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Resonant-Grid Quasi-Optical Diplexers

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Experimental results are reported concerning the transmission, reflection, and depolarization of metal grids that reflect an upper band, centered at 30 GHz, and transmit a lower band, centered at 20 GHz. With a single grid, the transmission loss is less than 0.1 dB, and the rejection exceeds 44 dB. Depolarization is measured under realistic conditions with direct dual-mode feed excitation. In a 10-percent band, depolarization is below 34 dB at all scanning angles for both the transmitted and reflected waves. Experiments with two parallel resonant grids are also reported.

I. INTRODUCTION

In many millimeter-wave systems associated with communication-satellite antennas or Hertzian cables,[1] quasi-optical filters and diplexers are attractive. Because of their large areas, quasi-optical devices have large power-handling capability and the multimoding problem is, in a sense, avoided. The ohmic losses can be small, and the grids are easy to manufacture by photographic techniques.

A simple type of diplexer is the plane-parallel Fabry-Perot resonator, incorporating parallel inductive grids and operating under oblique incidence.[2] The transmission of a plane-parallel Fabry-Perot resonator is essentially the same as that of a single-pole filter. This type of diplexer, however, suffers from the walk-off effects associated with the diffraction and lateral displacement of the incident beam.[3,4] This effect is aggravated if more than two grids are used to obtain a maximally flat response. The number of grids required, and therefore the walk-off effects, are minimized if the grids have resonant properties of their own. Narrow-band resonant crosses have been used in the far-infrared region.[5] No diplexing operation, however, was considered. The special features of the grid patterns considered here are their broadband characteristics and their capability of operating under oblique incidences. Preliminary results were reported in Ref. 6. In the present paper, new experimental results concerning the transmission, reflection, and depolarization of resonant-grid diplexers are reported. We give special attention to the depolarization of incident waves because, in some applications, it is required that two orthogonally polarized channels be transmitted, and depolarization introduces crosstalk. An ideal grid does not cause depolarization of incident plane waves under normal incidence when the pattern is invariant under a 90-degree rotation. This useful property of square array grids does not hold, in general, for waves under oblique incidence. However, we have observed that the depolarization remains small for a particular orientation of the grid in its own plane.

Although the main features of the diplexer response are easy to understand, some details are not yet fully understood. This is the case for the sharp spurious dips in transmission observed at certain angles of incidence and for the coupling through evanescent fields. In most of our experiments we tried to avoid the evanescent coupling by misorienting the grids. This coupling mechanism would perhaps be useful if it were precisely understood.

II. TRANSMISSION AND REFLECTION

The transmission characteristics of resonant grids are described here in an essentially qualitative manner. We assume that the grid periods, $\mathbf{p}_1$, $\mathbf{p}_2$, have equal magnitude and are perpendicular to one another. If the wavefronts are plane, unlimited, and parallel to the plane of the grid, the electric field can be decomposed into two components, one parallel to $\mathbf{p}_1$ and one parallel to $\mathbf{p}_2$. If the grid pattern is invariant under a 90-degree rotation, the transmission is the same in amplitude and phase for these two components, and therefore no depolarization is suffered. The transmission curves in Fig. 1 are applicable to such grids. Because the periods are much smaller than the wavelength, the grids can be represented by lumped circuit elements. If the filter incorporates more than one grid, the fine field structure of one grid (space harmonics) is assumed to be negligible at the other grid location. When the grid spacing is small, it is advisable, as indicated before, to set the grids at a small angle to one another, of the order of 2 degrees, to prevent a spurious coupling from taking place.

Figure 1a shows a simple mesh. This grid can be represented by an inductance in parallel on a transmission line representing free space. The smaller the opening areas, the smaller the grid reactance. When two such grids are parallel to each other, with a spacing slightly less than $l\lambda_o/2$, where l denotes an integer, a resonance takes place that can be pictured as resulting from the wave being reflected back and forth between the two grids. At each reflection, a wave of small amplitude is transmitted. Because the round-trip path length is of the order of $l\lambda_o$, the waves transmitted at the successive passes are in phase and add up. If the system has a plane of symmetry and the losses can be

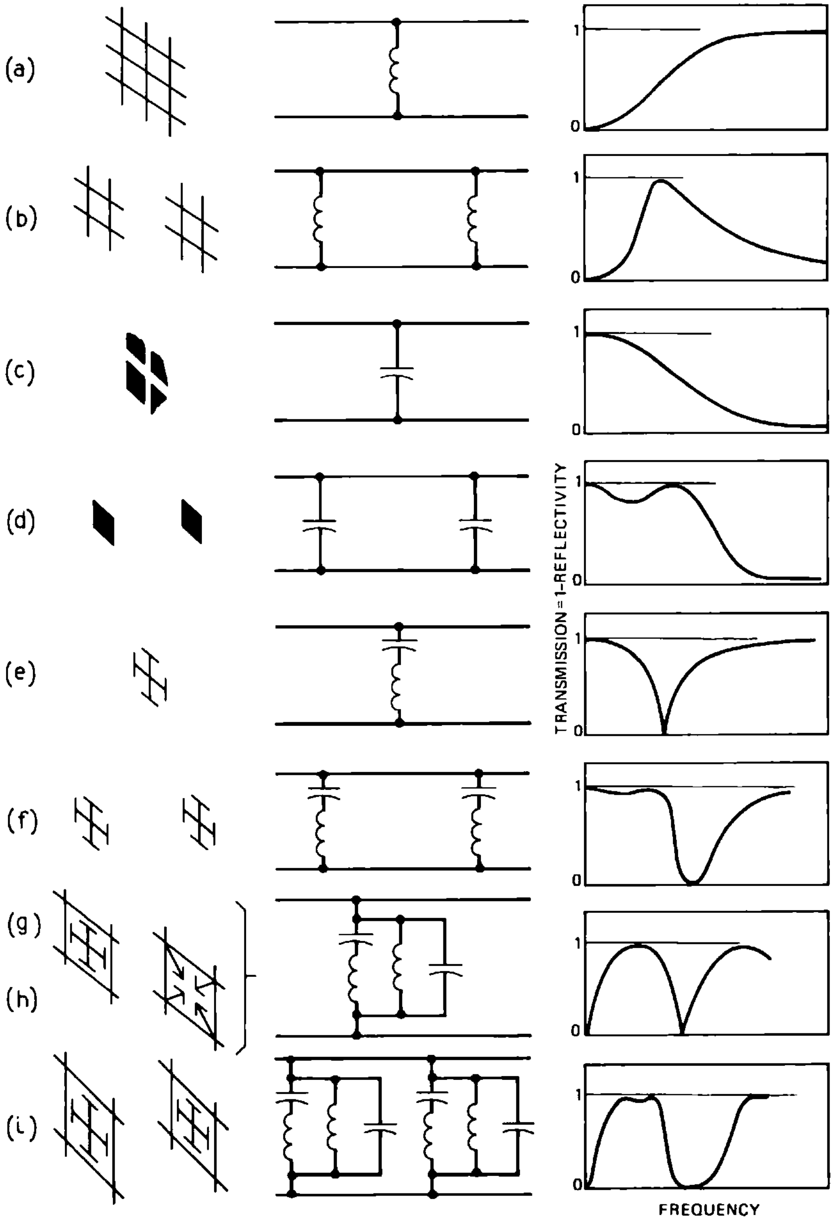

Fig. 1—Schematic representation of single-grid and double-grid diplexers. (a) Inductive grid. (b) Double-inductive grid. (c) Capacitive grid. (d) Double-capacitive grid. (e) Jerusalem cross (with rejection frequency). (f) Double Jerusalem cross. (g) Gridded Jerusalem cross. (h) Self-supported diplexer. (i) Double-gridded Jerusalem cross.

