January/February 1982

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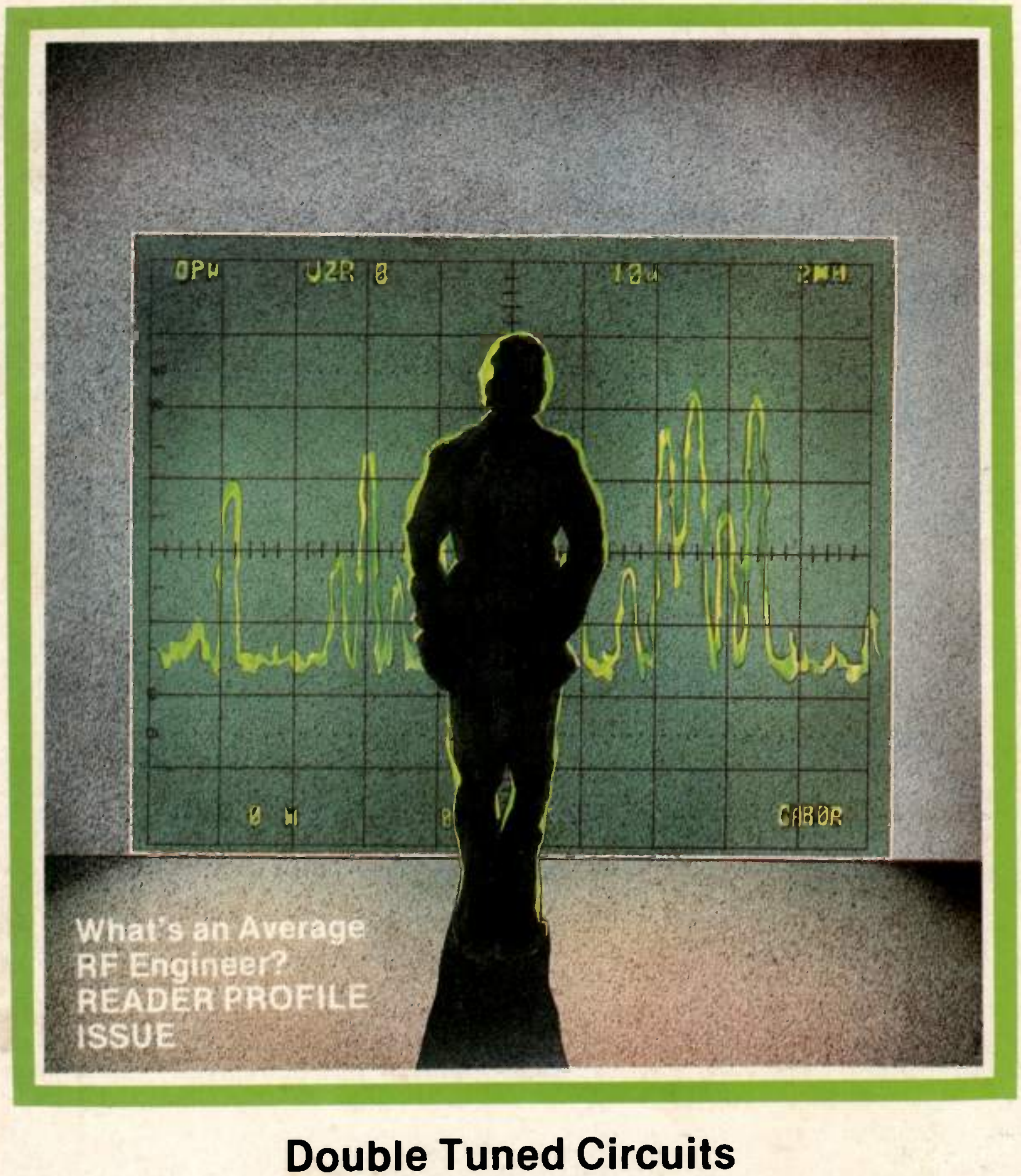
rf design



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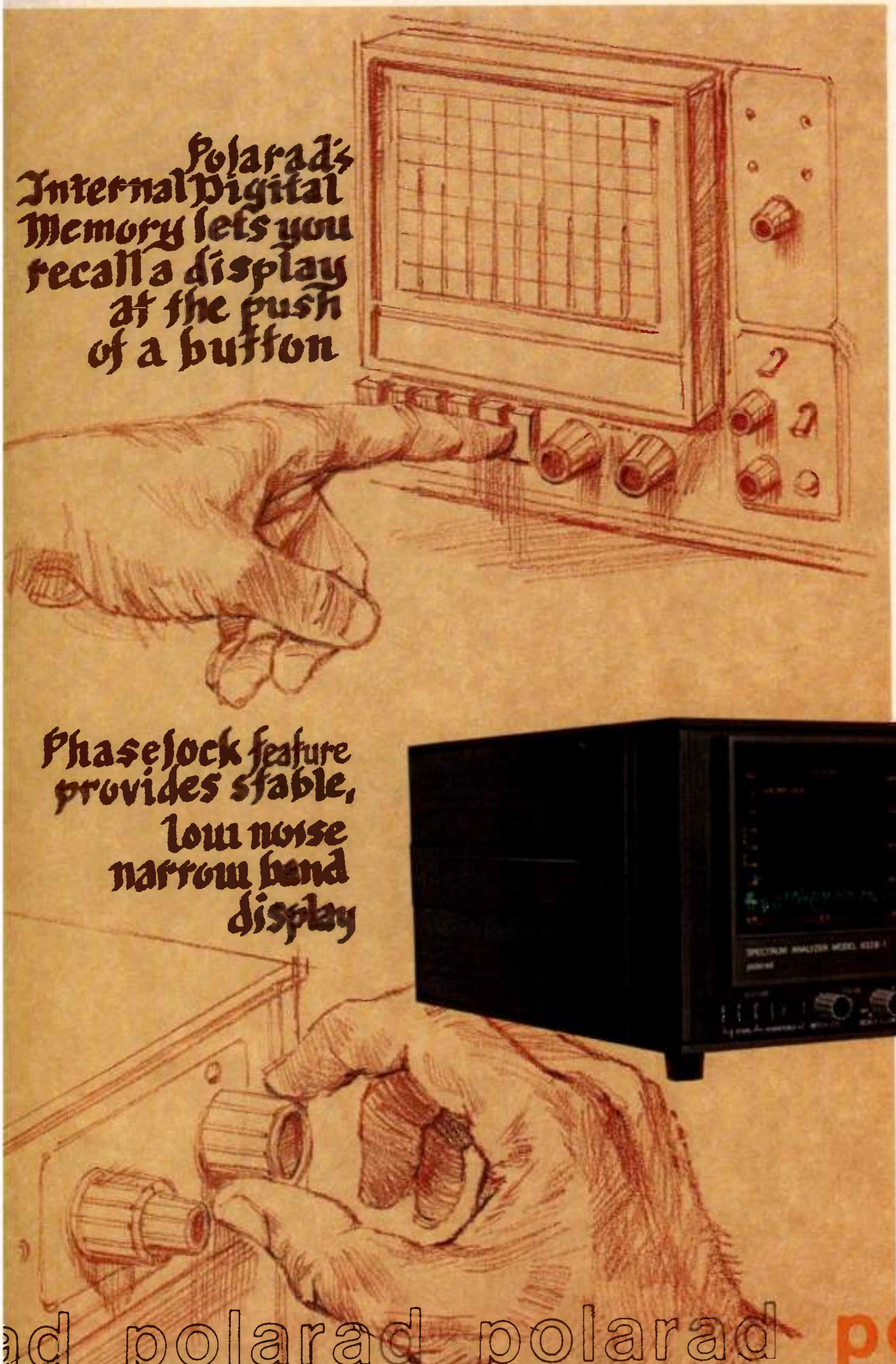

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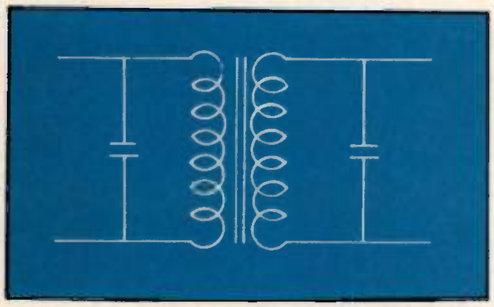

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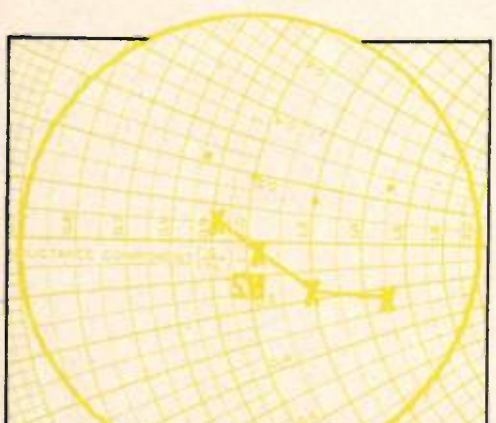

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IF	DC-500	DC-1000	DC-500	.5-500	DC-600
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one octave bandedge	6.5	6.0	7.5	7.5	7.0
total range	8.5	7.0	8.5	9.0	8.5
ISOLATION, dB, L TO R					
lower bandedge	50	50	45	45	50
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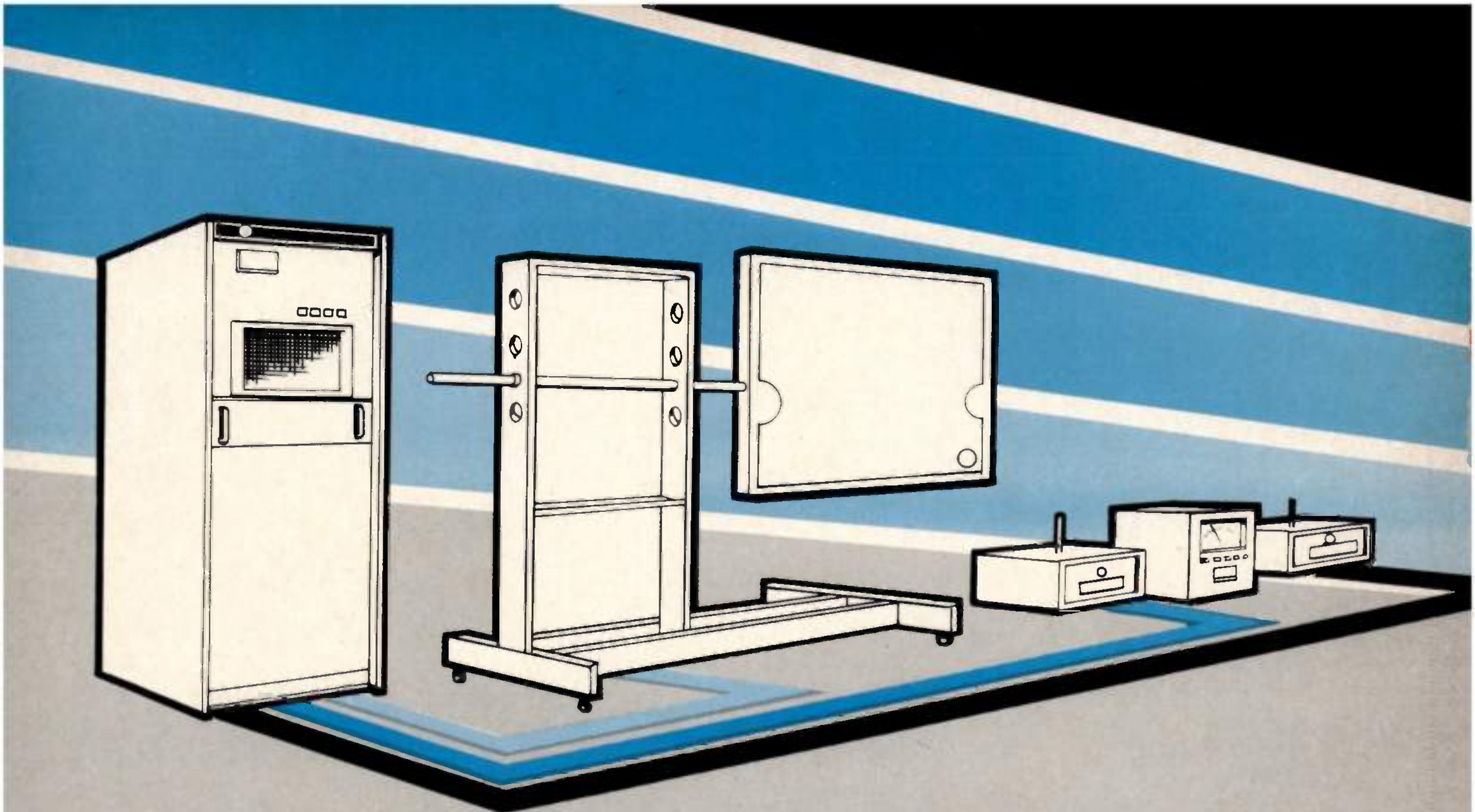

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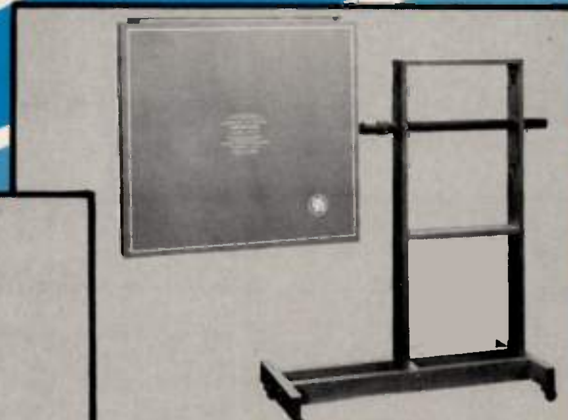
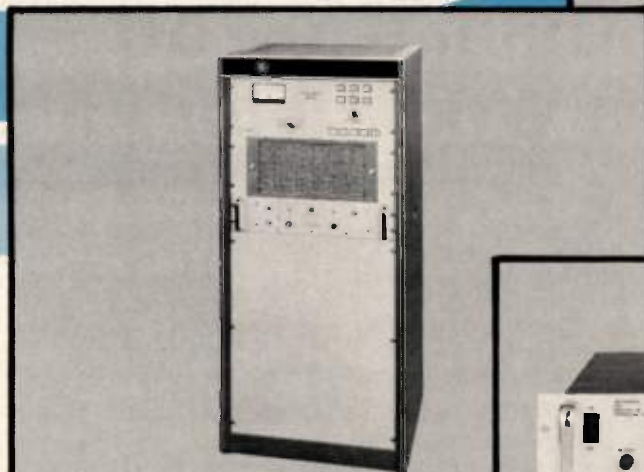
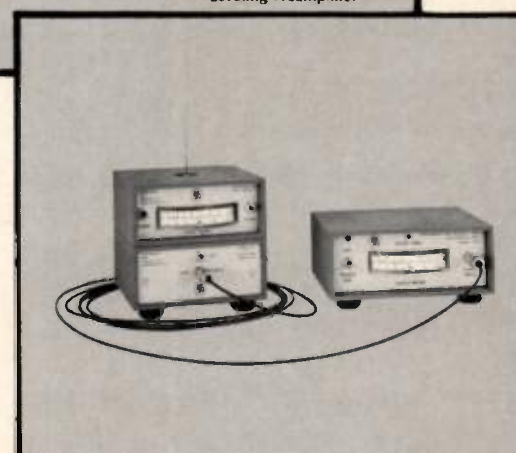


E-Field Generating Antenna—



Broadband Power Amplifiers—

Leveling Preamplifier—



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Thanks to You, It Worked for All of Us

In our last issue we announced our intention to prepare a reader profile. We told you that we were counting heavily on your participation to present a successful report. We're happy to say that your response was excellent, the report is a success, and we're proud to present the First Annual r.f. design Reader Profile in this issue.

Thanks to your help our report is full of interesting and, hopefully, useful information on RF engineers. Have you ever wondered if your salary is commensurate with your experience and educational level? Can you make more money with a different size company? Can a higher education lead to management? Is RF engineering a growing field? This report should provide you with some answers.

The results from this survey will also help us to present the kinds of articles you will find most useful. We are pleased to note that the majority of those who responded to our survey already find *r.f. design* to be the most helpful magazine they use for reference on their jobs.

Statiticians generally agree that a five percent return of a random survey will produce significant results. They also agree that a five percent return is all that the surveying company can reasonably expect. We sent out 1500 surveys with an intolerably quick deadline for return — and received well over a 19 percent response! Apparently you care as much as we do about the information in this report. We thank you for your cooperation.

The r.f. design Reader Profile will be presented annually. You should be able to detect trends in the RF engineering field. You should also be able to tell if you are keeping pace in your career. We are already looking forward to next year's report.

Bill W. Childs

Bill W. Childs
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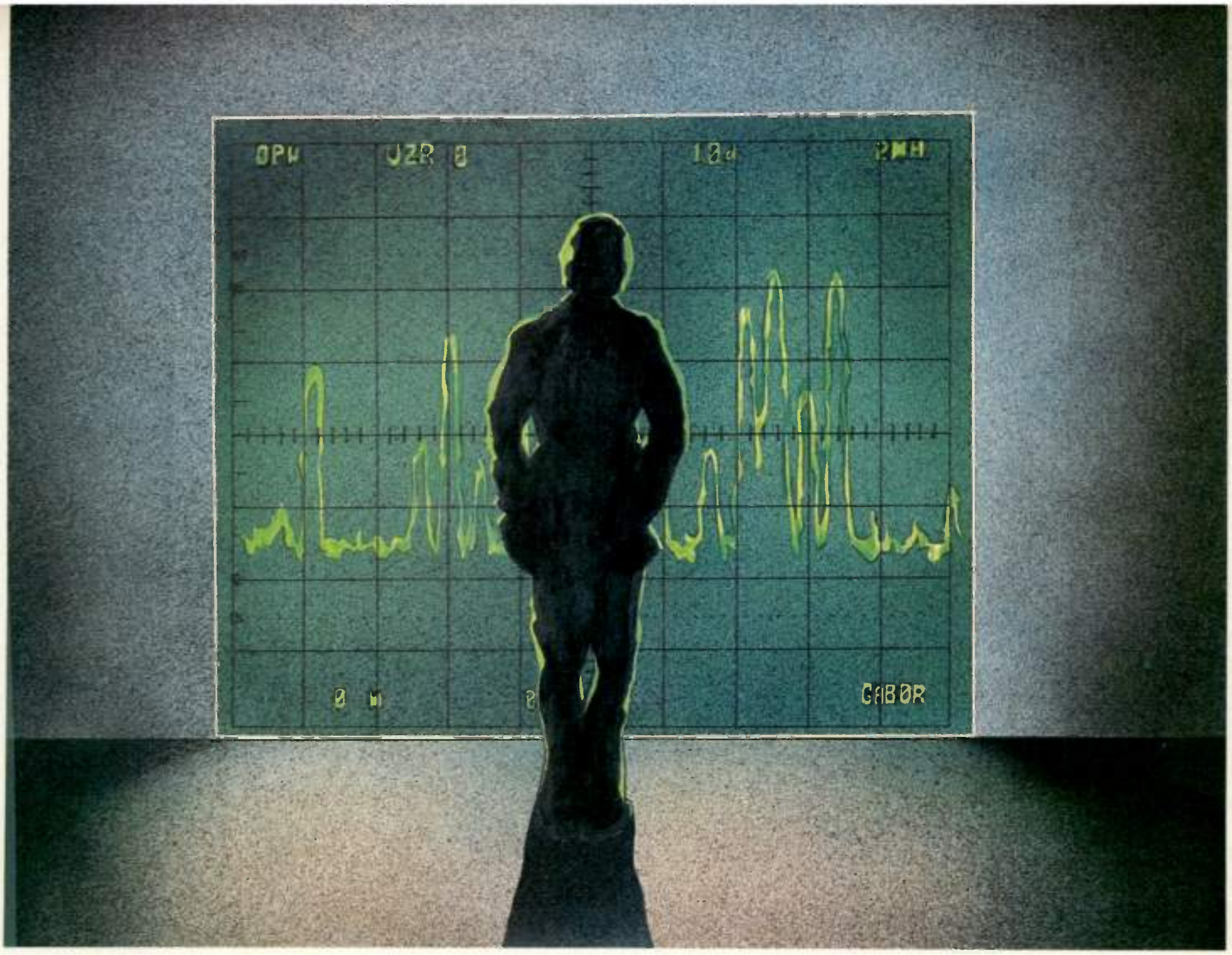


The First Annual rf design Reader Profile

Where do you stand in your career and what moves do you need to make to improve your position? The first in what will be an annual series of reports based on reader surveys helps you find the answers to these and other important questions.

By Doug Lumsden
Staff Editor

If you are typical of the readers who responded to our survey, you have been an engineer for nearly 15 years, and have been directly involved in rf design engineering for nearly 12 of those years. You received a bachelor of science degree and joined the ranks of the employed. Chances are, you didn't settle into your current position right away, but instead experimented a little, changing companies two or three times until you were satisfied. You work with nearly 1,000 other engineers in a company that grosses over $100 million dollars annually.

You are an important person in your company, with at least some influence, if not out-and-out control in the design and selection of test equipment, as well as active and passive components for your company.

You have an annual salary of about $30,000-35,000, and you feel like you are underpaid. Sometimes you wonder why you didn't become a doctor or a lawyer, but then you realize that your work is challenging, and, all-in-all, you enjoy it. You believe that the growth of rf engineering will be very significant within the next five years. Still, salary is important to you, and you wonder what you can do to improve it. For some of you, possible answers may be found in this report.

The Survey Method

We sent a limited number of questionaires to rf engineers at random in varying geographical groups across the country. The information in this report is based on the several hundred

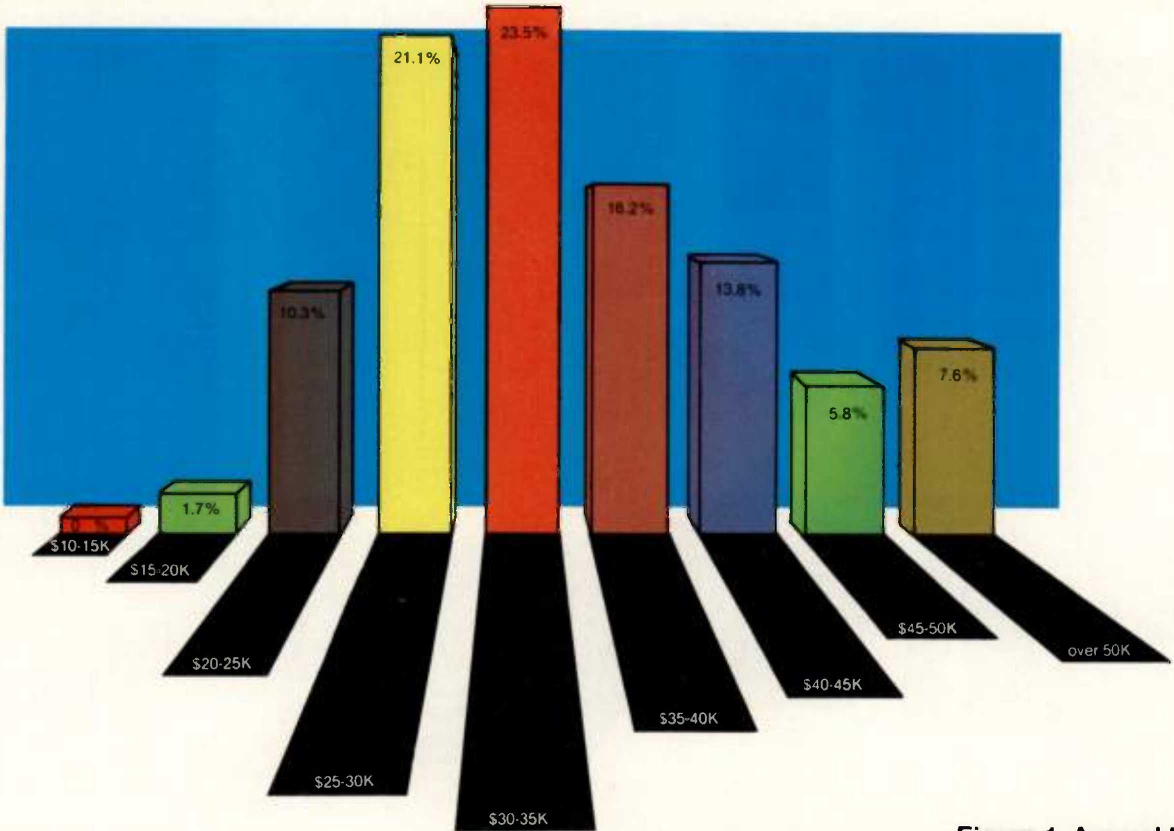


Figure 1. Annual Income

anonymous responses we received. Each response was used to represent many other engineers.

Obviously in a survey of this type we are dealing with statistical averages. To relate the findings to a specific engineer's job situation, he must take into account his own unique background, performance, community, and company factors.

The vast majority of responses to our questionnaire came from engineers, but we also heard from engineering managers, company owners, and corporate managers. The managers in our survey are responsible for an average of 16 people.

Who Has the Money?

Figure 1 shows the average annual income for all those who responded to our survey. In future reports, we will note with interest any changes in these figures.

Figure 2 shows the average annual income according to job function. These figures seem to indicate that an engineer can improve his income by becoming an engineering manager. The managers in our survey have a 25 percent larger income than the engineers.

How does an engineering manager maximize his income? By managing

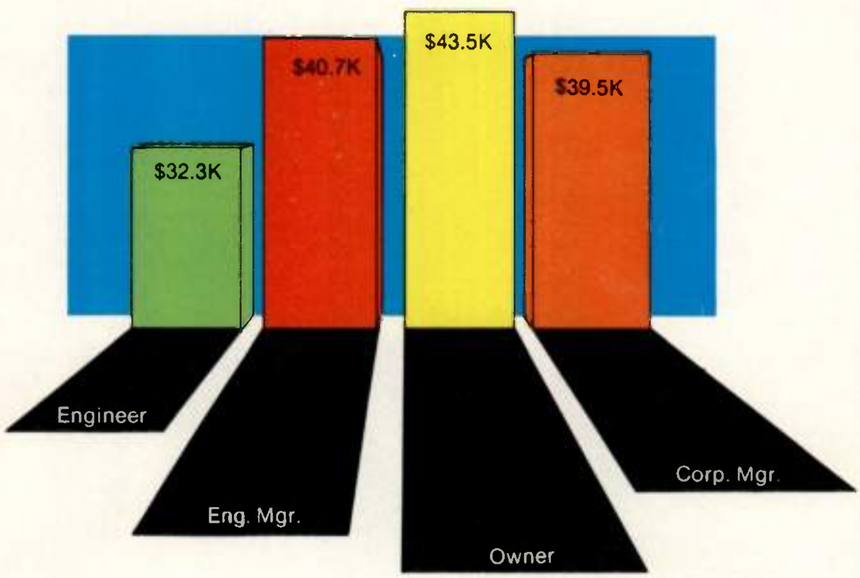


Figure 2. Income vs. Job Function

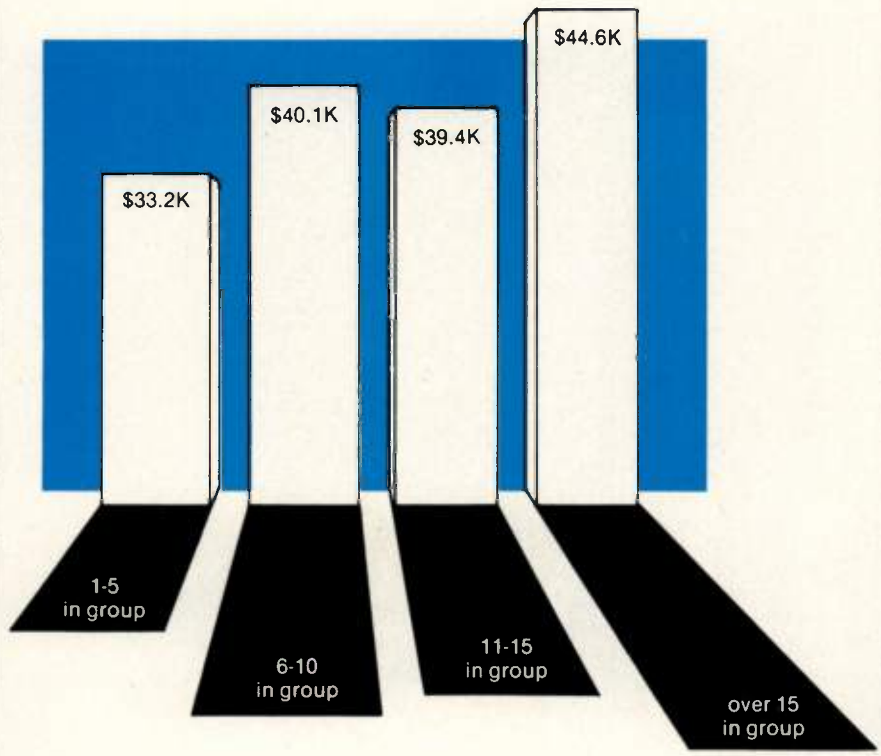


Figure 3. Income vs. Size of Group Managed

a larger group. This fact should surprise no one. But the difference in the salary of a small group manager and the manager of a large group is a staggering 33 percent!

The Education Factor

How important is education to an engineer? Comments in our survey seem to indicate that most engineers consider education to be a very significant factor. Feelings on type of degree and related study courses were mixed, however.

Some of the people who responded encouraged young engineers and engineering prospects to get an MS. Some said don't stop for anything short of a Ph.D. Others advised prospects to combine an engineering education with business, marketing, or extra math. A common piece of advice was to keep taking courses.

Does an advanced degree really make a significant difference in income? Figure 4 indicates that the route from a BS to an MS to a Ph.D seems to add $3,000 per degree level annually. A business degree seems to have little effect on income when held in conjunction with a BS.

Figure 5 indicates that engineers are basically a well-educated group,

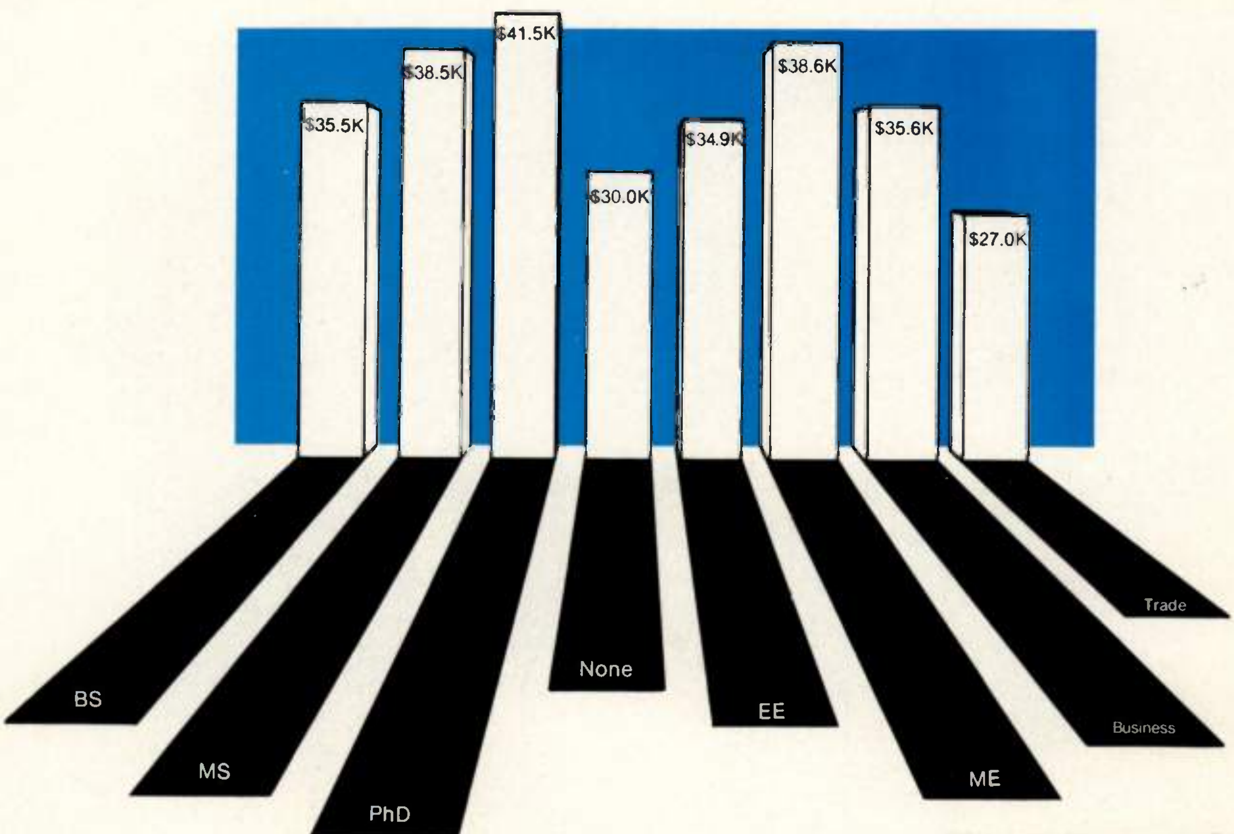


Figure 4. Income vs. Degree

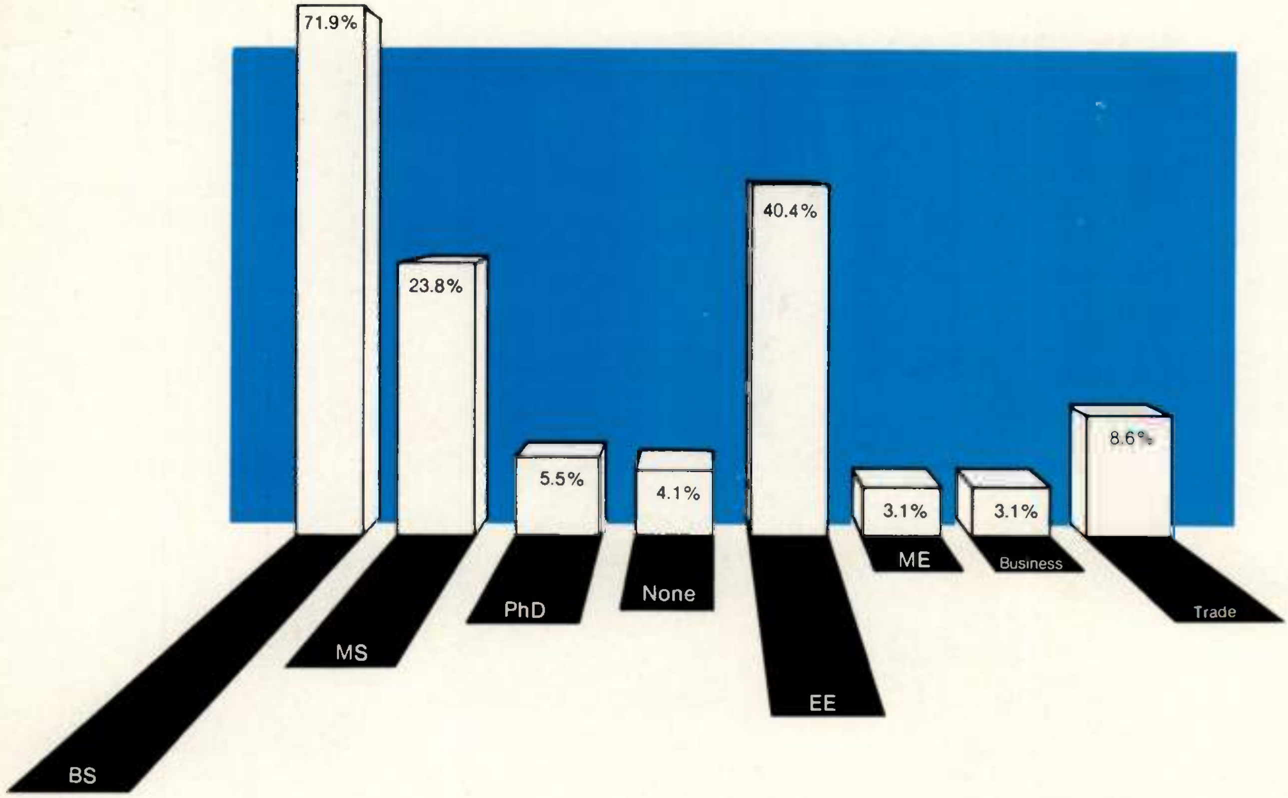


Figure 5. Degrees Held

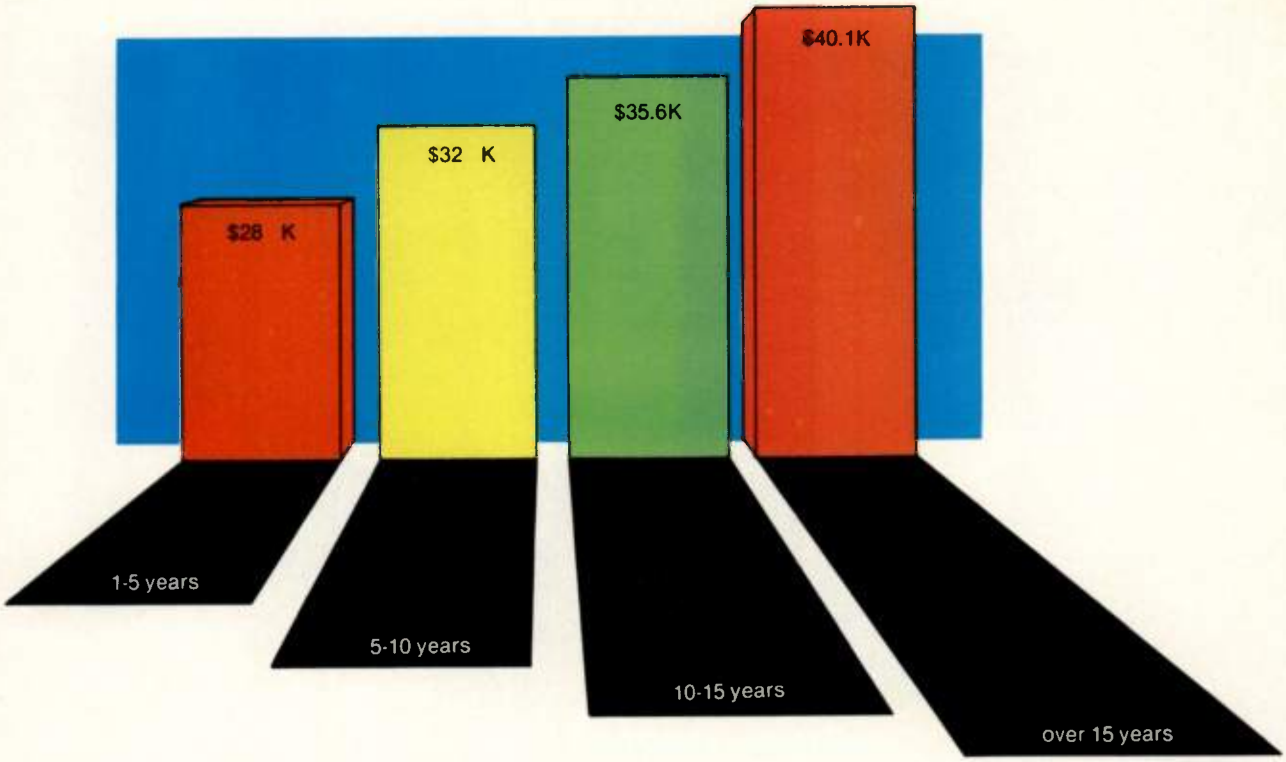


Figure 6. Income vs. Years in Field

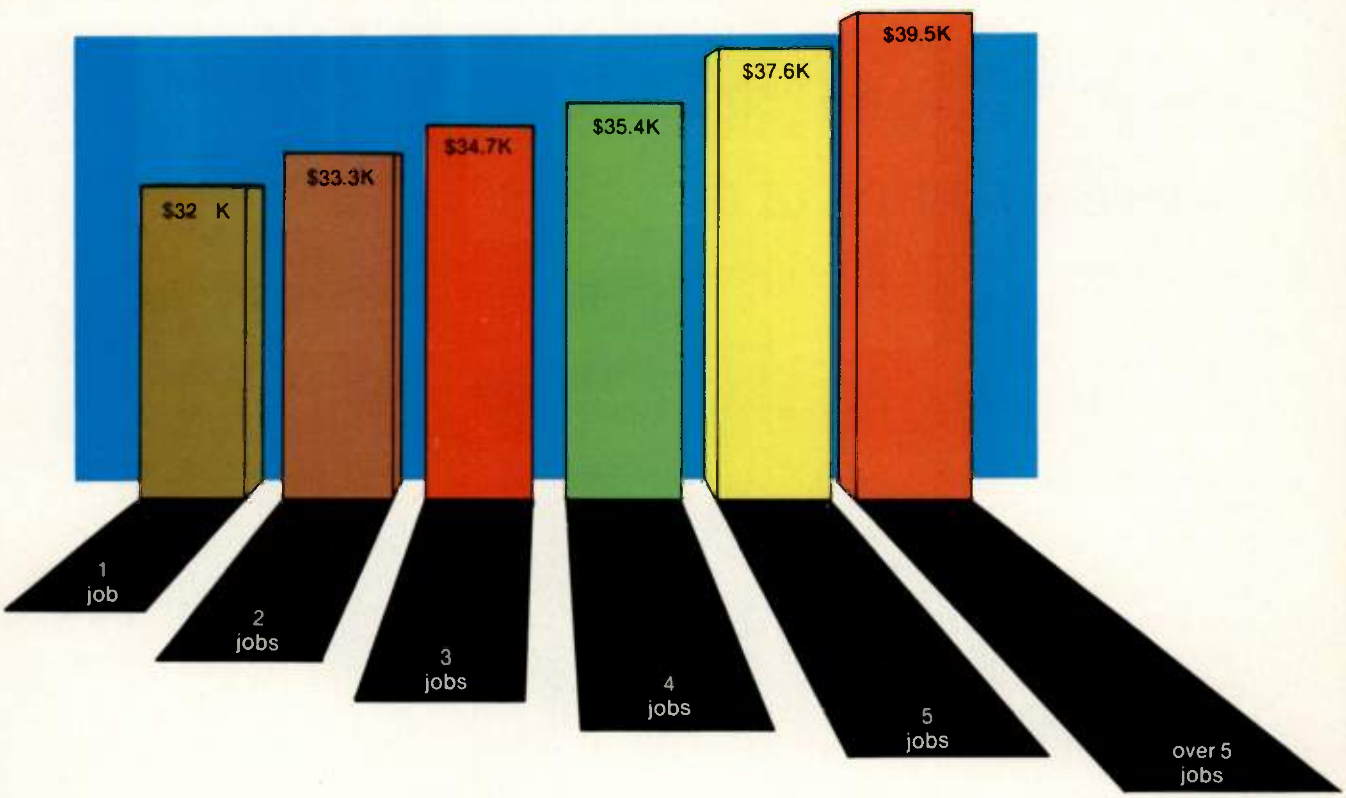


Figure 7. Income vs. Number of Jobs in Career

with 72 percent holding bachelor of science degrees and nearly one quarter having achieved an MS.

