Coupling Through a Slit in a Flanged Parallel-Plate Waveguide with a Conducting Strip

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1. Introduction

The radiation characteristics of a flanged parallel-plate waveguide (FPPW) with objects in the vicinity of the aperture have been investigated by some authors [1,2], because one may obtain the physical insights of the radiation properties of more practical structures such as waveguide-fed slot antenna[3], short backfire antenna[4], etc. In the most of the studies, however, it seems that the role of a nearby (conducting) scatterer on the radiation properties of the structure and the maximum coupling (or impedance matching) mechanism have not been discussed throughly.

In this study, the problem of coupling through a slit in a FPPW to a conducting strip near the aperture is considered for the purpose of gaining an insight into the impedance matching (or maximum coupling) phenomena through the investigation of the equivalent admittance properties of the coupling slit and complex powers in the geometry under consideration. Furthermore, the effects of the guide height and the slit width on the radiation properties have been examined along with discussions about the maximum coupling phenomena through a narrow slit due to a nearby conducting strip.

2. Theory

Fig. 1 shows a cross sectional view of a flanged parallel-plate waveguide (FPPW) opening into the half (z>0) free space (μ_0, ϵ_0) where is located a conducting strip parallel to the ground plane (z=0). The y-component TEM magnetic field incident on the aperture region can be given by $H_y^{\rm inc} = V_{\rm inc}/(\eta h)\exp[-jkz]$ in which $V_{\rm inc}$ is the potential difference across the plates, $k(=k_0\sqrt{\epsilon_r} = 2\pi/\lambda,\ k_0 = \omega\sqrt{\mu_0\epsilon_0} = 2\pi/\lambda_0)$ is the wavenumber, and $\eta(=\eta_0/\sqrt{\epsilon_r},\ \eta_0 = \sqrt{\mu_0/\epsilon_0})$ is the intrinsic impedance of the dielectric inside the FPPW.

Following Butler *et al.* [2], the present problems can be formulated in terms of coupled integro-differential equations for the induced electric current density $J_x(x)$ on the conducting strip and the magnetic current density $M_y(x)$, equivalent to the slit electric field $E_x^A(x) \equiv E_x(x,0)$, over the shorted slit as

$$(H_y^{J_0} + H_y^{M_0})|_{z=0_+} = (H_y^M + H_y^{sc})|_{z=0_-}\ ,\ \ |x - X_a| < a/2 \quad \text{(across the slit)} \quad (1)$$

$$E_x^{J_0}(\rho) + E_x^{M_0}(\rho) = 0\ ,\ \rho \in C \quad \text{(on the surface of strip).} \quad (2)$$

Here, the subscript x (y) indicates an x (y) component, $H_y^{J_0}$ $(H_y^{M_0})$ is the magnetic field in the free space due to J_x $(-M_y)$, H_y^M is the magnetic field inside the FPPW due to M_y over the shorted slit, and H_y^{sc} represents the short circuit magnetic field given by $2V_{\rm inc}/(\eta h)\cos kz$. The detailed expressions for the terms in (1) and (2) are given in [2]. In order to solve the equations (1) and (2) numerically using moment method, the unknown distributions $J_x(x')$ and $E_x^A(x')$ are expanded in terms of piecewise sinusoidal functions and pulse functions, respectively, and Galerkin's scheme is employed to reduce (1) and (2) to a linear equation system.

From knowledge of the distributions, one may obtain all the quantities of interests such as the voltage reflection coefficient Γ_V from the slit, the equivalent slit admittance $Y_s(=G_s + jB_s)$, the reflected power P_r from the slit, the coupled power P_s through the slit into the exterior free space region, and radiation pattern. All the power quantities are normalized to the incident TEM mode power P_i, for convenience, by setting $P_i = 1$.

Equivalent circuit representation [5] for the TEM mode in the guide is depicted in Fig. 2 where

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$Y_C(=1/\eta h)$ is the characteristic admittance and $y_s(=Y_s/Y_C)$ is the normalized equivalent slit admittance obtained from the voltage reflection coefficient Γ_V as

$$y_s = g_s + jb_s = (1-\Gamma_V)/(1+\Gamma_V) \ . \tag{3}$$

The equivalent slit susceptance b_s is associated with the nonpropagating reactive fields in the vicinity of the slit and so it could be expressed as the sum of b_{si} and b_{se} , as shown in Fig. 2, in which the reactive powers inside and outside the guide regions are taken into account, while the coupled (or radiated) power $P_s(=P_i-P_r)$ through the slit into the free space reflected in the conductance g_s .

