Electromagnetic Coupling Reduction Techniques

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By

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FOREWORD

This report was prepared by The University of Michigan, Ann Arbor, Michigan, under the direction of Professor Ralph E. Hiatt and Professor John A. M. Lyon and on Air Force Contract AF 33(615)-3371 under Task No. 435709 of Project 4357 (U) "Electromagnetic Coupling Reduction Techniques". The work was administered under the direction of the Air Force Avionics Laboratory, Electronic Warfare Division, Research and Technology Division, Wright-Patterson Air Force Base Ohio. The Task Engineer was Mr. Olin E. Horton, the Project Engineer Mr. Herbert Bartman.

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ABSTRACT

Significant improvements have been recently obtained for various methods of decoupling two antennas. The RF bridge reduction method has been analyzed to determine the decoupling possible with phase and amplitude errors present. Careful attention to matching the phase and amplitude characteristics of the auxiliary link to the coupling path has made it possible to extend the decoupling to a relatively broad range of frequencies.

The method utilizing circumferential corrugations surrounding either one or both of the antennas involved has been modified to include tapered corrugations. Improved performance has been achieved.

Further studies have been made on the use of metal picket fences as a decoupling means. Such a fence between two antennas has been very effective. Studies have been made utilizing both the metal fence and corrugations. The total decoupling is the sum of that achieved by each method.

In the near-field coupling of Archimedean spirals there is a practical problem in terminating the outside ends of bifilar spirals because the terminations occupy considerable space and influence the electromagnetic field. This report summarizes the studies to date on such near field coupling.

Some progress has been made on studying the effect of shift of phase center on coupling. Quantitative results are still being obtained. Experimental studies are being performed using two broadband antennas, a scimitar and a log-periodic monopole.

TABLE OF CONTENTS

LIST OF FIGURES

List of Figures (cont'd)

I

INTRODUCTION

Various specific technical efforts have been made on several topics of great importance in the reduction of electromagnetic coupling. Each one of these efforts is described in some detail in Chapter II of this report. Throughout all of the work during the past quarter, emphasis has been placed upon coupling reduction methods useful over a broadband of frequencies.

Further studies on the use of a bridge link to provide cancellation of unwanted electric fields in a receiving antenna due to a transmitting antenna located nearby have been conducted. As previously described, the bridge utilized to provide the destructive interference is one that must retain appropriate balance over a broadband of frequencies. Analytical studies have been advanced showing how the requirement of balance can be maintained over a range of frequencies. These studies also show the individual dependence for balance on both phase and attenuation. Although bridges are often considered as useful at one particular frequency, it has been possible to match the elements composing the bridge, including the direct path and the compensating path so that the parameters of these two paths tend to 'track' with frequency. It is quite remarkable that as the coupling between two antennas decreases with increasing frequency (associated with the change of spacing in terms of wavelength) it is possible to have, in the compensating bridge link, a combination of coaxial cable and waveguide which offers increasing attenuation as frequency increases. A certain appropriate choice of coaxial cable makes this possible. Information is presented on the limit of decoupling possible with a bridge for a given imbalance of power or phase angle. Because of these advances through analysis of the compensating bridge, it has been possible to make this method a truly broadband decoupling method. It is possible to obtain 15 db reduction in coupling over the entire X-band through the bridge link method if proper selection of compensating elements has been made.

Considerable new information is presented on the use of corrugated surfaces to reduce the coupling between two antennas. It has been found that corrugations surrounding a slot or horn antenna can be used to give substantial reduction in coupling over a frequency range of approximately 2:1. The best design of corrugations appears to have the corrugations 1/4 wavelength deep at the low frequency end of the band. Then as frequency increases, the depth of the corrugations will become greater in terms of wavelength. At a frequency where the depth of corrugations corresponds to 1/2 wavelength, the effect of the corrugations will be zero. If the frequency is such that the depth is somewhat greater than 1/2 wavelength, then the corrugations will be detrimental and the coupling level will be higher than it was without the corrugations being present. Other studies of corrugations have involved tapered corrugations or trenches. It has been found advantageous to have the bottom of the corrugations narrower than the width at the top of the corrugations. This means that the corrugations viewed from the top and looking vertically downward appear as a non-uniform transmission line. The analysis of tapered lines is applicable. With this tapering, the walls come up to the mounting surface almost in a sharp edge, with very little spacing from one trench to another. Such construction enables the coupling to remain lower as frequency is increased from the value corresponding to the 1/4 wavelength depth. This means that the coupling curve tends to be concave upwards, although finally, when the frequency corresponds to 1/2 wavelength, the influence of such corrugations is lost.

Studies have been made on the utilization of metal fences between two antennas in order to reduce coupling. The studies indicate the dependence of coupling reduction on the diameter of the rods used in such a fence. Thicker rods provide decoupling on a broader frequency basis. Although the metal fences used protruded from the metal mounting plane, it is possible to anticipate that the results achieved can be applied to a two antenna arrangement with a metal fence all immersed in a dielectric material; the whole configuration is then submerged so that the top dielectric interface is flush with the surrounding metal. It has been found that a metal foil fence is less effective in decoupling than is a metal picket fence.

Some further studies have been made on the near-field coupling associated with the outermost ends of the windings of Archimedean spiral antennas. There is some indication of the importance of eliminating standing waves on these distant ends. The experimental data so far obtained are not very definitive as far as establishing quantitatively the effects of these ends together with the orientation and closeness of these ends. The work is somewhat complicated by the need of an appropriate true resistor for a termination at the frequencies involved. The available experimental results for this report period are included in the technical description of this particular study.

Another basic study has been undertaken at the suggestion of the contract monitor, and this one is a study which can underlie the effectiveness of various types of coupling reduction. The study is that of the movement of a phase center of an antenna as frequency is varied. This is particularly important with certain types of broadband antennas. The effect of the shift of phase center, in itself, can represent a change in coupling as the frequency of operation changes, quite aside from the changing of spacing of antennas in accordance with the change in wavelength. Studies have been made of antennas which will be suitable to demonstrate the effects of the shift of phase center. Some information is given with results on the scimitar antenna. A second antenna which is essentially a log monopole array has been designed to continue the studies on the motion of phase center. It is recognized that the motion of phase center will also be involved in the maintainance of bridge balance in the bridge link compensation method.

II

EXPERIMENTAL STUDIES

2.1 Decoupling by Means of an RF Bridge

Analytic studies were made on the RF Bridge method of decoupling two slot antennas. This method basically cancels the electric field of the offending coupled signal with the electric field of the bridge signal. For perfect cancellation, which corresponds to infinite decoupling expressed in db, the phase angle θ between these two fields must be 180° and the amplitudes must be equal. This is easy to achieve at a single frequency. Unfortunately, in a broad-band system, some phase angle and amplitude errors will be introduced. The transmitted power varies with frequency and the coupled power falls 6 db per octave. The phase characteristics of the network elements also vary with frequency. It is important to know how much decoupling is possible for an RF Bridge with a given phase angle error and with a given amplitude error.

The following equation (2.1) was derived to answer this question.

$$D = m - 10 \log_{10} \left| 10^{m/10} + 1 + 2\sqrt{10^{m/10}} \cos \theta \right| \qquad (2.1)$$

D = decoupling in db,

m = difference between bridge and coupled power expressed in db,

θ = phase angle between bridge and coupled electric fields.

If m is greater than zero, the coupled power is greater than the bridge power, while if m is less than zero, the coupled power is less than the bridge power. This equation gives the additional isolation provided by the RF Bridge from the coupled power level. Typically for E-plane coupling, slot antennas spaced at 11.4 cm are already at a level of -30 db due to spacing. Thus if D = 20 db in Eq. (2.1), the total level would be -50 db using the RF Bridge for this case.

This equation is plotted in Fig. 2-1. The vertical axis is the decoupling expressed in db. The horizontal axis is the amount that the bridge power is less than the coupled power expressed in db. The phase angle error is the parameter designating individual curves. Thus for a 10° phase error and bridge power .2 db less than coupled power the decoupling is 15 db. If the bridge power were greater than the coupled power the same graph can be used. Merely read the proper decoupling value from Fig. 2-1 and subtract the amount by which the bridge power is greater from this value. For example consider the case that the bridge power is 1 db greater than the coupled power with 20° phase error. Reading Fig. 2-1 at 1 db one obtains D = 10 db. After subtracting the 1 db for the excess bridge power, D = 9 db.

An experiment was performed to check this equation. A bridge was unbalanced at a fixed frequency by specified amounts, both in amplitude and phase. The decoupling was recorded. The experimental data check Fig. 2-1 very closely.

A compensated decoupling bridge is being developed for use in X-band. The phase variations are compensated by using a "squeeze" section of waveguide. This is a piece of waveguide whose width can be varied. As the width changes, the propagation constant changes and the phase characteristic can be varied.

The amplitude variations are compensated by using a piece of RG-55 cable in the bridge link. The coupling power falls at 6 db per octave, and the attenuation of RG-55 cable increases with frequency at a similar rate in X-band. Thus, the bridge and coupled powers are more closely matched in level for a range of frequency. The X-band line stretcher, described in

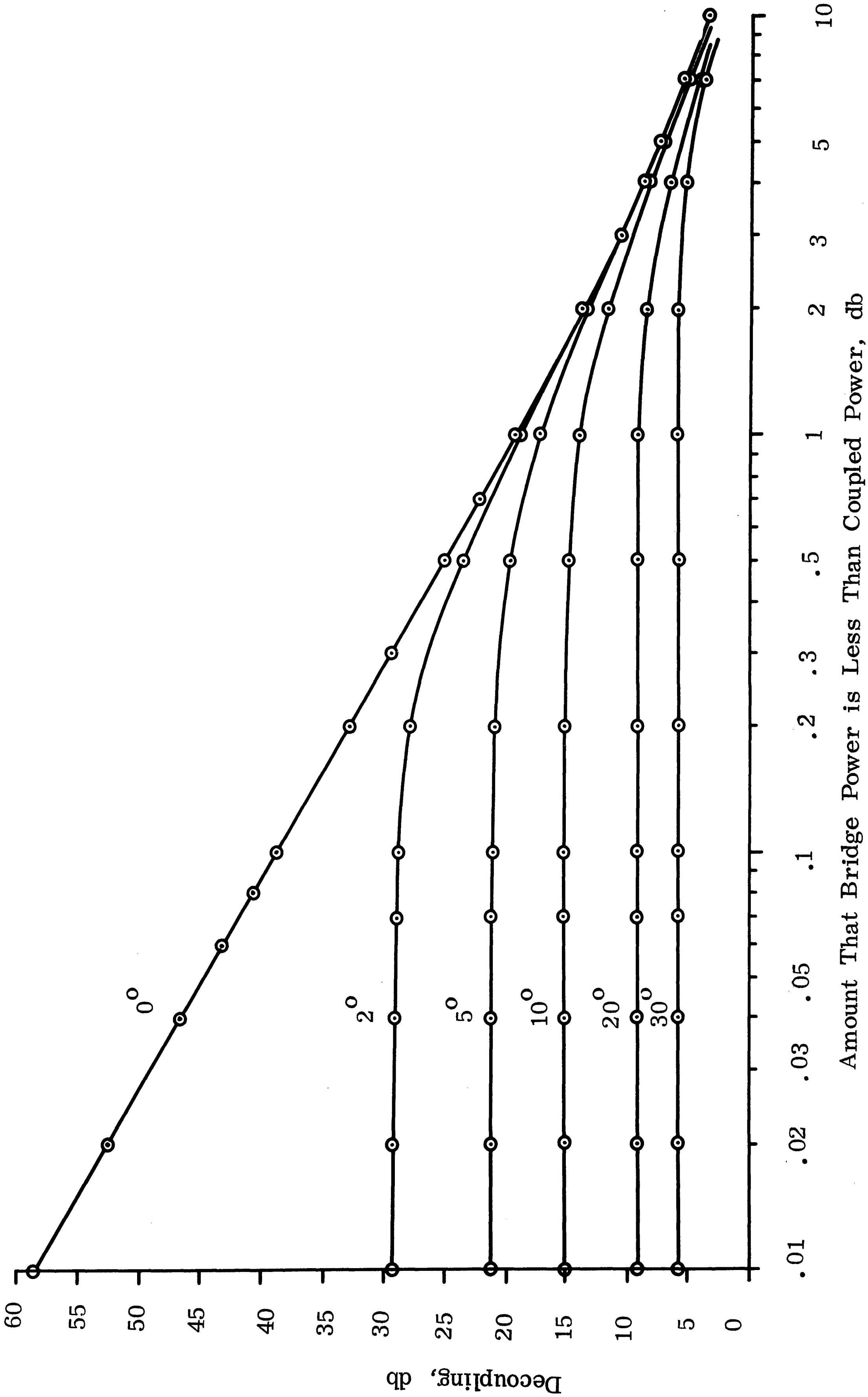


FIG. 2-1: DECOUPLING VERSUS DIFFERENCE IN POWER LEVELS WITH PHASE ERROR AS A PARAMETER.

