Magnetic Loop Antenna: Calculation, simulation, equivalent circuit representation, measurement, and improved understanding of operation

Klaus Solbach, DK3BA, klaus.solbach@uni-due.de

RF-Technology, Department of Digital Signal Processing DSV, Faculty of Engineering, University Duisburg-Essen

I. Introduction

After moving to a flat on the second floor, there was initially no antenna option for me as a radio amateur. A long wire for the shortwave bands across the street in front of the house was successfully tried out but had to be dismantled immediately because the use of public space is not permitted. Since the apartment has a second, small balcony, a Magnetic Loop Antenna (MLA) was an alternative. After some preliminary tests with different sizes and shapes, the proven design according to Ch. Käferlein, DK5CZ was selected: A ring (loop) with a diameter of 1.7 m was bent from 22 mm thick copper tube and the open ends were attached to a motor-driven butterfly Variable capacitor (modified kit from TA1LSX). In Fig. 1 you can see the antenna with mast attachments in front of the balcony railing; the variable capacitor is protected from the weather in a piece of HT pipe and at the lower end you can see a coupling loop for connecting to the coax line to the transceiver in the shack.

Figure 1 MLA at the balcony

In the picture you can see a large conductor loop, which forms a strong magnetic field in its immediate vicinity when fed from a high-frequency transmitter - hence the name magnetic loop - antenna; in contrast, in the reactive near field of wire dipoles, the electric field dominates. Design formulas for the MLA have existed since the 1970s. In the CQ DL, the first publications of DL2FA /1,2/ with their own formulas appeared in 1983, while at the same time DK5CZ developed and produced the "Tuneable Magnetic Antenna" (AMA). The most widespread are the formulas from W5QJR, which were adopted for the first time in the ARRL Antenna Handbook 15th Edition /3/; these or similar formulas were later used frequently and repeated in many articles, see the bibliographies with German-language articles in /4/ and international articles in /5,6/. When dimensioning my antennas, however, I largely worked with the Electro-Magnetic (EM) simulation based on EZNEC+ and, where possible, compared the corresponding results with values from the formulas. With the help of the simulation, the understanding of the function and the characteristics of the MLA could be improved. In this way, a complete equivalent circuit diagram of the MLA with a coupling conductor loop could be created and, in particular, conspicuous deviations from the theoretical bandwidths of the realized antenna could be modelled and explained.

In this article, the basic concept of the MLA is first briefly explained and the essential design parameters for an MLA with a diameter of 1.7 m are calculated. Results from calculations using formulas and from the EM simulation for an antenna in free space are compared. A coupling conductor loop is inserted in the simulation and an equivalent circuit diagram is derived for the antenna, which also includes the coupling conductor loop. After that, the theoretical predictions, in particular of the complex antenna impedance or the reflection factor are compared to the measurements on the realized MLA. Important insights into the different concepts of bandwidth and the power losses of the MLA in a real environment are gained here.

II. The theoretical concept of the MLA

The basic concept of the magnetic loop antenna in free space can be described with Fig. 2: A conductor loop with the inductance *L* is brought into resonance at the frequency f_{res} by a capacitor with the capacitance *C*. The current in this resonant circuit is limited by a series resistance *R*. At first glance, this resistance is made up of the loss resistance R_L of the conductor loop and capacitor and a radiation resistance R_R, which represents the radiated power of the conductor loop.

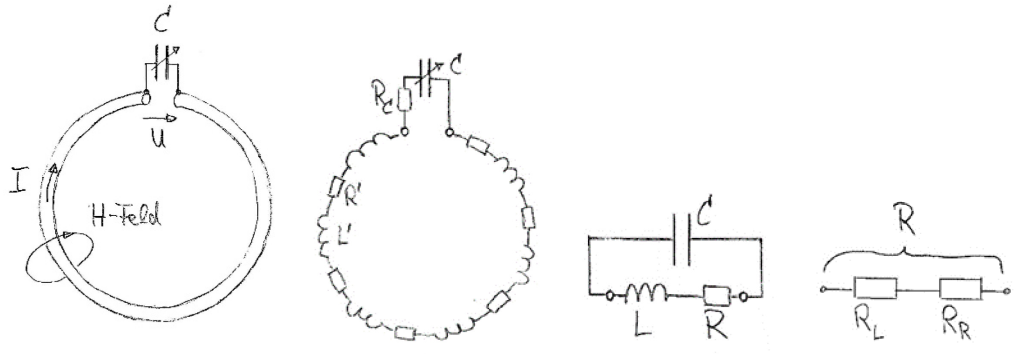

Figure 2 The MLA conductor loop in free space with capacitor tuning and corresponding equivalent circuit diagrams

A closer look reveals that the conductor loop of the MLA is brought to resonance with capacitor tuning such that a strong current I flows through the conductor and creates a magnetic field *H* around the conductor. The stored magnetic energy can be represented by distributed inductances with the dimension H/m. Due to the skin effect, the AC resistance of the conductor is significantly higher than the DC resistance of the conductor. This resistance along the conductor loop is represented by distributed resistances (dimension Ω/m) in series with the distributed inductances. In the simplified equivalent circuit diagram, both are represented by a lumped inductance *L* and a lumped loss resistance R_L; another part of the loss resistance comes from the connected capacitor, which in practice is not completely lossless and whose losses can be described by a series resistance, known as equivalent series resistance (ESR). However, there is also a distributed parallel capacitance between the left and right part of the conductor loop - this is a few pF and is neglected below. As a result, we see the MLA as a simple RLC resonant circuit, which is determined by an "ohmic" resistor in addition to its resonance frequency and reactance. In contrast to resonance circuits in filter or oscillator circuits, only part of the resistance represents the conversion of currents in the conductor loop ("coil") and the capacitor into heat, since a second part represents the conversion into radiated power (in the far field). These two parts of the resulting resistance also appear with every other wire antenna. Their ratio determines the efficiency of the antenna, since an antenna radiates 100% of the transmit power only if it is without losses. For an MLA much smaller than a quarter wavelength in circumference, the unwanted loss resistance is typically much larger than the desired radiation resistance, making the efficiency much less than 100%.

