Theoretical Analysis of Coupling and Cross Polarization of Perpendicular Slot Antennas on a Dielectric Half-Space

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***Abstract*— In this paper, design curves are given for minimizing the coupling between two perpedicular slot antennas on a dielectric half space. Such antennas may be utilized in polarimetric receivers in which coupling must be minimized to ensure polarization purity and in polarization diplexed quasi-optical receivers in which the local oscillator (LO) is received in a polarization perpendicular to that of the RF signal. For this analysis, Galerkin's method in the spectral domain is applied along the length of the slots and point matching across their widths. At the second resonance, for equal length slots and for constant width/length (w/L) of the slots, there is a decrease of about 2-dB coupling for a factor of three increase in dielectric constant ($\epsilon_1 = 1.0 \rightarrow$ 3.8 and $\epsilon_1 = 3.8 \rightarrow$ 12.8). For fixed dielectric constant there is a 1–2-dB increase in coupling for a factor of two increase in w/L. For slots of unequal length ($L_2 = L_1/2$), the changes are even smaller. Design curves are shown for various relative positions of the slots in two dimensions. A strong correllation between coupling levels and peak cross-polarization levels is found.**

***Index Terms*—Moment methods, slot antennas.**

I. Introduction

PREVIOUSLY, work has been done on mutual impedance between arbitrarily positioned slots on a semi-infinite dielectric substrate that are aligned parallel to each other [1]. In this paper, the analysis is extended to multiple slots that are aligned perpendicular to each other and may be positioned arbitrarily with respect to each other (Fig. 1). This type of analysis is useful for determining the mutual impedance between slot antenna elements that are intended to receive or transmit different polarizations. Such geometries may be encountered in polarimetric systems as well as high-frequency quasi-optical receivers that utilize different polarizations for the RF and local oscillator (LO) signals.

This analysis is done assuming an infinite dielectric since substrate lenses, which approximate an infinite dielectric, provide an attractive method of mounting slot antennas in receiver systems [2], [3]. The object is to obtain the self and mutual impedance of the slots from which the coupling can be calculated. The electric field in the slot for a given excitation can also be obtained. These fields are used to obtain radiation patterns.

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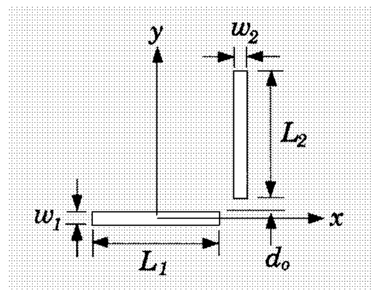


Fig. 1. Slots on a dielectric half-space. Slot 2 is centered at $(x_o, y_o, 0)$ and slot 1 is centered at the origin. The minimum space between the slots in the y-direction is defined as $d_o = y_o - L_2/2 - w_1/2$. The dielectric region ϵ_1 is $z < 0$.

First, a theory of fields for slot antennas on a dielectric substrate is given. The result is an algebraic equation in the Fourier domain for the electric field in the slots. These currents are expanded using fast converging basis functions. Galerkins method is applied along the length of the slots and point matching for its narrow dimension. An admittance matrix of inner products results. Coupling between slots is then calculated and compared with experimental results. Copolarized and cross-polarized radiation patterns will be presented to examine the correllation between slot-to-slot coupling and radiatied cross-polarization levels. Experimental verification of the calculated coupling is also presented.

II. Theory

The approach taken here is to write the electric vector potential in terms of the equivalent magnetic currents in the slots—a two-dimensional (2-D) integral equation. This can be reduced to a purely algebraic equation when transformed to the spectral domain [1]. After the method of moments is applied, double infinite integrals result again. The singularity in the Green's function can be removed from the problem analytically and the integral transformed such that they are semi-infinite in one dimension and finite in the other. We thus avoid having to work with quadruple integral's resulting from the application of the method of moments in the space domain. However, the order of the integral in the space domain approach can also be reduced analytically and has effectively been applied to folded-slot antennas [4].

After matching the tangential magnetic and electric fields at $z = 0$ in the slots, writing the magnetic field in term of the electric vector potential and transforming the problem into the spectral domain, we obtain (1) for the electric source currents $\tilde{J}$ in terms of the admittance matrix and the electric field in the slots

$$\begin{bmatrix} \tilde{J}_x \\ \tilde{J}_y \end{bmatrix} = \begin{bmatrix} Y_{xx} & Y_{xy} \\ Y_{yx} & Y_{yy} \end{bmatrix} \begin{bmatrix} \tilde{E}_x \\ \tilde{E}_y \end{bmatrix}_{z=0} \quad (1)$$

in which

$$Y_{xx} = \frac{2}{jk_o\eta_o}[(k_y^2 - k_o^2)Q_0 + (k_y^2 - k_1^2)Q_1] \quad (2)$$

$$Y_{yy} = \frac{2}{jk_o\eta_o}[(k_x^2 - k_o^2)Q_0 + (k_x^2 - k_1^2)Q_1] \quad (3)$$

$$Y_{yx} = Y_{xy} = \frac{-2k_xk_y}{jk_o\eta_o}(Q_0 + Q_1) \quad (4)$$

and in which Q_0 and Q_1 are the Fourier transforms of the 2-D Green's function in free-space and in the dielectric, respectively. Details of the derivation of this equation have been reported elsewhere and will not be repeated [1], [4]–[6].

This result is general to any arbitrary opening in a ground plane on a dielectric substrate. Antenna geometry is determined by where the basis functions used to expand the tangential electric fields are placed. Where there is no basis function, the tangential electric field is zero due to the ground plane. We also note that equations (1)–(4) are essentially identical to those used for periodic planar screens in which k_x and k_y are quantized [6].

