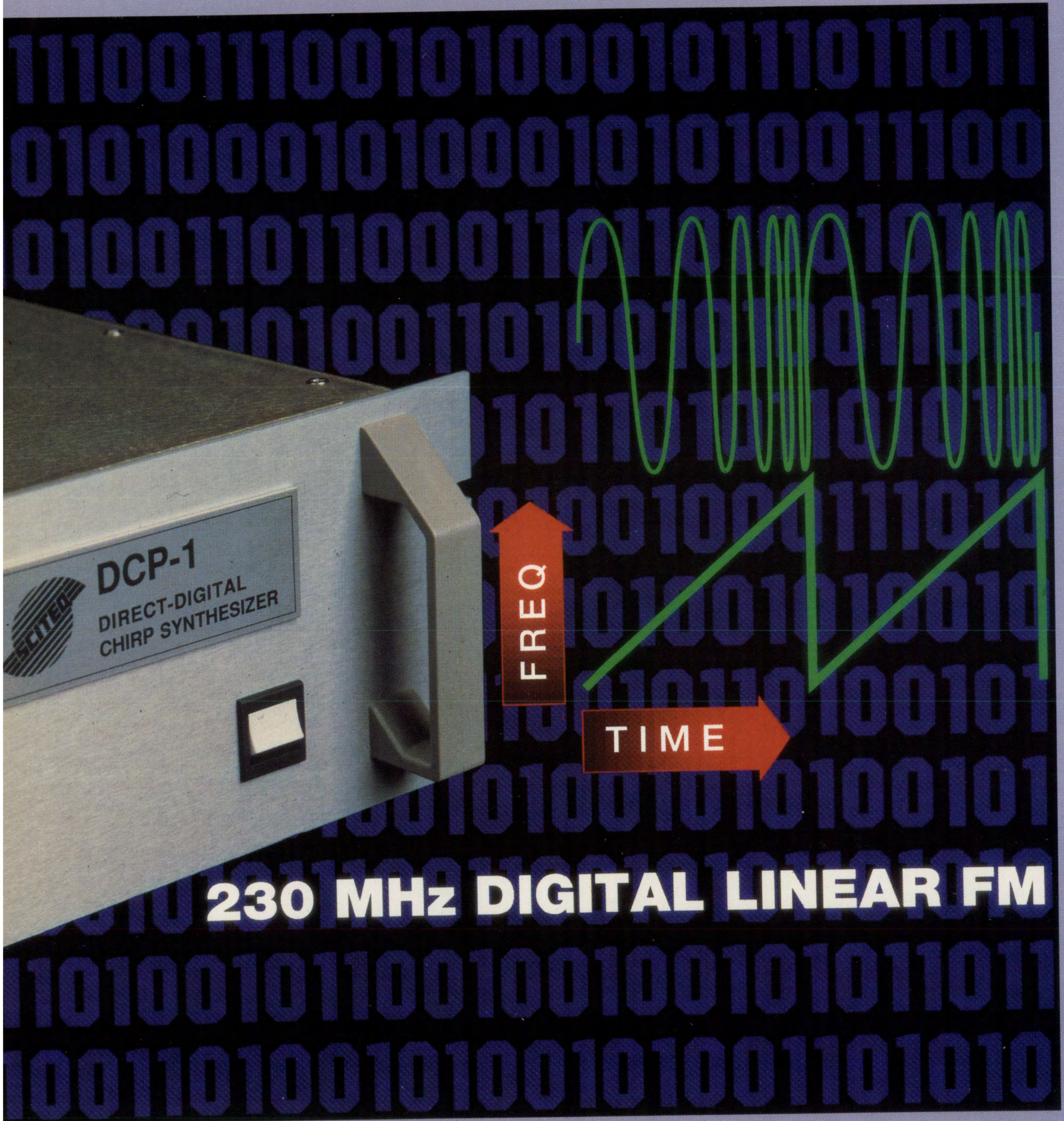
RF design™

engineering principles and practices

September 1993



Featured Technology
Direct Digital Synthesis

Cover Story
DDS Drives New Chirp Synthesizer

Inside — RF Expo East Technical Program Preview

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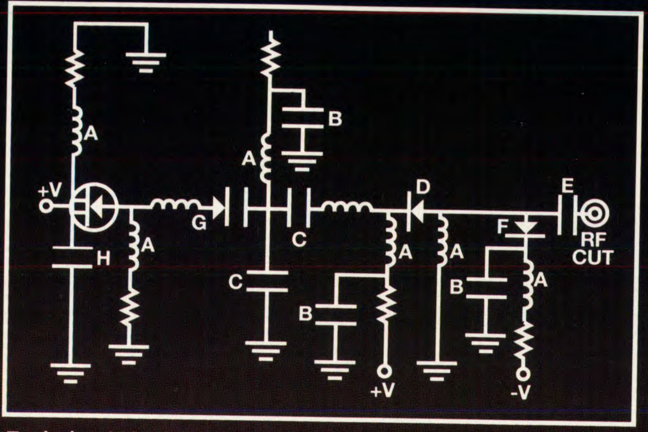

Typical switch with FSI semiconductors. (A) GC9002 Spiral Inductor; (B) GC84020 MNS Chip Capacitor; (C) GC86001 MNS Chip Capacitor; (D) GC4901 Beam Lead; (E) GC84030 