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The prediction of crosstalk in ribbon cables is investigated. Experimental results are obtained for a 20 wire ribbon cable and compared to the predictions of the multiconductor transmission line (MTL) model. Based on the experimental configuration tested, it would appear that accurate predictions of crosstalk can be achieved in controlled characteristic cables such as these. The prediction accuracies are typically within $\pm$ 1 dB for frequencies such that the line is electrically short ($\mathcal{L} < \frac{1}{10}\lambda$) and $\pm$ 6 dB for frequencies such that the line is

electrically long ($\mathcal{L} > \frac{1}{10}\lambda$). It was found that the parasitic wires in the cable can have a significant effect (as much as 40 dB for all frequencies) on the coupling between a generator circuit and a receptor circuit in the cable. Therefore to achieve accurate predictions in ribbon cables, one must consider the interactions between all wires in the cable. The wire insulation evidently can be ignored when the line is electrically short but cannot be ignored for higher frequencies. Conversely, the impedance of the reference wire cannot be ignored for low frequencies where the common impedance coupling dominates the electromagnetic field coupling.

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PREFACE

This effort was conducted by the University of Kentucky under the sponsorship of the Rome Air Development Center Post-Doctoral Program for RADC's Compatibility Branch. Mr. Jim Brodock of RADC was the task project engineer and provided overall technical direction and guidance.

The RADC Post-Doctoral Program is a cooperative venture between RADC and some sixty-five universities eligible to participate in the program. Syracuse University (Department of Electrical Engineering), Purdue University (School of Electrical Engineering), Georgia Institute of Technology (School of Electrical Engineering), and State University of New York at Buffalo (Department of Electrical Engineering) act as prime contractor schools with other schools participating via sub-contracts with the prime schools. The U.S. Air Force Academy (Department of Electrical Engineering), Air Force Institute of Technology (Department of Electrical Engineering), and the Naval Post Graduate School (Department of Electrical Engineering) also participate in the program.

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Further information about the RADC Post-Doctoral Program can be obtained from Mr. Jacob Scherer, RADC/RBC, Griffiss AFB, NY, 13441, telephone Autovon 587-2543, commercial (315) 330-2543.

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I. INTRODUCTION

Ribbon cables are groups of wires (cylindrical conductors with cylindrical, dielectric insulations) which are bonded together in a flat, linear array. These types of cables are sometimes referred to as flatpack cables although the flatpack designation includes a larger number of types of flat cable, e.g., the conductors may be rectangular rather than cylindrical. In other words, all ribbon cables are flatpack cables but not all flatpack cables are ribbon cables. All wires in the ribbon cable are identical and the spacing of the wires is carefully controlled with the separation between all adjacent wires being identical. A typical 40 wire ribbon cable is shown in Figure 1-1. All wires in a ribbon cable are typically identical, i.e., equal conductor radii, equal dielectric thicknesses and identical dielectric insulations. A typical cross-section of a ribbon cable is shown in Figure 1-2. The conductors are characterized by radii r_w, free space permittivity, ε_v, free space permeability, μ_v, and conductivity, σ. The dielectric insulations are characterized by thicknesses, t, permittivity, ε, and free space permeability (as is typical of dielectrics), μ_v. The dielectrics will be considered to be lossless, i.e., their conductivities are zero. The separations between all adjacent wires is denoted by d and the cable is surrounded by free space.

Ribbon cables are frequently being used to interconnect electronic systems such as minicomputers. The unintentional coupling of electrical signals from one wire to another (crosstalk) in these cables can degrade the performance of the electronic devices which are connected to the cable at its end points. For example, logic errors can occur in minicomputers by the inadvertent coupling of a logic bit into an unintended circuit. In the design

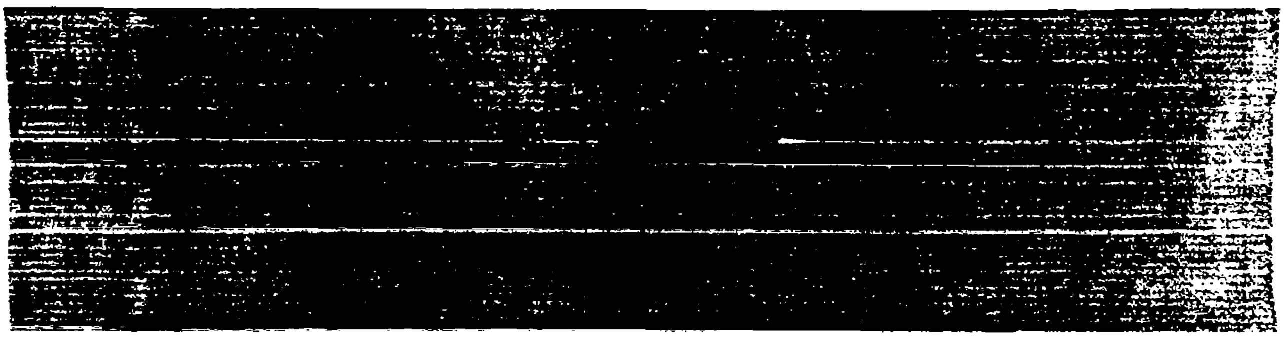


Figure 1-1. A typical 40 wire ribbon cable.

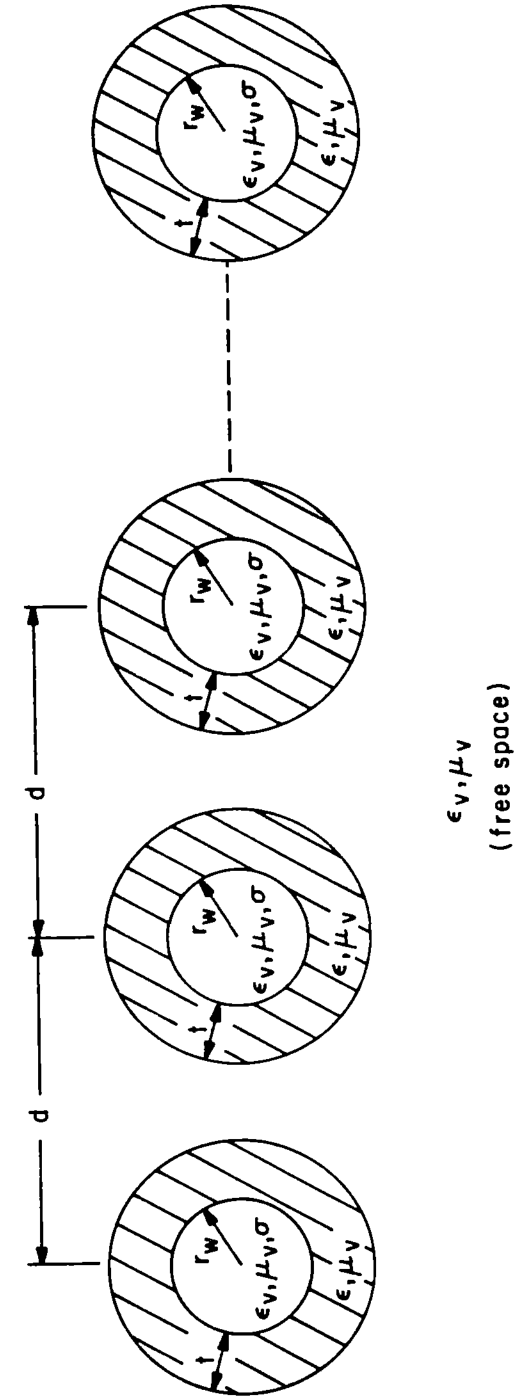


Figure 1-2. The ribbon cable cross section.

of an electronic system, it is of paramount importance to be able to predict and correct potential interference problems of this type.

In recent years, techniques for characterizing this type of cable as a multiconductor transmission line, (MTL),have been developed [1-8]. With the aid of matrix analysis, these techniques are straightforward extensions of the familiar results for two-wire lines to multiconductor lines. The multiconductor transmission line (MTL) model as specialized to ribbon cables will be described in Chapter II.

The purpose of this report is to compare experimental results to predictions of the MTL model to determine whether accurate predictions of crosstalk can be made for these controlled characteristics cables. In [8], it was shown that for random cable bundles in which relative wire separation along the cable is unknown and uncontrolled, accurate predictions of crosstalk could not be consistently achieved. Therefore, it is of interest here to determine if it is worthwhile to expend the computational effort required by the MTL model in attempting to predict crosstalk in ribbon cables.

A series of digital computer programs which implement the MTL model are described in [7] and will be used to provide the computed results. Four programs are described in [7], XTALK, XTALK2, FLATPAK, FLATPAK2. Although all fo programs implement the MTL model, each one considers or neglects certain factors such as conductor losses in order to provide an efficient computational program. XTALK neglects the presence of any wire dielectric, i.e., considers the wires to be bare, and also neglects the conductor losses, i.e., the conductors are considered to be perfect conductors. XTALK2 also neglects the presence of wire insulation but includes conductor losses. FLATPAK includes consideration of wire dielectrics as in ribbon cables but considers the conductors to be lossless. FLATPAK2 includes consideration of wire dielectrics and also includes the conductor losses. XTALK2 requires more array storage and computation time than XTALK. FLATPAK2 requires more array storage and computation time than FLATPAK. Similarly, FLATPAK requires more computational difficulty than XTALK. Therefore rather than writing one general MTL program to consider all factors, we have four programs, each of which are efficient for the specific problem being investigated. It is vitally important that the reader be intimately familiar with the factors which are included or neglected in each program, i.e., presence of wire dielectrics and consideration of conductor losses, since the computed results will be referred to by the program name so that repeated elaboration on the program capabilities will not be necessary. To assist the reader, the following table is provided which summarizes the program capabilities:

MTL Prediction Program Capabilities

Program Name	Presence of wire dielectric considered?	Conductor losses considered?
XTALK	NO	NO
XTALK2	NO	YES
FLATPAK	YES	NO
FLATPAK2	YES	YES

Note that none of the programs consider insulation dielectric losses. This seems to be a reasonable assumption and its validity will be determined when we compare the program results to the experimental results.

We will consistently have four objectives or questions to be answered:

[#1] Can crosstalk in ribbon cables be predicted and if so, what types of prediction accuracies can be expected?

[#2] Do we need to consider the presence of the dielectric insulation to achieve accurate predictions or can the wires be considered to be bare?

[#3] Do we need to consider the conductor losses or can the conductors be considered lossless?

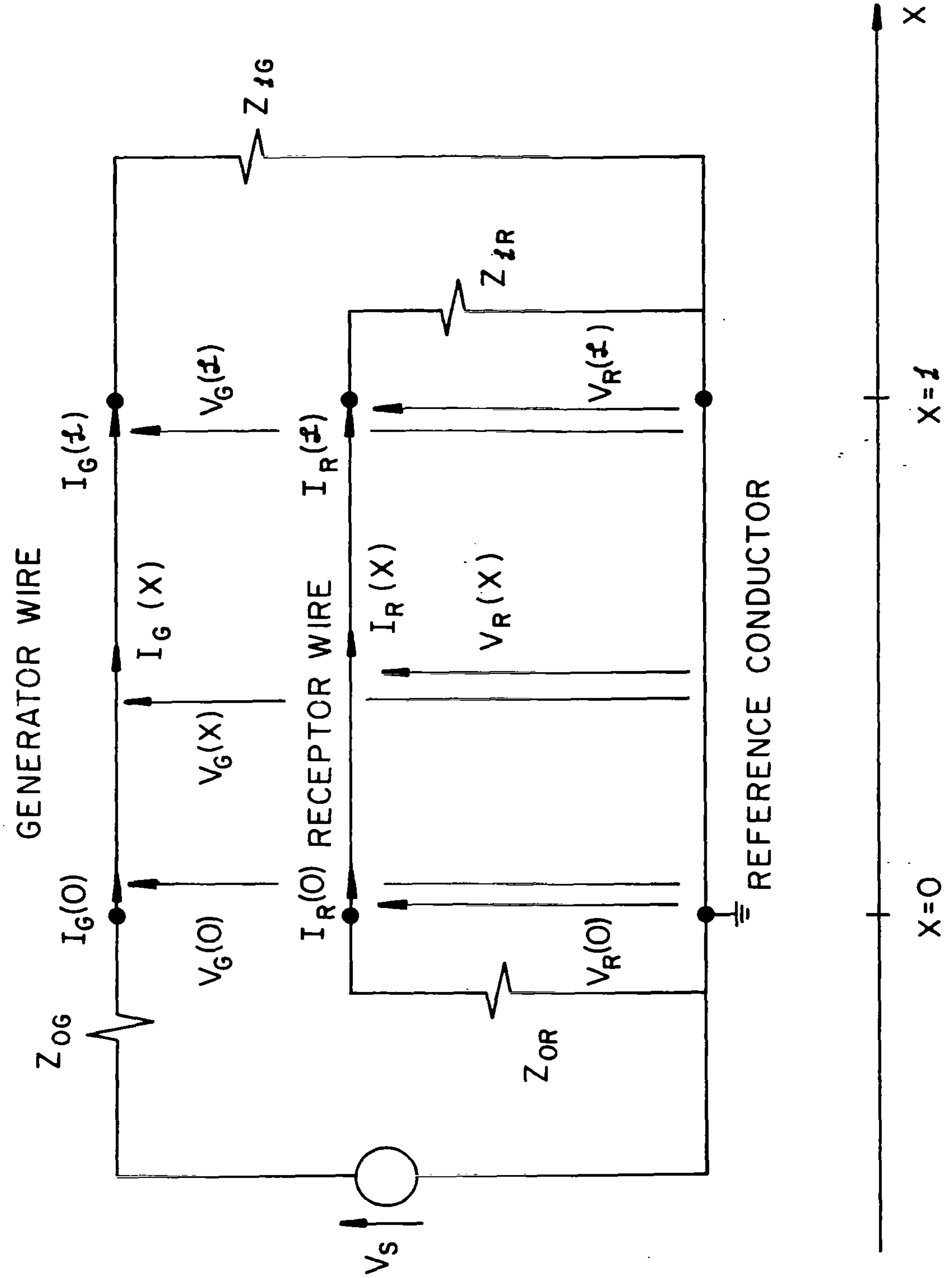
[#4] Do we need to consider the presence of the parasitic circuits in the cable when attempting to predict the coupling between two circuits in the cable?

To answer these questions with a firm YES or NO would require the investigation of an unlimited number of cases since there are such a large number of parameters involved, i.e., wire radius, dielectric types, insulation thickness, number of wires in the cable and specific load impedance configurations and values. Instead, we will investigate a typical 20 wire ribbon cable with randomly selected load impedances. We will drive one wire in the cable and measure the coupling to another wire in the cable.

To investigate the first question, we will consider the predictions of FLATPAK2 which neglects only the losses of the insulation dielectric. To investigate the second question, we will compare the results of FLATPAK and FLATPAK2 to XTALK and XTALK2. To investigate the third question, we will compare the results of XTALK2 and FLATPAK2 to XTALK and FLATPAK. Since the frequency response of the cable is generally required at a large number of frequencies, it is desirable to determine the most efficient computational del that will yield acceptable prediction accuracies. Since consideration of the presence of wire dielectric and conductor losses each effectively require programs with larger array storage and longer computation times (per-frequency) than if these factors are neglected, our interest centers on the investigation of the effect of these two parameters on the prediction accuracy of the MTL model.

