May/June 1983

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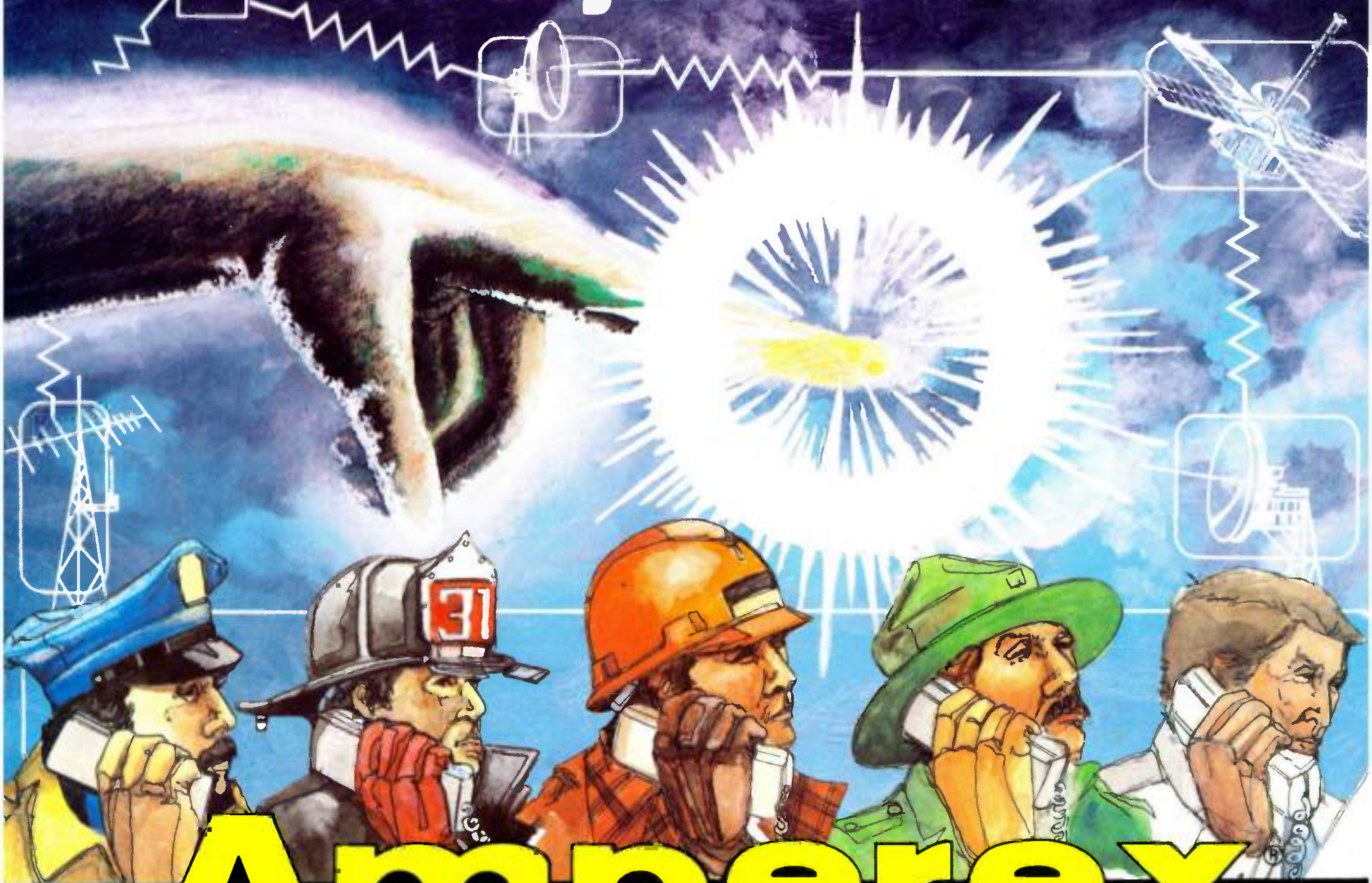

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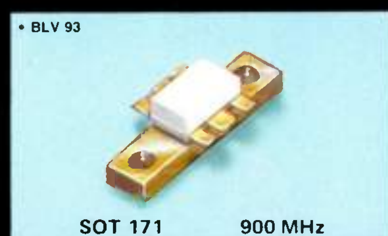
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

INFO/CARD 1

APPLICATION: 900 MHz Radio Telephone
POWER: 2 Watts
GAIN: ≥ 6.5 dB
FREQUENCY: 900 MHz
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

INFO/CARD 2



APPLICATION: 900 MHz Radio Telephone
POWER: 8 Watts
GAIN: ≥ 6 dB
FREQUENCY: 900 MHz
V_{CE}: 12.5 Volts
EFFICIENCY: ≥ 50%

INFO/CARD 3

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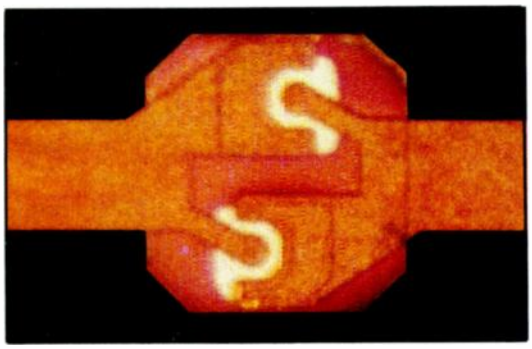
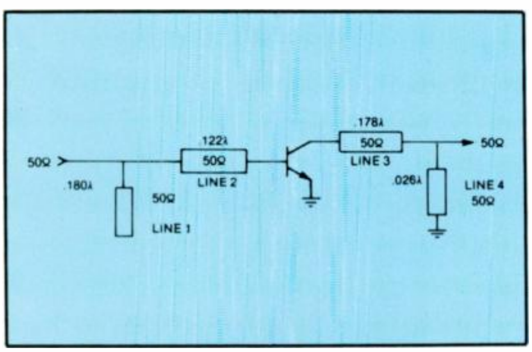
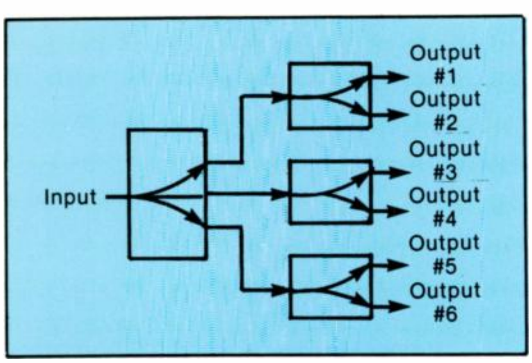


Cover Story, p. 76.



Calculator Aided Stub Matching, p. 30



"Magic Tee", p. 44

Cover

Features

Departments

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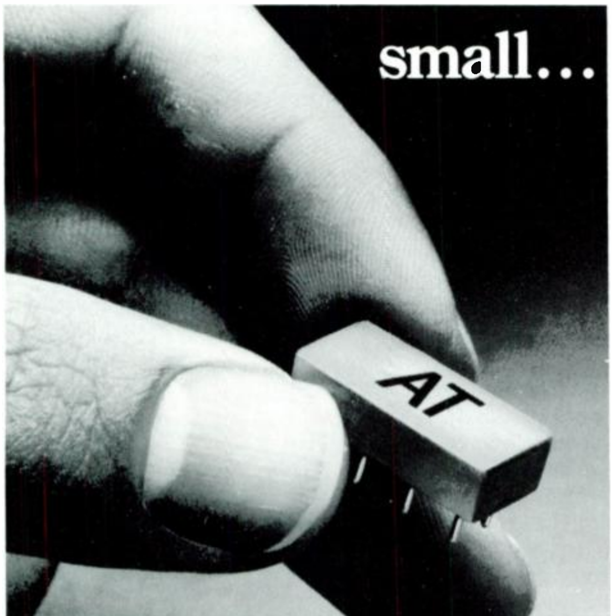

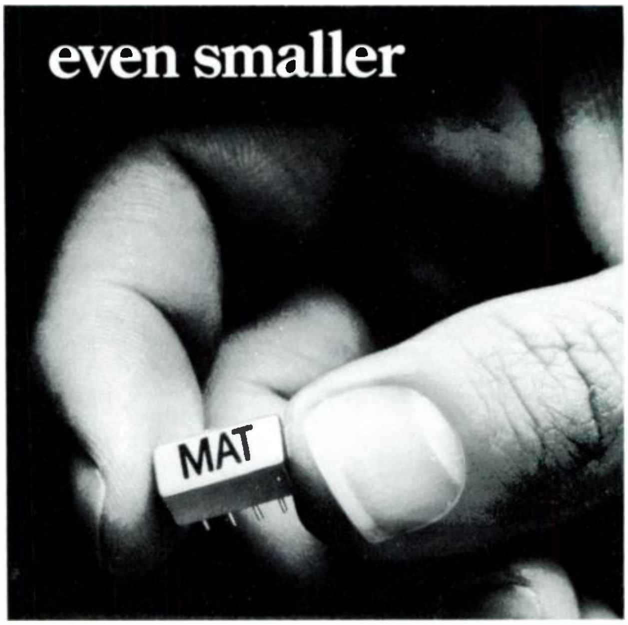

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AT-3	MAT-3	3	±0.2dB	0.6dB	1.0dB	1.3:1	1.5:1
AT-6	MAT-6	6	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-9	9	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
AT-10	MAT-10	10	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-12	12	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-15	15	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
AT-20	MAT-20	20	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
AT-40*		40	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1

*AT-40 Frequency Range DC-500 MHz
Frequency Range: DC-1500 MHz, power (max.) MAT = 0.5W, AT = 1W

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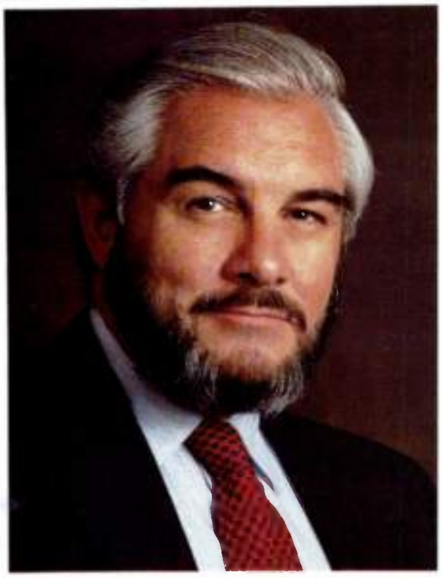
INFO/CARD 5

C-96 REV A

Bigger and Better



By Bill W. Childs
Publisher

It's been just a year now since I last wrote an editorial to you. In that last editorial I discussed our hiring of Larry Brewster as editor and prior to that an introduction of our new staff. We as a team pledged to make some positive changes in our magazine format and circulation. We also promised to maintain and if possible exceed our previous efforts in editorial content.

Looking to the future I would like to report that our July/August issue will address the issues of radiated and conducted radio frequency and electromagnetic interference under the general heading of EMC/EMI. Moving to our September/October issue we will focus on computer modeling and the use of calculator programs for increasing the efficiency of the RF design engineer. We will then wrap the year up with November/December covering RF Components. We believe these to be important topics and look forward to bringing this editorial focus to our readers.

This issue of *r.f. design* magazine represents the biggest and one of the finest magazines we have ever published for our readers. I am very pleased with the significant efforts of our editor Mr. Brewster and our sales staff Ruth Breashears, Blair Sutfin and Susan Myers along with our editorial assistant Tami Baird. Thanks, however would not be complete without giving considerable praise to our authors and our editorial review board.

As most of you know, any undertaking like this is a continuing process over time. Any time we can, it's perhaps wise to take snap shots of where we are and then think about where we are going.

In our September/October issue of last year we contracted with READEX a national readership research group to contact our readers and bring back a sample of your comments, needs and thoughts on the subject of *r.f. design*. The following is a small sample from their findings.

- 94 percent of our readers responded that *r.f. design* contains information useful to the job that is not found in any other publication.
- 57 percent of our readers save the entire magazine for future reference and 42 percent clip articles and save them.
- *r.f. design* magazine is passed on to an average of 2.3 other interested readers.
- 94 percent of our readers feel that the technical level of articles is just about right.

Other responses covered future article needs and frequency ranges that our readers work with on a daily basis. We were very pleased with the READEX responses and we look forward to incorporating new ideas in the magazine as we progress. We shall also be very conscious of maintaining the things about *r.f. design* which you the reader have come to expect.

My special thanks to you the reader who have loyally followed *r.f. design* and told us about it, and also our advertisers which make it all possible. We look forward to a bigger and better rest of 1983 and an even greater 1984.

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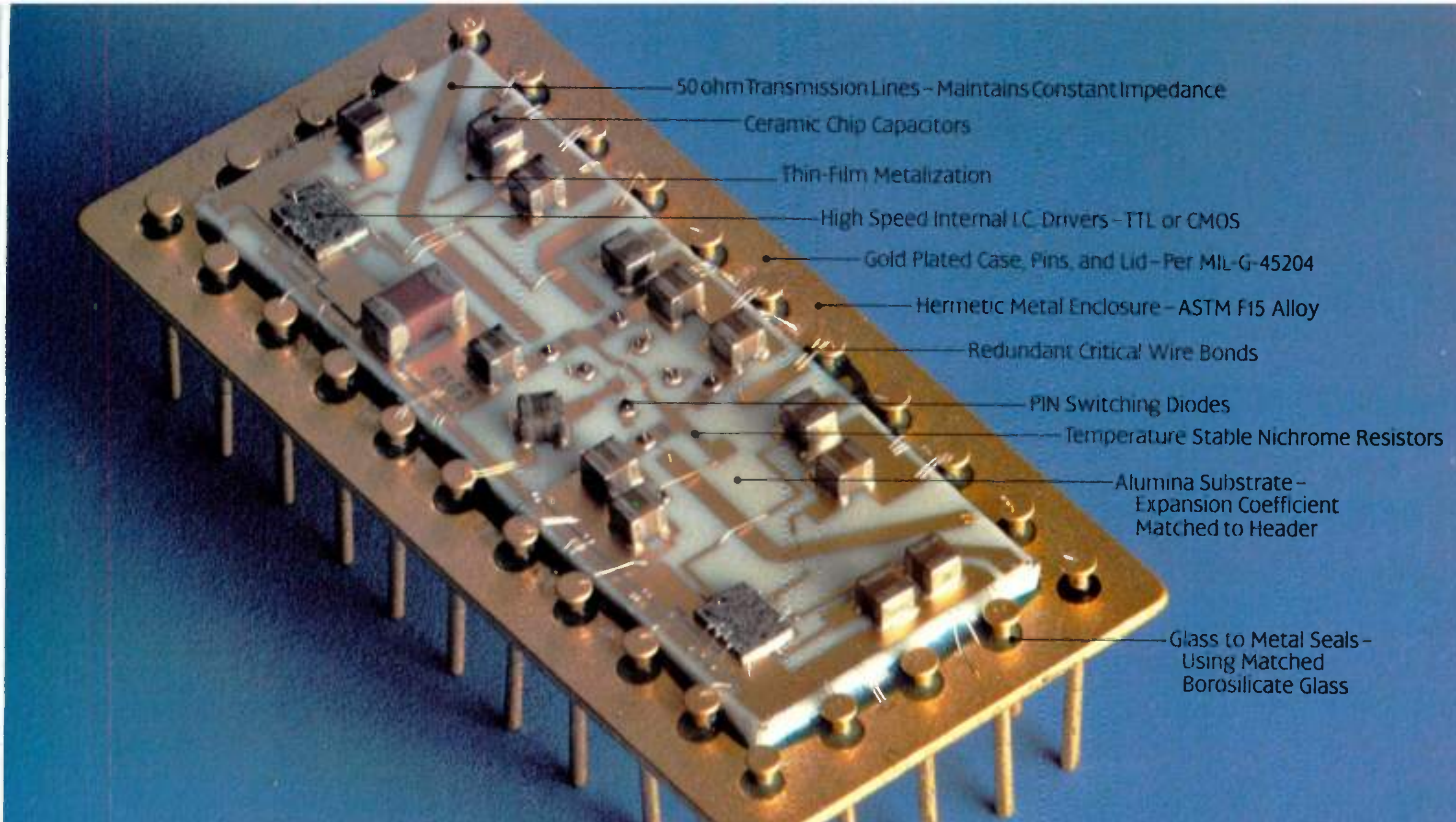


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INFO/CARD 6

Dear Larry:

I have had several phone calls from engineers who have been constructing The Linear Phase Filter I described in your May/June and July/August 1982 issues. The complaint was that the filter was almost but not quite working properly. I was able to trace the problem to an error in the equation for filter impedance level in Step 4 (Page 31 of the July/August 1982 issue). The equation reads:

$$R(\text{filter}) = (4/\pi)\,(Xc)\,|\,f = 2\,\Delta\,f$$

It should read:

$$R(\text{filter}) = (\pi/4)\,(Xc)\,|\,f = 2\,\Delta\,f.$$

I have carefully reviewed the rest of the equations and they are correct. Thank you.

Robert W. Sellers

Errata

The following are corrections to the "RF Amplifier Design Program" article in the March/April 1983 issue.

Example I, page 34, corrected sequence.

XEQ "AMP"

ENTR CMND
1-LSF
2-OPT TERM
3-GAINS
4-YIN/OUT
5-CHNG YS/L
6-CHNG YPAR
7-K-STERN
8-YS/L FOR K
9-S/Y CONV

CMND?

1. RUN

ENTER Y-PARAMS

GI = ?

22-03 RUN

BI = ?

9-03 RUN

GR = ?

0 RUN

BR = ?

– 2.2-03 RUN

GF = ?

40-03 RUN

BF = ?

– 185-03 RUN

GO = ?

1-03 RUN

BO = ?

8-03 RUN

L.S.F. = 923.3E-3

CMND?

2. RUN

OPTIMUM VALUES

L.S.F. = 923.3E-3

GS = 86.61E-3
BS = – 53.00E-3

GL = 3.937E-3
BL = – 10.00E-3

GMAX = 17.59E0 dB

CMND?

4. RUN

GIN = 86.61E-3
BIN = 53.00E-3

GOUT = 3.937E-3
BOUT = 10.00E-3

CMND?

5. RUN

ENTER YS AND YL
GS = ?

20-03 RUN

BS = ?

—50-03 RUN

GL = ?

1.5-03 RUN

BL = ?

5-03 RUN

CMND?

3. RUN

OPERATING GAIN
GP = 9.509E0 dB

AVAILABLE GAIN
GA = 16.27E0 dB

TRANSDUCER GAIN
GT = 4.952E0 dB

CMND?

Appendix I, Main Programming "AMP", page 38, program reproduction.

```
01*LBL "AMP"
02 CF 04
03 CF 03
04 ADV
05 "ENTR CMND"
06 AVIEW
07 "1-LSF"
08 AVIEW
09 "2-OPT TERM"
10 AVIEW
11 "3-GAINS"
12 AVIEW
13 "4-YIN/OUT"
14 AVIEW
15 "5-CHNG YS/L"
16 AVIEW
17 "6-CHNG YPAR"
18 AVIEW
19 "7-K-STERN"
20 AVIEW
21 "8-YS/L FOR K"
22 AVIEW
23 "9-S/Y CONV"
24 AVIEW
25*LBL 00
26 ADV
27 FIX 0
28 "CMND?"
29 PROMPT
30 ENG 3
31 GTO IND X
32*LBL 01
33 FC? 03
34 XEQ "PARA"
35 ADV
36 XEQ "LINV"
37 GTO 00
38*LBL 02
39 FC? 03
40 XEQ "PARA"
41 XEQ "OPT"
42 GTO 00
43*LBL 03
44 FC? 03
45 XEQ "PARA"
46 FC? 04
47 XEQ "YS/L"
48 XEQ "YI/O"
49 XEQ "GP"
50 XEQ "GA"
51 XEQ "GT"
52 GTO 00
53*LBL 04
54 FC? 03
55 XEQ "PARA"
56 FC? 04
57 XEQ "YS/L"
58 XEQ "YI/O"
59 "GIN = "
60 ARCL 23
61 AVIEW
62 "BIN = "
63 ARCL 22
64 AVIEW
65 ADV
66 "GOUT = "
67 ARCL 25
68 AVIEW
69 "BOUT = "
70 ARCL 24
71 AVIEW
72 ADV
```

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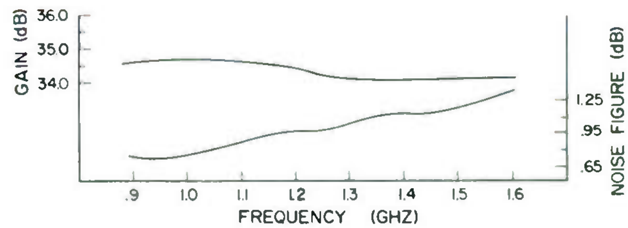
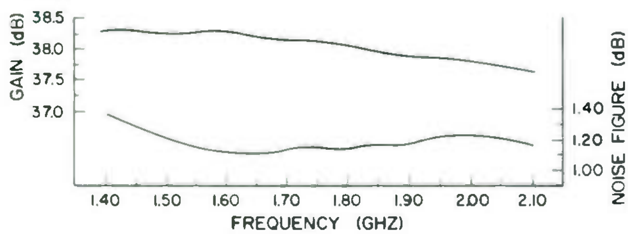
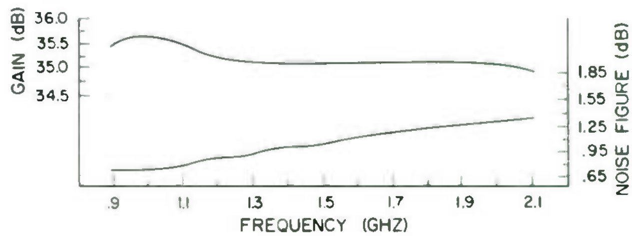
Frequency: 1.0-1.5 GHz
Gain: 33 dB
Gain flatness: ±.5 dB
Noise figure: 1.3 dB max.
Input/output VSWR: 1.33:1
Dynamic range: +10 dBm

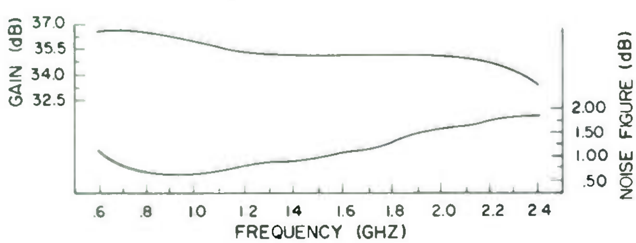




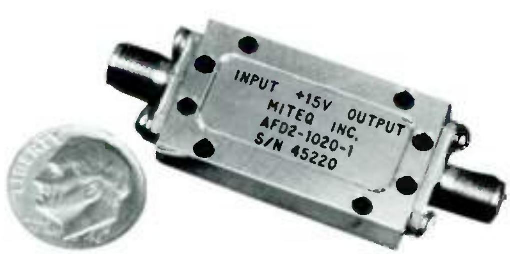
Frequency: 1.5-2.0 GHz
Gain: 32 dB
Gain flatness: ±.5 dB
Noise figure: 1.5 dB max.
Input/output VSWR: 1.33:1
Dynamic range: +10 dBm

Frequency: 1-2 GHz
Gain: 30 dB
Gain flatness: ±.6 dB
Noise figure: 1.6 dB max.
Input/output VSWR: 1.50:1
Dynamic range: +10 dBm





Frequency: .8-2.2 GHz
Gain: 30 dB
Gain flatness: ±.75 dB
Noise figure: 2.0 dB max.
Input/output VSWR: 2.0:1
Dynamic range: +10 dBm



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INFO/CARD 30

(Errata . . . Continued from page 8)

```
73 GTO 00
74*LBL 05
75 XEQ "YS/L"
76 GTO 00
77*LBL 06
78 XEQ "PARA"
79 GTO 00
80*LBL 07
81 FC? 03
82 XEQ "PARA"
83 FC? 04
84 XEQ "YS/L"
85 XEQ "STN"
86 ADV
87 "K-STERN = "
88 ARCL 17
89 AVIEW
90 GTO 00
91*LBL 08
92 FC? 03
93 XEQ "PARA"
94 XEQ "ITR"
95 ADV
96 XEQ "S/LPR"
97 GTO 00
98*LBL 09
99 XEQ "SPAR"
100 XEQ "PAR1"
101 GTO 00
102 END
```

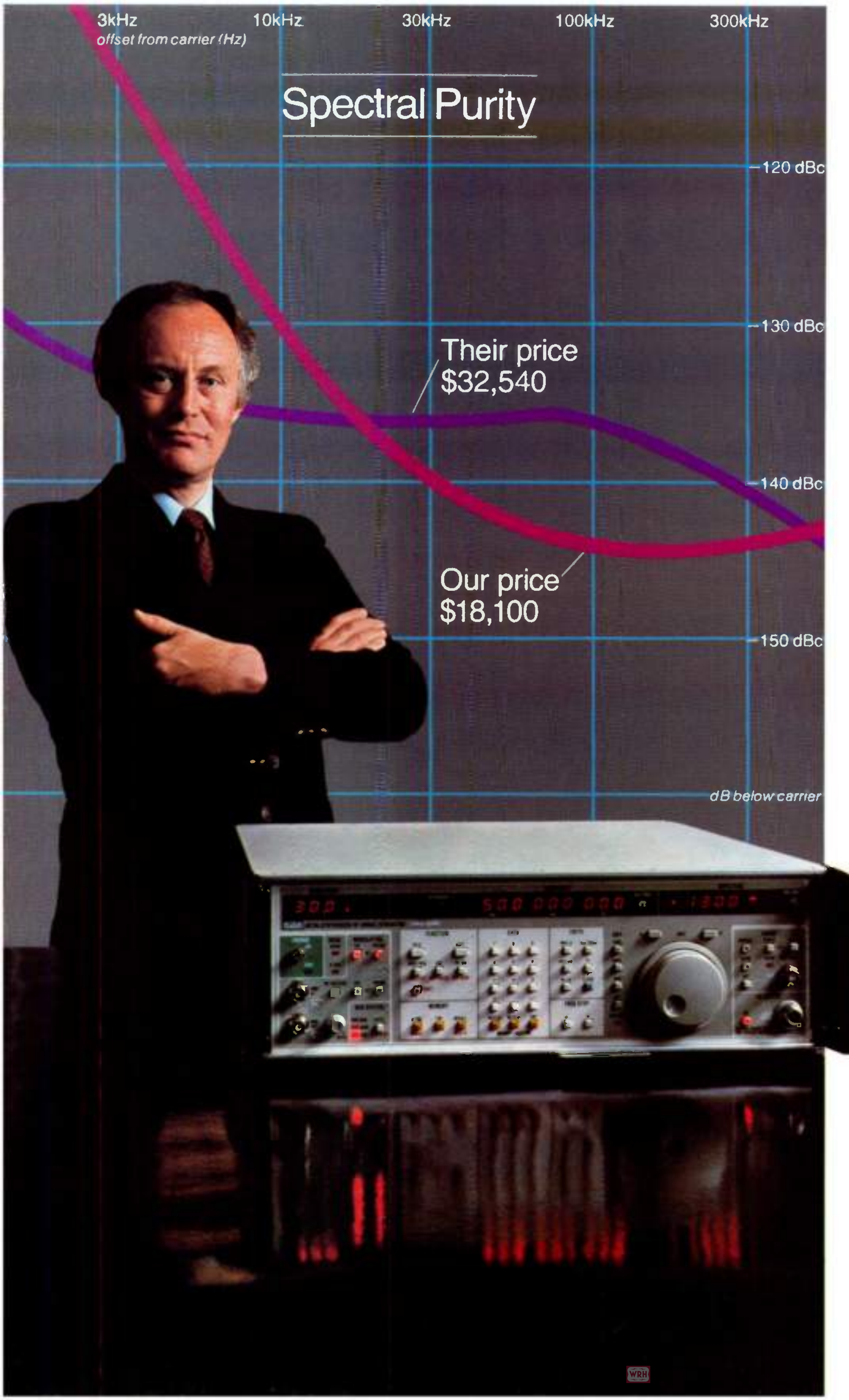
Appendix I, page 38, program listing "LINV", program reproduction.

```
01*LBL "LINV"
02 CF 00
03 RCL 03
04 X<0?
05 GTO 01
06 RCL 05
07 X<0?
08 GTO 01
09 RCL 12
10 RCL 14
11 /
12 STO 16
13 X<0?
14 SF 00
15 1
16 X<>Y
17 X>Y?
18 SF 00
19 "L.S.F. = "
20 ARCL X
21 AVIEW
22 ADV
23 RTN
24*LBL 01
25 SF 00
26 "UNSTABLE"
27 AVIEW
28 "GI/O<0"
29 AVIEW
30 END
```

Program listing "SPAR", page 45, omitted numbers
Line 127 RCL IND 26
Line 129 STO IND 26

IEEE-488

Superior spectral purity at half the price. Is there really any other choice?



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INFO/CARD 9

PLL Primer

Part I began the discussion of phase lock loops. Part II now considers some design parameters and possible problems.

By Andrzej B. Przedpelski
A.R.F. Products, Inc.
R&D Laboratory
Boulder, Colorado

From equations 4, 5 and 6 (Part I) it can be seen that T2 T3 are functions of desired phase margin (stability) and bandwidth; both constant. Time constant, T1, however, is also a function of K_p, K_v and N. The phase comparator gain, K_p, is essentially constant; thus, only the ratio K_v/N should be kept constant for stable operation. The problem is obviously most complicated in high frequency synthesizers, since usually both K_v and N vary. If you are one of the lucky ones, the two will compensate for each other, but the rest of us have to use some form of compensation. Two possible methods involve either varying R1 (Figure 6), since it affects T1 only, or compensating K_v by using varying gain circuit in the VTO input. This second method is usually used and several non-linear gain circuits are described in the literature[11]. The circuits have to have adequate bandwidth not to introduce additional poles below the lowest VTO modulation response pole.

Sidebands

If N is not unity, sidebands at the reference frequency, F_R, will be present in the output. The mechanism is quite simple. As can be seen in Figure 1, only the phase comparator and the integrator stand between the input and the VTO. Thus, any leakage of F_R will frequency modulate the VTO directly. The phase comparator does not have any frequency discrimination within the PLL range of frequencies. However, it has inherent isolation between the F_R input and its output. If a linear phase comparator (double balanced mixer) is used, the usual 25-35 dB isolation can be expected. It should be remembered that the same isolation also has to exist between the NF_R/N port and the comparator output. If a digital phase comparator is used, approximately the same order of magnitude of isolation can be expected. Theoretically, the isolation in either case is infinite, if the circuits are properly balanced, the switching of the phase comparator is ideal and an ideal active integrator is used. Since this is not the case in real life, there is not much that can be done to reduce F_R leakage through the phase comparator, other than the usual "idealizing" of the circuit. The integrator performance, on the other hand, is under the designer's control. From the integrator response (Figure 7), it is apparent that as the phase margin of the loop is decreased, the attenuation of the reference frequency increases. Unfortunately, the phase margin cannot be decreased indiscriminately, since it affects other loop parameters, such as stability and step response. However, a considerable amount of additional attenuation can be obtained, even with comfortable phase margins, over the usual second order loop (T3 = 0); in the above example over 20 dB. It can also be shown that decreasing loop bandwidth also increases F_R attenuation. Again, this parameter has to be a compromise, since it affects loop step response and acquisition time. Adding broadband gain before or after the integrator does not have any overall effect. While the effective K_p or K_v will change, T1 will have to change accordingly (for the same bandwidth and phase margin) and the overall attenuation to the reference frequency will remain the same.

The only avenue left open is the addition of frequency discriminating circuits between the integrator and the VTO. This has to be done very carefully not to reduce loop stability excessively. This technique is described in more detail in Reference 12.

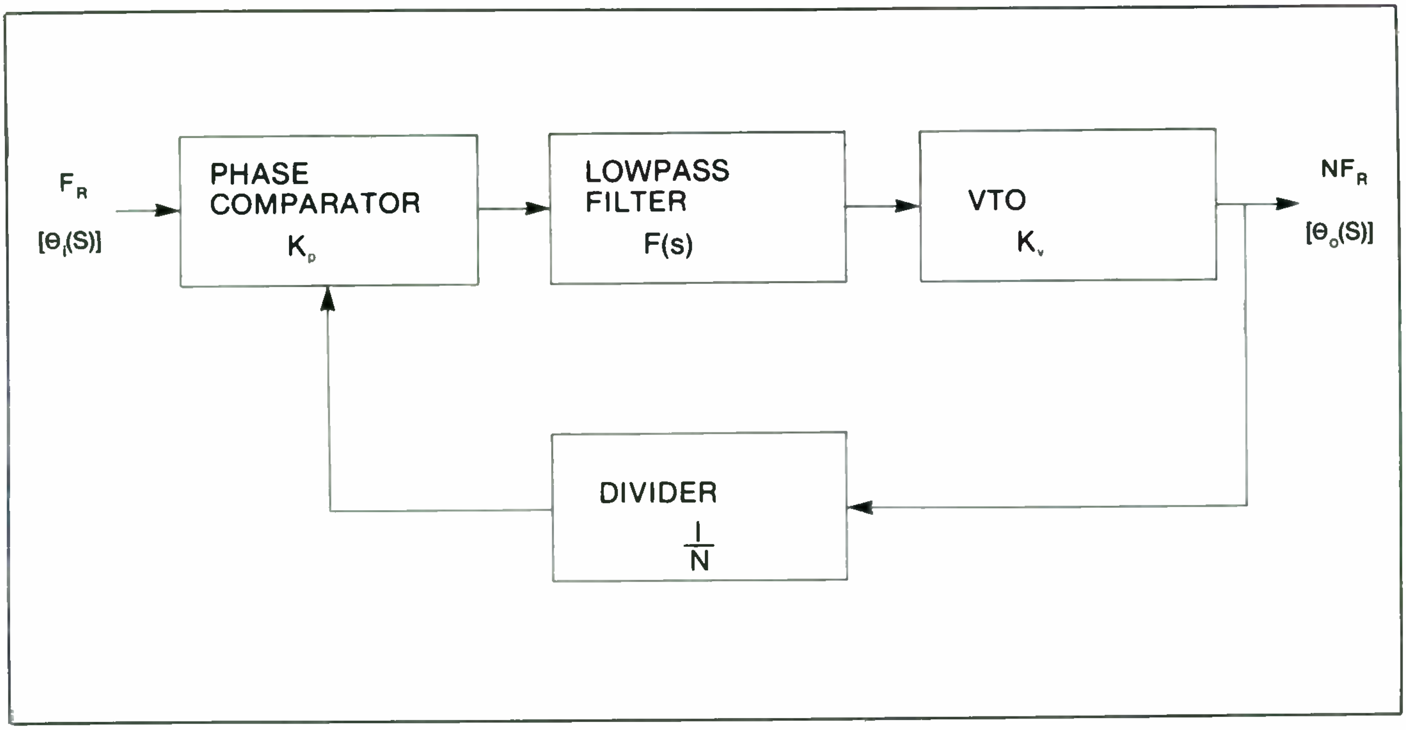

Figure 1. General Phaselock Loop Configuration.