Proven calculation formulas are known from /3/ for all elements of the equivalent circuit diagram and are reproduced here after conversion to metric dimensions:

(a) The following applies to the inductance of a conductor loop with a circumference *U* and a conductor diameter *d* in cm

$$L[\mu H] = 6{,}2 \cdot 10^{-4} \cdot U \cdot \left[7{,}353 \log\left(\frac{2{,}55 \cdot U}{d}\right) - 6{,}386\right]$$

(b) The following applies to the loss resistance per cm of a copper conductor with a diameter *d* in cm at the frequency *f* in Hz

$$R_L[\Omega/\mathrm{cm}] = 8{,}3 \cdot 10^{-8} \cdot \frac{\sqrt{f}}{d}$$

(c) The following applies to the radiation resistance R_R of a conductor loop with the area *A* in m^2 at the wavelength λ in m

$$R_R[\Omega] = 3{,}12 \cdot 10^4 \cdot \left(\frac{A}{\lambda^2}\right)^2$$

(d) The efficiency of the antenna is given by the loss resistance and the radiation resistance as

$$\eta = \frac{R_R}{R_R + R_L}$$

(e) The "unloaded" quality factor Q of an RLC series resonant circuit generally results from the ratio of the reactance X_L of the inductance (or X_C of the capacitance) at the resonant frequency f_{res} and the series resistance *R* and it determines the so-called -3dB bandwidth Δf of the resonator (the spacing of the frequencies at which the reactance magnitude is equal to the series resistance)

$$Q = \frac{X_L}{R} = \frac{2\pi f_{res} \cdot L}{R} = \frac{f_{res}}{\Delta f}$$

(f) At the resonant frequency, the capacitive reactance X_C is equal to the inductive reactance X_L

$$|X_C| = \frac{1}{2\pi f_{res} \cdot C} = 2\pi f_{res} \cdot L = X_L$$

This results in the resonant frequency depending on the square root of the product of inductance and capacitance

$$2\pi f_{res} = \frac{1}{\sqrt{L \cdot C}}$$

In contrast to the 1970s, with the derivatives of the Numeric Electromagnetic Code (NEC), simulators have long been available to radio amateurs that can reliably calculate currents on electrical conductors and are particularly suitable for the rapid calculation of "wire antennas". My MLA simulations were done with EZNEC+ v.6.0.3 thanks to the development work of W7EL, who now makes the latest upgraded version available for free download /7/.

The MLA shown in Figure 1 was initially modelled as a closed conductor loop without a capacitor (and without a coupling loop), see Figure 3. A current source as an excitation sits in the lower wire #1; EZNEC calculates the impedance at its terminals as a resistance *R* (very small, as expected) in series with an inductive reactance X_L as $Z = (0.122 + j220)\ \Omega$. This reactance should be compensated by a capacitor in the upper conductor (wire #9), so that the impedance at the terminals only contains the resistance *R*. You can calculate the capacitance required for this using the resonance condition from the inductive reactance or determine it by trial. Figure 3 shows the conductor loop with source and capacitor as well as the Smith Chart as the calculation result for a selected capacitance of 103 pF. One sees the behaviour of a series resonance at 7.05 MHz, where the reflection factor is close to the short circuit point and the impedance is (nearly) purely resistive because the imaginary part vanishes at this frequency. This resistance contains both the loss component R_L due to the calculated AC resistance of the conductor loop made of copper tubing and the radiation resistance R_R due to the calculated radiated power. To obtain the pure radiation resistance, set the wire loss to “zero” instead of “copper” and repeat the simulation run.

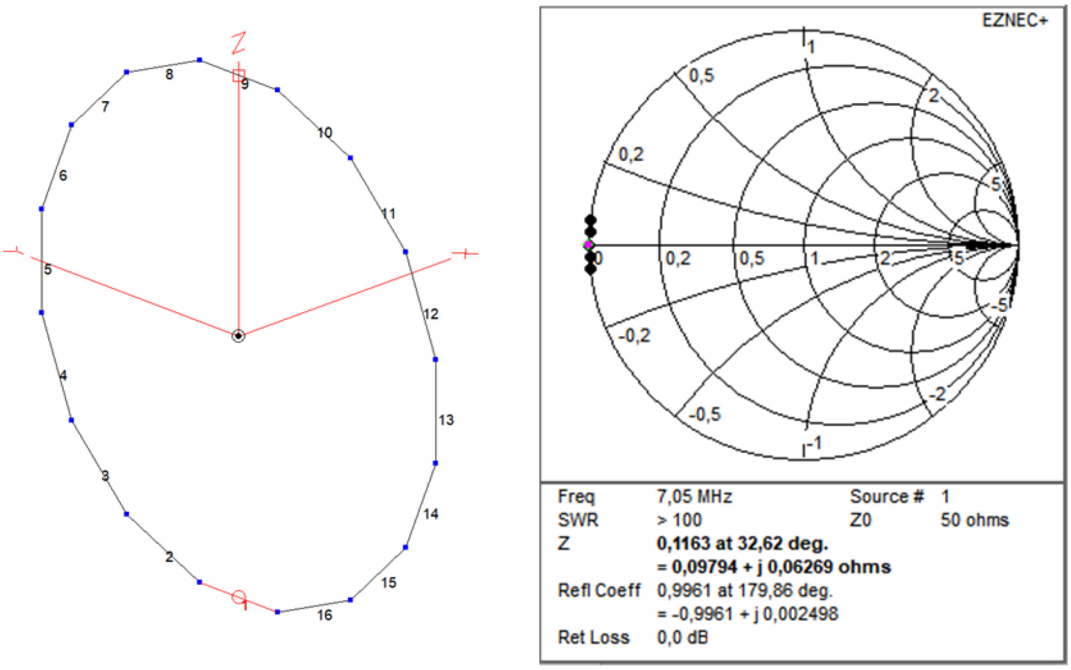

Figure 3 The conductors of the MLA with the source in wire 1 below and the capacitor in wire 9 above and the reflection factor in the Smith Chart as a simulation result. Note: Model with only 1 seg/wire.

With these results of the simulator one can determine the other variables and make a comparison with the results from the formulas, see Table 1. It can be seen that the simulation provides satisfactory agreement with the formula results, even without considering a parallel - capacitance of the conductor loop and although the accuracy can still be increased with more segments per wire; therefore, you can trust the simulation and continue working with it.

	L	C	R_L	R_R	η	Q	Δf
Formulas	4.65 µH	109 pF	0.0531 Ω	0.0493 Ω	-3.16 dB	2011	3.5 kHz
EZNEC	4.96 µH	103 pF	0.0494 Ω	0.0485 Ω	-3.05 dB	2238	3.14 kHz

Table 1 Calculated data of the MLA with U = 531 cm and d = 2.2 cm for 7.05 MHz.

As you can see, the excitation of the MLA by cutting the conductor loop results in a series resonant circuit with a very low resistance value, which is far from match to the usual transmission line characteristic impedance. Instead of direct feeding, most antenna implementations use inductive coupling of the large loop through a smaller conductor loop, as seen in the antenna in Figure 1, where the shielded inner conductor of a coaxial cable forms the coupling loop.

A corresponding model in EZNEC is shown in Figure 4: The conductor loop lies just above the lower segment of the large loop (wire 1) and is modelled as a copper conductor with a diameter of 2.6 mm. The current source is introduced into this conductor loop so that the impedance and the reflection factor at these terminals can be calculated. A diameter of the coupling loop can be found by trial and error, which leads to a match to a 50 Ω feed line. Figure 4 shows the corresponding conductor configuration and a Smith Chart as the simulation result.