Quarterly Report No. 5 (Lyon et al) under this contract has been abandoned. The power level coming out of it has proved to contain too many variations for easy compensation.

A sample decoupling chart for the compensated bridge is shown in Fig. 2-2. There is no abrupt drop in power at high frequencies. The decoupling is approximately 15 db over all of the X-band. Also, there is no problem in repeating experimental results.

2.2 Corrugations

2.2.1 Uniform Width Trenches.

The circumferential corrugations designed to reduce the coupling between two X-band slots have been tested and found effective over the entire X-band. The results have been reported in previous quarterly reports of this project. Since, in some applications, it is of interest to obtain high isolation between two systems operating over frequency ranges of 2 : 1 or even 3 : 1, the measurements have been extended to the waveguide band immediately above the X-band covering the range of frequencies from 12.4 GHz to 18 GHz. One of the names by which this frequency range is known, is K_u-band and it will be referred to as such in the following.

The same slots fed by X-band waveguide were used in both frequency ranges, with tapered transitions to the K_u-band waveguide when the K_u-band system was replacing the X-band system. Figure 2-3 shows the effect of one set of circumferential corrugations on the E-plane coupling between two slots in the X- and K_u-band. The maximum coupling between the two slots in a 2 : 1 frequency range (8.7 GHz to 17.4 GHz) is reduced by 11 db (from -29.5 db to -40.5 db). It is expected that the amount of decoupling (in db) would be doubled if both slots were surrounded by corrugations.

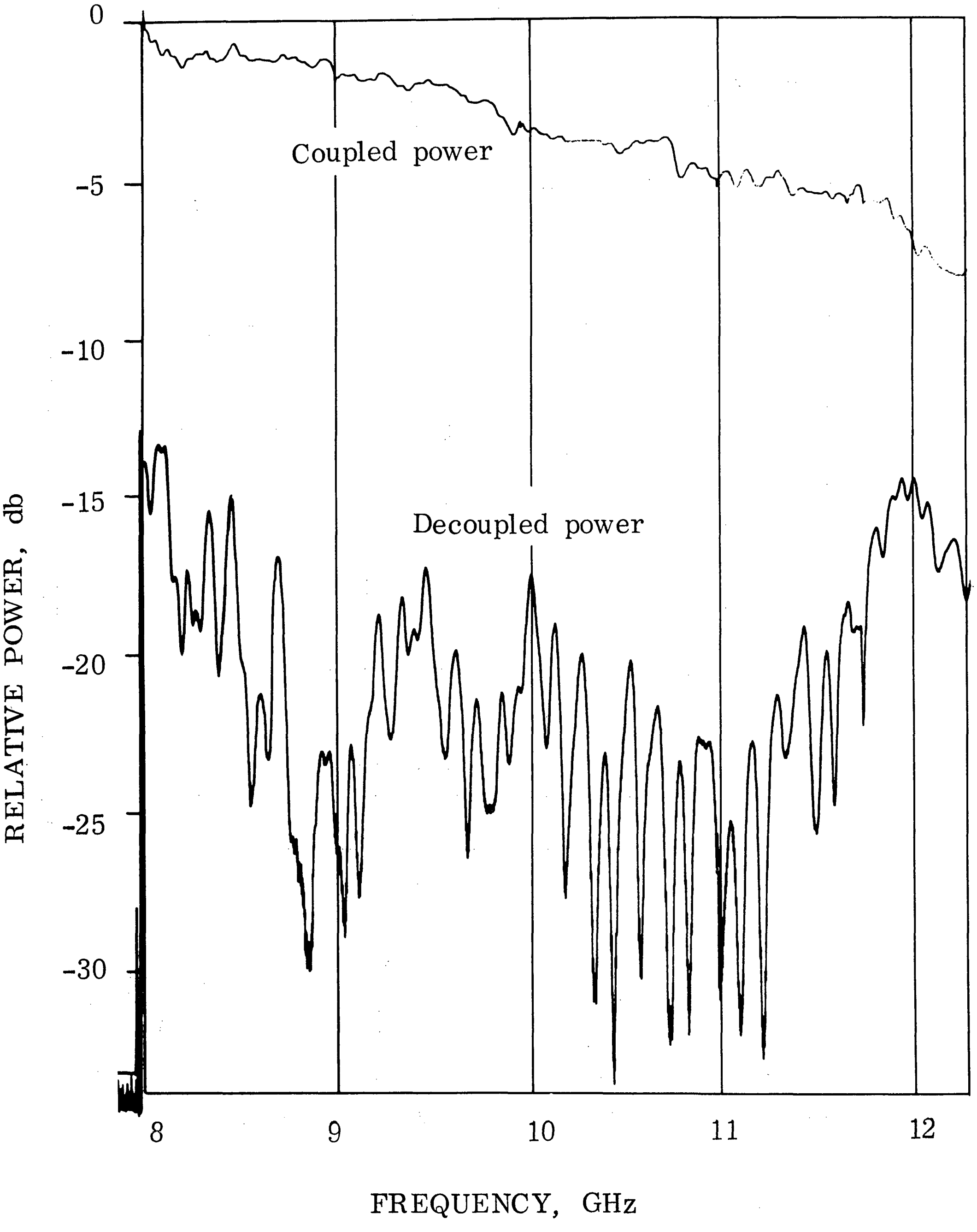


FIG. 2-2: SLOT ANTENNA DECOUPLING BY A COMPENSATED BRIDGE.

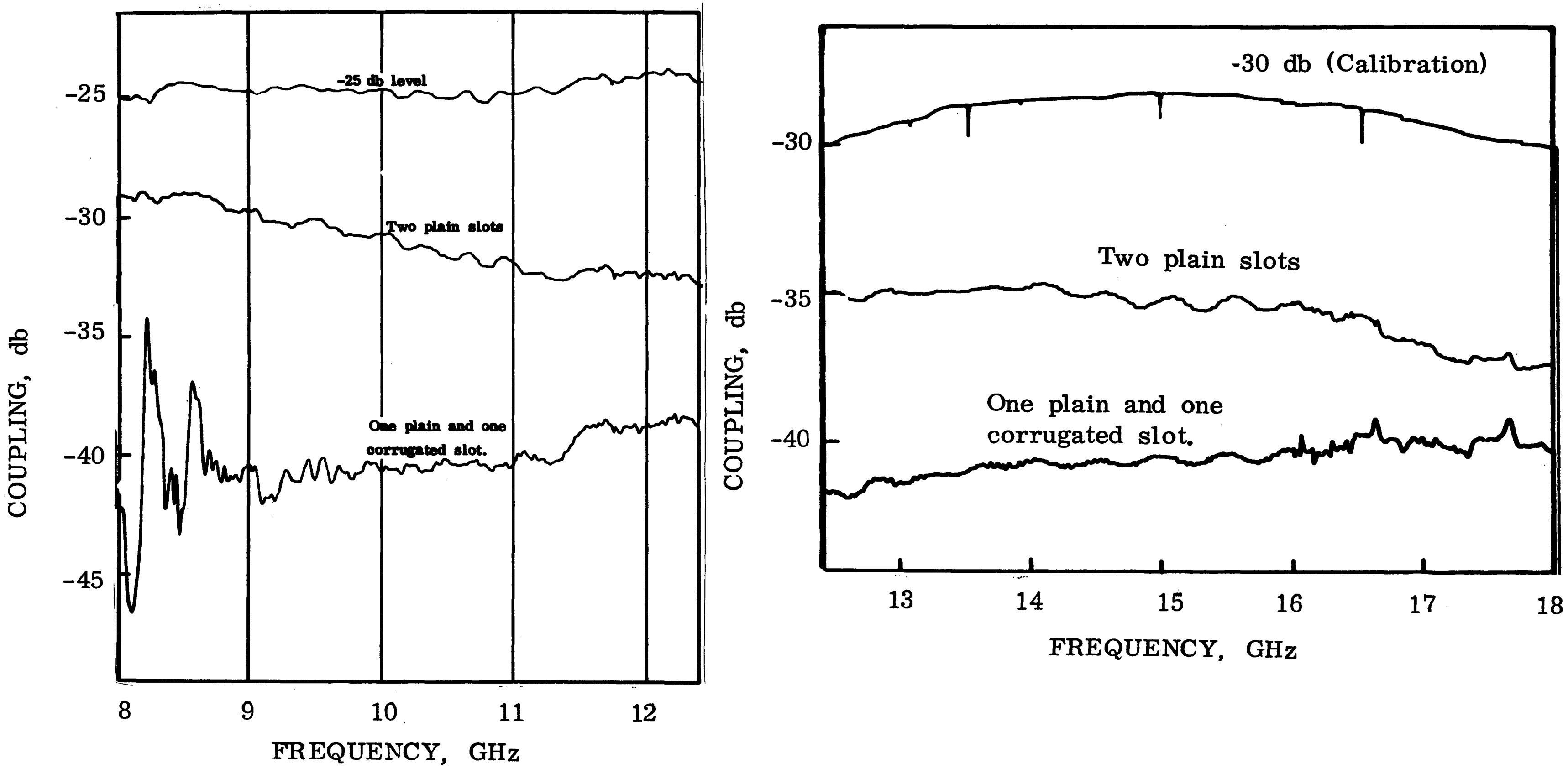


FIG. 2-3: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 11.4 CM. (Corrugations R-1).

It has been shown (Lyon et al 1967, 7692-5-Q) that a capacitive surface impedance is required in order to reduce the coupling between two antennas mounted on a common surface, while an inductive surface impedance has the opposite effect. In the case of a surface wave propagation normal to the corrugations the surface reactance is (by an approximation neglecting the high order modes which are rapidly attenuated in a narrow trench)

$$X_s = \frac{b}{b+t} \eta \tan kd, \tag{2.2}$$

where

b = trench width,
t = wall thickness,
d = trench depth,
k = propagation constant,
η = wave resistance in vacuum = 376.7 Ω .

Let

$$y = \frac{f}{f_o} = \frac{\lambda_o}{\lambda}, \tag{2.3}$$

$$d = \frac{\lambda_o}{4} \tag{2.4}$$

where λ_o is the cut-off wavelength. Here "cut-off" means the frequency at which X_s changes from positive to negative. Then from (2.2) it follows that when $y = 2$ or $f = 2f_o$, X_s becomes positive again and the decoupling at this point should be zero. This, however, is not the case in Fig. 2-3 which shows a decoupling of 3 db at $2f_o$. The apparent disagreement is due to the fact that in the case of a slot surrounded by circumferential corrugations, the surface wave is not always normal to the trenches, an assumption made in the derivation of Eq. (2-2). This turns out to be beneficial to decoupling in that it extends the frequency range over which the corrugations are useful.

