ANTENTOP

ANTENTOP 01 2009 # 011

ANTENTOP is *FREE* e-magazine devoted to ANTENna's
Theory,
Operation, and
Practice
Edited by hams for hams

1-2009

In the Issue:
Antennas Theory!

Practical design of HF-VHF- UHF Antennas!

Home brew Technique!

Propagation!

QRP!

And More….

RU3ARJ Compact Cage Antenna for 435 MHz

DEWD – Famous invention of UR0GT

EDITORIAL:

Well, my friends, new ANTENTOP – 01 -2009 come in! ANTENTOP is just authors' opinions in the world of amateur radio. I do not correct and re-edit yours articles, the articles are printed "as are". A little note, I am not a native English, so, of course, there are some sentence and grammatical mistakes there… Please, be indulgent!

ANTENTOP 01 –2009 contains antenna articles, some describtion of antenna patents, QRP- Stuff. Hope it will be interesting for you.

Our pages are opened for all amateurs, so, you are welcome always, both as a reader as a writer.

73! Igor Grigorov, VA3ZNW

ex: RK3ZK, UA3-117-386, UA3ZNW, UA3ZNW/UA1N, UZ3ZK
op: UK3ZAM, UK5LAP, EN1NWB, EN5QRP, EN100GM

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Welcome to ANTENTOP, FREE e - magazine!

ANTENTOP is ***FREE e- magazine***, made in **PDF**, devoted to antennas and amateur radio. Everyone may share his experience with others hams on the pages. Your opinions and articles are published without any changes, as I know, every your word has the mean.

Every issue of ANTENTOP is going to have 100 pages and this one will be paste in whole on the site. Preview's files will be removed in this case. I do not know what a term for one issue will need, may be 8-10 month or so. A whole issue of ANTENTOP hold nearly 10 MB.

A little note, I am not a native English, so, of course, there are some sentence and grammatical mistakes there… Please, be indulgent!

Preview: Some articles from "cooking" issue will be pasted for preview on this site, others no. Because, as I think, it must be something mysterious in every issue.

Publishing: If you have something for share with your friends, and if you want to do it ***FREE***, just send me an email. Also, if you want to offer for publishing any stuff from your website, you are welcome!

Your opinion is important for me, so, contact if you want to say something!

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I have a big collection of pictures, I have got the pictures and stuff in others ways, from ***FREE websites***, from commercial CDs, intended for ***FREE using***, and so on... I use to the pictures (and seldom, some stuff from closed websites) in ANTENTOP. ***If the owners still are alive***, please, contact with me, I immediately remove any Copyright stuff, or, if it is necessary, all needed references will be made there.

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73! ***Igor Grigorov***, VA3ZNW

ex: UA3-117-386, UA3ZNW, UA3ZNW/UA1N, UZ3ZK, RK3ZK

op: UK3ZAM, UK5LAP,EN1NWB, EN5QRP, EN100GM

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Feel Yourself a Student!

Dear friends, I would like to give to you an interesting and reliable antenna theory. Hours searching in the web gave me lots theoretical information about antennas. Really, at first I did not know what information to chose for ANTENTOP. Finally, I stopped on lectures “Modern Antennas in Wireless Telecommunications” written by Prof. Natalia K. Nikolova from McMaster University, Hamilton, Canada.

You ask me: Why?

Well, I have read many textbooks on Antennas, both, as in Russian as in English. So, I have the possibility to compare different textbook, and I think, that the lectures give knowledge in antenna field in great way. Here first lecture “Introduction into Antenna Study” is here. Next issues of ANTENTOP will contain some other lectures.

So, feel yourself a student! Go to Antenna Studies!

I.G.

My Friends, the above placed Intro was given at ANTENTOP- 01- 2003 to Antennas Lectures.

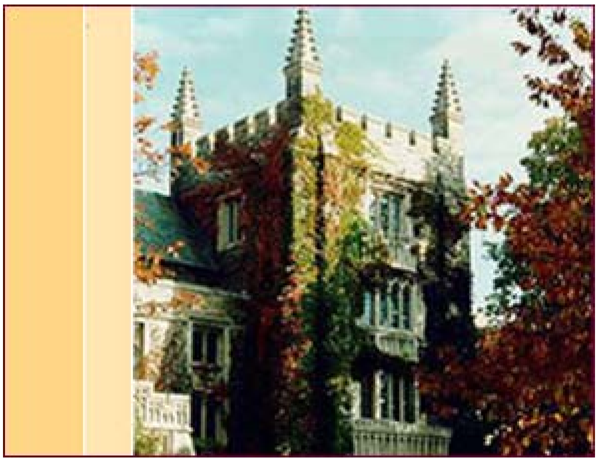
Now I know, that the Lecture is one of popular topics of ANTENTOP. Every Antenna Lecture was downloaded more than 1000 times!

Now I want to present to you one more very interesting Lecture - it is a Lecture Aperture Antennas – Part I. I believe, you cannot find such info anywhere for free! Very interesting and very useful info for every ham, for every radio- engineer.

So, feel yourself a student! Go to Antenna Studies!

I.G.

McMaster University Hall



Prof. Natalia K. Nikolova

Aperture Antennas – Part I

(*The uniqueness theorem. The equivalence principle. The application of the equivalence principle to aperture problem. The uniform rectangular aperture. The tapered rectangular aperture.*)

by Prof. Natalia K. Nikolova

LECTURE 12: Aperture Antennas – Part I

(*The uniqueness theorem. The equivalence principle. The application of the equivalence principle to aperture problem. The uniform rectangular aperture. The tapered rectangular aperture.*)

Introduction

Aperture antennas constitute a large class of antennas, which emit electromagnetic wave through an opening (or aperture). These antennas have close analogs in acoustics: the megaphone and the parabolic microphone. The pupil of the human eye is a typical aperture receiver for optical EM radiation. At radio and microwave frequencies, horns, waveguide apertures and reflectors are examples of aperture antennas. Aperture antennas are of common use at UHF and above. It is because aperture antennas have their gain increase as $\sim f^2$. For an aperture antenna to be efficient and have high directivity, it has to have an area comparable or larger than λ^2. Obviously, these antennas would be impractical at low frequencies. Another positive feature of the aperture antennas is their near-real valued input impedance and geometry compatibility with waveguide feeds.

To facilitate the analysis of these antennas, the equivalence principle is applied. This allows us to carry out the far-field analysis in the outer (unbounded) region only, which is external to the radiating aperture and the antenna. This requires the knowledge of the tangential field components at the aperture, as it follows from the equivalence principle.

1. Uniqueness theorem

A solution is said to be unique if it is the only one possible among a given class of solutions.

The EM field in a given region $V_{[S]}$ is uniquely defined if

- *all sources are given;*
- *<u>either</u> the tangential $\vec{E}_\tau$ components <u>or</u> the tangential $\vec{H}_\tau$ components are specified at the boundary S.*

The uniqueness theorem is proven by making use of the Poynting's theorem in integral form:

$$\oiint_S \left(\vec{E}\times\vec{H}^*\right)d\vec{s} + j\omega\iiint_{V_{[S]}}\left(\mu\,|\,\vec{H}\,|^2 - \varepsilon\,|\,\vec{E}\,|^2\right)dv + \iiint_{V_{[S]}}\sigma\,|\,\vec{E}\,|^2\,dv = -\iiint_{V_{[S]}}\left(\vec{E}\cdot\vec{J}^{i*} + \vec{H}^*\cdot\vec{M}^i\right)dv \tag{12.1}$$

Poynting's theorem states the conservation of energy law in EM systems.

One starts with the supposition that a given EM problem has two solutions (due to the same sources and the same boundary conditions): $\left(\vec{E}^a,\vec{H}^a\right)$ and $\left(\vec{E}^b,\vec{H}^b\right)$. The difference field is then formed:

$$\left|\begin{array}{l}\delta\vec{E} = \vec{E}^a - \vec{E}^b\\ \delta\vec{H} = \vec{H}^a - \vec{H}^b\end{array}\right. \tag{12.2}$$

Since the difference field has no sources, it will satisfy the source-free form of (12.1):

$$\oiint_S \left(\delta\vec{E}\times\delta\vec{H}^*\right)d\vec{s} + j\omega\iiint_{V_{[S]}}\left(\mu\,|\,\delta\vec{H}\,|^2 - \varepsilon\,|\,\delta\vec{E}\,|^2\right)dv + \iiint_{V_{[S]}}\sigma\,|\,\delta\vec{E}\,|^2\,dv = 0 \tag{12.3}$$

Since both fields satisfy the same boundary conditions at *S*, then $\delta\vec{E} = 0$ and $\delta\vec{H} = 0$ over *S*. This leaves us with

$$j\omega\iiint_{V_{[S]}}\left(\mu\,|\,\delta\vec{H}\,|^2 - \varepsilon\,|\,\delta\vec{E}\,|^2\right)dv + \iiint_{V_{[S]}}\sigma\,|\,\delta\vec{E}\,|^2\,dv = 0, \tag{12.4}$$

which is true only if

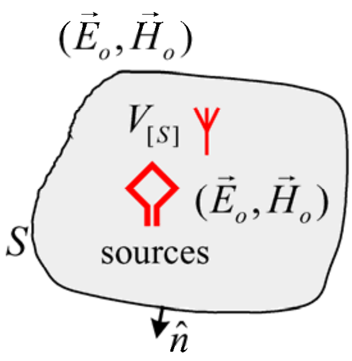
$$\left|\begin{array}{l}\omega\displaystyle\iiint_{V_{[S]}}\left(\mu\,|\,\delta\vec{H}\,|^2 - \varepsilon\,|\,\delta\vec{E}\,|^2\right)dv = 0\\ \displaystyle\iiint_{V_{[S]}}\sigma\,|\,\delta\vec{E}\,|^2\,dv = 0\end{array}\right. \tag{12.5}$$

If we assume some dissipation, however slight, equations (12.5) are satisfied only if $\delta\vec{E} = \delta\vec{H} = 0$ everywhere in the volume $V_{[S]}$. This implies the uniqueness of the solution. If $\sigma = 0$, which is a physical impossibility, but is often used approximation, multiple solutions $(\delta\vec{E},\delta\vec{H})$ may exist in the form of self-resonant modes of the structure under consideration. In open problems, resonance is impossible in the whole region.

Notice that the uniqueness theorem holds if *either* $\delta\vec{E}=0$ *or* $\delta\vec{H}=0$ *is true on any part of the boundary*.

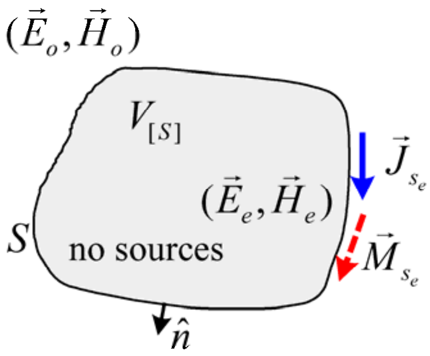
2. Equivalence principles

The equivalence principle follows from the uniqueness theorem. It allows us to build simpler to solve problems. As long as the equivalent problem preserves the boundary conditions of the original problem for the field at *S*, it is going to produce the only one possible solution for the region outside $V_{[S]}$.



(a) Original problem

$\Rightarrow$

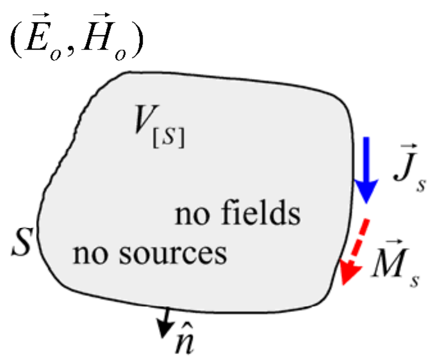


(b) General equivalent problem

$\Downarrow$

$$\begin{cases}\vec{J}_{s_e}=\hat{n}\times\left(\vec{H}_o-\vec{H}_e\right)\\ \vec{M}_{s_e}=\left(\vec{E}_o-\vec{E}_e\right)\times\hat{n}\end{cases}\qquad(12.6)$$

$$\begin{cases}\vec{J}_s=\hat{n}\times\vec{H}_o\\ \vec{M}_s=\vec{E}_o\times\hat{n}\end{cases}\qquad(12.7)$$



(c) Equivalent problem with zero fields

The zero-field formulation is often referred to as *Love's equivalence principle.*

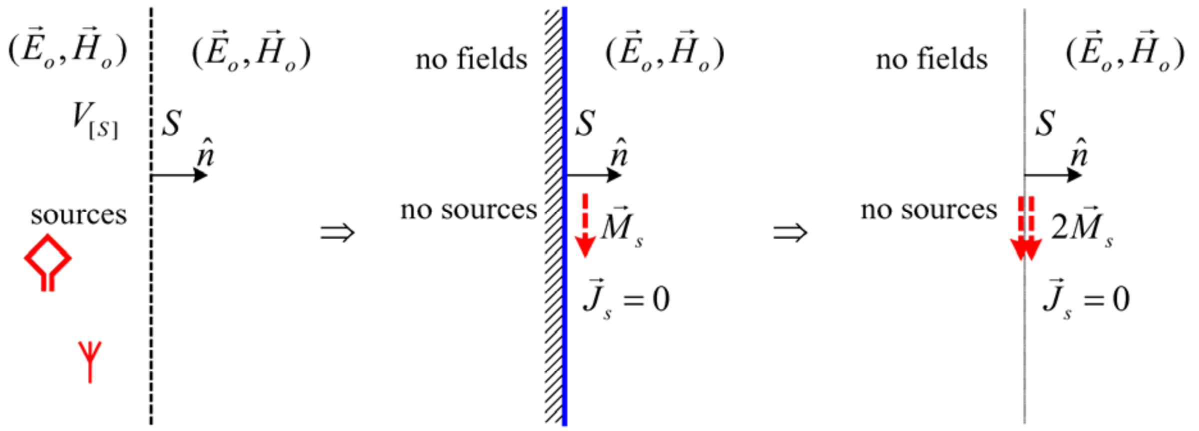
One can apply Love's equivalence principle in three different ways:

(a) One can assume that the boundary *S* is a perfect conductor. This eliminates the surface electric currents, i.e. $\vec{J}_s = 0$, and leaves just surface magnetic currents $\vec{M}_s$, which radiate in the presence of a perfect electric surface.

(b) One can assume that the boundary *S* is a perfect magnetic conductor. This eliminates the surface magnetic currents, i.e. $\vec{M}_s = 0$, and leaves just surface electric currents $\vec{J}_s$, which radiate in the presence of a perfect magnetic surface.

(c) Make no assumptions about the materials inside *S*, and define both $\vec{J}_s$ and $\vec{M}_s$ currents, which are radiating in free space (no fictitious conductors behind them). It can be shown that these equivalent currents create zero fields inside $V_{[S]}$.

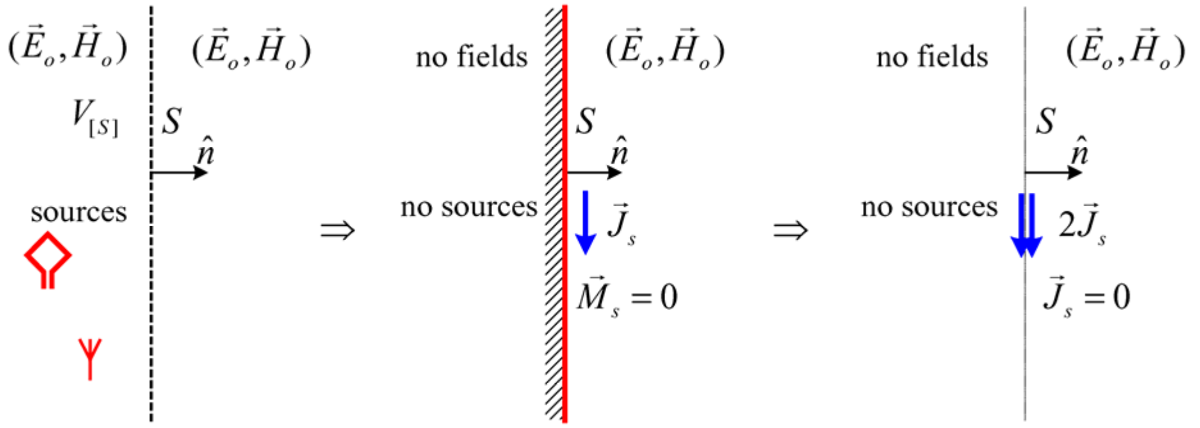
All three approaches lead to the same field solution according to the uniqueness theorem. The first two approaches are not very useful in the general case of curvilinear boundary surface *S*. However, in the case of flat infinite planes (walls), the image theory can be used to reduce the problem to an open one. Image theory can be successfully applied to curved surfaces provided the curvature's radius is large compared to the wavelength. Here is how one can implement Love's equivalence principles in conjunction with image theory.



(a) Original problem

(b) Equivalent problem - electric wall

(c) Equivalent problem - images



(a) Original problem

(b) Equivalent problem - magnetic wall

(c) Equivalent problem - images

The above approach is used to evaluate fields in half-space as excited by apertures. The field behind *S* is assumed known. This is enough to define equivalent surface currents. Using image theory, the open-region far-zone solutions for the vector potentials, $\vec{A}$ (resulting from $\vec{J}_s$) and $\vec{F}$ (resulting from $\vec{M}_s$), are found from:

$$\vec{A}(P) = \mu \frac{e^{-j\beta r}}{4\pi r} \iint_S \vec{J}_s(\vec{r}\,') e^{j\beta \hat{r}\cdot\vec{r}\,'} ds' \tag{12.8}$$

$$\vec{F}(P) = \varepsilon \frac{e^{-j\beta r}}{4\pi r} \iint_S \vec{M}_s(\vec{r}\,') e^{j\beta \hat{r}\cdot\vec{r}\,'} ds' \tag{12.9}$$

Here, $\hat{r}$ denotes the unit vector pointing from the origin of the coordinate system to the point of observation *P*. The integration point *Q* is specified through the radius-vector $\vec{r}\,'$. In the far zone, it is assumed that the field propagates radially away from the antenna. It is convenient to introduce the so-called *propagation vector*:

$$\vec{\beta} = \beta \hat{r}, \tag{12.10}$$

which characterizes both the phase constant and the direction of propagation of the wave. The vector potentials can then be written as:

$$\vec{A}(P) = \mu \frac{e^{-j\beta r}}{4\pi r} \iint_S \vec{J}_s(\vec{r}\,') e^{j\vec{\beta}\cdot\vec{r}\,'} ds' \tag{12.11}$$

$$\vec{F}(P) = \varepsilon \frac{e^{-j\beta r}}{4\pi r} \iint_S \vec{M}_s(\vec{r}') e^{j\vec{\beta}\cdot\vec{r}'} ds' \quad (12.12)$$

The relations between the far-zone fields and the vector potentials are rather simple.

$$\vec{E}_A^{far} = -j\omega(A_\theta \hat{\theta} + A_\varphi \hat{\varphi}) \quad (12.13)$$

$$\vec{H}_F^{far} = -j\omega(F_\theta \hat{\theta} + F_\varphi \hat{\varphi}) \quad (12.14)$$

Since

$$\vec{E}_F^{far} = \eta \vec{H}_F^{far} \times \vec{r}, \quad (12.15)$$

the total far-zone electric field is found as:

$$\vec{E}^{far} = \vec{E}_A^{far} + \vec{E}_F^{far} = -j\omega\left[\left(A_\theta - \eta F_\varphi\right)\hat{\theta} + \left(A_\varphi + \eta F_\theta\right)\hat{\varphi}\right] \quad (12.16)$$

Equation (12.16) involves both vector potentials as arising from both types of surface currents. Computations are reduced in half if image theory is used in conjunction with an electric or magnetic wall assumption.

3. Application of the equivalence principle to aperture problems

The equivalence principle is widely used in the analysis of aperture antennas. To calculate exactly the far fields, the exact field distribution at the aperture is needed. In the case of exact knowledge of the aperture field distribution, all three approaches given above will produce the same results. However, such exact knowledge of the aperture field distribution is usually impossible, and certain approximations are used. Then, the three equivalence-principle approaches produce slightly different results, the consistency being dependent on how accurate our knowledge about the aperture field is. Usually, it is assumed that the field is to be determined in half-space, leaving the feed and the antenna behind a infinite wall S (electric or magnetic). The aperture of the antenna S_A is this portion of S where we have an approximate knowledge of the field distribution based on the type of the feed line or the incident wave illuminating the aperture. This is the so-called *physical optics* approximation, which certainly is more accurate than the *geometrical optics* approach of ray tracing. The larger the aperture (as compared to the wavelength), the more accurate the approximation based on the incident wave.

Let us assume that the fields at the aperture are known: $\vec{E}_a, \vec{H}_a$, and they are zero everywhere else at S. The equivalent current densities are:

$$\left|\begin{array}{l} \vec{J}_s = \hat{n}\times\vec{H}_a \\ \vec{M}_s = \vec{E}_a\times\hat{n} \end{array}\right. \tag{12.17}$$

Using (12.17) in (12.11) and (12.12) produces:

$$\vec{A}(P) = \mu\frac{e^{-j\beta r}}{4\pi r}\hat{n}\times\iint_S \vec{H}_a e^{j\vec{\beta}\cdot\vec{r}'}ds' \tag{12.18}$$

$$\vec{F}(P) = -\varepsilon\frac{e^{-j\beta r}}{4\pi r}\hat{n}\times\iint_S \vec{E}_a e^{j\vec{\beta}\cdot\vec{r}'}ds' \tag{12.19}$$

The radiation integrals in (12.18) and (12.19) will be denoted shortly as:

$$\vec{\mathfrak{I}}^H = \iint_S \vec{H}_a e^{j\vec{\beta}\cdot\vec{r}'}ds' \tag{12.20}$$

$$\vec{\mathfrak{I}}^E = \iint_S \vec{E}_a e^{j\vec{\beta}\cdot\vec{r}'}ds' \tag{12.21}$$

One can find general vector expression for the far-field $\vec{E}$ vector making use of equation (12.16) written as:

$$\vec{E}^{far} = -j\omega\vec{A} - j\omega\eta\vec{F}\times\hat{r}, \tag{12.22}$$

where ***the longitudinal*** A_r ***component is to be neglected***. Substituting (12.18) and (12.19) yields:

$$\vec{E}^{far} = -j\beta\frac{e^{-j\beta r}}{4\pi r}\hat{r}\times\iint_{S_A}\left[\hat{n}\times\vec{E}_a - \eta\hat{r}\times\left(\hat{n}\times\vec{H}_a\right)\right]e^{j\vec{\beta}\cdot\vec{r}'}ds' \tag{12.23}$$

This is the full vector form of the radiated field resulting from the aperture field, and it is referred to as the *vector diffraction integral* (or *vector Kirchhoff integral*).

We shall now consider a practical case of a flat aperture lying in the $x - y$ plane with $\hat{n} \equiv \hat{z}$. Then:

$$\vec{A} = \mu \frac{e^{-j\beta r}}{4\pi r}\left(-\mathcal{J}_y^H \hat{x} + \mathcal{J}_x^H \hat{y}\right) \tag{12.24}$$

$$\vec{F} = -\varepsilon \frac{e^{-j\beta r}}{4\pi r}\left(-\mathcal{J}_y^E \hat{x} + \mathcal{J}_x^E \hat{y}\right) \tag{12.25}$$

The integrals in the above expressions can be explicitly written for this case in which $\vec{r}' = x'\hat{x} + y'\hat{y}$:

$$\mathcal{J}_x^E = \iint_{S_A} E_{a_x}(x', y') e^{j\beta(x'\sin\theta\cos\varphi + y'\sin\theta\sin\varphi)} dx'dy' \tag{12.26}$$

$$\mathcal{J}_y^E = \iint_{S_A} E_{a_y}(x', y') e^{j\beta(x'\sin\theta\cos\varphi + y'\sin\theta\sin\varphi)} dx'dy' \tag{12.27}$$

$$\mathcal{J}_x^H = \iint_{S_A} H_{a_x}(x', y') e^{j\beta(x'\sin\theta\cos\varphi + y'\sin\theta\sin\varphi)} dx'dy' \tag{12.28}$$

$$\mathcal{J}_y^H = \iint_{S_A} H_{a_y}(x', y') e^{j\beta(x'\sin\theta\cos\varphi + y'\sin\theta\sin\varphi)} dx'dy' \tag{12.29}$$

Note that the above integrals are exactly the double inverse Fourier transforms of the aperture field's components.

The vector potentials in spherical terms are:

$$\vec{A} = \mu \frac{e^{-j\beta r}}{4\pi r}\left[\hat{\theta}\cos\theta\left(\mathcal{J}_x^H \sin\varphi - \mathcal{J}_y^H \cos\varphi\right) + \hat{\varphi}\left(\mathcal{J}_x^H \cos\varphi + \mathcal{J}_y^H \sin\varphi\right)\right] \tag{12.30}$$

$$\vec{F} = -\varepsilon \frac{e^{-j\beta r}}{4\pi r}\left[\hat{\theta}\cos\theta\left(\mathcal{J}_x^E \sin\varphi - \mathcal{J}_y^E \cos\varphi\right) + \hat{\varphi}\left(\mathcal{J}_x^E \cos\varphi + \mathcal{J}_y^E \sin\varphi\right)\right] \tag{12.31}$$

By substituting the above expressions in (12.16), one obtains the far $\vec{E}$ field components as:

$$E_\theta = j\beta \frac{e^{-j\beta r}}{4\pi r}[\mathcal{J}_x^E \cos\varphi + \mathcal{J}_y^E \sin\varphi + \eta\cos\theta(\mathcal{J}_y^H \cos\varphi - \mathcal{J}_x^H \sin\varphi)] \tag{12.32}$$

$$E_\varphi = j\beta \frac{e^{-j\beta r}}{4\pi r}[-\eta(\mathcal{J}_x^H \cos\varphi + \mathcal{J}_y^H \sin\varphi) + \cos\theta(\mathcal{J}_y^E \cos\varphi - \mathcal{J}_x^E \sin\varphi)] \tag{12.33}$$

For apertures mounted on a conducting plane, the preferred equivalent model is the one with electric wall with magnetic current density

$$\vec{M}_s = 2\cdot\left(\vec{E}_a \times \hat{n}\right) \qquad (12.34)$$

radiating in open space. The solution, of course, is valid only for $z \geq 0$. In this case, $\vec{\mathcal{J}}^H = 0$.

For apertures in open space, the dual current formulation is used. Then, a usual assumption is that the aperture fields are related as in the TEM-wave case:

$$\vec{H}_a = \frac{1}{\eta}\hat{z}\times\vec{E}_a \qquad (12.35)$$

This implies that

$$\vec{\mathcal{J}}^H = \frac{1}{\eta}\hat{z}\times\vec{\mathcal{J}}^E \quad \text{or} \quad \mathcal{J}_x^H = -\frac{\mathcal{J}_y^E}{\eta}, \mathcal{J}_y^H = \frac{\mathcal{J}_x^E}{\eta} \qquad (12.36)$$

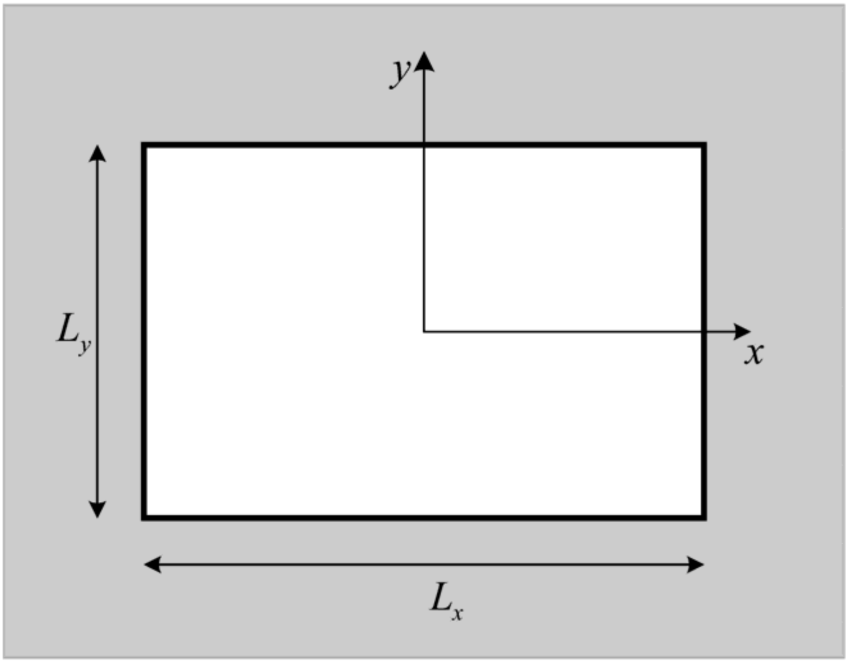
This assumption is valid for moderate and high-gain apertures; therefore, the apertures should be at least a couple of wavelengths in extent. The above assumptions reduce (12.32)-(12.33) to:

$$E_\theta = j\beta\eta\frac{e^{-j\beta r}}{4\pi r}\frac{(1+\cos\theta)}{2}\left[\mathcal{J}_x^E\cos\varphi + \mathcal{J}_y^E\sin\varphi\right] \qquad (12.37)$$

$$E_\varphi = j\beta\eta\frac{e^{-j\beta r}}{4\pi r}\frac{(1+\cos\theta)}{2}\left[\mathcal{J}_y^E\cos\varphi - \mathcal{J}_x^E\sin\varphi\right] \qquad (12.38)$$

4. The uniform rectangular aperture on an infinite ground plane

A rectangular aperture is defined in the $x-y$ plane as shown below.



If the fields are uniform in amplitude and phase across the aperture, it is referred to as a *uniform rectangular aperture*. Let us assume that the aperture field is *y*-polarized.

$$\vec{E}_a = E_0\hat{y}, \quad |x| \le \frac{L_x}{2} \ \text{ and } \ |y| \le \frac{L_y}{2} \qquad (12.39)$$

According to the equivalence principle, we assume an electric wall at $z=0$, where the equivalent magnetic current density is given by $\vec{M}_{s_e} = \vec{E}\times\hat{n}$. Applying image theory, one can find the equivalent sources radiating in open space as:

$$\vec{M}_s = 2\vec{M}_{s_e} = 2E_0\hat{y}\times\hat{z} = 2E_0\hat{x} \qquad (12.40)$$

The only non-zero radiation integral is:

$$\mathcal{J}_y^E = 2E_0 \int_{-L_x/2}^{L_x/2} e^{j\beta x' \sin\theta\cos\varphi} dx' \cdot \int_{-L_y/2}^{L_y/2} e^{j\beta y' \sin\theta\sin\varphi} dy' =$$

$$= 2E_0 L_x L_y \frac{\sin\left[\frac{\beta L_x}{2}\sin\theta\cos\varphi\right]}{\frac{\beta L_x}{2}\sin\theta\cos\varphi} \cdot \frac{\sin\left[\frac{\beta L_y}{2}\sin\theta\sin\varphi\right]}{\frac{\beta L_y}{2}\sin\theta\sin\varphi} \qquad (12.41)$$

It is appropriate to introduce the pattern variables:

$$\left| \begin{aligned} u &= \frac{\beta L_x}{2}\sin\theta\cos\varphi \\ v &= \frac{\beta L_y}{2}\sin\theta\sin\varphi \end{aligned} \right. \qquad (12.42)$$

The complete radiation fields are found by substituting (12.41) in (12.32) and (12.33):

$$\left| \begin{aligned} E_\theta &= j\beta \frac{e^{-j\beta r}}{2\pi r} E_0 L_x L_y \sin\varphi \frac{\sin u}{u}\frac{\sin v}{v} \\ E_\varphi &= j\beta \frac{e^{-j\beta r}}{2\pi r} E_0 L_x L_y \cos\theta\cos\varphi \frac{\sin u}{u}\frac{\sin v}{v} \end{aligned} \right. \qquad (12.43)$$

The total-field amplitude pattern is, therefore:

$$|\bar{E}| = F(\theta,\varphi) = \sqrt{\sin^2\varphi + \cos^2\theta\cos^2\varphi} \cdot \frac{\sin u}{u}\frac{\sin v}{v} =$$

$$= \sqrt{1-\sin^2\theta\cos^2\varphi} \cdot \frac{\sin u}{u}\frac{\sin v}{v} \qquad (12.44)$$

The principal plane patterns are:

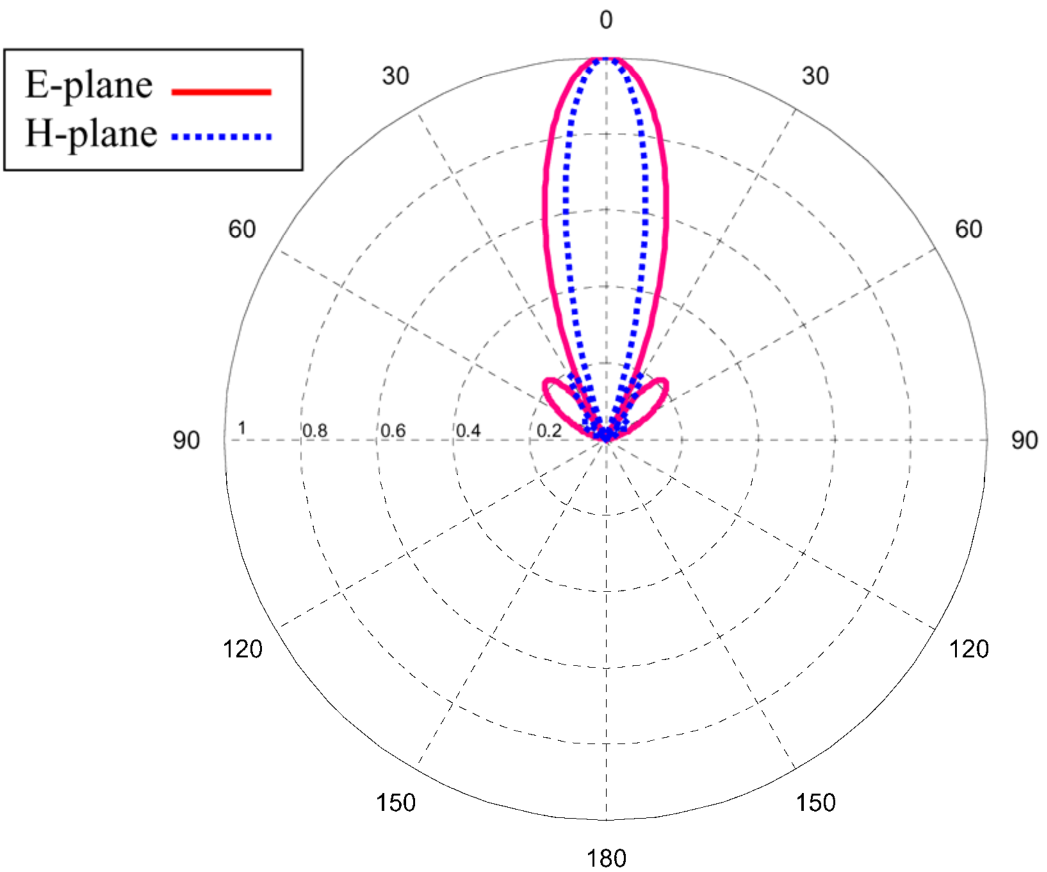
<u>*E*-plane pattern</u> $(\varphi = \pi/2)$

$$\bar{E}_\theta = \frac{\sin\left(\frac{\beta L_y}{2}\sin\theta\right)}{\left(\frac{\beta L_y}{2}\sin\theta\right)} \tag{12.45}$$

<u>*H*-plane pattern</u> $(\varphi = 0)$

$$\bar{E}_\varphi = \cos\theta \frac{\sin\left(\frac{\beta L_x}{2}\sin\theta\right)}{\left(\frac{\beta L_x}{2}\sin\theta\right)} \tag{12.46}$$

Principle patterns for aperture of size: $L_x = 3\lambda,\ L_y = 2\lambda$

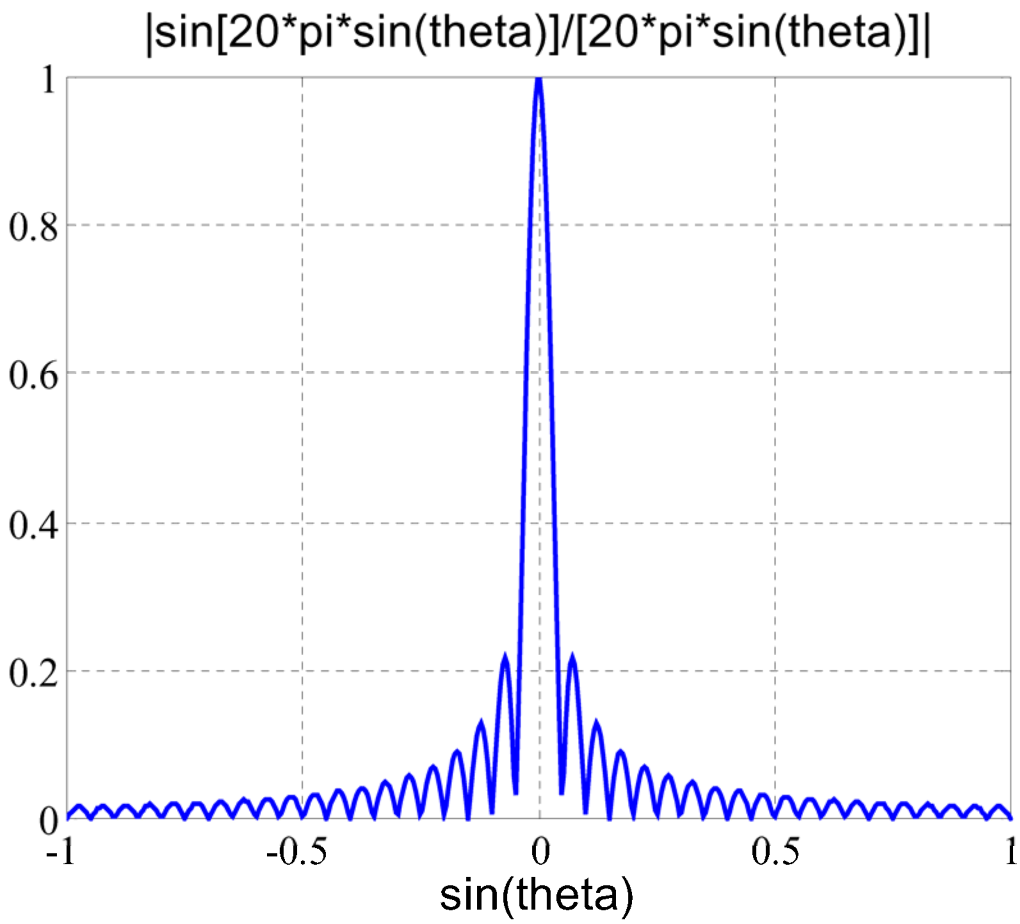


For electrically large apertures, the main beam is narrow and the $\sqrt{1-\sin^2\theta\cos^2\varphi}$ in (12.44) is negligible, i.e. it is roughly equal to 1 for all observation angles within the main beam. That is why, in the theory of large arrays, it is assumed that the amplitude pattern of a rectangular aperture is:

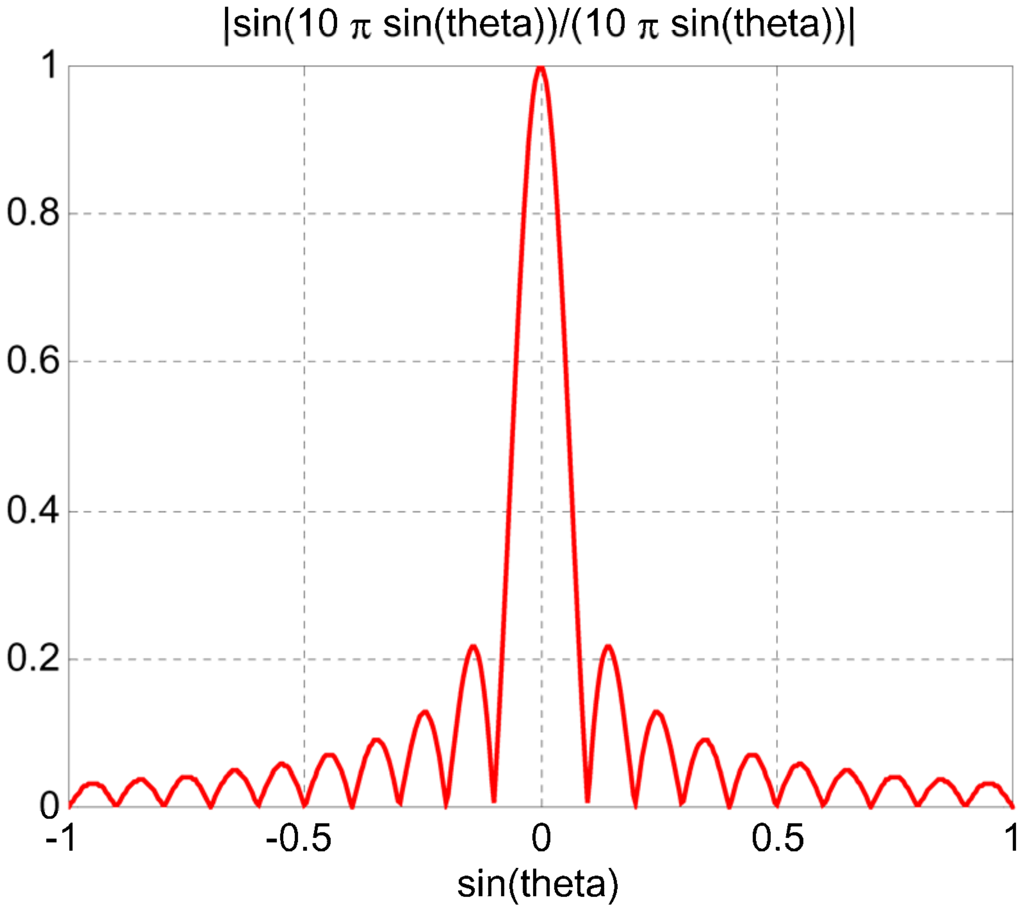
$$f(u,v) \simeq \left| \frac{\sin u}{u} \frac{\sin v}{v} \right| \tag{12.47}$$

where $u = \frac{\beta L_x}{2}\sin\theta\cos\varphi$ and $v = \frac{\beta L_y}{2}\sin\theta\sin\varphi$.

Here is a view of the $|\sin u/u|$ function for $L_x = 20\lambda$ and $\varphi = 0^\circ$ (H-plane pattern):



Here is a view of the $|\sin v / v|$ function for $L_y = 10\lambda$ and $\varphi = 90^\circ$ (E-plane pattern):



<u>Beamwidths</u>

(a) first-null beamwidth

One needs the location of the first nulls in the pattern in order to calculate the FNBW. The nulls of the E-plane pattern are determined from (12.45) as:

$$\frac{\beta L_y}{2} \sin\theta_{/\theta=\theta_n} = n\pi, \quad n = 1, 2, \ldots \tag{12.48}$$

$$\Rightarrow \theta_n = \arcsin\left(\frac{n\lambda}{L_y}\right), \text{ rad} \tag{12.49}$$

The first null occurs at $n = 1$.

$$\Rightarrow FNBW_E = 2\theta_n = 2\arcsin\left(\frac{\lambda}{L_y}\right), \text{ rad} \tag{12.50}$$

In a similar fashion, $FNBW_H$ is determined to be:

$$FNBW_H = 2\arcsin\left(\frac{\lambda}{L_x}\right), \text{ rad} \tag{12.51}$$

(b) half-power beamwidth

The half-power point in the E-plane occurs when

$$\frac{\sin\left(\frac{\beta L_y}{2}\sin\theta\right)}{\left(\frac{\beta L_y}{2}\sin\theta\right)} = \frac{1}{\sqrt{2}} \tag{12.52}$$

or

$$\frac{\beta L_y}{2}\sin\theta_{/\theta=\theta_h} = 1.391 \tag{12.53}$$

$$\Rightarrow \theta_h = \arcsin\left(\frac{0.443\lambda}{L_y}\right), \text{ rad} \tag{12.54}$$

$$HPBW_E = 2\arcsin\left(\frac{0.443\lambda}{L_y}\right) \tag{12.55}$$

A first-order approximation is possible for very small arguments in (12.55), i.e. when $L_y \gg 0.443\lambda$ (large aperture):

$$HPBW_E \simeq 0.886\frac{\lambda}{L_y} \tag{12.56}$$

The half-power beamwidth in the H-plane is analogous:

$$HPBW_H = 2\arcsin\left(\frac{0.443\lambda}{L_x}\right) \tag{12.57}$$

Side-lobe level

It is obvious from the properties of the $|\sin x / x|$ function that the first side lobe has the largest maximum of all side lobes, and it is:

$$|E_\theta(\theta=\theta_s)| = \left|\frac{\sin 4.494}{4.494}\right| = 0.217 = -13.26\text{, dB} \qquad (12.58)$$

When evaluating side-lobe levels and beamwidths in the H-plane, one has to include the $\cos\theta$ factor, too. The smaller the aperture, the less important this factor is.

