

$$\epsilon_{\nu} = 20 \left( \frac{E_p}{10^{17} eV} \right) \cdot \sin\alpha \cdot \cos\theta \cdot exp \left( -\frac{R}{R_0(\nu, \theta)} \right) \qquad [\mu V m^{-1} M H z^{-1}]. \tag{3.12}$$

The amplitude  $\epsilon_{\nu}$  is given for a certain frequency  $\nu$  which denotes the center frequency of the used receiver. Furthermore  $\alpha$  is the angle between the shower axis and the magnetic field. Air showers with large zenith angles  $\theta$  have a more distant shower maximum as described in section 3.2.1. The resulting reduction of the pulse amplitude is considered with the  $\cos\theta$  term. Finally the pulse amplitude depends on the primary particles energy  $E_p$  and the distance to the shower axis R.

#### **Energy Dependence:**

Allan et al. found that the pulse amplitude scales linear with the primary particle energy. This is confirmed by recent measurements of the LOPES experiment presented in fig.3.6 [25]. The measured amplitudes are corrected for the radial dependence. Due to low statistics and uncertainties within the experimental setup, no fit to the data has been performed. However, the data can be described by a power law with an index close to one which agrees with the predictions of the theoretical models and simulations.

### Lateral Dependence:

The field strength should decrease exponentially with the distance to the shower axis according to Allan's parameterization. He then measured a scaling parameter  $R_0$  of 110 m. A recent investigation of the LOPES collaboration found  $R_0 = 230 \pm 51$ 



Figure 3.6: Dependence of the pulse amplitude on the primary particles energy  $E_0$  measured by the LOPES experiment [25].

m. The measurement with the fit to the data is shown in fig.3.7 in comparison to Allan's result. Nevertheless, all experimental results show an exponential behavior which is expected by the mentioned theories and the corresponding simulations.

#### Azimuthal Dependance:

Allan's parameterization considers the dependence of the pulse amplitude on the angular distance  $\alpha$  from the shower axis to the magnetic field vector. This has been recently investigated by the CODALEMA collaboration [26]. Fig.3.8 (left) shows the distribution of the measured arrival directions. A significant asymmetry between events coming from north compared to events from the south can be observed. This behavior is explained by an electric field strength scaling linearly with  $sin\alpha$ :

$$\epsilon_{\nu} \propto \sin\alpha = \frac{|\vec{v} \times \vec{B}|}{|\vec{v}| \cdot |\vec{B}|},\tag{3.13}$$

with the particle velocity  $\vec{v}$  pointing along the shower axis. The geomagnetic field vector at the CODALEMA site  $\vec{B}$  points to the south with an zenith angle of 27°. The emitted field strength should be maximal when  $\vec{v}$  is perpendicular to  $\vec{B}$ , thus events coming from the northern directions give stronger signals. As the experiment uses threshold triggers on the pulse amplitude, the event rate from southern directions is reduced.

Compatible results were recently reported by the Auger Radio collaboration [27]. In their measurements of arrival directions in coincidence with the Auger surface detectors (see fig.3.8 (right)), an excess of events coming from the south is obvious.



Figure 3.7: Lateral decrease of the pulse amplitude with growing distances to the shower axis measured by the LOPES collaboration. The signal is rescaled with the primary energy. The red line indicates a fit to the data, the green line shows the result of Allans work in comparison [25].

As the magnetic field vector at the Auger site is orientated to the north ( $58^{\circ}$  zenith), the same phenomena was observed.

This results give further support to the geomagnetic emission mechanisms discussed above in section 3.3.1.



Figure 3.8: 10° Gaussian smoothed map of arrival directions of cosmic rays measured by the CODALEMA experiment (left) [26] and the Pierre Auger Radio collaboration (right) [27] in local spherical coordinates. The left map contains 619 events, the right 36. The zenith is at the center, the north to the top and east is on the right. The direction of the magnetic field is marked with a red dot.
# 4. The Pierre Auger Observatory

The Pierre Auger Observatory (PAO) is the largest experiment for the detection of cosmic rays. It was specially designed to investigate the highest energy particles above  $10^{18}$  eV. The project developed from a suggestion by Jim Cronin and Alan Watson in 1992 is hosted by the Pierre Auger colaboration including more than 250 scientists from 17 countries.

To observe the full sky, 2 sites, one on the southern and one on the northern hemisphere have been selected. Auger North is still in the design phase and will be installed in Lamar, Colorado. Auger South is located in the Argentinian Pampa near Malargüe in the province Mendoza and is taking data since 2004. 9080 events with energies above 10<sup>19</sup> eV and al least 500000 events above 10<sup>18</sup> eV have been collected until may 2009. They are measured via two independent techniques, merging a hybrid detector. The surface detector (SD) monitors secondary particles of EAS reaching the ground. The fluorescence detector (FD) observes the shower development directly in the sky by tracking the emitted fluorescence light along the showers path.

The four FD stations of Auger South are arranged as visible in fig.4.1. They survey an area of  $\sim 3100 km^2$  filled with 1600 surface detector stations. The hybrid concept allows high precision measurements of EAS seen in both detector types. Further cross calibrations can be performed which improve the accuracy of both detection methods.

# 4.1 The Surface Detector

The surface detector consists of 1600 autonomous stations covering an area of ~  $3100 \ km^2$ . They are arranged in a hexagonal grid with a spacing of 1.6 km. These stations, a photo and a sketch is presented in fig.4.2, are basically tanks filled with 12  $m^3$  ultra pure water. A secondary particle of an EAS will emit Cherencov radiation while traversing the tank since its velocity exceeds the speed of light in water. The cherencov pulse is diffusely reflected to three photomultipliers situated at the upper side of the tank. The signal of every photomultiplier is processed and stored for a certain time in the station electronics. A timestamp provided by an internal GPS clock will be attached to the event data. If at least tree neighboring tanks receive a signal within a adequate time window, the data is transmitted wireless to a central readout station located within one of the four FD stations. They forward the signals to the central data acquisition where they are processed further. Each station is powered by batteries charged by solar panels.



Figure 4.1: Overview of Auger South near Malargue, Argentina (adopted from [28]). The red dots mark the positions of the 1600 water cherencov tanks. The blue lines show the field of view of the four 'eyes' of the FD. Additionally the positions of the extension HEAT and the planned enhancements AMIGA and AERA are marked. Note the scale on the right.

The secondary particles arising from a UHECR are distributed over tens of square kilometers when they reach ground level. A  $10^{19}$  eV shower usually triggers ten stations. For inclined showers or even higher energies the number of stations involved increases. This affects the angular and the energy resolution as both improve with more measuring points considered in the reconstruction. The design resolution is better than 1.1° in direction and 12 % in energy. For UHECRs above  $10^{20}$  eV it is expected to be better than 0.6° and 10 % respectively. For the energy reconstruction (see chapter 3.2.2), knowledge of the shower development is necessary. That information is provided by the fluorescence detector yielding a calibration of the SD.



Figure 4.2: Foto (l) and sketch (r) of an Auger surface detector station [29].



Figure 4.3: Sketch of an Auger fluorescence telescope [29].

# 4.2 The Fluorescence Detector

Each of the four FD stations shown in fig. 4.1 is composed out of 6 fluorescence telescopes. The latter have an opening angle of  $30^{\circ}$  in azimuth and  $28.6^{\circ}$  in zenith, together forming an 'eye' with  $180^{\circ}$  azimuthal coverage. The construction of a single telescope is illustrated in fig. 4.3. The fluorescence light induced by an EAS passes a filter which blocks wavelengths not according to the fluorescence emission. The

light is focused into a camera by a spherical mirror. The camera consists of 440 photomultipliers each with an angular view of  $1.5^{\circ}$ . A light trace in the sky will be projected onto a spot running over the cameras photomultiplier pixels. The signal amplitude from each photomultiplier can be translated into the energy, deposited in fluorescence light in the corresponding angular window, with an accuracy better than 5 %. The fluorescence processes in the atmosphere require proper calibration in order to achieve optimal energy resolution. The Airlight experiment [30] is dedicated to this issue and should lead to a good understanding of the emission properties. As the emitted fluorescence light is very weak, these measurements are only possible in dark moonless nights. That limits the duty cycle of the FD to about 10%.

Nevertheless, a large amount of events has been observed in both detector types simultaneously, so called 'hybrid' events. The event display of a  $6.0 \cdot 10^{19}$  eV hybrid event is offered in fig. 4.4. The shower hits the array of surface detectors almost central and triggers 19 tanks. Despite their distance to the shower axis of more than 20km, the event is observed by 3 FD stations as well.

Beside the installed detectors, tree extensions will complement Auger South. The High Elevation Auger Telescopes (HEAT) [31] were installed beside the Coihueco FD station (see fig. 4.1) and recently started operation. HEAT consists of three telescopes constructed in the same way as the FD telescopes, but they are tilted by  $30^{\circ}$  in the zenith angle. Thereby, they have an elevated field of view from  $30^{\circ}$  to  $60^{\circ}$  above the horizon. This should allow to detect showers with primary energies below  $10^{18}$  eV. Their shower maximum lies higher in the atmosphere above the field of view of the other fluorescence telescopes.

Additionally, these lower energy showers will be detected by the planned infill array AMIGA (Auger Muons and Infill for the Ground Array) [32]. It will be sited in the field of view of HEAT covering an area of 23.5  $km^2$ . A smaller spacing between the SD tanks will improve the resolution to the lateral particle distribution. Underground muon scintillator counters will investigate the lateral profile of the muon component explicitly.

The third extension called AERA (Auger Engineering Radio Array) will introduce the novel radio detection technique to the Pierre Auger Observatory. It will be discussed in detail below as it is essential for this thesis.

# 4.3 The Radio Detector

With AERA (Auger Engineering Radio Array), a great effort for a large self-triggered antenna array within the Pierre Auger Observatory has recently been decided [24]. It will be co-located with the AMIGA infill array at the western side of the observatory (marked in fig. 4.1). This site provides the unique possibility to analyze the air showers in close conjunction with Augers baseline detectors and extensions AMIGA and HEAT. Thus cross calibrations and a benchmarking of the radio setup can be performed.

Fig. 4.5 illustrates the layout of the AERA detector. It will cover an area of about 20  $km^2$  with 156 antenna stations arranged with tree different spacings. It has a



Figure 4.4: Auger hybrid event seen by 19 SD stations and in three eyes of the FD. The gray points mark running SD stations. The triggered stations are indicated by colored circles, their color referring to the timing, their size to the signal strength. The triggered pixels in the FD cameras are represented by the time-color-coded rays. The red lines are the reconstructed shower axis by the FDs, the blue by the SD. The energy was reconstructed to  $6.0 \cdot 10^{19}$  eV.

dense core of 24 stations deployed on a triangular grid with a baseline of 150 m. Around the core 60 stations with 250 m distance will be placed. The outer region consists of 72 stations with 375 m spacing. This layout allows to study the impact of different spacings on the detection efficiency. The autonomous stations will send their data to a central data acquisition located within the array.

Assuming the described layout, the expected events per year for different primary energies can be calculated. The results presented in fig. 4.6 base either on extrapolations of CODALEMA or LOPES data or on simulations with REAS2 and RDAS [34]. At high energies above  $10^{18}$  eV all three approaches predict similar event rates. For lower energies the expectations from the different approaches diverge. However, a conservative energy threshold of the array would be around  $10^{17.3}$  eV. The total number of events per year with zenith angles up to  $60^{\circ}$  can be estimated to about



Figure 4.5: Overview of the planned AERA radio detector. The red squares mark the positions of antenna stations, the yellow dots SD tanks. A central radio station (CRS) collects the data of the autonomous stations. The yellow shaded hexagons indicate the contour of AMIGA and its infill array [33].

5000.

The antenna stations used for AERA will be similar to the station presented in fig. 4.7 (left). They consist of two crossed logarithmic periodic dipole antennas, one aligned in north-south direction, the second in east-west direction. This allows to measure the north-south and the east-west component of the radio signal independently. The signals are first amplified by 20 dB by a preamplifier directly attached to the pole. The preamplifier is discussed in detail in [35]. It is powered by a Bias-T inside the electronics box (see fig. 4.7 (right)). The amplified signals pass a filter chain and are amplified again by 31 dB. A scope computes and digitizes the signals until they are send to the central data acquisition via a wireless network. Each station is powered by batteries charged by solar panels.

The station presented in fig. 4.7 is one of four stations of the actual radio setup. It is located near the Balloon Launching Station (BLS) (see fig. 4.1). Antennas, different self-trigger systems and the electronics are tested here under realistic conditions. Recently four additional antennas have been installed to enhance the testing setup.

Earlier setups also used logarithmic periodic dipole antennas of a different construction. They were cabled with a central DAQ and externally triggered by scintillators. The setup succeeded in measuring more than 300 Events in coincidence with the nearby SD tank 'Olaia' [34]. Although this setup consisted of only three antenna stations, 26 events could be reconstructed with an average angular resolution of about  $6^{\circ}$  in azimuth and zenith.

For larger radio detection approaches like AERA, angular resolutions comparable to the surface detectors accuracy are realistic. Furthermore radio detection is sensitive to the longitudinal shower development since the emitted radio pulse is a result of the whole shower evolution. Contrary to the FD, radio detection has a duty cycle



Figure 4.6: Expected events per year for the AERA radio array depending on the primary energy. The calculated event rates are based either on extrapolations of LOPES or CODALEMA data or Monte Carlo simulations with REAS2 [24].

of 100 % as it is not limited by wheather conditions. Together with the SD, a hybid detection with 100 % duty cycle could be achieved. Thus, radio detection will contribute to a better understanding of EAS and cosmic rays.

Radio detection of cosmic rays is a technical challenge. One major issue is to cope with the background noise from various sources, of both man made and natural origin. The success of a radio detector strongly depends on the used antenna as it is the first link in the detection chain. It must be sensitive to a broad frequency range and angular distribution of the incident radio pulse. On the other hand it plays an important role in handling the noise issue and thus is essential to detect preferably pure signals.

For the further treatment of the received data, a precise knowledge of the antenna characteristics is necessary. A detailed characterization of the mentioned logarithmic periodic dipole antennas is one major objective of this thesis.



Figure 4.7: Antenna station (l) and electronics (r) of the actual radio setup. The electronics are mounted in a box behind the solar panels.

# 5. Antenna and Transmission Line Theory

In principle every conducting structure can work as an antenna. Antennas convert electromagnetic waves in currents and vice versa. The response of an antenna to an electromagnetic wave depends primarily on the geometry of the structure.

Starting with the response, basic antenna properties will be introduced in this chapter. In conjunction, transmission lines are discussed since they are important to understand complex antennas. Further, simulation software for antennas is presented and the deduced antenna properties will be illustrated by example simulations.

# 5.1 Antenna Theory

Antennas can be used to transmit or receive electromagnetic waves. The properties in both cases are the same due to the principle of reciprocity. Hence it is legitimate to switch the perspectives for convenience. The space around a transmitting antenna can be classified into a near field and a far field region. The far field begins at a distance r from the antenna depending on the definition. In a conservative approach we can assume the far field for  $r > 4 \cdot \lambda$ , where  $\lambda$  is the wavelength of the emitted wave [36].

In contrast to the near field region, the electric and magnetic components are in phase in the far field region. The radiated wave can be treated as a spherical wave in the far field region. Both, the direction of propagation and the power density of the wave can than be described by the Poynting vector  $\vec{S}$ :

$$\vec{S} = \vec{E} \times \vec{H} = \frac{1}{\mu_0} \vec{E} \times \vec{B} \qquad [Wm^{-2}],$$
 (5.1)

with the electric and magnetic fields  $\vec{E}$  and  $\vec{B}$  and the permeability of free space  $\mu_0$ . It is suggestive to choose spherical coordinates with the antenna in the center. The poynting vector will then be aligned parallel to the unity vector  $e_r$ , the field components  $\vec{E}$  and  $\vec{B}$  will be contained in the  $e_{\theta}$  -  $e_{\phi}$  - plane.

