THE BELL SYSTEM TECHNICAL JOURNAL

DEVOTED TO THE SCIENTIFIC AND ENGINEERING
ASPECTS OF ELECTRICAL COMMUNICATION

Volume 54 | December 1975 | Number 10

Quasi-Optical Polarization Diplexing of Microwaves

By T. S. CHU, M. J. GANS, and W. E. LEGG

(Manuscript received July 17, 1975)

The feasibility of a microwave quasi-optical polarization diplexer has been demonstrated using photo-etched copper strips with thin mylar backing as a practical example. The insertion losses of the principal polarization and the cross polarization have been calculated and measured. The conducting strips must be aligned in a preferred direction, namely, perpendicular to the plane of incidence, to minimize cross-polarized radiation, whereas orientation of the plane of the grid with respect to the beam direction is not restricted. The measured cross-polarized radiation agrees with predictions from simple theoretical models of a magnetic current sheet for the transmission mode and an electric current sheet for the reflection mode. This type of diplexer has been successfully employed in studying the polarization properties of the 20-GHz signal from the ATS-6 satellite.

I. INTRODUCTION

To achieve frequency reuse by employing orthogonal polarizations in a radio communication system, it is essential to avoid cross polarization in the feed patterns that illuminate the antennas; this relies upon the diplexing of two orthogonal polarizations with high isolation. Waveguide-type polarization couplers perform diplexing well where a carrier with an associated bandwidth of about 10 percent is involved. However, it is difficult to provide a low-loss ($\lesssim$0.1 dB) waveguide diplexer to separate effectively and simultaneously the two polarizations in each of two widely separated common carrier bands, such as 18 and 30 GHz. The difficulty stems from the vulnerability of an oversized waveguide to higher-order modes in the higher-frequency band. Contamination by only 1 percent of the power in higher-order modes may cause unacceptable cross-polarized radiation in the feed.

The necessity for overcoming this problem has led to the suggestion of using a closely spaced wire grid as a quasi-optical polarization diplexer. The purpose of this paper is to describe feasibility studies of this quasi-optical approach. Not only should the insertion loss be small and the feed pattern distortion slight in the principal polarization, but minimization of the cross-polarized radiation should also be achieved. Our measurements have shown that the wires must be oriented in a *preferred direction* to minimize the cross-polarized radiation. This property is explained theoretically by utilizing a magnetic current sheet for transmission through the grid and an electric current sheet for reflection from the grid. This preferred direction requires that the wires be perpendicular to the plane of incidence determined by the beam axis and the grid normal, as shown in Fig. 1b. One notes that the classical application of a polarizer, which consists of a wire grid parallel to an aperture plane, always satisfies this condition.

In Section II we calculate the insertion loss and cross-polarized radiation using simple theoretical models. Section III describes the measurements and the comparison between the calculated and measured cross-polarized radiation of wire grids in various configurations. Section IV discusses applications and includes concluding remarks.

To avoid confusion about the definition of cross polarization,[1] the following two explicit expressions

$$\hat{p}_1 = \hat{\theta} \cos \phi - \hat{\phi} \sin \phi \tag{1}$$

$$\hat{p}_2 = \hat{\theta} \sin \phi + \hat{\phi} \cos \phi \tag{2}$$

are defined as the two orthogonal polarization vectors which are the cross polarization of each other. The carat "^" indicates unit vector, and (θ, ϕ) are spherical coordinates as shown in Fig. 1a; $\hat{p}_1$ and $\hat{p}_2$ within the immediate vicinity of the Z axis are in the nominal X and Y directions, respectively. The usual antenna pattern measurements yield directly the patterns for the two polarization components under this definition. If a feed pattern with the polarization vector (1) or (2) (often called a balanced-feed radiation[2]) illuminates a paraboloid with its axis oriented in the Z direction, the reflected field in the aperture of the paraboloid is free of cross polarization.

II. THEORETICAL CALCULATIONS

2.1 Insertion losses

Implementation of quasi-optical polarization diplexers requires prediction of the insertion losses for both the principal and the unwanted