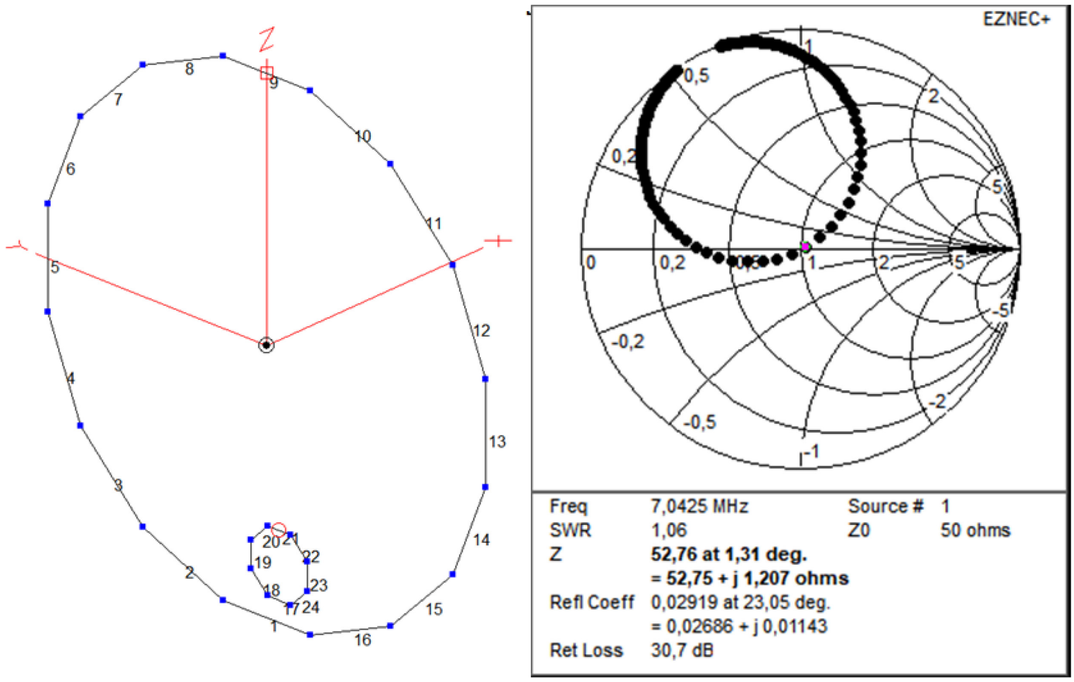

Figure 4 Conductor configuration of the MLA with feed in the lower conductor segment of a coupling loop with a diameter of 23.3 cm and a Smith Chart showing the calculated antenna impedance over the frequency range from 7.01 to 7.07 MHz.

The reflection factor now runs in a circle which touches the matching point at 7.0425 MHz and which is shifted by an inductive reactance, see the shift to above 0.5 on the perimeter of the SC. The resonator bandwidth Δf of about 3.5 kHz is measured between the frequencies where the VSWR is 2.62 - these frequencies correspond to the -3 dB bandwidth of the RLC tank circuit. In order to be able to represent the locus of the antenna impedance by an extended equivalent circuit diagram of the MLA, the coupling of the conductor loops must first be modelled:

In a first approximation, the coupling of the conductor loops can be understood as a lossless transformer, as shown in Figure 5: A current through the large conductor loop generates a magnetic field H and thus a flux of magnetic induction B via the loop area A. A small part of this magnetic flux also penetrates the much smaller coupling loop and, corresponding to area A_2, induces a correspondingly smaller voltage at its terminals 2 - 2' than at terminals 1 - 1'. Although we are only dealing with two conductor loops, each with one turn, an inductive transformer has to be assumed with a turns ratio of 1:N with $N \gg 1$.

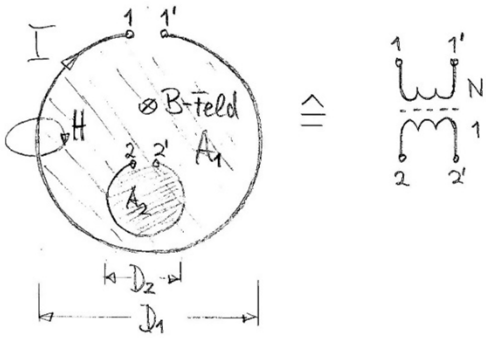

Figure 5 Equivalent circuit representation of the coupling of the large loop and the coupling loop

In the extended equivalent circuit diagram of the MLA, Figure 6, the inductive coupling of the conductor loops is represented accordingly by a lossless transformer, which connects the RLC circuit on the side with the large number of turns, in such a way that the resonant circuit now appears as a parallel RLC resonator. Accordingly, a parallel resonant circuit also appears at the input terminals of the coupling loop where the inductance L_C of the coupling loop is found in series with this impedance. The coupling loop inductance can either be calculated from the reactance read off the shift on the Smith Chart or determined using the loop inductance formula to be approximately 0.68 µH.

The transformer turns ratio 1:*N* can be calculated approximately from the area ratio of the conductor loops as

$$N \approx \frac{A_1(large\ loop)}{A_2(coupling\ loop)} = \frac{{D_1}^2}{{D_2}^2} = \frac{1.7^2}{0.233^2} = 53$$

However, *N* is underestimated because of the inhomogeneous field strength distributions in the conductor loops, but the correct value can be found by trial and error with the help of a circuit simulator; in the present case we set *N* = 85 for perfect matching. Figure 5 also shows the impedance calculated with a circuit simulator (I use the Advanced Design System, ADS). In the Smith Chart, the reflection factor rotates as in the antenna simulation with EZNEC in Figure 4. The markers m1 and m2 entered in the Smith Chart at VSWR ≈ 2.6 have a frequency spacing of about 3.2 kHz, somewhat lower than the resonator bandwidth in the antenna simulation.

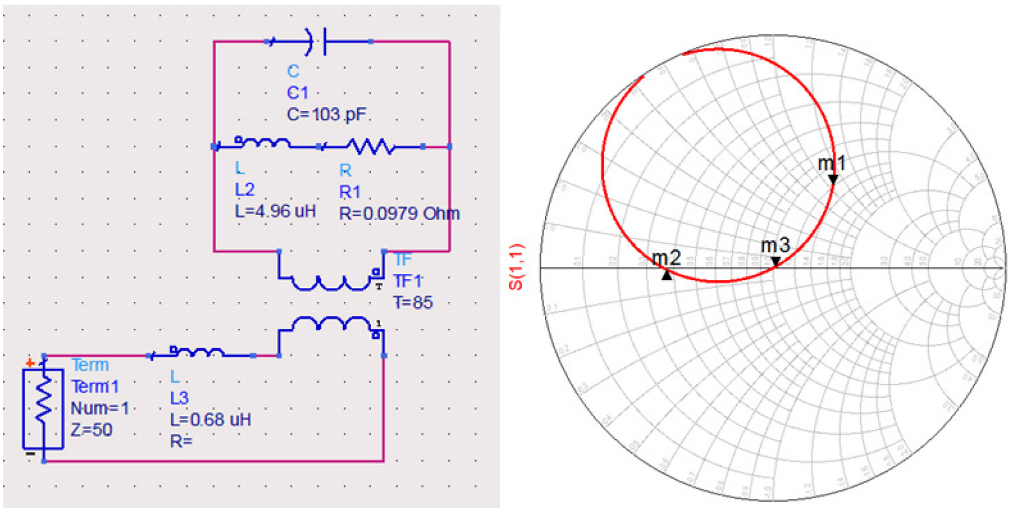

Figure 6 Equivalent circuit diagram of the MLA with coupling loop for matching to 50 Ω and Smith Chart as the result of the simulation in ADS.

