Propagation of High Speed Digital Signals in Printed Circuit Board Systems - Phase I

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Abstract

This technical report is a report on the first phase of a project entitled *Propagation of high speed digital signals in printed circuit board systems* funded by Bell Northern Research as an enhancement project within the Center for Communications and Signal Processing in the Electrical and Computer Engineering Department at North Carolina State University. The goal of the first phase, which lasted 6 months, 1/1/87 — 6/30/87, was to identify the problems encountered in printed circuit design for circuits with high speed digital signals and to develop a strategy for addressing them. This working paper reports on preliminary printed circuit board measurements comparing printed circuit boards with solid and with lattice ground and supply planes, initial developments of a transmission line simulator for efficiently simulating transmission line structures, and a literature review of techniques for the simulation of transmission line structures found on printed circuit boards.

Contents

1 Introduction

This is a report on an investigative project looking at digital signal propagation in systems incorporating printed circuit boards (PCBs). This project is concerned with reflection, cross-talk, and ground and supply noise phenomena in digital systems. In systems with digital devices that respond to fast rise- and fall-time signals, these phenomena can result in false responses and in some cases can render a system unusable. The project is concerned with establishing criteria, that minimize and control these effects, for the design of printed circuit boards and thus developing computer aided design tools to augment existing tools for their automated design.

The project is in three phases with the work progressing from literature and product review to development and integration of a computer aided design tool. Phase I lasted six months and involved a detailed literature and product search with the aims of identifying the problems to be addressed, those products which would assist in this program, requirements for integrating the results of this study with existing circuit board design systems, and outlining a work plan for the last two phases of the project.

This report is organized into sections which describe the work of the first phase. The problem is defined in section 2 so that we can properly address the high speed digital signal propagation in printed circuit board systems. The results of the literature and product review of transmission line models and simulators is presented in section 3 and appendix A. This and private communication with leaders in the RF and microwave CAD community have helped shape the plan of work presented in section 7. We propose to focus on developing a table based printed circuit board simulator which can use the results of existing software for predicting coupling and EMI, use experimentally derived data, and use heuristicly or analytically developed models. It is highly unlikely that any one type of approach to modeling printed circuit board structures would be feasible. Preliminary experimental characterization of printed circuit boards specifically fabricated for this project are presented in section 4. A particular aim of the experimental characterization work was determining the relative merits of printed circuit boards with solid and with lattice ground plans. Results are presented showing that there is a significant difference so that boards with solid ground planes are preferred for controlled impedance applications. A printed circuit board simulator — CAPNET — currently under development in the Center for Communications and Signal Processing is discussed in section 5. This simulator will be computationally efficient and is being developed as an assistant to an intelligent auto-router in laying out printed circuit board tracks. and as a stand alone simulation tool. In section 6 we summarize the tangible technology that has been transferred to Bell Northern research (BNR).

2 Problem Definition

The problem addressed is how to design a large digital system so that reflections, cross-talk and other distributed phenomena are sufficiently controlled that digital circuits function

correctly. The systems being considered have clock speeds of up to 50 MHz but respond to pulses with rise- and fall-times of $\frac{1}{2}$ — 3 ns (corresponding to requiring the circuit board tracks and associated components to have 3 dB bandwidths of 700 and 100 Mhz respectively). In view of projected developments in digital technology it was decided that characterization to 3 GHz was required.

In a PCB system digital signals will propagate at a speed of about $\frac{1}{2}c$ (assuming ϵ_r for PCBs of about 5). — or about 15 cm per nanosecond. Branching, impedance changes of the conductor paths, device loading and finite device drive capability will significantly affect rise- and fall-times $\frac{1}{2}$ — 3 ns for conductor lengths in excess of 2 — 12 cm.

Controlled impedance design of a PCB requires that a ground conductor near the signal-carrying conductor be used. The proximity of the conductors localizes fields so that cross-coupling effects and nonuniform propagation characteristics are minimized. While not practical in large PCB systems, using a microwave-type circuit would solve the problem and it should be possible to adapt to printed circuit board design many of the techniques developed for the design of microstrip circuits [2] [3] [4] [9].

3 Literature Review

A preliminary search of the literature has not uncovered any significant discussion of the propagation of fast rise- and fall-time digital signals in printed circuit board systems. However there is a large body of literature related to the subject that can be broadly categorized as general transmission line theory or as propagation of signals in planar circuits. Basic transmission line theory is well established [1]. Recent work, relevant to pulse propagation in PCB systems, has discussed digital signal propagation in telephone, computer and power line distribution networks [5,6]. Printed circuit boards are a class of planar circuits with multi-level metallization lines but without a clearly defined reference ground plane. However studies of propagation in planar circuits have largely been restricted to structures with single [4,7] or multi-level [2,8,9] metallization lines above infinite ground planes. These structures are typical of UHF and microwave circuits and the effects of discontinuities and of coupling are reasonably well characterized. While most of the work cited is confined to analog signals, the work can be extended to pulse propagation.

We have undertaken an extensive literature review of microwave transmission lines, coupling and discontinuities. We report on this review in Appendix A. The emphasis of this review is on models of transmission line structures and strategies for simulating printed circuit boards.

We have had several discussions with commercial vendors and investigated several RF and microwave simulation and CAD packages. There is nothing suitable for our purposes but the most promising — both because of user support and modeling accuracy — are SUPERCOMPACT [11] and TOUCHSTONE [10]. These are linear frequency domain analysis programs and do not contain the types of circuit elements typicly found in printed circuit boards. However they do provide simulation capability for some simple structures and aid in interpreting measurements and in model development. TOUCHSTONE is used

at BNR and at NCSU for this purpose.

4 Experimental Characterization

The aims of the initial experimental characterization of phase I were establishing the measurement requirements for characterizing circuit board structures, determining the modeling capability required, and comparing the characteristics of boards with solid and with lattice ground planes with respect to their applicability to controlled impedance design.

Transmission line models of four layer printed circuit board transmission lines, with track widths from 8 mil to 100 mil, were developed by fitting the physical microstrip transmission line model of TOUCHSTONE [10] to measured data. Measurements were made using the test setup of Figure 1 and the test board shown in Figure 2. The nomenclature denoting the tracks is described in Table 1. The nomenclature is of the form mxyn were ms indicates that we are dealing with a microstrip line, x is either s indicating that the printed circuit board has a solid ground plane or, if l, that the board has a lattice ground plane. If y is b then there is a bend in the middle of the transmission line. The final character is the number denoting the width of the track.

The boards were cut to length using a diamond saw and electrical contact between the adapter tab and the printed circuit board track was made using silver paste and the holders were held in place using clamps. The track lengths in all cases were 248 mm with metallization thickness of 25 μm and dielectric thickness of 0.35 mm. The ground plane closest to a track was exposed by first grinding the side of the printed circuit board opposite the track until the metal first showed, and then hand sanding until the entire ground plane was visible. Electrical contact between a holders and the ground plane was again ensured by again using silver paste. This measurement structure is the most accurate method for characterizing printed circuit board traces as it minimizes parasitics associated with the connection to the board. However it is destructive and and time consuming and is not recommended for Phase II measurements. The measurements reported here provide a reference for the nondestructive measurement procedure to be used in phase II.

Measurements and the results of modeling the 8 mil wide line are shown in Figures 3-6. The microstrip line model was obtained by fitting the model of Figure 7 to measured s parameters. Model fitting was achieved using TOUCHSTONE [10]. The measurements are modeled reasonably well up to 10 GHz. However there is a major discrepance between the measured and calculated magnitude of S_{21}. S_{21} is the transmission S parameter and its magnitude is the ratio of the voltage wave arriving at port 2 to the incident voltage at port 1, all measured in a 50 Ω measurement system. With the microstrip models in TOUCHSTONE the only loss mechanism accounted for is metallic resistivity modified by the skin effect. This results in a loss, expressed in dB/m, increasing as the square root of frequency up to 10 GHz as is seen in Figure 5. The actual loss to 10 GHz — expressed in dB/m — has almost linear dependence on frequency indicating a substantial loss due to radiation or dielectric loss.

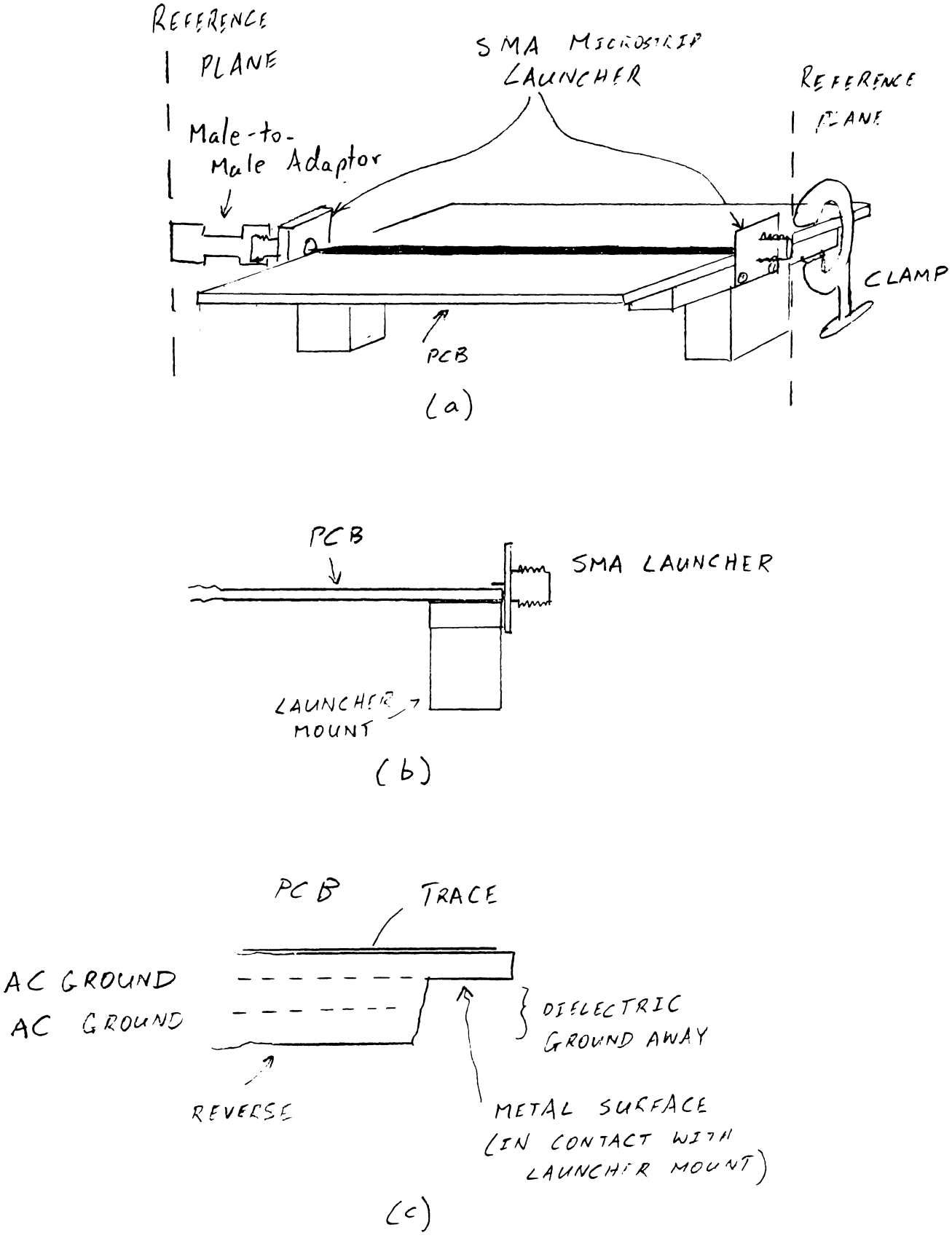
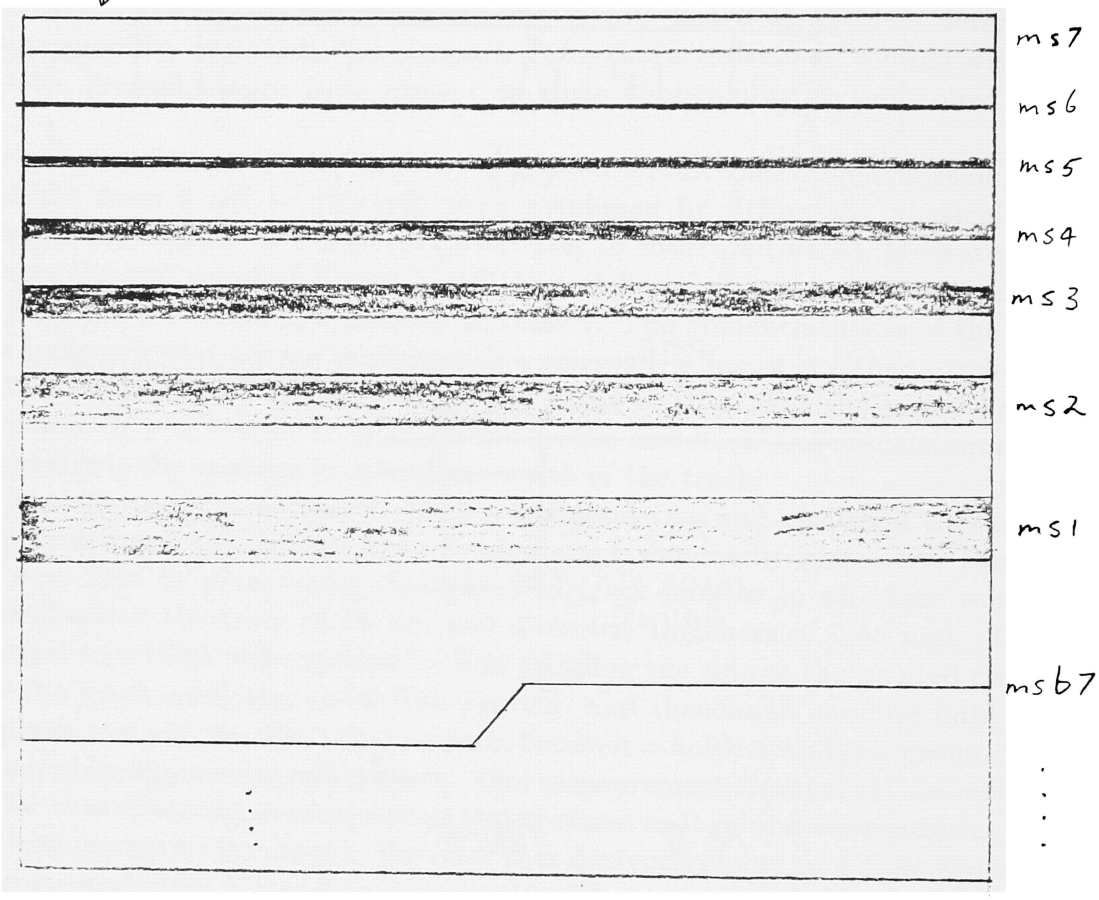
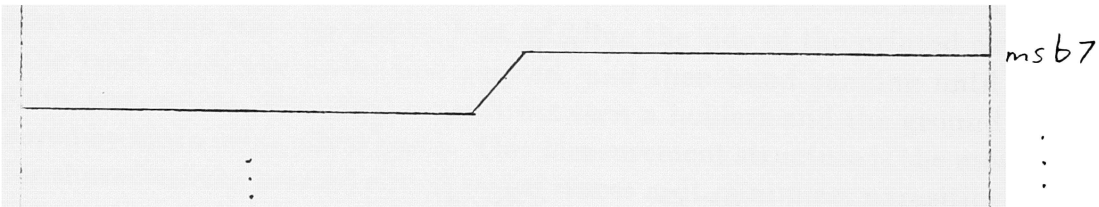


Figure 1: Test setup for printed circuit board measurements, (a) test structure showing reference planes, (b) detail of adaptor mounting, and (c) detail of electrical contact to ground plane of printed circuit board.

SECTION CUT FROM
BOARD USING DIAMOND SAW





Label	Nominal Width (mils)	DESCRIPTION
ms7	8	Microstrip, Straight, Solid Ground
ms6	13	
ms5	16	
ms4	20	
ms3	25	
ms2	50	
ms1	100	
ml7	8	Microstrip, Straight, Lattice Ground
ml6	13	
ml5	16	
ml4	20	
ml3	25	
ml2	50	
ml1	100	
msb7	8	Microstrip, Bend, Solid Ground
msb6	13	
msb5	16	
msb4	20	
msb3	25	
msb2	50	
msb1	100	
mlb7	8	Microstrip, Bend, Lattice Ground
mlb6	13	
mlb5	16	
mlb4	20	
mlb3	25	
mlb2	50	
mlb1	100	

Table 1: Description of nomenclature describing tracks.

8 mil Line on Solid Ground Plane

SET 3 5.13.87

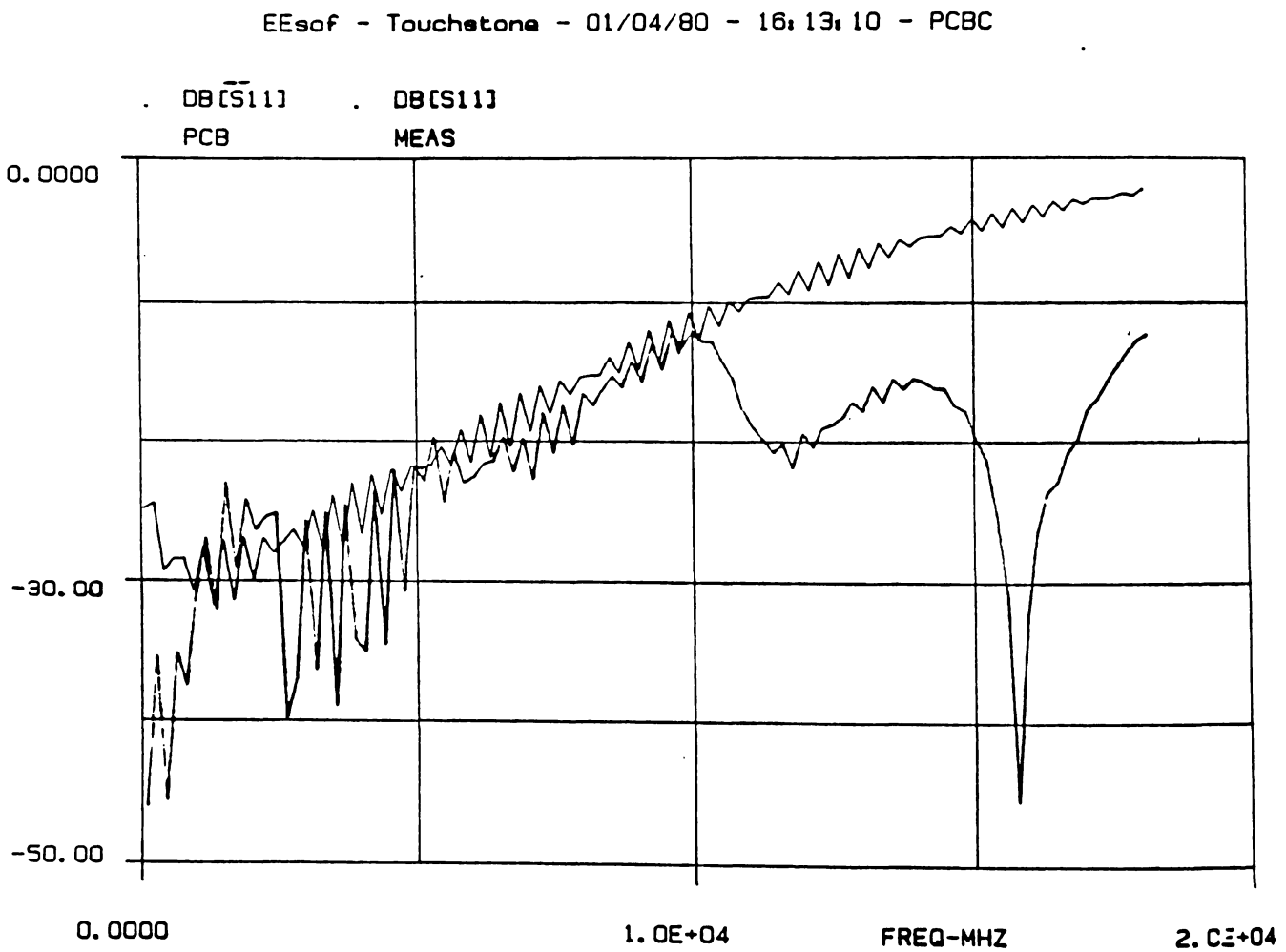


Figure 3: Comparison of measured and modeled S_{11} magnitude versus frequency for an 8 mil line above a solid ground plane. (ms7 SET 3)

8 mil Line on Solid Ground Plane

SET 3 5.13.87

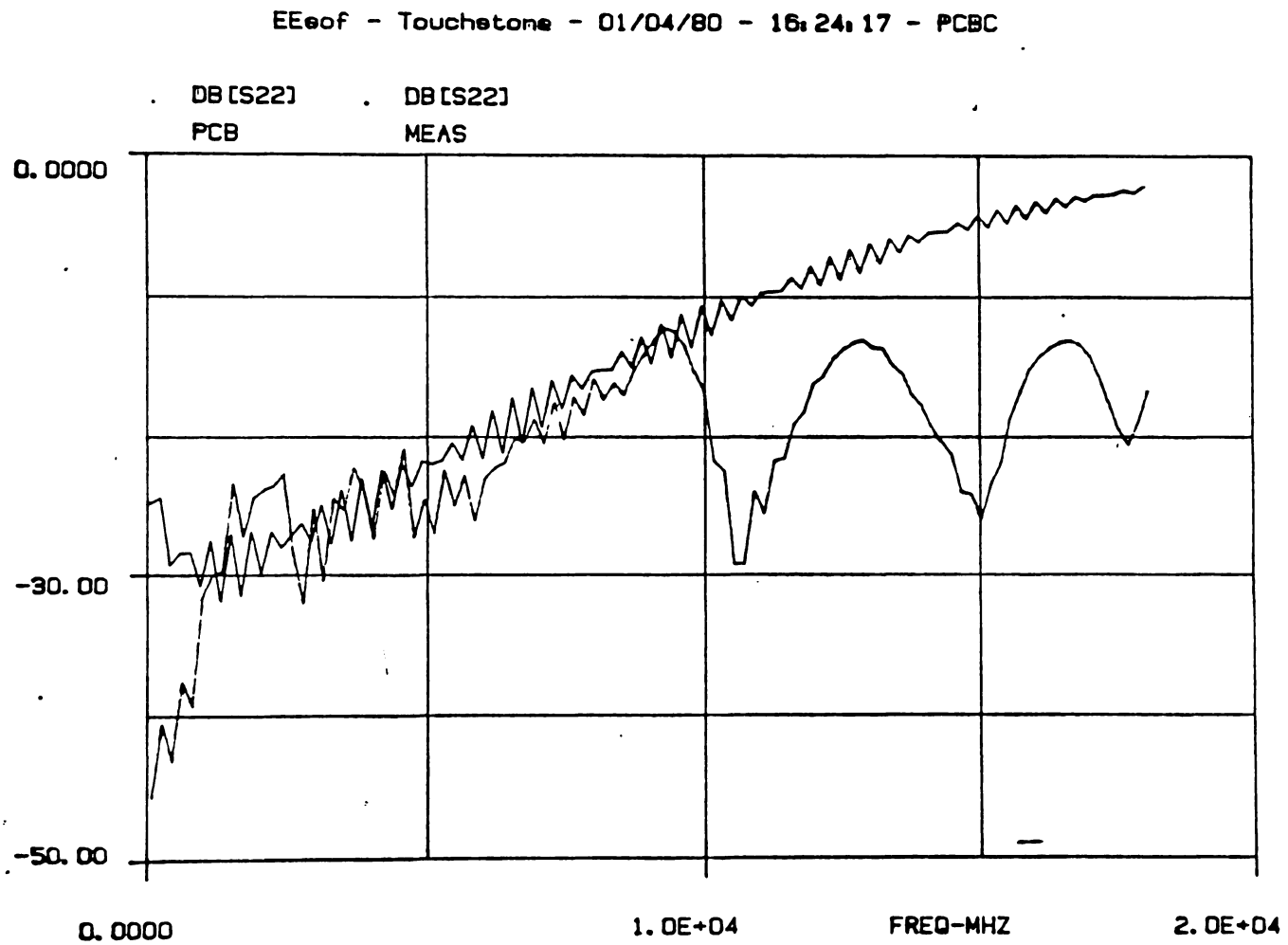


Figure 4: Comparison of measured and modeled S_{22} magnitude versus frequency for an 8 mil line above a solid ground plane. (ms7 SET 3)

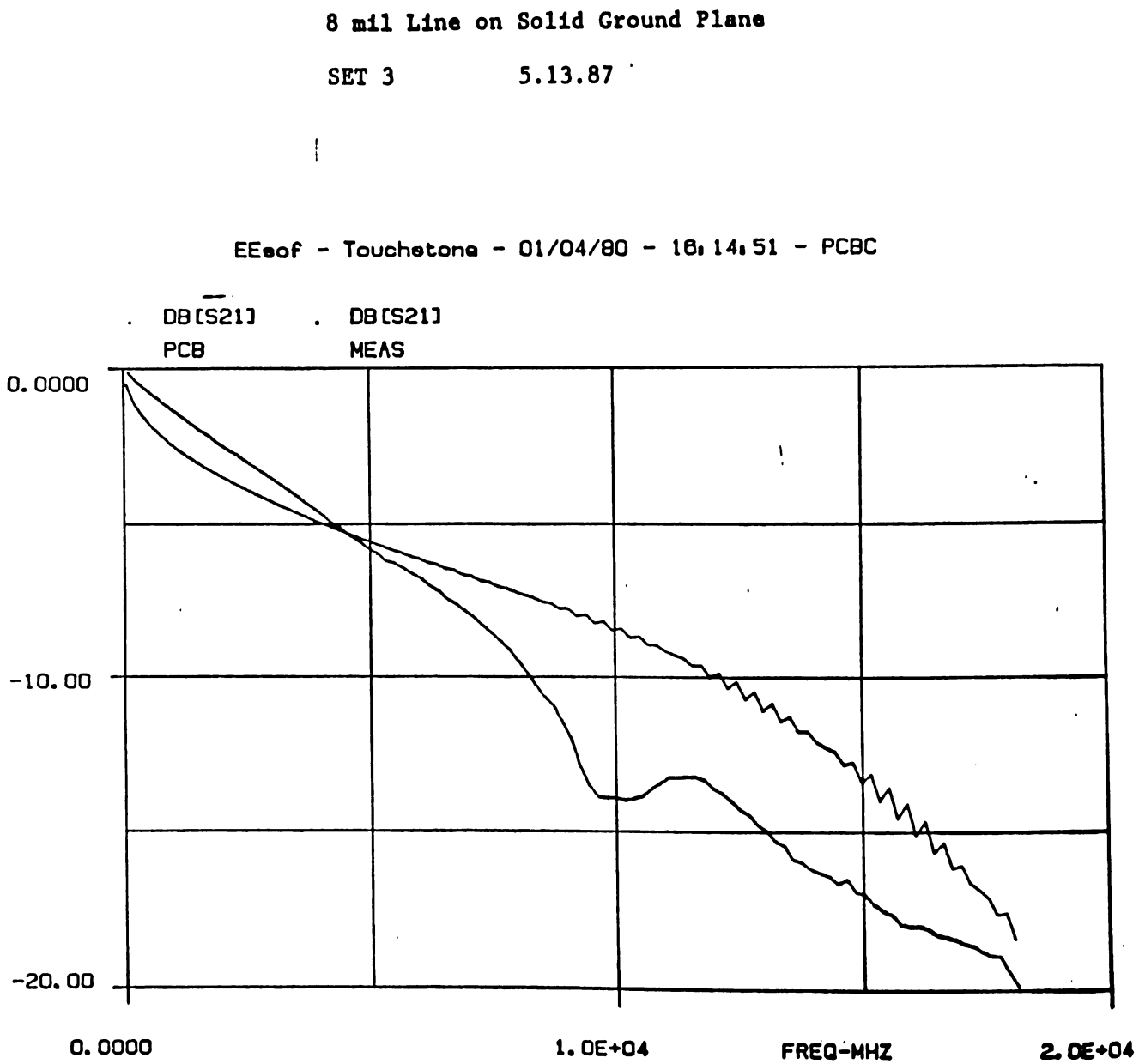


Figure 5: Comparison of measured and modeled S_{21} magnitude versus frequency for an 8 mil line above a solid ground plane. (ms7 SET 3)

8 mil Line on Solid Ground Plane

SET 3 5.13.87

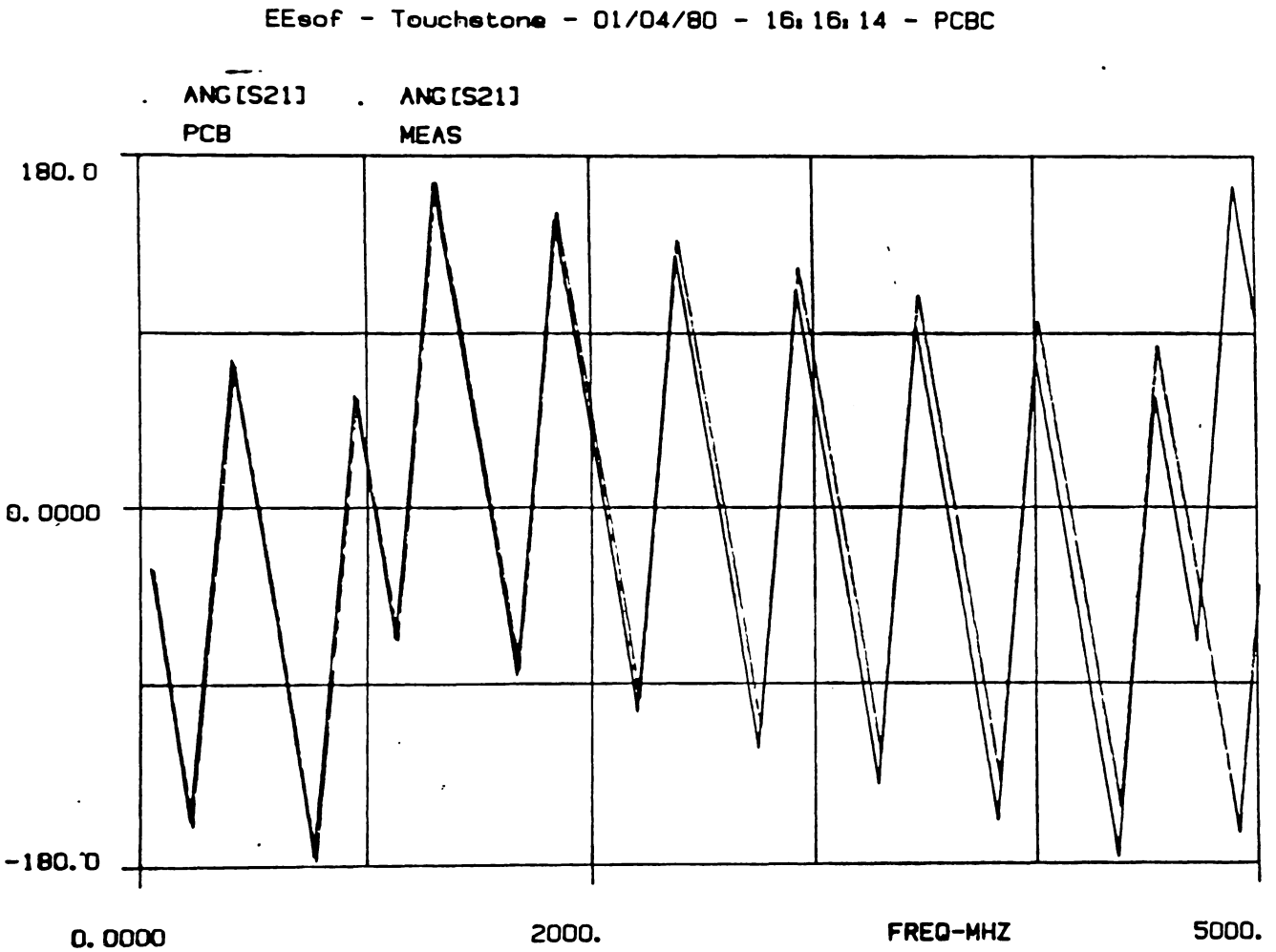


Figure 6: Comparison of measured and modeled S_{21} phase versus frequency for an 8 mil line above a solid ground plane. (ms7 SET 3)

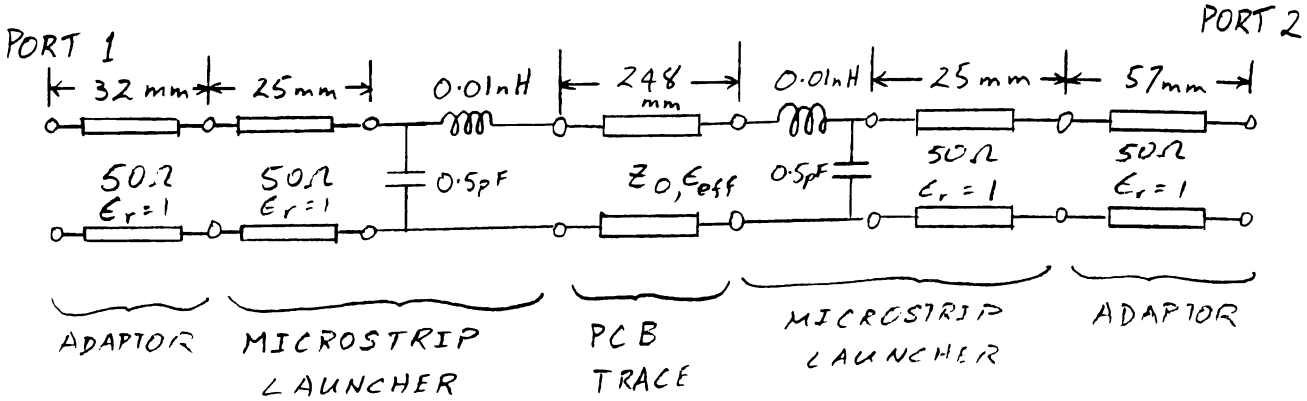


Figure 7: Model of measurement set up.

Model fitting is not a straightforward procedure and the modeling accuracy described here could not be obtained by simply using the TOUCHSTONE optimizer. The procedure followed here was to use the measured physical length of a track and then to optimize the phase match of S_{21}. The magnitude match of S_{21} was achieved and finally a match to all the measurements was obtained. Measurements and the modeling procedure were performed for track widths from 8 to 100 mil over solid and lattice planes and for straight lines and lines with a single bend. A summary of the modeling results are presented in Figure 8 and Tables 2 and 3. The calculated values were obtained using the program Z0STRP described in Appendix C. The calculated results are for tracks with solid ground planes and it appears impossible to develop a physical model for a track over a lattice ground plane. The calculated characteristic impedance is reasonable agreement with that measured for the solid ground plane test board. However the calculated loss is much less than that measured. The extra loss is either dielectric or radiative loss. One of the tasks of Phase II will be determining the relative importance of each type of loss.

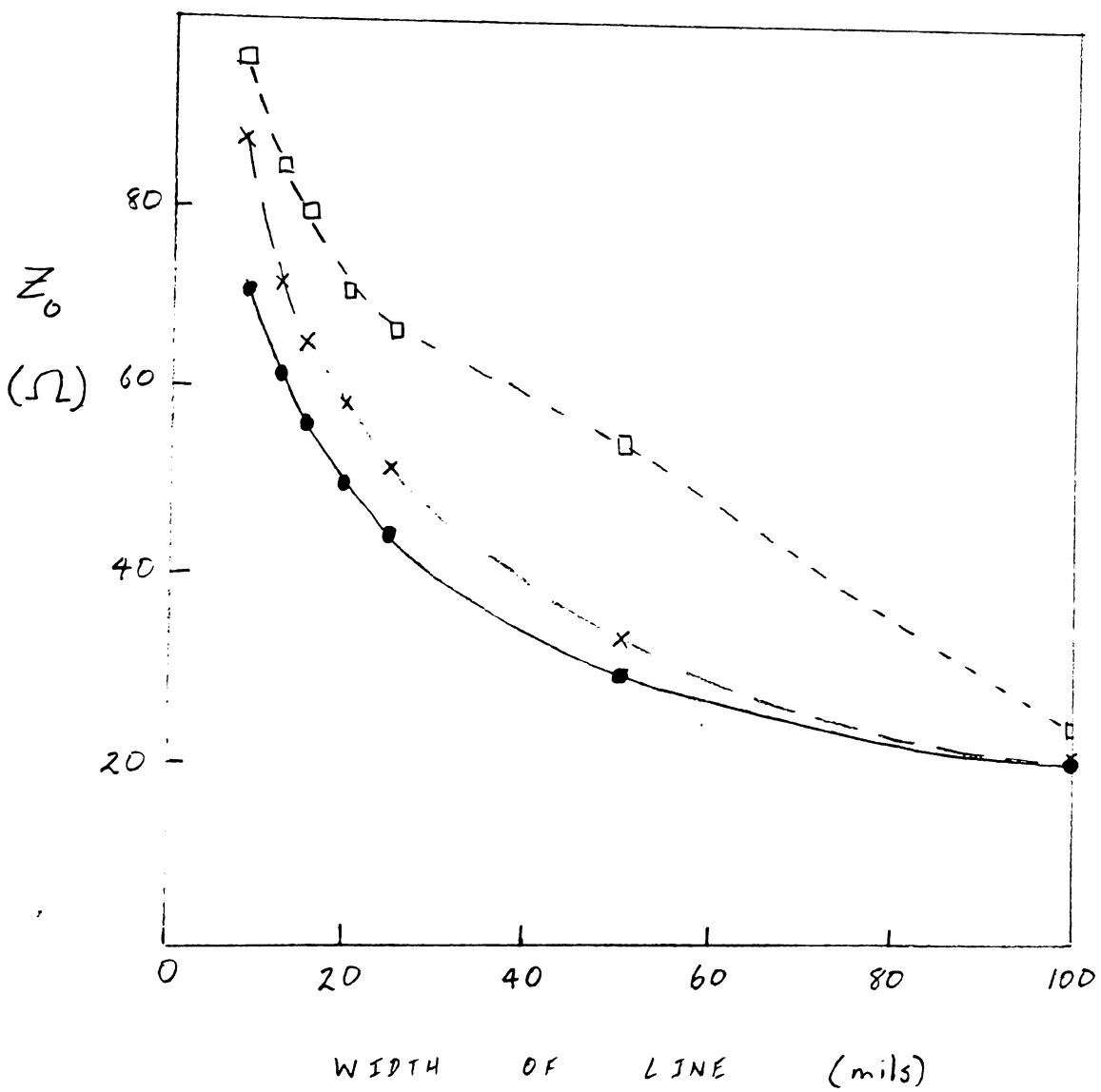


Figure 8: Characteristic impedance versus track width w for (a) calculated using Z0STRP, (b) modeled from measurements with a lattice ground plane, and (c) modeled from measurements with a solid ground plane.