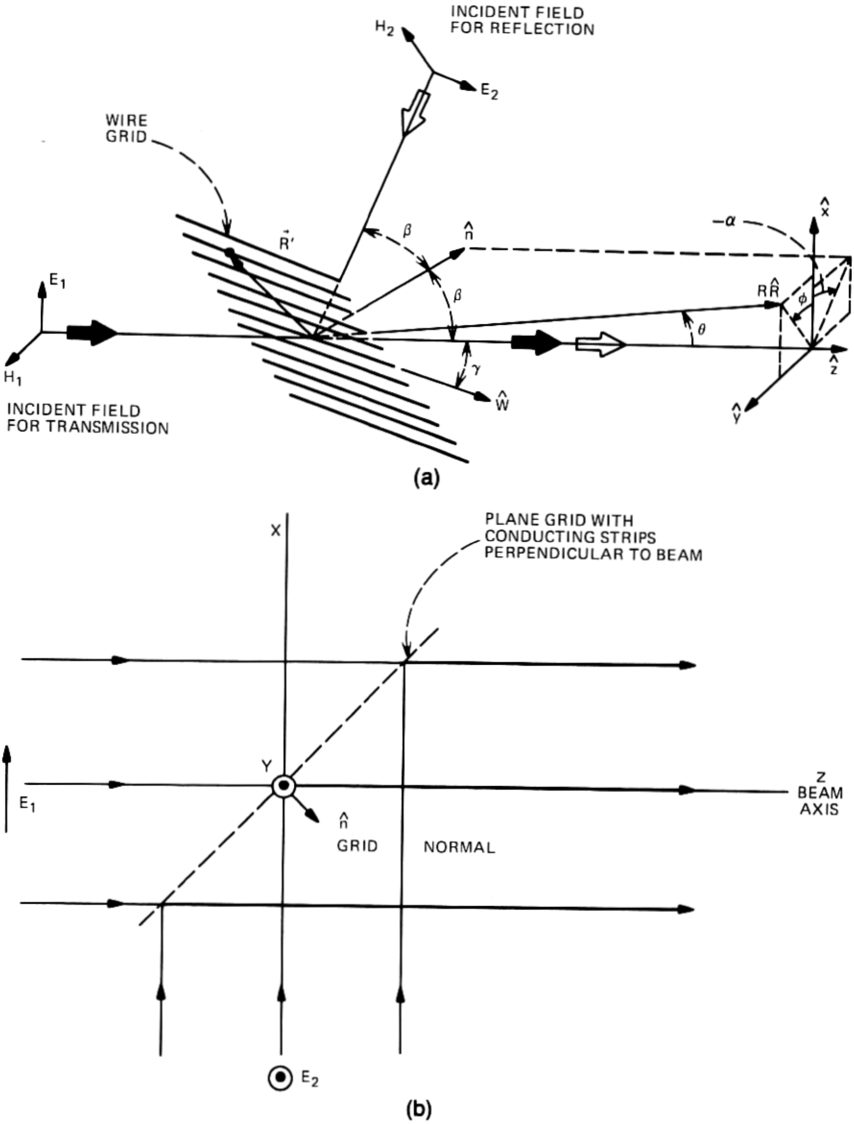

Fig. 1—(a) Geometry of a quasi-optical polarization diplexer. (b) Preferred configuration of quasi-optical polarization diplexing.

cross polarization. If the polarization grid is made of uniformly spaced copper strips with a thin mylar backing, the calculation is facilitated by the equivalence between the periodic grating and the capacitive diaphragm in a parallel-plate waveguide.[3] Neglecting the effect of the thin mylar layer, the power reflection coefficient for an incident wave polarized perpendicular to the strips is approximately given by[3,4]

$$R_{\perp} = \frac{B^2 \cos^2 \theta}{4 + B^2 \cos^2 \theta}, \tag{3}$$

where

$$B = \frac{4b}{\lambda} \ln \sec \left(\frac{\pi d}{2b} \right)$$

is the shunt susceptance, b the grating period, d the width of the conducting strip, λ the wavelength, and θ the angle of incidence between the direction of propagation and the normal to the grating plane. Equation (3)* is based upon the low-frequency approximation ($b \ll \lambda$) for a grating of infinitely thin perfectly conducting strips. Using Babinet's principle, the power transmission coefficient "$T_{\parallel}$" for an incident wave, polarized parallel to the plane determined by the strip and the propagation direction, is also approximately given by eq. (3) provided the strip width d in the expression for B is replaced by the spacing $(b - d)$.

If we employ the numerical value, $b = 0.5$ mm, $d = 0.2$ mm, $\lambda = 1.05$ cm, and $\theta = 45°$, as used in the experiment† later, substitution into eq. (3) yields the insertion loss of the cross polarization

$$-10 \log_{10} R_{\perp} = 36.8 \text{ dB}$$
$$-10 \log_{10} T_{\parallel} = 28.8 \text{ dB},$$

which is equivalent to an insertion loss of only 0.001 dB and 0.006 dB for the transmitting and reflecting principal polarizations, respectively. Typically, the cross polarization in a Cassegrain feed aperture that illuminates a polarization diplexer is of the order of −20 dB or less; thus, the improvement provided by use of a quasi-optical polarization diplexer reduces the residual cross-polarized components in the grid aperture to negligible values.

However, the residual cross polarization discussed above is only part of the possible cross-polarized radiation; the following calculations show that the co-polarized field in the grid aperture may also give rise to off-axis cross-polarized radiation if the strips are not oriented in a preferred direction.

* Equation (3) is known to be accurate when the strips are parallel to the plane of incidence for all strip widths. However, to the authors' knowledge, the rigorous demonstration, in the literature, of its accuracy for other strip directions is restricted to the cases $(b - d)/b \ll 1$, or $d/b \ll 1$.

† Equation (3), for the idealized grid, shows that $d = b/2$ provides the minimum value of the quantity, max $[R_{\perp}, T_{\parallel}]$. Therefore, the grid was designed to have both copper strip width and gap spacing equal to 0.25 mm. However, this specification was close to the resolution limit of the fabrication process in use at the time, which resulted in a grid with the above measured dimensions.

2.2 Magnetic current sheet

For the case of transmission through an arbitrarily oriented wire grid, a magnetic-current-sheet model is used to calculate the radiation.

If a wire grid is placed in front of a radiating aperture that produces a wave collimated in the Z direction, the on-axis radiation is linearly polarized perpendicular to the conducting direction of the wire grid. We shall find the polarization properties of the off-axis radiation.

