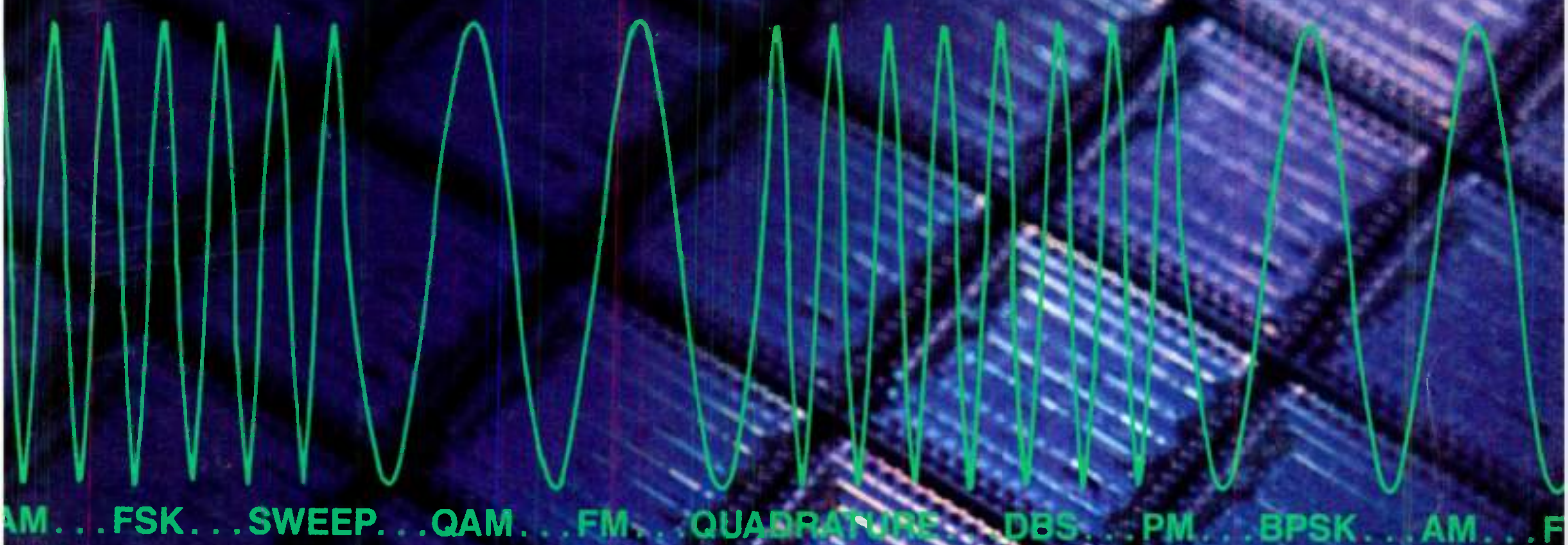
rf design

engineering principles and practices

March 1988

1978–1988 10th ANNIVERSARY



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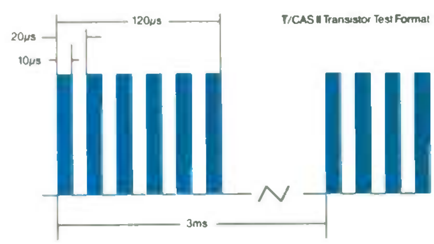

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INFO/CARD 1

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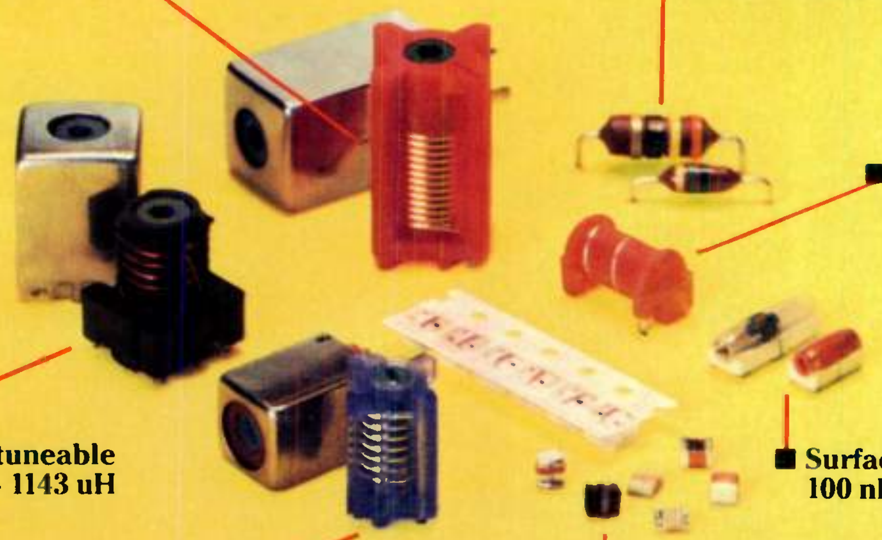

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INFO/CARD 17

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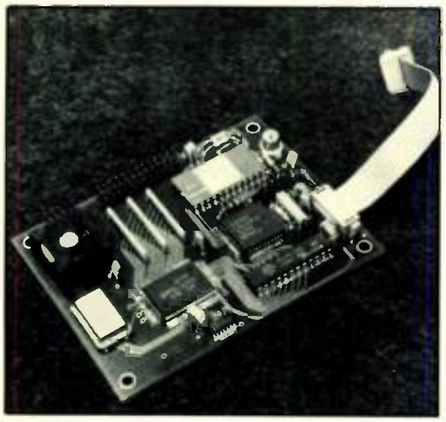
March 1988



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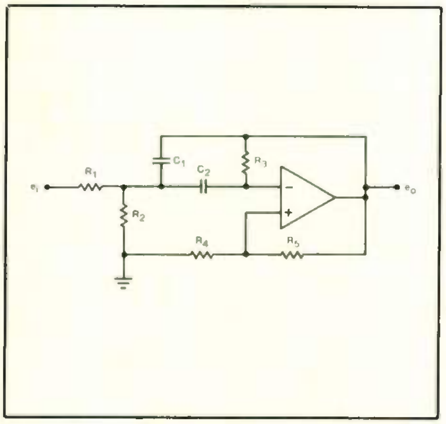

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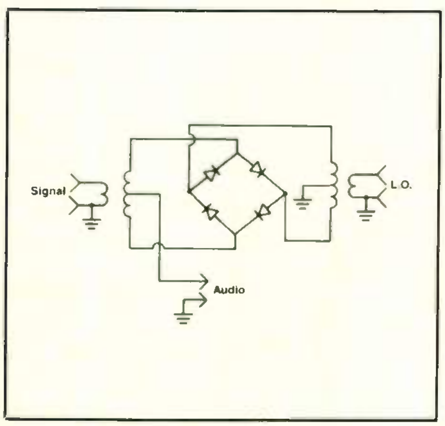

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Cover Story

Featured Technology Section

Departments

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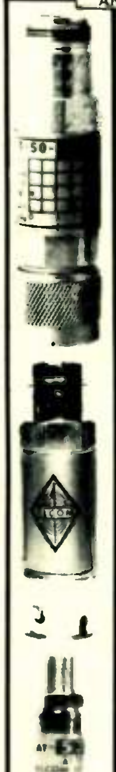
LOWEST PRICED, HIGHEST QUALITY ATTENUATORS – BNC $11.00 1-9 EA, SMA $14.40 10 EA AND TERMINATIONS – BNC $5.60 10 EA., SMA $5.60 10 EA., MIL. HI-REL. NETWORKS



Model Number (2)	Impedance Ohms (Power W)	Frequency Range	BNC	TNC	N	SMA	UHF	PC
			UNIT PRICE (4) EFFECTIVE 9-15-86					
Fixed Attenuators 1 to 20 dB								
AT 50(3)	50 (.5W)	DC 1.5GHz	14.00	20.00	20.00	18.00	—	—
AT 51	50 (.5W)	DC 1.5GHz	11.00	15.00	15.00	14.00	—	12.00
AT 52	50 (1W)	DC 1.5GHz	14.50	20.50	20.50	19.50	—	—
AT 53	50 (.25W)	DC-3.0GHz	14.00	17.00	—	15.00	—	—
AT 54	50 (.25W)	DC-4.2GHz	—	—	—	18.00	—	—
AT 55	50 (.25W)	DC 4.2GHz	—	—	—	14.40 (10 Pc)	—	—
AT 75 or AT 90	75 or 93 (.5W)	DC 1.5GHz (750MHz)	11.50	20.00	20.00	18.00	—	—
Detector, Mixer, Zero Bias Schottky								
CD-51, 75	50, 75	01-4.2GHz	54.00	—	—	54.00	—	—
DM-51	50	01-4.2GHz	—	—	—	64.00	—	—
Resistive Impedance Transformers, Minimum Loss Pads								
RT 50/75	50 to 75	DC 1.5GHz	10.50	19.50	19.50	17.50	—	—
RT 50/93	50 to 93	DC 1.0GHz	13.00	19.50	19.50	17.50	—	—
Terminations								
CT 50 (3)	50 (.5W)	DC 4.2GHz	11.50	15.00	15.00	17.80	—	—
CT 51	50 (.5W)	DC 4.2GHz	9.50	12.00	10.50	9.50	—	—
CT 52	50 (1W)	DC 2.5GHz	10.50	18.00	18.00	13.00	18.50	—
CT 53/M	50 (.5W)	DC 4.2GHz	5.60 (10 Pc)	—	—	5.60 (10 Pc)	—	—
CT 54	50 (2W)	DC 2.0GHz	14.00	15.00	18.00	17.50	—	—
CT 75	75 (.25W)	DC 2.5GHz	10.50	15.00	15.00	13.00	18.50	—
CT 93	93 (.25W)	DC 2.5GHz	13.00	15.00	—	15.00	15.50	—
Mismatched Terminations 1.05:1 to 3:1, Open Circuit, Short Circuit								
MT 51	50	DC 3.0GHz	45.50	45.50	45.50	45.50	—	—
MT 75	75	DC 1.0GHz	—	—	45.50	—	—	—
Feed thru Terminations, shunt resistor								
FT 50	50	DC 1.0GHz	10.50	19.50	19.50	17.50	—	—
FT 75	75	DC 500MHz	10.50	19.50	19.50	17.50	—	—
FT 90	93	DC 150MHz	13.00	19.50	19.50	17.50	—	—
Directional Coupler 30 dB								
DC 500	50	250-500MHz	60.00	—	84.00	—	—	—
Resistive Decoupler, series resistor or Capacitive Coupler, series capacitor								
RD or CC-1000	1000 (1000PF)	DC 1.5GHz	12.00	18.00	18.00	17.00	—	—
Adapters								
CA 50 (N to SMA)	50	DC-4.2GHz	13.00	13.00	13.00	13.00	—	—
Inductive Decouplers, series inductor								
LD-R15	0.17uH	DC-500MHz	12.00	18.00	18.00	17.00	—	—
LD 6R8	6.8uH	DC-55MHz	12.00	18.00	18.00	17.00	—	—
Fixed Attenuator Sets 3, 6, 10, and 20 dB, in plastic case								
AT 50-SET (3)	50	DC 1.5GHz	80.00	84.00	84.00	78.00	—	—
AT 51 SET	50	DC 1.5GHz	48.00	64.00	64.00	60.00	—	—
Reactive Multicouplers, 2 and 4 output ports								
TC 125-2	50	1.5-125MHz	64.00	—	67.00	67.00	—	—
TC 125-4	50	1.5-125MHz	67.00	—	81.50	81.50	—	—
Resistive Power Dividers, 3, 4 and 9 ports								
RC 3-50	50	DC 2.0GHz	64.00	84.00	—	64.00	—	—
RC 4-50	50	DC 500MHz	64.00	64.00	—	64.00	—	—
RC 8-50	50	DC 500MHz	—	—	—	84.50	—	—
RC 3-75, 4-75	75	DC 500MHz	64.00	84.00	—	64.00	—	—
Double Balanced Mixers								
DBM 1000	50	5-1000MHz	61.00	—	71.00	61.00	—	34.00
DBM 500PC	50	2-500MHz	—	—	—	—	—	34.00
RF Fuse, 1/8 Amp and 1/16 Amp								
FL 50	50	DC 1.5GHz	12.00	18.00	45.50	17.00	—	—
FL 75	75	DC 1.5GHz	12.00	18.00	—	17.00	—	—

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Publisher
Robert A. Searle

Editor
Gary A. Breed

Technical Editor
Mark Gomez

Consulting Editors
Andy Przedpelski
Robert J. Zavrel, Jr.

National Sales Manager
Kate Walsh

Advertising:
West Coast District Manager
Mary Bandfield
1341 Ocean Ave., Ste. 58
Santa Monica, CA 90401
(213) 458-6683

Midwestern States
Kate Walsh
Main Office

Eastern Sales Manager
Joseph Palmer
36 Belmont Rd. S.W. 3
West Harwich, MA 02671
(617) 394-2311

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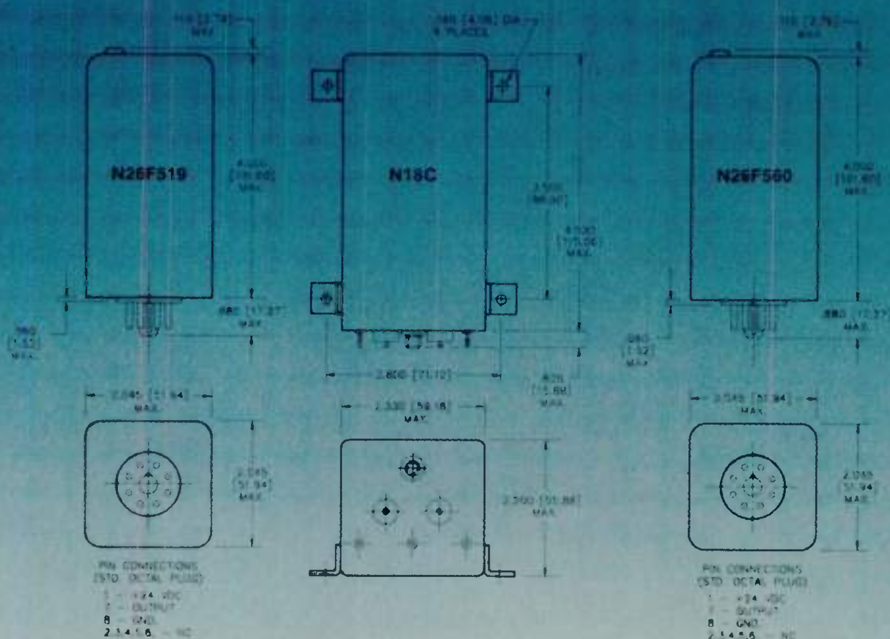

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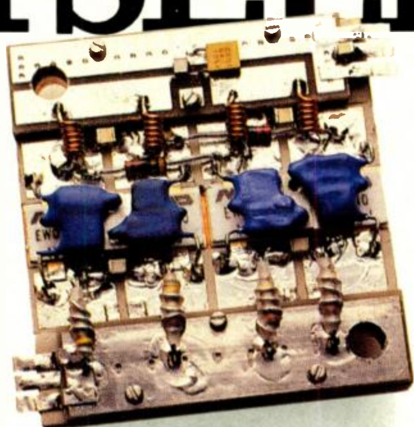

You're looking at the highest power density long pulse UHF amplifier we've ever built. And as far as we know, no one else has even come close.

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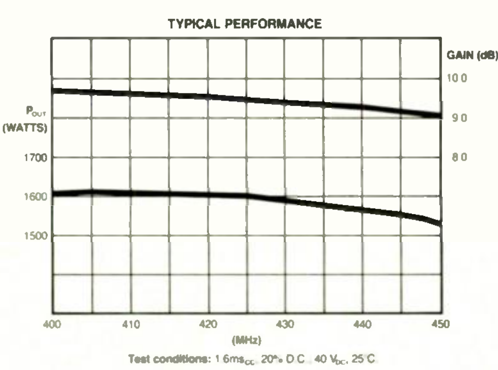

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INFO/CARD 9 **Literature**
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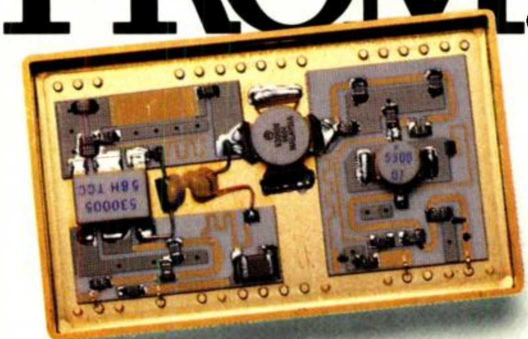
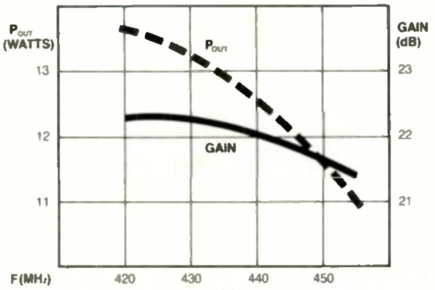


Two hundred and fifty a month, that's the number of this custom amplifier that MMD currently ships to Hughes.

Which is no small feat because when Hughes Aircraft approached MMD, they had the knotty problem of producing this sophisticated amplifier in volume.

MMD took a unique approach to solve this challenge. Our high degree of integration enabled us to respond with precisely what Hughes needed for its PLRS system: a thick film UHF pulsed amplifier with 10 watts of output power (-55 to $+85$ degrees C) at $+19$dBm input, with consistent lot to lot performance through production.



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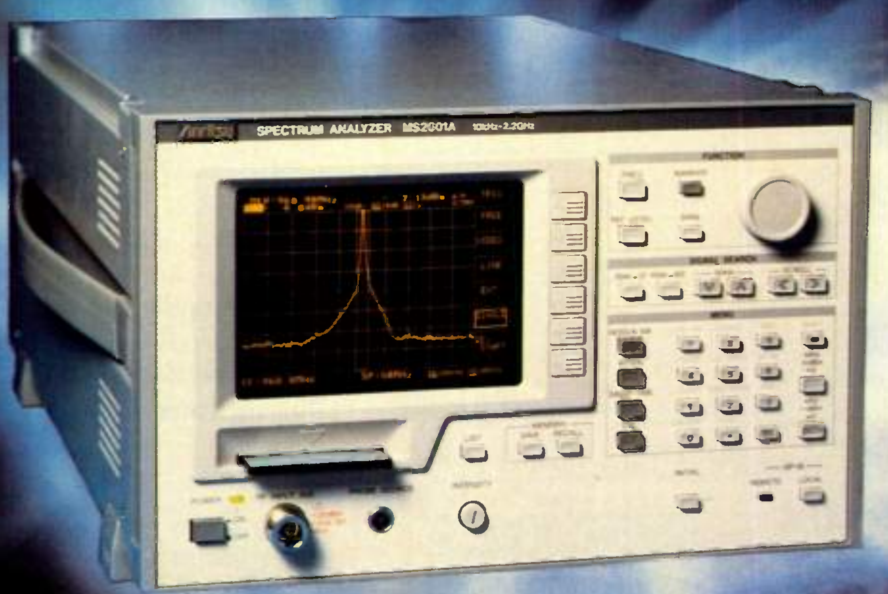

INFO/CARD 13

In Support of CAE

Editor:

I have just read Mr. Woodward's letter in the December issue, which you captioned "Man vs. Machine." It's not man vs. machine, it's man uses machine. When I was in college in 1950s, we had to work out complex circuit problems using a slide rule, pencil and paper. This was not my idea of fun. I am really glad those days are gone, I don't look back on them with nostalgia. Personal computers today can give you answers in seconds that took hours or days back then.

Mr. Woodward is right when he says engineers should build and tweak real circuits, but I say, first build the circuit in a computer. It takes 1/10 as long, and the computer will tell you if the circuit is going to work. "Not tweaking trash" means you know the circuit will work before you start building it. Many an engineer has spent hours tweaking a circuit, only to find out that there was a slight error in his calculations.

In your September editorial you seemed to assume that young engineers prefer using computers and older engineers prefer using older methods of circuit design. Good engineers keep up with the times, regardless of their age.

Robert Stanton
Norwich, NY

(The September editorial was only intended to warn engineers of over-dependence on computers, particularly younger engineers who may not have experienced the slide rule, pencil and paper ordeal. — Editor)

A Note on Filter Delay Analysis

Editor:

In the October 1987 issue, Robert Kane has given a program, with technical discussion, which accurately computes the envelope delay and delay distortion of various Butterworth and Chebyshev lowpass filters. I would like to offer a few comments which may be of value to readers who have an interest in this type of analysis.

The computation of envelope delay from the unfactored form of a transfer function is unnecessarily complicated, from a mathematical point of view. Since the delay contributions of the roots of a transfer function are additive, the usual procedure is to sum the contributions of the individual roots. For minimum-phase functions, such as reactive ladders, only the poles need be considered, since zeros on the jω axis contribute no delay. The delay produced by the pole at $-A \pm jB$, at a normalized frequency ω, is given by the simple expression:

$$T = \frac{A}{A^2 + (\omega \pm B)^2}$$

Mr. Kane's program contains a great deal of specific numerical information in the form of polynomial coefficients which tends to limit its versatility. The poles of a Butterworth lowpass function lie equally spaced on the unit circle in the left-half s-plane, and can be calculated by simple trigonometric functions. The poles of a Chebyshev function can be derived from these by multiplying the real and imaginary parts by a pair of factors which I have called SINH and COSH (ref. 1). These names have theoretical significance, but they can be computed by simple algebraic procedures without explicit reference to hyperbolic functions.

The enclosed routine, written in

Need an unusual mixer?

If we haven't already made it we'll invent it for you!

We have a full catalog and an impressive design file for Double Balanced Mixers, 10 KHz to 12 GHz. One of them will almost certainly fill your bill. Should you require one that we have not made before, we'll design it to your specifications, commensurate with the state of the art.

Here are just a few examples of some standard mixers and a few "specials."

Characteristics	Frequency Range	Conversion Loss Max (dB)	L.O. Power (dBm)	Isolation dB LO–RF	Isolation dB LO–IF	Package	Model
Low Level	0.05-200 MHz	6.5	0	50	45	P,F,C	FC-193Y / FC-194Y
Wide Band	2-1250 MHz	8.0	+7	35	30	P,C	FC-200Z / FC-201Z
General Purpose	10-1000 MHz	7.5	+7	30	25	F	FC-200ZF
Wide Band	10-3000 MHz	8.0	+10	30	25	F,C	FC200ZF-30 / FC-201ZF-30
Low Loss*	4.4-5.0 GHz	5.5	+10	30	25	C	FC-325D
Low Loss,* Low Distortion	7.9-8.4 GHz	5.5	+17	28	27	C	FC-327F
Wide Band	1.9-9.5 GHz	8.5	+7	20	20	C	FC-304SX
Low Distortion	2-1250 MHz	8.5	+13	35	30	P,F,C	FC-217Z / FC-218Z
Ultra Low Dist.	2.0-1000 MHz	8.0	+20	35	30	P,C	FC-234Z / FC-235Z
High Intercept Point (+35 dBm)	25-1000 MHz	7.0	+27	30	30	F,C	FC244Z / FC-245Z
Hi Compression Point (+20 dBm)	10-1000 MHz	7.5	+27	30	30	P,C	FC-253Z / FC254Z

P= P.C. Package F=Flatpack C=Connector Version *Available from 0.7 GHz to 12 GHz.

If any of these mixers look interesting, send for our catalog. If they don't quite fit your application—give us a call or send us your specs. We'll do the rest.

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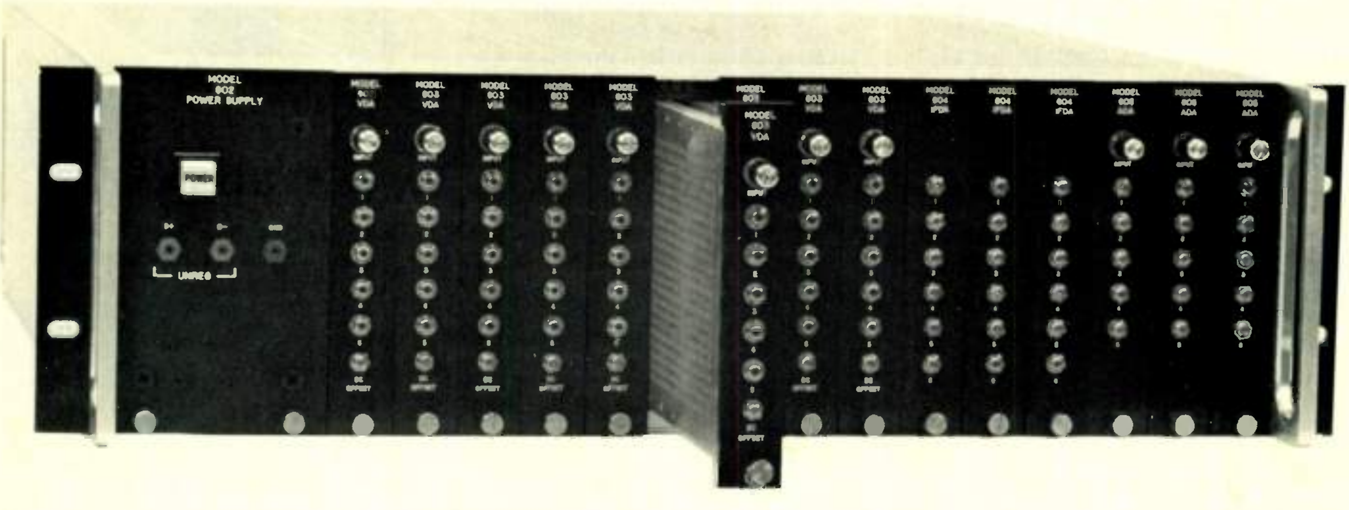
INFO/CARD 14

REACTION INSTRUMENTS'

MODEL 600
MODULAR SIGNAL PROCESSOR

Reaction Instruments' Model 600 Modular Signal Processor is a compact, highly flexible system in a single, stand-alone chassis, which provides frequency conversion, distribution and switching of signals in the Audio, Video, IF and RF range. Special function modules are available.



