Application Report
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ISM-Band and Short Range Device Antennas

Matthew Loy / Iboun Sylla (Editors)

High-Speed and RF

ABSTRACT

This application report discusses antenna fundamentals and the various types of antennas used for short range devices. Fundamentals are presented along with practical design principals.

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1 Antenna Basics

1.1 Fundamental Definitions

Antennas are the connecting link between RF signals in an electrical circuit such as a PCB and an electromagnetic wave propagating in the transmission media between the transmitter and the receiver of a wireless link.

In the transmitter, the antenna transforms the electrical signal into an electromagnetic wave by exciting either an electrical or a magnetic field in its immediate surroundings, the near field. Antennas that excite an electrical field are referred to as electrical antennas; antennas exciting a magnetic field are called magnetic antennas correspondingly. The oscillating electrical or magnetic field generates an electromagnetic wave that propagates with the velocity of light c. The speed of light in free space c_0 is 300000 km/s. If the wave travels in a dielectric medium with the relative dielectric constant ε_r, the speed of light is reduced to:

$$c = \frac{c_o}{\sqrt{\epsilon_r}}$$

We can calculate the wavelength from the frequency f of the signal and the speed of light c using the formula:

$$\lambda = \frac{c}{f}$$

Using common units, the equation:

$$\text{wavelength in meters} = \frac{300}{\text{frequency in MHz}}$$

is often used for the wavelength in free space. If the wave travels in a dielectric medium, for instance in the PCB material, the wavelength has to be divided by the square root of ε_r.

We can distinguish three field regions where the electromagnetic wave develops: reactive near field, radiating near field and far field:

- In the reactive near field, reactive field components predominate over the radiated field. This means that any variations in the electrical properties (for electrical antennas) or magnetic properties (for magnetic antennas) have a strong influence on the antenna’s impedance at the antenna feed point. The distance from the antenna to the boundary of the reactive near field region is commonly assumed as:

$$R_1 = \frac{\lambda}{2 \times \pi}$$

- In the radiating near field the radiated field predominates, the antenna impedance is only slightly influenced by the surrounding media in this region. But the dimensions of the antenna can not be neglected with respect to the distance from the antenna. This means that the angular distribution of the radiation pattern is dependent on the distance. For measurements of the radiation pattern, the distance from the antenna should be larger than the radiating near field boundary, otherwise the measured pattern will be different from that under real life conditions. The diameter of the radiating near field is $R_2 = \frac{2 \times D^2}{\lambda}$ with D as the largest dimension of the antenna.
- For distances larger than R_2, the radiation pattern is independent on the distance; we are in the far field region. The distance between transmitter and receiver antennas in a practical application is usually in this region.

In the receiver, the antenna gathers energy from the electromagnetic wave and transforms it into an electrical voltage and current in the electrical circuit. For better comprehension, the antenna parameters are often explained on a transmit antenna, but in most cases, if no nonlinear ferrites are involved, the characteristics of an antenna are identical in receive and transmit modes.

1.2 Antenna Characteristics

Polarization describes the trace that the tip of the electrical field vector builds during the propagation of the wave. In the far field, we can consider the electromagnetic wave as a plane wave. In a plane electromagnetic wave, the electrical and the magnetic field vectors are orthogonal to the direction of propagation and also orthogonal to each other. In the general case, the tip of the electrical field vector moves along an elliptical helix, giving an elliptical polarization. The wave is called right-hand polarized if the tip of the electrical field vector turns clockwise while propagating; otherwise it is left-hand polarized.

If the two axis of the ellipse have the same magnitude, the polarization is called circular. If one of the two axis of the ellipse becomes zero, we have linear polarization, vertical if the electrical field vector oscillates perpendicularly to ground, horizontal if its direction of oscillation is parallel to the ground plane.

A transmission system has the best performance (ideal case) when the polarization of the transmitter and the receiver antenna are identical to each other. Circular polarization on one end and linear polarization on the other gives 3-dB loss compared to the ideal case. If both antennas are linearly polarized but 90° turned to each other, theoretically no power is received. The same phenomenon happens if one antenna is right-hand circularly polarized and the other one left-hand circularly polarized.

In an indoor environment, reflections in the transmission path may change the polarization, which makes the polarization of the received wave difficult to predict. If one of the antennas is portable, we have to make sure that it works in any position. Circular polarization at one end and linear polarization at the other end results in a principal loss of 3 dB, but avoids the case of a total blackout, where no power is received.

For the description of the radiated power and the gain of the antenna often the concept of the isotropic radiator is used. The isotropic radiator is a hypothetical antenna, which radiates the supplied RF power equally in all directions. The power density at a distance r from the isotropic radiator is therefore the supplied power divided by the area of a sphere with the radius r.

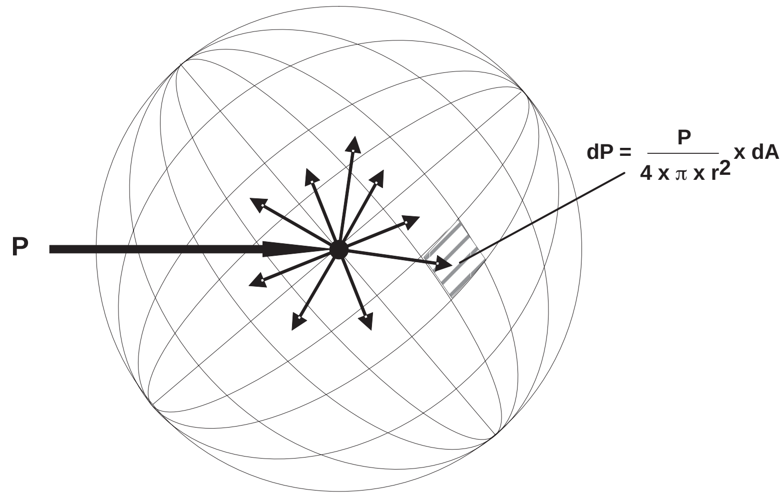


Figure 1. Isotropic Radiator

If we measure the power density in some distance from a device under test, the effective isotropic radiated power, EIRP, is the power which we would have to supply to an isotropic radiator in order to get the same power density in the same distance. The EIRP describes the power radiation capability of a device including its antenna.

From the EIRP, we can calculate the electrical field strength at a given distance from the radiator, which is specified in some government or regional regulations. The density of the radiated power D (in W/m^2) measured in the distance r from an isotropic radiator radiating the total power EIRP is the radiated power divided by the surface area of the sphere with the radius r:

$$D = \frac{dP}{dA} = \frac{EIRP}{4 \times \pi \times r^2}$$

The relationship between the electrical field strength and the power density is the same as between voltage and power in an electrical circuit $V = \sqrt{P \times R}$.

With the impedance of free space $Z_0 = 377\ \Omega = \pi\ 120\ \Omega$, the rms value of the electrical field strength is then:

$$E = \sqrt{D \times Zo} = \sqrt{D \times \pi \times 120\ \Omega}$$

This gives:

$$E = \sqrt{\frac{EIRP \times \pi \times 120\ \Omega}{4 \times \pi \times r^2}} = \frac{1}{r} \times \sqrt{30\ \Omega \times EIRP}$$

Or:

$$\mathrm{EIRP} = \frac{E^2 \times r^2}{30\ \Omega}$$

Taking the logarithm on both sides gives the EIRP value in dBm:

$$\mathrm{EIRP}_{[dBm]} = E_{[dB\mu V/m]} + 20\ \log r_{[meters]} - 10 \times \log 30 - 90\ \mathrm{dB}$$

In standard test setups, the electrical field strength is often measured at a distance of 3 m. In this case we can use the simple formula:

$$\mathrm{EIRP}_{[dBm]} = E_{[dB\mu V/m]} - 95.23\ \mathrm{dB}$$

As opposed to the (only hypothetical) isotropic radiator, real antennas exhibit more or less distinct directional radiation characteristics.

The radiation pattern of an antenna is the normalized polar plot of the radiated power density, measured at a constant distance from the antenna in a horizontal or vertical plane.

The isotropic gain G_{iso} of an antenna indicates how many times the power density of the described antenna in the main direction of propagation is larger than the power density from an isotropic radiator at the same distance. Antenna gain does not imply an amplification of power; it comes only from the bundling of the available radiated power in certain directions.

The radiation resistance (R_r) relates the power radiated from the antenna to the RF current fed into the antenna. For the same RF current, a resistor with the resistance R_r would dissipate exactly the same power into heat that the antenna radiates. R_r can be calculated from:

$$R_r = \frac{P_{radiated}}{I^2}$$

The radiation resistance is part of the impedance of the antenna at its feed point. Additionally, we have the loss resistance R_{loss} which accounts for the power dissipated into heat as well as reactive components L and C. Figure 2 has an equivalent circuit that describes the antenna around its resonant frequency.

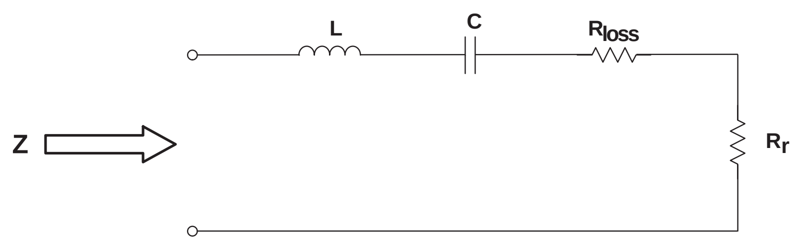


Figure 2. Antenna Equivalent Circuit

The inductor and the capacitor in the equivalent circuit build a series resonant circuit. The antenna impedance Z is:

$$Z = R_r + R_{loss} + j\left(2 \times \pi \times f \times L - \frac{1}{2 \times \pi \times f \times C}\right)$$

At the frequency of resonance, $f_{res} = \frac{1}{2 \times \pi \times \sqrt{L \times C}}$, the reactance's of the capacitor and the inductor cancel out each other; only the resistive part of the antenna impedance is left over. The inductance L and the capacitance C in the equivalent schematic are determined by the antenna geometry. If we want to build an antenna for a given frequency, we have to find a geometry (for example a wire with a certain length) that is resonant at the frequency of operation.

At the frequency of resonance the antenna input impedance equals $R_r + R_{loss}$. The antenna efficiency η in resonance is the ratio of the radiated power to the total power accepted by the antenna from the generator:

$$\eta = \frac{R_r}{R_r + R_{loss}}$$

At frequencies other than the resonant frequency, the antenna input impedance is either capacitive or inductive. This phenomenon is why it is possible to tune an existing antenna by adding a series capacitor or inductor.

The L-to-C ratio determines the bandwidth of the antenna for given radiation and loss resistances. For the same resistance values, a larger L-to-C ratio means a higher quality factor Q and a smaller bandwidth. The values of L and C in the equivalent schematic depend on the antenna geometry; often we can deduct intuitively how a variation of the geometry influence L and C. The quality factor is influenced by a contribution Q_{rad} from the radiation resistance and Q_{loss} from the loss resistance. The overall Q of the antenna is:

$$\frac{1}{Q} = \frac{1}{Q_{rad}} + \frac{1}{Q_{loss}}$$

Chu /1/ and Wheeler /2/ gave the theoretical limit for the quality factor Q and the fractional bandwidth of a lossless antenna as:

$$BW_{lossless} = \frac{1}{Q_{rad}} = \left(\frac{2 \times \pi \times a}{\lambda}\right)^3$$

with a as the radius of the smallest circumscribing sphere surrounding the antenna.

The selectivity of the antenna can help to suppress unwanted out of band emissions; but not always a small bandwidth is desirable. A small bandwidth means stringent requirements on the tolerances of the matching components and the antenna itself. For a given dimension of a small antenna, we can only increase the bandwidth if we introduce intentional losses. The bandwidth of an antenna with the efficiency η is then:

$$BW = \left(\frac{2 \times \pi \times a}{\lambda}\right)^3 \times \frac{1}{\eta}$$

The product of the bandwidth and the efficiency is a constant for a given antenna dimension. If we want to gain one, we have to sacrifice from the other.

1.3 Reflection, Matching, and Tuning

What happens if we connect a transmit antenna to a transmission line with the characteristic impedance $\overline{Z}_O$ (usually 50 Ω) and send a signal with the amplitude $\overline{V}_{IN}$ into the transmission line? In most cases, the antenna impedance $\overline{Z}$ will not be exactly the same as the transmission line impedance $\overline{Z}_O$. Then only a part of the incident wave will be transmitted to the antenna with an amplitude of $\overline{V}_{accept}$, while the remaining part will be reflected back to the generator with an amplitude of $\overline{V}_{refl}$.