Phase Noise

The noise in the output of a PLL signal source is a very important parameter. Fortunately, methods exist[13] to evaluate the noise performance of a PLL if the noise characteristics of the VTO and the reference are available. Basically, the phase noise in the output signal, within the loop bandwidth, is equal to the reference noise multiplied by N (the division factor within the loop and overall frequency multiplication). The noise level, above the loop bandwidth, is essentially that of the VTO. While the above may be an oversimplification, the overall concept is very useful in obtaining very low phase noise signal sources. This is done by combining, in a PLL, a reference frequency source having a high long term stability (low noise at low Fourier frequencies) with a VTO with good short term stability (low noise at high Fourier frequencies), as shown in Ref. 10.

Instability

For narrowband PLLs, the program of Table I calculates values for T1, T2 and T3 which practically guarantees stable loop operation, providing reasonable values are used for phase margin (above 45 degrees). For wider band applications, the operational amplifier open loop time constant (open loop bandwidth) and DC gain, as well as the VTO control time constant (modulation bandwidth), have all to be considered. The program of Table II takes all these circuit characteristics into account and provides a convenient method of circuit design verification.

Open loop gain = $G(j\omega)\,H(j\omega)$ =

$$= \frac{K_pK_vA_o}{N\omega}\left[\frac{j\omega C + 1}{j[\omega^2(\omega^2 D - F) + 1] + \omega(\omega^2 E - L)}\right] \qquad (7)$$

where $C = T_2 + T_3$

$D = T_oT_vT_1T_3$

$E = T_v[T_o(T_1 + T_2 + T_3) + T_1T_3(A_o + 1)] + T_oT_1T_3$

$F = T_v(A_oT_1 + T_o + T_1 + T_2 + T_3) + T_o(T_1 + T_2 + T_3) + T_1T_3(A_o + 1)$

$L = A_oT_1 + T_o + T_1 + T_2 + T_3 + T_v$

$T_1 = C_1 \times R_1; T_2 = C_1 \times R_2; T_3 = C_2 \times R_2$

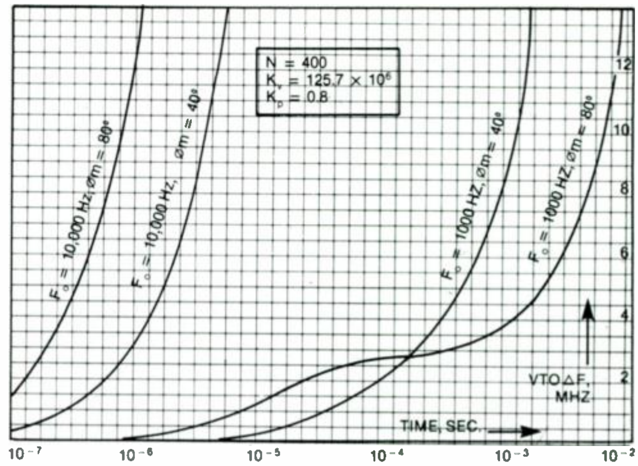

Figure 11. Typical VTO Slew Rate.

$$\text{Closed loop gain} = \frac{G(j\omega)}{1 - G(j\omega)\,H(j\omega)} \qquad (8)$$

$$\text{VTO noise reduction} = \frac{1}{1 + G(j\omega)\,H(j\omega)} \qquad (9)$$

The open loop gain and phase margin vs. frequency can be calculated and, if the stability margin is inadequate, the calculation of T1, T2 and T3 can be repeated (Table I program) using a larger nominal phase margin value and/or a lower value for F_o, to compensate for these parasitic effects. It may prove to be impossible to obtain stable operation using a given operational amplifier and VTO (for a given bandwidth). It will be then necessary to either obtain more desirable devices (larger bandwidths and higher amplifier gain) or reduce loop bandwidth (lower the value of F_o). The above analysis assumes that the standard precautions against stray pick-up and spurious reactances are taken. Any such parasitic effects will reduce loop stability and possibly cause oscillations. If a frequency synthesizer design (variable N) is being analyzed, additional factors have to be taken into account: variable N and K_v. If K_v/N is not a constant, the calculations have to be performed at several frequencies to ascertain that stability at all possible frequencies is assured. This means not only the steady state values within the operating range, but also all other values within the total VTO range (stop to stop), in all possible combinations.

Modulation

Conventional methods of AM modulation can be used. It is prefered that the VTO is not directly modulated, for obvious reasons. Either absorption modulation or modulation of a succeeding stage can be employed.

FM modulation is somewhat more complicated. If the range of modulating frequencies lies above F_o, the modulating voltage can be summed with the control voltage at the VTO input. Modulating frequencies below F_o will be attenuated in accordance with noise reduction response as calculated using Table II program. Obviously, modulation

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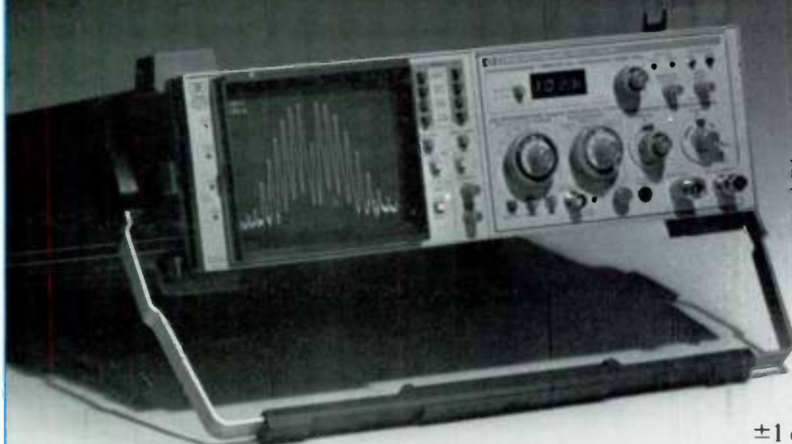

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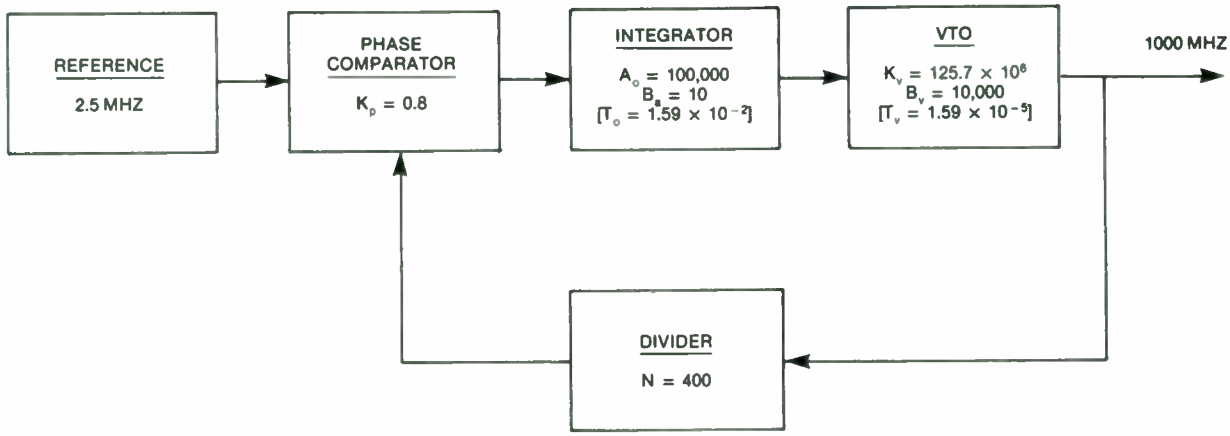

ORDER		T_1	T_2	T_3	CURVE
3	$\phi_m = 40°$	1.37×10^{-2}	2.67×10^{-4}	7.42×10^{-5}	A
3	$\phi_m = 80°$	7.28×10^{-2}	1.81×10^{-3}	1.39×10^{-5}	B
2	$\alpha = 0.707$	1.54×10^{-2}	3.49×10^{-4}	0	C

Figure 12. Typical PLL Example.

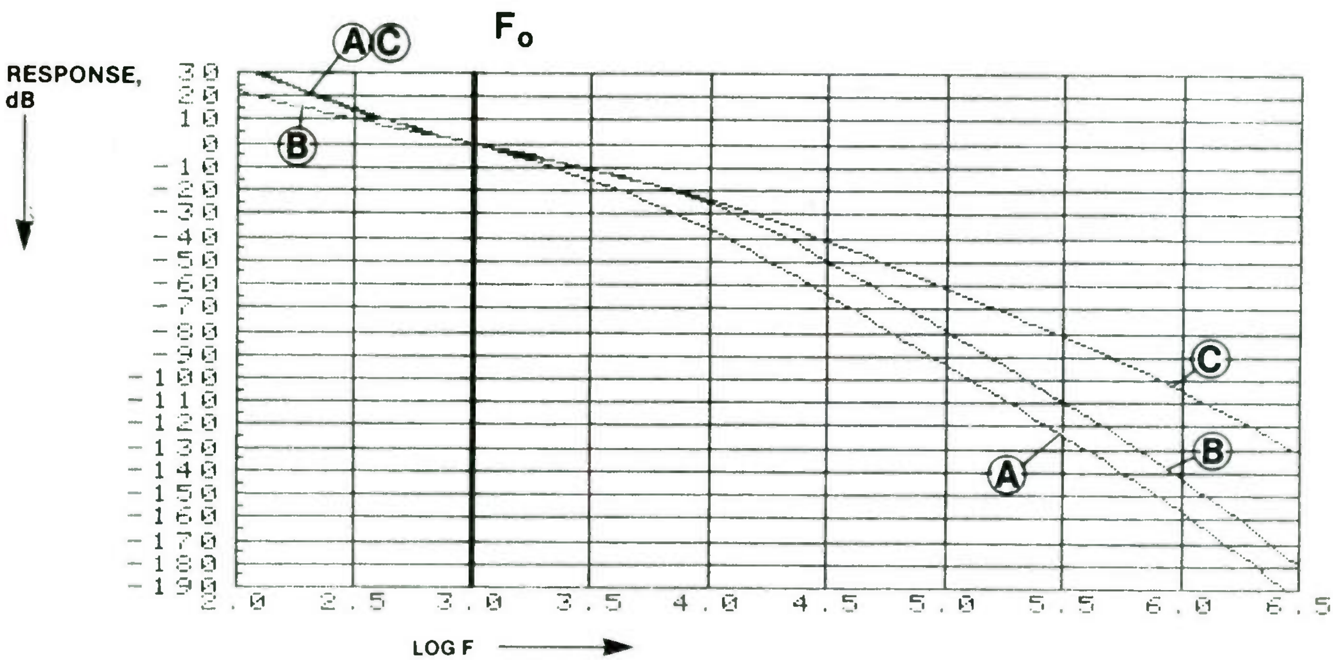

Figure 13. Open Loop Response.

down to DC is not possible using this method. If low modulation frequencies have to be accommodated, the reference has to be directly modulated. In this case the higher modulation frequencies will be attenuated following the closed loop response as calculated using Table II program. While the deviation of the reference (if it is the usual crystal oscillator) is limited, the output deviation will be multiplied by the loop division factor, N. A combination of both methods can be used, as described in Ref. 14. A third method, applicable to any unmodulated signal source, is the *serrodyne technique*. This method allows modulation down to DC and has no effect on (and is not affected by) the PLL design.

Frequency Acquisition

The speed of frequency acquisition (the agility of the PLL) is of interest where fast frequency switching is required. Here, the phase/frequency comparator is clearly superior, since it provides maximum correcting voltage when the loop is unlocked and off frequency. Thus, it will be the only one considered in detail in this discussion. A similar analysis can be made using the linear type of comparator. To simplify the analysis, the acquisition process can be divided into two parts: frequency acquisition and phase acquisition. It is apparent that, during frequency acquisition, when the phase detector output is "against the wall" (either in the positive or negative direction), the maximum rate of the VTO frequency sweep is directly proportional to the control voltage sweep rate and is mainly determined by the slew rate of the integrator. The integrator output voltage (for a constant input) is a nonlinear function of time. It can be obtained by applying a step function to the integrator transfer function and using a Laplace transformation. Figure 11 shows a typical integrator output vs. time for a constant input voltage, and for different values of loop bandwidth and phase margin. As anticipated, the larger bandwidth provides a faster response. The effect of the phase margin is not so obvious. For narrow loop bandwidth and small changes in frequency, the higher phase margin provides a faster response. However, for large frequency excursions, a loop with a lower phase margin is faster. The phase acquisition is a function of the loop step response. The third order step response is very similar to that of the second order, which is shown in Reference 2. While the above discussion of frequency/phase acquisition may be somewhat naive, it does provide some insight into the acquisition mechanism. A more detailed analysis will be presented at a future date.

An example of a typical frequency synthesizer PLL will illustrate some of the points previously discussed. Figure 12 shows such a circuit in a block diagram form. While the exact circuit is not shown, the circuit constants are taken from actual suitable components used previously in similar applications. Three PLL configurations are considered:

A — a third order PLL with a 40 degree phase margin
B — a third order PLL with an 80 degree phase margin
C — a second order PLL with a damping factor of 0.707

The second order loop is included to show the superiority of the third order design. A unity gain frequency, F_0, of 1000 Hz is used in all three examples.

First, the correctness of the three time constants (two for the second order loop), as calculated using program of Table I, can be confirmed. Figure 13 shows the open loop frequency response of the three loop configurations, as calculated using the program of Table II. Indeed, the 0 dB gain occurs at 1000 Hz for all three. Next, the phase margin

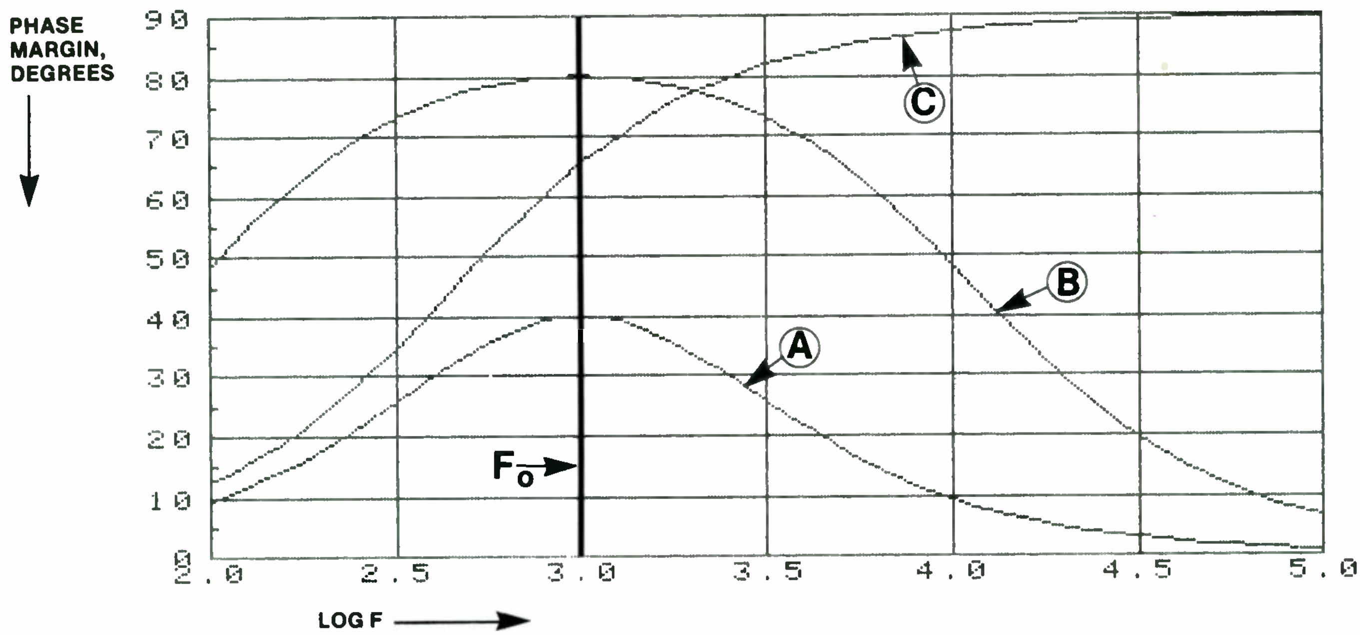

Figure 14. Ideal 2nd and 3rd Order PLL Phase Margin.

```
01♦LBL "HO PLL"
02 FS? 01
03 GTO 00
04 FS? 02
05 GTO 00
06 FS? 03
07 GTO 00
08 FS? 04
09 GTO 00
10 SF 21
11 "FREQ,HZ=? "
12 PROMPT
13 ARCL X
14 AVIEW
15♦LBL 00
16 PI
17 ST+ X
18 *
19 STO 20
20 X↑2
21 RCL 21
22 *
23 RCL 22
24 *
25 RCL 12
26 *
27 RCL 14
28 *
29 RCL 12
30 RCL 13
31 +
32 RCL 14
33 +
34 STO 26
35 RCL 21
36 *
37 RCL 23
38 1
39 +
40 RCL 12
41 *
42 RCL 14
43 *
44 +
45 RCL 23
46 RCL 12
47 *
48 RCL 21
49 +
50 RCL 26
51 +
52 RCL 22
53 *
54 +
55 -
56 RCL 20
57 X↑2
58 *
59 1
60 +
61 RCL 23
62 1
63 +
64 RCL 14
65 *
66 RCL 12
67 *
68 RCL 26
69 RCL 21
70 *
71 +
72 RCL 22
73 *
74 RCL 21
75 RCL 12
76 *
77 RCL 14
78 *
79 +
80 RCL 20
81 X↑2
82 *
83 RCL 23
84 RCL 12
85 *
86 RCL 21
87 +
88 RCL 26
89 +
90 RCL 22
91 +
92 -
93 RCL 20
94 *
95 CHS
96 R-P
97 X<>Y
98 RCL 13
99 RCL 14
100 +
101 RCL 20
102 *
103 1
104 R-P
105 RDN
106 +
107 FS? 04
108 RTN
109 FS? 01
110 GTO 05
111 FS? 02
112 GTO 05
113 FS? 03
114 GTO 05
115 "PM ∡= "
116 ARCL X
117 AVIEW
118♦LBL 05
119 180
120 -
121 STO 24
122 RDN
123 /
124 RCL 15
125 *
126 RCL 16
127 *
128 RCL 23
129 *
130 RCL 17
131 /
132 RCL 20
133 /
134 STO 25
135 LOG
136 20
137 *
138 FS? 01
139 RTN
140 FS? 02
141 GTO 06
142 FS? 03
143 GTO 06
144 "OL MAG= "
145 ARCL X
146 "⊦ DB"
147 AVIEW
148♦LBL 06
149 STO 26
150 RCL 24
151 RCL 25
152 P-R
153 1
154 +
155 R-P
156 1/X
157 LOG
158 20
159 *
160 FS? 02
161 RTN
162 FS? 03
163 GTO 07
164 "NOISE= "
165 ARCL X
166 "⊦ DB"
167 AVIEW
168♦LBL 07
169 RCL 26
170 +
171 RCL 17
172 LOG
173 20
174 *
175 +
176 FS? 03
177 RTN
178 "CL RESP= "
179 ARCL X
180 "⊦ DB"
181 AVIEW
182 ADV
183 END
```

Registers

R12	T_1
R13	T_2
R14	T_3
R15	K_p
R16	K_v
R17	N
R20	[ω]
R21	T_o
R22	T_v
R23	A_o
R24	[<GH]
R25	[IGHI]
R26	[$T_1 + T_2 + T_3$]

Table II.

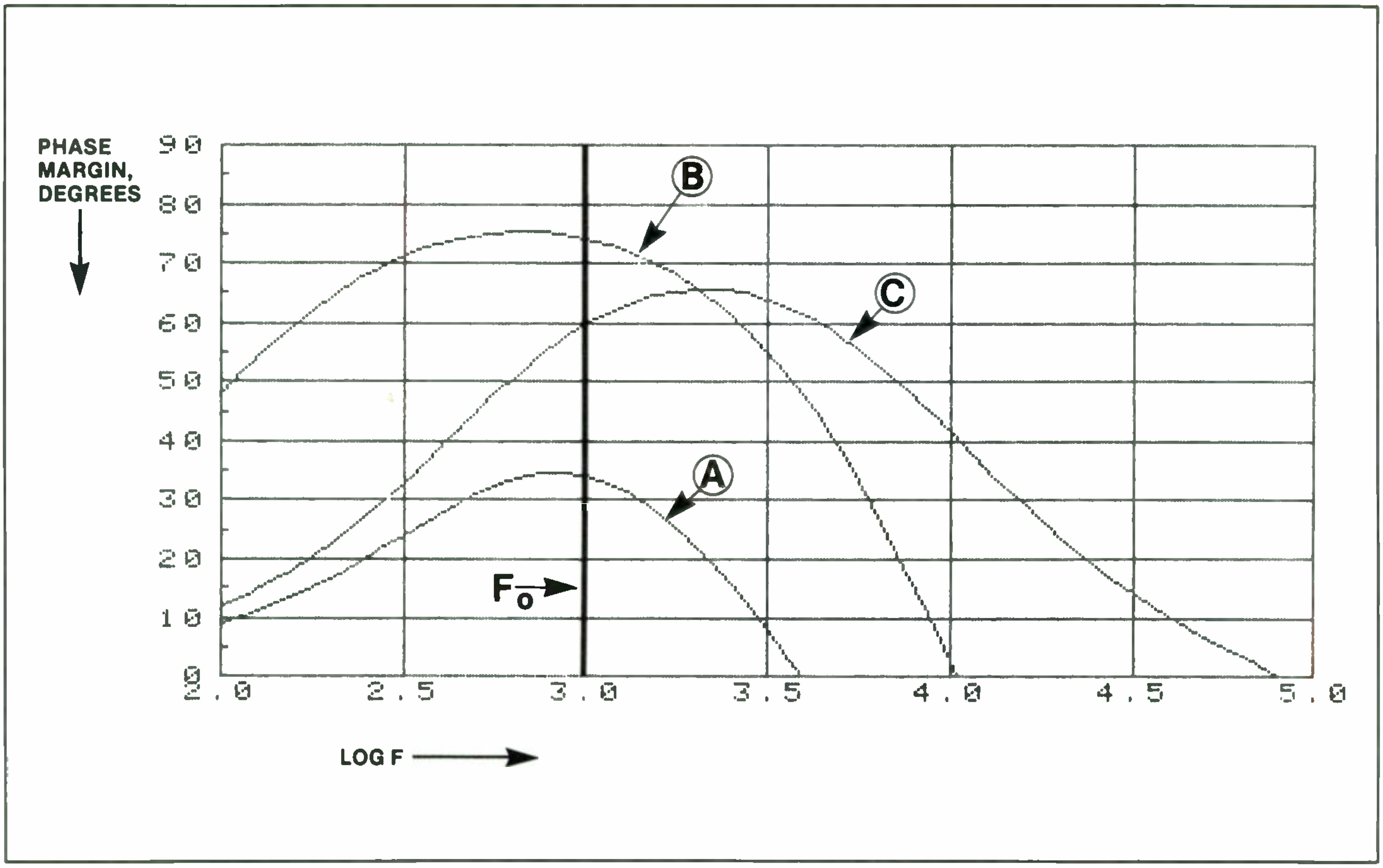

Figure 15. Actual 2nd and 3rd Order PLL Phase Margin.

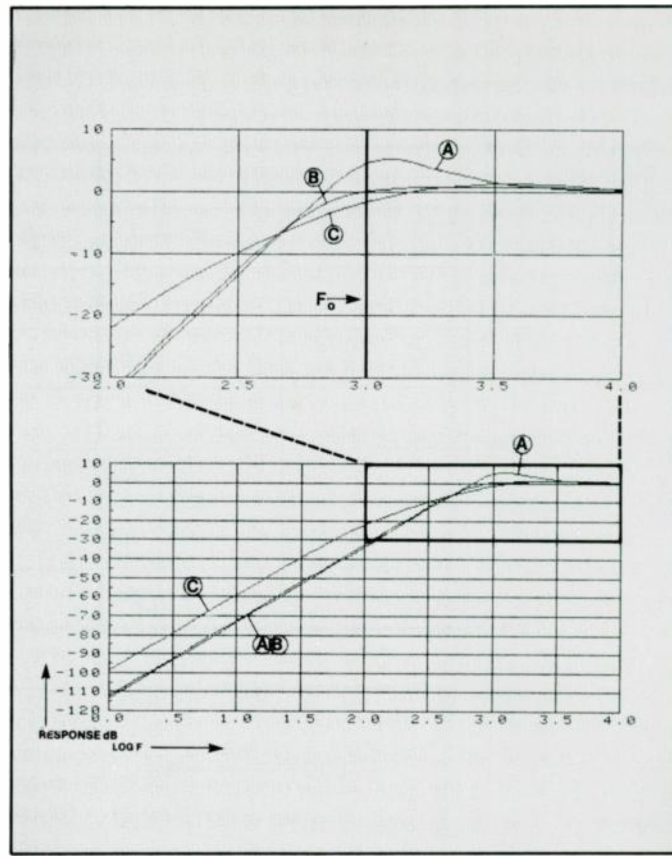

Figure 16. VTO Noise Attenuation.

can be similarly checked, as shown in Figure 15. The phase margin is about 5 degrees less than expected. This decrease in phase margin is caused by the imperfect integrator and the finite modulation response of the VTO, which this program can evaluate. To confirm, the phase margin can be calculated using an ideal circuit (A_o, B_a and B_v are all infinity). Figure 14 shows the expected phase margins. Figure 15, as compared to Figure 14, also shows that there is no such thing as a "second order loop", when actual components are used. While Figure 14 shows the classical second order curve with a 90 degree assymptote (indicating that the loop is unconditionally stable), actually, the phase margin decreases to 0 at some higher frequency. Thus, it is possible to design an "unconditionally stable second order loop" which oscillates. We could now use a higher (by about 5 degrees) phase margin in the original calculation to obtain a higher phase margin in the actual circuit. However, since both third order loops still have adequate phase margins and their gain margins are 20 and 26 dB (usually considered adequate), let's proceed with further circuit analysis of the three examples.

An important consideration in PLL design (especially when used in synthesizers) is the ability of the loop to attenuate VTO phase noise. This attenuation occurs at frequencies within the loop bandwidth and is inversely proportaional to frequency. Figure 16 shows this response. Both third order loops provide over 10 dB improvement in noise reduction within the loop bandwidth. It should be noted that a low phase margin produces an overshoot (an actual gain in noise). The same overshoot occurs for second order loops with low damping factors. However, the more damped third order loop has less overshoot than the second and still provides better noise attenuation.

Finally, the closed loop response can be calculated. The closed loop response indicates how much the reference contributes to the output noise. Here again, the response

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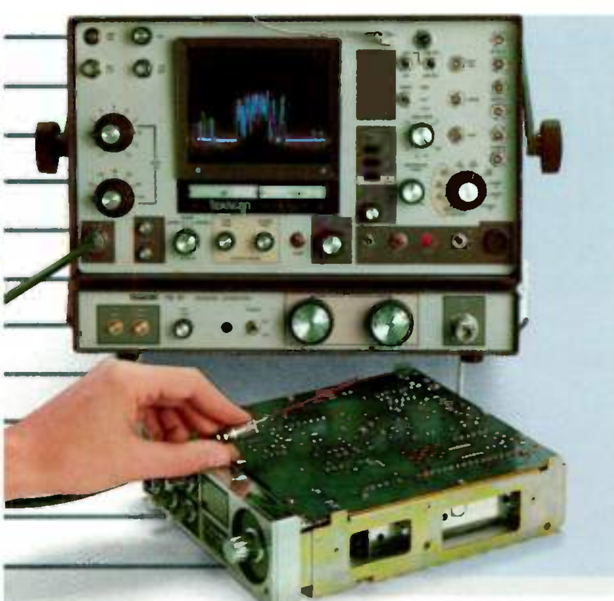

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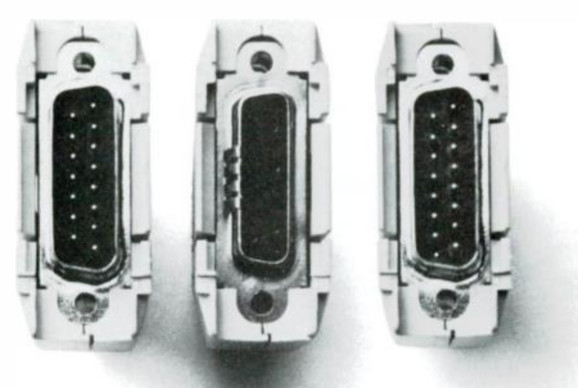

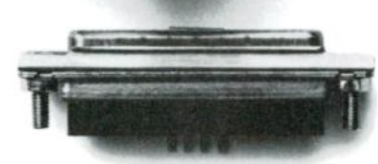

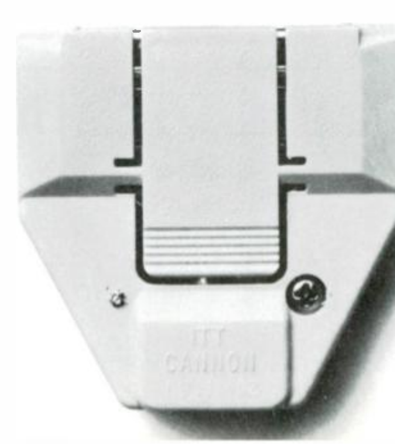

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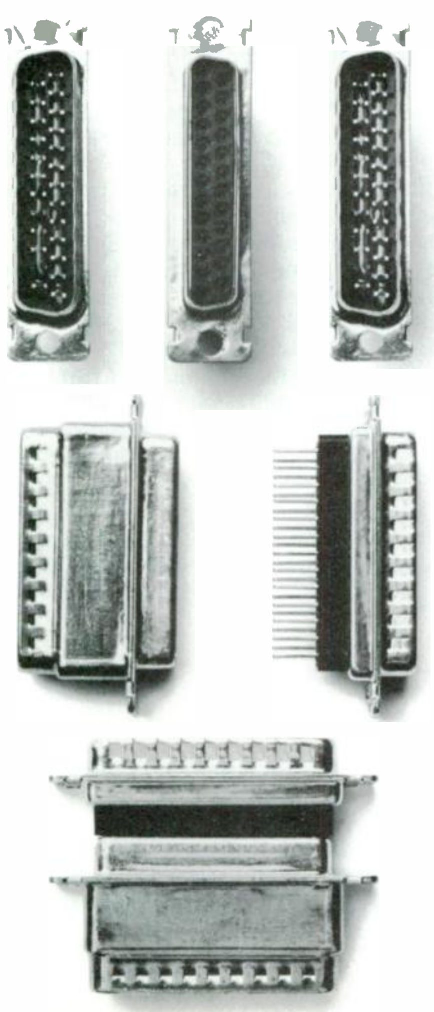

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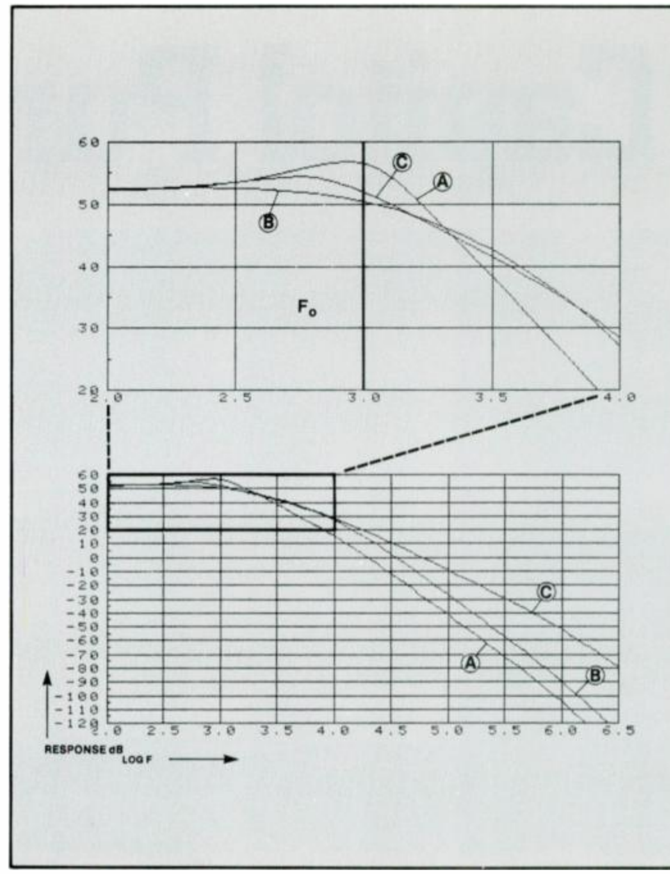

Figure 17. Closed Loop Response.

has an overshoot for lightly damped loops (both second and third order), as shown in Figure 17. However, the highly damped third order PLL has lower overshoot and still higher reference noise attenuation. Here, two factors should be noted: the reference noise is attenuated above the loop bandwidth and is proportional to frequency; and the entire reference source noise is increased by a factor equal to the multiplication factor, N (20 log 400 = 52 dB). These are important considerations in high frequency synthesizers using a low reference frequency.

In conclusion, two factors should be stressed:

— The third order loop analysis provides a superior and more true-to-life circuit performance.

— From equation (7) it is apparent that the loop evaluated by program of Table II is in reality of fifth order (not intentionally, perhaps, but still undeniably fifth) instead of the desired third. The two additional poles within the loop are the integrator pole and the VTO modulation response pole. These are parasitic poles and are generally ignored in conventional PLL analyses, but can be quite significant.

The last part of the PLL Primer (Part III) will be in the July/August issue. It will cover further refinements in PLL design, such as delay in the frequency divider, alternative integrator configuration, VTO modulation bandwidth and others.

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13. A.B. Przedpelski, "Programmable Calculator Computes PLL Noise, Stability", Electronic Design, March 31, 1981.

14. T.G. Giles, "Versatile LSI Frequency Synthesizer", Electronic Components and Applications, Vol. 2, No. 2, N.V. Philips, Eindhoven, the Netherlands, February 1980. □

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MAKING ADVANCED TECHNOLOGY WORK

Calculator Aided Stub Matching

A calculator-aided program presented which solves the single stub matching problem for a real-to-complex match.

By Alan Victor
Motorola Communications Div.
Plantation, FL 33322

Evaluation and design of high frequency components, both active and passive, usually requires some form of impedance matching. This is necessary in order to permit the device to operate properly in a system impedance which is perhaps different than its own characteristic impedance. In the high frequency range, (say) below several hundred megahertz, various lumped impedance matching techniques exist. Unfortunately, at much higher frequenices lumped element techniques become difficult to implement, control, and adjust with any degree of accuracy.

On the other hand the use of a transmission line affords the designer an additional degree of flexibility. Controlling the physical dimensions of the line by altering the electrical length, the characteristic impedance or the terminating impedance yields a flexible impedance matching element.

The transmission line properties are determined by the dielectric constant of the substrate material in the design. If the technique is to print the transmission line, a "stripline" or "micro-strip" design of some form is used.

