

RF design

engineering principles and practices

May 1991



Featured Technology
Test and Measurement
Crystal Oscillators
Quadrature Detectors

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TEXAS INSTRUMENTS INC
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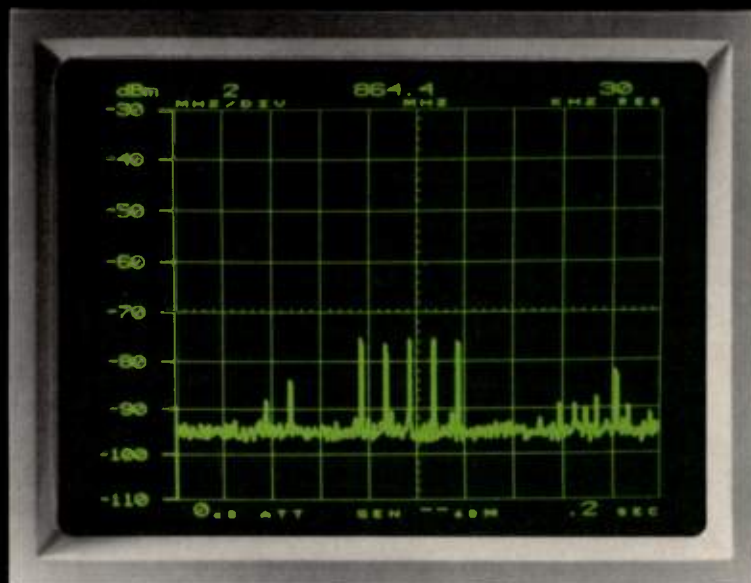
Specifications – Wavecom SP2T coaxial switches

Frequency Range (GHz)	DC - 3	3 - 8	8 - 12.4	12.4 - 18	18 - 26.5
VSWR (Max)	1.2:1	1.3:1	1.4:1	1.5:1	1.6:1
Insertion Loss (Max dB)	0.2	0.3	0.4	0.5	0.6
Isolation (Min dB)	80	70	60	60	50



LORAL
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If you're not using a spectrum analyzer...



Spectrum analyzer display of cellular telephone transmissions.

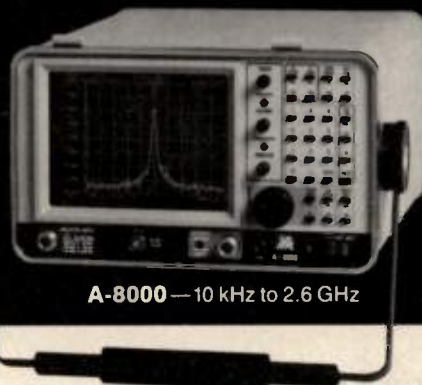
maybe it's time to take a closer look.



Spectrum Analyzers



A-7550 — 10 kHz to 1.0 GHz



A-8000 — 10 kHz to 2.6 GHz

You can't hear them or see them, but there are probably thousands of electromagnetic signals being transmitted through the air around you at this very moment.

Many of these signals originate from radio and television transmission. However, many other signals are emitted from computers, CRT display terminals, and other electrical devices.

Each of these signals can be a potential source of harmful interference to sensitive electrical equipment or critical systems you are operating.

The IFR A-7550 and A-8000 Spectrum Analyzers provide a full set of features to speed the identification and measurement of signals over a wide frequency range.

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DS0699	SPST	5-1500	1.0	48	23.0	TO-8	TTL
DS0990	SPST	10-2000	1.5	62	25.0	14 Pin DIP	TTL
DS0850	SP2T	DC-2000	0.4	47	5.0	TO-5	0/-7
DS0813	SP2T	DC-2000	0.7	40	140.0	TO-5	TTL
DS0812	SP2T	10-1000	0.5	54	50.0	TO-8	TTL
DS0860	SP2T	10-1000	0.5	47	50.0	.380 sq	TTL
DS0602	SP2T	5-9000	1.3	65	26.0	14 Pin DIP	TTL
DS0842	SP2T	5-1500	1.0	75	50.0	14 Pin DIP	TTL
DS0864	SP4T	5-2000	1.3	49	28.0	16 Pin DIP	TTL
DS0874	SP4T	DC-2000	2.2	60	55.0	14 Pin DIP	TTL
DS0833	SP8T	5-1000	2.2	40	40.0	24 Pin DIP	TTL
DS0918	SP16T	DC-4000	3.8	30	150	58 Pin DIP	TTL

*Typical performance measured mid-band

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featured technology

26 A Channel Sounder for Indoor Communication Systems

The authors describe a test transmitter and receiver for characterizing indoor radio propagation at L-band. Path performance measurements are required if reliability of factory or other radio systems is to be assured.

— Fernando Casadevall and Xavier Barba

40 High Power Bi-Directional Coupler for Space Applications

Control and protection of high reliability circuits such as spaceborne power amplifiers is critical. This article describes a high power coupler designed to meet specific coupling and directivity requirements.

— Pascal R. Goux

emc corner

56 Surge and Transient Considerations

A review of the causes and nature of surges and transients on power utility and common carrier signal lines, plus notes on protection of electronic equipment from their effects.

— Gary A. Breed

design awards

58 A Simple Clock and PSK Carrier Recovery Circuit

This month's featured contest entry is a simple circuit that simultaneously recovers clock and data information from a PSK modulated signal.

— Francois Methot

61 VHF and UHF Crystal Oscillators — Part II

Another note from *RF Design's* Consulting Editor on his experiments with some remarkably simple high frequency fundamental and overtone oscillator circuits.

— Andrzej B. Przedpelski

68 Design of Quadrature Detectors

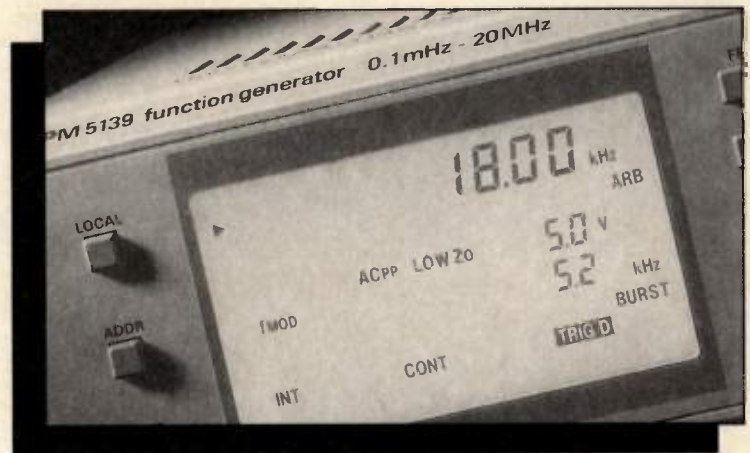
A review and analysis of the quadrature detector, typically the lowest cost option for demodulation of frequency and phase modulated information.

— Edward A. Richley

73 CAD of Single and Tandem Lange Couplers

The authors show how two Lange couplers connected in tandem can solve the problem of small dimensions at microwave frequencies. They provide a computer program for design of both single and tandem couplers.

— Timothy O'Connell and Patrick Murphy



departments

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Our cover, provided by John Fluke Mfg. Co., represents our emphasis on engineering test and measurement. More information on this product can be found on page 48.

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D2655	174-216 MHz

TYPICAL SPECIFICATIONS

INS. LOSS	0.15db
ISOLATION	25db
VSWR	1.2:1
POWER	2Kw
Connectors	
input	N
output	1-5/8 EIA

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decades ahead

RF editorial

Engineering Productivity Part II — Test Instruments

By Gary A. Breed
Editor



Last month, I noted some of the organizational changes that are occurring in the U.S. electronics industry. To effectively improve efficiency, these changes require that companies get the maximum results from their employees' efforts. For RF companies, this is an especially difficult assignment, since experienced RF designers are in relatively short supply and many are already extremely busy.

One method of improving an RF engineer's productivity is to provide support that allows him more time for creative engineering. Recent developments in the world of test and measurement certainly fit this description, with instruments that not only offer improved performance, but have operating features that minimize setup time.

The principal enhancement in newer test instruments is the incorporation of computing power. This allows the user to store standard setups, eliminating all the knob-twisting that used to be needed. Errors in those dial settings are eliminated, too. Computations based on the measured data can be performed as part of the pre-programmed test procedure. For example, a spectrum analyzer can list the harmonic levels relative to the fundamental, or identify spurious signal levels and frequencies.

All this power can be utilized in a multi-unit test setup. A modulated signal generator, spectrum analyzer, power supply and DC multimeter could be connected into a power amplifier test system, controlled by a host computer. A little extra time to program the system eliminates lengthy and tedious manually-operated measurements. You might not need a computer, either. The controller might be inside one of these instruments, since a few already have a keyboard option that is all they need to

be stand-alone PCs!

Performance features improve the accuracy and reliability of measurements, especially in high performance testing. Improved measurement confidence reduces the need to repeat a test "just to make sure." Lower noise synthesizers in signal generators, spectrum analyzers and network analyzers have increased the dynamic range of these instruments' measurements. Direct digital synthesis and digital modulation allows generation of complex waveforms that previously required combinations of several sources. Baseband FFT analysis in spectrum analyzers dramatically increases the precision of their measurements. Internal calibration routines make sure that these powerful functions are operating as intended when you need them.

These advantages are not limited to the designer's bench — production line testing may receive an even bigger benefit. Faster, more accurate measurements are exactly what the production test manager wants! Pre-programmed setup takes most of the human error out of the instrument operation and data collection process. To meet the needs of manufacturing, a new generation of instruments with "practical" rather than "ultimate" performance specifications is beginning to show up.

So, the next time you find yourself wishing for some new test equipment, take the time to evaluate how much time you spend preparing for tests, running them by hand, and manually tabulating the results. If you are under pressure to deliver your designs on increasingly shorter timetables, this information might be all that is needed to justify the investment. Your company will get more out of your talents, and you'll get greater satisfaction doing it.

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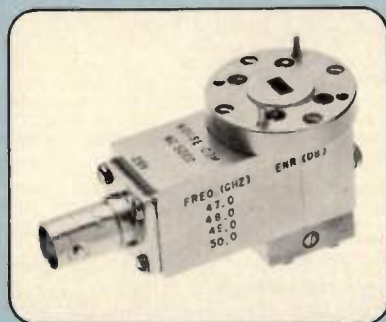
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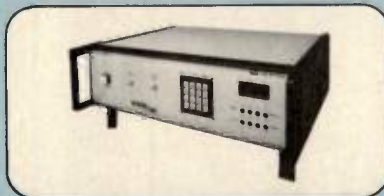


TYPICAL STANDARD MODELS

NC 6101	up to 20 kHz
NC 6107	up to 100 MHz
NC 6108	up to 500 MHz
NC 6109	up to 1 GHz
NC 6110	up to 1.5 GHz
NC 6111	up to 2 GHz
NC 6218	up to 18 GHz

Other standard models available
MOST ARE IN STOCK

PROGRAMMABLE
IEEE-488 (GPIB), MATE (CII)
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TYPICAL STANDARD MODELS

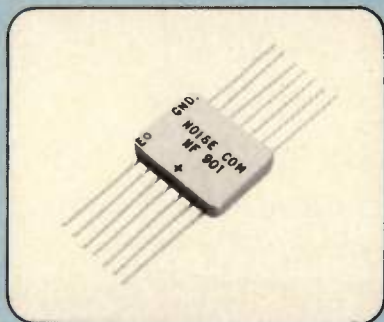
NC 7101	up to 20 kHz
NC 7107	up to 100 MHz
NC 7108	up to 500 MHz
NC 7109	up to 1 GHz
NC 7110	up to 1.5 GHz
NC 7111	up to 2 GHz
NC 7218	up to 18 GHz

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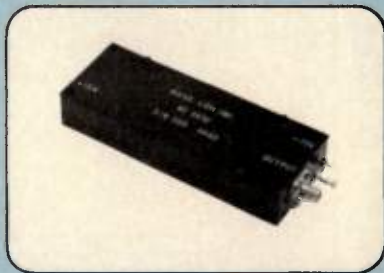
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thereafter

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0° - +50°C, ± 5 PPM	*CMOS = 1KHz to 10MHz
0° - +70°C, ± 5 PPM	ECL = 4MHz to 32MHz
-30° - +70°C, ± 10 PPM	SINE = 4MHz to
-30° - +85°C, ± 15 PPM	

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AT 50 (3)	50 (5W)	DC-1.5GHz	17.50	29.00	22.00	20.60				12.00
AT 51	50 (5W)	DC-1.5GHz	15.00	26.00	19.50	17.40				
AT 52	50 (1W)	DC-1.5GHz	20.50	29.00	26.00	39.00				
AT 53	50 (1.25W)	DC-3.0GHz	20.50	26.00		15.00			18.00	
AT 54	50 (2.5W)	DC-4.0GHz				20.00				
AT 55	50 (2.5W)	DC-4.0GHz				19.00				
AT 75 or AT 90	75 or 93 (5W)	DC-1.5GHz	17.50	26.00	45.50	19.50				
Detector Mixer, Zero Bias Schottky:										
DM 51 75	50 75	01-4.2GHz	84.00			84.00				
DM 91	50	01-4.2GHz				84.00				
Resistive Impedance Transformers, Minimum Loss Pads:										
RT 50 75	75 in 75	DC-1.8GHz	17.50	20.00	43.30	17.50				
RT 50 93	93 in 93	DC-1.0GHz	17.50	26.00	45.50	17.50				
Terminations:										
CT 10 (3)	50 (5W)	DC-4.2GHz	11.50	15.00	15.00	17.50				
CT 11	50 (5W)	DC-4.2GHz	8.50	12.00	14.00	9.50			9.00	
CT 12	50 (1W)	DC-2.5GHz	10.50	15.00	15.00	13.00		15.50		
CT 13	50 (1.5W)	DC-4.0GHz	5.50			5.80				
CT 14	50 (2W)	DC-2.0GHz	14.00	15.00	15.00	17.50				
CT 15	75 (2.5W)	DC-2.5GHz	10.50	15.00	15.00	13.00		15.50		
CT 16	93 (2.5W)	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					
MT 75	75	DC-1.0GHz				45.50				
Feed thru Terminations, shunt resistor:										
FT 50	50	DC-1.0GHz	17.50	26.00	19.50	17.50				
FT 75	75	DC-500MHz	17.50	26.00	45.50	17.50				
FT 93	93	DC-150MHz	17.50	26.00	45.50	17.50				
Directional Coupler, 30dB:										
DC 300	50	250-500MHz	80.00	84.00	84.00					
Reciprocal Decoupler, series resistor or Capacitive Coupler, series capacitor:										
TD or CC-1000	1000 (1000PF)	DC-1.5GHz	17.50	26.00	19.50	17.50				
Adapters:										
GA 10 (N to SMA)	50	DC-4.2GHz	17.50	26.00	19.50	17.50				
Inductive Decouplers, series inductor, Bias T:										
LD 115	0.17uH	DC-500MHz	17.50	26.00	19.50	17.50				
LD 116	6.0uH	DC-500MHz	17.50	26.00	19.50	17.50				
LD 50	1.8uH	15-300MHz	84.00	84.00	84.00	84.00				
Fixed Attenuator BNC, 3, 6, 10 and 20 dB, in plastic case:										
AT 50 SET (3)	50	DC-1.5GHz	78.00	120.00	99.00	84.00				
AT 51 SET	50	DC-1.5GHz	84.00	108.00	82.00	74.00				
Relative Multipliers, 2 and 4 output ports:										
TC 125-2	50	1.5-125MHz	84.00			84.00				
TC 125-4	50	1.5-125MHz	94.00			104.00				
Resistive Power Dividers, 3, 4 and 8 ports:										
RD 150	50	DC-1.0GHz	84.00	84.00	84.00	84.00				
RD 4-50	50	DC-300MHz	84.00	84.00	84.00	84.00				
RD 9-50	50	DC-500MHz	84.00	84.00	84.00	104.00				
RD 3-75, 4-75	75	DC-300MHz	84.00	84.00		84.00				
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	17.50	26.00	45.50	17.50				
FL 75	75	DC-1.5GHz				26.00				

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For Most People, It Was Just Another Warm September Day.

For design engineers, it was the day mixed analog/digital design came of age.

The event was the mixed-signal design demonstration at the IEEE Bipolar Circuits and Technology Meeting (BCTM). The goal was to give credence to mixed-signal simulation and to benchmark companies in the marketplace. The results were conclusive.

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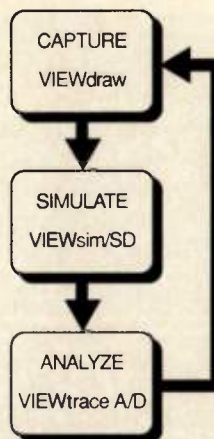


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.3 - 100MHz	100-250MHz	200-400MHz	200-500MHz	100-500MHz
100 Watts CW	100 Watts CW	100 Watts CW	100 Watts CW	100 Watts CW

MODEL LA200H	MODEL LA200F	MODEL LA200U	MODEL LA200UE	MODEL LP300H
.5 - 100MHz	100-250MHz	200-400MHz	250-500MHz	.3 - 100MHz
200 Watts CW	200 Watts CW	200 Watts CW	200 Watts CW	300 Watts Pulse

MODEL LA400U	MODEL LA500H	MODEL LA500V	MODEL LA500U	MODEL LA1000H
200-400MHz	5-50MHz	10-100MHz	200-400MHz	2-32MHz
400 Watts CW	500 Watts CW	500 Watts CW	500 Watts CW	1000 Watts CW

MODEL LA1000V	MODEL LA2000H1	MODEL LA3000HS	MODEL LP4000HV2	MODEL LP12000HV
10-100MHz	2-30MHz	5-45MHz	2-50MHz	2-50MHz
1000 Watts CW	2000 W pk, 1000W CW	3000 Watts CW	4000 Watts pk. pulse	12KW peak pulse

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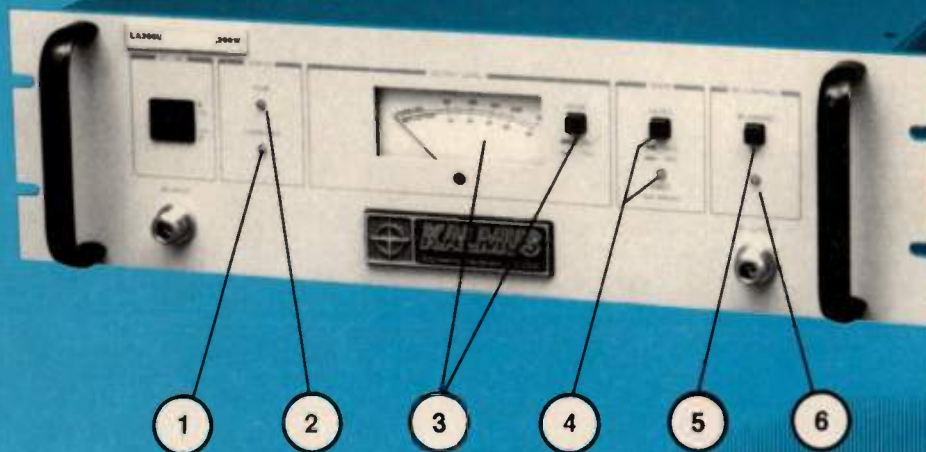
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100W to 12KW ALL MOS-FET RF POWER AMPLIFIERS

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- 47-440Hz AC power capability from 97 to 265VAC.
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LA 3000 AND 12000HV
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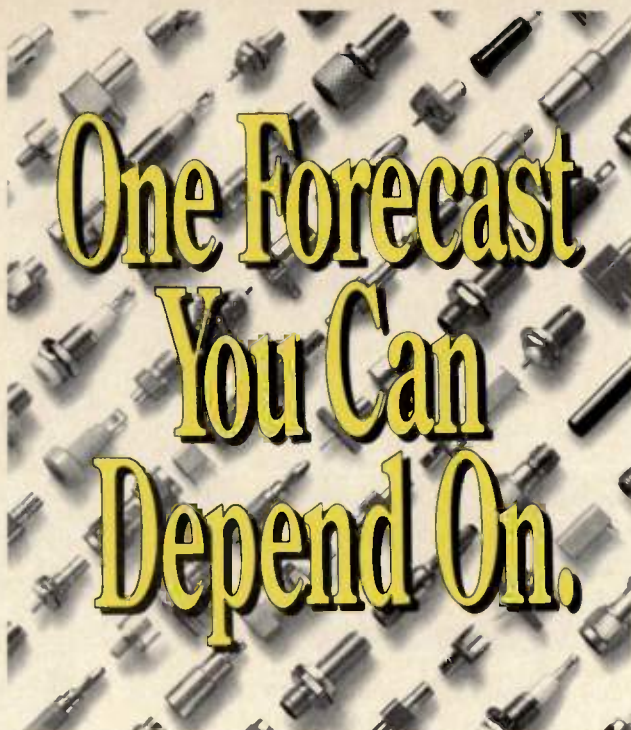


LA 500/600/1000
AND 4000HV



LA 2000





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INFO/CARD 10

RF letters

Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

Obsolete or Absolute?

I read your editorial on obsolete technologies with much interest and a bit of puzzlement.

I think you are way off base, 180 degrees out of phase with reality, about personal communications making "inconvenience obsolete." The plain truth is that such a system would make inconvenience *complete*. If you doubt this point of view, I challenge you to make a simple test. Just give all the secretaries around you a week off and for one week, answer your phone yourself every time it rings. And, of course, you will do this with no lessening of your regular work output.

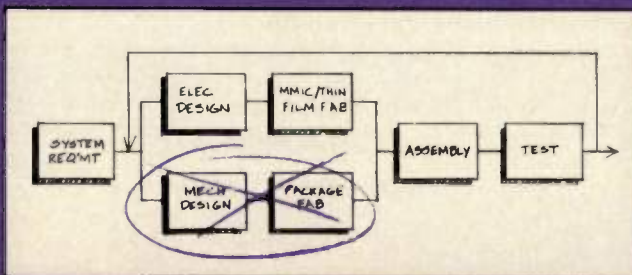
Hanging one of those infernal machines around your neck so that people could call you up and talk about a subject of their choosing at a time of their choosing, with absolutely no thought for your convenience would be — well, the inner circle of hell! The phone is getting to be an invasion of privacy already, but what would it be like if there were nowhere you could go to get away from it?

Add this to your list of pet peeves, and contemplate that we now have the technology to bring all of them to pass.

C.A. Chrestien
Sunnyvale, CA

Short cycle your circuits.

If you need a faster way to get your designs to market, try the new HP 83040 Series Microcircuit Package. Compared to custom packaging and test fixtures, it can cut weeks off your design time. And thousands of dollars off your development costs.



Pre-designed and off-the-shelf, the HP 83040 is ready for prototype and IC testing. Its modular format is compatible with CAE systems and network analyzer calibration techniques. Call your local HP sales office or circle the inquiry number below.

There is a better way.



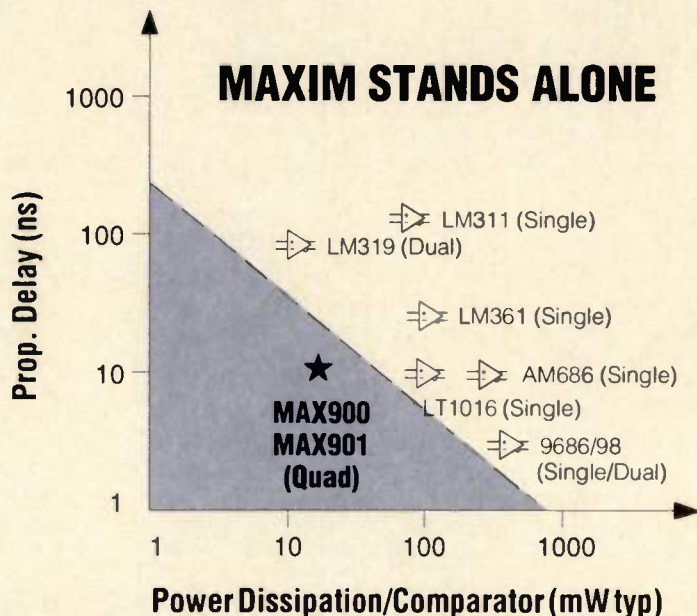
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8ns, 18mW—ONLY \$1.50*

- ◆ +5V Operation Cuts Power Drain and Heat Dissipation
- ◆ Input Range Includes Ground and Eliminates Need For Negative Supply
- ◆ Four Comparators in One Package Saves Board Space and Reduces Cost
- ◆ Separate AGND and DGND Minimizes Noise, Improves System Performance
- ◆ Dual Supply $\pm 5V$ Capability Allows Bipolar Input Range



One Stop Shopping For High Speed/Low Power

Maxim's new MAX900 and MAX901 quad comparators can deliver both high speed *and* low power at a price that compares favorably to high speed single or dual equivalents. MAX900/MAX901 offer you propagation delays of only 8ns with a 5mV overdrive, power dissipation of 18mW per comparator (when powered from a +5V supply) and space saving 0.3" DIP or small outline (SO) packages - for only \$1.50 per comparator.

Maxim's High Speed Comparator Family

Part Number	# Comps	Logic	Delay (typ)	Latch	Package	Price [†]
MAX900	4	TTL	8.0ns	Yes	DIP, SO	\$7.01
MAX901	4	TTL	8.0ns	No	DIP, SO	\$5.98
MAX9685	1	ECL	1.3ns	Yes	DIP, SO, Can	\$3.38
MAX9686	1	TTL	6.0ns	Yes	DIP, SO, Can	\$2.31
MAX9687	2	ECL	1.4ns	Yes	DIP, SO	\$5.12
MAX9690	1	ECL	1.3ns	No	DIP, SO	\$3.29
MAX9698	2	TTL	6.0ns	Yes	DIP, SO	\$3.92

* MAX901, 1000-up F.O.B. USA price per comparator † 1000-up F.O.B. USA

Call your Maxim representative or distributor today for applications information, datasheets and samples. Or, write Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, (408) 737-7600, FAX (408) 737-7194.

MAXIM

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RF calendar

May

14-16

IEEE Instrumentation and Measurement Technology Conference

Omni Hotel, Atlanta, GA
Information: Robert Myers, 3685 Motor Avenue, Ste. 240, Los Angeles, Ca 90034. Tel: (213)-1463. Fax (213) 287-1851.

29-31

45th Annual Symposium on Frequency Control

Los Angeles Airport Marriott, Los Angeles, CA
Information: Annual Symposium on Frequency Control, PO Box 826, Belmar, NJ 07719.

June

3-5

First Virginia Tech Symposium on Wireless Personal Communications

Virginia Tech, Blacksburg, VA
Information: Prof. Theodore S. Rappaport, Director, Mobile & Portable Radio Research Group, Virginia Polytechnic & State University, Blacksburg, VA 24061 -0111.

4-6

'91 International Communications Association Expo

Anaheim Convention Center, Anaheim, CA
Information: Angela Gallagher, ICA, 12750 Merit Drive, Suite 710, LB-89, Dallas, Texas, 75251. Tel: (214) 716-4124.

11-13

MTTS International Microwave Symposium and Expo

Boston, MA
Information: Tel: (617) 769-9750.

13-14

ARFTG

Boston, MA
Information: ARFTG, c/o Henry Burger, 1061 E. Frost Drive, Tempe, AZ 85282. Tel: (602) 839-6933.

24-27

Test Engineering Conference

Georgia World Congress Center, Atlanta, GA
Information: Miller Freeman Expositions. Tel: (800) 223-7126 or (617) 232-3976.

25-27

EMC Expo '91

Walt Disney World Resort, Orlando, FL
Information: Ellen Lunsford, Registration. Tel: (703) 347-0030.

July

15-18

Communication Networks West '91

Moscone Center, San Francisco, CA
Information: Conference Sales, Tel: (800) 225-4698.

22-24

Fifth International Conference on HF Radio: Systems and Techniques

Edinburgh Conference Centre, Edinburgh UK
Information: Secretariat, Conference Services, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 240 1871 ext. 222.

August

13-15

IEEE 1991 International Symposium on Electromagnetic Compatibility

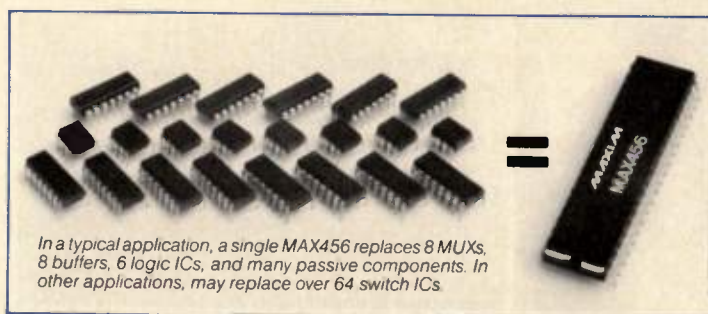
Cherry Hill, NJ
Information: IEEE International Symposium on EMC, PO Box 609, Lincroft, NJ 07738. Tel: (201) 386-2378.

8×8 VIDEO CROSSPOINT SWITCH WITH BUFFERS -ONLY \$2.50*/CHANNEL

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Maxim's new **MAX456** is the **first** monolithic 8×8 video crosspoint switch that routes standard video signals (NTSC, PAL, SECAM). With a digitally controlled 8×8 switch matrix, control logic, and eight 35MHz output buffers together in a 40-pin DIP or 44-pin PLCC, the MAX456 significantly reduces component count, board space and cost over discrete designs. Applications include video surveillance, imaging, visual automation, and video editing.

MAX456 Eliminates Over 20 Components

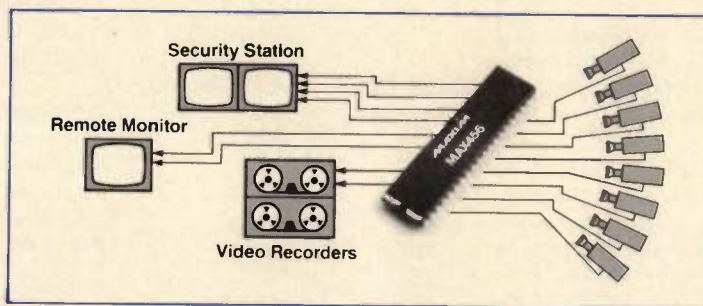


- Reduces Board Space up to 5X
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- Reduces Design and Layout Time
- Reduces Stray Capacitances
- Improves Reliability

Build Larger Crosspoint Arrays

Each MAX456 buffer output can be disabled under logic control. With three-state outputs, multiple MAX456s can be paralleled to form larger switch networks.

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250V/ μ s Slew Rate
35MHz Bandwidth
Buffer Disable Saves Power
- ± 5 V Power Supplies
- 80dB Off Isolation at 5MHz
- 70dB Crosstalk at 5MHz
- Serial or Parallel μ P Interface



MAX456 and MAX457s Drive 75 Ω Loads

Maxim also offers the MAX457, a dual 70MHz unity gain stable video amplifier. The MAX456 teams up with the MAX457 to drive 75 Ω loads efficiently. Special pricing is available for MAX456/MAX457 combination purchases.

Call your Maxim representative or distributor today for applications information, data sheets, and free samples. Or, contact Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, (408) 737-7600, FAX (408) 737-7194

* MAX456CPL, \$1998 1000-up
F.O.B. U.S.A price



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RF courses

Fundamentals of Radar Cross Section

May 20-24, 1991, San Diego, CA
Information: Kelly Brown, SCEEE, Tel: (407) 892-6146.

Microwave/Millimeter Wave Monolithic Integrated Circuits

June 4-7, 1991, Los Angeles, CA
Radiation Hardening of Electronic Systems
June 10-14, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Frequency Hopping Signals and Systems

May 20-22, 1991, Washington, DC
Principles of High Frequency Radio Communications: Applications for Operators and Managers
May 20-23, 1991, Washington, DC
New HF Communications Technology: Advanced Techniques

June 3-7, 1991, Washington, DC
Spread Spectrum Communications Systems
June 10-14, 1991, Washington, DC
Introduction to Modern Radar Technology
June 26-28, 1991, Washington, DC
Ionospheric Radio Propagation for System Planners
July 8-11, 1991, Washington, DC

Radio Frequency Spectrum Management
July 8-12, 1991, Washington, DC
Synchronization in Spread-Spectrum Systems
July 15-19, 1991, Washington, DC

Information: The George Washington University, Continuing Engineering Education, Merrill A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Fiber Optic Communications

June 10-12, 1991, Sunnyvale, CA
Information: Center for Professional Development, Arizona State University. Tel: (602) 965-1740.

Digital Signal Processing Workshop

June 12-14, 1991, Norwood, MA
July 16-18, 1991, Campbell, CA

Mixed Signal Design Seminar

May 15, 1991, Pleasanton, CA
May 16, 1991, Bellevue, WA
May 17, 1991, Beaverton, OR
May 20, 1991, Woburn, MA
May 21, 1991, Montreal, Canada
May 22, 1991, Bloomington, MN
May 23, 1991, Houston, TX
May 24, 1991, Dallas, TX
May 28, 1991, Phoenix, AZ
May 29, 1991, Denver, CO
May 30, 1991, Arlington, Heights, IL
May 31, 1991, Rochester, NY

Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Touchstone/Academy

June 3-7, 1991, Gilching, Germany

Libra/Academy

June 17-21, 1991, Baltimore, MD
June 17-21, 1991, Gilching, Germany
Information: EEsof. Tel: (818) 991-7530. Fax: (818) 991-7109.

Frequency-Time and Spatial-Time Signal Processing

June 10-14, 1991, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

Modern Power Conversion Design Techniques

May 20-21, 1991, San Rafael, CA
July 15-19, 1991, Chicago, IL
Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

ELINT Interception

May 14-16, 1991, Syracuse, NY
ELINT/EW Applications of Digital Signal Processing
May 14-16, 1991, Syracuse, NY
Integrated EW

May 21-22, 1991, Syracuse, NY
ELINT/EW Data Bases

May 21-23, 1991, Syracuse, NY

Radar Vulnerability to Jamming

June 4-5, 1991, Syracuse, NY

Electromagnetic Propagation

June 4-6, 1991, Syracuse, NY
Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Basic Network Measurements Using the 8510B Network Analyzer

May 29-31, 1991, Los Angeles, CA
June 11-13, 1991, San Francisco, CA
June 18-20, 1991, Atlanta, GA
June 25-27, 1991, Los Angeles, CA
Designing for EMC
July 11-12, 1991, Atlanta, GA
Information: Hewlett-Packard Company. Tel: (714) 999-6700.

Introduction to Fiber Optic Communications

June 4-7, 1991, San Francisco, CA
Digital Signal Processing: Techniques & Applications
June 4-7, 1991, San Francisco, CA
June 18-21, 1991, Washington, DC

Hands-On Datacomm Troubleshooting

June 4-7, 1991, Washington, DC
June 25-28, 1991, San Francisco, CA

Introduction to Telecommunications

June 11-14, 1991, Los Angeles, CA
June 25-28, 1991, Washington, DC
Information: Learning Tree International. Tel: (800) 421-8166, (703) 893-3555, (203) 417-8888.

Introduction to EMI/RFI/EMC

May 21-23, 1991, Washington, DC
June 24-25, 1991, Orlando, FL
EMI Control Methodology and Procedures
May 21-24, 1991, Philadelphia, PA

Practical EMI Fixes

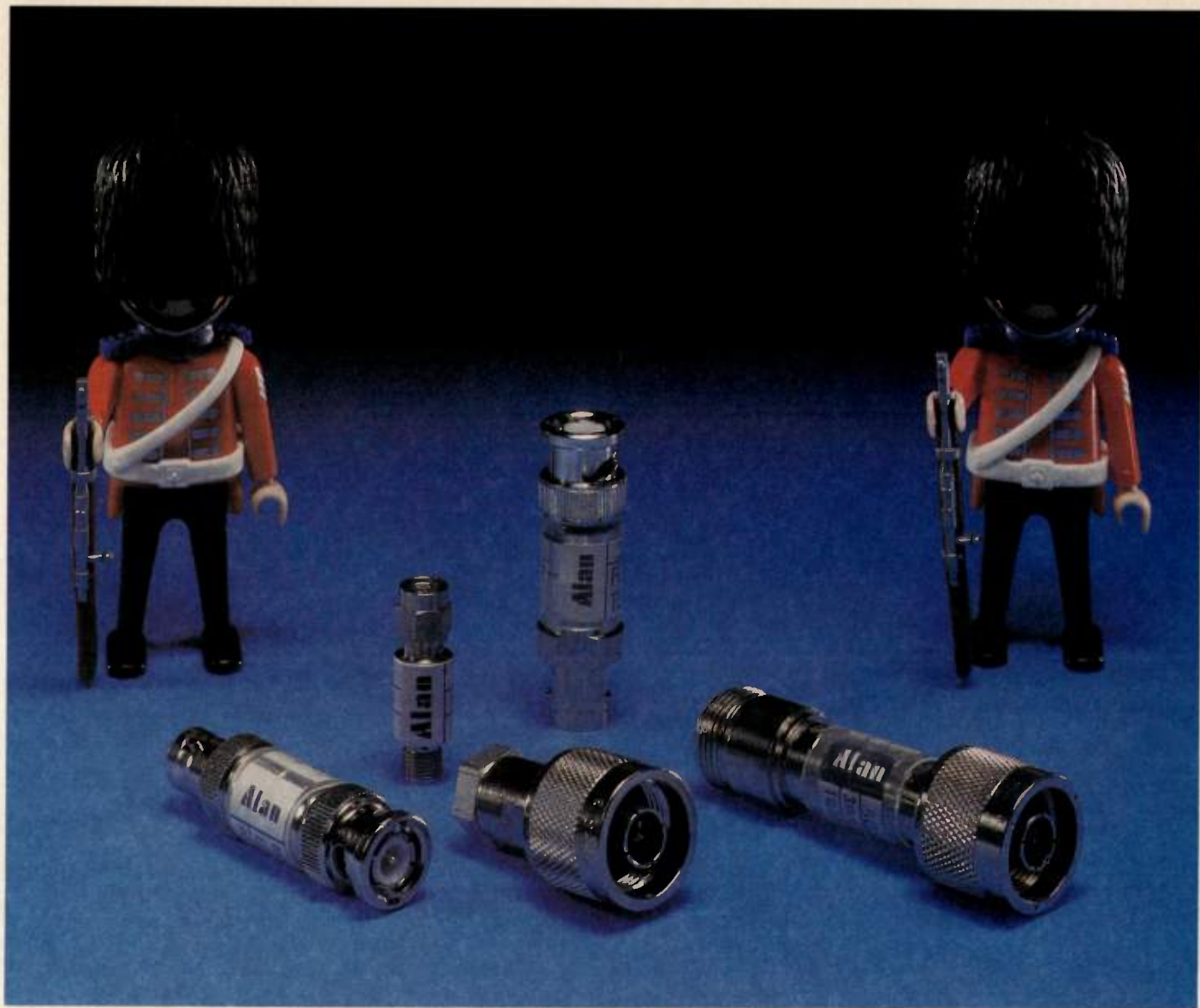
June 3-7, 1991, Las Vegas, NV

Grounding and Shielding

June 24, 27, 28, 1991, Orlando, FL

System Integration and Design for EMC

June 24, 25, 28, 1991, Orlando, FL
Information: Interference Control Technologies, Registrar. Tel: (703) 347-0030.



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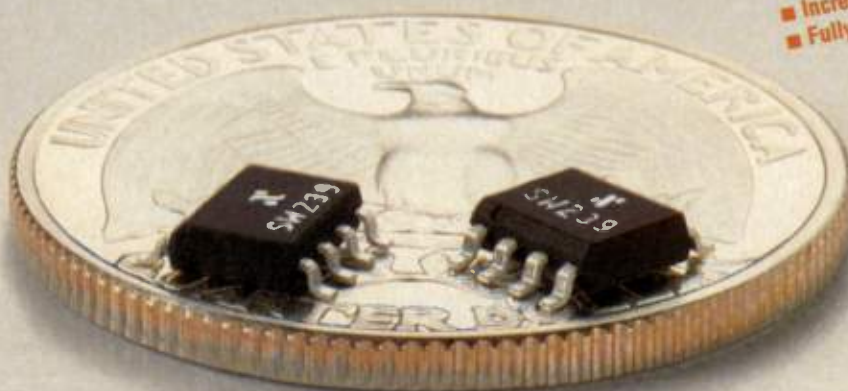
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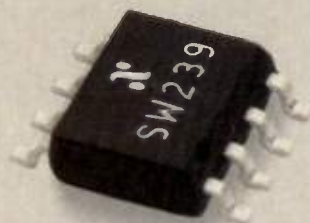


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The SW-239 SPDT RF switch features nanosecond speed with only microwatts of DC power.

Utilizing an 8-Lead SOIC package these units come ready for full automatic insertion.



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MODEL SW-239 Performance Characteristics	
Frequency Range	DC-2000 MHz
Operating Temperature	-40°C to +85°C
Insertion Loss	0.5 dB
VSWR	1.1:1
Isolation	50 dB
1 dB Compression	+27 dB
Switching Speed	4 nsec TYP
Package	8 Lead SOIC (3.8mm x 4.8mm)

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INFO/CARD 16

Bell Atlantic to Expand Microcell Technology

— Bell Atlantic Mobile has announced that it will expand its use of microcellular technology to enhance cellular reception indoors in public locations. The first location will be Amtrak's Union Station in Washington, DC. Microcell technology utilizes the cellular concept on a smaller scale than the traditional car usage. It will use extremely low-power cell sites in heavily trafficked areas. A cell site, housing computer-controlled transmitters and receivers, is a key piece of the cellular system, serving as a gateway to the traditional telephone network. In Union Station, a suitcase-sized transmitter and antenna will be placed in the Amtrak passenger waiting area. The initial equipment was developed by NovAtel Communications, Inc. Regardless of what network equipment is in use at a given time, customers will be able to make calls on their cellular phones while within the microcell's coverage area. Initially, the microcell will not be connected to the outside "macrocellular" system. Later this year, the two will be interconnected and customers will be able to make and receive calls, as well as hand-off calls as they leave the station.

Tektronix Announces Equipment Grant

— Tektronix Inc. recently presented a grant to the University of Minnesota's Institute of Technology School totaling approximately \$130,000 in the form of state-of-the-art engineering equipment. The contribution is comprised of test and measurement equipment manufactured by Tektronix and includes such instruments as digital oscilloscopes, spectrum analyzers, digital voltmeters, function and pulse generators, and color plotters. The equipment is dedicated for use at the newly-constructed senior design lab at the school.

USSR Air Traffic Control Modernization

— AT&T, Westinghouse, IBM, C. Itoh and Daimler-Benz have committed themselves to an approximately \$10 billion modernization of the Soviet air traffic control system. The plan is slated for completion by the year 2005. The plan calls for merging of Soviet national airspace plans with air traffic control plans of the United States, Canada, Europe and Asia.

Researchers Develop New Class of Antenna

— Engineers at the Georgia Tech Research Institute have

developed a new class of microstrip antenna that combines the broad-band performance typical of spiral and sinusoidal antennas with the surface mount capabilities, efficiency and low cost of microstrip antennas. The new antenna design was developed through a research program supported by the U.S. Air Force's Wright Research & Development Center. Though designed for mili-

tary use, the new antenna could have widespread commercial applications such as cellular telephones, identification, satellite communications, inter and intra-building communications and the tracking of ground shipments. The antenna uses existing technology and therefore could be mass-produced at a low-cost.