Directivity

In a general approach to the calculation of the directivity, the total radiated power Π has to be calculated first using the far-field pattern expression (12.44).

$$D_0 = \frac{4\pi}{\Omega_A} = 4\pi \frac{U_{\max}}{\Pi_{rad}} \qquad (12.59)$$

Here,

$$U(\theta,\varphi) = \frac{1}{2\eta}\left[|E_\theta|^2 + |E_\varphi|^2\right] r^2 = U_{\max} |F(\theta,\varphi)|^2 \qquad (12.60)$$

$$\Omega_A = \int_0^{2\pi}\int_0^{\pi} |F(\theta,\varphi)|^2 \sin\theta d\theta d\varphi \qquad (12.61)$$

However, in the case of an aperture illuminated by a TEM wave, one can use a simpler approach. Generally, for all aperture antennas, the assumption of a uniform TEM wave at the aperture ($\vec{E} = \hat{y}E_0$),

$$\vec{H}_a = -\hat{x}\frac{E_0}{\eta}, \qquad (12.62)$$

is quite accurate (although η is not necessarily the intrinsic impedance of vacuum). The far-field components in this case were already derived in (12.37) and (12.38). They lead to the following expression for the radiation intensity:

$$U(\theta,\varphi) = \frac{\beta^2}{32\pi^2\eta}(1+\cos\theta)^2\left[|\mathcal{J}_x^E|^2 + |\mathcal{J}_y^E|^2\right] \qquad (12.63)$$

The maximum value of the function in (12.63) is easily derived after substituting the radiation integrals from (12.26) and (12.27):

$$U_{\max} = \frac{\beta^2}{8\pi^2\eta}\left|\iint_{S_A}\vec{E}_a ds'\right|^2 \tag{12.64}$$

The integration of the radiation intensity (12.63) over a closed sphere is in general not easy. It can be avoided by observing that the total power reaching the far zone must have passed through the aperture in the first place. In the general aperture case, this power is determined as:

$$\Pi_{rad} = \oiint_S \vec{P}_{av}\cdot d\vec{s} = \frac{1}{2\eta}\iint_{S_A}|\vec{E}_a|^2\, ds \tag{12.65}$$

Substituting (12.64) and (12.65) in (12.59) finally yields:

$$D_0 = \frac{4\pi}{\lambda^2}\frac{\left|\iint_{S_A}\vec{E}_a ds'\right|^2}{\iint_{S_A}|\vec{E}_a|^2\, ds'} \tag{12.66}$$

In the case of a uniform rectangular aperture,

$$\Pi = L_x L_y \frac{|E_0|^2}{2\eta} \tag{12.67}$$

$$U_{\max} = \left(\frac{L_x L_y}{\lambda}\right)^2 \frac{|E_0|^2}{2\eta} \tag{12.68}$$

Thus, the directivity is found to be:

$$D_0 = 4\pi\frac{U_{\max}}{\Pi} = \frac{4\pi}{\lambda^2}L_x L_y = \frac{4\pi}{\lambda^2}A_p = \frac{4\pi}{\lambda^2}A_{eff} \tag{12.69}$$

The physical and the effective areas of a uniform aperture are equal.

5. The uniform rectangular aperture in open space

Now, we shall examine the same aperture when it is ***not*** mounted on a ground plane. The field distribution is the same as in (12.39) but now the $\vec{H}$ field must be defined, too, in order to apply the general form of the equivalence principle with both types of surface currents.

$$\left.\begin{aligned} \vec{E}_a &= \hat{y}E_0 \\ \vec{H}_a &= -\hat{x}\frac{E_0}{\eta} \end{aligned}\right\}, \quad \begin{aligned} -L_x/2 \le x' \le L_x/2 \\ -L_y/2 \le y' \le L_y/2 \end{aligned} \tag{12.70}$$

Above, again an assumption was made that there is a direct relation between the electric and the magnetic field components.

To form the equivalent problem, an infinite surface is chosen again to extend in the $z = 0$ plane. Over the entire surface, the equivalent $\vec{J}_s$ and $\vec{M}_s$ must be defined. Both $\vec{J}_s$ and $\vec{M}_s$ are not zero outside the aperture in the $z = 0$ plane because the field is not zero there. Moreover, the field is *not known a priori* outside the aperture. Thus, the exact equivalent problem cannot be built in practice (at least, not by making use of the infinite plane model).

The usual assumption made is that $\vec{E}_a$ and $\vec{H}_a$ are zero outside the aperture in the $z = 0$ plane, and, therefore, so are the equivalent currents $\vec{J}_s$ and $\vec{M}_s$:

$$\left.\begin{aligned} \vec{M}_s &= -\hat{n}\times\vec{E}_a = \underbrace{-\hat{z}\times\hat{y}}_{\hat{x}}E_0 \\ \vec{J}_s &= \hat{n}\times\vec{H}_a = \underbrace{\hat{z}\times(-\hat{x})}_{-\hat{y}}\frac{E_0}{\eta} \end{aligned}\right\}, \quad \begin{aligned} -L_x/2 \le x' \le L_x/2 \\ -L_y/2 \le y' \le L_y/2 \end{aligned} \tag{12.71}$$

Since the equivalent currents are related via the TEM-wave assumption, only the integral $\mathcal{J}_y^E$ is needed for substitution in the far field expressions derived in (12.37) and (12.38).

$$\mathcal{J}_y^E = 2E_0 \int_{-L_x/2}^{L_x/2} e^{j\beta x' \sin\theta \cos\varphi} dx' \cdot \int_{-L_y/2}^{L_y/2} e^{j\beta y' \sin\theta \sin\varphi} dy' =$$

$$= 2E_0 L_x L_y \frac{\sin\left[\frac{\beta L_x}{2}\sin\theta\cos\varphi\right]}{\frac{\beta L_x}{2}\sin\theta\cos\varphi} \cdot \frac{\sin\left[\frac{\beta L_y}{2}\sin\theta\sin\varphi\right]}{\frac{\beta L_y}{2}\sin\theta\sin\varphi} \quad (12.72)$$

Now, the far-field components are obtained by substituting in (12.37) and (12.38):

$$E_\theta = C \sin\varphi \frac{(1+\cos\theta)}{2} \frac{\sin u}{u} \frac{\sin v}{v}$$
$$E_\varphi = C \cos\varphi \frac{(1+\cos\theta)}{2} \frac{\sin u}{u} \frac{\sin v}{v} \quad (12.73)$$

where:

$$C = j\beta L_x L_y E_0 \frac{e^{-j\beta r}}{2\pi r};$$
$$u = \frac{\beta L_x}{2}\sin\theta\cos\varphi;$$
$$v = \frac{\beta L_y}{2}\sin\theta\sin\varphi.$$

The far-field expressions in (12.73) would be identical to those of the aperture mounted on a ground plane if $\cos\theta$ were replaced by 1. Thus, for small values of θ, the patterns of both apertures are practically identical.

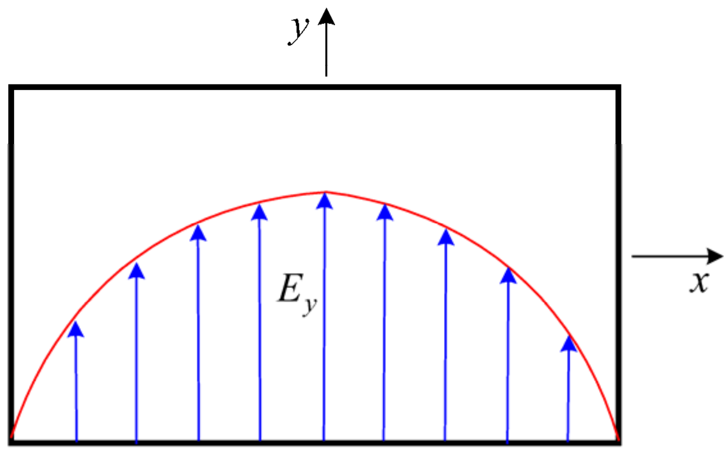
An exact analytical evaluation of the directivity is difficult. However, according to the approximations made, the directivity formula derived in (12.66) should provide accurate enough value. According to (12.66), the directivity is the same as in the case of the aperture mounted on a ground plane.

6. The tapered rectangular aperture on a ground plane

The uniform rectangular aperture has the maximum possible effective area (for an aperture-type antenna) equal to its physical area. This also implies that it has the highest possible directivity for all constant-phase excitations of a rectangular aperture. However, directivity is not the only important factor in the design of an antenna. A factor that frequently comes into a conflict with the directivity is the side-lobe level (SLL). The uniform distribution excitation produces the highest SLL of all constant-phase excitations of a rectangular aperture. It will be shown that the reduction of SLL can be achieved by tapering the equivalent sources distribution from a maximum at the aperture's center to zero values at its edges.

One very practical aperture of tapered source distribution is the open rectangular waveguide. The dominant (TE_{10}) mode has the following distribution:

$$\vec{E}_a = \hat{y} E_0 \cos\left(\frac{\pi}{L_x} x'\right), \quad \begin{cases} -L_x/2 \le x' \le L_x/2 \\ -L_y/2 \le y' \le L_y/2 \end{cases} \tag{12.74}$$



The general procedure for the far-field analysis is the same as before (in Section 4). The only difference is in the field distribution. Again, only the integral $\mathcal{J}_y^E$ is to be evaluated.

$$\mathcal{J}_y^E = 2E_0 \int_{-L_x/2}^{L_x/2} \cos\left(\frac{\pi}{L_x}x'\right) e^{j\beta x' \sin\theta\cos\varphi} dx' \cdot \int_{-L_y/2}^{L_y/2} e^{j\beta y' \sin\theta\sin\varphi} dy' \quad (12.75)$$

The integral of the y variable was already encountered in (12.41):

$$I(y) = \int_{-L_y/2}^{L_y/2} e^{j\beta y' \sin\theta\sin\varphi} dy' = L_y \frac{\sin\left[\frac{\beta L_y}{2}\sin\theta\sin\varphi\right]}{\frac{\beta L_y}{2}\sin\theta\sin\varphi} \quad (12.76)$$

The integral of the x variable is easily solved:

$$I(x) = \int_{-L_x/2}^{L_x/2} \cos\left(\frac{\pi}{L_x}x'\right) e^{j\beta x' \sin\theta\cos\varphi} dx' =$$

$$= \int_{-L_x/2}^{L_x/2} \cos\left(\frac{\pi}{L_x}x'\right)\left[\cos\left(\beta x'\sin\theta\cos\varphi\right) + j\sin\left(\beta x'\sin\theta\cos\varphi\right)\right] dx' =$$

$$= \frac{1}{2}\int_{-L_x/2}^{L_x/2} \left\{\cos\left[\left(\frac{\pi}{L_x} - \beta\sin\theta\cos\varphi\right)x'\right] + \cos\left[\left(\frac{\pi}{L_x} + \beta\sin\theta\cos\varphi\right)x'\right]\right\} dx' +$$

$$+ \frac{j}{2}\int_{-L_x/2}^{L_x/2} \left\{\sin\left[\left(\beta\sin\theta\cos\varphi - \frac{\pi}{L_x}\right)x'\right] + \cos\left[\left(\beta\sin\theta\cos\varphi + \frac{\pi}{L_x}\right)x'\right]\right\} dx'$$

$$\Rightarrow I(x) = \frac{\pi L_x}{2} \frac{\cos\left(\frac{\beta L_x}{2}\sin\theta\cos\varphi\right)}{\left(\frac{\pi}{2}\right)^2 - \frac{\beta L_x}{2}\sin\theta\cos\varphi}$$

$$\Rightarrow \mathcal{J}_y^E = \pi E_0 L_x L_y \frac{\cos\left(\frac{\beta L_x}{2}\sin\theta\cos\varphi\right)}{\left[\left(\frac{\pi}{2}\right)^2 - \underbrace{\frac{\beta L_x}{2}\sin\theta\cos\varphi}_{u}\right]} \frac{\sin\left[\frac{\beta L_y}{2}\sin\theta\sin\varphi\right]}{\underbrace{\left(\frac{\beta L_y}{2}\sin\theta\sin\varphi\right)}_{v}} \quad (12.77)$$

To derive the far-field components, (12.77) is substituted in (12.32) and (12.33).

$$E_\theta = -\frac{\pi}{2} C \sin\varphi \frac{\cos u}{\left[u^2 - \left(\frac{\pi}{2}\right)^2\right]} \frac{\sin v}{v}$$

$$E_\varphi = -\frac{\pi}{2} C \cos\theta \cos\varphi \frac{\cos u}{\left[u^2 - \left(\frac{\pi}{2}\right)^2\right]} \frac{\sin v}{v} \tag{12.78}$$

where:

$$C = j\beta L_x L_y E_0 \frac{e^{-j\beta r}}{2\pi r};$$

$$u = \frac{\beta L_x}{2} \sin\theta \cos\varphi;$$

$$v = \frac{\beta L_y}{2} \sin\theta \sin\varphi.$$

Principle plane patterns

In the E-plane, the aperture is not tapered. As expected, the E-plane principal pattern is the same as that of a uniform aperture.

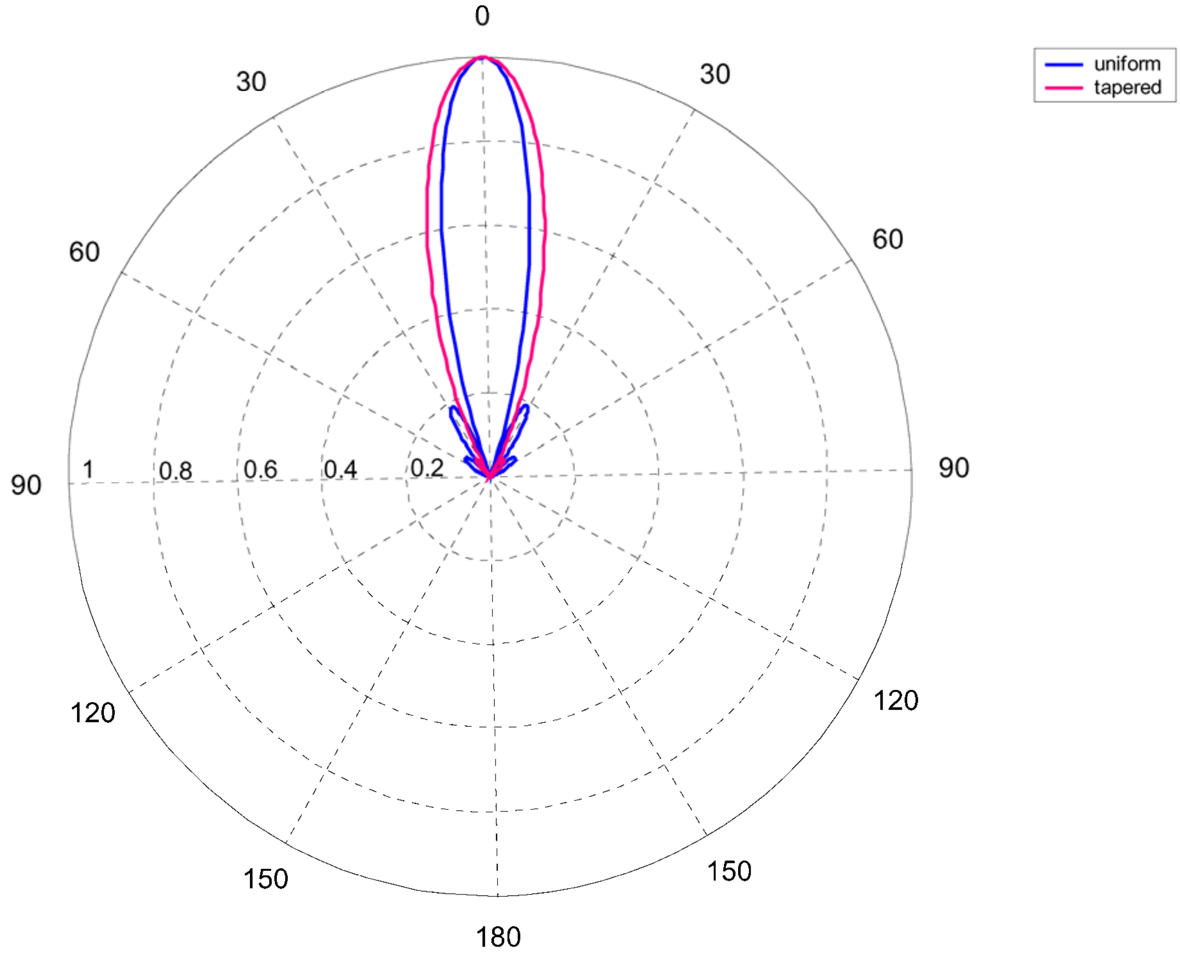
E-plane ($\varphi = 90^\circ$):

$$\overline{E}_\theta = \frac{\sin\left(\frac{\beta L_y}{2}\sin\theta\right)}{\left(\frac{\beta L_y}{2}\sin\theta\right)} \tag{12.79}$$

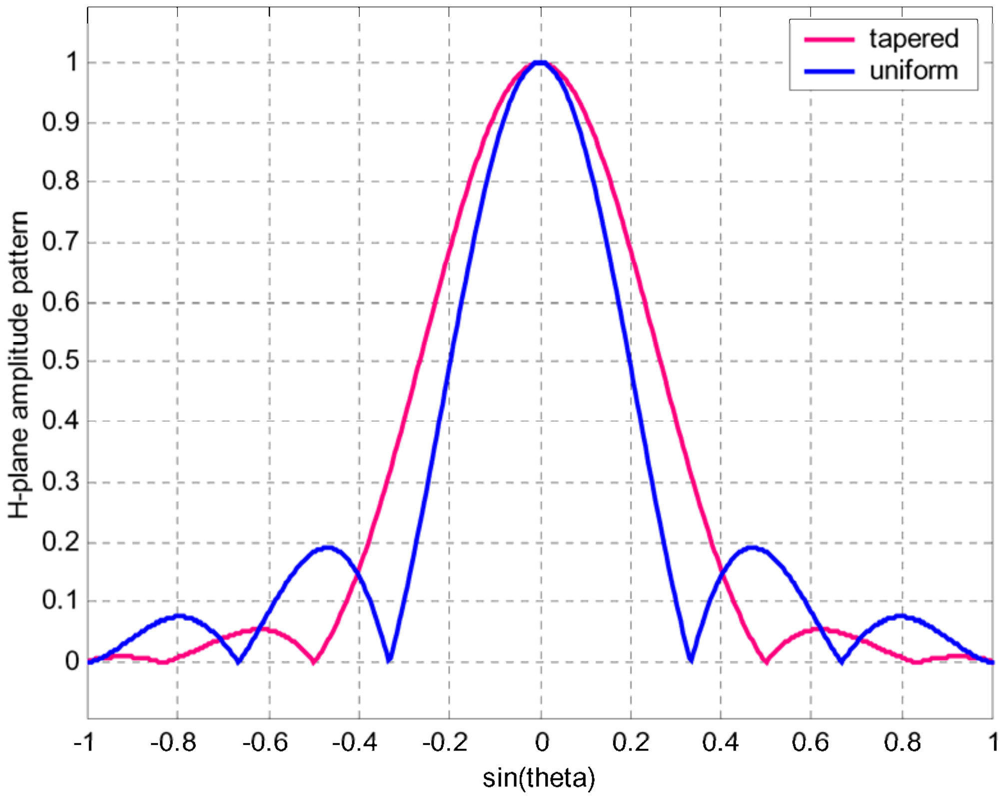
H-plane ($\varphi = 0^\circ$):

$$\overline{E}_\varphi = \cos\theta \frac{\cos\left(\frac{\beta L_x}{2}\sin\theta\right)}{\left(\frac{\beta L_x}{2}\sin\theta\right)^2 - \left(\frac{\pi}{2}\right)^2} \tag{12.80}$$

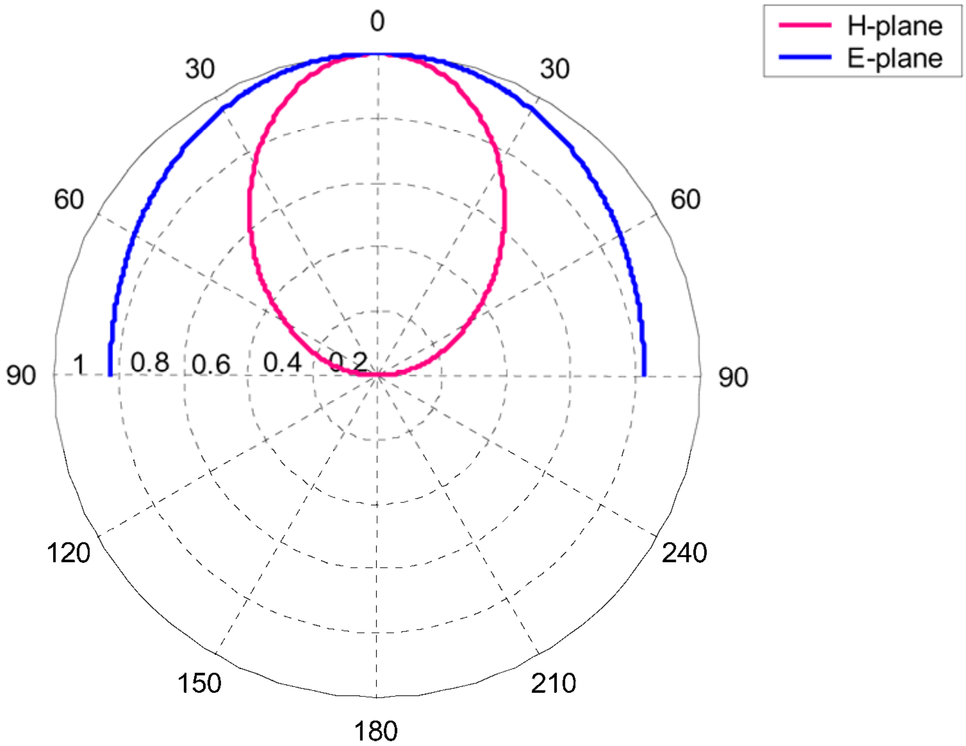
H-plane pattern – uniform vs. tapered illumination ($L_x = 3\lambda$):



The lower SLL of the tapered-source far field is obvious. It is better seen in the rectangular plot given below. The price to pay for the lower SLL is the decrease in directivity (the beamwidth of the major lobe increases).



The above example of $L_x = 3\lambda$ is illustrative on the effect of source distribution on the far-field pattern. However, a more practical example is the rectangular-waveguide open-end aperture, where the waveguide operates in a dominant mode, i.e. $\lambda_0 / 2 < L_x < \lambda_0$. Here, λ_0 is the wavelength in open space ($\lambda_0 = c / f_0$). Consider the case $L_x = 0.75\lambda$. The principal-plane patterns for an aperture on a ground plane look like this:



In the above example, a practical X-band waveguide was considered whose cross-section has the following sizes: $L_x = 2.286$ cm, $L_y = 1.016$ cm. Obviously, $\lambda_0 = 3.048$ cm, and $f_0 = 9.84$ GHz.

The case of a dominant-mode open-end waveguide radiating in free space can be analyzed following the approaches outlined in this Section and in Section 5.

The calculation of beamwidths and directivity is analogous to previous cases. Only the final results will be given here for the case of the x-tapered aperture on a ground plane.

Directivity: $$D_0 = \frac{8}{\pi^2}\left(\frac{4\pi}{\lambda^2} L_x L_y\right) \quad (12.81)$$

Effective area: $$A_{eff} = \frac{8}{\pi^2} L_x L_y = 0.81 A_p \quad (12.82)$$

Note the decrease in the effective area.

Half-power beamwidths:

$$HPBW_E = \frac{50.6}{L_y/\lambda}, \text{ deg. } (= HPBW_E \text{ of the uniform aperture}) \quad (12.83)$$

$$HPBW_H = \frac{68.8}{L_x/\lambda}, \text{ deg. } (> HPBW_H \text{ of the uniform aperture}) \quad (12.84)$$

The above results are approximate. Better results would be obtained if the following factors were taken into account:

- the phase constant of the waveguide β_g is not equal to the free-space phase constant $\beta_0 = \omega\sqrt{\mu_0\varepsilon_0}$; it is dispersive;
- the abrupt termination at the waveguide open end introduces reflection, which affects the field at the aperture;
- there are strong fringe currents at the waveguide walls, which contribute to the overall radiation.

Compact Cage Antenna

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Credit Line: www.cqham.ru

Everyone knows "classical" quad antenna. Its perimeter is equal to wavelength. Each side of the quad is 1/4 wavelength.

Amplification of the "classical" quad antenna is 3-dBi, input impedance (in a free space) is close to 120-Ohm **(Figure 1)**. Good antenna but the size!

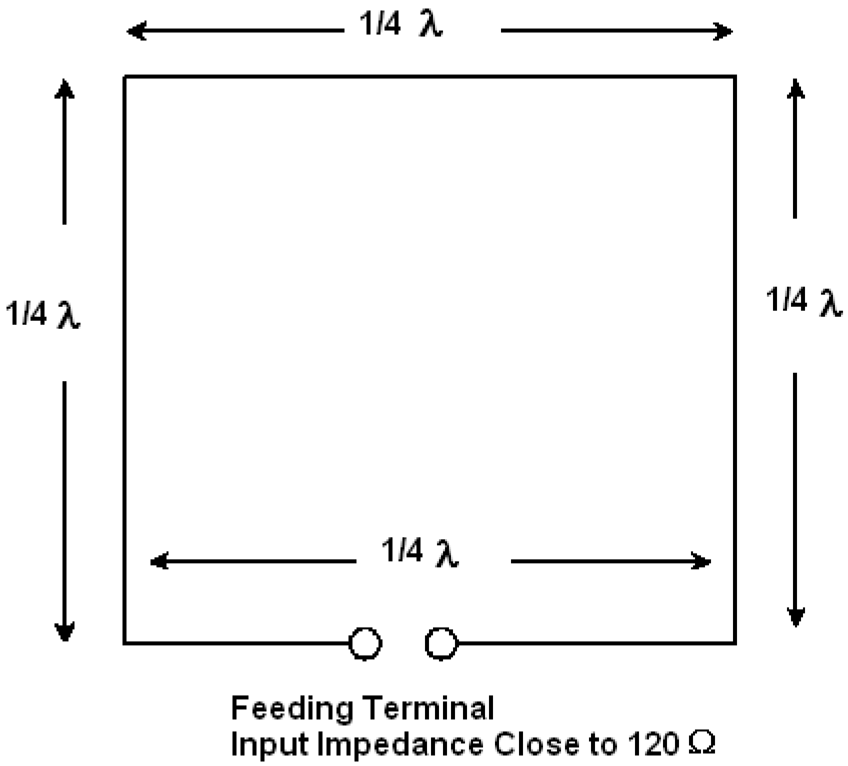


Figure 1. "Classical" quad antenna

It is possible to reduce its size by folding it in a "cage" shape. The simplest cage in which the antenna may be transformed is a cage of 10 sections 0.11 wavelength each. The Compact Cage Antenna (CCA) gain is 1.5-dBi and input impedance is close to 50-Ohm. **Figure 2** shows the CCA grabbed from MMANA. There are shown antenna currents and the source. MMANA file for the antenna (named Trihat- quad) may be found at MMANA "Antenna Library" (**References 1**) or loaded by the link below.

Figure 3 shows free space radiation patterns for Compact Cage Antenna, **Figure 4** shows Compact Cage Antenna radiation patterns at 1/3 wavelength height above the ground, **Figure 5** shows Compact Cage Antenna radiation patterns at 1.0 wavelength height above the ground.

Compact Cage Antenna may be fed at the low corner with the twin line (see **Figure 6**). The antenna may be fed with coaxial cable through a balun as well.

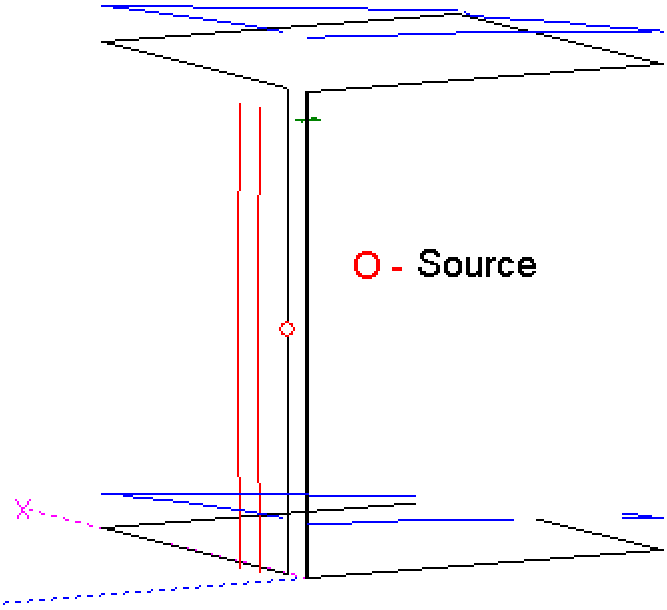
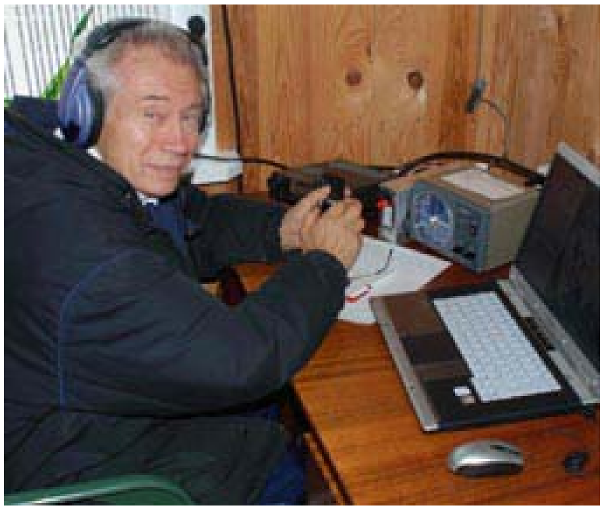
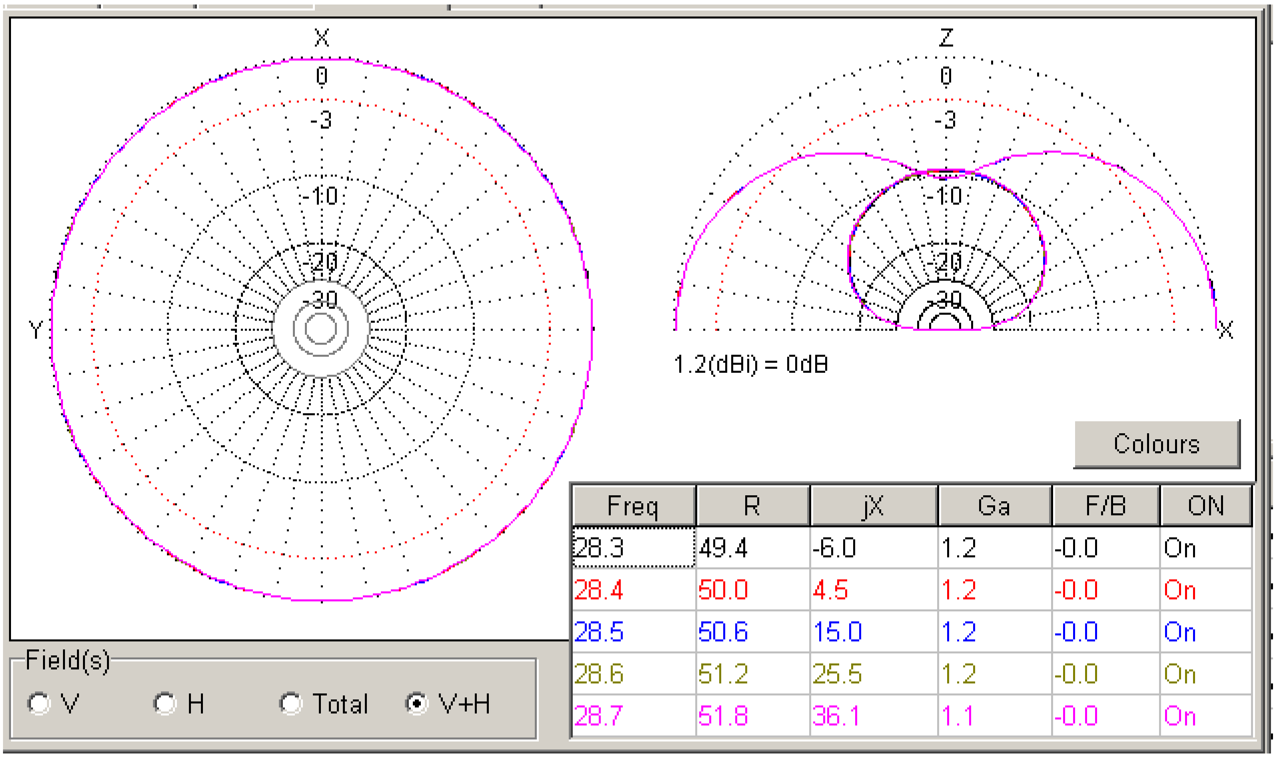


Figure 2 Compact Cage Antenna

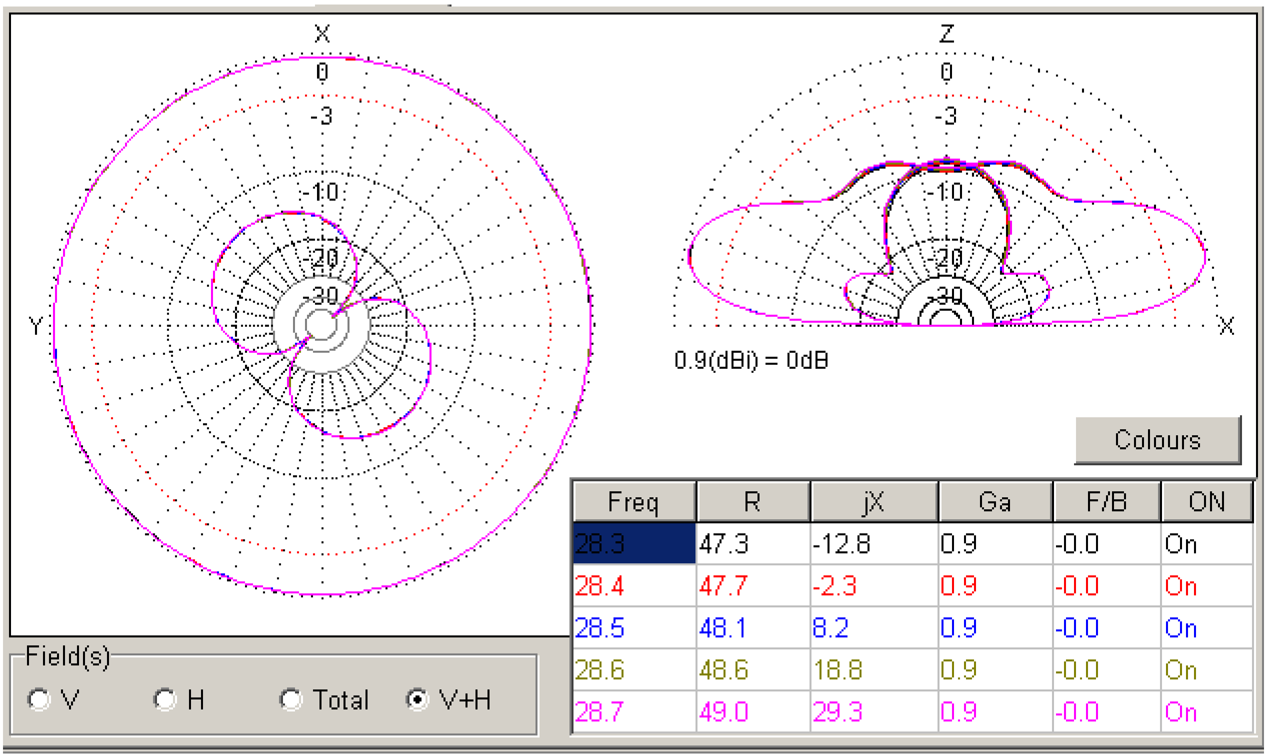


Vladislav Shcherbakov, RU3ARJ



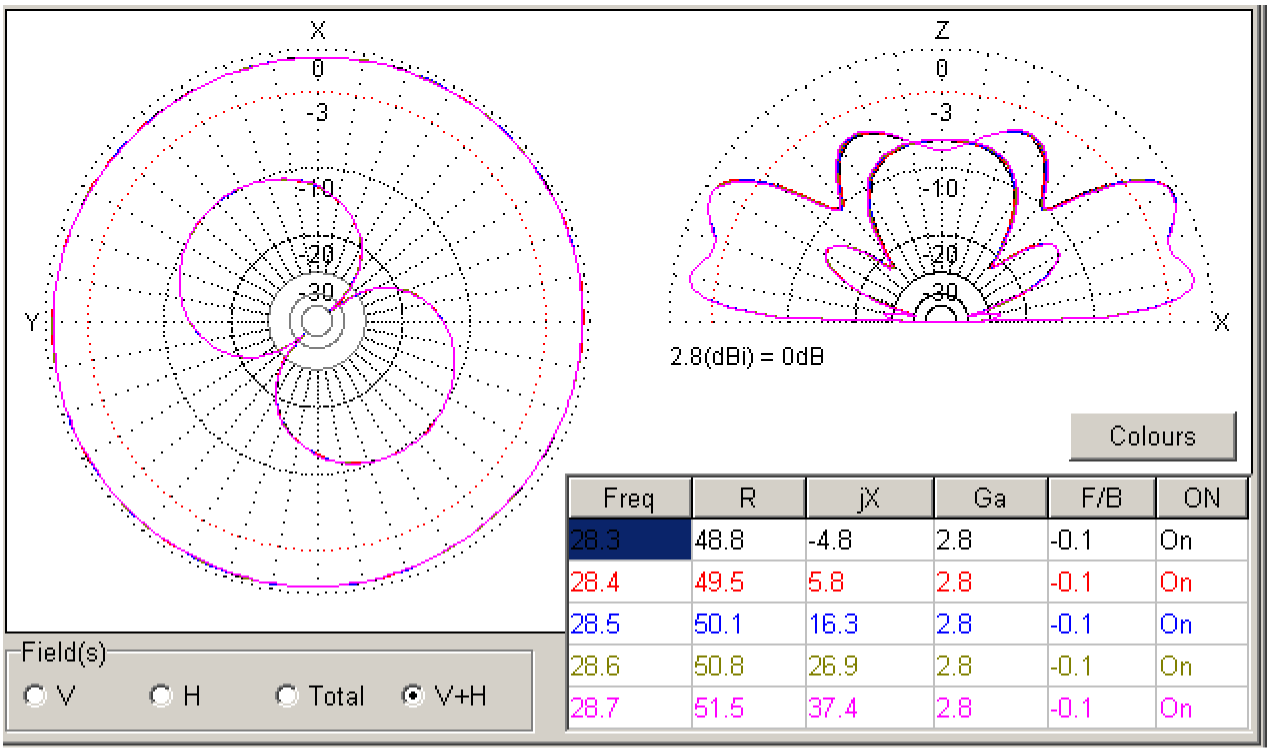
Freq	R	jX	Ga	F/B	ON
28.3	49.4	-6.0	1.2	-0.0	On
28.4	50.0	4.5	1.2	-0.0	On
28.5	50.6	15.0	1.2	-0.0	On
28.6	51.2	25.5	1.2	-0.0	On
28.7	51.8	36.1	1.1	-0.0	On

Figure 3 Free space radiation patterns for Compact Cage Antenna



Freq	R	jX	Ga	F/B	ON
28.3	47.3	-12.8	0.9	-0.0	On
28.4	47.7	-2.3	0.9	-0.0	On
28.5	48.1	8.2	0.9	-0.0	On
28.6	48.6	18.8	0.9	-0.0	On
28.7	49.0	29.3	0.9	-0.0	On

Figure 4 Compact Cage Antenna radiation patterns at 1/3 wavelength height above the ground



Freq	R	jX	Ga	F/B	ON
28.3	48.8	-4.8	2.8	-0.1	On
28.4	49.5	5.8	2.8	-0.1	On
28.5	50.1	16.3	2.8	-0.1	On
28.6	50.8	26.9	2.8	-0.1	On
28.7	51.5	37.4	2.8	-0.1	On

Figure 5 Compact Cage Antenna radiation patterns at 1.0 wavelength height above the ground

All the characteristics of corner fed CCA are almost the same as for usual CCA shown on **Figure 2**. MMANA file for the antenna (named Trihat- quad-1) may be found at MMANA "Antenna Library" (**References 1**) or loaded by the link below.

The modeling of both these antennas shows that their geometry may be modified for better match to 50-Om feeding line.

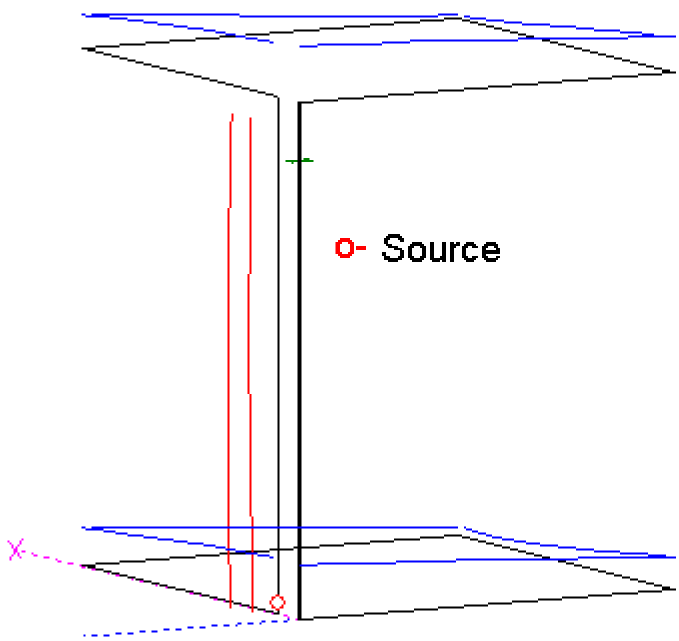


Figure 6 Compact Cage Antenna with corner feeding

It can be reached by "opening" the upper loop a bit (**Figure 7**) or bend the corner of the upper loop by 45-degrees upward (**Figure 8**). MMANA file for the antennas (**Figure 7**- cca_open_corner, **Figure 8**- cca_bended) may be loaded by link below. You may recalculate the dimensions (using function "Scale" at MMANA) for any frequency you need.

Compact Cage Antenna is a symmetric antenna, so, it is required to feed the antenna using a BalUn. Otherwise coaxial cable feeder would affect antenna input impedance and radiation patterns. As a BalUn we use 5-10 ferrite cores on the coax near the feeding point. You may wind 3-6 turns of the coax on a larger ferrite ring as well (near the feeding point).

Here are some building notes: for example, sides of the CCA for the 10-meters are near 1 meter in length. So, the antenna can be made from a copper or aluminum tube 5-10-millimeters in diameter. For lower bands you may use a copper wire 1-3 mm, stretched by some dielectric frame. For UHF-VHF bands the antenna may be made of any bimetallic conductor.

Some advantages of the CCA:
1. Tiny, toy-like size.
2. Bandwidth similar to usual full sized loop.
3. Gain is close to dipole's gain.
4. All-directional radiation pattern in the horizontal plane.

5. Low elevation angle.
6. Low-noise and no electrostatic.
7. Good match to 50-Ohm coaxial.

It could be very interesting to design a phase antenna arrays either with passive or active feeding using several CCA's.

However, I leave this for the readers

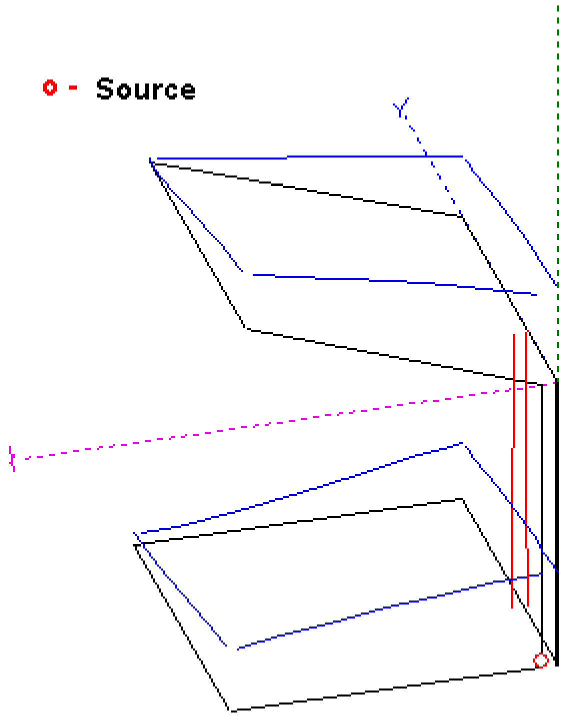


Figure 7 Compact Cage Antenna with "opened" the upper loop

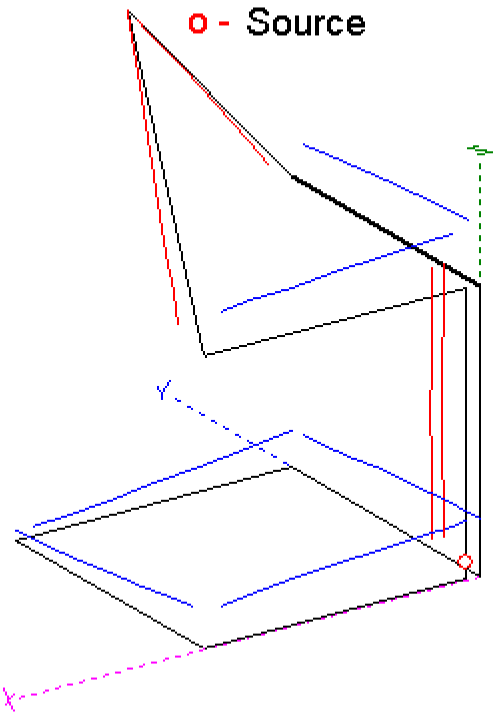


Figure 8 Compact Cage Antenna with a corner bended upward

References: www.dl2kq.de

Files MMANA:

http://www.antentop.org/011/cca_011.htm

Antennas by Nikolay Kudryavchenko, UR0GT

The publications devoted to memory UR0GT.

Antentop (from the issue) begin to publish antennas designed by Nick Kudryavchenko, UR0GT. No changes (compare to previous their publishing in the Internet) are added to the description and MMANA files of the antennas.

Some words about the history of the publication.