A parallel point of interest is question #4. For an (n+1) wire cable, the MTL model requires a solution of a minimum of n complex, simultaneous equations at each frequency [1]. The natural question is whether we need to consider the interactions between all wires in the cable or do we only need to consider the driven wire and the "pickup" wire. To illustrate this further, consider Figure 1-3. In the ribbon cable, one of the wires is designated as the reference wire for the line voltages. In Figure 1-3, we have shown the driven circuit or "generator" circuit consisting of the generator wire and the reference wire and the "pickup" or receptor circuit consisting of the receptor wire and the reference wire. The remaining (n-2) circuits consisting of (n-2) wires each with the reference wire will be referred to as parasitic circuits. The obvious question is whether these parasitic circuits measurably affect the coupling between the generator and receptor circuits. If they do not, then there is no need to consider them and our computational model is reduced from the solution of n complex, simultaneous equations at each frequency to only two (2) simultaneous, complex equations at each frequency. When the response at many frequencies is desired and/or n is large, this would result in an obvious savings in computation time as well as required array storage.

Chapter II will formulate the MTL model specialized to ribbon cables.



GENERATOR WIRE
RECEPTOR WIRE
REFERENCE CONDUCTOR
Z_{OG}
$Z_{\ell G}$
Z_{OR}
$Z_{\ell R}$
V_S
$I_G(O)$
$I_G(X)$
$I_G(\ell)$
$I_R(O)$
$I_R(X)$
$I_R(\ell)$
$V_G(O)$
$V_G(X)$
$V_G(\ell)$
$V_R(O)$
$V_R(X)$
$V_R(\ell)$
X=O
X=ℓ
X

Figure 1-3.

The reader need not be intimately familiar with the details of Chapter II in order to use the computer programs or understand the conclusions of this work. Chapter III provides the experimental results and the computer program predictions.

II. THE MULTICONDUCTOR TRANSMISSION LINE (MTL) PREDICTION MODEL

In this Chapter, we will give a brief description of the MTL model as specialized to the case of ribbon cables. The reader is referred to [1] and [7] for further details of the model. Computer programs XTALK, XTALK2, FLATPAK and FLATPAK2 which are described in [7] implement the results of this section with certain parameters, i.e., conductor losses and presence of wire dielectric, either considered or neglected. FLATPAK and FLATPAK2 are written specifically for consideration of ribbon cables. Although XTALK and XTALK2 allow consideration of more general types of cables, these programs can be used for our purposes here.

2.1 The MTL Model

If the line conductors are immersed in a homogeneous medium, e.g., bare wires in free space, the fundamental mode of propagation is the TEM (Transverse Electro-Magnetic) mode. If the line conductors are immersed in an inhomogeneous medium, e.g., dielectric insulations surrounded by free space as with ribbon cables, the fundamental mode of propagation is taken to be the "quasi-TEM" mode. The essential difference between these two cases is as follows. For the TEM mode, the electric and magnetic field vectors lie in a plane transverse or perpendicular to the line (x) axis. For lines in an inhomogeneous medium, the TEM mode cannot exist except in the limiting case of zero frequency (DC). However, for the inhomogeneous case, the assumption is made that the electric and magnetic field vectors are almost transverse to the line axis, i.e., the mode of propagation is "quasi-TEM".

With the assumption of the TEM or "quasi-TEM" mode of propagation, line voltages and currents may be defined. One of the conductors in the (n+1) conductor transmission line is designated as the reference conductor or zero-th conductor for the line voltages. Thus the line conductors are numbered from 0 to n, i.e., 0, 1, 2, ---, n. Throughout the report, the longitudinal ·· s of the line is the x direction and the line is considered to be uniform in the sense that all (n+1) conductors and surrounding dielectric insulations have no cross-sectional variation along the line and all (n+1) conductors are parallel to each other. For sinusoidal, steady-state excitation of the line, the line voltages, $\mathcal{V}_i(x,t)$, (with respect to the reference or zero-th conductor) and line currents, $\mathcal{I}_i(x,t)$, are

$$\mathcal{V}_i(x,t) = V_i(x)\, e^{j\omega t} \tag{2-1a}$$

$$\mathcal{I}_i(x,t) = I_i(x)\, e^{j\omega t} \tag{2-1b}$$

for i=1, ---, n where $V_i(x)$ and $I_i(x)$ are the complex, phasor line voltages and currents and ω is the radian frequency of excitation, $\omega = 2\pi f$. The current in the reference conductor satisfies

$$\mathcal{I}_0(x,t) = -\sum_{i=1}^{n} \mathcal{I}_i(x,t) \tag{2-2a}$$

$$I_0(x) = -\sum_{i=1}^{n} I_i(x) \tag{2-2b}$$

The per-unit-length equivalent circuit of the ribbon cable is shown in Figure 2-1. The conductors are assumed to be identical and have internal resistance and internal inductance per-unit-length of r_c and ℓ_c, respectively. The per-unit-length capacitance between the i-th and j-th wires is designated by c_{ij}. The per-unit-length self inductance of the i-th wire is designated

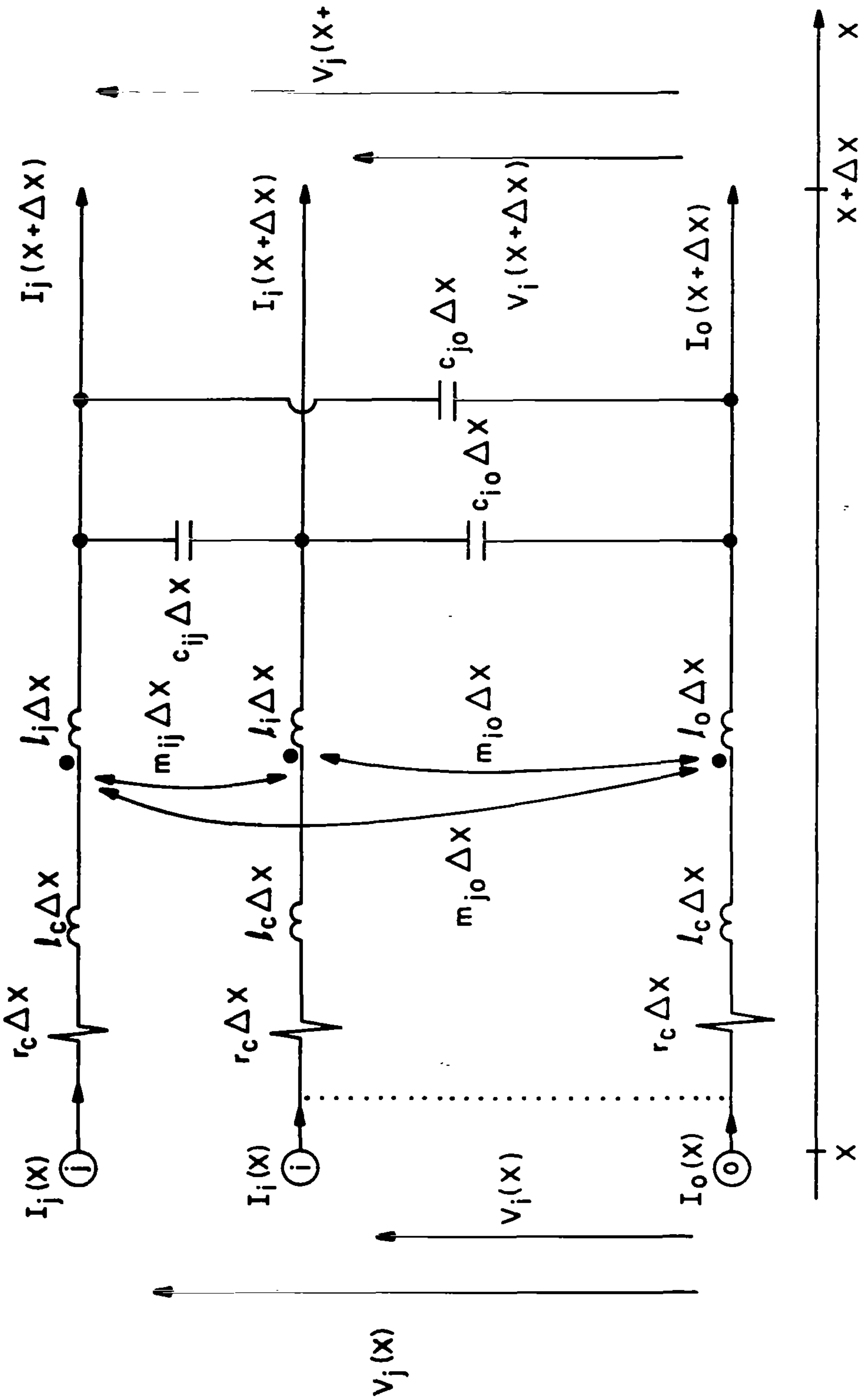


Figure 2-1. The per-unit-length model for a ribbon cable.

by ℓ_i and the per-unit-length mutual inductance between the i-th and j-th wires is designated by m_{ij}.

The MTL equations can be derived from the per-unit-length circuit of Figure 2-1 in the limit as $\Delta x \to 0$ as a set of 2n, complex-valued, first order, ordinary differential equations [1]

$$\frac{d}{dx}\begin{bmatrix}\underline{V}(x)\\ \underline{I}(x)\end{bmatrix} = -\begin{bmatrix}\underset{\sim}{0}_{n n} & \underset{\sim}{Z}\\ \underset{\sim}{Y} & \underset{\sim}{0}_{n n}\end{bmatrix}\begin{bmatrix}\underline{V}(x)\\ \underline{I}(x)\end{bmatrix} \tag{2-3}$$

A matrix $\underset{\sim}{M}$ with m rows and p columns is said to be mχp and the element in the i-th row and j-th column is designated by $[\underset{\sim}{M}]_{ij}$ with i=1,---,m and j=1, ---,p. An nχ1 vector is denoted with a bar, e.g., $\underline{V}$, with the entry in the i-th row denoted by $[\underline{V}]_i = V_i$. The matrix $\underset{\sim}{0}_{mp}$ is the mχp zero matrix with zeros in every position, i.e., $[\underset{\sim}{0}_{mp}]_{ij} = 0$ for i=1,---,m and j=1,---,p. The complex-valued phasor line voltages with respect to the reference conductor (the zero-th conductor), $V_i(x)$, and the line currents, $I_i(x)$, are given by $[\underline{V}(x)]_i = V_i(x)$ and $[\underline{I}(x)]_i = I_i(x)$.

The nχn complex-valued, symmetric matrices, $\underset{\sim}{Z}$ and $\underset{\sim}{Y}$, are the per-unit-length impedance and admittance matrices of the line, respectively. Since the line is assumed to be uniform, these matrices are independent of x and may be separated as [1]

$$\underset{\sim}{Z} = (r_c + j\omega\ell_c)[\underset{\sim}{1}_n + \underset{\sim}{U}_n] + j\omega\underset{\sim}{L} \tag{2-4a}$$

$$\underset{\sim}{Y} = j\omega\underset{\sim}{C} \tag{2-4b}$$

where $\underset{\sim}{1}_n$ is the nχn identity matrix, i.e., $[\underset{\sim}{1}_n]_{ii} = 1$, $[\underset{\sim}{1}_n]_{ij} = 0$ for i, j=1, ---, n and i$\neq$j and $\underset{\sim}{U}_n$ is the unit matrix with ones in every position, i.e., $[\underset{\sim}{U}_n]_{ij} = 1$ for i, j=1, ---, n. The nχn, real, symmetric matrices $\underset{\sim}{L}$ and $\underset{\sim}{C}$ are the per-unit-length external inductance and capacitance matrices, respectively. The entries in these matrices can be straightforwardly determined from the per-unit-length equivalent circuit in Figure 2-1 as

$$[\underset{\sim}{L}]_{ii} = \ell_i + \ell_0 - 2\, m_{i0} \tag{2-5a}$$

$$[\underset{\sim}{L}]_{\substack{ij \\ i\neq j}} = \ell_0 + m_{ij} - m_{i0} - m_{j0} \tag{2-5b}$$

$$[\underset{\sim}{C}]_{ii} = c_{i0} + \sum_{\substack{j=1 \\ i\neq j}}^{n} c_{ij} \tag{2-5c}$$

$$[\underset{\sim}{C}]_{\substack{ij \\ i\neq j}} = -\, c_{ij} \tag{2-5d}$$

The solution to (2-3) for a line of total length $\mathcal{L}$ is [1]

$$\begin{bmatrix} \underline{V}(\mathcal{L}) \\ \underline{I}(\mathcal{L}) \end{bmatrix} = \underbrace{\begin{bmatrix} \underset{\sim}{\Phi}_{11} & \underset{\sim}{\Phi}_{12} \\ \underset{\sim}{\Phi}_{21} & \underset{\sim}{\Phi}_{22} \end{bmatrix}}_{\underset{\sim}{\Phi}} \begin{bmatrix} \underline{V}(0) \\ \underline{I}(0) \end{bmatrix} \tag{2-6}$$

where $\underset{\sim}{\Phi}$ is the 2nχ2n chain parameter matrix and $\underset{\sim}{\Phi}_{ij}$ are nχn. Once the chain parameter matirx of the line is determined, the 2n unknown voltages, $\underline{V}(\mathcal{L})$ and $\underline{V}(0)$, and the 2n unknown currents, $\underline{I}(\mathcal{L})$ and $\underline{I}(0)$, can be determined by enforcing the equations of the terminal networks at the two ends of the line. The terminal networks are considered to be linear and can be characterized by Generalized Thevenin Equivalents as [1]

$$\underline{V}(0) = \underline{V}_0 - \underset{\sim}{Z}_0 \, \underline{I}(0) \tag{2-7a}$$

$$\underline{V}(\mathcal{L}) = \underline{V}_{\mathcal{L}} + \underset{\sim}{Z}_{\mathcal{L}} \, \underline{I}(\mathcal{L}) \tag{2-7b}$$

where $\underset{\sim}{Z}_0$ and $\underset{\sim}{Z}_{\mathcal{L}}$ are $n \chi n$ complex impedance matrices and $\underline{V}_0$ and $\underline{V}_{\mathcal{L}}$ are $n \chi 1$ vectors of the open circuit voltages of the termination networks (produced by independent sources within these networks).