In order to explain the role of the conducting strip in achieving impedance matching (or maximum coupling) in the geometry of Fig. 1, radiation and equivalent admittance properties of a slit in a FPPW in the absence of the nearby conducting strip should be examined. The reactive power $Q_{\rm in}$ ($Q_{\rm ex}$) in the region $V_{\rm in}$ ($V_{\rm ex}$), see Fig. 3, inside (outside) the PPW can be computed by integrating complex Poynting vector ($\underline{E}\times\underline{H}^*$) over the surface $S_{\rm i}$ ($S_{\rm e}$) enclosing $V_{\rm in}$ ($V_{\rm ex}$) as

$$Q_{\rm in(ex)} = \frac{1}{2}{\rm Im}\left\{\oint_{\rm s_{i(e)}} \underline{\rm E}\times\underline{\rm H}^* \cdot \underline{\rm ds}\right\}. \tag{4}$$

Then the reactive powers in the lossless medium $V_{\rm in}$ and $V_{\rm ex}$ could be expressed by

$$Q_{\rm in} = \frac{1}{2}{\rm Im}\left\{\sum_{m=1}^{N}\sum_{n=1}^{N} V_m (V_n Y_{mn}^h)^*\right\} \tag{5}$$

$$Q_{\rm ex} = \Gamma_V - \Gamma_V^* - Q_{\rm in} \tag{6}$$

in which

$$Y_{mn}^h = j\Delta^2 \frac{2k}{\eta h}\sum_{q=1}^{\infty}\frac{\cos[(x_m+0.5h)q\pi/h]\cos[(x_n+0.5h)q\pi/h]}{\sqrt{(q\pi/h)^2-k^2}}\left[\frac{\sin(q\pi\Delta/2h)}{(q\pi\Delta/2h)}\right]^2 \tag{7}$$

with $\Gamma_V = \sum_{n=1}^{N} V_n\Delta/V_{\rm inc}$, $x_n = X_a - a/2 + (n-0.5)\Delta$, $\Delta = a/N$, and total segment number N. Accordingly total reactive power in the vicinity of the slit is $Q_{\rm slit} = Q_{\rm in} + Q_{\rm ex} = 2{\rm Im}\{\Gamma_{\rm V}\}$. Now the equivalent slit susceptance b_{si} and b_{se} could be given as

$$b_{si(se)} = \frac{Q_{\rm in(ex)}}{Q_{\rm slit}} b_s. \tag{8}$$

If the slit is narrow relative to the guide height h as well as the guide wavelength λ and the electric field distribution is assumed to be uniform over the slit, then the uniform electric field E_s can be obtained as

$$E_s = \frac{2V_{\rm inc}/a}{A+j(B_i+B_e)} = \frac{2V_{\rm inc}/a}{A+jB} \tag{9}$$

in which

$$A = 1+\frac{k_o\eta h}{2\eta_o},\ B_i = 2k\sum_{q=1}^{\infty}\frac{\cos^2[q\pi(X_a+h/2)/h]}{\sqrt{(q\pi/h)^2-k^2}}\left[\frac{\sin(q\pi a/2h)}{(q\pi a/2h)}\right]^2,\ B_e = \frac{k_0\eta h}{\eta_0\pi}\left(\log\frac{2}{k_0 a 1.781}+1.5\right).$$

Now the equivalent admittance y_s of a narrow slit is

$$y_s = g_s + jb_s = (1-\Gamma_V)/(1+\Gamma_V) = (A-1) + j(B_i + B_e) \tag{10}$$

where $\Gamma_V = E_s a/V_{\rm inc} - 1$. From (10), one finds that the impedance matching ($y_s = 1 + j0$) for the case of narrow slit requires the conditions $A=2$ and $B=0$ (or $B_e = -B_i$), although both of them could not be satisfied without the nearby scatterer (conducting strip). Hence, in order to achieve the impedance matching, the arrangement of a scatterer in the vicinity of the narrow slit should bring a new complex term $A_s + jB_s$ to be appended to the denominator $A+jB$ in (9) as

$$E_s = \frac{2V_{\rm inc}/a}{(A+A_s)+j(B+B_s)} = \frac{2V_{\rm inc}/a}{A'+jB'}. \tag{11}$$

The equivalent slit admittance, owing to a nearby scatterer, would be changed to $y_s = (A' - 1) + jB'$ where b_{si} remained to be B_i while b_{se} is modified to $B_e + B_s$. For the case that $A' = 2$ and $B' = 0$ in (11), the slit admittance is matched to the characteristic admittance of the feed line (i.e., $y_s = 1 + j0$) and the maximum coupling ($P_s \simeq P_i$) would occur. In order for the slit to be resonated, B' in (11) would vanish and then B_s should be negative (inductive) because both B_i and B_e are positive (capacitive), which means that the nearby scatterer should append the required inductive susceptance B_s to the geometry having capacitive susceptance component $B(=B_i + B_e)$ and small conductance $A - 1$ for the case of no nearby scatterer.