Let the plane of the wire grid be oriented in an arbitrary direction, as shown in Fig. 1a, with the following unit normal

$$\hat{n} = \hat{x} \sin \beta \cos \alpha + \hat{y} \sin \beta \sin \alpha + \hat{z} \cos \beta. \tag{4}$$

In order that an incident electric field, $\mathbf{E}_1 = E_1\hat{x}$, may pass freely through the wire grid, the direction of the conducting wires

$$\hat{w} = \sin \gamma\, \hat{y} + \cos \gamma\, \hat{z} \tag{5}$$

must be the same as that of the equivalent magnetic current density $2\hat{n} \times \mathbf{E}_1$ (see the appendix):

$$\hat{n} \times \mathbf{E}_1 = E_1\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha} \left[\frac{\cos \beta \hat{y}}{\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha}} - \frac{\sin \beta \sin \alpha \hat{z}}{\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha}} \right], \tag{6}$$

i.e.,

$$\sin \gamma = \frac{\cos \beta}{\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha}}$$

and

$$\cos \gamma = \frac{-\sin \beta \sin \alpha}{\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha}}.$$

The far-zone electric-field radiation of a magnetic current sheet can be written

$$\mathbf{E}_1 = -\frac{jk}{2\pi R} e^{-jkR} \int [\hat{R} \times (\hat{n} \times \mathbf{E}_1)] e^{jk\mathbf{R}' \cdot \hat{R}} dA, \tag{7}$$

where k is the free-space phase constant, and R and $\hat{R}$ are the distance to the far-field point and the corresponding unit direction vector. The points on the magnetic current sheet are defined by $\mathbf{R}'$. The polarization is determined by the bracketed vector product in eq. (7),

$$\mathbf{P}_1 = \hat{R} \times (\hat{n} \times \mathbf{E}_1) = E_1\sqrt{\cos^2 \beta + \sin^2 \beta \sin^2 \alpha}\, [\hat{\theta}(-\cos \phi \sin \gamma) + \hat{\phi}(-\sin \theta \cos \gamma + \cos \theta \sin \phi \sin \gamma)]. \tag{8}$$

The dot product of eqs. (1) and (8) gives the principal polarization component

$$\mathbf{P}_1 \cdot \hat{p}_1 = -E_1 \cos \beta [1 - \sin^2 \phi (1 - \cos \theta) - \sin \theta \sin \phi \cot \gamma]. \quad (9)$$

If we substitute the above product for the bracket in eq. (7), it is seen that the on-axis radiation is the same as that without the wire grid, while the off-axis radiation is only slightly perturbed. Here the factor $\cos \beta$ accounts for the larger slanted area of the grid.

Now the dot product of eqs. (2) and (8) gives the cross-polarization component

$$\mathbf{P}_1 \cdot \hat{p}_2 = -E_1 \cos \beta [\sin \phi \cos \phi (1 - \cos \theta) + \sin \theta \cos \phi \cot \gamma]. \quad (10)$$

The cross polarization on axis vanishes, as expected. When the direction of the conducting wire is perpendicular to the beam axis, i.e., $\gamma = 90°$, only second-order cross polarization, as represented by the first term in eq. (10), is present with maxima in the $\phi = 45°$ planes. This second-order cross polarization is negligibly small for narrow feed patterns. However, this residue can become a considerable item for broad feed patterns, as will be demonstrated by an experiment described later.

When the direction of the conducting wire is not perpendicular to the beam axis, i.e., $\gamma \neq 90°$, the second term in eq. (10) represents first-order cross-polarization lobes with maxima in the $\phi = 0$ plane. This term is kept small if both θ and $(90° - \gamma)$ are small. Therefore, fine adjustment to reduce the residual cross polarization can be accomplished by rotation of the grid in its own plane in the case of a narrow feed pattern.

2.3 *Electric current sheet*

Next, consider the case of reflection from the wire grid. The reflected field is entirely due to the fields radiated by electric currents flowing in the wires. If the grid is fine enough, these currents will flow only in the direction of the wires. Thus, we may use an electric current sheet model to compute the field reflected from the wire grid. To obtain perfect reflection from the grid, the direction of the conducting wires must be the same as the induced electric current direction $\hat{n} \times \mathbf{H}_2$, where $\mathbf{H}_2$ is the incident magnetic field for this case. The far-zone electric field of an electric current sheet can be written

$$\mathbf{E}_2 = \frac{jkZ_0}{2\pi R} e^{-jkR} \int \{\hat{R} \times [\hat{R} \times (\hat{n} \times \mathbf{H}_2)]\} e^{jk\mathbf{R}' \cdot \hat{R}} dA, \quad (11)$$

where Z_0 is the free-space impedance. The polarization is determined by the vector product inside the bracket in eq. (11):

$$\mathbf{P}_2 = \hat{R} \times \{\hat{R} \times (\hat{n} \times \mathbf{H}_2)\} = |\hat{n} \times \mathbf{H}_2|[\hat{\theta}(\sin\theta\cos\gamma - \cos\theta\sin\phi\sin\gamma) + \hat{\phi}(-\cos\phi\sin\gamma)]. \quad (12)$$

The above polarization is orthogonal to that of eq. (8). The principal and cross-polarization components are obtained by the dot products of eq. (12) with eqs. (2) and (1), respectively.

$$\mathbf{P}_2 \cdot \hat{p}_2 = -H_2\cos\beta[1 - \sin^2\phi(1 - \cos\theta) - \sin\theta\sin\phi\cot\gamma] \quad (13)$$

$$\mathbf{P}_2 \cdot \hat{p}_1 = H_2\cos\beta[\sin\phi\cos\phi(1 - \cos\theta) + \sin\theta\cos\phi\cot\gamma]. \quad (14)$$

Here the relation $|\hat{n} \times \mathbf{H}_2|\sin\gamma = H_2\cos\beta$ follows the symmetry with respect to $\hat{n}$ between the incident wave and reflected wave of which the magnetic field is equal to $H_2\hat{x}$.