2.2.2 Tapered Wall Trenches

It has been thought of interest to investigate the properties of a different type of corrugated surface, from the one shown in Fig. 2-4a. One motive in this investigation is that circular trenches of this profile can be easily machined in contrast to the parallel wall trenches.

Let us consider a single tapered trench as in Fig. 2-4b which extends to infinity in the z and r directions. This can be viewed as a non-uniform transmission line in the r-direction. The impedance of this line can be determined as follows.

Let $\bar{E} = E_\phi \hat{\phi}$ and assume no dependence upon the z-coordinate. Then Maxwell's equations reduce to ($j\omega t$ time dependence assumed and suppressed)

$$\left.\begin{aligned} j\omega\epsilon E_\phi &= -\frac{\partial H_z}{\partial r} , \\ j\omega\mu r H_z &= -\frac{\partial (r E_\phi)}{\partial r} . \end{aligned}\right\} \qquad (2.5)$$

The solution in terms of the Hankel functions is:

$$\left.\begin{aligned} H_z &= A H_0^{(1)}(kr) + B H_0^{(2)}(kr) \\ E_\phi &= -j\eta \left[A H_1^{(1)}(kr) + B H_1^{(2)}(kr) \right] . \end{aligned}\right\} \qquad (2.6)$$

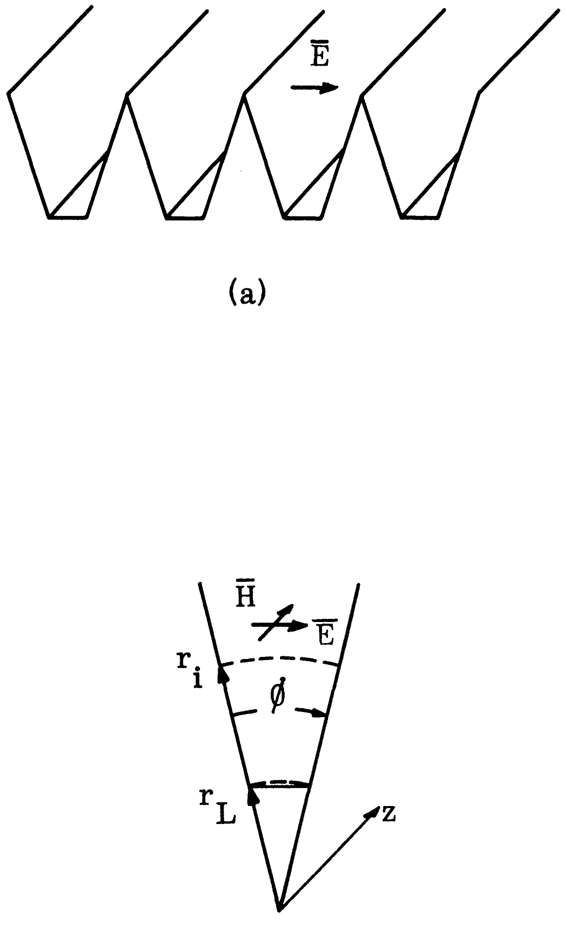


FIG. 2-4: CORRUGATIONS WITH TAPERED TRENCHES

Let

$$\left.\begin{aligned} H_n^{(1)}(x) &= H_n(x)\,e^{j\theta_n(x)} \\ \text{and} \qquad & \\ H_n^{(2)}(x) &= H_n(x)\,e^{-j\theta_n(x)} \end{aligned}\right\} \tag{2.7}$$

where

$$\left.\begin{aligned} H_n(x) &= \left[J_n^{(2)}(x) + Y_n^2(x)\right]^{1/2} \\ \theta_n &= \tan^{-1}\left[\frac{Y_n(x)}{J_n(x)}\right] \quad . \end{aligned}\right\} \tag{2.8}$$

Then Eq. (2.6) can be written:

$$\left.\begin{aligned} H_z &= H_o(kr)\left[A\,e^{j\theta_o(kr)} + B\,e^{-j\theta_o(kr)}\right] \\ E_\phi &= -j\eta\,H_1(kr)\left[A\,e^{j\theta_1(kr)} + B\,e^{-j\theta_1(kr)}\right] . \end{aligned}\right\} \tag{2.9}$$

The impedance of the line, looking towards the apex, is defined as

$$Z_i = -\left.\frac{E_\phi}{H_z}\right|_{r=r_i} \quad . \tag{2.10}$$

Let also

$$Z_L = -\left.\frac{E_\phi}{H_z}\right|_{r=r_L} \quad . \tag{2.11}$$

Substituting (2.9) and (2.11) into (2.10) one finally obtains the expression

$$Z_i(r_i) = Z_o(kr_i) \cdot \frac{Z_o(kr_L)\sin\left[\theta_1(kr_L) - \theta_1(kr_i)\right] + jZ_L\sin\left[\theta_o(kr_L) - \theta_1(kr_L)\right]}{Z_L\sin\left[\theta_o(kr_L) - \theta_o(kr_i)\right] - jZ_o(kr_L)\sin\left[\theta_1(kr_L) - \theta_o(kr_i)\right]} \tag{2.12}$$

where

$$Z_o(kr) = \eta\,\frac{H_1(kr)}{H_o(kr)} \quad . \tag{2.13}$$

For $Z_L = 0$ (short at $r = r_L$) this reduces to:

$$Z_i(r_i) = j\,Z_o(kr_i)\,\frac{\sin\left[\theta_1(kr_i) - \theta_1(kr_L)\right]}{\sin\left[\theta_o(kr_i) - \theta_1(kr_L)\right]} \quad . \tag{2.14}$$

Assuming that the line is narrow enough so that the high order modes are attenuated rapidly one may still use (2.14) to predict the input impedance of a tapered trench whose walls terminate at $r = r_i$. Then for a series of parallel trenches as shown in Fig. 2-4a considerations similar to the ones used to derive Eq. (2.2) yield for the surface reactance

$$X_s(r_i) = \frac{b}{b+t}\,Z_o(kr_i)\,\frac{\sin\left[\theta_1(kr_i) - \theta_1(kr_L)\right]}{\sin\left[\theta_o(kr_i) - \theta_o(kr_L)\right]} \tag{2.15}$$

where b and t have the same meaning as in Eq. (2.2) and are measured at the aperture. From (2.15) it is seen that the reactance changes from positive to negative when:

$$\theta_o(kr_i) - \theta_1(kr_L) = m\pi \;;\quad m = 1,\ 2,\ \dots\ . \tag{2.16}$$

The solution of (2.16) with $k = 2\pi/\lambda_o$ is plotted in Fig. 2-5. It is seen that this solution tends asymptotically to the corresponding solution of Eq. (2.2)

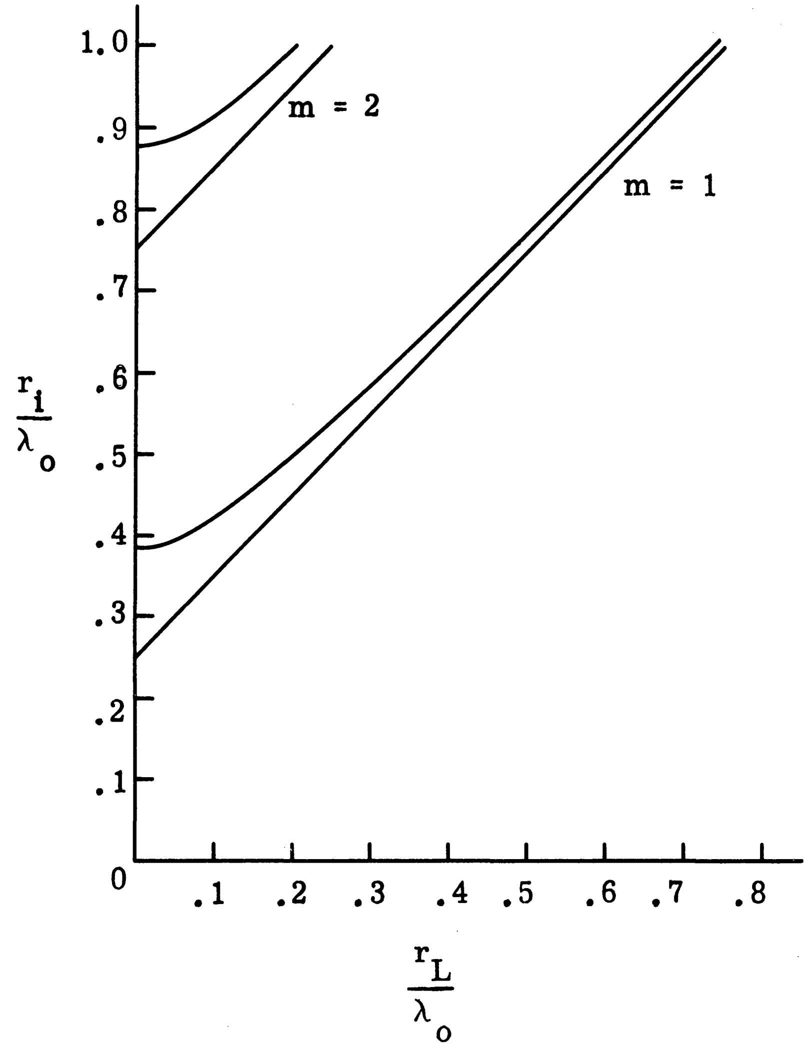


FIG. 2-5: SOLUTION OF $\theta_o(2\pi r_i/\lambda_o) = \theta_1(2\pi r_L/\lambda_o) + m\pi$ and ASYMPTOTE $r_i - r_L = (2m - 1)\lambda_o/4$.

as expected. With the help of Fig. 2-5 one can determine r_i once r_L and λ_o have been chosen.

When $r_L = 0$ from Eq. (2.15) it is seen that X_s will become positive when $f \cong 1.6\,f_o$. In general for $0 \leq r_L \leq \infty$ the crossover point will be at $1.6\,f_o < f < 2.0\,f_o$. So the tapering restricts somewhat the effective range of the corrugations but on the other hand it is capable of offering larger decoupling. near the "cut-off" frequency f_o.

Three sets of circumferential corrugations with different cut-off frequencies and profiles were manufactured and tested. Each set had six trenches; their dimensions are shown below in Table II-1.

TABLE II-1: Dimensions of Tapered Corrugations. (In cm)
(Tolerance of r_L, r_i: $\pm$ 0.03 cm).

	r_L	r_i	b	t
Set R4	0.25	0.97	0.50	$\leq$ 0.01
Set R5	0.05	0.98	0.50	$\leq$ 0.01
Set R6	0.36	1.36	0.50	$\leq$ 0.01

The corrugations were surrounding an X-band slot, being flush with the ground plane. The E-plane coupling with and without corrugations around one slot are shown in Figs. 2-6 and 2-7 for two of the above mentioned sets.