Although the formulation in (2-7) allows for general termination networks we will consider a special case which will be sufficient for our purposes. Consider the termination networks in Figure 2-2 in which each line conductor at each end of the line is connected to the reference conductor through an impedance, Z_{0i} or $Z_{\mathcal{L}i}$, and a voltage source, V_{0i} or $V_{\mathcal{L}i}$. The entries in (2-7) are easily determined for this form of the termination networks as

$$[\underset{\sim}{Z}_0]_{ii} = Z_{0i} \tag{2-8a}$$

$$[\underset{\sim}{Z}_0]_{\substack{ij \\ i \neq j}} = 0 \tag{2-8b}$$

$$[\underline{V}_0]_i = V_{0i} \tag{2-8c}$$

$$[\underset{\sim}{Z}_{\mathcal{L}}]_{ii} = Z_{\mathcal{L}i} \tag{2-8d}$$

$$[\underset{\sim}{Z}_{\mathcal{L}}]_{\substack{ij \\ i \neq j}} = 0 \tag{2-8e}$$

$$[\underline{V}_{\mathcal{L}}]_i = V_{\mathcal{L}i} \tag{2-8f}$$

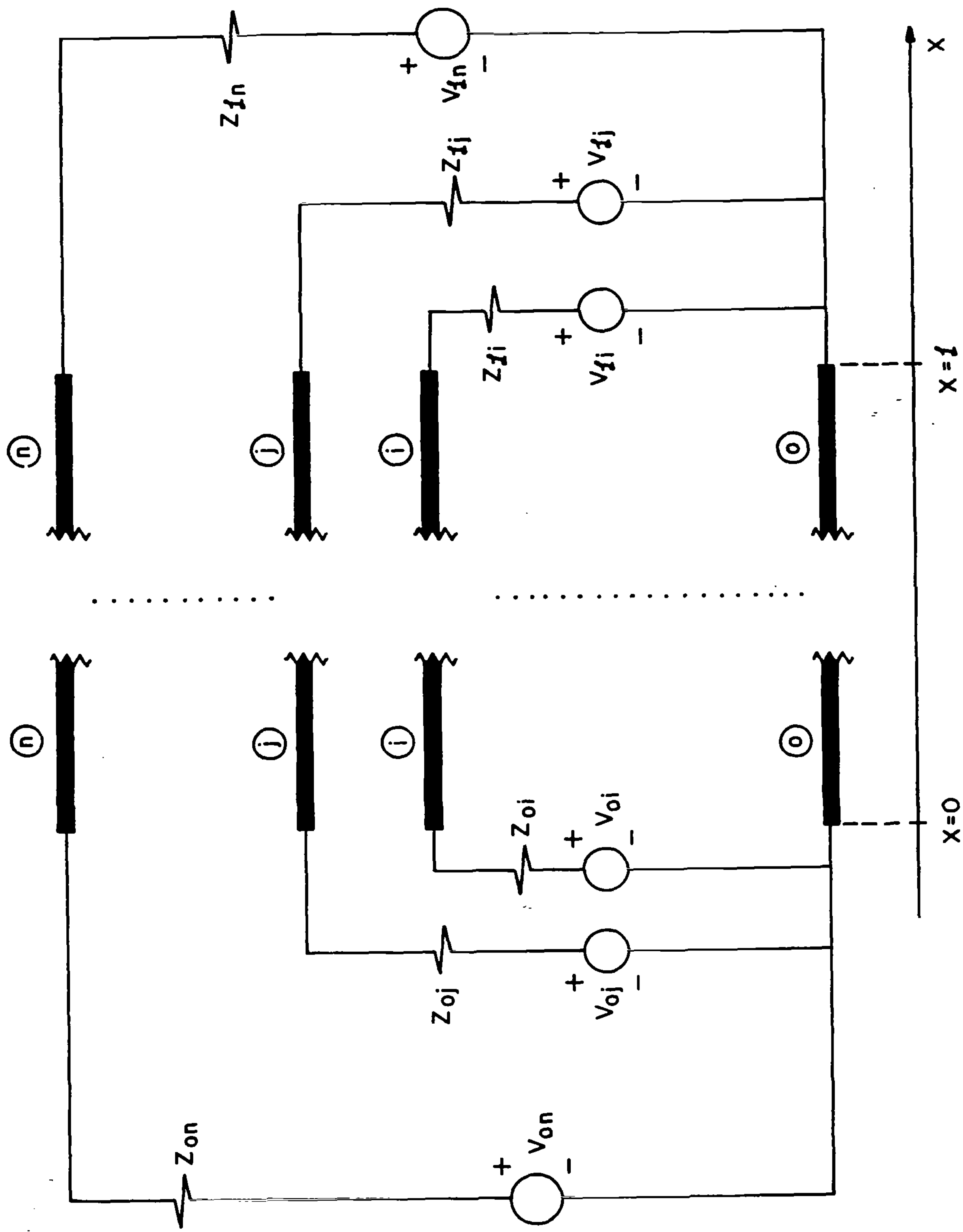


Figure 2-2. The termination structure for the ribbon cable.

and the terminal impedance matrices, $\underset{\sim}{Z}_0$ and $\underset{\sim}{Z}_{\mathcal{L}}$, are diagonal. This form of the terminal impedance networks will be used in obtaining the experimental results.

To determine the terminal voltages $\underline{V}(0)$ and $\underline{V}(\mathcal{L})$, substitute (2-7) into (2-6) to obtain [1]

$$[\underset{\sim}{Z}_{\mathcal{L}} \underset{\sim}{\Phi}_{22} - \underset{\sim}{Z}_{\mathcal{L}} \underset{\sim}{\Phi}_{21} \underset{\sim}{Z}_0 - \underset{\sim}{\Phi}_{12} + \underset{\sim}{\Phi}_{11} \underset{\sim}{Z}_0] \underline{I}(0) = [\underset{\sim}{\Phi}_{11} - \underset{\sim}{Z}_{\mathcal{L}} \underset{\sim}{\Phi}_{21}] \underline{V}_0 - \underline{V}_{\mathcal{L}} \quad (2\text{-}9a)$$

$$\underline{I}(\mathcal{L}) = \underset{\sim}{\Phi}_{21} \underline{V}_0 + [\underset{\sim}{\Phi}_{22} - \underset{\sim}{\Phi}_{21} \underset{\sim}{Z}_0] \underline{I}(0) \quad (2\text{-}9b)$$

Once the n equations in (2-9a) are solved for $\underline{I}(0)$, the currents $\underline{I}(\mathcal{L})$ can be found from (2-9b). The terminal voltages are then obtained from (2-7).

The remaining problems are the determination of the entries in the per-unit-length inductance and capacitance matrices, $\underset{\sim}{L}$ and $\underset{\sim}{C}$, respectively, the per-unit-length resistance and internal inductance of the wires, r_c and ℓ_c, respectively, and the entries in the chain parameter matrix, $\underset{\sim}{\Phi}$, (which obviously involve $\underset{\sim}{L}$, $\underset{\sim}{C}$, r_c and ℓ_c). These items will be discussed in the following sections.

2.2 The Chain Parameter Matrix

In this Section we will briefly describe the calculation of the chain parameter matrices which have been calculated elsewhere [1,7]. This will also serve to illustrate why there are different degrees of computational complexity (total computation time and required array storage) involved in programs XTALK, XTALK2, FLATPAK and FLATPAK2.

First consider the all inclusive case considered by FLATPAK2 in which conductor losses and the presence of dielectric insulations are included in the model. For this case, the entries in the chain parameter matrix are [1]

$$\Phi_{11} = \frac{1}{2} Y^{-1} T(e^{\gamma \ell} + e^{-\gamma \ell}) T^{-1} Y \tag{2-10a}$$

$$\Phi_{12} = -\frac{1}{2} Y^{-1} T \gamma(e^{\gamma \ell} - e^{-\gamma \ell}) T^{-1} \tag{2-10b}$$

$$\Phi_{21} = -\frac{1}{2} T(e^{\gamma \ell} - e^{-\gamma \ell}) \gamma^{-1} T^{-1} Y \tag{2-10c}$$

$$\Phi_{22} = \frac{1}{2} T(e^{\gamma \ell} + e^{-\gamma \ell}) T^{-1} \tag{2-10d}$$

where the inverse of an nχn matrix M is denoted by M^{-1} and $e^{\gamma \ell}$ is an nχn diagonal matrix with $[e^{\gamma \ell}]_{ii} = e^{\gamma_i \ell}$ and $[e^{\gamma \ell}]_{ij} = 0$ for i, j=1, ---, n and i≠j. The matrix T is an nχn complex matrix which diagonalizes the matrix product YZ as (see (2-4) for the forms of Y and Z)

$$\begin{aligned} T^{-1} Y Z T &= T^{-1} [j\omega C\{(r_c + j\omega \ell_c)(1_n + U_n) + j\omega L\}]T \\ &= \gamma^2 \end{aligned} \tag{2-11}$$

where γ^2 is an nχn diagonal matrix of propagation constants, γ_i, with $[\gamma^2]_{ii} = \gamma_i^2$ and $[\gamma^2]_{ij} = 0$ for i, j=1, ---, n and i≠j. The columns of T and the entries in γ are the eigenvectors and eigenvalues of the matrix product YZ,respectively. Thus an eigenvector- eigenvalue subroutine which handles complex matrices must be used in FLATPAK2. Furthermore, these eigenvalues and eigenvectors must be recomputed at every frequency being considered.

Now consider the reduction in complexity of the above calculation when conductor losses are neglected, i.e., $r_c = \ell_c = 0$ as is done in FLATPAK. In this case, the chain parameter matrix in (2-10) remains the same but the eigenvector-eigenvalue calculation in (2-11) becomes

$$\begin{aligned} \underset{\sim}{T}^{-1}\, \underset{\sim}{Y}\, \underset{\sim}{Z}\, \underset{\sim}{T} &= \underset{\sim}{T}^{-1}\, [j\omega \underset{\sim}{C}\; j\omega \underset{\sim}{L}]\, \underset{\sim}{T} \\ &= -\omega^2\, \underset{\sim}{T}^{-1}\, \underset{\sim}{C}\, \underset{\sim}{L}\, \underset{\sim}{T} \\ &= \underset{\sim}{\gamma}^2 \end{aligned} \qquad (2\text{-}12)$$

In this case, the eigenvectors are independent of frequency and need be computed only once. The propagation constants in $\underset{\sim}{\gamma}^2$ at each frequency are determined quite simply from the eigenvalues of $\underset{\sim}{C}\, \underset{\sim}{L}$. Clearly FLATPAK realizes a significant reduction in total computation time when more than one frequency is being considered.

The final two programs, XTALK2 and XTALK, neglect the wire insulation dielectric, i.e., the conductors are considered to be bare. The simplification that results from this assumption is that for this case (homogeneous medium), the product of the per-unit-length external inductance and capacitance matrices is diagonal, i.e.,

$$\underset{\sim}{L}\, \underset{\sim}{C} = \underset{\sim}{C}\, \underset{\sim}{L} = \mu\varepsilon \; \underset{\sim}{1}_n \qquad (2\text{-}13)$$

where ε and μ characterize the surrounding homogeneous medium (which is free space). For XTALK2 (which includes conductor losses) we insert (2-13) into (2-11) and obtain

$$\underset{\sim}{T}^{-1}\, \underset{\sim}{Y}\, \underset{\sim}{Z}\, \underset{\sim}{T} = j\omega(r_c + j\omega \ell_c)\, \underset{\sim}{T}^{-1}\, \underset{\sim}{C}\, (\underset{\sim}{1}_n + \underset{\sim}{U}_n)\, \underset{\sim}{T} - \omega^2 \mu_v \varepsilon_v\, \underset{\sim}{1}_n \qquad (2\text{-}14)$$

Therefore, we only need the eigenvalues and eigenvectors of the matrix product $\underset{\sim}{C}(\underset{\sim}{1}_n + \underset{\sim}{U}_n)$ which is frequency independent.

XTALK which neglects both insulation dielectric and conductor losses also utilizes the identity in (2-13) and we have

$$\underset{\sim}{T}^{-1} \underset{\sim}{Y} \underset{\sim}{Z} \underset{\sim}{T} = -\omega^2 \mu_v \varepsilon_v \underset{\sim}{1}_n \qquad (2\text{-}15)$$

In this case, there is no need to compute eigenvalues or eigenvectors of a matrix even once: an obvious computational savings.

2.3 Calculation of the Per-Unit-Length Parameters

In this section we consider calculation of the per-unit-length parameters of conductor resistance, r_c, conductor internal inductance, ℓ_c, and the entries in the per-unit-length inductance, matrix, $\underset{\sim}{L}$, and capacitance matrix, $\underset{\sim}{C}$. In calculating r_c and ℓ_c, we will assume that these parameters can be calculated for each wire as if the wires were isolated from each other. This is then a standard calculation which includes the skin effect for a solid cylindrical conductor. The result is given in [7] and is stored in XTALK2 and FLATPAK2 as approximations to the true result. These equations are obtained by first defining the skin depth, δ,

$$\delta = \frac{1}{\sqrt{\pi f \mu_v \sigma}} = \frac{1}{2\pi\sqrt{\sigma f \times 10^{-7}}} \qquad (2\text{-}16)$$

the D-C resistance of each wire, r_0,

$$r_0 = \frac{1}{\pi \sigma r_w^2} \qquad (2\text{-}17)$$

and the D-C internal inductance of each wire, ℓ_0,

$$\ell_0 = \frac{\mu_v}{8\pi} = .5 \times 10^{-7} \qquad (2\text{-}18)$$

In terms of these parameters, the resistance, r, and internal inductance, ℓ, of a solid, cylindrical conductor of radius r_w is

$$\text{(I)} \quad r_w \leq \delta$$
$$r = r_0 \qquad (2\text{-}19a)$$
$$\ell = \ell_0$$

$$\text{(II)} \quad \delta < r_w < 3\delta$$
$$r = \frac{1}{4}\left(\frac{r_w}{\delta} + 3\right) r_0 \qquad (2\text{-}19b)$$
$$\ell = \left[1.15 - .15\left(\frac{r_w}{\delta}\right)\right] \ell_0$$

$$\text{(III)} \quad r_w \geq 3\delta$$
$$r = \frac{r_w}{2\delta} r_0 \qquad (2\text{-}19c)$$
$$\ell = \frac{2\delta}{r_w} \ell_0$$

In typical ribbon cables, the wires are not solid conductors but are stranded. To calculate the total per-unit-length resistance, r_c, and internal inductance, ℓ_c, of each wire which consists of S strands of identical wires, we compute with (2-16) - (2-19) the resistance and self inductance of each strand and divide the result by the number of strands composing the wire. Thus we assume that all strands in each wire are identical (certainly a reasonable assumption) and all S strands in a wire are connected in parallel. The result is $r_c = r/S$ and $\ell_c = \ell/S$.

Calculation of $\underset{\sim}{L}$ and $\underset{\sim}{C}$ for programs XTALK and XTALK2 (the homogeneous medium case) are quite simple and are given in [1,7] as

$$[\underset{\sim}{L}]_{ii} = \frac{\mu_v}{\pi} \ \ell n \ (\frac{d_{i0}}{r_w}) \tag{2-20a}$$

$$[\underset{\sim}{L}]_{\substack{ij \\ i \neq j}} = \frac{\mu_v}{2\pi} \ \ell n \ (\frac{d_{i0} \ d_{j0}}{r_w \ \ d_{ij}}) \tag{2-20b}$$

where d_{i0} is the separation between the i-th wire and the reference wire and d_{ij} is the separation between the i-th and j-th wires. The per-unit-length capacitance matrix, $\underset{\sim}{C}$, for this homogeneous medium case is found from using (2-20) in the identity of (2-13), i.e.,

$$\underset{\sim}{C} = \mu_v \varepsilon_v \ \underset{\sim}{L}^{-1} \tag{2-21}$$

It should be noted that the simple result in (2-20) is made possible not only by neglecting the wire insulation dielectric but also it assumes that the minimum ratio of wire separation to wire radius, d/r_w, is "not too small". Quantitatively, this means that d/r_w should be greater than approximately 5 [6]. For a ribbon cable, if the insulation thickness, t, is equal to the wire radius (which is typically the case), then even if the wire insulations are touching, the conductor separation, $d \geq 2(t + r_w) = 4 \ r_w$, will be such that $d/r_w \geq 4$ and the approximate parameters should yield relatively accurate results.

