

rfdesign

engineering principles and practices

Fall 1989

1990 directory

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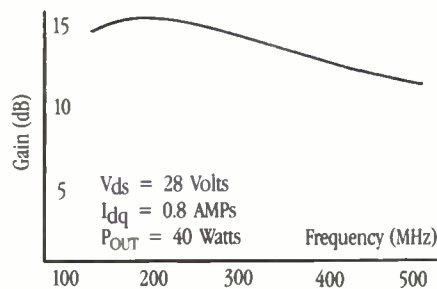
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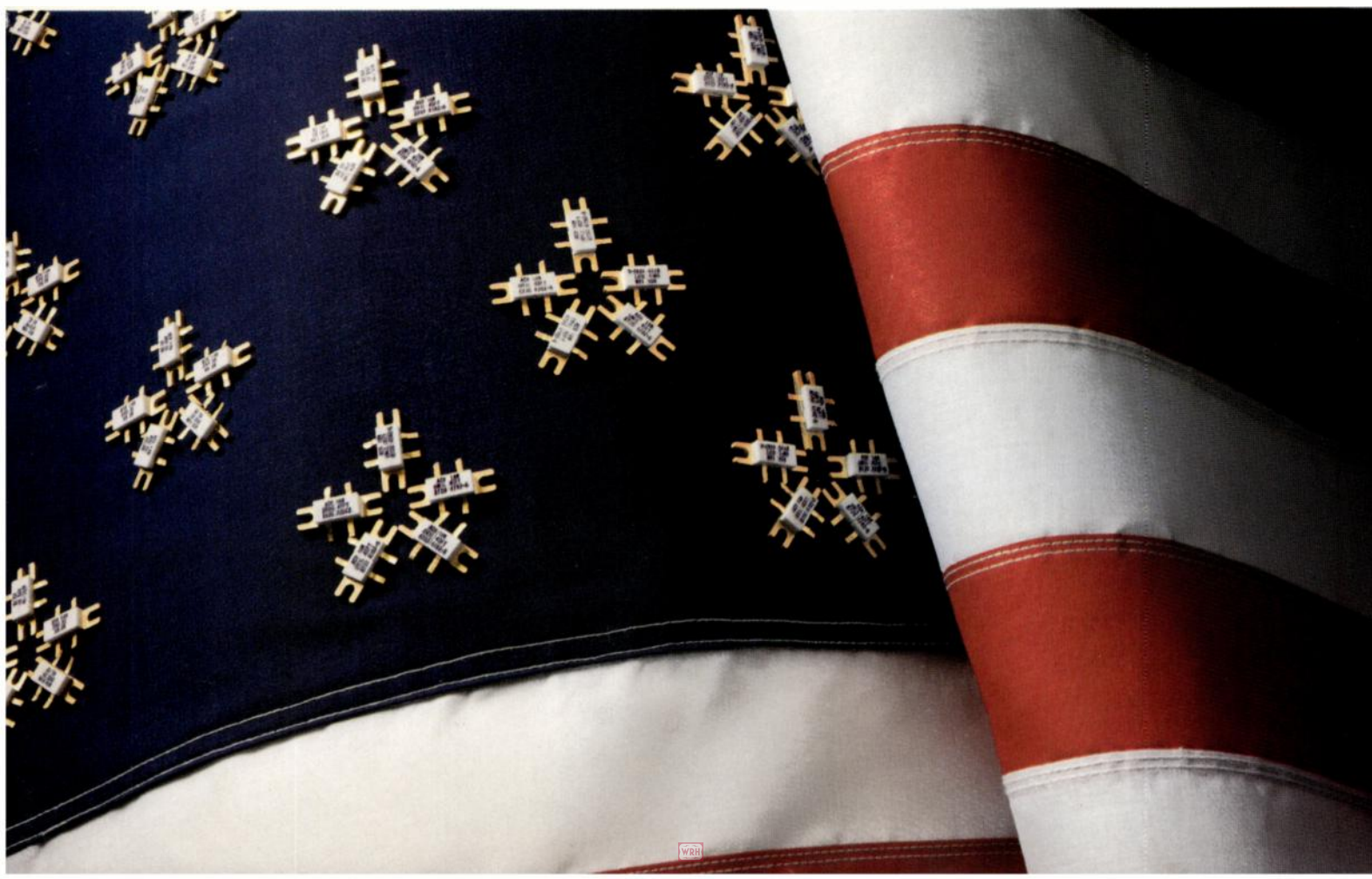
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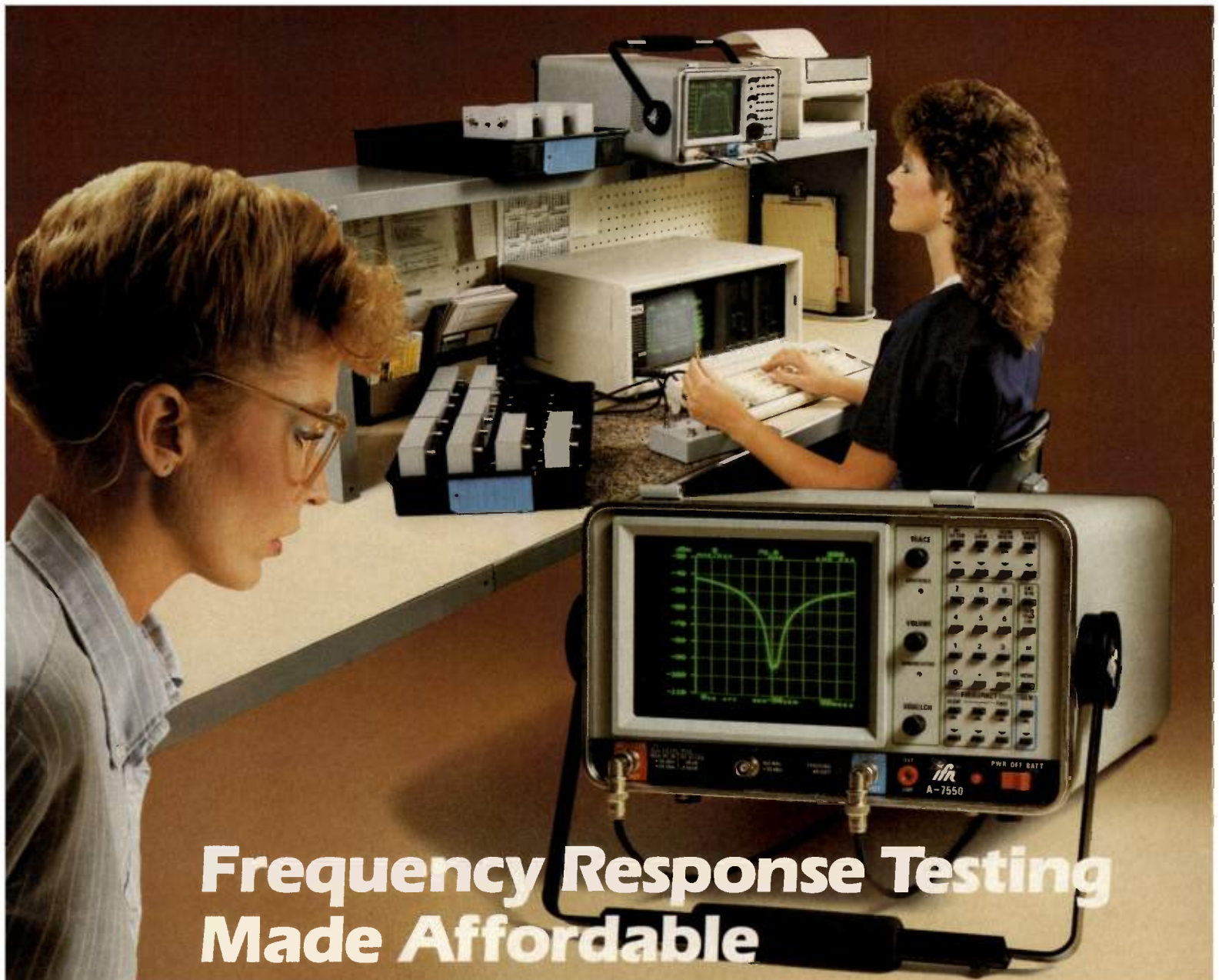
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6 Editorial — About the 1990 RF Design Directory

9 Components

Capacitors	9
Crystals	11
Diodes	12
Inductors	14
Integrated Circuits	14
Resistors	14
Transformers	15
Transistors	15
Vacuum Tubes	16

16 Modules and Hybrids

Amplifiers	16
Detectors	17
Filters	17
Frequency Multipliers	22
Limiters	22
Mixers	22
Modulators	22
Noise Sources	24
Oscillators	24
Resonators	29

29 System Blocks and Subassemblies

Amplifiers	29
Frequency Converters	30
Frequency Synthesizers	31
Motion/Proximity Sensors	31
Receivers	31
RF Modems	32
Tone Encoders/Decoders	32
Transmitters	32

34 Transmission Components

Antennas	34
Antenna Positioners	34
Attenuators	35
Coaxial Cable	36
Directional Couplers	36
Equalizers	38
Filters	38
Isolators/Circulators	43
Loads	43
Multicouplers	43
Multiplexers/Diplexers	43
Phase Shifters	44
Power Dividers/Combiners	44

Relays	45
RF Connectors	45
Switches	48
Terminations	49
Tuners	49

49 CAD/CAE Software & Systems

Software	49
Systems—Hardware	52

52 Materials and Hardware

Cabinets and Enclosures	52
Ceramics	52
Coatings — Environmental	52
Component/Hybrid Packages	52
Crystal Blanks	52
Electronic Hardware	52
Equipment Racks	52
Ferrites and Other Magnetics	52
Heat Sinks	53
Insulating Materials	53
PC Boards	53
Sockets	53
Substrates	53
Tools	53

53 Test Equipment

Amplifiers (Lab)	53
Antenna Measurement Systems	54
Calibrators/Standards	54
Dummy Loads	54
Field Intensity Meters	55
Frequency Counters	58
Modulation Meters	58
Network Analyzers	58
Noise Figure Meters	59
Noise Generators	59
Oscilloscopes	59
Power Meters	59
Probes	60
Pulse Generators	60
RLC/Impedance Bridges	60
Signal Level Meters	62
Signal/Sweep Generators	62
Spectrum Analyzers	62
SWR Meters	62
Test Fixtures	62
Wave/Distortion Analyzers	62

64 Services

Antenna Measurement	64
Circuit Assembly	64
Consultants	64
Custom Hybrids	64
Custom ICs and ASICs	64
EMC Testing	66
Equipment Calibration	66
Instrument Rental/Leasing	66
PC Layout/Fabrication	66
Plating/Coating	66
Software Development	66
Training/Education	66

68 EMC Products

Absorbers	68
AC Line Filters	68
Anechoic Chambers	68
Conductive Adhesives/Lubricants	68
Conductive Coatings	68
ESD Control Devices	68
Gaskets/Finger Stock	68
Hum Stop Filters	68
Shielded Enclosures	68
Shielding	69
Surge/Transient Protectors	69
Test Cells	69
Test Equipment & Systems	69

70 Design Guide

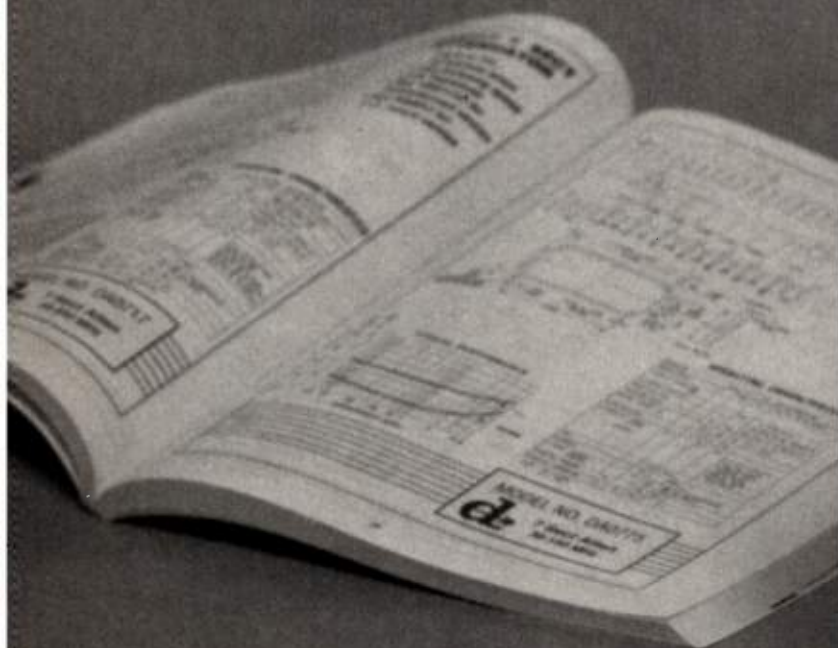
Chebyshev Lowpass Filter Design	70
Elliptic Lowpass Filter Design	73
Principles of Transistor Interactive Mapping	81

87 Company Index

122 Advertisers' Index

123 Info/Card

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About the 1990 RF Design Directory



By Gary A. Breed
Editor

This is the sixth annual *RF Design* Directory Issue. With each one, we have refined our data base of RF supplier companies. For example, this issue added over 200 company names and deleted close to 100. The Company Index now includes about 1300 firms who supply products or services to the RF industry.

For the third year, we continue our Design Guide section, with tutorial material that is the tradition of *RF Design*'s regular issues. Two articles on filter design by David Greene of NAP Consumer Electronics include excellent reviews of Chebyshev and elliptic filter design mathematics. Computer programs that accompany these two notes have been combined with Greene's Butterworth filter program (March 1989 *RF Design*) in the "disk-of-the-month" available from the RF Design Software Service.

Charles Ruelke has contributed another design tutorial on a graphical method of determining input and output reflection coefficients of transistor circuits. Oscillator and amplifier designers can benefit from his method of analysis. We invite your comments on these articles, and also want your suggestions for future Design Guide subject matter.

How to Use the Directory

There are nine major groupings of product/service types, from components to EMC products. Within each of these groups are a number of sub-categories. The Table of Contents identifies the location of all these categories. In the product listings, additional detailed breakdowns are included.

For example, to find out what companies to contact when you're looking for a GaAs FET, you should find the Components category, and the Transistors section under that heading. Turning to the page listed, you will find separate listings for Small Signal Bipolars, FETs, RF Power Transistors, and GaAs FETs. Make a note of the companies identified as providing GaAs FETs, and go to the back of the issue (after the Design Guide section) to the pages marked with black edges. This is the Company Index, an alphabetical list of RF suppliers. The address, phone, and contact person are included for each company. We have some FAX numbers, too, and are encouraging companies to add this information to their listings.

We have tried to come up with the "right" level of detail and complexity in our Directory. We don't have dozens of similar product categories like some directories, and we don't have the same sales offices listed a hundred times. Certainly, there will be times when more detail would be useful, but overall, readers tell us that the relative simplicity makes finding potential suppliers quick and easy.

Finally, some of you may be looking for an article index like we had in last year's Directory. It's not here — wait for the December issue. From now on, every December *RF Design* will have an index to articles published during the previous two years.

Keep this issue nearby during 1990. Then, when you need that special widget to meet your design specs, look it up and call the company. When you locate what you need, tell them you found it here!

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Switching: 5-20 μ s
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Spurious Outputs: -70dB

Phase Noise: -63dBc (0.15KHz)
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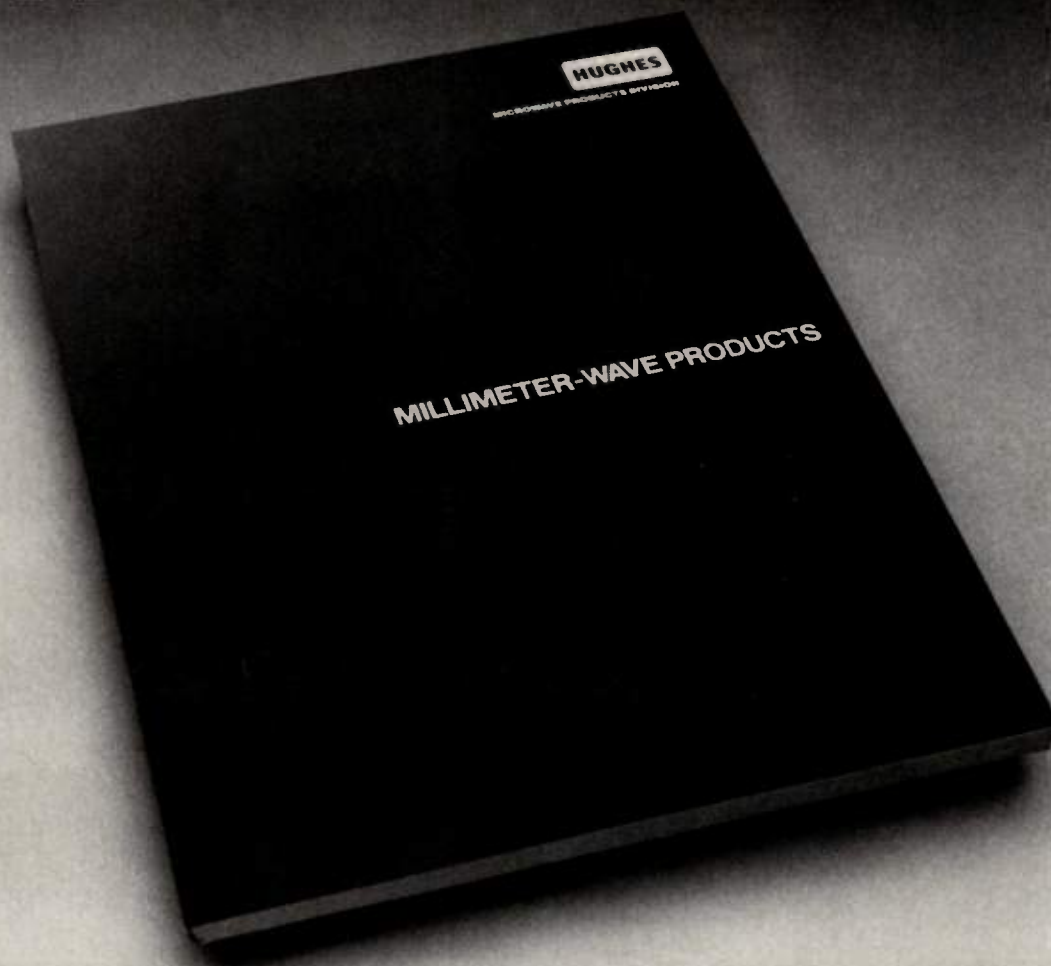
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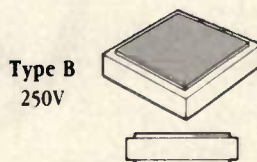
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32 bits, 14 milli-Hz @ 60 MHz

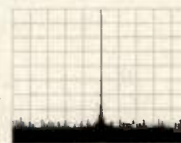
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HIGH SPECTRAL PURITY