The antenna was developed using



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Why Not Consider The New ATC STA-TUNE™ Adjustable Multilayer Capacitor?

STA-TUNE™ Capacitors, which brought you all the advantages of trimmers and multilayer capacitors in one 1/8 inch surface mount package, are now available in kit form. STA-TUNE is built with an external electrode employing a unique four-pad design, which allows incremental capacitance increases. In-circuit value adjustment is accomplished by connecting one of six (6) different capacitance combinations. The Kit includes a conductive epoxy pen for attachment. Kit Number ATC ST2001 is available from stock through ATC's QUIK-PICK 48 HOUR SYSTEM™. ATC ST2001 contains 3 pieces each of 22 capacitance values from 1.0 to 68 pF.

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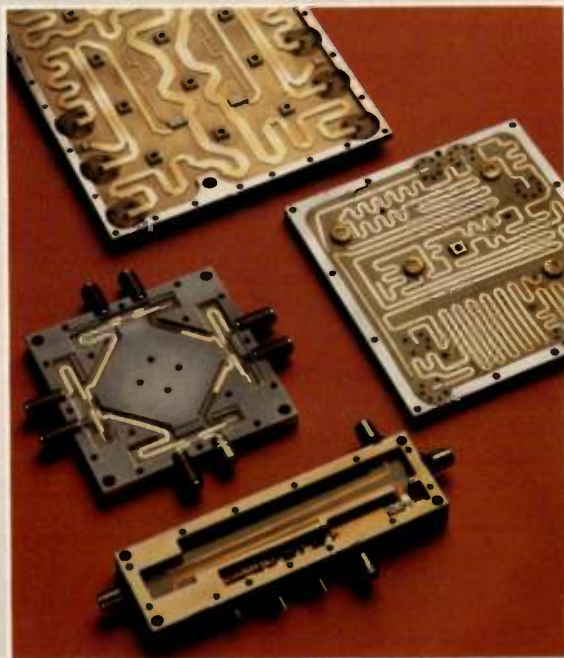
INFO/CARD 17

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- Customized and Multi-Function Designs
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21 Continental Blvd., Merrimack, NH 03054
Tel: 603-424-4111 Fax: 603-424-6580

two different antenna technologies: the broad-band spiral technology and the microstrip technology. While most microstrip antennas offer a bandwidth of just 10 percent, the new design offers bandwidths as high as 500 to 600 percent.

Direct Conversion Technique Relocates — Direct Conversion Technique (DCT) recently announced the opening of its new headquarters facility in Des Plaines, Illinois. Their new address is 1245 Forest Avenue, #12, Des Plaines, IL 60016. Tel: (708) 827-2282. Fax: (708) 827-2280.

MIC Technology Acquires Thin Film Coating and Patterning Business — MIC Technology has acquired the U.S. portion of the thin film coating and patterning business of Materials Research Corporation's Hybrid Products Division. MIC will operate this business as the Thin Film Division of MIC Technology. Materials Research Corporation will retain its European based Hybrid Products Division and continue to service the European market.

EEsof Users' Group to Meet at 1991 IEEE MTT Symposium — EEsof Inc. will sponsor the MTT-S meeting of its Users' Group, June 11 from 5:00 p.m. to 7:30 p.m. in room 311 at the Hynes Convention Center, Boston, MA. The EEsof Users' Group invites both members and non-members to attend the meeting. Technical papers will be presented investigating the use of CAE/CAD software applications in microwave and RF circuitry design. Proceedings will be available at the meeting.

Magnavox to Provide Pilot Automatic Vehicle Location System — Magnavox/Nav-Com has won a contract to provide the Swedish Road Administration with a pilot Automatic Vehicle Location System (AVLS), which will be used to track the location and movements of vehicles at a central facility. The vehicle navigation system uses a small multi-channel GPS receiver, which determines the vehicle's location from signals broadcast from the satellite based Global Positioning System. The Swedish Road Administration will provide the data link for the vehicles, based on the established Mobitex radio data communications network. The vehicles will be equipped with Volvo VTS

MC100 data terminals, which will be interfaced with the Magnavox navigation system, and will transmit AVLS messages to the central facility. There, using Magnavox-developed software, the vehicle positions are displayed on a 19 inch color graphics monitor.

IITS Purchases Dash, Straus & Goodhue — Inchcape Inspection and

Testing Services recently announced the purchase of Dash, Straus & Goodhue (DS&G). DS&G is involved with compliance testing for electrical and electronic products, particularly in the area of electromagnetic compatibility, including testing for electromagnetic interference and radio frequency interference. DS&G will become a member of IITS's Manufactured Goods Testing Group.



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Microwave Hybrid Circuits Call For Papers

— A call for papers has been issued for the 1991 Microwave Hybrid Circuits Conference. The conference will be held October 13-16 in Sedona, Arizona. The conference theme is "Designing for Production — Issues in the Transition from the Drawing Board to the Factory Floor." Papers describing new work will be considered on the

following subjects: innovative MIC and MMIC design; MIC processing and quality issues; advancement in MIC materials; MIC materials testing; MMIC packaging techniques; unsolved problems in MIC design or implementation; and Taguchi methods and design centering techniques. Abstracts are due June 1, 1991. Abstracts should be sent to: MHCC, c/o Toshikazu Tsukii, Pro-

gram Chairman Raytheon ESD, 6380 Hollister Avenue, Goleta, CA 93117. Tel: (805) 967-5511. Fax: (805) 964-0470.

New Report on Future of Broadcast Transmission Systems

— A new report issued by Elsevier Advanced Technology discusses the future of broadcast transmission systems to the year 1995 and beyond. The report, "The World Market for Broadcast Transmission Systems to 1995," discusses the effects of slow economic growth on the market, the changing political climate, increased competition and long term trends. Areas of special interest include the continuing expansion of alternative broadcasting technologies such as satellite and cable used for broadcasting transmission. The report predicts a doubling of the number of households who receive satellite or cable within a few years. Direct broadcasting by satellite is also expected to increase rapidly. The falling demand for LF and HF transmissions is also discussed and the subsequent changes in companies who will serve that market. On the brighter side, the growth of FM and TV transmission systems are expected to show significant growth especially in less developed countries. Problems are expected though in areas such as southern Europe where stricter spectrum management policies will be put into place.

Federal Rules on Antennas Should Supersede Local Rules

— The NAB has called for broader federal authority over local zoning restrictions on satellite dishes and other types of radio and TV antennas. The NAB made these comments to the Federal Communications Commission because many of the nation's municipal rules on radio and TV antennas are "undefined and unreasonable." These requests were made as a result of a Deerfield, New York zoning ordinance on satellite dishes.

Arlon Announces Plant Expansion

— Arlon, Microwave Materials Division has announced a 75,000 square foot building addition and equipment acquisition. The new facility will contain production, quality control, and testing laboratories for the company's PTFE laminated substrate materials. The addition is expected to be completed in February 1992 and will complement the company's existing 60,000 square foot facility.

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RF Technologies Announces Opening

— A new company has opened who will provide high power and super high power waveguide and coax networks, as well as transmission systems and components. The company, RF Technologies Corporation, will also offer a line of RF instrumentation and control equipment. Their address is 482 Congress Street, Suite 101, Portland, Maine 04101. Tel: (207) 773-7778 or (800) 634-4075.

Motorola and Northern Telecom to Develop Open Cellular System Standards

— Motorola and Northern Telecom recently announced an agreement to develop major enhancements to their existing cellular products for global markets, based on new open system standards. The two companies intend to jointly specify, develop and implement, in connection with the appropriate standards bodies, open cellular system interface standards which will permit operators to purchase cellular system components from a number of vendors. The standards will be based

on existing international switching and cellular standards, such as CCS7, GSM, ISO and ISDN.

New Cellular Network Antenna

— CAL Corporation and Bell Cellular recently introduced a new cellular antenna to improve network performance for cellular users. The low profile sector-beam cellular antenna was developed through a joint venture. The antenna tilt, used to adjust antenna frequency and scope, can be controlled remotely and adjustments no longer need to be done manually.

Antenna Specialists Relocates

— Antenna Specialists Company have announced the moving of its Site Management Products Research and Development Group to new facilities in Reno, Nevada. The new facility, with 50 percent more space, will include additional test equipment and environmental test chambers.

Varian Completes Restructuring

— Varian Associates recently com-

pleted the initial phase of its financial turnaround with the sale of its last remaining non-core business, a broadcast transmitter operation in England, to the Harris Corporation. The sale of this unit and four other non-core businesses and six product lines have generated combined proceeds in excess of \$60 million. The funds will be used for various purposes, including the second phase of Varian's turnaround which involves achieving operational excellence in each of its core businesses — electron devices, analytical instruments, medical equipment and semiconductor equipment.

S.T. Research Receives HPI Receiver System Contract

— S.T. research has been awarded a contract for the development and product of the High Probability of Intercept (HPI) Receiver System from the Naval Sea Systems Command. The contract, depending upon options exercised and ancillary system and support, will exceed \$50 million.

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A66GA	2	1-500	1.5:1	.7	20	±.25		
		2.5-400	1.1:1	.5	40	±.15		
A66L	2	.3-100	1.5:1	.5	35	±.2		
		1-50	1.1:1	.2	40	±.06		
A66U	2	5-1000	1.2:1	1.0	30	±.3		
A67	4	1-500	1.5:1	1.0	20	±.25		
		2.5-300	1.2:1	.5	30	±.1		

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HDTV: The World is Waiting

By Gary A. Breed
Editor

High definition television (HDTV) has clearly reached critical mass in its development. Regular developmental broadcasts are taking place in Europe and Japan, studio video equipment is readily available (although expensive), and component manufacturers are actively developing products for HDTV receivers and monitors.

The components industry is nearly ready for high-volume HDTV manufacturing, although production will wait until the market requires it. Thomson-ICS, Sawtek, and other SAW device houses have announced HDTV IF bandpass filters. Video amplifier and switching components from Maxim, National Semiconductor, Comlinear, Elantec, Burr-Brown and Analog Devices are capable of handling the required bandwidths. Direct satellite broadcast of HDTV is attractive to GaAs front-end suppliers such as Pacific Monolithics, Triquint and Anadigics.

Also required are high performance analog and digital signal processing components. A-to-D and D-to-A converters with at least 10-bit resolution must accurately handle greater than 25 MHz video bandwidth. Digital signal processing circuits have the same constraints. To put the required performance in perspective, current NTSC studio video requires 6 to 10 MHz bandwidths, depending on the ultimate quality level desired, while HDTV requires 25 to 50 MHz.

Studio equipment is already in place, and manufacturers are clamoring to supply networks, stations and video production houses with new systems. Test equipment, cameras, recorders, graphics and special effects, signal routing equipment, signal processing systems, and monitor makers are waiting for the doors to open on a huge market. However, the list of current suppliers is a Who's Who of Japanese electronics. NHK, the Japanese broadcasting agency, sponsored an HDTV "open house" at the recent National Association of Broadcasters convention, and the list of participating companies included 20 major firms such as Sony, Canon, NEC, Panasonic, and Toshiba.



Nikon offers HDTV imaging for microscopic observation.

Non-broadcast applications are helping keep enthusiasm high while broadcast standards evolve. Medical imaging, industrial monitoring and inspection, photo archives, and military display systems are but a few of the more promising applications. For example, Nikon is promoting high resolution imaging systems for microscopy in medical, metallurgical and microelectronic applications. Although alternative high resolution display technologies are available now, many potential applications will not become widespread until the large quantity manufacturing for consumer HDTV receivers brings prices down.

Terrestrial Broadcast

The selection of a method to bring HDTV to homes in the U.S. is the single major problem area at this time. FCC testing of the five applicants for consideration as the broadcast standard has been delayed again, this time due to delivery problems with a standards converter being built by Tektronix. Final development of this complex unit has required hand-picked A to D converters because standard components cause visible defects in the picture.

The goal of the FCC is to complete on-the-air testing by Spring 1992, in order to have evaluation ready for

Commission consideration by September of that year. Selection of a standard is currently planned for early 1993. Many industry and government observers believe that this is the latest possible date that is acceptable, and any further delay will certainly jeopardize the U.S. market position in the world HDTV market.

One industry participant, choosing to remain anonymous, muses, "Why are terrestrial standards holding things up when DBS and fiber are perfectly capable of delivering HDTV?" The strength of the broadcast lobby is one part of the answer to his question, but the situation is not quite that simple. First, neither fiber nor DBS is in widespread use in North America, and the time required to bring those technologies to "universal" status is unknown, perhaps longer than that required for a changeover to a broadcast system that at least approaches the quality available with HDTV.

With HDTV VCRs already developed, and given the current popularity of rented video programming, it will be extremely interesting to see how the average consumer reacts to broadcast television that has somewhat lesser picture quality than taped programming. Currently, the opposite is true.

Which leads us to the biggest question — will the consumer accept HDTV at all? The quality difference is analogous to that of the compact audio disc versus vinyl records, and CDs are enormously successful. However, the improvement offered by stereo AM broadcasting and quadraphonic FM has been generally ignored. Even stereo television sound has been under-utilized.

Will reruns of Gilligan's Island become more interesting on HDTV, or will program producers react to the quality of the medium and produce higher quality programs? The answer to this question is probably the key to HDTV's future, not the technical standards or hardware capabilities.

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INFO/CARD 22

WRH

A Channel Sounder For Indoor Communications Systems

By Fernando Casadevall and Xavier Barba
Department of Signal Theory and Communications
Catalunya Polytechnic University

From the beginning of the 80s the possibility of giving total mobility to the increasing number of communications terminal equipment in large buildings like factories, hospitals, warehouses, commercial centers etc., has emerged strongly due to the important advantages it brings. Even though an infrared optical system could be considered for achieving such an objective, it becomes inoperative when obstructed. For this reason, radio systems must be considered.

In radio communication systems, the multiple reflections of the transmitted signal from the structure of the building and surrounding inventory create multipath interference. Multipath has historically been identified as the most important factor limiting mobile and portable radio communication systems. For narrow-band systems (where the baseband digital symbol duration is several times greater than the extent of multipath-induced propagation delays, that is, flat fading conditions), multipath causes large fluctuations (fading) in the received signal voltage due to the changing phasor sum of the signal components arriving at the receiver antenna via different paths. Temporal variations of the channel as well as changing multipath geometry seen by a mobile user are the mechanisms for the fading. Therefore, an accurate characterization of the operating channel is a mandatory prerequisite for the development of reliable indoor radio systems.

Usually we could distinguish between a large scale model, used in determining coverage areas and co-channel interference levels, and a small scale model that describes the fading due to temporal variations of the received signal and used in determining the bit error rate and outage probability.

The expression that characterizes the large scale model is the Path Loss (L) given by $L = Kd^n$ where K is an appropriate constant, d is the distance between

transmitter and receiver in meters and n is the mean path loss exponent. For stationary transmitters and receivers, the small scale model is usually characterized by a Rician distribution. In this case the most relevant parameter is the ratio of the received power of the specular radio signal to the average power of the scattered signal.

Measurements reported in References 1 and 2, indicate that n usually ranges between 2 to 6 and K between 1 to 16 dB depending on the building structure. So, there is a great value dispersion and it could be convenient to characterize the operating channel in each case.

In this paper we present a simple channel sounder appropriate for characterizing narrow-band indoor communications channels. The first part is devoted to analyzing the transmitter structure, the second studies the receiver system and the last presents some practical results and conclusions.

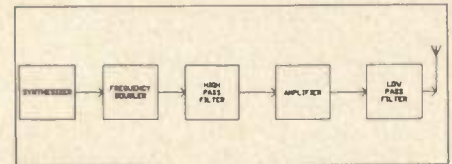


Figure 1. Schematic drawing of the CW transmitter.

Transmitter

The schematic drawing of the CW transmitter is shown in Figure 1. We can distinguish five parts. The synthesizer produces a sine signal whose carrier frequency ranges between 800-850 MHz. The frequency doubler changes the carrier frequency up to the 1.6-1.7 GHz band. This signal is not synthesized directly in the 1.6-1.7 GHz area because the voltage controlled oscillator (VCO) and counters (Prescalers) are cheaper and more readily available in the 800 MHz band than in the 1.6 GHz band. The high pass filter eliminates the

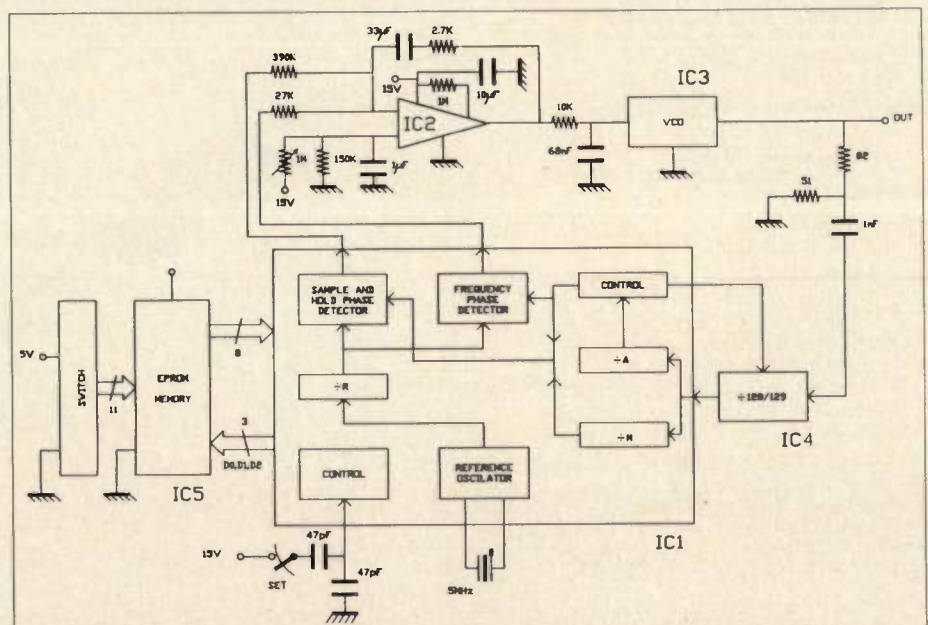


Figure 2. Schematic diagram of the synthesizer.

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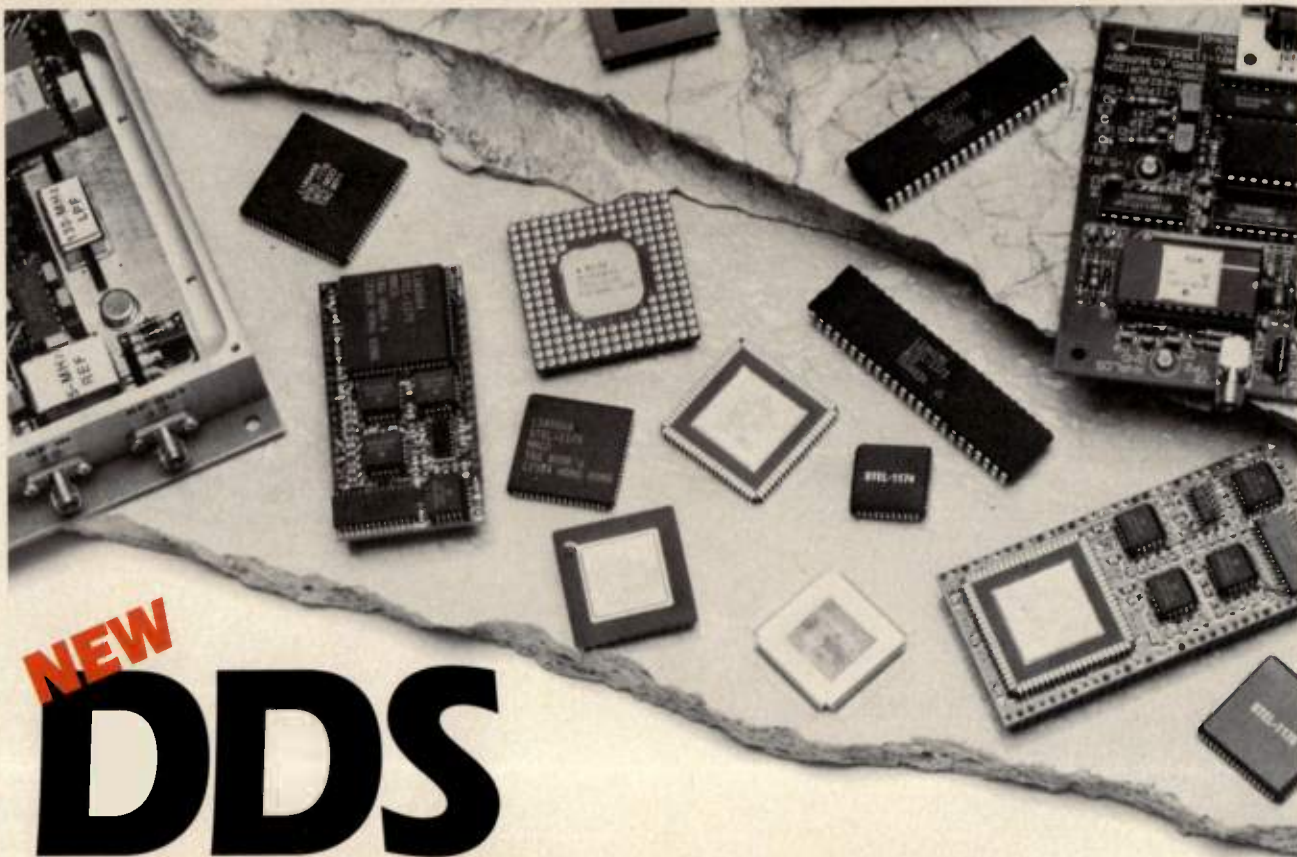
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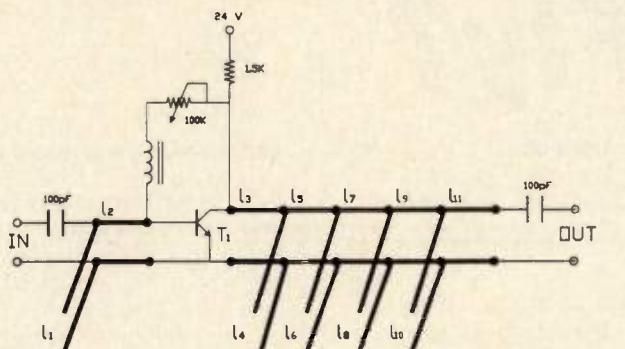
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 l_3 - 5 mm, 50 Ω l_7 - 48 mm, 115 Ω l_{11} - 48 mm, 115 Ω
 l_4 - 12 mm, 50 Ω l_8 - 48 mm, 45 Ω T_1 - HXTR3615
 Epoxy Substrate ($E_r = 4.75$, thickness 0.75 mm)

Figure 3. Frequency doubler.

spurious signals around 800 MHz that appear at the frequency doubler output. The output amplifier increases the useful signal level up to 20 dBm. Finally, the low pass filter attenuates the level of the harmonics present in the output signal.

In the paragraphs below we describe the most relevant aspects of each part.

Synthesizer - The synthesizer is built around the integrated circuit NJ8820 from Plessey. This circuit contains a reference oscillator, an 11 bit programmable reference divider, two phase comparators: one digital and one sample and hold type, two programmable dividers: one, called A, that counts down to zero from the programmed input, with a 7 bit length, and the other, called M, that counts up from zero to the programmed input and with a 10 bit length, and finally the necessary control and latch circuitry for accepting and latching the input data. By means of an external dual modulus counter (N/N+1), the synthesizer is able to program a total division ratio given by NM+A, where A is between 0 and 127, M is always greater than or equal to A and the minimum division ratio is $N^2 - N$. However, the value of N must also be large

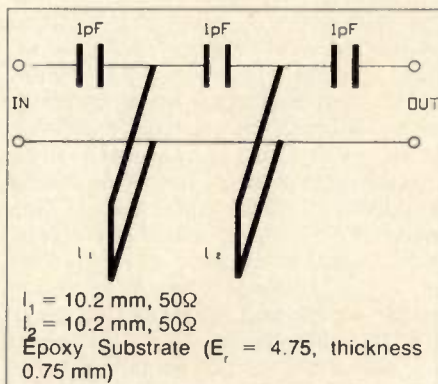
enough to guarantee that the output frequency from the dual modulus counter not exceed the maximum input frequency of the integrated circuit.

The frequency synthesizer is completed by an active loop filter and a VCO, as shown in Figure 2.

For each synthesized frequency, the NJ8820 circuit reads eight 4 bit words from an external memory. The timing signals needed are provided by the same integrated circuit. With this structure the synthesizer can produce a carrier frequency that ranges from 800 MHz to 850 MHz, with a frequency step of 25 KHz and whose level is around -5 dBm.

Frequency Doubler - The schematic of the frequency doubler is shown in Figure 3. It uses an HXTR3615 transistor as active device. The input matching network is composed of an open circuit microstrip stub and a transmission line, whereas the output matching network uses two open circuit microstrip stubs with two sections of transmission line between them (3). In this case the stub lengths are designed in order to present a 1/4 length at the second and the third harmonic of the output signal, respectively. In this way the network not only matches the output impedance but also introduces a selective effect to the undesired signals. A band stop filter, tuned at 800 MHz, is placed behind the output matching network. This filter is built using two open circuit microstrip stubs and three transmission line sections. Obviously this filter also eliminates the undesired signals placed around the third, fifth and seventh harmonics of the input signal.

It is important to emphasize that this circuit, and everything described below, have been designed using computer aided design (CAD) techniques. In particular, we used the SuperCompact



l_1 = 10.2 mm, 50 Ω
 l_2 = 10.2 mm, 50 Ω
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Figure 4. High pass filter.

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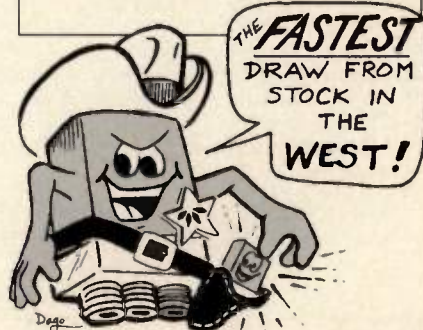
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Disk RFD-0391: March 1991

This disk contains "NOVA" by Robert Stanton, a nodal circuit analysis program for AC, time domain, and S-parameter analysis. Numerous models are included, and operating instructions are included in a READ ME file. This is offered by the author as shareware; a co-processor version, additional documentation, and future updates are provided to users who wish to pay for registration.

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program for this purpose.

High Pass Filter - In order to attenuate the 800 MHz spurious signal present at the output of the frequency doubler, we have placed a high pass filter before the amplifier. This filter has been designed using a fifth order Butterworth structure (4) and made using transmission lines and short-circuit stubs as shown in Figure 4. The 3 dB cut-off frequency is 1.56 GHz and the insertion loss is 1.34 dB. This structure has a theoretical attenuation value of 50 dB at the 800 MHz band and it guarantees sufficient attenuation of the spurious signals.

Amplifier - The signal level at the output of the frequency doubler is -5 dBm. Therefore, in order to obtain an useful signal of 20 dBm at the transmitter output, the amplifier gain must be approximately 25 dB. This amplifier has three stages with power gains of 10, 10 and 5 dB respectively. The first two stages make use of a HXTR3102 transistor as an active device whereas the output stage uses a BFG34 transistor able to produce the required output level. The matching network between stages is built by means of transmission lines and microstrip stubs in open and short circuit.

In Figure 5 we show the schematic diagram the amplifier. The last two microstrip stubs, placed at the output of the amplifier, are designed in order to attenuate the second and third harmonic respectively. It is important to emphasize that we have chosen for the second stage a short-circuit microstrip stub to diminish the low frequency gain of the amplifier and consequently to increase its stability.

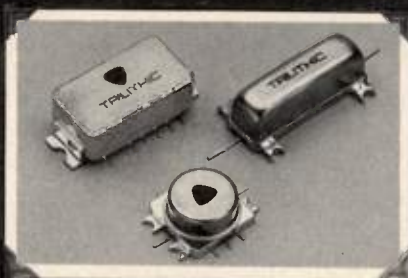
Low Pass Filter - The low pass filter is devoted to attenuation, as much as possible, of the harmonics and other spurious signals present at the output of the amplifier. For this purpose we have chosen a seventh order Butterworth filter, built again using transmission lines and open-circuit microstrip stubs. In Figure 6 shows a schematic diagram of the filter. The 3 dB cut-off frequency is 1.9 GHz and the insertion loss is 0.7 dB. This structure guarantees a theoretical attenuation value of 46 dB at 3.2 GHz (second harmonic of the output signal) and more than 70 dB at 4.8 GHz (third harmonic). These values are sufficient to obtain an output rejection to harmonics and spurious signals greater than 60 dB with respect to the

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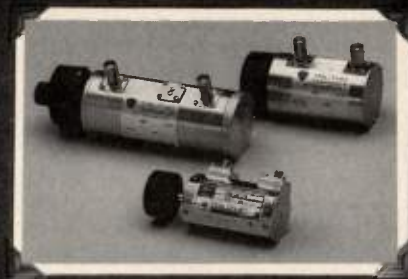
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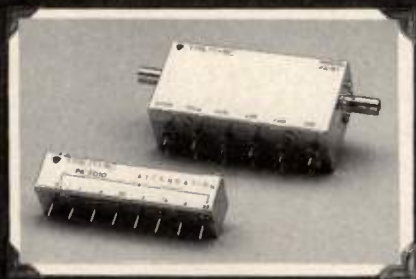
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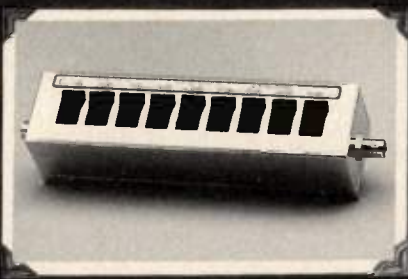
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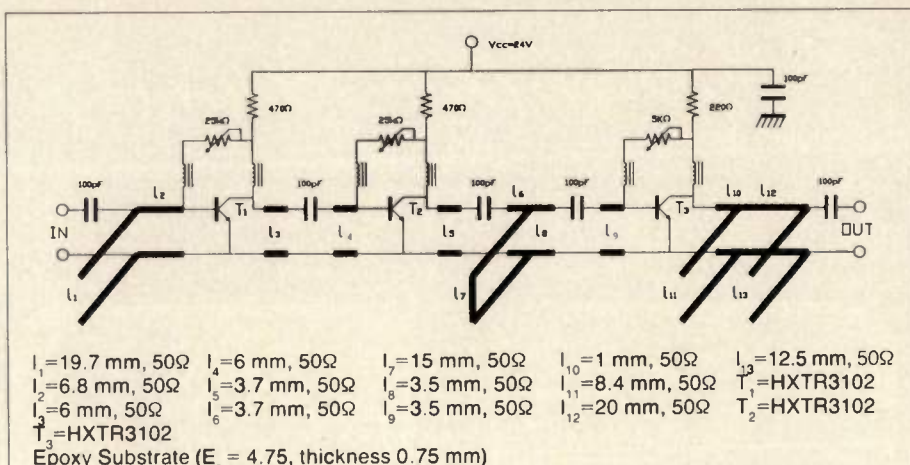


Figure 5. RF amplifier.

useful signal.

Transmitter Performance - Once the transmitter was made, a complete set of measurements was taken for its characterization. The most relevant features are described below:

- Output level: 18.5 dBm - 0.5 dB
- Frequency Range: 1.6 GHz to 1.7 GHz, with 50 kHz steps
- Spurious and harmonic rejection: Greater than 60 dBC
- SWR: 1.5 at 1.65 GHz, <1.85 between 1.6-1.7 GHz

Receiver

The schematic drawing of the CW receiver is shown in Figure 7. We can distinguish six different parts: an RF amplifier that increases the level of the weak signal present at the antenna output; an RF attenuator that generates an appropriate level at the input of the mixer; the mentioned mixer that produces a frequency translation of the received signal to a new carrier frequency of 70 MHz; the local oscillator associated with the mixer; an IF amplifier, and finally a logarithmic amplifier that produces an output signal proportional to the received power at the

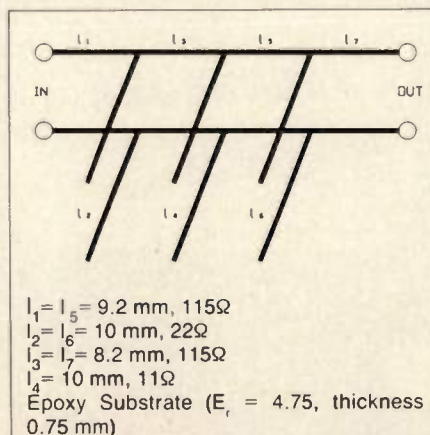


Figure 6. Low pass filter.

antenna port. This signal is delivered to a data acquisition card, controlled by a personal computer (PC), in order to be processed properly.

As with the transmitter, all the above mentioned subsets of the receiver have been designed using CAD techniques.

RF Amplifier - The schematic of the RF amplifier is shown in Figure 8. It is composed of two amplifier stages and a high pass filter. The first amplifier stage is designed to optimize the noise figure whereas the linearity aspects are emphasized in the second amplifier. Both use common emitter structure with transmission lines and open circuit microstrip stubs for the matching circuits. The transistor HXTR3615 from Hewlett-Packard is used as an active device. The power gains are 12.5 dB for the first stage and 13.5 dB for the second, and the 3 dB bands are 1.5-2 GHz and 1.5-1.9 GHz respectively. A high pass filter is placed between both stages which eliminates the noise and all the spurious signals below 1 GHz that could appear in the antenna port. The cut-off frequency is 1.3 GHz and the insertion loss is 0.3 dB. This high pass filter is placed after the first amplifier in order to optimize the noise figure of the receiver.

RF Attenuator - A simple attenuator can be made using a resistive pi structure. If the resistor values are controlled properly this circuit is capable of giving variable attenuation over a wide bandwidth and provide matched impedances in both ports. On the other hand, for

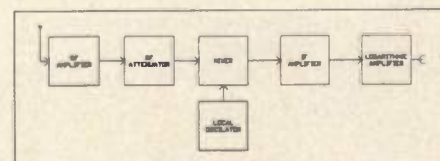


Figure 7. Schematic drawing of the CW receiver.

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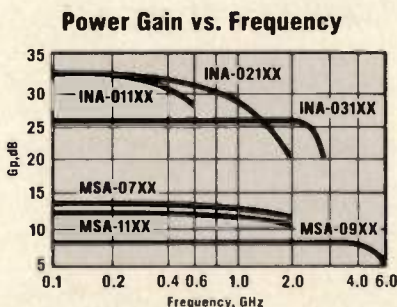
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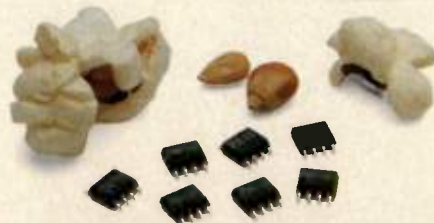
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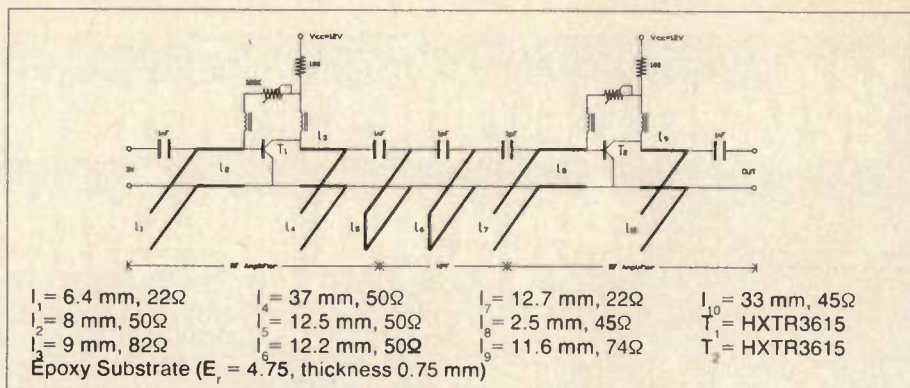


Figure 8. Receiver part: RF amplifier.

small signals at RF frequencies, the PIN diode, when it is conducting, appears essentially as a pure resistor whose value is roughly inversely proportional to the bias current. Moreover, the quality of this resistor, i.e. its linearity and lack of distortion products, can be very good if the diode is used under the right conditions. So, for these reasons the RF attenuator has been made using a PIN diode attenuator, as shown in Figure 9, where the series and parallel bias currents are properly controlled by means of a single control voltage (5). The insertion losses are 1.5 dB, the attenuation reaches up to 18 dB and the SWR ratio is lower than 2 dB in the full range of the attenuation values.

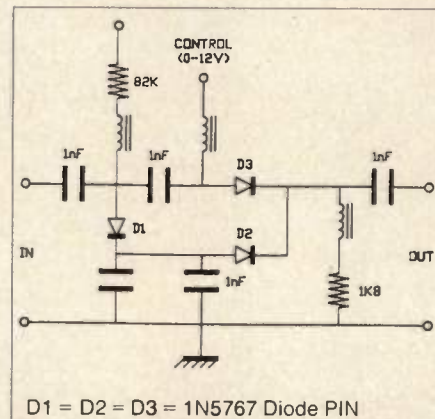


Figure 9. Receiver part: RF attenuator.

Mixer - This circuit is built around the SRA-2000 double balanced mixer from Mini-Circuits. In order to guarantee that the RF port and the IF port have 50 ohms at any frequency, special matching networks have been used (6), as shown in Figure 10. The input network is composed of an impedance transformer, made by means of a transmission line, and a short circuit microstrip stub in series with a 50 ohm resistor. At

the working frequency the stub appears as an open circuit. Then, the impedance transformer changes the output impedance of the RF amplifier (50 ohms) to the nominal mixer impedance of 200 ohms. At lower frequencies the stub appears as a very small inductive impedance. A 50 ohm resistor is placed at the input of the impedance transformer regardless of the source impedance value. Really, at these frequencies, the 10 pF capaci-

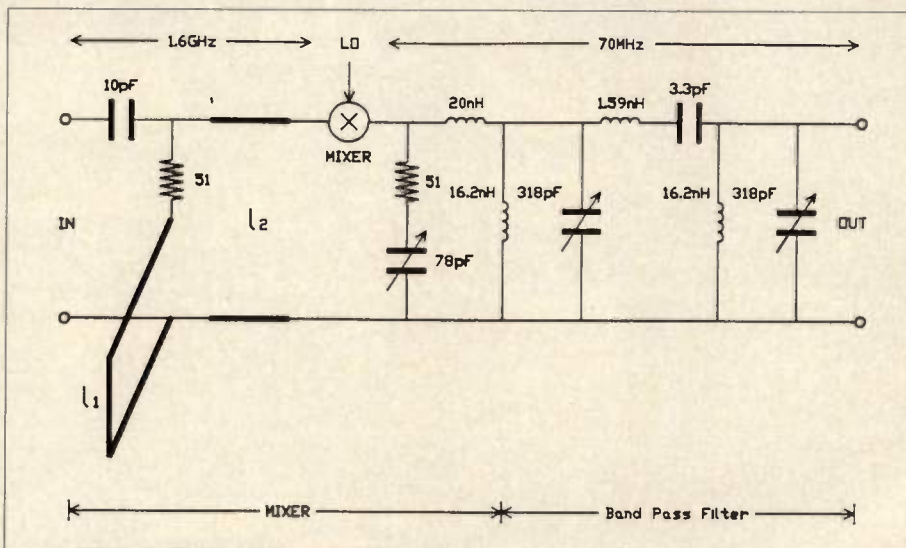


Figure 10. Receiver part: mixer.

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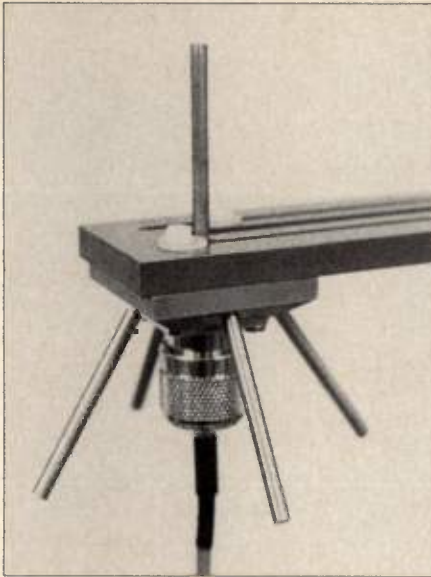


Figure 14. Photograph of the antenna.

is around 12 dB whereas for the SL623B it is 24 dB. The former provides a logarithmic characteristic over a 10 dB range and the later over 20 dB. As the integrated circuits provide a current whose level is proportional to the logarithm of input signal (with maximum level equal to 2.1 mA), the common base transistor and the operational amplifier

act as a current-voltage transformer with appropriate gain and DC offset.

The input signal must range between $0.2 \text{ mV}_{\text{RMS}}$ and 1 V_{RMS} for an accurate behavior of the logarithmic amplifier, but it could withstand input voltages up to $1.8 \text{ V}_{\text{RMS}}$ without damage. Within the input range, the dynamic error of the logarithmic characteristic is lower than 0.5 dB and the DC output signal ranges between 0 to 5 Volts.

Receiver Performance - Once the receiver was built, we obtained a complete set of measured characteristics, whose most relevant aspects are described below:

- Sensitivity: -97 dBm
- Maximum input level: 0 dBm
- IF Bandwidth: 10 MHz
- Logarithmic amplifier output: $0\text{-}5 \text{ Volts}$
- CAG: 18 dB manually controlled
- SWR: 1.9 at 1.65 GHz , <2.2 between $1.6\text{-}1.7 \text{ GHz}$

Antennas

In order to complete the channel sounder, two monopole antennas with ground planes were constructed. We used a $1/4$ monopole with four bars inclined 45 degrees to simulate the ground plane. The measured SWR at the antenna port is lower than 1.2 in the

$1.6\text{-}1.7 \text{ GHz}$ working band. A photograph of the antenna is shown in Figure 14.

Preliminary Measurements and Conclusion

Using the above described channel sounder, some preliminary measurements of the propagation conditions inside buildings were done. We recorded the received signal during 20 minute periods and using 20 samples/sec. In Figure 15 a typical example of the recorded measurements of the received signal amplitude in two standard cases is shown. The first case is line of sight between the transmitter and receiver and the second when obstructed. In both cases we can see large fluctuations (fading) in the received signal voltage. As was explained in the introduction, these fluctuations are due to the changing phasor sum of the signal components arriving at the receiver antenna via different paths. Presently a complete set of measurements is in process in order to characterize the transition channel in different buildings constructed using different materials.

In conclusion, a simple narrow-band channel sounder, appropriate for characterizing narrow band indoor communications channels, has been described.

-150 dBc Phase Noise...

Acknowledgements

This work has been supported by CICYT (Spain) under Grant TIC880543.