One year ago, in May 2008, I contacted with UR0GT by e- mail and asked him about permission for publishing his antennas design (translated in English in e-magazine AntenTop) that were already posted by him in some forums in Russian. Nick answered to me, since the stuff was already published in Russian and the stuff were accessible to public in the internet, he did not mention about my publishing the stuff in English. Nick conditions for publishing were simple: must be indicated his authorship and antennas MMANA files must be unchangeable.

However lots of antennas design of UR0GT (that were published in the Internet forums) had very brief commentary. Sometimes Nick posted in a forum just an antenna file (without any commentary). Therefore I dared ask Nick to write some commentary to those files and rework some antennas MMANA files. Nick answered to me that he had no time at this summer and ever this autumn to do such work but probably next winter (winter- 2009) he would try to do this. So, we agreed to get in touch in the winter 2009.

To regret of all radio- amateurs Nick died at June 2008. Half of year left to his pension, where he planned to work on his antennas and may be on his Antenna Book…. Radio- amateurs lost very talented antennas designer….

So, I decided to publish his antenna design like we agreed with him before at his life, i.e. no changes added to his antenna files and his commentaries.

DEWD – Famous invention of UR0GT

Credit Line:
http://www.radioscanner.ru/forum/topic25617-12.htm

All antennas design from UR0GT may be pasted to websites/media at above mention conditions.

DEWD Vertical for the 80- meters

The publications devoted to memory UR0GT.

By: **Nikolay Kudryavchenko, UR0GT**

Variant of a DEWD Vertical Antenna for 3.5 – 4.0 MHz looks astonishingly simply. Just one wire in diameter of 2- mm and 18.04 meters long is installed on the distance 74- cm near a Vertical Antenna with diameter of 60- mm and height of 20.6 meters.

This additional wire should be connected to counterpoises or to other (common with the Vertical Antenna) grounding. The simplified model of the antenna (file *DV3.5_4MHz.maa*) without counterpoises ("ground" does the job) may be loaded:

http: // www.antentop.org/011/vertical_011.htm

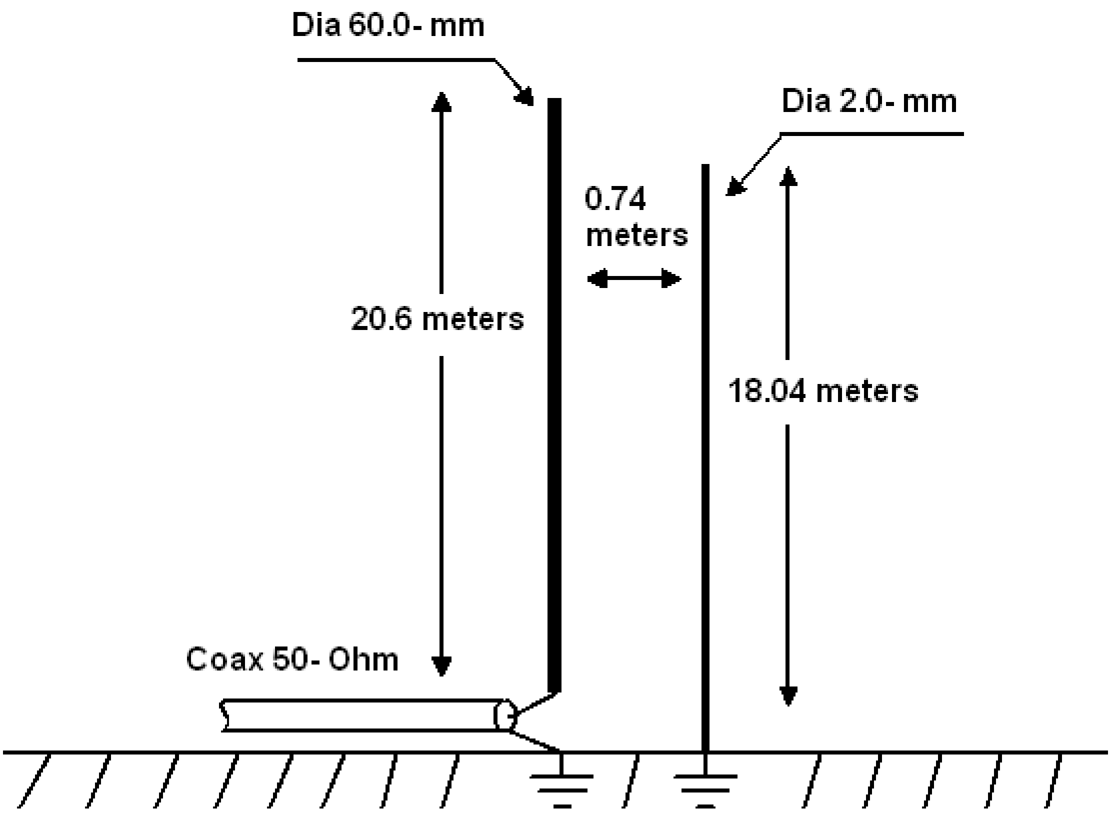


Figure 1 Simplified DEWD Vertical for the 80- meters

SWR for the DEWD Vertical is not exceeded 1.45:1.0 at the range 3.5 – 4.0- MHz.

The rest parameters for the antenna are almost similar to ¼-lambda Vertical Antenna.

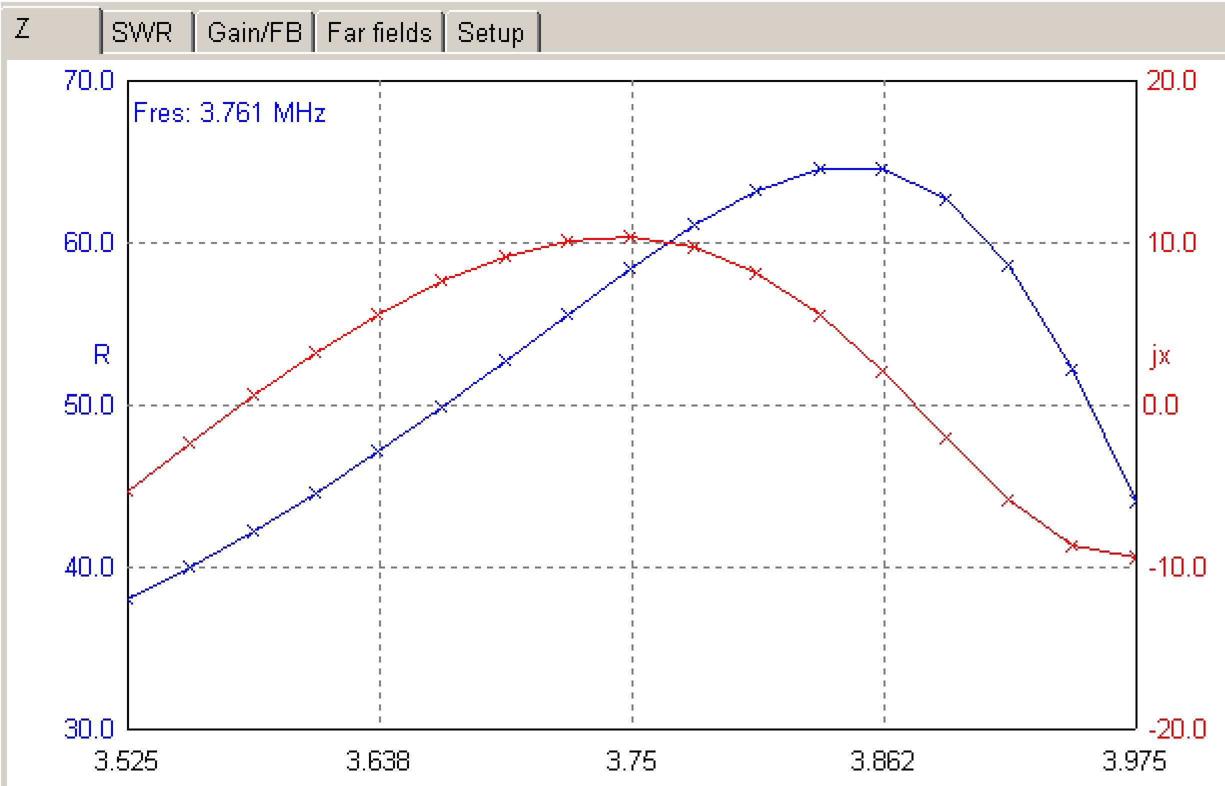


Figure 2 Z of the Simplified DEWD Vertical

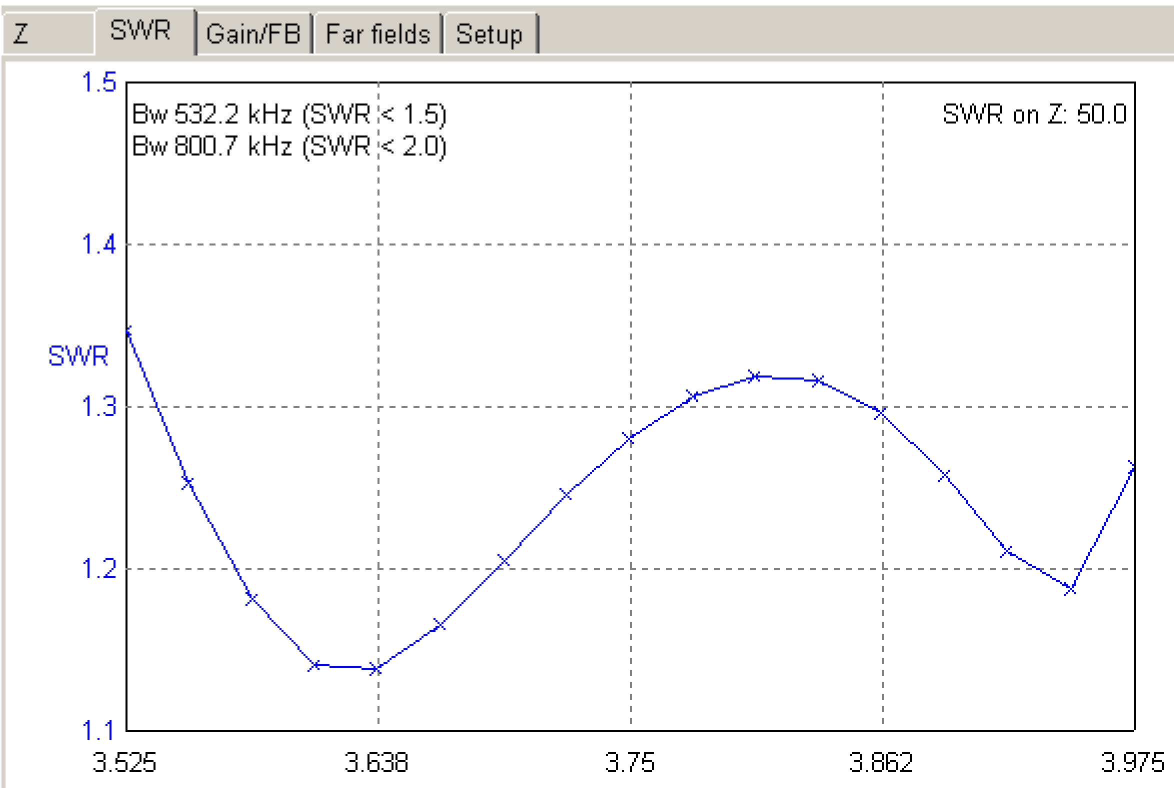


Figure 3 SWR of the Simplified DEWD Vertical

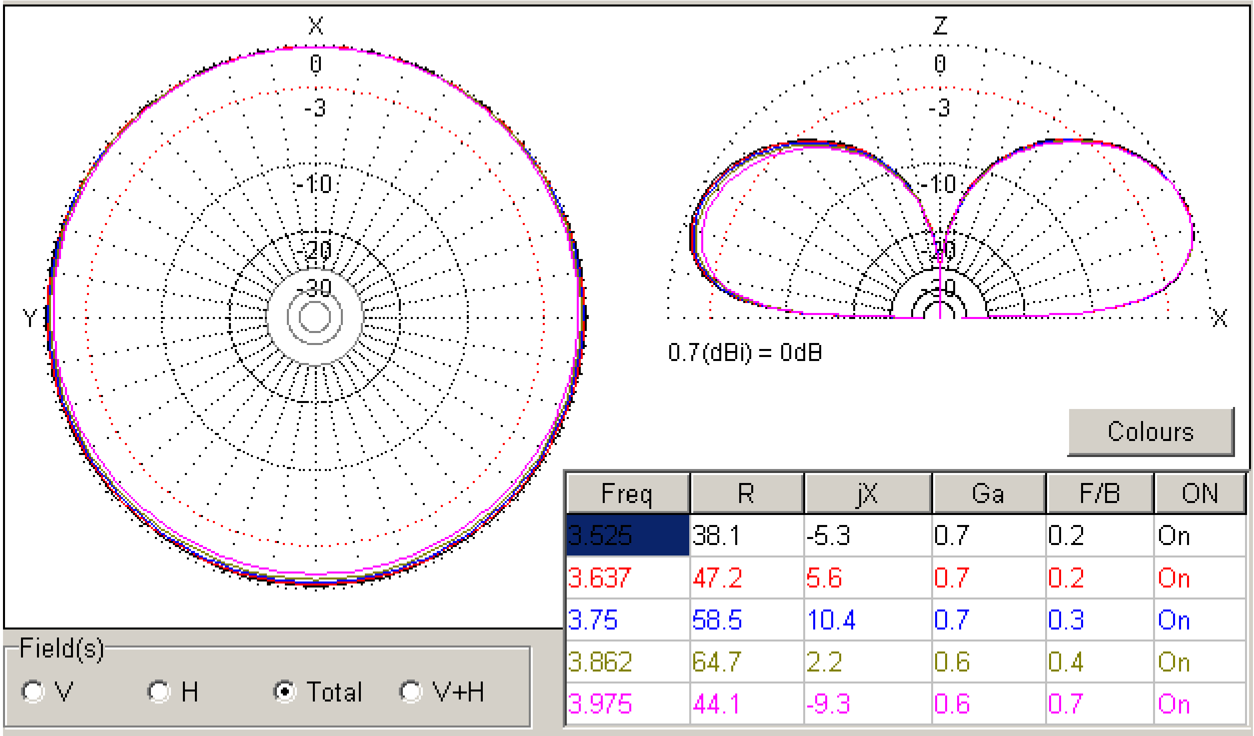


Figure 4 DD of the Simplified DEWD Vertical

Model of the “Real” DEWD Vertical with counterpoises (file *DV3.5_4MHz.2.maa*) placed at 1-meter above ground may be loaded from: http: // www.antentop.org/011/vertical_011.htm

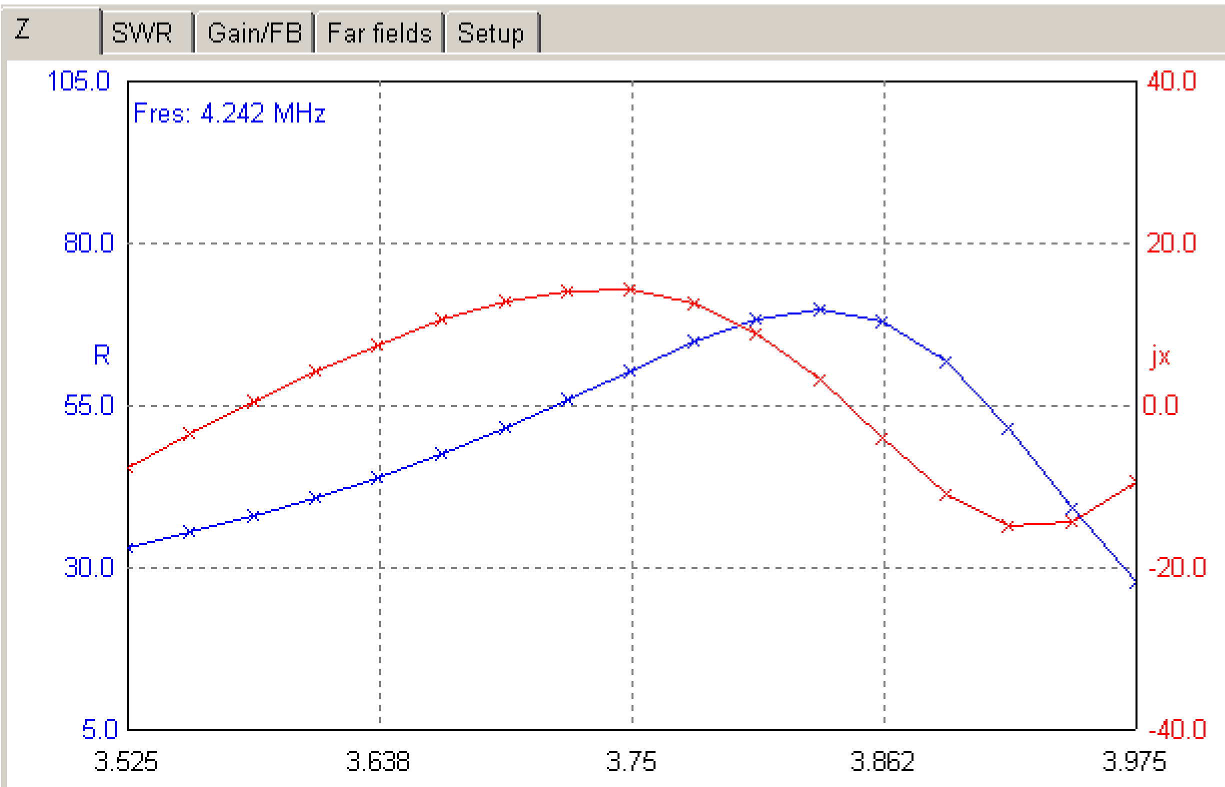


Figure 6 Z of the Real DEWD Vertical

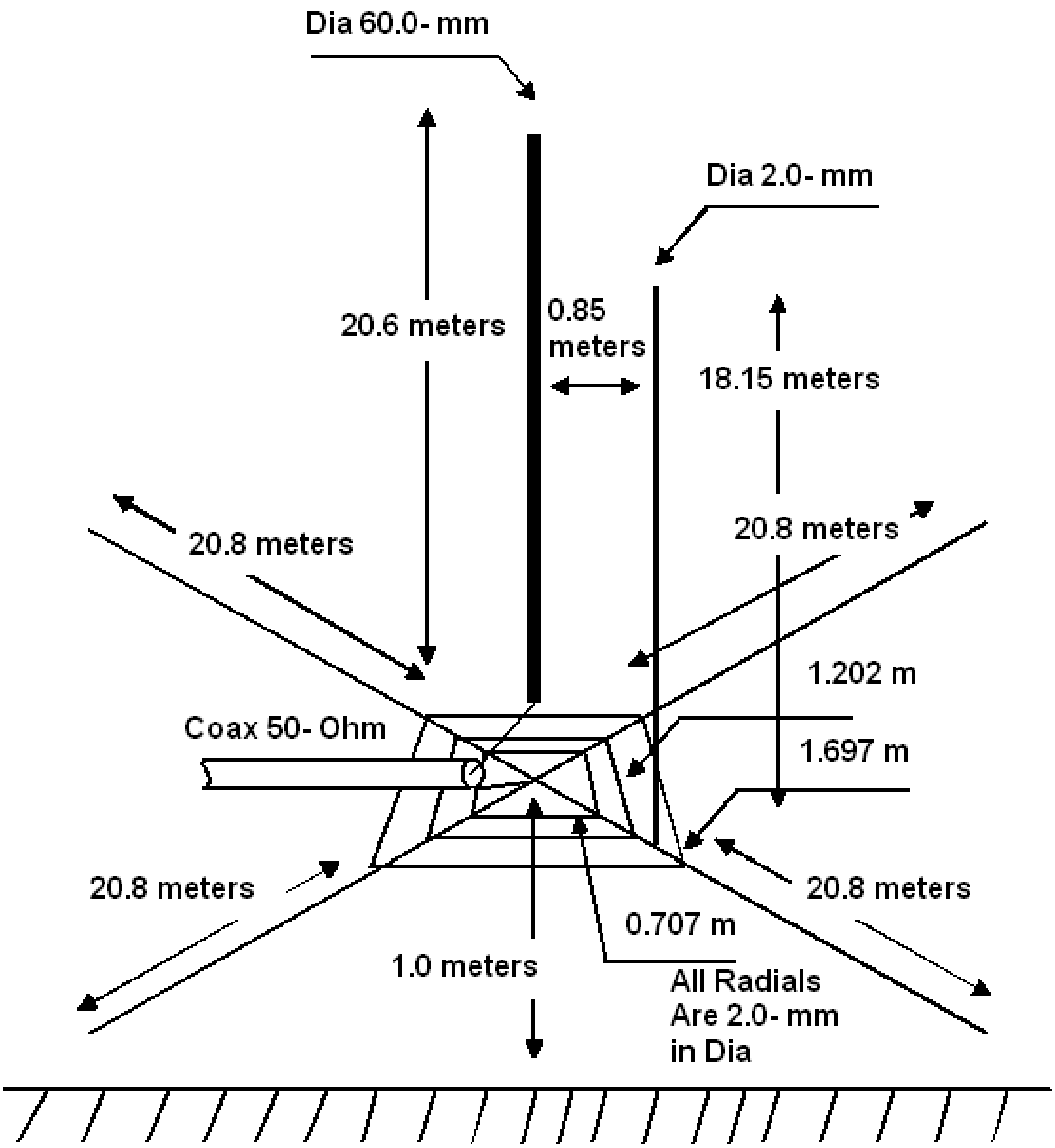
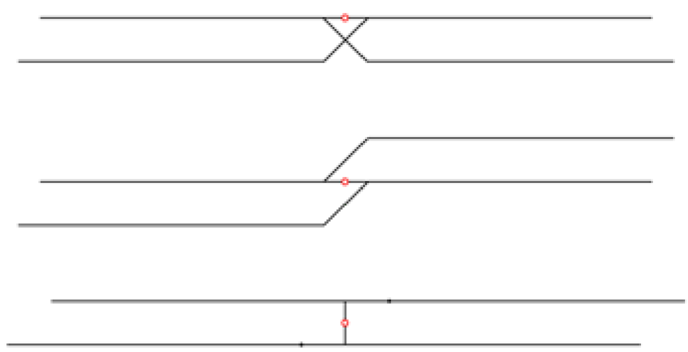


Figure 5 Real DEWD Vertical for the 80- meters



The Figure shows turning LPDA into DEWD. As a result of it, DD is changed to circular, cross line became a part of the DEWD antenna.

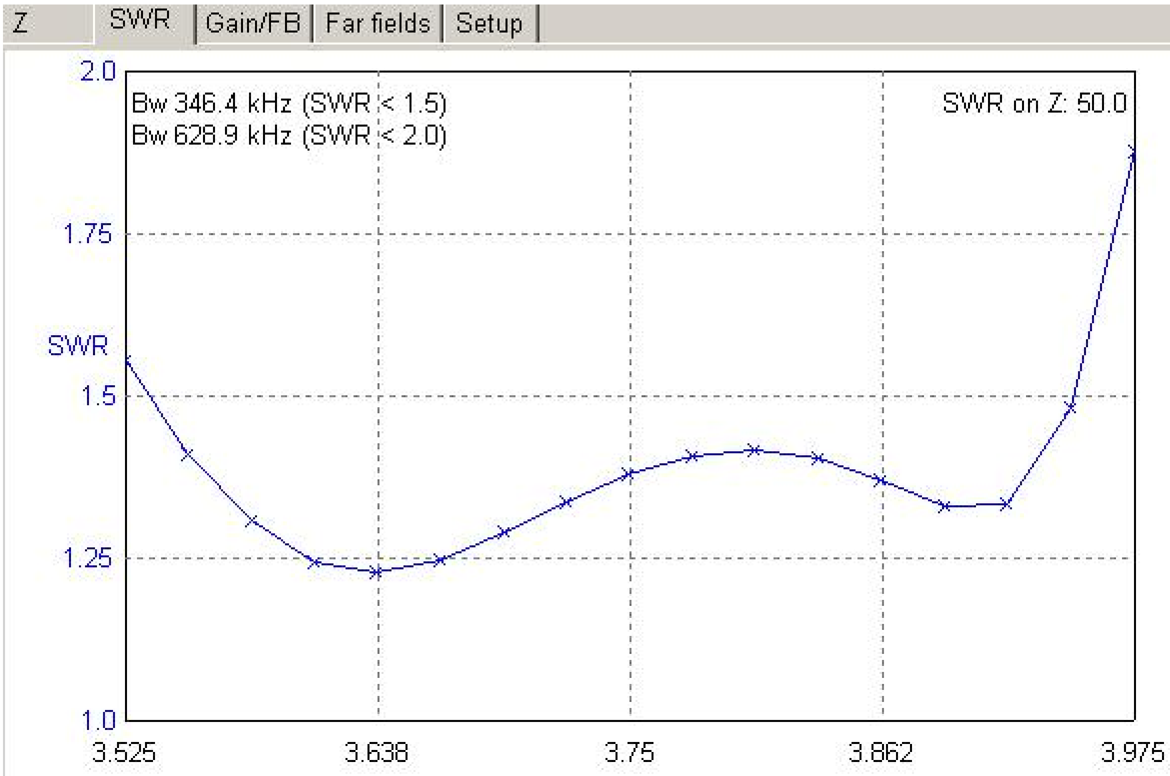
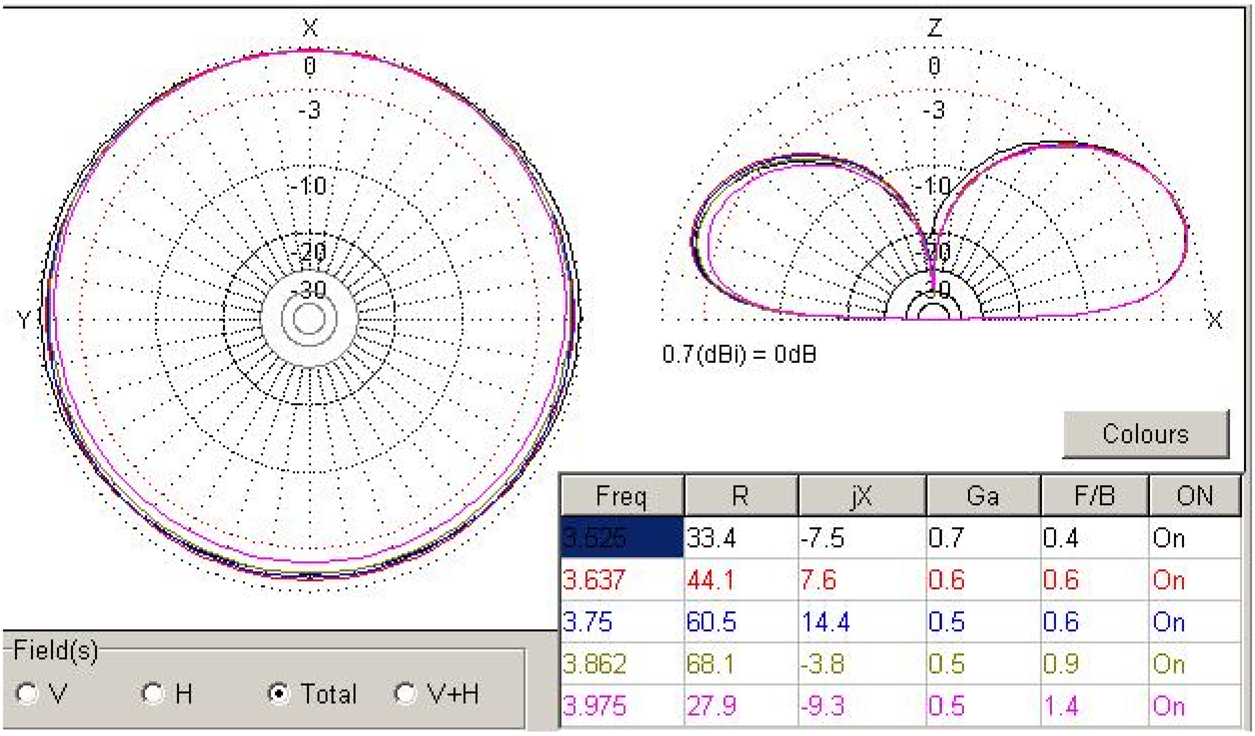


Figure 7 SWR of the Real DEWD Vertical



Freq	R	jX	Ga	F/B	ON
3.525	33.4	-7.5	0.7	0.4	On
3.637	44.1	7.6	0.6	0.6	On
3.75	60.5	14.4	0.5	0.6	On
3.862	68.1	-3.8	0.5	0.9	On
3.975	27.9	-9.3	0.5	1.4	On

Figure 8 DD of the Real DEWD Vertical

73 Nick

Credit Line: Forum from: www.cqham.ru

DEWD Dipole Antenna for the 80- meters

The publications devoted to memory UR0GT.

***By:* Nikolay Kudryavchenko, UR0GT**

The antenna has SWR less the 1.2:1.0 in the band 3.5- 3.8- MHz.

Model of the antenna (file: Broadband 80m.maa) may be loaded:

http: // www.antentop.org/011/dewd_dipole_011.htm

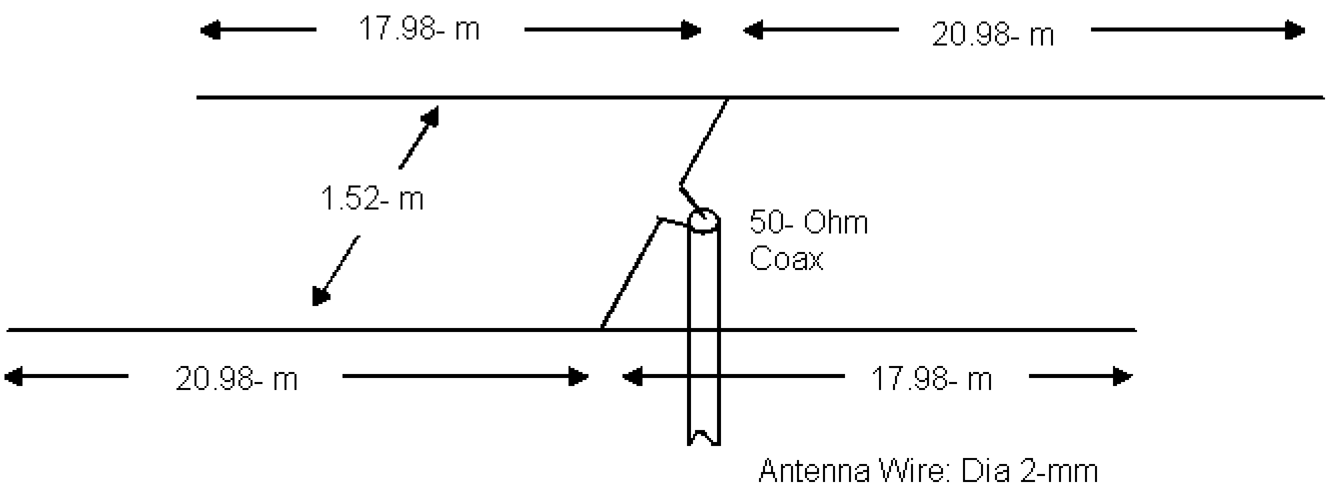


Figure 1 DEWD Dipole Antenna for the 80- meters

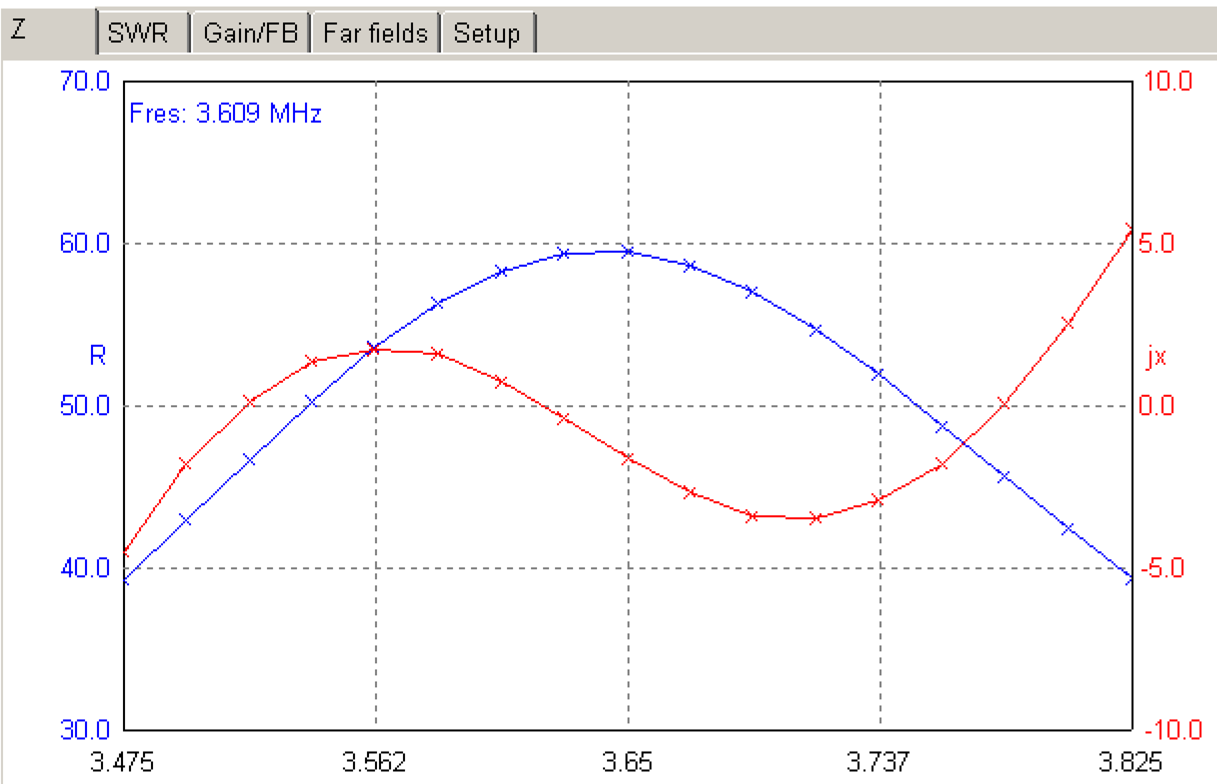


Figure 2 Z of the DEWD Dipole Antenna (placed in the free space)

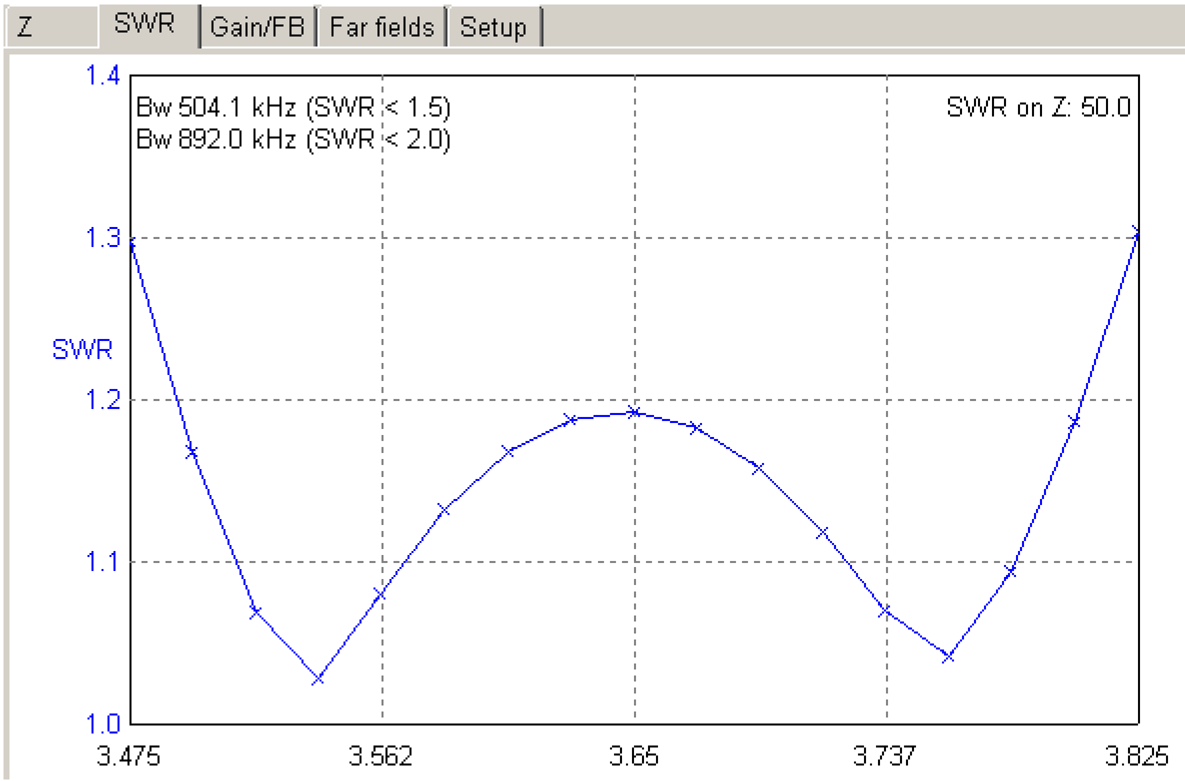


Figure 3 SWR of the DEWD Dipole Antenna (placed in the free space)

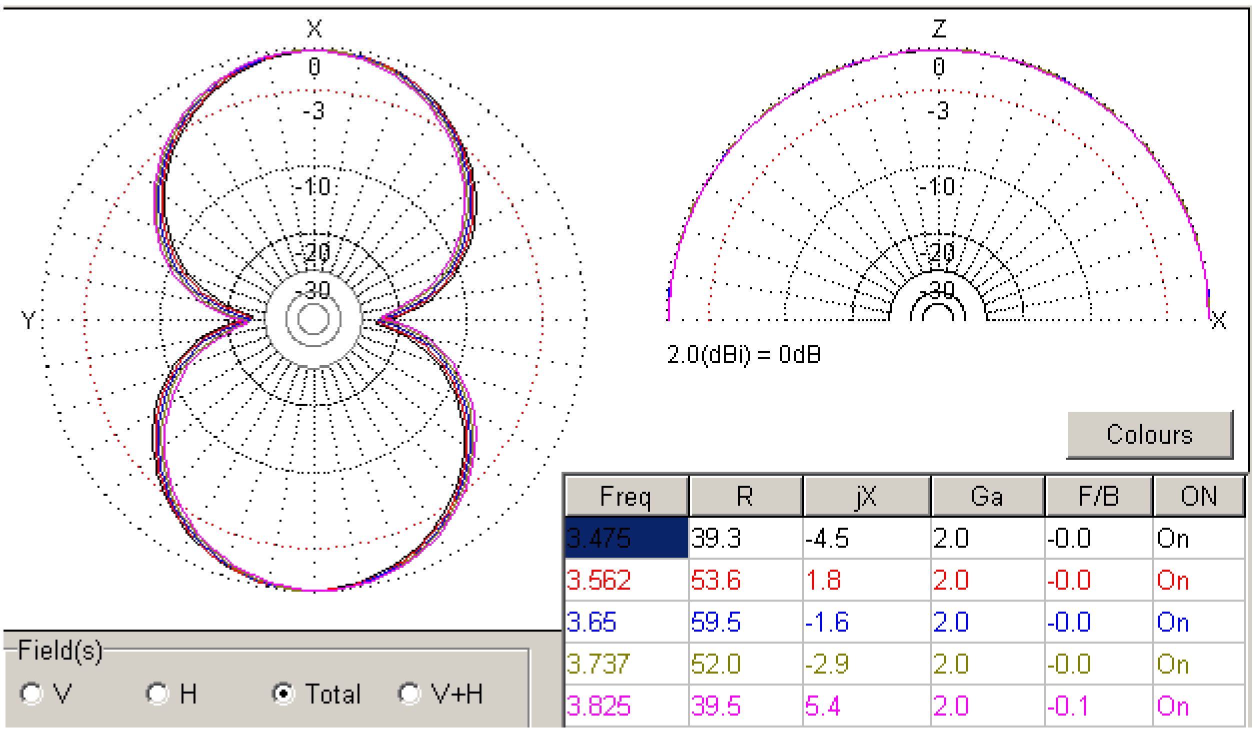


Figure 4 DD of the DEWD Dipole Antenna (placed in the free space)

Model for a shortened DEWD Dipole Antenna (file: Broadband 80m.1.maa) may be loaded:

http: // www.antentop.org/011/dewd_dipole_011.htm

Tuning of the antenna: To move the resonance needs to change the length of the "short" and "long" parts of the antenna. To shift the reactance- change the width of the antenna.