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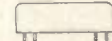
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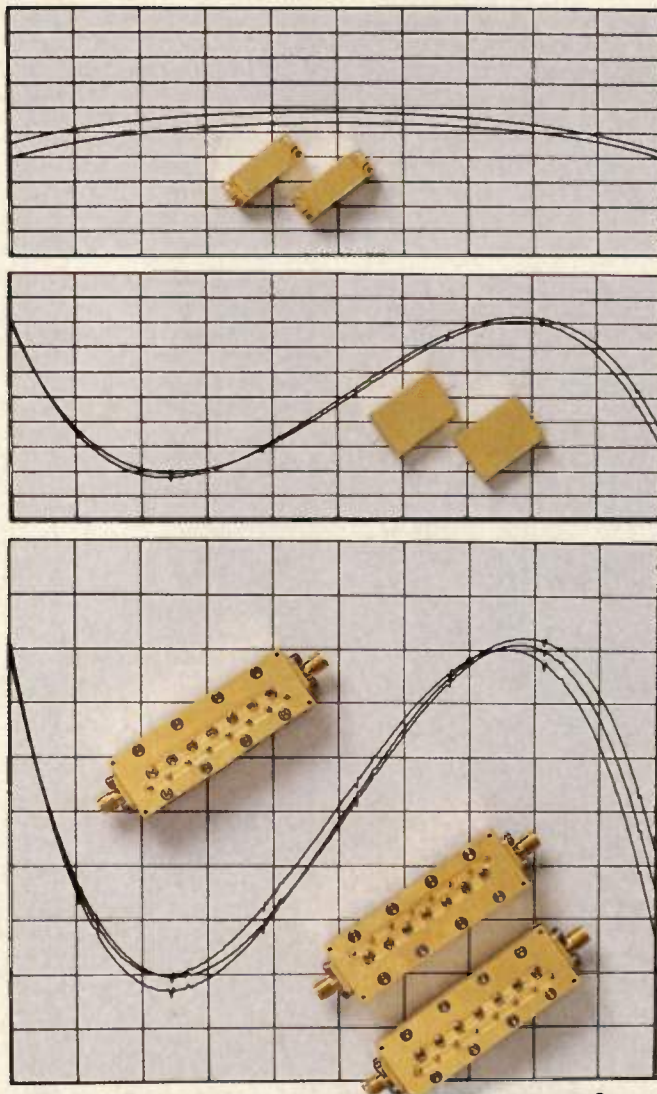
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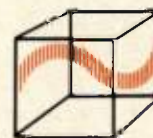
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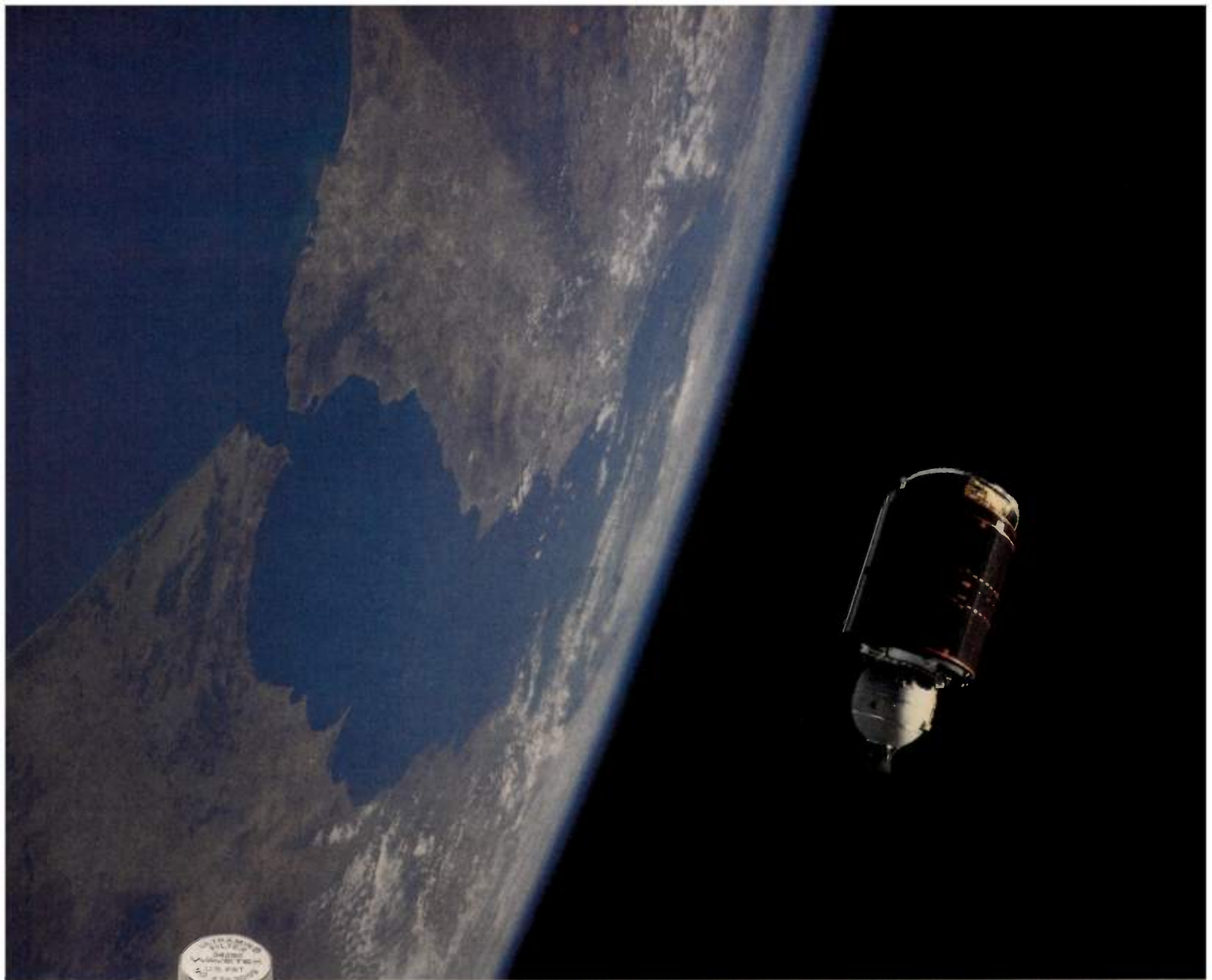
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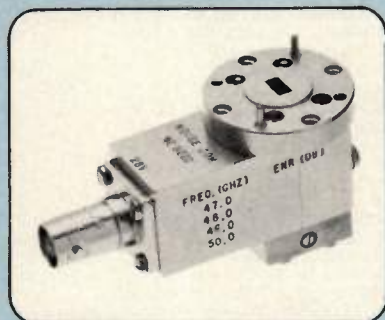
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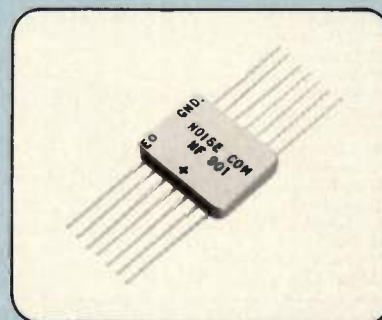
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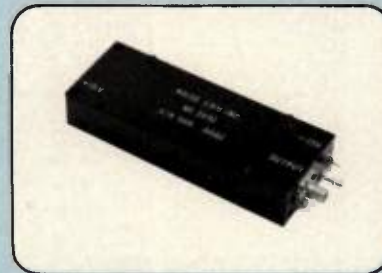
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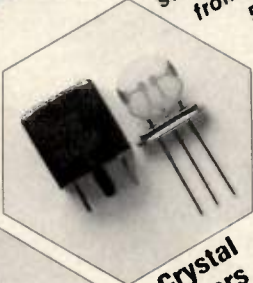
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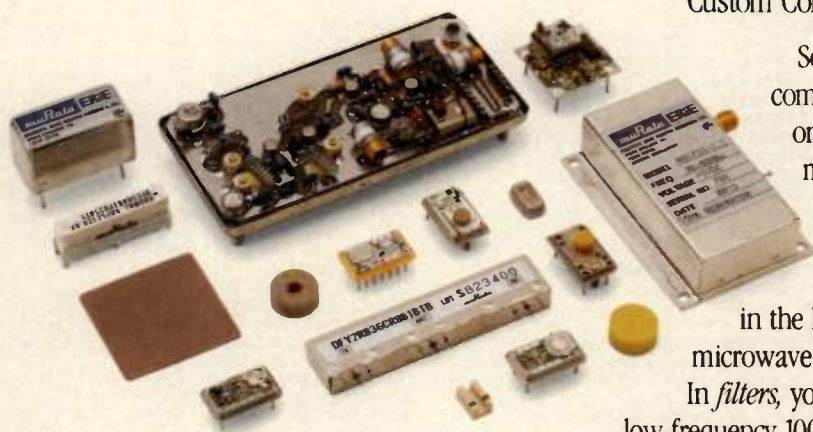
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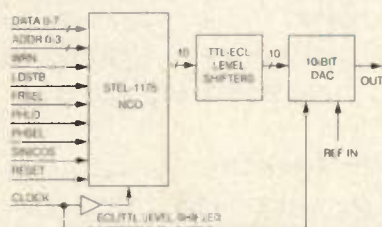
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TELEDYNE MICROWAVE
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VDS-3125/VME	110	<1	-40	-110	.15
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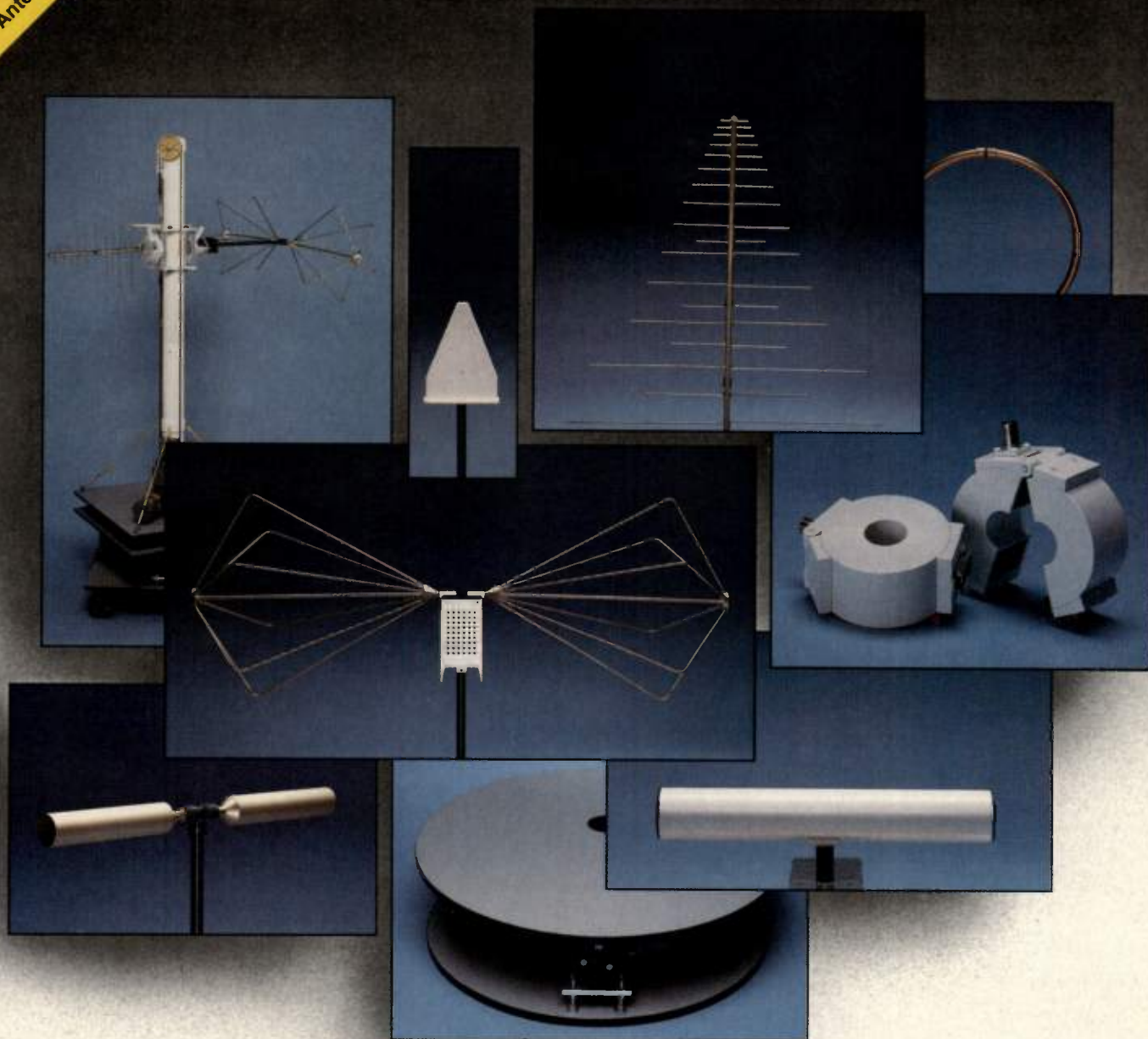
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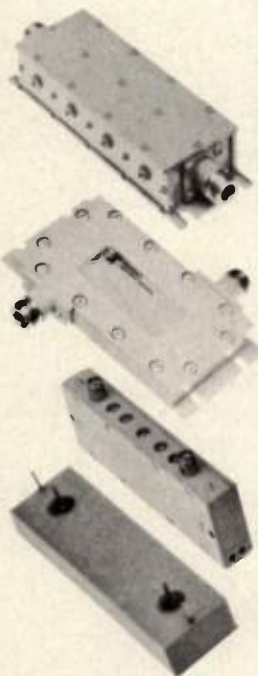
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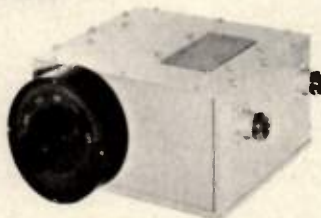
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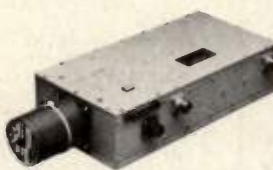


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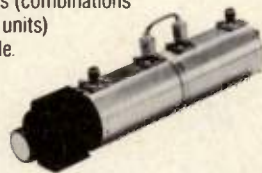
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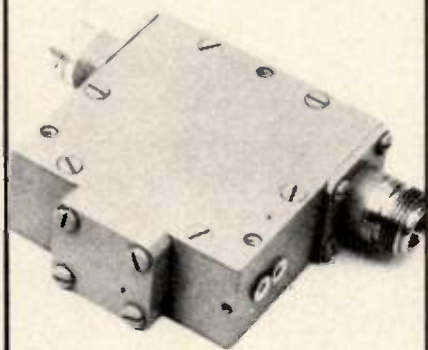
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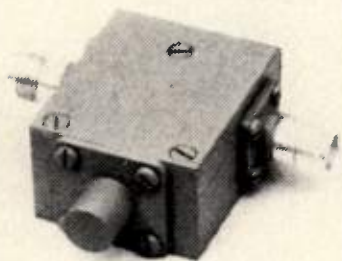
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HIGH POWER COMBINERS



**2-WAY 600 WATTS
MODEL D1994**

**4-WAY 600 WATTS
MODEL D1995**

**4-WAY 1500 WATTS
MODEL D1996**

**SPECIFICATIONS
above models**

Freq. Range	20-90 MHz
Insertion Loss	0.3 dB max.
Isolation	20 dB min.
VSWR	1.3:1 max.

HIGH POWER COUPLERS

**DUAL DIRECTIONAL
COUPLER
MODEL C1997**

Freq. Range	20-90 MHz
Coupling	40 dB
Directivity	20 dB min.
Power (watts)	1500

See Gold Book and Microwave/RF Handbook, for additional products.



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**See
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Ad on
Coaxial Switching**

Page 37

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**See
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Ad on
Coaxial Switching**

Page 37

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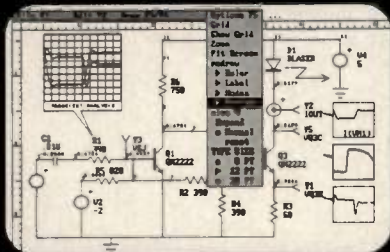
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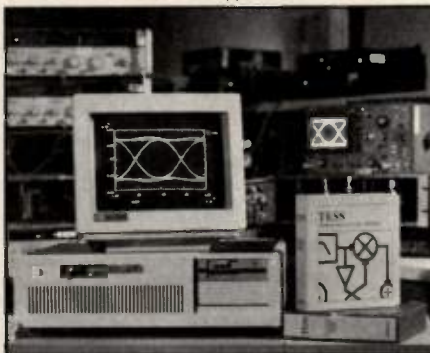
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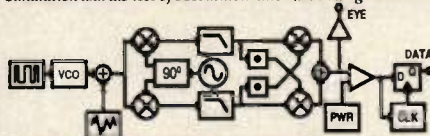
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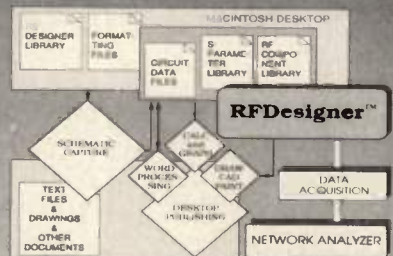
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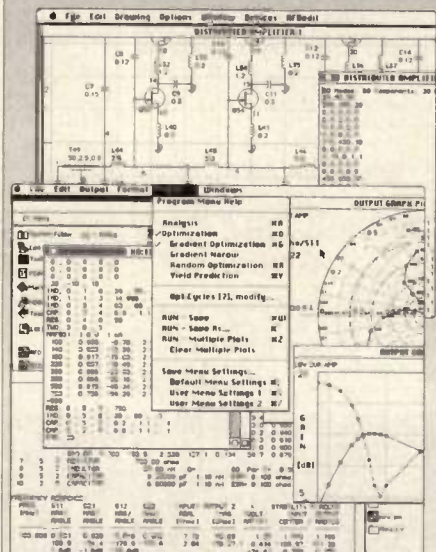


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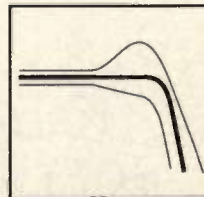
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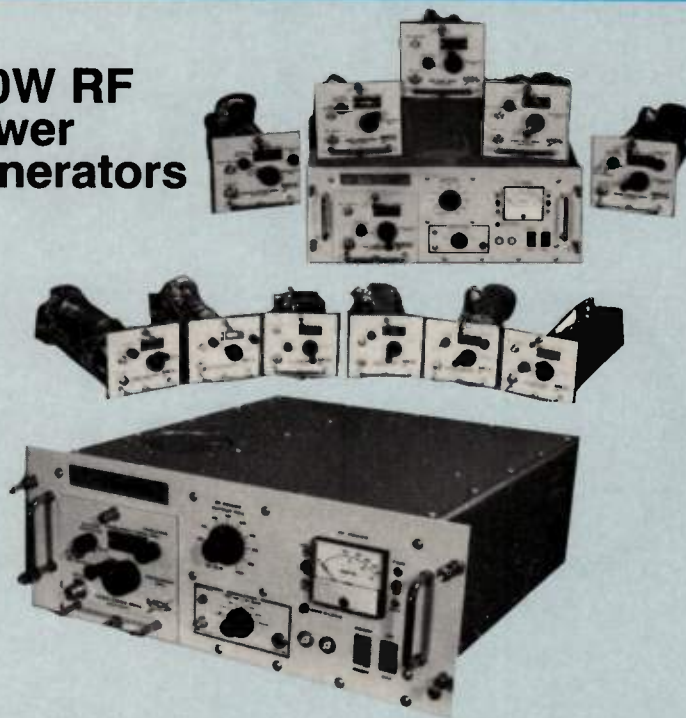
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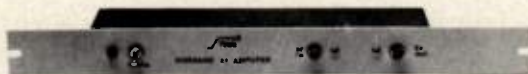
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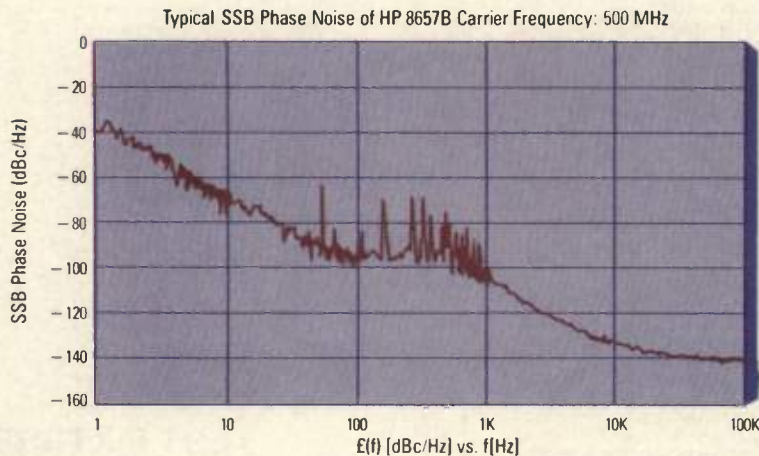
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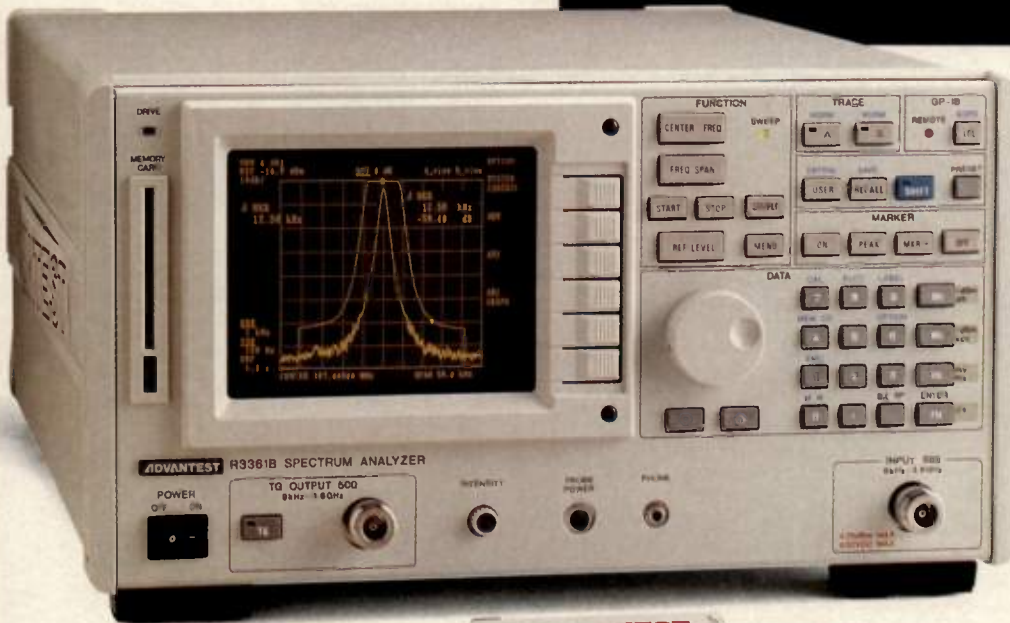
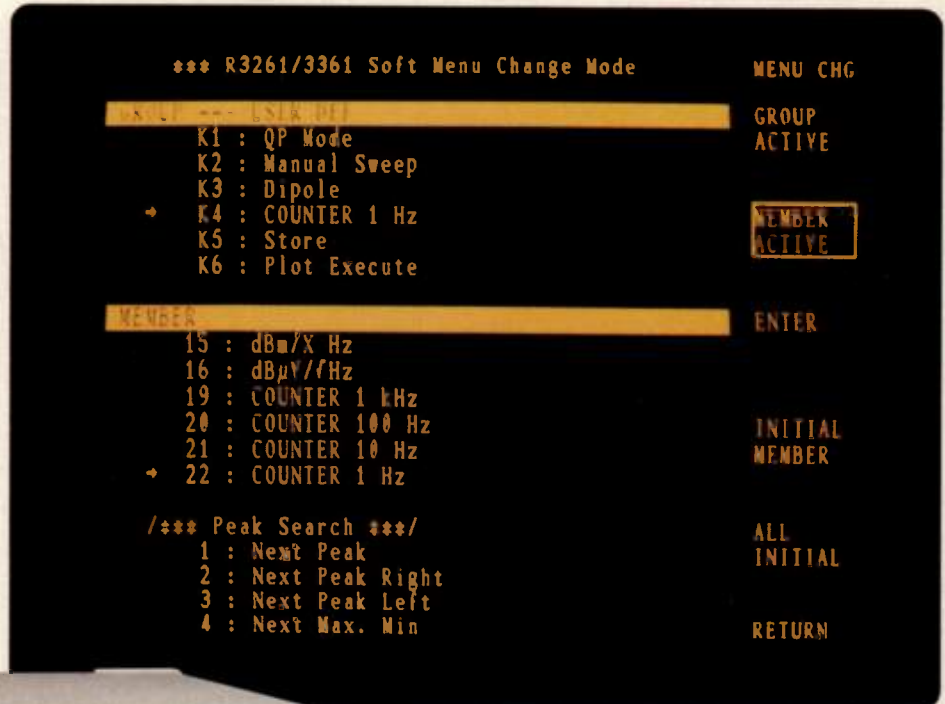
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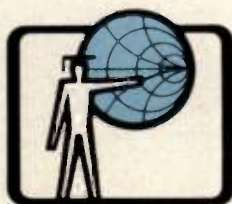
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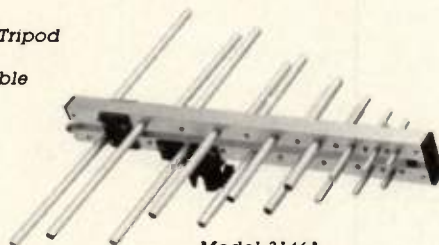
Model 3115 Double Ridge Guide
1-18 GHz
Model 3116 Double Ridge Guide
18-40 GHz