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SIGNAL SWITCHES □ DIGITAL CONTROLLER

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INSTRUMENTS
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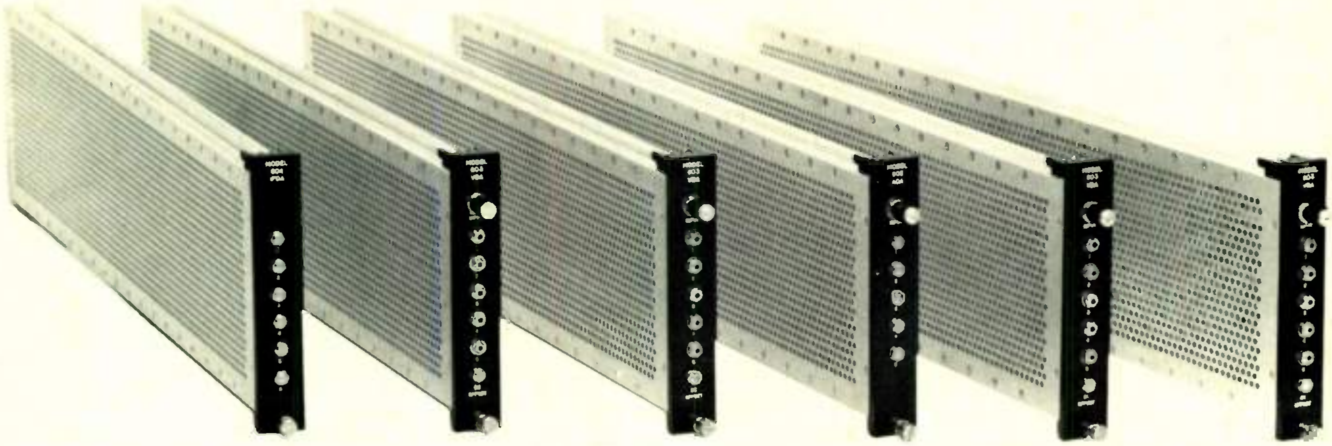


CONFIGURE THE SYSTEM OF YOUR CHOICE

MODEL 601 MAINFRAME
Dimensions: 5.25″ High by 20″ Deep
Suitable for 19″ Rack Mounting

MODEL 602 POWER SUPPLY

MODEL 603
VIDEO DISTRIBUTION AMP
Range: DC-30 MHz
No. of Outputs: 5 (6 optional)
Gain: +10 to −10 dB

MODEL 604 IF DISTRIBUTION AMP
Range: 5-300 MHz
No. of Outputs: 5 (6 optional)
Gain: 0 dB

MODEL 605
AUDIO DISTRIBUTION AMP
Range: 200-6000 Hz
(10 Hz to 100 kHz optional)
No. of Outputs: 5 (6 optional)

MODEL 606 IF-TAPE CONVERTER
Input: 21.4 MHz (160 MHz optional)
Output: 1.075 MHz (3.225 MHz optional)
Delay Equalized

MODEL 608 IF TO IF CONVERTER
Input: 21.4 MHz (160 MHz Optional)
Output: 160 MHz (21.4 MHz optional)
External Reference

MODEL 609 RF SIGNAL SWITCH
Range: 20-1250 MHz
No. of Inputs: 6
No. of Outputs: 1

MODEL 613 VIDEO SIGNAL SWITCH
Range: DC to 30 MHz
No. of Inputs: 6
No. of Outputs: 1

ALL UNITS DESIGNED TO MEET TEMPEST REQUIREMENTS

For further information and detailed specifications contact
Reaction Instruments
475 Springpark Place
Herndon, VA 22070
PHONE: 703/471-6060 FAX 703/471-4847
INFO/CARD 15

Microsoft BASIC (shown below), computes the Butterworth poles, modifies them to the elliptic Chebyshev locus if the non-zero ripple is requested, lists the poles, sums the delay contributions of the poles at each requested frequency, and scales the delay in accordance with the specified cutoff frequency. The basic formulas are all normalized to the ripple bandwidth. A sample computation using Kane's example shows agreement within 0.2 percent.

There are two advantages to computing the delay from the poles in this way: First, the poles of various types of lowpass functions have been extensively tabulated (ref. 2); and second, there are various procedures (e.g. ref. 3) for transforming these lowpass poles into the poles of corresponding highpass, bandpass and band-reject functions, so the envelope delays of these configurations can be directly computed. The algorithms given here could be used with Mr. Kane's plotting routine, etc., to enhance the versatility of his program.

David C. Bidwell
Dowty RFL Industires, Inc.
Boonton, NJ

```
PROGRAM ENVDEL                          11-06-1987

1   REM  PROGRAM ENVDEL (Chebyshev Envel
     ope Delay) - DCB 11/5/87
10 INPUT "DEGREE, RIPPLE (DB)";N,R
20 INPUT "CUTOFF FREQ. (RIPPLE BW)";FC
30 LPRINT USING "N = ##.  RIPPLE = #.##
     ## DB";N,R
40 LPRINT
50 LPRINT "FC ="FC : LPRINT
60 IF R<>0 THEN 80
70 SINH=1 : COSH=1 : GOTO 140
80 XISQ=10^(R/10)-1
90 XI=SQR(XISQ)
100 T1=(SQR(1/XISQ+1)+1/XI)^(1/N)
110 T2=1/T1
120 SINH=(T1-T2)/2
130 COSH=(T1+T2)/2
140 PI=3.14159
150 FOR J%=0 TO N-1
160 A(J%+1)=SIN((2*J%+1)*PI/2/N)*SINH
170 B(J%+1)=COS((2*J%+1)*PI/2/N)*COSH
180 LPRINT USING"POLE ## = ##.###### +J
      ##.######";J%+1,-A(J%+1),B(J%+1)
190 NEXT J%
200 LPRINT
210 LPRINT "W/WC          DELAY" : LPRINT
220 FOR W=.01 TO 3 STEP .02
230 T=0
240 FOR J%=1 TO N
250 T=T+A(J%)/(A(J%)^2+(W+B(J%))^2)
260 NEXT J%
270 T=T/2/PI/FC
280 LPRINT USING "##.##      ##.######
      ";W,T
290 NEXT W
300 END
```

References

1. Louis Weinberger, *Network Analysis and Synthesis,* McGraw-Hill Book Co., New York, 1962.

2. E. Christian and E. Eisenmann, *Filter Design Tables and Graphs,* John Wiley & Sons, New York, 1966.

3. D.C. Bidwell, "Nested Functions Transform Filter Roots," *EDN,* Mar. 18, 1987.

An Omission

The figure below was omitted from page 80 of the February 1988 issue ("Equal-Ripple LC Filter Synthesis" by Robert E. Kost):

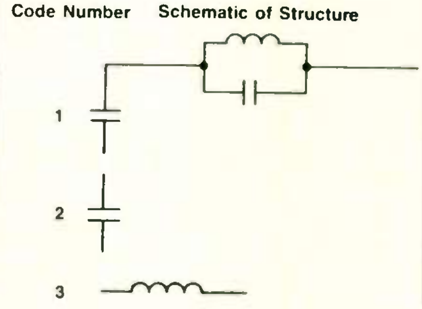


Figure 5. Circuits along with their code numbers.

WBE

MINIATURE RF AMPLIFIERS

Series A82 satisfies applications that require the following specifications in a rugged miniature package. Series A82 is ideal for direct antenna mounting. Applications include airborne equipment, test instrumentation, sweep equipment, spectrum analysers, frequency counters, telemetry, radio astronomy, wide band communications, or a convenient all around test bench amplifier.

Each stage is transformer coupled for highest power output with minimum distortion. Selective feedback is used to maintain flatness and impedance match over the entire bandwidth.

MODELS AVAILABLE

Models A82 and A82A are amplifiers of ultra wide bandwidth, extreme flatness, and excellent impedance match.

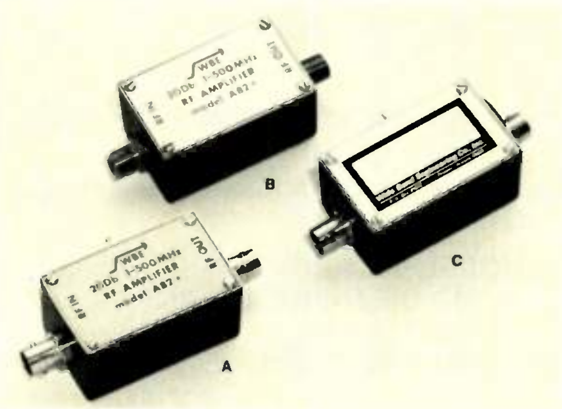
Models A82L and A82LA offer a more limited or specific bandpass with less stringent flantess specifications.

Models A82/RP, A82A/RP, A82L/RP, and A82LA/RP are remote power option amplifers which may be remotely powered through the output connector.

O.E.M. and quantity pricing available. Consult factory with specifications.

PICTURED

A) Model A82A B) Model A82A/RP C) O.E.M. Special Amp.



Model	Freq Range (Full Spec)	Approximate 3 dB points (MHz)	Gain	Gain Flatness	Output Capability in V output for 1 dB Compression	Power Requirements +12 VDC @__mA	VSWR	Noise Figure	Reverse Atten-uation	Weight oz.
A82	1-500	.3-650	20 dB Stable ±.5 dB -40 - 170°F	±.15	.7	28	1.5:1 max	7 dB max	-30 dB typical	2 1/2
A82A	1-500	.3-650		±.15	.7	28				3
A82L	.1-50	.050-150		±.5	1.0	50	1.1:1 typical	4.5 dB typical		3
A82LA	.4-30	.3-100		±.5	1.0	50				3

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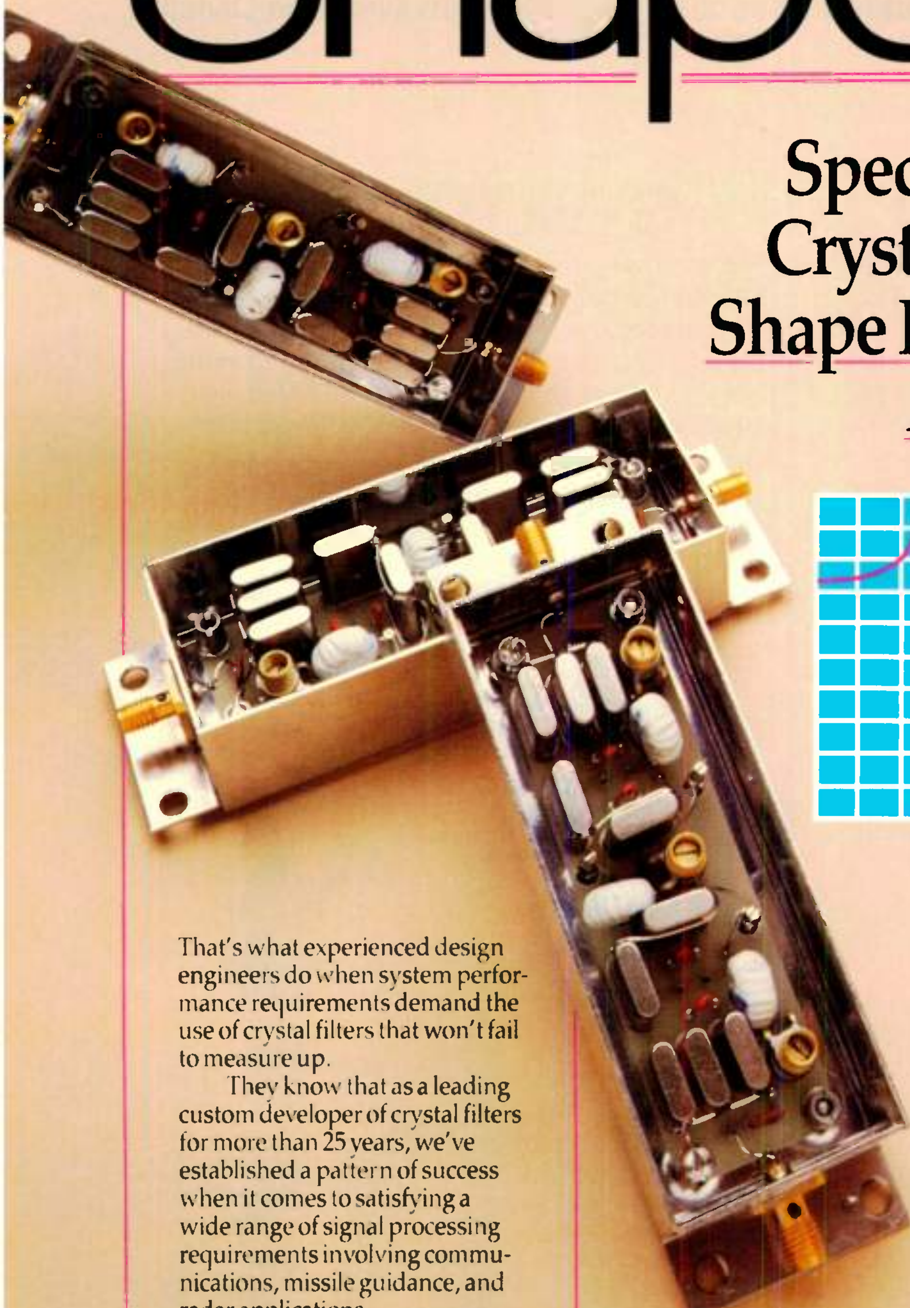
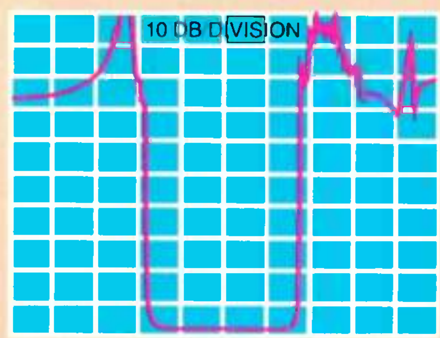
INFO/CARD 16

Shape Up!



Specify DAMON Crystal Filters with Shape Factors As Steep As 1.05:1.



That's what experienced design engineers do when system performance requirements demand the use of crystal filters that won't fail to measure up.

They know that as a leading custom developer of crystal filters for more than 25 years, we've established a pattern of success when it comes to satisfying a wide range of signal processing requirements involving communications, missile guidance, and radar applications.

And that in the process, we've earned a reputation for meeting some rather imposing specifications in the 10KHz–150MHz range with a set of equally imposing accomplishments that include: Shape factors as steep as 1.05:1 with 80dB attenuation levels... Phase and amplitude matching within the tightest tolerances... And passbands that remain constant over a wide temperature range.

As a system designer it means that you can optimize incoming signal processing, thus assuring better selectivity at the front end.

All of which adds up to increased design flexibility in terms of both performance and packaging, as well as improved reliability at the overall system level.

Behind this claim is Damon's continuing commitment to the latest in manufacturing resources, which includes a facility that's equipped with the latest in CAD and CAE technology, as well as provisions for in-house environmental testing (MIL-Std-202), crystal fabrication (MIL-C-3098), and soldering (WS-6536E). And that's only the beginning.

For more information, call us today to discuss your application (1-800-348-0028). And send for our complete catalog.

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KLYSTRONS

Beginning in the mid 70s Varian provided multi-megawatts of RF power for lower hybrid. This was achieved by combining half-megawatt klystrons at S band and quarter-megawatt klystrons at C band. This contributed significantly to the major United States effort in PLT and Alcator-C.

The reliability and long life of these tubes has led to their use by other facilities throughout the world.

Of significant importance is the advanced tube technology required to produce these products. This expertise is available for the design and production of tubes to meet your special requirements, whether they be plasma heating or particle accelerators.

VKS-8269/8270

VKC-7849 KLYSTRON

Operating Frequency (GHz)	CW Power (kW)	Typical Gain (dB)	Typical Efficiency (%)	Type Number
2.440–2.460	500	57	52	VKS-8269A
4.6	250	55	43	VKC-7849

For further information on Klystrons for lower hybrid heating call Varian Microwave Tube Division at (415) 424-5668.

INFO/CARD 18

GYROTRONS

From a very modest beginning 12 years ago Varian has become the world leader in Gyrotron technology. CW tubes up to 200 kW are available in frequencies from 28 GHz to 140 GHz. Current design efforts will soon lead to one Megawatt devices at 140 GHz and even higher frequencies.

Varian's experience in developing and producing high power microwave sources has created the capability to produce tubes for a multitude of plasma heating applications, ion source development and materials processing research.

VGE-8060
60 GHz
PULSED
GYROTRON IN
MAGNET

VGT-8014
140 GHz
GYROTRON

OSCILLATORS

Frequency (GHz)	Type	Output Power Pulse/CW (kW)
8	VGH-8001	500/ —
28	VGA-8000	— /200
35	VGA-8003	— /200
53	VGE-8053	200/ —
53	VGE-8005A2	— /200
56	VGE-8005	— /200
60	VGE-8006	— /200
70	VGE-8007	— /200
106	VGB-8106A1	400/ —
140	VGT-8014	200/100
250	VGT-8025	25/ —

For further information on Gyrotron Oscillators and Amplifiers call Varian Microwave Tube Division at (415) 424-6984.

INFO/CARD 19

edge of fusion science.

active participants in TFTR, Doublett III, JET, JT-60, FTU, Heliotron, Culham, Wendelstein and other leading fusion energy experiments.

POWER GRID TUBES

Varian Eimac's half century of experience in developing state-of-the-art power grid tubes has brought them to the forefront in Multi Megawatt Amplifier Tetrodes and Switch/Regulator Tetrodes for ICRH and Accelerator Amplifiers; NBI, ECRH, and LHH Switch/Regulators.

X2242
2.5 MEGAWATT
AMPLIFIER TETRODE

4CPW 100K
100 KV
SWITCH/REGULATOR

TYPICAL AMPLIFIER TUBES

Type	Maximum CW Output (MW)	Maximum Frequency (MHz)
8973	1.5	200
8974	3.0	50
X2242	2.5	150

TYPICAL SWITCH/REGULATOR TUBES

Type	Voltage	Maximum Current	Anode Dissipation
4CPW 100K	110 KV	100A	100 kW
4CPW 300K	100 KV	150A	300 kW
4CPW 1000KA	150 KV	150A	1000 kW
4CPW 1000K	175 KV	100A	1000 kW

For further information on Power Grid Tubes call Varian Eimac at (415) 594-4006.

INFO/CARD 20

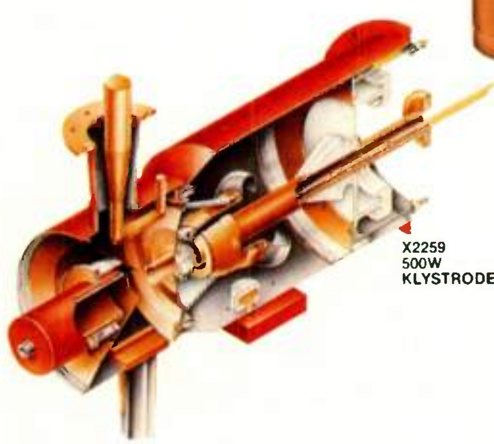
KLYSTRODES

Varian Eimac, long a leader in high power broadcast tube technology, has recently developed a series of high-power high-efficiency amplifiers for UHF-TV service based on the first new configuration of high-power electron tubes in over a decade.

This technology has led to the development of High Power UHF Amplifier Klystrodes for Accelerator and Fusion Service.

- RF Drive Pulsed
- High Efficiency
- Compact

X2254
KLYSTRODE
UHF AMPLIFIER



X2259
500W
KLYSTRODE CUTAWAY

KLYSTRODE POWER OUTPUT

Type	CW (kW)	Peak (kW)	Frequency
X2254	15	30	470–820 MHz
X2252	30	80	600–820 MHz
X2253	30	80	470–600 MHz
X2259	50	500	425 MHz
Dev.	500	500	425 MHz

For further information on Klystrodes call Varian Eimac at (415) 594-4000.

INFO/CARD 21

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INFO/CARD 22

rf courses

Besser Associates

Microwave Circuit Design I: Linear Circuits
June 20-24, 1988, Los Angeles, CA
August 15-19, 1988, Baltimore, MD

Microwave Circuit Design II: Non-linear Circuits
August 22-26, 1988, Baltimore, MD

Information: Les Besser, Besser Associates, Inc., 3975 East Bayshore Road, Palo Alto, CA 94303; Tel: (415) 969-3400

UCLA Extension

Digital and Analog Modulation
April 4-8, 1988, Los Angeles, CA

Modern Microwave Techniques
April 25-28, 1988, Los Angeles, CA

Introduction to Automatic Testing and ATE Engineering
May 9-12, 1988, Los Angeles, CA

Advanced Topics in Automatic Test Equipment
May 16-20, 1988, Los Angeles, CA

Microwave Circuit Design I
June 20-24, 1988, Los Angeles, CA

Information: UCLA Extension, P.O. Box 24901, Los Angeles, CA 90024; Tel: (213) 825-1901; (213) 825-1047; (213) 825-3344

Test Systems, Inc.

MIL-STD-1553
May 10-11, 1988, Phoenix, AZ

Information: Leroy Earhart, Test Systems, Inc., 217 W. Palmaire, Phoenix, AZ 85021; Tel: (602) 861-1010

Interference Control Technologies, Inc

Grounding and Shielding
March 21-25, 1988, Hilton Head, SC
April 11-15, 1988, San Diego, CA
April 25-29, 1988, Chicago, IL
May 11-13, 1988, Washington, DC
May 16-20, 1988, Toronto, Canada
June 13-17, 1988, Las Vegas, NV
June 20-24, 1988, Anaheim, CA
July 11-15, 1988, Boulder, CO

TEMPEST Facilities
April 11-15, 1988, Washington, DC
July 25-29, 1988, Palo Alto, CA

MIL-STD-461
May 3-5, 1988, Washington, DC
July 12-14, 1988, Palo Alto, CA

Intro to EMI/RFI/EMC
May 11-13, 1988, Washington, DC

TEMPEST Design
June 14-17, 1988, Palo Alto, CA

Information: Penny Caran, Registrar, Interference Control Technologies, Inc., State Route 625, P.O. Box D, Gainsville, VA 22056; Tel: (703) 347-0030

Compliance Engineering

Compliance Seminars: EMI, Safety, ESD, Telecom
April 19-22, 1988, Chicago, IL
June 7-10, 1988, Boston, MA

Information: Compliance Engineering, 593 Massachusetts Avenue, Boxborough, MA 01719; Tel: (617) 264-4208.

Georgia Institute of Technology

Elements of Phased Array Radar System Design
March 29-April 1, 1988, Atlanta, GA

Antenna Engineering
April 5-8, 1988, Atlanta, GA

Information: Deidre Mercer, Education Extension Services, Georgia Institute of Technology, Atlanta, GA 30332-0385; Tel: (404) 894-2547.

Southeastern Center for Electrical Engineering Education

Antennas: Principles, Design, and Measurements
August 2-5, 1988, San Diego, CA

Information: Ann Beekman, SCEEE, 1101 Massachusetts Ave., St. Cloud, FL 32796; Tel: (305) 892-6146

R & B Enterprises

Understanding & Applying MIL-STD-461C
March 28-29, 1988, Philadelphia, PA
April 25-26, 1988, Los Angeles, CA
June 7-8, 1988, Washington, DC

MIL-STD-461C/462 Test Workshop
March 30-March 31, 1988, Philadelphia, PA
April 27-28, 1988, Los Angeles, CA

Electromagnetic Pulse (EMP) Design & Test
April 11-12, 1988, Washington, DC
June 6-7, 1988, Philadelphia, PA

Identification & Control of Microwave/RF Hazards
May 11-13, 1988, Philadelphia, PA

TEMPEST-A Detailed Design Course
May 16-20, 1988, Philadelphia, PA

Grounding, Bonding & Shielding
June 16-17, 1988, Philadelphia, PA

Worst Case Circuit Analysis
June 20-22, 1988, Philadelphia, PA

EMI Suppression Methods
June 28-30, 1988, Philadelphia, PA

The R & B EMI Training Institute
August 8-19, 1988, Philadelphia, PA

Information: Greg Gore, R & B Enterprises, 20 Clipper Road, West Conshohocken, PA 19428; Tel: (215) 825-1684

Design & Evaluation, Inc.

The Worst Case Circuit Analysis Training Seminar
April 4-6, 1988, Orlando, FL
April 18-20, 1988, San Diego, CA
May 9-11, 1988, Washington, DC
July 11-13, 1988, Honolulu, HI
September 12-14, 1988, Boston, MA
October 17-19, 1988, San Francisco, CA

Information: Design & Evaluation, Inc., 1000 White Horse Road — Suite 304, Voorhees, NJ 08043; Tel: (609) 770-0800

Technology Service Corporation

Radar Signal Processing and Clutter
March 22-25, 1988, Atlanta, GA

Information: Lynda S. Epstein, Technology Service Corp., 962 Wayne Ave. — Suite 600, Silver Spring, MD 20901; Tel: (800) 638-2628 or (301) 565-2970

rf calendar

April 17-20, 1988
7th Annual Microwave Integrated Circuit Workshop
San Diego, CA
Information: Jerry Renken, Honeywell TSLO MN67-1B17, 6300 Olson Memorial Highway, Golden Valley, MN 55424; Tel: (612) 593-4385

April 17-20, 1988
1988 Array Conference
The Grand Kempinski Hotel, Dallas, TX
Information: Doug Giese, Vice-President, Array AP Labs, 4411 Morena Blvd., Suite 150, San Diego, CA 92117; Tel: (619) 272-8890

April 19-22, 1988
IEEE Instrumentation/Measurement Technology Conference
San Diego Princess Hotel, San Diego
Information: Bob Myers, IMtc, 1700 Westwood Blvd., Los Angeles, CA 90024; Tel: (213) 457-4571

April 20-27, 1988
Hannover Fair Industry 88
Hannover, West Germany
Information: Hannover Fairs USA, Inc., P.O. Box 7066, 103 Carnegie Center, Princeton, NJ 08540; Tel: (609) 987-1202

April 24-28, 1988
Association of Old Crows Technical Symposium of the Mountain and Western Regions
Red Lion Inn, Colorado Springs, CO
Information: Dan Odum, SIERRA Technical Group Inc., 43 Inverness Drive East, Englewood, CO 80112; Tel: (303) 790-1700

May 10-12, 1988
Electro '88
Bayside Exposition Center, Boston World Trade Center, Boston, MA
Information: Electronic Conventions Management, 8110 Airport Boulevard, Los Angeles, CA; Tel: (213) 772-2965

May 10-12, 1988
EMC Expo 88
Washington Hilton, Washington, DC
Information: Karen Smith, EMC Expo 88, P.O. Box D, Gainesville, VA 22065; Tel: (703) 347-0030

May 25-27, 1988
1988 IEEE MTT-S International Microwave Symposium
Javits Auditorium, New York City, NY
Information: Charles Buntschuh, Narda Microwave Corp., 435 Moreland Road, Hauppauge, NY 11788; Tel: (516) 231-1700

June 1-3, 1988
42nd Annual Frequency Control Symposium
Stouffer Harborplace Hotel, Baltimore, MD
Information: Raymond L. Filler, Frequency Control and Timing Branch, Department of the Army, Electronics Technology and Devices Laboratory, Fort Monmouth, NJ 07703-5000

POWER ATTENUATORS!

JFW Fast Delivery

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INFO/CARD 23

RF Expo East Presents Latest Technology to Over 1,600 RF Engineers

Anaheim, Calif., was again host to RF Technology Expo, putting engineers in touch with RF manufacturers and suppliers, and offering 48 technical papers on practical engineering topics. The vitality of the RF industry was demonstrated by a high level of activity on the exhibit floor as attendees examined the latest components and equipment.