## 5.1.1 Gain

If we switch to the receiving perspective, an antenna will remove power from an incident wave. The received power  $P_{rec}$  depends on an effective area of the antenna  $A_{eff}$ :

$$P_{rec} = A_{eff} \cdot |\vec{S}| = \frac{A_{eff}}{\mu_0 c} |\vec{E}|^2 = \frac{A_{eff}}{Z_{vac}} |\vec{E}|^2, \qquad (5.2)$$

with  $|\vec{B}| = |\vec{E}|/c$  and the vacuum impedance  $Z_{vac} = \mu_0 c = \sqrt{\mu_0/\epsilon_0} \sim 120\pi \ \Omega$ . The effective area depends on the frequency and the arrival direction of the incoming wave and can be expressed by a fundamental antenna property, the gain G:

$$A_{eff}(\theta,\phi) = \frac{\lambda^2}{4\pi} G(\theta,\phi).$$
(5.3)

The gain indicates the differential emitted power into a certain solid angle normalized by the power emitted by an hypothetical isotropic antenna:

$$G(\theta, \phi) = \frac{\frac{\partial P}{\partial \Omega}}{\frac{P_{acc}}{4\pi}}.$$
(5.4)

Both emitted powers refer to an equal accepted power  $P_{acc}$  at the input terminals of the antennas. An isotropic antenna would emit the accepted power equally into the complete solid angle of  $4\pi$ . As the gain depends on the spatial angles  $\theta$  and  $\phi$ , it forms a surface called gain sphere or gain pattern. The gain pattern displays the directional characteristic of the antenna.

The total radiated power  $P_{rad}$  by the antenna is given by integrating over the complete solid angle. With eqn. 5.4 it can be related to the gain:

$$P_{rad} = \int_{\Omega} \frac{\partial P}{\partial \Omega} d\Omega = \frac{P_{acc}}{4\pi} \int_{\Omega} G(\theta, \phi) d\Omega < P_{acc}.$$
 (5.5)

The total radiated power can not be larger than the accepted power. In fact, for real antennas the radiated power is always smaller than the accepted power due to structural losses which will be discussed in the next section. Hence the integral over the gain sphere is limited to a maximum value of  $4\pi$ . Consequently an antenna with a large gain in a certain direction will be less sensitive to the other directions.

## 5.1.2 Radiation Efficiency

The power applied to a real antenna will not be completely radiated due to the mentioned structural losses within the antenna structure. The power which is not radiated will mainly dissipate into heat. Structural losses are considered by the dimensionless conduction efficiency  $e_c$  and dielectric efficiency  $e_d$ . The conduction efficiency takes the ohmic resistance of the structure into account. Dielectric losses occur for instance in insulators within the antenna structure. As it is usually not possible to seperate  $e_c$  and  $e_d$  by measurements it is more convenient to summarize them with the antenna radiation efficiency  $e_{cd}$ :

$$e_{cd} = e_c \cdot e_d \qquad e_c, \ e_d \le 1. \tag{5.6}$$

With the radiation efficiency we are able to reconsider the gain. The total radiated power will be reduced compared to the accepted power due to the structural losses:

$$P_{rad} = e_{cd} P_{acc}.$$
(5.7)

Insertion into eqn. 5.4 delivers:

$$G(\theta, \phi) = e_{cd} \cdot \frac{\frac{\partial P}{\partial \Omega}}{\frac{P_{rad}}{4\pi}} =: e_{cd} \cdot D(\theta, \phi).$$
(5.8)

D is the directivity of an antenna. The difference between directivity and gain is that the directivity is normalized with the total radiated power, whereas the gain is related to the accepted power. By an integration over the whole solid angle we find the total radiated power which relates to the directivity as:

$$P_{rad} = \int_{\Omega} \frac{\partial P}{\partial \Omega} d\Omega = \frac{P_{rad}}{4\pi} \int_{\Omega} D(\theta, \phi) d\Omega.$$
 (5.9)

The integral over the directivity sphere thus has always a value of  $4\pi$ . Hence the directivity gives the pure dependence of the emitted power on the direction of the emission. The gain also considers the frequency dependent structural losses. Without any structural losses, the accepted power is completely radiated by the antenna. Eqn. 5.8 implies that the gain is equal to the directivity in this case. Furthermore it implies that both patterns have the same shape, only the size of the surface of the gain sphere varies with the radiation efficiency.

Structural losses can be minimized by using good conductors and by avoiding dielectrica within the antenna structure. The losses are then usually in the order of a few percent compared to the accepted power if the antenna works within its operational range. Precise values for the structural losses can be obtained through antenna simulations.

#### 5.1.3 Reflection Coefficient

So far we considered losses only within the antenna structure. Fig. 5.1 gives an overview of the different losses and their location with respect to the antenna structure. Additional losses may occur at the footpoint<sup>1</sup> of the antenna when it is connected to a transmission line or a source.

Both, the footpoint of the antenna and the connected source can be described by a characteristic impedance. We can now consider a signal delivered to the antenna's footpoint by a source with a certain power  $P_{in}$ . If the footpoint impedance does not match the impedance of the connected source, only a fraction of the power delivered by the source will be accepted by the antenna. The other fraction  $P_{ref}$  will be reflected at the footpoint back to the source yielding a loss for the antenna. As the footpoint impedance varies with the frequency of the input signal, the amount of the reflected power is also frequency dependent. This behavior is described by the reflection coefficient.

The reflection coefficient R is defined<sup>2</sup> as ratio of the input  $V_{in}$  and reflected voltage ampitude  $V_{ref}$ :

$$R = \frac{V_{ref}}{V_{in}} = \frac{\sqrt{P_{ref}}}{\sqrt{P_{in}}} = \frac{\sqrt{P_{in} - P_{acc}}}{\sqrt{P_{in}}}.$$
(5.10)

<sup>&</sup>lt;sup>1</sup>The footpoint of an antenna indicates the terminal where the signal is read out.

<sup>&</sup>lt;sup>2</sup>There are different definitions for the reflection coefficient. Another common definition is to compare powers directly:  $R = P_{ref}/P_{in}$ . This leads to: R + T = 1.



Figure 5.1: Antenna losses and their position with respect to the antenna structure. Structural losses  $e_c$  and  $e_d$  occur within the antenna structure whereas reflection losses R evolve at the footpoint. The reference powers for directivity, gain and absolute gain are denoted at the corresponding terminals (Based on [37]).

The reflected voltage amplitude depending on the frequency can be measured directly at the footpoint and is an important quantity to characterize antennas. The according transmission coefficient is then given by:

$$T = \frac{\sqrt{P_{acc}}}{\sqrt{P_{in}}} = \frac{\sqrt{P_{in} - P_{ref}}}{\sqrt{P_{in}}} = \sqrt{1 - R^2}.$$
 (5.11)

The reflection is commonly expressed in the decibel scale or as voltage standing wave ratio (VSWR). The decibel scale is a logarithmic scale used to describe power ratios. The reflection in dB,  $R_{dB}$  can be translated into the reflection coefficient R via:

$$R_{dB} = 10 \cdot \log\left(\frac{P_{ref}}{P_{in}}\right) = 20 \cdot \log\left(\frac{V_{ref}}{V_{in}}\right) = 10 \cdot \log(R^2).$$
(5.12)

The VSWR is defined as:

$$VSWR = \frac{V_{in} + V_{ref}}{V_{in} - V_{ref}} = \frac{1+R}{1-R}.$$
(5.13)

The reflection at the footpoint of the antenna can be related to the gain. Insertion of eqn. 5.10 into eqn. 5.4 delivers:

$$G(\theta,\phi) = \frac{\frac{\partial P}{\partial \Omega}}{\frac{P_{in}(1-R^2)}{4\pi}} = \frac{1}{(1-R^2)} \frac{\frac{\partial P}{\partial \Omega}}{\frac{P_{in}}{4\pi}} =: \frac{1}{(1-R^2)} G_{abs}(\theta,\phi).$$
(5.14)

 $G_{abs}$  is the absolute gain of an antenna. The absolute gain is normalized by the input power  $P_{in}$  delivered by a source. All possible losses, the reflection at the footpoint as well as the structural losses are included within the absolute gain. It relates to the previous defined gain and directivity as:

$$G_{abs}(\theta,\phi) = (1-R^2) \ G(\theta,\phi) = e_{cd}(1-R^2) \ D(\theta,\phi).$$
(5.15)

The different reference powers for directivity, gain and absolute gain are indicated in fig.5.1.

As the reflection coefficient and the absolute gain depend on the characteristic of the devices connected to the antenna, they are strictly speaking no intrinsic antenna properties. Nevertheless they are of great interest as they describe antenna properties under operating conditions.

## 5.1.4 Polarization

The polarization of the incident wave has also an influence on the response of the antenna. It is intuitive that a horizontally polarized electric field of a wave cannot accelerate the electrons in a vertically aligned dipole. The wave will only induce currents in the dipole if the electric field vector has a component along the dipole. In spherical coordinates, the electric field vector can be split in a polarization in the  $e_{\theta}$ -direction and in the  $e_{\phi}$ -direction:

$$\vec{E} = E_{\theta}e_{\theta} + E_{\phi}e_{\phi}, \qquad (5.16)$$

with the unit vectors in spherical coordinates  $e_{\theta}$  and  $e_{\phi}$ . The radial component  $e_r$  will vanish for an antenna in the center of the coordinate system. The total gain can be defined as the sum of the gain for the  $e_{\theta}$  polarization and the  $e_{\phi}$  polarization:

$$G = G_{\theta} + G_{\phi}.\tag{5.17}$$

Both will yield a received power via eqn. 5.2 and eqn. 5.3:

$$P_{\theta} = \frac{\lambda^2}{4\pi Z_{vac}} G_{\theta} E_{\theta}^2, \qquad P_{\phi} = \frac{\lambda^2}{4\pi Z_{vac}} G_{\phi} E_{\phi}^2. \qquad (5.18)$$

They will sum to the total received power in the antenna:

$$P_{rec} = P_{\theta} + P_{\phi} = \frac{\lambda^2}{4\pi Z_0} (G_{\theta} E_{\theta}{}^2 + G_{\phi} E_{\phi}{}^2).$$
(5.19)

# 5.2 Transmission Line Theory

## 5.2.1 Transmission Lines

To transfer signals in the radio frequency regime (1-100 MHz) preferably lossless, special cables are needed. There are two common types of radio frequency (RF) cables. The coaxial cable and parallel conductors, also called twin leads or Lecher lines.

For low frequencies, the behavior of electromagnetic fields in wires can be described by the resulting voltage, current and the ohmic resistance of the wire. For frequencies in the radio regime, this description is not sufficient. For the properties of RF wires, the wave nature of the electromagnetic fields becomes important. Therefore RF wires are called waveguides or transmission lines. Waves propagating along a transmission line can be reflected and may interfere with each other. This behavior is described



Figure 5.2: Equivalent circuit of a lossless transmission line.

by a homogenic linear differential equation with solutions being superpositions of incoming and returning waves. The ratio of voltage to current for these superimposed waves is found to be equal to a characteristic impedance  $Z_0$  of the transmission line. Finally the characteristic impedance is defined in analogy to a resistance in Ohm's law.

A lossless<sup>3</sup> transmission line can be approximated by a network of infinitesimal capacities and impedances as visible in fig.5.2. The characteristic impedance can be defined with a capacitance per unit length C' and an inductance per unit length L':

$$Z_0 = \sqrt{L'/C'}.$$
 (5.20)

Note that these impedance is a pure resistance with the unit  $\Omega$ , although it results from capacitances and inductances.

A transmission line terminated with a load equal to its characteristic impedance will transfer an applied pulse to the termination without reflection. In that case all the power in the signal is transferred to the load. The transmission line is called 'matched' with the load. If the load's impedance does not match the characteristic impedance of the line, a significant fraction of power can be reflected. This impedance matching is a general issue in RF-techniques and must be considered for all RF devices such as antennas and measuring instruments. It will be discussed in a general way in section 5.2.3.

A coaxial cable consists of an inner conductor embedded in a solid dielectric. This is finally surrounded by a flexible outer conductor. The energy is transported as a transverse electric and magnetic mode (TEM) of the wave, running between the inner and the outer conductor only within the dielectric. That implies that no energy is radiated to the outside and that there is no pickup of external signals - the cable is completely shielded. The dielectric usually has a high dielectric constant (common coaxial cables use Polyethylene with  $\epsilon_r = 2.3$ ) which decreases the propagation velocity v of the wave to:

$$v = c/\sqrt{\epsilon_r}.\tag{5.21}$$

For the mentioned cable with Polyethylene dielectric the signal velocity is then v = 2/3c. For the given geometry, the characteristic impedance of a coaxial cable can be calculated to be:

$$Z_0 = \sqrt{L'/C'} = \frac{60}{\sqrt{\epsilon_r}} \cdot \ln\frac{D}{d}, \qquad (5.22)$$

<sup>&</sup>lt;sup>3</sup>lossless means that the ohmic resistance can be neglected

depending on the diameter d of the inner and of the outer conductor D. The signal attenuation in a coaxial cable is caused by the ohmic resistance of the conductors and by losses in the dielectric. Both increase with the signal frequency. The properties of the widely used RG58 C/U coaxial cable can be taken from appendix A.1. Coaxial cables are usually operated as an unbalanced transmission line which means that the outer conductor is grounded.

Twin leads or Lecher lines consist of two parallel wires placed in a certain distance. Unlike the coaxial cables, they are operated in a balanced mode with both wires being on the same electric potential. The characteristic impedance is given by:

$$Z_0 = \sqrt{L'/C'} = \frac{120}{\sqrt{\epsilon_r}} \cdot \operatorname{arcosh} \frac{D}{d}$$
(5.23)

with the distance D between the two conductors of diameter d. Parallel conductors can be build only with air insulation between the two wires ( $\epsilon_r = 1$ ). The signal speed in these wires is then close to the speed of light,  $v \sim c$ . The signal attenuation is lower compared to coaxial cables as there are no losses in the dielectric. The disadvantage of parallel conductors is that they are not completely shielded and thus can pickup external signals.

## 5.2.2 Scattering Parameters

Scattering parameters (S-parameters) are used to express electrical properties of networks or components. In the context of S-parameters, scattering refers to the way in which the traveling currents and voltages in a transmission line are affected when they meet a discontinuity caused by the insertion of a network into the transmission line. This is equivalent to the wave meeting an impedance differing from the line's characteristic impedance. For instance, we can imagine the connection of a coaxial cable with  $Z_0 = 50 \ \Omega$  to an antenna with a differing footpoint impedance  $Z_A \neq 50 \ \Omega$ as a two port network. The reflection and transmission for that connection is then described by the S-parameters. Fig.5.3 illustrates the four S-parameters of a two port network which are:

- $\underline{S}_{11}$  is the input port voltage reflection coefficient
- $\underline{S}_{12}$  is the reverse voltage transmission
- $\underline{S}_{21}$  is the forward voltage transmission
- $\underline{S}_{22}$  is the output port voltage reflection coefficient

Each parameter is a ratio of complex voltages and thus is itself a complex number<sup>4</sup>. Furthermore each parameter is frequency dependent.  $\underline{S}_{11}$  and  $\underline{S}_{22}$  are usually expressed in so called Smith charts, where the real part is plotted against the imaginary part of the complex S-parameter for various frequencies. A smith chart is discussed in detail in chapter 5.3.

