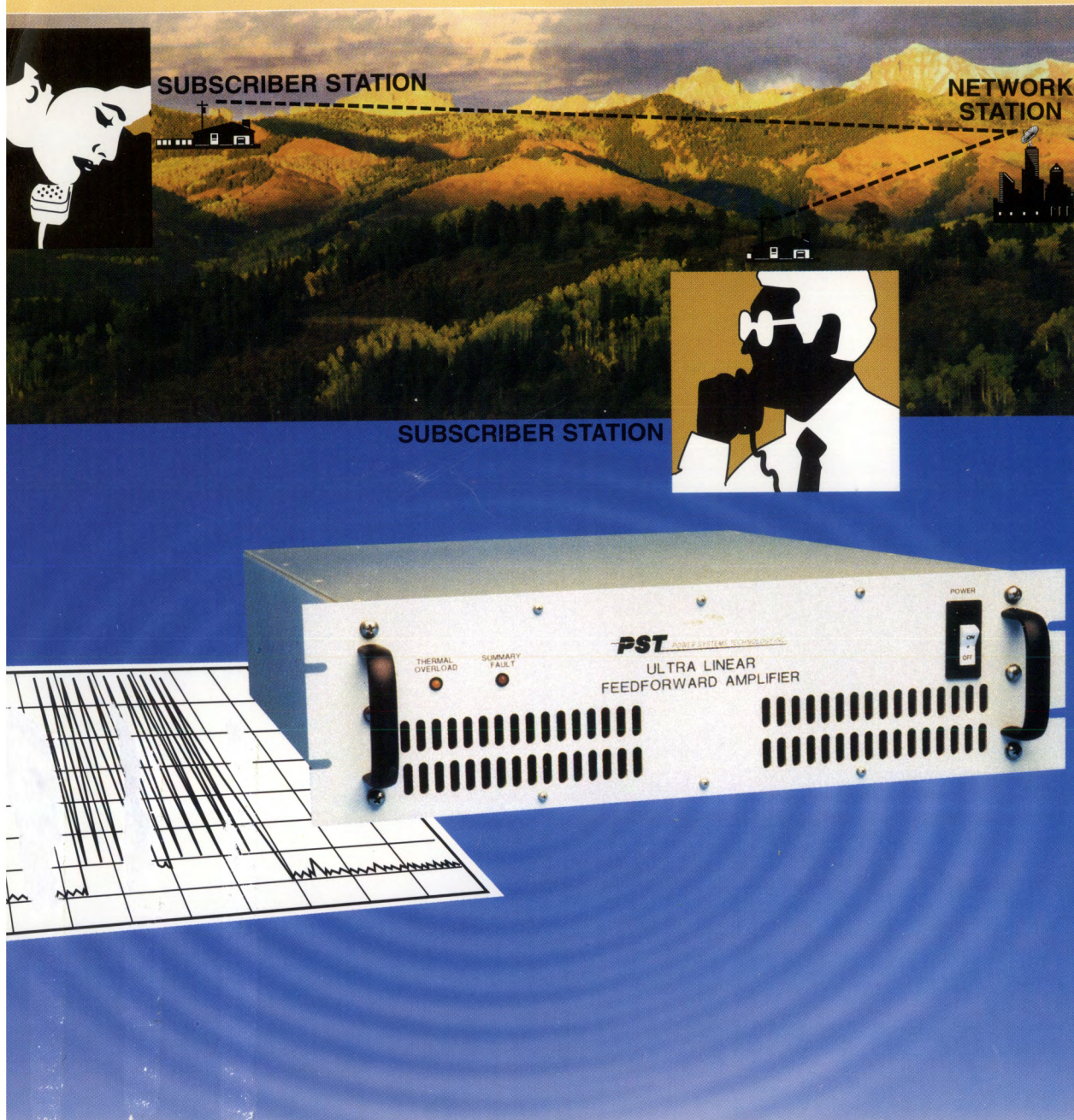


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engineering principles and practices

July 1994



*Cover Story*

**Power Amplifier Cuts Distortion**

*Featured Technology*

**Open-Loop Digital Modulation**

**A Fast Envelope Detector**



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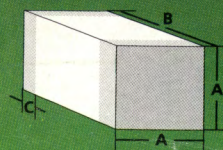
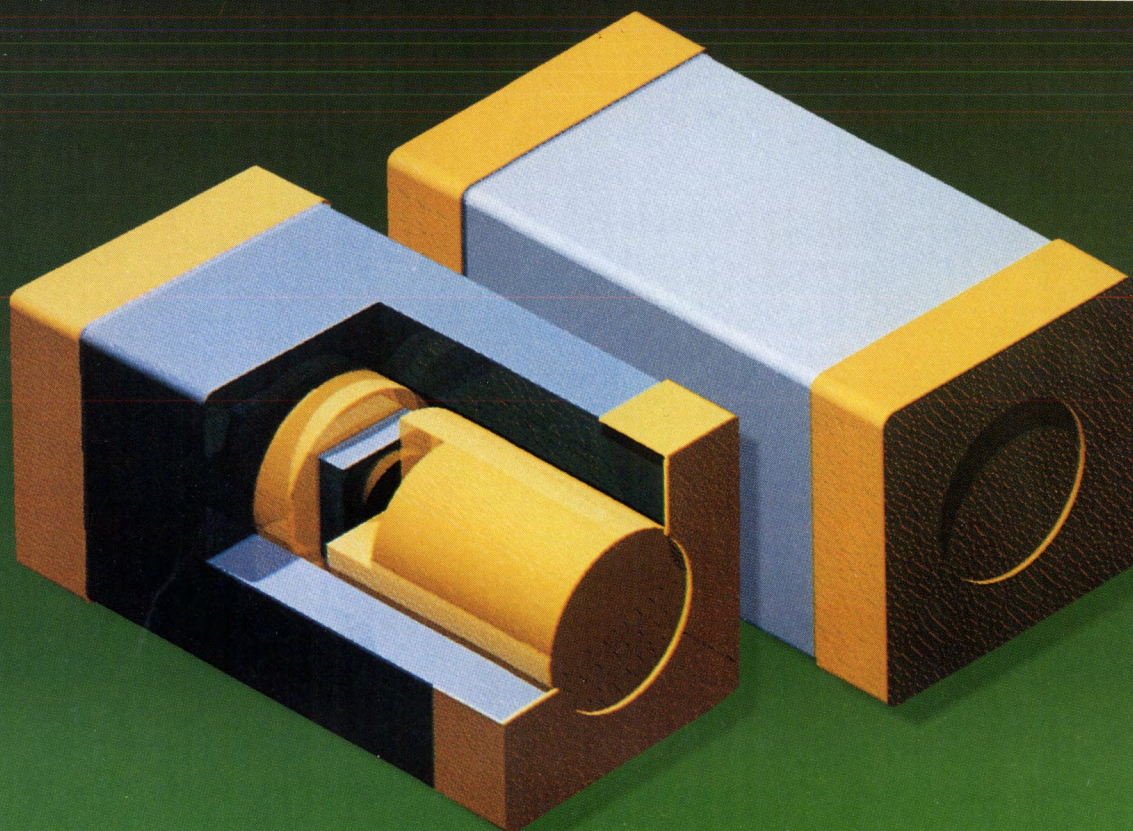
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Part No.	Suggested Case Style	Reverse Voltage Rating (Vr) – min  Ir < 10µA	Total Capacitance (pF) – max F=1MHz Vr=50v		Series Resistance (RS)–ohms F=100 MHz		Carrier Lifetime (TI) µsec (typ)  If=10mA	Thermal Resistance (Θj)  C/W
			M1	M2	If=100mA (max)	If=200mA (typ)		
SM0502	M1 or M2	500	0.45	0.50	0.70	0.55	1.00	35
SM0504	M1 or M2	500	0.55	0.60	0.60	0.45	1.50	20
SM0508	M1 or M2	500	0.85	0.95	0.40	0.25	2.00	15
SM0509	M1 or M2	500	1.00	1.25	0.35	0.20	2.50	12
SM0511	M1 or M2	500	1.30	1.45	0.30	0.15	3.00	9
SM0512	M2	500	N/A	1.50	0.25	0.12	3.50	7
SM0812	M2	800	N/A	1.30	0.40	0.25	4.00	7
SM1001	M2	1000	N/A	1.30	0.35	0.20	4.50	7

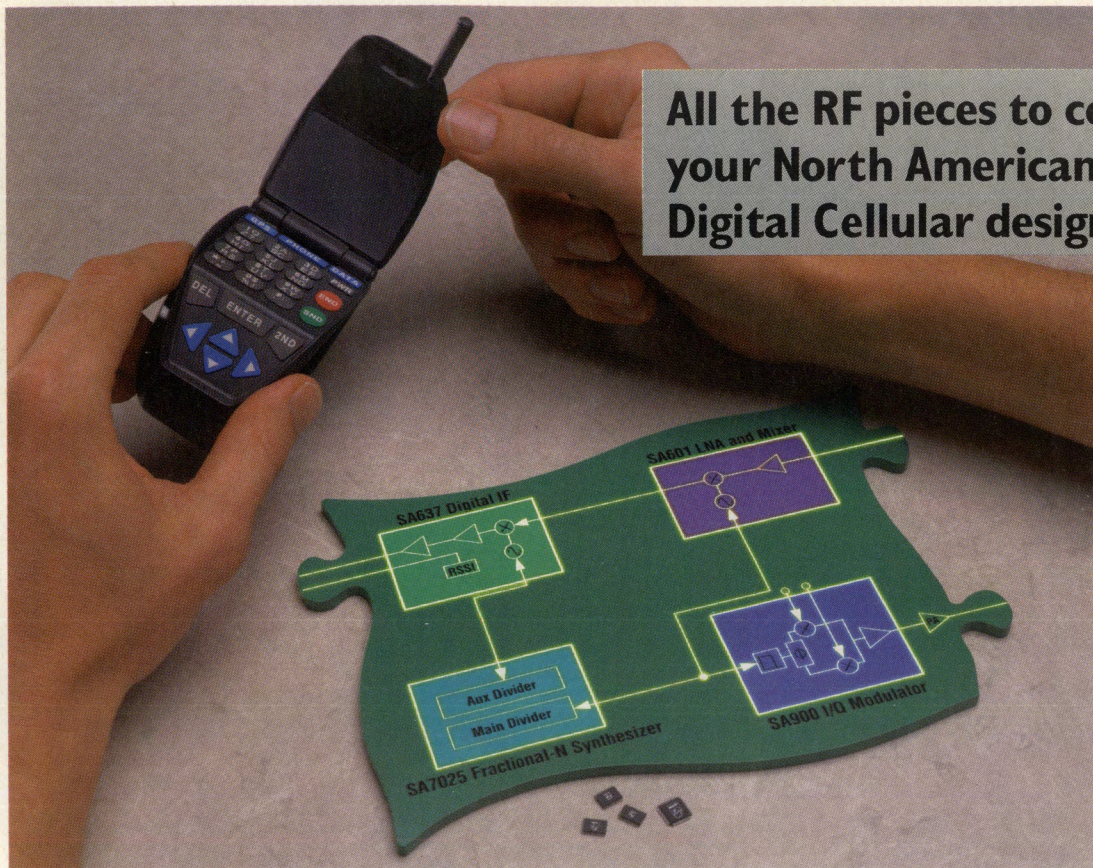


Case style	M1 (inches)		M2 (inches)	
Dimension	Min	Max	Min	Max
A	0.08	0.10	0.10	0.12
B	0.12	0.14	0.19	0.21
C	0.01	0.03	0.01	0.03

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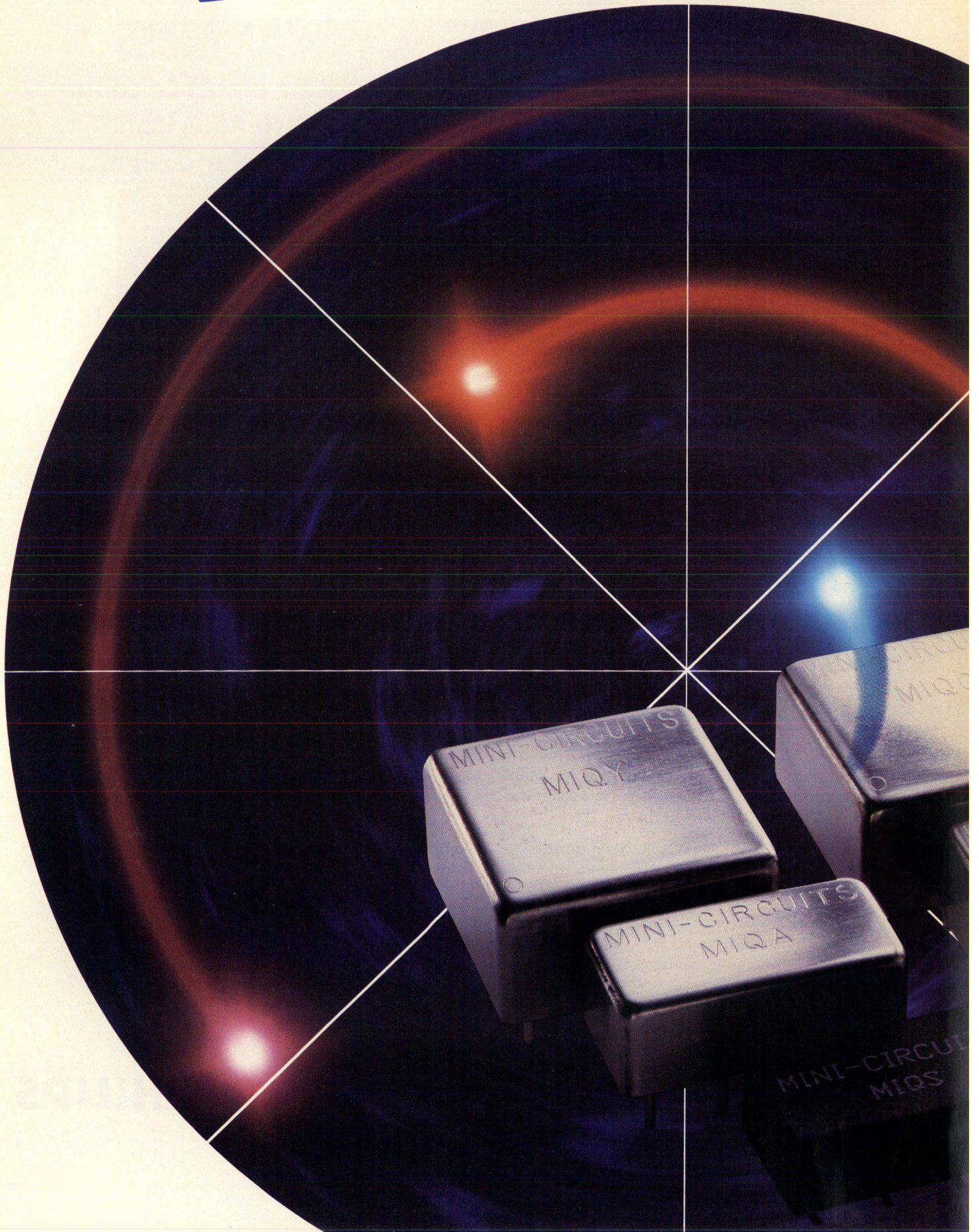
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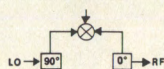
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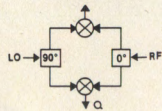
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	$f_L$	$f_U$	$x$	$\sigma$			3xl/Q	5xl/Q	
MIQA-10M	9	11	5.8	0.20	41	40	58	68	49.95
MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
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MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
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MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95



## I/Q DEMODULATORS

MODEL NO.	FREQ. (MHz)		CONV. LOSS (dB)		AMP. UNBAL. (dB) Typ.	PHASE UNBAL. (Deg.) Typ.	HARM. SUPPRESS (dBc) Typ.		PRICE \$ QTY (1-9)
	$f_L$	$f_U$	$x$	$\sigma$			3xl/Q	5xl/Q	
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQC-895D	868	895	8.0	0.20	0.15	1.5	40	55	99.95
MIQY-1.25D	1.15	1.35	5.0	0.10	0.15	1.0	59	67	29.95
MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
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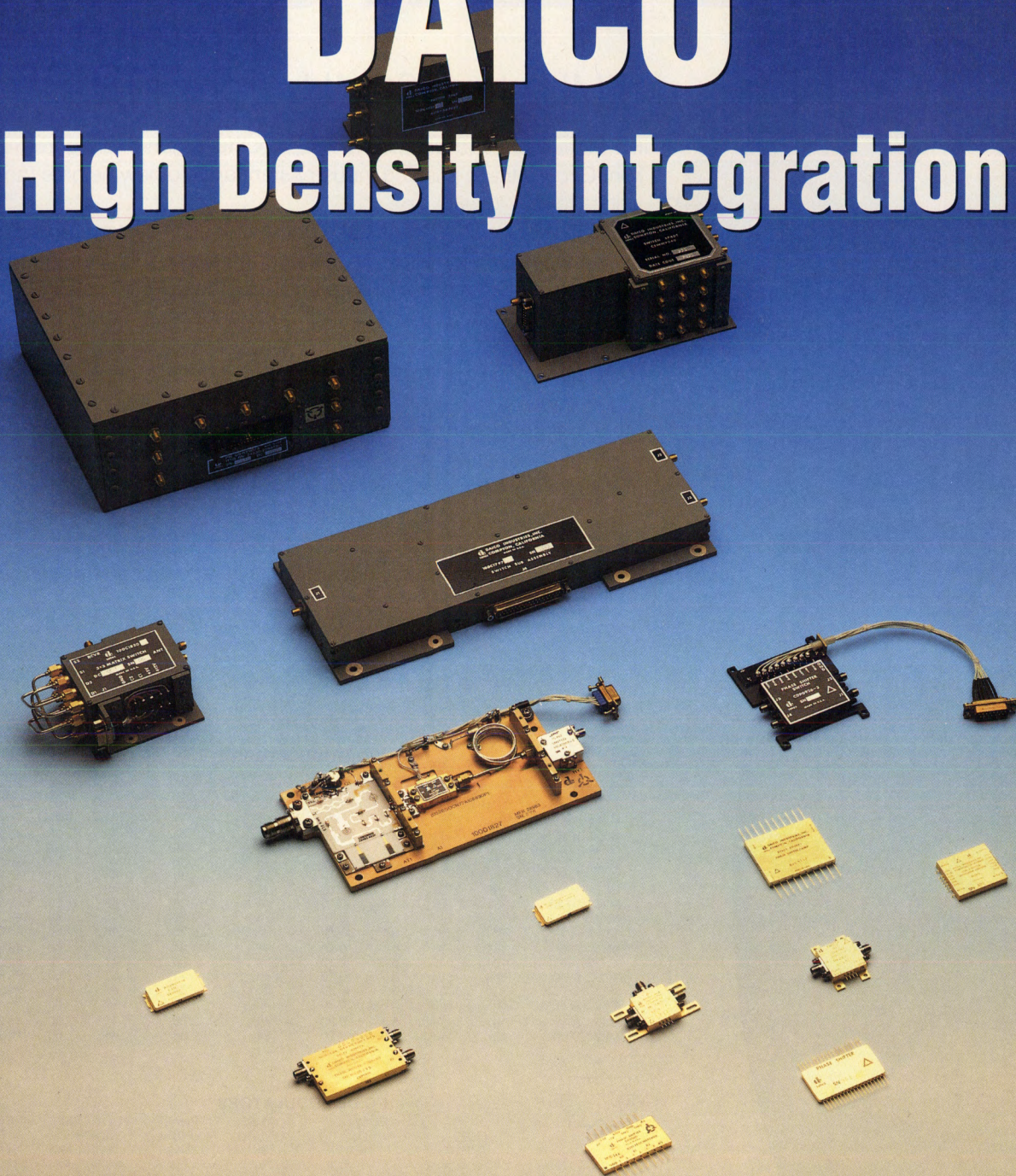
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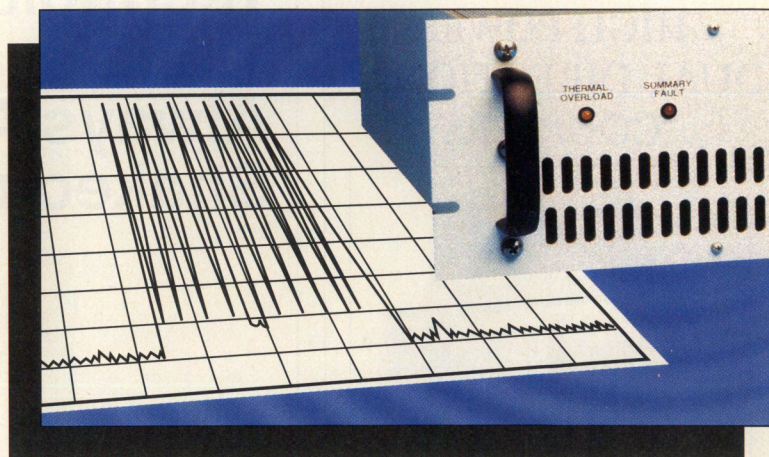


### featured technology

#### 26 Open Loop Modulation of VCOs for Cordless Telecommunications

New digital cordless phones such as those for Digital European Cordless Telecommunications (DECT) and the upcoming North American digital standard require wideband FM modulation during brief bursts. The PLL loop filter bandwidth in these systems is much less than the modulation bandwidth, but by briefly opening the loop before a transmission burst, the modulation can be accomplished.

— Daniel E. Fague



#### 34 A Fast Envelope Detector

As fast as quadrature signals can be squared, added and the sum's square root taken, this detector can demodulate AM signals. The response time of the detector described here is on the order of 1  $\mu$ s.

— Dominic J. Ciardullo

#### 44 An AM Detector Compatible with SSB-SC

AM signals are demodulated with this detector by multiplying carrier and sideband components of an IF signal. If the carrier signal is at least 7 dB above noise, this detector can demodulate SSB-SC as well.

— Chase P. Hearn

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### cover story

#### 48 A High Power, Low Distortion Feed-Forward Amplifier

Class AB amplifiers are more efficient than class A amplifiers, but they are less linear. A high-power, multichannel amplifier using a class AB final stage achieves its high linearity by cancelling intermodulation products

— Walter Koprowski

### tutorial

#### 56 Transmission Line Fundamentals

Transmission line behavior can seem mysterious if the equations that describe transmission line phenomena seem to come from nowhere. This tutorial starts from a basic transmission line model and ends with the commonly encountered transmission line equations.

— Andy Kellett

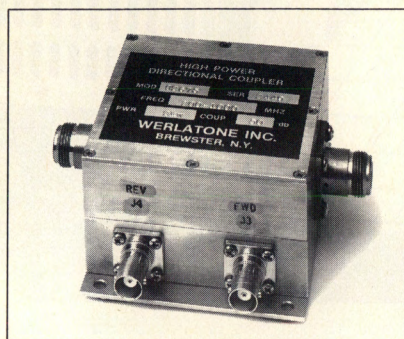
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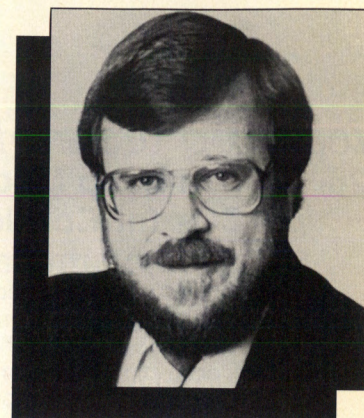
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## RF editorial

# With New Technology, There's Always A Shakeout

By Gary A. Breed  
Editor



One of the most fascinating aspects of the so-called "wireless revolution" is the wonderfully creative proliferation of new ideas. Engineering and marketing staffs at RF manufacturing companies have come up with some outstanding capabilities, and their customers (and potential customers) are thinking up new ways that RF linked communications can help them be more cost-effective in their business operations.

New players are getting involved, too. Component manufacturers that did little business in the realm of RF applications are investing millions of dollars to develop RF products. Computer and data communications companies are adding RF capabilities in addition to their twisted-pair or coaxial cable networking expertise. Customers are coming from new directions, too, mainly from the world of computers because the bulk of new ideas are directed toward data transmission.

With many new ideas, the potential market for RF products has grown dramatically. The possibility of a huge boom, similar to the personal computer, seems apparent to most observers (including me). But there is one nagging thought that keeps coming back...

How many of these new RF applications will become successful products?

Debate rages over regulations, standards, and frequency allocations. While resolution of these matters is certainly essential to opening new markets, the ultimate judgement comes from the customers — the "marketplace."

As new RF products go into large-scale production, a shakeout is inevitable. It happens whenever new

technology is brought to market. It may be slow, it may be rapid, but it *will* happen. Millions will be made and lost as products are accepted or ignored.

We can look at the personal computer industry for proof. There is no more S-100 bus and no CP/M operating system. Even mainframes and their smaller mini-computer counterparts have been replaced by PCs for many applications.

We have seen similar development within our own industry. GaAs technology, SAW devices and power transistors have all seen huge successes and agonizing failures as the market decided which approaches and which applications needed these products.

Which RF products are going to be as successful as the DOS-based PC? Which will rise and fall like the Commodore 64? Which will die on the loading docks before they even get shipped?

The mystery is part of the fun! It's sad to see innovative companies fail because of a fickle market, or simply because they were a little too early or too late with their product. But, hey, that's how business works.

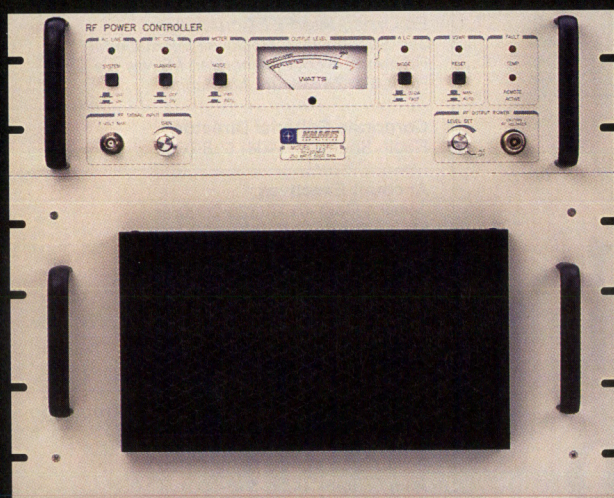
On the other hand, we are occasionally astounded by the success of a little company that was "in the right place at the right time." It's also fun to watch a product take off when market research never predicted that kind of success.

We will define the RF boom by the success of products that are accepted by customers and generate profits for their manufacturers. But, we need to remember that many innovations, companies, and people will be counted as failures, too. Business is a battleground, and battles always have casualties.



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**Editorial and Advertising Offices**  
6300 S. Syracuse Way, Suite 650  
Englewood, CO 80111  
(303) 220-0600 Fax: (303) 267-0234

**Vice President and Group Publisher**  
David Premo (303) 220-0600

**Editor and Associate Publisher**  
Gary A. Breed (303) 220-0600

**Technical Editor**  
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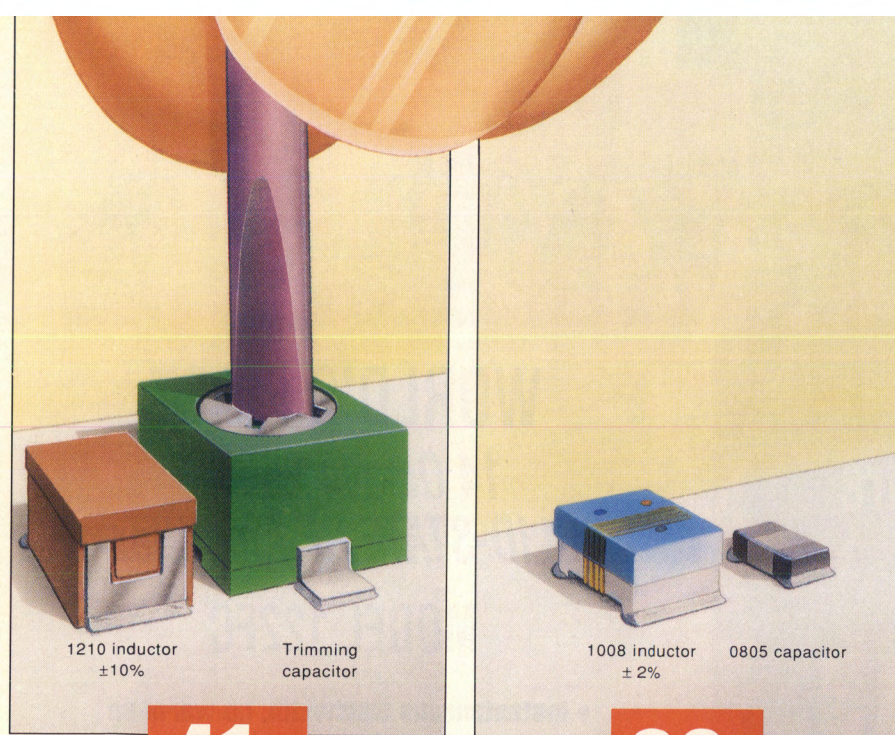
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# Giga-tronics Presents The Alternative To \$30,000 And \$40,000 RF Synthesizers.

Hewlett-Packard makes a couple of very good RF synthesizers. And if you can afford the luxury of paying \$30,000 or \$40,000 for the name, by all means, call HP right now. They'll be happy to take your order, and your money.

However, if you're looking for an RF synthesizer with outstanding performance and proven reliability for about half the price, you'd better call Giga-tronics.

Here's why:

## Performance.

Check the charts. In virtually every category, the Giga-tronics 6080A and 6082A RF Synthesizers meet or exceed the specs of the HP machines. And they use the same GPIB command set, for direct replacement without expensive new software.

## Experience.

Granted, Hewlett-Packard has been around a long time. But, Giga-tronics

is no Johnny-come-lately.

Giga-tronics has a 14-year history of building test and measurement gear for the most demanding requirements. We've shipped thousands of instruments for use in the testing of radar, EW and communications systems.

## Reliability.

Making reliable RF synthesizers is usually no fluke.

However, in this case, it is.





Both the 6080A and 6082A were originally introduced in 1990 by John Fluke Manufacturing Company. To date, thousands have performed flawlessly in the field.

For added confidence, the instruments incorporate self-testing, internal diagnostics and modular design for easy fault isolation and repair.

#### Service.

If a problem occurs, Giga-tronics technical support staff can often help you find and fix the problem over the phone.

If you need to return an instrument for repair, we can service it at our factory in California, or at one of our worldwide sales and service centers.

But at Giga-tronics, customer service starts even before you become a customer.

Whether you're looking to buy one unit or one hundred, you'll get the same assistance, including a demonstration at your facility.

#### Price.

Considering all this, the real question is not why Giga-tronics is so much less, but rather, why Hewlett-Packard wants so much more?

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Spurious	<-100 dBc	<-100 dBc	<-94 dBc	<-94 dBc
Subharmonics	None	None	<-45 dBc	<-45 dBc
Phase Noise*				
@ 20 kHz offset	<-134 dBc/Hz	<-131 dBc/Hz	<-125 dBc/Hz	<-125 dBc/Hz
Residual FM*				
(.3 to 3 kHz BW)	<2 Hz	<1.5 Hz	<5 Hz	<3 Hz
Output Range*	+16 to -140 dBm	+17 to -140 dBm	+16 to -140 dBm	+13 to -140 dBm
Accuracy	±1 dB >-127 dBm	±1 dB >-127 dBm	±1 dB >-127 dBm	±1 dB >-127 dBm
Reverse Power Protection	50 Watts/50 Vdc	50 Watts/50 Vdc	25 Watts/25 Vdc	25 Watts/25 Vdc
Amplitude Modulation				
Depth	0-99.9%	0-99.9%	0-99.9%	0-99.9%
Distortion @ 30%	<2%	<1.5%	<2%	<1.5%
Frequency Modulation				
Max. Deviation*	3 MHz	4 MHz	3 MHz	8 MHz
Distortion	<2%	<1% @ 50% Dev.	<2%	<1% @ 50% Dev.
Phase Modulation				
Max. Deviation*	100 Rad.	40/400 Rad.	200 Rad.	80/800 Rad.
Pulse Modulation				
On/off	>40 dB	>40/60 dB	>40/80 dB	>80 dB
Rise/fall time	<400 ns	<15 ns (Typ 7.5 ns)	<400 ns	<15 ns (Typ 7.5 ns)
Minimum Pulse Width	<2 µs	<30 ns	<2 µs	<30 ns
Internal Modulation Source	20 Hz to 100 kHz	0.1 Hz to 200 kHz	20 Hz to 100 kHz	0.1 Hz to 200 kHz
Level Range	0 to 3 Vpk	0 to 4 Vpk	0 to 3 Vpk	0 to 4 Vpk
Waveforms	Sine	Sine/Sq/Tri/Pulse	Sine	Sine/Sq/Tri/Pulse
Programmable	Yes	Yes	Yes	Yes
Memory Locations (NVM)	51 Full Function	50 Full Function	51 Full Function	50 Full Function
U.S. List Price	\$30,340	\$16,950	\$41,680	\$22,950

The question is not why Giga-tronics is so much less,

but rather, why Hewlett-Packard wants so much more.

