Short-Ended Electromagnetically Coupled Coaxial Dipole Array Antenna

Joong-Pyo Kim[1] · Chang-Won Lee[2]

Abstract

A short-ended electromagnetically coupled coaxial dipole array antenna is investigated theoretically. The antenna has an advantage of structural simplicity. The integral equations are derived for the proposed structure by use of the Fourier transform and mode expansion, and the simultaneous linear equations are obtained. The slot electric field and strip current are obtained by solving the simultaneous linear equations. The effects of slot and strip numbers on the radiation efficiency, beamwidth and directivity gain of the antenna are presented.

Key words : Collinear Antennas, Coaxial Dipole Array, Electromagnetically Coupled, Short-Ended, Strip

Ⅰ. Introduction

Collinear antenna has been considered as an array element to replace two-dimensional dipole arrays because of the simple feeding structure. There exist some kinds of collinear antenna, i.e., the coaxial collinear (COCO) antenna and the coaxial dipole array (CDA). COCO antenna employs a collinear arrangement of coaxial cables where the feeding structures are inverted in a half-wavelength step to produce in-phase excitations[1]~[3]. The radiating dipole of CDA[4] is fed by an annular ring slot which extends radically from the outer conductor of the feeding coaxial cable.

Recently, the coaxial-collinear antenna and slotted coaxial antenna were studied through transmission line method to check the possibility of them as the base-station antenna[5]. Electromagnetically coupled coaxial dipole (ECCD) array antenna which is composed of a half-wavelength metallic circular pipe fed electromagnetically by an annular ring slot on the outer conductor of the feeding coaxial cable was analyzed by Wiener-Hopf technique and equivalent transmission line analysis[6]. ECCD array antenna is another modification of CDA, which has an advantage of structural simplicity due to a novel use of electromagnetically coupled feed for CDA. More recently, radiation characteristics of finite strip-grating loaded dielectric-coated coaxial waveguide with finite periodic thick slots were investigated in a viewpoint of leaky wave[7]. Strip-grating of this structure is electromagnetically coupled with coaxial slot which is exactly same with the feeding mechanism of the ECCD array antenna.

In this paper, a short-ended electromagnetically coupled coaxial dipole array is investigated theoretically. It is derived by changing the matched termination of coaxial line of the previous configuration[7] with short-ended termination. By controlling the length of the short-ended termination from the center of the last slot, which totally reflects the transmitted waves, the impedance matching can be made without a quarter wavelength impedance transformer[6], and high radiation efficiency is also obtained. This characteristic gives more simplicity of design.

The analysis method and procedure are the same as [7]. The integral equations are derived for the proposed structure by use of the Fourier transform and mode expansion, and the simultaneous linear equations are obtained. The slot electric field and strip current are obtained by solving the simultaneous linear equations. The effects of slot and strip number on the radiation efficiency, beamwidth and directivity gain are presented.

Ⅱ. Formulation

A short-ended electromagnetically coupled coaxial dipole array structure is circular symmetry, therefore the cross-section geometry in the longitudinal direction is shown in Fig. 1, where O_s is the distance from the center of last slot to the short-ended termination.

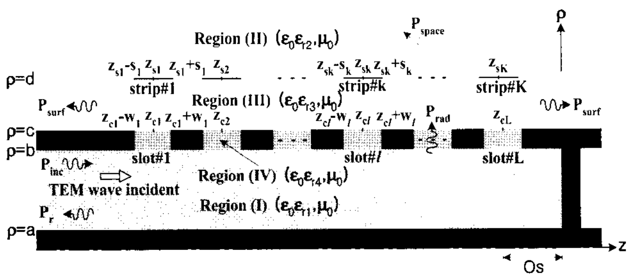

Fig. 1. Geometry of the proposed antenna.

