

# The Design of Calculable Standard Dipole Antennas in the Frequency Range of 1 - 3 GHz

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## Abstract

This paper presents the design of a calculable standard dipole antenna with a 3 dB hybrid balun in the frequency range of 1 GHz to 3 GHz. A new formula of an antenna factor is derived by using the power mismatch-loss concept. The antenna factors derived in this paper are in good agreement with the results calculated from S-parameters. The design results show that the calculable dipole antenna with a hybrid balun can be characterized by the power mismatch-loss component factors.

**Keywords :** Calculable dipole antennas, Antenna factor, Method of moments, Hybrid balun

## 1. Introduction

Standard dipole antennas are necessary to measure antenna factors and to generate the standard field. The calculable dipole antennas have been developed and used as the standard antennas [1], [2]. The half-wavelength resonance dipole is one of the most basic antennas because the accurate calculation of the antenna characteristics can be done usually by using method of moments (MoM). Extensive works have been carried out in the development of a calculable standard dipole antenna [3],[4].

The previous works dealt with the analysis of a calculable dipole antenna factor by using S-parameters [1]-[4]. In this paper, a new formula of the antenna factor including the power loss component factors for a calculable dipole antenna with a 3 dB 180 degree hybrid balun is derived and the design of a calculable dipole antenna in the frequency range of 1 GHz to 3 GHz is presented. The formulation presented in this paper is derived by using the power mismatch-loss concept [5],[6]. The antenna factors obtained by employing the power mismatch-loss concept are in good agreement with the factors derived by S-parameters [1]. From the results of theoretical analysis presented in this paper, we can readily design the calculable dipole antenna including the power loss components. Calculated input impedances of the dipole antenna are compared with the measured results.

## 2. Description of the Dipole with a Hybrid Balun

Fig. 1 shows the basic structure of the calculable dipole antenna with a 3 dB hybrid coupler and two phase-matched coaxial lines. The dipole antenna with a length of  $L$  and a radius of  $a$  is placed along the  $z$ -axis. The balun is designed so that its complex S-parameters can be easily measured. Two semi-rigid cables with a length of  $L_B$  of the 3 dB hybrid balun are connected to the antenna terminal as shown in Fig. 1. A  $50 \Omega$  load is connected to the sum port ( $\Sigma$ ) of the hybrid, and a matched measuring instrument is connected to the other port ( $\Delta$ ) via the coaxial cable with a length of  $L_R$ . The inner conductors of the two coaxial lines are connected to the balanced dipole elements. The outer conductors of the coaxial cables are in contact with each other electrically, i.e. short-circuited. Since this structure is perfectly symmetrical, the two matched output voltages have the same amplitude and a phase difference of  $\pi$  radian.

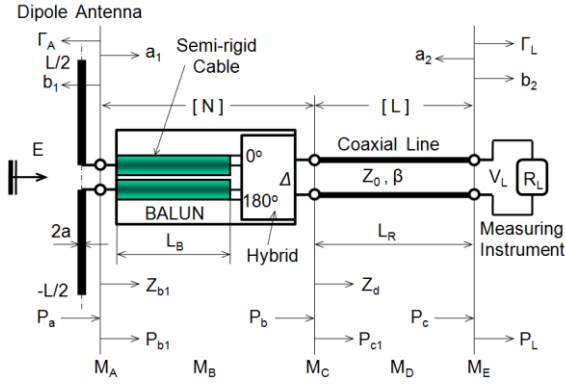


Figure 1: The basic structure of the dipole antenna with a 3 dB hybrid balun.

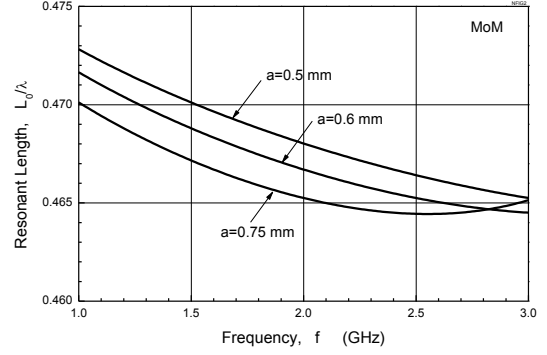


Figure 2: The resonance length of the standard dipole antenna.

### 3. Antenna Factor Expressions

A plane wave is incident on the calculable dipole antenna as shown in Fig. 1. If  $h_e$  is the effective length of a receiving antenna with a hybrid balun, and  $Z_a = R_a + jX_a$  is the antenna input impedance, then the power available from the receiving antenna can be expressed as

$$P_L^{ava} = \frac{|E|^2 |h_e|^2}{4R_a} \quad (1)$$

where  $|E|$  is the incident electric field.

