A LOOK AT A NOVEL ANTENNA DESIGN

Prof. Dr.-Ing habil. Ulrich L. Rohde (N1UL)

Department of Electrical Engineering, School of Computation, Information and Technology, Technical University of Munich

Prof. James K. Breakall (WA3FET)

Department of Electrical Engineering, School of Electrical Engineering and Computer Science, Penn State University, University Park, PA 16802

INTRODUCTION

- This discussion is triggered in part by the book [1], and the need to have small antennas with good front to back ratios. Needed for this are modern mathematical tools as well as test equipment.
- As an example of mathematical tools, Hallén wrote his famous integral equation to give an exact treatment of antenna current wave reflection at the end of a tube shaped cylindrical antenna in 1956, but his first work on this subject [2] probably goes back to 1938. This equation enabled him to show that on a thin wire the current distribution is approximately sinusoidal and propagates with nearly the speed of the light. For this treatment that is an important contribution.
- In general, following the common tradition we assume all ideal conditions for a better understanding of the topic!
- Antennas are key for communication, both for receive and transmit and there is a certain fascination with their design

- [1] Schaefer, CL.: Einführung in die theoretische Physik. (Introduction in to the theoretical Physics), Bd. 3/1 S. 331, Berlin: W.de Gruyter 1932
- [2] Burstyn W.: Die Strahlung und Richtwirkung einiger Luftdrahtformen im freien Raum (Radiation and Direction of Different styles of Open Wire Antennas), Jb. drahtl. Telegr. Bd. 13 (1918) S. 362]

ELECTRICAL DIPOLE

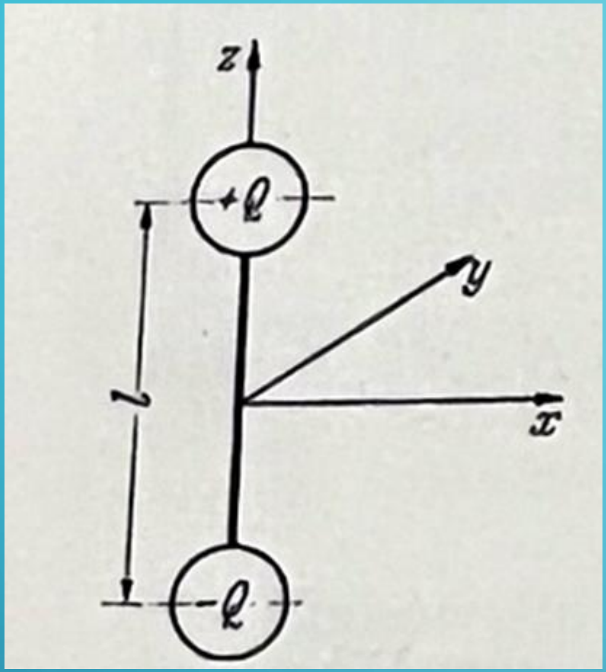


Fig. 1. (a) Electrical Dipole, charges +q and -q are changing periodically and harmonically as a function of T (sinusoidal)

- It all started with the Hertzian dipole and its understanding and mathematical treatment [1].
- We will learn here that by optimizing the effective length of a rod antenna, it will become directional and will have a radiation pattern that provides gain.
- Any antenna has a physical dimension, and the most relevant one is the antenna height.
- For a ground mounted vertical antenna, the mirror image in the ground is also considered, and the total length is called the effective length.
- The definition of "effective length" is going to be the key point and its value can generate some "interesting effects".
- In this discussion, only the magnitude of l_{eff} is of interest. For the definition of 'l', we need to look at Fig. 1(a).

$$i = -\frac{dQ}{dt} = -\frac{1}{l}\frac{df(t)}{dt} = +\frac{2\pi Q_0}{T}\sin\frac{2\pi t}{T} = +2\pi v Q_0 sin2\pi vt$$

ANTENNA DESIGN

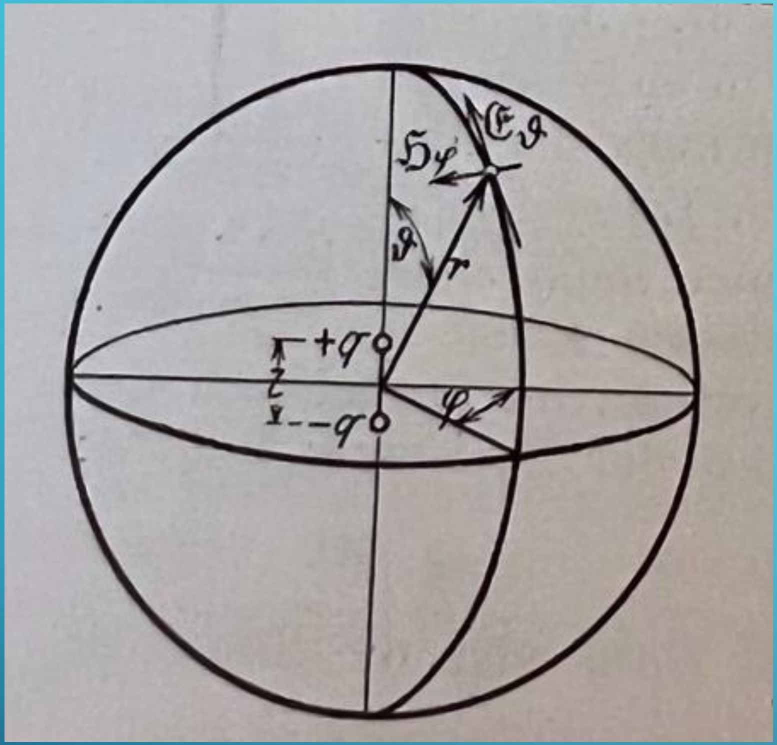


Fig. 1 (b): Polar Co-ordinates around the Dipole, see the charges +q and –q

Now let's look at the real-life situation.

- An electric field (E_i) multiplied with the effective length of the antenna ($l$eff) generates an EMF (electromotorical force).

EFFECTIVE LENGTH

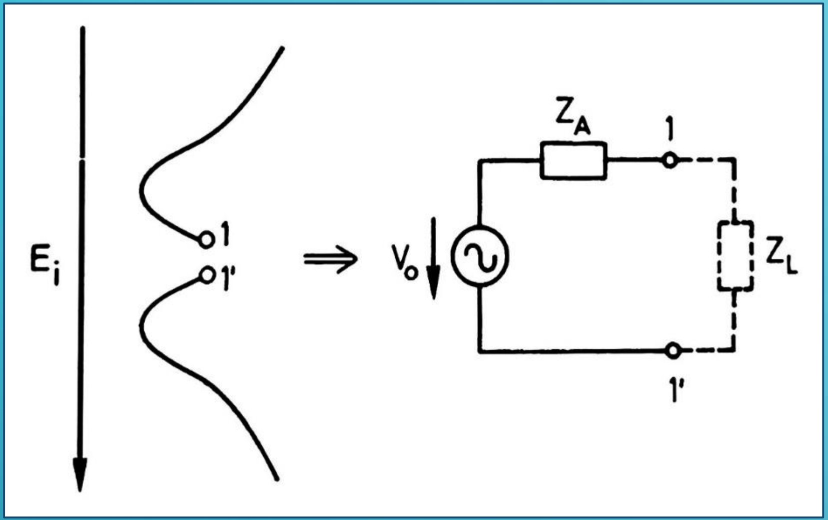


Fig. 2: The equivalent circuit of an antenna in electric field

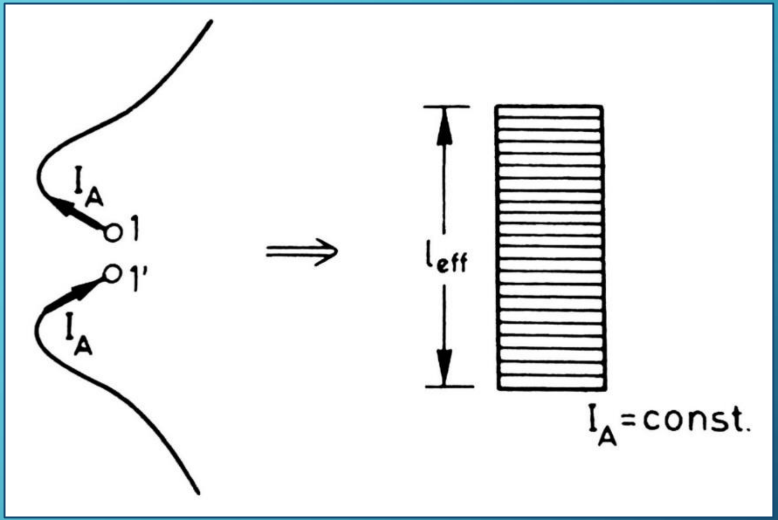


Fig. 3. Showing the effective length l_{eff}

We also notice that the antenna wire is not straight but has a specific shape, this will be the topic of this presentation. The effective length of an antenna multiplied with the electrical field E_i results in an EMF V_0

CALCULATION SHOWING THE VECTORS OF THE ELECTRICAL AND MAGNETIC FIELDS RELATIVE TO THE ISOTROPIC ANTENNA

The following derivation is based in part on [1]

$$\mathfrak{E}x = \frac{3xz}{r^5}f + \frac{3xz}{cr^4}f' + \frac{xz}{c^2r^3}f'' = \left[\frac{3}{r^3}f + \frac{3}{cr^2}f' + \frac{1}{c^2r}f''\right]\cos\varphi\cos\vartheta\sin\vartheta$$

$$\mathfrak{E}y = \frac{3yz}{r^5}f + \frac{3yz}{cr^4}f' + \frac{yz}{c^2r^3}f'' = \left[\frac{3}{r^3}f + \frac{3}{cr^2}f' + \frac{1}{c^2\gamma}f''\right]\sin\varphi\cos\vartheta\sin\vartheta$$

$$\mathfrak{H}z = \frac{3z^2 - r^2}{r^5}f + \frac{3z^2 - r^2}{cr^4}f' + \frac{z^2 - v^2}{c^2r^3}f''$$

$$= -\left[\frac{f}{r^3} + \frac{1}{cr^2}f' + \frac{1}{c^2r}f''\right] + \left[\frac{3}{r^3}f + \frac{3}{cr^2}f' + \frac{1}{c^2r}f''\right]\cos^2\vartheta$$

$$\mathfrak{H}x = \frac{y}{cr^3}f' - \frac{y}{c^2r^2}f'' = -\frac{1}{c}\left[\frac{f'}{v^2} + \frac{1}{cr}f''\right]\sin\varphi\sin\vartheta$$

$$\mathfrak{H}y = \frac{x}{cr^3}f' + \frac{x}{c^2r^2}f'' = +\frac{1}{c}\left[\frac{f'}{v^2} + \frac{1}{cr}f''\right]\cos\varphi\sin\vartheta$$

$$\mathfrak{H}z = 0$$

$$\mathfrak{E}_\vartheta = -\left[\frac{f}{r^3} + \frac{1}{cv^2}f' + \frac{1}{c^2r}f''\right]\sin\vartheta$$

$$\mathfrak{E}_r = 2\left[\frac{f}{r^3} + \frac{f'}{cr^2}\right]\cos\vartheta$$

DERIVATION CONT'D…

$$\mathfrak{E}_\varrho = \sqrt{\mathfrak{E}_x^2 + \mathfrak{E}_y^2} = \frac{z \cdot \varrho}{r^3}\left[\frac{3f}{v^2} + \frac{3f'}{cv} + \frac{f''}{c^2}\right]$$

$$\mathfrak{E}_z = \frac{2}{v^2}\left[\frac{f}{r} + \frac{f'}{c}\right] - \frac{\varrho^2}{r^3}\left[\frac{3f}{v^2} + \frac{3f'}{cv} + \frac{1}{c^2}f''\right]$$

$$\mathfrak{H}_\varphi = -\frac{\varrho}{r^2}\left[\frac{f'}{cv} + \frac{1}{c}f''\right]$$

$$where\ \varrho = \sqrt{x^2 + y^2}$$

Here are the derivatives referred to above:

$$f = \frac{I_0 l}{\omega}\cos\,\omega\left(t - \frac{r}{c}\right) = Q_0 l\cos\,\omega\left(t - \frac{r}{c}\right)$$

$$f' = -I_0 l\sin\,\omega\left(t - \frac{r}{c}\right) = -Q_0 l\omega\sin\,\omega\left(t - \frac{r}{c}\right)$$

$$f'' = -I_0 l\omega\cos\,\omega\left(t - \frac{r}{c}\right) = -Q_0 l\omega^2\cos\,\omega\left(t - \frac{r}{c}\right).$$

The part (t-r/c), where t=time, r = length of the vector and c=speed of light, indicates that there is a delay of the electromagnetic wave which is delayed at the distance 'r' at the time 't' so the value of the function cannot be achieved at the time 't', the delay is r/c, called retarded.

