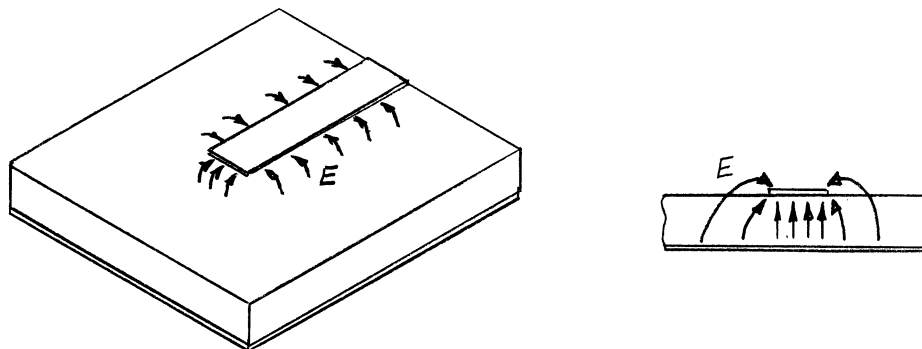


MICROSTRIP ANTENNAS

Microstrip antennas are narrowband antennas which can be conformally mounted on the sides of vehicles and provide an antenna with minimal impact on the vehicle. For this reason they have become important antennas. Second these antennas can be fabricated using printed circuit etching techniques which means they are cheap to produce. Third non-radiating microstrip circuitry such as power dividers and phase shifters can be mounted between the radiating elements and provide the feed circuit. Because of these reasons many people are working on these antennas today.

We will use a transmission line model to describe the microstrip patch antenna. On page 495 is a diagram of the patch antenna. In this figure the antenna is the rectangular patch which is fed by a microstrip transmission line. We will call the side with the feed line the width, W , of the patch and the other side the length, L . In general the input impedance of the antenna will not be matched to the microstrip transmission line so that there is a quarterwave transformer connecting the antenna to the transmission as shown in the figure. The radiator is a thin foil, usually copper, which is bonded to the dielectric substrate. On the other side of the dielectric slab there is another metal foil bonded to the substrate to form a ground plane under the antenna and the transmission line.

Consider an open circuited microstrip transmission line as shown below.

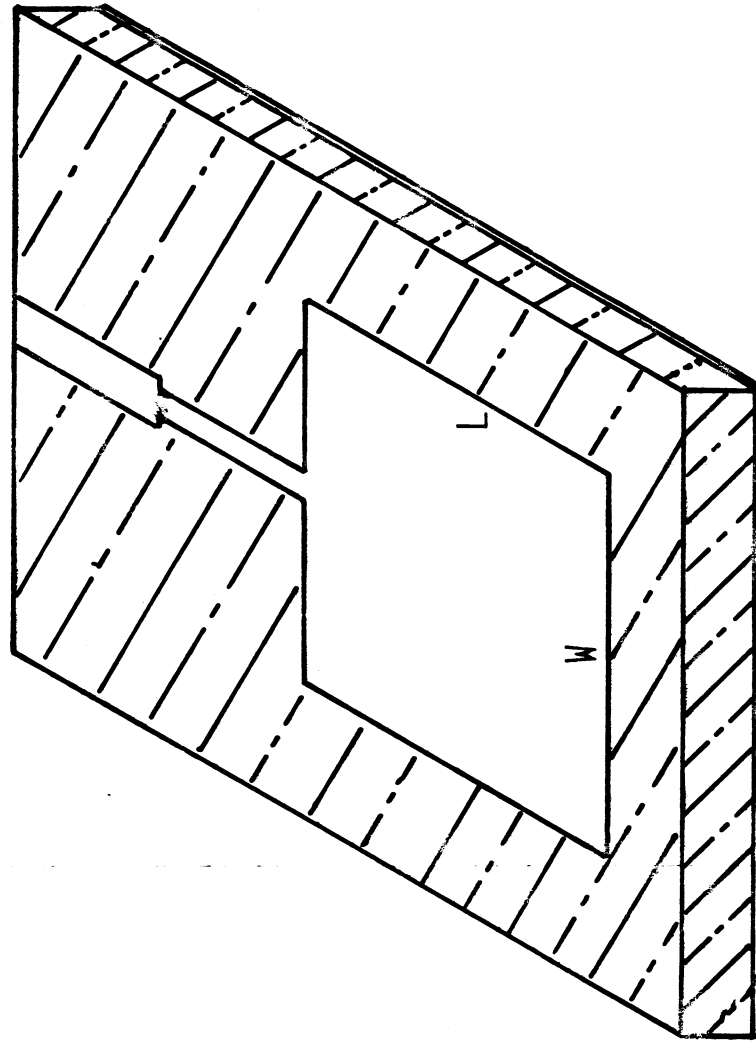


The microstrip line has fringing fields along the edges of the line to the ground plane below as shown. These fields do not radiate because there are equal and opposite fields on the two sides of the line and they are close together in wavelengths. But consider the open circuit. It does not have antisymmetric fields to cancel its radiation from the fringing fields. As long as the width of the microstrip line is small, the radiation conductance is small and little power is delivered to the radiator; most of it is reflected.

The microstrip circuit does not see an open circuit at the end of the line, but a fringing capacitance to ground. The microwave circuit designer handles this problem by decreasing the length of the open circuited stub to account for this shunt capacitance. Hammerstad¹ has found a formula for this length in

Hammerstad, E. O., "Equations for Microwave Circuit Design", Proc. 5th European Microstrip Conference, Hamburg, Sept. 1975, pp. 268-272.

MICROSTRIP PATCH ANTENNA



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terms of the thickness of the substrate, H , the width of the line, W , and the effective dielectric constant.

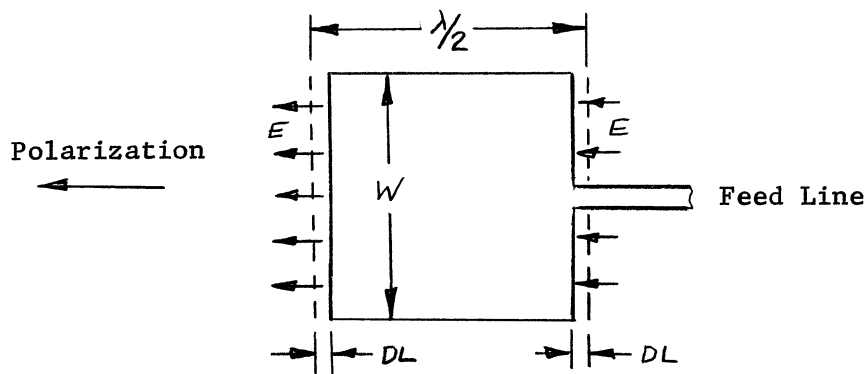
$$DL/H = 0.412 ((\epsilon_{\text{eff}} + 0.3)/(\epsilon_{\text{eff}} - .258)) ((W/H + .262)/(W/H + .813))$$

with the effective dielectric constant of the microstrip line given by

$$\epsilon_{\text{eff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{10 H}{W} \right)^{-\frac{1}{2}}$$

where ϵ_r is the dielectric constant of the substrate material. The effective dielectric constant of a microstrip transmission line is a function of the width (or impedance) of the line.

Consider a microstrip patch which has an effective length of one half wavelength.

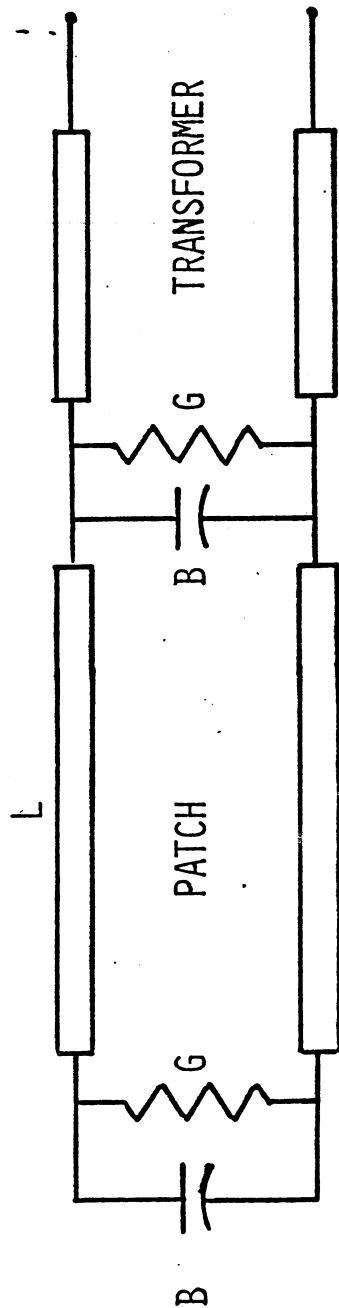


The fringing fields from the ground to the patch along the width of the patch are shown referenced to the input phase. Because the effective length of the patch is a half wavelength, the phase on the edge opposite the feed has changed sign as shown. These edges will radiate as slots since we can equate the electric field of the fringing field to a magnetic current. The edges form a two element array. The fields are in phase which gives a pattern with the maximum normal to the substrate. The slots are formed between the ground plane and the patch edge and are excited by the fringing fields.

The patch can be considered a microstrip transmission line with a low impedance. The effective dielectric constant of this transmission line is given by the formula above using the width, W , of the patch. The length, L , of the patch will be less than a half wavelength by $2 DL$, the effective length of the fringing capacitance on both ends. This length will give a resonant circuit, zero susceptance or reactance, at the feed input. We can represent this by the transmission line model on page 497.

The radiation is represented as a conductance, G , in the circuit; the fringing capacitance as a susceptance, B , and the patch as a microstrip transmission line of impedance Z with an effective dielectric constant of ϵ_{eff} . The length of the patch is less than a half wavelength so that the susceptance, B , of the further side slot from the feed will be transformed to a negative susceptance and cancel the slot susceptance at the input. The conductances

TRANSMISSION LINE MODEL OF PATCH ANTENNA



$$DL = .412 H (EFF + .3) (W/H + .264) / ((EFF - .258) (W/H + .8))$$

$$B = 2\pi DL \sqrt{EFF} / (\lambda Z)$$

$$L = \lambda / (2 \sqrt{EFF}) - 2 DL$$

$$G = W / (120\lambda)$$

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will add at the feed. If the patch length, L , was a half wavelength, then the transformed susceptance would retain the same sign and add to the shunt susceptance at the input. Any impedance (admittance) transformed through a half wavelength transmission line remains the same. The radiation conductance of the edge of the patch can be found from Harrington's formula for the parallel plate slot found on page 188.

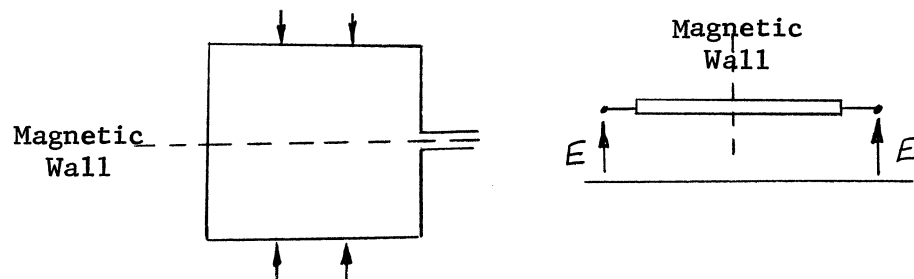
$$G = \frac{\pi W}{\eta \lambda_0} \left(1 - \frac{(\beta H)^2}{24} \right) \quad \eta = 377$$

The shunt susceptance can be found from the length DL .

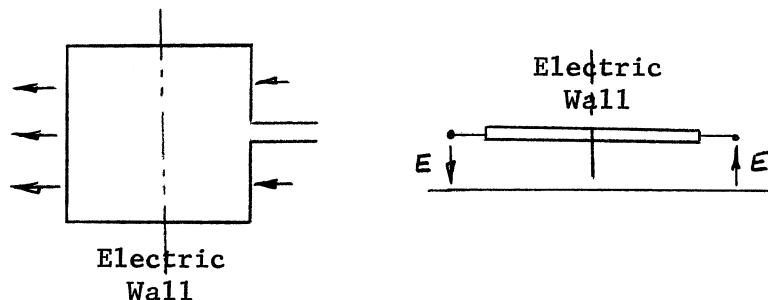
$$B = 0.01668 (DL/H) (W/\lambda_0) \epsilon_{\text{eff}}$$

We will use this transmission model to calculate the impedance bandwidth of the antenna. The antenna radiates power only because it is matched to the input transmission line and the energy is not reflected at the input.