MNS Chip Capacitor; (F) GC4221 PIN Diode; (G) GC15004 Linear Tuning Varactor; (H) GC84001 MNS Chip Capacitor.

Sample requests filled promptly. Comprehensive, 120-page catalog free on request. For more information, call or write Donna Langan, Semiconductor Sales Manager, Loral Microwave-FSI, 16 Maple Road, Chelmsford, MA 01824. Tel: (508) 256-4113. Fax: (508) 937-3748.

INFO/CARD 1

LORAL

Microwave - FSI

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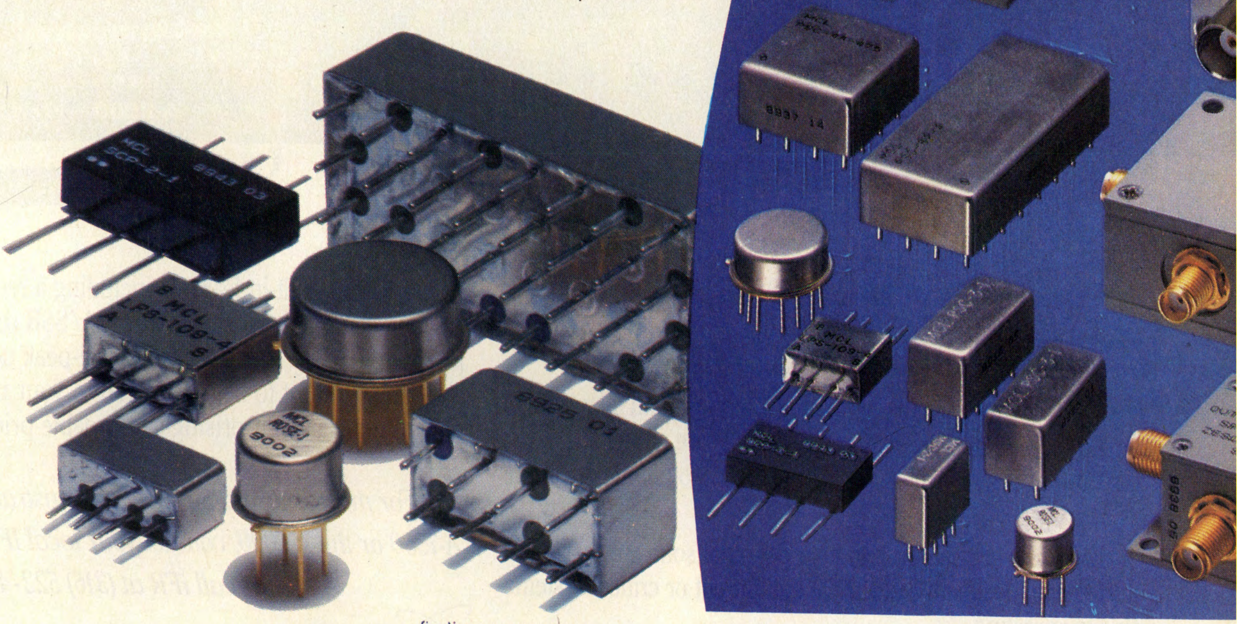