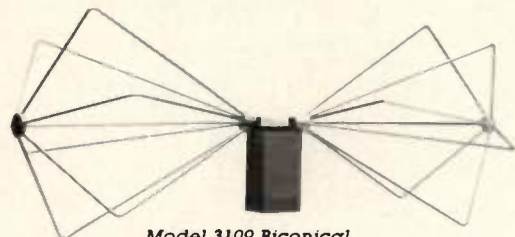
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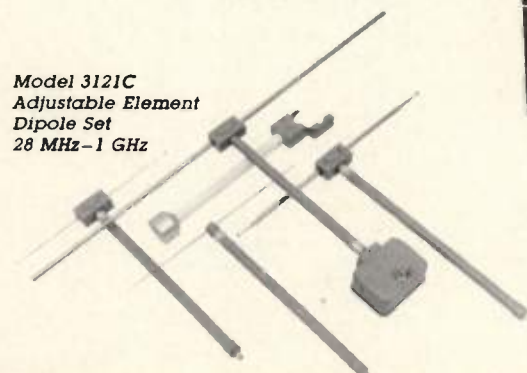
Model 1060 Turntable
with controller and IEEE-488 bus option
1-7.6 meter diameter available
three other models available



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Log Periodic
300 MHz-1 GHz
four other models available



Model 3109 Biconical
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Model 3121C
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four other models available



Model 3925 PLISN
5 kHz-1 GHz
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Chebyshev Lowpass Filter Design

By David C. Greene
Philips Consumer Electronics Co.

This article presents a computer program for use in designing Chebyshev equal ripple lowpass filters. A previous article (1) described a Butterworth approximation which gave a filter that was maximally flat at the origin. That approximation, while offering mathematical simplicity, produces a slow transition between passband and stopband. In situations where one is concerned with a band of frequencies, the Chebyshev approximation is often more appropriate. This approximation minimizes the magnitude of the largest error. It has the important property of being equi-ripple, with the error oscillating between minimum and maximum values of equal amplitude.

The N th-order Chebyshev polynomial $T_n(\omega)$ is defined as having the following properties (2):

- T_n is even (odd) if n is even (odd)
- T_n has all its zeros in the interval $-1 < x < 1$
- T_n oscillates between ± 1 in the interval $-1 \leq x \leq 1$
- $T_n(1) = +1$

This polynomial was discovered by P.L. Tschebyscheff while he was studying the construction of steam engines. He used these functions to describe various problems encountered in the steam engine design. His 1899 paper (3) was used by Dr. W. Causer to design a class of filters now known as Chebyshev type filters. (The spelling Tschebyscheff is the German translation of the Russian and has often been changed to the English spelling Chebyshev.)

From the above,

$$|T_n| \leq 1 \text{ for } -1 \leq x \leq 1$$

and outside this interval, the magnitude of the polynomial becomes very large in comparison with unity. It is precisely these properties that make the Chebyshev polynomial so useful for the approximation problem. The practical lowpass filter characteristic function given in Figure 1 is used. The maximum passband loss is:

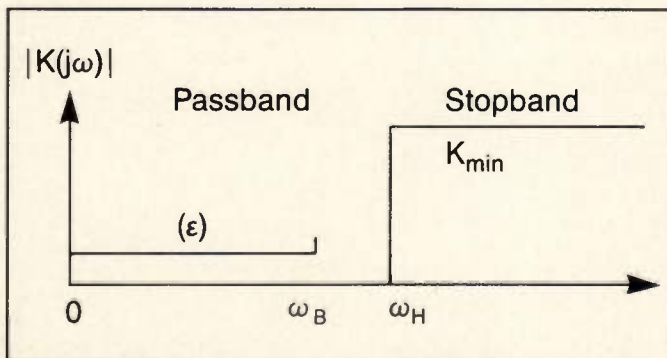


Figure 1. Lowpass filter characteristic function.

$$A_{\max} = 10 \log(1 + \epsilon^2)$$

and the minimum stopband loss is:

$$A_{\min} = 10 \log(1 + K_{\min}^2)$$

Since,

$$|K(j\omega)|^2 = \epsilon^2 T_n^2(\omega/\omega_B)$$

$$\text{and } [H(s)][H(-s)] = 1 + [K(s)][K(-s)]$$

the Chebyshev lowpass approximation can be written:

$$A(\omega) = 10 \log\{1 + [\epsilon T_n(\omega/\omega_B)]^2\}$$

$$\text{And } \epsilon^2 = 10^{0.1A_{\max}} - 1$$

$$\text{thus } \epsilon = \sqrt{10^{0.1A_{\max}} - 1}$$

If $(\omega/\omega_B) > 1$, then

$$T_n(x) = \cosh [N \operatorname{arccosh} (x)]$$

```

ENTER PASSBAND FREQUENCY (FC) IN Hz ? 4E6
ENTER PASSBAND RIPPLE (Amax) IN DB ? .5
DO YOU KNOW THE ORDER OF THIS LOWPASS (Y/N) ? Y
WHAT IS THE ORDER (MUST BE LESS THAN 20) ? 7
ENTER THE STOPBAND FREQUENCY (FS) IN Hz ? 5.5E6

THE 3 DB BANDWIDTH OF THIS LOWPASS WILL BE: 41.204E+05 HZ
ATTENUATION OF THIS LOWPASS AT (FS) WILL BE: 35.97983 DB
N = 7

ENTER TERMINATION RESISTANCE IN OHMS ? 75
SELECT TERMINATION TYPE ( 1 ) EQUAL OR ( 2 ) EXTREME ? 1
ZERO FREQUENCY GROUP TIME DELAY IN SECONDS = 25.091E-08

TSCHEBYSCHIEFF LOWPASS FILTER

****R****      R= 75.000E+00(OHMS)
*              *
****C****      C=0.92166E-09(FARADS)
*              *
*              L=L=0.37548E-05(HENRYS)
*              *
****C****      C=0.13997E-08(FARADS)
*              *
*              L=L=0.40117E-05(HENRYS)
*              *
****C****      C=0.13997E-08(FARADS)
*              *
*              L=L=0.37548E-05(HENRYS)
*              *
****C****      C=0.92166E-09(FARADS)
*              *
****R****      R= 75.000E+00(OHMS)

TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>
    
```

Figure 2. Program example.


```

5 REM ===== Chebyshev Lowpass Filter Design Program "CHEB.BAS" 1-31-88 =====
10 REM
20 REM
50 REM
60 REM
70 REM ===== Input Data =====
80 CLS
90 DIM G(20), C(20), L(20), A(20), B(20), D(20)
100 PRINT "THIS PROGRAM DETERMINES ALL THE NECESSARY PARAMETERS TO DESIGN"
110 PRINT "EQUAL OR EXTREME TERMINATION TSCHIBYSCHIEFF LOWPASS FILTERS"
120 PRINT "DAVID C. GREIME, PHILIPS CONSUMER ELECTRONICS CO., KNOXVILLE, TN"
130 PRINT "PRINT : INPUT "ENTER PASSBAND FREQUENCY (FC) IN HZ " : FC
131 IF FC <= 0 THEN 130
135 INPUT "ENTER PASSBAND RIPPLE (Amax) IN DB " : A1
136 IF A1 < 0 OR A1 > 3 THEN 135
140 INPUT "DO YOU KNOW THE ORDER OF THIS LOWPASS (Y/N) " : Z5
150 IF Z5 = "N" OR Z5 = "n" GOTO 190
160 IF Z5 = "Y" OR Z5 = "y" GOTO 170 ELSE GOTO 140
170 INPUT "WHAT IS THE ORDER (MUST BE LESS THAN 20) " : N
175 IF N <= 0 OR N > 20 THEN 170
180 GOTO 220
190 INPUT "ENTER THE MINIMUM STOPBAND ATTENUATION (Amin) IN DB " : A2
200 IF A2 < 0 THEN A2 = -1 * A2
210 IF A2 <= A1 THEN PRINT : PRINT "NOT MUCH OF A LOWPASS!!...Amin<Amax"
220 GOTO 130
230 INPUT "ENTER THE STOPBAND FREQUENCY (FS) IN HZ " : FS
240 IF FS < FC THEN PRINT "LOOKS LIKE A HIGHPASS!!...FS<FC" : GOTO 130
240 PRINT : PRINT
250 IF FS = FC THEN PRINT "YOU ARE NOT LOWPASSING ANYTHING...FS=FC!!" : GOTO 130
260 F1 = 1 : F2 = FS / FC
300 REM ===== Tschebyscheff order or Attenuation =====
310 DE = .43429448198 : DEF FNLOG(X) = LOG(X) * DE : PI = 4 * ATN(1)
320 Y = F2 / F1 : YV = LOG(Y + SQR(Y^2 - 1))
330 IF Z5 = "Y" OR Z5 = "y" THEN 370
340 E = 10 ^ (.1 * A1) - 1 : F = 10 ^ (.1 * A2) - 1
350 X = SQR(F / E) : XX = LOG(X + SQR(X^2 - 1))
360 N = XX / YV : IF INT(N) < N THEN N = INT(N) + 1
370 T = (EXP(N * YV) + EXP(-N * YV)) / 2 : E = 10 ^ (.1 * A1) - 1
380 A = 10 * FNLOG((1 + E^2) / 2) : IF Z5 = "Y" OR Z5 = "y" THEN A2 = A
390 YBW = SQR(1 / E) : ABW = LOG(YBW + SQR(YBW^2 - 1)) / N
384 COHB = (EXP(ABW) + 1) / (EXP(ABW) - 1) : BWT = COHB * FC
386 PRINT "THE 3 DB BANDWIDTH OF THIS LOWPASS WILL BE " : PRINT USING "###.##"
390 PRINT "ATTENUATION OF THIS LOWPASS AT (FS) WILL BE: " : A : "DB " : PRINT
395 IF N > 20 THEN PRINT "PROGRAM CANNOT ACCOMMODATE THIS HIGH OF AN ORDER!!" : GOTO 130
396 IF A < A2 THEN PRINT "ORDER IS TOO HIGH !!!" : GOTO 130
400 INPUT "ENTER TERMINATION RESISTANCE IN OHMS " : TER
410 IF TER <= 0 OR TER > 10000 THEN 400
420 INPUT "SELECT TERMINATION TYPE ( 1 ) EQUAL OR ( 2 ) EXTREME " : TEX
430 IF TEX < 1 OR TEX > 2 THEN 420
440 GOSUB 5000
450 CR = 1 / (TER * 2 * PI * FC) : LR = TER / (2 * PI * FC)
460 IF TEX = 2 AND INT(N / 2) = N / 2 THEN GOSUB 1500 ELSE 490
470 TEX = 1 : GOSUB 5000 : TEX = 2 : TD = 0
480 FOR K = 1 TO N : TD = TD + G(K) : NEXT K : TD = TD / (4 * PI * FC) : GOTO 500
490 IF TEX = 2 AND INT(N / 2) < N / 2 THEN GOSUB 1600 ELSE 495
491 TEX = 1 : GOSUB 5000 : TEX = 2 : TD = 0
492 FOR K = 1 TO N : TD = TD + G(K) : NEXT K : TD = TD / (4 * PI * FC) : GOTO 500
493 IF TEX = 1 AND INT(N / 2) < N / 2 THEN GOSUB 1600 ELSE 495
494 FOR K = 1 TO N : TD = TD + G(K) : NEXT K : TD = TD / (4 * PI * FC) : GOTO 500
495 FOR K = 1 TO N : TD = TD + G(K) : NEXT K : TD = TD / (4 * PI * FC)
496 GOTO 1700
500 REM ===== Printing routine =====
510 PRINT "ZERO FREQUENCY GROUP TIME DELAY IN SECONDS = " : PRINT USING "###.###"
520 PRINT : PRINT "TSCHIBYSCHIEFF LOWPASS FILTER " : PRINT
530 PRINT "*****R***** R=" : PRINT USING "###.###" : TER : PRINT
540 PRINT "*****C***** C=" : PRINT USING "###.###" : C(K) : PRINT
550 PRINT "*****L***** L=" : PRINT USING "###.###" : L(K+1) : PRINT
560 PRINT "*****H***** H=" : PRINT USING "###.###" : L(K) : PRINT
570 PRINT "*****F***** F=" : PRINT USING "###.###" : L(K+1) : PRINT
580 PRINT "*****B***** B=" : PRINT USING "###.###" : L(K) : PRINT
590 PRINT "*****D***** D=" : PRINT USING "###.###" : L(K+1) : PRINT
600 PRINT "*****G***** G=" : PRINT USING "###.###" : L(K) : PRINT
610 PRINT "*****A***** A=" : PRINT USING "###.###" : L(K+1) : PRINT
620 PRINT "*****N***** N=" : PRINT USING "###.###" : L(K) : PRINT
630 PRINT "*****T***** T=" : PRINT USING "###.###" : L(K+1) : PRINT
640 PRINT "*****Y***** Y=" : PRINT USING "###.###" : L(K) : PRINT
650 PRINT "*****Z***** Z=" : PRINT USING "###.###" : L(K+1) : PRINT
660 PRINT "*****X***** X=" : PRINT USING "###.###" : L(K) : PRINT
670 PRINT "*****W***** W=" : PRINT USING "###.###" : L(K+1) : PRINT
680 PRINT "*****V***** V=" : PRINT USING "###.###" : L(K) : PRINT
690 PRINT "*****U***** U=" : PRINT USING "###.###" : L(K+1) : PRINT
700 PRINT "*****S***** S=" : PRINT USING "###.###" : L(K) : PRINT
710 PRINT "*****R***** R=" : PRINT USING "###.###" : L(K+1) : PRINT
720 PRINT "*****Q***** Q=" : PRINT USING "###.###" : L(K) : PRINT
730 PRINT "*****P***** P=" : PRINT USING "###.###" : L(K+1) : PRINT
740 PRINT "*****O***** O=" : PRINT USING "###.###" : L(K) : PRINT
750 PRINT "*****N***** N=" : PRINT USING "###.###" : L(K+1) : PRINT
760 PRINT "*****M***** M=" : PRINT USING "###.###" : L(K) : PRINT
770 PRINT "*****L***** L=" : PRINT USING "###.###" : L(K+1) : PRINT
780 PRINT "*****K***** K=" : PRINT USING "###.###" : L(K) : PRINT
790 PRINT "*****J***** J=" : PRINT USING "###.###" : L(K+1) : PRINT
800 PRINT "*****I***** I=" : PRINT USING "###.###" : L(K) : PRINT
810 PRINT "*****H***** H=" : PRINT USING "###.###" : L(K+1) : PRINT
820 PRINT "*****G***** G=" : PRINT USING "###.###" : L(K) : PRINT
830 PRINT "*****F***** F=" : PRINT USING "###.###" : L(K+1) : PRINT
840 PRINT "*****E***** E=" : PRINT USING "###.###" : L(K) : PRINT
850 PRINT "*****D***** D=" : PRINT USING "###.###" : L(K+1) : PRINT
860 PRINT "*****C***** C=" : PRINT USING "###.###" : L(K) : PRINT
870 PRINT "*****B***** B=" : PRINT USING "###.###" : L(K+1) : PRINT
880 PRINT "*****A***** A=" : PRINT USING "###.###" : L(K) : PRINT
890 PRINT "*****Z***** Z=" : PRINT USING "###.###" : L(K+1) : PRINT
900 PRINT "*****Y***** Y=" : PRINT USING "###.###" : L(K) : PRINT
910 AS = INKEY$ : IF LEN(AS) = 0 THEN 940
920 IF ASC(AS) = 13 THEN 130
930 IF ASC(AS) = 81 OR ASC(AS) = 113 THEN 950
940 GOTO 910
950 END

```

Figure 3. Chebyshev lowpass filter design program listing.

Solving for $A(\omega)$:

$$A(\omega) = 10 \log\{1 + [\epsilon \cosh(N \operatorname{arccosh}(\omega/\omega_0))]^2\}$$

Solving for N:

$$N = \frac{\operatorname{arccosh}\sqrt{(10^{0.1A_{\min}} - 1)/(10^{0.1A_{\max}} - 1)}}{\operatorname{arccosh}(\omega/\omega_0)}$$

These equations are solved by the computer program in the following manner. The inverse hyperbolic cosines and hyperbolic cosine are found using:

$$\operatorname{ARCCOSH}(X) = \ln(X + \sqrt{X^2 - 1})$$

and

$$\operatorname{COSH}(X) = (\exp(X) + \exp(-X))/2$$

Lines 320-360 solve the equation for N. If N is not an integer value, the next larger integer value is used for N. Lines 370-380 solve the equation for the loss $A(W)$ at the stopband frequency FS. Lines 382-384 solve for the 3 dB down frequency BWT using:

$$\text{BWT} = \text{FC} \cdot \operatorname{COSH}[(\operatorname{ARCCOSH}((1/(10^{0.1A_{\max}} - 1))^{0.5}))/N]$$

For more information on the above equations, see Reference 2.