neglected, the transmission reaches 100 percent at one frequency at least. When the bandwidth is large, the response curve exhibits some dissymmetry, with a slower decay above resonance than below. If the metal and opened parts are exchanged, as shown in Figs. 1c and 1d, a capacitive grid is obtained, whose reflectivity is precisely equal to the inductive grid transmissivity. The square capacitive elements need, of course, to be supported by a dielectric sheet, perhaps in mylar. When the array period is small compared with the wavelength, it is very difficult to obtain even a moderately large reflectivity, such as $R = 0.9$. This is because the gap between squares required to provide such a reflectivity with a small period-to-wavelength ratio is of the order of a few micrometers. The power transmissivity is, in general,

$$T = [1 + (B/2)^2]^{-1} \tag{1}$$

for a susceptance, B, normalized to free space. For infinitely thin strips with a gap g and period p, we have ($p \ll \lambda$, thickness $\ll g$)

$$B = (2p/\lambda) \log_e [1/\sin (\pi g/2p)]. \tag{2}$$

It follows from eqs. (1) and (2) that if, for example, $\lambda = 10$ mm, $p = 3$ mm, a 10-percent transmissivity ($R = 0.9$, $B = 6$) requires a gap as small as 0.1 μm. This was one of our motivations for proposing a modified capacitive grid with an inductance in series with the capacitance, shown in Fig. 1e. These capacitive elements resemble "Jerusalem" crosses. At the resonance frequency, the Jerusalem-cross grid is perfectly reflecting and behaves as a plain sheet of copper. In a typical case, the measured transmission at the rejection frequency, 30 GHz, is at least 44 dB below the incident power.

Perfect transparency is obtained only at very low frequencies. To obtain a transmission band, two arrangements are considered. One consists in assembling two Jerusalem-cross grids parallel to each other. The response curve in Fig. 1f is obtained. A second possibility consists in introducing inductive elements in parallel with the crosses, as shown in Fig. 1g. The resulting grid is called a "gridded Jerusalem cross." An alternative configuration that does not require a mylar support but has essentially the same characteristics as the gridded Jerusalem cross is shown in Fig. 1h. These grids (type g or h) are very simple and attractive. Most of the experiments that we report were made on these types of grids. Finally, two Jerusalem-cross grids can be used, parallel to each other. Broad, uniform, transmission bands and broad, uniform, rejection bands are then obtained.

A grid, whatever its design, can be represented by circuit elements that are found empirically by fitting the measured response curve to the one calculated from the equivalent circuit. Sometimes, circuit elements that cannot be localized on the pattern need to be added to match the experimental curve. For instance, in the grid in Fig. 1g, a parallel capacitance needs to be added to account for a second, unexpected, transmission band above the rejection band. Even for a simple grid such as the one shown in Fig. 1a, a parallel capacitance needs to be added that tunes the inductance at the frequency c/p, p being the array period and c the speed of light in free space. Just below that frequency, the decay of the energy density of the space harmonics is very slow. This additional stored energy is represented by a capacitance. Above the frequency c/p, the transmission becomes a complicated function of frequency because of the excitation of grating lobes, and a circuit representation is of limited usefulness.

The transmission at resonance does not reach exactly 100 percent because of dissipation losses (ohmic losses in the metal and dielectric losses if mylar backings are used) and of scattering losses caused by the array not being perfectly periodic. The walk-off losses discussed in the introduction result from the incident beam being finite in cross section. For a single grid, the term "walk-off loss" is not applicable, but a similar physical effect exists. A beam with finite cross section has a finite angular spread, and a loss is suffered if the grid response depends significantly on the angle of incidence.

In a Fabry-Perot resonator incorporating conventional grids, the ohmic losses are small, particularly if the spacing between the grids is large (that is, if the axial mode number l is large). In contrast, high Q-factor resonant grids (e.g., narrow slits in a metal sheet) have rather high ohmic losses. Thus, resonant grids should be used only for broadband applications. Such grids are ideal, for example, for separating two channels widely separated in frequency, such as 20 and 30 GHz, or 4 and 6 GHz. Low-Q resonant grids are also useful in conjunction with conventional grids to eliminate side resonances in narrow-band filters. In any case, they provide greater flexibility in the filter design.

For convenient reference, let us give the expression for the susceptance, B, of the circuit shown in Fig. 1h (series L, C and parallel L', C')

$$B = (-L\omega + 1/C\omega)^{-1} + C'\omega - 1/L'\omega. \qquad (3)$$

In terms of the reactances at the resonance angular frequency of the LC circuit, $\omega_o = (LC)^{-\frac{1}{2}}$; that is, with $X \equiv L\omega_o = 1/C\omega_o$, $X' = L'\omega_o$, $X'' = 1/C'\omega_o$, and with $f \equiv \omega/\omega_o$, B is

$$B(f) = X^{-1}(-f + f^{-1})^{-1} + f/X'' - 1/X'f. \qquad (4)$$

This expression was used to generate the theoretical response curve in Fig. 7. For a single grid, this expression is to be substituted in eq. (1) to obtain the transmission T.

III. DEPOLARIZATION

The characterization of the depolarization introduced by a grid is discussed in this section. Let us first assume that the grid is an infinite plane and that the incident wave is plane. The normal to the wavefront of the incident wave and the normal to the grid plane define the angle of incidence i (= reflection angle). The orientation of the grid wires can be defined by the angle ν that they make with the normal to the incident plane in the clockwise direction for the direction of propagation. For a linear incident polarization, the polarization is defined by the angle Ω that the electric field makes with the normal to the incident plane, again in the clockwise direction. Arbitrary incident polarizations can be defined by their two components, along the normal to the incident plane and along the perpendicular direction. A wave is said to be E-polarized (or TM) if the magnetic field is linearly polarized and perpendicular to the incident plane ($\Omega = 90°$), and H-polarized (or TE) if the electric field is perpendicular to the plane of incidence ($\Omega = 0$). For given i, ν, the grid response is characterized by the complex transmission coefficients t_{EE}, t_{EH}, t_{HE}, and t_{HH}. These four parameters are, in general, functions of frequency. Because of linearity we have

$$\begin{aligned} e'_E &= t_{EE}e_E + t_{EH}e_H, \\ e'_H &= t_{HE}e_E + t_{HH}e_H. \end{aligned} \tag{5}$$

The parameters can be considered the elements of a complex 2-by-2 matrix $\mathbf{t}$ defined as

$$\mathbf{t} \equiv \begin{bmatrix} t_{EE} & t_{EH} \\ t_{HE} & t_{HH} \end{bmatrix}. \tag{6}$$

In terms of this matrix, (5) is written

$$\mathbf{e}' = \mathbf{te}. \tag{7}$$

Ideally, we would like to have $t_{EE} = t_{HH} \equiv t(\omega)$ and $t_{EH} = t_{HE} = 0$, in which case $\mathbf{e}' = t\mathbf{e}$, t being a scalar. We may require the less stringent condition that the state of polarization of the incident field be preserved to within an arbitrary, but fixed, rotation angle: $\mathbf{e}' = t\mathbf{Re}$, where $\mathbf{R}$ denotes a fixed rotation matrix and t a scalar function of ω. For this to happen, $\mathbf{t}$ must have the form $\mathbf{t} = t\mathbf{R}$. Reflection by an even number of plane mirrors, for example, preserves polarization in that sense. An even less stringent requirement is that orthogonal incident polarizations remain orthogonal. It can be shown[7] that this is the case if and only if $\mathbf{t}$ has the form $t\mathbf{U}$, with t a complex or real number and $\mathbf{U}$ a unitary matrix (that is, $\mathbf{U}^\dagger\mathbf{U} = \mathbf{1}$, where $\dagger$ denotes transposition and complex conjugation).

