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TITLE

ANALYTICAL STUDY OF THE CIRCULAR POLARIZATION SELECTIVE SURFACES - 7.75 AND
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Final Report

Analytical Study of the Circular Polarization
Selective Surfaces - 7.75 and 15GHz.

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WRI Project No. 1468101

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March 13, 1990

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Summary

Dr. W.V. Tilston invented the principle of the circularly polarization sensitive surface (CPSS) in 1986. The surface is a large planar array of crossed dipoles. Each crossed dipole pair is separated by $\lambda/4$ distance but is connected between the two feed points by a $\lambda/2$ transmission line. Dr. Tilston constructed a CPSS surface [1] with incomplete success since he had difficulty in fitting the $\lambda/2$ transmission line length to the $\lambda/4$ separation of the dipole pair.

Loading of the transmission line by high dielectric constant material to relieve such difficulty is ruled out because of the construction difficulty and expenses involved for a large array of over 1000 dipole pairs. The purpose of this contract is to give a new cross dipole design to overcome such difficulty without the above loading of transmission line.

This design has been done and the left hand CPSS surfaces constructed has been measured and found to give the desired polarization selective property. The construction and measurement were done by David and Steven Tilston of Til-tek Inc. of Kemptville, Ontario, under a separate contract from DREO [2].

This report gives the details of the design and the underlying principle. Briefly, the principle is to take the transmission line length to be $\lambda/4$, instead of $\lambda/2$, but to embed the remaining lengths of transmission line in the dipole pair, that is: $\lambda/8$ in each dipole.

The dipole array is printed on a dielectric slab of $\lambda/4$ in thickness for mechanical support. The effect of the dielectric slab is accounted for in the design.

The original contract calls for a design for 20 GHz. After the initial design studies, it is found that the dipole pairs resulted are too small and make the construction difficult. Therefore, with the agreement from the Scientific Authority, Dr. Gilbert Morin of DREO, the design was constructed instead for two frequencies, 7.75GHz and 15GHz.

Based on the discussion of this report and a reference search, we believe that the design is novel, and there is a need for this design in industry. Therefore, we believe that this design be patented and published.

I Introduction

Many satellite communication (SATCOM) and navigating systems use circular polarization to avoid polarization mismatch due to rotations of the satellite or the vehicle or both.

It is desirable to design antennas of such polarization with fast scanning to cover a vast area. Parabolic reflections are not suitable for fast scanning because of their large moment of inertia. Phased arrays with no moving parts are well suited for fast scanning except for the high expenses of their electronic phase shifting systems. The compromise of the two, with small moment of inertia and low expenses, are the scanning mirror systems with polarization selective surfaces.

Such type of surfaces reflects one polarization and transmits the other polarization with linear polarizations. With linear polarizations such surfaces are well known [3]. With circular polarizations such surfaces are unknown until recently. In 1984, W.V. Tilston, of Til-tek Ltd., Kemptville, Ontario, proposed a novel design for such circular polarization selective surface (CPSS). Because of construction difficulty, a modified version of such design was constructed and tested (CPSS contract serial No. 2SV84-00198). The results are reasonably good, but it was generally agreed that it could be improved.

The construction difficulty lies on the design requirement of having a $\lambda/2$ transmission line length through a $\lambda/4$ dielectric substrate. This means that the transmission line must have a propagation velocity half of that in the dielectric substrate. This requires the loading of the substrate near the transmission line with dielectric of 4 times the substrate dielectric constant. For a surface requiring say, a thousand of these transmission lines, this can present a serious problem. To side step this problem the transmission lines were not loaded in the above mentioned modified version.

The purpose of this contract is not to side step this problem after it has been tried, by Til-tek Inc., but to face the problem squarely. This is done through incorporating the extra $\lambda/4$ transmission line, beyond the $\lambda/4$ substrate, into the two adjacent dipoles. The arrangement of the dipoles and the transmission shall be clear in the next section.

II The Original Design by Tilston

The original design of Dr. W.V. Tilston is shown in Fig. 1. The dipoles and the transmission line (Twin Lead) are connected for scattering of the LHCP waves.

The mechanism of the scattering can be understood by considering a LHCP plane wave incident along the negative z-axis. Because of the LHCP, the $\lambda/4$ spacing between the two perpendicular dipoles and the $\lambda/2$ transmission line introducing a 180° phase shift, as shown in Fig.2. The induced voltages at the terminals of the two dipoles add in phase to bring about a large current flow which in turn causes the scattering.

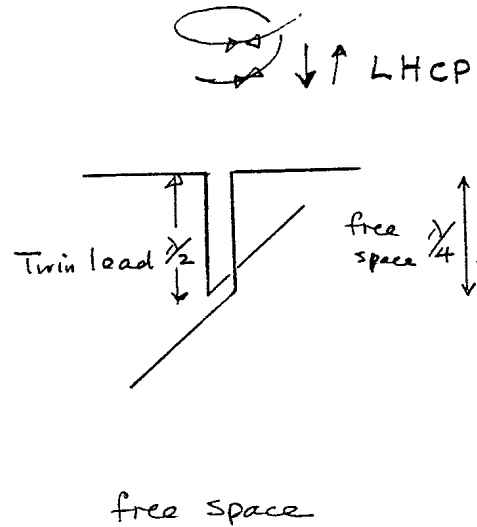
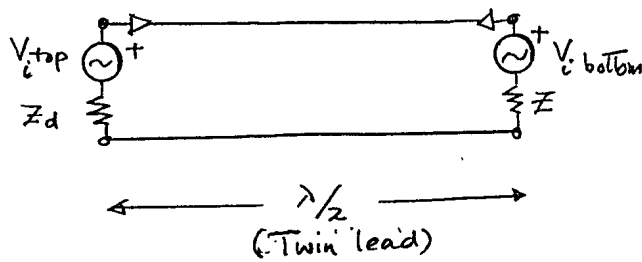


Fig.1. Two perpendicular dipoles on both sides of a $\lambda/4$ substrate but connected by a $\lambda/2$ twin lead.

With this understood, it is easy to see that with a RHCP plane wave instead, the induced voltages cancel and give zero current and therefore no scattering. The RHCP plane wave then propagates without seeing the dipoles and transmission line.

Fig. 2. The equivalent circuit of the two LH polarized receiving dipoles and their $\lambda/2$ transmission line under a RH incident wave. Here V_i is the open circuit voltage and Z_d is the radiation impedance of each dipole.



III The New Design - a Free Space Version

The principle of design here is to follow the original design in concept but to incorporate the extra $\lambda/4$ transmission line, of the $\lambda/2$ line through the $\lambda/4$ substrate, into the two dipoles, i.e. $\lambda/8$ in each dipole.

The dipole with an extra $\lambda/8$ twin lead transmission line as shown in Fig. 3 shall be called the staggered line dipole.

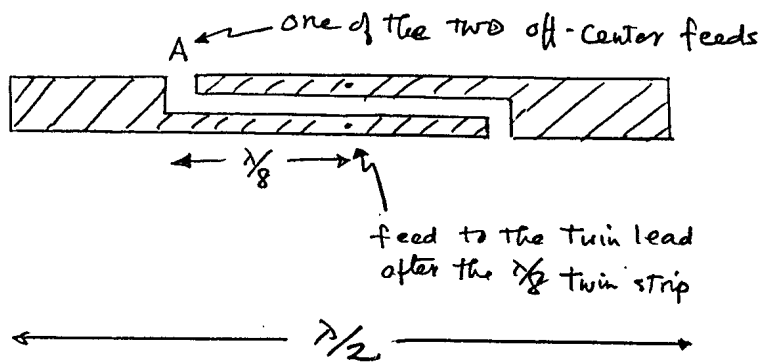
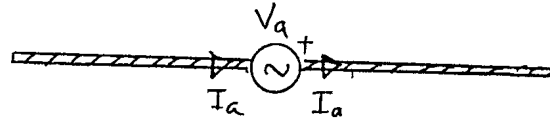


Fig.3. The staggered line dipole (in strip form for printing on substrate).