73, UR0GT, Nikolay

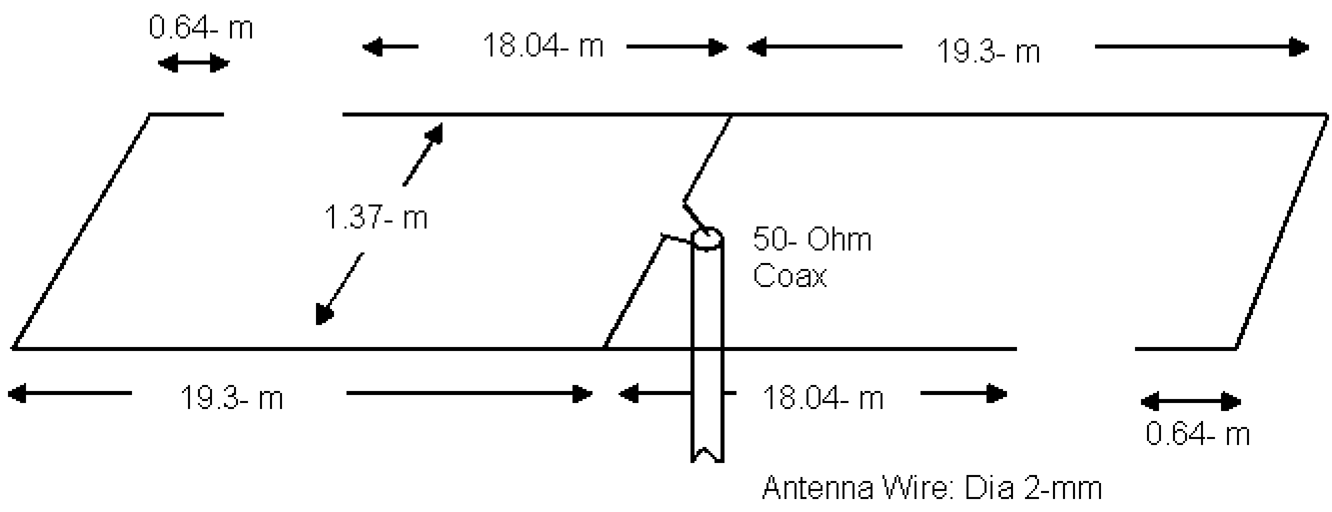


Figure 5 Shortened DEWD Dipole Antenna for the 80- meters

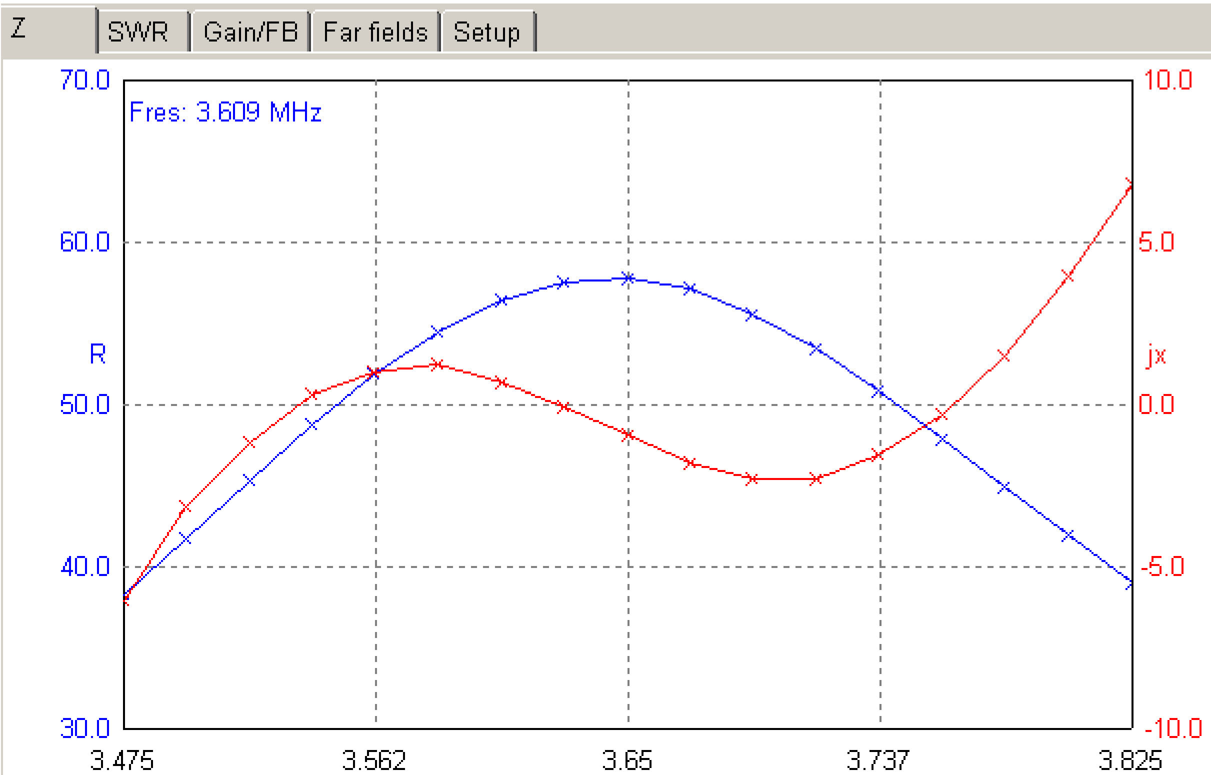


Figure 6 Z of the Shortened DEWD Dipole Antenna (placed in the free space)

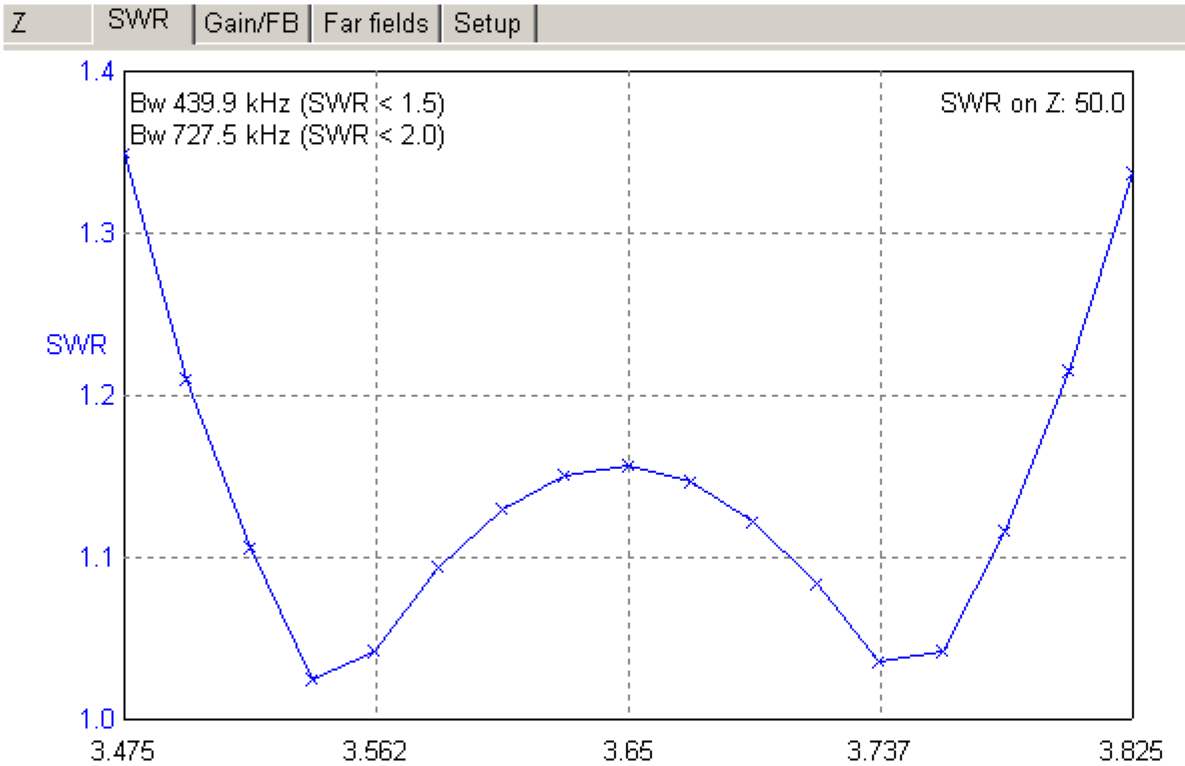


Figure 7 SWR of the Shortened DEWD Dipole Antenna (placed in the free space)

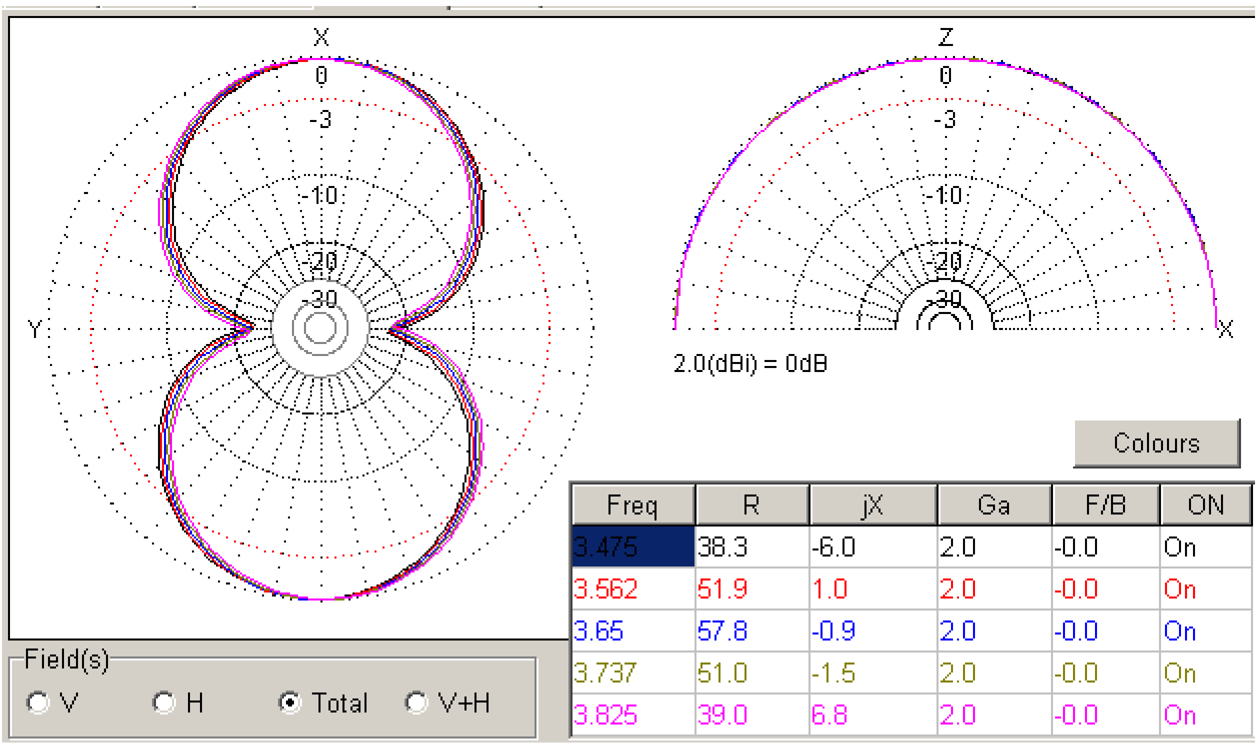


Figure 8 DD of the Shortened DEWD Dipole Antenna (placed in the free space)

DEWD I.V. for the 80- meters

The publications devoted to memory UR0GT.

By: Nikolay Kudryavchenko, UR0GT

The model of the DEWD IV antenna (file Broadband IV 80m. maa) in the free space may be loaded: http: // www.antentop.org/011/dewd_iv_011.htm

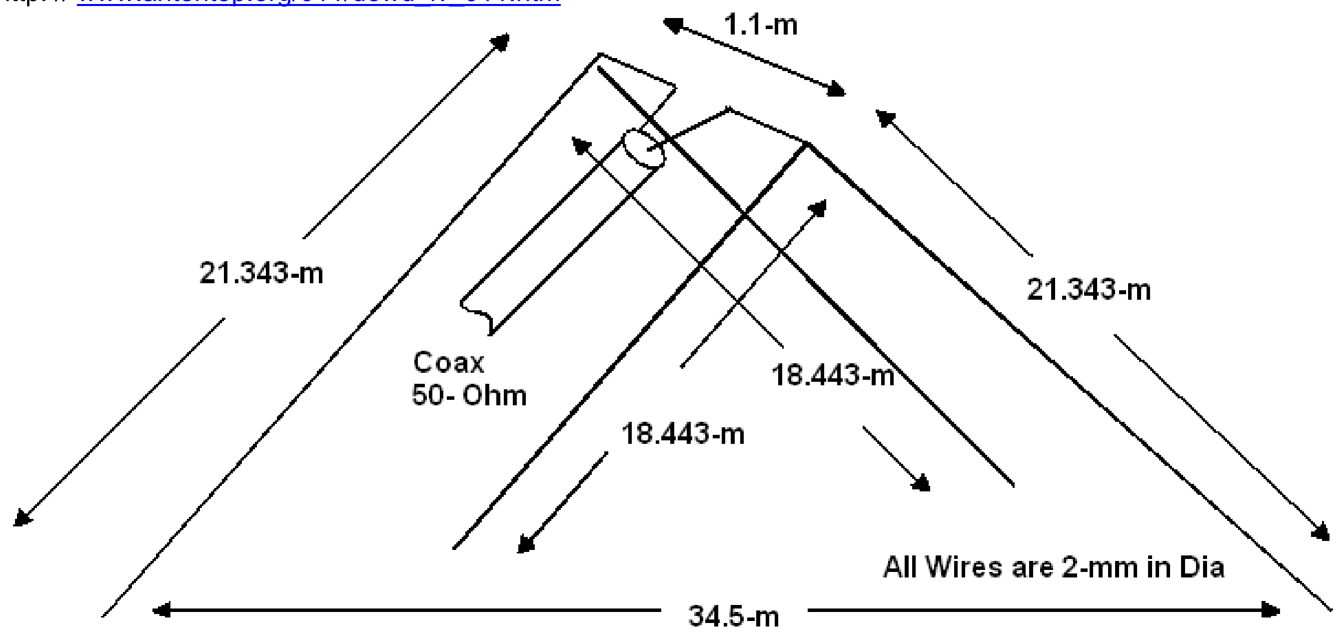


Figure 1 DEWD IV for the 80- meters

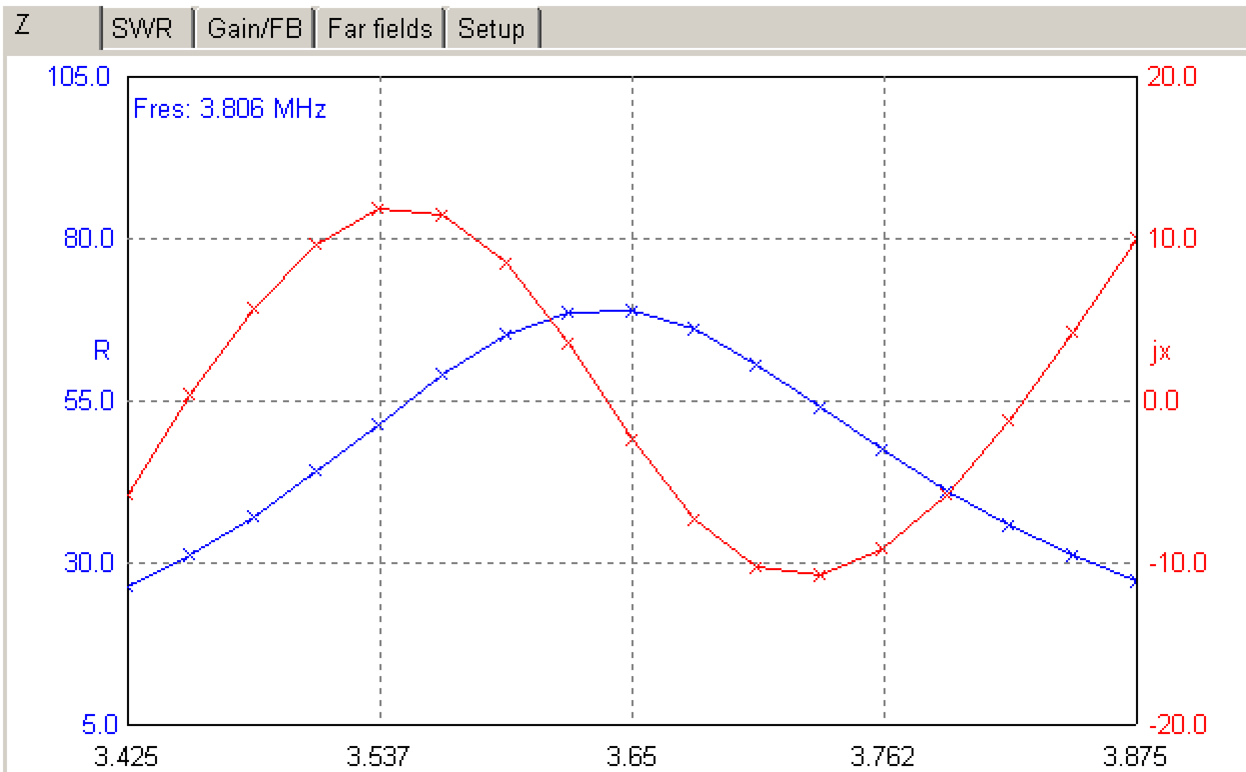


Figure 2 Z of the DEWD IV for the 80- meters (placed in the free space)

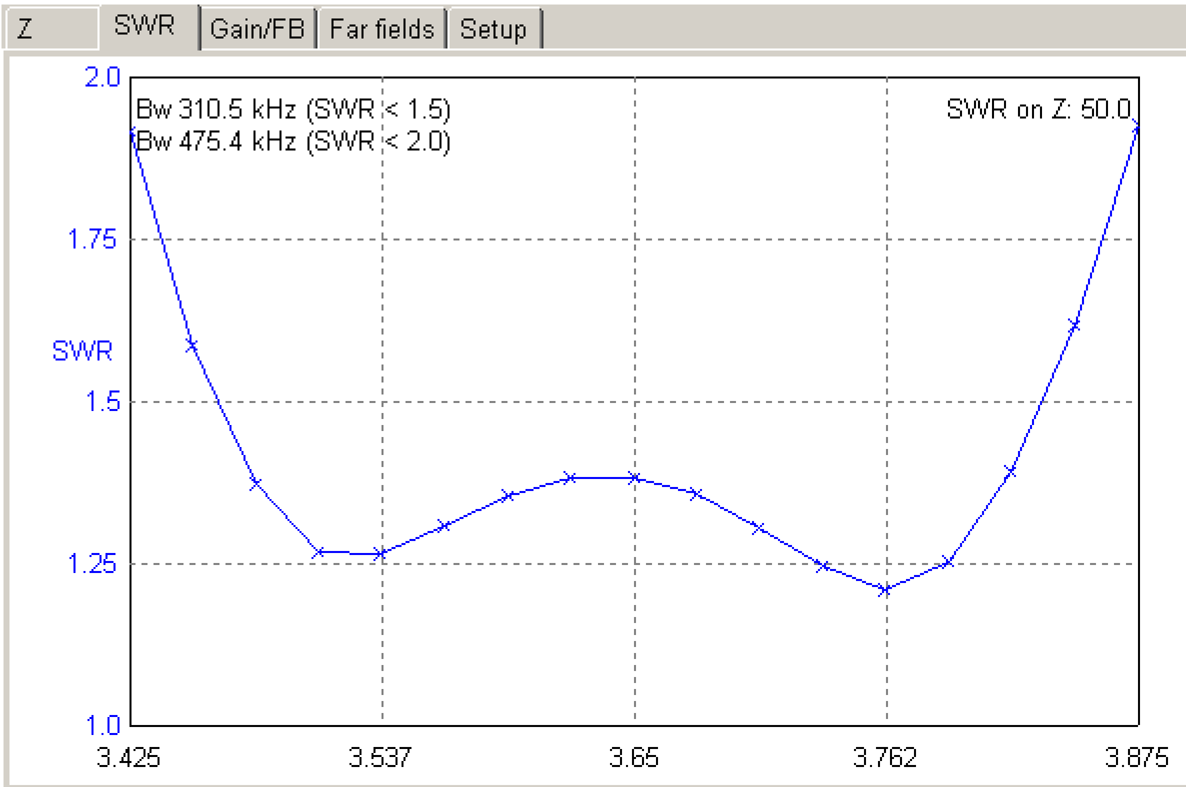


Figure 3 SWR of the DEWD IV for the 80- meters (placed in the free space)

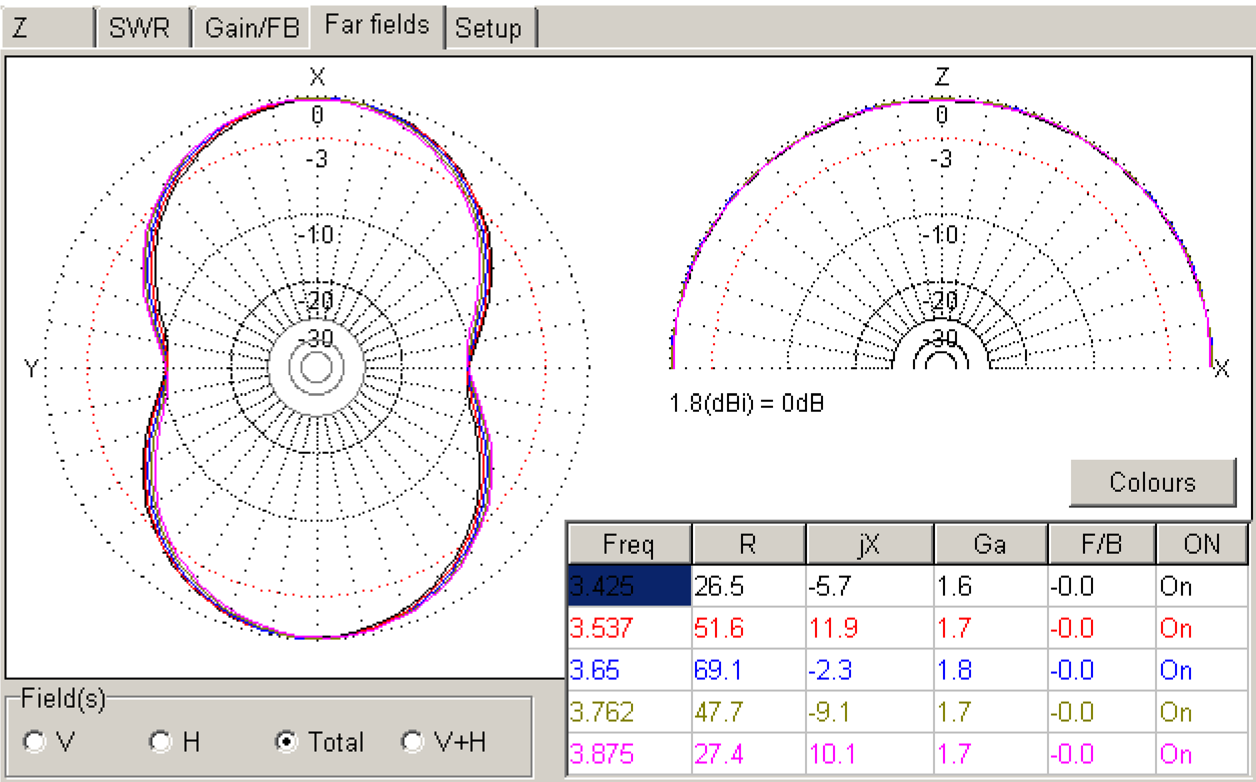


Figure 4 DD of the DEWD IV for the 80- meters (placed in the free space)

The antenna may be optimized. At optimization it needs to get the same curves of R and X (like for antenna from **Figure 1**) that are shown at **Figure 2**.

The antenna has three resonance (where X=0) at its band-pass. Interval between the resonance frequencies grows (and R grows) when the difference between “short” and “long” wires is grows up.

X-curve goes up (down) at increasing (decreasing) the width of the antenna. The model of the optimized DEWD IV antenna (file Broadband IV 80m.1. maa) in the free space may be loaded
http: // www.antentop.org/011/dewd_iv_011.htm

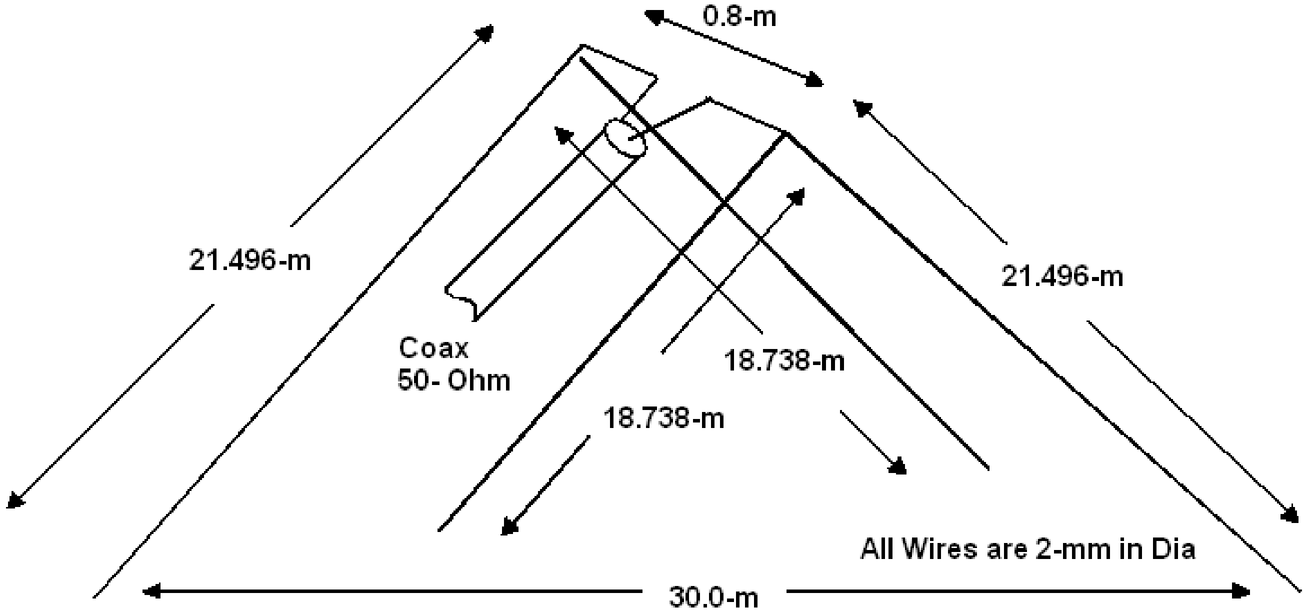


Figure 5 Optimized DEWD IV for the 80- meters

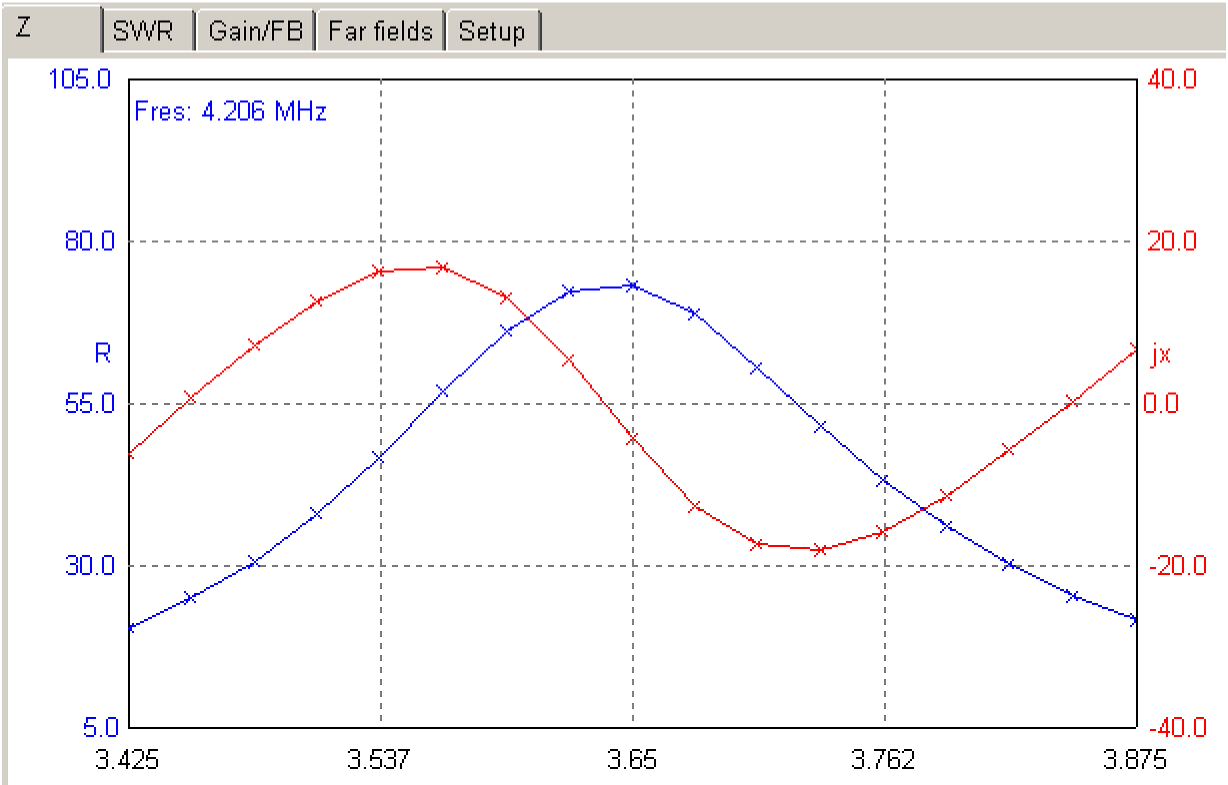


Figure 6 Z of the optimized DEWD IV for the 80- meters (placed in the free space)

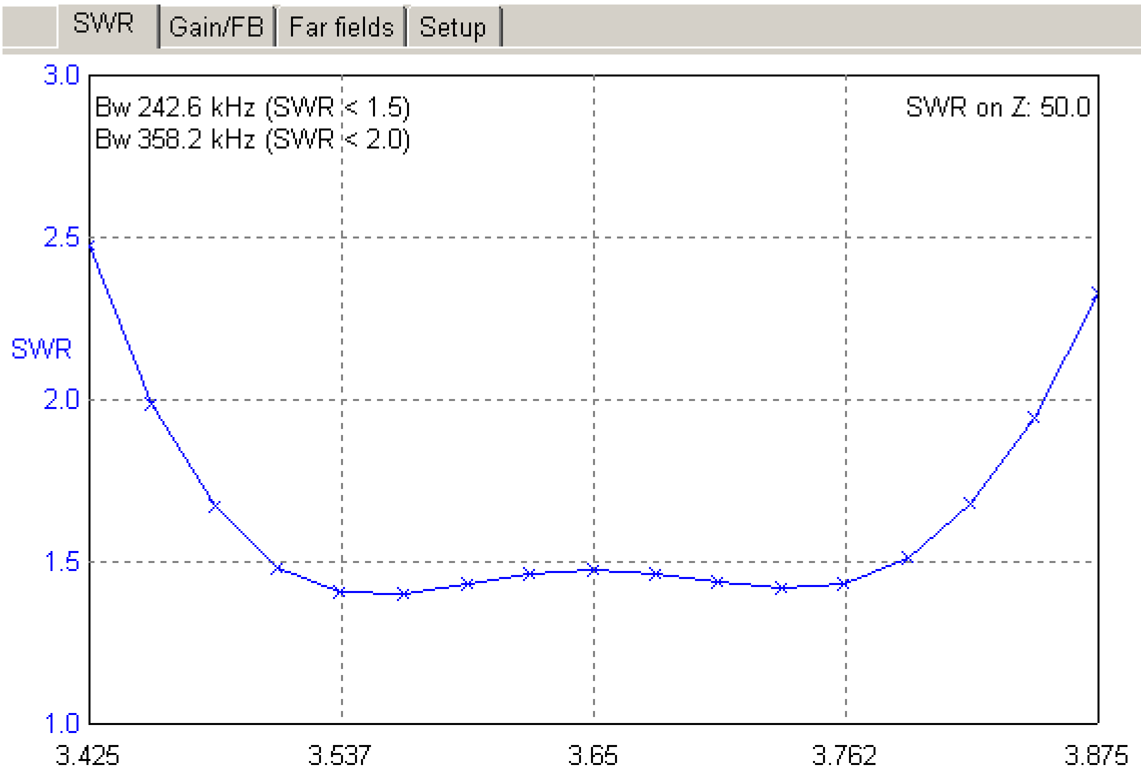


Figure 7 SWR of the optimized DEWD IV for the 80- meters (placed in the free space)

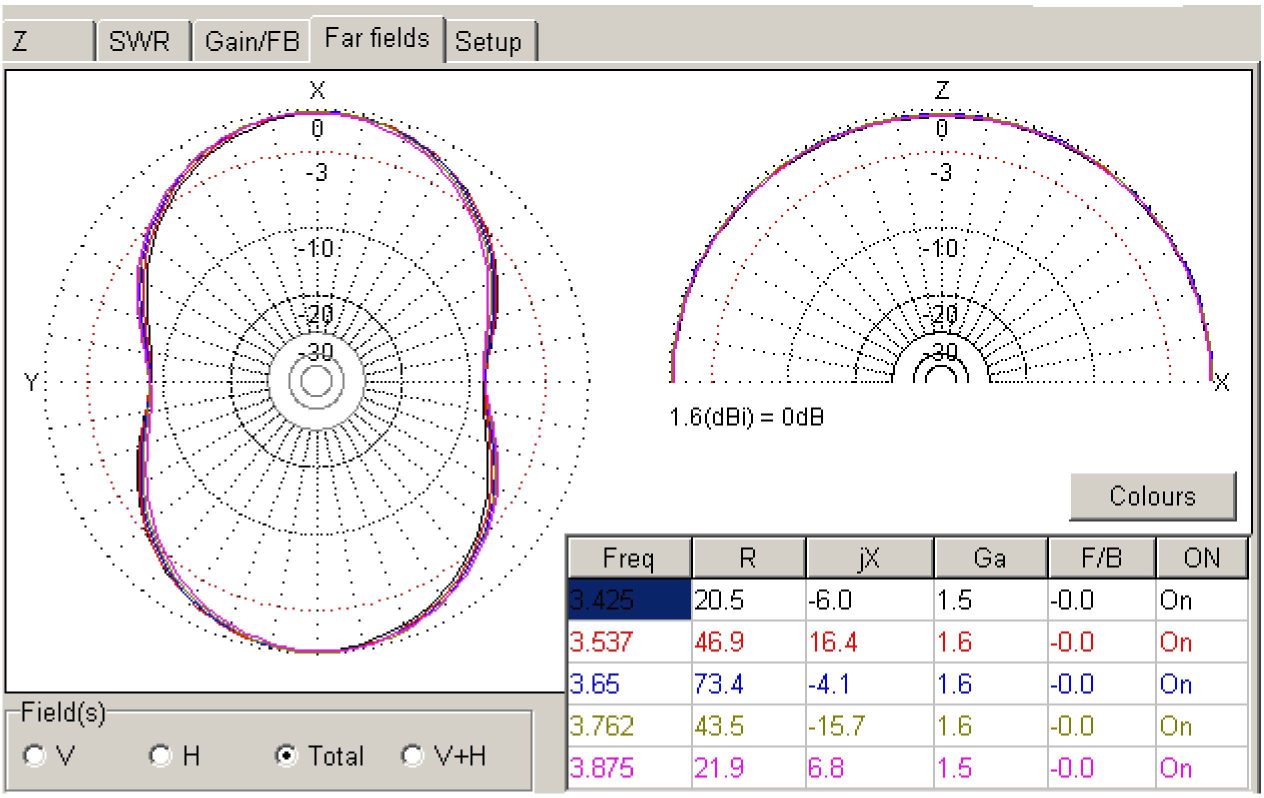


Figure 8 DD of the optimized DEWD IV for the 80- meters (placed in the free space)

73, Nikolay

Credit Line: Forum from: www.cqham.ru

DEWD Broadband YAGI Antenna for the 80- meters

The publications devoted to memory UR0GT.

By: Nikolay Kudryavchenko, UR0GT

The model of the antenna (file 4el_Broadband 80m_2.maa) may be loaded:

http: // www.antentop.org/011/yagi_dewd_011.htm

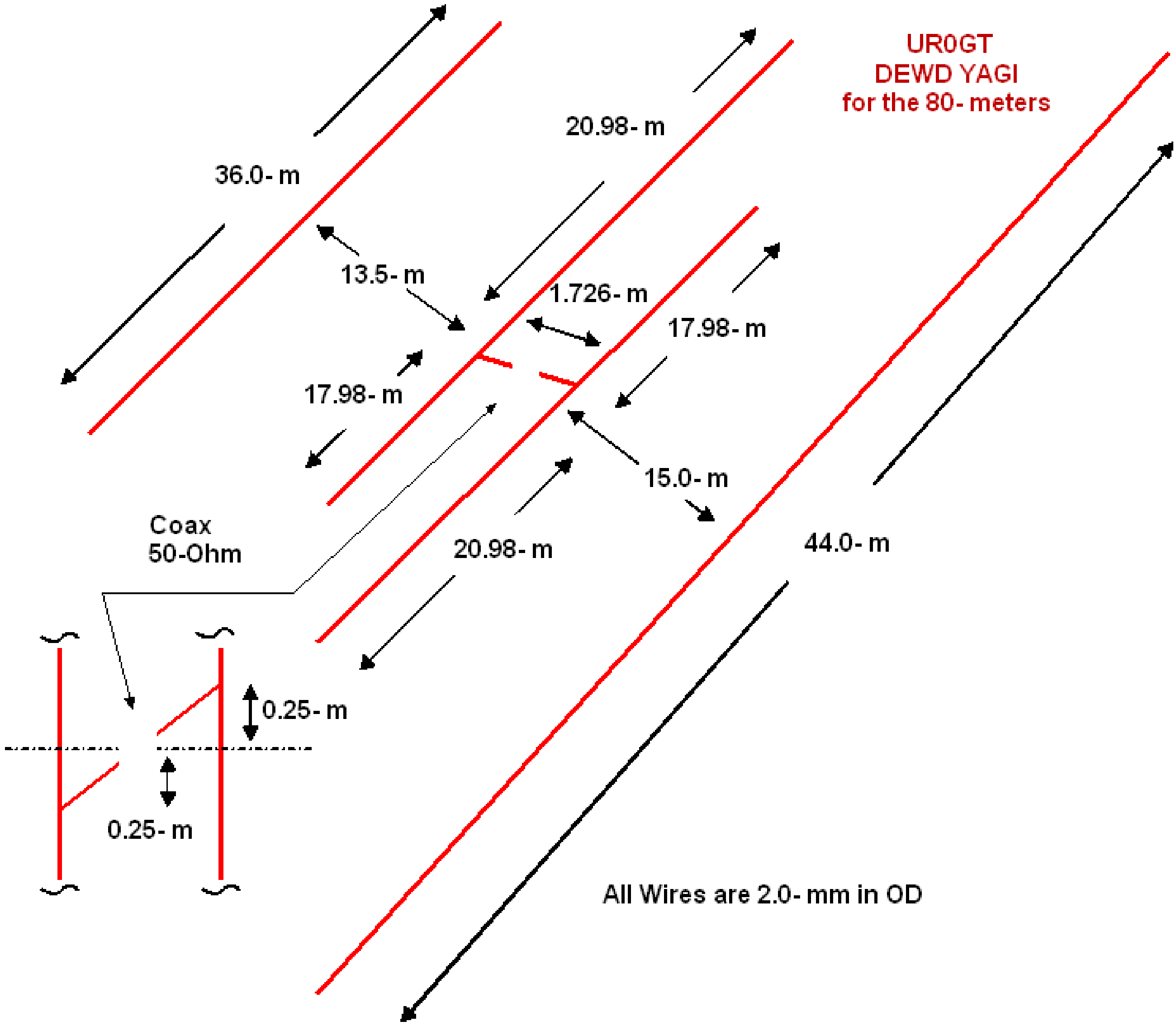


Figure 1 DEWD Broadband YAGI Antenna for the 80- meters

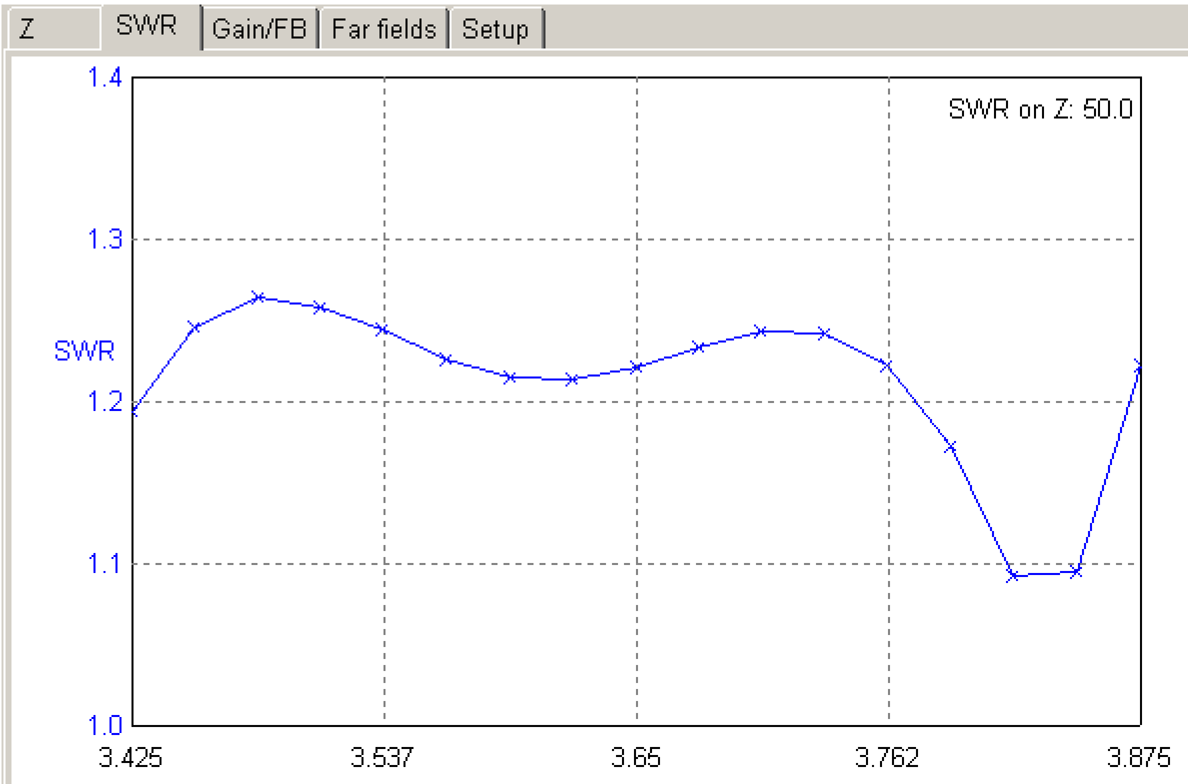
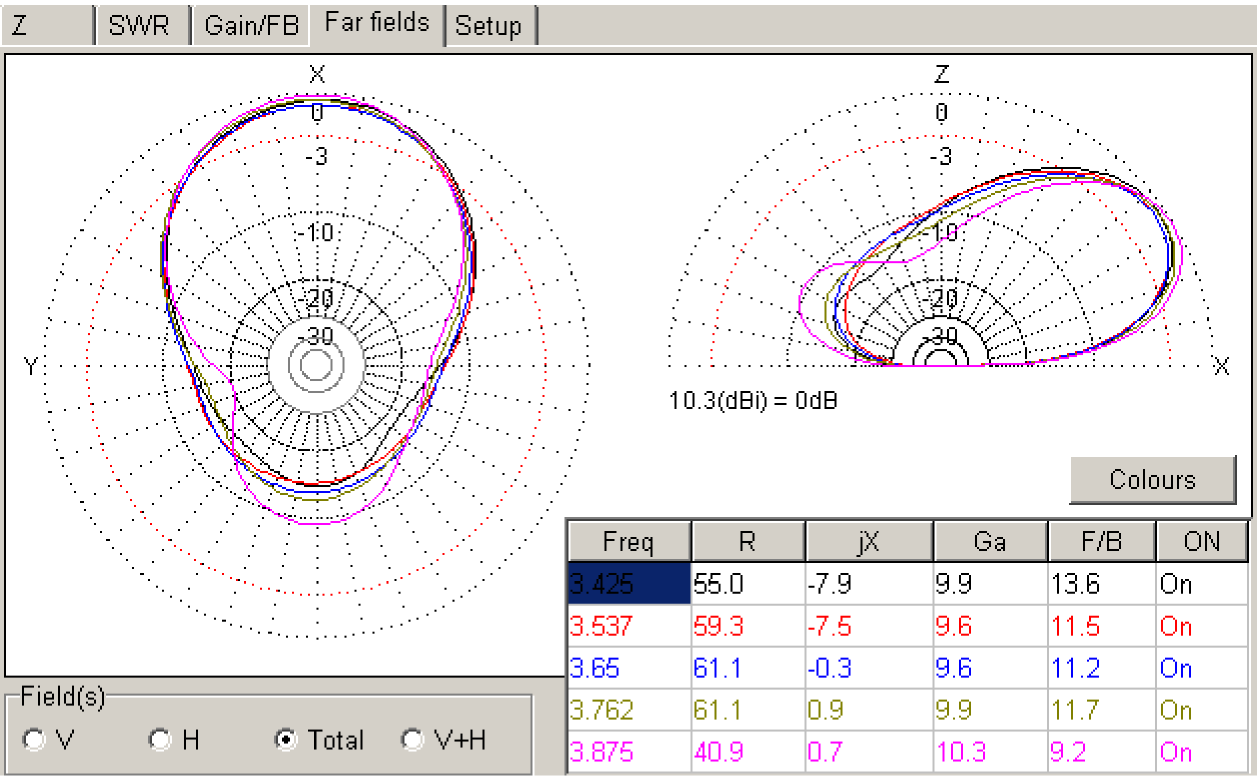


Figure 2 SWR of the DEWD Broadband YAGI Antenna (placed at 30 meters above real ground)



Freq	R	jX	Ga	F/B	ON
3.425	55.0	-7.9	9.9	13.6	On
3.537	59.3	-7.5	9.6	11.5	On
3.65	61.1	-0.3	9.6	11.2	On
3.762	61.1	0.9	9.9	11.7	On
3.875	40.9	0.7	10.3	9.2	On

Figure 3 DD of the DEWD Broadband YAGI Antenna (placed at 30 meters above real ground)

Credit Line: Forum from:
www.cqham.ru

Top Fed Five Band Vertical Antenna

By: **Yuri Medinets, UB5UG, Kiev**

Credit Line: Radio #1, 1984, p.24

It is a simple solution increase the efficiency of a vertical antenna. Just to remove all high currents parts of the antenna from ground. The losses at nearby subjects are decrease and the overall performance of the antenna increase. So, we come to a Top Fed Vertical Antenna.

Figure 1 shows one band Top Fed Vertical Antenna. Upper horizon part of the antenna stretched with help of a fishing rope by 1.0-mm in diameter.

To the middle of the horizon wire the inner conductor of the coaxial cable is connected. The outer sheath of the coaxial is unconnected. Coaxial cable is going straight down or with some angle to the horizon wire. An RF-Choke is installed at the distance lambda/4 from the top end of the coaxil. The RF-Choke made like a resonant LC-circuit. The circuit has resonant at the working frequency of the antenna.

RF- Choke practically made by coiling around of one side of a ferrite ring (having low losses at HF) with large OD/ID one or two turns by the antenna coaxial cable. A Resonance Coil coiled at another side of the ring. The circuit (coil and bridged to it capacitor) has resonance at the working frequency.

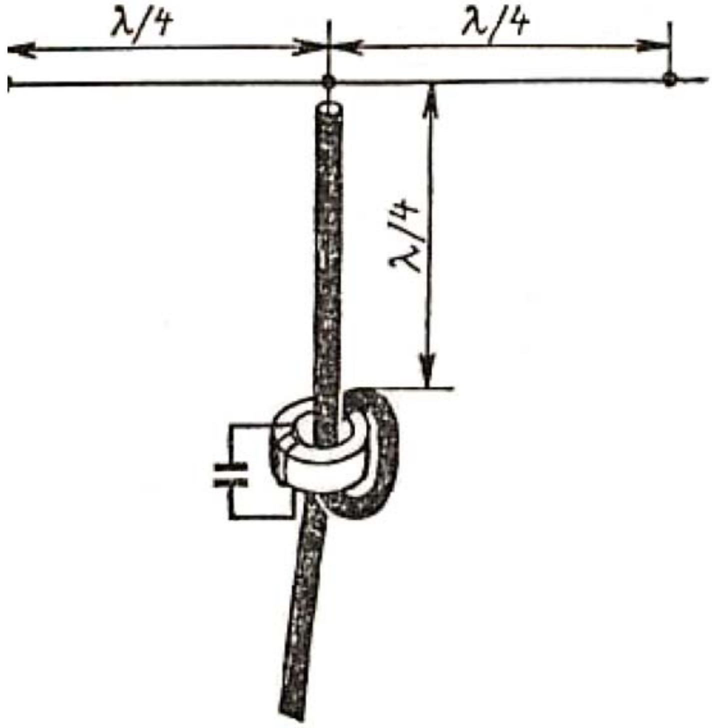


Figure 1 One band Top Fed Vertical Antenna

The Choke may be made without a ferrite ring. Choke looks like two- three large turns of wire in 2-4- mm in diameter bridged to capacitor. Coaxial cable in the needed place does one turn around the coil. The coax and the coil are tight by a Scotch. Capacitor should have high Q and high working voltage. TX with 100- Wtts output power may induce across the capacitor 400-500- V RF.

Data (approximately) for the Choke (made by wire in 2-mm in diameter) are given in **Table 1**.

The Choke does not influence to the coaxial going below the Choke. Below the Choke coaxial (any length) may go in any way- lay on the ground or roof, go near a home wall, etc.

It is very possible to create a multi- band Top Fed Choke Antenna. Use vertical wires that have resonance for each band and just place at needed places on to coaxial cable the special resonance Chokes. **Figure 2** shows the schematic for 5-band Top Fed Choke Antenna.