Experience is the Key

Our survey does not indicate that an advanced degree is necessary for promotion to management. The engineering manager actually has about the same amount of schooling as the people who work for him. As you might expect, job experience seems to be the real key for promotion to management. The typical engineering manager has six and a half more years of job experience than the engineer.

Figure 6 shows that an engineer can expect his salary to increase steadily as he gains experience. These figures are undoubtably affected by promotions and by the fact that many engineers are continuing their education while working.

Many of the engineers responding to our survey advised young engineers to change jobs a few times early in their careers as a method of increasing income. Does it work? Figure 7 indicates that a little job-hopping works very well. Income increases steadily as a function of number of jobs. The average engineer or engineering manager has had three jobs, and the average corporate manager or company owner has had a couple more. But overall, engineers seem to be very stable, spending 11.1 of their average 14.6 working years with one company. (See Figure 8.)

Larger Companies: Do They Mean Larger Income?

To some of the engineers we surveyed, it seems logical to expect a higher salary with a larger company. Others advised that smaller companies were more beneficial. Where can an engineer earn the most for his efforts?

Over 40 percent of the engineers in our survey work for large companies with over $100 million in gross sales annually. The average salary for this group is $36,200. The average salary for those engineers in our survey who work for smaller companies is $33,600.

Seemingly, large companies offer higher salaries. However, if we break these figures down a little further, a contradictory trend develops.

Since it is hard to refer to, say, a $75 million company as "small," let's examine just those companies whose gross annual sales number under $1 million. About nine percent of the

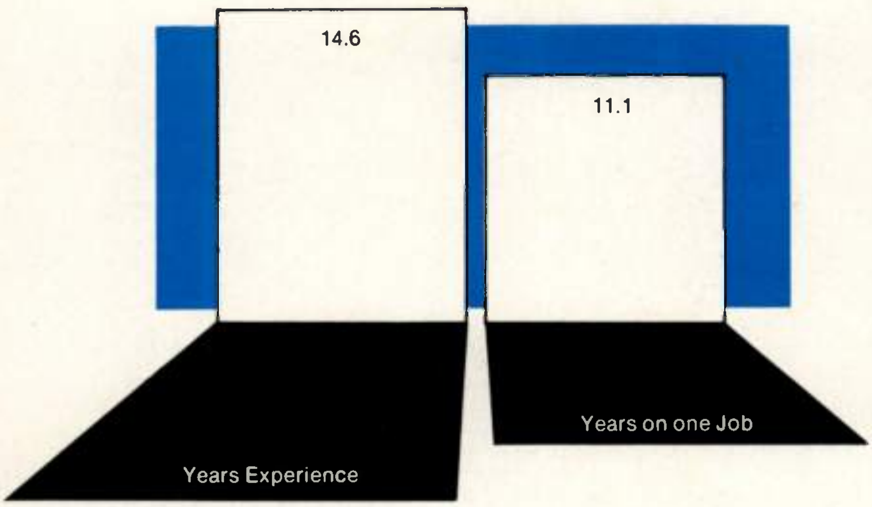


Figure 8. Years Experience and Most Years on One Job

engineers we surveyed work for companies in this category — a small but significant amount. The average salary in this group is $36,500. Clearly, working for a small company can be as beneficial as working for a giant, at least as far as salary is concerned.

Our survey seems to indicate no very significant relationship between an engineer's salary and annual gross sales of a company.

But can the number of engineers in a company affect the salary level of the individual engineer? Is it possible that with fewer engineers drawing from the engineering budget each would receive a larger share? Or does a larger department mean a larger budget with more for each? Our survey indicates that a larger department means a higher salary — up to a point.

Figure 9 shows a small but steady increase in the income levels of the engineers in larger groups up to those in groups of 500. The incomes of engineers in groups larger than 500 tail off slightly, but not significantly. In general, more engineers in a company means a higher salary for each, but not much higher. The difference between the highest and lowest average salaries in Figure 9 is less than nine percent.

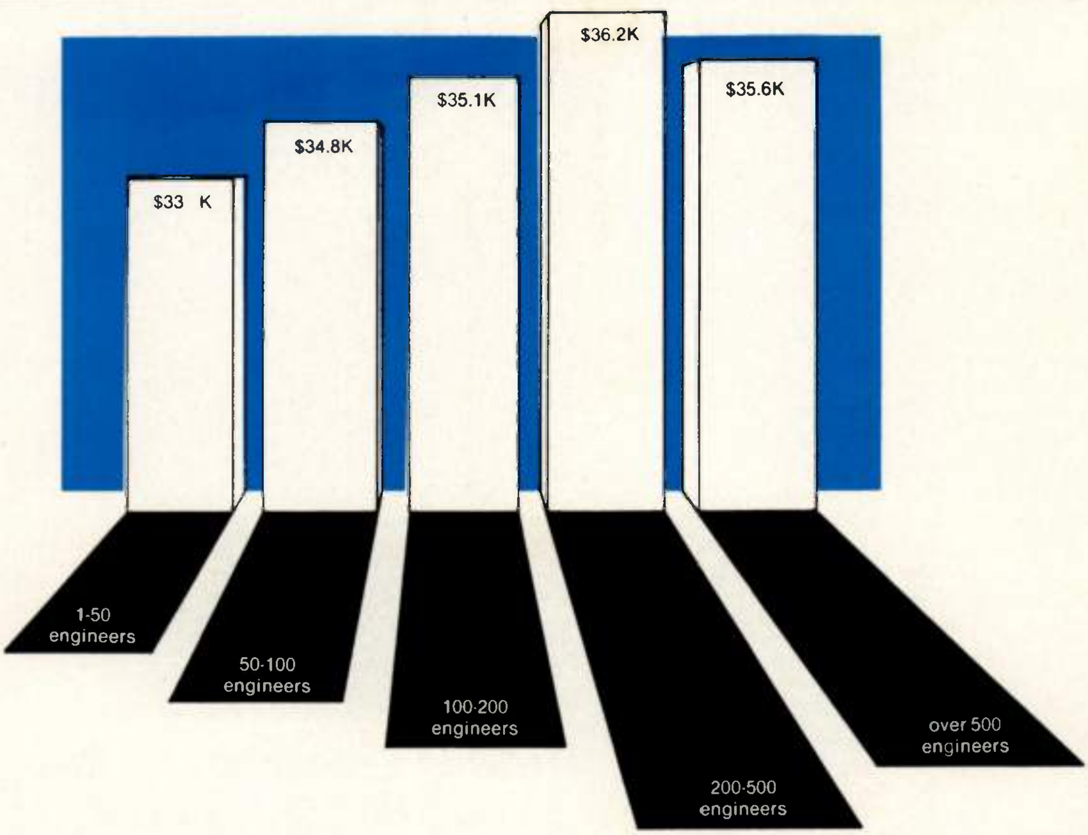


Figure 9. Income vs. Number of Engineers in Company

Why Do You Do It?

Many of the engineers we surveyed noted that money was not a reason to enter the engineering field. Asked to give advice to young engineers or engineering prospects, comments along the lines of "Don't expect excess dollars," "Don't do it for the money," and even "Go into law, medicine, or be a bookie" were common. Apparently rf engineers do not expect to get rich.

Yet, the vast majority of those surveyed like their work. "Challenging" was the word that appeared most often when the question "Why do you like your work?" was asked. "Fun" was another one. Or as one company owner put it: "Lifelong hobby." An rf engineer probably would dabble in electronics even if that wasn't his job. In fact, over 40 percent of those surveyed indicated that they are amateur radio operators.

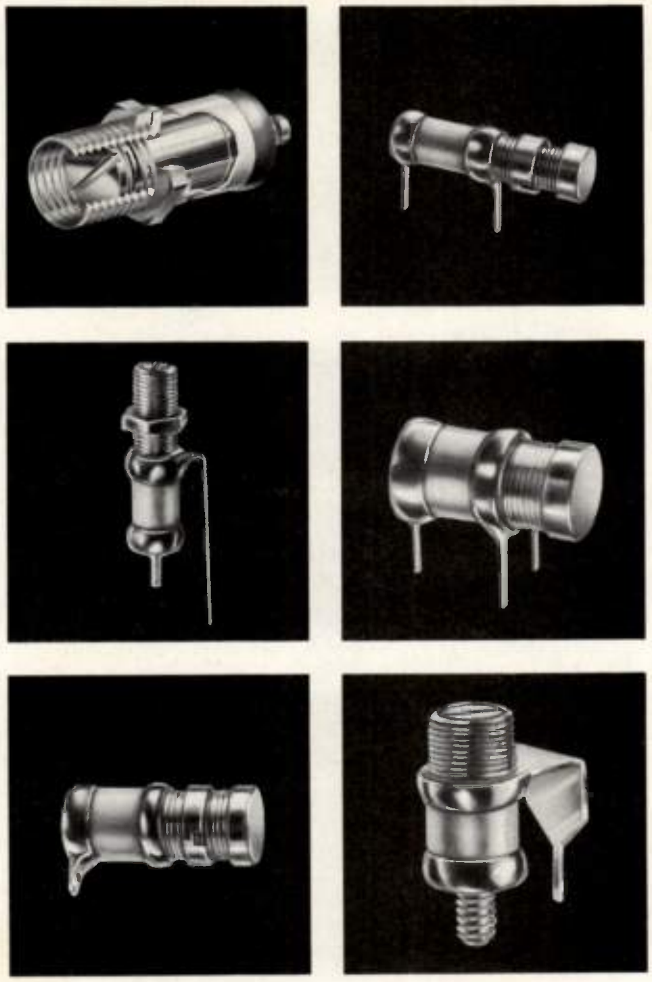
RF engineers also believe in the future of their field. Nearly 60 percent believe that the use of rf circuitry and systems for their own companies will increase more than 10 percent during the next five years. Or as one respondent advised young engineers: "Learn rf and you might survive 30 years."

First Annual Reader Profile

This year's survey and report has been the first in what will be an annual series of reader profiles. In future reports we will be able to compare current figures to past figures and pick up developing trends in the rf engineering field. These trends should provide valuable information to engineers and engineering prospects looking to advance themselves in their careers. At the very least, this information should be interesting to engineers who want to know where they stand in their careers compared with others in their field across the country.

The information in this report is statistically valid. However, an individual should take into account all relevant factors in his specific job situation before using these figures for comparison to arrive at generalized conclusions. We believe that this data can be used effectively to spot industry-wide averages and trends.

We wish to extend our appreciation to all those engineers, engineering managers, company owners, and corporate managers who participated in our survey. It is only through their cooperation that this reader profile could be compiled. □



DOUBLE TUNED CIRCUITS

. . . present some interesting design problems but the extra trouble can be worth the effort.

By Andrzej B. Przedpelski
A.R.F. Products, Inc.
R&D Laboratory
2559 75th Street
Boulder, Colorado 80301

Double tuned coupling circuits, either capacitively or inductively tuned, have been with us for a long time. (Reference 1.) Tube-type receivers used them almost exclusively in IF amplifiers. Transistors, with their odd impedances, necessitated some changes in configuration, but the old "double tuned IF transformer" survived fairly well. The inductively coupled transformer (Figure 1a) is more suitable for large production, since it eliminates one component.

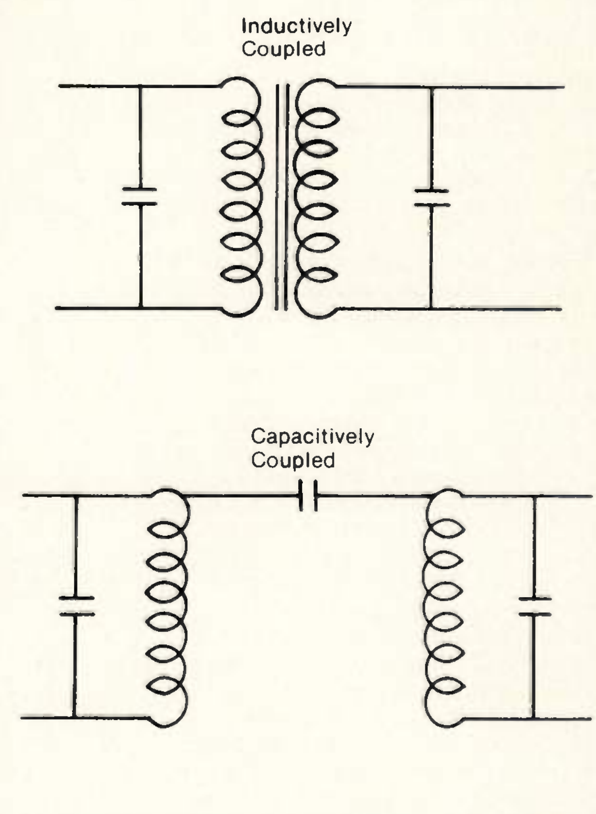


Figure 1. Basic Double Tuned Coupling Circuit

Its main disadvantage is that it is difficult to build and optimize when experimenting on breadboards. The capacitively coupled transformer (Figure 1b) on the other hand, while needing an extra capacitor to provide the required coupling, is very easy to construct from available components and is easier to optimize, since all values, including coupling, are easily adjustable.

The double tuned transformer is an almost ideal coupling network: it provides DC isolation, selectivity, low DC resistance load and source, and is easy to decouple. In the capacitively coupled version it is also very easy to design for breadboards and short production runs, and easy to optimize for critical applications. Therefore, only this type, in several versions, will be analyzed, but a similar analysis would also apply to the inductively coupled type.

Coupling Between Two Complex Unequal Impedances

The circuit shown in Figure 2 is suitable when maximum power transfer is required between two unequal complex impedances (a matched coupling circuit). Figures 2b and 2c show simplified equivalent circuits obtained by using series to parallel, parallel to series conversions and adding complex and real impedances between the conversions, wherever possible. (Reference 2.) Knowing only the input and output impedances, the Q of the inductor (Q_L) and desired bandwidth, (a function of Q_0), and basic relations shown in Table I, all the circuit components can be calculated. The calculation, using Figure 2, is as follows:

a. Convert series source impedance to equivalent parallel form: R_1 and X_5.

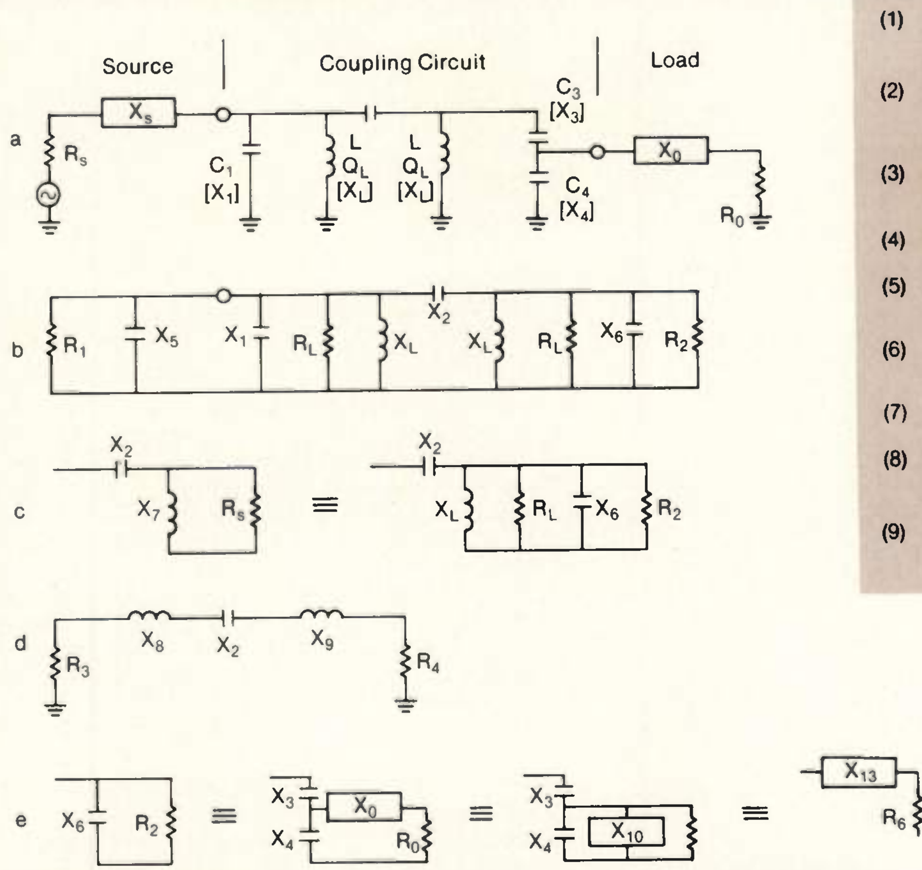
b. Combine (2), (3), and (4) to obtain:

$$(10)\quad X_L = R_1 \frac{Q_L - Q_0}{Q_L \cdot Q_0}$$

c. Calculate X_L and L.

d. Using parallel to series conversion and (6), (7), and (8) derive:

$$(11)\quad X_2 = -Q_0 \cdot X_L$$



$$(1)\quad R_1 > R_5$$

$$(2)\quad \frac{R_1 R_L}{R_1 + R_L} = R_P = R_s = \frac{R_2 R_L}{R_2 + R_L}$$

$$(3)\quad \frac{R_P}{X_L} = Q_P = Q_s = Q_0 = \frac{R_s}{X_L}$$

$$(4)\quad R_L = Q_L X_L$$

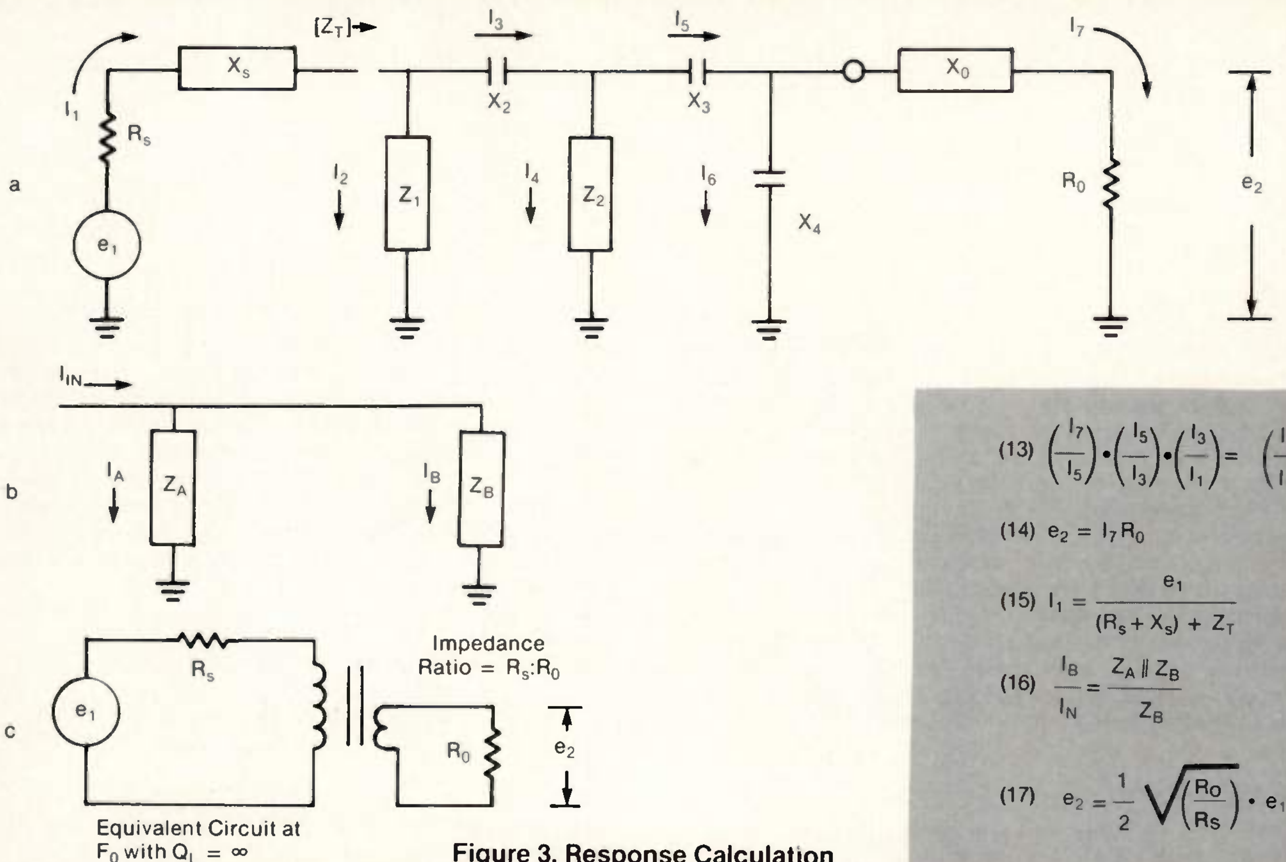
$$(5)\quad X_6 = X_1 \parallel X_5$$

$$(6)\quad k = \frac{1}{Q_0} = \frac{X_6}{X_6 + X_2} \quad \text{[For critical coupling]}$$

$$(7)\quad X_7 = X_6 \parallel X_L$$

$$(8)\quad X_2 = -2X_9 = -2X_8$$

$$(9)\quad X_4 = \frac{X_{11} X_{10}}{X_{10} - X_{11}}$$



$$(13)\quad \left(\frac{I_7}{I_5}\right)\cdot\left(\frac{I_5}{I_3}\right)\cdot\left(\frac{I_3}{I_1}\right) = \left(\frac{I_7}{I_1}\right)$$

$$(14)\quad e_2 = I_7 R_0$$

$$(15)\quad I_1 = \frac{e_1}{(R_s + X_s) + Z_T}$$

$$(16)\quad \frac{I_B}{I_N} = \frac{Z_A \parallel Z_B}{Z_B}$$

$$(17)\quad e_2 = \frac{1}{2}\sqrt{\left(\frac{Ro}{Rs}\right)}\cdot e_1$$

Figure 3. Response Calculation

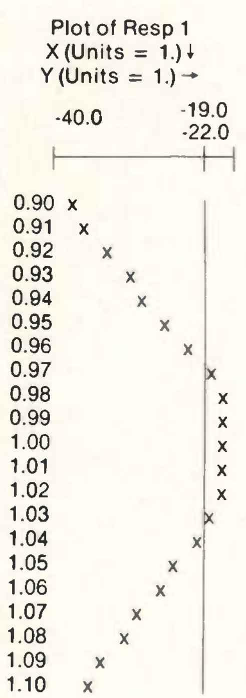


Fig. 4 Plot of Frequency Response

e. Calculate X_2 and C_2.
f. From (5) and (6) derive:

$$(12)\quad X_1 = \frac{X_2 \cdot X_5}{X_5(Q_0 - 1) - X_2}$$

g. Calculate X_1 and C_1.
h. Convert series load impedance to equivalent parallel form: R_5 and X_{10}.
i. Convert $R_2 \parallel X_6$ to series form: $R_6 + j(X_3 + X_{11})$.
j. Calculate X_{11}.
k. Using (9) calculate X_4 and C_4.
l. Calculate X_3 and C_3 ($X_3 + X_{11}$ is the series equivalent reactance of $X_6 \parallel R_2$).

This calculation is rather lengthy and is best left to a programmable calculator. A program for the HP-41C is available. Just send SASE or blank magnetic card to author and you will receive actual program.

It is usually desirable to know the frequency response of a given circuit. Several methods are available, but the one shown below is probably the easiest. The circuit of 2a can be redrawn, as shown in Figure 3a. Then, using (13), (14), (15) and (16) the response of the circuit can be calculated. The obtained gain, e_2/e_1, includes the effect of the impedance step-down, R_0/R_s, and the source resistance, R_s. Thus, at resonance and with infinite inductor Q_s (no dissipative loss), the output voltage will be as shown in equation (17).

As an example let us assume the following (Figure 2a):

$F_0 = 60 \times 10^6$ Hz	$X_0 = 20$
$R_s = 500$	$Q_0 = 20$
$X_s = -50$	$Q_L = 100$
$R_0 = 40$	

$$K = \frac{1}{\sqrt{Q_p Q_s}} = \frac{C_m}{\sqrt{(C_p + C_m)(C_s + C_m)}}$$

$$Q_p = \frac{\omega L_p}{R_p}$$

$$Q_s = \frac{\omega L_s}{R_s}$$

C_m = coupling capacity
C_p = total primary capacity
C_s = total secondary capacity

Table 1. Basic Relations

The Program calculates the circuit values:

$X_L = 20.2000$	$L = 0.0536$ UH
$X_1 = -213531$	$C_1 = 124$ PF
$X_2 = -404.0000$	$C_2 = 6.57$ PF
$X_3 = -14.6008$	$C_3 = 182$ PF
$X_4 = -6.3190$	$C_4 = 420$ PF

Using this Program we obtain an insertion loss at $F/F_0 = 1$ of 18.94 DB. From equation (17) we obtain an insertion loss (using lossless components) of 16.99 DB. Thus about 2 DB of the insertion loss can be attributed to the finite Q of the inductors. This loss is strictly dissipative, since the circuit is still matched because the program compensated for this loss in the circuit values calculation. Using the HP-41C printer PRPLOT function we can obtain the 3 DB bandwidth by placing the axis at −22 DB, as shown in Figure 4. The 3 DB points are approximately at F/F_0 of 0.967 and 1.039 or about 58 MHz and 62.3 MHz. A more detailed plot can be obtained by using smaller x-axis increments.

References

1. F.E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., Inc., 1943.
2. A.B. Przedpelski, "Simplify Conjugate Bilateral Matching of Complex Impedances," Electronic Design, March 1, 1978. □

Important!

The actual program is available from the author as described in the article. But we would like to know if you would rather we published all the programs for all the articles we publish. If you think we should, please circle #140 on RS card. If you think we should *not* then circle #139. If you don't care . . . don't do anything.

Part III

HF Transistor Amplifier Design

The third and last part of an investigation into practical Smith chart design.

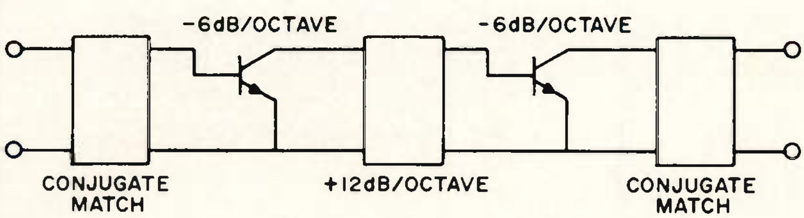


Figure 24. Two-Stage Gain Compensation.

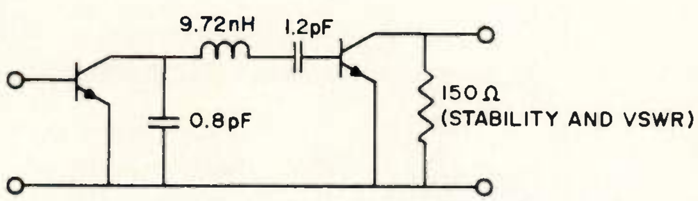


Figure 25. Interstage Gain Compensating.

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Parts I and II reviewed in some detail the theoretical aspects of Smith Chart design. This part concludes the series with a practical, typical design example of a two-stage amplifier.

Design Example #2 — Two Stage Amplifier

An amplifier design is required to meet the performance specification shown at right.

The amplifier input will be driven by an antenna via a bandbass filter, and the output will be used to drive a double-balanced mixer. This imposes two additional requirements on the design. Unconditional stability is a must when an amplifier is terminated by a filter, as these devices usually exhibit highly reative impedances outside their passbands. In order for a mixer to perform well, especially with respect to isolation and intermodulation, each of its ports must be reasonably well terminated at the L.O., I.F., and R.F. frequencies. The R.F. amplifier must, therefore, maintain decent output VSWR far out of band. For this reason, a broadband resistive network will be used for output match.

Two bipolar transistor stages will be used to achieve the 20 dB gain. The NE64535 is selected for use in both stages of the amplifier. The first stage will be biased at 7 mA Ic for low noise performance, while the

Frequency Range	1400 to 1700 MHz
Gain	20 dB min.
Gain Flatness	± 0.5 dB max.
VSWR (In and Out)	2.0:1 max.
Noise Figure	3.0 dB max.
Power Output at 1 dB Compression	+ dBm min.
DC Input	+ 15 Volts

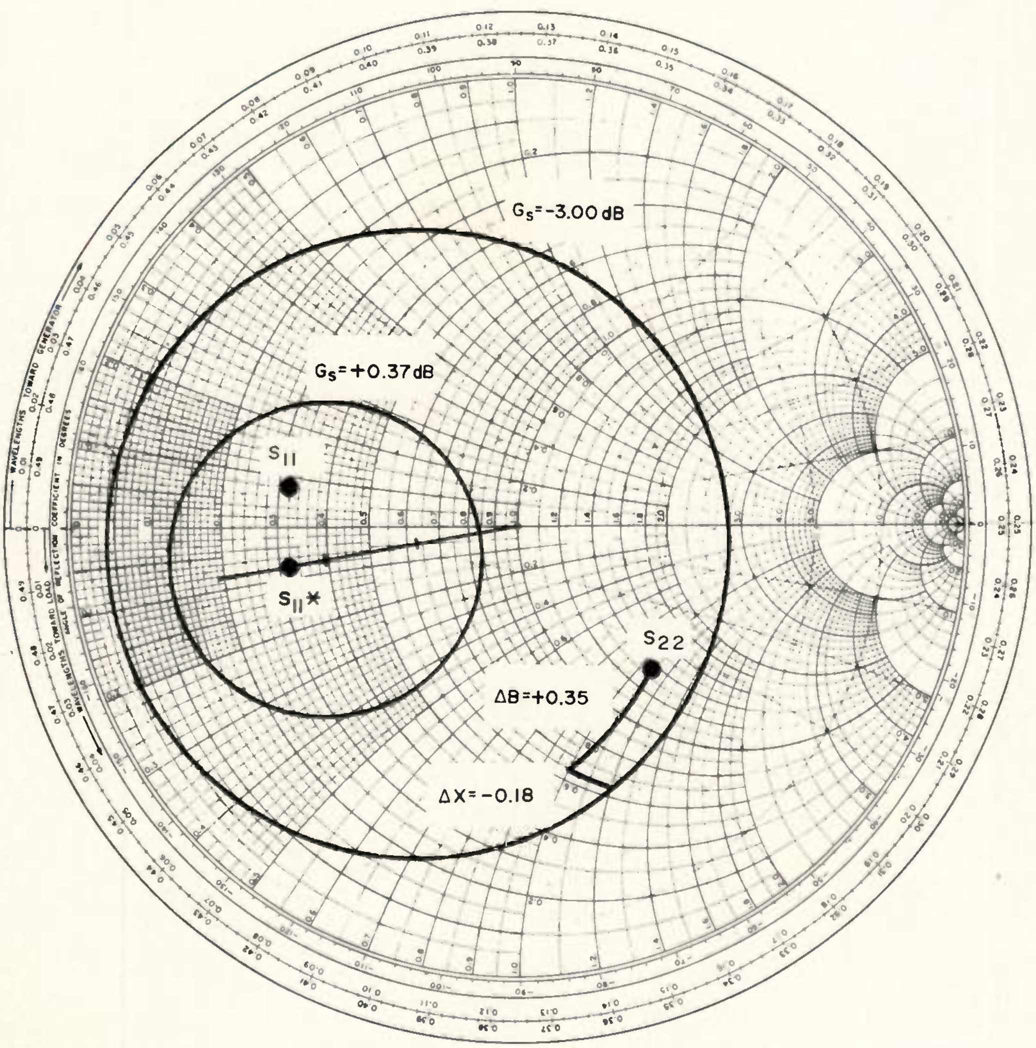

Figure 26. Interstage at 1400 MHz.

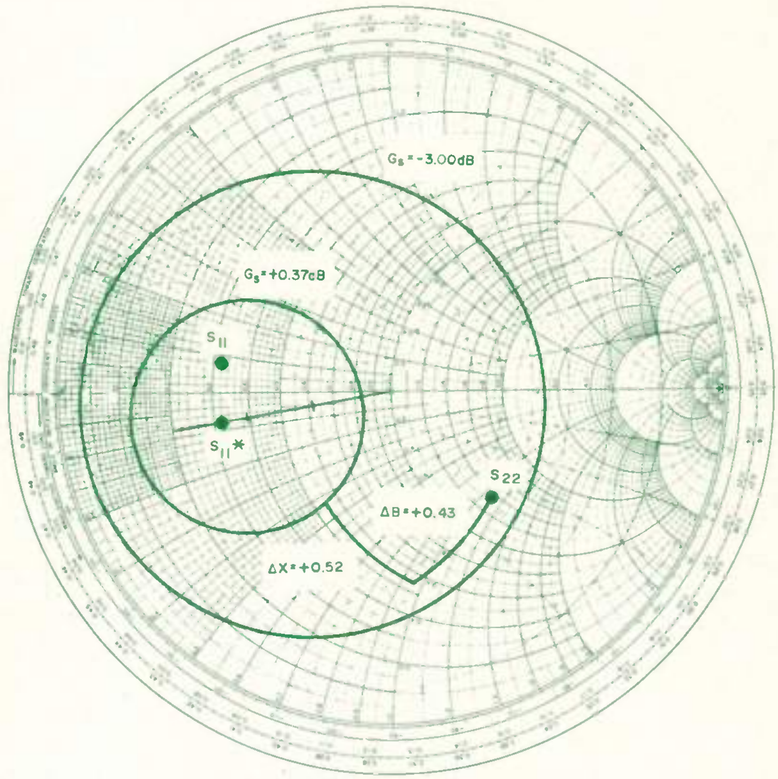

Figure 27. Interstage at 1700 MHz.

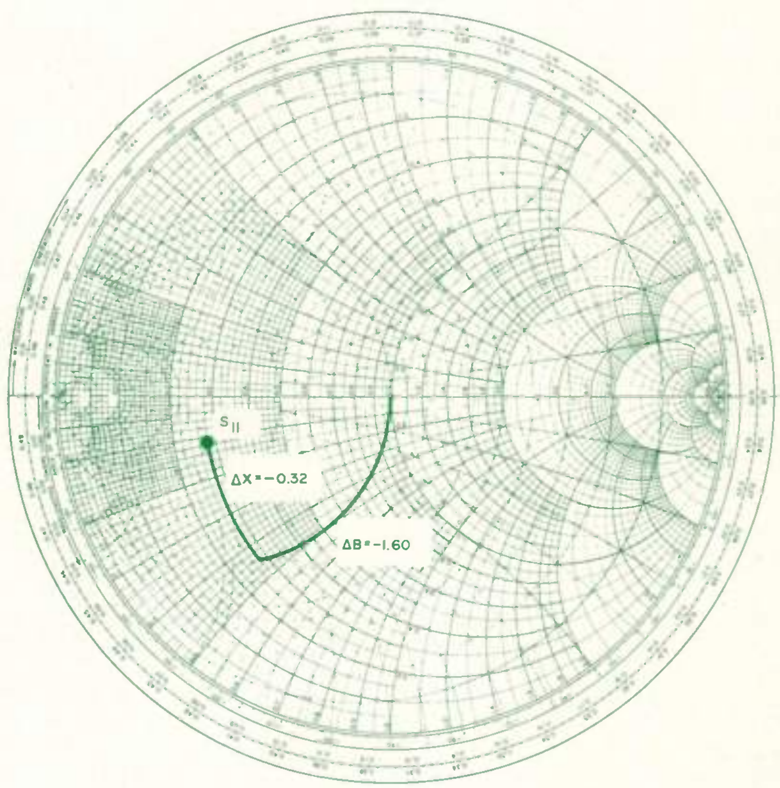

Figure 28. Input Match at 1550 MHz.

second stage will be biased at 20 mA for output power capability.

Device Stabilization

S-parameters of the two transistors are tabulated at right:

First Stage				
Frequency	S_{11}	S_{21}	S_{12}	S_{22}
1400 MHz	$.52\angle-171°$	$4.43\angle76.1°$	$.059\angle40.8°$	$.44\angle-47°$
1550 MHz	$.52\angle-176°$	$4.06\angle72.7°$	$.062\angle41.4°$	$.44\angle-48°$
1700 MHz	$.52\angle180°$	$3.79\angle69.7°$	$.064\angle42.0°$	$.44\angle-48°$

Second Stage				
Frequency	S_{11}	S_{21}	S_{12}	S_{22}
1400 MHz	$.52\angle172°$	$4.76\angle72.8°$	$.040\angle61.6°$	$.37\angle-39°$
1550 MHz	$.52\angle170°$	$4.34\angle70.0°$	$.045\angle63.0°$	$.38\ -40°$
1700 MHz	$.52\angle168°$	$4.08\angle67.1°$	$.049\angle63.6°$	$.38\angle-41°$

K-factors are calculated as approximately 1.1 for the first stage and 1.5 for the second stage. Both devices are unconditionally stable in the band of operation. Recall from Design Example #1, however, that this device becomes potentially unstable at lower frequencies. Unless both stages are stabilized at low frequency, the amplifier may exhibit out-of-band oscillation.

Stability of the second stage device will be provided by a shunt resistor, which is already required for the broadband output match. First stage stability will be ensured by careful selection of reactive networks. Figures 14 and 15 from Design Example #1 show that all potentially unstable source and load impedances and inductive (located in the upper half of the chart). By designing the reactive networks surrounding the first stage to present capacitive impedances at low frequency, unconditional stability may be preserved without introducing in-band loss.

Gain Compensation

The gain compensation method illustrated in Figure 24 will provide both flat gain and low input/output VSWR. Expected active device gain variation over the operating bandwidtn is:

$$\Delta G(\mathrm{dB}) = 40 \log \left[\frac{1400}{1700}\right] = -3.37 \text{ dB}$$

The interstage network must therefore have 3.37 dB more gain at 1700 MHz than at 1400 MHz.

For $S_{11}{}^*$ of the second device, G_{SMAX} = 1.37 dB at both 1400 and 1700 MHz. G_S will be designed for −3.00 dB at 1400 MHz and +0.37 dB at 1700 MHz. Location and radius of constant-gain circles are calculated from the equations given previously.