III**. Measurements and findings from them**

As shown in Figure 1, the examined antenna is not in "free space" but is attached to the metal railing on a small balcony on the second of three floors and its centre is only 1.7 m away from the wall of the house. In this respect, deviations in the antenna properties from the theoretical data must be expected, especially with regard to the losses and the radiation pattern. In addition, to achieve a 50 Ω match at the 40-meter band, a significantly larger coupling loop was required than in the simulation, with a loop length of 100 cm instead of 71 cm.

The most important measurement that characterizes the MLA is the reflection factor at the coupling loop terminals. The full information is only discovered with a Vector Network Analyzer (VNA), so that a comparison with the theoretical results is possible. My measurement with a NanoVNA-F.V2 is shown in Figure 7. Here you can see the circular locus of the reflection factor representative of a parallel resonant circuit, same as known from the simulation. However, its bandwidth is obviously much higher than expected from the simulation: The distance between the two markers with a

VSWR of around 2.6 is 24 kHz here instead of the 3.5 kHz in the EZNEC simulation model. Within the 100 kHz sweep bandwidth, the reflection factor circle does not close completely, so the shift in the start and end points of the circle must be estimated. This shift roughly corresponds to an inductance of $L_C \approx 1$ µH, appropriate to the length of the coupling loop used.

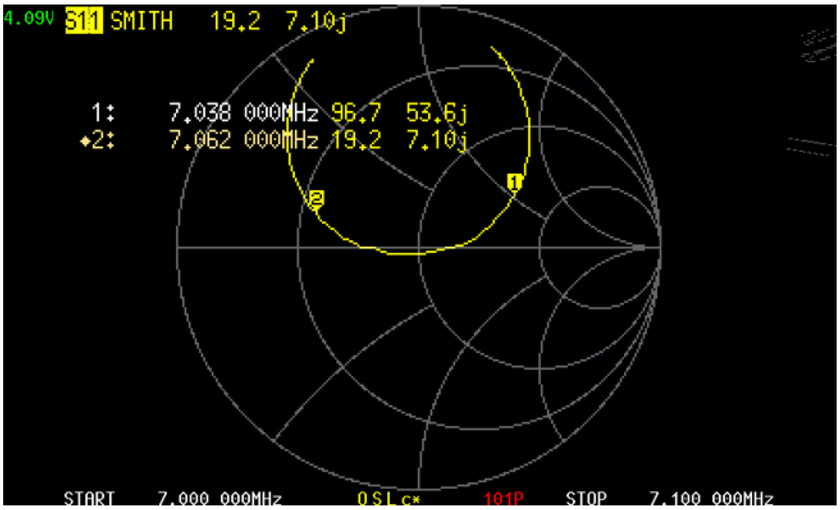

Figure 7 Reflection factor of the realized MLA from 7.0 to 7.1 MHz in the display of the NanoVNA-F.

How can the greatly increased bandwidth be explained? At least one modification of the simulation model was easily found, which also results in this bandwidth in the simulation model: We add a value of about 0.6 Ω to the loss resistance and increase the coupling loop length for impedance matching and thus get a good approximation in Figure 8 through the simulation result.

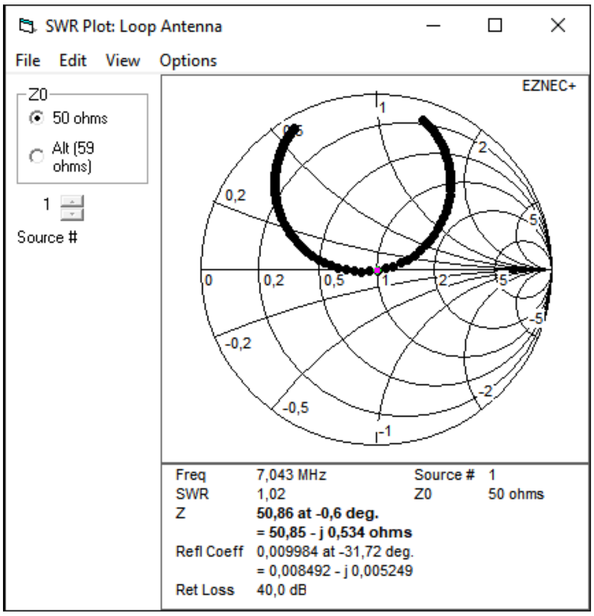

Figure 8 Smith Chart of the simulation model of the MLA with enlarged coupling loop and additional resistance of 0.6 Ω

To compare the bandwidth with that of other antennas of the same size, the bandwidth usually measured at VSWR = 2.0 can be used. With Δf = 17 kHz, our own antenna is significantly worse than the bandwidth of 12 kHz mentioned in an earlier data sheet /8/ for the proven AMA 82 (construction according to DK5CZ, manufacturer: WiMo), for which, however, an additional resistance of around 0.4 Ω must be applied. Other information from ham projects of MLA of the same size is not available. However, information on measured bandwidths can be evaluated for smaller MLA projects, with similarly large deviations from the theoretically expected bandwidths. For example, in /5/ Frank Dörenberg shows an MLA with a diameter of 1 m, for which he measures a bandwidth of 10 kHz, but which only comes to around 3 kHz in the simulation without additional resistance; to achieve the measured bandwidth, an additional resistance of 0.2 Ω would have to be used in the simulation. In /9/, Alan Boswell and colleagues also examine a 1 m MLA and arrive at a similar bandwidth, for which they blame an additional resistance of 0.25 Ω. An AMA82 with a diameter of 0.8 m would have to have an additional resistance of 0.11 Ω in order to achieve the bandwidth of 10 kHz given in /8/ as an empirical mean.