Previous articles (1-4) have discussed the single line and non-synchronous transmission line matching networks. Unfortunately, for some cases these techniques lead to line dimensions which are prohibitive in size for a given substrate material and dielectric constant.

The transmission line matching technique makes use of the fact that a particular complex impedance is provided by adding a series line of length x between a given source and load. Furthermore by adjusting the line length and line impedance, an impedance match between source and load is possible. A stub matching technique adds another degree of flexibility to the single line. By adding a shunt line (the stub) off the main series line and adjusting the terminating impedance of the stub, various series line lengths are available to implement an impedance match between source and load. In particular, if a short or open stub termination is selected, two possible series line lengths are found. In addition, by adjusting the reference impedance value of the network or transmission line characteristic impedance Z_0, additional flexibility is provided. Finally, the single-stub matching section allows some degree of "tweaking". These networks are readily solved on the Smith Chart, but can be tedious and lead to inaccuracies especially after renormalizing the chart for a different impedance.

The program outlined in this article handles the case of a real-to-complex match. All possible series line lengths and open/short circuited stubs are displayed. The system impedance z_0, and the reflection coefficient ρ_T associated with the terminal impedance measured in that system are the inputs. The series lines and stub lengths are the outputs. Furthermore, if microstrip is the design medium the program outputs line dimensions. The user now has the option of specifying a new system reference impedance z_0, new source or load resistance and calculating a new set of series line lengths and stubs or load resistance and calculating a new set of series line lengths and stubs which might be physically more realizable. Programs for the HP-41C are included and pertinent equations developed.

Equations Developed (5)

A transmission line with characteristic impedance Z_0, or $1/Y_0$ is terminated in a load admittance $Y_T = G_T + jB_T$. Using the definition of reflection coefficient ρ_T, we have:

$$\frac{Y_T}{Y_0} = \frac{(1 - \rho_T)}{1 + \rho_T} \tag{1}$$

and

$$\rho_{TNEW} = \frac{1 - Y_T/Y_{oNEW}}{1 + Y_T/Y_{oNEW}} \tag{1a}$$

where ρ_T is a complex number with magnitude $|\rho_T|$ and phase angle φ_T as defined in equation 2.

$$\rho_T = |\rho_T|\, e^{j\varphi_T} = |\rho_T|\,(\cos\varphi_T + j\sin\varphi_T) \tag{2}$$

The value ρ_{TNEW} in equation (1a) represents the NEW value of the reflection coefficient referred to a new characteristic impedance or admittance, Y_{oNEW}.

After rationalizing (1) and separating the real and imaginary parts we have:

$$\frac{G_T}{Y_0} = \frac{1 - |\rho_T|^2}{|\rho_T|^2 + 1 + 2|\rho_T|\cos\varphi_T} \tag{3}$$

$$\frac{B_T}{Y_0} = \frac{-2|\rho_T|\sin\varphi_T}{|\rho_T|^2 + 1 + 2|\rho_T|\cos\varphi_T} \tag{3a}$$

Equations (1), (1a), (3), and (3a), are used in the normalization portion of the program to convert the given value of ρ_T from the system reference impedance "z_0" to another impedance "NEW z_0".

The series line lengths and stub lengths are obtained from the definition of the admittance along a lossless line:

$$\frac{Y(x)}{Y_0} = \frac{Y_T/Y_0 + j\tan\beta_x}{1 - j(Y_T/Y_0)\tan\beta_x} \tag{4}$$

where x is the coordinate point on the line measured from the terminal end of the line, $\beta = 2\pi/\lambda$, and Y_T/Y_0 is the normalized load admittance. Letting $Y_T/Y_0 = G_T/Y_0 + B_T/Y_0 = G_{Tn} + jB_{Tn}$, substituting into (4) and expanding we find the real part as:

$$\frac{G(x)}{Y_0} = \frac{G_{Tn}(1 + \tan^2\beta x)}{(1 - B_{Tn}\tan\beta x)^2 + (G_{Tn}\tan\beta x)^2} \tag{5}$$

Using equation (5) and some algebra we solve for the series transmission line length x, in terms of the normalized load conductance $G(x)/Y_0$.

$$x = \frac{\lambda}{2\pi}\tan^{-1}\frac{\frac{G(x)}{Y_o}B_{Tn} \pm \sqrt{G_{Tn}\left[\frac{G(x)}{Y_o}|Y_{Tn}|^2 - \left(\frac{G(x)}{Y_o}\right)^2 G_{Tn} - G_{Tn} - \frac{G(x)}{Y_o}\right]}}{\frac{G(x)}{Y_o}\left(|Y_{Tn}|^2\right) - G_{Tn}} \qquad (6)$$

At the point x (in equation 6) the real part of the load admittance is equal to the generator or source admittance and a net susceptance remains. In equation (6) if the characteristic admittance of the line (Y_o) is assigned the same value as the system reference impedance (z_o) then $B(x)/Y_o$ is unity.

For a line impedance different than the system reference impedance, a NEW value of ρ_T and $B(x)/Y_o$ must be calculated using equation (1a), equation (3), and equation (5), with B_T/Y_o replaced with B_T/Y_{oNEW}.

Equation (6) yields the series line lengths (λ_{calc} or x) $\pm n\lambda/2$. If the result for x is negative, then we add a ½ wavelength of line to the result. The equation is general in that Y_o need not be the same as the system load or source admittance. This allows the designer to try a different impedance line, $1/Y_o$. The required susceptance of the line is then found in a similar manner using (4) and separating out the real and imaginary terms. After some algebra we have:

$$\frac{B(x)}{Y_o} = \pm\sqrt{\frac{G(x)}{Y_o}\left[\frac{1 - |Y_{Tn}|^2 - G_{Tn} - \left(\frac{G(x)}{Y_o}\right)^{-2} G_{Tn}}{G_{Tn}}\right]} \qquad (7)$$

from which either an open or shorted stub length is found using:

$$z_{in} = j\, z_o \tan\beta x \quad :\text{SSTB}$$
$$z_{in} = -j\, z_o \cot\beta x \quad :\text{OSTB} \qquad (8)$$

and the sign of the imaginary part of equation (4) as shown in (9):

$$\frac{B(x)}{Y_o} = -\tan\beta x\left[G_{Tn}{}^2 - B_{Tn}{}^2\right] + B_{Tn} + \tan\beta x - B_{Tn}\tan^2\beta x \qquad (9)$$

If (9) is positive, the program is directed to output "USE SSTB" (Shorted STuB) otherwise if negative the output displayed is "USE OSTB" (Open STuB).

After series lines and stub lengths are calculated, the program can be directed to the microstrip calculator (USTRIP) section. The only inputs required are z_o and ϵ_r. Outputs are ϵ_{eff} and W/H from which physical dimensions are calculated after entering the substrate height, frequency in MHz, and the desired electrical wavelength (λ_{calc}). Using (10) we have:

$$1(\text{physical}) = \frac{\lambda_{calc}C}{f\sqrt{\epsilon_{eff}}} = \lambda_{calc}\frac{300}{f(\text{MHz})\sqrt{\epsilon_{eff}}}\ (\text{meters}) \qquad (10)$$

Program Use

Load the programs, *STUB MATCH* and *USTRIP*. To run press XEQ "MAT" or XEQ "USTRIP". The program is self-promoting and RUN/STOP commands can be inserted to allow program control. Total steps for both programs are 598 and storage registers required are 11. The memory size need only be set to 12 and the separate programs can be readily modified to run on the HP67/97 or T158/59 machines.

The input prompts for the system reference impedance (z_o) which is also the initial series line and stub characteristic impedance Z_o. The program then prompts for the load or termination reflection coefficients measured in that reference impedance system and the load resistance. The program outputs the series line length, then tells you if an open series line length and stub is required. The next is the other series line length and stub requirement. Finally, the stub lengths are displayed and their type (open or shorted). The next input prompt is for a new Z_o, i.e., a new transmission line characteristic impedance and again the load resistance. New line dimensions are then displayed. If the calculated series line length is negative, then the program automatically adds a half-wave section and outputs HALF-WAVE ADDED.

The program "USTRIP" prompts for Z_o and ϵ_{eff}. Then physical dimensions of the previous calculated lines from "MAT" can readily be found.

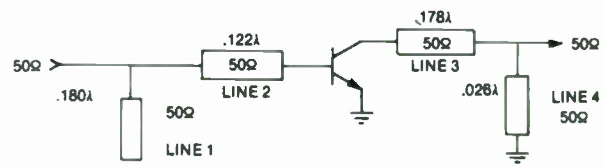
Example

Given a 2N3570 at 750 MHz. The 2 port scattering parameters characterized in a 50 Ω system impedance yield a simultaneous conjugate match condition for:

$$\rho^*_{input} = .730\ \angle 135.4°$$

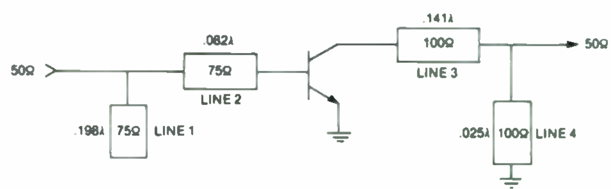
$$\rho^*_{output} = .951\ \angle 33.8°$$

Using "MAT" with 50 Ω lines we obtain the following input and output networks for $\rho_{T\ OUTPUT}$ of .951 $\angle -33.8°$ and $\rho_{T\ INPUT}$ of .730 $\angle -135.4°$.



Other networks are also listed for a Z_o of 75Ω and 100Ω. It is desireable to use those networks which yield the shortest required electrical lengths to realize a match. This selection not only reduces the physical layout area, but reduces the frequency sensitivity of the matching networks (5).

Therefore, we might try the following network:



INPUT			
z_o(Ω)	series line λ	OSTB λ	SSTB λ
50	.122 .002	.180 —	— .070
75	.082 .001	.198 —	— .052
100	.062 .001	.207 —	— .043

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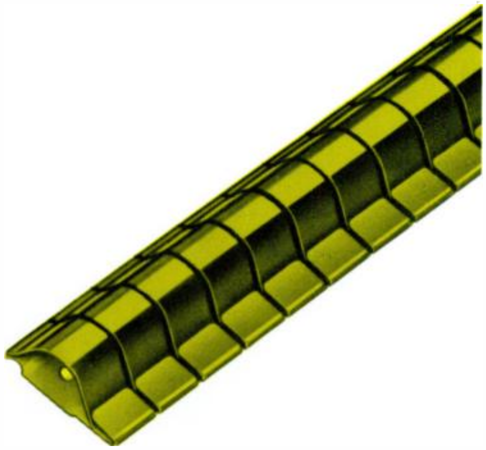
This narrow strip...



		OUTPUT	
$z_o(\Omega)$	series line λ	OSTB λ	SSTB λ
50	.178	—	.026
	.228	.224	—
75	.205	.225	—
	.158	—	.025
100	.185	.225	—
	.141	—	.025

Finally, we use "USTRIP". This program is adapted from (6) and handles a fairly wide range of Z_0 and ϵ_r values with good accuracy. Let's assume glass-teflon is the substrate material with $\epsilon_r = 2.50$ and 62.5 mil thickness. Then we have the following ϵ_{eff}, W/H and physical line lengths.

$z_o = 75\Omega$	$\epsilon_r = 2.50$	$\epsilon_{eff} = 2.007$	W/H = 1.443
$z_o = 100\Omega$	$\epsilon_r = 2.50$	$\epsilon_{eff} = 1.941$	W/H = .798

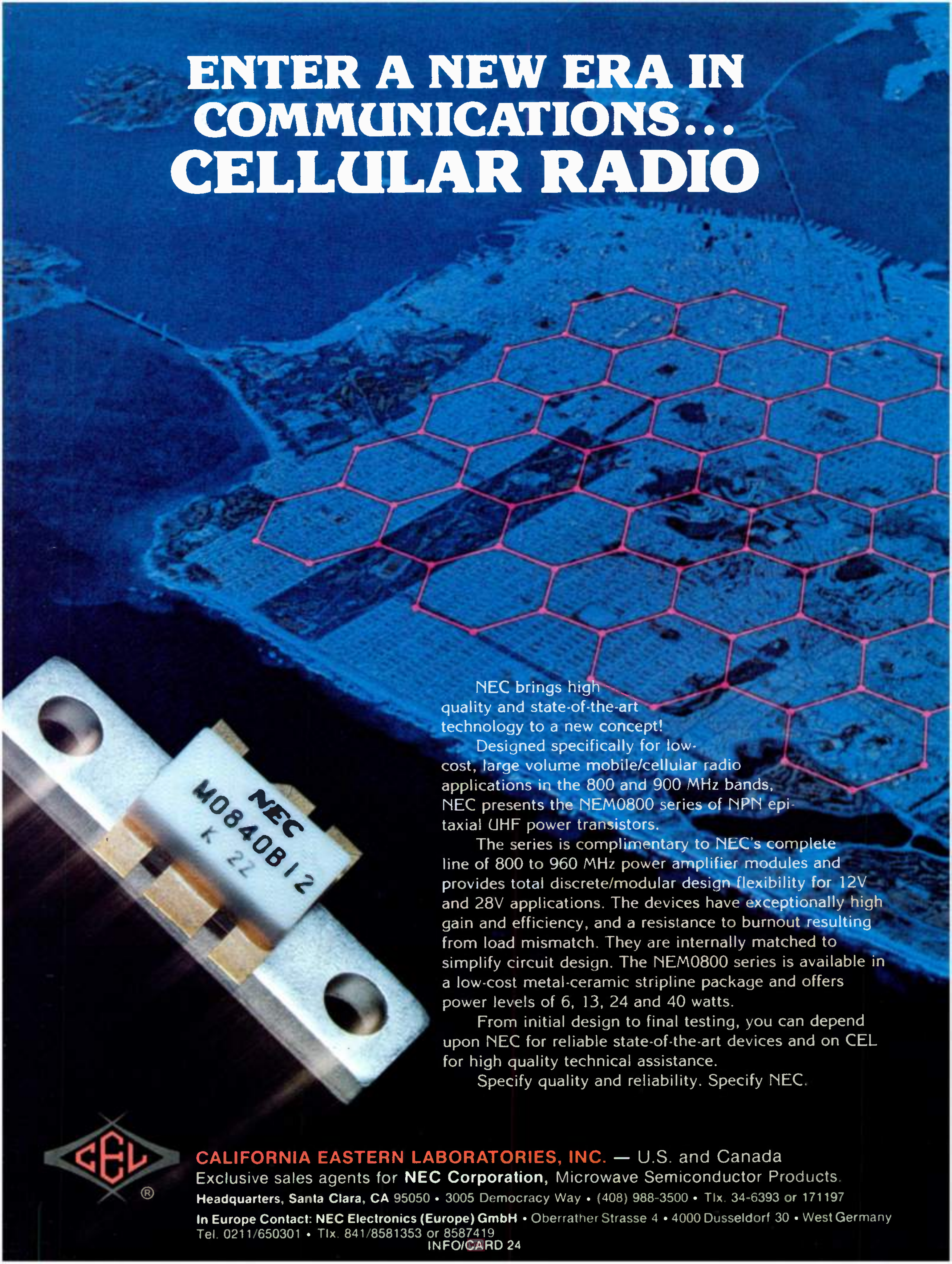
LINE 1 = 2200 mils	Width = 90 mils
LINE 2 = 911 mils	Width = 90 mils
LINE 3 = 1593 mils	Width = 49.8 mils
LINE 4 = 283 mils	Width = 49.8 mils

Conclusions

A calculator aided program has been presented which solves the single stub matching problem for a real-to-complex match. The program is enhanced by handling any characteristic line impedance thus reducing the tedious chore of renormalizing the Smith Chart for each solution. A micro-strip calculator routine is included which handles the conversion of electrical length to physical length for a wide range of Z_0 and ϵ_r materials.

References:

1. Regier, R.A., "Impedance Matching With A Series Transmission Line Section", Proceedings of the IEEE, pp. 1133-1134 (July 1971).
2. Hayes, Albert E., "Eighth-Wave Line Matches Impedances", EDN pp. 73-75 (January 5, 1979).
3. Przedpelski, Andrzej B., "Solve High-Frequency Impedance-Matching Problems With Transmission Lines", ED 19, pp. 144-146, (September 13, 1978).
4. Kurt P. Schwan, "Matching: When Is A Single Line Sufficient?", Microwaves, pp. 58-63, (December 1975).
5. Collin, R.E., *Foundations For Microwave Engineering*, McGraw-Hill, New York 1966, Chapter 5.
6. Edwards, T.C. *Foundations for Microstrip Circuit Design*, John Wiley and Sons, New York 1981.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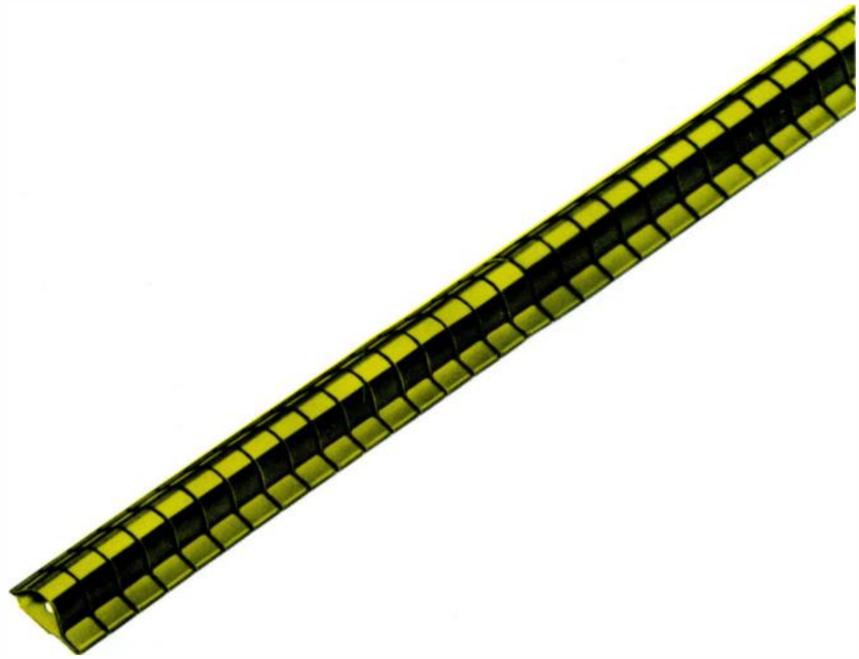
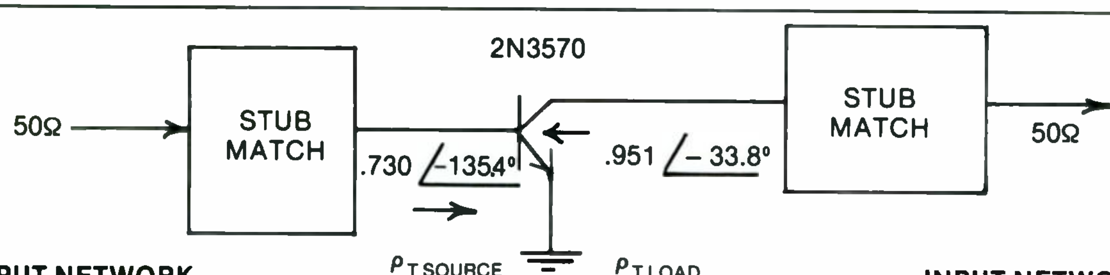




OUTPUT NETWORK

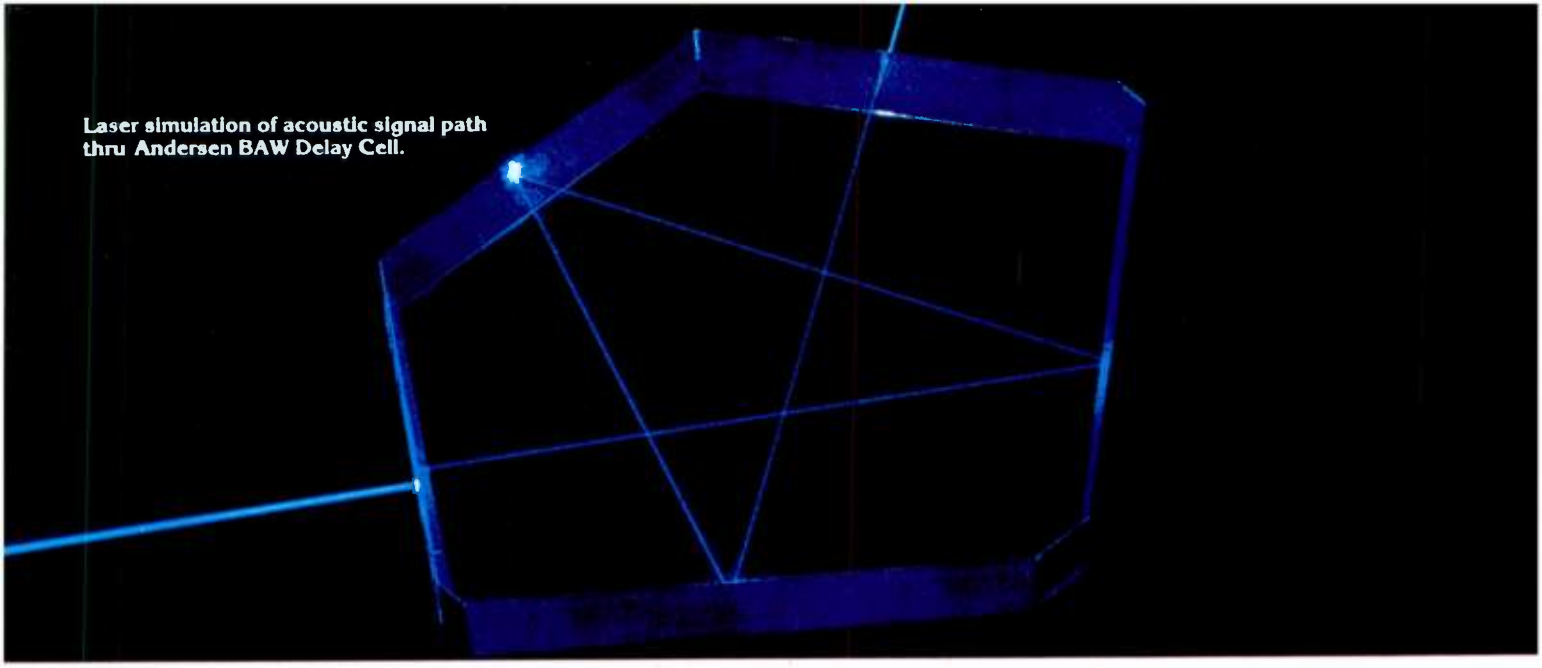
```
           XEQ "MAT"
*STUB MATCH*
Z0?
        50.000    RUN
RHOMAG LOAD?
          .951    RUN
RHOANG LOAD?
       -33.800    RUN
RSOURCE?
        50.000    RUN
SERIES LINES
SERIES LINE
L=0.228LAMBDA
USE OSTB
SERIES LINE
L=0.178LAMBDA
USE SSTB
STUBS
SHORTED STUB
L=0.026LAMBDA
OPEN STUB
L=0.224LAMBDA
```

```
NEW Z0?
        100.000    RUN
RSOURCE?
         50.000    RUN
SERIES LINES
SERIES LINE
L=0.185LAMBDA
USE OSTB
SERIES LINE
L=0.141LAMBDA
USE SSTB
STUBS
SHORTED STUB
L=0.025LAMBDA
OPEN STUB
L=0.225LAMBDA
```

INPUT NETWORK

```
NEW Z0?
           XEQ "MAT"
*STUB MATCH*
Z0?
        50.000    RUN
RHOMAG LOAD?
          .730    RUN
RHOANG LOAD?
      -135.400    RUN
RSOURCE?
        50.000    RUN
SERIES LINES
SERIES LINE
L=0.122LAMBDA
USE OSTB
SERIES LINE
L=0.002LAMBDA
USE SSTB
STUBS
SHORTED STUB
L=0.070LAMBDA
OPEN STUB
L=0.180LAMBDA
NEW Z0?
```

Calculated Transmission Line for Transistor Amplifier Example.



Laser simulation of acoustic signal path thru Andersen BAW Delay Cell.

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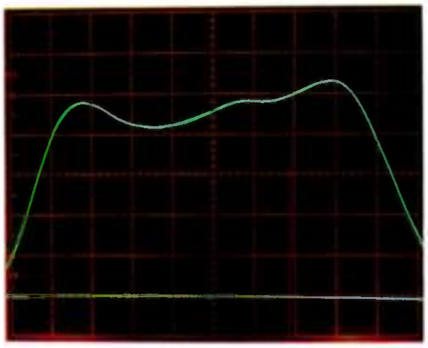

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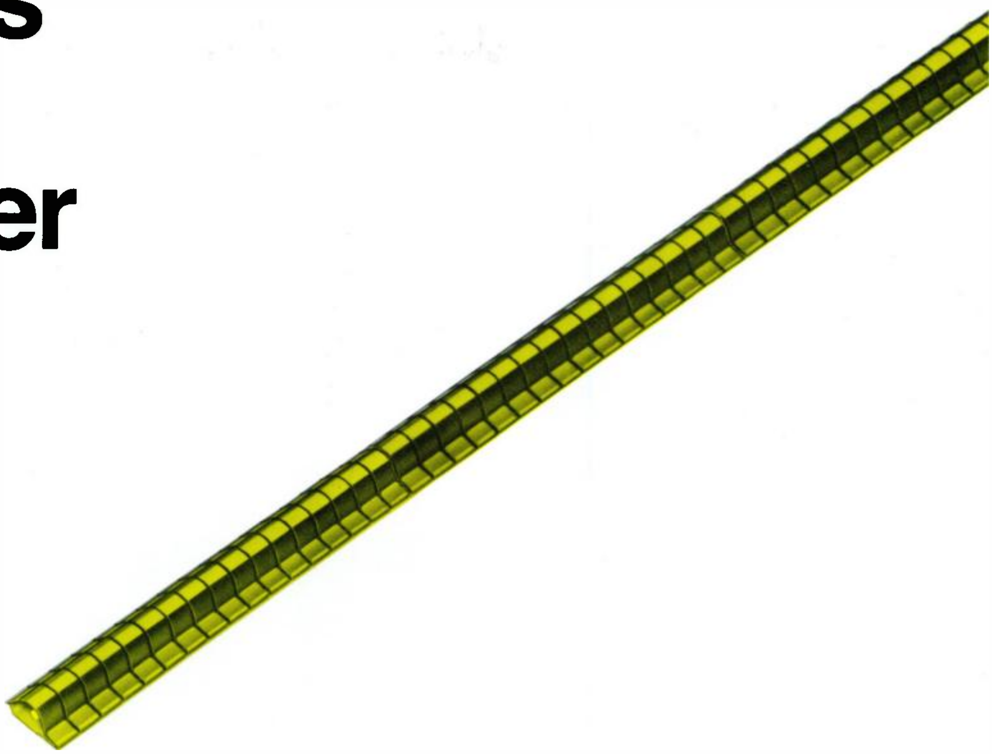
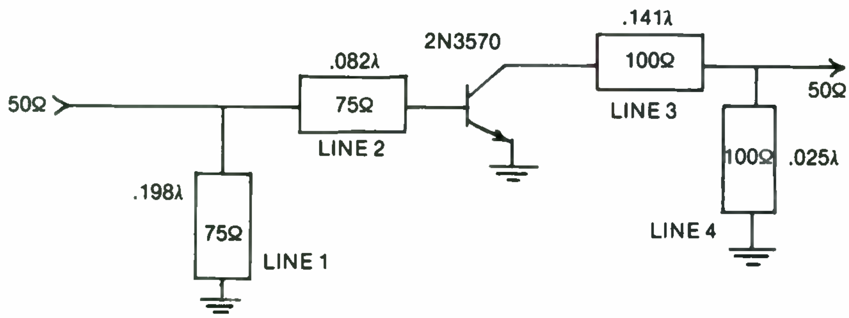




INPUT NETWORK

```
        XEQ "USTRIP"
Z0=?
         75.000   RUN
e REL=?
          2.500   RUN
W/H=1.443
e EFF=2.007
SUBST HT?
         62.500   RUN
FREQ-MHZ?
        750.000   RUN
LINE LENGTH?
           .198   RUN
W=90.165
L=2,200.825MILS
```

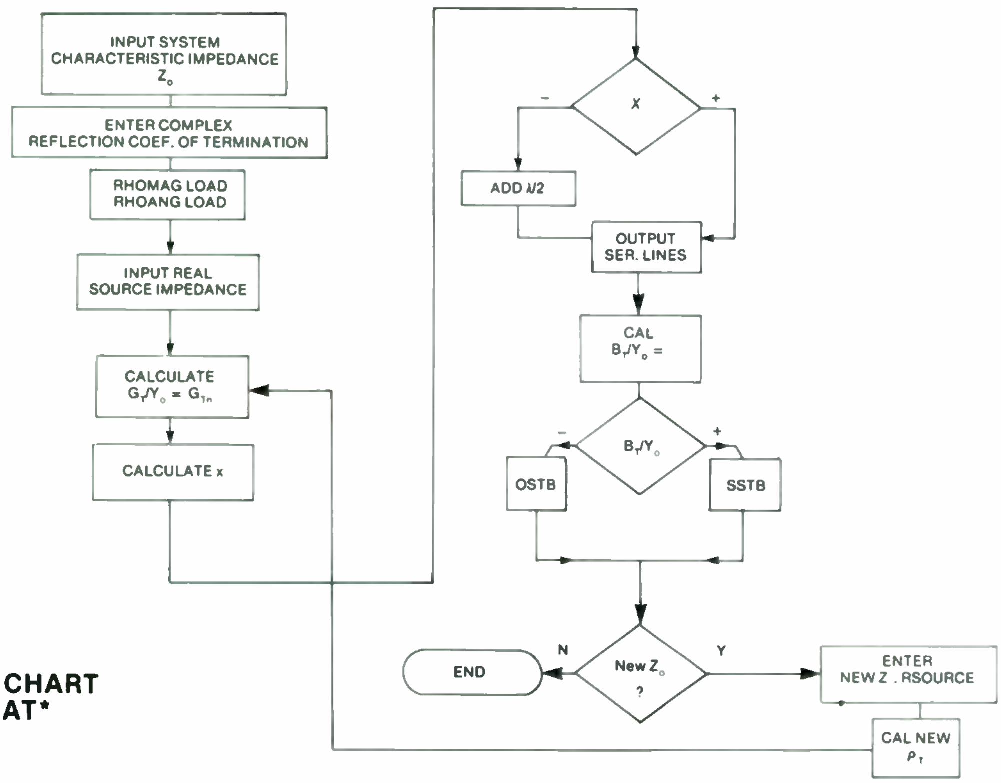
```
        XEQ "USTRIP"
Z0=?
         75.000   RUN
e REL=?
          2.500   RUN
W/H=1.443
e EFF=2.007
SUBST HT?
         62.500   RUN
FREQ-MHZ?
        750.000   RUN
LINE LENGTH?
           .082   RUN
W=90.165
L=911.453MILS
```

OUTPUT NETWORK

```
        XEQ "USTRIP"
Z0=?
        100.000   RUN
e REL=?
          2.500   RUN
W/H=0.798
e EFF=1.941
SUBST HT?
         62.500   RUN
FREQ-MHZ?
        750.000   RUN
LINE LENGTH?
           .141   RUN
W=49.891
L=1,593.664MILS
```

```
        XEQ "USTRIP"
Z0=?
        100.000   RUN
e REL=?
          2.500   RUN
W/H=0.798
e EFF=1.941
SUBST HT?
         62.500   RUN
FREQ-MHZ?
        750.000   RUN
LINE LENGTH?
           .025   RUN
W=49.891
L=282.565MILS
```

Calculate Output for Physical Line Dimensions



FLOW CHART
MAT

FLOW CHART
USTRIP

INPUT Z_o, ϵ_{REL}

↓

COMPUTE ϵ_{REL} RANGE

↓

CAL. ϵ_{EFF}, W/H

↓

INPUT FREQ., LINE LENGTH (λ)

↓

OUTPUT WIDTH, LENGTH (PHYSICAL)

INFO/CARD 29

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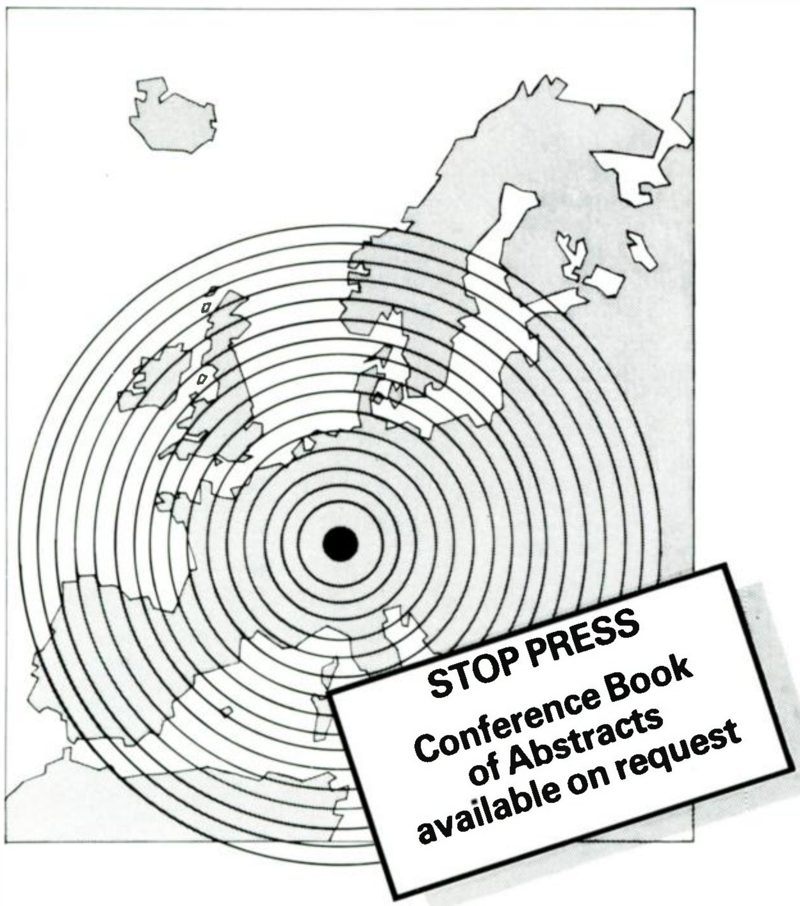