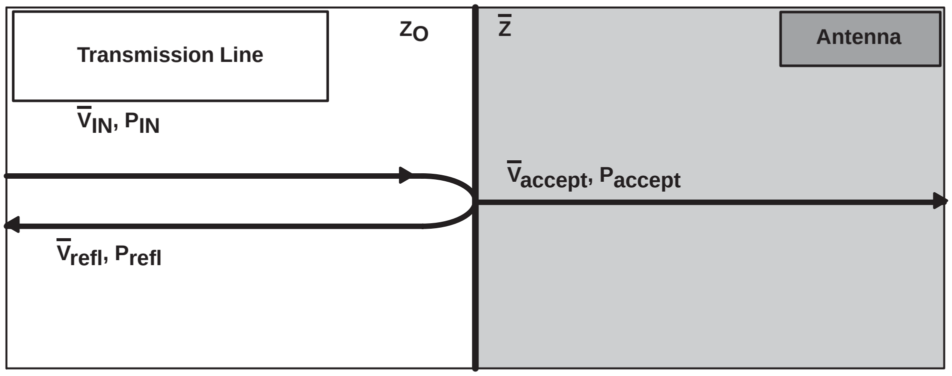


Figure 3. Reflection at a Discontinuity

The complex reflection coefficient $\overline{\Gamma}$ is defined as the ratio of the reflected wave's amplitude (e.g. voltage, current, or field strength) to the amplitude of the incident wave. We can calculate the reflection coefficient from the impedances of the antenna $\overline{Z}$ and the transmission line $\overline{Z}_O$:

$$\overline{\Gamma} = \frac{\overline{Z} - Z_O}{\overline{Z} + Z_O}$$

For an arbitrary complex load impedance $\overline{Z}$, the phase difference between the reflected and the incident wave may be anywhere in the range between 0 and 2π. The reflection coefficient is therefore a complex quantity. If we want to minimize the reflection loss, we must know the magnitude and phase angle of the reflection coefficient. To measure this, a vector network analyzer is needed. If the source is not a transmission line but the output of an IC, then the source impedance can be a complex quantity. The reflection coefficient is zero if $\overline{Z}$ equals $\overline{Z}^*_O$, the complex conjugate of the source impedance. In this case, all incident energy is accepted by the antenna; so we call the antenna perfectly matched.

The power ratio of the reflected to the incident wave is called the return loss (RL). The return loss tells us how many dB the power of the reflected wave is below the power of the incident wave. A perfectly matched antenna has an infinite return loss, because no power is reflected, all the power is accepted. The power of the accepted wave is smaller than the power of the incident wave by the amount of the mismatch loss (ML). The mismatch loss directly describes the impact of the usually unwanted reflection on the power radiated by the antenna. We can calculate return loss and mismatch from the reflection coefficient using the formulas:

$$RL = 10 \times \log \frac{P_{in}}{P_{refl}} = -10 \times \log |\overline{\Gamma}|^2 = -20 \times \log |\overline{\Gamma}|$$

$$ML = 10 \times \log \frac{P_{in}}{P_{accept}} = -10 \times \log(1 - |\overline{\Gamma}|^2$$

If we measure the voltage on a transmission line, we cannot distinguish between the incident and the reflected waves; we only see the sum of both. At some locations, both waves interfere constructively, at some other locations they partially cancel out each other.

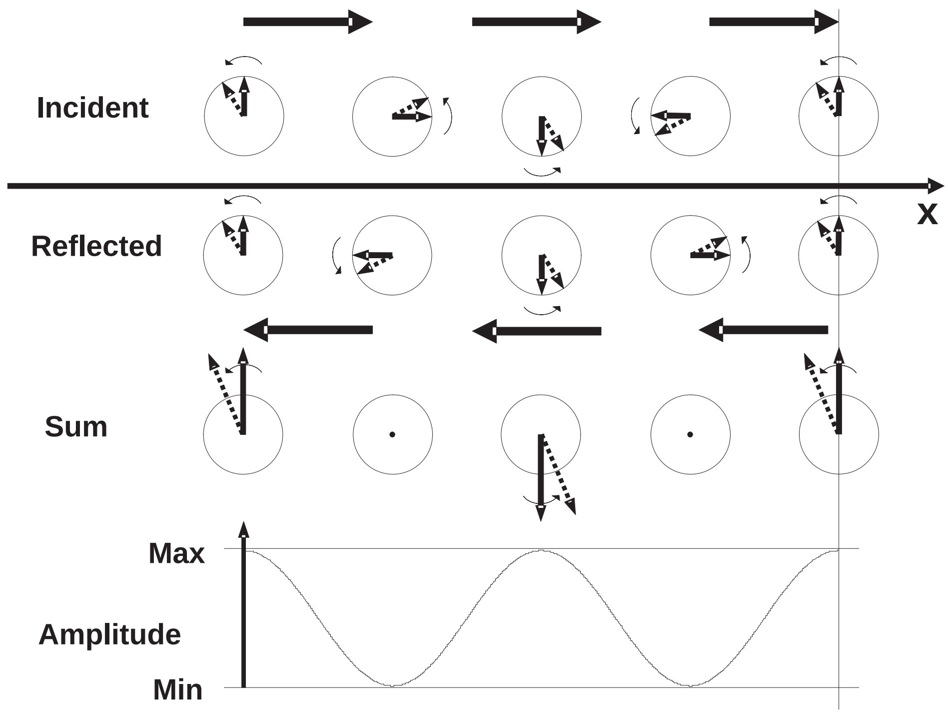


Figure 4. Standing Waves Due to Reflection

As we can see from Figure 4, the locations where maximum and minimum in the amplitude of the sum occur do not move; the incident and the reflected wave build a standing wave. The larger the amplitude of the reflected wave is the more pronounced the standing wave pattern will be. The voltage standing wave ratio (VSWR) is defined as the ratio of the maximum to the minimum voltage of the standing wave pattern and can be calculated from the magnitude of the reflection coefficient:

$$VSWR = \frac{V_{max}}{V_{min}} = \frac{1 + |\overline{\Gamma}|}{1 - |\overline{\Gamma}|}$$

The numerical value of VSWR is in the range between 1 (ideally matched load, no standing wave) and ∞ ($|\overline{\Gamma}| = 1$, total reflection or complete mismatch).

VSWR, $\overline{\Gamma}$, RL, and ML describe the same phenomenon of reflection and can be transformed into each other. While VSWR and RL are related to the amplitude of the reflected wave only, $\overline{\Gamma}$ contains the phase information too, as $\overline{\Gamma}$ is a complex quantity.

Often the antenna has an impedance different from that of the feeding transmission line. To minimize the mismatch loss, we have to transform one impedance to the complex conjugate of the other. A powerful tool that helps to determine the needed matching circuit is the Smith Chart. Basically, the Smith Chart plots the reflection coefficient $\overline{\Gamma}$ in the complex plane. For passive circuits, the length of the $\overline{\Gamma}$-phasor varies between 0 (ideal match) and 1 (complete mismatch). The phase difference f between the reflected and the incident wave may assume any value between 0 and 2π. Therefore, all possible $\overline{\Gamma}$-phasors (for passive circuits) are within a circle with the radius 1, which defines the outer boundary of the Smith Chart.

The reflection coefficient is +1 if the end of a transmission line is left open, −1 for a short at the end of the transmission line. An inductive load gives a reflection coefficient in the upper half, a capacitive load in the lower half of the Smith Chart. Any capacitors or inductors added to a given load move the reflection coefficient in the Smith Chart on circles: series components on a circle that goes through the open point at +1, parallel components on circles through the shortcut point at −1. Inductors shift the reflection coefficient upwards towards the inductive half; capacitors shift it downwards towards the capacitive half. Figure 5 shows how series or parallel inductors and capacitors influence a given reflection coefficient.

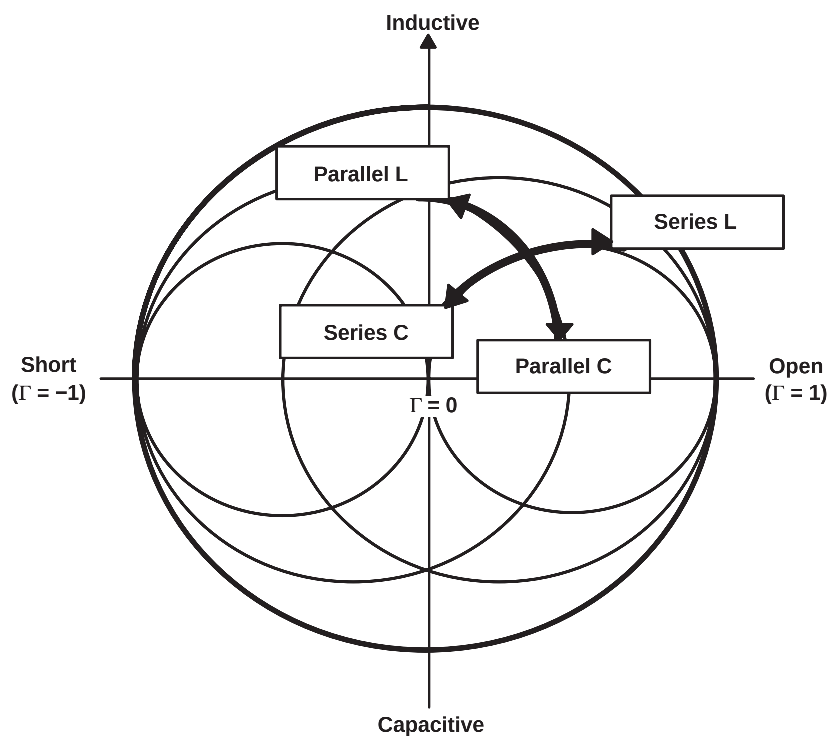


Figure 5. Series and Parallel Capacitors and Inductors in the Smith Chart

Using the Smith Chart we can find out what kind of components are needed to minimize the reflection coefficient for a given antenna impedance. In Figure 5 for example, a series capacitor could move the reflection coefficient on the circle through the open point (because it is a series component) towards the lower half of the Smith Chart (because it is a capacitor). The center of the Smith Chart (where the reflection coefficient is zero) can be reached by a proper capacitance value giving the perfect match.

For a system environment normalized to 50 Ω, the center of the Smith Chart is 50 Ω.

2 Antenna Types and Their Features

2.1 Half-Wave Dipole

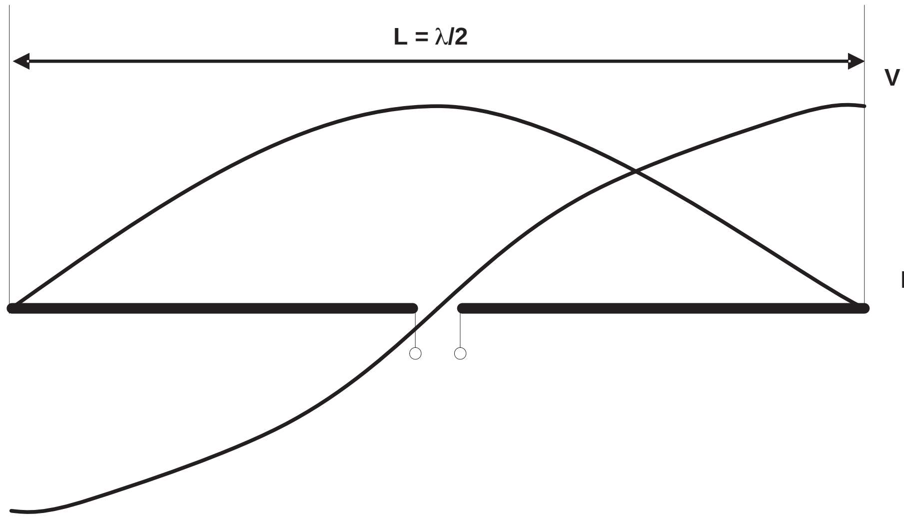


Figure 6. Half-Wave Dipole Antenna

The half-wave dipole antenna is the basis of many other antennas and is also used as a reference antenna for the measurement of antenna gain and radiated power density.

At the frequency of resonance, i.e. at the frequency at which the length of the dipole equals a half-wavelength, we have a minimum voltage and a maximum current at the terminations in the center of the antenna; the impedance is minimal. Therefore, we can compare the half-wave dipole antenna with a series RLC resonant circuit as given in Figure 2. For a lossless half-wave dipole antenna, the series resistance of the equivalent resonant circuit equals the radiation resistance, generally between 60 Ω and 73 Ω, depending on the ratio of its length to the diameter.