Using standard moment-method procedures, the electric fields are expanded in terms of piecewise sinusoidal basis functions along the length of the slots and a potential well [7, ch. 5] across their widths. Testing functions chosen correspond to Galerkins technique along the length and point matching across the width of the slots. We configure the basis function such that there is only a field directed across the width of the slots and none oriented along the length. We therefore limit our analysis from this point on to narrow slots.

Admittance matrix elements are determined in a similar manner as those in [1]. The admittance matrix integrals are oscillatory but have no singularities. Standard Quadpack [8] automatic quadrature routines specifically intended for oscillatory integrands obtained high accuracy in this paper.

The S-parameter matrix is calculated via the matrix equation

$$[S] = ([z] - [I])([z] + [I])^{-1} \quad (5)$$

in which $[z]$ is the normalized impedance matrix $[Z]/Z_o$ and $[I]$ is the identity matrix. Elements of the impedance matrix are calculated via

$$Z_{ij} = 4\pi^2 \frac{V_i}{I_j}_{I_k=0, k\neq j} \quad (6)$$

in which V_i is the coefficient of the basis function in the center of antenna element i and I_i is the result of the inner product of a testing function in the center of slot i and the source current $\tilde{J}$ (which is in the center of slot i). The $4\pi^2$ is required by Parseval's relation in the transformation to the space domain. Both $[Z]$ and $[S]$ are symmetric. In calculating Z_{11} we obtain the same results as those previously published [1], [9], [10].

III. Results

In this section, the coupling between two slots oriented at right angles is investigated numerically and confirmed experimentally. The coupling is defined as the magnitude of the S parameter $|S_{21}| = |S_{12}| = \sqrt{P_{\text{Load}}/P_{\text{Source}}}$ in which P_{Load} is the power received at the port of one of the slots terminated in $Z_o = 50\,\Omega$ due to a 50 Ω source (P_{Source}) at the port of the other slot. The coupling is the same regardless of which slot the source is in since the perpendicular slots constitute a reciprocal antenna system [11]. We choose to show coupling as opposed to mutual impedance because it is easy to relate coupling to cross-polarization levels.

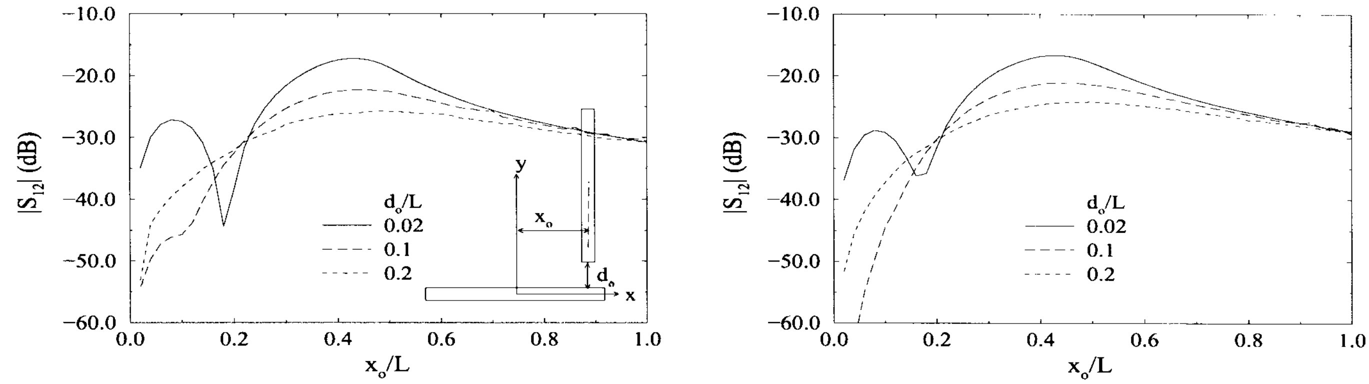
The physical layout is depicted in Fig. 1. First we consider cases in which $L_2 = L_1 = L$ and $w_2 = w_1 = w$. In a polarimetric receiver application each slot is intended to receive a different polarization of radiation in the same frequency range. Thus, equal length antennas are considered. Then cases will be considered in which $L_2 = L_1/2$ and $w_2 = w_1/2$. These are for quasi-optical subharmonic receiver design (of which we give an example) in which the smaller antenna receives the RF signal information and the larger antenna receives the LO signal, which is at approximately half the frequency of the RF signal. The use of different polarizations aids in simplification of the multiplexer design. The last part of this section compares our calculations to measurements.

A. Calculations—Equal Length Slots

The slot-to-slot coupling (S_{12}) has been calculated for equal length slots on three commonly used dielectrics: free-space, which would include the approximation of a very thin low permittivity dielectric; fused quartz ($\epsilon_1 = 3.8$); and GaAs ($\epsilon_1 = 12.8$). Figs. 2–4 were all computed at the second resonance of a single slot on the given dielectric. The second resonance is defined as the frequency value for which the imaginary part of the self impedance of a slot becomes zero for the second time. For slots, the second resonance input impedance (30–40 Ω) is much lower than that of the first resonance (≈400 Ω) and is, therefore, much more commonly used [10]. The second resonance free-space wavelength is λ_2 and is a function of slot width for constant slot length. For a given dielectric and for an increase of in width/length (w/L) of 0.02–0.04, there is a slight increase (1–2 dB) in coupling for $x_o/L > 0.2$ and a slight decrease (2–3) in coupling for $x_o/L < 0.2$—referring to peaks above and below $x_o/L = 0.2$. There is also an overall decrease of about 2 dB between plots of constant w/L and increasing relative permittivity, ϵ_1, again referring to peaks [i.e., Figs. 2(a), 3(a), and 4(a)].