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It is assumed that TEM mode is incident in the coaxial waveguide and the waveguide height $(b-a)$ is small enough that only TEM mode is the propagating mode in the coaxial waveguide. The incident field is

$$H_{\phi}^{i(I)}(\rho,z)=\frac{Ik_1}{\omega\mu_0\ln(b/a)\rho}e^{-jk_1z}, \tag{1}$$

The scattered field in region (I) which satisfies the boundary condition at $\rho=a$ and $z=0$ is expressed as

$$H_{\phi}^{s(I)}(\rho,z)=\frac{2}{\pi}\int_0^{\infty}\tilde{H}_{\phi}^{s(I)}(\zeta) \left[H_1^{(2)}(k_{\rho1}\rho)-\frac{H_0^{(2)}(k_{\rho1}a)}{H_0^{(1)}(k_{\rho1}a)}H_1^{(1)}(k_{\rho1}\rho)\right]\cos\zeta z d\zeta \tag{2}$$

where $k_{\rho1}=\sqrt{k_1^2-\zeta^2}$ and $H_n^{(1)}(\cdot)$ is the n order Hankel function of the first kind.

The field representations in region (II), (III) and, (V) are the same as those of [7], and also the boundary conditions imposed at every interface are identical with those of [7].

After imposing the boundary conditions, we can derive the following equations

$$\begin{aligned}&\sum_{l=1}^{L}\sum_{m=0}^{\infty}\frac{\varepsilon_{r2}k_{\rho4m}}{2\pi\varepsilon_{r4}}N_m^l(b)[I_{nm}^{rl}+E_{nm}^{rl}]\\&-\left[B_n^r\frac{H_1^{(1)}(k_{\rho4n}^r b)}{H_1^{(1)}(k_{\rho4n}^r c)}+C_n^r\right]\alpha_n w_r\\&=\frac{-Ik_1}{k_0\eta_0\ln(b/a)b}[Q_n^r(k_1)+Q_n^r(-k_1)]\end{aligned} \tag{3}$$

$$\sum_{k=1}^{K}\sum_{m=1}^{\infty}A_m^k a_m^k\frac{-1}{2\pi}H_{nm}^{jk}+\sum_{l=1}^{L}\sum_{m=0}^{\infty}\frac{k_{\rho4m}^l}{2\pi\varepsilon_{r4}}N_m^l(c)K_{nm}^{jl}=0, \tag{4}$$

$$\begin{aligned}&\sum_{k=1}^{K}\sum_{m=1}^{\infty}A_m^k a_m^k\frac{-1}{2\pi}K_{nm}^{rk}+\sum_{l=1}^{L}\sum_{m=0}^{\infty}\frac{\varepsilon_{r3}k_{\rho4m}^l}{2\pi\varepsilon_{r4}}N_m^l(c)J_{nm}^{rl}\\&-\left[B_n^r+C_n^r\frac{H_1^{(2)}(k_{\rho4n}^r c)}{H_1^{(2)}(k_{\rho4n}^r b)}\right]\alpha_n w_r=0.\end{aligned} \tag{5}$$

Here E_{nm}^{rl} and $Q_n^r(-k_1)$ in Eq. (3) are terms generated by the short-ended termination, and the remaining terms are given in [7] and [8], and E_{nm}^{rl} are represented as

$$E_{nm}^{rl}=\int_{-\infty}^{\infty}\frac{H_1^{(2)}(k_{\rho1}b)H_0^{(1)}(k_{\rho1}a)-H_0^{(2)}(k_{\rho1}a)H_1^{(1)}(k_{\rho1}b)}{k_{\rho1}[H_0^{(2)}(k_{\rho1}b)H_0^{(1)}(k_{\rho1}a)-H_0^{(2)}(k_{\rho1}a)H_0^{(1)}(k_{\rho1}b)]}\cdot Q_m^l(\zeta)Q_n^r(\zeta)d\zeta.$$