The  $\underline{S}_{11}$  parameter is of great interest for the further discussion. The  $\underline{S}_{11}$  parameter

<sup>&</sup>lt;sup>4</sup>from now on all complex values are indicated by an underline



Figure 5.3: S-Parameters of a two port network

for a single antenna is the ratio of the complex reflected voltage to the input voltage:

$$\underline{S}_{11} = \frac{\underline{V}_{ref}}{\underline{V}_{in}} \tag{5.24}$$

In chapter 5.1.3 we defined the reflection coefficient for an antenna as:

$$R = \frac{V_{ref}}{V_{in}} = \frac{|\underline{V}_{ref}|}{|\underline{V}_{in}|} = \left|\frac{\underline{V}_{ref}}{\underline{V}_{in}}\right| = |\underline{S}_{11}| = |\underline{r}|.$$
(5.25)

Thus the reflection coefficient for an antenna is equal to the absolute value of  $\underline{S}_{11}$ .  $\underline{S}_{11}$  describes the complex reflection of the antenna, including the phase information. Despite a coaxial cable with a constant resistive impedance of e.g.  $Z_0 = 50 \ \Omega$ , an antenna has a frequency dependent reactive footpoint impedance. It can be observed by measuring the complex reflection coefficient  $\underline{r} \ (= \underline{S}_{11})$  at the footpoint using a vector network analyzer. The complex reflection coefficient can be translated into the footpoint impedance via the following consideration. Fig. 5.4 shows an equivalent circuit of the measuring setup. The vector network analyzer has a pure resistive source impedance of  $Z_0 = 50 \ \Omega$  and delivers an input voltage  $\underline{V}_{in}$  and an input current  $\underline{I}_{in}$  to the load. In our case the load is an antenna with its footpoint impedance  $\underline{Z}_L$ . A fraction  $\underline{V}_L$  of the input voltage drops over the load, the other fraction  $\underline{V}_{ref}$  will be reflected to the source where it is measured by the network analyzer over the source impedance  $Z_0$ :

$$\underline{V}_L = \underline{V}_{in} + \underline{V}_{ref}, \qquad (5.26)$$

$$\underline{V}_{ref} = Z_0 \cdot \underline{I}_{ref}. \tag{5.27}$$

The reflected current  $\underline{I}_{ref}$  is given by:

$$\underline{I}_L = \underline{I}_{in} - \underline{I}_{ref}.$$
(5.28)

Further is:

$$\underline{V}_{in} = Z_0 \cdot \underline{I}_{in}, \qquad \underline{V}_L = \underline{Z}_L \cdot \underline{I}_L.$$
(5.29)

A combination of these equations (5.29 in 5.26, 5.28, 5.29 & 5.27) leads to:

$$\underline{V}_{in} + \underline{V}_{ref} = \underline{V}_L = \underline{Z}_L \cdot \underline{I}_L \tag{5.30}$$

$$= \underline{Z}_L \left( \frac{\underline{V}_{in}}{Z_0} - \frac{\underline{V}_{ref}}{Z_0} \right)$$
(5.31)

$$\iff \underline{V}_{in} \left( 1 - \frac{\underline{Z}_L}{Z_0} \right) = -\underline{V}_{ref} \left( 1 + \frac{\underline{Z}_L}{Z_0} \right).$$
(5.32)



Figure 5.4: Equal circuit of a reflection measurement of a complex load impedance  $\underline{Z}_L$ . The source impedance (e.g. a network analyzer) is  $\underline{Z}_0$ .

The complex reflection coefficient  $\underline{r}$  can then be found to be:

$$\underline{r} = \frac{\underline{V}_{ref}}{\underline{V}_{in}} = -\frac{1 - \frac{\underline{Z}_L}{Z_0}}{1 + \frac{\underline{Z}_L}{Z_0}} = \frac{\underline{Z}_L - Z_0}{\underline{Z}_L + Z_0}.$$
(5.33)

Hence, we can determine the footpoint impedance of the antenna or any other load impedance with a measurement of the magnitude  $|\underline{r}|$  and the phase  $\Phi$  of the complex reflection coefficient  $\underline{r}$ :

$$\underline{Z}_L = Z_0 \frac{1+\underline{r}}{1-\underline{r}}, \qquad \underline{r} = |\underline{r}| \cdot exp(i\Phi) = R \cdot exp(i\Phi). \tag{5.34}$$

#### 5.2.3 Impedance Matching

We figured out that every mismatched connection between RF-devices such as cables, measuring instruments and antennas cause reflections and therewith a loss of power. Measuring devices and cables are standardized to avoid this issue. A common standard is the 50  $\Omega$  system.

A priori, antennas and other RF-devices can have any impedance. They have to be tuned by impedance matching components to the impedance of the used system as far as possible. For narrow band applications this can be done e.g. by networks of capacitances and inductances. For broadband applications, so called transmission line transformers (TLT) are common. They consist of transmission lines wound on a ferrite core, suitably interconnected. TLTs provide various impedance transformations with excellent broadband performance. Special TLTs, named baluns (balanced - unbalanced) additionally allow to transform between unbalanced lines (e.g. coaxial cables) to balanced transmission lines (e.g. a Lecher line of an antenna). The circuit of a 4:1 Guanella balun is presented in Fig.5.5. The term 4:1 refers to the impedance transforming ratio (e.g from 200  $\Omega$  to 50  $\Omega$ ), Guanella to its inventor. The two transmission lines are connected in parallel at the low impedance side and in series at the high end. This leads to an transforming ratio of 4:1 [38].



Figure 5.5: Circuit of an 4:1 Guanella Balun

# 5.3 Antenna Simulations

The measurement of antenna properties like gain patterns are only possible within great efforts. Thus antenna simulations provide an important tool to develop antenna designs.

In this thesis a common simulation program called Nec2 is used. The code was developed by G.J. Burke and A.J. Poggio in 1977 [39]. First the user has to model an antenna structure which is then divided into segments. The current in each segment is calculated for a given voltage source at the footpoint of the antenna model. The radiated electric field is than specified by integrating over the current distributions for each segment. The resulting integral equations are solved numerically, using a technique called Method of Moments. Finally Nec2 generates an output file which describes the properties of the radiated electric fields.

These properties can be displayed with a graphical user interfaces available for Nec2, for instance 4Nec2 [40]. 4Nec2 also offers a convenient way to construct antenna models and to edit environment parameters for the simulation.

To demonstrate the possibilities of 4Nec2 and to illustrate the antenna properties introduced in the previous chapters, a simulation of a dipole antenna will be discussed in detail in the following. The  $\lambda/2$ -dipole is probably the most simple antenna, but is nevertheless of great interest as it is the basic element from which many complex antennas like the mentioned logarithmic periodic dipole antenna are constructed.

#### Gain

4Nec2 delivers three values for the gain. The vertical gain corresponds to inclining waves polarized in the  $e_{\theta}$ -direction, the horizontal gain to the  $e_{\phi}$ -polarization. The naming fits with our definition from chapter 5.1.3 for waves propagating parallel to the ground. They both add up to the third value, the total gain via eqn. 5.17.

Fig. 5.6 shows the gain pattern for waves polarized either in  $e_{\phi}$ -direction (left) or in the  $e_{\theta}$ -direction (right) for a  $\lambda/2$ -dipole. The patterns are simulated in free space at 30 MHz. The antenna is modeled with perfect conductors which means that no structural losses are considered. Furthermore the gain patterns are equal to the directivity patterns in this case (see. chapter 5.1.2). The dipole with a length of 5m is aligned along the x-axis of the cartesian coordinate system. The value of the



Figure 5.6: Horizontal (left) and vertical (right) gain pattern of a 5 m  $\lambda/2$ -dipole simulated with 4Nec2 at 30 MHz. The horizontal pattern refers to incident waves polarized in  $e_{\phi}$ -direction, the vertival pattern to waves polarized in  $e_{\phi}$ -direction.

gain is the distance from the point of origin of this coordinate system to the surface for any direction. It is given in a logarithmic decibel scale normalized to the gain of an isotropic antenna, according to the definition in eqn. 5.4. This unit is called dBi, where i stands for isotropic. Additionally the values are color coded.

For a wave polarized in the  $e_{\phi}$ -direction the maximum gain is reached for any arrival



Figure 5.7: Simulated total gain pattern (l) and directional diagram of a dipole (r). The total gain pattern is the sum of the horizontal and the vertical pattern. The directional diagram is a cut through the total pattern and indicates the beamwidth which equals to  $76^{\circ}$  in this case.

direction perpendicular to the dipoles axis. The electric field vector is then parallel to the dipole and induces the maximum current. With decreasing angle to the dipoles axis, the gain is reduced until, for arrival directions parallel to the dipoles axis, the antenna gets completely insensitive.

A different pattern evolves for waves polarized along  $e_{\theta}$ . In this case, the antenna will be blind for any arrival directions contained in the x-y-plane. It will also be insensitive for directions lying in the y-z-plane, except waves propagating exactly along the z-axis. Due to the spherical coordinates, both gain patterns refer to the same polarization of the wave for this direction.

The sum of the horizontal and the vertical pattern is the total gain presented in fig.5.7. It has the typically toroidal shape of a radiating dipole. The gain has a maximum of 2.17 dBi which fits the common value of 2.15 dBi for thin  $\lambda/2$ -dipoles [36]. The diagram in Fig.5.7 (right) is a cut along the x-z-plane through the total gain pattern and is called directional diagram. It is normalized to the maximum gain. The two lobes are called major lobe and minor lobe<sup>5</sup> and indicate the directivity of the antenna.

To define a quantity of directivity, the so called beamwidth can be specified from these diagrams. The black lines are determined by the intercept points of the major lobe with the -3dB circle. At these points the emitted power is reduced to approximately half the maximum value<sup>6</sup>. The angle between these lines is defined as the beamwidth. In this case we have a beamwidth of 76° which is close to a value of 78° given in [36].

#### **Reflection Coefficient**

4Nec2 also provides the reflection coefficient for every desired frequency. The user has to select a characteristic impedance  $Z_0$  to which the reflection will be related.  $Z_0$  corresponds to the impedance of the device to which the antenna is connected, e.g. a coaxial cable with 50  $\Omega$ . 4Nec2 calculates the complex impedance  $\underline{Z}_L$  of the antenna and converts it into the reflection using eqn. 5.33.

The magnitude of the reflection is displayed in fig. 5.8 for a characteristic impedance of 50  $\Omega$  in the decibel scale. The curve has a resonance dip around a frequency matching to the dipole length. This is typical for a dipole. At this frequency the reflection is very small as the antenna is able to radiate almost the total input power. Below and above this resonance, the input frequency does not match the antenna geometry and thus most of the power is reflected to the source. For a ideal  $\lambda/2$ dipole we would expect a resonance for  $\nu = c/\lambda = c/10$  m = 29.97 MHz.

For a real dipole, the resonance frequency is slightly lower. In real dipoles the waves propagate with a velocity lower than the speed of light. The propagation velocity depends in a non linear way on the ratio of the length of the dipole to its diameter. This causes the so called shortening of the resonance length. It can be considered

<sup>&</sup>lt;sup>5</sup>In this case the major lobe is identical to the minor lobe. Directive antennas usually have a distinctive major lobe pointing to the direction of maximum sensitivity and smaller minor lobes to other directions.

<sup>&</sup>lt;sup>6</sup>A power ratio in dB is given by  $P[dB] = 10log(P/P_{reference})$ . Hence -3 dB corresponds to 0,5012  $P/P_{reference}$ 



Figure 5.8: Simulated reflection coefficient in the decibel scale of a  $\lambda/2$ -dipole for frequencies from 20 to 40 MHz.

with the shortening factor V. The shortening factor for  $\lambda/2$ -dipoles depending on the ratio of the dipole length to its diameter d can be taken from appendix A.2 [36]. With the shortening factor we can estimate the resonance frequency of a  $\lambda/2$ -dipole with length l to:

$$\nu_{res} = \frac{c}{2l} \cdot V(l/d). \tag{5.35}$$

The simulated dipole with a length of 5 m and a diameter of 1 mm has a shortening factor of  $V \sim 0.97$  according to A.2. The resonance frequency can then be calculated to be  $\nu_{res} = 29.08$  MHz which fits the center frequency of the resonance dip in fig.5.8.

The width of the resonance dip of the reflection curve defines another important antenna property, the bandwidth of an antenna. It indicates the frequency range to which an antenna is sensitive to. This range is given as the difference of a lower frequency  $\nu_l$  and an upper frequency  $\nu_u$  which correspond to a certain threshold of reflection. The threshold is commonly set to a reflection coefficient of 1/3 which is equivalent to a value of -9.54 dB in the decibel scale and to a VSWR<sup>7</sup> of 2. Thus we can write the bandwidth *BW* as:

$$BW = \nu_u (VSWR = 2) - \nu_l (VSWR = 2). \tag{5.36}$$

For the simulated dipole we can identify:  $\nu_u(VSWR = 2) = 28.2$  MHz and  $\nu_l(VSWR = 2) = 29.7$  MHz in fig. 5.8. This results in a bandwidth of 1.5 MHz. However, the definition of the bandwidth depends on the needs of the particular ap-

plication. In [37] the bandwidth is defined as: 'The range of frequencies within which

 $<sup>^7\</sup>mathrm{The}$  Voltage Standing Wave Ratio (VSWR) was defined in chapter 5.1.2

the performance of the antenna, with respect to some characteristic, conforms to a specified standard'. This standard can be specified through the mentioned threshold of reflection depending on the application. The previous definition with a threshold of R = 1/3 is common if the antenna is used as a transmitter. Transmitting applications require low reflections as the reflected signals may overload the connected transceiver.

For receiving applications, a larger reflection might be acceptable. In those cases it is often sufficient if less than half the received power is reflected which corresponds to a threshold of  $R = \sqrt{1/2}$  or R = -3 dB respectively. For the simulated dipole this would yield a larger bandwidth of approximately 5.0 MHz according to fig.5.8.

#### **Complex Reflection and Impedance**

The complex reflection coefficient  $\underline{r} = R \cdot exp(i\phi)$  is usually expressed in a Smith chart. Smith charts offer a convenient way to investigate impedances and impedance matching in transmission line theory.

Complex numbers like impedances are commonly illustrated in a cartesian coordinate system. By the conformal map:

$$\underline{Z}_L \to \underline{r} = \frac{\underline{Z}_L - Z_0}{\underline{Z}_L + Z_0},\tag{5.37}$$

the cartesian coordinate system can be projected onto a unit circle. This conformal map is inspired by the definition of the reflection coefficient (see eqn. 5.33). The result is the Smith chart. The mapping implies a normalization to the characteristic impedance  $Z_0$  which is in the center of the Smith chart. Inductances lie in the upper sphere, capacitances in the lower sphere and pure resistances on the horizontal axis. Lines of equal real part become circles and lines of equal imaginary part become arcs with different radii. Lines of equal reflection are concentric circles around the center where the reflection is R = 0. The outer circle indicates R = 1. The phase is counted counterclockwise starting from the right. The frequency is color coded.

The complex reflection of the simulated dipole describes a circular trace in the smith chart. The complex footpoint impedance  $\underline{Z}_L$  can be read off directly from the Smith chart for the observed frequencies. The trace starts at 10 MHz in the right corner on the outer circle. That means that the input power is completely reflected by the antenna (see also fig. 5.8). The footpoint impedance has a value of  $\underline{Z}_L = 5.7 - i1432 \ \Omega$  at this frequency. Due to the high capacitive reactance, the power is mainly transformed into a reactive power.