There are two possible reasons for increasing the bandwidth, i.e., the attenuation of the antenna resonator: On the one hand, the loss resistance of the capacitor has not yet been taken into account in the MLA model. Even variable capacitors with air dielectric have AC resistances on the conductors and losses due to contact resistances in the connecting contacts to the conductor loop and in the contacts within the plate packs of the capacitor. In addition, there are dielectric dissipation losses in the insulating materials that hold the rotor and the static plate packs together. Good capacitors can achieve a Q-factor in the thousands, but even a practically achievable Q of 2000 would leave an ESR of 0.11 Ω in series with our 103 pF capacitor (with a reactance of 225 Ω). To estimate the capacitor Q-factor, the variable capacitor installed in the MLA, Figure 9, was replaced by a leaded ceramic high-voltage capacitor which resulted in a bandwidth of the antenna that was even 1 kHz lower. This means that the variable capacitor probably has a Q-factor of only 1000 and with a loss resistance of around 0.2 Ω contributes significantly to the attenuation of the antenna resonator.

Figure 9 The variable capacitor (upper part) of the MLA with drive motors (below)

On the other hand, the interaction of the antenna's near fields with the immediate surroundings of the antenna can also have a major influence on the Q-factor of the MLA resonance. The reactive near fields of an antenna in free space store magnetic and electric energy loss-free in a volume around the antenna. This is limited by a radius of about λ/2π for electrically small antennas, i.e., a radius of about 6 m at a wavelength of about 40 m. If material is introduced into this space around the antenna, currents or displacement currents can be induced therein, which extract power from the stored energy, and this has the same effect on the power balance as an additional resistance in the antenna resonant circuit.

A look at the radiation resistance R_R shows, for example, that installing the MLA close to an electrically conductive surface can mean an increase of up to twofold, since the reflection on the "image plane" changes the radiation pattern considerably and the effective aperture of the antenna appears to be enlarged. However, the placement over real ground is rather disadvantageous, since the ground is not a perfect conductor, so that fields can penetrate the conductive ground. Thus, the electric and magnetic near fields of the large conductor loop induce currents in the ground below the MLA, which heat up the ground like in a microwave oven. This means that part of the transmission power is lost there - represented in the model by a significant increase in the loss resistance in practice when the MLA sits close to the ground. In a setup of our MLA with the lower edge approx. 0.8 m above the ground of a meadow in front of the house, the antenna bandwidth was about the same, i.e., the same additional losses as the setup on the balcony. A simulation with EZNEC shows that the near fields of the MLA decrease steeply with the distance from the large conductor loop, but the increase in the loss resistance only comes down to the order of the radiation resistance above about 6 m height of the MLA over ground. The great importance of the possible power losses in the ground under the MLA for the bandwidth of the antenna and its efficiency has already been examined in various articles. E.g., in /10/, Owen Duffy uses simulation to show quantitatively the influence of the MLA height over ground on the additional resistance, which he calls equivalent ground loss resistance.

However, the situation of our MLA on the balcony is dominated by the galvanized steel rods of the railing and the steel support structure, where conductor currents are mainly excited by the magnetic near field of the antenna. In the simulation, these conductor structures and additional metallic frames of the windows and the steel reinforcement of the concrete ceilings of the house were modelled as interconnected conductor bars. However, EZNEC+ cannot use different conductivity for different conductors, so these conductor structures are modelled as copper conductors. For comparison, this configuration was also simulated with EZNEC Pro 2+, where lower conductivity conductor bars could be defined. Both simulations did not result in a clear difference in the bandwidth compared to the free space case, since the induced currents in the conductors were only up to 50 mA with a current in the large conductor loop of the MLA of around 8 A; this ratio is important because the power converted into heat loss in the conductors steeply decreases with the square of the conductor current. Even after inserting 1 Ω resistors in the conductors with the highest currents, no increase in bandwidth was discernible. From this it can be concluded that conductor structures in the vicinity of an MLA do not necessarily lead to significant attenuation loss of the antenna, unless the conductors form resonant structures, such as a dipole antenna with a terminating resistor, in which high currents are excited and correspondingly high power losses can be generated. The simulation of the MLA on the balcony also does not show that the metallic conductor structures would significantly change the far-field radiation pattern compared to an MLA in free space. This also corresponds to experience in practical operation of the antenna, where good omni-directional radiation characteristics are found for Europe connections. The proximity of the antenna to the building can therefore hardly have a significant influence on the radiation resistance of the MLA.

With this result, the interaction of the near fields of the antenna with the building remains as the major reason for the attenuating influence of the environment of the antenna. Since the near field still has very high field strengths up to a distance of a few meters, especially the electric fields can induce dielectric losses (heating due to displacement currents) in the nearby house wall, the concrete ceilings and nearby interior walls and apartment furnishings. Unfortunately, these losses cannot be simulated with EZNEC, since only conductor structures can be represented. An indication of the attenuating properties of the building was already obtained during the test operation of the MLA in the shack: resonance tuning of the MLA was also possible inside the apartment, but the bandwidth was almost twice as large as after installing the antenna outside at the balcony. A multiple comparison of the received field strengths at a Web-SDR station from transmissions by the MLA resulted in at least one S-level less signal from the room than with transmissions from the balcony.

However, it remains to be checked, ideally by means of an experiment, whether the measured increased bandwidth actually is properly represented by an additional resistance in the antenna resonator. In this case, when fed with a given transmit power, the current in the large conductor loop would also have to be significantly smaller than in the case of the MLA without an additional resistance. A measurement without loading and distorting the antenna resonator is by determining the magnetic field strength H using a probe at a distance of a few meters in the near field, as shown in Figure 10: The measurement is made with a shielded conductor loop of D = 14 cm diameter and area $A=\pi\, D^2/4$ in which a voltage $U_{ind}= \omega\, \mu_0\, H\, A$ is induced by the magnetic flux $\phi = B\, A = \mu_0\, H\, A$ of the MLA according to the induction law. This voltage is transmitted via a coaxial line to an oscilloscope with an input resistance of 50 Ω. A voltage drop due to the inductance of the conductor loop must be taken into account as well as the attenuation of the line. The conductor loop must be aligned parallel to the large conductor loop of the MLA and positioned exactly in the x-axis of the arrangement. Even if a shielded conductor loop is used, there is still a coupling of the strong y-directed E-field generated by the MLA due to asymmetries in the probe structure, so that a second measurement must be made for compensation with a 180° rotation of the probe around its axis and averaging the two readings. The result of the measurements can also be seen in Figure 10: With a transmission power of 50 W, four measured values at distances between around 2 and 5 m were recorded and entered in the diagram. For comparison, the expected variation of the magnetic field strength as a function of the distance from the MLA for the simulation model with an additional resistance of R_{zus}= 0.6 Ω were also drawn; a satisfactory agreement can be seen here, so the assumption of the additional resistance is plausible. Much higher field strengths would be expected if the MLA would not suffer from an additional resistance, as can be seen from the curve for R_{zus}= 0, which is also drawn for this case.