RF

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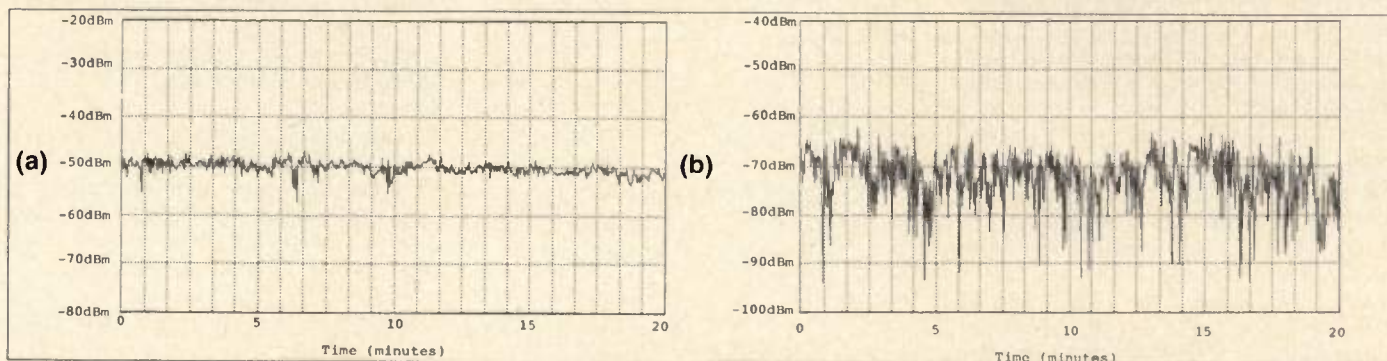


Figure 15. Signal level at the antenna port versus the time, (a) line of sight, (b) obstructed.

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This paper describes the RF portion of the control circuitry associated with the L-Band High Power Amplifier designed for such service by Spar Aerospace of Montreal.

Each amplifier has a 27 dB bi-directional coupler at its output to allow measurement of transmitted and reflected signal power levels. The resulting detector signals are used to limit output power, hence power consump-

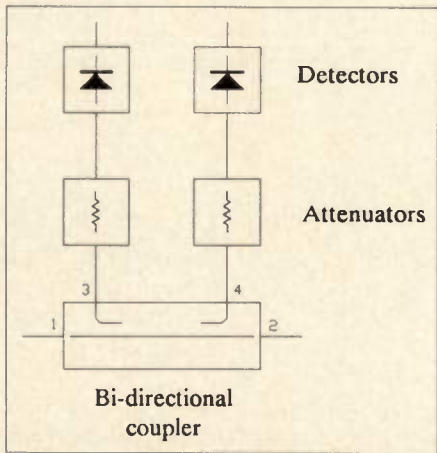


Figure 1. Block diagram.

tion without degrading linearity, and to protect the amplifier against output mismatch. In order to obtain sufficient accuracy, high directivity in the coupler is essential. A block diagram is shown in Figure 1.

The design of the coupler is presented together with a description of the method by which high directivity is obtained. Artwork and measured results are presented including results from high power testing in vacuum. The objective is to design, build and test a coupler with the specifications of Table 1.

Choice of Configuration

In order to be compatible with the rest of the circuitry making up the amplifier, the coupler must be fabricated using microstrip techniques on metallized alu-

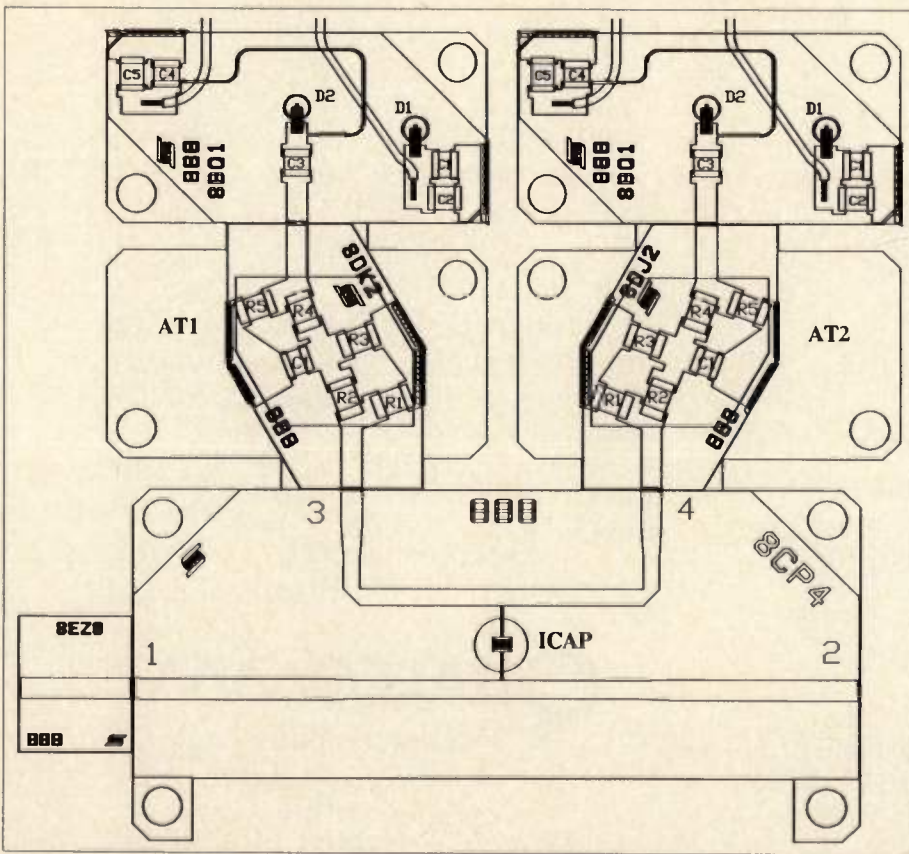


Figure 2. Artwork.

Operating Frequency	1 to 2 GHz
Low Insertion Loss (to minimize impact on amplifier efficiency)	≤0.15 dB
Low Coupling Value (to minimize insertion loss)	≤-25 dB
High Directivity (to differentiate forward & reflected signals)	≥25 dB
High Isolation (to achieve high direc- tivity with low coupling)	≥50 dB
Return Loss (to guarantee the direc- tivity)	≥25 dB
Must sustain high CW power and be tempera- ture insensitive	≥100W -60°C +120°C

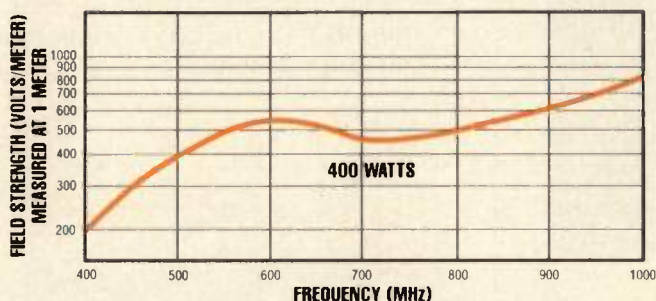
NOTE: Must also be bi-directional, symmetrical and easy to fabricate (preferably not to require tuning).

Table 1. Design specifications.



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mina substrate (0.050" thick, $\epsilon_r = 9.9$).

The problem of designing a high directivity coupler in this medium, (compared to an ideal TEM design), is due to the inhomogeneity of the dielectric structure. The isolation of the coupler is the most difficult characteristic to control, requiring wave cancellation and equal phase velocities for the even and odd modes of propagation.

The quarter-wave edge coupled design approach is chosen for its symmetry and simplicity, however the non-TEM propagation must be compensated.

Design Details

In order to solve the problem of non-TEM propagation, various techniques have been proposed. References 1 and 2 describe the Wiggly

coupled line but the available design information and mathematical relationships are not well suited for loose coupling. The electromagnetic field analysis is quite complex and none of the known CAD tools are helpful in this regard. References 3-5 describe the dielectric overlay coupler, but show poor repeatability due to the overlay boundary's position affecting matching, coupling and isolation. This solution is also somewhat cumbersome to produce and may be sensitive to high temperatures and high power levels. A configuration that is simple, repeatable and requires no tuning would be preferable.

References 6-10 present the design of an edge coupler with lumped capacitors at each end to compensate for the unequal phase velocities of even and odd modes at the isolated port. In the even mode, the capacitors have no effect, but they increase the odd mode phase length (transit time, ϵ_{reff} odd), thus equalizing the waves.

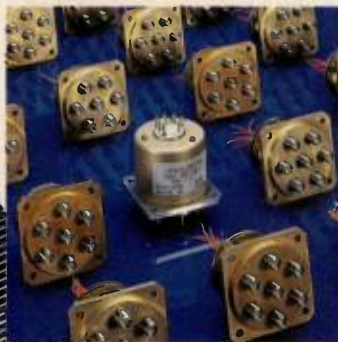
Due to the small value of the capacitance required, it can be realized as an interdigital structure (ICAP) formed on the same microstrip substrate as the coupler. From a mathematical approach, a capacitor at each end of the coupled

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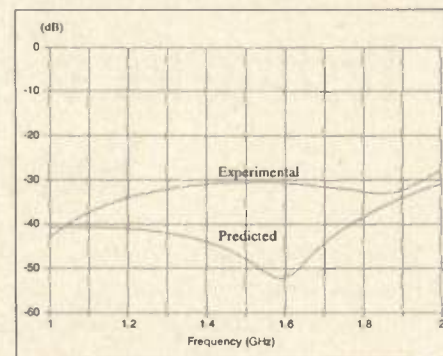


Figure 3. Predicted vs. experimental return loss (S_{11}).

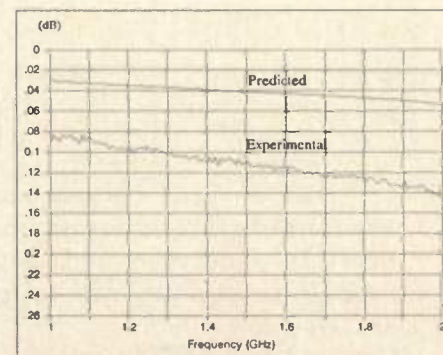


Figure 4. Predicted vs. experimental insertion loss (S_{21}).

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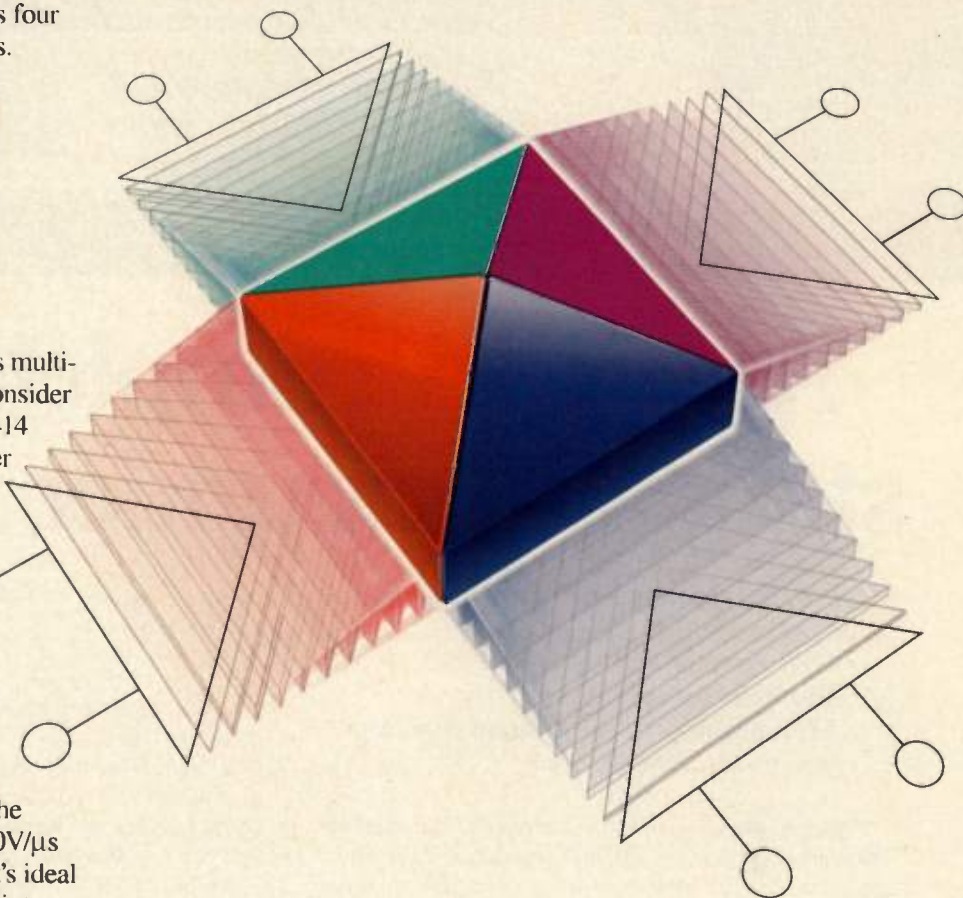
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INFO/CARD 41

line is used, (they must be separated by a quarter-wave length in order not to generate scattered waves to all ports). As mentioned in Reference 11, the capacitor at either end is satisfactory, but in order to have a good return loss, the capacitor should be located in the middle of the coupler to preserve the double symmetry of the circuit. The possibility of using just one capacitor located in the middle of the coupler was then investigated.

With this configuration, some characteristics tend to conflict and it is theoretically impossible to obtain both coupling flatness and optimum isolation. The selection of the capacitor requires evaluating the major performance parameters in terms of the intended service. In our case, the slope in frequency response of the coupled port can be conveniently equalized with a capacitor in the middle of the double- π attenuator. Note that if the ICAP varies, the slope does not change, only the level changes. The design therefore is made for maximum isolation and the capacitor value is calculated to be 0.11 pF. The model of the ICAP was derived from Reference 12. It is important to note that the ICAP has a significant shunt component that can affect the even mode. In order to reduce this undesired shunt capacitance, the ground plane below the ICAP is removed. Figure 2 details the artwork and layout of the coupler, attenuators and detectors. Note the ICAP and compensating capacitors (C1) in the attenuators (AT1 and AT2).

The coupling value was chosen for the best tradeoff between low insertion loss (loose coupling) and high directivity (tight coupling because of fixed isolation). Optimum power levels and slope for the Schottky diode detectors (D2) is then adjusted with the appropriate attenuator. The computed prediction and experimental results are shown in Figures 3-6.

Results

Due to the double symmetry of the coupler (verified experimentally), only four parameters are shown: the return loss (S_{11}) in Figure 3, the insertion loss (S_{21}) in Figure 4, the coupling (S_{31}) in Figure 5 and the isolation (S_{41}) in Figure 6. Comparing these experimental results and computer predictions, we can see some small explainable differences.

In Figure 3, we see that the predicted return loss is better than the one obtained experimentally. This is attributed to the connector-to-microstrip dis-

continuities.

In Figure 4, we see that the experimental insertion loss is more than expected. This is due to the coax-microstrip transition which was not included in the simulation.

In Figure 5, we see a small difference in the slope and level of the coupling. The slope difference is due to the presence of the capacitor between the

two coupled lines in the coupled region resulting in an incomplete confining of the fringing field through the dielectric. The level difference is due to the ICAP stray capacitance to ground.

In Figure 6, we see that the isolation is as expected, the small difference being due to a 0.1 pF over-etching of the ICAP. This could be fixed with a dielectric slab to increase the serial ICAP

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870	75 Ω	DC-1GHz	0-132dB	1dB
4440	50 Ω	DC-1.5GHz	0-130dB	10dB
4450	50 Ω	DC-1.5GHz	0-127dB	1dB
1/4450	50 Ω	DC-1GHz	0-16.5dB	1dB
4467	75 Ω	DC-1GHz	0-31dB	1dB
0/400	50 Ω	DC-500MHz	1-13dB	—
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For more information or to place an order call **Kay's Product Specialist at (201) 227-2000**.

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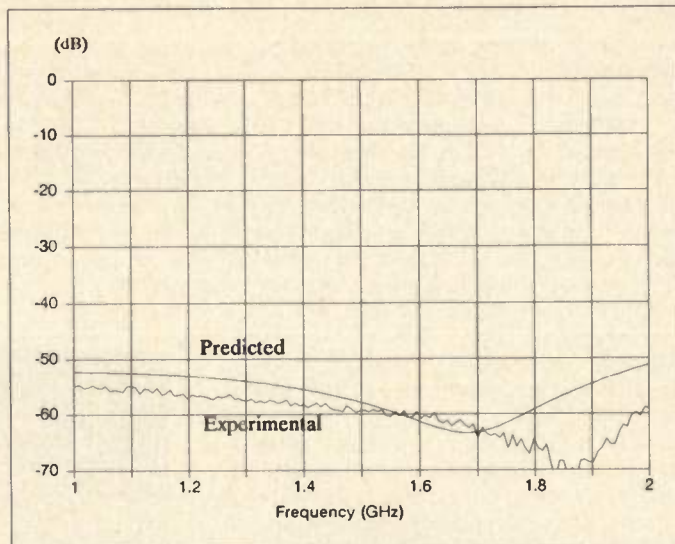
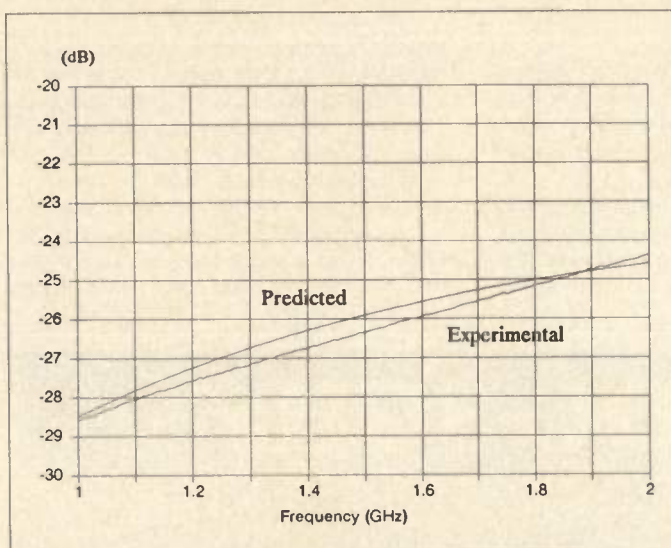


Figure 5. Predicted vs. experimental coupling (S_{31}).

Figure 6. Predicted vs. experimental isolation (S_{41}).

value (note that all results are without tuning).

space as an RF/Microwave engineer.

Acknowledgement

This paper was written while the author was employed by Spar Aero-

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About the Author

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$\div 16/17$  UPB553 UPB556 UPB555 UPB571	$\div 20/21$  UPB572	$\div 20/22$  UPB551 UPB552 UPB554	$\div 32/33$  UPB555 UPG569 UPB571	$\div 40/41$  UPB572
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John Fluke Mfg. Co. introduces the PM 5139 from Philips Test and Measurement, a 0.1 mHz to 20 MHz function generator combining ten general purpose test waveforms with arbitrary function capability. Sonar, mechanical testing, brain and heart waves, and baseband video or RF signals are just a few examples of arbitrary waveform applications. Using WaveCAD, a waveform generator and editing software package available from Fluke, the PM 5139's memory can be loaded with up to six arbitrary waveforms. Also, a real signal can be captured with a Philips digital storage oscilloscope, edited in software, and downloaded to the PM 5139.

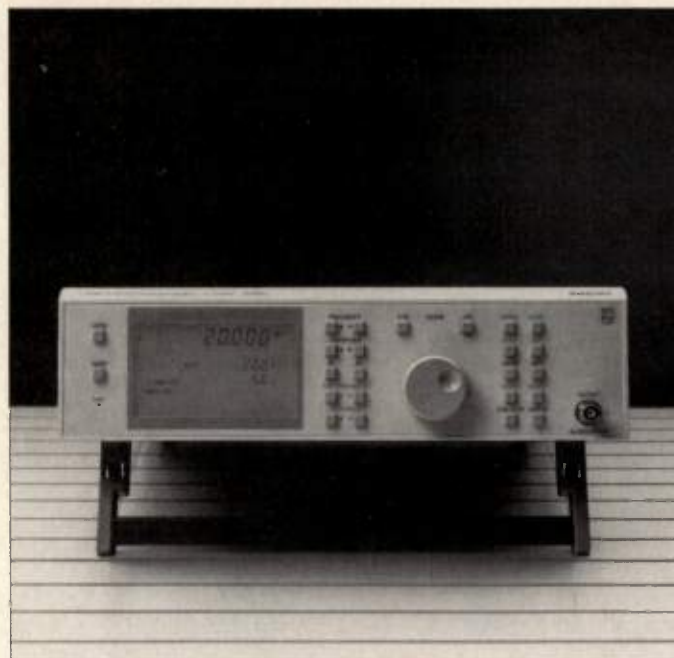
Standard waveforms include sine, triangular and square waves, positive and negative rectangular pulses, positive and negative sawtooth waves, haversine, half-period sine, and triangular pulses. Modulation modes avail-

able are: AM, FM, gating, bursts, PSK, and linear or logarithmic sweeps. Frequency accuracy is within ± 2 ppm over the entire range, with low jitter and phase noise, and very flat amplitude response. In the low frequency range, five of the standard waveforms can be varied in duty cycle from 1 to 99 percent with 0.1 percent accuracy. All manual settings can be performed remotely using an IEEE-488 bus interface.

The panel has a single rotary knob for numerical setting of parameters, with a logical array of parameter selection and control keys. Each key activates the rotary knob for dialing in the desired setting. A large central backlit LCD display shows all instrument settings at a glance.

The price of the PM 5139 is \$4190, or \$4690 with the IEEE-488/GPIB interface.

John Fluke Mfg. Co., Inc.
INFO/CARD #230



Six-Octave Bandpass Filter

Optoelectronics announces a new receiver bandpass filter that separates closely spaced radio signals. The Model APS-204 has a constant 4 MHz bandwidth, tunable over a 20 MHz to 1000 MHz frequency range.

The unit is an active circuit with no loss and very small size ($4 \times 1 \frac{1}{2} \times 7$ inches). At the heart of the circuit is a 4 MHz bandwidth, 4-pole resonant cavity filter with a Q of 325. A unique double-heterodyne technique currently being patented routes the signal to the filter and back to the operating frequency.

Noise figure is specified at 10 dB, and the third-order intercept is 15 dBm, typical. Power requirements are 12 VDC with 6 watts consumption. Price of the APS-204 is \$995.

Optoelectronics Inc.
INFO/CARD #229



RF Modem for Wireless Networking

The RF Modem, using uncoded spread spectrum tech-



niques defined by FCC Part 15, has been introduced by Proxim. Applications include hospital bedside monitors, point-of-sale devices, portable data recorders, traffic management, and security systems.

Usable for distances up to 500 feet, the RF Modem interfaces to a computer using a standard RS-232 or RS-485 port. Packages and unpackaged versions are available.

Two unpackaged version are available for designing into OEM products. One operates at 121 kbps, while the other operates at 242 kbps.

Proxim, Inc.
INFO/CARD #228

Farnell Instruments Now Available in U.S.

Wayne Kerr announces the introduction of the Farnell line of RF test instruments to the U.S. and Canadian markets. Farnell equipment has previously been marketed in the U.K. and the rest of Europe.

The product line includes the following signal generator products: The PSG1000 Synthesized Signal Generator (pictured below) covering 10 kHz to 1 GHz, the ESG Signal Generator which also covers 10 kHz to 1 GHz, the 100 kHz to 2 GHz SSG2000 Signal Generator. Other current product models include the 352C Spectrum Analyzer, covering 300 kHz to 1 GHz, and the AMM2000 Modulation Meter with 250 kHz to 2.4 GHz coverage.

Wayne Kerr Inc.
INFO/CARD #227

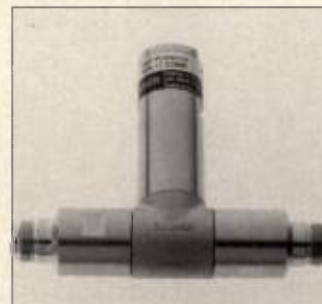


Lightning Protector for Cellular Radios

Huber+Suhner has introduced a 1/4-wave shorted stub specifically designed for cellular radio base stations, with an operating frequency range of 797-1016 MHz. Features include lower residual pulse than other protection methods and no intermodulation effects. The unit is offered with either type N or 7/16 connectors.

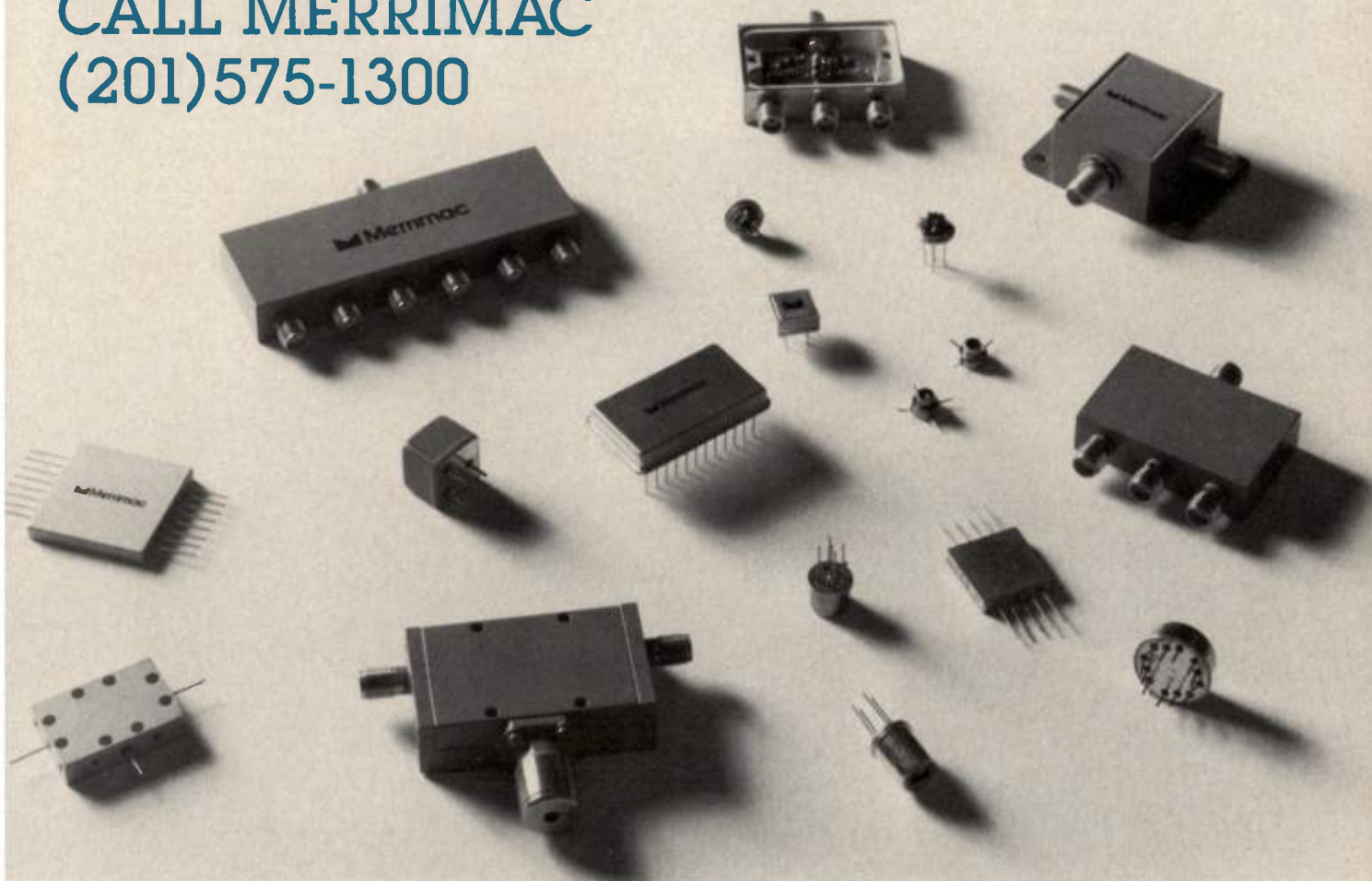
Mechanically, the lightning protector features zinc diecast construction, waterproof design for outdoor installation, and easy mounting with standard plumbing brackets or clamps.

Huber+Suhner AG
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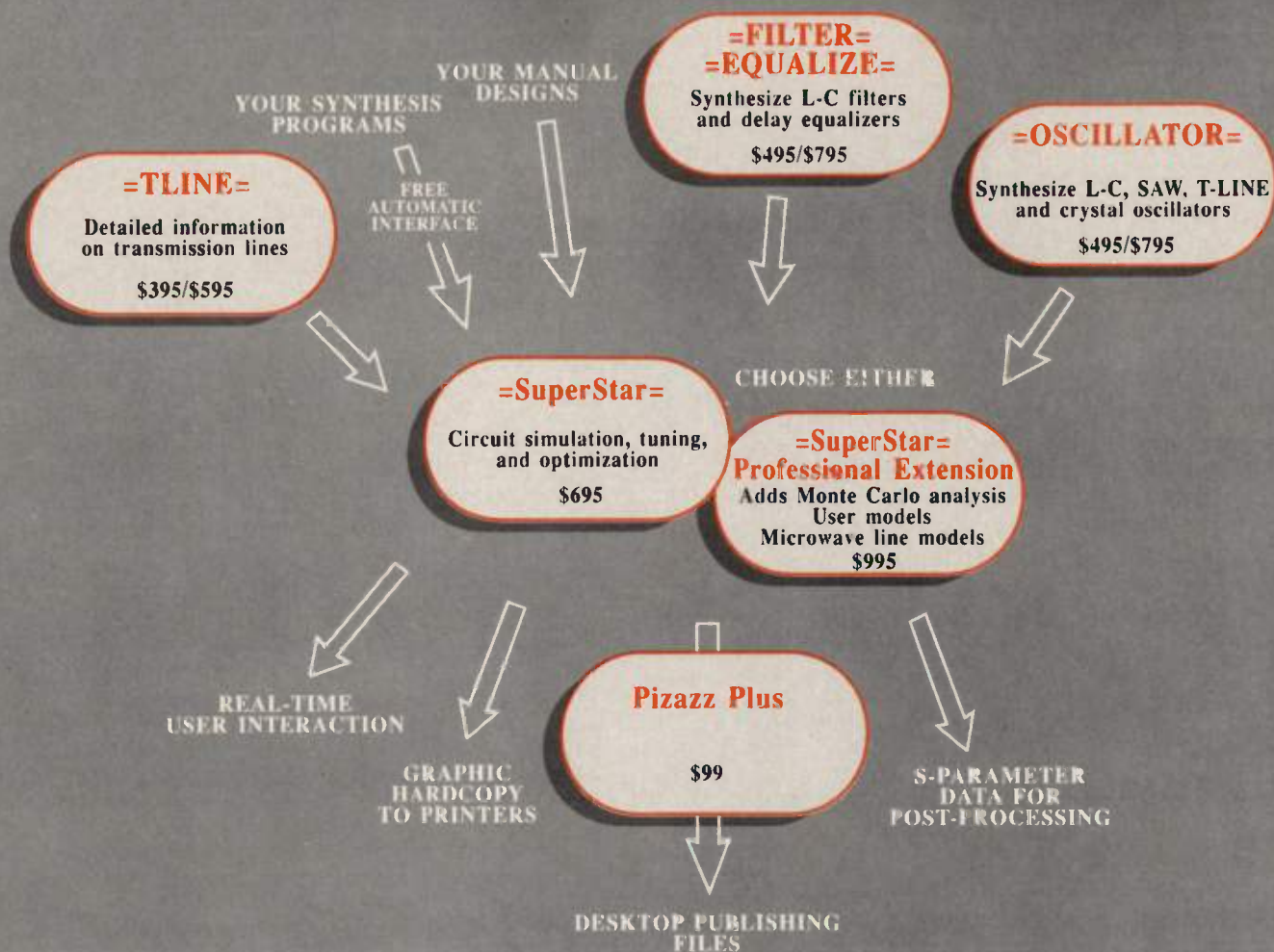
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Smaller SMT DIP Package

Signetics announces the first two integrated circuit products in their new SSOP small-footprint SMT package. The package is small enough to replace chip-and-wire assembly in some applications. The NE/SA575DK Compa-dor and the NE/SA605DK FM IF System are available in the 1.5 x 4.5 x 6.75 mm SSOP. In 100s, these device are priced at \$2.04 and \$3.54, respectively.

Signetics Company
INFO/CARD #226

GPS Precise Positioning Module

Stanford Telecommunications has received National Security Agency endorsement of its STEL-9300 Precise Positioning Service - Security Module (PPS-SM), allowing authorized users to realize the most accurate capabilities of GPS. Anti-Spoof and Selective Availability techniques deny access to the Precise Positioning Service (PPS) to unauthorized users. With this capability, GPS equipment can provide access to PPS for authorized users without making the equipment classified.

Stanford Telecommunications
INFO/CARD #225

Standard Power Amplifier Line

Alpha Industries announces the new CMA Series of Standard Power Amplifiers for terrestrial and satellite microwave communications systems. Models cover 5.9-6.4 GHz with 60 dB linear gain (minimum) and power outputs ranging from 40 to 44 dBm (minimum), utilizing +12 or +15 volt power supplies.

Alpha Industries, Inc.
INFO/CARD #224

Power Measurement Upgrade

High speed CW power measurement capabilities have been added to the 8003 Precision Scalar Analyzer from Wavetek Microwave. These new measurement modes provide power readings at greater than 100 per second over GPIB. The upgrade provides a 10 to 100 times improvement in speed over thermocouple CW power meters. This speed is particularly attractive in automated measurement systems.

Wavetek Microwave, Inc.
INFO/CARD #223

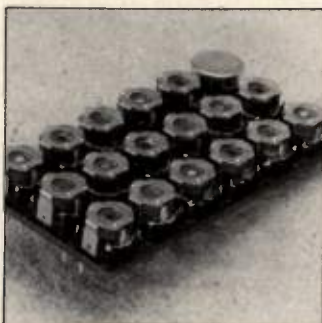
1000 Watt Amplifier

Model BHE4819-1000 from Power Systems Technology offers 1000 watts CW power output in AB linear service over 400-1000 MHz. RF input is 0 dBm, in CW, AM, Pulse, FM, or phase modulation. Protective circuits prevent damage from excessive VSWR, and the unit offers graceful degradation in the event of transistor failure. Applications include EMI testing, EW and ECM, troposcatter, television broadcasting, and laboratory amplification.

Power Systems Technology Inc.
INFO/CARD #222

Video Brickwall Filters

Miniature filters for NTSC and PAL video systems are available from Allen Avionics, offering cut-



off frequencies from 2 to 8.4 MHz. P.C. board or packaged versions are available. The typical ABW-4P20-P has passband flatness of 0.15 dB to 4.2 MHz, less than ± 10 nsec delay distortion, and 1.066 stopband ratio. These filters reduce interference from high frequency subcarriers and other out-of-band signals.

Allen Avionics, Inc.
INFO/CARD #221

ECL Oscillators With Enable/Disable

MF Electronics announces the M1900 ECL oscillators for graphics and other high speed circuits requiring clocks of 10 to 225 MHz. A unique feature is a "tristate" enable/disable feature which effectively disconnect the output. System test, selectable clock frequencies, and noise reduction by shutoff of unused sources are all benefits of this feature.

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The new HP 11757B multipath-fading simulator/signature test set features an internal printer, firmware and error counters to produce permanent records of M-curves. These signatures are used in installation and maintenance of digital radio systems. The HP 11757B is \$19,000,

and an upgrade to the earlier HP 11757A is \$4000.
Hewlett-Packard Co.
INFO/CARD #219

MIL-STD-2000 Capability

Bliley Electric announces that it is qualified and certified to design and manufacture soldered electrical and electronic assem-

blies to meet all requirements of MIL-STD-2000. This capability is used in custom-designed crystal oscillators, and is also available for contract assembly.

Bliley Electric Co.
INFO/CARD #218

Feedthrough Terminations

Coaxial feedthrough termina-

tions designed to terminate inputs to oscilloscopes, frequency counters, RF voltmeters, etc., have been introduced by Elcom Systems. They are available in 50, 75 and 93 ohms, with BNC, TNC, type N or SMA connectors. Prices start at \$17.50 in small quantities.

Elcom Systems, Inc.
INFO/CARD #217

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OCXO from 5 MHz to 20 MHz
 TCXO from 0,1 Hz to 40 MHz
 XO from 0,1 Hz to 140 MHz

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Hybrid Amplifiers

Two new 10 MHz - 1.2 GHz high linearity amplifiers have been introduced by Motorola RF Devices. Models CA5900 and CA5915 are designed for operation in 50 ohm linear systems, offering 28 or 15 volt supply, 15 dB gain, and 1 watt power output at 1 dB compression. Packaging is in the CATV style case 714P-01, and they are also available with a low profile heat sink package. Pricing is \$91.30 (U.S. delivery, 100s).

Motorola RF Devices
INFO/CARD #216

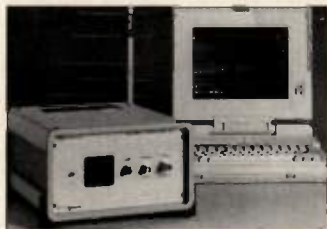
3-Way Divider

New three-way power dividers/combiners are available from RLC Electronics. Featuring 5 watt dissipation and either pin or SMA connector configurations, standard units are offered in octave or narrow band ranges from 500 MHz to 12 GHz. Amplitude balance is ± 0.5 dB, phase balance is ± 3 degrees, and isolation is 18 dB minimum. Prices start at \$225.

RLC Electronics, Inc.
INFO/CARD #215

EMI Measuring System

The Vianello PMM-8030 from IBEX Group is a complete emission measuring and monitoring system in the 30 to 1000 MHz band, meeting CISPR, VDE, FCC and EN specifications. External PC control offers ease of use and



flexibility. A complete package is available, including receiver, antennas, computer and software.

IBEX Group Inc.
INFO/CARD #214

SMT RF Chokes

Vanguard Electronics is offering surface-mount RF chokes designed and built to MIL-C-15305 or MIL-T-27. The chokes are fully encapsulated; suitable for various soldering techniques. Products are available for the 1 to 500 MHz frequency range with impedances up to 10 kohms. 1000-

piece pricing ranges from \$3.50 to \$4.50 each.

Vanguard Electronics Co.
INFO/CARD #213

EMC/TEMPEST Antennas

Two new models from Antenna Research Associates are designed for shield room testing requirements of MIL-STD 285 and NSA 65-6. Each antenna is a 12-inch tuned loop operating from 1 kHz to 30 MHz. The PLA-130/A is a two-band transmit/receive antenna capable of handling 1 kW input power, and the Model ALA-130/A is a receive antenna containing a low noise preamplifier. Each is individually calibrated.

Antenna Research Assoc., Inc.
INFO/CARD #212

New SMT Filters

Trilithic has introduced two new surface mount filter packages, which extend center frequencies to the 10-2000 MHz range and increase ultimate rejection to 70 dB. Designs available include Chebyshev, Butterworth, Linear Phase, Bessel, Gaussian and Elliptic with -3 dB bandwidths from 2 to 150 percent.

Trilithic, Inc.
INFO/CARD #211

Crystal Products

New products from KS Electronics include a 70.0 MHz VCXO with pullability of over ± 1500 ppm and +10 dBm output. Higher frequencies are also available. Also announced are 70.0 MHz crystal filters with 1.0 MHz bandwidth, with other frequencies available. The company has the capability of manufacturing crystal filters with third order intercept of +30 dBm.

KS Electronics
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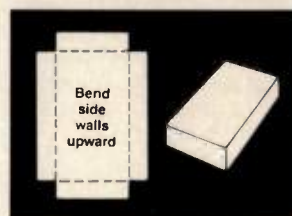
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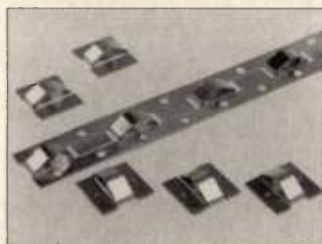
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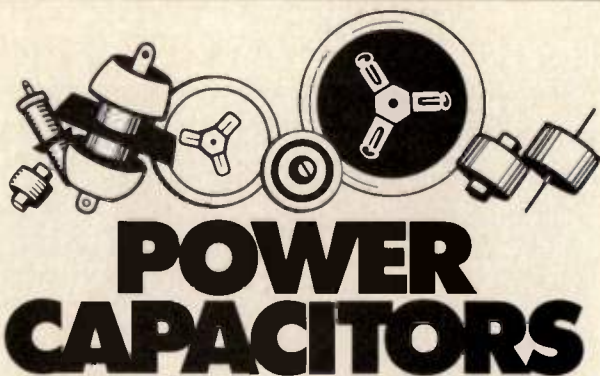
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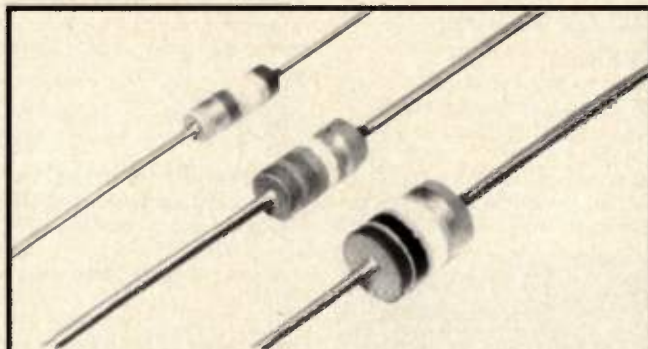
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Surge and Transient Considerations

By Gary A. Breed
Editor

Surges and transients on power and signal lines are major EMC problems in electronic equipment of all types. It is common to see these events occur at magnitudes of thousands of volts, with a considerable amount of total energy. The problem is a particular nuisance at radio frequencies, where even small amounts of energy can interfere with the reception of desired signals. Eliminating, or at least limiting the effects of this unwanted energy usually requires a combination of several standard protection components and circuits.

Surges

Surges are by common definition a low frequency phenomenon, whether they occur on power distribution circuits or signal lines. Intuitively, they can be considered as low frequency modulation of the desired electric or electronic energy normally found on the line. Most signal line surges can be traced to lightning, while power lines are also subject to the inductive kick of machinery or sudden changes in load.

Lightning surges (not direct hits) are usually the result of the discharge path being inductively coupled to a power or telecommunications line. Rural locations are particularly susceptible, since there are long, uninterrupted stretches of distribution lines available for coupling. This type of surge can be very substantial, but is largely dissipated by a combination of spark-gap discharge units and metal-oxide varistors (MOVs). However, equipment must be designed to withstand surges that are below the protection threshold of these devices.

Inductive kick and distribution circuit load changes are somewhat more predictable. Unless they are accompanied by switching transients, they tend to be smooth changes in voltage, with a damped swing of voltage on the line. Their effects are almost always limited to power supply variations. Linear power supplies must be designed to have an adequate raw DC voltage and capacitive storage to remain above the regulator's dropout threshold. Switching supplies



react much like they would to a change in load. If the supply is near its maximum current output capability, the voltage swing of the surge can modulate the regulated output.

Transients and Noise

Variations with very short time duration or with repetitive spikes are called transients, and they come from a large number of sources. In power distribution, they occur when circuits are switched in and out of the power grid, or when nearby lightning results in energy from close coupling or by direct contact of charged ions to the lines. Physical damage causing arcing or shorts contribute to the problem, as well. On signal lines, the potential effects due to lightning are the same as on power lines. AC line filters will remove most of the RF noise conducted by the lines, but

radiation from nearby wiring will still be present, unless shielded by conduit.

AC or DC potentials on signal lines can be another source of trouble. DC and ringing voltages are present on telephone circuits, and DC power is present on CATV distribution coaxial cable, as well the RF signals. If a leakage path is present, or if balanced lines become unbalanced with respect to ground, the resulting currents can cause problems with hum pickup from longitudinal currents coupled from AC power distribution circuits. These currents are normally suppressed by the balanced nature of the signal lines.