3. Results and Discussions

The equations (1) and (2) have been solved numerically using moment method by expanding J_x and E_x^A in terms of piecewise sinusoidal functions and pulse functions, respectively, and employing Galerkin's scheme. In Fig. 3, the normalized slit susceptances $b_s(=b_{si} + b_{se})$ are plotted for different aperture-to-guide height ratios ; a/h = 0.1, 0.4, and 1. From the results for b_{si} and b_{se} in Fig. 3, one finds that the reactive power $Q_{\rm in}$ in the waveguide region is much smaller than $Q_{\rm ex}$ in the free space region as far as the ratio a/h is near unity. However, as the ratio a/h is decreased, the portion of b_{si} as well as the susceptance b_s is increased. It is interesting to note that, for the case of thin waveguide (i.e., $h/\lambda_0 \ll 1$), the susceptance b_{se} is much greater than b_{si}, irrespective of the ratio a/h, as shown in Fig. 3. Another interesting point is that, within the limit of $ka \rightarrow 0$, b_s logarithmically approached to infinity so as to give the voltage reflection coefficient $\Gamma_v = -1$ (as it should, for a short ended circuit) while g_s remained constant [see (10)], regardless of the slit width a, though the divergent behaviors of b_s depend on the guide height h, similarly as those of a narrow transverse slit in a PPW [5,7].

As the slit width approaches to zero, the unnormalized conductance $G_s(= g_s Y_C)$ of a thin slit converges to, irrespective of a/h, $1/120\lambda_0$ which has also been observed in the problems of the equivalent series admittance of a narrow transverse slit and the equivalent load admittance of a wide transverse slit in a PPW, in the prior works [5,7].

The effect of slit offset X_a on the slit susceptances b_{si} and b_{se} for the case of ϵ_r=1, a/h=0.1, and h/λ_0 =0.4 is shown in Fig. 4 where b_{si} is increased along with the slit offset X_a while b_{se} is remained as it is.

The equivalent slit admittances for the case that ϵ_r=1 and a/h=1 are plotted in Fig. 5 where our results agree fairly well with the data in [8] which have been obtained under the assumption that the field distribution in the slit is of the form of the incident TEM wave.

From the above results for the slit admittance, one finds that, in most cases, the slit is impedance mismatched with the feed waveguide especially when the guide height is low or the slit is narrow in a thick waveguide, since the slit has small conductance g_s and its susceptance b_s does not vanish for the impedance matching(g_s=1 and b_s=0). Hence it is needed to place some types of scatterers in the vicinity of the slit so that it may be impedance matched to the feed waveguide, as discussed in the foregoing section.

However, it should be noted that the FPPW with large guide height itself behaves as an efficient radiator without conducting strip [for reference, $P_s \geq$ % when $k_0 h \geq$ 7, ϵ_r=1, and a/h=1], though the impedance matching performance could be improved due to the conducting strip. As an example, maximum coupling(i.e., $P_s \simeq P_i$ or $y_s \simeq$1+j0) has been observed for the case that ϵ_r=1, h=0.29λ_0, a/h=1, L=0.29λ_0, X_0=0, and Z_0=0.227λ_0, where the antenna gain(4.91dBi) is improved from 3.35dBi without strip (P_s=91.6% and y_s=0.77+j0.48). In this case, the ratio of -10dB impedance bandwidth Δf, over which P_s/P_i is greater than 90%, to the maximum coupling frequency f_0 is $\Delta f/f_0 \simeq$63% and the antenna gain is greater than 3.65dBi over the bandwidth. Hence the maximum coupling structure of this type is useful for wideband application.

On the contrary, the maximum coupling structures which consist of a narrow slit, regardless of a/h, and a nearby conducting strip have been found to be useful for the high gain antenna application at the expense of the frequency bandwidth. Impedance matching has been achieved, for example, when $h = 0.29\lambda_0$, a/h=0.1, $X_0 = 0$, $Z_0 = 0.093\lambda_0$, and $L = 0.519\lambda_0$, in which case the structure has the antenna gain greater than 6.63dBi over the -10dB impedance bandwidth $\Delta f/f_0 \simeq 11.4\%$.

