NPS 62-90-007

NAVAL POSTGRADUATE SCHOOL

Monterey, California

CAD MODELS FOR INDUCTIVE STRIPS
IN HOMOGENEOUS FINLINE:
THE METHODOLOGY

Jeffrey B. Knorr
March 1990

Interim Report for the Period
1 October 89 - 31 March 90

Prepared for: Naval Ocean Systems Center
San Diego, CA 92152

NAVAL POSTGRADUATE SCHOOL
MONTEREY, CALIFORNIA 93943

Rear Admiral R. W. West, Jr.
Superintendent

Harrison Shull
Provost

This report was prepared for the Microwave/Millimeter Wave Branch, Naval Ocean Systems Center and funded by the Naval Postgraduate School.

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This report was prepared by:

Jeffrey B. Knorr
Professor of Electrical
and Computer Engineering

Reviewed by:

John P. Powers
Chairman, Department of
Electrical and Computer
Engineering

Released by:

Gordon Schacher
Dean of Faculty and
Graduate Studies

UNCLASSIFIED
SECURITY CLASSIFICATION OF THIS PAGE

REPORT DOCUMENTATION PAGE

1a REPORT SECURITY CLASSIFICATION: UNCLASSIFIED

1b RESTRICTIVE MARKINGS:

2a SECURITY CLASSIFICATION AUTHORITY:

2b DECLASSIFICATION / DOWNGRADING SCHEDULE:

3 DISTRIBUTION / AVAILABILITY OF REPORT: APPROVED FOR PUBLIC RELEASE; DISTRIBUTION UNLIMITED

4 PERFORMING ORGANIZATION REPORT NUMBER(S): NPS-62-90-007

5 MONITORING ORGANIZATION REPORT NUMBER(S):

6a NAME OF PERFORMING ORGANIZATION: NPS

6b OFFICE SYMBOL (If applicable):

7a NAME OF MONITORING ORGANIZATION: NAVAL OCEAN SYSTEMS CENTER

6c ADDRESS (City, State, and ZIP Code): MONTEREY, CA 93943

7b ADDRESS (City, State, and ZIP Code): SAN DIEGO, CA 92152

8a NAME OF FUNDING / SPONSORING ORGANIZATION: NPS

8b OFFICE SYMBOL (If applicable):

9 PROCUREMENT INSTRUMENT IDENTIFICATION NUMBER: O&MN, DIRECT FUNDING

8c ADDRESS (City, State, and ZIP Code): MONTEREY, CA 93943

10 SOURCE OF FUNDING NUMBERS

PROGRAM ELEMENT NO	PROJECT NO	TASK NO	WORK UNIT ACCESSION NO

11 TITLE (Include Security Classification): (U) CAD MODELS FOR INDUCTIVE STRIPS IN HOMOGENEOUS FINLINE: THE METHODOLOGY

12 PERSONAL AUTHOR(S): JEFFREY B. KNORR

13a TYPE OF REPORT: INTERIM

13b TIME COVERED: FROM 1 OCT 89 TO 31 MAR 90

14 DATE OF REPORT (Year, Month, Day): MARCH 1990

15 PAGE COUNT: 34

16 SUPPLEMENTARY NOTATION:

17 COSATI CODES

FIELD	GROUP	SUB-GROUP

18 SUBJECT TERMS (Continue on reverse if necessary and identify by block number): FINLINE, DISCONTINUITY, INDUCTIVE STRIP

19 ABSTRACT (Continue on reverse if necessary and identify by block number)

This report presents the derivation of a circuit model for homogeneous finline. It also describes several models for inductive strips in homogeneous finline. The models are simple, accurate and completely compatible with existing CAD software.

(computer aided design)

20 DISTRIBUTION / AVAILABILITY OF ABSTRACT: ☐ UNCLASSIFIED/UNLIMITED ☒ SAME AS RPT ☐ DTIC USERS

21 ABSTRACT SECURITY CLASSIFICATION: UNCLASSIFIED

22a NAME OF RESPONSIBLE INDIVIDUAL: JEFFREY B. KNORR

22b TELEPHONE (Include Area Code): (408) 646-2815/2082

22c OFFICE SYMBOL:

DD FORM 1473, 84 MAR — 83 APR edition may be used until exhausted. All other editions are obsolete

SECURITY CLASSIFICATION OF THIS PAGE

☆ U.S. Government Printing Office 1986—606-2

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CAD MODELS FOR INDUCTIVE STRIPS

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THE METHODOLOGY

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Accession For		
NTIS GRA&I		☒
DTIC TAB		☐
Unannounced		☐
Justification		
By		
Distribution/		
Availability Codes		
Dist	Avail and/or Special	
A-1		

by
JEFFREY B. KNORR
March 1990

ABSTRACT

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I. INTRODUCTION

A. BACKGROUND

Finline is a printed circuit transmission structure first described by Meier [1] in 1974. A cross-sectional view of a finline is shown in Figure 1. In its most general form, finline consists of metal fins printed on a dielectric substrate which is supported in the E-plane inside a rectangular waveguide shield. Normally, the shield has the dimensions of a standard rectangular waveguide which permits easy interfacing with waveguide circuitry. The dielectric is typically thin and has a low dielectric constant, its primary purpose being to support the printed fin structure. In some cases, the fin structure is simply printed from thin metal with no substrate at all. Filters, for example, are frequently built this way. Since its introduction, finline has been widely used as a printed circuit medium at millimeter wavelengths. Studies contributing to its effective utilization are therefore of significant interest.

Practical application of finline requires the development of circuit models suitable for use in computer aided design. Existing commercial software products do not, in general, have finline element libraries. However, numerical and experimental studies carried out over the past 15 years provide a basis for development of some useful models. Figure 2, for example, shows a finline cavity containing an inductive strip of length T which spans the space between the metal fins at the center of the cavity. These strips are the principal elements used in the design of finline filters. Special purpose programs have been developed for the design of such filters [2], but no model has been developed for use with general purpose CAD software. The inductive strip has been studied numerically and experimentally, however, so data exist to support development of an appropriate model.

B. OBJECTIVE

The objective of the work reported here was to develop simple and accurate equivalent circuit models for lossless, homogeneous finline and for inductive strips in lossless, homogeneous finline.

C. RELATED WORK

The model development work described here is based on original work carried out by the author and his thesis students at the Naval Postgraduate School, Monterey, CA. The first complete frequency dependent analysis of finline was published by Knorr and Shayda in 1980 [3]. Detailed equations and a computer program listing appear in the 1980 thesis by Shayda [4]. The field equations were later reformulated by Knorr and Kim to achieve better numerical stability. A 1984 thesis by Kim presents the equations and a computer program listing [5]. This work utilized the spectral domain method and the program presented in Kim's thesis was used to generate numerical data for development of the model for homogeneous finline. The inductive strip in finline was studied experimentally by Miller [6] and was subsequently analyzed by Knorr and Deal [7] -[9], again using the spectral domain method. The computer program described in Deal's thesis was used to generate numerical scattering coefficients for inductive strips in finline. These data were used with the finline model to develop the models for inductive strips in homogeneous finline.

The equivalent circuit model for an inductive strip in homogeneous rectangular waveguide (homogeneous finline with $W/b=1$) was invented by the author and first described in a 1988 technical report [10]. The inductive strip models described in this report cover cases where $W/b < 1$. They are based on the original model concept but require, in addition, a model for homogeneous finline with $W/b < 1$. The finline model was developed by the the author and is presented herein for the first time. Three models for the inductive strip are described in this report. Two of these are new and are presented here for the first time.

It should be noted here that, in general, electromagnetic analyses based on Maxwell's equations are not well suited for CAD application. Such analyses are generally cumbersome and time consuming. For design purposes we would like to find simple and efficient models from which structural parameters can be found when electrical behavior is specified. The spectral domain analyses referenced above are very accurate but the method is mathematically complex and implementation requires significant computer resources. The circuit models described in this report are simple and accurate.

II. A MODEL FOR HOMOGENEOUS FINLINE

In general, finline has a large number of variable parameters which makes it difficult to model in any simple way. In practice, however, there are commonly used configurations which can be modeled. One such configuration has the fins centered ($h_1 = h_2 + D = a/2$) and printed on a thin dielectric substrate with $e_r = 2.22$. This configuration can be simply modeled. Here we will describe the modeling process for the homogeneous case ($D = 0$). Preliminary results indicate that the same modeling process can be extended to the inhomogeneous case ($D > 0$) with high accuracy.

When there is no dielectric present, finline is just a homogeneous cylindrical waveguide. For any homogeneous guide, the wavelength ratio and voltage power impedance are given by

$$(\lambda'/\lambda) = [1 - (\lambda/\lambda_c)^2]^{-1/2} \tag{1a}$$

$$Z_{ov} = Z_{ov\infty}(\lambda'/\lambda). \tag{1b}$$

These expressions require only a knowledge of λ_c and $Z_{ov\infty}$ which, once impedance is defined, depend only on the geometry of the structure. For simple geometries, λ_c and $Z_{ov\infty}$ can be found analytically while for more complicated structures such as homogeneous finline, these parameters must be computed numerically.

Rectangular waveguide is a special case of cylindrical waveguide for which the parameters λ_c and $Z_{ov\infty}$ can be found analytically. It is well known that

$$\lambda_c = 2a \tag{2a}$$

$$Z_{ov\infty} = 2(b/a)120\pi. \tag{2b}$$

Thus, if a finline and a rectangular waveguide have the same cutoff wavelength and voltage-power impedance limit their dominant modes will have exactly the same wavelength and impedance at any given frequency. This makes it possible to establish an equivalence between finline and rectangular waveguide.