Outside sources of trouble are common on signal lines. Failed distribution equipment, faulty customer equipment and physical damage are relatively common. The result may not be restricted to disturbances in the desired signals

Metallic Surge (across wires):	100 A, 800 V, 10 x 560 msec rise/fall
Longitudinal Surge (common mode):	200A, 1500 V, 10 x 160 msec
Surge Protection Devices:	Breakdown voltage greater than 250 V
Leakage Current (tip or ring to ground):	10 mA, 1000 VRMS
On-Hook Impedance (tip or ring to ground):	100 kohms
Signal-to-Noise Ratio (voice grade):	greater than 30 dB, typical
Signal-to-Noise Ratio (digital loops):	depends on data rate
Longitudinal-to-Metallic Balance:	60 dB minimum

Table 1. Some telephone circuit performance specifications.

and voltages. A broken shield on a coaxial line or unbalance due to inadvertent grounding of one side of a telephone line increases susceptibility to RFI from nearby transmitters. Illegal telephone and CATV connections are notorious culprits for increased RFI because they create mismatch, unbalance or compromise shielding integrity.

Noise in the user's equipment is the result of all the above effects, and more. Other sources of impairment on signal lines include crosstalk, distortion and excess attenuation. Line balance, its physical condition, and the status of transmission equipment all affect overall signal transmission performance. Table 1 shows some key performance specifications for telephone line performance, including surge and overvoltage protection standards, as well as signal path performance.

Protection and Prevention

Transients and noise are insidious problems. They often result in intermittent problems that are extremely difficult to diagnose. Methods of dealing with

voltage-current surges and transients include filtering techniques to minimize high-frequency components of the disturbances, along with fuses, spark gap, varistor and zener protection against more dramatic spikes.

Standard practices in the construction of these facilities include separation of signal and power distribution conductors whenever possible, metallic sheathing on telephone trunk lines, periodic grounding to minimize the accumulation of a potential on the shield, and static discharge wires on certain power lines. CATV systems, with a completely shielded system, usually include a ground block with a spark-gap arrester at the entrance to the customer's location. Telephone lines have fuses in both tip and ring circuits, which may be at a distribution pedestal or at the customer service entrance.

Noise, hum, distortion and interference that appear along with the desired signal require different techniques. Most areas of noise performance are dependent on the condition of the transmission equipment and the transmission me-

dium. Designers of user equipment must be certain that they are compatible with the wire or coax system they utilize, and must allow sufficient design margin to be sure that the transmission medium is the weakest link in the system. Minimizing noise and distortion requires control of signal levels to maintain exactly what is specified for the system used.

Above all, familiarity with system requirements, and the nature of surges, transients and noise helps make a good relationship with the power, telephone and CATV companies. The best service always seems to be given to knowledgeable customers.

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A Simple Clock and PSK Carrier Recovery Circuit

By Francois Methot
RadioSysteme, Inc.

Today, analog designers have to interface with the digital world by recovering the clock from their modem (electric or optic) for high speed data transmission.

The interface between analog and digital is critical, especially when it must be done at a low cost. I propose a different approach to obtain PSK demodulator and clock recovery with the same kind of circuit.

The circuit is based on a PLL with a "D" flip-flop as the phase comparator, coupled with a VCXO to avoid locking on undesired frequencies.

The carrier and/or clock is recovered by using the positive edge of the incoming PSK signal or data stream to sample the carrier or clock and create a DC voltage to control the VCXO. Figure 1 depicts the interconnection between the "D" flip-flop and the VCXO.

However, there is a compromise to be made in using this design. The phase detector does not have a linear transfer

function versus phase error. The gain of the phase detector ($K\phi$) in frequency is low but $K\phi$ for the phase is very high. Figure 2 shows an example of the transfer function of this design.

In the absence of data, the output of the "D" flip-flop can be at a logical "0" or "1" depending where the data did the last sampling in the clock, so the VCXO can be at the maximum or the minimum frequency. When the PLL begins the search for the frequency, it requires many samples to converge on the correct frequency, but the search for the phase is very fast.

For a low bit rate, I recommend the use of a CMOS "D" flip-flop because it has the same threshold for a logical "1" or logical "0". Most of the time a passive filter will do, when it is connected on the Q or Q output, depending if it is the positive edge or negative edge that is required at the center of the data bit.

For a high bit rate, an ECL "D" flip-flop must be used because it has only 1 nsec of propagation delay. However, an active filter (integrator) must be implemented to increase the range on the diode varicap in the VCXO circuit.

This circuit can be implemented to demodulate a PSK signal with some compromise. The carrier must be an exact multiple of the bit rate. Moreover, the phase transition of the carrier must be synchronized with the data. These requirements are cost efficient and easily implemented on the demodulator.

When feasible or when given the choice of IF frequency, the carrier to be modulated must be an exact multiple of

the data clock to be transmitted. A simple PLL can multiply the clock synchronously and form the carrier to be modulated. Figure 3 shows an example of a simple PSK modulator. Now the PSK carrier can be seen as a data stream of "1", "0", "1" ..., with one extra "0" or one extra "1" at each transition of the data. Refer to Figure 4 for the timing diagram.

A carrier can be regenerated at twice the PSK carrier frequency with the same circuit for the clock recovery and later, by dividing the carrier by two. A simple exclusive "or" can demodulate the PSK signal. Because of the use of the divider, there exists a phase ambiguity of 180 degrees. By using a differential code or NRZI encoding, the demodulated data will always be the same polarity.

In Figure 5, one ECL IC is used for limiting the PSK signal, another IC has the phase comparator and the frequency divider, and a final IC to demodulate the PSK signal and produce the VCXO. An NRZI to NRZ must be added after the low pass filter. This technique, compared to the conventional approach, requires no treatment on the signal (X^2) to rebuild the carrier information.

The component values of the integrator were calculated using Communica-

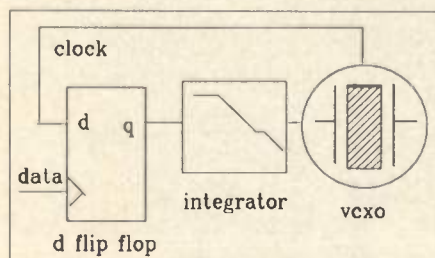


Figure 1. Interconnection between "D" flip-flop and VCXO.

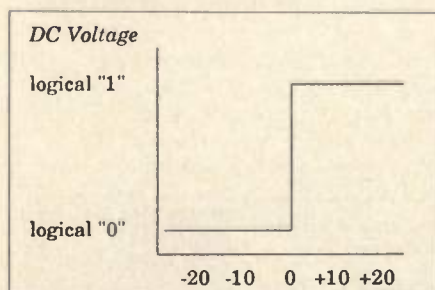


Figure 2. Transfer function from design in Figure 1.

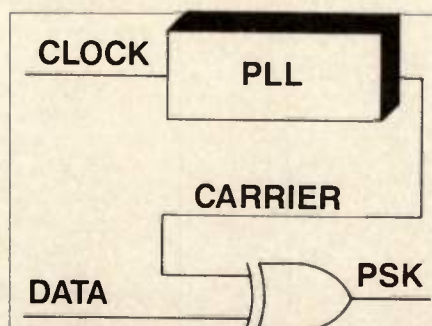


Figure 3. A simple PSK modulator.

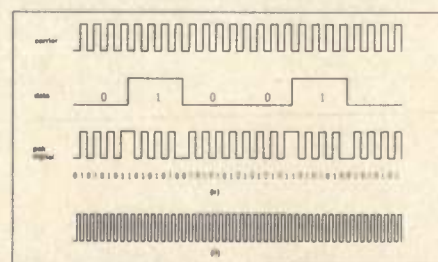


Figure 4. Timing diagram (a) the PSK signal can be seen as a data stream of 1, 0, 1, ... with one extra 0 or 1 at each transition of the data, (b) a carrier can be regenerated at twice the PSK carrier with a simple clock recovery and when it is divided by two, can demodulate the PSK signal.

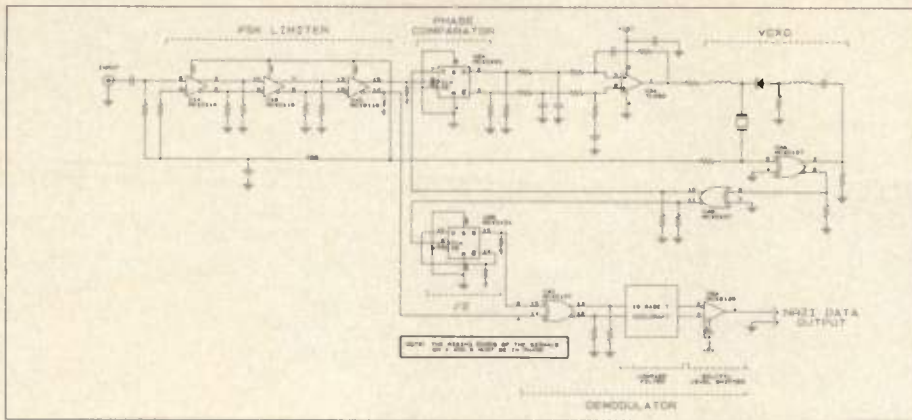


Figure 5. PSK demodulator circuit.

tions Consultation Corporation's (now Compact Software) PLL Design Kit.

Figure 6 shows my RF Design Contest entry: clock recovery based on "D" flip-flop phase detector. Figures 7 and 8 show the performance of the circuit's clock and data recovery.

Conclusion

The circuit shown in Figure 6 uses only one IC to both perform the phase comparator and sample the data. The VCXO was made with a crystal in fundamental at 139.264 MHz to eliminate frequency multiplication and achieve a larger range. A ΔF of 50 kHz was obtained with ± 15 kHz of capture range. The data stream comes from the optical detector from a source at -25

dBm. The clock recover still locks up to -31 dBm even if the data is at 10^{-2} of error rate. For the PLL, the parameters are: $W_n = 10$ kHz, $K_{VCO} = 1$ kHz/volt, $K_o = 10$ volts/radian, $\zeta = 0.7$. **RF**

About the Author

Francois Methot is President of RadioSysteme Inc., and has 15 years experience in CATV, digital telephony, radio communications and fiber optic networks. He can be reached at (514) 426-0773.

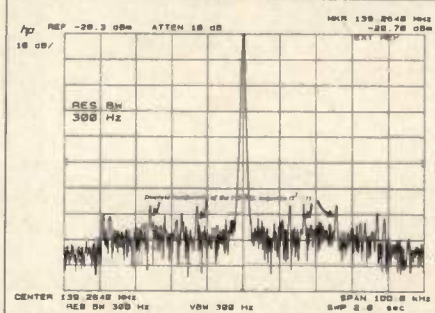


Figure 7. Spectrum analyzer display of the recovered clock signal.

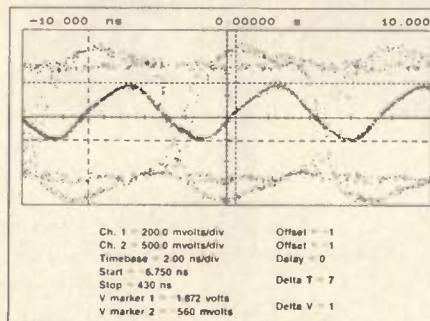


Figure 8. Digital oscilloscope display of clock and recovered data.

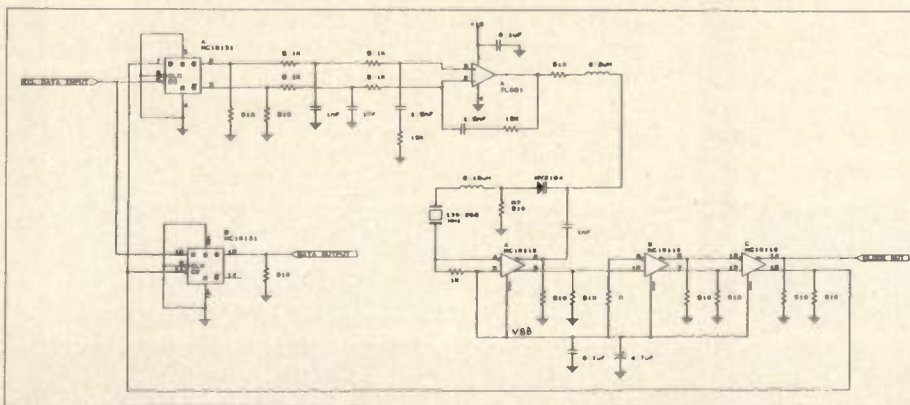
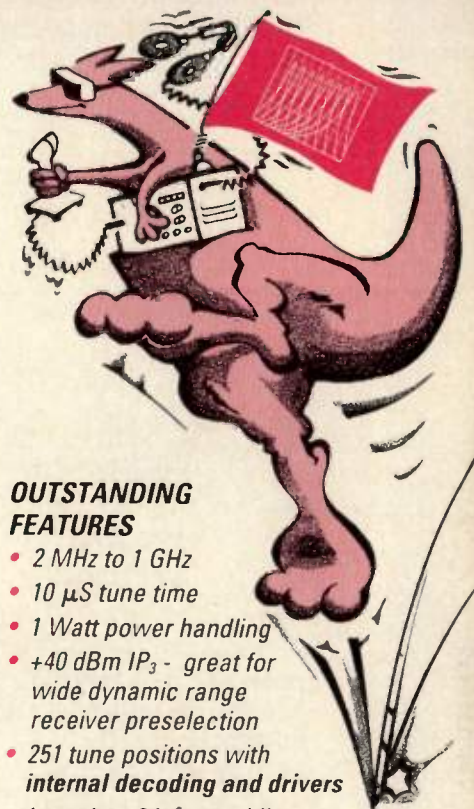


Figure 6. Clock recovery based on "D" flip-flop phase detector.

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VHF and UHF Crystal Oscillators — Part II

By Andrzej B. Przedpelski
A.R.F. Products, Inc.

This note contains some theory, more circuits, some suggestions and further considerations of 5th, 7th, and 9th overtone crystal oscillators (up to 630 MHz) using simple easy to build circuits.

I should have started Part I, July of 1990 (1), with some applicable theory, but I got carried away by the ease of constructing and adjusting these high frequency crystal oscillators and their reliable performance. To rectify this omission let's consider the requirements for an oscillator. Barkhausen tells us that for a circuit to oscillate, two conditions have to exist.

- the open loop gain $(A) \times (\beta)$ of Figure 2) has to be more than unity, and
- the phase of this gain has to be $360n$, where n is an integer (including 0).

In the circuits described, the gain, A is supplied by the MMIC amplifier and the feedback, β consists of the crystal, in series with a length of transmission line, and the capacitor(s). Using an ideal non-inverting amplifier (such as Figure 1a) (1), the feedback can have 0 degrees phase shift and no phase shifting network is thus necessary. The crystal can, theoretically, operate in series resonance (presenting only a resistance in the feedback path) and becomes an attenuator in conjunction with the 50 ohm input impedance of the MMIC amplifier. If this attenuation is

lower than the gain of the amplifier, the circuit will oscillate. The oscillations start slowly (relatively speaking) by amplifying circuit noise at the resonant frequency until the amplifier saturates (gain of the amplifier becomes equal to the feedback attenuation).

In the inverting amplifier configuration (such as Figure 1b) (1), the feedback has to provide the (theoretically) required 180 degree phase shift to allow oscillations to take place.

In practice, of course, the non-inverting amplifier has some phase shift at the operating frequency, and the inverting amplifier phase shift is usually somewhat less than 180 degrees. Thus, the feedback circuit has to compensate for these small phase shift discrepancies. The inverting amplifier oscillator configurations are generally preferred, since they have less tendency to oscillate at undesired frequencies. Any "simple" undesired feedback usually does not have the required 180 degree phase shift.

To provide the needed feedback phase shift, different methods can be used. The crystal itself can provide some phase shift. However, it is better to use it in a series resonance mode to reduce losses through the crystal. Thus, a phase shifting network providing almost 180 degree phase shift has to be used. At lower frequencies, a pi or tee phase shifter can be used. However, at the higher frequencies, accurate inductors

are difficult to construct, and spurious reactances cause additional uncontrolled phase shifts. To overcome this, and to provide very simple implementation of reliable circuits, a transmission line can be used to provide the main required feedback phase shift. This approach is feasible at the higher frequencies and provides easily reproducible phase shifts. To demonstrate this, the circuit shown in Figure 3 was built. While feasible and very simple it was not considered very practical since the length of the transmission line was difficult to optimize. A more feasible method using variable capacitors to fine-adjust the feedback phase shift was used in the more practical circuits. This method had the additional advantage of allowing some "pulling" of the oscillator frequency.

Now back to some more circuits

Circuit # 1 (350 MHz fifth overtone)

The circuits presented in Part I used hybrid type RF amplifiers. To reduce size still more and allow the use of surface mounted components, monolithic RF amplifiers were then tried. A typical inverting type was used (Avantek MSA-0735). Two possible circuit configurations were tried (Figures 4 and 5). Both gave an output of 6-8 dBm with either a 3.6 inch or 6 inch length of transmission line. The capacitors had to be retuned, of course. The circuit of Figure 5 seemed to have lower harmonic content (35-40 dB below fundamental for second and third harmonics). While monolithic amplifiers usually have the disadvantage of requiring input and output capacitors, one capacitor can be omitted, since the crystal provides DC isolation. These amplifiers also require a biasing resistor. In the case of the circuits shown, this was actually taken advantage of to provide a circuit which worked reliably directly from an airborne power source of 22-30 VDC. This was accomplished by the use of the high value (1000 ohm) biasing resistor.

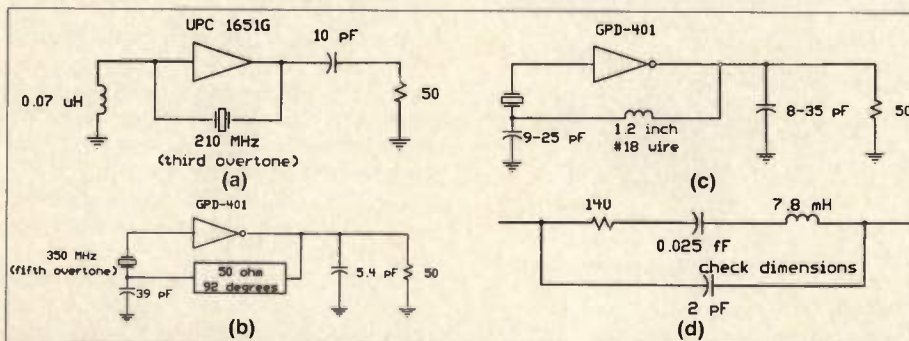


Figure 1. Figures from Part I: (a) third overtone circuit, (b) 350 MHz fifth overtone, (c) overtone oscillator with L/C phase shift, (d) approximate crystal equivalent circuit at fifth overtone.

Circuit # 2 (490 MHz seventh overtone)

The basic circuit of Figure 5 could be easily modified to provide reliable operation on the seventh overtone of the crystal. The same crystal was used with an 8.2 or 7.2 inch length of transmission line, as shown in Figure 6. An output of 6-8 dBm was obtained, but the harmonic content was high (-10 to -15 dB).

Otherwise, the circuit performed reliably.

Comments

Output was always observed on a spectrum analyzer to ensure that the proper overtone operation was obtained. Power output was measured at the desired output frequency.

All circuits were checked to ensure

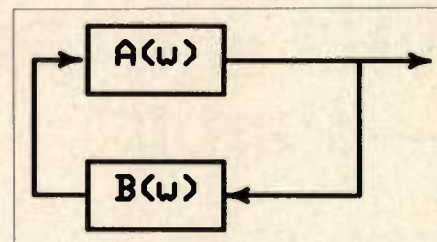


Figure 2. Oscillator equivalent circuit.

that no mode switching could be induced. DC power was turned on instantaneously or was brought up gradually while output was observed on the spectrum analyzer. The oscillators always started at the desired frequency.

50 ohm coaxial transmission lines were used, since they are readily available and easily changed. Once the circuit is finalized, the transmission line can be part of the PC board. There is no apparent reason why other impedances cannot be used. However, only 50 ohm coax cables were used.

An amplifier gain of about 12 dB at the desired frequency seems adequate. Higher gain amplifiers would allow a higher loss in the feedback circuit. This could reduce the crystal drive and improve stability. However, too much gain may make the circuit more susceptible to oscillations at undesired frequencies.

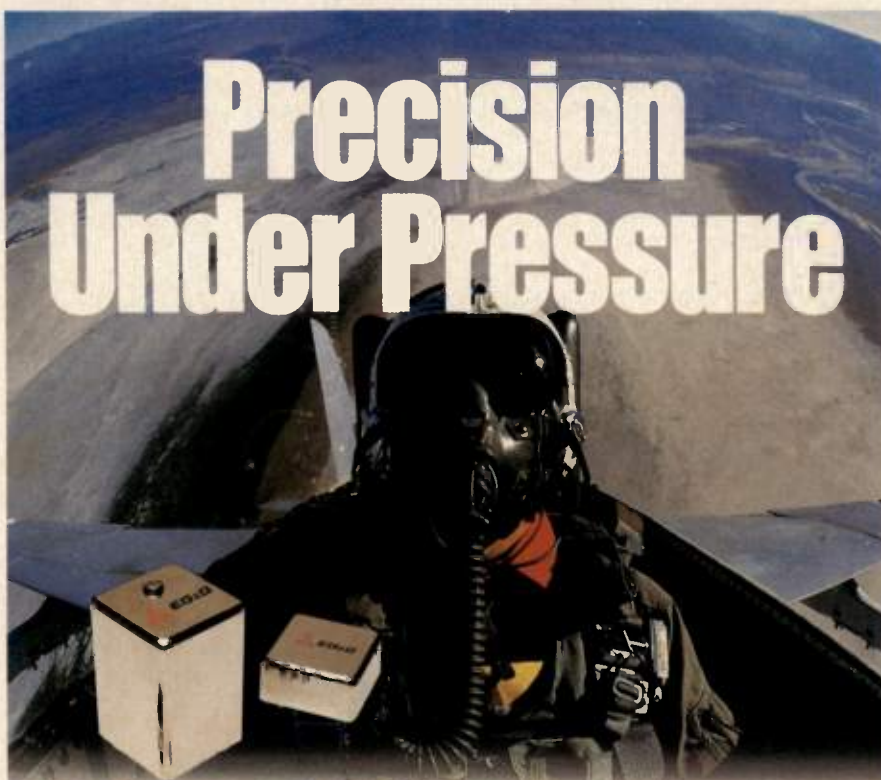
The phase shift in the amplifier changes with frequency. The typical inverting amplifier actually becomes non-inverting at higher frequencies. This occurs at about 4.5 GHz for the MSA-0735 used in the tests. The amplifier still has gain at this frequency, however, and this should be considered.

The crystal resistance does not seem to be critical. The crystal used had a resistance of about 140 ohms at the fifth harmonic (Figure 10). Higher resistance crystals can be used, if the amplifier gain is adequate.

The feedback path loss has to be less than the amplifier gain to sustain oscillations. The input impedance to the feedback path should be higher than 50 ohms to allow most of the amplifier power output to appear as useful power in the load. This is not unreasonable, since the feedback path can have appreciable insertion loss.

A quick temperature check was made using the circuit of Figure 5. The frequency vs. temperature curve was smooth and no mode jumping was observed.

The circuits were very reproducible.



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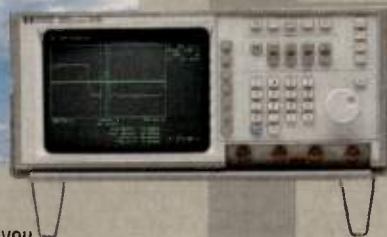
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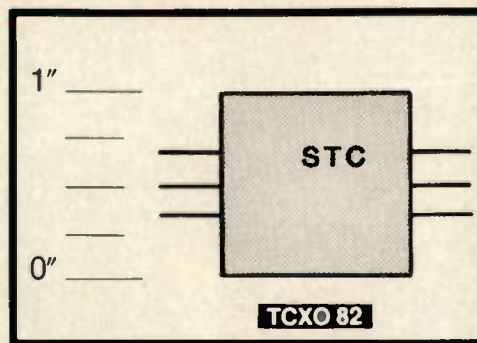
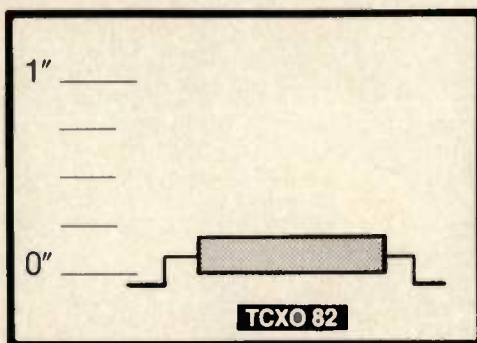
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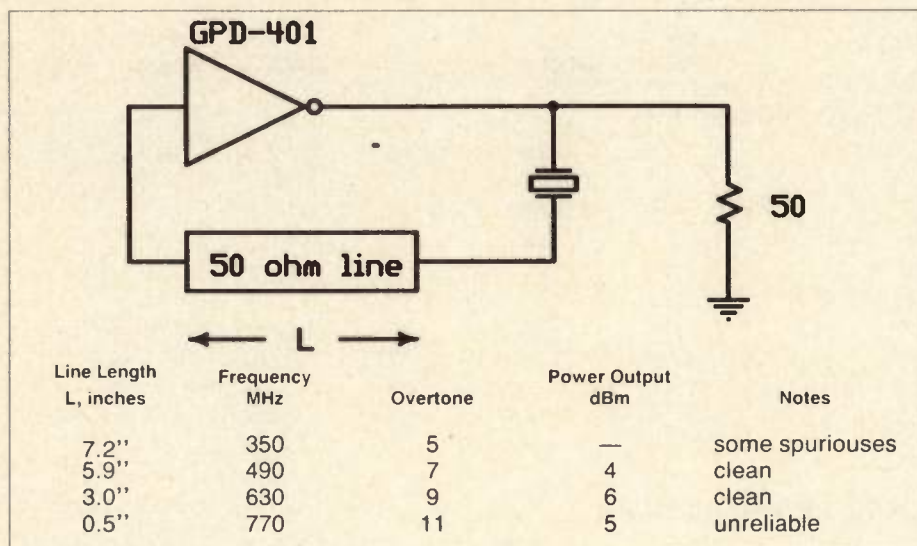


Figure 3. The simplest circuit.

The same circuit was built using direct interconnections (no chassis), using a ground plane or a PC board with all surface mount components. The performance was essentially the same.

Increasing feedback loss (reducing drive to the amplifier) reduces harmonic content (as expected). This reduces power output somewhat. For a power output reduction of 1-2 dB, the harmonic content can be reduced by 15-25 dB.

The initial tuning procedure is quite simple. Using a spectrum analyzer (to make sure that the right overtone operation is achieved) set one capacitor and tune the other (if two are used). If proper operation is not obtained, retune the first capacitor and repeat tuning with the other. Once the proper frequency is obtained, the output can be maximized using both capacitors.

The crystal can oscillate at the fundamental or an odd overtone. Circuit #10 (Figure 1c) (1), for instance could be made to oscillate at the third, fifth and seventh overtone by retuning the two capacitors. At these different frequen-

cies, the phase shift through the amplifier is different. While at very low frequencies it is the theoretical 180 degrees, it decreases with frequency. The crystal also looks different at these frequencies. Mr. David Smith of Innovative Frequency Control Products, Inc. kindly supplied typical equivalent circuit values for the crystal used (see Figure 1d) (1). Series R is about the same for the overtones, but is about 1/6 this value for the fundamental operation. The motional inductance (series L) is about the same for the fundamental and overtones. The motional capacitance varies to provide the proper series resonance.

While operation only up to the 9th overtone (630 MHz) was confirmed, there were indications that the 11th overtone was also feasible. The transmission line length became very short and was difficult to optimize. However sporadic operation at 730 MHz was obtained using the Figure 3 circuit and a transmission length of about 0.5 inch. As results of Figure 3 indicate, there is some spurious phase shift caused by

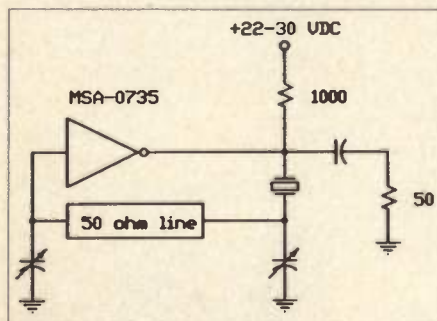


Figure 4. 350 MHz fifth overtone oscillator.

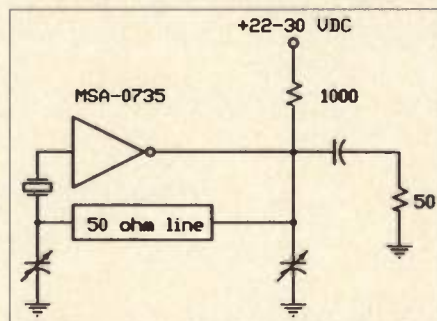


Figure 5. 350 MHz fifth overtone oscillator.

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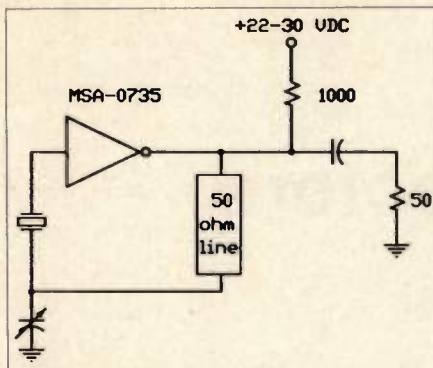


Figure 6. 490 MHz seventh overtone oscillator.

the stray inductances and capacitances in the circuit. With the breadboard used it was also somewhat difficult to accommodate the very short transmission line lengths. A better layout may make the 11th overtone operation quite feasible.

Another problem at the higher overtones is the stray pin-to-pin crystal capacity. While small (about 2 pF) its reactance at the higher frequencies becomes comparable or smaller than the crystal resistance. It may then provide the necessary feedback path to sustain oscillation at the wrong frequency. No effort was made to "tune-out" this reactance, as is sometimes done at lower frequencies to obtain true

series resonant operation and to reduce the risk of spurious oscillations.

While the above treatment of UHF crystal oscillators is by no means complete, the main effort was directed toward the simplest low cost reproducible designs using non-critical components. It was also meant as a starting point for further investigations using this general overall approach. The non-inverting amplifier circuits were also not adequately explored. One reason was that only one device was available at the time of the test; several manufacturers make suitable low cost inverting types.

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1. Andrej B. Przedpelski, "VHF and UHF Crystal Oscillators," *RF Design*, July 1990, pp. 63-65.

About the Author

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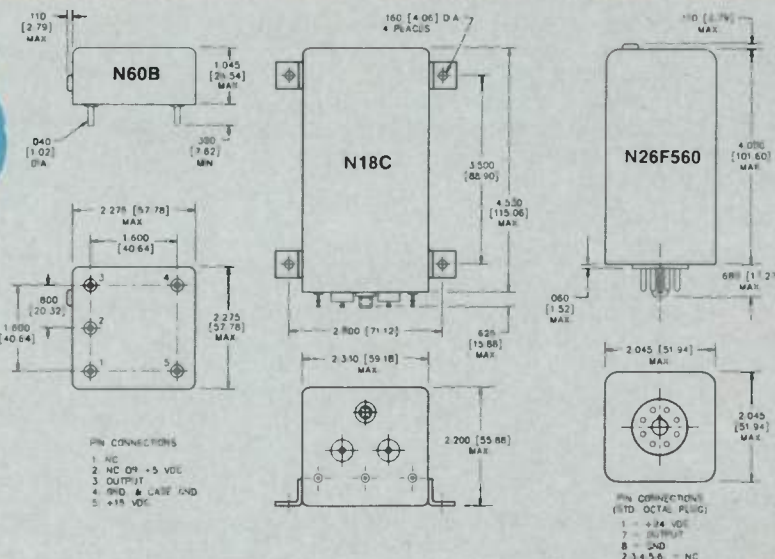
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Design of Quadrature Detectors

by Edward A. Richley
Xerox Palo Alto Research Center

The quadrature detector is a very common circuit used for the demodulation of both wide band and narrowband FM. However, there is a general misconception about its method of operation, and little information on its proper design. After careful analysis, it can be seen that the engineering trade-offs involved in the design of a quadrature network are easy to understand and very interesting. This article will describe such an analysis, and will show the important considerations for proper design.

The basic single-tuned quadrature detector network is shown in Figure 1. This network is typically used with FM receiver ICs such as the Motorola MC3357 (and related parts), Signetics NE604, or the RCA CA3189. The network has one input and one output. The input generally comes from the output of a limiting IF strip. This input, V_{IN} and the output of the network, V_{OUT} , are connected to the inputs of an analog multiplier.

The basic purpose of the quadrature network is to produce two signals which have a nominal phase difference of 90 degrees (hence the term "quadrature detector"), but with a variation in phase which depends on instantaneous frequency. The multiplier then serves as a phase detector, and can detect the lead or lag in phase caused by the interaction of the frequency modulated signal with the network.

In some versions of the circuit, the coupling capacitor, C_1 of Figure 1, is replaced by an inductor. The operation of the circuit is conceptually the same in both cases. Many texts give an explanation for the operation of this circuit which is based on the following assumption (1,2):

$$X_1(\omega_0) \gg R_L \quad (1)$$

where X_1 is the magnitude of the reactance of the coupling component, either a capacitor or an inductor, and ω_0 is the center frequency of the frequency modulated signal. ω_0 is also the resonant frequency of the network formed by L_L and C_L . With the assumption of equation 1, the nominal 90 degree phase shift is provided by the reactance of X_1 in series with a much smaller resistance, R_L . The subsequent phase lag or lead about the nominal will result from changes in the received signal frequency which cause the reactance of the R-L-C circuit to swing inductive or capacitive.

A quick look at some application notes, however, shows that the condition of equation 1 is often not met in practice. In fact, it is often strongly violated (3,4). Thus, the question arises as to how the nominal 90 degree phase shift is provided, and how the phase shift varies with frequency. In order to address this point of confusion, the network of Figure 1 must be examined more closely.

It is instructive to look at the network of Figure 1 in an alternate form. The dashed portion of Figure 2 can be transformed into a Norton equivalent, as shown in Figure 3. The resultant parallel capacitors combine as a single element of value $(C_1 + C_L)$. From Figure 3, it is evident that the nominal 90 degree phase shift can be guaranteed at the frequency ω_0 by simply making the combination of L_L and $C_L + C_1$ resonant

at ω_0 . Thus, the condition of equation 1 is not necessary. The size of C_1 relative to C_L does not affect the phase characteristics of the circuit at all, since the two capacitors are effectively in parallel. Their relative sizes only affect the amplitude of the output, and, as shall be seen later, the input impedance of the network. Thus, the condition of equation 1 is not necessary at all.

The Transfer Function

The circuit of Figure 1 has a transfer function, $H(j\omega)$, defined as the ratio of the output voltage to the input voltage, when both are expressed in phasor notation:

$$H(j\omega) = \frac{V_{OUT}(j\omega)}{V_{IN}(j\omega)} \quad (2)$$

From the equivalent circuit of Figure 3, it is easy to show that $H(j\omega)$ is determined by the impedance of the R-L-C circuit, and is given by:

$$H(j\omega) = - \frac{\omega^2 C_1 L_L R_L}{R_L (1 - \omega^2 L_L (C_1 + C_L)) + j\omega L_L} \quad (3)$$

There are four main concerns regarding the design of the quadrature network. These are as follows:

1. $|H(j\omega_0)|$ should be as large as possible.
2. The variation in $|H(j\omega)|$ over the desired range of frequency deviation should be as low as possible.
3. The phase variation in $H(j\omega)$ over the desired range of frequency deviation should be as linear as possible.
4. The input impedance of the network must be sufficient so as to not load the driving circuit.

In order to simplify the investigation of each of these requirements, it is useful to define some quantities. The first of these is the loaded Q of the network:

$$Q = \frac{R_L}{\omega_0 L_L} \quad (4)$$

In this analysis, any load resistance due to the phase detector input is lumped into the resistance R_L .

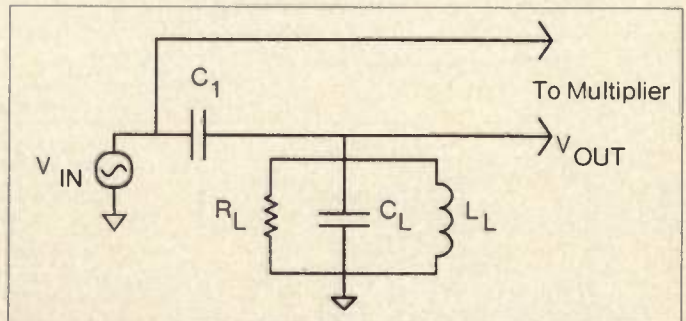


Figure 1. Basic single tuned quadrature detector network.

Another useful quantity is the center frequency, ω_0 . As described earlier, component values are chosen for resonance at ω_0 , so that:

$$\omega_0^2 = \frac{1}{L_L(C_1 + C_L)} \quad (5)$$

Finally, the frequency, ω , can be normalized to ω_0 by defining a new quantity, γ , as follows:

$$\gamma = \frac{\omega}{\omega_0} \quad (6)$$

In terms of these quantities, the transfer function is given by:

$$H(j\omega) = -\frac{C_1}{C_1 + C_L} \frac{\gamma^2 Q}{(1 - \gamma^2)Q + j\gamma} \quad (7)$$

From equation 7 it follows that the magnitude of the transfer function at the center frequency ($\gamma = 1$) is given by:

$$|H(j\omega_0)| = \frac{C_1 Q}{C_1 + C_L} \quad (8)$$

From equation 8 it is clear that for greatest magnitude of $H(j\omega_0)$, both Q and C_1 should be as large as possible. However, once C_1 is made comparable to C_L there is not much to be gained from increasing it further.

It is also useful to define the normalized frequency, γ , in terms of the deviation, ξ :

$$\gamma = 1 + \xi \quad (9)$$

In terms of ξ , the transfer function is:

$$H(j\omega) = -\frac{C_1}{C_1 + C_L} Q \frac{(1 + \xi)^2}{(-\xi(2 + \xi))Q + j(1 + \xi)} \quad (10)$$

The Amplitude Response

In order to examine the flatness of the amplitude response with frequency, it is helpful to look at the magnitude squared of the transfer function:

$$|H(j\omega)|^2 = \left(\frac{C_1}{C_1 + C_L}\right)^2 Q^2 \frac{(1 + \xi)^4}{Q^2 \xi^2 (2 + \xi)^2 + (1 + \xi)^2} \quad (11)$$

This expression can be expanded in terms of ξ about the center frequency, $\xi = 0$:

$$|H(j\omega)|^2 = \left(\frac{C_1}{C_1 + C_L}\right)^2 Q^2 (1 + 2\xi + (1 - 4Q^2)\xi^2 + \dots) \quad (12)$$

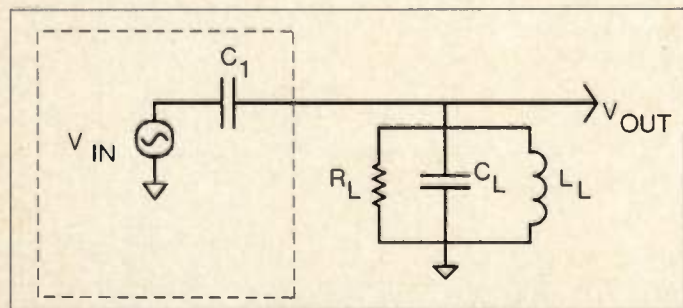


Figure 2. The quadrature network redrawn.

From equation 12 it is clear that in order to prevent AM distortion from becoming induced on the detector, it is necessary that:

$$|\xi| \ll \frac{1}{2} \quad (13)$$

and that:

$$4Q^2|\xi|^2 \ll 1 \quad (14)$$

In general, $|\gamma|$ is quite small (0.01 is typical) so that both of these requirements will be met for modest values of Q .

The Phase Response

The phase response can be examined in a similar manner. Equation 10 leads to the following expression for ϕ , the phase of the transfer function, $H(j\omega)$:

$$\phi = \arctan\left(\frac{(1 + \xi)}{\xi(2 + \xi)Q}\right) \quad (15)$$

Since the nominal value of ϕ is known to be 90 degrees ($\pi/2$), it is most useful to look at the deviation in ϕ about this mean. This deviation, θ , is given by:

$$\theta = \phi - \frac{\pi}{2} = -\arctan\left(\frac{\xi(2 + \xi)Q}{(1 + \xi)}\right) \quad (16)$$

The linearity of the phase response of the network can be evaluated by looking at the quantity $\partial\theta/\partial\xi$. From equation 16, it follows that:

$$\frac{\partial\theta}{\partial\xi} = -Q \frac{2 + 2\xi + \xi^2}{1 + 2\xi + (4Q^2 + 1)\xi^2 + 4Q^2\xi^3 + Q^2\xi^4} \quad (17)$$

This expression can be expanded in terms of ξ to give:

$$\frac{\partial\theta}{\partial\xi} = -2Q(1 - \xi + \frac{1}{2}(3 - 8Q^2)\xi^2 + \dots) \quad (18)$$

In order to maintain a linear phase response, it is important for the constant term in equation 18 to be the dominant term. Thus, the requirement for phase linearity can be expressed as follows:

$$|\xi| \ll 1 \quad (19)$$

and

$$4Q^2|\xi|^2 \ll 1 \quad (20)$$

These requirements can be seen to be almost identical to the requirements of equations 13 and 14.

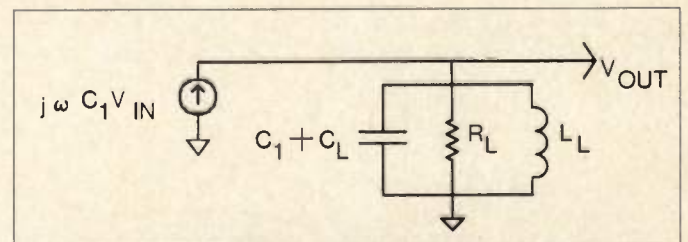


Figure 3. Norton equivalent of the quadrature network.

The Input Impedance

The remaining requirement has to do with the input impedance of the network. The magnitude of the impedance of the network of Figure 1, at the center frequency, ω_0 , as seen by V_{IN} , is:

$$|Z_{IN}(\omega_0)| = \omega_0 L_L \frac{\sqrt{Q^2 + \left(\frac{C_1 + C_L}{C_1}\right)^2}}{1 + Q^2 \left(\frac{C_1}{C_1 + C_L}\right)^2} \quad (21)$$

For even modest values of Q , and for values of C_1 which approach C_L (which is encouraged from equation 8), $|Z_{IN}(\omega_0)|$ approaches the following form:

$$|Z_{IN}(\omega_0)| \rightarrow \frac{\omega_0 L_L}{Q} \quad (22)$$

Clearly, a large enough value of Q and a small enough value of L_L could lead to an input impedance which is too low to be driven from the IF output. This consideration places a limit on the size of Q or the size of C_1 relative to C_L .