The bandwidth of the resonant circuit (or the antenna) is determined by the L-to-C ratio. A wire with a larger diameter means a larger capacitance and a smaller inductance, which gives a larger bandwidth for a given series resistance. That’s why antennas made for measurement purposes have a particularly large wire diameter.

As opposed to the (only hypothetical) isotropic radiator, real antennas such as the half-wave dipole have a more or less distinct directional radiation characteristic.

The radiation pattern of an antenna is the normalized polar plot of the radiated power density, measured at a constant distance from the antenna in a horizontal or vertical plane.

Figure 7 shows the radiation pattern of a half-wave dipole antenna.

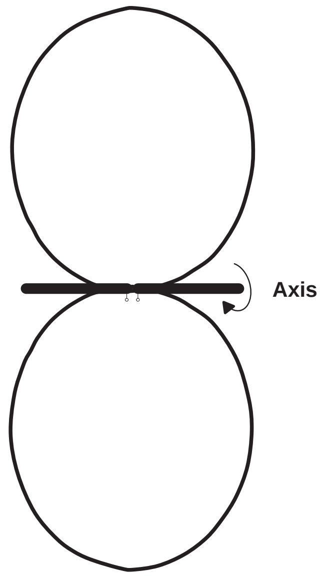


Figure 7. Radiation Pattern of a Half-Wave Dipole Antenna

As the dipole is symmetrical around its axis, the three-dimensional radiation pattern rotates around the wire axis.

The isotropic gain of a half-wave dipole antenna is 2.15 dB, i.e. in the direction perpendicular to the wire axis the radiated power density is 2.15 dB larger than that of the isotropic radiator. There is no radiation in the wire axis. The half-wave dipole produces linear polarization with the electrical field vector in line with, or in other words parallel to, the wire axis.

Because the half-wave dipole is often used as a reference antenna for measurements, sometimes the gain of an antenna is referenced to the radiated power density of a half-wave dipole instead of an isotropic radiator. Also the effective radiated power ERP, which is the power delivered to an ideal dipole that gives the same radiation density as the device under test, is used instead of the EIRP. The relations $G_{dipole} = G_{isotropic} - 2.15$ dB and ERP = EIRP − 2.15 dB can be applied.

The half-wave dipole needs a differential feed because both of its terminations have the same impedance to ground. This can be convenient if the transmitter output or the receiver input have differential ports. A balun will be used along with the half-wave dipole in case of single ended transmitters or receivers or if an antenna switch is used.

For external ready-manufactured half-wave dipoles, the balun is visually built-in to the antenna and provides a single-ended interface.

The half-wave dipole is an electrical antenna. This means that it is easily detuned by materials with a dielectric constant larger than one within its reactive near field. If for instance the housing of a device is in the reactive near field, the housing has to be present when the antenna is matched. The human body has a large dielectric constant of approximately 75; if an electrical antenna is worn on the body or held in the hand, it can be strongly detuned.

If the antenna is built as two traces on a PCB, the dielectric constant of the PCB material has to be considered. The electrical field in the reactive near field region spreads out partially into the PCB material, partially into the surrounding air. This gives an effective dielectric constant e_{eff} between that of the air and the PCB material /3/:

$$\epsilon_{eff} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \times \left[\frac{1}{\sqrt{1 + \frac{12 \times h}{w}}} + 0.04 \times \left(1 - \frac{w}{h}\right)^2 \right]$$

Where h is the thickness of the PCB material, w is the trace width of the dipole arms. The required length of the half-wave dipole is then:

$$L_{HW\_PCB} = \frac{L_{FreeSpace}}{\sqrt{\epsilon_{eff}}} = \frac{\lambda/2}{\sqrt{\epsilon_{eff}}}$$

Underneath the dipole and within the reactive near field, no ground plane is allowed.

2.2 Quarter-Wave Monopole

In many cases, the half-wave dipole is just too large. Also, the needed differential feed is often a disadvantage. If we replace one branch of the dipole antenna by an infinitely large ground plane, due to the effect of mirroring, the radiation pattern above the ground plane remains unaffected. This new structure is called a monopole antenna.

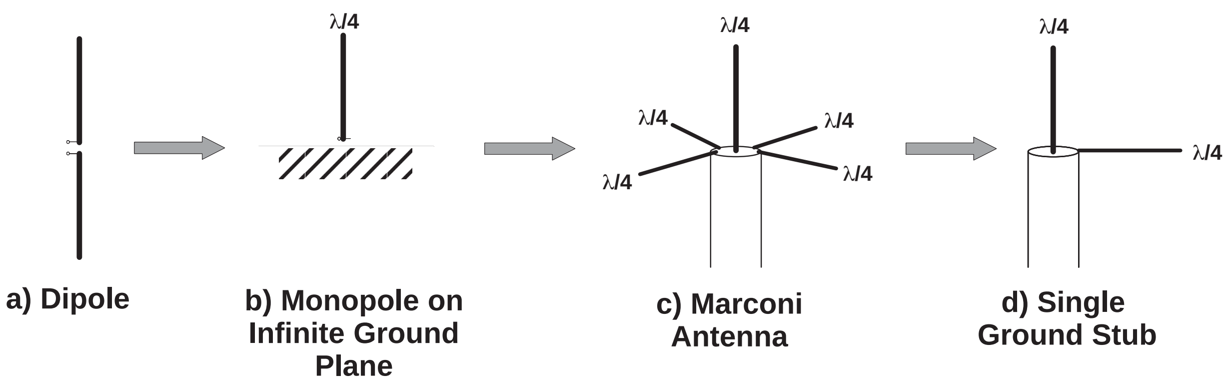


Figure 8. Building Up the Quarter-Wave Monopole

Because all the radiated power is now concentrated in the half-space above the ground plane, the gain of the monopole is 3 dB larger than the gain of the dipole.

Often a large ground plane is not feasible. The Marconi antenna replaces the (not realizable) infinitely large ground plane by several open-ended λ/4-Stubs, called the counterpoise.

A further reduction to only one stub gives a structure that looks like a bent dipole antenna. When designing a monopole antenna, the radiator should go as long as possible perpendicular to the ground stub or the ground plane. Bends close to the feeding point reduce the radiation resistance and the efficiency of the antenna.

The ideal quarter-wave monopole has a linear polarization with the vector of the electrical field in the wire axis. If the ground plane becomes unsymmetrical, the direction of polarization will be tilted towards the larger part of the ground plane, but still remains linear.

The radiation resistance of an ideal quarter-wave monopole is half of that of a dipole; depending on the ratio of length to diameter of the radiator between 30 Ω and 36.5 Ω.

Like the half-wave dipole, the quarter-wave monopole is an electrical antenna; it is influenced by the dielectric constant of the material in the reactive near field. The same formulas for the effective dielectric constant and the required length as for the half-wave dipole hold for the quarter-wave monopole.

Table 1 gives the length of half-wave dipoles and quarter-wave monopoles in free space and on a PCB for commonly used short range frequencies. For the PCB antennas, a PCB thickness of h = 1,5 mm and a trace width of w = 1 mm has been assumed; the PCB material is FR4 with $\varepsilon_r = 4.2$. This gives an effective dielectric constant of $\varepsilon_r = 2.97$.

Table 1. Length of Half-Wave Dipoles and Quarter-Wave Monopoles inFree Space and On the PCB

Frequency	Half-Wave Dipole L in Free Space	Half-Wave Dipole L on a PCB	Quarter-Wave Monopole L/2 in Free Space	Quarter-Wave Monopole L/2 on a PCB
315 MHz	476 mm	276 mm	238 mm	138 mm
434 MHz	346 mm	201 mm	173 mm	100,5 mm
868 MHz	173 mm	100,3 mm	86,5 mm	50,15 mm
915 MHz	164 mm	95 mm	82 mm	47,5 mm

It has to be mentioned that parasitic components such as capacitance to ground, inductance introduced by bends in the antenna as well as the influence of the package alter the antenna impedance. For monopole antennas, the ground plane is sometimes smaller than a quarter-wave length or not perpendicular to the radiator. In practice, the exact length of the dipole or the monopole has therefore to be determined by measuring the feed impedance with a vector network analyzer.

Sometimes the available space limits the length of an antenna; the antenna is made as long as the geometry permits, which can be smaller than one quarter wavelength. A monopole shorter than a quarter-wave length can be considered as a quarter-wave monopole, which is used at a frequency lower than the frequency of resonance. According to the equivalent schematic given in Figure 2, the input impedance at the frequency of operation will then be a series connection of a resistor and a capacitor. The series capacitance can be resonated out by a series inductor. A monopole antenna shorter than λ/4 with a series inductor is also referred to as a loaded stub antenna.

The radiation resistance of a loaded stub decreases with decreasing length. The smaller radiation resistance and the larger L-to-C ratio increase the quality factor and make the bandwidth smaller than for a quarter-wave monopole. Approximations for the radiation resistance of a monopole antenna are:

$$R_r = 395\,\Omega \times \left(\frac{L}{\lambda}\right)^2 \qquad \text{for } 0 < L < \frac{\lambda}{8}$$

$$R_r = 1.22\,k\Omega \times \left(\frac{L}{\lambda}\right)^{2.5} \qquad \text{for } \frac{\lambda}{8} < L < \frac{\lambda}{4}$$

At the frequency of operation (i.e., resonance), the impedance of the short stub will be that of a small resistor (radiation resistance plus loss resistance) with a series capacitor. From the Smith Chart in Figure 9 we can see that matching to a 50-Ω source can be achieved by a series inductor and a parallel capacitor.

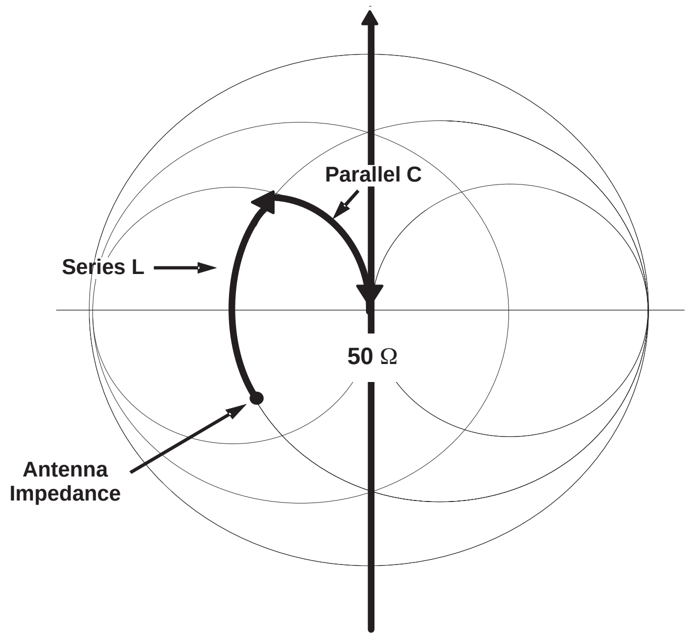


Figure 9. Matching of a Short Loaded Stub Antenna

Figure 10 shows an example of a loaded stub PCB antenna with matching components.