FOR THE FAR FIELD THE RESULTS ARE:

$$\mathfrak{E}_x = \frac{1}{c^2 r} f'' \cos\varphi \cos\vartheta \sin\vartheta, \; \mathfrak{H}_x = -\frac{1}{c^2 r} f'' \sin\varphi \sin\vartheta,$$

$$\left.\begin{aligned} \mathfrak{E}_y &= \frac{1}{c^2 r} f'' \sin\varphi \cos\vartheta \sin\vartheta, & \mathfrak{H}_y &= +\frac{1}{c^2 r} f'' \cos\varphi \sin\vartheta \\ \mathfrak{E}_z &= -\frac{1}{c^2 r} f'' \sin^2\vartheta, & \mathfrak{H}_z &= 0 \end{aligned}\right\}$$

$$\mathfrak{H}_\varphi = -\frac{1}{c^2 r} f'' \sin\vartheta,$$

$$\mathfrak{E}_r = 0,$$

$$\mathfrak{H}_\varphi = -\frac{1}{c^2 r} f'' \sin\vartheta$$

$$\mathfrak{E}_\varrho = \frac{2\varrho}{c^2 r^3} f'' = \frac{1}{c^2 r} f'' \cos\vartheta \sin\vartheta,$$

$$\mathfrak{E}_z = -\frac{\varrho^2}{r^3 c^2} f'' = -\frac{1}{c^2 r} f'' \sin^2\vartheta,$$

$$\mathfrak{H}_\varphi = -\frac{1}{c^2 r} f'' \sin\vartheta$$

AS A RESULT OF THIS WE CAN CALCULATE THE FIELD STRENGTH AT VARIOUS REGIONS

$$\left.\begin{aligned}
&\mathfrak{E}_\partial = \frac{2\pi I_0 l}{cr\lambda}\sin\ \vartheta\cos\ \omega\left(t-\frac{r}{c}\right) = \frac{4\pi^2 Q_0 l}{\lambda^2 r}\sin\ \vartheta\cos\ \bar{\omega},\\
&\mathfrak{E}_r = 0,\\
&\mathfrak{H}_\varphi = -\frac{2\pi I_0 l}{cr\lambda}\sin\ \theta\cos\ \omega\left(t-\frac{r}{c}\right) = -\frac{4\pi^2 Q_0 l}{\lambda^2 r}\sin\ \vartheta\cos\ \bar{\omega},\\
&\mathfrak{E}_\varrho = -\frac{2\pi I_0 l}{cr\lambda}\sin\ \vartheta\cos\ \vartheta\cos\ \omega\left(t-\frac{r}{c}\right) = -\frac{4\pi^2 Q_0 l}{\lambda^2 r}\sin\ \vartheta\cos\ \vartheta\cos\ \bar{\omega},\\
&\mathfrak{E}_z = \frac{2\pi I_0 l}{cr\lambda}\sin^2\ \vartheta\cos\ \omega\left(t-\frac{r}{c}\right) = \frac{4\pi^2 Q_0 l}{\lambda^2 r}\sin^2\ \vartheta\cos\ \bar{\omega},\\
&\mathfrak{H}_\varphi = -\frac{2\pi I_0 l}{6r\lambda}\sin\ \vartheta\cos\ \omega\left(t-\frac{r}{c}\right) = -\frac{4\pi^2 Q_0 l}{\lambda^2 r}\sin\ \vartheta\cos\ \bar{\omega}.
\end{aligned}\right\}$$

RADIATION CALCULATION

$$\mathfrak{E}_z = \frac{4\pi^2}{\lambda^2}\frac{T_0 l}{c}[\underbrace{\frac{2-3sin^2\vartheta}{a^3}cos\,\omega\left(t-\frac{a\lambda}{2\pi c}\right)}_{\text{Near field}}$$

$$+ \underbrace{\frac{3sin^2\theta-2}{a^2}sin\,\omega\left(t-\frac{a\lambda}{2\pi c}\right)}_{\text{Middle field}} + \underbrace{\frac{sin^2\vartheta}{a}cos\,\omega}_{\text{Farfield}},$$

with $a = 2\pi r/\lambda$.

- The total Radiation is

$$N_{S_T} = \frac{I_0^2\omega^2 l^2}{4\pi c^3}\int_0^{2\pi}\int_0^{\pi}\int_0^{T} cos^2\frac{2\pi}{T}\left(t-\frac{r}{c}\right) d\varphi sin^3\vartheta d\vartheta d$$

$$N_{S_T} = \frac{I_0^2\omega^2 l^2 T}{3c^3} = \frac{4\pi^2 I_0^2}{3c}\left(\frac{l}{\lambda}\right)^2 \cdot T = \frac{16\pi^4 l^2 Q_0^2}{3\lambda^3}.$$

T= time period

CALCULATION AND RADIATION PATTERN

- The energy radiated $N_S = \frac{4\pi^2 I_0^2}{3c}\left(\frac{l}{\lambda}\right)^2 = \frac{8\pi^2 I_{\text{eff}}^2}{3c}\left(\frac{l}{\lambda}\right)^2 = \frac{16\pi^4 l^2 Q_0^2 c}{3\lambda^4}$
- Expressed I_{eff} in Ampere we get $N_S = 80\pi^2\left(\frac{l}{\lambda}\right)^2 I_{\text{eff}}^2$ Watt
- Or the Radiation resistance is $R_S = 80\pi^2\left(\frac{l}{\lambda}\right)^2 \text{Ohm}$
- The resulting radiation pattern as a function of time is shown below

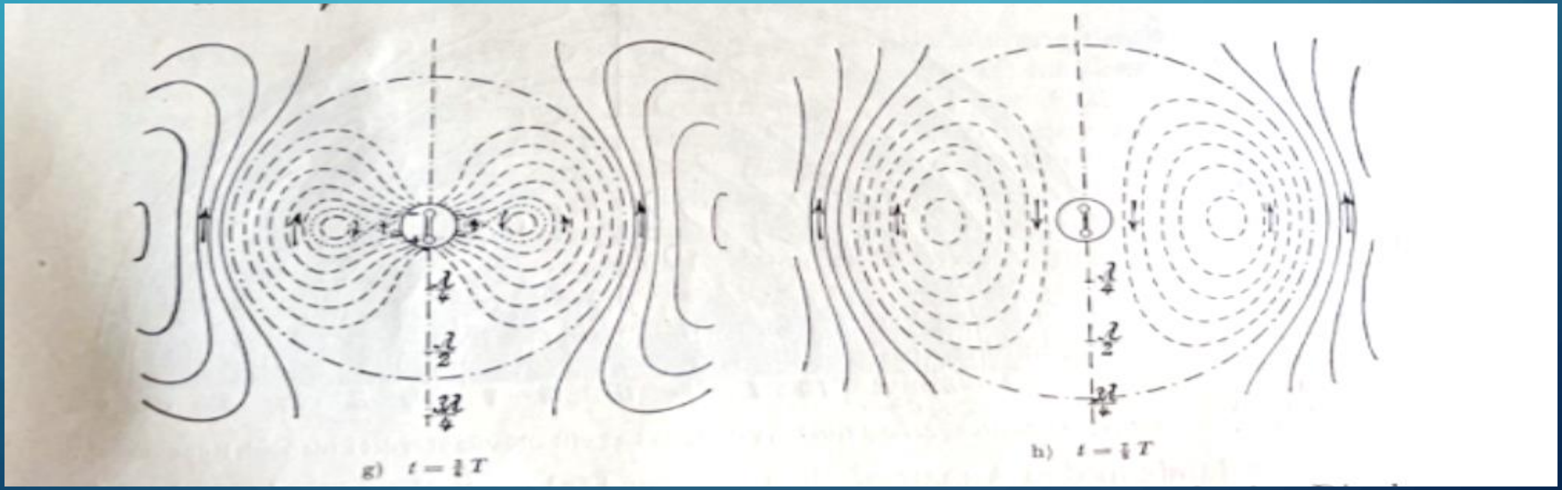


Fig. 4: Radiation pattern with reference to T

SOME INTERESTING FACTS…

- Important: ℓ is not the mechanical length but the electrical length,
- L(effective) $\sim \ell/1.56$
- $R_S = 80\pi^2 \left(\frac{l\,x\,0.64}{\lambda}\right)^2 = 36.6\ \text{Ohms}$
- Typically a factor of 2 is used, instead of 1.56, depending upon the definition of the length (mirror image in the ground).
- If the actual antenna at resonance due to the end effect is slightly inductive, typically it will be cut short and the error at the feeding point (imaginary part) will be compensated by a suitable matching network.
- The exact reason for the 0.64 correction factor is that the voltages and currents are sinusoidal ONLY when the wire is very thin

RADIATION RESISTANCE VS. WAVE LENGTH

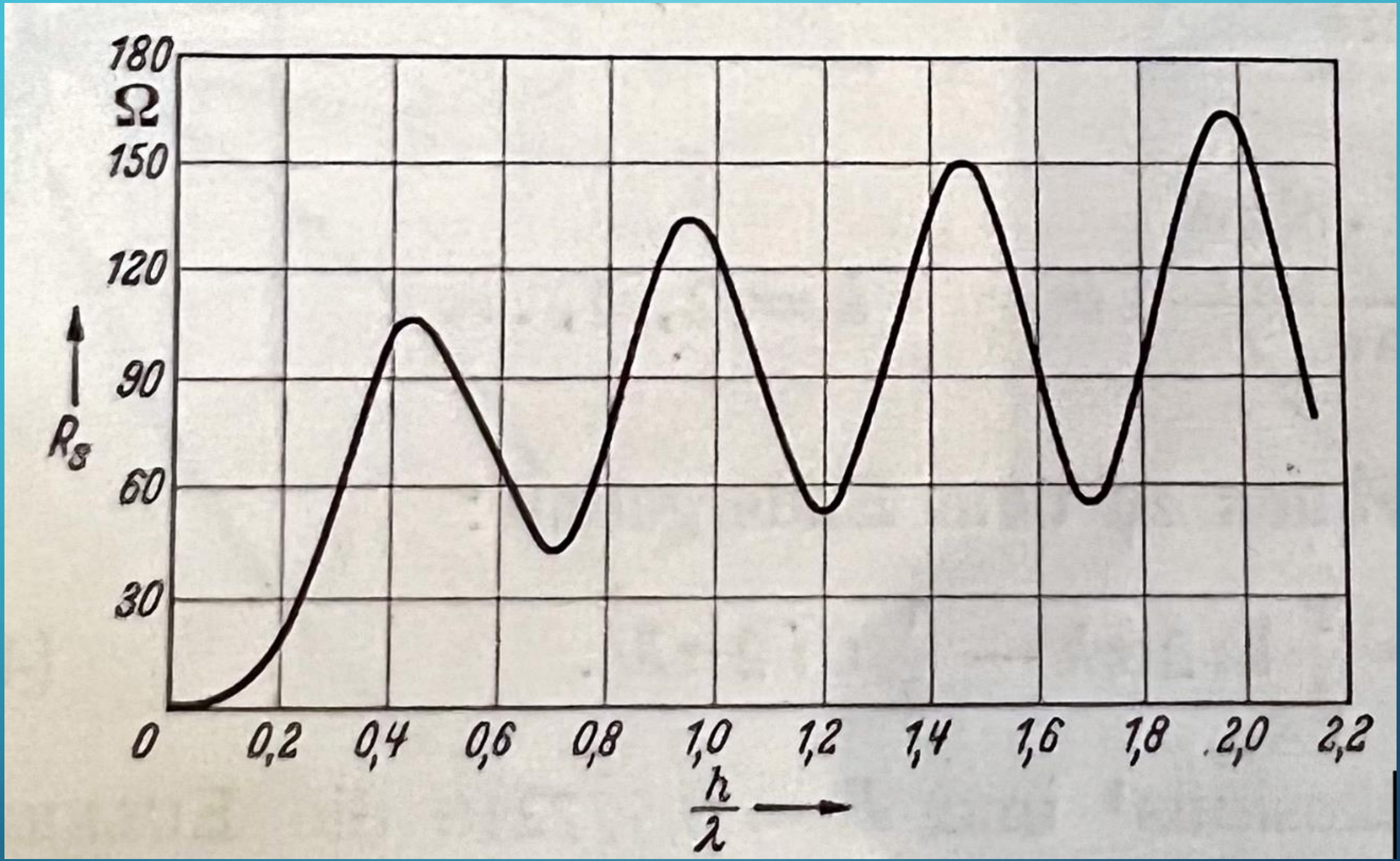


Fig. 5. Plot showing the radiation resistance as a function of the length to wave length ratio.