With increasing frequency the reactive component gets smaller until it vanishes for a frequency of 29.0 MHz. Here the footpoint impedance is pure resistant,  $\underline{Z}_L = 72.0 \Omega$ . This is also the frequency of minimal reflection as the distance to the center of the Smith chart is minimal for this point. A comparison with fig.5.8 shows that this is the resonance frequency of the dipole. The reflection coefficient has a value of R = 0.18 at the resonance, which corresponds to R = -14.9 dB in the decibel scale. This can also be observed in fig.5.8. The simulated footpoint impedance of  $\underline{Z}_L = 72.0 \Omega$  is typical for a  $\lambda/2$ -dipole at its resonance. For comparison, a value of  $\underline{Z}_L = 67 \Omega$  is given in [36].

For higher frequencies, the footpoint impedance becomes a capacitive reactance and the reflection increases. At 50 MHz the footpoint impedance is:  $\underline{Z}_L = 1074 - i1394 \Omega$ , which means that it is strongly mismatched to the characteristic impedance of 50  $\Omega$ .



Figure 5.9: Simulated complex reflection of a dipole for frequencies from 10 MHz to 50 MHz displayed in a Smith chart. The red sqare marks the resonance frequency of 29 MHz.

# 6. Conception and Design of an Antenna for the Radio Detection of EAS

Radio detection of cosmic rays induced air showers makes great demands on the used antenna. These demands are found to be fullfilled to a large extend by the mentioned logarithmic periodic dipole antenna (LPDA). The principle and the conception of an LPDA for this special purpose will be discussed. The hardware design will be compared for different types of LPDAs.

# 6.1 Demands on an antenna for the radio detection of EAS

The demands on an antenna for the detection of cosmic rays result from three different factors: the properties of the emitted radio pulses, the conditions at the site of the experiment and the resource intensity for the production of the antennas.

The properties of the emitted radio pulses have been discussed in detail in chapter 3.3. An important result was the predicted frequency spectrum of the pulse summarized in fig. 3.4. A priori the pulses consist of frequencies from a few MHz to several hundred MHz. With increasing distance to the shower axis the frequency spectra shifts to the lower frequencies. Thus the antenna should be sensitive to this low frequencies to detect most distant showers. The maximum distance to which a shower can be detected directly determines the number of antenna stations needed to cover a certain area and thus the costs of the experiment. On the other hand, every observed frequency contributes to the total received power in the antenna and thus a broad band sensivity is preferred.

Unfortunately noise sources constrain the usable frequency range for the radio detection. Fig. 6.1 shows a measurement of the radio background at the Auger site. Although the environment of the site is very sparsely populated, there are some man made radio sources. The strong peaks below 20 MHz result from short wave transmitters. They are omnipresent as the short waves are reflected by the atmosphere and can be received in distances of thousands of kilometers away from their source. Above 85 MHz local radio stations broadcast at about ten different frequencies. The spectrum between 20 MHz and 85 MHz results from the omnipresent galactic radio background [41] folded with the antenna characteristics. This will be discussed in



Figure 6.1: Radio background at the Auger site. The spectrum was recorded with a resulution bandwidth of 100 kHz. The noise floor of the spectrum analyzer lies at -103 dB for this setting.

detail in chapter 7. Continuous measurements of the radio background at the Auger site have been performed in [42]. Here, additionally transmitters of less power in the range from 20 MHz to 30 MHz were reported.

Finally, in a conservative approach, a frequency window from 30 to 85 MHz remains for the observation of radio pulses from cosmic ray air showers at the Auger site. Thus the antenna should be tuned to this frequency range.

An important criteria to successfully detect radio pulses from EAS is the signal to noise ratio of the measured events. The noise of the measurement setup has to be reduced as far as possible to be able to separate even weak signals from the noise background. The noise of the measurement setup results from the noise of the electronics, the antenna and the mentioned irreducible galactic noise. The noise of the antenna is mainly due to the white noise in the conductors of the antenna. White noise evolves from the thermal agitation of the electrons in a circuit. The power of the white noise  $P_{wn}$  depends on the temperature T and the observed bandwidth BW:

$$P_{wn} = k_B T \cdot BW. \tag{6.1}$$

Thus it is irreducible for our purpose as the bandwidth should not be reduced. However, the fact that the galactic noise can be observed, e.g. in fig. 6.1, implies that the noise of the measurement setup must be below the galactic noise. This

can be formulated as a requirement for the antenna: the antenna should be able to observe the galactic noise as its intrinsic noise must be sufficiently low in this case. Furthermore the antenna must have a sufficient sensitivity as the galactic noise is irreducible.

Radio pulses should preferably be detectable from air showers with any arrival directions. Therefore the antenna has to be sensitive to 360° in azimuth and 90° in zenith angle. On the other hand most of the man made noise comes from the very horizontal directions. Hence it is favourably to surpress signals from that directions with the antenna characteristics. Also the electronics in the antenna stations cause noise in the radio regime. This can be limited by an antenna being mostly unsensitive to the directions from the ground.

For a precise reconstruction of cosmic rays properties from the measured radio pulses, the underlying antenna properties should be mostly independent of the observation conditions. In particular the gain pattern must be stable over the interesting frequency range as far as possible. Beside the frequency, the ground below the antenna has a strong influence on the gain pattern. Incoming waves are reflected on the ground causing a transformation of the gain sphere. This transformation depends on the height of the antenna above the ground and on the ground conditions which change with the weather. An antenna characteristic being mostly independent of these changing conditions is preferred.

The weather conditions at the argentinian pampa alternate from hot dry summers to cold snowy winters. An antenna has to cope with temperatures from -20 to  $50^{\circ}$ C, it has to be water and frost proof and resistant against UV-radiation. Probably the biggest issue are extreme wind speeds up to 150 km/h at the site. The antenna exposed on a pole in several meters high has to withstand the permanent oscillations induced by the wind. These oscillations can be reduced by a preferably small windload and a low mass of the antenna. The antenna must survive these conditions without any damage for years as maintenance is extensive in this environment.

To establish the radio detection technique on large scales, the used antennas have to be produced in large quantities. For this purpose, the material costs and the production time have to be optimized, keeping a constant production quality in mind. Furthermore the antennas have to be shipped to the site. The costs for the transportation scale mainly with the volume. The volume of the antennas can be drastically reduced e.g. by integrated folding mechanisms. Finally they have to be manually installed on poles of several meters hight, which makes a light antenna favourably.

# 6.2 Logarithmic Periodic Dipole Antennas

Logarithmic periodic dipole antennas are a suitable choice for the discussed demands. They belong to the category of logarithmic periodic structures. These antenna types consist of repetitive elements with element sizes following a geometric series. In the receiving case, an input signal with a certain frequency will traverse from the footpoint along the structure until it finds an element with a resonance frequency matching the input frequency. At this point the power is radiated. With changing frequency the resonant region moves along the structure whereas the geometric ratios cause approximately equal antenna characteristics for the different frequencies. The gain pattern changes only slightly with the frequency compared to other types of broad band antennas. Furthermore the antennas footpoint impedance stays stable within certain limits over the whole bandwidth. This behavior allows to match the antenna's impedance to a single characteristic impedance.

## 6.2.1 Principle and Design Parameters of LPDAs

In case of a LPDA, the resonant elements are  $\lambda/2$ -dipoles of different lengths. Fig. 6.2 illustrates the construction of an LPDA consisting of n dipoles. The dipoles are connected to a central waveguide in an alternating way. The waveguide is a balanced Lecher line which was discussed in section 5.2.1. The footpoint of the antenna is located at the upper end of the Lecher line close to the shortest dipole. The lower end of the waveguide is short-circuited by the so called stub.

The power of an incoming monochromatic wave will be primarily absorbed by the dipole with a length  $l_i$  of about half the incident wavelength. In chapter 5.3 we found out that even a single dipole has a not vanishing bandwidth. Therefore also the neighboring dipoles will absorb a significant amount of power from the wave. If we assume an incoming broad band radio pulse e.g. from an EAS, every dipole will feed the waveguide with a different wavelenght  $\lambda_i$ . The signals from the single dipoles should interfere preferably constructively at the footpoint to gain a maximum total amplitude. This is realized by a constant ratio of the wavelengths to the distances  $R_i$  from the corresponding dipole to a virtual footpoint:

$$\frac{\lambda_i}{R_i} = const. \qquad i = 1...n. \tag{6.2}$$

This is fullfilled for constant ratios of sequenced distances and wavelengths:

$$\frac{R_i}{R_{i-1}} = \frac{\lambda_i}{\lambda_{i-1}} =: \tau \qquad \Rightarrow \frac{\lambda_i}{R_i} = \frac{\lambda_{i-1}}{R_{i-1}}.$$
(6.3)

The design parameter  $\tau$  sets the ratio of the lengths of sequenced dipoles if we assume  $\lambda_i \propto l_i$ . We can calculate the dipole lengths  $l_i$  and the distances  $R_i$  successive from the resulting geometric series:

$$l_i = l_1 \tau^{i-1}, \qquad R_i = R_1 \tau^{i-1}, \qquad i = 1...n,$$
(6.4)

whereas  $l_1$  and  $R_1$  are the initial values. These equations imply that the antenna structure becomes periodic in the logarithm of both the dipole length and the distances from the virtual footpoint. With given distances  $R_i$  the spacing  $S_i$  between the dipoles follows:

$$S_i = R_i - R_{i-1} = R_i(1-\tau).$$
(6.5)

The ratio of the spacings to their corresponding dipole lengths is also constant:

$$\frac{S_i}{\lambda_i} \approx \frac{S_i}{2l_i} = const. =: \sigma.$$
(6.6)



Figure 6.2: Principle construction of a LPDA (adapted from [36]).

It defines the second design parameter  $\sigma$  which indicates the relative spacing between the dipoles. The approximation is due to the shortening factor described in chapter 5.3 which is not considered here. The parameter  $\sigma$  relates to  $\tau$  and the opening angle  $\alpha$  of the structure via:

$$\sigma = \frac{1-\tau}{4} \cot\left(\frac{\alpha}{2}\right). \tag{6.7}$$

With the two design parameters  $\sigma$  and  $\tau$  and the choice of the operational frequency range, the complete antenna design is described. The operational frequency range from  $\nu_{min}$  to  $\nu_{max}$  is configured by the length of the shortest and the longest dipole via:

$$l_{max} \approx \frac{\lambda_{max}}{2} \qquad \Rightarrow \qquad \nu_{max} \approx \frac{c}{2l_{max}},$$
 (6.8)

$$l_{min} \approx \frac{\lambda_{min}}{3} \qquad \Rightarrow \qquad \nu_{min} \approx \frac{c}{3l_{min}}.$$
 (6.9)

The design parameters can then be choosen within reasonable limits.  $\tau$  affects the number of dipoles between  $l_{min}$  and  $l_{max}$ . More dipoles in this range will cause a more constant sensitivity as the resonances of sequenced dipoles will overlap to a larger extend. Assuming a constant relative spacing  $\sigma$  the antenna height will also increase with a rising value of  $\tau$ . The height can be reduced by choosing a smaller value of  $\sigma$ . From eqn. 6.7 we can conclude that a smaller  $\sigma$  results in a wider opening angle  $\alpha$ . The opening angle directly affects the directivity of the antenna: the wider the opening angle, the lower the directivity.

#### 6.2.2 Assembling a LPDA

LPDAs achieve their broad band properties by a combination of narrow band dipoles. To reproduce the transition from a single dipole to a broad band antenna, a LPDA has been assembled successively. The electrical layout of the used LPDA is identical to the antenna discribed in chapter 6.3. It consists of 9 wire dipoles with lengths from 1.47 m to 4.25 m. The outer structure of the antenna was replaced by a wooden frame to provide an easy access to change componets for different studies.

Starting with the shortest dipole, the further dipoles were attached one by one to the Lecher line. After each added dipole, a reflection measurement of the structure has been performed using a vector network analyzer<sup>1</sup> connected to the footpoint of the antenna. The series of reflection measurements in a frequency range from 10 MHz to 120 MHz is presented in fig. 6.3. The measured reflection coefficient has been converted into the reflected power using eqn. 5.10. For negligible structural losses, the power that is not reflected is radiated by the antenna.

In case of only one attached dipole (upper left in fig. 6.3), a resonance dip similar to the simulation of a  $\lambda/2$ -dipole can be observed (see. fig. 5.8). This dip seems to be much broader compared to the simulation which is due to the linear scale. The resonance frequency of the 1.47 m long dipole can be calculated with eqn. 5.35 and a shortening Factor of V = 0.96 to be 97.89 MHz. This value is slightly lower than observed in fig. 6.3 probably due to the attached Lecher line.

If the second dipole is attached, an according dip evolves also close to the expected frequency of 85.65 MHz. The resonance dip caused by the first dipole is apparently affected by the attachment of the second dipole as the reflection increases from zero to 50 % - the resonances seem to interfere each other.

In the further assembly every added dipole leads to a new resonance dip with a center frequency approximately equal to the resonance frequency expected from calculations. If the eighth dipole is attached (lower middle in fig. 6.3), simultanously a dip at approximately 110 MHz occurs. This is probably the  $3\lambda/2$ -resonance of the eighth dipole expected to be at 117.23 MHz. As the shortening factor for  $3\lambda/2$ -dipoles is lower than for  $\lambda/2$ -dipoles [36], its resonance frequency may fit to the observed 110 MHz.

With the ninth attached dipole the antenna is completely assembled. In this case a resonance dip can not be allocated clearly to the corresponding dipole anymore. However, a broad band sensivity of the antenna evolves. If we relate the threshold frequencies defining the bandwidth to half the reflected power (see chapter 5.3) we can observe a bandwith of approximately 52 MHz, from 31 to 83 MHz.

The sensivity within this bandwith is relatively discontinuous. This can be explained by the measuring conditions: the measurement was performed in the factory hall of the RWTH-Aachen University. In closed rooms, the radiated waves are reflected by walls etc. and will be picked up by the antenna again. These reabsorbed signals contribute to the measured reflection and thus influence the measurement. In further measurements performed outside, we will observe a much more constant sensivity within the bandwidth.

In summary, we can conclude that the broadband properties of an LPDA evolve

 $<sup>^1\</sup>mathrm{The}$  used network analyzer will be discussed in detail in chapter 6



Figure 6.3: Measurements of the reflected power during a successive assembly of a LPDA. The measurements from upper left to the lower right refer to one further attached dipole respectively. The calculated resonance frequency for the newly attached dipole is denoted with a red dot on the horizontal axis.

from a complex interaction of the single dipoles. A simple addition of the dipole resonances can not describe the observed characteristics. Furthermore we can reconsider the equations to determine the operational range of the antenna, eqn. 6.8 and eqn. 6.9. With the shortest (1.47 m) and longest dipole (4.25 m) of the measured antenna, these equations would predict a bandwidth from 35.26 MHz to 67.97 MHz. Espeacially the upper threshold frequency was observed to be about 15 MHz above the predicted value. Thus these equations only deliver a coarse estimate of the expected bandwidth. For more precise predictions special antenna design programs and simulations have to be taken into account.

# 6.3 The Black Spider LPDA

A LPDA was specially designed for the Auger Radio detector by Matthias Leuthold and Stefan Fliescher in 2008. The antenna is referred to as 'Black Spider LPDA'. A photo of the Black Spider is presented in fig. 6.4. The Black Spider was build in a mini series of ten antennas at the RWTH-Aachen University. Eight of these antennas are currently used in the Auger Radio measurement setups (see chapter 4.3). The antenna design evolved from a series of test antennas, among them the 'Flagship' antenna (see fig. 6.9 (right)). The Flagship antenna introduced a new wire design on which the Black Spider is based. Before the mechanical and electrical design features are discussed in detail, the overall conception as a consequence of the demands (see chapter 6.1) will be introduced.