No.	Nominal Width (mils)	CHARACTERISTIC IMPEDANCE calculated (Ω)	meas. solid (Ω)	meas. lattice (Ω)	ϵ_{eff} calculated	meas. solid	meas. lattice
7	8	87	72	97	2.85	2.95	3.32
6	13	71.6	60	85	2.96	2.97	3.40
5	16	65.0	56	80	3.01	2.98	3.50
4	20	58.2	48	72	3.06	3.05	3.60
3	25	51.5	44	67	3.12	3.10	3.70
2	50	33.1	28	55	3.32	3.20	3.92
1	100	19.6	20	23	3.54	3.45	3.60

Table 2: Comparison of calculated and modeled impedance, effective dielectric constant and loss of tracks.

No.	Nominal Width (mils)	LOSS dB/m calculated	meas. solid	meas. lattice
7	8	1.42	5.0	5.5
6	13	1.25	5.0	6.0
5	16	1.17	6.0	7.0
4	20	1.14	10.	8.0
3	25	1.11	5.0	6.0
2	50	1.04	5.0	6.0
1	100	1.01	8.0	5.0

Table 3: Comparison of calculated and modeled losses.

No.	Nominal Width (mils)	LOADING PER METER r (Ω)	l (nH)	c (pF)
7	8	248	489	65
6	13	178	410	80
5	16	152	376	89
4	20	132	339	100
3	25	114	303	114
2	50	69	201	184
1	100	40	123	320

Table 4: Calculated loading effects of lines at 1 GHz.

5 Simulator Development

In this section CAPNET a computer program for the graphical capture, modeling and analysis of transmission line networks is introduced. This program can be used to solve transmission line networks so that the propagation of high speed digital pulses may be investigated. A complete description of the technique used to solve the transmission line network and a description of the program is presented in Appendix B. Basically, the program uses recursive tree structure representations of the transmission line networks, and recursive routines to traverse the tree in order to solve the network. For a sinusoidal excitation of the network, the program first computes and updates the propagation constant and characteristic impedance for each section of transmission line. Next, it recursively computes the impedances throughout the network storing the impedances in the tree data structure. Finally, the current and voltage at all nodes are computed recursively starting from the source.

The program does allow for cascade sections in the network representing structures such as vias and bends as complex ABCD parameters. Time domain results are obtained by computing the frequency response of the network at discrete frequencies and then using FFT's to compute time domain responses.

The program currently models loads for various technologies (TTL, LSTTL, STTL, etc.) using functions which return impedance as functions of frequencies. The characteristic impedance and the propagation constant are calculated at each frequency directly from the geometry of the microstrip line representing the printed circuit board trace. The effects of losses due to skin effect and lossy dielectrics and fringe effects are taken into consideration.

The program reads data about different traces from the file TRANS_TYPES.DAT which contains a string for the name of the trace ("thin" for example) and the thickness, height, width and dielectric constant of the material (see Figure 9. Thus a library of such traces and other transmission lines may be used.

```
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C ROUTINES DESCRIBING LOADS

#include "graphpcnet.h"
#include math
#define pi 3.1415926
#define RESIST 100.0

complex ttl(freq)
float freq;
{
    complex z_load;
    float r,c,tmp,w;

    r=2000.0;
    c = 15e-12;
    w=2.0*pi*freq;
    tmp = w*r*c;
    tmp= 1 + tmp*tmp;
    z_load.re = r/tmp;
    z_load.im = (-1.0)*w*c*r*r/tmp;
    return(z_load);
}
complex sttl(freq)
float freq;
{
    complex z_load;
    float r,c,tmp,w;

    r=280.0;
    c = 10e-12;
    w=2.0*pi*freq;
    tmp = w*r*c;
    tmp= 1 + tmp*tmp;
    z_load.re = r/tmp;
    z_load.im = (-1.0)*w*c*r*r/tmp;
    return(z_load);
}
complex ttlls(freq)
float freq;
{
    complex z_load;
    float r,c,tmp,w;

    r=15000.;
    c = 10e-12;
    w=2.0*pi*freq;
    tmp = w*r*c;
    tmp= 1 + tmp*tmp;
    z_load.re = r/tmp;
    z_load.im = (-1.0)*w*c*r*r/tmp;
    return(z_load);
}
complex resist(freq)
float freq;
{
    complex z_load;

    z_load.re = RESIST;
    z_load.im = 0.0;
    return(z_load);
}

complex via(freq)
float freq;
{
    complex z_load;

    z_load.re = 1.0e+8;
    z_load.im = 0.0;
    return(z_load);
}


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FILE DESCRIBING TRACES USED IN PCB
SIMULATION

thin 0.002032 2.0e-5 0.00381 5.5
via  100.0 0.0 0.0 1.0


--------------------------------
```

Figure 9: Trace file for CAPNET.

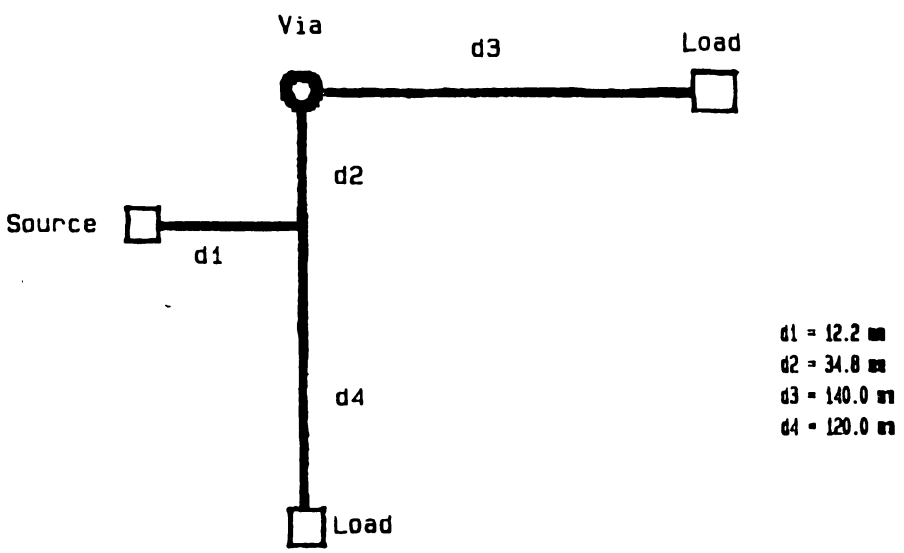

Figure 10: Example printed circuit board trace.

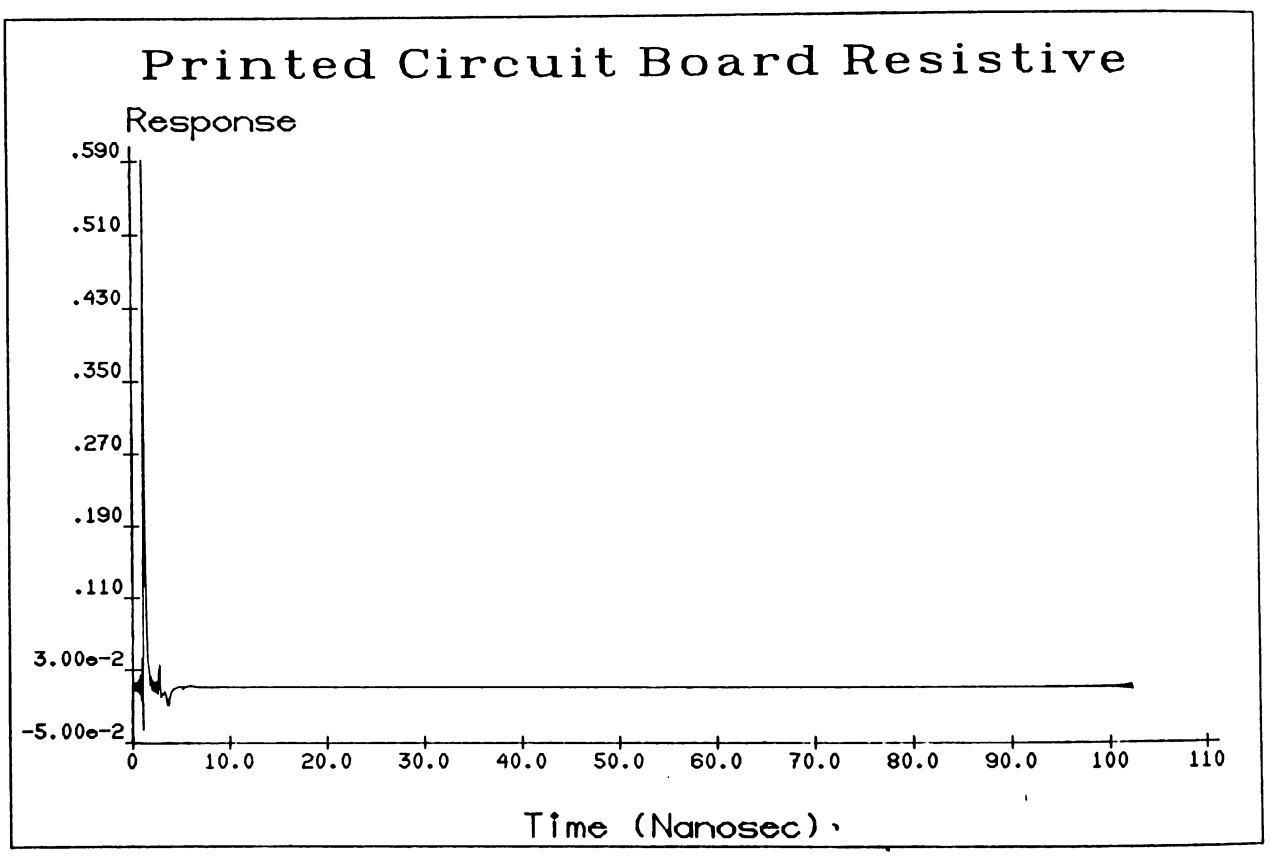

Figure 11: Impulse response.

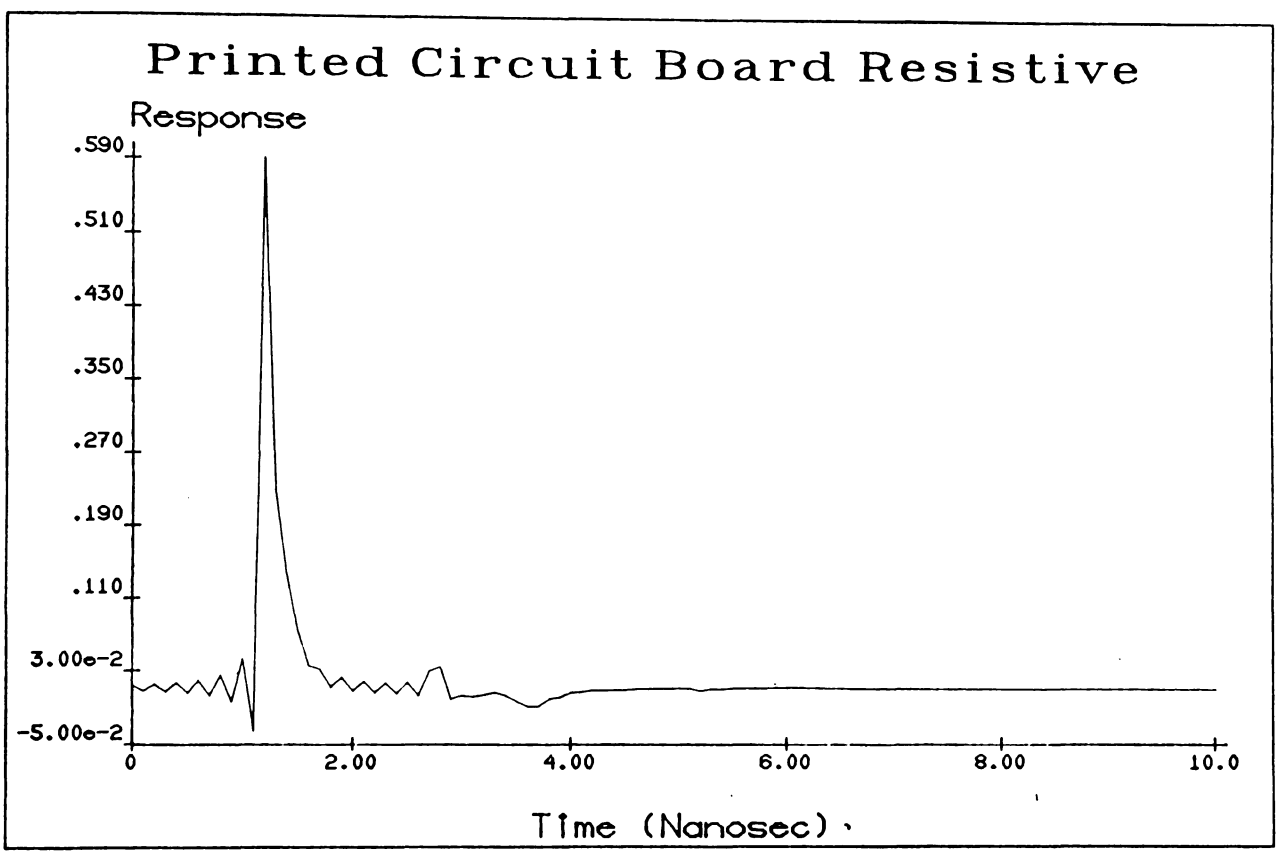

Figure 12: Impulse response with expanded time scale.

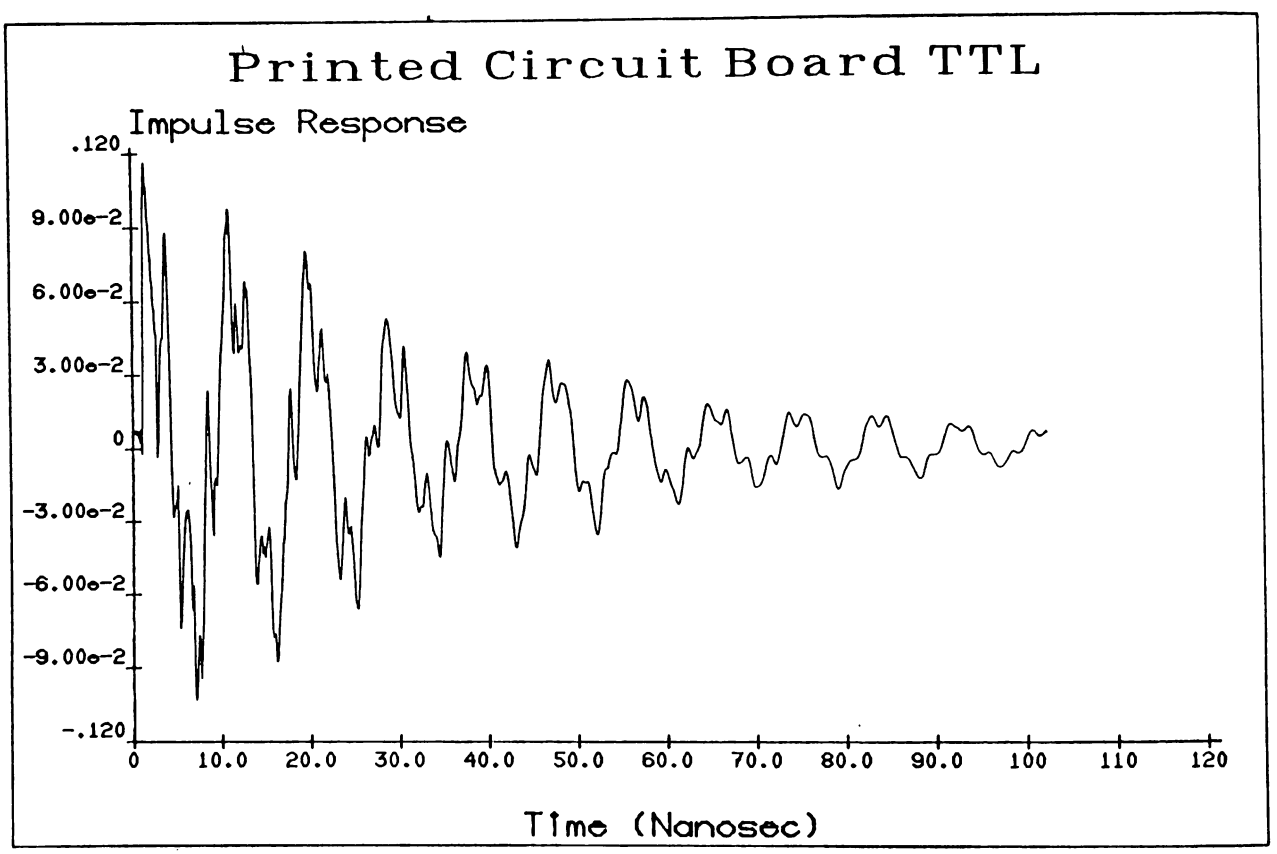

Figure 13: Impulse response with TTL load.

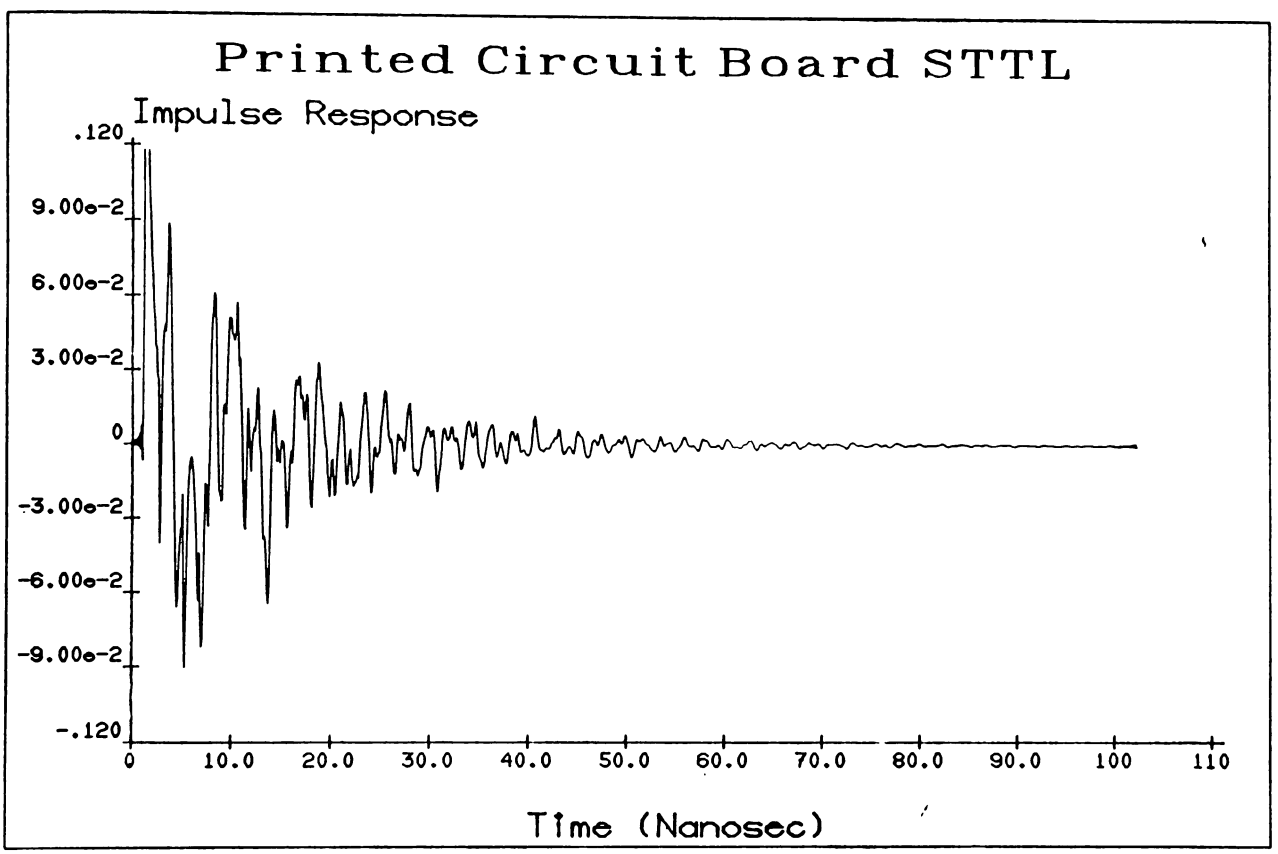

Figure 14: Impulse response with STTL load.

6 Technology Transfer

The products of this project delivered to BNR include a computer program for simulating lossy microstrip transmission lines embedded in possible lossy dielectric, establishment of a measurement capability, and establishment of a modeling capability. Major contributions to BNR include clarification of the problems that needs to be solved and formulation of a systematic plan to address them. The principal investigator of this project at North Carolina Sate University has on many occassions effectively acted as a consultant on printed circuit board and system design issues that relate to transmission lines and electromagnetic fields. Furthermore he has acted as an academic advisor to various members of the BNR staff undertaking graduate education. One result of university/industrial projects such as this is change in the orientation of course content to help train people in issues that are relevant to industry.

7 Plan of work

This study involves the investigation of pulse propagation in printed circuit board systems with emphasis on reflections, cross-talk, and ground and power supply noise — phenomena that have been identified as having limiting effects on PCB performance. With reference to these phenomena the following issues are to be addressed.

- What software tools are available for the simulation and CAD of signal propagation in PCB systems?
- What lessons can be learned from the well established procedures for the design and operation of small planar analog circuits at microwave frequencies?
- What design methodologies can be adopted to improve the rise- and fall-time performance of new and existing PCB systems?
- What effect do multi-level boards have on pulse propagation performance?
- Experimental characterization of pulse propagation in PCB systems.
- What new PCB materials are available with improved high frequency performance?
- Development of a computer program for simulating the propagation of digital signals in PCB systems.
- Development of design rules for the automatic routing design of PCB systems for high frequency performance.
- What is the effect of coupling and cross-talk on propagation performance? What procedures can be followed to minimize this effect?

- What effect does the backplane have on intra-board and inter-board propagation performance? How can this performance be improved?
- What is the effect of device loading and finite drive capability? What can be done to minimize the effect on performance of interchanging devices.

Development of the plan of work for phase II was a major aim of the first phase of the project. It includes electromagnetic characterization of the PCB systems, development of circuit models for various transmission line discontinuities, and the development of linear transmission line simulation and CAD tools. As agreed upon the majority of the measurements will be undertaken at BNR. The thrust of phase II is to produce usable results, although incomplete, in the first six months and to complete, document and augment the work in the final six months.

Another goal of phase II will be developing methodologies for extending the transmission line simulator to nonlinear loads and drivers. It will be possible to adapt this work to SCAMPER to improve its ability to simulate coupled transmission lines and other PCB transmission line structures.

The Plan of work for phase II is as follows

- Static Characterization of 4 layer boards
- Development of models for printed circuit board configurations for four layer boards.
- Experimental RF characterization of 4 layer boards.
- Report on PCB layout extraction.
- Experimental RF characterization of 6 layer boards.
- Development of models for printed circuit board configurations for 6 layer boards.
- Report on decoupling requirements based on experimental characterization and simulations.
- Development of user friendly transmission line program for PCBs with linear loads.
- Report on efficient simulation of transmission line structures by numerical integration programs like SCAMPER with prototype program and documentation enabling incorporation in SCAMPER.
- Report on future directions of project.
- Enhancements to the transmission line program for PCBs
- Final report on Phase II prepared in cooperation with BNR.

The final PCB CAD tool will be written in the C programming language and will be transportable so that it can be used by engineers at BNR. Our current thinking is to produce a program with inputs and outputs as follows:

Inputs

- Characteristics of conductor tracks (e.g. width, resistivity) on PCB and backplane. These may change with frequency.
- Physical topology. This will include the specifications of line lengths and physical layout.
- Line terminations and characteristics of drivers.
- Performance specifications including rise- and fall-time, crosstalk levels (noise margin), overshoot and undershoot limits etc.

Outputs

- Characteristics of signals at the end of conductor lines — e.g. rise-time, fall-time, overshoot, ringing.
- Circuit loading.
- Crosstalk, noise margin levels.
- Flag indicating whether specifications at an individual node have been met.
- Other outputs as yet unspecified.

The third phase of this project is envisaged to to concentrate on integrating the software developed as part of this project in existing PCB CAD tools — particularly those at Bell Northern Research. At this stage it is planned to extend the the CAD tool developed in Phase II two modules — one module being a design rule generator and design checker prior to auto-routing and the other module performing simulation of the circuit board system. It anticipated that the second module will function as a pre- and post-processor to a general purpose circuit simulation package such as SCAMPER or SPICE. As much as possible the modules developed in the third phase will be interfaced directly to Bell Northern Research's current circuit board design system. A more detailed plan of work for Phase III will be formulated when Phase I is completed and Phase II is in progress.

There are two main research aspects in Phase II. The first is experimental characterization of digital systems incorporating printed circuit boards with the aim of identifying and modeling those structures that result in reflections, cross-talk, and ground and supply noise. Proposed solutions will also be verified experimentally. The second aspect is development of Computer Aided Design (CAD) tools for the design of the transmission networks in PCB systems. This CAD tool will be written in the C programming language and will be transportable so that it can be used by engineers at BNR.

8 Conclusions

The major results of phase I of the project were identification of the problems encountered in the design of printed circuit boards with high speed digital signals, a literature review of the printed circuit board simulation approaches, and development of a plan of work for phase II of the project. As well it was concluded that circuit boards with solid rather than lattice ground and supply planes should be used for controlled impedance applications.

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[10] TOUCHSTONE, EEsof Inc., Westlake Village, CA.

[11] SUPERCOMPACT, Communications Consulting Corp., Paterson, NJ.

Appendix A. Review of microstrip transmission line studies and simulations

An Overview of Microstrip Transmission Line Systems

Countless papers have been written on almost every aspect of microstrip transmission line systems (MTL) in the last decade. While many provide helpful information for specific problems, few address the system level problem. Simulating special cases is one thing but handling an entire MTL system such as found in a printed circuit board (PCB) or a VLSI chip is outside of their range. This paper reviews what's been studied in the areas of MTLs, coupling and discontinuities. We'll also look at what useful information can be gleaned from these scores of papers and which ones can provide useful tools for further work.

A Review of Microstrip Transmission Line Studies and Simulations

There have been countless papers written on transmission lines and microstrip transmission lines in particular in recent years ranging from detailed electromagnetic fields analysis to experimental studies of just about anything one is interested in. In this paper we'll look at several majors aspects of these publications and in particular those dealing with simulation of microstrip and coupled microstrip lines.

The significant papers of interest to this review and the microstrip print circuit board (PCB) transmission line simulator project can be lumped into three general categories:

1. Microstrip Transmission Line Theory
2. Coupled Transmission Lines
3. Discontinuities and Characterization of Microstrip Lines

The papers on General Transmission Line Theory won't be discussed in this review except as they pertain to the other papers. The papers dealing with general Microstrip Theory are "only" important as they provide background information and ideas that the simulation and characterization papers use as their basis. Of most interest are those papers dealing with coupling phenomenon in microstrip lines; especially those that simulate multiple coupled lines. The handling of coupled lines is the single most difficult task in simulating an entire system of interconnect lines, whether they be on an IC or a printed circuit board. Since each line can couple to any other line in the system, the problem becomes horrendous if a simple brute force SPICE simulation using lumped element models is used. We'll see how many people have attempted to handle this problem in different

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ways, each having its' own tradeoffs, advantages and weaknesses. The last group of papers is important in developing realistic models and for measuring situations in the real world if the result of the work is to have any credibility. Too many papers compare their theoretical results with other's theoretical calculations which doesn't have the same meaning as comparison to real experimental measurements.

In a recent database search of all abstracts concerning microstrip transmission lines, over 700 papers were found to have been published over the past few years on this subject. Everything from measurement methods to design guides for microstrip transmission line (MTL) layouts were retrieved. Many of these papers were geared towards analysing in detail the various phenomenon and problem areas involved in using MTLs on circuit boards, VLSI chips and monolithic microwave integrated circuits (MMICs). The publications of particular interest for this project were those studying the characterization of MTLs, discontinuities in MTLs (bends, vias, etc.), and coupling between lines. To summarize very briefly, many of these papers were well written and had sound methods and conclusions. The handling of discontinuities was probably the most straightforward of the methods since a discontinuity is usually very localized (spatial) and deriving either an RLC lumped model or an S parameter representation of it is fairly simple. The characterization of the lines, including dispersion, characteristic impedance, attenuation and radiation as functions of frequency were also handled fairly well.

The real nagging problem is coupling/crosstalk between lines. At first glance there seems to be several very different methods for handling this problem but on closer exami-

nation we see that there are only two major approaches in use. The first is to model the coupling as either a lumped or "distributed lumped" RLC system and use a circuit analysis program like SPICE to calculate the system response. The variation between papers of this genre concern the method of calculation used to compute the [L] and [C] matrices. As might be expected, the tradeoff in accuracy vs computation time is present with this method; the more "distributed" the model the higher the accuracy but at a price of longer computation time. Also, for systems of MTLs of more than a few lines, the computation times become horrendous and methods to relieve this burden were the next area investigated by researchers in this field. The other method for handling coupled lines involves coming up with a network of some sort to include before and after the system of uncoupled MTLs that handles the coupling while the lines themselves are modeled as simple uncoupled lines. The most common network for this method consist of resistors and voltage and current controlled sources. The differences in the papers using this general method concerns how the decoupling is performed and the control equations that govern the dependent sources.

All of the papers either assume strict TEM mode propagation or at least quasi-TEM propagation. Most make the assumption that the "n" coupled lines under study are all parallel and many assume further that they are all in the same plane. Almost all of the MTL papers concern themselves with just the transmission lines themselves and few, if any, take into account the effect of nonlinear loads on the signals on the lines. In fairness, it's noted that methods using SPICE or similar programs can incorporate these effects separately but this just further increases the amount of computation time required.

For a circuit simulator of any type to be of practical use, the response time must be kept as fast as possible for the user to keep "on track" their line of investigation. Furthermore, since changes in a circuit design or layout are usually incremental (the slate is seldom wiped clean every time a problem is encountered or a change is made to a design) there should be a method of using this to the simulators advantage and not have to recompute everything, including those areas far removed from the place of change. Many VLSI CAD tools in use today use this fact in changes layouts to insert or remove parts of the design and simulators such as ASTAP tradeoff initial computation time for faster response calculation of small changes in the initial circuit and conditions.

What needs investigation in this area is better methods of handling coupling and better methodologies in approaching a complex system that allows for incremental changes in the design without requiring the recalculation of every parameter involved. Some use of the hierarchy of a layout should be incorporated; hierarchy is exploited in the design of the circuits themselves so why not use it in the analysis of the MTL problem?

Coupling of microstrip transmission lines; what's been accomplished ?

To understand what still needs to be done, one must examine what has already been accomplished. How well do some of these methods work? How do they work and is there a better way?

The first step is to decide what's important and what isn't. Ideally, we'd like to handle the coupling between all the lines in a system, regardless of layout and physical distance between lines. In practice with hundreds if not thousands of lines we need to ask; is this REALLY necessary? Do we indeed need to know the effects of a line on one side of a system to one on the other side? Common sense must also be incorporated into the design of such a simulator but we must be careful that "common sense" doesn't get us into trouble; common sense tells us that the sun revolves around the earth, doesn't it?

Obviously lines that are immediately adjacent to each other are going to have the most problem with coupling and perhaps that is where we should look first. The worst case condition is when two lines run parallel for long runs, allowing the highest degree of crosstalk between them. Lines crossing at right angles have less crosstalk and lines that are far removed from the line in question have (probably) the least effect. Many of the theoretical studies done have assumed that the coupled lines run parallel to each other and some assume further that they have the same dimensions; width, length, etc. For strictly theoretical studies this could be considered a valid set of assumptions but for the practical circuit designer creating electronic systems for a living, these are extremely limiting rules. Designing a PCB, for example, with no bends, no vias and no crossing lines would be difficult at best... especially if the lines must be in the same plane!

Several papers have been published by Vijai Tripathi and his associates at Oregon State University modeling coupling and crosstalk in microstrip transmission line systems. Most of their work involves generating lumped element models of one sort or another to

be used in SPICE or a similar circuit analysis tool. In one such paper (Tri 87) the case of multilevel interconnections as commonly found in integrated circuits and PCBs is analyzed for their time domain characteristics. Their method involves the straightforward solution of the general transmission line equations:

$$\partial \frac{[v]}{\partial z} = -[R][i] - [L]\partial \frac{[i]}{\partial t}$$

$$\partial \frac{[i]}{\partial z} = -[G][v] - [C]\partial \frac{[v]}{\partial t}$$

The result is used in a SPICE model utilizing voltage controlled voltage sources and voltage controlled current sources together with linear uncoupled transmission line models to model the system of interconnects. An example of this for N parallel transmission lines is shown in Figure 1.

A 1985 paper (Tri 85) concerns much the same things as the previous paper but provides a better background of the solution of the transmission line equations and the construction of the model used in SPICE.

Another approach to the problem is found in a paper by F.Y Chang (Cha 70) where he introduces the use of "congruence" transformers to model the effects of cross talk. Actually this is an intermediate step since the system of transmission lines is ultimately modeled by a resistive network and voltage controlled voltage sources... the congruence transformers help to visualize and set up the problem. This transformer is shown in Figure 2; the right hand side or primary side represents the voltage and current (e and J)

found at the input terminals to the system. The left hand or secondary side represents the applied voltages and currents (V and I) to the system of transmission lines. The turns ratios for the transformer are contained in the matrix [X] and they represent the amount of coupling from each line to each of the other lines.

It is this turns ratio matrix, [X], that's the key to analyzing the coupling problem and unfortunately it is probably the least talked about in the paper. To understand this paper better and in particular the calculations of the turns ratio matrix, two other papers are almost required reading (Wee 70) and (Gre 65). Furthermore, a key assumption that structure supports TEM mode propagation is made, allowing the inductance matrix to be calculated directly from the capacitance matrix. If we assume TEM propagation, then the velocity of propagation is related to the inductance and capacitance per unit length by; $v_0 = \sqrt{LC}$ and thus $L = 1/v_0^2 \; C^{-1}$.

A third step is necessary to find the transient response of this system, requiring the creation of a resistive network representation of the n-conductors. After additional manipulation, the network indicated in Figure 3 is arrived at which contains only voltages controlled sources and resistors. The terminal voltages and currents are immediately available from outside the network.

This is an interesting method but it too involves considerable computation especially in calculating the capacitance matrix, inverting it to find the inductance matrix, and calculating the turns ratio matrix. Any changes to the topology of the circuit would require recalculation of all these matrices and it is doubtful that this method would have

sufficient speed to be a credible tool for designers of complex interconnect systems.

A follow-up paper to Tripathi's work on SPICE models for coupled transmission line systems is an article by Romeo and Santomauro (Rom 87) in which they derive a simplified model from the general model purposed by Tripathi. In doing so, they've had to make several simplifying assumptions and restricted the type of transmission line systems that can be modeled with their method. The first of the assumptions, like many others, assumes strict TEM mode wave propagation. Thus they solve the telegraphists equations relating voltages and currents as functions of position and time. Their method consists of calculating an equivalent network for the system of transmission lines that consists of n uncoupled lines and a transformation network at the beginning and end of the lines containing dependent sources which model the coupling effects.

Their algorithm (in outline) is:

1. Compute eigenvalues of [L] and [C] given the number of lines, n.
2. Compute [M], the matrix of right eigenvectors of [L] & [C]
3. Derive the "control law" for each dependent source in the transformation network.
4. Calculate the characteristic impedance, Z and the time delay per unit length, W for each line.

In calculating these various parameters and control laws, two key assumptions have been made on the physical layout and the effects of near/far coupling on a given line. In order to keep [L] and [C] tridiagonal (for calculation simplicity) it is assumed that each line is coupled only to those immediately adjacent to it. Any other lines more than one removed from the line in question are assumed therefore to have no effect on the

line. The lines are also assumed to be of identical length and equally spaced. The reasoning behind this assumption underlies the fact that worst case coupling and crosstalk occurs for long runs of parallel lines such as might be seen in a data bus system. By making these assumptions, they have reduced [L] and [C] to tridiagonal, symmetric Toeplitz matrices which have nice computational properties.

Using the assumption of TEM mode propagation and uniform, equal length lines, Das and Prasad (Das 84) have performed an analysis of the coupled line problem using a Green's function formulation. The coupling capacitance and inductance on a per unit length basis were calculated for the configuration shown in Figure 4. Closed form equations were derived and the coupling coefficients due to the line capacitance/inductance can be found directly for the case of:

1. Symmetric striplines; coplanar conductors appearing midway between ground planes.
2. Single layer microstrip structures
3. Offset parallel striplines; parallel lines of equal width on different layers.

The chief disadvantage to their method is that only two lines at a time can be handled. The main advantage is the use of closed form equations requiring no iterations to calculate the coupling coefficients. If the "immediately adjacent" assumption is used (i.e. only lines right next to each other are important) than this technique could prove very useful and potentially faster than other, more general methods.

A unique approach to the coupled line problem that is more realistic than most is one used by Razban in a paper concerned with the transient analysis of coupled lines. The

parallel lines are grouped into bundles called "tubes" where all the lines in a tube are equally spaced and of the same length. The coupling between lines in a tube are analyzed using the method of characteristics (Bra 76) and each tube is modeled as an equivalent circuit consisting of resistors and dependent voltage sources (see Figure 5). When tubes of different orders meet (i.e. tubes with different numbers of coupled lines), the interface is handled as shown in Figure 5. The resistors and sources are found as a result of applying "dilatation" and "shrinkage" operators. The figures show the case where an interface between two sets of tubes, p and q ($N_p > N_q$) is modeled. The dilatation operator is used to handle signals running from Tp to Tq and the shrinkage operator handles the signals going the other way.

Providing both a theoretical and experimental approach to the problem, Seki and Hasegawa have written a paper concerned with coupling/crosstalk in high speed VLSI circuits (Sek 84). Even though their main concern was in IC the same theories and conclusions should hold for PCB microstrip transmission lines as they made no scale assumptions that would limit the work to micron size structures. What is unique about their method is the use of a periodic, infinite boundary condition. In other words, they assume that the n-conductor MTL circuit under study is repeated infinitely many times on either side. They show later in the paper that this is not such a bad assumption and the solutions give very good first order results and lend themselves to transient analysis well. The other assumption made concerns the method of excitation of the n-lines; they assume that the phase angle difference for voltage and current from one line to the next is a constant.

The solution for voltage and current as a function of distance for any of the lines allows the pulse response to be calculated directly by performing a numerical inverse Laplace transform of these two equations.

What makes this paper all the more interesting is their conclusions that they draw as a result of several "experiments" run using this method for various circuit topologies. In particular they calculated the pulse response of a terminated line adjacent to a driven line for cases where all other lines were terminated or left open. Their results match this authors experience with such cases which show much higher crosstalk amplitude for the unterminated case versus the terminated lines.

The last paper that will be discussed here concerning coupling in MTL systems is one by Yang, Kong and Gu which is a perturbational analysis of nonuniformly coupled lines. In their case, only pairs of lines were studied but the lines could have nonuniform coupling coefficients as a function of distance down the lines. Instead of performing a numerical integration of the coupled transmission line equations, the authors elected to perform a codirectional-contradirectional transformation for the decomposition of coupled-mode equations. The method of characteristics is then used to find a set of differential equations along a characteristic curve. Rather than integrate this equation numerically (tedious since the d.e.'s are coupled) a perturbational series was introduced. The result of all this is a set of zero order equations that have closed form solutions. Furthermore, all higher mode terms are generated from a single integration over the previous terms. The key advantage then is a set of first order, closed form

equations governing the coupling phenomenon between the lines. The other significant advantage to their approach that the higher order terms can be directly interpreted as reflections along the lines.