Suppose we have a square patch. Why does not the antenna radiate from the other two edges? We can equally say that the patch is a transmission line in the other direction. The transmission line is the same length to the other two edges and we can expect equal phase fringing fields to the ground plane on both sides.



Let us look at this from even and odd mode analysis. We will have a magnetic wall or virtual open circuit in the plane containing the centered feed line. The input is a shunt combination of the transmission line to the opposite side and another pair to the other two sides. The transmission lines to the non-radiating sides look like an open circuit to the feed line and receives no power. We can also expand the fields in the odd mode to the radiating edges.



This mode will give an electric wall or virtual short circuit half way through the patch. This is the normal radiating mode. We can put a shorting pin in the center of the patch to reinforce this virtual short with no effect on

the input impedance or pattern. The shorting pin will give the antenna some lightning protection.

Example. Design a square microstrip patch antenna at 3 GHz on a 1/16 inch dielectric substrate, dielectric constant = 2.5 .

The patch will be approximately a half wavelength long in the dielectric so we can assume at first that the width is $\lambda/2$ in the dielectric.

$$W = \frac{c}{2 f \sqrt{\epsilon_r}} = 1.244 \text{ inches}$$

Using this width we can find the effective dielectric constant of the microstrip transmission line between the radiating edges (slots).

$$\epsilon_{\text{eff}} = \frac{3.5}{2} + \frac{1.5}{2 \left(1 + \frac{.62}{1.244} \right)^{\frac{1}{2}}} = 2.363$$

We will use this effective dielectric constant in the formula for the fringing capacitance effective length,

$$W/H = 20.06 \quad DL = 0.031$$

The length of the patch is found using this cut back on both edges.

$$L = \frac{c}{2 f \sqrt{\epsilon_{\text{eff}}}} - 2 DL = 1.280 - .062 = 1.218 \text{ inches}$$

We can use this length for the width of the patch and recalculate the effective dielectric constant.

$$\epsilon_{\text{eff}} = 2.361$$

This gives us the same length as the first iteration. The input conductance of the patch which is fed on the edge will be twice the conductance of one of the edge slots.

$$G = \frac{1.218}{120 (3.934)} = .00258$$

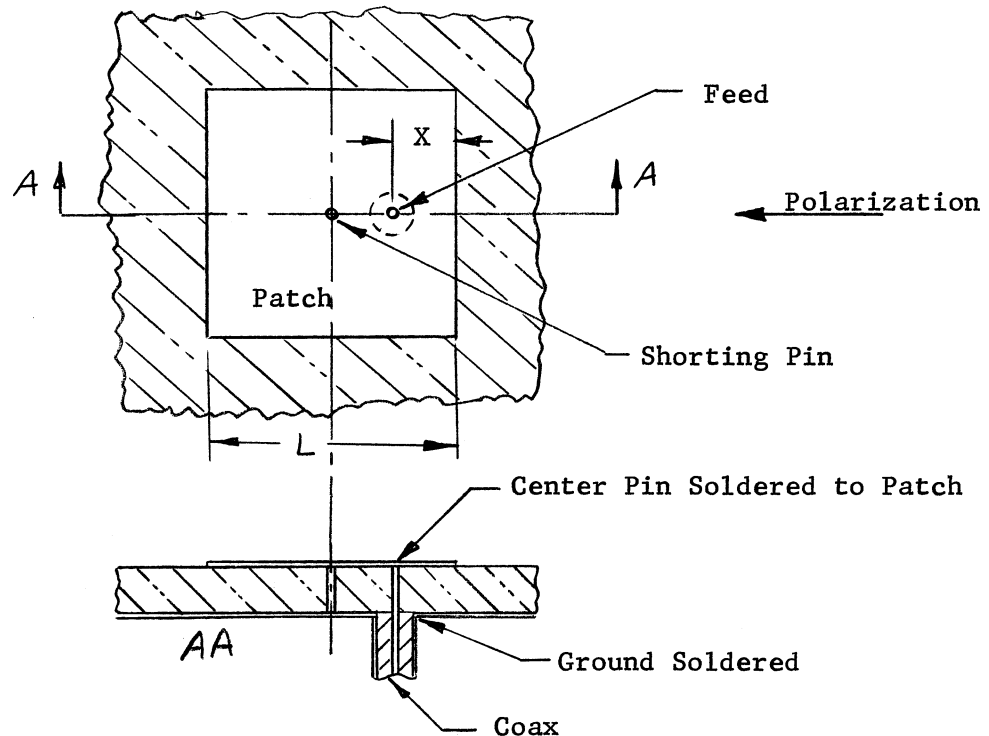
$$R = 1/(2 G) = 194 \text{ ohms}$$

The usual transmission line used on microstrip antennas is 100 ohms. We need a transformer to decrease the input impedance to the transmission line. The impedance of the transformer is given by

$$Z_T = (100 \cdot 194)^{\frac{1}{2}} = 139 \text{ ohms}$$

The microstrip patch can also be analyzed as a resonant cavity. The walls are impedance walls along the edges instead of metal walls. Using this method of analysis we come to the conclusion that the bandwidth is almost independent of the feed point when the impedance is transformed to the same value. The feed is just exciting a lossy cavity. We are lead to this analysis because the antenna is quite narrowband which says the antenna is a high Q circuit.

If the antenna is fed internal to the edge, lower values of input resistance will be found. The antenna has a virtual short at the plane half-way between the radiating slots. If we feed the antenna at that point, we can expect the input impedance to be zero at the resonant frequency. As we move the feed toward the edge the input impedance will increase until the impedance at the edge is obtained. Below is a diagram of the patch fed from underneath by a coax.



The impedance at the feed is given by the following approximate formula.

$$R_i = R_e \cos^2\left(\frac{\pi X}{L}\right) \quad 0 < X < L/2$$

The distance X is measured from the edge of the patch and L is the length of the patch. R_e is the input impedance at the edge of the patch. Using this formula we can find the feed point for the coax for a given input impedance.

$$X = \frac{L}{\pi} \cos^{-1} \sqrt{\frac{R_i}{R_e}}$$

Find the feed point in the example above for a 50 ohm input.

$$X = \frac{1.218}{180} \cos^{-1} (50/194)^{\frac{1}{2}} = 0.403$$

This distance is measured from the edge of the patch. The round rod across the gap of the microstrip patch will be a series inductor at the input.

The bandwidth of the patch antenna can be found from the transmission line model. The input impedance can be calculated over a band of frequencies and

the VSWR bandedges can be identified. The impedance is plotted on a Smith chart on page 502. Off resonance the antenna looks like a small inductor. The size of the circle increases as the input impedance increases. If the antenna is fed from below, then the circle diameter increases as the feed point is moved toward the edge of the patch. On page 503 is a plot of the 2:1 VSWR bandwidth of a square microstrip patch antenna on a woven teflon fiberglass substrate for various thicknesses of the substrate. The percent bandwidth increases with frequency for a fixed substrate thickness. It also increases for increasing substrate thickness. The percent bandwidth decreases for increasing substrate dielectric constant which can be seen from the plot on page 504 for a constant thickness. All these plots are for a square patch. The effect of changing the width to length ratio on the percent bandwidth can be seen on the plot on page 505. The bandwidth changes quite slowly as the ratio is changed.

We can change the input impedance at the edge of the patch by increasing the width of the patch. The impedance is 50 ohms for a width to length ratio of about 3.70. On page 506 is a plot of this function.

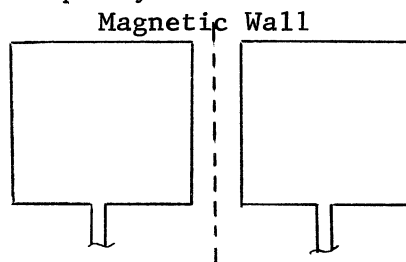
The bandwidth of the antenna can be increased slightly by changing the feed point or input transformer so that it is larger than 50 ohms. On page 507 is a plot of VSWR versus frequency for two feed points. The bandwidth at both the 2:1 and 3:1 VSWR points has increased. The antenna is no longer matched perfectly at the center frequency but the total VSWR bandwidth has increased.

The efficiency can be found by calculating the microstrip transmission line losses and using the transmission line model. Using transmission line theory the transmission matrix is calculated from the input to each slot including the other slot in the transmission matrix. The power delivered to each slot is calculated, summed, and compared to the input power. The difference is the efficiency which is plotted on page 508 for woven teflon fiberglass substrate. Note that the losses increase rapidly for low frequencies and thin substrates.

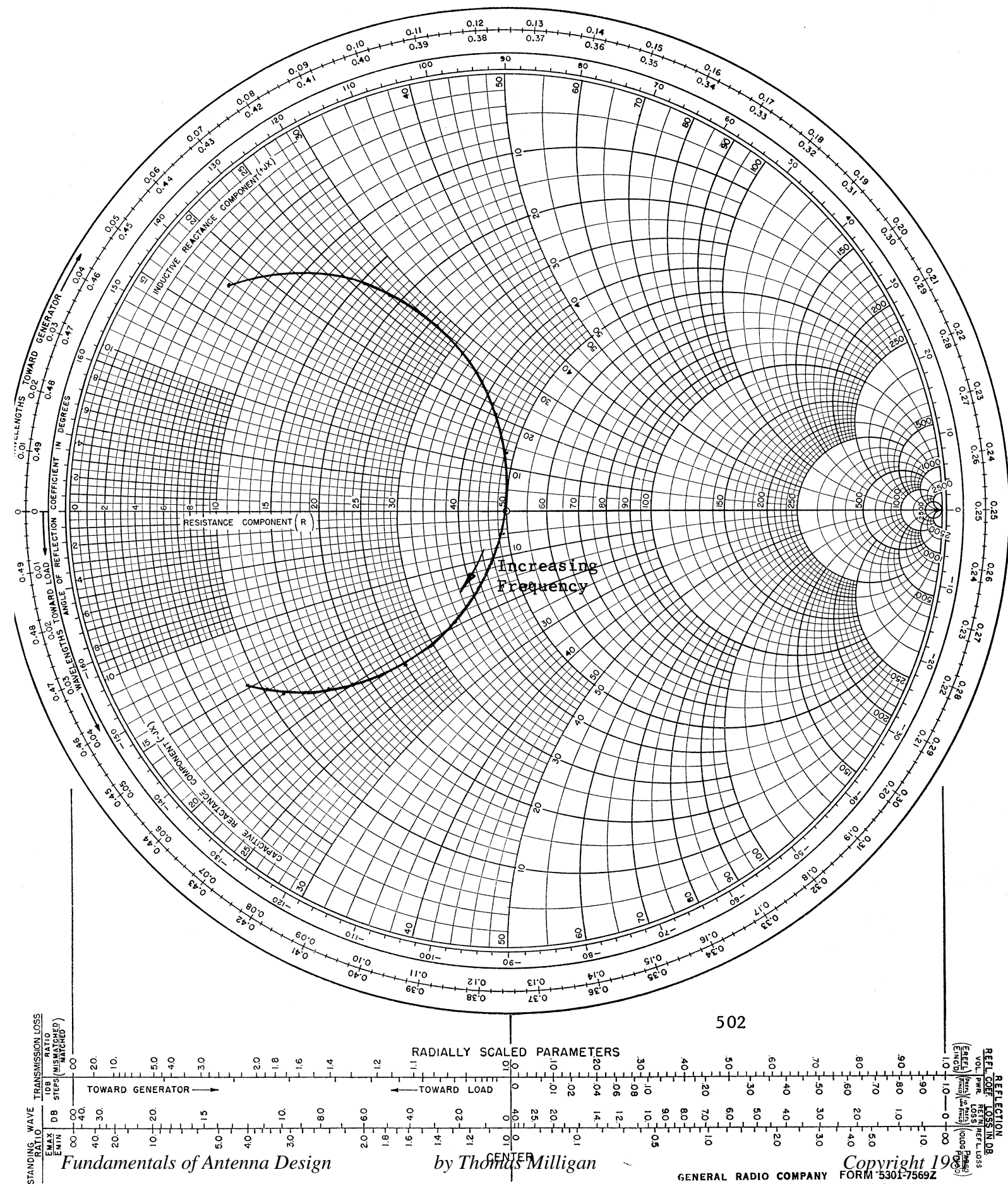
The field along the edges is approximately uniform with some tapering near the ends. This can be established from near field probing of the edges. The probe also reveals that the slot nearest the feed point has slightly larger fields. The pattern of a single patch is quite broad. The control of beamwidth is achieved by arraying the elements with the feed network etched on the same substrate between the patches.