Symmetry considerations sometimes show that depolarization should not be present. Let us, for example, assume that the grid is a rectangular mesh, with a period vector perpendicular to the incident plane (that is, $\nu = 0°$, modulo 90°). By reason of symmetry, we must have $t_{EH} = t_{HE} = 0$. In general, however, $t_{EE} \neq t_{HH}$. Thus, for that case, E waves and H waves are not depolarized, but waves with any other polarization, for instance, a wave with linear polarization at 45 degrees, will be depolarized and acquire an elliptical polarization. To avoid depolarization, it is not sufficient that $|t_{EE}| = |t_{HH}|$. The phases of t_{EE} and t_{HH} must be equal, too. However, if the magnitudes of t_{EE} and t_{HH} are found equal over a large band of frequency, the phases of

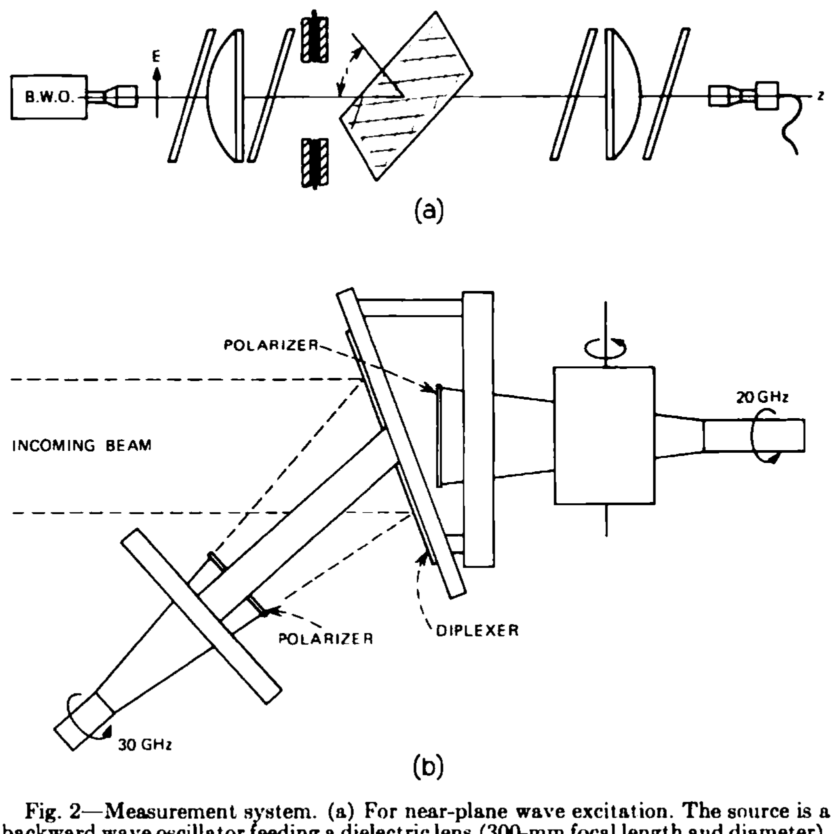

Fig. 2—Measurement system. (a) For near-plane wave excitation. The source is a backward wave oscillator feeding a dielectric lens (300-mm focal length and diameter). The system exhibits low depolarization (< -50 dB) when properly aligned. The incident electric field is vertical or horizontal. The grid under test can be rotated in its own plane. The angle ν refers to the angle between the grid wires and the normal to the incident plane. The angle of incidence (i) can be varied, and the grid assembly can be rotated by steps of 45 degrees about the z-axis (angle Ω) with the help of a 45-degree wedge. (b) Diplexer mounted with the feed.

t_{EE} and t_{HH} are most likely equal because of the integral relations existing between phase and amplitude for minimum phase circuits. Furthermore, if the grid can be considered very thin and lossless, there exists a simple relation between the phases and the moduli of the transmitted (t) and reflected (r) fields. The relation is

$$\begin{aligned} r &= -\cos\phi \exp(i\phi), \\ t &= -i \sin\phi \exp(i\phi). \end{aligned} \tag{8}$$

Thus, for thin grids, the phase of t can be obtained from the modulus of t at any frequency.

Another important result applicable to thin grids is that two-dimensional scaling applies. That is, nothing is changed if the wavelength and the grid dimensions in the plane are multiplied by the same factor. For thin grids, the Babinet principle also applies, which says that the reflectivity of a grid is equal to the transmissivity of its complement.

Let us now describe the experimental setup for plane wave measurement. The source (a backward wave oscillator) radiates through a dual-mode feed. The beam is collimated by a dielectric lens and sub-

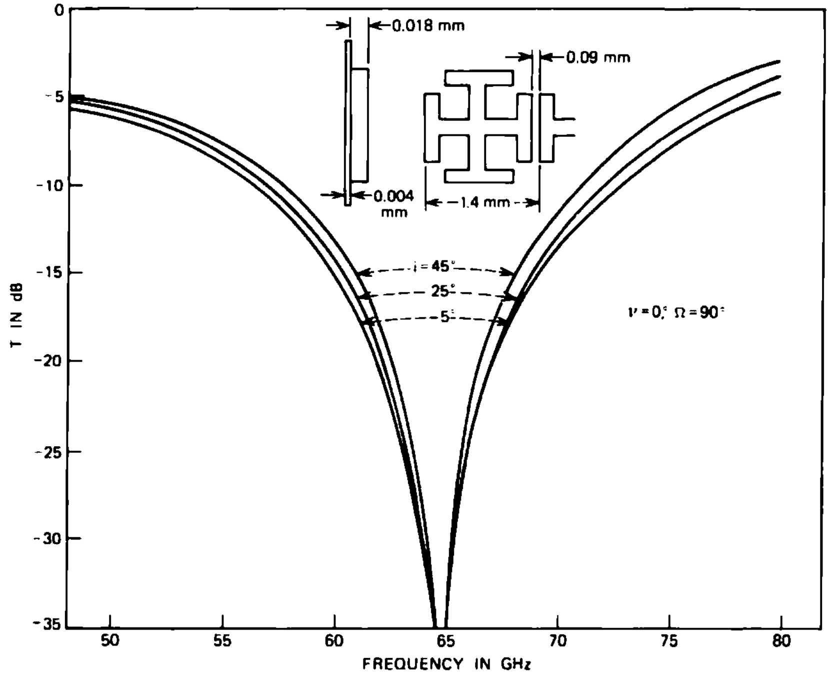

Fig. 3—Dimensions and response curve of a Jerusalem-cross grid. i denotes the angle of incidence ($\nu = 0°$, $\Omega = 90°$: horizontal electric field).

sequently focused by a second lens back on a collecting dual-mode feed. The size of the beam between the two lenses is sufficiently large that the field can be well approximated by that of a plane wave. To check the depolarization of the system, the collecting feed is rotated about its axis until a null is obtained. Because the system has a plane of symmetry, the cross-polarized components should be equal to zero. These components are found to be less than -50 dB. To damp the reflections that take place from the lenses and the feeds and between lenses, attenuating glass plates are introduced. These glass plates are tilted, but symmetry in the vertical plane is preserved. In this arrangement, the transmitted electric field must remain either vertical or horizontal. Otherwise, the tilted glass plates would depolarize the beam because of differential loss. Thus, the filter under test, rather than the source, needs to be rotated about the system axis (z) to measure its response for various polarization angles. In practice, it is sufficient to measure the depolarization for three angles, $\Omega = 0$, $45°$, and $90°$, and pick up the worst number. This maximum depolarization (X) is a function of the angle of incidence (i), of the angle that the grid wires