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Ustrip Program

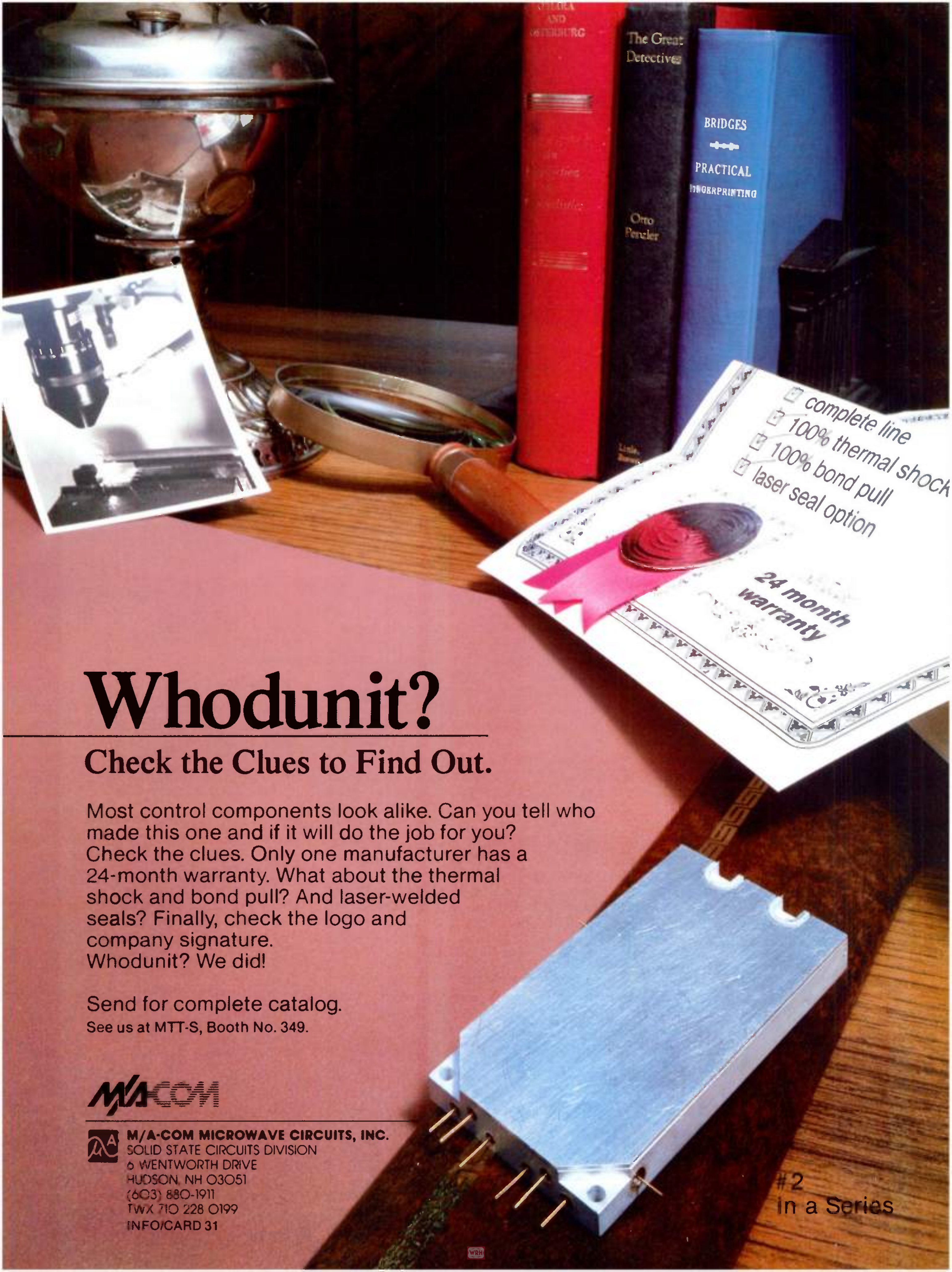
```
PRP "MAT"

01♦LBL "MAT"
02 FIX 3
03 "♦STUB MATCH♦"
04 AVIEW
05 PSE
06 "ZO?"
07 PROMPT
08 1/X
09 STO 00
10 "RHOMAG LOAD?"
11 PROMPT
12 STO 01
13 "RHOANG LOAD?"
14 PROMPT
15 STO 02
16 "RSOURCE?"
17 PROMPT
18 RCL 00
19 *
20 1/X
21 STO 09

22♦LBL 02
23 RCL 01
24 2
25 *
26 RCL 02
27 COS
28 *
29 RCL 01
30 X↑2
31 +
32 1
33 +
34 1/X
35 STO 03
36 RCL 01
37 X↑2
38 CHS
39 1
40 +
41 *
42 STO 04
43 RCL 00
44 *
45 RCL 03
46 2
47 *
48 CHS
49 RCL 01
50 *
51 RCL 02
52 SIN
53 *
54 STO 05
55 RCL 00
56 *
57 "SERIES LINES"
58 AVIEW
59 PSE
60 RCL 04
61 X↑2
62 RCL 05
63 X↑2
64 +
65 RCL 09
66 *
67 RCL 09
68 X↑2
69 RCL 04
70 *
71 CHS
72 +
73 RCL 04
74 CHS
75 +
76 RCL 09
77 +
78 RCL 04
79 *
80 SQRT
81 STO 06
82 RCL 04
83 X↑2
84 RCL 05
85 X↑2
86 +
87 RCL 09
88 *
89 RCL 04
90 -
91 1/X
92 STO 07
93 RCL 09
94 RCL 05
95 *
96 RCL 06
97 +
98 RCL 07
99 *
100 ATAN
101 360
102 /
103 X<0?
104 XEQ 01
105 "SERIES LINE"
106 AVIEW
107 PSE
108 "L="
109 ARCL X
110 "⊢LAMBDA"
111 AVIEW
112 XEQ 03
113♦LBL 02
114 RCL 09
115 RCL 05
116 *
117 RCL 06
118 -
119 RCL 07
120 *
121 ATAN
122 360
123 /
124 X<0?
125 XEQ 01
126 "SERIES LINE"
127 AVIEW
128 PSE
129 "L="
130 ARCL X
131 "⊢LAMBDA"
132 AVIEW
133 XEQ 03
134 "STUBS"
135 AVIEW
136 PSE
137 RCL 05
138 X↑2
139 RCL 04
140 X↑2
141 +
142 RCL 04
143 -
144 RCL 09
145 X↑2
146 RCL 04
147 *
148 CHS
149 +
150 1
151 +
152 RCL 04
153 /
154 RCL 09
155 *
156 SQRT
157 1/X
158 STO 08
159 ATAN
160 360
161 /
162 "SHORTED STUB"
163 AVIEW
164 PSE
165 "L="
166 ARCL X
167 "⊢LAMBDA"
168 AVIEW
169 PSE
170 RCL 08
171 1/X
172 ATAN
173 360
174 /
175 "OPEN STUB"
176 AVIEW
177 PSE
178 "L="
179 ARCL X
180 "⊢LAMBDA"
181 AVIEW
182 PSE
183 RCL 00
184 1/X
185 STO 06
186 "NEW ZO?"
187 PROMPT
188 STO 04
189 1/X
190 STO 00
191 "RSOURCE?"
192 PROMPT
193 *
194 1/X
195 STO 09
196 RCL 01
197 2
198 *
199 RCL 02
200 COS
201 *
202 CHS
203 RCL 01
204 X↑2
205 +
206 1
207 +
208 STO 05
209 RCL 01
210 X↑2
211 CHS
212 1
213 +
214 RCL 05
215 /
216 RCL 06
217 *
218 STO 07
219 RCL 02
220 SIN
221 RCL 01
222 *
223 2
224 *
225 RCL 05
226 /
227 RCL 06
228 *
229 STO 03
230 RCL 07
231 RCL 04
232 -
233 RCL 03
234 X<>Y
235 R-P
236 RCL 07
237 RCL 04
238 +
239 RCL 03
240 X<>Y
241 R-P
242 X<>Y
243 RDN
244 /
245 RDN
246 X<>Y
247 -
248 R↑
249 STO 01
250 X<>Y
251 STO 02
252 RCL 04
253 1/X
254 STO 00
255 GTO 02
256♦LBL 01
257 .5
258 +
259 "HALFWAVE ADDED"
260 AVIEW
261 RTN
262♦LBL 03
263 360
264 *
265 STO 11
266 TAN
267 X↑2
268 RCL 05
269 *
270 CHS
271 RCL 11
272 TAN
273 +
274 RCL 05
275 +
276 RCL 04
277 X↑2
278 RCL 05
279 X↑2
280 +
281 RCL 11
282 TAN
283 *
284 CHS
285 +
286 X<0?
287 GTO 04
288 GTO 05
289♦LBL 04
290 "USE OSTB"
291 AVIEW
292 PSE
293 RTN
294♦LBL 05
295 "USE SSTB"
296 AVIEW
297 PSE
298 RTN
299 END
```

```
PRP "USTRIP"

01♦LBL "USTRIP"
02 "ZO=?"
03 PROMPT
04 STO 00
05 "e REL=?"
06 PROMPT
07 STO 01
08 2
09 *
10 CHS
11 44
12 +
13 ABS
14 RCL 00
15 X>Y?
16 GTO 01
17 GTO 02
18♦LBL 01
19 4
20 ENTER↑
21 PI
22 /
23 LN
24 RCL 01
25 1/X
26 *
27 PI
28 ENTER↑
29 2
30 /
31 LN
32 +
33 STO 02
34 RCL 01
35 1
36 -
37 RCL 01
38 1
39 +
40 /
41 RCL 02
42 *
43 STO 03
44 2
45 /
46 STO 02
47 RCL 01
48 1
49 +
50 2
51 *
52 SQRT
53 119.9
54 /
55 RCL 00
56 *
57 RCL 02
58 +
59 STO 02
60 E↑X
61 4
62 *
63 1/X
64 CHS
65 RCL 02
66 E↑X
67 8
68 /
69 +
70 1/X
71 "W/H="
72 ARCL X
73 AVIEW
74 PSE
75 STO 05
76 GTO C
77♦LBL 02
78 RCL 01
79 SQRT
80 RCL 00
81 *
82 1/X
83 59.95
84 *
85 PI
86 X↑2
87 *
88 STO 04
89 RCL 01
90 1/X
91 .517
92 *
93 CHS
94 .293
95 +
96 RCL 04
97 1
98 -
99 LN
100 +
101 PI
102 /
103 RCL 01
104 /
105 RCL 01
106 1
107 -
108 *
109 STO 02
110 RCL 04
111 2
112 *
113 1
114 -
115 LN
116 CHS
117 RCL 04
118 1
119 -
120 +
121 2
122 *
123 PI
124 /
125 RCL 02
126 +
127 "W/H="
128 ARCL X
129 AVIEW
130 PSE
131 STO 05
132 GTO C
133♦LBL C
134 45
135 ENTER↑
136 RCL 00
137 X<=Y?
138 GTO 03
139 GTO 04
140♦LBL 03
141 8
142 ENTER↑
143 RCL 00
144 X<>Y
145 X<=Y?
146 GTO 05
147 GTO 04
148♦LBL 05
149 RCL 00
150 10
151 +
152 LOG
153 1
154 -
155 STO 02
156 RCL 01
157 4 E-3
158 *
159 CHS
160 .109
161 +
162 RCL 01
163 *
164 RCL 02
165 *
166 .96
167 +
168 1/X
169 RCL 01
170 *
171 X<0?
172 GTO 09
173 "e EFF="
174 ARCL X
175 AVIEW
176 PSE
177 STO 06
178 GTO "DIMEN"
179♦LBL 04
180 63
181 ENTER↑
182 RCL 01
183 2
184 *
185 CHS
186 +
187 ABS
188 RCL 00
189 X<Y?
190 GTO 06
191 GTO 07
192♦LBL 07
193 63
194 ENTER↑
195 RCL 01
196 2
197 *
198 CHS
199 RCL 00
200 X>Y?
201 GTO 08
202 "ERROR"
203 AVIEW
204 STOP
205♦LBL 06
206 GTO 05
207♦LBL 08
208 RCL 01
209 1
210 +
211 1/X
212 2
213 *
214 SQRT
215 29.98
216 *
217 RCL 00
218 /
219 RCL 03
220 *
221 1
222 +
223 X↑2
224 2
225 /
226 RCL 01
227 1
228 +
229 *
230 "e EFF="
231 ARCL X
232 AVIEW
233 PSE
234 STO 06
235 GTO "DIMEN"
236♦LBL 09
237 RCL 05
238 1/X
239 10
240 *
241 1
242 +
243 -.555
244 Y↑X
245 RCL 01
246 1
247 -
248 2
249 /
250 *
251 RCL 01
252 1
253 +
254 2
255 /
256 +
257 STO 06
258 "eEFF="
259 ARCL X
260 AVIEW
261 PSE
262♦LBL "DIMEN"
263 "SUBST HT?"
264 PROMPT
265 STO 07
266 "FREQ-MHZ?"
267 PROMPT
268 STO 08
269 "LINE LENGTH?"
270 PROMPT
271 STO 09
272 RCL 05
273 RCL 07
274 *
275 "W="
276 ARCL X
277 AVIEW
278 PSE
279 RCL 06
280 SQRT
281 1/X
282 RCL 09
283 *
284 300
285 *
286 RCL 08
287 /
288 100
289 *
290 2.54
291 /
292 1 E3
293 *
294 "L="
295 ARCL X
296 "⊢MILS"
297 AVIEW
298 STOP
299 END
```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Taking The Magic Out Of The "Magic Tee"

Understanding the principles of Hybrid Power Splitters and Combiners.

By Alex J. Burwasser
Consulting Engineer
Fair-Rite Products Corp.

Of all the passive circuits employed by the RF engineer, there are probably very few so misunderstood as the "magic tee" hybrid power splitter/combiner (the terminology itself is a good clue that the principles involved are subtle). In this paper, we will attempt to unveil some of the "magic tee's" mystery, discuss some practical design considerations, and present two practical, repeatable, and inexpensive designs, along with performance data.

Resistive Versus Hybrid Power Splitters

Suppose that we have a source of RF power that we wish to drive two loads simultaneously. Let us further assume that the RF source is a 75 ohm TV cable, and the loads are two 75 ohm input TV receivers, and that the splitter to be used for this purpose should:

(1) Present a low VSWR load at any port when the other two ports are terminated in 75 ohms.

(2) Impose low insertion loss.

(3) Provide isolation between the two output ports (so that local oscillator leakage from one TV receiver will not cause interference to the other).

(4) Exhibit characteristics (1) - (3) above over a multi-octave bandwidth.

A circuit sometimes employed to meet these requirements is the two-way 75 ohm matched resistive power splitter of Figure 1.

Note that when any two ports are resistively terminated in 75 ohms, the third port will also appear as a 75 ohm load. Thus, this circuit meets requirement (1) above. Insertion loss of this circuit from the input to either output is 6 dB, which is to say that the use of this circuit imposes a 3 dB penalty over and beyond the inherent 3 dB "split" that necessarily occurs when power is equally divided two ways. Requirement (2) above then is only marginally met. Isolation between the two output ports is likewise only 6 dB which does not meet requirement (3) above very well. With sufficient care in construction, the bandwidth of this circuit can extend from dc to well beyond the UHF range, so requirement (4) above is well met. The most attractive feature of the resistive power splitter is its low cost, and it is this feature alone that prompts many designers to forgive its shortcomings in terms of isolation and insertion loss.

Resistive power splitters may be designed for any number of outputs simply by adding additional ports, each one connected to the summing junction by a suitable resistor. For the two-way 75 ohm matched power splitter of Figure 1, the three resistors must all be 25 ohms as shown. For a three-way 75 ohm matched power splitter, the four resistors must each be 37.5 ohms, and for an N-way matched resistive power splitter in general, the required value would be:

$$R = R_T\left(\frac{N-1}{N+1}\right) \qquad (1)$$

where R = the required value of all resistors
R_T = the desired termination resistance at all ports (75 ohms in the examples used so far)
N = the number of ways the input power is to be split (N is always one less than the total number of ports)

Insertion loss for the N-way matched resistive power splitter is given by the equation:

$$\text{dB insertion loss} = -20\log\left(\frac{1}{N}\right) \qquad (2)$$

A power splitter that better meets our four above requirements is the hybrid "magic tee" power splitter. Operation of this device is such that, for a two-way splitter, power applied to the input port is equally divided and routed to the two output ports. Thus, in its ideal form, the only insertion loss is that caused by the inherent 3 dB "split" mentioned previously. This device also exhibits theoretically perfect isolation between the two output ports, provided that the input port is correctly terminated. A signal source connected to one of the output ports, for example, will not result in any power output at the other output port. Similarly, changes in load impedance at either output port will not change the impedance of the other output port. All ports are matched to the system impedance, and the bandwidth of the device, while not able to extend all the way down to dc, has no high frequency limitation.

Of course, a practical hybrid power splitter does not perform as well as the ideal one discussed above. Insertion loss is invariably greater than the 3 dB split, although in a well de-

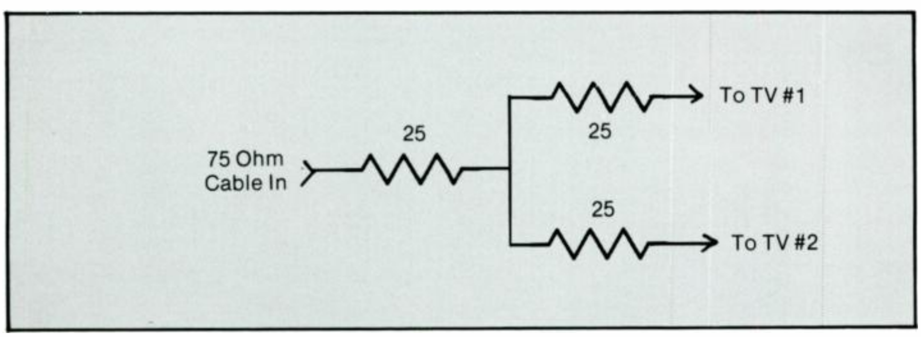

Figure 1. Two-way 75 Ohm Matched Resistive Power Splitter.

signed unit, this additional loss can be held to less than 1 dB. Output port-to-port isolation is not perfect, but isolation in excess of 20 dB is quite attainable. VSWR is generally good over the useable bandwidth of the device (typically in excess of two decades). Even in its practical form then, the hybrid "magic tee" power splitter has significant advantages over the resistive power splitter.

Two-way hybrid power splitters can be designed for inphase, out-of-phase, or quadrature outputs (i.e., "zero degree," "180 degree," or "90 degree" power splitters). Hybrid power splitters can also be designed for any number of outputs. Figure 2a illustrates the use of three two-way splitters to obtain a four-way splitter, while Figure 2b illustrates the use of one three-way splitter and three two-way splitters to obtain a six-way splitter.

Insertion loss of N-way hybrid power splitters due to the "split" may be obtained from the equation:

$$\text{dB insertions loss} = -10 \log \left(\frac{1}{N}\right) \quad (3)$$

Table I compares the losses of resistive and hybrid power splitters for selected N values from 2 to 32 (computed from equations (2) and (3) respectively). Notice that the insertion loss penalty of the resistive splitter increases with N. This additional insertion loss penalty is caused by the fact that as more resistors are added to accommodate higher N values, their values must increase to maintain the correct impedance at each port. This value increase then results in more dissipative loss. In the hybrid splitters, of course, the only theoretical loss is that caused by the "split."

Theory Of The Magic Tee

The hybrid power splitter in its most basic form comprises the "magic tee" circuit shown inside the dotted rectangular box in Figure 3, and consists of nothing more than a center-tapped transformer and a bridging resistor. If the loads connected to output ports A and B are to be 50 ohms as shown (R_a and R_b), then the bridging resistor must be 100 ohms and the source impedance (R_x) of generator E_x driving input port X must be 25 ohms for matched operation.

Generator Ex causes equal currents I_1 and I_2 to flow as shown. I_1 then creates voltage drop E_1 across the transformer left-hand winding, while I_2 creates voltage drop E_2 across the right-hand winding. By transformer action, however, E_1 induces E'_1 into the right-hand winding which cancels E_2, since E'_1 and E_2 are of equal magnitude but opposite polarity. In the same fashion, E_2 induces E'_2 into the left-hand winding cancelling E_1.

Thus, there is no net voltage drop across either transformer winding. If currents I_1 and I_2 are flowing through the windings with no net voltage drop, then the winding impedances each must effectively be zero. Thus, as far as generator Ex and its 25 ohm source resistance Rx are concerned, the effective load is simply 50 ohm resistors Ra and Rb in parallel, or a net load of 25 ohms. The system is therefore matched, and the source power equally divides between the two 50 ohm load resistors with no loss over and beyond the 3 dB "split." The 100 ohm bridging resistor can be ignored in this instance, since the equal voltages at points A and B allow no current to flow through this resistor.

The most confusing aspect of the "magic tee" for many people is the means by which isolation exists between ports A and B. The nature of this isolation is easily explained. Referring to Figure 4, we have the same "magic tee" as in Figure 3, but with the generator (E_a) placed in series with R_a (one of the two 50 ohm load resistors) rather than R_x.

E_a causes current I_1 to flow, resulting in voltage drop E_1 across the left-hand winding and voltage drop E_2 across the 25 ohm resistor. Since the effective impedance of the left-hand winding is also 25 ohms (the 100 bridging resistor is stepped down to 25 ohms by 4:1 autotransformer action) E_1 must equal E_2. By transformer action, E_1 results in E'_1 appearing across the right-hand winding. Since E'_1 is equal to E_1, E'_1 must also equal E_2. However, since E'_1 and E_2 are of opposite polarity, resulting currents I'_1 and I_2 in the right-hand loop cancel, and there is thus no net current flow through R_b. Since no power from generator E_a has reached R_b, we therefore have isolation between ports A and B.

Furthermore, since no current flows through R_b, the isolation exists for any value of R_b, and the amount of power transferred to 25 ohm resistor R_x remains unchanged. Perfect isolation can only exist if the 100 ohm bridging resistor is four times the value of R_x. For any other ratio, E_2 will no longer equal E_1 and E'_1, and right-hand loop currents I_2 and I'_1 will no longer cancel, thus resulting in some current flow through R_b.

N	dB Loss, Resistive Splitter	dB Loss, Hybrid Splitter	dB Penalty, Resistive Splitter
2	6.02	3.01	3.01
3	9.54	4.77	4.77
4	12.04	6.02	6.02
6	15.56	7.78	7.78
8	18.06	9.03	9.03
9	19.08	9.54	9.54
12	21.58	10.79	10.79
16	24.08	12.04	12.04
18	25.10	12.55	12.55
24	27.60	13.80	13.80
32	30.10	15.05	15.05

Table 1.

The power delivered to R_x in Figure 4 is simply:

$$P = \frac{E_2^2}{25} \text{ WATTS} \quad (4)$$

The power delivered to the 100 ohm bridging resistor is:

$$P = \frac{(E_1 + E'_1)^2}{100} \text{ WATTS} \quad (5)$$

Since $E_1 = E'_1 = E_2$, we can rewrite equation (5) as:

$$P = \frac{(E_2 + E_2)^2}{100}$$

$$= \frac{4\,E_2^2}{100}$$

$$= \frac{E_2^2}{25} \text{ WATTS} \quad (6)$$

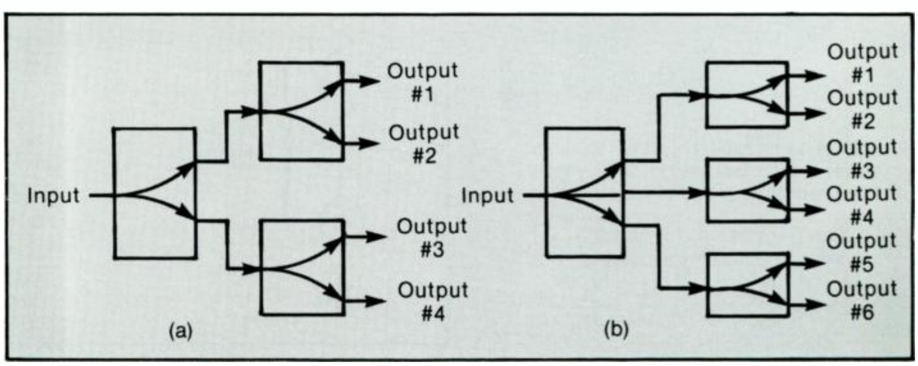

Figure 2. Four-and Six-way Hybrid Power Splitters.

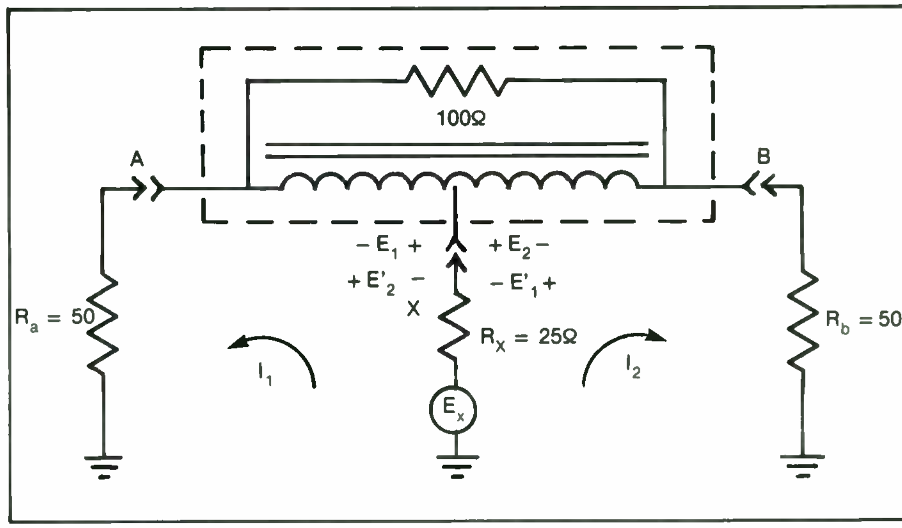

Figure 3. Basic "Magic Tee" Hybrid Power Splitter.

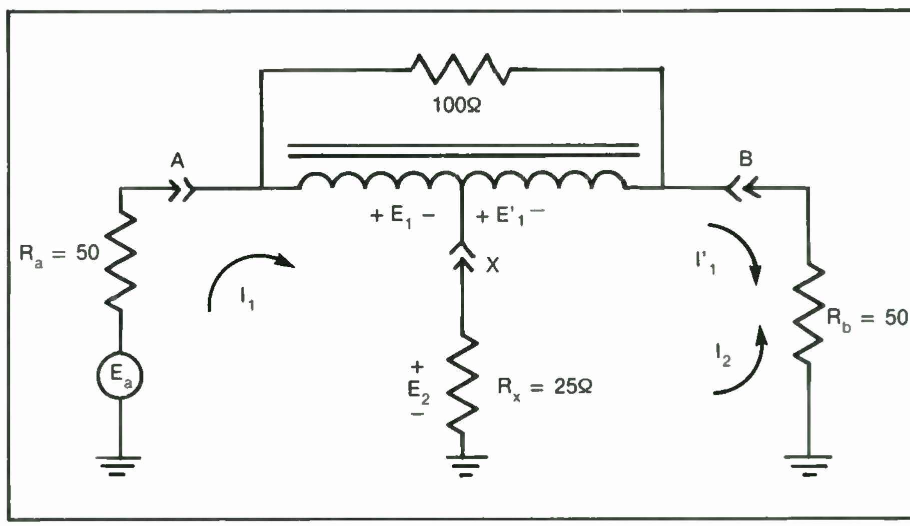

Figure 4. Isolation from port A to Port B occurs due to cancellation of currents I'1 and I2 in right-hand loop.

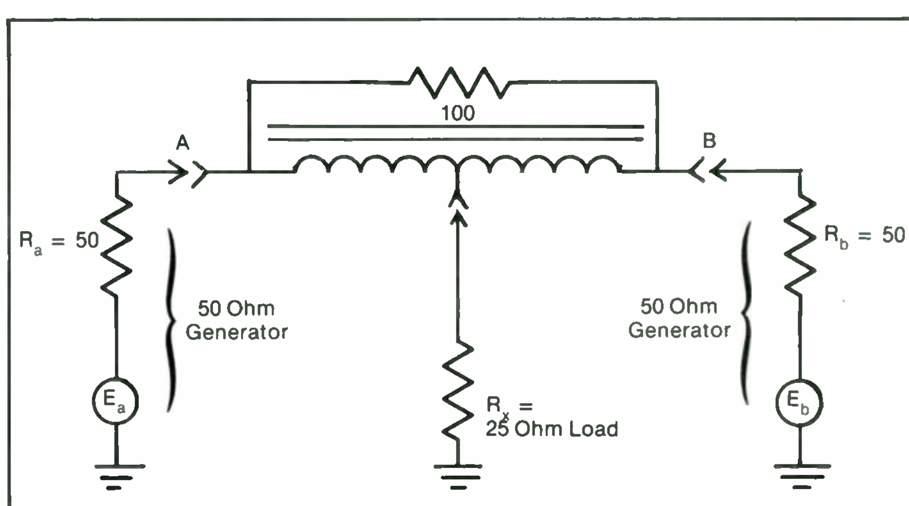

Figure 5. Basic "Magic Tee" Hybrid Power Combiner.

Since equation (6) is identical to equation (4), the power delivered to the 100 ohm bridging resistor must equal the power delivered into R_x. Thus, half the input power at port A is delivered to R_x and the remaining half dissipated in the bridging resistor.

By adding a second voltage source (placing it in series with R_b as illustrated in Figure 5) and considering R_x as the load resistor, the "magic tee" becomes a two-way hybrid power combiner. In this case, consider E_a and R_a, along with E_b and R_b to be 50 ohm signal generators, and R_x as the desired 25 ohm load. Isolation still exists between ports A and B, but the amount of power that either 50 ohm signal generator delivers to the 25 ohm load is influenced by the frequency, phase, and amplitude of the signal delivered by the other 50 ohm signal generator. The rules for determining the amount of power delivered to the 25 ohm load are as follows:

Case 1 — The two inputs are at different frequencies: Loss from either input port to the 25 ohm load is 3 dB. Circuit analysis is identical to that presented for Figure 4.

Case 2 — The two inputs are at the same frequency, of the same amplitude, and in phase: Since the instantaneous voltages at ports A and B of Figure 5 are always the same, no power is dissipated in the 100 ohm bridging resistor. Thus all power delivered to the input ports is transferred to the 25 ohm load.

Case 3 — The two inputs are at the same frequency and of the same amplitude, but not in phase: The insertion loss from either input port to the 25 ohm load is found by the equation:

$$\text{I.L.} = -10 \log \cos^2(\theta/2) \qquad (7)$$

where I.L. is the dB insertion loss and θ is the phase difference between the two input signals. The two extremes occur when the two inputs are exactly in phase (which results in no insertion loss as in Case 2 above) and when the two inputs are exactly out-of-phase (which results in infinite insertion loss due to all the power being dissipated in the bridging resistor).

Case 4 — The two inputs are the same frequency and in phase, but

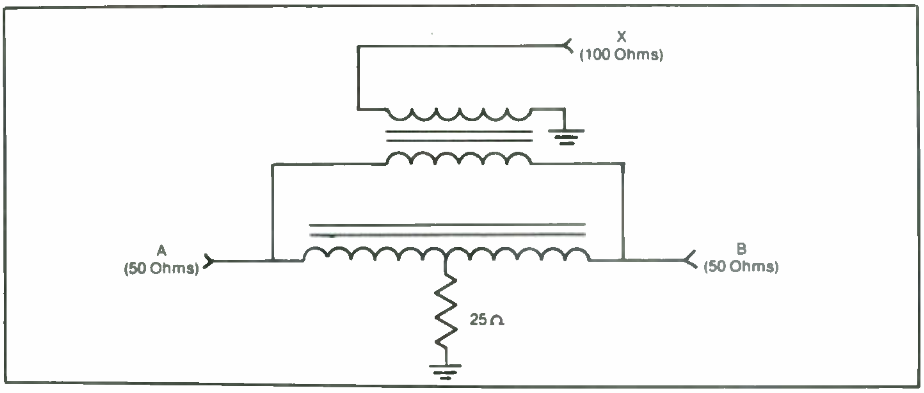

Figure 6. Basic 180 degree "Magic Tee".

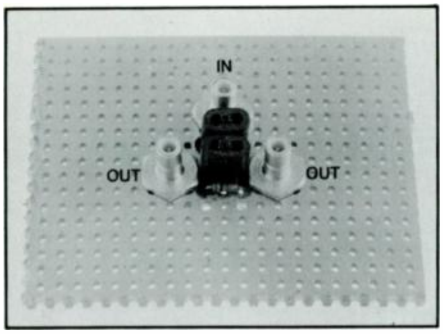

Figure 7. Photograph of 2-300 MHz ("HF/VHF") Hybrid Power Splitter. The compensation capacitor is mounted on the far side of the vector board.

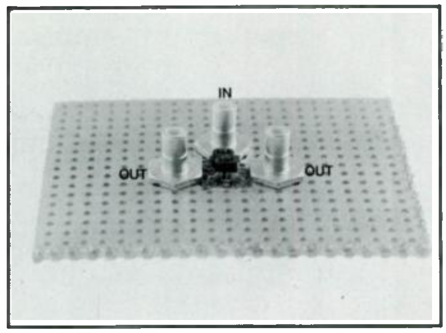

Figure 8. Photograph of 5-600 MHz ("VHF/UHF") Hybrid Power Splitter. The compensation capacitor is mounted on the far side of the vector board.

of different amplitudes: The insertion loss from either input port to the 25 ohm load is found by the equation:

$$\text{I.L.} = -10\log\left(0.5 + \frac{\sqrt{P1\ P2}}{P1 + P2}\right) \qquad (8)$$

where I.L. is the dB insertion loss, P_1 is the power delivered into either input port, and P_2 is the power delivered into the other input port. The insertion loss extremes range from zero dB (when P_1 and P_2 are equal) to 3 dB (when either P_1 or P_2 is zero).

Case 5 — The two inputs are at the same frequency, but are not in phase, nor of equal amplitudes: The insertion loss from either input port to the 25 ohm load is found by the equation:

$$\text{I.L.} = -10\log\left(0.5 + \frac{\cos\theta\sqrt{P1\ P2}}{P1 + P2}\right) \qquad (9)$$

where I.L., θ, P_1, and P_2 are previously defined.

So far, our discussion has dealt with zero degree (in-phase) hybrid power splitters and combiners. As it happens, the same "magic tee" circuit can be employed as a 180 degree two-way hybrid power splitter/combiner by simply terminating the 25 ohm output port in a 25 ohm resistor and replacing the 100 ohm bridging resistor with a 100 ohm output port as illustrated in Figure 6. Notice that the 100 ohm port employs a 1:1 isolation transformer to permit this port to be ground referenced. In some designs, this isolation transformer is combined with the "magic tee" transformer by simply adding a secondary winding.

As a 180 degree hybrid power splitter, a signal applied to the 100 ohm port divides into two equal amplitude (but out-of-phase) outputs at ports A and B (assuming both ports are terminated in 50 ohms). Under these circumstances, no power is lost in the 25 ohm termination resistor, and half the input power is delivered to each 50 ohm load. Thus, the only loss from the input to each output is that caused by the 3 dB "split," just as is the case for the zero degree hybrid splitter.

As a 180 degree hybrid power combiner, the signal sources are applied in ports A and B, and the combined output appears at port X. As is the case for the zero degree combiner, the amount of power that either 50 ohm signal source delivers to the load (the 100 ohm port in this case) is influenced by the frequency, phase, and amplitude of the signal delivered by the other 50 ohm signal source. The rules for determining the amount of power delivered to the 100 ohm output port are as follows:

Case 1 — The two inputs are at different frequencies: Loss from either input port to the output is 3 dB.