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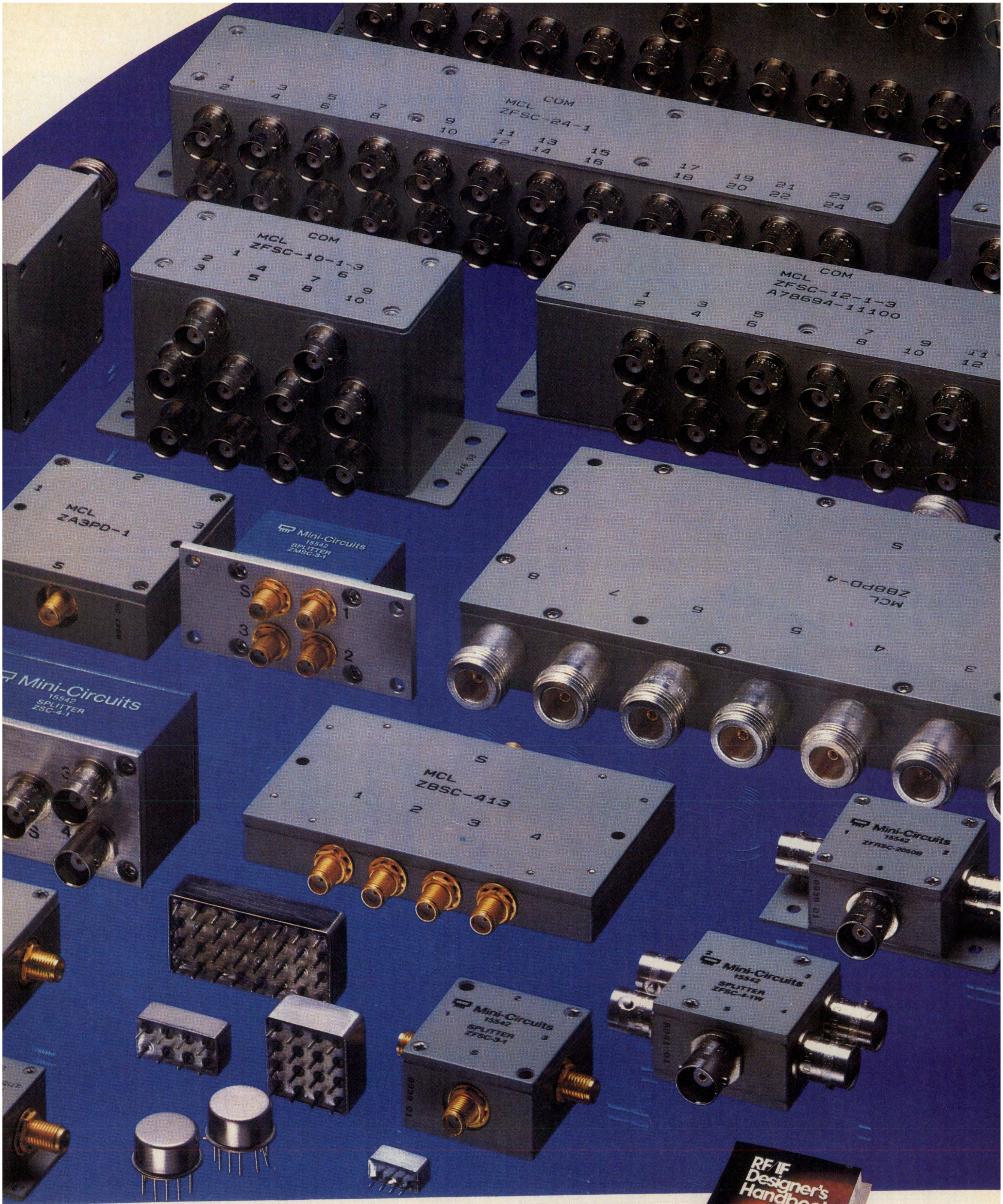
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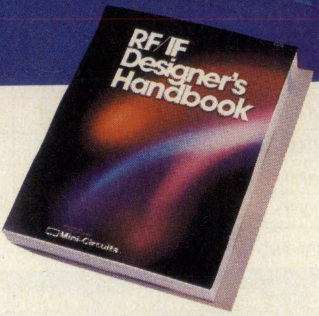
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SP2T	5-4000	1.0	79	0.035	TTL	SMA	CDSO882
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SP2T	10-2000	0.60	82	1.000	TTL	14 Pin DIP	DSO052
SPST	10-2000	1.8	67	0.035	TTL	14 Pin SMP	DSO790
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SP4T	DC-2000	1.7	70	0.075	TTL	14 Pin DIP	DSO874