For the simulation model, EZNEC also calculates the rms current in the large conductor loop. Without an additional resistance, the result would be around 20 A and with R_{zus}= 0.6 Ω only around 8 A. The product of the current and the reactance of the capacitor (here around 230 Ω) results in the effective voltage at the capacitor, i.e., "only" around 1840 V (2600V - peak) in the real situation instead of 2.5 times the value. The plate spacing of the variable capacitor of 1.5 mm was therefore completely adequate.

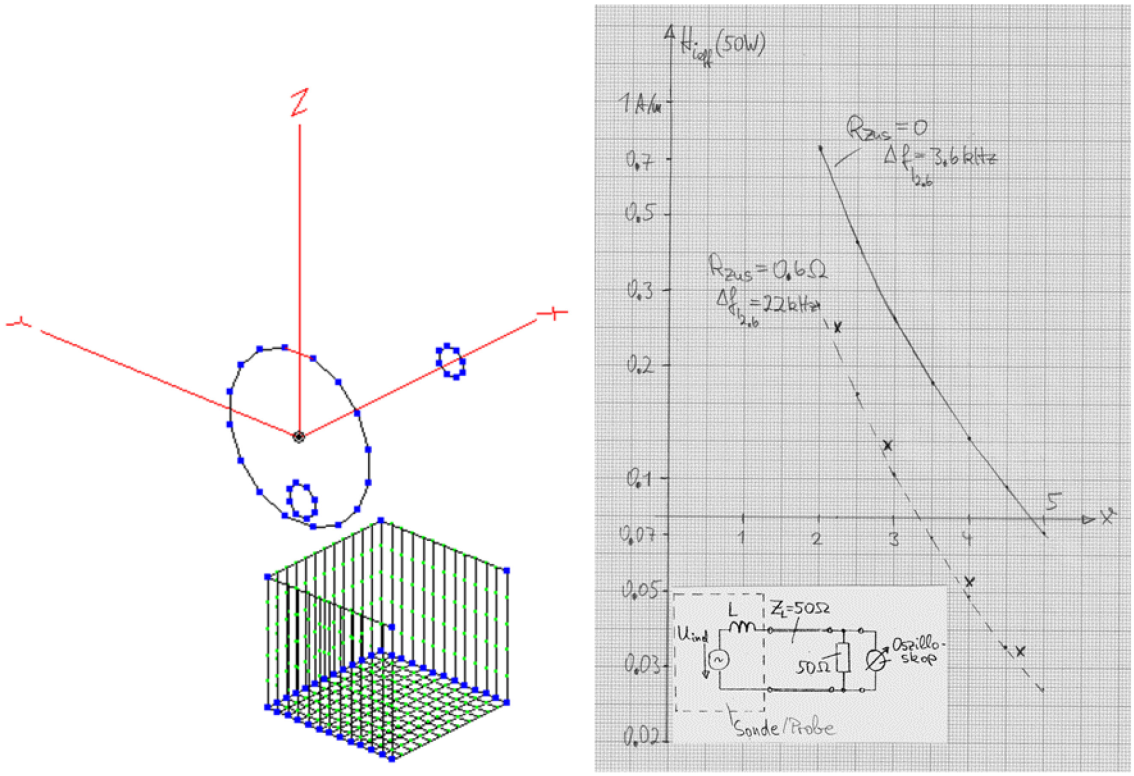

Figure 10 Model of the MLA in front of the balcony and test probe at a distance of 2 m on the x-axis as well as simulation and measurement results of the magnetic field strength when powered with 50 W

The safety distance on the x-axis can also be read from the plot of the measured values: With the present limit value of around 0.1 A/m, there is a safety distance of around 3.1 m for a transmit power of 50 W. This result for the safety distance fits well with the measurement results from an earlier publication /11/, in which a 1.7 m MLA also was examined. In Figure 9 you can see that the simulation of our antenna without additional resistance would result in a much larger safety distance of around 4.4 m. The attenuation of the antenna due to losses in the capacitor and in the immediate vicinity not only ensure that a larger part of the transmit power is lost as heat, but also that the current in the conductor loop at resonance does not reach the level that it would reach without the additional losses; with a lower current, the field strengths then drop proportionally and the safety distance to be maintained from the antenna is reduced (note: but not proportionally).

If one assumes that the additional resistance in the MLA is essentially due to losses in the contacting resistance to the capacitor, in the capacitor itself and through power loss from the near fields of the antenna, the additional resistance also has an effect on the efficiency of the MLA, since this lost power does not contribute to the far-field power. In the simulation with EZNEC+, the efficiency is determined by calculation of the "average gain"; with a lossless antenna, this value is 1 or 0 dB. In the examined antenna, the efficiency decreases from 0.45 (R_{zus} = 0) to 0.075 (R_{zus} = 0.6 Ω), correspondingly from -3.5 dB to -11.3 dB. This means that the power actually radiated decreases from 45% of the transmitter output power to just 7.5% due to the additional loss resistance; in receive operation, the reception level would thus be about 1.5 S-levels lower. Figure 11 shows the over-all distribution of the transmit power into the various losses and the radiation.

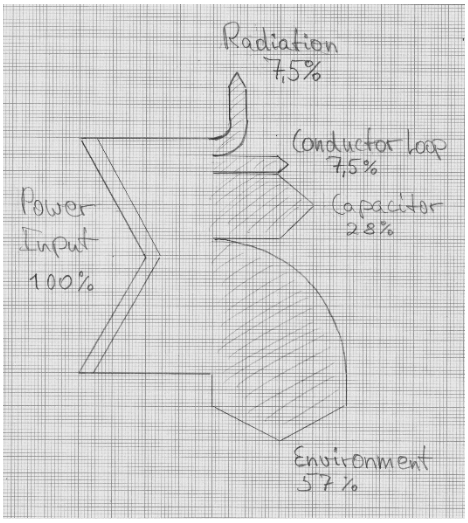

Figure 11 Distribution of input power into dissipation loss and radiation components for the realized MLA

The measured bandwidth of the MLA is therefore decisive for determining the actual efficiency of the MLA, without it having to be clear where exactly the additional losses take place in the antenna system: In a first step, the actual “unloaded” quality factor Q can be determined from the measured impedance bandwidth Δf (at VSWR=2.62) and the operating frequency. The inductance L and the inductive reactance X_L as well as the radiation resistance R_R can be calculated from the diameter of the large conductor loop using the simulation or the formulas. The actual resistance $R = (R_R + R_L)$ in the resonant circuit then results from the Q and the reactance, and thus η can be determined directly from formula (d). You can use Owen Duffy's calculator /12/ without having to do the math yourself.

The loss of antenna gain due to a deterioration in efficiency is of course bad news, but this loss also applies to the reception of the particularly high level of interference noise from our apartment building ("electro-smog"), so that the signal - to - noise power ratio (S/N) does not suffer in normal reception operation, since distant signals and local noise are attenuated to the same extent. Another positive side effect, as shown above, is the reduction of the safety distance, which is particularly helpful when installing the antenna close to the house.