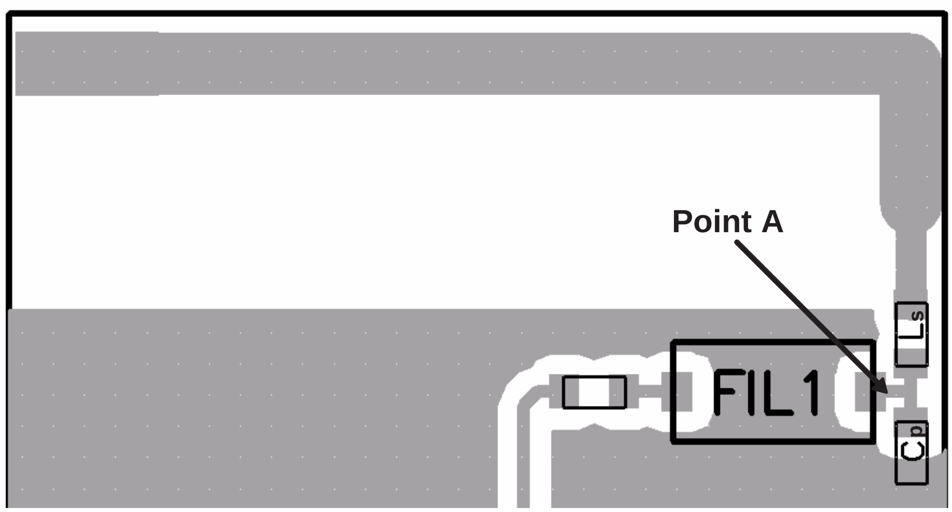


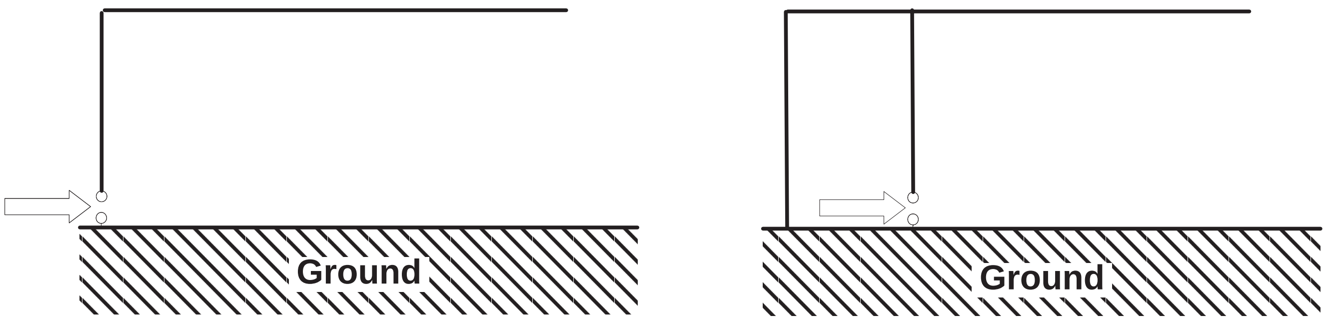
Figure 10. Loaded Stub PCB Antenna With Matching Components

The series inductor and the parallel capacitor transform the antenna impedance to 50 Ω, the input impedance of the filter (FIL1).

For dipole or monopole antennas, the component values for the series inductor and the parallel capacitor (C_P) have to be determined by measuring the feed impedance at point A in Figure 10 (with L_S = 0-Ω resistor and C_P left unpopulated) with a vector network analyzer.

Once this is determined, we can use a Smith Chart to assist in matching the antenna to 50 Ω using L_S and C_P.

Derivatives of the monopole are the inverted-L and inverted-F antennas as shown in Figure 11.



a) Inverted-L Antenna b) Inverted-F Antenna

Figure 11. Inverted-L Antenna and Inverted-F Antenna

In the inverted-L antenna, the monopole does not run perpendicularly to the ground plane over its whole length but is bent parallel to the ground plane after some distance. This helps to save space, but decreases the radiation resistance because the radiator comes closer to the ground plane. An additional matching circuit is needed to match the low-feed impedance to the usual transmission line impedance of 50 Ω.

If we proceed from the feed point of the inverted-L antenna to the end, we notice that the voltage increases (while the current decreases) from a maximum voltage value at the feeding point to almost zero at the end. This means, that the antenna impedance has its minimum if we feed the antenna as shown in Figure 11a) and increases if we move the feeding point towards the end. The inverted-F antenna in Figure 11b) is an inverted-L antenna with a feeding tap that gives larger antenna impedance. If the antenna is tapped at the right location, no additional matching circuit is required.

The structure of inverted-F antennas and in particular the location of the tap is usually determined by electromagnetic simulations.

2.3 Transversal Mode Helical Antenna

Another option to reduce the size of a monopole is to coil it up into a helix as shown in Figure 12.

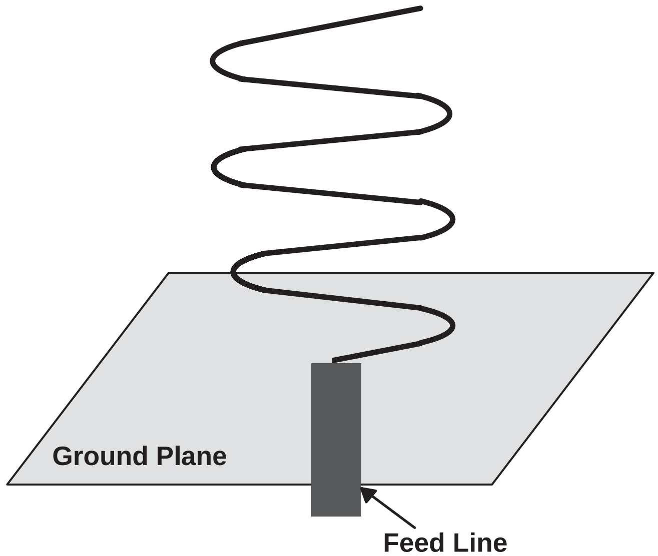


Figure 12. Helix Antenna on a Ground Plane

When the coil circumference and the spacing between adjacent turns are comparable to the wavelength, the antenna radiates a circular polarized beam in the axis of the helix. These antennas are called axial mode helicals.

In small short-range applications, the helix diameter and the spacing between turns are much smaller than a wavelength; the result is a normal mode helical antenna. The radiation pattern of a normal mode helix is similar to that of a monopole; the maximum radiation occurs perpendicular to the helix axis. Due to the shape and the size of the ground plane, radiation patterns of practical antennas can show deviations from this idealized form. The radiation from a normal mode helix is elliptically polarized; usually the component having the electrical field vector parallel to the antenna axis is stronger than the component which is parallel to the ground plane.

The exact calculation of transversal mode helical antennas is not as simple as for dipole and monopole antennas. Usually they are designed empirically: Start with a wire that is half a wavelength long, wind it up into a helix, and measure the antenna impedance using a vector network analyzer. Then cut it back until nearly real input impedance at the frequency of operation is obtained. Real input impedance means that the antenna is in resonance. Fine-tuning of the frequency of resonance is possible by compressing or stretching the helix.

Even if the antenna is in resonance, it will not be matched to 50 Ω yet. The input impedance will be the sum of the radiation and loss resistances, usually smaller than 50 Ω. For the design of the needed additional matching circuit, we can use the Smith Chart as described in Section 1.3.

Chu’s and Wheeler’s limit on the bandwidth for a given dimension also holds for helix antennas. A small transversal mode helix therefore has tight bandwidth and is sensitive to tolerances of the matching components.

2.4 Small Loop Antennas

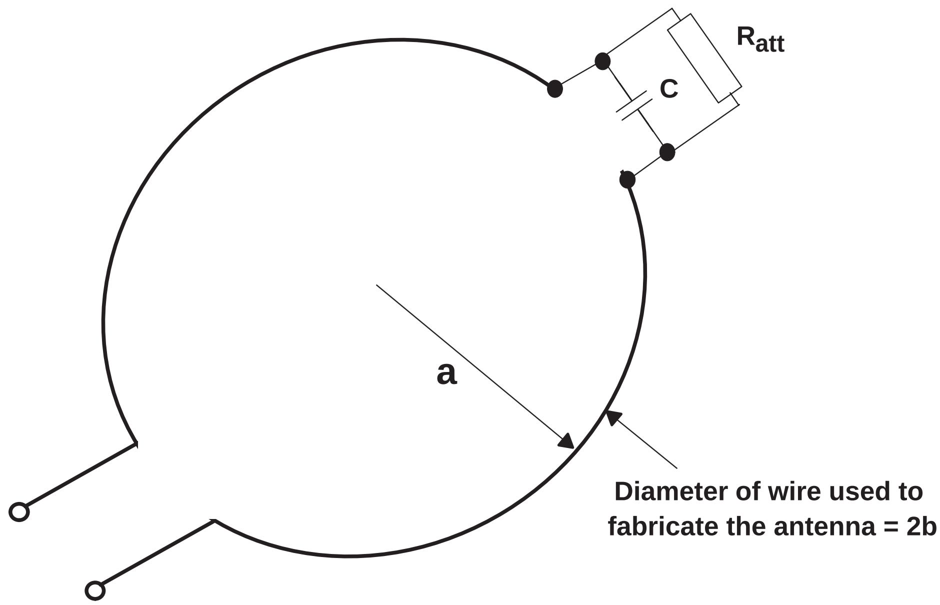


Figure 13. Small Loop Antenna With Differential Feed

The loop antenna shown in Figure 13 has a differential feed. Often a ground plane is made part of the loop, giving a single-ended feed as shown in Figure 14.

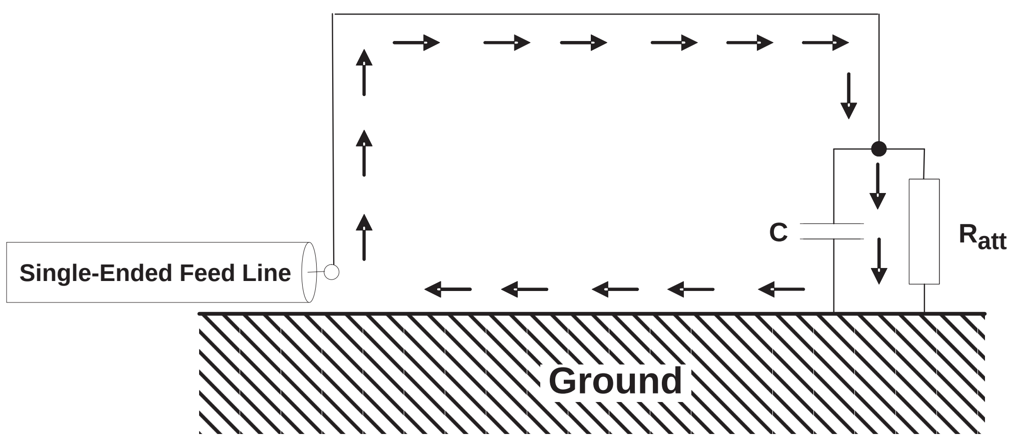


Figure 14. Single-Ended Loop Antenna

The small arrows indicate the current flow through the loop. On the ground plane, the current is mainly concentrated on the surface. The electrical behavior of the structure in Figure 14 is therefore similar to that of the loop with differential feed shown in Figure 13.

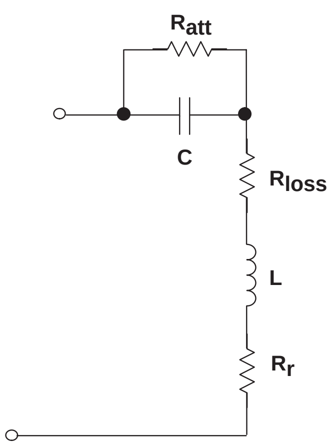
The following considerations on small loop antennas are based on /4/ and assume that the current is constant over the loop. This means that the circumference must be smaller than one tenth of a wavelength. Although this is rarely the case, the given formulas describe the principal behavior and can be used as a starting point for the loop antenna design.

If the current is constant over the loop, we can consider the loop as a radiating inductor with inductance L. Where L is the inductance of the wire or PCB trace. Together with the capacitor C, this inductance L builds a resonant circuit. Often a resistor R_{att} is added to reduce the quality factor of the antenna and to make it less sensitive to tolerances. Of course, this resistor dissipates energy and reduces the antenna's efficiency.

The following calculations hold for circular loops with the radius a for square loops with the side length a. A rectangular antenna with the sides a_1 and a_2 will be approximated by an equivalent square with the side length $a = \sqrt{a_1 a_2}$. The length (circumference) of the wire building the loop will be called U, where U = 2πa for a circular loop or 4a for a square loop.

For the calculation of the inductance, the wire radius b, where b is 1/2 the diameter of the actual wire used to fabricate the antenna, is needed. In the frequent case where a loop antenna is realized by a trace on a PCB, b = 0.35·d + 0.24·w can be used, where d is the thickness of the copper layer and w is the trace width.

Figure 15 shows the equivalent schematic of a small loop antenna.



$$R_{loss} = \frac{U}{2 \times w} \times \sqrt{\frac{\pi \times f \times \mu_0}{\sigma}} + R_{ESR}$$

$$L = \mu_0 \times a \times \left(\ln\left(\frac{a}{b}\right) + 0.079\right) \quad \text{(Circular Loop)}$$

$$L = \frac{2 \times \mu_0}{\pi} \times a \times \left(\ln\left(\frac{a}{b}\right) - 0.774\right) \quad \text{(Square Loop)}$$

$$R_r = \frac{A^2}{\lambda^4} \times 31.17\ \text{k}\Omega$$

Where: w = PCB trace width
b = wire radius
a = radius of the circular loop = side length of a square loop or equivalent side length of a rectangular loop.