As we've seen, there are numerous papers discussing the problem of coupling in MTL systems but virtually none concerned with the overall MTL problem. As we noted earlier, most of these papers are good for largely academic studies but have so many limitations and assumptions made that they are not real useful for studying a complex MTL systems such as found in a PCB. As tools for analyzing specific situations and for creating a database to be used by a simulator they are important, but not for handling a complex MTL problem by themselves.

Microstrip Discontinuities

When operating at high frequencies (synonymous with fast rise/fall times in digital circuits) the effect of simple bends or 'T' junctions become increasingly important. Bends (especially right angle bends), vias, tees, step-widths and other shapes all have the unpleasant effect of adding unwanted inductances and capacitances to the circuit. What appears to be a simple interconnection on a schematic can become a meandering line on a PCB or IC. The extra components "hidden" among the bends and vias can have a significant role in the undoing of an otherwise sound design. By themselves a few discontinuities may have little effect on the transmitted signal but in conjunction with all the other pitfalls of transmission lines, the integrity of the system can be dis-

rupted.

There are two philosophies when it comes to evaluating the effects of discontinuities (particularly microstrip discontinuities) on circuit performance. One can calculate the parasitic elements of a discontinuity and use a lumped parameter model in a SPICE type simulator. This by nature is a sort of "after the fact" approach since one needs to know the physical layout to calculate the equivalent circuit. If some form of autorouting of the interconnect lines is used, then the router must be run and its output passed to a parasitic element extractor to obtain the SPICE models. On the other hand, knowledge of the relative discontinuity of various features can be used before hand to guide the routing process. A critical subcircuit could be simulated in part on SPICE and the sensitivity to various discontinuities could be examined. These results could then be used by the router, along with discontinuity design rules, to select appropriate routing.... should four bends or two vias be used?

There are pros and cons to both sides but in one form or another discontinuities msut eventually be taken into account. As digital circuit speeds increase to ever faster rates, the adverse effects of such simple structures as bends and vias will draw closer to the surface, waiting for the unsuspecting designer to stumble across them.

Several papers have been written in recent years concerning various aspects of microstrip transmission line discontinuities, ranging from theoretical calculations to experimental techniques and measurements. As many of them are similar in their results,

only a representative few will be discussed here.

In almost all circuit layouts, most of the conducting lines lie in roughly the same plane and thus the parallel plate approximations for capacitance becomes very inaccurate and of little use. In response to this need for better accuracy for complex geometries, circuit designers have turned to electromagnetics and microwave circuit solutions. As in many such calculations, the tradeoff between accuracy and speed is still here, requiring careful considerations of the requirements. The various methods range from integral-equation numerical solutions of three dimensional conductor problems (Rue 73) to statistical modeling of parasitics (Mic 84).

One of the more interesting attempts at simplifying the magnitude of the required calculations is the use of statistical models for various interconnect geometries and shapes (Mic 84). This method uses conventional two-dimensional capacitance simulators to determine fringe and coupling capacitances under a variety of conditions and using these results along with statistical regression techniques to generate corresponding capacitance equations. For these initial simulations, the capacitance structure shown in Figure 6 were used where W, T and S are the width, thickness and separation respectively of the conductors and H is the height above the substrate. C_{pp} is the well worn parallel plate capacitance given by: $C_{pp} = \epsilon_0 \epsilon_r \frac{W}{H}$.

The conductor to ground capacitance is given by :

$$C_0 = C_{pp} + 2C_{f0}$$

$$C_1 = C_{pp} + C_{f0} + C_{f1}$$

$$C_2 = C_{pp} + 2C_{f1}$$

for the case of zero, one or two adjacent conductors respectively.

The total capacitance for one or two adjacent conductors is:

$$C_{tot1} = C_{pp} + C_{f0} + C_{f1} + C_c$$

$$C_{tot2} = C_{pp} + 2C_{f1} + 2C_c$$

Numerous simulations were run for varying widths, separation distances and thicknesses and the results used as the data base for a regression analysis. Models were found for C_{f0}, C_{f1} and C_c using multiple linear regression. The predictor variables included the height, width, thickness and separation of the conductors. Details of the method and the reasoning behind it can be found in (Mic 84) but the results for several values of W, T, H and S can be found in Table 1. The final equations for C_{f0}, C_{f1} and C_c (actually the natural log of these variables) are shown below.

$$Ln(C_{f0}) = 3.805 - 0.299Ln(H) + 0.0257(W) + 0.0524(T) - 0.00101W^2 - 0.0469(T\ x\ Ln(H)) - 0.0181(Ln(H))^2$$

$$Ln(C_{f1}) = 3.952 - 2.20(1/S) - 0.394Ln(H) - 0.533(Ln(H)/S) + 0.932(1/S^2) + 0.030W - 0.0187(W\ x\ Ln(H)/S) + 0.0846T - 0.00125W^2 - 0.00907(W/S) - .0776(T/S$$

$$
\begin{aligned}
Ln(Cc) = {} & 4.343 - 0.651Ln(S) + 0.193Ln(H) + 0.487Ln(T) - 0.879(1/W) \\
& - 0.212(Ln(S))^2 - 0.167(Ln(T)xLn(H)xLn(S)) \\
& + 0.104(Ln(H)xLn(S)/W) - 0.144(Ln(S)/W) + 0.0619(Ln(H)xLn(S)) \\
& + 0.470(1/W^2) - 0.144(Ln(T)xLn(S)/W) + 0.232(Ln(T)/W) \\
& + 0.111(Ln(T))^2 + 0.470(Ln(T)\ x\ Ln(H)\ x\ Ln(S)/W)
\end{aligned}
$$

Though lengthy, these equations provide a sort of empirical, closed form solution to some of the more common capacitance problems associated with integrated circuits. As with any regression model, one should never try to extrapolate these results beyond the limits used for W, T, H and S in the initial 2D simulations.

Another simple, closed form procedure for calculating two and three dimensional capacitances of interconnect lines has been proposed by T. Sakurai and K. Tamaru of Toshiba Corporation (Sak 83). The procedure is similar to the statistical modeling used by Michael in that detailed methods were used to calculate the "exact" capacitance by finite element methods and the correct capacitance was assumed to have been found when a doubling of the number of sub-areas produced less than a 1% change in the computed capacitance. From these calculations, the following empirical formulas were obtained for the cases of one, two and three conductors.

$$C_1 = \epsilon_{ox}[1.15(W/H) + 2.80(T/H)^{0.222}]$$

$$C_2 = C_1 + \epsilon_{ox}[.03(W/H) + .83(T/H) - .7(T/H)^{0.222}](S/H)^{-1.34}$$

$$C_3 = C_1 + 2\epsilon_{ox}[.03(W/H) + .83(T/H) - .7(T/H)^{0.222}](S/H)^{-1.34}$$

The wiring capacitance by various methods, including Sakurai's are shown in Figure 7.

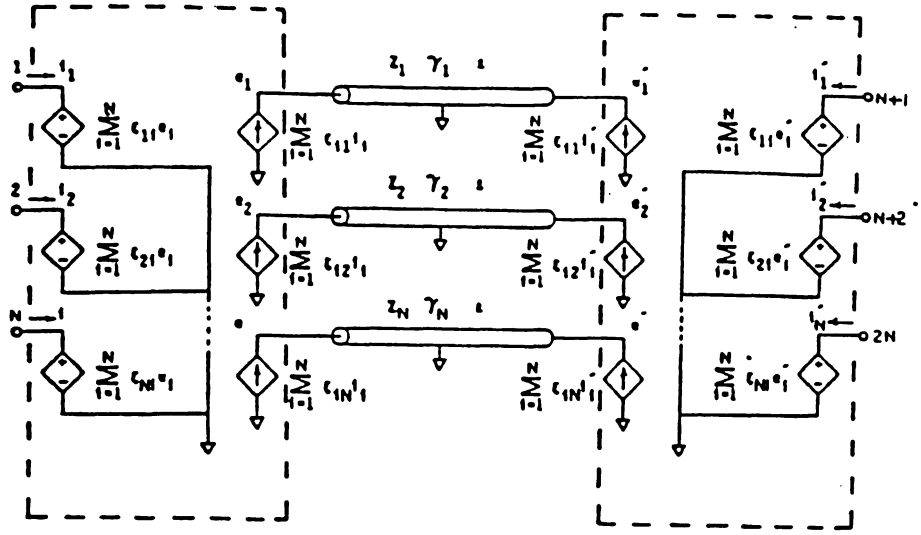
The problem of calculating inductive discontinuities is not as straightforward as finding capacitive elements and this is reflected in the small number of papers written on this subjects as compared to capacitive effects. Thomson and Gopinath have tried to fill this gap in a paper that is strictly concerned with the inductive nature of common microstripline structures (Tho 75). In the paper, the authors use a variation on the finite element method that minimizes computer memory requirements and some of the inherent inaccuracies of finite element methods. Their results are fairly accurate up to about 5 Ghz but are limited to strictly manhattan shapes (horizontal and vertical rectangles). Figure 8 shows the models used in their investigations and the normalized inductances for right angle bends and step-width junctions at low frequencies

In his paper on equivalent circuits for microstrip discontinuities, Easter provides experimental evidence to support his models of bends, T-junctions and crossover connections (Eas 75). His graphs and tables provide data for the models used to represent various structural shapes. Most of the models come in two forms; lengths of uniform transmission line with appropriate lumped susceptances and simple LC models.

Conclusion

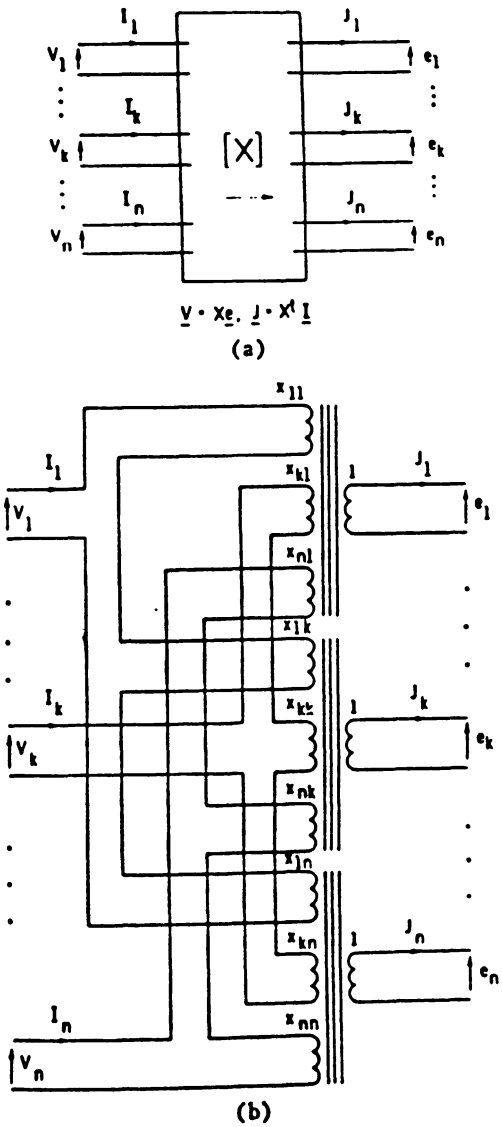
Though there have been many papers published and much research done in the transmission line area and particularly microstrip lines, much of the work is highly theoretical and deals with specialized topics. To date, none of the methods deals with the complex problem of an entire MTL system. Many of the methods discussed do, however, provide useful tools in analyzing specific cases and provides a means of developing a

database for use in a more sophisticated simulator. In general, the methods that develop SPICE models are either too simplistic or too slow for large scale studies. Another approach is needed that moves away from this dependence on SPICE and "traditional" approaches.

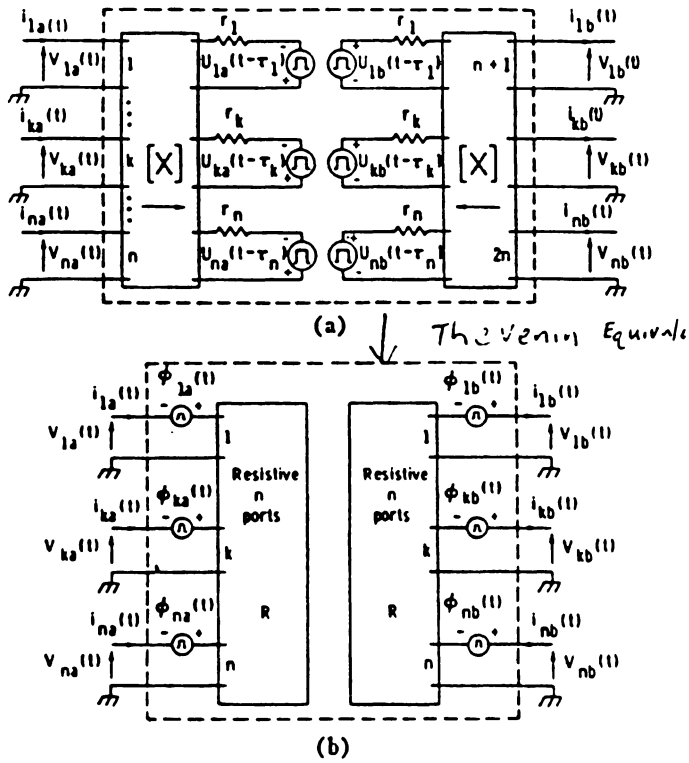


The parallel coupled interconnection model. All ξ's, Z's, and γ's are in general complex.

Figure 1 (Tri 87)



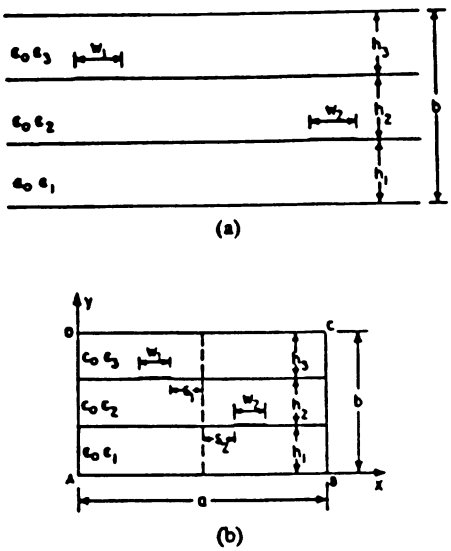
2. (a) Schematic notation. (b) Physical structure of congruence transformer. Arrow in (a) always points to primaries of congruence transformer.



(a) Equivalent circuit of an n-conductor system. (b) Equivalent resistive network of an n-conductor system.

Figure 2 (Cha 70)

Figure 3 (Cha 70)



(a) Offset parallel coupled strips between parallel planes filled with layered dielectric. (b) Configuration of (a) terminated by electric walls at finite distances from the strip.

Figure 4 (Das 84)

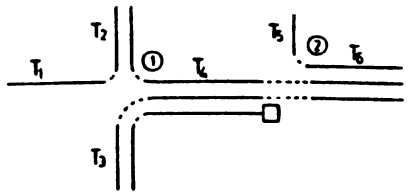


Fig. 1. An example of discontinuous coupled lines.

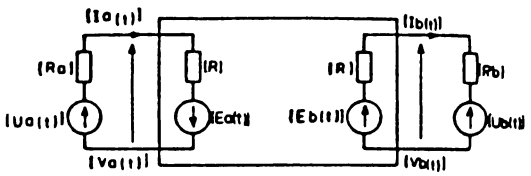
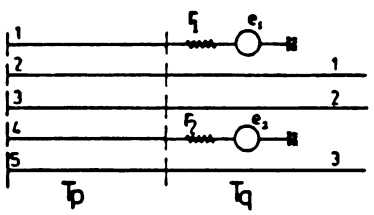


Fig. 2. Equivalent circuit of a tube.



An interface between two tubes of different orders ($np > nq$).

Figure 5 (Raz 87)

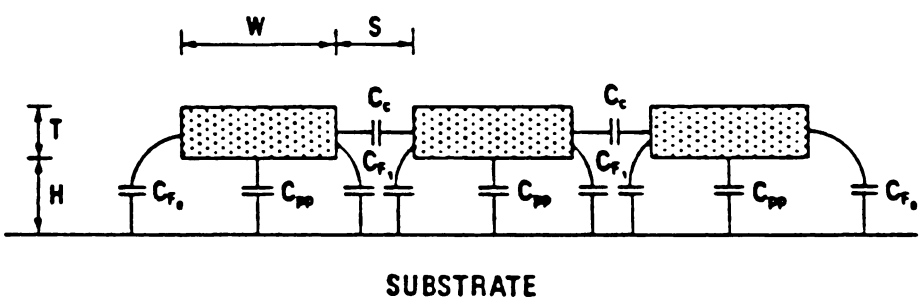
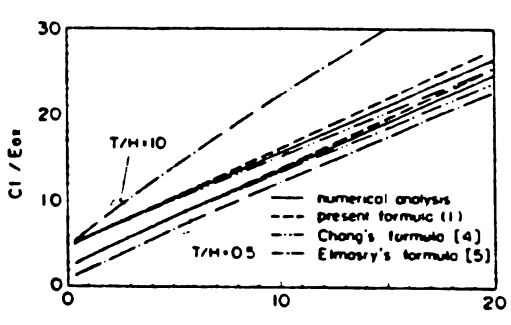
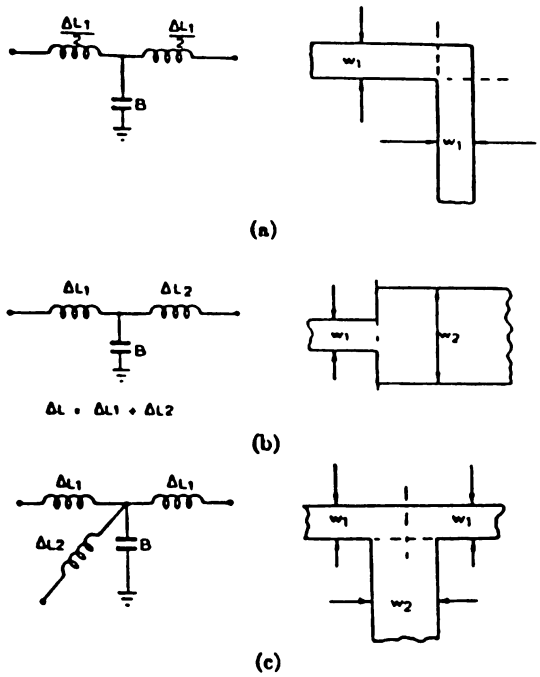


Figure 6

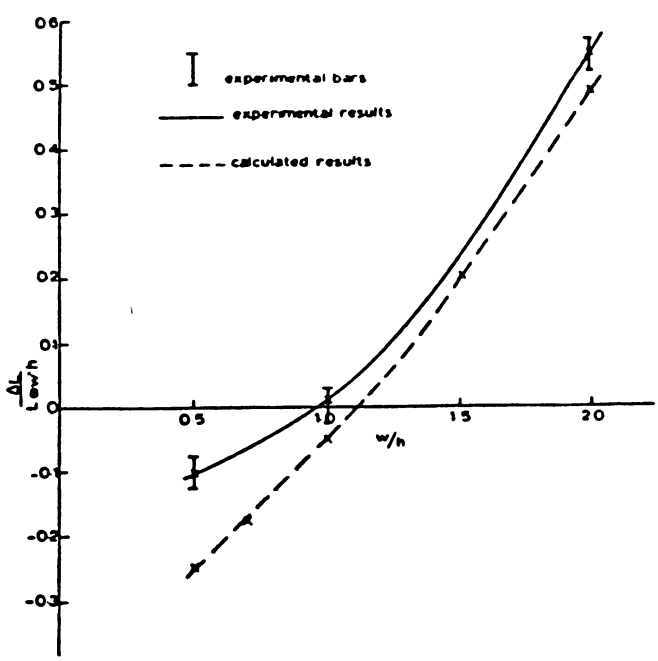


Calculated wiring capacitance by various formulas.

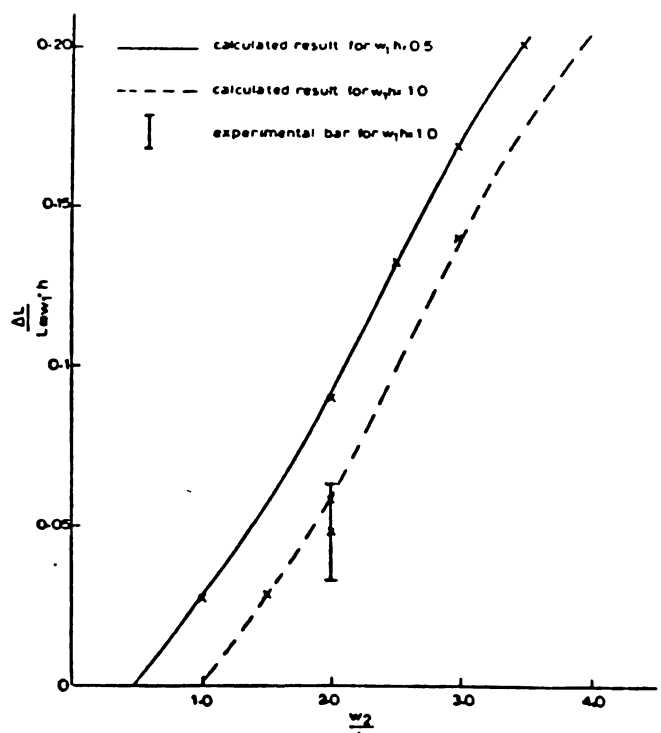
Figure 7 (Sak 83)



Equivalent circuits of three microstrip discontinuities with the positions of the reference planes. (a) Symmetric right-angled bend, strip w_1 wide. (b) Step-width change from strip w_1 width to w_2 width. (c) Symmetric T junction, straight arms w_1 wide; vertical, w_2 wide. Substrate thickness kept constant at $h = 1$. The inductances given in these circuits are plotted in Figs. 3–6.



Normalized inductance of a symmetric right-angled bend $\Delta L/(L_{w=1}h)$ plotted for different w/h ratios. Comparison is with experimental results from Easter [9].



Normalized inductance for step-width change $\Delta L/(L_{w=1}h)$ for $w_1/h = 0.5$ and 1.0 for different w_2/h. Experimental results for one value are marked.

Figure 8 (Tho 75)

Conditions				Regression			2-D Simulator		
W	T	H	S	C_C	C_{F1}	C_{TOT}	C_C	C_{F1}	C_{TOT}
1.0	0.8	0.8	1.0	40.56	18.41	161.12	44.24	18.54	168.75
1.0	0.8	0.8	9.0	2.51	49.82	147.85	2.33	47.69	143.24
5.0	1.6	1.6	1.0	93.05	10.78	315.63	91.64	10.30	311.87
5.0	1.6	1.6	5.0	17.67	34.56	212.45	18.62	32.81	210.84
5.0	1.6	1.6	9.0	7.33	42.35	207.33	7.48	45.35	213.65
13.0	1.6	1.6	13.0	4.44	48.80	387.19	3.44	49.33	386.29

Conditions			Regression		2-D Simulator	
W	T	H	C_{FO}	C_{TOT}	C_{FO}	C_{TOT}
1.0	0.8	0.8	51.30	145.79	51.70	146.59
1.0	1.6	0.8	53.95	151.08	55.01	153.21
1.0	1.6	1.6	40.35	102.28	43.42	108.44
5.0	1.6	1.6	43.64	195.24	43.93	195.84
13.0	1.6	1.6	46.35	373.38	47.28	375.31

Table 1 (Mic 84)

Appendix B. CAPNET — a transmission line simulator simulator

Abstract

A data structure and algorithm for representing, and solving complex transmission line systems is presented. The method is based on representing the network as a recursive tree structure and solving for the voltage, current, and impedance at each node using recursive programming techniques. First, all frequency dependent parameters within the tree structure are updated, then in a post-order traversing of the tree, the impedances at each node are computed followed by a pre-order traversing of the tree to compute node voltages and currents. The technique is useful in networks with many branches and mixed transmission line characteristics. Applications include the modeling and simulation of pulse propagation in distribution line carrier networks, local area networks, and the digital subscriber loop with bridged taps.

I. Introduction

In many digital communication systems, complex transmission line networks are encountered which contain mixed transmission lines with different characteristics and many branches. Examples include the distribution line carrier network, the digital subscriber loop with bridge taps, and certain local area network configurations. The transmission line network usually introduces severe magnitude and phase distortion resulting in the degradation of bit error rate performance in digital transmission systems.

In this paper we present a program for computer modeling and simulation of complex transmission line networks. The network is represented in the computer by a recursive binary tree data structure. Using recursive programming techniques, the node voltage current, and impedance at each node within the tree structure is computed. In this manner the frequency response of the network, from the source node to the receiving node is computed. The impulse response or the pulse response of the network is then calculated from the frequency response using Fast Fourier Transforms.

Computer programs for modeling transmission line networks have been written using ABCD parameters [1]. In this paper a new technique in which the frequency response at all nodes within the network are obtained simultaneously is presented. The technique is also suitable for the computer aided design and modeling of digital communication systems, with complex transmission line networks. A CAD tool is currently being developed for this purpose and a technical report describing the

graphical capture, editing, and analysis of networks is being prepared. The interactive graphics is based on the graphical kernel system (GKS) [4].

II. Transmission Line Networks

Consider the basic problem of simulating pulse transmission through a loaded transmission line. Assuming that the pulse is bandlimited with a cutoff frequency of f_c, we can obtain the pulse response by computing the inverse FFT of the complex multiplication of the frequency response of the pulse and the transmission line network. Therefore, as a first step in calculating the frequency response of the network, we analyze the network response to a single sinusoid of frequency f_0. Consider the loaded transmission line connected to the generator E_g through a source impedance Z_s as shown in Fig. 1 [2].

$$E(x) = \frac{E_g Z_0}{Z_0 + Z_s} e^{-\gamma x} \frac{1 + \Gamma_r e^{-2\gamma(l-x)}}{1 - \Gamma_r \Gamma s e^{-2\gamma l}} \tag{1}$$

$$I(x) = \frac{E_g}{Z_0 + Z_s} e^{-\gamma x} \frac{1 - \Gamma_r e^{-2\gamma(l-x)}}{1 - \Gamma_r \Gamma_s e^{-2\gamma l}} \tag{2}$$

In the above expressions

$$\gamma = \sqrt{(R + jwL)(G + jwC)} \tag{3}$$

is the propagation constant and

$$Z_0 = \sqrt{\frac{(R + jwL)}{(G + jwC)}} \tag{4}$$

is the characteristic impedance of the transmission line. The expressions for the source and load reflection coefficients are,

$$\Gamma_s = \frac{Z_s - Z_0}{Z_s + Z_0} \tag{5}$$

$$\Gamma_r = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{6}$$

The expression for $E(x)$ includes the superposition of all waves reflecting from the source and load mismatches. This can be seen by a Taylor series expansion of (1)

$$E(x) = \frac{E_g Z_0}{Z_0 + Z_s} \Big[e^{-\gamma x} + \Gamma_r e^{-\gamma(l-x)} \tag{7}$$

$$+ \Gamma_r \Gamma_s e^{-\gamma(2l+x)} + \Gamma_r^2 \Gamma_s e^{-\gamma(3l-x)} + \Gamma_r^2 \Gamma_s^2 e^{-\gamma(3l+x)} + \cdots \Big]$$

To obtain the shape of the pulse at the load we evaluate $E(l)$ at frequencies from $f=0$ to $f=fc$ in discrete steps where fc is the cutoff frequency of the bandlimited pulse. The number of points must be a power of 2 such that the inverse FFT may be used to obtain the sampled pulse response at the load.

Consider now the case where the boundary voltage and current are known on a section of transmission line. See Fig. 2.

Evaluate $E(0)$ in (4) and then compute

$$\frac{E(x)}{E(0)} = e^{-\gamma x} \frac{1 + \Gamma_r e^{-2\gamma(l-x)}}{1 + \Gamma_r e^{-2\gamma l}} \tag{8}$$

Also

$$\frac{I(x)}{I(0)} = e^{-\gamma x} \frac{1 - \Gamma_r e^{-2\gamma(l-x)}}{1 - \Gamma_r e^{-2\gamma l}} \tag{9}$$

Thus using (8) and (9) the voltage and current can be evaluated at any point on the transmission line given the boundary voltage and current.

With the above preliminaries, we will examine the simple network in Fig. 3 and present a methodology for its solution. In Fig. 3, the nodes have been labeled $n1$ through $n5$. To solve this network, that is to obtain the voltage and current at each node and at any location within the network, consider equation (4). This equation suggest that if the impedance at node $n1$ was known then the voltage and current at node $n1$ can be calculated from the generator and source impedance. Thus the first step is to obtain the impedance at $n1$. This impedance is seen to consist of the parallel combination of the impedance looking into $n5$ and $n2$ from $n1$. These impedances can be obtained by noting that the impedance at a distance l away from a loaded transmission line is given by (Fig. 4),

$$Z_{in} = \frac{1 + \Gamma_r \, e^{-2\gamma l}}{1 - \Gamma_r \, e^{-2\gamma l}} \, Z_0 \tag{10}$$

Thus, the first step is to calculate the impedances looking into $n3$ and $n4$ from $n2$. The parallel combination forms the impedance at $n2$. The impedance at $n1$ is thus calculated by the parallel combination of the impedances looking into $n2$ and $n5$

Thus, the following methodology is suggested for solving the network. In the first pass, starting from the three loaded end nodes, the impedances are calculated and the parallel combination of these impedances at the parent node forms the parent node impedance. Working backward in this manner, the impedance at the root note ($n1$ in the example) is calculated. Using (4) the voltage and current at the root node ($n1$) is calculated. Using (8) and (9) and the boundary voltages and currents, calculated at the parent node, the voltage and current at each node in the network can be calculated. Note that the current at each node is split into two

currents flowing into each node.

In the next section, a computer program will be introduced which uses recursion and recursive data structures available in C to solve complex transmission line networks.

III. Recursive Programming and Data Structures

To introduce the algorithm for solving a complex transmission line network, we first consider the case where the network is limited to the binary tree structure shown in Fig. 5. In the figure, the generator is connected to the root of the tree through a source impedance Z_s. The tree consists of nodes which are either parents or leaves. A leaf is a node which is terminated on a load. For example, $n4$, $n5$, $n7$, $n9$, $n10$, $n12$, $n16$, $n14$ and $n15$. Parent nodes have two branches. A left branch and a right branch. Nodes $n1$, $n2$, $n3$, $n6$, $n8$, $n11$, $n13$ and $n17$ are parent nodes. In general each branch represents a transmission line with different characteristics and lengths. Each section of transmission line is associated with the node on which it terminates. Thus the section of transmission line from the generator to the root node $n1$ is described in the data structure pointed to by $n1$. This concept is described below.

Each node has an associated data structure which occupies memory locations. A pointer can be defined which points to the data structure in memory. As nodes are added to the tree, memory is dynamically allocated for the data structure and a pointer is defined. Thus, for the nodes of the network in Fig. 5 the following data

structure can be defined in C.

```
struct      node
  {
     struct       node *left;
     struct       node *right;
     struct       node *parent;
     char         name [16];
     float length
     float        r,l,c,g;
     float length
     complex      Z - Left;
     complex      Z - Right;
     complex            Z_L
     complex      node- voltage;
     complex      left - current;
     complex      right - current;
     complex      input_current;
     complex      Z_0;
     complex      gamma;
  }
```

Within the data structure definition are three pointers to data structures of the same type. Thus the data structure is recursive. Two pointers point to the left and right nodes while the third pointer points to the parent node. Three cases are immediately evident. If the node is a leaf then the left and right node pointers are NULL. Otherwise, they will point to the left and right child nodes attached to the node. If the node is the root node, then the pointer to the parent will be NULL.

The other data types within the structure represent data necessary to describe the node. These can be classified into two groups. One group defines the name of the node and the characteristics of the transmission line (e.g., r,l,c,g and Z_0 and γ). The other group represents data which are calculated and depend on the network. These include the voltage at the node, the current flowing into the right and left

nodes, and the impedances looking into the nodes.

It is very convenient to access data in a data structure using pointers to data structures. For example, to assign the variable Z the value of the characteristic impedance at the node pointed to by np we write,

$$Z = np \rightarrow Z_0$$

To access the characteristic impedance of the left child node of the node pointed to by np we write,

$$Z = np \rightarrow left \rightarrow Z_0 \; ;$$

At this stage, we will describe recursive functions which are used to compute the impedances, voltage and currents at each node. First we introduce two methods for traversing tree data structures.

Traversing Trees

There are general methods for traversing trees [3]. We will apply two of these methods to solve the tree network.

Postorder Listing

Postorder traversing of trees is illustrated in Fig. 6. This method is useful in the first pass needed prior to solving for the voltages and currents of the network. As pointed out earlier the impedances at each node must be computed. Thus in Fig. 6, the impedance at 3 is the parallel combination of the impedances of the loaded transmission lines 1 and 2. Similarly, the impedance at 6 is computed from 4

and 5. Once the impedance of 3 and 6 are computed, the impedance at 7 can be calculated and so on. Careful study of the figure will show that the numbering schemes corresponds to the order in which the impedance calculations must be carried out. This order of traversing the tree is termed postorder listing. The method is summarized below [3]:

(1) If a tree is composed of only a single node, the post order listing consists of just that single node

(2) If a tree consists of more than one node, the postorder listing consists of the postorder listing of each subtree, in left-to-right order, followed by the root.

Preorder Listing

This method is used in the calculation of the voltages and currents at each node once the impedances have been determined. Preorder listing is illustrated in Fig. 7. Thus, once the boundary voltage at node 1 is known, the voltage at node 2 can be computed (since the impedance at 2 is also known from the first postorder traversing in computing the impedance). From 2, the voltage at 3 and 6 can be computed and so on. The preorder listing method is summarized below [3]:

(1) If a tree is composed of a single node, the preorder listing consists of just that single node.

(2) If a tree consists of more than one node, the preorder listing consists of the root, followed by the preorder listing of each subtree is left-to-right order.

IV. Recursive Calculation of Network Impedances

The following C code presents a recursive function that calculates the impedance at each node using a post order traversing of the tree structure.

```
① Complex calculate_impedance (root)
  struct node * root;
  {
②         if ((root) → left == NULL) && (root → right == NULL)) {
      /* we are at a leaf */
      /* calculate impedance a distance root → length away from the load, root → Z_L */
            /* calculate reflection coefficient at load */
            rcl = calc_rcl (root→ Z_L, root → Z_0);
            return (line_impedance (rcl, root → Z_0, root → gamma, root → length);
      }
③     root → left_impedance = calculate - impedance (root → left);
      root → right_impedance = calculate - impedance (root → right);
      /* calculate parallel combination of left and right impedance */
④     root → Z_L = Z impedance _parallel (root → left_impedance, root → right_impedance);
      rcl = calc_rcl (root → Z_L, root → Z_0);
⑤     return ((line_impedance(rcl, root → Z_0 , root → gamma, root → length));
  }
```

Function explanation:

① The function argument, root, is a pointer to a node within the tree. The function returns the impedance <u>looking into a node,</u> and it is complex.

② The function checks to see if the node is a leaf. If it is, then it calculates and returns the impedance looking into the leaf node. This is one of the terminating condition of the recursive function. In other words, once a leaf node is reached, the function returns. The function line_impedance () calculates the impedance based on equation (10).

③ If the node is not a leaf, then the function recursively calls itself by passing the left branch node pointer. After returning the impedance looking into the left node, the fraction is recursively called to calculate the right branch node impedance.

④ Once the left and right impedances at the node are known, the parallel combination is computed. This produces the total node impedance.

⑤ At this point the impedance looking into the node is computed and returned by the function. This is also another terminating condition.

After this function is called, the impedances at all nodes are stored in the data structures pointed to by each node within the network.

V. Recursive Calculation of Node Currents and Voltage

After the function calculate_impedance (root) is called, each node contains the terminating or left and right branch impedance. The next recursive function, calculates the total current flowing into each node including the left and right branch currents. The function makes use of equation (9) to compute the node current $I(l)$ based on knowledge of the boundary current $I(0)$. The function performs a preorder traversing of the tree network.

```
① calc_current (root, i_input)
  struct node * root;
  complex i_input;
  {

②     complex line_current(); /* evaluates boundary current based on Eq. (9) */
```

```
*   complex calc_left_current ();
*   complex calc_right_current ();
        /* calculate the total node current based on  */
        /* equation (9) i_input is the boundary       */
*       /* current corresponding to I(0) in the       */
        /* equation                                   */
③       root → input_current = line_current (root, i_input, root → Z_L , root → length);
④       If (root → left ! = (struct node *) NULL  {
                root → left_current = calc_left_current (root → left_impedance,
                                root → right_impedance, root → input_current);
                calc_current (root → left, root → left_current);
    }

⑤       if (root → right ! = (struct node  *) NULL) {
                root → right_current = calc_right_current (root → left_impulse,
                                root → right_impulse, root  input_current);
                calc_current (root →, root_right → right_current);
    }
}
```

Explanation:

① The root argument is a pointer to the current node, i_input is the complex boundary current corresponding to $I(0)$ in equation (9). That is, it is the total current flowing into the transmission line associated with the node.

② The function line_current () computes the current $I(l)$ in equation (9). The functions calc_left_current () and calc_right_current (), calculate the currents flowing into the left and right branches.

③ This operation calculates $I(l)$ and stores it in the data structure of the node. Note that since this is a recursive procedure for traversing all the nodes of the tree, this operation will occur for each node.

④ If the left branch is not a termination, then recursively call the function by

moving to the node attached to the current node, root. However, first compute the boundary current $I(0)$ flowing into the node, in this case left_current.

⑤ The same as in step ⑤ but for the right branch.

After this procedure is called, all node data structures will contain the total current, and the left and right currents. Note that when the procedure is first called, i_input is the total current flowing from the generator into the network. It is computed based on equation (2).

VI. Computer Simulation Results

In the previous sections, we showed how the voltage and currents at each node can be computed. In general, for bandlimited digital transmission over a complex transmission line network, the frequency response of the network and the input impedance looking into the network over a frequency range is required to completely model the network. Thus, the response of the network to a generator with arbitrary source impedance may be computed.

Fig. 8 shows a test network consisting of twisted pair transmission lines with different characteristics. We are interested in obtaining the impulse response from the sending station to the receiving station. The computer program computed the voltage at the receiving station at discrete frequency intervals up to 50 MHz. The program then performed a 1024 point inverse FFT to produce the impulse response at a sampling rate of f_s = 100 MHz. Due to skin effect at higher frequencies the frequency response of the network is effectively bandlimited. The computed impulse

response is displayed in Fig. 9.

Notice the pure delay corresponding to traveling a distance of 180 m from the sending station to the receiving station. Also, immediately following the main impulse, a large second impulse appears with a delay corresponding to traveling a distance of 40 m. This second impulse is due to reflections from the two open circuit transmission lines of 20 m length each (#26 gauge) which constructively add and propagate to the receiving station.

In Fig. 10, the voltage distribution along the 50 m cable connecting the receiving station to the main #19 gauge transmission line is shown at a frequency of 1 MHz.

VII. Conclusions

In this technical report, a computer program is described which solves complex networks of transmission lines. A tree data structure was introduced for representing the network in the computer. Recursive procedures were presented for traversing the tree data structure to compute the impedance, voltage and current at each node within the network. Simulation results were then presented in which the impulse response of a test network composed of transmission lines of various characteristics and lengths was computed. The impulse response was then related to the network in terms of the predicted reflections and delays.