\*Specifications for both the 6080A and the HP 8642A are at 1GHz. Specifications for both the 6082A and the HP 8642B are at 2GHz. Prices and specifications for the HP 8642A and HP 8642B are from the Hewlett-Packard 1993 catalog. Prices for the Giga-tronics 6080A and 6082A are U.S. list prices.

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# RF calendar

## July

**4-7 HF Radio Systems and Techniques**  
York, UK  
Information: HF '94 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, UK. Tel: 44-071-344 5478/5477. Fax: 44-071-497 3633.

**6-8 2nd International Conference on ADDA**  
Cambridge, UK  
Information: ADDA '94 Secretariat, IEE Conference Services, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 344 5478/5477. Fax: 071 497 3633.

## August

**9-12 IREECON 1994 - International Electronics and Telecommunications Exhibition**  
Melbourne, Australia  
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

**25-28 IEEE Electromagnetic Compatibility Symposium**  
Chicago, IL  
Information: Thomas Braxton, Vice-Chair, AT&T Bell Laboratories, Room 2B-217, 2000 N. Naperville, Road, Naperville, IL 60566. Tel: (708) 979-1299. Fax: (708) 979-5755.

**29-1 Surface Mount International**  
San Jose, CA  
Information: Institute for Interconnecting and Packaging Electronic Circuits, 7380 N. Lincoln Avenue, Lincolnwood, IL 60646. Tel: (708) 677-2850. Fax: (708) 677-9570.

## September

**5-8 The European Microwave Conference 1994**  
Cannes, France  
Information: Jacqueline Baron, Sales Manager, 24th EuMC, Nexus Business Communications Ltd., Warwick House, Azalea drive, Sawanley, Kent BR8 8HY, UK. Tel: 44 322 660070. Fax: 44 322 667633.

**27-29 16th Piezoelectric Devices Conference**  
Kansas City, MO  
Information: Electronic Industries Association, 2001 Pennsylvania Avenue, N.W., Washington, DC 20006. Tel: (202) 457-4930. Fax: (202) 457-4985.

## October

**3-7 Antenna Measurement Techniques Association**  
Long Beach, CA  
Information: 1994 AMTA Symposium, School of Engineering and Computer Science, Center for Research and Sciences, California State University, Northridge, 18111 Nordhoff St. - SECS, Northridge, CA 91330. Tel: (818) 885-2146. Fax: (818) 885-2140.

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# RF courses

## **Introduction to Electromagnetic Compatibility Design Practices**

July 7-8, 1994, Milwaukee, WI

Information: University of Wisconsin - Milwaukee, Non-Credit Registration Office, Drawer No. 491, Milwaukee, WI 53293.  
Tel: (414) 227-3200 or (800) 638-1828. Fax: (414) 227-3146.

## **Design for Testability and for Built-in Test**

July 11-14, 1994, Los Angeles, CA

## **Wavelet Transform: Techniques and Applications**

September 12-16, 1994, Los Angeles, CA

Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825-1047. Fax: (310) 206-2815.

## **Avionics & Weapons Systems Flight Test**

August 22-26, 1994, San Diego, CA

## **Navstar/GPS: Design Applications**

August 1-3, 1994, Washington, DC

## **High Speed & Microwave Devices & Applications**

October 24-27, 1994, Boston, MA

Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436.  
Tel: (818) 995-6335. Fax: (818) 995-2932.

## **Applied RF 1**

August 22-26, 1994, Los Altos, CA

## **Wireless Systems**

August 29-September 2, 1994, Los Altos, CA

Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949-3300. Fax: (415) 949-4400.

## **BiCMOS Process Integration and Engineering**

August 1-5, 1994, Troy, NY

Information: Rensselaer Polytechnic Institute, Office of Continuing Education, Troy, NY 12180-3590. Tel: (518) 276-8351.

## **Digital Cellular and PCS Communications - The Radio Interface**

October 10-14, 1994, Spain

## **RF/MW Circuit Design: Linear/Non-Linear, Theory and Applications**

October 10-14, 1994, Spain

## **Active and Passive RF Components: Measurements, Models, and Data Extraction**

October 12-18, 1994, Spain

Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

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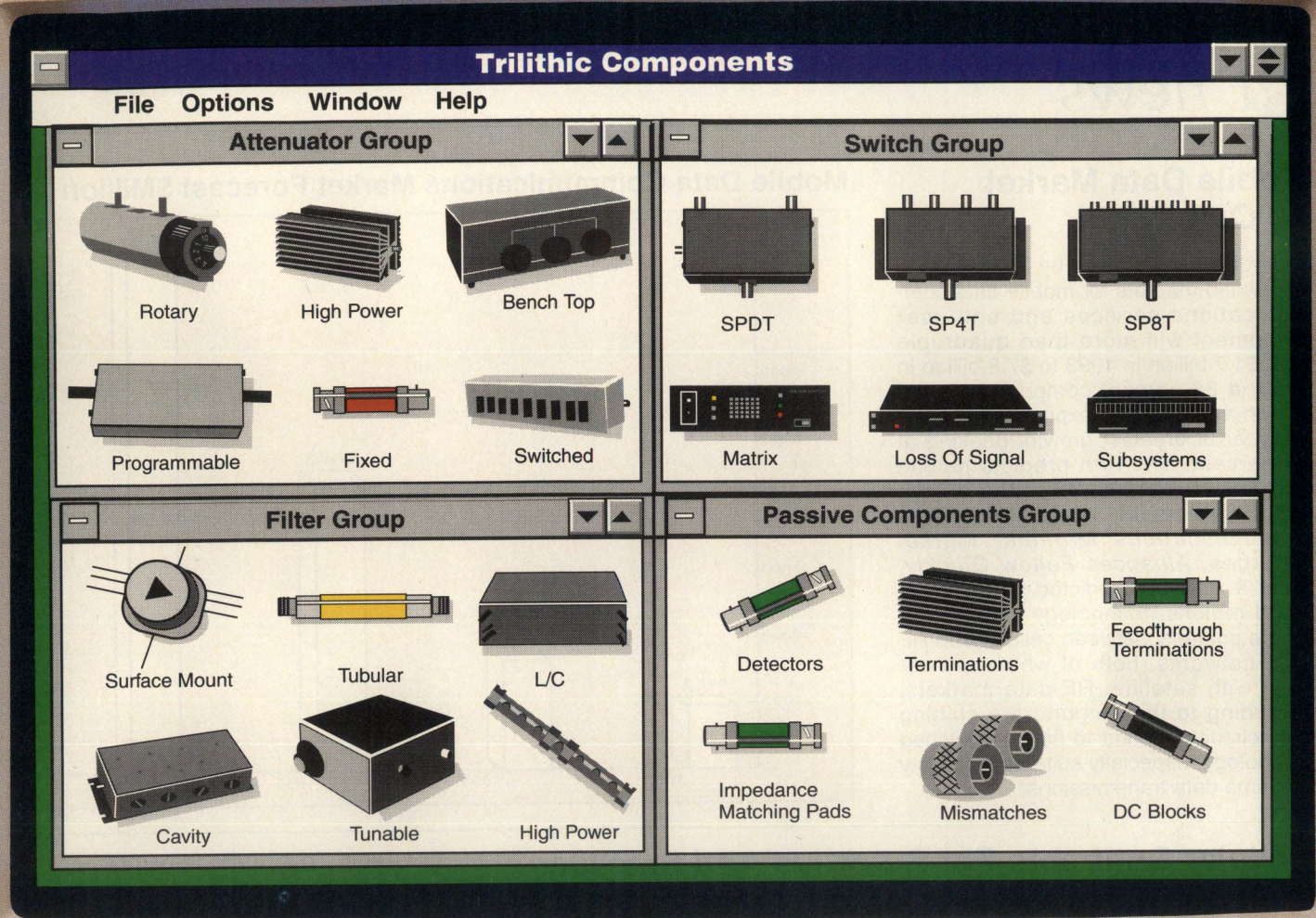
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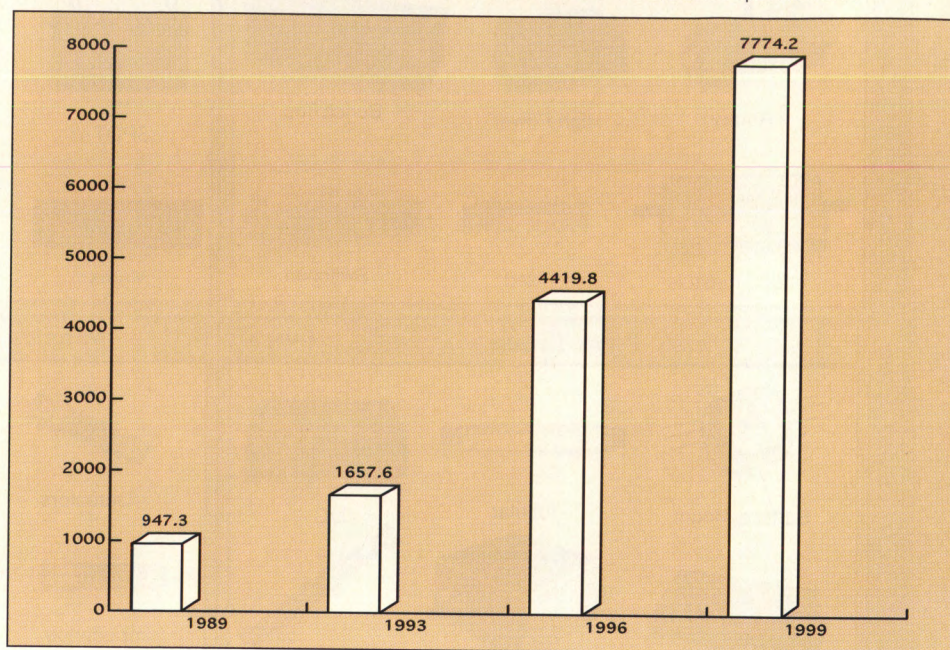
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## Mobile Data Market To Quadruple

According to a forecast by Frost & Sullivan, world markets for mobile data communications services and end-user equipment will more than quadruple from \$1.7 billion in 1993 to \$7.8 billion in 1999, a 28 percent compound annual growth rate. The firm expects 1996 to be the year of greatest growth, peaking at 48 percent. The firm predicts mobile data will surpass voice as the driving factor. In a report, *World Mobile Data Communications Markets: Market Emerges, Alliances Follow Quickly*, Frost & Sullivan predicted that in most world regions, technology competition will be primarily between cellular and RF data networks, both of which will co-exist with satellite. RF data markets, according to the authors, are shifting from circuit-switched to packet-switched technology, especially suitable for bursty real-time data transmissions.

## Mobile Data Communications Market Forecast \$Million



## Cellular Contracts For Subways And Satellites

**Schaeffer Magnetics Builds Satellite Hardware** — The Chatsworth, Calif.-based company has completed the design verification model of the solar array drive assemblies that will be used on the Iridium, Inc. satellite-based mobile telephone system. The mecha-

nism (photo above) provides two-axis positioning of the satellite's solar panels. Space Co. The satellite constellation will be comprised of 66 satellites. The global system is scheduled to be fully operational in 1997.

**Bell Atlantic Mobile** — is expanding its service in Washington to include the subway system. Commuters will have access to cellular phone use in underground stations and while riding the trains. Transmitters will be placed in underground stations. Antennas, provided by Cleveland-based Allen Telecom Group, will be placed in the tunnels. By early 1995, the service will be available in 48 stations and along 55 miles of subway track.

The system will use a series of micro-cells. Initial service is available in six stations covering 3-1/2 subway miles, with 17 additional stations scheduled for startup by April 1995.

**Avnet To Acquire Penstock** — Subject to applicable regulatory approval, Avnet, Inc. has entered into a contract to acquire all the shares of Penstock, Inc. Sunnyvale, Calif.-based Penstock is a distributor of several RF/microwave product lines, mostly to OEM customers in wireless and fiber wireless and fiber optic markets. Sales for the year ending March 31, 1994 exceeded \$45 million. Avnet is a \$3.5 billion company.

**Compact Gets Two Grants** — Compact Software has received two Phase I Small Business Innovative Research grants. One is for the development of accurate bias-dependent noise models for Heterojunction Bipolar Transistors (HBTs). HBTs have potential applications in monolithic microwave integrated circuits (MMICs). Existing noise models are based on approximations. The second grant is to develop accurate CAD models for Coplanar Waveguide (CPW) discontinuity structures in MMICs.

**Wavetek Licenses HP A-to-D** — Wavetek Corp. has licensed Hewlett-Packard Company's patented analog-to-digital conversion technology. Wavetek will use the technology in instruments manufactured by its Calibration Division (formerly Datron Ltd.) in the United Kingdom. The multislope A-to-D conversion technique provides higher speeds than similar conversions at equivalent resolution, according to the company.



nism (photo above) provides two-axis positioning of the satellite's solar panels. Under a \$14 million contract, Schaeffer will begin delivery of flight models in May 1995 to Lockheed Missiles and



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**AT&T Merges Flash Memory With DSPs** — This family of digital signal processors (DSPs) with on-chip flash memory from AT&T Microelectronics is intended to help consumer electronics manufacturers speed products to market. The ability to load, test and reload development software on the same DSP chip is projected to cut months from a typical development cycle, according to

the company. The FlashDSP 1616 processor, with a capacity of 50 million instructions per second (MIPS) at five volts and 30 MIPS at three volts is already being used by several cellular phone manufacturers. The FlashDSP 1616 processor is the first DSP on which the program data storage area can be electronically erased and reprogrammed, eliminating the need for each

software version to be permanently coded into a new chip's ROM.

**Scientific-Atlanta DAMA Network In Philippines** — International Communications Corporation, one of the Philippines' licensed private telecommunications providers, has selected Scientific-Atlanta to install a demand assigned multiple access (DAMA) satellite network. The network will overlay an existing nationwide network and include a master control earth station and 22 remote sites. The remote sites will include a mixture of eight-channel DDS (digital DAMA System) and two-channel MicroDama chassis for thin to medium route telephone traffic. The network is scheduled for completion later this summer. Scientific-Atlanta original VSAT hub in 1991.

**Motorola Expands** — Harvard, Ill. will be the site of a new manufacturing, engineering and administrative facility of Motorola Cellular Subscriber Group. The one million square feet is scheduled to be operational by early 1996. Similar to the company's headquarters in Libertyville, Ill., the new facility will create 2,000 to 3,000 new jobs.

**Omnipoint and Celeritek in PCS Venture** — Omnipoint Corporation and Celeritek, Inc. have agreed to work together in the development of a MMIC chip set for use in Omnipoint's Personal Communication Service (PCS) telephones and base station equipment. Under the joint agreement, Celeritek will provide production quantities of the chip set operating in the 1.9 GHz PCS spectrum. Omnipoint was recently awarded a 30 MHz PCS operator license in the New York Major Trading Area (MTA).

**Battelle To Automate Autos** — The Federal Highway Administration awarded a \$1.7 million contract to Battelle as part of the Department of Transportation plans for an automated highway system. Battelle will focus on systems analysis of malfunction management and safety, operational issues and other topics.

**Briefcase Earth Stations** — Rockwell Defense Electronics has two briefcase mobile earth station products that offer dial-up voice, fax and data services. The two products are similar; both use the INMARSAT satellite network. One version, the Secure Satcom Terminal provides secure voice, fax and data services.

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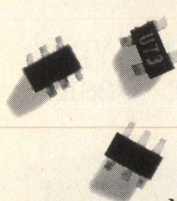
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**UPC2710**  
**Wide Band Amplifier**  
DC to 1000 MHz  
33 dB Gain  
3.5 dB NF

**UPC2713**  
**Wide Band Amplifier**  
DC to 1200 MHz  
29 dB Gain  
3.2 dB NF

**UPC2723**  
**AGC Amplifier**  
100 MHz to 1100 MHz  
38 dB AGC  
13 dB Gain

**UPC2721**  
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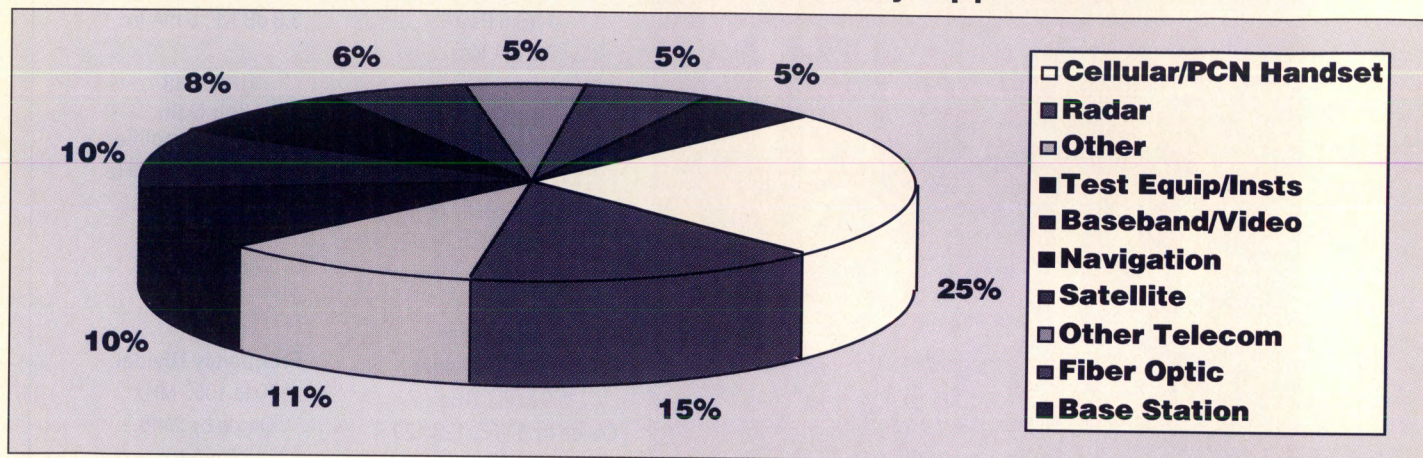
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## 1993 U.S. GaAs MMIC Market By Application



**GaAs MMIC Market To Grow** — Over the next five years, the market for gallium arsenide monolithic microwave integrated circuits will grow 25 percent annually, according to Allied Business Intelligence of Oyster Bay, N.Y. The company produced a report titled "Amplifiers: U.S. Markets, Applications & Competitors: 1993 to 1998 Analysis."

Low cost wireless telecommunications products are projected to grow substantially, prompting the growth of the GaAs MMIC market. The firm predicts the greatest opportunities at 5.8 GHz, 2.4 GHz and above 30GHz. The report includes sales and market share figures for 39 amplifier, 24 transistor and 21 MMIC manufacturers. The analysis was

conducted according to end use, selling prices, operating frequencies, output power, device technology, packaging type and bandwidth. The research firm predicts growth and lower prices in wireless telecommunications in both commercial and consumer markets. (Data for the graph above was provided by Allied Business Intelligence, Inc.)

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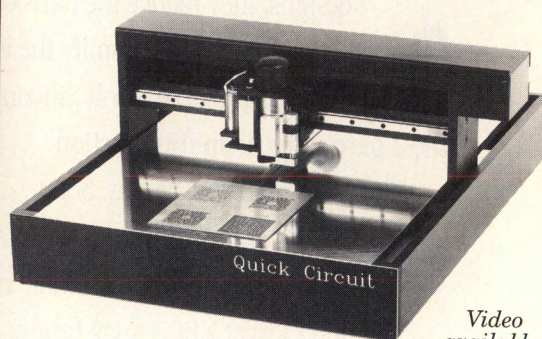
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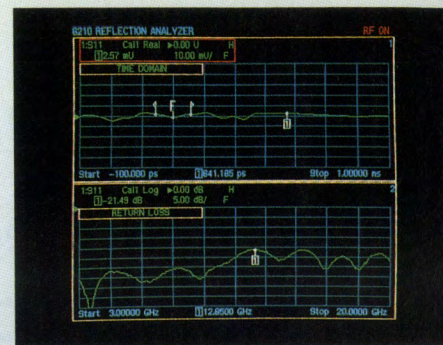
Now a powerful & economical new addition to the family, the **6210 Reflection Analyzer** is here. It uses the highly accurate "6-port coupler" technique to measure phase and amplitude characteristics of network inputs ( $S_{11}$ ). The Test Set thus provides accurate return-loss measurements, vector

measurements, and time domain measurements. The frequency range of the Model 6210 starts at 250 MHz and extends to 26.5 GHz (or as limited by the host 6200 Series Microwave Test Set).

The Reflection Analyzer is housed in an add-on adaptor that fits below the 6200 Series Microwave Test Set thereby retaining its compact profile for portable and field use. This adaptor technique also provides an easy upgrade route for users of the 6200 Test Set now and at any time in the future. All existing features of the 6200 Series Test Set are retained.

Key features of the 6210 include:

- Higher accuracy and wider range reflection measurements as compared to the RF bridge technique.
- Both vector and time domain analysis so that the causes of reflections can be diagnosed. Especially useful with fault location.
- Smith Chart presentation for easier impedance matching adjustments.



- Simultaneous time domain and frequency domain measurements for full characterization of an input port.

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INFO/CARD 18



# Antenna Technology Takes on the Challenge of Wireless Applications

By Gary A. Breed  
Editor

Radio signals must have an antenna in order to radiate and be received. Depending on the application, the antenna may be simple or complex, efficient or inefficient, large or small. For new wireless communications equipment, the requirements are often contradictory, requiring tradeoffs or unusual design solutions.

The first consideration is physical. What is the shape and volume of the device, is it fixed, handheld, or body-worn. Is it going to be well-treated or must it survive rough handling.

The next consideration is system performance. What range is required? How sensitive is the receiver, what is the transmitter power, and what are the signal-to-noise specifications? What is the frequency of operation?

With a picture in mind, an antenna is selected from available products, or uniquely designed to fit the requirements. The rest of this report will look at some of the antenna solutions that have been developed for wireless products.

### Dipoles and Monopoles

The common "whip" antenna is still in widespread use in those cases where the protrusion of a slender rod or mast can be accommodated. This has been the most common antenna used in traditional mobile radio, both for vehicular installation, and for handheld equipment. Antennas for handheld products are often physically shortened by the use of a loading inductance: the well-known "rubber ducky" antenna.

Because they are widely used, the cost of this type of antenna is low. Adaptation from standard products is also possible. For example, Part 15 devices operating in the 902-928 MHz band can use antennas designed for cellular telephones with little or no modification.

Collinear arrays, like the common cellular vehicle antenna with its distinctive phasing coil, can be used when greater antenna gain is useful. In particular, such antennas can provide increased range in locations where the path losses are significant.

At frequencies of 900 MHz and above, a dipole or monopole may be included within the structure of a products. In one case, a 915 MHz dipole has been incorporated into the plastic enclosure of a notebook computer.

### Other Configurations

Besides the simple dipole or monopole configurations, loop and patch antennas are commonly used. Each has physical characteristics that may be used to advantage.

Some pagers make use of loop antennas comprising an etched copper trace around the perimeter of the unit's circuit board. This allows the small belt-worn units to avoid external antennas that might break or get in the way. At the VHF frequencies most often used, the loop will be smaller than the full wavelength of a resonant loop, and matching is required. As reported by Dr. Ian Dilworth at RF Expo West 1994, the compensation for a short electrical length can be accomplished by capacitive, linear or inductive loading, including inductance contributed by ferrite materials.

Patch antennas are well-known in radar and some satellite applications. As such, they have thoroughly analyzed and characterized. Physically, a patch antenna constructed on common glass-PTFE substrate material will be most useful at frequencies above 2 GHz. However, some high-dielectric-constant ceramic substrates have been used to make compact antennas for GPS at in the 1.5-1.6 GHz range.

Diversity antennas are another family, where sufficient physical separation is present to combat fading due to multipath. At 2.4 GHz, two antennas can be placed in a single compact package, but at lower frequencies, other solutions are required. One is to actually use two antennas located at least one-quarter wavelength apart. Another option is to use polarization diversity, which can be easily accomplished by mounting two dipoles or monopoles at right angles to one another. Polarization diversity, in general, is not as effective as spatial

diversity, but it can provide a dramatic improvement over a single antenna.

### Future Applications

If development of products continues as expected, new antennas must be developed for severe environments, reduced size, and flat form factors. For applications involving Part 15 devices in the microwave ISM bands at 3.4 and 5.7 GHz, multipath is more severe and diversity becomes much more important. Shorter wavelengths at microwave frequencies can be an advantage, allowing full-size resonant structures to be constructed within a product's package.

It may be necessary for new antennas to cover more than one frequency, if multiple-use products appear. For example, a PCS cordless phone operating at 1.9 GHz may be combined with 900 MHz cellular circuitry in the same unit. Other products may utilize two or more of the unlicensed ISM bands in a single product, as well.

By definition, wireless RF products radiate and receive signals! To get the job done efficiently, the advanced circuitry inside a product needs a properly-designed antenna. In today's competitive market, no weak links can be tolerated.

RF

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# Open Loop Modulation of VCOs for Cordless Telecommunications

By Daniel E. Fague  
National Semiconductor Corporation

Cordless telecommunications systems are rapidly emerging, and low cost radio front ends are paramount to their success. Frequency modulation (FM) has traditionally been the choice for low cost radio systems, and indeed it is used in today's cellular telephones and analog cordless phones. The Digital European Cordless Telecommunications (DECT) standard specifies the use of a digital FM technique, Gaussian filtered Frequency Shift Keying (GFSK). A relatively simple technique to implement the modulator for DECT is open loop modulation of a VCO. This article will examine the concept of open loop modulation as used in DECT. Measured results from a hardware implementation of the technique for a full frequency (2 GHz) VCO are presented to demonstrate its robustness. Also, measurements from a half frequency VCO implementation are used to show open loop modulation's performance in an actual DECT phone.

Frequency modulators are the current choice for most mobile telephones. Two reasons for this include the low cost of both the modulator and demodulator (limiting amplifier and FM discriminator) and the relative ease of implementation of FM systems. While both analog cellular and analog cordless telephones utilize FM methods, the complexity of the modulators and radio front ends is different. Cordless applications typically can use less complex modulator structures because they operate in less "radio-hostile" environments. The less complex modulators (and radio front ends) translate directly to lower cost telephones. Since digital communications methods are emerging as the solution for future telecommunications systems, the need to find low cost implementations of high performance modulators is ever greater.

The pan-European DECT standard combines a high bit rate (1.152 Mb/s), high frequency operation (2 GHz), and fast switching between frequencies (~30 ms). A digital cordless telephone sold in

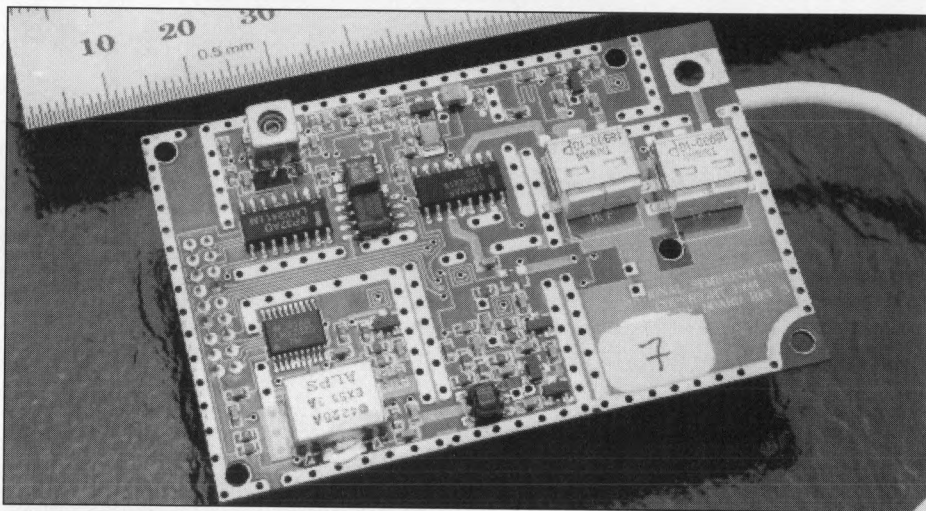


Photo of the test circuit utilizing an open-loop modulated half-frequency VCO, plus a frequency doubler.

the United States will have similar constraints to the DECT system. These stringent requirements must be met while keeping the overall cost of the radio low enough to be competitive in the consumer market.

The DECT standard calls for a (digital) FM technique. This allows the use of FM discriminators in the receiver, but other requirements in the DECT system preclude the use of conventional FM modulators. A solution to the design of a simple FM modulator for cordless systems

such as DECT is open loop modulation of a VCO. Open loop modulation simplifies the modulator because the loop filter in the phase locked loop will not affect the modulation, since the PLL is actually opened, or unlocked, when the modulation is applied.

Several key factors allow for the loop to be opened. First, the DECT bursts are short (approx. 0.5 ms). This reduces the time available for the VCO to drift. In addition, very low leakage charge pumps in PLL integrated circuits such as

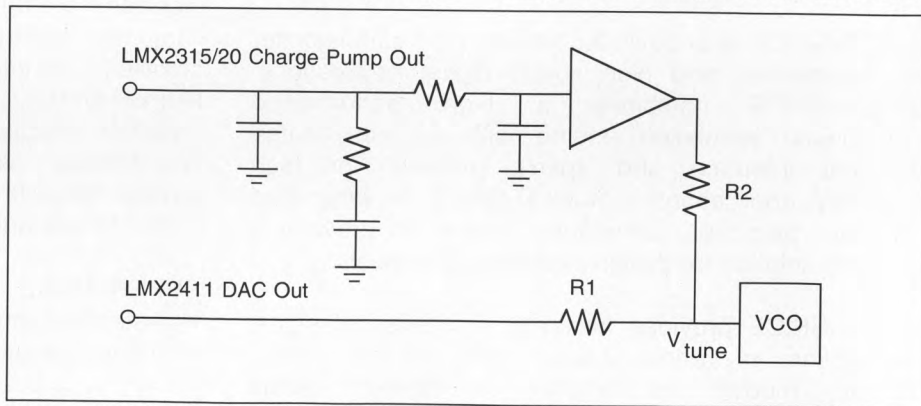


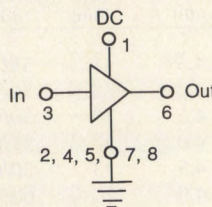
Figure 1. Block diagram of open loop modulation.





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the National Semiconductor Corp. (NSC) LMX2315 and LMX2320 prevent the loop filter voltage from sagging during unlocked bursts. Finally, any limitations that can occur with open loop modulation can be overcome. This article investigates those limitations and concentrates on methods to remove them. Measured results of these solutions are provided to illustrate their effectiveness.

These results include measurements from both a full frequency (1.9 GHz) and half frequency (950 MHz) VCO implementation.

### Open Loop Modulation

The DECT system uses a Time Division Multiple Access/Time Division Duplex (TDMA/TDD) access method. There are 24 time slots in a 10 ms

frame, or 420 ms per time slot. Ten frequency channels are available, each spaced 1728 kHz apart. This means a total of 120 full duplex channels are available for use. The modulation method for DECT is Gaussian filtered frequency shift keying (GFSK) with a pre-modulation lowpass filter bandwidth of half the bit rate ( $B_b T = 0.5$ ) and nominal peak frequency deviation of a quarter of the bit rate ( $\pm 288$  kHz). Because this is a frequency modulation technique, a voltage controlled oscillator (VCO) can be directly modulated by the baseband signal.

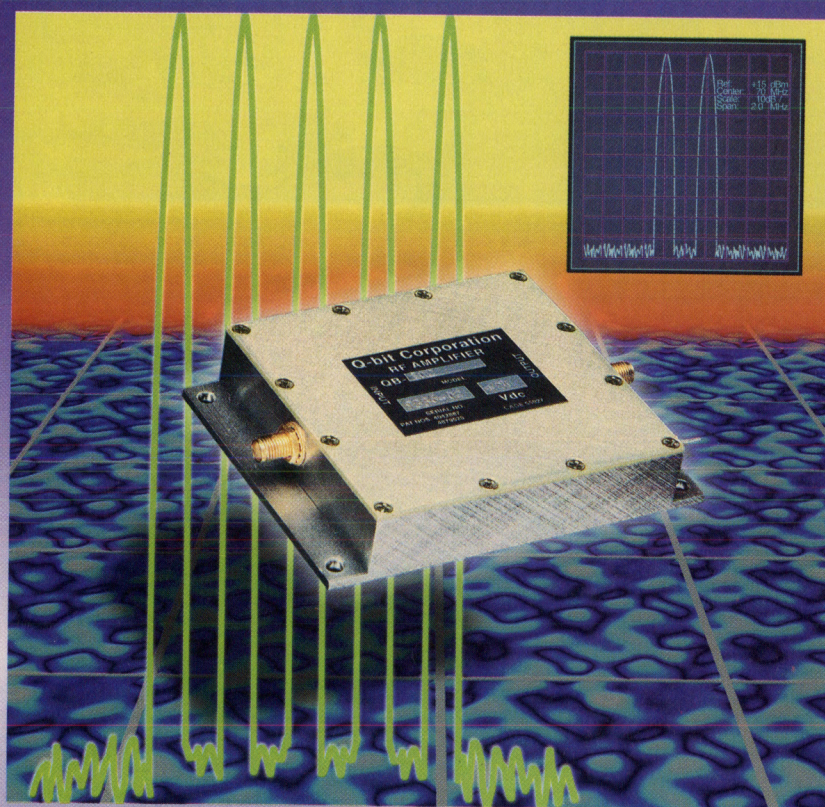
There are several techniques that can be used to modulate a VCO. The most common are modulation "in the loop" and modulation "over the loop". In general, modulation in the loop can be used when the output signal is narrowband with respect to the loop filter. Modulation over the loop can be used when the time required to switch frequencies is relatively long.