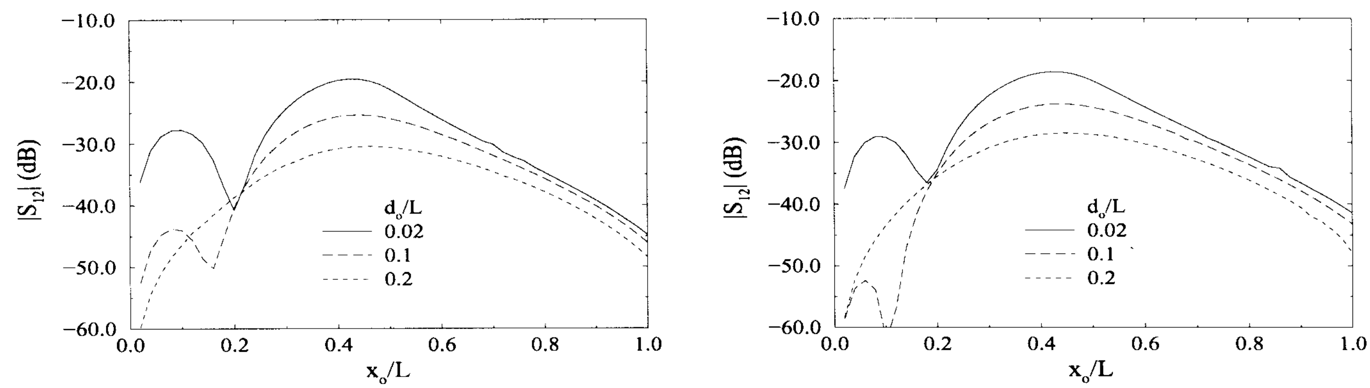
Although not shown here to conserve space, the case of $w/L = 0.08$ was also calculated. Curves shapes were similar except that the dip at $x_o/L = 0.2$ for $d_o/L = 0.02$ was not as deep. Also, the coupling below $x_o/L = 0.2$ was 5–6 dB lower for the $w/L = 0.08$ case when compared with the $w/L = 0.04$ case. Above $x_o/L = 0.2$ it was 1–2 dB higher at the peak.

Knowing the fourier transform of the electric field in the slots, it is easy to obtain the far-field radiation pattern [9]. For the copolarized and cross-polarized field references we use



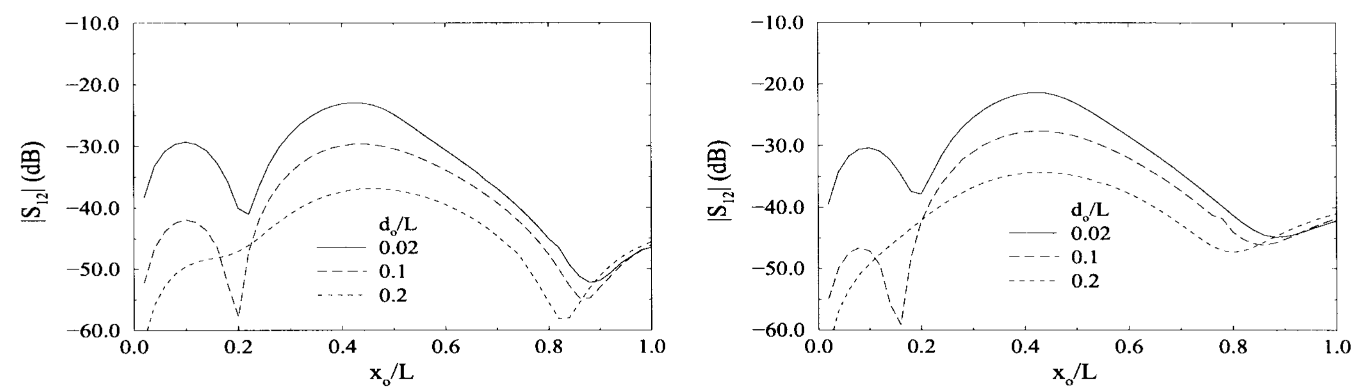
(a) (b)

Fig. 2. Coupling between equal length slot antennas on free-space dielectric ($\epsilon_1 = 1.0$). In all cases, the slot length is chosen so as to be at the second resonance and λ_2 is the free-space wavelength at the second resonance. (a) $w/L = 0.02, L = 0.840\lambda_2$. (b) $w/L = 0.04, L = 0.794\lambda_2$.



(a) (b)

Fig. 3. Coupling between equal length slot antennas on fused quartz ($\epsilon_1 = 3.8$). In all cases, the slot length is chosen so as to be at the second resonance. (a) $w/L = 0.02, L = 0.546\lambda_2$. (b) $w/L = 0.04, L = 0.515\lambda_2$.



(a) (b)

Fig. 4. Coupling between equal length slot antennas on GaAs ($\epsilon_1 = 12.8$). In all cases, the slot length is chosen so as to be at the second resonance. (a) $w/L = 0.02, L = 0.325\lambda_2$. (b) $w/L = 0.04, L = 0.305\lambda_2$.

a slightly modified Ludwig's "Definition 2" [12]. We use an x-directed magnetic dipole as the reference field for copolar calculations and an x-directed electric dipole for cross-polar calculations when the source is in the x-directed slot. In case the y-directed slot is excited, such as in Section III-B2), then the reference fields are due to y-directed dipoles. The source-fed slot induces fields in the adjacent perpendicular slot. These induced fields then radiate with a polarization perpendicular to the source-fed slot. Therefore, we conjecture that there should be a correllation between the coupling level and the peak cross-polarization level.