IV. Confusion about “bandwidths”

The most important role in characterizing the MLA up to this point has been the "bandwidth" of the antenna. First, we used the so-called -3 dB - bandwidth Δf , which characterizes RLC resonant circuits and corresponds to the quotient of resonant frequency and resonator “unloaded” quality factor. With the -3 dB bandwidth Δf, the magnitude of the impedance of a parallel resonant circuit decays to a factor of $1/\sqrt{2}$, i.e., by 3 dB. In turn, the resonator Q is the ratio of the reactance of the inductance or the capacitance at the resonant frequency and the loss resistance in the circuit. If a resonator is impedance matched to the generator impedance, e.g., by a transformer, the standing wave ratio increases towards the bandwidth limits to VSWR = 2.62 and the reflection factor to 0.447. The more common definition of antenna bandwidth uses the frequencies where VSWR = 2.0; this bandwidth is smaller than the -3 dB bandwidth by a factor of $1/\sqrt{2}$. Both definitions of the bandwidth refer exclusively to the impedance of the antenna, related to the generator impedance (usually 50 Ω). Accordingly, these bandwidths are measured with a directional coupler, standing wave meter or vector network analyzer (VNA).

But there is a second bandwidth in the operation of an antenna when the impedance matched antenna is connected to a receiver. A wave from a distant transmitter incident on the antenna induces a voltage U_{ind} in the antenna, so that in this case the antenna acts as a generator. Due to impedance match, the antenna transmits the maximum possible power of the received signal to the receiver, which then measures and displays the corresponding received voltage U_{Rx}. Figure 12 shows an equivalent circuit representation of this situation. The antenna impedance at the terminals of the coupling loop is represented as a parallel resonant circuit with a parallel resistance of R_P = 50 Ω and a voltage source in series with the inductance, and the receiver is defined only by a volt-meter and its input impedance, the characteristic impedance of the line, as $R = Z_L$= 50 Ω. Put simply, the receiver can be understood as a volt-meter that measures the received voltage U_{Rx} at its input. You can see that the two same resistors are now connected in parallel, which means that the attenuation of the antenna resonant circuit is doubled by the receiver input resistance; in the case of filter circuits, this is referred to as "critical coupling". The resulting quality factor of the antenna resonant circuit is called "loaded"-Q or Q_L and is only half the "unloaded"-Q of the antenna. This is the reason why the resistance is doubled in the calculation of the quality factor Q_L of the resonant circuit similar to formula (e): Q_L=X_L/2R. This quotient can be found in /4/ and many later articles, as well as the calculation of the "bandwidth" on this basis, without mentioning that the "unloaded" quality factor Q of the antenna is twice as high and the impedance matching bandwidth Δ*f* is only half as large.

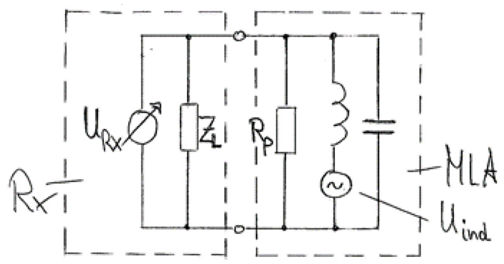

Figure 12 Simplified equivalent circuit for receiving a signal with the MLA

With twice as large a bandwidth 2Δ*f* following from Q_L, the standing wave ratio of the antenna at the corner frequencies increases to VSWR = 5.83 and the reflection factor to $1/\sqrt{2}$ or -3 dB. This means that the antenna reflects half the input power at these frequencies when transmitting. When receiving, the received power also drops by 3 dB, which is easy to show, e.g., with a two-port transmission measurement using a VNA or with the reception of a wide noise spectrum by an SDR receiver. Figure 13 shows the result of a two-port VNA measurement, in which the transmit signal of the VNA (port 1) is emitted near the MLA from a weakly coupled small probe and the signal received by the MLA is fed into the second port of the VNA, which acts here as a matched receiver. The antenna is tuned at about 7.05 MHz, where transmission is maximum, and the -3 dB frequencies in the transmission characteristic are about 47 kHz apart; this bandwidth is therefore twice as large as the measured impedance match bandwidth of the MLA. We get a similar result with the reception of a wide noise spectrum from a noise generator, which in turn is radiated from the probe into the MLA and its receive signal recorded with an SDR receiver. An attenuator at the SDR input must be used to ensure that the receiver actually offers an input resistance of 50 Ω. Figure 13 also shows the result of this measurement when the antenna is tuned to around 7.085 MHz, where the reception level is at its maximum. Some strong 40-meter band reception signals are still above the noise signal, especially those of the digital modes at 7.074 MHz. The spacing of the frequencies at which the received level of the noise signal has dropped by 3 dB is around 48 kHz, which is also slightly more than a doubling of the impedance match bandwidth of the MLA.

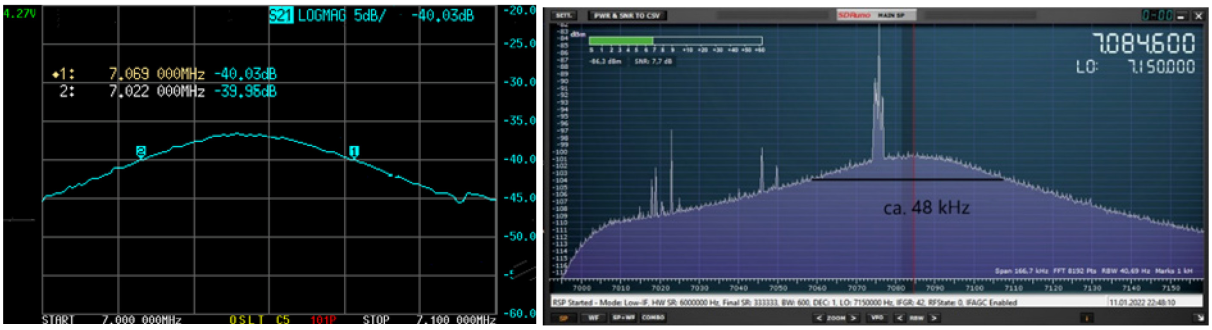

Figure 13 Transmission measurement (S21) of the MLA with the VNA (left) and received noise spectrum of a "noise source" displayed by an "SDR Play" receiver (right).

V. What still needs to be explained

The MLA can be detuned over a large frequency range by adjusting the capacitor, the tuning range essentially depending on the realizable capacitance range. In many MLA construction projects, two octaves can be swept in one MLA, e.g., 3.5 MHz to 14 MHz. However, optimal impedance matching requires "fine tuning" of the transformer ratio by deforming the coupling loop; for example, the transformer ratio *N* is increased with a smaller loop area (compressed circular shape) but reduced with a bending of the coupling loop closer to the large conductor loop. Thereafter, acceptable matching is maintained across the entire frequency range without having to change the coupling loop for each band.