In DECT, however, neither of these conditions is valid. The lock times of the phase locked loop must be very short (~30 ms) to avoid a blind slot, and the output signal's spectrum is much wider (approximately 50 times) than the loop filter bandwidth. In certain radio architectures such as single conversion receivers [3,4], there is a need for wide-band VCOs to cover the IF bandwidth plus the system bandwidth. In DECT, this amounts to a 130 MHz bandwidth at 1.77-1.90 GHz. Open loop modulation is an exciting technique that allows for fast switching speeds and wide output signal bandwidths while delivering high performance at the 130 MHz spread.

In open loop modulation, the PLL is actually unlocked ("opened") for a brief time to allow the modulation to occur. The modulating voltage is added to the loop filter voltage (at the center frequency) at either a modulation port or the tuning port (via a resistive adder). Figure 1 shows a block diagram of open loop modulation. One of the key reasons that open loop modulation can be used in DECT is that the time slots are of very short duration (420 ms), so the loop is open only for that short time.

The sequence for open loop modulation is as follows: Just prior to transmission, the loop is closed to lock the VCO to the correct carrier frequency. The modulating signal is then turned on, and the loop remains closed to re-lock to the center frequency.

At this point, the modulation signal is simply the mean DC point of the modu-



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QB-7205	2-70	22	1.5:1	4.0	27	51/65
QB-7223	2-32	22	1.5:1	4.5	33	50/65
QBS-101	2-70	12	1.5:1	4.0	20	60/100



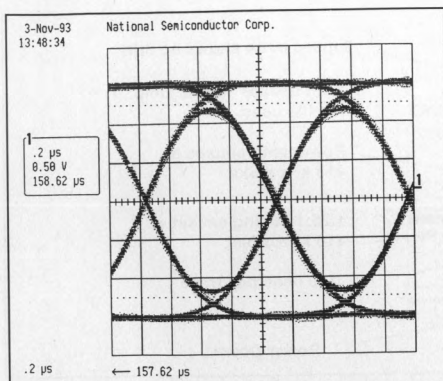
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**Figure 2.** Eye diagram of the demodulated output of the open loop VCO (full frequency), modulated by Gaussian filtered data.

lation. The loop is then opened, and modulation by the transmit data occurs. Once the modulation is finished, the loop can be closed again, and the PLL tuned to the receive frequency.

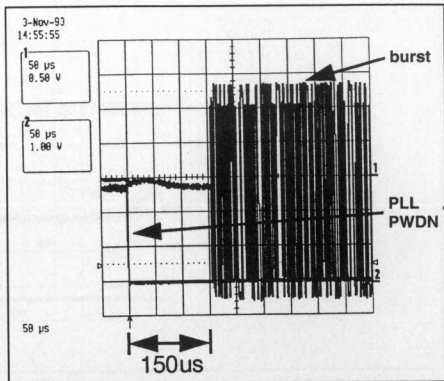
Since KV of a wideband VCO is on the order of 70 MHz/V for a 3 V VCO, it is important that the loop have very low drift since relatively small voltage changes can have a big impact on the center frequency.

There are some elements of open loop modulation that can limit performance, as listed below:

- **Frequency pushing:** A change in VCO output frequency caused by a change in the VCO supply voltage.
- **Load pulling:** A change in VCO output frequency that is caused by a change in the load that the VCO output buffer sees.
- **Frequency drift:** A change in the VCO output frequency that is caused by RF coupling or by droop in the VCO tuning voltage. Droop in the VCO tuning voltage can be caused by leakage from the PLL charge pump, the loop filter components, or the resistive adder at the tuning port.

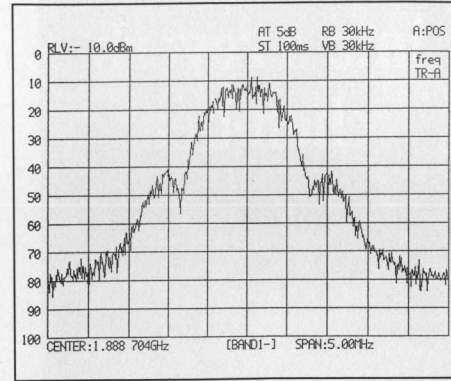
### Full Frequency Experimental Setup

The circuit used for the full frequency open loop modulation measurements utilized the LMX2320 2.0 GHz PLL with a TRI-STATE™ feature on the charge pump output, a 130 MHz wideband VCO, a 3.0 V regulator, and the Gaussian filter ROM-DAC on the LMX2411 to shape the transmit data. The data rate used was 1.152 Mb/s (DECT bit rate).



**Figure 3.** Demodulated output of the open loop VCO (full frequency), modulated by Gaussian filtered data.

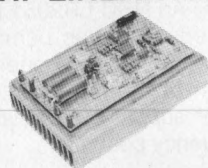
The carrier frequencies that were used to demonstrate open loop modulation performance were DECT channels 0, 5, and 9. The circuit was implemented on a printed circuit board. The power amplifier load was simulated by a 10 ohm resistor and a P-channel switch (for cur-



**Figure 4.** Plot of the open loop modulated DECT signal in burst mode as taken by a time gated spectrum analyzer.

rent draw), and a PIN diode and 220 ohm resistor to ground (for load impedance change). Measurements were made by using a Rohde and Schwarz CMT-55 DECT system tester to demodulate the output of the circuit (from the PIN switch), and the frequency offsets

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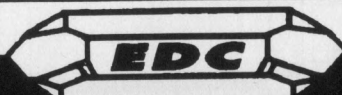
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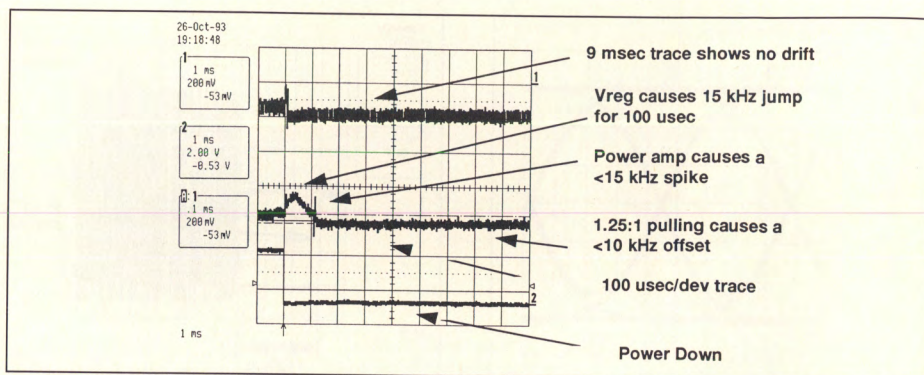


Figure 5. Measurement of open loop frequency pushing, load pulling, and frequency drift for DECT channel 0 at 25° C.

were determined from the demodulated signal on a digitizing oscilloscope.

The circuit featured the National Semiconductor LMC6482 CMOS rail-to-rail operational amplifier that was used to control leakage due to the resistive adder at the tuning port. The low output impedance of the op amp allowed the summing node to be a simple resistive adder. An RF buffer with good reverse isolation (>30 dB) was used at the output of the VCO to effectively control the load pulling caused by the power amplifier and T/R switch. An NSC LP2951 low dropout voltage regulator was used to supply the 3.0 V  $V_{CC}$  to the components, and this regulation of the input battery voltage plus power supply filtering limited noise and frequency pushing.

## Full Frequency Experimental Results

The first set of measurements shows the experimental results of modulating a full frequency (i.e., 1770 to 1900 MHz) VCO. The performance is seen to be very good. Figure 2 shows the demodulated eye diagram of the open loop modulated DECT signal. The virtually perfect

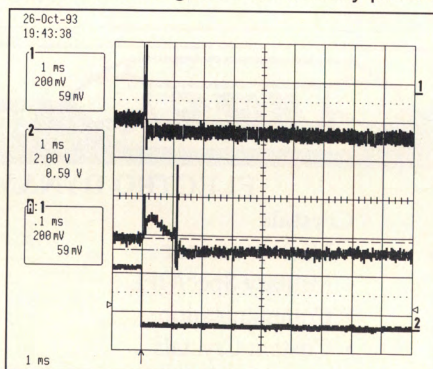


Figure 6. Measurement of open loop frequency pushing, load pulling, and frequency drift on DECT Channel 0 at -20° C.

eye diagram of Gaussian  $B_bT = 0.5$  filtered data can be seen. Figure 3 shows a time domain view of the demodulated signal after the loop was opened. A small 15 kHz jump that settles out can be seen in this time domain plot. Figure 4 shows the open loop modulated DECT signal operating in burst mode (i.e., TDMA/TDD operation). The measurement was taken with a time gated spectrum analyzer to capture a number of bursts. Note that the output spectrum of the signal is a DECT-compliant GFSK signal.

The next three plots show the small imperfections that can occur in open loop modulation. Figure 5 shows open loop frequency pushing, pulling, and drift at 25° C on Channel 0 in the DECT band (~1.9 GHz). The lower trace is the power down signal sent to the PLL to TRI-STATE the charge pump. The upper trace shows that there is imperceptible frequency drift over a 9 ms burst. The middle trace is a 10x zoom of the upper trace, and it shows a 15 kHz pushing jump while the voltage regulator recovers from the PLL powering down. This lasts about 100 msec. The (simu-

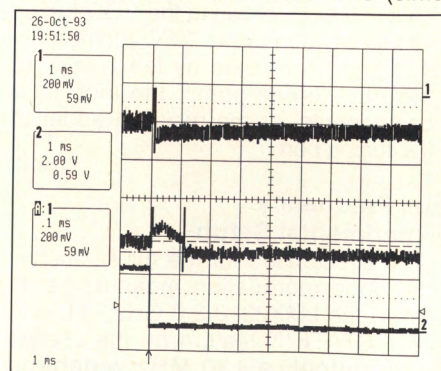
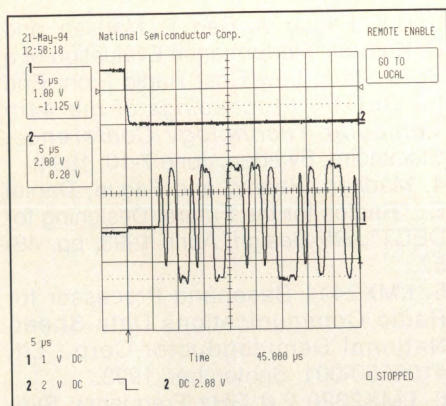


Figure 7. Measurement of open loop frequency pushing, load pulling, and frequency drift for DECT channel 0 at 60° C.





**Figure 8. Beginning of a demodulated DECT burst. Note small (30 kHz) jump after power amplifier power up signal goes low (upper trace). Horiz. 5 ms/div., vert. 200 kHz/div.**

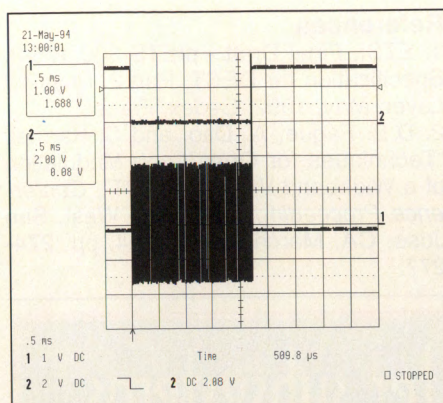
lated) power amplifier turning on causes a 15 kHz pushing spike, and 1.25:1 load pulling causes a <10 kHz offset. Similar curves are seen at Channel 9, the lower end of the DECT band (~1.88 GHz).

Figure 6 shows the performance of open loop modulation at -20°C. It can be seen that the only difference in performance is that the power amplifier pushing creates a slightly larger spike at the beginning of the burst, but the rest of the performance is nearly identical to that of 25°C. Figure 7 shows the performance at 60°C. In this case, the power amplifier pushes a slightly smaller spike. However, a small frequency drift is now noticed, and this is due to a small leakage from the charge pump to the VCO.

### Half Frequency Experimental Results

In a second experiment, a half frequency VCO and doubler circuit was implemented. This variation was tested because a half frequency VCO has a greater inherent immunity to RF radiation than the full frequency VCO. This higher tolerance for RF radiation exists because the VCO operates at half the frequency that the power amplifier radiates, and thus is less likely to couple to the power amplifier's output. A standard balun was used, although a printed balun can be used to save cost. The next three figures show the performance of the half frequency VCO open loop modulation used in the DECT phone.

Figure 8 shows the start of a demodulated burst from the DECT phone. After the power amplifier is powered up (top trace), a small (30 kHz) jump is seen to

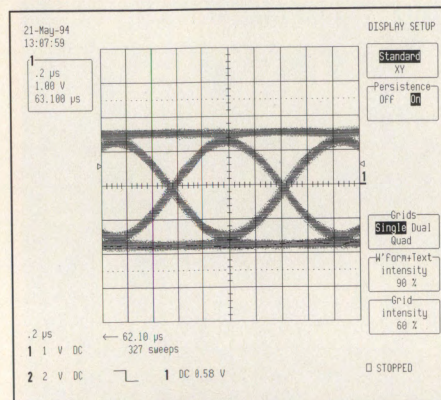


**Figure 9. Complete demodulated DECT burst. There is no center frequency drift, and the deviation is about 588 kHz (peak-to-peak). Horiz. 0.5 ms/div. (expanded 5x to 0.1 ms/div.), vert. 200 kHz/div.**

occur. Then, the demodulated random data appears with no further drift. The entire demodulated burst is shown in Figure 9. Here, it can be seen that the frequency deviation is approximately 588 kHz (peak to peak). Both Figure 8 and Figure 9 have vertical scales of 200 kHz/div. Finally, the received eye diagram is shown in Figure 10, and it can be seen that the demodulated signal is Gaussian filtered ( $B_b T = 0.5$ ) data. A photograph of the 950 MHz RF front end board is shown on the first page of this article.

### Summary

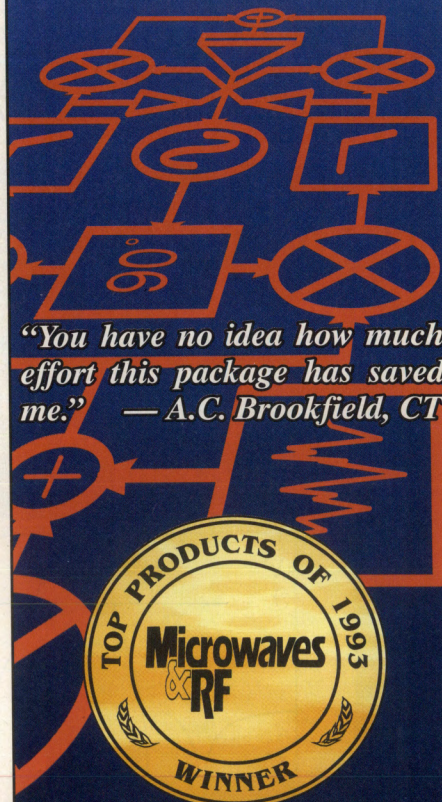
As demonstrated in this article, the principle of open loop modulation for wideband VCOs can be implemented and used in a practical and cost-effective manner. The measured results show that the methods used to reduce



**Figure 10. Received eye diagram from the DECT phone. Bit rate is 1.152 Mb/s.**

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or eliminate frequency pushing, load pulling, and frequency drift are effective. These methods are also robust, as the circuit shows virtually no degradation in performance over a wide range of temperatures. Open loop modulation can therefore provide a proven, low cost solution to designers who are implementing digital FM modulators in systems with short data burst durations. **RF**

## References

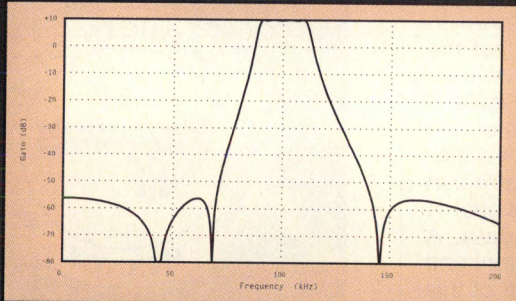
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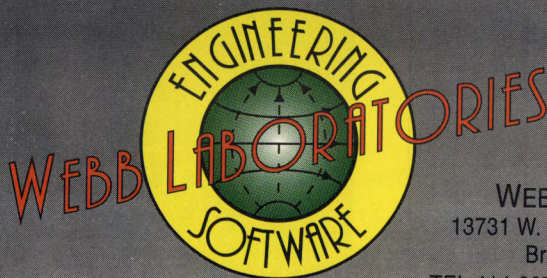
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## About the Author

Daniel E. Fague obtained his BSEE in 1989 at Gonzaga University and his MSEE (Signals and Systems) in 1991 at the University of California at Davis. Since 1991, he has worked at National Semiconductor Corporation in the Wireless Communications Group, specifying integrated circuits for radio transceivers and building radio boards with them. He is a member of the IEEE, and his interests include digital radio modems, indoor and outdoor propagation for mobiles, and digital signal processing. He can be reached at National Semiconductor Corporation, P.O. Box 58090, M/S A-1500, 2900 Semiconductor Drive, Santa Clara, CA 95052-8090.

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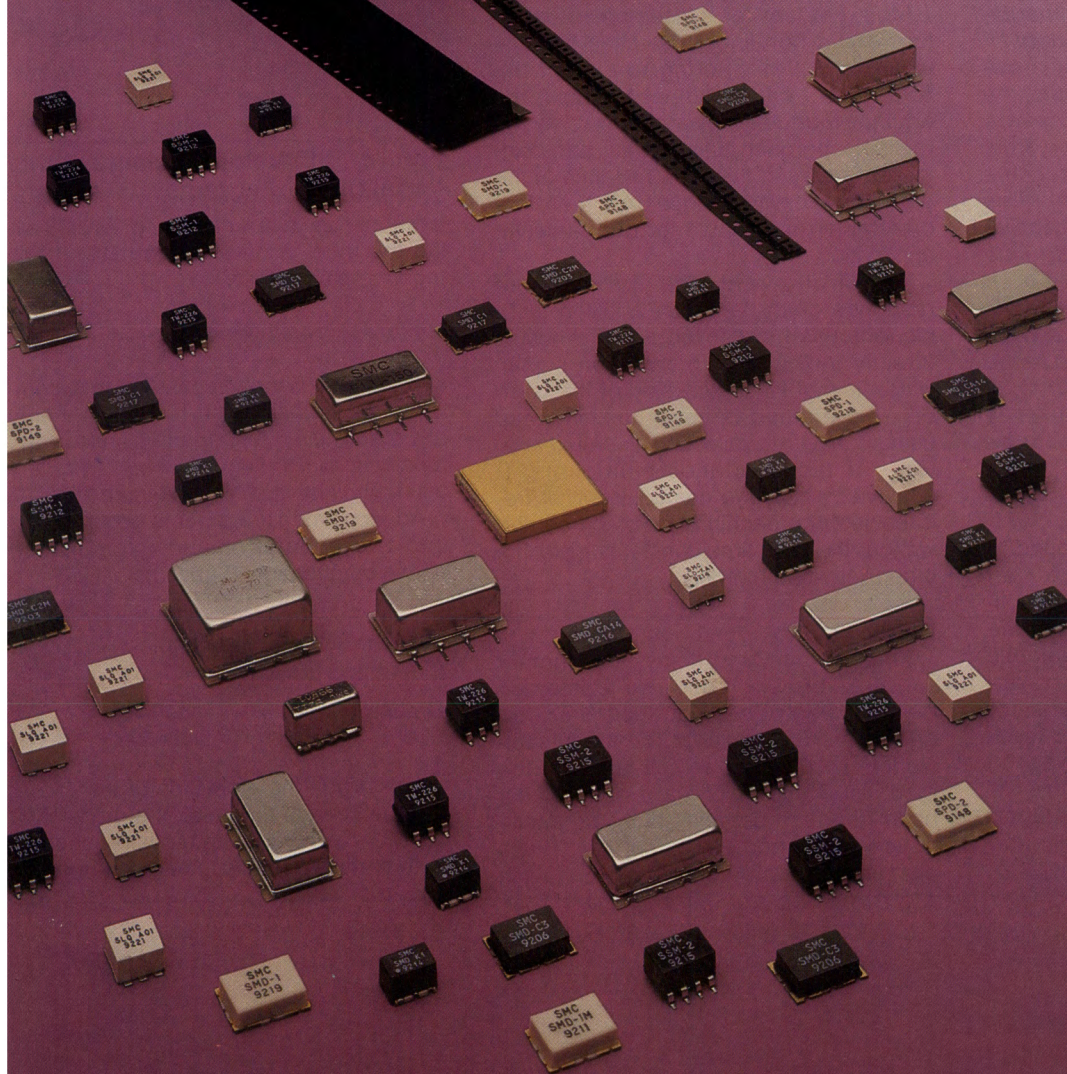
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## A Fast Envelope Detector

By Dominic J. Ciardullo  
Brookhaven National Laboratory

Traditional AM demodulating methods include both synchronous (product detector) and non-synchronous (rectification) techniques. While each has its merits, both methods require some amount of filtering to accomplish their goal of envelope detection. In certain applications, the resulting circuit delays present severe limitations to the system's response time, (for example, when the detection scheme is utilized within a feedback loop). This paper describes a wideband, active envelope detector with a response time on the order of 1  $\mu$ s. The circuit described has a demodulation bandwidth of DC-350 kHz, and provides a high degree of linearity over a dynamic range in excess of 40 dB. The device is designed for broadband carrier operation (1 MHz to 5 MHz), but can easily be adapted to demodulate carriers in other ranges (up to VHF).

The circuit described was developed to provide a real-time monitor for an amplitude modulated RF signal swept in frequency from 1 to 5 MHz. Although the specific accuracy, linearity and dynamic range requirements could have been satisfied using more conventional detector techniques, its inclusion as part of a feedback loop within a larger system was the ultimate motivation for developing this circuit. Standard envelope detectors generally utilize a signal rectification/filtering combination which requires a time con-

stant much greater than one RF period (at the lowest carrier frequency) to maintain reasonable amplitude accuracy.

Most product (or synchronous) detectors use a non-linear detection scheme which requires filtering to extract the "DC" term from its output to obtain the original modulation waveform. Like the rectifier approach, this filtering adds significant delay to the response time of the detector; This technique was therefore determined unsuitable for our particular application.

To circumvent the problem of response delay, a vector addition technique is used to convert an RF sinusoid directly into a "DC" modulation signal, with minimal RF processing of the carrier. This method is easily implemented using currently available high speed analog transconductance multipliers. It achieves amplitude demodulation at frequencies approaching that of the carrier, without the long time constants typical of other detection methods.

### Basic Theory

Since minimizing delay is a primary concern, it is desired to use a detection method which does not depend on low pass filtering to extract the modulation signal. To do this, we will first split the modulated carrier into two equal parts, each of which contains both the intelligence signal and the carrier RF. These elements may then be combined in such

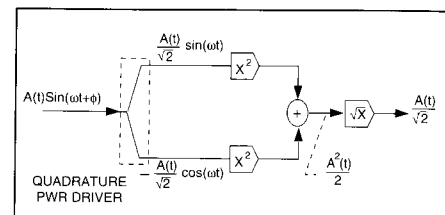


Figure 1. Envelope detector functional block diagram.

a manner as to null out the RF while leaving the modulation waveform intact. The trigonometric identity

$$\sin^2(\omega t) + \cos^2(\omega t) = 1 \quad (1)$$

shows that it is indeed possible to accomplish this task if the two components are made to be 90° apart in phase. The functional block diagram of such a detection technique is shown in Figure 1.

An incoming RF carrier is decomposed into two equal amplitude, quadrature phase components (I and Q):

$$I \text{ component: } \frac{A(t)}{\sqrt{2}} \sin(\omega t) \quad (2)$$

$$Q \text{ component: } -\frac{A(t)}{\sqrt{2}} \cos(\omega t)$$

where A(t) is the amplitude of the carrier as a function of time (i.e., the modulation waveform) and  $\omega$  is the carrier frequency. Note that we have assumed here the use

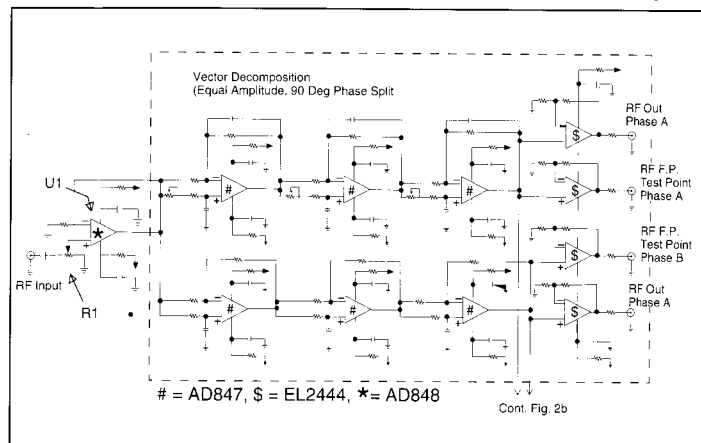


Figure 2a. Functional diagram of the demodulator. Portion of circuit in dotted box performs vector decomposition.

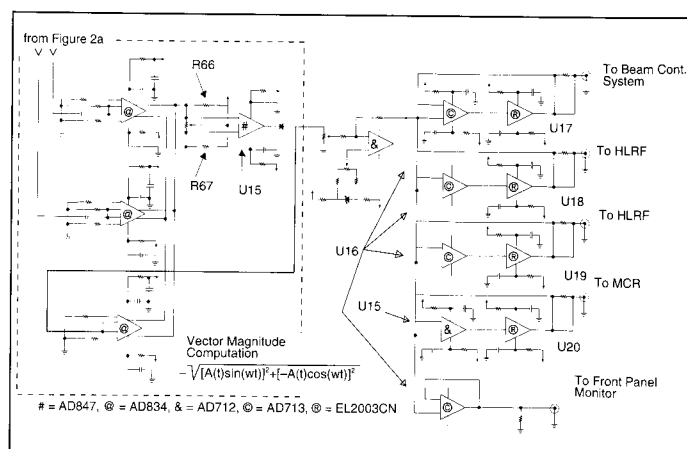


Figure 2b. Continuation of functional diagram. Circuitry in dotted box performs vector magnitude computation.



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of a power divider to obtain the quadrature phase split, with some insertion phase  $\phi$  (which is dependent upon the RF frequency). The vector sum of the two components is explicitly carried out to obtain a linearly scaled version of the amplitude modulating function,  $A(t)$ .

To obtain the vector sum, both components of the original carrier are squared, then added together. The square root of the sum of the squares is then computed in real-time to obtain:

$$\sqrt{\left[\frac{A(t)}{\sqrt{2}}\right]^2 \sin^2(\omega t) + \left[-\frac{A(t)}{\sqrt{2}}\right]^2 \cos^2(\omega t)} = \frac{A(t)}{\sqrt{2}} \quad (3)$$

using the trigonometric identity of equation 1. An important consequence of this result is the fact that there is no RF term in the solution; Although we started out with two RF (amplitude modulated) sinusoids, the result is a scaled version of the modulation waveform only. The significance of this outcome is two-fold; First, even though two non-linear operations

were involved, no filtering is required to separate the carrier from the original modulation waveform. Since no filtering is required, there are no long time constants to slow down the overall detector's response time. In addition, the modulation frequency can approach the actual carrier RF, since there is no need to allow room for filter "skirts". The second implication of this result is one of practicality; Any analog processing which follows this portion of the electronics need only operate at the modulation frequency (as opposed to at RF). Time delay and slew rate now become the chief design considerations from this point on in the circuit, rather than RF bandwidth.

### Implementing The Block Diagram

A schematic for the envelope detector as constructed is shown in Figures 2a and 2b. The drawing is divided into three basic sections; vector decomposition, vector magnitude computation and post processing gain/offset/distribution. Amplifier U1 is used to scale the maximum amplitude of the modulated RF to 2 Vpp. This upper limit is a constraint imposed

by the maximum input voltage to the vector magnitude section (the vector decomposition section has unity gain). Potentiometer R1 is calibrated such that the peak carrier voltage applied to the input of the circuit results in +1 Vpk at the output of U1. This adjustment matches the input signal to the AD834s to make full use of their dynamic range.

### Quadrature Phase Division

The 90° phase split can be accomplished using a good quality quadrature power divider, many of which are available commercially. Using such a device provides a quick, easy solution for frequencies above approximately 100 kHz, particularly if wideband carrier reception is not a main concern. Active all-pass networks [1] are a good choice for carriers in the HF frequency range and below, especially for applications where a 90° splitter doesn't exist "off the shelf" or when its cost is prohibitively expensive. The advantage to using an active all-pass network as opposed to a passive one is the amplitude match afforded by using operational amplifiers. The use of

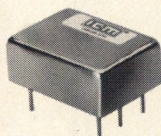


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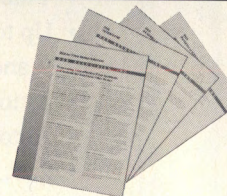
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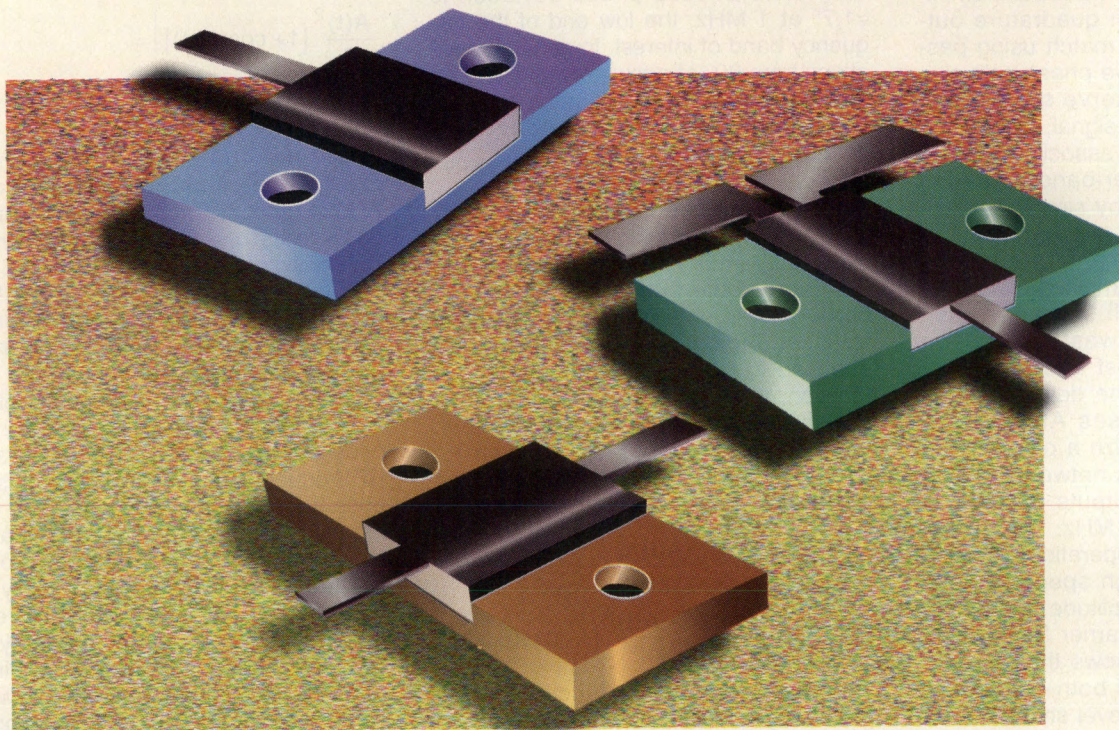
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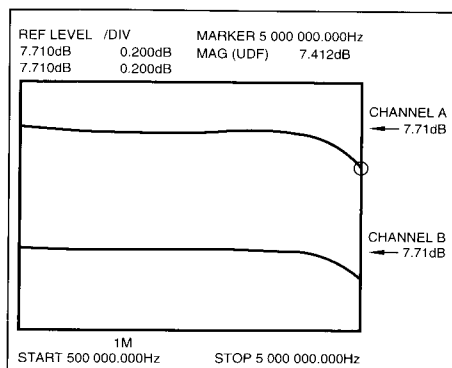


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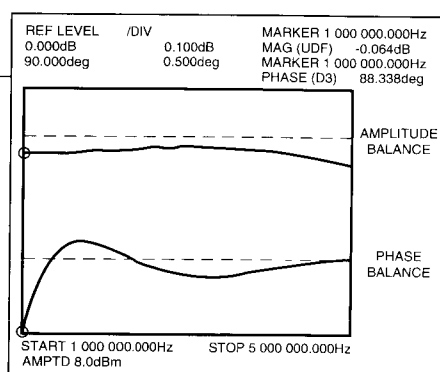


**Figure 3. Absolute amplitude response of both quadrature outputs.**

feedback provides a very flat amplitude response, resulting in a broadband amplitude balance between quadrature outputs that is difficult to match using passive networks. An active phase split also provides gain to preserve the original amplitude of the input signal, eliminating the  $1/\sqrt{2}$  scaling factor associated with a power divider. Broader-band operation can be accomplished by simply cascading additional sections, without incurring an appreciable loss in carrier signal or a significant increase in amplitude ripple.