Since eqs. (13) and (14) are identical to eqs. (9) and (10) except for a proportionality constant, the properties of the off-axis cross-polarized radiation described in the previous case are also valid for this orthogonal case in reflection.

III. EXPERIMENT

3.1 *Insertion loss*

The insertion loss of the combinations of a dual-mode horn and a wire grid were measured at 28.5 GHz and 19 GHz in both transmission and reflection. Figure 2 shows the sketch of the 28.5-GHz experimental model. The dual-mode horn has been described elsewhere.[5] The wire grid was made by photo-etching a copper-covered mylar sheet; copper

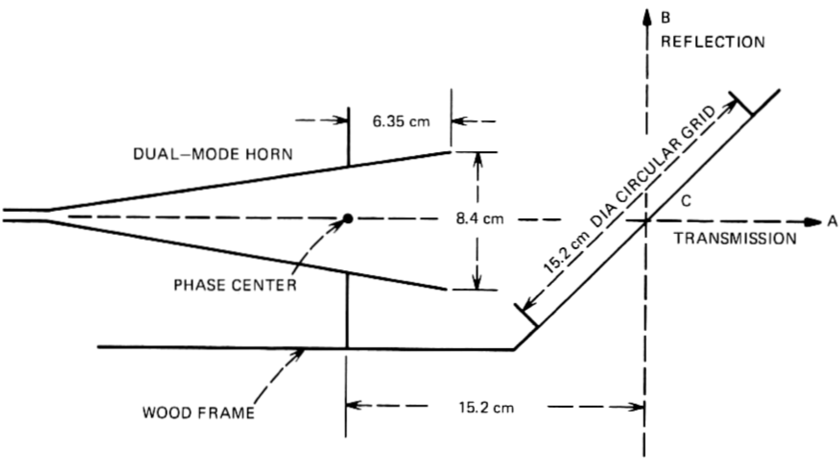

Fig. 2—Schematic of experimental assembly for 28.5 GHz.

strips 0.2 mm wide and 0.018 mm thick are spaced 0.3 mm apart on a mylar sheet 0.013 mm thick. The grid, mounted in a circular wooden rim, can be rotated in its own plane, which is oriented at 45° with respect to the horn aperture. The 19-GHz experiment uses a scaled dual-mode horn and 24.7-cm-diameter circular grid with the same copper strips and the same mylar sheet described earlier.

The measured minimum insertion loss on axis, for both transmission and reflection, was found to be only about 0.1 dB for the principal polarization; the discrepancy with the calculated 0.001 and 0.006 dB can be explained by measuring error and slight pattern distortion due to diffraction around the grid.

The maximum insertion loss of the cross-polarized field, on axis, at 19 and 28.5 GHz for both transmission and reflection is shown in Table I. The symbols $\|$ and $\perp$ indicate that the grid wires are parallel and perpendicular, respectively, to the plane of incidence.

The measured data are only in qualitative agreement with the approximate prediction from eq. (3). However, the effect of the mylar sheet (0.013 mm thick with a dielectric constant of 3), imperfect polarization of the horn radiation, and diffraction around the grid have been neglected in the approximate calculation. It was observed that the measured insertion loss of the cross polarization depends somewhat upon the spacing between the horn and the grid.

3.2 Radiation patterns at 28.5 GHz

The measured cross polarization in the radiation patterns is found to be negligible if the conducting strips are aligned in the preferred direction normal to the beam. But for the conducting strips in non-preferred directions, such as those parallel to the plane of incidence, maximum cross-polarized radiation is obtained in the transverse planes—AC for transmission and BC for reflection—both perpendicular to the plane of Fig. 2. To illustrate the predictions of the theoretical models in the preceding section, we present the measured

Table I — Measured insertion loss of cross-polarized fields

Grid-Wire Position	28.5 GHz		19 GHz	
	Transmission	Reflection	Transmission	Reflection
$\|$*	24 dB	25	32.5	30
$\perp$†	28 dB	30	38	34
Eq. (3)	28.8	36.8	32.3	40.3

* Conducting strips are parallel to the plane of incidence.
† Conducting strips are perpendicular to the plane of incidence.

transverse plane patterns at 28.5 GHz for four combinations of horn polarization and conducting strip directions.

The transverse plane patterns in Figs. 3 and 4 were measured with the radiation transmitted through the grid. In Fig. 3 the horn polarization is perpendicular to, and the conducting strips parallel to, the plane of Fig. 2. The average of the cross-polarization lobe maxima is about 20 dB below that of the principal polarization in the same direction ($\theta = 6°$), and agrees well with the prediction of eq. (10) [relative to eq. (9) with $\gamma = 45°$] as shown by the dotted curves. In Fig. 4,