Using (3)~(5), we can obtain the following simultaneous equations

$$\begin{bmatrix}\Psi_1&\Psi_2&\Psi_3\\\Psi_4&\Psi_5&\Psi_6\\\Psi_7&\Psi_8&\Psi_9\end{bmatrix}\begin{bmatrix}\mathbf{A}\\\mathbf{B}\\\mathbf{C}\end{bmatrix}=\begin{bmatrix}\mathbf{S}\\\mathbf{0}\\\mathbf{0}\end{bmatrix} \tag{6}$$

Once the unknown mode coefficients A_m^k, B_m^l and C_m^l are determined by solving (6), $\tilde{H}_{\phi}^{s(I)}$, $\tilde{H}_{\phi}^{(II)}$, $\tilde{H}_{\phi}^{(III)}$ and $\tilde{H}_{\phi}^{(III)}$ are computed. Hence, the fields in each region can be calculated. The waves radiated from the slots are divided into two wave types, i.e., one is the space wave which is radiated into the region (II), and the other the surface wave which is trapped in and above the dielectric and guided in the $\pm z$-directions. The far-zone magnetic field radiated into the region (II) as a space wave is calculated by using an asymptotic evaluation while the surface wave in each region calculated by applying the Cauchy residue theorem.

The time-averaged incident power (P_{inc}) is divided into the reflected power (P_r) and the transmitted power from slots (P_{rad}) to regions (II) and (III). Here the transmitted power from slots P_{rad} is also divided into the space wave power (P_{space}) and the surface wave power (P_{surf}).

Ⅲ. Numerical Results

The validity of the numerical results in this paper is assured by a check of the power conservation law

$$P_{inc}=P_r+P_{rad},$$
$$P_{rad}=P_{space}+P_{surf}.$$

We put the slot and strip period as T_c and T_s, respectively, for all examples below. First of all, we investigated the effect of short-ended distance O_s for a antenna structure having $a=0.05\lambda$, $b=a+0.1\lambda$, $c=b+0.01\lambda$, $d=c+0.1\lambda$, $T_c=1.0\lambda_1$, $w=0.032T_c$, $T_s=1.0\lambda_1$, $s=0.27T_s$, $\varepsilon_{r1}=2.5$, $\varepsilon_{r2}=1.0$, $\varepsilon_{r3}=\varepsilon_{r4}=2.0$, where λ_1 and λ are the wavelength of Region (I) and free space, respectively, when the number of slot and strip is one ($N_{slot}=N_{strip}=1$). The S_{11} and normalized input admittance $\underline{Y}_{in}$ versus the short-ended distance O_s are shown in Fig. 2. It is seen that the $|S_{11}|$ and $\underline{Y}_{in}$ is periodic, repeating for multiples of $\lambda_1/2$, which is the general property for a short circuit load. The second and third minimum points of $|S_{11}|$ are slightly smaller than the first, thus the minimum short-ended distance is selected the second minimum point of $|S_{11}|$, around $0.53\lambda_1$. At the minimum short-ended distance $O_s=0.53\lambda_1$, $\underline{Y}_{in}(=\underline{G}_{in}-j\underline{B}_{in})$ is (0.986+j0.025) where the real part $\underline{G}_{in}$ is very close to 1 while the imaginary part $\underline{B}_{in}$ is comparatively close to 0, which means the input admittance is well matched and the input power is almost coupled into the radiation power as shown in Fig. 2 and 3. On the other hand, for the open-ended case, $\underline{Y}_{in}$ is (0.495+j0.051), which indicates the admittance matching is worse than the short-ended case, thus the normalized radiation power is 0.44, nearing half of short-ended case.