Case 2 — The two inputs are at the same frequency and of the same amplitude, but 180 degrees out-of-phase: All input power is delivered to the output.

Case 3 — The two inputs are at the same frequency and of the same amplitude, but are not 180 degrees out-of-phase: The insertion loss from either input port to the output port is found by the equation:

$$\text{I.L.} = -10 \log \sin^2(\theta/2) \qquad (10)$$

where I.L. and θ have been previously defined. The two extremes occur when the two inputs are 180 degrees out of phase (which results in no insertion loss as in Case 2 above), and when the two inputs are exactly in phase (which results in infinite insertion loss due to all the power being dissipated in the 25 ohm resistor).

Case 4 — The two inputs are at the same frequency and are 180 degrees out-of-phase, but are of different amplitudes: The insertion loss from either input port to the output port is given by equation (8) above. The insertion loss extremes range from zero dB (when P_1 and P_2 are of equal amplitude) to 3 dB (when either P1 or P2 is zero).

Case 5 — The two inputs are at the same frequency, but are not 180 degrees out-of-phase nor of equal amplitudes: The insertion loss from either input port to the output port is found by the equation:

$$\text{I.L.} = -10 \log \left(0.5 - \frac{\cos\theta \sqrt{P1 \ P2}}{P1 + P2} \right) \qquad (11)$$

where I.L., 0, P1, and P2 are previously defined.

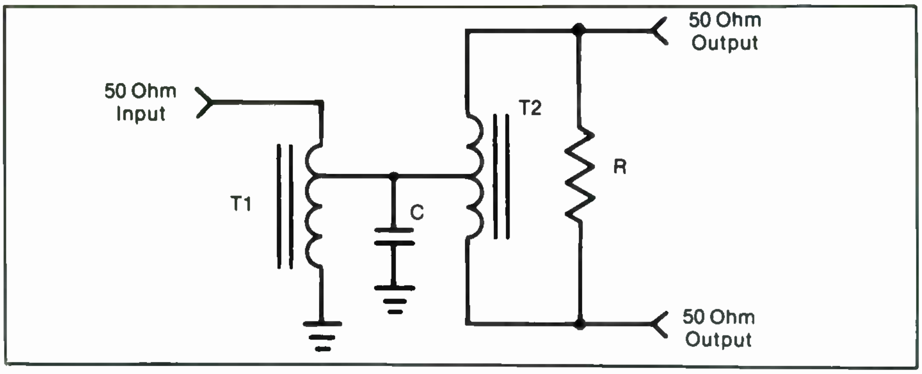

Figure 9. Practical two-way 50 Ohm Hybrid Power Splitter (See text for component valves).

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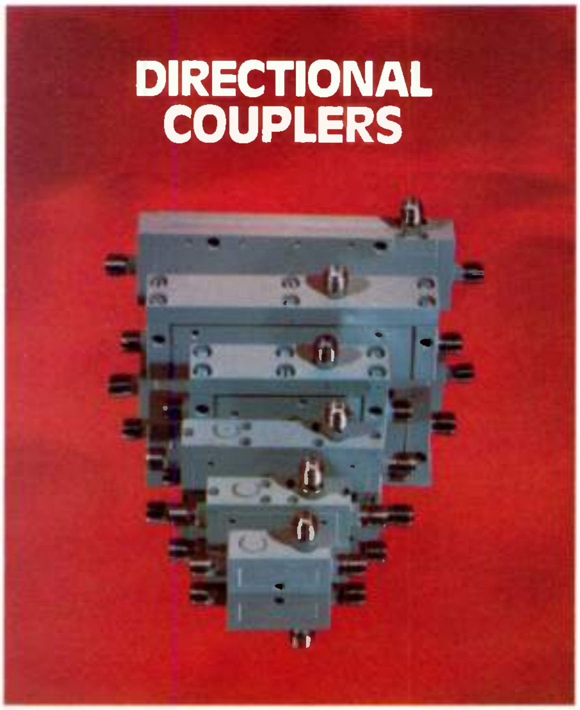

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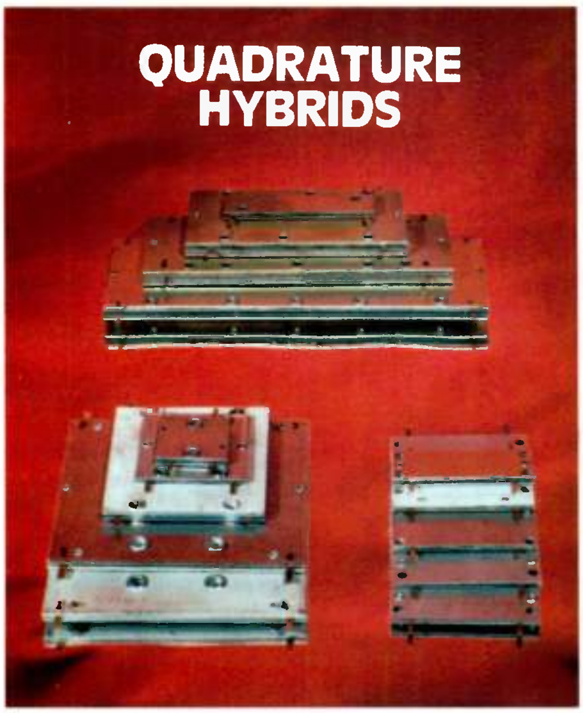

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As is true for the zero degree hybrid power splitter/combiner, isolation exists between ports A and B in the 180 degree hybrid power splitter/combiner. Once again, maximum isolation between ports A and B requires that the ratio of the bridging resistor (i.e. the load seen by the 100 ohm port and the 25 ohm summing resistor) be exactly 4:1.

By providing matched output ports in place of both the 25 ohm summing resistor and the 100 ohm bridging resistor, a θ/180 degree hybrid can be obtained. As a power combiner, if two equal amplitude in-phase signals are applied to input port A and B, all the power will appear at the sum (25 ohm) port. If the two signals are of equal amplitude but 180 degrees out of phase, all the power will appear at the difference (100 ohm) port.

For convenience of explanation, we have so far left the 35 ohm sum port and the 100 ohm difference port at their respective "natural" impedances. In practice, where it is usually more convenient for all ports to be of the same impedance, RF transformers are employed to provide the required impedance transformation. This technique is employed in the practical hybrid power splitters/combiners to be described.

Prototype Hybrid Power Splitters

Two prototype 50 ohm two-way hybrid power splitters were constructed in order to demonstrate performance characteristics obtainable using readily available and inexpensive components. The first of these prototypes was designed around Fair-Rite Products 2843002402 two-hole miniature ferrite balun cores, and yields good performance from 2 to over 300 MHz (this prototype is dubbed the "HF/VHF" splitter). The second splitter (dubbed the "VHF/UHF" splitter) was designed around Fair-Rite Products 2843002302 two-hole subminiature ferrite balun cores, and provides good performance from 5 to 600 MHz. Figures 7 and 8 are photographs of these splitters.

The schematic for these two splitters is shown in Figure 9 (the component values for the two respective prototypes are discussed below). The only difference between this schematic and that of the basic "magic tee' presented earlier is the addition of 2:1 autotransformer T1 at the input (to transform the desired 50 ohm input down to the 25 ohm level necessary for proper operation) and compensation capacitor C. Subminiature coaxial connectors are employed at all ports to facilitate performance measurements.

The secret to good performance in a wideband hybrid power splitter is in the proper selection of the transformers and very "tight" construction so as to minimize lead lengths. See references (1) and (2) for detailed information concerning the design and construction of wideband ferrite transformers. Another important factor is the use of a compensation capacitor (C is Figure 9) to extend the high frequency bandwidth. For any given impedance level, bandwidth requirement, and physical layout, the precise configuration of T1 and T2, and the value of C are most easily established empirically.

In the "HF/VHF" prototype splitter, T1 is wound with 3 turns #32 single polyurethane wire through both holes of a Fair-Rite Products 2843002402 miniature two-hole ferrite balun core, with the tap brought out 2 turns from ground (see Figure 10). T2 is wound on the same type of core with 3 turns #32 bifilar wire (single polyurethane, 10 twists/inch) with the windings connected series aiding (this effectively results in a 6 turn center-tapped 1:4 autotransformer). The use of bifilar wire in T2 results in better transformer balance, and therefore better port-to-port isolation. R is a 100 ohm 1/8 watt 5 percent carbon film resistor, and C is a 15 pF disc ceramic capacitor.

Performance of the "HF/VHF" splitter is graphically illustrated by the solid lines of Figures 11, 12, and

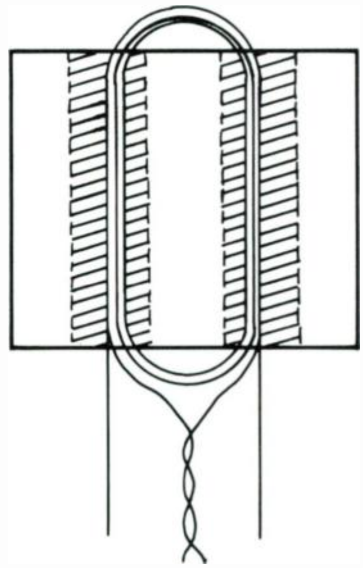

Figure 10. Cutaway drawing of transformer.

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13. Figure 11 shows insertion loss (in excess of the 3 dB "split"). Notice that this insertion loss is under 1 dB from 2 to over 300 MHz. Figure 12 shows the input VSWR being well under 1.5:1 in the 2 to 300 MHz range. Figure 13 shows output port-to-port isolation in excess of 20 dB over the same frequency range.

Some comments concerning the measurement techniques are in order. For insertion loss measurements, the

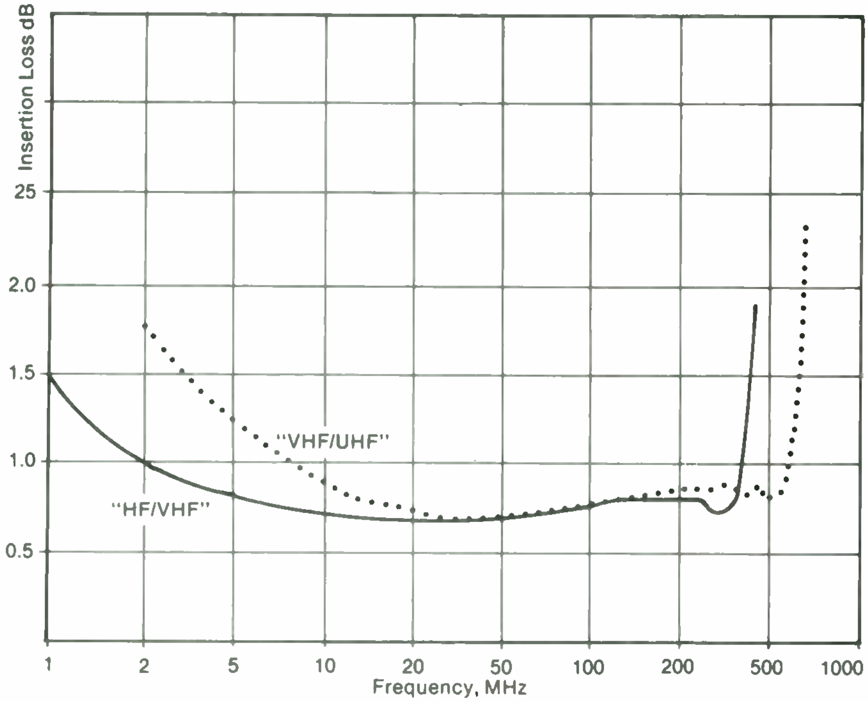


Figure 11. Insertion loss of prototype Hybrid Power Splitters.

RF generator is applied to the splitter input while the detector is connected to either output port with the other output port terminated in a 50 ohm load. For input VSWR measurements, the rho bridge is connected to the input port with the output ports both terminated in 50 ohm loads. For isolation measurements, the RF generator is applied to either output port while the detector is connected to the other output port with the input port termi-

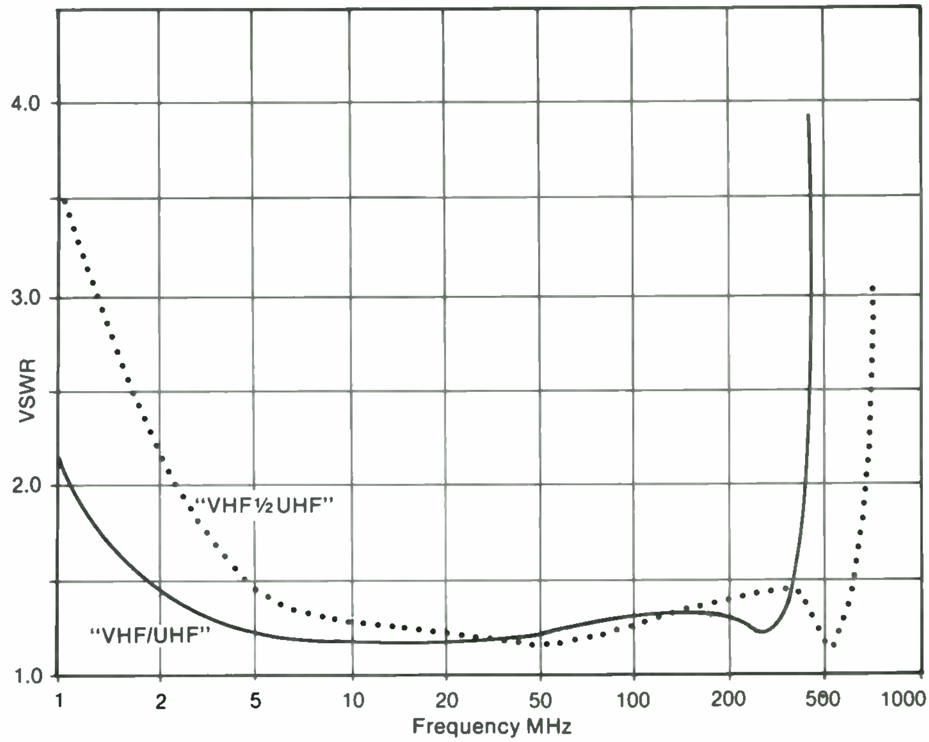
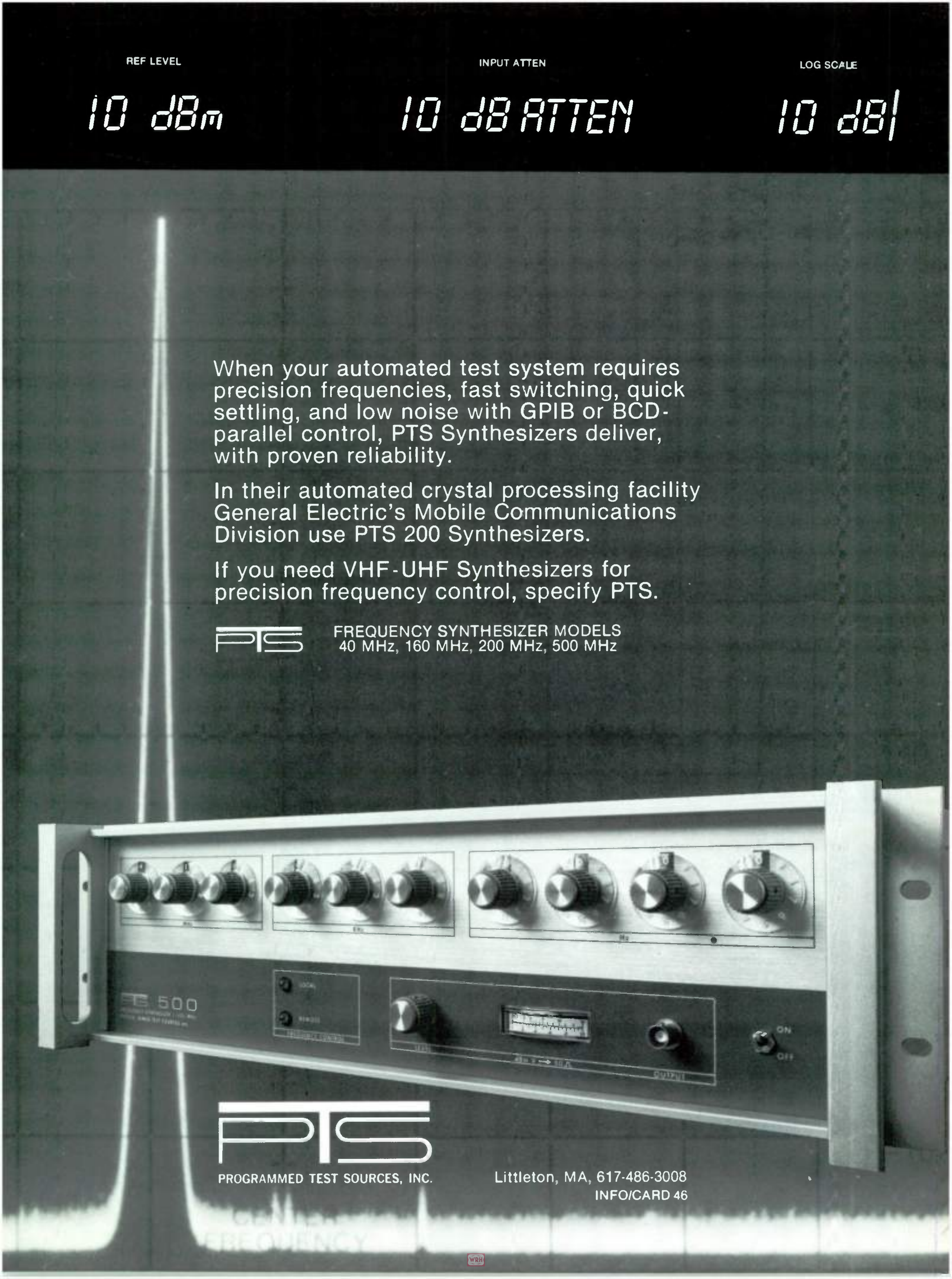


Figure 12. Input VSWR of prototype Hybrid Power Splitters.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nated in a 50 ohm load. See reference (2) for additional information concerning transmission and reflection measurements.

For best measurement accuracy, the 50 ohm terminations employed must have very low VSWR over the desired frequency range. This is particularly true for the isolation measurement. As we have seen, the degree of isolation is impaired when the T2 center tap does not see precisely R/4 to

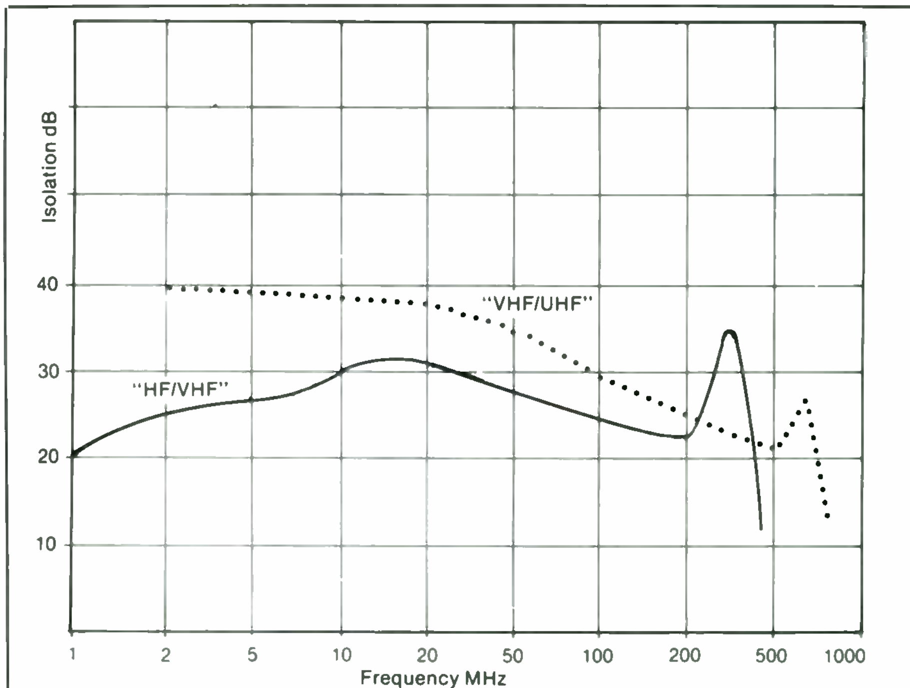

Figure 13. Isolation of prototype Hybrid Power Splitters.

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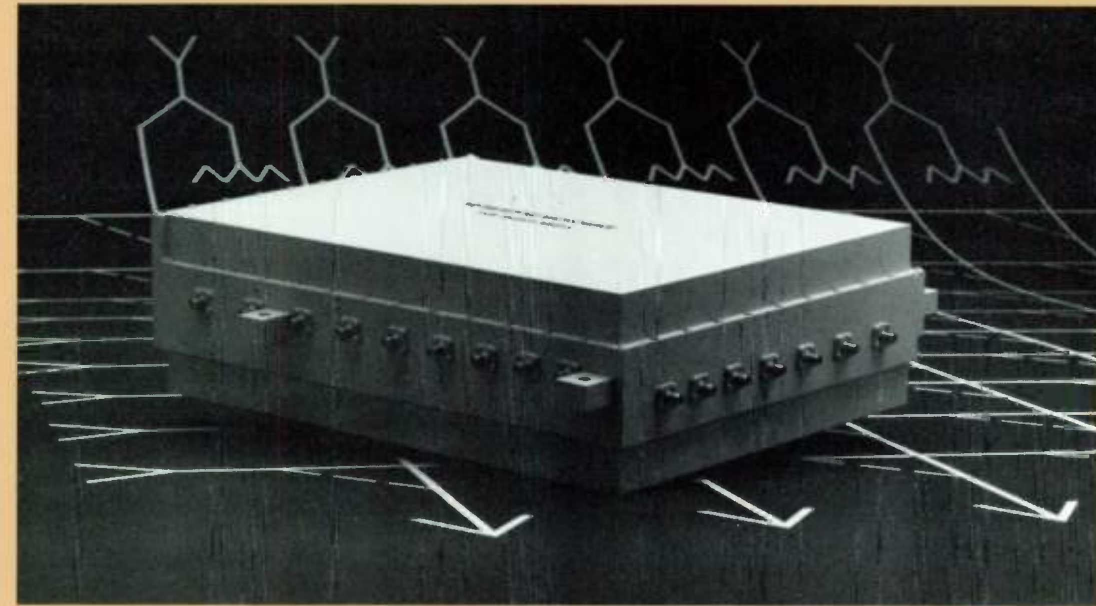

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ground. In an isolation measurement, we are attempting to measure the *inherent* isolation of the splitter rather than the degraded isolation resulting from a mismatched termination on the input port.

Performance of the "VHF/UHF" splitter is graphically illustrated with the dotted lines of Figures 11, 12, and 13, showing good performance between 5 and 600 MHz. Since smaller ferrites are employed, high frequency performance is improved at the expense of low frequency performance. T1 is wound with 3 turns #36 single polyurethane wire through both holes of a Fair-Rite Products 2843002302 subminiature two-hole ferrite balun core with the tap brought out 2 turns from ground. T2 is wound on the same type of core with 2 turns #36 bifilar wire (single polyurethane, 10 twists/inch) with the windings connected series-aiding. R is again a 100 ohm 1/8 watt carbon film resistor, and C is a 7 pF disc ceramic capacitor.

Although the focus on these two prototype hybrids has been on their use as splitters, they also function equally well as combiners. Output port VSWR (not illustrated) is as good as or better than the input VSWR.

Conclusion

"Magic tee" hybrids provide an efficient and economical means of splitting or combining power with the added bonus of providing substantial port-to-port isolation. Although the principle of the "magic tee" may at first appear subtle, the circuitry is very simple and can be designed around easily constructed RF transformers using inexpensive and readily available ferrite cores. This simplicity and good performance was demonstrated in the construction and test results of two prototype hybrid power splitters. Although one's specific requirements may not necessarily be met by either of the two prototypes presented, the designer should, aided by the information provided by this paper, be able to construct a satisfactory and economical hybrid power splitter/combiner to satisfy his own particular bandwidth and impedance level requirements. □

References

1. Martin, W.A., "Use of Ferrites for Wide Band Transformers." Application Note published by Fair-Rite Products Corporation, Wallkill, NY 12589.
2. Burwasser, A.J., "Wideband Monofilar Autotransformers," *r.f. design* magazine, January and March, 1981. Vol. 4, No. 1, Jan/Feb 1981, Vol. 4 No. 2, Mar/Apr 1981.

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Crystal Oscillator Circuits for VHF

Which crystal Oscillator circuit should you select for your application? This article offers some suggestions.

By Robert J. Matthys

Do you need a VHF crystal oscillator? This article describes how to design and use three good but relatively unknown VHF oscillator circuits. Each one works better than Butler's Common Base circuit, which is the only VHF circuit mentioned in most articles, and whose performance is only fair at best.

VHF crystal oscillators are available at frequencies up to at least 200 MHz. They oscillate at one of a crystal's odd harmonic frequencies, and use an LC circuit to select which crystal harmonic to oscillate at.

The basic schematics of the three VHF oscillator circuits and the Common Base circuit are shown in Figure 1, along with a short description of their primary characteristics. Only two things will be said about Butler's Common Base circuit: 1) it suffers badly from parasitics, and 2) the transistor's collector current must be held within a relatively narrow range or the circuit won't oscillate properly. Now on to the other circuits.

Butler's Emitter Follower

A schematic of Butler's Emitter Follower at 100 MHz is shown in Figure 2. This circuit has good performance without any parasitics because its emitter follower amplifier has a gain of only one, with built-in negative feedback to stabilize its gain. The circuit operates as follows; C_2 acts as a capacitive load for the crystal, and its impedance is set equal to or slightly less than the crystal's internal series impedance R_s, which is 40 Ω in this case. C_1 is set near series resonance with L_1 providing both a loop gain of about 2X for the oscillator and a phase lead to compensate the phase lag of the crystal with its capacitive load C_2.

R_1 gives the emitter follower a low output impedance for driving the crystal. Diodes D_1-D_4 provide a simple regulated bias source. Diodes D_5-D_6 limit the oscillation amplitude, 1) to keep the crystal's power dissipation within its limit, 2) to prevent the transistor from shutting off or saturating at any point over the oscillation cycle, thus giving a relatively constant source impedance to the crystal, and 3) to provide a sinusoidal drive waveform without any spikes to the crystal.

Above about 50-70 MHz, the impedance of the crystal's shunt terminal capacitance C_o becomes low enough to start "shorting out" the crystal. An inductance L_o is frequently placed across the crystal, as shown in Figure 2, to parallel resonate with the crystal's terminal capacitance C_o and tune it out of the circuit. The crystal has a relatively high in-circuit Q, since its load (C_2) is capacitive and not resistive, and its 25 Ω emitter source impedance is relatively low with respect to the crystal's 40 Ω internal resistance R_s.

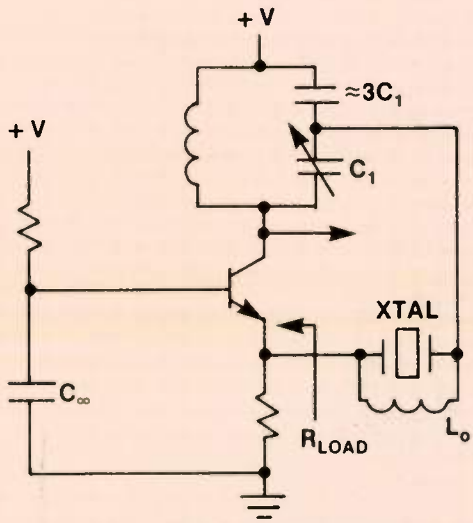
Figure 3 shows a schematic for 20 MHz operation, along with circuit waveforms. Note the characteristic phase lag from point E to point A in Figure 3, and the phase lead from point A to point B. This circuit shares one property in common with Butler's Common Base circuit, in that the oscillator's loop gain varies with the value of C_1 and with the value of the inductance L_1. Figure 4 shows the calculated loop gain versus C_1 for a typical Butler Emitter Follower circuit[1]. The loop gain is shown as a function of the circuit's oscillation frequency, which changes as C_1 is changed. For a given capacitance load C_2 on the crystal, there is an optimum value of inductance L_1 for maximum loop gain. Picking the optimum inductance value of 1 μH for L_1, as shown in Figure 4, the value of C_1 must be kept within 60-100 pF or the loop gain will become less than one and the circuit will stop oscillating.

At frequencies of 50 MHz or less, transistor Q_1 should be a 2N5179. At 100 MHz and above, transistor Q_1 should be an MRF904. And for frequencies between 50-100 MHz, both transistors should be tried to see which one works best.

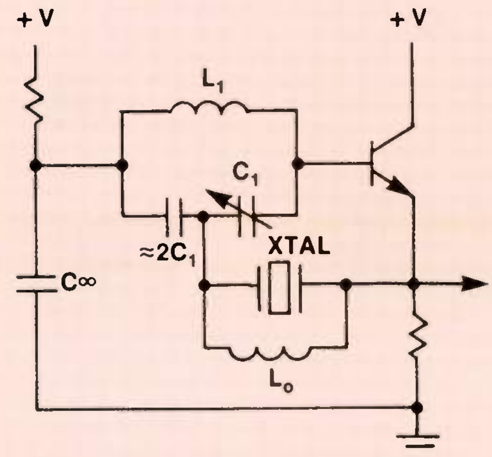
In practice, Butler's Emitter Follower circuit has good frequency stability and works very well. It also has a low output impedance, which may be useful in some applications. Its upper frequency limit is higher than those of the other circuits, primarily because its amplifier only has to provide a gain of one.

Colpitts Harmonic

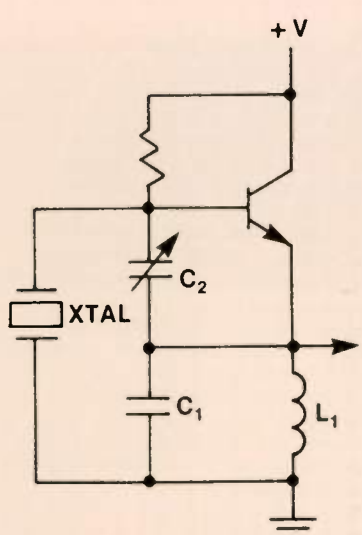
The Colpitts fundamental circuit is well known, but the harmonic version is not. Figure 5 shows a Colpitts Har-



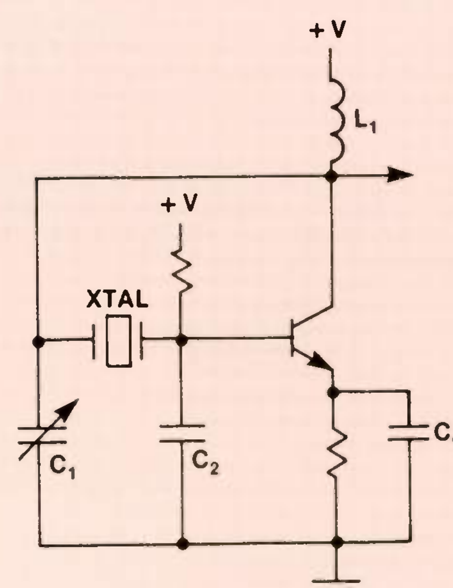
a) Butler Common Base. Operates at or near series resonance. Fair to poor circuit design. Has parasitics, touch to tune. Fair frequency stability.



b) Butler Emitter Follower. Operates at or near series resonance. Good circuit design. No parasitics, easy to tune. Good frequency stability.



c) Colpitts Harmonic. Operates 30-200 PPM above series resonance. Physically simple, but analytically complex. Low cost. Fair frequency stability.



d) Pierce Harmonic. Operates 10-40 PPM above series resonance. Good circuit design. Good to very good frequency stability.

Figures 1a-d. Basic Circuits for VHF Crystal Oscillators.