Design Procedure

Consideration of all of these constraints leads to a reasonable approach to the design of these single tuned quadrature networks. As with most engineering problems, there are trade-offs to be made. In general, the following procedure should suffice:

1. Choose frequency deviation or center frequency such that $|\xi| \ll 1/2$.
2. Choose Q as large as possible, without violating the requirement: $4Q^2 |\xi|^2 \ll 1$.
3. Choose the largest value for L_L which can be obtained within the limits of physical size, self-resonance, and the effects of its unloaded quality factor on Q .
4. If $(\omega_0 L_L / Q)$ is an acceptable value for input impedance, choose a value for C_1 which about is half the capacitance needed for resonance with L_L at ω_0 . Let C_L be the remaining capacitance needed for resonance.
5. If $(\omega_0 L_L / Q)$ is not acceptable, choose a value for C_1 which provides an acceptable value for $|Z_{IN}(\omega_0)|$, from equation 21. Choose a value for C_L which completes the requirement for resonance with L_L at ω_0 (equation 5).

This procedure should work well for most designs. It will provide the largest possible output level without sacrificing other parameters. Equations 8, 9, 12, 18, and 21 provide some quantitative insight into the behavior of signal tuned quadrature detectors. Although this analysis has been carried out for

capacitive coupling, it must be emphasized that a similar exercise could be performed for a circuit using an inductor in place of C_1 .

The Double Tuned Quadrature Detector

There is a higher performance version of the quadrature network which is often described in application notes (4,5). This is the double-tuned network which is shown in Figure 4. Much better linearity is claimed for this network.

This network uses a weakly coupled transformer having a tuned primary and a tuned secondary. In most implementations, both windings are identical. In Figure 4, both windings are shown to have an inductance, L . The amount of mutual inductance, M , is always a fraction of L . Thus:

$$M = kL, \quad k < 1 \quad (23)$$

where k is known as the coupling coefficient.

Instructions in various application notes indicate that the coupling coefficient, k , should be at least 70 percent of critical coupling, k_{cr} , where k_{cr} is defined in terms of the primary Q , Q_p , and the secondary Q , Q_s , as follows:

$$k_{cr} = \frac{1}{\sqrt{Q_p Q_s}} \quad (24)$$

Instructions also indicate that the primary is to be tuned for maximum signal, while the secondary is to be tuned for maximum linearity.

An analysis of the double tuned network will now be presented. This analysis will focus on the phase response of the network, as that is the important advantage of a double tuned network when compared with the single tuned network. An analysis of the amplitude response and the input impedance can be derived easily from this phase response analysis, and so will not be presented.

Some Approximations

The double tuned network is significantly more complicated and difficult to analyze than the single tuned network. Consequently, some assumptions and approximations will be used. These approximations are extremely plausible and serve to greatly enhance the understanding of the analysis. In fact, the analysis of the single tuned network would be greatly simplified by these approximations and would yield exactly the same result as equation 20.

The first assumption is simply that the maximum frequency deviation is very small. Thus, it is assumed that:

$$|\xi| \ll 1 \quad (25)$$

In comparison with terms of order 1, ξ can be ignored. Of course, one must be very careful to not ignore ξ when subtraction of large numbers will result in a term of the order ξ .

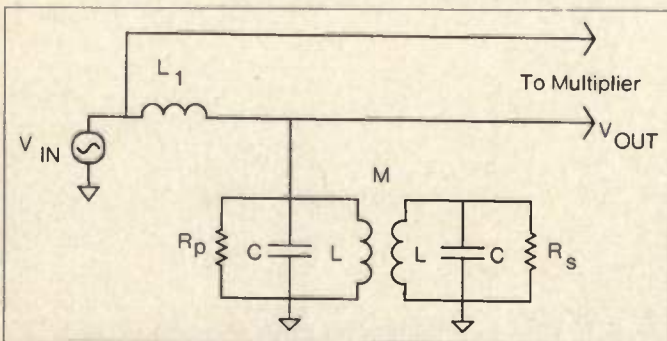


Figure 4. The double tuned quadrature network.

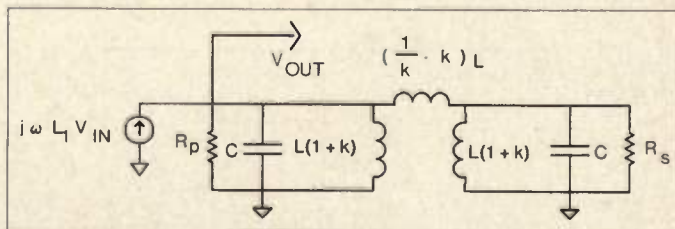


Figure 5. Equivalent of the double tuned quadrature network.

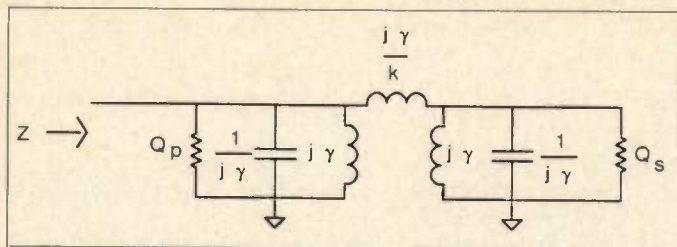


Figure 6. Approximation of the double tuned quadrature network.

Furthermore, it is assumed that the Q of both primary and secondary are large:

$$Q_s \gg 1, \quad Q_p \gg 1 \quad (26)$$

From these assumptions, and the definition of k_{cr} , it is clear that:

$$k \ll 1 \quad (27)$$

Equivalent Circuit

An equivalent circuit can be made for the double tuned network of Figure 4. This is shown in Figure 5. This equivalent will yield exactly the same output as the original circuit. The transfer function is determined once again by the input impedance of this network.

Only an analysis of the phase response of the double tuned network will be given here. Clearly the variation in the phase of the transfer function, about the 90 degree point, is given by the phase of the input impedance of the network. Just as with the single tuned network, the coupling element provides the initial 90 degrees of phase shift while the tuned network provides the deviation. Thus, the phase of interest is simply the negative of the phase of Y .

Figure 6 shows this input impedance as Z , with the impedance of all elements normalized to $\omega_0 L$. Simplification is provided by the use of equation 27. This simplified approximate network clearly consists of two networks in parallel. Figure 7 shows these two networks distinctly. Clearly, the admittance of the combined network Y , is simply the sum of the two admittances, Y_1 and Y_2 . The admittance, Y_1 , is just as useful for determining the phase response of the network as the impedance, Z , because Y and Z are reciprocals of each other. Thus, the variation component of phase response is simply the negative of the phase of Y .

Using equation 25, it is easy to show that, to a very good approximation sufficiently close to the center frequency, the admittance, Y_1 , is given by:

$$Y_1 = \frac{1}{Q_p} + 2j\xi \quad (28)$$

Similarly, the admittance, Y_2 , can be well approximated by:

$$Y_2 = k \frac{1 + 2jQ_s\xi}{Q_s(k - 2\xi) + j} \quad (29)$$

These forms for the admittances, Y_1 and Y_2 , assume that the tuned circuits of Figure 7 are tuned for resonance at $\xi = 0$. This is not necessarily the case. In fact, some degree of mistuning will yield the desired results. This mistuning can be represented by replacing ξ with $(\xi - \xi_1)$ in equation 28 and $(\xi - \xi_2)$ in equation 29. Thus, the total admittance, Y , is given by:

$$Y = \frac{1}{Q_p} + 2j(\xi - \xi_1) + k \frac{1 + 2jQ_s(\xi - \xi_2)}{Q_s(k - 2(\xi - \xi_2)) + j} \quad (30)$$

This detuning accounts for several effects. Slight differences in winding inductance, the loading effects of the coupling

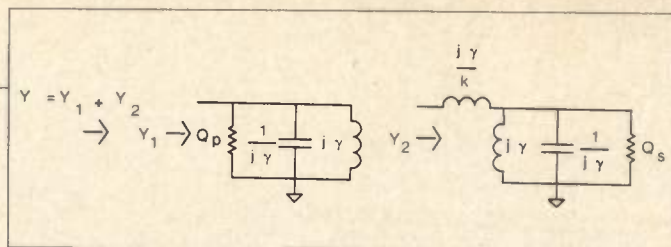


Figure 7. Simplified double tuned quadrature network.

element, L_1 , and the approximation made by replacing $L(1 + k)$ if Figure 5 with L in Figure 6 are a few of these effects. In general, the quantities ξ_1 and ξ_2 allow two more degrees of freedom for optimizing the transfer function.

The Phase Response

The trouble with the single tuned network is that once it has been tuned for nominal 90 degree phase shift, the transfer function is completely determined. Basically, the phase shift in a single tuned network is determined to be $\arctan(2Q\xi)$. There is no such limitation in the double tuned network. If the argument of the $\arctan()$ function could be made to more closely resemble the $\tan()$ function, a more linear phase relationship would be possible. This is exactly what the double tuned circuit allows. In order to see this more clearly, it is important to look at the Taylor expansion of the function $\tan(ax)$:

$$\tan(ax) = ax + \frac{a^3x^3}{3} + \frac{2a^5x^5}{15} + \dots \quad (31)$$

where a is some constant. For the purpose of this analysis, it is clearly desirable to have:

$$\frac{\text{Im}(Y)}{\text{Re}(Y)} = \tan(a\xi) \quad (32)$$

In practice, the best that can be obtained is an approximation to equation 32 for some range of ξ . If this relation can be approximated, the \arctan function performed by the phase detector would more closely reflect a linear relation with input frequency. Several requirements on Y are immediately clear:

$\text{Im}(Y)$ must vanish as ξ goes to zero. The ratio $\text{Im}(Y)/\text{Re}(Y)$ must have no even order terms. The coefficients of $(a\xi)$ should match those in the expansion of equation 31 through as many terms as possible.

The first two of the above requirements can be met by proper choice of ξ_1 and ξ_2 . Although not easy to derive, it is easy to verify that the choice of $\xi_1 = \xi_2 = -k/2$ satisfies both. Inserting this condition into equation 30 leads to:

$$\frac{\text{Im}(Y)}{\text{Re}(Y)} = 2\xi Q_p \left(\frac{(1 - k^2 Q_s^2)}{(1 + k^2 Q_s Q_p)} \right) \left(\frac{(1 + \frac{4Q_s^2}{1 - k^2 Q_s^2} \xi^2)}{(1 + \frac{4Q_s^2}{1 + k^2 Q_s Q_p} \xi^2)} \right) \quad (33)$$

This expression clearly has no terms in even powers of ξ . It also goes to zero as $\xi \rightarrow 0$. The polynomial quotient can be expanded to give the complete power series:

$$\frac{\text{Im}(Y)}{\text{Re}(Y)} = 2\xi Q_p \frac{(1 - k^2 Q_s^2)}{(1 + k^2 Q_s Q_p)} \left[1 + 4Q_s^2 \frac{k^2 Q_s (Q_s + Q_p)}{1 + k^2 Q_s (Q_p - Q_s) - k^4 Q_s^3 Q_p} \xi^2 + \dots \right] \quad (34)$$

The coefficient of the third order term in the Taylor expansion of $\tan(ax)$ can be matched by making:

$$\frac{1}{3} 4Q_p^2 \frac{(1 - k^2 Q_s^2)^2}{(1 + k^2 Q_s Q_p)^2} \quad (35)$$

$$= 4Q_s^2 \frac{k^2 Q_s (Q_s + Q_p)}{1 + k^2 Q_s (Q_p - Q_s) - k^4 Q_s^3 Q_p}$$

This defines a relation between k , Q_s , and Q_p . For example, if $Q_s = Q_p = Q$, the requirement of equation 35 can be found to be simply that $kQ = 0.327$. For $Q_s \neq Q_p$, the opportunity to make more terms in the expansion coincide, at least approximately, is available.

The data sheet for the RCA CA3189 (4) indicates that a preferred circuit uses $Q_s = 20$, $Q_p = 55$, and k is at least 70 percent of k_{cr} . Application of equation 35 leads to a value of 0.0229 for k . This is roughly 76 percent of k_{cr} , precisely as indicated by the data sheet. Although this choice does not give a good match for the term in ξ^5 , it gives a much better coefficient for that term than could be obtained from the choice of $Q_s = Q_p = Q$.

In general, for quadrature network design, the phase response, amplitude response, and input impedance are the important considerations. Of these quantities, the one most often overlooked is input impedance. It is clear from this analysis that as the output impedance of the limiting IF is lowered, its drive capability is increased. With proper design, a greater output signal from the quadrature network can be provided. Finally, the double tuned network can be used to greatly improve the linearity of the demodulator. Proper design of the double tuned network can be a bit tedious, as several

terms in the power expansion of equation 34 should be carefully evaluated. However, it is fairly easy to get a good match through the third order term.

Although the quadrature detector is a very popular device for the demodulation of FM, the literature contains very little information on its operation or design. This analysis provides the fundamental relations needed for the proper design of quadrature networks. The relations derived here provide the RF engineer with a means for optimizing the design of quadrature detectors. **RF**

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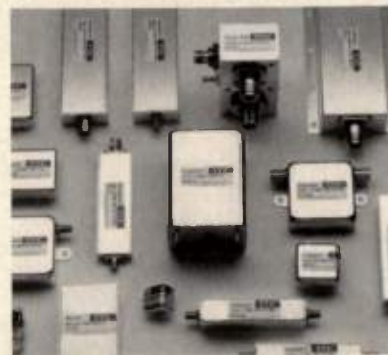
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CAD of Single and Tandem Lange Couplers

Timothy O'Connell and Patrick Murphy
University College Cork, Ireland

In recent times the 90 degree hybrid has become one of the most important passive elements to be used in microwave integrated circuits (MICs), because of its important function in balanced mixers, balanced amplifiers, and other circuits. Although there are many types of hybrids available that provide equal power division with 90 degree phase shift, the Lange coupler has been the most widely used because of its superior broadband performance. This article describes a CAD program which calculates the design parameters of Lange couplers. Since the main disadvantage of Lange couplers is their difficulty in fabrication due to the very small dimensions of the interdigitated fingers, the concept of combining couplers in tandem is presented, and this option is incorporated into the CAD program.

Since its introduction by Julius Lange in 1966 the Lange coupler (1) has been the most widely used 90 degree hybrid at microwave frequencies because of its very wideband -3 dB coupling over octave bandwidths. The structure of the device, shown in Figure 1 is relatively simple, with alternate interdigitated fingers coupled together using bond wires. For this structure:

S is the spacing between interdigitated fingers.

W is the width of the interdigitated fingers.

L is the width of the coupler; 1/4 wavelength at the center frequency.

h is the substrate thickness of the microstrip.

To date, the accepted design approaches are those reported by Presser (2) and Osmani (3), which are similar in nature and are suitable for CAD implementation. Their design approach may be summarized as follows:

The interdigitated directional coupler may be viewed as a multiconductor transmission line of K (K even) elements, not including the ground plane. However, coupling between non-adjacent sections may be neglected with little error. Using

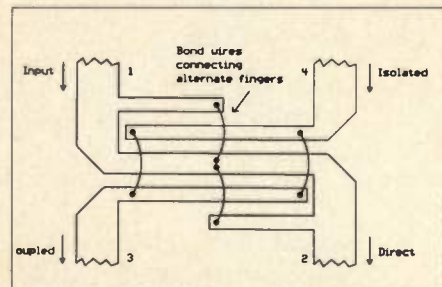


Figure 1. Diagram of a Lange coupler.

this simplification, Ou (4) derived the following set of equations for the even and odd mode impedances for such an array of parallel coupled lines.

$$Z_{oo} = Z_o \left(\frac{1 - C}{1 + C} \right)^{1/2} \frac{(K - 1)(1 + q)}{(C + q) + (K - 1)(1 - C)} \quad (1)$$

$$Z_{oe} = Z_{oo} \frac{C + q}{(K - 1)(1 - C)} \quad (2)$$

where,

K = even number of strips

Z_o = terminating impedance (usually 50 ohms)

C = voltage coupling coefficient

$$q = [C^2 + (1 - C^2)(K - 1)^2]^{1/2}$$

Next, the shape ratios are calculated from the synthesis equations of Akhtarzadm et al (5):

$$S/h = \frac{2}{\pi} \cosh^{-1} \left[\frac{\cosh\left(\frac{\pi}{2}(W/h)_{se}\right) + \cosh\left(\frac{\pi}{2}(W/h)'_{so}\right) - 2}{\cosh\left(\frac{\pi}{2}(W/h)'_{so}\right) - \cosh\left(\frac{\pi}{2}(W/h)_{se}\right)} \right] \quad (3)$$

$$W/h = \frac{1}{\pi} \cosh^{-1} \left[\frac{g + 1}{2} \cosh\left((W/h)_{se} \frac{\pi}{2}\right) + \frac{g - 1}{2} \right] - \frac{S}{2h} \quad (4)$$

where,

$$(W/h)'_{so} = 0.78(W/h)_{so} + 0.1(W/h)_{se} \quad (5)$$

and $(W/h)_{se,so}$ are given by Wheeler's equations for wide and narrow microstrip lines:

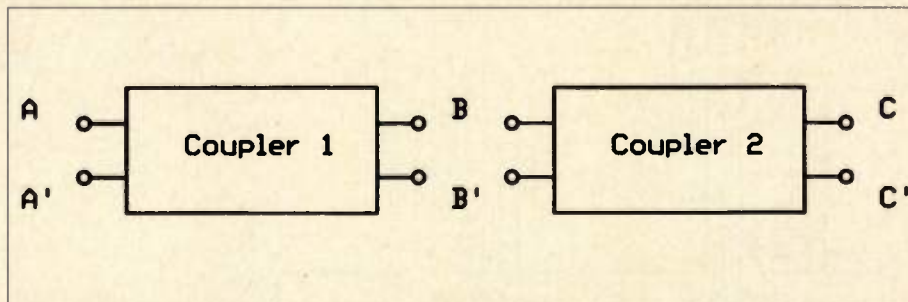


Figure 2. Couplers connected in tandem.

$$(W/h)_{se,so} = \frac{8}{p} \left[\frac{p(7 + 4/\epsilon_r)}{11.0} + \frac{11/\epsilon_r}{0.81} \right]^{1/2} \quad (6)$$

$$\text{where, } p = \left[e^{\left(\frac{Z_{0,so}}{84.8} \sqrt{\epsilon_r + 1} \right)} - 1 \right] \quad (7)$$

Tandem Lange Couplers

Although the design of Lange couplers is relatively straightforward using the above equations, their main disadvantage lies in the fact that the dimensions of the finger width (W) and spacing (S) are often so small that fabrication of the circuit is difficult. Typical dimensions for Lange couplers are of the order of 10 μm , and this can pose difficulties unless the fabrication process has very good tolerances. A novel solution is to combine two Lange couplers in tandem, so that the overall coupling coefficient is higher than the individual coupling coefficients. Therefore, since the dimensions are inversely related to the value of the coupling coefficients, the etching constraints are greatly reduced.

Consider two couplers connected in tandem as shown in Figure 2. Assuming the signal level at input A is 1, and is 0 and A', and that the coupling coefficient of each section is k, the signal level at the other ports are:

$$B = k \quad (8)$$

$$B' = j \sqrt{1 - k^2} \quad (9)$$

$$C = k^2 - (1 - k^2) \quad (10)$$

$$C' = j2k \sqrt{1 - k^2} \quad (11)$$

Here we can see that the 90 degree phase shift between C and C' is still preserved. Now for -3 dB power division, signals at C and C' must be equal. Therefore,

$$K^2 - (1 - k^2) = 2k \sqrt{1 - k^2} \quad (12)$$

Defining $l = k^2$ yields:

$$8l^2 - 8l + 1 = 0 \quad (13)$$

By solving this equation and discarding one of the solutions, we get $k = -8.34$ dB. Thus, to achieve an overall coupling factor of -3 dB, each of the two couplers connected in tandem

must have a coupling factor of -8.34 dB.

Therefore, two Lange couplers with coupling factors of -8.34 dB can be combined to make a -3 dB coupler with the same overall characteristics as a single coupler. The more loosely coupled individual sections will relax the etching constraints by increasing the dimensions of the coupled fingers. Figure 3 shows two Lange couplers connected in tandem.

The CAD Program

A FORTRAN CAD program has been written which enables the user to design Lange couplers using the equations above. Single or tandem configurations can be selected, depending on the dimensions obtained and the quality of the etching process available. Figure 4 shows the simulated Touchstone® response of the single and tandem couplers. While the response of the single coupler is marginally better, in some applications this slight deviation is much less than the discrepancy in results (with a single coupler) due to the etching tolerances of the process available.

Finally, the CAD program gives the user the option of producing an output file directly in the format of a Touchstone input file. This allows great flexibility in incorporating the Lange coupler into a larger circuit, or allows the mask of the circuit to be generated directly from software such as MICAD® or GAS-STATION®.

The operation of the program TLANG is described by the flowchart in Figure 5. First, the design parameters are entered; substrate dielectric constant, substrate thickness, center frequency, and overall coupling coefficient. If a tandem arrangement is selected, the program then calculates the coupling coefficient of each section. Next, the dimensions of the Lange coupler are calculated using the Osmani/Presser method, and finally, the results are produced either directly to the screen, or in the format of a Touchstone input file. Figure 6 shows a typical output of the TLANG program. Here a six finger tandem Lange coupler with a center frequency of 4 GHz has been designed on 5880 RT-Duroid® (0.79 mm substrate thickness).

The program is available from the RF Design Software Service, including a compiled version and source code listing. See page 30 for more information. **RF**

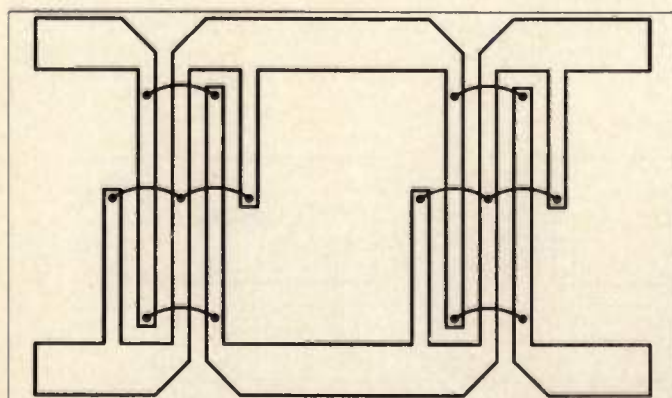


Figure 3. Tandem Lange coupler arrangement.

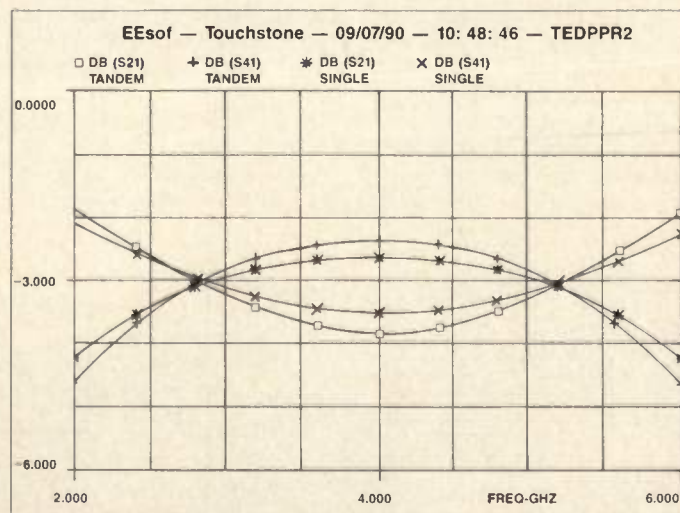


Figure 4. Simulated results of a single and tandem Lange coupler.

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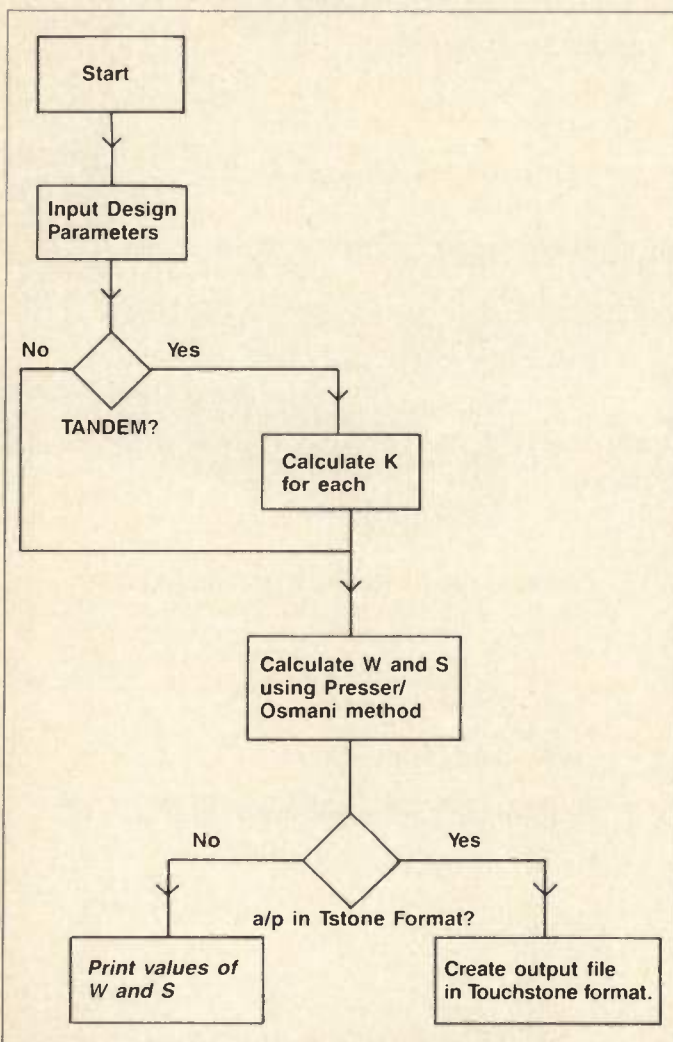


Figure 5. Flowchart of the TLANGE program.

```

DIM
FREQ GHZ
RES OH
IND NH
CAP PF
LNG MM
TIME PS
COND /OH
ANG DEG
  
```

```

CKT
MSUB ER= 2.20 H=0.79 T=.005 RHO=1 RGH=0
MLANG6 1 2 3 4 W=.253 S=.406 L=13.500
MLANG6 2 6 4 7 W=.253 S=.406 L=13.500
DEF4P 1 6 3 7 LNGE
  
```

```

OUT
LNGE DB[S21] GR1
LNGE DB[S41] GR1
  
```

```

FREQ
SWEEP 2.5 4.0 5.5
  
```

```

GRID
RANGE 2.5 4.0 5.5
GR1 0 -20 2
  
```

Figure 6. Touchstone file output from the TLANGE program.

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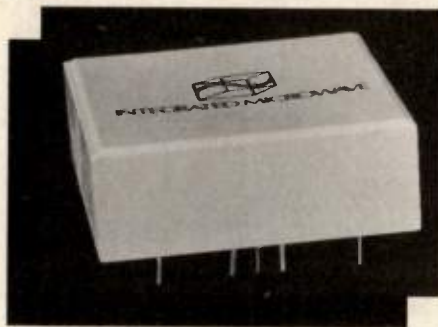
By Liane G. Pomfret
Associate Editor

The market for RF filters is stable and growing with minor variations that can be linked to the current recession. As usual, the commercial market is enjoying greater success than the military market. Despite a worldwide economic slump, some of the best business is being generated internationally and has buoyed relatively flat domestic sales. Surface mount technology is also contributing to continued profits. With a host of new applications and new technology, there's little to stand in the way of a successful future for RF filters.

It comes as no great surprise that the military market for RF filters has suffered because of congressional cutbacks. While the cutbacks have not been disastrous, they have been noticeable and companies involved in military have felt the pinch. For companies who focus on the commercial market, business is much better. The recent explosion of commercial RF applications has opened up new business opportunities. The list of applications — wireless communications, PCNs, mobile and land mobile, data transfer, medical systems, and satellites — continues to grow. While much of this technology was developed for the military, it has found success within the commercial arena, making it doubly profitable for the manufacturers.

Not all of the technology evolved from military contracts. Some developments came from the commercial sector. Surface mount and miniaturization of RF filters are the results of requests for smaller commercial devices — the 2 pound cellular phone as opposed to the 7 pound cellular phone. Like much of the RF industry, surface mount technology is now working its way into the filter industry. Yet the military still lags behind. Arie Verhagen, Vice President of Marketing for Murata-Erie comments, "The industry as a whole, has to do their homework to bring surface mount to the military's attention." Qualifying new technology for military specifications is a long, expensive process and the industry is reluctant to invest time and money without assurances that the technology will be accepted.

In the commercial sector, suppliers



A switched filter module from Integrated Microwave.

have been reducing their prices while maintaining or improving quality. As a result, manufacturers have more latitude to experiment with manufacturing techniques and materials. Steve Sodaro, Vice President of Marketing and Sales at TTE notes, "Components now are not as costly. It affords us the opportunity to do miniaturization. We're always trying to miniaturize without sacrificing performance." However, while a large firm has the resources to experiment with technologies such as surface mount, smaller companies are often restricted because of the lack of available packaging technology within their company. Packaging technology is available from outside vendors but cost and availability are a stumbling block to in-house development.

The communications industry offers a large number of uses for filters. Voice communications, data transfer and acquisition, LANs, and voice over data are all demanding more sophisticated filtering techniques. "The applications for filters in general are increasing, mainly due to enhancements in telecommunications, LANs, and various radio systems including cellular and land mobile beyond the classic military market," says Dave Distler, Director of Sales and Marketing for Trilithic. Manufacturers are demanding extremely sophisticated filters capable of high reliability, high performance filtering.

Every company has its own area of specialty and responses will vary regarding the active areas for filter products. However, the general consensus is that communications equipment is the hot area right now. Randy Rhea, President of Eagleware notes, "We see a definite

increase in interest in filters for data communications systems and that often results in a greater emphasis in the delay characteristics of filters." Al Lowenstein, National Sales Manager, for Motorola's RF filter products components division, sees interest for their products in other areas as well. "The new activity is in the 915 MHz range, especially spread spectrum, new generation portable telephones, and remote data acquisition, meter reading and things along that line." He adds, "The other thing that is taking off is GPS and the satellite industry in general." Wavetek's focus has been high reliability communications such as those found in avionics communications systems. Even with the slowdown in the domestic military market, Sara Mussman, Components Business Unit Manager for Wavetek comments, "I see a significant amount of business out there."

A lot of this business is coming from the international sector. Mussman notes, "Any country that is aggressively pursuing an upgrade will be a good source of business." For George Szentirmai, President of DGS, business has been fairly flat. "The only thing is that our international sales are pretty strong especially in Europe." His company has had a great deal of interest from eastern European countries, but since there's no money available, there's been no sales. Werner Mueller, Vice President of KVG, North America comments, "I would hope that due to the reunification of Europe and due to the communication lack in eastern Europe, that there will be an increase in the European market." Toyocom's market situation is different. They currently hold less than 10 percent of the U.S. market and are looking to expand within the United States according to Dennis Reifel, Marketing Manager at Toyocom.

With the majority of the RF filter business on the upswing, the outlook for the future is optimistic. The advent of surface mount technology and constant improvements in price and performance make filters an excellent prospect for profit in the coming year. **RF**

For reprints of this report, contact Cardiff Publishing Company at (303) 220-0600; ask for the Circulation Department.

SPICE Transistor Library

A comprehensive library of RF transistor SPICE models is available from Intusoft. Models include 36 popular bipolar and JFET transistors, usable with any Berkeley SPICE compatible simulator. The models use a subcircuit approach that accounts for package parasitics, and matches published S-parameter magnitude and phase data up to 5 GHz. The library is priced at \$99, and is available in PC and Macintosh formats.

Intusoft
INFO/CARD #233

Dielectric Resonators

Two programs with easy data input are available for analyzing dielectric resonators. The program DRESV3 computes the resonant frequency and the Q factor of the TE_{01d} mode. Program FOAM3 estimates the resonant frequencies of sixteen higher-order modes. Diskettes for the IBM PC or compatibles are priced at \$25.

Kajfez Consulting
INFO/CARD #235

Version 3.0 Introduced

EEsof announces the new Version 3.0 product suite, upgrading their line of high-frequency software tools. Version 3.0 links all

of EEsof's simulators under a single design environment; linear and nonlinear frequency domain, nonlinear transient SPICE analysis, and system/subsystem analysis. Its graphical design environment provides block diagram and schematic entry, multi-simulator control, documentation, and optional IC and board layout capability. Version 3.0 runs on UNIX-based workstations and PCs using the OS/2 operating system.

EEsof, Inc.
INFO/CARD #234

Software Catalog

A new eight page catalog presents simulation and synthesis CAE tools from Webb Laboratories. Included are the SysCad and Receiver Advantage system simulators, FilSolv and TransCad for filter and transmission line structure synthesis, and SANA for microwave network signal and noise analysis and optimization.

Webb Laboratories
INFO/CARD #232

Graphics-Based Analysis

DADISP is a visually-oriented graphics software package designed for scientific and technical applications. Among the 200 data processing functions are signal arithmetic and editing, waveform generation, FFTs, and

peak-finding. Extended memory support is an option, as is an IEEE-488 bus driver and an software module to interface with popular data acquisition boards. Versions are available for a wide range of PCs, workstations and larger computers.

DSP Development Corp.
INFO/CARD #231

SPICE Library for Linear ICs

SPICE macromodels are now available for nearly every Burr-Brown op amp, difference amp, and instrumentation amplifier. Two levels are offered: PSpice[®] Parts[™] and Enhance Parts[™] macromodels. A 5 1/4 inch diskette with more than 75 models is free from Burr-Brown.

Burr-Brown
INFO/CARD #230

Thermal Analysis

Version 1.33 of SAUNA, a 3-D thermal analysis package has been released by Tatum Labs. Larger models are possible; up to 700 nodes and 2750 thermal resistors. Reduced air density due to altitude is also accounted for. SAUNA runs on IBM PC compatible and Macintosh II and SE/30 computers.

Tatum Labs, Inc.
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RF/Microwave Product Guide

Avantek offers a 48-page Product Selection Guide to give engineers a quick overview of the company's standard products. Summary specifications are provided for transistors, integrated circuits, amplifiers, signal processing and control products, LNAs, uplink amplifiers, cellular base amplifiers, and YIG oscillators. Most items are in stock at Avantek distributors.

Avantek, Inc.
INFO/CARD #250

Ceramic Filter Catalog

Murata Erie announces a design manual and catalog for their line of ceramic filters for communications. Single and multiple element filters are listed, covering standard frequencies such as 455 kHz, 3.58 MHz, 4.5 MHz and 10.7 MHz. Ceramic discriminators and signal detectors are also included, as are surface mount filters. Application information includes theory of operation and circuit design considerations.

Murata Erie North America
INFO/CARD #249

Frequency Control Products

Comprehensive data on products available from Piezo Technology, Inc. is included in a new 96-page handbook. Monolithic and discrete crystal filters, LC filters, cavity band-pass filters, combline filters, interdigital filters, crystal and non-crystal oscillators, precision crystal resonators, RF subassemblies, and custom hybrid circuits are listed.

Piezo Technology, Inc.
INFO/CARD #248

RF Data Book

Motorola's RF Products Division has released their new RF Data Book, DL110/D Rev. 4, a two-volume edition covering power FETs, power bipolar and small-signal transistors, plus RF power modules. 84 new products have been included since the previous version, along with new applications information.

Motorola, Inc.
INFO/CARD #247

Coaxial Switches

A new Standard Products Brochure is available from Dow-Key Microwave, listing standard electromechanical coaxial switches and relays in configurations from SPDT to SP12T. Transfer switches, bypass switches and low cost RF relays are also included. Photographs, specifications and schematic diagrams are provided for these products.

Dow-Key Microwave Corp.
INFO/CARD #246

1991 Test Instruments

Tektronix' 1991 customer catalog is now available, highlighting more than 3000 products in three broad areas: electronic test and

measurement instruments, professional broadcast equipment, and computer peripherals. New products include communication signal analyzers, VXIbus products, high resolution display monitors, HDTV test equipment and low cost broadcast spectrum analyzers.

Tektronix, Inc.
INFO/CARD #245

Coatings Brochure

A line of coating and ink systems is described in a new brochure from Emerson & Cuming. One and two-component systems are presented, used in making membrane switches, molded and flexible circuits, polymer thick film circuitry, and for modifying circuit boards. The coatings provide protection against electrical contact during component assembly.

Emerson & Cuming, Inc.
INFO/CARD #244

Frequency Chart Update

The RF Products and Linear Integrated Circuits Divisions of Motorola have published a new U.S. frequency allocation chart covering the RF spectrum up to 4000 MHz. Added to the previous version's allocation categories is a segment for Part 15 unlicensed devices. The chart is free, and is suitable for convenient wall mounting.

Motorola, Inc.
INFO/CARD #243

SAW Filter Note

A new Application Note discussing SAW coupled-resonator (CR) filters is available from RF Monolithics. SAW CR filters are suitable for many narrow-band applications from 70 MHz to 1500 MHz, and can be built with fractional bandwidths from 0.03 to 0.5 percent. The filters exhibit low loss (3 dB typical) under matched conditions. The note covers theory, operational performance limits, and a comparison with other filter technologies.

RF Monolithics, Inc.
INFO/CARD #242

Reconditioned Equipment

A catalog featuring reconditioned Philips/Fluke test and measurement equipment is announced by Rubytron. Oscilloscopes, synthesizers, frequency counters/timers, sweepers, function generators and pulse generators are among the equipment types available. All instruments are calibrated and warranted.

Rubytron Instruments
INFO/CARD #241

Distributor Catalog

Tektronix 1991 Distributor Products Catalog contains data on analog and digital oscilloscopes, frequency counters, digital multimeters and function generators. These products are offered through authorized distributors.

RAG Electronics, Inc.
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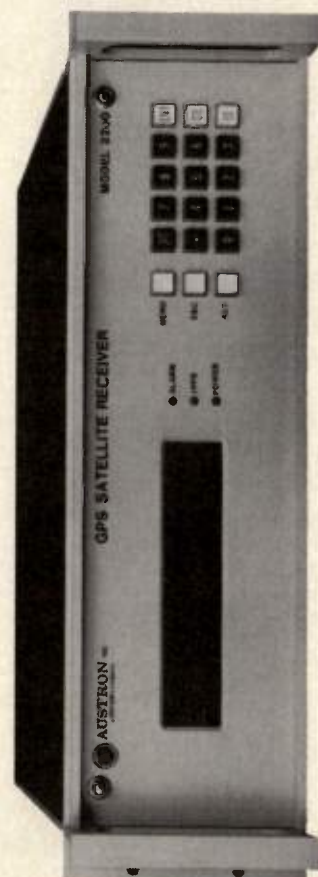
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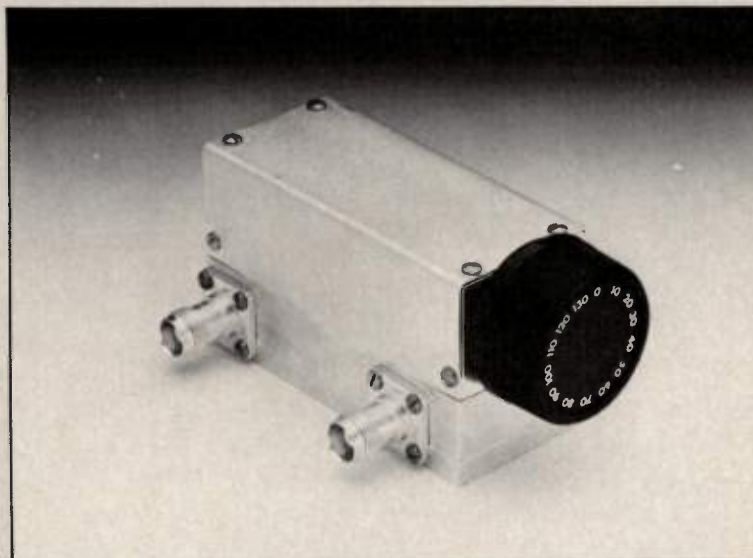
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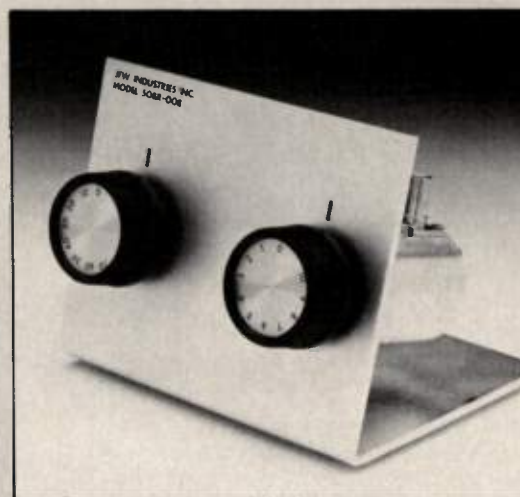


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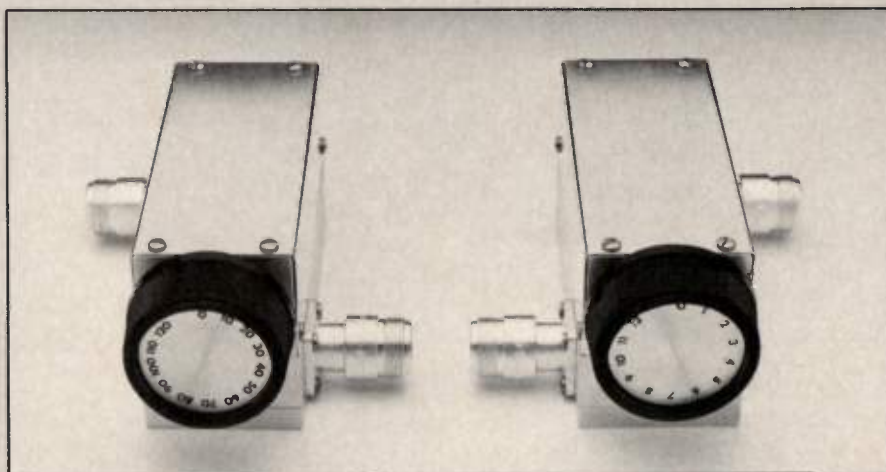
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50R-080 (right)

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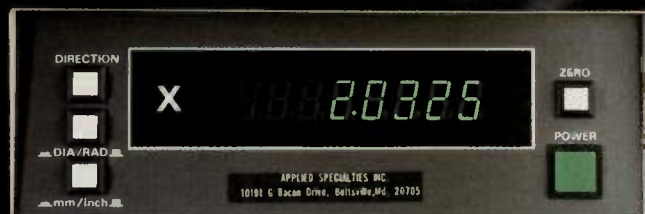
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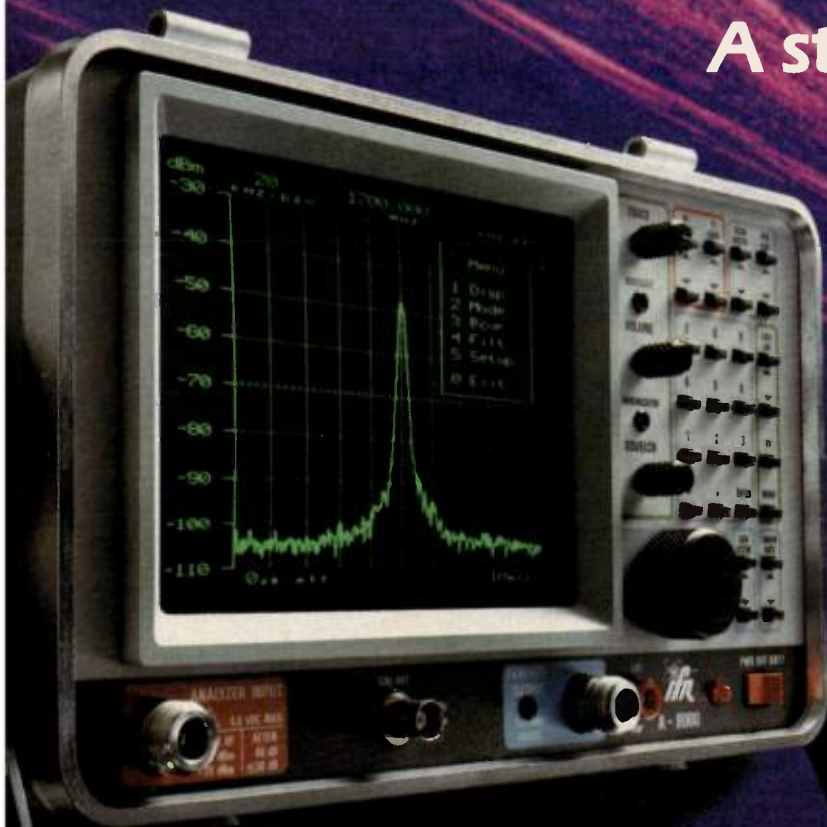
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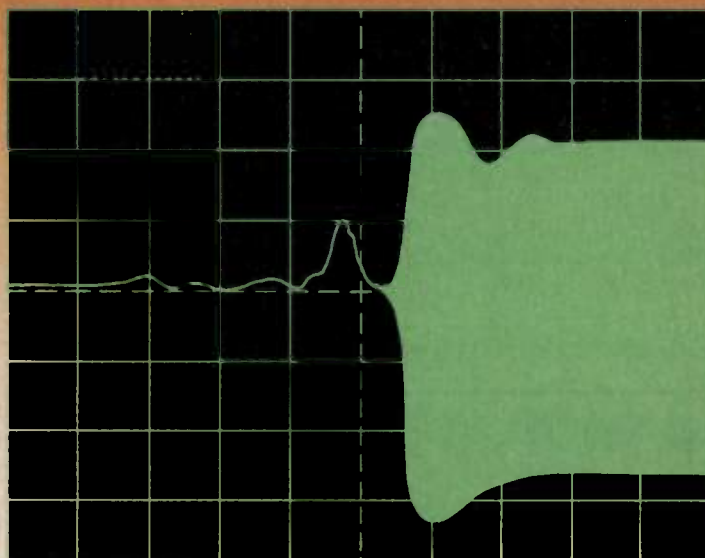
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featured technology

31 Designing a Direct Sequence Spread Spectrum Secure Communication System

Design considerations for a direct sequence spread spectrum system are described in this article, along with software simulation of system data error performance in a jamming or interference environment.