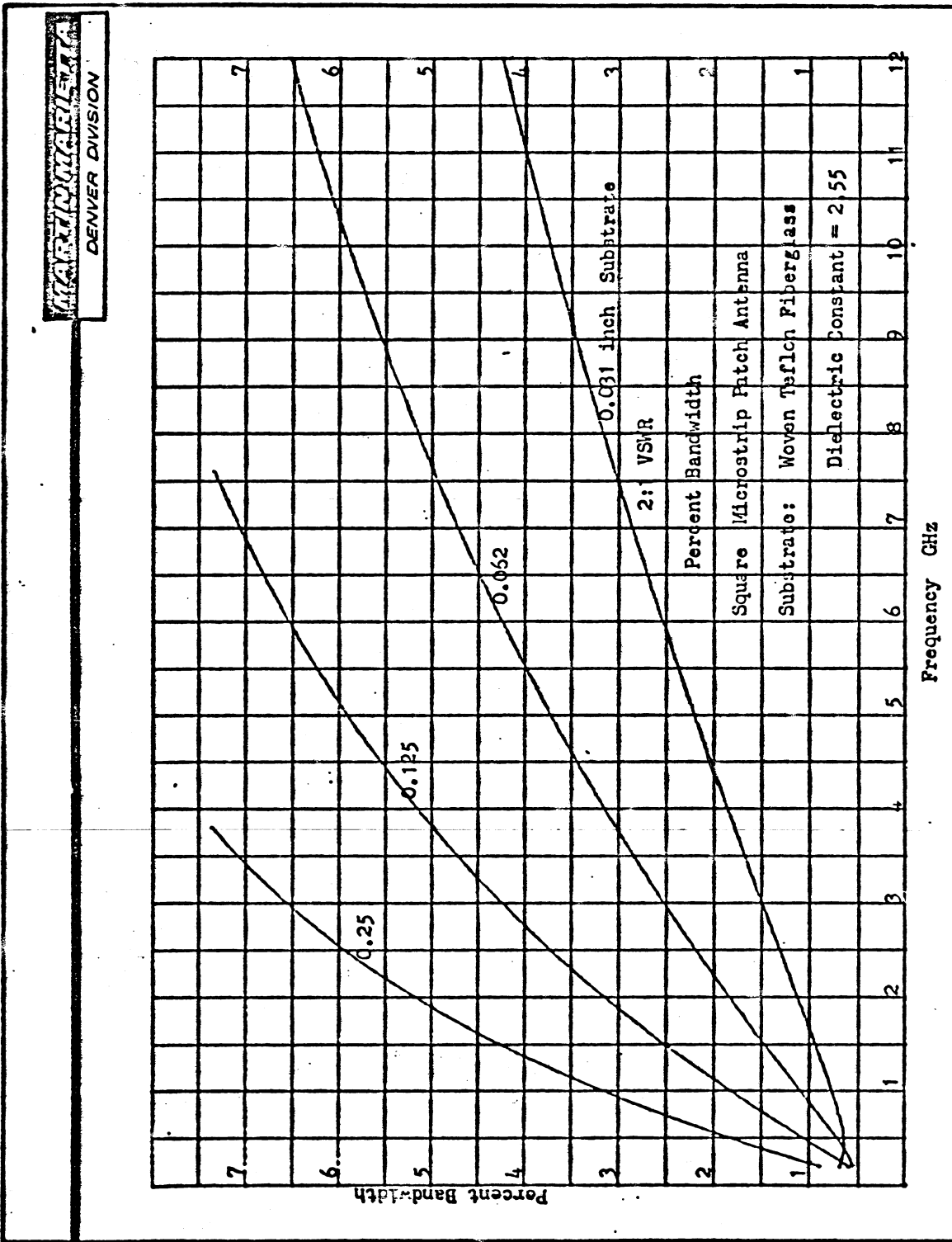
CONTINUOUS STRIP MICROSTRIP ANTENNA

Suppose we have two patches which are side by side in the H plane and equally fed. Because they are equally fed there will be a magnetic wall between the

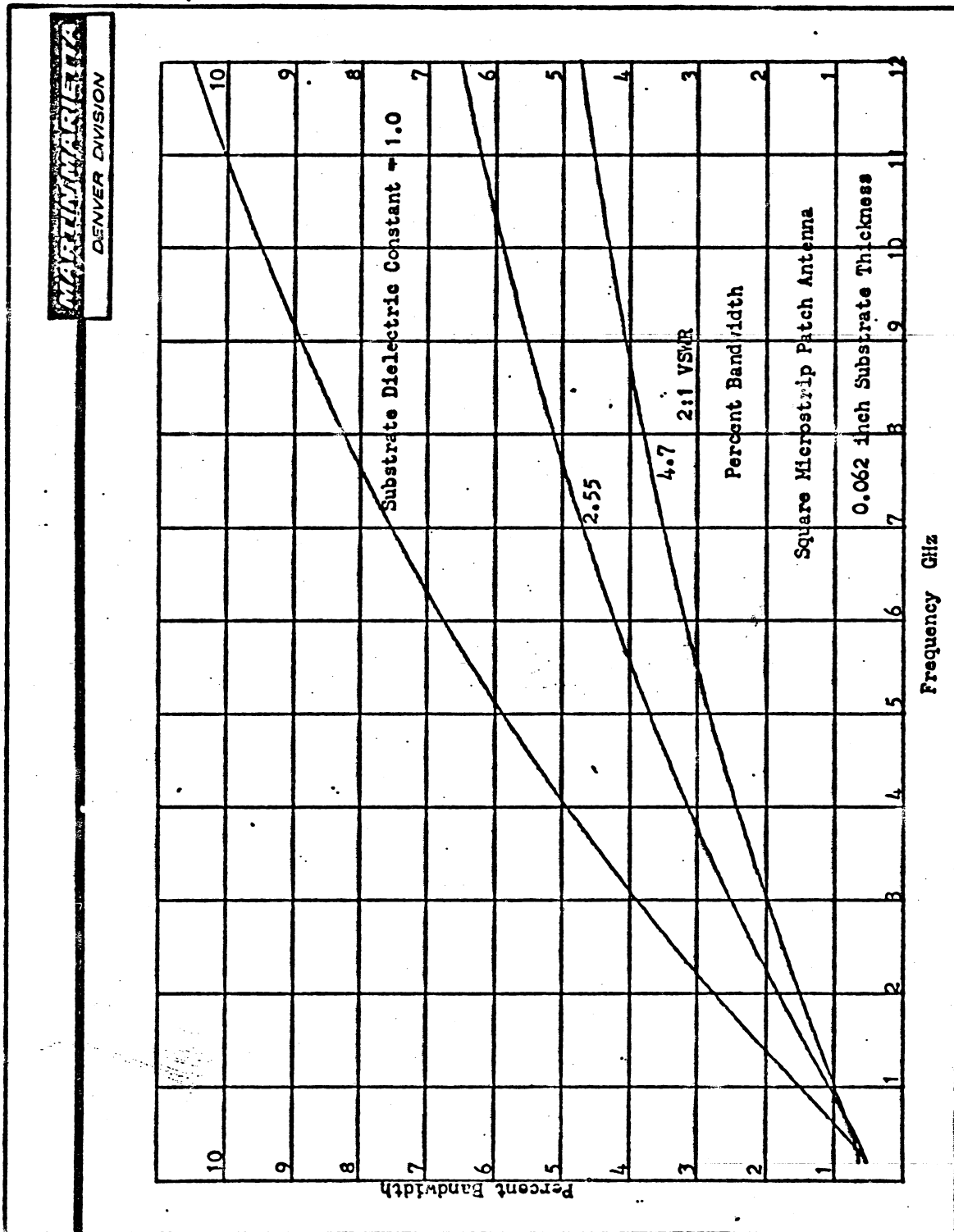


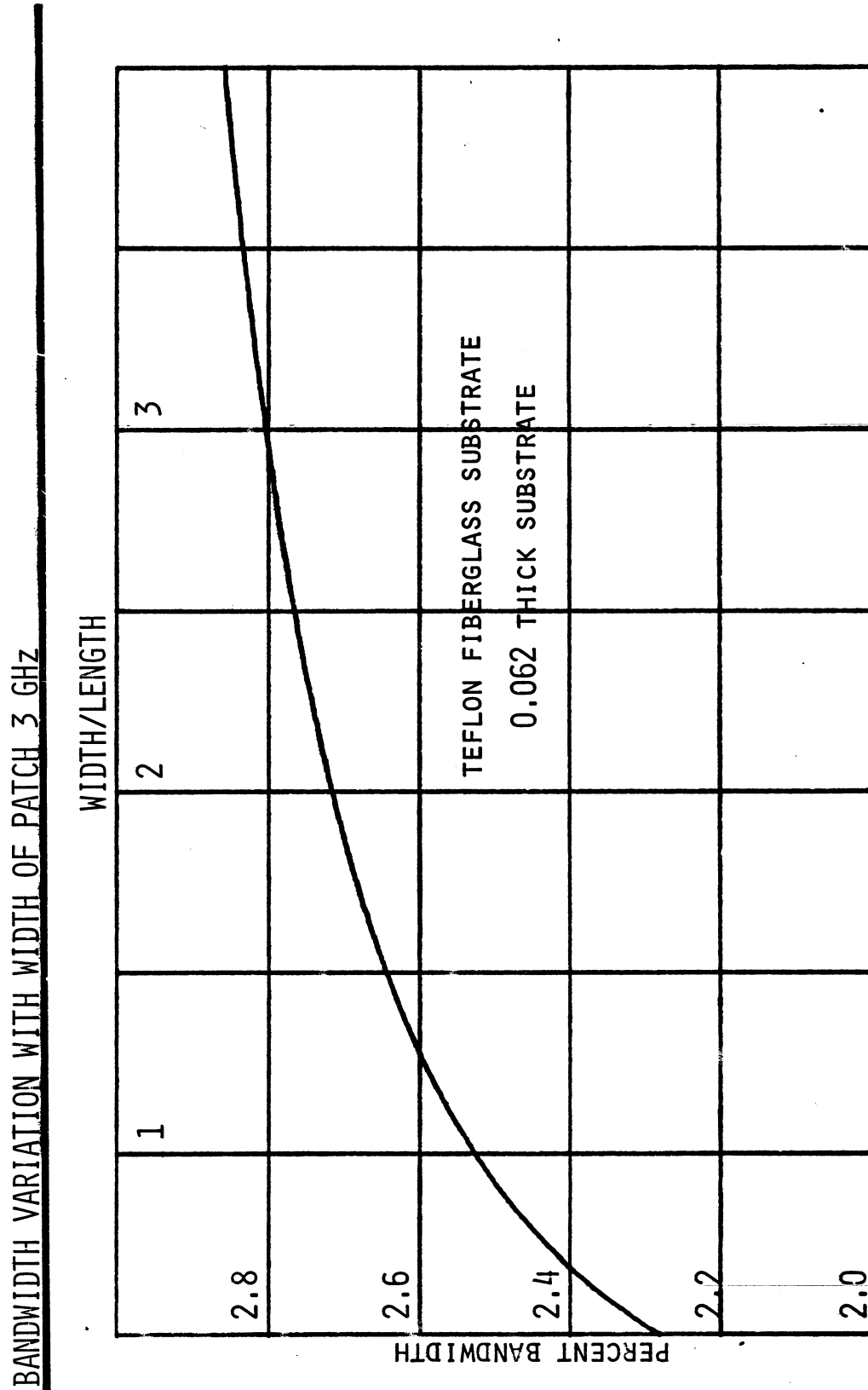
IMPEDANCE COORDINATES—50-OHM CHARACTERISTIC IMPEDANCE