Dr. H.J. Orchard has derived equations to solve implicitly the natural modes for the Chebyshev approximation. These equations directly yield the normalized LC component values for both equally terminated and extreme termination Chebyshev lowpass filters. The equations for the doubly terminated, normalized values are (4):

$$G_0 = 1.0$$

$$\omega_1' = 1.0$$

$$\beta = \ln \left[\coth \left(\frac{A_{\max}}{40 \log(e)} \right) \right] \quad \text{where } e = 2.718282$$

$$\gamma = \sinh \left(\frac{\beta}{2N} \right)$$

$$A_k = \sin \left[\frac{(2k-1)\pi}{2N} \right] \quad \text{for } k = 1, 2, \dots, N$$

$$B_k = \gamma^2 + \sin^2(k\pi/N) \quad \text{for } k = 1, 2, \dots, N$$

$$G_1 = 2A_1/\gamma$$

$$G_k = \frac{4A_{k-1}A_k}{B_{k-1}G_{k-1}} \quad \text{for } k = 2, 3, \dots, N$$

$$G_{N+1} = 1.0 \quad \text{for } N \text{ odd}$$

$$G_{N+1} = \coth^2(\beta/4) \quad \text{for } N \text{ even}$$

The program solves for β and γ in lines 5010 and 5020. The hyperbolic cotangent is found using the relation:

$$\text{COTH}(X) = ((\text{EXP}(X) + (1/\text{EXP}(X)))/(\text{EXP}(X) - (1/\text{EXP}(X))))$$

A_k and B_k are solved for in line 5040, G_1 in line 5080, and G_k in line 5100. G_{N+1} for N even is solved for and denormalized in line 1710.

For the extreme termination case, normalized values are (4):

$$D_k = \left[\gamma^2 + \sin^2 \left(\frac{k\pi}{2N} \right) \right] \cos^2 \left(\frac{k\pi}{2N} \right) \quad \text{for } k = 1, 2, \dots, N-1$$

$$G_1 = A_1/\gamma$$

$$G_k = \frac{A_k A_{k-1}}{D_{k-1} G_{k-1}} \quad \text{for } k = 2, 3, \dots, N$$

$$G_{N+1} = \infty$$

D_k is solved for in line 5050 of the program, G_1 in line 5120, and G_k in line 5140.

The actual component values for even-order filters are calculated in lines 1510-1530. They are:

$$L(K) = LR \cdot G(K) \quad \text{for } K = 1, 3, 5, \dots, N$$

$$C(K+1) = CR \cdot C(K+1) \quad \text{for } K = 1, 3, 5, \dots, N$$

$$LR = \text{TER}/(2 \cdot \text{PI} \cdot \text{FC})$$

$$CR = 1/(\text{TER} \cdot 2 \cdot \text{PI} \cdot \text{FC})$$

where, TER = termination resistance, PI = 3.14159265, and FC = passband frequency.

The actual component values for the odd-order filters are calculated in lines 1610-1630. The values are:

$$L(K+1) = LR \cdot G(K+1) \quad \text{for } K = 1, 3, 5, \dots, N$$

$$C(K) = CR \cdot G(K) \quad \text{for } K = 1, 3, 5, \dots, N$$

For the calculation of group delay at zero frequency, if RS = RL (the network terminations are equal), then the group time delay as $(\omega/\omega_B) \rightarrow 0$ is as follows:

$$T'(d0) = (d\phi/d\omega')|_{\omega' \rightarrow 0}$$

$$= (1/2) \sum_{k=1}^N G_k \text{ seconds} \quad (\text{normalized delay})$$

The actual delay, calculated in line 494 or 495, is given by:

$$\text{Actual Delay} = T'(d0)/(2 \cdot \text{PI} \cdot \text{FC})$$

For extreme termination, the program goes back and calculates the values of G_k for equal termination and then computes the time delay. For more information on these equations, see Reference 4.

The Program

A program has been written in Microsoft QuickBasic 4.5 to solve the above expressions and calculate the component values for a given filter. The program listing is given in Figure 3. The program prompts for the passband frequency FC and the maximum passband ripple in dB. It then asks if the order of the filter is known. If it is, the program asks for the stopband frequency FS and calculates the attenuation $A_{\min}$ at FS. If the order N is not known, the program prompts for the minimum stopband attenuation $A_{\min}$ and the stopband frequency FS. It then calculates the required order N .

Next, the program prints out the 3 dB bandwidth of the filter, the attenuation at FS, and the order N . It prompts for the termination resistance in ohms and the termination type, equal or extreme. The normalized component values are calculated. From these, the normalized group delay at DC is calculated and then denormalized to give the actual group delay at DC for the selected passband frequency FC. Then, the component values are denormalized to yield actual component values. Finally, these values are printed out in a format that forms the schematic connection diagram for the filter network being designed.

Figure 2 presents an example in which the program is used to design a seventh-order Chebyshev lowpass filter with equal 75 ohm terminations, a passband frequency of 4 MHz, a passband ripple of 0.50 dB, and a stopband frequency of 5.5 MHz.

For this example, the program calculated an attenuation of 35.98 dB at the stopband frequency of 5.5 MHz, with a 3 dB down frequency of 4.12 MHz and a DC group delay of 251 ns. The calculated component values are as shown in Figure 2 with the indicated schematic connection diagram and termination resistances.

This program is available on disk from the RF Design Software Service. See Page 80 for details. 

References

1. David C. Greene, "A Design Program for Butterworth Lowpass Filters," *RF Design*, March, 1989.
2. Richard W. Daniels, *Approximation Methods for Electronic Filter Design*, McGraw-Hill Book Co., 1974.
3. "Theorie des Mecanismes Connus Sous le Nom de Parallelogrammes," *Oeuvres*, Vol. 1, 1899.
4. Matthaei, Young and Jones, *Microwave Filters, Impedance-Matching Networks, and Coupling Structures*, McGraw-Hill Book Co., 1964.

About the Author

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Elliptic Lowpass Filter Design

By David C. Greene
Philips Consumer Electronics Co.

This article presents a series of computer programs written in Microsoft QuickBasic for the design of elliptic function lowpass filters using the IBM PC. Two polynomial filters, the Butterworth or maximally flat response and the Chebyshev equal ripple passband response, have been discussed previously (1,2). Since both of these are polynomial filters, they contain no finite poles of attenuation. The Chebyshev approximation spreads its ability to flatten its error in an equi-ripple manner across the entire passband instead of at the origin (as the Butterworth approximation does). Thus, it has a faster rate of cutoff. This rate of cutoff becomes larger as the passband ripple is allowed to increase.

Adding attenuation poles to an equal ripple passband polynomial filter can produce a stopband that has equal minima. If the filter described by such a polynomial has an equal ripple passband and an equal minima stopband, it is referred to as an elliptic filter, since these filter functions are found using elliptic functions. Elliptic filters are also known as Caue or Zolotarev filters since much of the design work was done by Dr. W. Caue, Zolotarev, and also Dr. S. Darlington.

Elliptic filters have equal loss maxima in the passband and equal loss minima in the stopband, and are referred to as equal ripple passband and stopband. They are a special case of filters that have an equal ripple passband and an arbitrary stopband. Iterative computer programs can be written for this more general type of filter. The Butterworth and Chebyshev filters can be considered special forms of this general form. For instance, if the passband ripple is set to zero, and the first loss pole and $A_{\min}$ (the minimum stopband loss) are both set to infinity, the Butterworth filter is the result. However, if only

$A_{\min}$ and the first loss pole are set to infinity, the result is the Chebyshev filter.

The Chebyshev Rational Function

The Chebyshev polynomial $T_n(x)$ was discussed in the treatment of the Chebyshev lowpass filter (2). The Chebyshev rational function will now be looked at to design the elliptic filter. The following discussion summarizes work by R.W. Daniels, described in more detail in Reference 3.

The Chebyshev rational function $R_n(x, L)$ is analogous to the Chebyshev polynomial $T_n(x)$ discussed previously. While $T_n(x)$ is a polynomial, $R_n(x, L)$ represents a ratio of two polynomials. It is defined as a rational function of x with the following properties:

- R_n is even (odd) if n is even (odd)
- R_n has all its n zeros in the interval $-1 < x < 1$ and all its n poles outside that interval
- R_n oscillates between ± 1 in the interval $-1 \leq x \leq 1$
- $R_n(1, L) = +1$
- $1/R_n$ oscillates between $\pm 1/L$ in the interval $|x| < x_1$

where x_1 is the first value of x for which $R_n(x, L) = L$. The characteristic function for an elliptic filter is:

$$|K(j\omega)|^2 = [\epsilon R_n(\omega/\omega_B, L)]^2$$

$$\text{where } \epsilon^2 = 10^{0.1A_{\max}} - 1$$

The equation

$$A(\omega) = 10 \log\{1 + [\epsilon R_n(\omega/\omega_B, L)]^2\}$$

describes the n th-order elliptic lowpass filter. For the Chebyshev rational function $R_n(x, L)$,

$$R_n(x, L) = L/R_n(x_1/x, L) \quad (1)$$

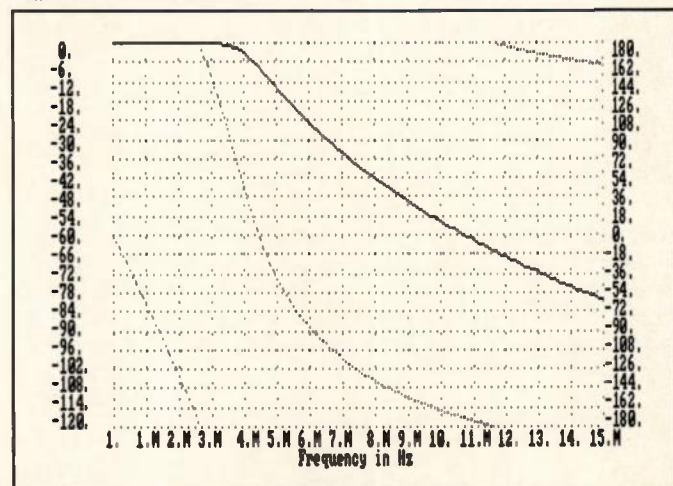


Figure 1(b). Butterworth lowpass filter.

```

Amax = 3DB, ENTER PASSBAND FREQUENCY (FC) IN HZ ? 4e6
DO YOU KNOW THE ORDER OF THIS LOWPASS (Y/N) ? Y
WHAT IS THE ORDER (MUST BE LESS THAN 20) ? 7
ENTER THE STOPBAND FREQUENCY (FS) IN HZ ? 6e6
ATTENUATION OF THIS LOWPASS AT (FS) WILL BE: 24.64707 DB N= 7
ENTER THE TERMINATION RESISTANCES IN OHMS ? 75
SELECT TERMINATION TYPE ( 1 ) EQUAL OR ( 2 ) EXTREME ? 1
ZERO FREQUENCY GROUP TIME DELAY IN SECONDS = 17.881E-08

```

BUTTERWORTH MAXIMALLY FLAT LOWPASS FILTER

```

****R****      R= 75.000E+00(OHMS)
*              *
****C****      C= 0.23610E-09(FARADS)
*              *
*              L= 0.37212E-05(HENRYS)
*              *
****C****      C= 0.95596E-09(FARADS)
*              *
*              L= 0.59683E-05(HENRYS)
*              *
****C****      C= 0.95596E-09(FARADS)
*              *
*              L= 0.37212E-05(HENRYS)
*              *
****C****      C= 0.23610E-09(FARADS)
*              *
****R****      R= 75.000E+00(OHMS)

```

TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>

Figure 1(a). Butterworth program MAXF run.


```

ENTER PASSBAND FREQUENCY (FC) IN Hz ? 3.88e6
ENTER PASSBAND RIPPLE (Amax) IN DB ? .5
DO YOU KNOW THE ORDER OF THIS LOWPASS (Y/N) ? y
WHAT IS THE ORDER (MUST BE LESS THAN 20) ? 7
ENTER THE STOPBAND FREQUENCY (FS) IN Hz ? 6e6

THE 3 DB BANDWIDTH OF THIS LOWPASS WILL BE: 39.967E+05 HERTZ
ATTENUATION OF THIS LOWPASS AT (FS) WILL BE: 45.81599 DB N = 7
ENTER TERMINATION RESISTANCE IN OHMS ? 75
SELECT TERMINATION TYPE ( 1 ) EQUAL OR ( 2 ) EXTREME ? 1
ZERO FREQUENCY GROUP TIME DELAY IN SECONDS = 25.867E-08
  
```

TSCHEBYSCHOFF LOWPASS FILTER

```

****R****      R= 75.000E+00 (OHMS)
*              *
****C****      C=0.95017E-09 (FARADS)
*              *
*              L=0.38709E-05 (HENRYS)
*              *
****C****      C=0.14429E-08 (FARADS)
*              *
*              L=0.41358E-05 (HENRYS)
*              *
****C****      C=0.14429E-08 (FARADS)
*              *
*              L=0.38709E-05 (HENRYS)
*              *
****C****      C=0.95017E-09 (FARADS)
*              *
****R****      R= 75.000E+00 (OHMS)
  
```

TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>

Figure 2(a). CHEBY program run.

This characteristic is a desirable one since, from equation 1 and the properties already described,

$$|R_n(x, L)| \geq L \text{ for } |x| > x_i$$

Thus, an equal ripple passband will automatically give an equal ripple stopband. Equation 1 also shows that if $x = x_i$ is a pole of the Chebyshev rational function, then $x = x_i/x_i$ is a zero. In addition, if x_i is a pole of the function, $-x_i$ is also a pole. It can then be written:

for n even —

$$R_n(x, L) = C_1 \prod_{i=1}^{n/2} \frac{x^2 - (x_i/x_i)^2}{x^2 - x_i^2}$$

for n odd —

$$R_n(x, L) = C_2 x \prod_{i=1}^{(n-1)/2} \frac{x^2 - (x_i/x_i)^2}{x^2 - x_i^2}$$

$$\text{where } C_1 = \sqrt{L} \prod_{i=1}^{n/2} x_i^2/x_i$$

$$\text{and } C_2 = \sqrt{L/x_i} \prod_{i=1}^{(n-1)/2} (x_i/x_i)^2$$

The Jacobian Elliptic Functions

Elliptic integrals and elliptic functions are used to determine a differential equation for the Chebyshev rational function $R_n(x, L)$. With the Jacobian elliptic sine function $\text{sn}(u, k)$, the poles and zeros of the above expressions can be described. $R_n(x, L)$ can then be expressed as:

for n odd —

$$R_n(x, L) = C_1 x \prod_{v=1}^{(n-1)/2} \frac{x^2 - [\text{sn}(2vK/n)]^2}{x^2 - [x_i/\text{sn}(2vK/n)]^2}$$

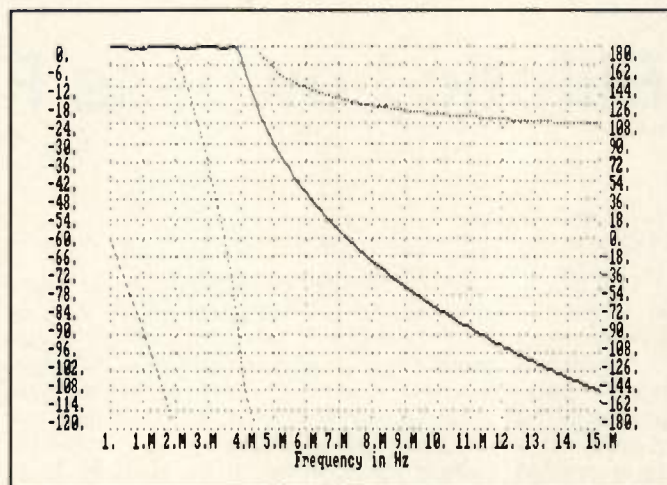


Figure 2(b). Chebyshev lowpass filter.

for n even —

$$R_n(x, L) = C_2 \prod_{v=1}^{n/2} \frac{x^2 - (\text{sn}[(2v-1)K/n])^2}{x^2 - \{x_i/\text{sn}[(2v-1)K/n]\}^2}$$

where the modulus k is equal to $1/x_i$. The modulus k affects the complete elliptic integral K as well. This parameter K can now be written as:

$$K = K(k) = K(1/x_i)$$

Now that an expression for $R_n(x, L)$ has been obtained, the loss of an elliptic filter at any frequency can be evaluated.

The elliptic integral of the first kind is given as:

$$u(\phi, k) = \int_0^\phi \frac{1}{\sqrt{1 - k^2 \sin^2 x}} dx$$

The modular angle θ is related to k by the definition:

$$k = \sin \theta$$

and the elliptic sine function is defined as $\text{sn}(u, k) = \sin(\phi)$. The complete elliptic integral of the first kind is defined as:

$$u(\pi/2, k) \text{ is } K = \int_0^{\pi/2} \frac{1}{\sqrt{1 - k^2 \sin^2 x}} dx$$

Similarly, the complementary modulus k' is given in terms of the modulus k , where:

$$k' = \sqrt{1 - k^2}$$

and the complementary elliptic integral K' is defined in terms of the complementary modulus k' , as follows:

$$u(\pi/2, k') \text{ is } K' = \int_0^{\pi/2} \frac{1}{\sqrt{1 - (1 - k^2)(\sin^2 x)}} dx$$

The following equation is used to solve for n :

$$n = \frac{K(1/x_i)K'(1/L)}{K'(1/x_i)K(1/L)} \quad (\text{program line 1440})$$


```

ENTER PASSBAND RIPPLE (AMAX), & STOPBAND LOSS (AMIN) IN DB
? .5,81.2
ENTER PASSBAND FREQ.(FB), & STOPBAND FREQ.(FH) IN (HZ)
? 3912e3,6e6
.....Calculating Elliptical Functions.....

DEGREE (N) REQUIRED IS : 7

LOSS POLE FREQUENCIES ARE :
1.242585E+07
7302214
6116355
FREQUENCY (F) TO CALCULATE LOSS AT= ? 4e6
DB LOSS= 3.03825
TO CONTINUE PRESS <ENTER> :TO GO TO START PRESS <1> :TO EXIT PRESS <Q>

```

Figure 3(a). ELT1 program run.

```

PASS BAND EDGE (KHZ)= ? 3912
STOP BAND EDGE (KHZ)= ? 6000
NUMBER OF FINITE ZEROS (1-15)= ? 3
CRITICAL Q= 20.40297
STOP BAND REJECTION (DB)= ? 81.2
PASSBAND RIPPLE (DB)= .5067881
3 DB (KHZ) ABOUT = 4004.905190377026
NOMINAL OHMS RESISTANCE = ? 75

```

** LOW-PASS FILTER **

```

****R****      R= 75.000E+00      (OHMS)
****C****      C= 0.90694D-09      (FARADS)
****          (N)= 1
****          C= 0.45305D-10      (FARADS)
****          L= 0.36211D-05      (HENRYS)
****          F= 1242.6D+01      (KHZ)
****          C= 0.12217D-08      (FARADS)
****          (N)= 3
****          C= 0.21912D-09      (FARADS)
****          L= 0.30900D-05      (HENRYS)
****          F= 6116.4D+00      (KHZ)
****          C= 0.11450D-08      (FARADS)
****          (N)= 2
****          C= 0.15138D-09      (FARADS)
****          L= 0.31380D-05      (HENRYS)
****          F= 7302.2D+00      (KHZ)
****          C= 0.81490D-09      (FARADS)
****R****      R= 75.000E+00      (OHMS)

```

PRECISION TEST: 5.904122E-12

Figure 4. Results of the elliptic design program ELFIL5.EXE run.