Among the most active product areas were: software design tools, direct-digital and PLL synthesizers, power amplifiers, high performance components and many new test instruments. The interest in these areas was evident in the technical sessions, too. Although there were twelve papers covering various aspects of frequency synthesis, *all* were well attended. On Thursday, papers by Dave Badger of Wavetek, Mike Black of Texas Instruments and Jeff Blake of Fairchild Weston Systems were presented before a standing-room-only crowd of more than 250 engineers.

As would be expected, some papers with very specific focus drew small but interested audiences. However, papers on component applications, antennas and propagation, test systems and CAE/CAD topics received strong support by the attending engineers. The demand for information and education has never been more evident.

The Fundamentals of RF course represented a different area of education. Les Besser's third presentation at Anaheim was attended by more engineers than any of the previous courses. As part of the course's continued evolution, this year's session included new material on toroidal transformer design and filter design. Also new were hands-on examples of Smith chart matching network design.

Several companies introduced products to the RF public for the first time at RF Expo. Avantek introduced four new silicon MMIC amplifiers in their MSA series, the MSA-0386, MSA-0486, MSA-0686 and MSA-1120. These devices have 3 dB bandwidths from 800 MHz to 1.6 GHz with various gain and power output specifications.

Amplifiers of another sort were introduced by RF Power Labs, their 300 and 900 series of power amplifiers. The 900, 901 and 902 operate from 10 kHz to 250 MHz with power outputs of 5, 10 and 20 watts, respectively. The 300 series operates in the 200-500 MHz frequency range in 100 and 200 watt power output configurations. To measure power like this, Coaxial Dynamics introduced the new Model 7510 Frequency Counter/Wattmeter. Using the line section sampler and standard detecting elements, frequency and power can be measured to over 1 GHz and 5 kW.

California Eastern Laboratories presented the NEC NE25337 low cost dual-gate GaAs FET for under-1 GHz applications. This device offers 20 dB small-signal gain and 1.1 dB typical noise figure. CEL also showed a new series of linear power GaAs FETs for 3-7 GHz applications, with up to 15 watts power output.

Doty Scientific displayed three new products: The DSI-5W-A 5-watt amplifier for 4-200 MHz broadband applications; the LN-2 series protected preamplifiers covering 5-800 MHz with input protection for 5 us pulses at 1 kW; and the SW-1 MESFET MMIC with high speed driving circuitry, connectorized package and up to 1.2 watt power handling capability.

Passive components were on hand, as well. FL Jennings showed their new TOC series miniature DPDT relays, with 1.4:1 VSWR and 0.3 dB insertion loss at 3.2 GHz. Kay Elemetrics presented the 800 series step attenuators in 0-101 and 0-141 dB ranges, with 1 dB steps.

Engineers should be looking ahead to RF Expo East, October 25-27, 1988 at the Philadelphia Civic Center; and RF Technology Expo 89, February 14-16, 1989 at the Santa Clara Convention Center. In response to continued growth in the educational needs of RF engineers, future Expos will present at least two more classroom courses, in addition to the Fundamentals of RF, and will continue to encourage industry participation with contributed papers on important RF topics.

Scientific Communications to Develop THFDF/HFC

Scientific Communications, Inc. has been awarded a $4.1 million contract by the U.S. Army for the initial design and production of a Tactical High Frequency Direction Finding Collection System. The THFDF/HFC system will consist of three (expandable to six) HF/DF and HF collection stations integrated into S-250 shelters. The system will detect, collect, determine azimuth, line of bearing, elevation angle, height of ionosphere and report data on targets or emitters using ground wave or sky wave signals in the HF range. Scientific Communications is a wholly owned subsidiary of Andrew Government Products Group, and will be joined in the project by Andrew Antennas (Melbourne, Australia) for DF technology, Harris Corp. Government Information Systems for integrated logistics support, and Codem Systems for operations software.

Japan Orders $5 Million in Tracking Terminals

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rf news Continued

for two tracking terminals, to be supplied to the Radio Research Laboratory of Japan for use in Very Long Baseline Interferometry (VLBI) experiments. The terminals will be located at the Kashima Space Station near Tokyo and on Marcus Island in the Pacific. Among the applications of VLBI are radio astronomy and terrestrial plate tectonics measurements that may lead to earlier prediction of earthquakes.

W-J Receives $9.9 Million DF Award

Watkins-Johnson Company announces the receipt of a contract to deliver AN/PRD-11 mantransportable radio direction-finding systems, and other electronic components, to the Army. Work on the contract will be performed at the company's Maryland plants.

Varian Receives Contract for GaAs Development

Varian Associates' III-V Device Center has received over $1 million in contracts from Aerojet Electro Systems to develop low-cost and reliable methods for manufacturing gallium arsenide devices. The Gunn and varactor diodes to be developed are for use in the radar systems of the Army's SADARM anti-armor munitions program. The Aerojet program is aiming for as much as an 80 per cent reduction in unit costs, while maintaining 100,000 unit per year production levels.

Plessey and AMCC Pursue Low Power ECL

Plessey Semiconductors and Applied Micro Circuits Corp. (AMCC) have agreed to jointly develop a family of high performance ECL gate arrays. The arrays will range up to 14,000 gates and are expected to be the lowest power ECL devices yet offered. The co-development agreement will combine AMCC's design and development skills with Plessey's fabrication processes. The new array will use the recently announced HE1 bipolar process, using 1 micron, triple-layer metal technology. The process produces transistors with F_T over 14 GHz.

Penstock Acquires TMA/RF

Penstock Inc. has acquired TMA/RF, an RF and microwave distributor with a territory of New York and New Jersey. In mid-1988 the name will changed to Penstock East.

Curtis Acquires RFI Filter Line

Curtis Industries, Inc. has purchased certain assets from Standex Electronics. The acquisition will enhance particularly

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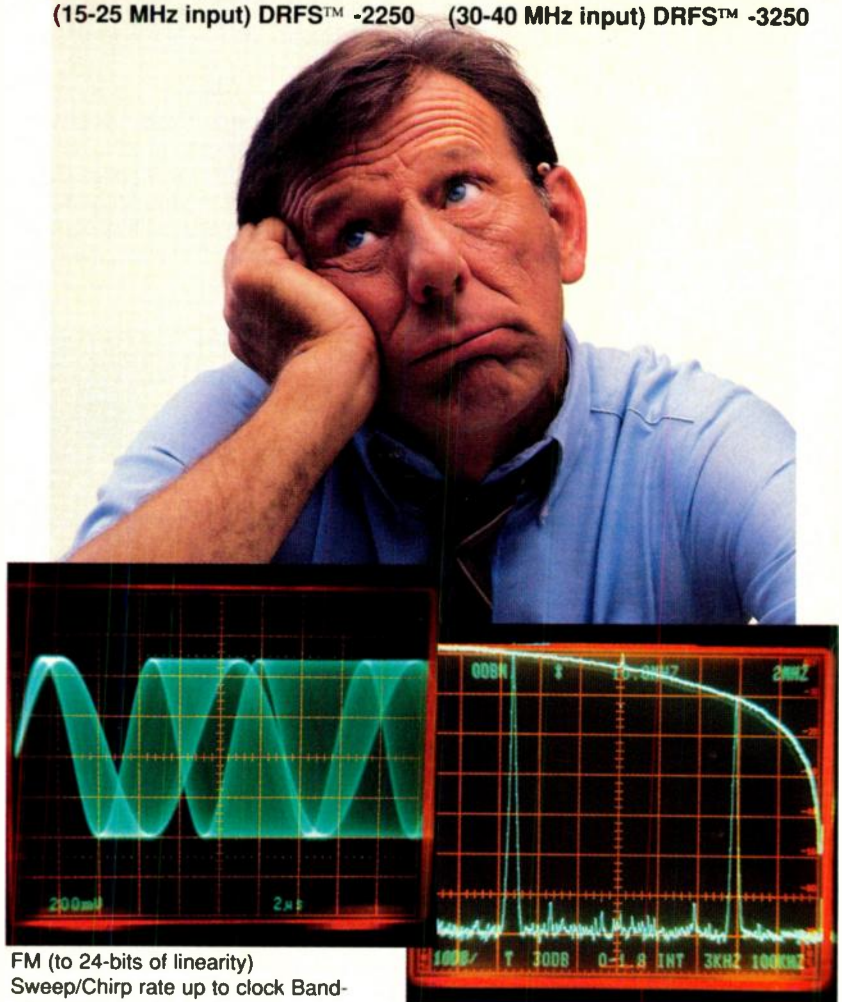

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the Curtis TEMPEST line and custom filter capabilities. Terms of the sale were not disclosed.

Book Describes RF Consulting

How To Become A Successful Consultant in RF Design contains a detailed discussion on locating clients, fees, ethics, professional advertising, developing business plans and writing contracts. The book costs $25.00 and is available from ATC Books, 804 Jordan Lane, Huntsville, AL 35816.

Singer, Dalmo Victor Awards $2.5 M Contract

M/A-Com Active Assemblies Division of Tempe, Arizona announced that it has been awarded a $2.5 million contract with options to design, develop and provide microwave assemblies for airborne receiver equipment from Singer, Dalmo Victor of Belmont, California.

Rockford Announces New Open Field Test Site

Rockford Engineering Services of Sunol, Calif., an engineering consulting firm, has completed its VDE 30-meter test site. Along with the Rockford services related to EMI are its traditional services in the areas of product safety testing (UL, ESL, CSA, TUV, VDE), power line susceptibility, electrostatic discharge test and analysis, IEEE 587 testing, special tests and calibration services and engineering consulting services.

Savoy and Hermann Lasonics Announce Joint Venture

Savoy Electronics, Inc., a subsidiary of Dixcom, Inc. has announced an agreement for a joint venture with Hermann Lasonics, Inc. optical fabrications of Taiwan, R.O.C. The agreement will provide manufacturing support for the growing involvement of Savoy in the optics and electro-optics field. With lapping and material handling experience, the agreement expands Savoy's capabilities to manufacture and market miniature optical windows, flats, and specialized substrates from crystal and optical glass, as well as aluminia and other ceramics.

Hermann Lasonics, on the other hand, will provide extensive production capabilities as well as expansion to products including lenses, mirrors, prisms and lens assemblies.

Dynascan to Buy Lloyd's Electronics

Dynascan Corp., a Chicago-based consumer electronics corporation, has entered into a contract to purchase certain assets of Lloyd's Electronics of Edison, N.J.

Lloyd's Electronics distributes a broad range of consumer electronics products domestically as well as in South America and in under the brand name Lloyd's.

Triax to Acquire COMSAT Unit

Triax Corp. and COMSAT Corp. announced that they have signed an agreement by which Triax will acquire the business of Amplica, Inc., a COMSAT subsidiary. Amplica manufacturers a wide range of microwave amplifiers and related subassemblies and subsystems for the defense electronics industry. Amplica will be operated as a wholly owned subsidiary of Triax and will continue to pursue its current business.

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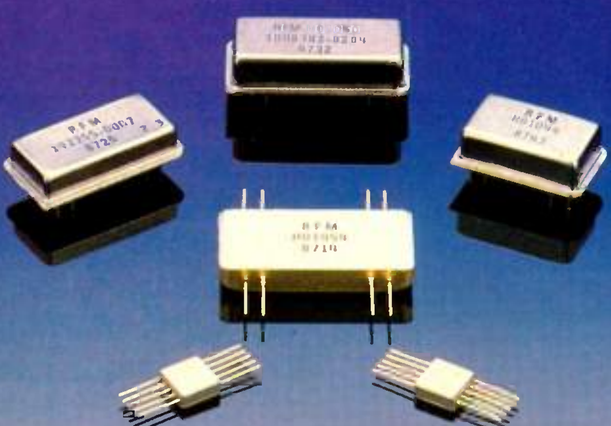

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Digital Modulation Using the NCMO™

Numerically Controlled Modulated Oscillator (NCMO) Creates Digitally Defined RF Waveforms.

By Robert J. Zavrel Jr.
Digital RF Solutions Corp.

The technique of direct digital signal synthesis (DDS) has recently become a viable technique for RF designs. Recent advances in digital-to-analog converter (DAC) technology, large scale integrated controllers and memory circuits have improved performance and raised operating frequencies. This article presents the fundamentals of an integrated controller that affords remarkable flexibility for synthesizer design.

The NCMO™ represents a highly integrated solution for configuring direct digital synthesis systems. Unlike many DDS approaches, the NCMO allows easy implementation of frequency, phase, and/or amplitude modulations. Combinations of these modulation techniques allow synthesis of nearly every kind of waveform because AM can be simultaneously combined with FM and/or PM. Consequently, the two great problems of RF synthesizer design, modulation and frequency agility, are solved in one monolithic CMOS device with low power consumption and easy interfacing capabilities.

Direct Digital Synthesis With the NCMO System

Unlike analog LC or crystal oscillators or phase-locked loop (PLL) synthesizers, all the parameters of the signal (ampli-

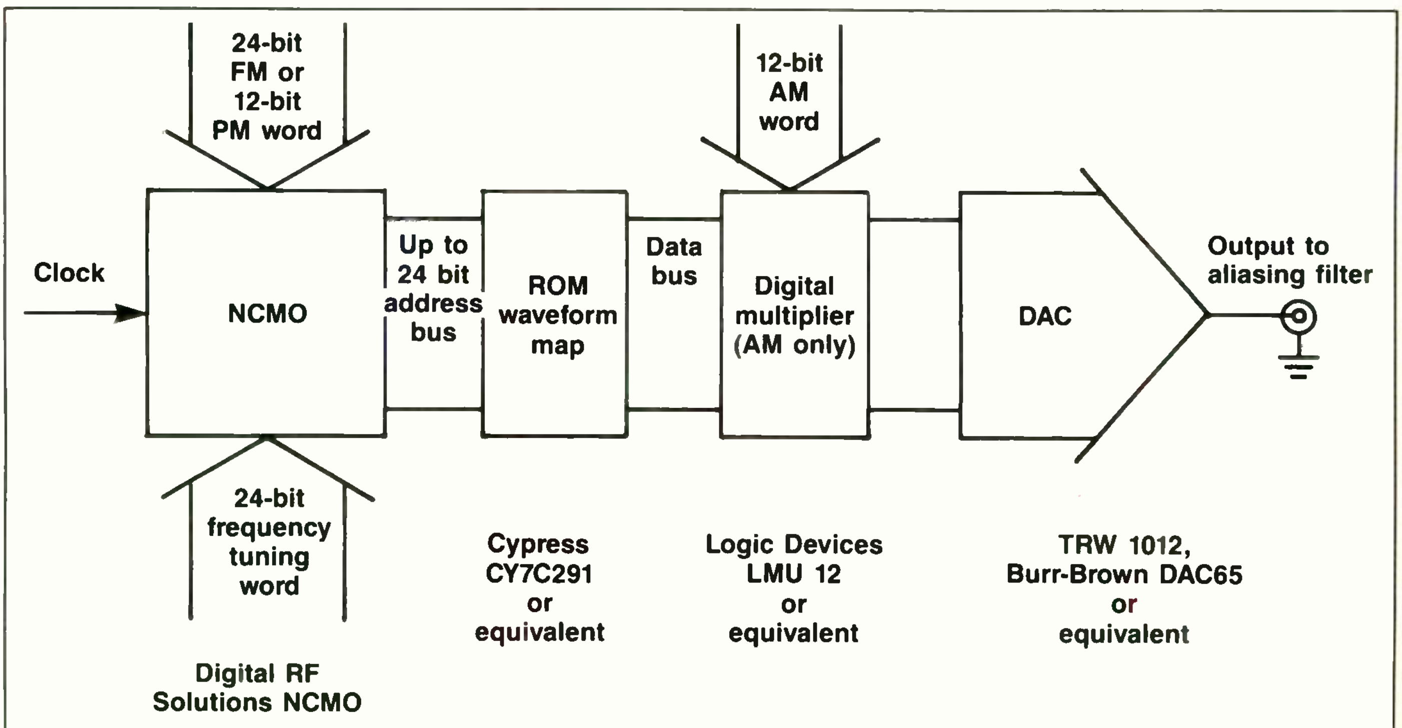

Figure 1. 4-chip DDS system using the NCMO.

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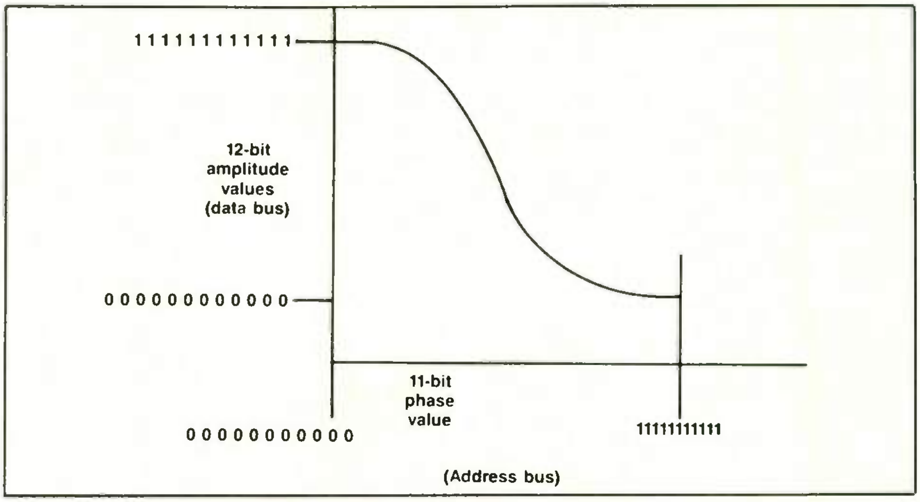

Figure 2. ROM waveform map plotting data bus values versus address locations.

tude, phase and frequency) are defined by digital numbers. Figure 1 shows a four IC DDS system implementation using the NCMO as the system core.

At the heart of the NCMO is the phase accumulator. For a fixed clock frequency (assume 20 MHz) there will be a corresponding fixed phase change for any desired output frequency (36 degrees for 2 MHz, 1.8 degrees for 100 kHz, etc.) every clock cycle. The phase accumulator accepts a 24-bit word which results in 24-bit frequency resolution. The NCMO then cyclically accesses ROM locations. The ROM acts as a "waveform map" (Figure 2), it simply converts the phase position (calculated in the NCMO) to the appropriate instantaneous sinusoidal amplitude value. Stated another way, the ROM *address* corresponds to time and phase, while ROM *data* corresponds to the appropriate amplitude value. The ROM is actually a digital phase-to-amplitude converter.

The ROM typically contains 180 degrees of phase information. Cosine data is preferred to sine data since it is better to have amplitude errors at "peaks and valleys" rather than zero crossing points. The NCMO clocks the ROM address locations up and then back down to complete a whole cycle. This data is applied to the DAC. A 12-bit DAC system requires a ROM with 11 address bits and 12 data bits.

Some of the interesting variations on the basic sinusoidal waveform include:

1. *Quadrature Synthesis:* Two ROMs with sinusoidal data in quadrature and two DACs can affect two signals which remain in quadrature at any output frequency.

2. *Random Phase Relationships:* Two signals of the same frequency but any phase relationship can be synthesized by using two NCMO systems driven off the same clock. The phase relationship between the two signals can then be stepped with 24-bit resolution and accuracy by clocking a single bit on the NCMO.

3. *Complex Waveform:* Finally by replacing the ROM with a RAM any waveform can be entered and realized, within the Nyquist limit set by the clock frequency. The NCMO can be configured to count through an entire 360 degree cycle for asymmetrical waveform synthesis.

Comparison with PLL Synthesizers

PLL synthesizer performance has approached theoretical operating limitations given loop bandwidths, settling times and phase noise trade-offs. In contrast, DDS techniques are limited mainly by DAC performance. Recent advances in the rapidly evolving DAC technology have sharply narrowed the PLL performance advantage. There are several clear technical advantages of NCMO DDS over PLL synthesizers and the remaining advantages of PLLs are subject to constant erosion. Some of the existing DDS advantages include:

1. *Near Zero Settling Time for Frequency Hops:* It takes two clock cycles for the new frequency word to be loaded into the NCMO. There are no loop filter time constants to contend with. The NCMO simply begins sequencing the new number values and the DAC converts these into the appropriate instantaneous analog signal. There is no signal "inertia" as in linear (PLL) systems.

2. *Phase Continuous Frequency Hops:* Undesirable glitches from short-duration pulses are eliminated as the new frequency sine wave begins from the last discrete amplitude value. Consequently the system is truly phase continuous or "syn-

chronous."

3. *24-Bit FM Modulation:* Input up to 24 bits of FM modulation either from a counter (for sweeps) or from ADCs to affect FM over the entire bandwidth with 24-bit linearity! These specs are simply not possible using any analog modulator. Center frequency and modulator parameters are now software problems.

4. *12-Bit Phase Modulation:* "Continuous" or discrete phase shifts can be created for analog PM or binary, QPSK, octal, 16, 32 or whatever point digital PM you desire. Simultaneous AM permits very complex data communication constellations.

5. *Frequency and Phase Deviation is Independent of Center Frequency:* Move the center frequency over octaves and the deviation remains constant. Modulation in the NCMO is an additive process, not a scaling multiplicative process.

6. *Phase Noise is Determined by the Clock Reference:* Below ¼ the clock frequency, all spurious signals are typically 70 dB below the desired signal using 12-bit DACs.

There are applications where PLLs still have clear advantages. The speed and resolution of state-of-the-art DACs limit the upper frequency operation of the basic NCMO system. Higher frequencies can be synthesized by combining PLL and DDS techniques, however. Also, PLLs still provide better performance in very low power applications, and where large frequency step sizes are required.

In many applications the advantages of both DDS and PLL techniques can be combined for optimum performance. For example, an NCMO can be used to generate an FM signal with 1 or 2 Hz resolution in the 5 MHz region. The 5 MHz signal is then used as the reference input to a PLL multiplier. The multiplier can have a fixed "N" value, say 128 for an output frequency of 640 MHz. The resolution of the synthesizer is now about 128 MHz, the low N value minimizes phase noise, modulation is accomplished, and PLL design is greatly simplified.

The combination of these two techniques can often optimize system design. DDS does not represent a direct replacement for PLLs, but does represent a better technique for many applications.

Tuning

The 24-bit frequency word can be presented to the NCMO by three methods: 24-bit parallel, three 8-bit words for microprocessor interface, or by two-bit quadrature serial sequencing allowing interface to rotary optical encoders. These encoders are typically mounted on a shaft and allow tuning similar to a radio dial. The tuning rate of the serial mode is selectable through a three-bit word. The 24-bit parallel mode is used where maximum speed is required for center frequency changes, as in frequency hopping radios. The eight-bit mode is used where easy interface to a microprocessor is desired. Finally, the serial mode is used where the synthesizer needs to be tuned by a dial.

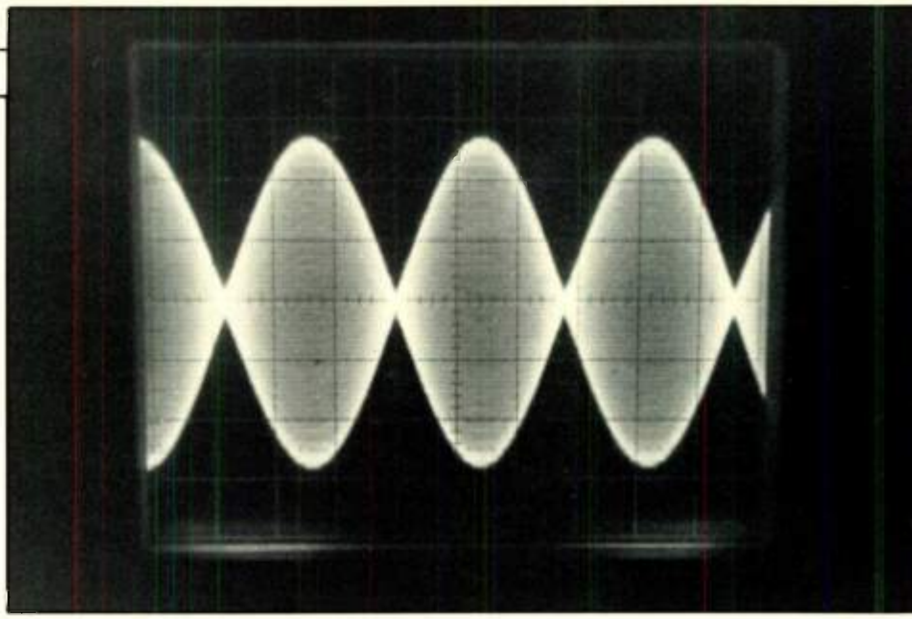

Figure 3. A DSB suppressed-carrier AM envelope.

Twenty-four-bit frequency resolution represents about 16.8 million possible frequency values. However, if the MSB on the tuning word is set high, the NCMO will attempt to synthesize a frequency above the Nyquist Frequency (one-half the clock frequency). This will produce a "negative" frequency response which will appear under the Nyquist value. Consequently, there are actually about 8.4 million possible frequency values under the Nyquist Frequency. The frequency resolution of the DDS synthesizer equals clock frequency/16.777216 MHz. If the clock frequency is 16.777216 MHz, the NCMO synthesizer will output a signal selectable in exactly 1 Hz steps. Finally, if this is not sufficient resolution, two NCMOs can be cascaded for nanohertz resolution with no consequence to phase noise performance.

Modulation

The NCMO has on-board FM and PM capabilities with 24-bit resolution. When the FM mode is selected, the 24-bit modu-

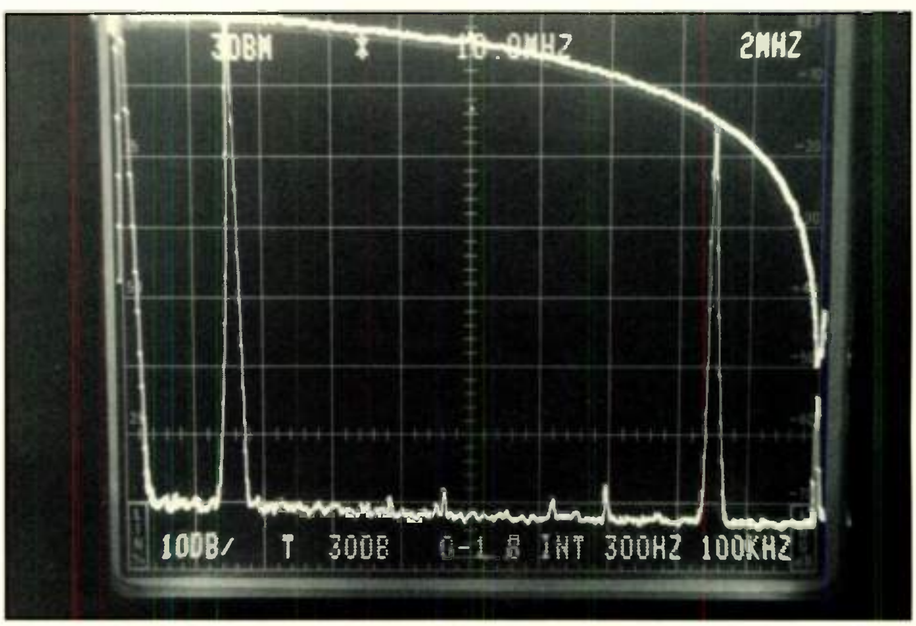

Figure 4. A 3 MHz signal with first alias frequency, using a 20 MHz clock. Note spectral purity using the TRW 1012 12-bit DAC.