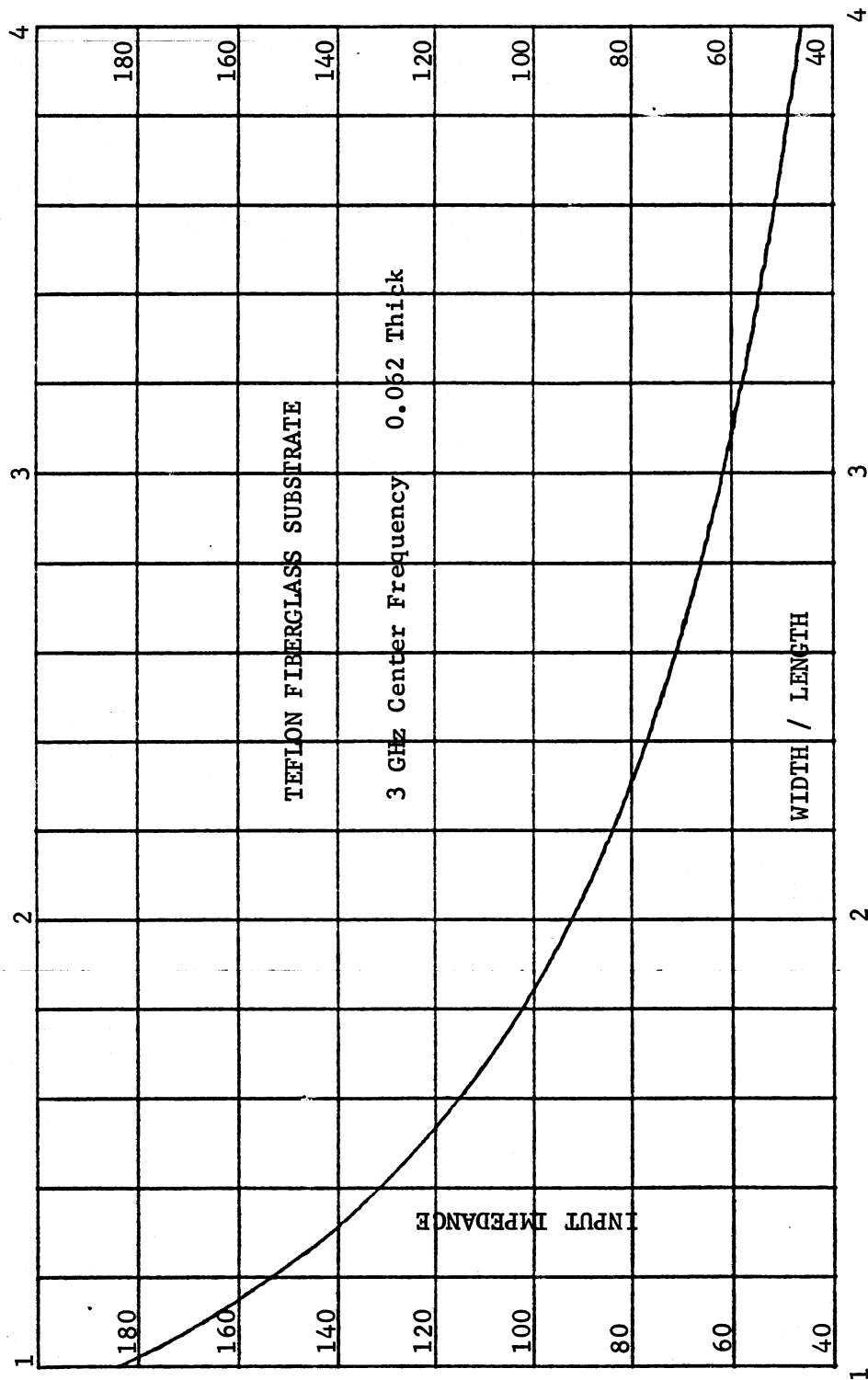
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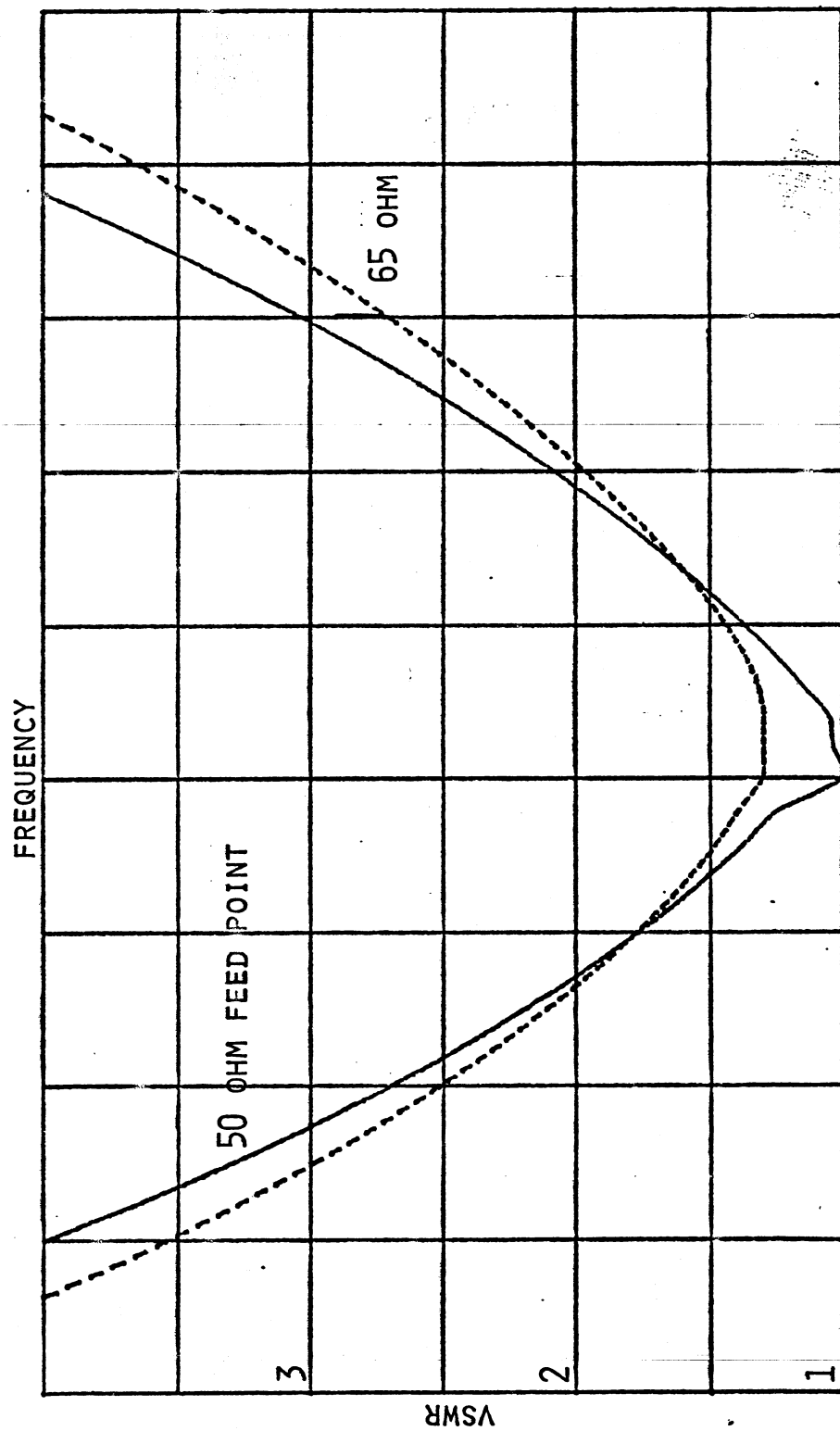
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IMPEDANCE VARIATION WITH PATCH WIDTH

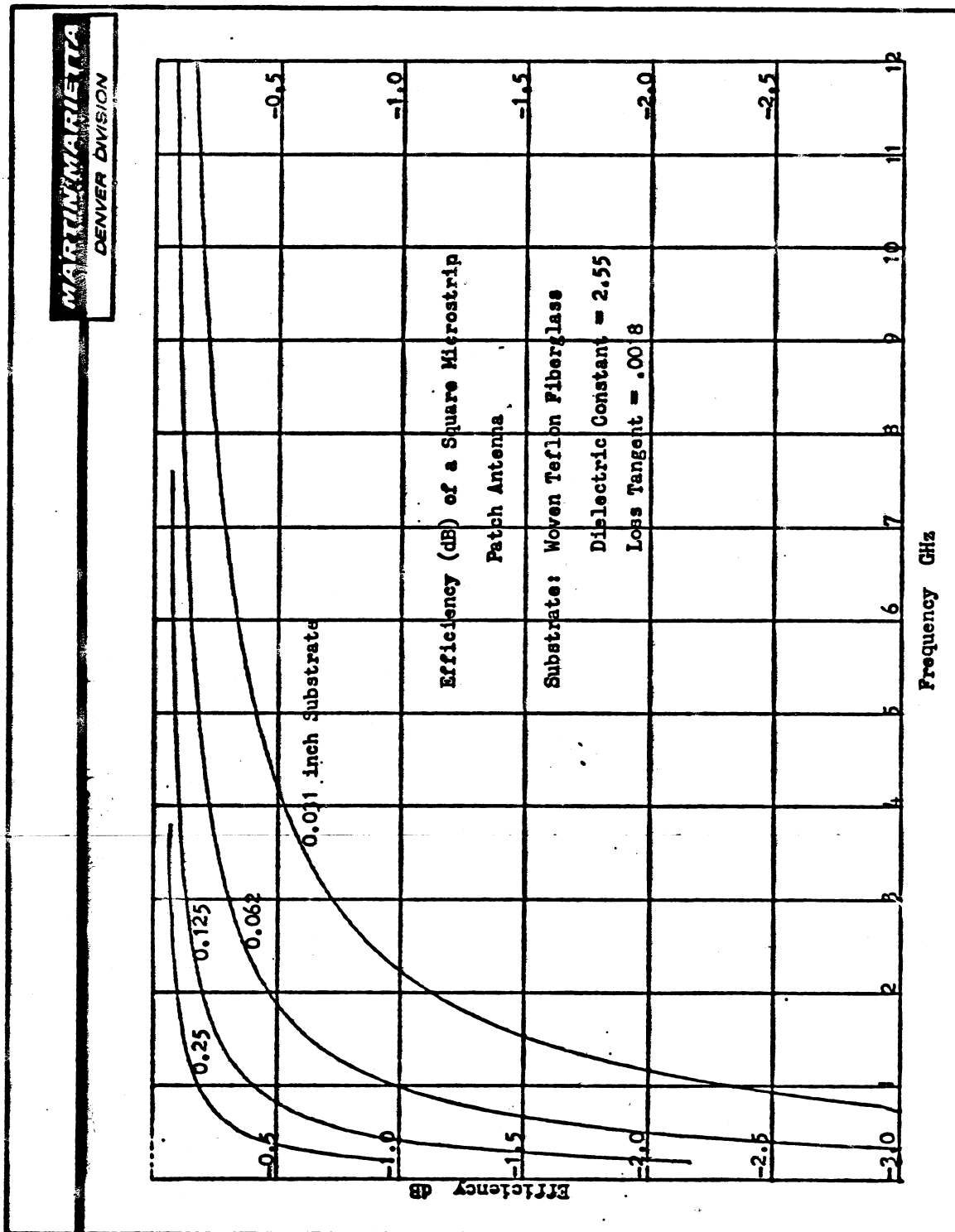


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BANDWIDTH VARIATION WITH FEED POINT IMPEDANCE



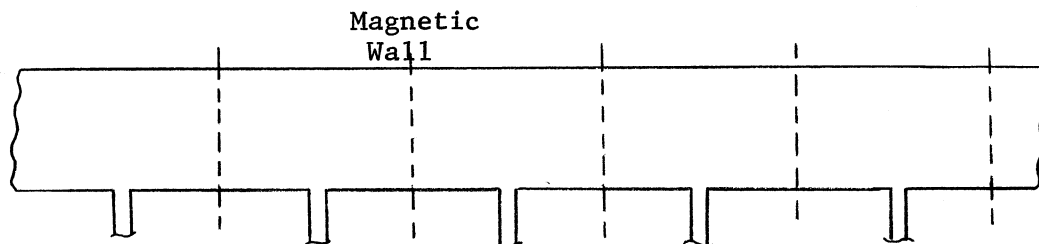
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elements. We can join the edges without any effect because the midpoint between the edges remains a virtual open circuit (magnetic wall). The antenna can be made of a continuous strip with a number of feeds. The feeds must be spaced close enough to provide uniform illumination along the edges and to prevent grating lobes.