Figure 6.4: Photograph of the Black Spider LPDA on the rooftop of the RWTH-Aachen Physics Institute.

## 6.3.1 Overall Conception

The Black Spider was designed for a frequency range from 36 MHz to 80 MHz. This fits the frequency range that is not polluted with man made noise sources at the experiment site. To deduce the dimensions of the antenna from the desired frequency range it is favourably to consider additional resources beside the equations given in chapter 6.2.1. Therefore a special software to design LPDAs has been used. The software is called LPCAD and was written by Roger Cox. The latest version, LPCAD 3.0 can be obtained as freeware from [43]. LPCAD delivers the length and spacing of each dipole for a LPDA design based on the upper and lower threshold frequency and the design parameters  $\tau$  and  $\sigma$ . Furthermore it creates an antenna model as an input file for the simulation software Nec2.

The LPCAD input parameter for the Black Spider LPDA are the following:

lowest frequency:	$36 \mathrm{~MHz}$
highest frequency:	$80 \mathrm{~MHz}$
wire diameter:	$1.4 \mathrm{mm}$
au:	0.875
$\sigma$ :	0.038

The choosen value of  $\tau$  is a good compromise between a constant sensivity within the bandwidth and a preferably low number of dipoles. More dipoles would increase the costs, the size and the weight of the antenna. The parameter  $\sigma$  was choosen relatively low compared to the optimum value of  $\sigma = 0.16$  given by LPCAD. This choice is reasonable due to the demand to achieve a compact antenna. The suggested  $\sigma$  would have resulted in an antenna with a height of 7.1 m. With the selected value the height from the lower to the upper dipole is 1.7 m, more than four times smaller. Furthermore a low sigma results in a wide opening angle  $\alpha$  of the structure. This yields a low directivity which is required for our purpose as the complete sky should be covered.

The result of the LPCAD calculation is a relatively compact LPDA with 9 dipoles. The exact dimensions for the wire lenght and spacing can be taken from the engineering drawing of the Black Spider presented in fig. 6.5.

## 6.3.2 Mechanical Layout

The Black Spider LPDA consists of two planes aligned perpendicular to each other as visible in fig. 6.4. These planes belong to two independent LPDAs unified in one mechanical structure. This is neccessary to cover the full azimuthal range of 360°. At the Auger site, one plane is aligned to the north-south-direction whereas the other plane points in east-west-direction. The planes are referred to as NS- and EW-polarization. Since both LPDAs can be read out separately, it is possible to investigate the dependance of the radio pulse properties on the geomagnetic field.

The construction of the Black Spider is based on the mentioned wire design. That means that the active elements are wires dipoles. The overall 36 dipole wires are guided through a central aluminium tube where they attached to the integrated waveguide. The central tube carries the whole structure and provides the housing for the integrated Lecher line at the same time. It has a length of 3.87 m and a outer diameter of 60 mm.

To tighten the wire dipoles in a horizontal position, an outer frame is neccessary. This frame consists of dynema wires supported by eight arms. These arms are made of fiberglass compound to reduce the amount of conducting materials within the antenna structure. The used compound material guarantees extremely light and rigid tubes. The arms are covered with a special finish to make them resistant against UV-radiation.

The outer wiring is tensioned from the top of the antenna over the ending of the arms to the bottom of the central tube. Additionally the four upper and four lower arms are connected with a expandable wiring. The tension in the outer wiring can be ajusted with tensioning levers at the bottom of the structure. They also allow to



Figure 6.5: Engineering drawing of the Black Spider LPDA. The dimensions of the wire dipoles and their spacing is given in mm. In the lower right a magnified view of a vertical cut through the central tube with a curcuit board is shown [44].

loose the outer wiring which is neccessary for the integrated folding mechanism. The eight arms are attached to the central tube through hinges. If the outer wiring is loosened, the arms can be folded towards the central tube. The folding mechanism reduces the size of the Black Spider antenna from  $3.87 \times 4.5 \times 4.5$  m to a packing size of only  $3.87 \times 0.4 \times 0.4$  m. Furthermore no parts have to be disassembled for this procedure. Toghether with a total weight of only 17 kg this provides an easy transportation and installation at the experimental site.

A big advantage of the wire design is the small windload of the antenna due to the thin dipoles. As the major fraction of the antenna mass is concentrated in the center, oscillations stimulated by the wind are not able to introduce high stresses in the antenna structure.

## 6.3.3 Electrical Layout

The dipoles are made of seven-strand chopper wires with a diameter of 1.4 mm. They are insulated with plastic to protect them against corrosion. The dipole wires are connected to the central Lecher line via circuit boards. A photo of a horizontal cut through the central tube of the Black Spider is presented in fig. 6.6 (left). One of the eleven circuit boards within the central tube is visible. The four small holes in the middle provide the guidance for the four wires of the two Lecher lines, one Lecher line for each polarization plane of the antenna. Every second circuit board has the mentioned cross connection of the dipoles to the waveguide which is visible on the photo. The four dipoles are soldered on the circuit boards close to the tube. As the footpoint of the antenna is located on the top, the connected coaxial cables have to lead to the bottom of the antenna. They are guided through two holes in the curcuit boards whithin the central tube.

The coaxial cables extent 3.2 m below the lowest point of the structure to be able to connect the antenna to the readout stations placed on the ground when it is mounted on a pole in about 2.5 m height. The four black rings visible in fig. 6.6 at the inner surface of the tube are rubber sealings to protect the circuits from water.

The wires of the lecher lines have a diameter of 0.5 mm and are placed in a distance of 12 mm. This leads to a characteristic impedance of the Lecher line of  $Z_0 = 464 \ \Omega$  according to eqn. 5.23. The connected dipoles also posses a characteristic impedance depending on the ratio of the dipole lenght  $l_i$  to its diameter  $d_i$  via [37]:

$$Z_0 = 120 \left( ln \left( \frac{l_i}{d_i} \right) - 2.25 \right). \tag{6.10}$$

For the given dimensions this leads to impedances of 564  $\Omega$  for the shortest to 676  $\Omega$  for the longest dipole. The impedances of the connected dipoles diminish the overall impedance of the Lecher line to ~ 200  $\Omega$ . Hence, the antenna has a footpoint impedance of 200  $\Omega$ . The footpoint impedance has to be reduced to the common standart of 50  $\Omega$  to be able to connect the antenna to a coaxial cable without reflection losses.



Figure 6.6: Horizontal cut through the central tube of the Black Spider LPDA (left). A circuit board with four soldered dipole wires and a cross connection to the central waveguide is visible. Guanella balun as used in the Black Spider (right).

Theoretically the Lecher line can be designed to directly achieve a footpoint impedance of 50  $\Omega$ . This would have yield a distance between the two wires of the Lecher line below their diameter - a demand which is not feasible. Furthermore a strong mismatch between the wire impedances and the impedance of the Lecher line would cause strong reflections via eqn. 5.33 at their connection point. Thus the Lecher line has to be tuned to a impedance approximately equal to the wire impedances.

The transformation between the footpoint impedance and the impedance of the coaxial cable is done by a 4:1 Guanella balun (see. chapter 5.2.3). A photo of the used balun is shown in fig. 6.6 (right). It is located at the upper circuit board directly at the footpoint of the antenna. It consists of a toroidal core made of a ferro-alloy with nine windings of insulated copper wire. It also provides the transformation from the balanced waveguide to the unbalanced coaxial cable.

The described antenna circuits are placed within the central tube where they are completely shielded agaist any electromagnetical interference from outside the structure.

## 6.3.4 Waveguide Measurements

In order to reach a better understanding of the electric transactions of an operating LPDA, measurements of the waveguide have been performed. The LPDA previously used for the measurement of the successive attachment of the dipoles (see chapter 6.2.2) has been investigated further. The voltage amplitude along the waveguide has been measured during transmitting operation.

For this purpose, a frequency generator has been connected to the antenna. It delivers a sinusoidal signal with adjustable amplitude and frequency. An amplitude of 5.0 V and three different frequencies within the bandwidth of the antenna, 30 MHz, 50 MHz and 70 MHz were selected. The voltage amplitude between the two wires of the Lecher line was measured with a digital oscilloscope for various distances to the stub. The stub (see fig. 6.2) represents the short circuit of the two wires of the waveguide. It is located 23 cm below the lowest dipole. A measurement was made at the height of each circuit board respectively and exactly between two boards in each case.

Fig. 6.7 summarizes the measurements of the voltage amplitude for the three different frequencies depending on the distance to the stub. The positions of the circuit boards are marked with black arrows on the horizontal axis. All measured curves start at the position of the stub with a voltage close to zero volts. This is evident as a direct short circuit is measured.

For the lowest frequency of 30 MHz the voltage steeply increases with the distance to the stub. A maximum voltage of 5.7 V is reached at the second dipole. With further increasing distance the voltage constantly drops to a minimum value at the sixth dipole. Finally it rises again towards the shortest dipole. The last two measuring points lie beyond the waveguide and therefore are seperated with a vertical line. The penultimate measuring point is taken directly in front of the balun, the last one behind the balun. For all frequencies a voltage drop over the balun to a constant voltage of  $\sim 3.5$  V is observed.

For the higher frequencies the developing of the curve is similar but shifted towards


Figure 6.7: Measurement of the voltage amplitude along the waveguide of the Black Spider for three different frequencies depending on the distance to the stub. The positions of the circuit boards are marked with black arrows on the horizontal axis. The measuring points right of the black line lie beyond the waveguide.

higher distances to the stub. At 50 MHz the maximum voltage is reached at the fifth dipole and a minimum between the seventh and eigth dipole. At 70 MHz the maximum lies at the eigth dipole.

The voltage distribution along the waveguide reflects the active regions of the LPDA. The power is radiated by the dipoles which are in resonance or close to resonance. These dipoles represent the active region of the antenna. The electromagnetic waves are radiated by alternating currents in the dipoles. These currents are driven by the voltages in the corresponding region of the transmission line. Therefore the voltage along the transmission line reaches a maximum at the active region.

At 30 MHz which is at the lower limit of the bandwidth, the three longest dipoles emit the most power as their length matches the input frequency. With increasing frequency the active region moves towards the shorter dipoles. This is clearly visible in the measured voltage distribution.

To investigate the corresponding current distribution of the antenna wires, a simulation with Nec2 (see chapter 5.3) has been performed. Nec2 calculates the radiation patterns based on the currents in the structure. These currents can be displayed with the mentioned graphical interface 4Nec2 directly on the antenna structure. The simulation is based on the antenna model of the Black Spider LPDA delivered by LPCAD (see chapter 6.3.1).

The result of the simulation is presented in fig. 6.8 for 30 MHz, 50 MHz and 70 MHz as in the measurement. In general, the current on a center feeded  $\lambda/2$ -dipole has a sinusoidal distribution. The current maximum is in the center, towards the ending the current ceases to zero. This progression can be observed with the color code on each single dipole in fig. 6.8. In this graphical presentation also the active regions of the antenna become very demonstrative. If we concentrate on the currents in the center of the dipoles, we are able to compare the distributions to the measured voltages along the Lecher line.

At 30 MHz we observe the strongest currents in the three lower dipoles. A minimum occurs at the sixth dipole followed by an increase towards the footpoint. This exactly reflects the measured voltages in fig. 6.7. The simulations for the other frequencies are also in good agreement with the measurements.

The good agreement between simulations and measurements inspires confidence to the simulations and the created antenna model. This is important as the investigations of the antenna's directivity discussed in the next chapter have to rely mostly on simulations.

### 6.4 Alternative Types of LPDAs

An LPDA for the needs of radio detection was first designed by Oliver Krömer for the LOPES<sup>STAR</sup> experiment within his PhD thesis [45]. The LOPES<sup>STAR</sup> experiment is an extension to the mentioned LOPES array and focuses on the realization of a Self Triggered Array of Radio detectors. The developed LPDA uses a mechanical layout commonly used for radio applications. A photograph of the antenna is shown in fig. 6.9 (left). It is referred to as Alu-LPDA as the complete structure is made from aluminium tubing. It also consists of two crossed LPDAs mounted on one pole. One of the planes is attached to the pole slightly higher, so that the dipole rods can cross each other without electrical contact. The mast is made of two square tubes for each plane which work as waveguides at the same time.

The Alu-LPDA is sensitive to a frequency range from 40 MHz to 100 MHz. The design parameters are the following:

$$\tau = 0.860, \qquad \sigma = 0.045, \qquad l_{min} = 1326mm, \qquad l_{max} = 3796.$$
 (6.11)

The Alu-LPDA was adopted for the Auger radio test setups and proved the ability of LPDAs to detect cosmic ray induced airshowers [46]. However, by now all Alu-LPDAs have been replaced by Black Spider antennas. The major issue of the Alu-LPDA is its mechanical instability. Due to the relatively thick dipole rods, the windload is much higher than for wire LPDAs. The wind induces strong oscillations, introducing high stresses to the structure due to the relatively high masses of the dipole rods. As a result the dipoles break at the anchorage point to the central masts. This issue was temporarily solved by connecting the dipole rods with plastic pipes at their ending as visible in fig. 6.9.

Another issue of the Alu-LPDA is the extensive installation at the experimental site. The antenna is too large to be shipped to the site in the mounted state - thus the dipoles have to be attached in the field. Furthermore the bandwidth is not optimal



Figure 6.8: Simulation of the current distribution along the dipoles of the Black Spider LPDA at 30 MHz, 50 MHz and 70 MHz. The current magnitude is color coded. In the lower right the model of the antenna structure is shown, including the crossed transmission line (blue) and the voltage source at the footpoint (pink).

matched to the radio background at the site (see. chapter 6.1) and thus can pick up man made noise in the frequency region from 85 MHz to 100 MHz.

The problems with the Alu-LPDA inspired Matthias Leuthold to construct a wire LPDA. The first realization of this concept was the Flagship antenna presented in fig. 6.9. The naming referrs to the remarkable size of the antenna with an extend of more than eight meters of one polarization plane. Dispite the size, the antenna proved the wind stability of the wire design in test setups at the site.



Figure 6.9: Photo of the Alu-LPDA (left) and the Flagship Antenna (right) at the Argentinian pampa.

# 7. Benchmarking the Black Spider LPDA

In this chapter the Black Spider LPDA will be investigated in detail. The basic antenna properties introduced in chapter 5 are quantified through measurements and simulations for the Black Spider. The ambition is to characterize those antenna properties which are essential for a precise reconstruction of measured radio signals from EAS. Furthermore different test setups, benchmarking the mechanical and electrical performance are discussed.

## 7.1 Basic Properties

### 7.1.1 Gain

Precise gain measurements of antennas for the radio frequency regime are only possible within great efforts. They are usually performed in special antenna measurement chambers, in which the distracting reflections of the radiated waves are absorbed through certain techniques. These measurements have to be made in the far field region of the antenna which is experimentally hard to realize due to wavelengths up to 10 m. However, antenna simulations deliver reliable predictions for the gain.

Simulations of the total gain of the Black Spider LPDA using Nec2 have been performed for various conditions. Fig. 7.1 (left) shows the gain sphere simulated in free space at 55 MHz. This frequency is in the center of the bandwidth of the antenna. The sphere has a smooth shape and reminds of the toroidal pattern of a dipole (see. fig 5.7). The major difference is that the radiation is beamed towards the upper directions. Regarding the color scale in fig. 7.1, almost the complete power is radiated into the upper hemisphere. A much higher maximal gain of 5.74 dBi compared to the dipole (2.17 dBi) is reached for directions towards the zenith. This can be specified in the directional diagram in fig. 7.1 (right) which is a cut through the y-z-plane. The gain in the backward direction is  $\sim -12$  dB lower than to the zenith. This refers to a surpression of the radiated power with a factor 0.06 compared to the maximal gain.