Frequency	G_S	Location (ds)	Radius (ps)
1400 MHz	−3.00 dB	0.23	0.70
1700 MHz	+0.37 dB	0.44	0.35

Illustrated in Figure 25 is a circuit which will transform S_{22} of the second stage onto the −3.00 dB circle at 1400 MHz and onto the +0.37 dB circle at 1700 MHz. At 1400 MHz, the normalized susceptance (ΔB) of the shunt

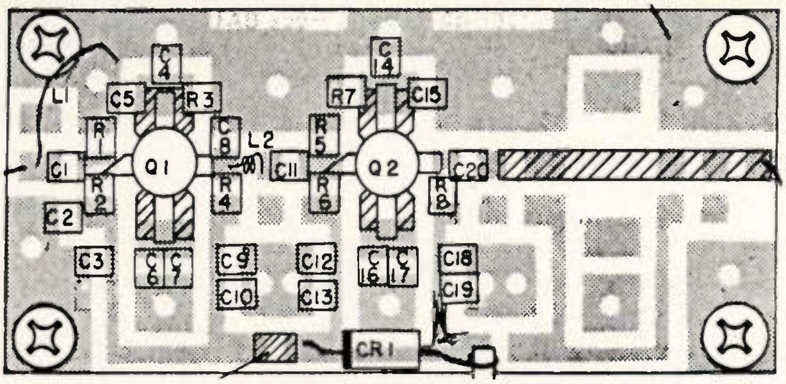

Figure 29. Two-Stage Circuit Layout.

capacitor is $+j0.35$ and the normalized reactance (ΔX) of the series L-C is $-j0.18$. at 1700 MHz, $\Delta B = +j0.43$ and $\Delta X = +j0.52$. Figures 26 and 27 illustrate the circuit behavior at 1400 and 1700 MHz. As desired for stability, the circuit appears entirely capacitive at all frequencies below 1400 MHz. Note: Due to the small spread in S_{11} and S_{22}, they have been assumed constant for graphical purposes.

A computer analysis was performed on the circuit of Figure 25, with results given below. Note: Included in the computer model, but not shown in Figure 25, were small shunt losses expected due to bias resistors.

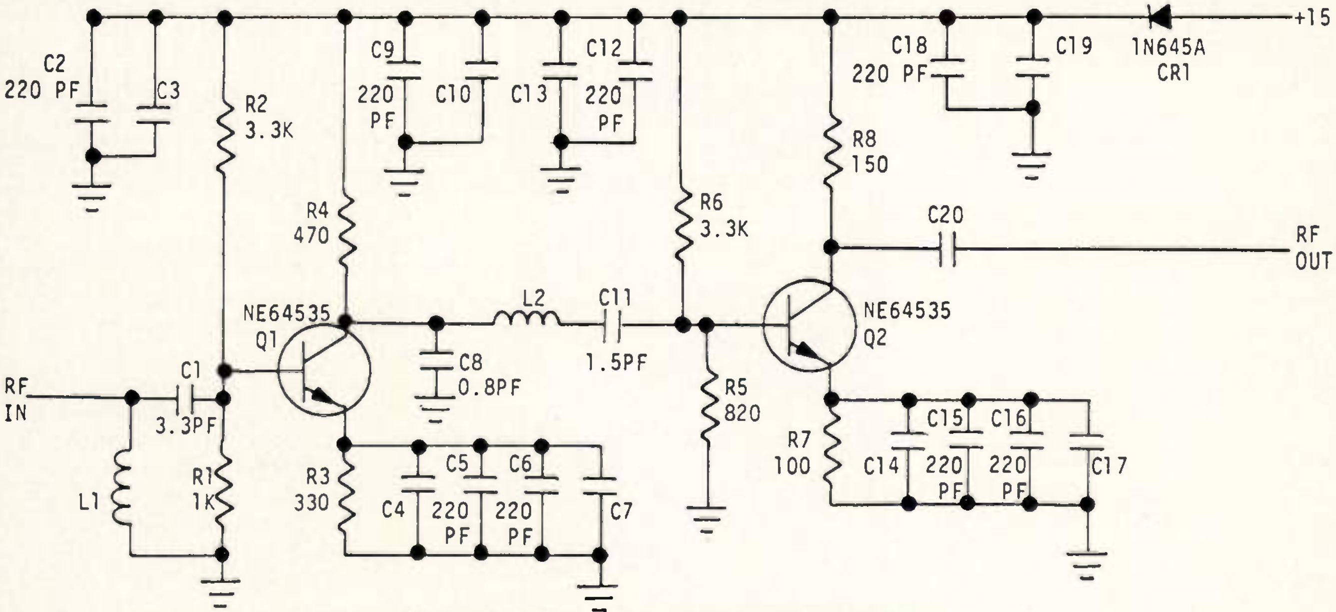

Figure 30. Two-Stage Amplifier Schematic.

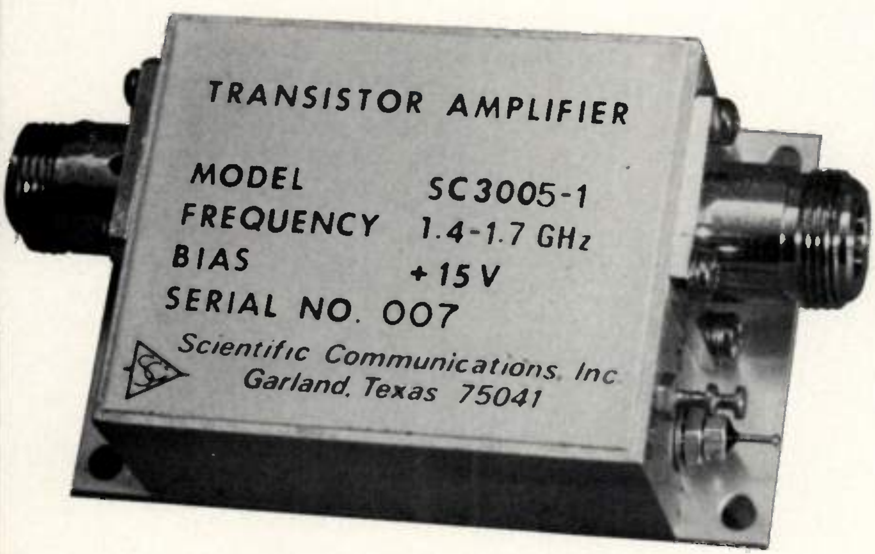

Figure 31. Two-Stage Amplifier.

Computer

Frequency	S'_{11}	Gain
1400 MHz	$.45 \angle -159°$	22.22 dB
1550 MHz	$.53 \angle -165°$	22.30 dB
1700 MHz	$.61 \angle -172°$	22.48 dB

Optimized-Gain

Frequency	S'_{11}	Gain
1400 MHz	$.48 \angle -159°$	22.97 dB
1550 MHz	$.57 \angle -166°$	23.11 dB
1700 MHz	$.65 \angle -178°$	22.91 dB

Final

Frequency	S'_{11}	Gain
1400 MHz	$.27 \angle 67°$	24.88 dB
1550 MHz	$.09 \angle -26°$	25.19 dB
1700 MHz	$.29 \angle -119°$	24.82 dB

The initial analysis shows gain to be slightly over compensated. Allowing the computer to optimize for gain flatness altered component values from 0.8 pF, 9.72 nH, and 1.2 pF to 0.92 pF, 10.36 nH, and 1.24 pF. Results are shown below.

Analysis

S'_{22}	K-Factor
$.13\angle-50°$	20.96
$.16\angle-49°$	13.11
$.20\angle-54°$	8.37

Matching

The predicted maximum output reflection coefficient (S'_{22}) of .22 is well within the 2.0:1 VSWR specification. A simple input matching network is designed in Figure 28. Component values are calculated as follows:

$$X_C = \Delta X(50) = .32(50) = 16 \text{ ohms}$$

$$C = \frac{1}{16 \times 2\pi \times 1.55 \times 10^9} F = 6.42 \text{ pF}$$

$$X_L = \frac{50}{\Delta B} = \frac{50}{1.6} = 31.25 \text{ ohms}$$

$$L = \frac{31.25}{2\pi \times 1.55 \times 10^9} H = 3.21 \text{ nH}$$

As required for stability, the topology was selected to present a capacitive impedance to the transistor at low frequency.

The input matching components are added to the computer model, and a final optimization performed. Optimized values are 3.32 nH and 5.45 pF for the input network. Optimized values for the interstage network are 0.78 pF, 8.20 nH, and 1.81 pF. Predicted performance is tabulated below.

Analysis

S'_{22}	K-Factor
$.14\angle-48°$	16.94
$.18\angle-50°$	10.04
$.22\angle-60°$	6.86

Results

Prototype units were constructed on fiberglass-epoxy board per the layout of Figure 29. After laboratory "tweaking," the final component values were documented as per the schematic of Figure 30. Notice that, with the exception of C1, the final component values are almost identical to those predicted by the graphical/calculator design (prior to computer optimization). Due to the inductance of a jumper wire to the RF input connector, the value of C_1 had to be reduced to preserve input match. The results shown above were measured on the prototype amplifiers over the 1400 to 1700 MHz frequency range.

Figure 31 is a photograph of the assembled amplifier. □

Analysis

S'_{22}	K-Factor
$.14\angle-43°$	13.16
$.19\angle-45°$	9.15
$.24\angle-56°$	7.03

Results

Gain	22 dB
Gain Flatness	± 0.2 dB
VSWR	1.6:1 max in, 1.5:1 max out
Noise Figure	2.3 dB max
Power Output at 1 dB Compression	+ 8.5 dBm min

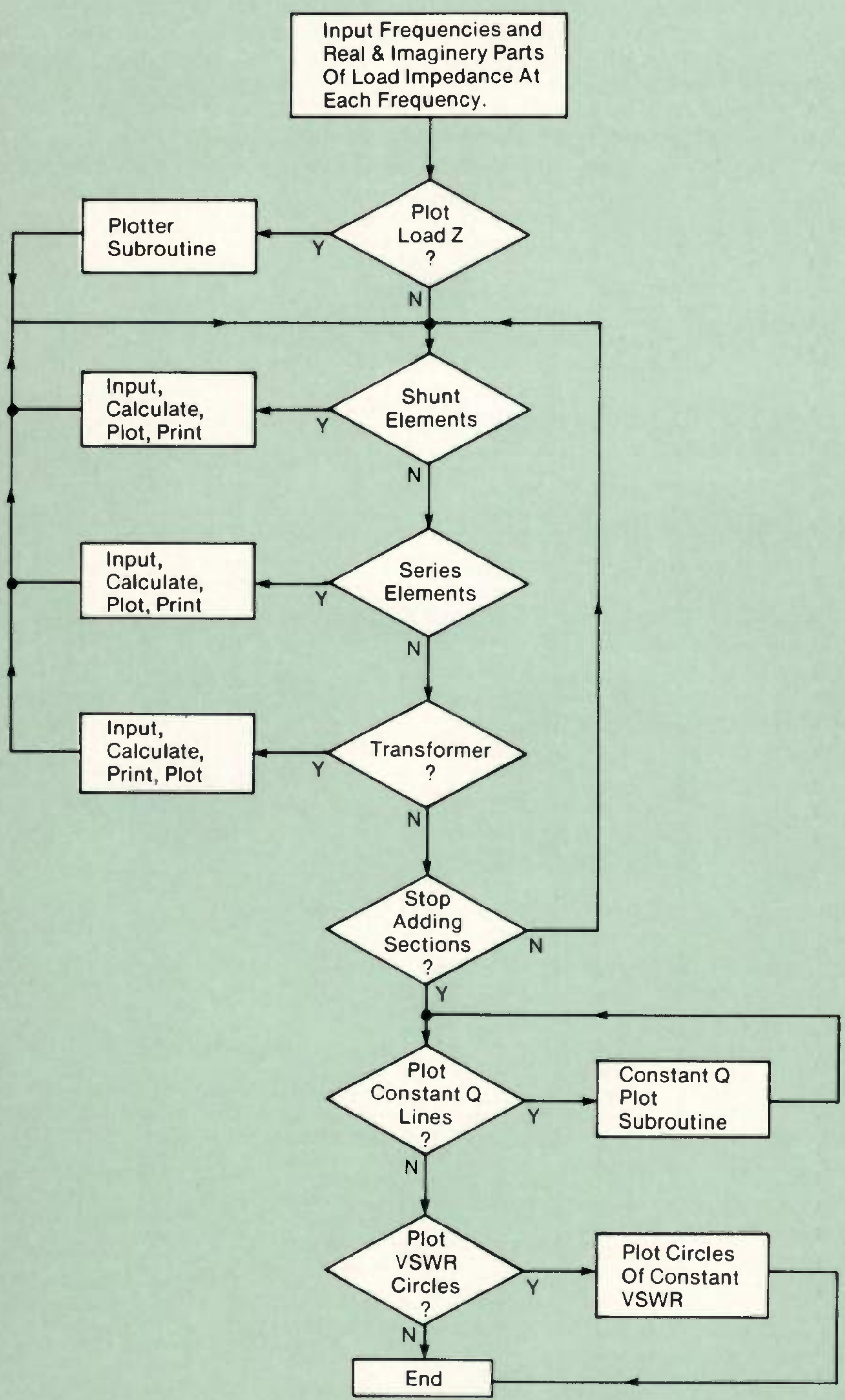

Figure 1. Flow Diagram for "Z-Match" Program.

Computer

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The Smith Chart has been used by RF engineers for many years as an aid in (1) designing impedance matching networks or (2) in determining the input impedances to various networks, etc. Many engineers prefer to use calculators or computers as design tools because of their speed and ability to give data in tabulated outputs. However, many RF engineers prefer to use the slower "graphical approach" because the Smith Chart plot provides a visual means of analization which is unsurpassed by columns of numbers.

This article describes the author's man-machine interface which utilizes a computer to perform high-speed calculations (performing in seconds those calculations which might take hours by hand with the graphical Smith Chart approach) and a plotter which automatically graphs the results on a Smith Chart for visual analization. First, this article describes several impedance matching network elements and lists the general equations for determining the resulting impedances in each case. These equations can be applied in various applications even if the author's program is not utilized. The second part of this article describes the author's program in BASIC as developed for use with an HP9830A desktop calculator and HP9862A calculator plotter. Details for plotter interface are discussed, and two simple design examples are given.

A Program For Calculating and Plotting

The equations for calculating input impedances to various networks are described in Appendix A. This section describes a program which was developed in BASIC for the HP9830A calculator and HP9862A plotter. First the general program will be discussed, then equations for plotting will be developed.

Interface For Smith Chart Calculations

After designing antenna matching networks for several years, the author developed the "Z-Match" program combining computers and Smith charts.

The Program

The flow chart for the program (called "Z-Match") is shown in Figure 1. Major sections of the program are listed by line number in Figure 2, and a complete program listing is included on pages 38-39.

The program is designed to be flexible. In the initial man-machine interface (program lines 10-210), the number of frequencies must be entered. This will accommodate the engineer who is only interested in a single frequency (or a few discrete frequencies) as well as the engineer who wants to plot several points across a frequency band as a continuous plot (lines 31, 32, 70, 80, 7284-7290). Next each frequency of interest and its corresponding initial or load impedance (Appendix A, Figure A) is entered (lines 100-140).

After the frequencies and initial impedances are entered, they are printed on the printer (lines 150-165) as a permanent record. At this point a check for initial errors can be made before the program is further exercised. Next a decision is made as to whether or not the load impedance will be plotted (lines 170-200). The plotting subroutine will be described later.

The main part of the program is contained in lines 300 through 6240. In lines 300 through 420 a selection is made as to whether the next matching element is a series element, a shunt element, or a transformer. These further branch to lines 600 through 760 where more detailed descriptions of the network are given. For example, if a series element is selected (lines 330-350), is it an inductor, capacitor, tuned circuit, or length of transmission line (lines 600-650)? From here the program branches to the particular equation for that network (see lines 2000-6240 in Figure 2) as described in the "Basic Equations" section in Appendix A.

At this point let us examine a particular network in detail rather than continuing in general terms. Assume that a series capacitor will be our next element. From lines 330-350 and 600, 610, 630, we have arrived at line 5000, and at this point we must enter the value of capacitance in pico-farads. In lines 5020-5050 a new value (at this point only a "trial" value) of network impedance is calculated. It is assumed that we are designing a network and do not know whether or not this will be a good value or even a good choice of element type. We branch to line 7000 where we can choose to plot our trial result on the Smith Chart. At this point the "plot" will be only light "tic" marks. If this was a bad choice (lines 7040-7060) the trial value is

LINE NOS.	PROGRAM CONTENT	
10-210	FORMAT, INPUT FREQUENCIES & LOAD IMPEDANCE AT EACH.	
300-500	SELECTION OF SERIES, SHUNT, OR TRANSFORMER ELEMENTS.	
9000-9990	SELECTION OF PLOT OF CONSTANT "Q" LINES, PLOT CONSTANT "VSWR" CIRCLES, END.	
2000-2130	CALCULATE SHUNT INDUCTANCE	
2500-2630	CALCULATE SHUNT CAPACITANCE	
3000-3270	CALCULATE SHUNT TUNED CIRCUIT	
3500-3750	CALCULATE SHUNT LINE (OPEN OR SHORTED STUB)	
4000-4170	CALCULATE TRANSFORMER	
4500-4610	CALCULATE SERIES INDUCTANCE	
5000-5110	CALCULATE SERIES CAPACITANCE	
5500-5730	CALCULATE SERIES TUNED CIRCUIT	
6000-6240	CALCULATE SERIES TRANSMISSION LINE	
7000-8770	PRINT & PLOT SUBROUTINES	
	7000-7100	KEEP OR IGNORE TRIAL SECTION?
	7200-7330	UPDATE (CASCADE SECTIONS) IF TRIAL VALUE IS TO BE KEPT. PRINT NEW IMPEDANCE AND MAKE SOLID PLOT.
	8000-8170	INITIAL PLOTTER ADJUST
	8300-8396	SELECT NORMALIZED CHARACTERISTIC IMPEDANCE
	8400-8680	CONSTANT "Q" LINES
	8700-8770	TRIAL PLOT (NOT SOLID LINE)

Figure 2. Program Content By Major Sections.

ignored and the program branches back to line 300. If we want to keep the element chosen, the new impedance is now transferred permanently as the value we will next build upon (lines 7070-7100), the value of capacitance we chose as a series element is printed (line 5080), and the program branches to the subroutine in lines 7200 through 7330 where the impedance at each frequency is printed and plotted. The program then branches back to line 300 for the next element.

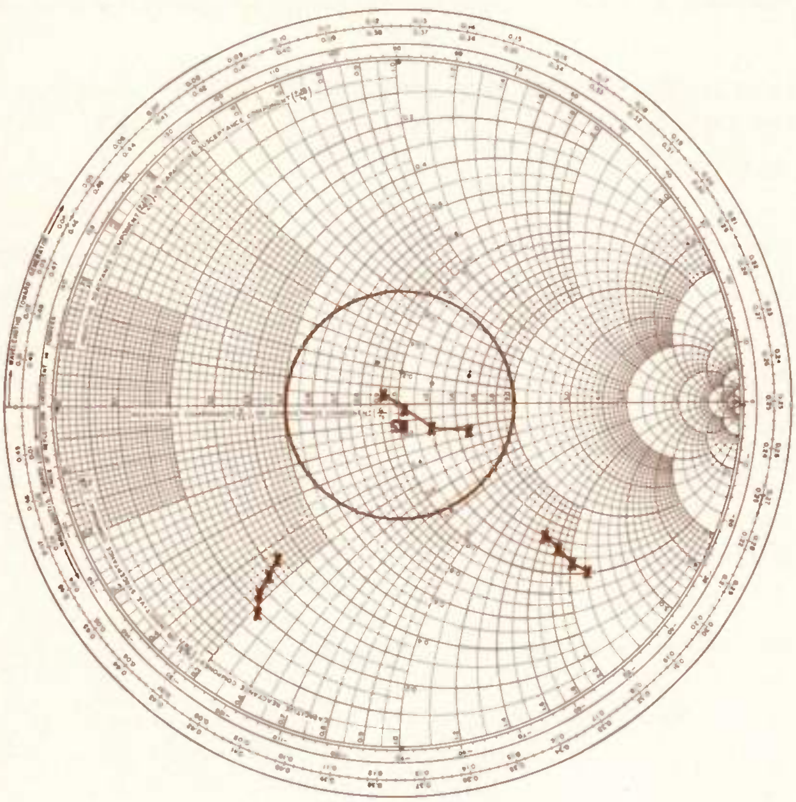
Plotter Interface

The Smith Chart is a *linear* plot of reflection coefficient ϱ; therefore, a linear X-Y plotter is used. For those readers who are not familiar with the transformation from impedance to reflection coefficient, the equations are developed in Appendix B at the end of this article.

Each time a plot is to be made, whether of initial load impedance (lines 170-200) or after a new impedance has been calculated (lines 7000-7030), a subroutine for initial plotter adjustment, lines 8000-8170, is called up. This permits the user to place a new Smith Chart on the plotter for each step of the design. Next the normalized characteristic impedance is selected, and the center of the Smith Chart is labeled if desired, in program lines 8300-8396.

If the plot is a new "trial" value, as discussed in the previous section, the plot is only "tic" marks at each frequency selected (lines 8700-8770). For the initial load impedance, and for each trial value which the user decides is "good," a large "X" is plotted for each frequency (lines 7260-7290), and a solid line is drawn between plotted points if a smooth plot was desired (lines 31-32).

At the end of the program, the user can have constant "Q" lines (program lines 9000-9040 and subroutine lines 8400-8680) or constant VSWR circles (program lines 9100-9180) plotted.



PROGRAM FOR IMPEDANCE MATCHING USING SMITH CHART

```
LOAD AT  7.000  HAS REAL PART 52.000  AND IMAJ PART  - 110.000
LOAD AT  7.100  HAS REAL PART 57.000  AND IMAJ PART  - 100.000
LOAD AT  7.200  HAS REAL PART 65.000  AND IMAJ PART  -  90.000
LOAD AT  7.300  HAS REAL PART 70.000  AND IMAJ PART  -  80.000
```

ADDED SERIES LINE 130.000 INCHES; Z0 = 50.000; VEL FACTOR = 0.670

```
NEW Z AT 7.000  HAS REAL PART  9.747  AND IMAJ PART  - 25.467
NEW Z AT 7.100  HAS REAL PART 11.601  AND IMAJ PART  - 23.891
NEW Z AT 7.200  HAS REAL PART 14.152  AND IMAJ PART  - 22.968
NEW Z AT 7.300  HAS REAL PART 16.492  AND IMAJ PART  - 21.892
```

ADDED SHUNT SHORTED STUB; LENGTH 100.000 INCHES; Z0 = 50,000 ;VP = 0.670

```
NEW Z AT 7.000  HAS REAL PART 74.465  AND IMAJ PART  - 11.641
NEW Z AT 7.100  HAS REAL PART 59.703  AND IMAJ PART  -  8.105
NEW Z AT 7.200  HAS REAL PART 51.398  AND IMAJ PART  -  1.240
NEW Z AT 7.300  HAS REAL PART 45.365  AND IMAJ PART     2.915
```

Example 1.

Example #1

After installing a 40-meter ham antenna and running a short coax into the operating room, the following impedances are measured:

7.0 MHz	52-j110 ohms
7.1 MHz	57-j100 ohms
7.2 MHz	65-j90 ohms
7.3 MHz	70-j80 ohms

We wish to match these impedances to within a 2:1 VSWR relative to 50 ohms using only lengths of 50-ohm coaxial cable (no discrete components). The following plot and printer outputs show the load, intermediate, and final impedance values. With a series line of 130 inches and a shunt shorted stub of 100 inches, we are well within a 2:1 VSWR.

Example #2

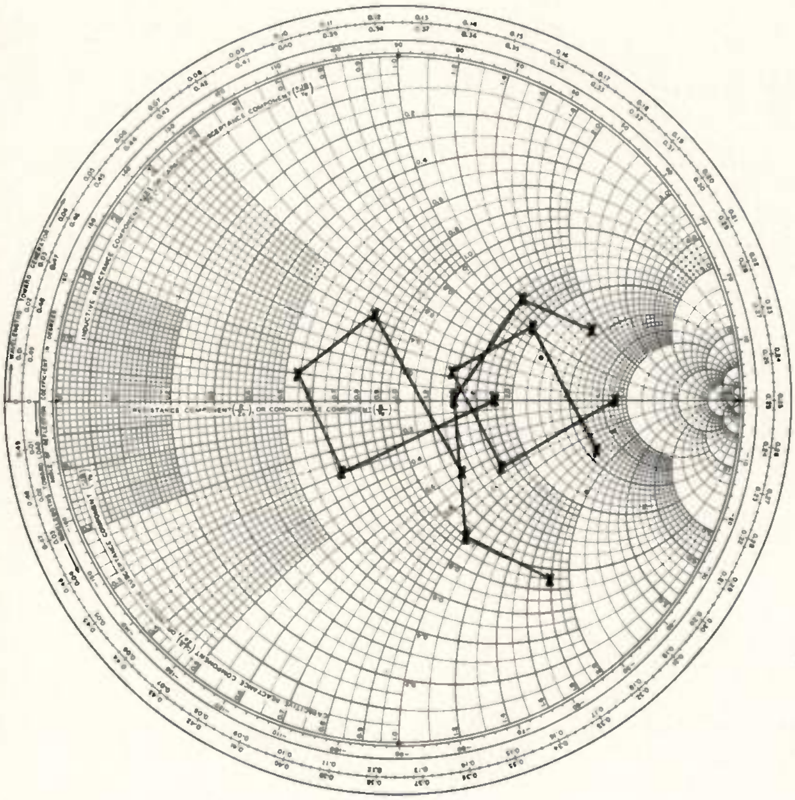
We need to match the following impedances to within a 2:1 VSWR of 50 ohms:

110 MHz	45-j90 ohms
120	50-j50
130	70 + jo
145	80 + j60
160	140 + j90

Although there are many approaches

we might take, experience tells us that if we place a shunt tuned circuit across the load which is resonant at the band center, we can "wrap" the endpoints into a smaller cluster without shifting the mid-point impedance. We will experiment with "trial" L-C values until we get the desired results, then we will transform the high impedances (centered around 100-150 ohms) to 50 ohms through a transformer. After several trial values, the results shown in Example 2(a) were obtained.

Since a broadband VHF transformer having a ratio of 0.4:1 might not be practical, we can "tap" the coil or use a split capacitor technique. The final values chosen are shown in Example 2(b), and the 2:1 VSWR objective has been met.



PROGRAM FOR IMPEDANCE MATCHING USING SMITH CHART

LOAD AT	110.000	HAS REAL PART	45.000	AND IMAJ PART	– 90.000
LOAD AT	120.000	HAS REAL PART	50.000	AND IMAJ PART	– 50.000
LOAD AT	130.000	HAS REAL PART	70.000	AND IMAJ PART	0.000
LOAD AT	145.000	HAS REAL PART	80.000	AND IMAJ PART	60.000
LOAD AT	160.000	HAS REAL PART	140.000	AND IMAJ PART	90.000

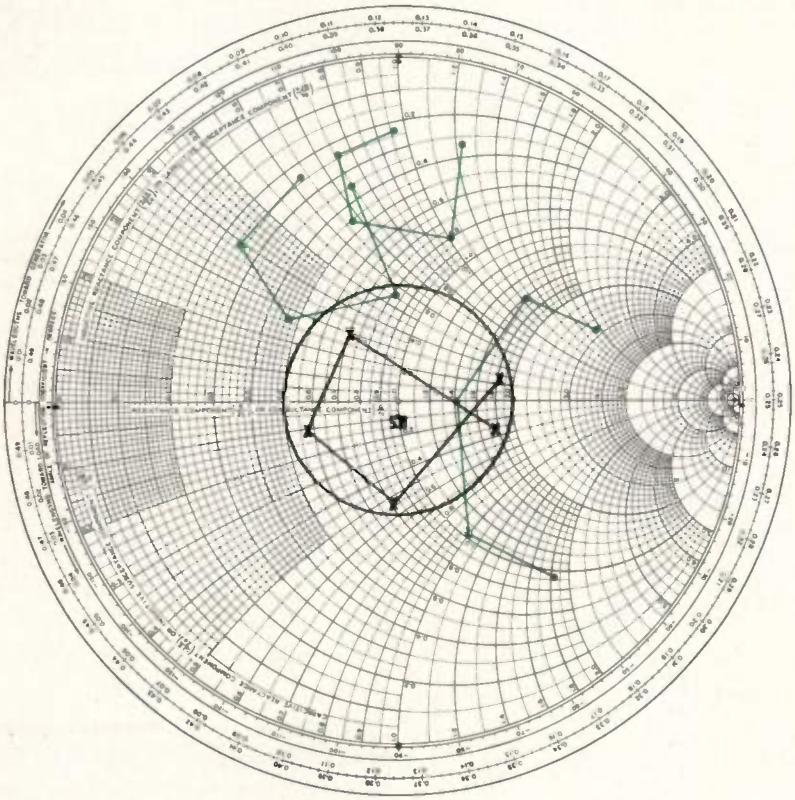
ADDED P 0.060 UH AND 22.000 PF IN SHUNT WITH NETWORK

NEW Z AT	110.000	HAS REAL PART	224.995	AND IMAJ PART	1.022
NEW Z AT	120.000	HAS REAL PART	83.267	AND IMAJ PART	– 37.327
NEW Z AT	130.000	HAS REAL PART	68.024	AND IMAJ PART	11.593
NEW Z AT	145.000	HAS REAL PART	97.484	AND IMAJ PART	51.792
NEW Z AT	160.000	HAS REAL PART	164.179	AND IMAJ PART	– 74.358

TRANSFORMED Z BY RATIO 0.400

NEW Z AT	110.000	HAS REAL PART	89.998	AND IMAJ PART	0.409
NEW Z AT	120.000	HAS REAL PART	33.307	AND IMAJ PART	– 14.931
NEW Z AT	130.000	HAS REAL PART	27.210	AND IMAJ PART	4.637
NEW Z AT	145.000	HAS REAL PART	38.993	AND IMAJ PART	20.717
NEW Z AT	160.000	HAS REAL PART	65.672	AND IMAJ PART	– 29.743

Example 2a.



Final value plotted is shown in black.

PROGRAM FOR IMPEDANCE MATCHING USING SMITH CHART

LOAD AT	110.000	HAS REAL PART	45.000	AND IMAJ PART	– 90.000
LOAD AT	120.000	HAS REAL PART	50.000	AND IMAJ PART	– 50.000
LOAD AT	130.000	HAS REAL PART	70.000	AND IMAJ PART	0.000
LOAD AT	145.000	HAS REAL PART	80.000	AND IMAJ PART	60.000
LOAD AT	160.000	HAS REAL PART	140.000	AND IMAJ PART	90.000

ADDED SHUNT L OF 0.060 UH

NEW Z AT	110.000	HAS REAL PART	17.667	AND IMAJ PART	60.522
NEW Z AT	120.000	HAS REAL PART	40.563	AND IMAJ PART	49.101
NEW Z AT	130.000	HAS REAL PART	23.026	AND IMAJ PART	32.888
NEW Z AT	145.000	HAS REAL PART	12.229	AND IMAJ PART	37.136
NEW Z AT	160.000	HAS REAL PART	12.072	AND IMAJ PART	47.357

ADDED SERIES C OF 60.000 PF

NEW Z AT	110.000	HAS REAL PART	17.667	AND IMAJ PART	36.408
NEW Z AT	120.000	HAS REAL PART	40.563	AND IMAJ PART	26.997
NEW Z AT	130.000	HAS REAL PART	23.026	AND IMAJ PART	12.483
NEW Z AT	145.000	HAS REAL PART	12.229	AND IMAJ PART	18.842
NEW Z AT	160.000	HAS REAL PART	12.072	AND IMAJ PART	30.779

ADDED SHUNT C OF 30.000 PF

NEW Z AT	110.000	HAS REAL PART	90.944	AND IMAJ PART	12.622
NEW Z AT	120.000	HAS REAL PART	40.831	AND IMAJ PART	– 26.883
NEW Z AT	130.000	HAS REAL PART	28.777	AND IMAJ PART	– 5.408
NEW Z AT	145.000	HAS REAL PART	35.247	AND IMAJ PART	14.559
NEW Z AT	160.000	HAS REAL PART	87.670	AND IMAJ PART	– 15.883

Example 2b.

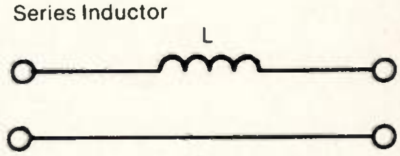
Appendix A

Basic Equations

A design frequently encountered in engineering is that of matching a given impedance to a second impe-

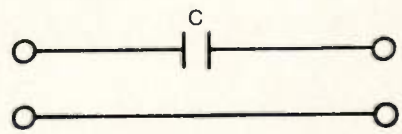
Series L-C Components

Series Inductor



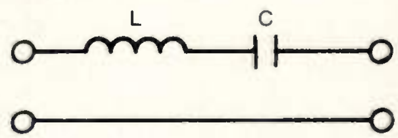
$$X = R$$
$$Y = 1 + 2\pi fL$$

Series Capacitor



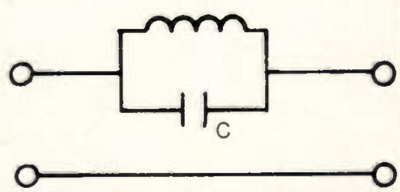
$$X = R$$
$$Y = 1 - \frac{1}{2\pi fC}$$

Series L-C in series



$$X = R$$
$$Y = 2\pi fL - \frac{1}{2\pi fC} + 1$$

Series L-C in parallel



$$X = R$$
$$Y = 1 + \frac{2\pi fL}{1 - (2\pi f)^2 LC}$$

Figure B.

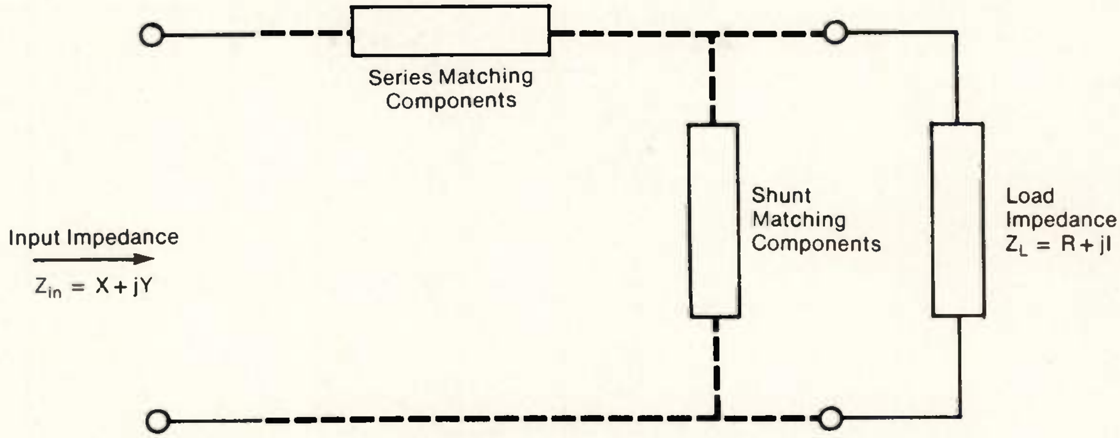


Figure A. Matching Network Configurations.

dance. An antenna design engineer may need to match his antenna to a 50-ohm transmission line. A circuit designer may need to match an amplifier stage to a crystal filter. The low reactive input impedance of a power amplifier stage may need to be transformed to a higher real value to be presented to the driver stage.

The circuit convention used in this article is shown in Figure A. The impedance which is to be transformed

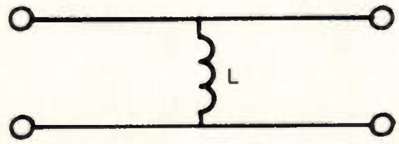
Shunt L-C Components

Shunt Inductor

$$W = 2\pi fL$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$

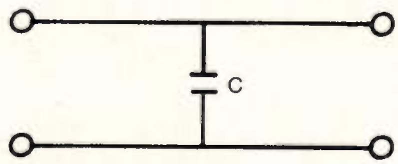


Shunt Capacitor

$$W = \frac{-1}{2\pi fC}$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$

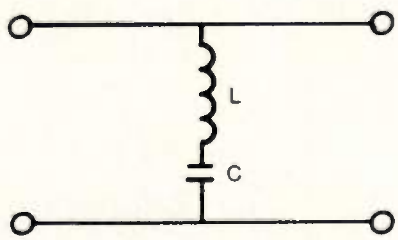


Shunt L-C in series

$$W = 2\pi fL - \frac{1}{2\pi fC}$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$

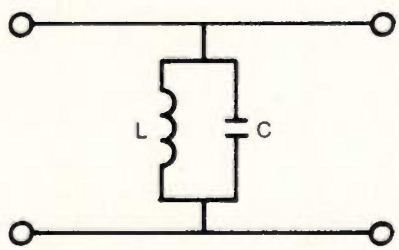


Shunt L-C in parallel

$$W = \frac{2\pi fL}{1 - (2\pi f)^2 LC}$$

$$X = \frac{RW^2}{R^2 + (I + W)^2}$$

$$Y = \frac{W(R^2 + I^2 + WI)}{R^2 + (I + W)^2}$$



Transformer

N = turns ratio
X = NR
Y = NI

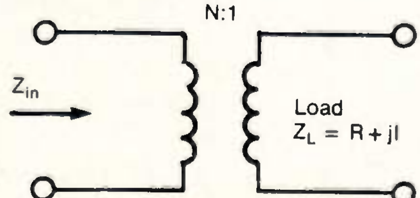
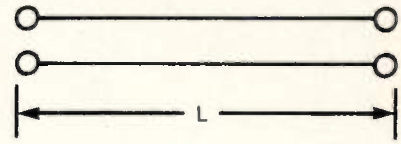


Figure C.

Series Transmission Line

$Z_{in} = X + jY$ $\quad$ $Z_L = R + jI$



Length = L (inches)
Characteristic Impedance = Z_1
Velocity Factor = V

$$W = \frac{1.2\,Lf}{39.37\,V}$$

$$D = \frac{2 Z_1 I}{(R + Z_1)^2 + I^2}$$

$$E = \frac{R^2 - Z_1^2 + I^2}{(R + Z_1)^2 + I^2}$$

$$Z = (R^2 + I^2)^{1/2}$$

$$T = \tan^{-1}(D/E) - 2W$$

Note: if calculator gives only 1st and 4th quadrant for $\tan^{-1}$ answers ($-90° < W < 90°$), add 180° to T if D was negative.

Figure D.

is given by $R + jI$. The transformation may be a shunt reactive element (or tuned circuit), a series element*, a transformer, or a length of transmission line. The resulting impedance is given as $X + jY$. Such impedance matching sections can easily be cascaded with the old $X + jY$ becoming the new $R + jI$ for the next network. For simplicity of calculations, lossless elements are assumed, but practical losses can be accounted for by calculating the associated series or shunt resistance and using the classical equations.

Equations for various matching sections, following the convention shown in Figure A, are shown in Figures B, C, and D.

The equations listed above for the various types of matching networks can be used individually on a small calculator, or they can be re-written in the reader's favorite computer language. In the program described in the following section, these equations, in BASIC, are found in lines 2000 through 6240 (see Figure 2).

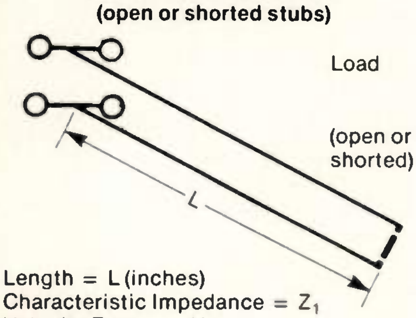
---

*Only single-ended or unbalanced circuits are discussed herein. Necessary modifications can be made to the equations if one works with balanced transmission lines.

$$X = \frac{Z_1\{1 - (Z\cos T)^2 - (Z\sin T)^2\}}{(1 - Z\cos T)^2 + (Z\sin T)^2}$$

$$Y = \frac{2 Z_1 Z\sin T}{(1 - Z\cos T)^2 + (Z\sin T)^2}$$

Shunt Transmission Line (open or shorted stubs)



Length = L (inches)
Characteristic Impedance = Z_1
Velocity Factor = V

$$W = \frac{1.2\ Lf}{39.37\ V}$$

If open stub: $T = Z_0 \tan(W + 90°)$
If shorted stub: $T = Z_0 \tan(W)$

$$X = \frac{RT^2}{R^2 + (I + T)^2}$$

$$Y = \frac{W(R^2 + I^2 + TI)}{R^2 + (I + T)^2}$$

Appendix B

The Smith Chart is nothing more than a plot of the reflection coefficient ϱ, defined as:

$$\varrho = \frac{Z - Z_0}{Z + Z_0}$$

where $Z = X + jY$ = the impedance to be plotted and Z_0 = characteristic impedance = value at center of chart. Note that the magnitude of ϱ can never exceed 1.0. Substituting $Z = X + jY$ into equation (A-1) above, we obtain:

$$\varrho = \frac{(X - Z_0) + jY}{(X + Z_0) + jY}$$

$$= \frac{(X + Z_0)(X - Z_0) + Y^2}{(X + Z_0)^2 + Y^2}$$

$$+ j \frac{2YZ_0}{(X + Z_0)^2 + Y^2}$$

Therefore,
x-axis value = real part of ϱ =

$$\varrho = \frac{X^2 - Z_0^2 + Y^2}{(X + Z_0)^2 + Y^2}$$

and y-axis value = imaginary part of

$$\varrho = \frac{2YZ_0}{(X + Z_0)^2 + Y^2}$$

Similarly, if x and y are the x-axis and y-axis coordinate values of ϱ, then solving equation (A-1) for Z, we find that:

$$Z = Z_0 \frac{1 + \varrho}{1 - \varrho}$$

and if $Z = R + jI$, then

$$R = Z_0 \frac{1 - X^2 - Y^2}{(1 - X)^2 + Y^2}$$

and

$$I = Z_0 \frac{2Y}{(1 - X)^2 + Y^2}$$

The equations above are the key to plotting on a Smith Chart overlay using a linear X-Y plotter. Equations as listed above are found in the program in lines 7220 through 7250 and 8700 through 8740.