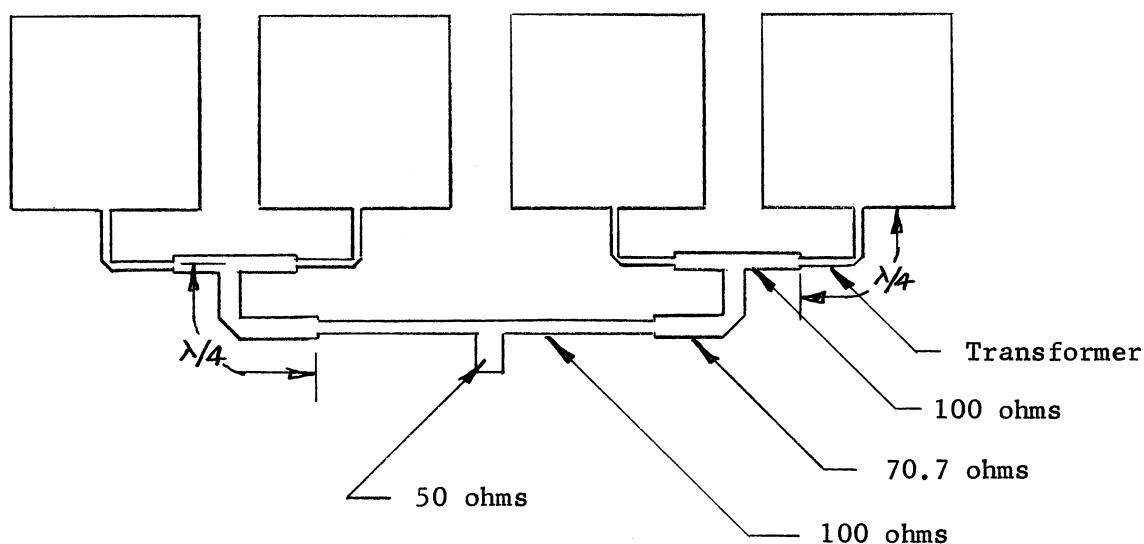
These antennas can be wrapped around missiles to provide omnidirectional coverage with no ripple in the pattern as the antenna spins about the axis of the missile. The antenna strip must be fed about every $3/4$ wavelength along the antenna.



The impedance at resonance at each feed will be the combination of the radiation conductances from the edges between the magnetic walls.

FEED NETWORKS

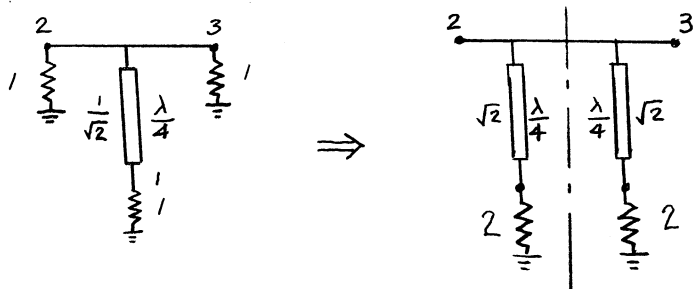
The feed network can be etched on the same substrate as the radiator with very little radiation from the feed. There is little radiation from the fringing fields on both sides of the microstrip lines because they are close together and in opposite directions. Let us consider equally fed microstrip patches or continuous strip. We must transform the resonant impedance of the patch to some microstrip characteristic impedance at each feed. The losses of lines from 50 to 100 ohms are about the same; increasing slightly with impedance. For substrates with dielectric constants around 2.5, the line widths must be quite large, $W/H = 2.82$, for 50 ohms lines. The usual characteristic impedance used is 100 ohms, $W/H = 0.76$. Consider a typical feed network drawn for a four element patch antenna array. This will demonstrate the general feed.



From the patch to the feed network is a quarterwave impedance transformer from the high impedance of the patch to a hundred ohm line. The power dividers used in the network are reactive power dividers. Compared to the patch they have sufficient bandwidth. The power division is determined by the impedances of the loads to the power divider. Since the patches are identical, the impedance and power division will be equal to the two patches. When we join the two 100 ohms lines in shunt as shown, the combination will be 50 ohms. We have decided to have a characteristic impedance of 100 ohms for the feed network lines. It is necessary to transform the 50 ohms to 100 ohms which is done by the 70.7 ohms quarterwave transformer. The lengths from the patches to the juncture must be equal so that we will have equal phases at the input.

After the power from the two antennas are joined together on a single line, the power is guided by the 100 ohms line and joined to the line coming from the other pair of radiators. In our example when these lines are joined in shunt, the combination is 50 ohms which matches directly to a coax input. If there were more patches, then the 50 ohms would be transformed again to 100 ohms and carried to another junction. In this method of feeding the antenna it is best to have 2^N outputs (2, 4, 8, 16, ... , etc.). Then all power dividers are the same.

The reactive power divider is matched at the input but not at the outputs. Since it is a lossless three port circuit, it cannot be matched at all ports without being a circulator which requires nonisotropic material in its construction. Consider the normalized power divider circuit below.



We can divide the transformer into a shunt connection of two lines each will twice the impedance which is loaded by twice the input impedance. With this network we can analyze the output reflection coefficient and isolation between the outputs. We will use the plane of symmetry with an even-odd mode analysis.

Suppose we have the even mode. There will be a magnetic wall (virtual open circuit) along the plane of symmetry. The circuit is reduced to a quarter-wave transformer between an impedance of one and two. The circuit is matched.

$$S_e = 0 \quad (\text{even mode reflection coefficient})$$

In the odd mode there is an electric wall (virtual short circuit) at the plane of symmetry. The short will short out the transformer and give us an odd mode reflection coefficient of a short circuit.

$$S_o = -1 \quad (\text{odd mode reflection coefficient})$$

The input reflection coefficient is given by

$$S_{22} = S_{33} = \frac{1}{2}(S_e + S_o) = -\frac{1}{2}$$

The transmission between the two output ports or isolation is given by

$$S_{23} = \frac{1}{2}(S_e - S_o) = \frac{1}{2}$$

From these voltage reflections and transmission we can find the return loss and isolation to be 6 dB.

If one of the antennas is damaged, then the power from that antenna will be reflected into the other antennas. In general this will change the power delivered to the other patches and change the distribution to the antennas. The effect will be greater than just having a missing antenna in the array. Power dividers with isolation resistors is the solution to this problem. Since it is difficult to mount the isolation resistors, most antennas are made without them. If the etching of the patches is good, then the antenna will work with the reactive power dividers.

One problem with feed networks is the coupling between different parts of the feed circuit. Many times these anomalies in the patterns have been explained as feed line radiation. Although it can occur, it is not the chief reason. We want to pack the feed network together into the smallest area. When we do this there is direct coupling between the feed lines. As the distance, S , in the feed above shrinks, the coupling between the feed lines increases and causes direct and coupled signals to be added in random phase. These random couplings appear as anomalies in the pattern. As the number of feeds increases the anomalies grow because there are more couplings. Unfortunately the coupling between microstrip lines falls off quite slowly.

MICROSTRIP TRANSMISSION LINE

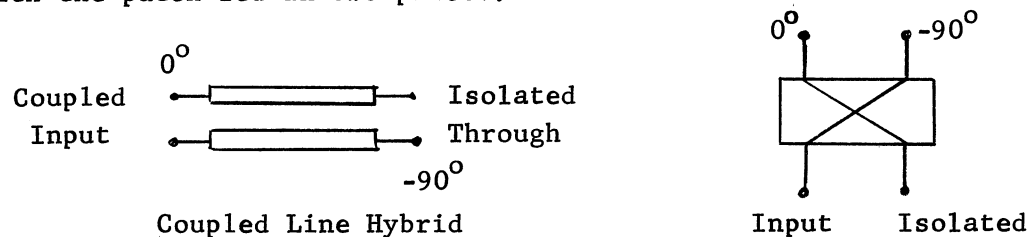
Finding the impedance of microstrip lines is not part of this course. A few curves are given here so that power divider networks can be designed for microstrip antennas. The impedance decreases with increasing stripwidth or dielectric constant. On page 512 is a plot of the characteristic impedance of microstrip versus the ratio of the stripwidth to the substrate thickness for various dielectric constants. A curve with various important impedances of stripwidth versus dielectric constant of the substrate is drawn on page 513. The effective dielectric constant of the line changes with stripwidth. A curve of the effective dielectric constant is drawn on page 514. The microstrip transmission line is also dispersive in frequency. This effect is so small that it may be ignored in most designs.

CIRCULARLY POLARIZED MICROSTRIP PATCHES

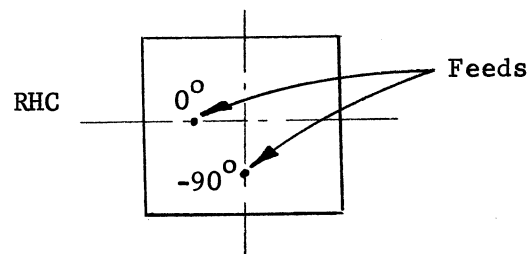
The microstrip patch antenna can radiate circular polarization if the patch is made square. The square patch can be excited with orthogonal linear polarizations at the same frequency. The edges work in pairs. If we excite the two linear polarizations 90 degrees out of phase, then the radiated wave will be circularly polarized. This is the same as crossed dipoles.

Hybrid Feed

The best method of feeding the patch is to use a 90° coupled line 3 dB power divider with the patch fed in two places.



Usually the patch is fed from below at the 50 ohm feed points or possibly

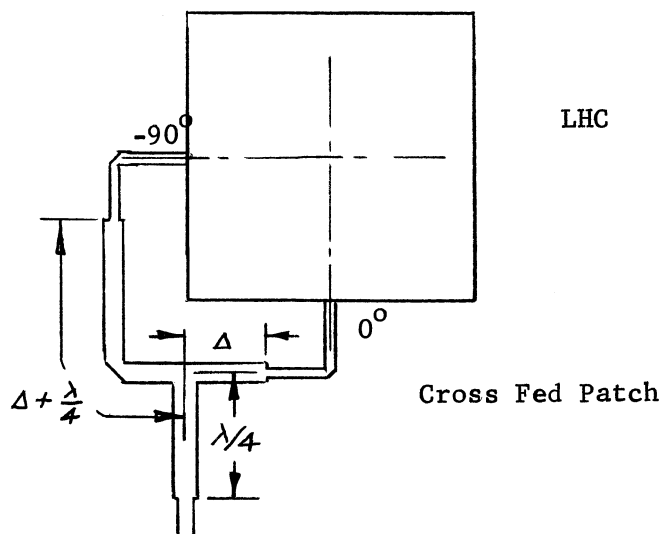


at a higher impedance point to achieve more bandwidth. When the input impedances to the patch are equal, the power division will be equal and 90 degrees out of phase. From symmetry the impedances will be equal at all frequencies. The boresight axial ratio is 0 dB at all frequencies for a perfectly square patch with equal feed points. The antenna shown above radiates right hand circularly polarized wave.

The symmetrical 90° coupled line hybrid couples the energy from equal output mismatches into the isolated port. Therefore the antenna will appear matched at the input to the coupler. The efficiency of the antenna will suffer due to the mismatch loss of the antenna. If we feed the antenna from the isolated port of the coupler, then we get the other sense of circular polarization.