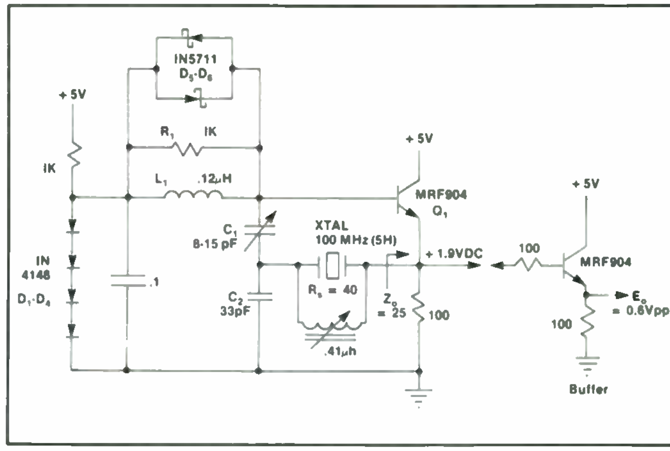


Figure 2. Butler's Emitter Follower At 100 MHz.

monic circuit at 100 MHz. L_1C_1 are selected to be resonant at a frequency below the desired crystal harmonic but above the crystal's next lower odd harmonic. C_2 should have a value of 30-70 pF, independent of the oscillation frequency. There is no requirement for any specific ratio of C1/C2, but practical harmonic circuits seem to work best when C_1 is approximately 1-3 times the value of C_2. Diodes D_1-D_3 provide a simple regulated bias supply. The resistance

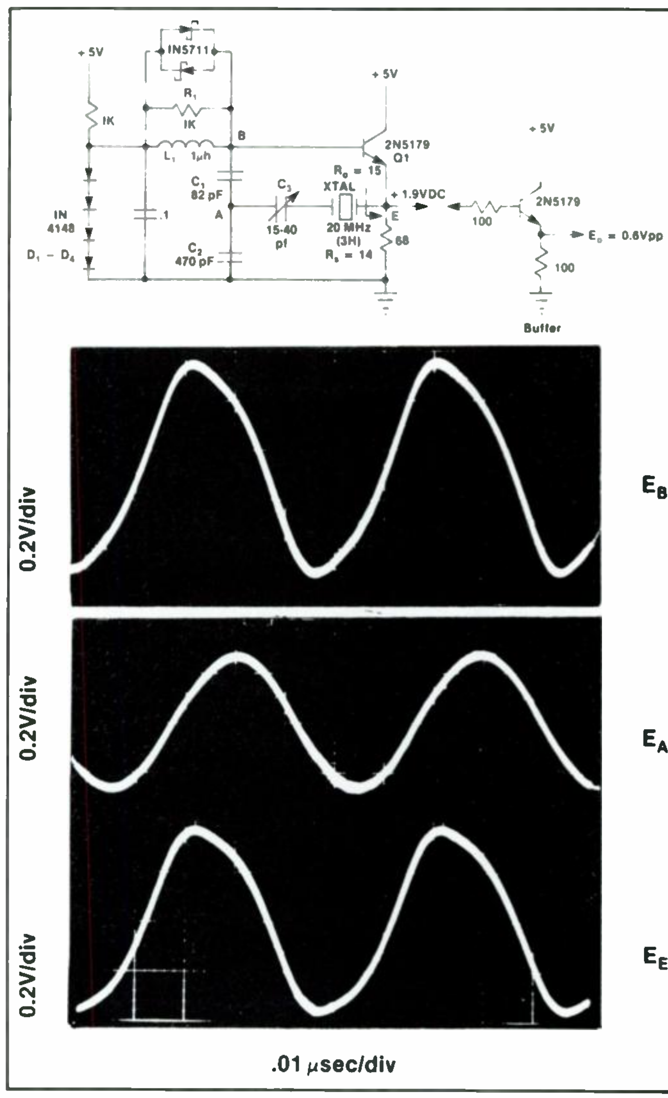
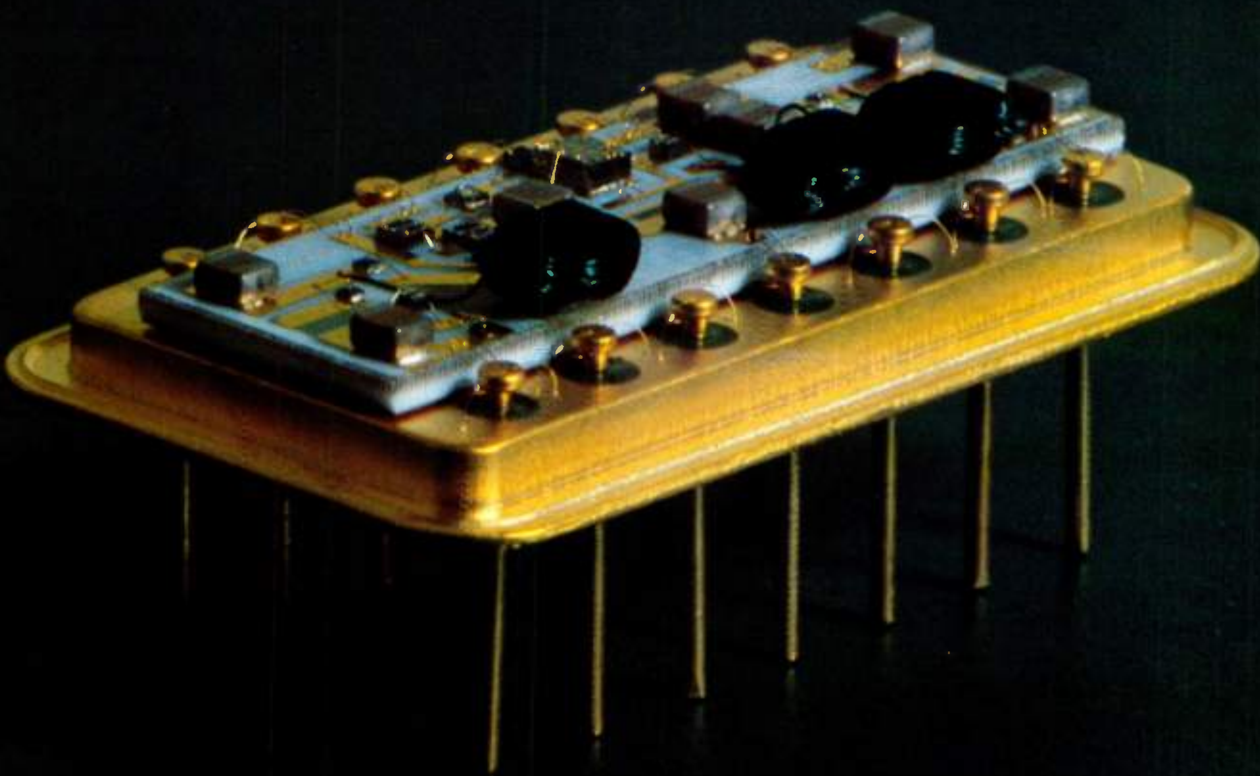


Figure 3. Butler's Emitter Follower at 20 MHz.



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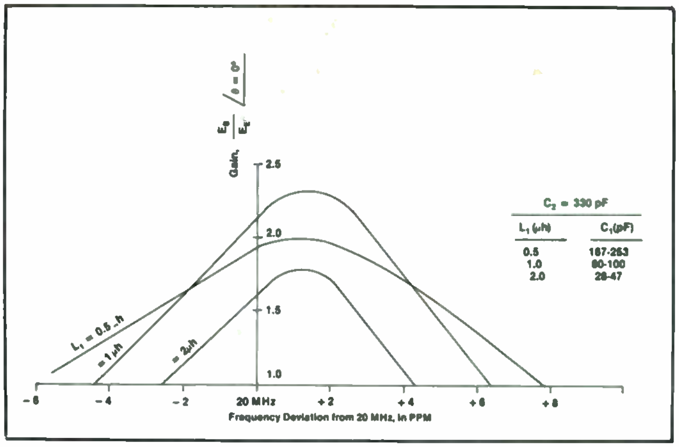


Figure 4. Calculated Loop Gain Vs. Frequency for Butler's Emitter Follower Circuit.

of R_1 should be as high as possible, as it affects the crystal's in-circuit Q.

Figure 6 shows a Colpitts Harmonic circuit for 20 MHz operation, along with circuit waveforms.

The Colpitts Harmonic has fair frequency stability. It is the easiest VHF circuit to design, because of its physical simplicity. It is a complex circuit to analyze, however, because the transistor conducts current only during a small portion of the oscillation cycle, creating two oscillation states whose boundary conditions must be matched to each other.

Pierce Harmonic

The Pierce fundamental circuit is also well known, but its harmonic version is not. The Pierce Harmonic circuit was developed by Belcher[2], with earlier work by Frerking[3], Figure 7 shows a circuit for 100 MHz operation. The output resistance R_o of the transistor's collector, together with the effective value of C_1, provides an RC phase lag of 30-50°. The crystal normally oscillates slightly above series resonance, where it is both resistive and inductive. Above series resonance the crystal's internal impedance (resistive and inductive) together with C_2 provides an RLC phase lag of 130-150°. The transistor inverts the signal, providing a total of 360° (or 0°, depending on your point of view) of phase shift around the loop. Inductor L_1 is selected to resonate with C_1 at a frequency between the crystal's desired harmonic and its next lower odd harmonic. Inductor L_1 is in parallel with C_1, and offsets part of the negative reactance of C_1 at the oscillation frequency. This reduces

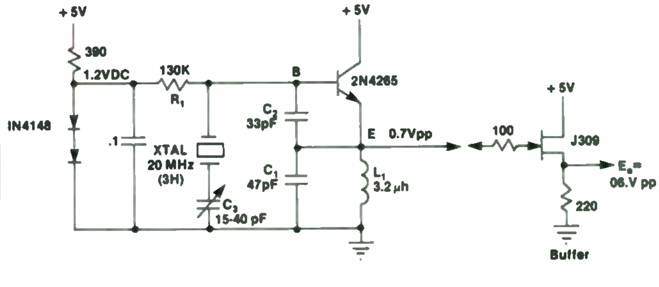
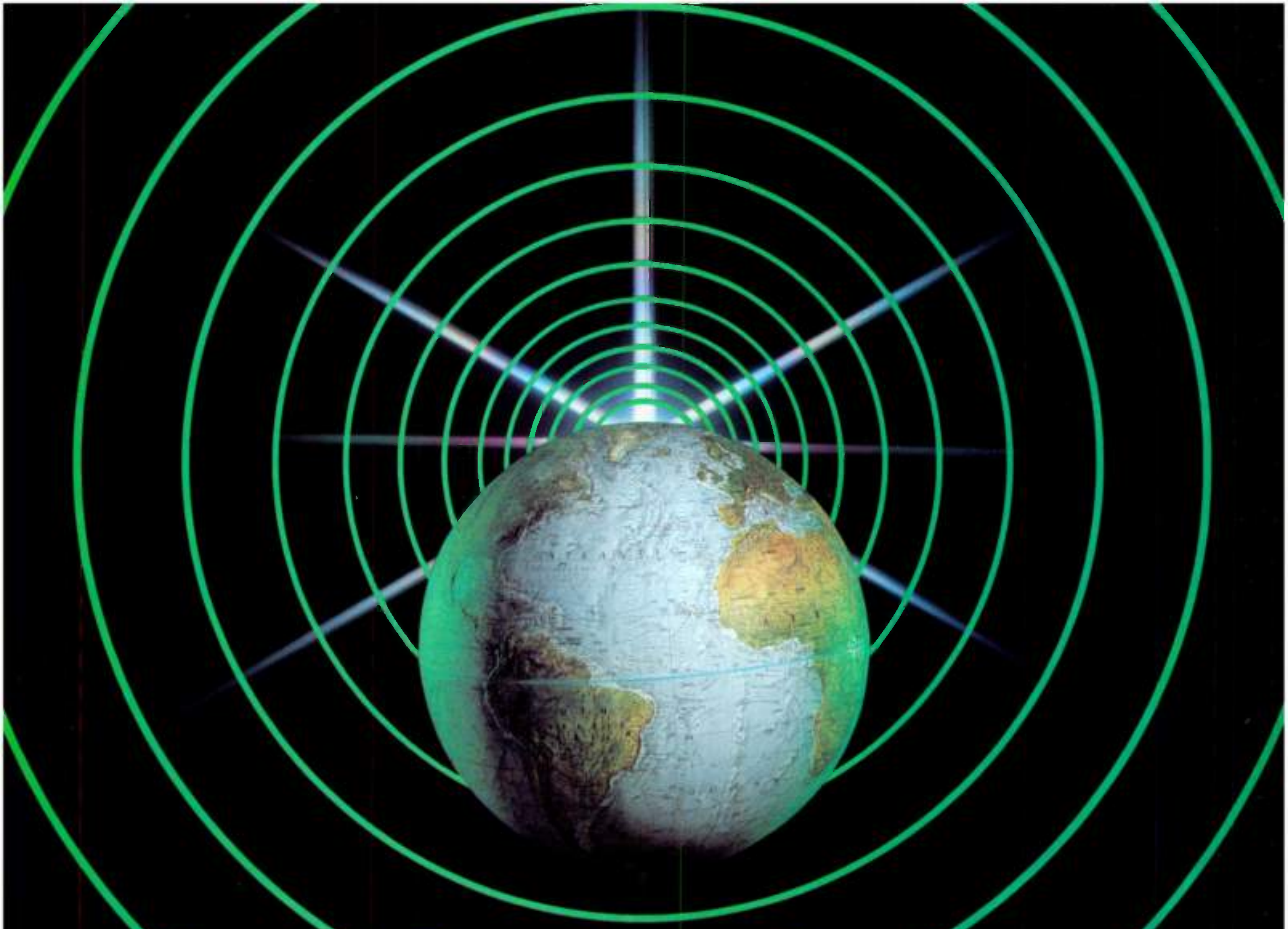


Figure 5. Colpitts Harmonic At 100 MHz.



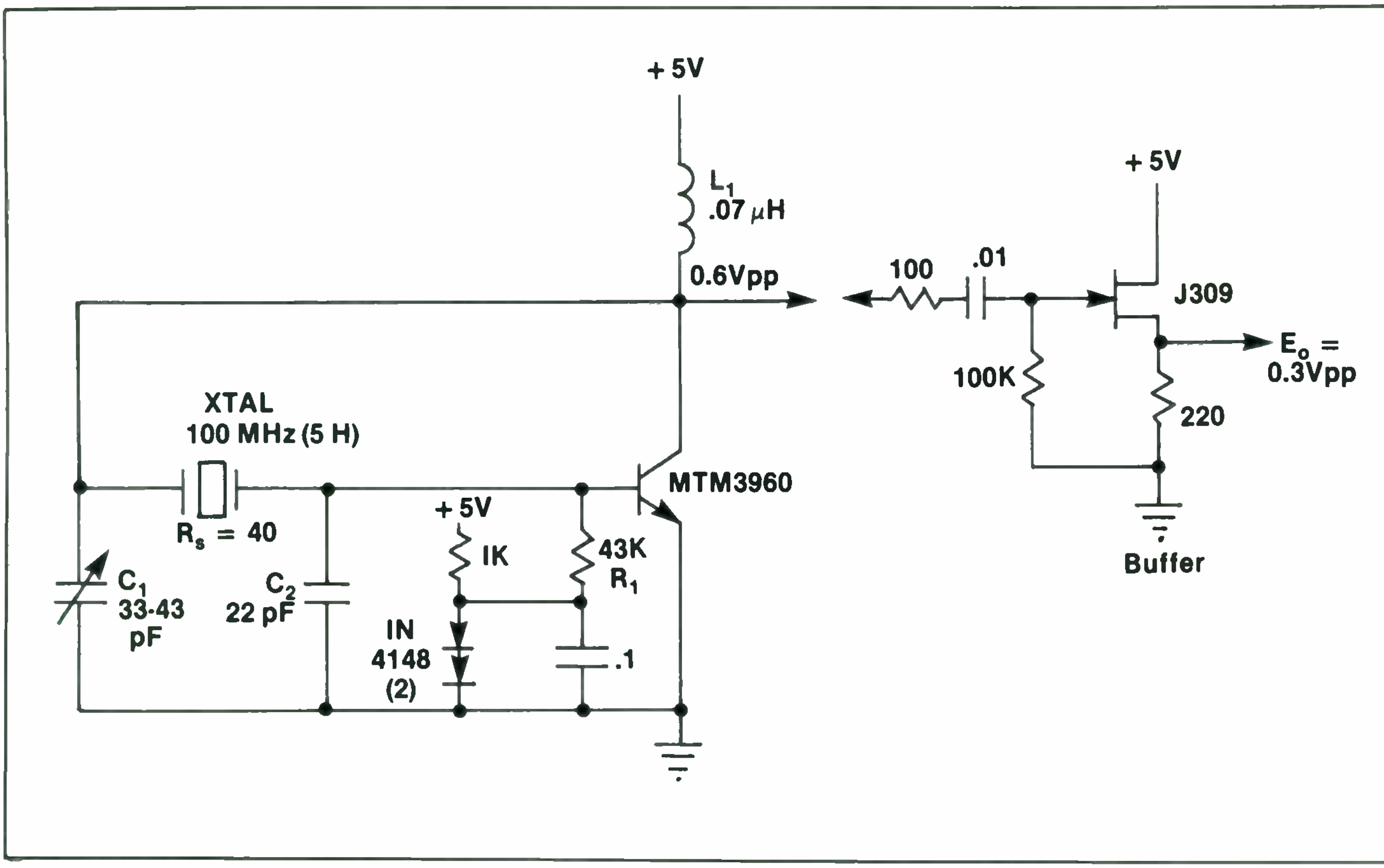


Figure 7. Pierce Harmonic at 100 MHz.

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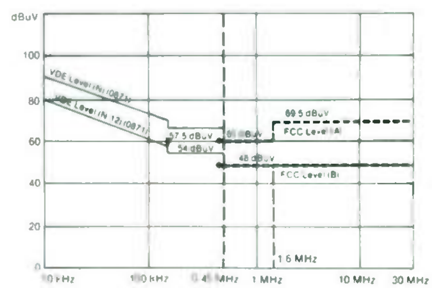

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the effective value of C_1 to something less than its actual value, in making its RC phase lag with R_o. The oscillation amplitude is sensitive to the transistor's base bias current, and is controlled by adjusting the bias resistor R_1.

Both C_2 and the effective value of C_1 should be made as large as possible. Their impedances should preferably be equal to or somewhat less than the crystal's internal series resistance R_s to lower the circuit's frequency sensitivity to changes in the capacitors or in the collector's output resistance R_o and increase the crystal's in-circuit Q, which is helped considerably by the non-resistive shunting of C_1 and C_2 on the crystal. The maximum values of C_1 and C_2 are limited by their loop gain losses, which means the transistor should be selected for high gain.

Figure 8 shows a Pierce Harmonic at 20 MHz, along with circuit waveforms. The RC phase lag of 29° (plus 180° for inversion) from point B to point C, caused by the collector's output resistance R_o and the effective value of C_1, can be seen in the waveform photographs. The waveforms also show the RLC phase lag of 151° caused by the crystal's internal impedance (resistive and inductive, above series resonance) combining with the capacitor C_2.

The Pierce Harmonic has the best short term frequency stability of the four VHF oscillator circuits, because the external load tied across the crystal is mostly capacitive rather than resistive, giving the crystal a high in-circuit Q.

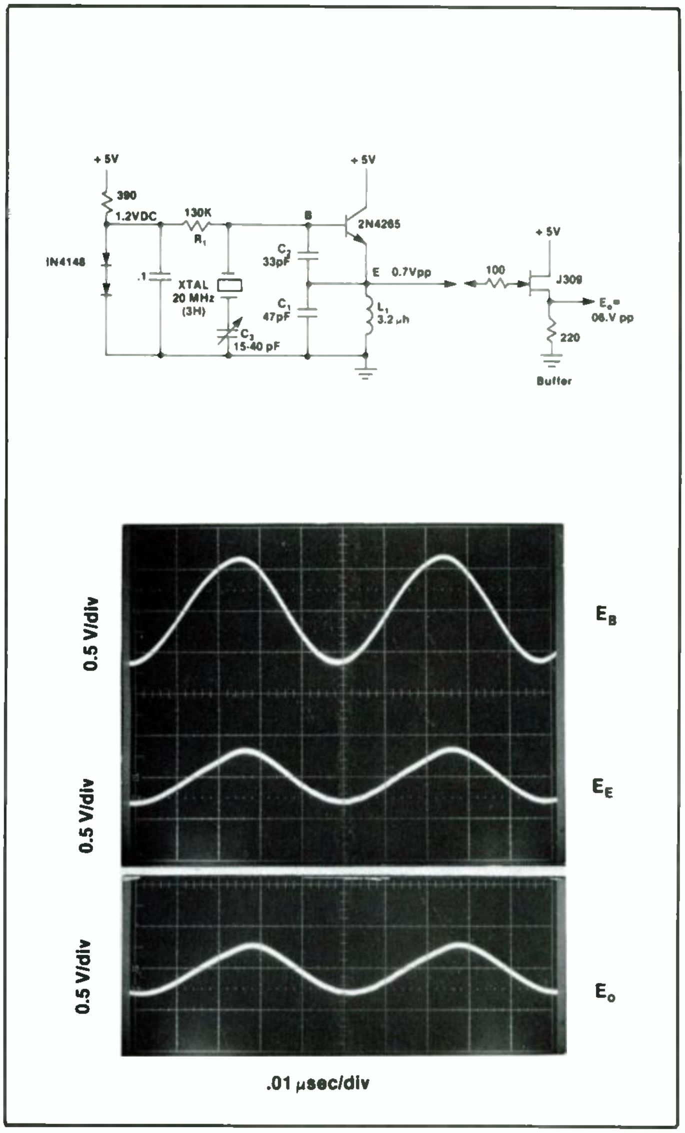


Figure 6. Colpitts Harmonic at 20 MHz.

Belcher's circuit arrangement, as shown in Figures 1d, 7 and 8, is an excellent blend of circuit simplicity and good frequency stability.

A drawback to the Pierce Harmonic circuits shown in Figures 7 and 8 is that the capacitance values of C_2, and particularly the effective value of C_1, are not as large as they might be for best operation. The maximum values of C_1 and C_2 are limited by the gain losses they introduce and by the transistor's gain.

The performance of the Pierce Harmonic, which is already very good, can be improved at the cost of increased circuit complexity[1]. The performance is improved by adding an emitter follower, as shown in the simplified circuit of Figure 9. Transistor Q_1's gain is increased by using a resonant L_1C_3 tank for the collector load, tuned to the oscillation frequency. The increased gain permits using larger shunt capacitance for C_1 and C_2, reducing their frequency sensitivity and increasing the in-circuit Q and the short term frequency stability. The emitter follower Q_2 raises the input impedance of the network surrounding the crystal, so as not to load the resonant L_1C_3 tank and reduce the gain of Q_1. The crystal is also driven at a lower excitation level, improving its long term stability. And by using the higher gain transistor MTM3960 (or MMCM3960) for both Q_1 and Q_2, one emitter follower should be sufficient rather than the two described in Ref.[1].

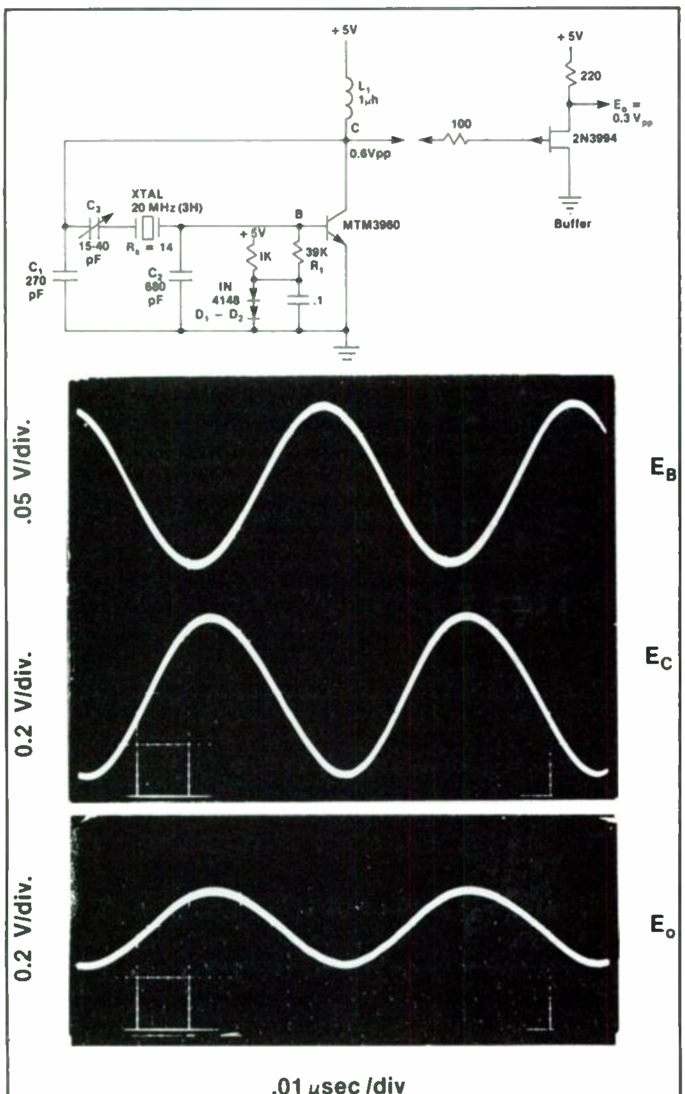


Figure 8. Pierce Harmonic at 20 MHz.

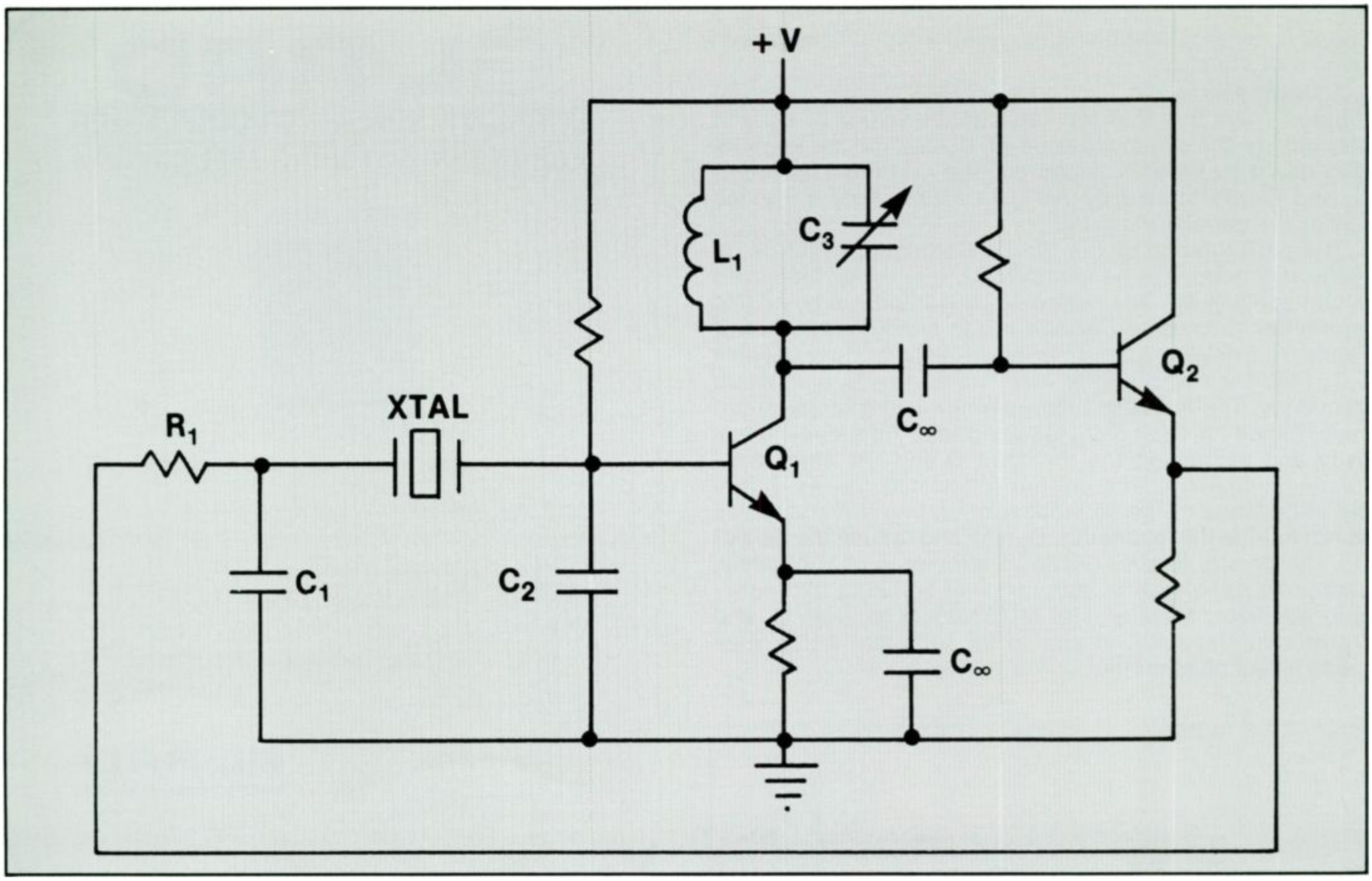
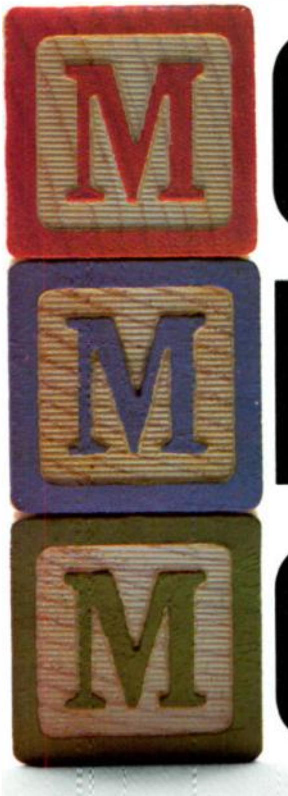


Figure 9. Higher Performance Version of Pierce Harmonic Circuit.



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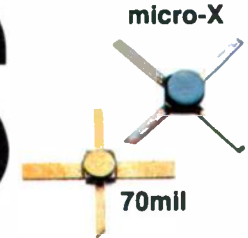

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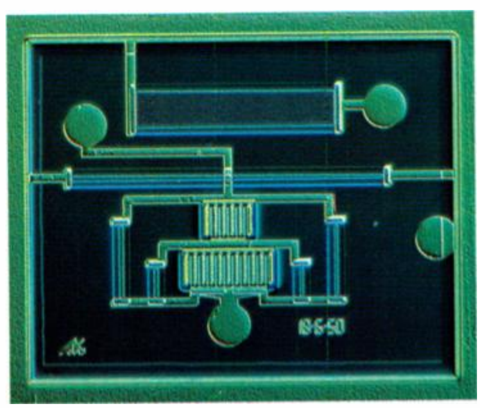
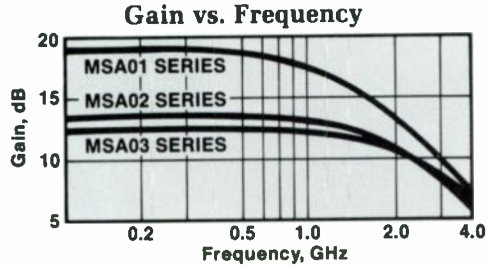


Gain vs. Frequency



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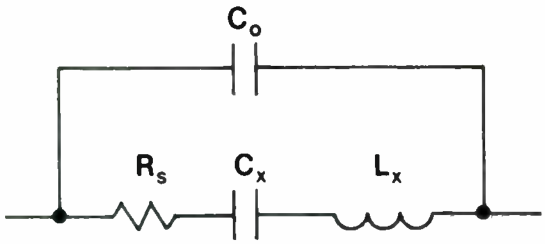
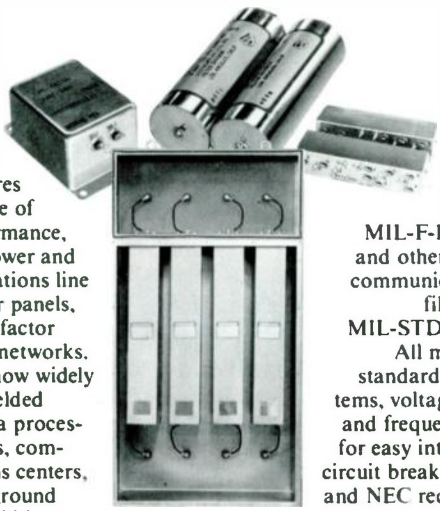


Figure 10. Equivalent Circuit for a Quartz Crystal.

Quartz crystals are operated harmonically at VHF, because above a fundamental frequency of 20-25 MHz* the crystal becomes too thin and delicate to be handled. Crystals oscillate on their odd harmonics, with the upper limit (usually the ninth harmonic) determined by the ability of the circuit's resonant LC tank to distinguish between adjacent odd harmonics. The crystal's equivalent electrical circuit is shown in Figure 10, where C_x and L_x represent the crystal's motional capacitance and inductance, R_s represents crystal's internal series resistance, and C_o is the shunt capacitance between the crystal's electrode terminals.

The crystal's internal series resistance R_s varies from 15-200 Ω over a harmonic frequency range of 20-200 MHz, respectively. The crystal's internal series resistance R_s is one of its most important parameters. The oscillator circuit is normally designed to drive a crystal load equal to R_s, and when R_s varies over a 13 to 1 range, the circuit design has to change to accommodate this.

The crystal's shunt terminal capacitance C_o (about 5 pF) is treated as part of the external load on the crystal. The crystal's external load affects both the in-circuit Q and the oscillation frequency. For maximum short term frequency stability, the crystal's in-circuit Q should be as high as possible, which means that the crystal's external load should have the lowest possible series resistance (or the highest possible parallel shunt resistance). Using a reactive load on the crystal (as in the Pierce circuit) is a very effective way to obtain a high in-circuit Q.



In general, a crystal oscillator's frequency can be trimmed by changing the circuit's phase shift at any point around the oscillator loop. One of the easiest ways is to insert a variable 15 to 40 pF capacitance in series with the crystal. This gives a 20-25 PPM (parts per million) frequency tuning range at frequencies below 40 MHz. The tuning range can be expanded by reducing the minimum series capacitance below 15 pF.

Above approximately 70 MHz, the capacitance values of C_1 and C_2 in the circuits drop to 30 pF or even less. The tuning range of a 15 to 40 pF capacitor in series with the crystal then drops to 2-5 PPM, which isn't enough. The frequency is now trimmed by directly varying either C_1 or C_2 in the circuit, rather than by varying a capacitor in series with the crystal. Varying C_1 or C_2 over a 10 or 20 pF range will give a 10-12 PPM tuning range (3 PPM for the Colpitts Harmonic) at these higher frequencies. The oscillator circuits shown here reflect this change in frequency trimming at the lower and higher oscillation frequencies. The tuning range of the improved Pierce Harmonic circuit in Figure 9, however, is 40 PPM at frequencies up to at least 100 MHz.

Performance

How good are these VHF oscillator circuits? Measurements of frequency stability and of frequency shifts due to temperature usually include the effects of both crystal and circuit, making it difficult to sort out those due to the circuit alone. The short term frequency stability of the circuits described here are at or better than the author's 0.1 PPM measurement capability. Indirect measurements indicate the Pierce Harmonic should have the best frequency stability, with Butler's Emitter Follower being second best. The Colpitts Harmonic and Butler's Common Base are tied for third place with about equal stability. Butler's Common Base also has the two undesirable characteristics mentioned earlier, so overall it is a less desirable circuit than the Colpitts Harmonic.

Applications

Which crystal oscillator circuit should you select for your application? If you need the best frequency stability, the Pierce Harmonic is the circuit to use. If you need a simple circuit that can be designed easily and quickly, then choose the Colpitts Harmonic. Or if you need a circuit with good frequency stability, a low output impedance, or one that will operate above 200 MHz, then Butler's Emitter Follower circuit is the one to use. All three circuits are free of parasitic oscillations.

For oscillator frequencies between 20-100 MHz, the circuit inductance and capacitance values can be interpolated between those given in the schematics. For frequencies above 100 MHz, the circuit inductance values should be reduced in approximately inverse proportion to frequency.

* Fundamental crystals up to 500 MHz have been reported recently[4], along with harmonic operation up to 1.5 GHz.

References

1. R.J. Matthys, *Crystal Oscillator Circuits*, Wiley Inter-Science, 1983.
2. D.K. Belcher, "A High Stability VHF Oscillator", *r.f. design*, Vol. 6, No. 1, pp. 26-36, Jan/Feb 1983.
3. M.E. Frerking, "Crystal Oscillator Design and Temperature Compensation", Van Nostrand Reinhold, 1978.
4. L. Bidart, "New Design of Very High Frequency Sources," Ferroelectrics, Vol. 40, pp. 231-236, 1982.

"Alpha Industries' New Aid

By Christine Peterson
Alpha Industries
Woburn, MA

Schottky diodes are monolithic devices used as mixers, detectors, low-level switches, and low-level limiters in a variety of applications ranging from a few megahertz (MHz) to hundreds of gigahertz (GHz).

Most applications up to 4 GHz require that the devices are supplied in standard microwave packages (see Figure 1) for convenient insertion into the circuit.

Fabrication

The use of monolithic fabrication techniques developed by Alpha Industries ensures the close matching of electrical characteristics among Schottky diodes in any arrangement, improving circuit performance. Less space is needed for each device and parasitics are reduced because the connections between the individual diodes are much shorter. Enabling the operator to transfer and position the entire diode arrangement as a unit rather than handling each device separately saves labor and gives higher yields because of less breakage.

The number of bonding operations required can be reduced from 25% to over 60% depending on the number of diode elements in the array, resulting in lower costs because of the labor hours saved. When the devices are packaged, cost savings are reflected in a lower price per unit.