The decision of which  $90^\circ$  phase shifting technique to use was based upon amplitude balance, cost and frequency range requirements. The detector circuit shown in Figure 2 uses AD847 high speed op-amps to realize a pair of three section active all-pass networks, which provide quadrature outputs over a frequency range of 1 to 5 MHz. The AD847 is a voltage feedback operational amplifier, selected for its high speed at unity gain, as well as its amplitude and phase characteristics at the carrier frequencies of interest. Figure 3 shows the absolute amplitude response of both quadrature outputs. The constant level shift between the two plots is primarily due to the non-exact component value match of the resistors used for reverse termination of each output. Note the response flatness out to just above 3 MHz, where the plots begin a slight roll off. Though down by only 0.3 dB at 5 MHz, this translates directly into a 0.3 dB amplitude error for the overall detector. The vector sum circuitry cannot distinguish between a change in amplitude at the input to the detector or changes in the response characteristic of the quadrature splitter. For this reason, it is important to minimize both the absolute response ripple and the relative amplitude imbalance between outputs of the splitter within the desired band of operation, especially if the envelope detector is to be used in a high accuracy application.

The relative amplitude imbalance and deviation from quadrature phase for the



**Figure 4. Relative amplitude and phase balance of the two quadrature outputs.**

active splitter outputs is shown in Figure 4. The worst-case phase deviation is  $\approx 1.7^\circ$  at 1 MHz, the low end of the frequency band of interest. In general, deviations from  $90^\circ$  of a few percent will have minimal effect on the vector sum [2,3] (and hence on the overall envelope detection error).

### Sum of Squares Function

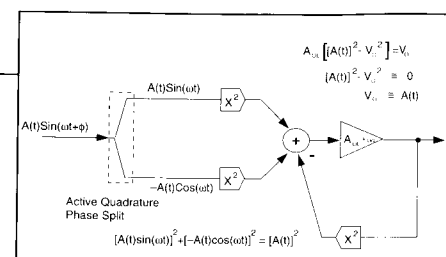
The next step in implementing the block diagram of Figure 1 involves squaring each of the quadrature modulated carriers. It is desirable to use variable transconductance type multipliers (as opposed to RF mixers) to carry out this function, to eliminate any need for filtering unwanted harmonics at the output. Each carrier is applied to both inputs of its associated multiplier. The resulting outputs are then

$$[A(t)\sin(\omega t)]^2 = \frac{A(t)^2}{2} [1 - \cos(2\omega t)] \quad (4)$$

$$[-A(t)\cos(\omega t)]^2 = \frac{A(t)^2}{2} [1 + \cos(2\omega t)]$$

where the modulation function  $A(t)$  is greater by a factor of  $\sqrt{2}$  than in equation 3; This is a consequence of using an active  $90^\circ$  phase split as opposed to a power divider. As can be seen by equation 4, the multiplier chosen should be capable of operating at twice the carrier frequency. In addition it should be a four quadrant device, since both inputs must be capable of accommodating a bipolar signal.

The Analog Devices AD834 transconductance multiplier was selected for this application due to its excellent combination of accuracy, dynamic range and DC to 500 MHz bandwidth. The AD834 also has differential current outputs which can be paralleled with one or more similar devices, providing a convenient method of summing their outputs. Since current addition is inherently wideband, use of these multipliers eliminates the need for an additional DC coupled broadband summing device. Adding the outputs from



**Figure 5. Block diagram of detector showing squaring function in feedback loop which performs square root extraction.**

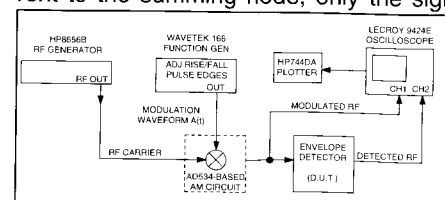
both AD834s we obtain

$$408\text{mV} \left[ \frac{A(t)^2}{2} [1 - \cos(2\omega t)] + \frac{A(t)^2}{2} [1 + \cos(2\omega t)] \right] = 408\text{mV} [A(t)]^2 \quad (5)$$

where the 408 mV scaling factor is a result of the +4 mA output current of the AD834 (from its transfer function) through  $51\Omega$  resistors R66 and R67 in the schematic of Figure 2.

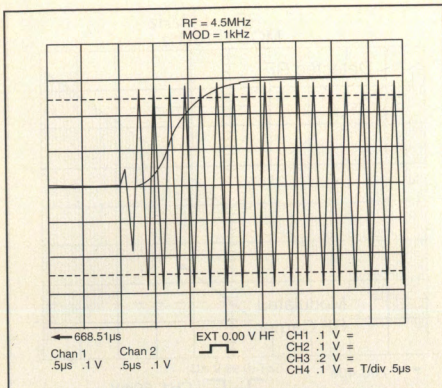
The differential nature of the AD834 outputs requires the use of a difference amplifier (e.g. an op-amp or an instrumentation amplifier) to convert the "sum of squares" into a single-ended voltage. The amplifier chosen for this purpose need not be capable of operation at the RF frequency; equation 5 illustrates that once the squares of the quadrature carriers are added together, all that remains is a scaled version of the square of the modulation term, devoid of any RF. Selection of an appropriate differential amplifier depends upon the delay and common mode input voltage handling of the amplifier, as well as on the slew rate and the bandwidth of the modulation. As indicated in the circuit schematic of Figure 2, an Analog Devices AD847 operational amplifier was chosen for this task.

Inspection of the circuit schematic in Figure 2b reveals that each AD834 is configured to output the negative of the square of its input. This is done as a practical matter to reduce crosstalk between the input and output pins. Since the same technique is applied to all three devices which contribute differential current to the summing node, only the sign



**Figure 6. Test set-up for time domain performance evaluation.**





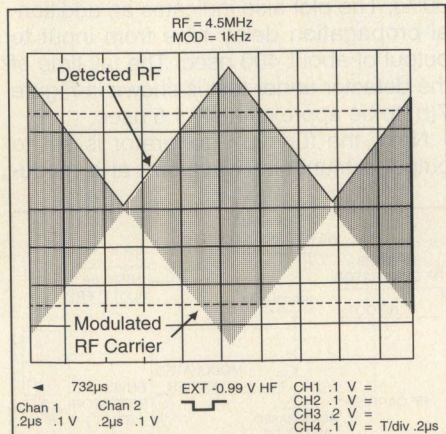
**Figure 7a. Risettime of the constructed envelope detector.**

of the final result is affected. The reader is directed to references [5] and [6] for information regarding the practical aspects of circuit design with the AD834 multiplier.

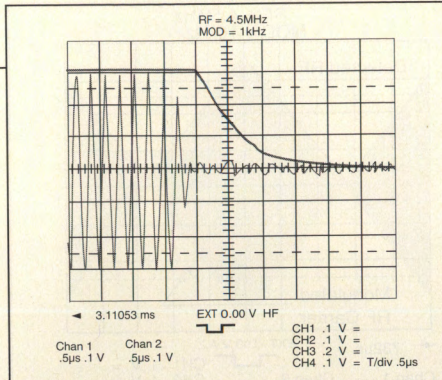
### Square Root Function

To complete the analog vector sum computation (and hence reconstruct the modulating waveform  $A(t)$ ), it is necessary to take the square root of the result found in equation 5. Several analog computational ICs capable of performing this calculation are available commercially. One should note, however, that this computational step presents significant limits on both the dynamic range and time delay specifications for the overall envelope detector. For example, a 40 dB dynamic range at the output of the square-root function (the output of the detector) necessitates an 80 dB range at its input. In addition, the device selected needs to have sufficient bandwidth to accommodate both the expected modulation frequency and the overall delay specifications for the circuit.

For the constructed envelope detector, a squarer is placed in the feedback loop of an operational amplifier to approximate the square root function shown in Figure

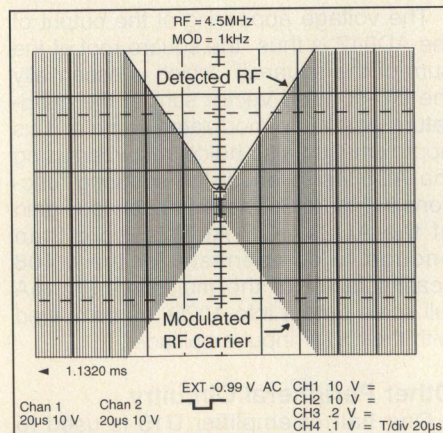


**Figure 8a. Overlay of modulated carrier and detected RF.**



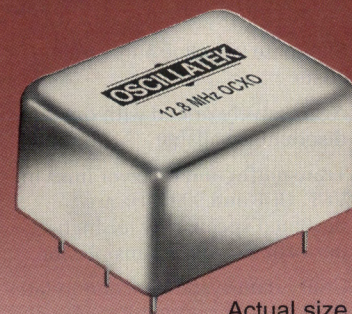
**Figure 7b. Falltime of the constructed envelope detector.**

1 [4]. This particular computational method was selected because it allows us to further capitalize on the benefits of the current summing technique used to implement the "sum of squares" in equation 5. In this case, a third AD834 multiplier operates as the squaring device. The output of the third AD834 is simply paralleled with the outputs from the two previous multipliers, but with its two output pins reversed. This has the effect of subtracting the differential output current of the third AD834 from the sum of the other two. [Note, however, that the common mode currents of all three multipliers still add]. The difference voltage across the summing node is amplified by the full open loop gain of the AD847 operational amplifier, which changes its output voltage until the voltage at both its inputs are equal (i.e., the differential summing node nulls to zero). A simplified block diagram representation of this feedback loop is illustrated in Figure 5. This diagram is meant as a generalized illustration of the envelope detector. As such, the scaling details of the AD834s used to effect the squaring function blocks have been left out for clarity. (It can be seen from the schematic of Figure 2b that each multipli-



**Figure 8b. Expanded view of modulated carrier and detected RF.**

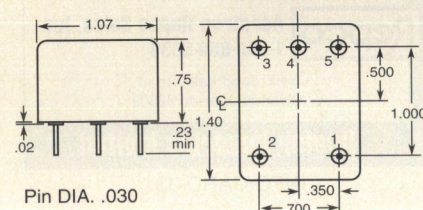
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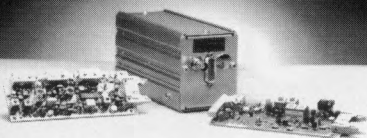
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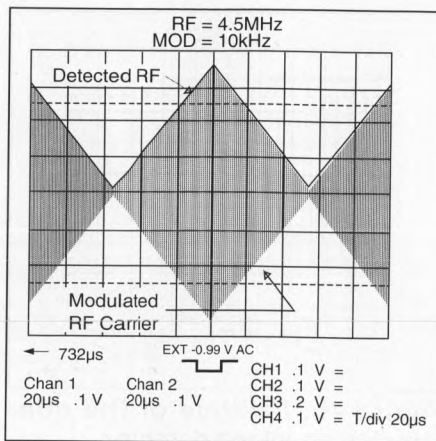


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**Figure 9a. Overlay of modulated carrier and detected RF.**

er is actually configured to output a scaled version of the negative square of its input).

As shown in the diagram, the output of the summing junction is multiplied by the open loop gain of the amplifier, resulting in

$$A_{OL} \left[ [A(t) \sin(\omega t)]^2 + [-A(t) \cos(\omega t)]^2 - V_0^2 \right] = V_0 \quad (6)$$

If the open loop gain is assumed to be large ( $A_{OL} \approx 70$  dB for the AD847), then

$$(A(t) \sin(\omega t))^2 + (-A(t) \cos(\omega t))^2 - V_0^2 \Rightarrow 0 \quad (7)$$

where each term on the left side of the equation represents the differential output from one of the three multipliers. Substituting equation 5 into 7 and solving for  $V_0$  results in

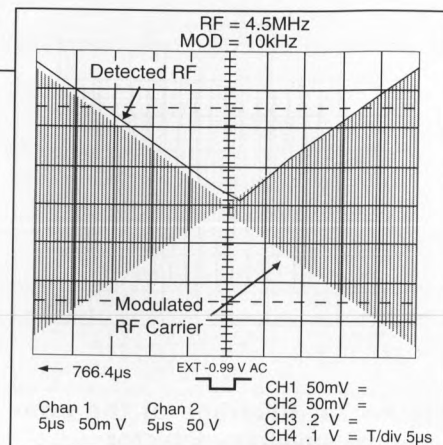
$$A(t)^2 - V_0^2 \approx 0 \quad (8)$$

$$V_0 \approx A(t)$$

The voltage appearing at the output of the AD847 is thus "the square root of the sum of the squares", which is essentially the DC-coupled vector sum of the quadrature amplitude modulated carriers. This approximation also holds true when using the AD834s to effect the squaring functions in Figure 5; The high open loop gain of the AD847 OP-AMP is more than enough to compensate for the 0.408 scaling factor of the multipliers ( $\pm 4$  mA full scale output into  $102\Omega$ , as compared with their  $\pm 1$  V input amplitude).

### Other Peripheral Circuitry

One half of amplifier U15 is used to adjust the gain and offset of the detected RF (this particular adjustment is application specific). U15-U20 serve to buffer and distribute the detected RF, with each



**Figure 9b. Expanded view of modulated carrier and detected RF.**

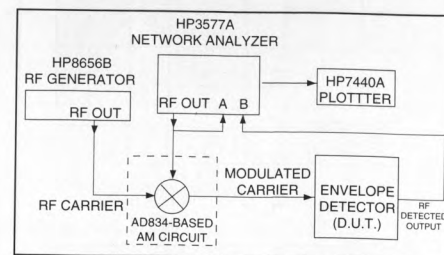
output capable of driving a  $50\Omega$  load. A regulated supply is important for this circuit because the AD 834s have current source outputs.

### Results

To evaluate the performance of the constructed envelope detector it is necessary to have some method of amplitude modulating an RF carrier. The performance of the modulator must meet or exceed that of the envelope detector if the test results are to be meaningful. For the results presented here, the device used to accomplish this task is a circuit based on the AD834 multiplier.

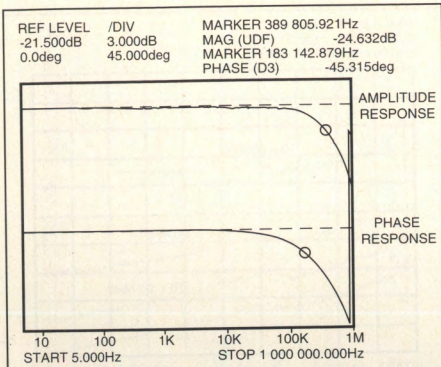
Figure 6 shows the test set-up used for the time domain plots of Figures 7a and 7b. The Wavetek function generator is first set to output a square pulse as the amplitude modulating function  $A(t)$  for the purpose of measuring the rise and fall times of the envelope detector. The RF source is set to output a carrier frequency of 4.5 MHz at the maximum amplitude acceptable to the modulator. Figure 7(a) shows the rise time of the envelope detector to be on the order of  $1\mu s$  (10%-90%); The plot also indicates an additional propagation delay time from input to output of about 400 nsec. The fall time of the detector under test is shown in Figure 7(b) to be approximately 1.5 usec.

Next, the function generator is set to output a triangular waveform at a modu-



**Figure 10. Test set-up for network analyzer based measurements.**





**Figure 11. Comparison of detected RF with modulation signal.**

lation frequency of 1 kHz. The amplitude of the waveform is adjusted to achieve close to 100% modulation of the carrier. Figure 8(a) shows the oscilloscope traces of both the modulated 4.5 MHz carrier and the detected RF (both traces overlay one another). Figure 8(b) is an expanded view, showing the amount of "failure to follow" distortion at low signal levels. These plots give a rough indication of the linearity and dynamic range for the detector, and can be used for initial circuit alignment. The lower bound on the dynamic range is minimized by adjusting the "zero input offset" potentiometer. Post processing gain and offset adjustments are determined by the particular application requirements for the circuit. Figures 9(a) and 9(b) are similar plots, at a modulation frequency of 10 kHz.

Figure 10 illustrates the set-up used to obtain the plots in Figures 11 through 14.

For these measurements, an HP3577A network analyzer is used to modulate the RF with a swept sinusoid. The output of the envelope detector is then compared with the original modulating waveform to determine its accuracy under a variety of carrier conditions. For the tests that follow, the AD834-based modulating circuit shown in the diagram is adjusted to provide approximately 20% carrier modulation. The reader should note that the following results represent the performance of the modulator/envelope detector pair, since the modulator is in fact inside the measurement loop of the network analyzer. The delay performance of the modulator is expected to be better than that of the detector because of its 30 MHz bandwidth. Amplitude accuracy and dynamic range, however, are expected to be similar, based upon these specifications for individual components utilized in the modulator circuitry.

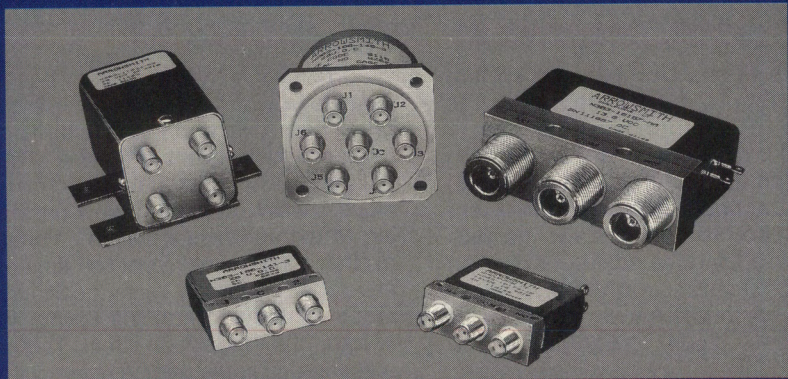
The plot of Figure 11 shows the amplitude and phase response of the envelope detector for a single carrier frequency of  $f_c = 4.5$  MHz, with the modulating signal swept from 5 Hz to 1 MHz. The top trace is the amplitude response (3 dB/div), indicating a -3dB bandwidth of just below

390 kHz. A high degree of amplitude accuracy is indicated by the response flatness, which is shown to be better than 0.5 dB (+0.25 dB) up to a modulation frequency of 100 kHz. For those applications requiring the detector to reside within a feedback loop, trace 2 is the phase response (45°/div) of the detector. The circuit exhibits 45° of phase shift at a modulation frequency of approximately 180 kHz.

Figure 12 consists of 10 individual amplitude response plots, overlaid on the same graph. The plots are intended to show the dynamic range of the detector with respect to the amplitude of the carrier signal. Each of the traces results from a 5 dB decrease in carrier amplitude. [In each case, the carrier has approximately 20% amplitude modulation]. The demodulation bandwidth does not start to significantly decrease until

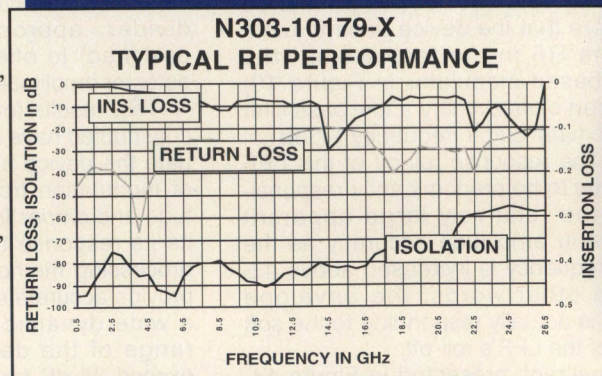
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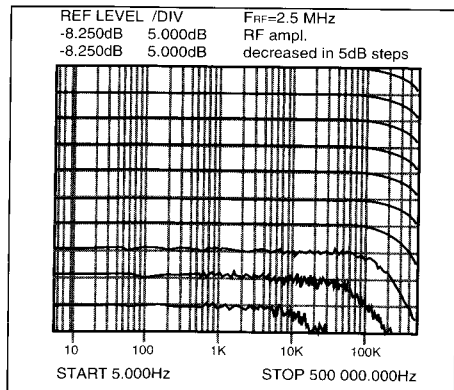
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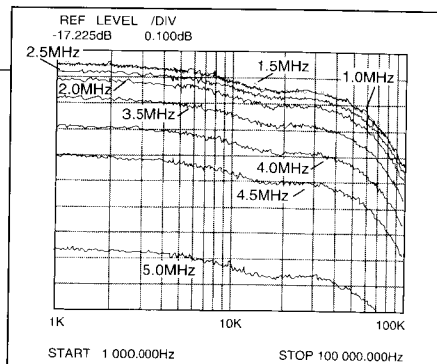
**Figure 12. Detected RF vs. modulation frequency for various carrier amplitudes.**

the carrier is approximately 35 dB down from its maximum amplitude. The detector circuit is still observed to be usable to -45 dB, but with modulation bandwidths only up to 10 kHz.

It is also desired to evaluate the envelope detector at different carrier frequencies, since the circuit constructed was designed to operate with carrier frequencies over a span of 1 MHz - 5 MHz. The results of such a measurement are shown in Figure 13. For this test, the modulation frequency is swept from 1 kHz to 100 kHz. Since we are looking for small (on the order of tenths of a dB) changes, the upper modulation frequency was limited to 100 kHz for the purpose of expanding the scale per division of the plot. Each trace in the figure represents a different carrier frequency, ranging from 1.0 MHz to 5.0 MHz in 500 kHz increments.

Note that the shape of all nine plots are similar. From the graph, carrier frequencies from 1 to 3 MHz are clustered within 0.07 dB of one another. Significant amplitude errors begin to be observed above approximately 3 MHz, some of which can be accounted for in Figure 3 (the amplitude response of the quadrature phase split circuitry). It should be noted here that the device used to modulate the RF in the test set-up (the AD834-based modulator in Figure 10) has as part of its circuitry a low pass filter with a nominal cutoff frequency (-3dB) of 6 MHz. The response roll-off of this LPF is believed to be predominantly responsible for the relatively large apparent decrease in amplitude accuracy as the carrier frequency is increased above 4.5 MHz. In other words, the envelope detector is actually responding to the soft portion of the LPF's roll-off.

The final plot, presented in Figure 14, shows the delay of the circuit for various carrier frequencies. In this measurement, the modulation frequency is swept from 10 Hz (the lowest frequency available from the network analyzer) to 500



**Figure 13. Envelope detector output versus modulation frequency at different carrier frequencies.**

kHz. The upper modulation frequency and lower carrier frequency were purposely chosen to coincide. It is interesting to note that the envelope detector is capable of demodulating frequencies approaching that of the carrier; this is possible because no explicit filtering is used during the AM detection process.

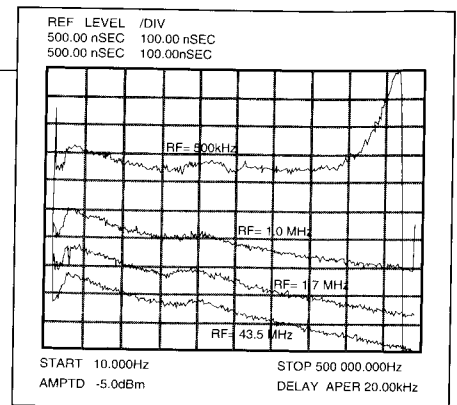
### Conclusion

An envelope detector circuit has been developed which operates over a carrier range of 1 MHz to 5 MHz. Specific delay requirements prevented the use of classical rectifier or product detector techniques, which depend on low pass filtering to extract the modulation signal. To minimize the overall response time of the envelope detector, a vector addition technique is used to null the carrier while keeping the modulation intact. The circuit presented is implemented using currently available high speed analog transconductance multipliers, and achieves amplitude demodulation at frequencies approaching that of the carrier, without the long time constants usually associated with other detection methods. The specific RF carrier bandwidth is easily modified by replacing the active all-pass section of the circuit with a 90° power divider, appropriately selected (or designed) to effect the desired overall detector amplitude accuracy.

Test results for the envelope detector constructed have been presented to evaluate the device under varying conditions of modulation frequency, carrier amplitude and carrier frequency. In addition to its  $\mu$ s response time, use of the vector processing method allows the device to provide accurate envelope detection over a wide dynamic range. The dynamic range of the device is measured to exceed 35 dB for modulation signals up to 100 kHz (45 dB for modulation below 10 kHz).

### Acknowledgments

The author wishes to express his



**Figure 14. Time delay for various carrier frequencies.**

appreciation to Dr. J.M. Brennan for his consultation and incitement throughout the design of this circuit, and for his review of the manuscript. This work was performed under the auspices of the U.S. Department of Energy. **RF**

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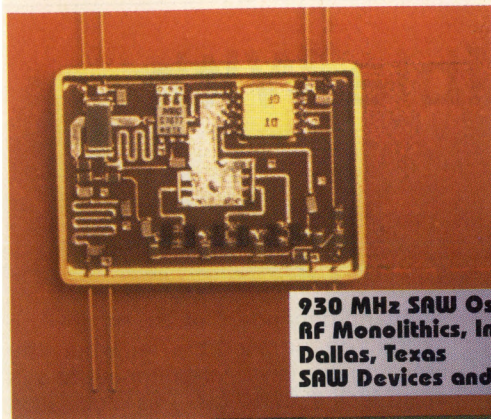
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### About the Author

Dominic J. Ciardullo is a Research Engineer in the Low Level RF group at Brookhaven National Laboratory. He received an Associates Degree in Engineering Science from Nassau Community College, BSEE from Rensselaer Polytechnic Institute and MSEE from the Polytechnic University. He is on the adjunct staff in the Ens./Phy./Tech. department at Nassau Community College. He can be reached at Brookhaven National Laboratory, Building 911B, Upton, NY 11973

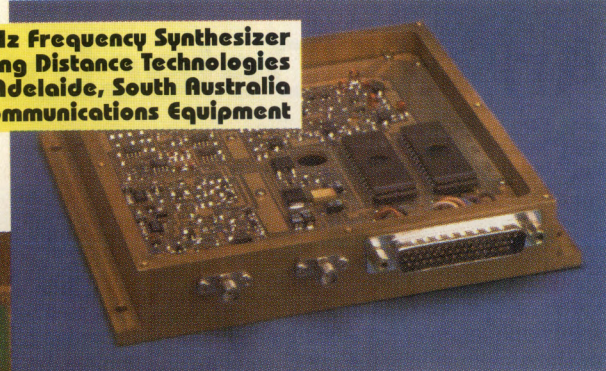


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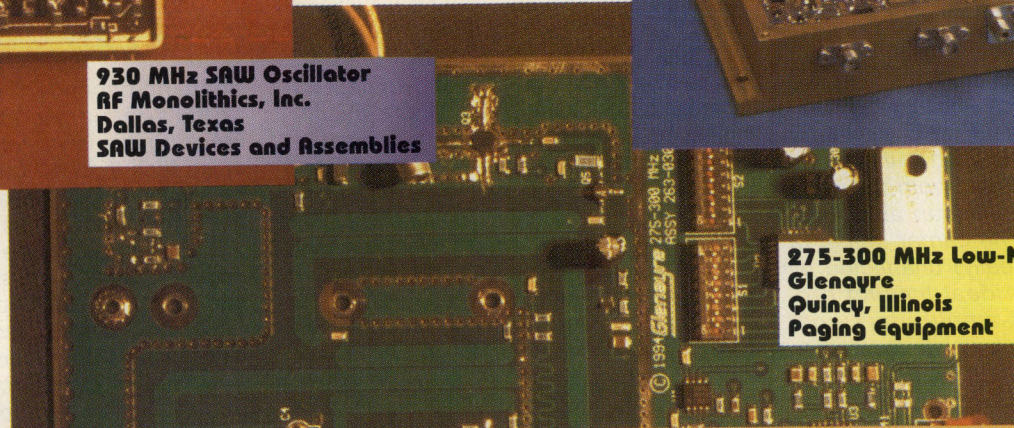


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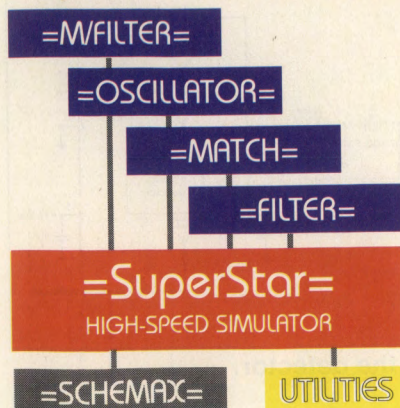
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## An AM Detector Compatible With SSB-SC

By Chase P. Hearn  
NASA

A recent article [1] presented an AM detection circuit which is basically a correlation detector. The local oscillator (LO) signal was generated with a zero-crossing detector driven from the input signal, which is basically an amplitude-limiting operation. The carrier phase of an AM signal is (ideally) unaffected by the sidebands, so ideal limiting produces a constant-amplitude signal which is phase coherent with the carrier. This LO signal actuated an analog gate to perform the required multiplication operation. This implementation produced the desired linear relationship between the input signal and the demodulated output over a 50 dB dynamic range.

Another approach to correlation detection is diagrammed in Figure 1. The LO is also derived from the input signal, but the carrier is extracted by a passive narrowband filter ( $B_c < 0.1 B_{IF}$ ) prior to limiting. This extends linear operation to input signal-to noise ratios (SNRs) lower than is possible with the usual envelope detector. This approach avoids the acquisition problems associated with a phase-locked-loop (PLL) carrier filter. The original application involved post-detection filtering of discrete sidebands in a bandwidth much less than  $B_{IF}$ . The design goal was linear detection at SNRs in  $B_{IF}$  as low as zero dB, and a dynamic range of 60 dB. These goals were fully met.

This approach is also applicable to the

detection of single sideband-suppressed carrier (SSB-SC) signals if there is sufficient residual carrier to be regenerated and maintain at least a 7-8 dB carrier-to-noise ratio (CNR) in the carrier-filter bandwidth,  $B_c$ . This was found to be possible except with very weak signals or unusually high carrier suppression. Carrier bandwidths between 300 and 30 Hz were evaluated, which provided CNR improvements of 14 to 24 dB with  $B_{IF} = 8$  kHz. When the CNR in  $B_c$  becomes too low, the detected audio quality suffers, and better performance is obtained with an independent LO signal, as normally used for SSB detection.

The operational behavior of this circuit is noticeably different from a diode AM detector. The difference between correct and incorrect tuning is more prominent, and is characteristic of the locking of a PLL demodulator. This is attributed to the narrow bandwidth of the carrier filter and the frequency-dependent differential phase shift between the signal and LO inputs to the multiplier produced by that filter. The detection gain of a correlation detector is dependent on that phase difference, so a phase-trim adjustment may be needed in one path to set the phase difference at the detector inputs to 0 or 180 degrees with the carrier centered in  $B_c$ .

Also, care should be taken to minimize level dependent phase shift (AM-PM conversion) through the limiter. In SSB reception, sideband energy can fall within  $B_c$  if it is too wide, and produce

sideband-dependent phase modulation of the regenerated carrier when the resulting carrier-to-interference ratio is much less than 10 dB. This modulation dependent LO "phase jitter" affects the demodulation process in the same way additive input noise affects a diode envelope detector: the optimum (coherent) phase relationship between the signal and LO is corrupted, and the demodulated output degraded.

A detailed circuit diagram of a "compatible linear demodulator," or CLD, operating at 500 kHz is shown in Figure 2 [2]. The CA-3089E is a multi-function IC designed as a quadrature FM detector; only the limiter and product detector are used in this application. The source-follower input amplifier was added to provide an input resistance compatible with vacuum tube circuitry. The carrier bandpass filter consisted of two series resonant crystals coupled capacitively to produce a  $B_c$  of approximately 75 Hz. A bandwidth of 300 Hz and a single-pole response was found satisfactory for AM operation, but smaller bandwidths and steeper skirt selectivity produced better SSB reception at the expense of more critical tuning. This is appreciated by noting that with  $B_c = 30$  Hz, a receiver (LO) drift of 15 Hz would produce a large signal-LO differential phase shift at the demodulator inputs.

Circuit evaluation consisted of quantitative dynamic range and distortion measurements and listening to AM broadcast, radio amateur and short-

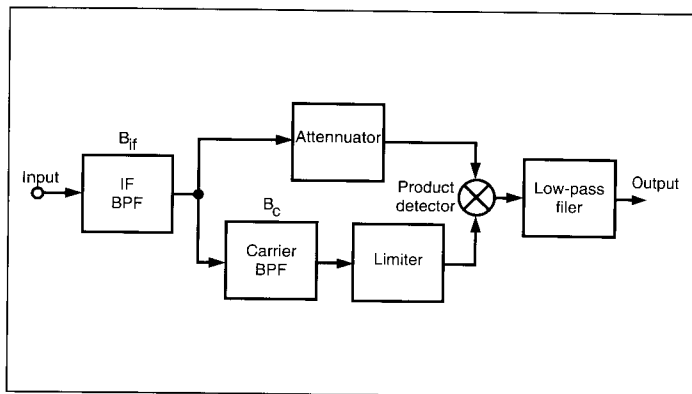


Figure 1. Block diagram of the detector.