Cross Feed

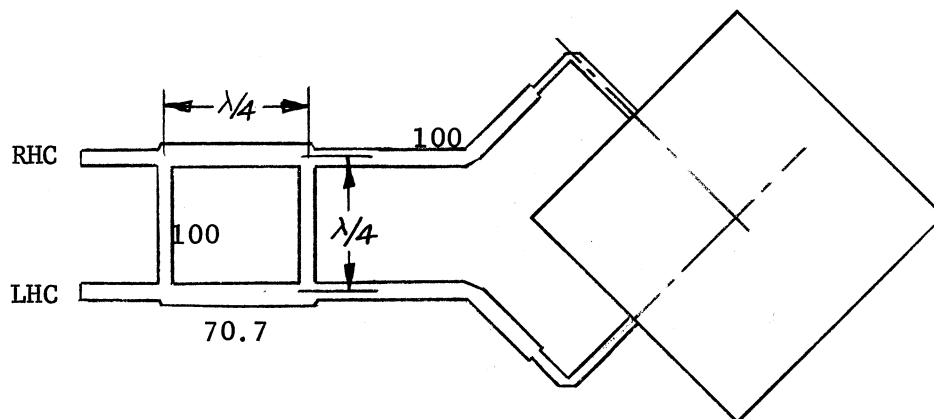
One of the most common methods of achieving circular polarization is to feed two edges of the patch with one of the feeds phased by 90 degrees at the center frequency. At each feed point there is a quarterwave transformer to change the impedance to the characteristic impedance of the feed network. The two feeds are then tied together at a reactive power divider as shown in the figure below. One of the feeds has an extra quarterwave section to delay



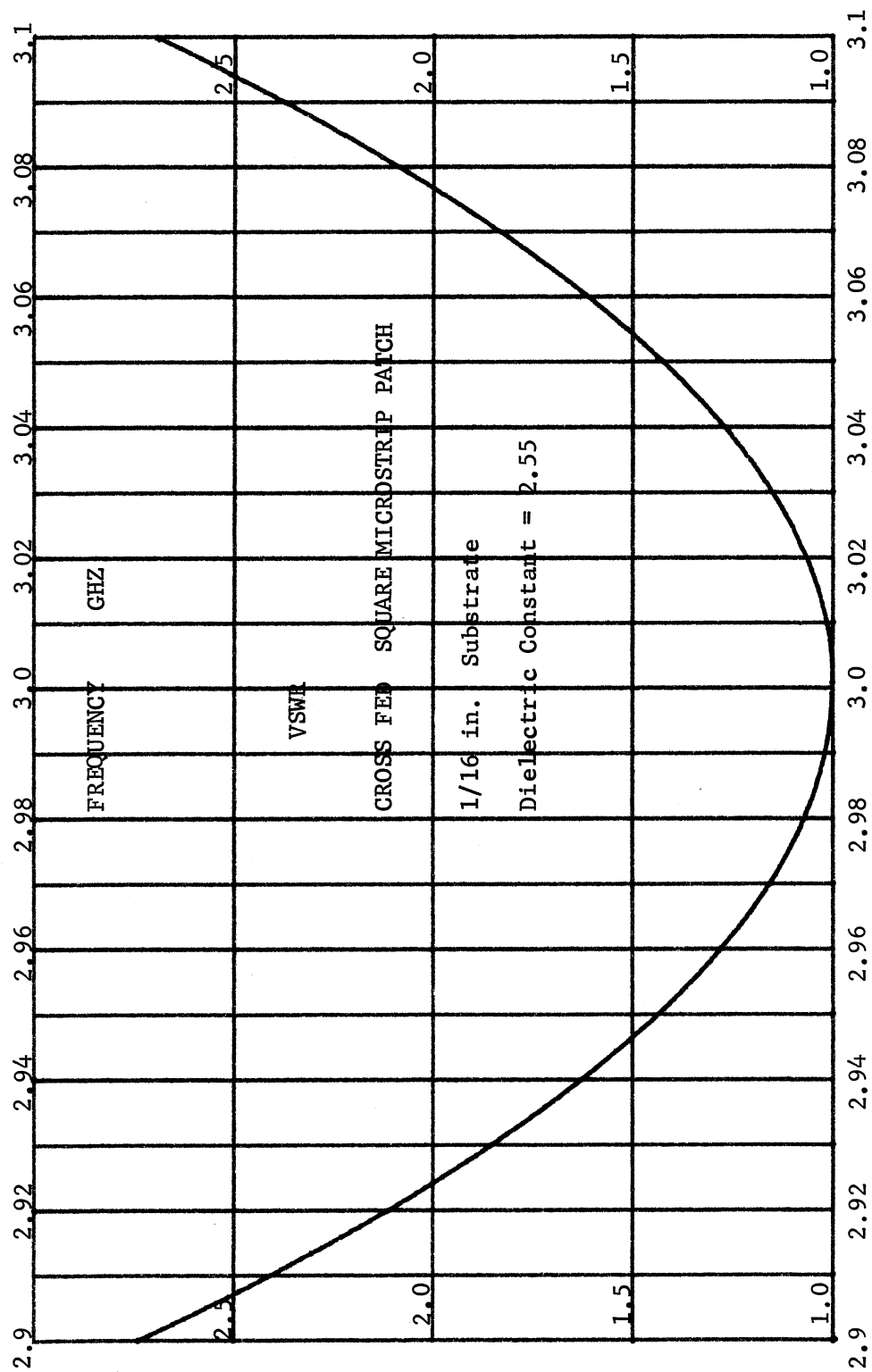
the phase by -90° . The patch in the figure radiates left hand circular polarization (LHC).

The response of an antenna was calculated for a 1/16 in thick woven teflon fiberglass substrate ($\epsilon_r = 2.55$) at 3 GHz for the cross fed square patch. On page 517 is a plot of the input VSWR which is approximately the same as the singularly fed patch. The boresight axial ratio response is plotted on page 518. From this plot we can see that the axial ratio bandwidth is quite small compared to the VSWR bandwidth. This antenna has the advantage that the feed network can be etched on the same substrate as the elements. The hybrid coupler feed above requires a coupler for each antenna.

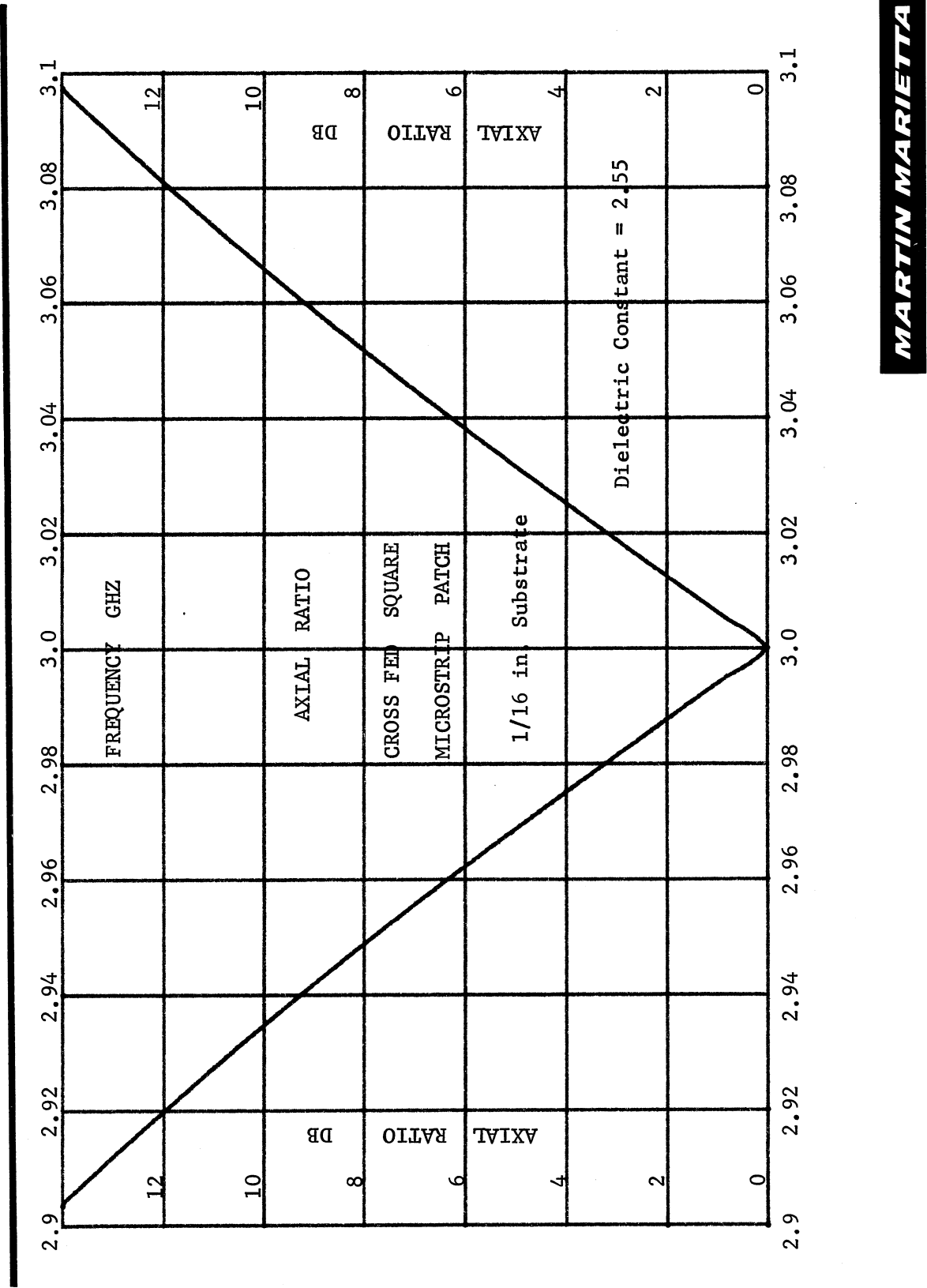
Branchline Hybrid Feed



The branchline hybrid consists of four transmission line connected in a square as shown above. The outputs have a phase difference of 90 degrees and equal amplitude at center frequency. The hybrid drawn above is for a 100 ohm system. The branchlines are the same impedance as the system and the through lines are 0.707 times the characteristic impedance of the system. The network is broadband compared to the microstrip patch antenna and provides for two inputs. The inputs will give different senses of circular polarization as shown in the figure above.



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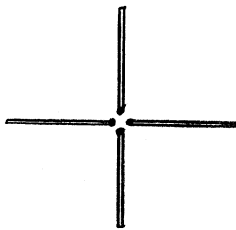
The input impedance of the microstrip patch must be transformed to the characteristic impedance of the feed network by quarterwave transformers as shown before it can be connected to the hybrid. For comparison with the cross fed antenna, the same antenna was analyzed as above. The antenna is on a 1/16 in. thick woven teflon fiberglass substrate at 3 GHz. The VSWR response of the antenna is plotted on page 520. Compared to the cross fed patch this antenna has a very large VSWR response. This is due to the hybrid. Near the center frequency equal mismatch reflections are reflected into the isolated port of the hybrid like the coupled line hybrid. The axial ratio response at boresight is plotted on page 521 and it too shows a wide bandwidth compared to the cross fed patch. Notice the change of scale on the two plots. The hybrid has a flat response over the band of the patch antenna and maintains the 90 degree phase shift and equal power division.

The antenna is not as good as it appears from the two plots given for VSWR and axial ratio. There is a load on the isolated port of the branchline hybrid which absorbs some of the energy when operating off center frequency. A plot of the total circuit losses is given on page 522 which shows that the efficiency of the antenna becomes the limiting factor for bandwidth. The same curve also holds for the coupled line hybrid. The loss is due to the reflected power of the antenna: mismatch loss.

TURNSTILE ANTENNA

We can also achieve circular polarization by feeding a rectangular patch at a corner. Before discussing the antenna directly, we need to backup and look at the dipole turnstile antenna.

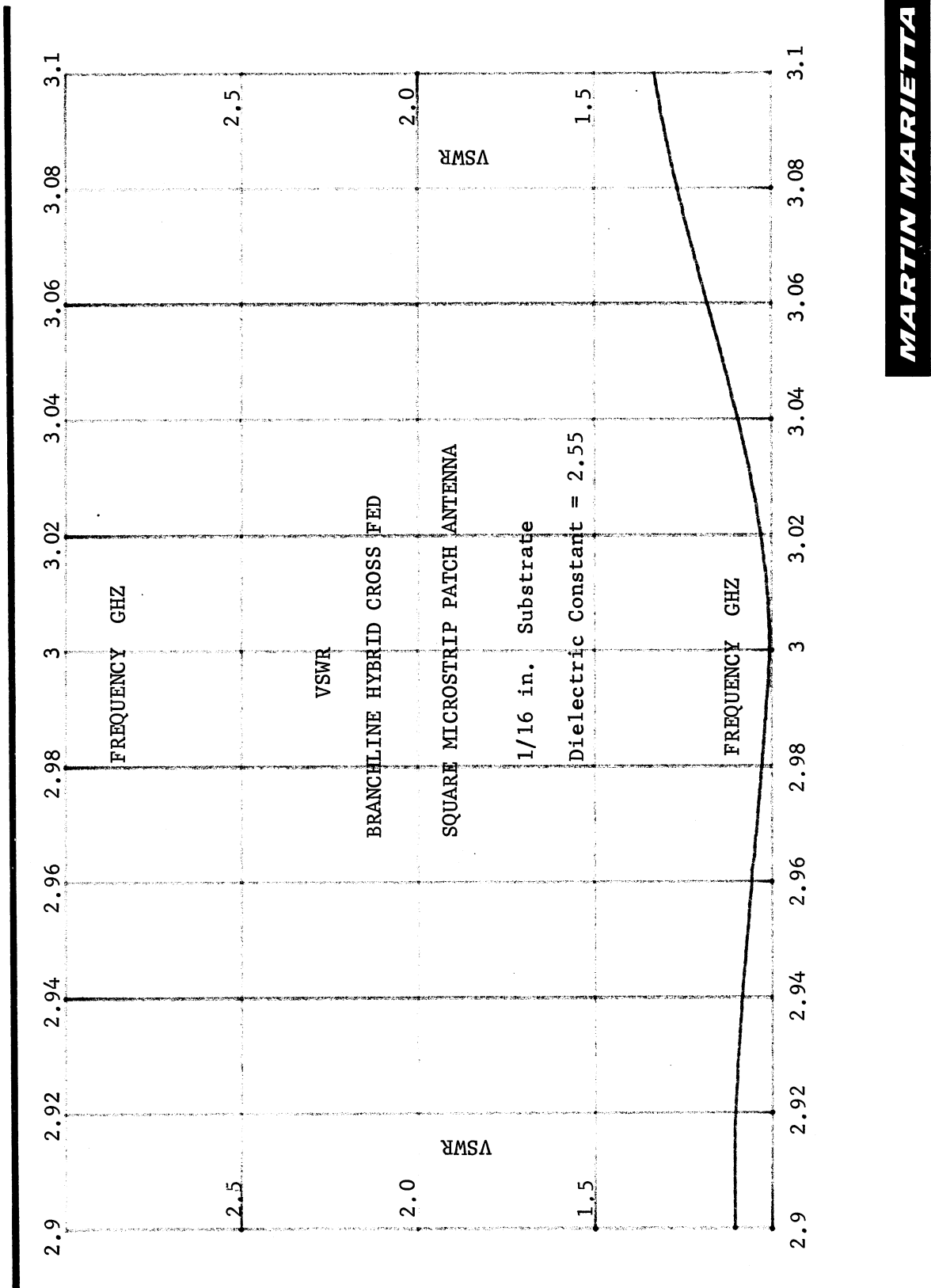
Suppose we have orthogonally placed dipoles as shown below with the centers at the same point. This is similar to the square patch. We can feed the two



