MAXIMUM ELECTROMAGNETIC COUPLING TO A NEARBY CONDUCTING STRIP THROUGH A NARROW SLIT IN THE PARALLEL PLATE WAVEGUIDE

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I. Introduction

The problem of maximum electromagnetic coupling to a conducting strip through a narrow slit in a parallel plate waveguide is considered in expectation of understanding more physics on the coupling mechanism of the aperture coupled microstrip antenna [1]. The emphasis of this article is centered on the description of two kinds of coupling mechanisms observed in the geometry.

The first treatment for a similar problem to the present one was given by Manniko et. al. [2]. But they did not look at radiative coupling mechanism, which we emphasize. As shown in Fig. 1, the geometry under consideration is chosen as a simplified model to 2-dimensional problem of the familiar aperture-coupled microstrip antenna.

II. Analysis

In Fig. 1 the center of the slit is chosen to be an origin of the coordinate. Each material of the upper free half-space, dielectric slab, and the inside of the parallel plate waveguide (PPW) is characterized by (ε_o , μ_o), ($\varepsilon_o\varepsilon_{rd}$, μ_o), and ($\varepsilon_o\varepsilon_r$, μ_o) respectively. The wavenumbers (intrinsic impedances) corresponding to each region are k_o, k_d, and k (η_o, η_d, and η) respectively.

It is assumed that only the *TEM* wave can propagate along the PPW and that the incident *TEM* magnetic field can be given by $H_y^{inc}=(V/\eta h)e^{-jkz}$, in which V is the potential difference across two plates of the PPW and h is the height of the PPW. The location (Z_s) of a shorting wall in the PPW is set to be about $\lambda/2$ from the slit center throughout such that this length (Z_s) of the short stub from the slit center plays the same role as that of the usual $\lambda/4$ open stub of the microstrip line used in the typical aperture coupled microstrip antennas. Here, λ ($=\lambda_o/\sqrt{\varepsilon_r}$, where λ_o = free space wavelength) denotes a wavelength in PPW.

Following the usual integral equation approach by use of methods in [2-4], we obtain a pair of coupled integral equations whose unknowns are the induced current density $\underline{J}_z$ on the strip and equivalent magnetic current density $\underline{M}_y$ over the shorted slit ($x=0^+$) as

$$\underline{E}_t^d(\underline{M}_y) + \underline{E}_t^d(\underline{J}_z) = 0 \qquad \text{on the strip} \tag{1}$$

and

$$\underline{H}_t^p(\underline{M}_y) + \underline{H}_t^d(\underline{M}_y) + \underline{H}_t^d(\underline{J}_z) = \underline{H}_t^{\,i} \quad \text{over the slit} \tag{2}$$

Here the superscript $d(p)$ denotes the dielectric slab (inside of PPW) region, the subscript t denotes the tangential component, and $\underline{H}_t^{\,i}=(2V/\eta h)\, e^{-jkz_s} \cos(k(z-Z_s))$.

From the knowledge of the unknowns obtained through the procedure of the conventional Method of Moments, the quantities of interest such as reflected power P_r, coupled power through the slit P_{slit}, surface wave power P_{sf} launched into the dielectric slab, space wave power P_{sp}, radiation pattern, and the normalized load admittance $\overline{Y_L}(=\overline{G_L}+j\overline{B_L})$ as shown in Fig. 2 can be obtained by use of the methods in [1-4]. In Fig. 2, $\overline{Y_1}$ is an equivalent circuit parameter for the slit (viewed upwardly toward the dielectric slab region) normalized with respect to the characteristic admittance Y_c ($=1/\eta h$) valid for narrow slit case as discussed in [4]. And $\overline{B}$ is a transformed susceptance to the slit location ($z=0$) of the shorting wall admittance ($=\infty$). When $Z_s=0.5\lambda$, $\overline{B}=\infty$. In this case, $\overline{Y_L}$ becomes $\overline{Y_1}$, which can be obtained by use of the methods in [4, 5].