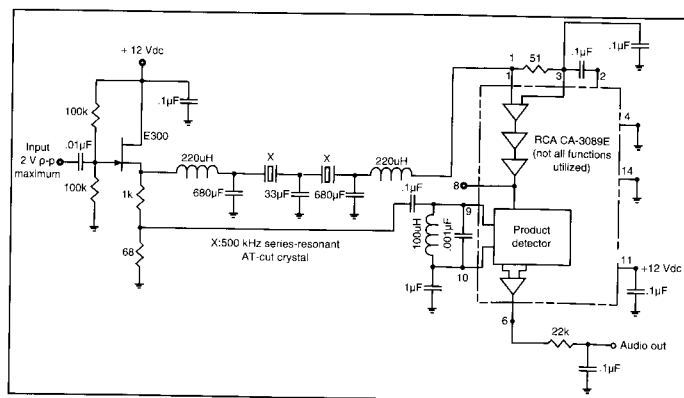
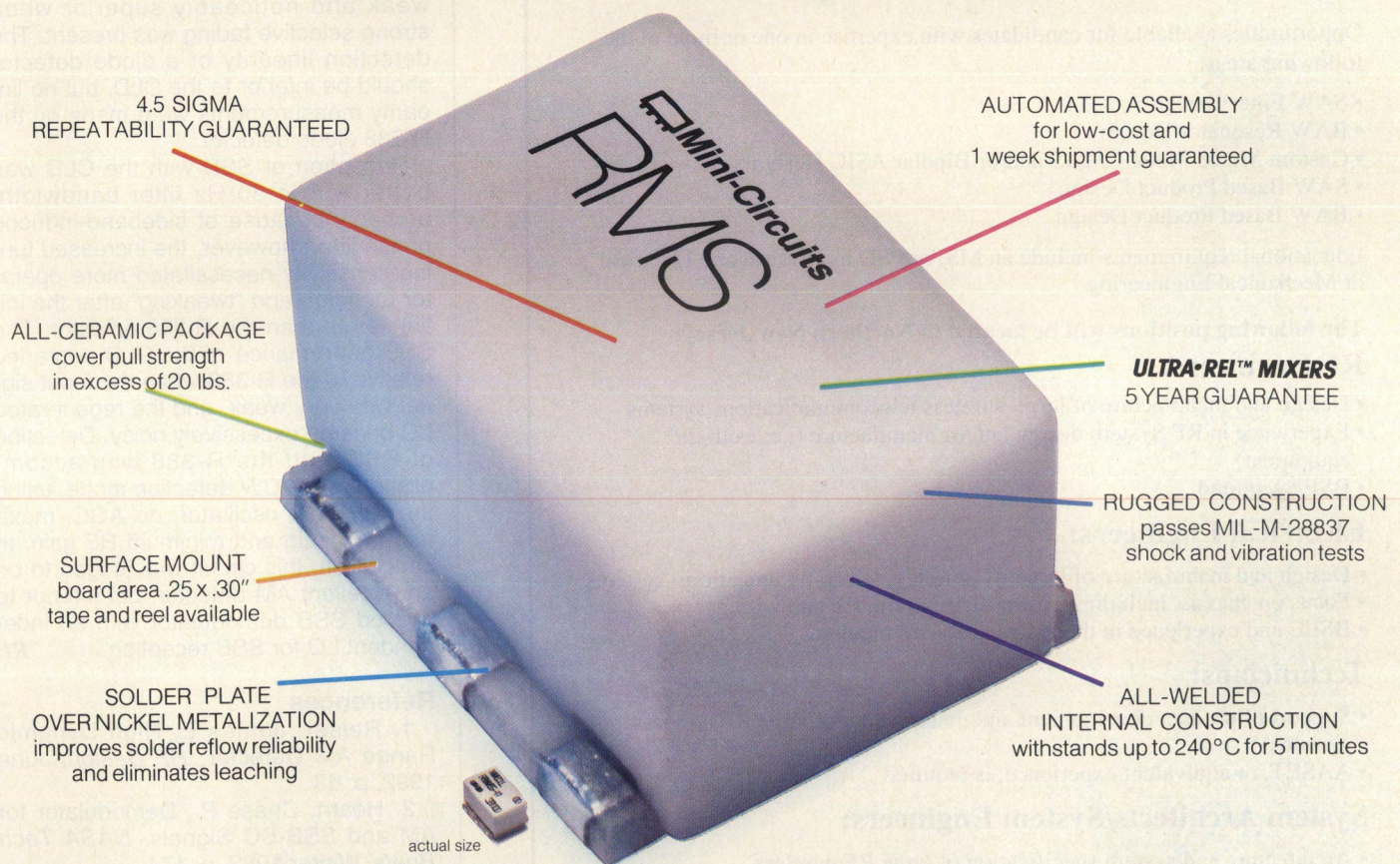


Figure 2. Schematic of the detector.



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wave transmissions, and subjectively comparing the performance to the internal diode detector in a military R-388 receiver.

The experimental detector was usable with AM inputs less than 10  $\mu\text{V}$  to 0.1 volt, a dynamic range of more than 60 dB. Total harmonic distortion (THD) was at least -40 dB. The unusual tuning characteristic of the CLD was not found to be an operational problem; no further adjustment was necessary once the signal was properly "tuned in".

The CLD was judged to be marginally better when an AM signal was very weak and noticeably superior when strong selective fading was present. The detection linearity of a diode detector should be inferior to the CLD, but no linearity measurements were made on the R-388 diode detector.

Reception of SSB with the CLD was better with a 30 Hz filter bandwidth, probably because of sideband-induced phase jitter; however, the increased tuning sensitivity necessitated more operator attention and "tweaking" after the initial set-up than the R-388 detector. The SSB performance of the CLD degraded relative to the R-388 when the input signal was very weak, and the regenerated LO became excessively noisy. Detection of SSB with the R-388 was accomplished in the CW detection mode, using the CW beat oscillator, no AGC, maximum AF gain and minimum RF gain. In conclusion, this circuit was judged to be an excellent AM detector, but inferior to a good SSB demodulator with an independent LO for SSB reception. *RF*

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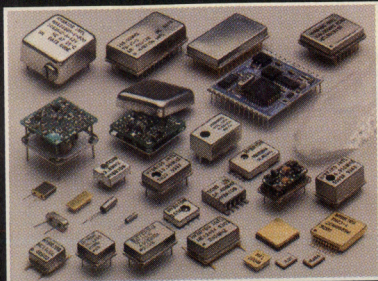
### **About the Author**

Chase P. Hearn has been a Research Engineer with NASA since 1961. His work there has involved analysis, design and development of circuits and systems spanning VLF to microwaves, and microwave measurement theory and practice. He has been a radio amateur since 1954 and currently enjoys "homebrewing", working with vacuum-tube equipment and the discrete-component technology of the forties and fifties. He can be reached at 104 Glenwood Dr., Williamsburg, VA 23185.



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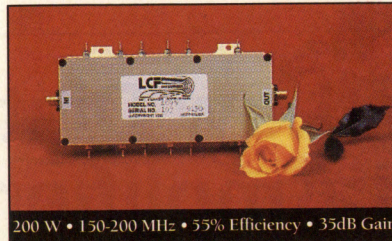
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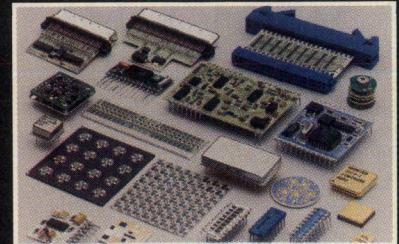
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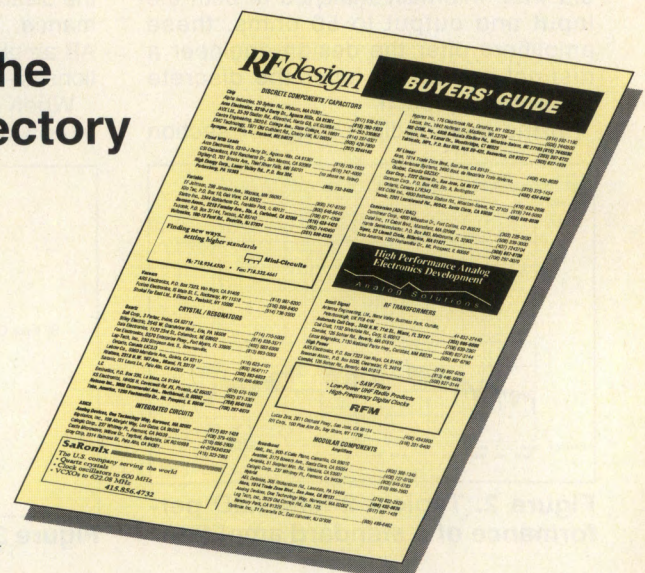
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By Walter Koprowski  
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The use of sophisticated radio technology to provide basic telephone service has created a need for cost-effective high power amplifiers. The technical requirements are that each amplifier will handle 1-24 RF channels and provide high peak power output with low intermodulation products ( $-55$  dBc). For this application, multiple amplifiers are installed in an unattended station that provides wireless telephone service to isolated subscribers and emergency services, and offers new choices for telephone service in developing countries.

**S**olid state amplifier technology has been in existence for over thirty years. The common design procedure is to choose the highest gain and power capability for the output transistors and then select the necessary driver stages for the overall gain required. A typical block diagram is shown in Figure 1.

The driver stages are class A biased, low cost, high performance hybrid amplifiers. These amplifiers were specifically designed for low distortion and wide dynamic range performance required in the CATV industry. The output capability of the hybrid amplifiers is approximately 0.5 watt. Internally matched at both the input and output to 50 ohms, these amplifiers offer the design engineer a distinct advantage over the discrete transistor approach.

In the intermediate amplifier section

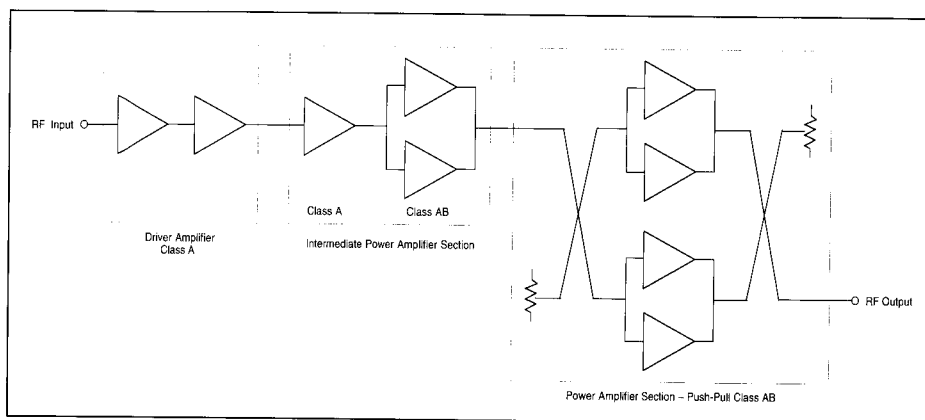


Figure 1. Standard high power amplifier design.

are two discrete bipolar transistors. The first stage biased class A, and the second stage is a push-pull device biased for class AB operation.

The PA section of a standard high power amplifier is designed utilizing two push-pull transistors biased class AB which are connected together using a 3 dB quadrature hybrid for balance and good input/output VSWR. The advantage of using class AB is improved power output capability at a lower power consumption than a class A. However, the disadvantage is lower linearity performance. Typical performance of a class AB amplifier under two-tone conditions is  $-30$  dBc, as shown in Figure 2.

When confronted with a specification of  $-55$  dBc intermodulation distortion at

a given power output, the designer must either use a very large class A biased amplifier, or use some means of distortion cancellation.

### Feed-Forward Cancellation

The concept of feed-forward cancellation has been used by manufacturers of hybrid CATV amplifiers for many years. The technique is to isolate the distortion of the power amplifier and re-inject it 180 degrees out of phase to cancel the distortion at the output. A standard feed-forward amplifier system is shown in the block diagram of Figure 3.

The RF input signals are equally split by means of a two-way divider. Half of the signal power goes into the power amplifier, and half goes through a delay

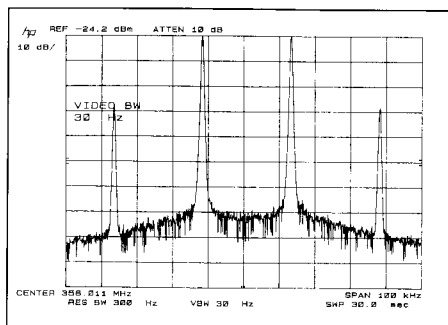


Figure 2. Typical 2-tone IMD performance of a standard amplifier.

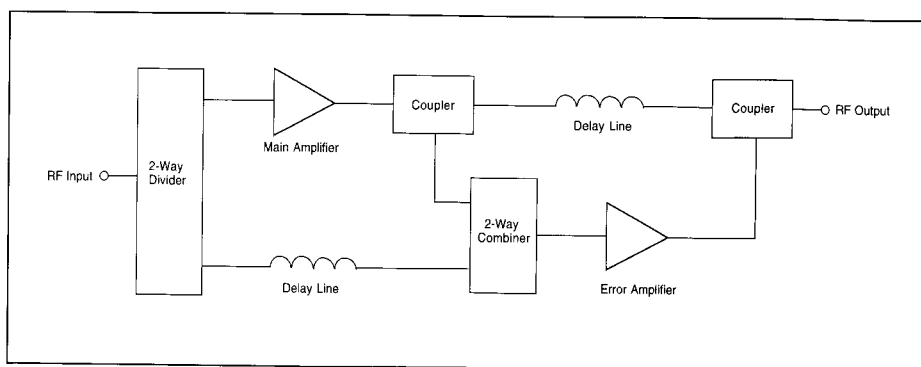
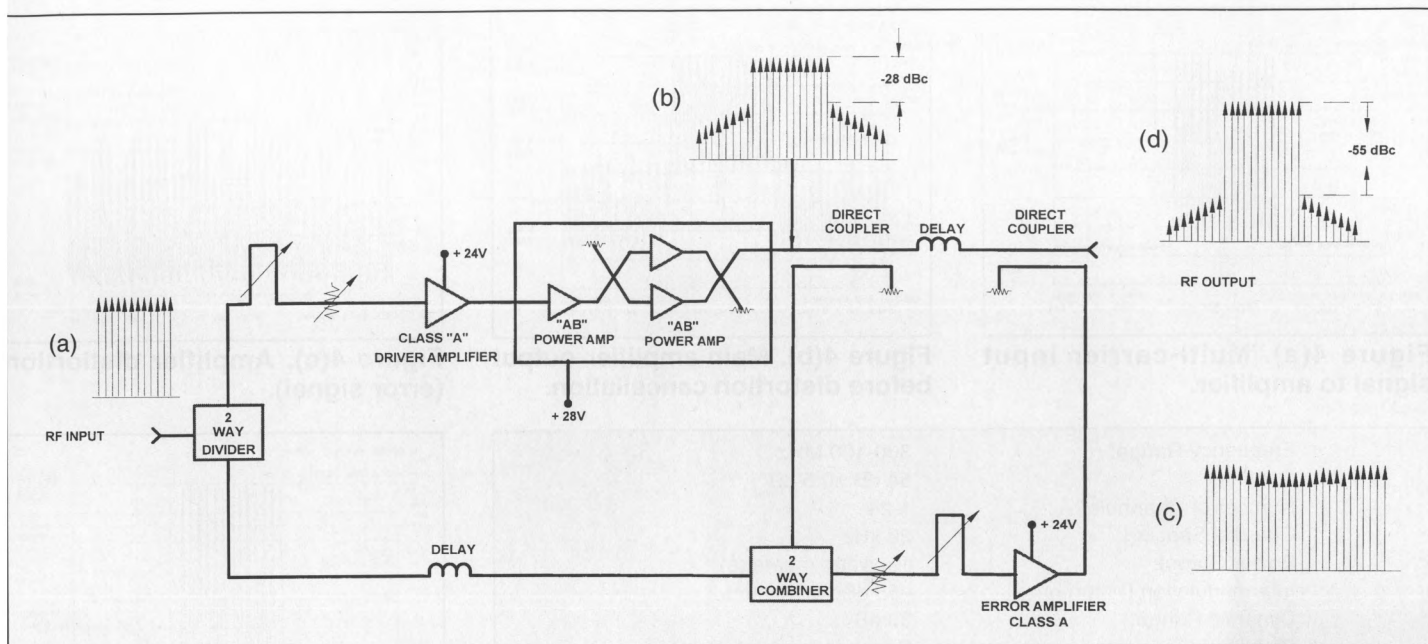


Figure 3. Standard feed-forward system block diagram.





**Figure 4. Block diagram of a 358 MHz multitone feed-forward amplifier for 24 carriers spaced 25 kHz.**

line. The multiple tones are amplified in the main amplifier, which also generates a certain amount of intermodulation distortion. A directional coupler is used to obtain low-level sample of the signal and its distortion, and provide it to one of the inputs of a two-way combiner that is part of the error signal path.

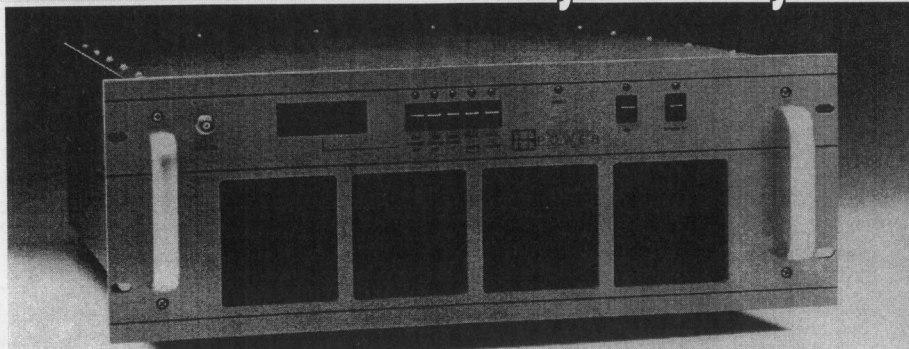
The 0 degree two-way combiner will algebraically sum the signals as long as both inputs are of equal phase and amplitude. The purpose of the delay line between the input divider and the error signal combiner is to assure that the input carriers and the sampled main amplifier carriers arrive at the two-way

combiner at the same time and 180 degrees out of phase. Fine amplitude and phase adjustment is accomplished via an attenuator and phase shifter in the main amplifier section. The signal carriers are canceled, and the output of the combiner, which now contains only the distortion products, is fed into the error amplifier. The error amplifier is a class A ultra-linear design, and its function is to amplify the distortion required in the cancellation loop to cancel the intermodulation products. The error amplifier output is connected to a directional coupler used for distortion cancellation. The choice of coupling factor is

important so as to maintain minimum main line loss as well as to limit the power capability of the error amplifier. For proper cancellation to take place at the output, it is imperative that the signals are equal in amplitude and 180 degrees out of phase when they reach the coupler. The delay line in the output main line compensates for propagation time in the error amplifier. Fine phase and amplitude adjustments are accomplished using a phase shifter and attenuator in the error amplifier.

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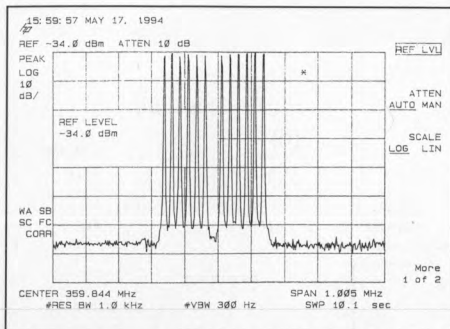
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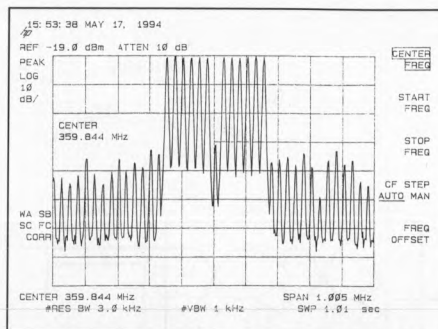
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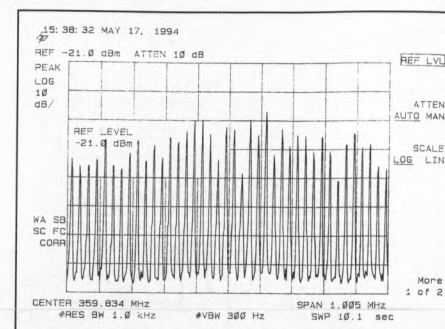




**Figure 4(a). Multi-carrier input signal to amplifier.**



**Figure 4(b). Main amplifier output before distortion cancellation.**



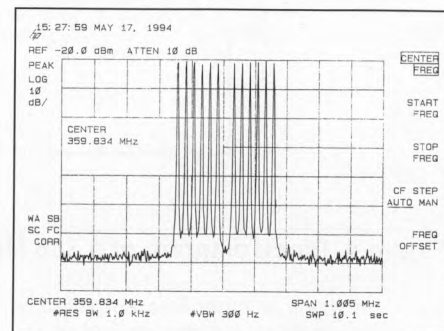
**Figure 4(c). Amplifier distortion (error signal).**

Frequency Range:	300-400 MHz
Gain:	54 dB $\pm$ 0.5 dB
Number of Channels:	1-24
Channel Spacing:	25 kHz
Power Output:	12 watts (average)
Intermodulation Distortion:	-55 dBc maximum
Dynamic Range:	30 dB
Stability:	Unconditionally stable for any source and load impedance
Environmental:	0 to +50°C
DC Power:	-42 to -54 VDC, 6 A maximum
Size:	15 x 21 x 5.25 inches

**Table 1. Feed-forward amplifier specifications.**

titone amplification of signals at 358 MHz is shown in Figure 4. The RF input accepts up to 24 carriers which are spaced 25 kHz apart. The input signals, shown in Figure 4(a), are divided in power to provide the drive to the main amplifier, and to the delay line that feeds the two-way combiner. The intermodulation distortion of the main amplifier is

-28 dBc, shown in Figure 4(b). The carriers are canceled in loop 1, and the residual distortion is amplified by the error amplifier. Figure 4(c) shows the distortion that feeds the output directional coupler used for final cancellation. The final output shown in Figure 4(d) is the result of a feed-forward cancellation which exceeds -55 dBc.



**Figure 4(d). Output after feed-forward cancellation.**

The amplifier system is housed in an enclosure 15x21x5.25 inches designed for rack mounting. The system contains internal heat sinks and fans that draw air from front to rear and provide proper cooling. Power is supplied by two high efficiency DC-DC converters. A fault/alarm board provides local and remote identifications of amplifier overdrive and excessive output load VSWR. Specifications for the amplifier are contained in Table 1.

## Conclusion

This article describes the overall operation of a multitone high power amplifier. The use of feed-forward cancellation techniques allows for significant intermodulation distortion improvements and encourages the use of class AB linear power amplifiers for applications in wireless communications equipment.

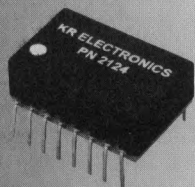
Readers may obtain more information about this amplifier by contacting the author at the address below, or by circling Info/Card #201. **RF**

## About the Author.

Walter Koprowski is Vice President, Amplifier Engineering at Power Systems Technology Inc., 105 Baylis Road, Melville, NY 11747. He can be reached at (516) 777-8900, or by fax at (516) 777-8877.

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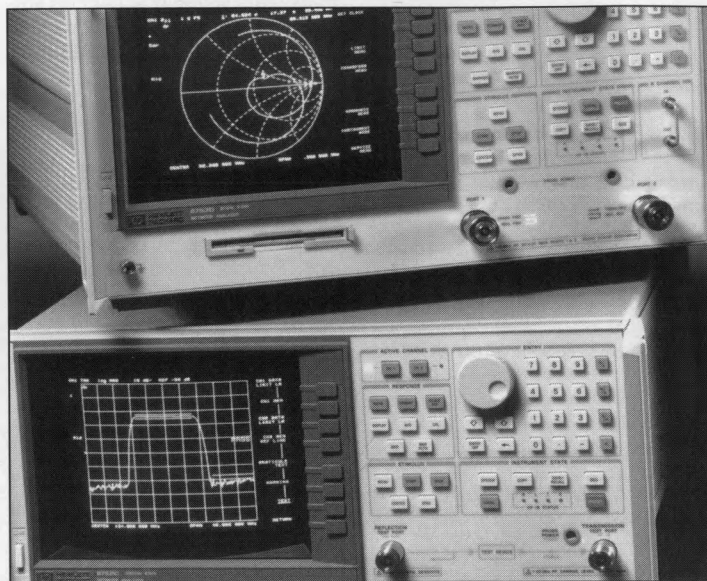


## Network Analyzers Upgraded

Hewlett-Packard has upgraded the performance and reduced the price on two of its high-performance network analyzers, the HP 8753 and 8752. The HP8753D now has an integrated S-parameter test set, improved dynamic range, TRL and LRM calibration, power specified at the test port and added power control features, 30 kHz start frequency, and an optional high-stability frequency reference. In addition, the HP 8753D has a new disk drive which supports both LIF and DOS formats, a new CPU which is 67% faster, and improved data and port-handler interfaces. Dynamic range is improved in both the standard 3 GHz source and optional 6 GHz

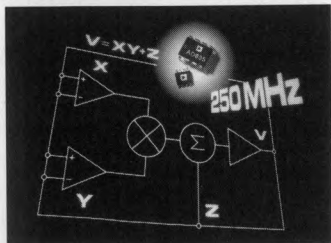
source by eliminating a doubler. Dynamic range for the 6 GHz source has been improved 25 dB to 105 dB, and the 3 GHz source has been improved 10 dB to 110 dB. The HP 8752C now has an optional 6 GHz source, built-in step attenuator and a 67% faster CPU. Base price of the HP 8753D RF network analyzer is \$34,500. Base price for the HP 8752C is \$22,000. Until November 30, 1994, HP offers customers a 20% discount on the HP 8753D and a 15% discount on the HP 8752C for any working vector network analyzer with an operating frequency that includes 1 GHz.

**Hewlett-Packard Co.**  
**INFO/CARD #250**



## Four-Quadrant $V_{out}$ Multiplier

Analog Devices has announced the industry's first 250 MHz, four-quadrant voltage-output monolithic analog multiplier. The AD835 can generate a linear product of its "X" and "Y" input voltages with a 3 dB bandwidth of 250 MHz. Small-signal rise-time is 1.0 ns; full-scale (-1 to +1 V) rise and fall time is 2.5 ns (with 150  $\Omega$  loads). Settling time to within 0.1% of full scale is typically 17 ns. High gains are achieved over wide bandwidths with low



noise contribution (44 nV/ $\sqrt{\text{Hz}}$ ). Few external components are required to apply the AD835. Its differential multiplication (X,Y) and summing (Z) inputs are high impedance nodes that do not require signal conditioning. Likewise, its low-impedance output signal needs no additional buffering to drive  $\pm 2.5$  V into loads as low as 50  $\Omega$ . The AD835 is priced at \$7.95 in 1000s and packaged in 8-pin mini-DIPs and 8-lead SOICs.

**Analog Devices**  
**INFO/CARD #249**

## GPS Antenna Module

Toko America has introduced the AMG series, a complete antenna module for GPS applications. This miniature module includes the antenna element, bandpass filter, and low noise

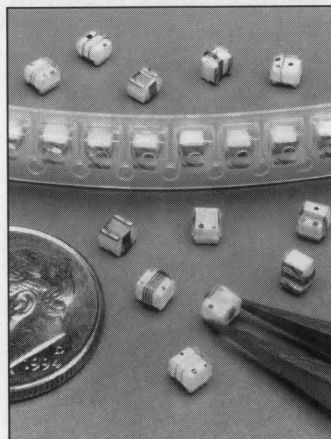


amplifier, sealed in a 38mm square package. The antenna is centered at 1575.42 MHz for GPS receivers, and operates on 4 to 5.25 V over the temperature range of -40 to +85  $^{\circ}\text{C}$ . The module provides excellent gain and directivity. Typical antenna gain is 4 dBi at 90 $^{\circ}$  angle of elevation, -4 dBi at 0 $^{\circ}$ . The maximum axial ratio is 3 dB. Typical gain of the low noise amplifier is 26 dB, with maximum current consumption of 25 mA and typical noise figure of 1.6 dB. The module can be provided with or without a radome. A cable and connector are also provided, and can be customized upon request. Samples are available from Toko America for \$150.

**Toko America, Inc.**  
**INFO/CARD #248**

## SMT Inductors

The 1008CX series from Pulse Engineering contains surface mount RF inductors with inductances from 4.7 nH to 4700 nH. The inductors are characterized at high frequencies and have high Q and high self resonant frequencies. The 1008CX series comes in the industry standard 1008-size footprint, with spot welded wire bonds that will not open under severe reflow conditions. Gold terminations are standard, with other types available.

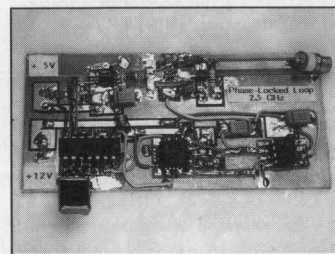


A sample kit is available for \$99 plus shipping and handling. An 0805-sized series will be available for sampling in September, with low-profile 1008-sized and ferrite core series to follow shortly afterward.

**Pulse Engineering, Inc.**  
**INFO/CARD #247**

## Prototyping System

The Wainright Solder-Mount System allows "freehand" construction of experimental circuits of all kinds - without the need for tools or services. Only diagonal cutting pliers and a soldering iron are needed. Solder-Mounts are self adhesive pieces of print-



ed circuit material (epoxy glass) with an etched and tinned pattern of soldering points - each laid out for one or more component types. To build a circuit, components are soldered to a suitable Solder-Mount, the paper cover removed from the adhesive, and the Solder-Mount with the component is then placed on a copper-clad and tinned groundplane. No layout or wiring plan is needed. The metal surface of the groundplane serves as electrical ground, providing a definite ground potential at every point in the circuit. Stray capacitance to ground is very small and can be compared to a printed circuit board.

**Wainright Instruments, Inc.**  
**INFO/CARD #246**



## SIGNAL SOURCES

### Miniature Oscillator

The Wenzel Associates Miniature Oscillator (WAMO) provides power consumption of less than 1 W at 25° C. At 1.33 x 1.33 x 1.33 inches, this 10 MHz oscillator rivals the performance of larger oscillators with temperature stability of  $\pm 5 \times 10^{-9}$  over 0 to 50° C and phase noise of -160 dBc/Hz at 1 kHz offset. Pinout is compatible with industry standard miniature designs. The WAMO is available as a MIL, Space or Commercial grade unit.

**Wenzel Associates, Inc.**  
INFO/CARD #245

### SC-Cut OCXO

Model 2930200 from Piezo Crystal Company utilizes Piezo's World Class "SC" cut crystals. The frequency range is from 80 to 110 MHz. Typical phase noise at 100 MHz is -90 dBc/Hz at 10 Hz, -120 dBc/Hz at 100 Hz, -140 dBc/Hz at 1 kHz and -153 dBc/Hz at 10 kHz. Frequency stability is  $\pm 2 \times 10^{-8}$  from 0 to +50° C, and the size of the 2930200 is 2.00 x 2.00 x 1.00 inches. The approximate price is \$350 to \$450 in quantities of 500 pieces.

**Piezo Crystal Company**  
INFO/CARD #244

### Low Profile OCXO

Piezo Technology has introduced its model XO5008, which features a maximum height of 0.75 inches, with a footprint of 1.5 x 1.5 inches. The standard center frequency is 10 MHz, with frequencies available from 3 MHz to 50 MHz. The unit features temperature stability of  $\pm 0.08$  ppm over -40 to +85° C, sinewave output, five minute warm-up, and moderate current consumption.

**Piezo Technology, Inc.**  
INFO/CARD #243

### SMT VCOs

A series of surface mount, voltage controlled oscillators is available from ST Microwave. The new design utilizes silicon bipolar

transistors with high-Q silicon varactor tuning diodes on advanced thin-film hybrid microwave integrated circuit substrates. GaAs FET buffer amplifiers are included to insure stable operation. Models are available at 2 to 6, 6 to 12, and 12 to 18 GHz, with power output of +10 dBm on the two lower frequency units and +5 dBm on the higher frequency unit. The packages measure 0.45 x 0.45 x 0.16 inches. Production quantities are available in 60 to 90 days.

**ST Microwave Corp.**  
INFO/CARD #242

## SIGNAL PROCESSING COMPONENTS

### Miniature Mixer

The ZP-11A mixer from Mini-Circuits offers space-saving miniature construction plus high reliability backed by a five-year guarantee. The RF and LO ports can operate from 1400 to 1900 MHz, while the IF operates from 40 to 500 MHz. Conversion loss specifications are statistically controlled to better than  $4.5\sigma$  from mean. Conversion loss is flat (0.6 dB typ.) over the full range, and 1 dB compression is 1 dBm typ. A miniature connectorized package houses the mixer.