The curve of Fig. 2-5could be used to predict the cut-off frequencies for each set of corrugations if a monopole antenna were used at the center so that the surface wave would be normal to the trenches. With a slot at

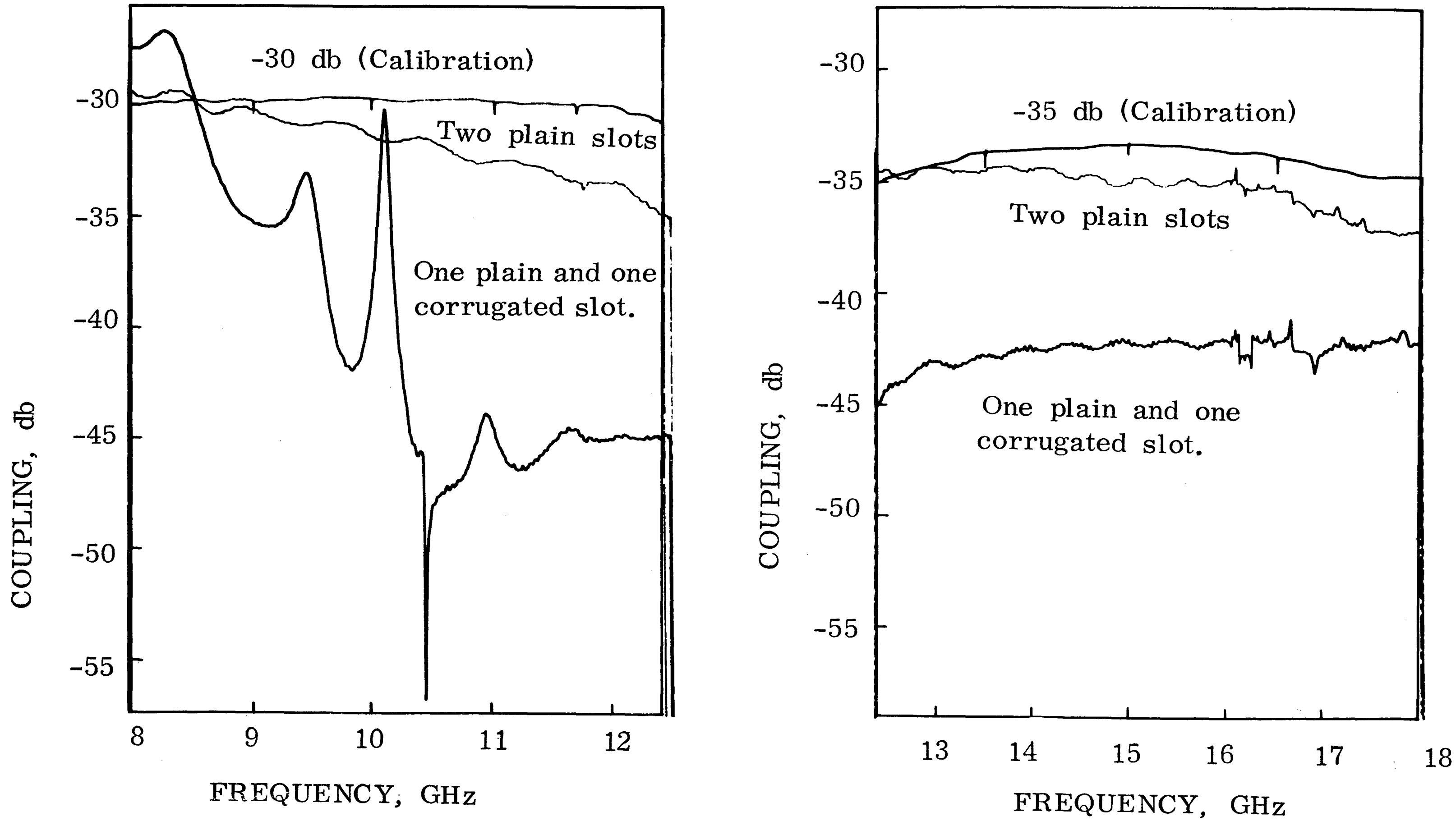


FIG. 2-6: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 11.4 CM (Corrugations R-5).

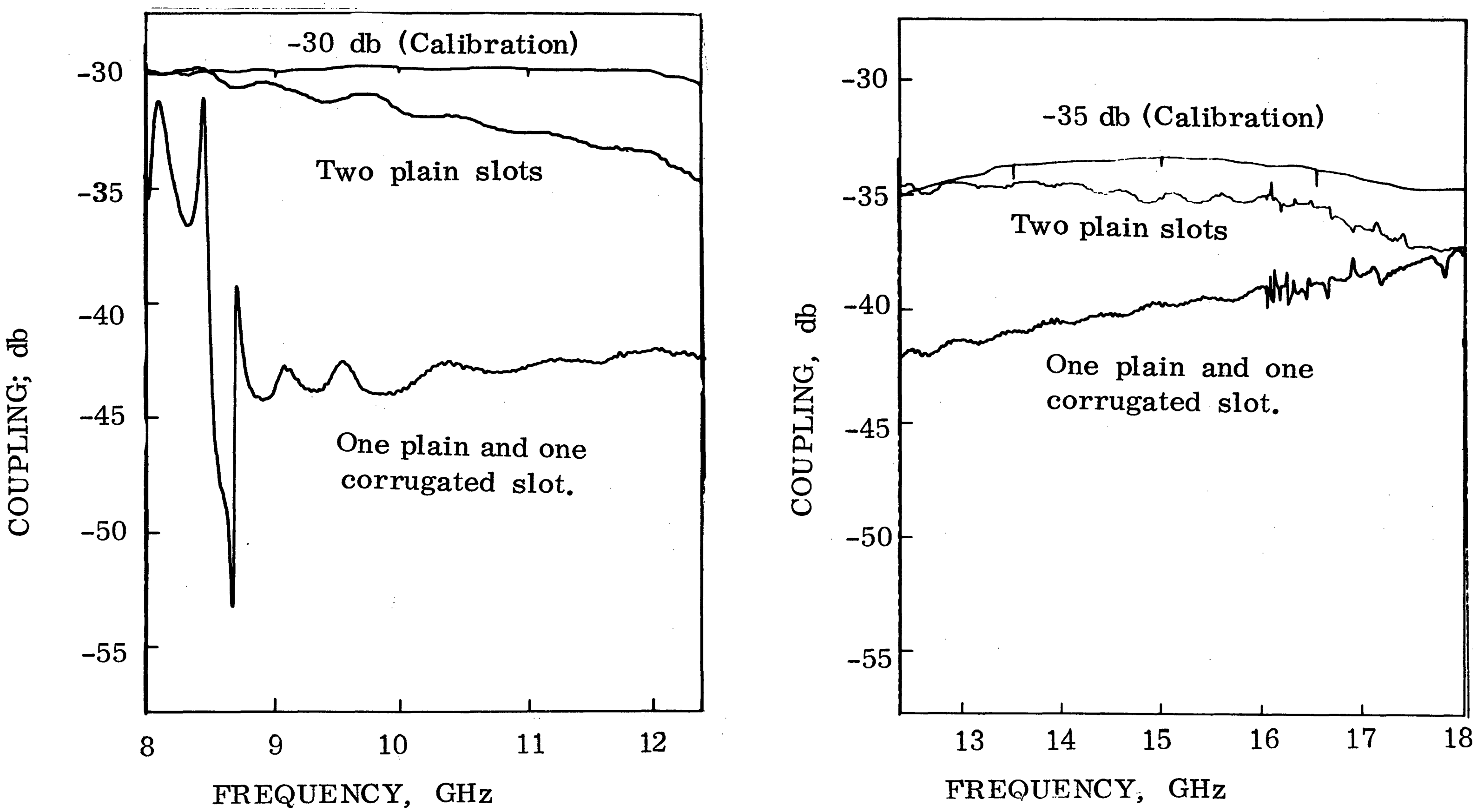


FIG. 2-7: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 11.4 CM (Corrugations R-6).

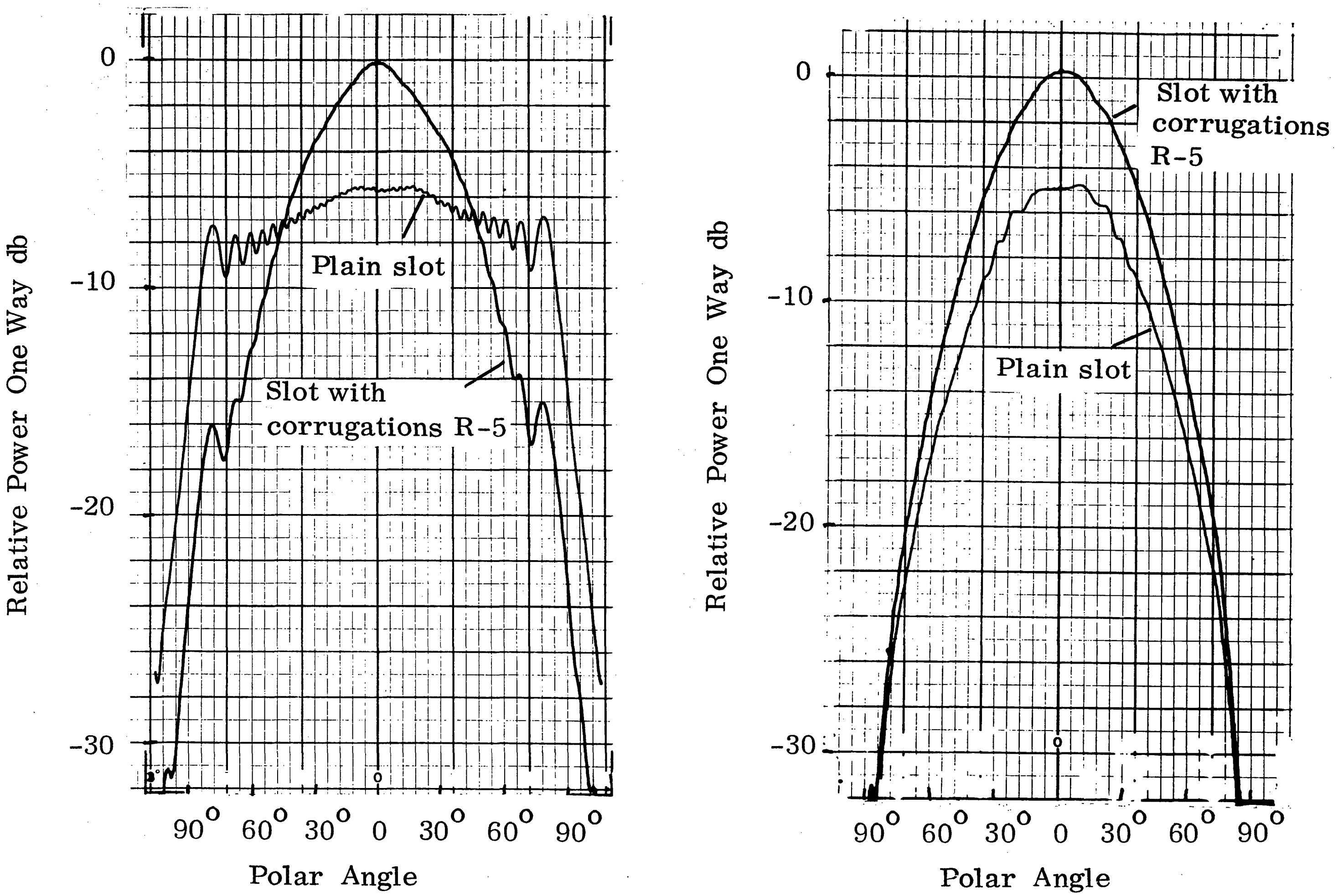
the center Eq. (2.16) can only be used for an estimate within 2 to 4 percent. Estimates using Eq. (2.16) are always high.

The E-plane decouplings obtained over a 1.5:1 frequency range with corrugations R-4, R-5 and R-6 are 11 db, 12 db and 11 db respectively. Only corrugations R-6 were measured over a 2:1 frequency band and there the decoupling obtained was 8 db. It should be noted that twice as much decoupling (in db) is expected when both antennas are surrounded by similar corrugations. Also somewhat greater decoupling can be obtained by increasing the number of trenches.