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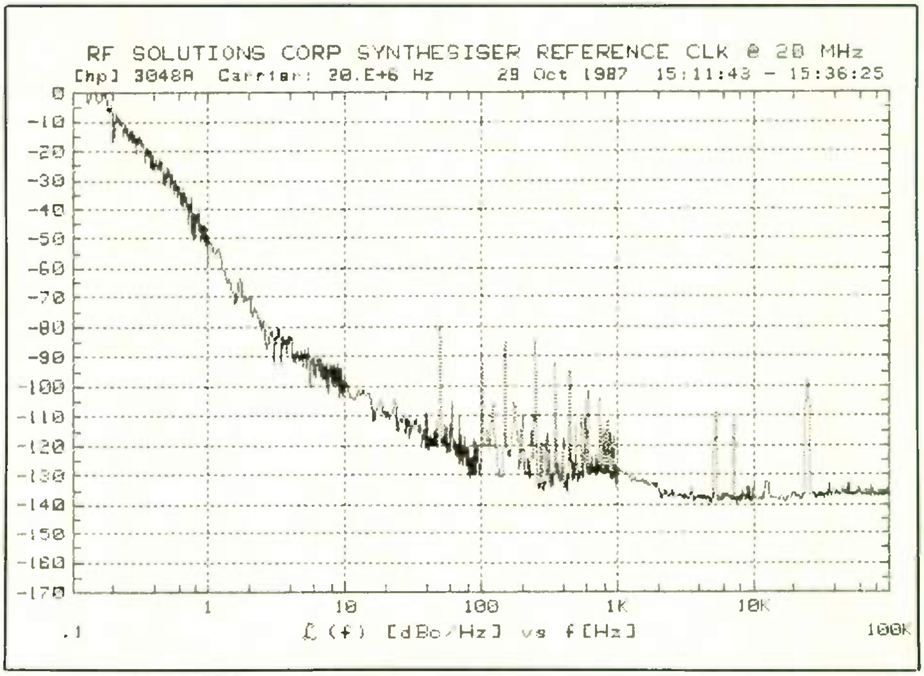

Figure 5(a). 20 MHz clock spectral noise.

lation port number is simply added to the tuning word before the phase accumulator, thus "pulling" the frequency "up." In the FM mode the modulation and tuning ports are interchangeable. Simultaneous FM and PM is possible by frequency modulating the tuning word and phase modulating the modulation port (if anyone would ever want to do such a thing). For phase modulation, 0 to 360 degree deviation is possible with 4096 phase steps. In the PM mode the modulation word is added to the output of the phase accumulator thus advancing the phase the appropriate number of degrees. The FM and PM modes are selected by a single pin level.

Digital control of modulation parameters has some very desirable advantages. Since the numbers sum to frequency values, frequency deviation can be held constant over the entire tuning range of the synthesizer. Bandwidth limitations are absolute, therefore preventing excessive deviations and overmodulation. Furthermore, performance will be identical from one unit to another eliminating many production line calibrations and real world operating digressions.

Finally, amplitude modulation can be affected by inserting a digital multiplier between the waveform map ROM and the DAC. The multiplier then adjusts the digital amplitude word applied to the DAC according to the modulating waveform. A four-quadrant multiplier will produce a DSB suppressed carrier signal, while a single-quadrant multiplier will produce a full carrier AM signal. Of course, AM can be accomplished in parallel with PM and/or FM for a wide range of vector modulation schemes. Figure 3 shows an envelope of a DSB suppressed carrier waveform using a sine wave modulator.

Spectral Purity

Figure 4 shows a typical spectral display of a 12-bit NCMO DDS synthesizer. There are four spurious categories to consider in DDS applications:

- broadband phase noise
- non-harmonic discrete spurs
- harmonic spurs
- aliasing responses

Broadband phase noise has been shown to be determined by the clock reference phase noise characteristics. Plot the phase noise of the reference in dBc/Hz versus frequency offset and then use the 20log(N) relationship to calculate broadband phase noise. Quantizing and DAC errors are less significant than the reference noise performance for medium-performance clock references. Figure 5 shows a spectal noise comparison of a 20 MHz clock reference and a 1 MHz DDS synthesized signal. The phase noise relationship between the clock and synthesized signals conform to 20log(N) function.

The most restricting spurious signals are the close-in discrete non-harmonic responses. As the phase accumulator runs up and down the address locations, corresponding DAC values are summoned. Eventually exact sequence values

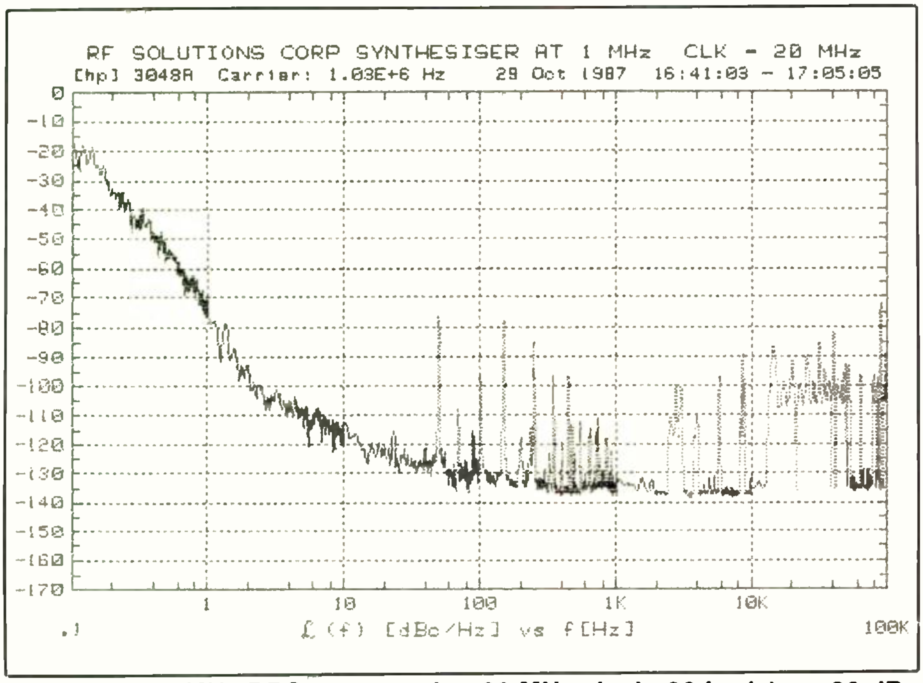

Figure 5(b). 1 MHz DDS output using 20 MHz clock. 20 log(n) = −26 dB, n = 1/20.

are repeated. Due to differential nonlinearity in the DAC these errors sum at the periods of this repetition. These errors are related to sub-harmonics of the carrier and appear as phase and/or amplitude sidebands or even separate signals. At this time we are quantifying the relationships of quantizing error, DAC error and spurious responses. We would certainly like to hear from anyone who has done similar work in this area.

Harmonic responses are related mainly to the DAC integral nonlinearity. They are generally not as important as the close-in spurs since they are more easily filtered. Often they are of little concern, as in switching mixer LO signal sources, for the levels typically encountered in DDS synthesis.

Aliasing signals will appear above the Nyquist Frequency with the same spacing above and below the clock frequency as the fundamental is above DC. For example, a system generating 3.6 MHz from a 20 MHz clock will produce a set of aliasing signals at 16.4 and 23.6 MHz. Figure 6 shows that the amplitude values of the aliasing signals conform nicely to a Sin(X)/(X) relationship.

Care must be taken when measuring spurious responses. Sometimes the levels of the DAC's output spectrum will be high enough to cause cross-modulation within the test instrument. Take care that you are measuring the device under test and not the test instrument! A step attenuator in-line will provide a quick check to see if there is a first or higher order response in effect. A common technique for suppressing spurs is to run the DDS output signal through a PLL employing a narrow loop filter.

Applications

The two most difficult problems facing the synthesizer system designer are modulation and frequency agility. Most of the performance limitations of synthesizers can be traced to these two requirements. The great power of the NCMO DDS system is that both functions are performed simultaneously. Furthermore, these two functions are accomplished with precision and linearity that is impossible with analog techniques. All known modulation techniques consist of AM, FM or PM or combinations of these. All three of these modulation techniques are simultaneously available with the NCMO system. Therefore, nearly any conceivable complex or simple modulated waveform is possible within the system bandwidth limitations. Some obvious RF applications might include:

- Radio receiver LOs
- RF test sweep signal generators
- Quadrature synthesis
- FM stereo broadcast generators
- Frequency hopping radios
- Video waveform monitors
- Video signal generators
- High speed data communications

Similar diverse applications exist for audio applications, where slower clock

The proof.

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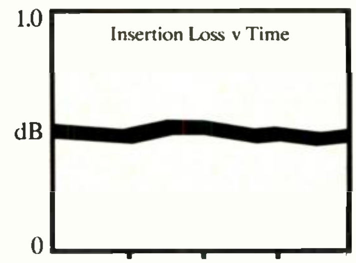

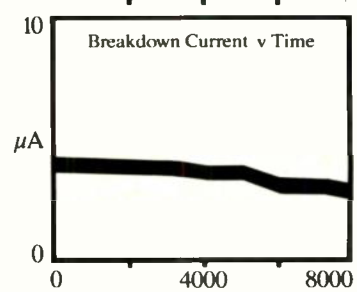

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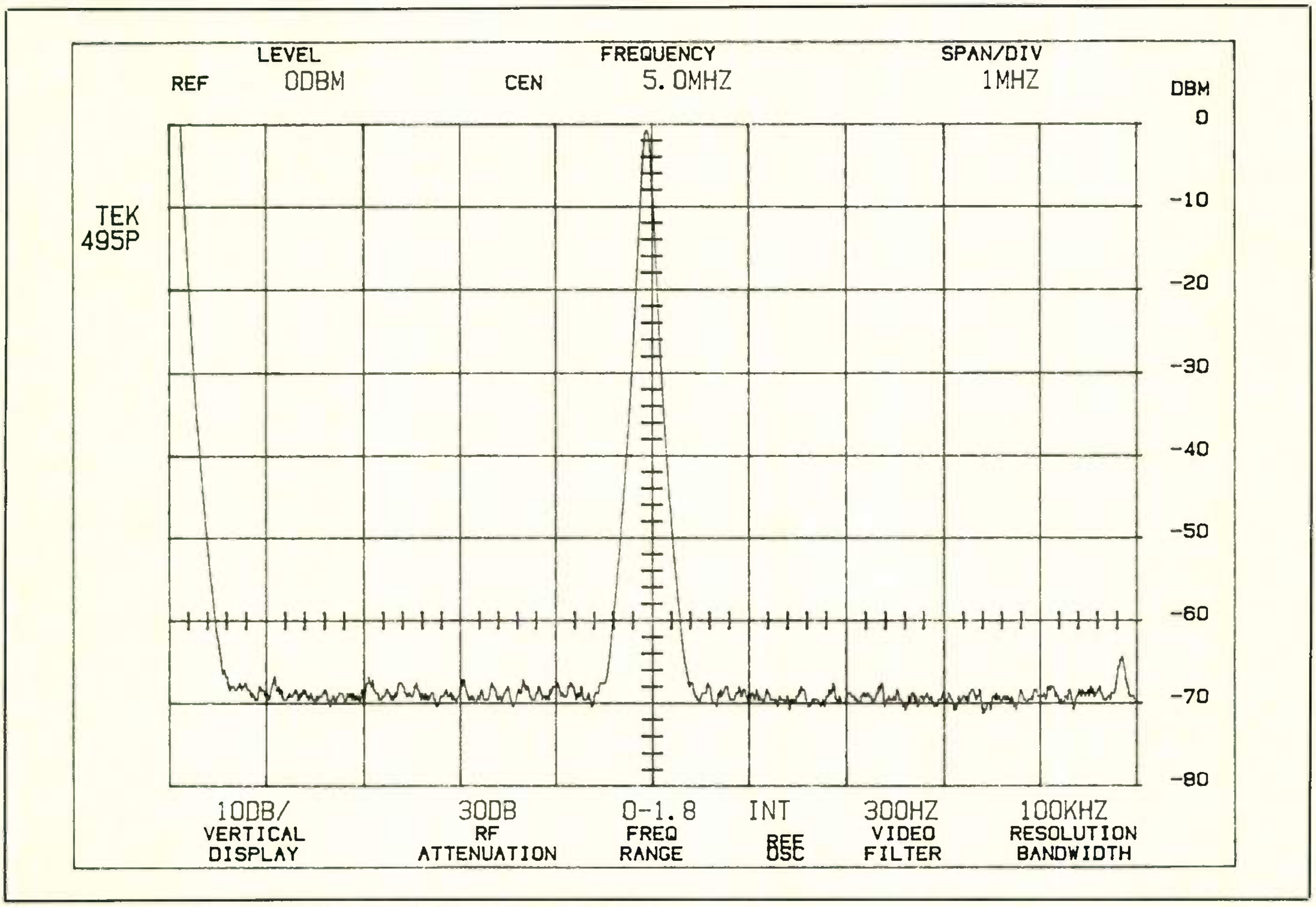

Figure 7. Typical spectra of a 12-bit NCMO DDS system.

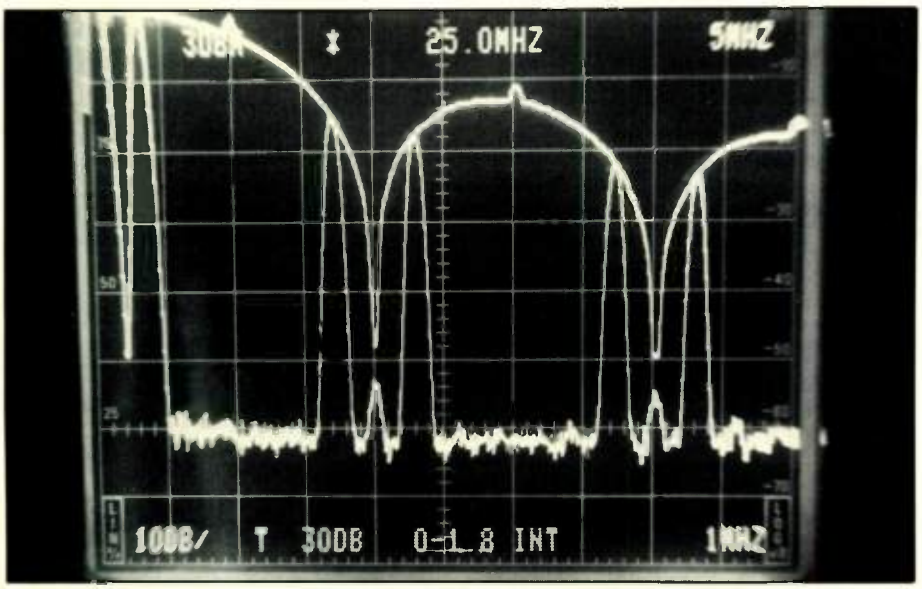

Figure 6. 3 MHz signal and aliases, showing Sin(X)/(X) response of the DDS. Glitches at the multiples of the Nyquist Frequency are due to summing of fundamental and alias signals.

frequencies allow for 16 and more bit DACs.

- Music synthesizers
- High performance audio test sweep oscillators

Conclusions

The NCMO represents a highly integrated solution to building direct digital synthesizers. The IC includes the necessary accumulator and interface logic to control a DDS system. Although still in a developmental stage, this device also includes a digital modulator thus affording the designer great flexibility. Advances in DAC technology have recently provided designers with devices that promise excellent performance even in demanding communications applications.

For more information about the NCMO DDS, contact Digital RF Solutions Corp. at 3080 Olcott St., Suite 200D, Santa Clara, CA 95054-3209. Their telephone number is (408) 727-5995. Information can also be obtained by circling INFO/CARD #250. rf

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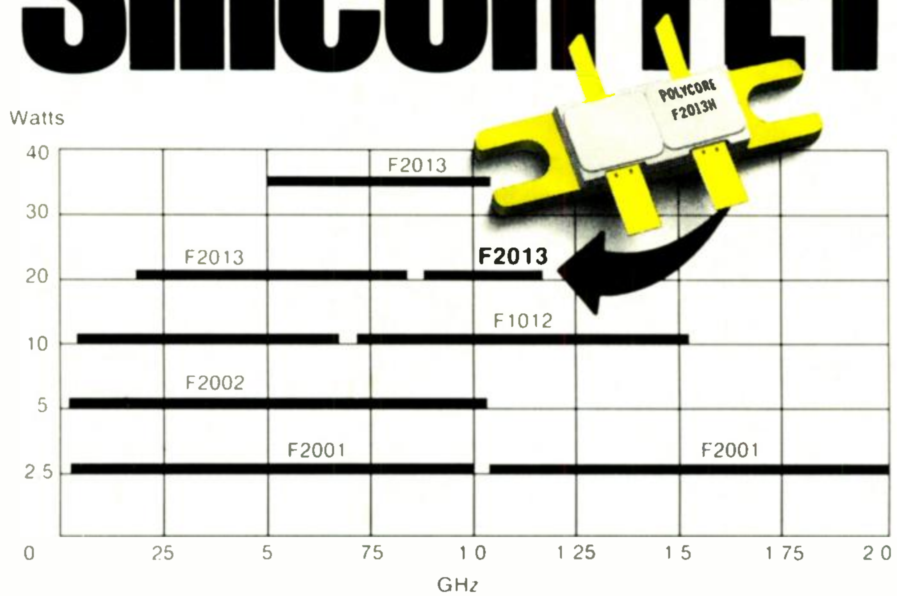

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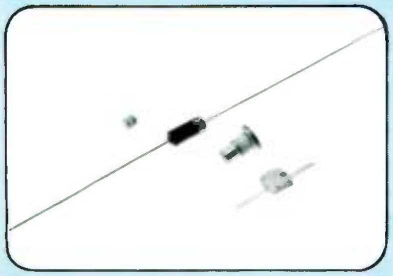

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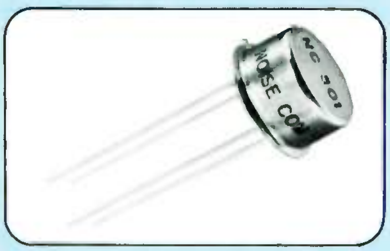

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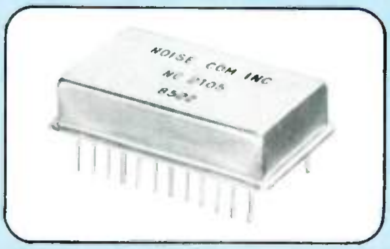

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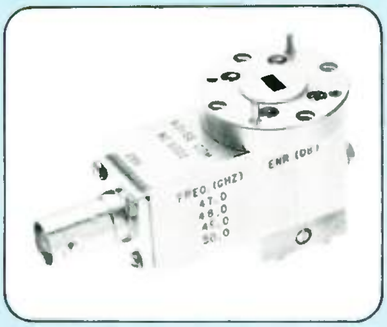

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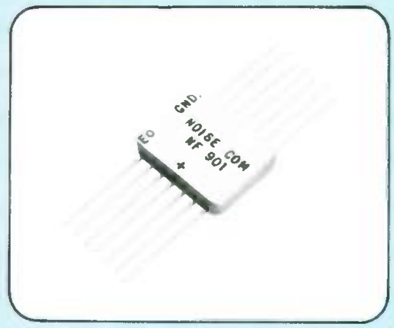

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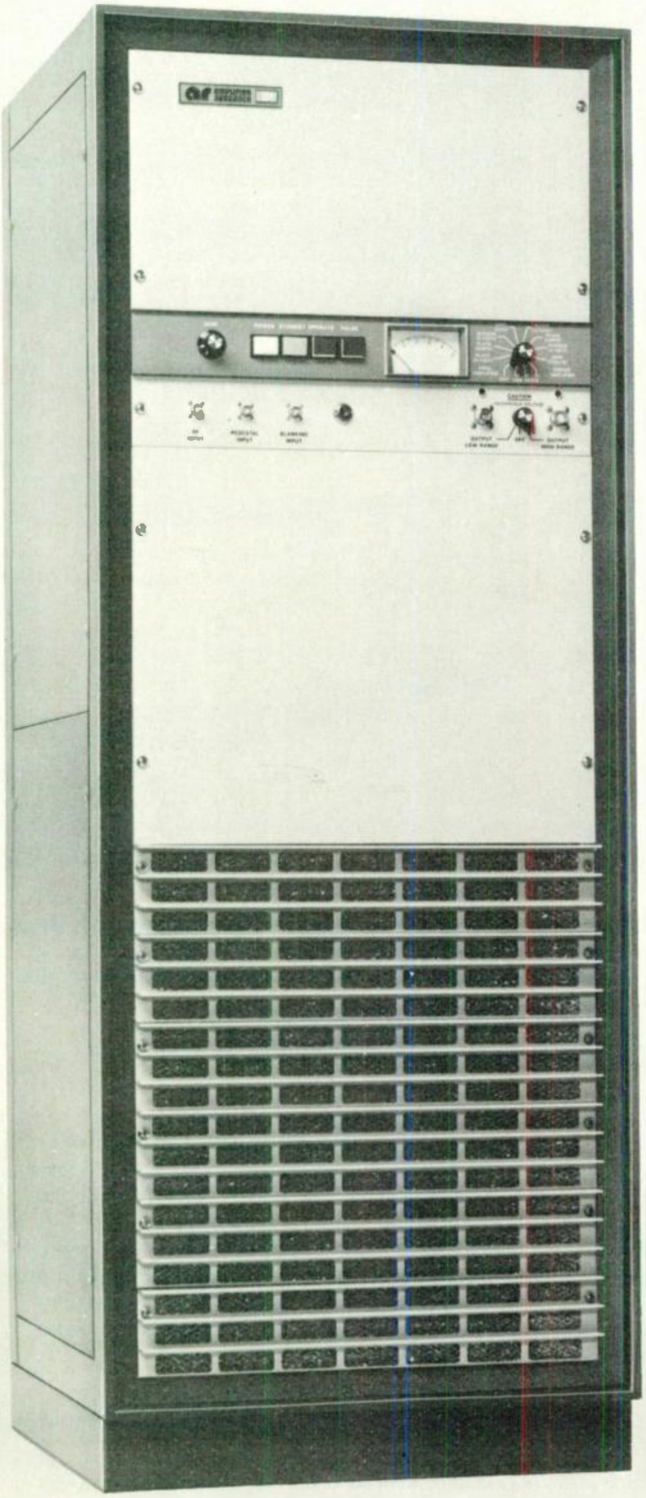

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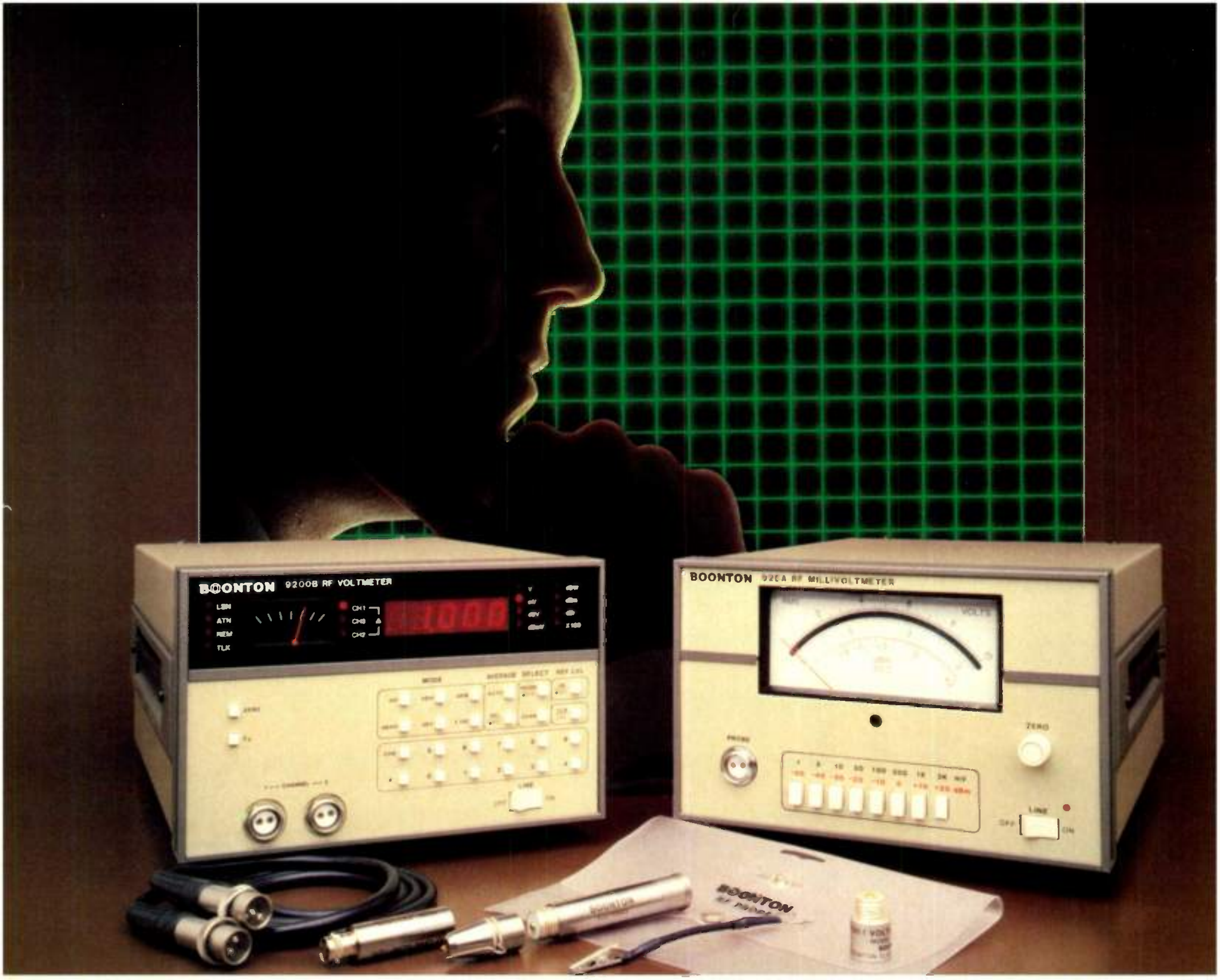

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INFO/CARD 41

Stagger-Tuned Bandpass Active Filters

By Jack Porter
Cubic Defense Systems

Although bandpass filters formed by cascading single-tuned amplifiers have been studied for more than half a century, (1, 2) they were not widely used until active filter circuits consisting of resistors, capacitors, and integrated circuit operational amplifiers became practical. Since each stage is tuned to a different frequency, such filter circuits are termed stagger- tuned.