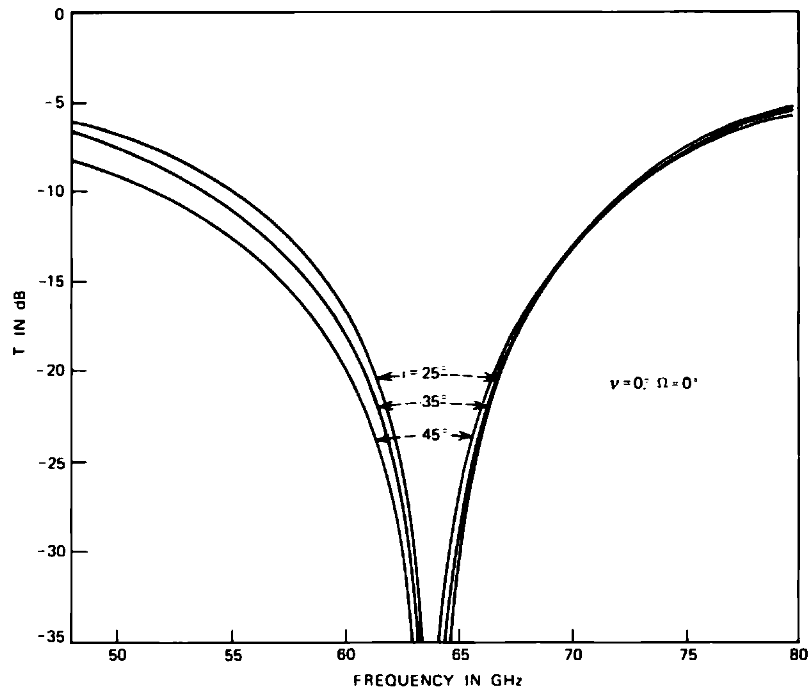

Fig. 4—Same as Fig. 3 with $\Omega = 0°$. For $\Omega = 45°$, the response (not shown) is found to be almost independent of i, for $i = 25°$ to $45°$.

make with the normal to the incident plane (ν), and, of course, of frequency. Thus, the plane-wave depolarization of a diplexer in transmission is characterized by a function

$$X_t = X_t(i, \nu, \omega) \qquad \text{(dB)}.$$

The depolarization in reflection is similarly defined, but the mechanical arrangement is more complicated.

In some applications, the incident wave is diverging rather than plane. This is the case when the diplexer is used to separate beams just before the feed of a primary feed antenna, as shown in Fig. 2b. The definition of what constitutes a perfect feed from the point of view of polarization is not obvious. It has been observed that, if the feed is intended to be used at the focal point of a parabolic dish, its polarization pattern should be the same as that of an Huygens source, a combination of electric and magnetic dipoles.[8] The feeds used in this paper have relatively narrow beam patterns (about ±4 degrees at the 3-dB points), and it seems that the ambiguities associated with the definition

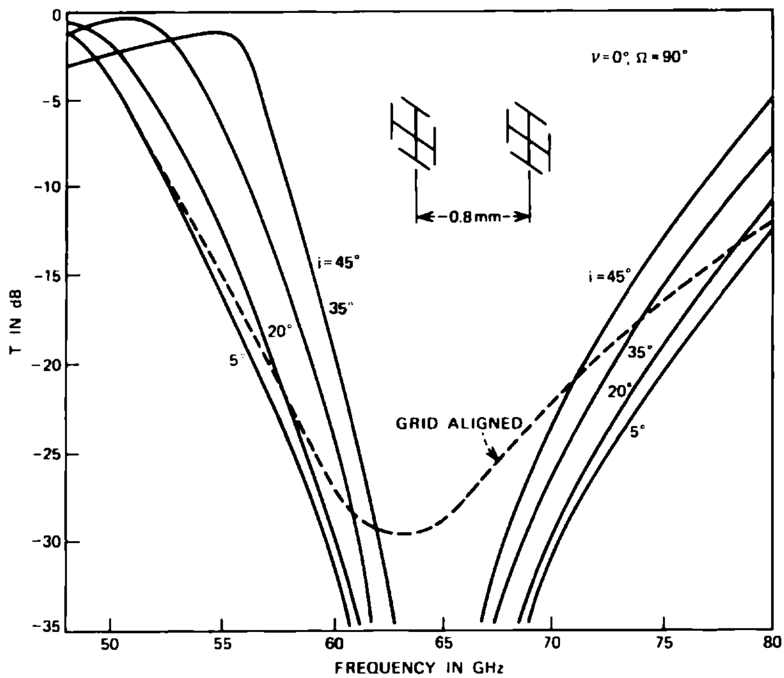

Fig. 5—Response curve of a filter incorporating two Jerusalem-cross grids as in Fig. 3, with a 0.8-mm spacing. The grids are at a small angle (~3°) to one another to avoid evanescent wave coupling. When the two grids are aligned, the response shown by a dashed line is obtained. For all curves, Ω = 90° (horizontal electric field).

of what constitutes a "good" feed can be ignored. The test thus simply consists of scanning the diplexer assembly in azimuth for various incident polarizations (0, 45, and 90 degrees) and measuring the cross-polarized component.

IV. GRID FABRICATION

The grids were made either in beryllium copper (75-μm thick) or from a sheet of copper bound on a mylar film. The fabrication consists of generating a mask from a computer program and using conventional photoetching techniques.

To generate the mask, three plotting techniques were used, all computer-driven. For simple grid patterns, the most convenient system seems to be the "Litehead" plot, which is a scanned focused beam of light. It was used for fabricating polarizers. For the complicated patterns considered in this paper, we used the "Rubylith" plot, which uses mechanical cuts. Because of computer limitations, only moderate array sizes can be obtained. To obtain large array sizes, it is necessary to use a "step-and-repeat" camera technique. This photographic technique has size limitations, and further mask joining is required.

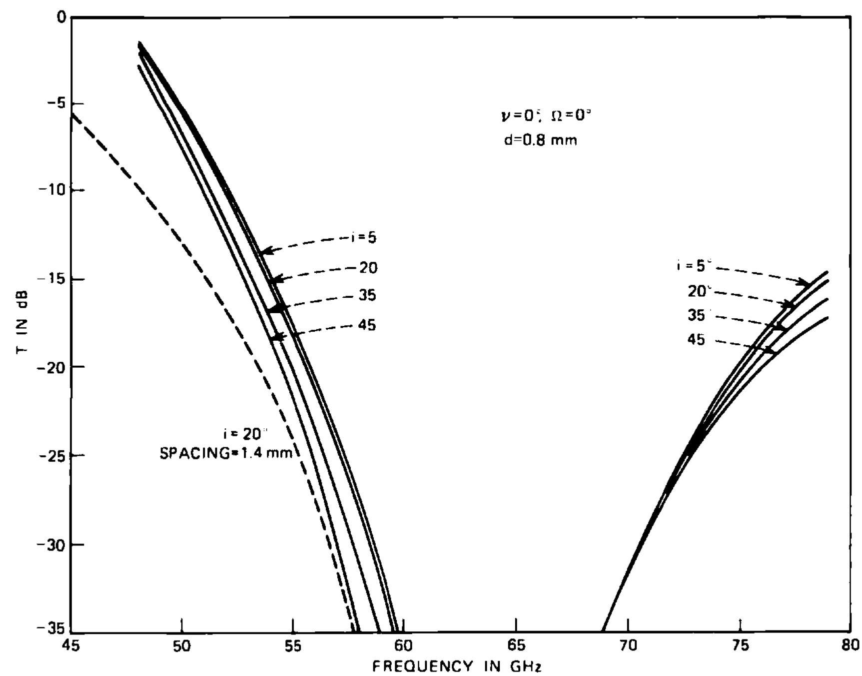

Fig. 6—Continuation of Fig. 5 with $\Omega = 0°$ (vertical electric field). The dashed line is for a 1.4-mm spacing.