APPROXIMATE ELECTRICAL FIELD CALCULATION FOR A CURVED RADIATING

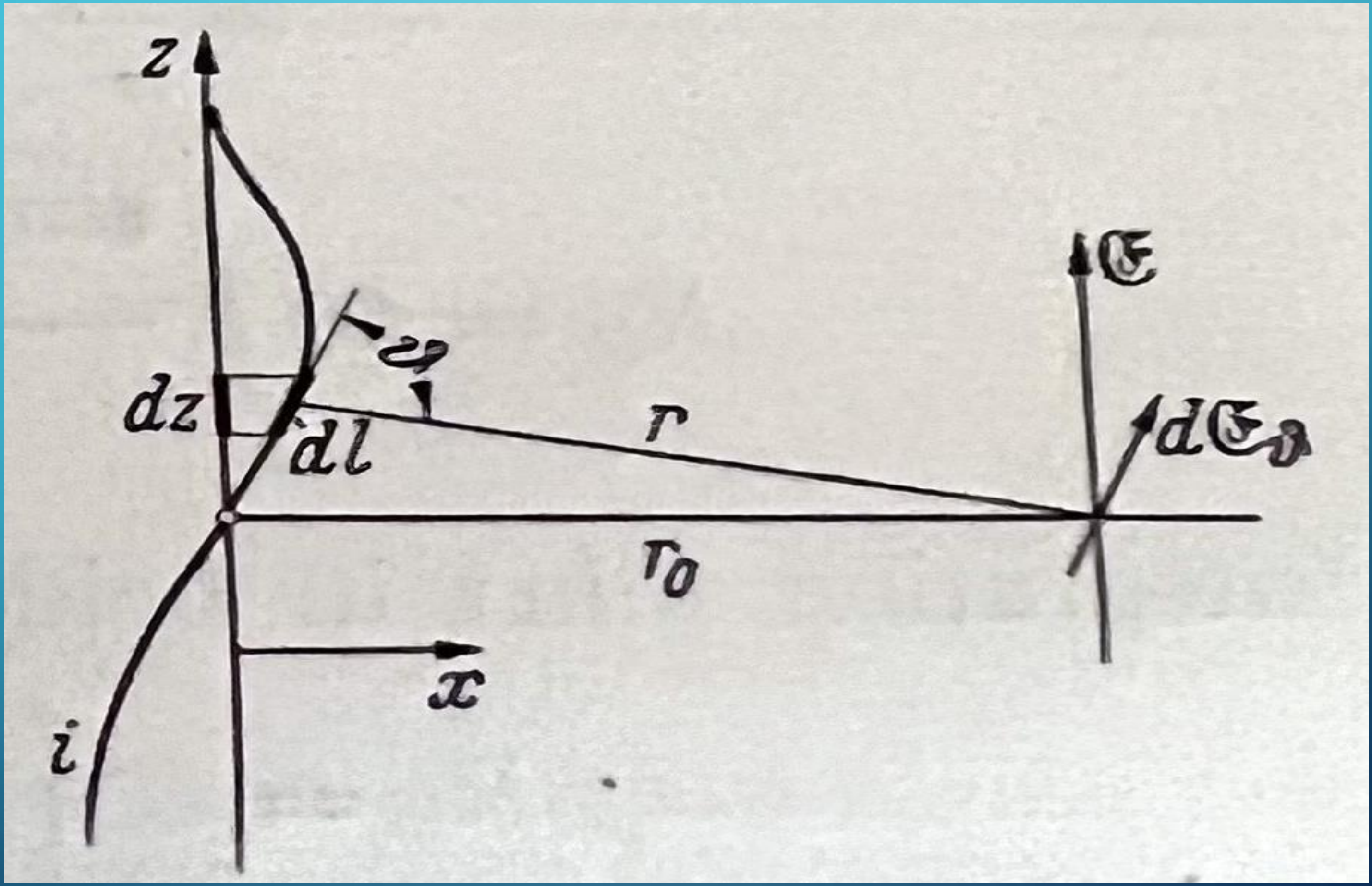


Fig. 6. Approximate electrical field calculation for a curved radiating rod

RADIATION PATTERNS/CURVES AS A FUNCTION OF DIFFERENT TIME ωT

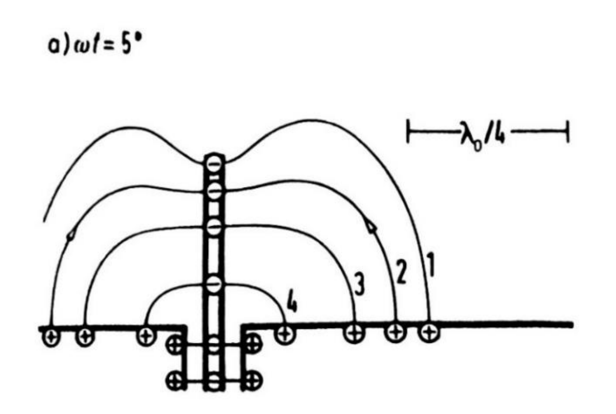


Fig. 7 (a): Radiation patterns as a function of different ωt (ωt=5° of 360°)

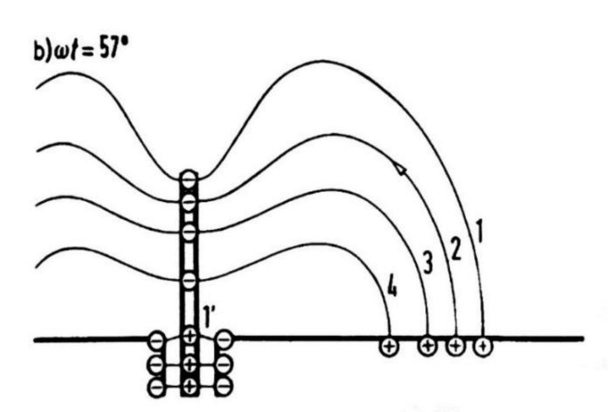


Fig. 7 (b): Radiation patterns as a function of different ωt (ωt=57°)

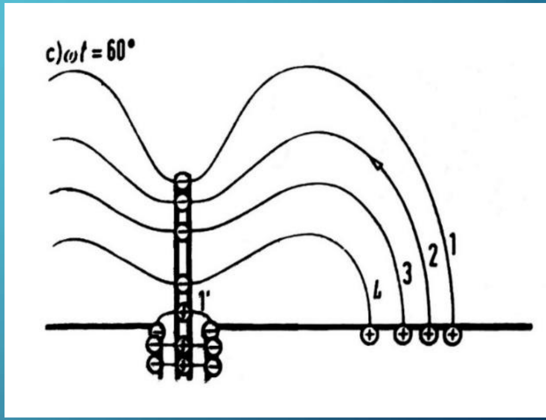


Fig. 7 (c): Radiation patterns as a function of different ωt (ωt=60°)

RADIATION PATTERNS/CURVES AS A FUNCTION OF DIFFERENT TIME ωT

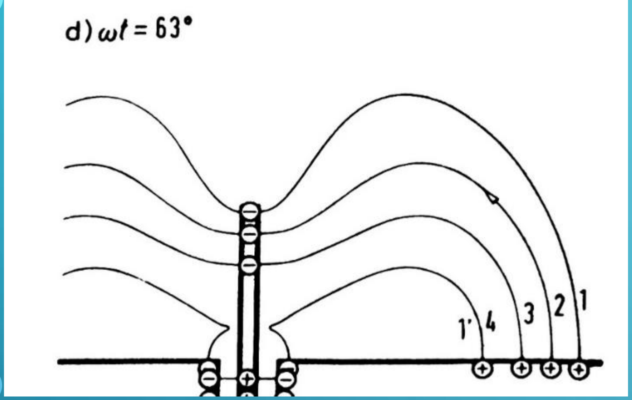


Fig. 7 (d): Radiation patterns as a function of different ωt (ωt=63°)

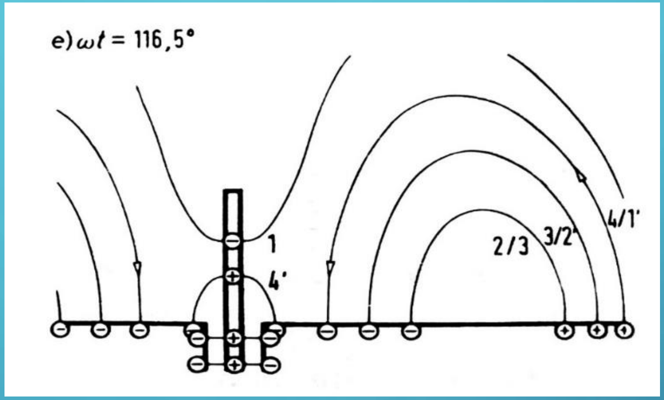


Fig. 7(e): Radiation patterns as a function of different ω t (ωt=116.5°)

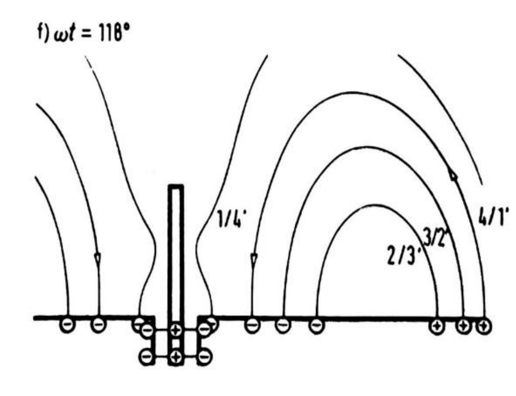


Fig. 7(f): Radiation patterns as a function of different ωt (ωt=118°)

EXAMPLE

- The example shown in Fig. 7 (a)-(f), was a vertical rod antenna, as we all know. There are circumstances where the antenna is not straight but curved. An early example is a trailing antenna from an airplane [2]
- As an example we will use the curved shape antenna shown in Fig. 8. It is interesting to know that the end effect of the antenna, as described by the Hallén's integral equation is equivalent to the end effect radiation in microstrips. This type of antenna is called a thin curvilinear monopole.
- This assumes that the ratio (l/d) is smaller than 0.01.
- The reason why the curved antenna is of interest is the military Manpack with the tilted antenna is shown in Fig. 8. If the antenna could be optimized in such a way that it shows gain, the effect of the building on the left side would be compensated for.

EXAMPLE

Fig. 8: Military Manpack with a straight whip antenna

- An antenna can also be described as an open-ended transmission line where the open end has the highest voltage field and the other side has the highest current.
- To be specific, this transmission line is lossy, so the hyperbolic tension function is applicable.
- This also means compared to a parallel tuned circuit, the antenna radiates at odd multiples (like 3, 5, 7…) of the fundamental frequency.
- The next relevant question is the input impedance

INPUT IMPEDANCE CALCULATION

- This input impedance can be determined from the following equations [4].
- Input impedance= $Z_{in} = (Z_a^2)/R_s$ (where R_s is the radiation resistance).
- $Z_a = \sqrt{L_a/C_a}$ *where 'l' is the antenna length in meters and,*
- *'d' is the wire diameter in meters*
- As an example for 4MHz, a 40m long dipole with 2mm thickness would be

$$L_a = \left(\frac{l\mu_0}{2\pi} \cdot ln\frac{4r}{D}\right) = 10\text{x}10^{-7}$$

$$C = \frac{20pF}{m}, \qquad so\ C_a = 20x20 = 400pF$$

$$Z_a = \sqrt{10\text{x}10^{-7}/400x10^{-12}} = 50\Omega$$

[4] Hans Ludwig Hartnagel, Rudiger Quay, Ulrich L. Rohde, Matthias Rudolph, "Fundamentals of RF and Microwave Techniques and Technologies", Ch. 6.