In Fig. 3 one sees two $\lambda/8$ twin leads forming a $\lambda/4$ slot. With the slot of the staggered lines the $\lambda/2$ dipole is symmetrically fed by two off-centre feeds. The principle of the design and the resulted radiation impedance can be illustrated from Fig. 4a to g. At this stage we shall assume that the dipoles and twin leads are all in free space.

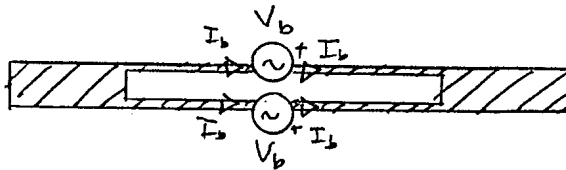
Fig.4. The Impedances of the staggered line dipole in free space.

- a) a simple dipole with one voltage source



$$Z_{ra} = \frac{V_a}{I_a} = 75\Omega$$

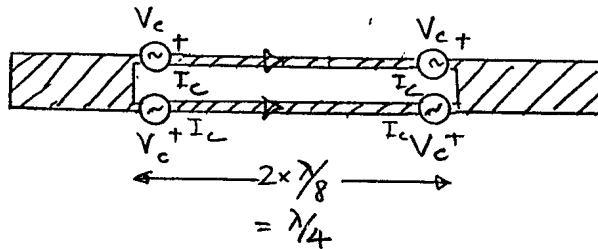
- b) a folded strip with 2 voltage sources



$$V_b = V_a, \quad I_b = I_a/2$$

$$Z_{rb} = \frac{V_b}{I_b} = 150\Omega$$

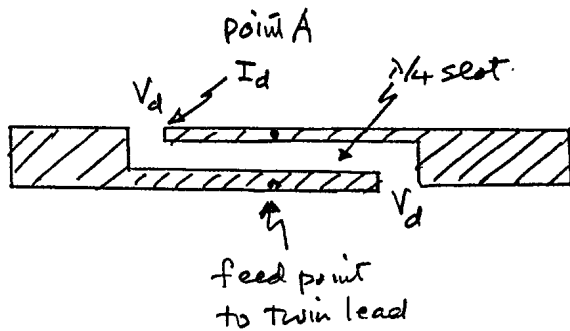
- c) a folded strip with 4 off set feeds



$$V_c = \frac{V_b}{\sqrt{2}}, \quad I_c = \frac{I_b}{\sqrt{2}}$$

$$Z_{rc} = \frac{V_c}{I_c} = 150\Omega$$

- d) the staggered line dipole with lambda/4 slot



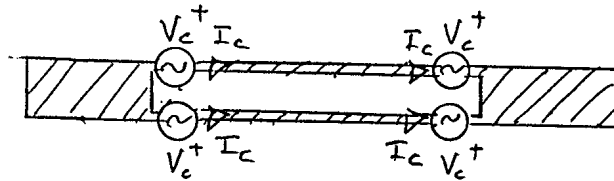
$$V_d = V_c + V_c = 2V_c$$

$$I_d|_A = I_c - I_c = 0$$

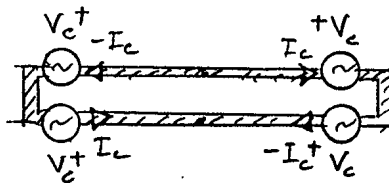
(see fig. e and f.)

⇓
Sum of e) + f)

e) radiation mode: same as c)

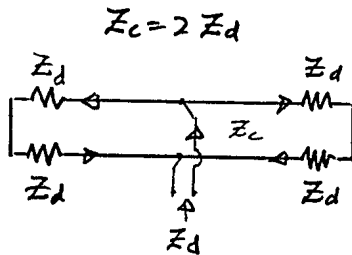


f) trans. line mode (twin strip)



$$Z_d = \frac{V_c}{I_t} = \frac{V_c}{I_c} = 150\Omega$$

g) equivalent loads Z_d



The impedance at the centre feed is therefore also $Z_d = 150\Omega$

Fig.4a shows a simple $\lambda/2$ dipole of radiation impedance of $Z_{ra} = 75\Omega$. Fig. 4b shows a folded version of the simple dipole. Unlike the regular folded dipole, there are not one but two voltage sources. For each source the radiation impedance is 150Ω , i.e. twice the 75Ω , in Fig.4a.

Fig.4c moves the two centre voltage sources sideways, $\lambda/8$ in distance, forming 4 voltage sources. Each voltage source sees a radiation impedance Z_{rc} of 150Ω . This impedance is obtained by considering the total radiated power of the dipoles in Fig.4c and 4a be equal. Then we must have

$$P = V_a I_a = 4V_c I_c$$

with

$$I_c = I_b/\sqrt{2} = I_a/(2\sqrt{2}).$$

Due to sinusoidal current distribution along the $\lambda/2$ dipole, we must have

$$V_c = V_a/\sqrt{2}$$

this means that

$$\frac{V_c}{I_c} = Z_{rc} = 2 \frac{V_a}{I_a} = 2Z_{ra} = 150\Omega$$

Fig.4d-g considers the current I_d and voltage V_d of the $\lambda/4$ staggered line dipole as the superposition of the I_c and V_c of the radiation mode and of the I_t and V_c of the transmission line mode, and gets converted into the radiated mode of Fig.4e through the boundary condition of $I_d = 0$ at the gap A of Fig.4d. The last figure, in Fig.4g, shows the equivalent load Z_d of the transmission line mode. It also indicates that the transmission line mode should have a transmission line characteristic $Z_c = 2Z_d$ and the centre feed should have an input impedance of Z_d .

With the input impedance known, the twin lead line connecting the two dipoles can easily be designed.

A word of caution is necessary here before ending this section. The impedance derivation given in Fig.4 assumes ideal conditions. That is the $\lambda/2$ dipole impedance is 75Ω in Fig.4a and b. The twin strip transmission line of Fig.4f has assumed an exact TEM wave propagating. In a practical device, the above assumption is not necessarily true. This is explained below:

- (i) The $\lambda/2$ dipole impedance can range from 72 to 80 ohms depending on the dipole width W .
- (ii) The moving of the feed point to the $\lambda/8$ position from the centre feed in Fig.4c, in principle doubles the impedance. In practice however, the resonant frequency also shifts slightly. It is found numerically that the moving of the

feed point also makes the change of the radiation resistance more rapid with respect to frequency. As a result the impedance $Z_{rc} = 150\Omega$ in Fig.4c may have a 10% error.

- (iii) The twin strip transmission line in Fig.4f assumes exact TEM wave. For a wide strip separation with a short $\lambda/8$ from the centre feed point to the slot at A , it is quite apparent that TEM wave may not be completely established. Hence the impedance Z_d of Fig.4g may not exactly be 150Ω . However, an early computation on a similar staggered line dipole in the next section, does indicate that the impedance is about 160Ω .

After such words of caution, now we have a word of encouragement. The change of Z_d occurs in both the top and the bottom dipole. Therefore, regardless of the change, the two dipoles are balanced. Since the transmission section is $\lambda/2$ connecting the two dipoles in Fig.2, the condition for scattering, of say LHCP and non scattering, of say RHCP in section II, still holds. In other words the desired property of the dipoles and the CPSS surface still holds regardless of the change in Z_d .

IV A Computer Simulation of an Early Staggered Dipole

An early version of the staggered dipole as shown in Fig.5 (completely in free space) was computed using the moment method.

The results are shown in Fig.6, 7 and 8. This work was done by a graduate student, Mr. John Haywood, in early 1988 before the project was approved. As a result, the specifications were not completely firmed and the design and results were not completely appropriate.

Since the design and results were not completely appropriate, there is little sense of discussing the details of Fig.6, 7 and 8 other than showing that the preliminary work was done.

In spite of this, the computations did indicate that we have been in the correct path and enabled us to come to the design of Fig.3 and the impedance considerations in Fig.4.

