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

Ki-Chai Kim¹ · Sang-Myeong Kim¹ · Jae-Yong Kwon² · Tae-Weon Kang² · Jeong-Hwan Kim²

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 the antenna factor for a dipole antenna with a hybrid balun is derived 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 power mismatch-loss component factors.

Key words: Calculable Dipole Antenna, Antenna Factor, Method of Moments, Hybrid Balun.

1. Introduction

Standard dipole antennas are necessary for measuring antenna factors and for generating a standard field. Calculable dipole antennas have been developed and used as 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 usually be achieved using method of moments (MoM). Extensive works have been carried out in the development of a calculable standard dipole antenna [3], [4].

Previous work has dealt with the analysis of a calculable dipole antenna factor by using S-parameters [1]~[4]. This paper presents the derivation of a new formula for an antenna factor composed of the power loss component factors for a calculable dipole antenna with a 3 dB 180-degree hybrid balun. The design of a calculable dipole antenna is presented that has a frequency range of 1 GHz to 3 GHz. The formulation introduced in this paper is derived by using the power mismatch-loss concept [5], [6]. The antenna factors obtained by employing this concept are in good agreement with the factors derived by S-parameters [1]. The results of the theoretical analysis presented in this paper, allow the design of a calculable dipole antenna using the power loss components. Calculated input impedances and antenna factors of the dipole antenna are compared with measured results.

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

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II. 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 from 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 (Σ) of the hybrid, and a matched measuring instrument is connected to the other port (Δ) 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 π radians.

III. 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 and $Z_a = R_a + jX_a$ is the antenna input impedance, then the power available from the receiving antenna can be expressed as

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

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 that only passive devices exist between the receiving antenna and the measuring receiver, and that the cable loss is zero, the total power loss can be represented as follows:

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

where

$$M_A = \frac{P_a}{P_{b1}} \quad (4)$$

$$M_B = \frac{P_{b1}}{P_b} \quad (5)$$

$$M_C = \frac{P_b}{P_{c1}} \quad (6)$$

$$M_D = \frac{P_{c1}}{P_c} \quad (7)$$

$$M_E = \frac{P_c}{P_L} \quad (8)$$

In (4)~(8), P_a is the power available from the antenna, P_{b1} is the power delivered to the hybrid balun (consisting of coaxial lines and a 180-degree hybrid), 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 obtained as follows:

$$AF = \sqrt{\frac{4R_a}{|h_e|^2 R_L}} \sqrt{M_A M_B M_C M_D M_E} = AF_0 K \quad (9)$$

where AF_0 is the antenna factor when the receiver is directly connected to the antenna and can be expressed as

$$AF_0 = \sqrt{4R_a / |h_e|^2 R_L} \quad (10)$$

K is the power mismatch-loss factor represented as

$$K = \sqrt{M_A M_B M_C M_D M_E} \quad (11)$$

and individual power mismatch-loss factors are as follows:

$$M_A = \frac{|Z_{b1} + Z_a|^2}{4R_a R_{b1}} \quad (12)$$

$$M_B = \frac{4R_T R_{b1}}{|Z_{b1} + Z_a|^2} \left| \frac{Z_{11} + Z_a}{Z_{12}} \right|^2 \quad (13)$$

$$M_C = \frac{|Z_T + Z_d|^2}{4R_T R_d} \quad (14)$$

$$M_D = \frac{4R_t R_d}{|Z_T + Z_d|^2} \left| \left(\frac{Z_0 + Z_T}{Z_0} \right) \frac{1 - \Gamma_S e^{-2\gamma L_R}}{2e^{-\gamma L_R}} \right|^2 \quad (15)$$

$$M_E = \frac{|Z_t + R_L|^2}{4R_t R_L} \quad (16)$$

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 co-

axial cable, and $\gamma = \alpha + j\beta$, where α and β are the attenuation coefficient and phase constant of the coaxial cable, respectively. These are expressed as follows:

$$Z_T = Z_{22} - \frac{Z_{12}Z_{21}}{Z_{11} + Z_a} \equiv R_T + jX_T \quad (17)$$

$$Z_t = Z_0 \frac{1 + \Gamma_S e^{-2\gamma L_R}}{1 - \Gamma_S e^{-2\gamma L_R}} \quad (18)$$