CALCULATION CONT'D…

- Let us assume a quarter wave antenna, with a radiation resistance of R_s of 36.6 Ω,
- Input impedance= $Z_{in} = (50)^2/36.6 = 68$, Z_a is an approximation.
- Now the exact solution [9]

$$Z_{11} = 30\int_0^{j2\pi n} \frac{1 - e^{-w}}{w} dw$$

- The integral i is an exponential integral with imaginary argument.
- In our case $y = 2\pi n$. This integral can be expressed in terms of the sine and cosine integrals.
- The input-impedance is

$$Z_{11} = R_{11} + jX_{11} = 30[\mathrm{Cin}(2\pi n) + j\mathrm{Si}(2\pi n)]$$
$$Z_{11} = 30[0.577 + \ln(2\pi n) - \mathrm{Ci}(2\pi n) + j\mathrm{Si}(2\pi n)]$$

- For a thin linear antenna half wave length see [10]

[9] Kraus, John Daniel, "Antennas", second edition, McGraw Hill, 1988, ISBN 0-07-035422-7

[10] S. A. Schelkunoff, Applied Mathematics for Engineers and Scientists, Van Nostrand, New York, 1948.p377)

CALCULATION CONT'D…

- $Z_{11} = j30 \int_0^L \left(\frac{e^{-j\beta z}}{z} + \frac{e^{-j\beta(L-z)}}{L-z}\right) \sin \beta z dz$
- Applying de Moivre's theorem to $\sin \beta z$,

$$Z_{11} = -15 \int_0^L \left[\frac{e^{-j2\beta z} - 1}{z} - \frac{e^{-j\beta L}\left(e^{j2\beta z} - 1\right)}{L - z}\right] dz$$

- For $L = n\lambda/2$ where $n = 1, 3, 5, \dots, e^{-j\beta L} = e^{-j\pi n} = -1$, so that (38) becomes

$$Z_{11} = -15 \int_0^L \left(\frac{e^{-j2\beta z} - 1}{z} + \frac{e^{j2\beta z} - 1}{L - z}\right) dz$$

$$Z_{11} = 15 \int_0^L \frac{1 - e^{-j2\beta z}}{z} dz + 15 \int_0^L \frac{1 - e^{j2\beta z}}{L - z} dz$$

$$- 15 \int_{2\pi n}^0 \frac{1 - e^{j(2\pi n - v)}}{v} dv$$

$$= 15 \int_0^{2\pi n} \frac{1 - e^{-jv}}{v} dv$$

- Equations (41) and (42) are definite integrals of identical form. Since their limits are the same, they are equal. Therefore

$$Z_{11} = 30 \int_0^{2\pi n} \frac{1 - e^{-ju}}{u} du$$

- If we now put $w = ju$, Z11 transforms to $Z_{11} = 30 \int_0^{j2\pi n} \frac{1 - e^{-w}}{w} dw$
- The integral is an exponential integral with imaginary argument and can be expressed in terms of the sine and cosine integrals

CALCULATION CONT'D…

- Hence, the input-impedance is

$$Z_{11} = R_{11} + jX_{11} = 30[Cin(2\pi n) + jSi(2\pi n)]$$
$$Z_{11} = 30[0.577 + \ln(2\pi n) - \mathrm{Ci}(2\pi n) + j\mathrm{Si}(2\pi n)]$$

- The more general situation, where the antenna length f is not restricted at an odd number of $\lambda/2$, has also been treated. The antenna is center-fed, and the current distribution is assumed to be sinusoidal The input impedance for this case is

$$R_{11} = 30\left[\left(1 - cot^2\frac{\beta L}{2}\right)\mathrm{Cin}\,2\beta L + 4cot^2\frac{\beta L}{2}\mathrm{Cin}\,\beta L + 2\cot\frac{\beta L}{2}\left(si\,2\beta L - 2s\beta L\right)\right]$$

- The above discussion of this section applies to balanced centered antennas [5].
- For a thin linear stub antenna of height l perpendicular to an infinite, perfectly conducting ground, the self-impedance is $\frac{1}{2}$ that for the corresponding balanced type.
- The general formula for input resistance can be converted for a stub antenna above a ground plane by changing the factor 30 to 15 and making the substitution $L = 21$.
- The formulas can be converted for a stub antenna with ground plane where the antenna is an odd number n of $\lambda/4$ long by changing the factor 30 to 15. In this case we get Z11=36.6 + 21 Ohm

[5] 'G. II. Brown and R. King "High Frequency Models is Antenna Investigations," Proc. IRE, 22, 457 – 480. April 1934.

DERIVATION

- Assuming this antenna which should resonate at about 4MHz, we can determine the combination of radiation losses and other losses. [6]
- A FoM (figure of Merit) is the standing wave ratio of 2.6, and the actual 3dB bandwidth will be about 200 kHz. This means that the operating Q of the antenna is 20. These are not common losses, but are radiation losses.
- The magnitude of the reflection coefficient rho is defined as:

$$\rho(f) := \left(\left| \frac{Z(f) - R}{Z(f) + R} \right| \right)$$

- If Zin = (50+j50), half power (-3dB) or (50-j50), other side of the 3dB point,
- Then the flection coefficient is at the bandwidth edges are:

$$\rho\mathrm{L} := \rho(Flow) \qquad \rho\mathrm{L} = 0.447$$
$$\rho\mathrm{H} := \rho(Fhigh\,) \qquad \rho\mathrm{H} = 0.447$$

- Now find the VSWR versus frequency: $\mathrm{VSWR(f)} := \frac{1+\rho(\mathrm{f})}{1-\rho(\mathrm{f})}$ and VSWR at the bandwidth edges:

$$\mathrm{VSW\,R}(Flow) = 2.618$$
$$\mathrm{VSWR}(Fhigh) = 2.618$$

[6]Kazimierz “Kai” Siwiak, (KE4PT), “Q and the Energy Around Antennas”, QST Feb 2013, pp-37

EXAMPLE

- Let us assume a quarter wave antenna, with a radiation resistance of Rs of 36.6 Ω
- Input impedance= Z_{in} = $(50)^2/36.6$ = 68, Z_a is an approximation
- Assuming this antenna which should resonate at about 4MHz, we can determine the combination of radiation losses and other losses.
- A FoM (figure of Merit) is the standing wave ratio of 2, and the actual 3dB bandwidth will be about 200kHz. This means that the operating Q of the antenna is 20. These are not common losses, but are radiation losses.
- If the bandwidth is much smaller, then there are some inductors or capacitors or a combination of (traps), inductors or capacitors, which make the antenna electrically longer at the expense of bandwidth.
- Compared to the tube amplifiers modern transistorized amplifiers cannot handle a larger the 1.5 VSWR and reduce drastically the output power.
- While the tube amplifiers had a Collins or Low Pass filter at the output and could tune out mismatch, the solid state amplifier needs held and resorts to an additional "line flattening circuit that most of the time can handle an VSWR of 3 and protects the output stage

CURVED ANTENNA

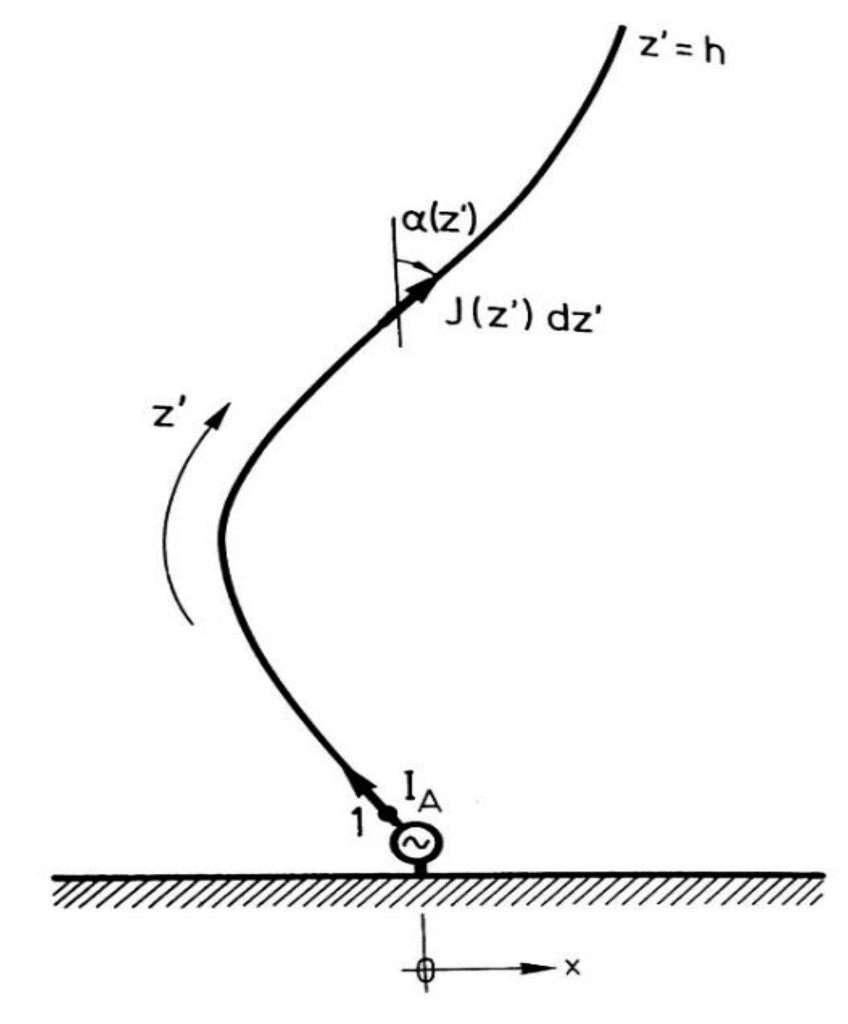


Fig. 9 (a): Curved antenna

- Landstorfer [3] notes that curvilinear antennas of general shape suffer from the fact that experimental investigations as well as numerical analysis are restricted to a limited number of antennas with arbitrary shapes. It is not possible to show all the possibilities of curvilinear wire antennas as a whole. The first mention of a quasi curvilinear antenna was found in the ref [2]. Here an antenna was dragged by an airplane. Let's assume that it had the curve shown in Fig 9(a).
- This curved antenna can be split in different sections (from 1, 2, 3, 4…. to n).

[3] F. M. Landstorfer and R. R. Sacher, "Optimisation of Wire Antennas", Research Studies Press Ltd., 1985, ISBN 0 863800254

CURVED ANTENNA SPLIT IN NUMEROUS SECTIONS FOR ANALYSIS

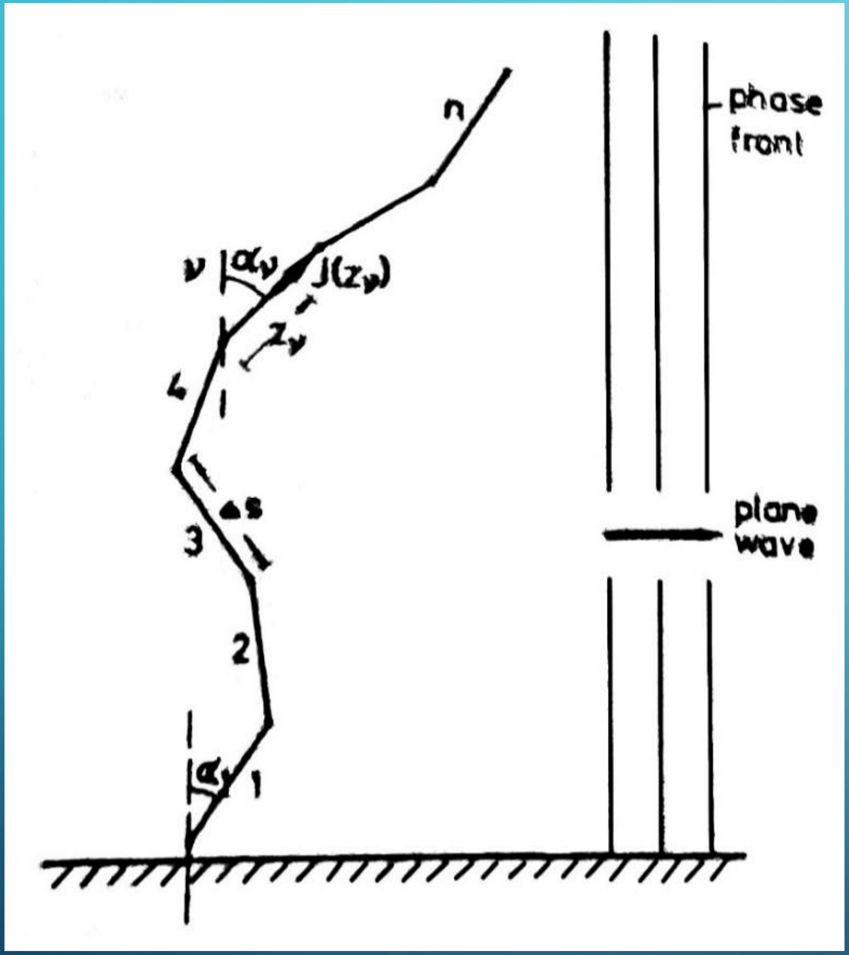


Fig. 9(b): Curved antenna split in numerous sections for analysis

- The following is a method to breakdown the antenna into small, straight, pieces.
- The current distribution along the antenna is sinusoidal and will not change even if the antenna shape changes, as long as the arc length 2h of the antenna remains constant.