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6	10-1000	1-63	4.2	0.050	TTL	24 Pin DIP	DAO886
7	DC-1000	0.5-63.5	7.1	0.100	TTL	38 Pin DIP	DAO897
7	10-1000	0.5-63.5	6.4	0.035	TTL	SMA	CDAO867

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RF design

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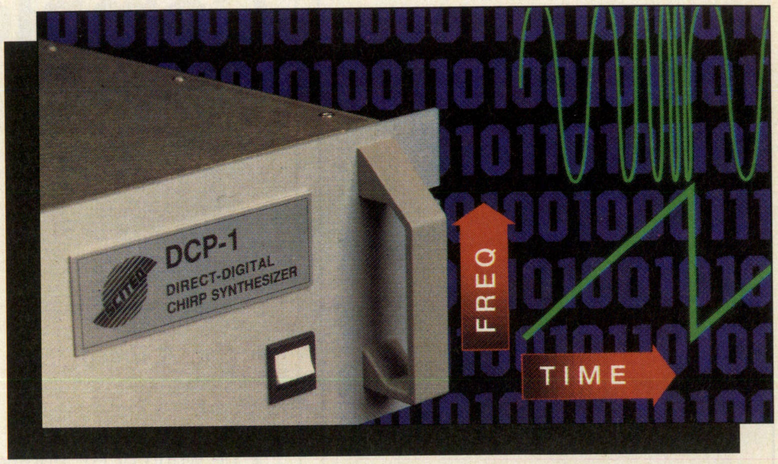

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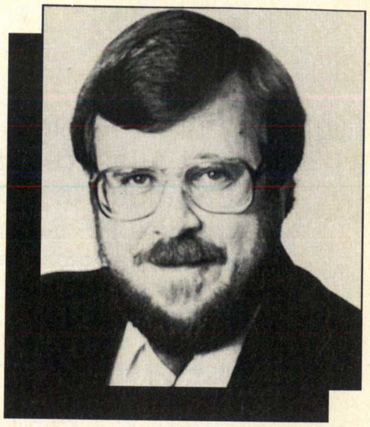
RF editorial

Coping With Complexity



By Gary A. Breed
Editor

The world of RF engineering is getting much more complex in a big hurry. Just when we thought we were getting a handle on BPSK, QPSK, π/4 DQPSK and MSK, we get direct sequence, frequency hopping, digital compression, forward error correction and adaptive signal processing. Instead of just worrying about simple downconversion, IF amplification, filtering and detection, RF engineers have to deal with microscopic circuit layouts, custom ICs, minimal power consumption, fast frequency synthesizers, and management that wants a product yesterday.