The program efficiently solves complex transmission line networks and has applications in the area of Computer Aided Design (CAD) of digital communication

networks. Specific applications include the Distribution Line Carrier Network, the digital subscriber loop, and Local Area Networks.

References

[1] D.G. Messerschmitt, "Transmission line modeling program written in C," *IEEE Journal on Selected Areas in Communications*, Vol. SAC-2, pp. 148-153, January 1984.

[2] Dworsky, Lawrence N, *Modern Transmission Line Theory and Applications*, John Wiley and Sons, Inc., 1979.

[3] Gerald E. Sobelman, David E. Krekelberg, *Advanced C Techniques and Applications*, Que Corporation, Indianapolis, Indiana,, 1985.

[4] F.R.A. Hopgood, D.A. Dull, J.R. Gallor, D.C. Sutcliffe, *Introduction to GKS*, Second Edition, Academic Press, 1986.

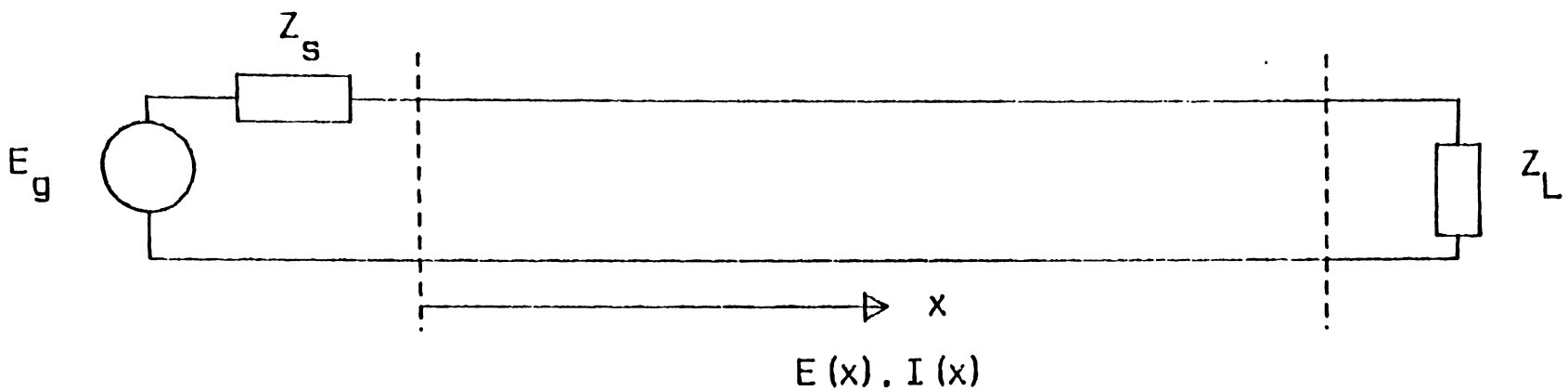


Fig. 1. Generator connected to loaded transmission line

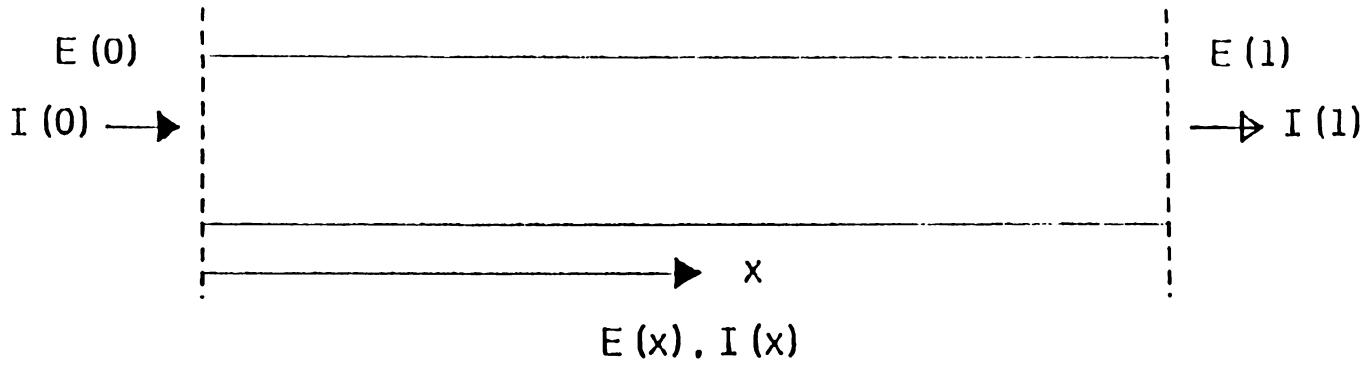


Fig. 2. Section of transmission line with boundary voltages and currents.

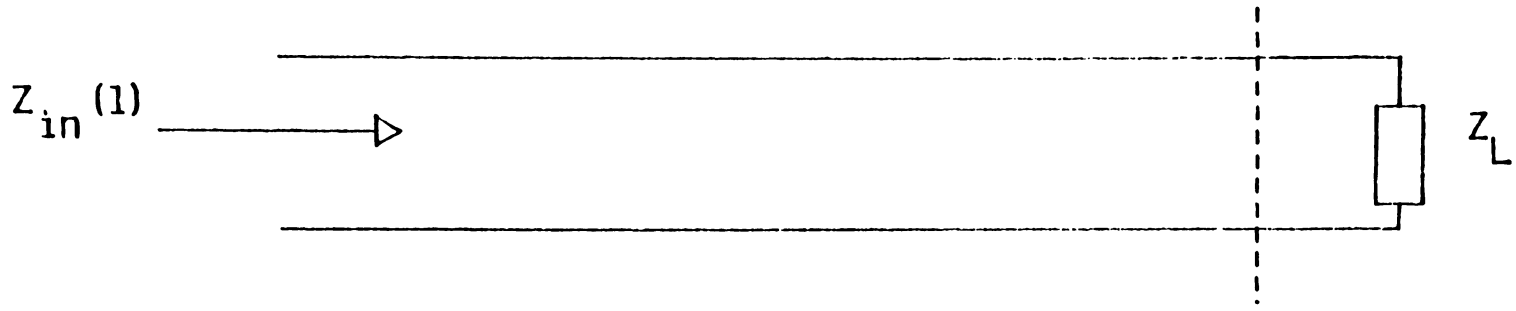


Fig. 3. Input impedance of loaded transmission line of length l

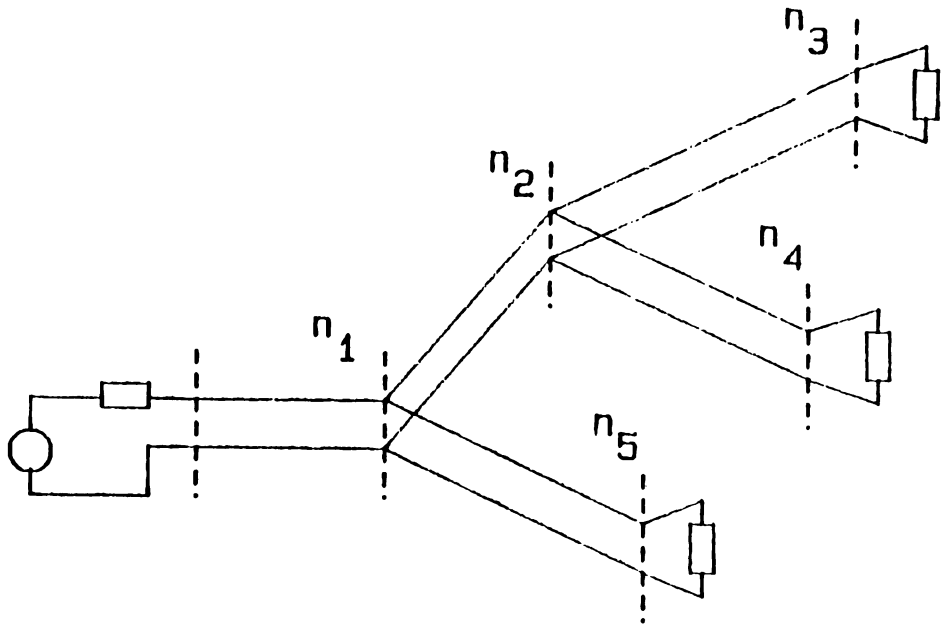


Fig. 4. Example transmission line network

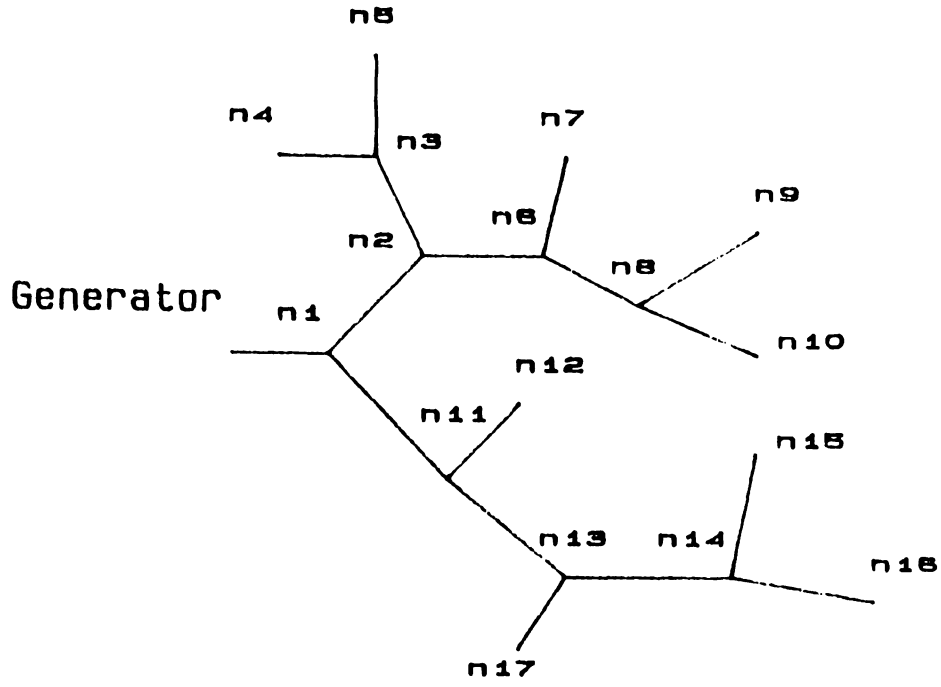


Fig. 5. Transmission line network as a tree structure

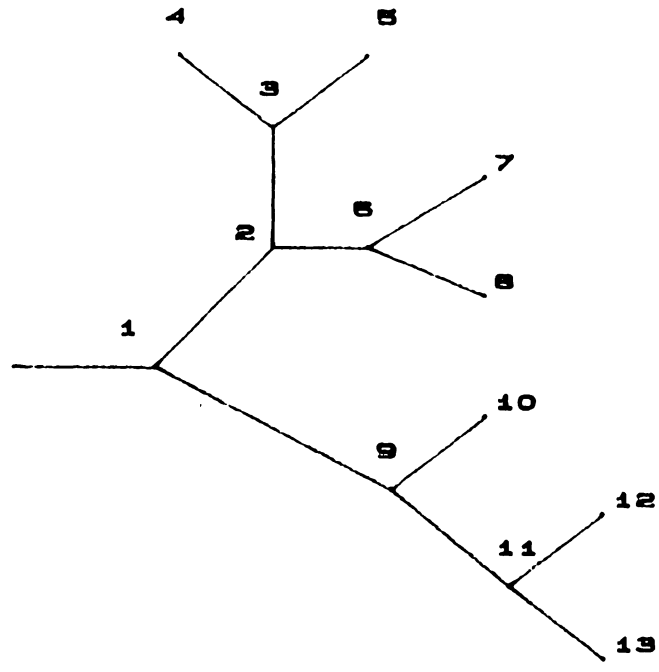


Fig. 6. Post order traversing of tree for impedance calculations

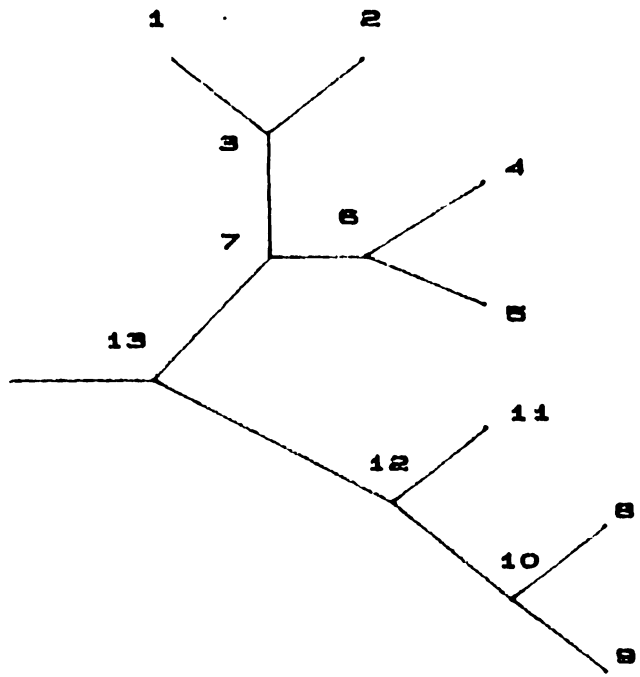


Fig. 7. Preorder traversing of tree for current and voltage calculation.

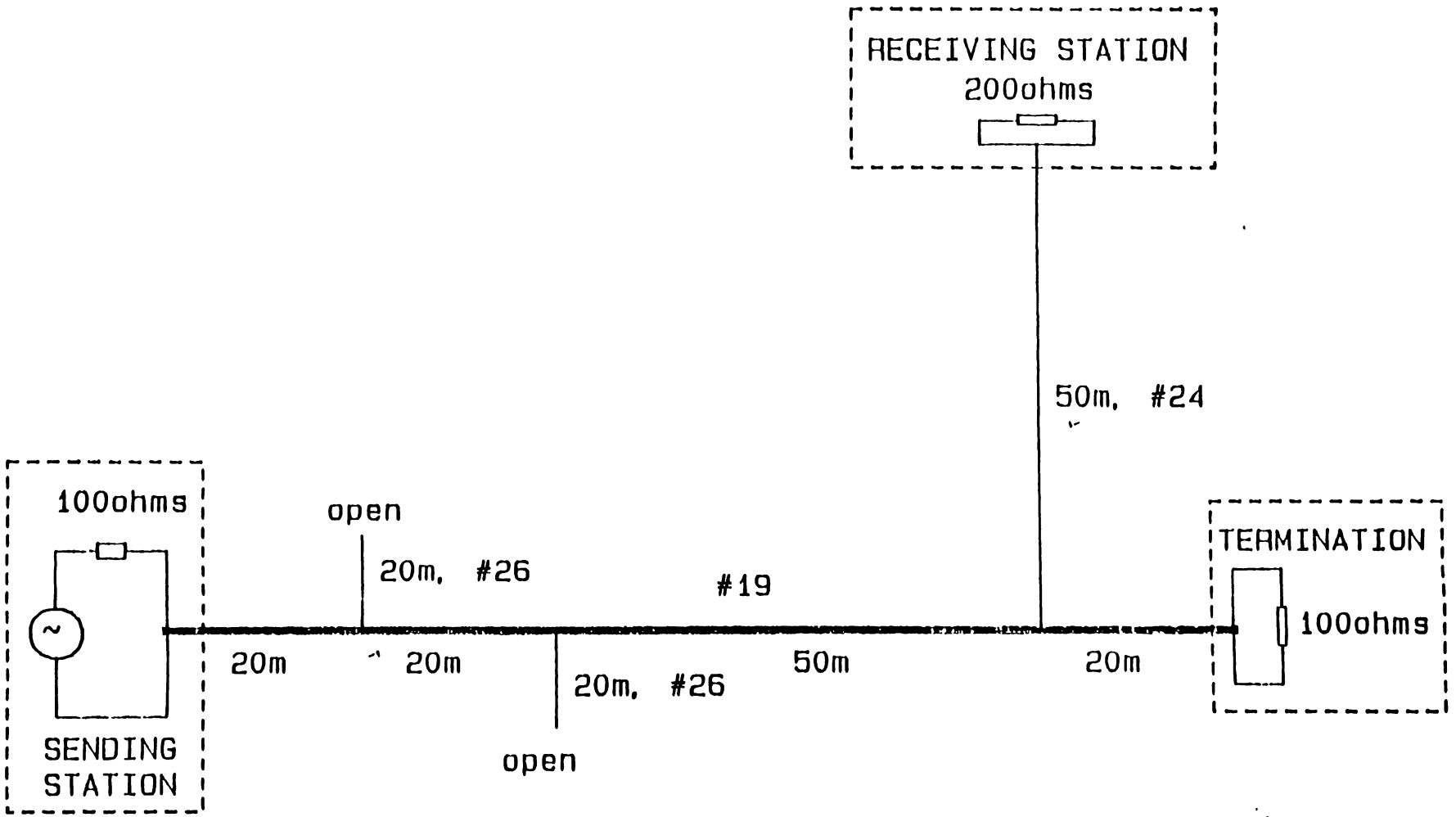


Fig. 8. Test network for simulation.

Fig. 9. Computed impulse response between sending and receiving stations of test network

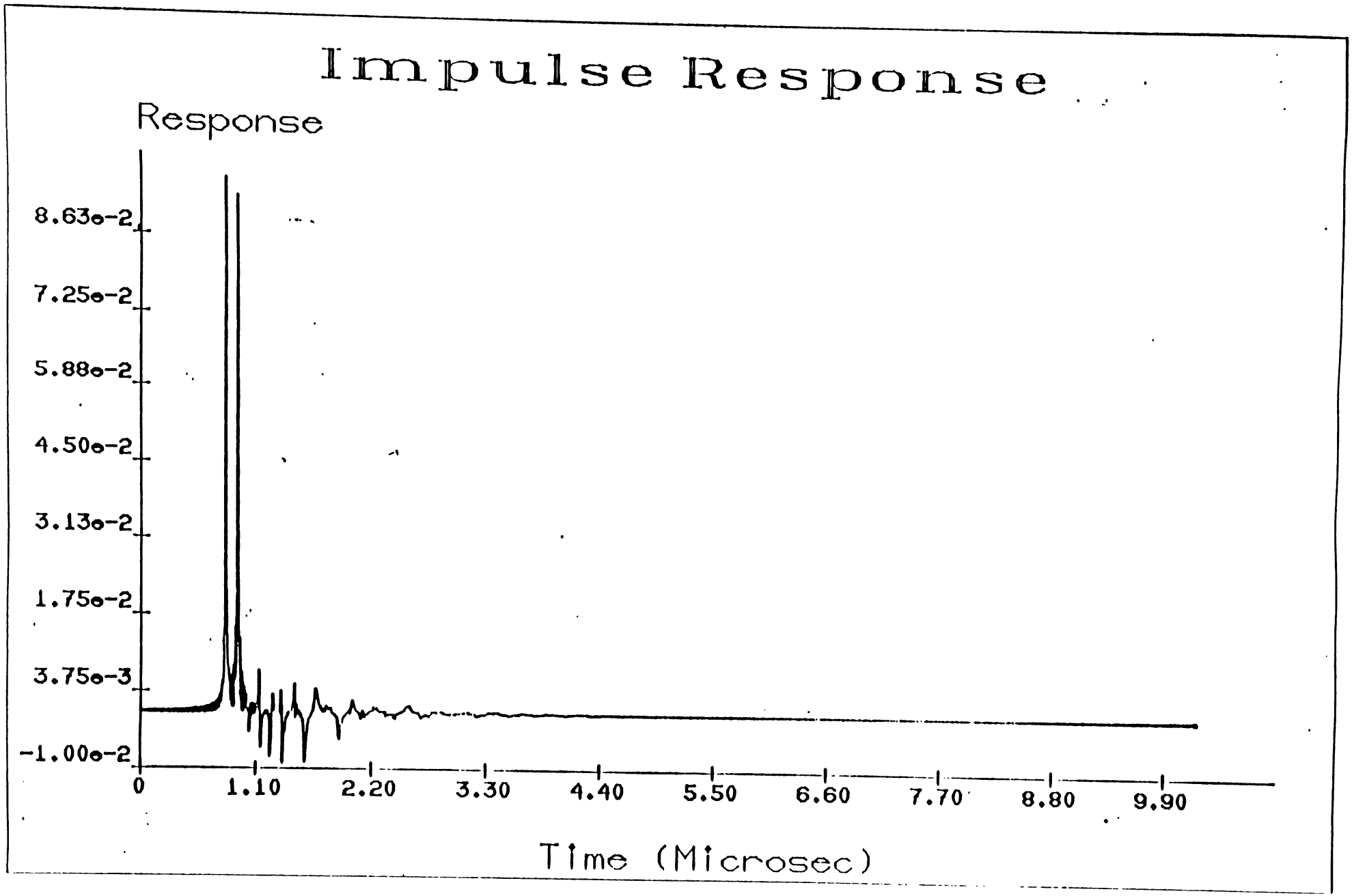
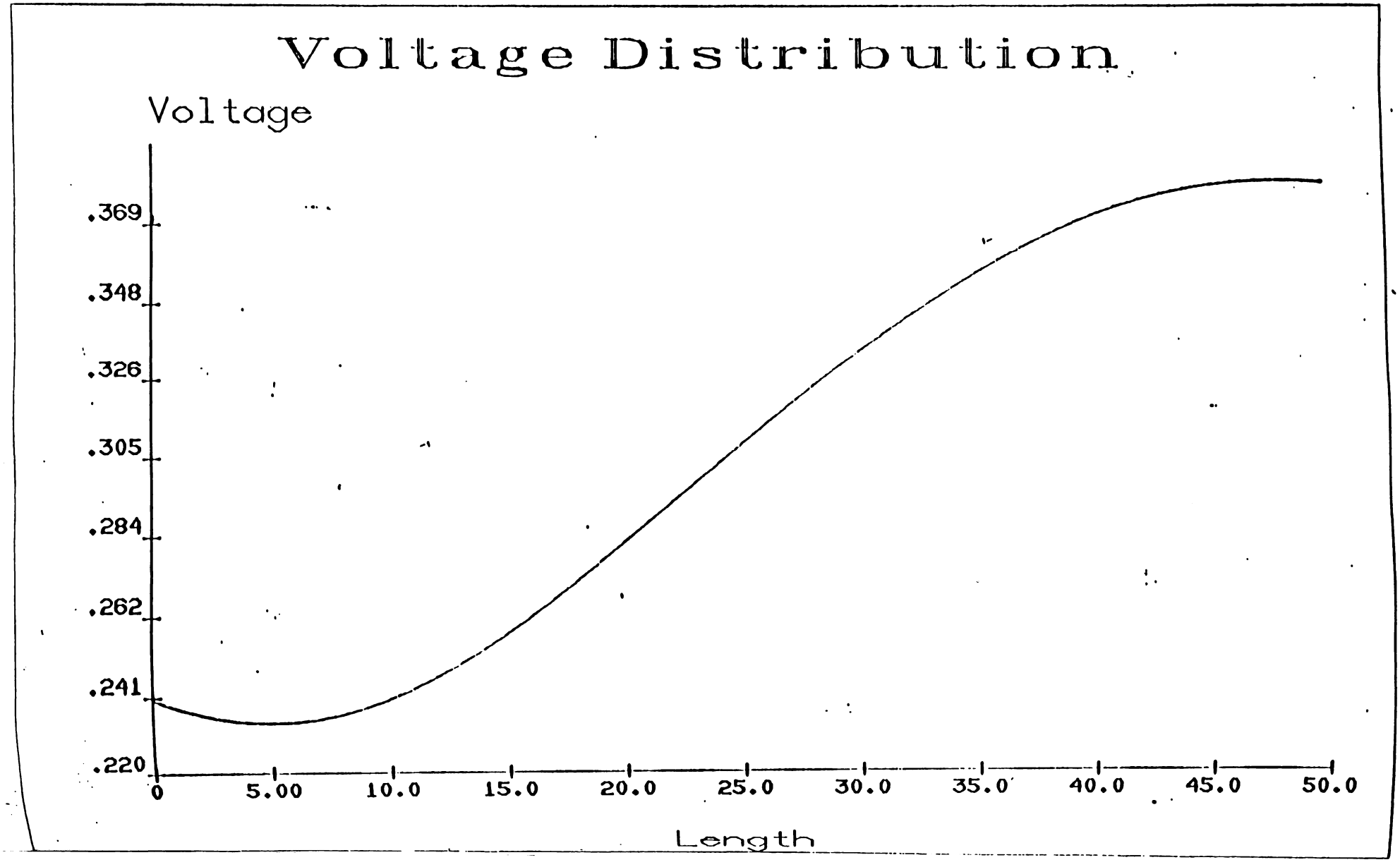


Fig. 10. Voltage distribution on the drop cable at the receiving station (1 MHz)



Appendix C. Z0STRP — Program for determining Microstrip Parameters

The program Z0STRP is a stand alone program for determining the transmission line parameters of a microstrip line imbedded in possibly lossy dielectric as shown in Figure C.1. The default values of the microstrip system parameters are given in Table C.1. The

```
Relative Permitivity of dielectric, epsr2   4.10000
Relative Permitivity of covering dielectric , epsr1   1.00000
Frequency in Hz,f   1.00000e+09
Thickness of microstrip, t   2.50000e-05
Thickness of covering dielectric, d  0.
Thickness of dielectric, h   3.50000e-04
Resistivity of metal gold=2.44E-8
copper= 1.7241e-8 , rhomi   1.72410e-08
Loss tangent of dielectric, tand2  0.
Loss tangent of covering dielectric, tand1  0.
```

Table 5: C.1: The default values of the microstrip system parameters.

results of a z0strp analysis for the tracks considered in this report are presented below.

```
MS7 microstrip with solid ground, 8 mil width
   Width of microstrip, w   2.03200e-04
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   2.03200e-04   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   87.0125,  -3.50403)
   Effective Permitivity, epseff=   2.84991
   Resistance per meter, r (Ohms)=   247.984
   Capacitance per meter, c (F)  =   6.47239e-11
   Inductance per meter, l (H)   =   4.89241e-07
   Loss in dB per meter, alpha   =   1.42499
   Wavenumber  (1/m) , beta      =   35.3855
  ------------------------------------------------
-------------------------------------------------------------------------------

MS6 microstrip with solid ground, 13 mil width
   Width of microstrip, w   3.33000e-04
  ------------------------------------------------
```

```
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   3.30000e-04   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   71.6058,  -2.46820)
   Effective Permitivity, epseff=   2.95540
   Resistance per meter, r (Ohms)=   177.842
   Capacitance per meter, c (F)  =   8.00750e-11
   Inductance per meter, l (H)   =   4.10088e-07
   Loss in dB per meter, alpha   =   1.24181
   Wavenumber  (1/m) , beta      =   36.0267
-------------------------------------------------------------------------------

MS5 microstrip with solid ground, 16 mil width
   Width of microstrip, w   4.06000e-04
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   4.06000e-04   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   65.0426,  -2.09314)
   Effective Permitivity, epseff=   3.00503
   Resistance per meter, r (Ohms)=   152.067
   Capacitance per meter, c (F)  =   8.88853e-11
   Inductance per meter, l (H)   =   3.75644e-07
   Loss in dB per meter, alpha   =   1.16898
   Wavenumber  (1/m) , beta      =   36.3252
  ------------------------------------------------
-------------------------------------------------------------------------------

MS4 microstrip with solid ground, 20 mil width
   Width of microstrip, w   5.08000e-04
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   5.08000e-04   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   58.1868,  -1.80353)
   Effective Permitivity, epseff=   3.06187
   Resistance per meter, r (Ohms)=   132.256
   Capacitance per meter, c (F)  =   1.00290e-10
   Inductance per meter, l (H)   =   3.39225e-07
```

```
   Loss in dB per meter, alpha    =   1.13648
   Wavenumber  (1/m) , beta       =   36.6657
  ------------------------------------------------
-------------------------------------------------------------------------------

MS3 microstrip with solid ground, 25 mil width
   Width of microstrip, w   6.35000e-04
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   6.35000e-04   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   51.5247,  -1.54007)
   Effective Permitivity, epseff=   3.12167
   Resistance per meter, r (Ohms)=   114.029
   Capacitance per meter, c (F)  =   1.14354e-10
   Inductance per meter, l (H)   =   3.03315e-07
   Loss in dB per meter, alpha   =   1.10655
   Wavenumber  (1/m) , beta      =   37.0208
-------------------------------------------------------------------------------

MS2 microstrip with solid ground, 50 mil width
   Width of microstrip, w   1.27000e-03
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
   f w t d h=      1.00000e+09   1.27000e-03   2.50000e-05  0.   3.50000e-04
   rhomi tand1 tand2 =   1.72410e-08  0.  0.

   Complex Characteristic     Z0 =   (   33.1267, -0.900268)
   Effective Permitivity, epseff=   3.32452
   Resistance per meter, r (Ohms)=   68.7836
   Capacitance per meter, c (F)  =   1.83538e-10
   Inductance per meter, l (H)   =   2.01262e-07
   Loss in dB per meter, alpha   =   1.03819
   Wavenumber  (1/m) , beta      =   38.2017
  ------------------------------------------------
-------------------------------------------------------------------------------

MS1 microstrip with solid ground, 100 mil width
   Width of microstrip, w   2.54000e-03
  ------------------------------------------------
   epsr2 epsr1=    4.10000   1.00000
```

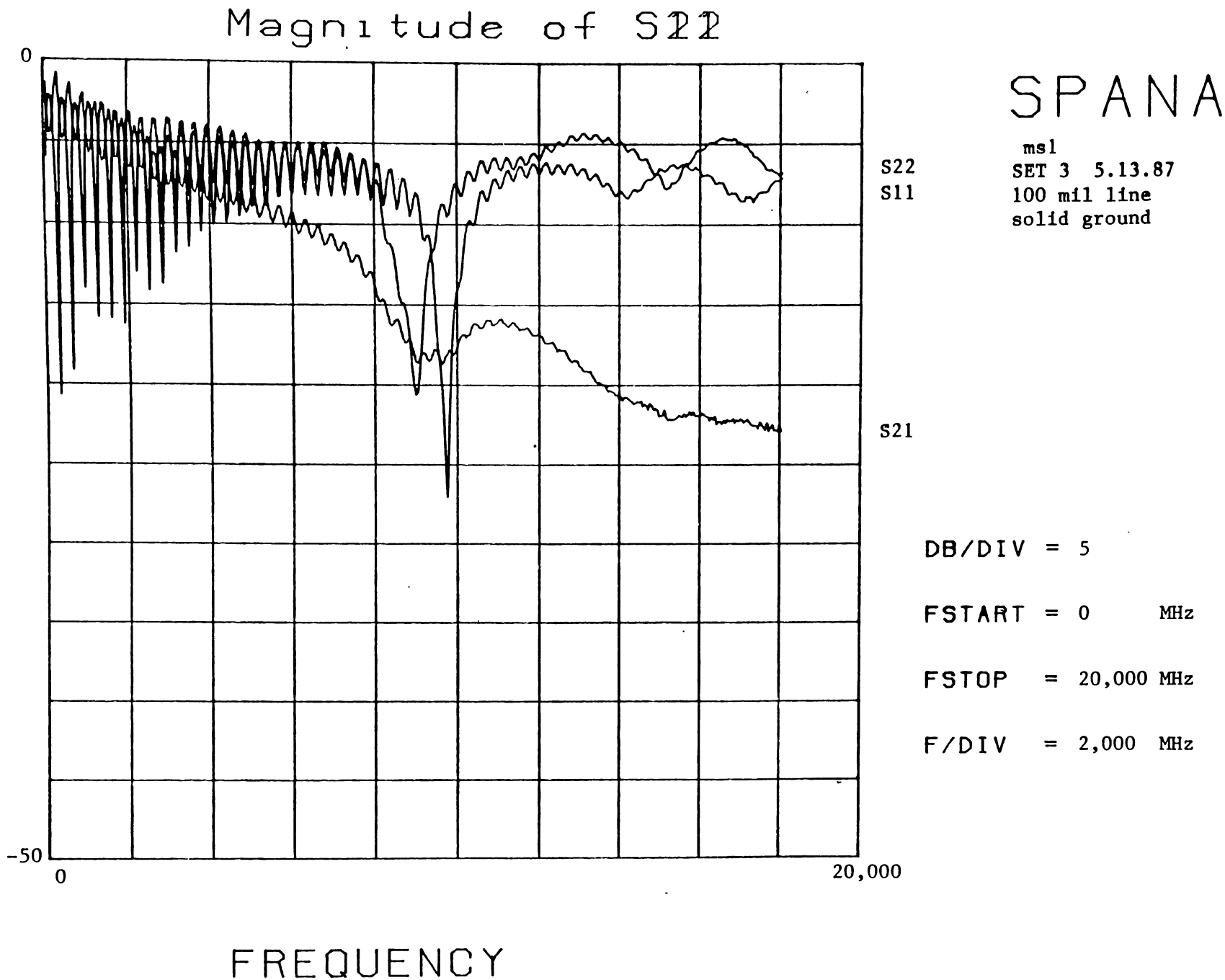
```
f w t d h=      1.00000e+09   2.54000e-03   2.50000e-05  0.   3.50000e-04
rhomi tand1 tand2 =   1.72410e-08  0.  0.

Complex Characteristic     Z0 =   (   19.6087, -0.503814)
Effective Permitivity, epseff=   3.53728
Resistance per meter, r (Ohms)=   39.7042
Capacitance per meter, c (F)  =   3.19822e-10
Inductance per meter, l (H)   =   1.22891e-07
Loss in dB per meter, alpha   =   1.01241
Wavenumber  (1/m) , beta      =   39.4036
-------------------------------------------------
```

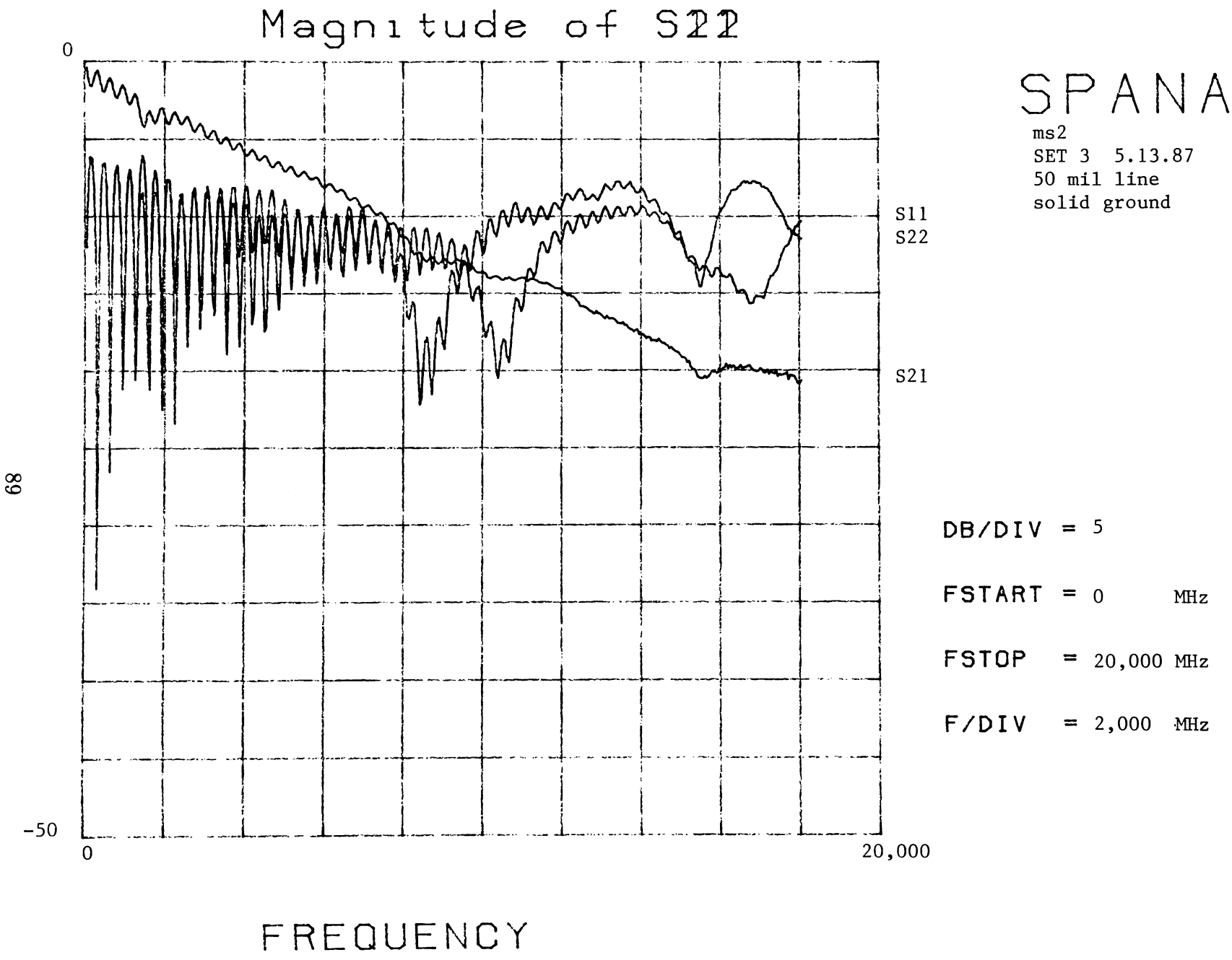
```
region 0                                    air

        ------------------------------------------
region 1                      covering dielectric d
                                                  |
                       v<-- w -->|                v
        ---------------xxxxxxxxxxx-----------------
                      t^                          ^
                               microstrip         |
region 2                                          h
                                                  v
        ------------------------------------------
        ////////////////////////////////////////////
                              ground plane
```

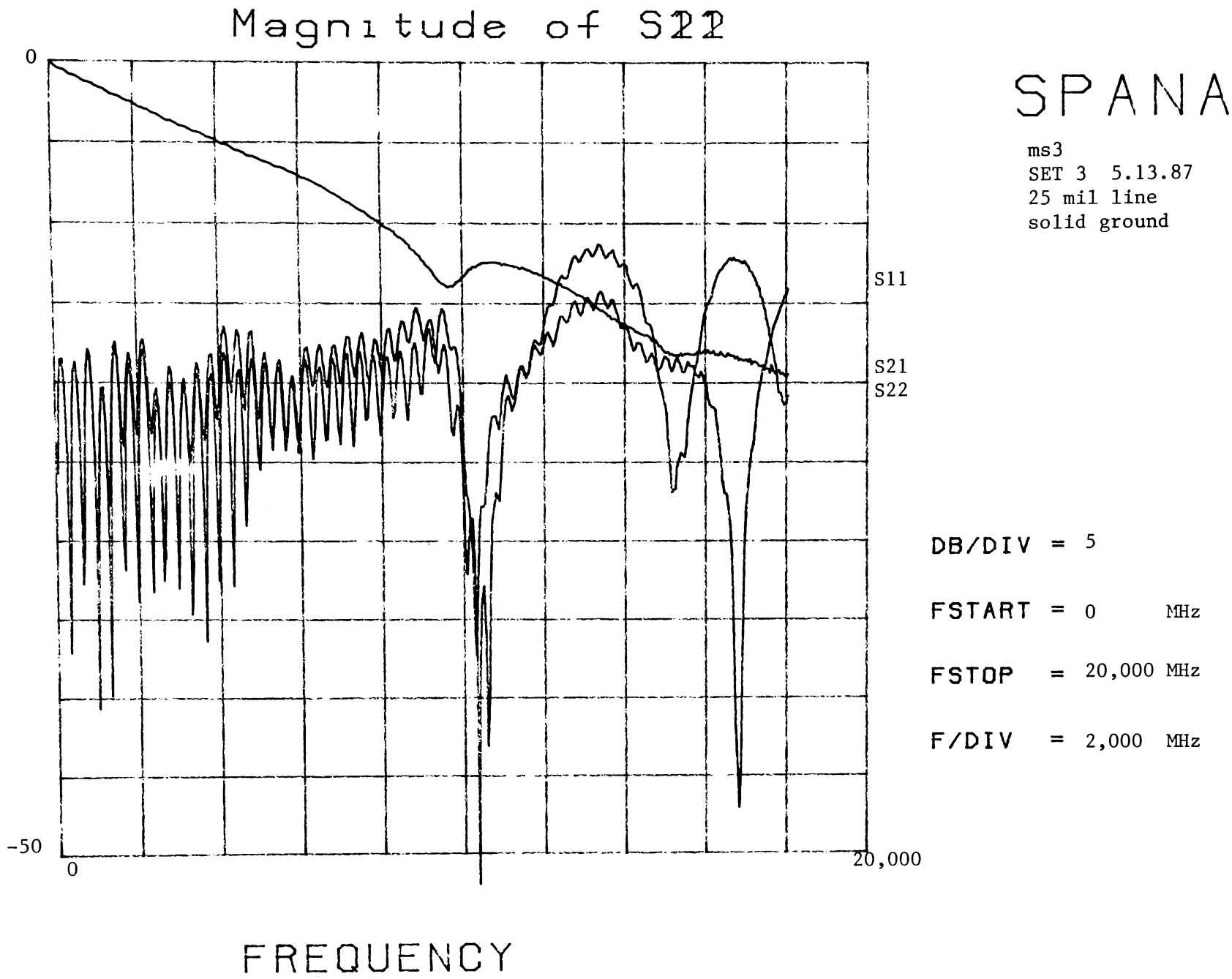
Figure 15: C.1: Microstrip line imbedded in possibly lossy dielectric with parameters epsr1, epsr2 the relative permitivity of regions 1,2; tand1,tand2 the loss tangent of the dielectric in regions 1,2; d,h the thickness of the dielectrics in region 1,2; t and w the thickness and width of the microstrip line; and rhomi the resistivity of the track and ground plane.