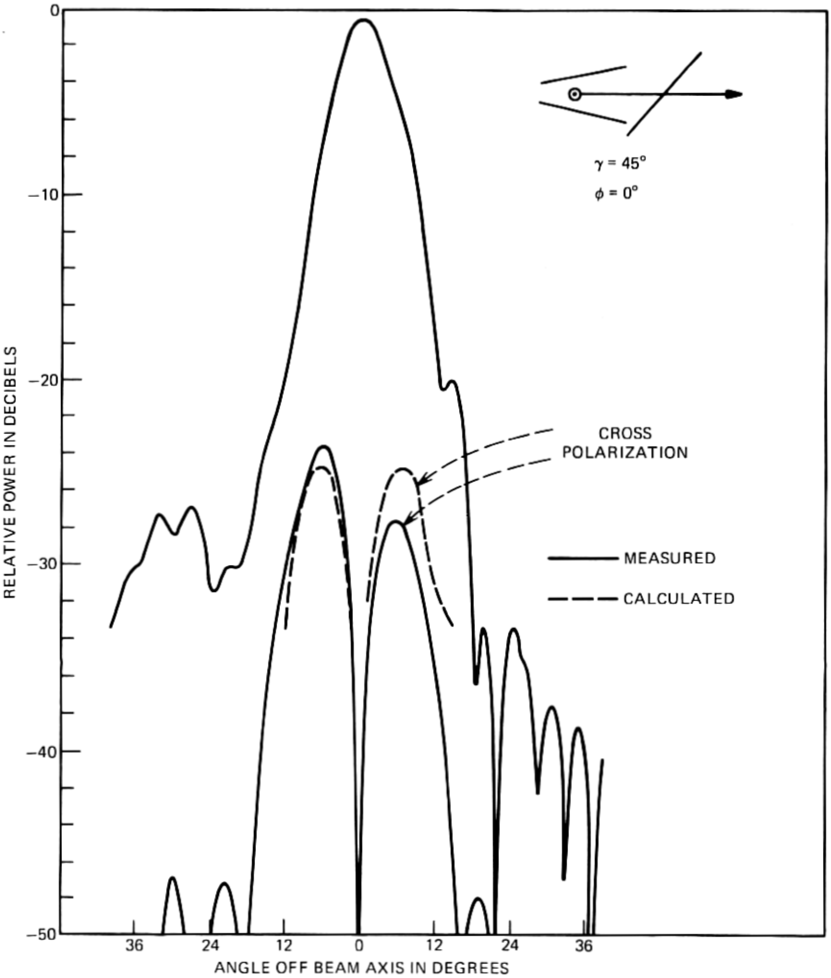

Fig. 3—Radiation patterns of a transmitting grid at 28.5 GHz with conducting strips parallel to the plane of incidence.

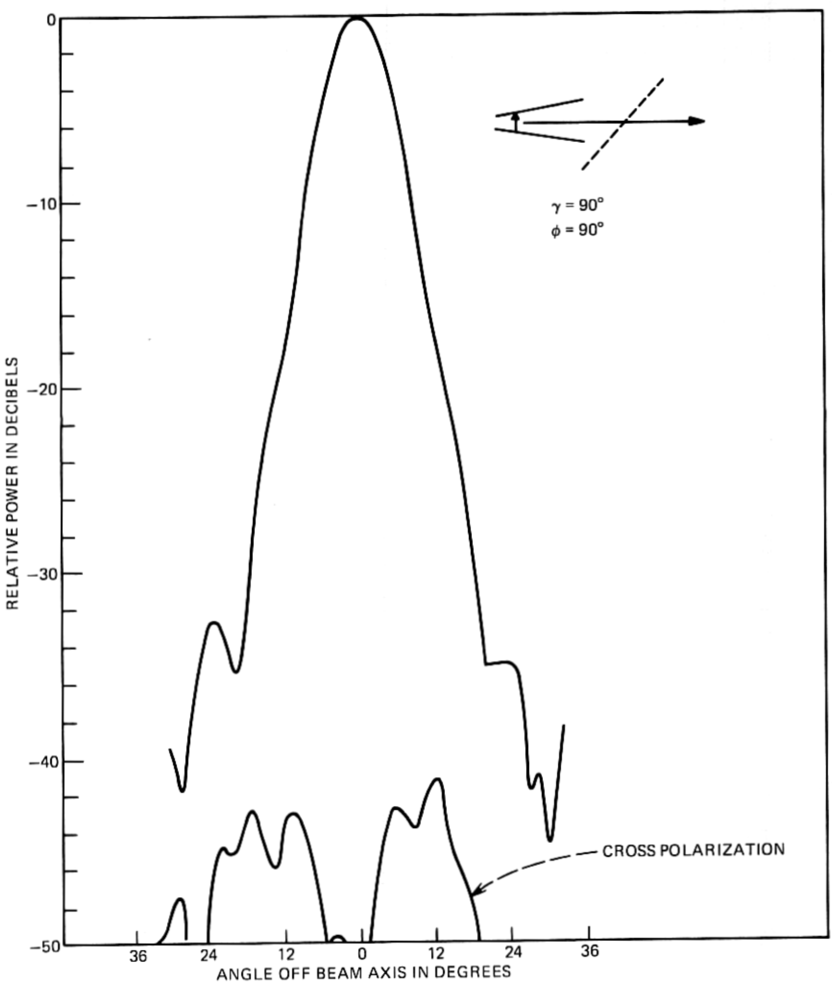

Fig. 4—Measured radiation patterns of a transmitting grid at 28.5 GHz with conducting strips perpendicular to the plane of incidence.

the horn polarization is parallel to, and the conducting strips perpendicular to, the plane of Fig. 2. The measured cross polarization of less than −40 dB essentially confirms the theoretical prediction of negligible cross polarization from eq. (10) ($\gamma = 90°$), since the measuring accuracy of the cross-polarization level is reliable down to about −40 dB.