An alternative technique is the PPG (primary pattern generator) system. This system is a raster-scan-type device that generates 190-mm by 240-mm arrays with a 7-μm by 7-μm resolution. This is the technique used to generate the grids whose depolarization is shown in Figs. 9 to 12, 15, and 16, except for the dashed line in Fig. 9.

V. THE JERUSALEM-CROSS GRID

The dimensions of the crosses of a typical Jerusalem-cross grid and the response curve are shown in Figs. 3 and 4 (E and H polarizations) for various angles of incidence. The resonance frequency of the series

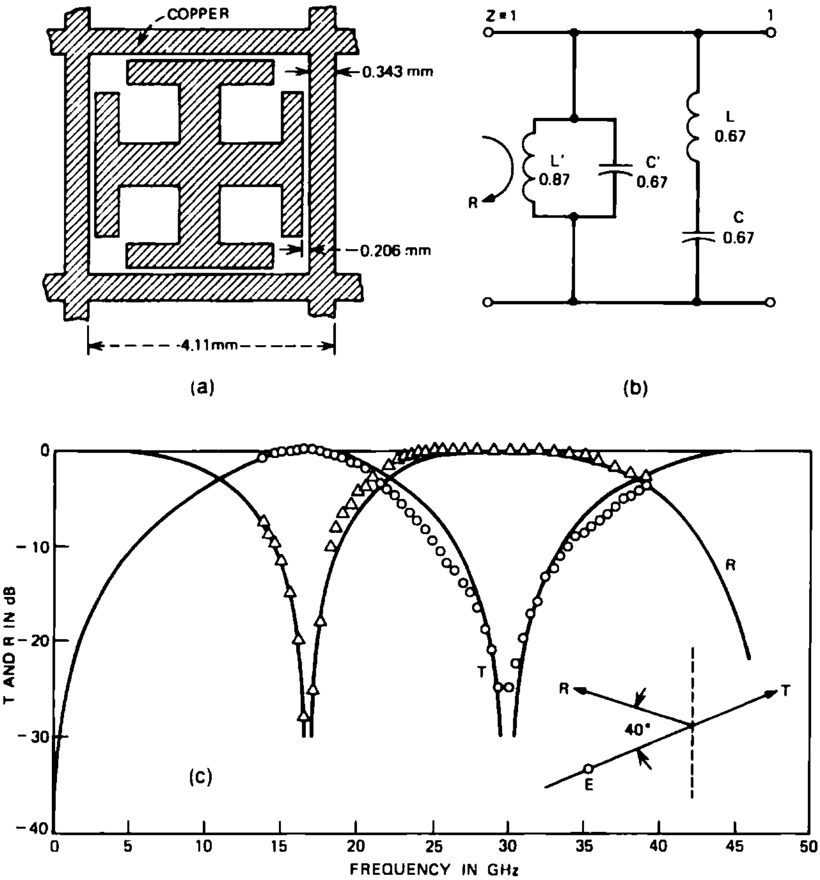

Fig. 7—Gridded Jerusalem-cross diplexer. The critical dimensions are shown in (a) and the equivalent circuit obtained by fitting the measured transmission and reflection curves in (c) is shown in (b). The peak transmission (loss < 0.1 dB) is at 16 GHz, and the peak rejection (> 44 dB) is at 30 GHz. The plain curve in (c) is theoretical from the equivalent circuit.

LC circuit is at 65 GHz. We note that the resonance frequency is almost independent of the angle of incidence.

The transmission characteristic of a pair of such grids, with a 0.8-mm spacing, at various angles of incidence, is shown in Figs. 5 and 6. At an angle of incidence of 35 degrees, for example, the transmission reaches its maximum at a frequency of 51 GHz. The rejection exceeds 30 dB from 61 to 69 GHz. In this experiment, the two grids are slightly rotated with respect to one another (about 3 degrees). The dashed curve in Fig. 5 shows that, when the grids have exactly the same orientation, the rejection is smaller. This probably results from a coupling through evanescent fields.

VI. THE GRIDDED JERUSALEM CROSS

A single grid exhibits a transmission band if parallel inductances are added to the series circuit. The grid dimensions and the measured response, both in transmission and in reflection, are shown in Fig. 7 (see also Ref. 6). The equivalent circuit has been selected to match the experimental response curve. The curve in Fig. 8 shows the effect of

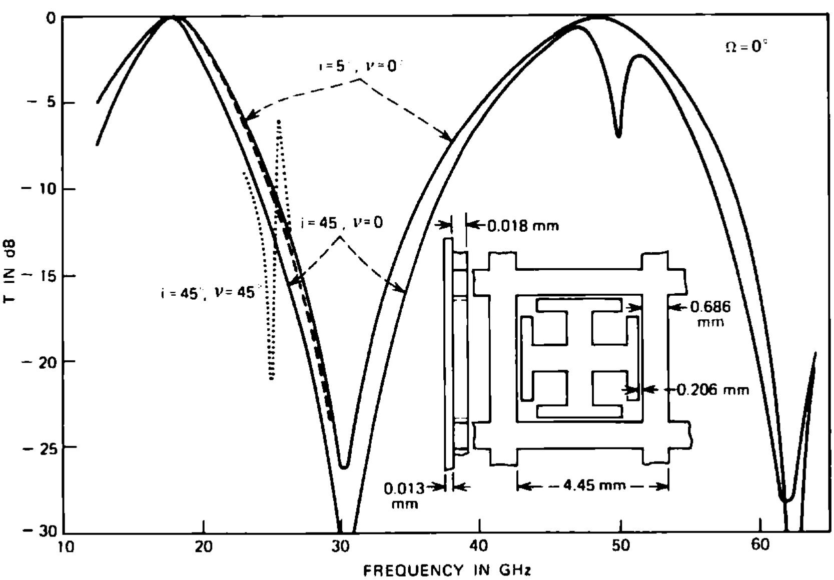

Fig. 8—Gridded Jerusalem-cross diplexer. This curve shows that the transmitted frequency can be raised from 16 to 18 GHz by increasing the width of the parallel wires. The transmitted curves are for angles of incidence i of 5 and 45 degrees, $\nu = 0°$. The small-dash curve is for the grid located between the feed (two wavelengths across) and the lens, that is, under strongly diverging wave excitation. The transmission curve is essentially unaffected. The curve for an angle of incidence of 45 degrees and the grid rotated in its own plane at $\nu = 45°$ exhibits a spurious dip.

increasing the width of the parallel strips from 0.343 to 0.686 mm. This brings the passband closer to the rejection band. By increasing the width of the parallel grids further to 0.840 mm and reducing the capacitive gap to 0.126 mm and the series inductance strip width to 0.336 mm, we found it possible to obtain a ratio of rejection frequency to transmission frequency as low as 1.5:1. An approximate theoretical analysis of the operation of the gridded Jerusalem cross under normal incidence will be reported shortly.[9] At a large angle of incidence and for some orientation of the grid, a sharp dip is observed. This dip is observed when the electric field is at 45 degrees to the series inductance wires. (The same rule applies to the "self-supported grid" discussed in Section VII.) A possible mechanism is the following. Consider as a simpler model an array of parallel strips with series capacitances. Because the phase velocity along the strips exceeds the velocity of light in free space, this system can, in principle, radiate. However, if the symmetry is perfect, the relevant space harmonic has zero amplitude. Thus, a very small lack of symmetry is needed to have radiation. The Q-factor of this resonance can be very high because the coupling to