To establish this equivalence between homogeneous finline and rectangular waveguide we must first compute the finline cutoff wavelength and voltage power impedance limit numerically for a given finline geometry (b/a, h_1/a, W/b). This can be done using the spectral domain method [4]-[5]. Once these finline parameters have been computed, we can use them to determine the dimensions a_{eq} and b_{eq} of a rectangular waveguide such that the wavelength and impedance of the dominant mode are the same for both structures. If λ_c and $Z_{ov\infty}$ are the computed finline cutoff wavelength and impedance, respectively, then the equivalent rectangular waveguide dimensions are given by

$$(a_{eq}/a) = (\lambda_c/2a) \tag{3a}$$

$$(b_{eq}/b) = (Z_{ov\infty}/120\pi)(a_{eq}/a)(a/2b). \tag{3b}$$

This equivalence is illustrated in Figure 3.

If one computes the finline cutoff wavelength and voltage-power impedance limit for many values of W/b, it is possible to plot a graph of (a_{eq}/a) and (b_{eq}/b) vs. W/b. Results for WR(90) finline are shown in Figures 4 and 5. On these graphs, the circles represent spectral domain data and the solid curves are analytical fits determined by Morua [11]. WR(90) has a=900 mils and b=400 mils (b/a = 4/9). Curves similar to those shown in Figures 4 and 5 can also be obtained for b/a = 1/2 which is the case for all millimeter wave rectangular waveguides. Morua has found that the following equations fit the spectral domain data for both b/a = 1/2 and b/a = 4/9 with a maximum error of 1.3% (in most cases the error is less than 1%) when $0.01 \leq W/b \leq 1.0$:

$$a_{eq}/a = 2 - [1 - (2b/a)^{0.77}(1 - W/b)^2]^{1/2} + 0.221(2b/a)^{-3.61}(1 - W/b)^{28} \tag{4a}$$

$$b_{eq}/b = 0.6 + [0.16 - 0.1347(2b/a)^{1.35}(1 - W/b)^2]^{1/2} - 0.17(2b/a)^{-1.15}(1 - W/b)^{10}. \tag{4b}$$

Although they have not been checked extensively, it appears that these equations are accurate for the larger range $0.25 < b/a < 0.6$ with approximately the same maximum error.

These equations permit homogeneous finline to be modeled as a rectangular waveguide with an error determined by the accuracy of the spectral domain data and the expressions for a_{eq}/a and b_{eq}/b. The model is completely compatible with existing CAD software. One simply uses the existing rectangular waveguide element from the software circuit element library and defines the dimensions a and b using Eqs.(4).

Preliminary results show that inhomogeneous finlines with thin, low dielectric constant substrates can also be modeled in the same way. The representation is no longer exact but the error is acceptable.

III. SEVERAL MODELS FOR INDUCTIVE STRIPS IN HOMOGENEOUS FINLINE

Three models for inductive strips in homogeneous finline with $W/b < 1$ are presented below. All are based on the concept first presented in [10]. The main difference between this work and the earlier work is that when $W/b < 1$ the strips are embedded in finline with $W/b < 1$. Thus the circuit model for the strip must be terminated at its input and output with a resistance equal to the voltage-power impedance of the particular finline under consideration. The two additional models presented here are somewhat more complicated than the one presented in [10] but they are also more accurate as will be shown.

The procedure used to determine unknown equivalent circuit element values is the same in all cases. First, spectral domain scattering data is generated for the strip of interest. This is used to create an S-parameter data file. The element values of the model are then optimized to minimize the error between the model scattering data and the spectral domain scattering data. For lossless, symmetric, reciprocal 2-port networks the scattering matrix has only 1 independent element [10] so it is sufficient to match the value of s_{11} predicted by the model to the value of s_{11} computed using the spectral domain method. The cutoff wavelength and voltage-power impedance limit must also be computed for finline of the appropriate a/b, h_1/a and W/b since impedance data are required to design the termination for the strip model.

A. MODEL 1

The first model is simply the one described in [10]. It follows from physical inspection of the inductive strip in homogeneous finline. The regions to either side of the strip are parallel sections of below cutoff waveguide. For a strip between centered fins these sections have width $a/2$, height b and length T. At each edge of the strip, the finline field sees a discontinuity which results in evanescent fields storing predominantly magnetic energy. This can be accounted for by including a parallel inductor in the equivalent circuit. The complete equivalent circuit is shown in Figure 6.

The equivalent circuit in Figure 6 is terminated in the finline impedance and the value of the inductor, L, is optimized to obtain the best match between predicted and computed values of s_{11} for a strip of specified length. To illustrate, consider a WR(90) finline with $W/b = 0.50$, fins centered and an inductive strip with a length $T = 40$ mils. Figure 7 shows a Smith chart plot of the predicted and computed values of s_{11} with $L = 12.49$ nH. Figure 8 shows the corresponding graphs of magnitude and angle of s_{11}. All data are for the 8 - 12 GHz frequency range. From Figure 8, it can be seen that the inductance, L, has been chosen to obtain a worst case error of 3.4% for both magnitude and angle of s_{11}.

B. MODEL 2

A slightly more complex model is shown in Figure 9. This model is identical to the previous model except for the capacitor which has been added in parallel with the inductor. Although the leading edge discontinuity effect results in a predominance of stored magnetic energy, there is both stored electric and magnetic energy associated with the evanescent field. The addition of the capacitor acknowledges this reality. From another point of view, it is clear that if we wish to match the predicted and computed values of s_{11} in both magnitude and angle we must, in general, have at least 2 degrees of freedom in the model. If the parameters of the below cutoff sections of waveguide are fixed according to physical reality (width = a/2, height = b, length = T) then the only way we can obtain a second degree of freedom is by introducing another variable element into the equivalent circuit. With the addition of the capacitor, the model can more accurately predict the scattering coefficients of the strip.

To illustrate the performance of model 2, again consider a WR(90) finline with W/b = 0.50, fins centered and a T = 40 mil strip. Figure 10 shows a Smith chart plot of the predicted and computed values of s_{11} and Figure 11 shows both the magnitude and angle of s_{11} over the 8 - 12 GHz frequency range. The values of inductance and capacitance for these Figures were L = 10.65 nH and C = .0028 pF. An inspection of Figure 11 shows that the worst case error is now 2% for for both magnitude and angle. Thus, an improvement in accuracy has been obtained at the expense of an increase in the complexity of the model.

C. MODEL 3

The third equivalent circuit model for the inductive strip is shown in Figure 12. The reasoning behind this equivalent circuit can be understood by thinking of the waveguide section containing the strip as a bifurcated, full height waveguide which is connected at each end to finned waveguide with W/b < 1. An impedance step occurs at the junctions where we pass from the bifurcated region to the finned region. The inductor and capacitor are included to account for the discontinuity effect and the ideal transformer models the impedance step from W/b = 1 to W/b < 1. The principal difference between model 2 and model 3 is that model 3 scattering coefficients are computed relative to the voltage-power impedance of full height waveguide as defined in Eqs.(1b) and (2b). This impedance is matched to the impedance of the finline with W/b < 1 by the ideal transformer with turns ratio $n = [Z1/Z2]^{1/2}$ where Z1 and Z2 are the voltage-power impedances of finlines with W/b = 1 and W/b < 1, respectively. This acknowledges the physical reality. It has been found that this model predicts scattering coefficients which are in excellent agreement with the computed scattering coefficients.

To illustrate, consider once more a WR(90) finline with W/b = 0.50, fins centered and a strip with length T = 40 mils.

Figure 13 is a Smith chart plot of predicted and computed values of s_{11}. Figure 14 shows the predicted and computed values of the magnitude and angle of s_{11}. The values of inductance and capacitance used here were L = 13.70 nH and C = .0015 pF. The transformer turns ratio n was computed using the equations from Figure 3 along with Eqs. (4) and varies with frequency. The error is so small that it is indeed difficult to determine from Figure 14. Inspection of the numerical values of s_{11}, however, show that it is less than about 0.7% for both magnitude and angle.

D. MODEL ELEMENT VALUES VS. STRIP LENGTH T

The procedure described above can be carried out for strips of various lengths. For each strip length an optimum value of inductance and capacitance can be found. These data can then be used to find an analytical expression which gives the element value as a function of strip length. For example, the author [10] found for W/b = 1 the model 1 inductance was given by

$$L = 16 + 14.4*EXP(-T/75) \text{ nH} \tag{5a}$$

and Morua [11] has determined that for W/b = 0.25

$$L = 3.81 + 4.02*EXP(-T/119) \text{ nH} \tag{5b}$$

where in each case T is in mils. It should be noted that the inductance in model 1 can be adjusted for a best fit in magnitude, a best fit in angle or a compromise. The inductance expressions given above result in a best fit to the angle of s_{11}.

E. MODEL ELEMENT VALUES VS. W/b

For model 1, exponential fits of the form

$$L = L1 + L2*EXP(-T/T0) \text{ nH} \tag{6}$$

can be found for any value of W/b. The last step in the modeling process is to find an analytical expression which properly predicts the constants L1, L2 and T0 in terms of W/b. This leads to a model which predicts scattering coefficients for a strip in terms of both T and W/b. Such a model is complete in all respects for a homogeneous finline with fixed b/a and h_1/a. Similar comments apply to the capacitance which appears in model 2 and model 3. No further examples can be given here since the necessary data has not yet been generated. The methodology, however, is clear and it is the methodology which we wish to present in this report.

