BRIEF COMMUNICATIONS

On the Effectiveness of Active Antennas

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One of the methods of miniaturization of receiving antennas at meter and longer wavelengths, which has been under intense investigation during recent years, involves the use of wideband transistor amplifiers, mounted directly at the antenna terminals [1, 2]. The references pertaining to active antennas are quite justified to point out the improvement in voltage developed across the low input impedance of the receiver which results from this method, as compared to a passive antenna, without an amplifier. This fact, however, is not an evidence of an improved effectiveness of the receiving system because the ultimate criterion of effectiveness which is the only valid criterion in the energy sense, is the ratio of powers P_s of the signal and P_n of the noise, referred to the input of the system, i.e., to the input of the amplifier in our case.

The amplifier noise in active antennas is frequently higher than the external (atmospheric) noise not only in short-wave band but also in lower frequency bands. This phenomenon is caused by both the low efficiency and the mismatch between the amplifier and the antenna proper, which is caused by small dimensions of the antenna.

The analysis of the noise characteristics, which is presented below, using the simple but most widely used active antenna (short, nonsymmetrical dipole with a high input impedance, wideband antenna amplifier (WAA), figure) as an example, indicated that great care must be exercised when using such small antennas in professional radio receivers.

For a specified voltage of the signal field and bandwidth of the receiver, the ratio P_s/P_n is proportional to the so-called sensitivity of the receiving antenna which is defined by the following formula [3] when the antenna impedance Z_A and the load impedance Z_L (in this case the amplifier input impedance Z_{amp}) are matched:

$$P = \frac{A_{ef}\eta}{T_A + T_L}, \tag{1}$$

where A_{ef} and η are the effective area and the efficiency of the antenna, T_A and T_L are the noise temperatures of the antenna and of the load, i.e., of the amplifier together with the noise of the feeder and of the receiver.

Let us extend this expression to the case when the antenna and the amplifier are mismatched. For this purpose we introduce the matching coefficients ζ_m which is equal to the ratio of power P_L, delivered by the antenna to any specified load, to the power $P_{L\,max}$, delivered to a matched load:

$$\zeta_c = \frac{P_L}{P_{L\,max}} = \frac{4R_A R_{amp}}{|Z_A + Z_{amp}|^2} \quad \frac{4G_A G_{amp}}{|Y_A + Y_{amp}|^2} \tag{2}$$

$$Y_A = 1/Z_A,\ Y_{amp} = 1/Z_{amp},\ G_A = \mathrm{Re}(Y_A),\ G_{amp} = \mathrm{Re}(Y_{amp}).$$

Taking into account the effect of mismatch on the power of signal and of external noise at the amplifier input, the expression for the sensitivity of an active antenna should be written in the following form

$$P = \frac{A_{ef}\eta\zeta_m}{T_A\zeta_m + T_L} = \frac{A_{ef}\eta}{T_A + T_L/\zeta_m} \tag{3}$$

$$T_A = T_{A\Sigma} + T_0(1-\eta), \tag{4}$$

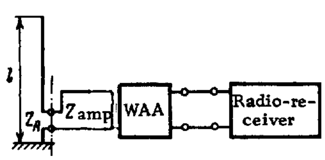
where T_0 is the standard temperature. The first term in (4) determines the contribution of external noise sources and the second term is due to ohmic losses.

In the short-wave band and lower frequency bands the dominant external noise source is the atmosphere, provided, of course, that there is no noise generated by the transmitter. Then

$$T_{A\Sigma} = \eta T_b, \tag{5}$$

where T_b is the brightness temperature of the atmospheric interference. When the power transfer coefficient of the amplifier is sufficiently high, we can neglect the feeder and the receiver noise, i.e., $T_L = T_{amp}$ (the noise temperature T_{amp} of the amplifier is determined by connecting a matched generator to its input terminals [4]). As a result of this we have

$$P = \frac{A_{ef}\eta}{T_b\eta + T_0(1-\eta) + T_{amp}/\zeta_m} = \frac{A_{ef}}{T_b + T_0\frac{1-\eta}{\eta} + \frac{T_{amp}}{\eta\zeta_m}}. \tag{6}$$