$$Z_{b1} = Z_{11} - \frac{Z_{12}Z_{21}}{Z_{22} + Z_d} \equiv R_{b1} + jX_{b1} \quad (19)$$

$$Z_d = Z_0 \frac{1 + \Gamma_L e^{-2\gamma L_R}}{1 - \Gamma_L e^{-2\gamma L_R}} \quad (20)$$

$$\Gamma_L = \frac{R_L - Z_0}{R_L + Z_0} \quad (21)$$

$$\Gamma_S = \frac{R_T - Z_0}{R_T + Z_0} \quad (22)$$

where Z_{11} , Z_{12} , and Z_{22} are equivalent T network impedances associated with the hybrid balun.

In this paper, we also derive the antenna factor using the scattering matrix (S -parameters), which was used earlier by NPL (National Physical Laboratory) [1]. From the Fig. 1(b), the antenna factor is expressed, using S -parameters, as

$$AF = \frac{(Z_a + Z_0) [(1 - T_{11}\Gamma_A)(1 - T_{22}\Gamma_L) - T_{12}T_{21}\Gamma_A\Gamma_L]}{Z_0 h_e (1 + \Gamma_R)T_{21}} \quad (23)$$

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 represented as follows:

$$\begin{bmatrix} T_{11} & T_{12} \\ T_{21} & T_{22} \end{bmatrix} = \begin{bmatrix} N_{11} & N_{12} \\ N_{21} & N_{22} \end{bmatrix} \begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix} \quad (24)$$

where N_{ij} and L_{ij} are the S -parameters of the hybrid balun and the coaxial cable, respectively, as shown in Fig. 1. We normalize N_{11} and N_{21} to 100 Ω , and N_{12} , and N_{22} to 50 Ω . In addition, Z_0 is the characteristic impedance of the coaxial cable, and Γ_A and Γ_L are the reflection coefficients of the antenna and the measuring receiver, respectively. The expression of the S -parameters of the hybrid balun for an ideal case are as follows:

$$\begin{bmatrix} N_{11} & N_{12} \\ N_{21} & N_{22} \end{bmatrix} = \begin{bmatrix} 0 & 2/\sqrt{2} \\ 1/\sqrt{2} & 0 \end{bmatrix} \quad (25)$$

The S -parameters of the coaxial cable for an ideal case are as follows:

$$\begin{bmatrix} L_{11} & L_{12} \\ L_{21} & L_{22} \end{bmatrix} = \begin{bmatrix} 0 & e^{-j\beta L_R} \\ e^{-j\beta L_R} & 0 \end{bmatrix} \quad (26)$$

Since $Z_0/(Z_a + Z_0) = (1 - \Gamma_a)/2$, the antenna factor of (23) 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 (23) derived from S -parameters.

IV. The Design of Standard Dipole Antennas

In the design of the standard dipole antenna, a thin wire kernel approximation with a segment length of 0.0125λ is used for the piecewise sinusoidal Galerkin's MoM analysis. The dipole radius was chosen to be less than 0.007λ (thin wire approximation) and the nominal value of 50 Ω is used for the characteristic impedance Z_0 . The coaxial cable with a length of 10 m is selected in the numerical calculations. In this paper, an induced electromotive force (EMF) method is also applied for comparison of the antenna input impedance to be used to calculation of the antenna factor.

The validity of the theoretical analysis for the dipole antenna design was checked by comparing, the antenna input impedances with those derived experimentally [7], as shown in Table 1. Table 1, also shows the input impedances calculated by the EMF method. The MoM results are in better agreement with the experiments than are the EMF results. Details of the comparison of antenna factor characteristics for the calculable dipole antenna by the MoM and EMF methods are explained in [8]. The MoM is used in all subsequent 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 as the frequency increases, but it increases gradually above 2.5 GHz when $a = 0.75$ mm. This is due to the fact that the dipole radius was chosen to be greater than 0.007λ (thin wire approximation).

Fig. 3 shows the frequency characteristics of the antenna input resistance at resonance length in terms of 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 small variation of less than 1.9 %. The

Table 1. Calculated and measured antenna input impedances.