Fig.5 LH polarization (reflection) dipole pair
(staggered line dipole)

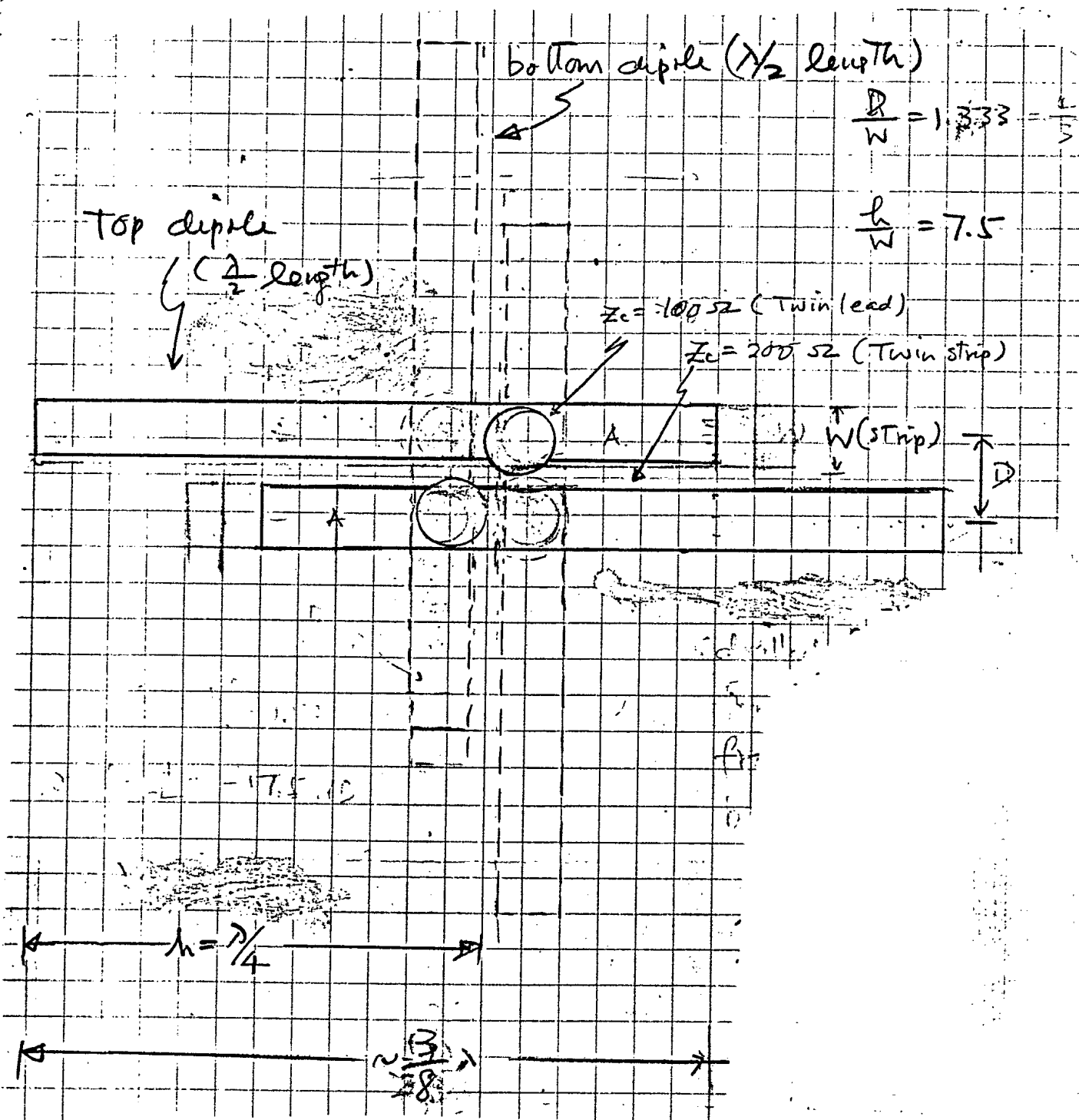


Fig.6 The resistance of a single staggered line dipole

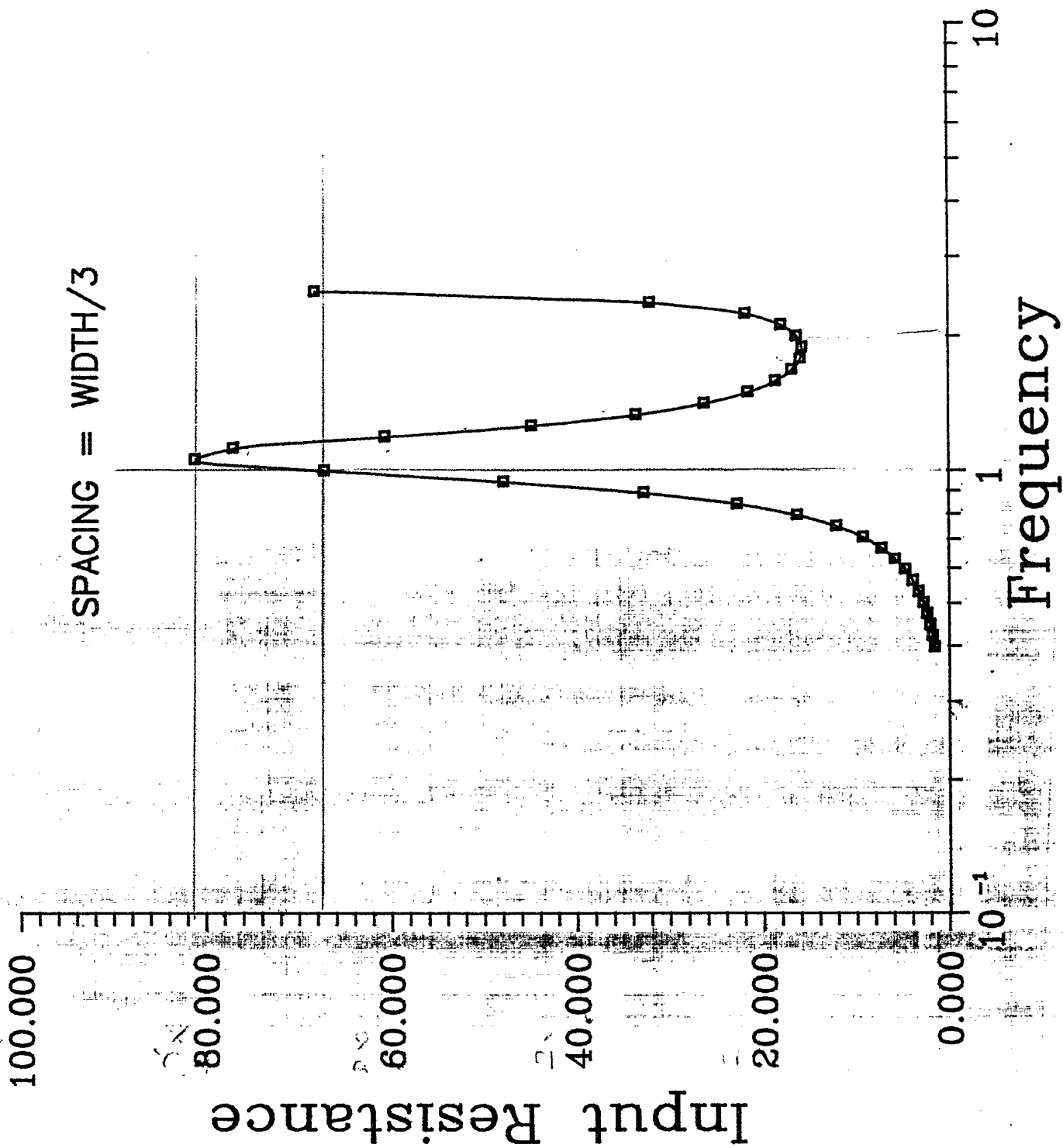


Fig.7 The input reactance of a single staggered line dipole with $D/W = 1.333$.

