EFFECTS OF A PARASITIC WIRE ON COUPLING BETWEEN TWO SLOT ANTENNAS

Takehiro MORIOKA* and Kazuhiro HIRASAWA
*Electrotechnical Laboratory
1-1-4, Umezono, Tsukuba, Ibaraki, 305-8568, Japan
University of Tsukuba
1-1-1, Tennohdai, Tsukuba, Ibaraki, 305-8577, Japan
E-mail tkmoriok@etl.go.jp

1. Introduction

With the development of wireless communications many antennas are needed for different purposes and it is often required to have these antennas close together in a limited area. Radiated power from one antenna is received by the other antennas around it and often makes system degradation problems. Studies on improvement of antenna characteristics by adding parasitic elements have been reported [1][2] where radiation patterns of the antenna are mostly focused on. Little researches on coupling between feeding ports of slot antennas have been done. We reported coupling characteristics between two monopole antennas that are designed to operate at different frequencies and proposed the method to reduce coupling at both frequencies by using a center-loaded slot [3].

In this paper coupling between two half-wavelength slot antennas on an infinite conducting plane is analyzed and a parasitic wire is introduced between the slots to reduce coupling. A coupling coefficient [4] is used to evaluate coupling between the slots. Then radiation characteristics of the antennas on a finite rectangular conducting plane are calculated by using the hybrid method of the moment method (MM)[5] and the geometrical theory of diffraction (GTD)[6] to include the diffracted electromagnetic fields.

2. Analytical Method

An infinite ground plane and two slots divide a space into two regions. The region above the ground plane is named region 1 and the other side of the ground plane is named region 2. Two slot antennas and a parasitic wire are located in region 1. It is assumed that only one slot is fed and the other slot is a scatterer with a load that equals the characteristic impedance of the feed line.

$\overline{E}^{reg1}_{slot1}$, $\overline{H}^{reg1}_{slot1}$ are the fields radiated by the excited slot (slot 1) and $\overline{E}^{reg1}_{slot2}$, $\overline{H}^{reg1}_{slot2}$, $\overline{E}_{wire}$ and $\overline{H}_{wire}$ are the scattered fields by the loaded slot (slot 2) and the parasitic wire in region 1. In region 2 $\overline{E}^{reg2}_{slot1}$, $\overline{H}^{reg2}_{slot1}$ are the electromagnetic fields radiated from the excited slot and $\overline{E}^{reg2}_{slot2}$, $\overline{H}^{reg2}_{slot2}$ are the scattered fields by the loaded slot as the slots radiate electromagnetic fields into both sides of the ground plane. Then by using the delta function d the necessary boundary conditions on the surface of the slots are written, respectively, as follows.

$$\left(\overline{E}^{reg1}_{slot1}+\overline{E}^{reg1}_{slot2}+\overline{E}_{wire}\right)\times\overline{n}=0 \quad \text{on the wire} \quad (1)$$

$$n\times\left(\overline{H}^{reg1}_{slot1}+\overline{H}^{reg1}_{slot2}+\overline{H}_{wire}-\overline{H}^{reg2}_{slot1}-\overline{H}^{reg2}_{slot2}\right)=I_{f,L}\,d\left(y-l_{f,L}\right) \quad \text{on the slots} \quad (2)$$

$$\left(\overline{E}^{reg1}_{slot1}+\overline{E}^{reg1}_{slot2}+\overline{E}_{wire}\right)\times\overline{n}=\left(\overline{E}^{reg2}_{slot1}+\overline{E}^{reg2}_{slot2}\right)\times\overline{n} \quad \text{on the slots} \quad (3)$$