F. A WORD ON ACCURACY

As seen above, it is possible to attain very close agreement between predicted and computed scattering coefficients. A difference of less than 1% can be achieved with model 3. At this point one must question the accuracy of the spectral domain data used to develop the models. Computationally, the accuracy of these data depends upon the quality of the approximations to the electric field in the plane of the fins (see Figs.1 and 2), the number of terms summed in several series, the accuracy of a root searching algorithm and lastly numerical errors. Thus, although spectral domain data is generally considered to be very accurate, it is difficult to determine exactly how accurate. If due care is exercised in developing the computer code, it is probably better than 1%.

Additionally, we must face the question of how accurately the electromagnetic model represents a real physical structure. A real finline has metal fins that are typically 0.5 - 2 mils thick where the EM model assumes zero thickness. A real finline can only be constructed with finite tolerance and once constructed, the actual dimensions as well as electrical parameters like dielectric constant can only be measured with finite precision. And there is always uncertainty when we measure the electrical performance of a structure, particularly at microwave and millimeter-wave frequencies. Thus even if we could model a finline precisely, we could not build it and know all its parameters precisely. Therefore, we must accept that there will always be some difference between performance predictions based on modeling and the performance we actually measure.

In the author's experience, it has been found that it is relatively easy to reduce the discrepancy between predicted and measured performance to a few percent. Beyond this the time, effort and expense required to reduce any discrepancy all increase dramatically as one attempts to reduce the discrepancy. An error of less than 1% is very good and in many cases, an error of a few percent is acceptable. For those cases where higher accuracy is desired, the benefit must be weighed against the cost.

IV. SCALING TO OTHER FREQUENCY BANDS

A methodology for developing equivalent circuit models for homogeneous finline and inductive strips in homogeneous finline has been outlined above. The finline model is valid for the range $4/9 < b/a < 1/2$ (and probably beyond these limits with little increase in error) and $0.01 \leq W/b \leq 1$. The model may be used in any frequency band. The models for the inductive strips, however, are at present limited to use only in the frequency band for which the scattering data computations were made. If the dependence of the circuit element values on W/b has been determined as described in Sec. III E then this parameter may be varied but b/a is constrained to the value for which the spectral domain scattering computations were made. The frequency band restriction can be removed by application of the scaling principle. The principle states that if all dimensions and wavelength are scaled by the same factor the electrical performance will remain unchanged.

A. GENERAL CASE (W/b < 1)

Application of the scaling principle requires that the normalized reactances,

$$x_L(\omega_c) = (\omega_c L/Z_{ov}) \tag{7a}$$

and

$$x_C(\omega_c) = (1/\omega_c C Z_{ov}) \tag{7b}$$

remain constant as a structure is scaled. For a given value of W/b we obtain expressions for inductance and capacitance of the form (see Eq. (6), for example)

$$L = f_1(T) \tag{8a}$$

$$C = f_2(T) \tag{8b}$$

where the functions f_1 and f_2 are appropriately determined for WR(90) finline. To scale the model for application in other frequency bands, we compute L′ and C′ using

$$L' = (b/400) f_1(900T/a) \tag{9a}$$

$$C' = (a/900) f_2(900T/a) \tag{9b}$$

where a and b are the shield (waveguide) dimensions in mils and L′ and C′ are the values of inductance and capacitance to be used when modeling an inductive strip in a finline with a given value of b/a and W/b (restriction on W/b is removed if the dependence has been determined as outlined in III E). The turns ratio for the ideal transformer must be determined from the voltage-power impedance of the finline with W/b < 1. Equations (4) provide the necessary information for calculating the voltage-power impedance. Thus, a model developed based on scattering data for strips in WR(90) waveguide operating in X-band can be scaled for use in any frequency band. It must be remembered, however, that

$b/a = 4/9$.

B. SPECIAL CASE ($W/b = 1$)

Scaling in the special case $W/b = 1$ was discussed in [10]. In this case, electrical performance is independent of finline shield height b and the restriction that b/a remain constant is removed. For this special case, the appropriate values of model inductance and capacitance are given by

$$L' = (b/400)f_1(900T/a) \tag{10a}$$

$$C' = (4/9)(a/b)(a/900)f_2(900T/a) \tag{10b}$$

where a and b are in mils and f_1 and f_2 are the analytic expressions for model inductance and capacitance for WR(90) finline. Examples of scaled results for $W/b = 1$ were presented in [10].

C. SENSITIVITY TO b/a

Most standard rectangular waveguides (and all millimeter band guides) have $b/a = 1/2$. Ref. [12] shows that there are only two standard rectangular waveguides that do not have aspect ratios in the range $4/9 \leq b/a \leq 1/2$. One is WR(112) which has $b/a = 0.443$ and the other is WR(42) which has $b/a = 0.405$. Thus, the homogeneous finline model presented here covers all the standard waveguide sizes with high accuracy. It is of interest to investigate the possibility that models for inductive strips in WR(90) finlines ($b/a = 4/9$) with $W/b < 1$ could be scaled to other frequency bands with relaxation of the restriction on b/a. If this restriction is ignored, the values of L and C are scaled as

$$L' = (b/400)f_1(900T/a) \tag{11a}$$

$$C' = (4/9)(a/b)(a/900)f_2(900T/a) \tag{11b}$$

where again a and b are in mils and f_1 and f_2 are the analytic expressions valid for inductive strips in WR(90) finline.

To determine the sensitivity of model 3 to b/a, Eqs.(11) were used to scale the model for a $T = 40$ mil strip in WR(90) finline with $W/b = 0.25$. The model was scaled to Ka band (26 - 40 GHz). The finline shield width was chosen equal to 280 mils, the width of WR(28) waveguide. Three shield heights were investigated; $b = 70$ mils ($b/a = 0.25$), $b = 140$ mils ($b/a = 0.5$ as for WR(28) guide) and $b = 168$ mils ($b/a = 0.6$). It should be noted that the last choice, $b/a = 0.6$, is probably not practical as it reduces the bandwidth relative to that obtained with WR(28) guide. The Ka band strip length was chosen as $T = 12.44$ mils to maintain the same value of T/a.

The scattering coefficients predicted by the scaled model 3 were compared to the correctly computed scattering coefficients.

Results for the three structures are shown in Figures 15 - 17. For b/a = 0.5, the maximum error in either magnitude or angle of s_{11} increases to 1.3%. Had the restriction on b/a not been violated, this error would have been 0.7%, the same as for WR(90) in X band. For b/a = 0.25 the magnitude error is a maximum of 6% and the angle error is a maximum of 2.5%. For b/a = 0.6, the magnitude error is a maximum of 4% and the angle error is a maximum of 2%. These results show that model predictions are not highly sensitive to b/a. Yet, the high accuracy obtainable with model 3 is sacrificed by violating the restriction on b/a which is necessary to adhere to the scaling principle. Thus, it seems that it would be best to modify the model to account for the dependence of L and C on b/a at some time in the future. In the interim, b/a can be varied over a small range with some degradation in accuracy.

POSTSCRIPT:

The results presented in the above paragraph for b/a = 0.6 should be disregarded. After this report was completed, it was determined that the CAD software produced erroneous results when using the waveguide model with b > a/2 as is required in the model for the inductive strip any time b/a > 0.5. It was found during testing of another situation that if b is assigned a value greater than a the software will disregard this and set a equal to the larger of the two values. Thus, at present, it is not possible to model inductive strips in finlines with b/a > 0.5 using the element RWG from the TOUCHSTONE library.

V. EXPERIMENTAL VALIDATION

It has been demonstrated that the models presented here are capable of accurately predicting scattering coefficients which agree with those computed using the spectral domain method. Thus, if the spectral domain data accurately predict what we observe experimentally, the models will also. The accuracy of the spectral domain data has been checked directly for W/b = 1 by measuring strip scattering coefficients in WR(90) waveguide [6]. Measured X-band filter performance was also found to be in excellent agreement with predictions made using model 1 as discussed in [10] for W/b = 1. For W/b < 1, fewer experiments have been done. Additional measurement data will therefore be needed to validate models developed using the methodology described in the previous sections.

For W/b < 1, direct measurement of scattering coefficients can be difficult and tedious even for homogeneous finline. Inhomogeneous finline presents even greater difficulties. However, the resonant frequency and Q of a simple resonator consisting of 2 identical strips depend upon the magnitude and angle of s_{11} in a relatively simple way. Thus, if the models correctly predict the insertion loss and return loss of a resonator or filter, it can be implied that the predicted strip scattering coefficients are in agreement with the scattering coefficients which would be observed experimentally <u>if</u> they were measured direcly. Since it is relatively easy to make the scalar insertion loss and return loss measurements on a resonator or a filter, the models actually provide a straightforward means for experimental validation. In addition, this approach is satisfying since the end use of the models will be for filter design in many cases. Thus, indirect verification by means of resonator and filter insertion and return loss measurements is attractive.

To illustrate the application of the above technique, the predicted and measured response of a WR(90) finline resonator with W/b = 0.25 will be presented. The resonator was constructed from 2 strips with length T = 63 mils separated by a distance of 678 mils. The resonator structure was etched from 2 mil beryllium copper with 3 inch linear tapers from W/b = 0.25 to W/b = 1.0 starting more than 1 inch away from the strip at each end of the structure. The purpose of the taper was to provide a transition from empty waveguide to the fin loaded region. Figure 18 shows the measured response of the resonator obtained using a HP 8756 Scalar Network Analyzer with a HP 8350 Sweep Oscillator. The measured resonant frequency is 8800 MHz. The frequency accuracy of the HP 83592B plug-in is specified as +/- 10 MHz. Since the 4000 MHz sweep is digitized at 401 points, the analyzer frequency resolution is also 10 MHz. Thus, an uncertainty of 10 - 20 MHz should be associated with the measured resonant frequency.