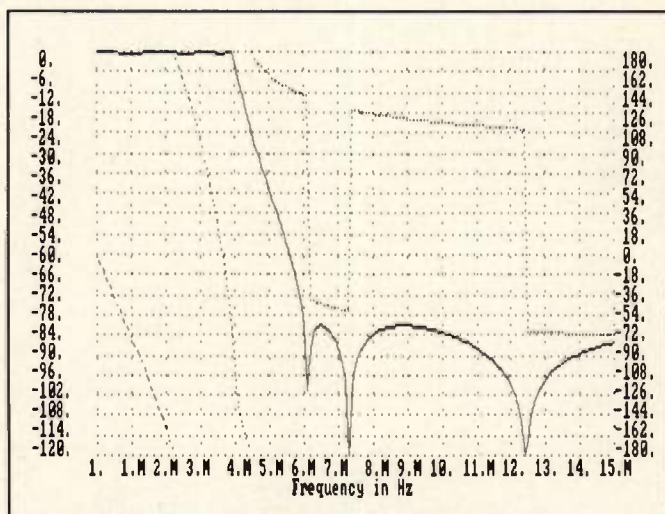


Figure 3(b). Elliptic lowpass filter.

where:

$x_i = FH/FB$ (the stopband/passband ratio)

$$L^2 = \frac{10^{0.1A_{\min}} - 1}{10^{0.1A_{\max}} - 1} \quad (\text{program line 1110})$$

Elliptic Lowpass Filter Loss Program

The Microsoft QuickBasic program ELT1 (Figure 8) has been written to solve these functions and calculate loss pole frequencies. This program asks for the passband ripple AMAX, the stopband loss AMIN, the passband frequency FB, and the stopband frequency FH. When this data is entered, the program calculates the order n (N) necessary in line 1440 of the program. Subroutine 5000 calculates the complete elliptic integral, called H4, and calls subroutine 11000, which performs the series expansion of $\text{SIN}(X)$. It also calls subroutine 12000, which performs the series expansion of $\text{COS}(X)$. With the order n known, $R_n(\omega/\omega_B, L)$ is found using:

$$R_n(x, L) = Cx^{1-NI} \prod \frac{x^2 - X_{zi}^2}{x^2 - X_{pi}^2}$$

where $NI = 0$ (or 1) if n is odd (or even). The poles, X_{pi} , and zeros, X_{zi} , of R_n are found in subroutine 7000. The zeros of $R_n(x, L)$ are at:

$$u = \frac{-2K(1/x_i)v}{nx_i} \quad \text{where } v = 0, 1, 2, \dots$$

Subroutine 3000 calculates $\text{CSN}(U, K)$. The loss poles are calculated in line 7200 of the program. Once the loss pole frequencies are calculated, they are printed out. At this point, the program asks for the frequency at which to calculate loss. When a frequency is entered, the loss in dB is calculated and printed out. Three choices remain: calculating another frequency's loss, starting over, or exiting the program. More information on these equations and the theory behind them can be found in Reference 3.

This program is a modified version of one by R.W. Daniels (3). His program suffers from round-off error due to the calculation of the complete elliptic integral by finite summation of finite differences. This leads to the program failing for very small passband ripple. The elliptic subroutines have been

PASS BAND EDGE (KHZ)= ? 3000
STOP BAND EDGE (KHZ)= ? 2500
NUMBER OF FINITE ZEROS (1-15)= ? 3
CRITICAL Q= 28.14362
STOP BAND REJECTION (DB)= ? 58.2
PASSBAND RIPPLE (DB)= .5049329
3 DB (KHZ) ABOUT = 2951.252434573005
NOMINAL OHMS RESISTANCE = ? 75

★★ HIGH-PASS FILTER ★★

```

****R****      R=  75.000E+00      (OHMS)
*              *
*              *
*      C      C=  0.44502D-09      (FARADS)
*              *
*      (N)=  1
*              *
*      L=  0.35820D-05      (HENRYS)
****L****C**   C=  0.39024D-08      (FARADS)
*              *
*      F=  1346.1D+00      (KHZ)
*              *
*              *
*      C      C=  0.37305D-09      (FARADS)
*              *
*      (N)=  3
*              *
*      L=  0.54492D-05      (HENRYS)
****L****C**   C=  0.76435D-09      (FARADS)
*              *
*      F=  2466.1D+00      (KHZ)
*              *
*              *
*      C      C=  0.43111D-09      (FARADS)
*              *
*      (N)=  2
*              *
*      L=  0.48941D-05      (HENRYS)
****L****C**   C=  0.11164D-08      (FARADS)
*              *
*      F=  2153.1D+00      (KHZ)
*              *
*              *
*      C      C=  0.56012D-09      (FARADS)
*              *
*              *
****R****      R=  75.000E+00      (OHMS)

```

```
PRECISION TEST: 1.534531E-13
TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>
```

Figure 5(a). Results of the ELFIL5 program run.

replaced by one which uses series expansion of sine and cosine functions, thus retaining numerical accuracy.

The power of the elliptic function is illustrated by designing a seventh-order lowpass filter with passband frequency $F_B = 4$ MHz and stopband frequency $F_H = 6$ MHz. The passband ripple, if it is equal ripple, is 0.50 dB. First, the Butterworth program MAXF is run, with the results shown in Figure 1(a).

As can be seen, the Butterworth polynomial has yielded a filter with about 24.65 dB of stopband loss at 6 MHz, with the passband frequency set to 4 MHz at the 3 dB down point. The order of this filter was set to $n = 7$. A plot of this filter is shown in Figure 1(b).

Continuing with this example, the same passband, 3 dB down frequencies, stopband frequency, and order n are used. This time, however, the Chebyshev polynomial is used for the design. The CHEBY program run is given in Figure 2(a). The advantage of the Chebyshev polynomial to spread its power over the passband is now seen. The order $n = 7$ is the same, the 3 dB down frequency is still 4 MHz, and the stopband frequency is still 6 MHz; however, the loss at 6 MHz is now 45.8 dB. Allowing 0.50 dB of passband ripple error has resulted in a much faster rate of cutoff compared to the Butterworth polynomial. A plot of this filter is shown in Figure 2(b).

The next example again uses the same passband and stopband frequencies, order n , and passband ripple as in the previous two cases. This time, the elliptic solution provided by the Chebyshev rational function is used. The ELT1 program run yields the results shown in Figure 3(a).

For the same 3 dB down frequency of 4 MHz, a passband ripple of 0.50 dB, and a stopband of 6 MHz, the elliptic solution gives 81.2 dB of loss at 6 MHz. The order is $n = 7$ as before, and the finite loss pole frequencies are calculated to be 6.116

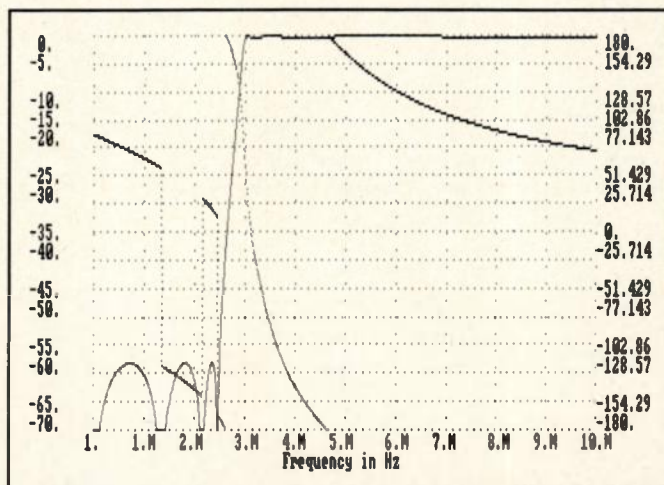


Figure 5(b). Seventh-order elliptic highpass filter designed by the program ELFIL5.

MHz, 7.3022 MHz, and 12.4258 MHz. This results in an equal ripple passband of 0.50 dB ripple and an equal ripple stopband of 81.2 dB. The much greater rate of cutoff for this elliptic function filter is evident. The price paid for this is that the stopband is not monotonic; it has points of infinite attenuation and between these points the attenuation decreases to a minimum. (That minimum is 81.2 dB in this example.) The passband is equal ripple and for this example is 0.50 dB. A plot of this filter is shown in Figure 3(b).

Amstutz Solution for Elliptic Filters

In the past, the actual design of the elliptic filter has been very difficult due to the numerical ill-conditioning of the natural modes of the Chebyshev rational function with respect to the synthesis of LC component values. This effect was especially critical in higher-order filters, since the zeros of the passband were crowded toward the edge of the passband. To alleviate this effect, bilinear transformations of the variable were used in conjunction with double or higher-order precision on mainframe computers. In 1978, Pierre Amstutz published a paper (4) in which he described an ingenious method for the design and synthesis of elliptic filters. He provided two FORTRAN programs for the design of both anti-metric and symmetrical filters. The program presented here is the Microsoft QuickBasic translation of the Amstutz FORTRAN program for symmetrical elliptic filters. This program can design both lowpass and highpass elliptic filters.

Amstutz recognized the need for a simplified means of designing elliptic filters using small computers. The elliptic approximation is used to construct the attenuation characteristic of the elliptic filter. Noting an inadequate normalization of the variables with Jacobi's sn function, Amstutz has tailored a function more suitable to the situation. He defines SNz as "the elliptic (i.e., meromorphic and doubly periodic) function which maps the rectangle $(-P+2iQ, -P, P, P+2iQ)$ of the complex z-plane on the upper half-plane," with the following correspondence:

SN0 = 0

$$\text{SN}(P + iQ) = 1$$

$$\text{SN}(2iQ) = \infty$$


```

PASS BAND EDGE (KHZ)= ? 9000
STOP BAND EDGE (KHZ)= ? 12000
NUMBER OF FINITE ZEROS (1-15) = ? 2
CRITICAL Q= 11.48639
STOP BAND REJECTION (DB)= ? 40
PASSBAND RIPPLE (DB)= .2443372
3 DB (KHZ) ABOUT = 9507.716409067003
NOMINAL OHMS RESISTANCE = ? 50

```

★★ LOW-PASS FILTER ★★

```

****R****      R=  50.000E+00      (OHMS)
*      *
*      *
***C****      C=  0.43553D-09      (FARADS)
*
*      *
*      *
(N)=  1
*      *
C=  0.78553D-10      (FARADS)
*      *
*      C L      L=  0.99027D-06      (HENRYS)
*      *
*      *
F=  1804.5D+01      (KHZ)
*      *
*      *
***C****      C=  0.60889D-09      (FARADS)
*
*      *
*      *
(N)=  2
*      *
C=  0.23345D-09      (FARADS)
*      *
*      C L      L=  0.70510D-06      (HENRYS)
*      *
*      *
F=  1240.5D+01      (KHZ)
*      *
*      *
***C****      C=  0.33311D-09      (FARADS)
*
*      *
****R****      R=  50.000E+00      (OHMS)

```

```
PRECISION TEST:  -4.804875E-14
```

TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>

****R****	R=	50.000E+00	(OHMS)
* * *			
* C	C=	0.51698D-09	(FARADS)
* *	(N)=	1	
* L	L=	0.14211D-05	(HENRYS)
LC**	C=	0.28663D-08	(FARADS)
* *	F=	2493.7D+00	(KHZ)
* * *			
* C	C=	0.36979D-09	(FARADS)
* *	(N)=	2	
* L	L=	0.19958D-05	(HENRYS)
LC**	C=	0.96448D-09	(FARADS)
* *	F=	3627.6D+00	(KHZ)
* * *			
* C	C=	0.67593D-09	(FARADS)
* * *			
****C****	C=	0.43553D-09	(FARADS)
* * *			
* * *	(N)=	1	
* * *	C=	0.78553D-10	(FARADS)
* C L	L=	0.99027D-06	(HENRYS)
* * *	F=	1804.5D+01	(KHZ)
* * *			
* * *			
****C****	C=	0.60889D-09	(FARADS)
* * *			
* * *	(N)=	2	
* * *	C=	0.23345D-09	(FARADS)
* C L	L=	0.70510D-06	(HENRYS)
* * *	F=	1240.5D+01	(KHZ)
* * *			
****C****	C=	0.33311D-09	(FARADS)
* * *			
****R****	R=	50.000E+00	(OHMS)

Figure 7(a). The bandpass network generated by cascading the highpass and lowpass networks.

tween the values of the central inductor calculated from both ports. Finally, the printing routines generate the output schematic connection diagram of the filter network.

Amstutz' method calculates more accurately the doubly periodic elliptic functions, especially the elliptic sine function. These are used to compute the poles and zeros of the elliptic filter transducer function H and characteristic function K . Then, the rational polynomial of the reflection coefficient in factored form is obtained. At the trap frequencies, the magnitude of the input reflection coefficient must be unity (i.e., complete reflection).

Amstutz ingeniously finds simple expressions for both input impedance and its derivative with respect to frequency or the group delay at the trap frequencies. Since $Z_{in} = sL_1 + M_1(s - s_1^2/s)$ at the trap frequency where M_1 is resonant, then $Z_{in} = sL_1$. Amstutz depends on the equivalence of any two networks having the same topology but permuted trap positions. He then permutes each trap to the input position and uses his recursive algorithm to yield all of the element values.

For more information on the above, as well as details of the synthesis techniques, refer to References 4 and 5.

Elliptic Filter Design Program

The elliptic filter design program ELFIL5 (Figure 9) begins by prompting for the passband and stopband frequencies in kilohertz. If the passband is lower than the stopband, a lowpass filter is designed. If the stopband is lower than the passband,

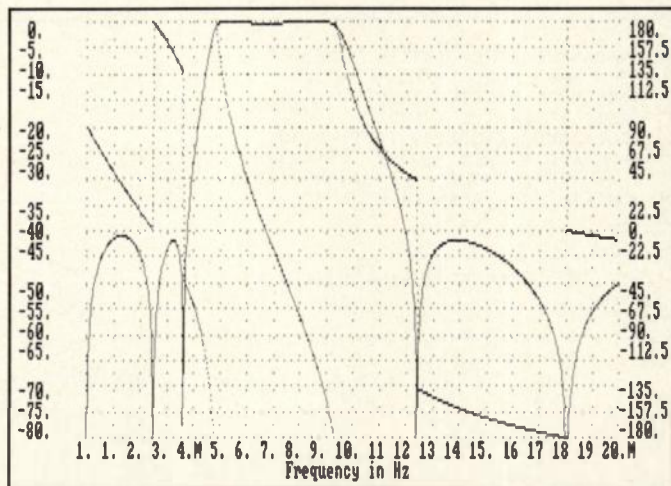


Figure 7(b). Highpass/lowpass cascade type bandpass filter.

a highpass filter is designed.

Then, the number of finite zeros n is requested. Note that the order of the filter $m = 2n + 1$. In the first part of the program, the elliptic functions are generated. This is the section where comparison to $1/E^8(S-P)$ is made (the addition formula). Next, the critical Q is calculated and printed out. This is the value for coil Q at which the first loss peak just does disappear. After input of data on stopband rejection, the passband ripple is calculated and printed out. Next, the 3 dB down frequency is calculated and printed out. The program prompts for the termination resistance value used to synthesize the component values for the filter. The component values are synthesized and the values printed out in such a form as to generate the component schematic connection diagram.

Another example program run, using the same input data as before, should yield results which agree with the values calculated by program ELT1. The results of the program run for the elliptic design program ELFIL5.EXE are shown in Figure 4. The values agree with the values calculated by program ELT1.EXE. A plot of this filter is given in Figure 3(b).

A second example shows the ability of the program ELFIL5 to design highpass filters. Again, a seventh-order filter is chosen with a passband frequency of 3.00 MHz, a stopband frequency of 2.5 MHz, and a passband ripple of 0.5 dB. Figure 5(a) presents the results of the ELFIL5 program run. As seen from the schematic connection diagram, the program has designed the seventh-order elliptic highpass filter requested. A plot of this filter is shown in Figure 5(b).

Highpass/Lowpass Cascade Type Bandpass Filter

A cascade connection of a highpass filter and a lowpass filter yields a bandpass filter. This technique can be used for wide and median bandwidth type bandpass filters. Consider the following case: A fifth-order highpass and fifth-order lowpass filter are to be designed and cascaded to yield a tenth-order bandpass filter. The lower band edge frequency is 5 MHz and the upper band edge frequency is 10 MHz, yielding a center frequency:

$$F_0 = \sqrt{F_1 F_2} = \sqrt{(5)(10)} = 7.071 \text{ MHz}$$

Choosing the upper stopband frequency as 12 MHz gives a stopband/passband ratio of $12/9 = 1.333333$. Since the

'This Program Computes the complete Elliptical integral and from this determines the loss pole frequencies (Transmission Zeros), the required order N to meet the specifications and the loss function for any specified frequency for equal ripple passband lowpass filters. This program is written in Microsoft QuickBasic by David C. Greene 29 November 1987. Copywrite 1987.

```

10 CLS
20 PRINT "Elliptical LowPass Filter Loss Program"
30 PRINT "David C. Greene, PHILIPS Consumer Electronics Co."
40 PRINT "Knoxville, TN ": PRINT : PRINT
50 PRINT "This Program Given PASSBAND Ripple, STOPBAND Loss"
60 PRINT "PASSBAND Frequency & STOPBAND Frequency, Computes"
70 PRINT "Transmission Zeros, Order (N) and Loss at Freq. (F)"
80 REM ----- 29 November 1987 -----
160 DEFINT D, H, L, R, U-X: DEFINT I
170 DIM F(5), DOO(30), DA(30), X(25), XZ(25)
190 DPI = 4 * ATN(14)
1050 REM -----
1060 PRINT : PRINT
1070 PRINT "ENTER PASSBAND RIPPLE (AMAX) & STOPBAND LOSS (AKIN) IN DB: INPUT AX, AN
1080 PRINT "ENTER PASSBAND FREQ. (FB), & STOPBAND FREQ. (FH) IN (HZ): INPUT FB, FH
1090 IF AX = 0 OR AN = 0 OR FB = 0 OR FH = 0 THEN 1100
1100 PRINT ".....Calculating Elliptical Functions....."
1110 HE = 10 ^ (AX / 10) - 1: HL = SQR(10 ^ (AN / 10) - 1) / HE: HK = FH / FB
1130 H = 1 / HL: GOSUB 5000
1140 KL = H4
1160 H = SQR(1 - 1 / HL / HL): GOSUB 5000
1170 LL = H4
1410 H = SQR(1 - 1 / HK / HK): GOSUB 5000
1420 H3 = H4
1430 H = 1 / HK: GOSUB 5000
1440 N = INT(LL / KL * H4 / H3) + 1
1450 PRINT : PRINT "DEGREE (N) REQUIRED IS : "; N
1460 NI = 0: RN = 1
1470 IF N = 2 * INT(N / 2) THEN NI = 1
1480 PRINT : PRINT "LOSS POLE FREQUENCIES ARE : "
1500 FOR I = 1 TO INT(N / 2): GOSUB 7000: NEXT I
1600 X = 1: GOSUB 9300
1700 C = 1 / RN
1800 PRINT "FREQUENCY (F) TO CALCULATE LOSS AT: "; INPUT F: GOSUB 6000
1810 PRINT "TO CONTINUE PRESS <ENTER> TO GO TO START PRESS <1> TO EXIT PRESS <0>"
1820 AS = INKEY$: IF LEN(AS) = 0 THEN 1860
1830 IF ASC(AS) = 13 GOTO 1800
1835 IF ASC(AS) = 49 GOTO 1060
1840 IF ASC(AS) = 81 GOTO 9900
1850 IF ASC(AS) = 113 GOTO 9900
1860 GOTO 1820
3000 REM ----- Calculate CSN(U,K) -----
3100 Q = EXP(-DPI * H3 / H4)
3200 V = DPI / 2 * U / H4: SNN = 0: J = 0
3300 W = Q ^ (J + .5): SNN = SNN + W * SIN((2 * J + 1) * V) / (1 - W * W): J = J + 1
3400 IF W > .0000001 GOTO 3300
3500 SNN = SNN * 2 * DPI / H / H4
3600 RETURN
5000 REM ----- Calculate H4 Complete Integral -----
5010 REM
5020 IF H = 1 THEN DA(0) = DPI / 2 ELSE DA(0) = ATN(H / SQR(1 - H ^ 2))
5030 DOO(0) = DPI / 2: HP = 1: I = 0
5040 DQ = DA(I): GOSUB 11000
5050 DV = 2 / (1 + DQ1) - 1
5060 DWN = DQ1
5070 DQ = DOO(I): GOSUB 11000
5080 DW = DWN * DQ1
5090 IF DV = 0 THEN DV = .0000001

```

```

5100 DA(I + 1) = ATN(SQR(1 - DV ^ 2) / DV)
5110 IF SQR(1 - DW ^ 2) = 0 THEN DOO(I + 1) = .5 * (DOO(I) + DPI / 2): GOTO 5130
5120 DOO(I + 1) = .5 * (DOO(I) + ATN(DW / SQR(1 - DW ^ 2)))
5130 DE = 1 - (DA(I + 1) * 2 / DPI): I = I + 1
5140 IF DE > .0000001 THEN 5040
5150 FOR J = 1 TO I
5160 DQ = DA(J)
5170 GOSUB 12000
5180 HP = HP * (1 + DQ1)
5190 NEXT J
5200 DV = DPI / 4 + DOO(I) / 2
5210 DQ = DV: GOSUB 11000
5220 DSN = DQ1: DQ = DV: GOSUB 12000
5230 DCN = DQ1: IF DCN < 0 THEN DCN = -1 * DCN
5240 DTN = DSN / DCN
5250 H4 = LOG(DTN) * HP
5260 RETURN
6000 REM ----- Calculate Loss -----
6100 X = F / FB
6200 RN = C * X ^ (1 - NI)
6300 GOSUB 9300
6400 A = 10 * LOG(1 + HE * RN * RN) / LOG(10)
6500 PRINT "DB LOSS= "; A
6600 RETURN
7000 REM ----- Calculate Poles and Zeros -----
7100 U = (2 * I - NI) * H4 / N: GOSUB 3000
7200 XZ(I) = SNN: X(I) = HK / SNN
7300 F(I) = FB * X(I): PRINT F(I)
7400 RETURN
9300 FOR I = 1 TO INT(N / 2)
9400 RN = RN * (X * X - XZ(I) ^ 2) / (X * X - X(I) ^ 2): NEXT I
9600 RETURN
9900 END
11000 REM ----- Series Expansion of SIN(X) -----
11010 DFACT = 1: SN = -1: DQ1 = DQ
11020 FOR K = 3 TO 15 STEP 2
11030 DFACT1 = K * (K - 1)
11040 DFACT = DFACT * DFACT1
11050 DE = 1
11060 FOR Q = 1 TO K
11070 DE = DE * DQ
11080 NEXT Q
11090 DQX1 = (DE / DFACT) * SN
11100 DQ1 = DQ1 + DQX1
11110 SN = -1 * SN
11120 NEXT K
11130 RETURN
12000 REM ----- Series Expansion of COS(X) -----
12010 DFACT = 1: SN = -1: DQ1 = 1
12020 FOR K = 2 TO 14 STEP 2
12030 DFACT1 = K * (K - 1)
12040 DFACT = DFACT * DFACT1
12050 DE = 1
12060 FOR Q = 1 TO K
12070 DE = DE * DQ
12080 NEXT Q
12090 DQX1 = (DE / DFACT) * SN
12100 DQ1 = DQ1 + DQX1
12110 SN = -1 * SN
12120 NEXT K
12130 RETURN

```

Figure 8. Elliptic lowpass filter loss program.

highpass filter is the reciprocal of the lowpass filter, the stopband frequency for the highpass filter is chosen as $1/1.333333$ or $(9/12)(5) = 3.75$ MHz. Running the synthesis program ELFIL5 yields the results given in Figure 6.

