

# *MARATHA*

USER'S MANUAL, 1973 ADDENDUM

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\$3.00

*MARTHA* users at M.I.T. will find the workspaces comprising *MARTHA* in public library no. 100 in the APL system that runs under TSO. This service is provided by the M.I.T. Information Processing Center.

For *MARTHA* users outside M.I.T., the means of accessing *MARTHA* are different. Users or prospective users may direct comments or inquiries to

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*MARTHA* users will require for reference both this 1973 Addendum and the original *MARTHA* User's Manual. Copies of *MARTHA* User's Manual (\$6.95, postpaid) may be obtained from The MIT Press, 28 Carleton Street, Cambridge, Massachusetts 02142. Copies of both the Manual and this 1973 Addendum are available from the representative listed above.

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## I. INTRODUCTION

*MARTHA* is a notation[1] for denoting electrical networks, and it is also a network-analysis computer program[2] which uses this notation.

As a notation[1], *MARTHA* serves the same purpose as a schematic diagram. It can be used for communication from one person to another (for example, for documentation purposes), or from a person to a computer (for example, the input to analysis programs), or from a computer to a person (for example, the output from synthesis or optimization programs). The *MARTHA* notation is concise and readable, and in a "character string" form well suited for communication to and from a computer. However, it lacks the "two-dimensional" graphic appeal of a schematic diagram, and so should not be regarded as a substitute for schematic diagrams in all circumstances.

As a computer program[2; 3], *MARTHA* is written in, and imbedded in, the general-purpose interactive language APL, and its syntax resembles that of APL in some ways. *MARTHA* is designed for ordinary use by a user who does not know any APL. Indeed, knowledge of circuits is far more important than knowledge of any computer language.

The program *MARTHA* has been available at M.I.T.[4; 5; 6] since 1970, and is also available on several commercial time-sharing computer services, with local telephone access in more than 70 cities in the United States, Canada, and Europe.

Different versions of *MARTHA* are identified by the last two digits of the year written on the top of all on-line documentation, and as part of a heading for *MARTHA* output. Versions dated 71 and 72 are adequately described in the *MARTHA* User's Manual[2]. Versions dated 73 include major improvements in capabilities, described in this 1973 Addendum to the *MARTHA* User's Manual. If you are using any version dated 73, you will need for reference both the *MARTHA* User's Manual[2] and this Addendum.

The changes discussed in this Addendum are all "upward compatible" in the sense that any normal usage of *MARTHA* which previously worked, will work now.\* The changes are in three categories. First are new options in defining networks and the network environment, described in Section II. Second are additions to existing *MARTHA* library workspaces, Section III. Third are two additional *MARTHA* library workspaces, Section IV.

Some behind-the-scenes improvements and tips on reducing computation cost are given in Section V, and finally some advanced examples illustrating the new features appear in Section VI. Advance notice of some future changes to *MARTHA* is given in Section VII so you can avoid features scheduled to be removed.

In the continuing evolution of the *MARTHA* notation and the program, the author has been helped by several people. Significant parts of the programming, or programming ideas, came from M. A. Picheny and R. R. Gillock. Ideas for improvements have come from many people, mostly *MARTHA* users; I am especially indebted to R. F. Bauer, L. M. Breed, W. B. Lurie, and D. F. Peterson[7]. During the work, discussions with

---

\* The only exception is that the two previous functions *WDUAL1* and *WDUAL2* are replaced by a single function *WDUAL*. Also, numerically-defined functions of frequency (FOF's) defined in earlier versions must be converted in form; for details, see Section IV-B.

R. F. Bauer[8], H. J. Carlin[9], R. E. Crochiere, and R. Spence were helpful. Among those who unwittingly helped define improvements were the following who participated in projects which sometimes demanded increased capabilities: R. W. Chick, M. Greenspan[10], P. Hirsohn[11], S. Lazović[12], B. E. Peetz, D. H. Steinbrecher, and K. I. Thomassen[10]. I am also indebted to several people for permission to use their designs as examples: J. A. Kong, S. Lazović, B. W. Leake, R. Levy, W. B. Lurie, D. F. Peterson, D. H. Steinbrecher, and G. Szentirmai. Computer facilities for this work were supported by the Department of Electrical Engineering, M.I.T.

## II. GENERAL IMPROVEMENTS

### A. Simplified Network Definitions

The expression for a network definition is unchanged, but the manner in which it is used may be different. It is no longer necessary to use an APL function when defining a network. Now you can simply execute the statements in question. (Previously, this was impossible before the frequency vector *F* was defined.)

For example, Example 2 on page 30 of the *MARTHA* User's Manual[2] starts off with the definition of the filter named *CHEB* as follows:

```
∇Z←CHEB
[1] Z←(C 34.817E-6) WC(WS L 0.07618) WC(C 45.381E-6) WC(WS L 0,07618) WC C 34.817E-6
[2] ∇
    ZG←100
    ZL←100
    F←8×125
    PLOT IG, DB IG OF CHEB
```

Although this is still allowable, it is faster and usually more convenient to dispense with the function definition, and start off that example as follows:

```
CHEB←(C 34.817E-6) WC(WS L 0.07618) WC(C 45.381E-6) WC(WS L 0,07618) WC C 34.817E-6
ZG←100
ZL←100
F←8×125
PLOT IG, DB IG OF CHEB
```

Because of this improvement, most *MARTHA* users do not need to define APL functions. For networks with an adjustable parameter, or an argument, function definitions are still convenient.

In earlier versions of *MARTHA*, the wiring functions such as *WC* actually performed the necessary calculations, and made use of the frequency vector *F*. Now, however, these functions have a "delayed action". Analysis is deferred until the results of the analysis are actually required. Analysis can be done early, however, if you want (and this may be useful to reduce computation time) by using the function *NDE*. An example of the use of this function is given in Section V-B.

A network in *MARTHA* now retains information about both the element values and the topology. This information can be displayed by the function *WHATIS* from the workspace 100 *MARTHAX*. This feature is useful in case *MARTHA* is used as a language in which to express the output of a network synthesis program. In that case the output is in a form which can be analyzed, displayed, or further modified or wired together with other networks.

Certain of the functions in *MARTHA*, namely *ZSCALE* for impedance scaling, and *WDUAL*, *WAD*, and *WCC*, which produce the dual, the adjoint, and the complex conjugate of a network, are referred to as wiring functions, even though they do not actually do any wiring. These functions now actually get "inside" the network and change element values and/or topology as necessary. For example, *ZSCALE* leaves the topology alone but changes element values appropriately. The result is a new network which again can be displayed (using *WHATIS*), analyzed, or further modified or wired. Some new "wiring functions" in the workspace 100 *MARTHAW* act sim-

ilarly; these include frequency scaling and other frequency transformations.

### B. Easier User-Defined Elements

A user-defined element is produced by the function *UDE* in the workspace 100 *MARTHAE*. Previously, any APL vector starting with 9 was assumed to be a user-defined element. This still works temporarily, but you are advised to avoid it because it will not work in future versions of *MARTHA*. The preferred usage is now as follows: prepare a vector containing all the parameters of the element, and, possibly, a code number denoting what type of element it is. Example:

```
ELT←UDE 2 17.35 12
```

defines a user-defined element, named *ELT*, with the parameter vector 2 17.35 12. The obsolete technique, which still temporarily works, is *ELT←9, 2 17.35 12*.

So far no calculations have been done. During analysis, when the user-defined element is encountered, a function named *NEWELEMENT* is automatically called. Its argument would be, for the example above, the original vector\* 2 17.35 12. This function, which you must write, is supposed to return the ABCD matrix of the element, when called during analysis. Now, you have more options of what to return. The value returned may be a matrix with as many rows as there are frequencies, and either 2, 6, or 8 columns. For eight-column matrices, as previously, the columns are to be the real and imaginary parts of A, the real and imaginary parts of B, the real and imaginary parts of C, and the real and imaginary parts of D, in that order. For reciprocal elements, the last two columns containing D may be omitted. And for one-port elements, a two-column matrix may be returned, containing the real and imaginary parts of the impedance.

The user-supplied function *NEWELEMENT* may, instead of returning a matrix of one of the shapes given above, return any *MARTHA* network, using any normal *MARTHA* wiring functions or elements, including other user-defined elements, and possibly incorporating one or more matrices of the forms given above as elements in that network.

A user-defined element is always assumed to be a two-port element\*\*, and if a two-column matrix is returned, the resulting one-port element is operated on by *WP*.

### C. General Forms for Network Environment

Previously, values of generator impedance *ZG*, load impedance *ZL*, and normalization impedances *ZN*, *ZNIN*, and *ZNOUT* could be numbers, either positive or negative, but could not depend on frequency, nor be complex. These restrictions are now removed, as follows. These five variables may each be any of the following:

1. A single number, positive, negative, or in some cases zero;

---

\* If the obsolete technique *ELT←9, 2 17.35 12* is used, then the 9 is included in the argument.

\*\* The only time this can possibly lead to difficulty is if a one-port user-defined element is operated on immediately by *WDUAL*, for example 2 *WDUAL UDE 2 73 5*. In that case *WDUAL* cannot tell that a one-port user-defined element is intended. Instead, use *WTO* explicitly, for example, 2 *WDUAL WTO UDE 2 73 5*.

the resulting impedance is independent of frequency.

2. A one-column or two-column FOF; the resulting impedance is frequency dependent; for one-column FOF's it is real and for two-column FOF's the two columns are interpreted as the real and imaginary parts of a complex impedance. The defining frequencies of the FOF need not coincide with  $F$ ; the values will be interpolated or extrapolated when necessary.

3. Any one-port network defined in *MARTHA* notation. In this case, the impedance of the network is evaluated and used. (If a two-port network is used, the output port is open-circuited, so that  $Z_{11}$  is used.) The network may contain any element or wiring function in *MARTHA*, including user-defined elements.

A form different from one of the above will lead to an error and, usually, the error message *NOT A SECTION*.

This new ability is useful in several ways. The most natural normalization impedance for waveguide systems is the frequency-dependent characteristic impedance of the guide. Example:

```
ZNIN←ZNOUT←WG 6.3E9 377
```

Actual generators and loads can be modeled more accurately than before. You can use *MARTHA* to find some voltages and currents inside the network, by incorporating a portion of the network into the load or generator. This applies also to calculation of power gain or power flow at interior points. Finally, experimental data characterizing a load or generator can be used.

The algorithms in *MARTHA* for calculating response functions account for complex, frequency-dependent values of  $Z_G$ ,  $Z_L$ ,  $Z_N$ ,  $Z_{NIN}$ , and  $Z_{NOUT}$ . The definitions of the wave variables  $A$  and  $B$  in terms of voltage  $V$  and current  $I$  are generalized to the form[13]

$$A = \frac{V + Z_n I}{2 \sqrt{|\operatorname{Re} Z_n|}} \quad (1a)$$

$$B = \frac{V - Z_n^* I}{2 \sqrt{|\operatorname{Re} Z_n|}} \quad (1b)$$

and similarly for  $A_1$  and  $B_1$  in terms in terms of  $V_1$ ,  $I_1$ , and  $Z_{nin}$ , and  $A_2$  and  $B_2$  in terms of  $V_2$ ,  $I_2$ , and  $Z_{nout}$ .

The function *SFOF*, which also makes use of  $Z_N$  or  $Z_{NIN}$  and  $Z_{NOUT}$ , interprets them in accordance with Equation (1) when they are negative or complex. No change is involved when they are real and positive.

#### D. Response Functions Are FOF's

The result of an output list is a numerically defined function of frequency (FOF) which can, of course, be printed or plotted. It can also be used to specify  $Z_G$ ,  $Z_L$ , etc., or be calculated upon by the functions in the new workspace 100 *MARTHAN* (see Section IV-B). Example:

```
ZL←Z12 OF NETWORK
```

sets the load impedance to the output image impedance of a network.



### III. LIBRARY ADDITIONS

#### A. New Elements

The workspace 100 *MARTHA*E contains a collection of elements for *MARTHA*. The new elements are described here. A complete list of *MARTHA* elements starts on page 78.

Waveguides. The functions *WGATTEN* and *WGISOLATOR* are similar to the functions *ATTENUATOR* and *ISOLATOR*, except that the first number in the argument of those functions is to be replaced with two numbers, namely the cutoff frequency in Hz and infinite-frequency characteristic impedance of the guide in ohms. Thus, the argument for *WGATTEN* must be of length 3, consisting of the cutoff frequency, infinite-frequency impedance, and attenuation in dB. The argument for *WGISOLATOR* must include the cutoff frequency and infinite-frequency impedance, followed by either no other number, or one or two, with the same interpretation as for the function *ISOLATOR*.

Waveguides with TM modes (propagating or cutoff) can now be analyzed by *MARTHA*. The element *TM* acts just like *WG* except that the characteristic impedance is proportional to the reciprocal of the guide wavelength, rather than proportional to the guide wavelength itself. In particular, if *TM* has an argument of length 2, the result, a one-port, is the characteristic impedance of the mode. If the argument has length 3, the result is a two-port length of waveguide. The functions *DEGREESAT*, *FORDIEL*, and *WAVESAT* may be used with *TM*.

Powers of Frequency. Four new functions are useful for models involving powers of frequency. Two of these, *ZPDE* and *YPDE*, can be used to generate one-port impedances (or admittances) of the form  $Ks^n$ , where  $s = j\omega = j2\pi f$ . If the argument for *ZPDE* has length 2, then it is assumed that  $\bar{Z}(s) = Ks^n$ , and the first number is taken to be  $n$ , and the next number  $K$ . The exponent  $n$  may be any integer between -5 and 5, inclusive. For  $n = 0$  the result is a resistor; for  $n = 1$  it is an inductor, and for  $n = -1$  it is a capacitor. Similarly, if the function *YPDE* has an argument of length 2, then the result is a one-port element with admittance  $Y(s) = Ks^n$  where  $n$  is the first number in the argument, and  $K$  the second.

For two-port elements of a similar nature, the functions *ZPDE*, *YPDE*, *HPDE*, and *ABCDPDE* are used with arguments with length 3, 4, or 5. In every case the first number in the argument is the exponent  $n$ , which must be an integer with magnitude  $\leq 5$ . Reciprocal and symmetric elements need only three numbers in their arguments; reciprocal elements require four numbers, and nonreciprocal elements require five, in each case including the exponent. Table I shows in detail how the arguments are interpreted in all possible cases.

#### B. New Wiring Functions

A complete list of the wiring functions in *MARTHA* appears on pages 81 and 82.

The "wiring functions" *WAD*, *WCC*, and *WDUAL* are not new. They have the same overall effect as before (*WDUAL* being the replacement for both *WDUAL1* and *WDUAL2*) but, like *ZSCALE*, they actually get "inside" the network definition and change element values, element type and topology as necessary. The resulting network definition can be displayed by the

function *WHATIS* from the workspace 100 *MARTHAX*.

There are three new "composite" wiring functions, and four other new functions that will be referred to as wiring functions even though they do not do any wiring. Of these four, three are frequency transformations, and the other offers a means to check the range of validity of models.

Composite Functions. The three new wiring functions are *WTSWN*, *WTOWN*, and *WROTWROT*. The first, *WTOWN*, Figure 1, has the same effect as

Table I. Interpretation of the functions *ZPDE*, *YPDE*, *HPDE*, and *ABCDPDE* for every allowable length argument X. Here  $n = X_1$ , and  $s = j2\pi f$ .

Length of argument X	2	3	4	5
Properties of result	One-port	Two-port Reciprocal Symmetric	Two-port Reciprocal	Two-port
For <i>ABCDPDE</i> :		$A = X_2$ $B = X_3 s^n$ $C = \frac{D^2 - 1}{B}$ $D = A$	$A = X_2$ $B = X_3 s^n$ $C = \frac{X_4}{s^n}$ $D = \frac{BC + 1}{A}$	$A = X_2$ $B = X_3 s^n$ $C = \frac{X_4}{s^n}$ $D = X_5$
For <i>HPDE</i> : ( $H_{21} \neq 0$ )		$H_{11} = X_2 s^n$ $H_{12} = X_3$ $H_{21} = -H_{12}$ $H_{22} = \frac{1 - H_{12}^2}{H_{11}}$	$H_{11} = X_2 s^n$ $H_{12} = X_3$ $H_{21} = -H_{12}$ $H_{22} = \frac{X_4}{s^n}$	$H_{11} = X_2 s^n$ $H_{12} = X_3$ $H_{21} = X_4$ $H_{22} = \frac{X_5}{s^n}$
For <i>YPDE</i> : ( $Y_{21} \neq 0$ )	$Y = X_2 s^n$	$Y_{11} = X_2 s^n$ $Y_{12} = X_3 s^n$ $Y_{21} = Y_{12}$ $Y_{22} = Y_{11}$	$Y_{11} = X_2 s^n$ $Y_{12} = X_3 s^n$ $Y_{21} = Y_{12}$ $Y_{22} = X_4 s^n$	$Y_{11} = X_2 s^n$ $Y_{12} = X_3 s^n$ $Y_{21} = X_4 s^n$ $Y_{22} = X_5 s^n$
For <i>ZPDE</i> : ( $Z_{21} \neq 0$ )	$Z = X_2 s^n$	$Z_{11} = X_2 s^n$ $Z_{12} = X_3 s^n$ $Z_{21} = Z_{12}$ $Z_{22} = Z_{11}$	$Z_{11} = X_2 s^n$ $Z_{12} = X_3 s^n$ $Z_{21} = Z_{12}$ $Z_{22} = X_4 s^n$	$Z_{11} = X_2 s^n$ $Z_{12} = X_3 s^n$ $Z_{21} = X_4 s^n$ $Z_{22} = X_5 s^n$

*WTO WN*, that is, as first applying *WN* and then *WTO*. In other words, this new function open-circuits the input of a two-port network and the output port then acts like a one-port network. The advantage of using *WTOWN* is, first, that it is many times faster than applying *WN* and *WTO* separately, and, second, it can be used on networks that are unilateral, or nearly so, for which *WN* may not work properly. (Two-port networks are represented by their ABCD matrices in *MARTHA*, and a network that is backwards-unilateral has no ABCD matrix.) The second function, *WTSWN*, Figure 1, acts like *WTS WN* except that it has the same advantages as *WTOWN*. The third function, *WROTWROT*, Figure 2, acts like *WROT WROT*; in other words, it rotates a three-terminal network counter-clockwise rather than clockwise. It is twice as fast as applying *WROT* twice. Example:

*WROTWROT ( WROT B) WC WROT A*

is the fastest way to realize a Darlington transistor connection.

Setting Frequency Restrictions. An important part of any model is its range of validity, and with the new "wiring function" *FLIMITS* you can specify a range of permissible frequencies. During analysis, if any frequencies in the vector *F* lie outside that range, the warning message *INVALID FREQUENCY* is printed, but analysis continues, even at the invalid

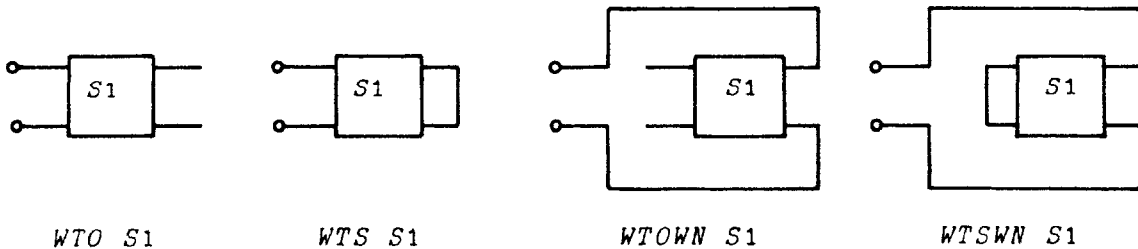


Figure 1. Two new wiring functions, *WTOWN* and *WTSWN*, compared with two previously defined wiring functions *WTO* and *WTS*.

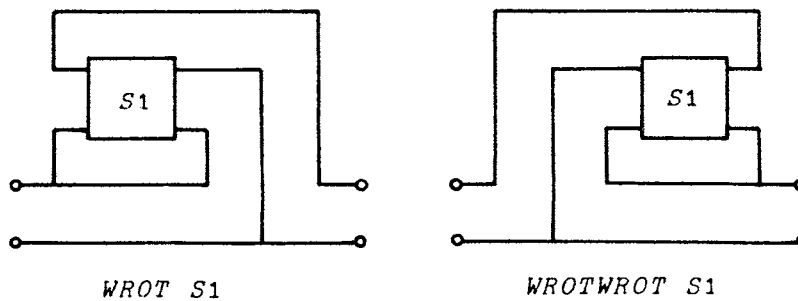


Figure 2. The new wiring function *WROTWROT* compared with the previously defined wiring function *WROT*.

frequencies. The calculations are not altered in any way by the use of *FLIMITS*. To use this function, place the limits in a vector with two numbers in it, to the left of *FLIMITS*, and the network to the right. The result is the same network with a tag giving its frequency limits. If desired, one limit may be 0 or the other any very large number such as 1E50. Example:

*NETWORK*←0 3E9 *FLIMITS NETWORK*

defines a network valid up to 3 GHz. If more than one pair of numbers appears in the left argument of *FLIMITS*, then the two lowest numbers constitute one allowed range, the next two lowest another range, etc. The setting of these limits is not undone by setting new limits. If more than one set of limits appears in a network definition, each is applied as it is encountered during analysis, and the warning message will be printed for as many sets as are violated.

Frequency Transformations. Three new "wiring functions" in the workspace 100 *MARTHAW* are the frequency transformations *FSCALE*, *FINVERT*, and *FBP*. These do not do any wiring, but instead get "inside" the network definition and alter element values, element types, and topology as necessary. The result of applying them is another network that can be analyzed, further modified or wired with other networks, or whose definition can be displayed by the function *WHATIS* in the workspace 100 *MARTHAX*. These functions, along with *ZSCALE*, are useful in generating highpass, bandpass, or bandstop filters from one-ohm, one-Hz (or one rps) prototype filters. Use of frequency transformations is covered in several reference books, e.g. [14, Sections 5.4 and 5.5; 15, Section 11-9].

The function *FSCALE* performs a frequency scaling. Its left argument is a number *N*, either positive or negative, which gives the scale factor. The topology of the network is unchanged, but capacitors and inductors are each reduced by the factor *N*. Transmission lines and waveguides are shortened by the factor *N*, and the cutoff frequency of waveguides multiplied by *N*. Numerically defined elements have their defining frequencies multiplied by *N*. The result is a network whose properties at frequency  $N \times F$  are the same as the properties of the original network at frequency *F*. This function does not work with user-defined elements. It works with both one-port and two-port networks.

The function *FINVERT* performs a frequency inversion, that is, a lowpass to highpass transformation. Its left argument is a pivot frequency *FP*. The topology of the network is left unchanged, and resistors and two-port resistive elements are not altered. Capacitors are turned into inductors and vice versa. Numerically defined elements have their defining frequencies replaced with *FP* squared, divided by the original defining frequencies. The network *FP FINVERT S1* has properties at frequency  $FP \times FP \div F$  that are numerically the same as the properties of *WCC S1* at frequency *F*. If the original network is a lowpass filter, the resulting network is a highpass filter. This function does not work with user-defined elements, transmission lines, or waveguides. It works with both one-port and two-port networks.

The function *FBP* performs a lowpass-to-bandpass transformation. Its left argument is a mid-frequency *FM* which can be either positive or negative. The topology of the network is changed, as additional elements are required. Resistors and resistive two-port elements are unchanged, but capacitors are replaced by parallel LC circuits resonant at frequency *FM*, and inductors by series LC circuits resonant at *FM*. Numerically defined elements have their defining frequencies replaced by new values above *FM* (if the defining frequencies are positive). The result is a network with properties at frequency  $(F + ((F \times F) + (4 \times FM \times FM)) \star 0.5) \div 2$  the same as the

properties of the original network at frequency  $F$ . If the original network is a lowpass filter, the resulting network is a bandpass filter. If the original network is a highpass filter, the resulting network is a bandstop filter. This function does not work with user-defined elements, transmission lines, or waveguides. It works with both one-port and two-port networks.

These frequency transformations are illustrated in the example in Section VI-C.

### C. New Response Functions

The workspace 100 *MARTHAR*, contains 34 new response functions, of which 28 are complex and 6 real. A complete list of *MARTHA* response functions appears on pages 82 to 84.

Voltages, Currents, Waves, Powers. Voltages and currents at the input and output of a network may now be calculated, with the complex response functions  $V1$ ,  $I1$ ,  $V2$ , and  $I2$ . These are all proportional to the Thevenin equivalent generator voltage  $EG$ , which can be set to any value. It is preset to one volt (rms), but you can set it to any (positive or negative) constant. Also, it can be set to any one-column (or two-column) FOF for a real (or complex) function of frequency. It may also be set to any *MARTHA* one-port network, even containing user-defined elements, in which case the voltage produced by a unit current will be used for  $EG$ . In short,  $EG$  can be specified in the same ways as  $ZG$ ,  $ZL$ ,  $ZN$ ,  $ZNIN$ , and  $ZNOUT$ . The response functions  $V1$ ,  $I1$ ,  $V2$ , and  $I2$  depend on the network and also on  $EG$ ,  $ZG$ , and  $ZL$ . The result contains more information than would be obtained by a voltmeter or ammeter, because the phase is included as well as the magnitude.

Besides the voltage and current, the input and output incoming waves  $A1$  and  $A2$  and outgoing waves  $B1$  and  $B2$  are available as complex response functions. These depend on the network,  $EG$ ,  $ZG$ ,  $ZL$ , and either  $ZNIN$  or  $ZNOUT$ . Also, the complex powers  $CP1$  and  $CP2$  entering the two-port at the input or output are available, as well as their sum  $CP$ . The complex power (real power plus  $j$  times reactive power) is the voltage times the complex conjugate of the current. For passive loads  $CP2$  will have a negative real part. The real part of  $CP$  gives the power dissipation, or the negative of the real part of  $CP$  gives the added power for amplifiers.

Special Terminating Impedances. There are six new response functions which are the terminating impedances of a two-port network under certain circumstances. These are the image impedances  $ZI1$  and  $ZI2$ , iterative impedances  $ZK1$  and  $ZK2$ , and conjugate-match impedances  $ZM1$  and  $ZM2$ . Their definitions are illustrated in Figures 3-5, or, equivalently,

$$z_{i1} = z_{11} - \frac{z_{12}z_{21}}{z_{22} + z_{i2}} \quad (2a)$$

$$z_{i2} = z_{22} - \frac{z_{12}z_{21}}{z_{11} + z_{i1}} \quad (2b)$$

$$z_{k1} = z_{22} - \frac{z_{12}z_{21}}{z_{11} + z_{k1}} \quad (3a)$$

$$z_{k2} = z_{11} - \frac{z_{12}z_{21}}{z_{22} + z_{k2}} \tag{3b}$$

$$z_{m1}^* = z_{11} - \frac{z_{12}z_{21}}{z_{22} + z_{m2}} \tag{4a}$$

$$z_{m2}^* = z_{22} - \frac{z_{12}z_{21}}{z_{11} + z_{m1}} \tag{4b}$$

In each case the impedances,  $z_{i1}$ ,  $z_{i2}$ ,  $z_{k1}$ ,  $z_{k2}$ ,  $z_{m1}$ , and  $z_{m2}$  are the terminating impedances necessary to achieve the conditions shown.

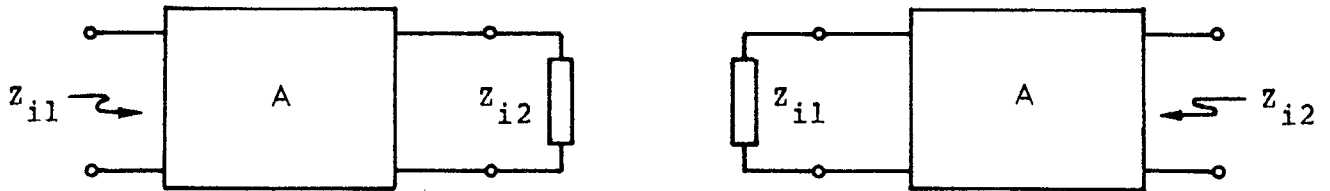


Figure 3. Definitions of image impedances for the two-port network A.

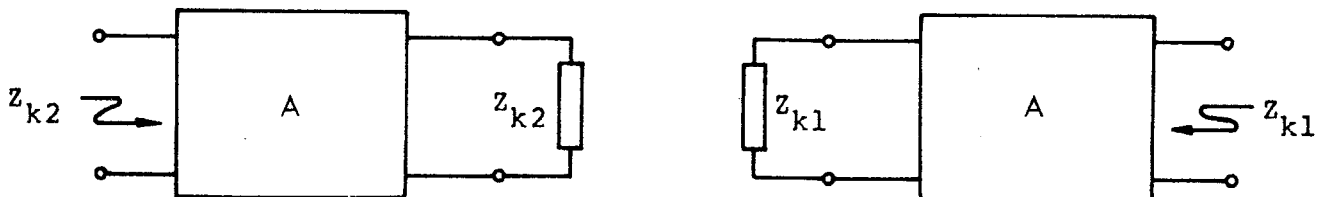


Figure 4. Definitions of iterative impedances for the two-port network A.

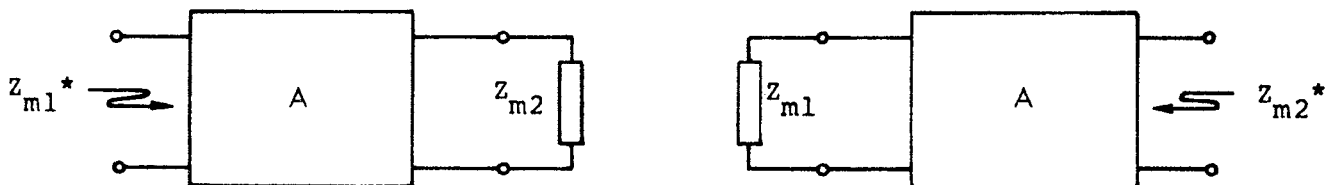


Figure 5. Definitions of conjugate-match impedances for the two-port network A.

This is the standard definition of image impedance (for example, [16, p. 269; 17, pp. 161-175]) and conjugate-match impedance [18], and is the definition of iterative impedance used by Carlin and Giordano [19, Section 4.12]. The definitions of  $Z_{k1}$  and  $Z_{k2}$  are interchanged from those given by some authors [17, p. 165; 20, p. 74].

When Equations (2), (3), or (4) are solved for  $Z_{i1}$ ,  $Z_{i2}$ ,  $Z_{k1}$ ,  $Z_{k2}$ ,  $Z_{m1}$ , or  $Z_{m2}$ , there appears a square root, which leads to an ambiguity in sign. The choice of sign made by *MARTHA* is based on a "convergence criterion" used by Johnson [21] for the iterative impedance. This criterion is derived as follows: The impedances are viewed as those resulting from an infinite cascade of networks as shown in Figure 6. This infinite cascade is imagined as replaced by a finite cascade terminated in some impedance. Consider, for example, the definition of  $Z_{K2}$ . The finite cascade for this case is shown in Figure 7. The input impedance of this finite cascade will be equal to  $Z_L$  if  $Z_L$  is either of the two possible choices for  $Z_{k2}$ , corresponding to the sign ambiguity mentioned earlier. The choice is made by inquiring about what happens for a load impedance  $Z_L = Z_{k2} + \delta Z_L$  slightly different from the exact value. The input impedance will then be  $Z_{in} = Z_{k2} + \delta Z_{in}$ , slightly different from  $Z_{k2}$ . In general, one choice of sign for the square root leads to an input impedance closer to the exact value than the load impedance, i.e.  $|\delta Z_{in}| < |\delta Z_L|$ , and the other choice to an input impedance farther away from the exact value. The choice made in *MARTHA* is for the case where the resulting input impedance is closer to the true value.\* That way, as the number of networks in the finite cascade is increased, the input impedance approaches more and more closely the *MARTHA* definition of  $Z_{K2}$ . Similar choices are made, using the "convergence criterion", for the *MARTHA* definitions of  $Z_{K1}$ ,  $Z_{I1}$ ,  $Z_{I2}$ ,  $Z_{M1}$ , and  $Z_{M2}$ . This

---

\* In a few cases, neither sign choice produces convergence of  $Z_{in}$ . The most important examples are the image and iterative impedances of lossless transmission lines, waveguides above cutoff, and lossless filters in the passband, and the conjugate-match impedances for potentially unstable networks. In those cases, *MARTHA* uses the following procedure for choosing the sign: For the image impedance  $Z_{I1}$ , its real part is chosen to be positive if the convergence test fails; if its real part is zero, then the imaginary part is arbitrarily chosen to be nonnegative. Then  $Z_{I2}$  is found from Equation (2b). For the iterative impedance  $Z_{K1}$ , the formula found by solving Equation (4a) is

$$Z_{k1} = (Z_{22} - Z_{11} \pm \sqrt{(Z_{11} + Z_{22})^2 - 4Z_{12}Z_{21}}) / 2.$$

When the convergence test fails, the radical is chosen to have positive real part, or, if its real part is zero, then the imaginary part is arbitrarily chosen to be nonnegative. The same procedure is used for  $Z_{K2}$ . Finally, the conjugate-match impedances  $Z_{M1}$  and  $Z_{M2}$  are undefined for lossless networks, and *MARTHA* produces the nonfatal error message *ATTEMPT TO DIVIDE BY ZERO*. For other networks, the formula for  $Z_{m1}$  is

$(-j \operatorname{Im} \alpha) \pm (\operatorname{Re} \alpha) \sqrt{1 - 1/k^2}$  where  $k$  is the invariant stability factor, and  $\alpha = Z_{11} - Z_{12}Z_{21}/(Z_{22} + Z_{22}^*)$ . For absolutely stable networks, for which  $|k| > 1$ ,  $Z_{m1}$  is complex and the convergence test always works. For potentially unstable networks,  $Z_{m1}$  is purely imaginary, and the convergence test always fails; in calculating  $Z_{M1}$ , *MARTHA* arbitrarily uses the positive sign above. Then  $Z_{M2}$  is calculated from Equation (4b).

criterion was discussed by Johnson[21] for the iterative impedances. Most discussions of image, iterative, and conjugate-match impedances ignore the question of sign ambiguity, or else resolve it only for passive networks.

Image, iterative, and conjugate-match impedances are supplied in *MARTHA*, but the corresponding admittances and reflection coefficients are not. However, they may be obtained easily by using appropriate modifiers. Example:

*PRINT REC Z11 , REC Z12 OF NETWORK*

produces the image admittances, where *REC* is a modifier in the workspace 100 *MARTHAM*. A new modifier *GAMMA* is described in Section III-D for converting impedances into reflection coefficients, and another new modifier *SWR* computes from reflection coefficients the resulting VSWR. Thus in effect the image, iterative, and conjugate-match impedances, admittances,

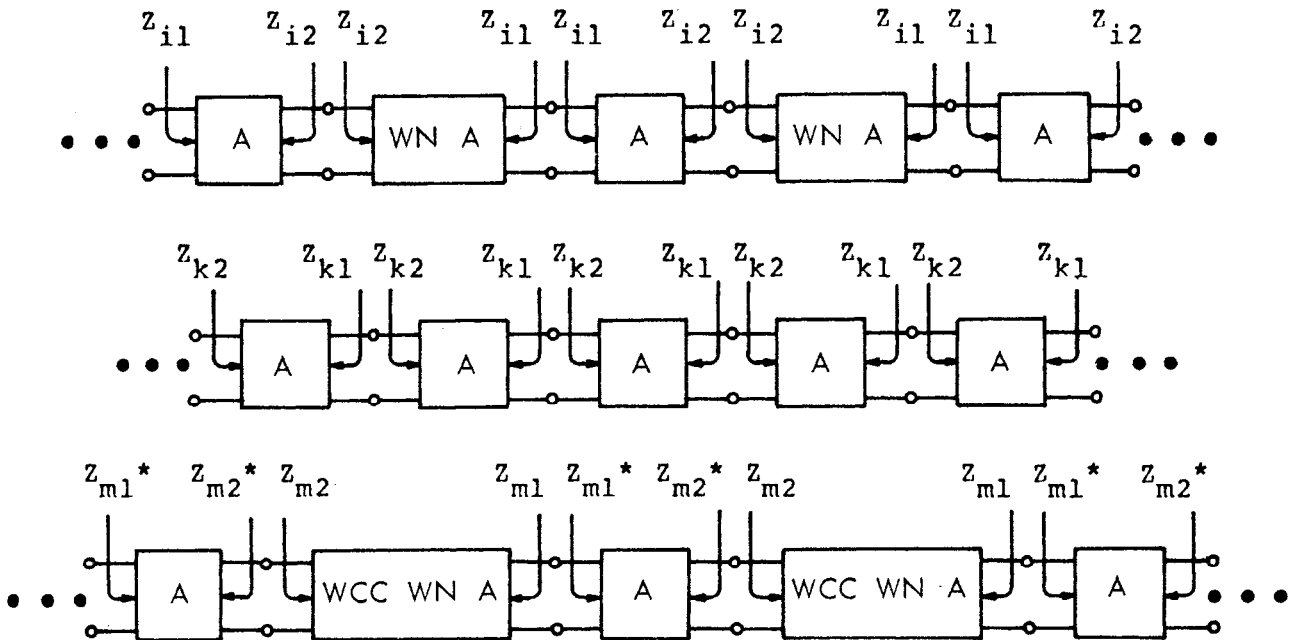


Figure 6. Alternate way of defining the six terminating impedances, as impedances seen at various reference planes in an infinite cascade of networks[17, Figures 51 and 53; 18]. In two of the three cases, networks related to *A* appear in the cascades.

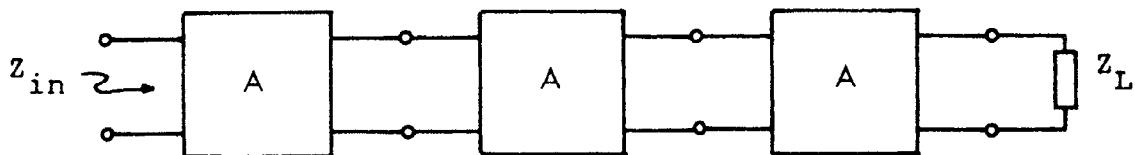


Figure 7. Finite cascade of the sort that defines  $Z_{k2}$ , shown in this case with three sections. If  $Z_L = Z_{k2}$ , then  $Z_{in} = Z_{k2}$  for any number of sections used. If  $Z_L = Z_{k2} + \delta Z_L$ , then  $Z_{in} = Z_{k2} + \delta Z_{in}$ , where  $\delta Z_{in}$  should approach zero as the number of networks in the cascade gets large. When calculating  $Z_{k2}$ , *MARTHA* chooses the sign of the square root in the formula for  $Z_{k2}$  to insure this, whenever possible.



reflection coefficients, and standing-wave ratios are all available.