The resonator is realized in rectangular waveguide, so it was necessary to use low VSWR coax to waveguide transitions at each end of the test fixture to interface to the test equipment.

These transitions, along with the linear tapers and the measurement system errors (principally source mismatch) are probably responsible for the "ripple" which is apparent in the measured response.

Figures 19 - 21 show the response obtained using model 1 with L selected for best magnitude fit, best angle fit and best overall fit to s_{11}, respectively. Figure 19 shows that with a best magnitude fit the correct Q is predicted but the resonant frequency is low by about 250 MHz, an error of 2.8%. The best angle fit in Figure 20 predicts the resonant frequency within about 20 MHz but the 10 dB bandwidth is about 200 MHz smaller than the measured 10 dB bandwidth of about 730 MHz. This is about 25% error in the bandwidth. Figure 21 shows that the best fit for both magnitude and angle results in a response between those shown in Figures 19 and 20. The discrepancy between the measured and predicted response is reasonable but not totally satisfying.

Using model 3, the response shown in Figure 22 is obtained. As might be expected, this model predicts the response more accurately than model 1. Overall, the predicted resonator response is very close to that which was measured. The predicted resonant frequency is 8836 MHz which is within 0.4% of the measured value. As stated earlier, this is satisfying since it is expected that these models will be used most frequently to design filters. Thus, the ability of the models to correctly predict the response of a filter is probably a fair test of their validity.

VI. CONCLUSIONS AND RECOMMENDATIONS

A. CONCLUSIONS

This report describes three models for inductive strips in lossless, homogeneous finlines. A finline model is presented for the case of centered fins, $.01 \leq W/b \leq 1$ and $.444 \leq b/a \leq .500$. This finline model is new. Two of the inductive strip models, 2 and 3, are also new. Both are more complex but also more accurate than the model published earlier by the author.

The finline model is based on spectral domain data. The methodology for determining the inductive strip model parameters is also based on spectral domain data. Some examples are given to illustrate the application of the methodology in the case of WR(90) finline. Validation procedures are discussed and some measurement data are presented to show that the model 3 results for inductive strips in WR(90) finline with W/b = 0.25 can be used to simulate the response of a resonator with very high accuracy. A procedure for scaling X band models to other frequency bands is also presented. This procedure permits models developed in one frequency band to be used in any other frequency band.

B. Recommendations

There is a considerable amount of work which must be accomplished in order to bring the strip modeling work to a conclusion. It seems relatively clear how that work should proceed:

1. The optimum values of inductance, L, and capacitance, C, need to be found for inductive strips of various lengths, T, in WR(90) finlines with a range of W/b's.

2. Analytical expressions for L(T) and C(T) need to be found for each chosen value of W/b.

3. The parameters of the equations for L(T) and C(T) must be expressed analytically in terms of W/b.

4. The complete model for strips in WR(90) homogeneous finline needs to be put in generalized form and demonstrated as to accuracy in other waveguide bands.

5. The steps above need to be repeated for other values of b/a, particularly b/a = 1/2. The dependence of model L and C on b/a needs to be determined and incorporated into the model.

6. Models for inhomogeneous finlines and inductive strips in inhomogeneous finlines must be developed. This can probably be accomplished by modifying the present models to account for the presence of dielectric substrate.

7. The effect of loss needs to be included in the model. Loss is not included in the TOUCHSTONE model for waveguide below cutoff. It can be anticipated that experimentally observed losses can be modeled as some combination of the metal loss in waveguide below cutoff plus additional loss associated with the currents generated by the strip edge discontinuity. For inhomogeneous finline, the effect of dielectric loss must also be included.

8. The dependence of model parameters on metal thickness must be determined. The spectral domain data used here applies in the case where the metal is infinitesimally thin. In general, however, the scattering coefficients of an inductive strip as well as the impedance and wavelength of a finline vary with metal thickness and this becomes an important consideration at higher frequencies. Thus, the dependence of scattering coefficients on metal thickness must be included in the modeling process or the use of the model at higher frequencies where t/a (metal thickness to guide width) cannot be assumed negligibly small will produce degraded results. (It should be noted that the X-band experimental structures discussed herein had metal thickness t = 2 mils or t/a = 2/900 = 0.0022. Thus, the metal was thin for these structures.)

9. Lastly, models must be validated. Little experimental data for this purpose currently exists.

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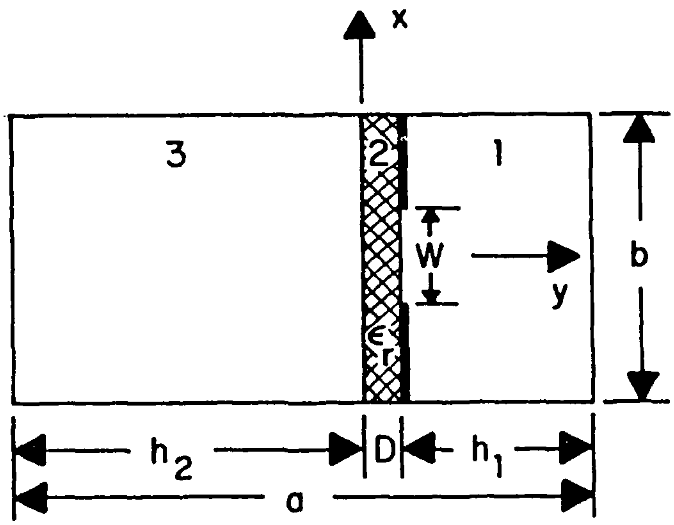


Figure 1. Cross-sectional view of a finline.

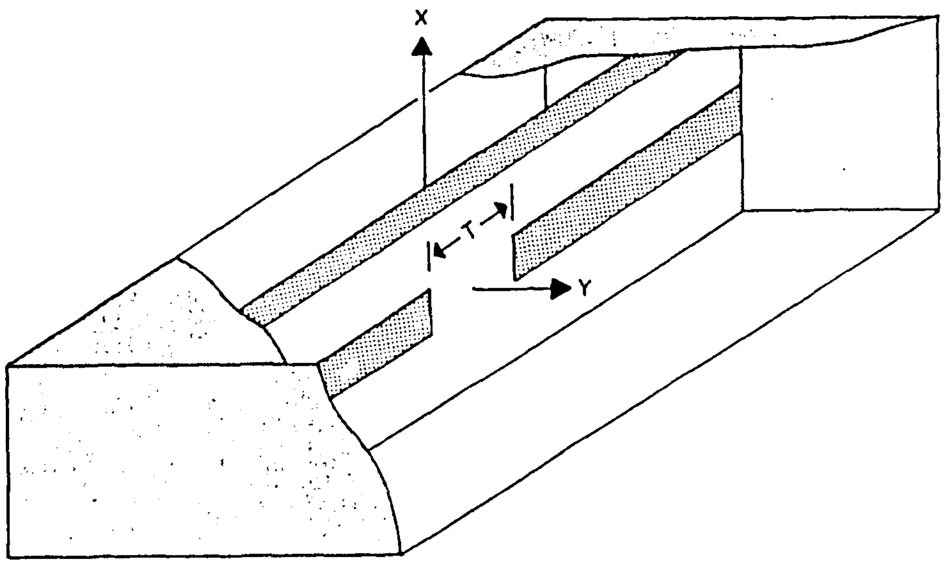
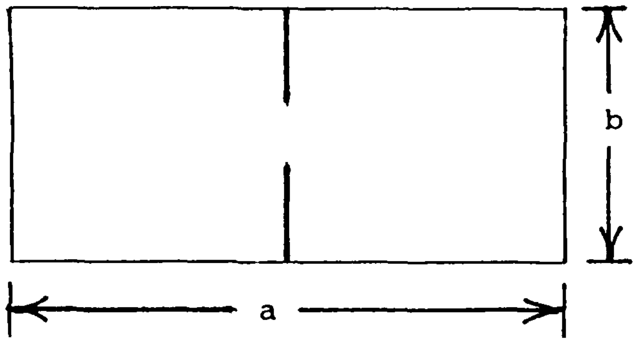


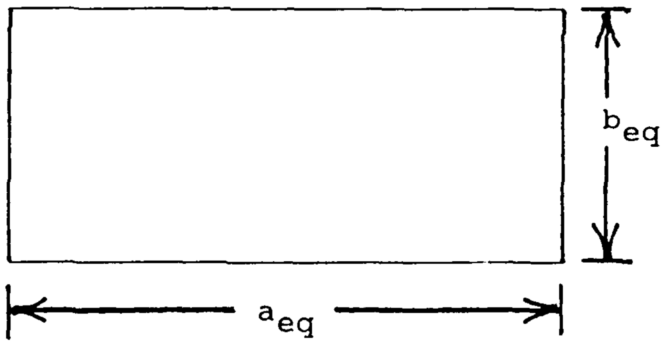
Figure 2. Cut away view of a finline cavity containing a metal strip of length T which spans the space between the fins at the center of the cavity.



$$(\lambda'/\lambda) = [1 - (\lambda/\lambda_c)^2]^{-1/2}$$

$$Z_{ov} = Z_{ov\infty}(\lambda'/\lambda)$$

(λ_c & $Z_{ov\infty}$ found numerically)



$$(\lambda'/\lambda) = [1 - (\lambda/2a_{eq})^2]^{-1/2}$$

$$Z_{ov} = 2(b_{eq}/a_{eq})120\pi(\lambda'/\lambda)$$

$$(a_{eq}/a) = (\lambda_c/2a)$$

$$(b_{eq}/b) = (Z_{ov\infty}/120\pi)(a_{eq}/a)(a/2b)$$

Figure 3. Homogeneous finline and equivalent homogeneous rectangular waveguide.