Table 1 Data for Loop Choke

Band, MHz	Numbers of turns	Length of the wire, cm	C, PF	Q
7	3	160	150	260
14	2	115	68	230
21	2	80	47	210
28	2	60	36	265

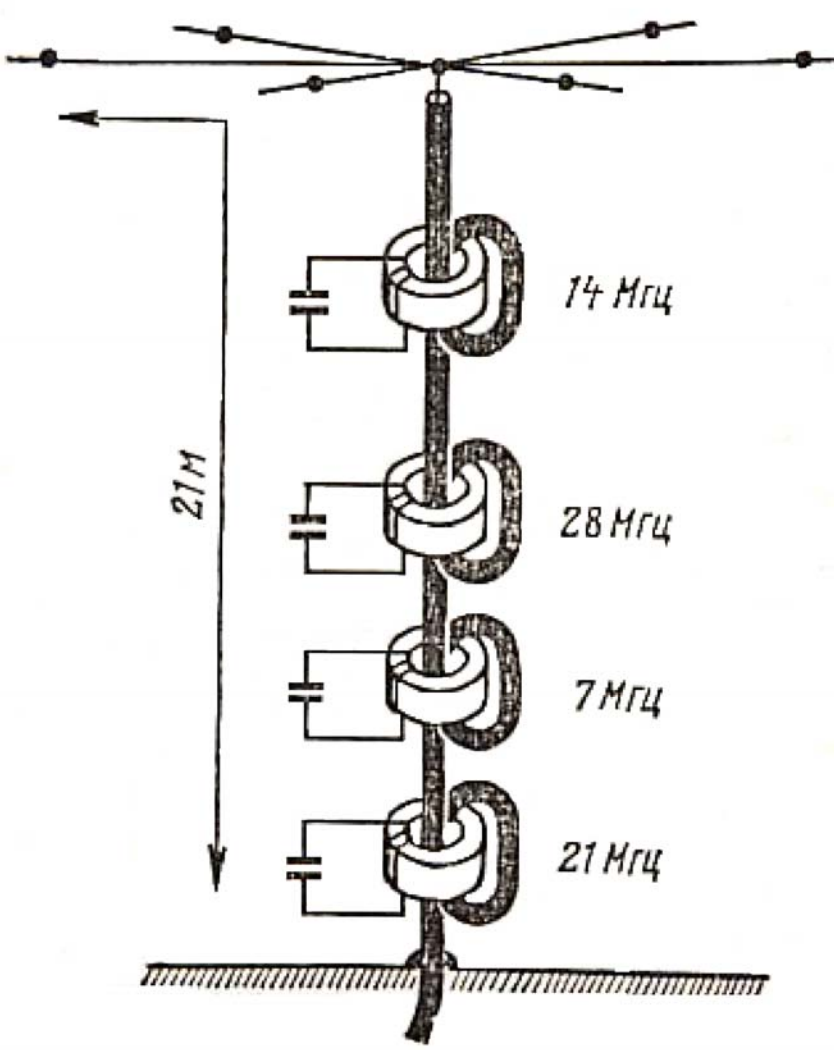


Figure 2 Top Fed Five Band Antenna

Horizon part of the antenna has three pair wires in length 10- m (for 7 and 21- MHz), 5- meters (for 14- MHz), 2.5- meters (for 28- MHz). Chokes are placed (from the Top End): 7.5- meters for 28- MHz, 3.5 or 10.5- meters for 21- MHz, 5- meters for 14- MHz, 10- meters for 7- MHz. The sizes are approximately and should be corrected at antenna tuning.

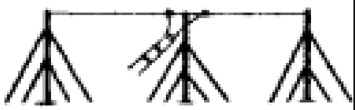
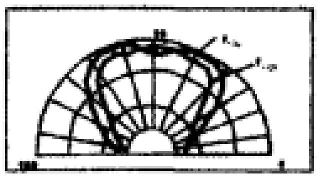
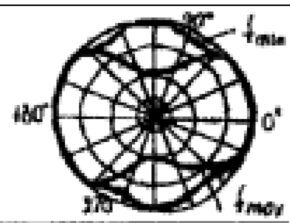
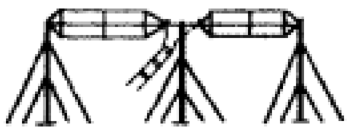
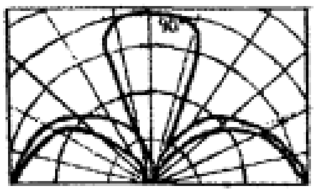
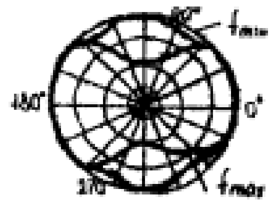
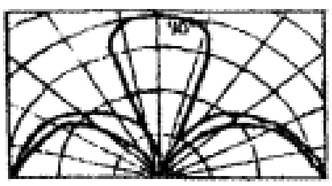
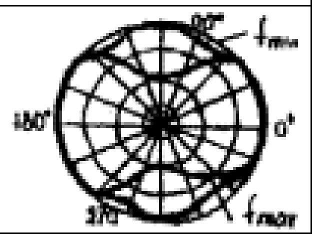
For band 3.5- MHz the outer sheath of the coaxial is grounded to the ground after last Choke.

The distance is approximately and should be corrected at antenna tuning. It is possible to tune the antenna at 3.5- MHz by the coil or capacity that is switching on between the outer sheath of the coaxial and the ground.

The input impedance of the antenna at the bands 7- 14- 21- 28- MHz is close to 50- Ohm. The input impedance of the antenna at the band of 3.5- MHz is close to 100- Ohm.

Russian Military Antennas Some Data

Credit Line: www.spvvius.ru

Antenna	Band, MHz	Communication Range	Comment
Horizon Dipole	0.1-20	0-60 200-2000	Fixed Antenna for Middle-Long Communication Range HF- Transmitters
ВГД 2x1	Design	DD in Vertical Plane	DD in Horizon Plane
			
Antenna	Band, MHz	Communication Range	Comment
Horizon Dipole	0.1-20	0-60 200-2000	Fixed Antenna for Middle-Long Communication Range HF- Transmitters
ВГД	Design	DD in Vertical Plane	DD in Horizon Plane
			
Antenna	Band, MHz	Communication Range	Comment
Horizon Dipole	0.1-20	0-60 200-2000	Fixed Antenna for Middle-Long Communication Range HF- Transmitters
ВГДШ	Design	DD in Vertical Plane	DD in Horizon Plane
			

To be continued…

Small Sized Helical Antennas

***By*: I.Kapustin, UA0RW, Yakutsk, USSR**

Small-Sized Helical Antenna – it is wire helix coiled on a dielectric rod. UA0RW used a wood rod boiled in the paraffin. **Figure 1** shows the design of the Vertical Small Sized Helical Antenna. Antenna for the 20- meters has rod in OD 2- cm, length of winding is 30 cm, antenna is coiled turn to turn) by enamel wire in 1- mm OD (18- AWG).

Credit Line: Radio #1, 1958, pp.:26-27

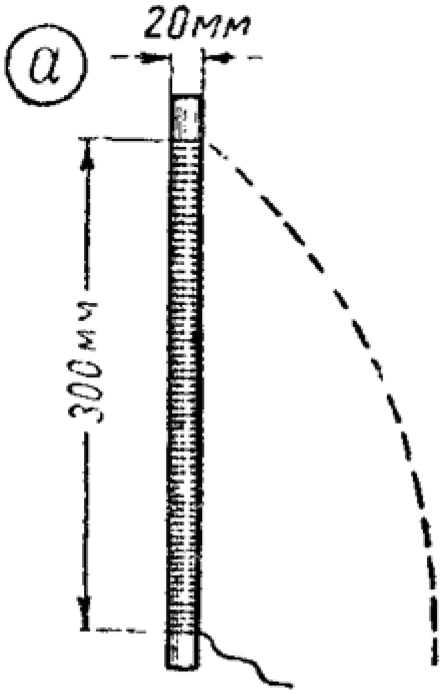


Figure 1
A. Design of the Vertical Small Sized Helical Antenna

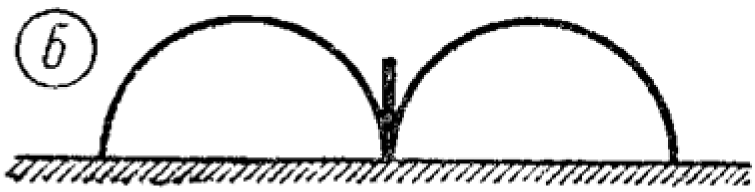


Figure 1
B. DD of the Vertical Small Sized Helical Antenna

Figure 2 shows the design of the Dipole Small Sized Helical Antenna. Antenna for the 20- meters has rod in OD 2- cm, each parts of the dipole has 220 coils of the enamel wire in 1- mm OD (18- AWG). Overall length of the two parts of the antenna is 60- cm.

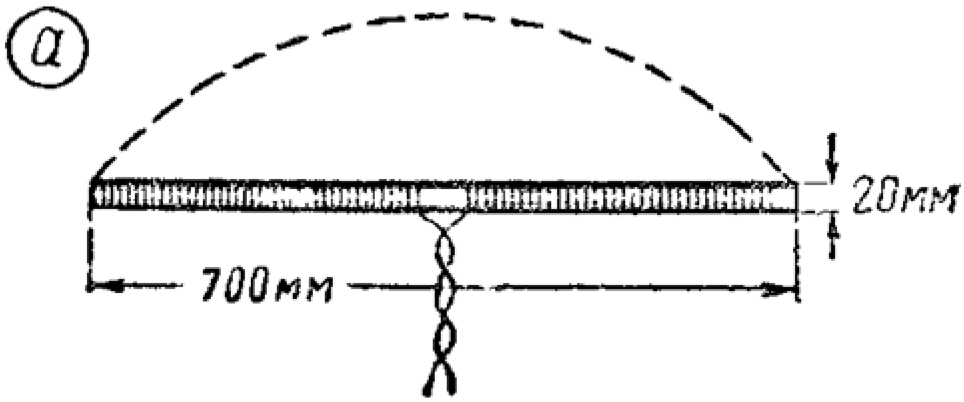


Figure 2
A. Dipole Small Sized Helical Antenna.

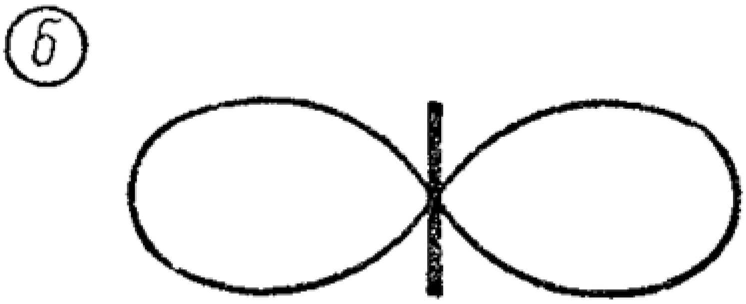


Figure 2
B. DD of the Dipole Small Sized Helical Antenna

At the test the Dipole Small Sized Helical Antenna was installed on the top of a- 5-meters mast. The antenna was fed by usual two- wire main cord. It was discovered, that the antenna has strong radiation directivity. Signals from receiving station almost disappeared when the antenna was turned around. The helical antenna loses 1- point at RS compare to Traveling-Wave Antenna pointed to correspondent.

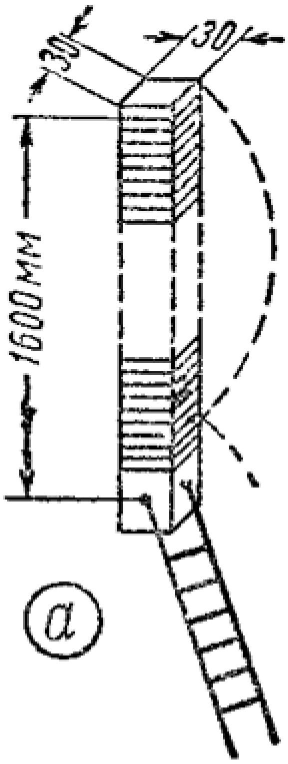


Figure 3
A. Design of the Vertical Small Sized Helical Antenna on a square rod

Note I.G.: At the article there is not explained what the Traveling-Wave Antenna is.

Figure 3 shows the design of the Vertical Small Sized Helical Antenna coiled by copper cord. The cord had 2.5- mm OD. Like a dielectric form for the antenna it was used a rod in length of 1.5- meters. The rod had the square- cross- section with side 3- cm. Antenna had 200 coils. Distance between coils was 7.5- mm.

At the test the antenna (**Fig.3**) was installed at 4.0- meters mast. Antenna was fed by 500- Ohm two- wire symmetrical line. At the antenna terminal one wire of the line was connected to the helical antenna, second wire was lived free. At the transmitter terminal one wire was connected to “Antenna” second one to the “Ground.” Antenna was tested at 20- meters with success.

Practice shows that it is wise to use a square rod for a helical antenna because coils sitting well at the form. Diameter/side of the square should be 1/50- 1/200 from the length of the Helical Antenna. Wires should be protected from atmospheric by strong paint. Wire for winding a helical should be have large diameter as possible but coils of helical should not short to each other.

Current distribution on the antenna may be found with the help of neon bulb. After that the tuning of the antenna was made by the current (checked by antenna milli- ampere- meter) going to the antenna.

Table 1 shows data for the Helical antenna for 20-40-80-160 meters. The antennas were made on a square rod with size 3- cm by wire in 2.5- mm OD. Gap between turns was 7.5- mm.

Table 2 shows data at testing the Helical Antenna compare to a Traveling-Wave Antenna.

Note I.G.: At the article there is not explained what the Traveling-Wave Antenna is.
At the article there is not explained what a Helical Antenna and at which band was tested.

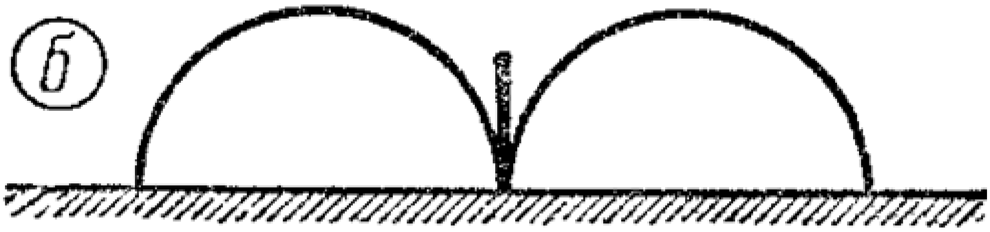


Figure 3
B. DD of the Vertical Small Sized Helical Antenna on a square rod

Table 1

	Numbers of turns for band:			
	20- meters	40- meters	80- meters	160- meters
Antenna from **Fig. 1**	90	180	360	720
Antenna from **Fig. 2**	90	180	360	720
Antenna from **Fig. 3**	200	400	800	1600

Note: For antenna Fig.2 the Data given for one part of the dipole

Table 2

Date	Time, msk	City, Callsign	RST Of the correspondent	RST for UA0RW when is used:	
				Traveling-Wave Antenna	Helical Antenna
22-V-1957	16-45	Sarapul, UA4WA	579	579	579
25-V-1957	16-00	Khabarovsk, UA0CJ	589	579	599
25-V-1957	18-30	Leningrad, UA1KAS	579	579	589
25-V-1957	21-15	Kaliningrad, UA2KAA	579	579	579
26-V-1957	14-00	Irkutsk, UA0SL	5(6/7)9	579	5(8/9)9
26-V-1957	14-45	Stalinabad, UJ8KAA	559	449	559
26-V-1957	16-00	579 Penza, UA4FC	579	569	579

Fire Antenna

By: Vladimir Polaykov, RA3AAE

Credit Line: ***www,qrp.ru***

Once Friday's day I was busy at my cottage with some deals to do. My backyard was pile upped with dry logs and branches. I made a fire in an old rim from a truck. Looking at the smoke going to the night sky, I remembered [1], about smoke that could improve receiving. I had a multimeter M830B and detector receiver described at [2]. So, I decided to make an experiment. I suppose that smoke above fire should collect current from atmosphere, like a broom antenna [3], so, smoke should work like antenna.

The first experiment was simple – I put on the fire a rusty hoop with wire connected to it. The wire was connected to one terminal of the multimeter, the second terminal was connected to the ground. Voltage flows from + 1.0 to - 0.2- V. I put the hoop on the ground- voltage at multimeter became + 0.3- V. I put the hoop to the fire- voltage at multimeter became + 0.1-V. I decided stopped the experiment because I believe the voltage may appear due electro- chemical reaction fire- ground- hoop.

In the next experiment I put an old metal grating above the fire. Multimeter was bridged to the grating and the ground. Current from the grating was very unstable. Depend on the wind and fire it often changed polarity, sometimes it was disappeared, maxima current was 1- 2 nanoAmpheres.

I tested the grating like antenna. Without fire it was possible to provide a weak reception of a nearest MW broadcasting from MAYAK (http://www.radiomayak.ru/), 549-kHz. I have got detectored voltage near 150- milliVolt. After the fire was started, the voltage risen up to 170- 180- milliVolt, then, after fire took full strength, the voltage drop to 100 and ever 75 – milliVolts with strong fluctuations.

This experiment proved influence fire to radio- reception, but in the other side. The more fire is the bad reception we have. However, thought about the experiment made me to decision that all were right. The maxima of ionization are inside of the fire. But fire lay on the ground. So the fire just shorts the antenna to the ground! To obtain benefits from the fire it required to insulate the fire from the ground.

Next day I found of three empty bottles made from a transparent glass. The bottles were installed on two bricks.

On the bottles I put the rim for fire. The rim had wire connected to it. (If you want to repeat my experiment- be careful! The design is not absolutely stable…) Like a ground I have used an iron sheet laying on the ground near the fire. **Figure 1** shows the schematic of the Fire Antenna and Detector Receiver

The experiment was made 5 times at different days. Were received very stable data. Current was zero at no fire. When the fire was burning, the current slowly increased up to + 7 nano Ampheres (at fire shown at **Picture 1**). Ever weak wind destroys the ionic column and decreased the current from the fire.

Picture 1 Typical Fire

I tested the fire antenna for radio- reception. No significant improvement at reception of a strong station. Broadcasting MW station MAYAK showed stable 110- milliVolt without fire. At fire there were fluctuations from 80- to 150- milliVolt. At headphones it looks like HF- reception with fading. It was found that the fire could change tuning of the detector receiver.

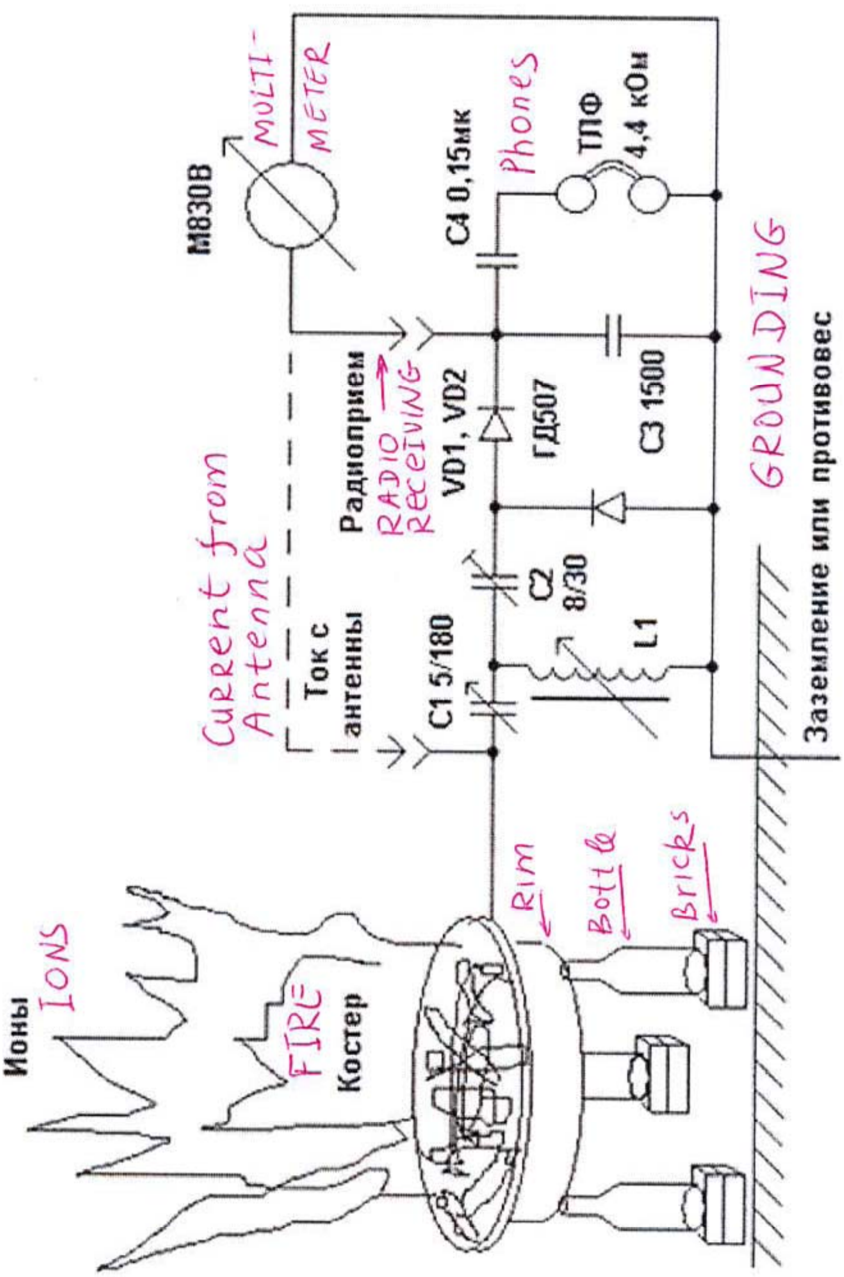


Figure 1 Schematic of the Fire Antenna and Detector Receiver

However, it was visa versa for a weak broadcasting radio- station. At my area I received a weak MW station PODMOSKOVIE, 846- kHz. Without fire it was only 15- 17 milliVolt from the station. After fire was burned the voltage increased up to 50- milliVolt. Fire was burned down the voltage drop to 15- milliVolt…

So, the experimenters proved that fire antenna gives effect at small signals, when RF-current is near the border of the atmospheric current that antenna may take from the atmosphere.

The experimenters are not finished. It should be tested receivers more sensitivity the simple detector ones. Different frequencies should be tested. Only after that we can say something about the Miracle Fire Antenna.

References

1. http://www.antentop.org/009/ra3aae009.htm

2. Receiving: Simple AM Receivers: Moscow, DMK-Press, 2001 (pp: 83.. 85)

3. http://www.antentop.org/010/ra3aae_010.htm

Compact Cage Antenna for 435 MHz

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ru3arj@mail.ru

Credit Line: www.cqham.ru

In article "Compact Cage Antenna" was discussed CCA for HF- ranges. However, that kind antenna may be used for UHF-ranges. At the article it discussed a CCA for 435 MHz.

I have made experimental CCA for 435-MHz **(Figure 1)**. MMANA file for the antennas (cca-435.maa) may be downloaded by the link below. Antenna was made of a bimetallic wire 2-millimeter diameter (9-AWG).

Balun- 4 ferrite cores (permeability 100, OD- 12 mm,

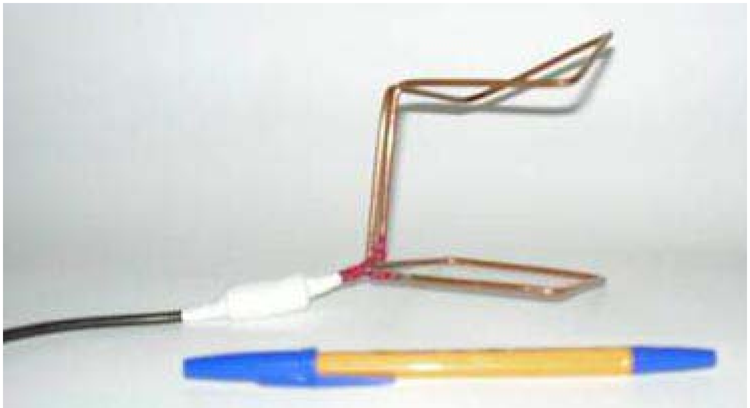


Figure 1 CCA for 430-MHz

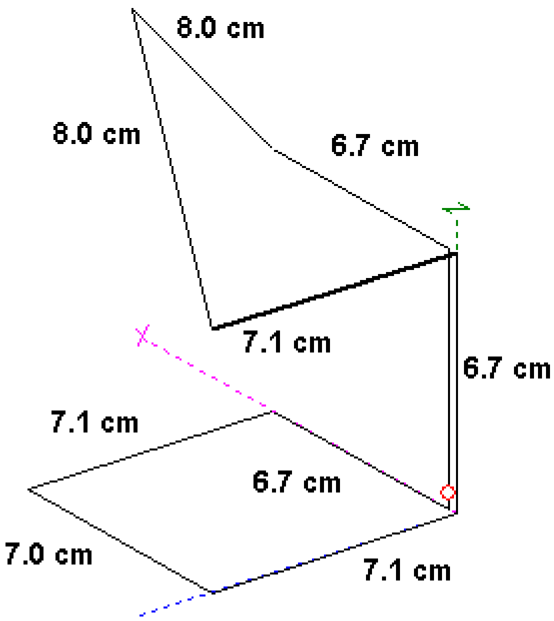


Figure 2 Design of the CCA for 430-MHz

ID- 6 mm, H-5 mm). The RG-58 coax is fed through the cores near the feeding point. The SWR is 1.2:1 on the whole 70-cm band. **Figure 2** shows the design of the antenna (taken from the MMANA file).

The antenna has much more gain in comparison with portable whips. The antenna may be easily placed anywhere and it won't catch somebody's eye because it does not look like usual antennas.

CCA can be wrapped in plastic film in order to look like a Chinese lamp or advertising decoration.

73! de RU3ARJ

Files MMANA:

http://www.antentop.org/011/cca435_011.htm

Broadband Avia Antenna (DEWD)

The publications devoted to memory UR0GT.

Credit Line: http://www.radioscanner.ru/forum/topic25617-3.html

Nikolay Kudryavchenko, UR0GT

The antenna has SWR less the 2.0:1.0 at the frequencies from 112 up to 165- MHz. Gain and pattern belong to the antenna are almost similar to a vertical lambda/2 dipole. The antenna has three resonance frequencies at the working band (112- 165- MHz). It explains why the antenna is such broadband.

Design: Two aluminum tubes, each is in 16-mm diameter, are placed in parallel plane with vertical shift compare to center each other. Crosspiece between the tubes made from a 50- Ohms coaxial cable. The crosspiece is in the plane where the tubes located. **Figure 1** shows the design of the Avia Antenna.

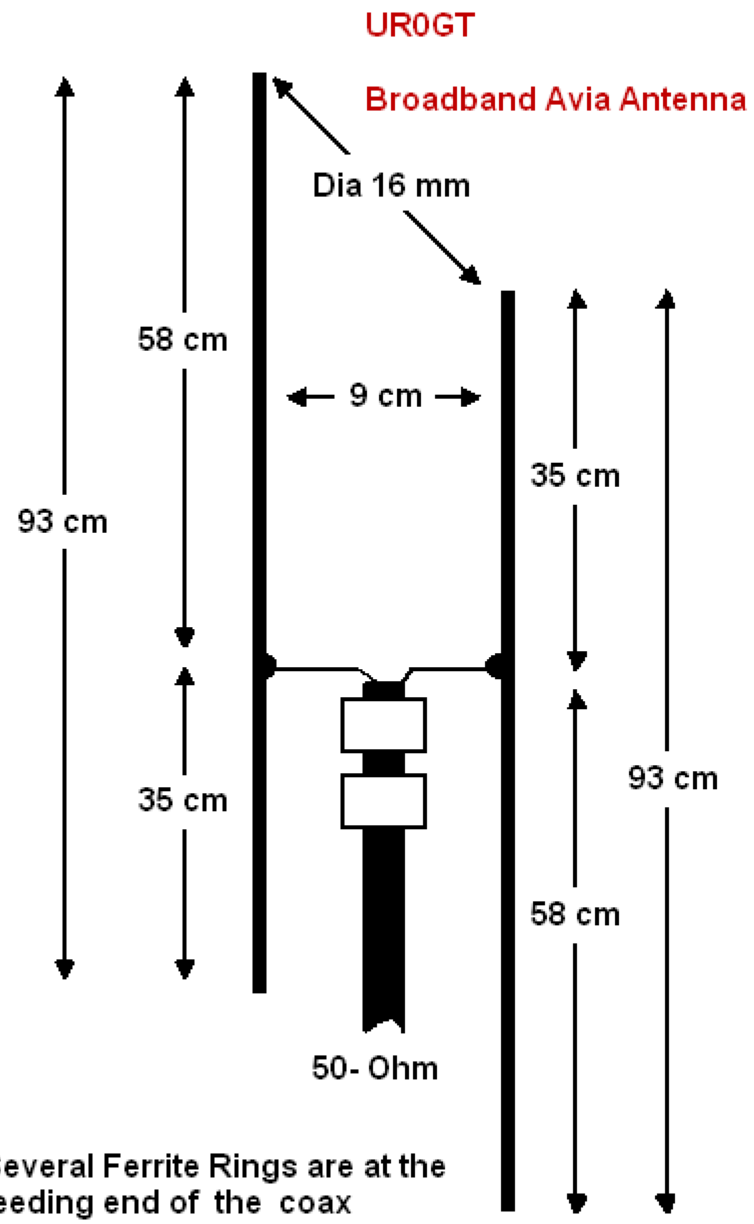
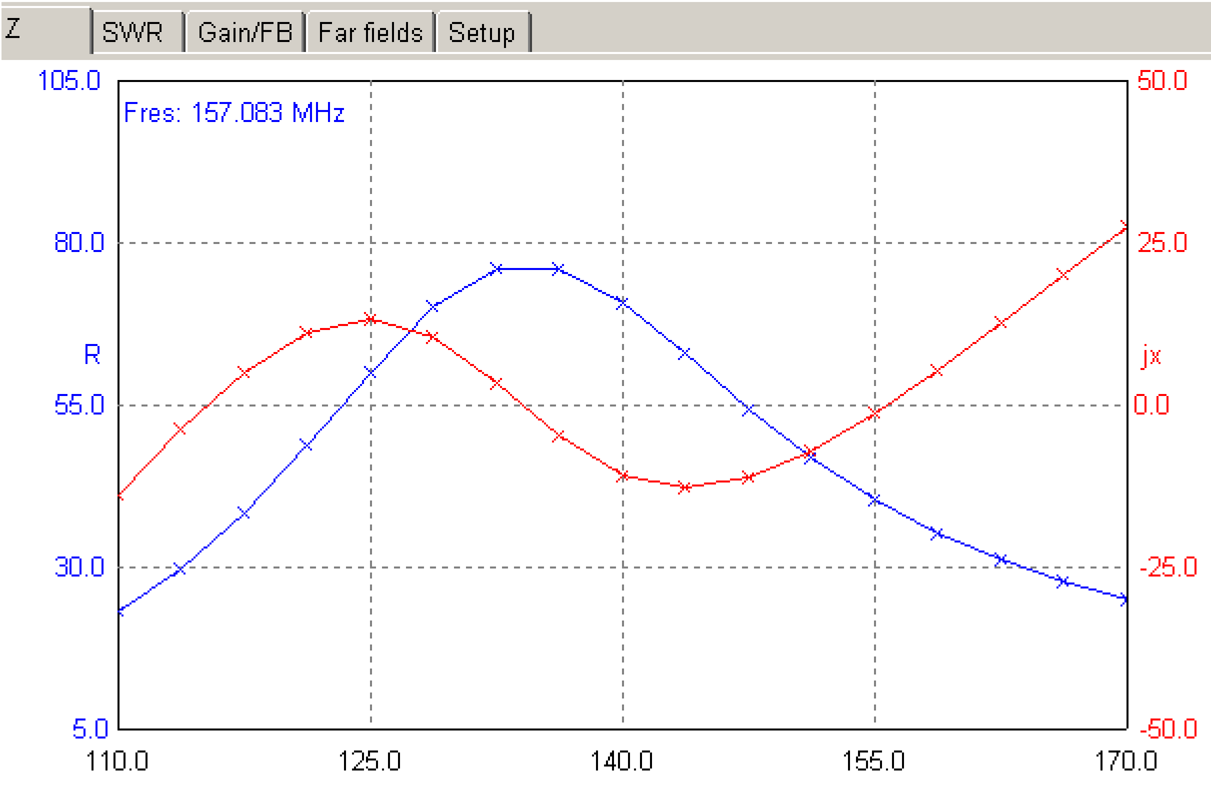
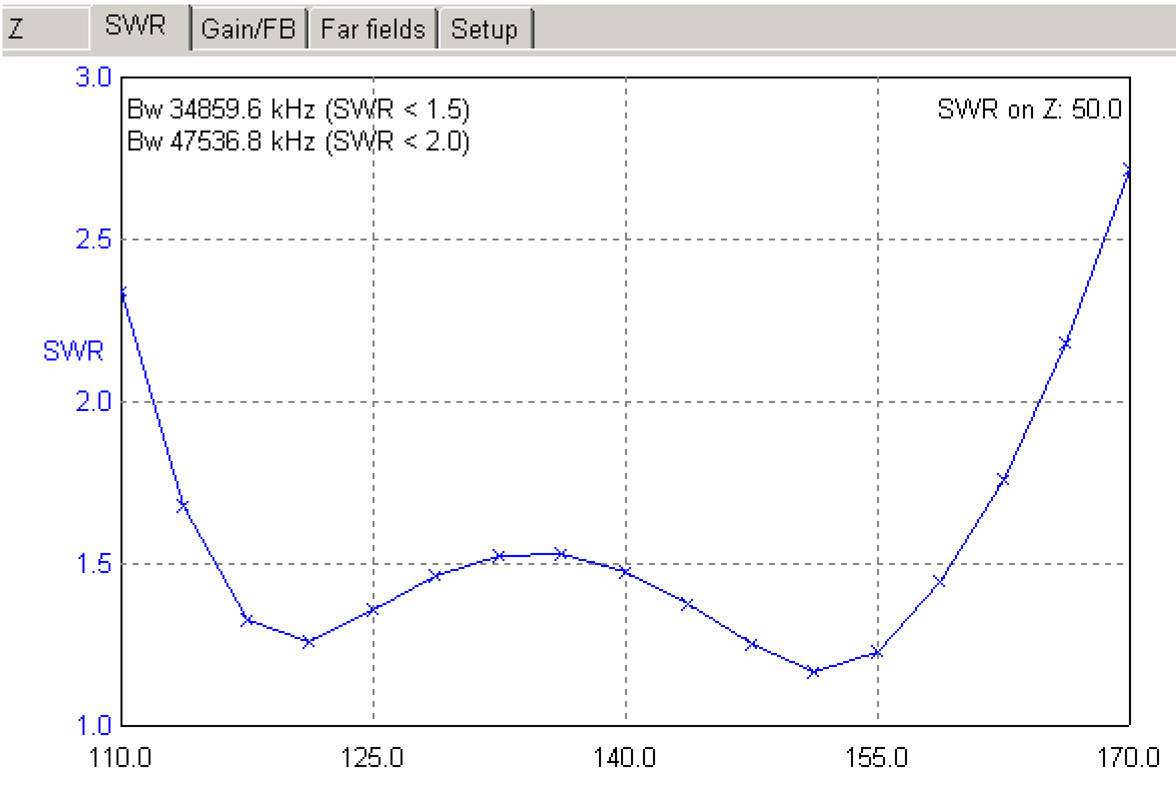


Figure 1 Design of the Avia Antenna

Antenna may be made from a tube less or more in diameter the 16- mm. However the less diameter the less working range, and visa versa, the more the diameter is the wider the working range of the antenna. Antenna may be made from L-Bar with width 14- 18- mm.



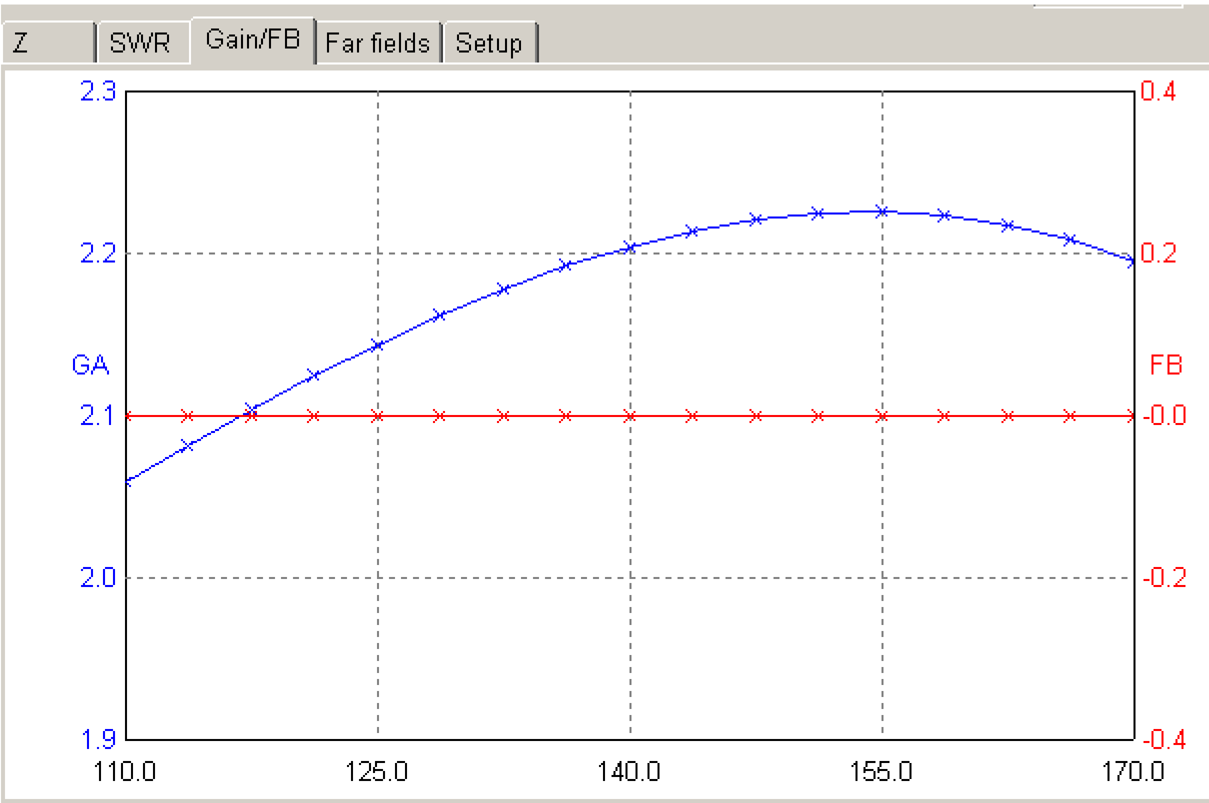
Z of the Avia Antenna



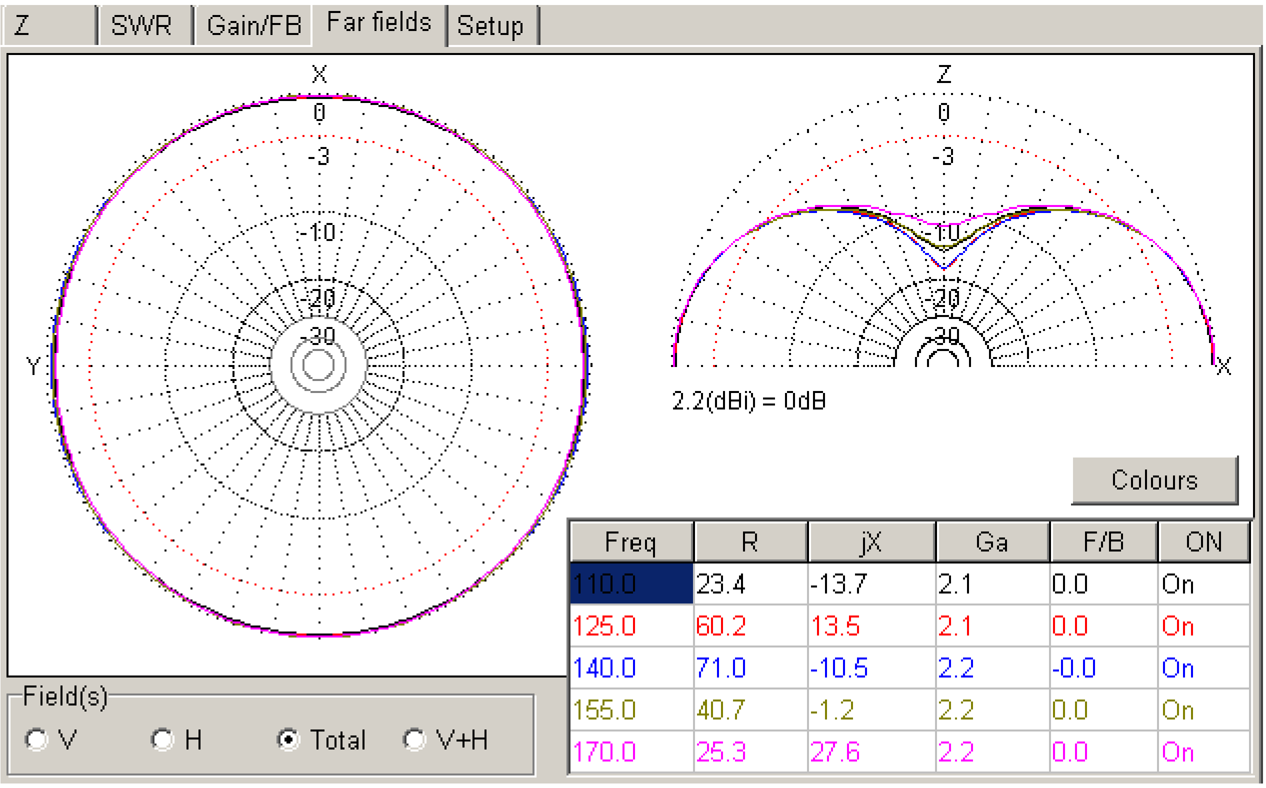
SWR of the Avia Antenna

The antenna is a symmetrical antenna, so, several ferrite rings at the feeding point will keep the symmetrical. Ferrite rings should be placed on the coaxial cable straight near "fork."