Transfer Constants. The image transfer constant  $ITC$  and the iterative transfer constant (also known as the propagation constant)  $KTC$  are new response functions given by the standard definitions[16, p. 269; 17, pp. 161-175; 20, Chapter 5]

$$ITC = \frac{1}{2} \left( \ln \frac{V_1}{V_2} + \ln \frac{I_1}{-I_2} \right) \quad (5)$$

when the load impedance is equal to  $Z_{i2}$ , and

$$KTC = \ln \frac{V_1}{V_2} \quad (6)$$

when the load impedance is equal to  $Z_{k2}$ .

New Gain Functions. *MARTHA* now has several new functions that give voltage or current gains according to various definitions. The voltage gain  $VG$  in *MARTHA* uses the customary definition, namely  $V_2/E_G$  when the network is terminated by the Thevenin equivalent source, and  $Z_L$  as a load. The voltage ratio  $VR$  is  $V_2/V_1$ , and is therefore independent of the source. The current ratio  $CR$  is, similarly,  $-I_2/I_1$ . The current gain  $CG$  is analogous to  $VG$ , that is, it is the ratio of  $-I_2$  to the Norton equivalent generator of the source. Finally, the insertion voltage gain  $IVG$  is the ratio of  $V_2$  to the value  $V_2$  would have if the source were connected directly to the load. All these are complex functions. Ghausi[22, Chapter 2] has a good discussion of these functions, along with  $TG$ ,  $PG$ ,  $AG$ , and  $MG$ . His definitions are the same ones used by *MARTHA*.

Amplifier Performance Functions. *MARTHA* now contains several response functions of interest to amplifier designers. First is the unilateral gain  $UG$  defined by Mason[23]. Next is a complex quantity called the reciprocity factor  $RF$ . It is the ratio  $Z_{21}/Z_{12}$ , and therefore equals one for reciprocal networks. Its magnitude often called the maximum stable power gain[24; 25; 26, p. 214], indicates to some extent the degree to which the network is unilateral, large values implying almost unilateral networks, and values close to zero meaning networks almost unilateral backwards. The return ratio  $RR$  is a complex quantity used by Miwa, Okuno, and Namekawa[27] for amplifier design. Example:

*PLOT DEG RR VS DB RR OF AMPLIFIER*

produces the Nichols charts those authors use. Two other amplifier design procedures, those of Linvill and Schimpf[28, 29, Chapter 11], and of Spence[26, Chapter 7], employ plots in the complex plane. The Linvill-Schimpf plot is of quantities they denote  $L$  vs  $M$ ; these are the real and imaginary parts of the new complex response function  $LM$  in *MARTHA*. Spence employs a "gain plane" with the complex quantity  $UG$  divided by  $RF$ . This ratio is the new *MARTHA* response function  $SGP$  (standing for "Spence gain plane"). Example:

*PLOT RE SGP VS IM SGP OF AMPLIFIER*

produces a plot of the type used by Spence.

The stability of amplifiers is important, and *MARTHA* can calculate the two major stability factors in use. One, Rollett's invariant stability factor[25]  $ISF$ , depends only on the network. The other, Stern's stability factor[30]  $SSF$ , depends on the network and the source and load im-

pedances. In each case the result is real, and values greater than one are generally indicative of stable networks.

A gain measure of great interest to amplifier designers is the conjugate-match gain  $MG$  and its reciprocal,  $ML$ . For absolutely stable networks (with  $ISF$  greater than one)  $MG$  is equal to the power gain when  $Z_L = Z_{m2}$ , and it is also equal to the available gain when  $Z_g = Z_{m1}$ , and it is also equal to the transducer gain when  $Z_g = Z_{m1}$  and  $Z_L = Z_{m2}$ . It is called maximum available power gain[25] or ultimate gain [18]; for attenuators the intrinsic attenuation[31] is given by  $DB ML$ . For potentially unstable networks,  $Z_{m1}$  and  $Z_{m2}$  are purely imaginary, and the maximum available gain is not well defined. In *MARTHA* the definition of  $MG$  is extended so that for potentially unstable networks it is the power gain if the load is slightly lossy, but approximately equal to  $Z_{m2}$ . It is also equal to the available gain if the source impedance is slightly lossy but approximately equal to  $Z_{m1}$ .

Other Response Functions. The new response function *OMEGA* calculates the frequency in radians per second. The result does not depend on the network being analyzed, but is useful if that is the desired independent variable. Example:

```
PRINT VS OMEGA ZI1, ZIN, IG OF FILTER
```

The response function *OUTFOF* places all columns of a FOF in the output list, treating a two-column FOF as complex and all others as real. This has been available previously in *MARTHA*, and its result, of course, does not depend on the network. A similar, new function is *OUTVAR*, which is preceded in the output list by the name  $ZG$ ,  $ZL$ ,  $ZN$ ,  $ZNIN$ ,  $ZNOUT$ , or  $EG$ . The appropriate variable is placed in the output list, as a complex quantity. This is useful for viewing complicated loads or sources.

Example:

```
PLOT EG OUTVAR, V2 OF NETWORK
```

produces a comparison of the generator and output voltages. *OUTVAR* is also useful if the generator or load admittance or reflection coefficient is desired. Example:

```
PLOT REC ZL OUTVAR, YOUT OF WCC AMPLIFIER
```

permits a visual inspection of how well the load admittance provides a conjugate match.

#### D. New Modifiers

*MARTHA* contains three new modifiers. Two of them, *NORM* and *GAMMA*, unlike the other modifiers in *MARTHA*, must be preceded in the output list by the name  $ZN$ ,  $ZNIN$ , or  $ZNOUT$ . A complete list of *MARTHA* modifiers appears on page 84.

Normalization. To produce normalized responses, it is only necessary to divide any of the response functions (with the dimensions of impedance) by either  $ZN$  (for one-port networks) or  $ZNIN$  or  $ZNOUT$  (for input or output port of two-port networks). The function *NORM* simply does this--its left argument is either  $ZN$ ,  $ZNIN$ , or  $ZNOUT$ , and its right argument is the output list generated so far. Example:

```
ZN-50
PRINT ZN NORM Z OF DESIGN
```

Of course *MARTHA* does not check whether the correct normalization impe-

dance was chosen by the user, and if you wish to normalize  $Z$  by something different from  $Z_N$  you may. Any response function may be normalized, although *NORM* was designed for the response functions  $Z$ ,  $Z_{11}$ ,  $Z_{22}$ ,  $Z_{IN}$ ,  $Z_{OUT}$ ,  $H_{11}$ ,  $G_{22}$ ,  $Z_{I1}$ ,  $Z_{I2}$ ,  $Z_{K1}$ ,  $Z_{K2}$ ,  $Z_{M1}$ , and  $Z_{M2}$ . If you want to normalize an admittance, you can do so by taking the reciprocal before and after. Example:

```
PRINT REC ZNIN NORM REC Y11 OF FILTER
```

Reflection Coefficient. The reflection coefficient corresponding to an impedance  $Z$ , is

$$\frac{Z - Z_n^*}{Z + Z_n} \quad (7)$$

where a complex, frequency dependent normalization impedance  $Z_n$  is possible. The new modifier *GAMMA* simply does this calculation. Its left argument is either  $Z_N$ ,  $Z_{NIN}$ , or  $Z_{OUT}$ , and its right argument is the output list produced so far, with the left-most response function an impedance. Example:

```
PRINT ZNOUT GAMMA ZM2 OF AMPLIFIER
```

produces the conjugate-match reflection coefficient to use as a load for an amplifier. Note that *MARTHA* does not check to see that the correct normalization impedance was chosen by the user, nor to see if the quantity being used is actually an impedance. You can calculate a reflection coefficient from an admittance if you first take its reciprocal.

Example:

```
PLOT ZNIN GAMMA REC Y11 OF NETWORK
```

VSWR. The standing-wave ratio corresponding to a reflection coefficient  $\Gamma$  is

$$\left| \frac{1 + |\Gamma|}{1 - |\Gamma|} \right|. \quad (8)$$

The modifier *SWR* makes this calculation. Example:

```
PRINT SWR S11, SWR ZNIN GAMMA Z11 OF NETWORK
```

produces the VSWR corresponding to both  $S_{11}$  and  $Z_{11}$ .

#### E. New Miscellaneous Functions

The workspace 100 *MARTHAX* contains extra miscellaneous functions to work with *MARTHA*. Two of these, *MAKEFOF* and *COLUMNSOF*, are used in connection with numerically defined functions of frequency, and have been moved to a new workspace, 100 *MARTHAN*, described in Section IV-B. A complete list of extra miscellaneous functions for *MARTHA* appears on page 87.

There are six new functions in 100 *MARTHAX* for use with waveguides. The existing two functions *RECT1* and *RECT2* calculate, for the dominant mode of rectangular waveguide, the cutoff frequency and infinite-frequency characteristic impedance from the waveguide dimensions. They differ only in the formula used for  $Z_{0\infty}$ . The new function *RECT* is similar to those two, and again differs in the formula for  $Z_{0\infty}$ . To summarize, the formulas used in the three cases are:

$$RECT: Z_{0\infty} = \sqrt{\mu_0/\epsilon} \quad (9a)$$

$$RECT1: Z_{0\infty} = 2(b/a)\sqrt{\mu_0/\epsilon} \quad (9b)$$

$$RECT2: Z_{0\infty} = b\sqrt{\mu_0/\epsilon} \quad (9c)$$

where  $\epsilon$  is the dielectric constant (values different from one can be specified with the *MARTHA* function *FORDIEL*), and  $a$  and  $b$  are the two inside dimensions, in meters ( $a \geq b$ ).

The other five new functions are for use with circular waveguides, for calculating cutoff frequency from the waveguide radius  $r$ . The formulas used for the five lowest-order modes (including the low-loss  $TE_{01}$  mode and two  $TM$  modes) are

$$CIRCTE11: f_c = (1/3.412 r) \sqrt{\mu_0/\epsilon} \quad (10a)$$

$$CIRCTE21: f_c = (1/2.057 r) \sqrt{\mu_0/\epsilon} \quad (10b)$$

$$CIRCTE01: f_c = (1/1.640 r) \sqrt{\mu_0/\epsilon} \quad (10c)$$

$$CIRCTM01: f_c = (1/2.613 r) \sqrt{\mu_0/\epsilon} \quad (10d)$$

$$CIRCTM11: f_c = (1/1.640 r) \sqrt{\mu_0/\epsilon} \quad (10e)$$

In each case the infinite-frequency characteristic impedance is  $\sqrt{\mu_0/\epsilon}$ . Note that *CIRCTM01* and *CIRCTM11* are to be used with the new element *TM*, not with *WG*, which produces *TE* waveguide modes only.

#### IV. NEW LIBRARY WORKSPACES

*MARTHA* has two new workspaces in its library, one (100 *MARTHAF*) devoted to format functions, and the other (100 *MARTHAN*) containing functions to work with numerically defined functions of frequency (FOF's).

##### A. New Format Functions

The new workspace 100 *MARTHAF* in the *MARTHA* library contains new functions for formatting the output of *MARTHA*. This includes a new Smith-chart plotter, as well as means for specifying the horizontal and vertical scales used in plots, and the printing precision of printouts. A complete list of *MARTHA* format functions appears on page 85.

Printing Precision. The precision of *MARTHA* prints (normally five significant figures) may be changed to another value by the format function *PLACES*. This is useful where five places are insufficient, or when fewer are enough and a faster or more compact output is needed. The word *PLACES*, preceded by a number, can appear anywhere in the output list.

Example:

```
PRINT 7 PLACES DB IG OF FILTER
```

Setting Scales. The horizontal and vertical scales of plots are normally chosen automatically by *MARTHA* so that all points fall within the plot and yet there is reasonable detail. The functions *HSCALE* and *VSCALE* can be used to set the limits if desired. Merely precede *HSCALE* (or *VSCALE*) by a vector containing two numbers, for the two extreme values. Place anywhere in the output list. To reverse a graph, the higher number can be put first in the vector. If *HSCALE* is used, the same scale is used for all dependent variables. Points that fall outside the graph are ignored. When used with *PLOG*, the left argument for *VSCALE* should be the powers of 10 between which the plot is to go. Examples:

```
PLOT 0 5000 VSCALE ZIN OF AMPLIFIER
```

```
PLOG 1 4 VSCALE 30 20 HSCALE DB TG, DB IG OF AMPLIFIER
```

Complex-Plane Plots. Normal *MARTHA* plots are vs. frequency, or (if the format function *VS* is used) vs. one of the calculated results. It is sometimes desired, however, to plot one variable against another, and a third against a fourth, etc. This is normally done for plots in the complex plane. The format function *PAIRS*, if placed in the output list, produces a plot where the first independent variable is plotted against the second, the third against the fourth, etc. The same scale is used on all vertical and all horizontal variables. Example:

```
PLOT PAIRS YIN, Y11 OF NETWORK
```

produces a plot of the admittance loci in the complex plane.

Smith Charts. The new function *SMITH* may be used instead of *PRINT*, *PLOT*, or *PLOG*, to produce a Smith chart. (The result is the same size as the Smith charts that are available in pads, so one of those can be laid directly over the *MARTHA* output for reading the impedance values.) Five of the circles of the Smith chart are used as background. More than one reflection coefficient can be plotted at one time. Examples:

```
SMITH S11, S22 OF AMPLIFIER
```

*SMITH SIN, ZNIN GAMMA Z12 OF FILTER*

Pads of Smith charts expanded about the middle are also available, and are useful for very nearly matched networks. The Smith chart plotter will be forced to make a chart to fit this paper if the word *EXPAND* appears in the output list. Example:

*SMITH EXPAND S11, SIN OF FILTER*

Other scales for the Smith chart and other sizes can be specified by using the format functions *HSCALE*, *VSCALE*, *WIDE*, and *HIGH*. These may, of course, be difficult to interpret, especially if the background circles do not happen to fall in the area considered.

Quick Prints. Since the output from a *MARTHA* analysis can be printed, plotted, or stored for later use, the general form of an output request in *MARTHA* is

```
(PRINT)
(PLOT )
(PLOG ) <output list> OF <network>
(SMITH)
(STORE)
```

If the first word is omitted, then a quick print without the heading is obtained, with the frequency in the last column rather than the first. To change the precision with which this "quick print" is printed, use the APL system command *)DIGITS*. For example, for only four-place precision, type *)DIGITS 4*. The workspace 100 *MARTHA* initially is set to *)DIGITS 5* so that quick prints have the normal *MARTHA* five-place precision. This will be retained if you obtain that workspace with the command

```
)LOAD 100 MARTHA
```

but if, instead, you

```
)COPY 100 MARTHA
```

the printing precision will revert to the APL standard of ten places.

The format functions *HSCALE*, *VSCALE*, *PAIRS*, and *EXPAND* are, like the other format functions in *MARTHA*, "one-shot" requests that apply only to the one print or plot. On the other hand, the *)DIGITS* command to set the precision of the quick prints will stay in force until changed by another *)DIGITS* command.

Displaying FOF's. FOF's can be displayed by *MARTHA* with the same format options available. Merely use the name of the FOF in place of the output list and network definition. Thus, to get a quick print without a heading, merely type the name of the FOF. All of the format functions except *VS* can be used.

Another way of displaying FOF's is to incorporate them in the output list by using the response function *OUTFOF*. In this case, they can be displayed alongside the results of a *MARTHA* analysis and all the *MARTHA* modifiers and format functions can be used.

Other Plotters. Some APL installations have general-purpose plotters, and in some cases these can be used with *MARTHA*. A specially written interface function *FROMMARTHA* is used. This function, if it exists on your installation, is in the workspace 100 *MARTHAF*, and instructions are in the variable *USAGE* in the workspace 100 *HOWMARTHA*.

Use of Tabs. Some APL installations can make use of tab settings on your terminal, to speed up *MARTHA* plotting. To see how to take advantage of this feature, if possible, refer to the variable *USAGE* in the workspace 100 *HOWMARTHA*.

## B. New FOF Functions

A new workspace 100 *MARTHAN* contains functions that are useful for working with numerically defined functions of frequency (FOF's). The existing *MARTHA* functions *MAKEFOF* and *COLUMNSOF* have been moved to that workspace from the workspace 100 *MARTHAX*. A complete list of FOF functions in *MARTHA* appears on page 86.

Warning. If you have any FOF's defined on versions of *MARTHA* dated 71 or 72 they must be changed so as to conform to the new way in which FOF's are now stored in *MARTHA*. This is easy to do:

```
NEWFOF ← 1 3 2 ⍉ OLDFOF
```

In case you are not sure whether a given FOF has been transformed, you can use the following test. The shape (obtained by typing the name of the FOF preceded by  $\rho$ ) should (after changing) be a set of three numbers. The first is 1, and the second is the number of defining frequencies, and the third is one plus the number of columns of the FOF.

Of the functions in 100 *MARTHAN*, all except the following rely for proper operation on the background function *FOF*. The following are each self-contained:

```
COLUMNSOF
FROMDBDEG
FROMFOF
FROMMAGDEG
INTERPOLATE
MAKEFOF
TOFOF
```

These will be described first.

Matrices and FOF's. The function *FROMFOF* has as an argument a FOF, and it returns a matrix containing all the values in the various columns. The matrix will have as many columns as there are columns in the FOF. This function is used to extract the numerical results from a *MARTHA* analysis or calculation, for use by other APL functions. In other words, it is a way of getting numerical results from *MARTHA*, for use outside *MARTHA*. It is also useful in the function *NEWELEMENT* for user-defined elements, since the matrix that function returns is the same form as that produced by *FROMFOF*.

The function *TOFOF* performs the reverse operation. It creates a FOF when its right argument is a matrix of the form produced by *FROMFOF*, and the left argument is a vector containing the defining frequencies to be used in the new FOF. There must be as many defining frequencies as there are rows in the matrix.

Complex Conversions. The function *FROMMAGDEG* is a convenient way to calculate FOF's with real and imaginary parts in pairs of columns, from a FOF with the same complex numbers in magnitude, degree form. For example, reflection coefficient data may be most easily obtained in magnitude-angle form. If a six-column FOF named *SM* is created this way, with three complex functions of frequency (magnitudes in columns 1, 3, and 5, and angles in degrees in columns 2, 4, and 6) then *FROMMAGDEG SM* is a six-column FOF with the same complex numbers stored in real-imaginary form.

The function *FROMDBDEG* is exactly the same as *FROMMAGDEG* except that the magnitudes are assumed expressed in dB. In each case the argument must be a FOF with an even number of columns.

Changing Independent Frequencies. The function *INTERPOLATE* provides a convenient way of changing the frequencies at which a FOF is defined. Its right argument is a FOF with any number of columns, and the left ar-

gument is a vector of the desired new defining frequencies. Each column will be interpolated (or extrapolated if necessary) linearly to create the new FOF. This function is useful because often it is more desirable to interpolate, say, the magnitude and phase than the real and imaginary parts, or to interpolate, say, the admittance to be used by *YFOF* rather than the ABCD-matrix values created by *YFOF*.

The functions above are all self-contained. The functions below all require the background function *FOF*.

Other Functions. The three functions *FCAT*, *FCOL*, and *FDF* can deal with *FOF*'s with any number of columns. *FCAT* is dyadic, and returns the FOF containing all columns of each of its arguments. Thus *A FCAT B* is a new FOF with the columns of *A* followed by the columns of *B*. The function *FCOL* is monadic, and returns the number of columns in the FOF. That is, if *B* is a five-column FOF, then *FCOL B* is equal to 5. The function *FDF* returns a FOF with one column, that column containing the defining frequencies of the FOF which is the argument for *FDF*. Thus if *B* is a three-column FOF with defining frequencies 2, 4, and 23 Hz, then *FDF B* is a one-column FOF with the same defining frequencies, and in its one column the numbers 2, 4, and 23.

The remaining seventeen functions, nine monadic and eight dyadic, expect as argument(s) a one-column or two-column FOF. A one-column FOF is interpreted as a real function, and a two-column FOF as a complex function, with real part in the first column, and imaginary part in the second column. In place of a one-column FOF a single number may be used, and in place of a two-column FOF a vector of length 2 may be used. Thus, 0 1 represents *j*.

The monadic functions, *FRE*, *FIM*, *FMAG*, *FRAD*, *FDEG*, *FCC*, *FSIG*, *FEXP*, and *FLN* all return a FOF with the same defining frequencies as the argument, but containing, respectively, the real part, imaginary part, magnitude, angle in radians, angle in degrees, complex conjugate, signum, exponential (base *e*), and natural log. The result is either a one-column (if real) or two-column (if complex) FOF. The angle is greater than -180 degrees and less than or equal to +180 degrees. The signum of a complex number is either 0 if the number is 0, or else has the same angle but unit magnitude. The natural log is, consistent with the angle range given above, calculated with a branch cut along the negative-real axis.

The dyadic functions, *FADD*, *FSUB*, *FMUL*, *FDIV*, *FPWR*, *FLOG*, *FEQ*, and *FNEQ*, all return a real or complex (real if possible) FOF. If the two arguments have the same defining frequencies, they are used in the FOF that is returned. The values are the result of, respectively, real or complex addition, subtraction, multiplication, division, exponentiation (like the APL *A★B*), logarithm evaluation (like the APL *A\*B*), or test for equality and inequality, respectively. The latter two functions apply an APL-like fuzz in the complex plane, and return either 1 if true or 0 if false.

If the two arguments for any of the dyadic functions (including *FCAT*) have different defining frequencies, then either one set or the other is used in the result, and the values of the FOF with the set not used are interpolated or extrapolated linearly to match the defining frequencies used. The defining frequencies of the left-hand argument are used unless (a) the right-hand argument has more defining frequencies, or (b) each has one defining frequency and that of the left argument is equal to zero.

Table II shows the FOF functions and their meaning for both real and complex arguments.

If you copy individual functions to the active workspace, be sure not to forget the background function *FOF*. The most used functions have



been gathered together in a group for easy copying--included in the group *FOF* are the functions *COLUMNSOF*, *FCOL*, *FCAT*, *FADD*, *FSUB*, *FMUL*, *FDIV*, *FPWR*, and the background function *FOF*.

Example. As an example of the use of these functions, consider the calculation of reflection coefficient from precision slotted waveguide (or slotted line) measurements. The formula used is

$$\Gamma = - \frac{V - 1}{V + 1} e^{\frac{j4\pi d}{c} \sqrt{f^2 - f_c^2}} \quad (11)$$

where  $V$  is the measured VSWR and  $d$  is the distance of the minimum in the standing-wave pattern in front of the reference plane. The speed of light is denoted  $c$ , and the cutoff frequency of the waveguide is  $f_c$ . Let us suppose we have a two-column FOF whose first column is the VSWR, and whose second column is the point of minimum. Let us write a function to convert this to reflection coefficient. First we should select the first column of the argument and call it, say,  $V$ , and then calculate the fraction in Equation (11). Then we should extract the defining frequencies of the FOF using the function *FDF*, and compute the argument of the exponential. Finally we can combine them. The overall function

Table II. Functions to perform arithmetic on real or complex FOF's. For the dyadic functions, if one argument is real and the other complex, the real one is treated as complex with an imaginary part of zero. The result is a real or complex FOF (real if possible). Many of the functions in the workspace 100 *MARTHAN* are not listed here; refer to page 86 for a complete list.

#### 'Scalar' FOF Functions

<u>Syntax</u>	<u>For Complex Arguments</u>	<u>For Real Arguments</u>
Z←FRE A	Real Part of A	A
Z←FIM A	Imaginary Part of A	0
Z←FMAG A	Magnitude of A	A
Z←FRAD A	Phase of A in Radians	0A<0
Z←FDEG A	Phase of A in Degrees	180×A<0
Z←FCC A	Conjugate of A	A
Z←FSIG A	Signum of A	×A
Z←FEXP A	e to the power A	★A
Z←FLN A	Natural Log of A	⊙A
Z←A FADD B	Sum	A+B
Z←A FSUB B	Difference	A-B
Z←A FMUL B	Product	A×B
Z←A FDIV B	Ratio	A÷B
Z←A FPWR B	A to the Power B	A★B
Z←A FLOG B	Log of B to Base A	A⊙B
Z←A FEQ B	Equality Test of A, B	A=B
Z←A FNEQ B	Inequality Test	A≠B

#### 'Mixed' FOF Functions

<u>Syntax</u>	<u>Result</u>
Z←A FCAT B	Catenation of All Columns of A and B; Like 'A,B'
Z←FCOL A	Number of Columns of A; Like 'ρA'
Z←FDF A	1-Column FOF Containing the Defining Frequencies of A

might look like this:

```

      ▽B←FC SLOTGUIDE A;V;D
[1]  V←1 COLUMNSOF A
[2]  B←(1 FSUB V) FDIV 1 FADD V
[3]  D←2 COLUMNSOF A
[4]  A← 0 1 FMUL 4 FMUL 3.1416 FMUL(D FDIV 3E8) FMUL(((FDF A) FPWR 2) FSUB FC★2) FPWR 0.5
[5]  B←B FMUL FEXP A
      ▽

```

The cutoff frequency is assumed to be the left argument of the function. For a TEM slotted line, a value of 0 would be used. The name A is used for the argument of the exponential after it is no longer necessary for the original FOF. Note the use of the single numbers 1, 4, 3.1416, 2, and 0.5 as arguments for the FOF functions, and the use of 0 1 to stand for j.

## V. PRACTICAL CONSIDERATIONS

### A. Behind-the-Scenes Improvements

One problem with *MARTHA* in the past has been the relatively small size of the APL active workspace, normally about 32,000 bytes. The message *WS FULL* is printed whenever the computer runs out of space. Execution then stops. This can happen during a *MARTHA* analysis if too many frequencies are used at once. The new version of *MARTHA* alleviates (but does not completely eliminate) this problem. *MARTHA* now checks the space available, and, if necessary, does the analysis more than once, with smaller batches of frequencies. The result is that *MARTHA* can now handle about twice as many frequencies as previously. If you are reasonably tidy in managing your workspace, you should not run into this trouble.

*MARTHA* now uses faster and more compact algorithms during analysis. For example, the common connections (*WP A*) *WC B* and (*WS A*) *WC B* are automatically detected and referred to fast one-step analyzers.

### B. Reducing Computation Cost

As an advanced *MARTHA* user, you may wish to try to minimize the cost of running *MARTHA*. Two strategies are particularly effective. The first is to use less expensive functions, and the second is to avoid unnecessary repeated calculations.

First, significant savings in running cost can be made by using the less expensive wiring functions, where possible. In general, those wiring functions that deal with one-port networks are many times faster than those for two-port networks. This is illustrated in Figure 8. If your network can be expressed in terms of the functions *S*, *P*, and *WT*, your analysis will be much faster. Figure 9 shows some examples. In particular, note that a terminated ladder network can be expressed as a series-parallel network.

Second, to plan your work so as to eliminate repeated calculations, you should know about the types of action necessary in using *MARTHA*. There are typically the following steps:

1. Element definition, e.g. with *MARTHA* functions *R*, *L*, *C*, *TEM*, etc. This is very inexpensive. The parameter values must be known, but frequency *F* and the environment variables *ZG*, *ZL*, *ZN*, *ZNIN*, *ZNOUT*, and *EG* need not be known.
2. Network wiring, e.g. with *MARTHA* functions *S*, *P*, *WC*, *WP*, *WS*, etc. This is more expensive. The element values and topology must be known, but not the frequency or environment variables.
3. Analysis, not including evaluation of response functions. This is the most expensive step. The network parameters, topology, and the frequency must be known, but not the environment variables.
4. Evaluation of Response Functions. This is less expensive. The environment variables must be known.
5. Formatting and output. This is relatively inexpensive.

These five steps can all be requested by a single *MARTHA* command, for example

```
PRINT ZIN OF (WS C .2E-6) WC L 1E-9
```

where element definition is done by the functions *L* and *C*; wiring by the functions *WS* and *WC*; analysis by the function *OF*; response function evaluation by the function *ZIN*; and output by the function *PRINT*. Alternatively, this procedure can be broken up into steps, and the intermediate results saved for use in the next step. The same example could proceed as follows:

```

C1←C 0.2E-6           (element definition)
L1←L 1E-9             (element definition)
NET←(WS C1) WC L1     (network wiring)
NETS←NDE NET          (network analysis)
RESP←ZIN OF NET      (evaluation of response)
PRINT RESP            (output)

```

The *MARTHA* function *NDE* produces a numerically defined element by analyzing the network which is its argument, at the frequencies in the vector *F*.

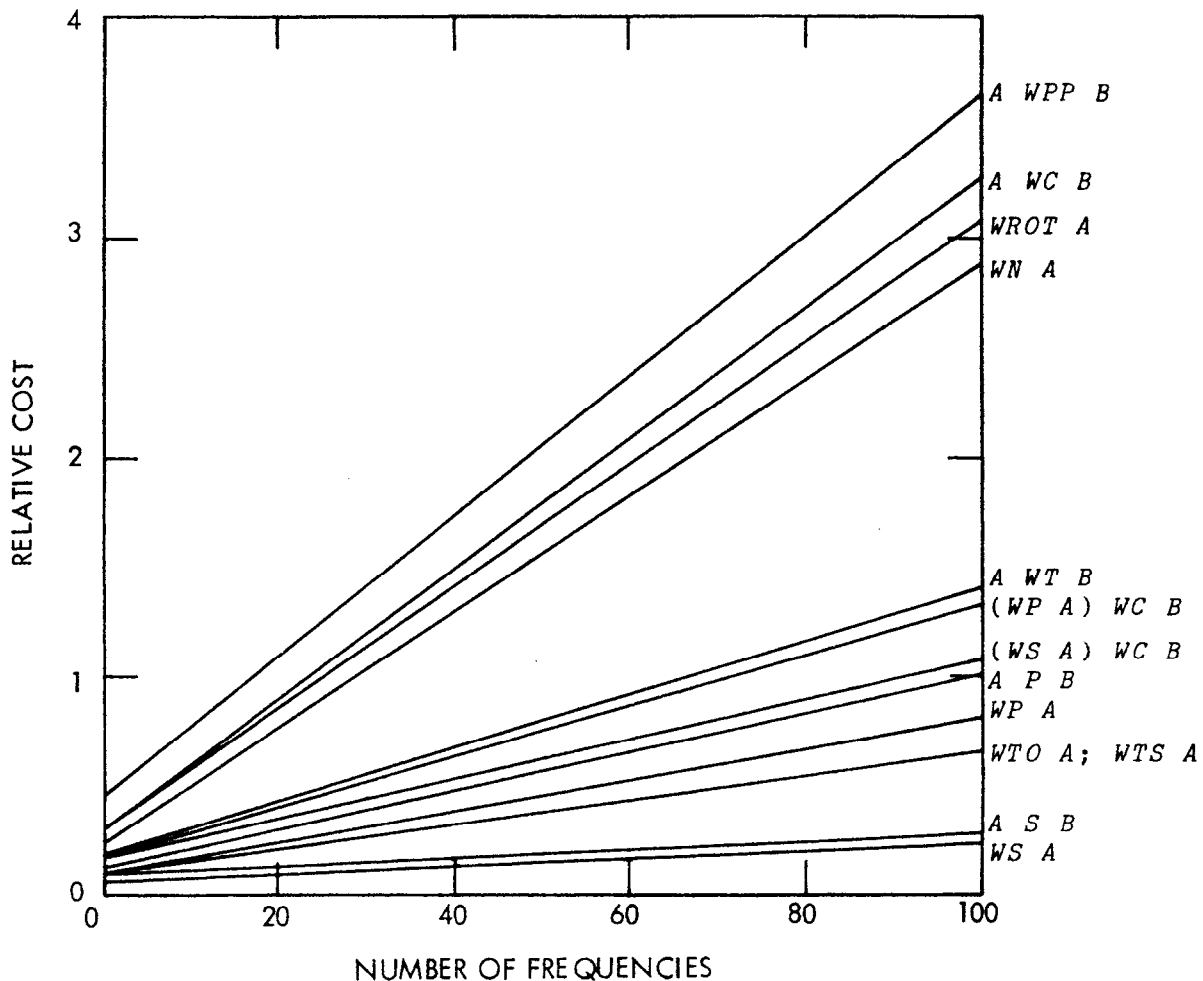


Figure 8. Approximate relative cost of some of the common wiring functions during analysis. The functions *WSP*, *WPS*, and *WSS* cost the same as *WPP*. The cost in each case consists of a fixed overhead plus an amount proportional to the number of frequencies. These results depend somewhat on the APL system used, and on some details of implementation.

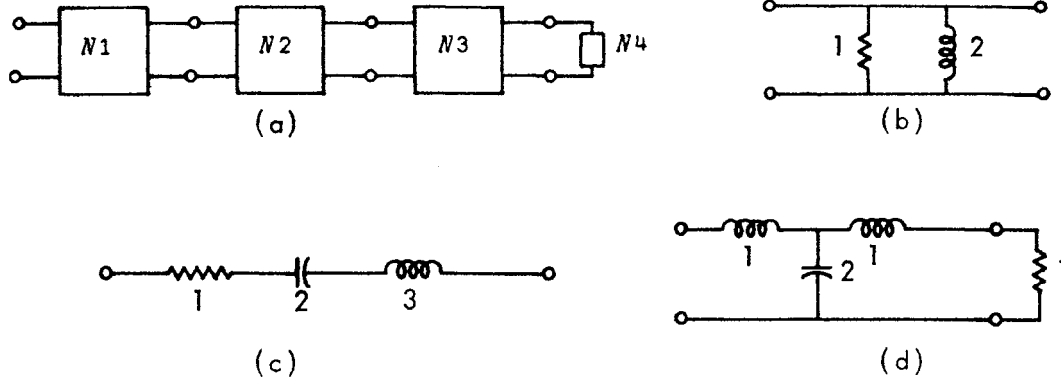


Figure 9. Networks with alternate MARTHA definitions, some less expensive to use during analysis than others. Expressions:

```

(a, more expensive) NET=(N1 WC N2 WC N3) WT N4
(a, less expensive) NET=N1 WT N2 WT N3 WT N4
(b, more expensive) (WP R 1) WC WP L 2
(b, less expensive) WP (R 1) P L 2
(c, more expensive) WTO (WP R 1) WSS (WP C 2) WSS WP L 3
(c, less expensive) WTS (WS R 1) WC (WS C 2) WC WS L 3
(c, least expensive) (R 1) S (C 2) S L 3
(d, more expensive) ZL=1
PRINT ZIN OF (WS L 1) WC (C 2) WC WS L 1
(d, less expensive) PRINT Z OF (L 1) S (C 2) P (L 1) S R 1

```

```

▽ N←CHEB
[1] N←(C 7E-6) WC(WS L 0.015) WC(C 9E-6) WC(WS L 0.015) WC C 7E-6
[2] ▽
ZG←100
ZL←100
F←80×\11
PRINT ZIN OF CHEB
F←40×\23
PRINT DB IG OF CHEB
F←20×\10
PRINT Z OF CHEB WT R 70

```

Example (a), more expensive

```

CHEB←(C 7E-6) WC(WS L 0.015) WC(C 9E-6) WC(WS L 0.015) WC C 7E-6
ZG←100
ZL←100
F←80×\11
PRINT ZIN OF CHEB
F←40×\23
PRINT DB IG OF CHEB
F←20×\10
PRINT Z OF CHEB WT R 70

```

Example (b), less expensive, by about 25% in this case

Figure 10. Cost of running MARTHA can be reduced by defining a network just once, even if  $F$  and the environment variables change.

To avoid repeated network definition, try to define a network by executing the wiring functions directly, rather than placing the network definition in an APL function to be repeatedly executed. Figure 10 illustrates this technique, which is useful if repeated analyses at different frequencies are contemplated.

In case the network has some parameter which you think you might want to change later, you can still cut your costs by defining the network as an APL function, but then executing it only once for each value of the parameter. Figure 11 illustrates this.

Repeated analysis can be avoided, if the frequency vector  $F$  is not changed between analyses, or if the new frequency vector is a subset of the previous one, by two techniques. First, for several output requests in a row, specify the network in the first request, and thereafter refer to the network as *SAME*. This variable, which is created during execution of the function *OF*, contains all the numerical values calculated by network analysis. It is, in fact, a numerically defined element, which can be used directly, or can be part of a more complicated network. Figure 11 illustrates this technique.

Another way of avoiding repeated analysis is to use the function *NDE* to create a numerically defined element, just like *SAME*. This is particularly useful if used to avoid repeated analysis of a subnetwork which is used two or more times in a larger network. Figure 12 illustrates this in the case of a symmetric filter. This is also useful if, for example, one transistor model is used several times in a circuit.

```
F←10★1+0,0.2×120
PLOG 40 HIGH VG, MAG VG OF AMP 100
PLOG 20 HIGH DB VG OF AMP 100
PRINT DB VG OF AMP 100
PLOG 20 HIGH (DB VG OF AMP 400), DB VG OF AMP 100
```

Example (a), more expensive.

```
F←10★1+0,0,2×120
AMP1←AMP 100
PLOG 40 HIGH VG, MAG VG OF AMP1
PLOG 20 HIGH DB VG OF AMP1
PRINT DB VG OF AMP1
PLOG 20 HIGH (DB VG OF AMP 400), DB VG OF AMP1
```

Example (b), less expensive, by about 20% in this case.

```
F←10★1+0,0.2×120
PLOG 40 HIGH VG, MAG VG OF AMP 100
PLOG 20 HIGH DB VG OF SAME
PRINT DB VG OF SAME
PLOG 20 HIGH (DB VG OF AMP 400), DB VG OF SAME
```

Example (c), least expensive, by about 52% over example (a).

Figure 11. Cost of running *MARTHA* can be reduced by eliminating redundant network definitions. It can be reduced still further by use of *SAME* to denote the most recently analyzed network, thus avoiding repeated analysis. This technique is useful if new response functions are calculated, or if the environment variables are changed, but  $F$  does not change. In this example, the function *AMP* is the same one used in Example 6 of *MARTHA* User's Manual[2, page 40].

### C. Common Errors

*MARTHA* has been in wide use by engineers who are not always experts at computer usage. The following tips are intended to help you avoid a few common errors. A complete list of *MARTHA* error messages appears on pages 87 to 89.