The transverse plane patterns in Figs. 5 and 6 were measured with the radiation reflected from the grid. In Fig. 5 both the horn polarization and the conducting strips are parallel to the plane of Fig. 2,

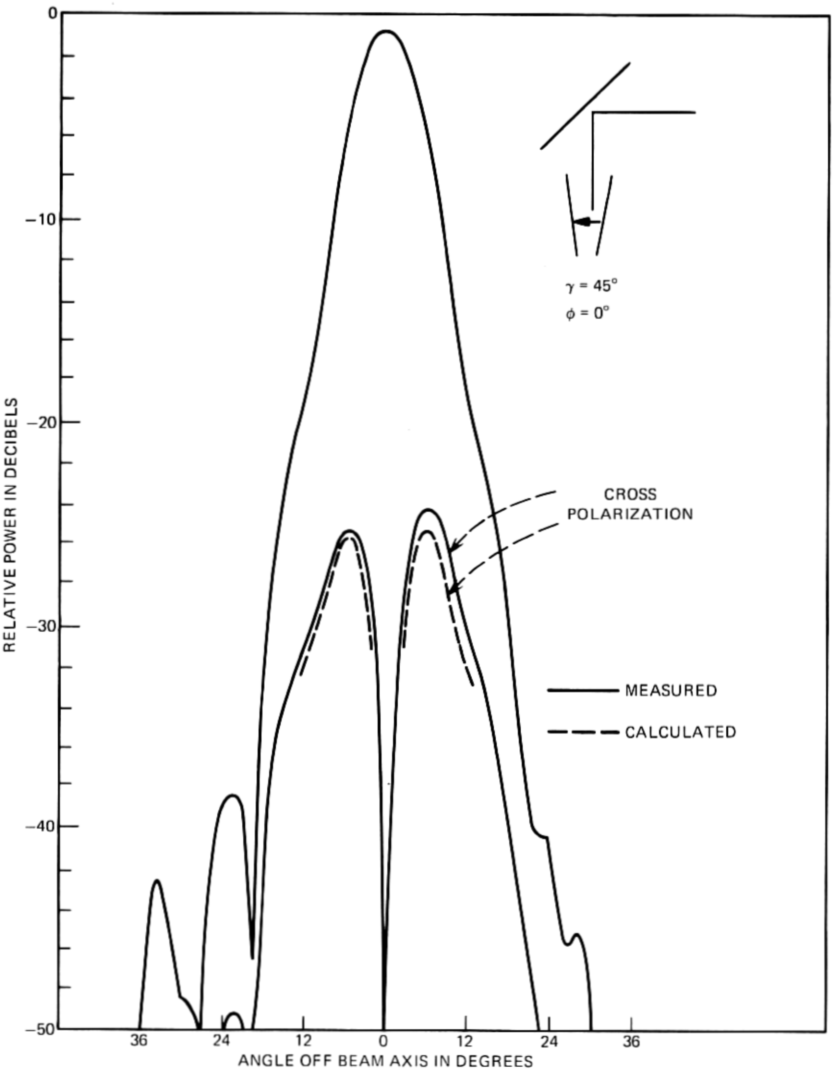

Fig. 5—Radiation patterns of a reflecting grid at 28.5 GHz with conducting strips parallel to the plane of incidence.

and the measured cross polarization is essentially the same as that of the transmitting case in Fig. 3. In Fig. 6, both the horn polarization and the conducting strips are perpendicular to the plane of Fig. 2, and the measured cross polarization of less than −40 dB is similar to that of Fig. 4. Thus, the results show that in employing quasi-optical polarization diplexers the off-axis cross-polarized radiation