III. Numerical results and discussion

Particular attention is given to investigating the radiative coupling mechanism in the present geometry for the case that all the geometrical dimensions and material constant are chosen to be comparable to those of the conventional microstrip antenna structures.

For this purpose as a first example of the case where the conducting strip is just above and very close to the slit, we have investigated the coupled power P_{slit} (normalized with respect to the incident power P_{inc}) and the load admittance against the offset position of the strip (Z_c/λ_d) with the location (Z_S) of the shorting wall fixed as $Z_s=0.5\lambda$. The results given in Fig. 3 have been calculated for the case that $\varepsilon_r=\varepsilon_{rd}=2.2$, $h=0.015\lambda$, $2a=0.002\lambda$, $h_d=0.01\lambda_d$, $W=0.482\lambda_d$, where $\lambda_d\,(=\lambda_o/\sqrt{\varepsilon_{rd}})$ means a wavelength in the dielectric slab region.

In Fig. 3(a), maximum radiative coupling is observed to occur at two values of $Z_c=\pm 0.226\lambda_d$. At this point the coupling efficiency amounts to 99% (almost perfect matching) and the slit admittance with inclusion of the loading effect of the strip conductor becomes about unity (1+j0) as shown in Fig. 3(b). The reason why there are two matching point locations at $Z_c=\pm 0.226\lambda_d$ can be explained as follows: for this case of $h_d=0.01\lambda_d$, the strip on the dielectric slab and the slitted upper plane of the PPW are considered as a transmission line terminated by open ends, which forms an electromagnetic cavity. So when the cavity is fed through the narrow slit by and incident TEM wave in the PPW whose frequency corresponds to the lowest cavity resonance, the fundamental TM_{01} mode is set up inside the cavity independent of the offset position Z_c of the strip. For the x-component electric field of the TM_{01} mode inside the cavity, we have one half-cycle variation along z direction (i. e., odd symmetrical electric field distribution about $z=0$). Therefore the maximum coupling through the slit is observed at two offset positions symmetrical about the zero offset position, which is also the case with the usual rectangular microstrip patch antenna case [6].

Note that the resonance length W (=0.482λ_d) for the maximum coupling is comparable to that ($\approx 0.464\lambda_d$) obtained by the conventional transmission line model [7] in the microstrip antenna theory. It is interesting to note also that the expression for strip resonant length for maximum radiation is exactly the same as that for resonant thickness for (maximum) transmission problem through narrow slit in thick conducting screens [8] except that the radiation susceptance for the open end of the strip in the former case is replaced by that ([8], p. 619, (43)) of a flanged PPW in the latter case. So the two problems are the same in essence. Only the difference is the location of the feeding point.

As seen from the enlarged view in Fig. 3(b), the conductance of the narrow slit itself with no strip present is very small and the susceptance is capacitive. On the other hand, it is observed that the presence of the nearby strip at the maximum coupling case increases the conductance to about 1 and provides the required inductive susceptance for resonance. It has been also observed that, once a cavity structure between the strip and the upper plate of the PPW is formed, the maximum value and the feature of the two main peaks for the coupled power P_{slit} through the slit are observed to be independent of the separation h_d between the strip and the upper plate of PPW for the range of small values of h_d. This is analogous to the observation that the transmission-slot width product is independent of the actual slot width for a narrow slot at resonance in the foregoing transmission resonance problem [8].

Based upon discussions so far, we will call the present maximum coupling “cavity type” of radiation in order to distinguish this one from the other (parasitic type of radiation) to be discussed.