Always clear your state indicator. When *MARTHA* gets into trouble, or when the APL system gets into trouble, execution is sometimes suspended "inside" one of the *MARTHA* functions or inside one of your defined functions. A good general rule, at least for new users, is to always clear the state indicator. This is done by typing a single right arrow  $\rightarrow$ . You can check to see if this is necessary by typing *)SI* and if anything gets printed, your state indicator is not clear. Failure to clear the state indicator can use up valuable space, and it can also lead to confusion about the meaning of some variables, including *F*.

Do not erase any of the *MARTHA* functions, even those which you do not think you will need. This is because some *MARTHA* functions rely on others which may at first appear to be unrelated. For example, the function *HYBRIDPI* uses some results calculated by the functions *R*, *C*, and *FET* (to see why, look at the equivalent circuits). If you suspect some *MARTHA* functions are missing, type

```
)COPY 100 MARTHA MARTHA
```

and they will be replaced, if missing, but your variables *F*, *EG*, *ZG*, *ZL*, *ZN*, *ZNIN*, and *ZNOUT* will be unchanged.

```
C1←C COAXDISCAP 0.5×0.01× 1.425 0.619 1.275
T1←TEM COAX 0.7125 0.6375 0.0003698
T2←TEM 50 0.020209
T3←TEM COAX 0.7125 0.6375 0.0021501 FORDIEL 2.03
T4←TEM 50 0.018843
F←1E9×7+0,0.1×130
FILTER←C1 WC T1 WC C1 WC T2 WC C1 WC T3 WC C1
FILTER←FILTER WC T4 WC WN FILTER
ZG←50
ZL←50
PRINT DB IG OF FILTER
```

Example (a), more expensive.

```
C1←C COAXDISCAP 0.5×0.01× 1.425 0.619 1.275
T1←TEM COAX 0.7125 0.6375 0.0003698
T2←TEM 50 0.020209
T3←TEM COAX 0.7125 0.6375 0.0021501 FORDIEL 2.03
T4←TEM 50 0.018843
F←1E9×7+0,0.1×130
FILTER←NDE C1 WC T1 WC C1 WC T2 WC C1 WC T3 WC C1
FILTER←FILTER WC T4 WC WN FILTER
ZG←50
ZL←50
PRINT DB IG OF FILTER
```

Example (b), less expensive, by about 17% in this case.

Figure 12. Cost of running *MARTHA* can be reduced by use of the function *NDE* to force early analysis of a subnetwork which appears more than once in a larger network.

Know which response function you want. *MARTHA* has over 100 response functions, including many that are similar-sounding. In case of doubt, check the definition, for example, by referring to the documentation on pages 82 to 84. For example, *G21*, *VG*, and *VR* are each a type of "voltage gain", but they are all different.

Don't forget to set *ZG*, *ZL*, etc. The insertion gain looks much different with the wrong load.

Remember the "precedence convention" for wiring functions. Thus *N1 S N2 P N3* means *(N1 S (N2 P N3))* and not *((N1 S N2) P N3)*. The rule is that each wiring function takes as its right-hand argument the entire network definition to its right. The left argument (if any) is the one object immediately to its left.

If you get an error message, it may be either a *MARTHA* error message, or an APL message. Look first in the list of *MARTHA* error messages on pages 87 to 89. The error message will often display a line that you yourself wrote, and point to the place where the error was found. The actual mistake may have occurred somewhere before and was not recognized right away.

You cannot name any variable the same as one of the *MARTHA* functions. If you try, the message *SYNTAX ERROR* will be printed. To see if this is the problem, type *)FNS* to get a list of function names.

If you encounter a bona fide bug in *MARTHA*, please report it. You might use the reader comment form at the end of this manual, or the form produced by the variable *COMMENTS* in the workspace 100 *HOWMARTHA*.



## VI. EXAMPLES

Eight fully worked out examples illustrate many of the new features. Two examples deal with circuits in the *MARTHA* User's Manual[2], analyzed in a simpler or more complete way.

### A. Example 11, Lossy Inductor and Dipole Antenna

Features illustrated: Delayed analysis; user defined elements often not necessary; *ZPDE*. As an illustration of the way that delayed analysis makes it easier to construct and use models, we treat here two examples from the *MARTHA* User's Manual[2, pp. 64-68] that were used there to illustrate how to set up user-defined elements. User-defined elements are not now necessary for these examples.

The first case is that of a lossy inductor, which is to be specified by its inductance  $L$  and quality factor  $Q$  at a reference frequency  $f_0$ . The model is a lossless inductor in series with a resistor with resistance

$$R = \frac{2\pi f_0 L}{Q} \quad (12)$$

The function *LOSSYL* creates such a network, calculating in the process the series resistance. This function may be run, to generate specific networks, before the frequency is specified. The definition of the network *LOSSYL1* does not depend upon  $F$ , as the example shows. In versions of *MARTHA* dated before 73, this procedure does not work because the wiring functions (in this case *S*) perform part of the analysis.

The second case studied is the radiation impedance of a short dipole antenna. This impedance is

$$Z = \frac{4\pi}{3} \sqrt{\frac{\mu_0}{\epsilon_0}} \frac{\ln\left(\frac{h}{a}\right) - 1}{2 \ln\left(\frac{2h}{a}\right) - 3} \left(\frac{hf}{c}\right)^2 - j \frac{\ln\left(\frac{h}{a}\right) - 1}{2\pi^2 f \epsilon_0 h} \quad (13)$$

where  $h$  is the half-length of the antenna,  $a$  is the radius, and  $c$  is the speed of light. The real part of  $Z$  is proportional to frequency  $f$  squared, and the imaginary part is represented by a capacitor. The function *SHORTDIPOLE* creates this model by first defining an element with impedance that goes as the square of  $s$  ( $2\pi jf$ ) and then wiring it together with the capacitor. The function *ZPDE* from the workspace 100 *MARTHAE* creates one-port or two-port elements with impedance proportional to  $s^n$  where  $n$  is an integer between -5 and 5. This element (and the related elements *YPDE*, *HPDE*, and *ABCDPDE*) are useful in many types of models. In the example shown, the radiation impedance is printed as a function of frequency.

In each of these cases, the definition of the elements required the ability to write APL expressions that correspond to the algebraic expressions above, but not much more knowledge of APL.

```

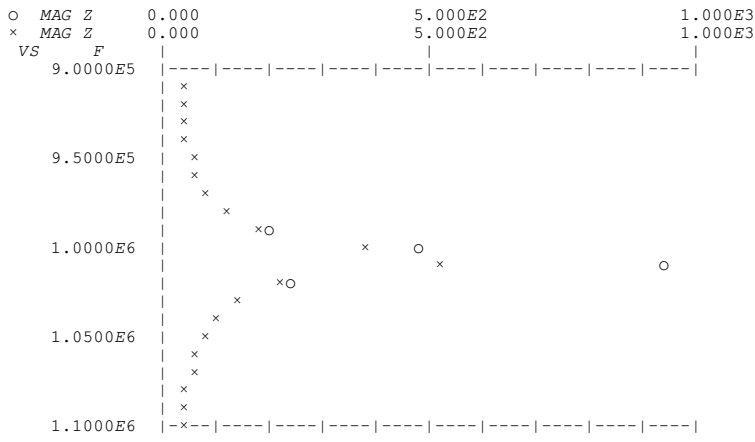
A BEGINNING OF EXAMPLE 11
)LOAD 100 MARTHA
SAVED 11.31.21 10/31/73

VB=LOSSYL A
[1] B=(L A[1]) S R 6.28*A[3]*A[1]+A[2]
[2] V

C1=C 0.025E-6
L1=L 1E-6
LOSSYL1=LOSSYL 1E-6 100 1E6
F=1E6*0.9+0.01*120

PLOT 20 HIGH SS (MAG Z OF C1 P L1), MAG Z OF C1 P LOSSYL1

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/13/73 16:17
MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY
    
```



```

VB=SHORTDIPOLE A;H
[1] H=A[1]
[2] A=A[2]
[3] B=ZPDE 2, -(+03)*377*((-1+@H÷A)÷-3+2*02×H÷A)×(H÷3E8)★2
[4] B=B S C 08.854E-12×H÷-1+@H÷A
[5] V

F=1E6* 0.5 1 2 4 8 16 32

)COPY 100 MARTHAE ZPDE
SAVED 11.32.12 10/31/73

PRINT Z, MAG Z, DEG Z OF SHORTDIPOLE 0.32 0.0049
PRINT Z, MAG Z, DEG Z OF SHORTDIPOLE 0.32 0.0049

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/13/73 16:21
    
```

F	RE Z	IM Z	MAG Z	DEG Z
5.0000E5	2.1173E-4	-1.1369E5	1.1369E5	-9.0000E1
1.0000E6	8.4692E-4	-5.6844E4	5.6844E4	-9.0000E1
2.0000E6	3.3877E-3	-2.8422E4	2.8422E4	-9.0000E1
4.0000E6	1.3551E-2	-1.4211E4	1.4211E4	-9.0000E1
8.0000E6	5.4203E-2	-7.1055E3	7.1055E3	-9.0000E1
1.6000E7	2.1681E-1	-3.5527E3	3.5527E3	-8.9997E1
3.2000E7	8.6725E-1	-1.7764E3	1.7764E3	-8.9972E1

A END OF EXAMPLE 11.

```

A BEGINNING OF EXAMPLE 12.

)LOAD 100 MARTHA
SAVED 11.31.21 10/31/73

X1=(C 8.60776E-12) P (C 0.0116013E-12) S (L 0.0341232) S R 0.1752
X2=(C 17.1140E-12) P (C 0.0177387E-12) S (L 0.0223008) S R 0.1121
X3=(C 18.7184E-12) P (C 0.0138144E-12) S (L 0.0286568) S R 0.1440
X4=(C 20.0690E-12) P (C 0.0147852E-12) S (L 0.0267554) S R 0.1345
    
```

### B. Example 12, Crystal Filter

Features illustrated: Format functions *HSCALE*; *SMITH*; *EXPAND*; *VSCALE*; *PLACES*; quick print. This crystal filter, Figure 13, is adapted, by frequency and impedance scaling, from a filter shown to the author by Mr. William B. Lurie, who ascribed the design to Professor G. Szentirmai [32, Figure 4-18]. There are two stages, each with a half-lattice construction which is modeled with an ideal transformer. The four crystals all have the same equivalent circuit, with different element values. They are defined, first, as X1 through X4; then the two stages are separately defined, and finally the crystal filter itself. The generator and a load impedance are each set to 500 ohms, and then a frequency band near 8 MHz is defined. The filter has a pass band about 4000 Hz wide, between 8 and 8.004 MHz. A plot of the insertion gain, expressed in dB, and angle of the voltage gain shows this passband. Next, we wish to observe the details of the passband by expanding the horizontal scale. The function *HSCALE* is copied from the library workspace 100 *MARTHAF*. This is used to obtain a plot between -2 and 0 dB, with considerable magnification of the passband. As in all cases where the functions *HSCALE* or *VSCALE* are used, points falling outside the specified scales are ignored. Next, a Smith chart is made of the input reflection coefficient of the filter. Although the stop-band points appear nicely on the outer rim of the Smith chart, the points corresponding to the passband all fall toward the center of the Smith chart, and so it is virtually unreadable. A new frequency band is chosen and another Smith chart is attempted. Again this shows insufficient detail. What is needed is an expanded Smith chart, so the function *EXPAND* from the workspace 100 *MARTHAF* is copied, and, after a new frequency vector is defined, an expanded Smith chart is produced. Both Smith charts, incidently, are the same size as standard Smith charts used by microwave engineers. Unfortunately the expanded Smith chart appears to be too much expanded, so a nonstandard, in-

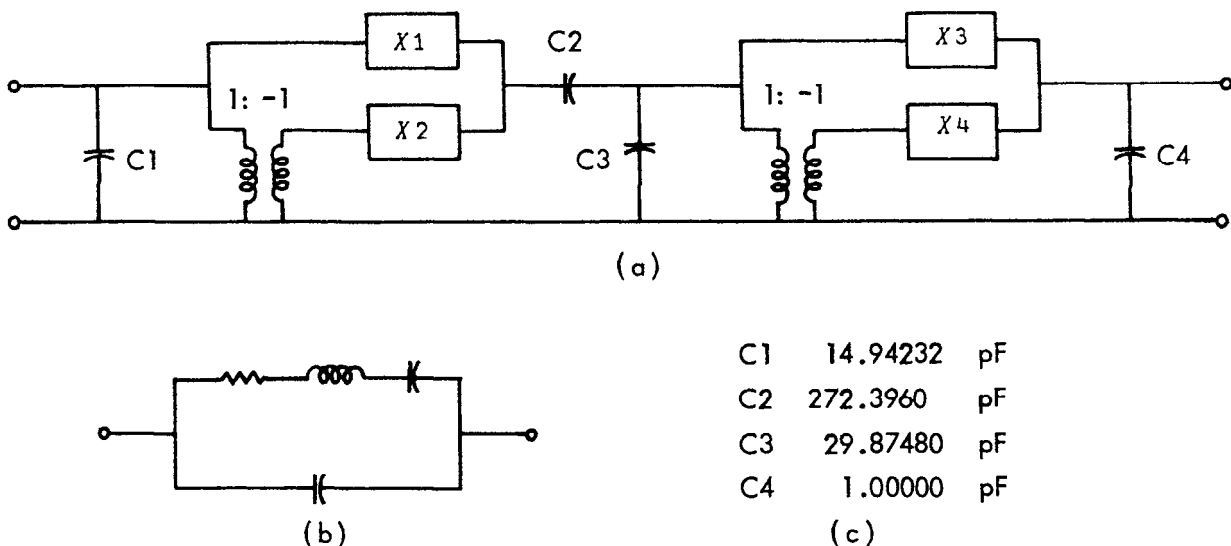


Figure 13. (a) crystal filter. (b) equivalent circuit for the crystals X1, X2, X3, and X4. (c) component values. The element values in the crystal models are not listed since they appear in the example. The resistances in the crystal model are calculated from assumed values of  $Q$ .

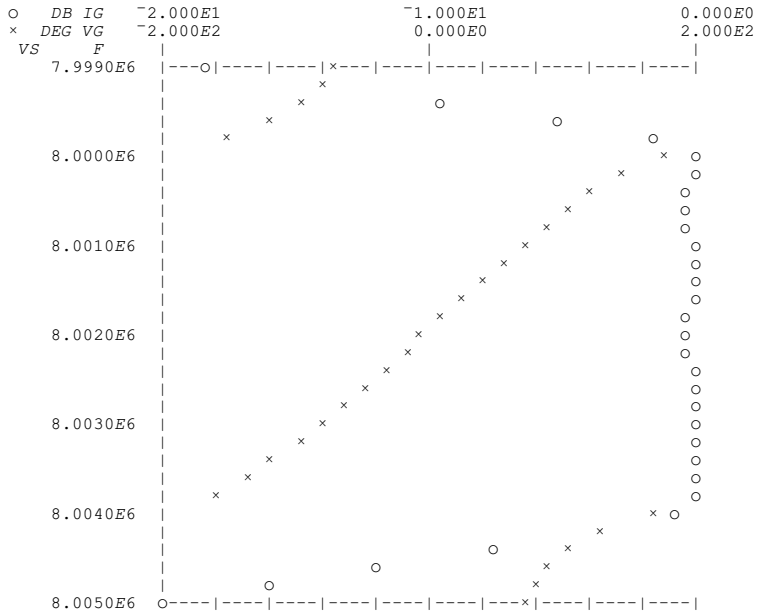
VI-B. Example 12, Crystal Filter

STAGE1-(C 14.94232E-12) WC (WS X1) WPP (IT -1) WC WS X2  
 STAGE2-(C 29.87480E-12) WC (WS X3) WPP (IT -1) WC WS X4  
 CRYSTALFILTER-STAGE1 WC (WS C 272.396E-12) WC STAGE2 WC C 1E-12

ZG-500  
 ZL-500  
 F-7.999E6+0,200x130

PLOT 30 HIGH DB IG, DEG VG OF CRYSTALFILTER

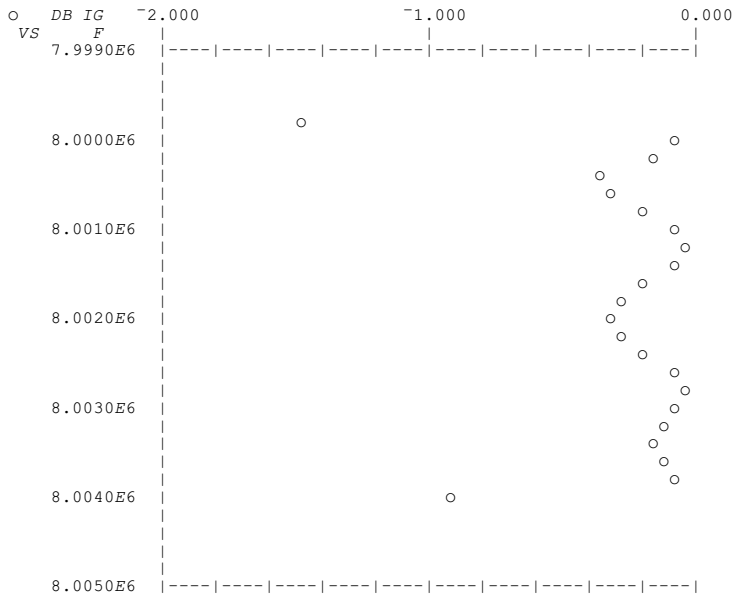
CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 15:58  
 MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY



)COPY 100 MARTHAF HSCALE  
 SAVED 11.32.43 10/31/73

PLOT 30 HIGH -2 0 HSCALE DB IG OF SAME

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:2



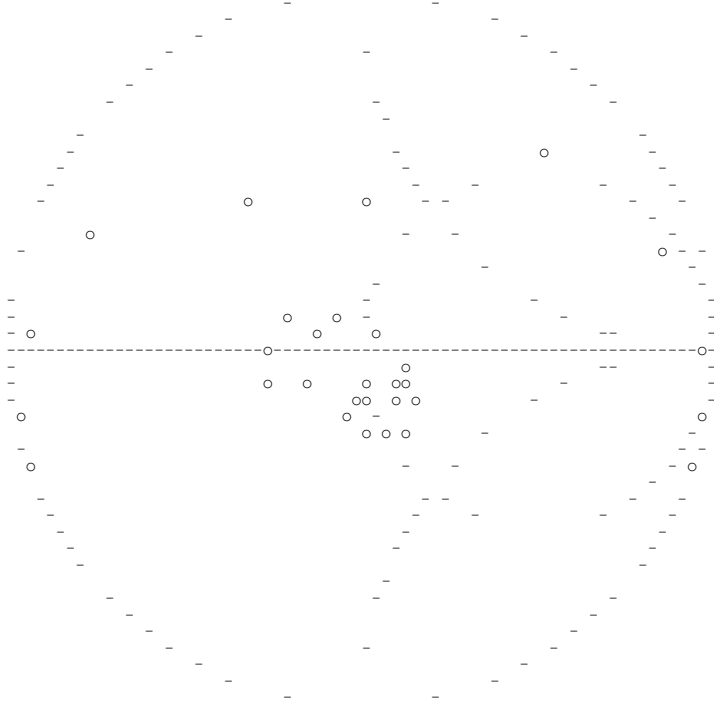
)COPY 100 MARTHAF SMITH  
 SAVED 11.32.43 10/31/73

ZNIN-ZNOUT=500

SMITH S11 OF SAME

CIRCUIT ANALYSIS BY MARTHA. 73\*D 12/14/73 16:5

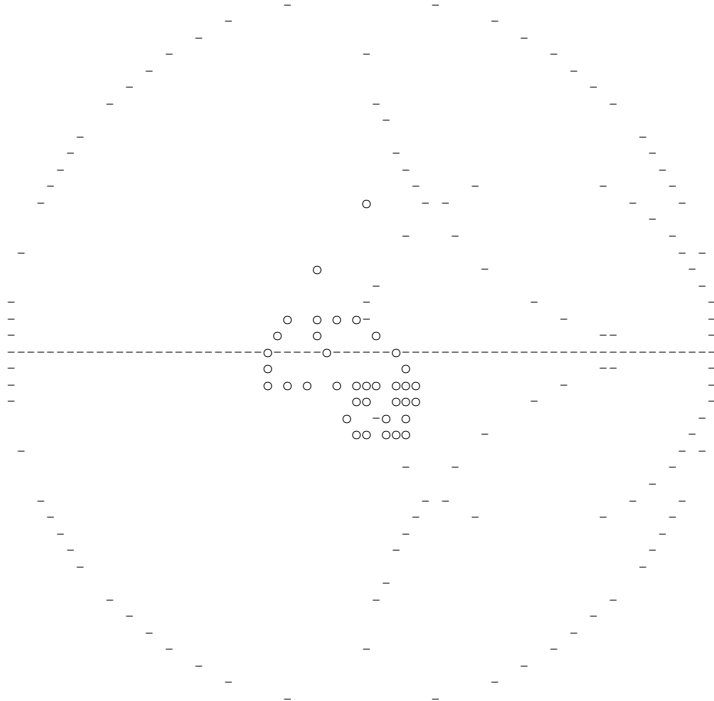
o RE S11 VS IM S11



F-8E6+0,0.001E6\*0.1\*140  
SMITH S11 OF CRYSTALFILTER

CIRCUIT ANALYSIS BY MARTHA. 73\*D 12/14/73 16:10

o RE S11 VS IM S11



intermediate size Smith chart, with specified scales between -0.28 and 0.28, is produced. The number 0.28 was chosen by measuring, on the second Smith chart produced, what radius would include all the points of interest.

Next, a printout of insertion gain, phase of voltage gain, and magnitude of input impedance is given for half of the passband of the filter, along with part of the stopband. An interesting effect is observed at approximately 8.007 MHz, where the phase changes abruptly by 180°. Let us quickly define a new frequency range, and again ask for the insertion gain and phase of the filter. This time, the word *PRINT* is omitted, and so a quick print is obtained, without a heading, and with the frequency in the right-hand column instead of the left. Again an abrupt change of 180° in phase is observed, indicating that the filter has a zero of transmission very close to the imaginary axis at this point. Since it may be important to know the frequency at which this occurs, and since the normal printing precision of five places is inadequate (all the frequency points are the same to five places) a more accurate printout to seven significant figures is made after the frequency is again expanded. Now a considerable amount of detail is visible.

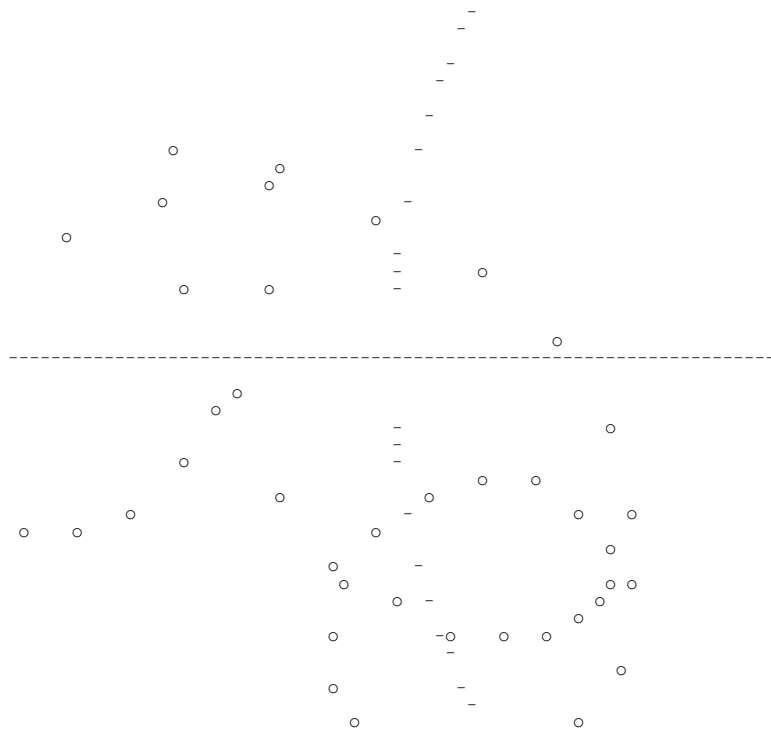
F=8E6+0, (100x139), 50 150 250 ,3300+ 50 150 250 350 450 550

)COPY 100 MARTHAF EXPAND  
 SAVED 11.32.43 10/31/73

SMITH EXPAND S11 OF CRYSTALFILTER

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:16

o RE S11 VS IM S11

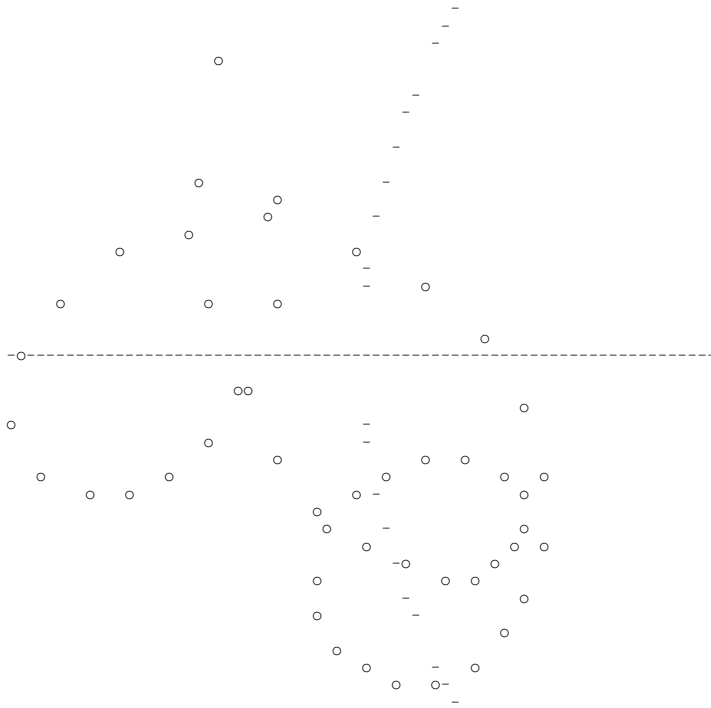


)COPY 100 MARTHAF VSCALE  
 SAVED 11.32.43 12/31/73

SMITH -0.28 0.28 HSCALE 0.28 -0.28 VSCALE S11 OF CRYSTALFILTER

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:20

O RE S11 VS IM S11



F=8.002E6+200x130  
PRINT DB IG, DEG VG, MAG ZIN OF CRYSTALFILTER

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:23

F	DB IG	DEG VG	MAG ZIN
8.0022E6	-2.7428E-1	-1.8240E1	5.0138E2
8.0024E6	-1.8555E-1	-3.2660E1	4.7038E2
8.0026E6	-8.9823E-2	-4.7854E1	4.7521E2
8.0028E6	-3.9720E-2	-6.3898E1	5.2096E2
8.0030E6	-6.3555E-2	-8.0654E1	5.9390E2
8.0032E6	-1.3452E-1	-9.7992E1	6.4698E2
8.0034E6	-1.7197E-1	-1.1624E2	6.1842E2
8.0036E6	-1.0630E-1	-1.3673E2	5.0307E2
8.0038E6	-8.4973E-2	-1.6221E2	3.8952E2
8.0040E6	-9.1322E-2	1.6514E2	5.0864E2
8.0042E6	-3.6485E0	1.3134E2	1.0891E3
8.0044E6	-7.7195E0	1.0583E2	2.8661E3
8.0046E6	-1.1993E1	8.8834E1	2.7605E4
8.0048E6	-1.6059E1	7.7184E1	4.9553E3
8.0050E6	-1.9882E1	6.8697E1	2.7793E3
8.0052E6	-2.3519E1	6.2179E1	2.0690E3
8.0054E6	-2.7039E1	5.6972E1	1.7124E3
8.0056E6	-3.0519E1	5.2684E1	1.4960E3
8.0058E6	-3.4040E1	4.9075E1	1.3495E3
8.0060E6	-3.7707E1	4.5983E1	1.2432E3
8.0062E6	-4.1680E1	4.3297E1	1.1621E3
8.0064E6	-4.6255E1	4.0941E1	1.0981E3
8.0066E6	-5.2139E1	3.8866E1	1.0461E3
8.0068E6	-6.2280E1	3.7103E1	1.0029E3
8.0070E6	-6.7171E1	-1.4507E2	9.6644E2
8.0072E6	-5.7815E1	-1.4640E2	9.3517E2
8.0074E6	-5.4557E1	-1.4777E2	9.0803E2
8.0076E6	-5.2904E1	-1.4905E2	8.8424E2
8.0078E6	-5.1997E1	-1.5022E2	8.6319E2
8.0080E6	-5.1516E1	-1.5131E2	8.4443E2

```

F-8.00685E6+10x110
DB IG, DEG VG OF CRYSTALFILTER
-6.8823E1 3.6769E1 8.0069E6
-7.0507E1 3.6761E1 8.0069E6
-7.2558E1 3.6793E1 8.0069E6
-7.5194E1 3.6907E1 8.0069E6
-7.8915E1 3.7229E1 8.0069E6
-8.5389E1 3.8437E1 8.0069E6
-1.0546E2 -1.6972E2 8.0069E6
-8.4100E1 -1.4631E2 8.0069E6
-7.8498E1 -1.4539E2 8.0069E6
-7.5168E1 -1.4511E2 8.0069E6

)COPY 100 MARTHA PLACES
SAVED 11.32.43 10/31/73

F-8.00691E6+110
PRINT 7 PLACES DB IG, DEG VG OF CRYSTALFILTER
CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:27
    
```

F	DB IG	DEG VG
8.006911E6	-8.639767E1	3.873101E1
8.006912E6	-8.753608E1	3.910974E1
8.006913E6	-8.884346E1	3.961350E1
8.006914E6	-9.037903E1	4.031436E1
8.006915E6	-9.223957E1	4.135279E1
8.006916E6	-9.459954E1	4.304337E1
8.006917E6	-9.782233E1	4.626150E1
8.006918E6	-1.028618E2	5.461697E1
8.006919E6	-1.119975E2	1.025601E2
8.006920E6	-1.054552E2	-1.697215E2

END OF EXAMPLE 12.

BEGINNING OF EXAMPLE 13.