Typical radiation patterns for a slot surrounded by tapered-trench corrugation (R-5) in a 2 ft by 3 ft ground plane are shown in Fig. 2-8. These patterns show that the sidelobe decrease is accompanied by a considerable increase in antenna gain (5.5 db in this case).

2.3 Fences

The study of fences with respect to reducing the coupling between two antennas continued on an experimental basis. The term "fence" as used here means a set of short wires erected on the metal plane between the antennas of the interfering systems. The fence is described by three parameters, height (h), spacing between adjacent wires (s) and wire diameter (2a). In this work two slot antennas oriented for E-plane coupling were used and the fence was always placed normal to the line joining the two slots. In this case only one parameter is needed to specify the position of the fence on the ground plane and as such the distance between the fence and the slot center (d) was chosen.



(a) E-plane

(b) H-plane

FIG. 2-8: E- AND H-PLANE RADIATION PATTERNS FOR A SLOT IN 2 FT BY 3 FT METAL PLANE AT 12 GHz.

The results obtained indicate that the best choice of parameters is $h \cong \lambda/2$, $s \cong \frac{3}{8}\lambda$, $\lambda < d < \frac{3}{2}\lambda$. The wire diameter, 2a, is also an important factor. Three different diameters were used $\frac{1''}{32}$, $\frac{1''}{16}$ and $\frac{1''}{8}$. Of these the larger diameter was found to be distinctly inferior to the other two of which the $\frac{1''}{16}$ seemed to be the best. A comparison chart is shown in Fig. 2-9. The chart shows the original level of coupling between two plain slots and the reduction of same with the addition of each one of the following (one at a time, all at the same distance $d = 3.5$ cm): a) a solid metal wall, height 1.7 cm length 7.3 cm; b) a fence of the same total length and height, $s = 1.2$ cm, $2a = \frac{1''}{8}$; c) as in (b) except for $2a = \frac{1''}{16}$.

In another series of experiments two fences were used, one near each slot of a symmetrical arrangement. Figure 2-10 shows that in this case a decoupling of 35 db was obtained over a range of 2.5 GHz and it seems only a matter of adjustment of the different parameters involved to extend this over at least the entire X-band. The geometry of this arrangement is shown in Fig. 2-11. The pertinent parameters are: $d = 3.5$ cm, $h = 1.8$ cm, $s = 1.2$ $2a = \frac{1''}{16} = 0.159$ cm.

The large reduction of coupling and the relatively simple structure of the fences make this method particularly attractive. A fence could also be considered in a flush mounted modification. This could be achieved by having the two antennas involved in interference mounted flush with a recessed part of the ground surface. The fence could also be mounted on this recessed part with the tops of the pins just even with the contour of the main ground surface. The recess could be filled with dielectric material. The exterior surface of the dielectric would be a smooth continuation of the metal surface of an aero-

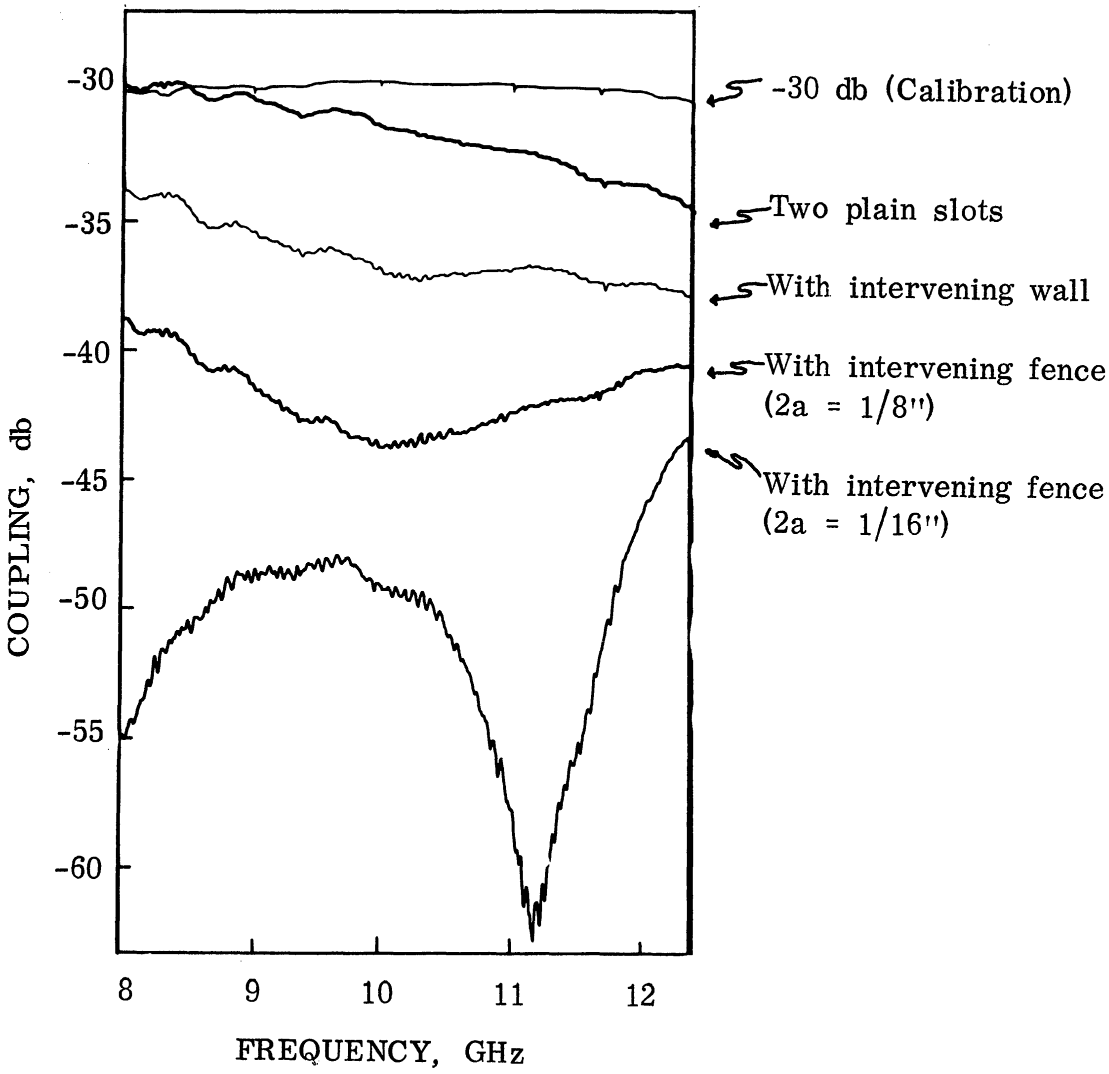


FIG. 2-9: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 11.4 CM WITH VARIOUS TYPES OF FENCES.

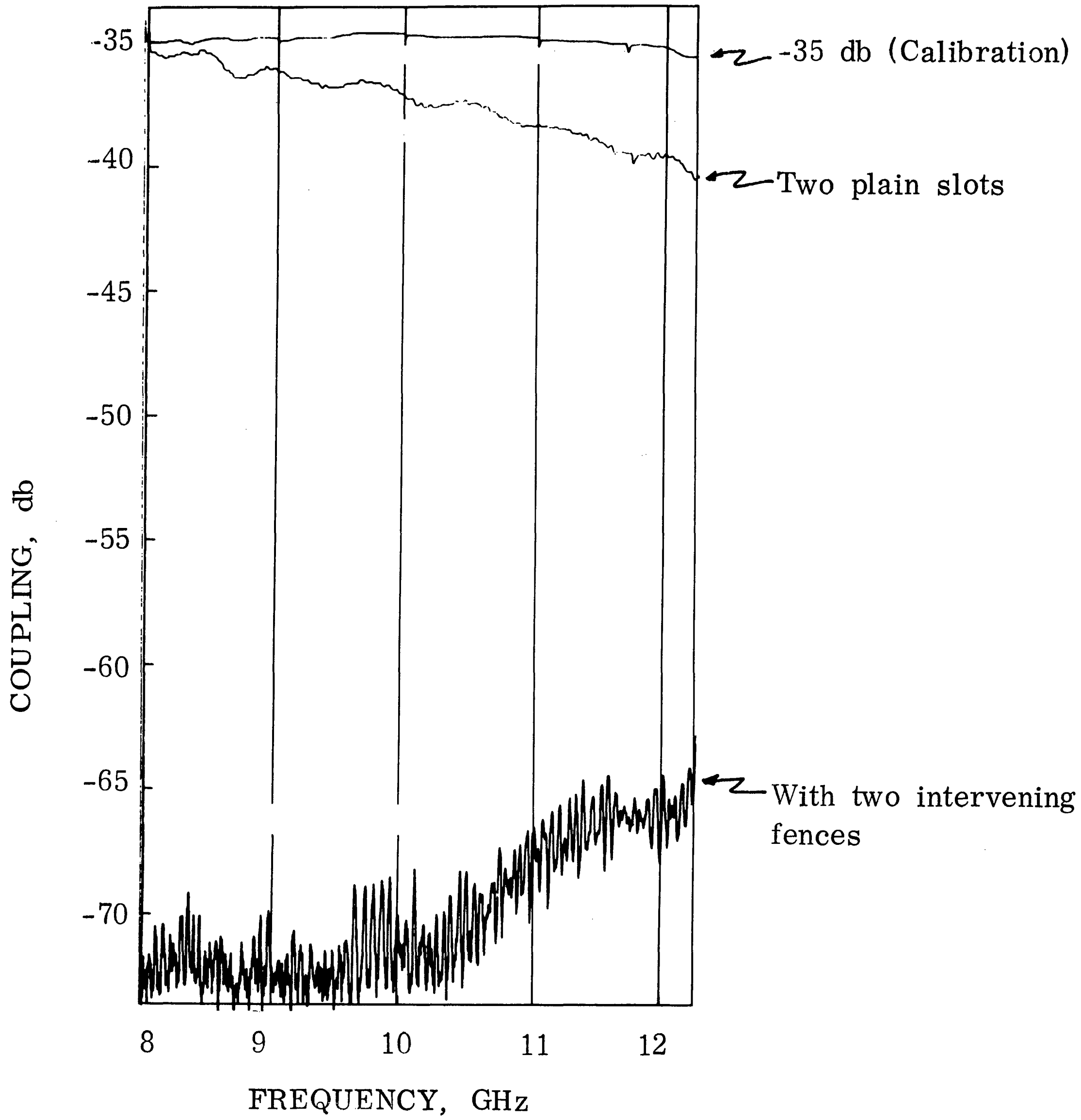


FIG. 2-10: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 22.8 CM WITH AND WITHOUT FENCES.

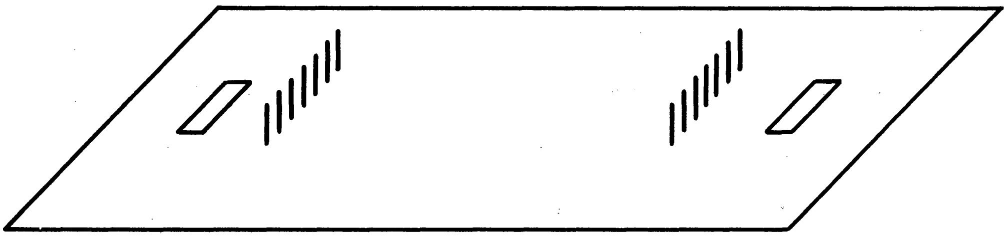


FIG. 2-11: GEOMETRY OF SLOTS WITH TWO FENCES.

space vehicle. Since the required half wavelength of the pins would be on the basis of the dielectric permittivity, the depth of the recessed part would be relatively small.