**Mini-Circuits**  
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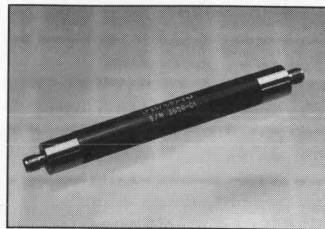
### Threshold Detector

Daico Industries introduces a flatpack threshold detector to its detector series. Model CTDO8014 has an operating frequency of 10 to 3000 MHz. Detection is over the dynamic range of -25 to -10 dBm, with 1.2 dB hysteresis. Transition time is 300  $\mu$ s, (10%/90% RF). Input flatness is  $\pm 1.5$  dB with VSWR of 1.5/1. The detector is comes in a 14-pin flatpack.

**DAICO Industries**  
INFO/CARD #240

### Bandpass and Lowpass Filters

In three different diameters, this new series of coaxial tubular filters offers either bandpass or lowpass characteristics from 30 MHz to 5000 MHz. The filters provide rejection of 70 dB and greater while VSWR is maintained at 1.5:1 or better. A wide choice of connectors is offered,



including SMA, N, TNC and BNC types as well as pins for PCB mounting.

**Atlantic Microwave Ltd.**  
INFO/CARD #239

### INMARSAT Phase Shifters

Vectronics introduces several phase shifters in the INMARSAT band for phased-array applications. Three-bit and four-bit units are available in switched-line designs (individual bit values vary linearly vs. frequency). All units feature low insertion loss and excellent phase accuracy. A number of different internal TTL driver options are available with switching times as low as 200 ns.

**Vectronics Microwave Corp.**  
INFO/CARD #238

### Diplexer

The GLDI-001 diplexer is a high performance, low cost solution to diplexing in the 30 to 152 MHz range. Maximum insertion loss is 0.7 dB, minimum isolation is 40 dB and VSWR is maintained at or below 1.5. The diplexer meets or exceeds military environmental requirements. Prices start at \$135 each in quantities of one to five.

**Geoffroy Labs**  
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## AMPLIFIERS

### Feedforward Amp

Model PA09005200-01R, a 200 W PEP, wideband, feedforward cellular base station amplifier provides intermodulation products of -60 dBc. This is accomplished over a 25 MHz instantaneous bandwidth at 20 W multichannel output power. Gain is 53 dB and the power requirement is 26 V, 9.5 A.

**AML Communications**  
INFO/CARD #236

### Broadband Amplifiers

**EM Research Engineering**

introduces its line of broadband modular amplifiers. These linear amplifiers have an instantaneous bandwidth of 10 to 1000 MHz and output powers ranging from 100 mW to 10 W with an input of 0 dBm. Variable gain control is available as an option in these amplifiers. Pricing starts at \$485 for a 1 W amplifier in small quantities (1 to 4).

**EM Research Engineering, Inc.**  
INFO/CARD #235

### 4W UHF Amplifier

ENI's model 604L power amplifier produces 4 W of linear class A output over 0.5 to 1000 MHz. With a gain of 40 dB, the 604L features harmonics more than 18 dBc below the fundamental and low intermodulation distortion. Other features include unconditional RF stability, +13 dBm overdrive protection, and infinite maximum load VSWR. The 604L is available for 30-day delivery at a cost of \$2850.

**ENI**  
INFO/CARD #234

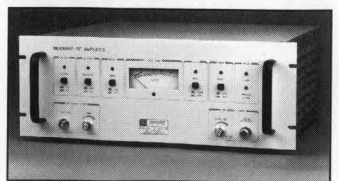
### 25-100 MHz, 100 W Amplifier

Model A005 from LCF Enterprises has an output power level of 100 W CW and operates from 25 to 300 MHz. The module measures 6.0 x 2.0 x 1.0 inches (excluding mounting feet and connectors), and delivers a minimum gain of 35 dB. Operation is from a 24 VDC supply. Customization to other frequencies and output powers is available.

**LCF Enterprises**  
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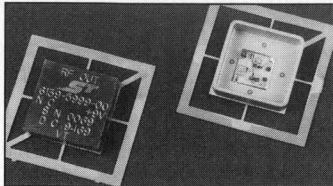
### Broadband, 100 W Amplifier

Model 7100LC is an instantaneous broadband RF power amplifier especially designed for



European EMC testing. Power output of 100 W CW covering 80-1000 MHz will offer solutions for many different applications. An IEEE-488 /RS-232 interface kit, model KEI IF-488, is available. Price for the 7100LC is \$19,500.

**Kalmus Engineering Inc.**  
INFO/CARD #232





## 10W Cellular Amp

The QBS-227 is a 10 W, 1 dB compression point, linear power amplifier operating over the 800 to 960 MHz frequency range. Having minimum gain of 40 dB, the QBS-227 uses a 24 V supply with an operating case temperature range of  $-20$  to  $+70^{\circ}\text{C}$ . With a  $+50$  dBm third order intercept point, the QBS-227 has  $-40$  dBc third order intermod product levels when two one watt tones are being amplified.

**Q-bit Corp.**  
INFO/CARD #231

## SEMI-CONDUCTORS

### SOT-143 NPNs

A family of NEC devices from California Eastern Laboratories is designed as an alternative to other manufacturers devices in mixers, oscillators and LNA applications. The NE85639R has  $f_T$  of 7 GHz and  $I_{C_{max}}$  of 100 mA, NE68139R has  $f_T$  of 9 GHz and  $I_{C_{max}}$  of 65 mA, NE68039R has  $f_T$  of 10 GHz and  $I_{C_{max}}$  of 35 mA, and NE68539R has  $f_T$  of 12 GHz. A range of performances are available from the family; noise figures from 1.1 dB at 1.0 GHz, insertion power gain to 15 dB at 1.0 GHz and outstanding low voltage/low current operation from VHF to 2.5 GHz.

**California Eastern Laboratories**  
INFO/CARD #230

### Evaluation Kits

A series of MMIC evaluation kits from Richardson Electronics contains the MMIC devices most widely used for low-noise amplifiers, general medium power applications, upconverter/downconverter and antenna switching for cellular, PCN and other applications. The kits are marketed as the RF Gain, Ltd. MMIC Evaluation Kits.

**Richardson Electronics, Ltd.**  
INFO/CARD #229

### IF ICs

AT&T Microelectronics has introduced a line of general purpose intermediate frequency (IF) integrated circuits. Included in the offering are a quadrature modulator, a programmable AGC amplifier and two quadrature demodulators with integrated AGCs. The W2009 quadrature modulator has I/Q input bandwidth of 4 MHz.

The W1466 amplifier provides up to 45 dB gain in selectable 3 dB steps; its 3 dB bandwidth is 100 MHz. Both the W1452 and W1575 combine a quadrature demodulator with digitally programmed IF AGC amplifier. Production quantities of the IC are available now, with pricing of \$2.50 for the W1466, \$4.50 for either the W1452 or W1575, and \$3.40 for the W2009.

**AT&T Microelectronics**  
INFO/CARD #228

### Power MOSFET

Motorola is announcing a new surface mount RF power MOSFET designed for broadband commercial and industrial applications at frequencies to 520 MHz. The MRF5003 device supplies up to 3 W output power at 7.5 V with minimum gain of 9.5



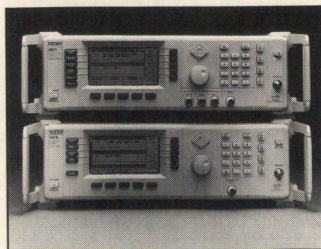
dB at 512 MHz. The MOSFET has low feedback capacitance ( $C_{rss} = 4.4$  pF typical) and can withstand 20:1 load VSWR at any phase angle when properly mounted. Pricing for the MRF5003 is \$7.91 in low volumes.

**Motorola Semiconductor**  
**Products Sector**  
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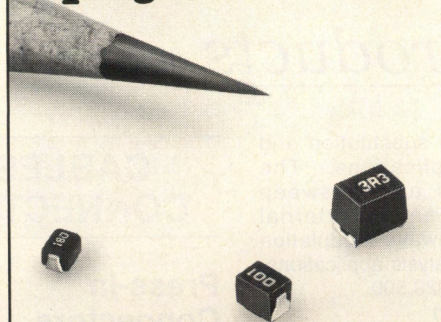
## TEST EQUIPMENT

### Synthesizers

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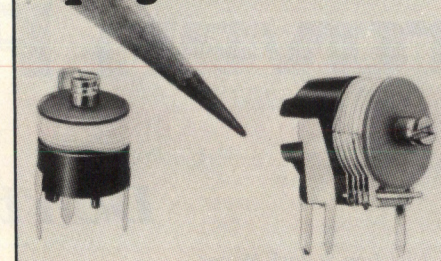
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## Synthesizer/Sweeper

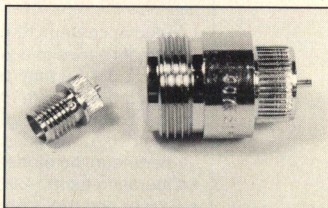
Giga-tronics has introduced 2 to 20 GHz models of its GT 9000 microwave synthesizer and GT 9000S synthesized microwave sweeper. Both the GT 9000 and 9000S have -95 dBc/Hz phase noise at 10 kHz offset from 2 GHz and output power of +13 dBm or greater. Internal pulse modulation is standard, while built-in amplitude, frequency and scan modulation are available options. The 9000S adds analog and digital sweep of both frequency and power. U.S. list prices for the 2 to 20 GHz GT 9000 and GT 9000S are \$22,950 and \$24,950, respectively.

**Giga-tronics, Inc.**  
INFO/CARD #225

## CABLES & CONNECTORS

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Delta Electronics' line of press-mount coaxial receptacle is designed to require only one through hole in the housing, as



opposed to the one hole plus two or four tapped holes for flange-mounted jacks. Delta's press-mount receptacles are currently available in type N and SMA series, and can be provided with tab, slotted, or solder pot contact ends.

**Delta Electronics**  
INFO/CARD #223

### TNC Jack

COAXICOM has announced the release of a "pre-assembled" TNC bulkhead jack for RG402 (0.141 dia.) semi-rigid cable. Model 4539CC-430-1 is directly soldered to the cable jacket after stripping back the cable's center contact and plugging it into the captive female contact at the rear of the connector. Part number 4539CC-431-1 is available for RG405 (0.085 dia) semi-rigid cable.

**Coaxial Components Corp.**  
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### High Isolation Cables

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**Tru-Connector Corp.**  
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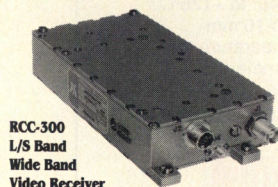
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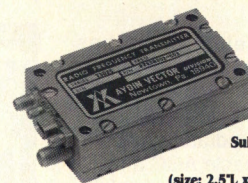


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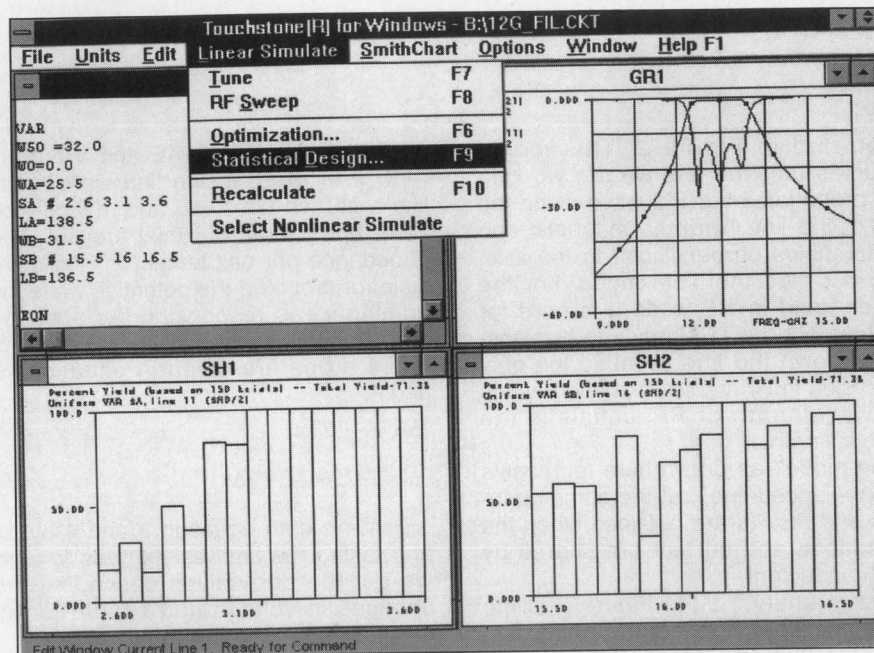
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1. Entries shall represent RF functions operating at frequencies from tens of kHz to 3 GHz.
2. Entries may be circuits, circuit design methods, test procedures, or design software programs.
3. If the entry is a circuit, it shall have a complexity equivalent to that of a circuit using 8-10 discrete active devices or 6-8 integrated circuits. The circuit may be a portion of a larger system.
4. If the entry is a design method, it must include an example of a circuit designed using the method described.
5. If the entry is a test method, it must include actual results of the measurement described.
6. If the entry is a computer program, it must operate on either an MS-DOS or Apple Macintosh system. It must be provided in a form that can be operated directly, without additional software (e.g., compiled). Programs must be submitted on disk, with supporting documentation provided in printed form.
7. Entries shall be the original work of the entrant, not previously published or publicly distributed. If developed as part of the entrant's employment, entries must have the approval of the entrant's employer.
8. Only one entry per person is permitted. An entry may have two or more co-authors.
9. Submission of an entry implies permission for publication in *RF Design*, and distribution of software submissions by the RF Design Software Service.
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11. Entries must be postmarked no later than July 29, 1994 and received no later than August 8, 1994.
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## Transmission Line Fundamentals

By Andy Kellett  
Technical Editor

Transmission lines are among the blackest of RF "black magic". At first glance, a transmission line appears to be nothing more than an RF garden hose, piping RF energy from one place to another. However, transmission lines exhibit a complex behavior which enables them to be used as impedance transformers and even filters. The purpose of this article is to quickly arrive at a few basic equations from which we can derive the quantities and formulas most often used in transmission line problems.

Figure 1 demonstrates the notation used in this article and most books. The distance between the end of the cable and some point along the cable is called  $x$ . The distance from the load end of the cable to the same point along the cable is called  $d$ . The total length of the cable is  $s$  so that  $x+d=s$ . The impedance the cable sees looking toward the signal generator is  $Z_g$ , and the impedance the cable sees looking into the load is  $Z_l$ . Quantities having to do with the end of the line connected to the signal source use the subscript  $s$  (for sending or source end), for instance, the potential at the sending end is called  $E_s$ . Quantities relating to the end of the line connected to load are denoted with the subscript  $r$  (for receiving). The potential on the line at the load-end of the line is called  $E_r$ .

The electrical model for a section of transmission line used in this article is

illustrated in Figure 2. This model assumes that the line we are working with propagates the EM wave along the axis of the line, with the magnetic and electric fields perpendicular to the axis. This is called the TEM mode. For the cables used in RF work, it is hard for anything but the TEM mode to be transmitted down the line because the only dimension in RF cables that approaches the wavelength of RF signals is the cable's length.

The model, as drawn here represents an unbalanced line, but the same quantities and results are obtained when the impedance and current are shared by both conductors.

The quantity, " $z$ ", in Figure 2 is the impedance per unit length of line, and " $y$ " is the unit admittance. Both of these quantities are complex, being made up of real and imaginary parts as shown in equations 1 and 2.

$$z = r + j\omega l \quad (1)$$

$$y = g + j\omega c \quad (2)$$

The real part of the impedance is the resistance of the line, the imaginary part is the familiar expression for inductive reactance. The admittance is made up of a conductance (the real part) and the inverse of the capacitive reactance (the imaginary part).

At the source end of the line there exists a potential  $E$  between the conductors and a current  $I$  flowing through a point at that end. On the other end there

will be a potential  $E+\Delta E$  and a current  $I+\Delta I$ . If a transmission line were really like a garden hose,  $\Delta E$  and  $\Delta I$  would be zero, but it turns out they are not. The impedance per unit length,  $z$ , is responsible for dropping the potential, while the admittance is responsible for shunting some of the current. The amounts of these drops are given in equations 3 and 4.

$$\Delta E = -I(z\Delta x) \quad (3)$$

$$\Delta I = -(E + \Delta E)(y\Delta x) \quad (4)$$

Dividing both equation 3 and 4 by  $\Delta x$  and taking the limit as  $\Delta x$  goes to zero we get the derivatives describing the changes in voltage and current for an infinitesimal section of cable:

$$\lim_{\Delta x \rightarrow 0} \left[ \frac{\Delta E}{\Delta x} = -Iz \right] \rightarrow \frac{dE}{dx} = -Iz \quad (5)$$

$$\lim_{\Delta x \rightarrow 0} \left[ \frac{\Delta I}{\Delta x} = -(E + \Delta E)y \right] \rightarrow \frac{dI}{dx} = -Ey \quad (6)$$

We can eliminate  $I$  from equation 5 and  $E$  from equation 6 by differentiating again with respect to  $x$  to get:

$$\frac{d^2 E}{dx^2} = -\frac{dI}{dx} z \quad (7)$$

$$\frac{d^2 I}{dx^2} = -\frac{dE}{dx} y \quad (8)$$

and substituting the expressions for  $dI/dx$  and  $dE/dx$  from equation 5 and 6 into the right side of equations 7 and 8. The result is a set of differential equations:

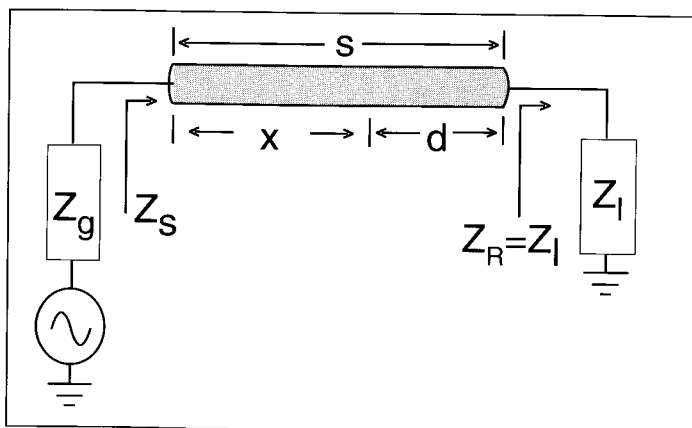


Figure 1. Diagram showing notations used to identify positions along cable of length  $s$ .

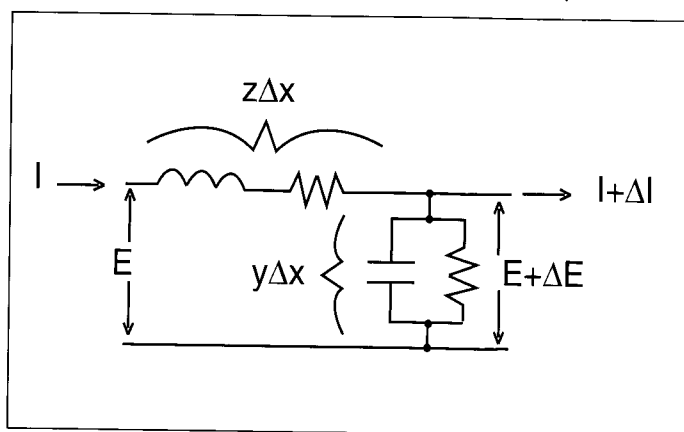
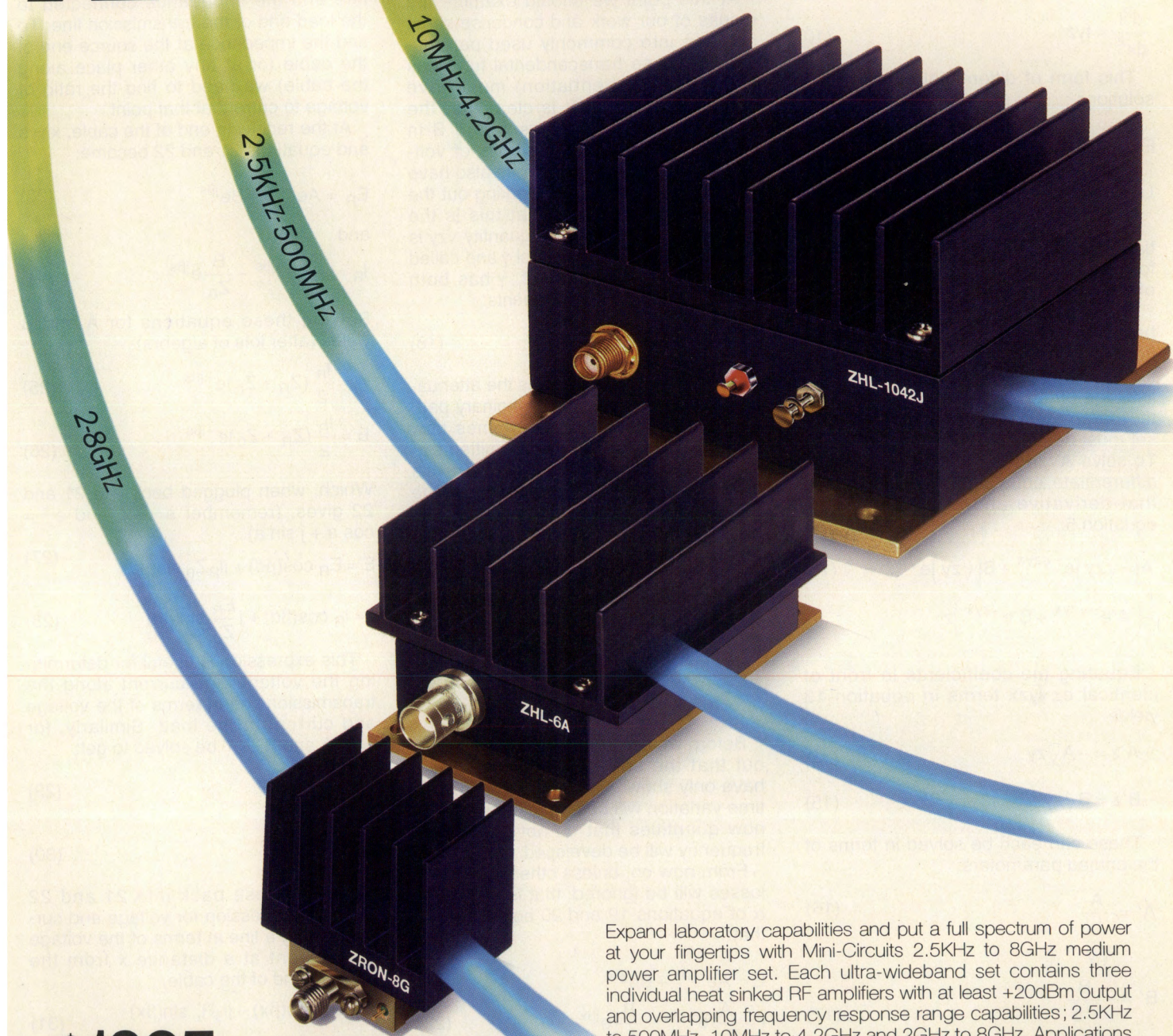


Figure 2. Electrical model of a section of lossy transmission line transmitting a signal in the TEM mode.



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$$\frac{d^2 E}{dx^2} = Eyz \quad (9)$$

$$\frac{d^2 I}{dx^2} = Iyz \quad (10)$$

This form of differential equation has solutions :

$$E = Ae^{-\sqrt{zy}x} + Be^{\sqrt{zy}x} \quad (11)$$

$$I = A'e^{-\sqrt{zy}x} + B'e^{\sqrt{zy}x} \quad (12)$$

Plug these solutions into the differential equations to verify them. So much for the easy part, the calculus, the rest of the work is messy algebra.

The parameters A, B, A' and B' are to be determined by the boundary conditions imposed on the particular transmission line situation, but ohm's law links the parameters of equation 11 with the parameters of equation 12, so one set must be solved in terms of the other. To solve A' and B' in terms of A and B, differentiate equation 11 and substitute that derivative and equation 12 into equation 5:

$$A(-\sqrt{zy})e^{-\sqrt{zy}x} + B(\sqrt{zy})e^{\sqrt{zy}x} = -\left(A'e^{-\sqrt{zy}x} + B'e^{\sqrt{zy}x}\right)z \quad (13)$$

Equating the coefficients in front of identical  $e^{\pm\sqrt{zy}x}$  terms in equation 13 yields:

$$-A'z = -A\sqrt{zy} \quad (14)$$

$$-B'z = B\sqrt{zy} \quad (15)$$

These can each be solved in terms of the primed parameters:

$$A' = \frac{A}{\sqrt{\frac{z}{y}}} \quad (16)$$

$$B' = -\frac{B}{\sqrt{\frac{z}{y}}} \quad (17)$$

The boundary conditions encountered by the line determine what form the constants A and B take. Even non-uniform lines can be described with these equations if they are piece-wise uniform, i.e. made up of segments of uniform line. Equations 11 and 12 are the starting point from which we derive all the really useful tools in transmission line analysis such as reflection coefficient and VSWR.

## Picking Familiar Parameters out of the Equations

At this point we should examine the results of our work and condense some notation into commonly used parameters. Because transcendental functions (such as exponentiation) must have unitless solutions, it is clear that the units for the parameters A and B in equation 11 should have units of voltage. Transcendental functions also have unitless arguments, and working out the units of  $\sqrt{zy}x$  shows that this is the case for our solution. The quantity  $\sqrt{zy}$  is commonly given the symbol  $\gamma$  and called the propagation constant.  $\gamma$  has both real and imaginary components:

$$\gamma = \sqrt{zy} = \alpha + j\beta \quad (18)$$

The real part, ( $\alpha$ ), describes the attenuation in the line, while the imaginary part, ( $\beta$ ), is the factor that determines how quickly the phase of the transmitted signal changes with the distance it travels down a line. The quantity  $\sqrt{z/y}$  is the characteristic impedance of the line, commonly called  $Z_0$ . Examining the units in  $\sqrt{z/y}$  should make this seem plausible. Equations 11 and 12 using the new quantities are written as

$$E = Ae^{-(\alpha+j\beta)x} + Be^{(\alpha+j\beta)x} \quad (19)$$

$$I = \frac{A}{Z_0} e^{-(\alpha+j\beta)x} - \frac{B}{Z_0} e^{(\alpha+j\beta)x} \quad (20)$$

Before we continue, it must be pointed out that the equations derived so far have only shown variation in space. The time variation will be added later, but for now quantities that do not depend on frequency will be developed.

From now on, unless otherwise noted, losses will be ignored, that is we will let  $\alpha$  of equations 19 and 20 equal zero so that:

$$E = Ae^{-j\beta x} + Be^{j\beta x} \quad (21)$$

$$I = \frac{A}{Z_0} e^{-j\beta x} - \frac{B}{Z_0} e^{j\beta x} \quad (22)$$

Removing losses makes the math much easier, and is an accurate assumption for short lengths of line. When this is done,  $\gamma$  becomes  $j\omega\sqrt{LC} = j\beta$  and  $Z_0$  becomes  $\sqrt{L/C}$ . Calculations that require loss terms are most commonly done using a computer or the Smith Chart.

## Impedances

A signal source looking into the end of a transmission line sees some imped-

ance. The impedance seen depends on both the characteristic impedance of the line and the impedance connected to the load end of the transmission line. To find the impedance at the source end of the cable (or at any other place along the cable) we need to find the ratio of voltage to current at that point.

At the receiving end of the cable,  $x = s$  and equations 21 and 22 become:

$$E_R = Ae^{-j\beta s} + Be^{j\beta s} \quad (23)$$

and

$$I_R = \frac{A}{Z_0} e^{-j\beta s} - \frac{B}{Z_0} e^{j\beta s} \quad (24)$$

Solving these equations for A and B yields (after lots of algebra)

$$A = \frac{I_R}{2} (Z_R + Z_0) e^{+j\beta s} \quad (25)$$

$$B = \frac{I_R}{2} (Z_R - Z_0) e^{-j\beta s} \quad (26)$$

Which, when plugged back into 21 and 22 gives, (remember  $s=d+x$ , and  $e^{ja} = \cos a + j \sin a$ ).

$$E = E_R \cos(\beta d) + jI_R Z_0 \sin(\beta d) \quad (27)$$

$$I = I_R \cos(\beta d) + j \frac{E_R}{Z_0} \sin(\beta d) \quad (28)$$

This expression is useful for determining the voltage and current along the transmission line in terms of the voltage and current at the load. Similarly, for  $x=0$ , A and B can be solved to get:

$$A = \frac{E_s + I_s Z_0}{2} \quad (29)$$

$$B = \frac{E_s - I_s Z_0}{2} \quad (30)$$

Plugging these back into 21 and 22 gives an expression for voltage and currents along a line in terms of the voltage and current at a distance x from the sending end of the cable.

$$E = E_s \cos(\beta x) - jI_s R_c \sin(\beta x) \quad (31)$$

$$I = I_s \cos(\beta x) - j \frac{E_s}{Z_0} \sin(\beta x) \quad (32)$$

Equations 27 and 28 can be used to find the impedance seen at the sending end of the cable. Using the expressions for I and E at the sending end ( $d=s$ ) in the relation  $Z = E/I$ , we get

$$Z_s = \frac{E_s}{I_s} = \frac{E_R \cos(\beta s) + jI_R Z_0 \sin(\beta s)}{I_R \cos(\beta s) + j \frac{E_R}{Z_0} \sin(\beta s)} \quad (33)$$



dividing through by  $I_R \cos \beta s$  yields the familiar equation

$$Z_s = Z_0 \frac{Z_R + jZ_0 \tan(\beta s)}{Z_0 + jZ_R \tan(\beta s)} \quad (34)$$

This relation is the basis for the quarter wave transformer. A length of line one-quarter a wavelength long has an input impedance given by:

$$Z_s = \frac{Z_0^2}{Z_R} \quad (35)$$

By selecting a quarter-wavelength line with the appropriate characteristic impedance, the input impedance of the transmission line/load combination can be tailored to any desired impedance.

Another special case of equation 34 puts  $Z_R = Z_0$  yielding:

$$Z_s = Z_0 \frac{1}{1} = Z_0 \quad (36)$$

Which means that any line and load whose impedances match always present that impedance at the sending end of the line, no matter the length.

### Time Variation

A true wave solution contains variations in both space and time. Substituting the expressions for A and B in equations 25 and 26 into equation 21 gives:

$$E = \frac{I_R}{2} (Z_R + Z_0) e^{j\beta(s-x)} + \quad (37)$$

$$\frac{I_R}{2} (Z_R - Z_0) e^{-j\beta(s-x)}$$

Because  $s-x=d$ , this can be rewritten as:

$$E = \frac{I_R}{2} (Z_R + Z_0) e^{j\beta d} + \quad (38)$$

$$\frac{I_R}{2} (Z_R - Z_0) e^{-j\beta d}$$

(remember  $d$  is a measure of distance from the *load end* of the line to some particular point.)

Assuming  $I_R$  has the form:

$$I_R = \hat{I}_R e^{j\omega t} \quad (39)$$

(where  $\hat{I}_R$  is a complex constant). The result is:

$$E = \frac{\hat{I}_R}{2} (Z_R + Z_0) e^{j(\omega t + \beta d)} + \frac{\hat{I}_R}{2} (Z_R - Z_0) e^{j(\omega t - \beta d)} \quad (40)$$

This is a solution to the wave equation. It may seem dubious that we arrive at a solution to the wave equation (a partial differential equation) without starting with the wave equation, (equations 9 and 10 are ordinary differential equations in  $x$ ). However, had we started the derivation in a way that led to the wave equation (as is done in many texts [1,2]) and used separation of variables to solve it, equations 9 and 10 pop out as the ordinary differential equations in  $x$  that must be solved. The ordinary differential equations in time that arise have solutions of the form given in equation 38.

The two terms in equation 40 represent two waves travelling in opposite directions. You can verify this by seeing how  $d$  has to change to keep the argument of each exponential term constant, (the argument of the exponential in each term is the phase of the wave it represents).

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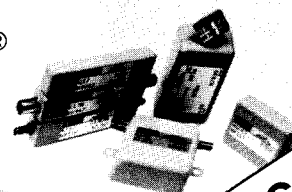
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For example, take the second term on the right hand side of equation 40. The real part of that term is (ignoring the factor in front of the exponential):

$$\cos(\omega t - \beta d) \quad (41)$$

When the argument,  $\omega t - \beta d$ , equals zero, equation 41 is at its maximum. Using the maximum as a "tracking point" we see what happens as  $t$  gets bigger (as time does when it moves forward). In order to maintain the argument's zero value as  $t$  gets larger,  $d$  must also get bigger. This means the maximum occurs at a point where  $d$  is larger, i.e. further from the load end of the line. Of course any other reference point besides the maximum could have been chosen and the results would be the same.