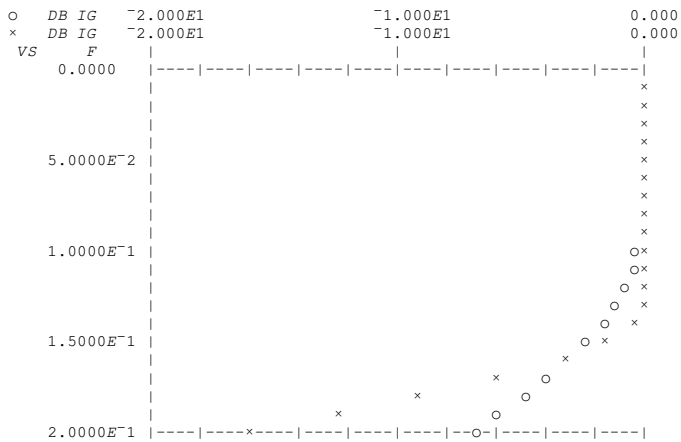
```

)LOAD 100 MARTHA
SAVED 11.31.21 10/31/73

VB=PROTO N;A;K
[1] A=2x100(0.5+1N)/N
[2] B=WP C A[K-N]
[3] LOOP:-(0>K-K-1)/0
[4] B=(WS L A[K]) WC B
[5] -(0>K-K-1)/0
[6] B=(WP C A[K]) WC B
[7] -LOOP
[8] V

F-0.01x120
ZG-1
ZL-1
PLOT SS 20 HIGH (DB IG OF PROTO 3), DB IG OF PROTO 8
    
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:31  
 MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY





### C. Example 13, Ladder Filter Design

Features illustrated: *WHATIS*; *PAIRS*; *ZSCALE*; *FSCALE*; *FINVERT*; *FBP*. *MARTHA* is, as a notation, useful not only in talking to a computer, but in learning from a computer. This example illustrates the use of *MARTHA* notation in a simple Butterworth filter synthesis program.

Prototype Filter. As a first step, the prototype filter *PROTO* is defined as a function of the order  $N$ . In the first line of *PROTO*, a vector of  $N$  element values is created, using the formula from Weinberg [15, Section 13-5). Next, the right-hand end element, in this case assumed to be a capacitor, is defined. The remainder of the function consists of a loop which is traversed as many times as is necessary to use all the elements in the vector  $A$ . The result is a prototype lowpass Butterworth filter of order  $N$ , normalized to 1 ohm and 1 radian per second cutoff frequency. The first plot shows, on the same scale, the insertion gain of third order and eighth order filters. Of course, the eighth-order filter has a sharper cutoff. In case we wish to find out what the element values are, we use the function *WHATIS* from the workspace 100 *MARTHAX*. This is applied to the network created by the function *PROTO*, and we see a display of the third-order Butterworth filter, from which Figure 14 is drawn. This, of course, is precisely the same prototype filter given in many handbooks of filter design, for example [14, page 312].

It is interesting to view, in the complex plane, the locus of impedance seen at the various reference planes in the third order prototype filter. The next plot shows the complex plane, with the circles indicating the input impedance of the filter itself, the crosses the impedance part way into the filter, and the equal signs representing the impedance seen at another reference plane. These three impedances are illustrated in Figure 14. The function *PAIRS*, from the workspace 100 *MARTHAF*, produces the three plots simultaneously. They all start out at low frequencies from the point  $Z = 1$ , and the first and third, if continued for higher frequencies, would approach 0 ohms, whereas the second would approach infinity.

Complete Filter. Next, a more ambitious function entitled *BUTTERWORTH* is written. This is useful because ordinarily the designer does not know the order of filter required, but only the desired performance. This performance might be specified as illustrated in Figure 15, as the maximum passband attenuation  $A_p$ , minimum stopband attenuation  $A_s$ , frequency at edge of passband  $f_p$ , and frequency at edge of stopband  $f_s$ . The function *BUTTERWORTH* expects as an argument a vector of length 5, consisting of  $A_p$  and  $A_s$  (expressed in dB),  $f_p$ ,  $f_s$ , and

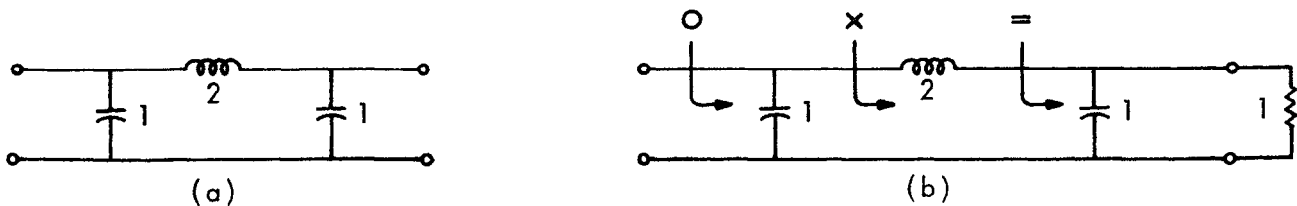


Figure 14. (a) Third-order Butterworth prototype filter *PROTO* 3. (b) same filter with reference planes at which impedances are calculated.

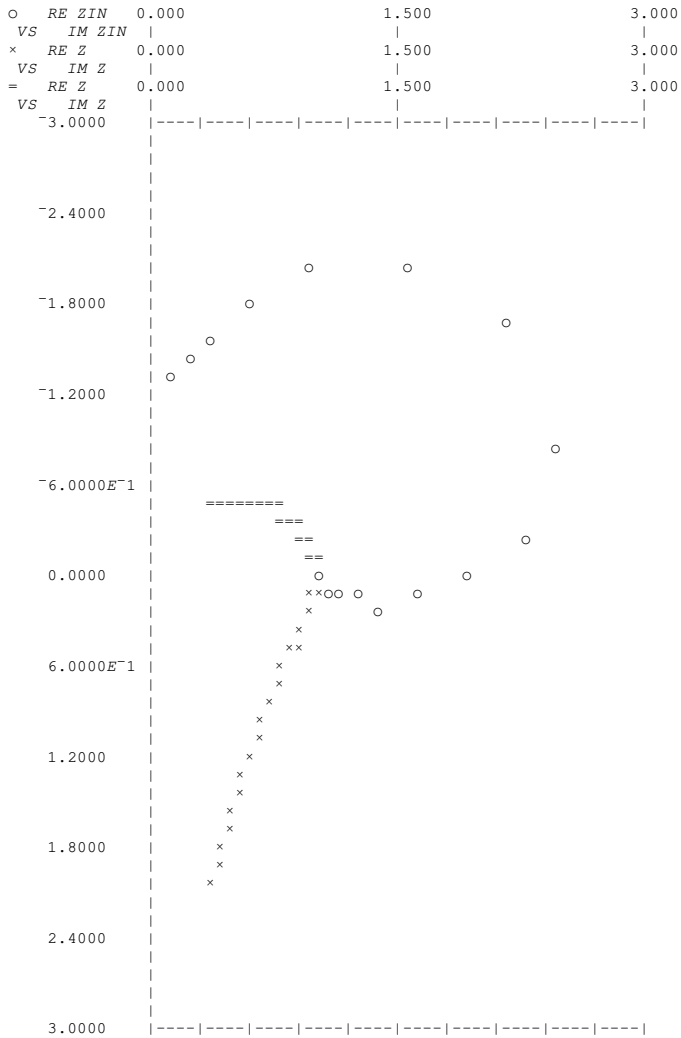
)COPY 100 MARTHAX WHATIS  
 SAVED 11.33.39 10/31/73

WHATIS PROTO 3  
 (WP C 1) WC (WS L 2) WC WP C 1

)COPY 100 MARTHAF PAIRS  
 SAVED 11.32.43 10/31/73

PLOT PAIRS (ZIN OF PROTO 3), (Z OF (L 2) S (C 1) P R 1), Z OF (C 1) P R 1

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:36



VB-BUTTERWORTH X;N;AP;AS;FP;FS;ZT  
 [1] AP=X[1]  
 [2] AS=X[2]  
 [3] FP=X[3]  
 [4] FS=X[4]  
 [5] ZT=X[5]  
 [6] N=Γ((10\*<sup>-1</sup>+10\*AS+10)-(10\*<sup>-1</sup>+10\*AP+10))/2\*10\*FS+FP  
 [7] B=PROTO N  
 [8] B=ZT ZSCALE B  
 [9] B=((O2\*FP)/(1+10\*AP+10)\*1+2\*N) FSCALE B  
 [10] ▽

)COPY 100 MARTHAW FSCALE  
 SAVED 11:25:38 10/31/73

LPASS-BUTTERWORTH 1 10 2500 4000 50  
 WHATIS LPASS  
 (WS L 0.0020576) WC (WP C 1.987E<sup>-6</sup>) WC (WS L 0.0049676) WC WP C 8.2305E<sup>-7</sup>  
 F=500\*110

$Z_t$ . (The filter is to work between equal generator and load impedances  $Z_t$ .) In line [6] the order  $N$  is calculated from a formula similar to one given by Weinberg[15, page 531]. Next, the prototype filter is created, and then it is scaled in impedance and in frequency. The result is the Butterworth lowpass filter. Before the function *BUTTERWORTH* is executed, the function *FSCALE* must be copied from the *MARTHA* library.

In the example, a filter named *LPASS* is defined, then analysed. Note that the way the function *BUTTERWORTH* is defined, the passband criterion is met exactly, and the stopband criterion is more than met. This is possible because the order  $N$  of the filter must be an integer, whereas the formula for  $N$  might produce a noninteger. Next, the function *FINVERT* is copied from the *MARTHA* library. This function creates a highpass filter from a lowpass filter, by replacing capacitors by inductors and vice versa. The filter *HPASS* is created from the filter *LPASS* by interchanging its performance above and below the pivot frequency, 2500 Hz. The definition of *HPASS* is displayed, and then its insertion gain and input impedance are calculated. Another frequency transformation, for creating a bandpass filter from a lowpass filter, is done by the *MARTHA* function *FBP*. The filter *BPASS* is created, then displayed, and next a filter *BSTOP* is created by doing the same transformation on a highpass filter. Note that this transformation replaces inductors by series LC circuits, and capacitors by parallel LC circuits. It requires a left-hand argument equal to the mean frequency of the filter. To be precise, this is the geometric mean of the band edges of the newly defined filter. These two filters, *BPASS* and *BSTOP*, are then analyzed.

Lossy Filter. Next, the effect of finite quality factor  $Q$  of the inductors of the filter is investigated. The function *PROTO* is altered so that inductors are defined with series resistors, and the altered function is displayed. The equivalent resistance is calculated from the  $Q$  and the inductor value, and it is assumed that the  $Q$  is specified at 1 radian per second, the cutoff frequency of the prototype filter. A new lowpass filter is defined, displayed, and analyzed. A comparison of its insertion loss with that of the original lossless filter shows that the finite  $Q$  somewhat degrades the performance.

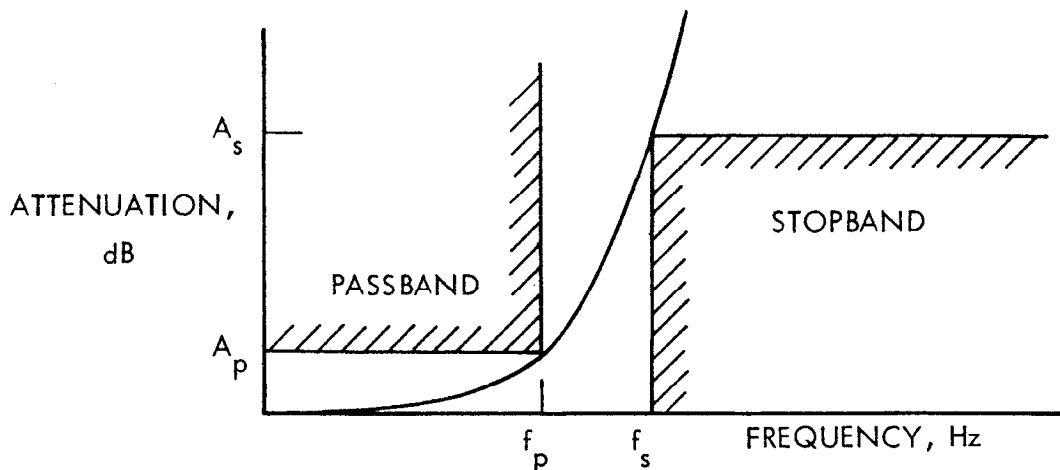


Figure 15. One way of specifying a lowpass filter is with the four numbers  $A_p$ ,  $A_s$ ,  $f_p$ , and  $f_s$ , along with the desired terminating impedance  $Z_t$ .

VI-C. Example 13, Ladder Filter Design

)COPY 100 MARTHA FSCALE  
 SAVED 11:25:38 10/31/73

ZG=50  
 ZL=50  
 PRINT DB IG, MAG ZIN OF LPASS

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/14/73 16:43

F	DB IG	MAG ZIN
5.0000E2	-2.8787E-6	5.0074E1
1.0000E3	-7.3689E-4	5.0818E1
1.5000E3	-1.8846E-2	5.1252E1
2.0000E3	-1.8468E-1	4.3679E1
2.5000E3	-1.0000E0	2.4165E1
3000	-3.2497	8.0396
3500	-6.8315	15.592
4000	-10.835	27.124
4500	-14.703	37.295
5000	-18.279	46.465

)COPY 100 MARTHA FINVERT  
 SAVED 11:25:38 10/31/73

HPASS=2500 FINVERT LPASS  
 WHATIS HPASS

(WS C 1.9697E-6) WC (WP L 0.0020397) WC (WS C 8.1586E-7) WC WP L 0.0049242

PRINT DB IG, MAG ZIN OF HPASS

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/14/73 13:45

F	DB IG	MAG ZIN
500	-50.049	155.09
1000	-25.978	66.968
1500	-12.153	30.653
2000	-4.054	7.9768
2500	-1	24.165
3.0000E3	-2.5395E-1	4.1024E1
3.5000E3	-7.5536E-2	4.8610E1
4.0000E3	-2.6103E-2	5.0978E1
4.5000E3	-1.0192E-2	5.1453E1
5.0000E3	-4.3904E-3	5.1361E1

)COPY 100 MARTHA FBP  
 SAVED 11:25:38 10/31/73

BPASS=102100 FBP LPASS  
 WHATIS BPASS

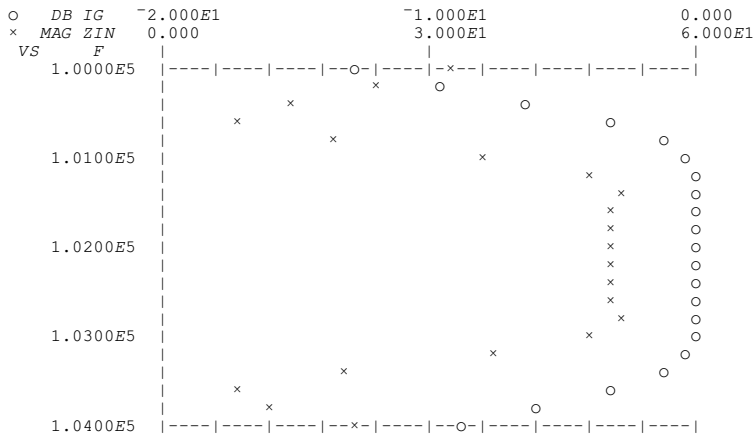
(WS (L 0.0020576) S C 1.1809E-9) WC (WP (C 1.987E-6) P L 1.2229E-6) WC (WS (L 0.0049676) S C 4.8915E-10) WC WP (C 8.2305E-7) P L 2.9523E-6

BSTOP=102100 FBP HPASS  
 WHATIS BSTOP

(WS (C 1.9697E-6) P L 1.2337E-6) WC (WP (L 0.0020397) S C 1.1913E-9) WC (WS (C 8.1586E-7) P L 2.9783E-6) WC WP (L 0.0049242) S C 4.9346E-10

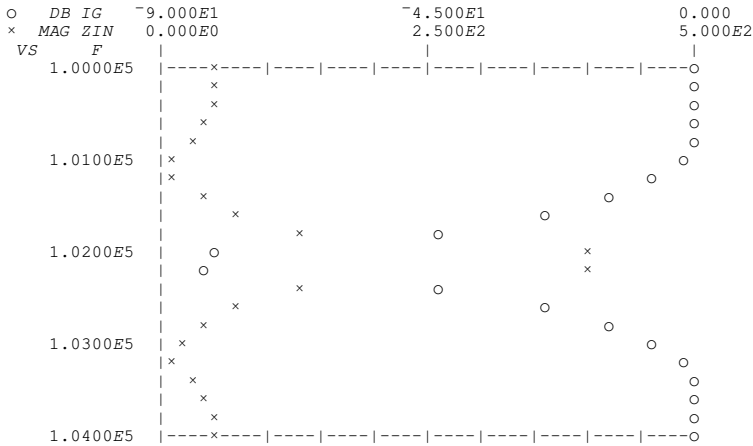
F=1E5+0,200x120  
 PLOT 20 HIGH DB IG, MAG ZIN OF BPASS

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/14/73 13:48



PLOT 20 HIGH DB IG, MAG ZIN OF BSTOP

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:50



```

∇PROTO[4]
[4] B=(WS(L A[K]) S R A[K]+Q) WC B
[5] ∇
    
```

```

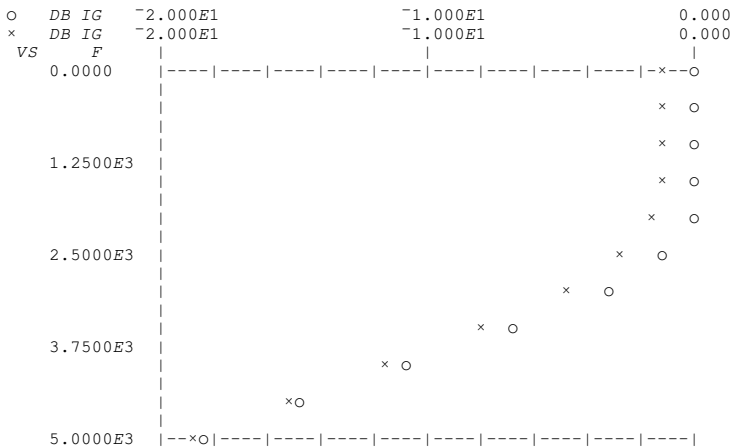
∇PROTO[0]∇
[0] B=PROTO N;A;K
[1] A=2*100(0.5+1N)/N
[2] B=WP C A[K-N]
[3] LOOP:-(0≥K-K-1)/0
[4] B=(WS(L A[K]) S R A[K]+Q) WC B
[5] -(0≥K-K-1)/0
[6] B=(WP C A[K]) WC B
[7] -LOOP
∇
    
```

```

Q=10
NEWLPASS=BUTTERWORTH 1 10 2500 4000 50
WHATIS NEWLPASS
(WS (L 0.0020576) S R 3.8268) WC (WP C 1.987E-6) WC (WS (L 0.0049676) S R 9.2388
) WC WP C 8.2305E-7
    
```

F=0,500x110  
PLOT SS 20 HIGH (DB IG OF LPASS), DB IG OF NEWLPASS

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 16:55



R END OF EXAMPLE 13.

D. Example 14, Waveguide Discontinuity Model

Features illustrated: *FLIMITS*; *RECT*; networks for  $Z_L$ ,  $Z_{NIN}$ . The equivalent circuit given by Marcuvitz[33, Section 5.10] for the window formed from a pair of semicircular obstacles along the sides of a waveguide, Figure 16, is shown in Figure 17, where the reactances have the same frequency dependence as inductors, with inductances

$$L_A = \frac{Z_\infty}{2\pi f_c} \left( \frac{c}{2\pi f_c D} \right)^2 \quad (14a)$$

$$L_B = - \frac{Z_\infty}{16\pi f_c} \left( \frac{2\pi f_c D}{c} \right)^4 \quad (14b)$$

Note that  $L_B$  is negative. In these formulas,  $Z_\infty$  is the characteristic impedance of the waveguide at infinite frequency,  $f_c$  the cutoff frequency, and  $c$  the speed of light. The function *WINDOW* in the example is written so as to expect an argument of length 3 containing the cutoff frequency in Hz, impedance  $Z_\infty$  in ohms, and the diameter  $D$  in meters. The inductances  $L_A$  and  $L_B$  are first defined, and then the network is wired. In the last line, the limits of validity of the model are specified to be  $f_c$  and  $3f_c$ . The function *FLIMITS*, from the workspace 100 *MARTHAW*, specifies limits but does not otherwise affect any calculations. In case, during analysis, any of the frequencies used is outside the range specified, a warning message is printed. This is useful because an important aspect of every model is the frequency range over which it is valid.

After the function is defined, the variable *WR90* is displayed. This contains the dimensions, in meters, of X-band waveguide. The function *RECT*, acting on that, produces the cutoff frequency and infinite-frequency impedance of the guide. These are in a form usable by the function *WG* to define a waveguide characteristic impedance. This is done

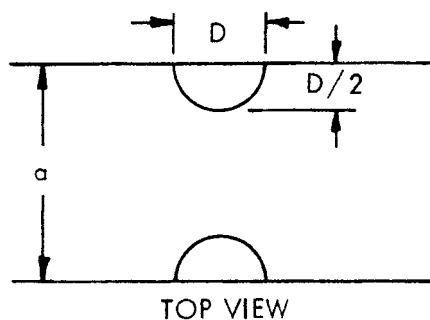


Figure 16. Window formed between two obstacles of circular cross section.

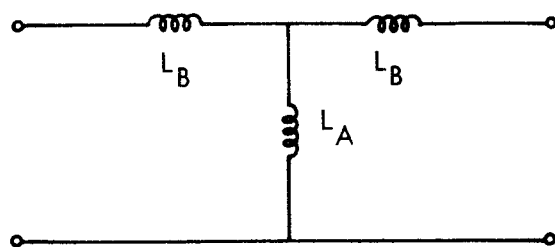


Figure 17. Equivalent circuit for the window of Figure 14, not including any of the waveguide proper. This equivalent circuit is valid for  $f_c < f < 3f_c$ , according to Marcuvitz[33, Section 5.10].

and  $Z_L$  and  $Z_{NIN}$ , the load and input normalization impedance, are set to this frequency-dependent characteristic impedance. Next, the device under investigation is defined, with an obstacle of diameter two millimeters. The input VSWR and admittance are printed, for a range of frequency all larger than  $f_c$ , which is 6.56 GHz. Next, the frequency vector is redefined to include some frequencies less than  $f_c$ , in order to investigate the action of the model below cutoff. Although Marcuvitz[33, Section 5.10] claims the model is not valid below cutoff, there seems to be a reasonable transition between the cutoff and propagating ranges. Nevertheless, because we have defined the limits of validity as given by Marcuvitz, the warning message *INVALID FREQUENCY* is printed.

#### E. Example 15, Waveguide Matching Filter

Features illustrated: *UDE*; *NORM*; two-column form for *NEWELEMENT*. This filter, Figure 18, is identical to the one analyzed as example 9, page 48, of *MARTHA* User's Manual[2]. As a first attempt to analyze it, the discontinuity capacitances of the steps are ignored, as in the previous analysis[2]. The filter is defined simply as the cascade of the three waveguide sections, and the load and input normalization impedances are defined to be the characteristic impedance of the terminating guides. Then the input VSWR is plotted. Note that this plot, which does not appear in the previous analysis, can be made because *MARTHA* now allows  $Z_{NIN}$  and  $Z_L$  to be frequency dependent. The frequency range was chosen to include only the passband, so as to magnify the passband ripple. The plot does not confirm the claim of the designers[34, pp. 302-304] of the filter that it has a nearly equal-ripple response. This is because the discontinuity capacitances were neglected in the analysis.

A better analysis would explicitly account for the discontinuity capacitances, which depend on the dimensions of the adjacent waveguides. In this example, a user-defined element is used. Such elements are generated by the function *UDE* from the library workspace 100 *MARTHA*E, in our case with an argument consisting of four numbers: first the cutoff frequency  $f_c$ ; second the infinite-frequency characteristic impedance  $Z_\infty$ ; and last the two waveguide dimensions,  $b'$  and  $b$ . Since these are given in the drawing in inches, they are converted to meters by multiplying

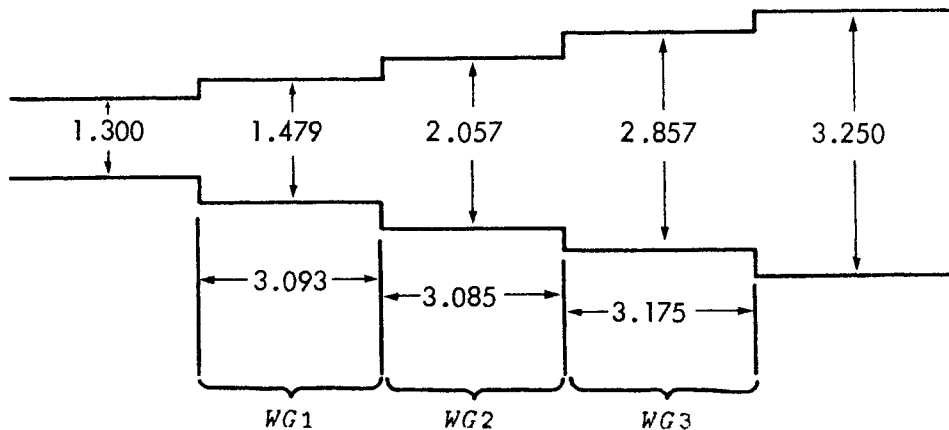


Figure 18. Waveguide matching filter, Example 15. Dimensions in inches. The guides all have the same width, 6.500 inches, and therefore the same cutoff frequency.

```

      * BEGINNING OF EXAMPLE 14

      )LOAD 100 MARTHA
SAVED  11.31.21 10/31/73
      )COPY 100 MARTHA FLIMITS
SAVED  11.25.38 10/31/73
      )COPY 100 MARTHA RECT
SAVED  11.33.39 10/31/73
      )COPY 100 MARTHA WR90
SAVED  11.33.39 10/31/73
      )COPY 100 MARTHA VSWRIN
SAVED  11.30.23 10/31/73

      *B-WINDOW A;X;LA;LB;FC;ZINF;D
[1] FC=A[1]
[2] ZINF=A[2]
[3] D=A[3]
[4] X=3E8+6.28*FC*D
[5] LA=ZINF*(X*2)+6.28*FC
[6] LB=ZINF*(X*4)+16*3.14*FC
[7] B=(WS L LB) WC (WP L LA) WC WS L LB
[8] B=(FC, 3*FC) FLIMITS B
[9] *

      WR90
0.02286 0.01016
      RECT WR90
6.5571E9 376.73

      ZNIN=ZL-WG RECT WR90
      DEVICE-WINDOW RECT WR90,0.002
      F=1E9+6+18

      PRINT VSWRIN, YIN OF DEVICE

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 17:3
MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY

```

F	VSWRIN	RE YIN	IM YIN
7.0000E9	1.2223E0	9.2925E-4	-1.8681E-4
8.0000E9	1.1126E0	1.5208E-3	-1.6238E-4
9.0000E9	1.0820E0	1.8184E-3	-1.4325E-4
1.0000E10	1.0659E0	2.0043E-3	-1.2784E-4
1.1000E10	1.0555E0	2.1315E-3	-1.1512E-4
1.2000E10	1.0481E0	2.2233E-3	-1.0442E-4
1.3000E10	1.0424E0	2.2922E-3	-9.5283E-5
1.4000E10	1.0380E0	2.3455E-3	-8.7368E-5

```

      F=6E9+0,0.1E9+110
      PRINT (Y OF WG RECT WR90), YIN OF DEVICE
INVALID FREQUENCY

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 17:5

```

F	RE Y	IM Y	RE YIN	IM YIN
6.0000E9	0.0000E0	-1.1702E-3	0.0000E0	-1.3895E-3
6.1000E9	0.0000E0	-1.0467E-3	0.0000E0	-1.2623E-3
6.2000E9	0.0000E0	-9.1385E-4	0.0000E0	-1.1258E-3
6.3000E9	0.0000E0	-7.6610E-4	0.0000E0	-9.7461E-4
6.4000E9	0.0000E0	-5.9182E-4	0.0000E0	-7.9693E-4
6.5000E9	0.0000E0	-3.5274E-4	0.0000E0	-5.5456E-4
6.6000E9	3.0201E-4	0.0000E0	3.0205E-4	-1.9861E-4
6.7000E9	5.4522E-4	0.0000E0	5.4528E-4	-1.9553E-4
6.8000E9	7.0306E-4	0.0000E0	7.0314E-4	-1.9254E-4
6.9000E9	8.2633E-4	0.0000E0	8.2642E-4	-1.8964E-4
7.0000E9	9.2915E-4	0.0000E0	9.2925E-4	-1.8681E-4

\* END OF EXAMPLE 14.

```

      * BEGINNING OF EXAMPLE 15.

      )LOAD 100 MARTHA
SAVED  11.31.21 10/31/73
      )COPY 100 MARTHA RECT1
SAVED  11.33.39 10/31/73
      )COPY 100 MARTHA VSWRIN
SAVED  11.30.23 10/31/73

```

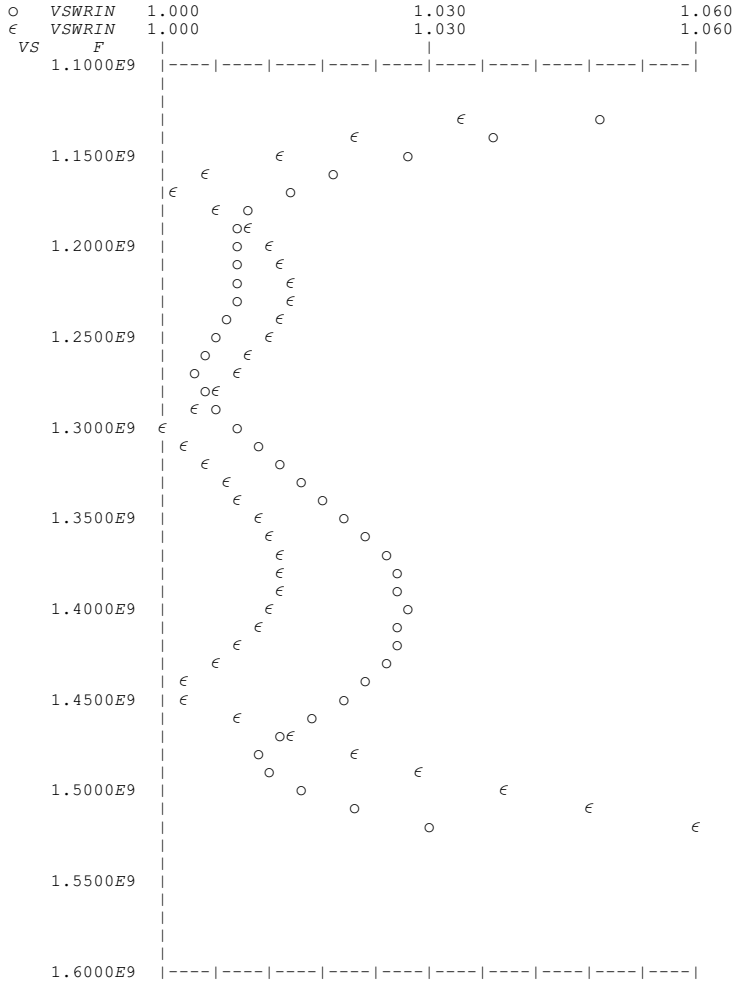




VI-E. Example 15, Waveguide Matching Filter

PLOT SS 'oε' SYMBOLS (VSWRIN OF FILTER), VSWRIN OF EXACT

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 17:24



F--F[1 10 20 30 40]  
PRINT (Y OF DC1), Y OF WGO

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 17:27

F	RE Y	IM Y	RE Y	IM Y
1.1300E9	0.0000E0	1.5833E-5	3.9508E-3	0.0000E0
1.2200E9	0.0000E0	2.1528E-5	4.4327E-3	0.0000E0
1.3200E9	0.0000E0	2.7523E-5	4.8170E-3	0.0000E0
1.4200E9	0.0000E0	3.3241E-5	5.1024E-3	0.0000E0
1.5200E9	0.0000E0	3.8739E-5	5.3222E-3	0.0000E0

)COPY 100 MARTHAM NORM  
SAVED 11:30:53 10/31/73

PRINT ZNIN NORM Z OF DC1

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/14/73 17:28

F	RE NORM	IM NORM
1.1300E9	0.0000E0	-2.4953E2
1.2200E9	0.0000E0	-2.0590E2
1.3200E9	0.0000E0	-1.7502E2
1.4200E9	0.0000E0	-1.5350E2
1.5200E9	0.0000E0	-1.3739E2

␣ END OF EXAMPLE 15.

by 0.0254, and  $f_c$  and  $Z_\infty$  are calculated by the function *RECT1*. The exact filter is then defined. Naturally, before analysis can be done, the actual formulas for this user-defined element must be programmed. The user must write a function entitled *NEWELEMENT*, which will, during analysis, automatically be called. Marcuvitz[33, Section 5.26] has given an equivalent circuit for this discontinuity, which consists of a shunt susceptance, with four formulas, of which the one most suited to this case is

$$\frac{B}{Y_0} = \frac{2b}{\lambda_g} \left( \frac{\delta}{2} \right)^2 \left[ \frac{2 \ln \frac{2}{\delta}}{1 - \delta} + 1 + \frac{17}{16} \left( \frac{b}{\lambda_g} \right)^2 \right] \quad (15)$$

where

$$\delta = 1 - \frac{b'}{b} . \quad (16)$$

The function *NEWELEMENT* in the example is simply a programming of Equations (15) and (16). The quantity returned by *NEWELEMENT* is a matrix with two columns, of which the first is the real part of the impedance, equal to zero, and the second the imaginary part; this is appropriate for one-port user-defined elements. As an illustration of the use of this function, the approximate (denoted  $\circ$ ) and exact (denoted  $\epsilon$ ) VSWR's are plotted. The exact analysis shows the equal-ripple response. It is interesting to note how small a value of shunt admittance is really involved in making the response equal-ripple. The admittance of one of the shunt elements is compared with the waveguide admittance for a few frequencies. This is even more evident in the printout of the normalized impedance, which is more than 100.

#### F. Example 16, Microwave Transistor Amplifier

Features illustrated: *MAKEFOF*; *FROMMAGDEG*; *ISF*; *ZM1*; *ZM2*; *MG*; *RF*; *UG*; *FOF*-functions; *OUTFOF*; *WAVESAT*; *OUTVAR*; use of *FOF*'s for *ZG* and *ZL*; response functions as *FOF*'s. This example deals with a microwave transistor characterized by its *S*-parameters, and the problem of matching this transistor for low noise at the input, and high gain at the output. The transistor is Texas Instruments model 35876E, and numerical values for its *S*-parameters and noise parameters as a function of frequency were obtained from the manufacturer. This example was conceived and originally run by Bernard W. Leake of Raytheon Company, and is used here with his kind permission.

Defining the Transistor. The first task is to define a *FOF* containing the *S*-parameters of the transistor. Since there are four *S*-parameters, each complex, an eight-column *FOF* will be used. Parameters were supplied by the manufacturer as a table of magnitude and angle in degrees, and they are entered into the computer in this form by use of the interactive program *MAKEFOF*. The transistor is named *TR*. After all eight columns have been defined, for the eight frequencies from 1 GHz to 8 GHz, the transistor definition is displayed to check for mistakes. The next task is to convert this data from magnitude-angle form to real-imaginary form. This is done with the function *FROMMAGDEG* from the *MARTHA* library workspace 100 *MARTHAN*. Just to verify that the *S*-parameters look reasonable, they are plotted on the Smith chart. The first two columns are plotted against each other using circles, and the

next two using crosses, etc. Thus the circles stand for  $S_{11}$ , the crosses for  $S_{12}$ , equal signs for  $S_{21}$ , and little circles for  $S_{22}$ . Since the transistor is nearly unilateral,  $S_{12}$  is much smaller than  $S_{21}$ . In fact, only three points of  $S_{21}$  appear on the graph; the others are all larger than one.

The next task is to convert this S-parameter data to an element, using the *MARTHA* function *SFOF*. This is done, and the element is given the same name *TR*.

Next, we look at the various parameters of the transistor, including the invariant stability factor *ISF*, the conjugate-match generator impedance *ZM1* and load impedance *ZM2*, and the matched gain *MG*, reciprocity factor *RF*, and unilateral gain *UG*. The matched gain is, for absolutely stable two-ports, the power gain when input and output are simultaneously conjugate matched. The reciprocity factor *RF* is by definition  $Z_{12}/Z_{21}$ ; its magnitude is the maximum stable gain of the amplifier. The unilateral gain and maximum stable gain are printed partly out of curiosity; no use is actually made of them in the subsequent design.

It is seen that the invariant stability factor is greater than one (and therefore the transistor is unconditionally stable) between 4 and 7 GHz. Outside this range, where the transistor is potentially unstable,

```

      A BEGINNING OF EXAMPLE 16.

      )LOAD 100 MARTHA
SAVED  11.31.21 10/31/73
      )COPY 100 MARTHA MAKEFOF
SAVED  11.33.15 10:31:73

      TR=MAKEFOF 8
      8 COLUMNS FOR 0 FREQUENCIES.
      FREQUENCIES TO CHANGE OR ADD:
      □:
        1E9*18

      COLUMN 1 FOR F = 1E9
      □:
        0.65 0.55 0.53 0.52 0.52 0.52 0.51 0.51

      COLUMN 2 FOR F = 1E9
      □:
        -112 -156 173 154 134 116 98 78

      COLUMN 3 FOR F = 1E9
      □:
        0.074 0.092 0.108 0.124 0.144 0.156 0.16 0.164

      COLUMN 4 FOR F = 1E9
      □:
        26 20 15 10 2 -8 -17 -28

      COLUMN 5 FOR F = 1E9
      □:
        5.21 2.91 2.1 1.43 1.19 0.96 0.82 0.77

      COLUMN 6 FOR F = 1E9
      □:
        100 67 47 21 3 -20 -40 -52

      COLUMN 7 FOR F = 1E9
      □:
        0.72 0.58 0.57 0.61 0.64 0.67 0.69 0.75

      COLUMN 8 FOR F = 1E9
      □:
        -43 -60 -82 -98 -113 -133 -152 -171

      TR
      6.5000E-1  -1.1200E2  7.4000E-2  2.6000E1  5.2100E0  1.0000E2  7.2000E-1  -4.3000E1  1.0000E9
      5.5000E-1  -1.5600E2  9.2000E-2  2.0000E1  2.9100E0  6.7000E1  5.8000E-1  -6.0000E1  2.0000E9
      5.3000E-1  1.7300E2  1.0800E-1  1.5000E1  2.1000E0  4.7000E1  5.7000E-1  -8.2000E1  3.0000E9
      5.2000E-1  1.5400E2  1.2400E-1  1.0000E1  1.4300E0  2.1000E1  6.1000E-1  -9.8000E1  4.0000E9
      5.2000E-1  1.3400E2  1.4400E-1  2.0000E0  1.1900E0  3.0000E0  6.4000E-1  -1.1300E2  5.0000E9
      5.2000E-1  1.1600E2  1.5600E-1  -8.0000E0  9.6000E-1  -2.0000E1  6.7000E-1  -1.3300E2  6.0000E9
      5.1000E-1  9.8000E1  1.6000E-1  -1.7000E1  8.2000E-1  -4.0000E1  6.9000E-1  -1.5200E2  7.0000E9
      5.1000E-1  7.8000E1  1.6400E-1  -2.8000E1  7.7000E-1  -5.2000E1  7.5000E-1  -1.7100E2  8.0000E9

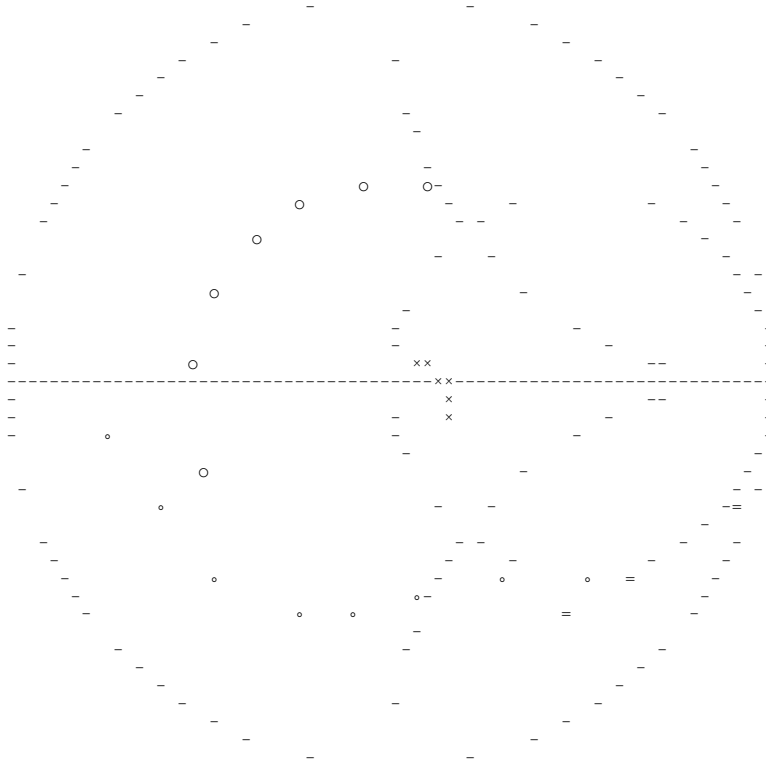
```

```
)COPY 100 MARTHAN FROMMAGDEG
SAVED 11.33.15 10/31/73
)COPY 100 MARTHA F SMITH
SAVED 11.32.43 10/31/73
```

```
TR-FROMMAGDEG TR
SMITH TR
```

```
CIRCUIT ANALYSIS BY MARTHA. 73*D 12/17/73 14:25
MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY
```

```
o VS
x VS
= VS
o VS
```



```
)COPY 100 MARTHA E SFOF
SAVED 11.32.12 10/31/73
)COPY 100 MARTHA R ISF
SAVED 11.30.23 10/31/73
)COPY 100 MARTHA R ZM1
SAVED 11.30.23 10/31/73
)COPY 100 MARTHA R ZM2
SAVED 11.30.23 10/31/73
)COPY 100 MARTHA R MG
SAVED 11.30.23 10/31/73
)COPY 100 MARTHA R RF
SAVED 11.30.23 10/31/73
)COPY 100 MARTHA R UG
SAVED 11.30.23 10/31/73
```

```
TR-SFOF TR
F-1E9*18
PRINT ISF, ZM1, ZM2, DB MG, DB MAG RF, DB UG OF TR
```

```
CIRCUIT ANALYSIS BY MARTHA. 73*D 12/17/73 14:35
```

F	ISF	RE ZM1	IM ZM1	RE ZM2	IM ZM2	DB MG	DBMAG RF	DB UG
1.0000E9	4.6417E-1	0.0000E0	4.1794E1	2.1641E-15	3.4017E1	2.1809E1	1.8476E1	2.2679E1
2.0000E9	8.2457E-1	0.0000E0	1.1878E1	-4.2264E-15	4.8354E1	1.5839E1	1.5001E1	2.0264E1
3.0000E9	9.1943E-1	0.0000E0	-1.6409E0	-1.4386E-15	4.3186E1	1.3253E1	1.2888E1	2.0958E1
4.0000E9	1.0930E0	4.9151E0	-1.4277E1	6.0353E0	3.9976E1	8.7608E0	1.0619E1	1.6371E1
5.0000E9	1.0226E0	2.6970E0	-2.3362E1	2.3127E0	3.1472E1	8.2493E0	9.1718E0	2.1462E1
6.0000E9	1.0753E0	5.4650E0	-3.2723E1	3.0677E0	2.0897E1	6.2164E0	7.8915E0	1.3510E1
7.0000E9	1.1914E0	1.0297E1	-4.2986E1	3.8804E0	1.2236E1	4.4507E0	7.0969E0	8.0757E0
8.0000E9	9.7214E-1	0.0000E0	-5.2472E1	-3.4156E-16	3.3200E0	6.8392E0	6.7165E0	1.4198E1

the conjugate-match impedances are imaginary. Our aim is to make an amplifier at 5 GHz, and that frequency is in the unconditionally stable region.

Noise Characterization. The transistor is characterized for noise, as microwave transistors often are, by its minimum noise figure  $F_{min}$ , optimum generator impedance  $Z_{g,opt}$ , and value of  $R_n$ , equivalent noise resistance, all as functions of frequency. The values used in the example were taken directly from the specification sheet for the transistor. All these depend on bias to some extent, and therefore, the best design would be obtained by using the numbers appropriate for the bias which is actually intended. Three separate FOF's are created, one for  $F_{min}$ , the next for  $R_n$ , and the third, complex, for  $Z_{g,opt}$ . Note that in the definition of the minimum-noise-figure FOF, the values are changed from dB by multiplying by 0.1 and then taking 10 to the resulting power.