CURVILINEAR MONOPOLE

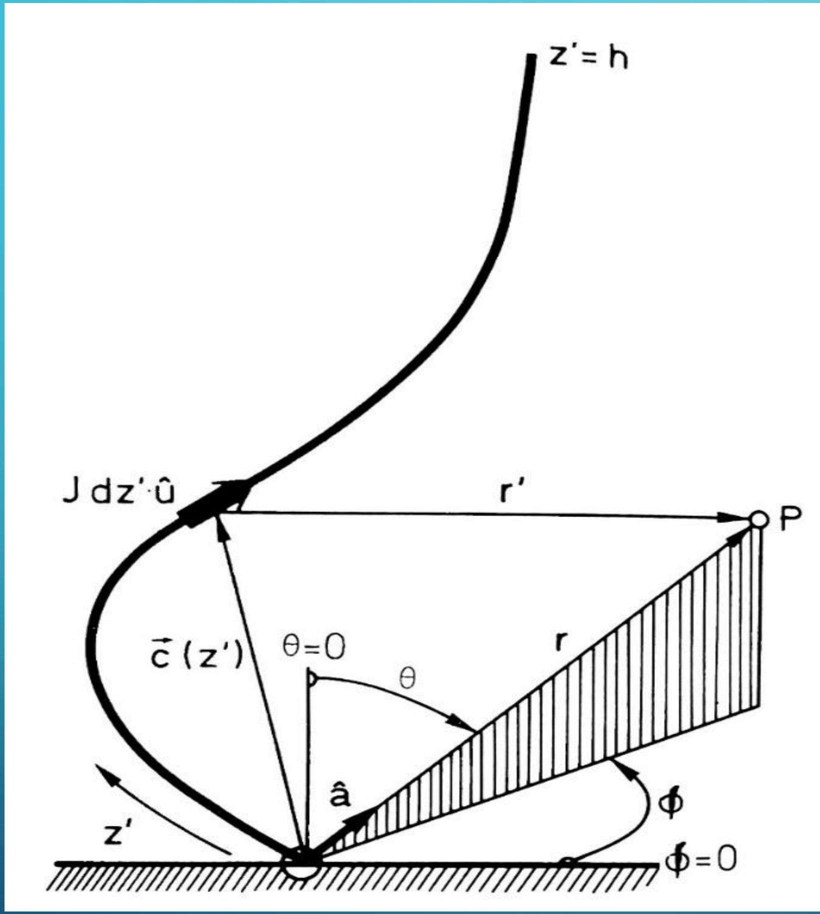


Fig. 10: 3-dimensional curvilinear monopole: designations

- The shape of the antenna of Fig 10.3 can be approximated by n straight-line sections of constant length Δs.

CURVILINEAR ANTENNA

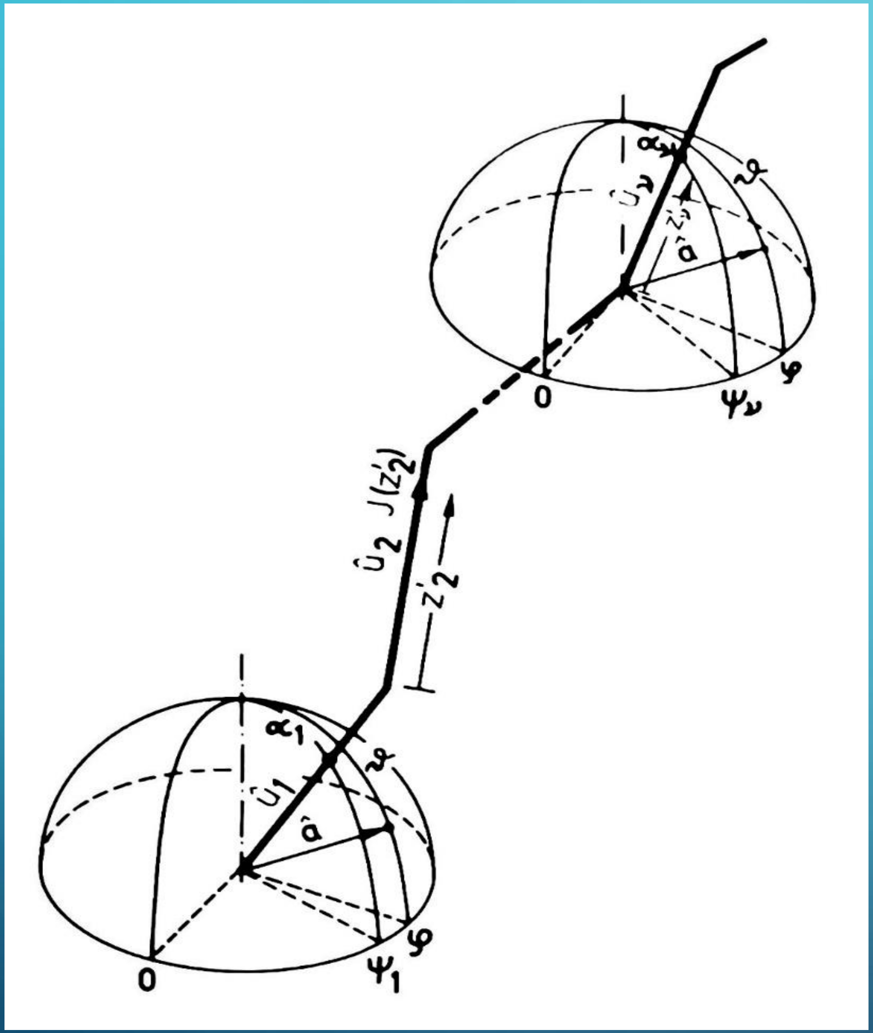
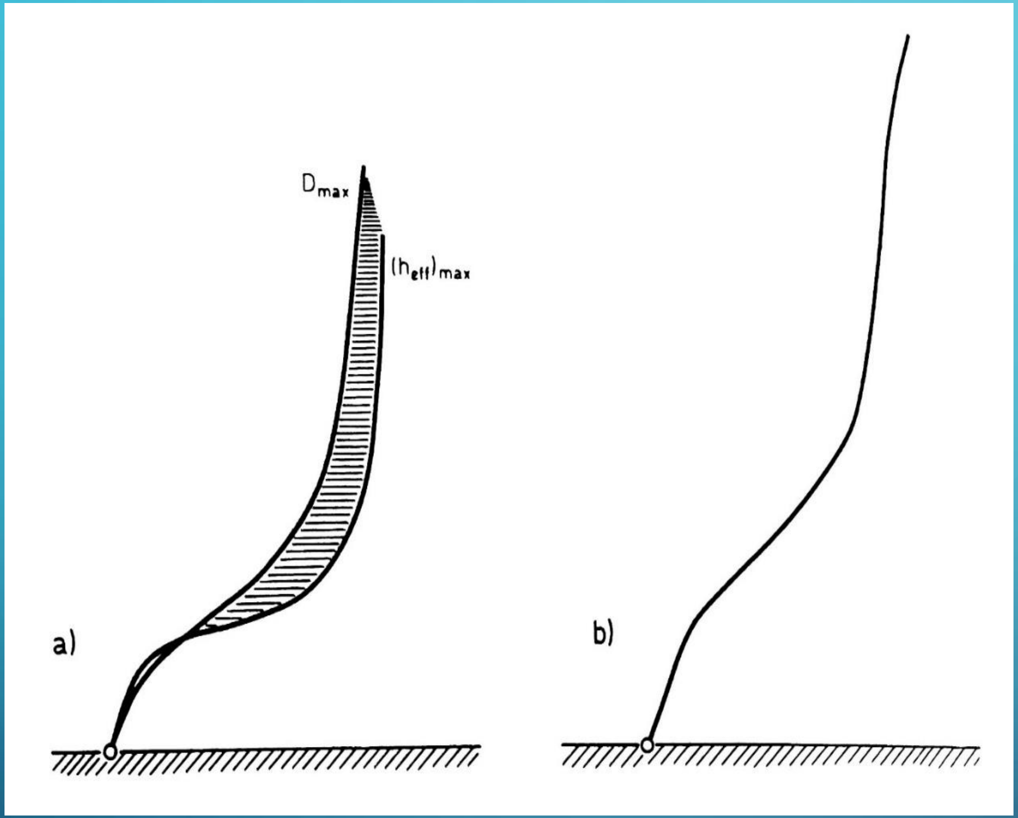


Fig. 11: 3-dimensional curvilinear monopole approximated by *n* straight line sections: designations

- As shown in Fig. 11, the antenna shape is given by the 2n angles $\alpha_1 .... \alpha_n$ and $\psi_1 ... \psi_n$ or the spatial directions of the different antenna sections as described by the unit vector $\hat{U}_1 ... \hat{U}_n$
- For a single-wire antenna, maximum directivity is obtained with a configuration completely restricted to the E-plane.

DIRECTIVITY



- **Fig. 12 shows that the profile for maximum directivity differs from that for maximum effective height in that it has a smaller tilt and is superior in directivity by about 0.6dB.**

Fig. 12: (a) Continuous profile for maximum directivity and maximum effective length (b) λ_0 - monopole optimized for maximum directivity

PIECEWISE LINEAR 3 SEGMENT MONOPOLE

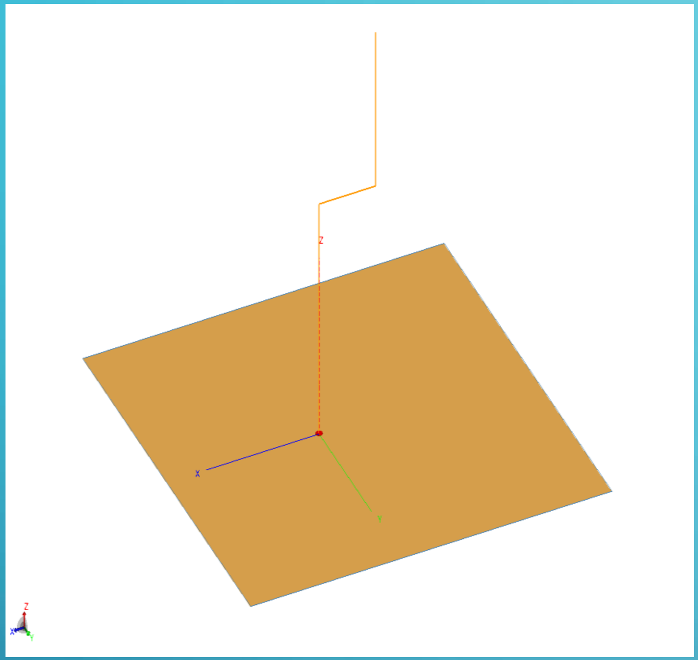


Fig. 13: Piecewise linear 3-segment monopole

- One can show that a simple stick model of the monopole can also be optimized for maximum gain with 3 piecewise linear sections as shown in Figure 13.
- That monopole is over a perfectly electrically conducting (PEC) ground plane.
- The optimization varied the lengths of each section to achieve the goal of the maximum gain along the ground plane (theta = 90 degrees) as well as to achieve a second goal of a VSWR less than 1.1 into 50 ohms.
- The optimized sections were as follows: Length of the fed section attached to the ground plane = .617 wavelengths, length of the horizontal section = .236 wavelengths, length of the top vertical section = .412 wavelengths, and the radius of the wires = .001 wavelengths.