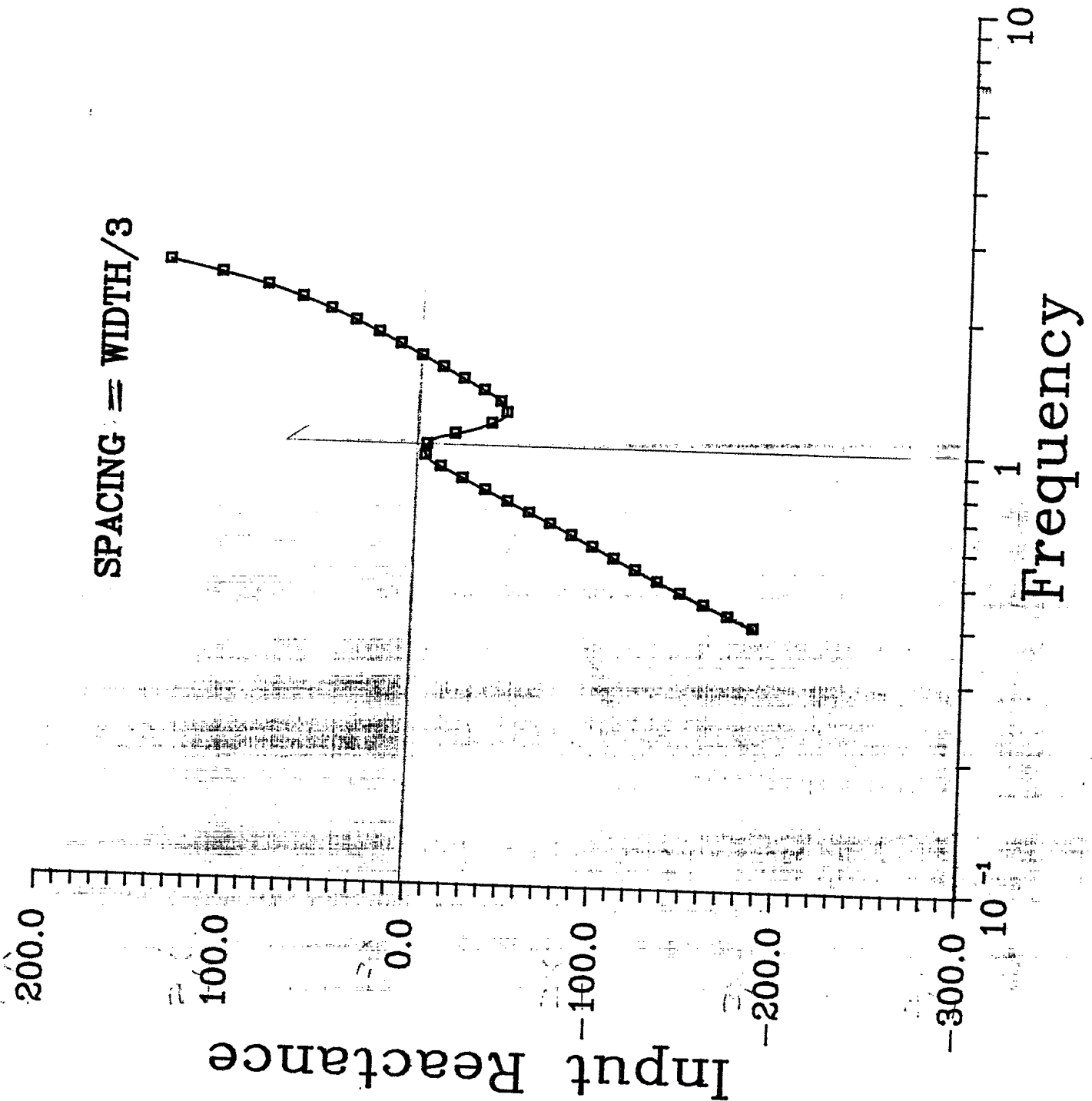
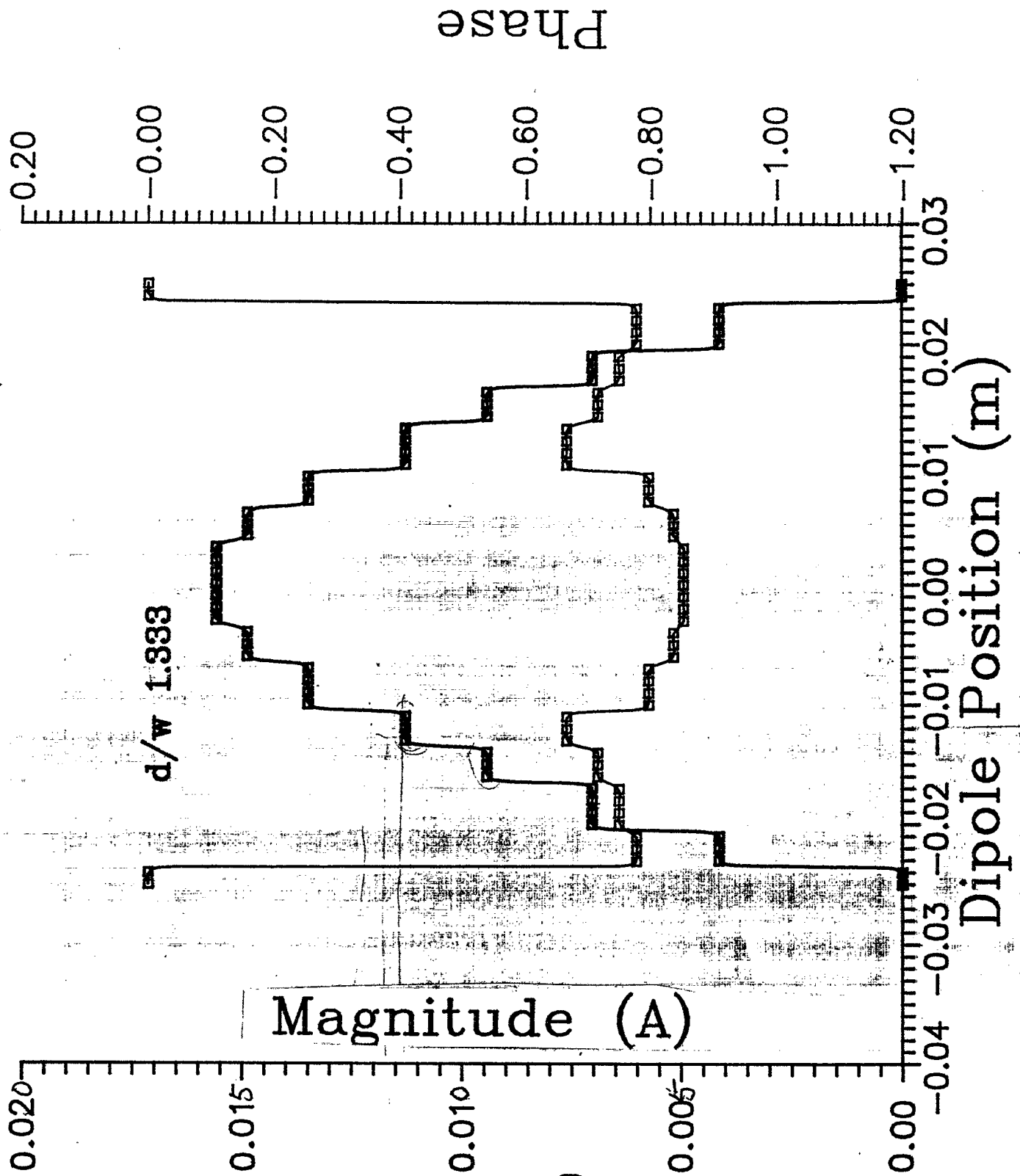


Fig.8 The current distribution of the radiation mode of a staggered line dipole of Fig.5.