File MMANA:
http://www.antentop.org/011/avia_011.htm



Gain of the Avia Antenna



Pattern of the Avia Antenna

4- Elements Directional Broadband Avia Antenna

The publications devoted to memory UR0GT.
Credit Line: http://www.radioscanner.ru/forum/topic25617-6.html

Nikolay Kudryavchenko, UR0GT

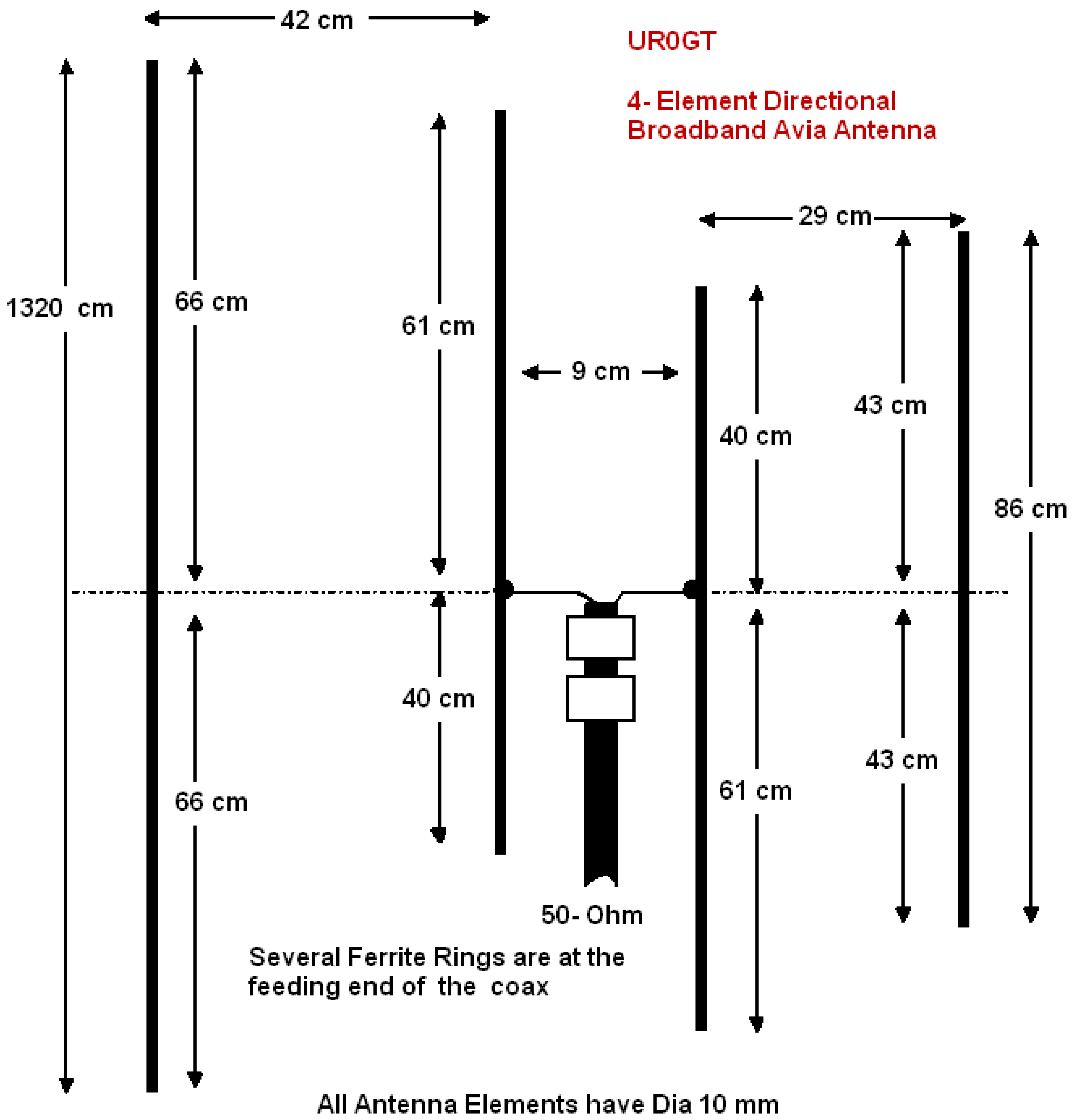
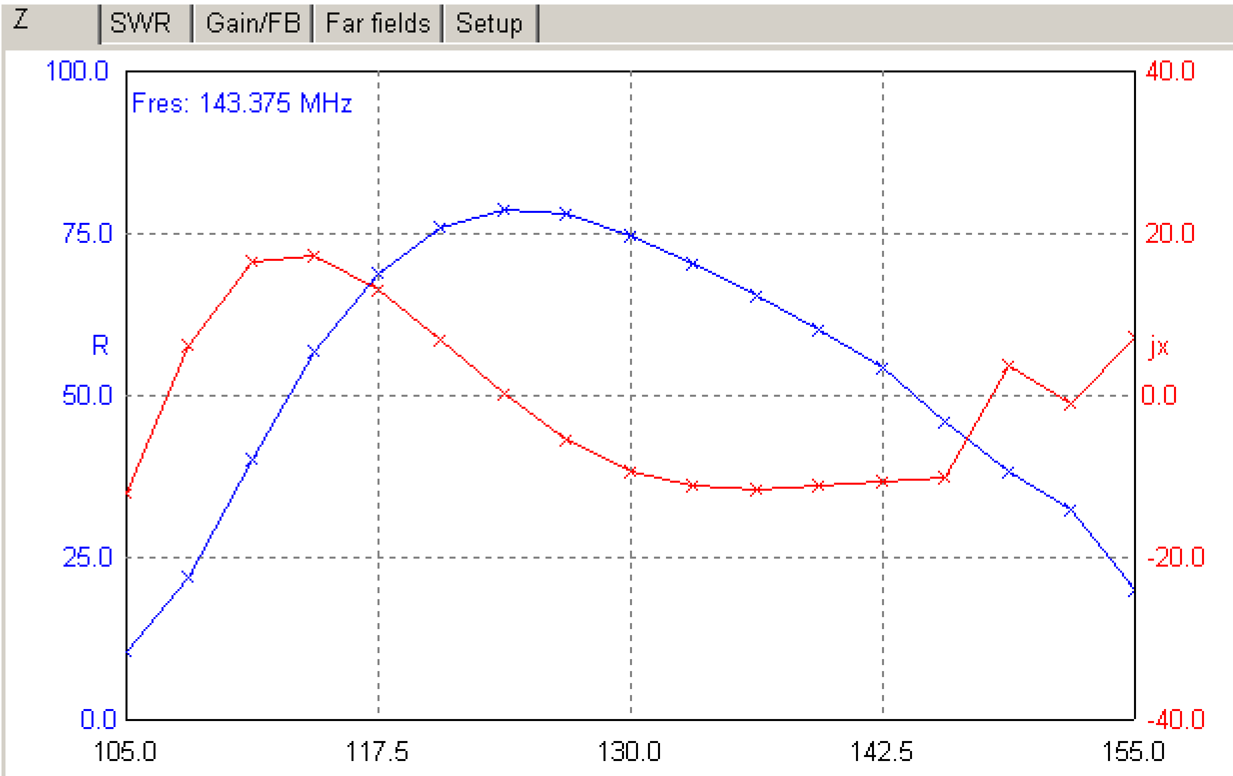
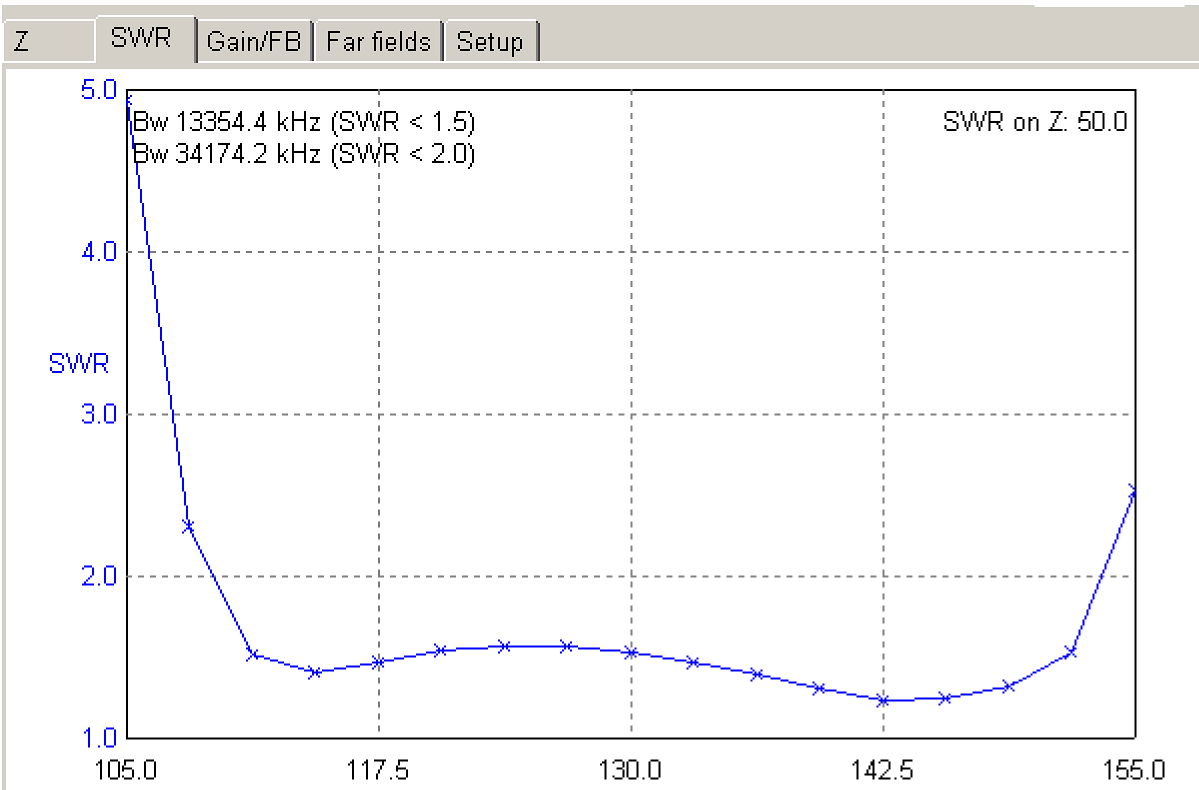


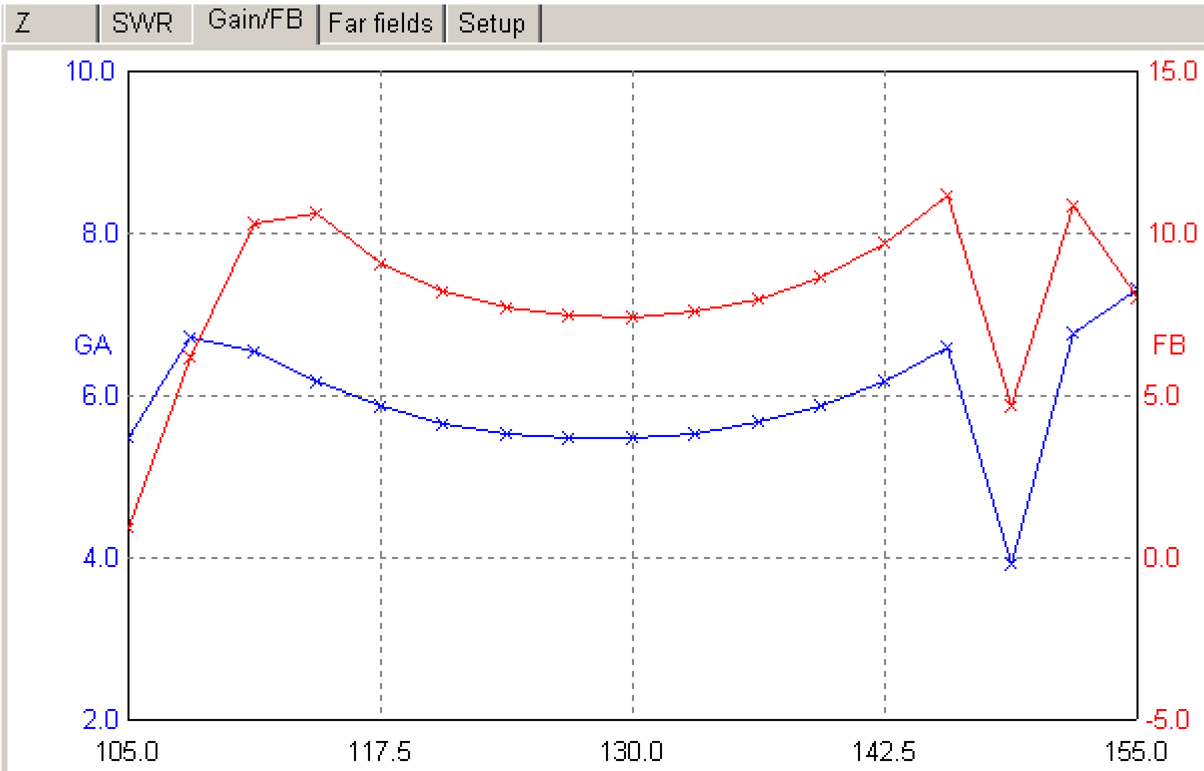
Figure 1 Design of the 4 Elements Directional Avia Antenna



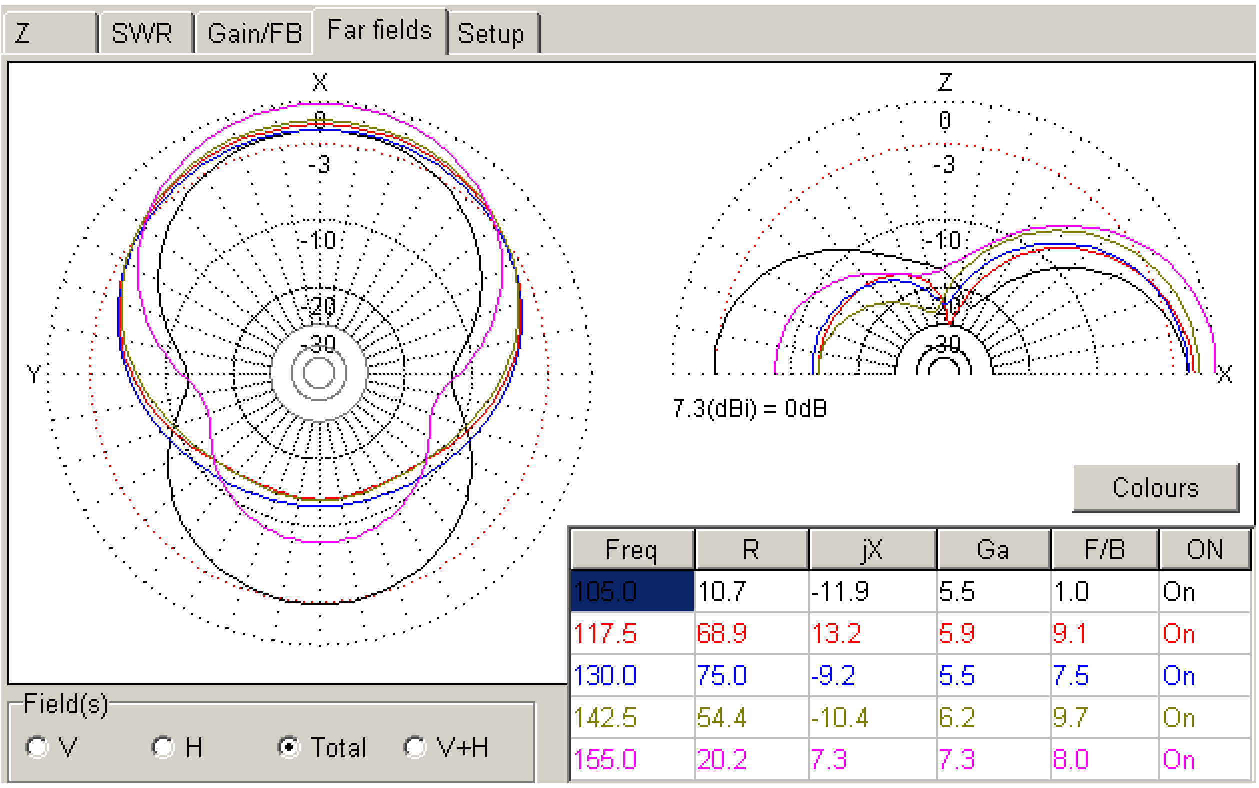
Z of the 4 Elements Directional Avia Antenna



SWR of the 4 Elements Directional Avia Antenna



Gain of the 4 Elements Directional Avia Antenna



Freq	R	jX	Ga	F/B	ON
105.0	10.7	-11.9	5.5	1.0	On
117.5	68.9	13.2	5.9	9.1	On
130.0	75.0	-9.2	5.5	7.5	On
142.5	54.4	-10.4	6.2	9.7	On
155.0	20.2	7.3	7.3	8.0	On

Pattern of the 4 Elements Directional Avia Antenna

File MMANA: http://www.antentop.org/011/dir_avia_011.htm

Bi- Quad Antenna for 2420-MHz

The publications devoted to memory UR0GT.
Credit Line: http://www.lan23.ru/forum/showthread.php?t=981

Nikolay Kudryavchenko, UR0GT

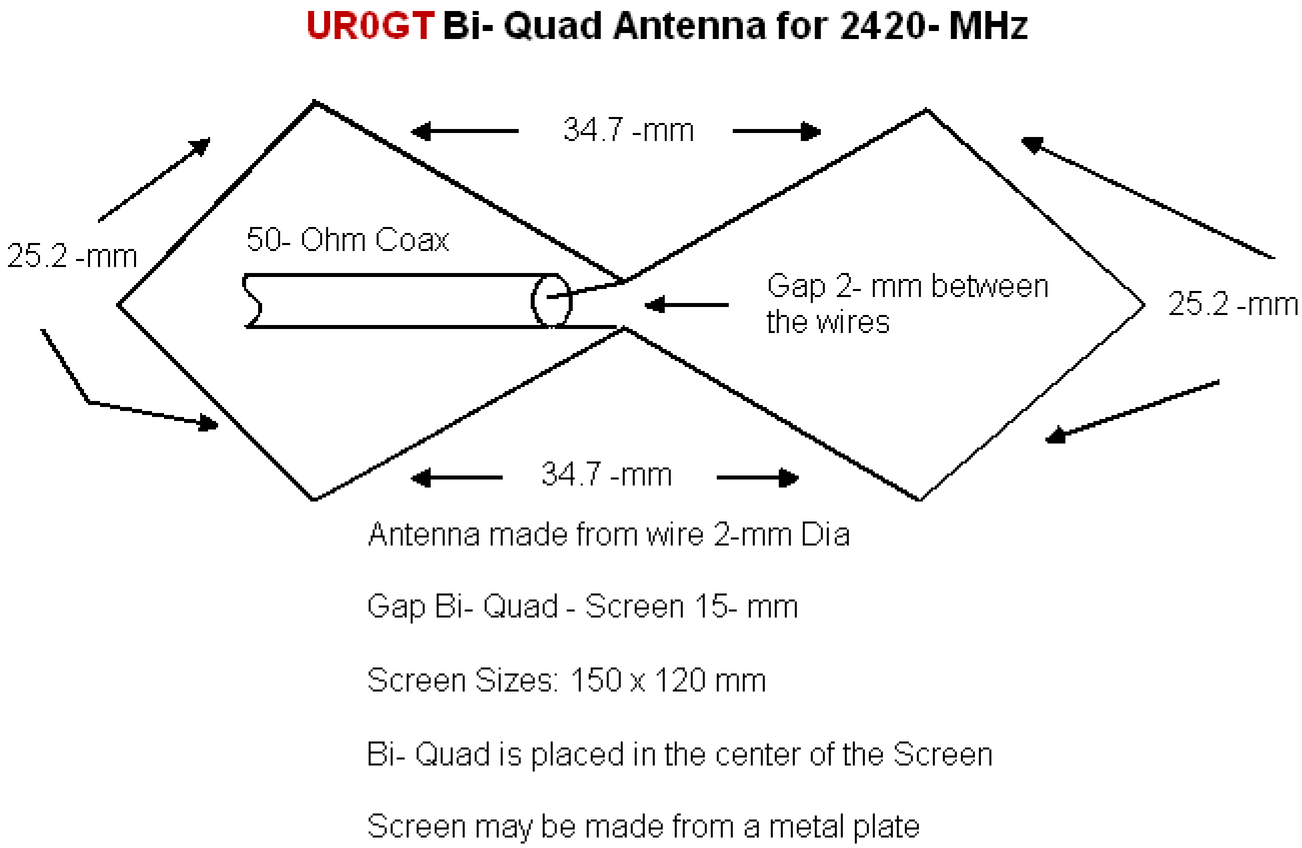
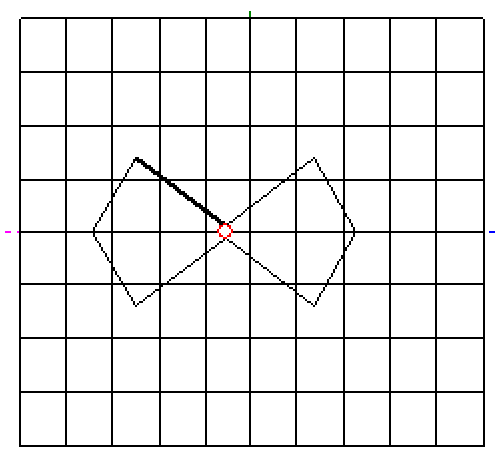
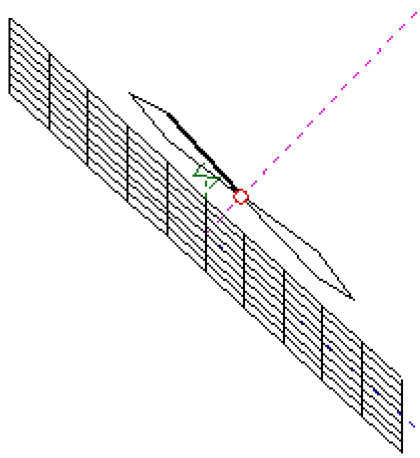
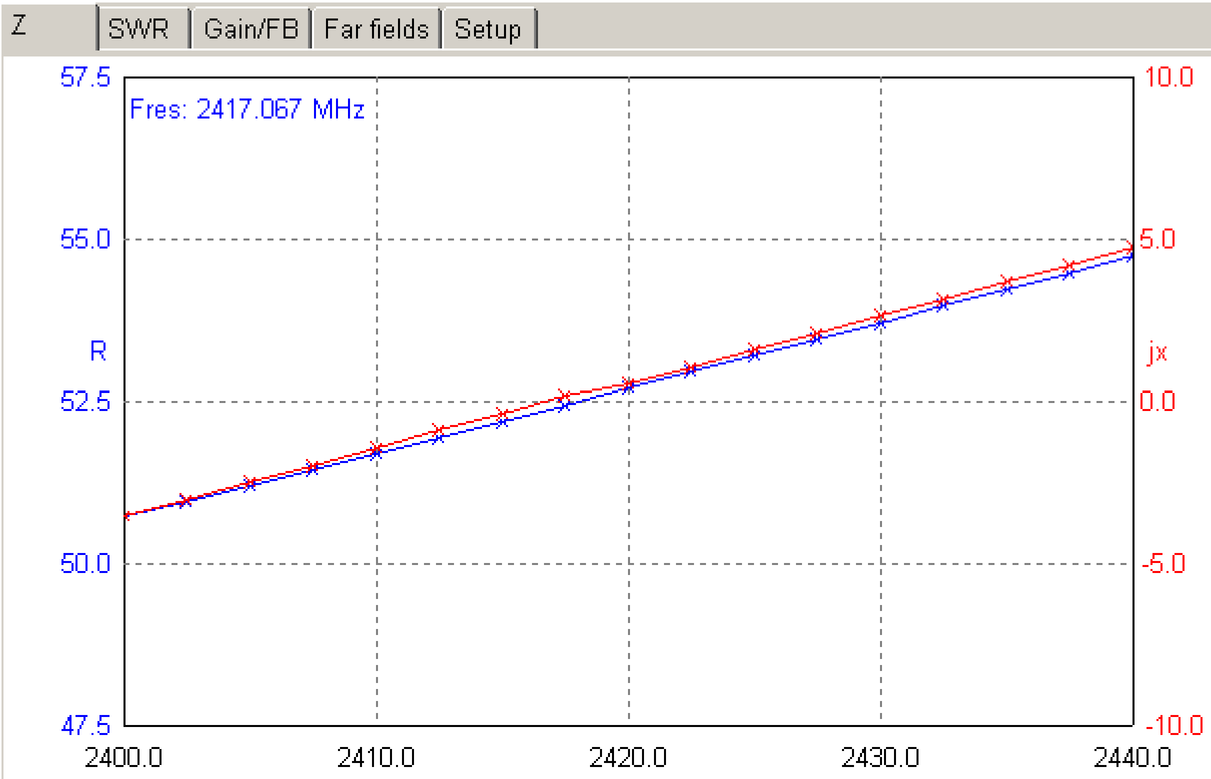


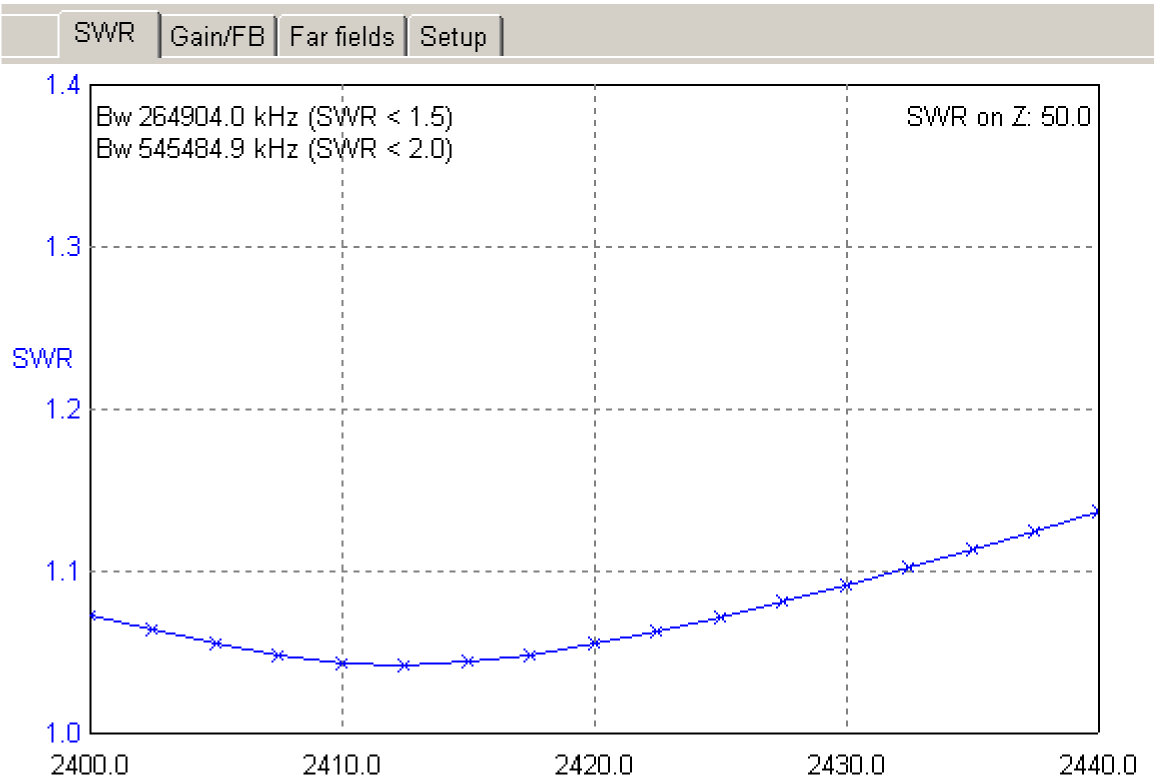
Figure 1 Design of the **Bi- Quad Antenna for 2420- MHz**



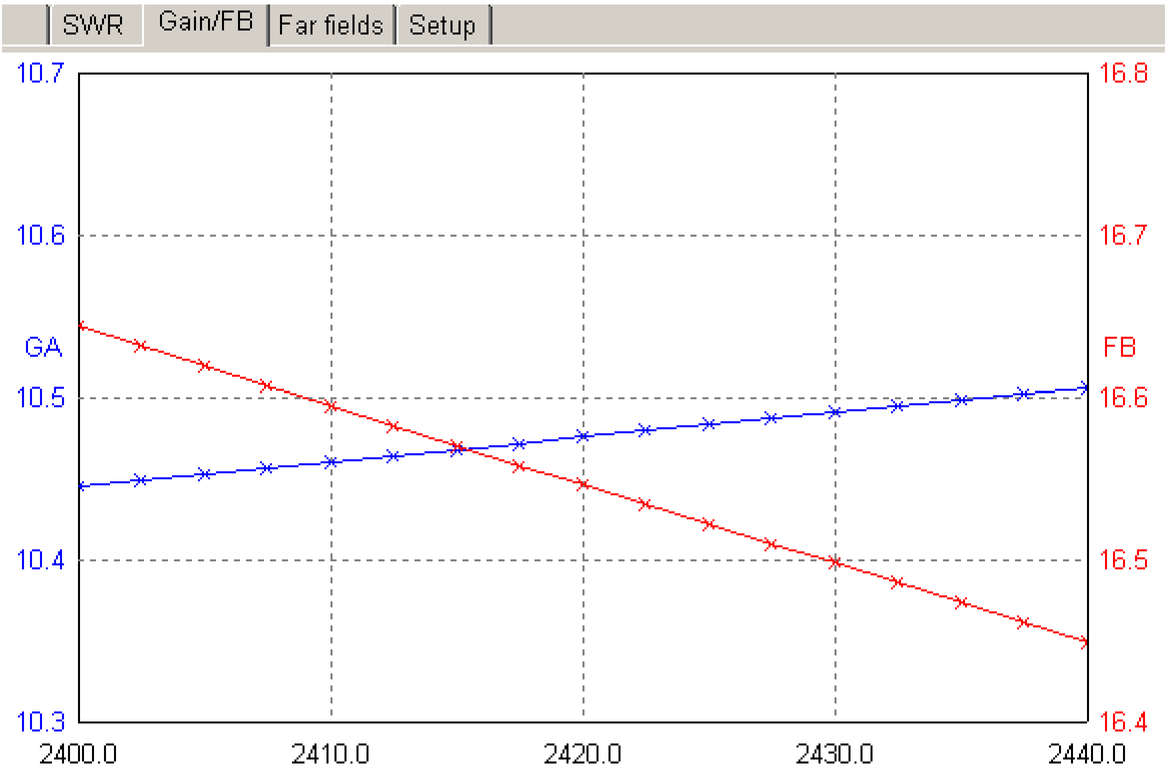




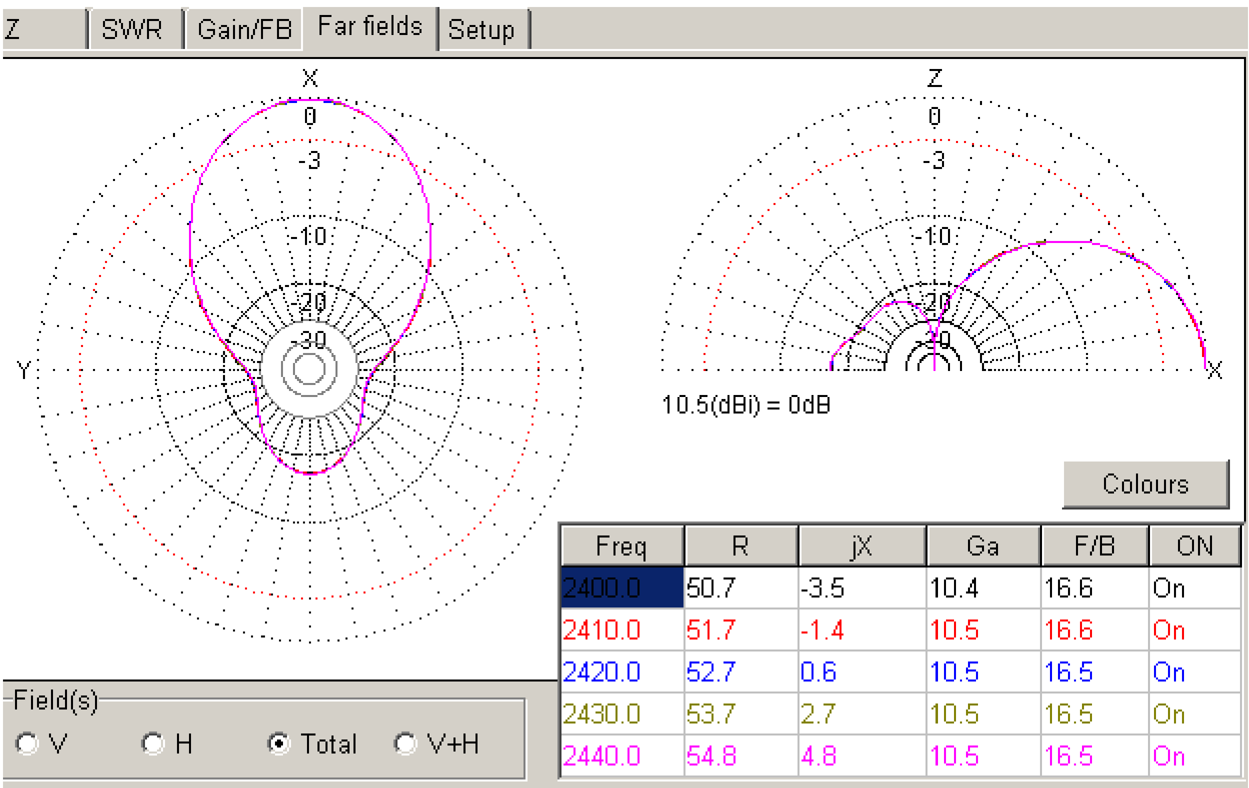
Z of the Bi- Quad Antenna for 2420- MHz



SWR of the Bi- Quad Antenna for 2420- MHz



Gain of the Bi- Quad Antenna for 2420- MHz



Freq	R	jX	Ga	F/B	ON
2400.0	50.7	-3.5	10.4	16.6	On
2410.0	51.7	-1.4	10.5	16.6	On
2420.0	52.7	0.6	10.5	16.5	On
2430.0	53.7	2.7	10.5	16.5	On
2440.0	54.8	4.8	10.5	16.5	On

Pattern of the Bi- Quad Antenna for 2420- MHz

File MMANA: http://www.antentop.org/011/biquad_011.htm

Universal RF - Amplifier of a QRP- Transceiver

By: Igor Grigorov, UZ3ZK

Credit Line: Radiokonstruktor # 7, 1999, pp. 2-3.

QRP design- it is very specific design, where ham wants to use as little parts as possible. The amplifier gives the possibility because it can be used twice- like RF- PA and like RF- RX- Amplifier.

Technical Data for the amplifier:

Supply Voltage: 12-V

Gain: near 15 (Pout/Pin) (at 28.0- MHz) – 20 (Pout/Pin) (at 1.8- MHz)

Input/Output Impedance: Close to 100- Ohm (amplifier was designed to work with 75-Ohm Coaxial Cable). **Figure 1** (p.71) shows schematic of the amplifier.

The schematic of the Universal RF Amplifier is simple. It is typical Push- Pull. For properly work of the amplifier it needs to install the same collector current (with R3 and R6). Value for the current is 15-25-mA. Transistors VT1 and VT2 should be a matched pair (have the same gain at third different collector currents- 10, 50 and 200-mA.

R4C3 and R5C4 are lowered the gain but rise the stability. Sometimes (when the Amplifier is used at range 1.8- 14.0- MHz and matched antenna is used) it is possible do not install them. Gain is raised up to 25 (Pout/Pin) at this case. With the transistors (KT606A, DATA for the transistors at: http://www.antentop.org/008/bip008.htm). The amplifier work out up to 1-Wtts output power. However, do not drive into the amplifier more the 50-mWtts because the output waveform signal may be distorted.

Figure 2 (p.72) shows the commutation between a QRP-transceiver and the amplifier. For switching RX/TX a two small relays are used. At TX mode the Amplifier should be matched with a QRP- Transceiver. Separate circuit (like L or pi- circuit) for matching transceiver with antenna is used for each band. At RX mode the matching circuit is used for filtering of the input signal. At my design the Amplifier at RX- mode was loaded on to balanced diode mixer.

Parts: All capacitors should have low losses at RF. All transformers are wound by pair of twisted wires (two twist on 1-cm length) in diameter 0.3- mm (29- AWG). Core-OD- 7…15 mm, height- 3- 7- mm, permeability- 400…600. Numbers of winding for each transformers is 20. It is uniform winding along all length of the core. However, for best efficiency of the amplifier the number of winding for each transformer should be experimentally chosen.

Design: The amplifier is assembled on a PCB by sizes 40x 50-mm. Two holes for VT1 and VT2 are drilled at the PCB. The transistors are installed on aluminum plate with sizes equal to PCB. Parts- resistors/transformers/capacitors are installed on the transistors pins and small circles cut on the PCB. Try to keep the leads from the parts as short as possible. Relays are installed near output and mixer of the QRP- Transceiver and connected with the Amplifier with a thin Coaxial Cable.

73! Igor Grigorov, va3znw

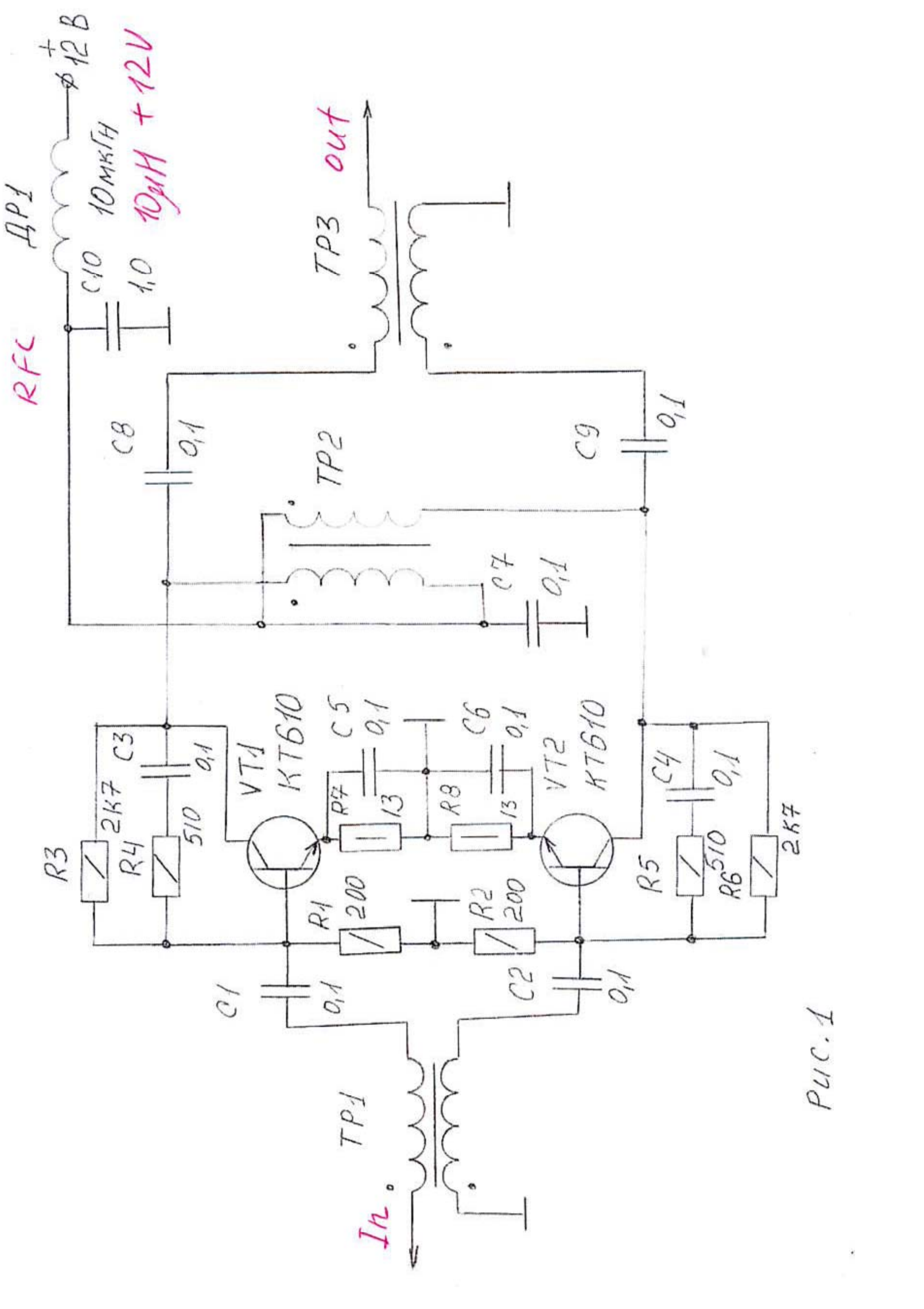


Figure 1 Schematic of the Universal RF Amplifier

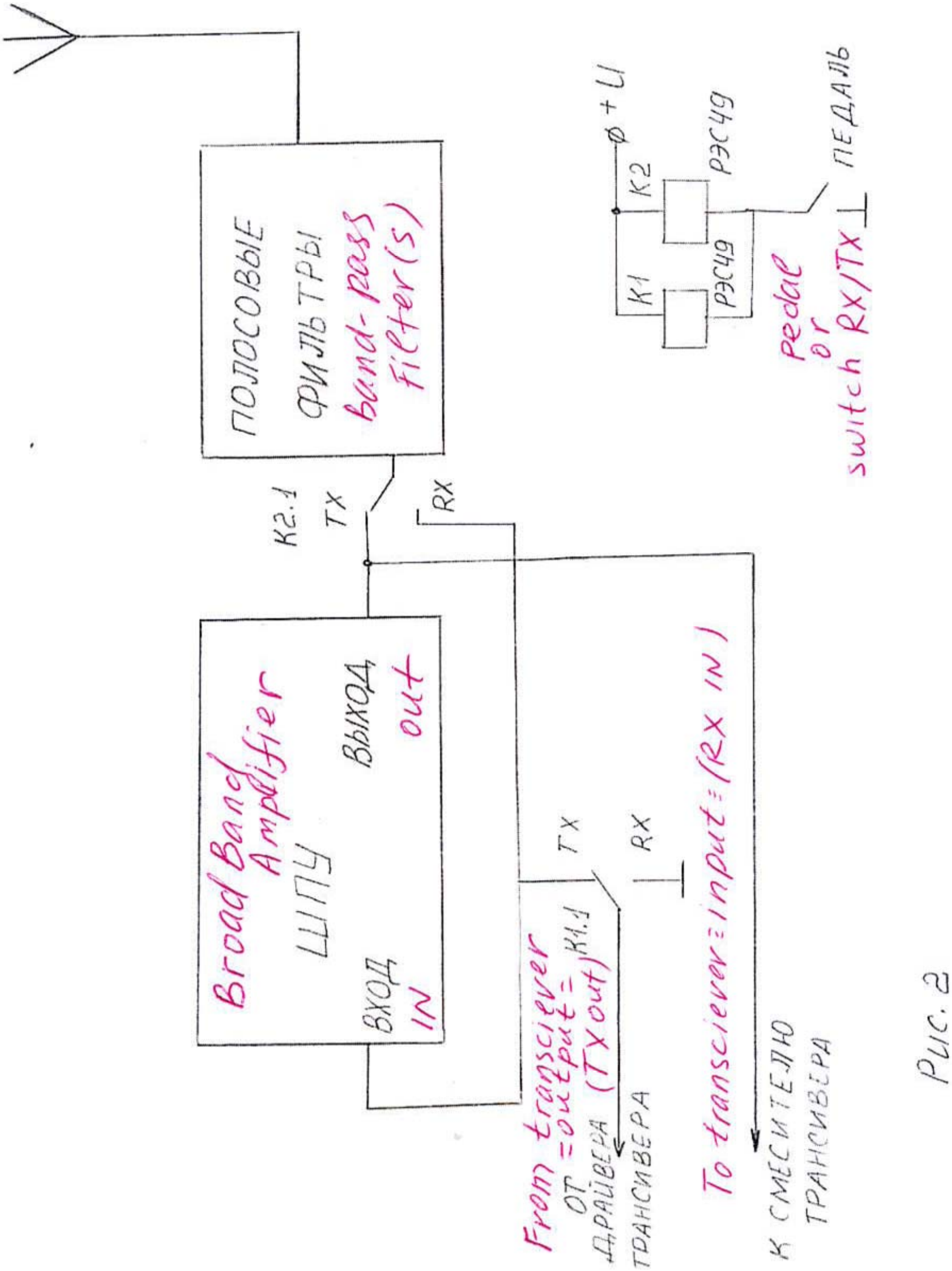


Figure 2 Commutation between a QRP- transceiver and the Universal RF Amplifier

Transceiver "POLEVIK"

By: Vladimir Polaykov, RA3AAE

Credit Line: ***CQ-QRP # 13*** (magazine of the UR- QRP-C)

The schematic of the transceiver is just a project and required the practical test. Author will be really appreciated those who test the transceiver.

Schematic of the transceiver is shown on **Figure 1**. It is DC transceiver.

Transistors VT1 and VT2: They work like a mixer at RX mode and like a PA and doubler (key down) at TX mode.

It is necessary to use MOSFET with the “right” gate/drain characteristic. At such transistor drain current is absent when gate connected to source and voltage is across drain/source. To open the MOSFET you need connect the gate with DC in positive polarity (relative to source). Cutting voltage for such transistors may be 0.5- 2.0- V.

If you have no MOSFET with the “right” gate/drain characteristic you may use usual MOSFET with the “left” gate/drain characteristic. However at the case the MOSFET should be closed by some negative polarity across gate- source. Transistors VT1 and VT2 should be matched pair (have the same parameters).

Transistor VT3: It is Qurtz- RF Generator. It works at the same mode at RX and TX. Quartz has resonance frequency twice lower the working frequency. At RX/TX it should be some frequency shift that allows work at the transceiver mode.

Transistors VT4 and VT5: They work at audio amplifier. Transformer Tp1 is an ordinary output transformer from an old transistor radio.

Tuning and Adjustment: At first at transmitting mode (key down) tune the PA to maximum power. It does with help symmetrical L3 (find necessary numbers of turns) and tuning L4C7 to resonance (F/2). Maximum power into antenna (or dummy load) is depend on L1 and C1. PA tuned to the maximum power should provide maximum sensitivity at RX mode.

Attention: It should not be any current through VT1 and VT2 when the Quartz is removed.

73/72! De RA3AAE

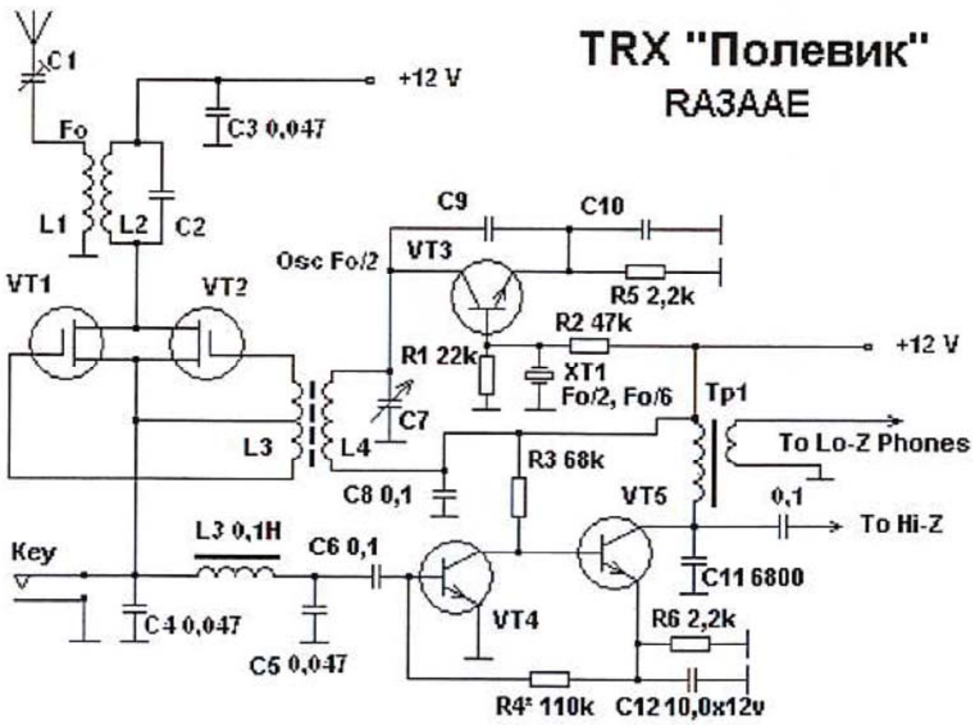


Figure 1 Transceiver “POLEVIK”

Tube AM Radio Station for the 160- meters

By: Igor Grigorov, RK3ZK

Technical Data:

TX: Output RF Power not less the 3 Wtts

RX: Sensitivity not less the 2 micro- V

The radio station consists of from a separate receiver and transmitter. **Figure 1** shows schematic of the AM radio station.

.Transmitter: It is made on VT1, VT2, VT3, VT7. VT1 is a voltage stabilizator for VFO (made on VT2). L1C1C2 should be covered frequencies 900-980-kHz. L2C5 is tuned on 1900-kHz. VT3 is PA for the transmitter. L3C2 and S3 is ATU for the PA. The circuit should match antenna having Z from several Ohms up to several kOhms with the PA. S1 is switch for tuning VFO to calling station. S2 is switch to change mode AM/CW. VT7 is audio modulator of the TX. At AM –mode the PA tube VT3 is switched to bridge with VT7. At TX relay K1 is switched high voltage to PA and shortened the 3-rd grid of the VT3 to the ground.