— Soo Kwan Eo, Chingwo Ma, Incheol Jang



40 IQ Sampling Yields Flexible Demodulators

The author describes a method which minimizes errors by using a single analog-to-digital converter in a front-end for a DSP-based system. The ADC functions as a multiplexed harmonic mixer and digitizer, with demultiplexed I and Q channel outputs.

— Thomas P. Hack

emc corner

65 Communications Range and Reliability of Part 15 Devices

Low power, unlicensed radio devices operating under FCC Part 15 have significant constraints on their performance. This article identifies areas of difficulty and means for obtaining desired performance.

— Bernard Kasmir

design awards

69 Parasitic Positive Frequency Acquisition in a PLL

This 1990 contest entry describes a method for dealing with frequency acquisition, one of the limitations of double balanced mixer phase detectors.

— Jonathon Y.C. Cheah

72 Biasing Solid State Amplifiers to Linear Operation

This note is a review of requirements for bias voltage and current, and their regulation and temperature compensation in linear bipolar and FET power amplifiers.

— H.O. Granberg

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Our cover, provided by Burr-Brown, represents increasing use of digital technology in RF applications. In the photo is the ZPA1000 Dynamic Signal Analysis software for digitized signals. Available digitizers and buffers will handle signals up to 10 MHz.

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RF editorial

Engineering Productivity Part I: The Team Concept



By Gary A. Breed
Editor

What does it take to be successful in an increasingly competitive marketplace? Usually, it means getting more done with less — fewer employees, lower manufacturing costs, lower overhead, plus fewer returned goods and service calls. In the next few months, I'll be commenting on some of the techniques that have been developed to meet these challenges, identifying the role of the engineer in each of them.

To meet the overall goal of increased productivity, many companies are adopting new operating methods. You have probably heard some of the terminology these methods use: world class, just in time, design for manufacture, integrated management, and total quality management. All these concepts have the single goal of creating a quality product without wasted time or material.

What these methods feature is involvement from start to finish by participants in the product's design, manufacture, costing, sale and service. Perhaps a management committee has been formed with representatives from the different departments. Or, an individual project manager might oversee cross-checking and review by the various departments. In either situation, this process should allow operating features and performance to be optimized for customer requirements, serviceability, and the most economical manufacturing processes. Putting this concept into practice can be extremely difficult, because it usually means a major change in the way a company operates.

I have to note that these aren't really new ideas. Consider a very small company made up of a few partners who want to develop an idea into a product. Every decision is made by the partners,

who represent every company function. Design tradeoffs are examined, costs evaluated, manufacturing methods explored, and plans are developed based on the group's consensus.

Implementing this cooperative process on a larger scale is the new part, and it presents some problems for those involved (such as design engineers). First, in a large company the participants do not have the urgency of a financial commitment like entrepreneurial partners. Also, there may be difficulty adjusting after many years of working on pieces of projects with no attention to overall goals. Another common problem is that each department feels it has to protect its "turf."

These problem areas point out how hard it is to make fundamental changes in the way a company operates. However, these changes will continue to take place, mandated by the need to achieve the highest possible efficiency. We can't (and in most cases, shouldn't) fight these changes, so we must adapt. In the next couple of editorials, I'll comment on some of the areas where engineers will find changes. I will also point out some areas where engineers can benefit significantly from their employers' commitments to improve competitiveness.

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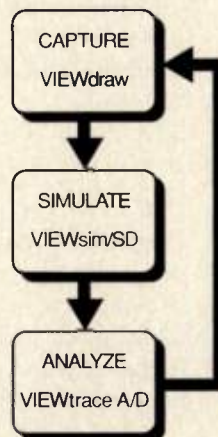


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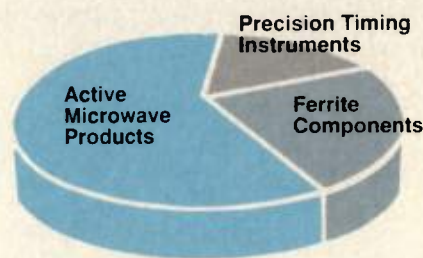
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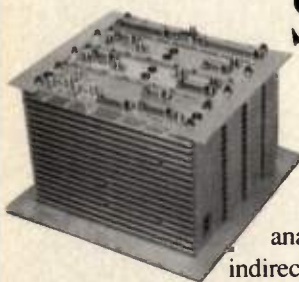


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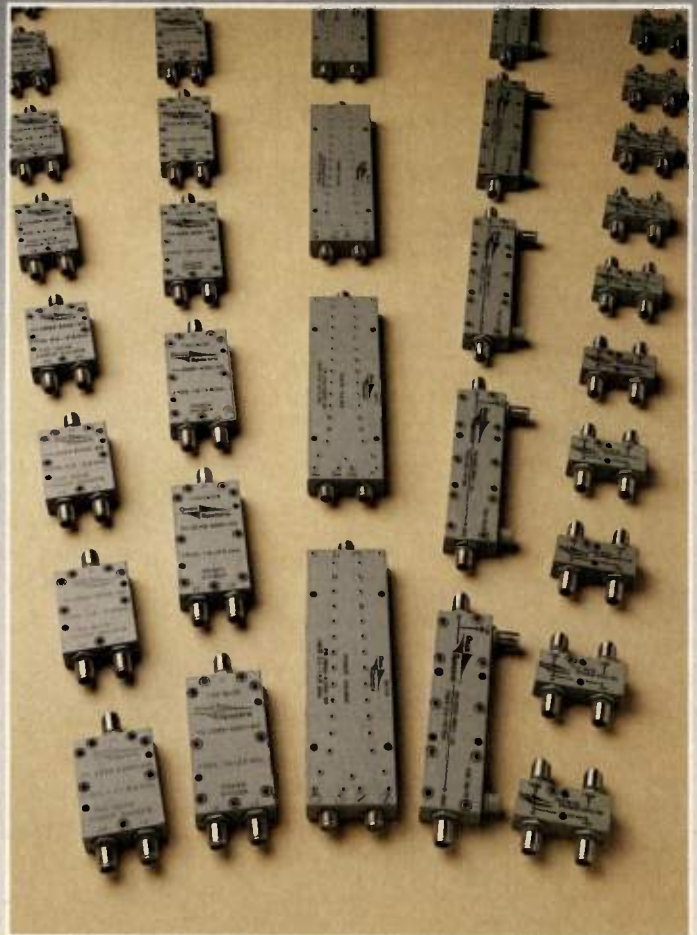
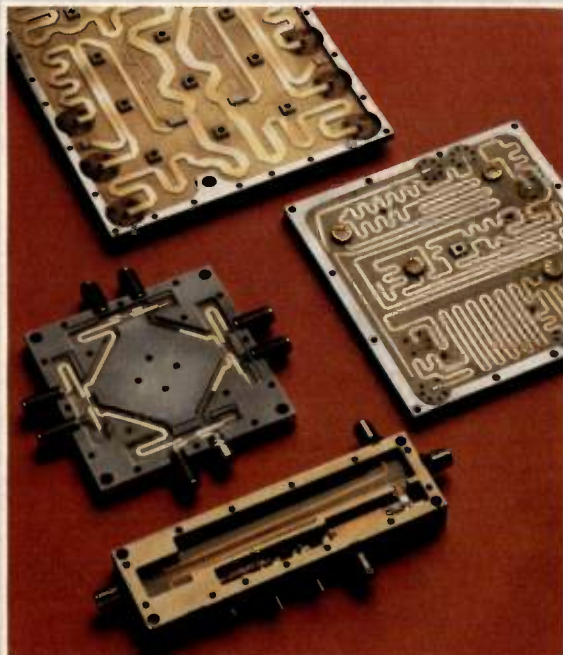
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Driftless VCO Comments

Editor:

The "Linear Driftless VCO" in the January 1991 issue of *RF Design* has a fatal flaw. The "frequency comparator" is a linear circuit and cannot actually compare frequency. Instead it compares the average DC levels of two variable duty cycle pulse streams.

If the 74HC74 flip-flops used as one shots and 4016s connected to the Q outputs are replaced with a true digital frequency comparator, such as phase comparator 2 in the 74HC4046, the circuit should work as described. This should bring the frequency drift down from the 5.2 kHz measured by the author to within calculated limits.

Jon Nelson
Uniplex Corporation
St. Paul, MN

The Author's Reply

Editor:

The frequency comparator used, as said by Mr. Nelson, "compares the average DC levels of two variable duty cycle pulse streams," but each pulse stream has as many pulses as the input frequency, so its duty cycle is proportional to the input frequency and the output will have an average perfectly proportional to frequency difference (with an offset of $V_{cc}/2$).

Actually, it does compare frequency:

$$AVG(V_{out}) = k(f_1 - f_2) + V_{cc}/2$$

As stated in the article, the design needs a frequency detector that keeps within the linear region of operation at any input frequency, to produce an error signal proportional to the difference of frequencies. Thus, a mean can be extracted and the VCO will remain locked on that mean. The suggestion to use a type 2 phase comparator such as the 4046 will fail to work.

At any instant the frequency comparator is dealing with two different frequencies, because the reference is switching between two different values and the prescaled VCO (assumed as stable) will be in between. A type 2 phase comparator is within the linear region of operation only at equal input frequencies. At different input frequencies, the phase comparator will usually saturate.

In a conventional PLL this will allow to search for the lock. In our case it produces a square wave that is a replica (but not exactly equal) of the multiplexing signal, because we are switching between one frequency above and one below the prescaled one. This way the circuit seems to work but in fact it is not locked. This circuit with a 4046 was tested in early stages of development, without success.

Luis Cupido
C&TC
Aveiro, Portugal

Comb Generator Source

Editor:

I was very interested in the article which appeared in the January issue's EMC Corner, "RFI Measurements Using a Harmonic Comb Generator."

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Two-Channel Receiver Correction

An alpha (α) was omitted from Figure 1 in "A Simple Two-Channel Receiver for 8-PSK," February, 1991, p. 29. The corrected figure is shown below. In addition, in Figure 3, an epsilon (ϵ) was omitted from the summation circle on the right side of the figure.

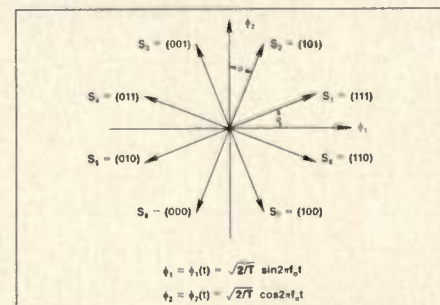
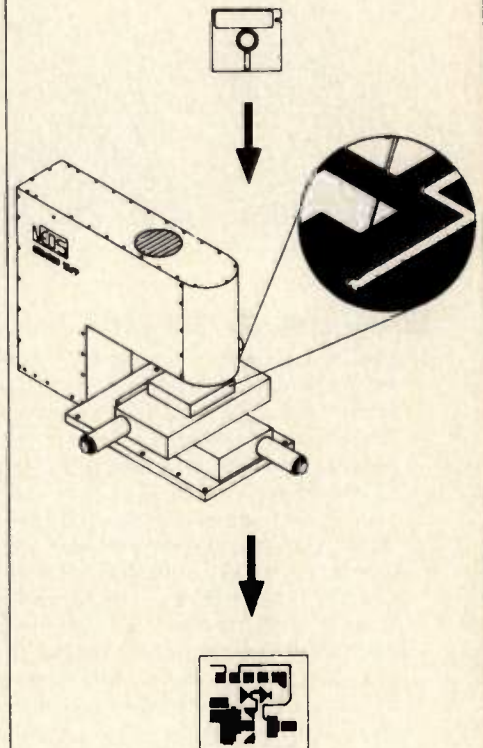


Figure 1. Signal space diagram showing gray level coding of symbols.

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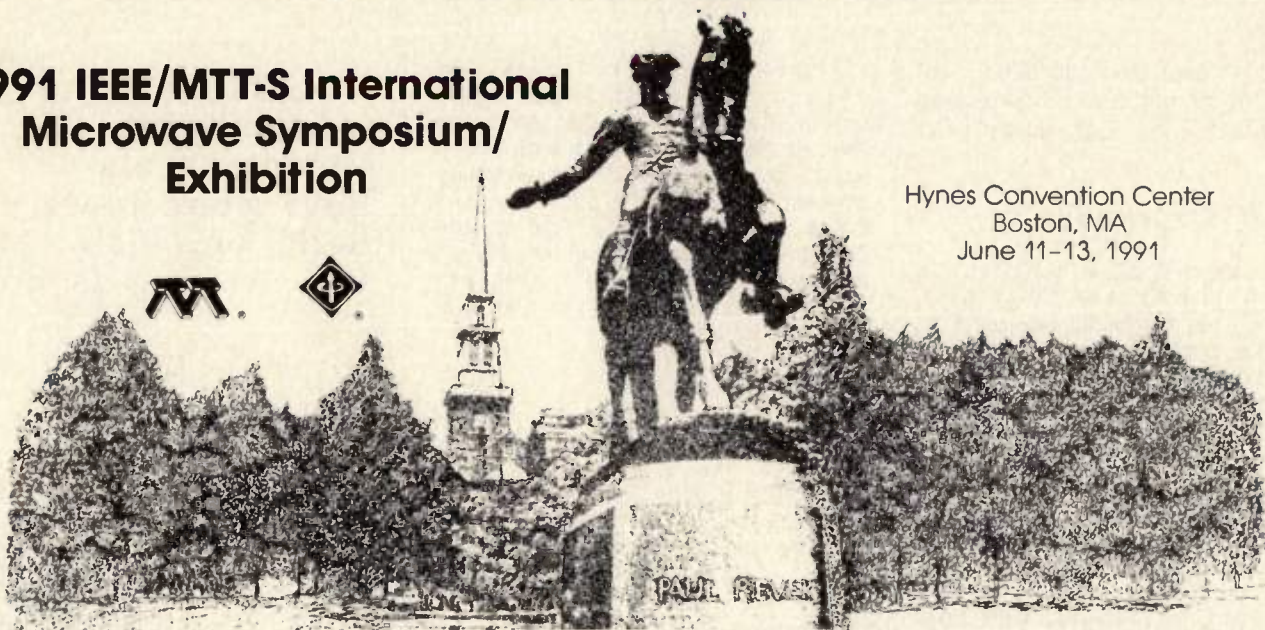
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San Diego Convention Center, San Diego, CA
Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616.
- 15-18** **Seventh International Conference on Antennas and Propagation**
University of York, UK
Information: Conference Services, IEE, Savoy Place, London, England WC2R 0BL. Tel: 44 071 240-2871 ext 222.
- 15-18** **NAB '91**
Las Vegas Convention Center, Las Vegas, NV
Information: NAB, 1771 N Street, N.W., Washington, DC 20036-2891. Tel: (202) 429-5350.
- 16-18** **Electro/International**
Jacob K. Javits Convention Center, New York, NY
Information: Electronic Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045-3194. Tel: (213) 772-2965.
- 28-1** **The 1991 MIC Workshop**
Kingsmill Resort & Conference Center, Williamsburg, VA
Information: Arlon, Microwave Materials Division, 1100 Governor Lea Road, Bear, DE 19701. Tel: (800) 635-9333.

May

- 13-15** **41st Electronic Components and Technology Conference**
Westin Peachtree Plaza, Atlanta, GA
Information: Electronic Industries Association. Tel: (202) 457-4930.
- 14-16** **IEEE Instrumentation and Measurement Technology Conference**
Omni Hotel, Atlanta, GA
Information: Robert Myers, 3685 Motor Avenue, Ste. 240, Los Angeles, CA 90034. Tel: (213)-1463. Fax (213) 287-1851.
- 29-31** **45th Annual Symposium on Frequency Control**
Los Angeles Airport Marriott, Los Angeles, CA
Information: Annual Symposium on Frequency Control, PO Box 826, Belmar, NJ 07719.

June

- 3-5** **First Virginia Tech Symposium on Wireless Personal Communications**
Virginia Tech, Blacksburg, VA
Information: Prof. Theodore S. Rappaport, Director, Mobile & Portable Radio Research Group, Virginia Polytechnic & State University, Blacksburg, VA 24061-0111.
- 11-13** **MTTS International Microwave Symposium and Expo**
Boston, MA
Information: Tel: (617) 769-9750.
- 13-14** **ARFTG**
Boston, MA
Information: ARFTG, c/o Henry Burger, 1061 E. Frost Drive, Tempe, AZ 85282. Tel: (602) 839-6933.

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ESD - Electrostatic Discharge

April 17-19, 1991, Madison, WI
Information: University of Wisconsin - Madison, Department of Engineering Professional Development. Tel: (608) 262-2061. Fax: (608) 263-3160.

Fundamentals of Radar Cross Section

May 20-24, 1991, San Diego, CA
Information: Kelly Brown, SCEE, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146.

Microwave/Millimeter Wave Monolithic Integrated Circuits

June 4-7, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Infrared/Visible Signature Suppression

April 16-19, 1991, Atlanta, GA
Information: Education Extension, Georgia Institute of Technology. Tel: (404) 894-2547.

Analog/RF Fiber-Optic Communications

April 24-26, 1991, Washington, DC

Microwave Systems Engineering

April 29-May 3, 1991, Washington, DC

Vulnerability of Spread Spectrum AJ and LPE Communication Systems

May 6-9, 1991, Washington, DC

Nonlinear Digital Signal Processing and Applications

May 6-10, 1991, Washington, DC

Electromagnetic Interference and Control

May 6-10, 1991, Washington, DC

Frequency Hopping Signals and Systems

May 20-22, 1991, Washington, DC

New HF Communications Technology: Advanced Techniques

June 3-7, 1991, Washington, DC

Spread Spectrum Communications Systems

June 10-14, 1991, Washington, DC

Information: The George Washington University, Continuing Engineering Education, Merrill A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Fiber Optic Communications

June 10-12, 1991, Sunnyvale, CA
Information: Center for Professional Development, Arizona State University. Tel: (602) 965-1740.

Digital Signal Processing Workshop

June 12-14, 1991, Norwood, MA
Information: Analog Devices, DSP Applications Department, Maria Butler. Tel: (617) 461-3672.

Design Seminar: Principles of EMC

May 7-9, 1991, Mariposa, CA
Information: CKC Laboratories, Registrar. Tel: (209) 966-5240. Fax: (209) 742-6133.

Worst Case Circuit Analysis

May 13-15, 1991, Washington, DC
Information: Design and Evaluation, Inc. Tel: (609) 228-3800.

Academy: Schematic/Layout

April 16-19, 1991, Westlake Village, CA

May 7-10, 1991, Baltimore, MD

Touchstone/Academy

April 15-19, 1991, Baltimore, MD

Libra/Academy

April 22-26, 1991, Westlake Village, CA

April 22-26, 1991, Gilching, Germany

Information: EEsof. Tel: (818) 991-7530. Fax: (818) 991-7109.

RF and Microwave Design II: Non-Linear Circuits

April 22-26, 1991, Pisa, Italy

Digital Microwave Systems: Theory and Applications

April 22-25, 1991, Pisa, Italy

Satellite Communication and Broadcasting

April 22-26, 1991, Italy

Frequency-Time and Spatial-Time Signal Processing

June 10-14, 1991, United Kingdom

Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

RF/MW Linear/Nonlinear Circuits and Applications

May 2-8, 1991, Washington, DC

Information: Besser Associates. Tel: (415) 949-3300, Fax: (415) 949-4400.

Modern Power Conversion Design Techniques

April 29-May 3, 1991, Phoenix, AZ

May 20-21, 1991, San Rafael, CA

Information: e/j Bloom Associates, Joy Bloom. Tel: (415) 492-8443. Fax: (415) 492-1239.

EW Receivers

May 7-9, 1991, Syracuse, NY

ELINT Analysis

May 7-9, 1991, Syracuse, NY

ELINT Interception

May 14-16, 1991, Syracuse, NY

ELINT/EW Applications of Digital Signal Processing

May 14-16, 1991, Syracuse, NY

Integrated EW

May 21-22, 1991, Syracuse, NY

ELINT/EW Data Bases

May 21-23, 1991, Syracuse, NY

Radar Vulnerability to Jamming

June 4-5, 1991, Syracuse, NY

Electromagnetic Propagation

June 4-6, 1991, Syracuse, NY

Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Basic Network Measurements Using the 8510B Network Analyzer

April 30-May 2, 1991, Los Angeles, CA

May 14-16, 1991, Boston, MA

May 29-31, 1991, Los Angeles, CA

Microwave Fundamentals

May 14-17, 1991, Los Angeles, CA

Information: Hewlett-Packard Company. Tel: (714) 999-6700.

Design for EMC

April 22-23, 1991, Philadelphia, PA

Practical EMC Retrofits

April 24-26, 1991, Philadelphia, PA

Information: R&B Enterprises. Tel: (215) 825-1960.

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45th Annual Symposium on Frequency Control

This year's Frequency Control Symposium, co-sponsored by the IEEE Ultrasonics, Ferroelectrics and Frequency Control Society and the U.S. Army Laboratory Command, Electronics Tech-

nology and Devices Laboratory, is being held May 29-31 at the Los Angeles Airport Marriott, Calif. The symposium serves as a leading international forum on all aspects of frequency control and precision timekeeping.

Major areas of interest include fundamental properties of piezoelectric crystals, theory, design and processing of piezoelectric resonators and filters, stable oscillators and synthesizers, atomic

and molecular frequency standards, frequency and time coordination and distribution. Special sessions have been scheduled this year for the increasing role of optics and lasers in frequency standards, and on high temperature superconductors.

The day before the start of the symposium, seven tutorials will be presented including: Introduction to Phase Noise, Introductions to Quartz Frequency Standards, and Introduction to Frequency Synthesis. For information contact: Clark Wardrip, Bendix Field Engineering Corporation, PO Box 6147, Vandenberg, AFB, CA 93437. Tel: (805) 865-3214.

International Telecommunications Union Starts On-Line Computer Based Communications Services

— The ITU recently announced the implementation of an on-line telecommunications related information exchange. The system will offer such facilities as electronic mail, bulletin boards, document interchange, computer conferencing, distributed access to ITU databases and notification of telecommunication information including a terminology infobase of 30,000 telecommunications terms in English, French and Spanish. The use of TIES (the ITU Telecom Information Exchange Services) is basically free of charge. The system is accessible through public telecommunications networks and is available 24 hours a day, seven days a week. For further information on TIES contact: Mr. L. Goelzer, Chief, Computer Department, ITU. Tel: (41) 22 730 5333. Fax: (41) 22 730 5337.

Wavetek to Enter Merger —

Wavetek Corporation has announced that its Board of Directors has entered into an agreement in principle with Torrey Investments Inc., whereby Torrey Investments would acquire Wavetek by merger, in which the stockholders of Wavetek would receive \$3.15 per share in cash. The value of the transaction is approximately \$28,300,000. The merger is subject to execution of a definitive agreement and its approval by the Board of Directors of both corporations. In addition, the proposed merger is conditional upon Torrey Investments obtaining certain financing.

ASEP Guides Electroless Plating on Plastics Standards — The American Society of Electroplated Plastics (ASEP) will hold two half-day electroless

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plating of plastics roundtable discussions April 23 and 25 respectively, in Boston and San Jose. Specifiers, applicators and industry suppliers will want to attend the discussions to establish standards for evaluating the performance of electroless plating on plastics for EMI shielding. Topics will include: electrical performance, shielding testing, adhesion tests, accelerated environmental aging methods, durability tests and appearance guidelines. For more information call ASEP headquarters at (202) 371-1323 or Fax (202) 371-1090.

Microstrip Patch Antenna Developed — NIST has recently come up with a new development that makes certain antenna measurements and electromagnetic interference/compatibility tests more convenient. NIST researchers have developed a small (20 cm square) microstrip patch antenna that can be used as a standard transmitting and receiving antenna at frequencies below 500 MHz in an anechoic chamber. Researchers proved that the resonant frequency, driving point impedance, antenna radiation pattern and radiated field strength of the microstrip patch antenna could be calculated theoretically from its geometry and are accurate to within three percent. Paper no. 2-91 describes the antenna in detail and is available from Jo Emery, Div. 104, NIST, Boulder, CO 80303. Tel: (303) 497-3237.

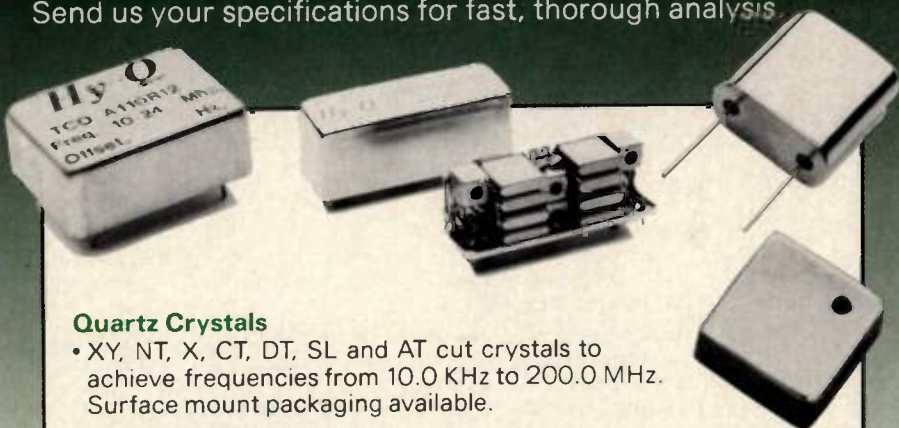
Piezoelectric Devices Call for Papers — The 13th Piezoelectric Devices Conference to be held October 9-11, 1991 has issued a call for papers. Papers may be tutorial or application-oriented, dealing with recent progress in design, development, processing, or manufacturing control in areas represented by the following topics: properties of natural and cultured quartz, design of quartz resonators, oscillator and filter design, CAD/CAM, manufacturing and process control, measuring and test techniques, packaging and hybrid techniques, quality and reliability assurance, and SAW devices and applications. An abstract of the proposed paper clearly describing the content, scope, organization key points and presentation time is required for paper selection and session assignment. Mail three copies of the abstract with a phone number by May 6, 1991 to: Components Group, Electronic Industries Association, 2001 Pennsylvania Ave., N.W., Washington, DC 20006.

Scholarships Available for Radio Amateurs — The Foundation for Amateur Radio, Inc., plans to award thirty-six scholarships for the academic year 1991-1992 to assist licensed Radio Amateurs. Licensed Radio Amateurs may compete for these awards if they plan to pursue a full-time course of studies beyond high school and are enrolled in or have been accepted for enrollment at an accredited university, college or

technical school. The awards range from \$500 to \$2000 with preference given in some cases to residents of specified geographical areas or the pursuit of certain study programs. Additional information and an application form can be requested by letter or QSL card, postmarked prior to May 31, 1991 from: FAR Scholarships, 6903 Rhode Island Avenue, College Park, MD 20740.

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Exhibitors Set for NAB '91 Convention — The National Association of Broadcasters has announced that more than 700 exhibitors will participate at its annual convention in Las Vegas, April 15-18. The NAB convention will offer exhibitors more than 430,000 square feet of show space at the Las Vegas Convention Center. The Engineering Conference is scheduled for April 14-18. In addition, NAB is sponsor-

ing the HDTV World Conference & Exhibition during the same dates.

Motorola and DSC to Develop Open Cellular System Standards

— Motorola, Inc. and DSC Communications Corp. have announced that they have reached an agreement in principle to cooperate in developing, in connection with appropriate standards bodies, open cellular system interface stan-

dards for the worldwide cellular market. The standards will be based on existing international switching and cellular standards such as CCS7, GSM, ISO and ISDN. Under the proposed agreement, Motorola and DSC will continue, on a non-exclusive basis, to market, enhance and support cellular systems worldwide. The agreement in principle, which is subject to preparation of definitive agreements, contemplates modifying the existing agreement between the two companies in the area of exclusivity, cross licensing of certain technology, marketing and manufacturing rights, and pricing and payment terms.

TriQuint Semiconductor and GigaBit Logic Merge

— TriQuint Semiconductor and GigaBit Logic recently announced the signing of a letter of intent that will allow the two companies to be merged into one corporation. The new company will be jointly owned by the existing investors. The name of the new company will be TriQuint Semiconductor Inc. A distinct digital product line, which will include digital standard products and ASICs, will be marketed under the GigaBit Logic name. Until the definitive agreement is signed and conditions for the merger are met, the two companies will continue to operate independently. The merger is expected to be completed by the end of the first quarter of 1991.

Report Available on HDTV and Related Systems

— A report released by the Office of Technology Assessment (OTA), *The Big Picture: HDTV & High Resolution Systems*, examines the scope of recent major technological advances and how the United States might benefit. The report evaluates the potential influence high definition television (HDTV) and related high resolution systems (HRS) may have in the consumer entertainment industry, and how they may affect communications, national security, research and education. The report also focuses on how HDTV and HRS are causing the U.S. to re-examine current national policies dealing with manufacturing, educational and training standardization, communications, military command, structural economic problems and the relationship between government and business. Finally, the report discusses the importance HDTV holds for the U.S. in rebuilding its leadership role in global and domestic electronic technology markets. The report, stock number 052-003-

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QBH-110	5-500	15.0	9.0	3.5	25.0	22/32	\$15
QBH-119	5-500	15.0	11.0	3.3	25.0	24/33	\$15
QBH-120	5-500	14.5	1.0	2.3	26.0	13/17	\$15
QBH-122	10-500	17.0	19.0	4.6	22.0	24/32	\$110
QBH-126	5-500	15.0	15.0	4.2	24.0	28/34	\$15
QBH-155	5-300	15.0	21.0	6.4	28.0	36/48	\$65
QBH-183	5-1100	10.3	14.0	6.5	12.0	27/38	\$80
QBH-184	5-1000	14.8	10.0	5.0	17.0	24/33	\$15
QBH-815	5-1000	11.4	20.5	8.0	12.5	30/45	\$95
QBH-822	10-2000	20.0	10.0	6.0	24.0	23/34	\$162

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NEC and CEL Expanding Design Center — The NEC Corporation and California Eastern Labs have announced plans to expand the NEC/CEL Design Center to include development of silicon MMIC devices. The program will concentrate on the development of both standard products and customer specific devices based on NEC's proprietary DNP III direct nitride passivation process.

Burr-Brown Offering Direct Marketing Services — In an effort to streamline the entire product selection and order placement process, Burr-Brown has established a direct marketing organization within its Customer Service Center. Customers can now call toll-free and receive real-time applications assistance, request and receive one-hour fax service on product literature, receive price and delivery quotations, and place their orders over the phone charging their purchases to VISA, MasterCard or Discover Cards. To implement the process, Burr-Brown has installed a toll-free number (800) 548-6132 and a telefax capability (602) 741-3895.

Symposium Dates Changed — The Third International Symposium on Recent Advances in Microwave Technology is now scheduled from August 18-21, 1991. The symposium will cover all the topics in Microwave Technology and its applications including components, circuits, antenna and radar, MICs and MMICs, remote sensing, biological effects, communication systems, CAD techniques, and propagation and measurements among others. For information on the symposium contact: Prof. Banmali Rawat, Dept. of Electrical Engineering, University of Nevada - Reno, Reno, Nevada 89557-0030. Tel: (702) 784-6927. Fax: (702) 784-1300.

Viewsonics Announces Move — Viewsonics recently announced that they will be moving to Florida from their present New York location. Their new address will be: Viewsonics, Inc., 6454 E. Rogers Circle, Boca Raton, FL 33487. Tel: (407) 998-9594. Fax: (407) 998-3712. The toll free number will remain (800) 645-7600.

13th Annual Antenna Measurement Techniques Symposium Call for Papers — The AMTA has issued a call for papers for the October 7-11 symposium. Topics include, advanced antenna measurement techniques, practical aspects of measurement equipment, systems and equipment interfacing, theory and applications of measurement techniques, range, design, automation and evaluation, near-field tech-

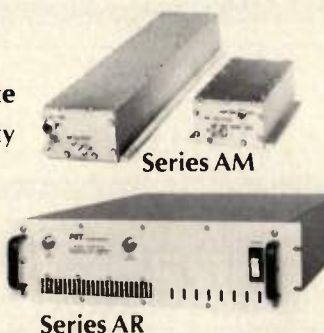
niques, anechoic chamber and absorber design and evaluation, phased array testing, and compact range design and evaluation. Authors should submit a 200 word abstract along with the contact author's phone and address by May 10, 1991 to: Prof. Roger C. Rudduck, AMTA Technical Coordinator, Electroscience Laboratory, Ohio State University, 1320 Kinnear Road, Columbus, OH 43212. Tel: (614) 292-6113.

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MODEL LA200H	MODEL LA200F	MODEL LA200U	MODEL LA200UE	MODEL LP300H
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200 Watts CW	200 Watts CW	200 Watts CW	200 Watts CW	300 Watts Pulse

MODEL LA400U	MODEL LA500H	MODEL LA500V	MODEL LA500U	MODEL LA1000H
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400 Watts CW	500 Watts CW	500 Watts CW	500 Watts CW	1000 Watts CW

MODEL LA1000V	MODEL LA2000H1	MODEL LA3000HS	MODEL LP4000HV2	MODEL LP12000HV
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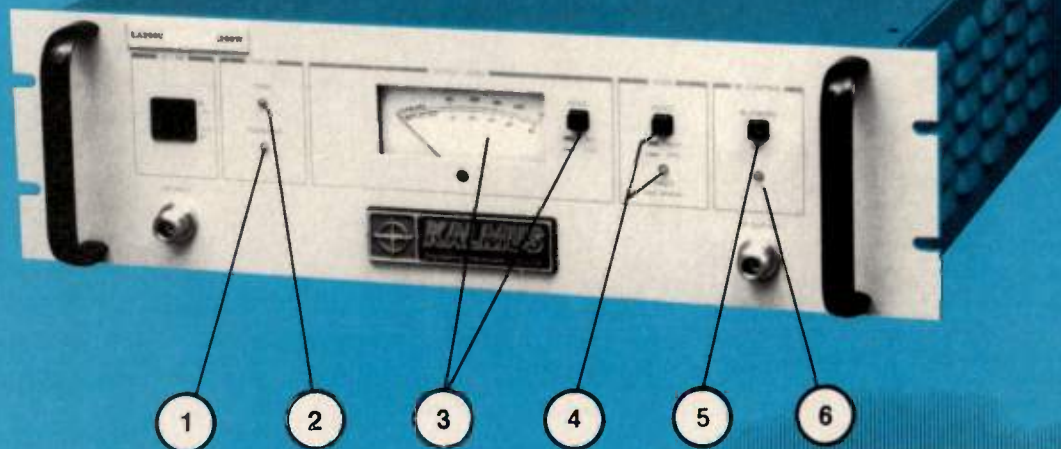
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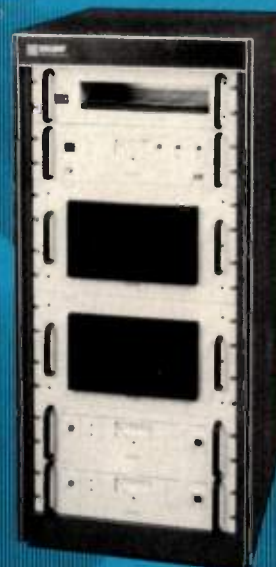


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AND 4000HV



LA 2000



RF Technology in Broadcasting and CATV

By Gary A. Breed
Editor

Broadcasting was one of the first commercial applications of RF technology, and radio, television and cable television continue to have a substantial influence on the development of new technologies. Studio, head-end, and transmitting facilities are still strong markets for advanced RF technology. This short report can only touch on some of the major developments.

In television broadcasting, Varian has had two major projects underway. The first is development of the Klystrode, as its name implies, a hybrid of the klystron and tetrode. This tube was created to improve the efficiency of the linear power amplifiers required for UHF transmitters. It doesn't take much math to estimate the cost of operating a 110 kilowatt transmitter using a 30 percent efficiency klystron, and then come up with the cost savings when that is improved to 40 or 45 percent. With development and test cooperation from TV transmitter manufacturer Comark, the Varian Klystrode has been proven to provide lower operating costs through higher efficiency.

The second project, also at Varian, is a higher efficiency klystron using a depressed-collector configuration. Because this tube is a true klystron, it has applications at much higher frequencies than are available with the Klystrode. Both this and the Klystrode have applications beyond broadcasting, in radar, EW and ECM, and satellite communications. This is why NASA is a participant in depressed-collector klystron development, as well as the broadcasting industry.

It would be unfair not to mention work on solid-state television power amplifiers. Several companies offer VHF transmitters with solid-state aural amplifiers (Class C) operating up to several kilowatts. The video amplifiers remain a challenge, mainly because television requires linear amplification with its necessary lower efficiency and higher

heat dissipation. Still, all current high power transmitters have solid state amplification through the driver stages.

Television production equipment utilizes new wideband linear integrated circuits, and has a major stake in high speed digital circuitry for image processing. Studio video for current video standards requires from 10 to 20 MHz bandwidth depending on the specific application. New operational amplifiers, buffers and analog multiplier ICs from companies like Elantec, Comlinear, Harris Semiconductor, Analog Devices and Burr-Brown are found in video distribution equipment, and in the analog input and output stages of digital video processing equipment. Analog switches from Maxim, Siliconix, and Harris are used for routing video and audio signals. In high performance applications, switches built with discrete FETs can still be found.

High speed digital technology had some of its first applications in television, with digital time base correctors, standards converters, and frame synchronizers. All of the special effects and graphics that television viewers take for granted are truly technological masterpieces. Pioneering these applications were companies like the Grass Valley Group, 3M, Ampex and others. Advanced image manipulation requires high speed analog-to-digital and digital-to-analog conversion plus high speed processors to perform the vector mathematics required for manipulating each pixel that makes up the image. Components for these applications are manufactured by Sony, TRW, Burr-Brown, Analog Devices, Datel and Comlinear.

Cable television continues to require medium power, low distortion amplifiers for signal distribution, with Motorola (ex-TRW) and Philips the main suppliers. High performance head end modulators are needed to minimize adjacent-channel interference, using SAW IF filters, and low distortion modulator and amplifier circuitry. Test equipment for

CATV has seen considerable improvement in recent years, primarily for real-time system performance monitoring, and fast troubleshooting of equipment failures.

Encryption technology is another area where high-speed digital processes are required. From simple remotely switched and monitored converters to high-security scrambler/descrambler units that continually change the encryption code, high speed video processing reaches many cable subscribers. Even simple set-top decoders require excellent dynamic range and intermodulation performance, as well as programmability over the wide bandwidth of CATV signals (as much as 500 MHz).

Radio broadcasting continues to change from vacuum tube power amplifiers to solid state devices. One transmitter manufacturer estimates that nearly 50 percent of the local AM radio stations (1 kW power or less) are now using solid state transmitters. At very high power, companies like Harris Corp., Continental, Nautel, and others have developed unique combining techniques for generating up to 50 kW at medium wave frequencies using multiple MOSFET amplifier modules. FM transmitters at the 3.5 kW level and under are available in solid state designs, but have only been sold for three years or so.

If you include the radios, stereo systems, television sets and VCRs used for the reception of broadcast and CATV programming, broadcasting is probably be the largest RF industry segment, worldwide. In the U.S., where consumer equipment is a more modest part of RF manufacturing, the transmission equipment for radio and television is still a considerable market for RF components and a source of significant technological advancements.