V The Effect of the Dielectric Substrate

The dipoles with $\lambda/4$ slot are not actually in free space but are printed on a $\lambda/4$ substrate of dielectric constant of 2.3. Since the thickness of $\lambda/4$ places one dipole (say on top of the substrate), in the radiation zone of the other dipole, (say at the bottom of the substrate), this means that as far as the inductive field is concerned, each dipole is sitting on a (semi-infinite) dielectric half space. The effective dielectric constant for a radiating source on the half space is the average of that dielectric constants of the half space and air, i.e. $\epsilon_{eff} = (\epsilon_r + 1)/2$. Thus the wave length of the slotted dipole is

$$\lambda = \lambda_0 / \sqrt{\epsilon_{eff}} \quad (1)$$

where λ_0 is the free space wave length.

The radiation impedances in Fig.4a to e may be similarly changed to

$$Z = Z_0 / \sqrt{\epsilon_{eff}} \quad (2)$$

where Z_0 is the original impedance in free space. Eq.(2) may have some error in its real part, the resistance, since the resistance represents the radiation field. This error however, does not matter because of the total transmission line length of $\lambda/2$ as discussed at the last paragraph in Section III.

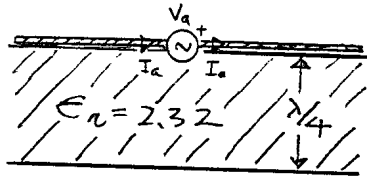
The total transmission line length has 2 $\lambda/8$ lengths formed by the slot of the dipole and a $\lambda/4$ twin lead connecting the two dipoles. This $\lambda/4$ twin lead (of circular wires) is completely inside the substrate, therefore, it has a wavelength the same as a plane wave in the substrate.

The characteristic impedances of this twin lead is that of a similar twin lead in free space divided by $\sqrt{\epsilon_r}$, when ϵ_r is the dielectric constant of the substrate.

Based on the above consideration of dielectric substrate, the impedances of the staggered line dipoles are calculated. Following the layout of Fig.4 for free space. Fig.9 gives the layout of impedance with a $\lambda/4$ dielectric substrate.

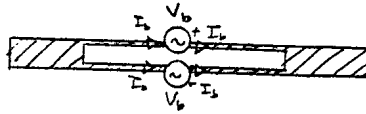
Fig.9. The impedance of the staggered line dipole printed on a $\lambda/4$ dielectric slab of $\epsilon_r = 2.32$ (to be compared to the corresponding parts a), b) etc. in Fig. 4).

a) single dipole



$$Z_{r1} = Z_{ro} \frac{1}{\sqrt{(\epsilon_r + 1)/2}} = 58.12\Omega$$

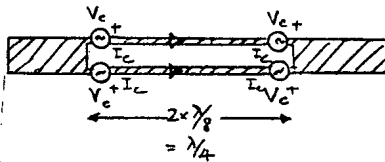
b) a folded dipole with 2 voltage source



$$V_b = V_a \quad I_b = I_a/2$$

$$Z_{rb} = \frac{V_b}{I_b} = 116.25\Omega$$

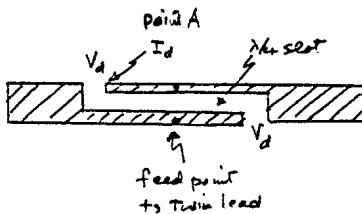
c) a folded dipole with off-set feeds



$$V_c = \frac{V_b}{\sqrt{2}}, \quad I_c = \frac{I_b}{\sqrt{2}}$$

$$Z_{rc} = \frac{V_c}{I_c} = 116.25\Omega$$

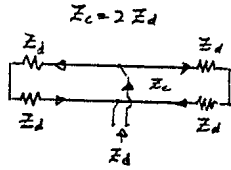
d) the staggered line dipole with $\lambda/4$ slot



$$V_d = 2V_c$$

$$I_d = 0$$

g) the equivalent load Z_d



$$Z_d = \frac{V_c}{I_c} = 116.25\Omega$$

$$Z_c = 232.5\Omega$$

VI The element Separating and Coupling of the Array Forming the Circular Polarization Sensitive Surface (CPSS)

It has been known and experimentally demonstrated by Til-tek Inc., in 1984, that a square grid of $\lambda/2$ separations of dipole elements approximates a perfect reflecting surface at the designed frequency of the dipoles. At this separation, there is no grating effect. Therefore, this is the grid separation we have chosen for this new CPSS surface.

There is some coupling between dipole elements in the array. The coupling changes the radiation impedance Z_r of each element. However, all elements are charged by the same amount. Therefore, the two dipoles in each linked pair on the top and bottom of the substrate are still balanced in excitation. This balance means the axial ratio of the reflected circularly polarized wave viewers to be very good despite the coupling. This maintenance of circular polarization has been mentioned in section III.

The dipole is $\lambda/2$ long and the grid work is also $\lambda/2$ in separation. There is a chance that the dipoles may touch and short. It is, therefore, recommended that the dipoles be rotated 45° away from the grid lines. In this case there is no worry on shorting the dipoles.

The wavelength λ for the grid separation is taken to be $\lambda_o / \sqrt{(\epsilon_r + 1)/2}$ when λ_o is the free space wavelength. This λ is a little shorter than λ_o and ensures that the surface approximates a solid surface. It may increase the coupling a little. However, as mentioned before, the coupling does not affect the circular polarization.

Based on the above, the array of staggered line antennas is designed in Fig. 10.