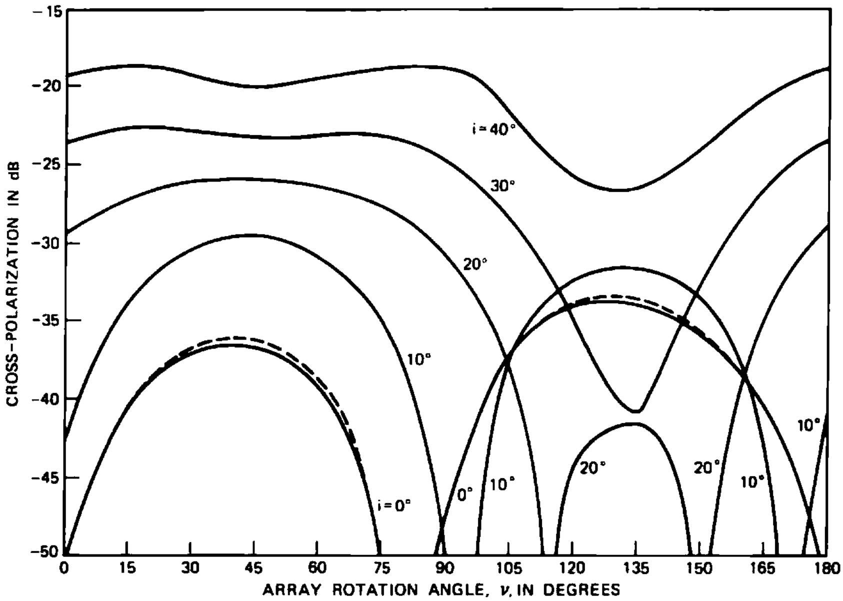

Fig. 9—Depolarization introduced by the gridded Jerusalem-cross diplexer (diplexer obtained from a PPG mask) in Fig. 8 under near-plane wave excitation for different angles of incidence (i), as a function of the rotation of the grid in its own plane (angle ν). The incident polarization is at 45 degrees to the incidence plane ($\Omega = 45°$). The dashed line is for joint masks (Rubylith plot).

incident plane waves is very small. Whether this model is applicable to the grids investigated remains to be seen.

The depolarization introduced by a grid of this type has been measured, both under plane wave excitation and under diverging wave excitation. The depolarization under plane wave excitation is shown in Fig. 9 as a function of the orientation of the grid in its own plane. It should be noted that, even under normal incidence, the depolarization is not zero and, under oblique incidence, the curve does not have a 90-degree period, contrary to our expectations based on the nominal symmetry of the grid. The first grid that we tested (whose response curve is shown as a dashed line in Fig. 9) was obtained from a composite mask made of two smaller masks. A slight discontinuity between the two masks was noted, which was thought to be responsible for the

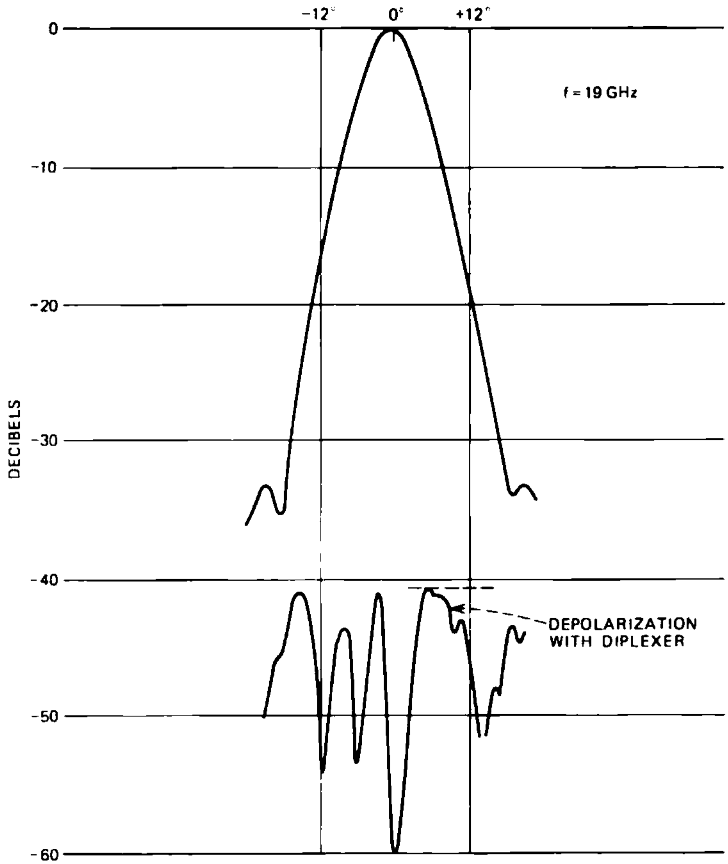

Fig. 10—Typical azimuth scan showing the transmission and depolarization of the diplexer in Fig. 8, at 19 GHz.

depolarization under normal incidence (34 dB maximum). However, the depolarization for a grid obtained from a single mask (PPG generator) is almost the same. The effect of diffraction at the edges of the filter has been eliminated as a possible cause for depolarization. No systematic lack of symmetry of the grid under a 90-degree rotation is noticeable under microscopic observation. The residual depolarization that we observe under normal incidence for the PGG grid, therefore, remains unexplained. A second observation is that excellent cross-polarization properties are obtained for angles of incidence less than 20 degrees for some orientations of the grid in its own plane (e.g., cross-polarization < -50 dB for $\nu = 110°$, $i = 20°$).

More complete tests were made in an anechoic chamber with dual-mode feed horns at 20 GHz (transmitted beam) and 30 GHz (reflected

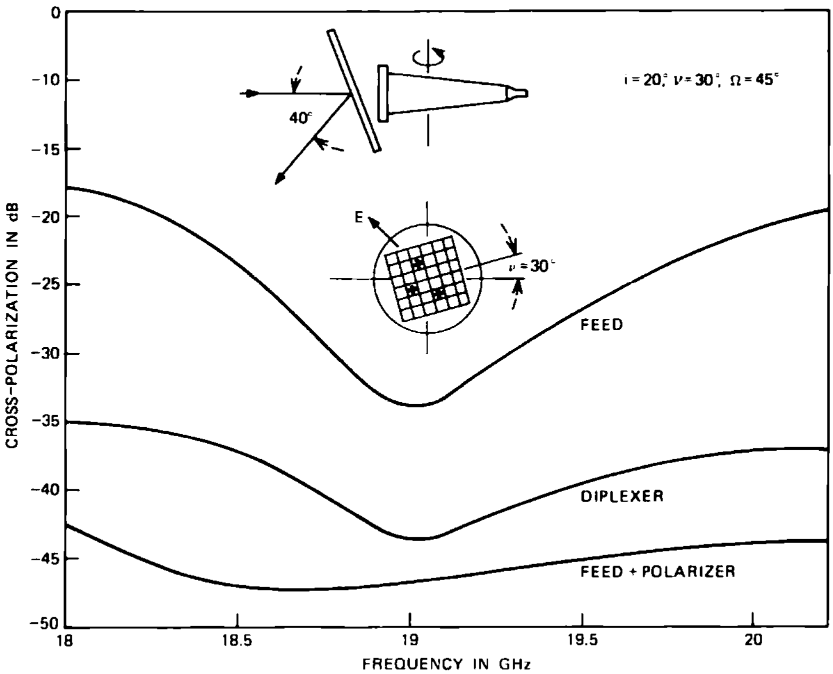

Fig. 11—Depolarization introduced by the gridded Jerusalem-cross diplexer in Fig. 8 under diverging wave excitation as a function of frequency. The feed diameter is 130 mm (3 dB points of the far-field pattern at ±5° at 19 GHz). The angle of incidence is 20 degrees. At each frequency, the diplexer (grid and feed together) is scanned in azimuth. The points shown correspond to the worst depolarization within a ±12° angle, corresponding to the −18-dB points of the far-field radiation pattern of the feed. The incident polarization is at 45 degrees to the incident plane. The orientation of the grid in its own plane ($\nu = 30°$) is shown in the figure. The upper curve is for the dual-mode feed alone. The lower curve is for the feed mounted with a grid polarizer. The central curve gives the depolarization introduced by the diplexer and feed combination.