As the separation h_d between the strip and the upper plate of PPW increases, the feature of two sharp peaks of the maximum p_{slit} (as shown in Fig. 3-(a)) becomes weaker and the strip length, which gives the maximum of p_{slit}, decreases gradually. That is, as the separation h_d increases, two sharp peaks approach to each other and at the same time, the region of the small coupling level of p_{slit} between two peaks swells up gradually. Consequently a different feature of the coupling is observed to occur as shown by the solid line marked with dots in Fig. 4 (calculated for the case that $\varepsilon_r=2.2$, $\varepsilon_{rd}=1.2$, $h=0.015\lambda$, $2a=0.002\lambda$, $h_d=0.044\lambda_d$, $W=0.425\lambda_d$, $Z_c=0$, $Z_s=0.5\lambda$). So p_{slit} becomes a maximum when the offset is zero. This behavior of the radiative coupling efficiency (meaning also the impedance matching characteristics) is similar to that in the aperture coupled microstrip antenna discussed in [1]. Besides the maximum coupling level of p_{slit} is insensitive to the offset near $Z_c=0$ and so almost constant over the range of $|Z_c/\lambda_d|\leq 0.1$. This is in contrast with the case of the foregoing cavity type of radiation. In Fig. 4, the admittance behavior of the slit for this case is also observed to be quite different from that for the cavity type case. That is, here the maximum radiative coupling through the slit occurs when the conductance ($\approx$1) and susceptance of the slit become stationary values, i.e., their own maximum and minimum, respectively, in contrast to the admittance behavior for the cavity type case. Note that $h_d=0.044\lambda_d$, $W=0.425\lambda_d$, and $\varepsilon_{rd}=1.2$ (low dielectric constant) for this case of Fig. 4. For these values, strong cavity mode field is not excited underneath the strip at the frequency of $f=c/(\lambda_d\sqrt{\varepsilon_{rd}})$(c: light velocity in free space). Rather these values are comparable to those corresponding to the parasitic director length and the spacing between driver and director respectively appearing as an example of a two-element Yagi-Uda structure in some textbook [9]. What is more, some enhancement in directivity has been observed in comparison with the foregoing cavity case. The magnitude of the induced current on the strip has been also found to be much smaller than that for the cavity case. So this type of coupling is designated "parasitic type of radiation" against the former cavity type.

Though the two types of coupling are quite different from each other as discussed so far, the maximum available coupling (radiation) efficiencies for both the types are observed to be almost the same, amounting to about 100% for appropriately chosen parameters as shown in Fig.3 and 4.

IV. Conclusion

We have considered the electromagnetic coupling problem to a nearby conducting strip through a narrow slit in a parallel plate waveguide as a simplified model of the familiar aperture-coupled microstrip antenna structure. It is found that there are two kinds of (maximum) coupling mechanisms, cavity and parasitic types, depending on the separation between the conducting strip and the upper plate of PPW. This study may be helpful in understanding more the radiation mechanism of the conventional aperture-coupled microstrip antenna.

Acknowledgment

This work was supported in part by KOSEF 97K3-0808-03-01-3, Marc, and BK21.

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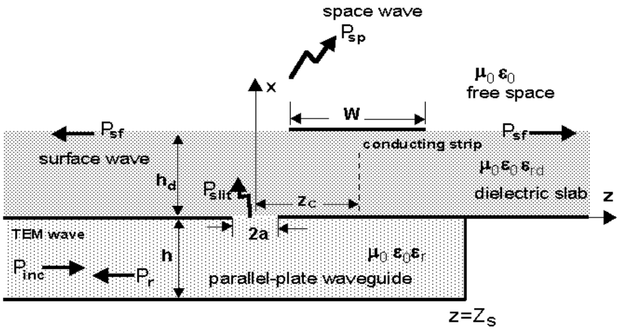