Top side of
12 x 12 crossed dipoles

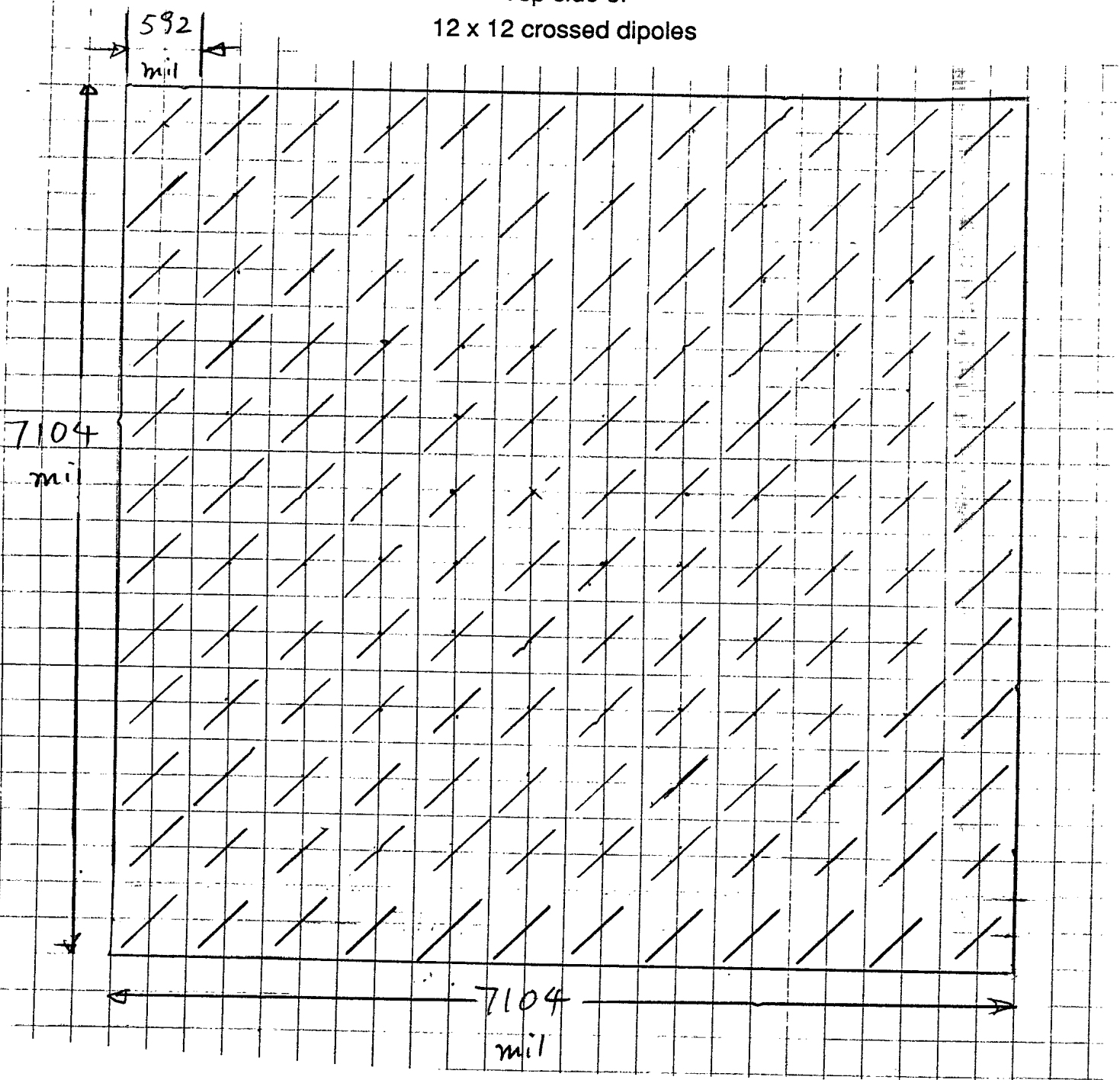


Fig.10. The 144 staggered line array on the top surface of the $\lambda/4$ substrate for 7.75 GHz. Substrate thickness is 1". The dipoles on the bottom surface are at right angle to those on top. The design for 15.5 GHz is half in size.

VII The Staggered Line Dipole Designs

Before the designs it is necessary to list the formula used.

1. The substrate thickness $t = \lambda/4$, for greater length twin lead transmission line between the two staggered line dipoles.

The centre frequency:

$$f = \frac{3 \times 10^8 \text{ m/sec}}{\lambda_o}$$

where in the free space

$$\lambda_o = \lambda \times \sqrt{\epsilon_r}$$

Therefore,

$$f = \frac{3 \times 10^8 \text{ m/sec}}{4t\sqrt{\epsilon_r}} \quad (i)$$

2. The dipole wavelength. Since the dipole is effective on the dielectric half space as far as the induction field is concerned. Therefore, the dipole length at centre frequency is

$$\frac{\lambda}{2} = \left(\frac{\lambda_o}{2} \right) / \sqrt{(\epsilon_r + 1)/2} \quad (ii)$$

3. The twin strip characteristic impedance of the staggered line dipole

$$Z_d = \frac{120}{\sqrt{(\epsilon_r + 1)/2}} \ln \frac{4D_w}{W} \quad (iii)$$

where W = strip width = 4 x effective radius of the dipole wire.

D_w = centre to centre separation of the twin strip.

The effective dielectric constant is $(\epsilon_r + 1)/2$ for the dipole in (ii) and (iii).

4. The twin lead characteristic impedance (between 2 dipoles)

$$z_c = \frac{120}{\sqrt{\epsilon_r}} \ln \frac{D}{r} \quad (iv)$$

where D is the centre to centre separation.

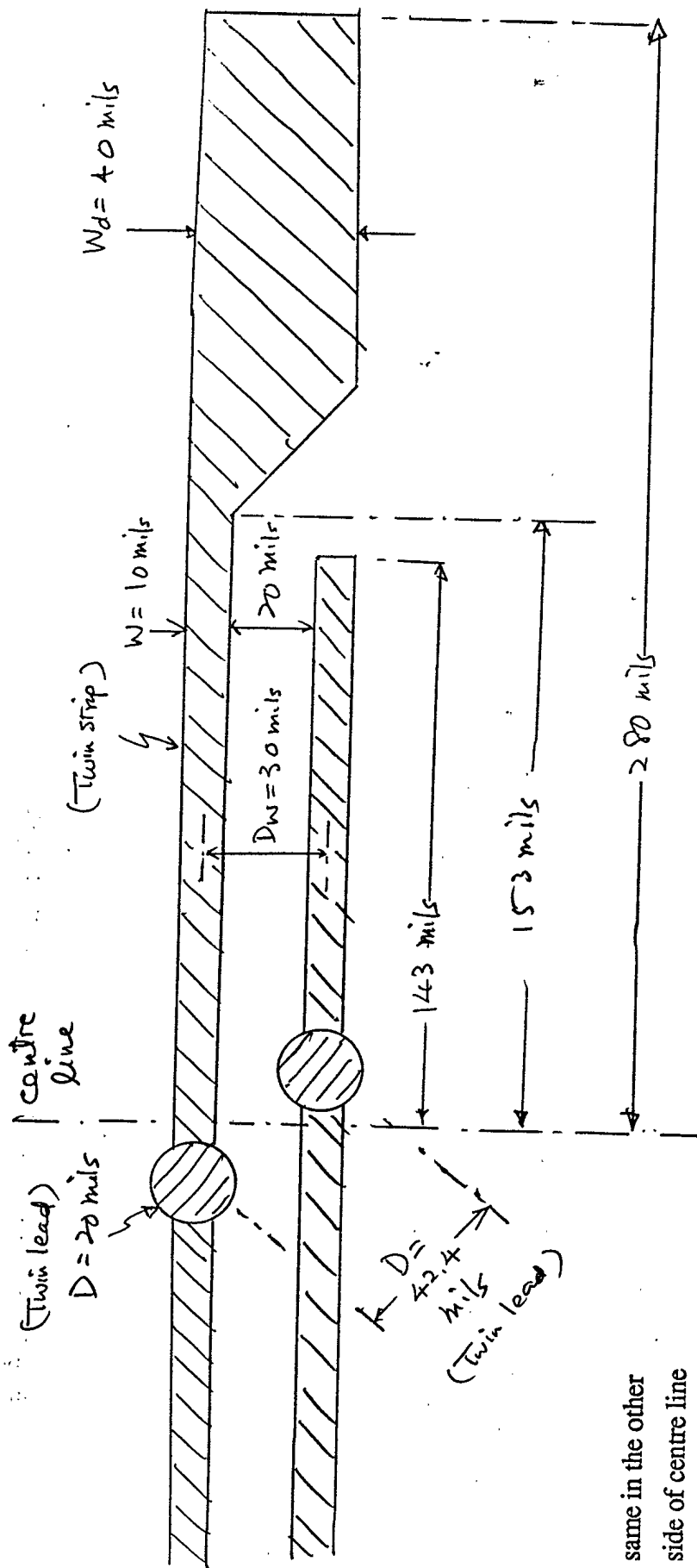
r is the radius of the twin lead.

ϵ_r is the dielectric constant in which the twin lead threads through.

The substrate thickness is either 1" or 1/2" from the available stock of duroid. The dielectric constant of duroid is 2.32. Based on these substrate parameters and the formulas

(i) to (iv), two CPSS surfaces are designed and constructed according to Fig.10 for the array layout and Fig.11 and 12 for the dipoles. Fig.11 is for the 7.75 GHz design for a substrate of thickness 1" and Fig.12 is for the 15.5 GHz design for the substrate thickness of 1/2".

Fig.11 assumes the dipole is $0.95 \lambda/2$. Fig.12 assumes the dipole is $\lambda/2$. The designs are restricted by the fact that only 20 mil holes can be drilled for the twin leads and they cannot be smaller.