The copolarized and cross-polarized radiation patterns were calculated for free-space and GaAs with the source in slot 1. As with all radiation patterns in this paper, radiation patterns are in the infinite dielectric side of the slots since we assume that the slots are to be mounted on a dielectric lens when $\epsilon_r > 1$. Two

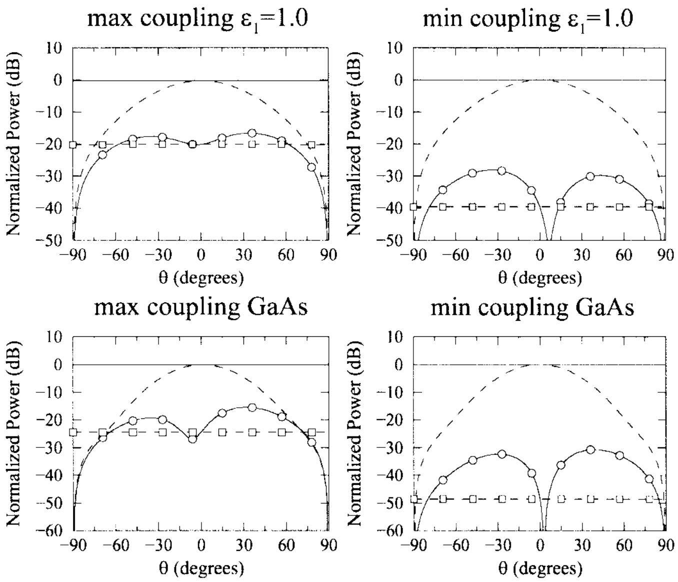


Fig. 5. Normalized radiation patterns for $L2 = L1$, $w/L = 0.04$, $d_o/L = 0.02$. ——— E plane copol, ○——○ E-plane cross pol, - - - -H-plane copol, □- - - -□ H-plane cross pol.

particular cases were examined for $w/L = 0.04$ and $d_o/L = 0.02$: x_o/L corresponding to maximum coupling and x_o/L corresponding to minimum coupling not including $x_o/L = 0$. We ignore $x_o/L = 0$ because there is not any coupling or cross-polarization levels. Because of the symmetry, there are no fields induced in slot 2 by slot 1 or vice versa when $x_o/L = 0$.

Fig. 5 shows the radiation patterns. For GaAs, the radiation patterns shown are those into the infinite dielectric. For free-space ($\epsilon_1 = 1.0$) and GaAs, minumum coupling is at $x_o/L = 0.16$ and $x_o/L = 0.20$, respectively. Maximum coupling is at $x_o/L = 0.42$ for both free-space and GaAs. In free-space there is a 10-dB decrease in peak cross-polarization levels in the E plane and 20-dB decrease in the H plane from the position of maximum coupling to the position of minimum coupling. For GaAs, the difference in cross-polarization levels is about 15 dB in the E plane and 25 dB in the H plane. The E plane is the y-z plane and the H plane is the x-z plane for Fig. 5. The D plane ($\phi = 45^o$) cross-polarization pattern is not shown here because it is almost exactly the same as the E-plane cross-polarization pattern.

B. Calculations—$L_2 = L_1/2$

1) Source in Long Slot: Also calculated was the mutual coupling between two slots for which $L_2 = L_1/2$ and $w_2 = w_1/2$ at the second resonance slot 1, the longer slot. The design curves for the case $L_2 = L_1/2$ at the second resonance of slot 1 have the same general shape and a slight shift in magnitude as compared to equal length slots. Rather than reproduce very similar figures, Table I shows the peak value of coupling for $L_2 = L_1/2$ at the second resonance of slot 1. This peak value occurs in the same place as for equal length slots ($x_o/L \approx 0.43$). A very good approximation to the coupling at any other value of x_o/L can be obtained by subtracting the value in Table I from peak value from Figs. 2-4 and then subtracting this difference from Figs. 2–4.

TABLE I
CALCULATED PEAK VALUE OF $|S_{12}|$ IN DECIBELS FOR $L_2 = L_1/2$ AND $L_2 = w_1/2$ AT THE SECOND RESONANCE OF SLOT 1. CURVES HAVE THE SAME SHAPE AS FIGS. 2–4

ϵ_1	w_i/L_i	d_o/L_1		
		0.02	0.1	0.2
1.0	0.02	-18.2	-23.1	-26.5
	0.04	-18.4	-22.9	-26.0
	0.08	-19.4	-23.3	-26.2
3.8	0.02	-18.1	-23.6	-28.4
	0.04	-17.9	-22.9	-27.1
	0.08	-18.5	-22.9	-26.6
12.8	0.02	-18.4	-24.7	-30.7
	0.04	-17.9	-23.5	-28.7
	0.08	-17.8	-22.4	-26.8

A few general observations can be made about Table I. The increased coupling when $x_o/L_1 > 0.1$ for increasing w_i/L_i is less significant than for equal length slots. In fact, there is a very slight decrease in coupling for $\epsilon_1 = 1.0$. For $w_i/L_i = 0.02$ the increase is only about 1 dB and for $w_i/L_i = 0.08$ there is an increase of about 1 dB between $\epsilon_1 = 1.0$ and $\epsilon_1 = 3.8$ and between $\epsilon_1 = 3.8$ and $\epsilon_1 = 12.8$. There is still a decrease of 2–6 dB for an factor of two increase in w_i/L_i for $x_o/L < 0.1$ with ϵ_1 held constant as in the equal slot length case.