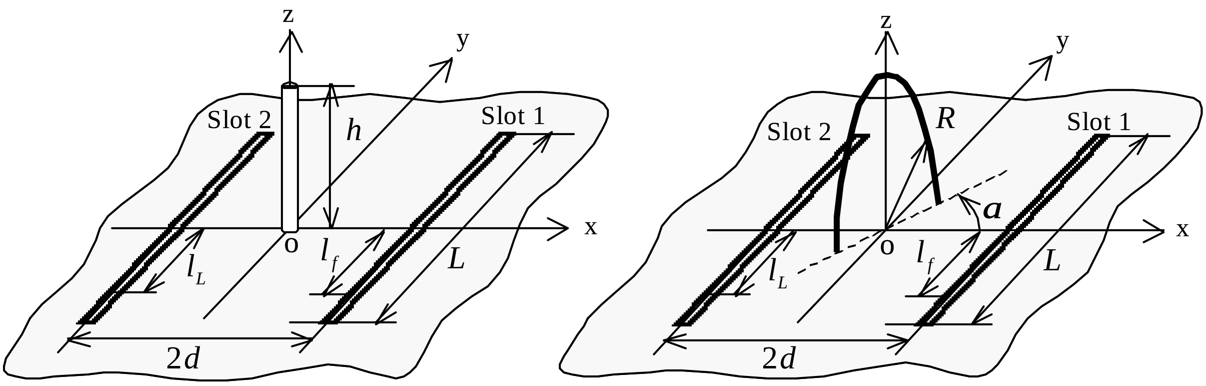
where $I_{f,L}$ is the feed current or the current at the load. In equations (2) and (3) $\overline{n}$ is the unit normal vector on the ground plane pointing from region 2 to 1. The equivalent theorem is applied and the boundary condition (2) on the slot can be rewritten as follows.

$$n\times\left(2\overline{H}_{slot1}^{reg1}+2\overline{H}_{slot2}^{reg1}+\overline{H}_{wire}\right)=I_{f,L}\,\delta\left(y-l_{f,L}\right) \tag{4}$$

where proper magnetic currents are used to satisfy (3) automatically. Thus (1) and (4) are solved by the MM. To take the diffracted electromagnetic fields $\overline{E}_{diff}$、$\overline{H}_{diff}$ into account they can be added to the total fields in (1) and (4).

3. Numerical Results

The calculation models are shown in fig. 1. Two offset-fed slot antennas are made on an infinite ground plane and a parasitic wire is located between the slots. Slots are parallel to the y-axis and their locations are $x=\pm d$. The length and width of the slot and the location of the feed point on the slot are shown as L, w and l_f, respectively. As shown in fig. 1(a) a parasitic monopole with the height h is located vertically at the origin. Figure 1(b) shows the antenna arrangement with a half-loop. The radius of the parasitic half-loop is R. The center of the half-loop is at the origin and the angle measured from the x-axis to the half-loop is assumed as α. In this paper L and w are fixed to 100mm and 2.0mm, respectively. The feed point on the slot is located at l_f=39.0mm where the minimum VSWR is obtained at 1.5GHz. The loaded position on the other slot is located to be $l_L=l_f$.



(a) two slots and a monopole. (b) two slots and a half-loop

fig. 1. Antenna arrangement.

3.1 Coupling Coefficients

A coupling coefficient is introduced to evaluate the coupling between two antennas. It is defined as the ratio of the received power at the load P_r to the input power of the transmitting antenna P_{in} as $R_{re}=10\log\left(P_r/P_{in}\right)$. The coupling coefficient without the parasitic element is –13.7dB at 1.5GHz.

3.1.1 Slots with a monopole

Figure 2 shows the coupling coefficients with respect to the monopole height at 1.5GHz. The minimum coupling coefficient is obtained when h is 49mm. Reduction of 14dB is obtained compared

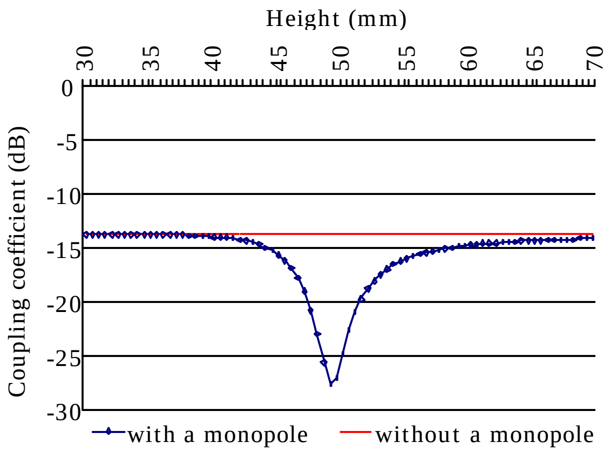


fig. 2 Coupling coefficients with respect to the monopole height.

with the coupling coefficient without the monopole. The height of the monopole becomes about a quarter wavelength at 1.5GHz and electromagnetic fields reradiated from the monopole cancel the fields on the loaded slot. Frequency characteristics of the coupling coefficient with this monopole are shown in fig. 3. A bandwidth of 60MHz is obtained where the coupling coefficient is reduced more than 10dB compared with the one without the monopole.