As an illustration, let us discuss the experimental electrical characteristics of one of the samples of an active antenna (l = 0.5 meters) at the frequency of 3 MHz. The input impedance of the antenna was Z_A = 4 - i 6900 ohms (the radius of the conductor a = 0.005 meters), i.e., the efficiency $\eta = R_{\Sigma A}/R_A = 2{,}5 \cdot 10^{-3}$. The resistive component of the input admittance of the amplifier was $2 \cdot 10^{-5}$ cm, the input capacitance was 5 pF and the noise temperature $T_{amp} = 3T_0$. The matching coefficient is $\zeta_m = 10^{-4}$ and consequently, $T_{amp}(\eta\zeta_m) = 10^{+7} T_0$. The brightness temperature in the European part of the USSR varies with time in the range 25-60 dB, with respect to T_0 [5], i.e., $T_b = 10^4 - 3 \cdot 10^8$K. Therefore the ratio $T_{amp}(\eta\zeta_m)$ will be greater than the brightness temperature T_b of atmospheric interference for a greater part of the year because the product $\eta\zeta_m$ is small. Then the ratio P_s/P_n will no longer depend only on the ratio of the directive gain to the external noise power, as it is commonly assumed for short-wave and lower frequency bands.

Expression (6) is applicable directly to the active antenna shown in the figure. Expressions for sensitivity of other circuit configurations (such as, for example, when a transistor is connected directly across the gap of the dipole) can be obtained by using the method for design of the entire receiving system, proposed in [2], and by taking into account efficiency of each radiating conductor and its match to a proper load.

Since a decrease of electrical dimensions of an antenna decreases its efficiency and the matching coefficient with the amplifier, the joint effect of these factors imposes a limit on the possible miniaturization of the antenna when the required energy potential of a radio line is specified. This should be a guiding factor in design of active antennas and in determination of the possible range of their applicability.

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Radiation from a Charged Particle, Passing Through a stack of Dielectric Plates in a Waveguide

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Reference 1 examined the radiation of a charged particle which passes through an infinite, stratified dielectric that fills out a waveguide. This radiation can serve as a basis for generation of microwaves and therefore it would be interesting to investigate the radiation obtained when a charged particle passes through a stack of a finite number of dielectric plates because such a configuration will most closely resemble a physical radiating system.

Let us consider a regular cylindrical waveguide of an arbitrary cross section, filled by a finite number of dielectric plates of thickness d ($\varepsilon = \varepsilon$), spaced at intervals a ($\varepsilon = 1$). The origin of coordinates is placed at the first boundary of the stack. Suppose that a charged particle q passes through the stack of plates with a velocity $\vec{v} = \vec{v}_z$ = const. We shall use the vector = potential $\vec{A}$ which, because of the symmetry of the problem, will have only one non-zero component $\vec{A}_z$ that satisfies the equation

$$\Delta A_{z,\omega} + \frac{\omega^2}{c^2}\varepsilon A_{z,\omega} = -\frac{4\pi}{c} j_\omega. \tag{1}$$

Here and below the subscript ω indicates the Fourier component of the corresponding quantity.

The solution of (1) will be obtained by expanding the vector-potential into a series of eigenfunctions $\Psi_n(x, y)$ of the first boundary problem for the transverse cross section of the waveguide. Writing the radiation field in each section in the form of waves propagating to the right and to the left and using the characteristic matrix of a single period [3], we obtain the following expressions for the coefficient of reflection $r_{n,N}$ and coefficients of transmission $t_{n,N}$ of the n-th mode through the stack of N plates:

$$r_{n,N} = \frac{1}{1+4(R_{n,N})^{-1}}; \qquad t_{n,N} = \frac{1}{1+\frac{1}{4}R_{n,N}}, \tag{2}$$

where

$$R_{n,N} = \left(\frac{\gamma_n}{\varepsilon\Gamma_n} - \frac{\varepsilon\Gamma_n}{\gamma_n}\right)\sin^2\gamma_n d\, U^2_{N-1}(\xi_n),$$

$$\xi_n = \cos\gamma_n d\cos\Gamma_n a - \frac{1}{2}\left(\frac{\gamma_n}{\varepsilon\Gamma_n} + \frac{\varepsilon\Gamma_n}{\gamma_n}\right)\sin\gamma_n d\sin\Gamma_n a,$$