RF

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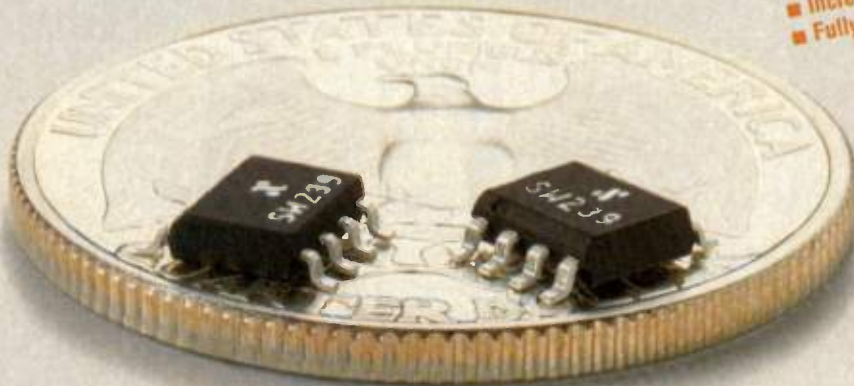
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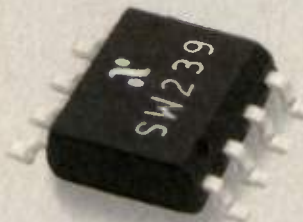


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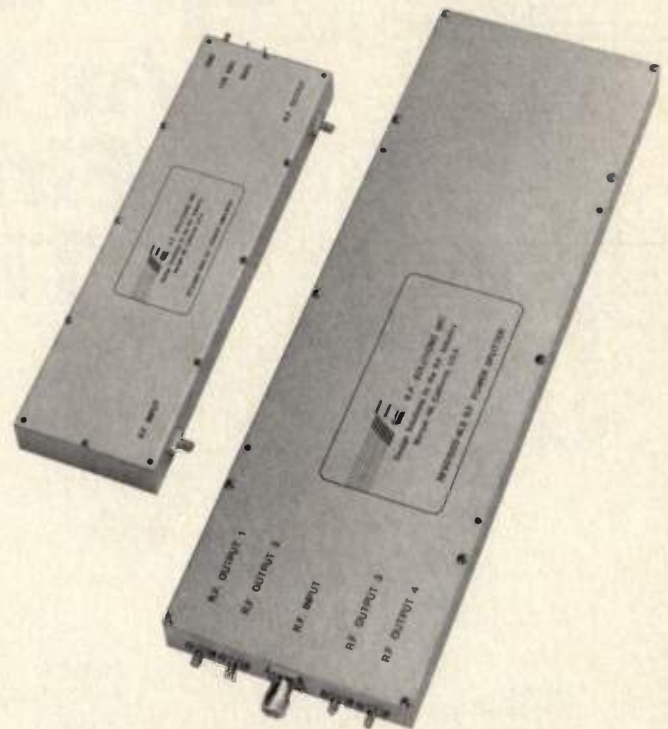
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RFP0550-1000	1000	16	50	\$5,040.00
FREQUENCY RANGE 50 – 100 MHz				
RFP0800-100P50	50	30	50	\$1,485.00
RFP0800-100P100	100	30	50	\$1,660.00
RFP0800-100P200	200	30	50	\$2,200.00
RFP800-100	600	16	50	\$2,424.00
FREQUENCY RANGE 1 – 100 MHz				
RFP01100-300	300	46	50	\$3,150.00
FREQUENCY RANGE 76 – 108 MHz				
RFP0810-600	600	16	50	\$1,780.00
FREQUENCY RANGE 75 – 150 MHz				
RFP0800-150P50	50	30	50	\$1,485.00
RFP0800-150P100	100	30	50	\$1,660.00
RFP0800-150P200	200	30	50	\$2,200.00
RFP800-150	500	14	50	\$2,424.00
FREQUENCY RANGE 100 – 200 MHz				
RFP0800-200P50	50	30	50	\$1,660.00
RFP0800-200P100	100	30	50	\$2,900.00
RFP800-200	400	13	50	\$3,636.00
FREQUENCY RANGE 225 – 400 MHz				
RFP0204-4	4	20	28	\$ 484.00
RFP0204-10	10	30	28	\$ 685.00
RFP0204-25	25	30	28	\$1,140.00
RFP0204-50	50	40	28	\$1,695.00
RFP0204-100	100	40	28	\$2,200.00
FREQUENCY RANGE 400 – 500 MHz				
RFP0405-4	4	20	28	\$ 435.00
RFP0405-10	10	30	28	\$ 616.00
RFP0405-25	25	30	28	\$1,026.00
RFP0405-50	50	40	28	\$1,525.50
RFP0405-100	100	40	28	\$1,980.00
FREQUENCY RANGE 1 – 500 MHz				
RFP00105-4	10	30	28	\$1,450.00
RFP00105-25	25	30	28	\$2,800.00
RFP00105-50	50	40	28	\$3,752.00
RFP00105-100	100	40	28	\$5,600.00



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RFS1000-5-50-4	4	0.3	1000	\$ 610.00
FREQUENCY RANGE 50 – 1000 MHz				
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RFS100-FL-FU-3	3	0.15	100	\$ 600.00
RFS100-FL-FU-4	4	0.3	100	\$ 800.00
RFS100-FL-FU-6	6	0.3	100	\$1,000.00
RFS1000-FL-FU-2	2	0.15	1000	\$ 750.00
RFS1000-FL-FU-3	3	0.15	1000	\$ 900.00
RFS1000-FL-FU-4	4	0.3	1000	\$1,200.00
RFS1000-FL-FU-6	6	0.3	1000	\$1,500.00

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Designing a Direct Sequence Spread Spectrum Secure Communication System

By Soo Kwan Eo, Chingwo Ma, Incheol Jang
Comdisco Systems

Spread spectrum signals in communications are used for the purpose of combatting the detrimental effects of interference due to jamming, interference arising from other users of the channel and self-interference due to multipath propagation. The spread spectrum signal can be hidden from unintended listeners by transmitting it at low power, thereby achieving message privacy in the presence of other listeners.

One form of spread spectrum, direct sequence spread spectrum (DS-SS) modulation, can be achieved by superimposing a pseudo-random number (PN) sequence upon the information bits. In the receiver, the received signal is cross correlated with the same PN sequence. The effect of this cross correlation is the reduction of the information spectrum bandwidth and, at the same time, the widening of the unwanted jamming signal and receiver noise in order to effectively increase the signal-to-noise ratio. A merit of this scheme is that power spectral dispersal can significantly improve the bit error rate of the information pulse train.

The performance of the PN sequence spread-spectrum system can be further increased, with respect to its immunity to the interfering signal, by incorporating a linear adaptive LMS filter before the de-spreading operation takes place. Some algorithms (such as that of Theodoridis (1)) employ LS linear phase FIR filters to suppress the interference. While they provide adequate performance, if the interference is a slowly time-varying signal these algorithms are ineffective. A better approach would be to adopt a time-recursive linear phase scheme such as the Widrow LMS algorithm (2,3).

The design of such a system requires careful analysis of three main concerns:

1) The reduction of cross-correlation of different pseudo-random sequences due to the lack of absolute randomness in the different sequences generated.

(Cross correlation close to zero is needed to generate a sequence that approximates white noise.)

2) The receiver's ability to synchronize the generation of the pseudo-random sequences in order to properly convolve the transmitted data.

3) The attributes of the interference suppression filter, (i.e., its ability to suppress interference of some predefined type as the relative strength of the data signal and the interfering signals vary.)

In this article we will present a design methodology that allows the designer of such a system to rapidly simulate the effects of changing system topology and parameters in a DS-SS system. The design software used is the Signal Processing WorkSystem, from Comdisco Systems. The overall system architecture is presented and the operation of the transmitter and receiver are discussed. For this example we will assume that the interfering signal is a high-power

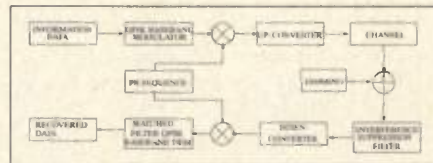


Figure 1. Direct spectrum spread signal system overview. A major design goal of this system is to recover the data signal in an environment with a jammer-to-signal ratio (JSR) of up to 30 dB.

narrow band signal that may be swept across the bandwidth of the data signal. The filter chosen for this example is a linear adaptive LMS filter that will be described and illustrated.

To demonstrate the ease of analysis provided by this design methodology, the injected narrow band jamming signal and the corresponding interference rejection filter response are shown. Designers of such systems want to know

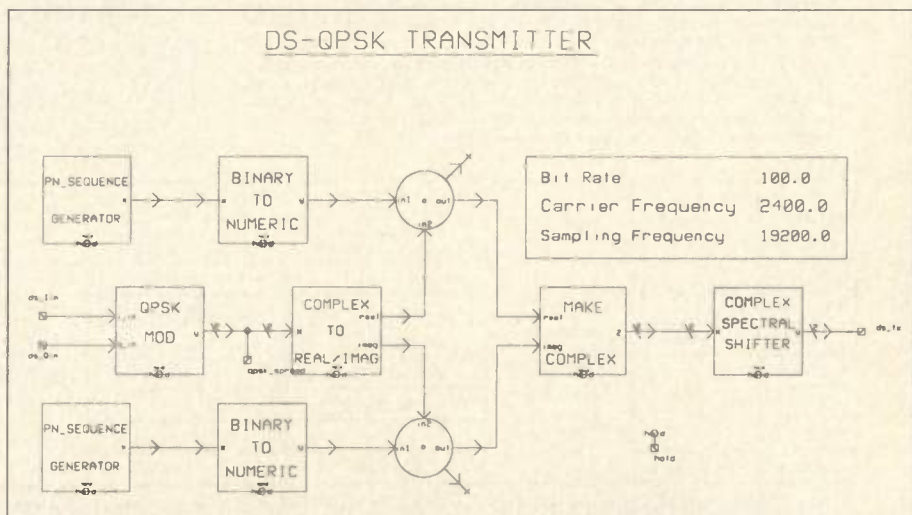


Figure 2. Direct sequence spread spectrum QPSK transmitter (as displayed on Comdisco System's Signal Processing WorkSystem). The QPSK output is at a carrier frequency of 2400 Hz with a sampling rate of 19200 samples per second and a bit rate of 100 bps.

Narrow Bandwidth Jammer

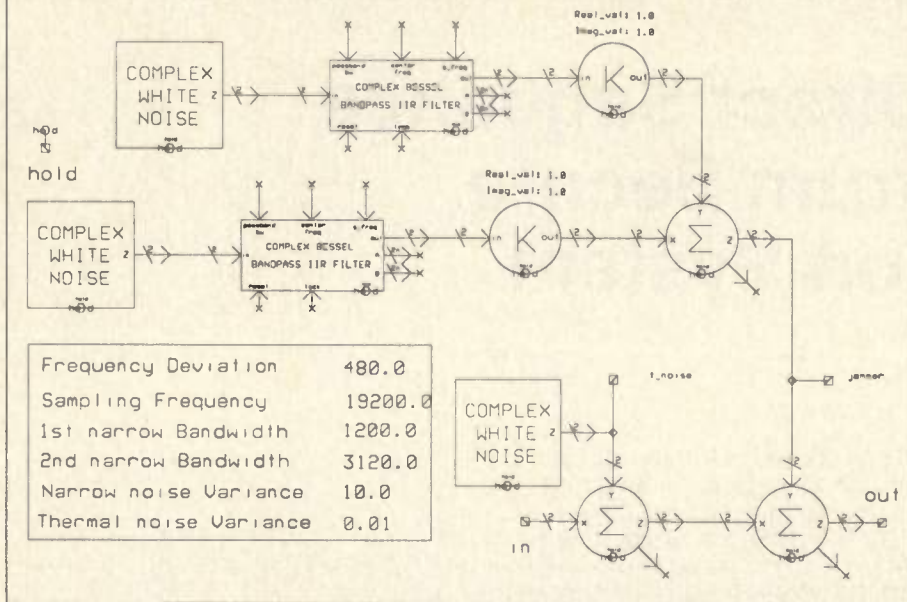


Figure 3. Narrow bandwidth jammer. This model is used to simulate interference that centers around two carrier frequencies (1200 Hz and 3120 Hz) with a frequency deviation of 480 Hz.

how the received symbol error rate varies as the jammer to signal ratio/chip varies. The techniques for performing this analysis are also demonstrated.

The Design Environment

The Signal Processing WorkSystem (SPW) provides simulation and analysis tools that automate digital signal proc-

essing system design. The user can select from a rich library of communications functional blocks, each of which can be represented hierarchically. The blocks are placed on the workstation display screen and then interconnected. To facilitate the simulation of the captured design, the user can either generate his own stimuli signals using a powerful signal editor, or, he can use real-world signals that have been digitized. The designer does not have to write any simulation code since the simulation program is automatically generated and compiled by SPW. The results of the simulation can be analyzed and displayed using FFTs to produce frequency-domain plots, eye diagrams, scatter plots, etc. Even the bit-error-rate (BER) of a digital communications system can be automatically calculated and displayed.

The software allows the designer to test the design at a high level of abstraction and to easily perform what-if trade-off analysis to optimize the design. It also lets the designer move to lower levels of abstraction so that an easy transition to hardware implementation

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A57TU	UHF Fixed		1-900 MHz	---				369.00
A57T/30	Low Frequency	Direct Reading Balun Null	30 KHz-30 MHz	---				311.00
A57TLS			300 KHz-100 MHz	---				258.00
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A56GA/6	VHF Variable		1-600 MHz	5-600 MHz			8 1/2 oz.	532.00

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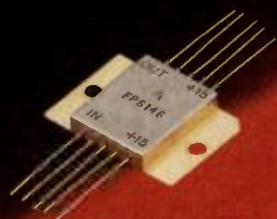
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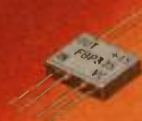
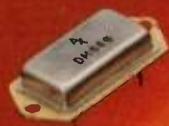
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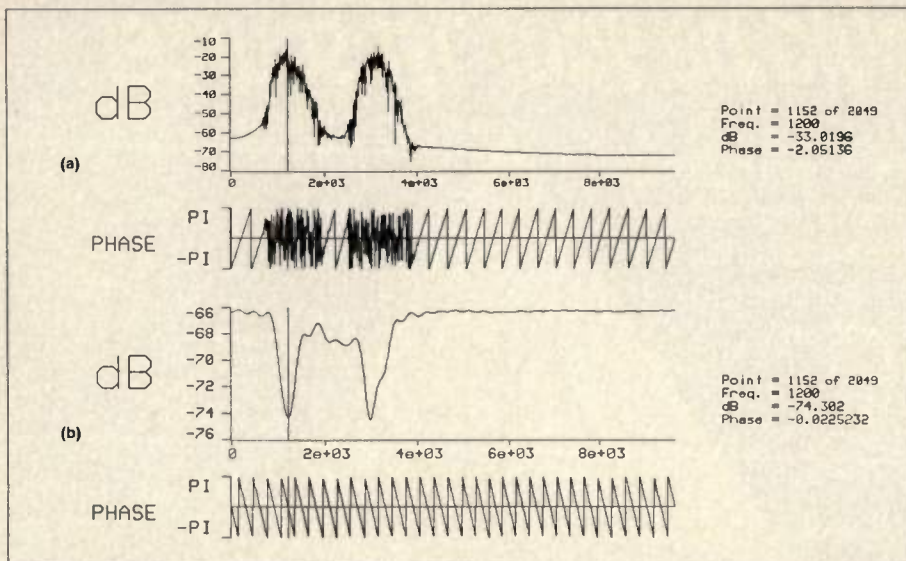


Figure 4. Interference suppression adaptive filter response. In (a) the two narrow band interference signals center at carrier frequencies of 1200 Hz and 3120 Hz with a 480 Hz frequency deviation. The response of the linear phase adaptive LMS filter is shown in (b).

is possible. Tools are provided that allow the user to implement his design in discrete integrated circuit standard parts, ASICs, or within many of the currently available digital signal processors utilizing Comdisco's Hardware Design System (HDS) for implementation

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The communications library includes a majority of the blocks a designer is likely to need. In fact, all of the blocks used in this design are contained in the

library. For those designs that require blocks not already in the SPW library, custom block building facilities are provided.

DS-SS QPSK System Overview

The direct sequence spread spectrum system presented here uses an adaptive suppression filter to minimize interference caused by time-varying jamming or interfering signals. A major design goal of this system is to recover the data signal in an environment with a jammer-to-signal ratio (JSR) of up to 30 dB. In the following sections, the direct-sequence spread spectrum (DS-SS) transmitter, jamming model, interference suppression filter model, and DS-SS receiver model are discussed.

The DS-SS QPSK system overview is illustrated in Figure 1. Sample information data is generated (or captured) within the software and a QPSK signal is output by the QPSK baseband modulator. The QPSK signal is multiplied by a binary sequence generated by the pseudo-random number (PN) sequence generator. The spectrum of the signal is further shifted to the carrier frequency

-150 dBc Phase Noise...

region by an up-conversion module and is transmitted into the communications channel. To receive the signal it is first processed by an interference suppression filter, then down-converted to the baseband. Then the signal is multiplied by the same PN sequence used in the transmitter. Finally a matched filter and QPSK demodulator reproduce the digital information.

For this example (see Figure 2), the transmitter uses a random data generator as its information source. Two sets of random data with equiprobable 0 and 1 bits are converted from binary to numeric form and are input to the QPSK modulator. Before the phase modulation takes place, the in-phase (I) and quadrature (Q) inputs are QAM (quadrature amplitude modulated) and EXORed with the pseudo-number sequence. The EXORing is performed with multiplication using the "1" or "-1" amplitude of the PN sequence. The spreaded information bit stream is not phase modulated before it is transmitted via the channel. In this example the bit rate is 100 bps, the carrier frequency is 2400 Hz, and the sampling rate is 19,200 samples per

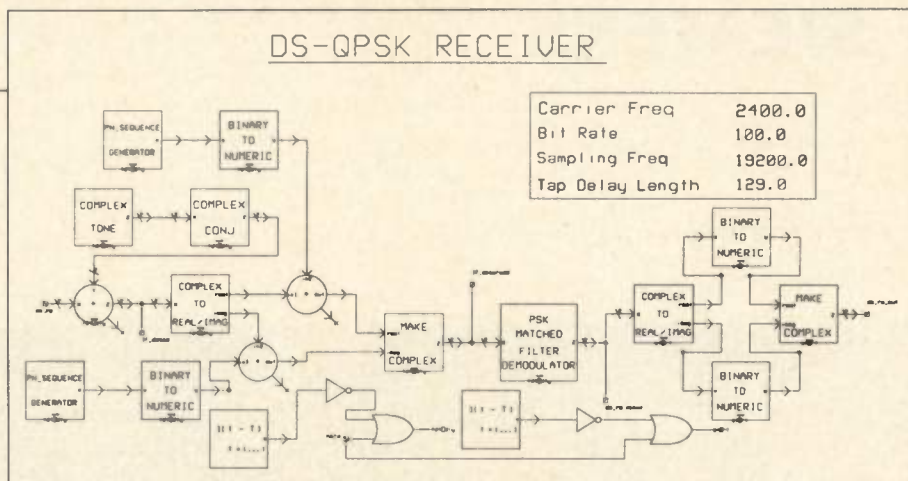


Figure 5. DS-QPSK Receiver.

second. The spreading is done by chip pulse and there are 24 chips per information symbol.

Narrow Bandwidth Jammer Model

Figure 3 shows the modeling of a subsystem that simulates a narrow bandwidth jammer. Other jamming modes, such as high-power CW jamming, or random power-variant jamming, are of course, possible. The system designer must decide which jamming modes will be modeled. He is also faced with the

need to determine whether or not a jamming signal is present (not a very difficult task) as well as the more challenging problem of identifying the jamming mode in order to activate the appropriate anti-jamming algorithm.

It is assumed that the jammer concentrates the jamming signal on the baseband data spectral region. In this example, a dual-band interfering signal is injected into the transmission channel (at 1200 Hz and at 3120 Hz) with a frequency deviation of 480 Hz using a

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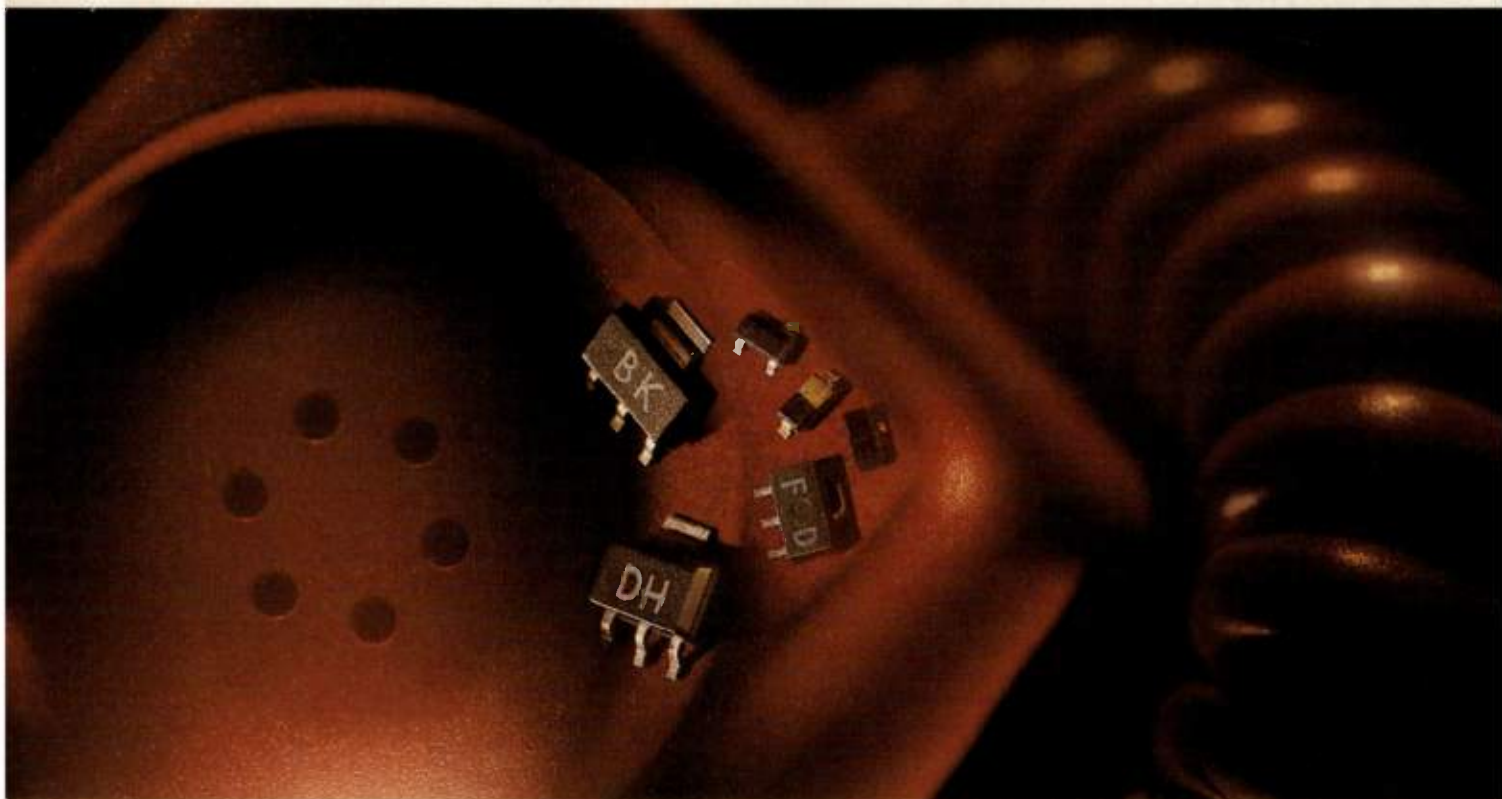
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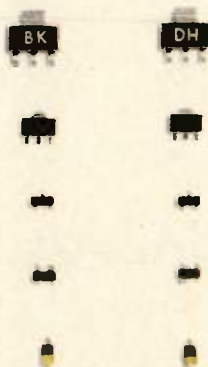


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	V	mA	900 MHz 8 V dB	900 MHz 8 V dB	$f = 0$ to above 3 GHz V/mA
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BFP 182	12	35	1.2	19	1 to 8/1 to 20 low-noise amplifier
BFP 183	12	65	1.2	19	1 to 8/2 to 28 low-noise amplifier
BFP 193	12	80	1.2	19	3 to 8/5 to 40 low-distortion output stage
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JSR/chip (dB)	-8.5	-1.5	3.7	13.3	15.7	18.6	21.5	23.2
W/O DS-SS	9.8×10^{-5}	.033	.05	0.3	N/M	N/M	N/M	N/M
W/O Filter	N/D	N/D	0.21	0.082	0.15	0.25	0.32	N/M
DS-SS, filter ($\times 10^{-4}$)	N/D	N/D	1.85	1.9	2.27	2.5	4.2	5.8

N/M : Not Measured
N/D : Not Detectable

Table 1. The symbol error rate for various JSR/chip ratios.

sampling rate of 19,200 with a level variance of 10 dB. Receiver thermal noise is modeled as white gaussian noise with a variance of 0.01 dB.

Interference Suppression Adaptive Filter

A linear phase adaptive LMS filter can be used to suppress the interference signal before the de-spreading operation is resumed at the receiver. In this example, an Equalizer Tap Coefficient block with backward error predicting scheme is used to implement the filter. Since the pseudo-random sequence is assumed to be almost white, the predictor predicts the high-power interference signal and the corresponding prediction error filter rejects the interference. It is assumed that the jamming signal is not synchronized to the signal's symbol rate but interferes on a bit-by-bit basis. The tap coefficients are appropriately updated to reject the interference. In Figure 4, the injected narrow band jamming signal and corresponding interference rejection filter response are shown.

DS-SS QPSK Receiver

The receiver operation is the exact inverse of the transmitter (see Figure 5). The DS-QPSK receiver recovers the data signal from the jamming signal by using a combination of linear adaptive filtering and spread spectrum techniques.

The transmitter sends a burst of synchronizing bits before it sends real data and the receiver uses this burst signal to synchronize the receiver (not shown in this current model).

Once interference is rejected, the received signal is phase demodulated and then de-spread using cross correlation between the received signal and the exact replication of the PN sequence that was used for transmission. This cross correlation further collapses the information signal to its original bandwidth. It also reduces the corresponding level of the narrow-band interference

by spreading it across the bandwidth occupied by the PN sequence.

After the received signal is collapsed to its original bandwidth and interference signal power is minimized, it is fed into the matched filter and PSK detector.

Simulation

The ultimate goal of this system is to recover a signal that is being jammed by narrowband interference. The best

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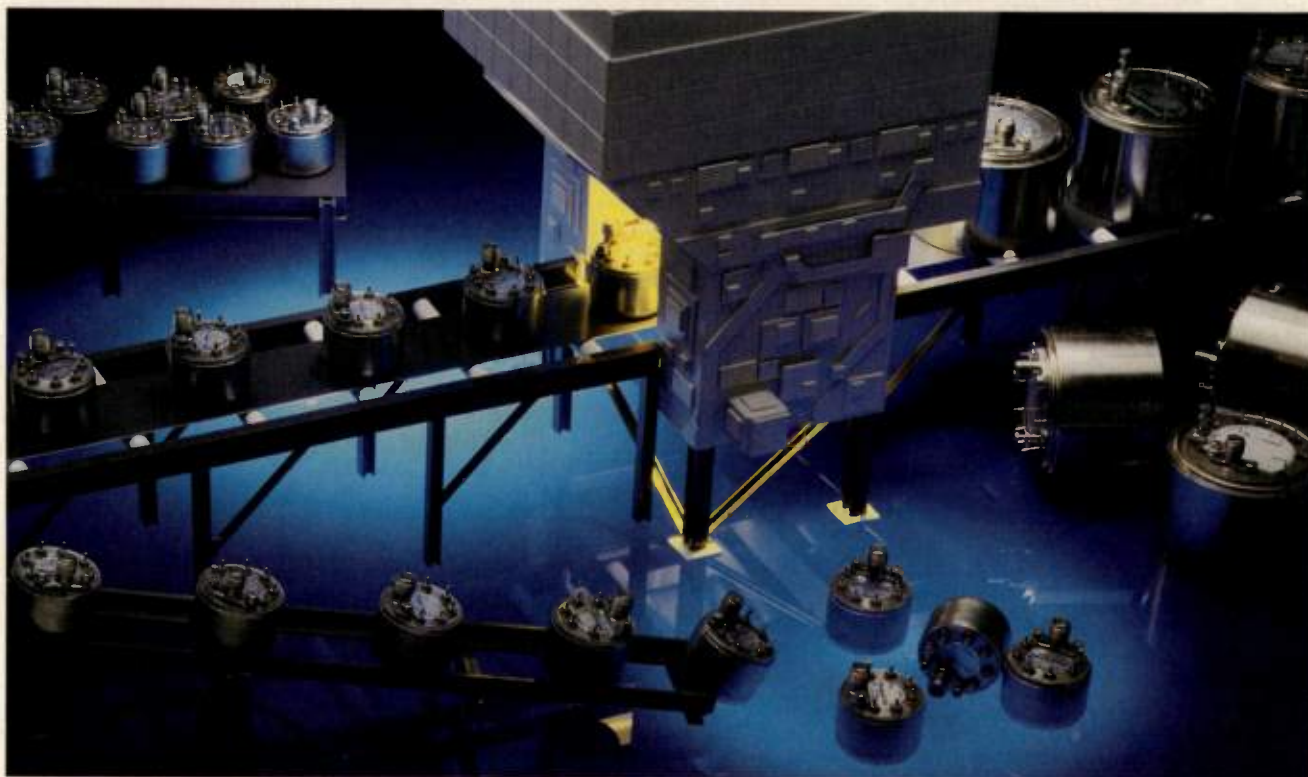
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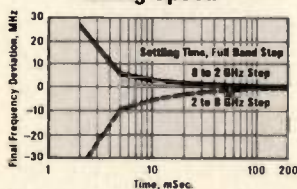
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measure of the effectiveness of the design is an examination of the received symbol error rate. Due to the complexity of this DS-SS QPSK system, determining this symbol error rate using hand coded simulation and analysis is nearly impossible. The Comdisco Signal Processing WorkSystem, however, allows the designer to accurately simulate the design, and automatically calculate the bit-error-rate to judge its effectiveness.

The simulation was configured with the assumption of a 129 tap filter, a jammer-to-signal ratio of 23 dB, and the two bands of narrow band interference described earlier in the article. Table 1 shows the symbol error rate for various JSR/chip ratios derived from the simulation. Each case was run with 2 to 3 million iterations. Once the symbol error is detected, the matched filter in the receiver loses phase synchronization with the incoming signal and tends to produce burst symbol errors. The symbol error rate is, therefore, calculated before the receiver experiences this burst error. Three scenarios were modeled and simulated. First, the system was simulated using no direct sequence spread spectrum techniques ("W/O DS-SS"). Then DS-SS was used but with no LMS adaptive filtering at the receive end ("W/O Filter"). Finally, the system was simulated using both DS-SS and filtering ("DS-SS, filter").

Note the multiplier of 10^{-4} for the DS-SS, filter case. As can be seen from this data, the combination of DS-SS and LMS adaptive filter produce greatly improved symbol error rates.

Conclusion

This design methodology offers the designer of a direct sequence spread spectrum system dramatically new options. He can now experiment with different filter schemes, interference patterns, PN generators, etc. to optimize system performance, minimize complexity, reduce cost, etc. before building any prototype. Within a workstation environment, the designer need only swap functional blocks, change interconnections, vary parameters within those blocks, and then analyze the data after the simulation is performed with the new specifications. **RF**

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3. L.B. Milstein, L.A. Iltis, "Signal Processing for Interference Rejection in Spread Spectrum Communications," *IEEE ASSP Mag.*, pp. 18-31.

About the Authors

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I/Q demodulation is commonly found in Doppler Radar and communications systems including AM, DSB, and QAM formats. With additional circuitry FM, PM, SSB, ISSB and other formats can also be added to the list. I/Q demodulators are quite flexible, but this flexibility comes with some problems. The use of modern high speed analog-to-digital converters (ADCs) allow traditional "analog" processing of I/Q demodulators to be carried out digitally, greatly reducing systematic errors. This article covers the practical implementation of a high-performance I/Q demodulator using modern ADCs.

A basic I/Q demodulator block diagram is given in Figure 1. An arbitrary IF signal enters the power divider, where it is split and fed to two mixers. Quadrature-phased local oscillators also feed these mixers. Low-pass filters remove the "sum" mixer products leaving the I and Q signals. When demodulating some signals (such as QAM) this is all that is needed. In others, ADCs and digital signal processing may follow.

There are several disadvantages to this approach. When implemented with conventional RF components, DC off-

sets, quadrature-phase errors, gain matching, carrier leakage, and impedance matching (among others) degrade demodulator performance. With advanced analog-to-digital converters such as Comlinear's CLC926, we can actually build a better demodulator.

Building a Better Demodulator

To illustrate the advantages of our approach, we'll start with a basic I/Q demodulator and modify it in stages showing the advantages after each modification. Since ADCs are very efficient harmonic mixers (see appendix), we can eliminate the mixers by replacing them with high-speed ADCs (Figure 2). All of the functions that follow the ADCs are now digitally implemented which eliminates matching errors in this part of the implementation and reduces most of the other errors.

Unfortunately ADC matching errors remain. Quadrature errors result from effective aperture delay mismatching between the two ADCs. ADC gain-matching errors are also present.

We could eliminate these errors if we somehow used only one ADC. Figure 3 provides us with a solution. Figure 3 shows the traditional local oscillator

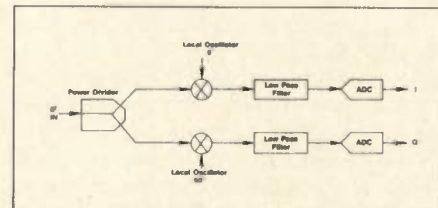


Figure 1. A basic I/Q demodulator.

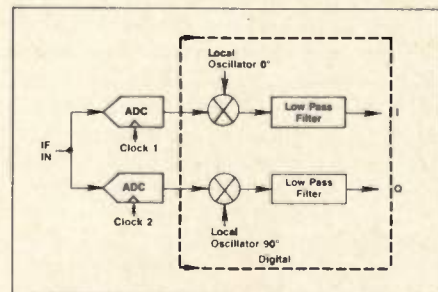


Figure 2. Another approach to I/Q demodulation.

signals required in I/Q demodulators using analog mixers. For the approach in Figure 2, all signals after the ADC are in discrete-time form. As a result, the local oscillator signals can be represented as impulses of strength -1 and 1. The samples from the ADCs coincide

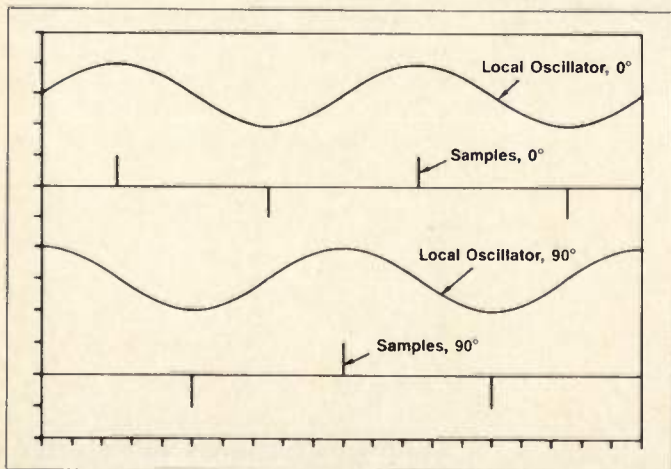


Figure 3. The upper 2 waveforms depict in-phase and the lower waveforms quadrature-phase local oscillators for the analog (Figure 1) and digital (Figure 2) approaches to I/Q demodulation.

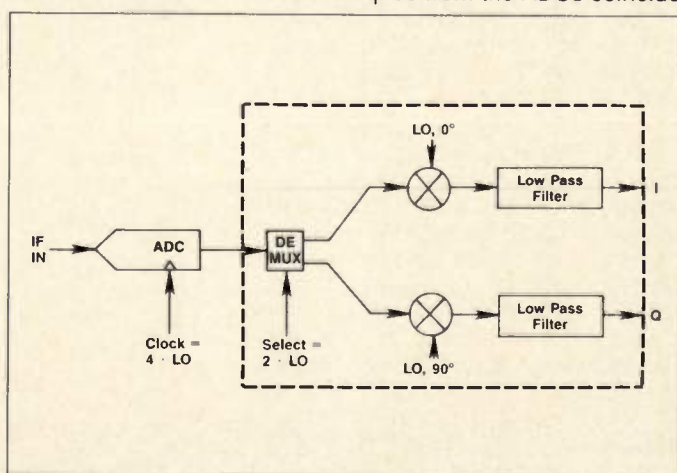


Figure 4. An improved I/Q demodulator.

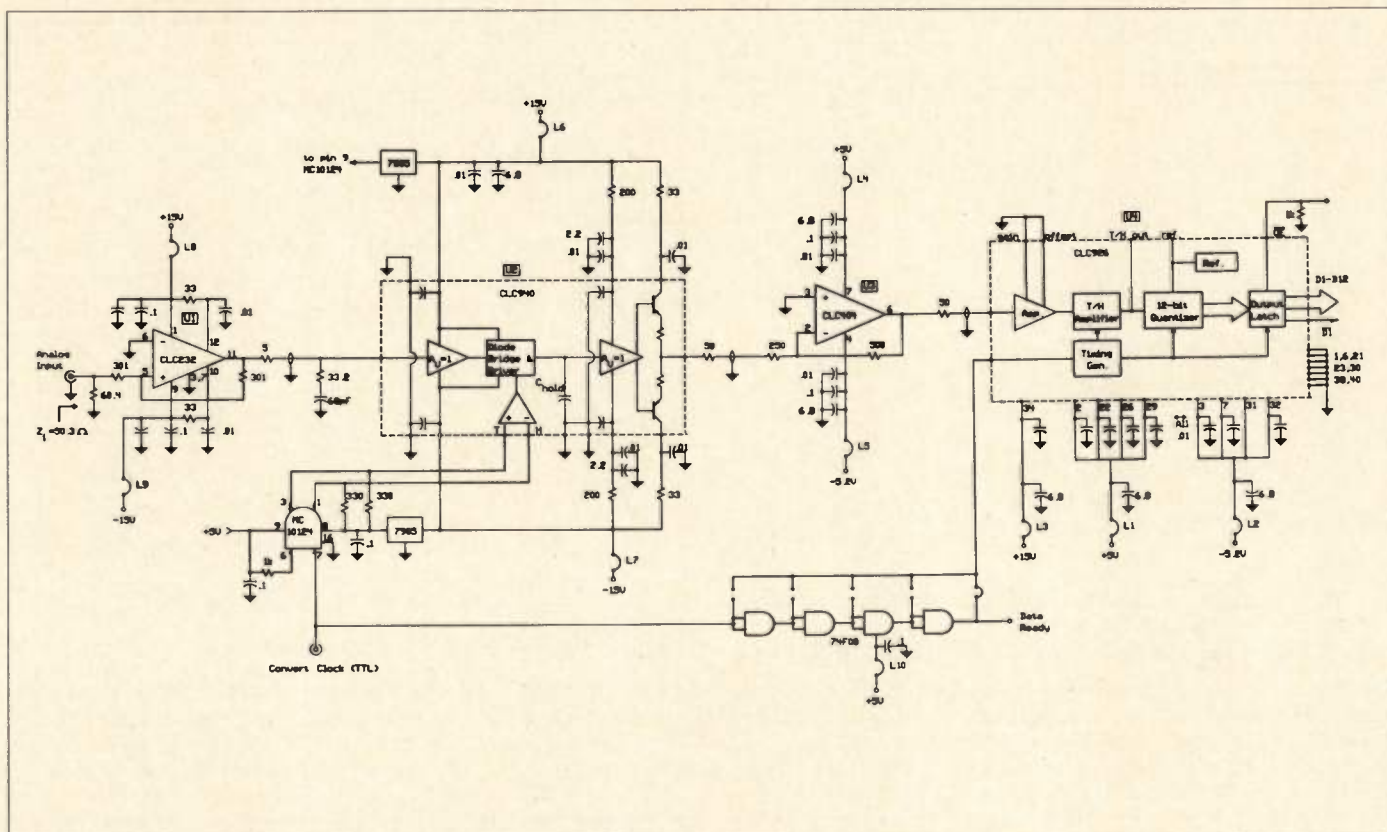


Figure 5. High performance IF sampling analog-to-digital convertor subsystem.

with the locations of these impulses.

From the diagram we can see that the in-phase ADC's samples occur exactly

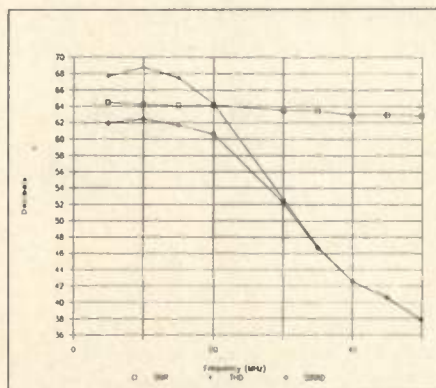


Figure 6. IF sampling performance.

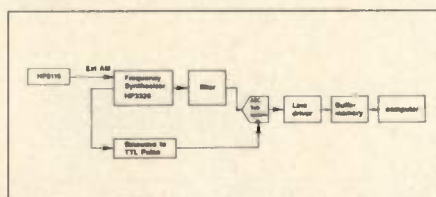


Figure 7. I/Q demodulation demonstration.

midway between the quadrature-phase ADC's samples. By using one ADC, doubling the sampling rate and sending alternate samples to the I and Q signal processing chain we can get the same effect. With this arrangement our sample clock will operate at four times the original local oscillator frequency. With both the I and Q samples coming from the same ADC, ADC quadrature and gain-matching errors are eliminated. The final block diagram is given in Figure 4.

Depending on speed requirements, discrete logic programmable DSP chips, or a mix of discrete logic and DSP chips could be used. For example, one DSP chip could be used after the ADC providing all demultiplexing, mixing, filtering, as well as any additional downstream signal processing. If more processing power is required, discrete logic can split the ADC data stream into separate I and Q signals with a DSP processor connected to each to speed things up. The same approach might also allow 2 cheaper DSP chips to replace one faster but more expensive chip. Mixing is also well suited to dedicated hardware, if desired. Since mixing is done by multiplying the ADC

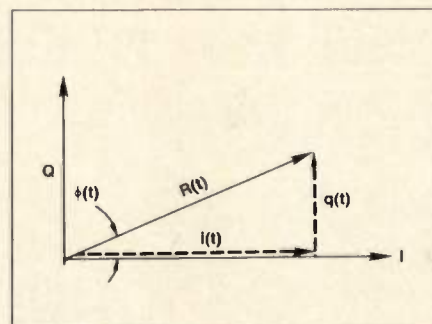


Figure 8. Quadrature carrier description of an IF signal.

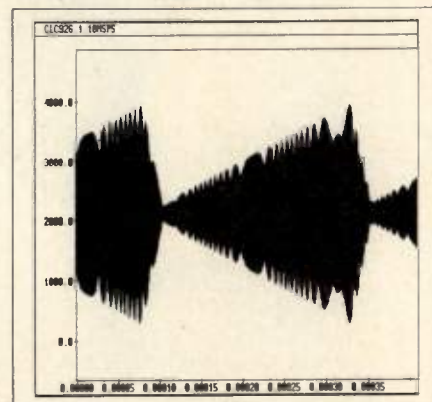


Figure 9. Sampled IF signal.