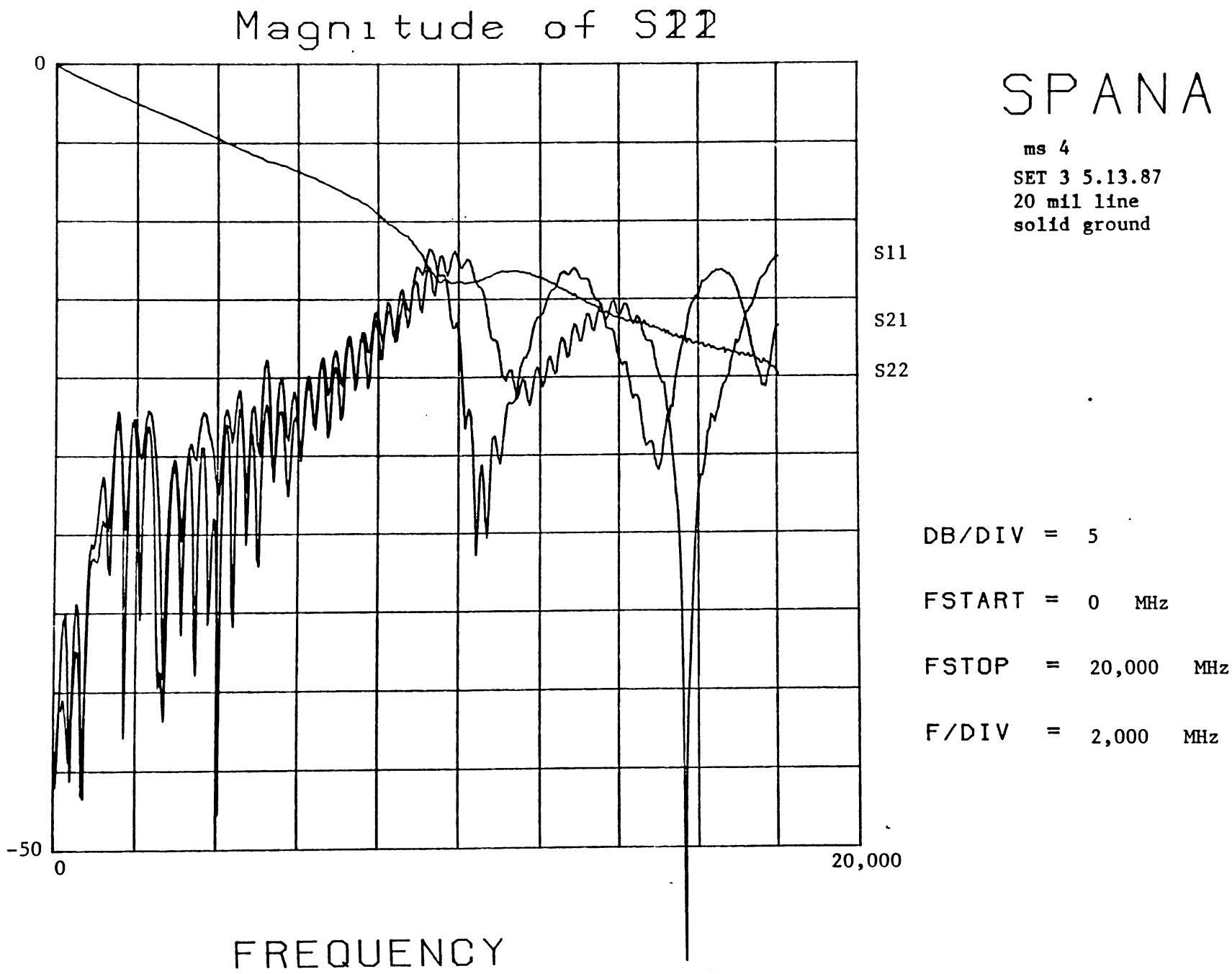
Appendix D. Experimental characterization of Test Board

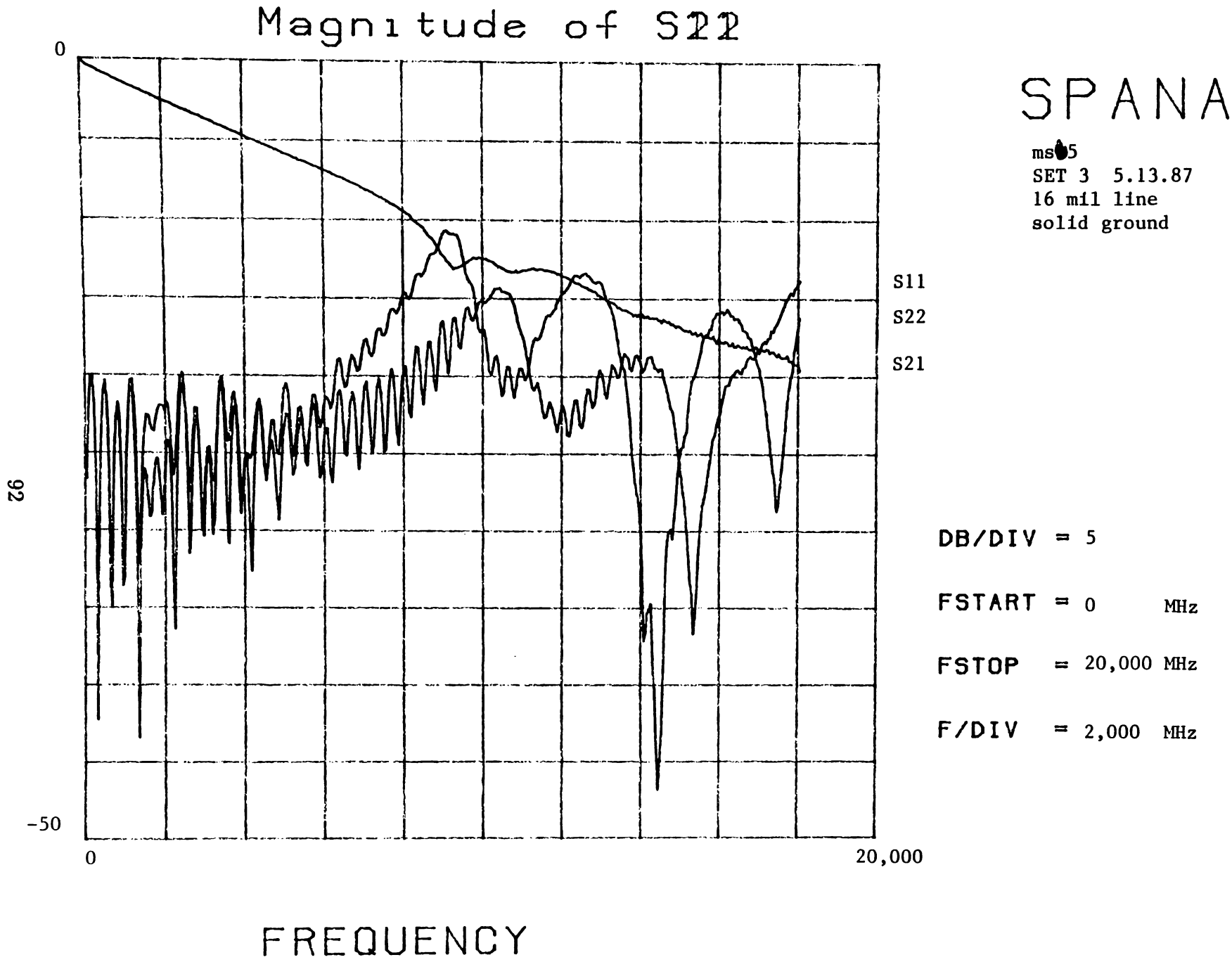
Here we present experimental characterization of the first BNR four layer (5D technology) test board. The test board tracks are shown in Figure 2. Details of the experimental setup are given in section 4.

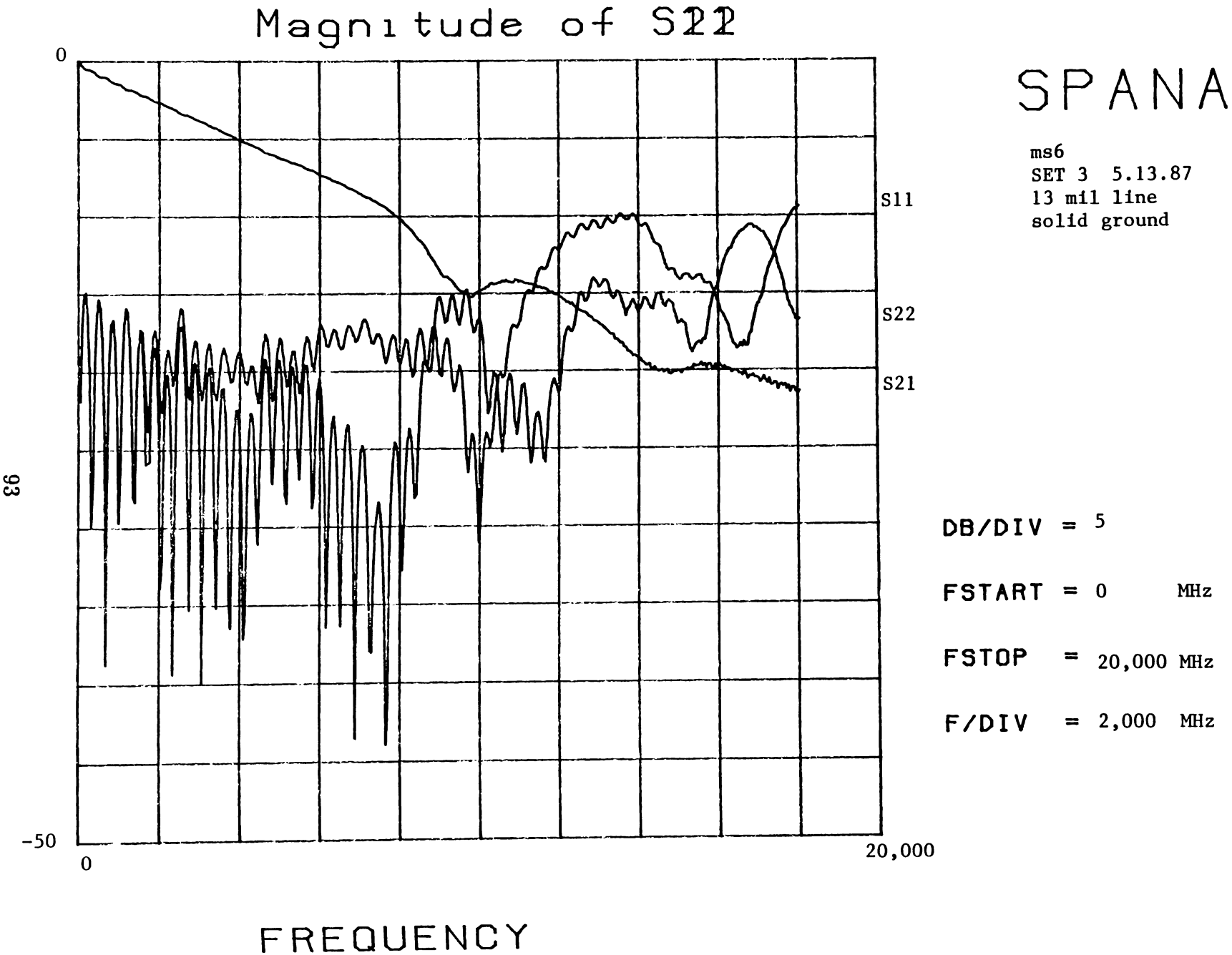

Magnitude of S22
0
-50
0
20,000
S22
S11
S21
SPANA
msl
SET 3 5.13.87
100 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz
FREQUENCY

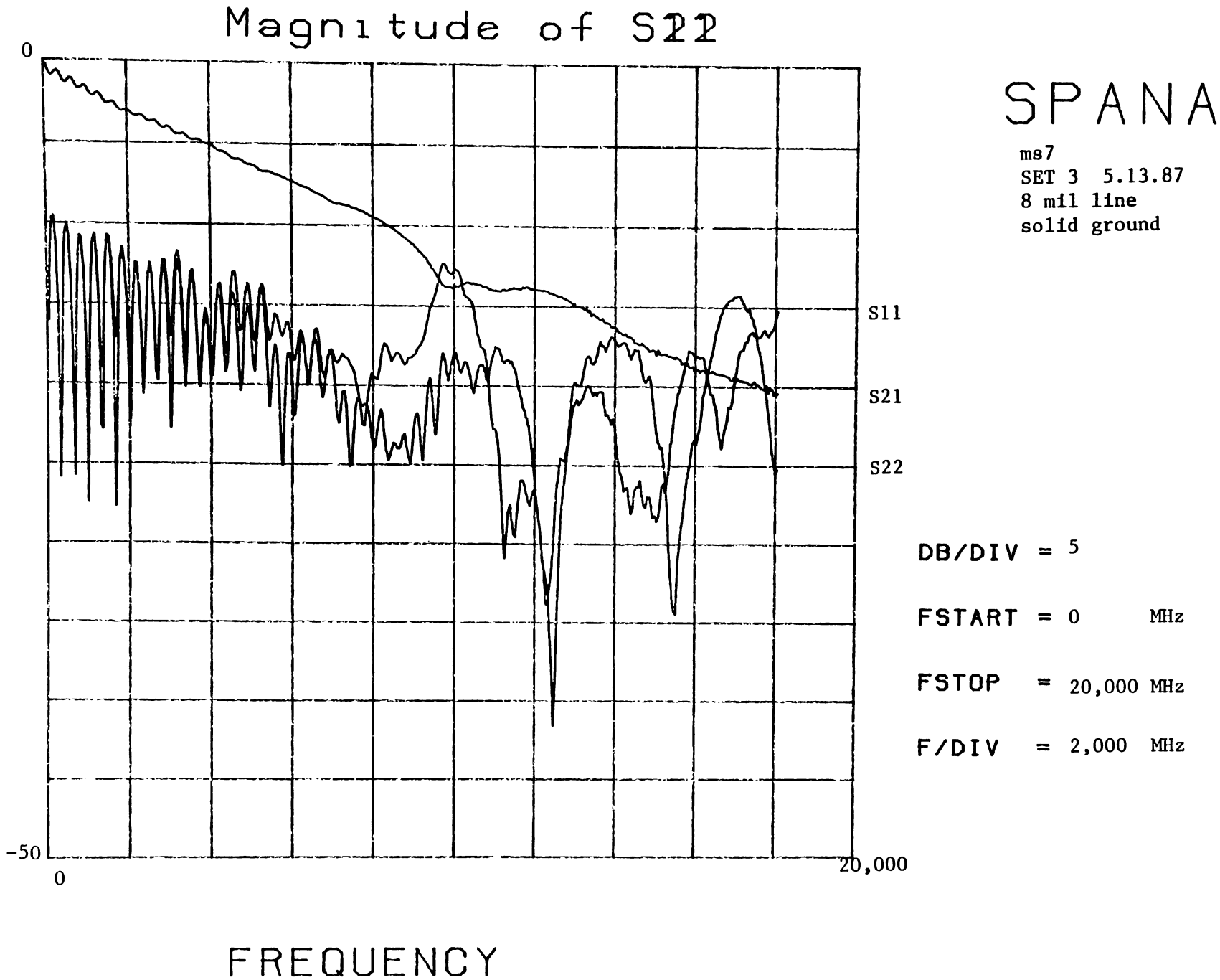

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
ms2
SET 3 5.13.87
50 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

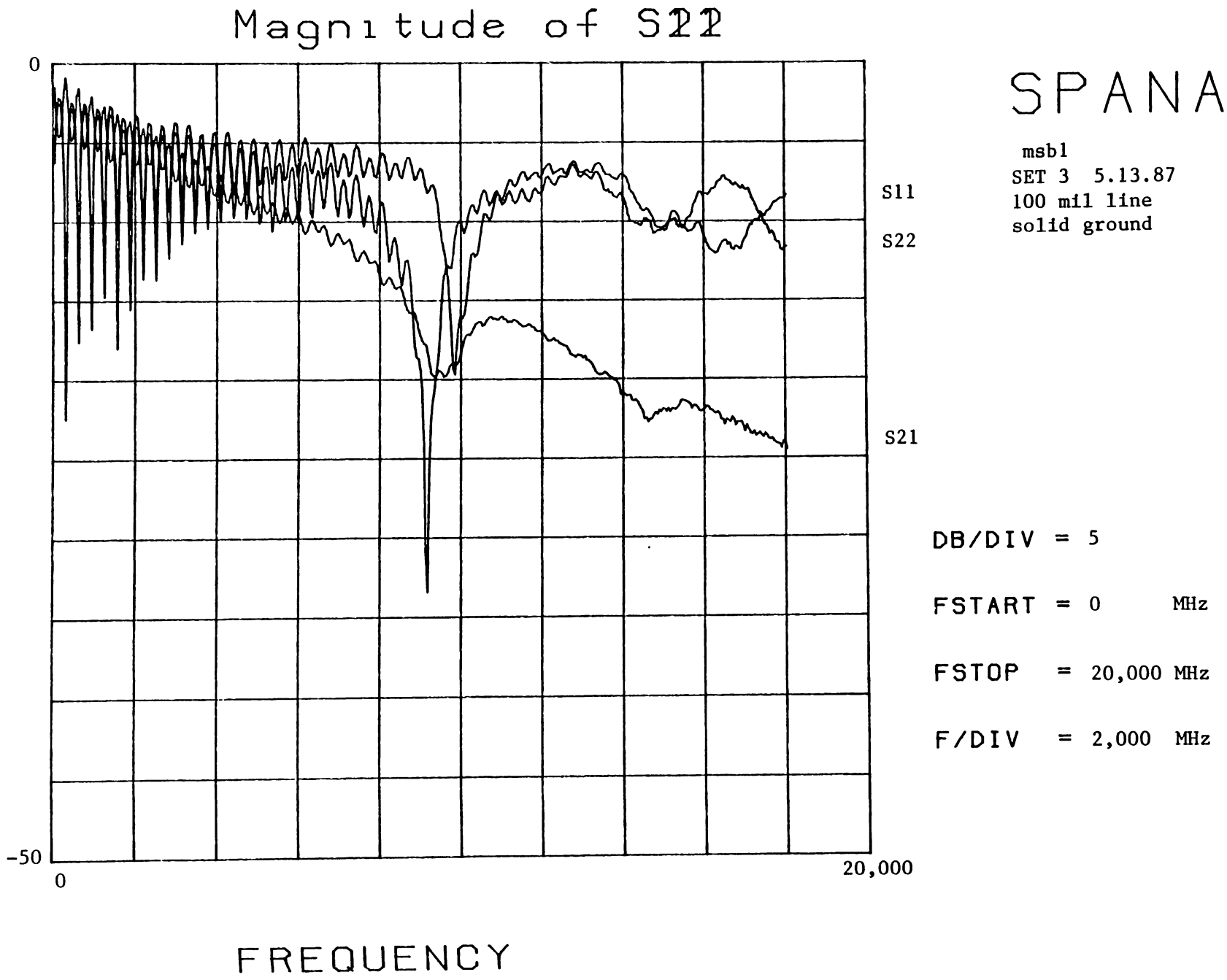

Magnitude of S21
0
-50
0
20,000
S11
S21
S22
FREQUENCY
SPANA
ms3
SET 3 5.13.87
25 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

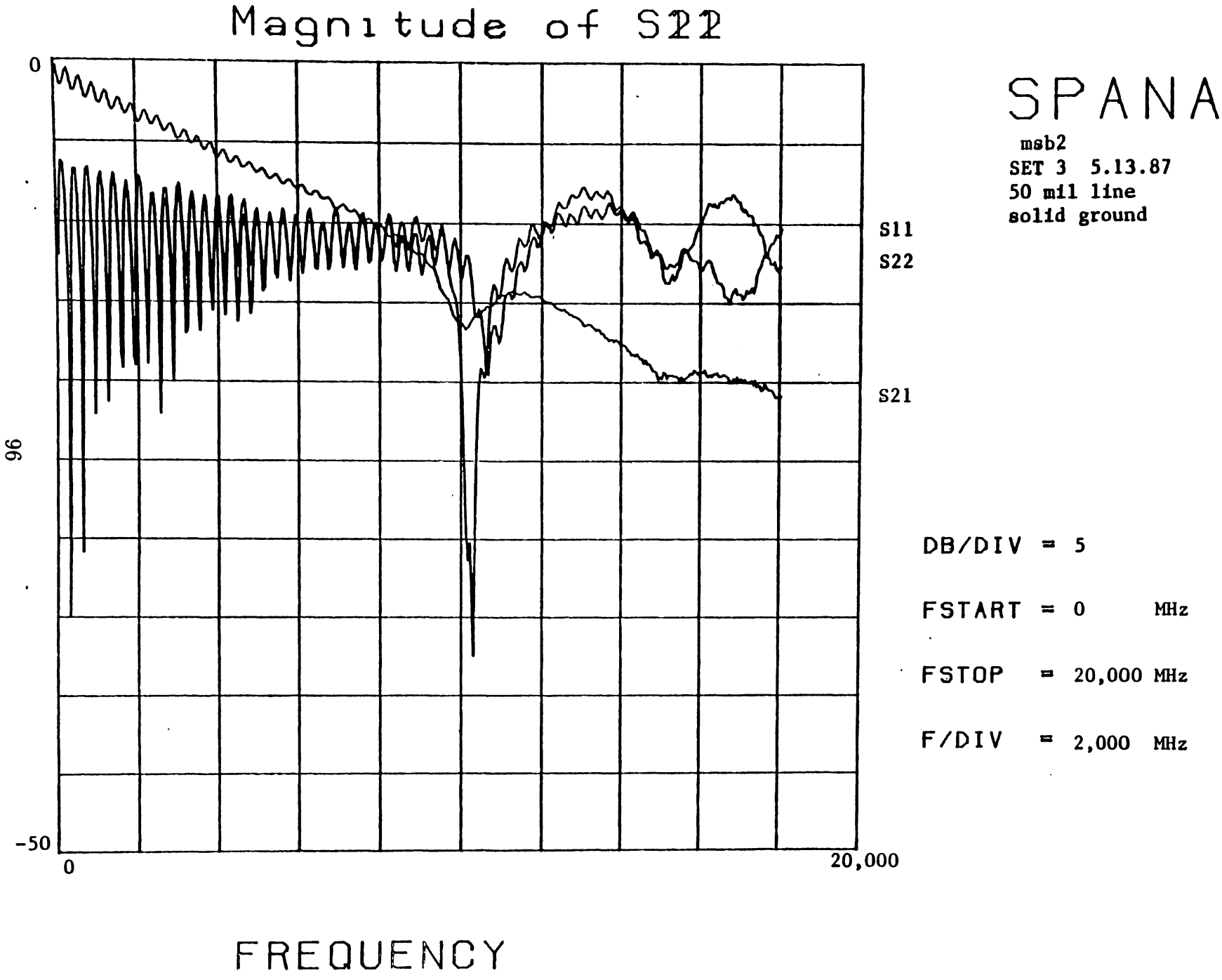

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S11
S21
S22
SPANA
ms 4
SET 3 5.13.87
20 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

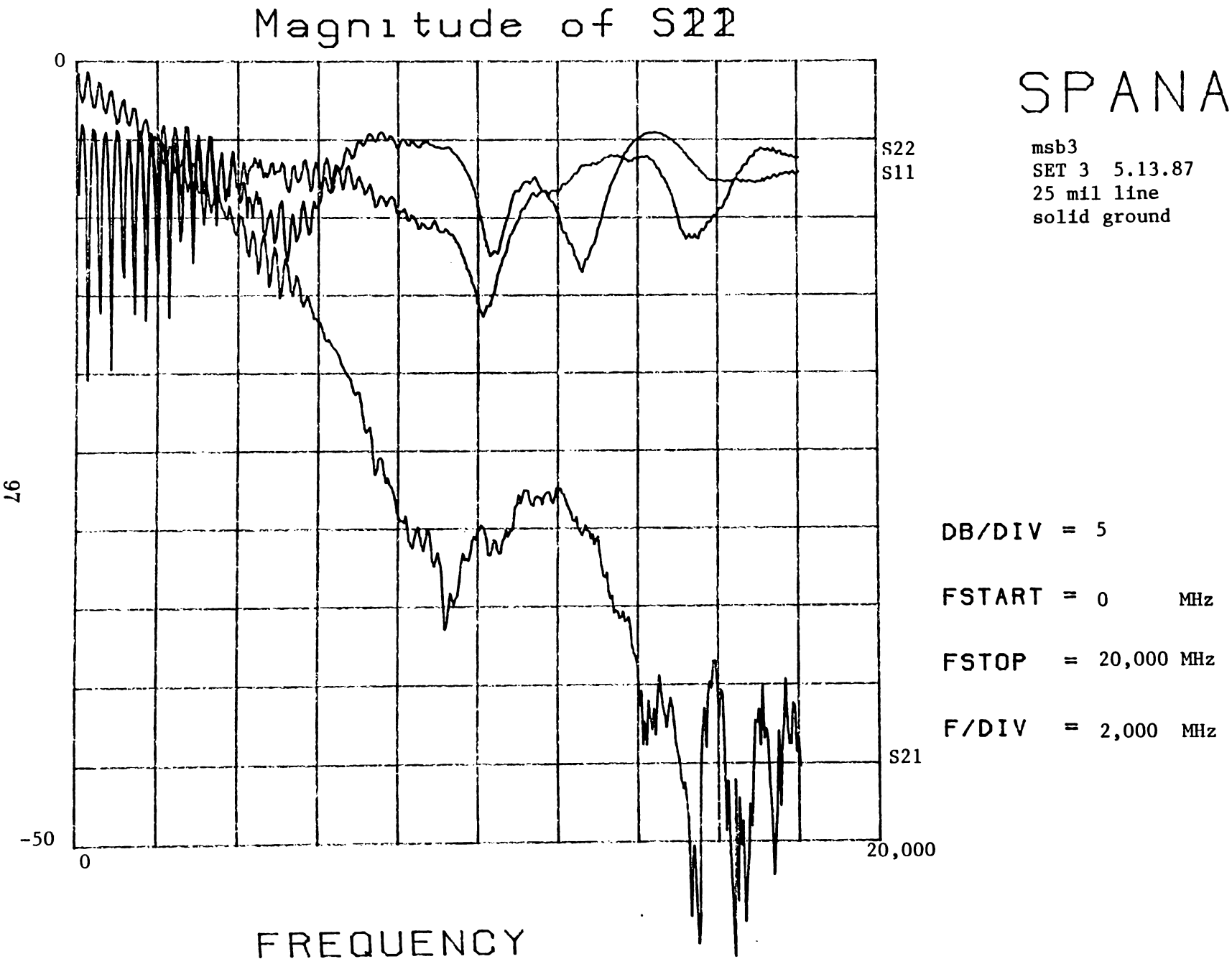

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
ms05
SET 3 5.13.87
16 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

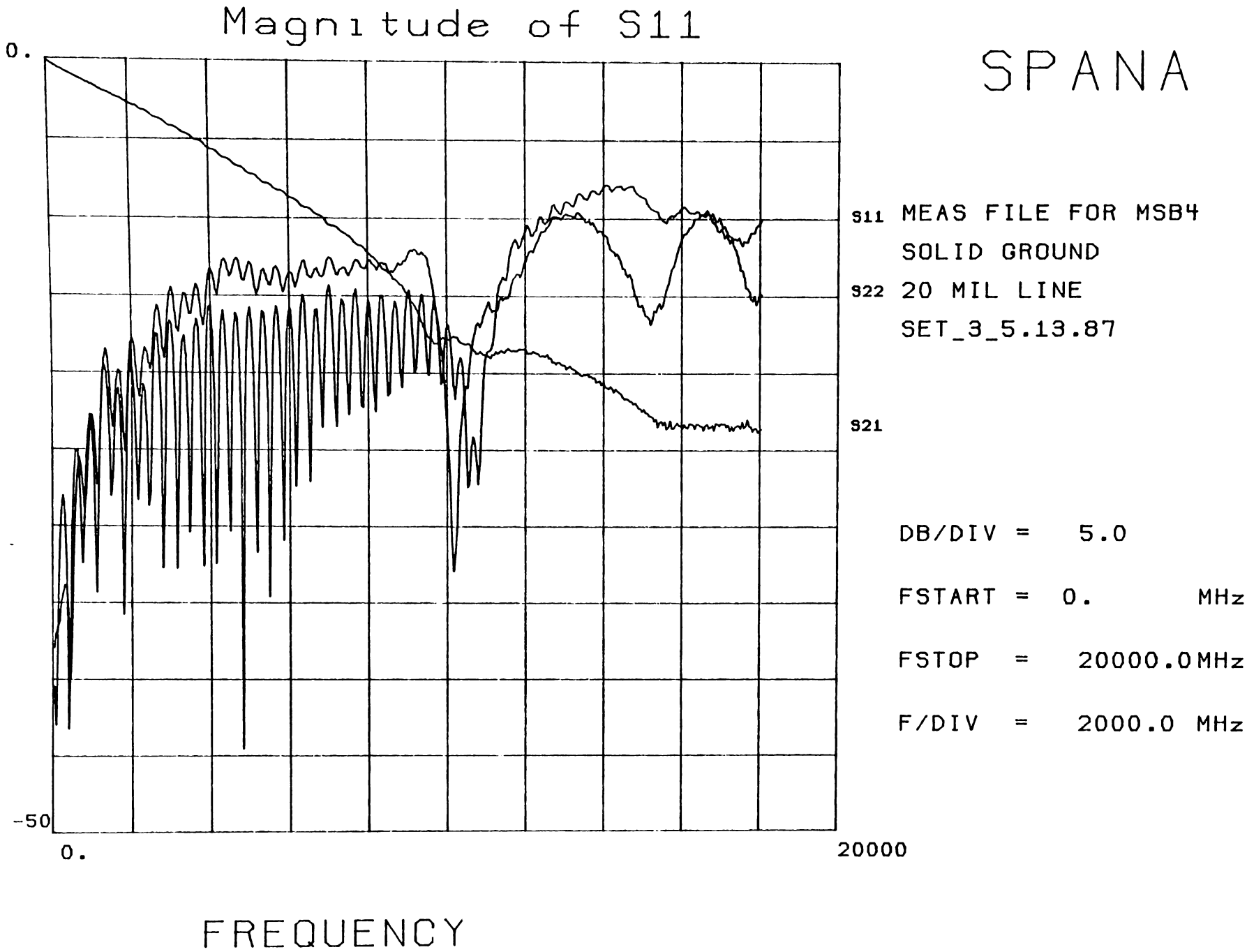

Magnitude of S22
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
ms6
SET 3 5.13.87
13 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

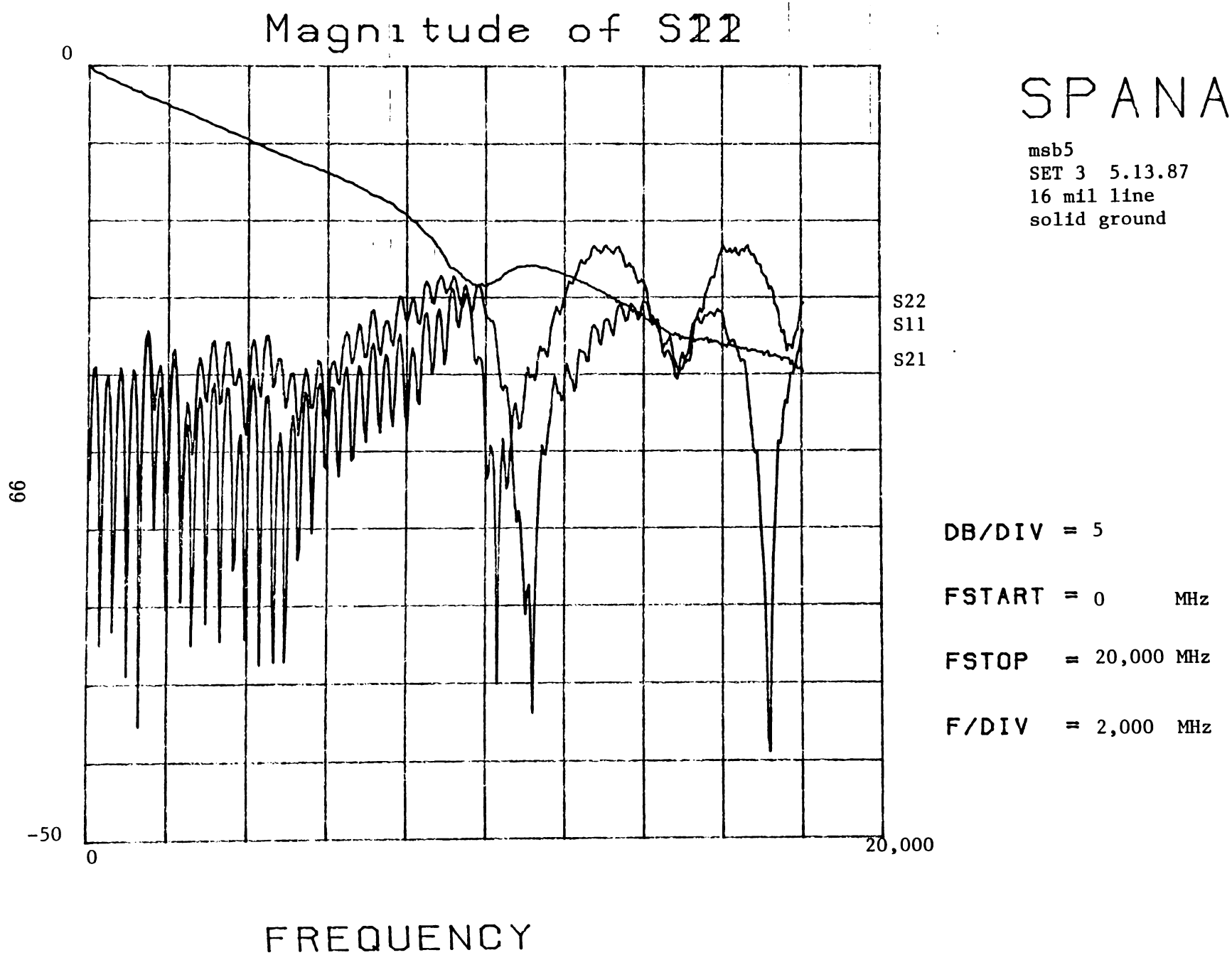

Magnitude of S22
0
-50
0
20,000
S11
S21
S22
SPANA
ms7
SET 3 5.13.87
8 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz
FREQUENCY

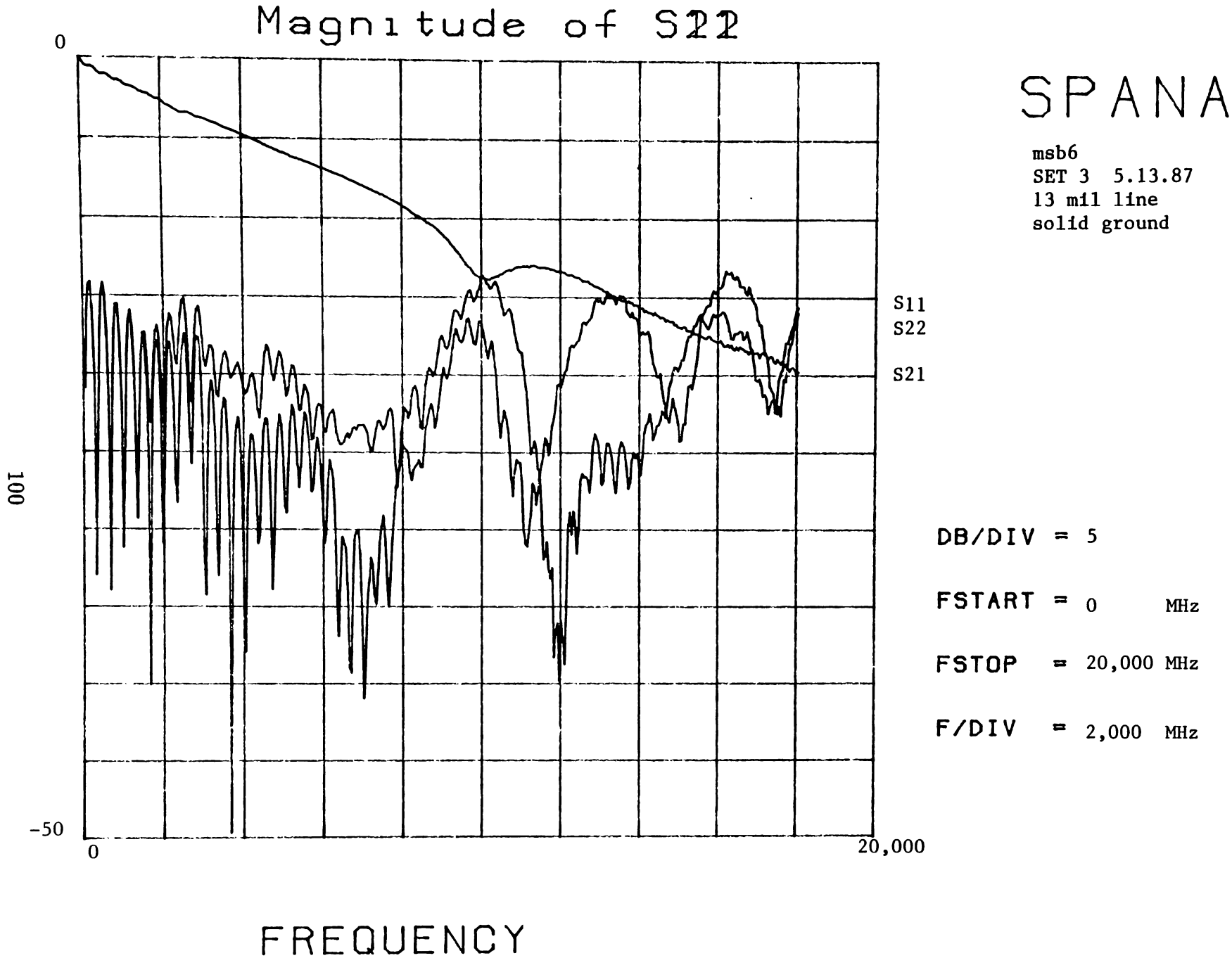

Magnitude of S11
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
msb1
SET 3 5.13.87
100 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