The program ELFIL5 has synthesized the highpass and lowpass networks shown, with the filter specifications which were supplied. The highpass network can be cascaded with the lowpass network to generate the bandpass network shown in Figure 7(a). A plot of this bandpass filter is given in Figure 7(b). When using this design technique, one must be careful not to put the highpass and lowpass sections too close together or distortion of the passband will result.

A final note about these programs is that with larger-order filters the data will scroll off the screen. It is also desirable to be able to have hardcopy of the data on a printer. Two utility programs are available to address these problems. The first is BACKSCRL, which saves data that scrolls off the screen. When activated the previous screens can be recalled. Used in conjunction with this program is another called SNIPPER, which when activated allows any portion of the viewed screen to be sent to a file or the printer. Both of these programs are of the terminate and stay resident type.

The elliptic filter design program ELFIL5 (Figure 9) appears

on page 80. The programs described in this article are available from the RF Design Software Service. See page 80 for details.

References

- David C. Greene, "A Design Program for Butterworth Lowpass Filters," *RF Design*, March 1989, pp. 43-45.
- David C. Greene, "Chebyshev Lowpass Filter Design," *RF Design 1990 Directory*, pp. 70-72.
- Richard W. Daniels, *Approximation Methods for Electronic Filter Design*, McGraw-Hill Book Co., 1974, pp. 51-85.
- Pierre Amstutz, "Elliptic Approximation and Elliptic Filter Design on Small Computers," *IEEE Trans. on Circuits and Systems*, Vol. CAS-25, No. 12, Dec., 1978, pp. 1001-1011.
- Thomas R. Cuthbert, Jr., *Circuit Design Using Personal Computers*, John Wiley & Sons, 1983, pp. 344-369.

About the Author

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"Symmetrical elliptical filter design program by Pierre Avestutz in his paper 'Elliptic Approximation and Elliptic Filter Design on Small Computers' published in 1972 Transactions on Circuits and Systems, Vol. CAS-25, No. 12 December 1978. Original Fortran program converted to Quickbasic and modified by David C. Greene, Philips Consumer Electronics Co., Knoxville, TN 37914

```
CLS
PRINT "SYMMETRICAL ELLIPTIC FILTER DESIGN PROGRAM BY P. AVESTUTZ, VOL. CAS-25, NO. 12, 12/78"
PRINT "Original Fortran Program converted and modified to this format by"
PRINT "DAVID C. GREENE, PHILIPS CONSUMER ELECTRONICS COMP., KNOXVILLE, TN 37914"
PRINT "This Program determines PASSBAND RIPPLE, 3dB DOWN FREQUENCY"
PRINT "LOSS POLES and SYNTHESIZES Component VALUES given"
PRINT "STOPBAND Freq., PASSBAND Freq., NO. of ZEROS and TERMINATION Impedances"
PRINT : PRINT
DEFINT B#,F#,M#,Z
DIM B(16), C(16), D(16), E(16), F(30)
IN = LOG(10) / 10: PI = 4 * ATN(1)
10 : PRINT "PASS BAND EDGE (REQ)= "; : INPUT FP
PRINT "STOP BAND EDGE (REQ)= "; : INPUT FS
IF ABS(FS - FP) <= 0 GOTO 10
PRINT "NUMBER OF FINITE ZEROS (1-15)= "; : INPUT N
IF N <= 0 GOTO 10
M = 2 * N + 1
PC = SQR(FS * FP)
R = PC + PC
FOR K = 1 TO 2
  S = FS + FP
  FOR J = 1 TO 6
    P = SQR(S * R)
    S = (S + R) / 2
    IF 18*06 * (S - P) < S GOTO 20
    R = P
  NEXT J
20 : IF K = 2 GOTO 30
  Q = M / S
  R = ABS(FS - FP)
NEXT K
30 : Q = Q * S
S = EXP(-PI / Q)
Y = S
PRINT "CRITICAL Q= "; Q / (4 * (1 - S) * S * N)
PRINT "STOP BAND REJECTION (DB)= "; : INPUT S
IF S <= 0 GOTO 10
S = EXP(S * IN / 2)
R = EXP(PI * Q)
P = (LOG(1 + (S * S - 1) / (R / 4 + 1 / R) ^ 2)) / IN
PRINT "PASSBAND RIPPLE (DB)= "; P
R = R / (2 * (S + SQR(S * S - 1)))
R = LOG(R + SQR(R * R + 1)) / (2 * Q)
R = SIN(R) / COS(R)
40 : W = R
PRINT "3 DB (REQ) ABOUT= "; FP + (FS - FP) / (1 + PC / (FP * R * R))
PRINT "NOMINAL OHM RESISTANCE= "; : INPUT R
IF R <= 0 GOTO 40
Z = Y * E(N) = W * W * W
FOR J = 1 TO N - 1
  F(J) = 1
NEXT J
K = 1
Calculation of Frequency of Zero
FOR J = 1 TO 1024
  F(K) = F(K) * (1 - Z) / (1 + Z)
  IF K <= M - 1 GOTO 50
  Z = Z * Y
  X = ((1 - Z) / (1 + Z)) ^ 2
  E(N) = E(N) * (M + X) / (1 + W * X)
  R = 0
50 : Z = Z * Y
IF Z < 2.5E-19 GOTO 60
' Comparison to zero for value
R = K + 1
NEXT J
60 : FOR J = 1 TO N
  F(J) = F(J) * F(M - J)
  F(M - J) = F(J)
NEXT J
FOR J = 1 TO N
  D(J) = F(2 * J) * (1 - F(J) ^ 4) / F(J)
  B(J) = E(N) * F(J)
NEXT J
C(1) = 1 / B(N)
FOR J = 1 TO N - 1
  C(J + 1) = (C(J) - B(N - J)) / (1 + C(J) * B(N - J))
  E(N - J) = E(N + 1 - J) + E(N) * D(J) / (1 + B(J) * B(J))
NEXT J
FOR J = 1 TO N
  B(J) = ((1 + C(J) * C(J)) * E(J) / D(J) - C(J) / F(J)) / 2
  C(J) = C(J) * F(J)
  D(J) = F(J) * F(J)
NEXT J
B(N + 1) = B(N) * C(N + 1) = C(N) * D(N + 1) = D(N)
IF N = 1 GOTO 100
L = 1
70 : FOR K = L + 2 TO N + 1 STEP 2
  FOR J = L TO K - 2 STEP 2
    Y = C(J) - C(K)
    Z = 1 / (Y / B(J) * (D(K) - D(J)) - 1)
    B(K) = (B(K) - B(J)) * Z * Z - B(J) * (1 + Z + Z)
    C(K) = Y * Z
  NEXT J
  NEXT K
IF L = 2 GOTO 80
L = 2: GOTO 70
80 : S = B(N) / B(N + 1) - 1
100 : Q = .0006 / (PI * PC)
P = Q * R: Q = Q / R
IF FS < FP GOTO 110
PRINT : PRINT : PRINT " ** LOW-PASS FILTER **": PRINT : PRINT
FOR J = 1 TO N
  C(J) = Q * C(J)
  D(J) = Q * B(J) * D(J)
  B(J) = P / B(J)
  F(J) = PC / F(J)
NEXT J
C(N + 1) = Q * C(N + 1)
GOTO 200
110 : PRINT : PRINT : PRINT " ** HIGH-PASS FILTER **": PRINT : PRINT
FOR J = 1 TO N
  C(J) = Q / C(J)
  D(J) = Q / (B(J) * D(J))
  B(J) = P * B(J)
  F(J) = PC * F(J)
NEXT J
C(N + 1) = Q / C(N + 1)
GOTO 200
```

```
150 : PRINT : PRINT "TO CONTINUE PRESS <ENTER> : TO EXIT PROGRAM PRESS <Q>"
160 : AS = INKEY$: IF LEN(AS) = 0 THEN 170
IF ASC(AS) = 13 THEN 10
IF ASC(AS) = 81 THEN 180
IF ASC(AS) = 113 THEN 180
170 : GOTO 160
180 : END
```

Lowpass Printing Routine

```
200 : PRINT " ***** R= "; : PRINT USING "000.000*****": TRES; : PRINT " (OHMS)"
FOR J = 1 TO N STEP 2
  PRINT " * "
  PRINT " ***** C= "; : PRINT USING "0.00000*****": C(J); : PRINT " (FARADS)"
  PRINT " * "
  PRINT " ***** B= "; : PRINT J
  PRINT " * "
  PRINT " ***** D= "; : PRINT USING "0.00000*****": D(J); : PRINT " (FARADS)"
  PRINT " * "
  PRINT " ***** L= "; : PRINT USING "0.00000*****": B(J); : PRINT " (HENRYS)"
  PRINT " * "
  PRINT " ***** F= "; : PRINT USING "00000.0*****": F(J); : PRINT " (HZ)"
  PRINT " * "
  PRINT " * "
  PRINT " ***** C= "; : PRINT USING "0.00000*****": C(N + 1); : PRINT " (FARADS)"
  IF N = 1 THEN GOTO 220
  L = (INT((N + 1) / 2)) * 2: K = M - 1 - L
  FOR J = L + 2 TO M - 1 STEP 2
    PRINT " * "
    PRINT " ***** N= "; : PRINT K
    PRINT " * "
    PRINT " ***** C= "; : PRINT USING "0.00000*****": D(K); : PRINT " (FARADS)"
    PRINT " * "
    PRINT " ***** C L= "; : PRINT USING "0.00000*****": B(K); : PRINT " (HENRYS)"
    PRINT " * "
    PRINT " ***** F= "; : PRINT USING "00000.0*****": F(K); : PRINT " (HZ)"
    PRINT " * "
    PRINT " * "
    PRINT " ***** C= "; : PRINT USING "0.00000*****": C(K); : PRINT " (FARADS)"
    K = K - 2
  NEXT J
220 : PRINT " * "
  PRINT " ***** R= "; : PRINT USING "000.000*****": TRES; : PRINT " (OHMS)"
PRINT : PRINT : PRINT "PRECISION TEST: "; S
GOTO 150
```

Highpass Printing Routine

```
300 :
PRINT " ***** R= "; : PRINT USING "000.000*****": TRES; : PRINT " (OHMS)"
FOR J = 1 TO N STEP 2
  PRINT " * "
  PRINT " ***** C= "; : PRINT USING "0.00000*****": C(J); : PRINT " (FARADS)"
  PRINT " * "
  PRINT " ***** N= "; : PRINT J
  PRINT " * "
  PRINT " ***** L= "; : PRINT USING "0.00000*****": B(J); : PRINT " (HENRYS)"
  PRINT " * "
  PRINT " ***** C= "; : PRINT USING "0.00000*****": D(J); : PRINT " (FARADS)"
  PRINT " * "
  PRINT " ***** F= "; : PRINT USING "00000.0*****": F(J); : PRINT " (HZ)"
  PRINT " * "
  PRINT " * "
  PRINT " ***** C= "; : PRINT USING "0.00000*****": C(N + 1); : PRINT " (FARADS)"
  IF N = 1 THEN 220
  L = (INT((N + 1) / 2)) * 2: K = M - 1 - L
  FOR J = L + 2 TO M - 1 STEP 2
    PRINT " * "
    PRINT " ***** N= "; : PRINT K
    PRINT " * "
    PRINT " ***** L= "; : PRINT USING "0.00000*****": B(K); : PRINT " (HENRYS)"
    PRINT " * "
    PRINT " ***** C= "; : PRINT USING "0.00000*****": D(K); : PRINT " (FARADS)"
    PRINT " * "
    PRINT " ***** F= "; : PRINT USING "00000.0*****": F(K); : PRINT " (HZ)"
    PRINT " * "
    PRINT " * "
    PRINT " ***** C= "; : PRINT USING "0.00000*****": C(K); : PRINT " (FARADS)"
    K = K - 2
  NEXT J
220 :
GOTO 220
```

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Figure 9. Elliptic filter design program.

Principles of Transistor Interactive Mapping

By Charles R. Ruelke
Motorola Portable Products Division

This article develops a method for graphically determining the input and output reflection coefficients for a transistor with an impedance shunting its input or output ports. A study of input and output stability circles for a given topology is also presented. When utilized in designing oscillators or amplifiers, this method provides an easy and effective means of calculating impedance values that can cause oscillations or induce stability.

The predisposition of a transistor under load to oscillate is the major consideration in oscillator design and is also an important consideration in amplifier applications. The tendency of a transistor to oscillate can be determined from the two-port S-parameter characterization of that transistor. In a two-port network, oscillations are possible when either the input or output port presents a negative resistance. This occurs when $|\Gamma_{in}| > 1$ or $|\Gamma_{out}| > 1$. For a unilateral device, this occurs when $|S_{11}| > 1$ and $|S_{22}| > 1$. Note that Γ = reflection coefficient = $(Z - Z_0)/(Z + Z_0)$. If the two-port is not unconditionally stable, it is potentially unstable and can produce a negative resistance for some passive load. The two-port is said to be unconditionally stable if the following conditions are met:

$$K = \{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2\} / \{2|S_{12}S_{21}|\} > 1$$

$$\text{and } |\Delta| < 1$$

$$\text{where } |\Delta| = |S_{11}S_{22} - S_{12}S_{21}|.$$

From a graphical point of view, the region of intersection between the Smith Chart and the two stability circles (input and output) is used to define those impedances which will induce oscillations (instability) or cause stability. These stability circles are generated from equations which will be examined later. Graphically, the condition for stability is that the input and output stability circles either fall completely outside of or completely enclose the Smith Chart (for $|S_{11}| < 1$ and $|S_{22}| < 1$). For a stability circle having a center at $|C| \angle \theta_c^\circ$ (polar form) with radius r , the condition for unconditional stability for passive loads is expressed in the form:

$$||C| - r| > 1 \quad \text{for } |S_{11}| < 1 \text{ and } |S_{22}| < 1$$

If the two-port is potentially unstable, either the input or output stability circle (or both) will intersect the border of the Smith Chart.

Another method for graphically determining the effects of an impedance on transistor stability is to map impedance circles onto the Smith Chart using the input and output reflection coefficient equations. Those impedance values lying outside of the Smith Chart ($|\Gamma| > 1$) will cause the two-port to oscillate. This method also gives the designer a feel for the "margin" of stability or instability he possesses for a given load, since the proximity of an impedance to the border of the Smith Chart becomes readily apparent.

Interactive Mapping of r and jx Into the Γ_{in}/Γ_{out} Plane

The Smith Chart is a graphical representation of the reflection coefficient plane, called the Γ plane, where:

$$\Gamma = (Z - Z_0)/(Z + Z_0) = (z - 1)/(z + 1)$$

where $z = Z/Z_0$ is the normalized value of Z .

Derivation of the Smith Chart is facilitated by setting $\Gamma = U + jV$ and $z = r + jx$ and substituting into the preceding equation, which yields:

$$\Gamma = U + jV = \frac{z - 1}{z + 1} = 1 - \frac{2}{z + 1} = 1 - \frac{2}{r + jx + 1}$$

Further manipulation gives the following equation:

$$r + 1 + jx = \frac{-2(U - 1 - jV)}{(U - 1)^2 + V^2} \quad (1)$$

The family of circles representing the resistance (real) part of z is derived from the real part of equation 1. Collecting the real part of equation 1 to one side of the equals sign, dividing through by $r + 1$, and completing the square for $U - 1$ yields the equation for the family of resistive circles as defined by:

$$\left(U - \frac{r}{r + 1}\right)^2 + V^2 = \left(\frac{1}{r + 1}\right)^2 \quad (2)$$

The family of circles representing the reactance (imaginary) part of z is derived from the imaginary part of equation 1. Collecting the imaginary part of equation 1 to one side of the equals sign, dividing through by x , and completing the square for V gives the equation for the family of inductive/capacitive circles as defined by:

$$(U - 1)^2 + \left(V - \frac{1}{x}\right)^2 = \left(\frac{1}{x}\right)^2 \quad (3)$$

Plotting the family of circles defined by equations 2 and 3 onto the same coordinate system and constraining the limits of the plot to $r \geq 0$ (i.e., $|\Gamma| \leq 1$) defines the Smith Chart. The vector from the origin (center) of the Smith Chart to a given impedance (as defined by the intersection of the corresponding real and imaginary circles) is equal to $\Gamma = |\Gamma| \angle \arg(\Gamma)$ for that impedance.

The effects of source and load impedances on a two-port network establish a new set of resistive/reactive circles in the Γ plane. The input and output reflection coefficients (Γ_{in} and Γ_{out} , respectively) for a given two-port terminated by source (Γ_s) and load (Γ_L) impedance are defined using two-port S-parameters by:

rf design guide

$$\Gamma_{in} = S_{11} + \frac{S_{12} S_{21} \Gamma_L}{1 - S_{22} \Gamma_L} = S_{11} + \frac{S_{12} S_{21}}{1/\Gamma_L - S_{22}} \quad (4)$$

where $\Gamma_L = (z - 1)/(z + 1)$

$$\Gamma_{out} = S_{22} + \frac{S_{12} S_{21} \Gamma_s}{1 - S_{11} \Gamma_s} \quad (5)$$

For demonstration purposes, the family of circles relating a load impedance of $r + jx$ to Γ_{in} will now be derived. Examining $1/(1/\Gamma_L - S_{22})$ of equation 4, it is evident that:

$$\begin{aligned} \frac{1}{1/\Gamma_L - S_{22}} &= \frac{z - 1}{(1 - S_{22})z + 1 + S_{22}} \\ &= \frac{1}{1 - S_{22}} - \frac{1 + (1 + S_{22})/(1 - S_{22})}{(1 - S_{22})z + 1 + S_{22}} \end{aligned}$$

Since $1 = (1 - S_{22})/(1 - S_{22})$, the above equation becomes:

$$\begin{aligned} \frac{1}{1/\Gamma_L - S_{22}} &= \frac{1}{1 - S_{22}} - \frac{2/(1 - S_{22})}{(1 - S_{22})z + 1 + S_{22}} \\ &= \frac{1}{(1 - S_{22})} \left(1 - \frac{2}{(1 - S_{22})z + 1 + S_{22}} \right) \end{aligned}$$

Dividing the z term by $(1 - S_{22})$ and noting that $(1 + S_{22})/(1 - S_{22}) = 1 + 2[S_{22}/(1 - S_{22})]$, equation 4 can be written as:

$$\Gamma_{in} = S_{11} + \left(\frac{S_{12} S_{21}}{1 - S_{22}} \right) \left(1 - \frac{2/(1 - S_{22})}{z + 1 + 2[S_{22}/(1 - S_{22})]} \right) \quad (6)$$

For reference purposes, it is helpful to define the following terms:

$$A = (S_{12} S_{21})/(1 - S_{22}) \quad (7)$$

$$B = z + 1 + 2[S_{22}/(1 - S_{22})] \quad \text{where } z = r + jx \quad (8)$$

Then:

$$\Gamma_{in} = U + jV = S_{11} + A \left(1 - \frac{2/(1 - S_{22})}{B} \right)$$

which can be written as:

$$B = \frac{-2 [A/(1 - S_{22})] [(\sim U) - j(\sim V)]}{(\sim U)^2 + (\sim V)^2} \quad (9)$$

where $(\sim U) = U - \text{Re}\{S_{11} + A\}$ and $(\sim V) = V - \text{Im}\{S_{11} + A\}$.