Figure 15. Equivalent Schematics of the Small Loop Antenna in Figure 13

The radiation resistance of loop antennas is small, typically smaller than 1 Ω. The loss resistance R_{loss} describes the ohmic losses in the loop conductor and the losses in the capacitor (expressed by its ESR). Usually, the ESR of the capacitor can not be neglected. Interestingly, the thickness of the copper foil is not needed for the calculation of the loss resistance because due to the skin effect, the current is confined on the conductor surface.

Together with the loop inductance L, which is the inductance of the wire, the capacitor C builds a series resonant circuit. In practice, the L-to-C ratio of this resonant circuit is large giving a high quality factor Q. This would make the antenna sensitive to tolerances. That's why often an attenuating resistor R_{att} is added to reduce the Q. To describe the influence of R_{att} on the loop antenna, we make a parallel to series conversion and use the equivalent series resistance R_{att_trans}. The resistance value of R_{att_trans} is determined by the acceptable tolerance of the capacitor and the geometry of the loop.

The maximum usable quality factor is calculated from the capacitance tolerances $\Delta C/C$:

$$Q = \frac{1}{\sqrt{1 + \frac{\Delta C}{C}} - 1}$$

The series transformed attenuation resistance then will be:

$$R_{att\_trans} = \frac{2 \times \pi \times f \times L}{Q} - R_r - R_{loss}$$

This gives the efficiency of the loop antenna:

$$\eta = \frac{R_r}{R_r + R_{loss} + R_{att\_trans}}$$

In most cases, the radiation resistance is much smaller than the loss resistance and the transformed attenuation resistance, giving a poor efficiency. In this case the approximation:

$$\eta = \frac{Q \times R_r}{2 \times \pi \times f \times L}$$

can be used. R_r is determined by the loop area, which is πa^2 for circular loops, a^2 for square loops, and $a_1 a_2$ for rectangular loops. Figure 16 shows the efficiency of small circular loop antennas versus their diameter for an assumed tolerance of 5%. The trace width has been assumed as 1 mm, the copper thickness as 50 μm; but both values have only a minor influence on the efficiency, which is mainly determined by the attenuation resistance R_{att}. As expected, the efficiency increases with increasing diameter.

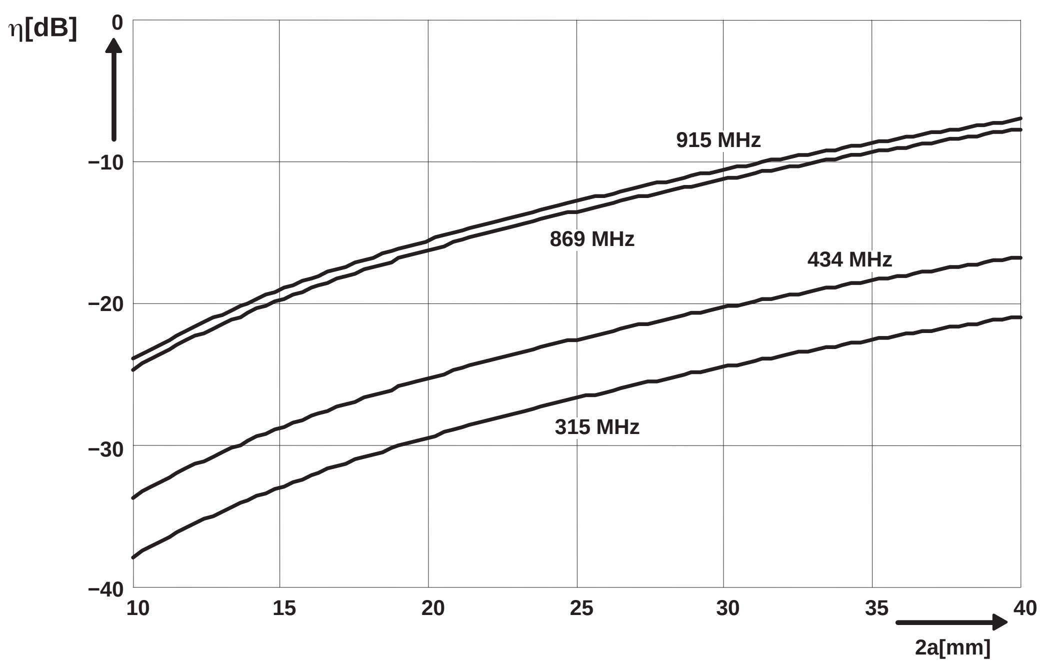


Figure 16. Efficiency of Small Loop Antennas for 5% Tolerance

If we feed the loop antenna as shown in Figure 14, the series equivalent circuit of Figure 15 describes the antenna impedance. Even including the effect of R_{att}, the total series resistance will be small, usually below 10 Ω. If we feed the antenna directly at the capacitor, the parallel equivalent circuit describes the antenna impedance. A small series loss resistor transforms into a large parallel resistor, usually several kΩs.

In both cases matching to 50 Ω will be difficult. That's why the loop antenna is often tapped, giving an impedance in between the too small and the too large values. Figure 17 shows an example.

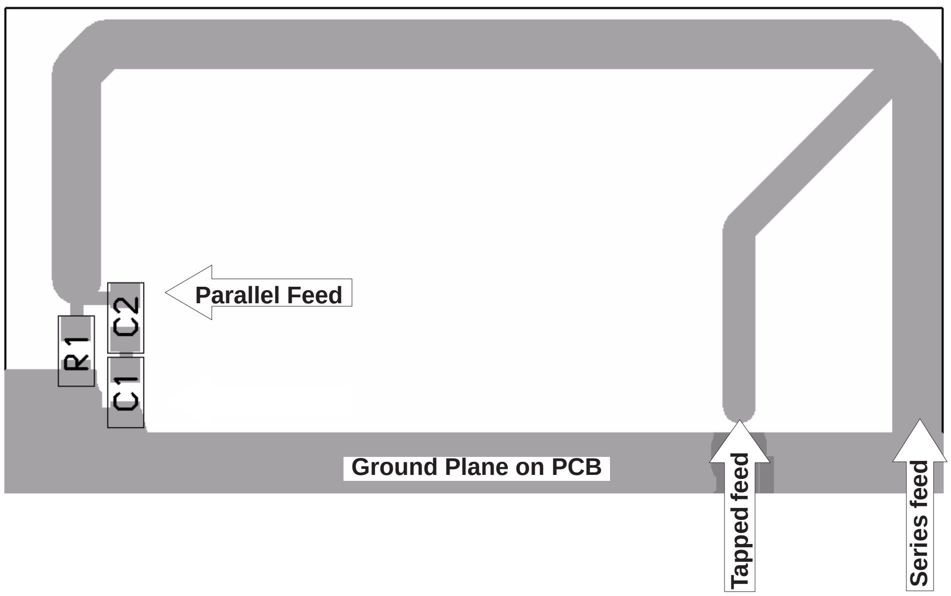


Figure 17. Example of a Tapped PCB Loop Antenna

A series feed (in the lower right corner) would give a small impedance; a parallel feed (directly at the capacitors) would give a much too large impedance. The tap provides an impedance close to 50 Ω in this example. The loop capacitor has been split into two series capacitors C1 and C2. This makes it possible to realize non-IEC capacitance values. R1 is the damping resistor which de-Q's the circuit, thus increasing the bandwidth and subsequently reducing the tolerance requirements.

Unfortunately, there are no easy formulas that describe the tapped structure and give the right location for the tap. The line from the antenna termination to the tap is not a transmission line and will disturb the field in the antenna. Therefore, we have to find out the optimal structure by electromagnetic simulations. Often a trial and error procedure is used as an alternative: Using a vector network analyzer we determine the capacitance value that gives the best return loss and the largest resistance value that gives the required bandwidth.

The loop antenna gives a linear polarization with the vector of the electric field oscillating in the plane built by the loop.

As opposed to all the antennas discussed so far, loop antennas are magnetic antennas. This means that they are not detuned by the dielectric constant of the material in their reactive near field. That's why loop antennas are often used for body-worn or hand-held equipment.

Table 2 shows a feature comparison of the discussed antennas.

Table 2. Features of Short Range Antennas

Antenna Type	Applications and Interest	Efficiency	Sensitivity to Detuning
Dipole	Large wire antenna; balanced feed	High	Moderate
Monopole	Large wire antenna; single-ended feed	High	Moderate
Loaded stub	Small wire and PCB antennas	Moderate	High
Transversal mode helical antenna	Small wire antennas	Moderate	High
Small loop	Body-worn antennas	Low	Low

2.5 Rules of Thumb for the Antenna Design

We can summarize the considerations made so far in the following rules of thumb:

- If the available space is sufficient, use a half-wave dipole (for differential feed) or quarter-wave monopole (for single-ended feed) antenna for best efficiency.
- If possible, keep the space around the antenna clear from conducting or dielectric materials, such as electronic components, the casing or the user's body.
- Sometimes, dielectric materials in the reactive near field are unavoidable. In these cases, measure the antenna impedance under real application conditions and match it to the needed characteristic impedance.
- Due to space limitations, the ground plane of quarter-wave monopoles is often too small. In these cases try to create as much ground plane around the feed point as possible, measure the resulting antenna impedance and match it to the needed characteristic impedance. Good performance can be obtained from a counterpoise made from a quarter wavelength conductor that is connected to ground in the vicinity of the antenna's feeding point. The counterpoise should run as long as possible perpendicular to the monopole radiator.
- When using premanufactured antennas keep in mind that their performance depends on the attached ground plane. The manufacturer's specifications are only achieved if the ground plane has the same size and shape as the manufacturer's evaluation board. In all other cases, you have to measure the impedance of the premanufactured antenna under application conditions and to match it to the needed characteristic impedance.
- Small loop antennas are insensitive to varying dielectric conditions in their reactive near field. They can be a good solution for portable and hand-held devices but have a much lower efficiency than electrical antennas. Only antennas with a circumference smaller than one tenth of a wavelength can be considered as purely magnetic antennas. Larger loops have a higher gain but also a higher sensitivity to the environmental conditions.
- Size matters: Always keep in mind that Chu's and Wheeler's limit determines the product of the bandwidth and the efficiency for a given antenna dimension. An extremely small antenna can not be efficient and tolerance-insensitive at the same time.

2.6 Antenna on the TRF6901 Reference Design

Figure 18 shows the structure of the antenna used in the TRF6901 reference design:

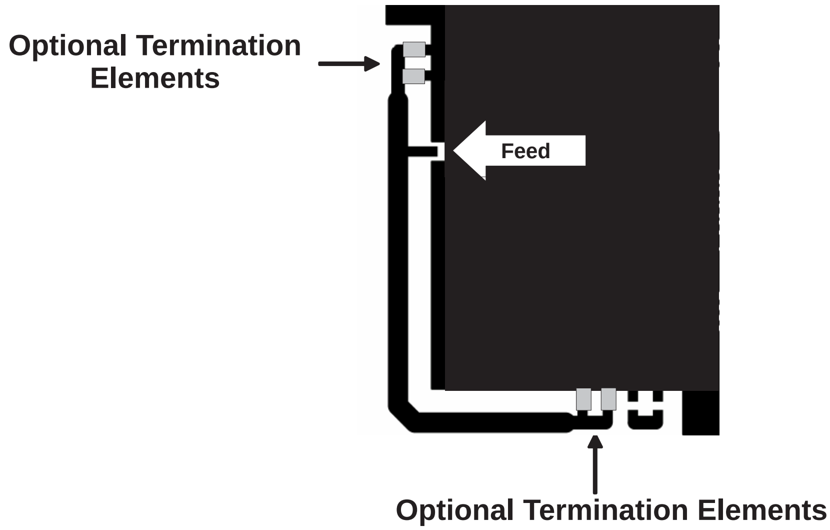


Figure 18. PCB Antenna Layout on the TRF6901 Reference Design

Beginning from the feeding point, two legs with different lengths run in parallel to the PCB ground plane and can be terminated at their ends by SMD components to ground. Depending on the used termination elements the layout can be used as an inverted-F antenna or a tapped small loop antenna.

The short leg is usually terminated by a 0-Ω resistor. If the long arm is left open, we get an inverted-F antenna. Terminating the long arm with a capacitor and a resistor in parallel gives a tapped loop antenna similar to the one shown in Figure 17; 1 kΩ and 0.5 pF bring the loop into resonance at 915 MHz.