Adding to the challenge of rapid engineering development is the regulatory process. The FCC, IEEE, ANSI, IEC and other standards and regulatory bodies are being asked to act faster than they ever have before. Sometimes, technology can take a major step ahead before they can act, making the work they just finished obsolete! This happened with HDTV, as analog systems gave way to digital systems, which were superseded by still other digital systems — it has the FCC pleading with developers to get together on a single system, anti-trust laws notwithstanding.

The question that arises from this recent acceleration in RF technology is, "How can I deal with the magnitude of it all?" There is no good answer, but I've gotten a few suggestions from some RF engineers who have been around for a long time:

"Go with the flow, work hard and do the best you can," is trite, but reliable advice. We wouldn't have the problem if we weren't right in the middle of a new RF revolution. Personally, I'd much rather be confused, frustrated and overwhelmed by success and growth than comfortably cruising through the same old stuff!

The other suggestion is simply to think; and to think simply. Too often, we automatically apply complex answers to complex problems, even when a simple solution will work. The first way to simplify our thinking is to avoid duplicating work that has already been done. This means we need to read, write and talk with our colleagues. There is an infinite supply of information that may be useful — learn to find your way through it.

Another part of thinking is just that — thinking — before you go to the computer, your favorite reference books, or to manufacturers' data books. Complex problems must be well-defined and well-understood before they can be solved; you can't just jump in and start designing. No matter how complex a task appears to be, it can be divided into simpler sections. Computer programmers have known this for years, using structured programming and modular software design.

Work from a block diagram first, pick an architecture that makes use of standard components, apply well-known design methods whenever possible, decide just how much of the system needs to be really high performance. For example, can that RF data link you're working on be done with a simple RF front end and some great digital signal processing; or will a superior RF chain make the back-end processing really easy?

So, when the magnitude of a design assignment overwhelms you, don't panic. Get organized, figure out all the angles, do your homework, then attack it one step at a time.

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Editorial and Advertising Office:
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111 • (303) 220-0600
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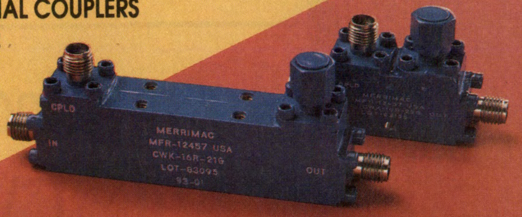

- Octave and multi-octave bandwiths to 65 GHz
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- High Power Dual Directional, 500W failsafe to 5 GHz
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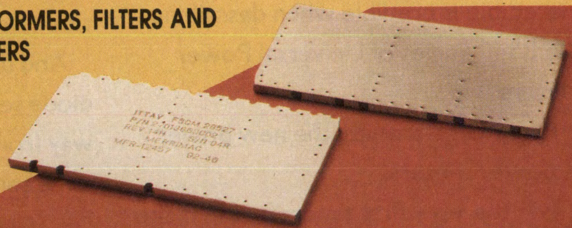

- Narrow and wideband Beamformers with low loss to 18 GHz
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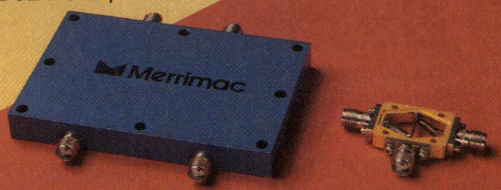

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INFO/CARD 9

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Incredible is credible when describing the 8540 Series of Universal Power Meters.

From Giga-tronics, the new power in power meters.

For the very first time, you can make CW *and* peak power measurements quickly and accurately with a single meter—a Universal Power Meter.

And all for about the same price you'd pay for the competitor's CW *only* power meter.

The two-line display also lets you set the desired resolution and select either Lin or Log readout for each line.

POWER MEASUREMENTS INSTANTLY

Imagine seeing display updates instantly: measurement speeds over the GPIB exceeding 200 readings per second and an exclusive Burst Mode capturing more than 2,000 readings in the same tick of a clock.