$\lambda/2 = 592 \text{ mil}$
 $\lambda/4 = 296 \text{ mil}$
 $0.95 \times \lambda/4 = 280 \text{ mil}$
 $\lambda/8 = 148 \text{ mil}$

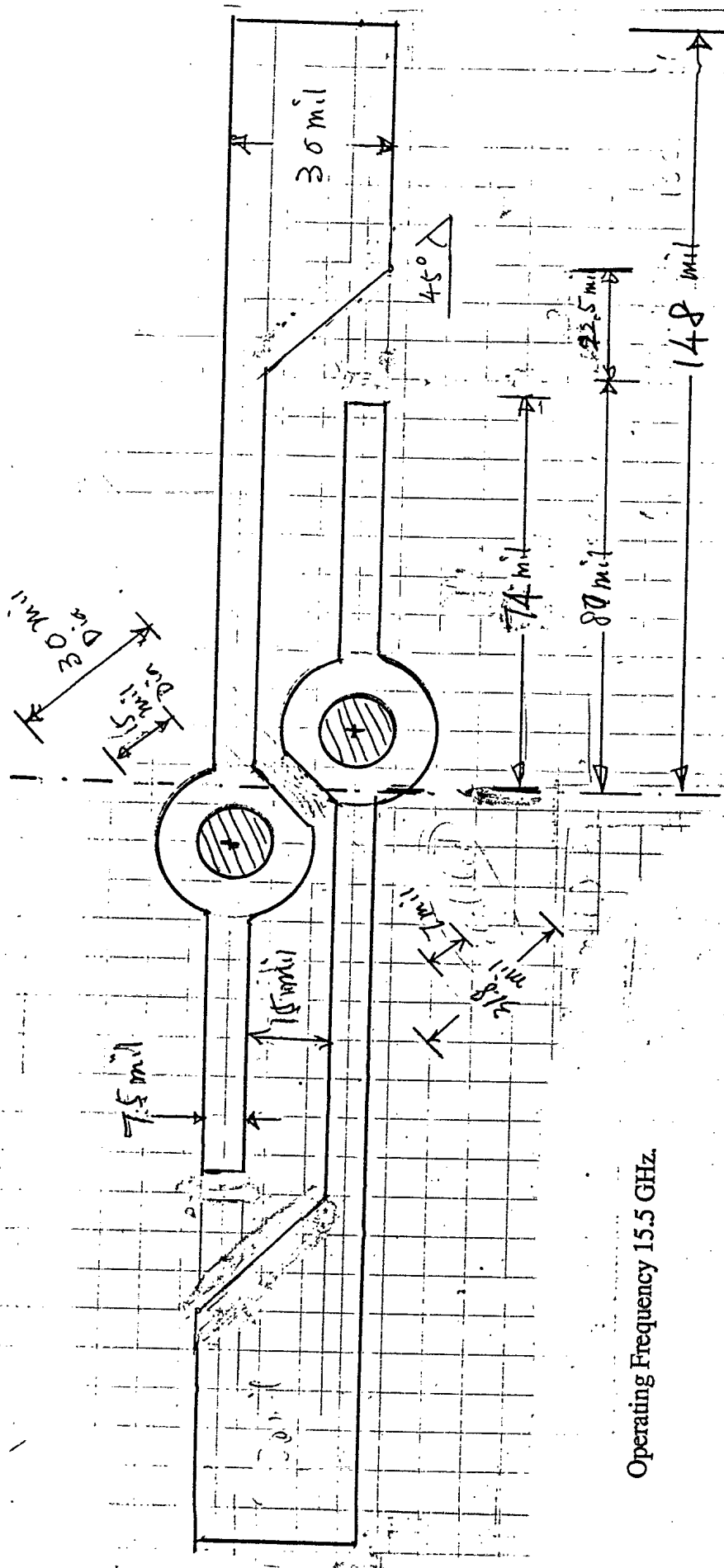
$$Z_c = \frac{120}{\sqrt{\epsilon_{eff}}} \ln \frac{4D_w}{W} \quad (\text{twinstrip})$$

$$= \frac{120}{\sqrt{1.66}} \ln \frac{4 \times 30 \text{ mil}}{10 \text{ mil}} = 231 \Omega$$

$$Z_d = \frac{120}{\sqrt{\epsilon_r}} \ln \frac{D}{r} = \frac{120}{\sqrt{2.32}} \ln \frac{42.4 \text{ mil}}{10 \text{ mil}} = 113.45 \Omega \quad (\text{twinlead})$$

Fig. 11. The detailed dipole structure (LH reflection) for the 7.75 GHz design.

Revised design for a 15.5 GHz dipole of the circular polarization selective surface, September 13, 1989.



Operating Frequency 15.5 GHz.

Fig. 12. The detailed dipole structure (LH reflection) for the 15.5 GHz design.

VIII The Interpretation of the Measurements

The measurements are thoroughly covered by the Til-tek report [2]. It is not necessary to cover them here. For interpretation however, two figures are included here.

The transmission of the CPSS is measured. If the surface reflects, say the LH polarization, at a certain frequency, the transmission is greatly attenuated there. If the surface does not reflect, the signal is unaffected, i.e. uniform transmission throughout the frequency range.

Fig. 13 shows the transmission of the LH polarization for the 7.75 GHz surface. It is observed there that the resonant frequency is at 8.25 GHz which is 6% above the design frequency. This is interpreted as that a mistake was made to design the dipole to be $0.95 \lambda/2$ long, since the numerical solutions of section IV shows that the resonant frequency should be exactly at $\lambda/2$ for the staggered line dipole.

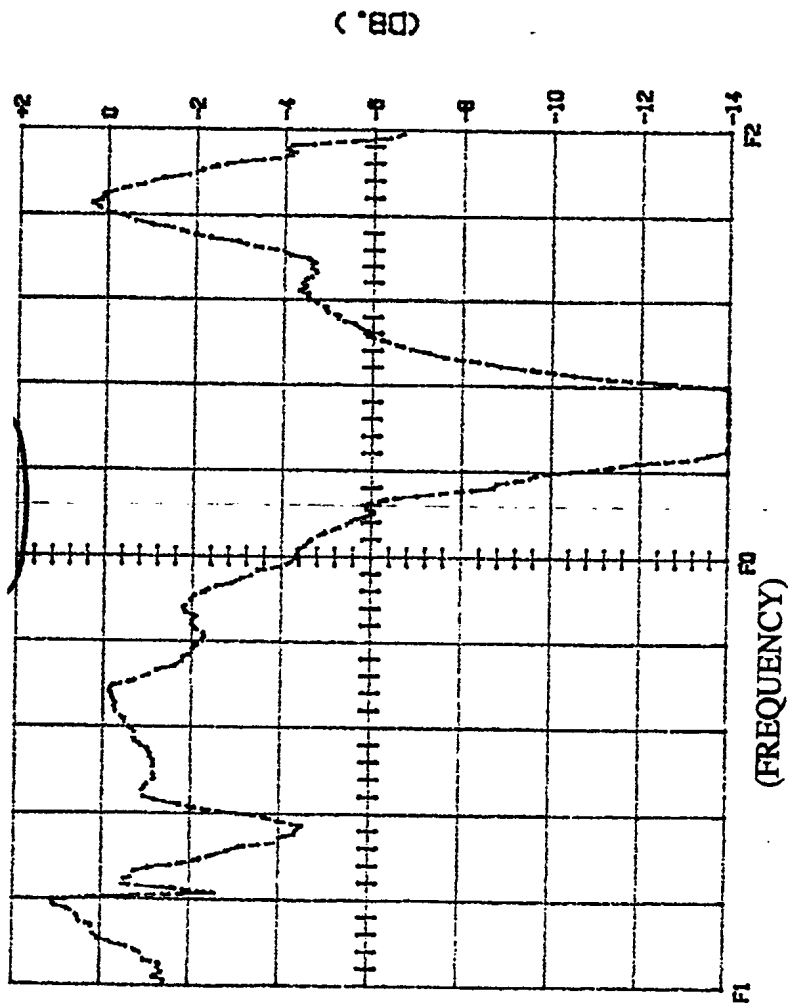
This mistake is corrected in Fig.12 for the dipoles of 15.5 GHz. Fig.14 shows the transmission again. The frequency of resonance in this case is right in 15.8 GHz, only about 2% above the designed 15.5 GHz.

Model #: PS #1 NO DIELECTRIC COVER
A: CHANNEL OFF
B: TRANSMISSION LOSS 2 DB/DIV. N.L. = -57.49 DB.

DATE: 26-+7-89
SERIAL #: N/A
TESTED BY: D.A.T.
APPROVED:

NORMAL INCIDENCE

RETURN LOSS: _____
TRANS. LOSS: _____



F0: 7500 MHZ.
F1: 5000 MHZ.
F2: 10000 MHZ.

M1:
M2:

POWER LEVEL: 0 DBM.