Calculation of $\underset{\sim}{L}$ and $\underset{\sim}{C}$ for programs FLATPAK and FLATPAK2 are considerably more involved due to the presence of the inhomogeneous medium dielectric insulation and free space) surrounding the conductors. Assuming the dielectric insulations to have free space permeability, μ_v, we have [1,7]

$$\underset{\sim}{L}\,\underset{\sim}{C}_0 = \mu_v \varepsilon_v \underset{\sim}{1}_n \qquad (2\text{-}22)$$

where $\underset{\sim}{C}_0$ is the per-unit-length capacitance with the dielectric insulations removed. Thus

$$\underset{\sim}{L} = \mu_v \varepsilon_v \underset{\sim}{C}_0^{-1} \qquad (2\text{-}23)$$

For the calculation of $\underset{\sim}{C}$, the identity in (2-13) does not hold. Therefore we need to compute the per-unit-length capacitance matrix with <u>and</u> without the dielectric insulations. This is still a formidable problem and generally no closed form results are obtainable. The approximation in (2-20), however, could be used for $\underset{\sim}{L}$. Calculation of $\underset{\sim}{C}$, remains difficult. Numerical approximation methods have been devised to very accurately compute $\underset{\sim}{C}$ and $\underset{\sim}{C}_0$ [1, 4, 5, 6]. For the specific case of a ribbon cable, a digital computer program has been written to compute $\underset{\sim}{C}$ and $\underset{\sim}{C}_0$. The program GETCAP is described in [5] and produces punched card output containing the entries in $\underset{\sim}{C}$ or $\underset{\sim}{C}_0$ for direct use in the FLATPAK and FLATPAK2 programs.

III. EXPERIMENTAL RESULTS AND MTL PREDICTION ACCURACIES

In this Chapter, the experimental test of a 20 wire ribbon cable will be described. The predictions of the four computer programs will be presented for comparison with the experimental results. The reader should refer to Chapter I for a complete discussion of the objectives of this study as well as the parameters which are included or neglected from consideration in each of the MTL prediction programs, XTALK, XTALK2, FLATPAK and FLATPAK2. A brief summary of the program capabilities is repeated here for convenience:

MTL Prediction Program Capabilities

Program Name	Presence of wire dielectric considered?	Conductor losses considered?
XTALK	NO	NO
XTALK2	NO	YES
FLATPAK	YES	NO
FLATPAK2	YES	YES

3.1 The Experiment

The cable tested was a 20 wire ribbon cable. The wires were #28 gauge which consisted of stranded conductors with 7 strands of #36 gauge copper wire. The center-to-center wire separation, d, in Figure 1-2 was .05 inches (1.27×10^{-3} meters). Using a conductor diameter of #28 gauge (diameter of 3.2004×10^{-4} meters), this resulted in an insulation thickness, t, in Figure 1-2 of 3.4798×10^{-4} meters. The insulation was polyvinylchloride and a relative dielectric constant of 3.5 is assumed.

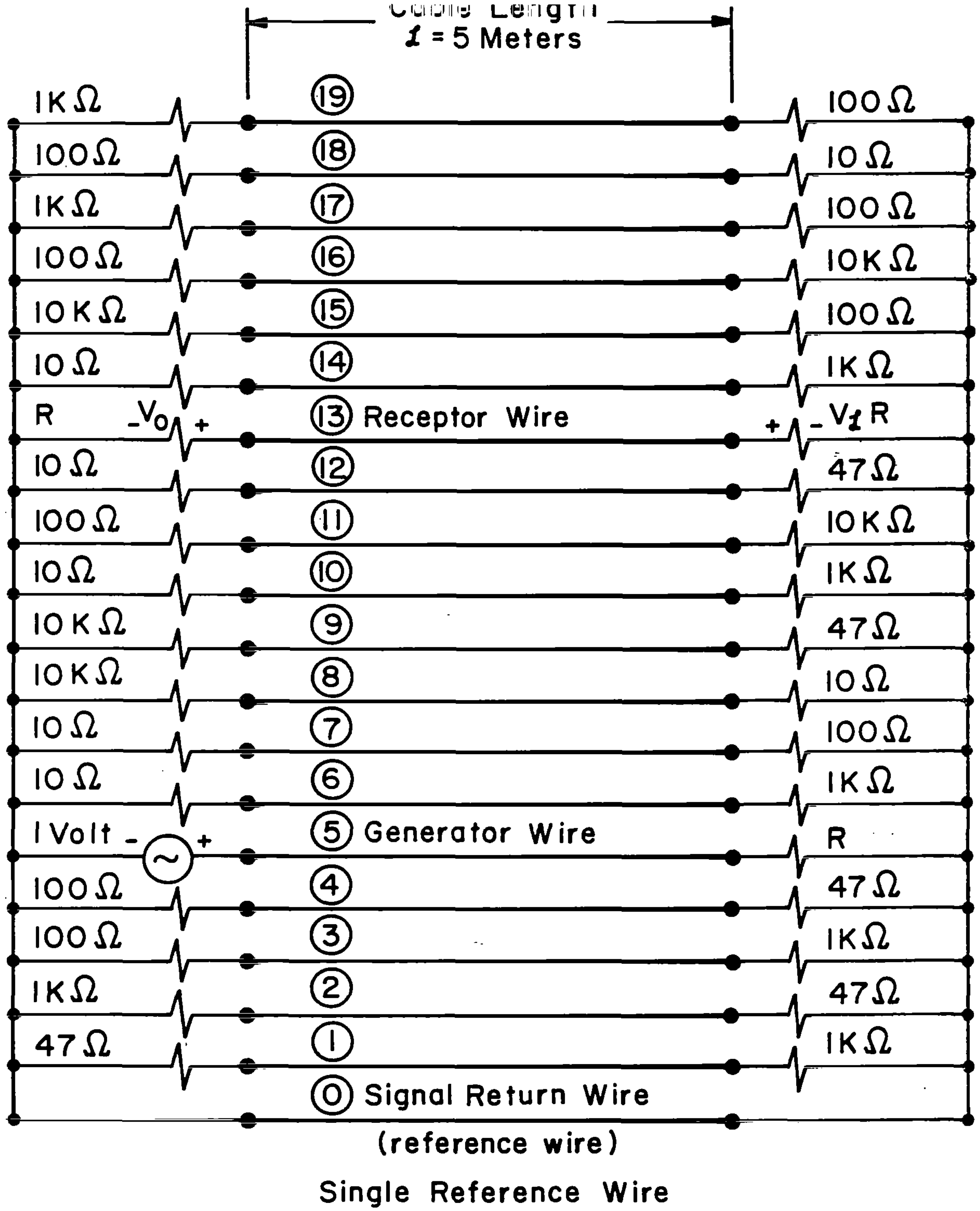
The total cable length was 5 meters. One of the outer wires was chosen as the reference wire and all other wires are terminated to this wire. The wires are numbered from 0 to 19 as shown in Figure 3-1. The generator wire is chosen to be wire #5 and the receptor wire is chosen to be wire #13. The generator wire is driven at x=0 by a 1 volt source (zero source impedance). This was accomplished in the experiment by monitoring the voltage (with respect to the reference wire) of wire #5 at x=0 and adjusting at each frequency the signal generator to achieve 1 volt at this point. The other end of the generator wire is terminated in a resistance R. The two ends of the receptor wire are terminated to the reference wire with a resistance R also. Two values of R will be investigated;

$$R = 50\,\Omega$$

$$R = 1000\,\Omega$$

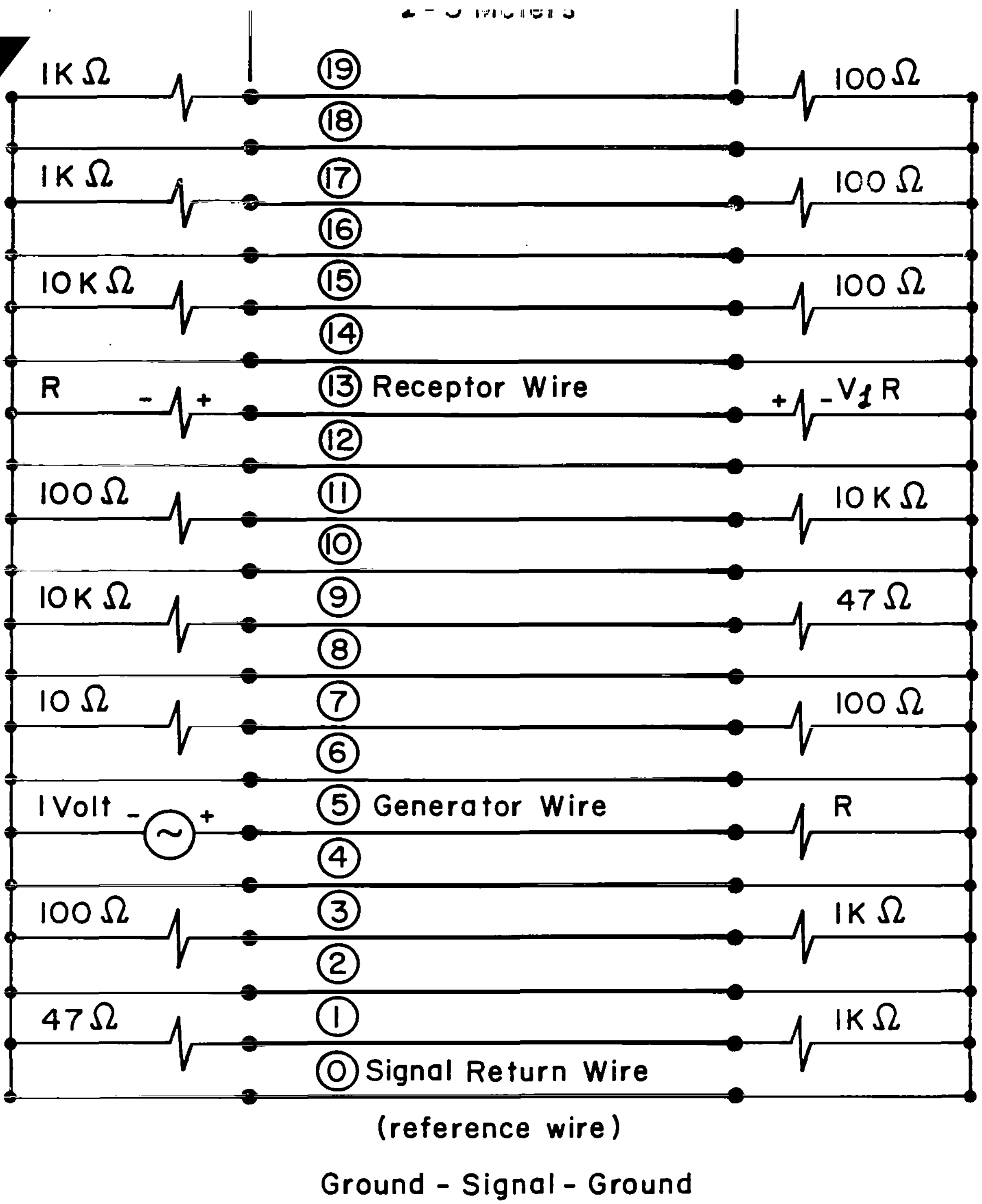
The rationale for selecting these two values of impedance for the generator and receptor wires is described in [8]. It would appear from the results in [8] that vastly different results would be obtained when R is a "high impedance load" (greater than the characteristic impedances of each isolated two wire circuit which are 442 Ω and 556Ω) as opposed to the case when R is a "low impedance load". R = 50 Ω represents, for this physical situation, a "low impedance load" and R = 1K Ω represents a "high impedance load". Various other load resistors are connected between the ends of each of the 17 parasitic wires and the reference wire as shown in Figure 3-1. The resistances remain the same whether R = 50 Ω or R = 1KΩ.

Two sets of load structures on the parasitic wires will be considered as shown in Figure 3-1. The first set in Figure 3-1a is termed the Single Reference Wire case. A quite common alternate scheme is the Ground-Signal-



(a)

Figure 3-1. The load structure for the ribbon cable (cont).



(b)

Figure 3-1. The load structure for the ribbon cable.

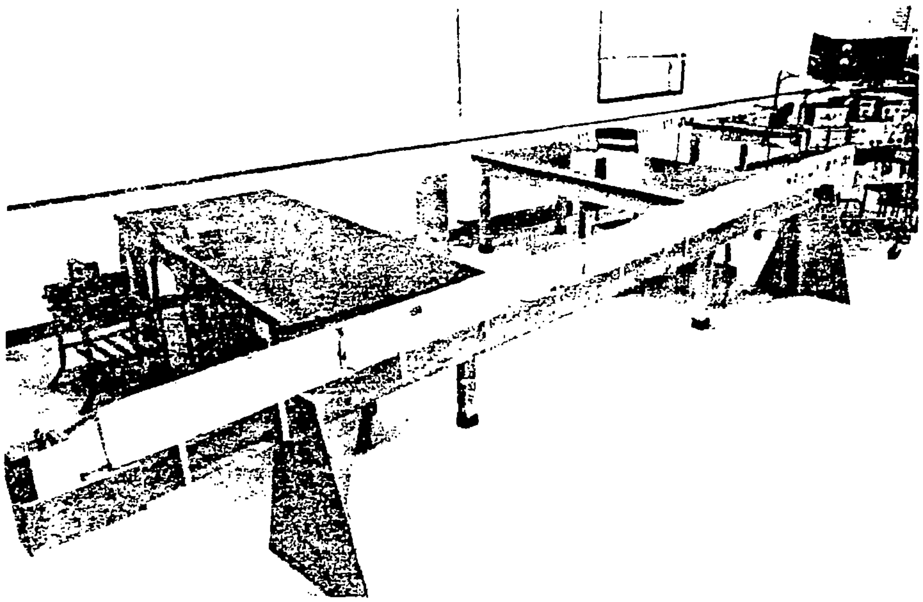
Ground configuration shown in Figure 3-1b in which alternate wires in the cable are tied together resulting in 10 wires of our 20 wire cable being reference wires. The remaining wires use the same resistances used in Fig. 3-1a.

The received voltage V_0 at x=0 for the receptor wire is the quantity which was measured. The frequencies investigated range from 10 KHz to 100 MHz. Based on free space calculation,the line is one wavelength (λ) long at 60 MHz. Therefore this range of frequencies should provide an investigation of the cable when it is electrically short ($\ell \ll \lambda$) as well as when it is electrically long ($\ell \gg \lambda$).

Discrete frequency measurements were made at numerical values of 1, 1.5, 2, 2.5, 3, 4, 5, 6, 7, 8, 9 in each decade from 10 KHz to 100 MHz. A frequency counter was used to control the excitation frequency to within .01% of the desired frequency. The measurement equipment consisted of:

		Frequency Range
(1)	HP 8405A Vector Voltmeter	1 MHz → 100 MHz
(2)	HP 3400A RMS Voltmeter	10 KHz → 1 MHz
(3)	HP 8601A Generator	1 MHz → 100 MHz
(4)	Wavetek 134 Generator	10 KHz → 1 MHz
(5)	Tektronix DC502 Frequency Counter	10 KHz → 100 MHz

Figures of the physical configuration are shown in Figure 3-2. A large "sawhorse" was constructed using a 16 foot 2" x 6" board placed on its edge as the central member and supported at each end by plywood legs. Styrofoam blocks were placed along the top edge of the board and the cable was placed on top of these blocks.