```

NFMIN-MAKEFOF 1
1 COLUMNS FOR 0 FREQUENCIES.
FREQUENCIES TO CHANGE OR ADD:
[]:
    1E9*16

COLUMN 1 FOR F = 1E9
[]:
    10*0.1* 2.15 2.25 2.65 3.3 4.1 5

RN-MAKEFOF 1
1 COLUMNS FOR 0 FREQUENCIES.
FREQUENCIES TO CHANGE OR ADD:
[]:
    1E9*16

COLUMN 1 FOR F = 1E9
[]:
    50* 0.56 0.37 0.22 0.18 0.47 1.02

ZGOPT-MAKEFOF 2
2 COLUMNS FOR 0 FREQUENCIES.
FREQUENCIES TO CHANGE OR ADD:
[]:
    1E9*16

COLUMN 1 FOR F = 1E9
[]:
    78 32 24 22 23 24

COLUMN 2 FOR F = 1E9
[]:
    48 27 12 ^4 ^20 ^34

^N-NF
[1] N-NFMIN FADD (RN FDIV FRE ZG) FMUL (FMAG 1 FSUB ZG FDIV ZGOPT) FPWR 2
[2] ^

)COPY 100 MARTHAN FOF
SAVED 11.33.15 10/31/73
)COPY 100 MARTHAN FRE FMAG
SAVED 11.33.15 10/31/73
)COPY 100 MARTHAN FMAG
SAVED 11.33.15 10/31/73
)COPY 100 MARTHAN OUTFOF
SAVED 11.30.23 10/31/73

ZG=50
ZL=50
PRINT DB NFMIN OUTFOF, DB NF OUTFOF, DB TG OF TR

CIRCUIT ANALYSIS BY MARTHA. 73*D 12/17/73 14:44

```

F	DB FOF	DB FOF	DB TG
1.0000E9	2.1500E0	2.6641E0	1.4337E1
2.0000E9	2.2500E0	2.7900E0	9.2779E0
3.0000E9	2.6500E0	3.2042E0	6.4444E0
4.0000E9	3.3000E0	3.8488E0	3.1067E0
5.0000E9	4.1000E0	4.9715E0	1.5109E0
6.0000E9	5.0000E0	6.2749E0	-3.5458E-1
7.0000E9	5.7451E0	7.2760E0	-1.7237E0
8.0000E9	6.3809E0	8.0891E0	-2.2702E0

In a similar way, the values of  $R_n$  are denormalized from 50 ohms.

When a generator impedance  $Z_g$  different from  $Z_{g,opt}$  is used, the noise figure is higher than the  $F_{min}$ . The expression is well known[35, Equation (31)]:

$$NF = F_{min} + \frac{R_n}{\text{Re } Z_g} \left| 1 - \frac{Z_g}{Z_{g,opt}} \right|^2 \quad (17)$$

An APL defined function named *NF* is defined for the noise figure, merely by writing Equation (17) with the various operations replaced by the corresponding FOF functions. These functions, which perform arithmetic on FOF's with one or two columns, are in the library workspace 100 *MARTHAN*. There are functions for adding, subtracting, multiplying, dividing, taking the magnitude, and taking the power, as well as several others that are not used in this particular example. The generator impedance *ZG* is assumed to be either a FOF or a constant.

After this function is defined, the various FOF functions are copied into the active workspace. Note that *FOF* is the name of a group which contains many of the FOF functions required. However, *FRE* and *FMAG* are not in *FOF*, so they are copied separately.

Analysis Without Matching. First we look at the noise figure and transducer gain of the transistor without an input or output matching network, Figure 19b, that is, with the generator and load impedances 50 ohms. The function *OUTFOF* is used in the output list with the quantity immediately to its left being a FOF. In this case *NFMIN* is a FOF which was created by typing in the actual numbers, and *NF* is the name of the APL function defined above which returns a FOF based on Equation (17). Thus, the first column (after the frequency) is the minimum noise figure, the second column is the actual noise figure if the generator impedance is 50 ohms, and the third column is the transducer gain. All these are expressed in dB.

Conjugate Matching. This obviously is not a good design. As a next step in the design we might simultaneously match the input and output, Figure 19c. The response function *ZM1*, the generator impedance which conjugately matches the transistor, is calculated and *ZG* set to it. Next *ZL* is set to the conjugate-match load impedance *ZM2*. Then the minimum noise figure, the actual noise figure with this generator impedance, and the transducer gain, available gain, and matched gain of the transistor are printed. Note that the error message calls attention to the fact that some of the results may be in error, because an attempt was made to divide by zero. A glance at Equation (17) shows that this happens when the generator impedance has zero real part. In the potentially unstable frequency range, *ZM1* has a zero real part, and this is what gives rise to the error report. Thus this report can be ignored since we are only interested in the range around 5 GHz. In the unconditionally stable range, between 4 and 7 GHz, the minimum noise figure (first column) is less than actual noise figure (second column), by as much as 4 dB. In this same range, with these generator and load impedances, the transducer gain, available gain, and matched gain are all equal to each other.

Noise Matching on Input. Since this generator impedance does not lead to low noise, we next analyze the transistor when *ZG* is set to  $Z_{g,opt}$ , Figure 19d. Now, we expect the minimum noise figure and the noise figure to be identical (which they are) and the transducer gain and available gain to be less (which they are). Since the amplifier is intended for a low-noise application, the noise improvement is judged to be

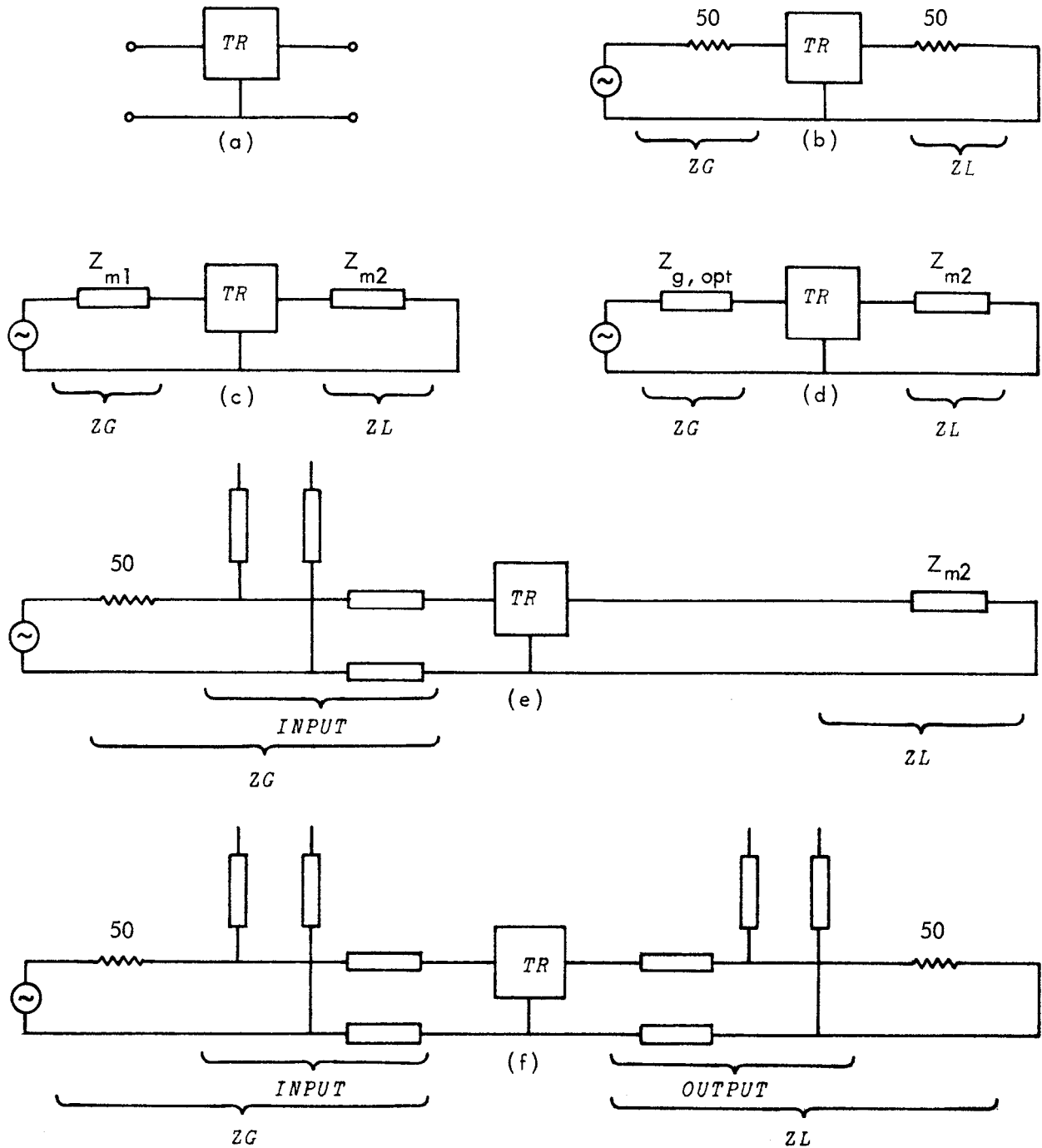


Figure 19. Stages in the design of the microwave transistor amplifier. (a) Transistor. (b) No matching. (c) Conjugate matching on input and output. (d) Noise matching on input. (e) Realizable noise matching on input. (f) Realizable noise match on input and gain match on output.



of greater importance.

A simple transmission-line network using 50-ohm transmission lines, which approximately matches the input for good noise figure, is defined as *INPUT*. Then, the new generator impedance is defined to be the output

```
ZG-ZM1 OF TR
ZL-ZM2 OF TR
PRINT DB NFMIN OUTFOF, DB NF OUTFOF, DB TG, DB AG, DB MG OF TR
ATTEMPT TO DIVIDE BY ZERO
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:46

F	DB FOF	DB FOF	DB TG	DB AG	DB MG
1.0000E9	2.1500E0	1.3440E1	-7.5860E2	-7.5860E2	2.1809E1
2.0000E9	2.2500E0	1.1731E1	-7.5860E2	-7.5860E2	1.5839E1
3.0000E9	2.6500E0	1.1298E1	-7.5860E2	-7.5860E2	1.3253E1
4.0000E9	3.3000E0	5.5555E0	8.7608E0	8.7608E0	8.7608E0
5.0000E9	4.1000E0	8.1575E0	8.2493E0	8.2493E0	8.2493E0
6.0000E9	5.0000E0	7.0089E0	6.2164E0	6.2164E0	6.2164E0
7.0000E9	5.7451E0	6.4170E0	4.4507E0	4.4507E0	4.4507E0
8.0000E9	6.3809E0	1.3489E1	-7.5860E2	-7.5860E2	6.8392E0

```
ZG-ZGOPT
PRINT DB NFMIN OUTFOF, DB NF OUTFOF, DB TG, DB AG, DB MG OF TR
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:48

F	DB FOF	DB FOF	DB TG	DB AG	DB MG
1.0000E9	2.1500E0	2.1500E0	-1.4455E2	1.5588E1	2.1809E1
2.0000E9	2.2500E0	2.2500E0	-1.4246E2	1.1168E1	1.5839E1
3.0000E9	2.6500E0	2.6500E0	-1.4703E2	9.0144E0	1.3253E1
4.0000E9	3.3000E0	3.3000E0	5.9299E0	6.9181E0	8.7608E0
5.0000E9	4.1000E0	4.1000E0	3.9247E0	6.2332E0	8.2493E0
6.0000E9	5.0000E0	5.0000E0	4.0208E0	4.9012E0	6.2164E0
7.0000E9	5.7451E0	5.7451E0	3.5363E0	3.7872E0	4.4507E0
8.0000E9	6.3809E0	6.3809E0	-1.5227E2	4.9071E0	6.8392E0

```
) COPY 100 MARTHAX WAVESAT
SAVED 11.33.39 10/31/73
) COPY 100 MARTHAR OUTVAR
SAVED 11.30.23 10/31/73
```

```
ZG-50
INPUT-(WP WTO TEM 50 0.1189 WAVESAT 5E9) WC TEM 50 0.0216 WAVESAT 5E9
ZG-ZOUT OF INPUT
PRINT ZG OUTVAR, ZGOPT OUTFOF OF TR
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:51

F	RE VAR	IM VAR	RE FOF	IM FOF
1.0000E9	4.8505E1	-7.2612E0	7.8000E1	4.8000E1
2.0000E9	4.4326E1	-1.3347E1	3.2000E1	2.7000E1
3.0000E9	3.8235E1	-1.7427E1	2.4000E1	1.2000E1
4.0000E9	3.1161E1	-1.9158E1	2.2000E1	-4.0000E0
5.0000E9	2.3920E1	-1.8591E1	2.3000E1	-2.0000E1
6.0000E9	1.7097E1	-1.5984E1	2.4000E1	-3.4000E1
7.0000E9	1.1071E1	-1.1631E1	2.5000E1	-4.8000E1
8.0000E9	6.0942E0	-5.7736E0	2.6000E1	-6.2000E1

```
PRINT DB NFMIN OUTFOF, DB NF OUTFOF, DB TG, DB AG, DB MG OF TR
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:52

F	DB FOF	DB FOF	DB TG	DB AG	DB MG
1.0000E9	2.1500E0	2.8117E0	-1.4552E2	1.7236E1	2.1809E1
2.0000E9	2.2500E0	3.2274E0	-1.4422E2	1.1040E1	1.5839E1
3.0000E9	2.6500E0	3.5559E0	-1.4852E2	8.7636E0	1.3253E1
4.0000E9	3.3000E0	3.6533E0	5.4097E0	6.5364E0	8.7608E0
5.0000E9	4.1000E0	4.1051E0	3.7259E0	6.1151E0	8.2493E0
6.0000E9	5.0000E0	5.8017E0	2.9701E0	4.1816E0	6.2164E0
7.0000E9	5.7451E0	8.7074E0	-5.4249E-1	4.3750E-1	4.4507E0
8.0000E9	6.3809E0	1.2561E1	-1.6318E2	-3.5157E0	6.8392E0

impedance of this network, when terminated on the input by 50 ohms. To verify that this design is correct, we print the resulting  $ZG$  next to  $ZGOPT$ . The resulting  $ZG$  is inserted in the output list of an analysis by the function  $OUTVAR$ , and  $ZGOPT$  is inserted in the output list by the function  $OUTFOF$ . Actually, in this case, either  $OUTFOF$  or  $OUTVAR$  would work with either one of these, since they are both FOF's.  $OUTVAR$ , on the other hand, works with any of the allowed forms for  $ZG$ ,  $ZL$ , etc., and  $OUTFOF$  can handle FOF's with any number of columns. It is seen that at 5 GHz, the two impedances are close.

What is more important, of course, is how well the noise figure approximates the minimum noise figure. This is shown in the next analysis, of Figure 19e, where again the first column is the minimum noise figure, next the actual noise figure with this input matching network, and finally the transducer gain, available gain and matched gain. At 5 GHz, the noise figure is very close to its minimum.

Gain Matching on Output. The transducer gain is more than 2 dB lower than the available gain. The available gain is already 2 dB less than the matched gain, but this is a penalty we must pay for matching for low noise at the input. On the other hand, a different output network can raise the transducer gain as high as the available gain without affecting the noise figure. An approximate network which does this, using 75-ohm and 25-ohm transmission lines, is defined as  $OUTPUT$ , and then  $ZL$  is defined as the input impedance of this network when it is terminated with a 50-ohm load at its output. If this is to be a good match, then this  $ZL$  should be the conjugate of  $ZOUT$ . At 5 GHz, it is seen to be a good conjugate match. Next, we print the minimum noise figure, noise figure, transducer gain, available gain, and matched gain of the overall design, Figure 19f. The transducer gain is about 2 dB under the maximum available gain, but the noise figure is very close to optimum. Just to summarize the success of this simple design, the noise figure and minimum noise figure are plotted together (the minimum noise figure, given by crosses, is of course always less than the actual noise figure, given by circles). Next, the transducer gain is compared with the matched gain.

Obviously this is not a complete or adequate design of a microwave transistor amplifier. However, this example does show how many of the advanced features of *MARTHA* can be brought to bear on the design process.

### G. Example 17, Skin Effect in Transmission Lines

Features illustrated: User-defined elements; Six-column matrix for  $NEWELEMENT$ . The loss due to skin effect goes as the square root of frequency. Since this effect is not exactly modeled by any *MARTHA* element, we must use a user defined element. After *MARTHA* is loaded from the public workspace 100 *MARTHA*, the user-defined-element function  $UDE$  is copied from the *MARTHA* library, along with other functions that will be needed.

Element Definition. We wish to model an air-filled transmission line that is either lossless, or is made out of silver, copper, aluminum, or brass. In the lossless case, we may use the *MARTHA* function  $TEM$ , but have the function  $COAX$  calculate the characteristic impedance.

First, the function  $LINE$  is defined as a dyadic APL function, whose left argument is either the letter 'L', or 'F', or 'C', or 'A', or 'B'. The first line contains a branch to one of the various lines in the function for each of the five cases. For the lossless case, the transmission line is calculated using the  $TEM$  function in *MARTHA*. For the other cases, a user-defined element is defined with an argument equal to  $R_s$ ,

ZL-50  
 OUTPUT--(TEM 75 0.1057 WAVESAT 5E9) WC WTO TEM 25 0.125 WAVESAT 5E9  
 ZL-ZIN OF OUTPUT  
 PRINT ZL OUTVAR, ZOUT OF TR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:55

F	RE VAR	IM VAR	RE ZOUT	IM ZOUT
1.0000E9	4.3696E1	-7.7126E0	5.4588E1	-1.1107E2
2.0000E9	3.1758E1	-5.9922E0	4.4255E1	-7.1856E1
3.0000E9	2.1888E1	3.1272E0	2.7257E1	-5.2967E1
4.0000E9	1.5276E1	1.5893E1	1.6917E1	-4.1573E1
5.0000E9	1.0956E1	3.1076E1	1.0908E1	-3.1112E1
6.0000E9	7.9978E0	4.9282E1	8.1970E0	-1.8437E1
7.0000E9	5.7727E0	7.2703E1	8.3659E0	-8.6648E0
8.0000E9	3.8356E0	1.0645E2	7.3622E0	-8.0712E-1

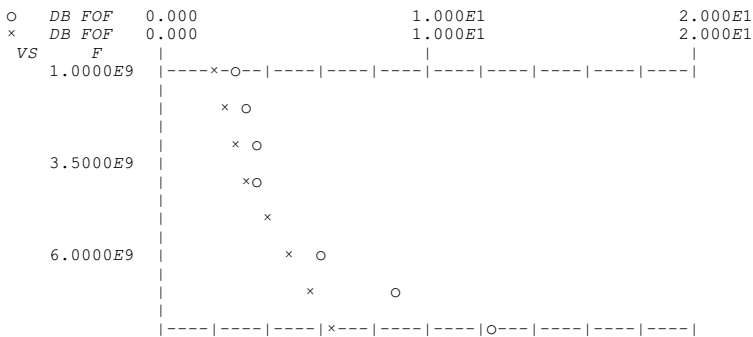
PRINT DB NFMIN OUTFOF, DB NF OUTFOF, DB TG, DB AG, DB MG OF TR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:56

F	DB FOF	DB FOF	DB TG	DB AG	DB MG
1.0000E9	2.1500E0	2.8117E0	1.3272E1	1.7236E1	2.1809E1
2.0000E9	2.2500E0	3.2274E0	7.8058E0	1.1040E1	1.5839E1
3.0000E9	2.6500E0	3.5559E0	5.6397E0	8.7636E0	1.3253E1
4.0000E9	3.3000E0	3.6533E0	4.3865E0	6.5364E0	8.7608E0
5.0000E9	4.1000E0	4.1051E0	6.1150E0	6.1151E0	8.2493E0
6.0000E9	5.0000E0	5.8017E0	-2.4727E0	4.1816E0	6.2164E0
7.0000E9	5.7451E0	8.7074E0	-1.3038E1	4.3750E-1	4.4507E0
8.0000E9	6.3809E0	1.2561E1	-2.3512E1	-3.5157E0	6.8392E0

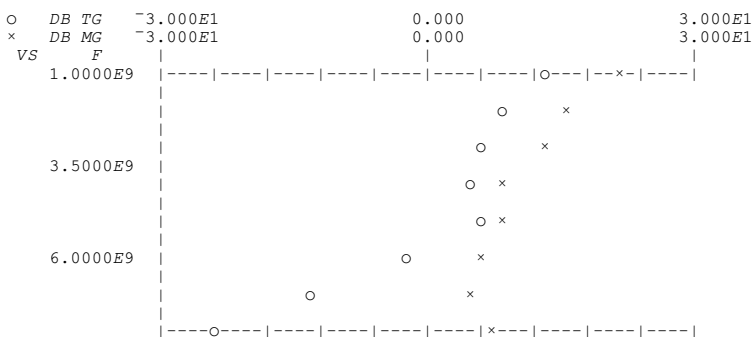
PLOT 14 HIGH SS DB NF OUTFOF, DB NFMIN OUTFOF OF TR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:57



PLOT 14 HIGH SS DB TG, DB MG OF TR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/17/73 14:59



R END OF EXAMPLE 16.

a measure of the resistivity of the material\*, catenated with the argument A. This argument should contain the outer radius of the coax line, the inner radius, and the length, all in meters.

The function *LINE* merely defines a transmission line, and does not perform any of the calculations necessary during analysis of a network containing that line. The function *NEWELEMENT*, written by the user, must be supplied to do the real calculations.

A length of transmission line, even if lossy, is reciprocal. The function *NEWELEMENT* may return a matrix with 6 columns for reciprocal elements, the 6 columns being the real and imaginary parts of A, the real and imaginary parts of B, and the real and imaginary parts of C. The ABCD matrix for this line is given by

$$A = D = \cosh \gamma \ell \quad (18a)$$

$$B = Z_0 \sinh \gamma \ell \quad (18b)$$

$$C = (\sinh \gamma \ell) / Z_0 \quad (18c)$$

where  $\ell$  is the length and

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (19a)$$

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (19b)$$

and where the per-unit-length quantities L, C, R, and G are given by[36, Table 8.09]

$$L = \frac{\mu_0}{2\pi} \ln \frac{r_o}{r_i} \quad (20a)$$

$$C = \frac{2\pi\epsilon_0}{\ln r_o/r_i} \quad (20b)$$

$$G = 0 \quad (20c)$$

$$R = R_s \sqrt{f} \frac{1}{2\pi} \left( \frac{1}{r_o} + \frac{1}{r_i} \right) \quad (20d)$$

where  $r_o$  and  $r_i$  are the outer and inner radii of the line. The function *NEWELEMENT* in the example incorporates these formulas. First, the resistivity measure  $R_s$ , outer and inner radii, and length are extracted from the argument A. Next, the capacitance, inductance, and resistance per unit length are found, with the resistance per unit length  $RR$  proportional to  $\sqrt{f}$ . Next, the real and imaginary parts of the characteristic

---

\*  $R_s$  is actually the coefficient of  $\sqrt{f}$  in the formula for surface resistivity[36, Table 5.14].



impedance are given, followed by the real and imaginary parts of  $\gamma l$ . Next, the output variable  $E$  is set up as a matrix with all columns 0. These are filled in, two columns at a time. Note the calculation of the complex cosh and sinh. The third and fourth columns originally are set to the hyperbolic sine, but then in lines [15]-[17] that result divided by  $Z_0$  is placed in columns 5 and 6. Finally, in line [18], columns 3 and 4 are multiplied by  $Z_0$ . This is a nontrivial program, and some knowledge of APL is necessary either to write it or to read it.

This much typing represents a considerable investment in time on the part of the user, and he may feel it advisable to save these functions just defined, to guard against accidental loss. The active workspace is saved under the name *SKINEFFECT*, but then the analysis continues.

Attenuation of Lossy Line. The attenuation of a 100-foot length of line is calculated next, mainly to check out the programs defined, by comparing the results with the known attenuation of coaxial lines. The function *HF*, which has as an argument the letter corresponding to the material to be used, that is either 'L', 'S', 'C', 'A', or 'B', is defined. Its return  $B$  is set to the insertion gain in dB of a section of line

```

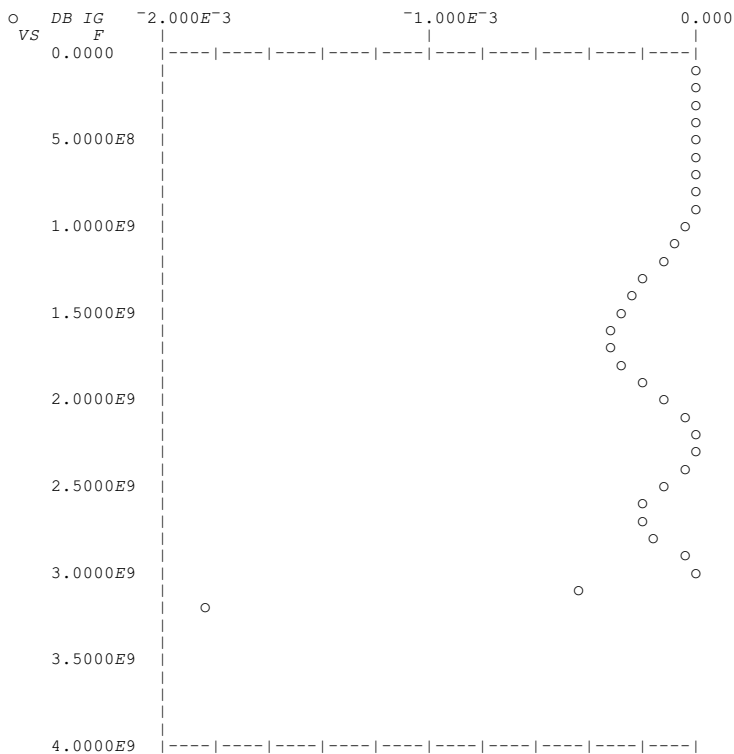
C1-C COAXDISCAP 0.5×0.0254× 0.561 0.244 0.233
C2-C COAXDISCAP 0.5×0.0254× 0.561 0.233 0.268
C3-C COAXDISCAP 0.5×0.0254× 0.561 0.268 0.197
C4-C COAXDISCAP 0.5×0.0254× 0.561 0.197 0.289

VN=FILT MATR;L1;L2;L3;L4
[1] L1-MATR LINE 0.0254× 0.2805 0.1165 0.491
[2] L2-MATR LINE 0.0254× 0.2805 0.1340 0.490
[3] L3-MATR LINE 0.0254× 0.2805 0.0985 0.482
[4] L4-MATR LINE 0.0254× 0.2805 0.1445 0.485
[5] N-C1 WC L1 WC C2 WC L2 WC C3 WC L3 WC C4
[6] N-N WC L4 WC WN N
[7] ∇
    
```

```

F=1E8×132
PLOT 40 HIGH DB IG OF FILT 'L'
    