These op-amp circuits, which reproduce the theoretical frequency response curves very accurately, can also provide voltage gain. They are small in size, even at low frequencies, and can be fabricated as hybrid integrated circuits. Any of the conventional all-pole filter responses, such as Butterworth, Chebyshev or Bessel, can be synthesized in this manner. Each pole of the low-pass prototype filter is transformed into a single-tuned bandpass filter section. The center frequency and Q of each bandpass section are uniquely determined by the center frequency and bandwidth of the filter and the location of the corresponding low-pass prototype pole. Filters containing more than five of these sections are rarely practical, since the passband shape is quite sensitive to changes in center frequency of the individual filter sections. If the filter requires high Q's and must operate over a wide temperature range, no more than three sections should be considered. Multiple-feedback (e.g., leapfrog) filters (3), which are more complex and difficult to align, should be used if more poles are necessary. Designing such filters consists of the following steps: 1. Specify the filter center frequency, bandwidth, and overall voltage gain. 2. Select the type of response and the number of poles. 3. Determine the low-pass prototype pole locations. 4. Calculate the center frequency and Q of each section. 5. Select a filter circuit and the type of operational amplifier to be used in it. 6. Determine the voltage gain for each section. 7. Select capacitor values and calculate resistor values.

Type and Number of Poles

Selection of the response type and number of poles is based on the required attenuation and group delay characteristics. Plots of these for various filters can be found in reference 4, and methods of calculating them in reference 5.

Low-Pass Prototype Poles

Butterworth and Chebyshev pole locations can be calculated using the formulas in Figure 1, or they can be obtained from tables. Bessel filter poles are shown in Table 1.

Center Frequencies and Q's

The formulas used in the center frequency and Q calculations are summarized in Figure 2. The frequency F_f is the center of the filter passband. The frequency F_c which is used in all of the calculations is slightly different, since the passband is geometri-

For an n-pole Chebyshev filter with A_m dB peak-to-peak passband ripple and 1 radian/sec ripple (A_m dB) bandwidth:

$$A_p = 10\,(A_m/10) \qquad \varepsilon = \sqrt{A_p - 1}$$

$$a = \frac{1}{n}\sinh^{-1}\left(\frac{1}{\varepsilon}\right)$$

$$b = \sinh a \qquad c = \cosh a$$

For an n-pole Butterworth filter with 1 radian/sec half-power (3 dB) bandwidth:

$$b = 1 \qquad c = 1$$

The low-pass prototype filter poles are:

$$P_k = -\sigma_k \pm j\omega_k \qquad k = 1, 2, \ldots, n_2$$

$$\sigma_k = b \sin\left[\frac{(2k-1)\pi}{2n}\right] \qquad n_2 = \frac{n}{2} \text{ for n even}$$

$$\omega_k = c \cos\left[\frac{(2k-1)\pi}{n}\right] \qquad n_2 = \frac{n+1}{2} \text{ for n odd}$$

The Chebyshev filter half-power bandwidth is:

$$\omega_H = \cosh\left[\left(\frac{1}{n}\right)\cosh^{-1}\left(\frac{\sqrt{A_H - 1}}{\varepsilon}\right)\right]$$

$$A_H = 2 \text{ for n odd}$$

$$A_H = 2A_p \text{ for n even}$$

Divide the ripple bandwidth values for σ_k and ω_k by ω_H to obtain the 3 dB bandwidth values.

Figure 1. Pole locations — Chebyshev and Butterworth filters.

cally symmetrical about F_c. Two filter sections, one tuned above F_c, the other below, correspond to each pair of complex conjugate low-pass prototype poles. Each pole on the real axis has a corresponding section tuned to F_c. The formulas for calculating the center frequency and Q of each filter section are derived in reference 6. A different algorithm due to Geffe (7, 8), is also often used. It requires about the same amount of computation and produces identical results.

Circuits and Amplifiers

The single-amplifier biquadratic circuit shown in Figure 4, usually referred to as the Deliyannis-Friend circuit (9, 11), is the one most commonly used for bandpass filters. More elaborate multiple-amplifier circuits are sometimes necessary to realize very high Q's, but stagger-tuned filters don't usually require them. The op-amps used in these filters should have approximately 90 degrees open-loop phase shift at the center frequency of the filter section and they should be stable when used as unity-gain amplifiers. The maximum frequency at which an op-amp can be used is determined by its slew rate and its open-loop gain at the operating frequency. Slew rate is usually the limiting factor, so a general rule is that the full-power bandwidth of the op-amp should be greater than the upper limit of the filter passband. Suitable op-amps vary greatly in full-power bandwidth, from about 50 kHz to several MHz.

Gain Distribution

There are various criteria for determining the optimum way of distributing the filter gain among the various stages. The most common one, which provides maximum dynamic range, is to adjust the stage gains for equal maximum output voltage at each one when an input signal is swept across the passband. This maximum output voltage occurs at or near the frequency to which that stage or one of the preceding stages is tuned, thus it is necessary to calculate the gain of each stage at its center frequency and at the center frequency of every other stage. The total gain of stage and all preceding stages is then calculated at its center frequency and at that of the preceding stages. This is followed by selecting the gain of that stage so that the highest gain at any of the frequencies is equal to the total gain of the filter. The gain of the final stage is selected to provide the specified total gain at F_f. The required calculations are summarized in Figure 3. It's not always practical to select the stage gains in this manner, since the gain of the first stage at its center frequency is equal to the total filter gain. If, for instance, a two-stage filter with a gain of 100 is required, it makes more sense to select a gain of 10 for each stage. The way in which the stages are ordered is not extremely important, but there is some advantage in placing stages with the same Q next to each other, since their center frequencies are symmetrical about F_c. The stages with highest Q are often put at the input, since these have inherently high gain, and the first stage usually has the highest gain.

$-\sigma_p \pm j\omega_p$: low-pass prototype poles.

F_f : Filter center frequency in Hz.

BW: Filter bandwidth in Hz.

$$F_c = \sqrt{F_f^2 - \left(\frac{BW}{2}\right)^2} \quad \text{(design center frequency)}$$

$$Q_F = \frac{F_c}{BW}$$

$$\sigma = \frac{\sigma_p}{Q_F} \qquad \omega = \frac{\omega_p}{Q_F}$$

$$t = \frac{4 - \sigma^2 + \omega^2}{2}$$

$$u = \sqrt{t + \sqrt{t^2 + \sigma^2 \omega^2}} \qquad \text{For } \omega_p = 0\text{:}$$

$$v = \frac{\sigma\omega}{u} \qquad F = F_c$$

$$Q_1 = Q_2 = \frac{1}{2}\sqrt{1 + \frac{u^2}{\sigma^2}} \qquad Q = \frac{1}{\sigma}$$

$$F_1 = F_c\, Q_1\, (\sigma - v)$$

$$F_2 = F_c\, Q_2\, (\sigma + v)$$

Figure 2. Center frequencies and Qs.

A_f: Gain of the entire filter at F_f, the filter center frequency

A_{kk}: Gain of stage k at F_k, its center frequency

A_{kj}: Gain of stage k at F_j, the center frequency of another stage

A_{k0}: Gain of stage k at F_f

$$\text{Define: } T_{kj} = \frac{Q_k\,(F_k^2 - F_j^2)}{F_k\, F_j}$$

$$G_{kj} = \frac{A_{kk}}{A_{kj}} = \sqrt{1 + T_{kj}^2}$$

for $k = 1$ to n, $j = 0$ to n; $A_{00} = 1$, $G_{kk} = 1$

n: Number of low-pass prototype poles.

If $n > 1$: $A_{11} = A_f$ $\qquad A_{10} = \frac{A_{11}}{G_{10}}$

If $n > 2$: For $k = 2$ to $n-1$:

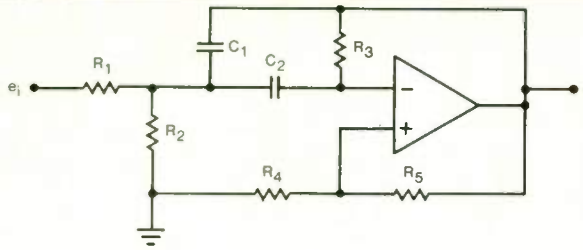
$$\text{Evaluate } G_j = \prod_{i=1}^{k} G_{ij} \qquad \text{For } j = 1 \text{ to } k$$

G_{min} is the minimum value of G_j

$$A_{kk} = \left[\frac{A_f}{\prod_{j=1}^{k-1} A_{jj}}\right] G_{min} \qquad A_{k0} = \frac{A_{kk}}{G_{k0}}$$

$$A_{n0} = \frac{A_f}{\prod_{j=0}^{n-1} A_{j0}} \qquad A_{nn} = A_{n0}\, G_{n0}$$

Figure 3. Stage gains for maximum dynamic range.



F_a = Op-amp gain bandwidth (MHz)

$$T_a = \frac{10^{-6}}{2\pi F_a} \qquad K = 1 + \frac{R_4}{R_5}$$

$$k = \frac{C_2}{C_1} \qquad \text{Usually } k = 1$$

$$\frac{e_o}{e_i} = \frac{-\left[\dfrac{Ky_1}{C_1 + KT_a\, y_T}\right] s}{s^2 + \left[\dfrac{(k+1)y_3 - k(K-1)y_T}{k\,(C_1 + KT_a y_T)}\right] s + \left[\dfrac{y_T y_3}{kC_1\,(C_1 + KT_a y_T)}\right]}$$

$$\frac{e_o}{e_i} = \frac{-A_o\, D\omega_o s}{s^2 + D\omega_o s + \omega_o^2}$$

$$\omega_o = 2\pi F_o \qquad D = \frac{1}{Q}$$

A_o: gain at ω_o

$\dfrac{1}{T_a\omega_o}$: Op-amp gain at ω_o

Calculate: $a = \dfrac{k + 1}{4\,k\,T_a\,\omega_o}$

$$K = 1 - \frac{D}{\sqrt{ak}} + \frac{k+1}{ak}$$

If $a > 40$, set $a = 40$

If $K < 1$, set $K = 1$, $R_4 = 0$, $R_5 = \infty$

Select C_1, C_2, R_5

Usually $C_2 = C_1$

Calculate: $a_T = (k + 1)\,\omega_o^2\, C_1^2$

$$b_T = [(k + 1)\, KT_a\omega_o - D]\,\omega_o C_1 \qquad c_T = 1 - k - kDT_a$$

$$y_T = \frac{2a_T}{\sqrt{b_T^2 - 4a_T c_T} - b_T}$$

$$y_1 = A_o\, D\omega_o \left(\frac{C_1}{K} + T_a y_T\right)$$

$$y_2 = y_T - y_1$$

$$y_3 = k\omega_o^2\, C_1 \left(\frac{C_1}{y_T} + KT_a\right)$$

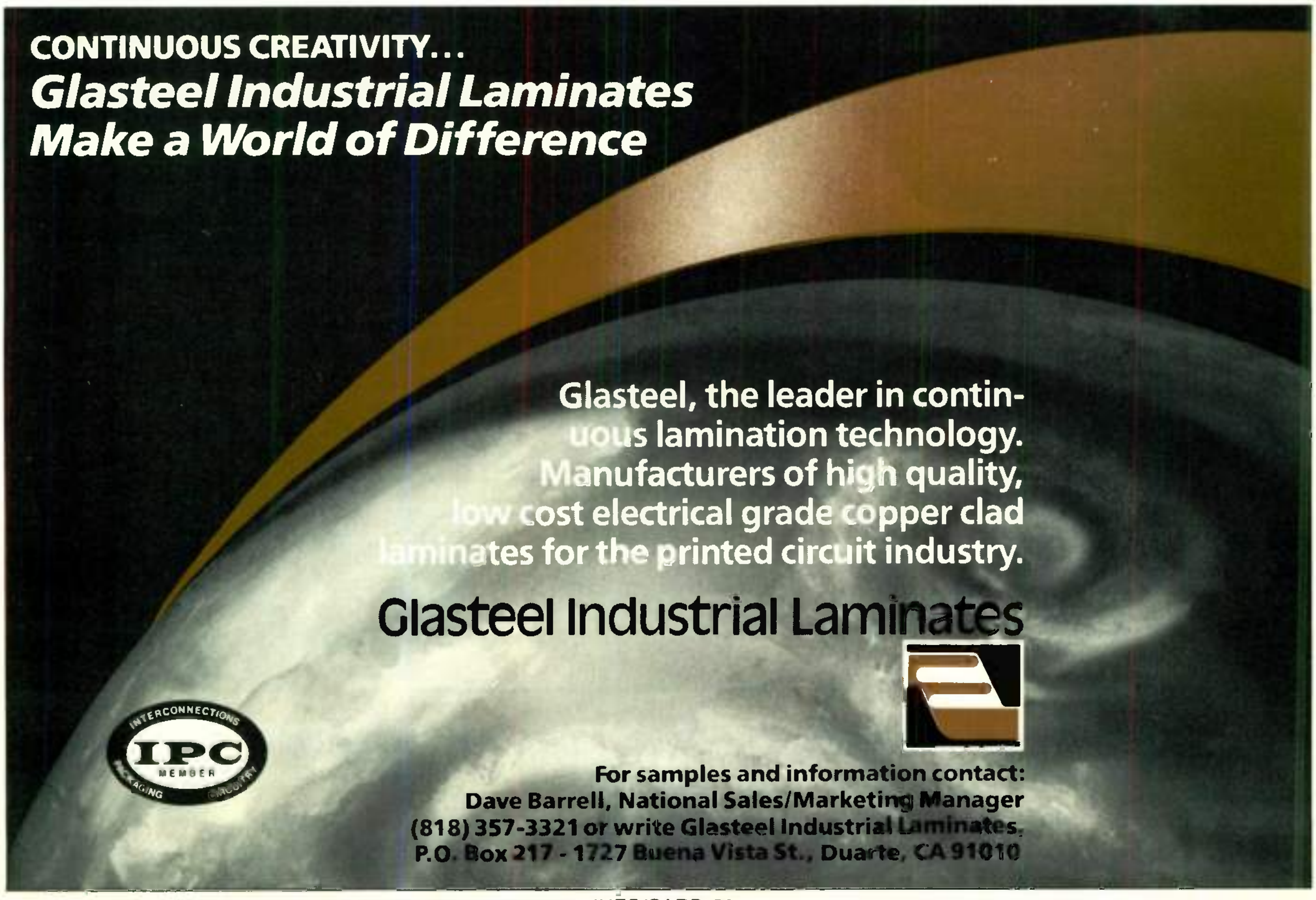
Approximate correction for capacitor dissipation factor:

$$y_3' = y_3 - .01\, D_c \left(\frac{k}{k+1}\right) \omega_o C_1$$

D_c: Capacitor dissipation factor (%)

$$R_1 = \frac{1}{y_1} \qquad R_2 = \frac{1}{y_2} \qquad R_3 = \frac{1}{y_3'} \qquad R_4 = (K - 1)\, R_5$$

Figure 4. Transfer function and component values.



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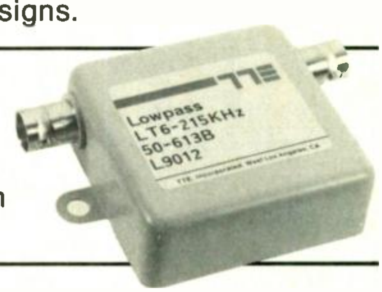

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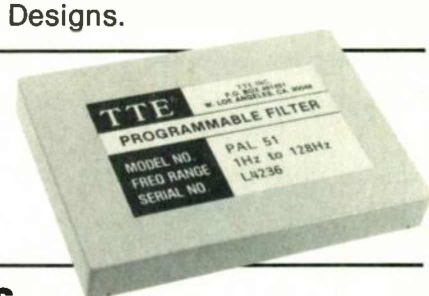

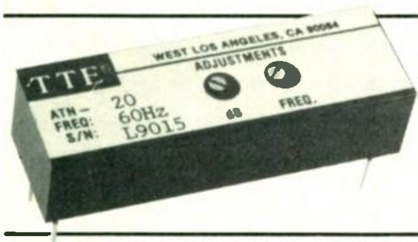

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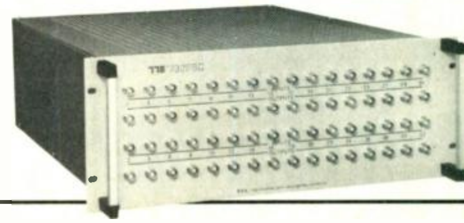

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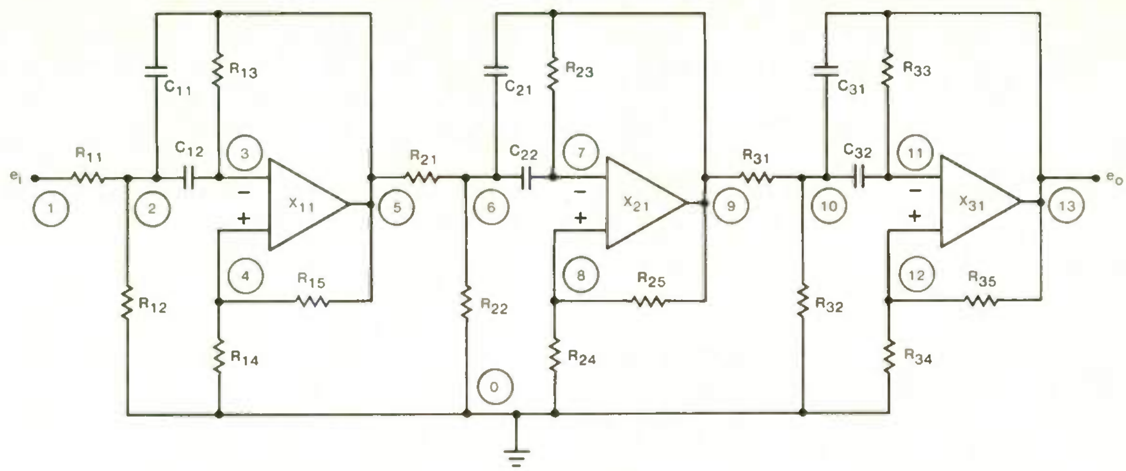

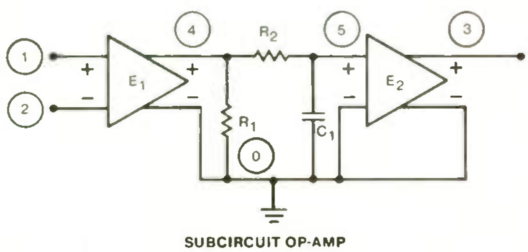

Figure 5. SPICE analysis schematic.

Component Values

Figure 4 shows the transfer function of a filter section and summarizes the calculations used to determine the resistor values. Since the transfer function includes the effect of finite op-amp bandwidth, it also appears in the resistor value calculations. Predistorted values of F_o and Q are often used to compensate for finite amplifier bandwidth (9); the component values are then calculated assuming the op-amp is perfect (i.e., has infinite gain and bandwidth). SPICE simulations indicate that this procedure is less accurate than including op-amp bandwidth in the component value calculations.

Sensitivity To Component Value Changes

Shifts in resonant frequency or Q of the filter sections cause the transfer function of the filter to be distorted. It is a factor of

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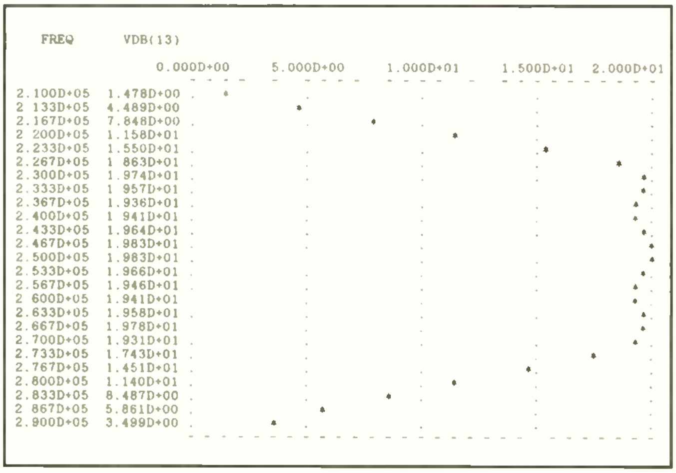

Figure 6. Filter design example.

Q more sensitive to frequency shifts than to Q shifts (5). Since the resonant frequency is mainly determined by the resistor and capacitor values, it is essential that these components have a temperature coefficient as close to zero as possible. The amount of positive feedback affects the variation in Q caused by passive component value changes and the variations in both resonant frequency and Q caused by changes in op-amp gain (10). The value of the feedback factor K calculated by the equations in Figure 4 was chosen to minimize these variations as much as possible.

BESSEL FILTER POLES

N	Re(P)	Im(P)
1	1	0
2	1.101601	.6360098
3	1.322676	0
	1.047409	.9992644
4	1.370068	.4102497
	.9952088	1.257106
5	1.502316	0
	1.380877	.7179096
	.9576765	1.471124
6	1.57149	.3208964
	1.381858	.9714719
	.9306566	1.661863
7	1.684368	0
	1.612039	.5892445
	1.378903	1.191567
	.9098678	1.836451
8	1.757408	.2728676
	1.636939	.8227956
	1.373841	1.388357
	.8928697	1.998326

Table 1. Bessel filter poles.