Receiver: It is made on VT4, VT5 and VT6. VT4 is a voltage stabilizator for the RX. VT6 is RF- amplifier for the RX. L7C23 is tuned to 1900- kHz. Serial L6C20 as well is tuned to 1900- kHz. Double triode VT5 works as regenerative detector (left triode) and Audio amplifier (right triode). L5C18C19 should be have high Q, so, coiled by quality copper wire and C18 and C19 should be air –gap capacitors. R14- RF Gain, R6- regeneration level.

Parts: For Data for inductors see **Table 1**. All resistors may have tolerance 30%. Wide range of tubes may be used at the radio- station. Any low power pentode may be used for VFO. Any pentode that can give out 5-10 Wtts may be used at PA. Tetrode as well may be used for PA. In that case RX input connected through capacitor in 10-pF to anode of VT7, contact K1.1 of relay K1 is connected to bridge with L7C23 (so at TX mode the input is shortened to the ground).

Table 1 Inductors Data of the Radio Station

#	inductance, microH	OD, mm	Length of winding, mm	Numbers of turns	Wire	Note
L1	50	20	20	60	0.2 mm/32-AWG	
L2	40	20	20	53	0.2 mm/32-AWG	
L3	30	34	40	38	0.8 mm/20-AWG	Tap to plate of VT3 from 30 turn from "cold end." 11 taps to S1 from each 3 turn beginning from 4-turn from the "cold end."
L4	5	36	10	6	0.5 mm/24-AWG	
L5	100	34	35	60	0.3 mm/28-AWG	
L6	20		10	25	0.3 mm/28-AWG	
L7	40	20	20	53	0.2 mm/32-AWG	
RFC1, RFC2				400	0.1 mm/37-AWG	Bulk on Russian resistor WS-2, resistance more the 51-kOhms. For the resistor see ***Transceiver SQT, Antentop # 1, 2008***.

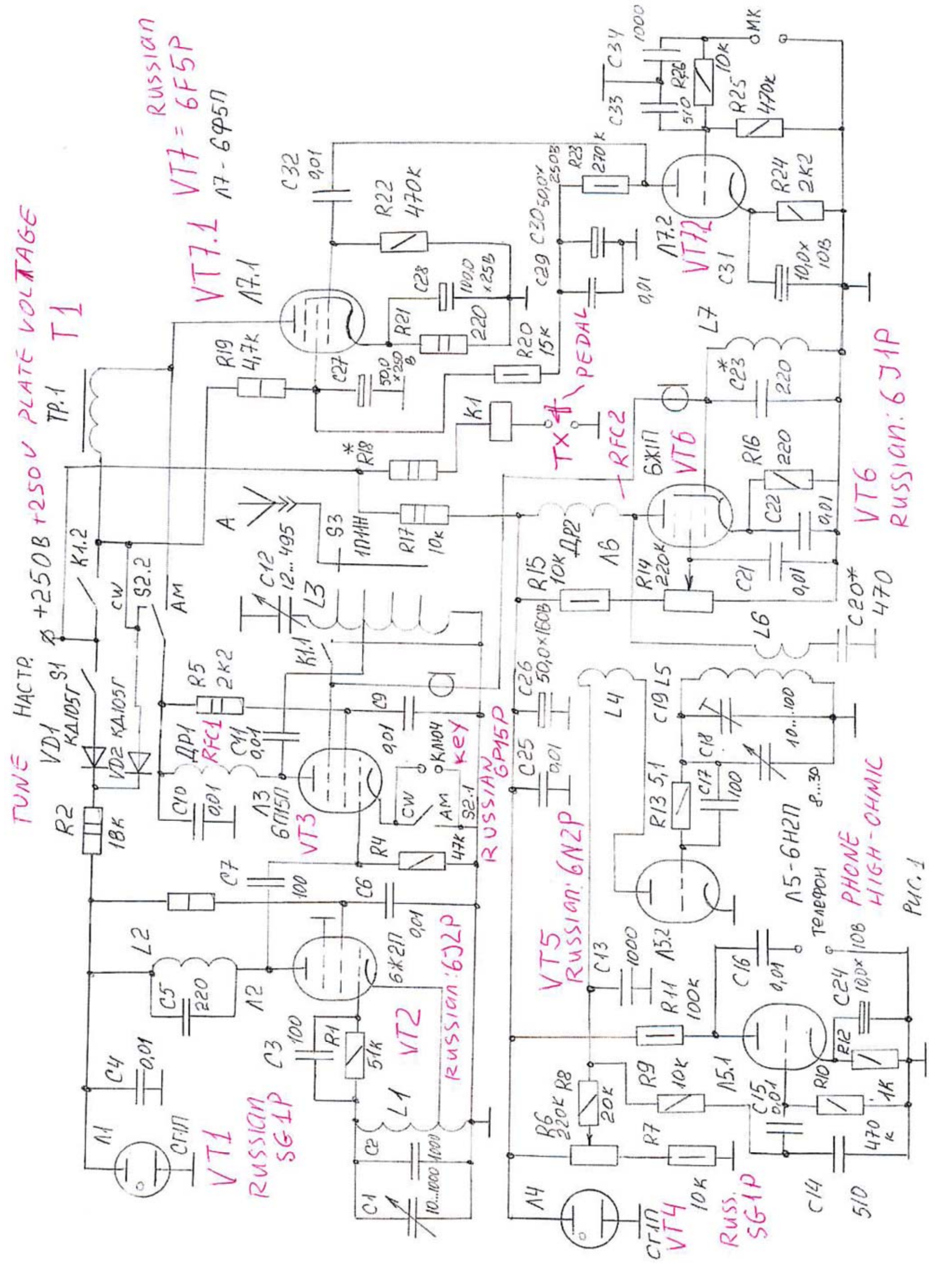


Figure 1 Tube AM radio station for 160-meters

Glow-discharge stabilitron VT1 and VT2 may have voltage 80- 150-V. Any low power double triode may be used at receiver (VT5). Any low power RF-pentode may be used at VT6.

Coil L4 may be moved along L5. It is need to get optimal regime for the regenerative receiver. **Figure 2** shows the design of the receiver's coil. Relay K1- any suitable relay. R18 limited current through the relay.

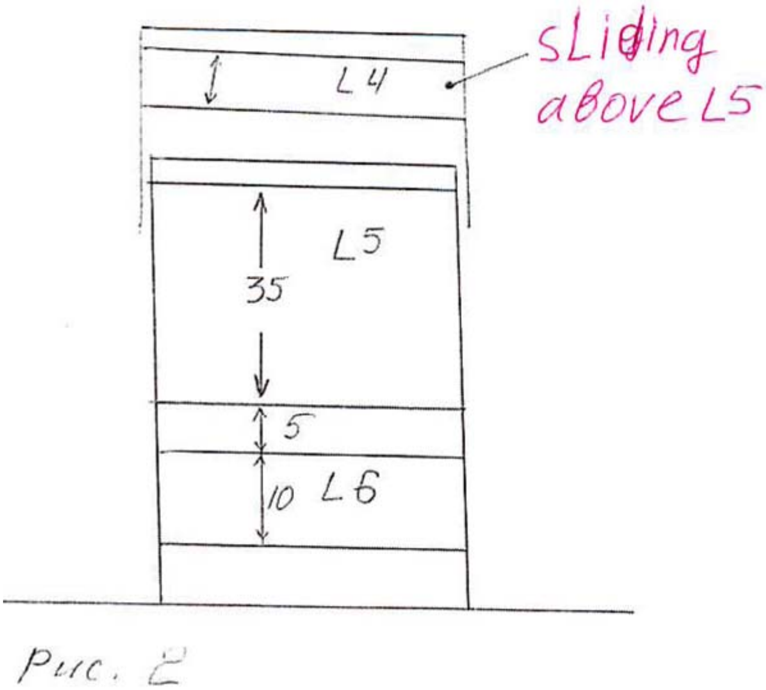


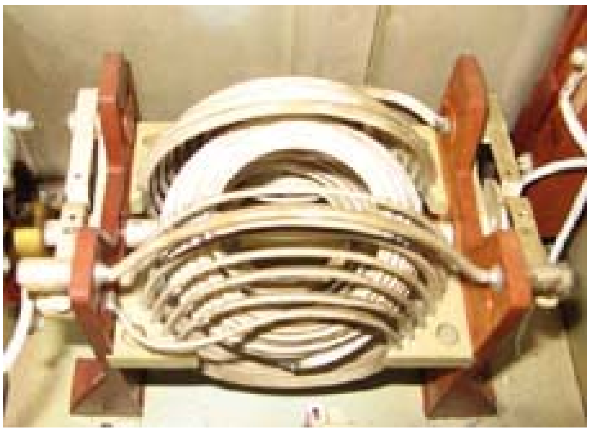
Figure 2 Design Coil of the Receiver

Tuning of the Radio Station

The tuning is simple and takes one evening if the radio station is made from right parts that are installed in correct way.

At first do tuning of the receiver. Receiver should receive something with antenna at least 3 meters long. Tune L3C12S3 for the best receiving. With help C19 set the receiver to the middle of the 160-meters band. Be sure, that C18 covers all 160-meters band. R6 should provide a smooth regeneration control. If not, change the distance along L4 and L5. If it is no regeneration, switch visa versa terminal of L4 or decrease distance between L4 and L5. Set the receiver in the middle of the 160s and tune L6C20 and then L7C23 on to maximal sensitivity.

The second, do set up of the transmitter. Begin from the VFO. Using MW receiver set the VFO to 900- kHz (300-meters). Load the transmitter to a 50-Ohm/10-Wtts resistor or bulb having resistance in range 50- 300- Ohms. Adjust doubler L2C5 on to maximum output power of the PA. Audio amplifier/modulator works straight away at right parts. Audio transformer T1 was used from an old tube receiver.



Design of the Radio Station

Design of the Radio Station is shown on the **Figure 3**.

However, it is very possible to use almost any old tube receiver to be remade in the Radio Station.

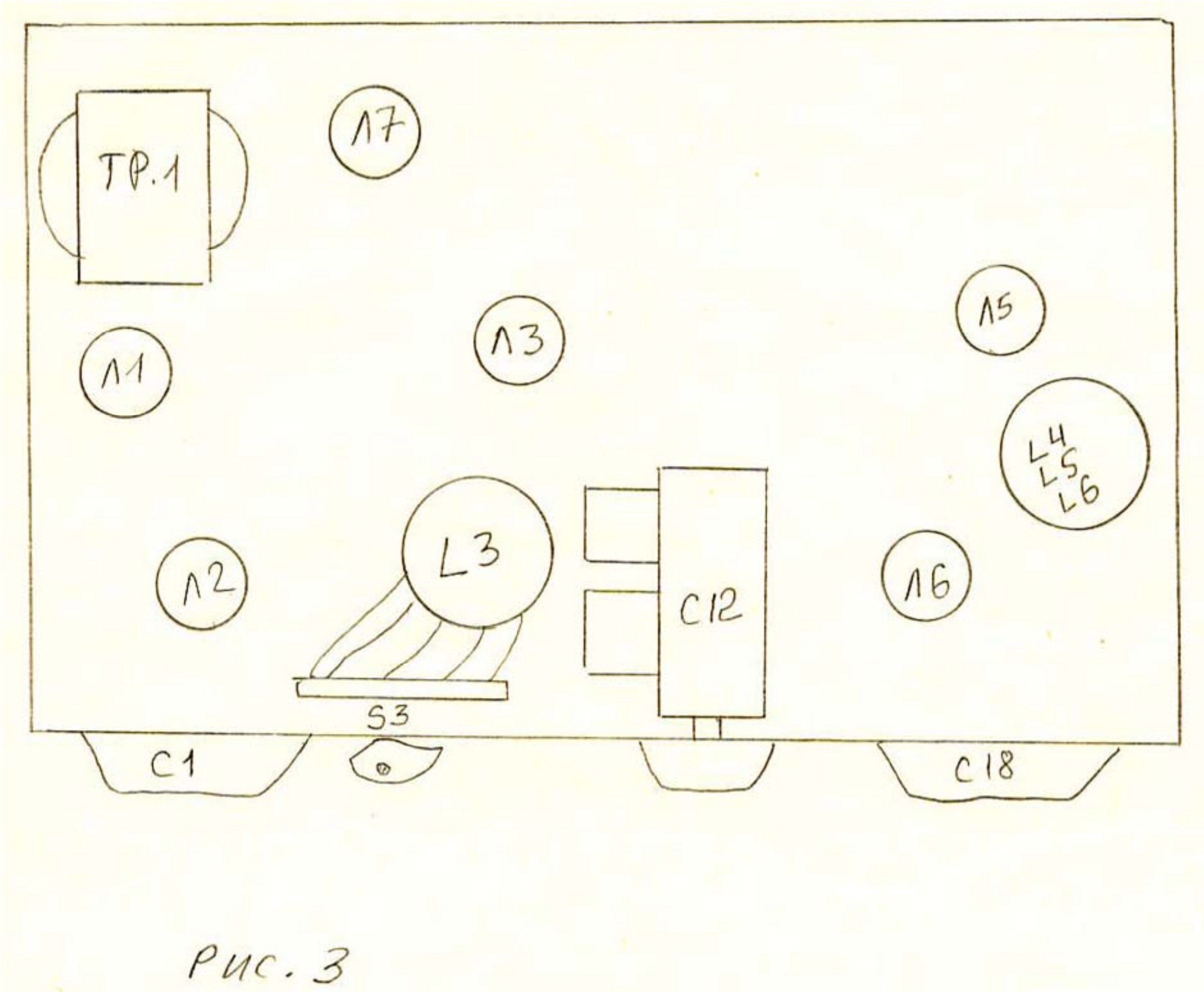


Figure 3 Design of the Radio Station

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2. Radio №9 1980, V. Grushin (RA3ANW), AM Transmitter for the 160- meters;
3. Radio №9 1979, V. Grushin (RA3ANW), Simple AM Transmitter
4. Radio №4 1980, Y. Lapovok, Transceiver for the 160- meters
5. Radio №10, №11 1982, V. Polaykov, DC Transceiver for the 160- meters

QRPP/QRP Transceiver by UB5UG

By: **Yuri Medinets, UB5UG, Kiev**

Credit Line: Radio #1, 1984, p.24

The transceiver is designed for the 10-meters Band. Schematic of this simple and reliable transceiver is shown on **Figure 1**.

Transceiver consists of the RF generator on VT1 (RX/TX), mixer on VD3 (RX) and audio amplifier on VT2 and VT3 (RX).

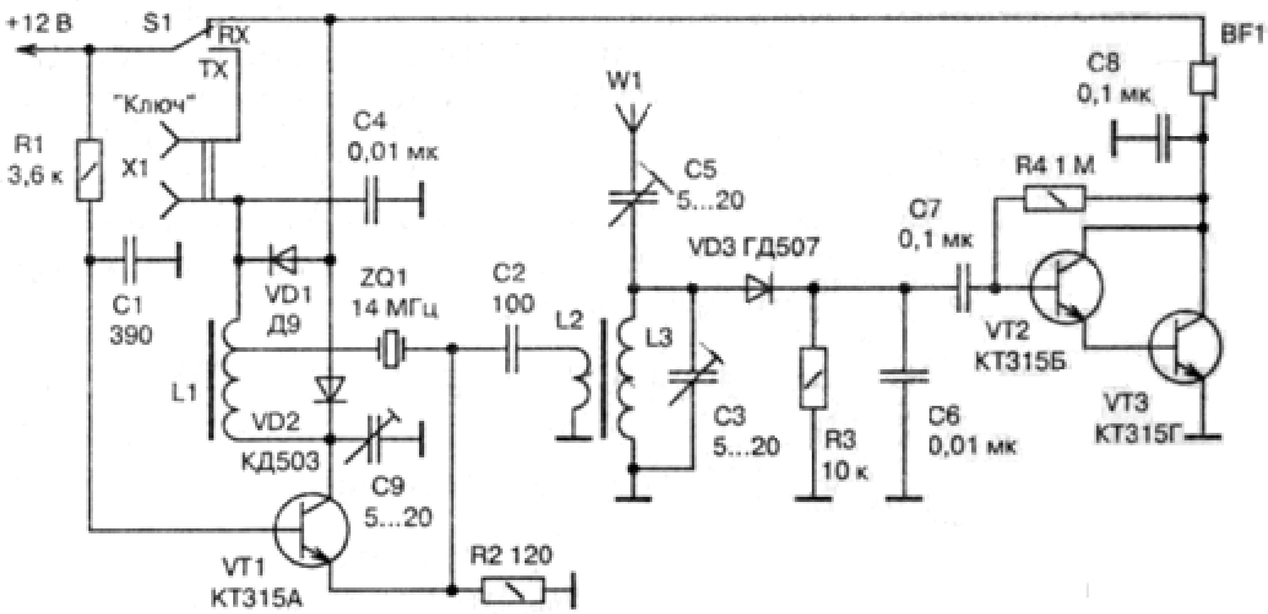


Figure 1 Transceiver by UB5UG

RX Mode: RF voltage at L1 is limited by VD1and VD2 up to a level of 0.3-V. The level (0.3-V) is chosen by amplitude of the second harmonic (that selected with L2L3C3). The amplitude should be sufficient for good job of the mixer on VD3.

TX Mode: Diodes VD1 and VD3 are electrically unplugged from L1C9. RF voltage is maximum across L1. TX gives full power into L2L3C3 and then into antenna. Frequency of the generator differs (because VD1 and VD3) at RX/TX on several hundreds Hz, that is needed transceiver mode. One –Transistor quartz generator may provide up to 50-mWts. Simple Push-Pull PA may increase the transceiver power up to 1-Wtts (and convert QRPP to QRP). **Figure 2** shows schematic for the PA. C2 and L2 do not use at the configuration.

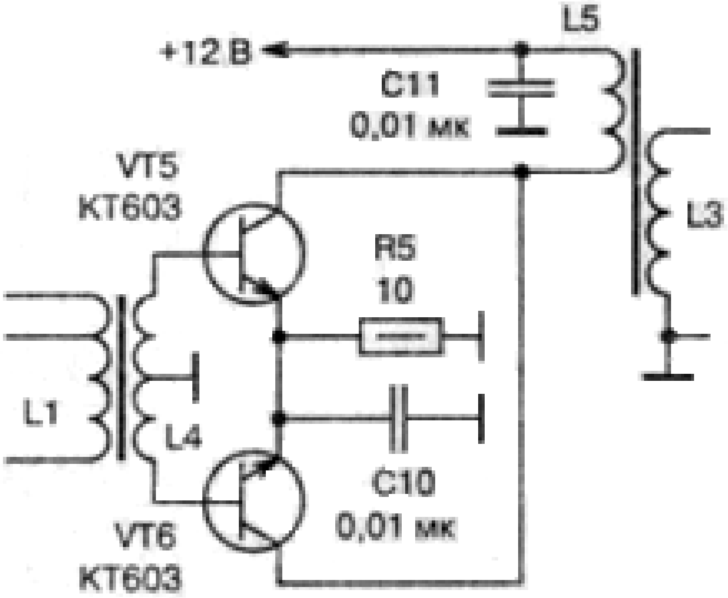


Figure 2 PA for QRPP Transceiver

All inductors were coiled on ferrite rings (permeability 30, OD- 7-mm, ID- 4-mm, H- 2-mm) by wire dia 0.27-mm (29-AWG).
L1: 2+ 22 turns
L2: 1 turn
L3: 12 turns
L4: 2x3 turns
L5: 4 turns

VD1: Germanium small power detector diode
VD2: Silicon small power high speed switch diode
VD3: Germanium mixer diode

VT1: Small power RF Transistor, Fmax- 250- MHz, Pmax- 250-mWtt
VT2, VT3: Any small power high gain transistors
VT5, VT6: Middle power RF Transistors, Fmax- 250- MHz, Pmax- 2.0-Wtt

BF1: High ohmic (more the 1000- Ohms) head-phone

X1: Connector for straight key

73!

Military Russian Radio-Operators

"Antennas" by Sergey Nadenenko

The Book Is in the Antentop Amateur Library

I'm pleased to announce the famous in ex- USSR book **“Antennas” by S. Nadenenko**. The book was published at 1959 in limited circulation. Lots students learning Electronics had read the book in their university library. As well as me when I was a student I prepared for testing on Antennas using the book.

Now the book is the real “hard- to- find” thing. However, I have got a “pdf copy of the book and now the book is in free access for everyone.

Copyright and distribution issues

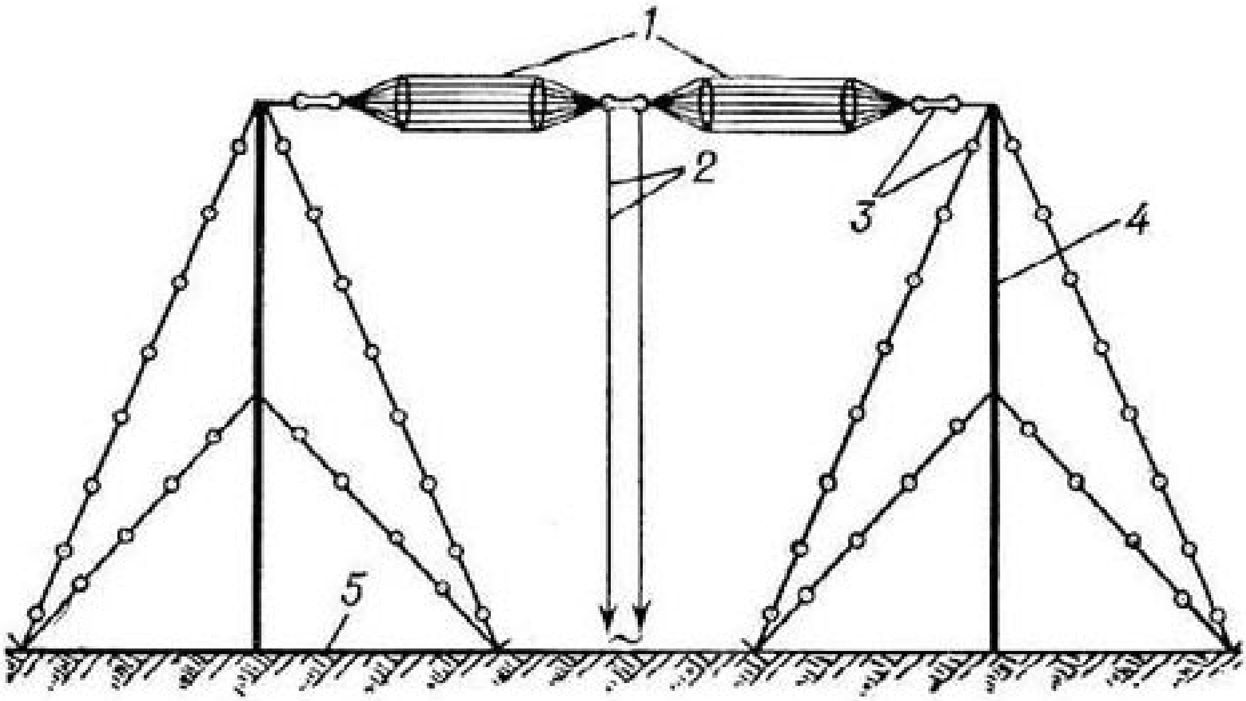
The issue is practically the same as for **Radio Antenna Engineering** by Lapport (see Antentop- 01- 2008).

“Antennas” by S. Nadenenko was published in Moscow in 1959 by Svyaz’izdat (the publisher). Under the laws for former USSR this initial copyright was valid for a term of 50 years, and expired in 2009. The copyright could be renewed in the last year (or before the term) of this initial term for a second term. Under the laws then in effect, this renewal was *not* automatic - the renewal had to be registered by the copyright holder in a timely manner and ISBN must be obtained.

However, I could not find out that the copyright of the book is renewed.

So, I believe that the copyright on **“Antennas” by S. Nadenenko** expired in the middle of 2009, and then the book passed into the public domain.

The E- book may be freely copied, distributed, pasted on the any site at the web BUT only without commercial purposes.



Dipole Nadenenko. First used by Nadenenko in 1935.

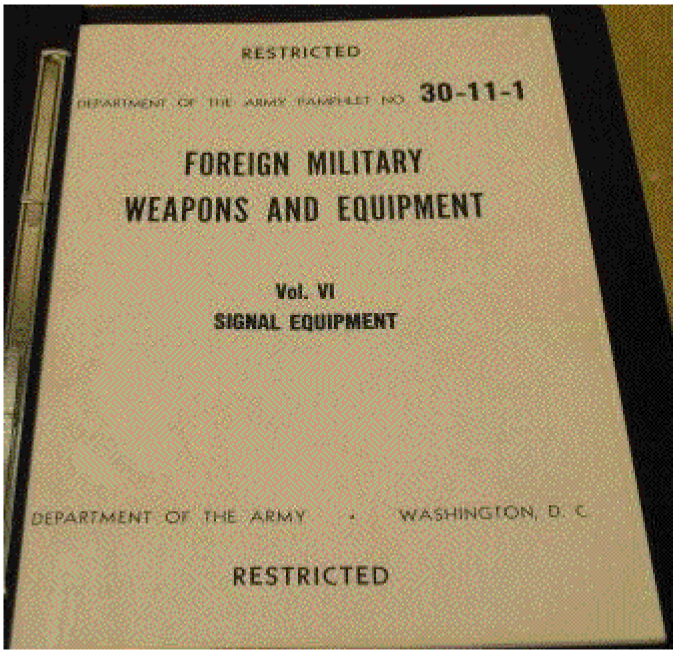
Foreign Military Weapons and Equipment (ex- USSR Euipment)

"Foreign Military Weapons and Equipment"
Volume VI
Signal Equipment
Department of the Army
Pamphlet no 30-11-1

The Pamphlet was published at 1951. At the time the book was a Secret, Restricted Book. However, at our times- 2009, there is no secret in the book. The book is freely sold at different points in the Internet. Just do googling and find out lots data.

I managed to find in the internet a part of the book in the pdf. The part is available in the Library at ***Antentop site***.

The parts contained 46 pages from original 81. The Pamphlet is very interested and as you can see it described the old USSR's Radio Equipment very well.



Credit Line for Picture:
http://www.wlhoward.com/id967.htm

You may found here:

Glossary of Russian Terms
Radio Set Type RBM-1
Radio Set Type RBM
Radio Set Type 9RS
Radio Set Type RSB-F
Radio Set Type A-7-A
Radio Set Type 13-R
Radio Set Type 12 RTM
Radio Set Type RSI-4
Radio Set Type RSI-4T
Radio Set Type RSI-6M-1
Radio Set Type US-4S
Radio Set Type US-4
Radio Set Type RB-45
Radio Set Type 9R
Radio Set Type RSI-3
Radio Set Type RSB
Radio Set Type 12 BR
Radio Set Type 4R (RBS)
Radio Set Type 5-AK-1M
Radio Set Type RBM-5
Radio Direction Finder PKV- 45

USSR's radio-operators at training (1960s)
(Photo from a newspaper)

Receiving Magnetic Loop Antennas

If you can read in ***Russian*** *you can download a free 56- pages e- book "Receiving Magnetic Loop Antennas" by Igor Grigorov, RK3ZK. The book is e-variant of a chapter from a paper book "Antennas for Radioamateurs printed in Russia". Some fragment of the book was translated in the English and was published (and, as I hope, will be published) at ANTENTOP. Other fragments of the book also going to translated in the English. Below you can see the Contents of the book and path to load the Russian variant.*

Receiving Magnetic Loop Antennas

By Igor Grigorov, RK3ZK.

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30. Magnetic Loop Antenna for the 160 meters

31. Magnetic Loop Antenna for the 160 - 40 meters

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33. Simplified VHF Magnetic Loop Antenna

34. Magnetic Loop TV Antenna for receiving in conditions of the strong reflected signal

35. Magnetic Loop Antenna – Low-Pass Filter

36. Magnetic Loop Antenna with a TV dipole antenna

37. Multichannel TV Antenna

38. TV Antenna with spiral vibrator

References

Cover of the book


Григоров И. Н.
(RK3ZK)

Приемные
Магнитные
Рамочные
Антенны

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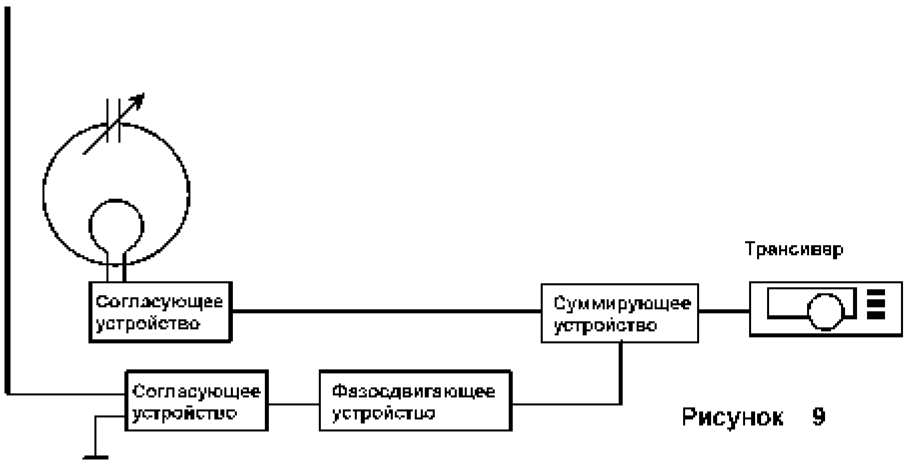


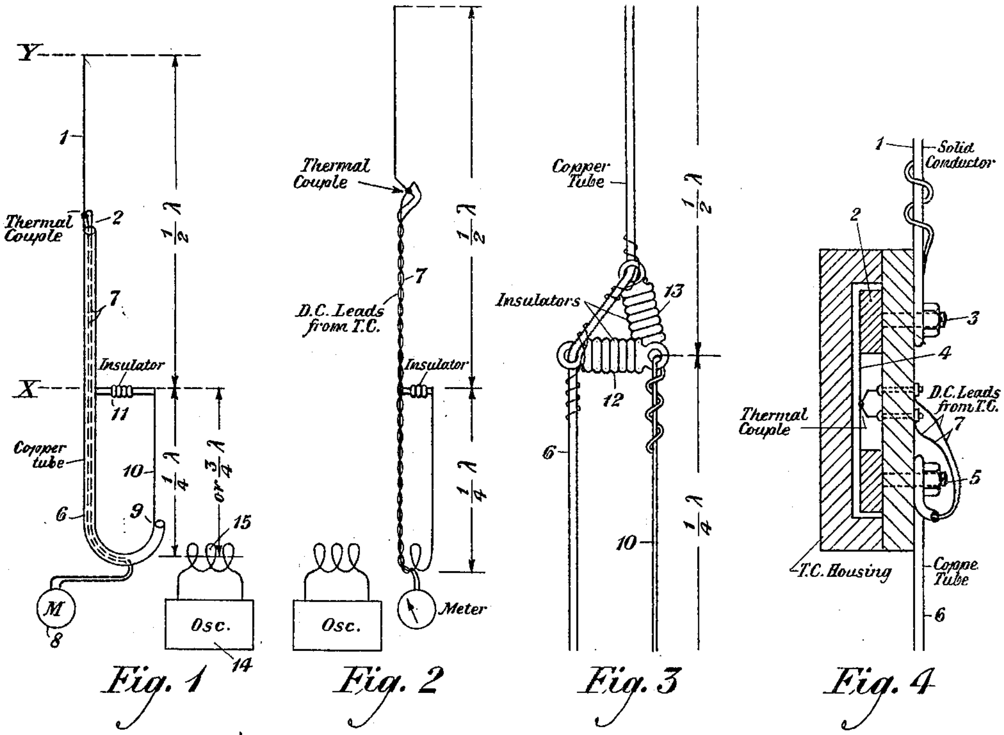
Рисунок 9

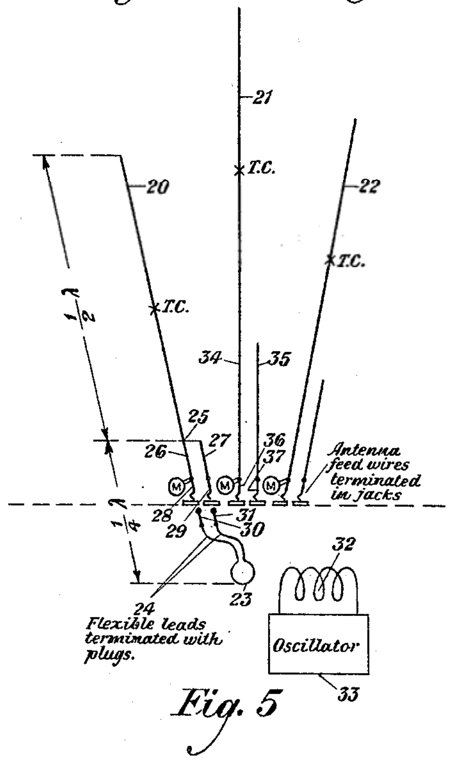
FIP8102 OR 1,831,921

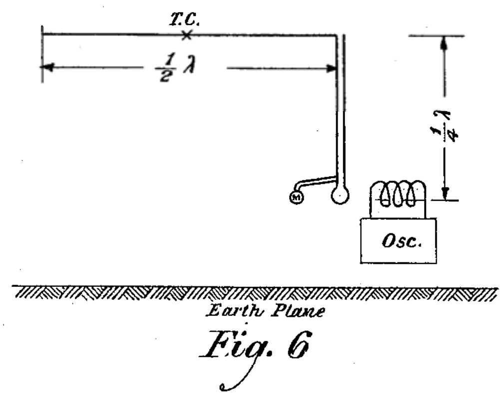
Nov. 17, 1931. DE LOSS K. MARTIN 1,831,921

SHORT WAVE RADIO ANTENNA SYSTEM

Filed May 24, 1927







INVENTOR.
D. K. Martin
BY
ATTORNEY

Patented Nov. 17, 1931 1,831,921

UNITED STATES PATENT OFFICE

DE LOSS K. MARTIN, OF WEST ORANGE, NEW JERSEY, ASSIGNOR TO AMERICAN TELEPHONE AND TELEGRAPH COMPANY, A CORPORATION OF NEW YORK

SHORT WAVE RADIO ANTENNA SYSTEM

Application filed May 24, 1927. Serial No. 193,892.

This invention relates to a short-wave radio antenna system, and particularly to means for determining the current at the midpoint of such a system and for connecting a plurality of such antennæ to a common terminating circuit.

In the operation of a short-wave antenna system, it is desirable to know the magnitude of the current at the midpoint of the system. In the past it has been customary to place a meter at that point, but since such antennæ are usually elevated well above the earth the reading of the meter has been difficult. In some instances a telescope has been employed.

One of the objects of this invention resides in a method by which the current strength at the midpoint of the short-wave antenna may be indicated at the station at which the signals are applied to the said antenna.

Another object of this invention consists in a method for terminating a plurality of antennæ, each of which is intended to operate upon a different wave length, the said method employing a terminating circuit which may be used with any of the antennæ without changing the effective length of the several antennæ.

Other objects of this invention will be apparent from the following description when read in connection with the attached drawings of which Figure 1 shows schematically one form of embodiment of the invention; Fig. 2 shows another form which differs from that of Fig. 1 in the manner in which the direct current leads are brought to the meter upon which the antenna current is indicated; Fig. 3 is a detail of construction of the form of the invention shown in Fig. 1; Fig. 4 shows the mode of connection of a thermocouple with the antenna of Fig. 1; Fig. 5 shows schematically a system comprising three antennæ, each designed to operate upon a frequency differing from that of the other antennæ, all of which antennæ are arranged for connection with a common terminating circuit; and Fig. 6 shows a horizontal type of short-wave antenna.

In the arrangement shown in Fig. 1, the conductor 1 may be supported in any manner that will insure proper insulation to ground. For example, it may be supported from a horizontal conductor by means of suitable insulators. The lower end of the conductor 1 is connected with one side of the heating element of the thermocouple 2, which is shown in detail in Fig. 4. This thermocouple is a well known translating device that has been found very useful in measuring high frequency currents by converting them into equivalent direct current potentials. As will be seen in Fig. 4, the conductor 1 is connected with the lug 3, which in turn is metallically connected with the heating element 4 of the thermocouple. The other end of the heating element is connected with a lug 5, with which is connected a tube 6 of copper or other suitable material which, as will be pointed out later, serves as part of the short-wave antenna and also as part of the transmission circuit, by means of which high frequency oscillations are carried from the operating room of the station to the antenna. The tube 6 also serves to carry the direct current leads 7 from the direct current side of the thermocouple to the meter 8 (shown in Fig. 1) and at the same time to effectively shield these direct current leads from extraneous interference. The lower end of the tube 6 is bent in the form of a U and is connected at the point 9 with a conductor 10 that serves as part of the transmission circuit. In order to effectively support and to insulate the upper end of the conductor 10 from the tube 6 an insulating device 11 is inserted therebetween. A form of the device 11 is shown in detail in Fig. 3. As will be seen from this figure, two or more insulators 12 and 13 are fixedly connected with the tube 6 and the conductor 10 and serve to maintain the two sides of the transmission circuit well insulated from each other. In order that a clear idea of the nature of this invention may be obtained it is important to note that the transmission circuit embraces that portion of the copper tube extending from the point X to the point 9 and the conductor 10 that extends upward from the tube at the point 9. This U-shaped circuit constitutes a loop by virtue of the distributed capacity and distributed inductance along the circuit, and oscillations will be set up in this loop by an oscillator 14, which is coupled with the loop by means of the coupling inductance 15. The antenna extends from X to Y and includes that portion of the tube 6 between X and the thermocouple 2 and the conductor 1. It is important to point out that this antenna may be one-half wave length long or any multiple thereof, such as one and one-half, two and one-half wave lengths. It is also important to point out that the length of the transmission circuit shall be one-fourth wave length or any other odd quarter, such as three-quarters, five-quarters etc. In the form of antenna and transmission circuit shown in Fig. 1 a loop of voltage is obtained at the ends of the antenna and a node of voltage at the midpoint thereof. This result is obtained in practical forms of the antenna when the far end of the transmission line is terminated in an input impedance which corresponds practically to an open circuited transmission line, and with the near end terminated in an impedance equal to the natural impedance of the line. As pointed out, to obtain that result the length of the transmission line may be any odd number of quarter wave lengths, and to obtain efficient radiation from the antenna its dimension should be exactly equal to a half wave length.

In the operation of the arrangement shown oscillations created by 14 will be impressed by the loop 15 upon the transmission circuit 6—10, and the voltage created therein will be impressed at the point X upon the antenna. By virtue of the relative dimensions stated, the voltage impressed upon the antenna by the transmission circuit at the point X will be the maximum voltage, and the wave set up in the antenna will be such that the node will occur at the midpoint of the antenna where the thermocouple is located. The ultra high frequency current flowing in the antenna will produce an equivalent direct current voltage by virtue of the heating effect produced in the thermocouple. This direct voltage will produce a current which will be transmitted over the leads 7 to the meter 8 and will indicate to the attendant at the station the magnitude of the current at the midpoint of the antenna.

The arrangement shown in Fig. 2 differs from that in Fig. 1 only in the mode of support of the direct current leads. In Fig. 2 they are twisted around the lower portion of the conductor that constitutes a part of the antenna and also a part of the transmission circuit.

Since an antenna of the form shown in Figs. 1 and 2 will radiate efficiently only when excited to the resonant frequency for which it is proportioned it is necessary, in order to be able to transmit on a plurality of different wave lengths, to employ antennæ of different lengths. On the other hand, since the oscillator at a station may be readily adjusted to produce any of a plurality of frequencies, it is desirable to use the same terminating circuit for each of the antennæ. The manner in which this may be accomplished is shown in Fig. 5, wherein 20, 21 and 22 represent antennæ, each of which differs in length and is intended to transmit a wave length differing from the others. The antenna 20 is connected at the point 25 with the conductor 26, which forms a part of one side of the transmission circuit. 27 forms part of the other side. These conductors are terminated in jacks 28 and 29 with which the plugs 30 and 31 are intended to operate. These plugs are connected by cords with the loop 23, which is inductively coupled with the loop 32 of the oscillator 33. In like manner, the conductors 34 and 35 connected with antenna 21 are terminated upon the jacks 36 and 37, and similarly, the conductors connected with antenna 22 are connected with jacks. All of these jacks are intended to cooperate with the plugs 30 and 31 so that oscillations from the source 33 may be applied to any of the antennæ by simply plugging the terminal circuit into the jacks of the desired antenna. The length of the portion of the conductors of the transmission circuit extending from the jacks toward the antennæ is such that when added to the length of the conductors of the terminal circuit 24 it renders the length of the entire transmission circuit equal to an odd quarter of the wave length desired to be transmitted over a particular antenna.

Fig. 6 shows that the invention may be employed in a system employing a horizontal antenna. In this case the transmission circuit is vertical and equal to an odd quarter of the wave length. In such an antenna the current may be measured by a thermocouple, as in the other forms shown, and the leads may be brought down to the measuring instrument either through a copper tube, as in Fig. 1, or by spiralling around the conductor, as shown in Fig. 2.

While the invention has been disclosed as embodied in particular forms, it is capable of embodiment in other and different forms without departing from the spirit and scope of the appended claims.

What is claimed is:

1. In a short-wave radio signaling system the combination with a transmission circuit, of a radiating member connected therewith, a translating device connected with the midpoint of the said radiating member to convert alternating current into equivalent direct current potentials, a meter and means to connect the said meter to the said translating device, the said connecting means being enclosed in a metallic tube constituting a part

of the said transmission circuit and part of the said radiating member.

2. In a short-wave radio signaling system the combination with a plurality of antenna systems each including a radiating member connected with a portion of a transmission circuit, of a loop having conductors for interchangeable connection with each of the transmission circuits, the length of the said loop being such that when added to that portion of each transmission circuit connected with the said radiating members makes the total length of the circuit thus created equal to an odd quarter of the wave length upon which the antenna connected therewith is intended to operate.

3. In a short-wave radio signaling system, the combination with a transmission circuit, of a radiating member connected therewith, a translating device connected with the midpoint of the said radiating member to convert alternating current into equivalent direct current potentials, a meter and means to connect the said meter to the said translating device, the said connecting means being effectively supported partly by the said transmission circuit and partly by the said radiating member.

4. In a radio signaling system, the combination with a source of oscillations of an antenna located at a considerable distance from the said source, a transmission circuit connecting the said source to the said antenna, a rectifying device connected to the said antenna to convert a portion of the alternating current into the equivalent direct current potential, a current measuring device associated with the said source of oscillations, and a circuit connecting the said current indicating device to the said rectifying device whereby the magnitude of the current in the said antenna may be indicated to an operator at the said source of oscillations.

In testimony whereof, I have signed my name to this specification this 23rd day of May 1927.

DE LOSS K. MARTIN.

Broadband High Frequency Sky- Wave Antenna

United States Patent [19]

Feigenbaum

[11] **Patent Number: 4,733,243**

[45] **Date of Patent: Mar. 22, 1988**

[54] **BROADBAND HIGH FREQUENCY SKY-WAVE ANTENNA**

[75] Inventor: **Bernard E. Feigenbaum**, Neptune, N.J.

[73] Assignee: **The United States of America as represented by the Secretary of the Army**, Washington, D.C.

[21] Appl. No.: **943,095**

[22] Filed: **Dec. 18, 1986**

[51] **Int. Cl.^{4}** **H01Q 11/10**

[52] **U.S. Cl.** **343/736;** 343/792.5; 343/860

[58] **Field of Search** 343/792.5, 731, 739, 343/829, 846, 859, 860, 806, 735–738, 740, 825, 828, 830, 845, 848

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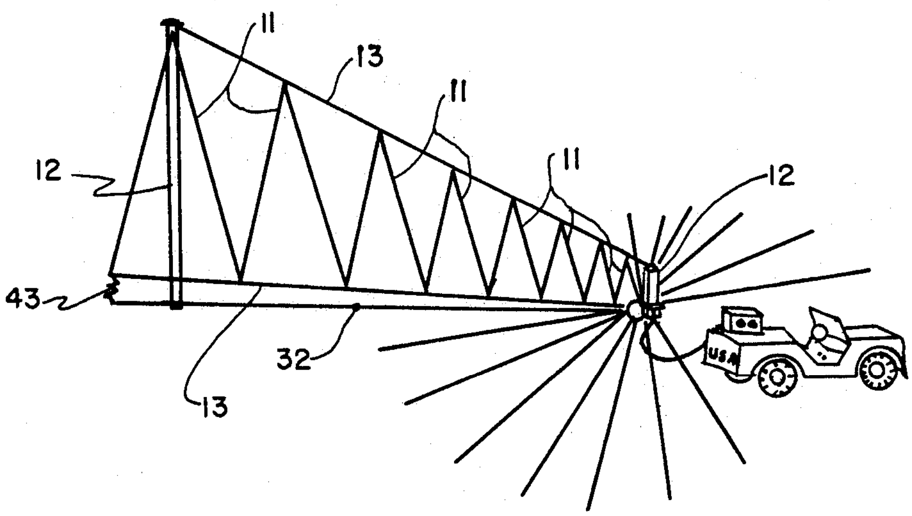
Primary Examiner—William L. Sikes
Assistant Examiner—Michael C. Wimer
Attorney, Agent, or Firm—Sheldon Kanars; Jeremiah G. Murray; John K. Mullarney

[57] **ABSTRACT**

A broadband H.F. directional sky-wave antenna system comprises a plurality of interconnected, vertical, zig-zag, antenna sections of predetermined increasing height. A counterpoise is utilized to balance the antenna. Power is coupled to the shortest zig-zag section via an impedance matching transformer. A resistance is used to terminate the antenna in its characteristic impedance.