Applying this type of verification to the other term in equation 40, it can be seen that there are two waves, each traveling in opposite directions. The wave traveling away from the source end is called the incident wave, and the wave traveling away from the load end is called the reflected wave.

### Phase velocity

As in the last section, we will follow a particular point on a wave to determine phase velocity, which is the speed at which the wave travels along the transmission line. Because we are following a fixed point on the waveform (again use the crest for a mental example), we set the phase equal to a constant:

$$(\omega t - \beta d) = \text{const.} \quad (42)$$

Taking the derivative of this with respect to time we get:

$$\frac{d}{dt}(\omega t - \beta d) = \frac{d}{dt} \text{const.} = 0 \quad (43)$$

$$\omega + \beta \frac{d(d)}{dt} = 0$$

The quantity  $d(d)/dt$  is the velocity we are looking for, so solving for that quantity we get:

$$\frac{d(d)}{dt} = \frac{\omega}{\beta} \quad (44)$$

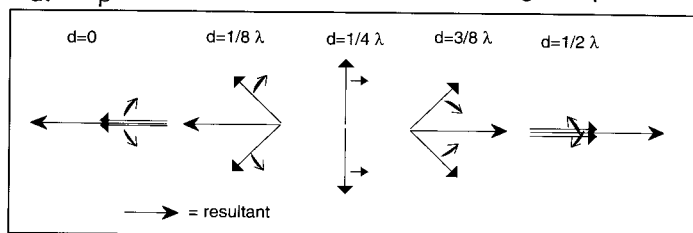


Figure 3. Phase relation between equal incident and reflected current waves as a function of distance from short-circuited end.

This is the phase velocity. Notice if we select the phase of the other term to set constant, the absolute value of the phase velocity is the same, but the sign is reversed, again demonstrating that the two terms represent waves travelling in opposite directions.

### Reflection Coefficients

While the existence of two waves has been established, the amplitudes of the waves relative to each other has not been discussed. The reflection coefficient is the ratio of the reflected and incident wave amplitudes *at the load end* of the line:

$$\Gamma = \frac{E_-}{E_+} \quad (45)$$

Where  $E_+$  represents the incident wave voltage and  $E_-$  represents the reflected wave voltage. Generally,  $\Gamma$  is a complex number. The magnitude of each wave is given by the coefficients in front of each exponential term in equation 38. Plugging these coefficients into equation 45 results in:

$$\Gamma = \frac{\frac{I_R}{2}(Z_R - Z_0)}{\frac{I_R}{2}(Z_R + Z_0)} = \frac{(Z_R - Z_0)}{(Z_R + Z_0)} \quad (46)$$

Some special cases help demonstrate  $\Gamma$ . If  $Z_R = Z_0$ ,  $\Gamma = 0$ , which means there are no reflections in a line terminated in its characteristic impedance. For  $Z_R = 0$ , (a short circuit), and  $Z_R = \infty$ , (an open circuit), the magnitude of  $\Gamma = 1$ , indicating the incident and reflected waves are equal. Figure 3 demonstrates the relationship between the two current waves in a short-circuited line.

### Standing Waves, VSWR

Except for lines terminated in open and short circuits, the reflected and incident waves do not have equal amplitudes, meaning there will never be total cancellation of the two waves (see Figure 4). The ratio of the maximum voltage amplitude along a line to the mini-

mum voltage amplitude along the line is called the voltage standing wave ratio (VSWR), (because voltage and current are proportional, the ratio of currents can also be used):

$$\rho = \frac{|E_{\max}|}{|E_{\min}|} = \frac{|I_{\max}|}{|I_{\min}|} \quad (47)$$

The maximum amplitude occurs when the maximum of the reflected wave coincides with the maximum of the incident wave, and the minimum amplitude occurs when the maxima are 180 degrees apart, so that:

$$\text{VSWR} = \frac{|E_+| + |E_-|}{|E_+| - |E_-|} \quad (48)$$

Notice that if both the numerator and denominator on right hand side of equation 48 is divided by  $|E_+|$ , the result is:

$$\text{VSWR} = \frac{1 + \left| \frac{E_-}{E_+} \right|}{1 - \left| \frac{E_-}{E_+} \right|} = \frac{1 + |\Gamma|}{1 - |\Gamma|} \quad (49)$$

### Conclusion

Of course nobody goes about solving a transmission line problem by starting with differential equations. This tutorial only touched upon the really useful tools in transmission line analysis, such as reflection coefficient and VSWR. However, it is hoped that this tutorial has helped to make use of those tools more intuitive and less mechanical by illustrating the principles that lead up to them. RF

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2. Sinnema, William, *Electronic Transmission Technology*, 2nd ed., Prentice Hall, 1988.

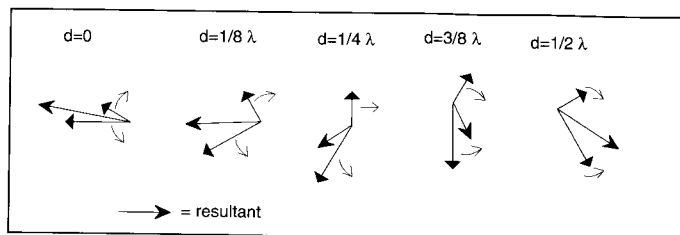


Figure 4. Incident and reflected waves at different distances from an arbitrary complex load. Notice phase of resultant changes gradually.



# RF product forum

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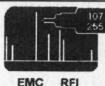
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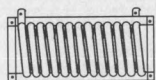
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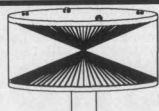
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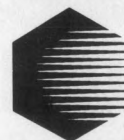
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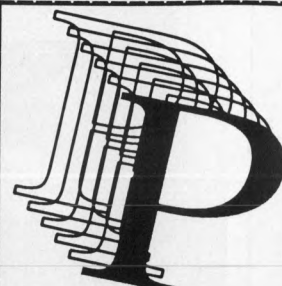
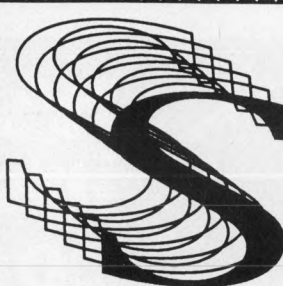
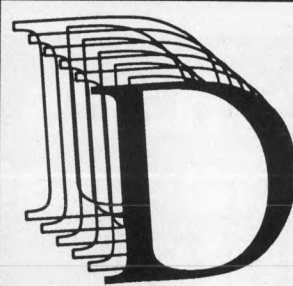
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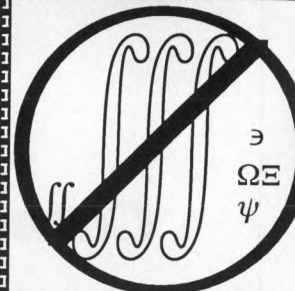
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12. RF Des. - EMC Prod. - 8 Loc's - Sal. 35-60K

### SALES AND MARKETING

1. Reg. Sales Mgr. - Components - 70-90K + Comm.
2. Nat. Sales Mgr. - EMC Prod's Sal. Open + Comm.
3. South Amer. Sales Mgr. - Must be bilingual comp. Sal. Neg. + Comm.
4. Acct. Rep. - Sys. Block & Sub Assemblies - To 85K + Comm.

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## JIM YOUNG & ASSOCIATES

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INFO/CARD 91

## Electronics Instructor:

B.S.E.E. or B.E.T. or related degree. 2 years experience in directly related field. Good communication and presentation skills required.

**Starting Date: August 1, 1994**

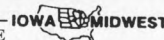
Send resume and transcripts to:

Dr. Veldon L. Law  
John Wood Community College  
150 South 48th Street  
Quincy, Illinois 62301  
Phone: 217/224-6500, Ext. 164  
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WORK WITH AN RF ENGINEER IN YOUR MOVE TO THE MIDWEST!

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40 YEARS OF RF ENGINEERING, MANAGEMENT, AND PLACEMENT OF ENGINEERS

INFO/CARD 92



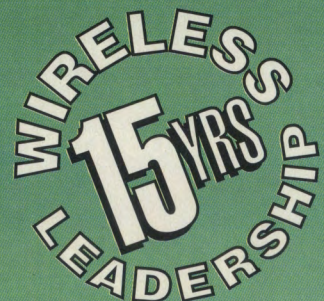
## July Special Section: Product Data Sheets

Featuring Products from:

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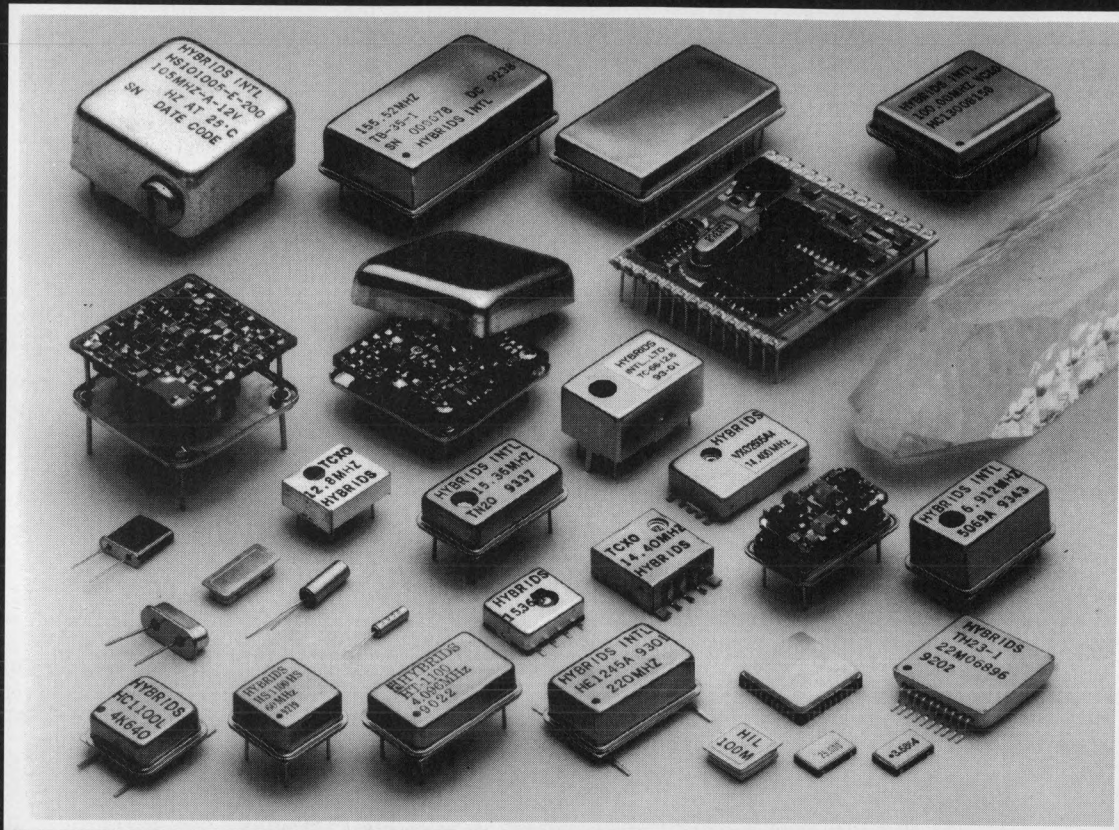
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# CRYSTAL OSCILLATORS AND THICK FILM HYBRID CUSTOMIZATION



- FREQUENCY RANGE 1Hz TO 1GHz.
- ECL, XO, VCXO, ECL, PECL, MECL, ECL IN PS, 10K, 100K DIFFERENTIAL OUTS FOR SONET, LAN, ATM, DDS, AND OTHER DATA RECOVERY APPLICATIONS.
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# RF POWER AMPLIFIERS

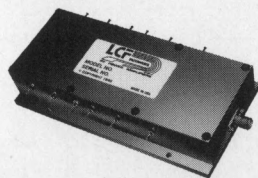
(Standard or Custom)

**1 MHz - 2 GHz • 1 W - 1 kW**

Part No.	Output	Gain	MODULES		Part No.	Output	Gain	MODULES	
	Power		DC Supply	PKG		Power		DC Supply	PKG
2 MHz - 30 MHz					400-225-50-10				
30-2-5-35	5W	35dB	24/28V	E3	400-225-50-18	50W	18dB	24/28V	A2
30-2-50-35	50W	35dB	24/28V	E3	400-225-50-35	50W	35dB	24/28V	A3
30-2-100-35	100W	35dB	28V	E3	400-225-100-10				
30-2-200-35	200W	35dB	28V	E3	400-225-100-18	100W	18dB	28V	A2
30 MHz - 100 MHz					400-225-100-35	100W	35dB	28V	A3
100-30-5-35	5W	35dB	24/28V	E3	225 MHz - 600 MHz				
100-30-100-35	100W	35dB	28V	E3	600-225-30-10	30W	10dB	24/28V	A1
50 MHz - 150 MHz					600-225-30-18	30W	18dB	24/28V	A2
150-50-2-40	2W	40dB	12/15V	C3	600-225-30-30	30W	30dB	24/28V	A3, B3
150-50-100-35	100W	35dB	28V	A3	400 MHz - 600 MHz				
150 MHz - 200 MHz					600-400-30-10	30W	10dB	24/28V	A1
200-150-100-10	100W	10dB	28V	A1	600-400-30-18	30W	18dB	24/28V	A2
200-150-100-18	100W	18dB	28V	A2	600-400-30-30	30W	30dB	24/28V	A3, B3
200-150-100-35	100W	35dB	28V	A3	925 MHz				
200-150-200-35	200W	35dB	28V	A3	925-1-8	1W	8dB	12/15V	D1
50 MHz - 250 MHz					100 MHz - 500 MHz				
250-50-200-10	160W	10dB	28V	A1	500-100-5-30	5W	30dB	24/28V	A3
250-50-200-18	160W	18dB	28V	A2	500-100-10-30	10W	30dB	24/28V	A3
250-50-200-35	160W	35dB	28V	A3	500-100-100-30	100W	30dB	28V	A3
50 MHz - 400 MHz					500 MHz - 1000 MHz				
400-50-1-30	1W	30dB	12/15V	C3	1000-500-10-8	10W	8dB	24/28V	A1
225 MHz - 400 MHz					1000-500-10-16	10W	16dB	24/28V	A2
400-225-1-35	1W	35dB	12/15V	C3	1000-500-10-30	10W	30dB	24/28V	A3
400-225-10-35	10W	35dB	24/28V	A3	10 MHz - 1200 MHz				
400-225-30-10	30W	10dB	24/28V	A1	1200-10-10-30	10W	30dB	24/28V	A3
400-225-30-18	30W	18dB	24/28V	A2					
400-225-30-35	30W	35dB	24/28V	A3, B3					

\* Test Fixture (Option "B") includes heat sink, fan, thermal shutdown, and electrical fuse protection  
 \*\* Rack-mount amplifiers: 120 Vac - 60 Hz / 240 Vac - 50 Hz

## MODULES - SMALL SIZE • HIGH EFFICIENCY

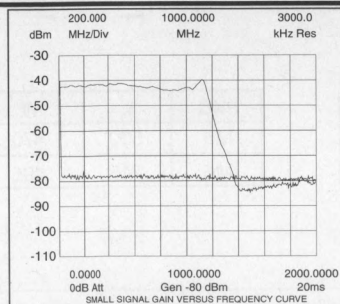


- 200 watts
- 150 - 200 MHz
- 55% overall efficiency
- 35 dB gain
- 4.84" x 2.0" x 1.0"



- 100 watts
- 225 - 400 MHz
- 45% overall efficiency
- 10 dB gain
- 3.0" x 2.0" x 1.0"

## SYSTEMS - LOW COST • RUGGED PROTECTION



- 10 MHz - 1200 MHz
- 10 watts CW
- 40 dB
- 19" x 7" x 18"
- 35 lbs maximum



# TEMEX STANDARD 10.7MHZ MONOLITHIC CRYSTAL FILTERS

## CHANNEL SPACING FOR 12.5 • 20 • 25 AND 50KHZ

ELECTRONICS, INC.

CRYSTALS • CRYSTAL FILTERS • L/C FILTERS

NO. POLES	TEMEX P/N	PASSBAND		STOPBAND				LOSS	RIPPLE	ULT. REJ	TERM. (Rp//Cp) OHM/PF
		dB	±KHz	dB	±KHz	dB	±KHz	dB	dB-MAX	dB-MIN	
2	TE5000	3	3.75	20	18.0	--	----	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	--	----	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	--	----	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.5	20	30.0	--	----	1	1.0	50	2700//0
4	TE5050	3	6.5	30	15.0	--	----	2	2.0	75	3100//0
6	TE5060	6	6.5	60	19.5	--	----	3	2.0	90	3100//0
8	TE5070	6	6.5	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.5	20	35.0	--	----	1	1.0	50	3000//0
4	TE5090	3	7.5	30	17.5	--	----	2	2.0	75	3300//0
6	TE5100	6	7.5	60	22.5	--	----	3	2.0	90	3300//0
8	TE5110	6	7.5	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	--	----	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	--	----	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	--	----	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

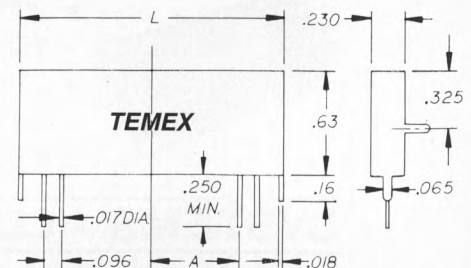
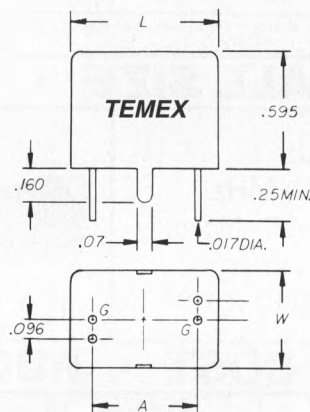
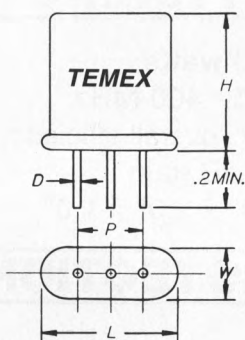
### NOTES:

1. Maximum inband ripple over temp. range -20°C to 70°C.
2. Parallel termination capacity is adjusted for optimum filter response. Nominal parallel capacity, Cp: ±3pf. Impedance, Rp specified above.
3. A tandem set is a combination of matched 2 pole filter units making up multipole filters [example: 4 pole response; (2) 2 pole units-matches.]
4. These models available in other packages not shown below.

**TO ORDER:** TEMEX P/N and PACKAGE TYPE: Example: TE5100M5.

# POLES	PACKAGE SELECTION
2	M3 or 4
4	M3 or 4(x2)M5, 6, 7, 8 or 9
6	M3 or 4(x3)M5, 6, 8 or 9
8	M3 or 4(x4)M6 or 9

(x2)=2 cases (x3)=3 cases (x4)=4 cases



PKG	L	W	H	P	D
M3	.435	.185	.45/.53	.192	.017
M4	.750	.350	.750	.486	.030

HC-18/3:M3  
HC-49/3:M3

HC-6/3:M4

PKG	L	W	A
M5	.590	.470	.354
M6	.745	.496	.528

PKG	L	A
M7	.89	.216
M8	1.32	.435
M9	1.75	.645

All specifications subject to change without notice.

Consult **TEMEX** for your custom crystal and filter requirements.

3030 W. Deer Valley Rd. • Phoenix, AZ 85027 • 602-780-1995 • (FAX) 602-780-2431

INFO/CARD 50



# TEMEX STANDARD 21.4 MHZ MONOLITHIC CRYSTAL FILTERS

## CHANNEL SPACING FOR 12.5 • 20 • 25 AND 50KHZ

ELECTRONICS, INC.

CRYSTALS • CRYSTAL FILTERS • L/C FILTERS

NO. POLES	TEMEX P/N	PASSBAND		STOPBAND				LOSS	RIPPLE	ULT. REJ	TERM. (Rp//Cp) OHM/PF
		dB	±KHz	dB	±KHz	dB	±KHz	dB	dB-MAX	dB-MIN	
2	TE5180	3	3.75	15	12.5	--	----	2	1.0	50	850// + 6
4	TE5190	3	3.75	30	12.5	--	----	3	2.0	70	850// + 5
6	TE5200	6	3.75	60	12.5	--	----	4	2.0	90	850// + 5
8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850// + 5
2	TE5220	3	6.5	15	20.0	--	----	2	1.0	50	1300// + 2
4	TE5230	3	6.5	30	22.5	--	----	3	2.0	70	1400//0
6	TE5240	6	6.5	60	22.5	--	----	4	2.0	90	1400//0
8	TE5250	6	6.5	60	17.5	80	22.5	4	2.0	100	1400//0
2	TE5260	3	7.5	15	25.0	--	----	2	1.0	50	1500//0
4	TE5270	3	7.5	30	25.0	--	----	3	2.0	70	1600//0
6	TE5280	6	7.5	60	25.0	--	----	4	2.0	90	1600//0
8	TE5290	6	7.5	60	20.0	80	25.0	4	2.0	100	1600//0
2	TE5300	3	15.0	15	50.0	--	----	2	1.0	45	3000//0
4	TE5310	3	15.0	30	45.0	--	----	3	2.0	60	3000// -1
6	TE5320	6	15.0	60	45.0	-	----	3	2.0	90	3000// -1
8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000// -1

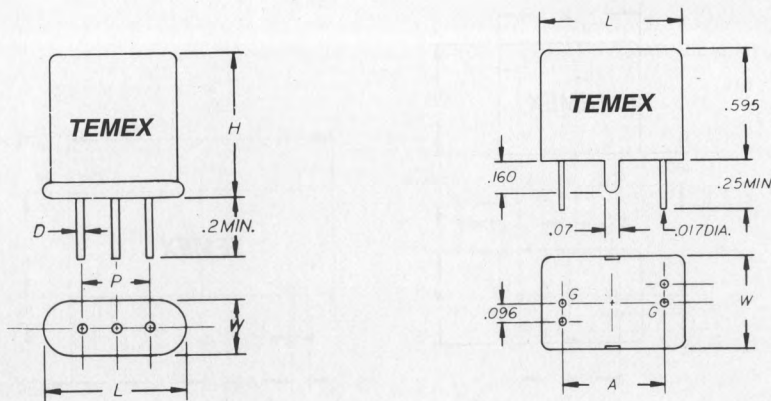
### NOTES:

- Maximum inband ripple over temp. range -20°C to 70°C.
- Parallel termination capacity is adjusted for optimum filter response. Nominal parallel capacity, Cp: ± 3pf. Impedance, Rp specified above.
- A tandem set is a combination of matched 2 pole filter units making up multipole filters [example: 4 pole response; (2) 2 pole units-matched.]
- These models available in other packages not shown below.

TO ORDER: TEMEX P/N and PACKAGE TYPE: Example TE5280M3

# POLES	PACKAGE SELECTION
2	M1, 2 or 3
4	M1, 2 or 3(x2)M5, 6, 7, 8 or 9
6	M1, 2 or 3(x3)M5, 6, 8 or 9
8	M1, 2 or 3(x4)M6 or 9

(x2) = 2 cases (x3) = 3 cases (x4) = 4 cases



PKG	L	W	H	P	D
M1	.300	.100	.310	.114	.017
M2	.310	.125	.320/.345	.148	.017
M3	.435	.185	.45/.53	.192	.017

HC-44/3:M1 HC-45/3:M2 HC-18/3:M3  
HC49/3:M3

PKG	L	W	A
M5	.590	.470	.354
M6	.745	.496	.528

PKG	L	A
M7	.89	.216
M8	1.32	.435
M9	1.75	.645

All specifications subject to change without notice.

Consult TEMEX for your custom crystal and filter requirements.

3030 W. Deer Valley Rd. • Phoenix, AZ 85027 • 602-780-1995 • (FAX) 602-780-2431

INFO/CARD 50



# TEMEX STANDARD 45.0 MHZ MONOLITHIC CRYSTAL FILTERS

ELECTRONICS, INC.

CRYSTALS • CRYSTAL FILTERS • L/C FILTERS

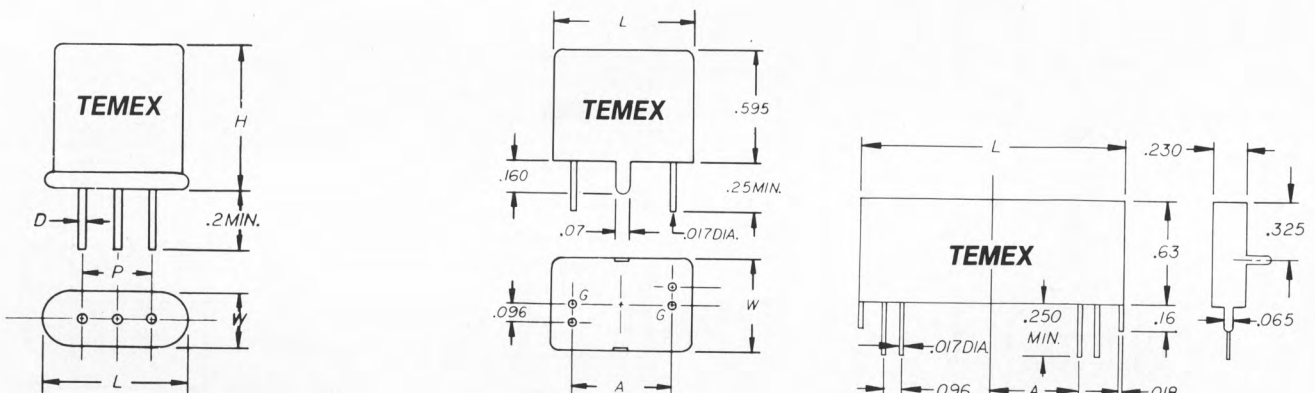
45 MHZ MONOLITHIC CRYSTAL FILTERS										
NUMBER OF POLES	MODE	TEMEX P/N	PASSBAND		STOPBAND		LOSS	RIPPLE	ULT. REJ.	TERMINATION
			dB	±KHz	dB	±KHz	dB	dB-MAX	dB-MIN	
2	30T	TE7420	3	7.5	18	28	2	1	40	3000// -1.0
4	30T	TE7430	3	7.5	40	30	3	1	70	3000// -1.0
2	30T	TE7440	3	15	15	47	2	1	40	8000// -1.5
4	30T	TE7450	3	15	30	50	3	1	70	8000// -1.5
2	FUND	TE7730	3	15	15	50	2	1	40	1100// +1.5
4	FUND	TE7740	3	15	40	60	3	1	70	800// +1.0

## NOTES:

- Maximum inband ripple over temp. range -20° C to 70° C.
- Parallel termination capacity is adjusted for optimum filter response. Nominal parallel capacity, Cp: ±3pf. Impedance, Rp specified above.
- A tandem set is a combination of matched 2 pole filter units making up multipole filters [example: 4 pole response; (2) 2 pole units-matched.]
- These models available in other packages not shown below.
- Standard package = M2.
- 50 Ohms Z I/O available in our M5 or larger packages.
- 30T = Third overtone crystals.  
Fund = Fundamental crystals.
- Other models available, consult factory.

# POLES	PACKAGE SELECTION
2	M1, 2 or 3
4	M1, 2 or 3(x2)M5, 6, 7, 8 or 9

(x2) = 2 cases



PKG	L	W	H	P	D
M1	.300	.100	.310	.114	.017
M2	.310	.125	.32/.345	.148	.017
M3	.435	.185	.45/.53	.192	.017

HC-44/3:M1

HC-45/3:M2

HC-18/3:M3  
HC-49/3:M3

PKG	L	W	A
M5	.590	.470	.354
M6	.745	.496	.528

PKG	L	A
M7	.89	.216
M8	1.32	.435
M9	1.75	.645

All specifications subject to change without notice.

Consult TEMEX for your custom crystal and filter requirements.

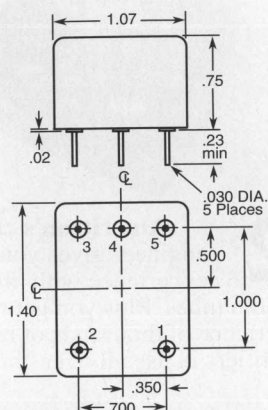
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INFO/CARD 50



**OSCILLATEK**

A DOVER TECHNOLOGIES COMPANY

620 N. Lindenwood Drive • Olathe, Kansas 66062  
FAX (913) 829-3505 • Phone (913) 829-1777**Solder Seal OCXO Small Package****RD Package****Output Options:**

TD	= TTL	= DC to 50 MHz
HC	= HCMOS	= DC to 50 MHz
RC	= +7 dBm 50Ω	= DC to 50 MHz

**Stability Options**

A18	= $\pm 1 \times 10^{-8}$	0° to +50°C
B28	= $\pm 2 \times 10^{-8}$	0° to +70°C
D27	= $\pm 2 \times 10^{-7}$	-40° to +85°C
S38	= $\pm 5 \times 10^{-8}$	-40° to +85°C

**Screening Options:**

C = Standard Commercial  
S = Class "B" per MIL-O-55310

**Typical P/N:**

HC RD K1 A 4 D27 C 10.000 MHz  
Screening  
Stability  
Aging  
Supply  
Predetermined  
Package  
Output

**Supply Options:**

A	= +5.0 VDC	3 Watts
B	= +12.0 VDC	3 Watts
C	= +15.0 VDC	3 Watts
D	= +24.0 VDC	3 Watts

**Aging Options:**

2	= $1 \times 10^{-8}$ /day	$2 \times 10^{-6}$ /year
3	= $5 \times 10^{-9}$ /day	$1 \times 10^{-6}$ /year
4	= $3 \times 10^{-9}$ /day	$5 \times 10^{-7}$ /year
5	= $1 \times 10^{-9}$ /day	$2 \times 10^{-7}$ /year

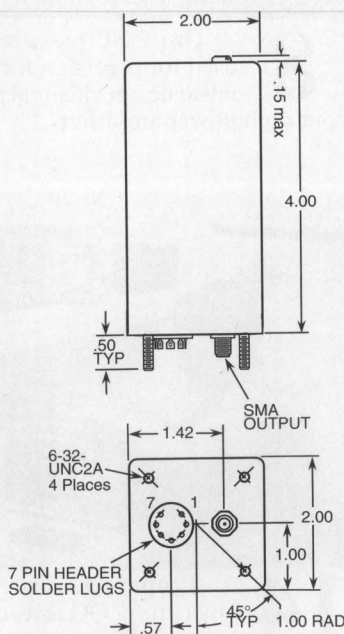
**Phase Noise (10 MHz Sinewave)**

At	10 Hz	= -100 dBc
	100 Hz	= -127 dBc
	1.0 KHz	= -148 dBc
	>10.0 KHz	= -155 dBc

**Pinouts:**

Pin 1 = Output  
Pin 2 = Gnd, Case  
Pin 3 = VCO Input\*  
Pin 4 = VCO Reference\*  
Pin 5 = Supply

\*When specified

**High Stability HF OCXO****V1 Package****Output Options:**

RC	= +7 dBm 50Ω
RE	= +13 dBm 50Ω

**Stability Options**

A50	= $\pm 5 \times 10^{-10}$	0° to +50°C
A30	= $\pm 3 \times 10^{-10}$	0° to +50°C
B70	= $\pm 7 \times 10^{-10}$	0° to +70°C
C29	= $\pm 2 \times 10^{-9}$	-20° to +70°C
C19	= $\pm 1 \times 10^{-9}$	-20° to +70°C
S29	= $\pm 2 \times 10^{-9}$	-40° to +70°C

**Screening Options:**

C = Standard Commercial  
S = Class "B" per MIL-O-55310

**Typical P/N:**

RC V1 M1 C 7 C29 C 10 MHz  
1 MHz 5 MHz 100 MHz  
Screening  
Stability  
Aging  
Supply  
Predetermined  
Package  
Output

**Supply Options:**

B	= +12.0 VDC	12 Watts
C	= +15.0 VDC	12 Watts
D	= +24.0 VDC	12 Watts

**Aging Options:**

6	= $5 \times 10^{-10}$ /day	$1 \times 10^{-7}$ /year
7	= $3 \times 10^{-10}$ /day	$5 \times 10^{-8}$ /year
8	= $1 \times 10^{-10}$ /day	$2 \times 10^{-8}$ /year
9	= $7 \times 10^{-11}$ /day	$1 \times 10^{-8}$ /year

**Phase Noise (5 MHz)**

At	10 Hz	= -120 dBc
	100 Hz	= -140 dBc
	1.0 KHz	= -155 dBc
	>10.0 KHz	= -157 dBc

**Pinouts:**

Pin 1 = Supply  
Pin 2 = Gnd, Case  
Pin 3 = No connect  
Pin 4 = Supply  
Pin 5 = VCO Reference\*  
Pin 6 = VCO Input\*  
Pin 7 = VCO Return\*

\*When specified

**CALL FOR OUR 100 PAGE CATALOG FOR MORE OPTIONS**



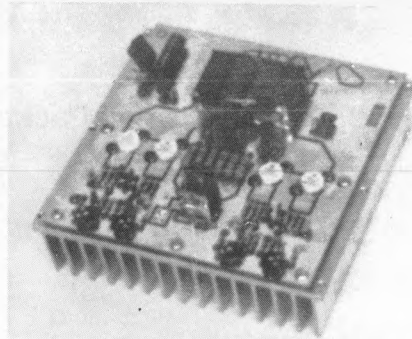
# High Power RF Amplifiers, Power Generators, Transmitters...