The beamwidth is observed to be 150° for the selected cut perpendicular to the antenna plane. This behavior fulfils the desired demands as a high sensivity to almost the complete sky and a low acceptance to signals from the ground and the very horizontal direction is observed. The directivity diagram for a cut through the antenna plane, will result in a lower beamwidth. Here, it is important to consider that the



Figure 7.1: Gain sphere of the Black Spider LPDA simulated in free space (left) and corresponding directional diagram (right). The antenna structure is shown above the pattern, the gain is color coded. The beamwidth of 150° is marked with the black lines.

shown gain sphere evolves only from one antenna polarization. For the real Black Spider two polarization planes complement each other: if one antenna plane has the minimal beamwidth, the other plane has its maximum beamwidth.

As the antenna will be operated in a certain height above a ground surface, the influence of the ground to the antenna gain has to be considered. Since electromagnetic waves in the radio regime are reflected on the ground, a significant change of the gain sphere is expected for every type of radio antenna. Fig. 7.2 shows the gain pattern of the Black Spider simulated in a height of 2.5 m above ground. The height relates to the distance of the lowest dipole to the surface. The choosen height above ground refers to the approximate height in which the antennas are installed at the Auger site.

The gain sphere is deformed since waves that would be radiated towards the ground are reflected to the side. As a consequence, the gain to the zenith and to high zenith angles increases, whereas it is slightly reduced for the zenith angles around 45°. For the horizontal directions the gain ceases to zero. Note that the color coding differs from the pattern in free space. This is reflected in the corresponding directional diagram. An equal maximum gain of 6.8 dBi is reached for both, a zenith angle of  $0^{\circ}$  and  $\sim \pm 70^{\circ}$ . The beamwidth can be identified to be 158°.

In the following, gain patterns of the Black Spider simulated over ground will be investigated with regard to different influences: the dependence on the frequency, the influence of different heights of the antenna above the ground and the variation with different ground conditions.



Figure 7.2: Gain sphere of the Black Spider LPDA simulated in a height of 2.5 m above ground (left) and corresponding directional diagram (right). The beamwidth can be determined to be 158°.

#### **Frequency Dependence**

To examine the frequency dependence of the gain pattern of the Black Spider, it was simulated in a height of 2.5 m above ground at 35 MHz, 50 MHz, 65 MHz and 80 MHz. The selected frequencies cover the whole bandwidth of the antenna. The properties of the gain patterns are compared based on their directional diagrams in fig. 7.3.

At 35 MHz which is at the lower limit of the bandwidth, the corresponding directional diagram has a smooth round shape. The gain is almost constant over an angular distance of 120°. The beamwidth can be estimated to 150°. At 50 MHz the round curve contracts at zenith angles of about 40°. With further increasing frequency, these contractions become more distinctive and shift to larger zenith angles. At 80 MHz additional contractions occur around 30° zenith angle. Thereby the beamwidth slightly rises to a value of about 170° at 80 MHz. The gain to the zenith stays almost constant in all cases.

Although the directional diagram changes in detail with the frequency, the overall shape stays stable. The sensitivity covers the complete sky exept the very horizontal directions.

#### Dependence on the Height above Ground

The height of the antenna above the ground surface also affects the gain pattern. This effect was studied with simulations of the Black Spider in heights of 1 m, 2,5 m, 5 m and 10 m above ground. The corresponding directional diagrams at a frequency of 55 MHz are compared in fig. 7.4.

A similar dependence as on the frequency can be observed. At 1 m above ground the directional diagram has smooth round shape. For larger heights above ground, ripples occur which become more distinctive and more frequently with increasing



Figure 7.3: Directional diagrams of the Black Spider LPDA simulated for different frequencies.

height. The beamwidth changes from  $150^{\circ}$  to approximately  $170^{\circ}$  at 10 m height. As for the frequency, the basic characteristic of the directional diagram stays similar for different heights.

#### Dependence on the Ground Conditions.

4Nec2 allows to configure the ground conditions for the simulations. The influence of the ground on the propagation of electromagnetic waves can be parameterized by the dielectric constant  $\epsilon_r$  and the conductivity  $\sigma$  of the ground. These two parameters depend on the type of the ground, e.g. sand or water, and on the condition, e.g. dry or moist.

4Nec2 allows to preselect different ground types and calculates the gain with respect to the corresponding values of  $\epsilon_r$  and  $\sigma$ . The most common ground condition during a year at the Auger site can be characterized by dry and sandy. Therefore the previous discussed gain patterns have been simulated with the ground 'dry, sandy'. At certain times of the year the ground may be covered with snow or can be very wet. To cover a broad range of possible ground conditions at the site, simulations with the grounds: 'dry, sandy', 'moist', 'water' and 'arctic' have been performed. They have been simulated at a frequency of 55 MHz in a height of 2.5 m above ground. A comparison of the according directivity diagrams is presented in fig. 7.5.

The simulations for the different ground conditions deliver very similar results as the different curves are almost congruent. Slightly variations at zenith angles around  $45^{\circ}$  and  $70^{\circ}$  are observed for the 'arctic' ground. However, the simulations predict that the Black Spider is almost independent on different ground conditions.



Figure 7.4: Simulated directional diagram of the Black Spider LPDA for different heights above ground at 55 MHz. The given heights refer to the distance from the lowest dipole to the ground surface.



Figure 7.5: Simulated directional diagram of the Black Spider LPDA for different ground conditions at 55 MHz.

#### Horizontal Directional Diagram

So far we focused on the vertical directional diagram of the antenna, as the

vertical beamwidth is of major interest for our purpose. From the three dimensional gain sphere (see fig. 7.2), also the horizontal directional diagram can be derived by a horizontal cut through the gain sphere at a certain zenith angle.

A measurement of the horizontal directional diagram has been performed for the Black Spider at the Auger site. The used measurement setup consists of a transmitter at a frequency of 40 MHz mounted in a height of 5.25 m above the ground. The emitted waves are polarized horizontally. A Black Spider antenna was placed on the ground in a distance of 100 m from the transmitter. With the given dimensions, the signal from the transmitter reaches the antenna at a zenith angle of  $87^{\circ}$ . The power of the signal received in one of the two polarization planes of the antenna is measured with a spectrum analyzer<sup>1</sup>. Afterwards the antenna is rotated by  $10^{\circ}$  in the azimuth angle and the measurement is repeated. In this way, a horizontal directional diagram, covering a azimuth angle of  $360^{\circ}$  evolves.

The measured diagram is presented in fig. 7.6 in comparison to the corresponding simulation. As the power of the used transmitter is not specified, no absolute values for the gain can be determined. Hence, both the measured and the simulated directional diagram were normalized to the maximum gain respectively. The maximum gain is reached for waves incoming from directions perpendicular to the antenna plane (90° and 270°). A minimum sensitivity is observed for those directions parallel to the antenna plane. The simulation shows very similar results as both curves have an almost identical shape. A difference of about 22 dB between the maximum and the minimum gain is measured, the corresponding value from the simulation is 20 dB.





Figure 7.6: Measured (left) and simulated (right) horizontal directional diagram for the Black Spider LPDA. Both diagrams are normalized to the maximum gain respectively. They refer to a zenith angle of 87° and a frequency of 40 MHz. The antenna plane is aligned parallel to the horizontal axis.

### 7.1.2 Reflection

The reflection describes the ratio of the reflected power to the input power at the footpoint of the antenna. The power that is not reflected is either radiated by the antenna or dissipated into heat due to the structural losses within the antenna structure (see chapter. 5.3). These losses are usually low compared to the accepted power at the footpoint. However, this can be verified through simulations. Fig. 7.7 shows the simulated radiation efficiency of the Black Spider LPDA for frequencies from 30 MHz to 100 MHz. The efficiency slightly varies around 99 %. The structural losses are thus close to 1 % compared to the accepted power at the footpoint of the antenna.

In the further discussion we will neglect the structural losses and assume that the complete power that is not reflected will be radiated by the antenna.

In the following various antenna measurements are presented. All measurements have been performed with the Rhode & Schwarz FSH4 spectrum analyzer. Details and data sheets can be taken from [47]. Beside the spectrum analyzer functions, the device provides a two port vector network analyzer which enables the measurement of all four S-parameters (see chapter 5.2.2).

The spectrum analyzer mode measures spectra from 100 kHz up to 3.6 Ghz with a resulution bandwidth from 100 Hz to 300 kHz. The network analyzer is widely used to perform reflection measurements in this thesis. During a reflection measurement, the analyzer delivers a defined input voltage to the device under test at a certain frequency. The reflected voltage is measured. The user can preselect a frequency range and a resolution bandwidth. The analyzer sweeps through the selected frequency range and records the reflection for every frequency bin.

Based on the performed measurement mode, the device is referred to as spectrum



Figure 7.7: Simulated Radiation Efficiency of the Black Spider LPDA in the frequency range from 30 MHz to 100 MHz.



Figure 7.8: Comparison of the measured and simulated reflection of the Black Spider. In the legend, the integrated radiated power within the bandwidth is denoted respectively.

analyzer or network analyzer.

A reflection measurement of the Black Spider LPDA was performed under realistic conditions at the Auger site. The antenna mounted in the typical height of 2.5 m was connected via its coaxial cable to the network analyzer. The coaxial cable of the antenna has a total length of 6 m. 2.8 m lie within the central tube, reaching from the footpoint at the top to the bottom of the tube. The cable extends 3.2 m below the antenna to enable a connection to the antenna station placed on the ground. Within the coaxial cable the signal is attenuated which is not negligible. The attenuation can be reduced by placing the low noise amplifier directly at the bottom of the central tube (see chapter 8.2).

As the signal from the network analyzer has to propagate to the antenna and back again to the analyzer, the signal is attenuated twice. The attenuation of the used cable can be taken from appendix A.1 depending on the frequency. An average attenuation of 0.1 dB/m within the bandwidth can be estimated. For the relevant cable length of 12 m, the attenuation is then 1.2 dB. The following measurement is corrected for this attenuation to be able to compare to simulations.

The reflection was measured from 10 MHz to 120 MHz. In fig. 7.8 the measured reflected power is compared to the simulated reflection. The simulation was adjus-

ted to the measurement conditions via the choice of a height of 2.5 m above a 'dry, sandy' ground.

Above about 30 MHz the measured reflected power steeply decreases. From 30 MHz to approximately 80 MHz a plateau evolves where the reflection stays below 20 % of the input power delivered by the network analyzer. The dips typical for an LPDA are visible within this range. Compared to fig. 6.3 these dips are less distinctive as the measurement was performed outside without the distracting reflections from walls etc.. Above 80 MHz the reflection increases - the antenna becomes less sensitive. The bandwidth can be determined with the criteria introduced in chapter 5.1.3 of a VSWR smaller than 2. This value corresponds to a reflected power of 11.1 %. Hence, a bandwidth of 52 MHz from 32 MHz to 84 MHz is observed. The total radiated power within the bandwidth can be calculated by integrating over the bandwidth. The according values for both, measured and simulated reflection are denoted in fig. 7.8 in percent of the input power.

The simulation delivers similar results. The simulated bandwidth is 52 MHz as well but shifted for about 1.5 MHz towards the higher frequencies compared to the measurement. A significant difference is observed at frequencies above 80 MHz. The simulation predicts a steeper decline of the sensitivity beyond the bandwidth. However, the observed bandwidth covers the complete frequency window available for the radio detection at the site.

### 7.1.3 Impedance

The impedance of the antenna can be observed if the phase of the reflected voltage is measured in addition to the magnitude. This was done for the measurement discussed in the previous section. Fig. 7.9 illustrates both, the measured magnitude and phase in a Smith chart for frequencies ranging from 20 MHz to 100 MHz. The Smith chart is normalized to the source impedance of the connected spectrum analyzer of 50  $\Omega$ .

The complex impedance of the antenna varies with the frequency, thus forming a trace which rotates several times around the center of the Smith chart. The trace starts in the left outer region at 20 MHz at a large reflection coefficient. It rotates a full turn and then begins to collapse towards the center of the Smith chart for frequencies above 30 MHz. Within the bandwidth of the antenna the trace rotates closely around the center. These small distances to the center of the Smith chart accord to low reflections already displayed in fig. 7.8.

Each full turn refers to one resonance dip in the reflection plot (compare to fig. 7.8). The impedance of the antenna has a reactive component which alternates between capacitive and inductive reactances. Above 80 MHz the turning radius slowly increases as the reflection rises again.

The observed behaviour of the antenna impedance is typical for LPDAs. A LPDA consists of dipoles which have a capacitive reactance below the resonance frequency and an inductive reactance above their resonance (see fig. 5.9). The arrangement of the dipoles provides that the reactances of the single dipoles compensate each other to a large extent within the bandwidth of the LPDA. However, a full compensation



Figure 7.9: Smith Chart of the measured complex reflection coefficient of the Black Spider in a frequency range from 20 MHz to 100 MHz. The center is normalized to the pure resistive source impedance of 50  $\Omega$ .

cannot be realized and thus the periodic variation of the antenna impedance cannot be avoided.

The performance of the impedance matching between the antenna and the source can be characterized through the turning radius of the trace in the smith chart and through the position of the center of the circular traces. If the average impedance within the bandwidth of the antenna differs from the source impedance, the center of the circular traces diverges from the center of the Smith chart.

An alternative representation of the impedance can be obtained if the reactance and the resistance are plottet seperately against the frequency. This is presented in fig. 7.10. Here, the periodic variation of both, the reactance and the resistance is more demonstrative. Beyond the bandwidth of the antenna, the amplitudes of the variation become very large. The values of the reactance and resistance within the observed bandwidth from 32 MHz to 84 MHz were filled in a histogram respectively. The RMS of both, the resistance and the reactance is about 16  $\Omega$ . It should be preferably low within the bandwidth. The mean values are 50.7  $\Omega$  for the resistance and 0.3  $\Omega$  for the reactance. This means that the antenna impedance is well matched to the pure resistive source impedance of 50  $\Omega$  within the bandwidth.



Figure 7.10: Measured resistance (top) and reactance of the Black Spider for frequencies from 20 MHz to 100 MHz. At the right, the according values within the bandwidth are filled into a histogramm respectively.

#### 7.1.4 Group Delay

A further antenna property which is of great interest for the purpose of radio detection is the group delay of the antenna. Signals with different frequencies will obtain different delays within the antenna structure.

One reason for the altering delays are the different distances the signals have to travel within the antenna structure. An electromagnetic wave with a frequency of 30 MHz will mainly be received by the lower dipoles. It has to travel along the whole waveguide before it reaches the footpoint of the antenna. In contrast, a wave with a frequency of 80 MHz is received by the upper dipoles directly at the footpoint and thus will have a lower delay. Consequently, radio pulses consisting of various frequencies will be distorted by the antenna. Since the pulse shape of the radio signal containes basic informations of the underlying air shower, a precise knowledge of the group delay caused by the receiving antenna is essential for the reconstruction.

The group delay  $\tau_G$  is defined as the derivative of the phase  $\Phi$  with respect to the angular frequency  $\omega$ :

$$\tau_G(\omega) = -\frac{d\Phi}{d\omega}.\tag{7.1}$$

The group delay of antennas can be determined by a transmission measurement performed with two identical antennas and a two port vector network analyzer. The antennas are placed facing each other. One of the antennas works as a transmitter, the other one as a receiver. The distance between the antennas must be large en-



Figure 7.11: Setup of a group delay measurement of the Black Spider LPDA. Two Black Spiders tilted by 90° are facing each other in a distance of 47.6 m.

ough, so that the receiving antenna is within the far field region of the transmitting antenna. In this case the received wave can be assumed as a plane wave, hitting each point of an antenna plane, perpendicular to the propagation direction at the same time. The network analyzer is connected with cables to both, the transmitting and the receiving antenna and a transmission measurement is performed. The network analyzer provides an input signal for the transmitting antenna for various frequencies. This signal is radiated and received by the other antenna. The magnitude and the phase of the received signal compared to the input signal is measured. This is a measurement of the  $\underline{S}_{11}$ -parameter, introduced in chapter 5.2.2. With the phase and frequency information, the group delay can be calculated via eqn. 7.1.