In Fig. 4, the variation of $|S_{11}|$ and $\underline{Y}_{in}$ versus the operating frequency f for an antenna matched at f_0 with the minimum

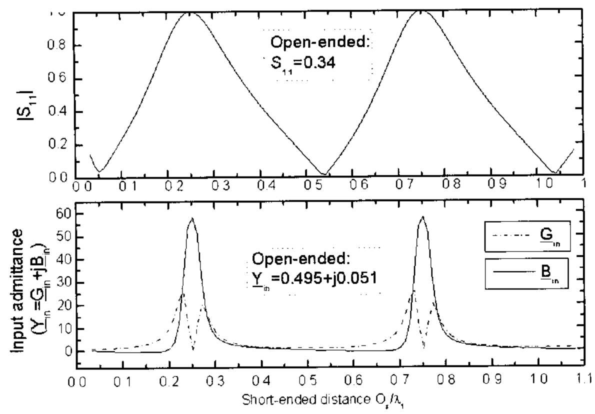

Fig. 2. $|S_{11}|$ and $\underline{Y}_{in}$ vs. short-ended distance O_s.

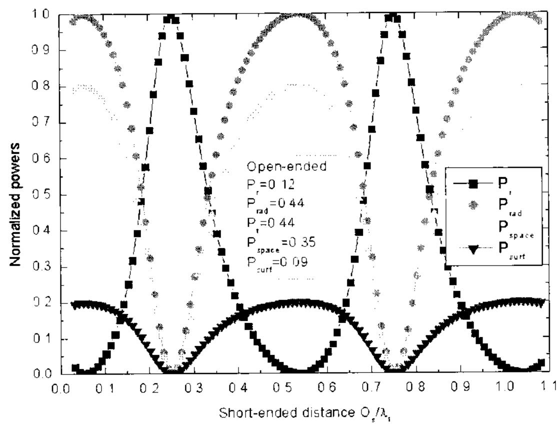

Fig. 3. Normalized powers vs. short-ended distance O_s.

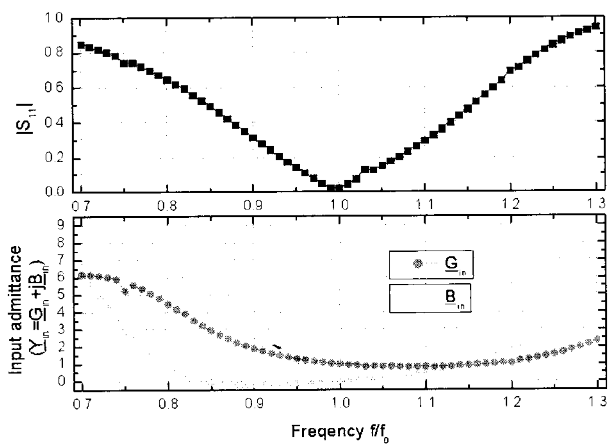

Fig. 4. $|S_{11}|$ and $\underline{Y}_{in}$ vs. frequency f/f_0.

short-ended distance $O_s = 0.53\lambda_1$ are shown. The range of the return loss(=20log$_{10}|S_{11}|$) less than −10 dB is approximately from 0.9 f_0 to 1.1 f_0, the variation of $|S_{11}|$ is on the whole similar

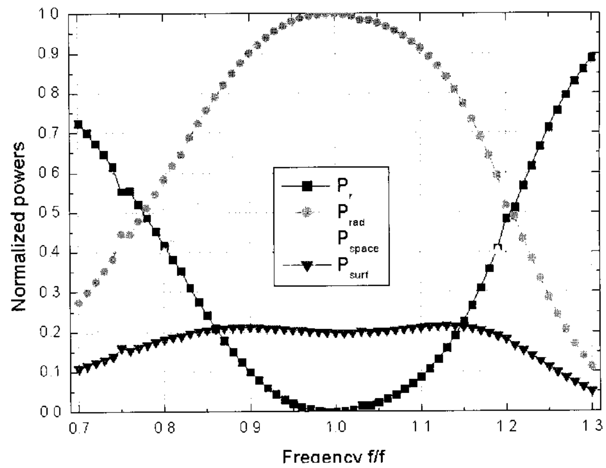

Fig. 5. Normalized powers vs. frequency f/f_0.

to that of Fig. 6 in [6]. It is also shown in Fig. 5 that more than 90 % of the incident power is radiated over the range of 0.9 f_0 through 1.1 f_0.