```

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 9:54



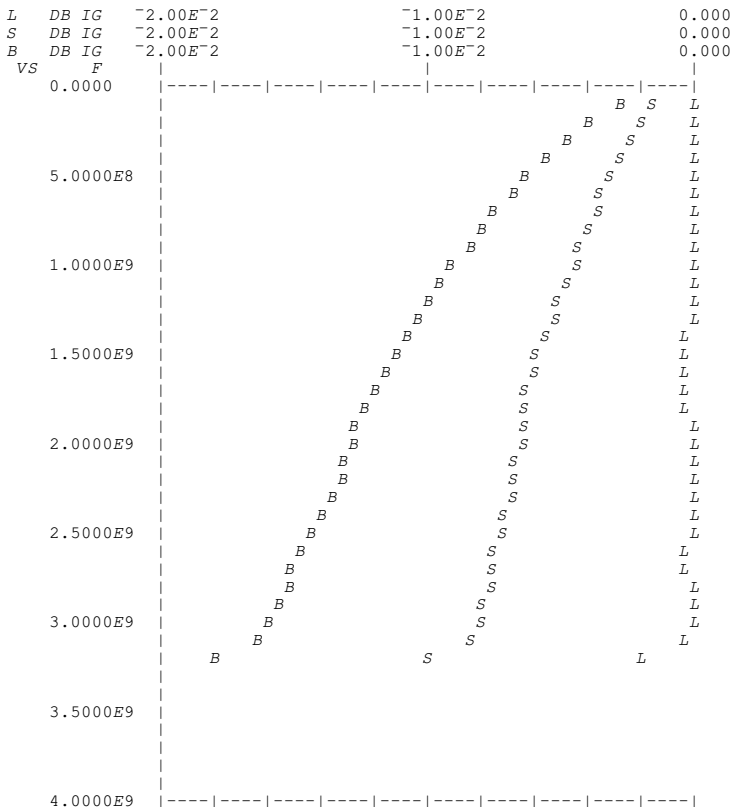
100 feet long, with inner radius 0.122 inches, and outer radius 0.2805 inches. All these dimensions in inches are converted to meters by the factor 0.0254. A plot of the insertion loss (in dB this is the negative of *DB IG*) is made for the five materials, including the lossless case. Naturally, the lossless transmission line has an insertion loss of 0 dB. The others have insertion loss, at 3 GHz, in the range from 2 to 6 dB. This is entirely consistent with published data on loss of coaxial transmission lines of comparable size[37, volume 1, p. 103].

Lossy Filter. Now that the functions are seen to yield reasonable results, we might inquire as to whether using lines made out of these various materials will introduce appreciable loss into coaxial filters. A simple coaxial filter, Figure 20, whose design was originally given by Levy and Rozzi[38] is used for this investigation. The model for the filter, Figure 21, includes the coaxial discontinuity capacitors, and these are defined as C1 through C4. Next, the APL function *FILT* is defined, with an argument specifying the material. Four lengths of transmission line are cascaded alternately with C1 through C4 to form the left half of the filter. Then, in line [6], the overall filter is wired.

The insertion gain of the filter made of lossless material is plotted, and it is seen that the insertion loss has some relatively small ripple, less than 0.001 dB. Next, the insertion gain (again expressed in dB) of both the lossless filter, and filters made out of silver and brass, are plotted. Note that the effect of the resistivity is to introduce a loss much greater than that of the lossless filter. Thus, at 3 GHz, the loss due to resistivity (in the curves labeled *B* and *S*) is

PLOT SS 40 HIGH 'LSB' SYMBOLS (DB IG OF FILT 'L'), (DB IG OF FILT 'S'), (DB IG OF FILT 'B')

CIRCUIT ANALYSIS BY MARTHA. 73\*D 12/18/73 9:58



more than ten times the loss due to reflection alone (curve L). Of course, outside the passband, the reflection loss is greater than the resistivity loss, and this is beginning to become evident in the last point plotted.

It is also interesting to see how the VSWR of the filter is altered by the loss. First, is plotted the VSWR at the input in the lossless case; over the passband, it is less than 1.02. Next, the lossless fil-

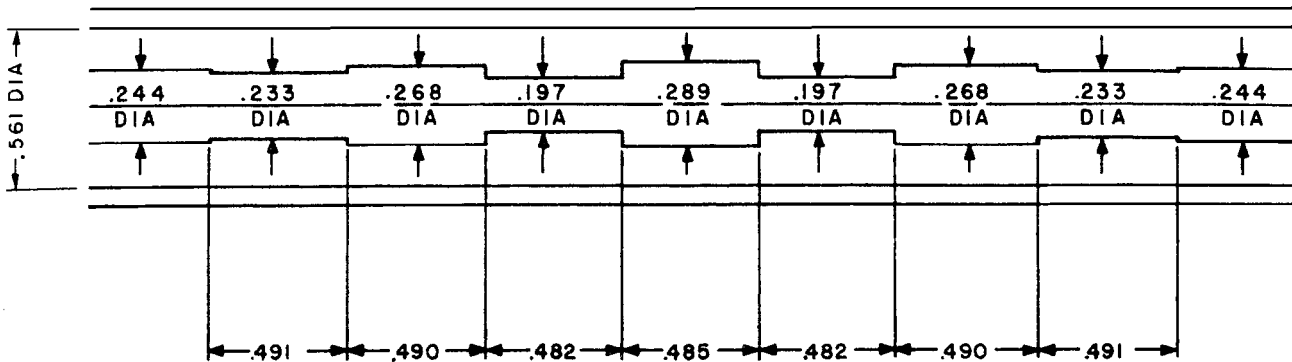
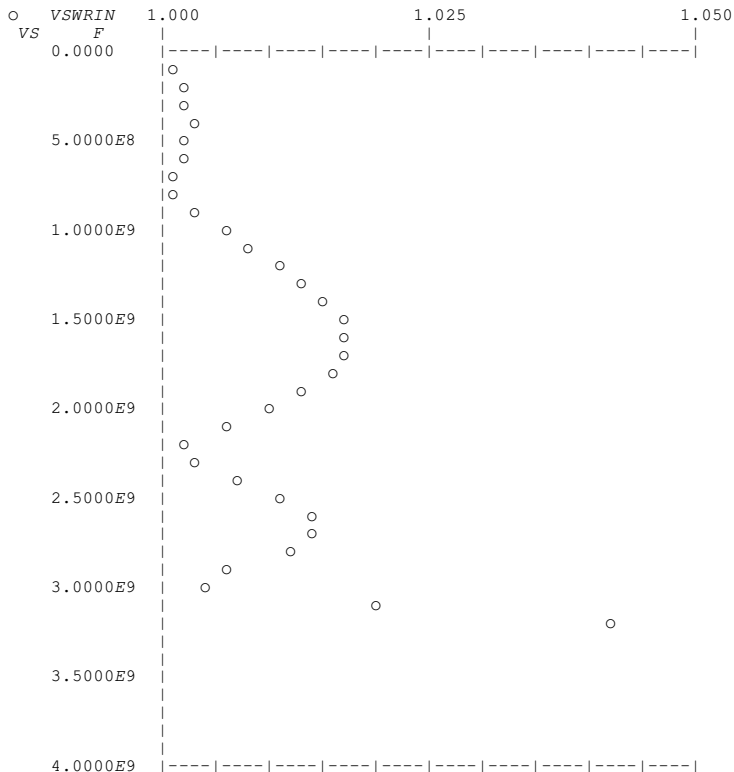


Figure 20. Seven-section coaxial low-pass filter. Dimensions are in inches, and the nominal characteristic impedance of the terminating lines is 50 ohms. The filter has a cutoff frequency of 3 GHz.

)COPY 100 MARTHA VSWRIN  
 SAVED 11.30.23 10/31/73

PLOT 40 HIGH VSWRIN OF FILT 'L'

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:2





ter is compared with a filter made from the most lossy material, brass. The differences in the two responses are very small. That is to say, the standing-wave pattern set up by the filters is relatively insensitive to the resistance in the lines.

H. Example 18, Parametric Amplifier

Features illustrated: *FROMFOF*; Eight-column matrices used as elements; Succession of models of increasing accuracy; Frequency shifting; Representation of three-port networks. This example, worked out by Dr. D. F. Steinbrecher, consists of a 10-GHz parametric amplifier with a 400-MHz passband.

Varactor Model. The varactor-diode model used in the parametric amplifier, Figure 22, is a series resistance and a time-varying capacitor, with an average capacitance  $C_0$ , and half-amplitude components  $C_1$  at the pump frequency, and, later in the example,  $C_2$  at twice the pump frequency. The first model for the parametric amplifier is a simple one with only the idler frequency. It is assumed that the upper-sideband frequency (the sum of the signal frequency and pump frequency) is terminated in a short circuit inside the series resistance. Although this cannot ever be exactly true, it may not be a bad approximation. The second model consists of both the idler and the upper sideband. The third model allows for a possibility of direct coupling between the idler and upper sideband by means of the capacitance coefficient  $C_2$ . This case is usually treated by means of an equivalent three-port network. Although *MARTHA* is restricted in its topology to 2-port networks, it can still handle an equivalent circuit for this three-port network terminated at the upper-sideband frequency.

The equations which govern the small-signal voltage and current at the idler, signal, and upper-sideband frequencies are, in general,

$$V_i^* = R_s I_i^* + U_i^* \tag{21a}$$

$$V_s = R_s I_s + U_s \tag{21b}$$

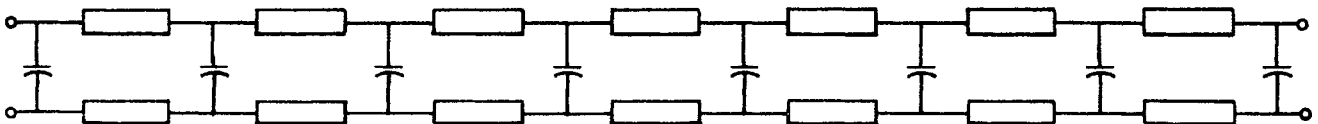


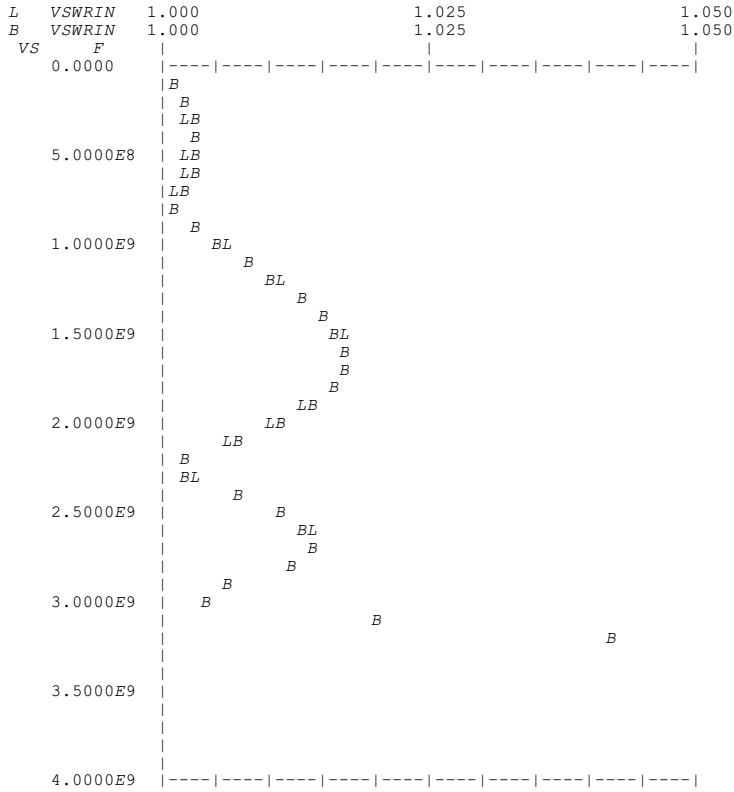
Figure 21. Model for the filter of Figure 18, showing the coaxial discontinuity capacitors.



Figure 22. Varactor diode model, the capacitance  $C(t)$  has components at the pump frequency  $f_p$  and at  $2f_p$ .

PLOT SS 40 HIGH 'LB' SYMBOLS (VSWRIN OF FILT 'L'), VSWRIN OF FILT 'B'

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:5



A END OF EXAMPLE 17.

A BEGINNING OF EXAMPLE 18.

)LOAD 100 MARTHA  
 SAVED 11.31.21 10/31/73  
 )COPY 100 MARTHAN FROMFOF  
 SAVED 11.33.15 10/31/73

∇E-CONV A  
 [1] E←( (ρ, F), 8) ρ0  
 [2] E[;3]←-A[4]÷6.28×(F+A[2])×+/A[3 4]★2  
 [3] E[;4]←-A[3]÷(6.28×F+A[2])×+/A[3 4]★2  
 [4] E[;5]←-A[4]×6.28×F+A[1]  
 [5] E[;6]←-A[3]×6.28×F+A[1]  
 [6] ∇

∇NET-VARACTOR;NI  
 [1] F←F-FP  
 [2] NI←FROMFOF Z OF (C C0) P (R RS) S IDLER  
 [3] F←F+FP  
 [4] NET←(CONV 0, (-FP), C1) WT NI  
 [5] NET←(R RS) S (C C0) P NET  
 [6] ∇

$$V_u = R_s I_u + U_u \quad (21c)$$

$$I_i^* = j(\omega_s - \omega_p) C_0 U_i^* + J_i^* \quad (22a)$$

$$I_s = j \omega_s C_0 U_s + J_s \quad (22b)$$

$$I_u = j(\omega_s + \omega_p) C_0 U_u + J_u \quad (22c)$$

where the newly introduced voltages and currents are related-by

$$J_i^* = j\omega_s C_1^* U_s + j(\omega_s + \omega_p) C_2^* U_u \quad (23a)$$

$$J_s = j(\omega_s - \omega_p) C_1 U_i^* + j(\omega_s + \omega_p) C_1^* U_u \quad (23b)$$

$$J_u = j(\omega_s - \omega_p) C_2 U_i^* + j\omega_s C_1 U_s \quad (23c)$$

We have chosen to write the equations in terms of capacitance coefficients. A similar set of equations using elastance coefficients[39, Section 4.5.1] may be somewhat more exact, and the corresponding *MARTHA* analysis is quite similar.

This full model is to be used in the last part of the example. Initially, we set  $U_u$  to 0, so at first this set of equations reduces to

$$\begin{bmatrix} J_i^* \\ J_s \end{bmatrix} = \begin{bmatrix} 0 & j\omega_s C_1^* \\ j(\omega_s - \omega_p) C_1 & 0 \end{bmatrix} \begin{bmatrix} U_i^* \\ U_s \end{bmatrix} \quad (24)$$

Two-Frequency Analysis. For the first analysis, the voltages and currents at the signal and idler frequencies are assumed to be at different hypothetical ports. Thus, the first model, Figure 23, consists of a two-port network with the input port at signal frequency, and the output port at the idler frequency, or more precisely, the negative of the idler frequency. The converter shown in Figure 23 has the admittance matrix shown in Equation (24).

The frequency dependence shown in the matrix in Equation (24) is not

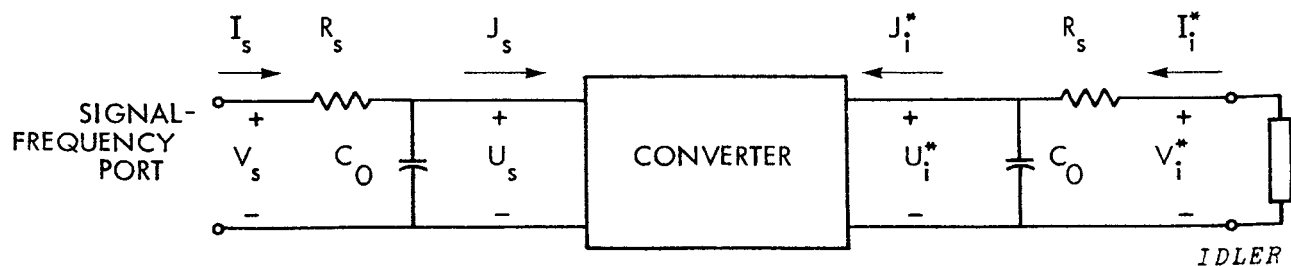


Figure 23. Equivalent circuit for the first analysis. During analysis, the terminating impedance at the idler frequency is denoted *IDLER*.

RS-4  
 CO-0.2E-12  
 C1-0.07E-12 0

F1-9E9+0,0.2E9x110  
 F-F1

PRINT Z OF (C 0.088E-12) P (TEM 10 4 DEGREESAT 10E9) WT (L 0.3E-9) S (C 0.2E-12) P (L 0.15E-9) S C CO

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:13  
 MARTHA COPYRIGHT (C) 1973 MASSACHUSETTS INSTITUTE OF TECHNOLOGY

F	RE Z	IM Z
9.0000E9	0.0000E0	-1.9072E1
9.2000E9	0.0000E0	-1.8112E1
9.4000E9	0.0000E0	-1.7173E1
9.6000E9	0.0000E0	-1.6251E1
9.8000E9	0.0000E0	-1.5346E1
1.0000E10	0.0000E0	-1.4455E1
1.0200E10	0.0000E0	-1.3576E1
1.0400E10	0.0000E0	-1.2708E1
1.0600E10	0.0000E0	-1.1850E1
1.0800E10	0.0000E0	-1.0999E1
1.1000E10	0.0000E0	-1.0154E1

F-10E9

PRINT Z OF WTS TEM 50 0.001

1.0000E10 0.0000 1.0635E1  
 50x14.45÷10.63

67.968

PRINT Z OF WTS TEM 68 0.001

1.0000E10 0.0000 1.4464E1

VUSE LEN;BIASLINE

- [1] BIASLINE-(TEM 10 4 DEGREESAT 10E9) WT (C 0.088E-12) P WTS TEM 68,LEN
- [2] IDLER-(L 0.15E-9) S (C 0.2E-12) P (L 0.3E-9) S BIASLINE
- [3] SIGNAL-(WS BIASLINE S L 0.3E-9) WC (C 0.2E-12) WC WS L 0.15E-9
- [4] V

USE 0.001

F-20E9+0,1E9x130

PRINT Y OF (C CO) P IDLER

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:18

F	RE Y	IM Y
2.0000E10	0.0000E0	5.5202E-2
2.1000E10	0.0000E0	6.6687E-2
2.2000E10	0.0000E0	8.3499E-2
2.3000E10	0.0000E0	1.1242E-1
2.4000E10	0.0000E0	1.8020E-1
2.5000E10	0.0000E0	6.1398E-1
2.6000E10	0.0000E0	-2.9027E-1
2.7000E10	0.0000E0	-9.0919E-2
2.8000E10	0.0000E0	-3.9834E-2
2.9000E10	0.0000E0	-1.4705E-2
3.0000E10	0.0000E0	1.5311E-3
3.1000E10	0.0000E0	1.3973E-2
3.2000E10	0.0000E0	2.4840E-2
3.3000E10	0.0000E0	3.5487E-2
3.4000E10	0.0000E0	4.7173E-2
3.5000E10	0.0000E0	6.1707E-2
3.6000E10	0.0000E0	8.2829E-2
3.7000E10	0.0000E0	1.2143E-1
3.8000E10	0.0000E0	2.3285E-1
3.9000E10	0.0000E0	-6.5888E0
4.0000E10	0.0000E0	-1.7615E-1
4.1000E10	0.0000E0	-7.4996E-2
4.2000E10	0.0000E0	-3.9124E-2
4.3000E10	0.0000E0	-2.0022E-2
4.4000E10	0.0000E0	-7.7356E-3
4.5000E10	0.0000E0	1.0966E-3
4.6000E10	0.0000E0	7.9273E-3
4.7000E10	0.0000E0	1.3489E-2
4.8000E10	0.0000E0	1.8192E-2
4.9000E10	0.0000E0	2.2284E-2
5.0000E10	0.0000E0	2.5926E-2

FP-55E9

F-F1

that of any *MARTHA* element. Therefore the converter in Figure 23 must have its properties calculated during analysis. This can be done either by making it a user-defined element, or, as is done here, by calculating a matrix of the same form that the user-defined function *NEWELEMENT* would ordinarily return. We note that four parameters are really necessary to specify this two-port network. First is the frequency which is added to  $f$  in  $Y_{21}$ , namely  $-f_p$ . Next is the frequency added to  $f$  in  $Y_{12}$ , namely  $0$ . Next are the real and imaginary parts of capacitance half-amplitude  $C_1$ . In the example we define a function *CONV* which expects an argument of length 4. (It happens that one of the parameters in this argument is zero, but later this same function will be used with nonzero values.)

Now let us write the function *CONV*. The ABCD matrix corresponding to the admittance matrix in Equation (24) is

$$A = D = 0 \quad (25a)$$

$$B = - \frac{1}{j\omega C_i} \quad (25b)$$

$$C = j(\omega - \omega_p)C_1 \quad (25c)$$

The function *CONV* returns an eight-column matrix containing the real and imaginary parts of  $A$ ,  $B$ ,  $C$ , and  $D$ . Note the provision for adding to  $f$  in the calculation of both  $B$  and  $C$ .

Next, the function *VARACTOR* is defined, to correspond to Figure 23. At first glance it might appear to be a simple case of terminating the converter in the idler network, and at the same time adding in the series resistance and average capacitance. That is, it might be thought that the expression would be

$$(R \text{ RS}) S (C \text{ CO}) P (\text{CONV } 0, (-FP), C1) WT (C \text{ CO}) P (R \text{ RS}) S \text{ IDLER}$$

However, the idler network has a special requirement, and that is that it be analyzed at frequency  $f - f_p$ , rather than  $f$ . But then the numerical results have to be used with *CONV*, which is analyzed at frequency  $f$ . Thus we must not only shift frequency before and after analysis at the idler frequency, but also preserve the numerical results while hiding from *MARTHA* the fact that they arise from an analysis at a different frequency. *MARTHA* is designed to automatically interpolate or extrapolate numerical answers from one set of frequencies to another, when the answers are in the form of numerically defined elements or FOF's. Thus, although the function *NDE* would force the analysis of the idler network, its result would automatically carry along the idler frequencies, and then undesirable extrapolation would be performed to signal frequency.

There are several ways around this dilemma, but possibly the simplest is to get from *MARTHA* some numerical results without any indication of the frequency, and then re-use them later. The *MARTHA* function *FROMFOF* serves this purpose. The argument is a FOF, in this case the one produced by the response function  $Z$ , and the result is a matrix containing the values of the FOF. The result of the response function  $Z$  is a two-column FOF, and the result of *FROMFOF* is then a two-column matrix.

The function *VARACTOR* is then easy to understand. First, the frequency is shifted down to the negative of the idler frequency, and a portion of the circuit is analyzed at that frequency. A two-column matrix of the resulting impedance is saved before the frequency is returned to

its original value. Next, the remainder of the network is defined, in two steps. Note that the matrix created by *FROMFOF* is the same form that might be generated by the function *NEWELEMENT* for a user-defined element, and therefore may be used as an element, much as is the result returned by the function *CONV*. This is because *MARTHA* expects as a part of a network either an element, or a previously defined network, or a matrix of the form that might be returned by the function *NEWELEMENT*.

The particular amplifier in question has a bias line which is modeled by the equivalent circuit shown in Figure 24. It is part of the design of the amplifier to determine the characteristic impedance of the line *T2*. The criterion is that the overall circuit, modeled in Figure 25, should present a match at the signal frequency of 10 GHz to the average capacitance of the diode. This is most easily done by printing the impedance seen looking back from that transmission line. We see that the impedance is imaginary (the structure is lossless), but in addition, the length must be such that its impedance when short-circuited is equal to  $j14.45$  ohms. The impedance of a 50-ohm line is found to be 10.63 ohms. A simple calculation then shows that a 68-ohm line is nearly correct.

Now that this portion of the design is done, the analysis can continue. It is desired to keep the length of this section of line as a variable, so that it can be adjusted at a later time to optimize the performance of the amplifier. The function *USE*, with an argument of the length to be used, sets up two circuits, denoted *IDLER* and *SIGNAL*, which

are equal to the pertinent portions of the circuit at signal and idler frequencies. Note that each of them uses the bias line as a portion of the circuit. The idler circuit is a one-port network including the lead inductance and capacitance, together with an inductance accounting for the section of waveguide. This is the network seen looking out from the diode at the idler frequency. The signal network, on the other hand, is a two-port network with an input at the point of access, and output inside the case capacitance and lead inductance of the varactor diode. These circuits are set up by executing the function *USE* with the length equal

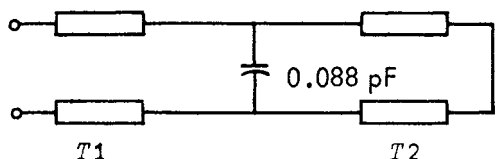


Figure 24. Bias line model. Line T1 has characteristic impedance 10 ohms and an electrical length 4 degrees at the mid frequency, 10 GHz, Line T2 has length 1 mm., and characteristic impedance to resonate the entire structure, modeled in Figure 23, at 10 GHz.

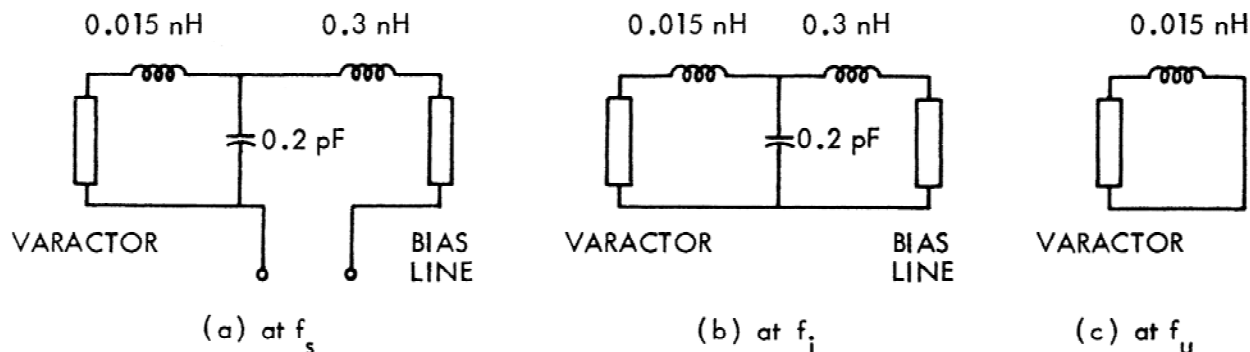


Figure 25. Parametric amplifier model at the various frequencies. The inductor of value 0.3 nH represents a waveguide shunting the mount, and the other two elements represent the varactor package.

to 1 mm.

The next portion of the design is to select the pump frequency. This is done by observing the idler response and selecting an idler resonance that is not tuned too sharply. In the range from 20 to 50 GHz, two resonances are observed, near 29 GHz and 45 GHz. The resonance at 29 GHz is sharper than that at 45 GHz. This is because at the higher frequency, the energy is substantially confined to the lead inductance and case capacitance, whereas at the lower frequency the energy extends out into the bias line. Because the 45 GHz resonance appears to offer more potential for a broad band amplifier, a pump frequency is specified as 55 GHz.

Next, as a test, the impedance of the signal circuit, terminated in the average capacitance of the diode, is printed, and the impedance is seen to pass through a resonance at 10 GHz. This, of course, is because the bias line was designed to accomplish that. Next, the impedance of the signal coupling network, terminated with the varactor, is printed. This circuit is no longer lossless, so its impedance has a real part, but because of the parametric-amplifier action the real part is negative. The device remains tuned at the desired signal frequency of 10 GHz.

The next step in the design is to devise a matching network to transform the negative impedance of approximately 12 ohms to a negative impedance of approximately 40 ohms. A quarter wave transmission line at an intermediate impedance (the geometric mean of the impedances should be used) is defined. Actually, a 22-ohm line would be very nearly exact, but a 20-ohm line is not far from being correct. The reflection coefficient magnitude and phase are calculated and printed.

An awkward feature of this amplifier is now evident. Passive circuits, as the frequency is raised, traverse the Smith chart in a clockwise direction. This response goes around the Smith chart in a counter-clockwise direction because at one point the phase shifts from  $180^\circ$  to  $-180^\circ$  as the frequency is raised. A circuit that does this will be unstable. To see this point another way, analysis is repeated for a smaller value of  $C_1$  than was originally used. The gain is higher, and for some even lower value of  $C_1$ , the gain will be infinite. Therefore, the circuit will oscillate if terminated in the characteristic impedance of the transmission line, 50 ohms.

A different matching network will get around this problem. Instead of matching to 40 ohms in one quarter-wave section, we try two quarter-wave sections. For optimum bandwidth, the impedance of the interface between the quarter-wave lines should be the geometric mean of the two impedances, 13 and 40 ohms, that is, 23 ohms. One transmission line should then have the geometric mean between 12 and 20 ohms, and the other the geometric mean of 40 and the result of the first transformation. One line is actually 16 ohms rather than 16.9 ohms, and the other line 29 ohms. The matching network *MATCH1A* is defined as the two-stage quarter-wave matching network. Now, calculations on this matching network show that the reflection coefficient proceeds from  $-180^\circ$  to  $180^\circ$  as frequency is increased. Therefore the reflection coefficient is moving clockwise around the Smith chart, and the circuit is stable. This is verified by slightly lowering  $C_1$  and finding that gain reduces slightly.

Three-Frequency Analysis. Next, let us consider the sum frequency as well as the idler frequency, but with  $C_2 = 0$ . Now Equations (21) and (22) are still valid, and Equation (23) reduces to

$$J_i^* = j\omega_s C_1^* U_s \quad (26a)$$

PRINT Z OF SIGNAL WT C C0

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:20

F	RE Z	IM Z
9.0000E9	0.0000E0	-9.1235E0
9.2000E9	0.0000E0	-7.2435E0
9.4000E9	0.0000E0	-5.3928E0
9.6000E9	0.0000E0	-3.5682E0
9.8000E9	0.0000E0	-1.7668E0
1.0000E10	0.0000E0	1.4260E-2
1.0200E10	0.0000E0	1.7777E0
1.0400E10	0.0000E0	3.5260E0
1.0600E10	0.0000E0	5.2617E0
1.0800E10	0.0000E0	6.9872E0
1.1000E10	0.0000E0	8.7048E0

PRINT Z OF SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:21

F	RE Z	IM Z
9.0000E9	-1.1686E0	-1.7020E0
9.2000E9	-1.2435E0	-1.3813E0
9.4000E9	-1.2942E0	-1.0393E0
9.6000E9	-1.3107E0	-6.8358E0
9.8000E9	-1.2869E0	-3.2634E0
1.0000E10	-1.2231E0	1.7865E-1
1.0200E10	-1.1264E0	3.3634E0
1.0400E10	-1.0084E0	6.2156E0
1.0600E10	-8.8154E0	8.7210E0
1.0800E10	-7.5610E0	1.0913E0
1.1000E10	-6.3896E0	1.2850E0

(12×40)★0.5

21.909

MATCH1-TEM 20 90 DEGREESAT 10E9

ZN-50

PRINT DB SC, DEG SC OF MATCH1 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:22

F	DB SC	DEG SC
9.0000E9	2.9141E0	1.5212E2
9.2000E9	3.9443E0	1.4878E2
9.4000E9	5.4326E0	1.4620E2
9.6000E9	7.6169E0	1.4613E2
9.8000E9	1.0734E0	1.5379E2
1.0000E10	1.3584E0	-1.7809E2
1.0200E10	1.1763E0	-1.4279E2
1.0400E10	8.2404E0	-1.3056E2
1.0600E10	5.6361E0	-1.2994E2
1.0800E10	3.8768E0	-1.3313E2
1.1000E10	2.6895E0	-1.3750E2

C1

7E-14 0

C1-6.9E-14 0

PRINT DB SC, DEG SC OF MATCH1 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:23

F	DB SC	DEG SC
9.0000E9	2.8805E0	1.5129E2
9.2000E9	3.9152E0	1.4769E2
9.4000E9	5.4229E0	1.4474E2
9.6000E9	7.6686E0	1.4410E2
9.8000E9	1.0983E0	1.5109E2
1.0000E10	1.4253E0	-1.7915E2
1.0200E10	1.2165E0	-1.4053E2
1.0400E10	8.3508E0	-1.2861E2
1.0600E10	5.6512E0	-1.2861E2
1.0800E10	3.8609E0	-1.3222E2
1.1000E10	2.6648E0	-1.3687E2



$$J_s = j(\omega_s - \omega_p)C_1 U_i^* + j(\omega_s + \omega_p)C_1^* U_u \quad (26b)$$

$$J_u = j\omega_s C_1 U_s \quad (26c)$$

An equivalent circuit for all these equations is shown in Figure 26. The function *VARACTOR* is edited to add the calculation of the response at sum frequency. The resulting sum network is viewed through another one of the conversion networks used earlier, and the result put in parallel with the result from the idler. A display of the entire function shows the equivalent circuit of the parametric amplifier to be just that shown in Figure 26. To verify that the new program works properly, we terminate the varactor at the sum frequency in a rather small impedance, and find that the results do not differ much from the previous results. Next, the sum-frequency equivalent circuit is assumed to be that shown in Figure 25, and it is evident that the results of our previous calculations are modified. The point of maximum gain has shifted to 10.2 GHz. To counteract this effect, we look at the impedance without the matching network, and note that it goes through a resonance at 10.2 GHz. A first-order correction for this is to merely change the pump frequency down 200 MHz. As a result, the point of resonance moves back closer to 10 GHz. A two stage quarter-wave matching network *MATCH2* is designed, and the gain and phase of the amplifier using this matching network are calculated. The gain is seen to peak at 10 GHz, and again be stable.

Three-Frequency Analysis with Nonsinusoidal Capacitance. Next, let us add the effect of the second-harmonic capacitance  $C_2$ . The necessary changes to the program are made, to conform to the new circuit in Figure 27, which is consistent with the original three-port Equations (23). Note that this equivalent circuit for the three-port, terminated at two of its ports, allows *MARTHA* to proceed using two-port networks only. A very small value of  $C_2$  is specified, and the impedance seen without the matching network is printed. This is done to check the program, and fortunately, the results are seen to be substantially the same as before. Next, a more reasonable value of  $C_2$  is defined, and the effect of the

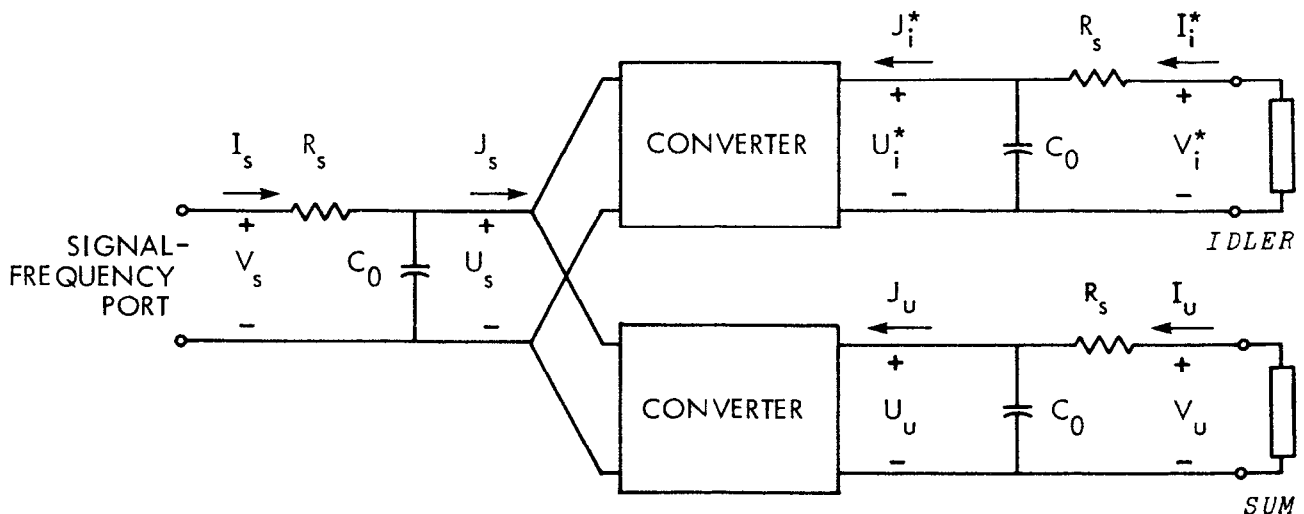


Figure 26. Equivalent circuit for the second analysis. The sum frequency is included, but  $C_2$  is neglected.

second matching network is observed. Although the response is not bad, we decide to go through the exercise of redefining the matching network. The result, *MATCH3*, is then used and the final design (as far as this example is concerned) shows a maximum gain of almost 20 dB at 10.1 GHz, and a bandwidth about 400 MHz.

Bias-Line Tolerance. In order to determine the effect of slight

```

C1=7E-14 0
(12x40)*0.5
21.909
(12x22)*0.5
16.248
(40x16x16x12)*0.5
29.212
MATCH1A-(TEM 29 90 DEGREESAT 10E9) WC TEM 16 90 DEGREESAT 10E9
PRINT DB SC, DEG SC OF MATCH1A WT SIGNAL WT VARACTOR
CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:26

```

F	DB SC	DEG SC
9.0000E9	5.8068E0	-2.1331E1
9.2000E9	7.6013E0	-3.6907E1
9.4000E9	1.0008E1	-5.4434E1
9.6000E9	1.3357E1	-7.6737E1
9.8000E9	1.7884E1	-1.1434E2
1.0000E10	1.9245E1	1.7621E2
1.0200E10	1.4797E1	1.2782E2
1.0400E10	1.1043E1	1.0200E2
1.0600E10	8.3592E0	8.2828E1
1.0800E10	6.3609E0	6.5902E1
1.1000E10	4.8215E0	4.9859E1

```

C1
7E-14 0
C1=6.9E-14 0
PRINT DB SC, DEG SC OF MATCH1A WT SIGNAL WT VARACTOR
CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:28

```

F	DB SC	DEG SC
9.0000E9	5.7163E0	-2.3038E1
9.2000E9	7.4766E0	-3.9063E1
9.4000E9	9.8137E0	-5.7241E1
9.6000E9	1.2987E1	-8.0500E1
9.8000E9	1.6978E1	-1.1834E2
1.0000E10	1.8082E1	1.7865E2
1.0200E10	1.4330E1	1.3155E2
1.0400E10	1.0823E1	1.0467E2
1.0600E10	8.2249E0	8.4756E1
1.0800E10	6.2651E0	6.7347E1
1.1000E10	4.7471E0	5.0970E1

```

C1=7E-14 0
A NOW TO CONSIDER THE SUM FREQUENCY ALSO.
VVARACTOR[0]
[0] NET=VARACTOR;NI;NS
[1] [3.1]
[3.1] F=F+FP
[3.2] NS=FROMFOF Z OF (C C0) P (R RS) S SUM
[3.3] F=F-FP
[3.4] [4.5]
[4.5] NET=NET P (CONV 0,FP,C1[1],-C1[2]) WT NS
[4.6] v
VVARACTOR[ ]v
v NET=VARACTOR;NI;NS
[1] F=F-FP
[2] NI=FROMFOF Z OF (C C0) P (R RS) S IDLER
[3] F=F+FP
[4] F=F+FP
[5] NS=FROMFOF Z OF (C C0) P (R RS) S SUM
[6] F=F-FP
[7] NET=(CONV 0,(-FP),C1) WT NI
[8] NET=NET P (CONV 0,FP,C1[1],-C1[2]) WT NS
[9] NET=(R RS) S (C C0) P NET
v

```

SUM-R 0.01  
PRINT DB SC, DEG SC OF MATCH1A WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:31

F	DB SC	DEG SC
9.0000E9	5.1234E0	-2.1813E1
9.2000E9	6.7657E0	-3.7790E1
9.4000E9	8.9332E0	-5.5788E1
9.6000E9	1.1835E1	-7.8250E1
9.8000E9	1.5494E1	-1.1254E2
1.0000E10	1.7348E1	-1.6980E2
1.0200E10	1.4354E1	1.3953E2
1.0400E10	1.0864E1	1.0981E2
1.0600E10	8.1927E0	8.8527E1
1.0800E10	6.1691E0	7.0339E1
1.1000E10	4.6049E0	5.3432E1

SUM-(L 0.15E-9) S C 0.2E-12  
PRINT DB SC, DEG SC OF MATCH1A WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:33

F	DB SC	DEG SC
9.0000E9	4.5374E0	-5.8624E0
9.2000E9	5.8654E0	-1.8929E1
9.4000E9	7.5970E0	-3.2641E1
9.6000E9	9.9236E0	-4.7500E1
9.8000E9	1.3243E1	-6.5103E1
1.0000E10	1.8496E1	-9.2428E1
1.0200E10	2.4025E1	-1.7028E2
1.0400E10	1.7561E1	1.2211E2
1.0600E10	1.2637E1	9.6322E1
1.0800E10	9.4509E0	7.7844E1
1.1000E10	7.1631E0	6.1217E1

PRINT Z OF SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:34

F	RE Z	IM Z
9.0000E9	-1.4341E1	-2.2761E1
9.2000E9	-1.5194E1	-1.9147E1
9.4000E9	-1.5736E1	-1.5293E1
9.6000E9	-1.5852E1	-1.1300E1
9.8000E9	-1.5484E1	-7.3217E0
1.0000E10	-1.4650E1	-3.5306E0
1.0200E10	-1.3449E1	-6.8132E-2
1.0400E10	-1.2024E1	2.9901E0
1.0600E10	-1.0519E1	5.6392E0
1.0800E10	-9.0479E0	7.9263E0
1.1000E10	-7.6860E0	9.9234E0

FP

5.5E10

FP-54.8E9  
PRINT Z OF SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:35

F	RE Z	IM Z
9.0000E9	-1.5503E1	-2.1279E1
9.2000E9	-1.6044E1	-1.7349E1
9.4000E9	-1.6152E1	-1.3284E1
9.6000E9	-1.5766E1	-9.2394E0
9.8000E9	-1.4908E1	-5.3914E0
1.0000E10	-1.3679E1	-1.8825E0
1.0200E10	-1.2225E1	1.2115E0
1.0400E10	-1.0691E1	3.8868E0
1.0600E10	-9.1945E0	6.1921E0
1.0800E10	-7.8101E0	8.2008E0
1.1000E10	-6.5745E0	9.9900E0

(15×40)★0.5  
 24.495  
 (15×25)★0.5  
 19.365  
 (40×20×20÷15)★0.5  
 32.66  
 MATCH2-(TEM 32 90 DEGREESAT 10E9) WC TEM 20 90 DEGREESAT 10E9  
 PRINT DB SC,DEG SC OF MATCH2 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:37

F	DB SC	DEG SC
9.0000E9	6.0528E0	-2.3890E1
9.2000E9	7.6790E0	-3.9483E1
9.4000E9	9.7068E0	-5.7903E1
9.6000E9	1.2114E1	-8.1890E1
9.8000E9	1.4336E1	-1.1602E2
1.0000E10	1.4666E1	-1.5892E2
1.0200E10	1.2783E1	1.6429E2
1.0400E10	1.0435E1	1.3805E2
1.0600E10	8.3964E0	1.1781E2
1.0800E10	6.7326E0	1.0034E2
1.1000E10	5.3767E0	8.4062E1

⊠ NOW TO ADD EFFECT OF C2.

```

VVARACTOR[7]
[7] NET=(CONV 0,(-FP),C1) WC NI WC CONV (-FP),FP,C2[1],-C2[2]
[8] NET=(NET WPP CONV 0,FP,C1[1],-C1[2]) WT NS
[9] V
VVARACTOR[⊠]V
V NET=VARACTOR;NI;NS
[1] F=F-FP
[2] NI=FROMFOF Z OF(C C0) P(R RS) S IDLER
[3] F=F+FP
[4] F=F+FP
[5] NS=FROMFOF Z OF(C C0) P(R RS) S SUM
[6] F=F-FP
[7] NET=(CONV 0,(-FP),C1) WC NI WC CONV (-FP),FP,C2[1],-C2[2]
[8] NET=(NET WPP CONV 0,FP,C1[1],-C1[2]) WT NS
[9] NET=(R RS) S(C C0) P NET
V
    