If  $V_L$  is the input voltage to a measuring receiver connected to the antenna, and  $Z_L = R_L + j0$  is the input impedance of the receiver, then the power delivered to the measuring receiver can be expressed as

$$P_L = \frac{|V_L|^2}{R_L} \quad (2)$$

Assuming there are only passive devices between the receiving antenna and the measuring receiver, and the cable loss is zero, the total power losses can be represented as follows:

$$\frac{P_L^{ava}}{P_L} = \frac{|E|^2 |h_e|^2 R_L}{4R_a |V_L|^2} = M_A M_B M_C M_D M_E \quad (3)$$

where

$$M_A = \frac{P_a}{P_{b1}}, \quad M_B = \frac{P_{b1}}{P_b}, \quad M_C = \frac{P_b}{P_{c1}}, \quad M_D = \frac{P_{c1}}{P_c}, \quad M_E = \frac{P_c}{P_L} \quad (4)$$

In (4),  $P_a$  is the power available from the antenna,  $P_{b1}$  is the power delivered to the hybrid balun,  $P_b$  is the power available from the hybrid balun,  $P_{c1}$  is the power delivered to the coaxial cable,  $P_c$  is the power available from the coaxial cable, and  $P_L$  is the power delivered to the receiver input terminal.

From (3), the desired antenna factor  $|E/V_L|$  can be expressed as

$$AF = \sqrt{\frac{4R_a}{|h_e|^2 R_L}} \sqrt{M_A M_B M_C M_D M_E} = AF_0 K \quad (5)$$

where  $AF_0 = \sqrt{4R_a/|h_e|^2 R_L}$  is the antenna factor when the receiver is directly connected to the antenna.  $K$  is the power mismatch-loss factor expressed as  $K = \sqrt{M_A M_B M_C M_D M_E}$  and individual power mismatch-loss factors are as follows:

$$M_A = \frac{|Z_{b1} + Z_a|^2}{4R_a R_{b1}}, \quad M_B = \frac{4R_T R_{b1}}{|Z_{b1} + Z_a|^2} \left| \frac{Z_{11} + Z_a}{Z_{12}} \right|^2, \quad M_C = \frac{|Z_T + Z_a|^2}{4R_T R_d} \quad (6a)$$

$$M_D = \frac{4R_t R_d}{|Z_T + Z_a|^2} \left| \left( \frac{Z_0 + Z_T}{Z_0} \right) \frac{1 - \Gamma_S e^{-2\gamma L_R}}{2e^{-\gamma L_R}} \right|^2, \quad M_E = \frac{|Z_t + R_L|^2}{4R_t R_L} \quad (6b)$$

where  $Z_T$  is the Thevenin's equivalent impedance seen from the output terminal of the hybrid balun into the antenna and  $Z_t$  is the Thevenin's equivalent impedance seen from the output terminal of the coaxial cable into the antenna as shown in Fig. 1. Also,  $Z_{b1}$  is the input impedance seen from the input terminal of the hybrid balun into the receiver and  $Z_d$  is the impedance of the coaxial cable seen from the output terminal of the hybrid balun into the receiver.  $L_R$  is the length of the coaxial cable, and  $\gamma = \alpha + j\beta$ , where  $\alpha$  and  $\beta$  are the attenuation coefficient and phase constant of the coaxial cable, respectively.

In this paper, we also derive the antenna factor by using the scattering matrix (S-parameters), which is used earlier by NPL [1]. The antenna factor using S-parameters is expressed as

$$AF = \frac{(Z_a + Z_0) [(1 - T_{11}\Gamma_A)(1 - T_{22}\Gamma_L) - T_{12}T_{21}\Gamma_A\Gamma_L]}{Z_0 h_e (1 + \Gamma_R) T_{21}} \quad (7)$$

where  $T_{ij}$  are the S-parameters of the hybrid balun and the coaxial cable used for connection to the measuring receiver. The cascading matrix  $[T]$  of the composite 2-port is expressed as follows:

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} N_{11} & N_{12} \\ N_{21} & N_{22} \end{bmatrix} \begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix} \quad (8)$$

where  $N_{ij}$  and  $L_{ij}$  are the S-parameters of the hybrid balun and the coaxial cable, respectively, as shown in Fig. 1, and  $N_{11}$ ,  $N_{21}$  are normalized to  $100 \Omega$ , and  $N_{12}$ ,  $N_{22}$  are normalized to  $50 \Omega$ .

$Z_0$  is the characteristic impedance of the coaxial cable,  $\Gamma_A$  and  $\Gamma_L$  are the reflection coefficients of the antenna and the measuring receiver, respectively. Since  $Z_0/(Z_a + Z_0) = (1 - \Gamma_A)/2$ , the antenna factor of (7) is identical with the antenna factor expression in [1]. The calculated results confirm that the antenna factor of (5) derived by power mismatch-loss concept gives the same result as the antenna factor of (7) derived from S-parameters.