3 Claims, 6 Drawing Figures



U.S. Patent Mar. 22, 1988 Sheet 1 of 2 4,733,243

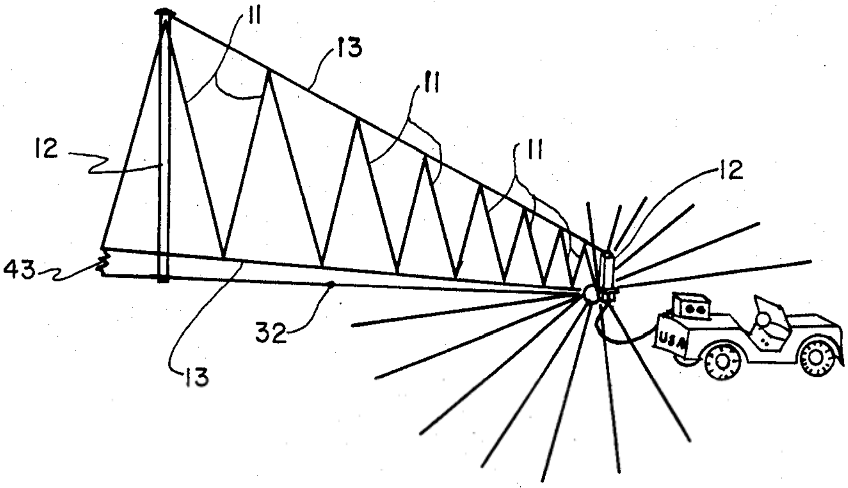


FIG. 1

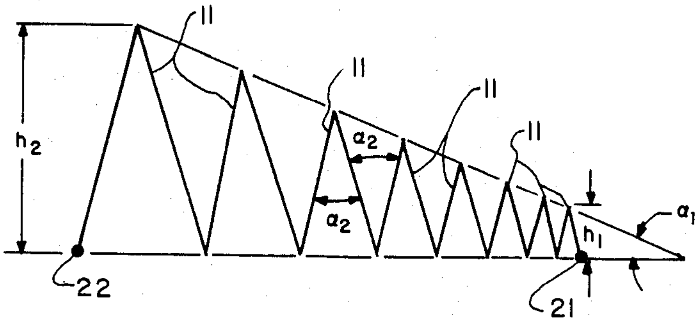


FIG. 2

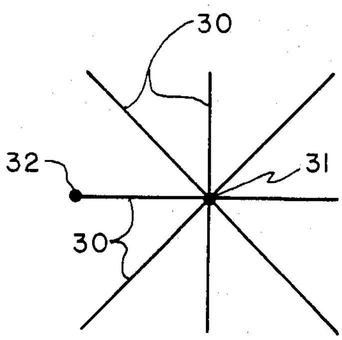


FIG. 3

U.S. Patent Mar. 22, 1988 Sheet 2 of 2 4,733,243

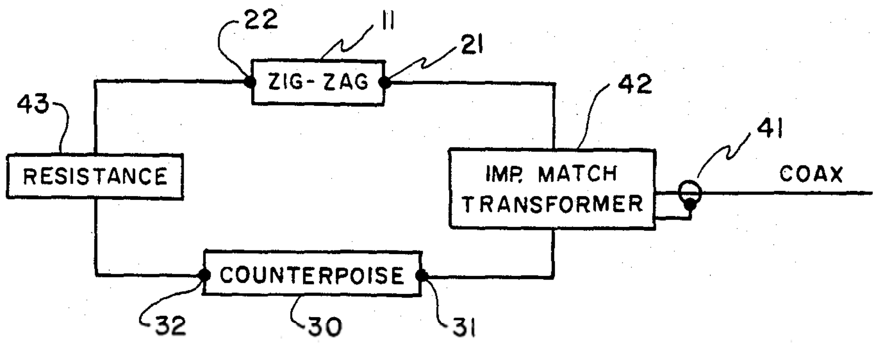


FIG. 4

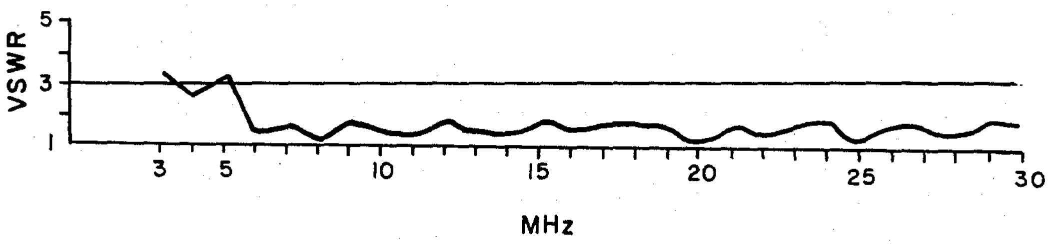


FIG. 5

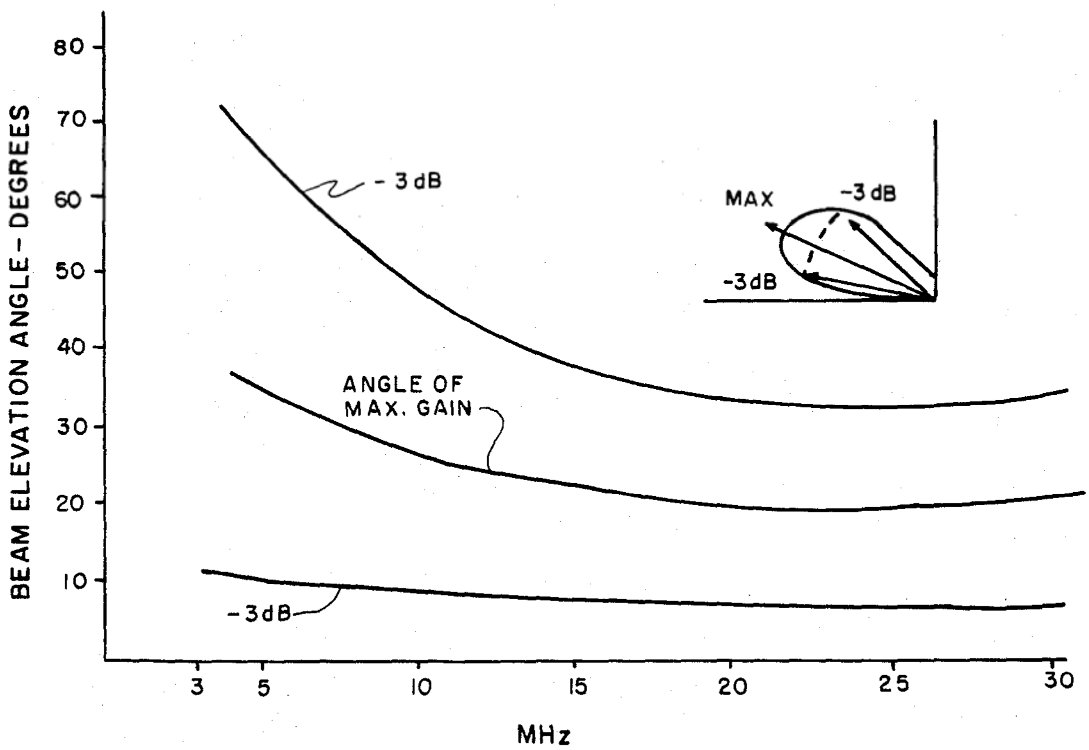


FIG. 6

BROADBAND HIGH FREQUENCY SKY-WAVE ANTENNA

The invention described herein may be manufactured, used and licensed by or for the Government for governmental purposes without the payment to me of any royalties thereon.

TECHNICAL FIELD

The present invention relates to a relatively small, broadband, high frequency, directional antenna for producing the low-angle radiation necessary for long-range communication by sky-wave propagation.

BACKGROUND OF THE INVENTION

Effective high frequency (HF) antennas, which are compatible with the varied and demanding requirements peculiar to military tactical communications, are by no means easily engineered. Military radios operate over a broadband (e.g., 3–30 MHZ) and vary in power from watts to kilowatts. The communication systems are often fixed, but must be transportable by vehicular, man-portable or airborne means and often necessitate ionospheric propagation over long range paths.

The required mobility would suggest small antenna size of light weight, but electrical performance will be compromised if the antenna is made too small in terms of the wavelength(s). The necessary broad frequency range which typically spans three octaves or more complicates the design of efficient antennas.

The conventional log-periodic dipole array (LPDA) generally offers good efficiency and broad bandwidth and has been used heretofore for military communication purposes. A commercially available log-periodic dipole antenna which has been utilized as the base station antenna for military communication purposes is the AS-317A/TSC-99 antenna made by Technology Communications International Co. (T.C.I.). While this antenna is satisfactory in the above-mentioned respects, it is unwieldy, difficult and time consuming to deploy, expensive, and because of its very large physical size the number of possible sites is restricted by political, economic, logistic, and other considerations (e.g., zoning approval).

SUMMARY OF THE INVENTION

It is the primary object of the present invention to achieve a broadband, high frequency (H.F.), directional, sky-wave antenna design that is small compared to current broadband H.F. sky-wave antennas.

A related object of the invention is to provide a sky-wave antenna that is relatively small, light-weight, inexpensive, and easy to deploy.

The foregoing and other objects are achieved in a preferred embodiment of the present invention wherein a given number (e.g., 10) of interconnected, vertical, zig-zag, antenna sections are of predetermined increasing height. The antenna is terminated by an appropriate resistance so as to maintain the characteristic impedance over the entire frequency range of interest. A counterpoise is utilized to balance the antenna system. Power is coupled to the shortest zig-zag section via an impedance matching transformer. The currents that are induced on the antenna cause power to radiate at low elevation angles.

It is an advantageous feature that, unlike the conventional vertical log-periodic antenna, the zig-zag antenna

of the present invention is a non-resonant, log-periodic, directional antenna which achieves a single main beam by virtue of satisfying a condition (i.e., Hanson-Woodyard) for significantly increased directivity.

BRIEF DESCRIPTION OF THE DRAWINGS

The invention will be more fully appreciated from the following detailed description when the same is considered in connection with the accompanying drawings, in which:

FIG. **1** shows a zig-zag antenna system in accordance with the present invention;

FIG. **2** shows the zig-zag antenna, per se;

FIG. **3** illustrates a typical counterpoise;

FIG. **4** is a block schematic diagram of the antenna system of the invention;

FIG. **5** shows the VSWR over the 3–30 MHz band for this zig-zag antenna system; and

FIG. **6** shows curves of take-off angle vs. frequency for the zig-zag antenna system of the invention.

DETAILED DESCRIPTION

Turning now to FIGS. **1** and **2** of the drawings, an antenna in accordance with the present invention is shown to comprise a plurality of vertical, zig-zag, antenna sections **11** of predetermined increasing height (h). While only eight interconnected zig-zag or triangular sections are shown in FIGS. **1** and **2**, the preferred embodiment of the invention comprises 10 sections±2 sections. However, it is to be understood that the invention is not so limited and might comprise even more or fewer zig-zag sections; the exact number of sections utilized will represent a compromise or trade-off between antenna size and performance. In the preferred embodiment, the shortest section has a height $h_1=1.0$ meter, while the height of the tallest section $h_2=10$ meters. The heights h_1 and h_2 are inversely proportioned to f_1 and f_2, the high and low frequencies, and thus are determined in great part by the frequency band to be propagated. As indicated in FIG. **2**, the increasing height of the successive zig-zag sections is defined by the angle α_1, which in the preferred embodiment is 30 degrees±5 degrees. The angle α_2 of the zig-zag sections is also preferable 30 degrees±5 degrees. However, once again, it is to be understood that these values for the angles α_1 and α_2 are given by way of example and the invention is not so limited. Further, α_1 is not necessarily equal to α_2. The angle chosen for α_2 should be a constant for a given antenna design. An antenna heretofore constructed in accordance with the invention had a length of approximately 20 meters.

A wire which can be readily utilized for the zig-zag antenna of FIGS. **1** and **2** is a commercially available 7-strand, No. 12 AWG (American Wire Gauge), made of Sn/Cu. The zig-zag antenna requires a pair of supporting masts **12** (FIG. **1**) and the antenna itself is strung between a pair of synthetic ropes **13**. Guying tensions are about 10 kilograms (22 lbs.). This compares quite favorably with a LPDA which typically requires a catenary tension of several hundred kilograms due to the large distance between supporting masts or towers.

A counterpoise such as shown in FIG. **3** is needed to balance the antenna system. The FIG. **3** counterpoise comprises four counterpoise wires **30**, the wire used being similar to the wire utilized for the zig-zg antenna. The counterpoise is preferably 20 or more meters square for good balance. However, as will be evident to those skilled in the art, the discrete counterpoise of

FIG. 3 may not be necessary in all instances. For example, if the chosen site consists of moist or damp soil then ground stakes may be substituted for the counterpoise of FIG. 3. For dry or sandy soil a discrete counterpoise is probably necessary.

Referrring now to FIG. 4, the zig-zag antenna **11** has a characteristic impedance of 600 ohms. If a 50 ohm coaxial transmission line **41** is used to deliver the power, a 12:1 BALUN (balanced-to-unbalanced) transformer **42** is used for impedance matching purposes. And, for a 75 ohm transmission line, an 8:1 transformer would be needed. The antenna is terminated with a 600 ohm resistance **43** (e.g., a power rated resistor or a load line) in order to maintain the characteristic impedance over the entire frequency range.

The zig-zag antenna receives power at the feedpoint **21** via the impedance matching transformer. The currents that are induced on the antenna cause power to radiate at low elevation angles. The radiating efficiency of the antenna is frequency dependent so the terminating resistance attenuates any portion of the power that is not radiated. The reference numerals **21** and **22** of FIG. 2 indicate the connection points of the zig-zag **11** to the transformer **42** and resistance **43** of FIG. 4. Also, in FIG. 4, the transformer **42** and resistance **43** are respectively connected to the counterpoise of FIG. 3 at points **31** and **32**.

Unlike all other known prior art antennas, the broadband (3–30 MHz), HF non-resonant, log-periodic, directional antenna of the present invention achieves a single main beam by virtue of satisfying the Hanson-Woodyard condition for increased directivity. The Hanson-Woodyard condition for an increased directivity array is one of the few instances where it is practical to obtain more directivity than would normally be obtained from an array of given size. For a brief discussion of the Hanson-Woodyard condition see the text entitled "Antennas" by J. D. Kraus, McGraw-Hill Book Co. (1950), page 81.

A broadband (3–30 MHz) zig-zag antenna constructed in accordance with the preferred embodiment of the invention exhibited the VSWR characteristic illustrated in FIG. 5 of the drawings. As is known to those skilled in the art, a low VSWR is desirable for maximum power transfer. Over the entire 3–30 MHz band the zig-zag antenna had a VSWR of about 3 or less—and from 6–30 MHz the VSWR$\leqq$2.

FIG. 6 shows curves of take-off angle vs. frequency for the preferred embodiment of the invention. For illustrative purposes, the curves have been smoothed from the raw data that was obtained. As will be evident to those skilled in this art, the curves indicate that the zig-zag antenna of the invention is excellent for medium to long-range (e.g., 4000 km) sky-wave communication.

The antenna of the present invention has constant impedance and radiation-pattern characteristics. Therefore, this zig-zag antenna is suitable for use with frequency hopping HF transmission, or other spread-spectrum techniques, as well as single frequency operation.

The performance of the zig-zag antenna of the invention is, at the least, comparable to that of the conventional vertical log-periodic antenna (VLP). Moreover, an antenna constructed in accordance with the present invention is only about 1/10 the size of a VLP; its weight is approximately 1/50 that of a VLP; and its cost is also about 1/50 of that of a VLP. Thus, unlike the VLP, the antenna of the invention is readily transportable (when collapsed), it can be set up by only one man, and it can be sited on a relatively small piece of land (e.g., $\leqq 100 \times 20$ feet). And, because of its weight it can be erected on even soft ground without the need of concrete reinforcements.

While a specific embodiment of the invention has been described in detail, it is to be understood that numerous modifications and variations therein may be devised by those skilled in the art without departing from the spirit and scope of the invention.

What is claimed is:

1. A broadband antenna system comprising a pair of support lines, one of said support lines being horizontal and the other at a vertical angle of 30°±5° to the horizontal, a length of antenna wire strung continuously between said support lines in a vertical zig-zag manner so as to present a plurality of interconnected inverted-V sections, said plurality being equal to ten±two, the legs of each inverted-V section meeting at the ends thereof with the ends of the legs of inverted-V sections immediately adjacent thereto, each inverted-V section providing a spatial phase reversal to the current from leg-to-leg, the angle between adjacent legs of the inverted-V sections being a constant equal to 30°±5°, impedance matching transformer means for coupling input signals to the shortest of the inverted-V sections, resistance means connected to the tallest of the inverted-V sections for terminating the antenna in its characteristic impedance, and counterpoise means coupled to said resistance means and said transformer means and serving to balance the zig-zag antenna.

2. An antenna system as defined in claim **1** wherein said zig-zag antenna is non-resonant.

3. An antenna system as defined in claim **2** wherein said zig-zag antenna is comprised of 7-strand, number 12 AWG wire having a copper core and tin cladding.

* * * * *

Multi Frequency Band Antenna

United States Patent [19] [11] **4,062,017**
Thompson [45] **Dec. 6, 1977**

[54] **MULTIPLE FREQUENCY BAND ANTENNA**

[76] Inventor: **Wallace T. Thompson,** 10528 Tomwood Ave., El Paso, Tex. 79925

[21] Appl. No.: **633,928**

[22] Filed: **Nov. 20, 1975**

[51] **Int. Cl.**^{2} **H01Q 1/14;** H01Q 21/28
[52] **U.S. Cl.** **343/722;** 343/727; 343/822; 343/886
[58] **Field of Search** 343/722, 820, 821, 822, 343/727, 816, 886

[56] **References Cited**

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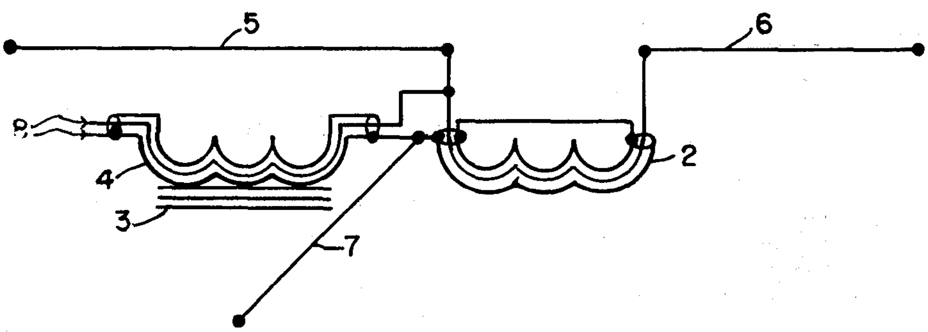
Primary Examiner—Eli Lieberman
Attorney, Agent, or Firm—Stevens, Davis, Miller & Mosher

[57] **ABSTRACT**

An antenna configuration in which multiple frequency band coverage is obtained in simple structures. This invention is most useful in the high-frequency spectrum where other techniques of providing multiple frequency operation become physically difficult to implement. The antenna configuration includes a multiplicity of conductor elements, a coaxial cable network for interconnection of the elements and connection to a feedline, and a high permeability core to aid decoupling to the feedline exterior from the antenna proper.

7 Claims, 9 Drawing Figures



U.S. Patent Dec. 6, 1977 Sheet 1 of 4 4,062,017

FIG. 1.

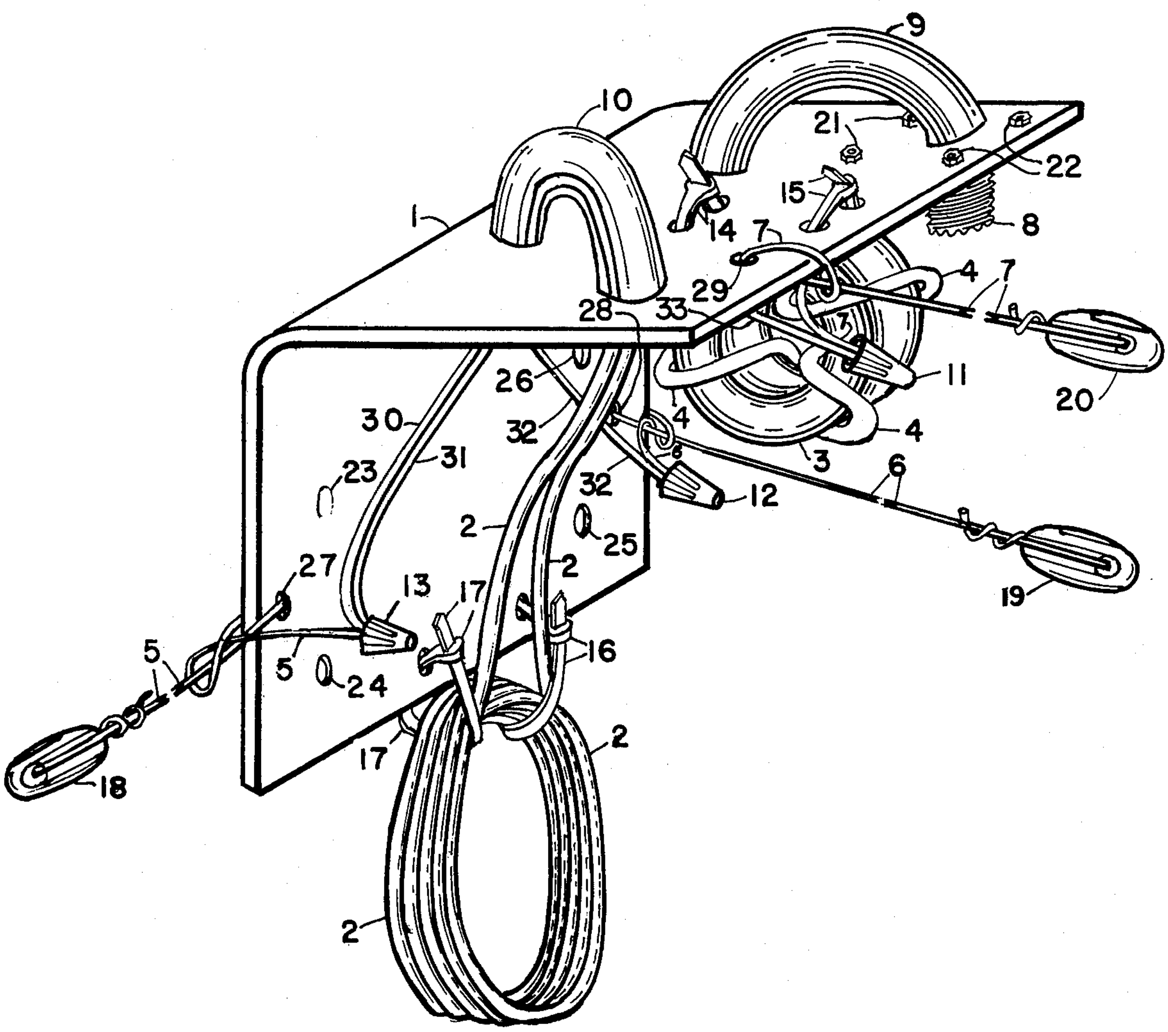


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FIG. 2.

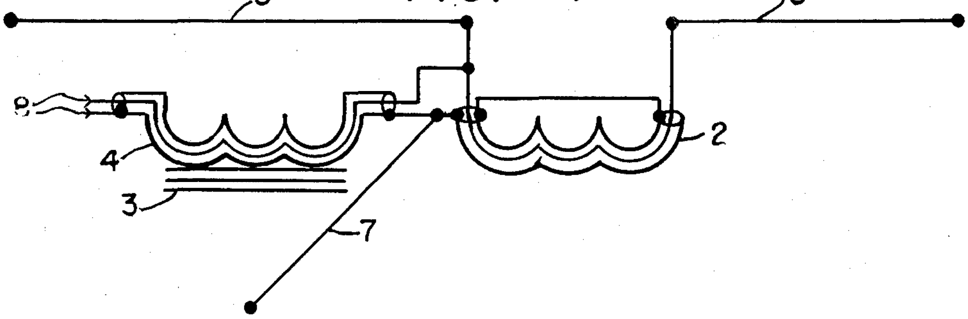


FIG. 3.

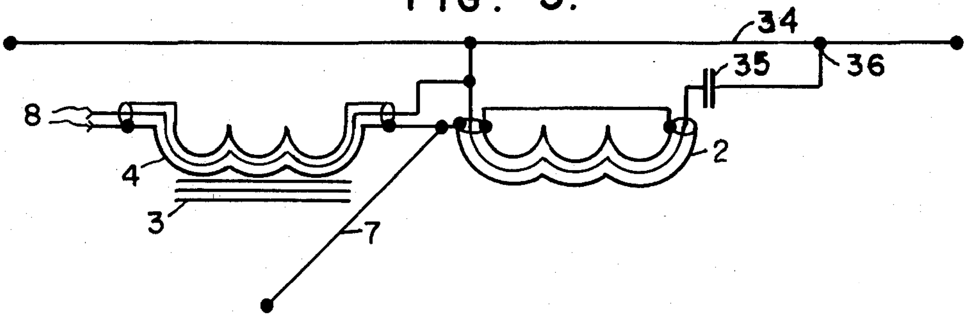


FIG. 4.

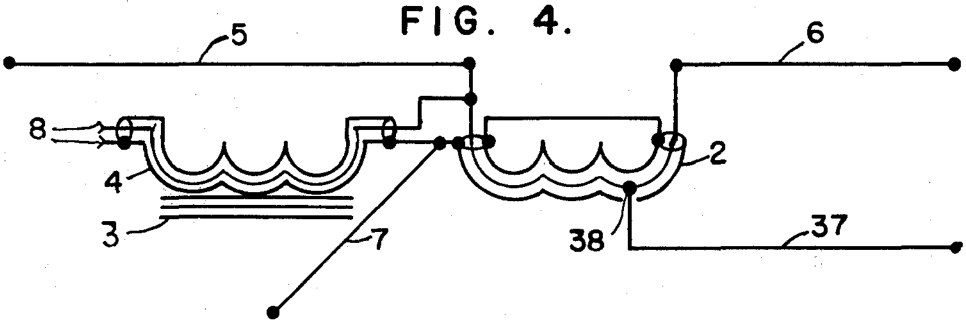


FIG. 5.

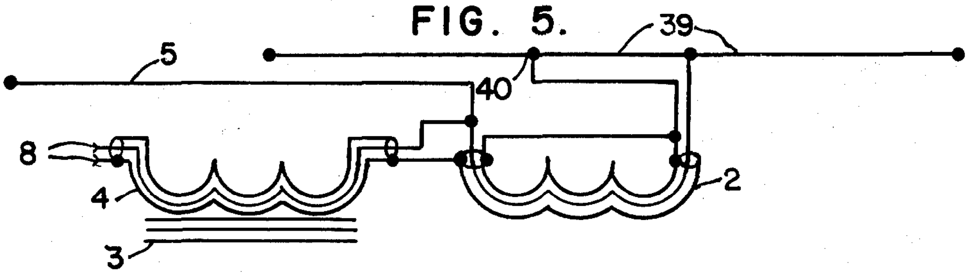


FIG. 6.

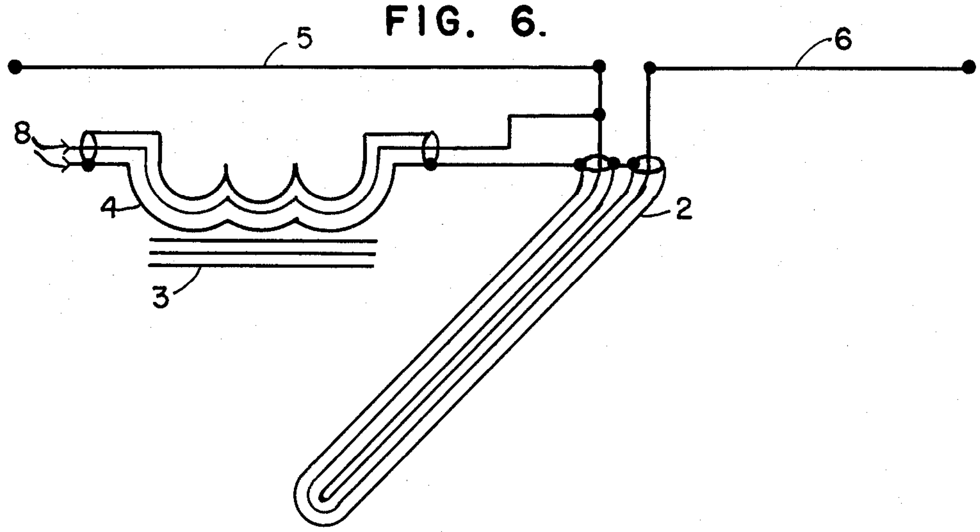


FIG. 7.

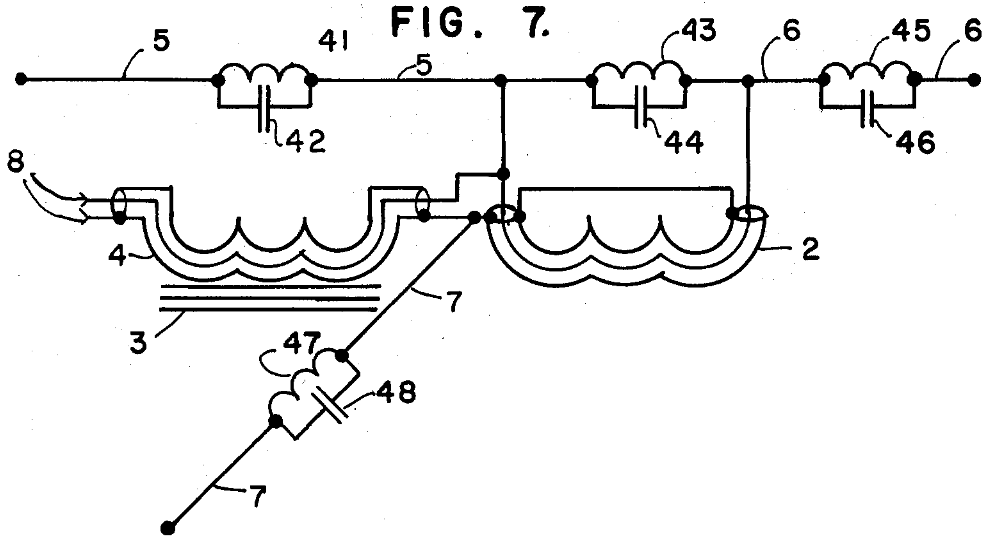


FIG. 8

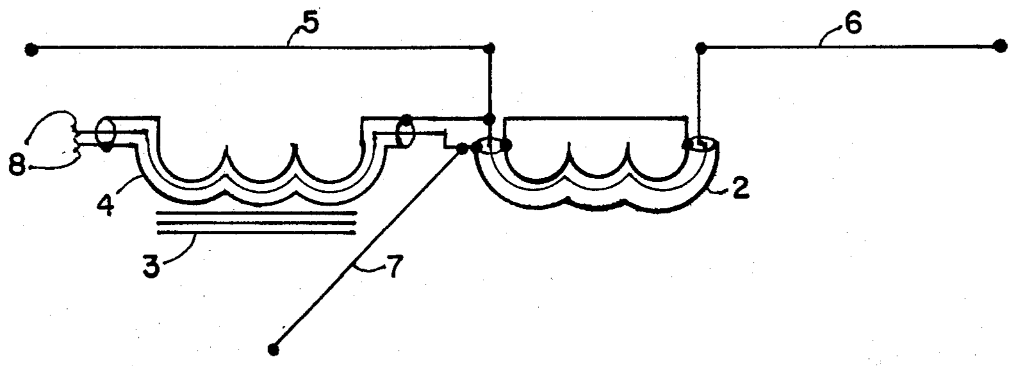
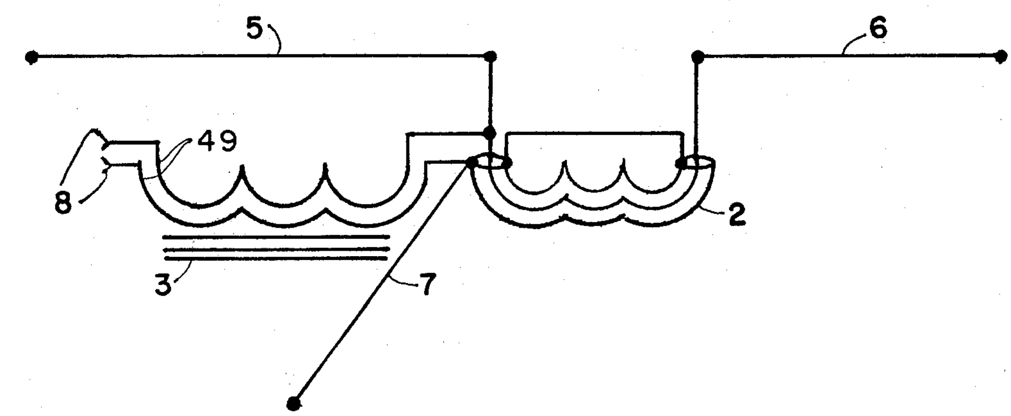


FIG. 9



MULTIPLE FREQUENCY BAND ANTENNA

BACKGROUND OF THE INVENTION

This invention relates to radio antenna apparatus and it particularly relates to an antenna configuration which is useful for providing multiple band and frequency coverage while employing a single feedline to supply and receive radio frequency energy.

One of the more difficult problems in radio antenna construction and performance is that of achieving a high degree of efficiency while maintaining a correct impedance match across a wide range of frequencies. An antenna element typically exhibits resonance wherein it absorbs energy from a source such as a transmission line and transmitter more readily at some frequencies than at others.

It is often desirable to have communications capability at various frequencies throughout the high frequency spectrum. A high frequency band antenna should be effective on as many frequencies as possible. Resonance behavior inhibits effective impedance matching between the transmitter and antenna. A structure which reduces the effects of resonance behavior will provide improved impedance matching. A typical definition of the useful bandwidth of an antenna would be those spectrum regions where the voltage-standing-wave-ratio is three-to-one or less. Some critical applications might call for two-to-one and some non-critical applications might well tolerate five-to-one or six-to-one values of voltage-standing-wave-ratio. Absolute perfection of a one-to-one voltage-standing-wave-ratio is rarely necessary and generally achieved only with antennas designed to operate on a single discrete frequency.

There are numerous versions of prior-art antenna configurations with wideband, multiple-band or all-frequency performance in the high-frequency spectrum.

Techniques which are employed include the use of manual tuning or servomechanism tuning so as to regain resonance and proper impedance matching. While effective, this technique is unwieldy or expensive.

Also included are multiple antennas, connected to a common feed point, each accepting radio frequency energy at its own resonant frequency. When the number of bands becomes high, this configuration becomes difficult to adjust and suffers from appearance and practical construction problems.

Further techniques are the use of tuned circuit elements or transmission line segments to act as bandstop networks which electronically disconnect portions of an antenna structure so that the remainder is resonant. These are commonly called "traps" and are effective for two or three band performance. Since the components employed are subjected to high voltage, care is required in the basic design to provide adequately rated components and humidity-resistant construction.

Prior art also includes the use of resistive termination elements, which, by sacrificing efficiency, can reduce reflected waves on the antenna and improve its behavior across frequency spans.

Recent technology includes the use of structures which show repetitive ratio construction with each fixed frequency change percentage. Known as "log-periodic" structures, these are probably the most successful in providing wide bandwidth performance. Such structures are generally neither simple nor inexpensive when implemented in the high-frequency region of the spectrum.

A review of available techniques shows there remains a need for a relatively simple, inexpensive structure which provides efficient radiation and moderately good impedance matching characteristics over a number of bands.

OBJECTS AND SUMMARY OF THE INVENTION

It is an important object of this invention to provide an antenna structure and design capable of useful operation in multiple radio service bands in the high-frequency spectrun.

It is also an object of this invention to provide an antenna structure characterized by simplicity, ease of construction and unobtrusive appearance.

It is also an important object of this invention to provide an antenna structure which is capable of operation in non-critical applications in the entire high-frequency spectrum of 3 Megahertz to 30 Megahertz.

It is an object of this invention to provide an antenna structure with a single feedline.

It is also an object of this invention not to use resistive elements as part of the antenna structure for the purpose of improving impedance matching by sacrificing efficiency.

It is also an important object of this invention to provide a design which achieves decoupling of the radio frequency energy in the antenna from the exterior of a coaxial-cable-feedline supplying that energy.

It is also an important object of this invention to provide a design which possesses the versatility of a degree of selection in frequency where the optimum operating bands are located, extending the useful frequencies above and below the high-frequency spectrum.

It is an object of this invention to allow a feed arrangement at a point nearer one end of the antenna structure.

It is also an important object of this invention to provide a design which is capable of adding mechanical guying strength to a radio tower.

Further purposes and objects of this invention will become apparent as the specification proceeds.

All of the foregoing objects are provided by an embodiment of my improved antenna design wherein the antenna comprises a multiplicity of elements connected to a common feed assembly and the feed assembly is comprised of a bracket providing mounting means, a coaxial-cable-network providing electrical connection means, a connector providing feedline attachment means, a high permeability core providing mechanical winding and electrical decoupling means, and associated hardware providing environmental protection means and means to secure the assembly together.

BRIEF DESCRIPTION OF THE DRAWINGS

Referring to the accompanying drawings, a particular embodiment of the present invention is illustrated, wherein:

FIG. 1 is an oblique perspective view showing my antenna invention.

FIG. 2 is a schematic diagram of my antenna invention.

FIG. 3 is a schematic diagram of one of the useful variations of the basic antenna design.

FIG. 4 is a schematic diagram of another useful variation of the basic antenna design.

FIG. 5 is a schematic diagram of another useful variation of the basic antenna design.

FIG. 6 is a schematic diagram of another useful variation of the basic antenna design.

FIG. 7 is a schematic diagram of another useful variation of the basic antenna design.

FIG. 8 is a schematic diagram of still another variation of the basic antenna design.

FIG. 9 is a schematic diagram of another variation of the basic antenna design.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENT

Referring to FIGS. **1** and **2**, the preferred embodiment of the antenna is shown in detail. FIG. **1** is an oblique view of the antenna while FIG. **2** is a schematic diagram of the antenna.

An insulating bracket **1**, made of a non-conducting, high-strength material suitable for exposure to weathering is employed to mount the components of the antenna. The insulator bracket **1** is attached to its support in turn by bolts or other means through mounting holes **23, 24, 25** and **26**. The wire radiating elements **5, 6** and **7** are secured to the insulator bracket **1** by connection through their respective tie holes **27, 28** and **29**. Connector **8** provides a means to attach a coaxial feedline to the antenna. A coaxial cable **4** exits connector **8**, passing through protective sleeve **9** and then is wrapped in multiple turns through toroidal core **3**. The toroidal core **3** is made of a material having high magnetic permeability. The toroidal core **3** and cable **4** assembly is secured to the insulator bracket **1** by two tie straps **14** and **15** which pass through appropriately located holes in the insulator bracket **1**. From the toroidal core **3**, the coaxial cable **4** continues through protective sleeve **10**. The center conductor **30** of coaxial cable **4** is connected to antenna wire **5** and conductor **31** inside wire connector **13**. The shield **33** of coaxial cable **4** exits protective sleeve **10** separately from the center conductor **30** and is connected to antenna wire **7** inside wire connector **11**. Conductor **31** is the center conductor of coaxial cable **2**. Coaxial cable **2** has an electrical length of substantially an integral multiple of one-half wavelengths at a frequency where antenna wires **5** and **6** together are half-wave resonant. The outer shield of coaxial cable **2** joins shield **33** of coaxial cable **4** inside protective sleeve **10**. Only shield **33** is brought out of sleeve **10**. Coaxial cable **2** passes through protective sleeve **10**, is coiled at the base of the insulator bracket **1** and the far end of coaxial cable **2** then returns into protective sleeve **10**. Coaxial cable **2** is attached to insulator bracket **1** by means of two tie straps **16** and **17**. The shield of coaxial cable **2**, at the far end, is also attached to shield **33** of coaxial cable **4**. The center conductor **32** of coaxial cable **2** at the far end is brought out of protective sleeve **10** and connects to antenna wire **6** inside wire connector **12**. The ends of antenna wires **5, 6** and **7** are terminated by insulators **18, 19** and **20** which are in turn used to attach support members which are not a part of the antenna proper. Connector **8** is secured to insulator bracket **1** by fasteners **21** and **22**.

It will be apparent to those skilled in the art that a number of equivalent construction techniques are apparent and included within the scope of this invention. These include, but are not limited to, such techniques as alternate fastening members to secure the components

in place, cylindrical rather than toroidal cores for the feedline decoupling choke, different lengths of coaxial cable **2** and rigid rather than wire antenna elements. The antenna may be also mounted with one or more elements vertical. The configuration of the preferred embodiment is chosen largely on the basis of low cost. Other configurations are suitable when needs dictate.

More specifically, a typical antenna constructed in accordance with my invention is believed to operate primarily as a half-wave dipole at higher wavelengths (such as the 80 and 40 meter bands) and as a **3, 5** and **6** half-wave antenna in the 20, 15 and 10 meter bands. At 160 meters, the antenna is believed to operate as a pair of monopoles fed by a quarter-wave phasing section and at 6 meters as a traveling wave antenna. While the theory of operation is not fully understood, it is believed that all three elements of the antenna are active and also interact on each band in such a way that there are generally two primary elements, the remaining element functioning as an impedance adjuster.

The following table shows which of the wire radiating elements **5, 6** and **7** are believed to function as the primary elements and summarizes what is believed to be their primary behavior:

BAND (METERS)	PRIMARY BEHAVIOR	PRIMARY ELEMENTS
160	Two monopoles 90° out of phase	5 and 6
80	Off-center fed "Windom" doublet	5 and 6
40	$\lambda/2$ Doublet	6 and 7
20	3 $\lambda/2$ Doublet	5 and 7
15	5 $\lambda/2$ Doublet	5 and 7
10	6 $\lambda/2$ Doublet	5 and 7
6	Traveling wave antenna	5 and 7

It will also be apparent to those skilled in the art that there are useful variations in the basic configuration of this device which are included within the scope of the invention. Thus, FIG. **3** illustrates a variation wherein two of the antenna wire elements **5** and **7** are combined into a single conductor **34** and energy is transferred into the structure through a capacitor **35** and tap **36** onto this single conductor. Such impedance matching schemes are commonly called gamma-matching sections. FIG. **4** illustrates a schematic configuration of the antenna wherein another conductor element **37** has been added to the basic structure by adding a tap **38** to an intermediate position of the coaxial cable **2**. Such taps allow the addition of specific conductor radiating elements to control the impedance characteristics at additional frequencies. FIG. **5** illustrates another configuration wherein two elements **6** and **7** are joined into a single element **39** and a tap **40** is used for energy transfer. This configuration allows the tap **40** to bring a direct-current ground to the antenna elements **39** and **5**. It will also be apparent to those skilled in the art that the configuration of FIG. **6**, wherein the coaxial cable **2** in FIG. **1** is uncoiled and positioned in place of wire element **7**, can be usefully employed to save construction material at the sacrifice of appearance considerations. Wire element **7** would be eliminated altogether in such a configuration. It will also be apparent to those skilled in the art that lumped reactive components **41, 42, 43, 44, 45, 46, 47** and **48** may be employed singly or together to allow a smaller size structure or to allow additional resonances, as shown in FIG. **7**. Traps or loading reactances may be inserted in the conductor elements of this antenna much as they are inserted in other, prior-art antenna structures. FIG. **8** illustrates the readily apparent variation wherein the connections between coaxial cable 4 and coaxial cable 2 are reversed from the connections shown in FIG. 2. Finally it is apparent to those skilled in the art that a parallel wire transmission line 49 may be substituted in place of coaxial cable 4 as shown in FIG. 9.

What is claimed is:

1. A multiple frequency band antenna for coupling to a feedline comprising:

a high permeability core;

a coaxial network including

first and second input terminals for connection to said feedline, and

first and second coaxial cables each comprising an electrically conductive shield having first and second ends and a center conductor having first and second ends, said first coaxial cable being wrapped about said high permeability core and having the first end of its center conductor and its shield connected to said first and second input terminals, the second end of the center conductor of said first coaxial cable being connected to the first end of the center conductor of said second coaxial cable, the conductive shields of said first and second coaxial cables being connected together; and

first, second and third conductive radiation elements, said first element being coupled to the junction between the first and second ends of the center conductors of said first and second coaxial cables respectively, said second element being coupled to the second end of the center conductor of said second coaxial cable and said third element being coupled to the junction between the shields of said first and second coaxial cables,

said second coaxial cable having an electrical length equal approximately to an integral multiple of one-half wavelengths at the frequency at which said first and second elements are half-wave resonant, said high permeability core decoupling energy in said antenna from the exterior of said feedline.

2. A multiple frequency band antenna as defined by claim 1 which further comprises a support structure and wherein said second coaxial cable is coiled and suspended from said support structure, said high permeability core also being suspended from said support structure.

3. A multiple frequency band antenna as defined by claim 1 which further comprises a fourth conductive radiation element attached between the first and second ends of the center conductor of said second coaxial cable.

4. A multiple frequency band antenna as defined by claim 1 which further comprises reactance components connected in series with said first, second and third conductive radiation elements.

5. A multiple frequency band antenna as defined by claim 4 wherein said reactance components comprise resonant circuits.

6. A multiple frequency band antenna as defined by claim 1 wherein the second end of the conductive shield of said first coaxial cable is connected to the first end of the conductive shield of said second coaxial cable.

7. A multiple frequency band antenna as defined by claim 1 wherein the first and second ends of the conductive shield of said second coaxial cable are connected together.

* * * * *

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