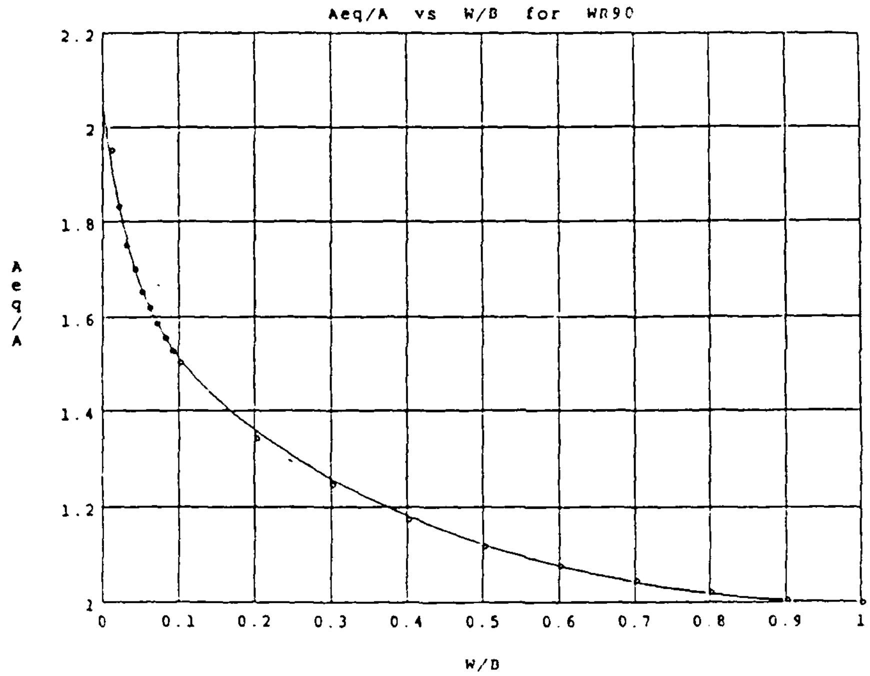


Figure 4. Graph of a_{eq}/a vs. W/b for a WR(90) homogeneous finline with fins centered.

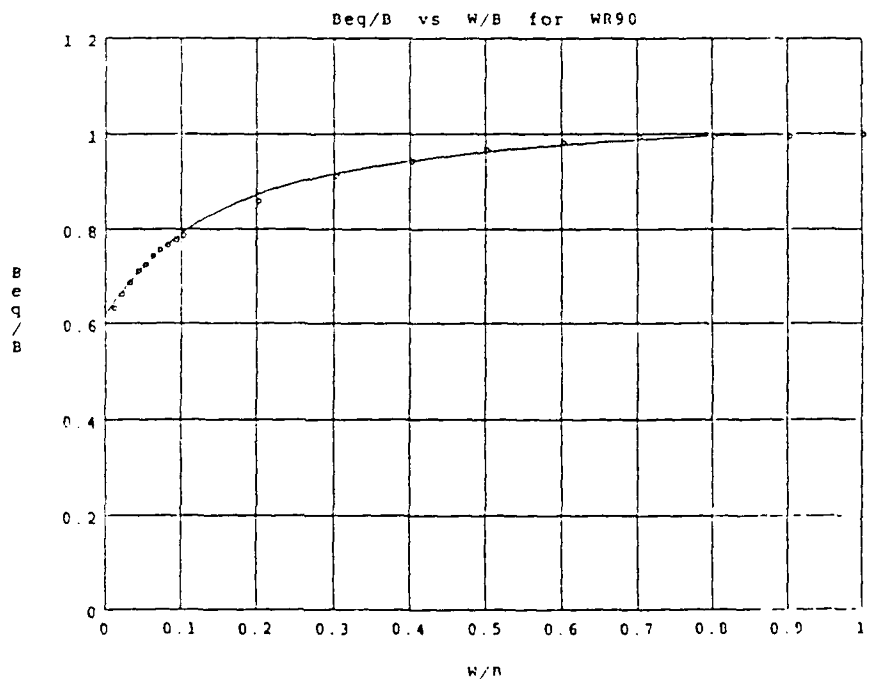


Figure 5. Graph of b_{eq}/b vs. W/b for a WR(90) homogeneous finline with fins centered.

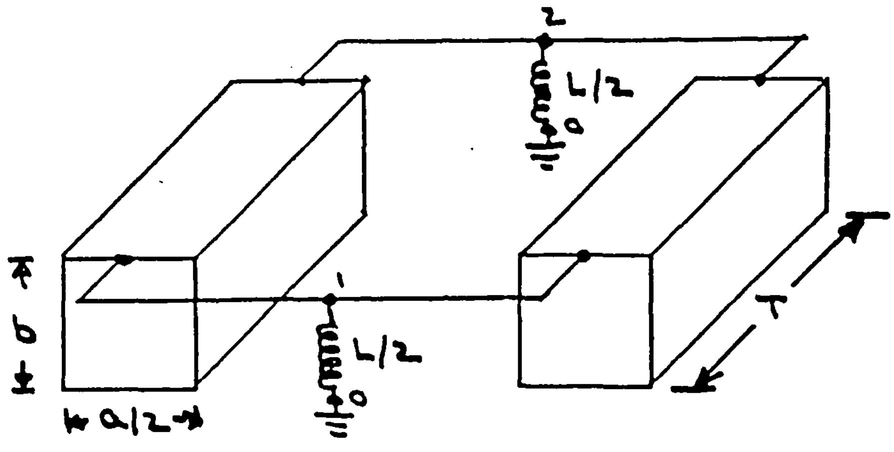


Figure 6. Circuit model 1 for an inductive strip in a homogeneous finline.

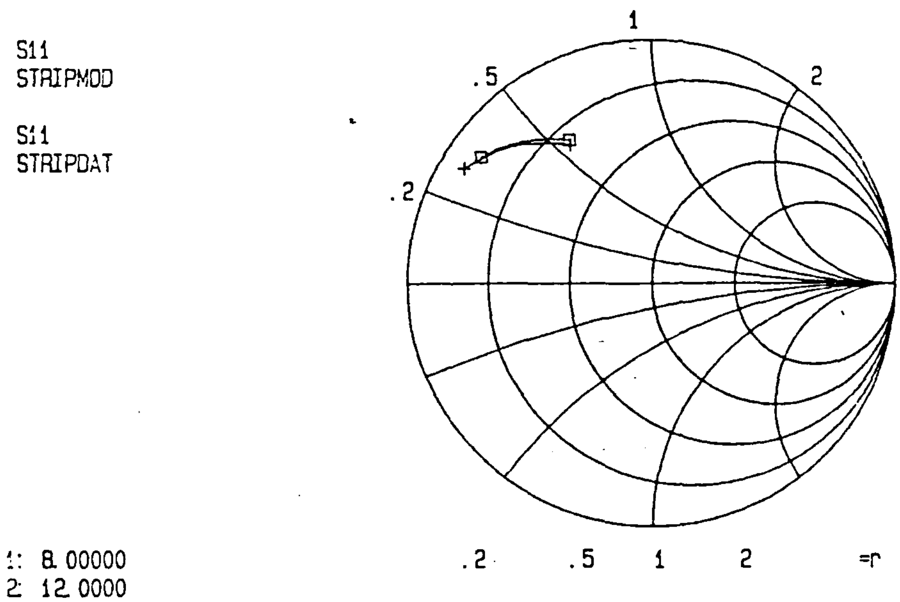


Figure 7. Smith chart plot of predicted and computed values of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 1.

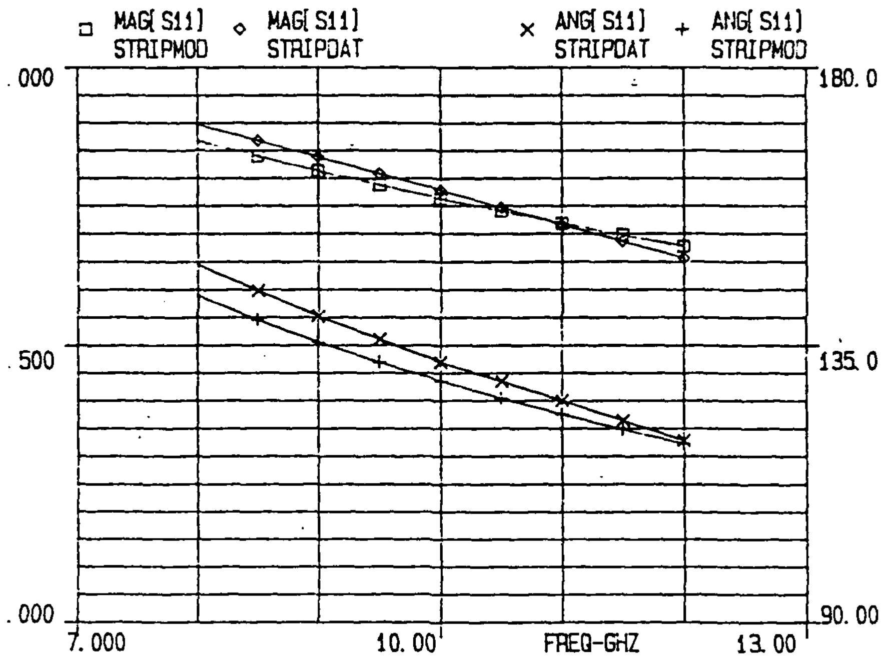


Figure 8. Predicted and computed values of magnitude and angle of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 1.