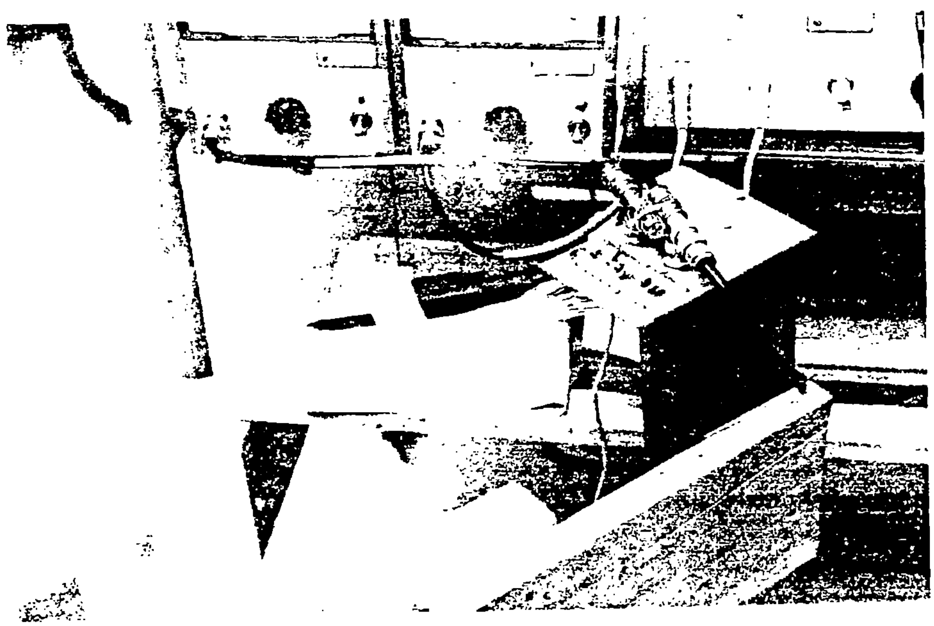


Figure 3-2. The physical configuration for the experiment (cont).

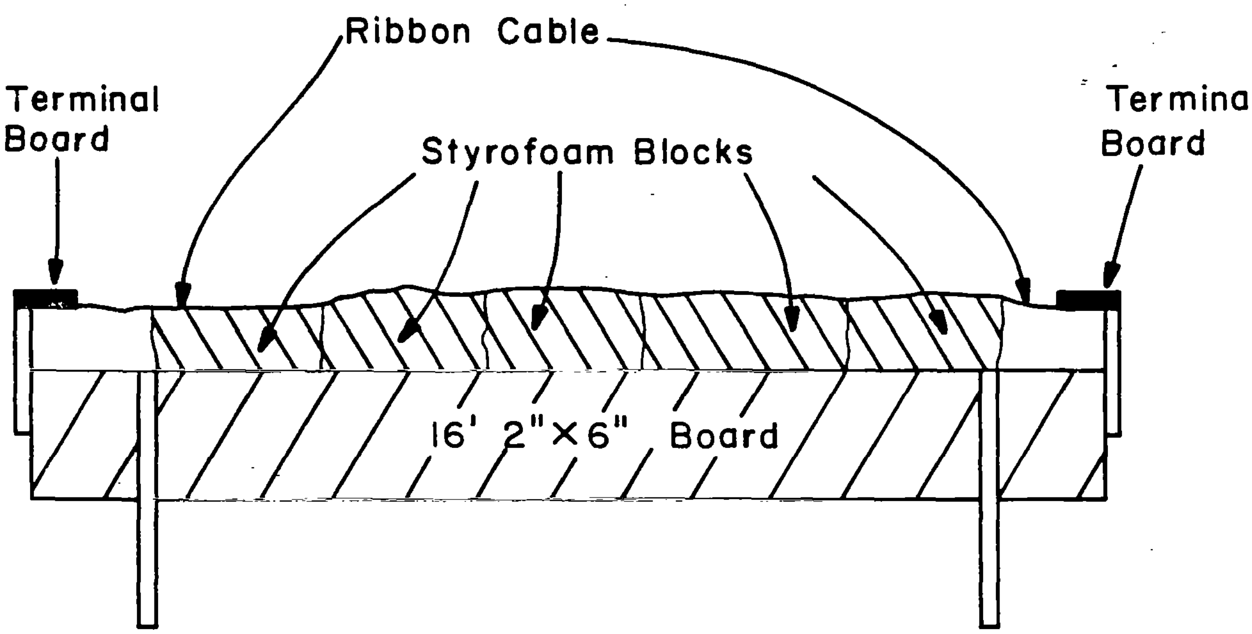


Figure 3-2. The physical configuration for the experiment.

3.2 Common Impedance Coupling

A very simple mode of coupling which is dominant at low frequencies is due to the reference wire (signal return) impedance which is common to both the generator and receptor circuits. This portion of the total coupling will be clearly demonstrated in the experimental results. In cases where the signal return path common to both generator and receptor circuits is a wire (as it is here), this form of coupling cannot be overlooked.

The primary effect of imperfect conductors is to introduce this common impedance coupling. Consider a transmission line in which there is no cross-coupling within the termination networks i.e., at each end of the line, each endpoint of a wire is terminated directly to the reference wire and is not physically connected to the endpoints of the other wires. In this case, clearly the voltages induced via electromagnetic field coupling at the ends of a "receptor" circuit consisting of one wire and the reference wire due to a "generator" circuit consisting of another wire and the reference wire will approach zero as the frequency of excitation is reduced to zero. However, the reference wire impedance can couple a signal into the receptor circuit even at D-C and this is usually termed common impedance coupling.

To illustrate this, consider Figure 3-3. In Figure 3-3a, a three-conductor transmission line is shown. The reference conductor has a certain total impedance, Z_0, which may be considerably smaller in magnitude than Z_{OR} or $Z_{\ell R}$. Consequently, the current in the generator wire at frequencies approaching D-C may be determined as

$$I_G = \frac{V_G}{Z_{OG} + Z_{\ell G}} \qquad (3\text{-}1)$$

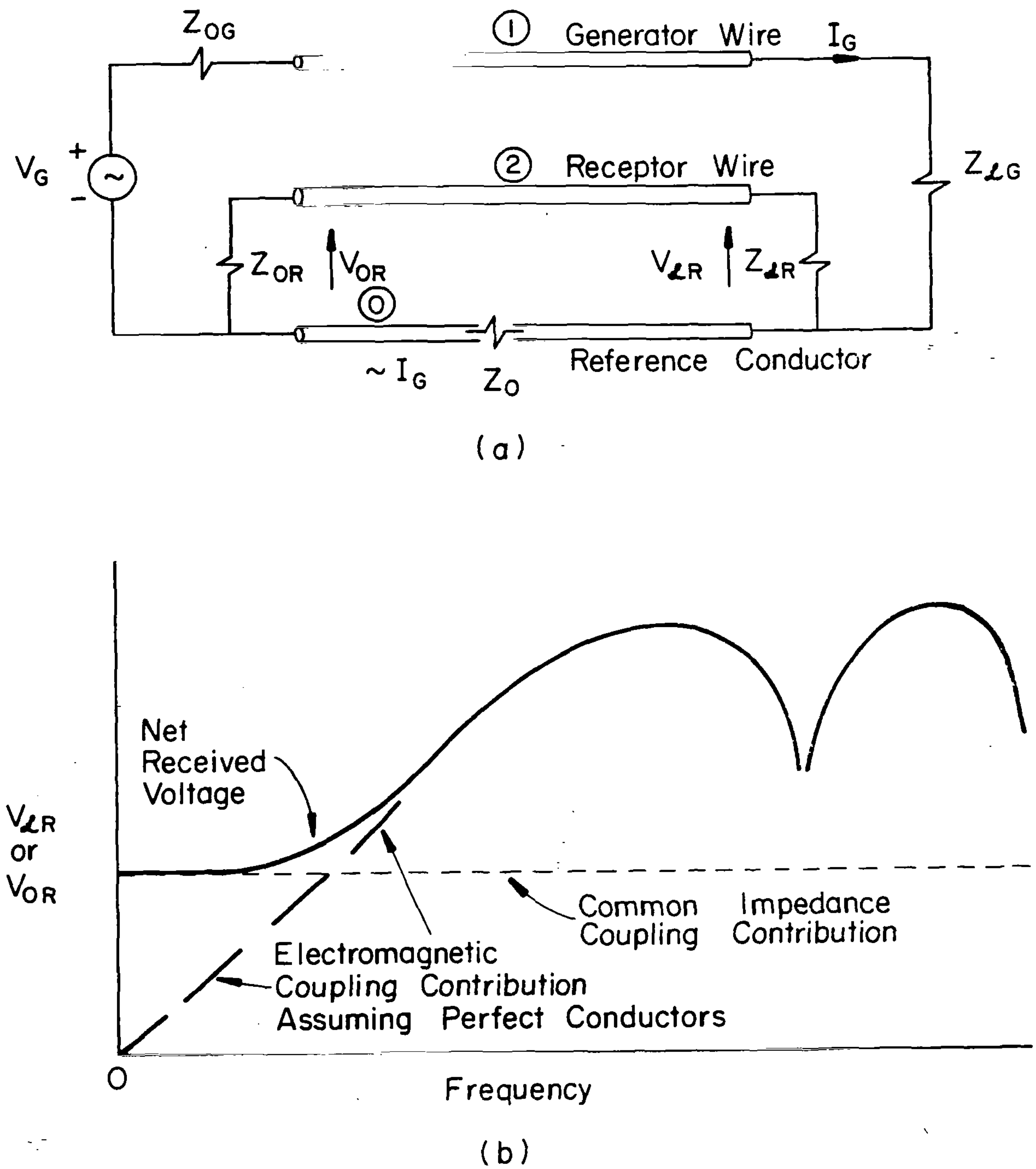


Figure 3-3. Illustration of common impedance coupling.

This is due to the fact that Z_0, Z_G (the impedance of the generator wire), and Z_R (The impedance of the receptor wire) are usually quite small and the net impedance seen by the source V_G is

$$Z_{0G} + Z_{\not{\ell}G} + Z_G + [(Z_{0R} + Z_{\not{\ell}R} + Z_R) || Z_0] \cong Z_{0G} + Z_{\not{\ell}G} \tag{3-2}$$

where || means "in parallel with". This current, I_G, will be divided between the parallel paths consisting of the reference wire impedance, Z_0, and the impedance of the receptor circuit $Z_{0R} + Z_{\not{\ell}R} + Z_R \cong Z_{0R} + Z_{\not{\ell}R}$. Consequently, virtually all of I_G will flow in the receptor wire. This current produces a voltage drop of

$$V_{CI} = Z_0 \; I_G \tag{3-3}$$

This results in received voltages

$$V_{\not{\ell}R} \cong \frac{-Z_{\not{\ell}R}}{Z_{\not{\ell}R} + Z_{0R}} \; Z_0 \; I_G \tag{3-4a}$$

$$V_{0R} \cong \frac{+Z_{0R}}{Z_{\not{\ell}R} + Z_{0R}} \; Z_0 \; I_G \tag{3-4b}$$

Although this portion of the total received voltage may be "small" it may nevertheless be larger than the contribution due to electromagnetic field coupling as shown in Figure 3-3b. Consequently, this common impedance coupling generates a "floor" of induced voltage where a solution assuming perfect conductors would indicate a perhaps negligably small received voltage at the lower frequencies.

The frequency at which this common impedance coupling becomes significant depends on many factors some of which are line geometry (which affects the level of the electromagnetic portion of the coupling) and type of reference conductor. Reference conductors consisting of a #36 gauge wire or a large, thick ground plane would certainly not produce the same level of common impedance coupling.

The above separation and superposition of the two coupling mechanisms is only correct when one dominates the other by a considerable amount. To obtain a quantitatively correct answer, one must include the conductor self impedances directly in the transmission line solution and this is done in XTALK2 and FLATPAK2.

However, we may calculate the D-C common impedance voltage for the two configurations in Figure 3-1 quite simply. The common impedance coupling contribution to V_0 is from (3-1) and (3-4)

$$V_{OCI} = \left(\frac{R}{R + R}\right) Z_0 \left(\frac{1\angle 0}{R}\right) = \frac{Z_0}{2R} \qquad (3\text{-}5)$$

where R is 50Ω or $1K\Omega$ (Z_{0G} is zero for the experiment.) For the 20 wire ribbon cable used, each wire consisted of 7 strands of #36 gauge copper wire. Each strand has a D-C impedance per meter of (see(2-17)) (the nominal radius of #36 gauge wire is 2.5 mils)

$$\begin{aligned} r_{DC} &= \frac{1}{\pi \sigma r_w^2} \\ &= \frac{1}{\pi(5.8 \times 10^7)(6.35 \times 10^{-5})^2} \qquad (3\text{-}6) \\ &= 1.361052\,\Omega/\text{m} \end{aligned}$$

Considering the 7 strands of each wire to be in parallel, the net common impedance is

$$Z_0 = (\frac{r_{DC}}{7})(5m) = .97218\ \Omega \qquad (3-7)$$

(Single Reference Wire)

For the Ground-Signal-Ground configuration in Figure 3-1b, the net common impedance consists of 10 wires with impedance given in (3-7) in parallel or

$$Z_0 = \frac{.97218}{10}$$

$$= .097218 \qquad \text{(Ground-Signal-Ground)} \qquad (3-8)$$

The common impedance coupling contribution to the voltage V_0 in (3-5) for the two configurations in Figure 3-1 are given in Table I.