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```
6210 PRINT "ADDED SERIES LINE"L"INCHES;  Z0="Z1"; 'VEL FACTOR="
6220 PRINT
6230 GOSUB 7200
6240 GOTO 300
7000 DISP "PLOT? (Y OR N)";
7010 INPUT A$
7020 IF A$="N" THEN 7040
7030 GOSUB 8000
7040 DISP "GOOD VALUE? (Y OR N)";
7050 INPUT A$
7060 IF A$="N" THEN 300
7070 FOR K=1 TO N
7080 R[K]=X[K]
7090 I[K]=Y[K]
7095 NEXT K
7100 RETURN
7200 FOR K=1 TO N
7210 PRINT "NEW Z AT"F[K]"HAS REAL PART"R[K]"AND IMAJ PART"I[K]
7220 D=(R[K]+Z0)↑2+I[K]↑2
7230 R=(R[K]↑2-Z0↑2+I[K]↑2)/D
7240 I=2*I[K]*Z0/D
7250 PLOT R,I
7260 CPLOT -0.3,-0.3
7270 LABEL (*)"X"
7280 IPLOT 0,0
7284 IF D$="Y" THEN 7290
7285 PEN
7290 NEXT K
7295 PEN
7300 PRINT
7310 PRINT
7320 PRINT
7330 RETURN
8000 DISP "PLOTTER ADJUST OK? (Y OR N)";
8010 INPUT A$
8020 IF A$="Y" THEN 8300
8030 SCALE -1,1,-1,1
8040 DISP "ADJUST LEFT HORIZ POS'N & CONT"
8050 PLOT -1,0,1
8050 STOP
8070 DISP "ADJUST RIGHT HORIZ SIZE & CONT"
8075 PLOT 1,0,1
8030 STOP
8090 DISP "ADJUST LOWER VERT POS'N & CONT"
8100 PLOT 0,-1,1
8110 STOP
8120 DISP "ADJUST UPPER VERT SIZE & CONT"
8130 PLOT 0,1,1
8140 STOP
8170 GOTO 8000
8300 Z0=50
8310 DISP "NORMALIZED TO 50 OHMS/(Y OR N)";
8320 INPUT A$
8330 IF A$="Y" THEN 8352
8340 DISP "WHAT Z0 DO YOU WANT?";
8350 INPUT Z0
8352 DISP "LABEL Z0? (Y OR N)";
8354 INPUT A$
8356 IF A$="N" THEN 8700
8360 PLOT 0,0,1
8365 STANDARD
8370 CPLOT -1.8,-1.5
8380 LABEL (*)Z0
8390 PEN
8395 FIXED 3
8396 GOTO 8700
8400 DISP "VALUE OF Q";
8440 INPUT Q
8450 GOSUB 8600
8460 B=Q/(1+SQR(1+Q↑2))
8470 PLOT 0,B,1
8480 CPLOT -2.8,0.25
8490 LABEL (8500)Q
8500 FORMAT "Q=",F5.1
8510 Q=-Q
8520 GOSUB 8600
8530 RETURN
8600 A=0.01/ABSQ
8610 D=A↑2*(1+Q↑2)+2*A+1
8620 R=(A↑2*(1+Q↑2)-1)/D
8630 I=2*A*Q/D
8640 PLOT R,I
8650 A=A+0.05*A
8660 IF A<100 THEN 8610
8670 PEN
8680 RETURN
8700 FOR K=1 TO N
8710 D=(X[K]+Z0)↑2+Y[K]↑2
8720 R=((X[K]-Z0)*(X[K]+Z0)+Y[K]↑2)/D
8730 I=2*Y[K]*Z0/D
8740 PLOT R,I
8742 IF E$="P" THEN 7260
8745 PEN
8750 NEXT K
8760 PEN
8770 RETURN
9000 DISP "CONST Q LINES? (Y OR N)";
9010 INPUT A$
9020 IF A$="N" THEN 9100
9030 GOSUB 8400
9040 GOTO 9000
9100 DISP "CONST VSWR CIRCLES? (Y OR N)";
9110 INPUT A$
9120 IF A$="N" THEN 9200
9130 DISP "VALUE OF SWR";
9140 INPUT S
9150 S=(S-1)/(S+1)
9160 FOR K=1 TO 370 STEP 5
9170 PLOT S*COS(K),S*SIN(K)
9180 NEXT K
9190 PEN
9200 PRINT
9980 STOP
9990 END
```

Program Listing