A fence can be combined with another decoupling method to give additional isolation. One slot was surrounded by corrugations (Set R-6) and a fence ($h = 1.7$ cm, $s = 1.2$ cm, $2a = 1/16''$) was placed in front of a second slot ($d = 3.5$ cm). The decoupling obtained when each method was used individually and when the two were combined is shown in Fig. 2-12. The combination resulted in an increase of isolation between the two systems by 27 db over a 2.7 GHz range. The frequency range over which the combination is effective could easily be extended to a full waveguide band by adjusting the fence parameters so that the decoupling regions of the individual methods overlap. Since the two methods have different attenuation and phase characteristics the combination of the two does not necessarily result in an algebraic addition of the isolations offered by each method separately. This is more apparent at the position of the nulls.

2.4 Effect of Termination on Near-Field Spiral Coupling

Attempts have been made to investigate the near-field coupling of spiral antennas by measuring the current distribution along the spiral windings. Since the current probing facility was designed for antennas in the VHF and UHF range, only approximate current distributions can be obtained for frequencies higher than 1 GHz. Also, since the apparatus allows probe motion in only one direction it was possible to measure the distribution diametrically only. Figure 2-13 shows the current distribution of an AEL spiral as measured with a small loop probe. The existence of the active region is readily seen from each part of the figure. Peaks and valleys exist beyond the active regions, corresponding respectively to spiral windings and the spaces between windings.

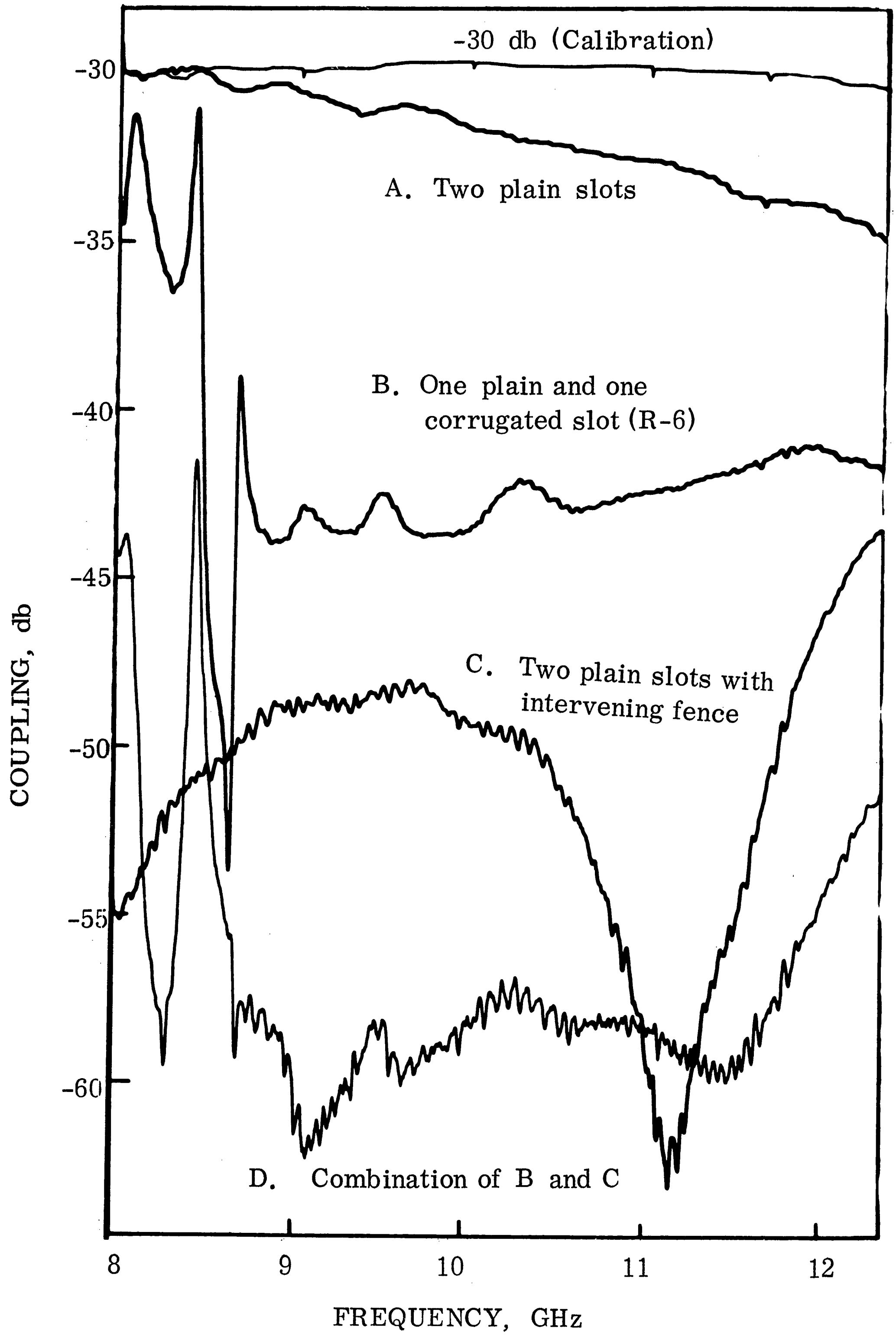
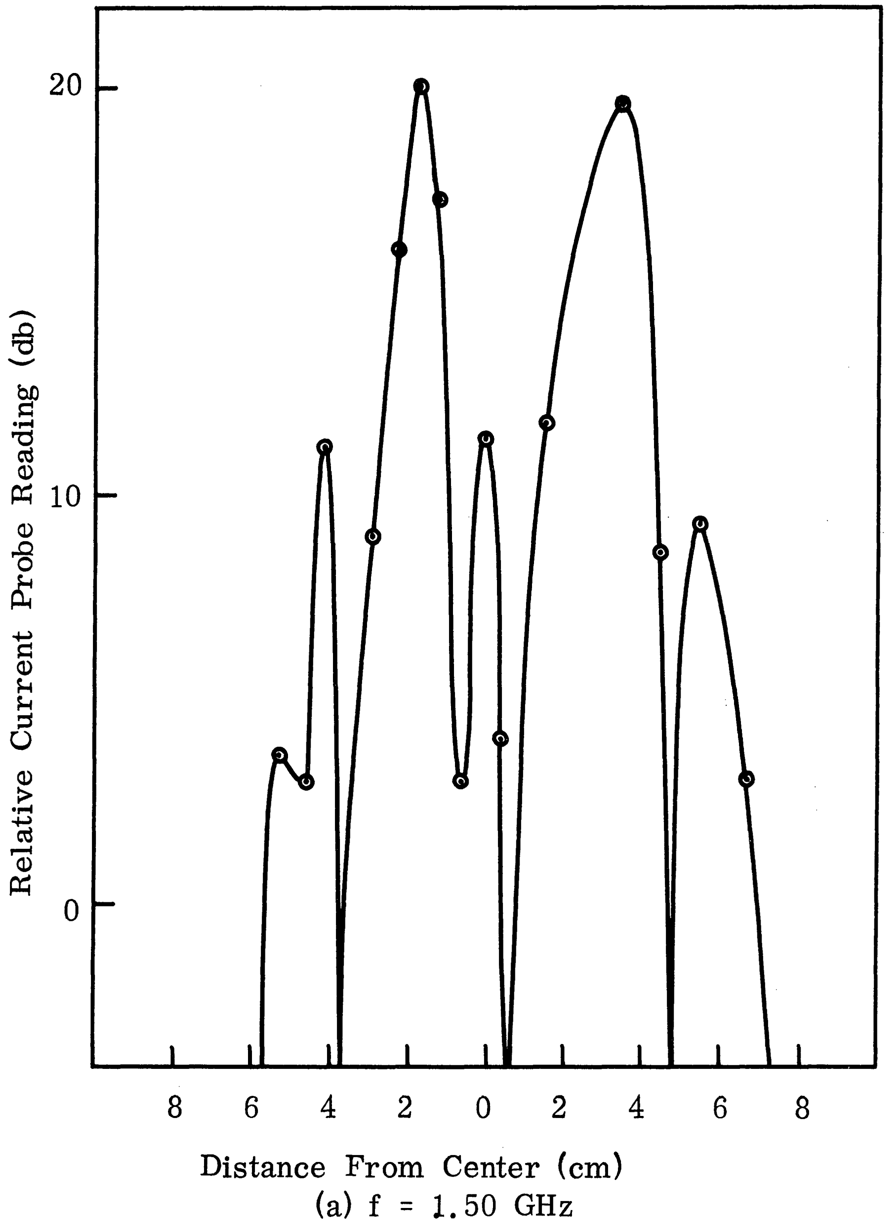
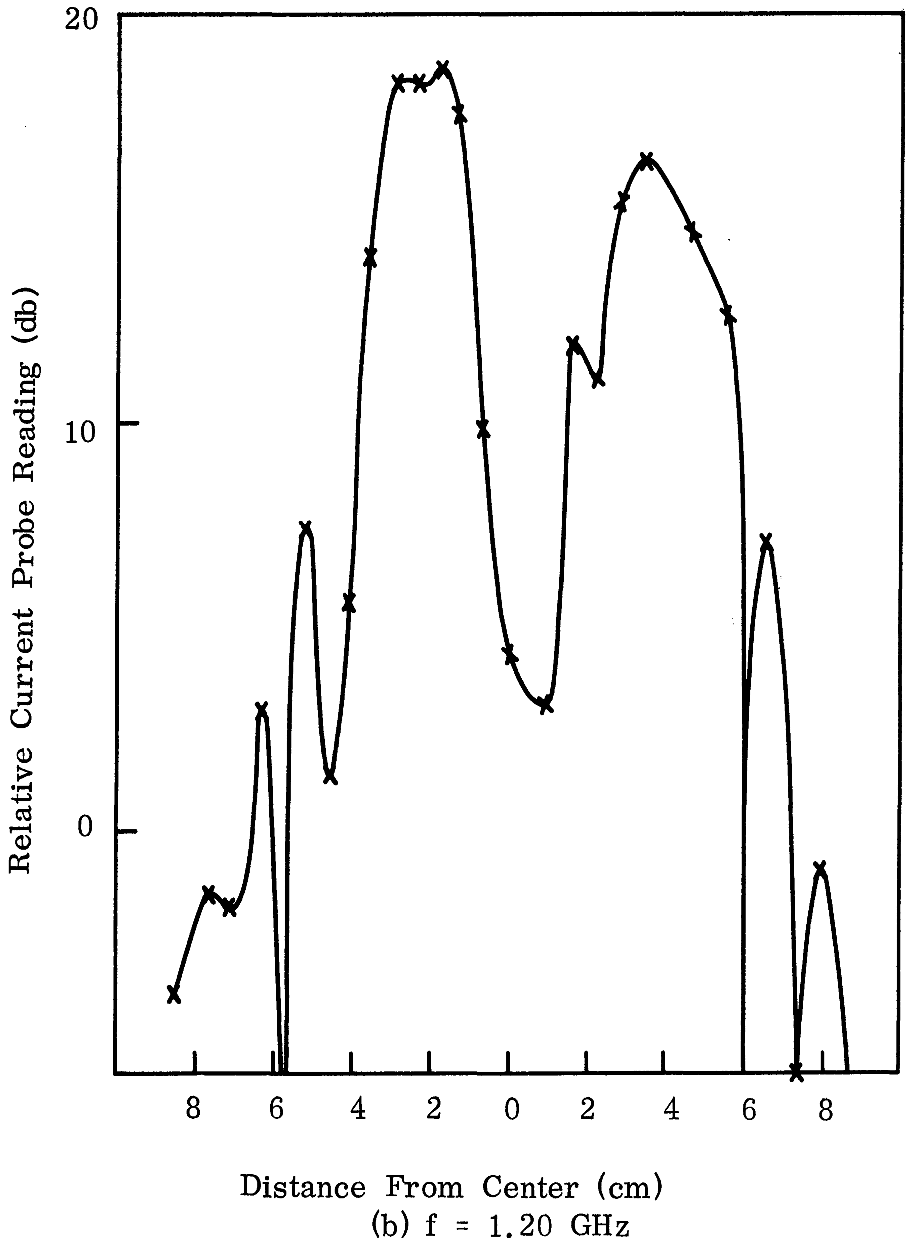


FIG. 2-12: E-PLANE COUPLING VS FREQUENCY FOR TWO SLOTS SPACED 11.4 CM.