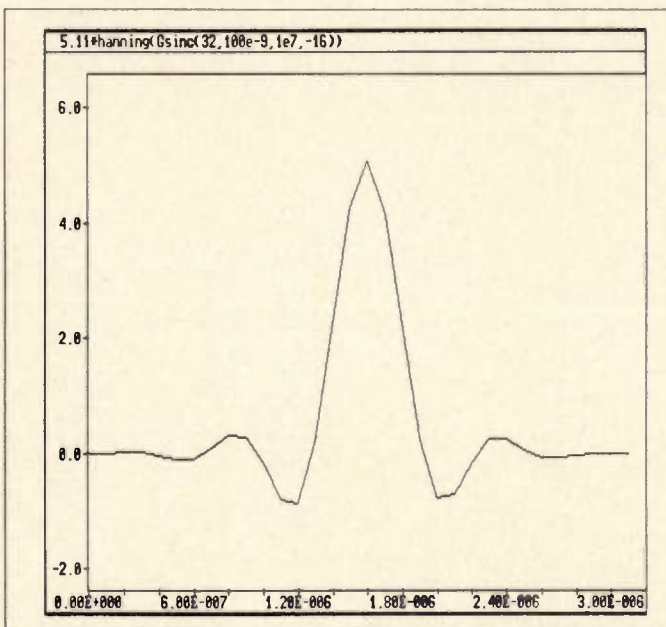


Figure 10a. Filter time domain response (impulse).

samples by alternating ones and minus ones (our local oscillator), simple hardware (not a full-blown multiplier) will suffice. The rest of the signal processing

can be investigated in similar fashion.

While the technique outlined isn't new, it hasn't had wide appeal because data converters didn't offer enough

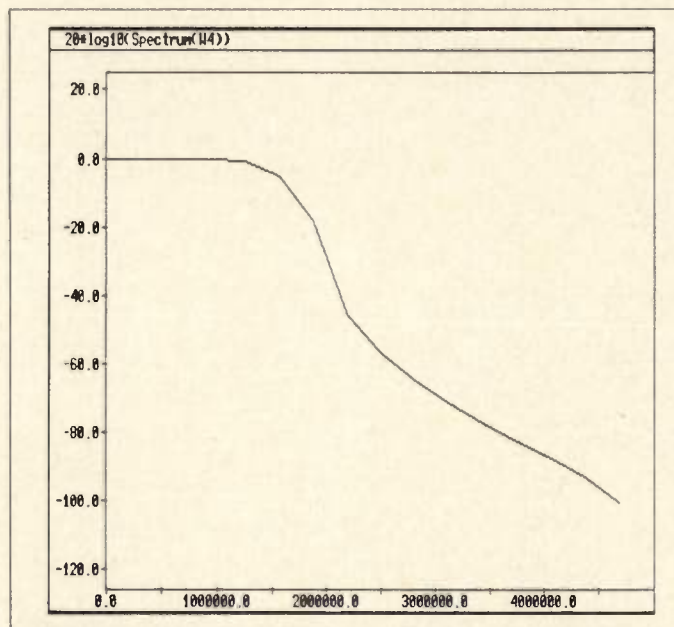
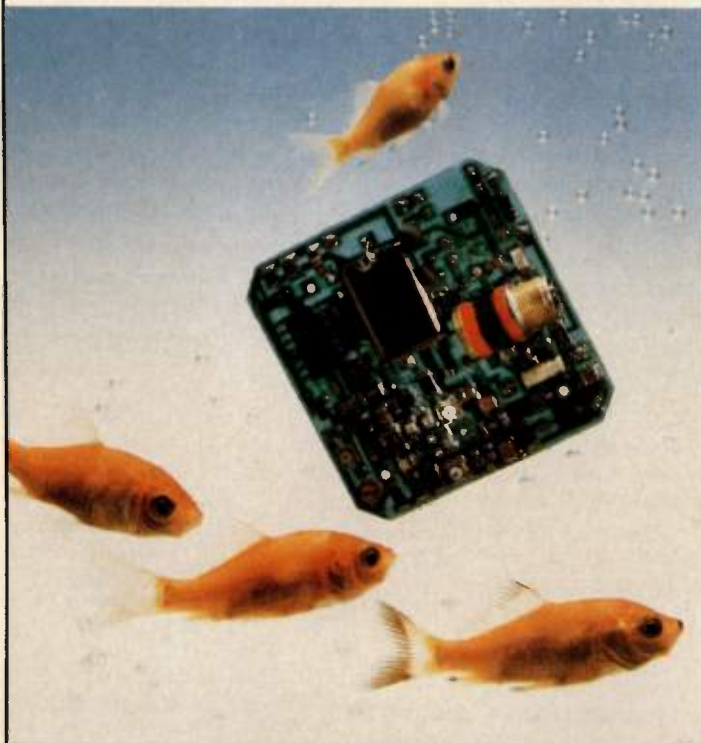


Figure 10b. Filter frequency response.

performance except at very low IF frequencies. When state-of-the-art ADCs and track-hold amplifiers such as the CLC926 and CLC940 are combined

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together, however, IFs up to 21.4 MHz and beyond become practical. Figure 5 is a schematic for a high-performance IF sampling analog to digital convertor subsystem. Analog-to-digital conversion is provided by U4, a CLC926 (a 10 MSPS, 12-bit ADC offering 70 MHz analog input bandwidth). The CLC926 is a good choice with its state-of-the-art spurious-free signal range, signal to noise ratio, SINAD, and THD — and all of these specifications are tested on every part.

Performance is extended at higher IF frequencies by placing track-hold amplifier U2 (a CLC940) in front of the ADC. Since the ADC does its conversion on a held signal its dynamic errors are reduced. U2 sets the aperture jitter to 1 ps rms typical so signal to noise ratio of the incoming signal is not degraded by the track-hold even at high IF frequencies.

To reduce distortion, the CLC940 track-hold is intentionally operated at reduced amplitude. With typical second and third harmonic distortion performance of -57 and -60 dBc at 20 MHz and 2 V_{pp} output amplitude, distortion isn't low enough for our purposes. But since the CLC940 has classic intercept-point behavior, dropping the amplitude 6 dB to 1 V_{pp} should reduce HD2 by 6 dB (or -63 dBc) and HD3 should drop 12 dB to -72 dB. A CLC404 wideband, low distortion amplifier (U3) boosts the 1 V_{pp} CLC940 output to 2 V_{pp} — the full-scale range of the CLC926 ADC (This insures highest ADC dynamic range). Since the CLC404 is after the track-hold, dynamic performance requirements on it are somewhat relaxed.

Several other tricks are used to maximize performance. U1 (a CLC232) buffers the input signal providing less than 10 ohms driving-point impedance over frequency-minimizing track-hold amplifier distortion products. Distortion is also slightly improved by using differential ECL track-hold control signals.

Looking at the results from single-tone tests (Figures 6), it is easy to see that our efforts have been rewarded. At 21.4 MHz IF (-1 dB full scale), SINAD (signal to noise + distortion) is better than 60 dB (or 9.8 effective bits). Total harmonic distortion is -64 dBc and in-band harmonics are less than -66 dBc or better. Signal-to-noise ratio when measured over the full Nyquist bandwidth is 64 dB.

A Simple Demonstration

I/Q demodulation has been demon-

strated with the circuit in Figure 5 using a PC-based data acquisition system (Figure 7). An HP 3326A synthesizer provides modulated signals which are filtered prior to being sent to the ADC circuit. Filtering removes synthesizer harmonics and wideband noise. A 10 MHz clock is also provided by the synthesizer by level shifting the 10 MHz reference output. This gives us complete control over the LO/carrier fre-

quency ratio (phase is not as easy to control, however).

ADC samples are sent through a line driver, down a cable into a buffer memory. The personal computer accesses the buffer and performs the digital signal processing algorithms using an off-the-shelf DSP package.

For our demonstration we externally amplitude modulated the synthesizer with a 4 kHz sawtooth wave from an



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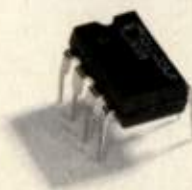
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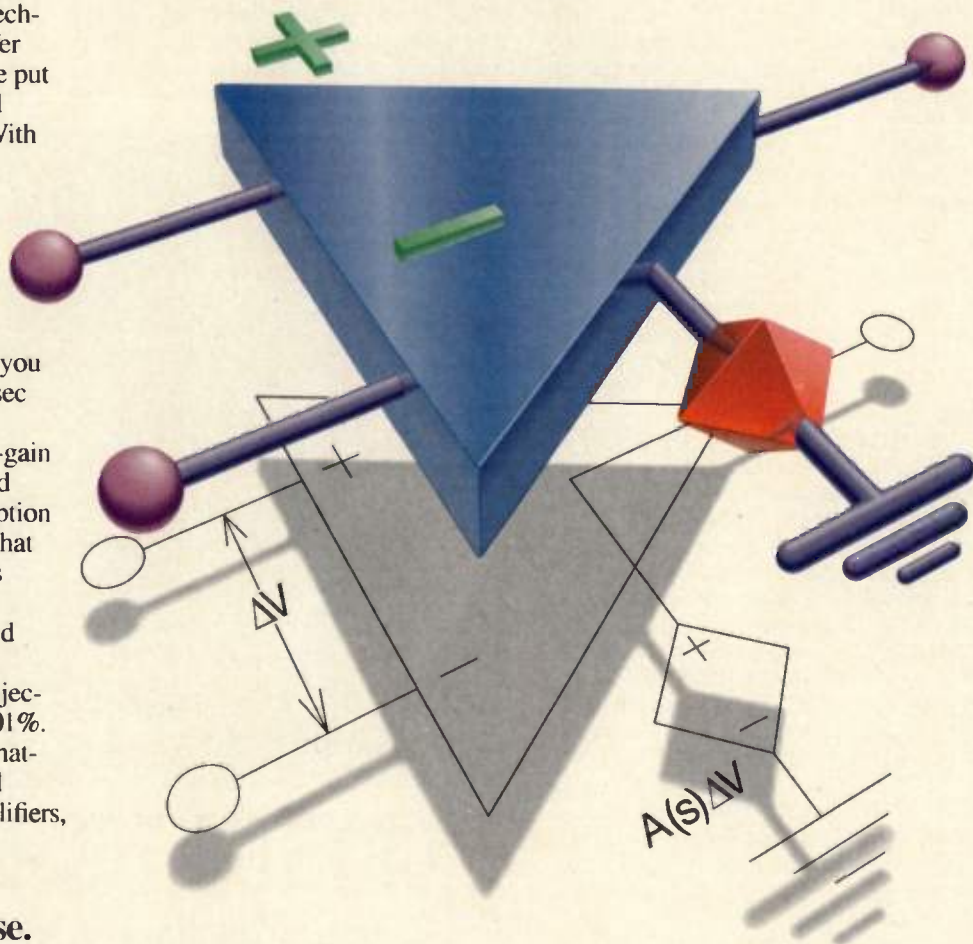
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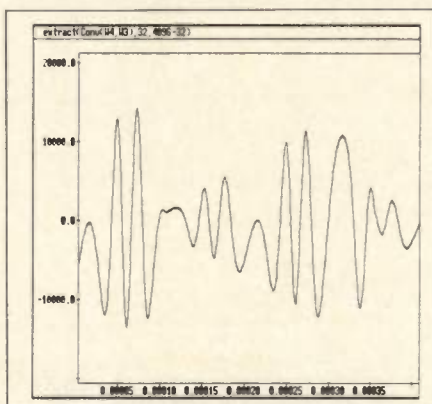


Figure 11a. I and Q signals.

HP8116 function generator and internally phase modulated with a 5 kHz sine wave (peak deviation 2π radians). The center frequency was set to 12.5 MHz.

To understand how to demodulate these signals, let's look at quadrature-carrier description of an arbitrary IF signal (Figure 8). We start by defining axes which represent our in-phase and quadrature local oscillators. An incoming IF signal can be resolved into components which lie along these axes — in fact this is what our mixers will do. The incoming signal has instantaneous amplitude $R(t)$ and phase $\phi(t)$.

To recover the AM signal, recognize that $R(t)$ represents the demodulated AM signal. We will compute $R(t)$ from:

$$R(t) = \sqrt{i(t)^2 + q(t)^2} \quad (1)$$

An important point to make is that we do not need to phase lock our sample clock (thus our local oscillator frequency) to the IF carrier frequency. If the IF carrier and local oscillator frequencies are at different frequencies, the vector representing the IF signal will be spinning relative to the I and Q reference axes — that is the instantaneous phase will be changing. However the equation for $R(t)$ is true no matter what value $\mu(t)$ assumes.

To recover the phase modulation, we could apply the relation:

$$\phi(t) = \arctan \left(\frac{q(t)}{i(t)} \right) \quad (2)$$

but this falls apart for large phase deviations.

A way out of this problem is to develop an FM detector:

$$\phi'(t) = \left(\frac{1}{1 + \left(\frac{q(t)}{i(t)} \right)^2} \right) \left(\frac{i(t)q'(t) - q(t)i'(t)}{i^2(t)} \right) \quad (3)$$

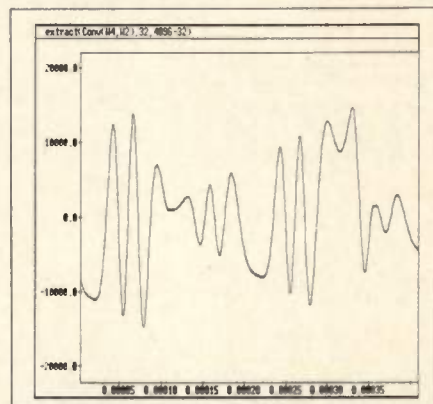


Figure 11b. I and Q signals.

$$= \frac{i(t)q'(t) - q(t)i'(t)}{R^2(t)}$$

The FM detector output is integrated to produce the PM signal. When implemented with DSP, a discrete-time approximation for the derivatives will be used, as will a discrete-time approximation for integration.

The results are given in Figures 9-14. The IF signal is sampled (Figure 9), and mixed as described earlier (by sending alternate samples to the I and Q arrays and alternately digitally multiplying the I and Q samples by 1 and -1). The results are filtered using a 32 order finite-impulse response filter whose response is given in Figure 10. The I and Q signals are shown in Figure 11. Finally, the amplitude modulation is recovered by using Equation 1 (Figure 12) and the phase modulation is recovered by FM detection using Equation 3 (Figure 13) and integrating (Figure 14).

In this example, signal to noise ratio could have been increased by lowering the bandwidth of the low pass filters after the mixers. Assuming a 10 MSPS sample rate, the I and Q channel each receive 5 MSPS for a Nyquist frequency of 2.5 MHz. If we had reduced the low pass filter bandwidths so that the noise equivalent bandwidth was approximately 25 kHz, signal-to-noise ratio would increase by approximately $10\log(100) = 20$ dB for a net signal to noise ratio of over 80 dB. This is based on the fact that noise power spectral density of ADCs are approximately uniform. For other modulation formats, signal to noise ratio improvements (through processing gain) may be available.

The example shows only a fraction of the potential of this technique.

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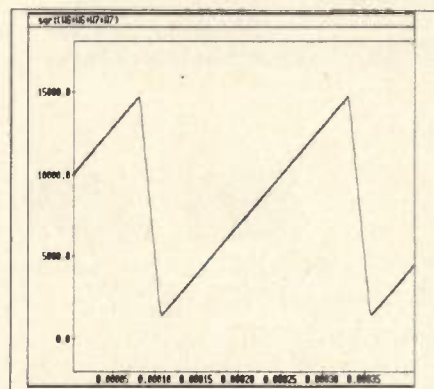


Figure 12. Recovered AM.

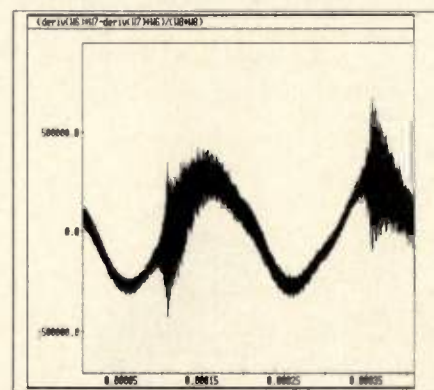


Figure 13. FM detection.

art track-hold amplifiers, such as the CLC940, previously unattainable ADC performance is reached. With this new potential comes the ability to digitally implement I/Q demodulators at common IF frequencies (such as 21.4 MHz).

CLC926-based I/Q sampling when teamed up with modern DSP has the potential for accurate, stable, repeatable, and flexible performance.

Acknowledgement

Thomas Hack would like to acknowledge DSP Development Corporation, Cambridge, Mass., for use of their DADiSP Software in this article. **RF**

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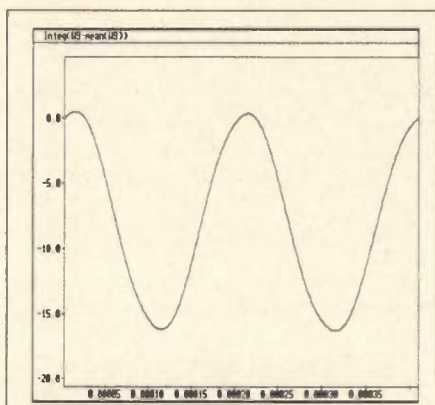


Figure 14. PM detection.

About the Author

Thomas Hack is an Applications Engineer with the Data Conversion Group of Comlinear Corporation. He received his BSEE from Cooper Union, his MSEE from Renaissance Polytechnic Institute, and his MBA from the University of Colorado. He can be reached at (303) 226-0500 and his address is 4800 Wheaton Drive, Fort Collins, CO 80525.

Appendix: A Quick Tour of IF Sampling

Figure 15 diagrams a common way of looking at the sampling behavior of an ADC. $x(t)$, a bandlimited analog signal (from DC to some arbitrary frequency), is multiplied by sampling function $s(t)$ simulating ADC sampling. At an assumed sample rate of 10 MSPS, the ADC aperture is substantially smaller than the sample period so $s(t)$ can be approximated by a series of unit impulses spaced 100 ns apart (Figure 16). The multiplier produces a series of weighted impulses corresponding to discrete-time samples of the analog input. The actual observable samples at the ADC output are of course digital, not analog.

Figure 17 shows the same signals, but in the frequency domain. We will initially assume that the bandwidth of the analog input $X(f)$ is less than 5 MHz (shown by the solid line). The sampling function described earlier produces a comb spectrum $S(f)$ with spikes at DC and integer multiples of 10 MHz. Since the samples' output is produced by

multiplying $x(t)$ by $s(t)$, the frequency domain representation of the output is obtained by convolving $X(f)$ with $S(f)$. The result is the repeating spectrum shown in Figure 17c — a result of harmonic mixing.

Nyquist's sampling theorem recognizes that alias-free sampling occurs when the sidebands at each of the harmonics of the sampling frequency don't share the same frequencies (overlap). For example, if the analog input bandwidth were higher as depicted by the dotted lines in Figure 17, aliasing occurs due to sideband overlapping (shown by the cross hatched

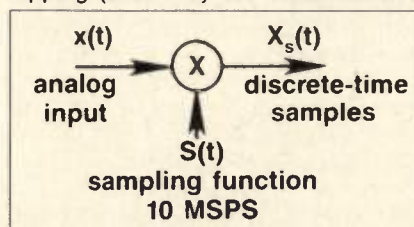


Figure 15. Sampling model.

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regions in Figure 17c).

From this concept comes Nyquist's sampling theorem which states that the sampling frequency must be twice the bandwidth of the signal to be sampled. This is often interpreted to mean that the highest signal frequency must be less than half the sampling frequency or 5 MHz in this example. For video sampling — where the signal extends from virtually zero frequency to some upper frequency, as it has in our example, this would be correct.

But what happens for narrow-band signals, such as we might find in a receiver IF? Figure 18 shows the results from passing a narrow-band IF signal into the same ADC. We start off with an IF signal centered at 17.5 MHz — well above half the sampling frequency. To make things general, the signal is assumed to have independent lower and upper sidebands. As a result of the same sampling action as before, the sampled spectrum has these sidebands repeated at 5 MHz intervals. One of these sidebands will always occur below Nyquist frequency. (Hence IF sampling can replace analog mixers in some instances.)

To prevent overlapping sidebands we must prevent the sidebands from extending greater than ± 2.5 MHz from each center frequency. Since each sideband is independent of the other, the net bandwidth available is the same as before, 5 MHz! In order to attain the alias-free bandwidth claimed by Nyquist's sampling theorem, carefully choose the sampling clock frequency based on the center frequency of the IF.

In the example, we never completely bring the sidebands down to baseband — the lowest mixer products are centered at 2.5 MHz (resulting from mixing of the 2nd harmonic of the sampling frequency with the IF signal). In many instances, there may not be a need to mix to baseband, or it may not

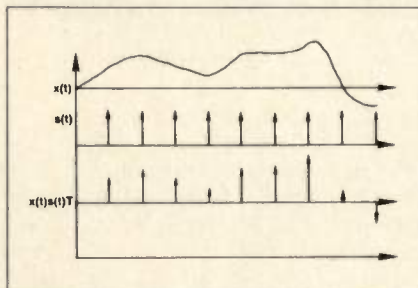


Figure 16. Sampling in the time domain.

be desirable.

A few final points: 1) we could have just as easily placed our signal at another frequency and produced the same output from the ADC without changing the clock frequency. 2) For the interested, a similar analysis can be used to explain sequential repetitive sampling in oscilloscopes.

Our conclusions:

1) IF sampling doesn't violate Nyquist's sampling theorem. The required sample rate is determined from the bandwidth of the IF signal, not the center frequency.

2) To get the most from the tech-

nique, sampling clocks should be carefully selected.

3) To select suitable sampling frequencies determine the lowest sampling frequency based on Nyquist's sampling theorem and the bandwidth of the IF signal. Then look at how the IF signal mixes with harmonics of the sampling clock and adjust the sampling frequency relative to the IF frequency so that sidebands don't overlap.

4) The results are slightly different in the case where the upper and lower sidebands are not independent and our conclusions are altered somewhat.

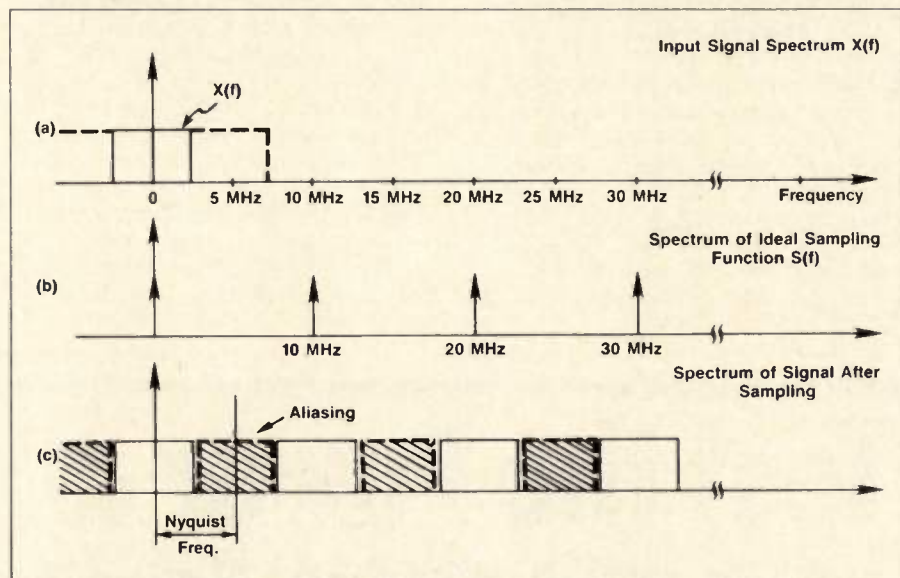


Figure 17. Video sampling: frequency domain representation.

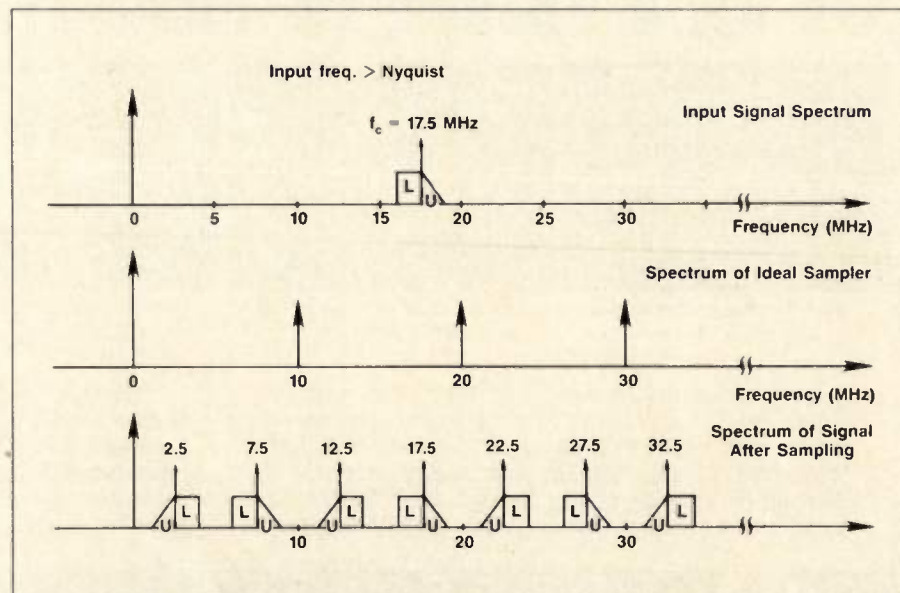


Figure 18. IF sampling: frequency domain representation.

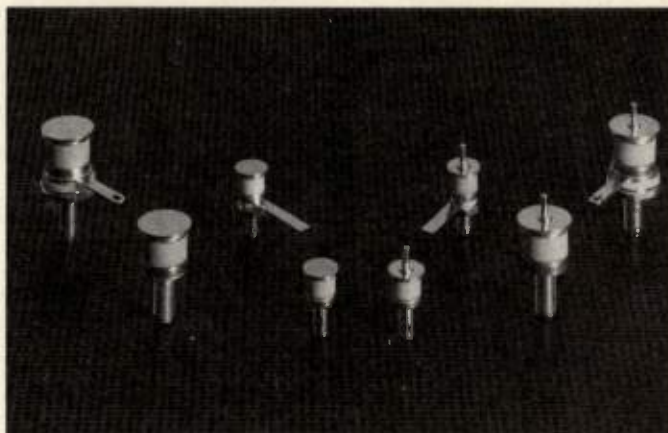
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M/A-COM introduces new KILOVOLT™ PIN diodes, designed for use as high power switching elements in multi-kilowatt HF and VHF applications. MRI switches, filter switches, antenna couplers and power amplifier bypass switching are some of the uses targeted for these components. Long carrier lifetime of 10 usec for the 2000-volt version and 20 usec for the 3000-volt model assures operation at frequencies as low as 1 MHz. Maximum series resistance (4 MHz, with 0.5 A forward current) for each version is .20 ohms and .25 ohms, respectively. The

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M/A-COM Semiconductor Products Div.

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Personal Radiation Safety Monitor

A non-ionizing radiation monitor, the Nardalert™ Personal Monitor, Model 8840B, has been introduced by Loral Microwave-Narda. With this monitor, workers in potentially hazardous areas can have a hands-free, full-time sensor to warn of illumination by high



power radar, communications systems or EW systems. The small, wearable, pocket-sized unit allows accurate monitoring from 2 to 18 GHz for use in areas where conventional meter and probe are not viable solutions for protection. Thermocouple-based detection is used for true RMS average field sensing, avoiding multiple signal or modulation errors which might generate false alarms. Built-In-Test circuitry verifies operation each time it is turned on, and a low battery indicator warns when replacement is required.
Loral Microwave-Narda
INFO/CARD #249

Function Generator Synthesizes Arbitrary Signals

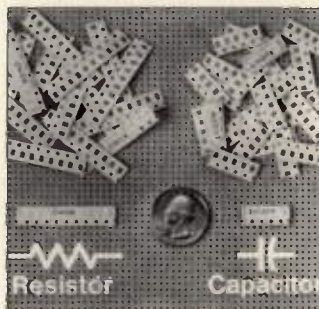
The DS345 Synthesized Function Generator from Stanford Research Systems offers digitally synthesized waveforms to 30 MHz with 1 uHz resolution. Outputs can be simple sine, triangle, ramp or square waves, or arbitrary signals with up to 16,300 (12 bits) and sampling times to 25 nsec can be generated. Internally synthesized modulation capabilities include phase continuous linear and logarithmic frequency sweeps, plus amplitude, frequency, phase and burst modulation. Complex applications include binary shift keying, phase shift keying and arbitrary waveform generation. Rear panel trigger outputs, external input for amplitude modulation, sweep and span markers, IEEE-488 and RS-232 interfaces and external timebase inputs are additional features. Spurious components are better than -55 dBc. Output level is 10 mV to 10 V_{p-p}. List price is \$1895, computer interface with Arbitrary Waveform Composer software is \$495, and a high stability timebase option is \$650.

Stanford Research Systems
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SMT Resistors and Capacitors Available in Small Quantities

Communications Specialists now offers surface mount resistors and components in small quantities and individual values. These components are offered to service shops and design labs which typically cannot meet the minimum purchase requirements of major component distributors. The unit of sale is per strip. Resistors are furnished in strips of 10 and sell for \$2.50 per strip. Capacitors are furnished in strips of 5 and sell for \$1.25 per strip.



Values are clearly marked on each strip. Minimum order is \$10.00. Available resistors and capacitors are the same as those provided in the company's CC-1 and CR-1 prototyping kits, which generally are standard 5 percent resistor values, and 10 percent ceramic capacitor values.

Communications Specialists, Inc.
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PC-Based Counter Operates With Windows 3.0

A new 10 Hz to 2.4 GHz counter has been introduced by Optoelectronics. The PC-10 Universal Frequency Counter-Timer is in the form of a 9-inch card for personal and laptop computers, using Windows 3.0 as a control panel and display window. Input signals of 10 mV or greater are accepted and displayed with up to 10 digit resolution. Input gating is continuously variable from 1 msec to 28 seconds for optimum measurement time and resolution. Measurement modes include discrete and average frequency, pulse width, time interval, period, and ratio between two frequencies. The counter can also control a radio receiver, such as the ICOM R7000, for operation faster than the typical scanner, or automatically tuning the radio to a detected signal frequency for surveillance applications. The PC-10 is priced at \$335 each.

Optoelectronics Inc.
INFO/CARD #246



Frequency Multipliers

KW Microwave announces a series of active and passive frequency multipliers with integrated amplifiers, featuring multiplication factors from 2 to 96, with minimum input frequency of 1 MHz and maximum output frequency of 12 GHz. An example multiplier with a multiplication factor of 16 has an input frequency of 450 MHz and an output of 7.2 GHz. Typical input power is +23 dBm and typical output power is 0 dBm. Spurious and harmonic signals are -50 dBc. Multipliers are available in plug-in, drop-in or connectorized versions.

KW Microwave Corp.
INFO/CARD #245

Large Area Diamond Heat Sinks

Thick, large area chemical vapor deposited (CVD) diamond heat sinks are now being produced for power device and high speed applications. In dimensions up to 25 mm square and 0.5 mm thick,

diamond heat sinks offer thermal conductivity about 2.5 times that of copper and 5 times that of beryllia at 100 C. For price sensitive applications, diamond-coated silicon substrates are also available.

Crystallume
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120 MHz Synthesizer IC

GEC Plessey Semiconductors introduces the NJ88C33 synthesizer, designed for operation from 2.5 to 5.5 volts with 3 mA current consumption. The device may be operated as a single-chip 120 MHz synthesizer, or with an external prescaler for operation to



frequencies beyond 2 GHz. On-chip are programmable 16-bit 'R', 12-bit 'N' and 7-bit 'A' counters which are addressed by an I2C bus. The bus will operate up to 2 MHz and can achieve channel loading in 20 us. In 100s, the price is \$7.87.

GEC Plessey Semiconductors
INFO/CARD #243

GaAsFET Switch Matrix

A 5-in, 2-out GaAsFET switch matrix is introduced by K&L Microwave, covering DC-2 GHz, with TTL compatible control inputs. Switching speed is typically 40 nsec, insertion loss is 4.5 dB maximum, isolation is 60 dB, and VSWR is a maximum of 1.5:1. The matrix is enclosed in a hermetically sealed 3 x 1.5 x 0.4 inch enclosure.

K&L Microwave, Inc.
INFO/CARD #242

Communications Test Set

Motorola announces the R-2600A Communications System Ana-

lyzer, with cellular testing capability. Available cellular formats include EAMPS, ETACS, JTACS and NTACS. Test functions include sweep generator, spectrum analyzer, receiver, digital storage oscilloscope, terminated wattmeter, counter, signalling generator and decoder, SINAD meter, distortion meter, AC/DC voltmeter, and audio generator. The display includes softkeys, windowing and fast autoranging scales for ease of operation.

Motorola Government Electronics Group
INFO/CARD #241

DC-4 GHz Differential Amplifier

The AM-4001, a stand-alone differential amplifier has been introduced by Tektronix. This pulse amplifier has a DC-4 GHz bandwidth with 18 dB gain and gain flatness of less than 1 dB from 1 MHz to 2.5 GHz. Available output level is ± 2 V into 50 ohms from each output. The amplifier is designed for applications requiring accurate time domain characteristics such as preamplification

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Vari-L Company, Inc.
INFO/CARD #237

PC-based Counter

The GT100 from Guide Technology is a low cost frequency counter for automated test systems. The counter includes a full-size PC card and a DOS-based software package which provides a virtual front panel on any XT, AT or compatible PC. 100 MHz or 1.3 GHz models are available, with selectable gate time, fast frequency measurement, time interval, pulse width, period, ratio, and totalize functions. Price of the GT100 is \$995 for the 100 MHz model and \$1745 for the 1.3 GHz version.

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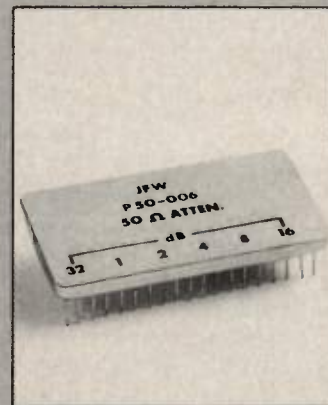
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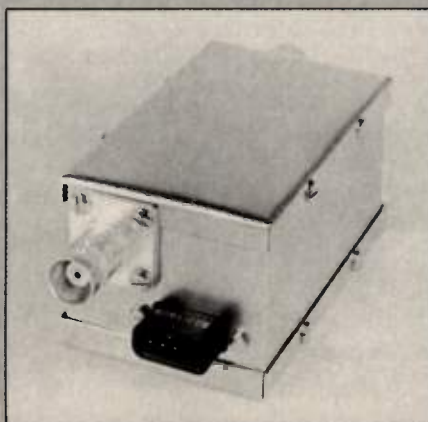
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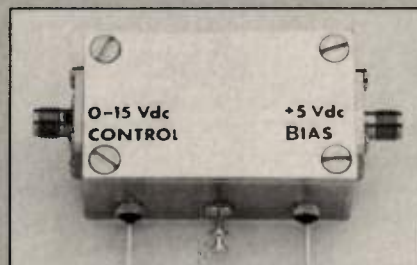
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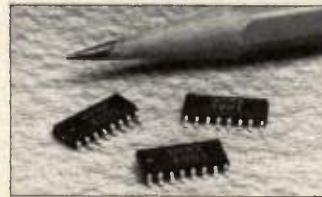
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The first of a new series of RF ICs, the HPMX-2001 QPSK modulator has been introduced by Hewlett-Packard Co. The device incorporates twin double balanced modulators cross-coupled into a summing amplifier, plus a standby circuit for controlling current sources. Applications include digital cellular and digital cordless telephone equipment, as well as other applications requiring



amplitude modulation, single-sideband transmission or vector generation. Pricing is \$7.25 for 1-99 units.

Hewlett-Packard Company
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1 MV Pulser System

A pulser system generating 250 kV to 1 MV pulses with a risetime of 65 nsec and duration of up to 10 ms is announced by Maxwell Laboratories. The Model 40404 Marx Generator can be used as in high power microwave sources, for accelerating electron beams, in pulsed lasers, flash X-ray units or for general high-voltage laboratory use. Low jitter and droop, and low overshoot are additional features of this design.

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INFO/CARD #229

Tunable Notch Filters

A new line of tunable notch filters from Trilithic covers a range of 25 to 1000 MHz with octave band units, offering notch depths up to 50 dB. 3 dB bandwidth varies from 1 to 16 MHz, depending on the octave band selected, while

passband loss is 1 dB maximum. Models are available in 50 or 75 ohm impedances, with type N, SMA, TNC or BNC connectors.

Trilithic
INFO/CARD #228

Mica Chip Capacitors

Type MC mica chip capacitors are now available from Cornell Dubilier. Natural mica dielectric retains high Q up to self-resonant frequencies exceeding 1 GHz. Capacitance ranging from 1 pF to 2000 pF (100 V) or 1200 pF (500 V) and tolerances as tight as ± 0.25 pF or ± 0.25 percent are offered. Bulk or reel packaging is available. OEM quantity pricing starts at \$0.20 each.

Cornell Dubilier Electronics
INFO/CARD #227

RF Prototyping Board

Ivanboard is a universal circuit board for quick, reliable construction of all types of electronic circuits. Using a grid of pads with 0.1 inch spacing on one side, and a solid low-impedance ground plane on the opposite side, stable

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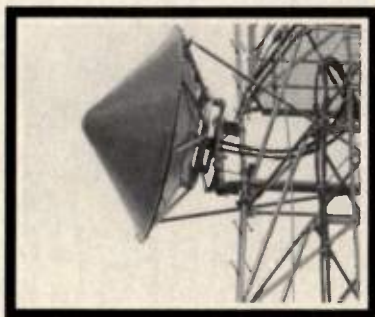
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INFO/CARD 42



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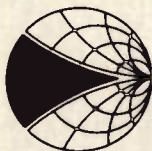


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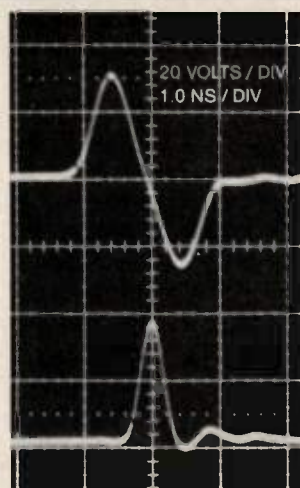
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INFO/CARD 78

RF, analog and digital circuits can be built.

Europe Marketing
INFO/CARD #226

Attenuator Uses ACT Drivers

KDI/Triangle's DAP0024 thick film attenuator uses advanced CMOS-TTL compatible (ACT) drivers to achieve 100 ns switching speed in a 0.-15.5 dB, 0.5 dB per step attenuator. Operating frequency is 20 MHz to 1 GHz, with other models operating to 2 GHz. Attenuation accuracy is ± 0.25 dB.

KDI/Triangle Electronics
INFO/CARD #225

SMT Inductors

The PM20 series SMT inductors from J.W. Miller cover 0.01 μ H through 220 μ H with current ratings from 450 mA to 50 mA (over the same inductance range). Operating temperature is specified as -20 C to $+100$ C. 1210 outline, precise dimensions and soldering terminal strength facilitate automated production.

J.W. Miller Div., Bell Industries

INFO/CARD #224

Microcomputer Compensated Oscillator

Frequency Electronics introduces its MCXO Microcomputer Compensated Crystal Oscillator for precision applications. The MCXO achieves compensation without the use of ovens or conventional temperature-compensation techniques. It provides a order-of-magnitude improvement in frequency stability for low power, high-accuracy timekeeping and frequency control. Stability with temperature is $\pm 3 \times 10^{-8}$ including hysteresis, over -55 to $+85$ C. Standard outputs are time-of-day, 3.2 MHz, and 1 pulse/sec. Optional spectrally pure sine wave outputs or square wave outputs in the range of 1-100 MHz are also available.

Frequency Electronics Inc.
INFO/CARD #223

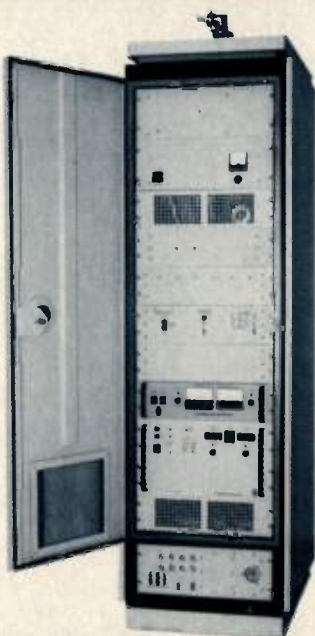
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INFO/CARD 33



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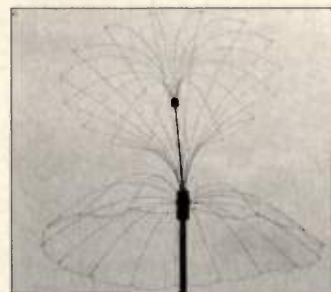
INFO/CARD 45

IVA-05208, a variable gain amplifier with DC to 1.5 GHz -3 dB bandwidth, 30 dB gain, and 30 dB gain control range. The device operates from a single 5 VDC, 35 mA supply, and uses a 0 to 5 V control voltage. Packaging is in a standard low cost SO-8 8-lead SMT package. The price is \$10.45 in 100-piece quantities. The IVA-05208 is in stock at Avantek distributors.

Avantek, Inc.
INFO/CARD #222

Rapid Deployment Antenna

The Astron Model WBJ-250B Broadband HF/VHF/UHF antenna covers 20 to 520 MHz



without the use of couplers or tuning networks. VSWR is 2.5:1 maximum, and the antenna will handle 50 watts input power. The antenna is rapidly deployed in the field for both transmit and receive operation. Deployed height is 12.3 feet.

Astron Corp.
INFO/CARD #221

Sealed Trimmers

Voltronics is now producing its smaller "K" line precision air dielectric trimmers with either an alumina or plastic case. Both designs are O-ring sealed, and are offered in 8 or 10 pF values. Q is over 5000 at 100 MHz, and self-resonant frequency is over 2 GHz at maximum capacitance. The case is 0.23 in. diameter by 0.48 in. long, provided in panel mount, strip line, surface mount and p.c. mounting. 1000-piece quantities are priced at \$4.23 for the 8 pF ceramic case (KP8) and \$3.96 for the 8 pF plastic unit (KEP8).

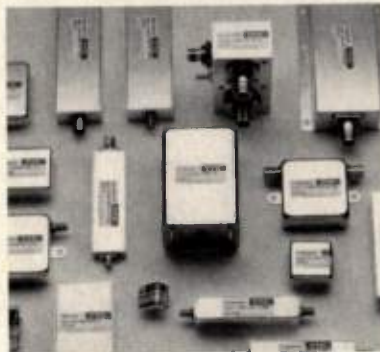
Voltronics Corp.
INFO/CARD #220

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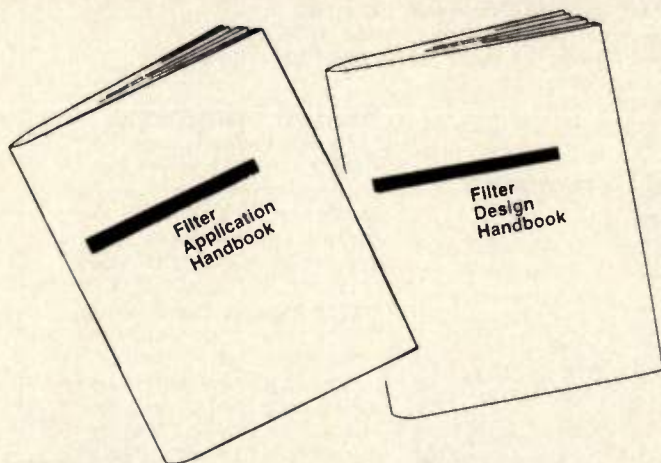


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