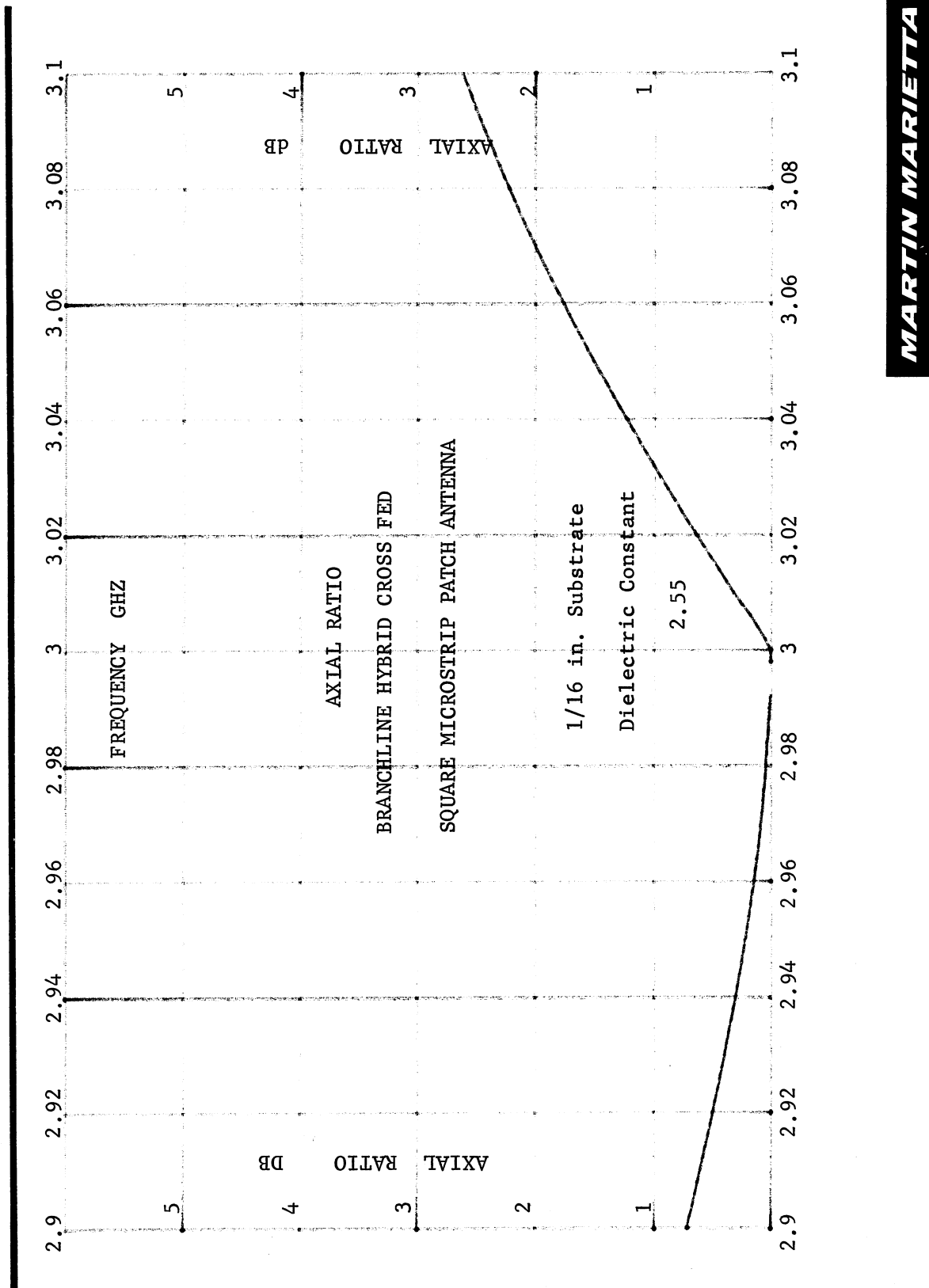
dipoles 90° out of phase and achieve circular polarization above the dipoles in the same manner as the patch. If the dipoles are equal and the inputs are fed from a coupled line hybrid or branchline coupler, then the antenna will have a large VSWR bandwidth at the input to the hybrid. Also the axial ratio bandwidth will be large. Each set of dipoles must be fed by a balanced line.

If one of the dipoles is at resonance and we increase the length which is the same as increasing the frequency, then the input impedance will become inductive. This can be seen from the plot on page 317. The current into the dipole is given by

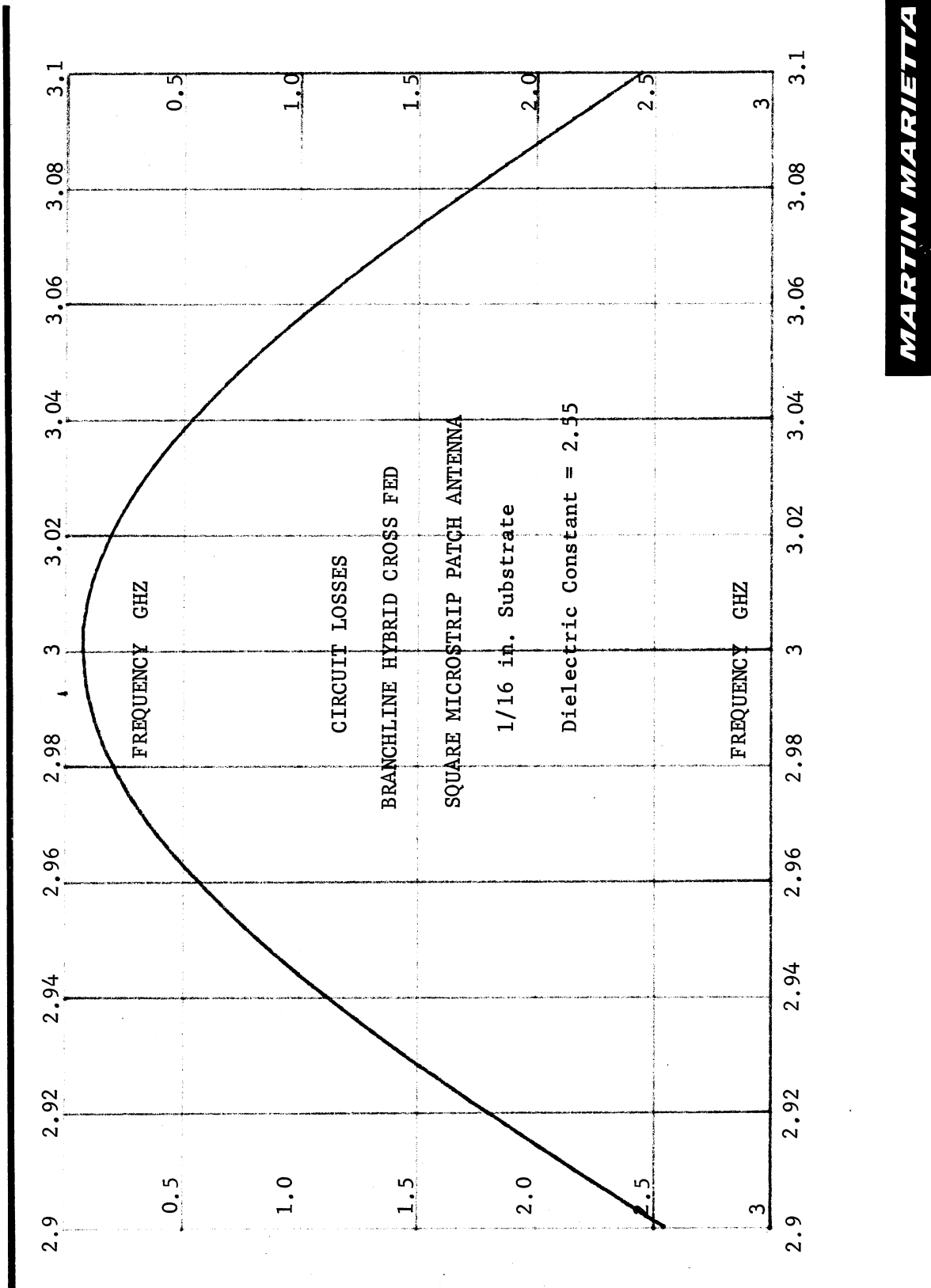
$$I_2 = \frac{V}{Z_2} = \frac{V}{R_2 + jX_2} = \frac{V(R_2 - jX_2)}{R_2^2 + X_2^2}$$



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MARTIN MARIETTA



The radiated electric field is proportional to the magnetic vector potential which in turn is proportional to the current. The phase of the field becomes more negative relative to the unchanged dipole.

$$I_2 = |I_2| e^{j\gamma_2} \quad \gamma_2 = \tan^{-1}(-X_2/R_2) = -\tan^{-1}(X_2/R_2)$$

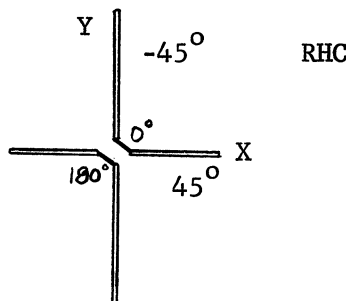
$$E_2 = k A_2 = K_1 I_2$$

Similiarly if we shorten a resonant dipole, it becomes capacitive (p. 317) and the phase of the input current increases.

$$I_1 = \frac{V}{R_1 - jX_1} = \frac{V(R_1 + jX_1)}{R_1^2 + X_1^2}$$

The phase of the electric field increases relative to the resonant dipole field.

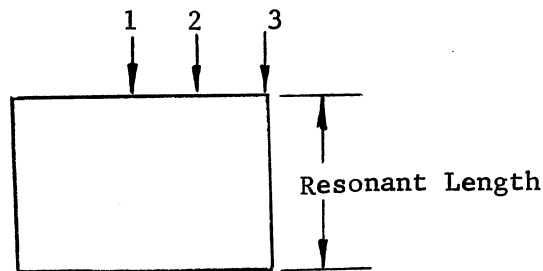
Let us connect the two dipoles together so that we have a single feed. One set of dipole element lengths will be increased while the other set will be decreased. The dipoles are connected in shunt at the input. The input admittance will be the sum of the individual dipole admittances taken separately. If the dipoles are orthogonal, then the mutual impedance between them is zero. We can adjust



the sizes so that the combination of the susceptances cancel, but this is not our intent. We are seeking a circularly polarized wave. If each dipole has its reactance equal to the resistance, then its phase will be plus or minus 45 degrees. This combination will give us a circularly polarized wave on boresight. The connection in the diagram above will give right hand circular polarization above the figure. From symmetry the wave will be left hand circularly polarized below. The antenna is usually mounted about a quarter-wave over a ground plane so that the reflected wave which changes sense of polarization will add with the directly radiated wave. The antenna is adjusted until the desired performance is achieved.