(a) f = 1.50 GHz

FIG. 2-13a: DIAMETRICAL CURRENT DISTRIBUTION OF AN AEL SPIRAL ANTENNA. The Diameter of the Spiral is about 14.5 cm.



(b) f = 1.20 GHz

FIG. 2-13b: DIAMETRICAL CURRENT DISTRIBUTION OF AN AEL SPIRAL ANTENNA. The Diameter of the Spiral is about 14.5 cm.

Such distinction is not possible toward the central region of the spiral, owing to the very small width of and the very small separation between the windings there (as compared with the dimension of the loop probe). It is seen that the active region does not quite reach the outermost windings even at very low frequencies (the specified low frequency limit of this antenna is 1.0 GHz). Thus no major change in current distribution and radiation pattern can be expected from terminating the ends of this spiral by resistors or other means.

For S-band square spirals, even approximate current distribution could not be obtained by the current probes available. The effect of termination of this type of antenna is studied through the measurements of SWR and coupling. Figure 2-14 shows the coupling between two closer-packed square spirals when the transmitting antenna is terminated with two carbon resistors. Figure 2-15 shows the SWR of the transmitting spiral 1L with and without a 1,000 Ω (d-c) carbon resistor termination. No change in either SWR or coupling has been observed throughout the S-band by terminating the ends of the spiral 1L within the accuracy limitations of the measurements. This is again an indication that at most a current of insignificant magnitude can reach the ends of the spiral and thus both the SWR and the coupling are quite independent of the end condition, whether the ends are open-circuited or terminated.

2.5 Effect of Phase Center Shift on Coupling

To find the effect of phase center shift on coupling two types of antennas with phase center shift were used:

(a) Scimitar antenna

(b) Log-periodic monopole over the ground plane.

In the scimitar antenna case use was made of a simple coupling experiment between two different types of scimitar antennas and a resonant monopole, to determine

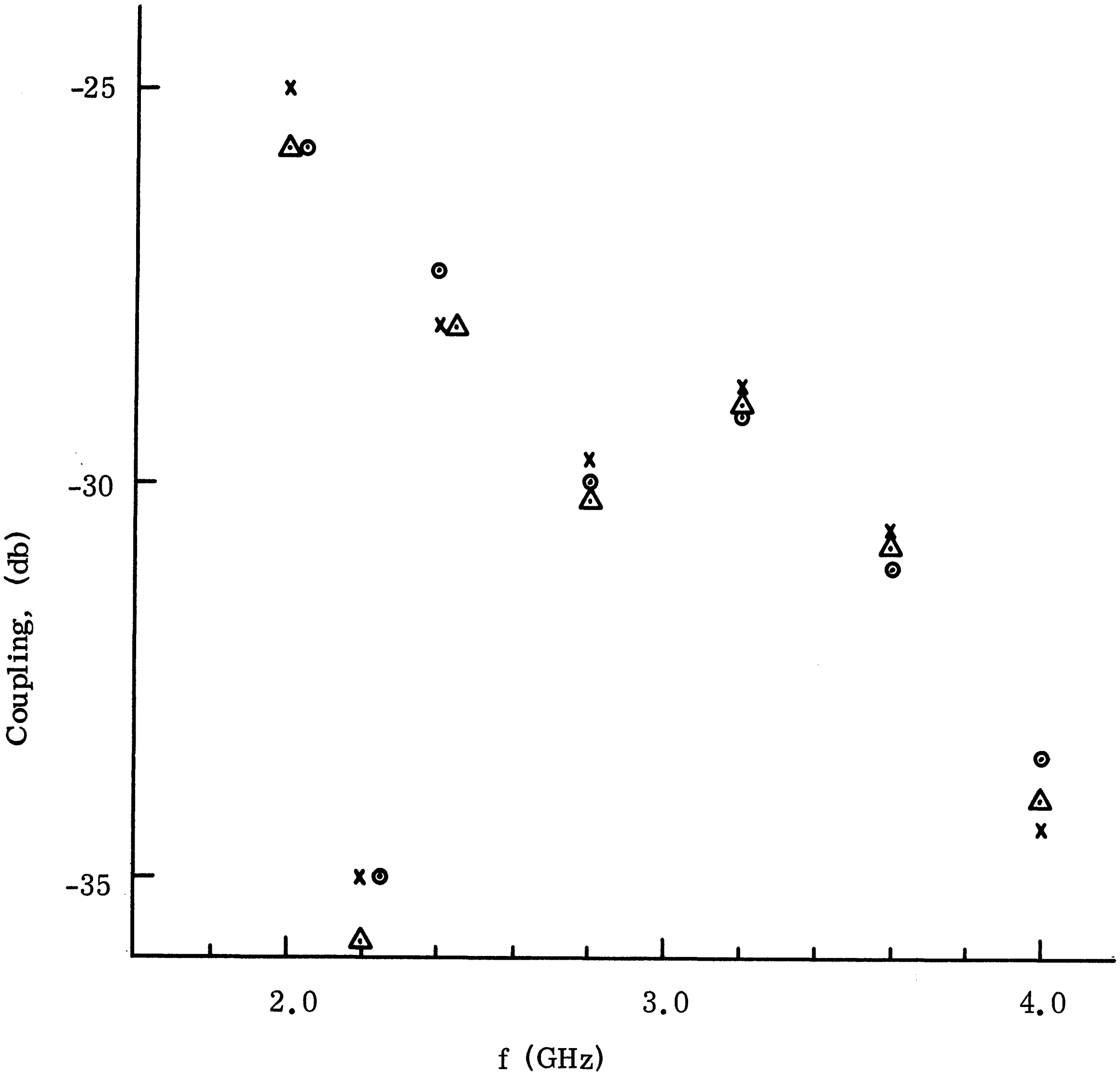


FIG. 2-14: COUPLING BETWEEN TWO CLOSED-PACKED SQUARE SPIRALS 1L AND 2L. (-x-) Plain (-⊖-) 1L Terminated with 56 Ω Carbon Resistors (-△-) 1L Terminated with 1,000 Ω Carbon Resistors.

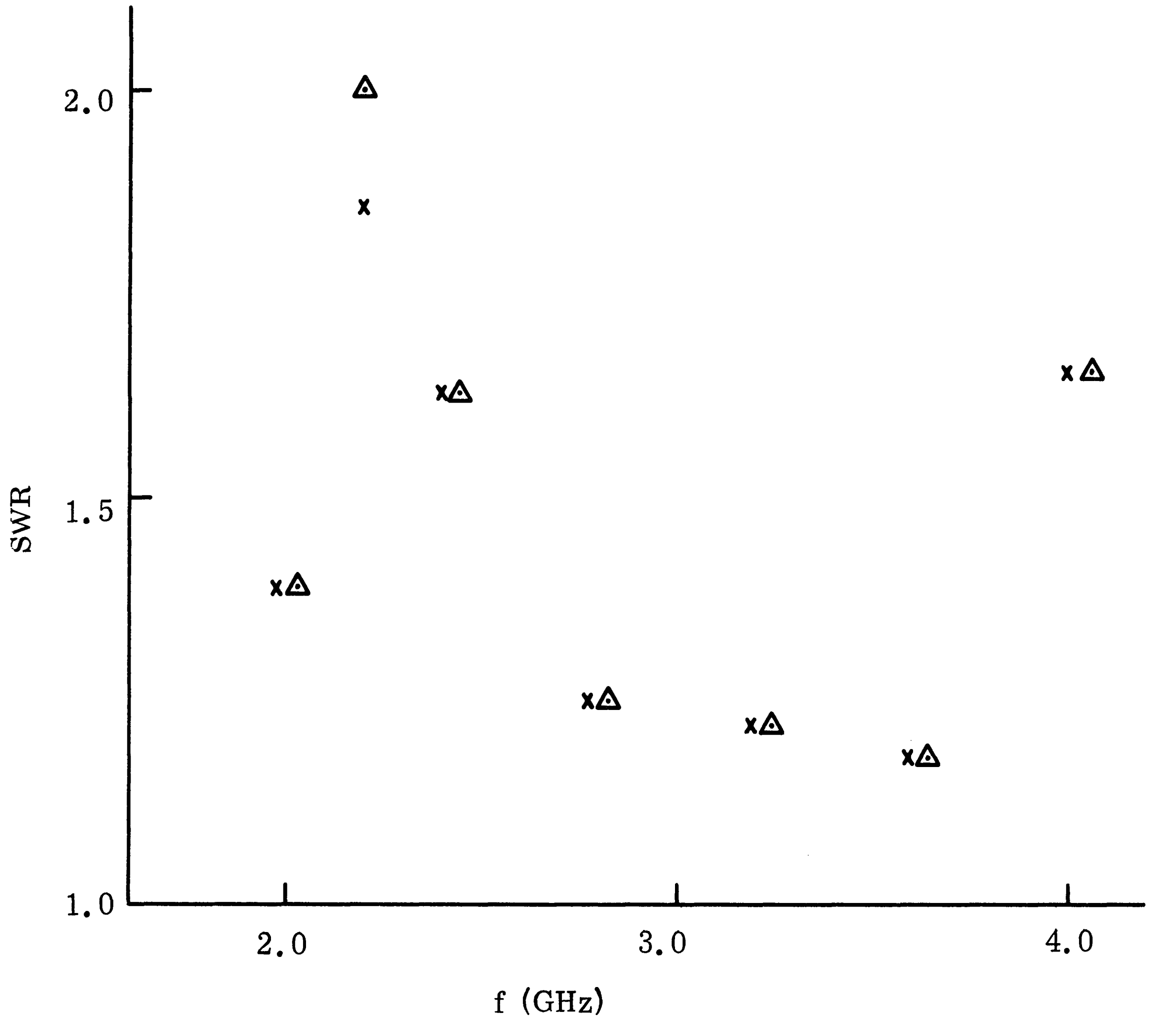


FIG. 2-15: EFFECT OF TERMINATION ON SWR OF SPIRAL ANTENNA 1L. (x) Plain (Δ) Terminating Both Ends with 1,000 Ω Carbon Resistors.

the horizontal location of the phase center and its change with the variation of frequency. Also, the coupling experiments were performed for different values of distances between the scimitar and the resonant monopole.