```

C2=1E-17 0  
 PRINT Z OF SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:41

F	RE Z	IM Z
9.0000E9	-1.5500E1	-2.1278E1
9.2000E9	-1.6041E1	-1.7349E1
9.4000E9	-1.6149E1	-1.3284E1
9.6000E9	-1.5763E1	-9.2398E0
9.8000E9	-1.4906E1	-5.3920E0
1.0000E10	-1.3678E1	-1.8833E0
1.0200E10	-1.2223E1	1.2106E0
1.0400E10	-1.0690E1	3.8860E0
1.0600E10	-9.1934E0	6.1912E0
1.0800E10	-7.8091E0	8.2001E0
1.1000E10	-6.5737E0	9.9892E0

C2=0.01E-12 0  
 PRINT DB SC, DEG SC OF MATCH2 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73°D 12/18/73 10:42

F	DB SC	DEG SC
9.0000E9	5.6120E0	-3.0923E1
9.2000E9	7.0484E0	-4.7909E1
9.4000E9	8.7269E0	-6.7866E1
9.6000E9	1.0484E1	-9.2674E1
9.8000E9	1.1776E1	-1.2385E2
1.0000E10	1.1805E1	-1.5852E2
1.0200E10	1.0593E1	1.7031E2
1.0400E10	8.9313E0	1.4521E2
1.0600E10	7.3332E0	1.2448E2
1.0800E10	5.9434E0	1.0619E2
1.1000E10	4.7679E0	8.9107E1

PRINT Z OF SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:43

F	RE Z	IM Z
9.0000E9	-1.3038E1	-2.0416E1
9.2000E9	-1.3564E1	-1.6808E1
9.4000E9	-1.3733E1	-1.3066E1
9.6000E9	-1.3484E1	-9.3164E0
9.8000E9	-1.2819E1	-5.7108E0
1.0000E10	-1.1815E1	-2.3800E0
1.0200E10	-1.0591E1	5.9903E-1
1.0400E10	-9.2768E0	3.2125E0
1.0600E10	-7.9787E0	5.4958E0
1.0800E10	-6.7675E0	7.5103E0
1.1000E10	-5.6806E0	9.3232E0

(11×40)★0.5

20.976

(11×21)★0.5

15.199

(40×15×15÷11)★0.5

28.604

MATCH3-(TEM 29 90 DEGREESAT 10E9) WC TEM 15 90 DEGREESAT 10E9  
PRINT DB SC, DEG SC OF MATCH3 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:45

F	DB SC	DEG SC
9.0000E9	4.4472E0	-5.3408E0
9.2000E9	5.8481E0	-1.9108E1
9.4000E9	7.7144E0	-3.4103E1
9.6000E9	1.0273E1	-5.1408E1
9.8000E9	1.3951E1	-7.4514E1
1.0000E10	1.8976E1	-1.1788E2
1.0200E10	1.8805E1	1.6551E2
1.0400E10	1.3785E1	1.2262E2
1.0600E10	1.0131E1	9.8308E1
1.0800E10	7.5546E0	7.8850E1
1.1000E10	5.6398E0	6.0983E1

USE 0.0011

PRINT DB SC, DEG SC OF MATCH3 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:47

F	DB SC	DEG SC
9.0000E9	4.8914E0	-7.9170E0
9.2000E9	6.5099E0	-2.1772E1
9.4000E9	8.7391E0	-3.6667E1
9.6000E9	1.1994E1	-5.3723E1
9.8000E9	1.7391E1	-7.8203E1
1.0000E10	2.5357E1	-1.5901E2
1.0200E10	1.7285E1	1.2125E2
1.0400E10	1.1899E1	9.6113E1
1.0600E10	8.6202E0	7.7698E1
1.0800E10	6.3470E0	6.0882E1
1.1000E10	4.6734E0	4.4618E1

USE 0.0009

PRINT DB SC, DEG SC OF MATCH3 WT SIGNAL WT VARACTOR

CIRCUIT ANALYSIS BY MARTHA. 73-D 12/18/73 10:48

F	DB SC	DEG SC
9.0000E9	4.1214E0	-3.1496E0
9.2000E9	5.3645E0	-1.6893E1
9.4000E9	6.9723E0	-3.2036E1
9.6000E9	9.0623E0	-4.9619E1
9.8000E9	1.1766E1	-7.2199E1
1.0000E10	1.4892E1	-1.0613E2
1.0200E10	1.6272E1	-1.5745E2
1.0400E10	1.3953E1	1.5562E2
1.0600E10	1.0906E1	1.2481E2
1.0800E10	8.4056E0	1.0175E2
1.1000E10	6.4383E0	8.1688E1

END OF EXAMPLE 18.

changes in the bias-line filter, two different lengths are used, each 10% off. The slightly longer length raises the maximum gain about 5 dB, and the slightly lower length lowers it by a few dB and shifts the point of maximum gain. The relatively high sensitivity of the gain to this parameter indicates that there might be tolerance difficulties in the construction of the parametric amplifier. Probably a new bias circuit should be designed that would more effectively exclude signal-frequency energy.

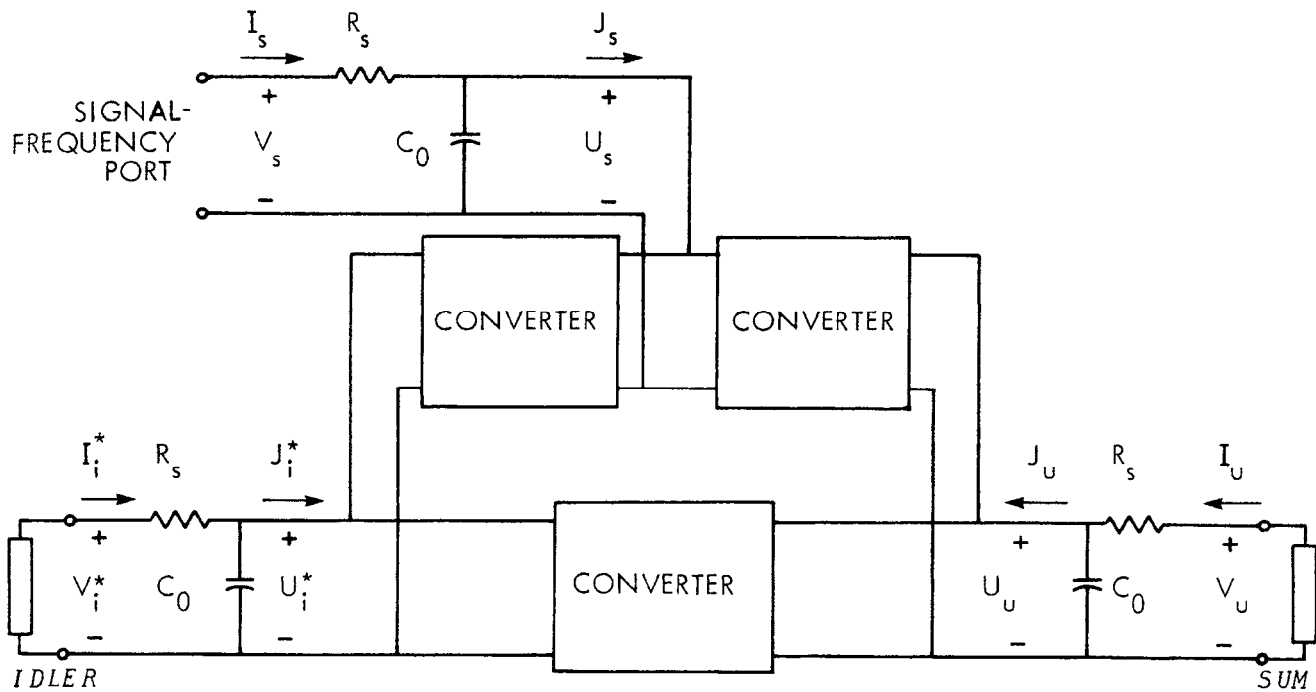


Figure 27. Equivalent circuit of varactor as terminated at idler and sum frequencies. The upper two converters are between frequencies separated by  $f_p$ , and therefore use  $C_1$ . The lower converter is between frequencies that differ by  $2f_p$ , and therefore uses  $C_2$ .

## VII. FUTURE CHANGES TO *MARTHA*

### A. Obsolete Functions

Experience has shown that the functions in Table III, from the *MARTHA* library, are seldom used. Most of them will probably be omitted in future versions of *MARTHA*. You are advised not to use them. In case there are any on this list which you have found useful, and if you feel that alternate functions do not serve your purposes as well, you are urged to communicate this fact to the author.

### B. User-Defined Elements

User-defined elements are now defined slightly differently. See Section II-B. The old method, which is that any vector starting with 9 is considered to be a user-defined element, still works temporarily, but it will not work in future versions of *MARTHA*.

### C. Options

A series of options which will expand the scope of services performed by *MARTHA* is now under development. The options will be numbered, with option 0 being normal *MARTHA*. Any options available on your computer will be described in the workspace 100 *HOWMARTHA*. Different options are called for by copying different functions named *OPT*. At any time you can type *OPT* to find out what option you are currently using.

Table III. Functions that are not guaranteed to be supported in future versions of *MARTHA*.

<u>WORKSPACE</u>	<u>FUNCTION</u>	<u>ALTERNATIVE</u>
100 <i>MARTHAE</i>	<i>RTEE, RPI</i> <i>LTEE, LPI</i> <i>CTEE, CPI</i>	Use <i>R, L, C, WP, WS, WC</i>
100 <i>MARTHAE</i>	<i>RROTATOR, RREFLECTOR</i> <i>LROTATOR, LREFLECTOR</i> <i>CROTATOR, CREFLECTOR</i> <i>VSCALOR, ISCALOR</i> <i>PSCALOR</i>	<i>ZPDE, YPDE, HPDE, ABCDPDE</i>
100 <i>MARTHAW</i>	<i>WTM</i>	<i>ZI1</i>
100 <i>MARTHAR</i>	<i>OCVG</i> <i>SCCG</i>	<i>G21</i> <i>-H21</i>

## VIII. DOCUMENTATION

The workspace 100 *HOWMARTHA* contains, on-line, extensive documentation about how to use *MARTHA*. The next 14 pages contain this documentation as of the date of publication of this Addendum.

*MARTHA* 73•D 1 NOVEMBER 1973

*MARTHA* IS A SET OF FUNCTIONS THAT ANALYZE LINEAR ELECTRICAL NETWORKS, NORMALLY AS A FUNCTION OF FREQUENCY. FOR A COMPLETE DESCRIPTION, SEE PAUL PENFIELD JR., 'MARTHA USER'S MANUAL,' THE MIT PRESS, CAMBRIDGE, MASS. 02142; 1971. ALSO SEE PAUL PENFIELD JR., 'MARTHA USER'S MANUAL, 1973 ADDENDUM,' THE MIT PRESS, CAMBRIDGE, MASS. 02142; 1974. FOR A SUCCINCT SUMMARY:

)LOAD 100 *HOWMARTHA*

THIS WORKSPACE CONTAINS THE FOLLOWING SUMMARIES (BUT NOT THE FUNCTIONS)

DESCRIBE	WIRING	FORMATS	HINTS	OBSOLETE	COMMENTS
USAGE	RESPONSES	FOF	EXAMPLES	ERRORS	AUTHORIZATION
ELEMENTS	MODIFIERS	EXTRA	CHANGES	ORDERFORM	

TO GET THE FUNCTIONS:

)LOAD 100 *MARTHA*

*MARTHA* INCLUDES SEVERAL VARIABLES AND FUNCTIONS THAT THE USER DOES NOT DIRECTLY USE. THESE ALL HAVE NAMES CONTAINING TWO UNDERLINED LETTERS. THERE ARE 7 VARIABLES THAT THE USER CAN CHANGE:

VARIABLE	PRESET TO	MEANING
<u>F</u>	1 2	FREQUENCY VECTOR, IN HZ
<u>EG</u>	1	R.M.S. GENERATOR VOLTAGE IN VOLTS
<u>ZG</u>	0	GENERATOR IMPEDANCE IN OHMS
<u>ZL</u>	1E25	LOAD IMPEDANCE IN OHMS
<u>ZN</u>	50	NORMALIZATION IMPEDANCE OF 1-PORT, IN OHMS
<u>ZNIN</u>	50	NORMALIZATION IMPEDANCE AT INPUT, IN OHMS
<u>ZNOUT</u>	50	NORMALIZATION IMPEDANCE AT OUTPUT, IN OHMS

*MARTHA* INCLUDES 76 OTHER VARIABLES AND FUNCTIONS. THE *MARTHA* LIBRARY CONTAINS MORE THAN 150 ADDITIONAL OBJECTS IN 7 WORKSPACES:

WORKSPACE	CONTENTS
100 <i>MARTHAE</i>	ELEMENTS
100 <i>MARTHAW</i>	WIRING FUNCTIONS
100 <i>MARTHAR</i>	RESPONSE FUNCTIONS
100 <i>MARTHAM</i>	MODIFIERS FOR RESPONSE FUNCTIONS
100 <i>MARTHAF</i>	FORMAT FUNCTIONS
100 <i>MARTHAN</i>	NUMERICALLY-DEFINED-FOF FUNCTIONS
100 <i>MARTHAX</i>	EXTRA FUNCTIONS TO WORK WITH <i>MARTHA</i>

TO GET A SPECIFIC FUNCTION, E.G. *VCCS* FROM 100 *MARTHAE*:

)COPY 100 *MARTHAE VCCS*

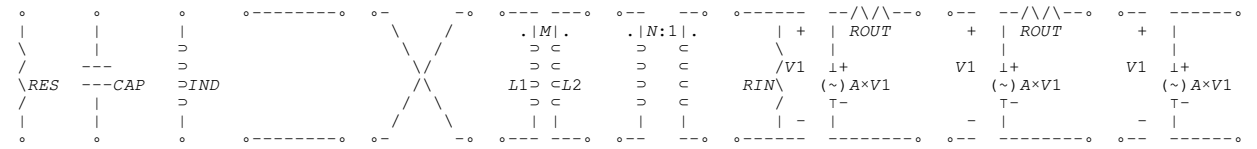
FOR A SUMMARY OF THE CONTENTS OF THE *MARTHA* LIBRARY, PRINT THE VARIABLE NAMED 'DESCRIBE' IN EACH WORKSPACE.



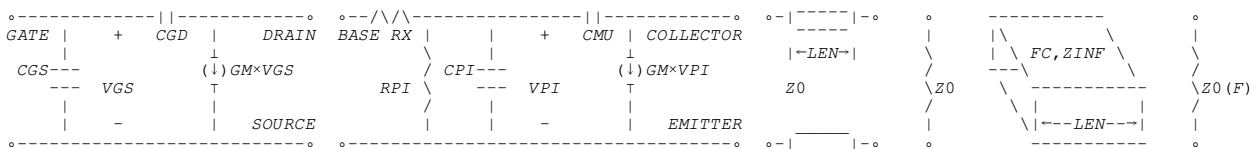
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ELEMENTS IN MARTHA:

ELEMENT	TYPE	NAME	ARGUMENT VECTOR	REQUIRED	EQUATIONS
RESISTOR	1-PORT	R	RESISTANCE RES IN OHMS		$V=RES \times I$
CAPACITOR	1-PORT	C	CAPACITANCE CAP IN FARADS		$I=S \times CAP \times V$
INDUCTOR	1-PORT	L	INDUCTANCE IND IN HENRIES	IND#0	$V=S \times IND \times I$
STRAIGHT-THROUGH CONNECTION	2-PORT	WTHRU	(NONE)		$V1=V2; I1=-I2$
POLARITY REVERSE	2-PORT	WR	(NONE)		$V1=-V2; I1=I2$
MUTUAL INDUCTOR	2-PORT	L	INPUT SELF-INDUCTANCE L1 IN HENRIES OUTPUT SELF-INDUCTANCE L2 IN HENRIES MUTUAL INDUCTANCE M IN HENRIES	M#0	$V1=S \times (L1 \times I1) + M \times I2$ $V2=S \times (M \times I1) + L2 \times I2$
IDEAL TRANSFORMER	2-PORT	IT	TURNS RATIO N	N#0	$V1=N \times V2; I1=-I2 \div N$
OPERATIONAL AMPLIFIER	2-PORT	OPAMP	OPEN-CIRCUIT VOLTAGE GAIN A OUTPUT IMPEDANCE ROUT IN OHMS INPUT IMPEDANCE RIN IN OHMS	A#0 RIN#0	$V1=RIN \times I1$ $V2=(A \times V1) + ROUT \times I2$
OPERATIONAL AMPLIFIER	2-PORT	OPAMP	OPEN-CIRCUIT VOLTAGE GAIN A OUTPUT IMPEDANCE ROUT IN OHMS	A#0	$I1=0$ $V2=(A \times V1) + ROUT \times I2$
OPER. AMPLIFIER	2-PORT	OPAMP	VOLTAGE GAIN A	A#0	$I1=0; V2=A \times V1$
FIELD-EFFECT TRANSISTOR MODEL, GROUNDED-SOURCE	2-PORT	FET	GATE-SOURCE CAPACITANCE CGS IN FARADS GATE-DRAIN CAPACITANCE CGD IN FARADS TRANSCONDUCTANCE GM IN MHOS	GM#0	$I1=S \times (CGS \times V1) + CGD \times V1 - V2$ $I2=(GM \times V1) + S \times CGD \times V2 - V1$
BIPOLAR-TRANSISTOR MODEL, GROUNDED-EMITTER	2-PORT	HYBRIDPI	RESISTANCE RX IN OHMS RESISTANCE RPI IN OHMS CAPACITANCE CPI IN FARADS CAPACITANCE CMU IN FARADS TRANSCONDUCTANCE GM IN MHOS	RPI#0 GM#0	$V1=VPI + RX \times I1$ $I1=(VPI \times (S \times CPI) + RPI) + S \times CMU \times VPI - V2$ $I2=(GM \times VPI) + S \times CMU \times V2 - VPI$
LOSSLESS TRANSMISSION LINE	2-PORT	TEM	CHARACTERISTIC IMPEDANCE Z0 IN OHMS LENGTH LEN IN METERS	Z0#0	$A=J \times \omega \times 2 \times LEN \times F + 3E8 \div DIEL \star .5$ $(V2 - Z0 \times I2) = A \times V1 + Z0 \times I1$ $(V1 - Z0 \times I1) = A \times V2 + Z0 \times I2$
TRANSMISSION LINE CHARACTERISTIC IMPEDANCE	1-PORT	TEM	CHARACTERISTIC IMPEDANCE Z0 IN OHMS	Z0#0	$V=Z0 \times I$
LOSSLESS WAVEGUIDE DOMINANT MODE	2-PORT	WG	CUTOFF FREQUENCY FC IN HERTZ INFINITE-FREQUENCY CHARACTERISTIC IMPEDANCE ZINF IN OHMS LENGTH LEN IN METERS	ZINF#0 ~FC<F	$Z0=ZINF \div (1 - (FC \div F) \star 2) \star .5$ $A=J \times LEN \times ZINF \times F + Z0 \times 3E8 \div DIEL \star .5$ $(V2 - Z0 \times I2) = (\star - \circ 2 \times A) \times V1 + Z0 \times I1$ $(V1 - Z0 \times I1) = (\star - \circ 2 \times A) \times V2 + Z0 \times I2$
WAVEGUIDE CHARACTERISTIC IMPEDANCE	1-PORT	WG	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS	ZINF#0 ~FC<F	$V=I \times ZINF \div (1 - (FC \div F) \star 2) \star .5$



R RES C CAP L IND WTHRU WR L L1,L2,M IT N OPAMP A,ROUT,RIN OPAMP A,ROUT OPAMP A



FET CGS,CGD,GM HYBRIDPI RX,RPI,CPI,CMU,GM TEM Z0,LEN TEM Z0 WG FC,ZINF,LEN WG FC,ZINF

FOR ELEMENTS IN 100 MARTHAE, PRINT THE VARIABLE DESCRIBE IN THE WORKSPACE 100 MARTHAE

MARTHA 73-D 1 NOVEMBER 1973

FOR A SUMMARY OF ELEMENTS IN MARTHA, SEE ELEMENTS IN 100 HOWMARTHA

ELEMENTS IN 100 MARTHA:

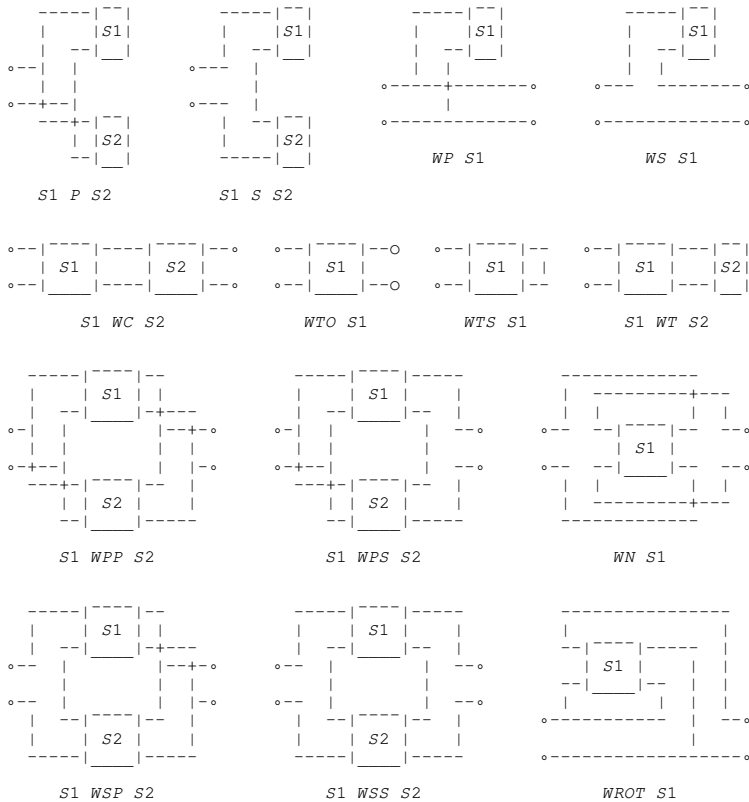
ELEMENT	TYPE	NAME	ARGUMENT VECTOR	REQUIRED	EQUATIONS
SIMPLE TRANSISTOR MODEL	2-PORT	BETAIC	CURRENT GAIN BETA COLLECTOR BIAS CURRENT IC IN AMPS	BETA#0 IC#0	$I_2 = -BETA \times I_1$ $V_1 = I_1 \times BETA + 40 \times IC$
ATTENUATOR	2-PORT	ATTENUATOR	CHARACTERISTIC IMPEDANCE Z0 IN OHMS ATTENUATION A IN DECIBELS	Z0#0 1000> A	$(V_2 + Z_0 \times I_2) = (10 \star A + 20) \times V_1 - Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star A + 20) \times V_2 - Z_0 \times I_2$
WAVEGUIDE ATTENUATOR	2-PORT	WGATTEN	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS ATTENUATION A IN DECIBELS	ZINF#0 ~FC#F 700> A	$Z_0 = ZINF \div (1 - (FC \div F) \star 2) \star .5$ $(V_2 + Z_0 \times I_2) = (10 \star A + 20) \times V_1 - Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star A + 20) \times V_2 - Z_0 \times I_2$
ISOLATOR	2-PORT	ISOLATOR	CHARACTERISTIC IMPEDANCE Z0 IN OHMS FORWARD LOSS B IN DECIBELS REVERSE LOSS A IN DECIBELS	Z0#0 1000> B 1000> A	$(V_2 + Z_0 \times I_2) = (10 \star A + 20) \times V_1 - Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star B + 20) \times V_2 - Z_0 \times I_2$
ISOLATOR WITH INF. REVERSE LOSS	2-PORT	ISOLATOR	CHARACTERISTIC IMPEDANCE Z0 IN OHMS FORWARD LOSS B IN DECIBELS	Z0#0 1000> B	$V_1 = Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star B + 20) \times V_2 - Z_0 \times I_2$
IDEAL ISOLATOR	2-PORT	ISOLATOR	CHARACTERISTIC IMPEDANCE Z0 IN OHMS	Z0#0	$V_1 = Z_0 \times I_1; (V_2 - Z_0 \times I_2) = V_1 + Z_0 \times I_1$
WAVEGUIDE ISOLATOR	2-PORT	WGISOLATOR	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS FORWARD LOSS B IN DECIBELS REVERSE LOSS A IN DECIBELS	ZINF#0 ~FC#F 700> B 700> A	$Z_0 = ZINF \div (1 - (FC \div F) \star 2) \star .5$ $(V_2 + Z_0 \times I_2) = (10 \star A + 20) \times V_1 - Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star B + 20) \times V_2 - Z_0 \times I_2$
WAVEGUIDE ISOLATOR WITH INF. REVERSE LOSS	2-PORT	WGISOLATOR	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS FORWARD LOSS B IN DECIBELS	ZINF#0 ~FC#F 700> B	$Z_0 = ZINF \div (1 - (FC \div F) \star 2) \star .5$ $V_1 = Z_0 \times I_1$ $(V_1 + Z_0 \times I_1) = (10 \star B + 20) \times V_2 - Z_0 \times I_2$
WAVEGUIDE IDEAL ISOLATOR	2-PORT	WGISOLATOR	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS	ZINF#0 ~FC#F	$Z_0 = ZINF \div (1 - (FC \div F) \star 2) \star .5$ $V_1 = Z_0 \times I_1; (V_2 - Z_0 \times I_2) = V_1 + Z_0 \times I_1$
GYRATOR	2-PORT	GYRATOR	CHARACTERISTIC RESISTANCE Z0 IN OHMS	Z0#0	$V_1 = -Z_0 \times I_2; V_2 = Z_0 \times I_1$
TM-MODE WAVEGUIDE	2-PORT	TM	CUTOFF FREQUENCY FC IN HERTZ INFINITE-FREQUENCY CHARACTERISTIC IMPEDANCE ZINF IN OHMS LENGTH LEN IN METERS	ZINF#0 ~FC#F	$Z_0 = ZINF \times (1 - (FC \div F) \star 2) \star .5$ $A = J \times LEN \times Z_0 \times F \div ZINF \times 3E8 \div DIELEL \star .5$ $(V_2 - Z_0 \times I_2) = (\star - O2 \times A) \times V_1 + Z_0 \times I_1$ $(V_1 - Z_0 \times I_1) = (\star - O2 \times A) \times V_2 + Z_0 \times I_2$
TM-MODE WAVEGUIDE CHAR. IMPEDANCE	1-PORT	TM	CUTOFF FREQUENCY FC IN HERTZ INF.-FREQ. CHAR. IMP. ZINF IN OHMS	ZINF#0 ~FC#F	$V = I \times ZINF \times (1 - (FC \div F) \star 2) \star .5$
NEGATIVE-IMPEDANCE CONVERTER	2-PORT	VNIC	(NONE)		$V_1 = -V_2; I_1 = -I_2$
NEGATIVE-IMPEDANCE CONVERTER	2-PORT	INIC	(NONE)		$V_1 = V_2; I_1 = I_2$
CURRENT-CONTROLLED CURRENT SOURCE	2-PORT	CCCS	CURRENT GAIN AC	AC#0	$V_1 = 0; I_2 = -AC \times I_1$
CURRENT-CONTROLLED FLUX SOURCE	2-PORT	CCFS	TRANSINDUCTANCE LM IN HENRIES	LM#0	$V_1 = 0; V_2 = S \times LM \times I_1$
CURRENT-CONTROLLED CHARGE SOURCE	2-PORT	CCQS	TRANSFER TIME TM IN SECONDS	TM#0	$V_1 = 0; I_2 = -S \times TM \times I_1$
CURRENT-CONTROLLED VOLTAGE SOURCE	2-PORT	CCVS	TRANSRESISTANCE RM IN OHMS	RM#0	$V_1 = 0; V_2 = RM \times I_1$
FLUX-CONTROLLED CURRENT SOURCE	2-PORT	FCCS	TRANSFER INVERSE INDUCTANCE HM IN INVERSE HENRIES	HM#0	$I_1 = 0; I_2 = -HM \times V_1 \div S$
FLUX-CONTROLLED FLUX SOURCE	2-PORT	FCFS	FLUX GAIN AF	AF#0	$I_1 = 0; V_2 = AF \times V_1$
FLUX-CONTROLLED CHARGE SOURCE	2-PORT	FCQS	TRANSCONDUCTANCE GM IN MHOS	GM#0	$I_1 = 0; I_2 = -GM \times V_1$
FLUX-CONTROLLED VOLTAGE SOURCE	2-PORT	FCVS	TRANSFER INVERSE TIME FM IN INVERSE SECONDS	FM#0	$I_1 = 0; V_2 = FM \times V_1 \div S$
CHARGE-CONTROLLED CURRENT SOURCE	2-PORT	QCCS	TRANSFER INVERSE TIME FM IN INVERSE SECONDS	FM#0	$V_1 = 0; I_2 = -FM \times I_1 \div S$
CHARGE-CONTROLLED FLUX SOURCE	2-PORT	QCFS	TRANSRESISTANCE RM IN OHMS	RM#0	$V_1 = 0; V_2 = RM \times I_1$
CHARGE-CONTROLLED CHARGE SOURCE	2-PORT	QCQS	CHARGE GAIN AQ	AQ#0	$V_1 = 0; I_2 = -AQ \times I_1$
CHARGE-CONTROLLED VOLTAGE SOURCE	2-PORT	QCVS	TRANSFER INVERSE CAPACITANCE SM IN DARAFS	SM#0	$V_1 = 0; V_2 = SM \times I_1 \div S$
VOLTAGE-CONTROLLED CURRENT SOURCE	2-PORT	VCCS	TRANSCONDUCTANCE GM IN MHOS	GM#0	$I_1 = 0; I_2 = -GM \times V_1$

VOLTAGE-CONTROLLED FLUX SOURCE	2-PORT	VCFS	TRANSFER TIME TM IN SECONDS	TM#0	I1=0; V2=S*TM*V1
VOLTAGE-CONTROLLED CHARGE SOURCE	2-PORT	VCQS	TRANSCAPACITANCE CM IN FARADS	CM#0	I1=0; I2=-S*CM*V1
VOLTAGE-CONTROLLED VOLTAGE SOURCE	2-PORT	VCVS	VOLTAGE GAIN AV	AV#0	I1=0; V2=AV*V1
NULLOR	2-PORT	NULLOR	(NONE)		V1=0; I1=0
RESISTOR TEE	2-PORT	*RTEE	FIRST RESISTANCE RIN IN OHMS SECOND RESISTANCE RMID IN OHMS THIRD RESISTANCE ROUT IN OHMS	RMID#0	V1=(RIN*I1)+RMID*I1+I2 V2=(ROUT*I2)+RMID*I1+I2
RESISTOR PI	2-PORT	*RPI	FIRST RESISTANCE RIN IN OHMS SECOND RESISTANCE RMID IN OHMS THIRD RESISTANCE ROUT IN OHMS	RIN#0 ROUT#0	I1=(V1+RIN)+(V1-V2)÷RMID I2=(V2÷ROUT)+(V2-V1)÷RMID
INDUCTOR TEE	2-PORT	*LTEE	FIRST INDUCTANCE LIN IN HENRIES SECOND INDUCTANCE LMID IN HENRIES THIRD INDUCTANCE LOUT IN HENRIES	LIN#0 LMID#0 LOUT#0	V1=S*(LIN*I1)+LMID*I1+I2 V2=S*(LOUT*I2)+LMID*I1+I2
INDUCTOR PI	2-PORT	*LPI	FIRST INDUCTANCE LIN IN HENRIES SECOND INDUCTANCE LMID IN HENRIES THIRD INDUCTANCE LOUT IN HENRIES	LIN#0 LMID#0 LOUT#0	I1=((V1÷LIN)+(V1-V2)÷LMID)÷S I2=((V2÷LOUT)+(V2-V1)÷LMID)÷S
CAPACITOR TEE	2-PORT	*CTEE	FIRST CAPACITANCE CIN IN FARADS SECOND CAPACITANCE CMID IN FARADS THIRD CAPACITANCE COUT IN FARADS		V1=((I1÷CIN)+(I1+I2)÷CMID)÷S V2=((I2÷COUT)+(I1+I2)÷CMID)÷S
CAPACITOR PI	2-PORT	*CPI	FIRST CAPACITANCE CIN IN FARADS SECOND CAPACITANCE CMID IN FARADS THIRD CAPACITANCE COUT IN FARADS		I1=S*(V1÷CIN)+(V1-V2)×CMID I2=S*(V2÷COUT)+(V2-V1)×CMID
R-ROTATOR	2-PORT	*RROTATOR	ANGLE T IN DEGREES RESISTANCE RES IN OHMS	RES#0	V1=(V2×2OT)+RES×I2×1OT I1=(V2÷RES÷1OT)-I2×2OT
L-ROTATOR	2-PORT	*LROTATOR	ANGLE T IN DEGREES INDUCTANCE IND IN HENRIES	IND#0	V1=(V2×2OT)+S×IND×I2×1OT I1=(V2÷S×IND÷1OT)-I2×2OT
C-ROTATOR	2-PORT	*CROTATOR	ANGLE T IN DEGREES CAPACITANCE CAP IN FARADS	CAP#0	V1=(V2×2OT)-I2÷S×CAP÷1OT I1=(-V2×S×CAP÷1OT)-I2×2OT
R-REFLECTOR	2-PORT	*RREFLECTOR	ANGLE T IN DEGREES RESISTANCE RES IN OHMS	RES#0	V1=(V2×2O2×T)-RES×I2×1O2×T I1=(V2÷RES÷1O2×T)+I2×2O2×T
L-REFLECTOR	2-PORT	*LREFLECTOR	ANGLE T IN DEGREES INDUCTANCE IND IN HENRIES	IND#0	V1=(V2×2O2×T)-S×IND×I2×1O2×T I1=(V2÷S×IND÷1O2×T)+I2×2O2×T
C-REFLECTOR	2-PORT	*CREFLECTOR	ANGLE T IN DEGREES CAPACITANCE CAP IN FARADS	CAP#0	V1=(-V2×2O2×T)-I2÷S×CAP÷1O2×T I1=(V2×S×CAP÷1O2×T)-I2×2O2×T
I-SCALOR	2-PORT	*ISCALOR	CURRENT SCALE FACTOR A		V1=V2; I1=-A×I2
V-SCALOR	2-PORT	*VSCALOR	VOLTAGE SCALE FACTOR A		V1=A×V2; I1=-I2
P-SCALOR	2-PORT	*PSCALOR	VOLTAGE SCALE FACTOR CURRENT SCALE FACTOR AV AI		V1=AV×V2; I1=-AI×I2
USER-DEFINED ELEMENT	2-PORT	UDE	VECTOR TO BE ARGUMENT FOR NEWELEMENT		
POWER-DEFINED ELEMENT	1-PORT	ZPDE YPDE	INTEGER N COEFFICIENT	5≥ N	SEE HOWPDE IN 100 MARTHAE
POWER-DEFINED ELEMENT	2-PORT	ZPDE YPDE HPDE ABCDPDE	INTEGER N 2, 3, OR 4 COEFFICIENTS	5≥ N	SEE HOWPDE IN 100 MARTHAE
NUMERICALLY DEFINED ELEMENT	1-PORT	SFOF YFOF ZFOF	FOF WITH 1 OR 2 COLUMNS (2 FOR SFOF)		SEE HOWFOF IN 100 MARTHAE
NUMERICALLY DEFINED ELEMENT	2-PORT	ABCDFOF HFOF SFOF YFOF ZFOF	FOF WITH 3, 4, 6, OR 8 COLUMNS (4, 6, OR 8 FOR SFOF)		SEE HOWFOF IN 100 MARTHAE

\* OBSOLETE. WILL PROBABLY BE OMITTED FROM MARTHA IN THE FUTURE.

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WIRING FUNCTIONS IN MARTHA:



IF A 2-PORT NETWORK IS EXPECTED AND A 1-PORT NETWORK APPEARS, WP IS AUTOMATICALLY INVOKED TO CHANGE IT INTO A 2-PORT NETWORK. IF A 1-PORT NETWORK IS EXPECTED AND A 2-PORT NETWORK APPEARS, WTO IS AUTOMATICALLY INVOKED TO CHANGE IT INTO A 1-PORT NETWORK.

N ZSCALE S1 IS A NETWORK OF THE SAME TOPOLOGY AS S1 BUT WITH EACH ELEMENT SCALED IN IMPEDANCE BY THE FACTOR N, WHICH MAY BE POSITIVE OR NEGATIVE. S1 IS A 1-PORT OR 2-PORT NETWORK.

WIRING FUNCTIONS IN 100 MARTHAW:

- WROTWROT S1 IS THE SAME AS WROT WROT S1 BUT TWICE AS FAST.
- WTOWN S1 IS THE SAME AS WTO WN S1 BUT MUCH FASTER. IT WORKS FOR UNILATERAL NETWORKS, WHERE WN S1 CANNOT BE CALCULATED.
- WTSWN S1 IS THE SAME AS WTS WN S1 BUT MUCH FASTER. IT WORKS FOR UNILATERAL NETWORKS, WHERE WN S1 CANNOT BE CALCULATED.
- RES WDUAL S1 IS A NETWORK WHICH IS THE DUAL OF S1. EACH ELEMENT IS REPLACED BY ITS DUAL, AND ALL WIRING FUNCTIONS BY THEIR DUALS. RES IS THE CHARACTERISTIC RESISTANCE OF THE OPERATION. S1 IS A 1-PORT OR 2-PORT NETWORK.
- WAD S1 IS A NETWORK WHICH IS THE ADJOINT OF S1. EACH ELEMENT IS REPLACED BY ITS ADJOINT. THE OPERATION HAS THE EFFECT OF TRANSPOSING THE Z OR Y MATRIX. IT HAS NO EFFECT ON RECIPROCAL NETWORKS. S1 IS A 1-PORT OR 2-PORT NETWORK.
- WCC S1 IS A NETWORK WITH EACH ELEMENT REPLACED BY ITS 'COMPLEX CONJUGATE.' THIS HAS THE EFFECT OF CONJUGATING THE Z OR Y MATRIX. S1 IS A 1-PORT OR 2-PORT NETWORK.
- ★WTM S1 IS A 1-PORT NETWORK WHOSE IMPEDANCE IS THE FORWARD IMAGE IMPEDANCE Z11 OF THE 2-PORT NETWORK S1.
- FL FLIMITS S1 IS THE SAME NETWORK S1 TAGGED WITH THE FREQUENCY LIMITS FL. FREQUENCIES OUT OF RANGE WILL BE REPORTED DURING ANALYSIS. THE TAG DOES NOT AFFECT ANY COMPUTATIONS. THE FREQUENCY LIST FL IS A VECTOR WITH AN EVEN NUMBER OF FREQUENCIES, THE SMALLEST 2 GIVING AN ALLOWED RANGE, THE NEXT SMALLEST 2 ANOTHER ALLOWED RANGE, ETC. S1 IS A 1-PORT OR 2-PORT NETWORK.
- N FSCALE S1 IS A NETWORK OF THE SAME TOPOLOGY AS S1 BUT WITH EACH ELEMENT SCALED BY THE FACTOR N, WHICH MAY BE POSITIVE OR NEGATIVE. THE NEW NETWORK HAS THE SAME PROPERTIES AT

FREQUENCY  $N \times F$  AS THE OLD ONE AT  $F$ .  $S1$  IS A 1-PORT OR 2-PORT NETWORK NOT CONTAINING A USER-DEFINED ELEMENT.  
 FP FINVERT  $S1$  IS A NETWORK OF THE SAME TOPOLOGY AS  $S1$  WITH CAPACITORS REPLACED BY INDUCTORS AND VICE VERSA. THE NEW NETWORK HAS THE SAME PROPERTIES AT FREQUENCY  $FP \times FP \div F$  AS (WCC  $S1$ ) AT  $F$ . THAT IS, FINVERT PERFORMS A LOWPASS-TO-HIGHPASS TRANSFORMATION.  $S1$  IS A 1-PORT OR 2-PORT NETWORK NOT CONTAINING TEM, WG, OR UDE.  
 FM FBP  $S1$  IS A NETWORK SIMILAR TO  $S1$  WITH CAPACITORS REPLACED BY PARALLEL L-C, AND INDUCTORS BY SERIES L-C, RESONANT AT FREQUENCY FM. THE NEW NETWORK HAS THE SAME PROPERTIES AT FREQUENCY  $(F + ((F \times F) + 4 \times FM \times FM) \star 0.5) \div 2$  AS THE OLD ONE AT  $F$ . THAT IS, FBP PERFORMS A LOWPASS-TO-BANDPASS TRANSFORMATION.  $S1$  IS A 1-PORT OR 2-PORT NETWORK NOT CONTAINING TEM, WG, OR UDE.

★ OBSOLETE. WILL PROBABLY BE OMITTED FROM MARTHA IN THE FUTURE. USE THE RESPONSE FUNCTION Z11 INSTEAD.

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RESPONSE FUNCTIONS IN MARTHA; 26 COMPLEX, 4 REAL:

NAME	C/R	MEANING	DEPENDS ON
Z	C	IMPEDANCE OF A 1-PORT NETWORK $V=Z \times I$	NETWORK
Y	C	ADMITTANCE OF A 1-PORT NETWORK $I=Y \times V$	NETWORK
SC	C	REFLECTION COEFFICIENT OF 1-PORT $B=SC \times A$	NETWORK, ZN
Z11	C	IMPEDANCE MATRIX	NETWORK
Z12	C	$V1=(Z11 \times I1) + (Z12 \times I2)$	
Z21	C	$V2=(Z21 \times I1) + (Z22 \times I2)$	
Z22	C		
Y11	C	ADMITTANCE MATRIX	NETWORK
Y12	C	$I1=(Y11 \times V1) + (Y12 \times V2)$	
Y21	C	$I2=(Y21 \times V1) + (Y22 \times V2)$	
Y22	C		
H11	C	HYBRID MATRIX	NETWORK
H12	C	$V1=(H11 \times I1) + (H12 \times V2)$	
H21	C	$I2=(H21 \times I1) + (H22 \times V2)$	
H22	C		
S11	C	SCATTERING MATRIX	NETWORK, ZNIN, ZNOUT
S12	C	$B1=(S11 \times A1) + (S12 \times A2)$	
S21	C	$B2=(S21 \times A1) + (S22 \times A2)$	
S22	C		
ZIN	C	INPUT IMPEDANCE $V1 \div I1$	NETWORK, ZL
YIN	C	INPUT ADMITTANCE $I1 \div V1$	NETWORK, ZL
SIN	C	INPUT REFLECTION COEFFICIENT $B1 \div A1$	NETWORK, ZL, ZNIN
ZOUT	C	OUTPUT IMPEDANCE $V2 \div I2$ WHEN OUTPUT EXCITED	NETWORK, ZG
YOUT	C	OUTPUT ADMITTANCE $I2 \div V2$ WHEN OUTPUT EXCITED	NETWORK, ZG
SOUT	C	OUTPUT REFLECTION COEFFICIENT $B2 \div A2$ WHEN OUTPUT EXCITED	NETWORK, ZG, ZNOUT
VG	C	VOLTAGE GAIN $V2 \div EG$	NETWORK, ZG, ZL
AG	R	AVAILABLE GAIN $POUT, AV \div PIN, AV$	NETWORK, ZG
IG	R	INSERTION GAIN $POUT(NETWORK) \div POUT(WTHRU)$	NETWORK, ZG, ZL
PG	R	POWER GAIN $POUT \div PIN$	NETWORK, ZL
TG	R	TRANSDUCER GAIN $POUT \div PIN, AV$	NETWORK, ZG, ZL

WAVE-VARIABLE DEFINITIONS 2-PORT NETWORK WITH GENERATOR AND LOAD  
 $A = (V + I \times ZN) \div 2 \times (|RE ZN) \star .5$        $- + ZG + I1 \rightarrow$        $\leftarrow I2 + ZL$   
 $B = (V - I \times CC ZN) \div 2 \times (|RE ZN) \star .5$        $--(\sim)--/\backslash \backslash -o-----|$        $-----o-/\backslash \backslash -$   
 $A1 = (V1 + I1 \times ZNIN) \div 2 \times (|RE ZNIN) \star .5$        $| \quad \leftarrow EG \quad V1 \quad |$        $| NETWORK |$        $V2 \quad |$   
 $B1 = (V1 - I1 \times CC ZNIN) \div 2 \times (|RE ZNIN) \star .5$        $-----o-----|$        $-----o-----$   
 $A2 = (V2 + I2 \times ZNOUT) \div 2 \times (|RE ZNOUT) \star .5$        $-$        $-$   
 $B2 = (V2 - I2 \times CC ZNOUT) \div 2 \times (|RE ZNOUT) \star .5$

VARIABLES EG, ZG, ZL, ZN, ZNIN, AND ZNOUT CAN BE ANY OF:  
 (1) A SINGLE CONSTANT (REAL FREQUENCY-INDEPENDENT IMPEDANCE OR VOLTAGE)  
 (2) A 1-COLUMN FOF (REAL FREQUENCY-DEPENDENT IMPEDANCE OR VOLTAGE)  
 (3) A 2-COLUMN FOF (COMPLEX FREQUENCY-DEPENDENT IMPEDANCE OR VOLTAGE)  
 (4) A 1-PORT MARTHA NETWORK, OR IF A 2-PORT NETWORK WTO ASSUMED. THE IMPEDANCE IS USED FOR ZG, ZL, ZN, ZNIN, OR ZNOUT, AND THE VOLTAGE PRODUCED BY A UNIT CURRENT IS USED FOR EG.

RESPONSE FUNCTIONS IN 100 MARTHAR; 61 COMPLEX, 12 REAL:

NAME	C/R	MEANING	DEPENDS ON
G11	C	HYBRID MATRIX	NETWORK
G12	C	$I1=(G11 \times V1) + (G12 \times I2)$	
G21	C	$V2=(G21 \times V1) + (G22 \times I2)$	
G22	C		
ABCD	C	ABCD TRANSMISSION MATRIX	NETWORK
$\bar{A}BCD$	C	$V1=(ABCD \times V2) - (ABCD \times I2)$	
$\bar{A}BCD$	C	$I1=(\bar{A}BCD \times V2) - (\bar{A}BCD \times I2)$	
$\bar{A}BCD$	C		
RVV	C	REVERSE TRANSMISSION MATRIX	NETWORK
RVI	C	$V2=(RVV \times V1) - (RVI \times I1)$	
RIV	C	$I2=(RIV \times V1) - (RII \times I1)$	
RII	C		
K11	C	INVERSE SCATTERING MATRIX	NETWORK, ZNIN, ZNOUT
K12	C	$A1=(K11 \times B1) + (K12 \times B2)$	
K21	C	$A2=(K21 \times B1) + (K22 \times B2)$	
K22	C		
U11	C	HYBRID SCATTERING MATRIX	NETWORK, ZNIN, ZNOUT
U12	C	$B1=(U11 \times A1) + (U12 \times B2)$	
U21	C	$A2=(U21 \times A1) + (U22 \times B2)$	
U22	C		
W11	C	HYBRID SCATTERING MATRIX	NETWORK, ZNIN, ZNOUT
W12	C	$A1=(W11 \times B1) + (W12 \times A2)$	
W21	C	$B2=(W21 \times B1) + (W22 \times A2)$	
W22	C		
TAA	C	SCATTERING TRANSMISSION MATRIX	NETWORK, ZNIN, ZNOUT
TAB	C	$A1=(TAA \times A2) + (TAB \times B2)$	
TBA	C	$B1=(TBA \times A2) + (TBB \times B2)$	
TBB	C		
RAA	C	REVERSE SCATTERING TRANSMISSION MATRIX	NETWORK, ZNIN, ZNOUT
RAB	C		
RBA	C	$A2=(RAA \times A1) + (RAB \times B1)$	
RBB	C	$B2=(RBA \times A1) + (RBB \times B1)$	
ZI1	C	IMAGE GENERATOR IMPEDANCE	NETWORK
ZI2	C	IMAGE LOAD IMPEDANCE	NETWORK
ITC	C	IMAGE TRANSFER CONSTANT	NETWORK
ZK1	C	ITERATIVE GENERATOR IMPEDANCE	NETWORK
ZK2	C	ITERATIVE LOAD IMPEDANCE	NETWORK
KTC	C	ITERATIVE TRANSFER CONSTANT	NETWORK
ZM1	C	CONJ.-MATCH GENERATOR IMPEDANCE	NETWORK
ZM2	C	CONJUGATE-MATCH LOAD IMPEDANCE	NETWORK
*OCVG	C	OPEN-CIRCUIT VOLTAGE GAIN $V2 \div V1$ WHEN $I2=0$	NETWORK
VR	C	VOLTAGE RATIO $V2 \div V1$	NETWORK, ZL
IVG	C	INSERTION VOLTAGE GAIN $V2$ (NETWORK) $\div V2$ (WTHRU)	NETWORK, ZG, ZL
CG	C	CURRENT GAIN $-I2 \times ZG \div EG$	NETWORK, ZG, ZL
*SCCG	C	SHORT-CIRCUIT CURRENT GAIN $-I2 \div I1$ WHEN $V2=0$	NETWORK
CR	C	CURRENT RATIO $-I2 \div I1$	NETWORK, ZL
RR	C	RETURN RATIO $-Y12 \times Y21 \div (Y11 + \div ZG) \times (Y22 + \div ZL)$	NETWORK, ZG, ZL
MG	R	CONJ.-MATCH (MAX AVAIL) GAIN	NETWORK
UG	R	MASON'S UNILATERAL GAIN	NETWORK
SGP	C	SPENCE'S GAIN PLANE $UG \times Z12 \div Z21$	NETWORK
LM	C	LINVILL'S L-M PLANE	NETWORK, ZL
ISF	R	ROLLETT'S INVARIANT STABILITY FACTOR	NETWORK
SSF	R	STERN'S STABILITY FACTOR	NETWORK, ZG, ZL
RF	C	RECIPROCITY FACTOR $Z21 \div Z12$	NETWORK
V1	C	INPUT VOLTAGE	NETWORK, ZG, ZL, EG
V2	C	OUTPUT VOLTAGE	NETWORK, ZG, ZL, EG
I1	C	INPUT CURRENT	NETWORK, ZG, ZL, EG
I2	C	OUTPUT CURRENT	NETWORK, ZG, ZL, EG
A1	C	INPUT INCOMING WAVE	NETWORK, ZG, ZL, EG, ZNIN
A2	C	OUTPUT INCOMING WAVE	NETWORK, ZG, ZL, EG, ZNOUT
B1	C	INPUT OUTGOING WAVE	NETWORK, ZG, ZL, EG, ZNIN
B2	C	OUTPUT OUTGOING WAVE	NETWORK, ZG, ZL, EG, ZNOUT
CP1	C	COMPLEX INPUT POWER $V1 \times CC$ $I1$	NETWORK, ZG, ZL, EG
CP2	C	COMPLEX OUTPUT POWER $V2 \times CC$ $I2$	NETWORK, ZG, ZL, EG
CP	C	TOTAL COMPLEX POWER $CP1 + CP2$	NETWORK, ZG, ZL, EG
AL	R	AVAILABLE LOSS $\div AG$	NETWORK, ZG
IL	R	INSERTION LOSS $\div IG$	NETWORK, ZG, ZL
ML	R	INTRINSIC ATTENUATION $\div MG$	NETWORK
PL	R	POWER LOSS $\div PG$	NETWORK, ZL
TL	R	TRANSDUCER LOSS $\div TG$	NETWORK, ZG, ZL
VSWR	R	VOLTAGE STANDING-WAVE RATIO OF 1-PORT $ 1 +  SC  \div (1 -  SC) $	NETWORK, ZN
VSWRIN	R	VOLTAGE STANDING-WAVE RATIO AT INPUT $ 1 +  SIN  \div (1 -  SIN) $	NETWORK, ZL, ZNIN
VSWROUT	R	VOLTAGE STANDING-WAVE RATIO AT OUTPUT $ 1 +  SOUT  \div (1 -  SOUT) $	NETWORK, ZG, ZNOUT

OMEGA	R	FREQUENCY IN RADIANS/SECOND	F
OUTVAR	C	VARIABLE EG, ZG, ZL, ZN, ZNIN, OR ZNOUT	VARIABLE
OUTFOF	R,C	ALL COLUMNS OF A FOF, USING INTERPOLATION OR EXTRAPOLATION 2-COLUMN FOF COMPLEX; OTHERS REAL	FOF, F

★ OBSOLETE. WILL PROBABLY BE OMITTED FROM MARTHA IN THE FUTURE. USE G21 FOR OCVG AND -H21 FOR SCCG.

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MODIFIERS IN MARTHA:

NAME	MEANING	COMMENTS
RE	REAL PART	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
IM	IMAGINARY PART	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
MAG	MAGNITUDE	FOR COMPLEX RESPONSES, THE MAGNITUDE. FOR REAL RESPONSES, THE ABSOLUTE VALUE.
RAD	PHASE ANGLE IN RADIANS	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
DEG	PHASE ANGLE IN DEGREES	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
DB	MAGNITUDE IN DECIBELS	FOR COMPLEX RESPONSES, $20 \times 10 \times \text{MAGNITUDE}$ . FOR REAL RESPONSES, $10 \times 10 \times \text{ABSOLUTE VALUE}$ .

IF NO MODIFIER IS USED, THE REAL AND IMAGINARY PARTS OF COMPLEX RESPONSES WILL RESULT.

MODIFIERS IN 100 MARTHAM:

NAME	MEANING	COMMENTS
PD	PHASE DELAY (MOD. PERIOD)	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
REC	RECIPROCAL	FOR REAL OR COMPLEX RESPONSES.
MAGRAD	MAGNITUDE AND PHASE (RADS.)	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
MAGDEG	MAGNITUDE AND PHASE (DEGS.)	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
DBRAD	MAG. (DB) AND PHASE (RADS.)	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
DBDEG	MAG. (DB) AND PHASE (DEGS.)	FOR COMPLEX RESPONSES. WILL BE IGNORED IF APPLIED TO A REAL RESPONSE.
SWR	STANDING-WAVE RATIO	FOR COMPLEX RESPONSES (REFLECTION COEFFICIENTS ONLY). PRODUCES THE VOLTAGE STANDING-WAVE RATIO; E.G. SWR S11 PRODUCES $(1+ S11 ) \div (1- S11 )$ .
NORM	NORMALIZED RESPONSE	FOR REAL OR COMPLEX RESPONSES. PRECEDE BY ZN, ZNIN, OR ZNOUT. RESULT IS THE RESPONSE FUNCTION DIVIDED BY ZN, ZNIN, OR ZNOUT.
GAMMA	REFLECTION COEFFICIENT	FOR COMPLEX RESPONSES (IMPEDANCES ONLY). PRECEDE BY ZN, ZNIN, OR ZNOUT. ACTS ON ANY IMPEDANCE TO PRODUCE A REFLECTION COEFFICIENT; E.G. ZNIN GAMMA ZI1 PRODUCES $(ZI1-CC ZNIN) \div (ZI1+ZNIN)$ .

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GENERAL FORM OF AN OUTPUT REQUEST IN MARTHA:

```
(PRINT)
(PLOT )
(PLOG ) <OUTPUT LIST> OF <NETWORK>
(SMITH)
(STORE)
```

IF THE FIRST WORD IS OMITTED, A PRINT WITHOUT THE HEADING IS OBTAINED, WITH THE FREQUENCY IN THE LAST COLUMN RATHER THAN THE FIRST.

TO DISPLAY A FOF, USE ITS NAME IN PLACE OF '<OUTPUT LIST> OF <NETWORK>' IN THE GENERAL FORM ABOVE; ALL FORMAT FUNCTIONS EXCEPT VS CAN BE USED.

FORMAT FUNCTIONS IN MARTHA:

NAME	PURPOSE
PRINT	PRINTS RESPONSE FUNCTIONS. WITH ONLY ONE FREQUENCY, HEADING IS OMITTED. 'PRINT 10' PRODUCES ONLY THE ONE-LINE HEADING.
PLOT	PLOTS RESPONSE FUNCTIONS VS FREQUENCY, AUTOMATICALLY CHOOSING SCALES.
PLOG	PLOTS RESPONSE FUNCTIONS VS FREQUENCY ON A LOGARITHMIC SCALE, AUTOMATICALLY CHOOSING SCALES.
TITLE	OPTIONAL. SET THE VARIABLE TITLE TO ANYTHING TO BE PRINTED AT THE TOP OF THE NEXT OUTPUT. (NORMALLY BLANK)
SS	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST. FORCES ALL DEPENDENT VARIABLES TO BE PLOTTED ON THE SAME SCALE. IGNORED IN PRINTED OUTPUT. (NORMALLY NOT ON SAME SCALE)
SYMBOLS	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY CHARACTERS TO BE USED AS PLOT CHARACTERS IN THE NEXT OUTPUT. IGNORED IN PRINTED OUTPUT. (NORMALLY 'OX=0+#*l÷v-□')
WIDE	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY NUMBER. NEXT OUTPUT WILL BE THAT NUMBER OF SPACES WIDE. IGNORED IN PRINTED OUTPUT. (NORMALLY 50 WIDE)
HIGH	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY A NUMBER. NEXT OUTPUT WILL BE THAT NUMBER OF LINES HIGH. IGNORED IN PRINTED OUTPUT. (NORMALLY 50 HIGH)
VS	OPTIONAL. PLACE IN OUTPUT LIST BEFORE DESIRED INDEPENDENT VARIABLE. NEXT OUTPUT HAS ALL OTHER RESPONSE FUNCTIONS AS DEPENDENT VARIABLES. IF USED WITH PLOT, ALL PLOTTED AGAINST INDEPENDENT VARIABLE ON A LINEAR SCALE. IF USED WITH PLOG, ALL PLOTTED AGAINST INDEPENDENT VARIABLE ON A LOGARITHMIC SCALE. IF USED WITH PRINT, INDEPENDENT VARIABLE APPEARS IN THE FIRST COLUMN. (NORMALLY VS F)
OF	REQUIRED. PLACE AT END OF OUTPUT LIST, BEFORE NETWORK DESCRIPTION.

FORMAT FUNCTIONS IN 100 MARTHAF:

NAME	PURPOSE
SMITH	PLOTS A SMITH CHART. USE INSTEAD OF PRINT, PLOT, ETC. NORMALLY STANDARD SIZE (UNLESS WIDE OR HIGH IS USED) AND SCALES FROM 1 TO 1 (UNLESS EXPAND, HSCALE, OR VSCALE IS USED). BACKGROUND IS THE DOMINANT CIRCLES FROM THE SMITH CHART, INCLUDING THE UNIT CIRCLE.
PLACES	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY A NUMBER. NEXT OUTPUT WILL BE PRINTED TO THAT PLACE ACCURACY. (NORMALLY 5 PLACES)
HSCALE	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY TWO NUMBERS. NEXT OUTPUT WILL USE THOSE LIMITS ON THE HORIZONTAL SCALE. POINTS FOR DEPENDENT VARIABLES OUTSIDE THE RANGE ARE NOT PLOTTED. IGNORED IN PRINTED OUTPUT. (NORMALLY AUTOMATIC SCALE SELECTION)
VSCALE	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST, PRECEDED BY TWO NUMBERS. NEXT OUTPUT WILL USE THOSE LIMITS ON THE VERTICAL SCALE. POINTS FOR INDEPENDENT VARIABLES OUTSIDE THE RANGE ARE NOT PLOTTED. IGNORED IN PRINTED OUTPUT. (NORMALLY AUTOMATIC SCALE SELECTION)
PAIRS	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST. NEXT OUTPUT WILL PLOT FIRST VARIABLE VS SECOND, THIRD VS FOURTH, ETC., ALL ON THE SAME SCALES (COMPLEX VARIABLES COUNT AS TWO). FOR PRINTED OUTPUT, EQUIVALENT TO 'VS LAST VARIABLE'. (NORMALLY ONLY ONE INDEPENDENT VARIABLE)
EXPAND	OPTIONAL. PLACE ANYWHERE IN THE OUTPUT LIST. WHEN USED WITH SMITH, SETS SCALES FOR STANDARD-SIZE EXPANDED SMITH CHART. EQUIVALENT TO 0.226 0.226 HSCALE 0.226 0.226 VSCALE.



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FOF FUNCTIONS IN 100 MARTHA:

1. 'SCALAR' FOF FUNCTIONS  
SYNTAX FOR COMPLEX ARGUMENTS FOR REAL ARGUMENTS

Z-FRE A	REAL PART OF A	A
Z-FIM A	IMAGINARY PART OF A	0
Z-FMAG A	MAGNITUDE OF A	A
Z-FRAD A	PHASE OF A IN RADIANS	0A<0
Z-FDEG A	PHASE OF A IN DEGREES	180×A<0
Z-FCC A	CONJUGATE OF A	A
Z-FSIG A	SIGNUM OF A	×A
Z-FEXP A	E TO THE POWER A	★A
Z-FLN A	NATURAL LOG OF A	⊙A
Z-A FADD B	SUM	A+B
Z-A FSUB B	DIFFERENCE	A-B
Z-A FMUL B	PRODUCT	A×B
Z-A FDIV B	RATIO	A÷B
Z-A FPWR B	A TO THE POWER B	A★B
Z-A FLOG B	LOG OF B TO BASE A	A⊙B
Z-A FEQ B	EQUALITY TEST OF A, B	A=B
Z-A FNEQ B	INEQUALITY TEST	A≠B

2. 'MIXED' FOF FUNCTIONS  
SYNTAX RESULT

Z-A FCAT B	CATENATION OF ALL COLUMNS OF A AND B; LIKE 'A,B'
Z-FCOL A	NUMBER OF COLUMNS OF A; LIKE 'oA'
Z-FDF A	1-COLUMN FOF CONTAINING THE DEFINING FREQUENCIES OF A

FOR THE 'SCALAR' FUNCTIONS ABOVE, ARGUMENT A (OR B) CAN BE A 1-COLUMN FOF (TREATED AS REAL) OR A 2-COLUMN FOF (TREATED AS COMPLEX, WITH THE FIRST COLUMN THE REAL PART, AND THE SECOND COLUMN THE IMAGINARY PART). FOR THE 'MIXED' FUNCTIONS ABOVE, THE ARGUMENTS CAN BE FOF'S WITH ANY NUMBER OF COLUMNS. A SINGLE NUMBER IS TREATED AS A 1-COLUMN FOF, AND A VECTOR OF LENGTH 2 AS A 2-COLUMN FOF, EACH WITH ONE DEFINING FREQUENCY EQUAL TO 0 HZ.

FOR THE 'SCALAR' FUNCTIONS ABOVE, THE RESULT IS A REAL OR COMPLEX (REAL IF POSSIBLE) FOF WITH THE SAME DEFINING FREQUENCIES AS THE ARGUMENT. FOR DYADIC FUNCTIONS, IF THE ARGUMENTS HAVE DIFFERENT DEFINING FREQUENCIES, THE ONE WITH MORE DEFINING FREQUENCIES IS USED (IF THE ARGUMENTS HAVE THE SAME NUMBER OF DEFINING FREQUENCIES, THOSE OF THE LEFT ARGUMENT ARE USED UNLESS THE LEFT ARGUMENT HAS ONE DEFINING FREQUENCY EQUAL TO 0 HZ). ALL COLUMNS OF THE OTHER FOF ARE INTERPOLATED OR EXTRAPOLATED AS NECESSARY, TO MATCH THE DEFINING FREQUENCIES OF THE RESULT.

EACH OF THE FUNCTIONS ABOVE ACTS INDEPENDENT OF THE OTHERS, BUT EACH USES THE BACKGROUND FUNCTION FOF. IF INDIVIDUAL FUNCTIONS ARE COPIED TO THE ACTIVE WORKSPACE, THE FUNCTION FOF MUST NOT BE FORGOTTEN. THE GROUP 'FOF' CAN BE COPIED TOGETHER; IT CONTAINS COLUMNSOF, FCOL, FCAT, FADD, FSUB, FMUL, FDIV, FPWR, AND FOF.

3. MISCELLANEOUS OTHER FUNCTIONS. REFERENCE IS TO SECTION IN THE MARTHA USER'S MANUAL.

NAME	LEFT ARG	RIGHT ARG	RESULT	PURPOSE	REF
MAKEFOF		NUMBER	FOF	INTERACTIVE FOF-CREATOR	3.9
MAKEFOF		FOF	FOF	INTERACTIVE FOF-EDITOR	3.9
INTERPOLATE	FREQ VECTOR	FOF	FOF	INTERPOLATES TO NEW DEFINING FREQUENCIES	
COLUMNSOF	VECTOR OF COL. NOS.	FOF	FOF	SELECTS COLUMNS OF FOF	3.9
FROMMAGDEG		FOF	FOF	CHANGES COLUMNS IN PAIRS FROM MAG,DEG TO RE,IM	
FROMDBDEG		FOF	FOF	CHANGES COLUMNS IN PAIRS FROM DB,DEG TO RE,IM	
TOFOF	FREQ VECTOR	MATRIX	FOF	URNS MATRIX INTO FOF	
FROMFOF		FOF	MATRIX	MATRIX TURNS FOF INTO MATRIX	

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CONTENTS OF 100 MARTHAX. REFERENCE IS TO SECTION IN MARTHA USER'S MANUAL.

NAME	LEFT ARG	RIGHT ARG	RESULT	PURPOSE	REF
WHATIS		ANYTHING		IDENTIFIES ARGUMENT	3.8
STORE		FOF		COMBINES FOF WITH STORED	3.10
STORE		OUTPUT LIST		COMBINES OUTPUT WITH STORED	3.10
ATATIME	SCALAR NUMBER	FREQUENCY VECTOR	FOF	MANY-FREQUENCY SWEEP	3.11
SWEPPAR		PARAMETER VECTOR	FOF	PARAMETER SWEEP	3.12
WAVESAT	NO. OF WAVELEN	REFERENCE FREQUENCY	(SPECIAL CODE)	SETS UP CALCULATION OF PHYSICAL LENGTH	4.5
COAX		TWO RADII LENGTH	Z0, LENGTH	CALCULATES Z0 OF COAXIAL LINE	4.5
MICROSTRIP		STRIP WIDTH SUBSTRATE THICKNESS	Z0, LENGTH	CALCULATES Z0 OF MICROSTRIP LINE	4.5
COAXDISCAP		COMMON RAD 2 DISC RAD	CAPACITANCE	CALCULATES CAP. OF COAXIAL DISCONTINUITY	4.5
RECT		2 DIMENS LENGTH	FC, ZINF, LENGTH	CALCULATES FC, ZINF $FC=1.5E8 \div A \times DIEL \star .5$ $ZINF=377 \div DIEL \star .5$	
RECT1		2 DIMENS LENGTH	FC, ZINF, LENGTH	CALCULATES FC, ZINF $FC=1.5E8 \div A \times DIEL \star .5$ $ZINF=753 \times B \div A \times DIEL \star .5$	4.5
RECT2		2 DIMENS LENGTH	FC, ZINF, LENGTH	CALCULATES FC, ZINF $FC=1.5E8 \div A \times DIEL \star .5$ $ZINF=377 \times B \div DIEL \star .5$	4.5
CIRCTE11		RADIUS	FC, ZINF, LENGTH	CALCULATES FC, ZINF	
CIRCTE21		LENGTH		$FC=3E8 \div N \times A \times DIEL \star .5$ $ZINF=377 \div DIEL \star .5$	
CIRCTE01					
CIRCTM01					
CIRCTM11					

VALUES OF N FOR CIRCULAR GUIDE:

TE11	TE21	TE01	TM01	TM11
3.412	2.057	1.64	2.613	1.64

WR2300 THROUGH WR3 -- SEE HOWWR IN 100 MARTHAX

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ERROR MESSAGES IN MARTHA (BESIDES APL ERROR MESSAGES):

START OF MESSAGE	MEANING
ATTEMPT TO DIVIDE BY ZERO	ATTEMPT TO DIVIDE BY COMPLEX ZERO OR TAKE ITS RECIPROCAL
NOT A CAPACITANCE	ARGUMENT FOR FUNCTION C HAS LENGTH $\neq 1$
NOT A CGS, CGD, GM	ARGUMENT FOR FUNCTION FET HAS LENGTH $\neq 3$

NOT A NETWORK      ATTEMPT TO USE RESPONSE FUNCTION WITH IMPROPER NETWORK, EG, ZG, ZL, ZN, ZNIN, OR ZNOUT

NOT AN FC, ZINF, LENGTH      ARGUMENT FOR FUNCTION WG HAS LENGTH  $\neq$  2 OR 3

NOT AN INDUCTANCE      ARGUMENT FOR FUNCTION L HAS LENGTH  $\neq$  1 OR 3

NOT AN RX, RPI, CPI, CMU, GM      ARGUMENT FOR FUNCTION HYBRIDPI HAS LENGTH  $\neq$  5

NOT ANY DEFINING FREQUENCIES      ATTEMPT TO USE AN NDE OR FOF WITH 0 DEFINING FREQUENCIES

NOT A REFERENCE FREQUENCY      ATTEMPT TO USE A REFERENCE FREQUENCY BELOW CUTOFF FREQUENCY OF A WAVEGUIDE

NOT A RESISTANCE      ARGUMENT FOR FUNCTION R HAS LENGTH  $\neq$  1

NOT A SCALE FACTOR      LEFT ARGUMENT FOR FUNCTION ZSCALE HAS LENGTH  $\neq$  1

NOT A SECTION      ATTEMPT TO USE A WIRING FUNCTION ON SOMETHING OTHER THAN A SECTION OR AN ELEMENT; OR IMPROPER FORM FOR EG, ZG, ZL, ZN, ZNIN, OR ZNOUT

NOT A TURNS RATIO      ARGUMENT FOR FUNCTION IT HAS LENGTH  $\neq$  1

NOT A VOLTAGE GAIN      ARGUMENT FOR FUNCTION OPAMP HAS LENGTH  $\neq$  1, 2 OR 3

NOT A Z0, LENGTH      ARGUMENT FOR FUNCTION TEM HAS LENGTH  $\neq$  1 OR 2

## ADDITIONAL ERROR MESSAGES FROM FUNCTIONS IN THE MARTHA LIBRARY:

START OF MESSAGE	MEANING
ATTEMPT TO EXPONENTIATE ZERO	ATTEMPT TO RAISE 0 TO A POWER WITH A NEGATIVE REAL PART
ATTEMPT TO TAKE LOG OF ZERO	ATTEMPT TO TAKE LOGARITHM OF 0 OR CALCULATE A LOG TO BASE 1
INVALID FREQUENCY	ONE OR MORE FREQUENCIES OUTSIDE RANGE SET BY FLIMITS
PLEASE USE	ATTEMPT TO USE OBSOLETE FUNCTION
PREVIOUS STORED RESULTS LOST	ATTEMPT TO APPEND WITH FUNCTION STORE WITH WRONG NUMBER OF COLUMNS
REPLACE LINE [3]	ATTEMPT TO USE FUNCTION ATATIME OR SWEEP PAR WITHOUT HAVING EDITED THEM
NOT A BETA, IC	ARGUMENT FOR FUNCTION BETAIC HAS LENGTH $\neq$ 2
NOT A CHARGE GAIN	ARGUMENT FOR FUNCTION QCQS HAS LENGTH $\neq$ 1
NOT A CIN, CMID, COUT	ARGUMENT FOR FUNCTION CPI OR CTEE HAS LENGTH $\neq$ 3
NOT A COMMON RADIUS, TWO	ARGUMENT FOR FUNCTION COAXDISCAP HAS LENGTH $\neq$ 3 OR COMMON RADIUS LIES BETWEEN DISCONTINUOUS RADII
NOT A CURRENT GAIN	ARGUMENT FOR FUNCTION CCCS HAS LENGTH $\neq$ 1
NOT A CURRENT SCALE FACTOR	ARGUMENT FOR FUNCTION ISCALOR HAS LENGTH $\neq$ 1
NOT A FLUX GAIN	ARGUMENT FOR FUNCTION FCFS HAS LENGTH $\neq$ 1
NOT A FOF	ARGUMENT FOR FUNCTION ABCDFOF, COLUMNSOF, HFOF, OUTFOF, SFOF, YFOF, ZFOF, OR A FUNCTION IN 100 MARTHAN IS NOT A FOF, OR HAS WRONG NO. OF COLUMNS
NOT A MATRIX	ARGUMENTS FOR FUNCTION TOFOF ARE NOT CONFORMABLE, OR RIGHT ARGUMENT NOT A MATRIX
NOT A MID FREQ	LEFT ARGUMENT FOR FUNCTION FBP HAS LENGTH $\neq$ 1
NOT A MIN VALUE, MAX VALUE	LEFT ARGUMENT FOR FUNCTION HSCALE OR VSCALE HAS LENGTH $\neq$ 2, OR HAS EQUAL VALUES.
NOT AN ANGLE, CAPACITANCE	ARGUMENT FOR FUNCTION CREFLECTOR OR CROTATOR HAS LENGTH $\neq$ 2
NOT AN ANGLE, INDUCTANCE	ARGUMENT FOR FUNCTION LREFLECTOR OR LROTATOR HAS LENGTH $\neq$ 2
NOT AN ANGLE, RESISTANCE	ARGUMENT FOR FUNCTION RREFLECTOR OR RROTATOR HAS LENGTH $\neq$ 2
NOT AN FC, ZINF, ATTENUATION	ARGUMENT FOR FUNCTION WGATTEN HAS LENGTH $\neq$ 3

NOT AN FC, ZINF, FWD LOSS, ARGUMENT FOR FUNCTION WGISOLATOR HAS LENGTH # 1, 2 OR 3

NOT AN FC, ZINF, LENGTH ARGUMENT FOR FUNCTION TM HAS LENGTH # 2 OR 3

NOT AN FMIN, FMAX LEFT ARGUMENT FOR FUNCTION FLIMITS HAS ODD LENGTH

NOT AN INVERSE CAPACITANCE ARGUMENT FOR FUNCTION QCVS HAS LENGTH # 1

NOT AN INVERSE INDUCTANCE ARGUMENT FOR FUNCTION FCCS HAS LENGTH # 1

NOT AN LIN, LMID, LOUT ARGUMENT FOR FUNCTION LPI OR LTEE HAS LENGTH # 3

NOT AN N, ABCD, ARGUMENT FOR FUNCTION ABCDPDE HAS LENGTH # 3, 4, OR 5, OR FIRST ELEMENT IS NOT AN INTEGER WITH MAGNITUDE  $\leq 5$

NOT AN N, H11, ARGUMENT FOR FUNCTION HPDE HAS LENGTH # 3, 4, OR 5 OR FIRST ELEMENT IS NOT AN INTEGER WITH MAGNITUDE  $\leq 5$

NOT AN N, Y11, ARGUMENT FOR FUNCTION YPDE HAS LENGTH # 2, 3, 4, OR 5, OR FIRST ELEMENT IS NOT AN INTEGER WITH MAGNITUDE  $\leq 5$

NOT AN N, Z11 ARGUMENT FOR FUNCTION ZPDE HAS LENGTH # 2, 3, 4, OR 5, OR FIRST ELEMENT IS NOT AN INTEGER WITH MAGNITUDE  $\leq 5$

NOT A RADIUS, LENGTH ARGUMENT FOR FUNCTION CIRCTE11, CIRCTE21, CIRCTE01, CIRCTM01, OR CIRCTM11 HAS LENGTH # 1 OR 2

NOT AN RIN, RMID, ROUT ARGUMENT FOR FUNCTION RPI OR RTEE HAS LENGTH # 3

NOT A PIVOT FREQ LEFT ARGUMENT FOR FUNCTION FINVERT HAS LENGTH # 1

NOT A REFERENCE FREQUENCY ATTEMPT TO USE A REFERENCE FREQUENCY BELOW CUTOFF FREQUENCY OF A TM WAVEGUIDE

NOT A RESISTANCE ARGUMENT FOR FUNCTION GYRATOR OR LEFT ARGUMENT FOR FUNCTION WDUAL HAS LENGTH # 1

NOT A SCALE FACTOR LEFT ARGUMENT FOR FUNCTION FSCALE HAS LENGTH # 1

NOT A STRIP WIDTH, SUBSTRATE ARGUMENT FOR FUNCTION MICROSTRIP HAS LENGTH # 2 OR 3

NOT A TRANSCAPACITANCE ARGUMENT FOR FUNCTION VCQS HAS LENGTH # 1

NOT A TRANSCONDUCTANCE ARGUMENT FOR FUNCTION FCQS OR VCCS HAS LENGTH # 1

NOT A TRANSFER INVERSE TIME ARGUMENT FOR FUNCTION FCVS OR QCCS HAS LENGTH # 1

NOT A TRANSFER TIME ARGUMENT FOR FUNCTION CCQS OR VCFS HAS LENGTH # 1

NOT A TRANSINDUCTANCE ARGUMENT FOR FUNCTION CCFS HAS LENGTH # 1

NOT A TRANSRESISTANCE ARGUMENT FOR FUNCTION CCVS OR QCFS HAS LENGTH # 1

NOT A VALID NETWORK RIGHT ARGUMENT FOR FUNCTION FSCALE, FINVERT, OR FBP CONTAINS A UDE OR (EXCEPT FOR FSCALE) A TRANSMISSION LINE OR WAVEGUIDE

NOT A VOLTAGE GAIN ARGUMENT FOR FUNCTION VCVS HAS LENGTH # 1

NOT A VOLTAGE SCALE FACTOR ARGUMENT FOR FUNCTION VSCALOR HAS LENGTH # 1 OR ARGUMENT FOR FUNCTION PSCALOR HAS LENGTH # 2

NOT A Z0, ATTENUATION ARGUMENT FOR FUNCTION ATTENUATOR HAS LENGTH # 2

NOT A Z0, FWD LOSS, REV LOSS ARGUMENT FOR FUNCTION ISOLATOR HAS LENGTH # 1, 2 OR 3

NOT TWO INSIDE DIMENSIONS, ARGUMENT FOR FUNCTION RECT, RECT1, OR RECT2 HAS LENGTH # 2 OR 3

NOT TWO RADII, LENGTH ARGUMENT FOR FUNCTION COAX HAS LENGTH # 2 OR 3

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CHANGES SINCE THE PUBLICATION OF MARTHA USER'S MANUAL (THESE ARE ALL EXPLAINED IN DETAIL IN THE 1973 ADDENDUM TO THE MANUAL):

\*\*\*\*\* NOTE: FOF'S DEFINED BEFORE 15 JULY 1973 DO NOT WORK NOW. \*\*\*\*\*  
 \*\*\*\*\* SEE ITEM 3 BELOW. \*\*\*\*\*

1. NETWORK DEFINITIONS NO LONGER MUST BE APL DEFINED FUNCTIONS. ACTUAL ANALYSIS IS AUTOMATICALLY DEFERRED UNTIL REQUIRED. THE FREQUENCY MAY BE SET OR CHANGED AFTER THE NETWORK IS DEFINED. MARTHA USERS DO NOT NEED TO DEFINE ANY APL FUNCTIONS. THE FUNCTIONS ZSCALE, WAD, WCC, AND WDUAL ALTER THE ELEMENT VALUES AND/OR THE TOPOLOGY OF THE NETWORK. THE FUNCTION WHATIS NOW REPORTS AN ENTIRE NETWORK DEFINITION. ANALYSIS CAN BE DONE EARLY, IF DESIRED, BY THE NEW FUNCTION NDE WHICH PRODUCES A NUMERICALLY DEFINED ELEMENT FROM A NETWORK DEFINITION, WITH THE FREQUENCY VECTOR F AS THE DEFINING FREQUENCIES.

2. THE RESULT OF AN OUTPUT LIST IS A NUMERICALLY DEFINED FUNCTION OF FREQUENCY (FOF) WHICH CAN, AS BEFORE, BE PRINTED OR PLOTTED. IT CAN ALSO BE USED TO SPECIFY ZG, ZL, ETC., OR SERVE AS AN ARGUMENT FOR THE FUNCTIONS IN THE NEW WORKSPACE 100 MARTHAN.

3. FOF'S DEFINED IN VERSIONS OF MARTHA EARLIER THAN '73' DO NOT WORK NOW. THIS IS BECAUSE OF A CHANGE IN THE FORM IN WHICH FOF'S ARE STORED. TO CONVERT YOUR OLD FOF'S TO THE NEW FORM IS EASY:  
 NEWFOF- 1 3 2 Q OLDFOF

4. THE NETWORK-ENVIRONMENT VARIABLES EG, ZG, ZL, ZN, ZNIN, AND ZNOUT MAY NOW BE: (A) A SINGLE CONSTANT, AS BEFORE; (B) A 1-COLUMN OR 2-COLUMN FOF FOR A REAL OR COMPLEX FUNCTION OF FREQUENCY; OR (C) ANY 1-PORT NETWORK DESCRIBED IN MARTHA (THE IMPEDANCE WILL BE USED FOR ZG, ZL, ZN, ZNIN, AND ZNOUT, AND THE VOLTAGE PRODUCED BY A UNIT CURRENT WILL BE USED FOR EG). FOR EXAMPLE, ZN-WG 2E9 377 NORMALIZES TO THE FREQUENCY-DEPENDENT CHARACTERISTIC IMPEDANCE OF THE WAVEGUIDE, OR ZL-ZM2 OF NET SETS THE LOAD TO THE CONJUGATE-MATCH TERMINATION.

5. FOR QUICK PRINTOUTS WITHOUT THE AUTOMATIC HEADING, THE WORD PRINT MAY BE OMITTED. THE FREQUENCY APPEARS IN THE LAST COLUMN RATHER THAN THE FIRST.

6. ABOUT TWICE AS MANY FREQUENCIES AS BEFORE CAN BE HANDLED WITHOUT WS FULL MESSAGES.

7. THE COMMON WIRING CONNECTIONS (WP A) WC B AND (WS A) WC B ARE NOW AUTOMATICALLY REFERRED TO A FAST ONE-STEP ANALYZER.

8. USER-DEFINED ELEMENTS ARE MADE BY THE NEW FUNCTION UDE RATHER THAN BY A VECTOR BEGINNING WITH 9. THE FUNCTION NEWELEMENT MAY RETURN A 2-COLUMN MATRIX CONTAINING RE Z; IM Z FOR 1-PORTS, OR A 6-COLUMN MATRIX CONTAINING RE ABCD; IM ABCD; RE ABCD; IM ABCD; RE ABCD; IM ABCD FOR RECIPROCAL 2-PORTS, OR (AS BEFORE) AN 8-COLUMN MATRIX.

9. CAPACITANCES SPECIFIED AS 0 (INCLUDING THOSE IN HYBRIDPI AND FET) ARE AUTOMATICALLY SET TO 1E<sup>-25</sup> FARAD.

10. FREQUENCIES SPECIFIED AS 0 ARE SET TO 1E<sup>-13</sup> RATHER THAN 1E<sup>-25</sup> HZ.

11. ONE FUNCTION WDUAL REPLACES WDUAL1 AND WDUAL2.

12. NEW LIBRARY WORKSPACE 100 MARTHAF CONTAINS FORMAT OPTIONS.

13. NEW LIBRARY WORKSPACE 100 MARTHAN CONTAINS FUNCTIONS TO PERFORM ARITHMETIC ON REAL OR COMPLEX FOF'S. THE EXISTING FUNCTIONS MAKEFOF AND COLUMNSOF ARE MOVED THERE FROM THE WORKSPACE 100 MARTHAX.

14. NEW FUNCTIONS IN MARTHA LIBRARY WORKSPACES:  
 100 MARTHAE: WGATTEN WGISOLATOR TM ZPDE YPDE HPDE ABCDPDE UDE  
 100 MARTHAW: FSCALE FINVERT FBP FLIMITS WROTWROT WTOWN WTSWN  
 100 MARTHAR: ZI1 ZI2 ITC ZK1 ZK2 KTC ZM1 ZM2 VR IVG CG CR  
 RR MG UG SGP LM ISF SSF RF V1 V2 I1 I2 A1  
 A2 B1 B2 CP1 CP2 CP ML OMEGA OUTVAR  
 100 MARTHAM: NORM GAMMA SWR  
 100 MARTHAF: HSCALE VSCALE PLACES PAIRS SMITH EXPAND  
 100 MARTHAN: FADD FSUB FMUL FDIV FPWR FLOG FEQ FNEQ FCAT FRE  
 FIM FMAG FRAD FDEG FCC FSIG FEXP FLN FDF FCOL  
 INTERPOLATE FROMMAGDEG FROMDBDEG TOFOF FROMFOF  
 100 MARTHAX: RECT CIRCTE11 CIRCTE21 CIRCTE01 CIRCTM01 CIRCTM11

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