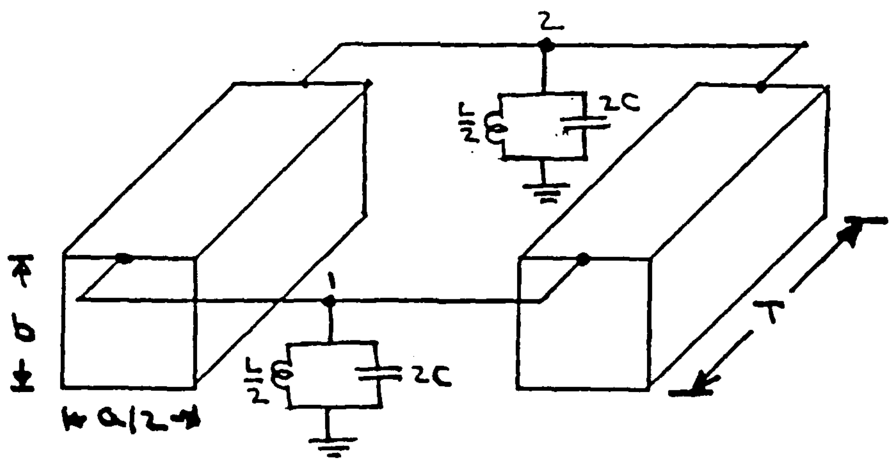


Figure 9. Circuit model 2 for an inductive strip in a homogeneous finline.

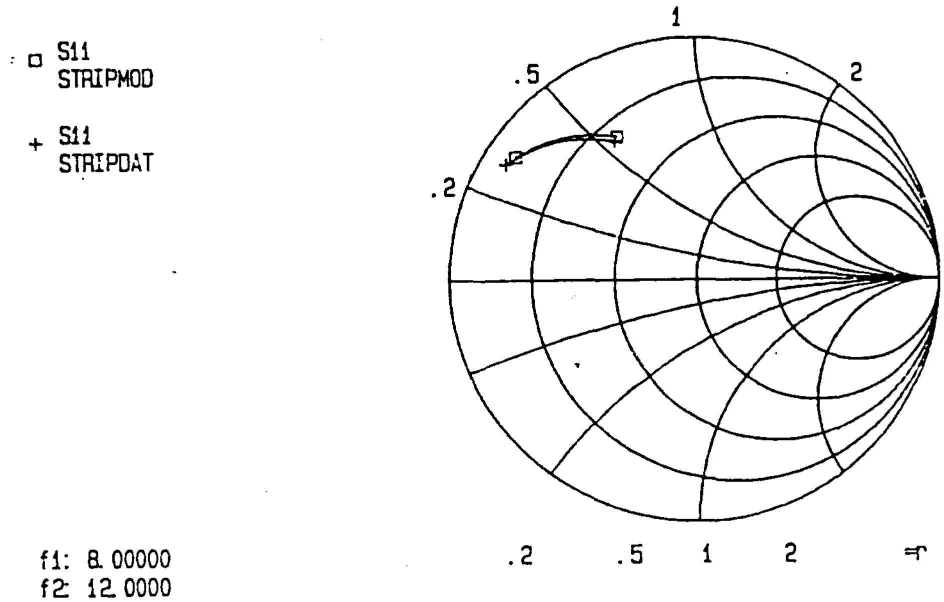


Figure 10. Smith chart plot of predicted and computed values of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 2.

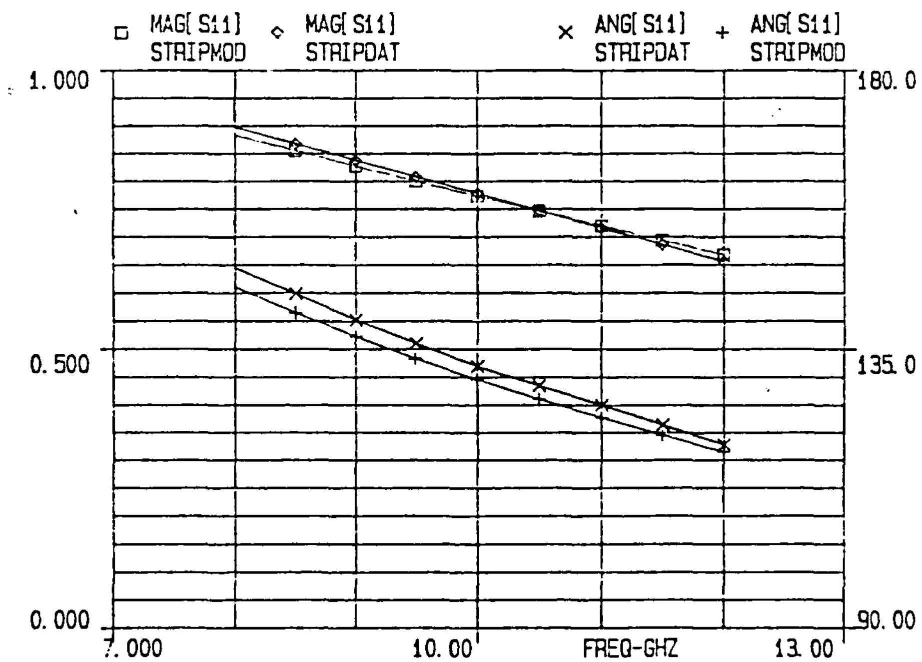


Figure 11. Predicted and computed values of magnitude and angle of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 2.

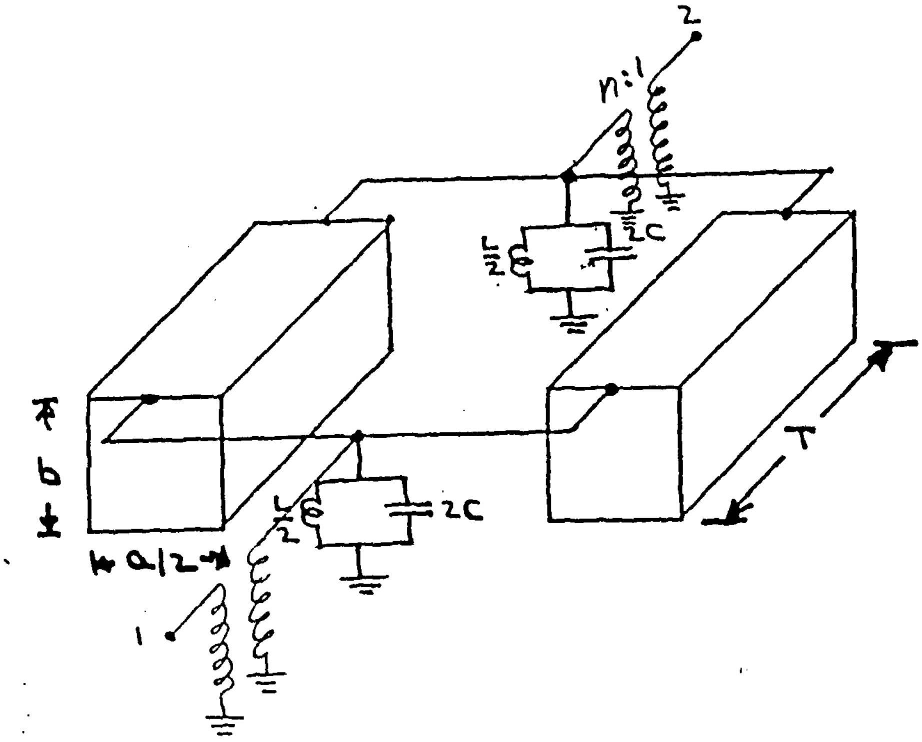


Figure 12. Circuit model 3 for an inductive strip in a homogeneous finline.

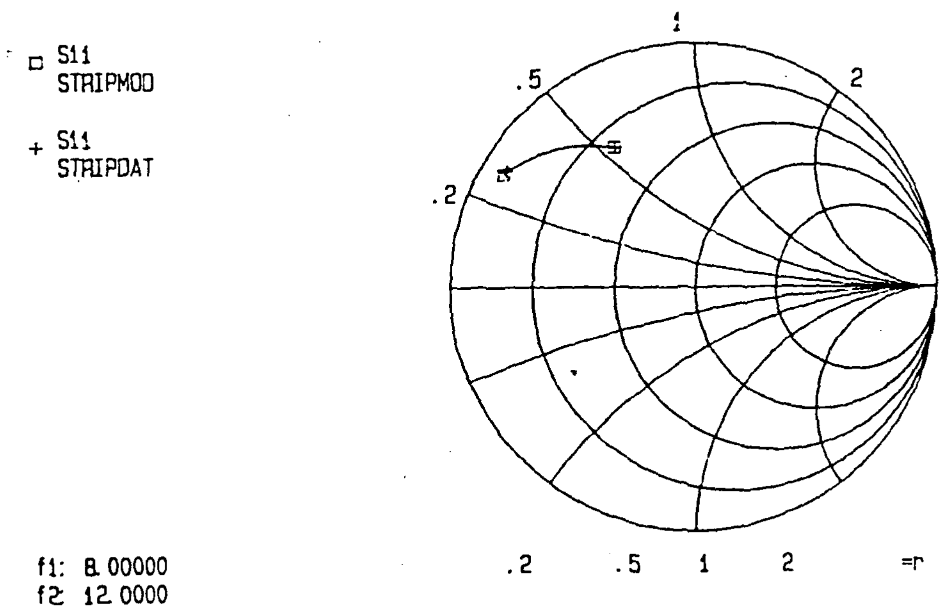


Figure 13. Smith chart plot of predicted and computed values of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 3.