From Fig. 3 and Fig. 5, respectively, it is ascertained that the power conservation law is well satisfied. At $O_s = 0.53\lambda_1$, the radiated power (P_{rad}) is 99.94 %, the reflected power (P_r) 0.06 %, and the space wave power (P_{space}) and surface wave power (P_{surf}) are 80.1 % and 19.9 % of the radiated power, respectively.

Fig. 6 (a) and (b) show that the amplitude and phase of slot magnetic current and strip electric current, respectively, and the electric field and strip electric current satisfy the edge conditions.

Fig. 6 (c) shows the radiation patterns for each of short-ended and open-ended case. The broadside radiation angle and beamwidth of the short-ended case are 90° and 80.76°, respectively, which is same as those of the open-ended case, but the directivity gain of the short-ended case is 1.04 dBi while that of the open-ended case −2.51 dBi. It is found that at the maximum radiation angle(90°), the radiation power level of the short-ended case is 3.5 dB larger than that of the short-ended case.

The amplitude and phase of slot electric fields and strip electric currents are shown in Fig. 7 (a) and (b), respectively, when the number of slot and strip is five ($N_{slot} = N_{strip} = 5$) and $a = 0.05\lambda$, $b = a + 0.1\lambda$, $c = b + 0.01\lambda$, $d = c + 0.1\lambda$, $T_c = 1.0\lambda_1$, $w = 0.25T_c$, $T_s = 1.0\lambda_1$, $s = 0.19T_s$, $\varepsilon_{r1} = 2.5$, $\varepsilon_{r2} = 1.0$, $\varepsilon_{r3} = \varepsilon_{r4} = 2.0$, $O_s = 0.5T_c$. Fig. 7 (c) shows the radiation pattern of the structure in which the beamwidth and the directivity gain are 17.22° and 7.67 dBi, respectively. For this antenna, the radiated power(P_{rad}) is 99.9 %, the space wave power(P_{space}) and the surface wave power (P_{surf}) are 94.4 % and 5.6 % of the radiated power, respectively. It is observed that the space wave portion of the radiated power becomes dominant as the number of slot and