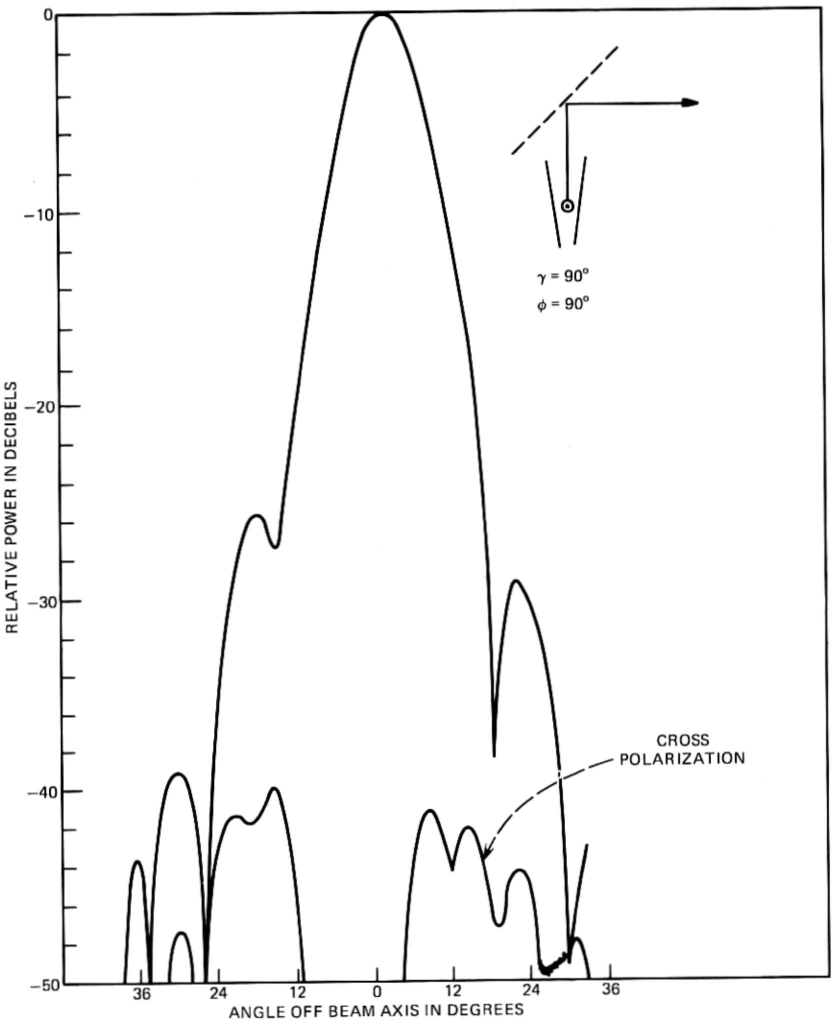

Fig. 6—Measured radiation patterns of a reflecting grid at 28.5 GHz with conducting strips perpendicular to the plane of incidence.

can be suppressed only if the conducting wires are perpendicular to the beam (i.e., perpendicular to the plane of incidence).

Owing to limitation of the measuring accuracy, it is difficult to measure the second-order cross polarization, the $(1 - \cos\theta)$ term in eqs. (10) and (14), of the wire grid for narrow feed patterns. Therefore, we conducted an experiment with a broad feed pattern to check this term, which grows rapidly when θ increases. The radiation patterns

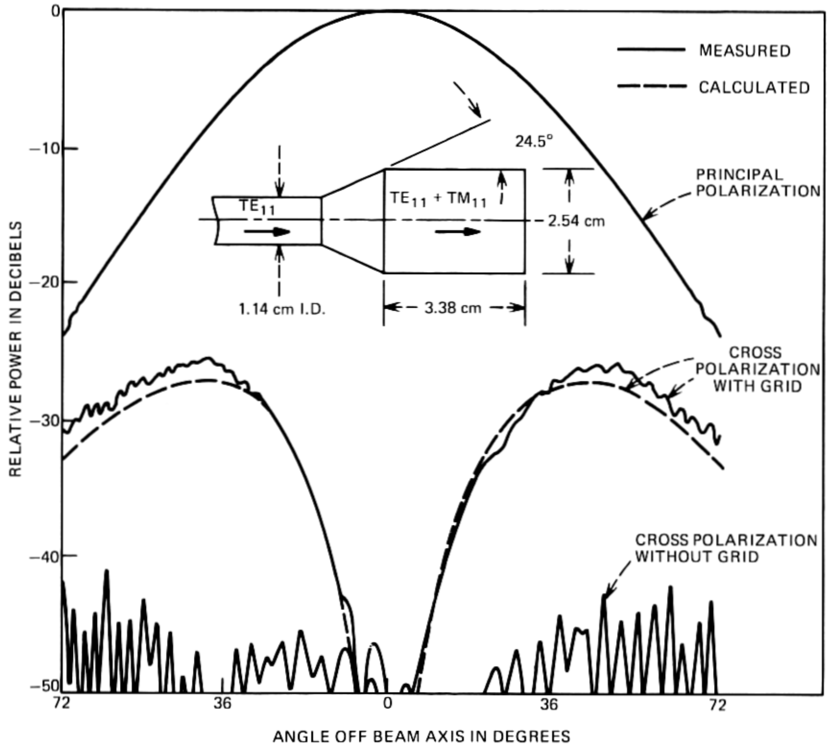

Fig. 7—Radiation patterns of a small dual-mode circular aperture feed ($D/\lambda = 1.39$ at 16.5 GHz).

of a small dual-mode horn[6] (see inset in Fig. 7) were measured with and without the wire grid.

In the absence of the grid, the measured 45° plane patterns of principal and cross polarization are as shown in Fig. 7. The measured patterns in other planes (not shown) exhibited circular symmetry in the co-polarized radiation pattern, and less than −40 dB in cross polarization everywhere. When the small dual-mode aperture was covered by a wire grid, the measured pattern in co-polarization remains essentially the same as without the grid; however, the cross polarization in the $\phi = 45°$ plane rises to −26 dB as shown in Fig. 7. The calculated cross polarization, which is plotted as a dotted curve, shows good agreement with the measured pattern. The cross-polarized radiation of a small grid-covered aperture is similar to that of a dipole.

Thus, the above results demonstrate that the wire grid is a good polarizer for large apertures, whereas improper use of the wire grid can even enhance the cross-polarized radiation of a small aperture.

IV. DISCUSSION

It has been observed[2] that there will be no cross polarization in the main reflector aperture of an offset near-field Cassegrainian antenna provided no cross polarization illuminates the subreflector. This ideal condition can be approximately realized by application of a quasi-optical polarization diplexer to the feed of an offset Cassegrainian antenna with large effective F/D ratio. The quasi-optical polarization diplexer can also be used on a symmetrical Cassegrainian antenna as demonstrated by its application in an earth-station receiver[7,8] for the 20-GHz ATS-6 signal.

The basic philosophy of the quasi-optical polarization diplexer can be simply stated as a cleaning up of the two orthogonal polarizations simultaneously just before illuminating the subreflector. This cleaning process is especially desirable if the feed is an offset reflector with relatively small F/D ratio. But the conducting wires must be oriented in a preferred direction, perpendicular to the plane of incidence, to avoid the off-axis cross-polarized radiation. For the broad feed pattern of a small dual-mode aperture, the second-order cross-polarized radiation from a classical polarizer may *exceed* that of the dual-mode aperture *without* the polarizer. The accuracy of the theoretical predictions demonstrates the utility of equivalent current sources for such analyses.