When the antenna is used as a loop antenna, we have to consider the circumference of the loop in relation to the wavelength on the PCB. The trace width of the antenna in the reference design is 1,5 mm and the thickness of the PCB material is also 1,5 mm. As an approximation, we calculate the effective dielectric constant according to Section 2.2 and get $\varepsilon_{eff} = 3.1$. The wavelength of a 915-MHz signal on the PCB is then:

$$\frac{\lambda_0}{\sqrt{\varepsilon_{eff}}} = 186 \text{ mm}$$

The circumference of the loop in the reference design is 91 mm, which is almost half a wavelength and much more than one tenth of a wavelength as assumed in Section 2.4. The antenna behavior will therefore be different from that of a pure magnetic antenna; the antenna will excite some electrical field in the reactive near field too. This gives a gain larger than predicted for the loop antenna in Section 2.4.

As we can see from Figure 18, the antenna is bent around the corner of the PCB. This gives radiation in both the x and y direction and helps to make the radiation pattern more omni directional.

3 RF Propagation

3.1 Path Loss

The basis for an estimation of the achievable distance in a communication link is the link budget. The link budget describes the relationship between the received power P_r and the transmitted power P_t as:

$$P_r = P_t \times G_t \times G_r \times \left(\frac{\lambda}{4 \times \pi}\right)^2 \times \left(\frac{1}{d}\right)^n$$

G_t and G_r are the gains of the transmitter and receiver antennas respectively. λ is the wavelength and d the distance between transmitter and receiver. Identical units must be used for λ and d. As we can see, the received power increases with the square of the wavelength (or decreases with the square of the frequency). This comes from the fact that an antenna with the same gain is larger at lower frequencies and therefore catches more power from the radiated field.

The path loss exponent describes the influence of the transmission medium. In free space, the path loss exponent is theoretically two, which describes the equal power distribution from an isotropic radiator on the surface of a sphere. $N < 2$ means that the medium bundles the wave, giving a path loss smaller than in free space. An attenuating medium gives a path loss exponent $n > 2$.

The path loss is the ratio of the powers between the transmitter and receiver antennas in logarithmic units:

$$PL(d) = 20 \times \log\left(\frac{4 \times \pi \times 1\ \text{meter}}{\lambda}\right) + 10 \times n \times \log\left(\frac{d}{1\ \text{meter}}\right)$$

For convenience, we can use the formula:

$$PL(d) = 32.4\ \text{dB} + 20 \times \log\left(\frac{f}{1\ \text{GHz}}\right) + 10 \times n \times \log\left(\frac{d}{1\ \text{meter}}\right)$$

In outdoor applications, we often have a direct line of sight between the transmitter and the receiver. In this case, we can use a path loss coefficient of two, if there are no obstacles in the first order Fresnel zone. The Fresnel zone is an ellipsoid, which has the transmitter and receiver antennas at its foci as shown in Figure 19.

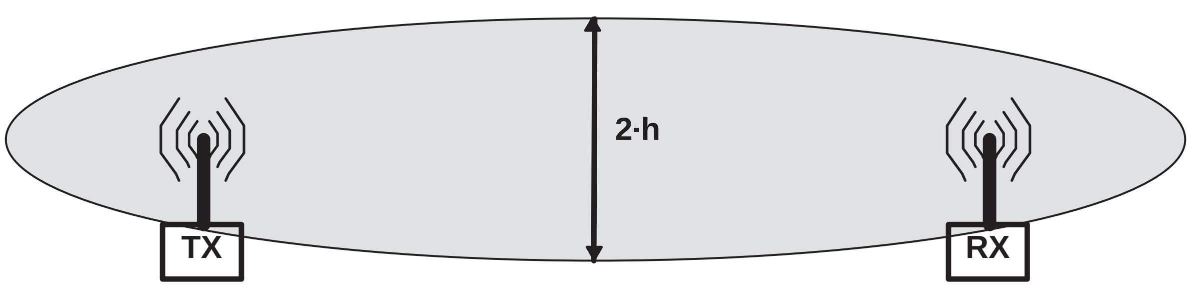


Figure 19. Fresnel Zone Between Transmitter and Receiver

In the middle between the transmitter and the receiver, the first Fresnel zone has the diameter $2h = \sqrt{d \times \lambda}$. Often it is assumed that a path loss coefficient of two still can be used, if at least 60% of this zone is free from obstacles.

Figure 20 shows the free space path loss (n = 2) for four frequently used short range bands:

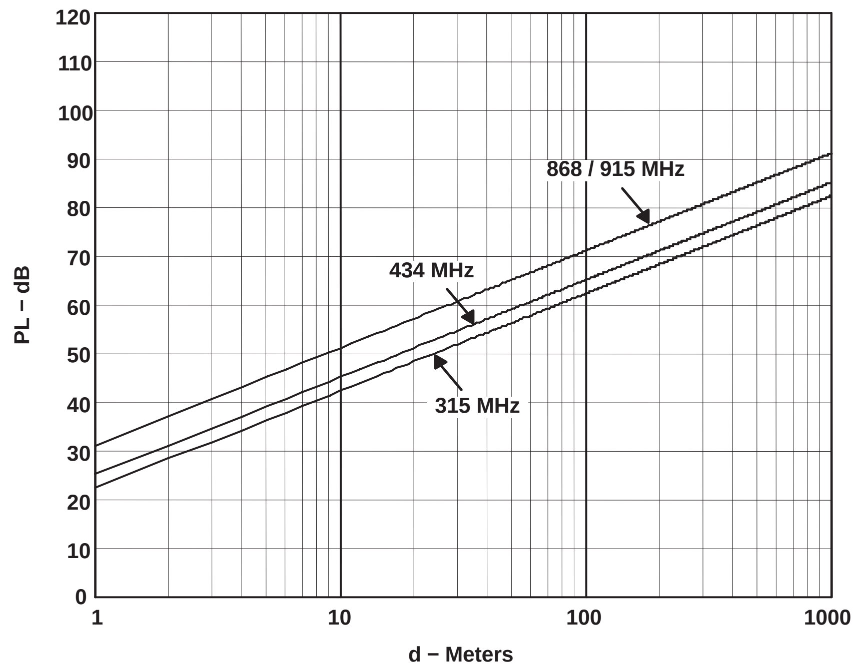


Figure 20. Free Space Path Loss (P_L) For Four Short Range Frequency Bands

If we have no line of sight conditions, there will be additional losses due to absorption, diffraction and refraction. These losses are described empirically by the path loss coefficient n. Table 3 has some measured values of the path loss coefficient together with the associated standard deviations /5/, /6/. It is assumed that the transmitter and the receiver are on the same floor.

Table 3. Measured Path Loss Coefficients and Standard Deviations /5/

Location	n	σ_n
Retail store	2.2	8.7
Grocery store	1.8	5.2
Office, hard partitions	3	7
Office, soft partitions	2.6	14.1
Metalworking factory, line of sight	1.6	5.8
Metalworking factory, obstructed sight	3.3	6.8

If the transmitter and the receiver are not on the same floor, a floor attenuation factor $L(N_f)$ with N_f as the number of penetrated floors, must be added. Table 4 has some typical floor attenuation factors according to /5/.

Table 4. Typical Floor Attenuation Factors /5/

Number of Floors N_f	$L(N_f)$ in dB	σ in dB
Through 1 floor	13	7
Through 2 floors	19	2.8
Through 3 floors	24	1.7
Through 4 floors	27	1.5

We can see that the standard deviations are extremely large; there will be a lot of uncertainty in the path loss prediction. An improvement is possible if we track the path from the transmitter to the receiver. This method is called ray tracing and accounts for all the individual partition losses at walls, doors, windows, etc. The estimated path loss is then:

$$PL = 20 \times \log\left(\frac{4 \times \pi \times d}{\lambda}\right) + \Sigma \text{ partition losses in dB} + L\left(N_f\right)$$

Table 5 has some typical partition losses according to /5/.

Table 5. Typical Values of Partition Losses

Material Type	Loss in dB
Concrete block wall	13 to 20
Moveable wall (cubicle)	1.4
Window	2
Metal foil insulation	3.9
Storage rack	4 to 6

The partition loss values depend on the individual construction of the particular obstacle and also on the frequency.

3.2 Multipath Propagation Effects

In a practical transmission system, the receiver does not only get the signal via the direct path, but also from reflections, diffracted, and scattered rays.

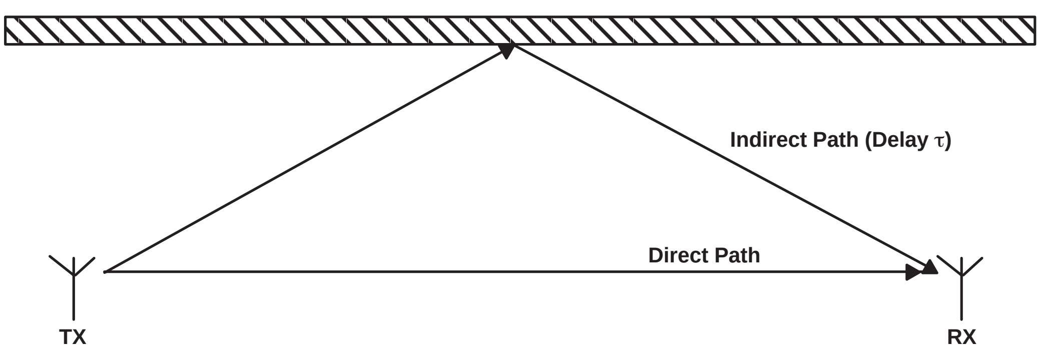


Figure 21. Multipath Propagation

Multipath propagation can cause two kinds of problems: fading and inter symbol interferences (ISI).

Fading occurs if the time difference between the arrival of the direct and the delayed (reflected) wave is in the order of magnitude of the RF period time.

If the time difference is an integer multiple of the period time, both waves interfere constructively; the received signal is stronger than without fading.

If the time difference is an odd multiple of the half-period time, the direct and the indirect component subtract from each other, in the worst case they totally cancel out.

If a receiver is moved into an environment with multipath transmission, there will be an alternating stronger and weaker signal. The damage from the mutual cancellation is much more significant than the advantage from the constructive interference at other locations.

If the time difference, τ, is on the order of magnitude of the bit duration, multipath transmission leads to inter-symbol interference. This principle is shown in Figure 22.

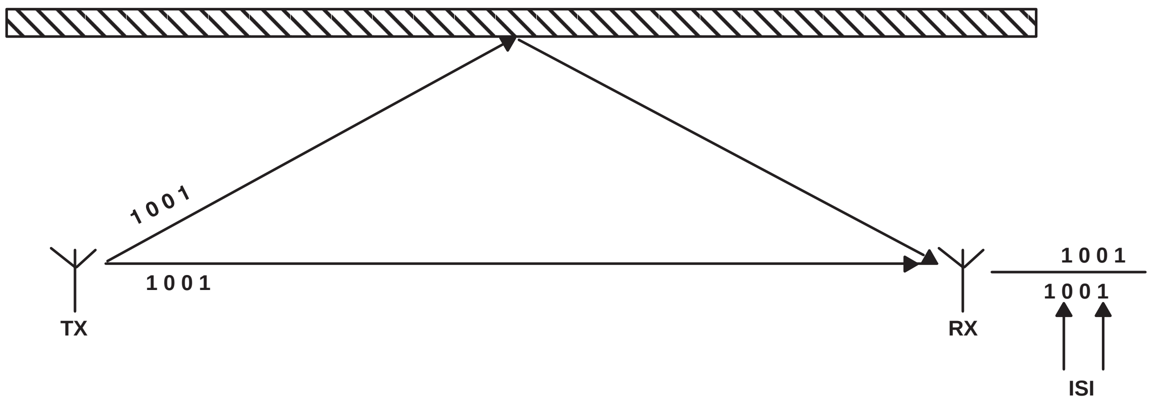


Figure 22. ISI Due to Multipath Propagation

Even with a strong RF signal at the receiver input, the transmission is disturbed by ISI. The influence of the ISI is particularly important for higher bit rates, because then a time difference in the order of magnitude of the bit duration can occur in smaller rooms.

To avoid fading problems, antenna diversity on the transmitter or the receiver side can be used. In the simplest realization, two or more antennas are combined by a passive divider. The probability that both antennas are in a deep fade is much smaller than for one antenna only. This simple structure can give a substantial improvement in the line reliability.

For better performance, an RF switch is used which connects either one or the other antenna to the IC device. During the preamble of the protocol, the RSSI signal can give information on which of both antennas performs better. This antenna is then used during the data package.

In some cases where the link reliability is important, *true diversity* can be implemented. This means that two complete receivers with separated antennas are used. Depending on the signal quality at the receiver outputs, one or the other is used.