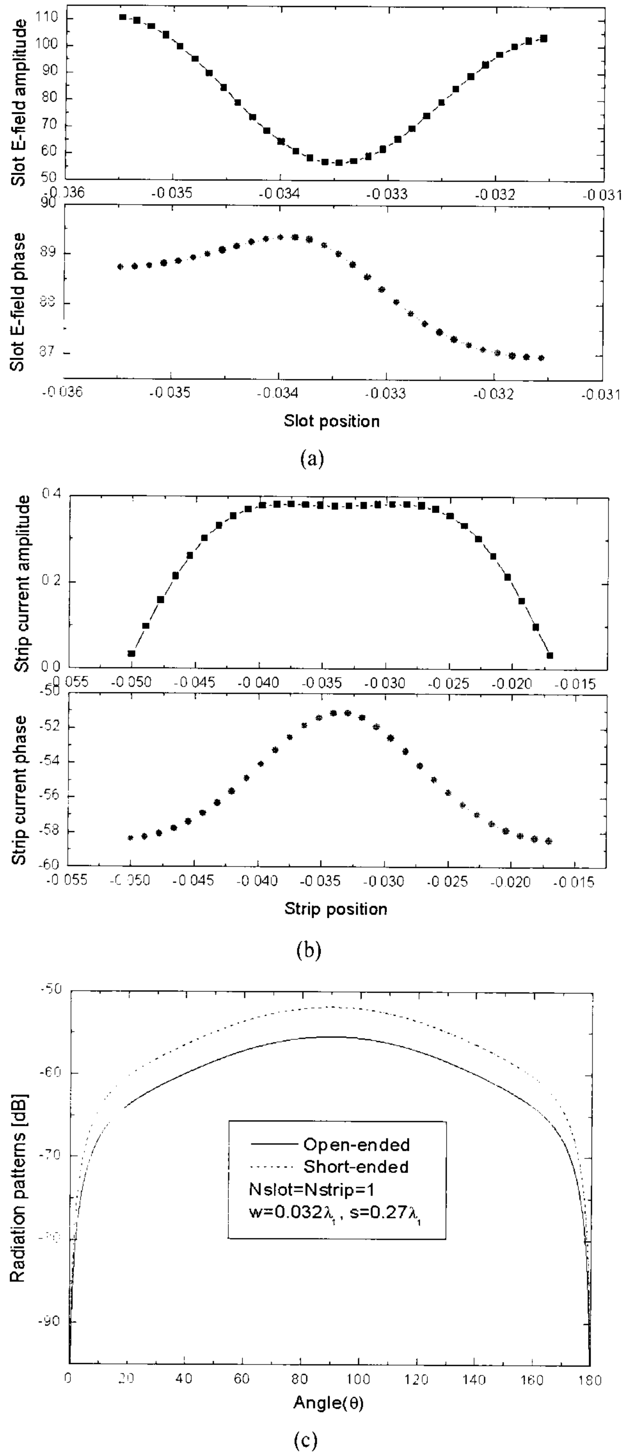

Fig. 6. (a) Slot E-field amplitude and phase, (b) Strip electric current amplitude and phase, (c) Radiation pattern(single element).