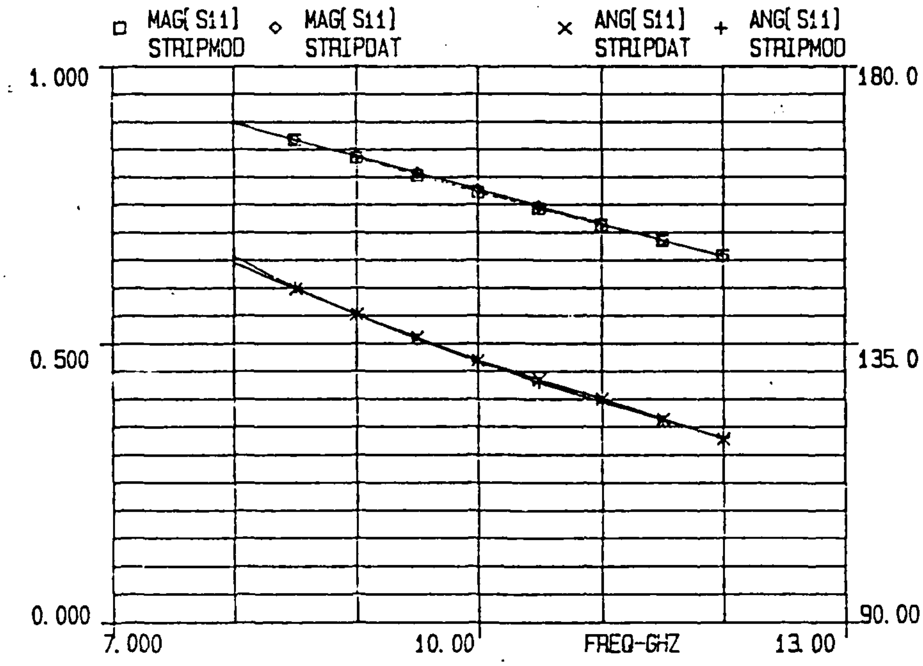


Figure 14. Predicted and computed values of magnitude and angle of s_{11} for a T = 40 mil strip in a WR(90) finline with W/b = 0.5 using model 3.

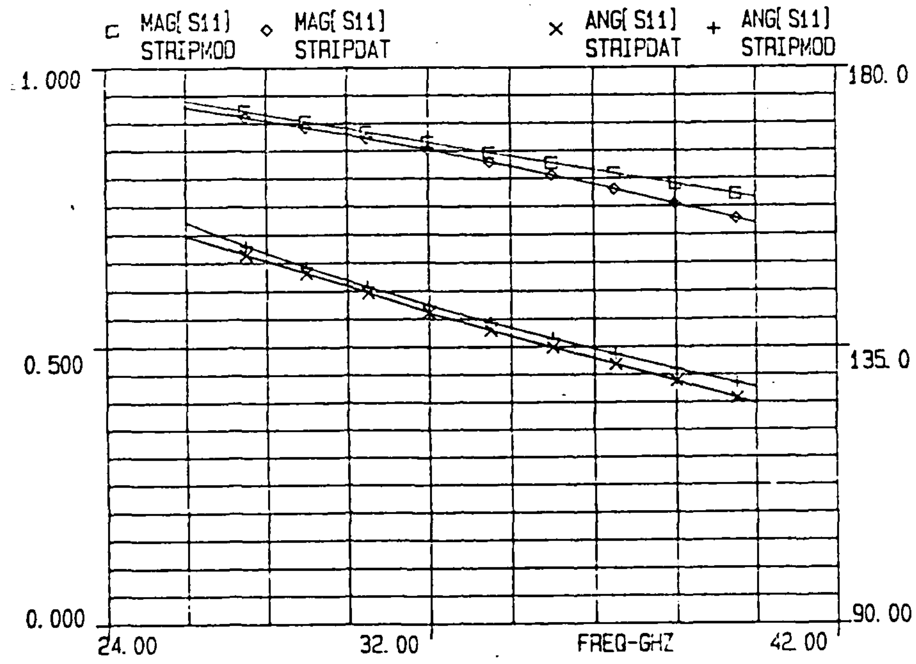


Figure 15. WR(90) model 3 scaled to Ka band. W/b = 0.25, T = 12.44 mils, a = 280 mils, b = 70 mils (b/a = 0.25).

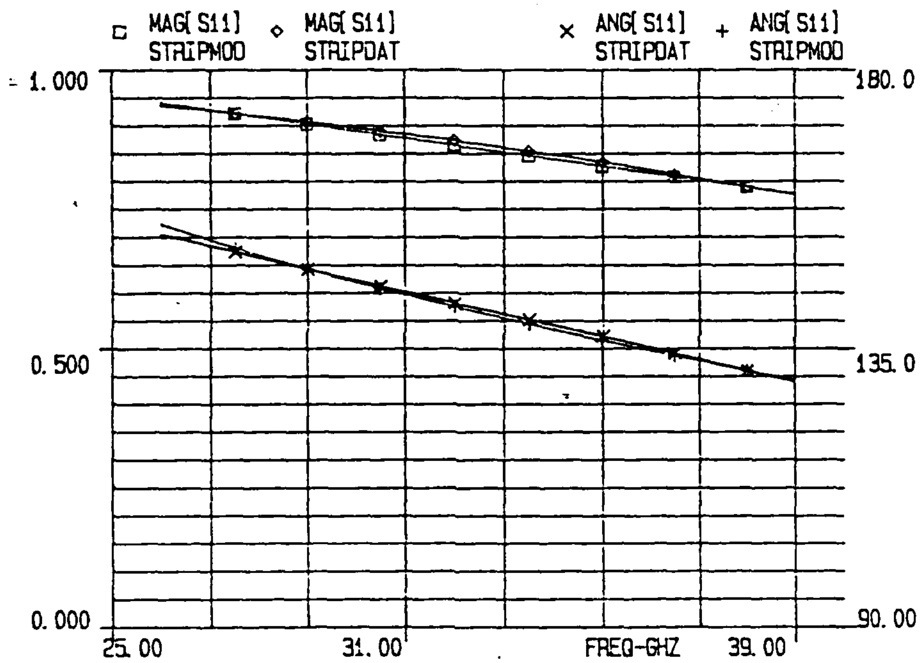


Figure 16. WR(90) model 3 scaled to Ka band. W/b = 0.25, T = 12.44 mils, a = 280 mils, b = 140 mils (b/a = 0.5).

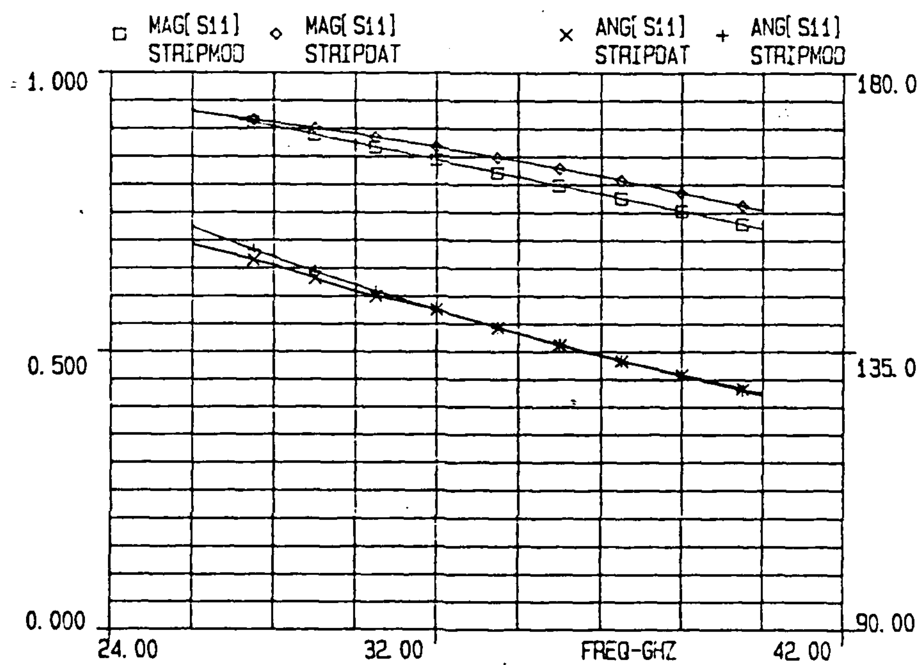


Figure 17. WR(90) model 3 scaled to Ka band. W/b = 0.25, T = 12.44 mils, a = 280 mils, b = 70 mils (b/a = 0.6).