Radius a/λ	Calculated Z_{in} (Ω)		Measured [7] Z_{in} (Ω)
	MoM	EMF	
2.98×10^{-3}	93.44 + j 4.86	73.13 + j 1.42	91.21 + j 43.26
3.97×10^{-3}	97.18 + j 43.58	73.12 + j 41.05	96.42 + j 43.54

Fig. 2. The resonance length of the dipole antenna.

Fig. 3. The input impedance at resonance length of the dipole antenna.

maximum variation of the input resistance is 5.7 % at 3 GHz.

Fig. 4 shows the frequency characteristics of the an-

Fig. 4. Frequency characteristics of the antenna factor.

tenna factor. The frequency characteristics of the theoretical antenna factor with respect to the value of the dipole radius are almost the same 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 within 0.11 dB.

Fig. 5 shows the frequency characteristics of the power mismatch-loss factors. Details of the power mismatch-loss factors are shown in Table 2 in the frequency range from 1 GHz to 3 GHz. As can be seen in Fig. 5 and 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 $\sqrt{M_E}$ which is the mismatch-loss factor of the coaxial cable and the measuring receiver.

Fig. 6 shows the measured and calculated antenna

Table 2. Calculated antenna factors by the MoM.

f (GHz)	a (mm)	L_0 (cm)	Z_{in} (Ω)	Mismatch loss factor K (dB)					Antenna factor (dB)	
				$\sqrt{M_A}$	$\sqrt{M_B}$	$\sqrt{M_C}$	$\sqrt{M_D}$	$\sqrt{M_E}$	AF_0	AF
1.0	0.75	14.10	72.53	0.111	0.111	0.111	0.111	0.111	28.09	28.20
1.2	0.75	11.72	72.77	0.109	0.109	0.109	0.109	0.109	29.67	29.78
1.4	0.75	10.02	73.06	0.107	0.107	0.107	0.107	0.107	31.01	31.12
1.6	0.75	8.751	73.40	0.103	0.103	0.103	0.103	0.103	32.17	32.27
1.8	0.75	7.765	73.81	0.100	0.100	0.100	0.100	0.100	33.19	33.29
2.0	0.75	6.979	74.30	0.095	0.095	0.095	0.095	0.095	34.10	34.20
2.2	0.75	6.338	74.88	0.091	0.091	0.091	0.091	0.091	34.93	35.02
2.4	0.75	5.806	75.59	0.085	0.085	0.085	0.085	0.085	35.68	35.77
2.6	0.75	5.359	76.44	0.078	0.078	0.078	0.078	0.078	36.38	36.46
2.8	0.75	4.978	77.49	0.070	0.070	0.070	0.070	0.070	37.02	37.09
3.0	0.75	4.651	78.79	0.062	0.062	0.062	0.062	0.062	37.62	37.68

Fig. 5. Frequency characteristics of the power mismatch-loss factors.

Fig. 6. The measured and calculated antenna factors of the calculable dipole antenna with 3 dB hybrid balun.

factors of the calculable dipole antenna with 3 dB hybrid balun (Model 5310A-104, Picosecond Pulse Labs, Inc.). The calculated antenna factors are in good agreement with the measured results.

Fig. 7 shows the measured amplitude flatness and phase flatness of the constructed antenna balun with 3 dB hybrid. The amplitude and phase flatness are defined as $(N_{12}/N_{13}) = (N_{21}/N_{31}) = Ae^{j\phi}$, where A and ϕ are the amplitude and phase of the ratio of the 3-port S-parameters, respectively. The measured results of the balun characteristics satisfy the requirements mentioned in CISPR regulation as $0.95 < A < 1.05$ and $178^\circ < 180\phi/\pi < 182^\circ$.

V. Conclusions

A new formula for the antenna factor is derived using the power mismatch-loss concept. A design is presented for a standard dipole antenna with a hybrid balun in the frequency range of 1 GHz to 3 GHz. The numerical re-

Fig. 7. The measured amplitude flatness and phase flatness of the constructed antenna balun.

sults show that the antenna factor of the calculable dipole antenna, whose formulation includes the power loss component factors, can be characterized by the power mismatch-loss concept and the MoM analysis.

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