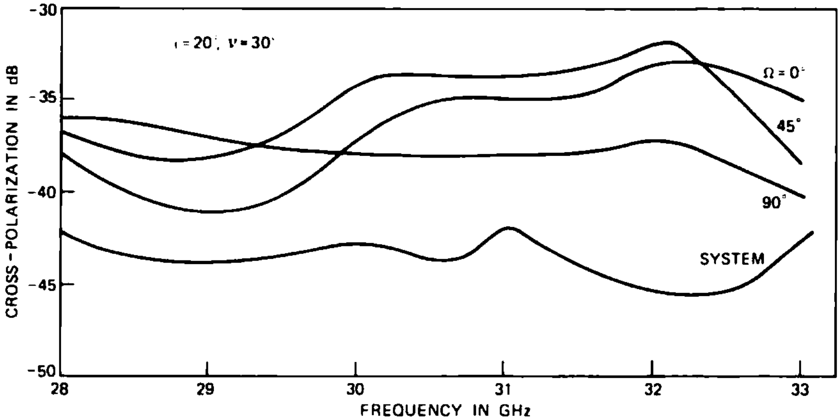

Fig. 12—Depolarization introduced on reflection in the 30-GHz band by the gridded Jerusalem-cross diplexer in Fig. 8. Measurements were made for the same grid orientation as in Fig. 11 for incident polarizations corresponding to angles $\Omega = 0$, 45°, and 90°. The worst depolarization within a $\pm 12°$ scanning angle, corresponding to the -22-dB point in the feed response at 30 GHz, is selected. The curve labeled "system" is for the dual-mode feed-mounted with a polarizer.

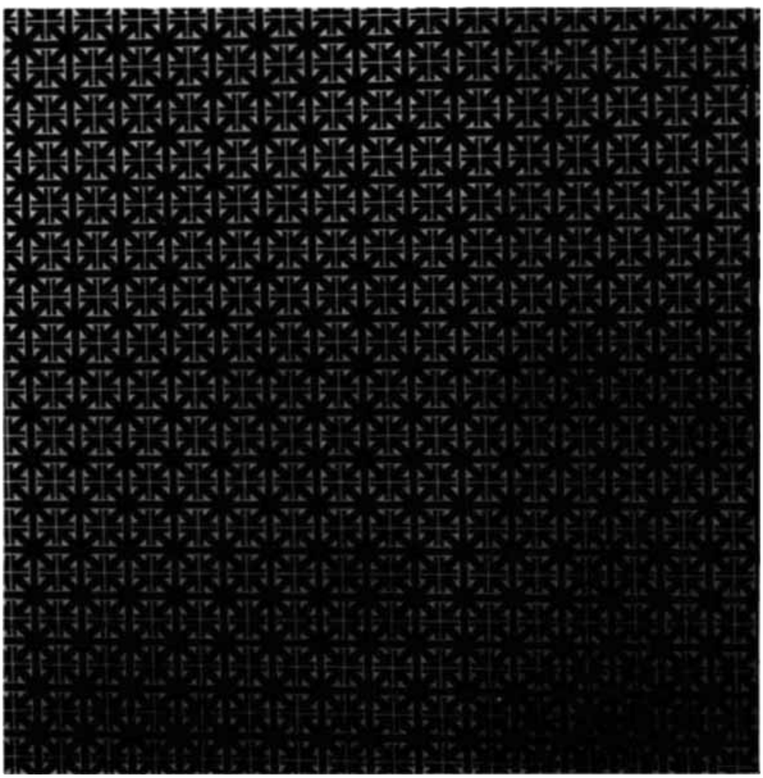

Fig.13—Photograph of a self-supported diplexer in beryllium copper.

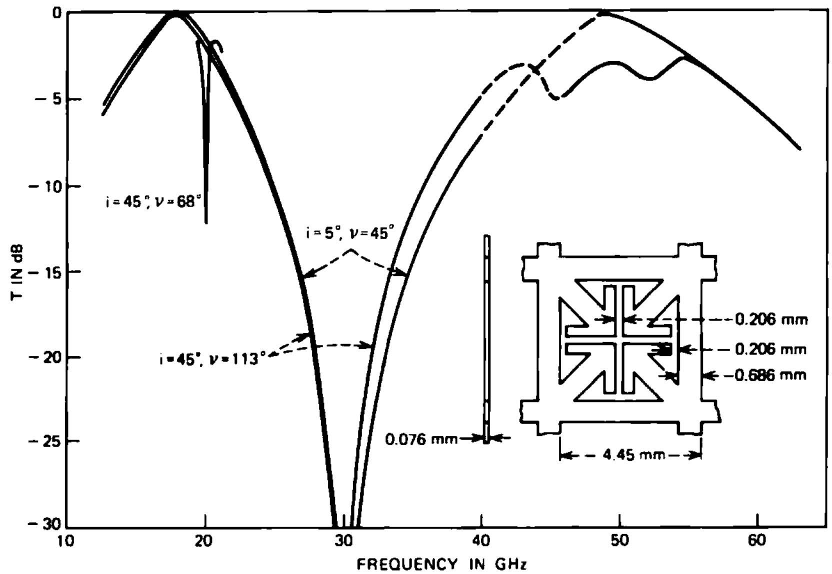

Fig. 14—Transmission of a self-supported grid in beryllium copper whose dimensions are shown on the figure. At large angles of incidence ($i = 45°$) and for some orientations of the grid, a large dip appears in the response.

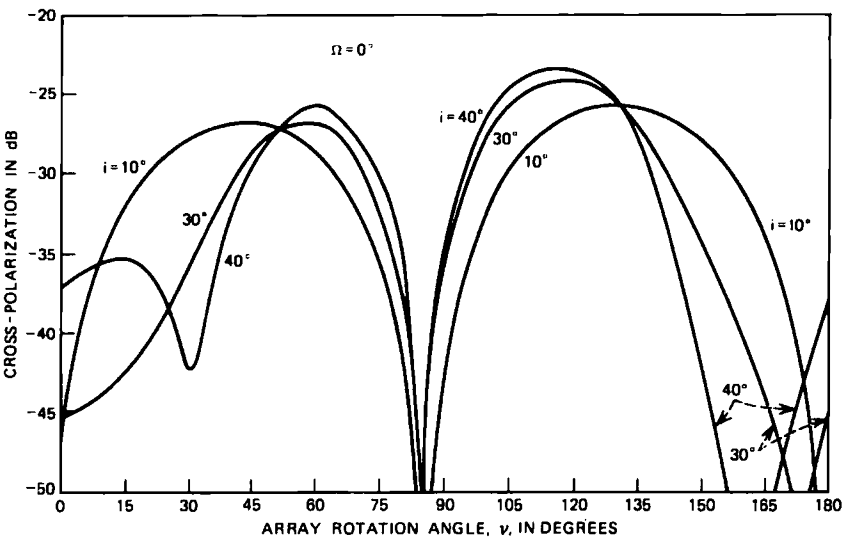

Fig. 15—Depolarization introduced by the self-supported grid in Fig. 13 as a function of the orientation of the grid in its own plane (angle ν) for various angles of incidence. The frequency is 19.5 GHz, the beam diameter is 130 mm, and the incident polarization is perpendicular to the incidence plane ($\Omega = 0°$).