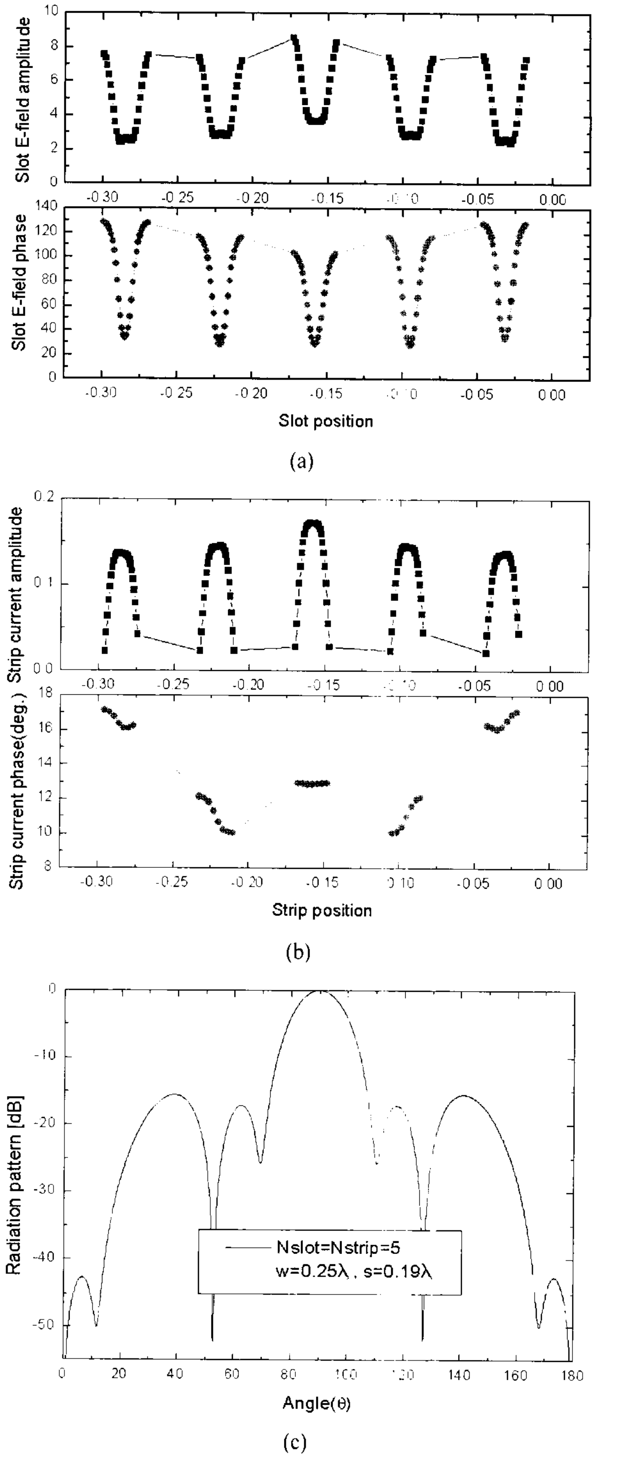

Fig. 7. (a) Slot E-field amplitude and phase, (b) Strip electric current amplitude and phase, (c) Radiation Pattern(5 elements).

strip increases. Fig. 7. (a) and (b) shows that the edge conditions at the slot and strip edge are well satisfied: the electric fields are goes to infinity as approaches to the slot edges and the strip electric current goes to zero at the strip edges. As

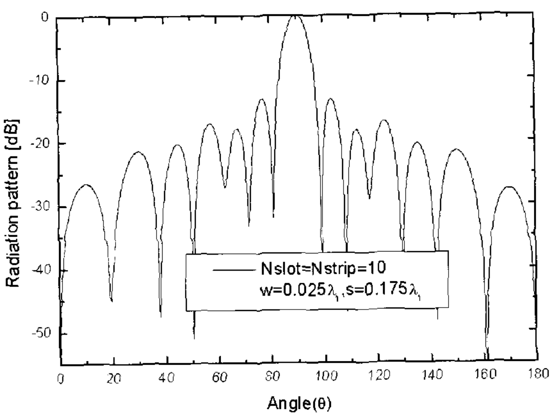

Fig. 8. Radiation pattern(10 elements).

Table 1. Comparison of our and approximate results.

Number of slot	Beamwidth (Degree)	Ours (dBi)	Approximate (dBi)[4]
5	17.22°	7.67 dBi	7.89 dBi
10	8.04°	10.84 dBi	11.08 dBi

shown in Fig. 7. (a) and (b), the slot electric field and strip electric current distributions are symmetric, as is often case with collinear arrays.

The radiation pattern is shown in Fig. 8 for a structure having $a=0.05\lambda$, $b=a+0.1\lambda$, $c=b+0.01\lambda$, $d=c+0.1\lambda$, $T_c=1.0\lambda_1$, $w=0.025T_c$, $T_s=1.0\lambda_1$, $s=0.175T_s$, $\varepsilon_{r1}=2.5$, $\varepsilon_{r2}=1.0$, $\varepsilon_{r3}=\varepsilon_{r4}=2.0$, $O_s=0.68T_c$, $N_{slot}=N_{strip}=10$. For this structure, the radiated power (P_{rad}) is 99.7 %, the space wave power (P_{space}) and the surface wave power (P_{surf}) are 96.4 % and 3.6 % of the radiated power, respectively. As shown in this figure, the beamwidth and the directivity gain of the antenna are 8.04° and 10.84 dBi, respectively.

The validity of the proposed method is checked by comparing antenna gains calculated from the proposed method with those obtained from approximate relationship between directivity and beamwidth for broadside collinear arrays[4],[9], and shown in Table 1. As seen in the table 1, our results are almost same with the approximate ones.

Ⅳ. Conclusion

The rigorous analysis method for short-ended electromagnetically coupled coaxial dipole array is proposed, and the validity of the proposed method is assured by checking the power conservation law and comparing antenna gain with the one obtained by approximate method. The impedance matching is well accomplished for the variations of slot number by controlling the distance between the center of the last slot and short-ended termination. As the number of strip increases, the radiation efficiency and gain are increased, and beamwidth is reduced. It is observed that the antenna loss due to the unexpected surface wave decreases as the number of antenna element increases.

It is thought that the proposed antenna can be applied to the base-station antenna, the surveillance radar for air traffic control or the omnidirectional antenna for the telemetry, command and ranging (TC&R) subsystem of communication satellite.

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