3.3 The Single Reference Wire Configuration

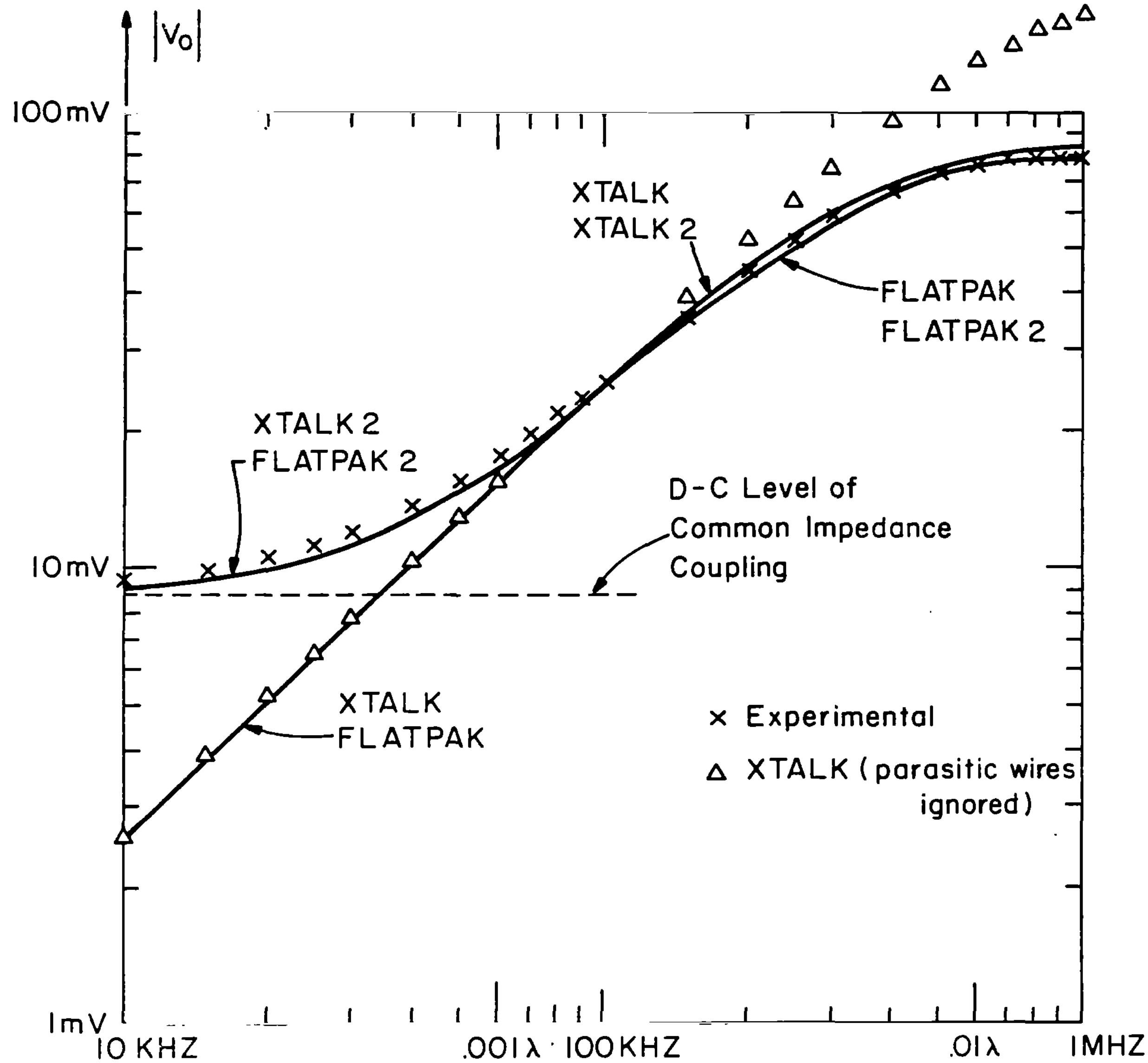
The experimental results for the Single Reference Wire Configuration in Figure 3-1a is shown in Figure 3-4 for R=50Ω(low impedance loads on the generator and receptor wires) and in Figure 3-5 for R=1KΩ(high impedance loads on the generator and receptor wires). Recall that the line is approximately one wavelength long at 60 MHz. We observe the usual rapid variation of the response with frequency for frequencies such that $\mathcal{L} > \lambda/10$ or f > 6 MHz which we refer to as the "standing wave region" of the response.

Note that the experimental results approach the D-C common impedance levels given in Table I as the frequency goes to zero (frequency $\lesssim$ 30 KHz). In all cases the D-C level was measured by reducing the signal generator

TABLE I

Common Impedance Coupling Contribution to V_0

	Single Reference Wire Fig. 3-1a	Ground-Signal-Ground Fig. 3-1b
R = 50	9.72 mV	.972 mV
R = 1K	.486 mV	.0486 mV



Single Reference Wire Configuration
R = 50Ω

Figure 3-4a

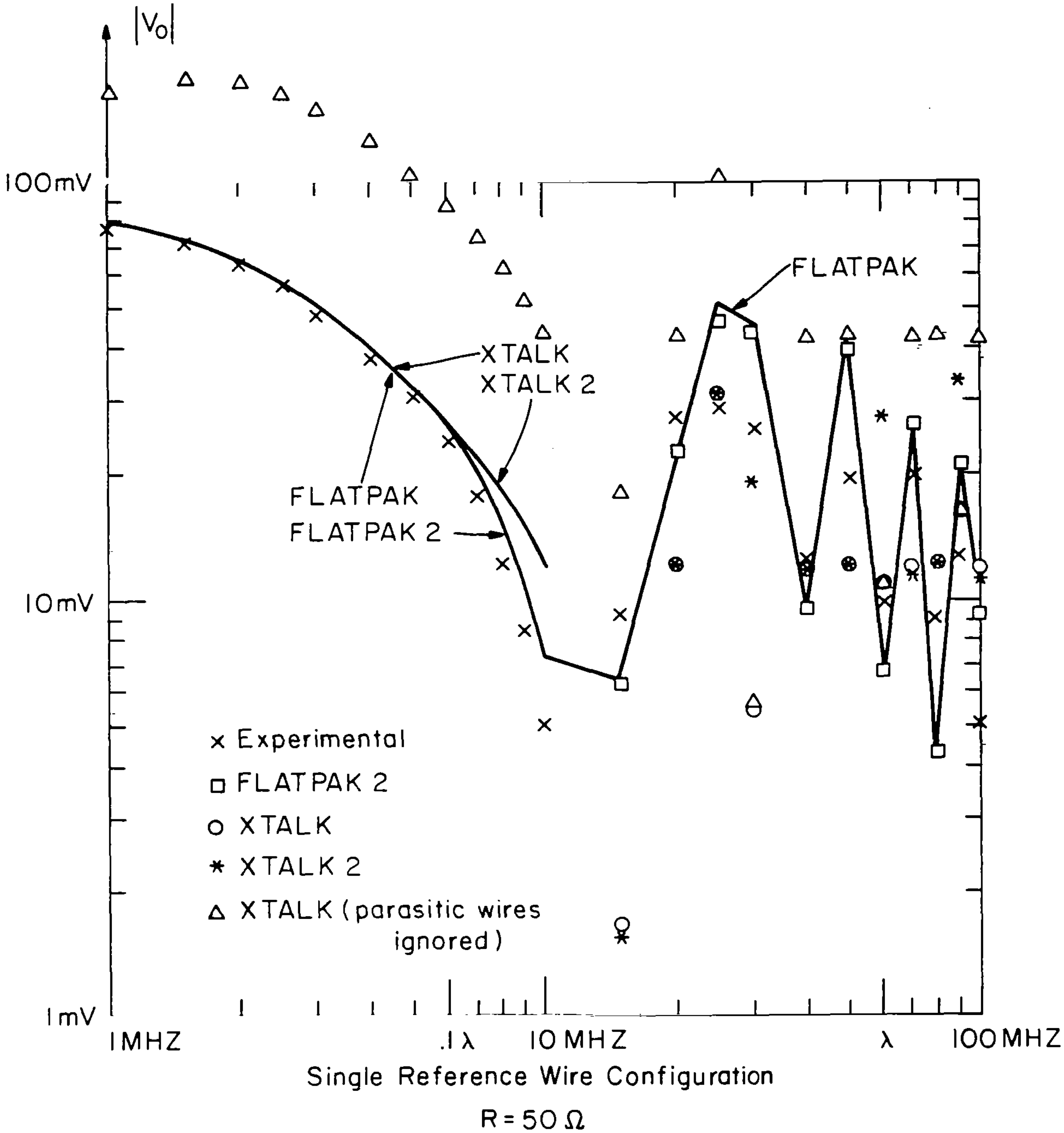

|Vo|
100mV
10mV
1mV
FLATPAK
XTALK
XTALK 2
FLATPAK
FLATPAK 2
× Experimental
□ FLATPAK 2
○ XTALK
* XTALK 2
△ XTALK (parasitic wires ignored)
1MHZ
.1 λ
10 MHZ
λ
100MHZ
Single Reference Wire Configuration
R = 50 Ω

Figure 3-4b.

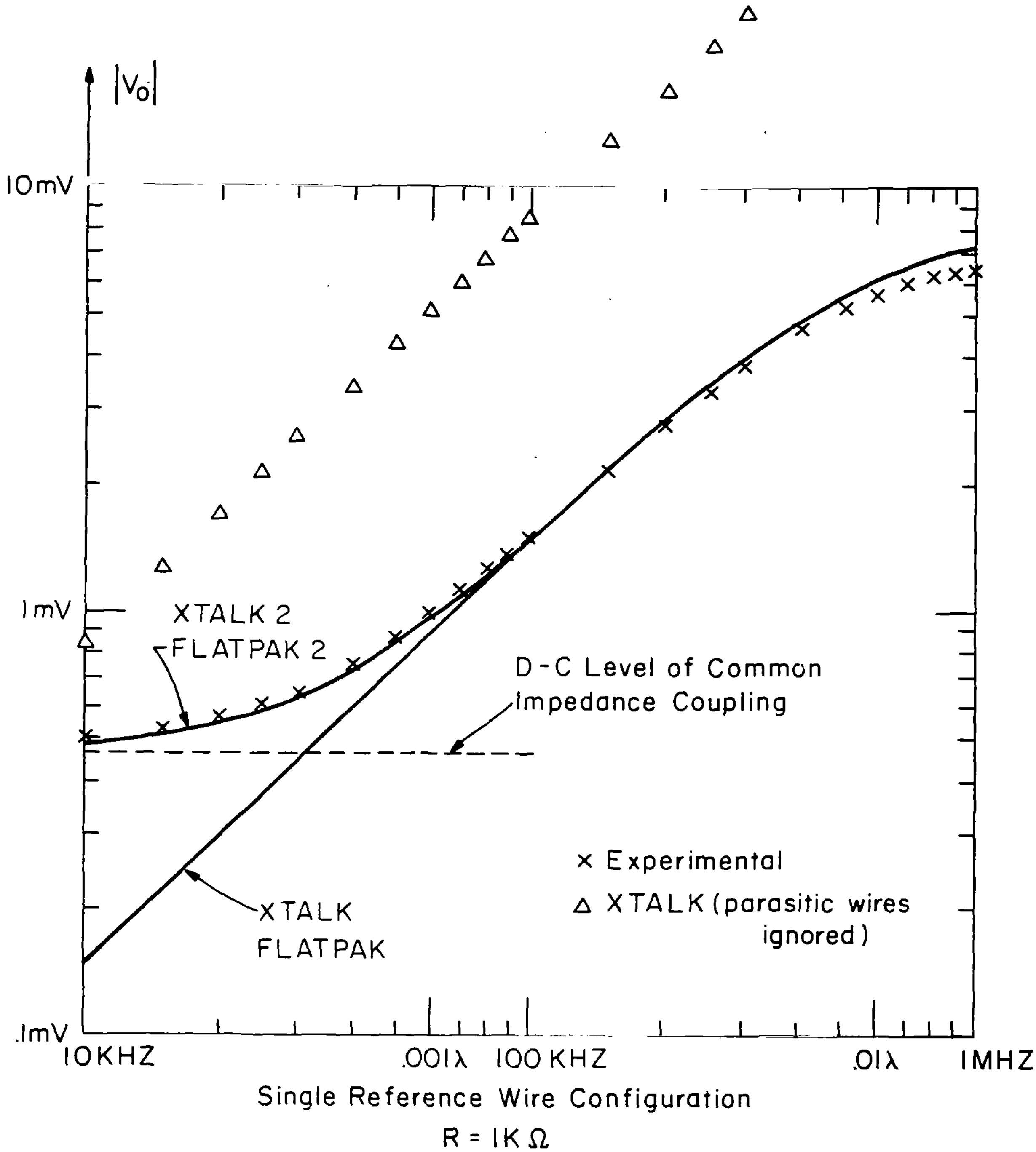
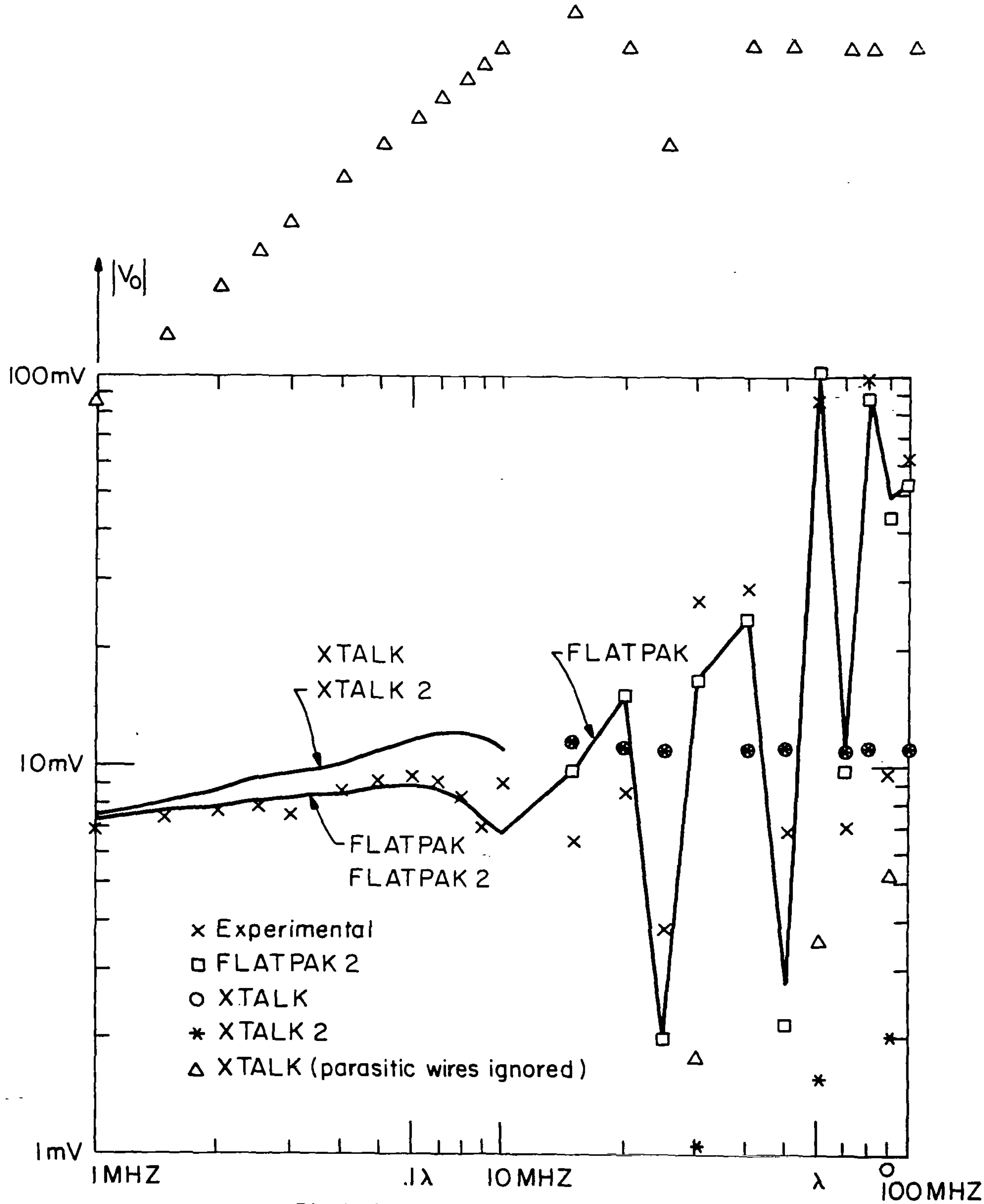

|Vo|
10mV
1mV
.1mV
10KHZ
.001λ
100KHZ
.01λ
1MHZ
XTALK 2
FLATPAK 2
D-C Level of Common
Impedance Coupling
XTALK
FLATPAK
× Experimental
△ XTALK (parasitic wires
ignored)
Single Reference Wire Configuration
R = 1K Ω

Figure 3-5a.