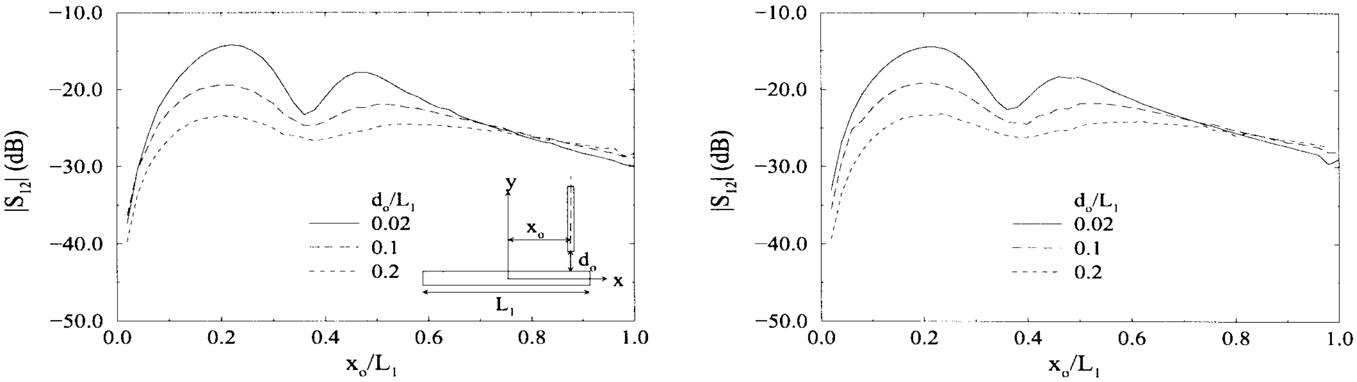
2) Source in Short Slot: Mutual coupling between two slots for which $L_2 = L_1/2$ and $w_2 = w_1/2$ at the second resonance slot 2 (the shorter slot) was also calculated (Figs. 6–8). As with the calculation at the second resonance of slot 1, the decrease in coupling for constant w_i/L_i and increasing ϵ_1 is less significant than for that of equal length slots. However, the shape of these calculated curves are quite different. In this case, the peak value of the coupling is about 3 dB higher and occurs at about half the value of x_o/L_1. This is significant because this value of $x_o/L_1 \approx 0.22$ is not far from a position of interest in subharmonic receivers/transmitters. This will be discussed as a design example in the next section.

Radiation patterns were calculated at the second resonance of slot 2. Since the source is in slot 2, we define the E plane of the short slot as the x-z plane for Fig. 9. Note that this is different than for Fig. 5. As for equal length slots, patterns were calculated for free-space and GaAs with $w/L = 0.04$ and $d_o/L_1 = 0.02$ at the first value of x_o/L for which the coupling is minimum and maximum excluding $x_o/L_1 = 0$. The maximum for both free-space and GaAs occurs at $x_o/L_1 = 0.22$. The minimum for free-space is $x_o/L_1 = 0.36$ and the minimum for GaAs is $x_o/L_1 = 0.34$.

Fig. 9 shows the radiation patterns with GaAs again on the dielectric side. There is a decrease of 5 dB in peak cross-polarization level for free-space and a decrease of 9 dB for GaAs. The D-plane cross-polarization levels are on the average less than or equal to those of the E plane at any given value of θ.

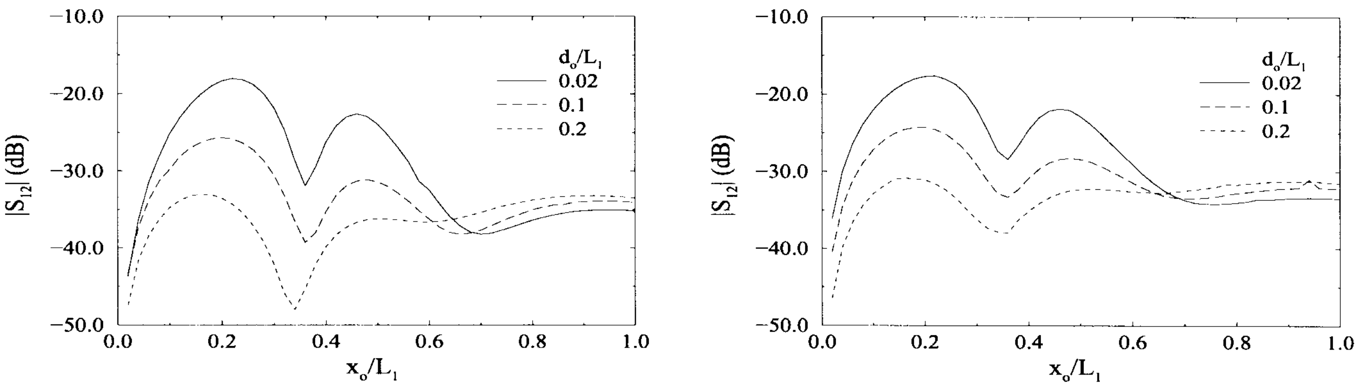
C. Design Example—Subharmonic Receiver

Double slots are often placed *parallel* on a substrate (Fig. 10) separated by a distance s_i such that the E-plane and H-plane patterns are approximately the same [2], [3]. For two equal length parallel slots with $w/L = 0.04$ on fused



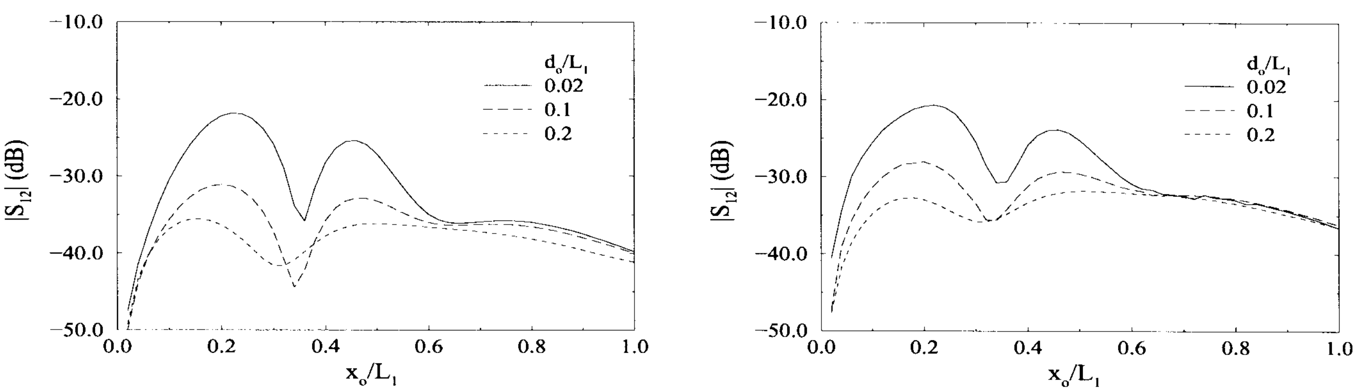
(a) (b)

Fig. 6. Coupling between slots for which $L_2 = L_1/2$ and $w_2 = w_1/2$ on free-space dielectric ($\epsilon_1 = 1.0$). λ_2 is the free-space wavelength at the second resonance of L_2. (a)$w_i/L_i = 0.02, L_2 = 0.840\lambda_2$. (b) $w_i/L_i = 0.04, L_2 = 0.794\lambda_2$.