And because the 8540 Series uses diode sensors, you can measure all the way from −70 to +20 dBm with the same sensor, and without range changing delays.

If you're worried about having to write new code for your computer controlled testing, don't be: The 8540 Series uses the same GPIB command set as HP's 436A, 437B and 438A.

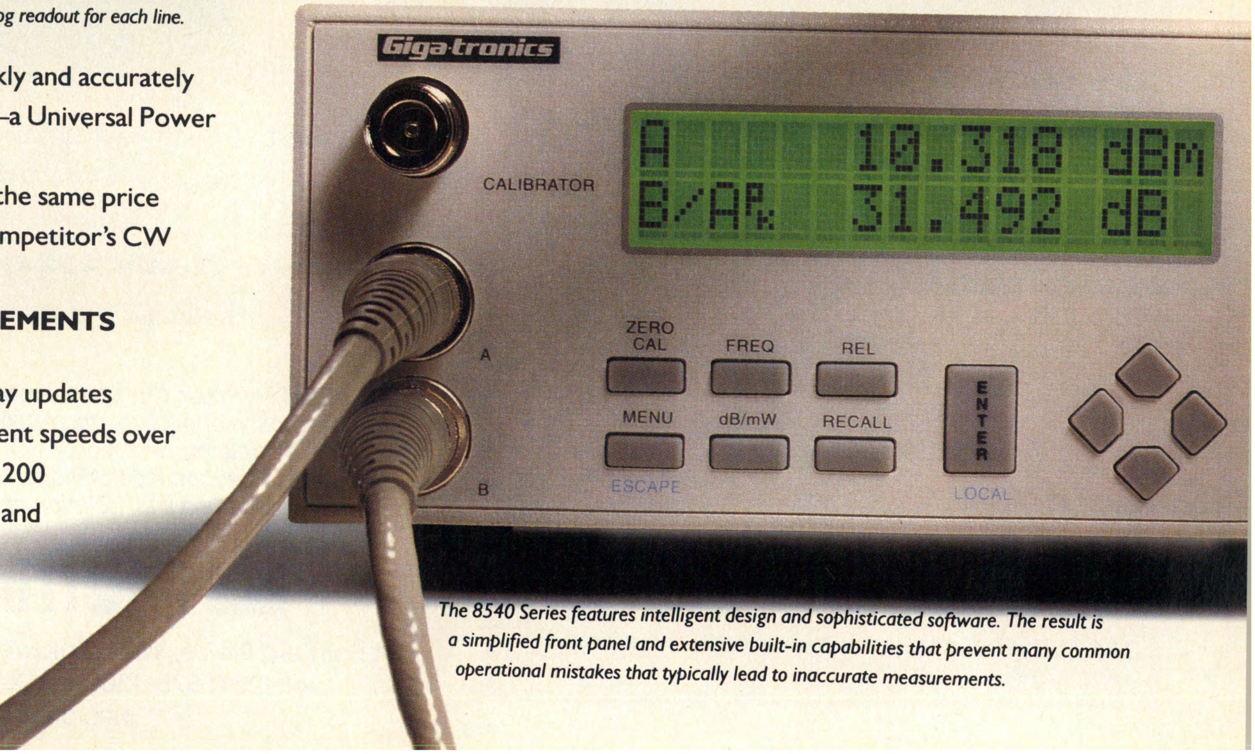
Think about what all this will do for your ATE productivity as well as for your company's bottom line.

FAST, EASY PEAK POWER MEASUREMENT

Now, an easy-to-use CW power meter can also measure pulsed RF signals with the simple addition of a peak power sensor.