To avoid excessive spill-over loss and pattern distortion, the quasi-optical diplexer should be made conservatively large, typically with edge illumination less than -20 dB. The main disadvantages of quasi-optical feed systems appear to be bulkier volume and heavier weight compared with conventional waveguide diplexing feed systems, especially when both polarization and frequency diplexing are performed by quasi-optical components. But these components have an advantage in handling high power without difficulty. Equation (10) indicates that fine tuning ($\gamma \approx 90°$) of the residual polarization response may be accomplished by rotation of the grid in its own plane.

APPENDIX

To calculate the field transmitted through the wire grid, a magnetic current equivalent source is chosen, the choice being governed by the following considerations.

Given the tangential electric and/or magnetic field on the bounding surface of a source-free region, we may place equivalent sources on the bounding surface to correctly reproduce the original field in the source-free region:[9]

(*i*) Magnetic current: $\mathbf{K} = \mathbf{E}_t \times \hat{n}$, backed by a perfect electric conductor on the bounding surface.

(*ii*) Electric current: $\mathbf{J} = \hat{n} \times \mathbf{H}_t$, backed by a perfect magnetic conductor on the bounding surface.

(*iii*) Combination:

$$\left\{\begin{array}{l}\mathbf{K} = \mathbf{E}_t \times \hat{n}, \text{ magnetic current} \\ \mathbf{J} = \hat{n} \times \mathbf{H}_t, \text{ electric current}\end{array}\right\} \text{ operating in free space.}$$

All three equivalent sources give identical results if they are based on the true fields, $\mathbf{E}_t$ and $\mathbf{H}_t$. However, in the case of transmission through a grid, we do not know the true magnetic field $\mathbf{H}_t$ on the source-free side of the grid. If the polarizer is fine enough, one can be sure, though, that the tangential electric field is perpendicular to the wires. Thus, the field transmitted through the grid can be predicted most accurately by the magnetic current equivalent source backed by an electric conducting plane on the grid. Since a tangential magnetic current imaged in an electric conducting plane is equal to itself, we may include the effect of the electric conducting plane by using twice the magnetic current, 2**K**, operating in free space.

The far field radiated by the magnetic current density, 2**K**, in free space is

$$\mathbf{E}_1 = -\frac{jk}{4\pi R} e^{-jkR} \int_{\substack{\text{grid}\\ \text{gaps}}} [(2\mathbf{K}) \times \hat{R}] e^{jk\mathbf{R}'\cdot\hat{R}} dA. \tag{15}$$

If the grid spacing is very small compared with wavelength, then, as the magnetic current $2\mathbf{K} = 2\mathbf{E}_t \times \hat{n}$ varies from zero on the grid wires to maximum in the space between, the other terms in the integrand of eq. (15) are essentially constant. Thus, we may replace the fluctuating **K** with its average value, $\mathbf{K}_{\text{avg}}$,

$$\mathbf{E}_1 = -\frac{jk}{4\pi R} e^{-jkR} \int [(2\mathbf{K}_{\text{avg}}) \times \hat{R}] e^{jk\mathbf{R}'\cdot\hat{R}} dA. \tag{16}$$

Although we know the direction of **K**, we do not know its magnitude unless the reflection coefficient of the grid is known. In the usual case, the grid is designed to introduce negligible insertion loss for the desired polarization, whence $\mathbf{E}_1$ on axis should equal that present when no grid is used. In this case, the magnitude of $2\mathbf{K}_{\text{avg}}$ would have to be such that

$$2\mathbf{K}_{\text{avg}} = 2\mathbf{E}_1 \times \hat{n} \text{ (negligible insertion loss)} \tag{17}$$

in order that eq. (16) will result in the correct on-axis value for $\mathbf{E}_1$. By substituting eq. (17) into eq. (16), we arrive at eq. (7), the desired equation for computing the field transmitted through the wire grid.

REFERENCES

1. A. C. Ludwig, "The Definition of Cross Polarization," IEEE Trans., *AP-21* (January 1973), pp. 116–119.
2. T. S. Chu and R. H. Turrin, "Depolarization Properties of Off-Set Reflector Antennas," IEEE Transactions, *AP-21* (May 1973), pp. 339–345.
3. R. E. Collin, *Field Theory of Guided Waves*, New York: McGraw-Hill, 1960, p. 366.
4. J. R. Wait, "Reflection at Arbitrary Incidence from a Parallel Wire Grid," Appl. Sci. Res., Sec. B, *4*, No. 6, 1954–1955, pp. 393–400.
5. M. J. Gans and R. A. Semplak, "Some Far-Field Studies of an Offset Launcher," B.S.T.J., *54*, No. 9 (September 1975), pp. 1319–1340.
6. R. H. Turrin, "Dual Mode Small-Aperture Antennas," IEEE Trans. Antennas and Propagation, *AP-15* (March 1967), pp. 307–308.
7. D. A. Gray, "Depolarization of ATS-6 Satellite 20 GHz Beacon Transmitted Through Rain," presented at the 1975 USNC-URSI Spring Meeting, June 3–5, Urbana, Illinois.
8. R. H. Turrin, personal communication.
9. V. H. Rumsey, "Some New Forms of Huygens' Principle," IRE Transactions, *AP-7* (Suppl.) (December 1959), p. S103.