(a) (b)

Fig. 7. Coupling between slots for which $L_2 = L_1/2$ and $w_2 = w_1/2$ on fused quartz ($\epsilon_1 = 3.8$). λ_2 is the free-space wavelength at the second resonance of L_2. (a) $w_i/L_i = 0.02, L_2 = 0.546\lambda_2$. (b) $w_i/L_i = 0.04, L_2 = 0.515\lambda_2$.



(a) (b)

Fig. 8. Coupling between slots for which $L_2 = L_1/2$ and $w_2 = w_1/2$, on GaAs ($\epsilon_1 = 12.8$). λ_2 is the free-space wavelength at the second resonance of L_2. (a) $w_i/L_i = 0.02, L_2 = 0.325\lambda_2$. (b) $w_i/L_i = 0.04, L_2 = 0.305\lambda_2$.

quartz to have equal 10-dB beamwidths, this separation is $s_1/L_1 = 0.54$. With this separation, the 10-dB beamwidths are 95.0°. Implementation of the mutual coupling between two parallel slots was accomplished in a similar manner to [1] and radiation patterns were computed from the slot electric fields with both slots excited with equal phase at their centers.

In a subharmonic receiver, it would be of interest to place two perpendicular slots of length $0.5L_1$ between the two slots that are separated by $0.54L_1$, as shown in Fig. 10. In this configuration, the smaller slots receive a signal at twice the frequency and at a different polarization as that of the long slots. There is just enough room for them so that $d_o/L_1 = 0$. The shorter $0.5L_1$ slots would be separated by $s_2 = 0.27L_1$ so that their E-plane and H-plane 10-dB beamwidths at their second resonance are equal at 95.0° also. Therefore, x_o/L_1 for any two perpendicular slots would be $x_o/L_1 = 0.135$.

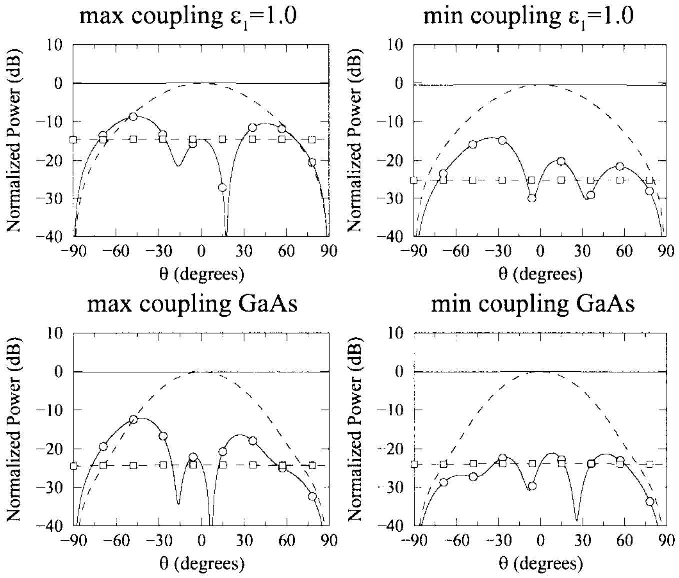


Fig. 9. Normalized radiation patterns with source in slot 2 at its second resonance for $L_2 = L_1/2, w_i/L_i = 0.04, d_o/L = 0.02$. E plane is x-z plane. ——— E-plane copol, ○——— ○ E-plane cross-pol, - - - - H-plane copol, □- - - -□ H-plane cross-pol.

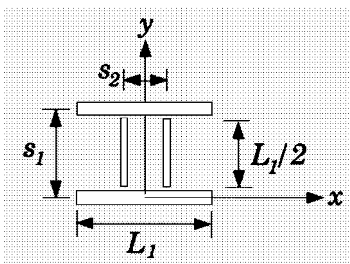


Fig. 10. Two sets of parallel dual slots in a subharmonic receiver design such that $s_i/L_i = 0.54$ so that 10-dB E-plane and H-plane beamwidths for each parallel set are equal on quartz for $w_i/L_i = 0.04$.

According to Fig. 7(b), the coupling will be approximately -20 dB between any two perpendicular slots at the second resonance of the short slots. The coupling at the second resonance of the long slots is smaller at about -30 dB as can be seen from Table I and Fig. 3(b). In this application at normal incidence, there would not be any total coupling since the two sets of parallel slots would induce opposite currents in their perpendicular partners. But for anything but normal incidence in Fig. 10 the coupling may have to be taken into account.