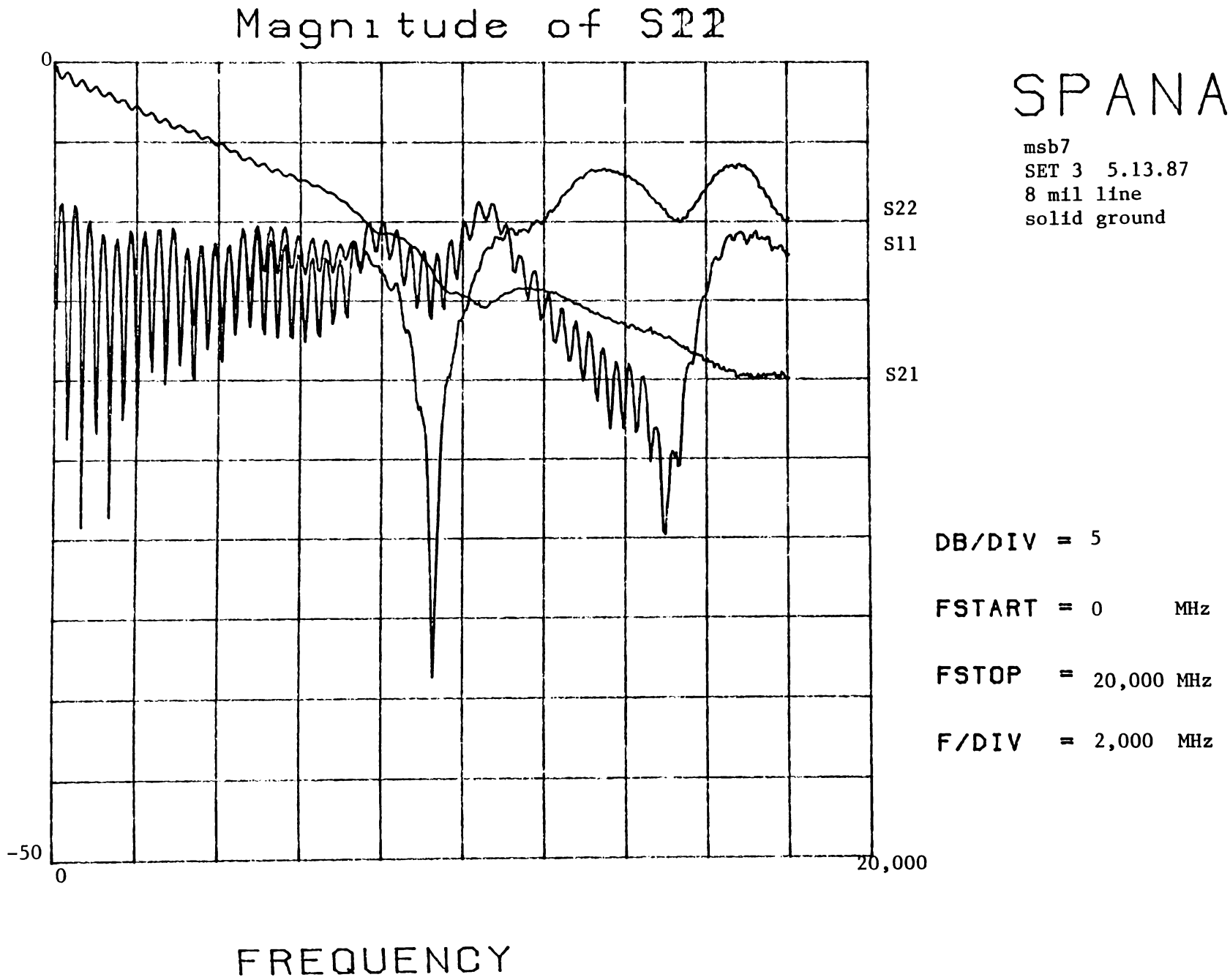

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
msb2
SET 3 5.13.87
50 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

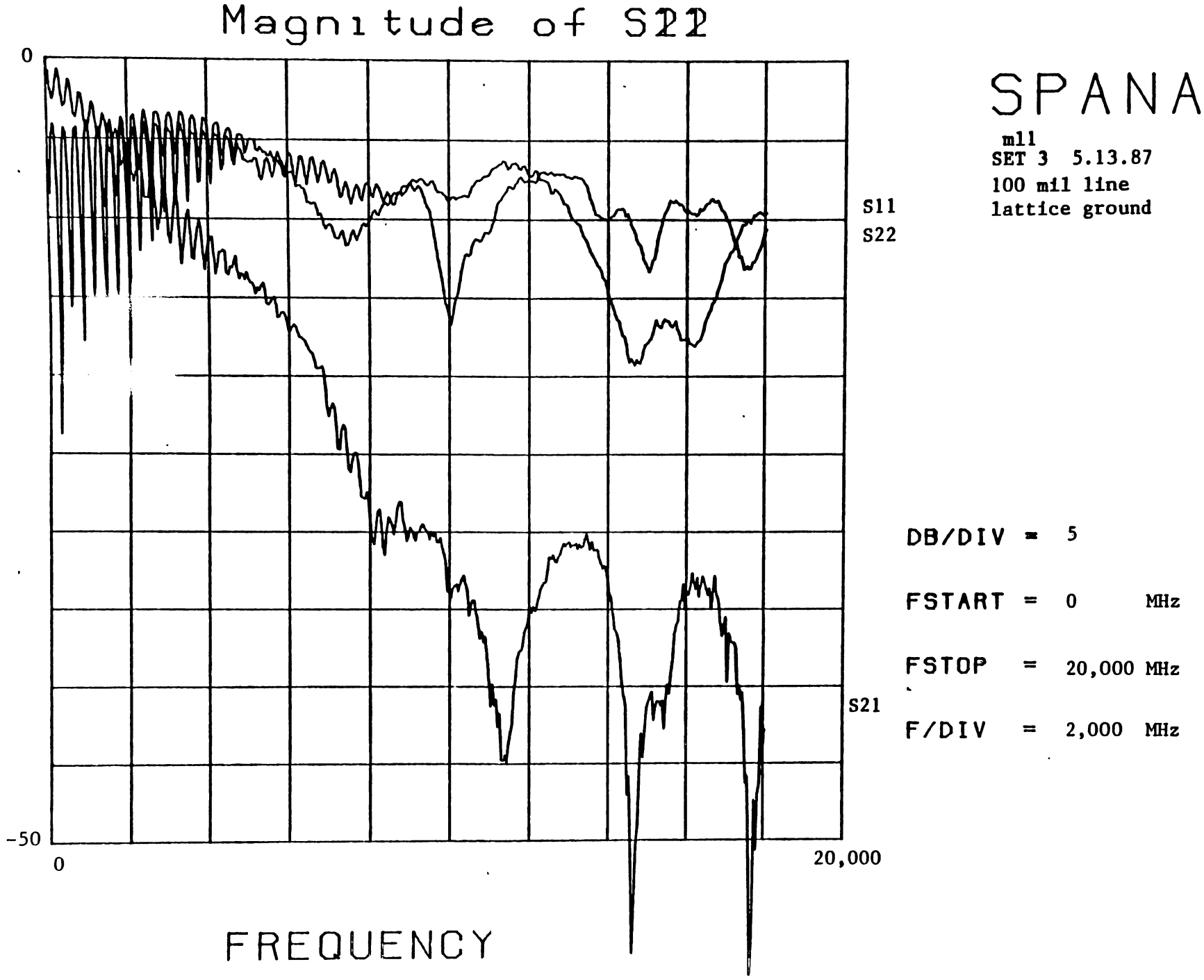

Magnitude of S22
0
-50
0
20,000
S22
S11
S21
FREQUENCY
SPANA
msb3
SET 3 5.13.87
25 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

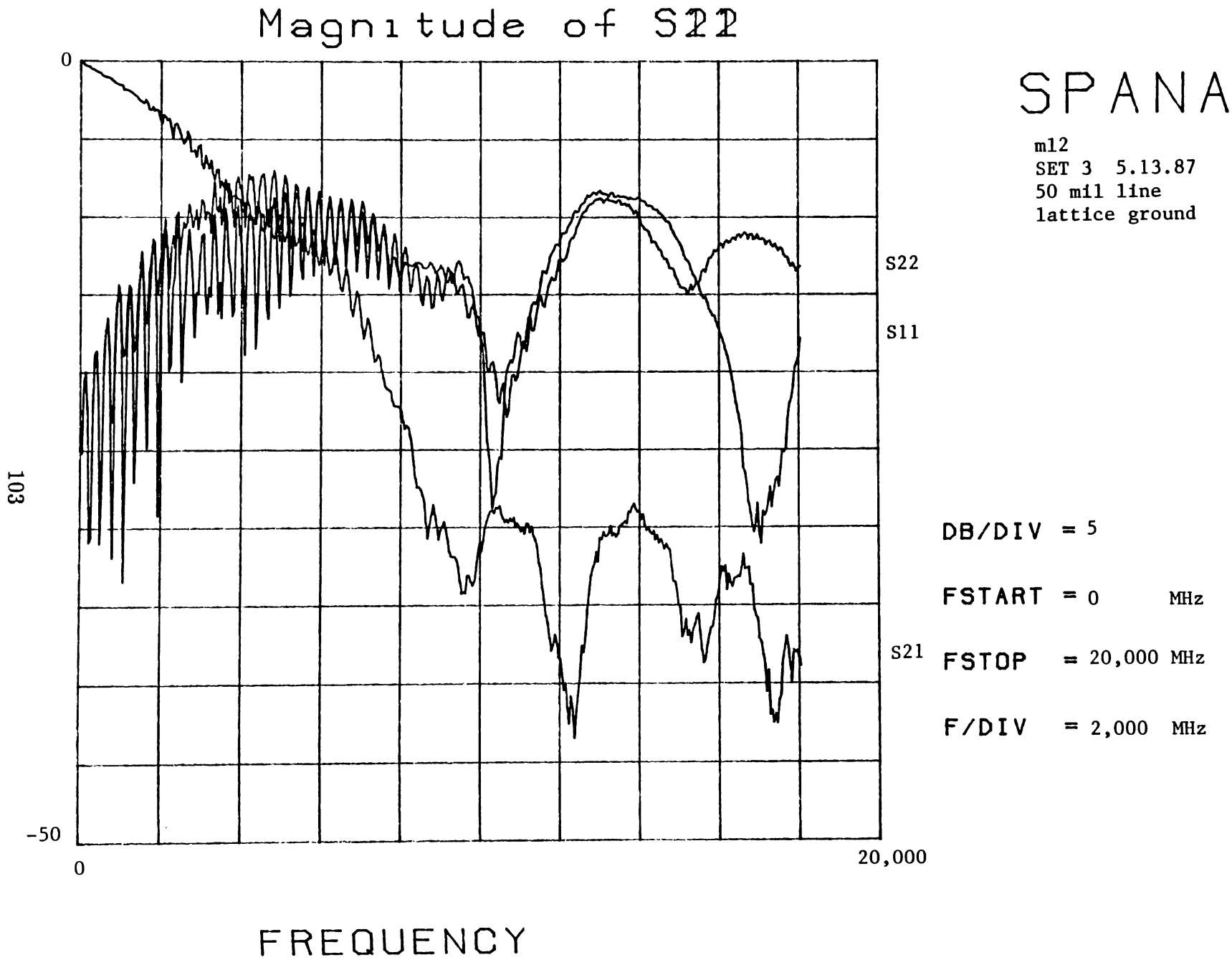

Magnitude of S11
SPANA
0.
-50
0.
20000
FREQUENCY
S11 MEAS FILE FOR MSB4
SOLID GROUND
S22 20 MIL LINE
SET_3_5.13.87
S21
DB/DIV = 5.0
FSTART = 0. MHz
FSTOP = 20000.0MHz
F/DIV = 2000.0 MHz

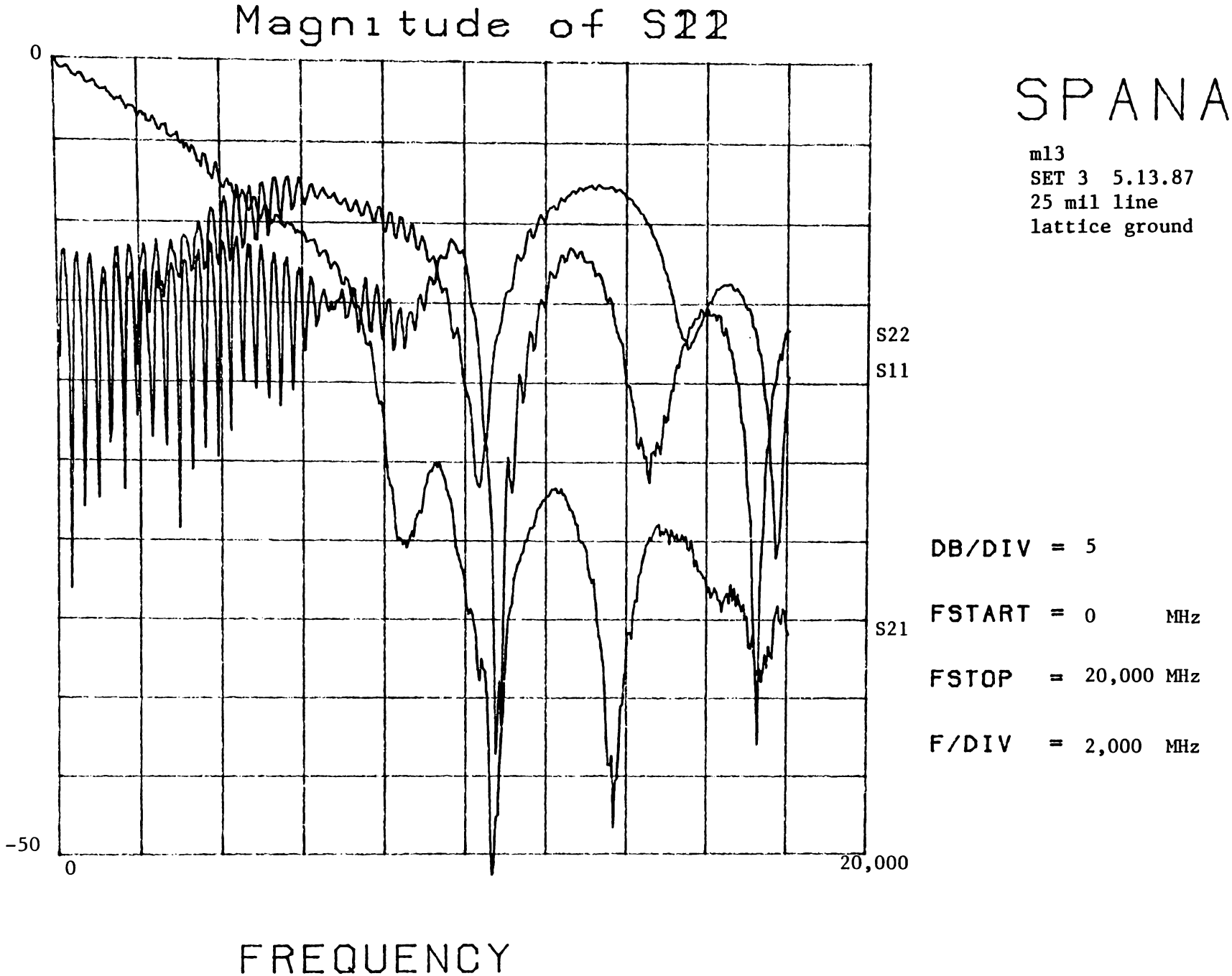

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S22
S11
S21
SPANA
msb5
SET 3 5.13.87
16 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

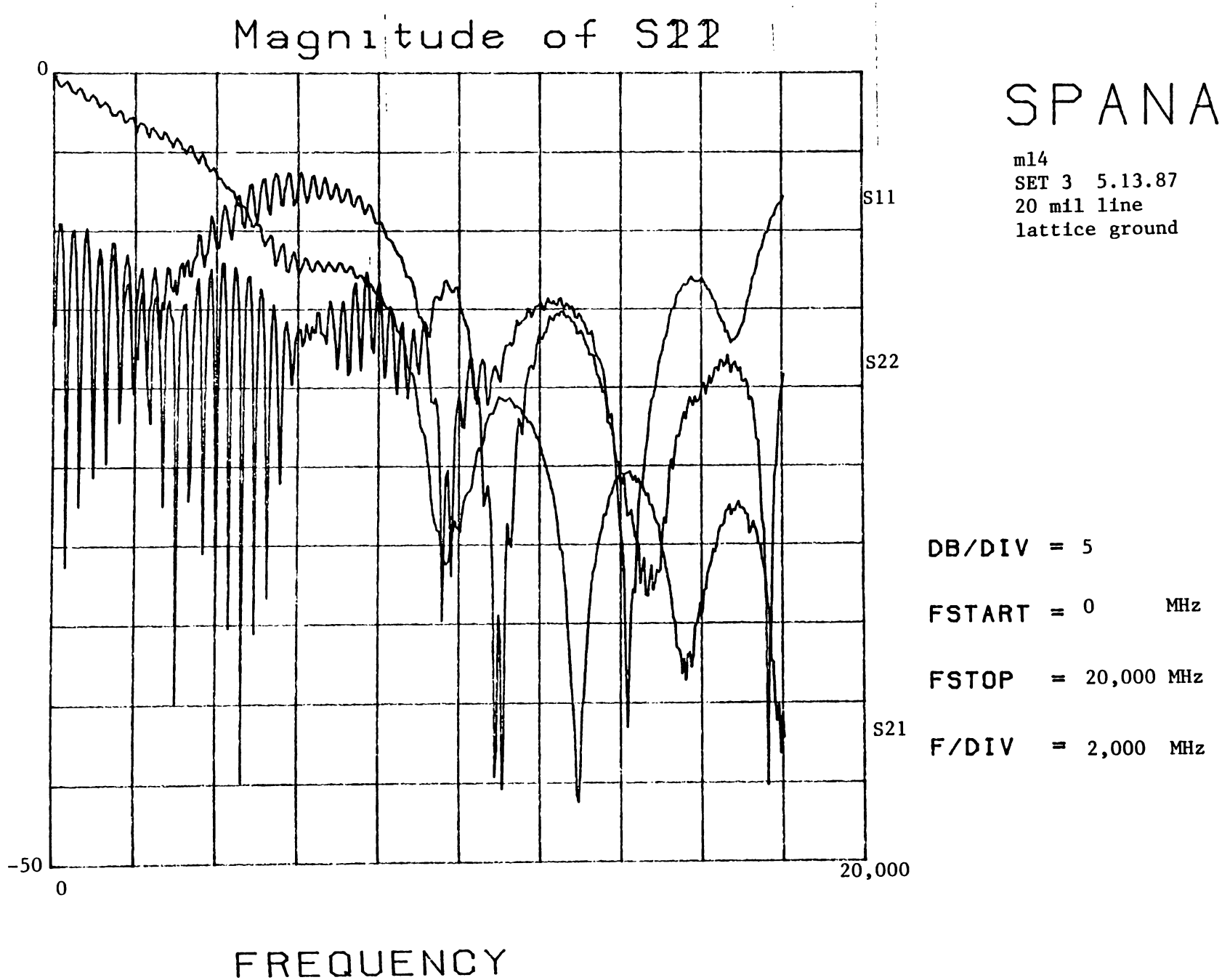

Magnitude of S22
0
-50
100
0
20,000
FREQUENCY
S11
S22
S21
SPANA
msb6
SET 3 5.13.87
13 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

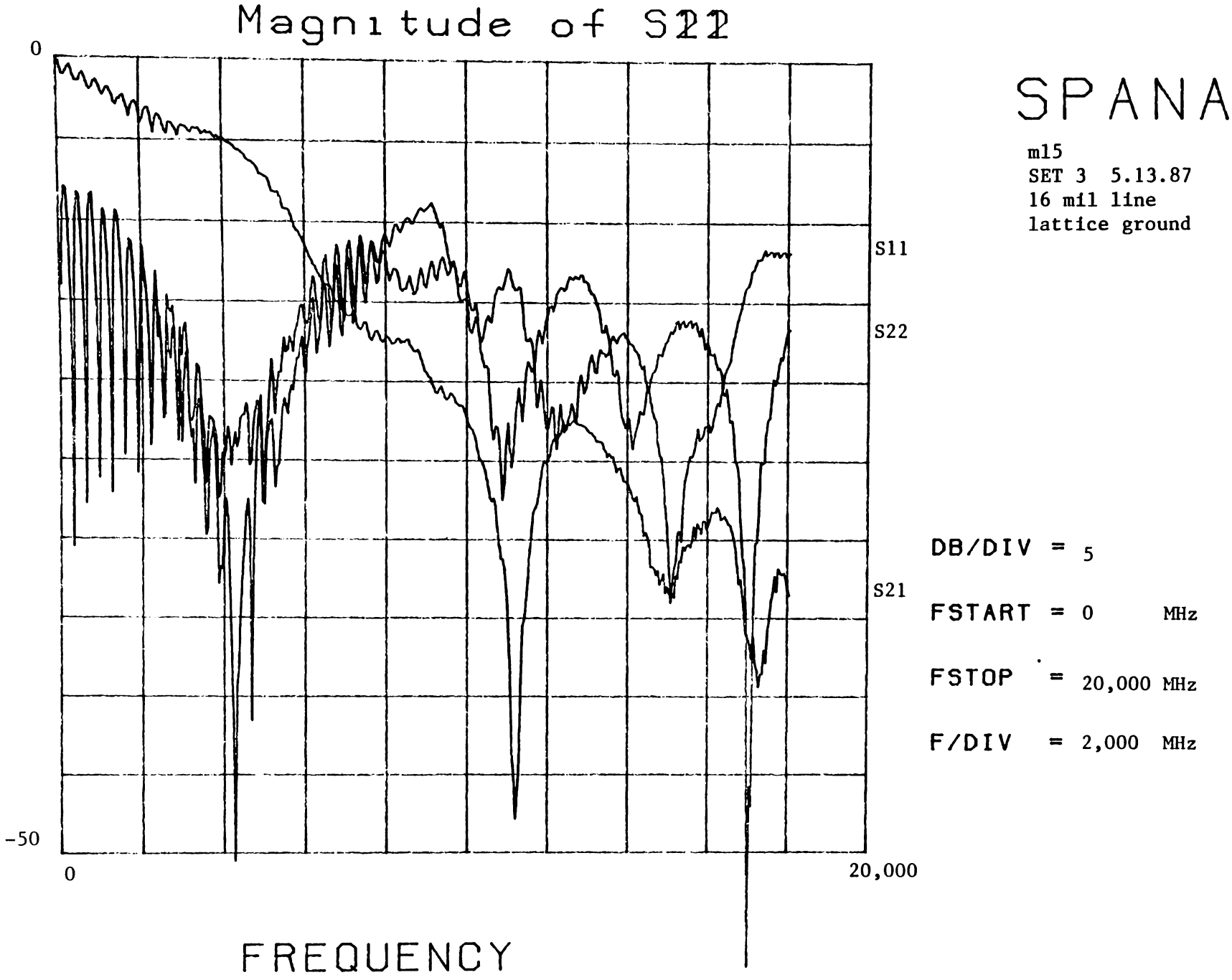

Magnitude of S22
0
-50
0
20,000
S22
S11
S21
FREQUENCY
SPANA
msb7
SET 3 5.13.87
8 mil line
solid ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

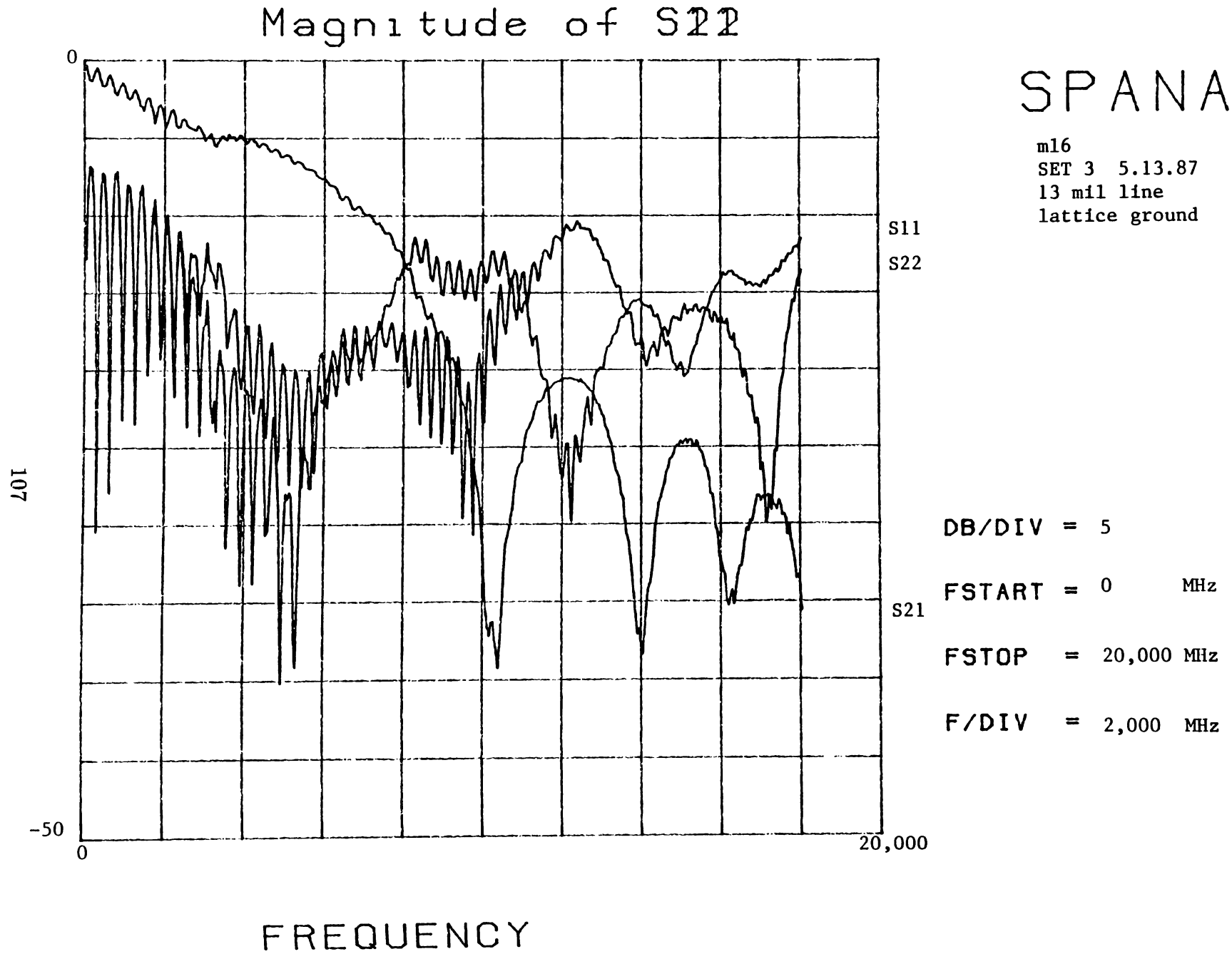

Magnitude of S21
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
m11
SET 3 5.13.87
100 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

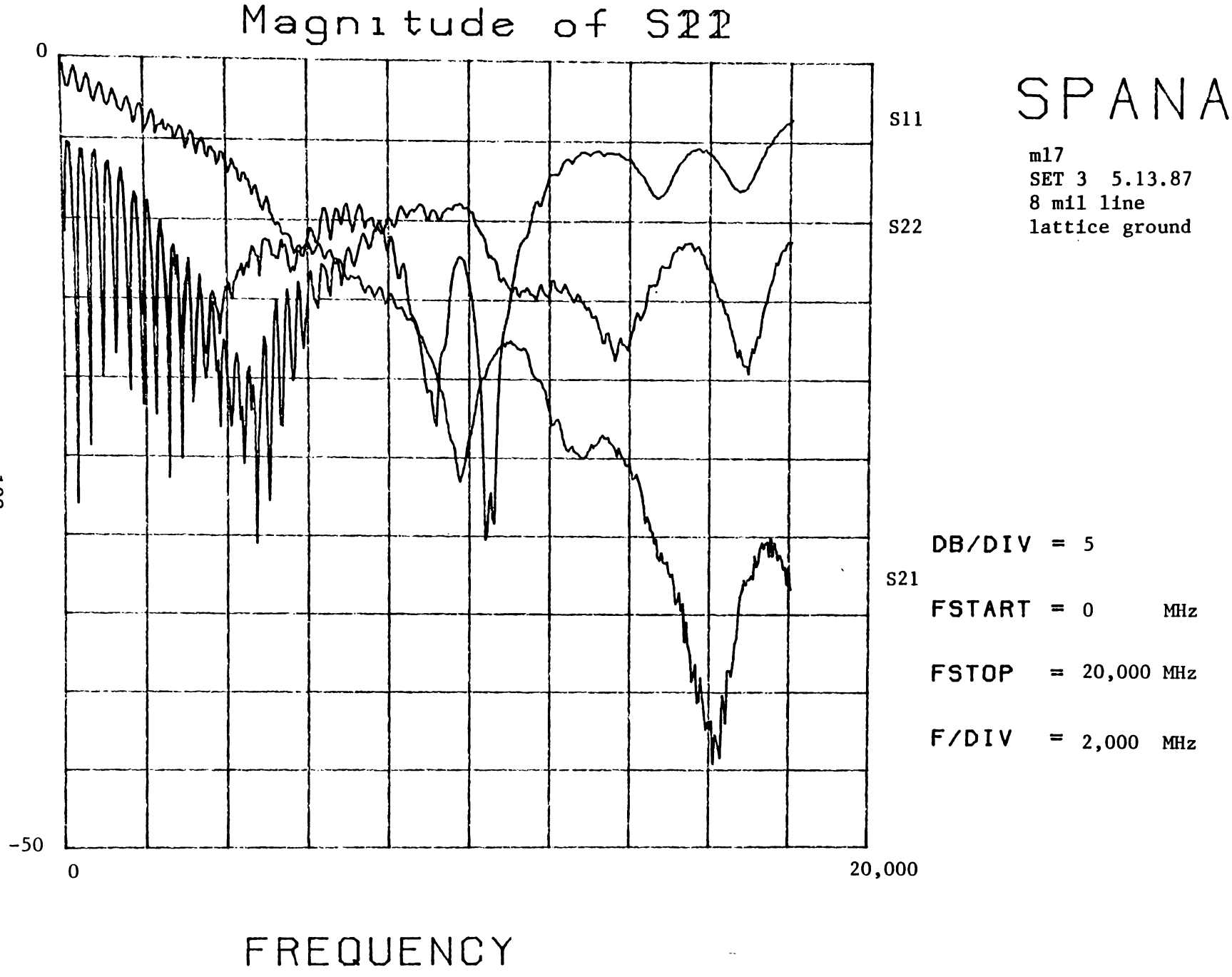

Magnitude of S21
0
-50
0
20,000
FREQUENCY
S22
S11
S21
SPANA
m12
SET 3 5.13.87
50 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

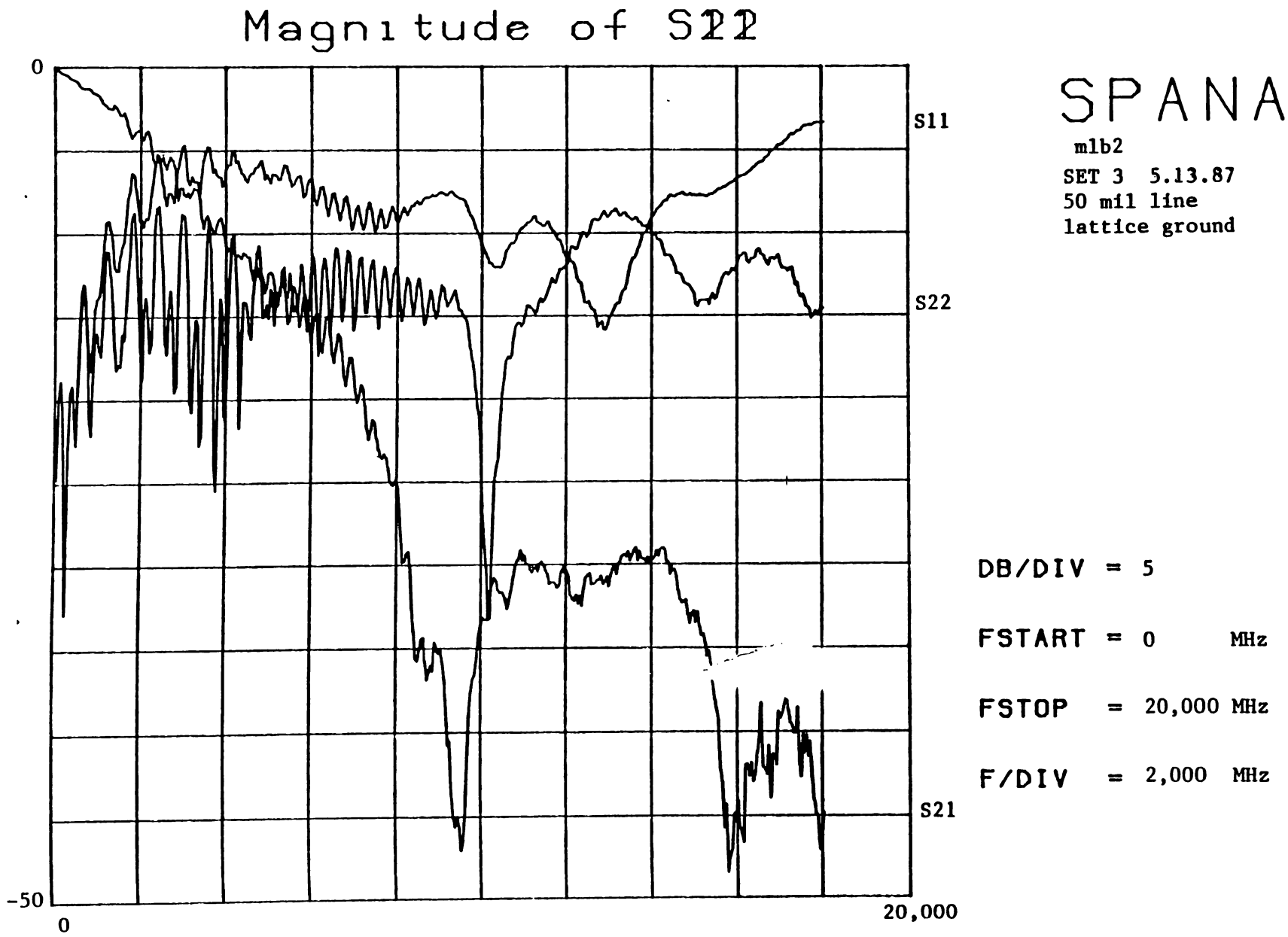

Magnitude of S22
0
-50
0
20,000
S22
S11
S21
FREQUENCY
SPANA
m13
SET 3 5.13.87
25 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

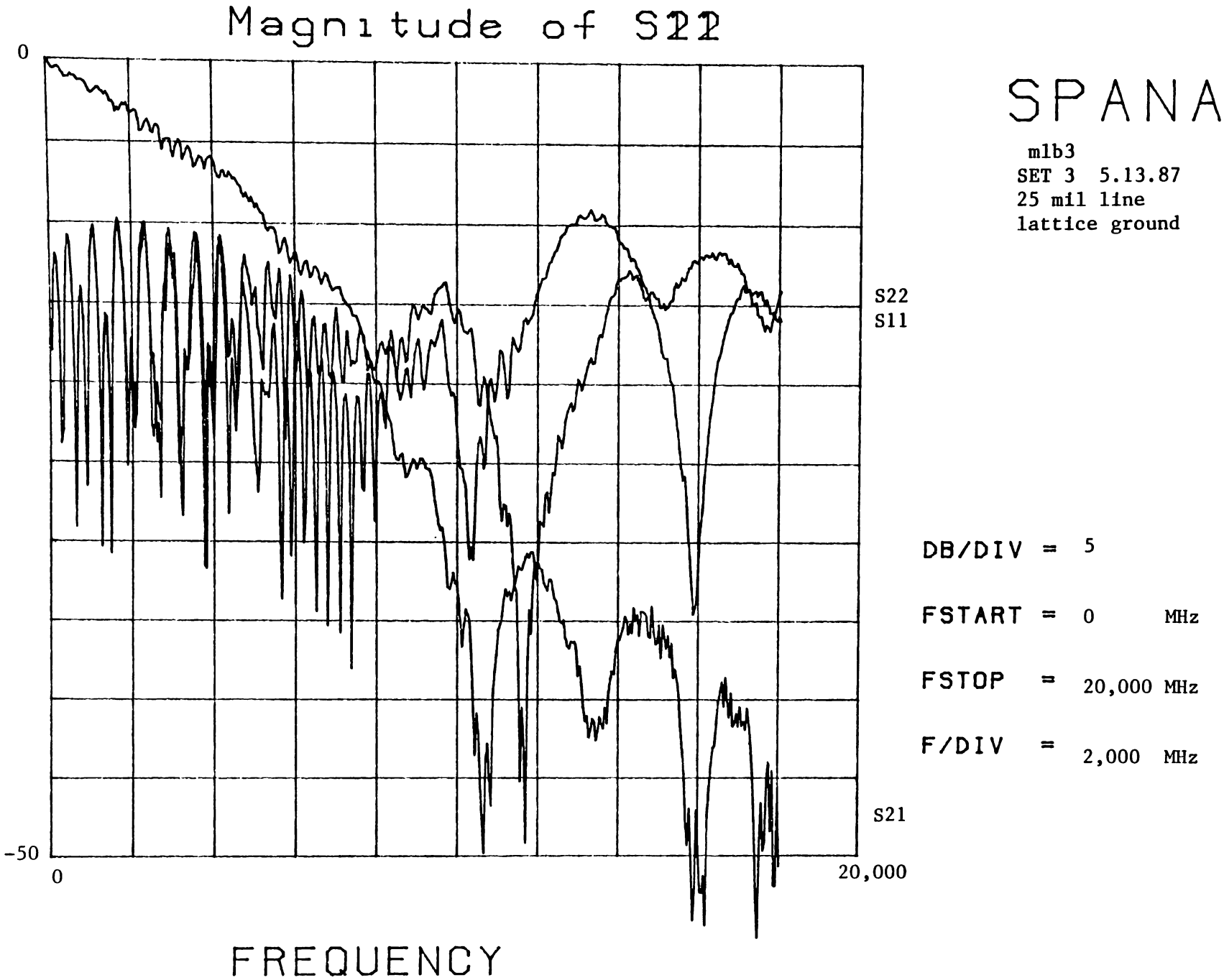

Magnitude of S11
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
m14
SET 3 5.13.87
20 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