Using the described measurement setup, the group delay of two Black Spider antennas has been measured at the Auger site. Fig. 7.11 shows a photograph of the setup. The antennas were mounted in a distance of 47.6 m referring to the longest dipole. In chapter 5 we defined the far field region for distances to the antenna greater than four times the observed wavelenght. For the choosen distance, this is the case for any frequency above 25.2 MHz.

The transmission measurement was performed in a frequency range from 10 MHz to 120 MHz. The measurement was calibrated for the cable delay of the coaxial cables used to connect the analyzer to the antennas. Fig. 7.12 shows the recorded magnitude (top) and phase (bottom) of the transmitted signal. The magnitude reflects the antenna characteristics. Below about 30 MHz the noise floor of the network analyzer related to the selected resolution bandwidth is observed. On the one hand, the transmitting antenna barely radiates any power at the frequencies below 30 MHz. On the other hand, the receiving antenna is almost insensitive for those frequencies



Figure 7.12: Magnitude (top) and phase (bottom) of a transmission measurement between two Black Spider LPDAs placed in a distance of 47.6 m. Below about 30 MHz the noise of the network analyzer is observed.

(see fig. 7.8). For heigher frequencies both antennas become sensitive according to their bandwidth. A strong signal of a power being in average 45 dB above the noise floor is received within the bandwidth. Above 80 MHz the signal strength slightly decreases.

Below 30 MHz the phase shows a chaotic behavior as only the noise is measured. Above 30 MHz the phase continuously decreases. A linear decrease is expected for a linearly increasing frequency and a fixed length of the signal path if the group delay is constant. Note that the decrease is not linear.

From the recorded phase the group delay was computed via eqn. 7.1. The result is presented in fig. 7.13. The curve has been corrected for the offsets caused by the constant delays of the transmitted wave travelling between the antennas and the delay within the antenna cables. This results in the pure group delay of two antennas. Therefore the values plotted in fig. 7.13 were devided by two to receive the group delay of a single Black Spider.

Due to the chaotic phase below 30 MHz, the group delay is undefined for these frequencies. As suggested by the non linear progression of the observed phase, the group delay is not constant above 30 MHz. It decreases towards higher frequencies. The group delay within the bandwidth from 32 MHz to 84 MHz decreases about 88 ns. This value cannot be explained only by the different signal paths within the antenna as mentioned. The additional traveled distance along the waveguide (1.7 m) would result in a delay of 5.7 ns for the lowest frequencies compared to the heighest

frequencies within the bandwidth. Since the antennas are aligned facing each other, the wave has to travel an equal additional distance in the air. As the observed decrease of the group delay is much higher than 11.4 ns we can conclude the following: either the signal velocity along the waveguide is much lower than the speed of light, or an other, unknown effect yields different signal velocities for different frequencies within the antenna structure.

However, for the reconstruction of EAS based on the measured radio pulses, a low decrease of the group delay within the bandwidth is not necessary. It is more important that the progression of the group delay is precisely known.



Figure 7.13: Measured group delay of the Black Spider LPDA in the frequency range from 10 MHz to 120 MHz.

### 7.1.5 Spread in Production

The Black Spider antenna is a complex device, consisting of a large amount of parts of both, the mechanical and the electrical structure. Variations during the production may cause different characteristics of the single antennas. Furthermore the antennas are exposed to permanent stresses at the site of the experiment which may change the antenna characteristics.

To investigate these uncertainties, all eight Black Spider antennas which are at present used in the measurement setup were compared based on a measurement of their reflection coefficient. Fig. 7.14 shows the measured reflection curves for the NS-polarizations. The corresponding comparison of the EW-polarizations is attached in appendix A.3. Very similar results occur as the curves are almost congruent. The variations are in the order of a few percent within the bandwidth.

The blue curve belongs to a Black Spider placed on a metal pole. All other antennas are mounted on wooden poles. As the wooden poles suffer from the weather conditions, it was considered to replace them by more resistant metal poles. The question was raised whether the conducting metal poles have any influence on the



Figure 7.14: Comparison of the reflection of the NS-polarization of eight different Black Spider LPDAs mounted at different poles of the Auger Radio setup.

porformance of the antenna. Since the blue curve does not diverge from the other recorded curves, it can be assumed that the influence of the metal pole is negligible.

## 7.2 Test Setups

Several test setups have been used to benchmark both, the mechanical and the electrical performance of the Black Spider antenna. Stress tests were performed in the laboratory and under operating condition at the Auger site. The electrical performance of the wire design was compared to other LPDA designs and different antenna types.

### 7.2.1 Stress Testing

A critical point of the wire design might be the soldered attachment of dipoles to the circuit boards within the central tube of the Black Spider (see fig. 6.6). The permanent oscillations of the dipoles induced by the wind might damage these soldered connections. This was investigated by a test setup in the laboratory, where such oscillations have been simulated. A photograph of the stress testing setup is presented in fig. 7.15. In the background of the photo, a short part of the central tube of the Black Spider with two attached dipoles is visible. It includes a part



Figure 7.15: Stress testing setup for the attachment of the dipole wires to the circuit boards within the central tube of the Black Spider.

of the waveguide and two circuit boards identical to those within the Black Spider. The dipoles are tensioned using spiral springs. In the foreground an electric motor with an attached arm is visible. The arm hits the dipoles while rotating and induces the oscillations of the dipoles. The amplitude of this oscillations can be adjusted by the revolution speed of the electric motor. An amplitude of more than 10 cm has been simulated, which is much larger than any amplitudes observed at the Auger site. The setup is running permanently since October of 2008 by now, which corresponds to more than 50 millions of oscillations. No damage of any kind could be observed at the antenna parts. However, the testing will be continued.

The most realistic test setup is the experiment site itself. The antennas are exposed to the weather conditions on a pole in several meters height. Four of the eight antennas, used at the moment in the measurement setup in Argentina, survived these conditions for more than a year by now. The other four Black Spiders were recently mounted at the site and were formerly tested under outside conditions in Germany and France. No damage of any kind on the mechanical structure was reported by now although no maintenance was performed to the antennas. Fig. 7.14 proves that the electric properties did not suffer either. These results are remarkable as the antennas were exposed to several storms with wind speeds above 150 km/h.

### 7.2.2 Galactic Noise Measurement in Argentina

In chapter 6.1 we found that the antenna should be sensitive to the galactic noise background to be able to detect radio pulses from EAS effectively.

A measurement of the galactic noise background was performed in [42] with the Alu-LPDA described in chapter 6.4. The result is presented in fig. 7.16. The measured signal was amplified by about 20 dB with a low noise amplifier (LNA). The blue curve represents a continuous measurement of the spectrum at the site. The transmitters of the short wave bands below 30 MHz and the transmitters of the FM-band above



Figure 7.16: Continuous measurement of the radio background at the Auger site (blue) and calculated progression of the galactic noise (red) [42].

about 85 MHz are visible.

In the range from 30 MHz to 85 MHz the galactic noise background is observed. The intensity of the galactic noise in the radio regime was parameterized by H.V. Cane in 1977. The spectral intensity  $I(\nu)$  was found to follow the parameterization:

$$I(\nu) = I_g \nu^{-0.52} \frac{1 - exp(-\tau(\nu))}{\tau(\nu)} + I_{eg} \nu^{-0.80} exp(-\tau(\nu)) \qquad [Wm^{-2}Hz^{-1}sr^{-1}],$$
(7.2)

with the frequency  $\nu$  in MHz. The parameter  $I_g = 2.48 \cdot 10^{-20}$  refers to the galactic contribution, whereas the parameter  $I_{eg} = 1.06 \cdot 10^{-20}$  refers to the extra-galactic contribution. The parameter  $\tau(\nu) = 5.0 \cdot \nu^{-2.1}$  denotes the opacity towards the polar direction [41].

Using this parameterization, the power of the galactic noise received by an antenna  $P_{rec}$  with an effective antenna area  $A_{eff}$  and an opening angle  $\Omega$  can be estimated to:

$$P_{rec} = I(\nu) \cdot A_{eff} \cdot \Omega. \tag{7.3}$$

With this consideration, the red curve in fig. 7.16 was computed. A similar progression compared to the measured background can be observed. The cutoff at above 35 MHz results from the antenna characteristic. The used antenna has a bandwidth from 40 MHz to 100 Mhz.

A comparable measurement has been performed with the Black Spider LPDA. It



Figure 7.17: Measurement of the radio background at the Auger site performed with the Black Spider LPDA. The blue curve represents the noise floor of the spectrum analyzer. The red curve was recorded without amplification, the black curve is amplified by 20 dB with an LNA.

is presented in fig. 7.17. The blue curve represents the noisefloor of the spectrum analyzer. The red curve was recorded only with the Black Spider whereas a LNA was interconnected to record the red curve. Already without the LNA the galactic noise can be observed. The received power is higher at lower frequencies as the galactic noise decreases exponentailly with the frequency as visible in fig 7.16. The progression of the galactic noise becomes more distinctive if the signal is amplified by 20 dB. The cuttoff at about 30 MHz matches the bandwidth of the Black Spider LPDA. The ripples within the bandwidth refer to the dips of the reflection curve (see fig. 7.8).

### 7.2.3 Galactic Noise Measurement in Nancay

As the Black Spider has proved its ability to detect the galactic noise background, we can now investigate whether the antenna is also capable to observe variations of the galactic noise. The radio background is not isotropic. An higher intensity can be observed towards the region of the galactic center. With an antenna of sufficient sensitivity, a variation of the galactic radio background due to the transit of the galactic center along the sky can be observed.

A measurement of this variation was recently performed at the Radio Observatory

in Nancay, France. The Black Spider LPDA toghether with 3 different prototypes of radio antennas recordet the galactic noise simultanously for 20 hours in a frequency range from 55 MHz to 57 MHz. The comparison of all four antennas is presented in fig 7.18. A significant variation of the recieved power is visible for the Black Spider, the Butterfly antenna and the Lwda antenna. For the Salla antenna the variation is less distingtive. The variation is mainly caused by the transit of Centaurus A. Due to other radio sources which contribute to the galactic noise, the variation is not pure sinusoidal.



Figure 7.18: Measured variation of the galactic noise due to the transit of the galactic center along the sky recorded by the Black Spider and three different prototype antennas.

### 7.2.4 Comparison to the Aluminum LPDA Design

The wire design of by the Black Spider yields a relatively extensive electrical layout compared to common constructions of LPDAs. Mostly LPDAs use a design similar to the layout of the Alu-LPDA discussed in chapter 6.4 which avoids the application of baluns and circuit boards within the antenna structure. To investigate how the more complex wire design influences the performance of the antenna, the Black Spider was benchmarked with a corresponding Alu-LPDA. For this purpose a Alu-LPDA with the design parameters of the Black Spider was build, reffered to as Big Alu-LPDA since it is larger compared to the previous discussed Alu-LPDA.

The performance of both LPDAs is compared on the basis of reflection measurements in fig. 7.19. The Big Alu-LPDA has an enlarged bandwidth of 57 MHz compared to the Black Spider LPDA, which is slightly shifted to the higher frequencies. The average sensitivity within the bandwidth is comparable. The reflection curve of the Big Alu-LPDA shows significant discontinuities, e.g at 27 MHz and 44 MHz, which are not due to errors in the measurement. They are caused by the interference of the two single polarization planes of the antenna. The dipoles of the two planes cross each other in a small distance so that the radiation emitted by the measured plane is disturbed. The discontinuities disapear if one plane of the antenna is measured seperately i.e. when the other plane is removed. The corresponding comparison of a single plane to the Black Spider is shown in appendix A.4.

In case of the Black Spider the crossing of the two antenna planes is realized internally on the circuit boards. This design suppresses the interferences between the to polarizations planes. This is an advantage since a continuous sensivity is preffered. In summary, it is observed that the complex electrical layout of the wire design does not reduce the performance, on the contrary, it has the advantage to be free from interferences between the polarization planes.



Figure 7.19: Comparison between the reflection of the Black Spider and the Big Alu-LPDA.

# 8. Antenna for AERA: Small Black Spider

As a first attempt to establish the radio detection technique on large scales, the AERA (Auger Engineering Radio Array) detector will be constructed within the Pierre Auger Observatory. For the first phase of the AERA project, a new antenna design based on the Black Spider LPDA will provide the antenna hardware. The complete AERA detector, will consist of ~ 160 antennas within an area of ~ 20  $km^2$ . Within this thesis an antenna design was developed which is optimized for an effective production in large quantities. Thereby a good reproducibility of the antenna properties as well as preferably low costs were focused for the mass production design.

In the previous discussion we verified that the properties of the Black Spider LPDA are well adjusted to the demands of radio detection at the Auger site. Therefore the overall concept and the electrical layout were adopted to the new antenna for AERA. Mainly the mechanical structure provides potential for an optimizing with regard to mass production. The developed antenna is referred to as Small Black Spider and will be discussed in detail in the following.

## 8.1 Mechanical Layout

The complete antenna design of the Small Black Spider was created in AutoCAD 3D, a Computer Aided Design software used in engineering. It allows to model three dimensional designs of complex structures. Each part of the designed structure can be extracted and delivers an input for the fabrication. As far as possible, the components of the Small Black Spider are fabricated with CNC (Computerized Numerical Control) machines which use the input of AutoCAD.

This process of production delivers a maximum precision and reproducibility of the components. Furthermore, if once the AutoCAD design is created, the production of the components is less expensive than for a manufacturing of the parts, especially when large quantities are needed. The AutoCAD design of the complete Small Black Spider is presented in fig. 8.1.

The Small Black Spider is constructed with the same design parameters used for the Black Spider LPDA. In consequence, the dipole lengths and their spacing are identical to its previous version. The name Small Black Spider refers to a reduction of the size of the outer structure. The central tube is 420 mm shorter compared to



Figure 8.1: AutoCAD 3D model of the Small Black Spider LPDA [48].

the Black Spider. The exact dimensions can be taken from fig. 8.3. The reduction of the outer dimensions is mainly achieved by using the lower arms of the outer frame as dipoles at the same time. In consequence the Small Black Spider has only 8 wire dipoles and one dipole made from aluminum tubing. These dipole arms are electrically insulated from the central aluminum tube through insets of rigid PVC. The active part of the dipole arms is connected via short wires to the central waveguide. This allows to integrate a folding mechanism as discussed for the Black Spider.

The upper arms are made from tubes of fiberglass compound. An application of aluminum tubes as upper dipoles, as done for the lower dipoles, would reduce the performance of the antenna. The impedances of the thicker tubes (see eqn. 6.10) compared to the thin wires affect the electrical operation of the waveguide. This is critical if these lower impedances are appended directly at the footpoint of the antenna.

It was proven by measurements that the lower dipoles of aluminum tubing have no negative influence on the performance of the antenna.

All eight arms of the Small Black Spider are connected to the central tube by hinges



Figure 8.2: Details of the AutoCAD model of the Small Black Spider: the connection of the outer wiring to the arms (left), the attachment of the lower arms to the central tube through aluminum profiles (center) and the plug with attached outer wires at the top of the central tube [48].

made of aluminum sheet plates bended to a U-profile. These profiles are attached through blind rivets to the central tube. A magnified view of the attachement of the arms is shown in fig. 8.2. This yields an efficient design of a major component of the antenna with regard to costs and production time.