```
10 PRINT " ","PROGRAM FOR IMPEDANCE MATCHING USING SMITH CHART"
20 PRINT
30 PRINT
31 DISP "SMOOTH PLOT? (Y OR N)";
32 INPUT D$
40 SCALE -1,1,-1,1
50 DEG
60 DIM B$[7]
70 DISP "HOW MANY FREQUENCIES";
80 INPUT N
90 FOR K=1 TO N
100 DISP "INPUT FREQ"K"IN MHZ";
110 INPUT F[K]
120 DISP "INPUT RE, IM OF LOAD AT "F[K];
130 INPUT R[K],I[K]
131 X[K]=R[K]
132 Y[K]=I[K]
140 NEXT K
145 FIXED 3
150 FOR K=1 TO N
160 PRINT " ";"LOAD AT "F[K]"HAS REAL PART"R[K]"AND IMAJ PART "I[K]
165 NEXT K
166 PRINT
167 PRINT
168 PRINT
169 E$="P"
170 DISP "PLOT LOAD IMPEDANCE? (Y OR N)";
180 INPUT A$
190 IF A$="N" THEN 300
200 GOSUB 8000
210 E$="N"
300 DISP "IS NEXT ELEMENT SHUNT? (Y OR N)";
310 INPUT A$
320 IF A$="Y" THEN 700
330 DISP "IS NEXT ELEMENT SERIES? (Y OR N)";
340 INPUT A$
350 IF A$="Y" THEN 600
360 DISP "TRANSFORMER ? (Y OR N)";
370 INPUT A$
380 IF A$="Y" THEN 500
390 DISP "STOP ADDING SECTIONS? (Y OR N)";
400 INPUT A$
410 IF A$="Y" THEN 9000
420 GOTO 300
500 GOTO 4000
600 DISP "SERIES L, C, BOTH, OR LINE";
610 INPUT B$
620 IF B$="L" THEN 4500
630 IF B$="C" THEN 5000
640 IF B$="BOTH" THEN 5500
650 IF B$="LINE" THEN 6000
660 GOTO 600
700 DISP "SHUNT L, C, BOTH, OR LINE";
710 INPUT B$
720 IF B$="L" THEN 2000
730 IF B$="C" THEN 2500
740 IF B$="BOTH" THEN 3000
750 IF B$="LINE" THEN 3500
760 GOTO 700
2000 DISP "VALUE OF SHUNT L IN UH";
2010 INPUT L
2020 FOR K=1 TO N
2030 W=2*PI*F[K]*L
2040 D=R[K]↑2+(I[K]+W)↑2
2050 X[K]=R[K]*W↑2/D
2060 Y[K]=W*(R[K]↑2+I[K]↑2+W*I[K])/D
2070 NEXT K
2080 GOSUB 7000
2090 PRINT
2100 PRINT " ","ADDED SHUNT L OF "L"UH"
2110 PRINT
2120 GOSUB 7200
2130 GOTO 300
2500 DISP "VALUE OF SHUNT C IN PF";
2510 INPUT C
2520 FOR K=1 TO N
2530 W=2*PI*C*1E-06
2540 D=(1-W*F[K]*I[K])↑2+(R[K]*W*F[K])↑2
2550 X[K]=R[K]/D
2560 Y[K]=(I[K]*(1-W*F[K]*I[K])-R[K]↑2*W*F[K])/D
2570 NEXT K
2580 GOSUB 7000
2590 PRINT
2500 PRINT " ","ADDED SHUNT C OF  "C"PF"
2510 PRINT
2520 GOSUB 7200
2530 GOTO 300
3000 DISP "S FOR SERIES LC, P FOR PARALLEL";
3010 INPUT C$
3020 DISP "VALUE OF L IN UH";
3030 INPUT L
3040 DISP "VALUE OF C IN PF";
3050 INPUT C
3060 IF C$="S" THEN 3090
3070 IF C$="P" THEN 3160
3080 GOTO 3000
3090 FOR K=1 TO N
3100 W=2*PI*F[K]*L-(1E+06)/(2*PI*C*F[K])
3110 D=R[K]↑2+(I[K]+W)↑2
3120 X[K]=R[K]*W↑2/D
3130 Y[K]=(W*(R[K]↑2+I[K]↑2+W*I[K]))/D
3140 NEXT K
3150 GOTO 3220
3160 FOR K=1 TO N
3170 W=(2*PI*F[K]*L)/(1-((2*PI*F[K])↑2)*L*C*1E-06)
3180 D=R[K]↑2+(I[K]+W)↑2
3190 X[K]=R[K]*W↑2/D
3200 Y[K]=(W*(R[K]↑2+I[K]↑2+W*I[K]))/D
3210 NEXT K
3220 GOSUB 7000
3230 PRINT
3240 PRINT "ADDED  "C$;L"UH AND "C"PF IN SHUNT WITH NETWORK"
3250 PRINT
3260 GOSUB 7200
3270 GOTO 300
3500 DISP "LINE LENGTH IN INCHES";
3510 INPUT L
3520 DISP "Z0 OF SHUNT LINE";
3530 INPUT Z0
3540 DISP "VEL FACTOR OF SHUNT LINE";
3550 INPUT V
3560 DISP "OPEN OR SHORTED STUB";
3570 INPUT B$
3580 FOR K=1 TO N
3590 T=L*F[K]*1.2/39.37/V
3600 IF B$="OPEN" THEN 3650
3610 IF B$="SHORTED" THEN 3630
3620 GOTO 3560
3630 W=Z0*TAN(T)
3640 GOTO 3660
3650 W=Z0*TAN(T+90)
3660 D=R[K]↑2+(I[K]+W)↑2
3670 X[K]=R[K]*W↑2/D
3680 Y[K]=W*(R[K]↑2+I[K]↑2+W*I[K])/D
3690 NEXT K
3700 GOSUB 7000
3710 PRINT
3720 PRINT "ADDED SHUNT "B$" STUB;  LENGTH"L"INCHES; Z0="Z0";VP="V
3730 PRINT
3740 GOSUB 7200
3750 GOTO 300
4000 DISP "TRANSFORM UP OR DOWN? (U OR D)";
4010 INPUT A$
4020 DISP "RATIO =";
4030 INPUT W
4040 IF A$="U" THEN 4080
4050 IF A$="D" THEN 4070
4060 GOTO 4000
4070 W=1/W
4080 FOR K=1 TO N
4090 X[K]=W*R[K]
4100 Y[K]=W*I[K]
4110 NEXT K
4120 GOSUB 7000
4130 PRINT
4140 PRINT " ","TRANSFORMED Z BY RATIO"W
4150 PRINT
4160 GOSUB 7200
4170 GOTO 300
4500 DISP "VALUE OF SERIES L IN UH";
4510 INPUT L
4520 FOR K=1 TO N
4530 X[K]=R[K]
4540 Y[K]=I[K]+2*PI*F[K]*L
4550 NEXT K
4560 GOSUB 7000
4570 PRINT
4580 PRINT " ","ADDED SERIES L OF"L"UH"
4590 PRINT
4600 GOSUB 7200
4610 GOTO 300
5000 DISP "VALUE OF SERIES C IN PF";
5010 INPUT C
5020 FOR K=1 TO N
5030 X[K]=R[K]
5040 Y[K]=I[K]-1/(2*PI*F[K]*C*1E-06)
5050 NEXT K
5060 GOSUB 7000
5070 PRINT
5080 PRINT " ","ADDED SERIES C OF"C"PF"
5090 PRINT
5100 GOSUB 7200
5110 GOTO 300
5500 DISP "S FOR SERIES LC, P FOR PARALLEL";
5510 INPUT C$
5520 DISP "VALUE OF L IN UH";
5530 INPUT L
5540 DISP "VALUE OF C IN PF";
5550 INPUT C
5560 IF C$="S" THEN 5590
5570 IF C$="P" THEN 5640
5580 GOTO 5500
5590 FOR K=1 TO N
5600 X[K]=R[K]
5610 Y[K]=I[K]+2*PI*F[K]*L-1/(2*PI*F[K]*C*1E-06)
5620 NEXT K
5630 GOTO 5680
5640 FOR K=1 TO N
5650 X[K]=R[K]
5660 Y[K]=I[K]+(2*PI*F[K]*L)/(1-((2*PI*F[K])↑2)*L*C*1E-06)
5670 NEXT K
5680 GOSUB 7000
5690 PRINT
5700 PRINT "ADDED   "C$;L"UH AND"C"PF IN SERIES WITH NETWORK"
5710 PRINT
5720 GOSUB 7200
5730 GOTO 300
6000 DISP "SERIES LINE LENGTH IN INCHES";
6010 INPUT L
6020 DISP "SERIES LINE Z0";
6030 INPUT Z1
6040 DISP "SERIES LINE VELOCITY FACTOR";
6050 INPUT V
6060 FOR K=1 TO N
6070 T=L*1.2*F[K]/39.37/V
6080 D=(R[K]+Z1)↑2+I[K]↑2
6090 R=(R[K]↑2-Z1↑2+I[K]↑2)/D
6100 I=2*Z1*I[K]/D
6110 Z=SQR(R↑2+I↑2)
6120 T=ATN(I/(R+(1E-90)*(R=0)))-2*T+180*(R<0)
6130 R=Z*COS(T)
6140 I=Z*SIN(T)
6150 D=(1-R)↑2+I↑2
6160 X[K]=Z1*(1-R↑2-I↑2)/D
6170 Y[K]=Z1*2*I/D
6180 NEXT K
6190 GOSUB 7000
6200 PRINT
```

Wideband Sweep Oscillator Plug-Ins to 26.5 GHz

With the widest frequency coverage yet for a microwave sweeper — 10 MHz to 26.5 GHz — Hewlett-Packard Company offers its new model 83595A plug-in, the latest RF unit for HP's recently introduced 8350A microprocessor-based sweep oscillator family.

Two other new wideband units are also being announced: the HP 83594A covering 2 to 26.5 GHz and HP 83590A for 2 to 20 GHz. (Last fall HP introduced a 10 MHz to 20 GHz plug-in, model 83592A.)

The wide frequency ranges suggest broadband component testing as a major application area for these sweepers. In addition, however, the precision of performance (as seen in the frequency accuracy, frequency stability, frequency linearity, and calibrated power output flatness) and the full programmability of the sweeper/RF plug-in combination make this instrument valuable for integration into automatic test systems and for signal simulation and system testing applications.

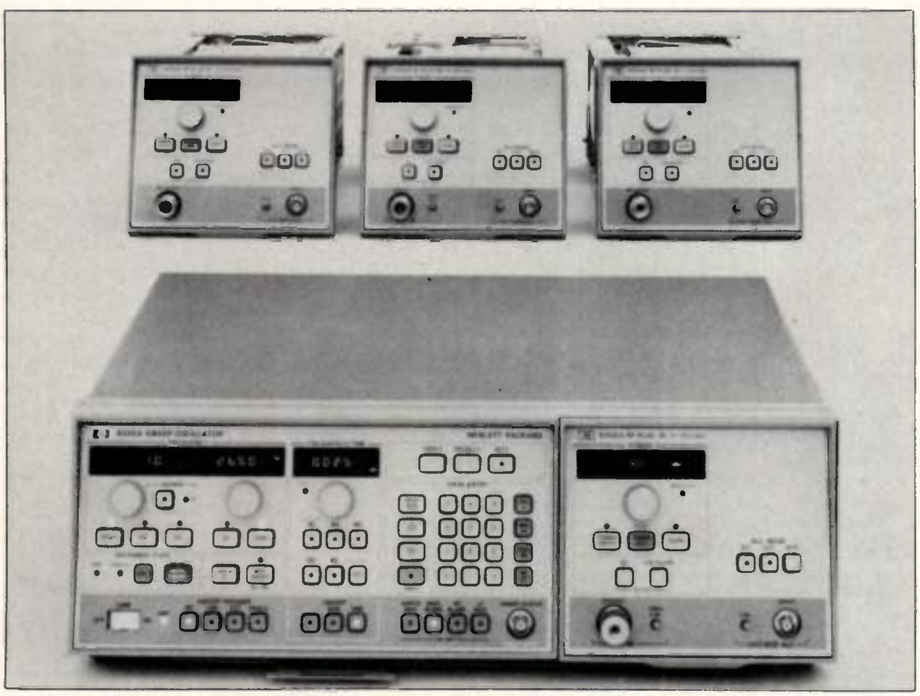
Its wide 10 MHz to 26.5 GHz frequency range is achieved by a combination of heterodyne and multiplication techniques using a single 2-7 GHz YIG oscillator.

HP 83595A Sweep Oscillator Plug-In

Output Band	Output freq. (GHz)	CW Freq. acc. (± MHz)	Swept Freq. acc. (± MHz)	CW residual FM (± kHz)	Lin. (± MHz)	Leveled Output (mW max)	Spurious Outputs Har.	Spurious (dBc) Nonhar.	Power variation (± dB)
0	0.01- 2.4	5	15	8	2	10	25	25	0.9
1	2.3 - 7.0	5	20	8	2	10	25	50	0.7
2	6.9 -13.5	10	25	15	4	10	25	50	0.7
3	13.4 -20.0	15	30	15	6	10	25	50	0.8
4	19.9 -26.5	20	35	20	8	4	20	50	0.9
full sweep	0.01-26.5	—	50		10	2.5	20	50	1.0

HP Model No.	Frequency Range	Domestic U.S.A. Price	Availability (Weeks)
83595A	0.01-26.5 GHz	$27,000	16
83594A	2 -26.5 GHz	$22,000	16
83590A	2 -20 GHz	$19,000	12

Mixing the YIG oscillator output with a cavity oscillator obtains the 10 MHz to 2.4 GHz low band. Amplifying, multiplying (by 1, 2, 3, or 4) and filtering the proper harmonic obtains the higher frequency bands (2.4-7 GHz, 13.5-20 GHz, and 20-26.5 GHz). Use of a single YIG oscillator results in outstanding frequency accuracy (better than ±20 MHz at 26.5 GHz), and residual FM less than 20 kHz peak at 26.5 GHz, as well as allowing tight control of frequency accuracy and overlap at band switch points.



Output power can be leveled to +10 dBm for frequencies up to 18.6 GHz and to +4 dBm for frequencies up to 26.5 GHz. Flatness is ±1 dB over the full band, and power is calibrated with 0.1 dB resolution and better than ±2.0 dB accuracy down to −5 dBm on all plug-ins. An optional internal step attentuator extends the calibrated range to −80 dBm for plug-ins operating to 20 GHz and −60 dBm for plug-ins operating to 26.5 GHz.

Specification Summary

Major specifications of the HP 83595A 10 MHz to 26.5 GHz plug-in are summarized in the table above.

Specifications for the other new wideband plug-ins are comparable over their relevant frequency ranges.

Price and Availability

Domestic U.S.A. prices and availability are summarized above.

Internal step attentuator, Option 002, add $900. (The HP 8350A Sweeper mainframe price is $4,250.) Circle INFO/CARD #138.

Solid State H.F. Amplifiers

Instruments For Industry, Inc. (IFI), recently introduced its new 1500 series of broadband amplifiers. These Class A solid state amplifiers incorporate IFI's proprietary broadband techniques to provide instantaneous bandwidth over a frequency range of 10 kHz to 20 MHz. Rated for between 5 and 130 watts of continuous RF output, the amplifiers are driven from a standard signal generator.

Self-contained, the IFI 1500 series of amplifiers are unconditionally stable with full protection against damage to internal circuitry and power supply. They combine high quality construction with the latest state-of-the-art design features for high reliability and performance. IFI's 1500s need no tuning or adjustment.

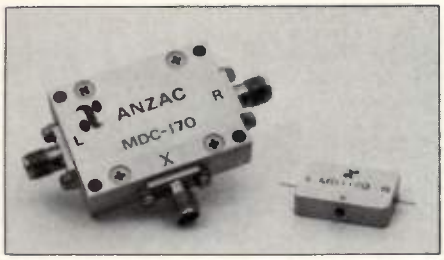
This new line of IFI amplifiers are designed for a variety of applications including communications, laser modulation, ultrasonics and general laboratory testing. Fully compatible with other IFI test equipment, these units range in size from 11" wide x 13" deep x 7" high to 19" wide x 17" deep x 7" high. Prices for these units start at $1250 and delivery is 30 days ARO.

For additional information on IFI's entire line of quality amplifiers and test equipment, contact: Ronald Richards, Instruments For Industry, Inc., 151 Toledo Street, Farmingdale, NY 11735, (516) 694-1414. INFO/CARD #137.

0.5-18 GHz Mixer Features + 12 dB Compression Point

Anzac Division of Adams-Russell Company has developed a 0.5-18 GHz Mixer for high level mixing or upconverter use. The MD-170 features a + 12 dBm 1 dB compression point and an 8 dB typical conversion loss. The mixer also has an IF bandwidth of 2-5000 MHz and a port-to-port isolation of better than 20 dB. These features make the MD-170 ideally suited for high dynamic range receiver applications and high level upconverters needed in modern EW systems.

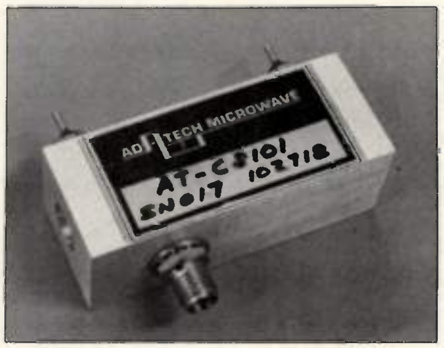
The MD-170, like all of Anzac's



microwave mixers, is designed for military environments and comes in a hermetic drop in module or a connectorized housing. Pricing in small quantity is $565 with delivery from stock. INFO/CARD #135.

Voltage Controlled Oscillators

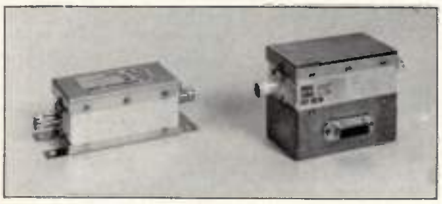
The Series C8000 voltage controlled oscillators from Ad-Tech Microwave, Inc. provide center frequencies between 8 and 13 GHz with an electronic tuning range of six percent minimum. Size is just over 1.0 cubic inches and output power is +10 dBm min. over a temperature range of −20 degrees C to 65 degrees C. DC input voltage is between −5 and −18 volts, depending upon frequency, with tuning voltage typically 0 to −25 VDC. Spurious outputs are −60 dBc maximum and phase noise is −85 dBc/Hz maximum 100 kHz from the carrier. Price for



small quantities begins at $1400. R.C. Havens, Ad-Tech Microwave, Inc., 7755 E. Redfield Rd., Scottsdale, AZ 85260, (602) 998-1584. INFO/CARD #134.

Logic Adapters

Kay introduces two low cost Logic Adapters (models 4000 and 4001) to interface their line of programmable attenuators directly to TTL circuits. These new Logic Adapters mount to the bottom of the attenuators. Inter-



connection to the 5 or 8 bit drive circuits is through a sub-miniature connector.

Both models are $150.00 in single piece quantities.

Contact Stephen Crump, sales manager, (201) 227-2000 for more information. Kay Elemetrics Corp, 12 Maple Avenue, Pine Brook, NJ 07058. Circle INFO/CARD #133.

Portable Spectrum Analyzer

Texscan Corporation introduces its new portable AL-57 Spectrum Analyzer with 0.1 to 1600 MHz coverage.

The AL-57 is configured with an on-board computer used to assist the

operator in controlling functions with ease. Single touch, audible feedback membrane switches activate such functions as dispersion, resolution, scan mode and markers. A high visability CRT, digital LCD readouts and LED function groups provide fast visual feedback for confident operation.

This analyzer does not compromise laboratory performance for field operation with an amplitude measurement range of +30 dBm to −120 dBm and noise sideband performance of −60 dB 10 kHz away. INFO/CARD #132.

A Power Blocked Mini-Sized Step Attenuator

Comsonics, Inc. of Harrisonburg, Va. announces the introduction of its 75 ohm model Mini-Sized Step Attenuator, with an attenuation range of 0-102.5 dB.

This 100 percent shielded unit allows control of insertion loss padding by steps in .5 dB increments, the control being provided by rugged RF High Frequency switches. Type F connectors standard, its Mini size is perfect for fulfilling .1-1000 MHz testing needs, whether service bench or laboratory set up. Other connectors available by request.

Features of this model are — RF gasket on the base to prevent signal leakage — Power Blocked — low reactive, low loss toggle switches.

A very compactly designed unit,

the 9 step model is 4.875" L (case) x 2.698" W x 7/8" D.

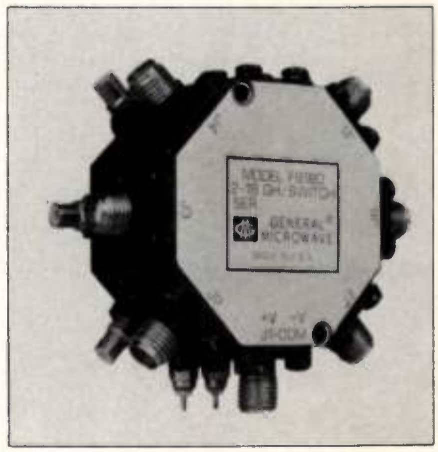
Contact Sales Department, ComSonics, Inc., Harrisonburg, VA, (800) 336-9681. INFO/CARD #131.

High Speed Terminated Multi-Throw Pin Diode Switches

From General Microwave Corporation, Farmingdale, NY; series 91HT multi-throw switches combine high speed performance, a broadband design, 1-18 GHz, and feature built-in terminations so that the off-ports are also non-reflective.

The response time, i.e., from 50 percent TTL-command to the 90 percent point of the RF pulse (going on) or 10 percent of the RF pulse (going off) is less than 35 nanoseconds; including the rise (and fall) time of 10 nanoseconds, maximum.

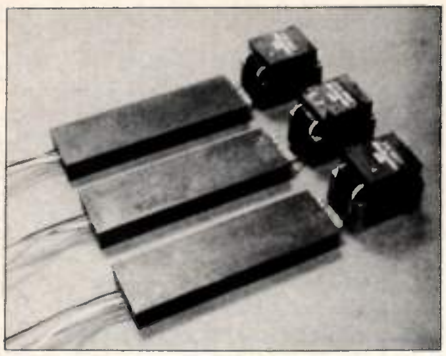
The units feature low VSWR in either the on or off ports and low



insertion loss. Rated isolation is 60 dB to 12.4 GHz, 50 dB to 18 GHz. They are available as standard items in SP2T to SP4T configurations and as options from SP5T through SP7T. Please circle INFO/CARD #130.

Miniature Power Supply For FET Amplifiers

The Microwave Components and Subsystems Division of Varian Associates has announced the development of a new miniature power supply to operate FET amplifiers. Cast in rigid thermal epoxy, the unit is designed to withstand shock, vibration, and



altitude, and can be customized to meet other environmental requirements.

The new power supply operates from a prime power of 115 VAC, 57-420 Hz, single phase, and requires 20 watts of input power. When loaded with an impedance so that output current is between 0.25 and 0.70 amperes, the unit will provide 4.2 to 6.5 VDC (±1 percent regulation) with less than 5 mV of output ripple (peak-to-peak). This output is provided at both high and low line over the frequency range, and over a baseplate temperature range of −30 degrees C to +85 degrees C. Maximum output power is 5 watts. Other output-voltage and input-power ranges are available upon request.

The power supply has a maximum length of eight inches, maximum width of 1.6 inches, and a height of 0.7 inches, except that the input transformer located at the input end of the supply is 1.7 inches high with a depth of 2.2 inches. In volume production, the unit can be delivered at a minimum rate of 50 per month.

For further information, literature is available from Varian Associates, Microwave Components and Subsystems Division, 3200 Patrick Henry Drive, Santa Clara, CA 95054, (408) 496-6273. INFO/CARD #129.

High Power, Coaxial Lowpass Filters

Cir-Q-Tel, Inc. is offering a new series of high power, coaxial lowpass filters that can handle up to 10 KW average input power. Designated as the HPL/KW Series, they feature an efficient 13-pole Zolotarev design for maximum performance and are available with cut off frequencies from 30 MHz to 450 MHz (> 2 GHz on lower power versions).

VSWR is ≤1.2:1 to 1.35:1 up to fc

and typical response for Model A versions with a signal passband of 0.6 to 1.0 maximum free signal passband (fsp) is 5:1; Model B with a signal passband of 0.2 to 1 fsp maximum is 5:1. Passband insertion loss is <0.3 dB (fsp) and spurious levels are typically >60 to 80 dB greater than 2.4 GHz. Average input power rating is: up to 10 KW (continuous) into a matched load; up to 5 KW (continuous) with a 3:1 VSWR load.

These filters are supplied with Type SC, HN or N right-angle connectors. Other models available with standard rigid flanges. Standard impedance is 50 ohms and nominal dimensions are 4.5" x 2.5" x 18" for models in the 30 to 35 MHz range. Hard mounting to heat sink via six 1/4/28 studs or 1/4/28 tapped holes.

For additional information contact the Sales Department, Cir-Q-Tel, Inc., 10504 Wheatley Street, Kensington, MD 20895-2695, (301) 946-1800, TWX: 710-828-0521. INFO/CARD #128.

Low Cost GaAs FETs

California Eastern Laboratories, Inc., introduces NEC's new NE700 and

NE720 GaAs FETs. These devices are designed and manufactured to provide excellent performance, reliability and uniformity at a substantially lower cost than previous GaAs FETs.

The NE70083, a 0.5 micron recessed gate device in a rugged hermetically sealed package features low noise figures of less than 1.0 dB at 4 GHz, 2.3 dB at 12 GHz and high associated gains through 18 GHz. Prices start at $55 (1-9).

The NE72089 is a 1.0 micron recessed gate GaAs FET which offers low noise figures of 1.3 dB at 4 GHz and 2.1 dB at 6 GHz. The device also features high gain through 8 GHz. The low price of $15 (1-9) puts this FET in competition with many bipolar transistors with less performance.

These devices are also available in chip form. The chips' gates and channels are glassivated with a thin layer of SiO_2 for mechanical protection.

The performance vs. price trade-off offered by these two new FETs will allow many new projects to move from the conception stage into production.

For further information, contact California Eastern Laboratories, Inc., 3005 Democracy Way, Santa Clara, CA 95050, (408) 988-3500. Please circle INFO/CARD #126.

Broadband, General Purpose RF Amplifier

Providing a typical gain of 13 dB in the 20 to 900 MHz range, TRW Semiconductors' LNA1100 exhibits a noise figure of only 3.0 dB at 900 MHz.

The LNA1100 is self-contained hybrid, general purpose amplifier with all bias, matching and decoupling components in a TO-8 package. Among its other features are: typical gain flatness of ±0.5 dB, maximum input and output VSWR of 2:1 in a 50Ω system, and typical power output at 1 dB compression of 6 dBm.

Typical applications for the LNA1100 include data communication systems, fiber-optic communications, radio relay, digital pre-scalers, signal processing, buffer amplifiers, instrumentation and IF amplifiers.

Price of the LNA1100 is $65 each (domestic US) in quantities of 1,000. Delivery is off-the-shelf. For further information, contact TRW Semiconductors, 14520 Aviation Blvd., Lawndale, CA 90260, (213) 679-4561, TWX: 910-325-6206, Telex: 67-7148. Circle INFO/CARD #125.

Safety Resistors For Circuit Protection

A new line of safety resistors to be used in series with circuit elements requiring short-circuit protection, is now available from Murata Erie North America, Inc. These new devices are

asically current/sensitive resistors with a very sharp increase in resistance at a specific current. This increase in resistance limits current thus protecting the circuit. When current levels return to normal, these devices automatically return to their low resistance state.

Ambient temperature resistance values range from 3.3 ohms to 220 ohms with trip currents ranging from 65 mA to 520 mA. All safety resistors in this line are rated 24 VDC with maximum currents from 1 to 2 A. Overall dimensions are .380 D x .200 T.

To receive complete technical information, write to Murata Erie North America, Inc., 1148 Franklin Road, S.E. Marietta, GA 30067, (404) 952-9777. INFO/CARD #124.

Economical Microwave Sweep Generator

Wavetek Indiana, announces a new generation of economical, easy-to-use, digital display generators. Wavetek has incorporated a number of innovations into this compact-sized unit.

The model 1084 has three operating modes: CW, ΔF and Full Sweep. Frequency in the CW mode is set by a 10-turn potentiometer and displayed with a resolution of 1 MHz on a three and a half digit display. In the ΔF mode, center frequency is selected by the 10-turn potentiometer; the sweep width range of 500 kHz to 1000 kHz is controlled by a 100 MHz/Step selector

and a 100 MHz vernier. The model 1084 features one percent display linearity.

In the Full Sweep mode, the start frequency is 3.5 GHz and the stop frequency is 4.5 GHz. The 10-turn potentiometer and three and a half digit frequency display operate as a variable marker. The marker produces a bright spot on the display by momentarily delaying the sweep ramp for approximately 2 msec. Accuracy is ±10 MHz. External marker input is standard.

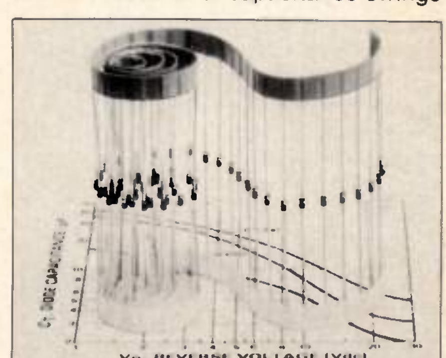
A front panel switch controls the birdy by-pass marker system and provides selection of harmonic markers at 1, 10, and 100 MHz. Differences in marker amplitudes make identification of markers easy.

The model 1084 has an output power range of +13 to −60 dBm. This unit also features a dB/Step attenuator and an 11 dB vernier for continuous adjustment of output. Output level is displayed on a 3 digit readout with a 0.1 dB resolution.

Price is $2795 (Domestic U.S.) and delivery takes 90 days A.R.O. For further information contact Wavetek Indiana, Inc., 5808 Churchman, P.O. Box 190, Beech Grove, IN 46107, (317) 787-3332, TWX: 810/341-3226. Circle INFO/CARD #123.

Low Cost Hyperabrupt Tuning Diode Family

A new family of low cost, hyperabrupt tuning diodes designed for use in military and commercial VHF-UHF communications circuits has been announced by Frequency Sources Semiconductor Division. Types KV 3201, 3901 and 3902 offer capacitance swings



as high as 8 to 1 from 3 to 25 volts and designers can select from 3 volt capacitance values of 11, 25 or 29 pF from the economical high Q, hermetic glass DO-34 package series.

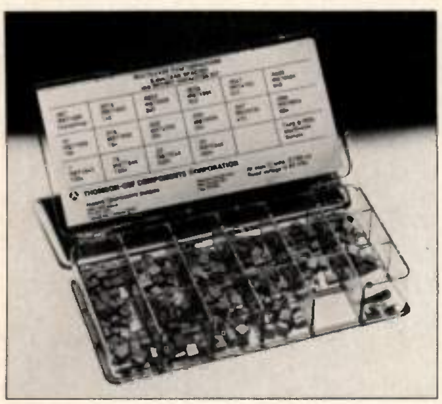
These low inductance devices have Q values up to 400 at 50 MHz and can be used up to 1 GHz in voltage control oscillators and filters. All types are available taped and reeled for automatic insertion which is not the case with the non-hermetic BB 105 and MV 109 types this family replaces.

Delivery in production quantities is 60 days with 100 to 999 pricing ranging from .64 cents to .70 cents each.

Contact Frequency Sources, Semiconductor Division, 16 Maple Road, Chelmsford, MA 01803, (617) 256-8101, TWX: (710) 343-0101. INFO/CARD #122.

IRD Multilayer Film Capacitor Sample Kit

Thomson-CSF Components Corporation, Passive Components Division, has announced that it is now selling IRD multilayer film capacitor sample kits containing a full complement of IRD capacitor values. The kits — a $55 value — sell for $25 and include 25 pieces each of 16 different values



ranging from .001 mFd to .33 mFd. In addition, the kit offers a tape and reel mechanical sample. All caps in the kit have 5 mm lead spacing, with 10 percent tolerance and 63 VDC.

For more information or to purchase a sample kit, contact: Thomson-CSF Components Corporation, 6660 Variel Avenue, Canoga Park, CA 91303, (213) 887-1010. INFO/CARD #127.

High-Intercept HF Amplifiers

Watkins-Johnson Limited has introduced two new high-intercept amplifiers for use in the 0.5 to 30 MHz frequency range. The amplifiers use VMOS field-effect transistors to achieve exceptionally wide dynamic range, high power-handling capability and excellent linearity.

Designated the WJ-7033-5 and WJ-7033-6, both amplifiers feature +43 dBm third order and +80 dBm second order output intercept points with +34 dBm minimum power output at 1 dB gain compression.

The WJ-7033-5 offers 13.5 dB minimum gain with a noise figure of 6.5 dB, maximum; the WJ-7033-6 provides 9.5 dB minimum gain with a maximum noise figure of 7.5 dB. Maximum gain flatness for both models is ± 1 dB.

The amplifiers are intended for use as low-noise, wideband preamplifiers, producing minimum intermodulation distortion in the presence of high levels of received noise. All units are supplied with protection against high input voltages and standard heat-sinking is adequate for operation in ambient temperatures up to 35 degrees C. Information on the WJ-7033-5 and WJ-7033-6 is available from Watkins-Johnson Limited, Dedworth Road, Oakley Green, Windsor, Berkshire SL4 4LH, England, (075-35) 69241, Telex: 847578. INFO/CARD #120.

5 Watt DC/DC Converters

The DCE Series of DC/DC conver-

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ters are a full line of compact modular power supplies which deliver 5 watts of power to a broad range of digital and analog applications.

The DCE Series is available in 23 model types. Inputs of 5, 12, 24, 28, and 48 VDC convert to single outputs of +5, +12, +15 VDC and dual outputs of ±12 and ±15 VDC. Current ratings, which depend on output voltage, range from 1 amp for the +5 VDC converters, to 150 mA for the dual output ±15 VDC units. All outputs are isolated from the input(s) to provide floating power which permits complete isolation of the output from the DC bus. Because the outputs are isolated, a positive or negative supply can be configured by the user. Each converter is packaged in a shielded case virtually eliminating radiated RFI/EMI.

The DCE Series offers accurate 0.02 percent error regulated output with very low noise, 1 mV RMS. Not only is the output noise extremely low, but a high attenuation input filter reduces kickback spikes, noise and reflected ripple caused by inverter switching, thus protecting other loads across the same DC power bus.

The operating temperature range is −25 degrees C to +71 degrees C for full load operation and −25 degrees C to +85 degrees C for 50 percent load. The module size is 2 x 2 x 0.38 inches (51.3 x 51.3 x 9.65 mm).

The price of the DCE Series is $87.00 (1-9 pcs.) each for single output units; and $73.00 (1-9 pcs.) each for dual output units. Delivery is stock to six weeks. INFO/CARD #119.

Medium Power Solid State Switch

A new SP2T solid state 20 watt CW switch is now available covering 20-500 MHz. Daico Industries Part Number 100C1592 contains an internal single line TTL driver.

Switching speed including driver delay is 10 Microseconds maximum. DC power required is +5 volts at 220 mA, −15 to −30 volts at 20 mA.

Other key specifications include: Isolation 65 dB minimum (20-300 MHz), 55 dB minimum (300-500 MHz). Insertion Loss 0.5 dB maximum (20-300 MHz), 0.75 dB maximum (300-500 MHz). VSWR 1.2/1 maximum (20-300 MHz), 1.25/1 maximum (300-500 MHz).

For complete specifications and information, contact Daico Industries, Inc., 2351 East Del Amo Blvd., Compton, CA 90220, (213) 631-1143. Please circle INFO/CARD #121.

Panel Mount Adapters For Rectangular Trimmers

Spectrol is now offering two versions of Panel Mount Adapters for the popular 3/4-inch Spectrol model 43 and the 1 1/4-inch Spectrol model 70 cermet trimming potentiometers. These new model 6 Panel Mount Adapters are designed to slip over standard, rectangular trimmers, converting them in seconds to rugged panel mounts. Both versions of model 6 Panel Mount Adapters are offered with shaft options including flush and extended shafts. These adapters are now available off the shelf from local authorized Spectrol distributors. For information, contact: Spectrol Electronics Corporation, P.O. Box 1220, City of Industry, CA 91749, (213) 964-6565. INFO/CARD #111.

Advertiser Index

New Literature

Programmable FM/AM Signal Generator Brochure

A new eight-page brochure describes the many unique features of the Boonton models 1020 and 1021 Programmable FM/AM Signal Generators, including outstanding FM and AM modulation specifications, low distortion, programmable modulation source, high speed sweep capability with intensity marker, and internal store and recall operation. Contact Scott Elkins, V-P Marketing, Boonton Electronics Corporation, P.O. Box 122, Parsippany, NJ 07054, (201) 887-5110. INFO/CARD #103.

RF Switching Catalog

Do you need your signals routed? A wide variety of switches and matrices which can be readily customized to meet your requirements are described in a new illustrated 10-page catalog issued by MarLee Company of Upland, Calif. Designed to provide the ultimate in versatility and capability, the MarLee coax switches, matrices and systems cover a broad frequency range DC to 100 MHz.

Complete turnkey switching systems are also available. These are made to customer specifications and can include computer compatible automatic remote control devices, lighted pushbutton controls, power supplies and interconnect cabling. Solid state control logic can be furnished with any unit. INFO/CARD #136.

Resistor Cross Reference Guide

RCD Components is offering a free six-page cross reference guide on resistor products. The guide includes 28 different manufacturers, both domestic and foreign, and covers over 100 different types of wirewound, metal film, carbon film, high voltage, and power oxide resistors, as well as resistor networks. The guide also references MIL part numbers and power dissipation. Copies can be obtained by writing to RCD Components Inc., 330 Bedford St., Manchester, NH 03101. Circle INFO/CARD #108.

New RF Amplifier Catalog

A new short form catalog issued by Aydin Vector, Newtown, Pa., describes hybrid RF amplifiers and cascaded amplifier assemblies that cover a frequency range of 1 to 700 MHz as well as voltage controlled attenuators from 1.0 MHz to 1.0 GHz. All listed models are screened to MIL-STD-883B and are shipped from stock. The catalog also tells about the company's capabilities to design and manufacture custom RF amplifiers to meet special customer requirements.

Complete specifications are provided such as frequency range, gain, noise figure, input and output power, IMP3 and VSWR. The single and multistage amplifiers are housed in TO-8, TO-12 or 4-pin DIP packages. Voltage controlled attenuators are in TO-8 (0-20 dB) and 14-pin DIP (0-40 dB) packages. Dimensions are also included.

For a copy of the new catalog, write Aydin Vector Division, Dept. C, P.O. Box 328, Newtown, PA 18940, (215) 968-4271. INFO/CARD #107.

Designing Lossless Feedback Amplifiers

Anzac Division of Adams-Russell has prepared an article on the application of lossless feedback technology to RF amplifiers for improved noise figure and output power performance. The brochure details design approach and expected performance improvements. The brochure also includes a list of Anzac's lossless feedback products. INFO/CARD #106.

Eliminate Terrestrial Interference

This 20-page catalog (MTV/82) tells how to eliminate terrestrial interference on earth stations, and lists a complete line of filters designed and tested for this purpose. The origins of each type of interference are explained, symptoms are given and specific filters are recommended. Products include fixed and tunable noise bandpass filters. IF traps, waveguide adapters, power dividers, and coax adapters. MTV/82 is designed for effective use by novice and experienced earth station operators alike.

Microwave Filter Co., Inc., 6743 Kinne St., East Syracuse, NY 13057, (800) 448-1666. INFO/CARD #105.

Marine Radar Products Brochure

A four-page brochure describing Varian's marine radar products has been published by the Beverly Division.

The product line includes marine radar magnetrons, TR receiver protectors, and TR limiters.

Varian Authorized Distributors maintain a shelf inventory of the marine products to assure immediate availability. For copies, write to Varian/Beverly Division, Salem Road, Beverly, MA 01915. INFO/CARD #104. □

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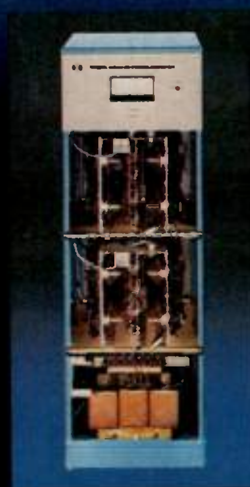

And mismatched loads can't cause problems because, like every ENI amplifier, the A-1000 is unconditionally stable and protected against both overload and overdrive.

For more information, or a full-line catalog, please contact us at ENI, 3000 Winton Road South, Rochester, NY 14623. Call 716/473-6900, or telex 97-8283 ENI ROC.

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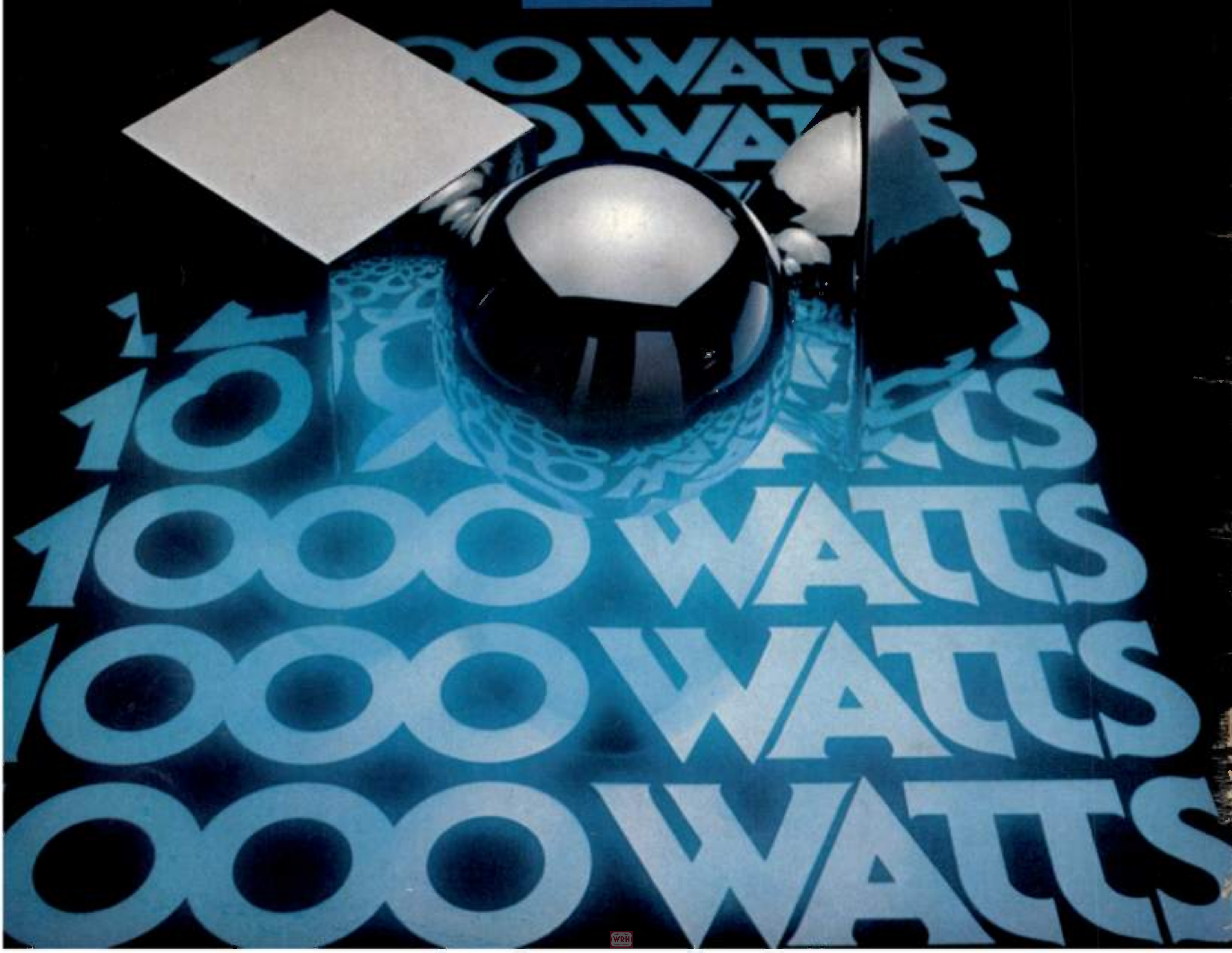
INFO/CARD 24



May/June 1982

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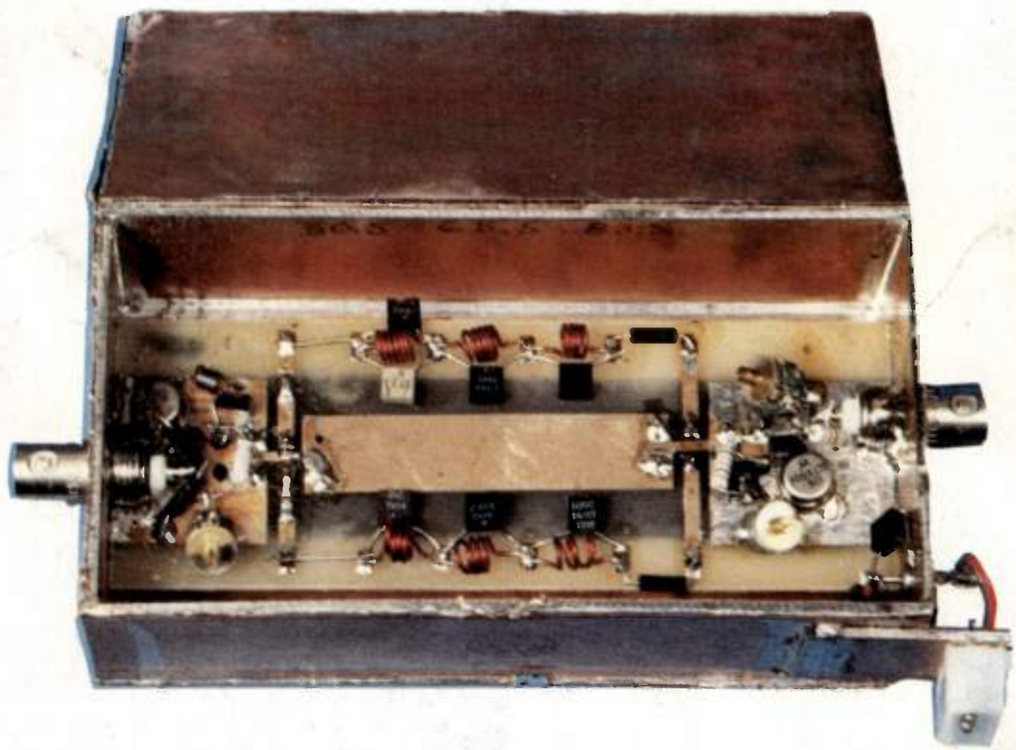
rf design



Bandpass Filter Design

Amplifier Design and Testing

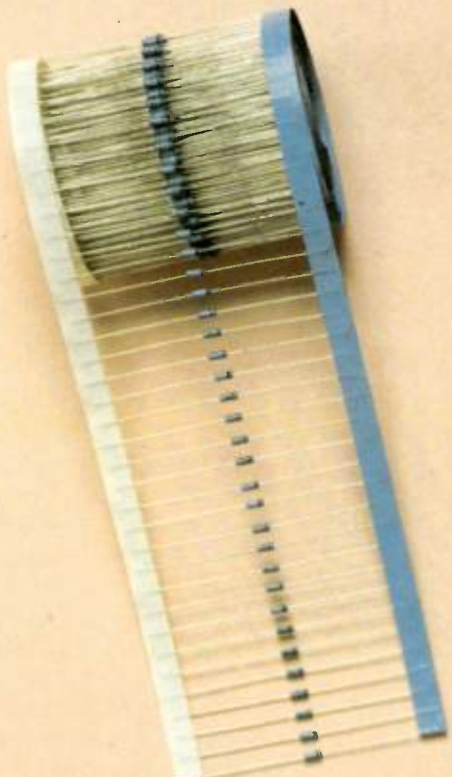
Double Tuned Circuits

Dual Directional Coupler

MILESTONES IN TECHNOLOGY

ONE OF A SERIES

High Volume Schottky and PIN Diodes



MOBILE COMMUNICATIONS • CATV • DIGITAL SWITCHING

These *low cost* axial-lead glass diodes are ideal for use as attenuators, switches, mixers, or detectors in commercial applications. Representative of today's microwave technology, these diodes exhibit performance characteristics far superior to UHF devices. Key typical characteristics for a few typical models are shown in the table.

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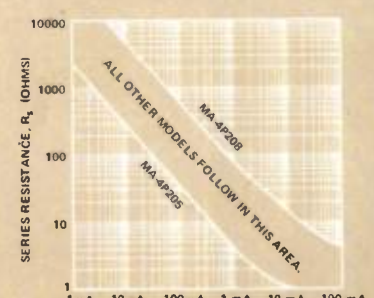
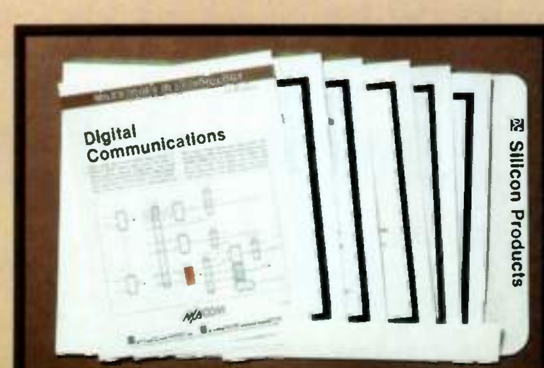


TYPICAL CHARACTERISTICS

Application	Type	Model MA-	V_B (Min.)	V_F (Max. @ 1 mA)	C_T (pF-Max.)	R_S (Ohms @ 100 mA)	R_S (Ohms @ 1 mA)	R_S (Ohms @ 0.01 mA)	T_L (μs)	Max. T_L (ps @ 5 mA)
Switch	PIN	4P205	100	—	1.0 @ 50V	0.4	4	250	1.0	—
Attenuator	PIN	4P208	100	—	0.4 @ 50V	5	90	5000	4.5	—
Detector/ Switch	Schottky	4E2800 (1N5711)	70	0.410	2.0 @ 0V	—	—	—	—	100
Mixer	Schottky	4E2810 (1N5712)	20	0.550	1.2 @ 0V	—	—	—	—	100
Detector	Schottky	4E2835	5	0.340	1.0 @ 0V	—	—	—	—	100

SEND FOR YOUR COPY of *Digital Communications* from our *Milestones in Technology* series along with specifications of *High Volume Schottky and PIN Diodes* — all packaged in a handy file folder.



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INFO/CARD 1

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Frequency Range	1-1600 MHz	0.1-1500 MHz	.01-1800 MHz
Dispersion	20 KHz-1000 MHz	50 KHz-1000 MHz	.5 KHz-1000 MHz
Frequency Accuracy	±.005%**	20% of Dispersion ± 5 MHz	20% of Dispersion ± 5 MHz
Amplitude Dynamic Range	70 dB	70 dB	80 dB
Average Noise Level	−117 dBm (10 KHz resolution)	−107 dBm (10 KHz resolution)	−105 dBm (10 KHz resolution)
Accuracy (worst case total)	±3.5 dB	±3.5 dB	±3.5 dB
Resolution (minimum)	.3 KHz	1 KHz	.03 KHz
Short Term Stability	.2 KHz (phaselocked)	1 KHz	.01 KHz (phaselocked)
Noise Sidebands	−60 dB 10 KHz away	−65 dB 50 KHz away	−75 dB 300 KHz away (10 KHz res.)
Operating Power	115/230V AC, 12V DC	115/230V AC	115V AC
Size In³	1367	2059	Not specified
Weight Lbs	40	40	Not specified
Internal Battery	Standard	Not offered	Not offered
External 12V DC oper	Standard	Not offered	Not offered
Phaselock	Standard	Not offered	Standard
Audio	Standard	Not offered	Not offered
Frequency Markers	Standard	Not offered	Not offered
Digital Storage	Optional	Optional	Standard
Preset Frequency Bands	Standard	Not offered	Not offered
Two Log Ranges	Standard	Standard	Standard
Rugged Carrying Case w/Front Panel Cover	Standard	Optional	Standard
Camera Mount	Standard	Standard	Standard
Portability	Fully portable	Requires an inverter for field use	Requires an inverter for field use

*Information obtained from manufacturer's published specifications **Using internal cyrstal comb markers

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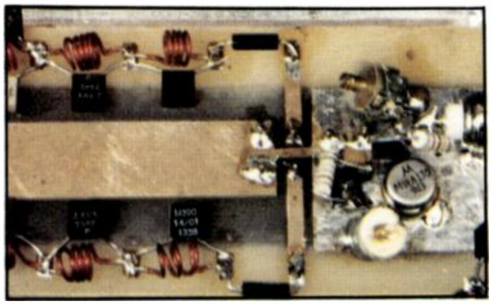
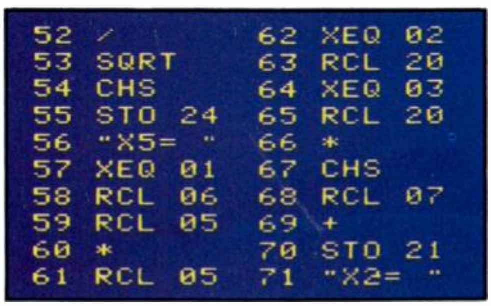
INFO CARD 2

r.f. design

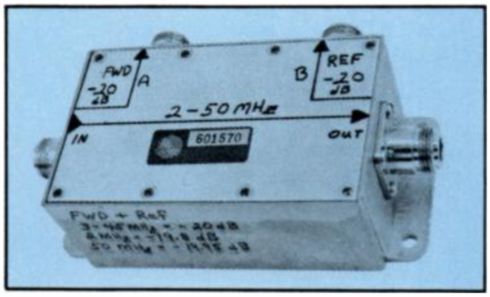
May/June 1982



Double Tuned Circuit



Bandpass Filters



Dual Directional Coupler

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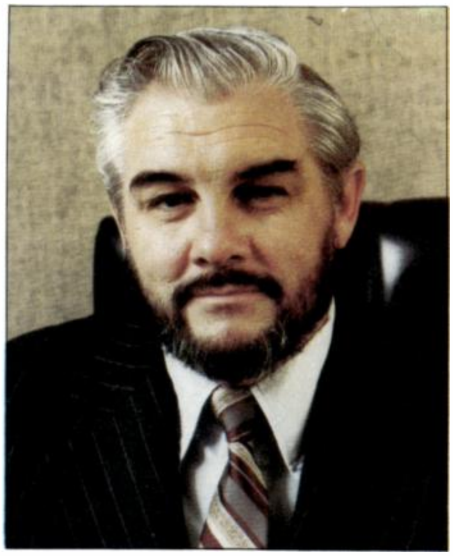
Editorial Excellence

Quality, well read articles are the goals of many editors and publishers around the country. This has always been the primary goal of the *R.F. Design* editorial staff. With each issue we receive confirmation that we are reaching toward this goal from our readers who write in and tell us about the articles.



Recently, Mr. Andrzej B. Przedpelski wrote our consulting editor, Mr. E. Patrick Wiesner, and stated that his (Mr. Przedpelski's) article on "Double Tuned Circuits" in January/February issue had exhausted him of his considerable supply of programs on a magnetic card. When he offered to send a copy of the program he had no idea the article would be so well read. We were pleased.

With Pat's help we have brought on a new editor effective this month. His name is Mr. Larry Brewster. Larry joins us having served the University of Texas, McDonald Observatory in Fort Davis, Texas for the last two years where he has been in charge of their observatory electronics. Prior to his stint there as a research engineer/scientist, Larry worked for Dixson Inc. as a microprocessor systems design engineer, Mostek Corp. as a senior electronic design engineer, Data 100 Corp. as a design engineer, and Control Data and National Cash Register as a design engineer. He is a member of IEEE, has his commercial first class radio telephone license and is a graduate of the University of Colorado, Boulder with a BSEE. You will be seeing and hearing from Larry in the next issue.

We plan for an even better editorial posture in the future with Larry as editor and Pat as consulting editor. Please continue to let us know how you the reader views our progress. We are here to serve your needs in editorial excellence.

Bill W. Childs

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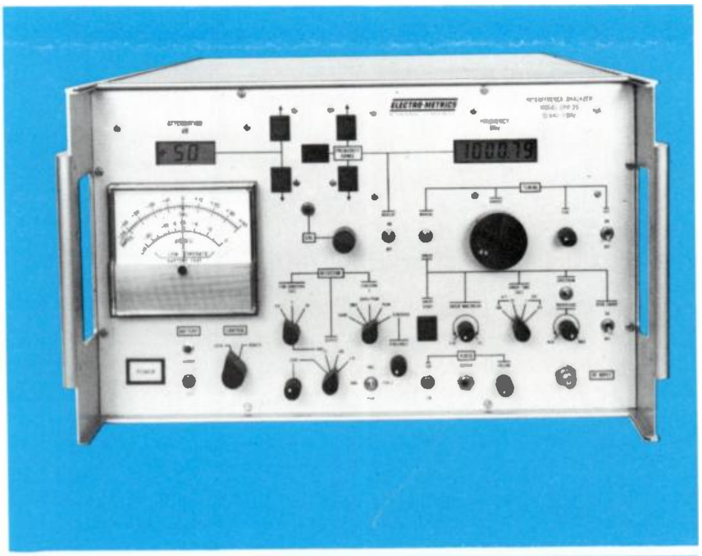

Model ESA-1000 shown with Optional Model DM-1000 Memory Unit.

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Filter Article Corrections

I enjoyed reading my article in a respected technical magazine. Reviewing the article I found a few corrections, which could amend the article, to help reader understanding. I'm listing them below:

- Programming steps 000 to 148 were omitted (I'm enclosing a copy)
- Figure 6 the −1 did not print for the inverse hyperbolic cosine function.
- Figure 7, the division / for AdB/10 did not print.
- Figure 8 steps 6, 7 and 8, the prime in A′, B′, and C′ did not print.
- Appendix A, 14th line, the equation should have read:

$$= 8.68\left[(N+1)\phi - \frac{1}{(N+1)\phi} - N\phi - \frac{1}{N\phi}\right]$$

- Appendix B, 12th line should have read:

$$10^{AM/10} \approx 1 + AM/10$$

I hope you find these corrections useful. A reviewing step, by the article author, might be worthwhile.

Marvin Kefer
49 Madison St.
Massapequa, NY 11758

000	76	LBL	037	01	1	074	04	4	111	92	RTN
001	23	LNX	038	95	=	075	85	+	112	76	LBL
002	43	RCL	039	65	×	076	43	RCL	113	14	D
003	04	04	040	43	RCL	077	08	08	114	42	STO
004	33	X²	041	01	01	078	33	X²	115	02	02
005	55	÷	042	95	=	079	95	=	116	92	RTN
006	43	RCL	043	85	+	080	34	√X	117	76	LBL
007	15	15	044	01	1	081	85	+	118	38	SIN
008	75	-	045	95	=	082	43	RCL	119	42	STO
009	43	RCL	046	28	LOG	083	08	08	120	01	01
010	15	15	047	65	×	084	95	=	121	33	X²
011	95	=	48	01	1	085	55	÷	122	75	-
012	50	I×I	049	00	0	086	02	2	123	01	1
013	42	STO	050	95	=	087	95	=	124	95	=
014	08	08	051	42	STO	088	42	STO	125	34	√X
015	92	RTN	052	09	09	089	13	13	126	85	+
016	76	LBL	053	92	RTN	090	75	-	127	43	RCL
017	16	A'	054	76	LBL	091	43	RCL	128	01	01
018	42	STO	055	28	LOG	092	08	08	129	95	=
019	15	15	056	43	RCL	093	95	=	130	23	LNX
020	71	SBR	057	07	07	094	42	STO	131	92	RTN
021	23	LNX	058	55	÷	095	14	14	132	76	LBL
022	92	RTN	059	43	RCL	096	92	RTN	133	39	COS
023	76	LBL	060	05	05	097	76	LBL	134	43	RCL
4	68	NOP	061	65	×	098	11	A	135	11	11
025	33	X²	062	43	RCL	099	42	STO	136	22	INV
026	42	STO	063	08	08	100	04	04	137	23	LNX
027	01	01	064	95	=	101	92	RTN	138	85	+
028	43	RCL	065	42	STO	102	76	LBL	139	43	RCL
029	03	03	066	10	10	103	12	B	140	11	11
030	55	÷	067	92	RTN	104	42	STO	141	94	+/-
031	01	1	068	76	LBL	105	03	03	142	22	INV
032	00	0	069	48	EXC	106	92	RTN	143	23	LNX
033	95	=	070	43	RCL	107	76	LBL	144	95	=
034	22	INV	071	04	04	108	13	C	145	55	÷
035	28	LOG	072	33	X²	109	42	STO	146	02	2
036	75	-	073	65	×	110	06	06	147	95	=
									148	92	RTN

Program steps omitted from Figure 9, March/April issue.

Wrong Wavelength

N.O. Sokal of Design Automation points out that in A. Przedpelski's article, "Quarter Wavelength and Exponential Taper Line Matching Transformers" on page 26 of the Nov./Dec. 1981 issue, that the "perfect match (τ = 0) is obtained only at frequencies whose wavelength is 4, 4/3, 4/5, 4/7 etc. times the line length. . ." instead of 4, 12, 20, 28 etc. as it is printed.

Thank you, Mr. Sokal—Ed.

You are right, or course. I appreciate your bringing it to my attention. I wrote it wrong in my first draft, and, since it was so simple, neither I nor other people reading it paid attention to it and noticed it.

In addition, the figures lost something in the translation (the x-axis seems to have shifted). I am enclosing copies of the correct ones.

Andrzej B. Przedpelski
Vice President, Development
A.R.F. Products, Inc.
Boulder, Colo.

Enjoyed

Dear Mr. Przedpelski:

We enjoyed your latest article "Double Tuned Circuits" in *R.F. Design* and would like to try out your program.

I've enclosed a self-addressed envelope for your convenience.

I have enjoyed your articles over the years and hope you continue your writing your innovative approches to RF problems.

Leon Weed
Micro-Wave Engineering
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- CE ☐ Consumer electronic mfr
- ME ☐ Marine Communication equipment mfrs.
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- CM ☐ Components mfrs.
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Double Tuned Circuits

The article ("Double Tuned Circuits" by Andrzej B. Przedpelski, Jan./Feb. issue) was interesting and informative. It would be useful to receive the calculator program for design calculations, and helpful if the program were printed with the article in future. The calculator is becoming a constant companion, and we may as well learn to accept its presence in our work as routine.

My only comment regards the choice of methods of coupling in "double tuned circuits" is that capacitive coupling may lead to unwanted high frequency responses that the inductive method tends to avoid (when properly applied). The near-in response of the circuit will be as shown, but there are usually strong responses periodically occuring an octave or more above the desired frequencies depending upon the circuit layout and parasitic impedances.

You can use capacitive coupling and enjoy its flexibility and simplicity without much likelihood of undesirable high-frequency responses if you will tap the inductors of each coupled resonant circuit and couple them at the taps. The capacitor size increases, but the series impedance of the coils now works to prevent high-frequency spurs in the response of an otherwise elegant circuit. An alternative which has been used is to couple through the *stunt* impedance of a common capacitor at the low side junction of the two tuning capacitors (with the inductors separately "grounded".

It pays to look well beyond the desired frequency response of the circuit. Please accept this advice from an engineer who has learned the "hard way".

Wm. D. Young
McLean, VA

"We're Glad You Asked" Dept.

Although the formula below has nothing to do with radio frequency, it certainly is the solution to the pronunciation of Andrzej (Andy) B. Przedpelski's last name. It is being reprinted by popular request from a "rebus" card used by Andy's father.

By Robert W. Sellers
Harris Corp., GESD
Melbourne, Florida

Editor's Note: *Every once in a while, a good engineer who is also a good writer, writes about his encounter with a problem in such a way that the account becomes, to other good engineers, an exciting search for a "solution" that is almost as much fun reading as Robert Ludlum. In our opinion, this two-part article is one of those times.*

Bandpass Filters With Self-Equalized Group Delay: Part 1

Mathematicians are scientists in the strictest sense, and are not impressed by close approximations to perfection; we engineers can be satisfied and, in fact, ecstatic over sufficiently good approximations.

It is generally known to be impossible to provide simultaneously a flat amplitude and group delay (linear phase) over the entire passband of a frequency selective network (the so-called rectangular response). This fact is well treated by the elegant theoretical arguments of Chirlian[1] and Paley-Wiener[2].

The bandpass filter implementation described in this article is not designed with filter theory, but with elementary LC circuit analysis. The performance of the filters shown in Figures 1 and 2 are surprising in several ways. First, the most obvious attribute is that the flat group delay response extends past the frequency where the amplitude has started to roll off. Second, the group delay "ears" normally associated with L-C filters are conspicuously absent. Not evident is a third attribute; namely, the design of the filter is deceptively simple, although successful implementation at 70 MHz requires patience and care.

These filters were developed to satisfy the needs of receivers and modems for low cost linear phase I.F. filters. Computer program data is presented at a 70 MHz center frequency with bandwidths ranging from 5 MHz to 36 MHz.

Test data on filters which were actually constructed is also presented. Some of the key details of the (actual) filter implementation have been intentionally omitted, because of pending patent litigation. Fortunately, these details are not necessary to understand the filter operation but will become painfully evident when one tries to implement the design with actual hardware. It is the author's intent to reveal these details at some future date. In particular, it was found that the filter is particularly susceptible to mutual inductance between adjacent coils, lattice arm cross-coupling and component tolerance. Proprietary techniques have been employed to overcome these difficulties, as evidenced by the test data presented for various bandwidths near the end of this article.

Background

The performance of S.A.W. type filters is evidence that it is possible to make close approximations to the ideal bandpass filter characteristics. S.A.W. filters, however, have certain features which make them very undesirable for use in wideband analog systems such as FM-FDM. The first is the large insertion loss necessitating high intercept point amplifiers to replace the system gain which is lost. Secondly, S.A.W.'s exhibit triple transient reflections, or echo, which we found limited the system NPR to about 35 dB for the 36 MHz bandwidth filters. A third shortcoming of S.A.W. filters is the temperature dependency of the narrow bandwidth units when fabricated in a low-cost material such as Lithium Niobate.

The conventional approach to linear phase filters, of course, is to provide delay equalization external to a minimum phase amplitude selective network by using all-pass elements. An equalization pole is typically required for each pole in the filter network.

This conventional approach has been considered a straightforward acceptable method for implementation of high quality linear phase filters. The main objection to this method is due to the large number of components required which increase the time needed to adjust the filters, and the quality and training of the personnel needed to perform the alignment. While reviewing the published literature on linear phase filter techniques, a unique approach to the design of linear phase filters was found. This was R.M. Lerner's classical article[3] which describes a half-lattice filter design which offered a means for delay self-equalization over a large portion of the filter's 3 dB bandwidth. Lerner's network is elegantly simple, composed of a half-lattice structure with series resonant circuits in parallel in each arm.

The filter design method to be described shortly is due in a large extent to the theories of Lerner, although considerable improvement in group delay has been achieved at the band edges, primarily due to a realization that the "residue" Lerner defines should not be equal when a wide percentage bandwidth is desired.

Experimental Evaluation

The results shown in the Lerner article are directly applicable to linear phase IF filters. It was quickly discovered, however, that the filters could not be easily constructed at high center frequencies.

Following Lerner, a design computation quickly led to inductor values at 70 MHz which were long past self resonance at any reasonable value of filter impedance. In addition, capacitors which were less than the strays associated with the best R.F. layout techniques were required. Fortunately the computer is not concerned by the fact that capacitors with a magnitude on the order of .01 pf cannot easily be produced. A simple program was written in COMPACT to evaluate the (unrealizable) components, and as expected, Lerner's theory proved correct.

The next task was to determine just how large an inductance we could fabricate at 70 MHz. Some experimentation resulted in a modified helical structure producing inductance of about 3.5 uHy. These would be needed for the correctors so inductance values of about 1.7 uHy were fabricated for the inband poles. This configuration produced a filter impedance on the order of 11 ohms at 10 MHz bandwidth.

Intrigued by the computer-predicted performance, and armed with some broadband transformers which produced

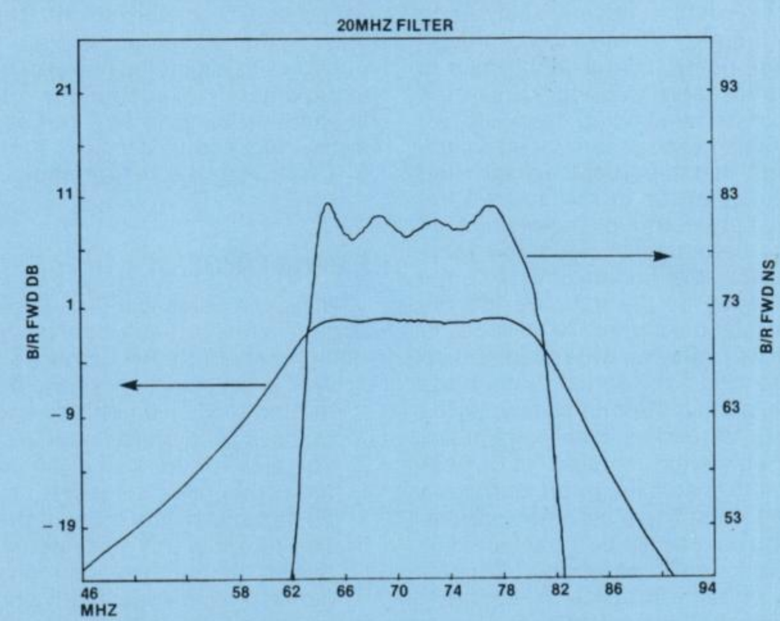

Figure 1. Amplitude and Delay Response of Original Breadboard Filter.

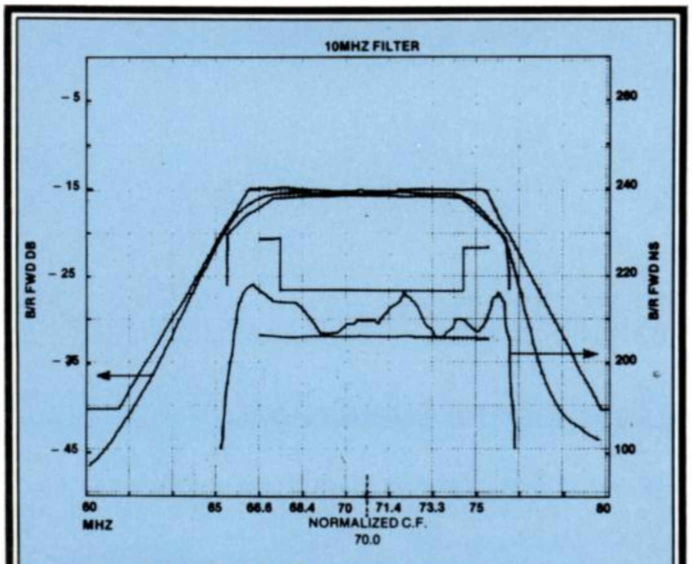

Figure 2a. Second Breadboard Filter.

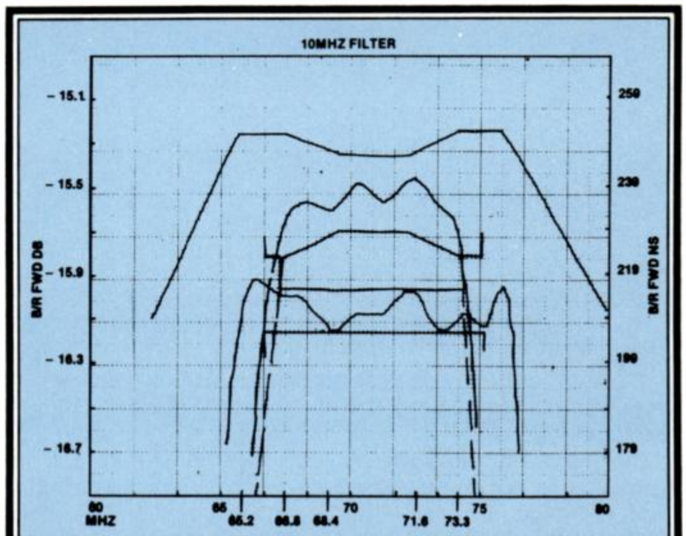

Figure 2b. Noise Response of Second Breadboard Filter.

the low impedance necessary to get into and out of the half lattice, we breadboarded a circuit using helical resonators to provide the required high L to C ratio. This initial breadboard was aligned on the H.P. 8507 network analyzer to produce the filter shown in Figure 1 (20 MHz BW), through sheer determination. The surprising lack of rabbit ears in the group delay characteristics were baffling, to say the least. The next step was to determine what, precisely, we had built. The tuned frequency of each resonator was carefully measured. It was found that they bore no apparent special relationship to each other, or to the desired frequencies as specified by the Lerner article. The network as evaluated by the computer definitely showed the large "rabbit ears" on the group delay. We had obviously built a device which was not described by the data we fed the computer. The breadboard was at the correct 70 MHz center frequency, but its bandwidth was twice as wide as its design value of 10 MHz. Attempts to model the actual filter on the computer resulted in failure, and an attempt to realign the filter to other bandwidths to determine some (unknown) interrelation between the elements resulted in further confusion. After some time, the first major implementation breakthrough for the high frequency filters became apparent.

The Lerner implementation could easily be z-transformed* from its y-equivalent. This resulted in an array of parallel resonant circuits in series in each lattice arm, and very reasonable component values at 70 MHz. A 250 ohm impedance 10 MHz BW filter, for example, would use capacitors in the 100 pf range and inductor values on the order of 50 nh.

Using a z-transformed Lerner design, a four pole (in band) filter was constructed. Alignment of the filter was also difficult, but was eventually accomplished. This filter was fabricated with four inband and two corrector poles and had the flat delay characteristics predicted by Lerner over the central 70 to 75% of the passband. The presence of large group delay ears in this filter and their (accidental) absence in the first try could not be forgotten, however. Imagine the elation when it was discovered that this filter, too, could be "misaligned" to provide a flat group delay characteristic simultaneously with flat amplitude. This could not be achieved at the precise design center frequency, but occurred at a frequency somewhat higher; ≅ 71 MHz. (See Figure 2) At this point, the frequency of each resonator was measured and found, as in the first filter, to have no apparent relationship to the desired (Lerner) frequencies. This time however, an obscure clue was evident. The frequencies to which the (inband) resonators were tuned were approximately exponentially, rather than linearly, spaced in each lattice arm. This turned out to be the second key in the successful design computation of the filter. Prior to continuing the reader should review the Lerner paper.

Lerner Paper

The same nomenclature will be used to describe the filter design in this article wherever possible. The main points to comprehend from the Lerner paper are as follows:

1) With a non-minimum phase design, simultaneous approximation to ideal amplitude and constant dealy characteristics is possible, at least in a bandpass network.
2) The array of (inband) circuits in each half-lattice arm of the Lerner filter is designed such that each has the same admittance as the other at the individual resonances.
3) The filters which were constructed by Lerner were observed to improve their phase characteristics at the nominal band edge (NBE) when the corrector resonators were tuned slightly out of band.

The concept presented by Lerner for designing his filters is repeated here to provide a basis for the discussion which follows.

"The basic algorithm for building these filters turns out to be so simple that the networks can be designed almost without recourse to pencil and paper. Figure 3 shows one form of such a network in (a) as a lattice and in (b) as a half-lattice with a transformer (a convenient practical form).

In both cases, Y_A and Y_B consist of a number of (lossless) series resonant circuits in parallel. The resonators are of two types: all but two are in-band resonators in which the inductances all have the same magnitude L; the other two are corrector resonators whose inductors are nominally 2L. The resonators are tuned to frequencies f_1, f_2. . . at equal intervals 2 Δ f across the desired pass band, alternate frequencies f_1, f_3, f_5. . . being assigned to Y_A and f_2, f_4. . . being assigned Y_B. See Figure 4. A frequency Δ f below f_1 is the nominal 6-dB band edge of this filter. One of the corrector resonators is tuned to this frequency and assigned to the network branch opposite to that of the f_1 resonator: Y_B in

* The terminology "z-transformed" merely means impedance as opposed to admittance in this article, and should not be confused with digital filtering theory.

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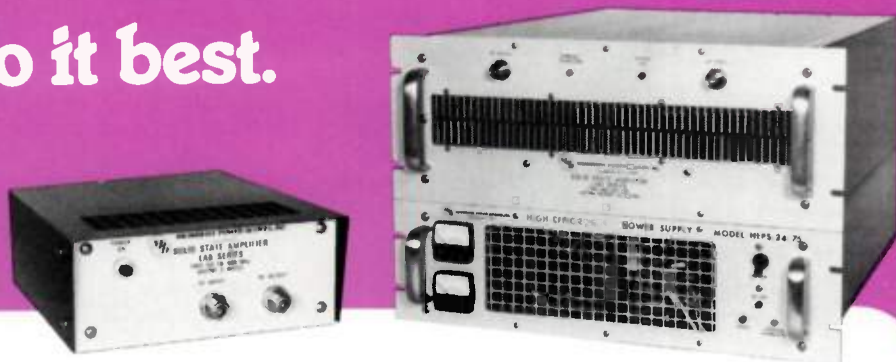

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Figure 4. Similarly, the other corrector resonator is tuned to a frequency Δ f above that of the upper most in-band resonator f_n and assigned to the opposite network branch. The resistance R is taken equal to $4/\pi$ times the (calculated) impedance of L at a frequency of 2 Δ f Hz. In addition, parallel-tuned LC (resonant in-band) and, in some cases, series-tuned LC circuits (resonant out-of-band) may be placed across the loads."

The design method for the modified z-transformed filter is not quite so simple, but is easily handled.

All of the designs presented in this article consist of 4 inband resonators and two corrector resonators, for a total of 3 circuits in each lattice arm. This quantity of resonators yields a 30/3 dB share factor of around 2.2 which was sufficient to meet the Intelsat masks. As with the Lerner filter, the ultimate rejection is a function of how well the lattice arms are matched in amplitude and phase and the common mode rejection of the differencing transformer. In practice, this turned out to be 30 to 35 dB. However, this low ultimate rejection is not of great concern because this rejection outside the passband can be vastly improved by the addition of added out of band zeros shunting the filter input or output lines. These networks have minimal effect on the filter passband. An even number of inband resonators was used for implementation to aid the symmetry of both layout and filter skirt attenuation. It is not known at this time how the filter would be implemented with an odd number of poles, such as would be necessary for a low pass equivalent. Possibly, active devices may prove useful toward implementing a successful low pass design, by extending the required 180° phase shift to D.C.

Practical Discussion

Complete understanding of poles and zeros is not one of the authors' best subjects, and reading the Lerner and Albershiem[4] papers caused many trips to the dusty bookcase for some of the finer points that were made. No amount of reasoning in filter theory alone could explain the existence of a stable network which would exhibit the simultaneous characteristics which were found. For this reason, the explanation which follows is based on intuitive reasoning, simple L-C circuit theory, and brief trips to the time domain.

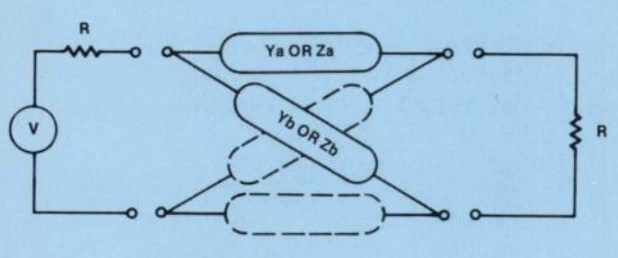

Figure 3a. Full Lattice.

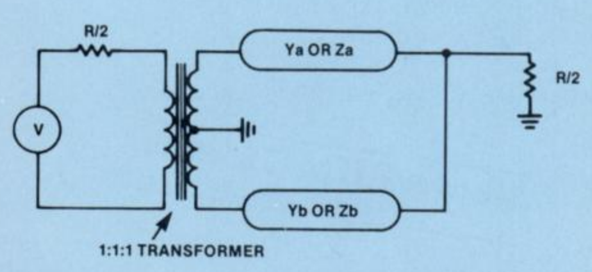

Figure 3b. Practical Half Lattice.

The schematic of the L.C. network of Figure 3 does not resemble a filter at all to the casual designer. Because of this, it is important to realize that the "bandpass" characteristic is brought about by 180° summation in the transformer. For example, at frequencies far from the resonances of the lattice arms, the input signal is split into equal amplitude and opposite sign (due to the transformer) portions which cancel each other in the load. When an input signal in frequency to one of the series resonances occurs, the lattice arm with the proper resonance transmits relatively more of the signal than does the other, producing an unbalance in the two amplitudes, and hence an output to the load.

Here it is important to note that the arm without the signal resonance has a relative null at the signal frequency. For example, reference to Figure 4 and a little careful thought will reveal that when both Y_a and Y_b are presented a signal frequency of say f_3, then L and C_3 produce maximum transmission between the ends of Y_b, while Y_a has an attenuation pole at f_3. This is because the circuit at f_2 in Y_a is past its resonant frequency and appears mostly inductive, while the network at f_4 (also in Y_a) is below resonance and appears capacitive. Due to the integral frequency spacing, and the choice of components, the result is a *parallel resonant* circuit at f_3 in Y_a. The poles and zeros then alternate between the lattice arms, with one neatly cancelling the residual of the other, across the passband. If the circuit components are selected correctly, the filter is self equalized over the inband portion of the passband. In the Lerner lattice, the resonator at each nominal band edge (NBE) is selected to have twice the value of inductance as the inband resonators, but with its frequency spacing only ½ that of the inbands. The result is a proper frequency location to cancel the closest inband resonator in the opposite arm, but alas, there is no corrector for the corrector pole itself, save the compensation by the external series resonances.

This filter network is designed to have a relative *phase difference* between the lattice arms of 90° across the passband, at frequencies between individual resonances. This is purposely achieved by the selection of the resonances to interlace with a "periodicity" of Δ f, equivalent to using a (very long) transmission line in each lattice, with one 90° longer than the other at a frequency of 2 ΔF. It is this phase difference that allows the *ratio* of the transfer functions to have both poles *and* zeros in the right-half plane! Here the right half plane poles imply that one of the networks has less phase than the other and not that the output of either network anticipates the input signal, an obvious impossibility. With a sufficient number of poles, it is theoretically[4] possible to approximate this 90° phase difference over a given range of frequencies to within any desired phase tolerance.

The concept of using the relative phase between two networks to achieve flat time delay is at least as old as single sideband A.M. Both Albersheim and Shirley, who introduced some very useful calculation methods, and the Rhodes[5] theory in 1968 recognized the need for more than one signal path between the input and output in order to allow time-invariant transmission through a two-port device.

Elementary Considerations

Since no compensation apparently exists in the Lerner lattice filter for the corrector poles, how is the modified device able to eliminate the group delay "ears" and provide such surprising results? For the answer, we must turn to the basic theory of the LRC circuit. Remember that we wish to "turn the corner" with the group delay characteristic with the same sort of authority as the amplitude response.

Since group delay is the derivative of phase shift versus frequency, we must select a function for phase shift change at the NBE which has the same "shape" as its derivative. We are then limited to selecting one of the transcendental functions. Lerner had selected the cosine shape, but the apparent exponential frequency spacing in the second breadboard filter offered another suggestion and with some additional thought, the following explanation is offered.

A purely resistive network is truly time invariant. However, without the inclusion of frequency selective components, no filter results. It is here we must realize that it is not necessary for the (composite) network to be time invariant at all frequencies, but we do require that it have different but controlled values of transmission (resistance) at particular frequencies in each lattice arm. At least, within the desired passband it is also required that each resonator be memoryless, or *critically damped* with *respect* to the other resonators. This infers for the complete (half) lattice that either:

1) The tuned frequency of each resonator be offset from the Lerner spacing, or
2) the magnitude of the individual inductances (capacitances in the lattice) be modified accordingly, i.e. they will not all be the same value, as dictated by the Lerner method.

In retrospect, I have noticed with interest that if a large number of in band resonators were employed, and a relatively narrow percentage bandwidth filter was required, there is little practical differences in the component values between Lerner's method and the one to be described shortly. This is due to the fact that the arithmetic and the geometric means of two different (inband) frequencies are essentially the same when the frequency difference is small, as would be required with large numbers of inband resonators. In his article, Lerner describes nine and 23 resonator filters of relatively narrow bandwidth. The method in this article is based upon 4 inband resonators used in relatively wide bandwidth filters.

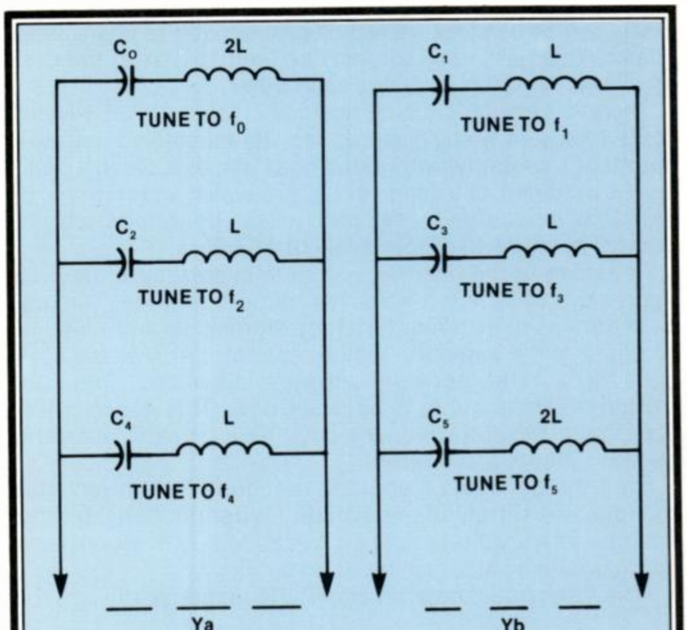

Figure 4. Form of Ya and Yb.
$f_0 = f_1 - \Delta f, f_2 = f_1 + 2\Delta f, f_3 = f_1 + 4\Delta f$, etc.
f_1 = lower filter band edge

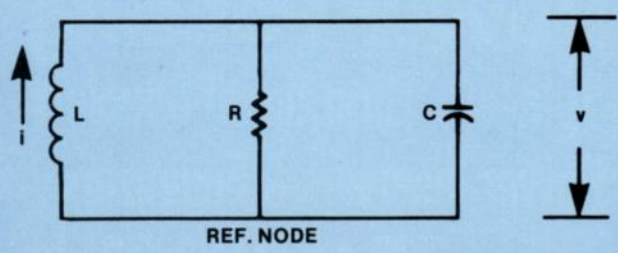

Figure 5. Elementary Parallel Circuit.

Theory Behind the Design Method

The above considerations are not possible with a minimum phase design since a circuit "Q" greater than unity is required to achieve selectivity. With the lattice design, the cancellation provided by the transformer phase inversion can be used to provide the selectivity, and if critically damped LC circuits are employed, the resultant network (loaded) Q for the individual arms is 0.5, or that of a single critically damped LC circuit.

It is well known that the general form of the natural response of a parallel resonant circuit is, (see Figure 5)

$$V_{(T)} = A_1e^{S_1t} + A_2e^{S_2t} \qquad (1)$$

Where $$S_{1,2} = -\alpha \pm \sqrt{\alpha^2 - \omega o^2} \qquad (1A)$$

and $A_{1,2}$ represent the arbitrary constants used to satisfy some specified initial conditions of i and V.

$\frac{1}{2RC} = \alpha$ is the exponential damping coefficient, or Neper frequency

while $\omega o = \frac{1}{\sqrt{LC}}$ is the resonant radian frequency of the circuit.

The circuit is over-damped when $LC > 4R^2C^2$, and under-damped when $LC < 4R^2C^2$. Critical damping occurs when $LC \equiv 4R^2C^2$. The single special case of critical damping produces a unique circuit. This can be seen by allowing ωo to be equal to α in eq. 1A. At this point, equation 1 apparently losses its meaning because $S_1 = S_2 = S$, and it can be rewritten as

$$V_{(t)} = A_1e^{St} + A_2e^{St} = A_3e^{St} \qquad (2)$$

Eq. 2 contains only one arbitrary constant, but there are *two* initial conditions (i and V) which must be satisfied. The solution to this apparent problem is given in several elementary circuits texts[6] and can be found by regression to the defining differential equation for Figure 5 which reduces to:

$$V_{(t)} = e^{-\alpha t}(A_1t + A_2) \text{ or} \qquad (3)$$

$$V_{(t)} = A_1te^{-\alpha t} + A_2e^{-\alpha t} \qquad (4)$$

It should be noted that:

1) This is *not* an overall exponential solution as is (1) and,
2) The solution is expressed as the sum of two terms, one being the negative exponential, but the second is t times the negative exponential.
3) $e^{-\alpha t}$ is a simple delay operator, exponential in time.

The meaning of the third statement may become more germain to the filter lattice arms with the following explanation.

The real and the imaginary parts of the complex frequency describe, respectively, the exponential and the sinusoidal variation of an exponentially varying sinusoid. Our ordinary concept of "frequency" actually carries with it another connotation in addition to "repetitions per second". It also tells us something about the rate of change of the function being considered. For example, if we take

$$f(t) = Ke^{St} \qquad (5)$$

and differentiate to obtain the time rate of change of f(t)

$$\frac{df}{dt} = SKe^{St},$$

and normalize by dividing by f(t) we have

$$\frac{df/dt}{f_{(t)}} = S.$$

This normalized rate of change is a constant, *independent* of time. It is moreover identically equal to the complex frequency, S. We may therefore interpret complex frequency as the "normalized time rate of change" of the complex exponential function (5). This alternate definition may lead to some curious results. For example, although the complex frequency associated with the function $e^{j\omega t}$ is $s = j\omega$, the normalized rate of change associated with cos ωt must be $(-\omega)$ tan ωt. If we try to treat the result, which is a function of time, as a complex frequency, then we are led to a complex frequency which is a function of time. The correct answer is obtained only when it is recognized that a conjugate *pair* of complex frequencies is required to characterize cos ωt. This pair, however, may be uniquely (as in the case of critical damping), equal to each other.

There is certainly nothing new here. The "curious results" could represent nothing more than the monotonic (resistive) nature of a properly terminated coaxial cable. Such a cable can be modeled as a series of L-C networks, and they will be found to be critically damped when reflections are not present because of proper cable termination.

Returning to the filter network, it is interesting to describe the means by which it achieves its performance. This can be more easily explained by first digressing and using an example of a network which cannot be realizable. In general, a filter network with the idealized amplitude response (rectangular) is not realizable. This is not due to the *steepness* with which the amplitude cuts off, but rather because it cuts off to *zero.*

Even the gaussian amplitude response filter attenuates too much to satisfy the realizibility criterion: consider that a series of N R-C networks are cascaded (with isolation) to approximate a gaussian error curve.

This filter would have an amplitude response of:

$$A_n(\omega) = \left(1 + \frac{\omega^2 \ln^2}{N}\right)^{-N/2} \text{ with an}$$

associated phase function of

$$\phi n(\omega) = N \tan^{-1}\left(\frac{\omega\sqrt{\ln^2}}{\sqrt{N}}\right) \quad (7)$$

Comparing (6) with a gaussian error curve having $\omega = 1$ as its 3 dB point (.707), e.g.,

$$A = e^{-\omega^2 \frac{\ln^2}{2}} \quad (8)$$

it is clear that equations (6) and (8) converge as $N \rightarrow \infty$. However $\phi(N)$ (7) diverges, becoming infinitely large for all ω.

Consider now two cases of amplitude characteristics whose phase functions tend to converge with the amplitude response, e.g.,

$$A(\omega) = (\sin \omega/\omega)\, e^{-j\omega} \text{ and} \quad (9)$$

$$A(\omega) = (\sin^2\omega/\omega^2) e^{-2j\omega} \quad (10)$$

Filters with these amplitude arguments will be recognized to have the familiar sinc and (sinc^2) impulse responses, which certainly insinuates flat time delay and amplitude response in the frequency domain.

In equations (9) and (10), it is easily seen that the associated phase functions converge (as a matter of fact converge to *linear* phase lags of ω and 2ω radians respectively.

The practical significance of these two cases is that although networks of the form of (9) have amplitude characteristic which satisfy the Paley-Wiener criterion, they are only exactly realizable by an *infinite* number of RLC elements. (Or by a finite number of lines with distributed constants.)

Network (10) is not exactly realizable at all over an infinite frequency range, but at the expense of increased time delay, the ideal amplitude and phase characteristics can be *approximated* arbitrarily closely over a finite frequency range, and as it turns out, with a very finite number of networks. The absolute value function for the filters described in this article is of the form of (10) as evidenced by their characteristics. In particular the phase slope shown in the figures at the end of this article is 2ω radians. The filter time delay absolute value yields a time-bandwidth product of mean 2, as expected. And finally, the extremely close convergence of amplitude response and time delay is conspicuously evident in the computer plots as will be shown in Part II.

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(5) See the dicussion surrounding Dr. Rhodes work as in "Linear Phase Filters and The Demise of External Equalization". Ralph Levy, Microwave Systems News, Aug/Sept. 1976.

(6) For example "Engineering Circuit Analysis", Hayt and Kemmerly, McGraw-Hill, 1962, second edition. □

HOW LOAD VSWR AFFECTS

If your amplifiers test out fine in the lab but fail QC testing, the testing environment

By Don Murray
TRW Semiconductors,
Lawndale, Calif.

Consider the following scenario: You're designing and implementing into production a broadband Class C power amplifier. During your design phase, you follow all the rules of science and also dig into your bag of electronic tricks to meet the design specification. Your design is fabricated and tested successfully in the lab. Twenty-five more units are built in the lab and they, too, test out fine.

Confident that both design and production procedures are satisfactory, you begin series production. But when the first units reach RF test, not one meets specification. Yet when you retrieve the units, they test OK in the lab.

What's wrong with these amps? Probably nothing. This scenario, in one form or another, is all too common in the design and manufacture of non-linear RF circuitry. The culprit is correlation of test systems. A difference of .5 dB is enough to fail units that are perfectly good, resulting in unnecessary and expensive retesting or even reworking. Still worse, a half dB error will pass units that don't meet specs and never should be shipped.

Such correlation errors will disrupt an even more important function, that of maintaining product continuity. A device built in 1982 should perform the same as an identical model number device built in 1976. Another way of saying this is that a device tested in a 1982 test system should produce the same results when tested in a 1976 system. The key, of course, is RF correlation.

What is RF correlation? Simply put, RF correlation occurs when target error limits are established *and* adhered to on a continous basis among two or more testing stations. Such correlation is essential to cost-effect production of non-linear RF and microwave power amplifiers, whose circuits are extremely sensitive to the impedance of their loads, either in test systems or equipment environments. It is easy to compensate for the insertion loss errors in an attenuator, but it is much more difficult to compensate for variations in the input impedance difference between attenuator pads, that is, the load VSWR.

Let's examine RF correlation on both an empirical and theoretical level.

Empirical Approach

The empirical approach is shown in Table I, where several test circuit loads (consisting of series attenuators, directional couplers and RF switches) were assembled. The insertion loss and input impedance of each load string was measured. Following this, the individual loads were connected to a given test circuit containing a common base microwave power transistor. The power meter used was also a constant.

Table I shows insertion loss, insertion loss corrections, indicated RF power, and actual power data of each load string. A maximum error of 0.52 dB was detected with a standard deviation of .19 dB. All these loads had a VWSR less than 1.1:1 at the frequency tested. A VSWR of 1.1:1 is better than the published specifications of commercially available attenuators, directional couplers, and RF switches from most leading manufacturers. A VSWR of 1.5:1 is a typical VSWR specification limit at 1.4 GHz. It must be noted that many users will gladly pay an additional nominal charge for components meeting a tighter VSWR spec.

Theoretical Approach

The vehicle for the theoretical discussion is the well known expression:

$$P_0 = \frac{(V_{CC} - V_{CESAT})^2}{2R_L}$$

Where: P_0 = Power output
V_{CC} = Collector supply voltage
V_{CESAT} = Collector-emitter saturation voltage
R_L = Load resistance.

This expression is valid for a narrow range of R_L (10% range maximum). Over a wider range of R_L, significant changes in V_{CESAT} occur as a function of R_L. Output power varies with the square of V_{CESAT}. V_{CESAT} is a very strong function of collector current and transistor die temperature.

The theoretical approach will evaluate the changes in amplifier output power (P_0) for a given change in load resistance (R_L).

For simplicity, let us assume the following hypothetical conditions, which are typical of today's RF power transistors.

NON-LINEAR CIRCUITS

—not the product—is likely at fault.

Hypothetical conditions:

$V_{CC} = 28V$
$V_{CESAT} = 1.5V$
$P_{OUT} = 50W$
Frequency = 1.0 GHz

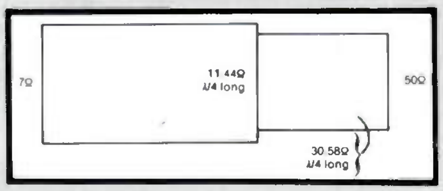
Solving for load resistance:

$$R_L = \frac{(V_{CC} - V_{CESAT})^2}{2P_0} = \frac{702.25}{100} = 7.02\Omega$$

Additionally, assume that a simple two-section impedance matching network matches the 7Ω to 50Ω. Let this two-section match consist of two λ/4 wave transformers.

Given the conditions we have hypothesized, the R_L of 7.02 Ω represents the collector load that will yield the best simultaneous satisfaction of device efficiency, device gain, gain transfer characteristics, and saturated power.

For minimum Q, with a 2 section match, the transformation ratio of each section is $\sqrt{\frac{50}{7}} = 2.67$.



$$Z_0 \text{ 1st section} = \sqrt{(7)(2.67)(7)} = 11.44\Omega$$

$$Z_0 \text{ 2nd section} = \sqrt{(7)(2.67)(50)} = 30.58\Omega$$

λ/4 @ 1 GHz = 2.95" = .075m

Table II shows the transformed impedance at the input of the matching network as a function of various load impedances. Our example utilizes a real-to-real impedance match for

Table I. Microwave Load Substitution Study

The vehicle used for this test was a production test fixture and correlation sample #2 for the TRW MRA-1417-6 broadband, high-gain transistor. Measurements were taken at 1400 MHz with input power of 1.1 W.

Load #	Measured Power Level	Circuit Return Loss	Collector Current	Measured Insertion Loss	Calibration Error	Actual Power	Delta from Reference	Load Input Return Loss	Impedance Angle	Real	Imaginary
1	1.1 W	35 dB	–	30.03 dB	+.03 dB	thru	calibration	–40.2	99.1	49.8	+1.0
1	7.7W	16 dB	.51 A	30.03 dB	+.03 dB	7.75 W	reference	–40.2	99.1	49.8	+1.0
2	7.6 W	15.5 dB	.5 A	39.66 dB	–.44 dB	6.87 W	–.52 dB	–30.5	–77.5	50.6	–3.0
3	7.65 W	15.5 dB	.51 A	39.68 dB	–.32 dB	7.10 W	+.38 dB	–34.1	–171.5	50.4	–2.0
4	8.0 W	15.5 dB	.51 A	39.8 dB	–.20 dB	7.63 W	–.07 dB	–34.1	68.1	50.7	–1.9
5	7.2 W	16 dB	.505 A	30.16 dB	+.16 dB	7.47 W	–.16 dB	–30.1	–128.0	51.1	–3.0
6	8.3 W	15.2 dB	.51 A	39.78 dB	–.22 db	7.89 W	+.08 dB	–31.7	–144.6	47.9	–1.5
7	7.75 W	16.2 dB	.505 A	39.73 dB	–.27 dB	7.28 W	–.27 dB	–32.7	11.9	49.0	–2.4
8	7.78 W	16.8 dB	.503 A	39.7 dB	–.30 dB	7.26 W	–.28 dB	–35.4	–111.9	49.1	–1.5

Largest Delta after calibration correction is 0.52 dB.
Mean value of the measured power = 7.41 W.
Standard Deviation = .34 W = .19 dB.

Note: – 30 dB RETURN LOSS = ρ of 0.03 and VSWR of 1.06:1.

Table II. R_L Effects on Output Power

Load Resistance (Ω)	Transformed Load Resistance (Ω)	Output Power (W)	ΔdB	Cumulative ΔdB
45	6.30	55.73		
			.095	.095
46	6.44	54.52		
			.093	.189
47	6.58	53.36		
			.091	.280
48	6.72	52.25		
			.090	.370
49	6.86	51.18		
			.087	.457
50	7.00	50.16		
			.086	.543
51	7.14	49.18		
			.085	.628
52	7.28	48.23		
			.083	.710
53	7.42	47.32		
			.081	.791
54	7.56	46.45		
			.080	.871
55	7.70	45.60		

Maximum Delta dB Vs. VSWR

VSWR	Maximum ΔdB
1.02	.17 (±.085)
1.04	.34 (±.17)
1.06	.51 (±.255)
1.08	.68 (±.34)
1.10	.87 (±.435)

Table III Notes

Suggestions to the Maintenance Of Correlation

1. Serialize and document all components (attenuators, directional couplers, power meters, detectors, etc.) of the test system. Do not disturb the system once calibration has been performed. Calibrate the system once a month.
2. Require that loads have a calibration return loss ⩾ −35 dB (VSWR of 1.05:1) in frequency band of interest.
3. Dedicate test systems to specific circuits or products. This is necessary for both correlation and product continuity.
4. The placement of transistors in the test fixtures must be uniform. For instance, flanged transistors should be placed in the test fixtures with the device pushed towards collector load circuitry.
5. Be selective when using cables in test systems. For example, the MIL-C-17 specification for "RG" cable types says that RG-58 can have a characteristic impedance from 48 to 52 Ω (maximum VSWR of 1.04:1) when terminated in a "perfect" 50 Ω load.
6. Be very selective when choosing RF switches. The VSWR of a mechanical switch will vary with time.
7. If possible, terminate the system with a 50 Ω load rather than an attenuator. Load manufacturers need only consider the VSWR of a load. However, for attenuator, tradeoffs must be made between VSWR and frequency response. Measure power and other performance parameters via calibrated directional couplers.

convenience. The analysis also is appropriate for an imaginary-to-real match in that center of the VSWR circle at the input to the matching network will be rotated but won't change in magnitude from the data presented.

Conclusion

The data presented in table represents the power variation into a load with a VSWR of 1.1:1 relative to 50Ω. The result is a power output of 50W ± 5.3W (± .435 dB). The total Delta is 10.3 W (.87 dB). This is enough to:

A) Make a good circuit look bad, or...

b) Make a bad circuit look good.

This analysis was done for a single frequency. The problem is compounded in a broadband environment by requirements for a good broadband load impedance.

Test Equipment Accurcy

Test equipment manufacturers have produced some very impressive equipment in recent years; however, the accuracy of a well constructed system using the latest equipment available is generally considered to be no better than ± 3%. Considering the number of variables in RF testing and the magnitude of the task faced by the test equipment manufacturers, ± 3% is no small achievement. However, ± 3% is ± .13dB. This ± .13 dB added to the ± .435 dB indicated earlier yields a total possible error magnitude of ± .565 dB. This adds up to a total possible error of ± 14% into a load with 1.1:1 VSWR. The output power range of our amplifier is now 50 W ± 7.05 W.

Now we see how bad things can be, a few comments on reality are in order.

The author believes that the correlation target for the test of RF power devices should be ± 0.2 dB, which we believe is the optimum tolerance for combining strict quality standards and the need for easy repeatability under series production conditions. If more than an occasional device fails this test, do not assume that the devices are at fault. Instead, first analyze the test circuit and then the test system to determine the reason for the additional error. Some suggestions on how to maintain a ± 0.2 dB correlation are shown in Table III.

The 0.2 dB target is an achievable target in broadband test systems. However, a constant awareness of the test system capabilities and potential problem areas is mandatory. RF correlation problems will never go away, but they can be made easier to handle. ☐

LOW IMPEDANCE DOUBLE TUNED

CIRCUIT

Przedpelski strikes again! Another calculator program to make our design life easier.

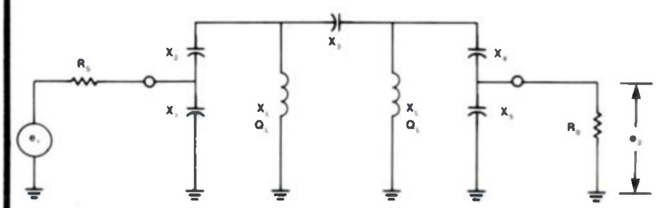
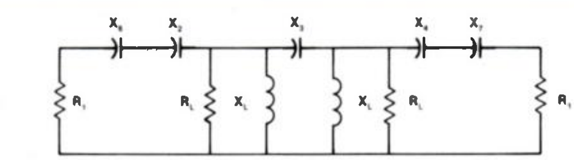


Figure 1. Low Impedance Coupling Circuit.

(a)



$$R_L = Q_L X_L \quad (1)$$

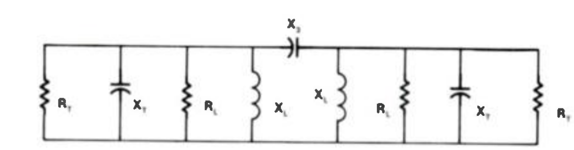
$$X_6 + X_2 = X_7 + X_4 \quad (2)$$

$$X_6 = X_1 \frac{\left(\frac{R_S}{X_1}\right)^2}{1 + \left(\frac{R_S}{X_1}\right)^2} \quad (3)$$

$$X_7 = X_5 \frac{\left(\frac{R_0}{X_5}\right)^2}{1 + \left(\frac{R_0}{X_5}\right)^2} \quad (4)$$

$$R_1 = \frac{R_S}{1 + \left(\frac{R_S}{X_1}\right)^2} = \frac{R_0}{1 + \left(\frac{R_0}{X_5}\right)^2} \quad (5)$$

(b)

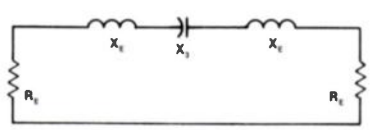


$$Q_0 X_L = \frac{R_T R_L}{R_T + R_L} \quad (6)$$

$$X_T = \frac{X_3}{Q_0 - 1} \quad (7)$$

$$R_1 = \frac{R_T}{1 + \left(\frac{R_T}{X_T}\right)^2} \quad (8)$$

(c)



$$X_3 = -2X_E \quad (9)$$

$$X_6 + X_2 = X_7 + X_4 = X_T \frac{(R_T/X_T)^2}{1 + (R_T/X_T)^2} \quad (10)$$

Figure 2. Equivalent Circuits.

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Using the circuit described previously[1] to match two low impedances can result in unreasonable component values, particularly for narrow bandwidths (high Q_O). The circuit shown in Figure 1 overcomes this problem. It is a combination of the standard double tuned circuit[2] and the capacitive divider[3]. In addition, to obtain correct results, the Q of the standard selected inductors is also considered.

To simplify the calculation it is assumed that input and output stray capacities (if any) can be combined with X_1 and X_5. Using the equivalent circuits of Figure 2, the following method can be used to obtain the circuit values:

a. Using (9), (6), and (7) obtain:

$$X_3 = -X_L Q_O \tag{11}$$

b. Using (7), (6), (8), (2) and (10) obtain:

$$X_1 = -\sqrt{\frac{R_1 R_S^2}{R_S - R_1}} \tag{12}$$

and

$$X_5 = -\sqrt{\frac{R_1 R_O^2}{R_O - R_1}} \tag{13}$$

c. Using (3), (4), (10) and (2) obtain:

$$X_2 = (X_6 + X_2) - X_6 \tag{14}$$

and

$$X_4 = (X_7 + X_4) - X_7 \tag{15}$$

The HP-41C/CV program of Table I will calculate the required circuit values. Store Q_O, R_S, R_O, Q_L and X_1 in registers R12, R13, R15, R17 and R19 respectively. XEQ "CIRKT 2" and the values of X_1, X_2, X_3, X_4 and X_5 will be stored in registers R20, R21, R22, R23 and R24 respectively, in addition to being shown in the display or printed out. These stored values can be then used to calculate frequency response using program of Table II.

The current ratio [1] is used to obtain the circuit response vs. frequency. The basic relationships are shown in Figure 3. Note that all the circuit reactances (shown in Figures 1 and 3a) are reactances of circuit components at the center frequency, F_O. The program calculates the actual reactance at the different frequencies by using the F/F_O input requested by the PROMPT display.

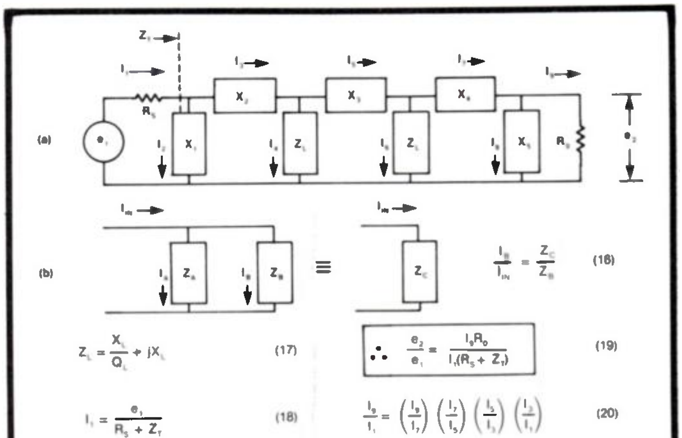


Figure 3. Basic Gain and Current Relationships.