The idea of the experiment was as shown in Fig. 2-16. Assume one phase center and that it is located at a distance L from the feeding axis. Let x be the distance from the phase center to the monopole. C is the distance between feeds. The distance from the phase center to the monopole will be C - L for case "a". Then rotate the scimitar 180° around its feed point as shown in the Fig. 2-16 for case "b". The distance from the phase center to the monopole will be C + L for case "b". Suppose that the difference in coupling between case "a" and case "b" in db is R. Assume far-zone field, perfect polarization match, and perfect impedance match. Then the coupling is

$$C_a = \left(\frac{\lambda}{4\pi\,(C - L)}\right)^2 \qquad \text{case "a"}$$

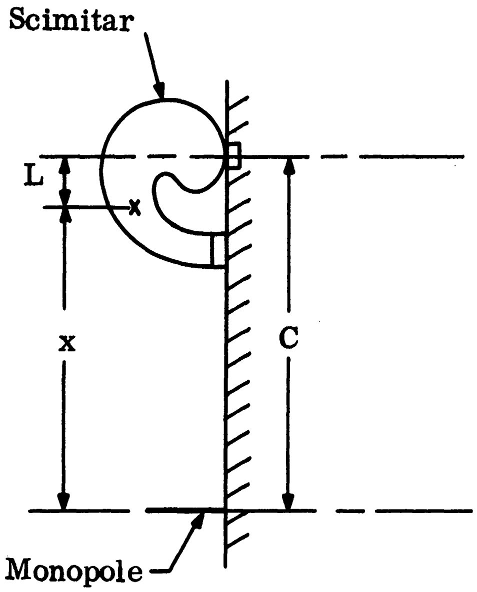
$$C_b = \left(\frac{\lambda}{4\pi\,(C + L)}\right)^2 \qquad \text{case "b"}$$

and,

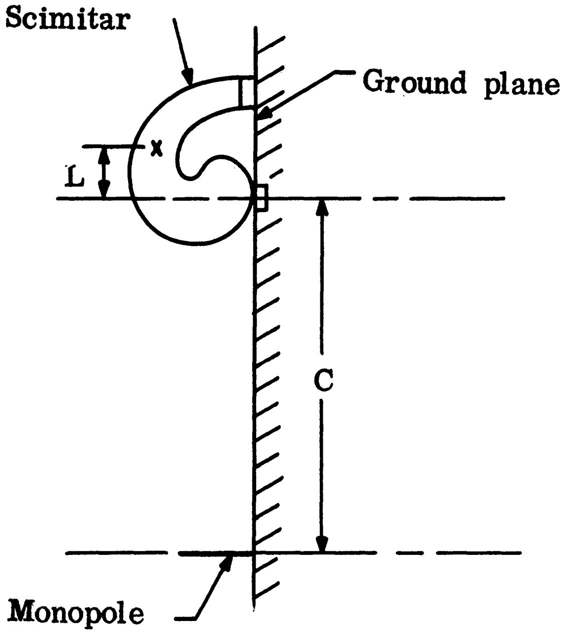
$$R = 20 \log_{10} \frac{C + L}{C - L} \ .$$

Solving for L we get

$$L = C \, \frac{10^{\frac{R}{20}} - 1}{10^{\frac{R}{20}} + 1} \ ,$$



Case "a"



Case "b"

FIG. 2-16: ARRANGEMENTS OF SCIMITAR ANTENNAS

It was expected by rotating the scimitar antenna 180^{o} around its connector that the change in coupling would be higher in case of smaller distance compared to the change in coupling for larger distance between the monopole and the scimitar at the same frequency. This was not the case according to the results obtained from the experiments on two different types of scimitars. The results were not very consistent.

The other set of experiments was on a log-periodic monopole structure mounted over the ground plane. This log-periodic monopole was chosen in spite of the fact it is not a flush mounted arrangement because of its simplicity in construction and the ease of location of the phase center as shown in Fig. 2-17:

A set of experiments was made for coupling measurements between the log-periodic monopole and a resonant monopole for two separation distances 22.8 cm and 45.7 cm (Fig. 2-18). Again the results didn't give any indication of the effect of the phase center shift on the coupling.

There are many factors involved in the change of coupling values with the change of frequency which makes it difficult to single out the effect of the phase center alone. These factors are:

(a) The phase center shift

(b) The distance between the two antennas in wavelength is variable with frequency.

(c) The variation of the radiation patterns with frequency.

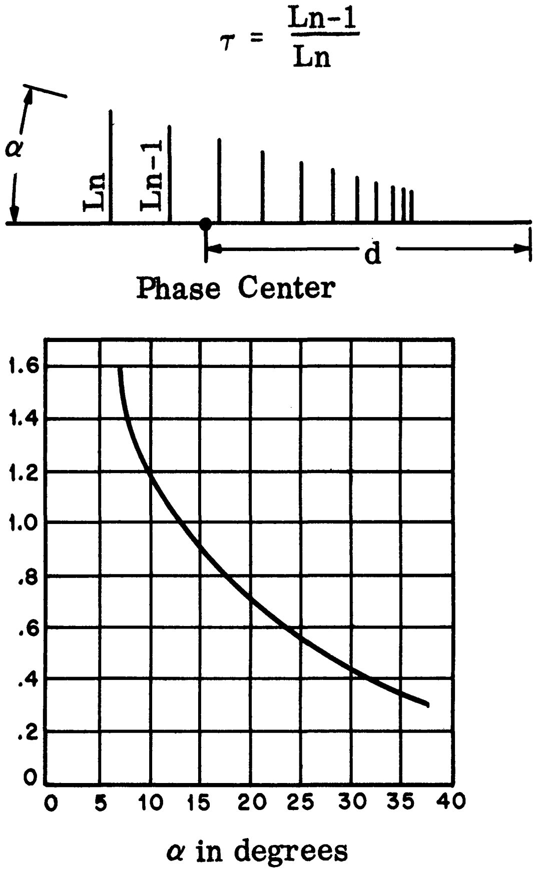


FIG. 2-17: DISTANCE FROM VERTEX TO PHASE CENTER IN WAVELENGTHS AS A FUNCTION OF *α*.

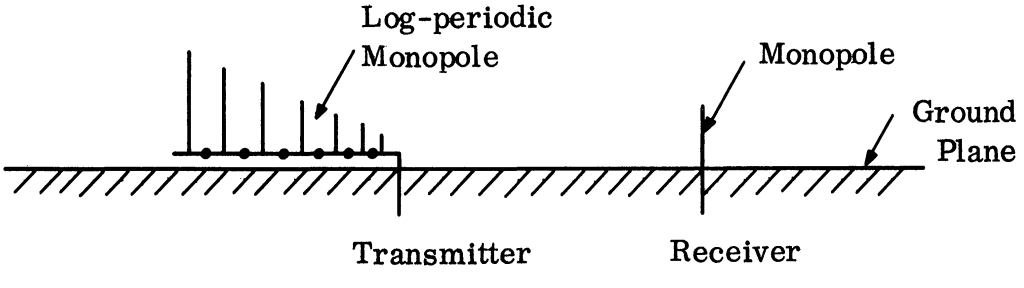


FIG. 2-18: ARRANGEMENT FOR COUPLING MEASUREMENTS FOR LOG-PERIODIC MONOPOLE.

There have been several factors which may have contributed to the inconsistency of the results; the most important ones are:

(a) The adjustment for matching of the antennas was critical

(b) The length of the monopole must remain $\lambda/4$ for different values of frequency.

III

CONCLUSIONS

The work described in this report has shown a great deal of promise in improving methods so that a broad range of frequencies can be adequately covered. Two nearby broadband systems present a more difficult problem of interference than two nearby narrowband systems. The decoupling means must be valid over the entire range of frequencies where these systems may interfere. Thus, to the extent that the frequency bands of the two systems do not overlap, the problem of interference is relatively simple, except for the excitation of higher order modes, and the occurrence of harmonics.

Results are reported showing that a properly compensated RF bridge provides broadband decoupling. It is necessary that the phase and attenuation characteristics of the direct coupled path "track" those of the auxiliary link with frequency. Any deviations from close tracking will result in less decoupling. Analytic studies are included which give the decoupling versus phase and attenuation errors.

A significant improvement has been brought about in the use of a series of concentric corrugations or slots. The idea of a tapered slot has been explored and has been found useful for extending the bandwidth of decoupling.

The method of using fences has been further improved. Also, this method has been used to supplement the decoupling obtained with corrugations. It has been shown that the two methods of decoupling may be effectively combined. The decoupling achieved by the corrugations and the decoupling achieved by the metal picket fence do not necessarily add algebraically.

The studies on Archimedean spirals have shown some general trends that have been previously recognized. The near-field phenomenon is primarily associated with the operation of spirals in the lowest frequency range for which the spirals were designed. At other higher frequencies, the coupling tends to be far-field coupling and is not greatly influenced by the position of the terminations of the windings at the outer periphery of a circular Archimedean spiral. It is still true that there may be some dependence of coupling on the orientation of either one of the Archimedean spiral antennas. The dependence on orientation is an indication that some modes other than the design mode number 1 exist. The design mode number 1, corresponding to a rotating dipole, causes a coupling which has complete independence of the orientation of the antenna.

IV

FUTURE EFFORT

4.1 Future Topics for Investigation

Presently the RF bridge method suffers from a 3db power loss both at the transmitting and receiving junctions. This loss is not inherent in the method, but is due to using hybrid junctions in the experimental setup. Alternative ways to supply power to the bridge link will be investigated to eliminate this loss.

More detailed studies will be made on the combination of decoupling methods. A combination of methods seems necessary to obtain high additional isolation (40 to 50 db). It is anticipated that the RF bridge will be combined with surface decoupling methods such as fences and corrugations. The decoupling obtained from such a combination may approach the sum of the decoupling of each method at certain frequencies.

Additional work must be done on the study of the shift of phase center for broadband antennas. Knowledge of this shift with frequency is necessary for accurate prediction of coupling levels and also for use of the RF bridge with these antennas.

Sophisticated problems of electromagnetic compatibility may exist in distance and angle measuring systems. These systems may dictate very low limits on coupling. For example, an interferometer using three Archimedean spirals was used in the rendezvous radar on the Gemini vehicle (Carl et al, 1966). In this case, coupling due to the scattering from a third element must be considered. Exciting a spiral antenna with an impinging wavefront presents a most challenging problem. An attempt will be made to properly assess such problems, and to indicate to what extent analysis and experiments are necessary to properly evaluate the coupling and decoupling needs of such systems.

Another possible coupling problems deals with the Van Atta array, where coupling between elements appreciably affects the performance. An element of the array, which is to be primarily fed from another element by a direct transmission link, may be adversely irradiated by nearby elements. There may be a need to greatly reduce this coupling to achieve good results.

4.2 Test Facility; Proposed Modifications

In the development and testing of decoupling methods, and in measuring the radiation characteristics of antennas, it has become apparent that some modifications and improvements in the antenna test facility were required. Two areas are involved; 1) improvements in instrumentation to make experimental methods faster and more accurate, and 2) improvements in anechoic chamber performance to increase measurement accuracy.

Plans have been made and work is now in progress to achieve some of the needed improvements. Work on these problems will continue.

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13. ABSTRACT
Significant improvements have been recently obtained for various methods of decoupling two antennas. The RF bridge reduction method has been analyzed to determine the decoupling possible with phase and amplitude errors present. Careful attention to matching the phase and amplitude characteristics of the auxiliary link to the coupling path has made it possible to extend the decoupling to a relatively broad range of frequencies.

The method utilizing circumferential corrugations surrounding either one or both of the antennas involved has been modified to include tapered corrugations. Improved performance has been achieved.

Further studies have been made on the use of metal picket fences as a decoupling means. Such a fence between two antennas has been very effective. Studies have been made utilizing both the metal fence and corrugations. The total decoupling is the sum of that achieved by each method.

In the near-field coupling of Archimedean spirals there is a practical problem in terminating the outside ends of bifilar spirals because the terminations occupy considerable space and influence the electromagnetic field. This report summarizes the studies to date on such near-field coupling

Some progress has been made on studying the effect of shift of phase center on coupling. Quantitative results are still being obtained. Experimental studies are being performed using two broadband antennas, a scimitar and a log-periodic monopole.

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