Another method for obtaining impedance matching of a narrow slit by use of a nearby conducting strip is to adjust the lateral displacement X_0 of the strip center for given strip length L and distance Z_0 from a slit. Maximum coupling phenomena similar to the cavity-type coupling phenomena [9] occurred in the coupling problem through a narrow transverse slit in a PPW with a conducting strip have also been observed in the present geometry of Fig. 1, as expected.

4. Conclusion

The problem of electromagnetic coupling through a slit in a flanged parallel-plate waveguide to a nearby conducting strip has been considered with main interests centering on the impedance matching mechanism or maximum coupling phenomenon. The equivalent slit susceptances associated with reactive powers near the coupling slit inside and outside the guide have been obtained. In addition, the effects of the conducting strip on the maximum coupling have been discussed from the viewpoints of equivalent admittance properties and complex powers.

Acknowledgement

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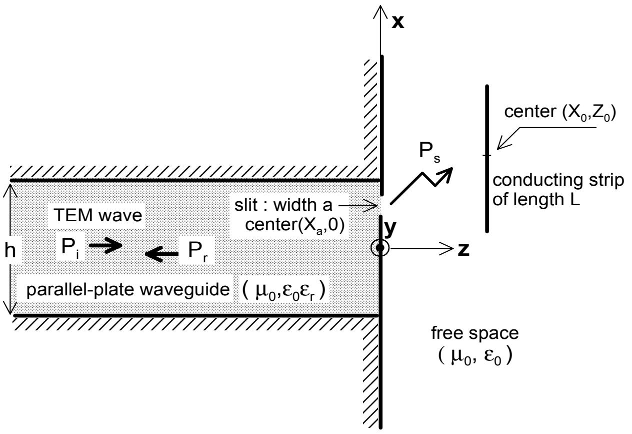

Fig. 1. Geometry.

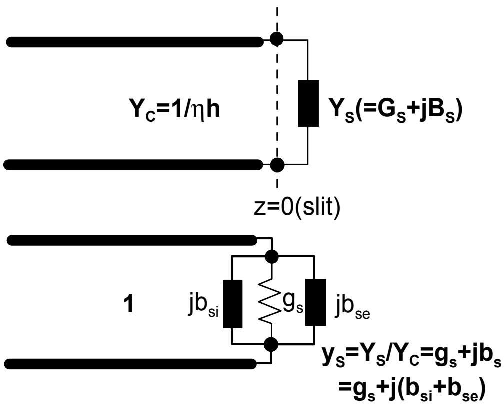

Fig. 2. Equivalent circuit.

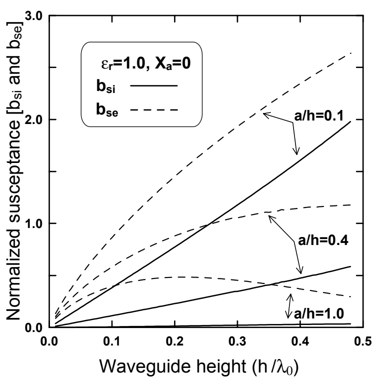

Fig. 3. Slit susceptance.

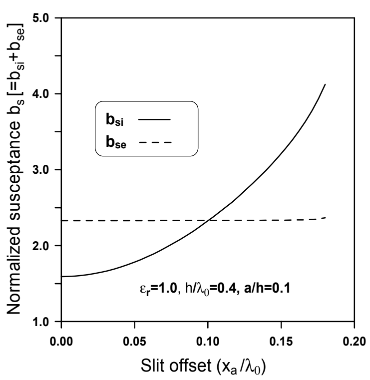

Fig. 4. Slit susceptance.

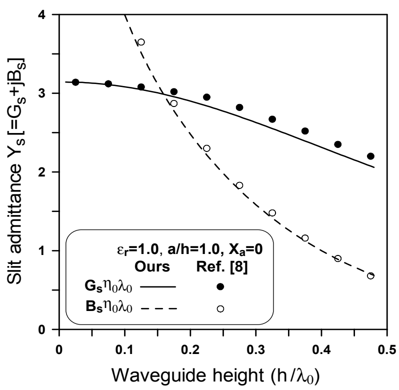

Fig. 5. Slit admittance.