ELEVATION PATTERN OF OPTIMIZED PIECEWISE LINEAR 3 SECTION MONOPOLE

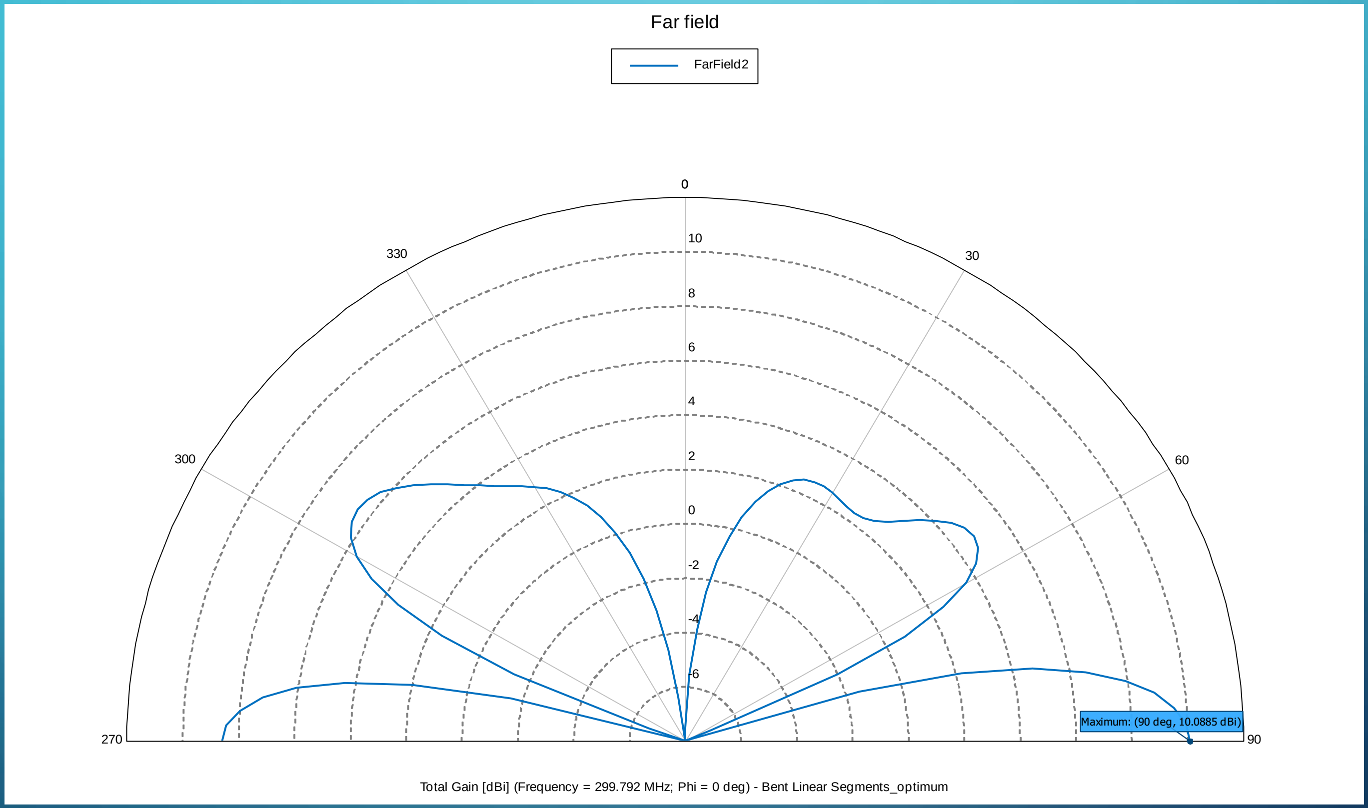


Fig. 14: Elevation pattern of optimized piecewise linear 3 section monopole

SIMULATION MODEL

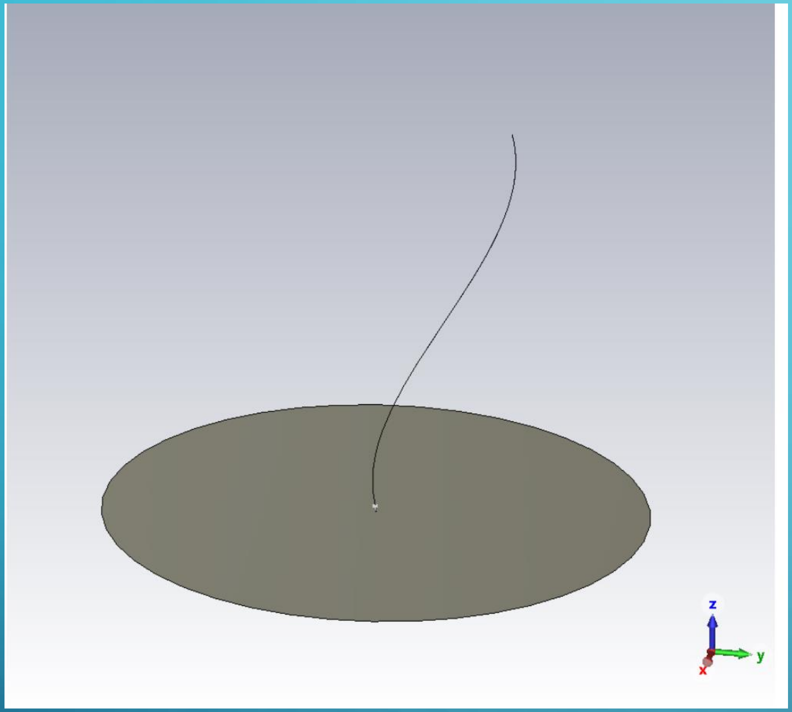


Fig. 15: ¾ λ antenna curvilinear antenna model for simulation

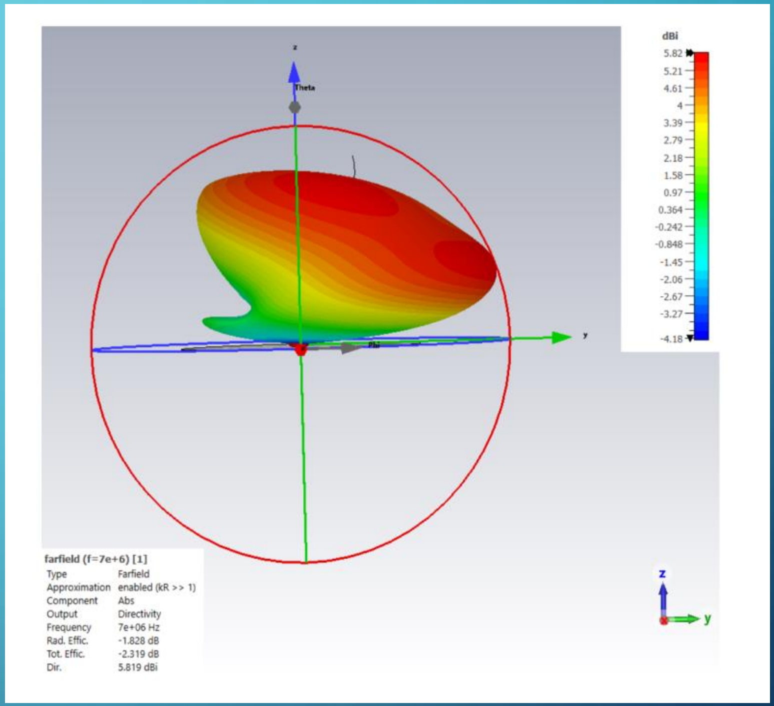
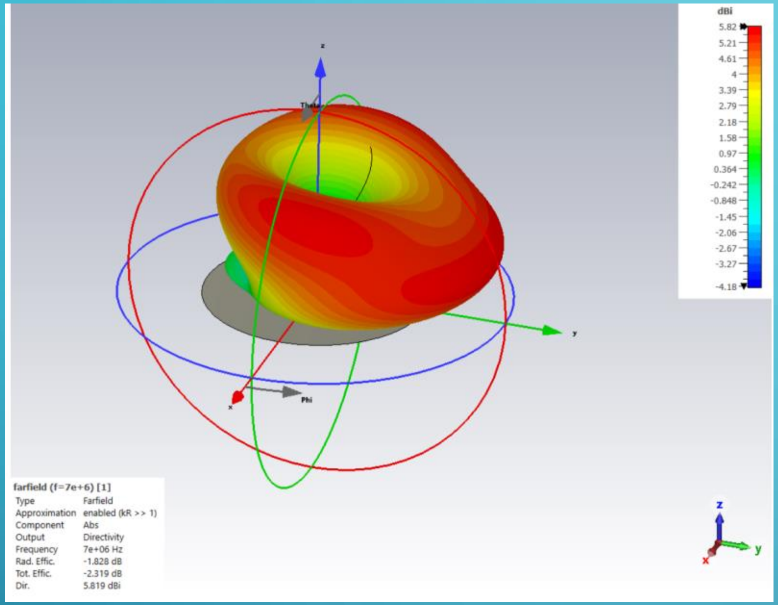


Fig. 16: Radiation Pattern Simulation

SIMULATION MODEL



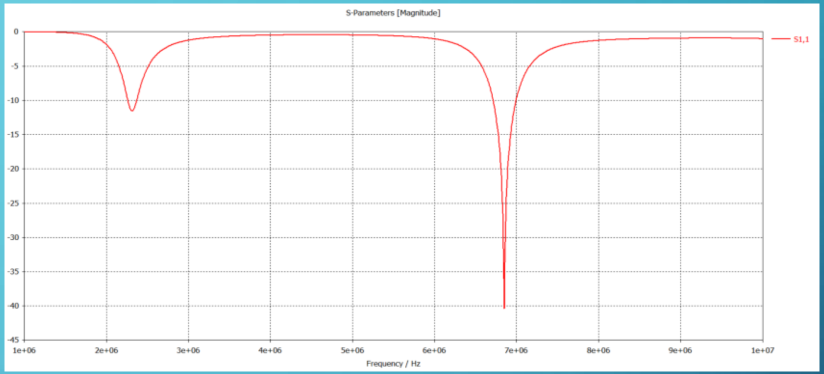


Fig. 18: Impedance Matching

Fig. 17: Simulation

Figures 15 to 19 shows modern modeling of the curved monopole and the results that can be obtained for further optimization and impedance matching.

SIMULATION

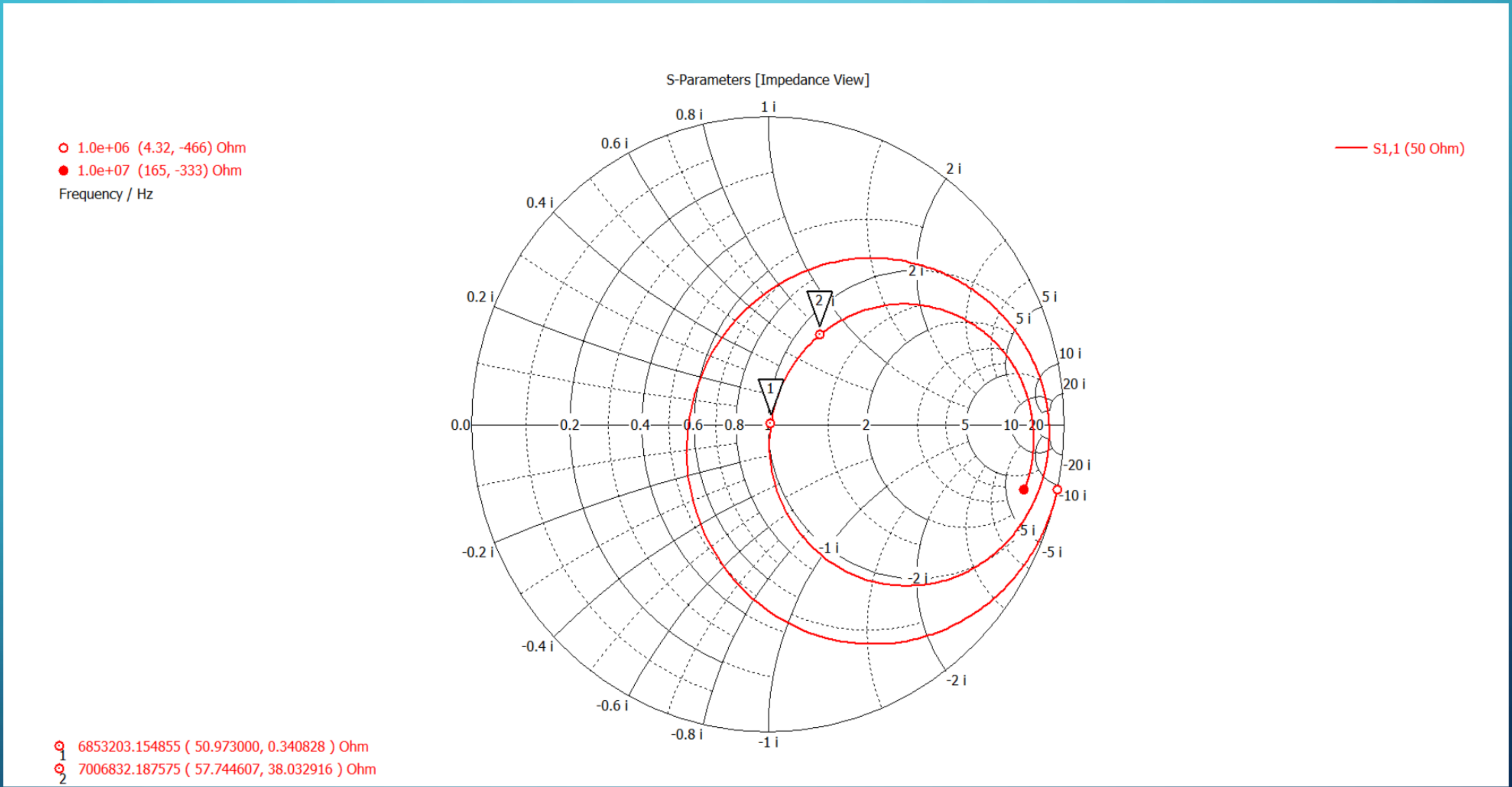


Fig. 19: Simulated Antenna Input Impedance

FEKO - OPTIMIZATION

- Furthermore, optimization in FEKO has been applied to the same 7 MHz curvilinear antenna with a height of 3/4 wavelengths over earth with a relative permittivity of 13 and a conductivity of 0.005 S/m (average ground conditions) ref [7-8].
- A spline representation of the curved wire is input to the antenna software with points at each 0.15 wavelengths of height.
- At each point, the width dimension can be stretched plus or minus 0.5 wavelengths maximum. These are the constraints of the Genetic Optimization. There are 5 such points to give a maximum height of the antenna of 3/4 wavelengths.
- Curvilinear segments form the wire for the Method of Moments formulation. The antenna is fed against a conduction ground plane 0.5 wavelengths in diameter above the earth ground.
- An optimization was run with maximizing the gain at a 20 degree takeoff angle above the ground while also minimizing the peak back lobe in the rear of the pattern from 0 to 90 degrees in elevation and 90 to 270 degrees in azimuth. This will find the peak back lobe in this 3D region of space and minimize it.
- This will give the global best front to back ratio that can be achieved for this specified region of space.