To study cross polarization of the antenna in Fig. 10, radiation patterns were calculated at the second resonance of the long slots with both of them excited in phase [Fig. 11(a)] and the second resonance of the short slots with both of them excited in phase [Fig. 11(b)]. When the source is in the long slots, the E-plane is the y-z plane; when the source is in the short slots, the E-plane is the x-z plane. The antenna was configured as previously described such that the E-plane and H-plane 10-dB beamwidths are equal in the quartz. There is no cross polarization in the E and H planes since the currents induced in a perpendicular slot by two parallel sources excited in phase and symmetrically placed on either side of

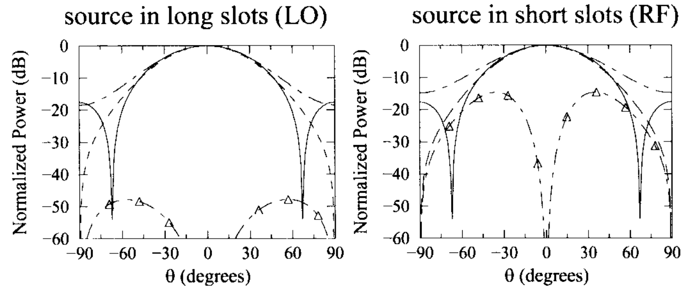


Fig. 11. Normalized radiation patterns for the four slot subharmonic receiver design. $L2 = L1/2, w_i/L_i = 0.04, d_o/L_1 = 0, x_o/L_1 = 0.27$. (a) E plane is x-z plane with the short slots excited at their second resonance. (b) E plane is y-z plane with the long slots excited at their second resonance and ——— E-plane copol, - - - -H-plane copol, - — - — -D-plane copol, △- - - -△ D-plane cross pol.

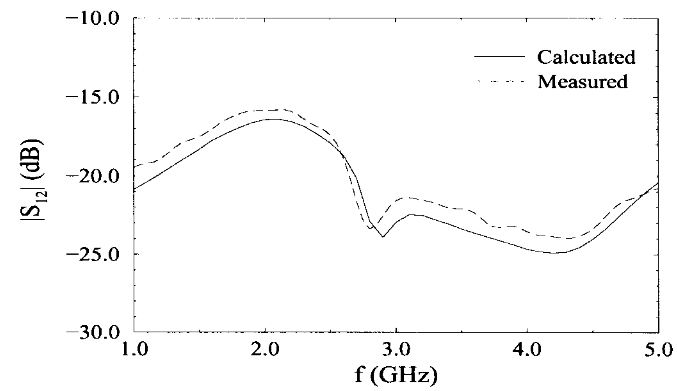


Fig. 12. Measured and calculated coupling levels versus frequency: $L_2 = L_1 = 100$ mm, $w_2 = w_1 = 4$ mm, $x_o/L = 0.48$, $d_o = 2$ mm, $\epsilon_1 = 1.0$.

the perpendicular slot, are in opposite directions. Of particular note is the right-hand plot, which shows that the D-plane peak cross polarization is quite high (-14 dB) even though the E- and H-plane cross polarizations are zero. This can be reduced by bringing the smaller slots closer together so that x_o/L_1 is smaller and the coupling is smaller [Fig. 7(b)]. However, bringing the two slots closer together to reduce the D-plane cross-polarization level will increase the E-plane beamwidth of the double slots. In effect, there is a tradeoff between polarization, purity, and equivalence of E- and H-plane 10-dB beamwidths for the pair of antennas.

D. Experimental Results

Experimental verification of the calculated coupling between the two slots was accomplished using an HP-8722C network analyzer. Fig. 12 shows a frequency scan for $L = 100$ mm, $w/L = 0.04, d_o/L = 0.02, x_o/L = 0.48$ and $\epsilon_0 = \epsilon_1 = 1.0$. The experimental antenna was constructed on a 0.25-mm woven glass dielectric with the woven glass removed from the slot region, thus approximating the case of free-space on both sides of the slots. The ground plane extended at least 40 cm from all edges of both slots, thus assuring that it was in the far field at the second resonance. At the second resonance of a single slot (2.38 GHz), the difference between experiment and theory is only 0.1 dB. Fig. 13 shows theory and experiment

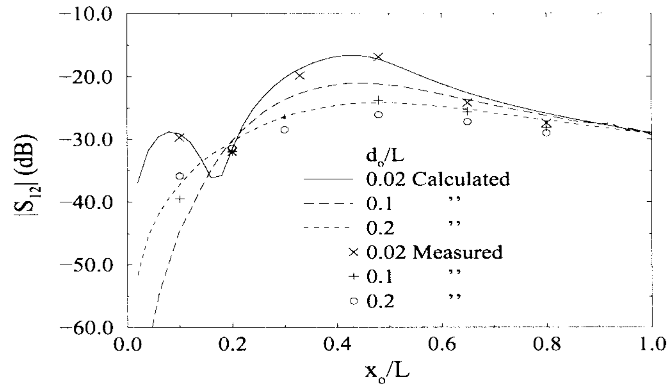


Fig. 13. Comparison of calculated and experimental coupling versus x_o and d_o at the second resonance: $L_2 = L_1 = L$, $w_2 = w_1 = w$, $\epsilon_1 = 1.0$, $w/L = 0.04$, $L = 50$ mm, $L = 0.794\lambda_2$.

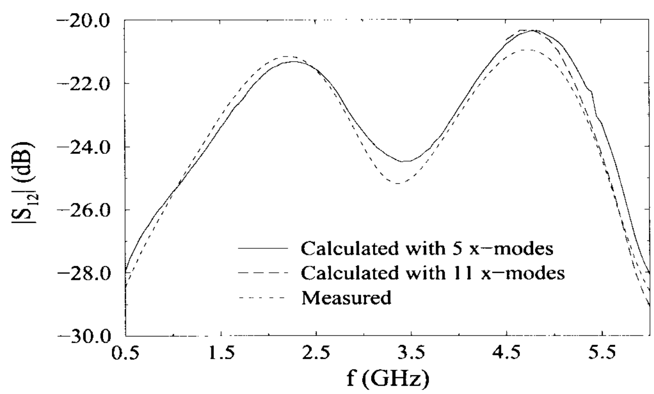


Fig. 14. Measured and calculated frequency scans: $L_2 = L_1/2$, $w_2 = w_1/2$, $\epsilon_1 = 1.0$, $w_i/L_i = 0.04$, $L_1 = 100$ mm, $x_o/L_1 = 0.48$, $d_o/L_1 = 0.06$.

for a scan of x_o for the same antenna. Again, the agreement is very good. Note that Fig. 13 has experimental data mapped onto Fig. 2(b).