For reference purposes, equation 9 can be written more simply as follows:

$$B = -2C[(\sim U) - j(\sim V)]/[(\sim U)^2 + (\sim V)^2] \quad (10)$$

where $C = A/(1 - S_{22})$

At this point, it is intuitively apparent that equation 10 for the Γ_{in} chart is similar in structure to equation 1, which was used in deriving the Smith Chart. The symmetry is even more apparent

when it is observed that in order for Γ_{in} to equal Γ_L in equation 4, set $S_{11} = S_{22} = 0$ and $S_{12} = 1/S_{21}$. Upon substituting these values for the S-parameters into A, B and C, equations 10 and 1 become identical. Therefore, one can conclude that the chart resulting from equation 10 will be similar to the Smith Chart, having a family of resistive and reactive circles bounded by the $r = 0$ circle. The family of resistance circles for Γ_{in} are found by examining the real part of equation 10. Collecting the real part of equation 10 to one side of the equals sign and dividing through by $\text{Re}\{B\}$ gives:

$$(\sim U)^2 + 2(\sim U) \frac{\text{Re}\{C\}}{\text{Re}\{B\}} + (\sim V)^2 + 2(\sim V) \frac{\text{Im}\{C\}}{\text{Re}\{B\}} = 0$$

Upon completing the squares for $\sim U$ and $\sim V$ the equation becomes:

$$\left(\sim U + \frac{\text{Re}\{C\}}{\text{Re}\{B\}} \right)^2 + \left(\sim V + \frac{\text{Im}\{C\}}{\text{Re}\{B\}} \right)^2 = \left(\frac{|C|}{\text{Re}\{B\}} \right)^2 \quad (11)$$

Equation 11 is the equation of a circle centered at:

$$\left(\text{Re}\{S_{11} + A\} - \frac{\text{Re}\{A/(1 - S_{22})\}}{\text{Re}\{B\}} \right), \left(\text{Im}\{S_{11} + A\} - \frac{\text{Im}\{A/(1 - S_{22})\}}{\text{Re}\{B\}} \right)$$

The radius is:

$$\frac{|A/(1 - S_{22})|}{\text{Re}\{B\}}$$

where $\text{Re}\{B\} = r + 1 + 2\text{Re}\{S_{22}/(1 - S_{22})\}$.

By inspection, it is evident that the radii of the resistive circles for the Γ_{in} plot are inversely proportional to r . The centers of the resistive circles are offset from the center of the Smith Chart by a constant factor of $S_{11} + A$ and by a term inversely proportional to r . For the limiting case of $\text{Re}\{B\} \gg |A/(1 - S_{22})|$ (i.e., $r \rightarrow \infty$ or $\Gamma_L \rightarrow 1$) the resistive circle becomes a point located at $S_{11} + A = S_{11} + (S_{12} S_{21})/(1 - S_{22})$. This agrees with the results obtained by substituting $|\Gamma_L| = 1$ into equation 4. The other limiting case of $r = 0$ defines the boundary for the interactive mapping which intersects the boundary of the Smith Chart for potentially unstable two-port networks. The reactive circles are derived from the imaginary part of equation 10. Collecting the imaginary part of equation 10 to one side of the equals sign and dividing through by $\text{Im}\{B\}$ yields:

$$(\sim U)^2 + \frac{2(\sim U) \text{Im}\{C\}}{\text{Im}\{B\}} + (\sim V)^2 - \frac{2(\sim V) \text{Re}\{C\}}{\text{Im}\{B\}} = 0$$

After completing the squares for $\sim U$ and $\sim V$, the above equation assumes the proper form:

$$\left(\sim U + \frac{\text{Im}\{C\}}{\text{Im}\{B\}} \right)^2 + \left(\sim V - \frac{\text{Re}\{C\}}{\text{Im}\{B\}} \right)^2 = \left(\frac{|C|}{\text{Im}\{B\}} \right)^2 \quad (12)$$

Equation 12 is an equation of a circle centered at:

$$\left(\text{Re}\{S_{11} + A\} - \frac{\text{Im}\{A/(1 - S_{22})\}}{\text{Im}\{B\}} \right), \left(\text{Im}\{S_{11} + A\} + \frac{\text{Re}\{A/(1 - S_{22})\}}{\text{Im}\{B\}} \right)$$

having a radius of:

$$\frac{|A/(1 - S_{22})|}{\text{Im}\{B\}}$$

where $\text{Im}\{B\} = x + 2\text{Im}\{S_{22}/(1 - S_{22})\}$.

Again note that the centers of the reactive circles are inversely proportional to x . Since x can assume either a positive or negative value, two reactive circles, representing the same magnitude but possessing different signs, are present. These properties parallel the characteristics of the Smith Chart. However, significant differences are also evident between the symmetry of the reactive circles for Γ_{in} and the Smith Chart. For example, $x = 0$ ($j0$) defines a circle for the Γ_{in} plot, whereas $j0$ for the Smith Chart is represented by a line (or a circle with infinite radius). Also, for the Γ_{in} plot, reactance circles of the same magnitude possess different radii. This is readily apparent in that if x is of the same sign as $\text{Im}\{S_{22}/(1 - S_{22})\}$, then $\text{Im}\{B\}$ is increased, resulting in a smaller radius.

An identical method can be employed in mapping the effects of source impedances on Γ_{out} . The advantage of inscribing the Γ plot onto the Smith Chart is that one can graphically determine the input and/or output reflection coefficient for a particular load. This is accomplished by measuring the distance and angle from the center of the Smith Chart to the point in the Γ plot representing the impedance of interest (defined by the intersection of the appropriate resistive and reactive circles). Thus, at a glance a designer can find how much margin he may have in altering a particular impedance before the two-port becomes stable or begins to oscillate. Exact values for the reflection coefficient can be calculated using equations 4 and 5 once a "ballpark" impedance value is chosen using the graphical analysis.

Figure 1 is an example of an output reflection plot (Γ_{out} where z is located at the input) for a two-port network possessing two-port S-parameters (measured at 100 MHz):

$$\begin{aligned} S_{11} &= 0.47 \angle -144^\circ \\ S_{12} &= 0.035 \angle 56^\circ \end{aligned}$$

$$\begin{aligned} S_{21} &= 19.34 \angle 107^\circ \\ S_{22} &= 0.42 \angle -100^\circ \end{aligned}$$

From Figure 1, the output reflection coefficient can be determined by measuring the magnitude and angle from the center of the Smith Chart to the impedance of interest. For example, by inspection, $5 + j10$ (point A in Figure 1) is located at about $1.1 \angle -52^\circ$ from the center of the Smith Chart. This indicates that for this particular load the transistor is unstable. This result can be verified by using equation 5 to calculate the exact output reflection coefficient for $Z = 5 + j10$ ($\Gamma_s = 0.824 \angle 157.166^\circ$, $Z_0 = 50$ ohms) which is:

$$\Gamma_{out} = S_{22} + S_{12}S_{21}/[(1/\Gamma_s) - S_{11}] = 1.114 \angle -52.26^\circ$$

A simpler and quicker way to check the results of the output reflection plot is to set $Z = 50 + j0$ ohms (point B in Figure 1) which makes $\Gamma_s = 0$. For $\Gamma_s = 0$, the overall reflection coefficient becomes $\Gamma_{out} = S_{22} = 0.42 \angle -100^\circ$. It is also interesting to note that the $R = 15$ ohms circle tangentially intersects the border of the Smith Chart. This indicates that a 15 ohm load placed in series (since reflection plot is an impedance representation) to the input port will guarantee stability for the transmitter's output port. Note that the $x = j0$ line has been technically augmented in Figure 1.

Interactive Mapping of g and jb Into the Γ Plane

Interactive mapping of the admittance circles can also be accomplished using equation 4 and noting:

$$\Gamma_L = \frac{z - 1}{z + 1} = \frac{(1/y) - 1}{(1/y) + 1} = \frac{1 - y}{1 + y}$$

where $y = Y/Y_0 = g + jb$.

Substituting the above equation into equation 4, and after some manipulation, the following fundamental equation is obtained:

$$B' = 2C'[(\sim U') - j(\sim V')]/[(\sim U')^2 + (\sim V')^2] \quad (13)$$

where, $B' = y + 1 - 2[S_{22}/(1 + S_{22})]$

$$C' = A'/(1 + S_{22})$$

$$A' = S_{12}S_{21}/(1 + S_{22})$$

$$(\sim U') = U - \text{Re}\{S_{11} - A'\}$$

$$(\sim V') = V - \text{Im}\{S_{11} - A'\}$$

Collecting the real portion of equation 13 to one side of the equals sign and completing the square for $(\sim U')$ and $(\sim V')$ gives the equation for the conductance circles as follows:

$$\left(\sim U' - \frac{\text{Re}\{C'\}}{\text{Re}\{B'\}}\right)^2 + \left(\sim V' - \frac{\text{Im}\{C'\}}{\text{Re}\{B'\}}\right)^2 = \left(\frac{|C'|}{\text{Re}\{B'\}}\right)^2 \quad (14)$$

Equation 14 is the equation of a circle centered at:

$$\left(\frac{\text{Re}\{S_{11} - A'\} + \frac{\text{Re}\{A'/(1 + S_{22})\}}{\text{Re}\{B'\}}\right) + j\left(\frac{\text{Im}\{S_{11} - A'\} + \frac{\text{Im}\{A'/(1 + S_{22})\}}{\text{Re}\{B'\}}\right)$$

and having a radius of:

$$\frac{|A'/(1 + S_{22})|}{\text{Re}\{B'\}}$$

where $\text{Re}\{B'\} = g + 1 - 2\text{Re}\{S_{22}/(1 + S_{22})\}$.

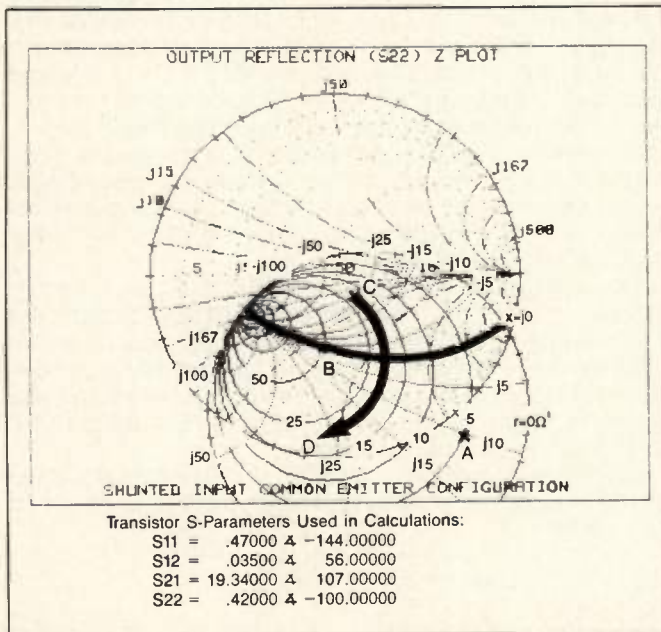


Figure 1. Shunted input common emitter configuration.

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For the limiting case of $g \rightarrow \infty$ (i.e., $\Gamma = -1$), the conductance circles become a point located at $\text{Re}\{S_{11} - A'\} + j\text{Im}\{S_{11} - A'\}$ which is equal to $S_{11} - (S_{12}S_{21})/(1 + S_{22})$. This agrees with substituting $\Gamma_L(y \rightarrow \infty) = (1 - y)/(1 + y) = -1$ into equation 4. It is also important to note that this point corresponds to $z = 0 + j0$ for the impedance interactive map. For $z = 0 + j0$, the resistive and reactive circles intersect at $S_{11} - (S_{12}S_{21})/(1 + S_{22})$, since $\Gamma_L(z = 0) = (z - 1)/(z + 1) = -1$. The other limiting case of $g = 0$ forms the boundary of the admittance in interactive mapping which intersects the Smith Chart for potentially unstable two-ports. Note that for $g = 0$, from equation 4:

$$\begin{aligned}\Gamma(g = 0) &= S_{11} + \frac{S_{12}S_{21}}{(1 + jb)/(1 - jb) - S_{22}} \\ &= S_{11} + \frac{S_{12}S_{21}}{(-jx + 1)/(-jx - 1) - S_{22}} \\ &= \Gamma(r = 0 \text{ \& } jx = -1/jb)\end{aligned}$$

Therefore, since the admittance representation for jx is $-1/jb$, it is evident that the $g = 0$ circle will possess the same radius and center as the impedance circle defined by $r = 0$. However, for g not equal to 0, the conductance circles and the reactive circles for $g = r$ will not have the same radii or center. This is intuitively apparent by comparing equations 11 and 14.

The susceptance circles are generated from the imaginary portion of equation 13. Collecting the imaginary part of equation 13 to one side of the equals sign and completing the square for $(\sim U')$ and $(\sim V')$ yields the following equation:

$$\left(\sim U' - \frac{\text{Im}\{C'\}}{\text{Im}\{B'\}}\right)^2 + \left(\sim V' + \frac{\text{Re}\{C'\}}{\text{Im}\{B'\}}\right)^2 = \left(\frac{|C'|}{\text{Im}\{B'\}}\right)^2 \quad (15)$$

Equation 15 is the equation of a circle centered at:

$$\left(\text{Re}\{S_{11} - A'\} + \frac{\text{Im}\{A'/(1 + S_{22})\}}{\text{Im}\{B'\}}\right) \left(\text{Im}\{S_{11} - A'\} - \frac{\text{Re}\{A'/(1 + S_{22})\}}{\text{Im}\{B'\}}\right)$$

and having a radius of:

$$\frac{|A'/(1 + S_{22})|}{\text{Im}\{B'\}}$$

where $\text{Im}\{B'\} = b - 2\text{Im}\{S_{22}/(1 + S_{22})\}$.

The susceptance circle for $b = 0$ possesses the same center and radius as the $x = 0$ circle as defined in equation 12. This can be proven by noting that:

$$\begin{aligned}\Gamma(b = 0) &= S_{11} + \frac{S_{12}S_{21}}{(1 + g)/(1 - g) - S_{22}} \\ &= S_{11} + \frac{S_{12}S_{21}}{(r + 1)/(r - 1) - S_{22}} \\ &= \Gamma(x = 0 \text{ \& } r = 1/g)\end{aligned}$$

It is also interesting to note that the admittance representation of the standard impedance Smith Chart can be generated using equations 14 and 15 for $S_{11} = S_{22} = 0$ and $S_{12} = 1/S_{21}$.

Stability Circles Using Two-Port S-Parameters

Sometimes the designer is not necessarily interested in the actual value of a given reflection coefficient, but only in which impedance value will induce oscillations. For this case, a much simpler procedure can be employed to inscribe stability circles onto the Smith Chart. In terms of reflection coefficients, the conditions for stability at a given frequency are:

$$|\Gamma_{in}| = |S_{11} + (S_{12}S_{21})/[\Gamma_L - S_{22}]| < 1 \quad (16)$$

$$|\Gamma_{out}| = |S_{22} + (S_{12}S_{21})/[\Gamma_s - S_{11}]| < 1 \quad (17)$$

The solutions of equations 16 and 17 give the conditions required for the two-port to be stable. Setting $|\Gamma_{in}| = 1$ and $|\Gamma_{out}| = 1$ and solving for Γ_s or Γ_L defines the two stability circles. It has been shown that the stability circle equations are given by reference 1:

$$\text{for input, } |\Gamma_{in}| = 1; \quad |\Gamma_L - (S_{22} - \Delta S_{11}^*)/\delta_1| = |S_{12}S_{21}/\delta_1| \quad (18)$$

$$\text{for output, } |\Gamma_{out}| = 1; \quad |\Gamma_s - (S_{11} - \Delta S_{22}^*)/\delta_2| = |S_{12}S_{21}/\delta_2| \quad (19)$$

where $\Delta = S_{11}S_{22} - S_{12}S_{21}$

$$\delta_1 = |S_{22}|^2 - |\Delta|^2$$

$$\delta_2 = |S_{11}|^2 - |\Delta|^2$$

and

$$S_{bb}^* = \text{Conjugate}(S_{bb})$$

$$(S_{ee} - \Delta S_{bb}^*)^* = \text{Conjugate}(S_{ee} - \Delta S_{bb}^*)$$

From equations 18 and 19, the radii and centers of the stability circles are obtained:

$$\text{Radius for } \Gamma_L \text{ values} = |S_{12}S_{21}/\delta_1|$$

$$\text{Center for } \Gamma_L \text{ values} = (S_{22} - \Delta S_{11}^*)/\delta_1$$

$$\text{Radius for } \Gamma_s = |S_{12}S_{21}/\delta_2|$$

$$\text{Center for } \Gamma_s = (S_{11} - \Delta S_{22}^*)/\delta_2$$

To determine which region of the stability circle (inside or outside) is stable, simply calculate Γ_{in} (or Γ_{out}) for any arbitrary impedance Γ_s (or Γ_L) and observe where the load is located with respect to the stability circle. For example, if $|\Gamma_{in}| > 1$ for a given load and the load falls within the stability circle, all impedances encompassed by that stability circle define $|\Gamma_{in}| > 1$. Therefore, all impedances falling outside of the stability circle will be $|\Gamma_{in}| < 1$. Similarly, if $|\Gamma_{in}| > 1$ and the impedance is located outside of the stability circle, then all impedances falling outside of the circle define $|\Gamma_{in}| > 1$, while $|\Gamma_{in}| < 1$ for those loads inside of the stability circle.

For convenience, the center of the Smith Chart is usually chosen, since $z = 1 \pm j0$ sets $\Gamma_{in} = S_{11}$ or $\Gamma_{out} = S_{22}$ ($\Gamma_L = 0$ or $\Gamma_s = \infty$, respectively). Therefore, a simple glance at the appropriate two-port S-parameter for a given topology, and the observation of where the center of the Smith Chart is located with respect to a given stability circle, defines which region of the Smith Chart is stable or unstable.

Figure 2 is an example of a stability circle corresponding to the output port of a transistor possessing the following two-port S-parameters:

$$S_{11} = 0.47 \angle -144^\circ$$

$$S_{12} = 0.035 \angle 56^\circ$$

$$S_{21} = 19.34 \angle 107^\circ$$

$$S_{22} = 0.42 \angle -100^\circ$$

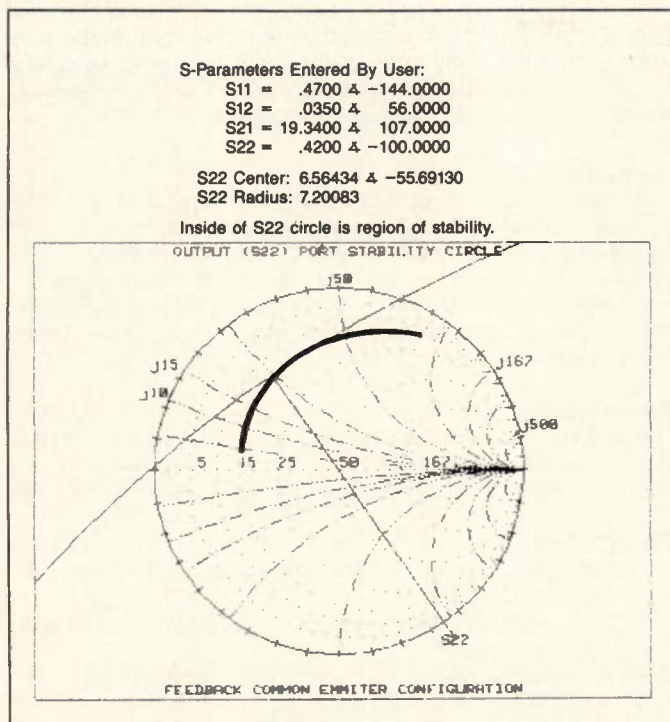


Figure 2. Feedback common emitter configuration.

The center and radius of the stability circle is calculated from equation 17 and the results are as follows:

Center: 6.564 $\angle$ -55.691°
 Radius: 7.201

Note that it is engineering convention to refer to the stability circle generated by looking into the output port (magnitude $\Gamma_{out} = 1$) as the input plane or Γ 's plane stability circle. To maintain compatibility with the interactive mapping already discussed, this article refers to the magnitude $\Gamma_{out} = 1$ stability circle as the output port stability circle.