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INFO/CARD 46

```
100 'Program BPAFST
110 'DEFINITIONS OF VARIABLES:
120 'A0: Gain of filter section at filter center frequency
130 'As: Gain of filter section at its center frequency
140 'BW: Filter bandwidth in Hz
150 'D1: Capacitor dissipation factor in %
160 'Am: Passband ripple in dB (Chebyshev filters);
170 '     for Butterworth filters Am=0
180 '     For other types of filters enter -1 for Am
190 'Ff: Filter center frequency in Hz
200 'Fc: Corrected center frequency in Hz
210 'Fs: Center frequency of filter section in Hz
220 'GBW: Op-amp gain bandwidth in MHz
230 'N: Number of filter sections
240 'Q: Q of filter section
250 'C1, C2, R1, R2, Etc.: Component values in pF or ohms
260 CLEAR
270 DEFINT I-J, N
280 DIM A(8,1), E$(4), F(8), G(8,8), Q(8), R(5), R$(5), U(4), V(4)
290 PI=3.141593: DP=10/LOG(10): D1=.1: N7=20
300 A1$="BANDPASS ACTIVE FILTER"
310 B1$="Prototype poles"
320 E$(0)="Butterworth": E$(1)="Chebyshev": E$(2)=E$(1)
330 E$(3)="Bessel": E$(4)=""
340 R$(1)="R1=": R$(2)="R2=": R$(3)="R3=": R$(4)="R4=": R$(5)="R5="
350 CLS: PRINT A1$; TAB(2*N7) DATE$
360 INPUT "N, Am(dB)"; N1, A1
370 IF N1<1 THEN 1980
380 N2=INT((N1+1)/2)
390 I3=0: I8=0: IF A1>=0 THEN 530
400 PRINT "1. "; E$(3); "  2. "; "Other";: INPUT I3
410 I3=I3+2: IF I3>3 THEN 490
420 READ J1: J3=INT((J1+1)/2)
430 FOR J2=1 TO J3
440 READ S1, W1
450 IF J1=N1 THEN U(J2)=S1: V(J2)=W1
460 NEXT J2
470 IF J1<N1 THEN 420
480 RESTORE: GOTO 690
490 FOR J1=1 TO N2
500 PRINT J1; "Sp, Wp";: INPUT U(J1), V(J1)
510 NEXT J1
520 GOTO 690
530 A3=1: A4=1: W1=1
540 IF A1=0 THEN 640
550 INPUT "Specify:  1. Ripple BW,  2. 3 dB BW"; I3
560 A2=2: A5=EXP(A1/DP): E2=1/SQR(A5-1)
570 T1=EXP(LOG(E2+SQR(E2*E2+1))/N1)
580 A3=(T1-1/T1)/2: A4=(T1+1/T1)/2
590 IF I3<2 THEN 640
600 IF N1=2*INT(N1/2) THEN A2=A2*A5
610 T2=E2*SQR(A2-1)
620 T1=EXP(LOG(T2+SQR(T2*T2-1))/N1)
630 W1=(T1+1/T1)/2
640 FOR J1=1 TO N2
650 J2=2*(N2+1-J1)-1: T1=J2*PI/(2*N1)
660 U(J1)=A3*SIN(T1)/W1: V(J1)=A4*COS(T1)/W1
670 NEXT J1
680 IF V(1)<.00001 THEN V(1)=0
690 PRINT CHR$(13); B1$; ":"
700 PRINT "Section", "   Re(P)", "   Im(P)"
710 FOR J1=1 TO N2: PRINT J1, U(J1), V(J1): NEXT J1
720 PRINT: INPUT "Print prototype poles (Y or N)"; I$
730 IF LEFT$(I$,1)="N" THEN 770
740 LPRINT CHR$(13): GOSUB 1770
750 LPRINT B1$; ":"; CHR$(13); "Section", "   Re(P)", "   Im(P)"
760 FOR J1=1 TO N2: LPRINT J1, U(J1), V(J1): NEXT J1
770 PRINT: INPUT "Ff, BW (Hz)"; F(0), B0
780 ON SGN(F(0))+2 GOTO 1980, 350, 790
790 B2=B0/2: A2=0: I8=0
800 INPUT "Correct Ff to center it in 3 dB BW (Y or N)"; I$
810 IF LEFT$(I$,1)="N" THEN F1=F(0): ELSE F1=SQR(F(0)*F(0)-B2*B2)
820 Q1=F1/B0
830 INPUT "GBW (MHz)"; G1
840 FOR J1=1 TO N1 STEP 2
850 J2=1+INT((N1-J1)/2)
860 S1=U(J2)/Q1: W1=V(J2)/Q1
870 T1=(4-S1*S1+W1*W1)/2: T2=S1*W1
880 U2=T1+SQR(T1*T1+T2*T2): V1=T2/SQR(U2)
890 Q(J1)=SQR(1+U2/(S1*S1))/2
900 F(J1)=F1*Q(J1)*(S1-V1)
910 IF J1=N1 THEN 940
920 Q(J1+1)=Q(J1)
930 F(J1+1)=F1*Q(J1+1)*(S1+V1)
940 NEXT J1
950 INPUT "Specify:  1. Total filter gain  2. Gain of each stage"; I4
960 IF I4<1 THEN 350
970 IF I4=1 THEN J4=N1: INPUT "Total gain"; A2: ELSE J4=0
980 FOR J1=1 TO N1
990 FOR J2=0 TO J4
1000 T1=Q(J1)*(F(J2)*F(J2)-F(J1)*F(J1))/(F(J2)*F(J1))
1010 G(J1,J2)=SQR(1+T1*T1)
1020 NEXT J2, J1
1030 IF I4>1 THEN 1210
1040 IF N1>1 THEN A(1,1)=A2: ELSE A(1,0)=A2: GOTO 1200
1050 A(1,0)=A(1,1)/G(1,0): A6=A(1,0)
1060 IF N1<3 THEN 1190
1070 FOR J1=2 TO N1-1
1080 A5=1: A(J1,1)=A5: G5=1000000!
1090 FOR J2=1 TO J1
1100 G6=1
1110 FOR J3=1 TO J1
1120 G6=G6*G(J3,J2)
1130 NEXT J3
1140 IF G5>G6 THEN G5=G6
1150 A5=A5*A(J2,1)
1160 NEXT J2
1170 A(J1,1)=A2*G5/A5: A(J1,0)=A(J1,1)/G(J1,0): A6=A6*A(J1,0)
1180 NEXT J1
1190 A(N1,0)=A2/A6
1200 A(N1,1)=A(N1,0)*G(N1,0)
1210 CLS: PRINT A1$; TAB(2*N7) DATE$; CHR$(13); "Fc="; F1;
1220 IF I4=1 THEN PRINT TAB(2*N7) "Total gain="; A2: ELSE PRINT
1230 PRINT "      ", "   Fs", "    Q", "  As/A0"
1240 FOR J1=1 TO N1: PRINT J1, F(J1), Q(J1), G(J1,0): NEXT J1
1250 PRINT: IF I4>1 THEN 1270
1260 INPUT "Section, C1(pF)"; I1, C1: GOTO 1280
1270 INPUT "Section, C1(pF), Gain"; I1, C1, A3
1280 IF I1<0 THEN 1980
1290 IF I1>0 THEN 1310
1300 IF A1<0 THEN 770: ELSE 350
1310 W0=2E-12*PI*F(I1): D2=1/Q(I1)
1320 IF I4=1 THEN A3=A(I1,0). A4=A(I1,1): ELSE A4=A3*G(I1,0)
1330 C2=C1: K2=C2/C1
1340 T0=1000000!/(2*PI*G1)
1350 A6=(K2+1)/(4*K2*T0*W0)
1360 IF A6>40 THEN A6=40
1370 K1=1-D2/SQR(K2*A6)+(K2+1)/(K2*A6)
1380 IF K1<1 THEN K1=1
1390 IF K1>1 THEN R(5)=10000!: R(4)=(K1-1)*R(5): ELSE R(4)=0: R(5)=0
1400 A7=(K2+1)*W0*W0*C1*C1
1410 B7=((K2+1)*K1*T0*W0-D2)*W0*C1
1420 C7=1-K1-K1*D2*T0*W0
1430 Y7=2*A7/(SQR(B7*B7-4*A7*C7)-B7)
1440 Y5=K2*W0*W0*C1*(C1/Y7+K1*T0)
1450 Y1=A4*D2*W0*(C1/K1+T0*Y7)
1460 Y5=Y5-.01*D1*W0*C1*K2/(K2+1)
1470 R(1)=1/Y1: R(2)=1/(Y7-Y1): R(3)=1/Y5
1480 PRINT "As="; A4; TAB(N7) "A0="; A3
1490 PRINT "C1="; C1; "pF"; TAB(N7) "C2="; C2; "pF"; TAB(2*N7) " a="; A6
1500 J2=0: FOR J1=1 TO 5
1510 IF R(J1)=0 THEN 1540
1520 PRINT TAB(J2) R$(J1); R(J1);
1530 J2=J2+N7: IF J2>2*N7 THEN J2=0: PRINT
1540 NEXT J1
1550 PRINT TAB(J2) " K="; K1;
1560 IF J2>0 THEN PRINT
1570 PRINT: INPUT "Print results (Y or N)"; I$
1580 IF LEFT$(I$,1)="N" THEN 1210
1590 IF I8=1 THEN 1640
1600 I8=1: GOSUB 1770
1610 LPRINT "Op-amp gain BW="; G1; "MHz", "Capacitor DF="; D1; "%"
1620 IF I4=1 THEN LPRINT "Total gain=", A2
1630 LPRINT "Ff="; F(0); "Hz"; TAB(N7) "Fc="; F1; "Hz", "BW="; B0; "Hz"
1640 LPRINT CHR$(13); "Section"; I1
1650 LPRINT "Fs="; F(I1); "Hz"; TAB(N7+1) "Q="; Q(I1)
1660 LPRINT "As="; A4; TAB(N7) "A0=";A3;
1670 LPRINT TAB(2*N7) "As/A0="; G(I1,0)
1680 LPRINT "C1="; C1; "pF"; TAB(N7) "C2="; C2; "pF"; TAB(2*N7) " a="; A6
1690 J2=0: FOR J1=1 TO 5
1700 IF R(J1)=0 THEN 1730
1710 LPRINT TAB(J2) R$(J1); R(J1);
1720 J2=J2+N7: IF J2>2*N7 THEN J2=0: LPRINT
1730 NEXT J1
1740 LPRINT TAB(J2) " K="; K1;
1750 IF J2>0 THEN LPRINT
1760 GOTO 1210
1770 IF I7=1 THEN 1810
1780 I7=1: LPRINT CHR$(27)+CHR$(77)+CHR$(10);:   'Left margin
1790 LPRINT CHR$(27)+CHR$(78)+CHR$(8);:    'Skip over perforation
1800 LPRINT CHR$(13); CHR$(14); A1$; CHR$(13); CHR$(14); TAB(7) DATE$
1810 LPRINT CHR$(13): IF I3<4 THEN LPRINT E$(I3),
1820 LPRINT "N="; N1,
1830 IF I3=0 OR I3>2 THEN 1870
1840 LPRINT "Am="; A1; "dB",
1850 IF I3=1 THEN LPRINT "Ripple BW",
1860 IF I3=2 THEN LPRINT "3 dB BW",
1870 LPRINT: RETURN
1880 DATA 1, 1, 0
1890 DATA 2, 1.101601, .6360098
1900 DATA 3, 1.322676, 0, 1.047409, .9992644
1910 DATA 4, 1.370068, .4102497, .9952088, 1.257106
1920 DATA 5, 1.502316, 0, 1.380877, .7179096, .9576765, 1.471124
1930 DATA 6, 1.571490, .3208964, 1.381858, .9714719, .9306565, 1.661863
1940 DATA 7, 1.684368, 0, 1.612039, .5892445
1950 DATA 1.378903, 1.191567, .9098678, 1.836451
1960 DATA 8, 1.757408, .2728676, 1.636939, .8227956
1970 DATA 1.373841, 1.388357, .8928697, 1.998326
1980 IF I7=1 THEN LPRINT CHR$(12)
1990 END
```

Figure 2. The BASIC listings perform the calculations for Figures 1 to 4.

```
Chebyshev       N= 3            Am= .5 dB      3 dB BW        R1= 12480.38         R2= 1288.381         R3= 38147.64
Prototype poles:                                              R4= 513.8839         R5= 10000             K= 1.051388
Section           Re(P)           Im(P)
 1              .5365863        0                             Section  2
 2              .2682932        .8753236                      Fs= 271599.2 Hz       Q= 18.61458
                                                              As= 6.653608         A0= 2.04951          As/A0= 3.246439
                                                              C1= 100 pF           C2= 100 pF             a= 27.61422
Chebyshev       N= 3            Am= .5 dB      3 dB BW        R1= 15746.81         R2= 1141.762         R3= 29167.89
Op-amp gain BW= 15 MHz          Capacitor DF= .1 %            R4= 622.0341         R5= 10000             K= 1.062203
Total gain= 10
Ff= 250000 Hz        Fc= 248746.9 Hz           BW= 50000 Hz   Section  3
                                                              Fs= 248746.9 Hz       Q= 9.271459
Section 1                                                     As= 1.766895         A0= 1.759274         As/A0= 1.004332
Fs= 227817.3 Hz      Q= 18.61458                              C1= 220 pF           C2= 220 pF             a= 30.15114
As= 10               A0= 2.773427          As/A0= 3.605648    R1= 14533.31         R2= 527.9815         R3= 15105.25
C1= 100 pF           C2= 100 pF              a= 32.92112      R4= 466.8987         R5= 10000             K= 1.04669
```

Table 3. The output of the program in Table 2.

Tuning Procedure

To tune each filter section, disconnect its input from the preceding stage and apply a signal at the calculated center frequency for that section from a low-impedance source. Adjust R2 for a 180 degree phase shift between the input and the output of that stage. Then, when all sections have been tuned, connect them together.

A Basic Program

Table 2 is a listing of a computer program which performs all of the calculations shown in Figures 1 thru 4. The language is a common version of Microsoft Basic which will run on IBM PC's and compatibles, TRS-80's, and most other small computers. The printer settings in lines 1780 and 1790 are for a Gemini-10X and may have to be changed for other printers.

An Example

A three-stage Chebyshev filter with 0.5 dB passband ripple, 250 kHz center frequency, 50 kHz 3 dB bandwidth, and a gain of 10 is required. LM318 op-amps with 15 MHz gain-bandwidth will be used. Capacitor values of 100 pF are selected for the first two stages and 220 pF for the last. The program shown in Table 2 is used to calculate the required resistor values. Line 1460 of the program, which corrects the value of R3 to compensate for the dissipation factor (.1 percent) of NPO ceramic capacitors, was deleted for these calculations. Table 3 is the printed output. A SPICE program was used to analyze the resulting filter. Figure 5 is the schematic described by the input listing in Table 4. The op-amp subcircuit models the LM318 open loop gain (106 dB DC gain and 15 MHz gain-bandwidth). An output plot is shown in Figure 6.

```
.OPTIONS LIST NODE
.WIDTH OUT=80
.SUBCKT OPAMP 1 2 3
E1 4 0 1 2 2E5
R1 4 0 1E5
R2 4 5 21221
C1 5 0 1E-7
E2 3 0 5 0 1
.ENDS OPAMP
VIN 1 0 AC
R11 1 2 12480
R12 2 0 1288.4
C11 2 5 100E-12
C12 2 3 100E-12
R13 3 5 38148
R14 4 0 513.88
R15 4 5 1E4
X11 4 3 5 OPAMP
R21 5 6 15747
R22 6 0 1141.8
C21 6 9 100E-12
C22 6 7 100E-12
R23 7 9 29168
R24 8 0 622.03
R25 8 9 1E4
X21 8 7 9 OPAMP
R31 9 10 14533
R32 10 0 527.98
C31 10 13 220E-12
C32 10 11 220E-12
R33 11 13 15105
R34 12 0 466.90
R35 12 13 1E4
X31 12 11 13 OPAMP
.AC LIN 25 210E3 290E3
.PLOT AC VDB(13)
.END
```

Table 4. Input listings for Figure 5.

References

1. S. Butterworth, "On the Theory of Filter Amplifiers", *Wireless Engineer,* Vol. 7, pp. 536-541, 1930.
2. G.E. Valley, H. Wallman, *Vacuum Tube Amplifiers,* (MIT Rad Lab Series, Vol. 18) Chapter 4, McGraw-Hill, 1948.
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9. A.S. Sedra, M.A. Ghorab & K. Martin, "Optimum Configurations for Single-Amplifier Biquadratic Filters", *IEEE Transactions on Circuits and Systems,* Vol. CAS-27, pp. 1155-1163, 12/80.
10. A.S. Sedra & L. Brown, "A Refined Classification of Single Amplifier Filters", *International Journal of Circuit Theory and Applications,* Vol. 7, pp. 127-137, 1/79. Reprinted in "Modern Active Filter Design", IEEE Press, 1981.
11. P.E. Fleischer, "Sensitivity Minimization in a Single Amplifier Biquad Circuit", *IEEE Trans. Circuits System,* Vol. CAS-23, pp. 45-55, 1/76. Reprinted in "Modern Active Filter Design."
12. G.S. Moschytz & P. Horn, "Reducing Nonideal Op-Amp Effects in Active Filters by Minimizing the Gain-Sensivity Product (GSP)", *IEEE Trans. Circuits System,* Vol. CAS-24, pp. 437-445, 8/77. Reprinted in "Modern Active Filter Design".

About the Author

Jack Porter is senior technical specialist at Cubic Defense Division, 9333 Balboa Ave., P.O. Box 85587, San Diego, CA 92138-5587; Tel: (619) 277-6780.

Product Detector Performance

Basic Information About An Often-Misunderstood Circuit.

By Fred Brown
Palomar Mountain, Calif.

Product detectors are widely used in communications but are often poorly understood. There have been only a few published comparative analyses of different configurations. This paper breaks ground for future quantitative comparisons of different PD circuits.

Product detectors (PDs) are essentially frequency converters that down convert signals in an IF passband to audio or video frequencies. Figure 1 is an example of the type of spectrum the PD must handle. Signal components in the IF passband are denoted by e_{S1} and e_{S2} and the local oscillator by E_0. The product detector must mix e_{S1} with E_0 and e_{S2} with E_0 — a requirement that is easily met by a nonlinear device. However, an ordinary detector will also mix e_{S1} with e_{S2} and pass this unwanted frequency on along with the desired frequencies. An easy way to measure product detector performance is to substitute the carrier and 2 sidebands of an amplitude modulated signal for e_{S1} and e_{S2} in Figure 1. The AM rejection ratio is then

AM rejection ratio =

$$20 \log \frac{\text{PD output due to carrier}}{\text{PD output due to AM detection}} \text{ dB} \qquad (1)$$

The carrier amplitude is the same in both numerator and denominator. An audio filter can be used to separate the detected AM from the heterodyned carrier and sidebands. This is easy if the signal frequency is separated from the LO frequency by several times the modulation frequency. If the PD truly responds only to the product of the signal voltage, e_s, and LO voltage, E_0, the output voltage becomes that of equation 1 with k being a constant.

$$E_{out} = kE_0 \text{Sin}\omega_0 t \cdot e_s \text{Sin}\omega_s t \qquad (1)$$

Using SinA · SinB = ½Cos(A − B) − ½Cos(A + B), (1) becomes

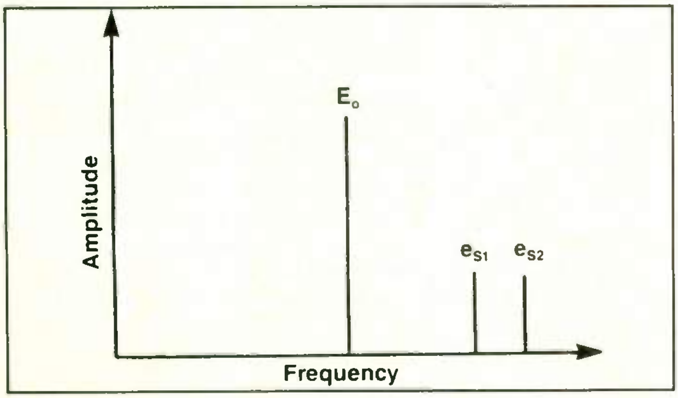

Figure 1. Spectrum encountered by product detector.

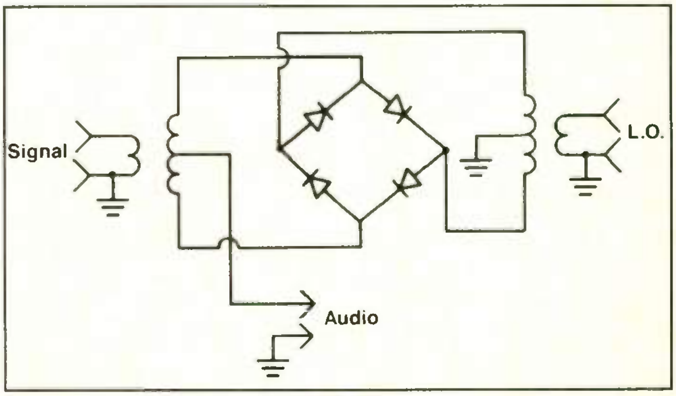

Figure 2. Classic ring modulator or double balanced mixer.

$$E_{out} = ½kE_oe_sCos(\omega_o - \omega_s)t - ½kE_oe_sCos(\omega_o + \omega_s)t \quad (2)$$

Equation (2) shows that a true product detector will produce only sum and difference frequencies in its output. Note that the sum frequency is easily removed with a low-pass filter. If the signal is amplitude modulated with modulation index, M,

$e_s = e_a(1 + MSin\omega_m t)$, then (1) becomes

$$E_{out} = kE_oe_aSin\omega_o t\,(1 + MSin\omega_m t)Sin\omega_s t \quad (3)$$

or

$$E_{out} = ½kE_oe_a(1 + MSin\omega_m t)\,Cos(\omega_o - \omega_s)\,t - ½kE_oe_a(1 + MSin\omega_m t)\,Cos(\omega_o + \omega_s)\,t \quad (4)$$

If we ignore the second term of (4), the sum frequency, and expand the first term as before, we have

$$E_{out} = ½kE_oe_a[Cos(\omega_o - \omega_s)\,t + ½MSin(\omega_m + \omega_o - \omega_s)\,t + ½MSin(\omega_m - \omega_o + \omega_s)\,t] \quad (5)$$

Notice in (5) that the three frequencies present in the output are the frequencies produced by the LO mixing with the carrier and the two sidebands. No frequency corresponding to ω_m is present; in other words, AM detection does not occur in a perfect product detector. For comparison the AM rejection ratio for a conventional square-law detector is computed. The output of such a detector is proportional to the square of the input voltage as shown in equation 6, with k being a constant.

$$E_{out} = k(e_{in})^2 \quad (6)$$

If a local oscillator and AM signals are applied to such a detector the output will be that of equation 7.

$$E_{out} = k[E_oSin\omega_o t + e_s(1 + MSin\omega_m t)Sin\omega_s t]^2 \quad (7)$$

$$E_{out} = kE_o^2Sin^2\omega_o t + 2kE_oe_s\,Sin\omega_o t\,(1 + MSin\omega_m t)\,Sin\omega_s t + ke_s^2\,(1 + MSin\omega_m t)^2\,Sin^2\omega_s t \quad (8)$$

The second term in (8) can be expanded as shown in equation 9.

$$kE_oe_s\,(1 + MSin\omega_m t)\,[Cos(\omega_o - \omega_s)t - Cos(\omega_o + \omega_s)\,t] \quad (9)$$

The amplitude of the difference frequency is now

$$kE_oe_s(1 + MSin\omega_m t) \quad (10)$$

The last term in (8),

$$ke_s^2(1 + MSin\omega_m t)^2Sin^2\omega_s t$$

now becomes that of equation 11 by using the following identity:

$$Sin^2A = ½ - ½Cos2A$$

$$ke_s^2(1 + MSin\omega_m t)^2(½ - ½\,Cos2\omega_s t) \quad (11)$$

This gives rise to an audio term as shown below.

$$½ke_s^2(1+MSin\omega_m t)^2$$

$$=½ke_s^2(1+2MSin\omega_m t+½m^2-½m^2Cos2\omega_m t) \quad (12)$$

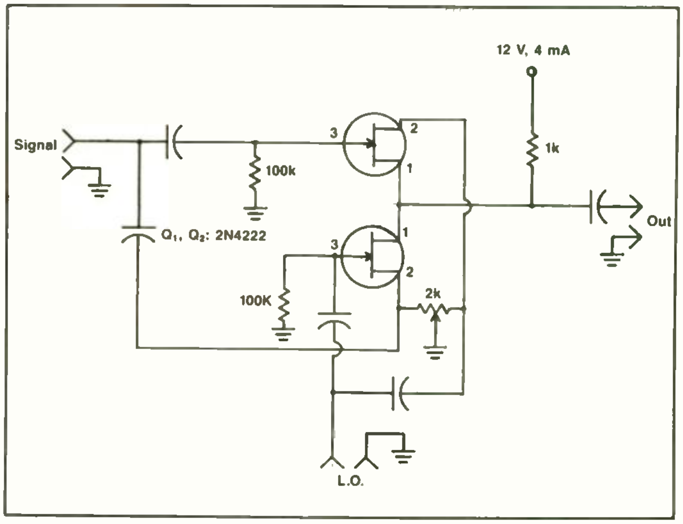

Figure 3. A doubly balanced FET product detector with unbalanced inputs.

The term $¼ke_s^2m^2Cos2\omega_m t$ in (12) represents second harmonic distortion caused by square-law detection, and will be ignored. The detected carrier, from (10), will have an amplitude of kE_oe_s and detected AM, from (12), of ke_s^2M. If the modulation index, M, is taken as 1 (100 percent modulation) the AM rejection ratio for a square-law will be seen to be simply E_o/e_s, that is the ratio of the local oscillator to signal amplitudes. So even a square-law detector will perform well as a product detector if the signal amplitude is kept small compared to the local oscillator amplitude. The mathematical analysis of a linear envelope detector is not as simple as the square-law detector but experimental results have shown that the linear detector also has an AM rejection ratio of approximately E_o/e_s. The classic PD is the ring modulator shown in Figure 2. With proper balance, such a detector can theoretically provide infinite AM rejection. Response will be linear for signals up to a level of about 10 dB below the LO input level. The ring modulator has the disadvantage of requiring large amounts of LO injection, balanced signal and LO inputs, and a conversion loss of about 6 dB.

A more convenient PD configuration is shown in Figure 3. This cross-coupled FET PD has a doubly balanced output and does not require a balanced LO or signal input. Only 1 volt or about 3 mW of LO injection is needed. Signal and LO impedances are 350 ohms, and the output impedance is essentially that of the drain load resistor, in this case 1000 ohms.

The 2K balance adjustment pot, R_1, can be adjusted for minimum LO voltage at the output. Input coupling capacitors should be chosen to have less than 100 ohms reactance at the signal and LO frequencies. The value of the output capacitor depends on the lowest desired output frequency and the load impedance. Conversion voltage gain will be about 2 dB. The AM rejection ratio for a 100 mV, 90 percent modulated signal is 27 dB.

About the Author

Fred Brown is a self-employed RF Consulting engineer. He has worked with RF for over 30 years and holds an MSEE from the University of Illinois. Mr. Brown can be contacted at Box 73, Palomar Mtn., CA 92060; Tel: (619) 742-1328.

A Review of Op Amp Specifications

High-Speed Op Amps Are Finding Applications in RF Circuits

By Mark Gomez
Technical Editor

The operational amplifier (op amp) represents one of the most versatile building blocks in analog design. Op amps can be used to implement circuits such as comparators, sample-and-holds, active filters, buffer, inverters and amplifiers to name just a few. This article defines some of the commonly used terms associated with these components.

An op amp is a very stable amplifier used to implement a wide variety of linear and non-linear operations by changing external components such as resistors, capacitors, inductors and diodes. The name is derived from its usage in performing a wide number of mathematical operations on input signals such as addition, subtraction, multiplication and division by a constant, differentiation, and integration as well as the generation of special non-linear functions. The op amp is a direct coupled gain amplifier that uses feedback to control its performance characteristics. The equivalent circuit for an op amp is shown in Figure 1. The output of the device is controlled by the two inputs V_1 and V_2 which are connected by the input resistance R_i. A voltage controlled source A_dV_d in series with output resistance R_o constitutes the output circuit of an op amp. A_d is usually very large (typically 10,000) in comparison with the gain of the circuit in which it is used. R_i is typically 10 k ohms and $V_d = V/A_d$. Note that for the ideal op amp, $V_d = V_1 - V_2 = 0$.

Op Amp Parameters

Unity Gain Bandwidth — The unity gain bandwidth of an op amp is the frequency where op amp voltage gain is unity. It can be computed by measuring the 10 percent to 90 percent rise time (t_r) and using equation 1.

$$f_u = 0.35/t_r \qquad (1)$$

This method is only valid when the op amp is connected as a unity gain non-inverting amplifier.

Closed Loop Bandwidth — When the circuit must supply large voltage swings, the closed loop bandwidth is of concern and this value is usually much less than f_u. Depending on the peak output voltage, the closed loop bandwidth may be only 1/10 or 1/100 f_u. This high level bandwidth, f_f, is the maximum frequency of full output power response. Most general purpose op amps will give adequate performance for the DC and audio frequency range where gain bandwidth product is linear. At high frequencies, gain bandwidth product merely represents a number since it is not linear. Hence it is best represented in graphical form for high speed and wideband op amps.

Small Signal Bandwidth is the frequency at which the op amps closed loop gain has dropped 3 dB with respect to the low frequency gain.

Power Bandwidth is the frequency at which slew rate limitations cause the large signal gain to drop by 3 dB.

Slew Rate — Slew rate is defined as the maximum rate at which op amp output voltage can change. Maximum slew rate is called slew rate limiting.

$$S = \frac{\Delta V}{\Delta t} \qquad (2)$$

Slew rate is higher for high gain circuits and can be reduced by reducing the input signal and increasing the gain of the circuit. When an AC signal is used, the maximum slew rate occurs when f_f is maximum resulting in maximum output voltage swing. The user should examine the expected output waveform and select an amplifier where the slew rate exceeds the maximum rate of change of output signal.

Input Offset Voltage — Input offset voltage (V_{io}) is the voltage required across input terminals to drive op amp output voltage to zero.

$$V_{io} = \pm V_o/(1 + \text{voltage gain}) \qquad (3)$$

where V_o is the output offset voltage. This equation holds true for both the inverting and non-inverting configurations. V_{io} is typically in the range of a fraction of a millivolt to several millivolts. In high gain circuits, this may result in an output offset voltage of several volts. In AC cir-

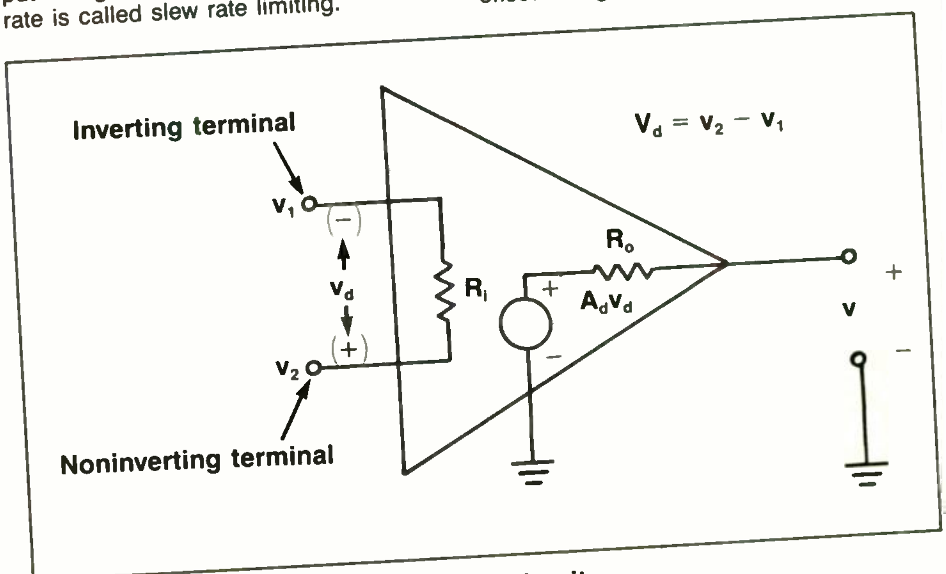

Figure 1. The basic op-amp equivalent circuit.