In most simple applications, antenna or time diversity are cost-prohibited and seldom used.

If neither the transmitter nor the receiver is mobile, antennas with a high directivity can suppress the unwanted reflections.

4 Examples and Measurements

4.1 Test Module Schematics

In many short range applications, the overall size of transmitter or receiver modules is small compared to the wavelength. The ground plane is therefore usually part of the antenna and radiates or receives energy also. To test the behavior of various antennas under practical conditions, test modules based on the TRF4903 have been built that deliver an output power of +8 dBm. The MSP430 microcontroller is used to program the TRF4903. Figure 23 shows the test module schematics.

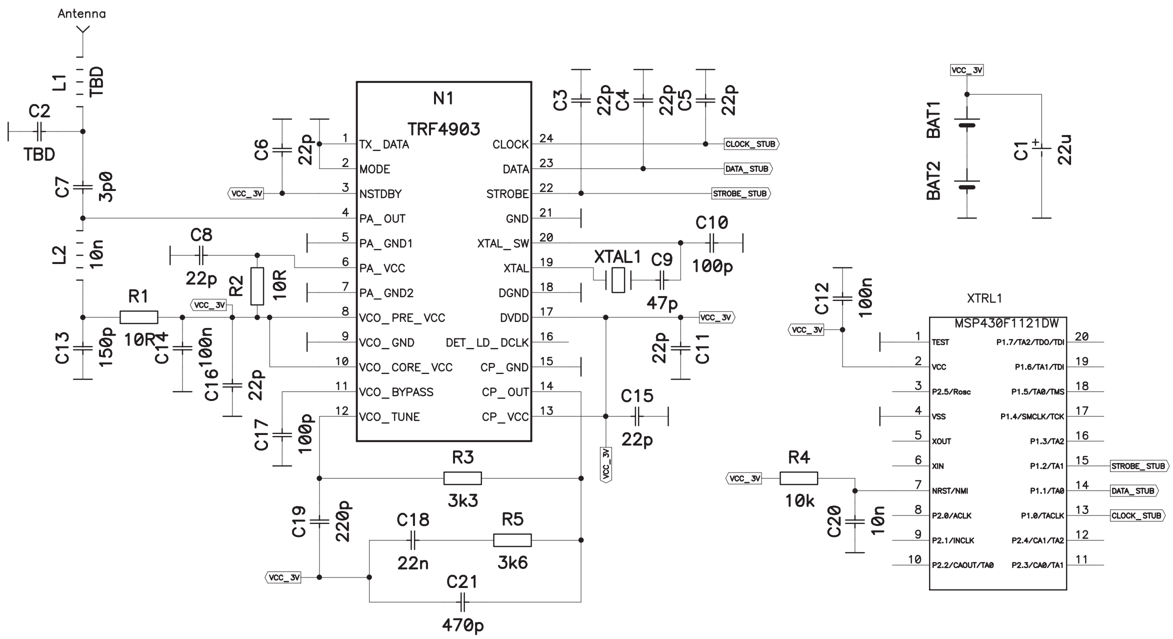


Figure 23. Test Module Schematics

The TRF4903 was programmed to send an unmodulated CW signal with +8 dBm at 915 MHz. The matching elements L1 and C2 have been optimized by network analyzer measurements to tune the particular antennas used.

4.2 PCB Monopole Antenna Module

Figure 24 shows the layout of the PCB monopole antenna. In order to save PCB space the monopole has been bent by 90 degree. In the PCB layout, the monopole was made longer than calculated according to Table 1. This should give some room for possible manufacturing tolerances.

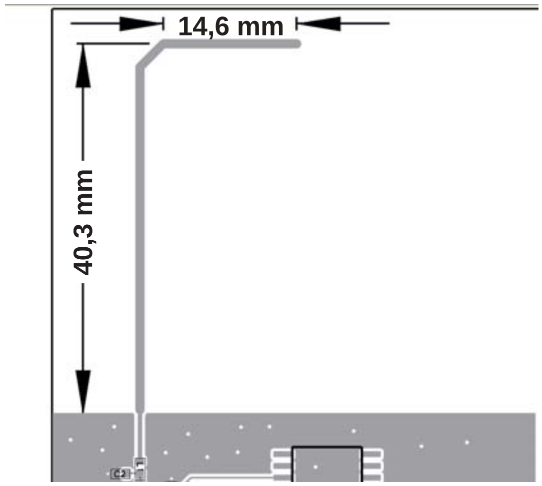


Figure 24. Layout of the PCB Monopole Antenna

Using a vector network analyzer, we measured the antenna impedance on the upper pad of L1 and cut back the monopole until real antenna impedance was achieved. The antenna impedance in resonance is 35.5 Ω, which is within the theoretical value range of 30 Ω to 36.5 Ω. The mismatch loss to 50 Ω is as low as 0.13 dB in this case. No further matching components have been used; inductor L1 was replaced by a 0-Ω resistor. C2 was left unpopulated.

The radiation characteristic of the antenna module was measured in an anechoic chamber with the test module upright (see Figure 25) and flat (see Figure 26) on the turntable.

The outer boundary of the radiation patterns given in this report correspond to an effective radiated power (dipole related) of ERP = + 10 dBm; the scale is 20 dB/division.

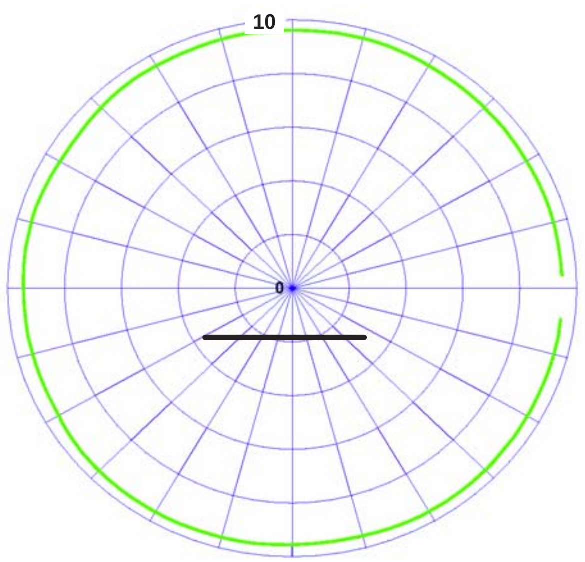


Figure 25. Vertical Radiation Pattern of the Stub Module (Upright)

The radiation pattern is almost angle-independent, the maximum ERP is + 6.5 dBm, corresponding to EIRP = 6.5 dBm + 2.15 dB = +8.65 dBm. The TRF4903 delivers +8 dBm of output power. The maximum antenna gain is therefore +0.65 dB.

As expected, the horizontal radiation pattern has a more pronounced radiation characteristic:

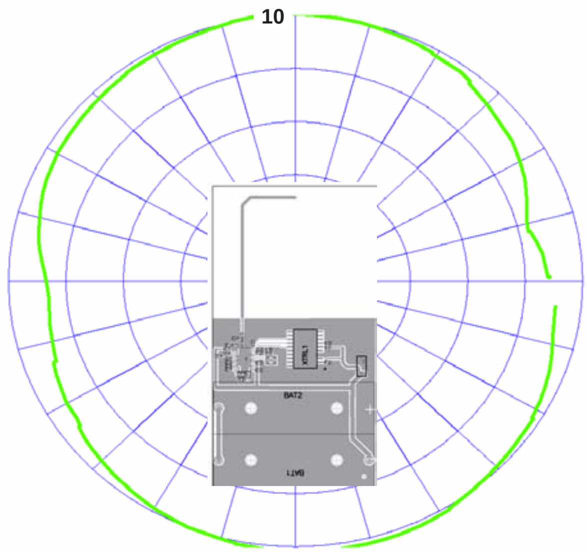


Figure 26. Horizontal Radiation Pattern of the Stub Module (Flat)

The maximum ERP is +10.85 dBm, corresponding to EIRP = +13 dBm. With +8-dBm transmit power, this gives an antenna gain of 5 dB.

Electrical antennas are sensitive to detuning by dielectric material in their reactive near field. Figure 27 shows the horizontal radiation pattern of the same stub module attached to the arm of a test person.

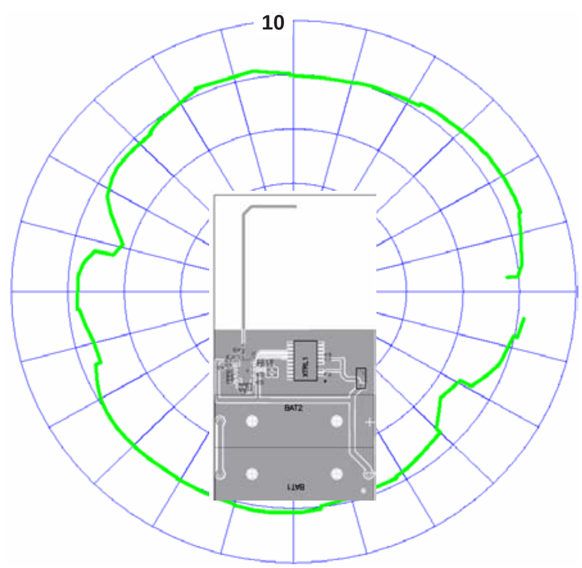


Figure 27. Horizontal Radiation Pattern of the Stub Module Close to the Human Body

The maximum ERP is −4.4 dBm, compared to +10.85 dBm measured on the free stub module. The loss due to detuning and absorption is as large as 15.25 dB in the direction of maximum radiation.

4.3 Transversal Mode Helical Antenna Module

For the helical antenna module, a commercially available transversal mode helix was chosen: The KUNDO 593266 is a SMT component that comes on tape and reel and can be used for 868 MHz as well as in the 915-MHz band. The layout is shown in Figure 28.

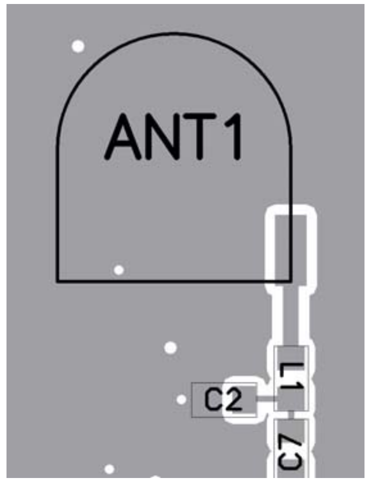


Figure 28. Layout of the SMT Helical Antenna

As opposed to the monopole, underneath and around the antenna there is a ground plane. The white dots in Figure 28 are vias connecting the top side ground with the internal ground layer of the PCB.

The antenna impedance was measured on the upper pad of C7. Without matching components and with L1 replaced by a 0-Ω resistor, the measured reflection coefficient was $\overline{\Gamma} = 0.91 \cdot e^{-143°}$. L1 = 5.6 nH and C2 = 16 pF give an acceptable reflection coefficient of $\overline{\Gamma} = 0.35 \cdot e^{147°}$.

Figure 29 has the radiation pattern. The test module was placed flat on the turntable.

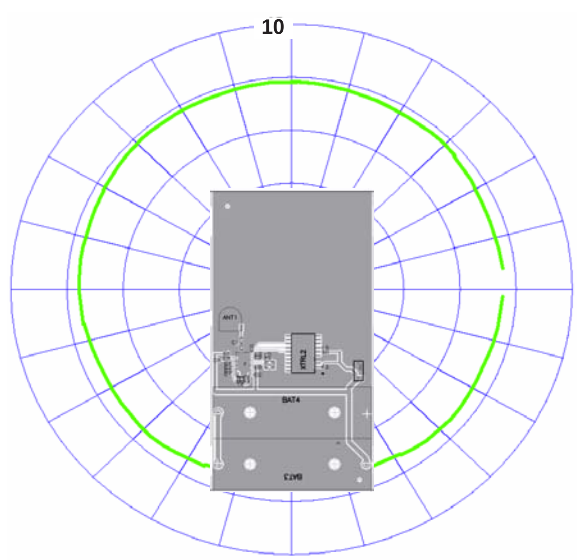


Figure 29. Horizontal Radiation Pattern of the Helix Module

The maximum ERP is −10.6 dBm, corresponding to EIRP = −8.45 dBm. With +8-dBm transmit power follows a gain as low as −16.45 dB. The small dimensions of the antenna lead to a small radiation resistance and a strong effect of losses in the PCB and the matching components. Also, the size of the PCB ground plane is not large compared to the wavelength. That's why the shape and the size of the ground plane have an effect on the radiation pattern.