#### 4. The Design of Standard Dipole Antennas

In the design of the standard dipole antenna, a thin wire kernel approximation with a segment length of  $0.0125\lambda$  is used for the piecewise sinusoidal Galerkin's MoM analysis. The dipole radius was chosen to be less than  $0.0075\lambda$  (thin wire approximation) and the nominal value of  $50 \Omega$  is used for the characteristic impedance  $Z_0$ . The coaxial cable with a length of 10 m is selected in the following numerical calculations.

Fig. 2 shows the frequency characteristics of the resonance length with respect to the dipole radius. As shown in Fig. 2, the resonance length of the dipole antenna almost linearly decreases with the frequency increases, but it increases gradually above 2.5 GHz when  $a = 0.75$  mm.

Fig. 3 shows the frequency characteristics of the antenna input resistance at resonance length for the dipole radius. As shown in Fig. 3, the input resistance of the antenna increases with the frequency. At the frequencies below 2 GHz, the input resistance of the antenna has very little variation less than 1.9%. The maximum variation of the input resistance is 5.7% at 3 GHz.

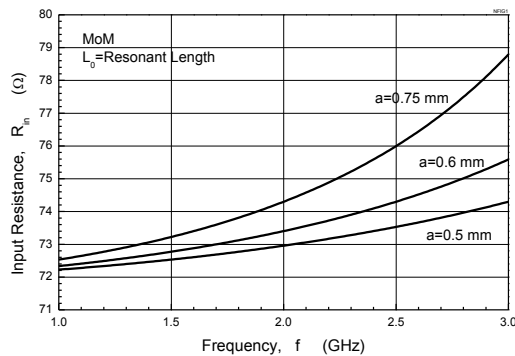


Figure 3: The input resistance of the standard dipole antenna at resonance.

Table 1: Calculated and measured antenna input impedances

Radius $a/\lambda$	Calculated $Z_{in}$ ( $\Omega$ )		*) Measured $Z_{in}$ ( $\Omega$ )
	MoM	EMF	
2.98 $\times 10E-3$	93.44 +j 4.86	73.13 +j 1.42	91.21 + j 43.26
3.97 $\times 10E-3$	97.18 +j 43.58	73.12 +j 41.05	96.42 + j 43.54

\*) R. W. P. King, "Tables of antenna characteristics", Plenum Press, 1971.

To check the validity of the theoretical analysis for the dipole antenna design, the antenna input impedances were compared with those of experiments as shown in Table 1. In Table 1, also shown are the input impedances calculated by an induced electromotive force (EMF) method. The results show that the MoM results are in better agreement with the experiments than EMF results.

Table 2 shows the theoretical antenna factor in the frequency range from 1 GHz to 3 GHz. In Table 2, the power mismatch-loss factors,  $K$ , are also shown for comparison with the desired antenna factor  $AF$ . As can be seen in Table 2,  $\sqrt{M_A}$ ,  $\sqrt{M_B}$ ,  $\sqrt{M_C}$ ,  $\sqrt{M_D}$ , and  $\sqrt{M_E}$  have the same value except for the sign. The mismatch-loss factor,  $\sqrt{M_E}$ , is only contributing to the desired antenna factor as shown in Table 2, i.e., the amount of  $AF - AF_0$  in dB is equal to the  $\sqrt{M_E}$  which is the mismatch-loss factor of the coaxial cable and the measuring receiver. The results show that the frequency characteristics of the theoretical antenna factor with respect to the value of the dipole radius are overlapped for the three dipole radii of 0.5 mm, 0.6 mm, and 0.75 mm. The differences between the antenna factors,  $AF$  and  $AF_0$ , are in within 0.11 dB.

Table 2: Calculated antenna factors by the MoM

$f$ (GHz)	$a$ (mm)	$L_0$ (cm)	$Z_{in}$ ( $\Omega$ )	Mismatch-Loss Factor					Antenna Factor (dB)	
				$K$ (dB)					$AF_0$	$AF$
				$\sqrt{M_A}$	$\sqrt{M_B}$	$\sqrt{M_C}$	$\sqrt{M_D}$	$\sqrt{M_E}$		
1.0	0.75	14.10	72.53	0.111	-0.111	0.111	-0.111	0.111	28.09	28.20
1.5	0.75	9.343	73.22	0.105	-0.105	0.105	-0.105	0.105	31.61	31.71
2.0	0.75	6.979	74.30	0.095	-0.095	0.095	-0.095	0.095	34.10	34.20
2.5	0.75	5.573	76.00	0.082	-0.082	0.082	-0.082	0.082	36.04	36.12
3.0	0.75	4.651	78.79	0.062	-0.062	0.062	-0.062	0.062	37.62	37.68

## 5. Conclusions

A new formula of an antenna factor is derived using the power mismatch-loss concept and the design of the standard dipole antenna with a hybrid balun in the frequency range of 1 GHz to 3 GHz has been presented. The numerical results show that the antenna factor of the calculable dipole antenna including the power loss component factors can be characterized by the power mismatch-loss concept and the MoM analysis.

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