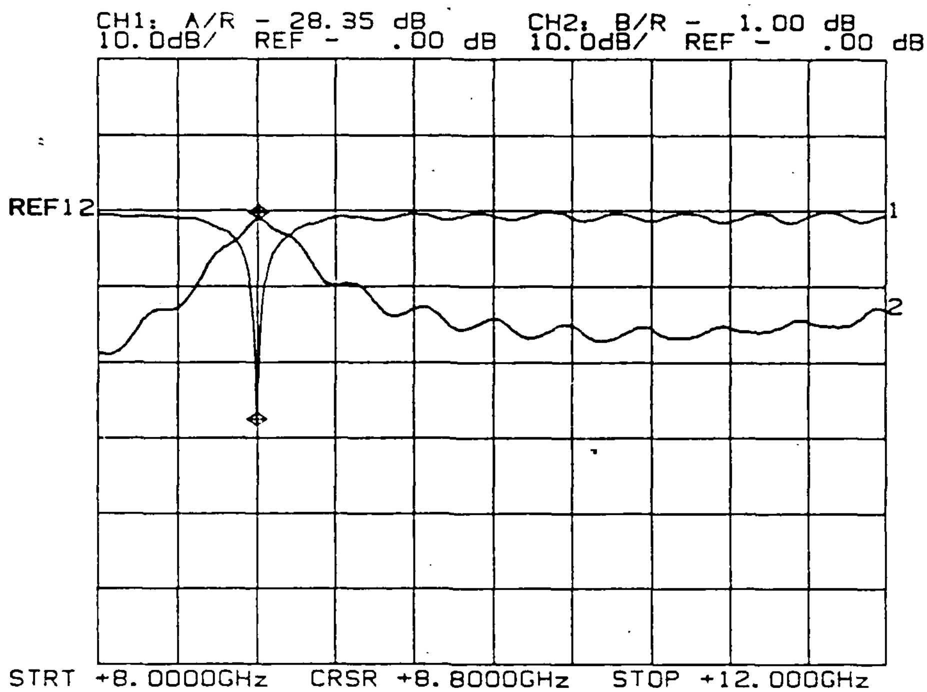


Figure 18. Measured response of a WR(90) finline resonator with W/b = 0.25, T1 = T2 = 63 mils and R= 678 mils.

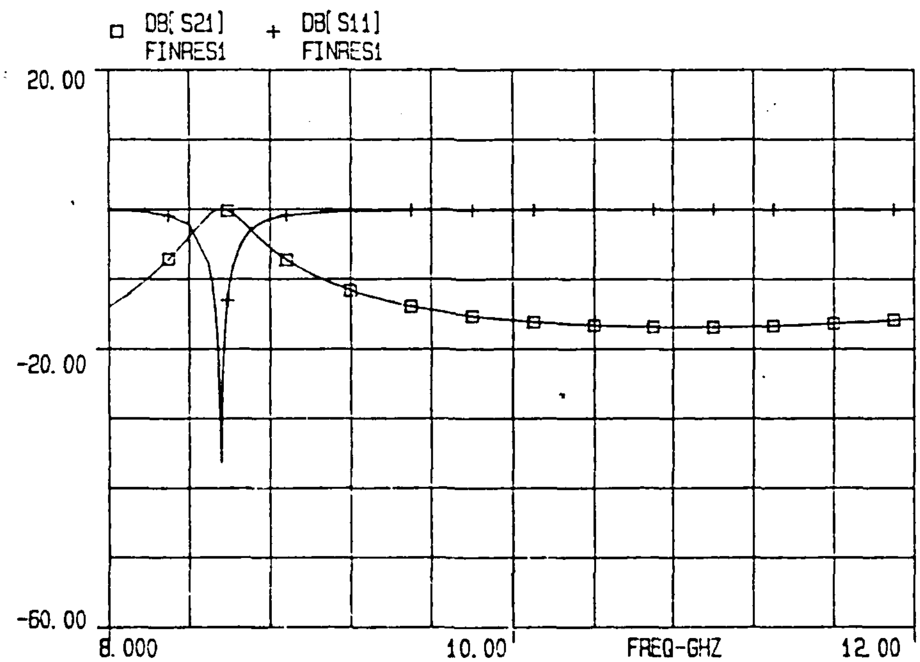


Figure 19. Predicted response of the WR(90) finline resonator using model 1 with L selected for best fit to magnitude of s_{11}.

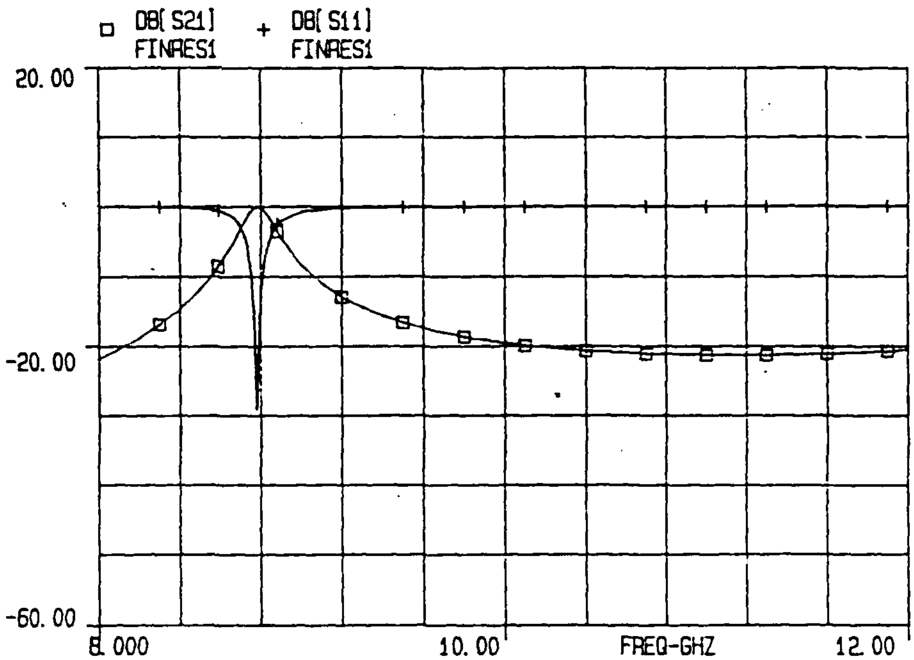


Figure 20. Predicted response of the WR(90) finline resonator using model 1 with L selected for best fit to angle of s_{11}.

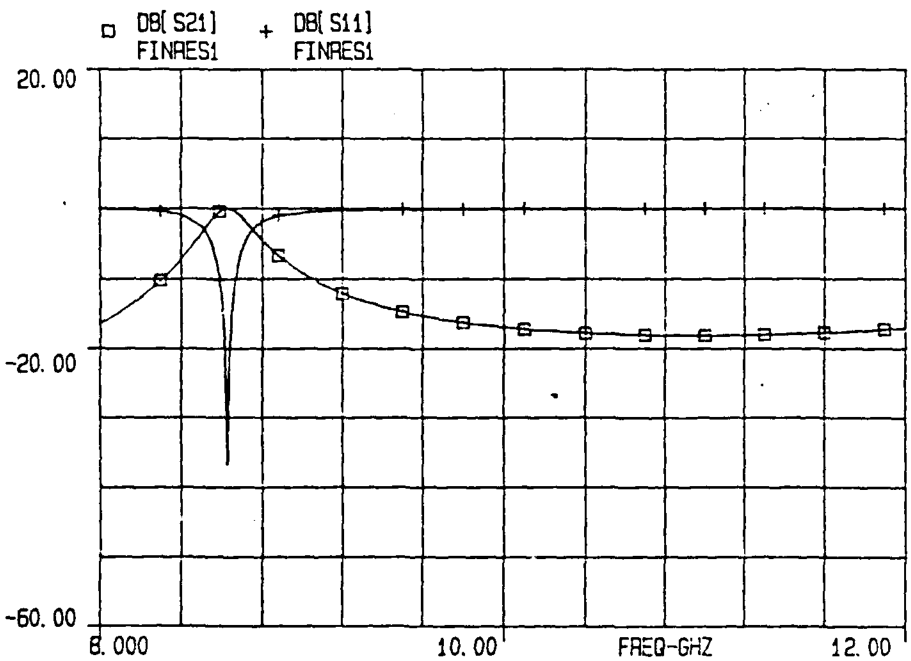


Figure 21. Predicted response of the WR(90) finline resonator using model 1 with L selected for best overall fit to s_{11}.

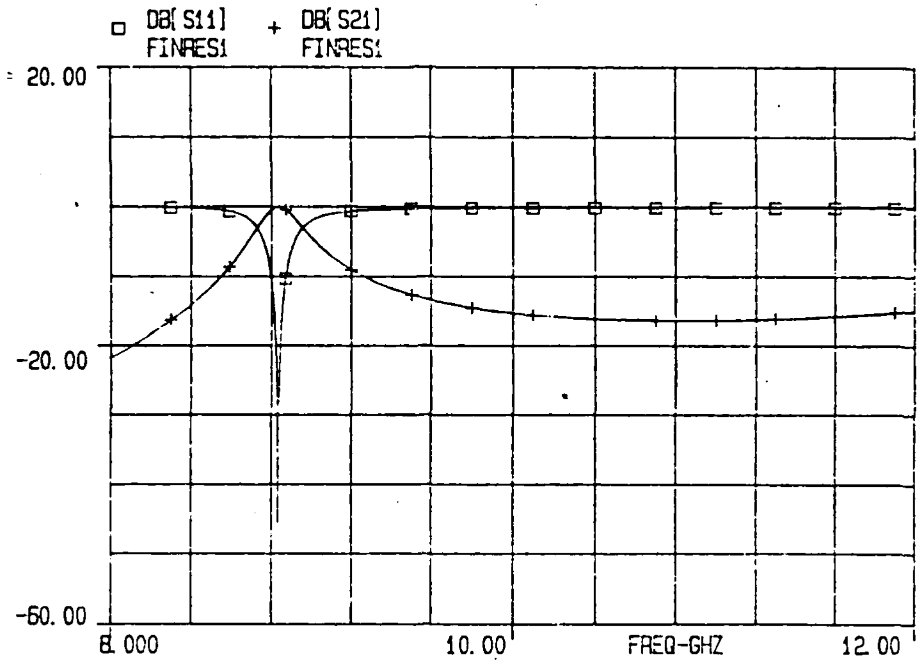


Figure 22. Predicted response of the WR(90) finline resonator using model 3.

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