Prior to the development of early monolithic technology, arrangements of multiple diodes necessitated the bonding of individual devices into the desired pattern. In the 1970s, the simple two-diode series pair and four-diode ring were developed. More recently, the variety of designs available has expanded to include four-diode bridges, eight-diode and twelve-diode rings, and the new types of diode designs pictured on the cover.

Each of the monolithic multiple Schottky diodes is designed for a specific application. The STAR Quad is used in a star mixer circuit, where four separate diodes were previously required. Excellent matching among the diodes is important in this type of design to ensure good isolation at the IF output port (see Figure 2). The Split Pair is ideal for DC-biased detector circuits in which one diode is in an RF circuit and the other is in a temperature compensation circuit. In this situation, close matching mini-

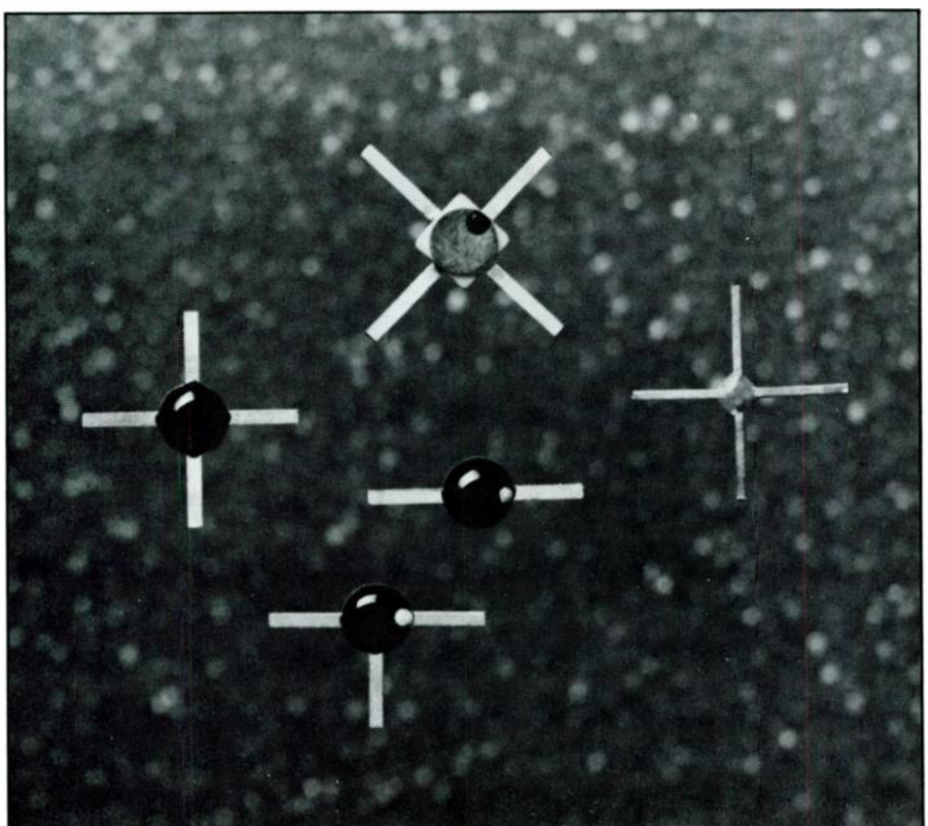


Figure 1.

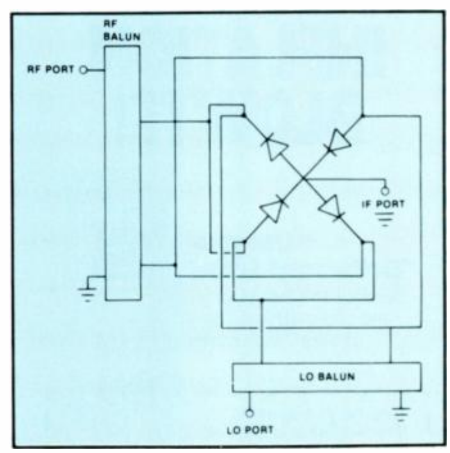


Figure 2. Star Mixer Circuit.

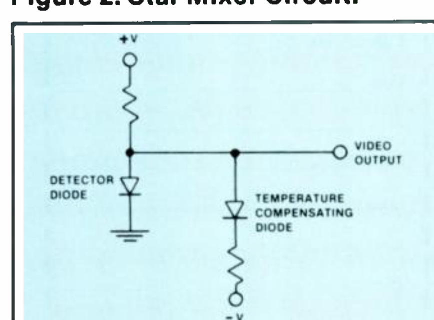


Figure 3. Temperature Compensated Detector Circuit.

Monolithic Devices RF Design"

mizes the change in bias voltage with temperature (see Figure 3). The Anti-parallel Pair offers significant noise figure improvement over a single diode in a subharmonically-pumped mixer, without the complexity and cost of a balanced mixer. Additional advantages of this device include the suppression of local oscillator odd harmonics and the elimination of any DC component at the IF terminal. Specifically designed to operate at a +20 dBm LO drive level, the Four-Junction Diode Pairs are used for many high local oscillator power level applications. The Common Cathode Pairs are particularly useful in video detector circuits involving signal comparison where close matching gives improved differential video output.

Early Schottky diodes were made in chip form only, with contacts on the top and bottom surfaces. Later Schottky diodes, called "beam lead" devices, were developed fabricating thick gold leads as part of the diode during the manufacturing process. In monolithic technology, the leads are formed so that they connect multiple silicon chips together to make up a simple diode "circuit." Virtually any arrangement that a designer can draw using diode symbols can be implemented with Alpha's flexible beam lead technology.

Characteristics

Schottky diodes are designed with various turn-on voltages, or "barriers," as shown in Figures 4a and 4b. These different characteristics are the result of varying the Schottky barrier metallization layer, which produces high, medium, low and very low or "zero bias" barriers as required for a given application. Very high barrier effects are produced by fabricating a number of high barrier diodes in series. An additional parameter to be optimized is the junction capacitance that is set according to the use frequency. By varying the barrier level, capacitance, and arrangement of diodes, Alpha produces a wide variety of custom-standard Schottky beam lead devices. New devices can be made in virtually any arrangement required by a circuit. The arrangements pictured were produced according to customer needs. Future designs will also be based on customer requests. □

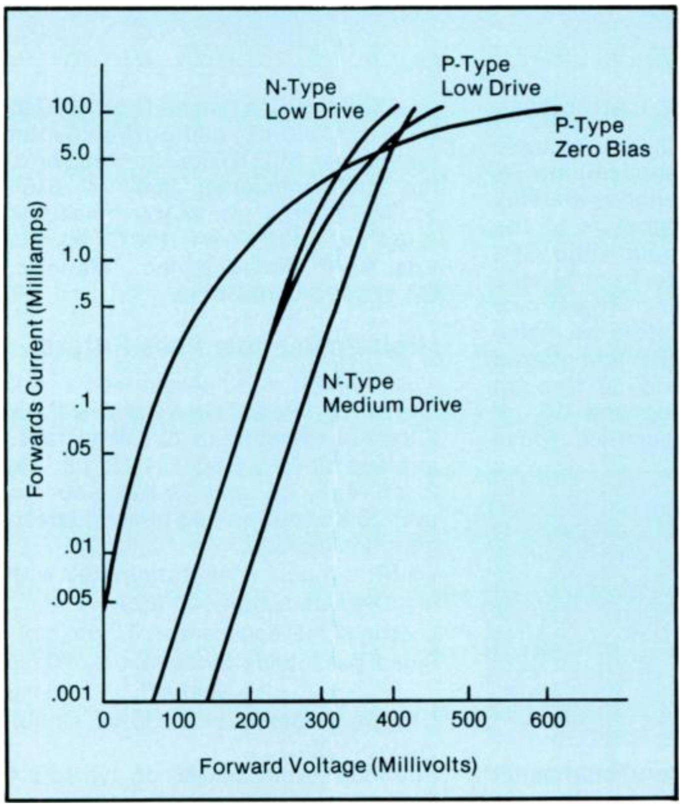


Figure 4a. Typical Forward DC Characteristic Curves — Voltage vs. Current.

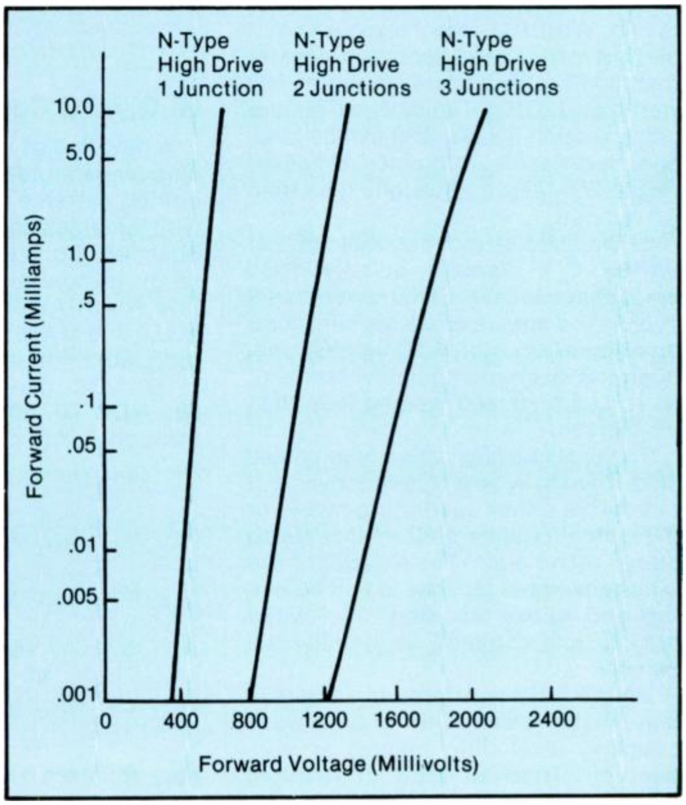


Figure 4b. Typical Forward DC Characteristic Curves — Voltage vs. Current.

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Flatpack 3dB hybrids and couplers have been limited in use in some applications because of low directivity and low power capability. To overcome these limitations Sage Laboratories is introducing an extention of their Wireline® technique in the form of WIREPAC™.

WIREPAC™ is the name of a new family of quadrature hybrids and couplers based on the Wireline concept. WIREPAC™ hybrids are recommended for operation over octave bandwidths in the frequency range of 150 MHz to 2 GHz. WIREPAC™ couplers are recommended for narrower band operation over a much wider frequency range, depending on the coupling value.

WIREPAC™ comes in two cross sections, either .25 inches square or .25 inches diameter round.

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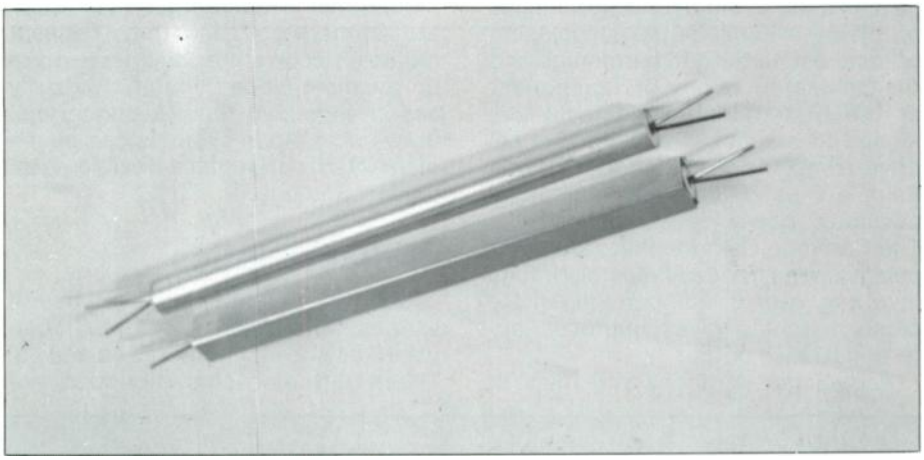
When used as a directional coupler WIREPAC™ displays equally good performance over a narrower band. Typically, a square cross section, 20dB directional coupler, 1.30 inches long, operates over a frequency range of 88 to 108 MHz and weighs less than 0.5 ounces.

The principle application is in printed circuit work, where it is inconvenient to achieve either quadrature hybrid or directional coupler performance using planar techniques. The exterior of the unit is tin plated for ease in soft soldering and epoxy bonding. The wires are cut and trimmed to simplify assembly.

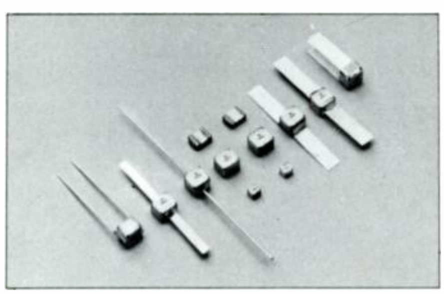
WIREPAC™ is offered in an octave bandwidth version with quarter-wave coupling (excluding losses) at midband of 2.70dB ± .15dB. In addition, a narrow band version is available for frequency bandwidths less than 30% with midband coupling (excluding losses) of 3.0dB ± .15dB. The modules are supplied cut to length. The length of a hybrid in inches is determined by dividing 1.97 by the center frequency in GHz.

Calculations of WIREPAC™ for directional coupler applications is more complicated and is described in an application note available from Sage Laboratories. To illustrate the versatility of WIREPAC™, a 4 inch length becomes a 10 dB directional coupler at 100 MHz. **Sage Laboratories, Natick, MA 01760, or INFO/CARD #117.**



HI-Q, NPO Ceramic Capacitors

A new line of high Q, NPO ceramic chip capacitors for applications requiring extreme capacitance stability with changes in temperature at frequencies from HF through the GHz range is now available from Murata Erie North America, Inc. These new MA50 and MA60 capacitors are manufactured to MIL-C-55681B and offered in 50 VDC (MA50) and 50 through 500 VDC (MA60) rating and 0.3 pF through 2,200 pF capacities. These



new units are available in chip pellet and leaded configurations including nickel barrier terminations. Additional specifications include a T.C. (NPO) of 0, ± 30 ppm/°C, a typical Q of 1,000 at 100 MHz (100 pF) and environmental testing per MIL STD's. In chip form, the MA50 measures .055" × .055" × .055" while the MA60 measures .110" × .110" × .100". **Murata Erie North America, Inc., Marietta, GA or INFO/CARD #113.**

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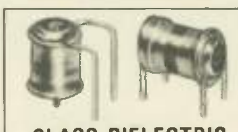

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Hybrid Crystal Oscillators

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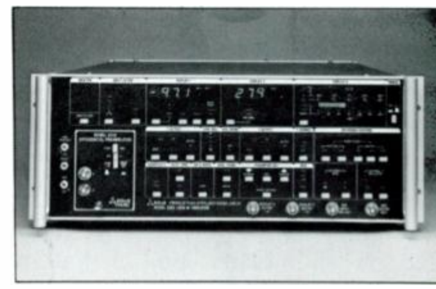

directly or via computer interface, and automatic modes are available for tuning, phase adjustment, and range setting. In addition to the standard exponential averaging technique, the Model 5301 offers a linear averaging mode for truer reproduction of signals which change rapidly with time. Other special features include plug-in preamplifiers, a multi-function signal channel for optimal noise rejection, in-phase and quadrature signal demodulation and an 80 dB dynamic reserve. The price for the main frame is $13,000. The Model 5316 High/Low Impedance Preamp and the Model 5317 High Impedance Preamp are priced at $750 and $500 respectively. The system is available for 45-day delivery. **EG & G Princeton Applied Research, Princeton, NJ 08540 or INFO/CARD #121.**

NPR Optimized IF Receiver Filters Sub LC and Crystal

Comstron Corporation has announced the FA-3900 Series of ultra-high performance FDM IF Filters optimized for high noise power ratio (NPR) receiver performance. The optimal design of this series is based on minimizing selected coefficients of power series expansions of the attenuation and group delay response as described in Garrison's classic paper on communication technology, (April 1968 PP 289-303). Data supplied with each filter includes: Passband

WHAT CTC CUSTOMERS SHOULD KNOW ABOUT TODAY'S ACRIAN

AN OPEN LETTER FROM
J. R. HARRIS, PRESIDENT,
ACRIAN INC.

On October 31, 1982, ACRIAN Inc. and CTC (Communications Transistor Corporation) became one company.

Some things have stayed the same. Some things are changing.

What's stayed the same is our total commitment to the CTC product line—and to CTC customers. ACRIAN has the capability to satisfy the complete range of RF and microwave power needs—from HF to 6 GHz. And you can rely on something else: ACRIAN commits to on-time delivery.

What's changing? The answer can be summed up in one word: *performance.* We're proud of our track record. But we're still not satisfied. Aspiring to higher performance and then making it happen is a watchword at ACRIAN. In consultation with our customers, ACRIAN is working toward a common purpose—upgrading existing products with higher technology—giving you more power, at less cost, in a smaller package.

At today's ACRIAN, we're supporting systems in telecommunications, radar, and avionics—with a level of integration that provides you with power, whatever your purpose.

ACRIAN

POWER WITH A PURPOSE

10060 Bubb Road
Cupertino, CA 95014
(408) 996-8522
Twx 910-338-2172

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INFO/CARD 73

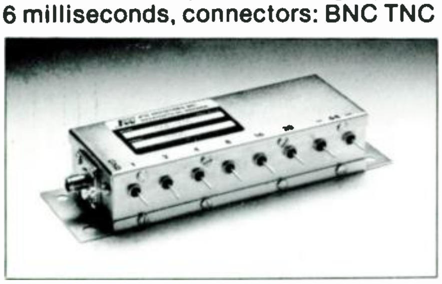
attenuation response; Passband group delay response; Stopband response; Coefficients of power series expansion of group delay and attenuation in passband; Intermodulation distortion at specified power level and frequencies. The FA-3900 Series provides uncompromised system performance when NPR is a primary consideration. Standard IF frequencies of 1.5, 10, 7, and 21.4 MHz are available with bandwidths of 1.5 Hz to 4 MHz. **Comstron Corporation, Freeport, N.Y. 11520, or INFO/CARD #84.**

Ferropac™ Isolators

M/A-COM Microwave Components, Inc. announces a new series of miniature Ferropac™ isolators for TACAN and IFF applications. Specially designed connectors and circuitry allow for miniaturization (1" × 1" × 3/8") without degradation of RF, temperature or power performance. Spanning the frequency range 960 MHz to 1215 MHz, the 5020 series offers an isolation of 18 dB, an insertion loss of 0.45 dB and a VSWR of 1.3:1. These Ferropac™ isolators able to withstand a mechanical shock of 20,000 G. They are also able to perform in temperatures ranging from −54°C to +95°C, and incorporate SMA female connectors. For the IFF frequency range, the 5020 series exhibits an isolation of 20 dB minimum, an insertion loss of 0.4 dB maximum and a VSWR of 1.25:1 maximum. Write for bulletin number 1152. **M/A-COM Microwave Components Inc., Burlington, MA 01803, or INFO/CARD #123.**

Programmable Attenuator

DC-1000 MHz, model 50 P-076, 0-127 dB in 1 dB steps, switching speed: 6 milliseconds, connectors: BNC TNC



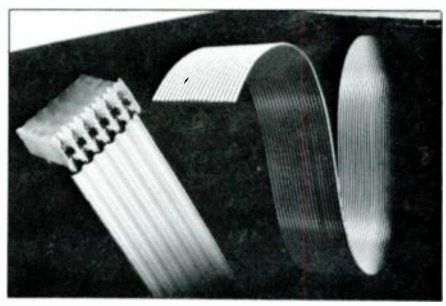
N or SMA. Delivery: 2-4 Weeks. JFW Attenuators are available 75 or 50 Ohm Impedance, price $360.00 (1-9 pcs.) **JFW Industries, Inc., Indianapolis, IN 46203 or INFO/CARD #124.**

Pre-test Limiter Mods

SDI's 9000 series of pre-test limiter mods offer superior low loss performance from 0.5 to 18.0 GHz. These revolutionary, positive-mounting devices have been developed by SDI to allow easy and efficient electrical testing of RF limiter modules. Expensive fixturing can be eliminated by electrically testing all limiters prior to connector removal and insertion into a stripline or a micro strip medium. Eight standard circuit variations are available. Power handing is from 150 watts to 3000 watts. Specified leakage power is as low 10 mW. Low loss is guaranteed by an extensive PIN diode and internal Epitaxial Material Capability. **SDI, Incorporated, North Billerica, MA 01862 or INFO/CARD #125.**

Shielded Cable

Molex announces the development of conductive polymer shielded planar cable providing its signal conductors with individual and overall shielding from audio frequency EMI. The cable, X40163 on .100" (2,54 mm) center spacings and X8833 on .156" (3,96 mm) centers consists of polyethylene-insulated signal conductors separated



and shielded by a conductive polymer layer. The polymer adheres to an uninsulated drain conductor which provides a means for grounding the shield. A layer of PVC coats the outside of the product. This cable can be mass terminated in insulation displacement connectors. The .100" center product accommodates 28, 26 and 24 AWG and is available in 2 to 36 conductor positions. The .156" center product accepts 26, 24 and 22 AWG and can be supplied with 2 to 24 conductors. Signal and drain wire locations are customer specified. Optional Teflon® insulation over the signal conductors is available. These products are targeted for audio applications, audio-level telecommunications, computers, business machines and low speed digital applications. **Molex Incorporated, Lisle, IL 60532 or please circle INFO/CARD #126.**

New Synthesizer

Designed especially for phase-coherent radars, FSK and frequency agile communications systems, the new Wavetek Rockland Model 5130A Frequency Synthesizer achieves spurious-free switching and true phase-continuity through patented direct digital frequency synthesis. It covers

100KHz to 160MHz with steps as fine as .001 Hz through program control or .1Hz manually. Phase continuous switching in 3 microseconds over .001 Hz to 99 KHz increments makes possible programming of smooth step sweeps, ramps, chirps or arbitrary frequency patterns without spurious between steps. Switching from any frequency to any other over the full 160 MHz range can be achieved in no more than 20 microseconds with amplitude changes in 50 microseconds. For use in automatic systems the Model 5130A, and its simpler companion, Model 5120A (without phase-continuous switching), are provided with remote control through a fast BCD interface or an optional IEEE-488. Special care in design realizes exceptional reliability for these computer-controlled systems with MTBF typically 13,000 hours. The signals from the 5120A and 5130A are exceptionally pure. SSB phase noise at only 100Hz from the carrier is a low −100dBc/Hz and amplitude noise less than −80 dBc. A high stability $\pm 4 \times 10^{-9}$/day reference of TCXO reference is optional. An external reference input or the synthesizer's reference output can be used to lock the unit and the test sytem together. Price: 5120A-$6475, 5130A-$8350, delivery: 30 days. **Wavetek Rockland, Inc., Rockleigh, NJ 07647 or please circle INFO/CARD #127**

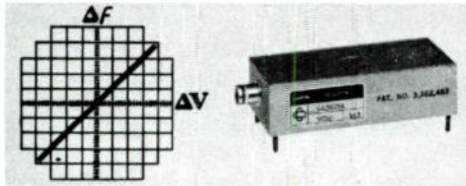
Digitally Controlled Varactor Filter

K & L Microwave is pleased to announce a breakthrough in voltage tuned filters. The new product series, designated VTB and VTR, are available in both bandpass and bandreject configurations respectively. In addition to being voltage tuned, the filters utilize microprocessor/EPROM technology to provide a variety of options to the user. These voltage tuned filters can be controlled by such devices as communications or test equipment, computers or various other control panels that put out various forms of digital logic. Logic which can be utilized includes serial in the form of RS232 or current loops or parallel in the form of BCD and GPIB (IEEE-488). The devices can be specified in tuning ranges up to an octave anywhere in the frequency

DAMON THE FORM FACTOR SPECIALISTS

VCXOs — 1 KHz to 500 MHz

Damon VCXOs provide high stability over a broad range of center frequencies. An optional square wave or pulse output signal compatible with I. C. logic circuits is available.

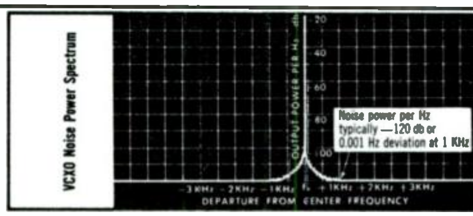


VCXO CHARACTERISTICS:

Center Frequency .. 1 KHz to 500 MHz
Deviation .... .01% to 0.5% of Center Frequency
Linearity 1% to 5% of Peak Deviation
Frequency Stability .... Determined by Peak Deviation and Operating Temp.
Modulation Rates .... DC to 2 MHz

Low Noise VCXOs

Damon low noise miniaturized VCXOs are all solid state quartz crystal stable oscillators whose frequencies can be varied linearly by an external DC or AC modulating signal. Only a source of power and control signal are required . . . no auxiliary compensating circuitry.

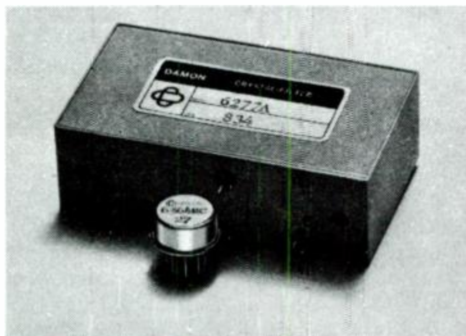


TYPICAL CHARACTERISTICS:

Center Frequency 10.000 MHz ±1 KHz
Deviation .... ±10 KHz
Linearity .... ±1%
Modulation Rate .... Flat to 4 KHz
Output Power .... >1 MW

Gaussian Crystal Filters

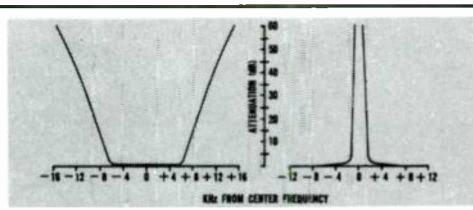
Gaussian Crystal Filters of either conventional or monolithic design. Useful in FM communications or radar systems where flat time delay is required. In a coherent pulsed radar, it can minimize return-pulse overshoot. Eliminates ringing problems in swept spectrum analysis. Available in 2 to 8 pole designs.



Center Frequency
Conventional .... 10 KHz - 35 MHz
Monolithic .... 3 - 30 MHz
3 dB Bandwidth .... 100 Hz - 30 KHz
Shape Factor, 30 dB to 3 dB 3:1 to 7:1
Ringing, Pulse or Impulse 25 - 50 dB down
Phase Linearity .... ±2%
No. of Poles
Conventional .... 2 - 8
Monolithic .... 2 - 4
Volume
Conventional .... 1 - 6 cu. in.
Monolithic .... 0.016 - 0.4 cu. in.

Band-Pass/Reject Crystal Filters

A wide variety of conventional crystal filters are produced at Damon. Standards designs such as Butterworth, Chebyshev, Gaussian, Bessel, and Legendre, as well as custom tailored designs are available.



RANGE OF PARAMETERS:

Center Frequency .... 10 KHz - 75 MHz
Bandwidth .... .01% - 3% of C.F.
Shape Factor .... Down to 1.2:1
Differential Phase Shift .... ±2°
Group Delay Uniformity .... ±5%

Surface-Wave Delay Lines

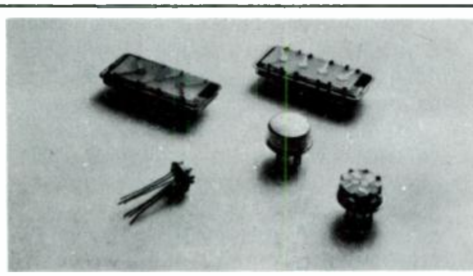
With signal transfer across the surface instead of through the body (a state-of-the-art development), the new Surface-Wave concept results in substantially smaller packages, lower cost in production quantities and superior performance over a 20 MHz to 300 MHz range.

RANGE OF PARAMETERS:

Frequency Range .. 20 MHz - 300 MHz
Time Delay Range 0.5 - 20 Microsecs.
Bandwidth at −3 dB .... 3% - 30% of Center Frequency
Insertion Loss .... 10 - 20 dB
Spurious Responses 25 - 30 dB below Desired Response
Environment .... Full Military Temp., Shock, Vibration Available
Typical Size .... < 1 cu. in.

Monolithic Crystal Filters

Damon Monolithic Filters utilize a single quartz plate for multiple resonators and the coupling medium between them. Advantages over conventional crystal filters include smaller size, superior stability, and lower cost. Available in a wide variety of computer-assisted designs from Chebyshev to Gaussian characteristics and from 2 to 8 poles.



MONOLITHIC CHARACTERISTICS:

Center Frequency .... 2.5 - 30 MHz
3 dB Bandwidth .... 0.3 - 30 KHz
No. of Coupled Resonators .... 2 - 6
Insertion Loss .... 1 - 6 dB
Ripple .... 0.5 dB Max.
Terminations .. 50 Ohms to 20 Kohms
Size .... TC-5, TO-8, or Flat-Pack

Write for technical bulletins.

DAMON/ELECTRONICS DIVISION

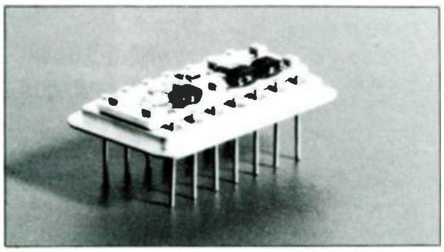
80 WILSON WAY, DEPT. G, WESTWOOD, MASS. 02090, TEL: (617) 449-0800

INFO/CARD 7

spectrum from 100 MHz to 6 GHz. Bandwidths can be as narrow as two percent to as great as 10 percent of the tuned center frequency. The devices are ideally suited for receive applications. The filters can be constructed with two through five resonant sections and can be manufactured to full military environmental requirements. Availability is from 6 weeks. **K & L Microwave, Inc., Salisbury, MD 21801, or INFO/CARD #21.**

New Switch Line

VARI-L Company, Inc. introduces a new RF, Solid State Switch line covering a frequency range of 3-2000 MHz. Devices are available with TTL and CMOS Compatible integral drivers.



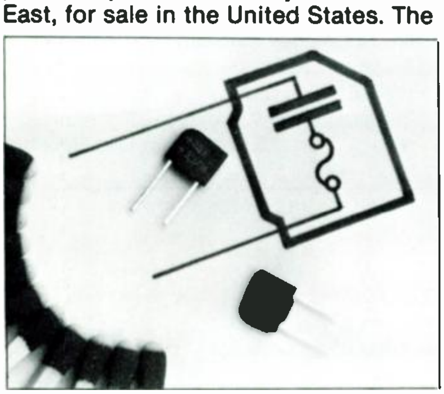
Package styles include TO-5, TO-8, flatpack and DIP platform configurations. Qualified to meet MIL-STD-883B, Methods 5004 and 5008, Class 8. **VARI-L Company, Inc., Denver, CO, 80239, or INFO/CARD #112.**

High Power Porcelain Capacitor

Monolithic, High Q, High Power, High Current Porcelain Capacitor known as Dielectric Laboratories, Inc. style 21HAH for use in RF power amplifiers, I.E., 200pf at a frequency of 100 MHz will handle six amperes of changing current without overheating. Many capacitance values are available from stock. **Dielectric Laboratories Inc., Cazenovia, NY 13035 or please circle INFO/CARD #114.**

Hybrid Capacitor-Fuse

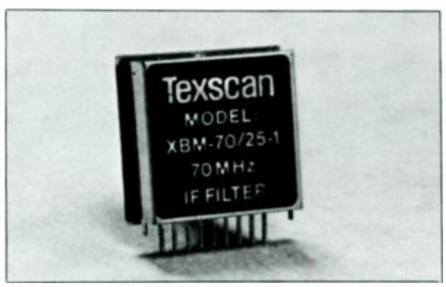
Matsuo Electronics of America, Inc. has announced the availability of its Type 245 solid tantalum capacitor-fuse, previously available only in the Far East, for sale in the United States. The



hybrid device features built-in circuit protection. An incorporated fuse prevents catastrophic failure of other expensive board components and adds a significant fire prevention safety factor. The component provides protection against total equipment destruction and satisfies "product liability" concerns. The fuse in the capacitor is designed to open at currents greater than 5 Amps. When current is less than 5 Amps, the device acts as a thermo-fuse to prevent element burning or resin decomposition. The Type 245 can be used as a fast and inexpensive method of adding circuit protection to an existing design by replacing an existing capacitor without board modification. Pricing for the Matsuo Electronics Type 245 capacitors is 39¢ in OEM quantities. **Matsuo Electronics of America, Inc., Garden Grove, CA 92641, or please circle INFO/CARD #111.**

70 MHz I.F. Bandpass Filter

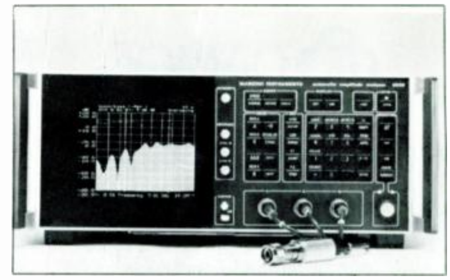
The Texscan XBM 70 series of bandpass filters is intended for use as a low cost, single piece I.F. filter in satellite receivers. It features small size, rugged construction and low insertion loss. The package is suitable for wave soldering direct into the master printed circuit board and is



supplied pre-aligned to the correct frequency and bandpass characteristics. **Texscan Instruments, Indianapolis, IN or INFO/CARD #110.**

Scalar Analyzer

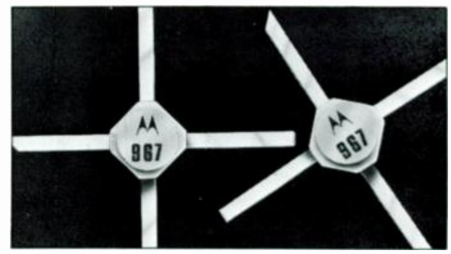
Marconi announces a new Automatic Amplitude Analyzer, Model 6500, designed to provide a flexible solution to the problems of testing Microwave systems and components. Unique on-screen graphics bring alive the displayed signal. Amplitude and frequency scaling can be automatically arranged for optimum display and a powerful bright line cursor shows frequency and amplitude across the band for ease of use and accurate measurements. Model 6500 has been designed to control any sweeper or RF source which may be driven externally by a voltage ramp eliminating the need to buy special RF sources. With its 10 MHz to 18 GHz detectors, it mea-



sures and displays absolute power, transmission and return loss and for the first time in a Scalar Analyzer, VSWR. Model 6500 priced at $9,325 is available in 90 days. **Marconi Instruments, Northvale, N.J. or please circle INFO/CARD #109.**

1 GHz Gallium Arsenide FETS

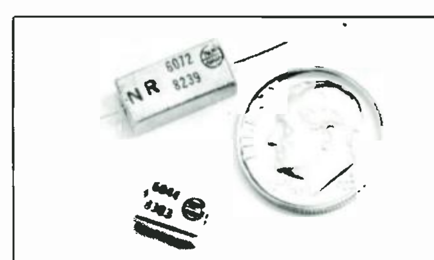
Motorola, which has been sampling the industry with state-of-the-art UHF GaAs FETs, is going into full scale production and is prepared to deliver



these in quantity. The new Motorola FETs offer 1 GHz performance, with a typical gain of 18 dB and noise figure of 1.2 dB (typ). This will significantly improve UHF and 800 MHz radio reception and high-frequency performance of electronic instruments. Available from warehouse stock, the transistors are being introduced under the part numbers MRF966 for plastic packaged devices, and MRF967 for ceramic packaged components. Prices for these parts are $3.10 and $9.85, respectively, in quantities of 100 to 999, but production quantities will be available at substantially lower costs. **Motorola Semiconductor Products, Inc., Phoenix, AZ 85036 or please circle INFO/CARD #108.**

Drop-In Switch Modules

New England Microwave Corp. is now offering a complete line of microstrip SPST drop-in modules covering the 10 MHz to 18 GHz frequency



Consider the source.