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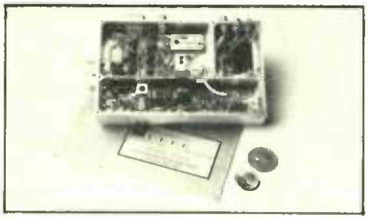

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- Remotely piloted vehicles
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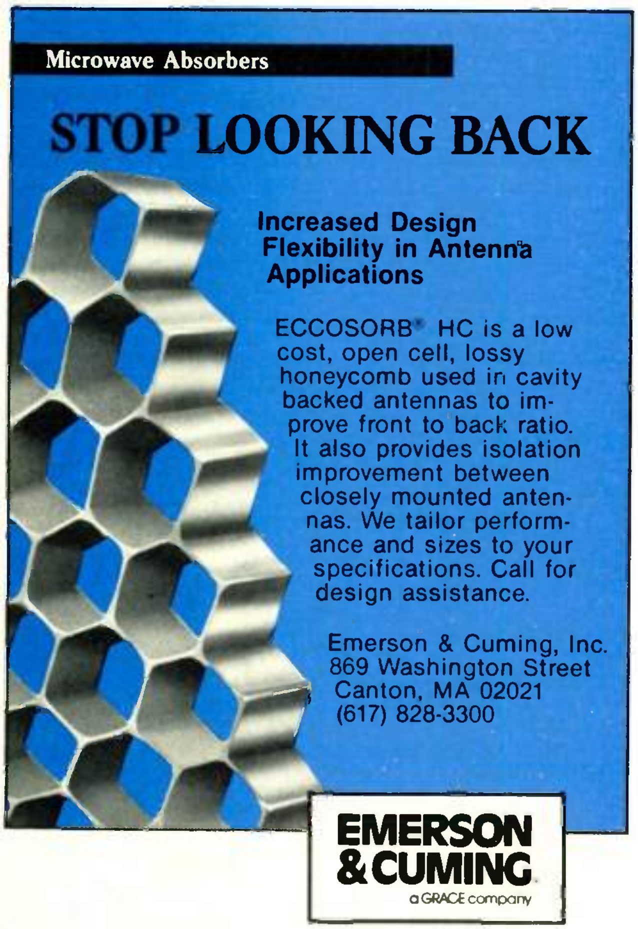

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cuits this is often neglected unless the output offset voltage and peak AC voltage add up to a voltage approaching either power supply voltage.

Common-mode rejection ratio — The common-mode rejection ratio (CMRR) is a measure of how much the common-mode signal is rejected relative to the desired differential mode signal. CMRR usually deteriorates at high frequencies. This is generally not a problem at low frequencies because the op amp runs out of bandwidth before common mode rejection drops significantly.

$$CMRR = A_v/A_{cm} \quad (4)$$

A_v is the op amp differential gain as a function of frequency and A_{cm} is the common mode gain as a function of frequency. The CMRR of an op amp can be optimized by choosing an op amp with a large minimum CMRR at DC and by making sure that the circuit in which the op amp is being used is at least 20 times larger than the op amp CMRR. Also, the op amp should have the largest possible CMRR values over the same frequency range as the circuit.

Power Supply Rejection Ratio — Power supply rejection ratio (PSRR) is the ratio of change in input offset voltage over the change in power supply voltage. If small signals are used this may pose as a problem. The problem can be reduced by choosing an op amp with a small PSRR, filtering the power supply, increasing the input current and decreasing the circuit gain.

This article defined some of the commonly used parameters associated with operational amplifiers. These versatile devices are finding their way into RF applications as the operating speeds go into the RF range. rf

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Microprocessor Interference to VHF Radios

FCC Regulations Do Not Always Protect Communications Equipment

By Daryl Gerke, PE
Kimmel Gerke & Associates, Ltd.

Interference to communications systems from microprocessors is not new. In fact, this interference is the basis for the current FCC regulations that limit emissions from digital equipment. Unfortunately, these regulations were designed to protect television receivers, and are often inadequate to protect more sensitive land mobile communications receivers. This paper examines the current FCC regulations and compares them to typical threshold sensitivity levels for communications equipment. It then discusses a case history of a microprocessor control system interfering with a land mobile communications receiver. In the final section several recommendations are made.

When analyzing and trying to understand an electromagnetic interference (EMI) situation, it is helpful to first divide the problem into categories. At the highest level, the three categories are the source, the victim, and the coupling path. Secondary categories typically address the coupling path — radiated, conducted, cross-talk, and common impedances to name a few.

In this paper, the following assumptions are made: the sources are microprocessor based digital computers, the victims are VHF/UHF communications receivers, and the primary coupling path is direct electromagnetic radiation which is picked up by the receiver antenna.

Microprocessors — The Source

In characterizing microprocessor based systems as a source of emissions, several alternate approaches are available. Predictions, measurements, or "system approach" together with existing equipment limits can be used as baseline.

In this paper, the systems approach is used together with Federal Communications Commission limits for digital computers, as specified in Part 15J of the FCC rules and regulations. This is a reasonable starting point since personal computer systems must meet those requirements, and also because they are being incorporated as controllers in larger systems.

In the late 1970s, it was apparent that the proliferation of home and business computers was creating a serious interference problem to licensed communications systems. Thus, the FCC established minimal emission requirements for digital computer equipment. Since the goal was to establish realistic requirements that were representative of typical problems, without being overly restrictive, several assumptions were made.

First, it was noted that below 30 MHz, the primary coupling path was conduction through the power lines, while above 30 MHz the primary coupling path was electromagnetic field radiation. This resulted in conducted emission limits from 450 kHz to 30 MHz, and radiated emission limits from 30 MHz to 1 GHz.

Second, most of the complaints were regarding interference to television receivers. This resulted in the radiated emission limits set to protect television receivers. The values were determined by assuming typical signal levels, typical signal/noise ratios, and typical antenna factors. The goal was to limit undesired noise to levels below those that would cause interference in normal metropolitan situations.

	Frequency MHz	Class A (30 meters)	Class B (3 meters)
Radiated	30-88	30 uv/m	100 uv/m
	88-216	50 uv/m	150 uv/m
	216-1000	70 uv/m	200 uv/m
Cond	0.45-1.6	1000 uv	250 uv
	1.6-30	3000 uv	250 uv

Table 1. FCC Limits.

Third, the computers were being installed in two environments, in businesses and in homes. In the home, it was assumed that a computer could be installed close to a television receiver, which resulted in a distance assumption of 3 meters. In business, it was assumed that the nearest television receiver would not be as close, so a distance of 30 meters was assumed.

The current emission requirements for digital computers, as specified in FCC Part 15J, are shown in Table 1. Although the Class A limits appear more stringent, when scaled to the same distance (assuming electric field intensity varies as 1/distance), it can be seen that Class A allows higher emanations. Thus, we can use the Class A limits as a "default" for computers.

The Victim — The VHF Receiver

The frequency and sensitivity characteristics of land mobile receivers are well defined. Land mobile receivers operate in three discrete frequency bands throughout the VHF/UHF range: 30-50 MHz, 150-170 MHz, and 450-470 MHz. Typical input sensitivities are 0.25 uv to 1 uv at the antenna input terminals, and typical antennas are 1/4 or 1/2 wavelength.

Table 2 shows the electric field intensities that result in 0.25 uv at the antenna terminals, assuming both 1/4 wavelength and 1/2 wavelength antennas with 50 ohm inputs. These were derived using the formulas:

$E=eF/33$ for a 1/4 wavelength antenna

$E=eF/39$ for a 1/2 wavelength antenna

where E is the field intensity in uv/meter, e is the input voltage in uv, and F is the frequency in MHz. The assumption is that if the interference is at or below the input threshold, then interference to even the weakest signals cannot occur.

A Comparison of Source and Victim Levels

An insight for the potential for problems can be gained by comparing the source characteristics of a digital computer meeting the Class A FCC requirements in terms of field intensity generated, and the victim characteristics of a VHF receiver at the antenna in terms of field intensity tolerated.

Table 3 shows the limits for field intensities, as derived in Table 2, versus the anticipated field intensity emissions from a Class A computer at several distances, based on a 1/distance decrease in field intensity. It can be seen that the potential for interference between the computer and the receiver exists if the two are within one kilometer of each other. (This ignores attenuation effects due to buildings or other materials. Nevertheless, it does show the gravity of the situation).

An alternate way of viewing this is to observe that additional shielding will be needed at the closer distances. This is shown in Table 4. At one meter, up to 72 dB of additional shielding would be needed to provide the necessary margin between the source and the receiver.

A Case Study

In this case history, the author participated in the integration of a system using a microprocessor based controller located three feet from a VHF receiver. The controller was originally designed using FCC

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	30 MHz	50 MHz	150 MHz	460 MHz
1/4 Wavelength	0.23 uv/m	0.38 uv/m	1.1 uv/m	3.5 uv/m
1/2 Wavelength	0.19 uv/m	0.32 uv/m	0.96 uv/m	2.9 uv/m

Table 2. Electric field intensity levels for 0.25 uV at antenna.

Class A limits as a guideline. Needless to say, there were problems.

The microprocessor based system used a MHz clock, divided internally by four. As a result, harmonics appeared at 2 MHz intervals with enough amplitude to cause problems on the 150 MHz-170 MHz band, as observed on both a radio receiver and a spectrum analyzer.

The computer system was housed in a rack which provided little shielding at 150 MHz. The computer consisted of three units, a central processor, a memory unit, and a display/control unit. The three were interconnected with parallel high speed digital input/outputs, plus there were over a hundred low speed input/output lines used for control and sensing.

The communications receiver had a sensitivity of 0.25 uv, and was connected to a 1/4 wavelength whip antenna. The signal received from the computer was enough to "break squelch", and was considered a major nuisance by the end user. The system was often used in fringe areas, so the maximum sensitivity was needed. Furthermore, the communications transceiver was installed for safety reasons, so its use was required.

Moving the communications receiver and/or antenna was not an option. For safety reasons, it was required that the operator have a VHF radio readily accessible. However, it was guaranteed that the radio antenna would not be located any closer than three feet.

Only a very narrow band of frequencies needed to be guarded, approximately 2 MHz at 150-170 MHz, and 1 MHz at 450-470 MHz. Fortunately, there were no discernable problems in the 450-460 MHz band.

First, an analysis was undertaken to assess the severity of the situation. Using the techniques described in this paper, it was determined that a Class A design still needed over 60 dB of isolation. Since isolation by distance was not an option, this meant additional shielding was needed. This was also compared with MIL-STD 461 levels, which was quite helpful in demonstrating the gravity of the situation to *management*.

Second, this following fixes were needed:

- Solid copper shields on all high speed I/O lines
- Routing of I/O lines as close as possible to case
- Shielded bulkhead connectors on all I/O, with circumferential shield terminations
- Filter pins on all I/O and power lines
- Solid shielded cabinet on three subassemblies, with RF gaskets at seams
- Screen shielding over the display unit

In addition, the clock frequency was changed to move the clock harmonics off one critical frequency. This was not con-

Class A Computer Electric Field Intensity Levels vs. Receiver Sensitivity Levels

	30 MHz	50 MHz	150 MHz	460 MHz
1/4 Wave Antenna	0.2 uv/m	0.4 uv/m	1.1 uv/m	3.5 uv/m
Class A-1 meter	900 uv/m	900 uv/m	1500 uv/m	2100 uv/m
3 meters	300	300	500	700
10 meters	90	90	150	210
30 meters	30	30	50	70
100 meters	9	9	15	21
300 meters	3	3	5	7
1000 meters	0.9	0.9	1.5	2.1
3000 meters	0.3	0.3	0.5	0.7

Table 3. Intensity versus sensitivity levels.

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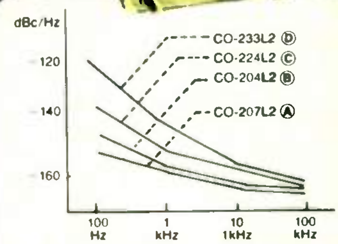

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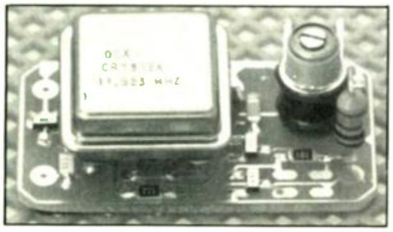

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Isolation Required Between Class A Computer and Receiver	30 MHz	50 MHz	150 MHz	460 MHz
1/4 Wave Antenna				
–uv/m	0.2	0.4	1.1	3.5
–dB uv/m	–13	–8.5	0.8	11
Class A Computer–1 meter				
–uv/m	900	900	1500	2100
–dB uv/m	59	59	63	66
–Isolation dB	72	68	62	55
Class A Computer–3 meters				
–uV/m	300	300	500	700
–dB uv/m	50	50	54	57
–Isolation dB	63	59	53	46

Table 4. Required isolation.

sidered a long term solution since harmonics still fell within the overall band to be used.

The results of these retrofits were a quasi-militarized system, at no small expense. It was definitely not a commercial personal computer type of design.

Recommendations and Conclusions

The engineer should always be aware that a computer meeting the FCC Part 15J requirements will not always work with a sensitive VHF radio receiver. The FCC limits were designed primarily to protect television receivers at prescribed distances. Also, analyze the situation and try to predict both the sensitivity level of the receiver and the emission level of the computer. If the computer meets the FCC limits, use that as a default, and adjust for distance.

Use space separation between the source (computer) and victim (receiver) where possible. Frequency separation (changing clock frequencies), if used, should be carried out carefully. This is successful only if you are guarding a single frequency; otherwise it is likely to fail, since digital systems are rich sources of many signals.

Another recommendation is that the engineer should be prepared to use much shielding and filtering if a microprocessor and a VHF receiver are co-located in a system. rf

About the Author

Daryl Gerke is a principal in Kimmel Gerke & Associates, Ltd., 1544 North Pascal, St. Paul, MN 55108. His career spans 20 years, and he has held engineering positions with Rockwell Collins, Sperry Defense Systems, Tektronix, and Intel Corp. He can be reached at (612) 330-3728.

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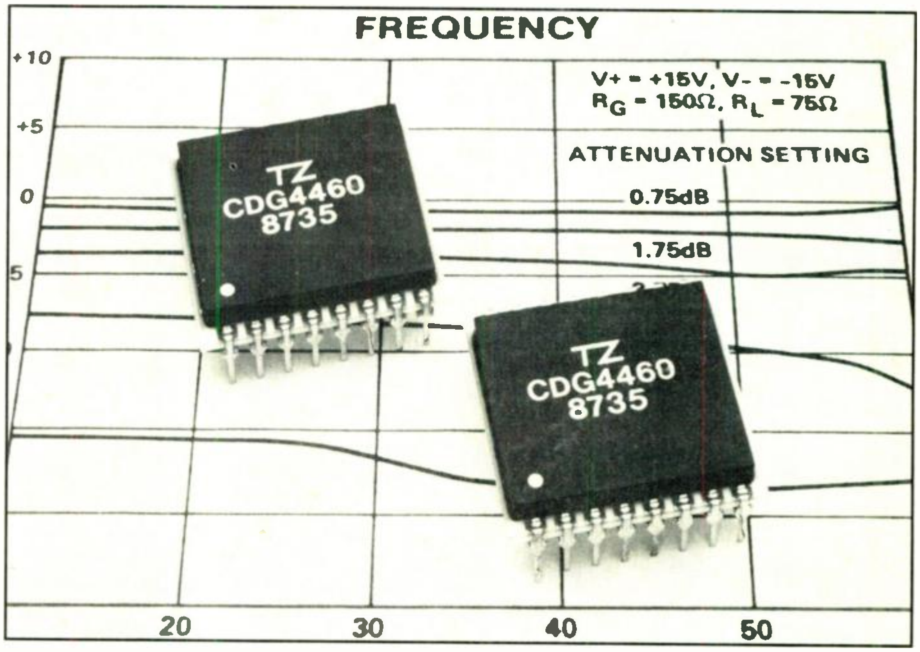
Topaz Introduces Digitally-Controlled Attenuators

The CDG4460J is a 6-bit, digitally-controlled attenuator for frequencies up to 40 MHz. It is a hybrid device comprised of three CMOS/D-MOS integrated circuits with 11 laser trimmed resistors mounted on a thick film ceramic substrate. The circuit is available in a hybrid package which requires one square inch of pcb space.

The device offers an attenuation range of 0 dB to 15.75 dB selectable in 0.25 dB steps. Designed for 75-ohm systems, the CDG4460J provides constant input and output resistance regardless of the attenuation setting. On-board latch allows the user to set pre-determined attenuation levels or to recall settings from memory.

These features allow the device to be used in applications requiring in-circuit attenuators for common 10.7 MHz to 30 MHz frequencies, modulator and demodulator attenuator strips, IF amplifiers and test instruments. Other possible applications include video attenuation, wideband amplifier gain control, variable burst generation, IF amplifier attenuation, and frequency synthesizers.

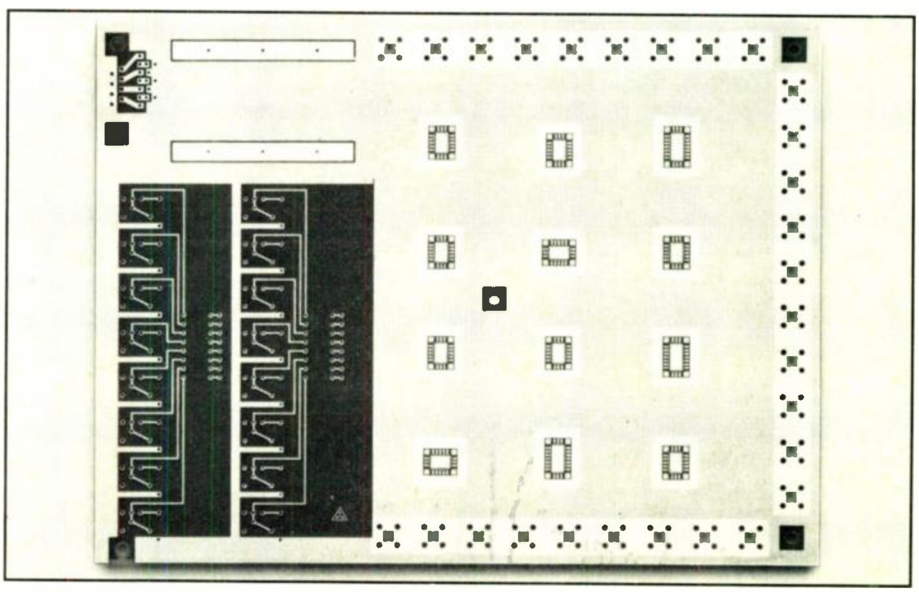


Insertion loss is measured at 6 dB (max). Maximum propagation delay is 250 ns for data to attenuation set and 220 ns for clear to attenuator reset. Set-up time for data to latch enable is 200 ns and hold time for latch enable to data off is the same. The high level input voltage is typically 3.4 V while the low level input voltage is 1 V (max). Maximum power dissipation is 600 mW and the junction temperature range is −55 to +125°C.

This digitally-controlled attenuator costs $39.50 when purchased in unit quantities. **Topaz Semiconductor, San Jose, CA. INFO/CARD #220.**

CEL Unveils Prototyping Board for GaAs Digital ICs

The MD12BRD four layer universal prototyping board is designed to provide an easy way to evaluate NEC high speed GaAs logic devices. The board has nine sites for 16-pin flatpack GaAs ICs, and allows access to power supply and RF interconnections. It was designed for device applications with clock inputs up to 5 GHz and 50 ohm impedance system operation.



The board can be used to build systems that run at speeds up to 4 GHz by using a variety of NEC GaAs digital ICs. Typical functions that can be performed with the board include input clock signal sensitivity, maximum operating frequency, propagation delay measurements, interfacing with multiple GaAs devices, rise and fall time measurements, output termination techniques and system prototyping. The MD12BRD is priced at $100 each. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #219.**

Logic Controlled Resistor Network

The FLJ-ACR1, -ACR2 are designed for use with resistor tuneable filters and offer four resistor networks with each network comprising of four resistors (R, R/2, R/4 and R/8). The R value of the ACR1 is 1.59 M ohms and ACR2 is 159 k ohms. The logic controlled resistor network products range from $36 to $79 in single quantities. **Datel, Inc., Mansfield, MA. INFO/CARD #218.**

Direct Digital Synthesizer

The Model SP2001 is a direct digital synthesizer that directly generates code required for a digital-to-analog converter to produce sinewave outputs over a range from 5 kHz to 100 MHz. This ECL 100K compatible numerically controlled oscillator accepts a 16-bit digital frequency setpoint and generates an 8-bit parallel output to drive a DAC and produce a quality sinewave. Time delays inherent in phase-locked loop synthesizers are eliminated, and hop time from frequency setpoint to frequency setpoint is only limited by DAC setting time. It is packaged in a 40-pin ceramic DIP and is priced at $675 in quantities of 100. **Plessey Semiconductors, Irvine, CA. Please circle INFO/CARD #217.**

Miniature Amplifier

JCA Technology introduces a miniature amplifier (Model JCA112- 215) that has a frequency range of 1 to 2 GHz. The minimum gain is 20 dB while gain flatness is ± 0.5 dB. Maximum noise figure is 2.0 dB and power output at the 1 dB compression point is +10 dBm. VSWR is 2:1 and the price is $350. **JCA Technology, Inc., Newbury Park, CA. INFO/CARD #216.**

RF Power Transistor

TRW RF Devices Division has introduced a Class C RF power transistor delivering 12 watts over a 2.3 to 2.7 GHz bandwidth with a typical gain of 6.8 dB. It is priced at $230 in quantities of 1 to 24. **TRW RF Devices Div., Lawndale, CA. INFO/CARD #215.**

DDS Module

The VDS-8 is a direct digital synthesizer module that covers from DC to 8 MHz in 0.10 Hz steps. The switching speed is under 500 ns with harmonics under −40

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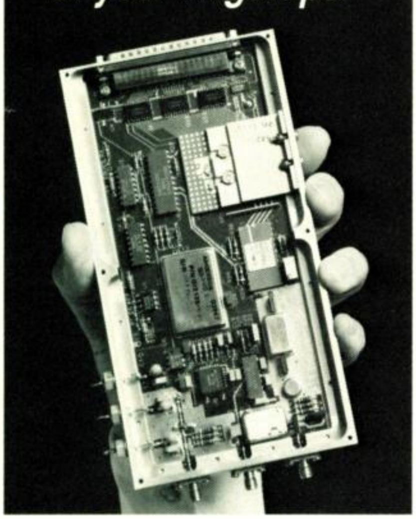
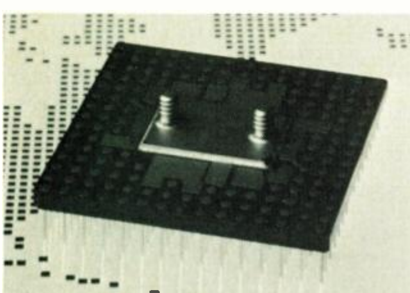


STEL-9272 Digital Direct Synthesizer provides a signal in the 2 KHz to 130 MHz band with a 1.1 Hz resolution. (Uses the STEL-2172 NCO.)

Features:
- 28-bit frequency resolution, 300 MHz clock
- 37 dB signal resolution at all frequencies up to 130 MHz
- Frequency changes as rapidly as every 125 nsecs
- Input signal levels TTL compatible
- Operating Temp: -20°C to +70°C.



The STEL-2172 ECL Numerically Controlled Oscillator generates sine waves at very precise frequencies at clock speeds of up to 300 MHz.
Features: 28-bit frequency resolution, 8-bit parallel sine output, 100K ECL outputs, TTL inputs.

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dBc and phase noise at 1 kHz measuring under −140 dBc/Hz. Divided by 100 to address the audio and sonar ranges, it can cover 80 kHz with under −100 dBc spurs. **Sciteq Electronics, Inc., San Diego, CA. INFO/CARD #214.**

Dual Output Synthesizer

PTS introduces the PTS 250 dual output frequency synthesizer. It has a frequency range of 1 to 250 MHz with resolution of 10 Hz to 100 kHz. The phase noise is -63 dBc at a frequency range of 0.5 Hz to 15 kHz. **Programmed Test Sources, Inc., Littleton, MA. INFO/CARD #213.**

Digital Attenuator

KDI announces the availability of a 0 to 103 dB attenuator in 1 dB steps which operates with a ±7 degree maximum phase shift. The Model DA-DO5S operates from 2.7 to 3 GHz with maximum input power of +27 dBm CW. Attenuation accuracy is ±0.5 dB up to 63 dB and ±0.7 dB over the full attenuation range. Other features include a maximum switching speed of 500 ns from 50 percent of input TTL logic to 0.25 dB of final selected value. The maximum insertion loss is 6.5 dB including loss of 2 input/output test couplers which are integrated in series with the attenuator. **KDI Electronics, Inc., Whippany, NJ. INFO/CARD #212.**

Sequence Detector/Descrambler

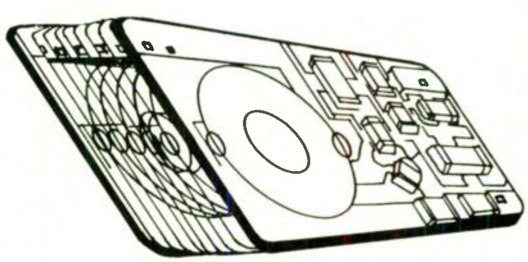
The Model 200 sequence detector/descrambler is a high speed, self-synchronizing sequence detector or data scrambler which operates from 10 MHz to 600 MHz. When used in companion with the Mode 100 sequence generator/scrambler, the Model 200 can measure bit error rate over high speed communication links. It is priced at $2,500. **Broadband Communications Products, Melbourne, FL. INFO/CARD #211.**

Reference Synthesizer

Operating at an output center frequency of either 5 or 10 MHz, the Model 3031B reference frequency synthesizer provides a digitally controlled ±4 kHz frequency deviation with 2 uHz resolution. Spurious components are below −100 dBc and phase noise is below −145 dBc at 100 Hz offset. The input reference can be either 5 or 10 MHz. Price is $14,950. **Pentek, Inc., Rockleigh, NJ. INFO/CARD #210.**

ECL CRYSTAL OSCILLATORS, NOW TO 350 MHz

AT CUT CRYSTAL OSCILLATOR WITH 10KH SERIES OUTPUT.



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FREQUENCY RANGE: 10 MHz TO 350 MHz
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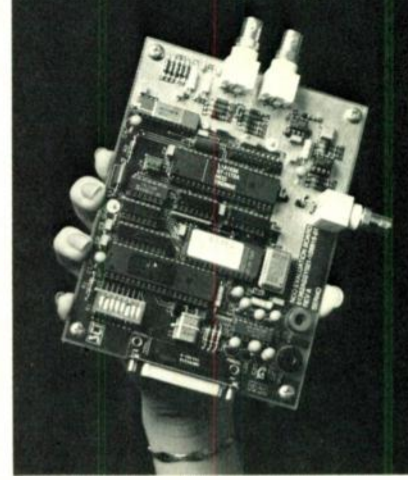
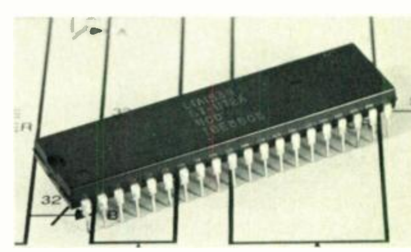


STEL-9172 Numerically Controlled Oscillator (NCO) Evaluation Board.