However, this is extremely astonishing in view of the strong frequency dependence of the resistances in our model. The radiation resistance of the large conductor loop increases with the fourth power of the frequency and the AC resistance of the conductor loop increases with the square root of the frequency. Without considering the additional resistance in our MLA, the resistance *R* of our MLA would increase over the 1:4 frequency ratio (3.5 MHz/14 MHz) from about 0.04 Ω to 0.8 Ω, i.e. at the ratio of 1:20. However, a resistance appears at the terminals of the coupling loop with a much lower frequency dependence. This is mainly because the resistance *R* is in series with the inductance *L* of the large conductor loop: At any given frequency, this series connection can be converted into an equivalent parallel circuit, as shown in Figure 14. By this, a parallel resistance *R*' appears, which depends on the square of the reactance of the inductor: $R' \approx \frac{(\omega L)^2}{R}$. It can be seen that the parallel resistance value, in contrast to the series resistance, remains within narrow limits, since the numerator increases by a factor of 16 with the square of the frequency, while the denominator increases by a factor of 20; this means that the resistance transformation of the coupling loop does not necessarily has to be changed. This conversion into a parallel resistance is the reason why the original series RLC tank circuit, from the transmitter's point of view, appears as a parallel RLC tank circuit, where the transformer only converts the impedance of typically hundreds of kΩ of *R*’ to the 50 Ω impedance to match the transmission line.

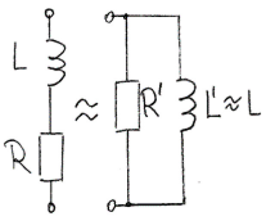

Figure 14 Equivalent conversion of a series circuit into a parallel circuit for small resistance values *R*

The impedance of a parallel resonant circuit, which is connected in series with the inductance of the coupling loop, is therefore measured at the terminals of the coupling loop. In the Smith Chart of the reflection factor, the circular graph belongs to the parallel resonant circuit, while the reactance of the coupling loop inductance rotates this circle further and further clockwise along the perimeter of the Smith Chart with increasing frequency. The diameter of the circle of the reflection factor becomes smaller and smaller by this transformation, although the real part (the resistance value) of the antenna impedance remains approximately the same. This pattern can be seen in Figure 15, where plots are shown side by side of the MLA for 3.5 MHz, 7 MHz and 14 MHz as calculated with realistic additional resistances. The coupling loop is obviously set optimally for the 40-meter band (critical coupling), so that the coupling appears overcritical in the 80-meter band and undercritical in the 20-meter band, but the VSWR remains below 1.65.

Unfortunately, the coupling loop inductance becomes problematic at the higher frequencies; to keep the effect of the inductance of the coupling loop as low as possible, the largest possible conductor cross-section should be used, and the coupling loop should be placed close to the large conductor loop in order to minimize the required length of the coupling loop and thus its inductance. If the effect of the series inductance of the coupling loop prevails in the upper frequency band, a series capacitance in the feed line can also help to reduce the mismatch in this frequency range by rotating back the circle of the reflection factor a bit; however, the series capacitance must be chosen carefully, since it also rotates the circles in the lower frequency bands many times more than in the upper frequency band (the capacitive reactance increases inversely to the frequency). Another effective way is to use a short coax line and series inductor as a matching circuit directly placed at the base of the antenna.

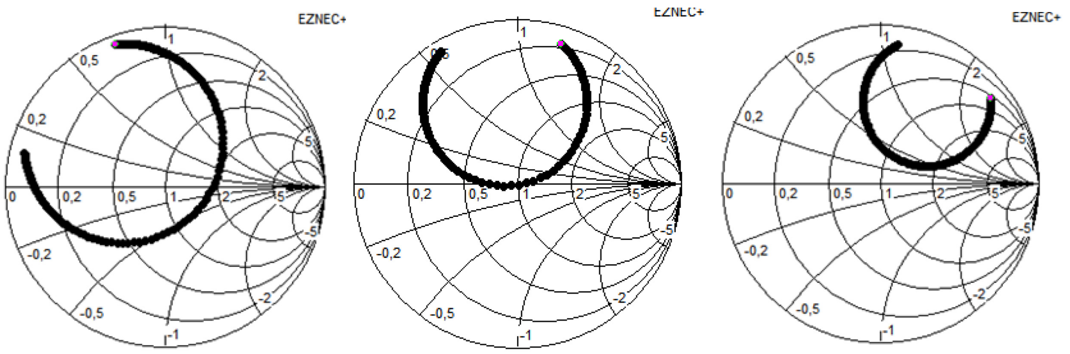

Figure 15 Simulated reflection factor plots in the 80-meter band (left), 40-meter band (middle) and 20-meter band (right) of the MLA with realistically assumed additional resistances.

References

/1/ Hans Würtz, "DX antennas with reflecting surfaces", Part 12, cq-DL 2/1983, pp. 64 - 67.

/2/ Hans Würtz, "DX antennas with reflecting surfaces", cq-DL 4/1983, pp. 170 - 171.

/3/ Rothammel's antenna book, chapter "Magnetic antennas", DARC Verlag

/4/ The American Radio Relay League (1988): “Small High Efficiency Loop Antennas for Transmitting”. The ARRL A Handbook, 15th Edition, pp. 5-11 to 5-17.

/5/ Exciting reports on the design and manufacture of MLAs as well as a large collection of references regarding MLAs can be found on the web page of Frank Dörenberg, N4SPP (accessed Feb. 4th, 2022): https://www.nonstopsystems. com/radio/frank_radio_antenna_magloop-small.htm

/6/ Theoretical and practical contributions to the construction and evaluation of MLAs as well as a list of the few scientific publications on MLA can be found on the web page of Steve Yates, AA5TB (accessed Feb. 4, 2022): http://www. aa5tb.com/loop.html

/7/ (as of Feb. 15, 2022): https://eznec.com

/8/ (as of Feb. 15, 2022): https://docplayer.org/47738314-tunable-magnetic-antennas.html

/9/ Alan Boswell et al., "Performance of a small loop antenna in the 3 – 10 MHz band", IEEE Antennas and Propagation Magazine, vol. 47, no. 2, April 2005, p. 51-56

/10/ Owen Duffy, “Small transmitting loop – ground loss relationship to radiation resistance”, in (accessed Feb. 4, 2022) https://owenduffy.net/blog/?p=4888

/11/ Th. Moliere, DL7AV, "Field strength measurements on a magnetic antenna", cq DL 4/99, pp.316-317

/12/ Owen Duffy,” Calculate small transmitting loop gain from bandwidth measurement”, in (accessed Feb. 4, 2022) https://owenduffy.net/calc/SmallTransmittingLoopBw2Gain.htm