Fig. 1. Geometry under consideration

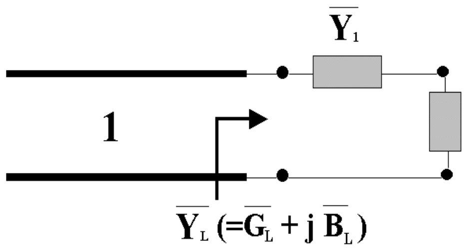
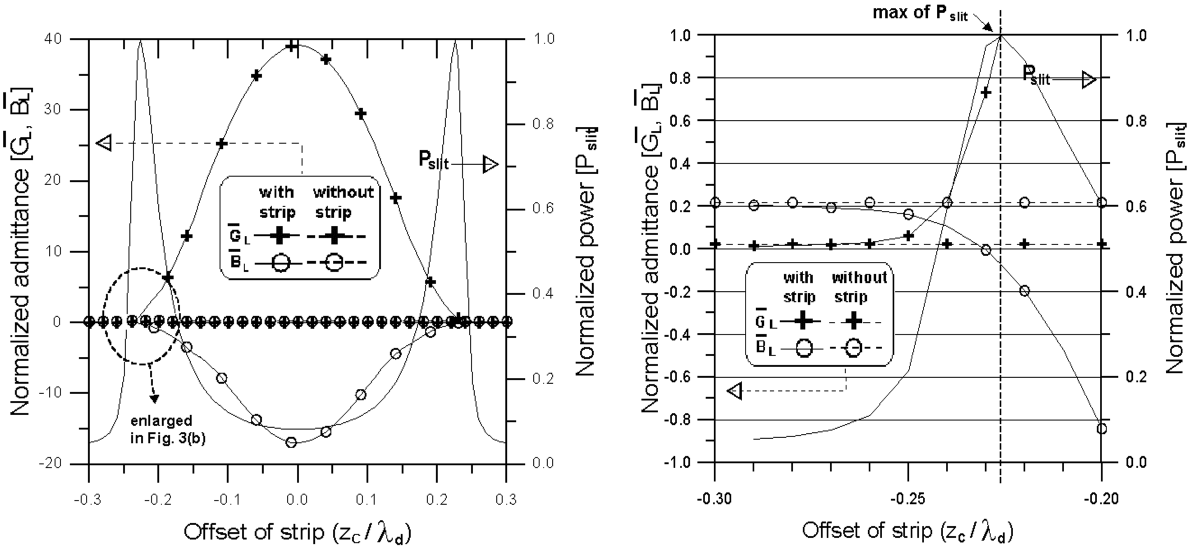


Fig. 2. Equivalent circuit representation



(a) (b)

(a) Normalized coupled power P_{slit} through the slit and load admittance $\overline{Y_L}\ (=\overline{G_L}+j\,\overline{B_L})$ versus offset (λ_c/λ_d)

(b) Enlarged view of curves for P_{slit} and $\overline{Y_L}\ (=\overline{G_L}+j\,\overline{B_L})$ near the offset position for maximum coupling

Fig. 3. Normalized coupled power through the slit and load admittance (cavity type)

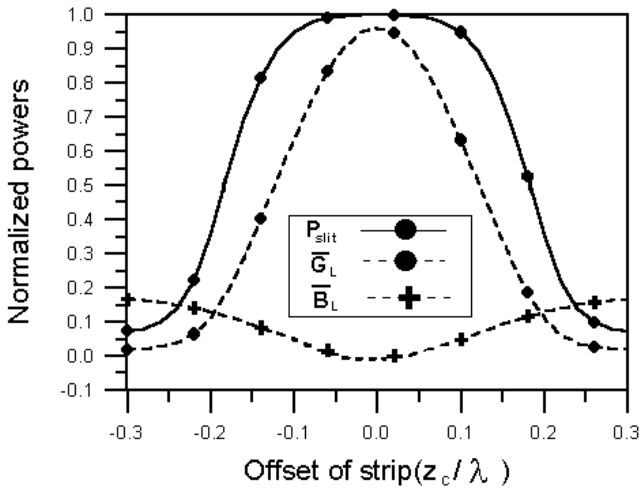


Fig. 4. Normalized coupled power P_{slit} through the slit and load admittance $\overline{Y_L}\ (=\overline{G_L}+j\,\overline{B_L})$ (parasitic type)