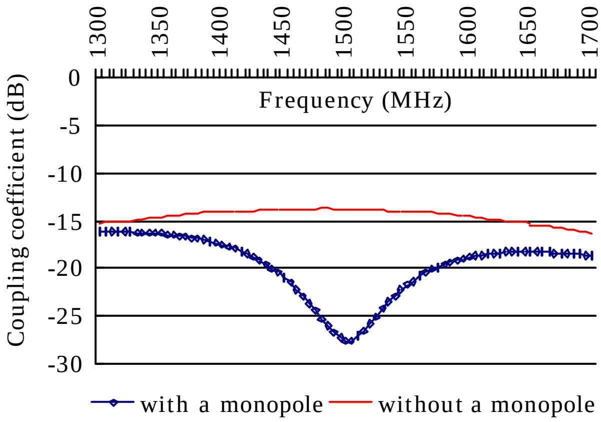


fig. 3 Coupling coefficients with a monopole.

3.1.2 <u>Slots with a half-loop</u>

Two antenna arrangements with R=34.5mm, α=90° and R=37.0mm, α=0° are focused on and named as type A and B, respectively. Frequency characteristics for types A and B are shown in fig. 4. For type A a bandwidth of 70MHz is obtained where the reduction of more than 10dB compared with the coupling coefficient without a half-loop and the relative bandwidth is 4.7%. The maximum reduction of 19.1dB is obtained at 1505MHz. For type B the maximum reduction of the coupling coefficient is 28.1dB at 1490MHz and the bandwidth is about 150MHz (the relative bandwidth is 10%).

It is shown that the reduction of the coupling coefficient can be obtained by using a parasitic monopole or a half-loop with a proper h, R and α. The electromagnetic fields radiated from the excited slot induce the electric current on the parasitic wire. This current radiates electromagnetic fields, which cancel the direct fields on the surface of the loaded slot. Thus the amplitude of the magnetic current on the loaded slot is reduced and the reduction is obtained.

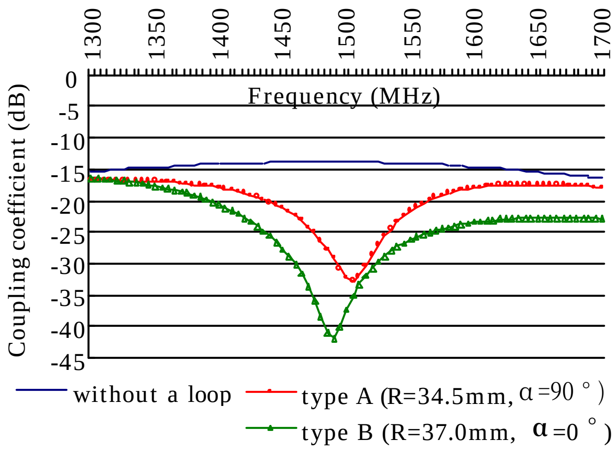


fig. 4 Coupling coefficients with a half-loop.

3.2 Radiation Patterns

Radiation patterns at 1.5GHz with a rectangular ground plane of 365mm(a side parallel to the y axis)×465mm are shown in fig. 5. Angles are measured from the z-axis to the x- or y-axis. These radiation patterns are normalized by the maximum electric field without a parasitic element. The strong electric field is radiated into the backside of the ground plane as the slots radiate electromagnetic fields into both sides of the ground plane. In the x-z plane of fig. 5(a) E_f does not exist, as the axes of the slot are parallel to the y-axis. E_q components at angles of 225° and 315° with a parasitic element are reduced to –20dB for type A. All radiation patterns with a parasitic element show the same tendency. For this reason reradiated and diffracted fields from the parasitic element disturb the radiation patterns in those directions. Figure 5(b) shows radiation patterns in the y-z plane. The maximum E_f is obtained in the z direction and the E_q

component that cannot be seen without a parasitic element is obtained in this plane. The E_q component is radiated from the parasitic element and the maximum radiation is obtained at an elevation angle of 10° from the ground plane by the effect of the diffracted field. These radiation patterns show that the antenna patterns around the +z-axis direction are almost the same with or without a parasitic element.