[7] FEKO, Altair Engineering Inc., 2022
[8] Siwiak K., Rohde U.L., " Tuning electrically short antennas for field operation", May 2019

OPTIMIZED CURVED WIRE SHAPE

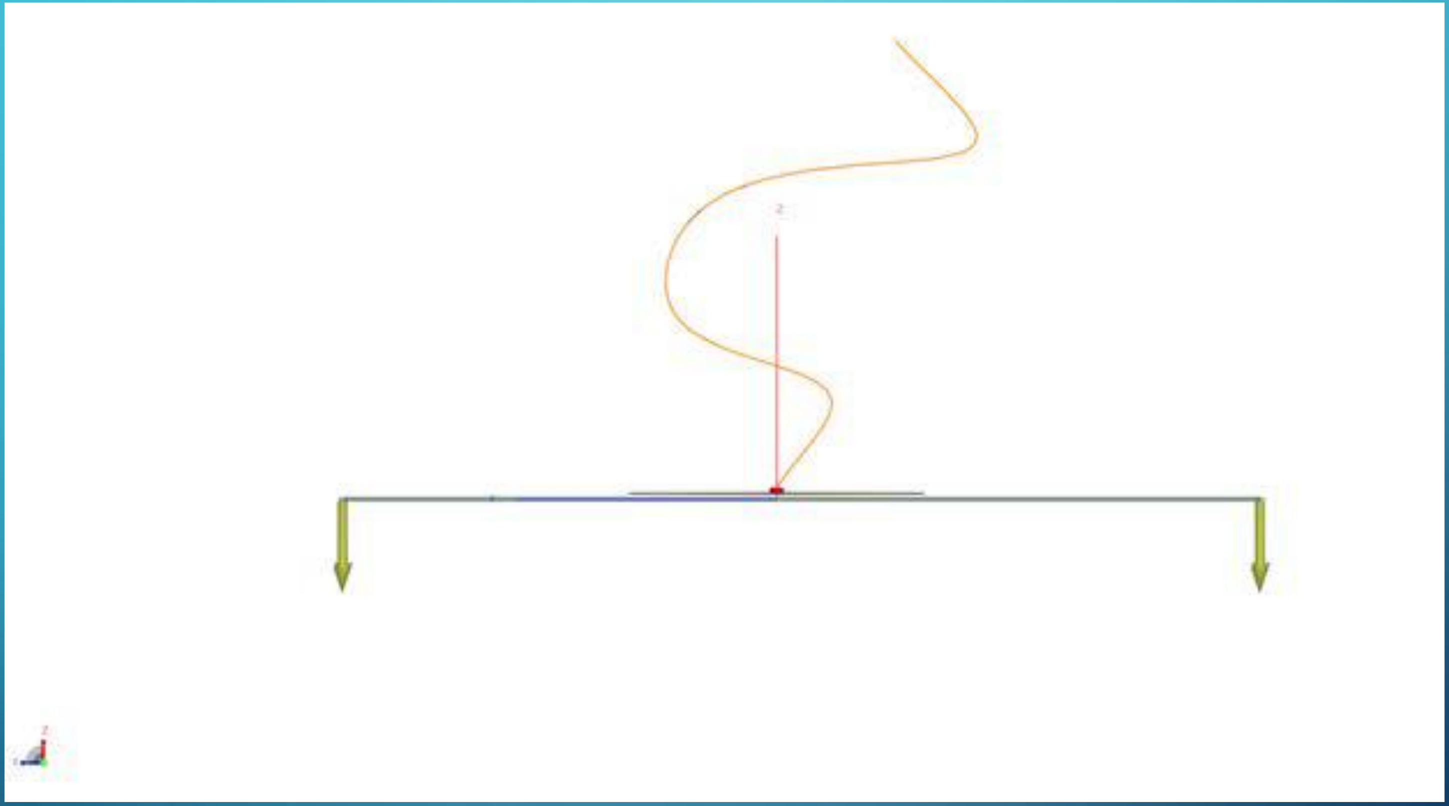


Fig. 20: Optimized curved wire shape

CURRENT DISTRIBUTION

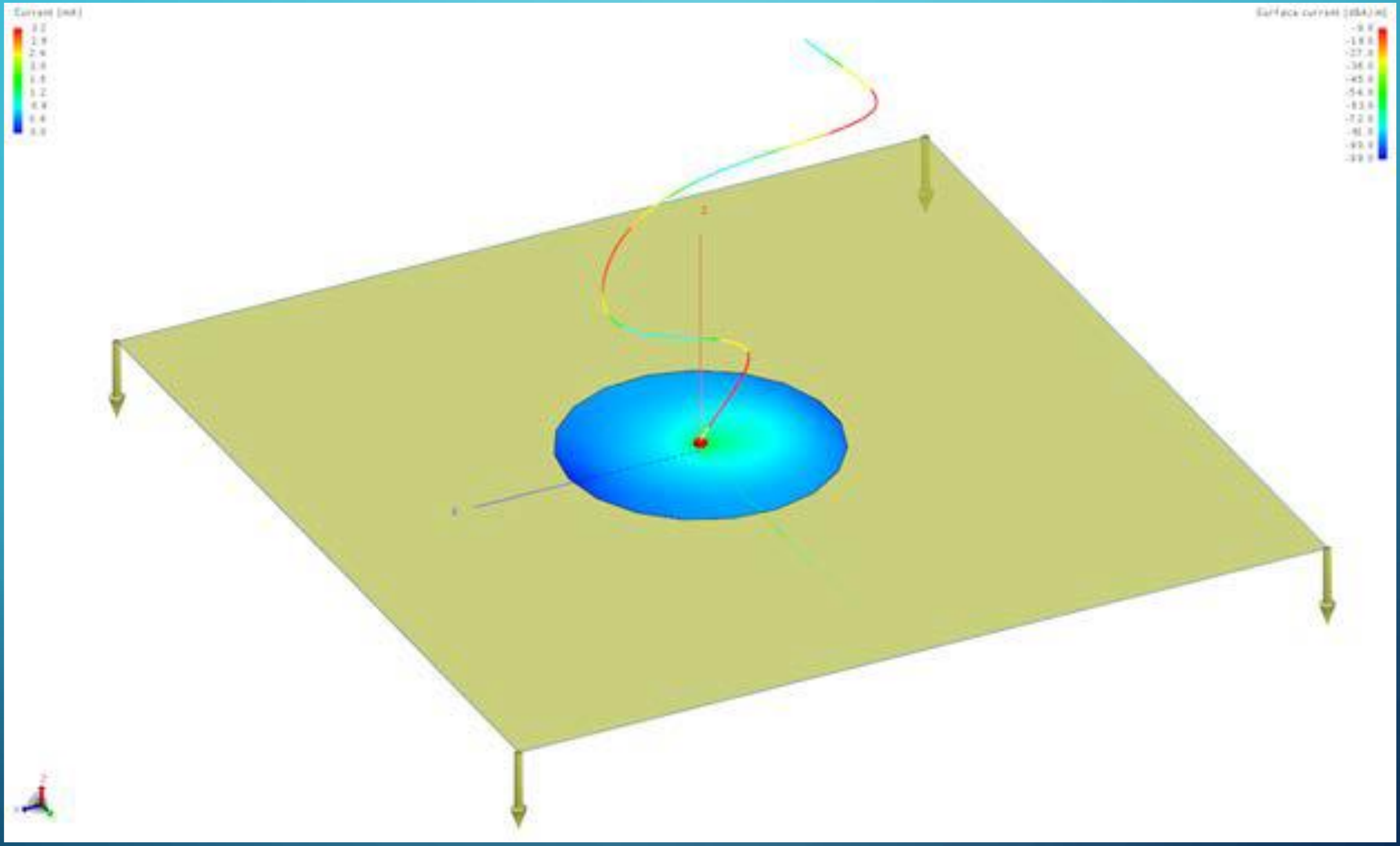


Fig. 21: Current Distribution

3D - PATTERN

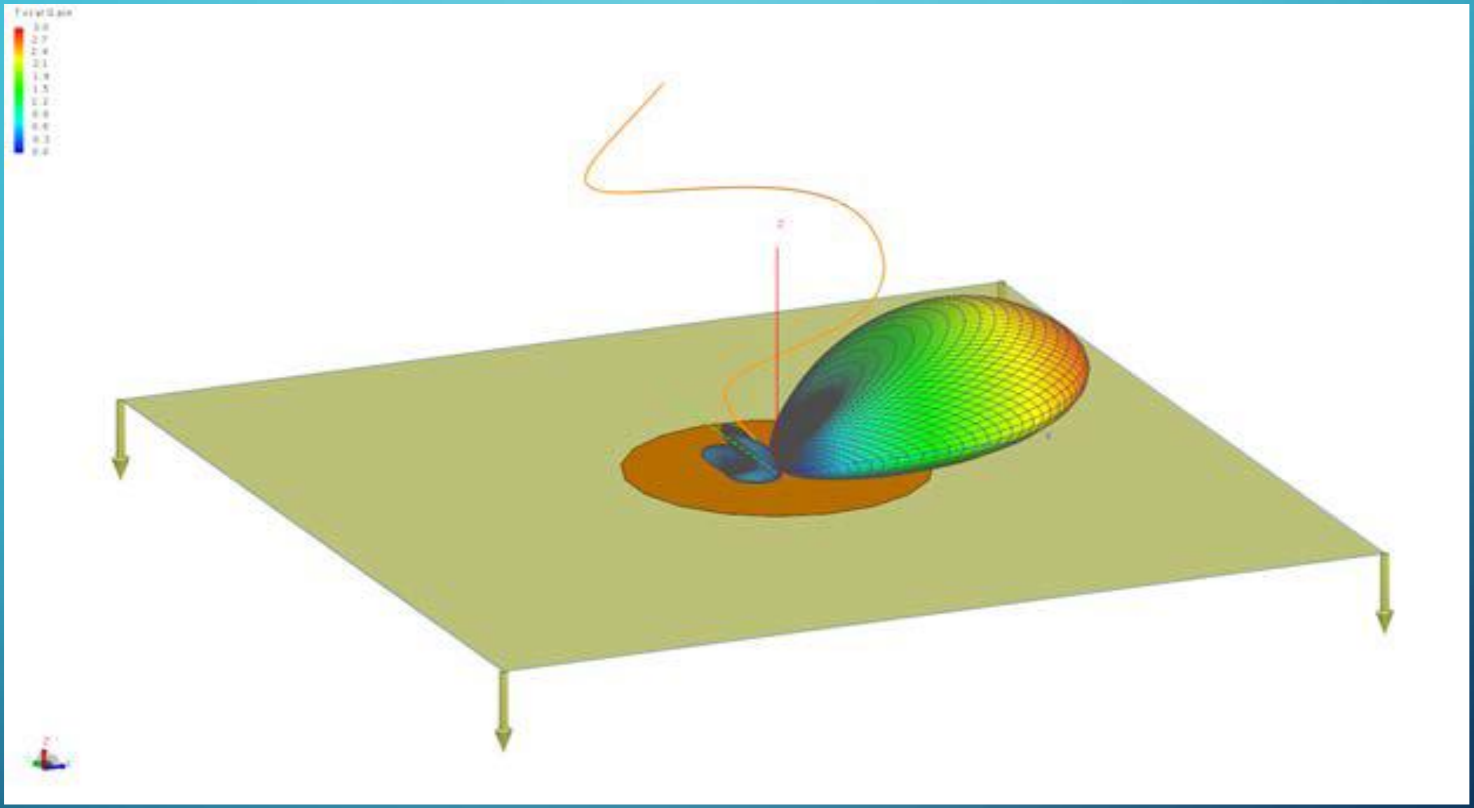


Figure 22: 3D Pattern

AZIMUTH PATTERN AT A TAKEOFF ANGLE OF 20 DEGREES

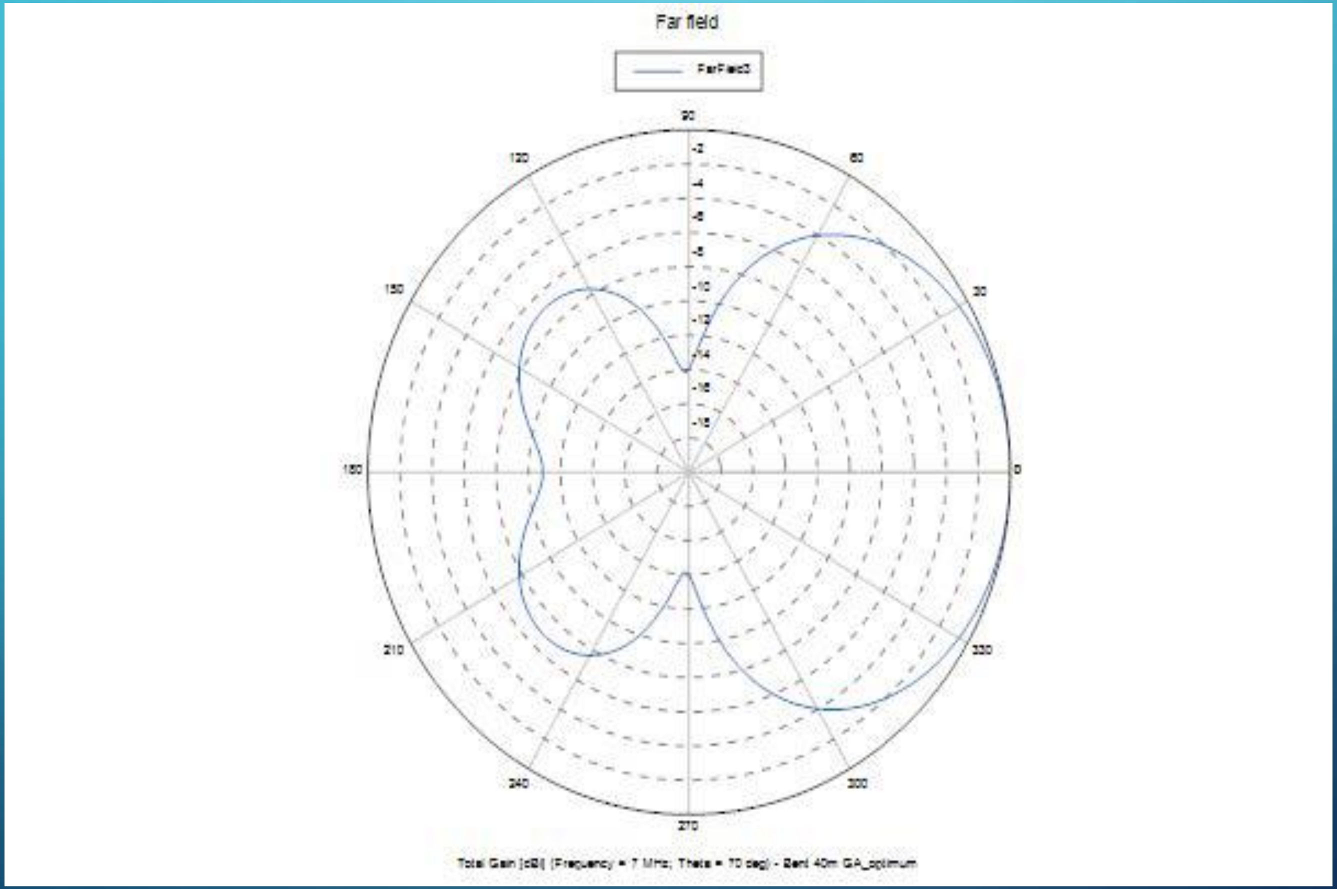


Figure 23: Azimuth Pattern at a takeoff angle of 20 degrees

ELEVATION PATTERN

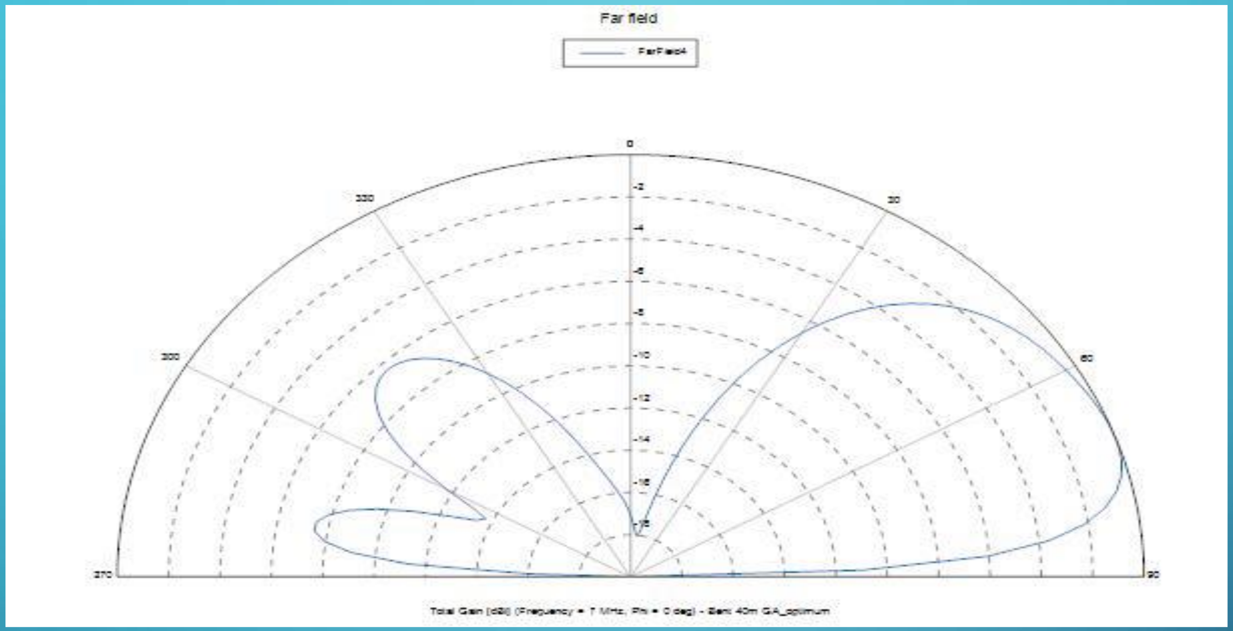


Figure 24: Elevation Pattern

Figures 20 to 24 show all the results from the global Genetic Optimization of the curved antenna where the optimized shape has been determined for this height of 3/4 wavelengths above the ground. The gain was 4.3 dBi which is about 4.3 dB of gain over a standard 1/4 wavelength monopole which has a gain of 0 dBi over earth ground with a good 0.5 wavelength diameter ground plane.

APPLICATIONS

R&S ® MR3000U manpack radio, covering 25 to 512 MHz

VHF/UHF (left) and HF/VHF (right) manpacks. The UHF/VHF manpack shown uses a thick vertical pole antenna, and the HF manpack uses a 5 m dipole (the yellow wire).

CONCLUSIONS

- This presentation has shown the historical methods of computing antenna currents on wires of any arbitrary shape.
- Furthermore, the modern Method of Moments technique has been used with the Genetic Optimization method to obtain the optimum curved shaped for the wire monopole type of antenna to achieve the best front to back ratio for a fixed height of 3/4 wavelengths.
- This is a very simple method to achieve such performance of just one simple wire compared to more complex arrays of multiple antennas necessary to achieve similar performance.

REFERENCES

1. Schaefer, CL.: Einführung in die theoretische Physik. (Introduction in to the theoretical Physics), Bd. 3/1 S. 331, Berlin: W.de Gruyter 1932.
2. Burstyn W.: Die Strahlung und Richtwirkung einiger Luftdrahtformen im freien Raum (Radiation and Direction of Different styles of Open Wire Antennas), Jb. drahtl. Telegr. Bd. 13 (1918) S. 362].
3. F. M. Landstorfer and R. R. Sacher, "Optimisation of Wire Antennas", Research Studies Press Ltd., 1985, ISBN 0 863800254
4. Hans Ludwig Hartnagel, Rudiger Quay, Ulrich L. Rohde, Matthias Rudolph, "Fundamentals of RF and Microwave Techniques and Technologies", Ch. 6.
5. 'G. II. Brown and R. King "High Frequency Models is Antenna Investigations," Proc. IRE, 22, $457-480$. April 1934.