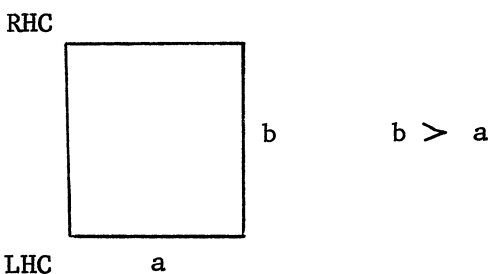
MICROSTRIP PATCH TURNSTILE ANTENNA

If we take a rectangular microstrip patch antenna and move an edge feed away from the center line, the input impedance remains constant near the resonant frequency determined by the distance to the opposite edge. The input impedance at the various feed points shown in the figure below is constant which can be verified by experiment. The feed is exciting the resonant cavity of the patch by injecting current into the top plate. If we have a square



Equal input impedance at points 1, 2, and 3 along the non resonant length.

patch and move the feeds on both edges to the same corner, then we can combine the feeds into a single feed. The phase of the injected current into each orthogonal mode is determined by the input admittance for the resonant lengths in each direction. The patch has duplicated the shunt connected dipoles of the turnstile antenna.



The patch in the figure above has a reduced 'a' dimension and increased 'b' dimension from a square patch to give a circularly polarized wave like the turnstile dipole antenna. The patch will radiate RHC or LHC depending on which corner is used to feed the antenna (shown above). The dimension 'b' is larger than 'a'.

The Q of a resonant circuit is associated with the VSWR bandwidth by

$$BW = \frac{VSWR - 1}{Q \sqrt{VSWR}}$$

The Q of a circuit is determined by the points where the reactance equals the resistance off resonance.

$$\Delta F = F_o / Q$$

Where $\Delta F = F_2 - F_1$ and $F_o = \sqrt{F_1 F_2}$

If we know the bandwidth ratio of the microstrip patch antenna singly fed, then we can find the two resonant frequencies of the two dimensions of the patch to give a circularly polarized wave.

Suppose we use the 2:1 VSWR bandwidth which is given on the plot on page 503, then

$$Q = \frac{1}{\sqrt{2} BW}$$

The larger dimension is resonant at a frequency

$$F_1 = F_0 \sqrt{1 - 1/Q}$$

The smaller dimension is resonant at the frequency

$$F_2 = F_0 \sqrt{1 + 1/Q}$$

Example: Design a rectangular corner fed patch to give circular polarization at 3 GHz. Use 1/16 inch woven teflon fiberglass substrate.

The 2:1 VSWR bandwidth of the singly fed patch is 2.53% which corresponds to a $Q = 27.96$. Using the equations above we can find the resonant frequencies.

$$F_1 = 2945.9 \text{ MHz}$$

$$F_2 = 3053.2 \text{ MHz}$$

The resonant lengths are found by the standard technique to be

$$L_1 = 1.223 \text{ inches}$$

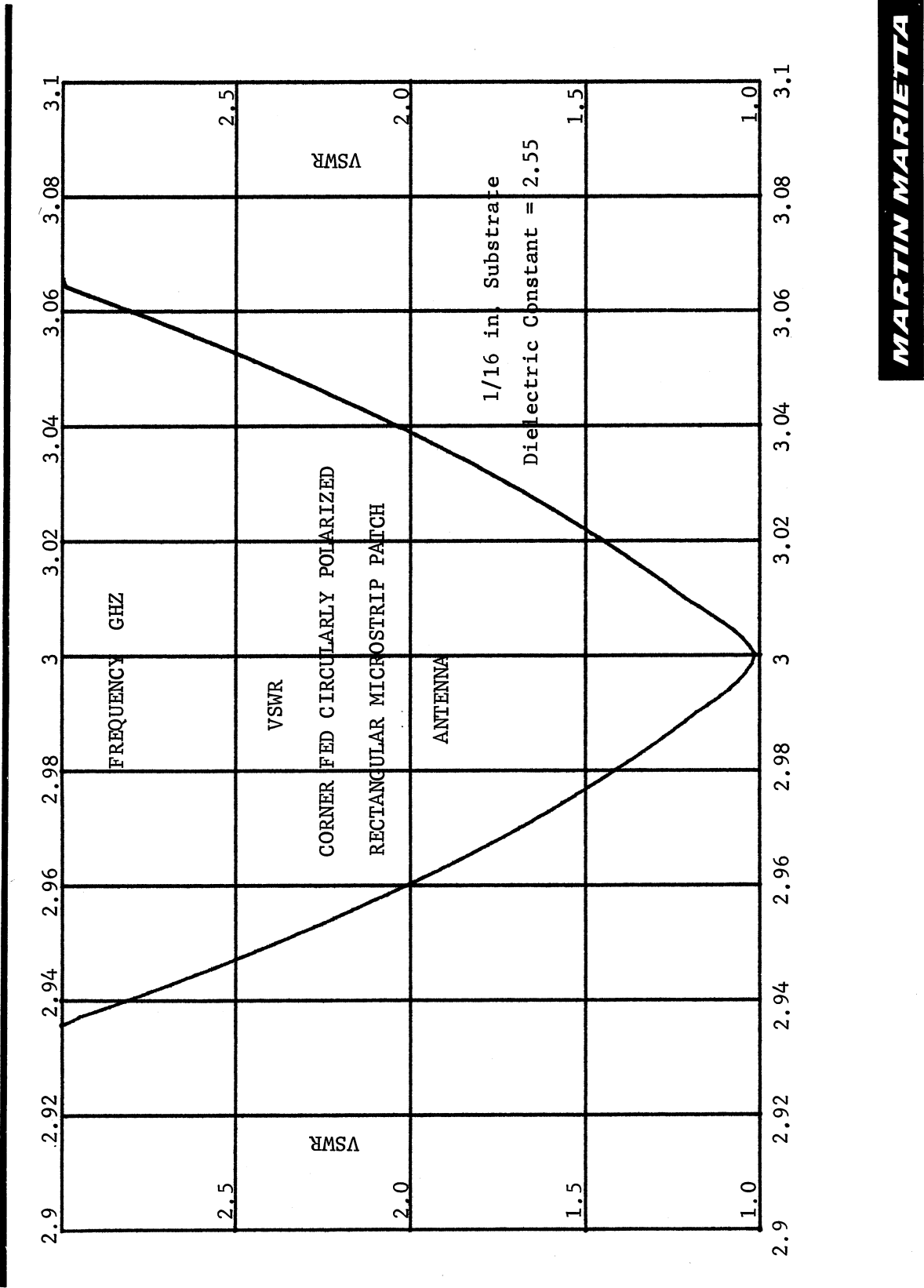
$$L_2 = 1.179 \text{ inches}$$

The input impedance will be about half the value of the singly fed antenna at the same size because it is feeding two cavities in shunt with approximately the same input resistance. The antenna may be fed from the corner or from a probe from below along the diagonal. The probe will excite modes in both directions.

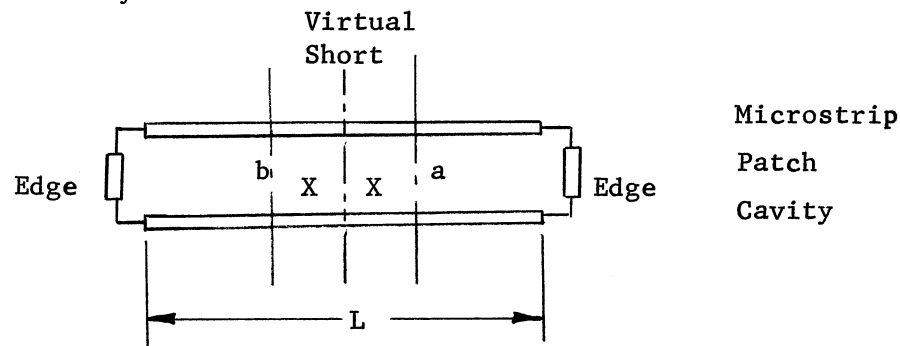
The design of the antenna was completed and the frequency response has been calculated. The design is on the same size substrate and at the same frequency as the previous designs so that we can compare them. The VSWR response is plotted on page 526. The 2:1 VSWR bandwidth of the cross fed patch is about twice the corner fed antenna. On page 527 is a plot of the axial ratio frequency response. When we compare this to the curve on page 518, we see that the axial ratio bandwidth has increased for the corner fed patch from the cross fed patch. The increase at the 6 dB level is about 1.5 times. Compared to the cross fed antenna, the corner fed antenna has a smaller VSWR bandwidth but a larger axial ratio bandwidth.

Microstrip patch antennas can be made with shapes other than rectangles. Circular, elliptical, and pentangular patches have been described in the literature. In these cases the size of the patch for resonance is found by solving for the resonant modes of the cavity. Most of these patches have been used for circular polarization (ellipse and pentangle). The circular patch can be fed with dual linear polarization like the square patch.

The microstrip antenna is a cavity from the circuit point of view. The transmission line impedance is low and the impedance of the edges are high in comparison. When radiating, the cavity is in an odd mode with a virtual short along the center line. We can put small susceptances in the cavity



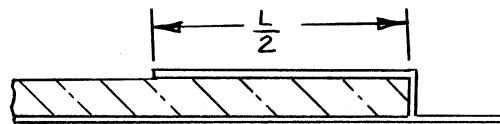
to shift the resonant frequency or improve the input match. But these susceptances must not dominate the end susceptances at the edges which establish the odd mode in the cavity.



If there is a shunt susceptance connected at point 'a' in the circuit of the patch as shown above, it will disturb the symmetry of the radiating mode and mismatch the input. But if another equal susceptance is connected at the symmetrical point 'b', then symmetry is maintained and the input remains matched. These susceptances must remain small in comparison to the edge admittance which establishes the resonance of the cavity.

HALF MICROSTRIP PATCH

The patch antenna is in the odd mode when radiating. This odd mode has a virtual short in the central plane of the antenna. We can reinforce the short by putting a shorting plane at the center and eliminate one-half of the microstrip patch. The effect is to remove one of the radiating slots.



The length of the cavity is about a quarterwave since the full patch is about a half wavelength long. The E plane pattern becomes that of a single slot: very broad. This slot antenna was discussed on page 189, but we will discuss the antenna in terms of the microstrip patch.

We only have the conductance and susceptance of one radiating edge. The input resistance at resonance will be twice that of the full patch. When the half patch is fed from the edge, the quarterwave transformer must be a higher characteristic impedance than for a full patch to transform the input impedance to the characteristic impedance of the feed network. This can present problems with thin substrates because the transformer line width will become too thin to practically make or reproduce.

The antenna is usually fed from underneath. The same equation given on page 500 for the location of the feed also holds for the half patch. There is however one other problem. Because the structure is no longer symmetrical, the input impedance no longer remains strictly real as the feed point is moved toward the short. The resonant frequency of the antenna will change as the feed point is moved. This means that the size of the patch will have to be changed while tuning the antenna.

The length of the half patch is found by the same method as the full patch. The capacitive susceptance of the slot is related to an equivalent length of transmission line. The cavity must be a quarterwave long at resonance. The length will be a quarter wavelength long in the effective dielectric constant of the microstrip line minus the cut back at the slot (edge). The width of the patch will determine the effective dielectric constant by the formula on page 496. Using this effective dielectric constant, the value of the cut back is found from the formula on page 496. The size of the half patch is determined from these.

$$L/2 = \lambda / (4 \epsilon_{\text{eff}}^{\frac{1}{2}}) - DL$$

The VSWR bandwidth of the half patch antenna is almost identical to the bandwidth of the full patch for antennas on the same thickness substrate, dielectric constant, and frequency. Because the transformer must be higher, the bandwidth is slightly less. For example the patches at 3 GHz on 1/16 in. woven teflon fiberglass have 2:1 VSWR bandwidths of 75.9 MHz for the full patch and 74 MHz for the half patch. Since these are so close, we can use the curves for the full patch bandwidth given on pages 503 through 506 for the half patch realizing that the bandwidth will be slightly less.

The short circuit of the antenna is quite critical. The microstrip impedance of the patch is quite low which means there are high currents on the patch. These high currents must be shorted by the shorting strap. If the impedance of the strap is not low, then the currents will not be properly shorted. These effects will detune the antenna and possibly cause problems in the radiation pattern of the antenna.