```
01♦LBL "CIR
KT 2"
02 RCL 12
03 RCL 19
04 *
05 STO 05
06 CHS
07 STO 22
08 "X3= "
09 XEQ 01
10 RCL 05
11 RCL 19
12 RCL 17
13 +
14 STO 06
15 *
16 RCL 06
17 RCL 05
18 -
19 /
20 STO 05
21 RCL 22
22 RCL 12
23 1
24 -
25 /
26 STO 06
27 RCL 05
28 RCL 05
29 RCL 06
30 /
31 X↑2
32 STO 05
33 XEQ 02
34 X↑2
35 +
36 RCL 13
37 RCL 07
38 -
39 /
40 SQRT
41 CHS
42 STO 20
43 "X1= "
44 XEQ 01
45 RCL 07
46 RCL 15
47 X↑2
48 *
49 RCL 15
50 RCL 07
51 -
52 /
53 SQRT
54 CHS
55 STO 24
56 "X5= "
57 XEQ 01
58 RCL 06
59 RCL 05
60 *
61 RCL 05
62 XEQ 02
63 RCL 20
64 XEQ 03
65 RCL 20
66 *
67 CHS
68 RCL 07
69 +
70 STO 21
71 "X2= "
72 XEQ 01
73 RCL 15
74 RCL 24
75 XEQ 03
76 RCL 24
77 *
78 CHS
79 RCL 07
80 +
81 STO 23
82 "X4= "
83♦LBL 01
84 ARCL X
85 AVIEW
86 FC? 21
87 STOP
88 RTN
89♦LBL 02
90 1
91 +
92 /
93 STO 07
94 RCL 13
95 RTN
96♦LBL 03
97 /
98 X↑2
99 RCL X
100 1
101 +
102 /
103 .END.
```

REGISTERS

R05	USED
R06	USED
R07	USED
R12	Q_0
R13	R_S
R15	R_0
R17	Q_L
R19	X_L
R20	$[X_1]$
R21	$[X_2]$
R22	$[X_3]$
R23	$[X_4]$
R24	$[X_5]$

SIZE: 025

Table I.

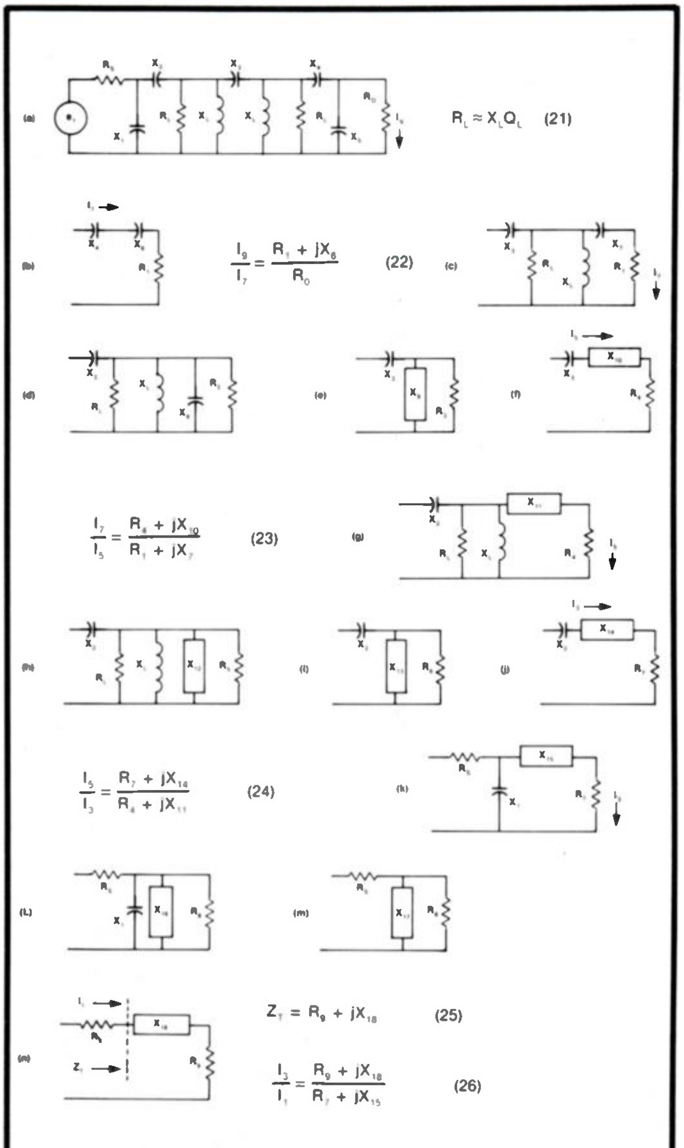


Figure 4. Current Ratios Calculations.

The calculation of the current ratios, shown in Figure 4 is quite tedious, since it involves numerous parallel-to-series, and series-to-parallel conversion, as well as complex division. The program, shown in Table II is relatively simple, however, since the same calculations are repeated and can be used as subroutines. The procedure is shown in detail, since it can be applied to other similar calculations.

To obtain the response, e_2/e_1, equation (19) is used. Since I_9/I_1 is the product of the four current ratios, as shown in equation (20), the four ratios calculated by equations (22), (23), (24) and (26) can be used. The other required unknown, Z_T, can be obtained from equation (25). The program of Table II displays the response in dB, or 20 log e_2/e_1.

Two more equations are required to proceed with the circuit design. It should be remembered that the center frequency is closer to the geometric rather than arithmetic mean, or:

$$F_0 = \sqrt{F_1 F_2} \tag{27}$$

and

$$Q_O \approx \frac{F_0}{0.7\,BW_{3db}} \tag{28}$$

An example will demonstrate the procedure and show some features of this method: a circuit is needed to couple a source resistance of 300 ohms to a load resistance of 50 ohms at 30 MHZ with bandwidth of 1 MHZ, using 1 UH inductors. Two types of inductors are available: with a Q_L of 150 and 45.

Since F_1 = 29.5 MHz and F_2 = 30.5 MHz, F_O = 29.996 MHz, using equation (27). Using equation (28) we obtain Q_O = 42.85. Using:

$$X_L = 2\pi F_O L \tag{29}$$

1 μh

we obtain X_L = 188.47, the last required input to proceed with circuit value calculation, using Table I program:

$Q_L = \infty$	$Q_L = 150$	$Q_L = 45$	
$X_3 = -8{,}075.940$	$X_3 = -8.075.940$	$X_3 = -8{,}075.940$	.659pF
$X_1 = -37.471$	$X_1 = -31.604$ 168pf	$X_1 = -8.133$	652.7
$X_5 = -15.932$	$X_5 = -13.276$ 399pf	$X_5 = -3.326$	1596p
$X_2 = -155.968$	$X_2 = -161.660$ 32.84pf	$X_2 = -184.847$	28.78p
$X_4 = -178.400$	$X_4 = -180.516$ 29.4pf	$X_4 = -189.662$	27.9

Using:

$$C = -\frac{1}{2\pi F_0 X_C} \tag{30}$$

the actual capacitor values can be calculated. For any circuit and inductor value there is a minimum Q_L required. In the above example, for instance, Q_L values below 40 do not give a solution (DATA ERROR at step 40). The value of the parallel resistance, R_L, is too low to provide a match.

Using the circuit values obtained, the frequency response can be calculated using the program of Table II. The pro-