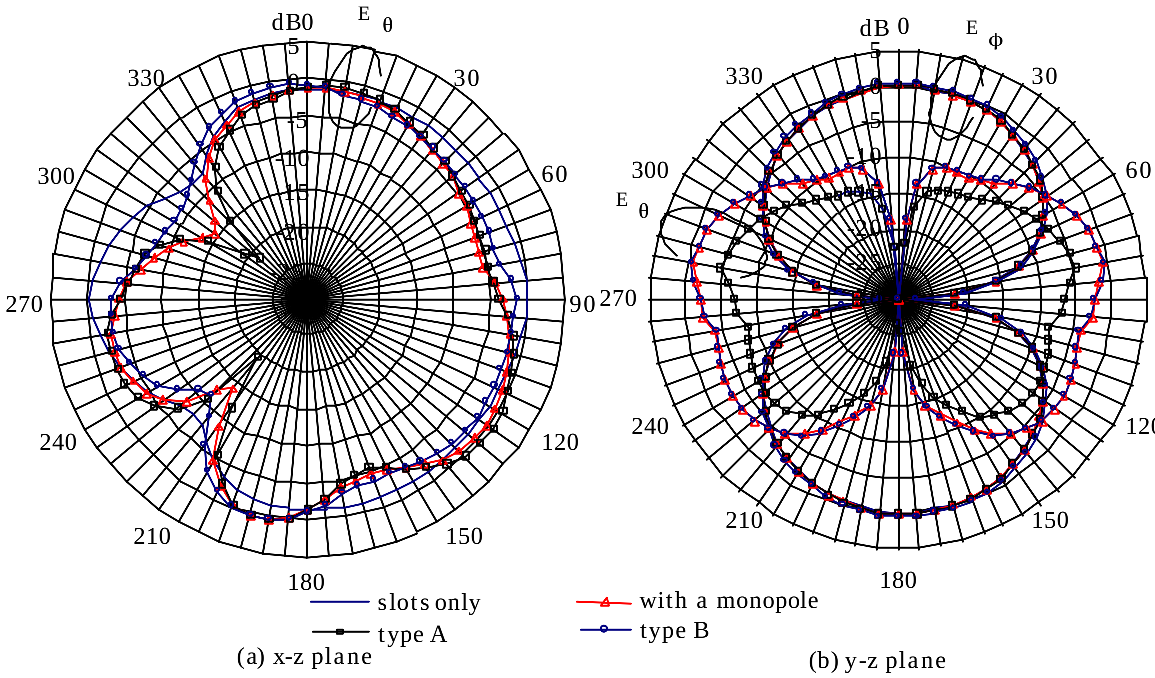


fig. 5 Radiation patterns with a finite ground plane at 1.5GHz.

4. Conclusions

In this paper coupling between two slot antennas on an infinite ground plane and radiation patterns on a finite ground plane are calculated. Numerical results show that the reduction of 13.9dB is obtained by adjusting a monopole height to a quarter wavelength of the operating frequency. Also the properly adjusted parasitic half-loop reduces the coupling coefficient by 24dB. Radiation patterns of the antennas on a 365mm ×465mm ground plane are calculated where the diffracted fields are taken into account shown. It is found that the parasitic elements little affect the antenna patterns around the z-axis direction.

References

[1] S. A. Long, "A combination of linear and slot antennas for quasi-isotropic coverage," IEEE Trans. on Antennas and Propagat., AP-23, pp.572-576, July 1975.

[2] A. Clavin, D. A. Huebner and F. J. Kilburg, "An improved element for use in array antenna," IEEE Trans. on Antennas and Propagat., AP-22, pp. 521-526, July 1974.

[3] T. Morioka and K. Hirasawa, "Reduction of coupling between two wire antennas using a slot," IEICE Trans. on Commun., E80-B, pp.699-705, May 1997.

[4] K. Hirasawa, "Bounds of uncertain interference between closely located antennas," IEEE Trans. on Electromagnetic Compatibility, EMC-26, pp.129-133, Aug. 1984.

[5] R. F. Harrington, "Field computation by moment methods," IEEE, New York, 1993.

[6] G. A. Thiele and T. H. Newhouse, " A hybrid technique for combining moment methods with the geometrical theory of diffraction," IEEE Trans. on Antennas and Propagat., AP-23, pp.62-69, Jan. 1975.