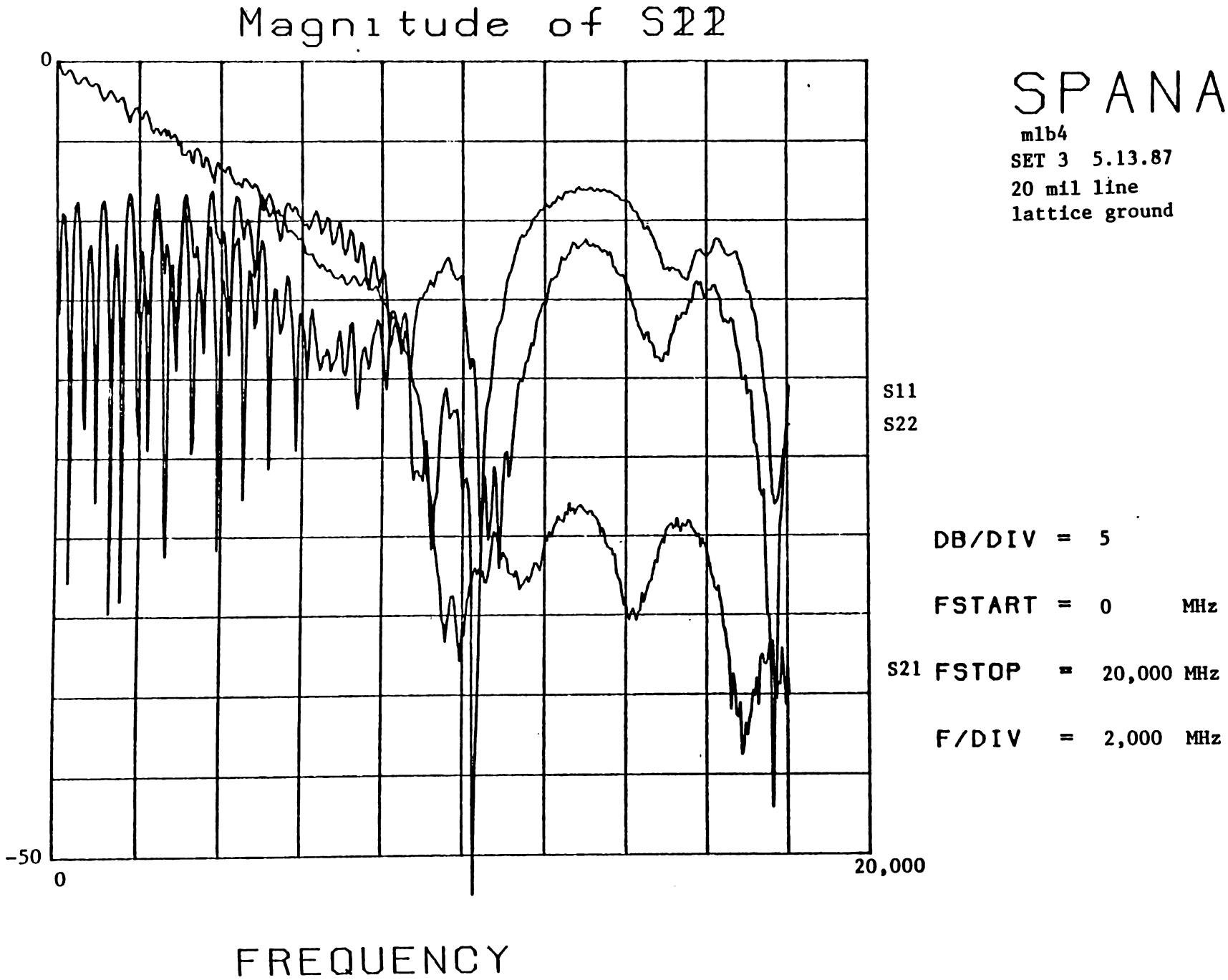

Magnitude of S11
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
m15
SET 3 5.13.87
16 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

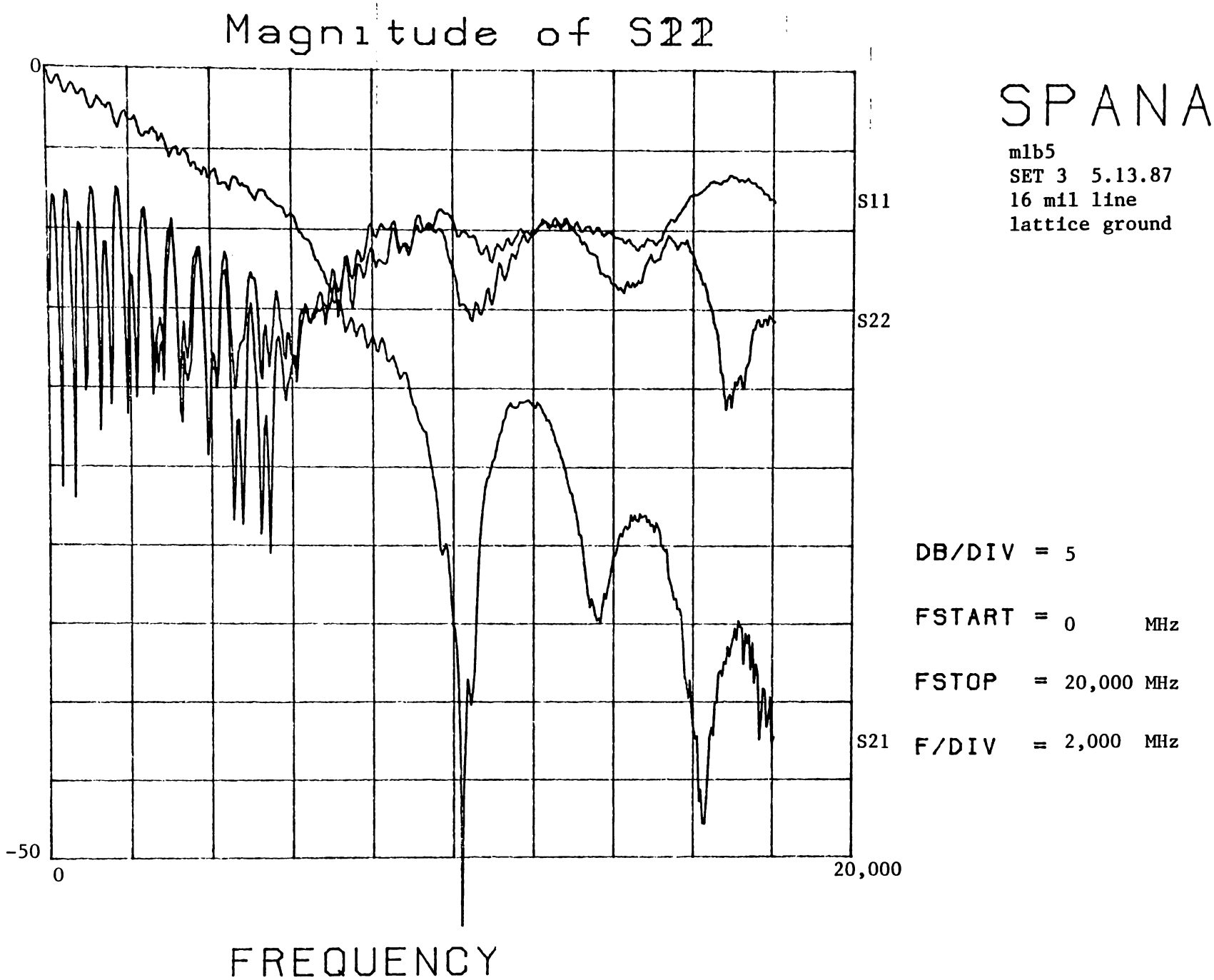

Magnitude of S11
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
m16
SET 3 5.13.87
13 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

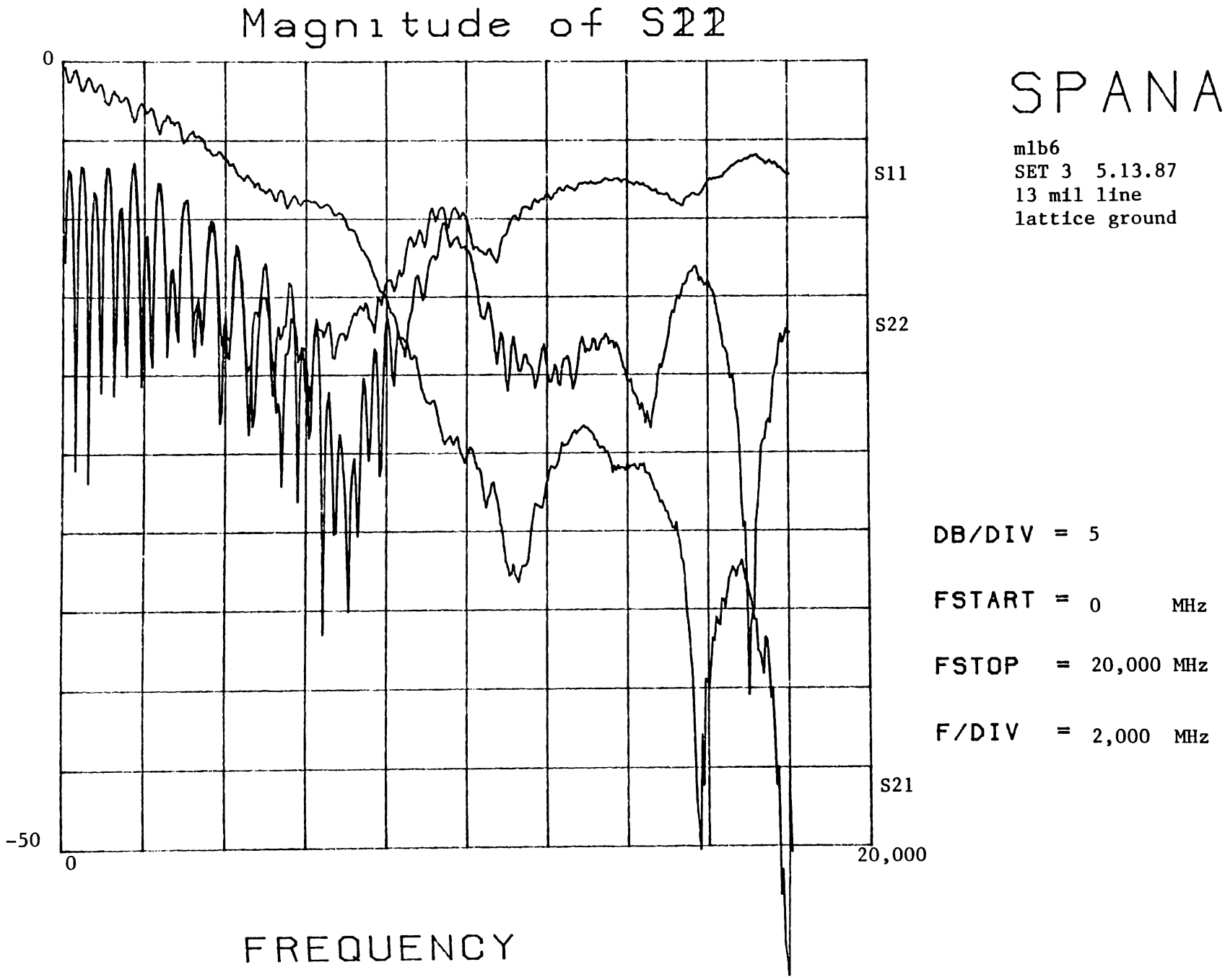

Magnitude of S11
0
-50
0
20,000
FREQUENCY
S11
S22
S21
SPANA
m17
SET 3 5.13.87
8 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

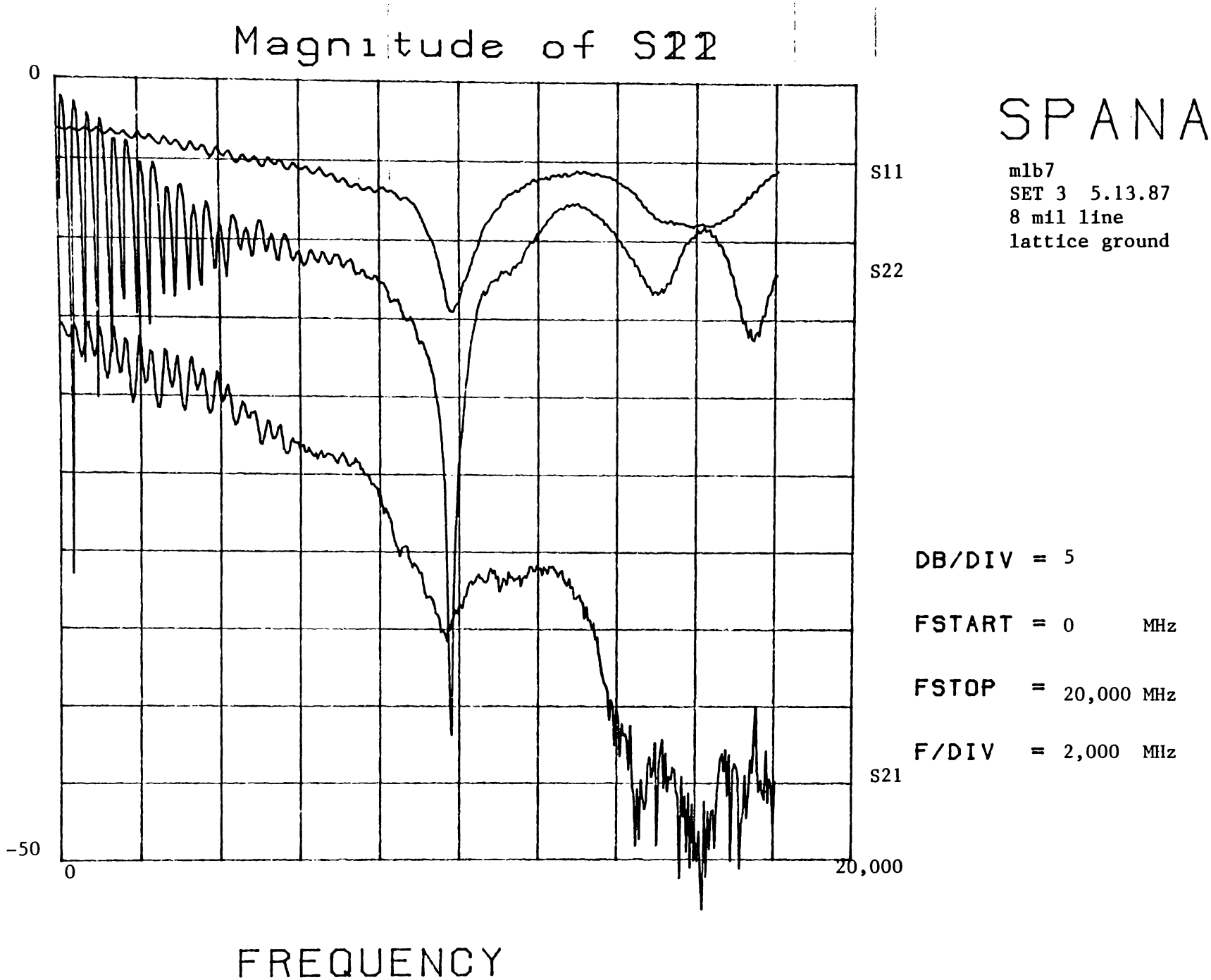

Magnitude of S11
0
-50
0
20,000
S11
S22
S21
SPANA
mlb2
SET 3 5.13.87
50 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

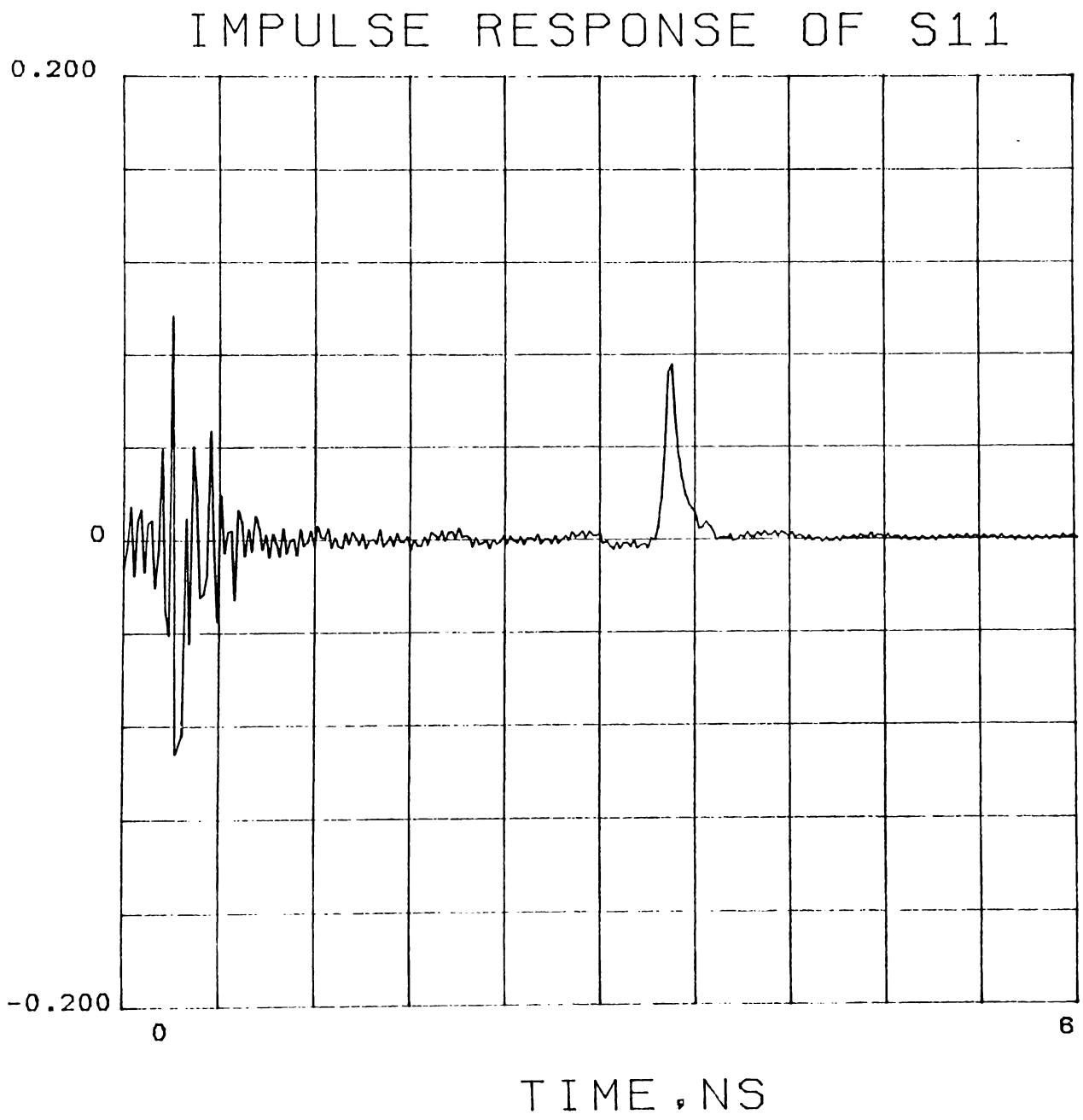

Magnitude of S22
0
-50
0
20,000
FREQUENCY
S22
S11
S21
SPANA
mlb3
SET 3 5.13.87
25 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz


Magnitude of S11
SPANA
mlb4
SET 3 5.13.87
20 mil line
lattice ground
0
-50
0
20,000
S11
S22
S21
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz
FREQUENCY


Magnitude of S11
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
mlb5
SET 3 5.13.87
16 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz


Magnitude of S22
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
m1b6
SET 3 5.13.87
13 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz

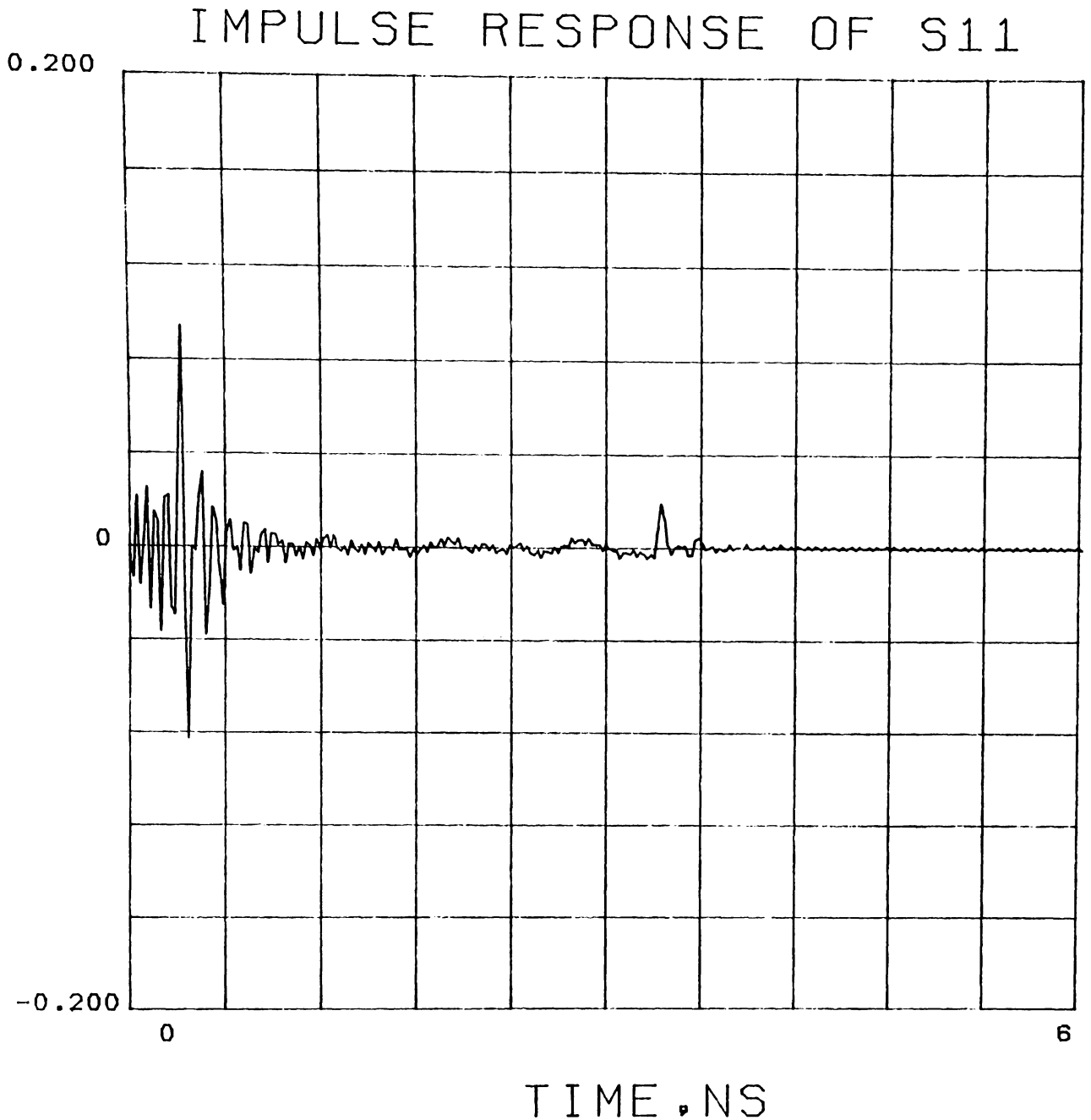

Magnitude of S11
0
-50
0
20,000
S11
S22
S21
FREQUENCY
SPANA
mlb7
SET 3 5.13.87
8 mil line
lattice ground
DB/DIV = 5
FSTART = 0 MHz
FSTOP = 20,000 MHz
F/DIV = 2,000 MHz



SPANA

MEAS FILE FOR MS2
SOLID GROUND
50 MIL LINE
SET_3_5.13.87

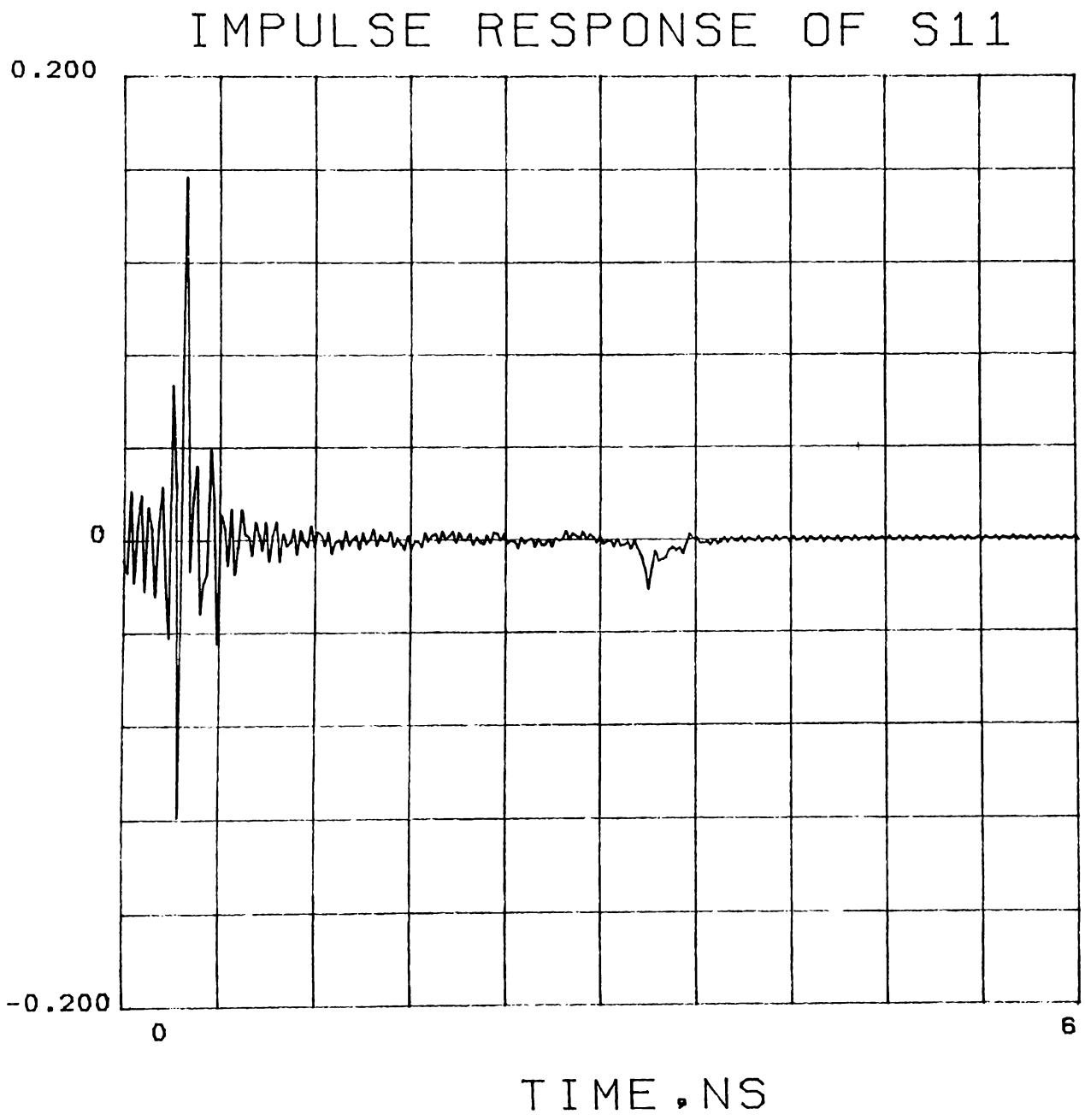
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MS4
SOLID GROUND
20 MIL LINE
SET_3_5.13.87

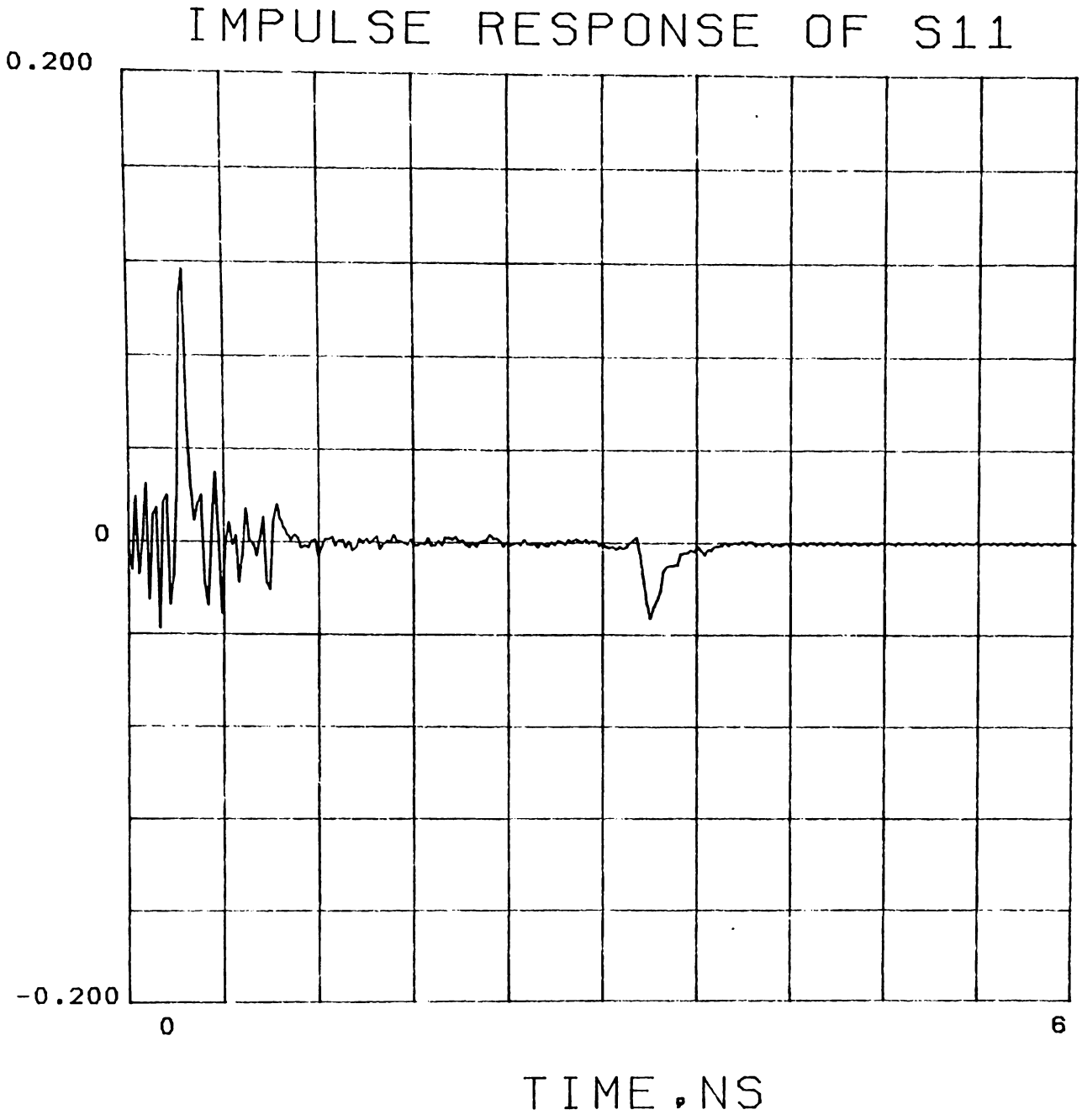
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MS6
SOLID GROUND
13 MIL LINE
SET_3_5.13.87

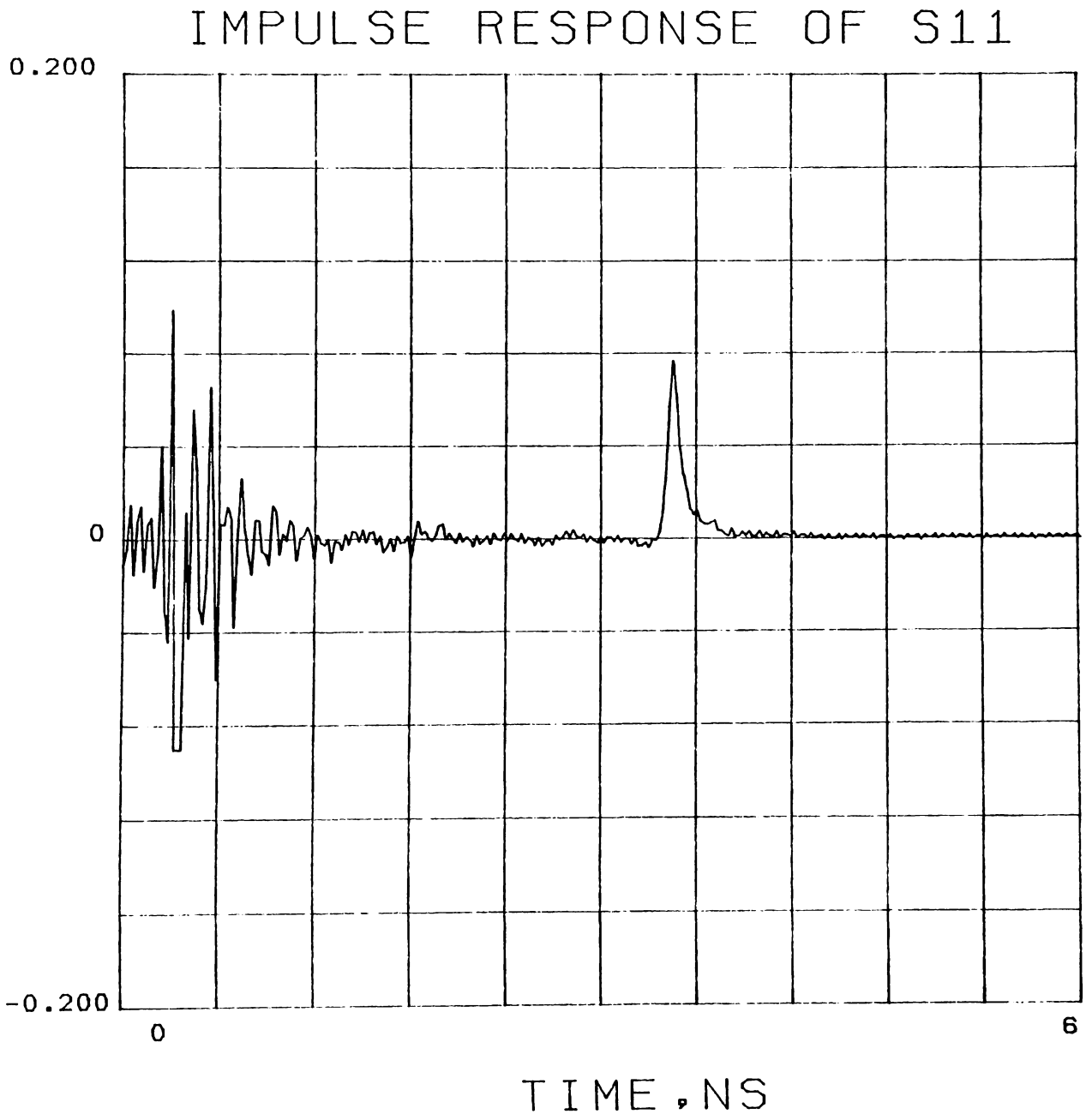
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MS7
SOLID GROUND
8 MIL LINE
SET_3_5.13.87

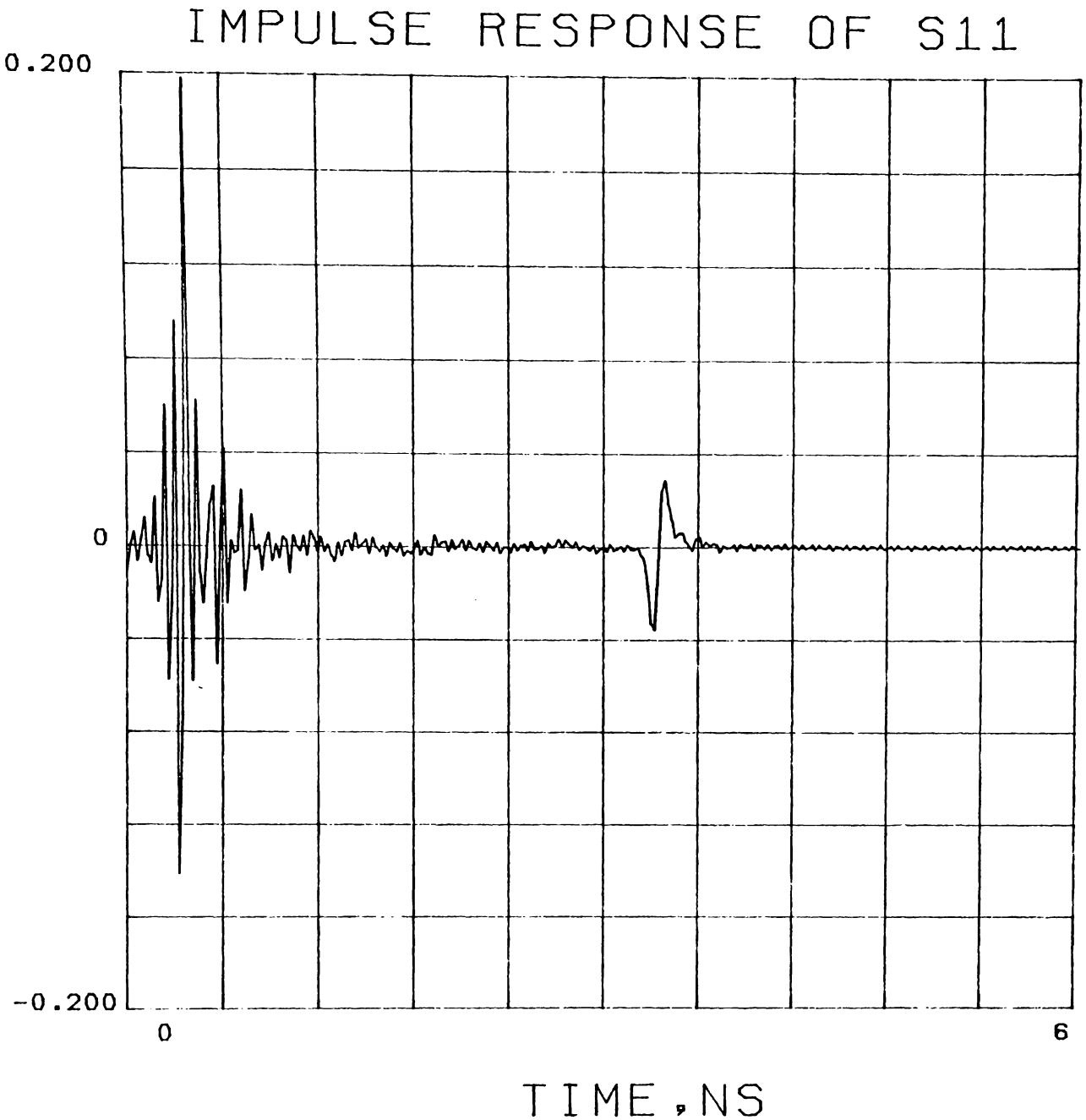
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MSB2
SOLID GROUND
50 MIL LINE
SET_3_5.13.87