Single Reference Wire Configuration

R = 1K Ω

Figure 3-5b

frequency to ~ 5 Hz and the levels correspond very closely to those in Table I. Notice that in both cases, for frequencies less than 100 KHz, XTALK2 and FLATPAK2 both accurately predict the experimental results (within 10% or approximately 1 dB) yet XTALK and FLATPAK do not predict the common impedance coupling and continue downward at 20 dB/decade. The prediction error in dB is given by

$$\text{ERROR (dB)} = 20 \log_{10} \frac{V_{0\ \text{Predicted}}}{V_{0\ \text{Experimental}}} \tag{3-9}$$

This clearly indicates that in the low frequency range, one must include the wire impedances. In the higher frequency range it appears unnecessary.

For frequencies between 100 KHz and 10 MHz, all four programs predict the response quite accurately.

In the standing wave region consisting of frequencies such that $\mathcal{L} > \frac{1}{10}\lambda$, only FLATPAK and FLATPAK2 provide any prediction of the experimental results and these are generally within 6 dB except at 90 MHz in Figure 3-5. Considering the sensitivity of the results when the line is not electrically short, this is quite remarkable. Note however that XTALK and XTALK2 provide virtually no prediction in this frequency range. This seems to indicate that accurate prediction of the results when $\mathcal{L} > \frac{1}{10}\lambda$, requires that one must include consideration of the wire insulation.

As a final point, we have shown the prediction of XTALK with the parasitic wires ignored. Notice for R = 50Ω in Figure 3-4, there is some overprediction yet in Figure 3-5 there is a very large error (on the order of 20dB - 30dB overprediction). This clearly shows that the parasitic circuits in the cable cannot be ignored in the prediction. This conclusion is also supported by the results in [8].

3.4 The Ground-Signal-Ground Configuration

Virtually the same conclusions can be reached for the Ground-Signal-Ground Configuration as were reached for the Single Reference Wire Configuration. The results for R = 50Ω are shown in Figure 3-6 and those for R = 1KΩ are shown in Figure 3-7.

Two outstanding differences are noted. For the Ground-Signal-Ground Configuration, the parasitic wires dramatically affect the results. The predictions of XTALK with the parasitic wires ignored are in both cases as much as 40 dB above the experimental results. Actually this is intuitively to be expected and seems to be one of the main reasons for using the Ground-Signal-Ground Configuration over the Single Reference Wire Configuraion.

The second difference between these results and the previous results is that the level of common impedance coupling has been substantially reduced by a factor of 10 (10 wires in parallel now form the "reference wire"). Note that the common impedance coupling level is precisely that shown in Table I. This is an added benefit of the Ground-Signal-Ground Configuration since the common impedance coupling could easily be the main contributor to interference problems, especially if the signals in a logic circuit have "slow" rise times, i.e., only low frequency components.

Again the influence of conductor losses and wire insulation on the prediction accuracies are the same as in the Single Reference Wire Configuration.

3.5 Summary

A summary of the prediction accuracies and influence of conductor losses, wire insulation and parasitic wires for both configuration is shown in Table II. The only difference between the two configurations as far as this

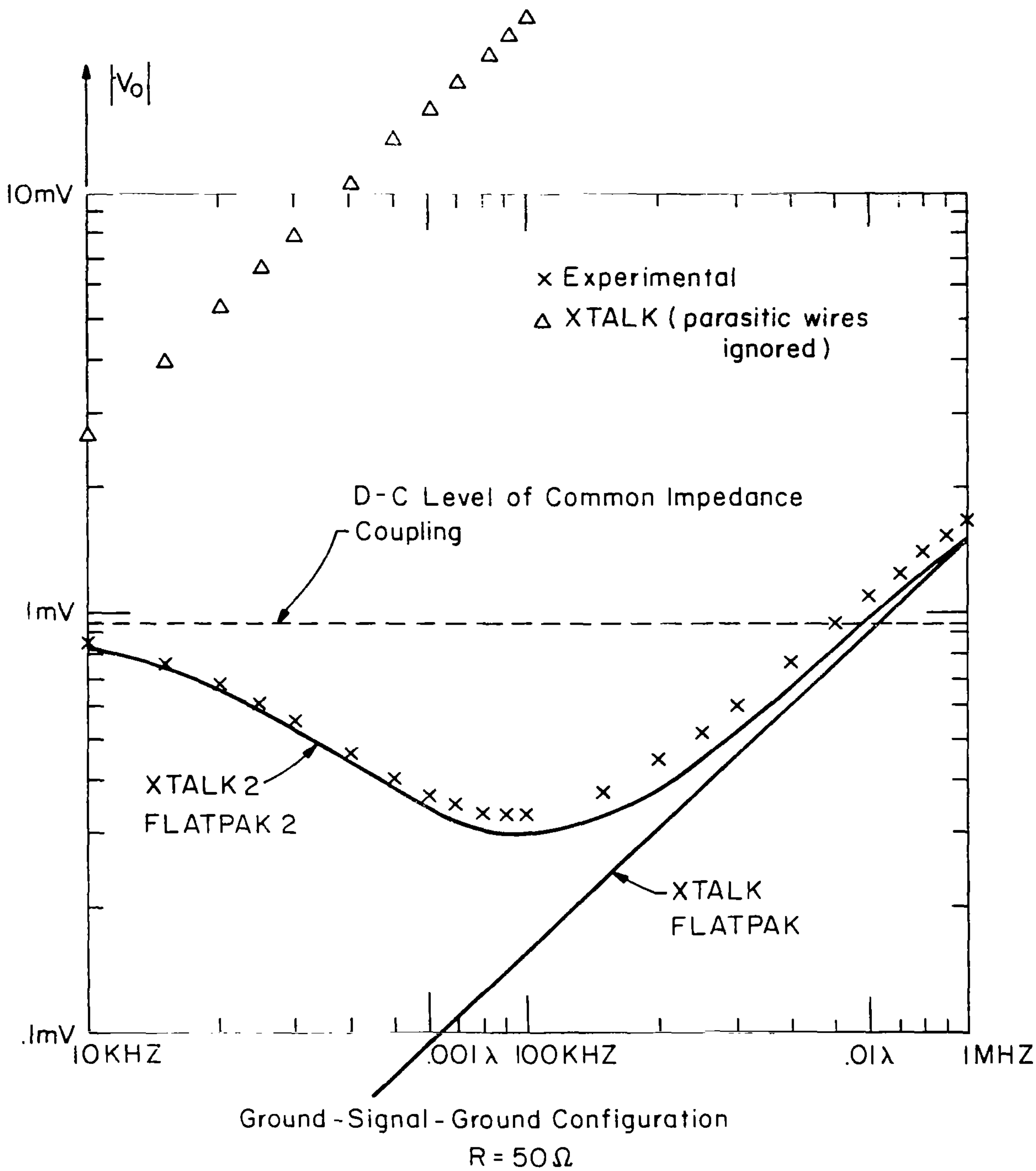
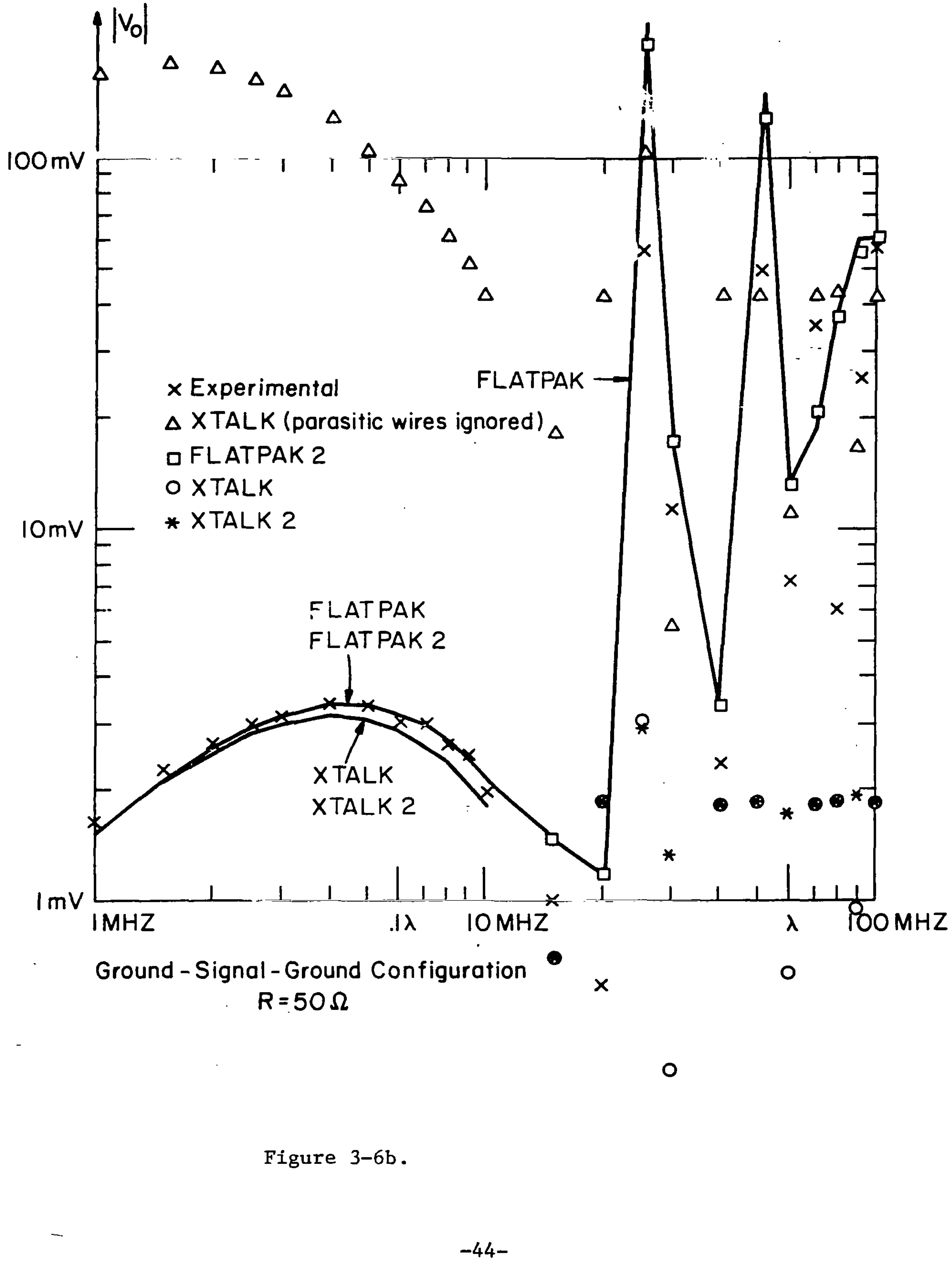

|Vo|
10mV
1mV
.1mV
× Experimental
△ XTALK (parasitic wires ignored)
D-C Level of Common Impedance Coupling
XTALK 2
FLATPAK 2
XTALK
FLATPAK
10KHZ
.001λ
100KHZ
.01λ
1MHZ
Ground - Signal - Ground Configuration
R = 50Ω

Figure 3-6a.



|Vo|
100mV
10mV
1mV
× Experimental
△ XTALK (parasitic wires ignored)
□ FLATPAK 2
○ XTALK
∗ XTALK 2
FLATPAK
FLATPAK
FLATPAK 2
XTALK
XTALK 2
1MHZ
.1λ
10MHZ
λ
100MHZ
Ground-Signal-Ground Configuration
R=50Ω

Figure 3-6b.

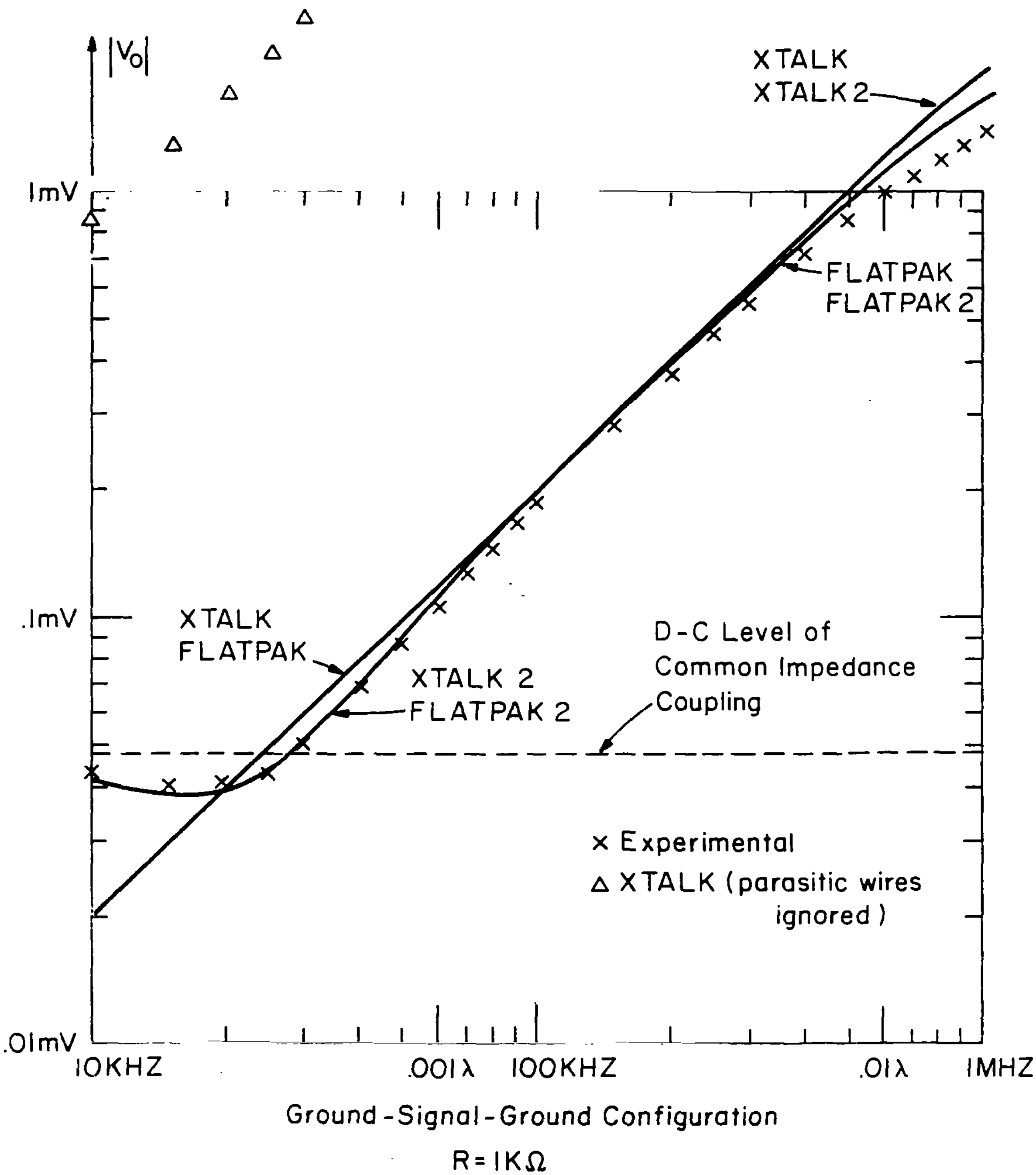

|Vo|
1mV
.1mV
.01mV
XTALK
XTALK 2
FLATPAK
FLATPAK 2
XTALK
FLATPAK
XTALK 2
FLATPAK 2
D-C Level of
Common Impedance
Coupling
× Experimental
△ XTALK (parasitic wires
ignored)
10KHZ
.001λ
100KHZ
.01λ
1MHZ
Ground-Signal-Ground Configuration
R = 1KΩ

Figure 3-7a.