0° incidence

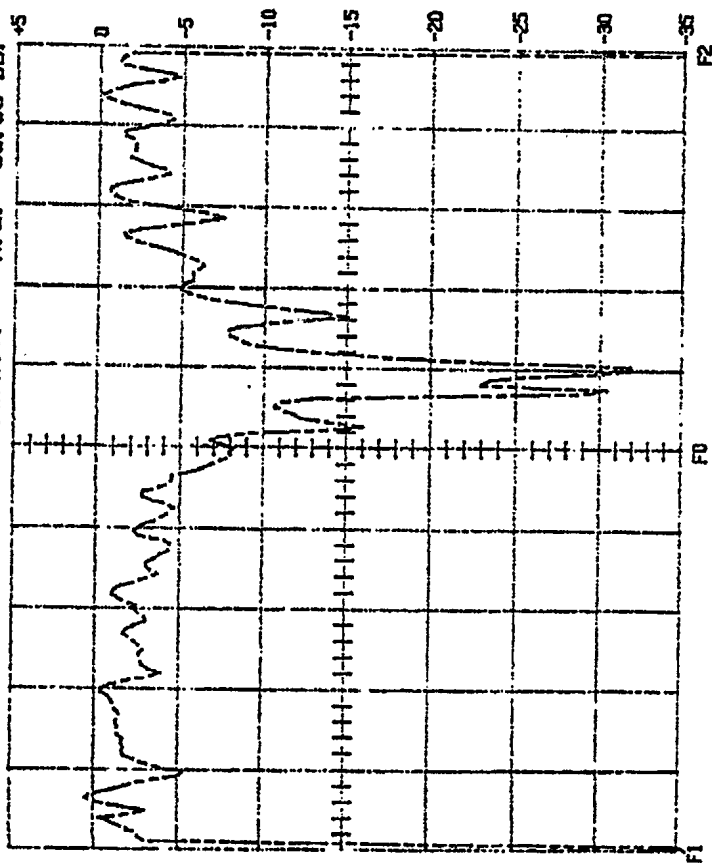
Fig. 13. The transmission measurement of the 7.75 GHz surface. Resonance at 8.25 GHz.

MODEL #: LHCP ATTENUATION DUE TO PSS 0 DEG.

A: CHANNEL OFF

B: TRANSMISSION LOSS 5 DB/DIV. N.L. --56.80 DB.

DATE: 20-10-89
 SERIAL #: N/A
 TESTED BY: D.A.T.
 APPROVED:



(89)

RETURN LOSS: -----
 TRANS. LOSS: -----

(FREQUENCY)

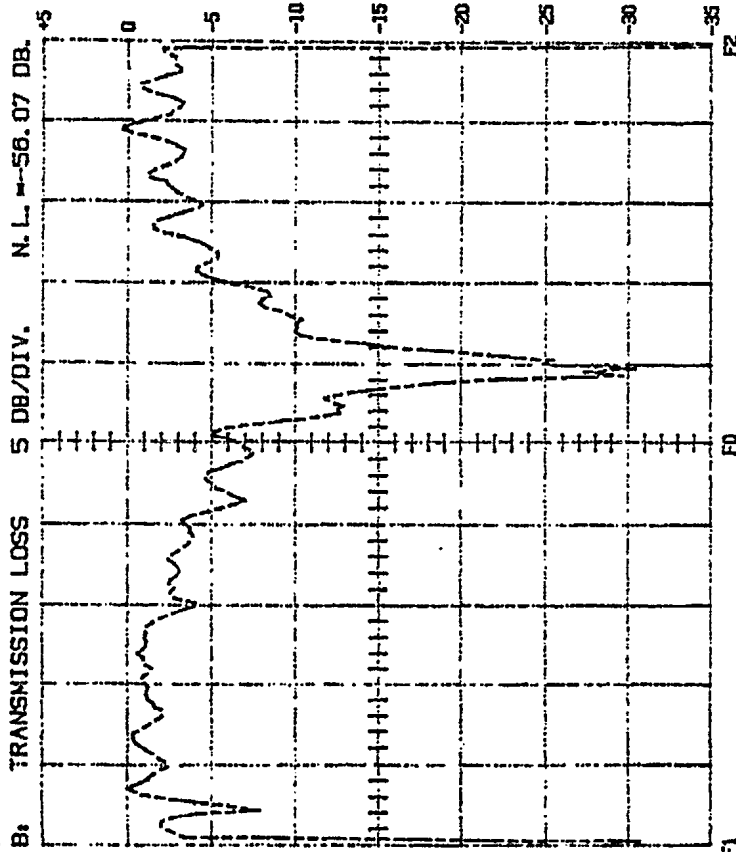
F0 : 15000 MHZ. M1 :
 F1 : 10000 MHZ. M2 :
 F2 : 20000 MHZ.

POWER LEVEL : 11 DBM.

Fig. 14 (b) 0° incidence

MODEL #: LHCP ATTENUATION DUE TO PSS 15 DEG.

A: CHANNEL OFF



N.L. = -56.07 DB

DATE: 20-10-89
 SERIAL #: N/A
 TESTED BY: D. A. T.
 APPROVED:

(89)

RETURN LOSS: -----
 TRANS. LOSS: -----

(FREQUENCY)

F0 : 15000 MHZ. M1 :
 F1 : 10000 MHZ. M2 :
 F2 : 20000 MHZ.

POWER LEVEL : 11 DBM.

(A) 15° incidence

Fig. 14. The transmission measurement of the 15.5 GHz surface.
 Resonance at 15.8 GHz.

IX Conclusion

The staggered line dipole was developed in the following sequence:

- (i) The computation of an early dipole.
- (ii) The derivation of impedances of the staggered line dipole in Fig.4.
- (iii) The considerations of the element separations, coupling and the effect of the dielectric substrate.
- (iv) The design as limited by the minimum hole size to form the twin lead connecting the dipoles. The minimum hole size is 20 mil at the present time.

As one sees in Fig. 11 and 12, one needs to have a flange for soldering the twin lead to the twin strip of the staggered line dipole. The larger is the minimum hole size, the larger is the minimum flange size. The flange disrupts the smoothness of the dipole and causes cross polarizations.

One sees in Fig. 12, the flange size in the 15.5 GHz design is quite large. Fortunately, despite such large flange, the measurement of Til-tek in Fig. 14 shows that the cross polarization is still over -30dB. This is quite acceptable since the contract specification is only -20dB.

The one draw-back, still in the design, is that the bandwidth of the maximum attenuation is quite narrow. Fig. 13 and 14 show that the bandwidth at -20dB is only 2%. This however, is acceptable in many applications.

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Dr. W.V. Tilston invented the principle of the circularly polarized selective surface (CPSS) in 1986. The surface is a large planar array of crossed dipoles. Each crossed dipole pair is separated by $\lambda/4$ distance but is connected between the two feed points by a $\lambda/2$ transmission line. Dr. Tilston constructed a CPSS surface with incomplete success since he had difficulty in fitting the $\lambda/2$ transmission line length to the $\lambda/4$ separation of the dipole pair.

Loading of the transmission line by high dielectric constant material to relieve such difficulty is ruled out because of the construction difficulty and expenses involved for a large array of over 1000 dipole pairs. The purpose of this contract is to give a new cross dipole design to overcome such difficulty without the above loading of transmission line.

This design has been done and the two left-hand CPSS surfaces constructed has been measured and found to give the desired polarization selective property. The construction and measurement were done by Til-tek Inc. under a separate contract from DREO (W7714-8-5651/01-SV).

This report gives the details of the design and the underlying principle. Briefly, the principle is to take the transmission line length to be $\lambda/4$, instead of $\lambda/2$, but to embed the remaining lengths of transmission line in the dipole pair, that is: $\lambda/8$ in each dipole.

The dipole array is printed on a dielectric slab of $\lambda/4$ in thickness for mechanical support. The effect of the dielectric slab is accounted for in the design. The design was constructed for two frequencies: 7.75 and 15 GHz.

The experimental measurements proved the validity of the design.

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