Features:
- Uses the ST-1172A NCO
- Sine Wave or Square Wave (TTL) Outputs
- 1 Hz to 10 MHz Operation
- Fixed Frequency Mode (Generates a fixed frequency from 1 Hz to 10 MHz)
- Swept Frequency Mode (Sweeps between a start and stop frequency at up to 10K steps/sec.)
- Stepped Frequency Mode (Steps through a user selected sequence of up to 8 frequencies at up to 10K steps/sec.)



The ST-1172A generates digital sine and cosine functions of very precise frequency at clock speeds of up to 25 MHz. Features: 32-bit frequency resolution, 8 bits sine and cosine amplitude resolution, 12-bit phase outputs, 300mw power dissipation at 25 MHz.

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rf products *Continued*

Two New Amplifiers Introduced

Model 3205 delivers 250 W linear power from 6 to 220 MHz with pulse widths over 250 ms. The pulse droop is less than three percent and blanking (turn off) is faster than 5 us. The Model 2060, on the other hand, delivers 15 W of Class A CW power with less than 1 dB compression from 20 to 200 MHz. **American Microwave Technology, Inc., Fullerton, CA. INFO/CARD #209.**

Frequency Synthesizer

Syntest introduces Model SI-102, a frequency synthesizer that operates from 0.1 Hz to 16 MHz. It features 5½ digits of resolution and a TTL output that is continuously adjustable into a 50 ohm load. Typical applications include calibration standard for test instrumentation, plotting and alignment of active and passive filters, precision variable clock for IC system testing, and radar range markers. It can also be used as a precision receiver local oscillator and is priced at $681. **Syntest Corp., Marlboro, MA. INFO/CARD #208.**

Wideband Op Amps

The CLC205 and CLC206 operational

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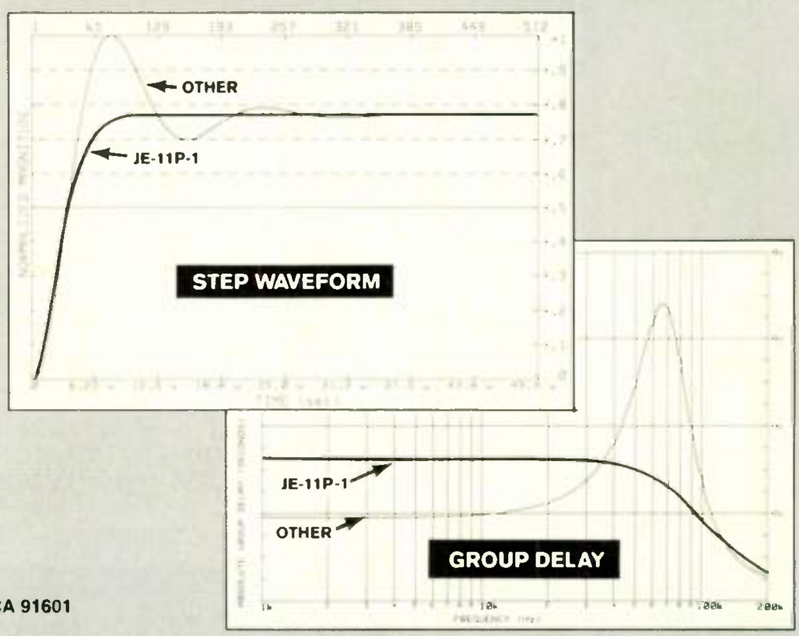

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amplifiers offer bandwidths of up to 180 MHz and rise times from 2.0 ns. The −3 dB bandwidth for the CLC205 is 170 MHz at a gain of +20 and settling time to 0.05 percent is 24 ns. The power consumption is 56 W at ±5 V. The CLC206 has a settling time of 19 ns and small signal bandwidth of 180 MHz. The drive capacity of the op amp is 100 mA — double that of the CLC205. The price of either device ranges from $56 to $138 depending on the version purchased. **Comlinear Corp., Fort Collins, CO. INFO/CARD #207.**

UHF Spectrum Analyzer

Penntek introduces the SA-500E 550 MHz spectrum analyzer. When connected to an X-Y display or oscilloscope, the instrument allows frequency domain signal viewing with an on-screen dynamic range of 70 dB. The display center frequency is adjusted from 1 to 500 MHz and is shown on the front panel meter. A 70 dB front panel RF input step-attenuator is included together with crystal controlled frequency markers at 5 to 50 MHz intervals. Uses include measuring harmonic signal levels, finding spurious signals, CATV signal level measurement, off-the-air signal analysis, production test and alignment, and two-way radio servicing. It is priced at $1495. **Penntek Instruments, Lewistown, PA. INFO/CARD #206.**

SC-Cut Crystal Filters

Piezo has developed crystal filters utilizing SC-cut crystals for the frequency range from 10 MHz to 80 MHz. Compared to AT-cut performance, power handling is as much as 10 dB higher and third order intercept is typically 6 dB. **Piezo Technology, Inc., Orlando, FL. Please circle INFO/CARD #205.**

GaAs Adder and Lookahead ICs

GigaBit Logic has released two additions to its family of Picologic™ GaAs digital ICs: the 10G100, 1.3 GHz 1200 ps delay expandable 4-bit adder and the companion 10G101, 1.4 GHz 675 ps delay carry lookahead generator. The carry lookahead can expand the adder's capability to handle up to 16-bit wide additions, and multiple 10G100s and 10G101s can implement fast adders of any larger word size. The 100 costs $59.50 and the 101 costs $55. **GigaBit Logic Inc., Newbury Park, CA. INFO/CARD #204.**

High Frequency Probe

A high frequency probe with low circuit loading, an accessory for spectrum and network analyzers, is being introduced by HP. The HP 85024A has 0.7 pF input capacitance shunted by an input resistance of 1 M ohm. The 3 dB bandwidth is nominally 100 kHz to 3 GHz. It is compatible with the HP 8568B, HP 8562A, HP 8590A and HP 4195A spectrum analyzers, and HP 8753A, HP 3577A and HP 4195A network analyzers. It comes with a 10:1 divider and is priced at $1,900. **Hewlett-Packard Company, Palo Alto, CA. Please circle INFO/CARD #199.**

Wideband Log Amps

RHG Electronics Laboratory introduces the Model LV750 log amplifier. It has a frequency range of 500 to 1000 MHz, input dynamic range of −60 to +5 dBm and input VSWR under 1.5:1. Video rise time is under 15 ns and video fall time is under 100 ns. **RHG Electronics Laboratory, Inc., Deer Park, NY. INFO/CARD #200.**

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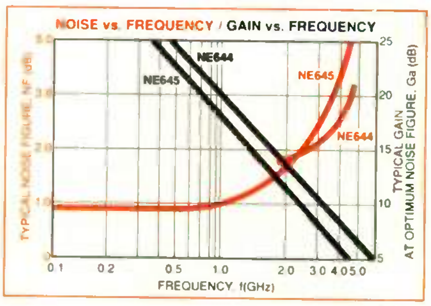

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Touchstone 1.6

EEsof introduces version 1.6 which features a sparse admittance matrix reduction capability, new full-feature screen editor commands, and increased variable and equation faculties. The sparse matrix reduction technique offers an alternative and powerful addition to the existing scattering parameter reduction analysis currently available in Touchstone. The user interface has been enhanced with new full-screen editor commands, including search and replace; repeat find command; file read/insert; and select, delete, move, copy, and undo block. Other additions include VSWR measurements, stripline cross and stripline curve elements, 512 frequency points, 10 two-port data files, sweep progress indicator, print and plot interrupt, Sony and NEC transistor data files, and the capability of acquire S-parameter measurement data form a network analyzer for direct use in its analyses and optimizations. Pricing starts at $9,900 depending on system and configuration. **EEsof, Inc., Westlake Village, CA. INFO/CARD #166.**

Transmission Line Analysis and Synthesis Software

=TLINE= handles microstrip, stripline, coplanar waveguide, coplanar waveguide with a ground, coaxial, coupled microstrip and coupled stripline. The routines used include the effects of dispersion, metalization thickness and surface roughness. The synthesis process optimizes impedances to the desired values using the analysis routines. Besides the normal dimensional and impedance information, TLINE gives line loss, effective dielectric constant, propagation velocity, highest frequency for accurate results, critical cover height for microstrip, coupling factor for coupled lines and even the unloaded Q when the line is used as a resonator. It is tailored for IBM or compatible machines and is priced at $395 for SuperStar owners and $595 otherwise. **Circuit Busters, Inc., Stone Mountain, GA. INFO/CARD #162.**

Schematic Entry Program

Intusoft introduces SPICE_NET, a schematic entry program for SPICE circuit simulators. With this program, a SPICE input deck is produced from a schematic description. It is designed for use with IBM PCs equipped with a mouse and graphics adapter. The price is $295. **Intusoft, San Pedro, CA. Please circle INFO/CARD #163.**

Analog Circuit Simulation Program

ECA-2 2.30 features SPICE compatible transistor models, graphics improvements and new transition function generators. The ECA-2 semiconductor models have been expanded to include Gummel-Poon BJT and Shichmann-Hodges FET models. It has a plotting swept values capability that allows the ECA-2 to simulate a transistor curve tracer. It costs $675 for IBM PC/XT/AT/PS-2. **Tatum Labs, Inc., Ann Arbor, MI. INFO/CARD #164.**

Program for Lumped Element Filter Synthesis and Analysis

LEFLTR performs electrical circuit synthesis, user-file circuit analysis, and two port frequency analysis. The program is capable of designing lowpass, highpass, bandpass, and bandstop filters. It is priced at $495. **Microwave Software Applications, Inc., Norcross, GA. Please circle INFO/CARD #167.**

New! 2.8 GHz Prescalers... in Silicon!

By directly synthesizing frequencies up to 2.8GHz our new low priced prescalers help you simplify your designs, minimize the number of components, and reduce board space.

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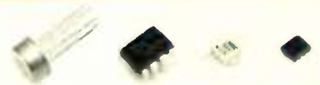
MODEL	FREQUENCY	V_{CC}	I_{CL}	DIVISION RATIO
UPB581	0.5–2.8GHz	5V	30mA	2
UPB582	0.5–2.8GHz	5V	45mA	4
UPB584	0.5–2.5GHz	5V	18mA	2
UPB585	0.5–2.5Ghz	5V	26mA	4
UPB587	0.05–1GHz	2.2 to 3.5V	5.5mA	2, 4, 8



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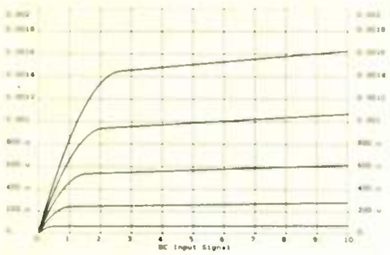
ANALOG CIRCUIT ANALYSIS
ECA-2



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ECA-2 IBM PC/XT/AT/PS-2 $675. PC Evaluation Kit $95.

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INFO/CARD 52

rf literature

RF/Microwave Products Catalog

This brochure gives information on RF and microwave filters, multiplexers, switch filter banks, and the design and production facilities of the company. The filters and multiplexers are listed by function, technology, and response. **Integrated Microwave, San Diego, CA. Please circle INFO/CARD #188.**

Signal Processing Handbook

This handbook contains information, detailed specifications, and performance curves for a line of RF, IF and microwave signal processing components spanning from DC to 6 GHz. Specifications for power splitters, frequency mixers, amplifiers, attenuators, switches, filters, frequency doublers, phase detectors, terminations, limiters, GaAs switches and drivers, RF transformers, and directional couplers are included. A technical article and outline drawings of the products are featured as well. **Mini-Circuits, Brooklyn, New York, NY. INFO/CARD #187.**

Digitizing Oscilloscope Brochure

Iwatsu Instruments has issued a brochure on its SAS-8130A digitizing oscilloscope. The functions of the scope together with illustrations, specification, and applications are included. **Iwatsu Instruments, Carlstadt, NJ. Please circle INFO/CARD #180.**

Data Conversion Devices Catalog

This catalog describes analog-to-digital converters and includes 12- to 16-bit resolution devices with conversion speeds ranging from 600 ns to 50 us. Video DACs mentioned are 4- and 8-bit devices with update rates as high as 125 MHz. Other devices include 12- to 16-bit DACs, sample/hold amplifiers, complete data acquisition subsystems, DC/DC converters, monolithic alarm circuits, and frequency to voltage converters. **Advanced Analog, Santa Clara, CA. Please circle INFO/CARD #172.**

Membership Directory

The Automatic Identification Manufacturers, Inc. (AIM) has published a directory that includes company names, addresses, phone numbers, and contacts for its 102 members. Included is a section that categorizes the products and services offered by the companies and in-

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PF865	20-400	16.5	8.0	+35
PF749-1	146-174	16.5	4.5	+35
PF829	406-512	16.5	4.5	+38
PF797A	800-960	19.5	5.0	+35
PF833	806-920	26.5	2.8	+34
PF845	890-915	18.0	2.0	+35
PF849F	825-851	16.0	1.0	+20

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cludes bar code scanners, printers and suppliers; printing vendors; optical character recognition, magnetic stripe, vision and voice systems, and supplies; RF identification and RF data communications equipment; software for automatic identification systems and systems integration/consulting services. **AIM, Pittsburgh, PA. INFO/CARD #184.**

Analog MMIC and Digital IC Brochure

A brochure covering NEC's GaAs analog MMIC and digital IC products is available from CEL. It outlines GaAs analog MMICs, GaAs prescalars, GaAs digital logic elements and GaAs fiber optic ICs together with a selection guide, packaging information, reliability data and product electrical specifications. **California Eastern Laboratories, Inc., Santa Clara. CA. INFO/CARD #182.**

Power MOSFET Data Book

The *MOSPOWER Data Book* from Siliconix provides a reference for industry-standard power MOSFETs and MOSPOWER products such as the dense-cell, large chip ELEFET and the low on-resistance MOSFETs. Detailed specifications for approximately 400 MOSPOWER transistors and small-signal MOSFETs can be found using part number cross-reference, selector guide and alphanumeric arrangement. Avalanche current ratings are included in the listings. There are sections devoted to die specifications, process flow, and package outlines. **Siliconix, Inc., Santa Clara, CA. Please circle INFO/CARD #176.**

Publications Catalog

A catalog of special publications, research publications, conference publications and technical papers (1977-1986) is available. The special publications are concerned with scientific and technical information derived from NASA programs. The research publications contain compilations of scientific and technical data. Conference publications record the proceeding of scientific and technical symposia and other professional meetings sponsored or cosponsored by NASA. Lastly, the technical papers present results of significant research conducted by NASA scientists and engineers. **National Technical Information Service, Springfield, VA. INFO/CARD #181.**

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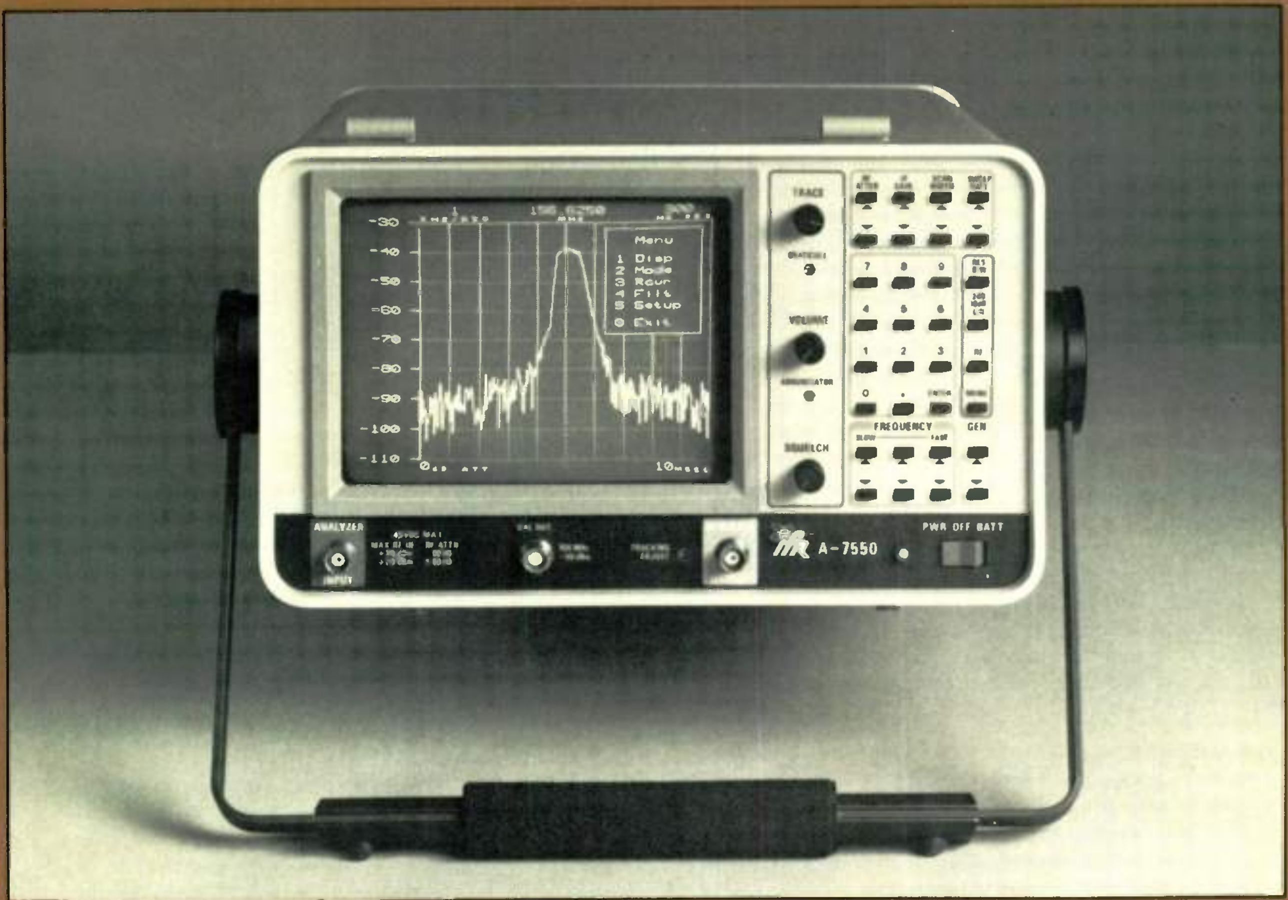


Runnerup Prize: The A-7550 Spectrum Analyzer from IFR, Inc. features 100 kHz to 1 GHz frequency coverage and microprocessor control that automatically selects and optimizes bandwidth, sweep rate, and slew rate of control functions. Single function keyboard entry and a menu driven display make operation easy and straightforward.

The runner-up prize is only slightly less grand than . . .

JUDGING CRITERIA

1. **Originality:** The purpose of the contest is to reward engineers for their unique design contributions. Each design will be evaluated according to its similarity to work by others, unusual application of a device or technique, and other judgments of its contribution to the advancement of the engineering craft.

2. **Engineering:** Engineering is the application of technology to solve a problem or meet a design goal. Entrants should clearly identify how their circuit was created in response to such a need. Judges will evaluate performance, practicality, reproducibility and economy.

3. **Documentation:** Communicating ideas to others is the business of *RF Design* and a necessary part of good engineering. Each entry will be judged on its description, analysis and graphical material. Each circuit should have a complete list of components, explanation of functions, and a summary of performance and test data.

ENTRY RULES

1. Entries shall be RF circuits containing no more than 6 single active devices (tubes or transistors), or 4 integrated circuits, or be passive circuits of comparable complexity.

2. The circuit must have an obvious RF function (as defined on page 6 of November 1987 *RF Design*) and operate in the below 3 GHz frequency range.

3. Circuits must be the original work of the entrant.

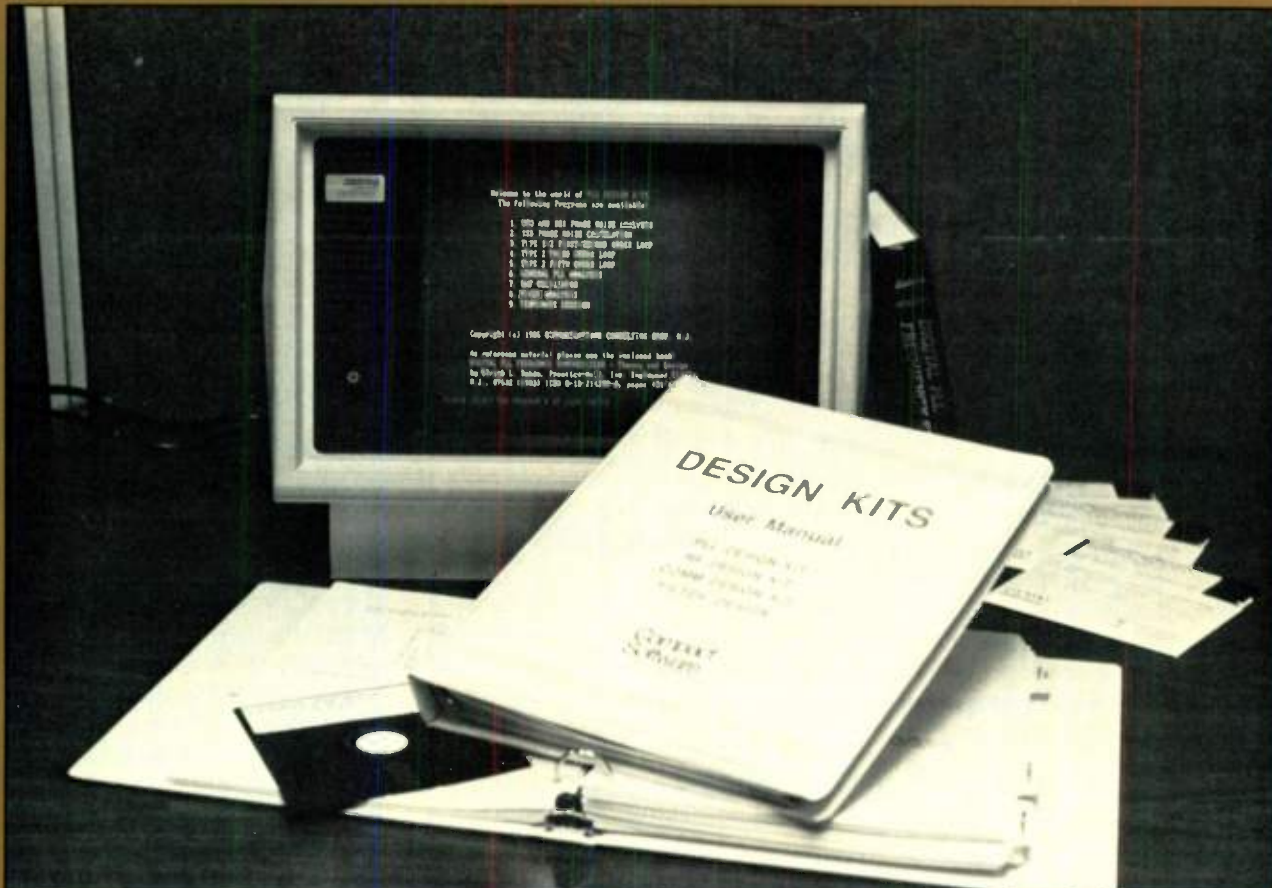
4. If developed as part of the entrant's employment, entries must have the employer's approval for submission.

5. Components used must be generally available, not obsolete or proprietary.

6. Submission of an entry implies permission for *RF Design* to publish the material. All prize winning designs will be published, plus additional entries of merit.

7. Winners shall assume responsibility for any taxes, duties or other assessments which result from the receipt of their prizes.

8. Deadline for entries: **March 31, 1988.**



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MAIL YOUR ENTRY BY MARCH 31, 1988 TO GARY BREED, EDITOR. . .

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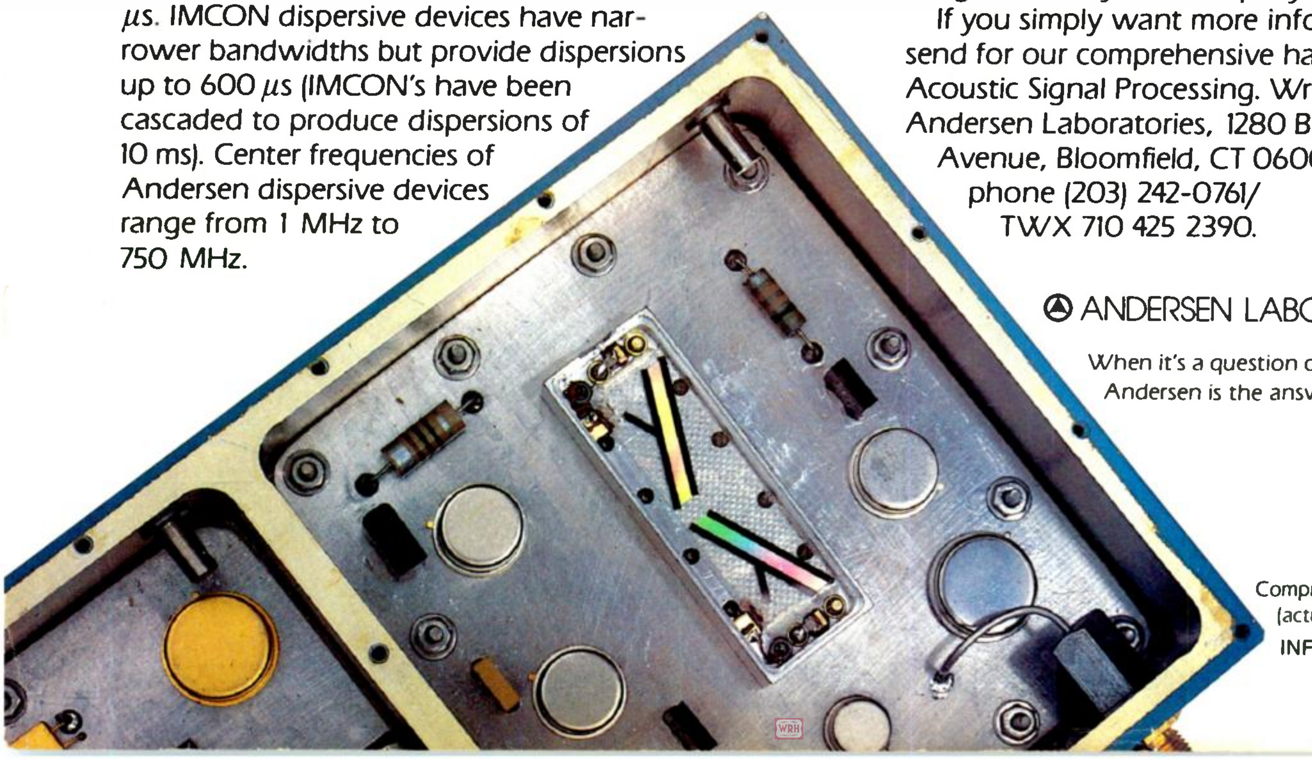

Compressive Receiver (actual size)

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