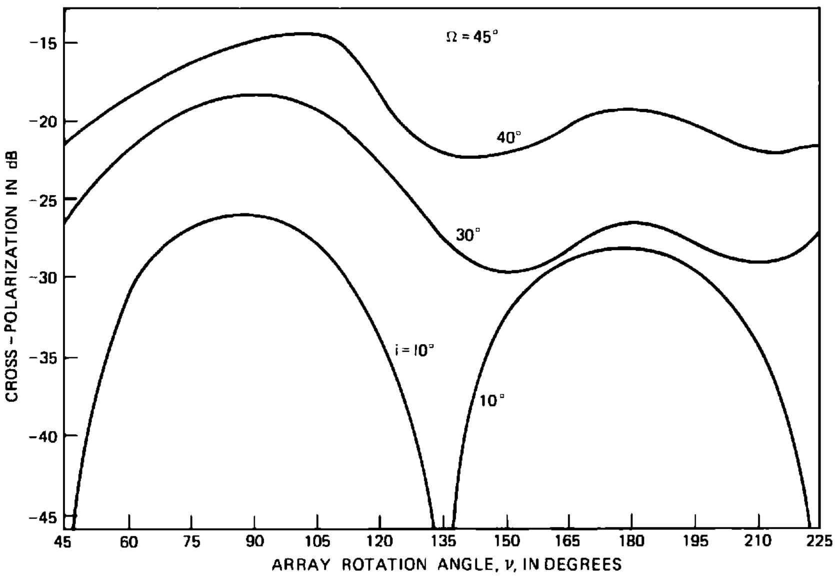

Fig. 16—Continuation of Fig. 14 with the incident polarization at 45 degrees to the normal to the incident plane (Ω = 45°).

beam). The orientation of the grid is selected for minimum depolarization in transmission at 19 GHz. It was noted that operation of a single-grid diplexer under diverging wave conditions does not significantly degrade the system response. The diplexer assembly was scanned within a ±12-degree angle (±12 degrees correspond to the −18-dB points of the feed response at 19 GHz). A typical scan is shown in Fig. 10. The worst depolarization is plotted in Fig. 11 as a function of frequency. The incident polarization was at 45 degrees to the incident plane. The depolarization is less than −35 dB from 18 to 20 GHz. The depolarization on reflection, shown in Fig. 12, turns out not to be sensitive to the orientation of the grid in its own plane. This is fortunate, since there is no particular reason to expect the optimum orientation of the grid to be the same on reflection and on transmission. The depolarization on reflection was measured for three incident polarizations, $\Omega = 0°$, $+45°$, and $90°$. Considering only the upper envelope of these curves, we find that the depolarization on reflection is less than −34 dB from 28 to 31 GHz.

In conclusion, we find that the Jerusalem-grid diplexer, operating with a 40-degree angle between reflected and incident beams, gives transmission and reflection losses less than 0.1 dB. The depolarization is less than −34 dB within 10-percent bands centered at 18 and 30 GHz.

VII. THE SELF-SUPPORTED DIPLEXER (Fig. 13)

The most critical dimensions of the self-supported grid are shown in Fig. 14. The equivalent circuit is almost the same as that of the gridded Jerusalem-cross grid previously discussed, but this new grid does not require a mylar backing. The grid was made of beryllium copper, 75-μm thick. The transmission characteristic is shown in Fig. 14 as a function of frequency. Here again, a sharp dip in transmission shows up for some orientations of the grid. The depolarization introduced by the self-supported diplexer is shown in Figs. 15 and 16 for quasi-plane wave excitation. We observe that a very small depolarization can be obtained for a proper orientation of the grid in its own plane, if the angle of incidence does not exceed 20 degrees.

VIII. DOUBLE-SELF-SUPPORTED DIPLEXER

An almost flat response can be obtained from 17 to 19.5 GHz by combining two self-supported grids (shown in Fig. 13) with a 6.3-mm

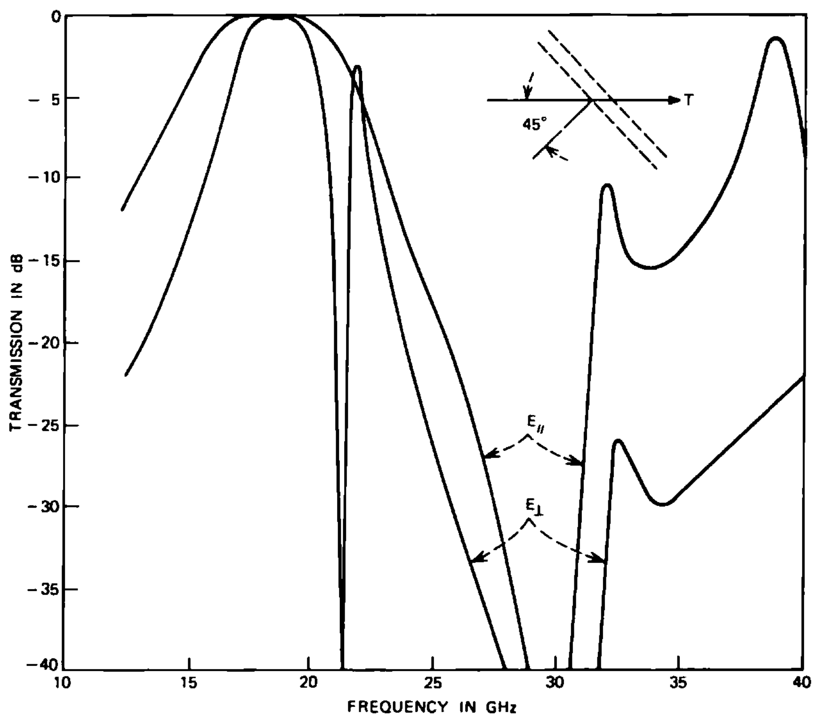

Fig. 17—Transmission of a double-self-supported diplexer with a 6.3-mm spacing at 45-degree angle of incidence, for two polarizations. The grid mask is as shown in Fig. 14. The grids were etched on 76-μm-thick pure copper. Grid size = 300 $\times$ 460 mm. Depolarization at 19 GHz is below -30 dB and the peak rejection at 30 GHz exceeds 60 dB.

spacing (Fig. 17). A broad rejection band is also obtained, with a peak rejection of −68 dB.

IX. CONCLUSION

We have shown that a single metallic grid may constitute an efficient diplexer to separate or combine frequency bands in the ratio 1.5:1 or more. The depolarization is low and could be reduced further by selecting unequal periods and dimensions. Because a very accurate theory is not presently available for the complicated patterns that we have investigated, an optimum design would require further testing. For a two-resonant grid diplexer, flat responses are obtained, and the depolarization remains acceptable for most applications. Some of the effects observed: depolarization under normal incidence, sharp dips in transmission at some orientations of the grids, and evanescent coupling, though understandable in principle, remain to be investigated.

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