Figure 30 shows the radiation pattern under the same conditions but with the helix module attached to the forearm of a test person. The PCB ground plane was between the helix and the arm; the forearm was held in a horizontal position.

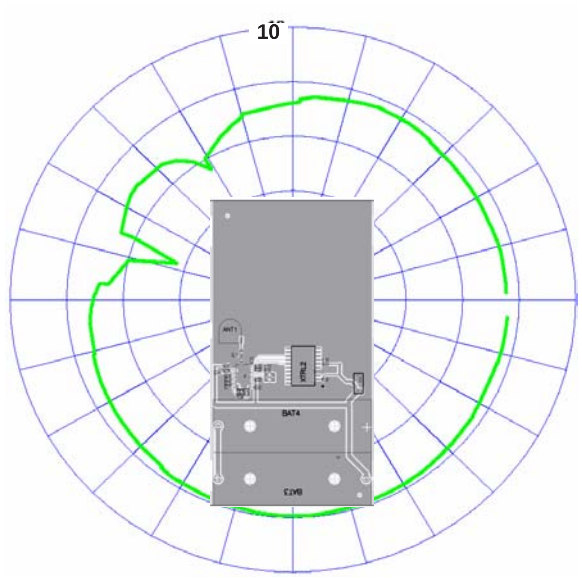


Figure 30. Horizontal Radiation Pattern of the Helix Module Close to the Human Body

Despite the lower efficiency in the upper right corner (which comes from the absorption by the chest of the test person), the radiation pattern is identical to that of the free helix module. The PCB ground plane acts as a shield and makes the radiation pattern independent on the tissue underneath it.

4.4 Loop Antenna Module

Figure 31 has the layout of the loop antenna module. The loop capacitor is a series connection of the two capacitors C40 and C44; this allows realizing small capacitance values in fine steps. R11 is the attenuation resistor; L5 and C46 should help to improve the matching to 50 Ω experimentally, through an iterative approach.

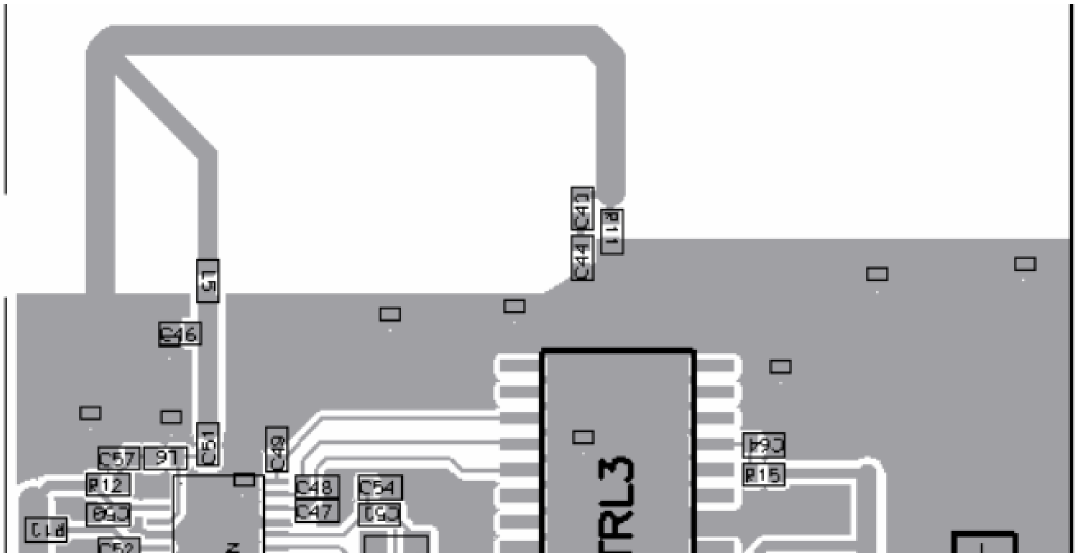


Figure 31. PCB Loop Antenna

The loop width is 25 mm, the height 11,5 mm, the trace width 1,5 mm with a copper thickness of 50 µm.

According to the formulas in Figure 15, the inductance is L = 40.9 nH and the radiation resistance R_r = 0.22 Ω. The calculated capacitance needed for resonance at 915 MHz is 0.74 pF.

Usual tolerance values for low cost capacitors are around 5%. For the small capacitance needed here, the effective tolerance would be larger because the parasitic capacitance of the damping resistor, the PCB pads, and even the solder material contributes to the total capacitance uncertainty. We assumed a total tolerance of 20%, which gives a maximum quality factor:

$$Q = \frac{1}{\sqrt{1.2} - 1} = 10.5$$

The impedance of the loop inductance is $Z_L = 2\pi \cdot x$ 915 MHz x 40.9 nH = 235 Ω. The damping resistance must therefore be not larger than R_{att} = 10.5 x 235 Ω = 2.47 kΩ. In the test module, a 2.2-kΩ resistor was chosen, giving Q = 2.2 kΩ/235 Ω = 9.36. The transformed series attenuation resistance is then $R_{att\_trans} = Z_L/Q$ = 235 Ω / 9.36 = 25.1 Ω. Compared to that, we can neglect the loss resistance of the copper trace on the PCB.

The theoretical antenna efficiency is then:

$$\eta = \frac{0.22\ \Omega}{0.22\ \Omega + 25.1\ \Omega} = 8.7 \times 10^{-3} = -20.6\ \text{dB}$$

As described in Section 2.4, the loop antenna has been tapped to increase the antenna impedance. The position of the tap was determined empirically by electromagnetic simulations.

The antenna impedance was measured on the assembled PCB on the upper pad of C51; L5 was replaced by a 0-Ω resistor at first. We varied the loop capacitors C40 and C44 to bring the loop into resonance. A series connection of 0.8 pF and 0.5 pF gives the reflection coefficient $\overline{\Gamma} = 0.22 \cdot e^{j35°}$, corresponding to a mismatch loss of 0.2 dB. No further matching was required, so C46 was left unpopulated.

The capacitance of the series connection of the 0.5-pF and 0.8-pF capacitors is 0.3 pF and thus smaller than the calculated value of 0.74 pF. One explanation is that the parasitic capacitance of the resistor also contributes to the loop capacitance. Also, the parasitic inductance of the capacitors themselves makes their capacitance look larger at high frequencies than their nominal value.

Figure 32 has the horizontal radiation pattern of the loop antenna module arranged flat on a turntable.

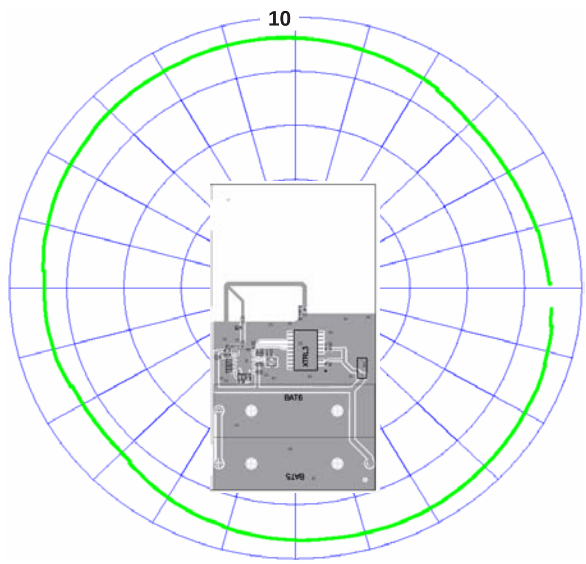


Figure 32. Radiation Pattern of the Loop Antenna Module

The maximum ERP is +4.15 dBm, corresponding to EIRP = +6.3 dBm. This gives an antenna gain of −1.7 dB, much more than calculated. The reason is that the circumference of the loop is not negligible with respect to the wavelength. All calculations in Section 2.4 were made under the assumption that the circumference is smaller than one tenth of the wavelength. The geometrical circumference of the loop antenna in the test module is 73 mm, the wavelength in free space for 915 MHz is 327 mm. Assuming the effective dielectric constant of a trace on FR4 material of 2.97 as in Section 2.2, the wavelength on the PCB is $327\text{mm}/\sqrt{2.97} = 190\text{mm}$. The circumference is therefore 73 mm / 190 mm = 0.38 times the wavelength. The loop does not act like a purely magnetic antenna any more; it will also excite the electrical field in its reactive near field region. As a result, the behavior is in between that of a small loop and an electrical antenna.

For a loop antenna, a smaller influence of dielectric material in the reactive near field on the tuning and the radiation characteristic is expected. Figure 33 shows the radiation pattern of the loop module attached to the forearm of a test person.

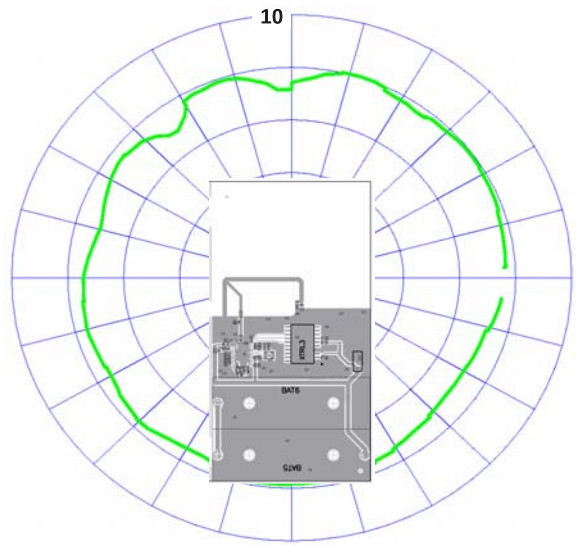


Figure 33. Radiation Pattern of the Loop Antenna Module Close to the Human Body

The ERP is −9.75 dBm, corresponding to EIRP = −7.6 dBm. Given the transmitter power of +8 dBm, the antenna gain is −15.6 dB, almost 14 dB worse than in free space. Note that the gain is still larger than the theoretical value of – 20.6 dB. This again comes from the large dimensions which make the antenna more similar to an electrical radiator.

In order to achieve greater independence from the influences of the surrounding material, the size of the loop antenna must be made smaller. This reduces the radiation resistance and L-to-C ratio and decreases the efficiency as discussed in Section 2.4.

5 References

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Appendix A Antenna Manufacturers

Various manufacturers offer ready-made antennas with 50-Ω input impedance. Most of them have a standard RF connector (usually SMA) and are either a quarter wavelength monopole or a stretched form of a transversal mode helical antenna covered by a protecting rubber. These antennas are sometimes referred to as rubber ducks. The user has to make sure that sufficient ground plane perpendicular to the antenna is available. If the ground plane dimensions are smaller than a quarter-wave length, the antenna impedance will not be 50 Ω any longer. This limits the usage of these antennas to equipment that is larger in size. Vendors of rubber ducks are (in alphabetical order):

- Maxrad www.maxrad.com
- ACT Components www.actcomponents.com
- Antenna factor www.antennafactor.com
- Antenex www.antenex.com
- Centurion www.centurion.com
- Comtelco www.comtelco.net
- CushCraft www.cushcraft.com
- EAD www.ead−ltd.com
- Galtronics www.galtronics.com
- HyperLink Technologies www.hyperlinktech.com
- Jema www.jema.com
- lm Electronics www.lm−electronics.com
- Mobile Mark www.mobilemark.com
- OKW Electronics www.okwelectronics.com
- R W Badland Ltd. www.badland.co.uk
- Sensor Systems www.sensorantennas.com
- ShenZhen Gerbole Electric Technology www.szgerbole.com
- SuperPass www.superpass.com
- Tyco www.rangestar.com

Other vendors offer PCB antennas which can be handled as SMD components. Often they are derivates of PCB monopoles or small loop antennas. These antennas have their specified performance only if the ground plane on the application PCB is identical or at least similar to that of the manufacturer’s evaluation board. In all other cases, additional matching components will be required.

Vendors of PCB antennas are:

- Antenna Factor www.antennafactor.com
- Corry Micronics www.cormic.com/products/Wireless_PatchAnt.cfm
- Mitsubishi www.mmea.com
- Tricome www.tricome.com

In some cases, when both the transmitter and the receiver are at fixed locations, a high directivity and a large gain in the main direction are desired. This helps to increase the usable link distance and to reduce the impact of multipath propagation effects. These antennas are usually equipped with a standard RF connector and have a ground plane independent impedance of 50 Ω. Examples of vendors are:

- HyperLink Technologies www.hyperlinktech.com
- Mars Antennas www.mars−antennas.com
- SuperPass www.superpass.com
- Winncom www.winncom.com

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