The outer wiring of the Small Black Spider consists of overall 20 dynema wires reinforced by thimbles at their endings. The endings are fixed through CNC-machined aluminum parts to the central tube e.g. the outer endings of the arms. Detail views of these componets designed with AutoCAD are presented in fig. 8.2. These components are plugged on the central tube or the arms respectively allowing a fast assembly and replacement of parts for maintenance. The outer wires are attached through bolts secured with splints.

The Small Black Spider uses stainless steel spiral springs to tension the dipoles. The springs are electrically insulated from the dipoles through plastic insulators. It was proven by reflection measurements that the applied springs have no negative influence on the performance of the antenna.

The springs are completely resistant against UV-radiation and low temperatures. The force introduced to the dipole wires scales almost linear with the elongation of the springs. That allows to abstain from tensioning levers to introduce tension in the outer wiring, resulting in a further reduction of costs for the Small Black Spider.

## 8.2 Integrated Low Noise Amplifier

In chapter 7.1.2 we discussed the power loss within the coaxial cable of the antenna during a reflection measurement. The measured reflection was corrected for the signal loss according to twice the cable lenght, as the signals propagate from the network analyzer to the antennas footpoint and back again to the analyzer.

If the antenna is operated in receiving mode, only the attenuation according to once the cable lenght has to be considered. However, the signal loss is still not negligible. Within the overall six meters of cable needed to reach the readout stations on the



Figure 8.3: Sideview including the exact dimensions of the Small Black Spider LPDA [48].

ground, the signal is attenuated by in average 0.6 dB (see. appendix A.1). This corresponds to a power loss of about 15 % of the input power.

This power loss can be reduced by amplifying the received signal as close to the footpoint as posssible.

In case of the Small Black Spider, the LNAs will be installed directly at the bottom of the central tube . This combines an relatively easy access to the LNAs with a lower power loss. Since the unamplified signals are only attenuated along the cable within the central tube with a lenght of about 2.8 m, the power loss can be estimated to be about 6.5 %.

A direct installation of the LNAs at the footpoint of the antenna within the central tube is leads to disadvantages. The maintenance of the LNAs would be extensive in this case, as the antenna has to be removed from the pole in order to access the footpoint at the top of the antenna.

Furthermore, in case of the direct installation at the footpoint, connectors have to be integrated at the upper circuit board to enable a replacement of the LNA. Detailed



Figure 8.4: AutoCAD design of the bottom of the Small Black Spider. The lower end cab provides the attachment of the outer wiring and the connection of the two LNAs. The LNAs are placed in separate housings. The cylindrical box protecting the LNAs from rain, wind etc. was removed for a demonstrative illustration [48].

investigations of different connectors were performed in [35]. Here, it was observed, that these connections result in a significant increase of the internal noise of the LNA.

The two LNAs are placed in seperate housings below the central tube of the Small Black Spider as illustrated in a magnified view of the AutoCAD model presented in fig. 8.4. These housings are screwed by male N-Type<sup>1</sup> connectors to the corresponding female connectors integrated in the lower end cab which also provides the attachement of the outer wiring. The female N-Type connectors are attached to the coaxial cables within the central tube. The connectors are electrically insulated from the end cab by plastic insets. This avoids ground loops and any pickup of noise which might be injected by the mechanical structure.

A cylindrical box shelters the LNAs and other readout components such as filters which can be interconnected in a modular way. This provides flexibility to e.g. filter out certain frequency contents of the received signals. The cylindrical box was removed in fig. 8.4 for a better illustration but is included in fig. 8.1.

<sup>&</sup>lt;sup>1</sup>N-Type connectors are waterproof RF-connectors widely used to connect coaxial cables and RF-devices.

## 8.3 Waveguide

The electrical layout of the waveguide of the Small Black Spider, i.e. the diameter and the distance of the parallel wires was not modified compared to its previous version. To enable an easier mounting to the central tube, a support structure was integrated.

A photograph of a prototype of the waveguide of the Small Black Spider is shown in fig. 8.5. The circuit boards are supported by two plastic pipes glued to the boards. The plastic pipes provide a guide for the coaxial cables running from the footpoint to the lower end cab of the antenna. The waveguide, including the baluns on the upper circuit board, is assembled in a mounting fixture which adjusts the exact spacing of the circuit boards. If the waveguide is completely assembled, it is removed in one piece from the mounting fixture and is slided into the central tube of the Small Black Spider. The dipole wires are then soldered from the outside of the central tube to the corresponding contacts of the circuit boards.

The new design using support pipes leads to an easy and fast mounting of the Small Black Spider.



Figure 8.5: Prototype of the waveguide of the Small Black Spider hold by a mounting fixture. A curcuit board with the cross connection of the dipoles to the Lecher line in the foreground and the support pipes are visible.

## 8.4 Baluns

The baluns used for the Black Spider LPDA were discussed in detail in chapter 6.3.3. They are handcrafted from insulated copper wires wound around a toroidal core. For the Small Black Spider, the application of commercial SMD-baluns (Surface Mounted Device) is evaluated. A suitable balun, providing the same transformation ratio of 4:1 as the baluns used in the Black Spider, is the Mini-Circuits type TC4-19+ SMD-balun. A photo and the equalent circuit of this balun is presented in fig. 8.6. These baluns provide a more constant performance compared to those used for



Figure 8.6: Photo (left) and equivalent circuit (right) of the Mini-Circuits type TC4-19+ SMD-balun. A fifth contact for an alternative transformation ratio is visible in the equivalent circuit (Adopted from [49]).

Black Spider LPDA. Furthermore they are much smaller with a size of  $3.81 \times 3.81 \times 4.06$  mm, which allows to directly integrate them to the upper circuit board.

To compare the performance of the SMD-baluns to the baluns of the Black Spider, reflection measurements were performed. For this measurement, the baluns were connected to the network analyzer with a source impedance of 50  $\Omega$  at their low impedance side. The high impedance side was terminated with a load of 200  $\Omega$ . A reflection measurement was performed in a frequency range from 0 MHz to 120 MHz. If the balun achieves the design impedance transformation ratio of 4:1, the signal will be totally consumed by the terminated load and thus no power will be reflected back to the analyzer.

The result of the reflection measurements is presented in fig. 8.7. Both curves show a steep decline of the reflection up to frequencies of about 10 MHz for the SMD-balun and about 17 MHz for the handcrafted balun respectively. At these frequencies the baluns reach their maximal performance. For higher frequencies, the reflection for the regular balun increases steeply, whereas the reflection for the SMD-balun increases only slowly. In the interesting frequency range according to the bandwidth of the antenna ( $\sim 30$  MHz - 80 MHz), the reflection is always lower for the SMD-balun within the bandwidth.

Thus the reflection measurements support the choice of the SMD-baluns for the Small Black Spider. However, it should be verified that they also yield better results when they are mounted in the antenna.

### 8.5 Comparison to the Black Spider LPDA

A prototype of the Small Black Spider was recently completed. A photograph of this prototype is presented in fig. 8.9. It complies to a large extend with the design specifications discussed above. A difference is that still the baluns of the Black Spider are used instead of the planned application of SMD-baluns.

A reflection measurement of the prototype was performed to give a first impression of the electrical performance of the new antenna. In fig. 8.8, the recorded reflection is compared to the measurement for the Black Spider discussed in chapter 7.1.2.

Overall, a similar progression of the reflection, but also variations are observed. An about 2 MHz larger bandwidth compared to the Black Spider, ranging from 29 MHz to 83 MHz can be determined. The resonance dips of both reflection curves refer



Figure 8.7: Reflection measurement of the Mini-Circuits type TC4-19+ SMD-balun, which is intended be used in the Small Black Spider, in comparison to the balun used in the Black Spider LPDA.

mostly to equal resonance frequencies. For the Small Black Spider these dips are more distingtive. This reduces the total radiated power within the bandwidth (see fig 8.8) for about 5 % compared to the Black Spider. However, still 89.5 % of the input power is radiated within the bandwidth.

The extend to the lower frequencies of about 3 MHz compared to the Black Spider can be partially explained by the smaller shortening factor which reduces the resonance frequency of the lowest dipole. For the lower dipole of the Small Black Spider made of aluminum tubes with a diameter of 30 mm and a lenght of 4.25 m, the shortening factor can be estimated to 0.93 with fig. appendix A.2. This yields a reduction of the resonance frequency of the lowest dipole of 1.4 MHz via eqn. 5.35 compared to the Black Spider.

However, the enlarged bandwidth of the Small Black Spider utilizes the frequency window for the observation of radio signals at the Auger site even better (see. fig. 7.16) than the Black Spider.

A significant difference in the reflection can be noticed for frequencies above about 83 MHz. The sensitivity of the Small Black Spider declines steeply above this frequency whether the Black Spider shows a flat decline. This steep decline agrees with the behaviour predicted by simulations (see fig. 7.8).

This is an advantage of the Small Black Spider as it will efficiently surpress the strong noise sources of the FM-transmitters between 88 MHz and 105 MHz.



Figure 8.8: Comparison of the reflection of Small Black Spider and Black Spider. The total reflected power within the bandwidth in percent of the input power is denoted in the legend respectively.



Figure 8.9: Photograph of the prototype antenna of the Small Black Spider.

# 9. Summary and Outlook

Radio detection of cosmic ray induced air showers has again become a vivid field of research during the last years. It promises detailed informations of EAS through calorimetric measurements and a duty cycle of 100 %. A great effort for a self-triggered antenna array within the Pierre-Auger-Observatory has recently been decided. The AERA detector, covering an area of 20  $km^2$ , is the first attempt to establish the radio detection technique on large scales in the EeV energy regime.

Since they are the first link in the detector chain, the used antennas play a key role for the successful detection of cosmic rays. The LPDAs have proven their ability to detect radio pulses induced by EAS in various test setups at the experiment site.

We introduced the design of the Black Spider LPDA as a consequence of the environmental demands of the radio detection at the Auger site.

The electrical layout of the Black Spider provides the basis for the new antenna for AERA.

In order to reach a deeper understanding of the mode of operation of the antenna hardware, we investigated the interactions of the single components of the Black Spider during a successive assembly of the antenna. A measurement of the voltage along the waveguide of the Black Spider gave an insight to the electric transactions within the antenna structure.

We determined those basic antenna properties, which are required for the reconstruction of EAS.

To investigate the gain of the Black Spider, various gain patterns were simulated with respect to relevant influences at the experiment site. The patterns were found to be overall stable with the frequency and the height of the antenna above ground. Furthermore, we observed that the variation of ground conditions is expected to have no negative influence on the quality of the measurements of EAS.

A vertical beamwidth between  $150^{\circ}$  and  $170^{\circ}$  was derived from the various gain simulations. Additionally the horizontal directional diagram was measured. A good agreement with the corresponding simulation was noticed.

By reflection measurements, a total bandwidth of 52 MHz ranging from 32 MHz to 84 MHz was observed. This range covers the complete frequency window required for the detection of EAS and remains essentially free from man made noise sources at the site.

The impedance of the antenna was measured to be on average  $50.7 + i0.3 \Omega$  within the bandwidth. Thus a good impedance matching to a pure resistive source impe-

dance of 50  $\Omega$  is reached.

We measured the group delay of the antenna and found a decline of 88 ns within the bandwidth.

The Black Spider was benchmarked in various test setups. A stress test of the dipole attachment to the waveguide was performed in the laboratory. No damage was observed after more than six month of testing.

Several Black Spiders were exposed to the weather conditions at the experimental site for more than a year without maintenance needed. No damage and no reduction of the electrical performance could be observed.

The variation of the performance between different Black Spiders was investigated through reflection measurements and was found to be in the order of only a few percent.

The sensitivity of the antenna has been investigated through measurements of the galactic noise in Argentina and Nancay, France. Here, the antenna has proven its ability to detect the galactic background in general and its variation due to the transit of the galactic center.

We presented a new antenna design optimized for the AERA-detector considering the requirements of mass production. The Small Black Spider employs the lower arms as dipoles at the same time, so that the size of the outer structure is reduced. CNC-machined components are used as far as possible which increases the reproducibility of the antenna properties and reduces the costs for a production in large quantities. The application of an improved waveguide and SMD-baluns enables a faster assembly which yields low costs as well.

Furthermore, the Small Black Spider provides an integrated housing for the low noise amplifiers and optionally the installation of filter chains directly within the antenna.

As the electrical layout of the Small Black Spider is basically identical to that of the Black Spider, the new antenna for AERA is expected to be consistent with the antenna properties determined in this thesis to a large extent.

# A. Appendix

SUHNER SWITZERLAND

### HUBER+SUHNER<sup>®</sup> DATA SHEET Coaxial Cable: RG\_58\_C/U

#### Description Single screened coaxial cable Technical Data

#### Construction

Centre conductor Dielectric Outer conductor Jacket Print	Material Copper, Tin plated PE (Polyethylene) Copper, Tin plated PVC (Polyvinyl Chloride) HUBER+SUHNER RG 58 C/U	Detail Strand-19 Braid, 96 % RAL 9005 - bk 50 Obm (PA no.)	Diame 0.9 2.95 3.6 4.95	eter mm mm mm mm +/- 0.15
Electrical Data	HOBER GOINER ROOD OF	50 Onim (PA 110.)		
Impedance Max. operating frequency Capacitance Velocity of signal propagation Signal delay Insulation resistance Min. screening effectiveness Max. operating voltage Test voltage			50 1 100.7 66 5.03 ≥1 x >38 2.5 5	Ω +/-2 GHz pF/m % ns/m 10 <sup>8</sup> MΩm dB (up to 1 GHz) kVrms (at sea level) kVrms (50 Hz/1 min)

Matrix

Attenuation [ formula: (a\*f\*0.5 + b\*f) ] and Power CW [ formula: (p\*/ f\*0.5) ]

Coe	fficients:						
a=	0.3862	b=	0.1935	f <sub>max.</sub> =	1	p <sub>at</sub> <sub>1GHz</sub> =	105
	Frequency	<u> </u>	Nom. attenuation		Nom. attenuation		Max. CW power
	(GHz)		(dB / m)		(dB / ft)		(watt)
			sea level 25° C ambient temperature		sea level 25° C ambient temperature		sea level 40° C ambient temperature
	0.05	-	0.10		0.030		470
	0.10		0.14		0.043		332
	0.15		0.18		0.055		271
	0.20		0.21		0.064		235
	0.25		0.24		0.073		210
	0.30		0.27		0.082		192
	0.35		0.30		0.091		177
	0.40		0.32		0.098		166
	0.45		0.35		0.107		157
	0.50		0.37		0.113		148
	0.55		0.39		0.119		142
	0.60		0.42		0.128		136
	0.65		0.44		0.134		130
	0.70		0.46		0.140		125
	0.75		0.48		0.146		121
	0.80		0.50		0.152		117
	0.85		0.52		0.158		114
	0.90		0.54		0.165		111
	0.95		0.56		0.171		108

Figure A.1: Datasheet of a RG58 C/U coaxial cable



Figure A.2: Shortening factor V of a  $\lambda/2$ -dipole, adapted from [36]



Figure A.3: Comparison of the reflection of the EW-polarization of eight different Black Spider LPDAs mounted at different poles of the Auger Radio setup.



Figure A.4: Comparison between the reflection of the Black Spider and the Big Alu-LPDA. The Big Alu-LPDA was measured demounted e.g. only one separate polarization plane was measured, the other plane was removed.

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# Erklärung

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Aachen, den 29. Juni 2009

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