Since the radius is larger than the magnitude representation of the stability circle center, the Smith Chart center is encompassed by the stability circle. Therefore, the inside portion of the stability circle is the stable region, since $|S22| < 1$. Note also that the Smith Chart's $R = 15$ ohms ($r = 0.3$, $Z_o = 50$ ohms) circle tangentially intersects the stability circle and is included in the stable region (see darkened portion). Therefore, a series 15 ohms at the input will guarantee stability at the output port. This agrees with the output reflection plot illustration shown in Figure 1.

Illustrated Examples

Example 1. A transistor possesses the following common emitter two-port S-parameters at 100 MHz:

S11 = 0.47 $\angle$ -144°
 S12 = 0.035 $\angle$ 56°
 S21 = 19.34 $\angle$ 107°
 S22 = 0.42 $\angle$ -100°

The designer wishes to determine the magnitude and configuration of a resistive load that will ensure unconditional stability for the output port of the transistor. The previous discussion

has shown that a series 15 ohm load at the input will guarantee stability. Figure 3 is used to determine the magnitude of the parallel load which will ensure stability.

Figure 3 is the admittance representation of Figure 1, with the stability circle shown in Figure 2 superimposed. The admittance reflection plot shows the $Y = 1/5$ circle ($y = 10$, $Y_o = 1/50$) tangentially intersecting the Smith Chart, which indicates that a 5 ohm load in parallel with the input port should guarantee stability at the output port. This agrees with the stability circle, which tangentially encompasses the 1/5 mho circle of the admittance Smith Chart.

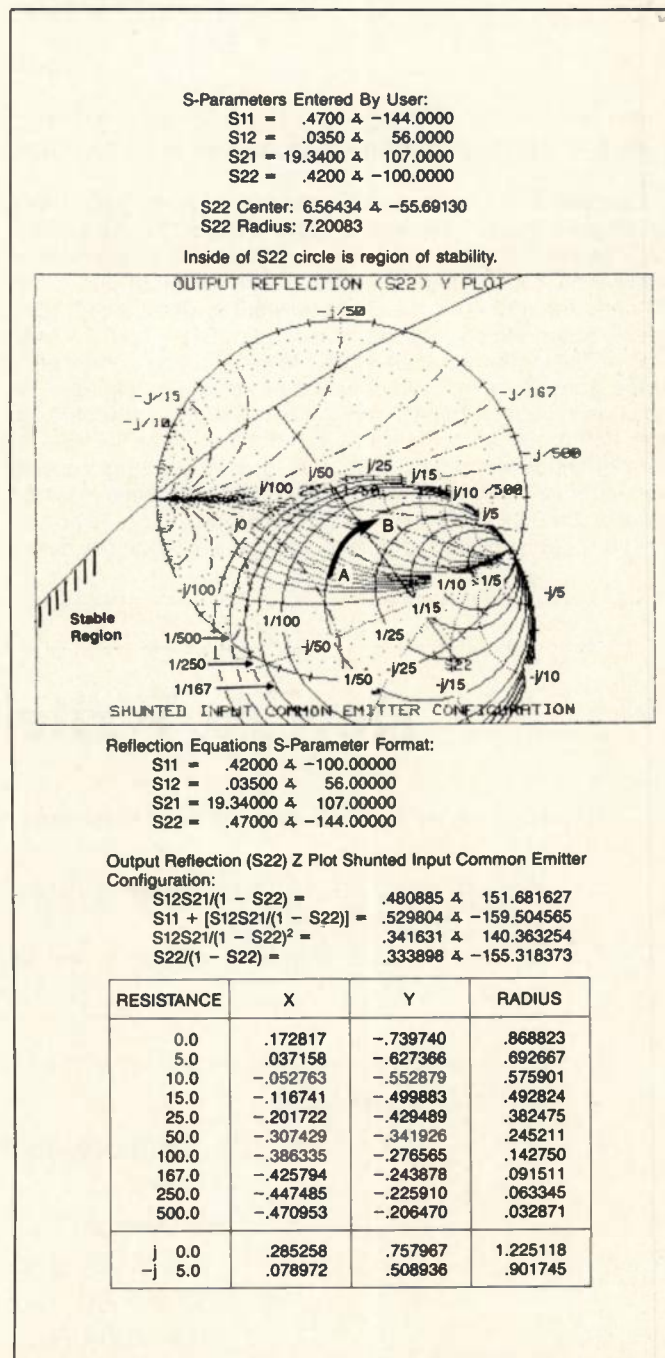


Figure 3. Shunted input common emitter configuration.

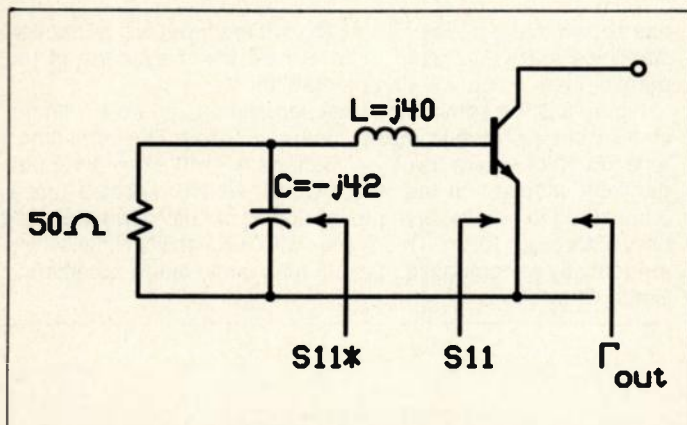


Figure 4. 50 ohm antenna matched to the transistor.

Example 2. A transistor in Example 1 is used to determine an amplifier stage in a receiver. The input of the transistor is conjugately matched for maximum transducer power gain. A passive network is then utilized to match the input port to the 50 ohm antenna. The matching network is determined using the reflection plots in Figures 1 and 3. For $S_{11} = 0.47 \angle -144^\circ$, $S_{11}^* = 0.47 \angle 144^\circ$ ($z = 0.4 + j0.3$, $Z_0 = 50$ ohms) corresponds to the transistor's input impedance of approximately $20 + j15$ which is located at point D in Figure 1. Starting at 50 ohms or $Y = 1/50$ (point A on Figure 3), the addition of $\Delta jB = +j/42$ of parallel capacitance moves point A to point B along the constant conductance circle of $G = 1/50$ ohms to achieve $Y = 1/50 + j/42$ or $Z = 20 - j25$ ($z = 0.4 - j0.5$ normalized to $Z_0 = 50$ ohms). $Z = 20 - j25$ is located at point C of Figure 1. Adding a series

inductance of $+j40$ moves point C to $S_{11}^* = 20 + j15$ ohms (point D on Figure 1). Note that the reflection coefficient at the transistor's output for any node of the input matching network is readily available simply by viewing Figure 1 or 3. The finalized matching network is illustrated in Figure 4.

Conclusion

A graphical method of determining the input and output reflection coefficients for a transistor with various source or load impedances was developed herein. This methodology can be effectively applied in the design of oscillator and amplifier circuits and has been shown to give accurate results. Therefore, the utilization of the interactive mapping technique can be an effective design tool.

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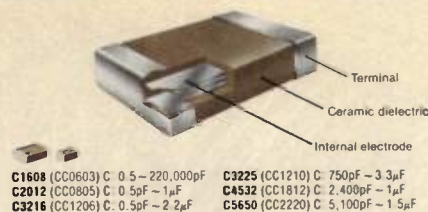
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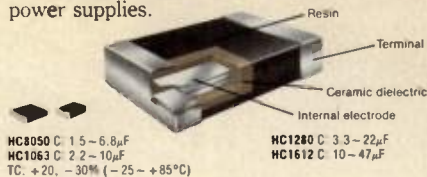
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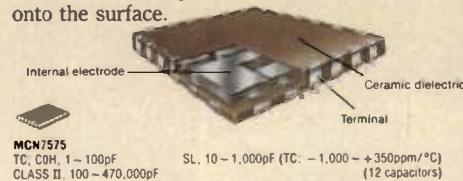
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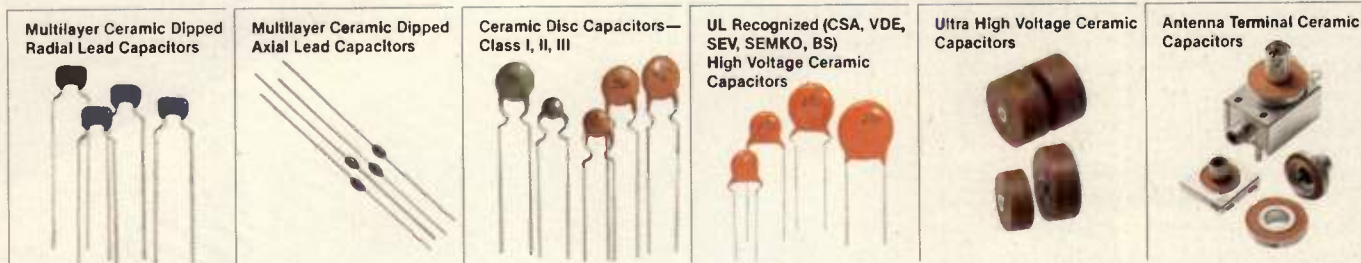


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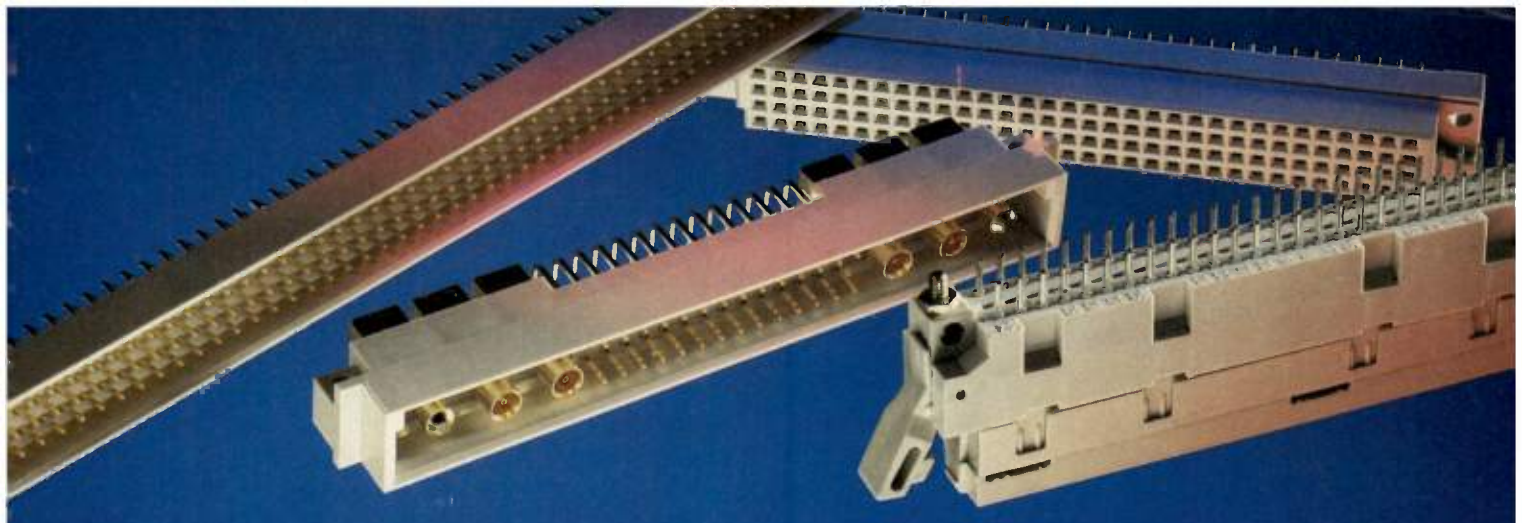
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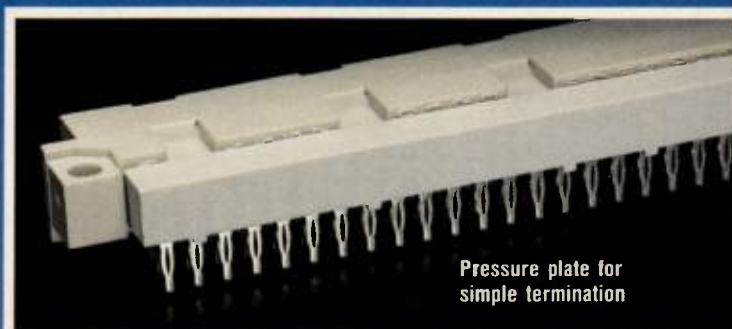
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MOLDED RADIAL 1pF-3.3μF

C052 (CK05) (CKR05) C062 (CK06) (CKR06) C056 (CKR05) C066 (CKR06)

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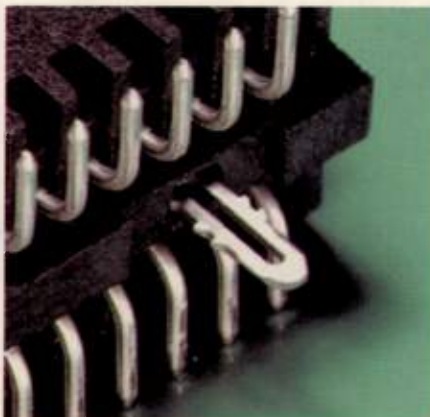
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Advertiser's Index

Acrian Inc.	2
Advanced Control Components	18
Advantest America, Inc.	63
AEG Corporation	127
A.H. Systems	33
Amphenol	47
AMP Incorporated	45
Amplifier Research	54
Applied Engineering Products	46
Aviel Electronics	46
Avtech	60
Aydin-Vector	16
Besser Associates	65
Brightline	26
Cal Crystal Lab., Inc.	24
Coaxial Dynamics, Inc.	60
Coilcraft	12
Colorado Crystal Corp.	11
Comtran Integrated Software	50
Daico Industries, Inc.	5
EMCO	67
ENI	128
Fair-Rite Products Corp.	51
Hewlett-Packard	61,123-124
High Voltage Components, Inc.	10
Hughes Aircraft/Microwave Division	8,31,35,58
Hybrids International	18
IFR Systems Inc.	3
ingSOFT Limited	49
Innovative Frequency Control Products, Inc.	59
Instruments for Industry	53
Integrated Microwave	19
International Crystal Manufacturing Co., Inc.	11
Intusoft	48
John Fluke Manufacturing Co. Inc.	56-57
Kay Elemetrics Corp.	36,125-126
KVG	26
M/A-COM Semiconductor Division	13
Matrix Systems Corporation	37,45
McCoy Electronics Company	25
MCL, Inc.	55
Noise Com, Inc.	23
Penstock Engineering Inc.	10
Programmed Test Sources, Inc.	7
Repco Incorporated	32
RF Power Systems	45
Sciteq Electronics, Inc.	31
Southwest Microwave	46
Sprague-Goodman Electronics, Inc.	9
Stanford Telecommunications	15,28
Surcom Associates, Inc.	9
Tecdia	9
Teledyne Microwave	48
Telonic/Berkeley	38,39,41
Tesoft	48
Trilithic, Inc.	20
TTE, Inc.	40
UTE Microwave, Inc.	42
Wavetek	21
Werlatone Inc.	44
Wide Band Engineering Company, Inc.	58
W.L. Gore & Associates, Inc.	36
XTALONIX	53

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2	U 862BS	2.4	40	250mV	TO 50
	U 822BS	2.1	31	250mV	TO 50
	U 2620BS	2.1	30	250mV	TO 50 Ceramic
4	U 864BS	2.4	45	250mV	TO 50
	U 824BS	2.1	37	250mV	TO 50
	U 4620BS	2.1	35	250mV	TO 50 Ceramic
64	U 810BS	1.2	40	ECL	TO 50
128	U 811BS	1.2	40	ECL	TO 50
256	U 847BST	1.3	35	TTL	DIP-8
	U 812BS	1.2	40	ECL	TO 50
64-128 -256	U 813BS	1.1	35	ECL	DIP8, SO8, SIP6
	U 813BSE	1.1	35	EF	DIP8, SO8, SIP6
	U 833BS	1.3	35	ECL	DIP8, SO8, SIP6
	U 833BSE	1.3	35	EF	DIP8, SO8, SIP6
	U 891BS	1.1	47	ECL	DIP8, SO8
	U 893BSE	1.3	48	EF	DIP8, SO8, SIP6

ICs for Tuners and Tuning

FUNCTION	TYPE	F _{max} GHz	FEATURE	PACKAGE
PLL	U 6316B	1.0	3 WIRE BUS	DIP20
	U 6358B	1.0	3 WIRE BUS	DIP20, SO20
	U 6359B	1.0	3 WIRE BUS	DIP20, SO20
	U 6202B-B	1.3	I ² C BUS	DIP18, SO20
	U 6204B-A	1.3	I ² C BUS	DIP18, SO20
FRONT END	U 632B	0.3	FM TUNER	DIP18
		0.47	TV MIXER	DIP14, SO16
		0.47	TV MIXER	DIP14, SO16
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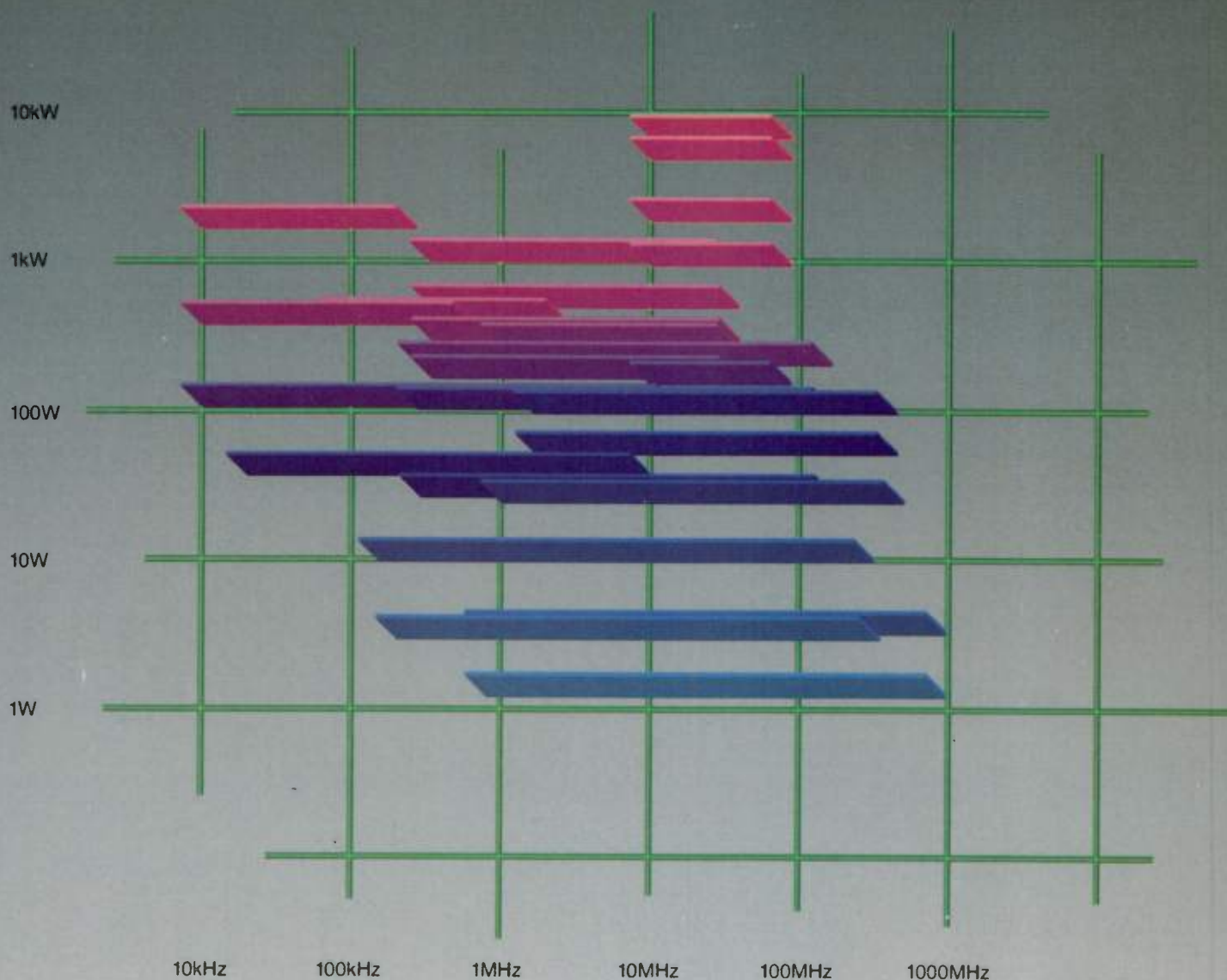
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