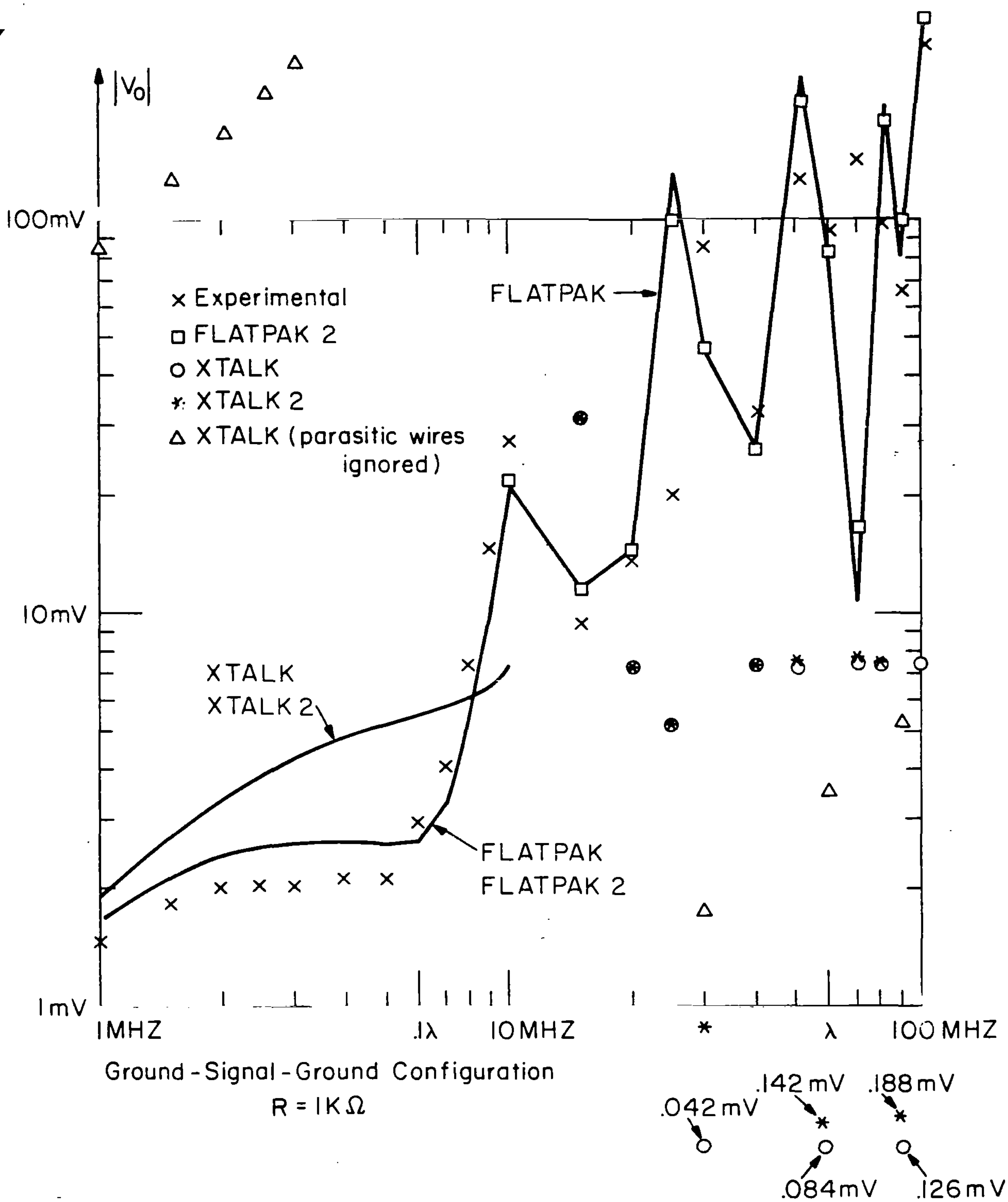

|V0|
100mV
10mV
1mV
× Experimental
□ FLATPAK 2
○ XTALK
✱ XTALK 2
△ XTALK (parasitic wires ignored)
FLATPAK
XTALK
XTALK 2
FLATPAK
FLATPAK 2
1MHZ
.1λ
10MHZ
λ
100 MHZ
Ground-Signal-Ground Configuration
R = 1KΩ
.042 mV
.142 mV
.188 mV
.084 mV
.126 mV

Figure 3-7b.

Table is concerned is in the frequency f_{CI} at which the D-C common impedance coupling level and the level of the electromagnetic coupling (assuming perfect conductors) are equal. This frequency is summarized by:

f_{CI}	Single Reference Wire	Ground-Signal-Ground
R = 50	35 KHz	600 KHz
R = 1K	30 KHz	25 KHz

For all the situations considered, we find that for $f < f_{CI}$ XTALK2 and FLATPAK2 yield virtually identical results and XTALK and FLATPAK yield virtually identical results. For $f > f_{CI}$, we find that XTALK and XTALK2 yield virtually identical results and FLATPAK and FLATPAK2 yield virtually identically results. (Actually, XTALK and XTALK2 differ at 30 MHz, 60 MHz, 90 MHz, i.e., multiples of $\lambda/4$, but this is to be expected since the variation of the response with frequency is most rapid at these frequencies [8].)

The results in Table II would suggest that an efficient computational procedure would not be to use one single program but to use several programs over the different frequency ranges. A possibility would be to first compute the level of common impedance coupling from (3-1), (3-4) and Section 2.3. Then compute the response using most efficient of the four programs, XTALK, for frequencies starting at $f|_{\mathcal{L} = \frac{1}{10}\lambda}$ and decreasing down to f_{CI} at which the XTALK result equals the common impedance coupling level. Above $f|_{\mathcal{L} = \frac{1}{10}\lambda}$ use FLATPAK. A more refined estimate would be to use XTALK2 to compute the response at frequencies around f_{CI}. (For our results this would only be truly necessary for R = 50 Ω and the Ground-Signal-Ground Configuration shown in Figure 3-6b.) This would seem to provide the most computationally efficient

TABLE II

A Summary of the Prediction Accuracies

Frequency Range $f \triangleq$ frequency	Common Impedance Coupling Dominant Region $0 < f < f_{CI}$	$f_{CI} < f < f\|_{\mathcal{L}=1/10\lambda}$	$f > f\|_{\mathcal{L}=1/10\lambda}$
Can Accurate Predictions be Achieved? (Representative Prediction Accuracy)	YES (± 1 dB)	YES (± 1 dB)	YES (± 6 dB)
Must Conductor Losses be Considered?	YES	NO	NO
Must Insulation Dielectric be Considered?	NO	NO	YES
Must Parasitic Circuits be Considered?	YES	YES	YES

scheme which would result in acceptable prediction accuracies and is summarized in Figure 3-8. The alternative is to use FLATPAK2.

To illustrate the relative complexities of the four programs, the compile and execution times and required memory for the object deck and the arrays for the 20 wire ribbon cable problem are shown in Table III. The programs used the IBM 370/165 machine and the WATFIV compiler and all computations were in double precision arithmetic. These times will vary with other machines and are also variable on the IBM 370 in that this machine operates in a multi-processing mode so that the same job run at different times will require slightly different cpu times. The memory requirements are shown in bytes. The compile and execution times are shown in seconds and the execution times are for a computation for 10 frequencies in the run. The per-frequency computation times can be approximately computed by dividing the execution times by the number of frequencies in the run (10).

It should be clear that the most efficient computation is neglecting the parasitic wires. We found, however, that virtually no valid predictions could be obtained so this must be discarded. Of the four programs which consider all the wire interactions, XTALK is certainly the most efficient in terms of computation time and required memory. Both XTALK2 and FLATPAK2 require approximately 25 times the per-frequency computation time for XTALK and double the required array memory. (The object decks require approximately three times the memory for XTALK or FLATPAK.) When many frequencies are being considered, this results in a substantial savings. For the 45 frequencies computed, XTALK would require a total computation

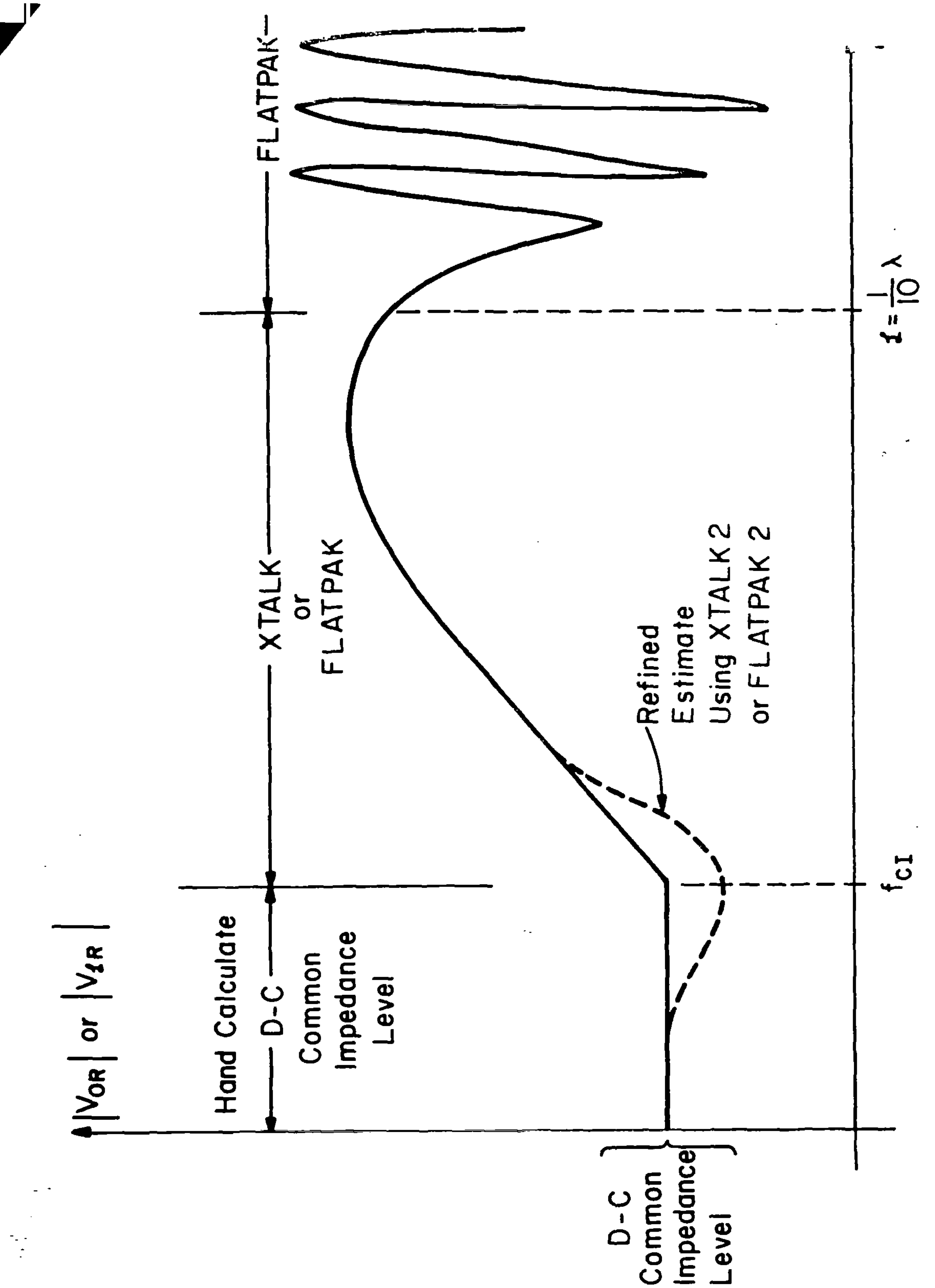


Figure 3-8. A summary of the optimum prediction process.

TABLE III

Execution Times and Memory Requirements for the

Four Programs and the 20 Wire Ribbon Cable Problem

Program	Object Code (bytes)	Array Area (bytes)†	Compile Time (sec.)	Execution Time (sec.)*
XTALK	20336	30864	.69	3.10
XTALK2	59696	61632	1.68	83.20
FLATPAK	23992	33304	.97	18.74
FLATPAK2	53296	64128	1.26	80.73

† The required array area depends on n where the cable consists of (n+1) wires. Here n = 19. For other problems, the required array area in bytes can be approximately computed from (P is the precision; P = 4 for single precision and P = 8 for double precision):

XTALK: $\text{Array} = P * \left[10n^2 + 12n\right]$

XTALK2: $\text{Array} = P * \left[20n^2 + 23n\right] + 2n$

FLATPAK: $\text{Array} = P * \left[11n^2 + 9n\right]$

FLATPAK2: $\text{Array} = P * \left[21n^2 + 21n\right]$

* The total number of frequencies run is 10. Therefore the approximate per-frequency computation time can be obtained by dividing these times by 10.

time of approximately 14 seconds whereas XTALK2 and FLATPAK2 would require approximately 370 seconds or 6 minutes. Notice that the FLATPAK program provides some savings over XTALK2 and FLATPAK2 in that the per-frequency computation time and required array memory are approximately 1/4 and 1/2, respectively, over those for the XTALK2 and FLATPAK2 programs. If one is not concerned about computation time, the author would suggest using only the FLATPAK2 program.

IV. SUMMARY AND CONCLUSIONS

The prediction of crosstalk in ribbon cables was investigated. Based on the experimental configurations tested, it would appear that accurate predictions of crosstalk can be achieved in controlled characteristic cables such as these. The prediction accuracies are typically within ± 1dB for frequencies such that the line is electrically short ($\mathcal{L} < \frac{1}{10}\lambda$).

It was found that the parasitic wires can have a significant effect (as much as 40dB for all frequencies) on the coupling between a generator and receptor circuit. Therefore to achieve accurate predictions in ribbon cables, one must consider the interactions between all wires in the cable.

The wire insulation evidently can be ignored when the line is electrically short ($\mathcal{L} < \frac{1}{10}\lambda$) but cannot be ignored for higher frequencies. Conversely the impedance of the reference wire cannot be ignored for low frequencies where the common impedance coupling dominates the electromagnetic field coupling.

In view of these observations, a suggested approach to the prediction of crosstalk which maximized computational efficiency was to use a program which neglected conductor losses but included wire insulation for frequencies where the line is electrically long ($\mathcal{L} > \frac{1}{10}\lambda$). Compute the D-C level of common impedance coupling by hand and use this as the level up to the frequency f_{CI} at which it equalled the electromagnetic field coupling. For frequencies between these two extremes, i.e., $f_{CI} < f < f|_{\mathcal{L} = \frac{1}{10}\lambda}$ use a program which ignores both conductor resistance and wire insulation.

Although these conclusions cannot be generalized absolutely, they are supported by other results [8]. The alternative is to use a more complicated program which does not neglect wire insulation and conductor losses such as FLATPAK2.

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