```
01•LBL "RES
P 2"
02 FS? 00
03 GTO 00
04 "F/F0=?"
05 PROMPT
06 VIEW X
07•LBL 00
08 STO 27
09 RCL 19
10 *
11 RCL 17
12 *
13 STO 18
14 RCL 15
15 RCL 24
16 RCL 27
17 /
18 XEQ B
19 STO 16
20 RCL 15
21 /
22 X<>Y
23 STO 14
24 RCL 15
25 /
26 R-P
27 STO 25
28 RDN
29 STO 26
30 RCL 14
31 RCL 16
32 RCL 23
33 XEQ D
34 XEQ A
35 RCL 14
36 RCL 16
37 RCL 22
38 XEQ D
39 XEQ A
40 RCL 14
41 RCL 16
42 RCL 21
43 RCL 27
44 /
45 +
46 STO 16
47 XEQ C
48 RCL 20
49 RCL 27
50 /
51 1/X
52 +
53 1/X
54 XEQ A
55 RCL 15
56 ST* 25
57 RCL 16
58 RCL 14
59 RCL 13
60 +
61 R-P
62 ST/ 25
63 RCL 25
64 LOG
65 20
66 *
67 FS? 00
68 RTN
69 "RESP= "
70 ARCL X
71 "⊢ DB"
72 AVIEW
73 ADV
74 RTN
75•LBL A
76 XEQ B
77 X<>Y
78 RCL Y
79 X<> 16
80 RCL Y
81 X<> 14
82 R-P
83 1/X
84 X<>Y
85 CHS
86 RDN
87 RDN
88 R-P
89 R↑
90 +
91 RDN
92 +
93 R↑
94 ST+ 25
95 RDN
96 ST+ 26
97 RTN
98•LBL B
99 1/X
100 X<>Y
101 1/X
102 R-P
103 1/X
104 P-R
105 X<>Y
106 RTN
107•LBL C
108 X<>Y
109 R-P
110 1/X
111 P-R
112 1/X
113 X<>Y
114 RTN
115•LBL D
116 RCL 27
117 /
118 +
119 STO 16
120 XEQ C
121 RCL 19
122 RCL 27
123 *
124 1/X
125 +
126 1/X
127 X<>Y
128 1/X
129 RCL 18
130 1/X
131 +
132 1/X
133 X<>Y
134 RTN
135 END
```

REGISTERS

R_{13}	R_s
R_{14}	[R]
R_{15}	R_0
R_{16}	[X]
R_{17}	Q_L
R_{18}	[R_L]
R_{19}	X_L
R_{20}	X_1
R_{21}	X_2
R_{22}	X_3
R_{23}	X_4
R_{24}	X_5
R_{25}	[MAG]
R_{26}	[<]
R_{27}	[F/F_0]

SIZE: 028

Table II.

"To provide stability in frequency output for our new cable television series 6700 set-top terminals, we felt we needed SAW resonators," explained Mr. Kelly. "But there were a couple of problems. We not only needed the resonators in high volume, but we needed them in a hurry.

"So, we spoke with RF Monolithics. They said they could solve the problem. That's one of the points we liked best about them, right away. They've got a unique problem-solving attitude.

"They began to work closely with our engineers to make sure that our requirements and costs were coordinated. They even went beyond the device and helped with circuit application assistance, too. Yet, they continued to make sure that everything was cost beneficial to both of us.

"We gave RFM some tough problems. But they met every challenge we gave them. They produced the amounts of high-quality resonators we needed and on schedule.

"RFM did what they said they could do. They helped solve a difficult problem for us. That's why we plan to continue with them."

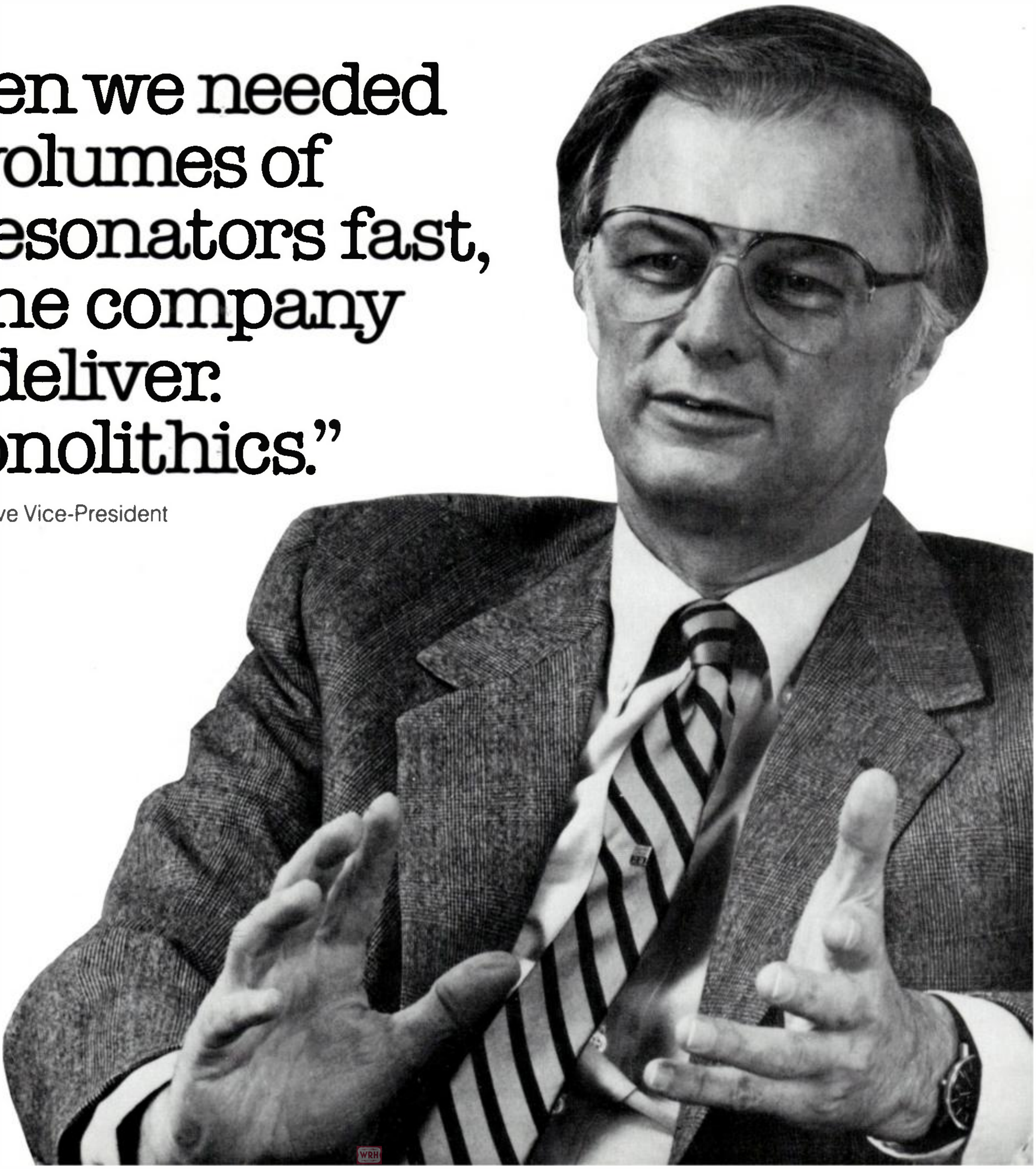
To find out more about our second generation resonators and low-loss filters in the UHF/VHF range, write RF Monolithics, 4441 Sigma Road, Dallas, TX 75234. Or phone (214) 233-2903.

RF Monolithics, Inc.
4441 Sigma Road
Dallas, Texas 75234
214 233-2903
INFO/CARD 12

F

"When we needed high volumes of SAW resonators fast, only one company could deliver. RF Monolithics."

—Jack Kelly, Executive Vice-President
Scientific-Atlanta



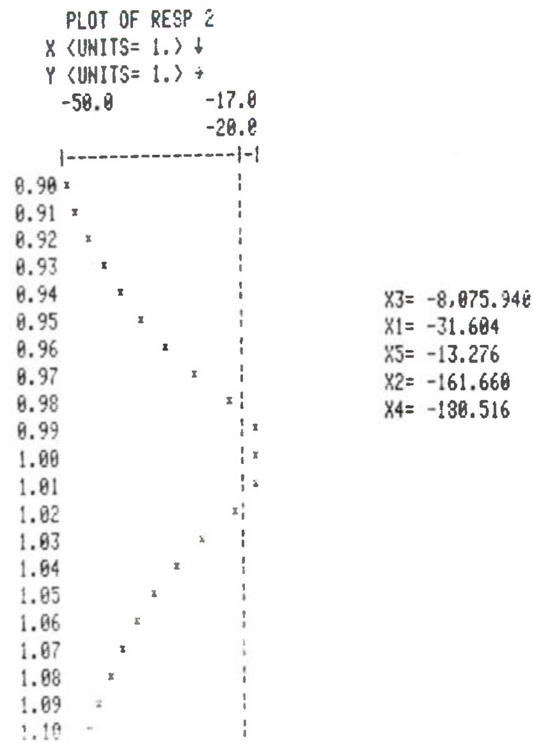


Figure 5. Frequency Response Plot.

gram is arranged to permit plotting (using PRPLOT) by setting flag 00. The frequency response shape is very similar for the three examples given. However, the insertion loss at center frequency ($F/F_O = 1$) will be different. This insertion loss includes the transformation loss (10 log R_O/R_S), the inherent loss of 6.02 DB and any losses due to finite Q_L of the inductors. Thus, the insertion loss using lossless inductors is:

$$IL = 20 \log 2\sqrt{R_O/R_S} \tag{30}$$

This is confirmed by the program and the calculated response is −13.802 dB at $F/F_O = 1$ for infinite inductor Q. The loss increases as the inductor Q decreases and the response becomes −16.724 dB for $Q_L = 150$ and −40.218 dB for $Q_L = 45$. When the critical minimum Q_L is reached (between 40 and 45 in this example) the insertion loss becomes infinite, but the circuit is still properly matched.

In this case it is obvious that high-Q inductors have to be used to provide reasonable insertion loss.

The plot of the response (using inductors with a Q of 150) is shown in Figure 5. The response is down 3.2 dB at 29.5 MHz and 2.9 dB at 30.5 MHz or close to the required 3 dB.

References

1. A. Przedpelski, "Double Tuned Circuits", r.f. design, Jan/Feb 82.

2. F.E. Terman, "Radio Engineer's Handbook", McGraw-Hill Book Co., N.Y., 1943, P. 164.

3. A. Przedpelski, "Program Impedance Matching with Capacitive Tap", Electronic Design, Nov. 22, 1980. □

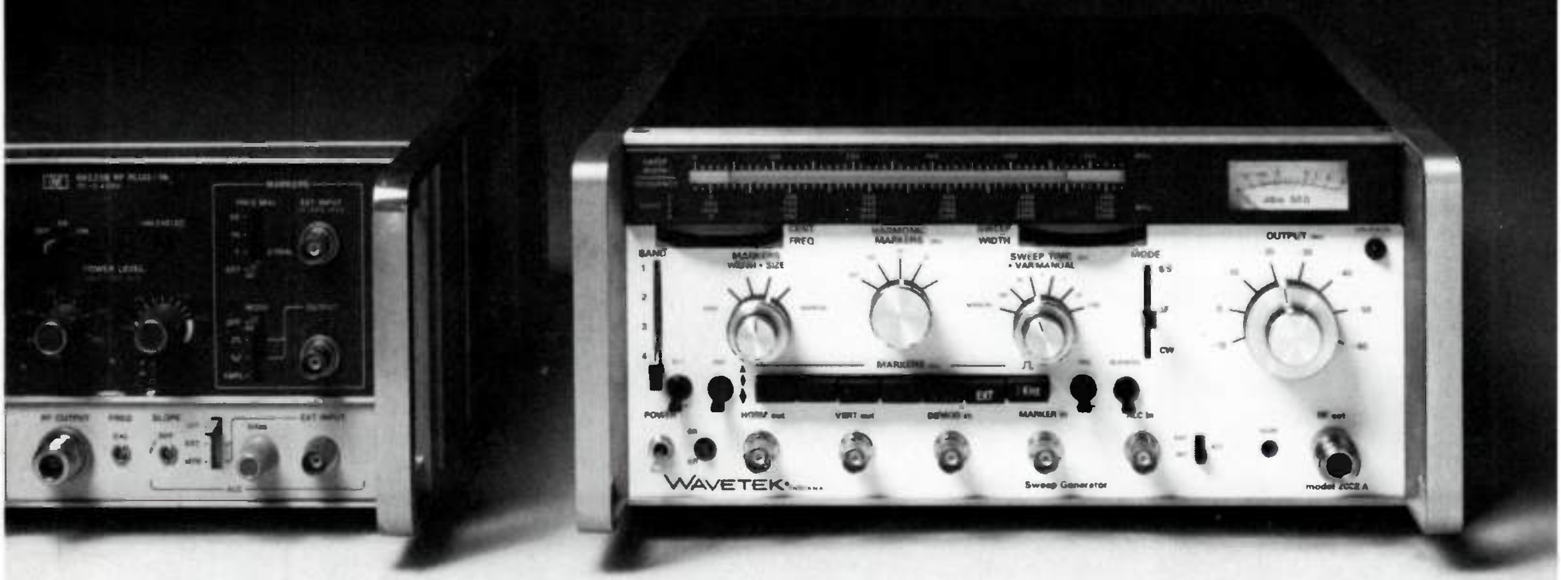
If you're looking for a sweeper in the 2.5 GHz range, the obvious choices are Wavetek's Model 2002A and HP's 8620C mainframe with an 86222B plug-in. But just look at the chart: the Model 2002A is a lot more instrument for about half the price. The only way your choice could be more clear would be if the HP instrument didn't exist. But then what could we compare our Model 2002A against? Wavetek Indiana, Inc., P.O. Box 190, 5808 Churchman, Beech Grove, IN 46107. Toll-free 800-428-4424; in Indiana, (317) 787-3332. TWX (810) 341-3226.

Demonstration: INFO/CARD 15
Literature: INFO/CARD 33

	Wavetek Model 2002A including Harmonic Markers	HP 8620C Mainframe with 86222B plug-in
Price	$4600	Plug-in $5750 + Mainframe $2850 = Total $8600
Convenience	Single, stand-alone unit	Two detachable units, mainframe & RF plug-in
Frequency Range	1 MHz to 2.5 GHz	10 MHz to 2.4 GHz
Non-harmonics at 13 dBm	None detectable (0.5 to 2.5 GHz) >35 dBc (1 to 500 MHz)	> 30 dBc (0.01 to 2.3 GHz) > 25 dBc (2.3 to 2.4 GHz)
Calibrated Output Level Meter	Standard	Not available
Step Attenuator	Standard	Optional at $400
Harmonic Markers	1, 10, 50 and 100 MHz (plus single frequency markers)	1, 10 and 50 MHz
Marker Range	1 MHz to 2.5 GHz	10 and 50 MHz markers to 2.4 GHz 1 MHz marker to 1.0 GHz
Marker Width	Adjustable from 15 kHz to 400 kHz	Fixed. Minimum width is 150 kHz
Marker Size	Adjustable	Fixed

WAVETEK®

Compare our 2.5 GHz sweeper with its closest competitor.



(The competition costs more and does less; we think that puts it out of the picture.)

Design Of RF Amplifiers

Part II: Using Potentially Stable Devices

Part I dealt with potentially unstable devices. Here, Part II, completes the discussion.

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In this part a TI 59 program is provided which first calculates the Linvill stability factor, C, for a device and then, if the device is inherently stable, calculates optimum device terminations and the transducer gain. With these three programs on hand, an amplifier can be designed using either an inherently stable or potentially unstable device whose Y or S parameters are known.

When a designer of an RF amplifier is fortunate enough to find an inherently stable device, optimum terminations which maximize transducer power gain, can be found. Solution of these terminations is straight forward but tedious since complex arithmetic is involved. The given calculator program, however, calculates the Linvill stability factor, C, the optimum terminations, and the transducer gain in approximately 20 seconds. The stability of a device as shown in Figure 1, whose Y parameters are known, can be investigated by calculating the Linvill stability factor C.

$$[y] = \begin{bmatrix} y_i \, y_r \\ y_f \, y_o \end{bmatrix} = \begin{bmatrix} g_i + jb_i & g_r + jb_r \\ g_f + jb_f & g_o + jb_o \end{bmatrix}$$

$$C = \frac{|y_f y_r|}{2g_i g_o - \mathrm{Re}(y_f y_r)} \qquad (1)$$

where | | and Re( ) denote the magnitude and the real part of the complex quantity respectively. If C < 0 or C ⩾ 1 the device is potentially unstable and optimum terminations which maximize transducer power gain do not exist. The designer can then use the procedure described in the article. If 0 ⩽ C < 1 the device is inherently stable and optimum terminations can be found as follows. The transducer gain, G_T, is given by

$$G_T = \frac{4G_S G_L |y_f|^2}{|(y_i + y_s)(y_o + y_L) - y_f y_r|^2} \qquad (2)$$

Each of the four optimum termination quantities, G_S, B_S, G_L, and B_L are found by setting the partial derivative of G_T with respect to each of the quantities to zero. Solution of these four equations gives the optimum terminations.

$$G_S = \frac{[(2g_ig_o - Re(y_fy_r))^2 - |y_fy_r|^2]^{1/2}}{2g_o} \tag{3}$$

$$B_S = -b_i + \frac{I_m(y_fy_r)}{2g_o} \tag{4}$$

$$G_L = \frac{[(2g_ig_o - Re(y_fy_r))^2 - |y_fy_r|^2]^{1/2}}{2g_i}$$

$$G_L = \frac{G_sg_o}{g_i} \tag{5}$$

$$B_L = -b_o + \frac{I_m(y_fy_r)}{2g_i} \tag{6}$$

Note that if $y_r = 0$, which is the unilateral case, Y_S and Y_L, as expected, become the complex conjugates of y_i and y_o respectively.

All that is required to run the program is the y parameters of a device. The program calculates C, and, if the device is inherently stable, G_S, B_S, G_L, B_L, and the transducer gain in db. If a printout is desired, label A' should be pressed before execution.

Example #1

Calculate C, optimum terminations, and G_T for a device with the following y parameters.

$$\begin{bmatrix} y_i\, y_r \\ y_f\, y_o \end{bmatrix} = \begin{bmatrix} 8.5 + j3.5 & 0 - j.2 \\ 40 - j20 & .35 + j.75 \end{bmatrix}$$

all in mmhos

Procedure:

1. Read side 1 and side 2 of the magnetic card.
2. Press E to initialize the program.
3. Enter y_i, y_r, y_f, and y_o in order using the following procedure.
   Enter g_i
   Press A Note: g_i, b_i . . . g_o, b_o are automatically printed as they are entered.
   Press A
   .
   .
   .
   Enter b_0
   Press A
4. Press A' if a printout is desired
5. Press B

The printout is shown at right. If the printer is not used, the calculator will stop with C displayed. Successive pressings of R/S will display G_S, B_S, G_L, B_L, and G_T in that order. Execution time for this example is approximately 20 seconds.

If S parameters are to be used the S → Y conversion program from Part I must be executed first. This program uses the same procedure for entering the device parameters and automatically places the calculated parameters in the proper registers.

```
         8.5-03
         3.5-03
          0.  00
         -2. -04
          4. -02
         -2. -02
         3.5-04
         7.5-04

   0.8989218        C
  .0062274081       GS
 -.0149285714       BS
  .0002564227       GL
 -.0012205882       BL
  21.45415076       GT
```

Label E must be pressed to initialize the program before entering the S parameters and the S parameters must be entered in polar form.

If the device is potentially unstable the printer will print only C. If the printer is not used the calculator will stop displaying C, however, pressing R/S will cause the display to flash.

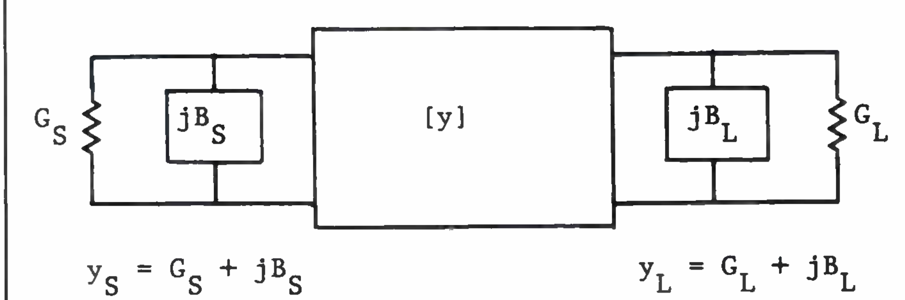

Figure 1. Amplifying Device with Terminations.

Program Listing

```
000 91 R/S
001 76 LBL
002 11 A
003 72 ST*
004 09 09
005 99 PRT
006 69 OP
007 29 29
008 91 R/S
009 76 LBL
010 16 A'
011 86 STF
012 01 01
013 91 R/S
014 76 LBL
015 12 B
016 43 RCL
017 12 12
018 42 STO
019 01 01
020 43 RCL
021 13 13
022 42 STO
023 02 02
024 43 RCL
025 14 14
026 42 STO
027 03 03
028 43 RCL
029 15 15
030 42 STO
031 04 04
032 36 PGM
033 04 04
034 13 C
035 42 STO
036 18 18
037 53 (
038 24 CE
039 33 X²
040 85 +
041 32 X:T
042 42 STO
043 29 29
044 33 X²
045 54 )
046 34 √X
047 42 STO
048 19 19
049 53 (
050 24 CE
051 55 ÷
052 53 (
053 02 2
054 65 ×
055 43 RCL
056 10 10
057 65 ×
058 43 RCL
059 16 16
060 75 -
061 43 RCL
062 18 18
063 54 )
064 54 )
065 42 STO
066 20 20
067 43 RCL
068 20 20
069 29 CP
070 22 INV
071 77 GE
072 23 LNX
073 01 1
074 32 X:T
075 43 RCL
076 20 20
077 77 GE
078 23 LNX
079 53 (
080 53 (
081 53 (
082 02 2
083 65 ×
084 43 RCL
085 10 10
086 65 ×
087 43 RCL
088 16 16
089 75 -
090 43 RCL
091 18 18
092 54 )
093 33 X²
094 75 -
095 43 RCL
096 19 19
097 33 X²
098 54 )
099 34 √X
100 55 ÷
101 02 2
102 55 ÷
103 43 RCL
104 16 16
105 54 )
106 42 STO
107 23 23
108 53 (
109 43 RCL
110 11 11
111 94 +/-
112 85 +
113 43 RCL
114 29 29
115 55 ÷
116 02 2
117 55 ÷
118 43 RCL
119 16 16
120 54 )
121 42 STO
122 24 24
123 53 (
124 43 RCL
125 23 23
126 65 ×
127 43 RCL
128 16 16
129 55 ÷
130 43 RCL
131 10 10
132 54 )
133 42 STO
134 25 25
135 53 (
136 43 RCL
137 17 17
138 94 +/-
139 85 +
140 43 RCL
141 29 29
142 55 ÷
143 02 2
144 55 ÷
145 43 RCL
146 10 10
147 54 )
148 42 STO
149 26 26
150 53 (
151 43 RCL
152 10 10
153 85 +
154 43 RCL
155 23 23
156 54 )
157 42 STO
158 01 01
159 53 (
160 43 RCL
161 11 11
162 85 +
163 43 RCL
164 24 24
165 54 )
166 42 STO
167 02 02
168 53 (
169 43 RCL
170 16 16
171 85 +
172 43 RCL
173 25 25
174 54 )
175 42 STO
176 03 03
177 53 (
178 43 RCL
179 17 17
180 85 +
181 43 RCL
182 26 26
183 54 )
184 42 STO
185 04 04
186 36 PGM
187 04 04
188 13 C
189 42 STO
190 27 27
191 32 X:T
192 42 STO
193 28 28
194 43 RCL
195 12 12
196 42 STO
197 01 01
198 43 RCL
199 13 13
200 42 STO
201 02 02
202 43 RCL
203 14 14
204 42 STO
205 03 03
206 43 RCL
207 15 15
208 42 STO
209 04 04
210 36 PGM
211 04 04
212 13 C
213 94 +/-
214 44 SUM
215 27 27
216 32 X:T
217 94 +/-
218 44 SUM
219 28 28
220 53 (
221 53 (
222 43 RCL
223 27 27
224 33 X²
225 85 +
226 43 RCL
227 28 28
228 33 X²
229 54 )
230 35 1/X
231 65 ×
232 04 4
233 65 ×
234 43 RCL
235 23 23
236 65 ×
237 43 RCL
238 25 25
239 65 ×
240 53 (
241 43 RCL
242 14 14
243 33 X²
244 85 +
245 43 RCL
246 15 15
247 33 X²
248 54 )
249 54 )
250 53 (
251 24 CE
252 28 LOG
253 65 ×
254 01 1
255 00 0
256 54 )
257 42 STO
258 32 32
259 87 IFF
260 01 01
261 24 CE
262 43 RCL
263 20 20
264 91 R/S
265 43 RCL
266 23 23
267 91 R/S
268 43 RCL
269 24 24
270 91 R/S
271 43 RCL
272 25 25
273 91 R/S
274 43 RCL
275 26 26
276 91 R/S
277 43 RCL
278 32 32
279 91 R/S
280 76 LBL
281 24 CE
282 22 INV
283 52 EE
284 98 ADV
285 98 ADV
286 01 1
287 05 5
288 69 OP
289 04 04
290 43 RCL
291 20 20
292 69 OP
293 06 06
294 02 2
295 02 2
296 03 3
297 06 6
298 69 OP
299 04 04
300 43 RCL
301 23 23
302 69 OP
303 06 06
304 01 1
305 04 4
306 03 3
307 06 6
308 69 OP
309 04 04
310 43 RCL
311 24 24
312 69 OP
313 06 06
314 02 2
315 02 2
316 02 2
317 07 7
318 69 OP
319 04 04
320 43 RCL
321 25 25
322 69 OP
323 06 06
324 01 1
325 04 4
326 02 2
327 07 7
328 69 OP
329 04 04
330 43 RCL
331 26 26
332 69 OP
333 06 06
334 02 2
335 02 2
336 03 3
337 07 7
338 69 OP
339 04 04
340 43 RCL
341 32 32
342 69 OP
343 06 06
344 91 R/S
345 76 LBL
346 23 LNX
347 98 ADV
348 22 INV
349 52 EE
350 01 1
351 06 6
352 01 1
353 07 7
354 04 4
355 02 2
356 02 2
357 04 4
358 01 1
359 05 5
360 69 OP
361 01 01
362 01 1
363 07 7
364 00 0
365 00 0
366 03 3
367 03 3
368 03 3
369 02 2
370 03 3
371 07 7
372 69 OP
373 02 02
374 01 1
375 07 7
376 03 3
377 01 1
378 03 3
379 07 7
380 02 2
381 04 4
382 01 1
383 03 3
384 69 OP
385 03 03
386 02 2
387 07 7
388 02 2
389 07 7
390 04 4
391 05 5
392 00 0
393 00 0
394 00 0
395 00 0
396 69 OP
397 04 04
398 69 OP
399 05 05
400 69 OP
401 00 00
402 04 4
403 01 1
404 03 3
405 01 1
406 03 3
407 06 6
408 03 3
409 07 7
410 01 1
411 03 3
412 69 OP
413 01 01
414 01 1
415 04 4
416 02 2
417 07 7
418 01 1
419 07 7
420 00 0
421 00 0
422 00 0
423 00 0
424 69 OP
425 02 02
426 69 OP
427 05 05
428 69 OP
429 00 00
430 98 ADV
431 01 1
432 05 5
433 69 OP
434 04 04
435 43 RCL
436 20 20
437 69 OP
438 06 06
439 91 R/S
440 00 0
441 35 1/X
442 91 R/S
443 76 LBL
444 15 E
445 01 1
446 00 0
447 42 STO
448 09 09
449 81 RST
```

Example #2

$$\begin{bmatrix} y_i & y_r \\ y_f & y_o \end{bmatrix} = \begin{bmatrix} 1.5 + j3.5 & 0 - j.3 \\ 56 - j11 & .1 + j.75 \end{bmatrix}$$

all in mmhos

Repeat the procedure of example #1 to yield the following printout.

```
           1.5-03
           3.5-03
            0.  00
          -3. -04
           5.6-02
          -1.1-02
            1. -04
           7.5-04