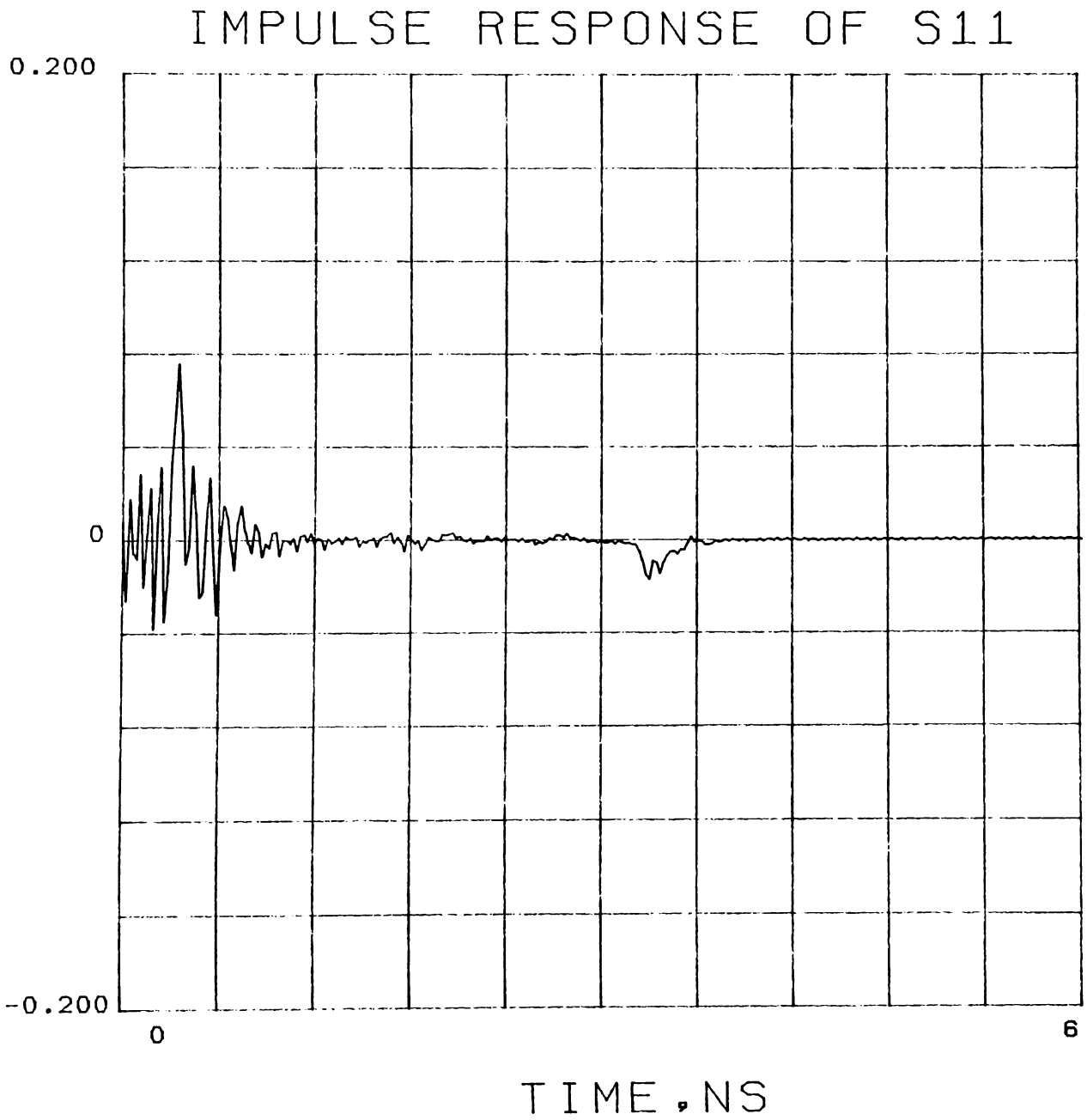
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MSB4
SOLID GROUND
20 MIL LINE
SET_3_5.13.87

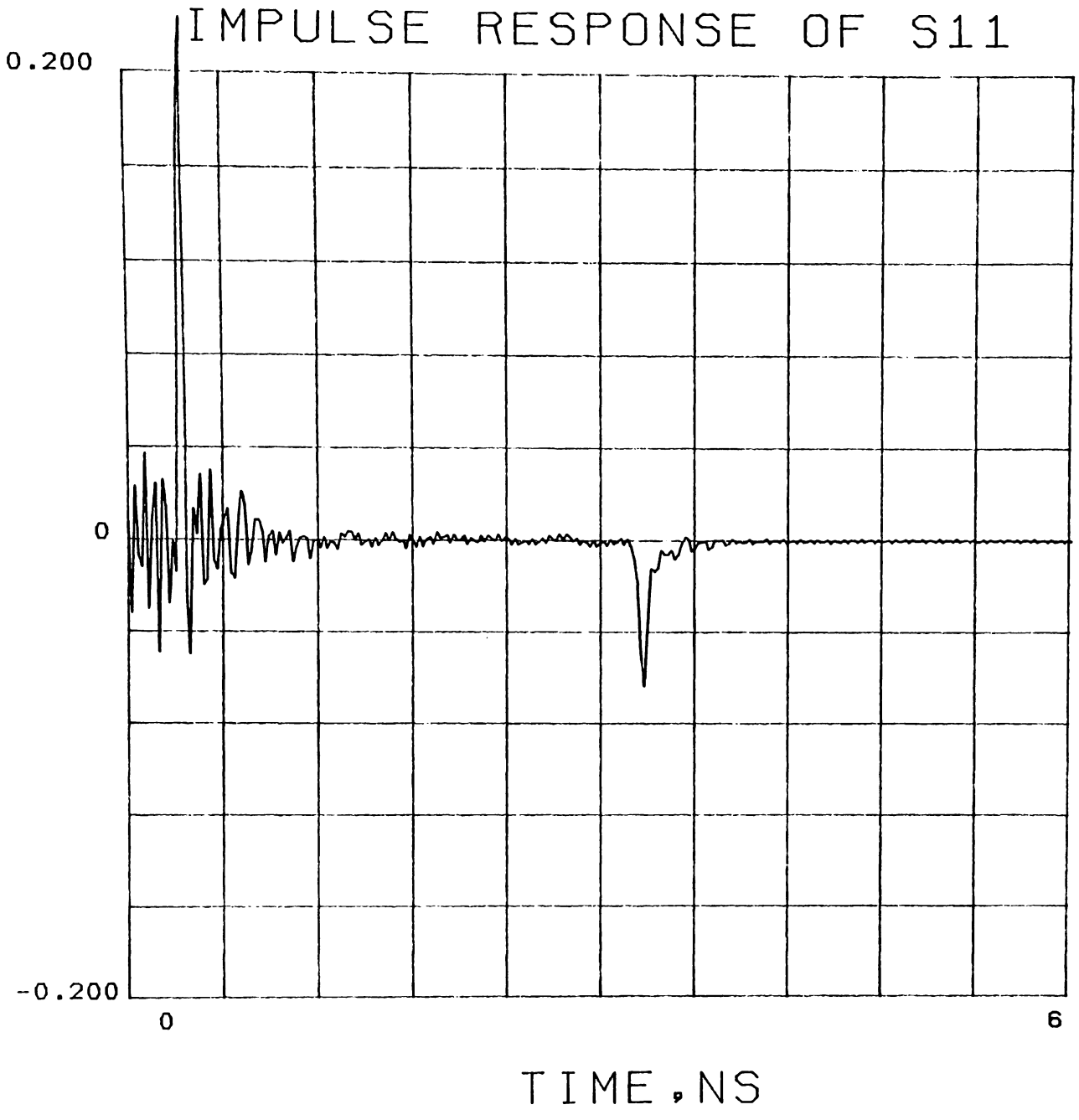
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MSB6
SOLID GROUND
13 MIL LINE
SET_3_5.13.87

/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MSB7
SOLID GROUND
8 MIL LINE
SET_3_5.13.87

/DIV = 0.04
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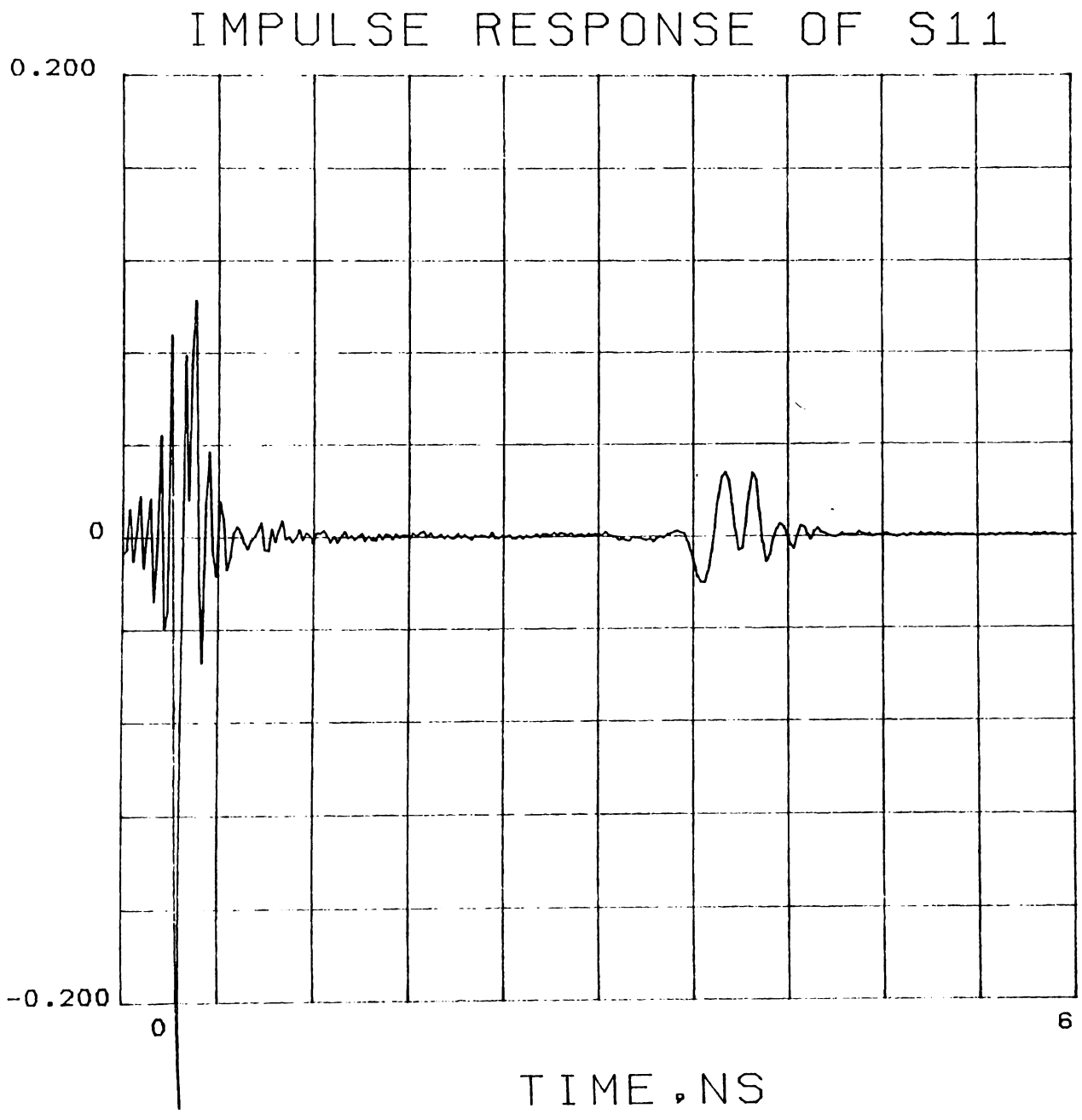
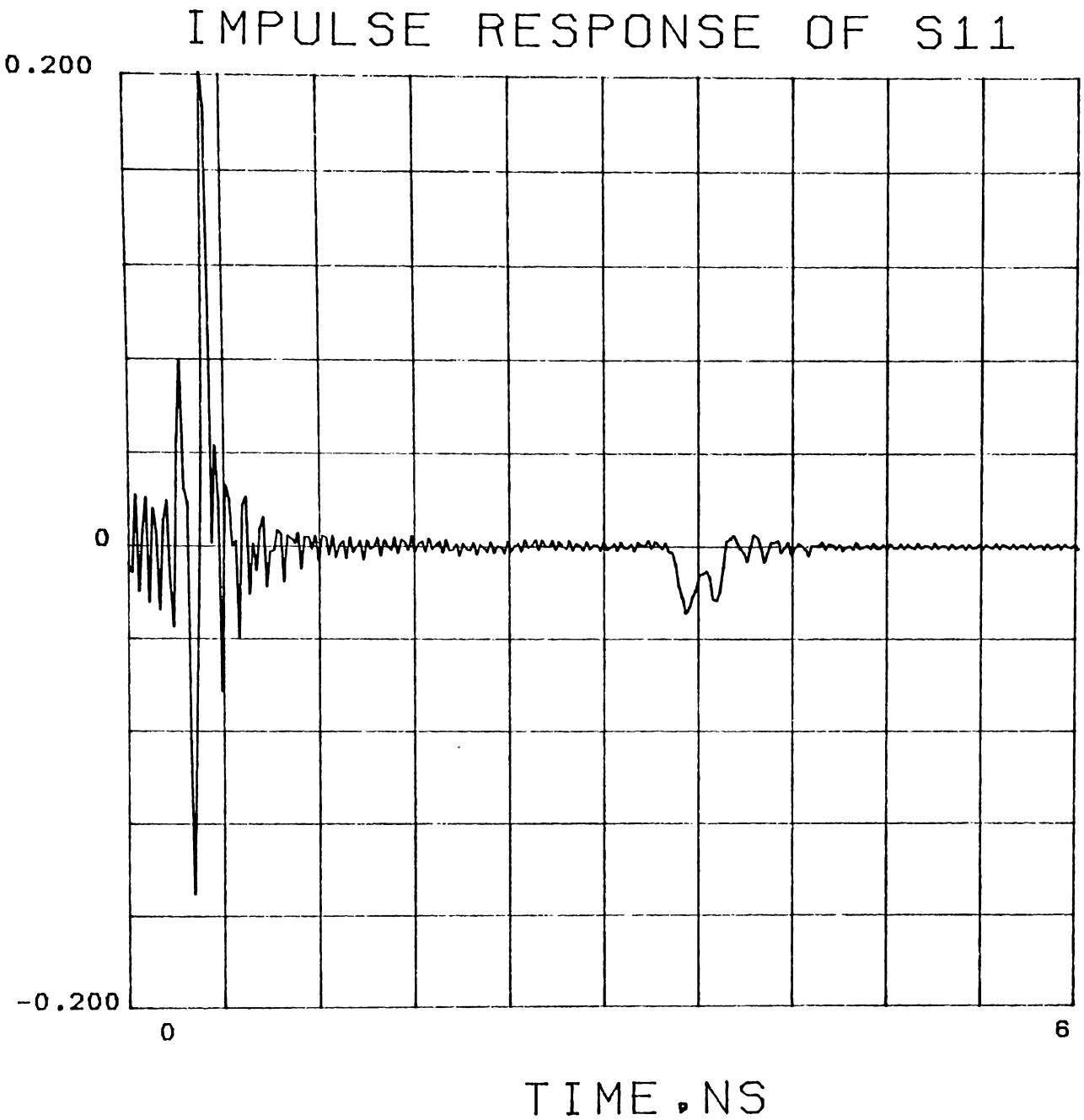


SPANA

MEAS FILE FOR ML2
LATTICE GROUND
50 MIL LINE
SET_3_5.13.87

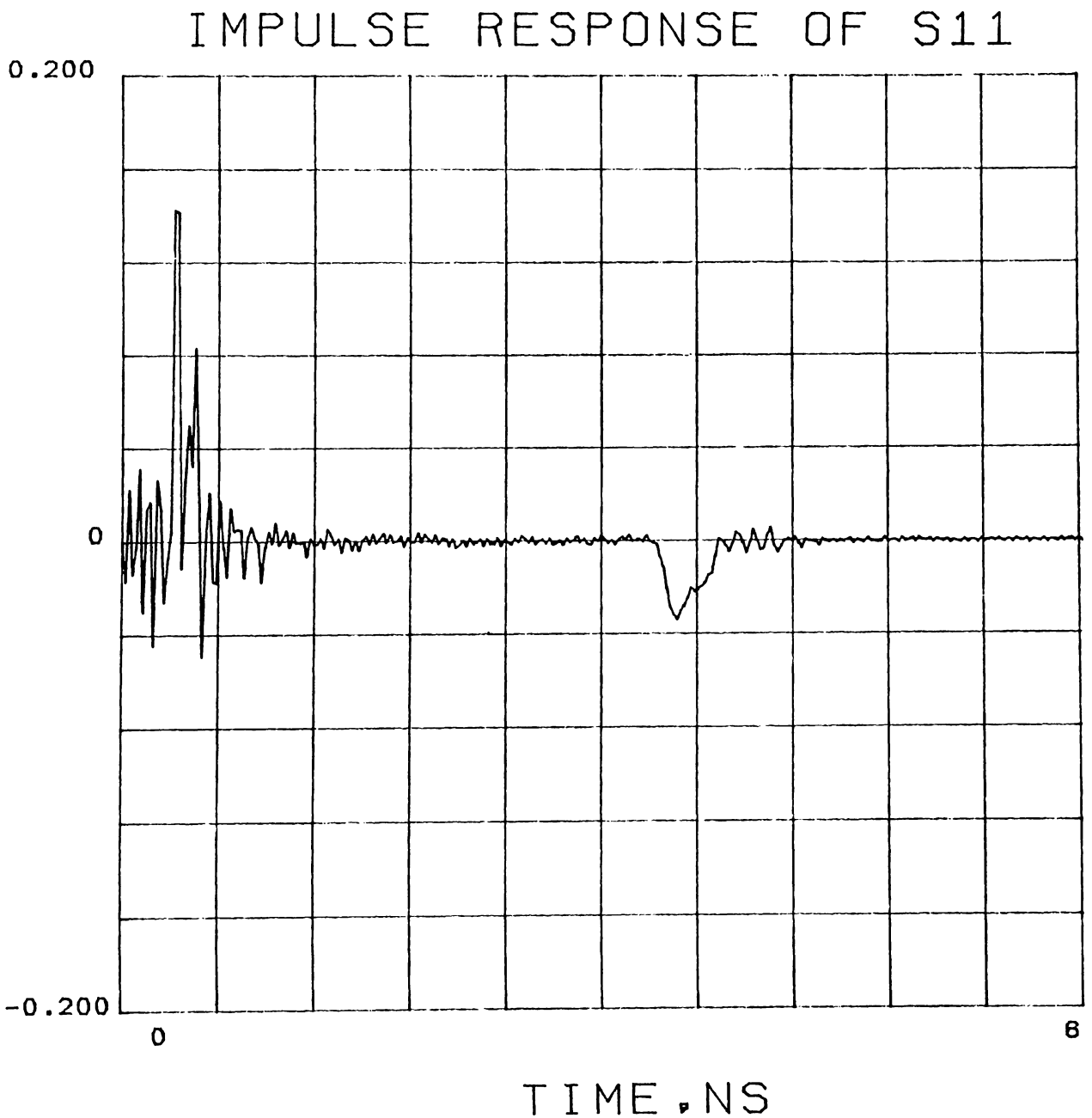
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR ML4
LATTICE GROUND
20 MIL LINE
SET_3_5.13.87

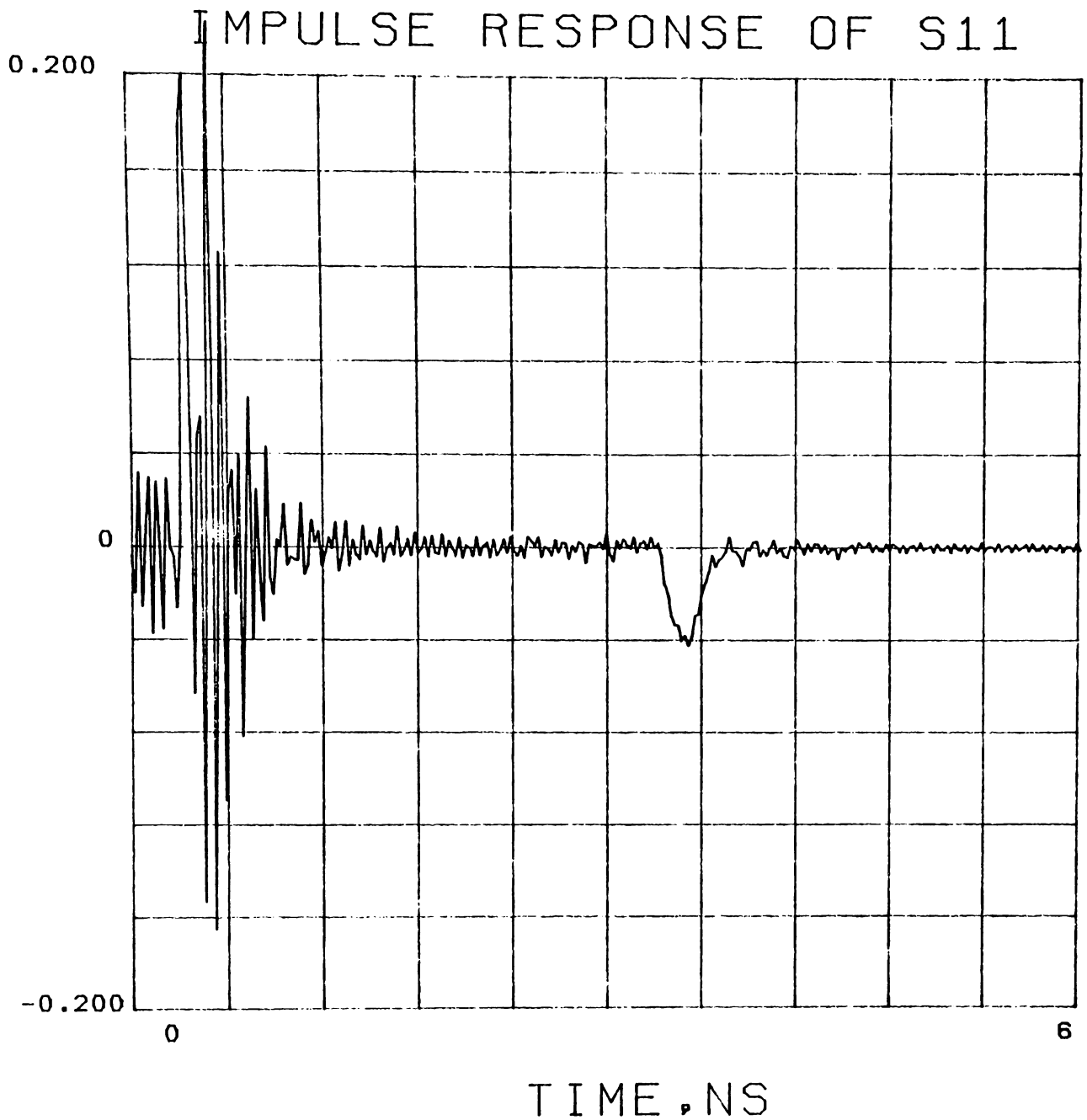
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR ML6
LATTICE GROUND
13 MIL LINE
SET_3_5.13.87

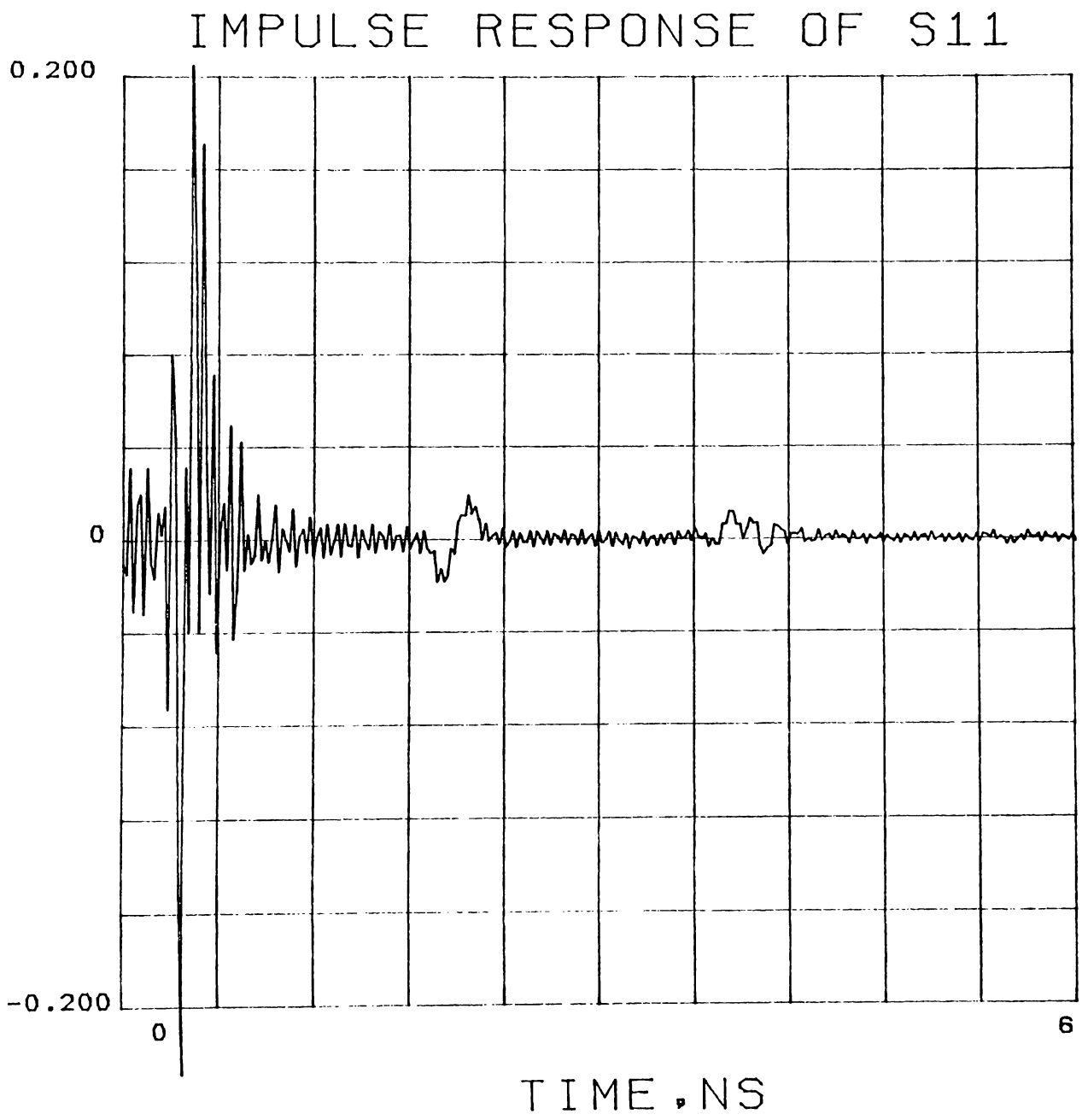
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR ML7
LATTICE GROUND
8 MIL LINE
SET_3_5.13.87

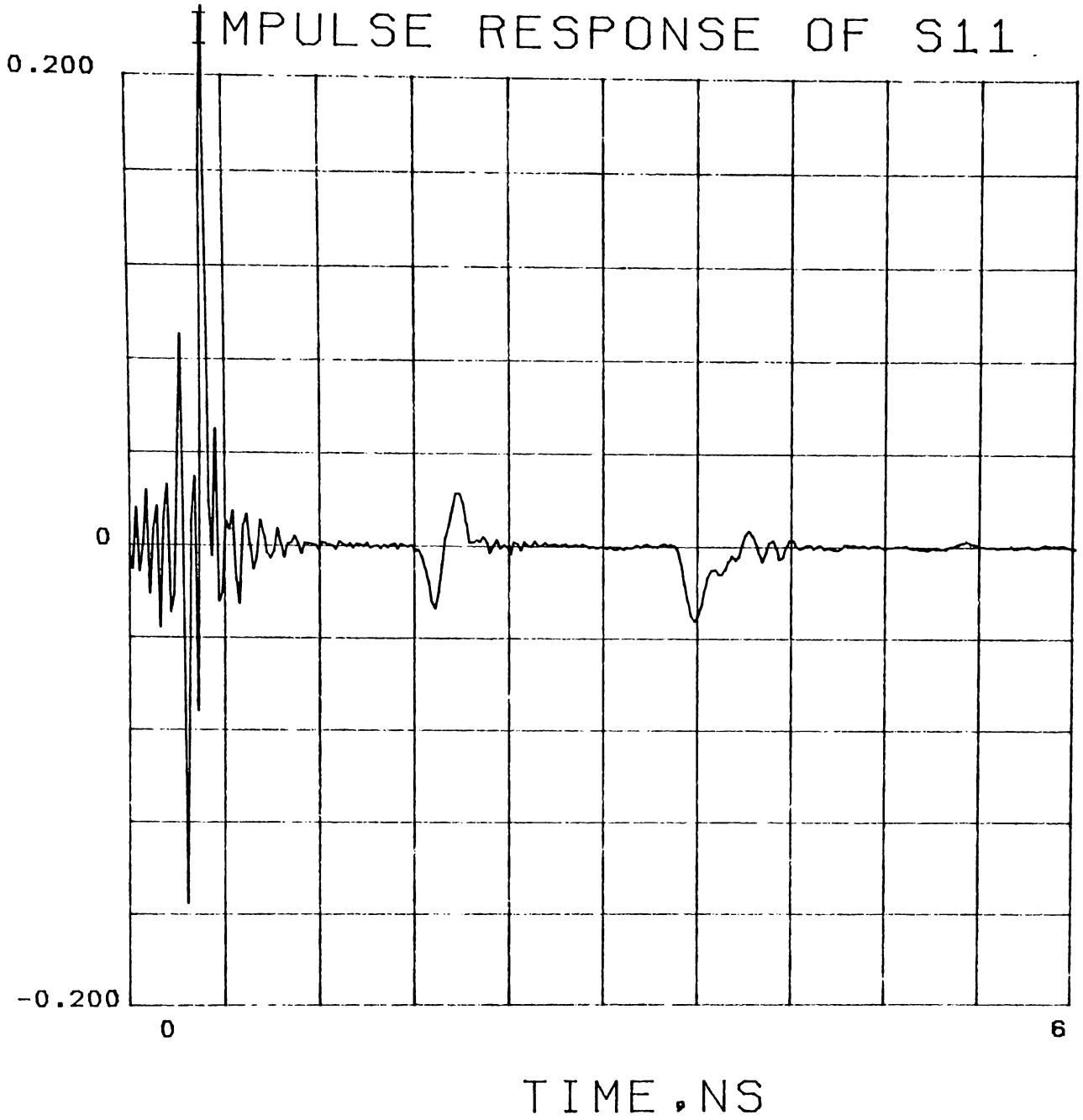
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MLB2
LATTICE GROUND
50 MIL LINE
SET_3_5.13.87

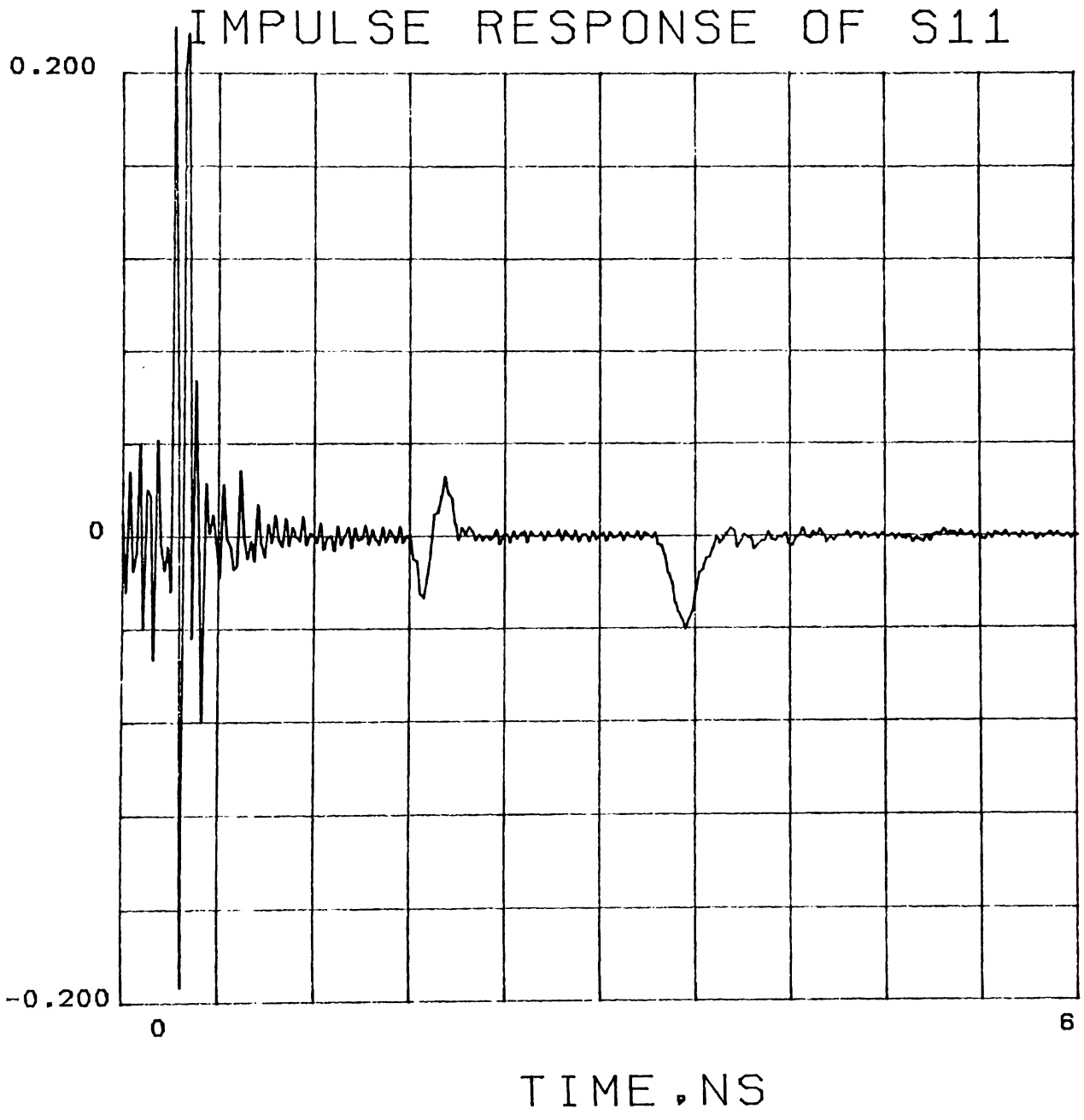
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MLB4
LATTICE GROUND
20 MIL LINE
SET_3_5.13.87

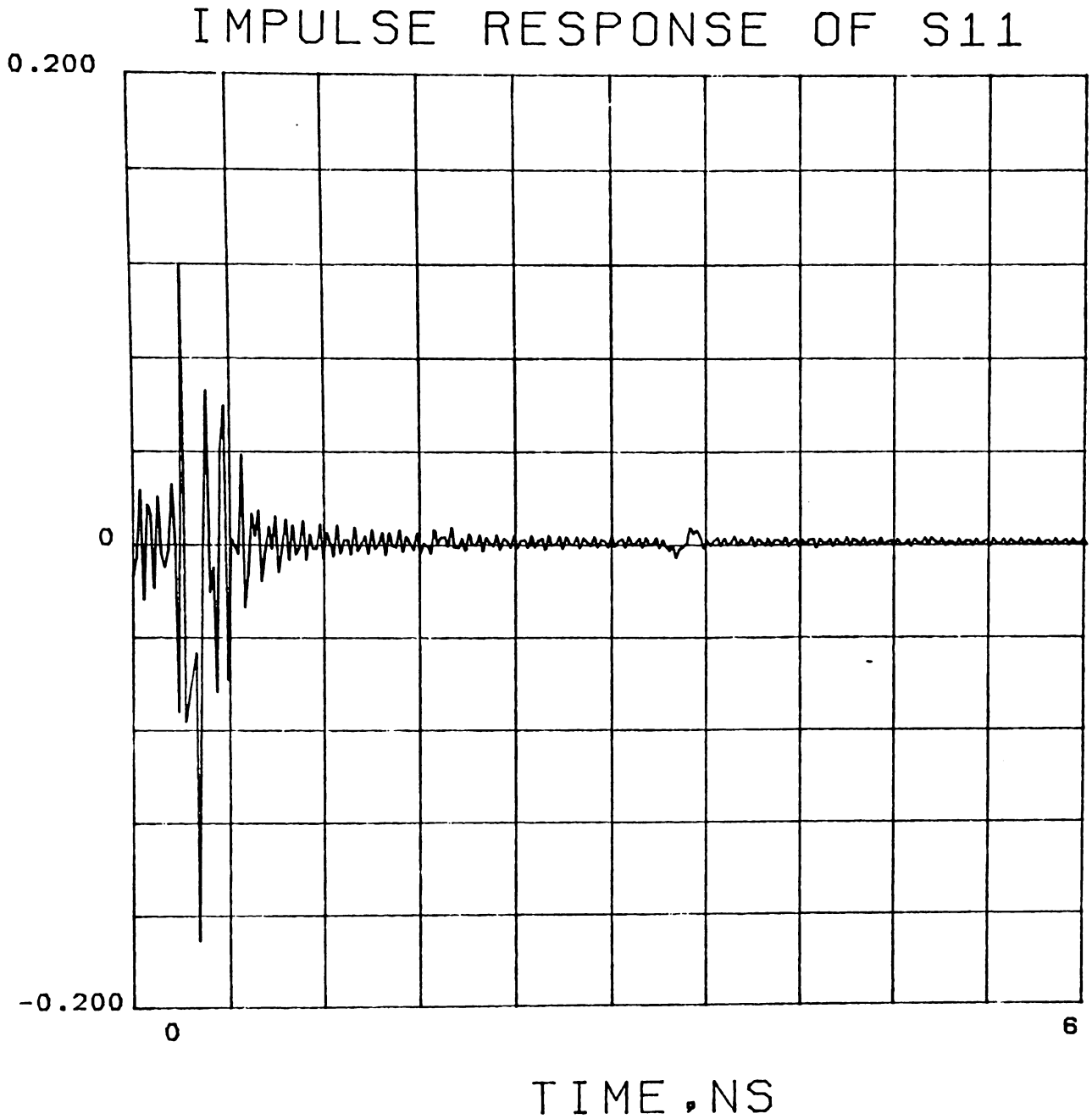
/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MLB6
LATTICE GROUND
13 MIL LINE
SET_3_5.13.87

/DIV = 0.04
NS/DIV = 0.60



SPANA

MEAS FILE FOR MLB7
LATTICE GROUND
8 MIL LINE
SET_3_5.13.87

/DIV = 0.04
NS/DIV = 0.60