Fig. 14 shows a frequency scan for the case $L_2 = L_1/2, w_2 = w_1/2, \epsilon_1 = 1.0, w_i/L_i = 0.04, L_1 = 100$ mm, $x_o/L_1 = 0.48, d_o/L_1 = 0.06$. It is noted that at higher frequencies, better convergence is obtained with more x modes (basis functions) in slot 1. It was not necessary to use any more than five x modes anywhere else in this paper to obtain convergence. All calculations were done with five y-modes (basis functions in slot 2).

Also included is Fig. 15, which shows the change in coupling for excitations of the slot away from the second resonance. Here we note that while we excited the antennas in Figs. 2–4 at the second resonance of a single slot, the deviation from resonance with the addition of a perpendicular slot was a maximum of about 1% of the second resonance. This maximum change occurs at the position of maximum coupling, where $x_o/L \approx 0.43$. However, the change in Fig. 15 is 10% of the second resonance. It is, therefore, concluded that if we were to have made the calculations at the second resonance of a slot in the presence of a second perpendicular slot that the difference in the curves would be only a small fraction of 1 dB. This includes the position where the coupling is minimum $(x_o/L = 0.15)$ since the change in the second resonance with

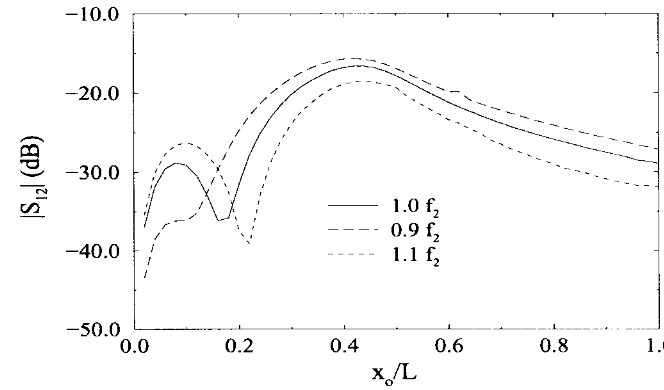


Fig. 15. Calculations showing the coupling away from the second resonance f_2: $L_2 = L_1 = L$, $w_2 = w_1 = w$, $\epsilon_1 = 1.0$, $w/L = 0.04$, $L = 0.794\lambda_2$.

the addition of a perpendicular slot is very small at that point.

IV. Conclusions

This paper has analyzed coupling between and radiation patterns of arbitrarily positioned perpendicular slots and found a strong correlation between coupling levels and calculated peak cross-polarization levels. It has been shown that there is not a significant increase in coupling when the slots are placed on an infinite dielectric; in some cases, coupling is reduced. As expected from geometrical symmetry, there is no coupling between perpendicular slots when $X_o/L = 0$. We have demonstrated that a very low value of coupling can be attained for $x_o/L \approx 0.2$ when $L_2 = L_1$ and $x_o/L \approx 0.4$ for $L_2 = L_1/2$. It is best to avoid $x_o \approx L_1/2$ for equal length slots when cross-polarization levels are important. We have also shown that increasing the y-separation between the slots (d_o/L) can have the effect of increasing the coupling at some values of x_o/L for equal length slots. This paper has also demonstrated that in some cases, an increase in slot width can decrease coupling between perpendicular slots. The moment-method technique used herein has been shown to be very accurate by experimental verification.

References

[1] G. V. Eleftheriades and G. M. Rebeiz, “Self and mutual admittance of slot antennas on a dielectric half-space,” *Int. J. Infrared Millimeter Waves*, vol. 14, no. 10, p. 1925, 1993.

[2] S. S. Gearhart and G. M. Rebeiz, “A monolithic 250 GHz schottky-diode reciever,” *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 2504–2511, Dec. 1994.

[3] D. F. Filipovic, S. S. Gearhart, and G. M. Rebeiz, “Double-slot antennas on extended hemispherical and elliptical silicon dielectric lenses,” *IEEE Trans. Antennas Propagat.*, vol. 41, pp. 1738–1749, Oct. 1993.

[4] T. M. Weller and L. P. Katehi, “Single and double folded-slot antennas on semi-infinte substrates,” *IEEE Trans. Antennas Propagat.*, vol. 43, pp. 1423–1428, Dec. 1995.

[5] C. M. Butler and K. R. Umashankar, “Electromagnetic penetration through an aperture in an infinite, planar screen separating two half spaces of different electromagnetic properties,” *Radio Sci.*, vol. 11, no. 7, pp. 611–619, July 1976.

[6] C. Scott, *The Spectral Domain Method in Electromagnetics.* Norwood, MA: Artech House, 1989.

[7] T. Itoh, Ed., *Numerical Techniques for Microwave and Millimeter-Wave Passive Structures.* New York: Wiley, 1988.

[8] R. Piessens, *Quadpack: A Subroutine Package for Automatic Integration.* New York: Springer-Verlag, 1983.

[9] M. Kominami, D. Pozar, and D. Schaubert, “Dipole and slot elements and arrays on semi-infinite substrates,” *IEEE Trans. Antennas Propagat.*, vol. 33, p. 600, June 1985.
[10] J. Zmuidzinas and H. G. LeDuc, “Quasi-optical slot antenna SIS mixers,” *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1797–1804, Sept. 1992.
[11] R. E. Collin, *Foundations for Microwave Engineering*. New York: McGraw-Hill, 1992.
[12] A. C. Ludwig, “The definition of cross polarization,” *IEEE Trans. Antennas Propagat.*, vol. AP-21, pp. 116–119, Jan. 1973.

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