DEVICE POTENTIALLY
UNSTABLE

 4.755844358        C
```

If it is still desired to use the potentially unstable device, the "RF Amp with Specified Stability" program from Part I can now be executed without reentering the Y parameters. Simply read the magnetic card, enter the desired Stern's stability factor, and press B.

The three programs given in the two parts are compatible and very simple to run. As in the previous programs, this program does not use direct addressing which allows the user to easily add to or modify the program.

References

1. Carson, R.S., High-Frequency Amplifiers, Wiley Intersciences, 1975.
2. Kraus, H.L., Solid State Radio Engineering, Wiley, 1980. □

Errata

In part 1 in the last issue one step was inadvertently omitted from the procedure in example 2 (page 25). The S → Y parameter conversion program must be initialized by pressing label E before entering the S parameters. Failure to initialize the program causes the S parameters to be stored in the wrong registers.

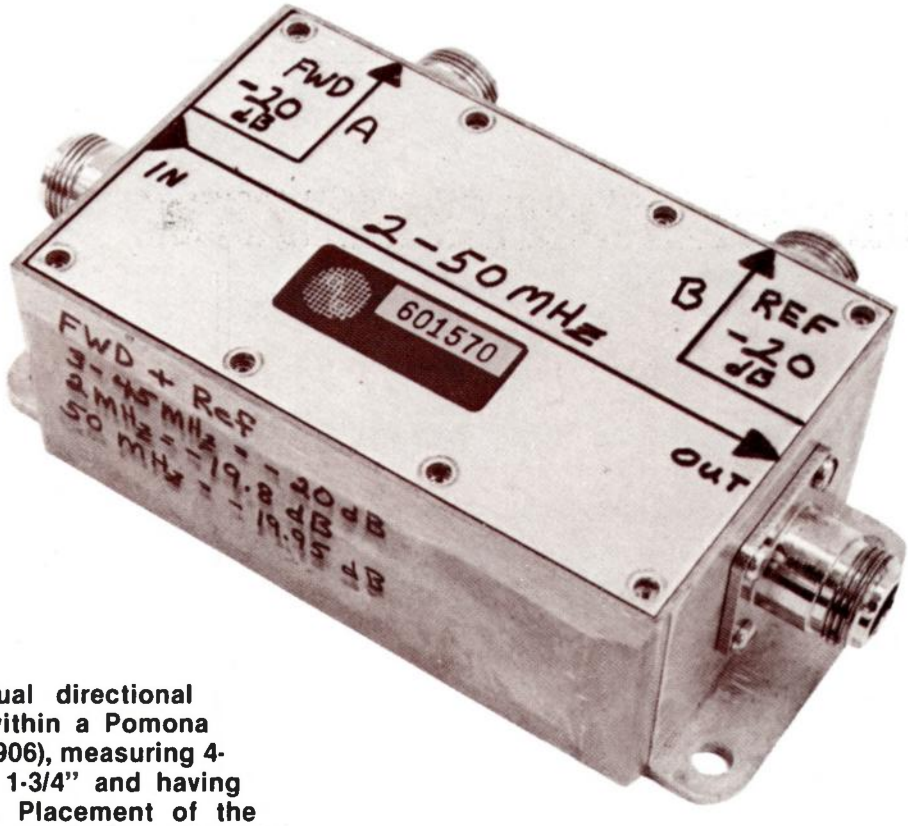

Figure 1. The dual directional coupler is built within a Pomona Electronics box (2906), measuring 4-1/4" × 2-5/8" × 1-3/4" and having 1/2" end flanges. Placement of the type—N receptacles is dedicated by the printed circuit board.

LOW COST, WIDEBAND DUAL DIRECTIONAL COUPLER

Described in this article is a low cost, dual directional coupler for use with a 50 ohm coaxial transmission line. It has a useful range of 500 kHz to over 150 MHz and a power handling capability of 1,000 watts, continuous service.

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A directional coupler is a device which samples RF power flowing in one direction but is insensitive to power flow in the reverse direction. Two directional couplers can be used "back-to-back" to make an inexpensive SWR meter. Choice of couplers is selected by the forward-reverse switch of the instrument. In other circumstances a single directional coupler will do the job.

The coupler is shown in Figure 1. The design was a outgrowth of the need of a high power dual device for laboratory testing, monitoring forward and reflected power in the drive line to an experimental RF power amplifier.

The schematic of the simple dual coupler is given in Figure 2. Two inexpensive toriod transformers in the lines provide inductive coupling. The unit is built on a printed circuit board which provides the necessary microstrip lines. All that is required for construction are the transformers, the board, the coaxial fittings and an inclosure.

Coupler Construction

An interior view of the device is shown in Figure 3. The coupler is built upon a two-sided circuit board measuring 4-1/8" long by 2-3/8" wide. The board just fits comfortably within the cast aluminum box. The input and output coaxial

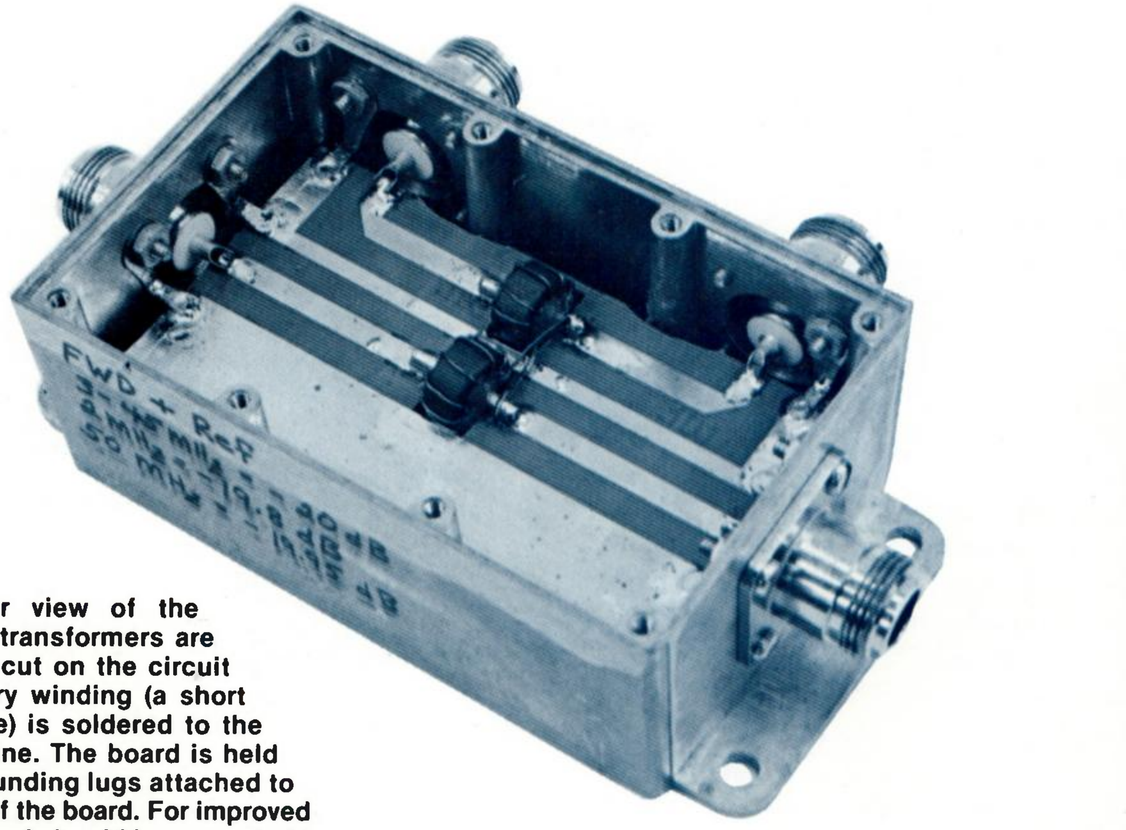


Figure 3. Interior view of the coupler. The two transformers are mounted in slots cut on the circuit board. The primary winding (a short length of #10 wire) is soldered to the ends of the stripline. The board is held in position by grounding lugs attached to the ground plane of the board. For improved vhf service the board should be grounded to the box around the entire perimeter.

receptacles are on the ends of the box and the coupling ports are on one side of the box.

A thin strip of shim copper is cut into strips and installed around the edges of the board, shorting the top and bottom copper ground planes together. Two slots are cut in the board so that the toroid transformers can fit directly into the striplines. A short length of #10 bare copper wire serves as the primary winding. It will be soldered in the gap of the stripline when the transformer is completed. The wire is passed through the center of the core after the secondary winding is in place. When the transformers are assembled, they are given a coat of epoxy to secure the windings in place.

The #10 wire is soldered to the ends of the stripline and the ends of the secondary winding are cross connected as shown in the interior view, the opposite end of each winding being grounded to the copper ground plane by a short connection. The free ends of the windings are connected to the opposite strip line as directly as possible.

The coaxial receptacles are mounted on the walls of the box to match the ends of the lines. The circuit board is suspended in the middle of the box, supported by the center terminals of the receptacles and by soldering lugs mounted beneath the receptacle bolts. The lugs are grounded to the board. Two lugs are used on the end receptacles and one lug on each of the side ports.

To calculate the amount of coupling at each port:

> Coupling = 20 Log N, where N = Number of turns on each of T_1 or T_2 where $T_1 = T_2$. Thus a range of coupling can easily be obtained.

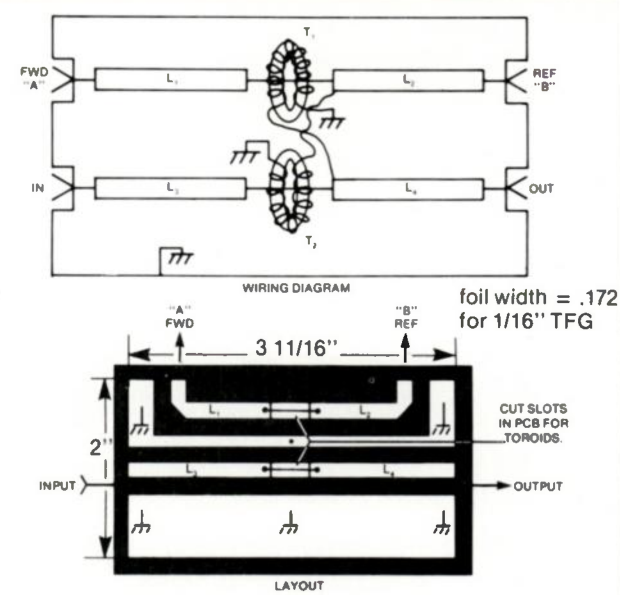


Figure 2. Coupler schematic. Lines L1, L2, L3 and L4 are 50 ohm microstrip lines on 1/16" thick TFG printed circuit board. T1 and T2 are each 10 turns #24 Formvar insulated wire evenly spaced on an Indiana General Toroid Core #F627-8-H (u_i = 850). Turns are counted on inside of core.

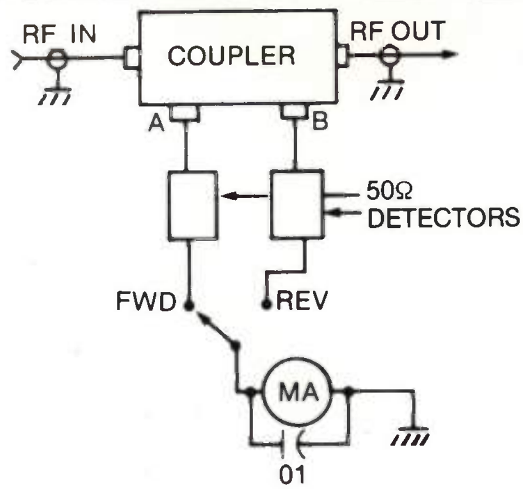


Figure 4. Block Diagram of the Coupler when used as an SWR meter.

The inductance of the grounding lugs limit the upper frequency of operation of this unit and it is suggested that a more complete ground connection be made between the board and the box if the coupler is to be mainly used for vhf operation.

Coupler Calibration

When completed, the coupler should be checked for proper operation. The output and "B" port of the unit are terminated in good 50 ohm loads and a known amount of power is applied to the input fitting. The amount of power applied is limited by the capacity of the load at the output port. The amount of power present is then measured at port "A". This power level should be at least 20 dB lower than the input power. For example, if 100 watts are applied to the coupler, less than 1 watt should appear at port "A".

The coupler is now reversed. The input and "A" ports are terminated in 50 ohms, power is applied to the output port, and the power measured at port "B". Again, this reading should be at least 20 dB less than the applied power. The representative readings of this coupler are summarized in Table 1. Directivity and coupling factor are quite good, comparing favorably with instruments many times the cost of this device.

A block diagram of the use of this coupler as an SWR meter is shown in Figure 4.

Thanks to Dave Wisherd (WA0DAW) and Jerry Stambaugh of Communications Transistor Corporation, and Bill Orr (W6SAI) of Varian EIMAC for their help in preparing this material.

Table 1

F(mHz)	FWD(dB)	REV(dB)	*IL(dB)
>1	19.72	20.27	0.27
3	19.88	20.32	0.22
10	19.92	20.21	0.27
30	19.91	20.10	0.13
100	19.78	19.93	0.19
150	19.44	19.14	0.15

*IL = Insertion Loss Input to Output.

Frequency Synthesized TV Tuning Systems

Motorola has introduced two new parts in a family of phase locked loop frequency synthesizers. The MC6195 and MC6196 are 20 pin IC's designed primarily for CATV or broadcast TV reception. These parts have combined many of the original functions to provide scan up or scan down channel selection, receiver on/off control, automatic active channel seek, automatic switching to AFT mode, BCD channel number output, and audio muting.

The intended applications are shown below.

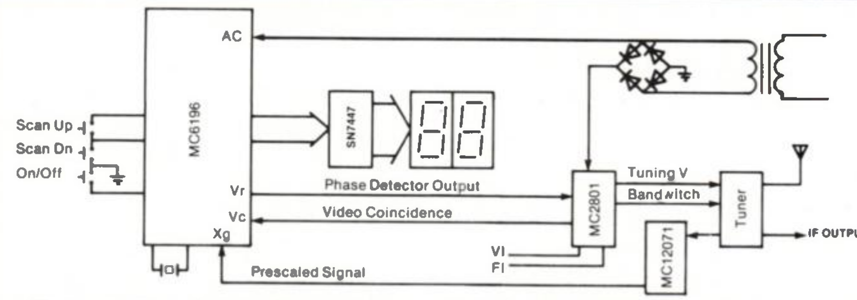


Figure 1. Complete tuning system would include the MC6195/96 Synthesizer, a linear control chip (MC2801) for the 96 or an Op-amp (MLM358) for the 95, ECL high speed prescaler (MC12071), LED decoder/driver (SN74LS47), a Varactor type tuner, and a minimum of external parts.

System Operation

A tuning system which uses an MC6195/96 is very easy to operate. Only three push buttons are required to control on/off, scan up, or scan down.

The on/off function is used to turn the receiver on or off and also blank the channel display. The scan function operates in one of two ways. First is manual operation: when the scan up or down button is pressed and held, the channels will step from one to the next at a two per second rate for the first four channels and then switch to seven and a half per second. This two speed scan is useful if there are a great many active channels. The second is an automatic mode: when up or down is pressed momentarily the channels will step from one to the next but will stop automatically when an active channel is detected.

Circuit Description

The MC6195/96 Frequency Synthesizer is the heart of the tuning system. It contains the control logic for scanning, on/off, and channel number display, along with the phase

Device	Ref OSC	Prescaler VHF	Prescaler UHF	Band	Lowest Channel	Highest Channel
MC6195	4 MHz	256	256	CATV	00	59
MC6196	4 MHz	256	256	USA	02	83

locked loop (PLL), programmable dividers, and internal channel ROM.

The prescaler divides by 256. This signal is inputted to the MC6195/96 which contains a 12 bit programmable divider. The divide number required for each channel is stored in the internal channel ROM. This divided signal is one input to the PLL phase detector. The reference input for the PLL comes from an external 4MHz crystal. This frequency is divided in the chip by 1024. The phase detector compares the reference frequency with the divider tuner frequency and outputs an error voltage which in turn controls the tuner.

Although this is a Phase Locked Loop tuning system Automatic Fine Tuning may be employed, and is controlled by the AFT input. If AFT is held low (Vss) it is defeated giving the PLL full control of the tuning voltage. If AFT is high (Vdd) or open, the PLL will tune the desired channel and when video coincidence is detected, will relinquish control to the AFT circuits in the receiver. This will continue until the loss of video coincidence or a channel change is entered. Note: for AFT to operate the automatic scanning mode must be implemented.

Channel scanning is a single pin input to the chip. If the input is taken to Vdd (5v) the channel counter will increment. If the input is taken to Vss (Gnd) the channel counter will decrement. As stated earlier, the manual mode has a two speed scan. This is useful if the system were used as a CATV converter, because there are a great many active channels. The two speeds allow the user to quickly go from one end to the other but still select the desired channel easily. The automatic mode looks for coincidence between the horizontal flyback pulse and video sync. This occurs only when an active channel is received. If the scan up or down button is pressed momentarily the channel will step from one to the next until video coincidence is detected. Vc is the input which determines the manual or automatic mode. If Vc is tied high (Vdd) the manual mode is implimented. For automatic mode Vc is tied to the video coincidence circuits. Vc is held low until an active channel is detected. This produces a high level which stops the channel advance.

The AC input is a zero crossing detector which is used as the clock for scanning and the multiplex rate for the BCD channel number output. Channels are scanned at either one for every 30 clock cycles (2/sec), one for every 8 clock cycles (7.5/sec), or one for each clock cycle (60/sec). Invalid channels (60-99 or 84-99) are advanced at the 60/sec rate.

The multiplexing for the BCD operates as follows. The MSD is outputted when AC is high (Vdd) and the

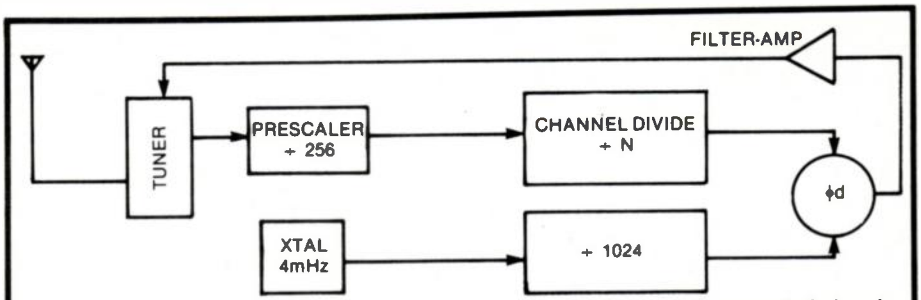


Figure 2. Tuner control is achieved by use of the PLL. One input is a prescaled signal derived from the local oscillator in the tuner.

LSD is outputted when the AC is low (Vss). This allows the common anode displays to be driven from the low side of the power transformer with out additional switching.

Audio muting is a very useful function. As the channels change a great deal of noise is produced from the speaker. SS, is an output which is normally high but goes low while scanning. This action allows muting of the audio during channel changes when the part is designed into the receiver.

Linear Control Chip

The MC2801, Linear Control Chip, is a part which incorporates many functions needed for a complete MC6196 tuning system.

First, this system requires two supply voltages. 5 volts is required for the Tuning Synthesizer and the display driver while 30 or so volts is required for the varactor tuning voltage. The MC2801 has these two regulators.

Next a filter amplifier is required to process the output of the phase detector in the PLL network. The output of this amplifier (Vt) drives the tuner varactor diodes.

The coincidence output (CO) will go high when the video sync signal, inputted to the video input (VI), and the flyback pulse, inputted to the flyback input (FI), are synchronized. "CO" indicates an active channel and is used for auto scanning and AFT switching.

Last is band switching. A tuner requires information as to which band it is tuning. These are, Low VHF (2-6), High VHF (7-13), and UHF (14-83). The synthesizer provides the information and provides a constant current ouput to external transistors for the switching operation.

The MC6195 is slightly different from the 96 in that it is designed to be used in a top of set converter. Here, there will be no video coincidence signals available or band switching information needed. Therefore, the MC2801 may be replaced with an op-amp (MLM358) which will act as the filter amplifier. The complete converter will still require two power supplies which may be implemented by separate voltage regulators.

For additional information on these parts contact: Gary Kloesz, 2-343, Motorola Semiconductor Products Sector, Box 20912, Phoenix, Arizona 85036, (602) 244-6945. Circle INFO/CARD #140.

Three New Series of Mixers

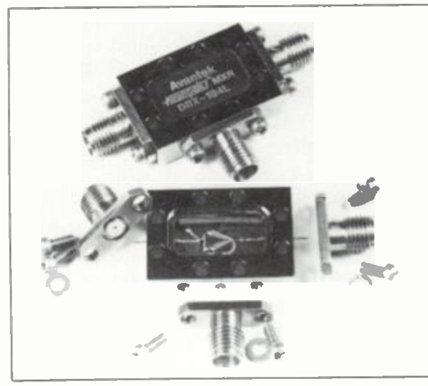
Avantek, Inc., Santa Clara, CA is expanding its line of high-performance double-balanced microwave mixers with three new series designed to fill gaps in available performance for EW and other broadband receiver applications. The basic series available are:

DBX-1824/-18212 Series, featuring 2-26 GHzLO and 2-18 GHzRF response with a choice of 1-12 GHz or .005-4.0 GHzIF range. These versions are designed for superhetrodyne receivers that fold large spectrum segments into high IF processing bands.

DBX-1221 Series, with 2-12 GHzRF/LO response, DC-1.3 GHzIF bandwidth. These mixers are especially suitable for 2-8 GHz downconversion or 5.9-6.4 GHz upconversion.

DBX-186 Series, with 6-18 GHzRF/LO and DC-7 GHzIF bandwidths. A unique design permits overlapping RF/LO and IF frequency coverage with greater than 25 dB RF-IF isolation and constant conversion loss. Useful in 6-18 GHz band-folding applications.

Many of the mixers are available for operation with low (+10 to +13 dBm), medium (+10 to +17 dBm) and high (+13 to +20 dBm) local oscillator drive levels, offering two-tone input intercept points of +15 dBm, +20 dBm and +22 dBm, respectively. All are built with precisely-matched Schottky-barrier diodes, with circuitry fabricated



on a flexible, low-dielectric constant substrate and use an unusual "quasi-planar" layout. This assures excellent electrical symmetry for excellent port-to-port isolation and low intermodulation distortion, as well as a good 50-ohm match at all ports.

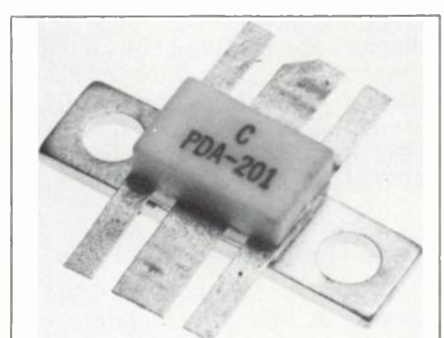
They are packaged in the *Avanpak*™ miniature microwave flatpack which is hermetically sealed and may be used with field-replaceable RF connectors or without connectors for stripline or microstrip drop-in applications. All versions operate over the −55° to +100°C temperature range, may be tested to MIL-STD-202E and are uniquely suited for application in satellites or other equipment requiring high MTBF components.

These *Avanpak*™ microwave mixers are now available, with deliveries from stock to 30 days ARO, and are carried by Avantek's nationwide network of stocking distributors, and handled by Avantek representatives.

For more information contact David Gray, Avantek, Inc., 3175 Bowers Ave., Santa Clara, CA 95051, (408) 496-6710, ext. 2528 or please circle INFO/CARD #139.

High Power RF Differential Amplifier

Communications Transistor Corporation introduced today what is said to be the first power differential amplifier for high frequency use in medium power RF circuits. This new device, called the PDA 201, combines state-of-the-art high power metal ceramic



packaging technology with microwave semiconductor devices to produce a hybrid integrated circuit.

According to Bruce C. Hoffman, CTC Product Marketing Manager, the PDA-201 is the highest-frequency high-gain diff amp on the market today, with uses in a wide variety of applications: as an amplifier (>25dB gain @ 100 MHz; >40dB AGC range); mixer (only 3dBm L.O. power required); frequency multiplier, receiver front end or transmitter (to 1000 MHz). The basic unit can also be combined with other external components to solve many RF circuit design challenges.

Price: $50 per 100. Delivery: In stock.

Communications Transistor Corporation, a subsidiary of Varian, manufactures an expanding line of transistors, resistors, attenuators and terminations. The company is located in San Carlos, California 94070 at 301 Industrial Way or circle INFO/CARD #138.

RF SP10T and SP8T Switches

Dynatech/UZ, a manufacturer of multiple position, electromechanical switches, has produced an engineering breakthrough in eight and ten position coaxial RF switches: a miniaturized SP10T and SP8T switch that operates to 18 GHz.

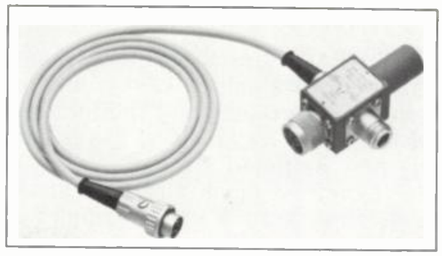
The units (W-10, W-8) have diameters slightly greater than two inches and boast not only compactness and weight savings, but a longer operating life as well. They are guaranteed for a minimum of one million cycles per position, the standard of excellence met by all Dynatech/UZ switches.

Described as a "systems designer's dream" by company executive vice-president, Norm Feigenbaum, the W-10 and the W-8 are engineered for MIL-E-5400, Class 2 environments.

Dynatech/UZ designs and manufacturers a complete line of R.F. switching devices, components and integrated assemblies to meet commercial and military requirements. In addition to off-the-shelf product lines, Dynatech/UZ also creates unique mechanical and electrical designs for special requirements.

For more information, contact Nancy Renfrow at (213) 392-9821 or circle INFO/CARD #137.

SWR Autotester For 75 Ohm Measurements



WILTRON announces another test component for the forgotten engineer who has to work at 75Ω. A new SWR Autotester expands the capability of WILTRON's 560 Scalar Network Analyzer to measure return loss (SWR) of 75Ω test devices. Operating over the 1 to 2000 MHZ range with 40 dB directivity, the unit is available with type N male (560-6N75) or female (560-6NF75) test port connectors. The accuracy resulting from the high directivity is enhanced by the 560 Network Analyzer, which automatically subtracts from test data residuals stored in memory during calibration. Measurement accuracy is 0.01 ± 0.06ρ^2, where ρ is the measured reflection coefficient of the test device.

The SWR Autotester integrates in one small package a broadband, high-directivity bridge, a detector, a precision termination, and a low reflection test port.

Price: WILTRON SWR AUTOTESTER (75Ω) model 560-6N75 or 560-6NF75, $450. Delivery: 30 days.

For additional information, contact: Walt Baxter, WILTRON Company, 805 East Middlefield Road, P.O. Box 7290, Mountain View, CA. 94042-7290, (415) 969-6500, TWX: 910-379-5478 or circle INFO/CARD #136.

AM/FM Telemetry Receiver

New to Aydin Vector's line of telemetry products, the RLS-2000 is a dual conversion superheterodyne AM/FM receiver specifically designed for use in telemetry ground stations. In addition to low distortion, wide bandwidth and excellent frequency response, the RLS-2000 features a synthesized 1st L.O. which enables front-panel thumbwheel-switch-controlled digital tuning from 1435 to 2390 MHz in 0.5 MHz steps.

The RLS-2000 meets the design objectives of flexibility and cost-effectiveness by including in the standard 19" rack-mount chassis all major operating functions—thereby

eliminating the need for plug-in function modules.

A spectrum display, the SD-2000, and an optional factory-installed Pre-D record/playback converter are available. Typical high-performance specifications include: 9 dB maximum noise in any band, 60 dB minimum spurious rejection, 80 dB minimum image rejection and 80 dB minimum dynamic range.

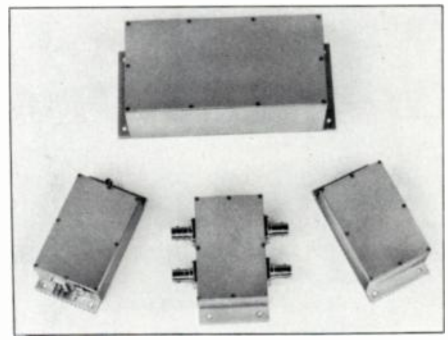
For more information, write or call Aydin Vector Division, P.O. Box 328, Newtown, PA 18940. Telephone: (215) 968-4271 or circle INFO/CARD #135.

RFI Shielded Cases

COMPAC, designers and manufacturers of shielded cases, has expanded its off-the-shelf product line of low cost RFI shielded cases with six new extruded sizes. They are in the COMPAC blank and RFT series and range from 1" × 1" square to 5" × 6" in five heights.

These new extrusions are available in either Iridite or Nickel finish and increase the quantity of basic COMPAC sizes from seventy to one hundred.

COMPAC shielded cases are effective from 60 to ≥ 100dB at 100MHz



and are used in military, industrial and commercial systems and equipment.

For additional information, contact Ken Bardon, Marketing Dept., COMPAC, 279 Skidmore Road, Deer Park, New York, 11729—Tel: 516-667-3933 or circle INFO/CARD #134.

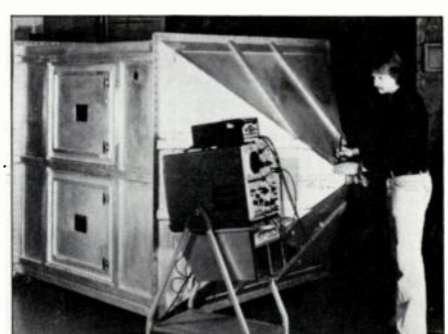
TEM Test Cell

Instruments For Industry, Inc. (IFI), announces its unique CC-101S, the largest Crawford Cell to be commercially produced for laboratory and R.F. susceptibility testing. Suitable for a wide variety of test applications ranging from dc to 100 MHz, the CC-101S provides a controlled environment for generating a homogeneous E-Field.

The IFI CC-101S features a unique square center section, which provides considerably more usuable test area than other standard TEM test cell designs. With dimensions of 148" × 72" × 72", this cell can accomodate objects up to 200 lbs., including large PC board assemblies and heavy black box sub-assemblies.

To prevent hazardous radiation exposures, a side door is provided for insertion of objects under test. Bulkhead transition plates are located at each end of the unit for easy access of cables and connections to the test sample. Provisions for mounting a dedicated E-Field monitoring system; and other custom features to meet customer-specified dimensions are also available in this IFI TEM cell.

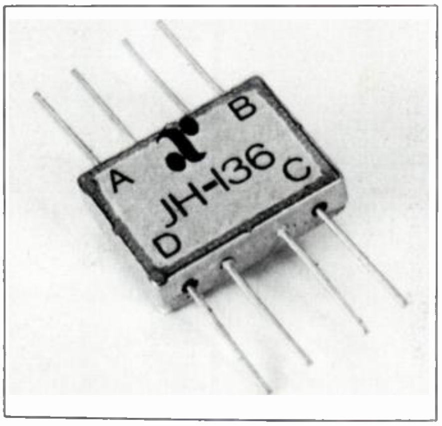
For additional information on the



new CC-101S or other IFI test equipment and broadband amplifiers, contact: Ronald Richards, Vice President, Instruments For Industry, Inc., 151 Toledo Street, Farmingdale, New York 11735; (516) 694-1414. Circle INFO/CARD #133.

Quad Hybrid

Anzac Division of Adams-Russell Company has developed a new quadrature hybrid to cover the 175 - 350 MHz band in a miniature (3/8 × 1/2") flatpack. The JH-136 features typically 25



dB isolation and 0.3 dB loss over the entire octave. Other features include a 1.2:1 VSWR and 2° phase and 0.3 dB amplitude balance.

The JH-136 is a hermetically sealed unit designed for military environments. As with all of Anzac's standard products, delivery is from stock with small quantity pricing of $85.00 each.

Contact Adams-Russell Co., 80 Cambridge Street, Burlington, Mass., 01803, Tel. (617) 273-3333. Circle INFO/CARD #132.

Frequency Synthesizers

A new series of special purpose frequency synthesizers, uses digital locking techniques to produce RF output frequencies in steps from 1 MHz to 10 MHz in bands from .3 to 19 GHz. The Series SLSR, fully automatic and the manually tunable SLSM Series are designed for applications in telecommunications, radar, telemetry and instrumentation systems.

The close-in phase noise, i.e. less than 20 KHz, is determined by the reference noise. The far out phase noise is determined by the high Q voltage tuned cavity oscillator. The phase lock loop bandwidths are set to optimize the overall phase noise performance. As an optional feature, CTI offers a built in 100 MHz VCOX reference that is phase lockable to either a 5 or 10 MHz external frequency standard. Adaptive filter techniques are utilized to reduce microphonic effects caused by shock and vibrations.

Each synthesizer is supplied with a multi-section Thumbwheel Switch. This switch can be supplied to read the RF output frequency, an offset frequency or a channel number.

Unit size is 4.5" × 3" × 4" with base plate mounting dimensions identical to standard phase locked sources.

CTI will customize units to meet a wide range of step sizes, power outputs and frequency bands.

For further information, contact Communications Techniques, Inc., 36 Route 10, East Hanover, N.J., 07936, (201) 884-2580. Circle INFO/CARD #129.

High Current Filter Choke Series

A new series of high current filter chokes in 24 values has been introduced by the J.W. Miller Division of Bell Industries in Compton, California.

Utilizing high saturation flux density rods, the chokes are rated from 5 uH/23 Amps to 250 uH/4 Amps.

The 5500 series chokes are ideal for

use in filtering, energy storage and switching power supply applications.

In addition, J.W. Miller Division produces an extensive line of interference filters and inductors.

Additional information may be obtained from Joe Johnson, J.W. Miller Division, Bell Industries, 19070 Reyes Avenue, Compton, CA 90221. (213) 537-5200. Circle INFO/CARD #131.

Dual Channel Downconverter

Input Frequency:	2.2 – 2.3 GHz
Output Frequency:	215 – 315 MHz
Gain:	25 dB, ± 0.5 dB
Noise Figure:	7 dB typical
Dynamic Range:	+ 30 dBm output intercept
Input/Output VSWR:	1.5:1 maximum
Image and IF Rejection:	> 75 dB

Optional features of model DN 5535 include low noise GaAs FET amplifiers (1.3 dB noise figure), gain and phase tracking, gain control and phase locking to externally supplied 5 MHz station reference.

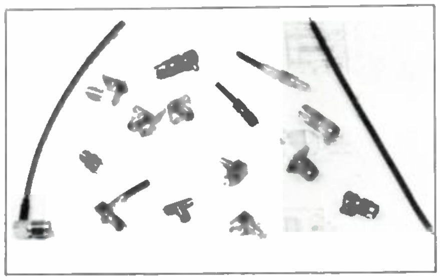
Price:	$7,500
Availability:	60-90 days

Contact MITEQ Inc., 100 Ricefield Lane, Hauppauge, New York 11788-2086, Tel. (516) 543-8873. Circle INFO/CARD #130.

Sub-Miniature RF Connectors

A wide selection of types SMA, SMB, SMC, Sub-Miniature RF Connectors are now being made available by Herman H. Smith, Inc.

Each of the new HH Smith subminiature connector types are made



to meet and exceed Mil-C-39012 and permit the user to make low cost attachments through compression or crimp cable methods in plant, laboratory or in the field.

The new sub-miniature connectors, which are available in more than 65 models, can interface with all other SMA, SMB and SMC types and include a full range of plugs and jacks, receptacles, bulk heads, printed circuits plus in-line and between series adapters.

Plating contacts for the 3 connector types are specified by the manufacturer as gold plated per Mil-G-45204. All SMA bodies are available in gold, passivated and nickel plating while the SMB and SMC type are gold plated per Mil-G-45204. The nickel plate units are also available on special order.

The new Smith Connector line carries a 50 Ohm impedance rating with SMA types having a DC frequency range of 26 GHz while SMB and SMC units have a 2 GHz DC freq. range. Variations on these and other Smith connector types are available on special order. Contact Herman H. Smith, Inc., 1913 Atlantic Ave., Manasquan, NJ 08736, (201) 223-9400. INFO/CARD #128.

High-Power Broadband Pin Switch

WINCOM Corporation has announced the completion of product development on the Model W-8005 High Power SPDT PIN Diode Switch for waveguide applications. The new switch features broadband operation over the 6.5 to 18.0 GHz frequency range and power handling capability of 200 watts CW.

Typical specifications include low insertion loss of 0.7 dB and high isolation of 25 dB. The new Model W-8005 switch is particularly well suited to airborne applications because of its compact size of just 4" × 1.4" × 2", and its ability to operate over an ambient temperature range exceeding 75°C. An integral driver operates from TTL logic commands. Power supply requirements are + 5 VDC and – 30 VDC.

Contact: Robert Antonucci, Wincom Corporation, 23 Shepard Street, Lawrence Industrial Park, Lawrence, Mass. 01842, (617) 685-3930 or circle INFO/CARD #127.

Broadband SP2T

Two new 20-2000 MHz SP2T MIC switches are now available from Daico Industries, Inc. Both devices are built using thin film technology and are available to MIL-STD-883 screening.

Part Number DS0352 is in a 14 Pin DIP while Part Number 100C1558 is in a SMA connectorized package.

Key parameters of both devices include internal TTL drivers with 2 microsecond maximum switching speed, 80 dB typical isolation to 50 MHz, 50 dB to 500 MHz and 30 dB at 2000 MHz. Insertion loss is less than 1 dB typical to 500 MHz and less than 2.5 dB at 2000 MHz. VSWR is better than 1.2/1 typical to 500 MHz and 1.35/1 to 2000 MHz. Maximum RF power is + 10 dBm, both ports are terminated into 50 Ohms in the off state. DC power is + 5 volts at 25 mA, – 5 volts at 25 mA.

For complete specification and information contact Daico Industries, Inc.; 2351 E. Del Amo Boulevard; Compton, CA 90220; telephone (213) 631-1143 or circle INFO/CARD #125.

Coaxial Bench Step Attenuator

Weinschel Engineering's Bench Attenuators, Series 3050, 3051 and 3052 are designed for the precision control of signal levels in the dc to 1.25/2 GHz frequency range.

Offered in six standard ranges and steps from 0 to 1 dB/0.1 dB increments to 0 to 140 dB/10 dB increments, the 3050 Series utilize standard Weinschel

Engineering 3000 Series OEM attenuators as the attenuating device.

The 3050 Series are single-drum units and the 3051 Series are dual-drum units with concentric shafts. The 3052 Series is a twin-drum unit drive in tandem with a special gear assembly.

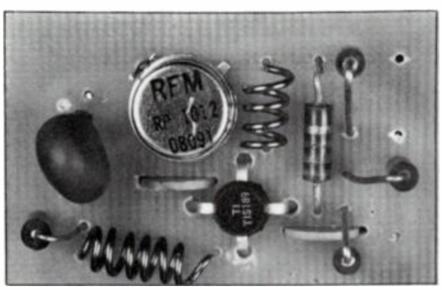
Weinschel patented* detent-mechanism tested to more than a million operations at +65°C is dependable even down to −40°C. Patented** resistive elements on ceramic substrates provide not only a uniquely flat frequency response but consistency of performance. Connectors are BNC female (Type N female optional).

Model 3050 sells for $295; Model 3051 sells for $450.

For more information contact Don Moore, Weinschel Engineering, One Weinschel Lane, Gaithersburg MD 20877, (301) 948-3434 or please circle INFO/CARD #126.

Experimental Oscillator Kit

RF Monolithics, Inc., a Dallas-based leader in the manufacture of Surface Acoustic Wave (SAW) components, has announced the availability of an experimental oscillator kit. The kit



was developed in response to inquiries about the latest oscillator technology from RF engineers throughout the country.

Contained in the kit are an oscillator at 674 MHz center frequency and a SAW resonator to control frequency and eliminate the need for crystal and multiplier chain. The result is a simplified circuit which is always repeatable, uses less power and has excellent phase noise.

The experimental oscillator kit is available immediately at $100 per kit for a quantities of 1-5. Applications/specifications, circuit and assembly instructions are included. Contact RFMonolithics, Inc., 4441 Sigma Road, Dallas, Texas 75234, (214) 233-2903 or circle INFO/CARD #124.

Broadband EMI Filters

U.S. Capacitor Corporation has recently introduced a complete line of 1/4-28 miniature broadband EMI filters, specifically designed to meet or exceed applicable sections of MIL-F-28861 and MIL-F-15733 on high reliability products.

The new line is available in three series, in either hermetic or epoxy fill styles. Case size for the 1/4-28 line is standardized at .375 diameter, although other sizes are available on special request.

All three series offer a voltage range 50WVDC to 300WVDC, 115VAC/400Hz. Standard thread lengths are .187" and .312".

All three series offer a voltage range from 50WVDC to 300WVDC, 115VAC/400Hz. Standard thread lengths are .187" and .312".

For additional information, contact Hosmer at USCC, 11144 Penrose Street, Sun Valley, CA 91352. Phone: 213/767-6770. Circle INFO/CARD #122.

80 Watt/175 MHZ "ISOFET" RF Power Transistor

ACRIAN, INC. announces a complete line of 175 MHz RF/POWER FET'S ranging from 20 watts to 80 watts. ACRIAN'S "ISOFET" technology boasts higher frequency capability, higher gain-bandwidth products, improved stability, increased efficiency while at the same time offering an improvement in VSWR performance over existing products. The "ISOFET" technology incorporates many of the well known process steps used on D-MOS and reduces many of the problems associated with V-MOS. These D-MOS FET's operate at 28 volts from DC to 175 MHz and have a minimum gain of 13db. Products are available now and delivery is from stock to two weeks. Product designations are VMIL20FT, VMIL40FT, VMIL60FT, VMIL80FT, which are 20, 40, 60 and 80 watt "ISOFETS" respectively. Contact Algis J. Juodikis, ACRIAN INC., 10131 Bubb Road, Cupertino, CA 95014, (408) 996-8522 or circle INFO/CARD #123.

Lab Portable Microwave Power Meter

Pacific Measurements, announces a portable R.F. Power Meter. Housed in a small, ¼ cubic foot moisture-proof case of sturdy drawn aluminum and weighing only 8½ pounds, the Model 1034A Microwave Power Meter

offers laboratory accuracy in a rugged, reliable, go-anywhere (AC as well as battery powered) instrument.

Features such as instant-on +10 dBm to −50 dBm measuring capability in seven 10 dB-spaced ranges, self-checking, 1 MHz to 18 GHz frequency range, 50 dB measuring range on one mirror-backed scale, 50 and 75 ohm impedance matching capability, and the ability to handle up to 200 milliwatts of continuously applied power are just a few of the hallmarks of this meter.

Contact Pacific Measurements, Incorporated, 488 Tasman Drive, Sunnyvale, California 94086, (408) 734-5780, or circle INFO/CARD #121.

New Literature

APP Notes

Application Note AN81301, "55 to 85 MHz, 75 OHM IF Amplifier," is now available from California Eastern Labs.

This application note covers typical specifications, circuit design, circuit tuning and discusses the results.

Photographs of the actual amplifier and diagrams are included.

Also available is Application Note AN81901 titled "FET Bias Supply."

This bias supply is a version of the supply described in AN80901 and is intended to be used with the NE137, NE218, NE694, NE695, NE700, NE720, NE8681 series, NE8682 series, NE8684 series and NE869 series.

Included in this application note are specifications and the circuit diagram.

For further information, contact California Eastern Laboratories, Inc., 3005 Democracy Way, Santa Clara, CA 95050, (408) 988-3500. Circle INFO/CARD #108.

Application Note for Users Of Low-Phase Noise Signals

Advancements in modern communications, avionics and radar systems require more signals with low, single-sideband (SSB) phase noise. A new Hewlett-Packard application note describes the techniques used to generate and measure these low-noise signals.

Application Note 283-1 covers both the general concepts of phase noise as well as measurement principles using the HP 8662A synthesized signal generator as a reference. This 45-page document contains 46 drawings and charts, four tables and two photographs.

Comprehensive detail is presented on SSB measurement methods through microwave frequencies including automated systems with software examples. It also illustrates how to use the HP 8662A to generate low-noise signals with multipliers.

Special chapters describe key design aspects and resultant phase noise performance of the HP 8662A; using it for receiver testing; how it can improve the stability parameters of the HP 8901A modulation analyzer, the HP 8505A network analyzer, the HP 8672A microwave signal generator and the HP 5390A frequency stability analyzer. Circle INFO/CARD #107.

Catalog from Anzac

Anzac Division of Adams-Russell announced a new master catalog featuring 352 pages of complete product specifications, application notes and helpful design aids. This new catalog introduces Anzac's next generation of components and many new product lines. RF switches, biphase and quadraphase modulators, digital attenuators, logarithmic RF amplifiers, microwave mixers, and GaAS FET amplifier product lines are introduced. The catalog also contains extensive additions to Anzac's traditional product lines of low noise amplifiers, RF mixers and passive devices such as power dividers, hybrids, and couplers.

As with all Anzac's standard products listed in the catalog, the catalog is now in stock. Contact Adams-Russell, Anzac Division, 80 Cambridge St., Burlington, Mass. 01803, (617) 273-3333 or please circle INFO/CARD #106.

AM/FM Processor Chip

Type ULN-2240A, a very efficient, AM/FM Signal Processor, is the newest Sprague Electric integrated circuit for use in AM/FM stereo systems, particularly in automative applications.

Complete technical information is given in Engineering Bulletin 27121.62 which is available from the Technical Literature Service section of Sprague Electric Company, Marshall St., North Adams, Mass., 01247 or circle INFO/CARD #105.

Millimeter Microwave Antenna Catalog

Alpha Industries, TRG Millimeter Components Division, is offering a 100-page Millimeter Microwave Antenna Catalog and Handbook which provides information to aid engineers in selecting antennas and in the specification of critical parameters.

The publication contains descriptive information and test data on Parabolic, Cassegrain, Lens and Conical Scan Antennas and Horns. Special designs are described including monopulse, arrays and scanning antennas. A glossary contains extensive waveguide information, atmospheric attenuation, definitions of terms for antennas and detailed waveguide flange information.

Contact Alpha Industries, Inc., 20 Sylvan Road, Woburn, MA 01801, (617) 935-5150, TWX 710-393-1236, TELEX 949436 or circle INFO/CARD #100. □

Advertiser Index

Classifieds

PRECISION, PROTECTION, CALIBRATION, SUPERIOR CONSTRUCTION.

INSTRUMENTS FOR INDUSTRY

State-of-the-Art E-Field Testing and Monitoring. Radiation Hazard Monitors, TEM Cells and E-Field Sensors. All from IFI.

The EFS and RHM series of E-Field Sensors provide direct readout in the near and far field for accurate measurements, both in and out of the laboratory environment. IFI TEM Cells, when used for RF susceptibility testing, permit safe generation of extremely high E-Field levels with a moderate amount of RF power.

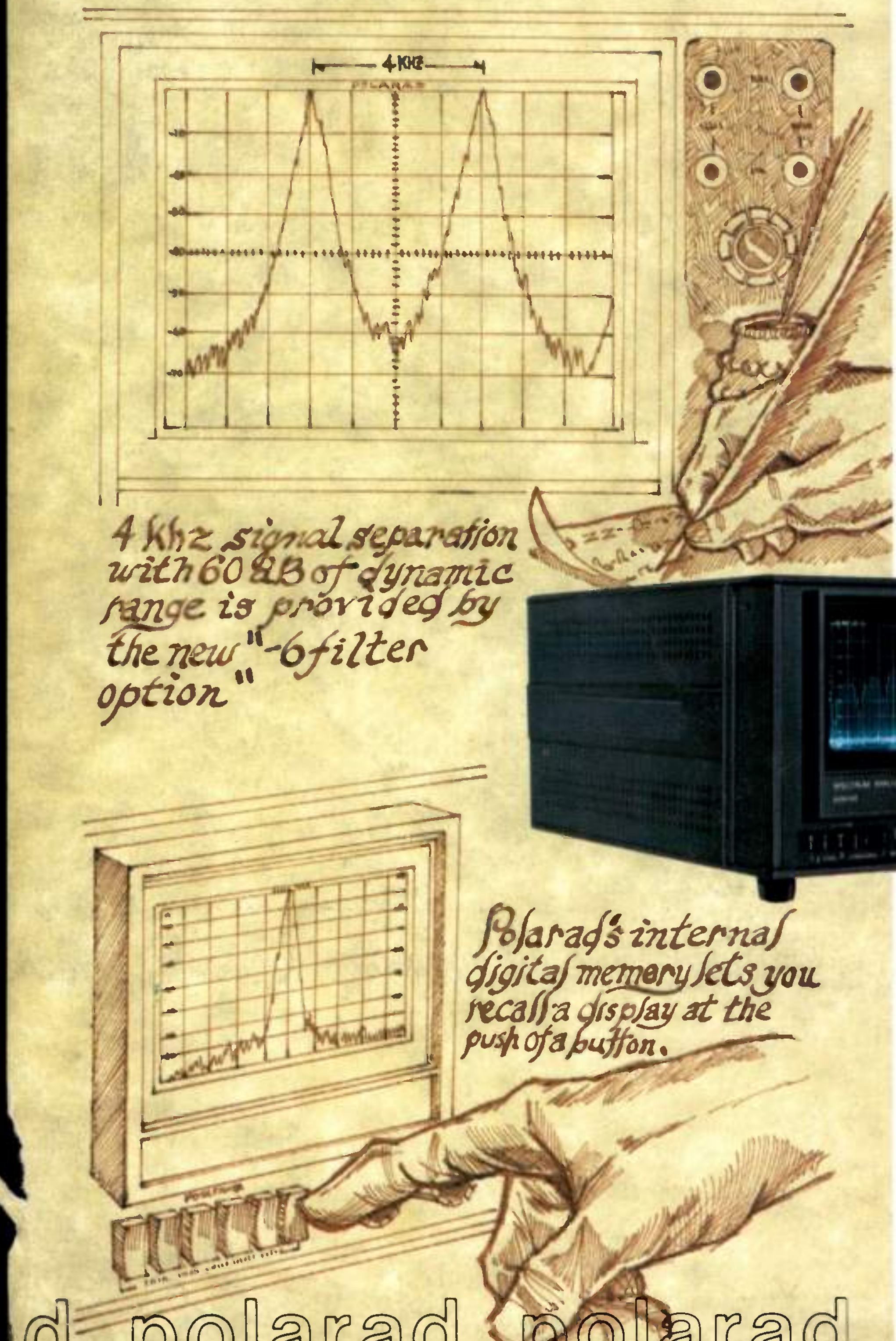
RADIATION HAZARD MONITORS		E-FIELD SENSORS		TEM CELLS	
FEATURES	**APPLICATIONS**	**FEATURES**	**APPLICATIONS**	**FEATURES**	**APPLICATIONS**
•No tuning or bandswitching •Broadband coverage •Neglible Field Distortion •Direct reading between 1-300 volts per meter •Isotropic response •Individual vector selection	For measurement of strong electric fields in the VLF through VHF regions of the spectrum. An optional dielectric handle will permit it to be used as a probing device for sources of radiation.	•No tuning or bandswitching •Compact, battery operated •System compatible •No calibration curves or tables •Five meter ranges for easy reading •Fiber optic remote readout available	Ideal for radiated susceptibility testing. A versatile, cost-effective instrument, the EFS-1 monitors electromagnetic environments, measures field strength, and more.	•Available in 10 different size models •Rugged, commercial construction •Easy access of of cables and connections to test sample •Side door allows for safe product insertion	Suitable for various test applications ranging from dc to 1 GHz, IFI has produced the first commercial TEM Cell for lab and RF susceptibility testing. The IFI TEM Cell offers an efficient means of accurate and broadband E-Field testing.

Instruments for Industry, Inc.
151 Toledo St., Farmingdale, N.Y. 11735 (516) 694-1414

"Depend on IFI"

INFO/CARD 31

CLEARER, SHARPER PICTURES.



Polarad's new 600B Series Spectrum Analyzers offer enhanced signal resolution for close-in analysis.

Polarad has developed a series of higher resolution filters for close-in measurements of carrier and/or signals offered as "Option –6".

And with Polarad, you get the digital memory that's far superior to variable persistence types... high resolution with continuously updated displays — without blooming, fading or smearing.

Polarad's 600B Series of Spectrum Analyzers...the performance, ease of use and versatility you need, at a price you can afford.

Call us for a demonstration and quotation today or write for further informaton and specifications.

INFO/CARD 32

Selection Guide

Model	Frequency
632B-1	100 kHz-2 GHz
630B-1	3 MHz-40 GHz
640B-1	3 MHz-40 GHz

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Polarad Electronics, Inc. 5 Delaware Dr. Lake Success, N.Y. 11042 Tel. 516-328-1100 TWX 510-223-0414