6. Kazimierz "Kai" Siwiak, (KE4PT), "Q and the Energy Around Antennas", QST Feb 2013, pp-37
7. FEKO, Altair Engineering Inc., 2022
8. Siwiak K., Rohde U.L., "Tuning electrically short antennas for field operation", Microwave Journal, May 2019,
9. Kraus, John Daniel, "Antennas", second edition, McGraw Hill, 1988, ISBN 0-07-035422-7
10. S. A. Schelkunoff, Applied Mathematics for Engineers and Scientists, Van Nostrand, New York, 1948.p377)

FURTHER READING

- U.L.Rohde, 'Die Anpassung von kurzen Stabantennen für KW-Sender', Funkschau 1974, heft 7.
- Match antenna over 1.5to 30MHz range with only two adjustable elements, electronic design 19, Sept 13 1975
- William E Sabin and Edgar O. Schoenike, 'Single sideband systems and circuits', McGraw hill second edition page 582-605.
- RCA, RF power transistor manual, chapter Design techniques for RF power amplifiers, page 49-63.
- The ARRL handbook for radio communications 2013 chapter 22 Antennas and chapter 18 page 18.6
- http://en.wikipedia.org/wiki/Antenna
- http://en.wikipedia.org/wiki/Antenna_tuner
- http://en.wikipedia.org/wiki/Active_antenna
- http://en.wikipedia.org/wiki/File:Spark_gap_transmitter_diagram.png.
- Christof Rohner, Rohde & Schwarz – Antenna Basics

FURTHER READING

- Ulrich. L. Rohde, "Active Antennas", RF Design, May/June 1981
- "Active Receiving Antennas", www.electronics-tutorials.com
- Ulrich. L. Rohde, "Receiver Measurements, How to evaluate Receivers", QEX, July/Aug 2005
- R. A. Sainati, D. E. Fessenden, "Performance of an Electrically Small Antenna Amplifier Circuit
- William E. Sabin, Edgar O. Schoenike, "Single sideband systems and Circuits", Ch-15, McGraw Hill Inc.
- Ashraf Ramadan, "Active Antennas with High Input Impedance Low Noise and Highly Linear Amplifiers", Ph.D Dissertation, July 2005
- Ulrich. L. Rohde, "Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Part 1, QEX, Nov/Dec 2002
- Ulrich. L. Rohde, "Theory of Intermodulation and Reciprocal Mixing: Practice, Definitions and Measurements in Devices and Systems, Part 2, QEX, Jan/Feb 2003
- Ulrich. L. Rohde, "From Spark Generators to Modern VHF/UHF/SHF Voltage Controlled Oscillators", QEX, July/Aug 2008
- Altshuler, E. E. and Linden, D. S., "Wire-Antenna Design Using Genetic Algorithms," IEEE Antennas and Propagation Magazine, 39, 2, April 1997, p. 33-43
- Rabelo. R. and Terada, M., "Analysis and Optimization of Wire Antennas over the Internet," IEEE Antennas and Propagation Magazine, Vol. 521, No. 1, February2010

QUESTIONS ????