HP's 8656A programmable signal generator

Consider the value.

Versatility.

The HP 8656A synthesized RF signal generator is fully HP-IB programmable. Even your most time-consuming tests can be readily automated, increasing productivity in the laboratory and throughput on the production test floor.

Performance.

With a frequency range of 100kHz to 990MHz, the 8656A's signals are accurate and stable with good spectral purity. You get 100 or 250Hz resolution, an output range of +13 to −127dBm with microprocessor-controlled unit conversion, ±1.5dB absolute accuracy and 0.1dB resolution. Low RFI leakage provides confidence in micro-volt level signals. AM and FM are available from both internal and external sources and simultaneous modulation modes are easily generated. All this and reverse power protection and HP-IB are standard.

Quality.

The 8656A incorporates HP's 30 years of experience in designing and manufacturing high performance signal generators. Our reputation for designed-in quality and reliability is backed world-wide by a responsive sales and service network.

Economy.

At $7150*, the 8656A is HP's lowest-priced programmable signal generator. It's a price-performance value that makes good economic sense for both laboratory and production environments in a broad range of applications.

DESIGNED FOR HP-IB SYSTEMS

Consider the exceptional value of the 8656A. You may find that it's just the right source for you. To put the 8656A to work in your particular application, or for help in choosing from our broad line of signal generators, contact your nearby HP sales office. Or write, Hewlett-Packard, 1820 Embarcadero Road, Palo Alto, CA 94303.

Consider HP.

*U.S. domestic prices only.

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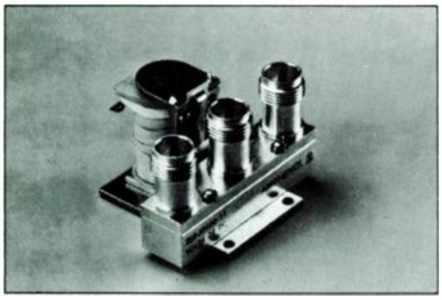
range. Included in the new product line are the Models 6044 and 6072. Laser-welded for hermeticity and screened to the requirements of MIL-STD-883, both models operate from .5 to 4 GHz with the 6044 providing 45 dB minimum isolation and the 6072 providing 60 dB minimum isolation. Availability for these new modules is off-the-shelf to four weeks, depending upon quantity requirements. **New England Microwave Corporation, Hudson, New Hampshire 03051, or INFO/CARD #107.**

75 Ohm Video Switches

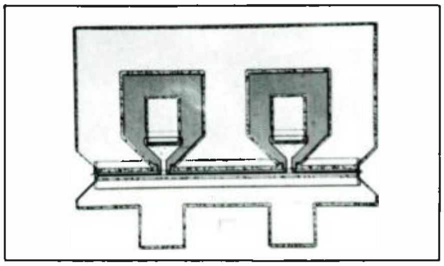
Switches designed for use in 75 ohm video circuitry are now available from Amphenol. Applications of the Amphenol® 318 Series include efficient signal routing in television broadcast and earth station/satellite equipment. The series comprises 1P2T shorting and non-shorting, remote-actuation failsafe video switches. RF performance figures for the non-



shorting units include insertion loss 0.2 dB maximum from 1 to 5 GHz. Isolation is rated at 35 dB at 5 GHz, 40 dB at 3GHz, 55 dB at 1 GHz and 60 dB at 0.4 GHz. Maximum VSWR values range from 1.15 at 0.4 GHz to 1.25 at 5 GHz. Switching time to operate is 15 milliseconds maximum, and release time is 10 milliseconds. The 318 Series uses Ampehnol connectors which mate with Ampehnol cable plug 86475 for RG-59, 62, 71, 140, 210 and 302/U cables. RF power ratings (non-switching) range from 100 watts CW at 400 MHz to 50 watts CW at 5 GHz. **Amphenol World Headquarters, Oak Brook, Illinois 60521 or INFO/CARD #106.**

GaAs FET Transistor

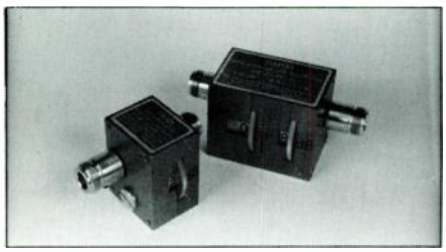
A new gallium arsenide, field-effect transistor for low-noise, high-gain operation through Ku-band has been introduced by Raytheon's Special Microwave Devices Operation. The Raytheon Type RLK048 is a drop-in replacement for the NE137 GaAs FET. Its 0.5 micron-long, recessed-aluminum gate results in an exceptionally-low noise figure. Operating at 8 GHz, the typical noise figure is 1.2 dB, while at 18 GHz it is 2.5 dB. Gain is 11.0 dB



typical at 8 GHz and 7.5 dB typical at 18 GHz. Applications for the RLK048 include low-noise amplifiers and oscillators; limiting amplifiers; mixer/down converters; and all FET front ends for TV satellites and CATV. The new device is available as a chip, in the 1.8 mm-square standard package, or on a carrier. Chips are available for immediate delivery from stock and are priced at $50 each in small quantities. **Raytheon Company, Northborough, MA 01532, or INFO/CARD #105.**

Step Attenuators

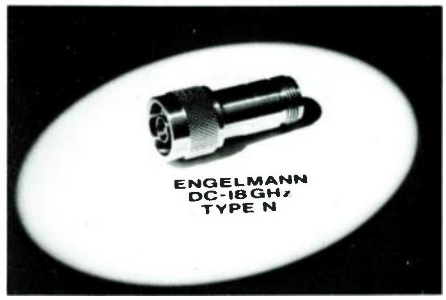
The Narda Microwave Corporation is offering a new series of precision step attenuators that cost less, last longer, and are smaller in size than any step attenuator previously available. A new concept in turret design combined with a proprietary contact design enabled Narda to produce this unique series. All models in the series operate from DC to 18 GHz and exhibit extremely flat frequency response. Three standard configurations are available today with two additional configurations* scheduled for availability in May 1983: Single Configur-



ations: 0-9 dB in 1 dB steps, 0-60 dB in 10 dB steps, 0-90 dB in 10 dB steps*. Dual Configurations: 0-69 dB in 1 dB steps, 0-99 dB in 1 dB steps*. All units are bidirectionsl with either type N or SMA female connectors provided on the standard units. Any combination of male and female connectors, including TNC and 7mm, are available on special order. All connectors comply with MIL-C-39012. **The Narda Microwave Corporation, Hauppauge, NY 11788 or INFO/CARD #104.**

DC-18 GHz Type N Attenuators

Englemann Microwave has announced availability of a new precision Type N Attenuator operating over the full frequency range from DC - 18 GHz. The A900N Attenuators have a maximum VSWR of 1.35 up to 18 GHz. Attenuation, variation and accuracy is better than ± .3 dB for value up to 10 dB, ±0.5 dB for values up to 20 dB, ±1 dB for values up to 35 dB over the full frequency range. Average power is rated at 2 watts CW and the units are specified to meet the requirements of MIL C39012 and other applicable environmental and MIL specifications. The A900 Series are priced



at $65.00 each for all standard values to 30 dB. Delivery stock to thirty days. **Englemann Microwave Co., Whippany, New Jersey, 07981, or INFO/CARD #103.**

Power Transistors

NEC's NEM0800 series of NPN epitaxial UHF power transistors is now available from California Eastern Laboratories, Inc. These transistors provide superior RF performance and are designed specifically for large volume mobile/cellular radio applications in the 800 and 900 MHz bands. The NEM0800 series provides high gain, high efficiency, and a high resistance to burnout with load mismatch. The NEM0800 series incorporates gold metallization for high reliability and is available in a low-cost metal-ceramic stripline package offering power levels of 6, 13, 24 and 40 watts. They are internally matched to simplify circuit design. These power transistors are complimentary to NEC's complete line of 800 to 960 MHz power amplifier modules offering total discrete-modular design flexibility. **California Eastern Laboratories, Inc., Santa Clara, CA 95050, or INFO/CARD #128.**

EAPROM Arbitrary Waveform Storage

An EAPROM (Electrically Alterable PROM) is now an available option on

Clean sweep to 1 GHz.

50 watts minimum is a lot of rf power throughout a bandwidth of 1 to 1000 MHz. But that's what our new Model 50W1000 delivers for all your broadband test needs.

As your hunger for power plus bandwidth grows, this year and next, our all-solid-state "W" series of 100-kHz-to-1000-MHz linear amplifiers should become more and more important in your plans. Today you may need only 1 watt (the little portable on the top of the pile), or 5, or 10—all with that fantastic bandwidth instantly available without tuning or bandswitching—the kind of bandwidth that lets you sweep clean through with no pausing for adjustment.

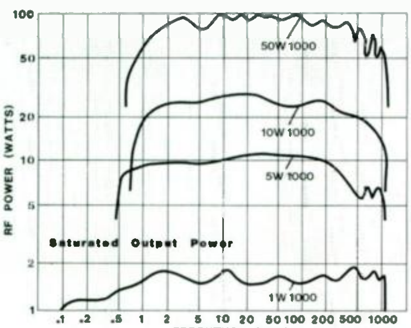
And next year?

But chances are good that next year you'll be moving up into higher-power work in the same bandwidth. Then you'll be glad you have a 50W, the *only* rf power amplifier in its power-to-bandwidth class. At that point, your smaller "W" series amplifiers can be freed for lower-power work around your lab.

What you can't see in the performance curves shown below is the unconditional *stability* of all Amplifier Research amplifiers—immunity to even the worst-case load mismatch or shorted or open cable with no fear of damage or system shutdown.

The "W" series is part of a complete line of amplifiers offering rf power up to 10,000 watts cw, for such diverse applications as plasma/fusion research, NMR, RFI susceptibility testing, and a host of other test situations that demand the very finest in broadband rf power.

Send for our free booklet, "Your guide to broadband power amplifiers."



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Wavetek reveals the best-kept price/performance secret in signal generator history:

Model 3010

Price: $5,295

Performance: 1 MHz to 1 GHz frequency range, 100 Hz resolution, complex or simultaneous modulation (AM on FM, FM on FM, or AM on AM), 0.001% accuracy (typical is ±0.0002%), programmable and agile (1 MHz step settling time less than 20 msec). Easy to use. Eight standard options available.

For the whole secret, phone toll-free 800-428-4424. Wavetek Indiana, Inc., 5808 Churchman, Beech Grove, IN 46107. Phone (317) 787-3332, TWX 810-341-3226.

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LITERATURE—INFO/CARD 75

WAVETEK®

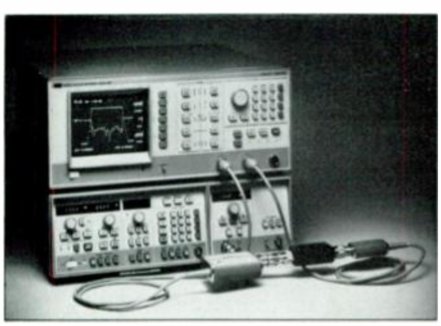
Krohn-Hite's Model 5910 0.0001Hz to 5MHz Arbitrary Waveform/Function Generator. This option now allows the storage and retrieval of the instrument's arbitrary waveforms, autoprogrammer programs and operating parameters in user transparent non-volatile memory. Up to 89 user programmed waveforms and programs may now be loaded into EAPROM manually via the front keyboard or the standard IEEE-488 bus. Delivery is 4 weeks ARO. The price of this option (option 003) is $400.00. The price of the 5910 is $4500.00. **Krohn-Hite Corporation, Avon, MA. INFO/CARD #129.**

Coaxial Crystal Detectors

The new Micronetics Coaxial Crystal Detectors cover a frequency range of 0.01 to 12.4 GHz and a frequency response within ± 0.5 dB absolute. Relative matching excluding bias sensitivity, is within ± 0.2 dB. Other specifications include: Output impedance; < 15 ohms shunted by 10 pf. Output polarity; Negative (positive optional). Connectors; RF Input, type N, male, VSWR; Values to 12.4 GHz maximum, Maximum power; 100 mW peak or coverage, **Micronetics, Inc., Norwood, NJ 07648, or INFO/CARD #116.**

Scaler Network Analyzer

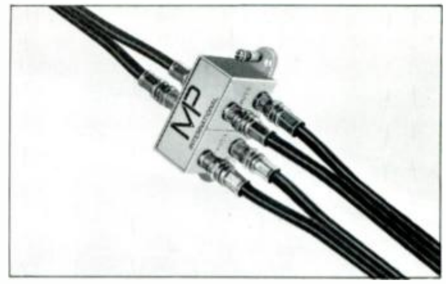
A new scalar network analyzer for microwave measurements from Hewlett-Packard brings operating features and measurement precision that the manufacturer believes will result in increased productivity in both manual and automated measurement applications. The microprocessor-managed HP 8756A scalar network analyzer is a two-channel measurement and display instrument that, combined with companion microwave bridges and detectors, measures transmission and return loss from 10 MHz to 26.5 GHz in coax, and to 40 GHz and higher in waveguide.



The HP 8756A uses microprocessor control to make operation simple, yet it gives the operator more flexibility in the measurement process than perhaps had ever been available before. Measurements and displays are digitally processed with CRT annotations to further assist in data analysis. Virtually every function of the analyzer can be run under program control via the Hewlett-Packard Interface Bus (HP-IB). **Hewlett-Packard Company, Palo Alto, CA 94303 or INFO/CARD #22.**

Dual Splitter

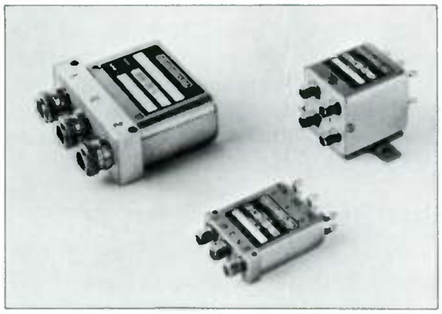
Multi Products International has introduced an innovative new Dual Cable Splitter design that features A/B Circuit Isolation greater than 110dB The Multi View MPI2202 provides an extremely fast and "foolproof" indoor/outdoor installation which eliminates



the crossing of cables. With fully machined non-tapered connectors, which are spaced to accommodate traps, the MPI2202 is ideal for all dual cable installations. All circuits are rated at 450 MHz and the housing provides a convenient grounding terminal. **Multi Products International, Cedar Grove, NJ 07009 or please circle INFO/CARD #26.**

Miniature Coaxial Transfer Switches

Model RST-M remote miniature coaxial transfer switches cover the DC - 18 GHz frequency range with an insertion loss of 0.5 dB max. SWR is 1.5 max, impedance, 50 ohms and isolation, 60 dB min. Swtiching time for the break before make switch is 15 ms. max. with 1M cycles min. life.



RF connectors are type SMA female. Weight: 6 oz. Price: $285.00. Delivery stock to 2 wks. **Micronetics, Inc., Norwood, NJ, or INFO/CARD #36.**

New TCXO Line

Dale Electronics, Inc. has just introduced the TCXO-32, the first in a new line of temperature compensated crystal oscillators with hermetically sealed metal cases. The TCXO-32 is adjustable with a multi-turn trimmer to an accuracy of ±0.1 PPM (at 25° C). Its stability vs. temperature rating is ±1 PPM, over the temperature range of −40°C to +85°C. The TCXO-32 is available over the frequency range of 3.0 MHz to 15.0 MHz, with 4.0, 5.0 and 10.0 MHz frequencies available from factory stock. The new

Dale oscillator is designed to drive 10 TTL loads. Its hermetically sealed metal case is 2" square and .53" high. Typical price of a TCXO-32 (10 MHz) in 25 quantity is $81.70 each **Dale Electronics, Inc., Tempe, AZ or INFO/CARD #96.**

High Power Attenuator

Weinschel Engineering is extending its high power attenuator line to 250 watts. Conservatively rated, the new model 45 can handle 250 watts CW or 10 KW peak bidirectionally from DC to 1.5 GHz. The model 45 is designed to meet high quality control standards and is able to withstand high levels of shock and vibration. A proprietary

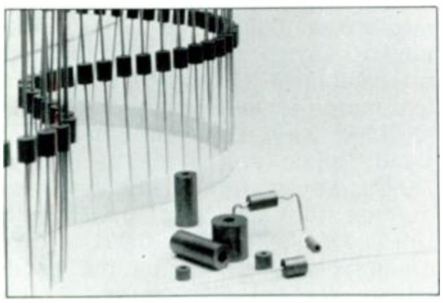
film process produces resistive elements with features such as low temperature coefficients, low aging rates and immunity to momentary overload. Use of troublesome, oil heat dissipative baths is also eliminated. Operating temperature range is from −55°C to +125°C. The model 45 is offered in a wide variety of attenuation values to satisfy both system and measurement applications. Prices: Model 45-3, 6, 10, 20 and 30: $435, delivery: 90 days ARO Model 45-40: $460. **Weinschel Engineering, Gaithersburg, MD 20877, or INFO/CARD #70.**

Flyweight Graphite Reflectors

Guaranteed precision microwave reflectors for higher frequency applications. The standard series of "Flyweight" reflectors have diameters from 10 inches to 6 feet, and each is a parabola of revolution with a F/D ratio of .33. Milliflect reflectors are presently in use at frequencies up to 140 GHz. **Milliflect, Newark, CA 94560 or INFO/CARD #55.**

Ferrite Beads

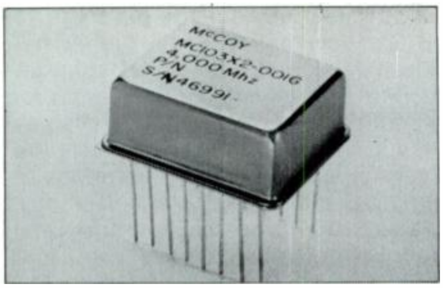
Stackpole's ferrite beads can inexpensively and conveniently prevent circuit components from receiving or transmitting electromagnetic interference (EMI). For shielding, RF decoupling and parasitic suppression applications, the soft-ferrite beads are placed near the noise source or device to be protected. When the magnetic field created by a noise transient passes into a bead, the ferrite's high losses at high frequency cause the



local impedance to rise rapidly. This chokes off unwanted current flow of the transient through the bead-shielded conductor. The beads, usually hollow cylinders which can be as small as .025" ID, .060" OD and .050" long, decouple spurious oscillations without adding another mechanical or electrical component to the system. Two grades of Ceramag® ferrites are widely used as beads. Ceramag 7D contains oxides of nickel-zinc and exhibits low initial permeability. Ceramag 24 contains oxides of manganese-zinc and exhibits high initial permeability. Three popular sizes are available on lead tape reel; others are in bulk pack. Prices range from $8 to $50 per thousand, depending on size, material requirements and quantity. **The Stackpole Corporation, St. Marys, PA 15857 or INFO/CARD #93.**

Thick Film Hybrid TCXO

Excellent frequency stability and small size are featured in this new temperature compensated crystal oscillator. Utilizing thick film hybrid techniques for both the oscillator and compensation circuits, ModelMC103X2 is smaller than other comparable discrete component devices. ±1 PPM



over 0 to +50°C and ±3 PPM over −30°C to +70°C are standard stabilities offered. Other specifications include Input Voltage +5 VDC ±2%, 15 ma maximum, TTL Output and an Output Frequency Range of 3.2 MHz to 25 MHz. The oscillators use a precision glass crystal for optimum aging performance. **McCoy Electronics Company, Mt. Holly Springs, PA 17065, or INFO/CARD #66.**

High Pass Filters

High pass filters that eliminate or greatly reduce interference on TV and FM receivers caused by amateur or CB transmitters are now available from the J.W. Miller Division of Bell Industries in Rancho Dominguez, California. Series C-513 high pass filters attenuate front end overload interference by a power factor greater than 1,000,000 to 1 when installed in the TV or FM receiver antenna input. The C-513-T2 model provides 75-ohm coax connectors and the C-513-T3 model provides 300-ohm twin lead connectors. **J.W. Miller Division, Rancho Dominguez, CA 90221 or INFO/CARD #122.**

Automatic RF/Microwave Test System

Watkins-Johnson Company has expanded its ADATE line of automatic test equipment with the introduction of the WJ-1550 RF/Microwave Test System. This automated, computer-intensive system for evaluation of RF parameters provides improved testing efficiency and significant cost savings for the user. Frequency coverage to 18 GHz is standard, with higher frequencies optional Menu-drive software facilitates efficient operator and programmer tasks. A touch-panel display or a graphics terminal provides the operator with a "friendly" interface to execute, edit or generate test programs. The graphics terminal and optional graphics printer present test results in an easy-to-understand format. Applications include production test, maintenance and support for all types of broadband and narrowband equipment, including receiver systems, transceivers, microwave relays, RF components and local-area network equipment. **Watkins-Johnson Company, San Jose, CA 95131 or please circle INFO/CARD #68.**

Ultra-Miniature Coaxial Conductor Cable and Connector Combinations

A new concept whereby cables have been designed in combination with connectors to achieve optimum miniaturization is announced by Microtech, Inc. In addition to ultra miniaturization, this design concept makes possible special built-in-features, heretofore unattainable, that assures the user of unbelievably simple, rapid and economical assembly of the cable to the connector without special tools and with inexperienced operators. The units are factory pre-assembled, with a minimum of loose parts so that the operator need only trim the cable to the proper length and screw it directly into the connector to complete a cable assembly in seconds. They close the gap for high density packaging and the extreme miniaturization required today in the computer, transducer, instrumentation, medical and IC fields. The line is precision built of the highest quality materials. The coaxial cables are made in either standard shielded or special low-noise treated types to prevent spurious output from vibration, shock or cable whip. The cable construction consists of teflon insulated solid AWG #30 wire, a silver plated copper shield and an extruded teflon outer jacket. The connectors utilize 10-32 and 4-48 threads for the A & C Series connectors respectively. Bodies, pins and sockets are gold plated brass, the dielectric inserts TFE teflon and the washers silicone rubber. Prices are exceptionally economical for the small quantity users considering no special tools are required and as low as $0.75 each for industrial and O.E.M. applications. Delivery is from stock. **Microtech Inc., Boothwyn, PA 19061 or INFO/CARD #119.**

New Literature

Microwave Component Catalog

Over 200 pages of in-depth information on the Sage passive component product line including rotary joints, switches, hybrids, couplers, phase shifters, power dividers, filters, attenuators, terminations, video detectors, diode holders, and WIRELINE® . Each section provides specifications, performance curves, schematic drawings, photographs, product features and special information on Sage's custom engineering capabilities in these product categories. **Sage Laboratories, Inc. Natick MA. INFO/CARD #140.**

Instrument Catalog

Krohn-Hite Corporation introduces their latest Short Form instrument catalog. Comparative specifications on all Krohn-Hite's — *Function Generators* from .0001HZ to 30 MHz with Lin/Log sweep, AM/FM, Marker, programmable GPIB compatable units and more. *Active Filters* from .001Hz to 3MHz, High, Low, and Band Pass, Band Reject and Tracking Filters. *RC Oscillators* with Sine/Square wave outputs from .001Hz to 10 MHz with Ultra-Low distortion to 0.001%. Programmable GPIB compatible oscillator units are also listed. *Digital Phasemeters* with up to 0.05 degree accuracy, at operating frequencies of 3Hz to 5MHz. *Distortion Analyzers* to measure total harmonic distortion down to 0.003%, and operating from 1Hz to 1MHz with auto-ranging and auto nulling, plus programmable GPIB compatibility built in. *Wideband Power Amplifiers* capable of handling DC to 1MHz and delivering a maximum of 75 watts output. **Krohn-Hite Corporation, Avon, MA or INFO/CARD #139.**

Microwave Component Catalog

Includes a broad line of waveguide switches, coaxial switches, dummy loads, bolometers and coaxial crystal detectors. **Micronetics, Inc., Norwood, N.J. 07648 or INFO/CARD #137.**

L-C Filters Catalog

Allen Avionics Catalog 19F presents the firm's extensive line of custom-built L-C Filters in frequencies from 20 Hz to 500 MHz. Included are bandpass, band reject, linear phase, highpass and lowpass filters, and video filters consisting of NTSC lowpass, reject and bandpass types. Also covered are sub-miniature lowpass, highpass and bandpass filters with frequencies up to 500 MHz. The company specializes in fast prototypes. The 24 page catalog contains a special Filters Facts Section and Glossary of Filter Terms. **Allen Avionics, Inc., Mineola, NY 11501 or INFO/CARD #135.**

Attenuator And Termination Catalog

Elcom Systems, Inc. catalog No. 831 has 32 pages listing a complete line of coaxial attenuators and terminations as well as minimum loss pads, multicouplers, double balanced mixers, detectors, and RF Fuses. They are available in BNC, N, TNC, and SMA connectors. These units operate over the frequency range of DC to 4.2 GHz. This free catalog is available from **Elcom Systems, Inc. Boca Raton, FL 33431, or INFO/CARD #138.**

Application Note

Anzac Division of Adams-Russell has available an article on the application of RF Amplifier Loss-Less Feedback Technology. Advantages to this technology include improved noise figures, increased output power performance and higher third order intercepts. In addition to design approaches, this application note contains a listing of Anzac's entire High Dynamic Range Amplifier product line. **Adams Russell, Anzac Division, Burlington, MA 01803, or please circle INFO/CARD #136.**

Designers Guide

Analogic announces the availability of the new fifth edition of its "Designers Guide and Handbook" for data conversion products. If offers a convenient source of information — formulae, definitions and application considerations — essential to signal processing systems which meet the user's real-world requirements. The 32-page book also includes condensed specifications and selection criteria for a broad range of A/D and D/A converters, sample-and-hold circuits, and ancillary devices (analog multiplexers, filters, isolation amplifiers, power supplies, and modular subsystems) useful in creating precision signal-handling circuits and systems. **Analogic Corporation, Wakefield, MA 01880, or please circle INFO/CARD #134.**

Microwave Coaxial Components Catalog

Gilbert Engineering's NEW G874 and G900 Catalogs offer a complete line of 14mm 50 ohm connectors, adapters, terminations, attenuators, airlines, power dividers, low pass filters, miscellaneous coupling elements and cable assemblies all utilizing the

G874, G890 and G900 connectors. In addition, many of the above items are also available in 75 ohm versions. **Gilbert Engineering, Phoenix, AZ or please circle INFO/CARD #133.**

Molded Bobbins Catalog

New 20-page catalog offers the most comprehensive listing of molded bobbins available with complete dimensional data in easy to use form. Included are round core, rectangular stack, reed switch, pot core, transformer, and square stack types. Listed are various materials such as nylon, glass reinforced nylon, polypropylene, Delrin, as well as numerous other substances. This catalog also provides a complete listing of solder lug styles and 70 different flange designs available. A separate section is devoted to a detailed explanation of fabricated bobbins suitable for low-volume applications. Illustrated are a wide variety of sizes, configurations, termination styles and materials. **Bobbins, Inc., Chicago, IL 60639 or INFO/CARD #132.**

Product Note

A new product note from Hewlett-Packard, "Bandwidth, Probing and Precise Time-Interval Measurements," provides a concise, comprehensive review of the factors that affect the accuracy of time-interval measurements. As a reference guide for test-system designers, the product note demonstrates how an understanding of system bandwidth and probing limitations can improve measurement repeatability. This 6-page note (part no. 5953-3926) has seven figures and three tables. It discusses waveform distortion that can be caused by measurement-system bandwidth and by probing techniques used in measuring small (picosecond) time intervals. It covers how this distortion can affect the time-interval measurement. The product note also gives guidelines to help determine if choosing a high-bandwidth instrument to eliminate errors is worth the increased waveform distortion caused by the high bandwidth. **Hewlett Packard Company, Palo Alto, CA 94303 or INFO/CARD #131.**

Components Catalog

Thor Electronics Corporation of Linden, New Jersey U.S.A. announces the released of their new twelve page catalog of "Microwave Semiconductors and Electron Tubes". This catalog lists over 3800 items and is the most detailed listing of microwave semiconductors and electron tubes available. **Thor Electronics Corporation, Linden, N.J. 07036 or please circle INFO/CARD #130.**

pick a mixer

standard level (+7dBm LO)

from 500 KHz to 1GHz... hi-rel and industrial miniature, flatpack, and low profile from $3.95

Choose from the most popular mixers in the world. Rugged construction and tough inspection standards insure MIL-M-28837/1A performance.*

Check these features...

SRA-1 the world standard, covers 500 KHz to 500 MHz, Hi-REL, 3 year guarantee, HTRB tested, MIL-M-28837/1A-03 S performance.* $11.95 (1-49).

TFM-2 world's tiniest Hi-REL units, 1 to 1000 MHz, only 4 pins for plug-in or flatpack mounting, MIL-M-28837/1A performance.* $11.95 (6-49).

SBL-1 world's lowest cost industrial mixers, only $3.95 1 to 500 MHz, all metal enclosure.

SBL-1X industrial grade, low cost, $4.95 (10-49) 10 to 1000 MHz, rugged all metal enclosure.

ASK-1 world's smallest double-balanced mixers, 1-600 MHz, flat-pack mounting, plastic case, $5.95 (10-49).

*Units are not QPL listed

INFO/CARD 85

MODEL	SRA-1	TFM-2	SBL-1	SBL-1X	ASK-1
FREQUENCY, MHz					
LO, RF	.5-500	1-1000	1-500	10-1000	1-600
IF	DC-500	DC-1000	DC-500	.5-500	DC-600
CONVERSION LOSS, dB					
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
mid range	40	40	35	30	35
upper bandedge	30	30	25	20	20

For complete specifications and performance curves refer to the 1980-1981 Microwaves Product Data Directory, the Goldbook or EEM.

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setting higher standards

Mini-Circuits

A Division of Scientific Components Corporation
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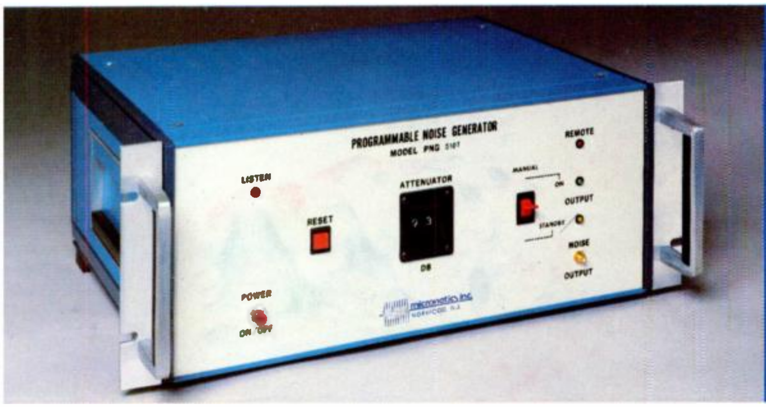
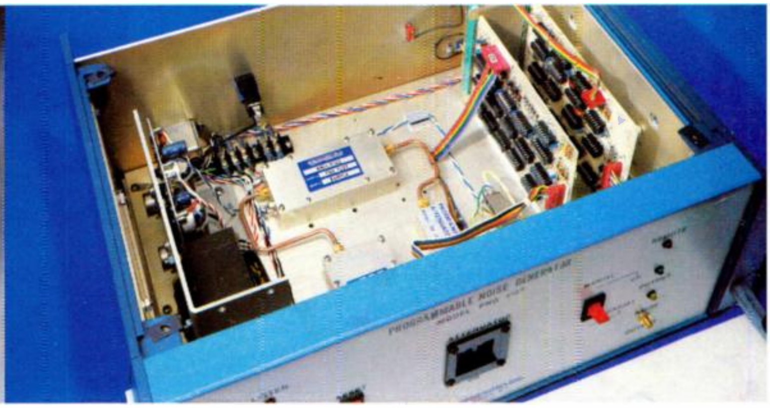
ADVERTISER INDEX

Meet some of our new noise family.





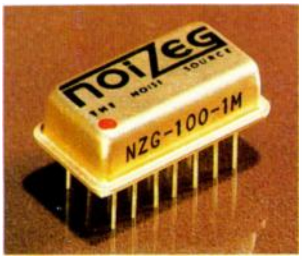
Micronetics new solid state broadband programmable noise generator.

MODEL NO.	FREQUENCY RANGE	FLATNESS	VSWR
PNG 5103	10 Hz–500 KHz	± 0.5 dB	1.5:1
PNG 5107	100 Hz–100 MHz	± 1.50 dB	1.5:1
PNG 5109	30 MHz–500 MHz	± 2.5 dB	1.5:1
PNG 5111	1 GHz–2 GHz	± 2.5 dB	2.0:1
PNG 5200	100 Hz–1000 MHz	± 2.5 dB	2.0:1

PNG 5000 Series

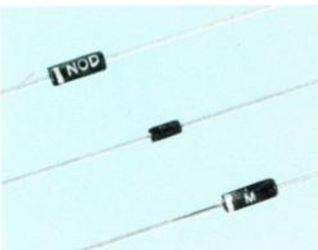
- Digitally programmable from remote source via GPIB (IEEE-488 bus line)
- Full noise source capability from 10 Hz–2 GHz
- High crest factor (peak to rms ratio higher than 5)
- Completely Solid State
- Output greater than 10 dBm (10 mw) across given frequency band
- 0-99 dB attenuation range in 1 dB steps
- Rack mounting available

Noise Generators



NZG Series/100 Hz to 5 MHz
Up to 150 mv output/White Gaussian Noise/14 Pin DIP
Small size and weight.
QUALIFIED PEAK FACTOR: 5:1 (min)
FREQUENCY FLATNESS: ± 0.5 DB

Noise Diodes



- Audio, VHF, UHF, RF and Millimeter
- White Gaussian Noise
- Hermetically Sealed
- Meet Mil Requirements
- 10 Hz to 40 GHz
- Glass and Pil Packages

Noise Sources

- Audio, VHF, UHF, RF & Millimeter
- White Gaussian Noise
- Hermetically Sealed
- Meet Mil Requirements
- Up to 5 Volt Output
- Qualified Peak Factor 5:1 (min)
- 10 Hz to 50 GHz

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We're introducing our latest development — the OSP™ Modular Plug-In Connector. Designed for cost effective electronic system packaging. More than just a connector, it's a unique packaging solution in answer to the demand for high density packaging. OSP™ connectors eliminate external cables — modules literally snap together. Multiple connectors mate simultaneously. And, they operate reliably to 26 GHz even when only partially mated. Components utilizing OSP™ connectors are also available.

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