**Ameritron** can deliver high power amplifiers from stock, custom build your design or design a product to your specs from 1 to 250 MHz at power levels up to 5000 watts.

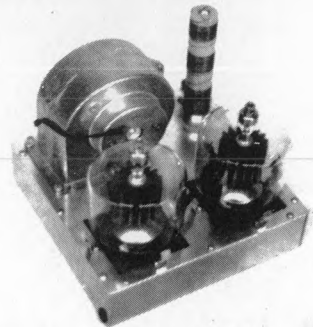
We manufacture from start to finish using modern CAD tools and CNC machinery.

Whether you need one or OEM quantities, you'll get the lowest prices and the fastest delivery in the industry.

Here's how...



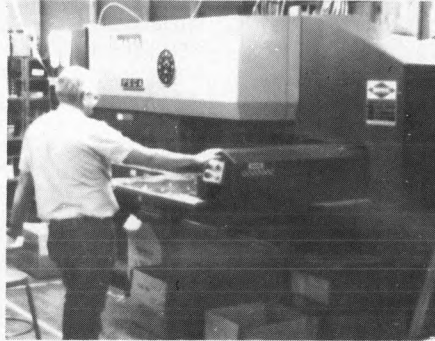
**1** Ameritron's 200 Watt FET Linear Amplifier Module is ready-to-use. Can be combined to produce several kilowatts. Modules available for 2-30 MHz; 30-120 MHz; 120-250 MHz.



**2** Ameritron's chief engineer gives you 20 years of experience with high power vacuum tubes. Plus you get the field experience of thousands of high powered amplifiers in use all over the world.



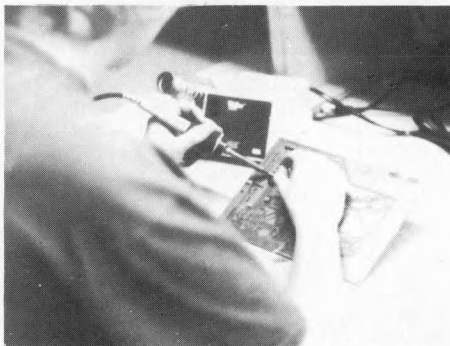
**6** CAD workstations are used for product mechanical design. Sheet metal layouts are downloaded to our Computer Numeric Controller (CNC) punch press for precision stamping.



**7** Our CNC punch press eliminates expensive tooling for prototyping your design. Changes can be made at a moment's notice *without* costly tooling charge.



**8** Our CNC press brake is used for precision forming of intricate mechanical pieces for your high power amplifiers.



**12** PC board assemblies are constructed in-house by our experienced assembly workers. Every PC board assembly is tested and quality checked by electronic techs so you'll get first-rate products every time.



**13** We have dedicated assembly lines made up of well trained workers to make sure you get quality products.



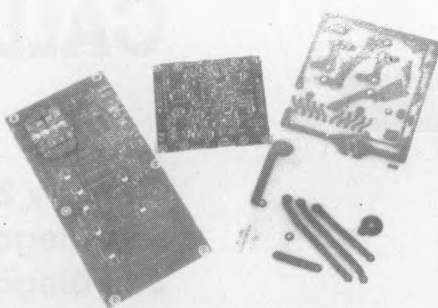
**14** Proper QC is stressed at Ameritron. QC tests during the assembly process permit better visual and electrical inspections than are possible after final assembly.

**Interested? Call Steven Pan, toll-free... 800-647-1801**

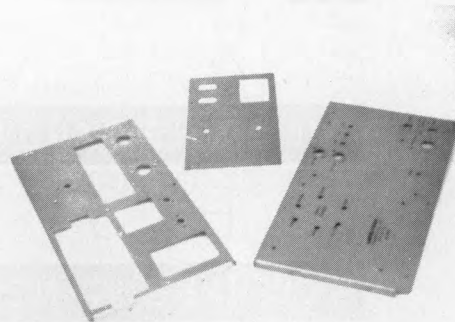




**3** PCB layouts are designed using CAD work stations for optimum component and trace placement. Our experienced engineers are familiar with the quirks of RF grounding for feedback-free operation.



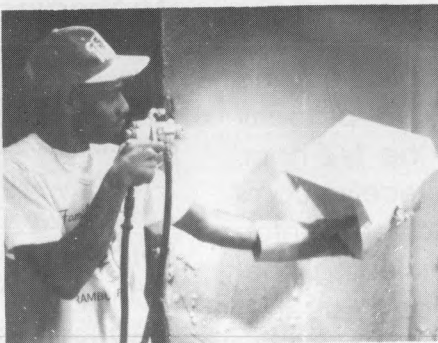
**4** Ameritron's in-house printed circuit board production facility can make prototype boards to rapidly confirm PCB layouts. In-house film and screen processing eliminates outside delays.



**5** Sheet metal is fitted and proofed in our prototype metal shop. Your changes in sheet metal design can be rapidly implemented.



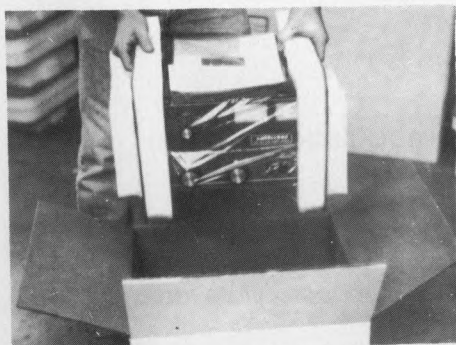
**9** Ameritron can chemically clean and etch sheet metal parts in house. Our careful metal preparation helps ensure a durable, attractive finish for your product.



**10** In Ameritron's paint department we can apply a tough *baked-on* paint to your product. We control the quality of your cabinets from raw material to finished product.



**11** Everything from artwork to silk-screening is done in house to give you *exactly* what you want. Our graphic arts department can create a cabinet design to your specifications.



**15** Every amplifier is protected by foam-in-place packing or re-usable die cut closed-cell foam inserts -- you'll never get a damaged amplifier because of rough shipping and handling.



**16** Your instruction manuals are printed in Ameritron's printing department on modern off-set printing presses -- you'll never be at the mercy of your printer's schedule again.



**17** You get excellent technical support from our group of friendly, well trained customer service technicians.

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Attn: Gary Breed

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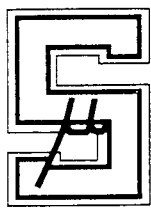
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*Please Note!...nearly any current application or useful RF technique is an appropriate topic!*

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**Stanford**  
Microdevices

**data sheet**

**SMM-210**  
**1.5-2.5 GHz, 1Watt**  
**GaAs MMIC**  
**June, 1994**

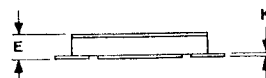
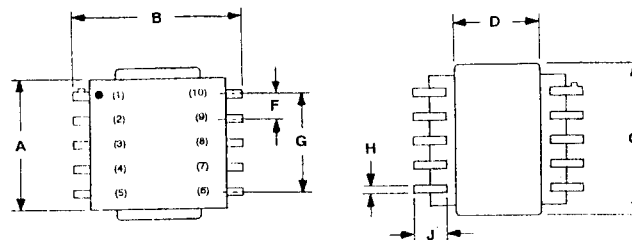
### Features

- 27dB Gain and 30dBm Output Power
- Excellent VSWR: 1.5:1 Typical
- High Power Added Efficiency
- Characterized at 3 volts for portables

### Description

Stanford Microdevices' SMM-210 is a high performance gallium arsenide monolithic-microwave-integrated-circuit (MMIC) housed in a low-cost surface mount package. Designed for operation in wireless systems operating in the 1500 to 2500 MHz frequency range; this amplifier has 30dBm of output power and 27dB Gain at P1dB.

Also available in die form, its small size (2.0 x 1.5mm) makes it suitable for use on thin and thick-film circuits. Proven gold-based metallization and die passivation add to the reliability and durability of this device.



PIN DESIGNATION TABLE	
PIN NO	DESCRIPTION
1	GROUND
2	GROUND
3	RF IN
4	GROUND
5	Vg = -5V
6	GROUND
7	GROUND
8	RF OUT
9	GROUND
10	Vd = +5V

DIMENSION TABLE		
LTR	DIMENSIONS	NOTES
A	0.270	SQUARE
B	0.355	
C	0.310	
D	0.175	
E	0.060	
F	0.050	TYPICAL
G	0.200	
H	0.015	TYPICAL
J	0.070	TYPICAL
K	0.010	TYPICAL
ALL DIMENSIONS ARE IN INCHES		
TOLERANCES: +/- 1% NLT 0.005		

### Electrical Specifications at Ta = 25c

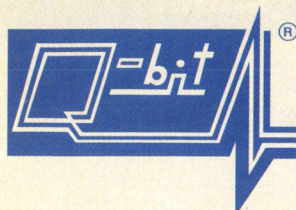
Symbol	Parameters: Test Conditions	Units	Min	Typ	Max
P1dB	Output Power at 1dB Compression Vdd = 7.0 V, Idd = 600 mA	f = 2.0 GHz	30	31	
		f = 2.4 GHz		29	
Ga	Associated Power Gain Vdd = 7.0 V, Idd = 600 mA	f = 2.0 GHz	25	27	
		f = 2.4 GHz		25	
Psat	Saturated Output Power Vdd = 7.0 V, Idd = 600 mA	f = 2.0 GHz	31	32	
		f = 2.4 GHz		30	
P1dB	Output Power at 1dB Compression Vdd = 3.0 V, Idd = 400 mA	f = 2.0 GHz	24	25	
		f = 2.4 GHz		23	
Ga	Associated Power Gain Vdd = 3.0 V, Idd = 400 mA	f = 2.0 GHz	22	24	
		f = 2.4 GHz		20	
VSWR	Input and Output Match			1.5:1	
IP3	Third Order Intermodulation Distortion	dBm	37	40	
PAE	Power Added Efficiency Vdd = 7.0 V, Idd = 600 mA	%	25	30	

### Other Products from Stanford (Late '94)

Model	Freq(GHz)	Description
SMM-228	1.5-2.0	5 Watt MMIC Amp, 34dB Gain
SHM-610	5-7	3 Watt MMIC Amp, 27dB Gain
SMM-910	8-10	5 Watt MMIC Amp, 15dB Gain
SMM-1410	13-15	3 Watt MMIC Amp, 22dB Gain
SMM-108	0.5-3.0	LNA, 2dB NF and P1dB of 17dBm
SMM-810	2-8	Broadband Amp, 25dB Gain, 17dBm Po

**Call 1(800) SMI-MMIC or fax (408) 746-3630**





# Commercial Radio & Cellular Band Amplifiers

800	820	840	860	880	900	920	940	960	980
MHz									
Cellular Base Station RX		Public Safety Radio		GSM Band					
Public Safety Radio				SMR		Japanese Cellular Band			
<ul style="list-style-type: none"><li>• QB-304</li><li>• QB-761</li><li>• QBH-1254</li><li>• QBS-110</li></ul>	<ul style="list-style-type: none"><li>• QB-304</li><li>• QB-761</li><li>• QBH-1254</li><li>• QBS-108</li><li>• QBS-110</li><li>• QBS-133</li><li>• QBS-135</li><li>• QBS-136</li><li>• QBS-137</li><li>• QBS-141</li><li>• QBS-142</li></ul>			<ul style="list-style-type: none"><li>• QB-304</li><li>• QB-761</li><li>• QBH-1254</li><li>• QBS-125</li><li>• QBS-146</li><li>• QBS-147</li></ul>		<ul style="list-style-type: none"><li>• QBS-104</li></ul>		<ul style="list-style-type: none"><li>• QBS-125</li><li>• QBS-126</li><li>• QBS-127</li></ul>	
SMR • Specialized Mobile Radio									

## Specification Summary

Guaranteed Specifications @ 25 °C

Model	Frequency Range MHz	Gain dB	Output P1dB dBm	VSWR	Noise Figure dB	Reverse Isolation dB	3rd OIP <sup>1</sup> dBm	DC Power Vdc/mA	Housing
QB-304	800-900	20.0	24.0	1.5:1	3.0	37	40/	15/300	19202
QB-761	806-870	23.0	18.0	1.5:1	3.5	42	32/	15/140	187-2
QBH-1254	804-901	12.5	23.5	1.5:1	3.0	26	37/	15/142	TO-8
QBS-104	896-925	27.0	26.0	1.7:1	2.0	37	41/	15/450	19069
QBS-108	824-849	8.0	25.0	2.0:1	5.0	12	40/	15/200	19134
QBS-110	806-849	27.0	26.0	1.7:1	2.0	37	41/	15/450	19069
QBS-125	896-960	40.0	28.0	1.5:1	1.5	51	42/	15/850	19130
QBS-126-1	925-960	40.0	28.0	1.5:1	1.5	51	42/	15/850	19130
QBS-126-2 <sup>2</sup>	925-960	40.0	28.0	1.5:1	1.5	51	42/	19-31/850	19105
QBS-127-1	925-960	33.0	22.0	1.5:1	1.5	41	38/	15/450	19130
QBS-127-2 <sup>2</sup>	925-960	33.0	22.0	1.5:1	1.5	41	38/	19-31/450	19105
QBS-133	824-849	33.0	19.0	2.0:1	1.3	41	35/	15/250	19121
QBS-135	824-849	40.0	24.0	2.0:1	1.3	51	39/	15/425	19121
QBS-136	824-849	14.0	25.0	2.0:1	5.5	21	40/	15/325	19134
QBS-137	824-849	26.0	15.0	2.0:1	1.3	36	28/	15/150	19134
QBS-141-1	824-849	40.0	28.0	1.5:1	1.5	51	42/	15/850	19130
QBS-141-2 <sup>2</sup>	824-849	40.0	28.0	1.5:1	1.5	51	42/	19-31/850	19105
QBS-142-1	824-849	33.0	22.0	1.5:1	1.5	41	38/	15/450	19130
QBS-142-2 <sup>2</sup>	824-849	33.0	22.0	1.5:1	1.5	41	38/	19-31/450	19105
QBS-146-1	870-915	40.0	28.0	1.5:1	1.5	51	42/	15/850	19130
QBS-146-2 <sup>2</sup>	870-915	40.0	28.0	1.5:1	1.5	51	42/	19-31/850	19105
QBS-147-1	870-915	33.0	22.0	1.5:1	1.5	41	38/	15/450	19130
QBS-147-2 <sup>2</sup>	870-915	33.0	22.0	1.5:1	1.5	41	38/	19-31/450	19105
<b>Power Amplifiers</b>									
QBS-227	800-960	40.0	40.0	2.0:1	-	60	50	24/2600	
QBS-230	800-960	40.0	43.0	2.0:1	-	60	53	24/4400	
QBS-233	1800-2200	40.0	40.0	2.0:1	-	60	52	12/4600	

SPECIFICATIONS ARE 25°C TYPICAL

<sup>2</sup> SOFT FAIL DESIGN WITH BUILT IN REGULATION

<sup>1</sup> OIP = OUTPUT INTERCEPT POINT

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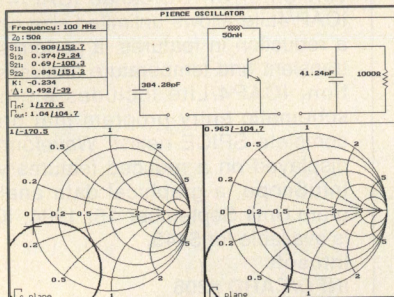




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## RF Amplifier Design

Optimization of Gain and Noise Figure  
Stability Analysis  
Maximum Unilateral Gain  
User-definable Transducer Gain  
Noise Figure Analysis  
Circuit Diagram Generator

## Wideband Amplifier Design

Optimizer  
Stability Analysis  
Circuit Diagram Generator

## RF Oscillator Design

Negative Resistance  
Oscillator Design  
Colpitts Oscillator Design  
Pierce Oscillator Design

## Lumped Element Matching

Computerized Smith™ Chart  
Interactive Design  
Circuit Diagram Generator

## Computer Requirements

IBM PC or compatible  
VGA graphics  
Math co-processor  
Laser printer

## RF POWER AMPLIFIERS

**HIGH PERFORMANCE**

**1 MHZ - 2GHZ • 1W - 1KW**

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4.84" x 2.0" x 1.0" • 200 Watts • 150 - 250 MHZ  
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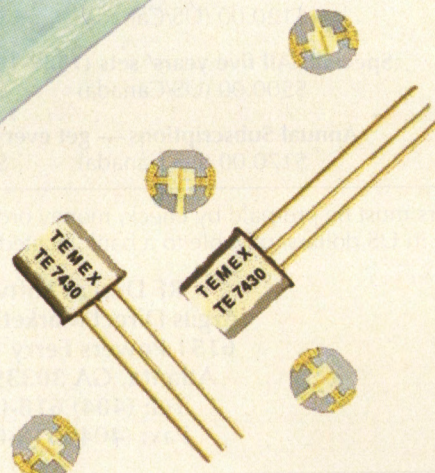
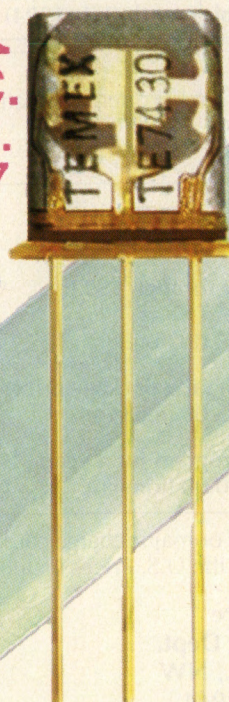
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# COAXIAL RELAYS

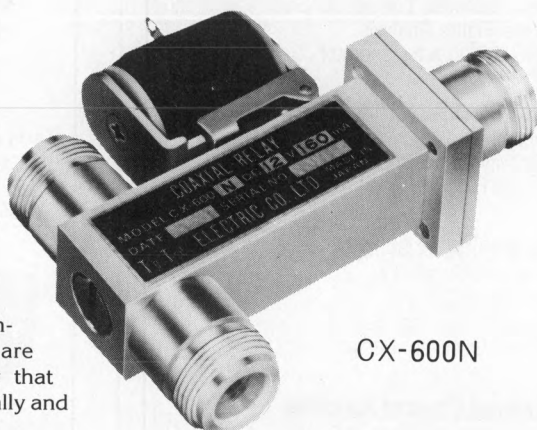
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"Coaxial relay reliability has always been a problem. About five years ago, we switched to TohTsu Coaxial Relays in our high-power amplifiers. Since that time we have not experienced a single relay failure."

"TohTsu offers traditional Japanese quality and workmanship. The contact designs are exceptional, making a relay that has proven itself both electrically and mechanically reliable."

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INFO/CARD 56

## RF Design Software

*Programs from RF Design provided on disk for your convenience*

**NOTE! There is no new program for this month (July 1994)**

### June Program Disk — RFD-0694

"Phase-Locked Loop Parameters and Filters" by Jack Porter. Program provides design and analysis of PLL active loop filters. (True Basic, compiled version and source code).

#### Monthly program disks:

\$25.00 (US/Canada) \$30.00 (foreign)

#### Yearly disk sets available for 1989 through 1993:

##### Order #RFD-(year)-SET

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## RF software

### Budget Version of SPICE Simulator

ICAP/4Lite is a low-cost analog and mixed-signal circuit simulation system. The software is based on Intusoft's professional version of SPICE called ICAP/4Windows. ICAP/4Lite allows unlimited size circuits with a reduction in features. It performs AC, DC, transient and temperature analyses. In addition, ICAP/4Lite includes an integrated schematic entry program that produces a complete SPICE netlist. Waveforms can be displayed on a reduced features version of IntuScope, a graphical data analysis program. Until September 15, 1994, ICAP/4Lite is available for \$445.

**Intusoft**  
INFO/CARD #206

### RF Library for DSP Simulation

Alta Group (formerly Comdisco Systems) has introduced its RF Library for the Signal Processing WorkSystem® (SPW™). The RF Library is a collection of blocks for modelling the RF portion of a signal processing system. The library includes laboratory-validated blocks for components such as: nonlinear amplifiers, switches, couplers, A/D converters and mixers. Alta Group's RF Library is priced beginning at \$5000 and will be available in the third quarter of 1994.

**Alta Group**  
INFO/CARD #207

### Filter Design

Filtroid from Geesof synthesizes, joins, edits and analyzes filter circuits. Lowpass, highpass, bandpass and bandstop filters can be designed using Tchebychev, Butterworth, and Elliptic transfer functions or, user-specified masks, user-specified G-values or user-specified K-Q values. Filtroid provides tabular and graphic output of analysis of VSWR, return loss, phase/phase linearity and other parameters. Mechanical dimensions for many distributed filters can also be calculated. Filtroid sells for \$2995.

**Geesaman Software, Inc.**  
INFO/CARD #208

### PC-Based 3D EM Simulation

Developed by Zeland Software and Marketed by Bay Technology, IE3D performs 3D electromagnetic simulation of electrical circuits ranging from DC to over 100 GHz on a personal computer operating MS Windows. IE3D supports all 2D and 3D simulation tasks previously done only on workstations. Using a full-wave integral equation method-of-moments algorithm, IE3D keeps the number of nodes to a minimum without any reduction in accuracy and with reduced compute time over workstation products. IE3D for MS-Windows operates on any 386/486 PC and is priced at \$5000 and \$10,000 (depending on model size) and \$20,000 for the Windows NT version.

**Bay Technology**  
INFO/CARD #209



# RF literature

## Crystal Oscillator Catalog and Handbook

Vectron Laboratories has published its new 1994 comprehensive full-line handbook and catalog. The 80-page publication details a complete line of crystal oscillators available from 0.01 Hz to beyond 2 GHz and highlights new products recently introduced.

**Vectron Laboratories, Inc.**  
INFO/CARD #220

## Fiberoptic Communications Guide

Ortel Corp. has released a four-page Product Use and Selection Guide for the Wireless Communications market. The guide includes photos, specifications, technical diagrams and applications for standard and specific application cellular radio fiberoptic transmitters. The guide is available free to all qualified specifiers.

**Ortel Corporation**  
INFO/CARD #219

## Antenna and Accessory Catalog

Antenna Research's 127-page catalog includes descriptions of the company's lines of omnidirectional, directional, GPS, microstrip, aerospace, D.F., and search and surveillance antennas, along with their antenna test equipment and accessories. Also described is their line of individually calibrated antennas for EMI/EMC testing.

**Antenna Research Associates, Inc.**  
INFO/CARD #218

## Multichannel PA Catalog

AML Communications announces the publication of its Cellular/PCS Multichannel Power Amplifier Catalog. The catalog details amplifiers with PEP powers up to 240 W with intermods as low as -60 dBc. Designs include feed-forward, class AB, and class A types. Operating frequencies cover the North American and International Cellular spectra.

**AML Communications**  
INFO/CARD #217

## Cables and Assemblies

Coaxitube is introducing its expanded product catalog featuring semi-rigid and flexible coaxial cables and assemblies. This 64-page multi-color publication contains a cable identification guide and updated descriptions of Coaxitube's hand-formable Al and Cu semi-rigid cables. Flexible cables and cable assemblies are also featured along with a wide choice of specialty type cables.

**Precision Tube Co., Inc.**  
**Coaxitube Div.**  
INFO/CARD #216

## Low-Price Circuit Material Literature

Rogers Corp. offers a literature package for

its RO3003™ High Frequency Circuit Materials for commercial microwave and RF printed circuit boards. The literature includes a preliminary data sheet and fabrication guidelines.

**Rogers Corp.**  
INFO/CARD #215

## Crystal Oscillator Catalog

Wenzel Associates' 1994 catalog Crystal Oscillators, RF Products outlines a variety of high performance crystal oscillators, frequency standards, multipliers, mixers and other RF products designed for MIL, space and commercial applications. The catalog updates performance and engineering capabilities for custom and standard products.

**Wenzel Associates, Inc.**  
INFO/CARD #214

## Updated Data Books

Motorola has released its revised RF Device Data book for 1994 (DL110/D Rev 5). The product families include bipolar, LDMOS, MOSFET RF power, and GaAs chip technologies. Motorola's 1994 RF Selector Guide and Cross Reference (SG46/D Rev 11) includes a new RF Integrated Circuits category. Either publication can be ordered by calling (800) 441-2447, or by faxing (602) 994-6430.

**Motorola Semiconductor**  
INFO/CARD #213

## Rental/Lease Guide

GE Rental/Lease announces its updated product guide. The 210-page catalog details many of GE Rental/Lease's more than 120,000 name-brand test and measurement instruments. In addition to equipment information, the catalog contains financing option information.

**GE Rental/Lease**  
INFO/CARD #212

## Detector/Limiter Brochure

A four-page, color brochure from Advanced Control Components describes their lines of tunnel diode detectors, limiter and pad detectors, schottky diode detectors, and limiters. Design considerations for both detectors and limiters are listed.

**Advanced Control Components**  
INFO/CARD #211

## Circuit Fabrication Capabilities

Polyflon has published a four-page, color brochure describing their circuit processing capabilities, including material selection, CAD, plating, plated through holes, machining, quality assurance and control, and testing. In addition, Polyflon offers a design data sheet describing their multilayer mixed media capability, which allows combination of microstrip and stripline circuits in one unit.

**Polyflon Co.**  
INFO/CARD #210

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# RF guide to editorial coverage

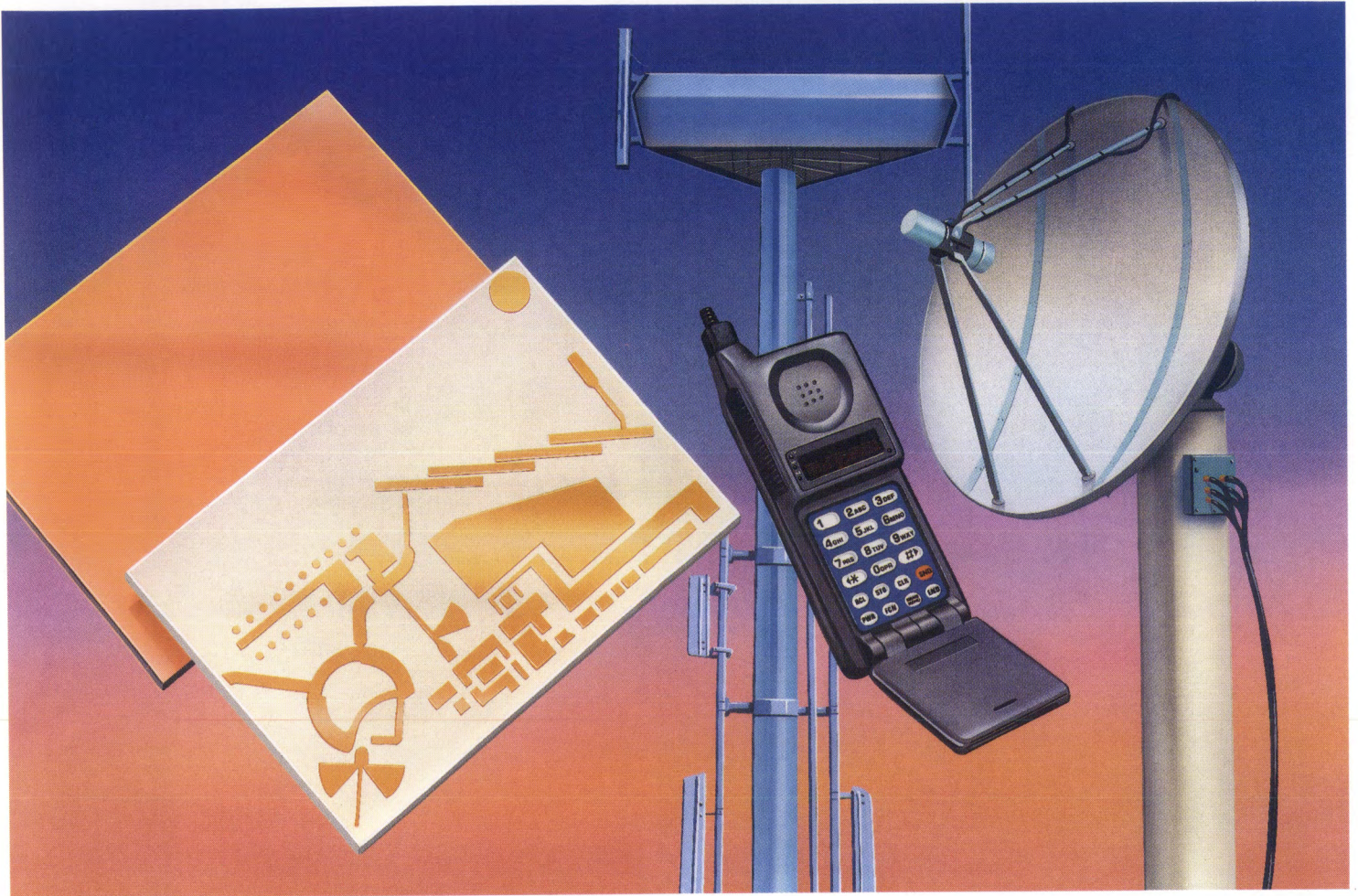
Advanced Control Components	83	EM Research Engineering	52	Power Systems Technology	48
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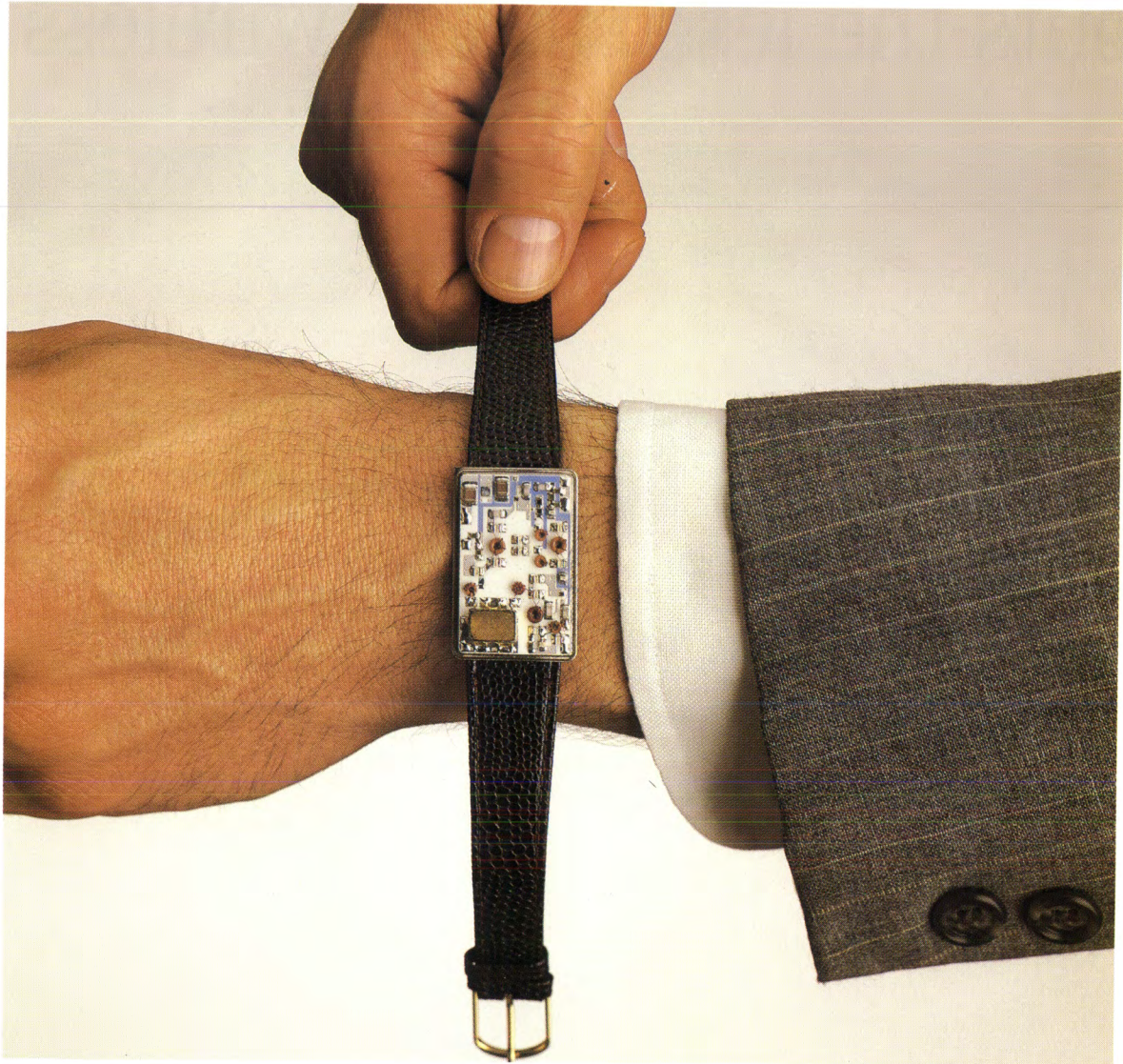
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