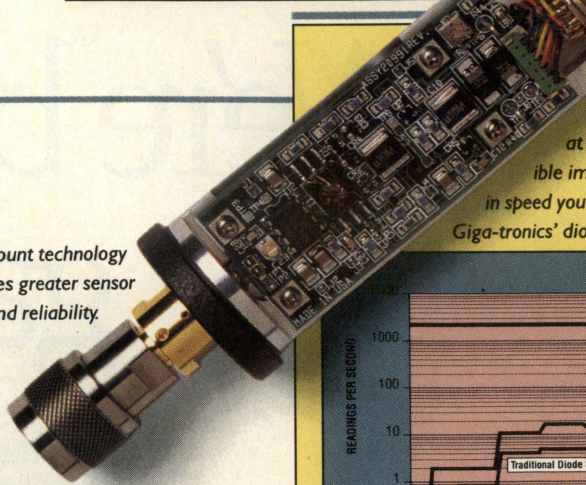
There are no time-consuming, unreliable duty cycle corrections, and you'll get the same accuracy and speed you'd get with a much-more-expensive dedicated peak power meter.

View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.



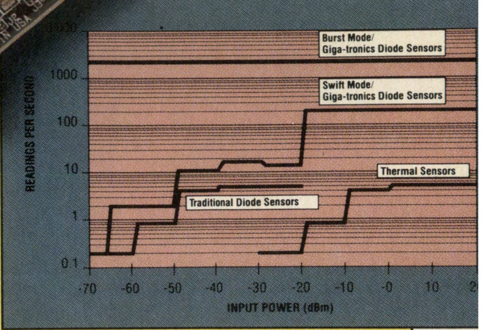
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.



The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

What's more, Giga-tronics has solved the challenge that previously limited diode sensors to the "square law" region—below −20 dBm—by utilizing a built-in power sweep calibration system. So you get speed and a full 90 dB dynamic range without sacrificing accuracy.

You'll be confident of your peak power readings, and still have all the benefits of an incredibly fast CW power meter.

ONE OR TRUE TWO CHANNEL OPERATION

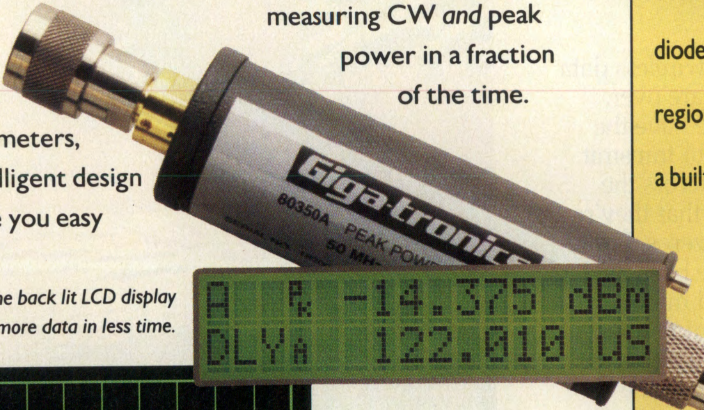
If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously.*

SIMPLE, INTELLIGENT OPERATION

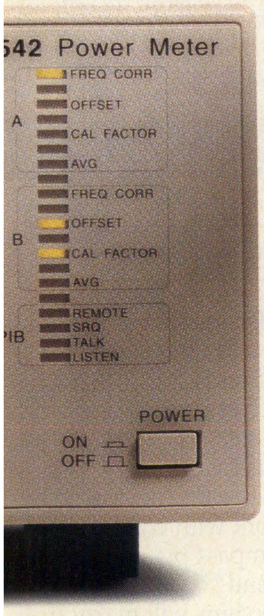
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

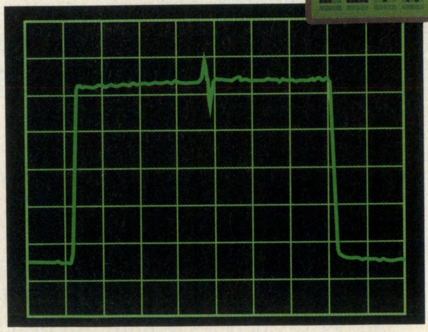
For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

Imagine all this power and performance. But why just imagine? Get the truly incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.



A two-line back lit LCD display provides you more data in less time.





The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulsed signals.

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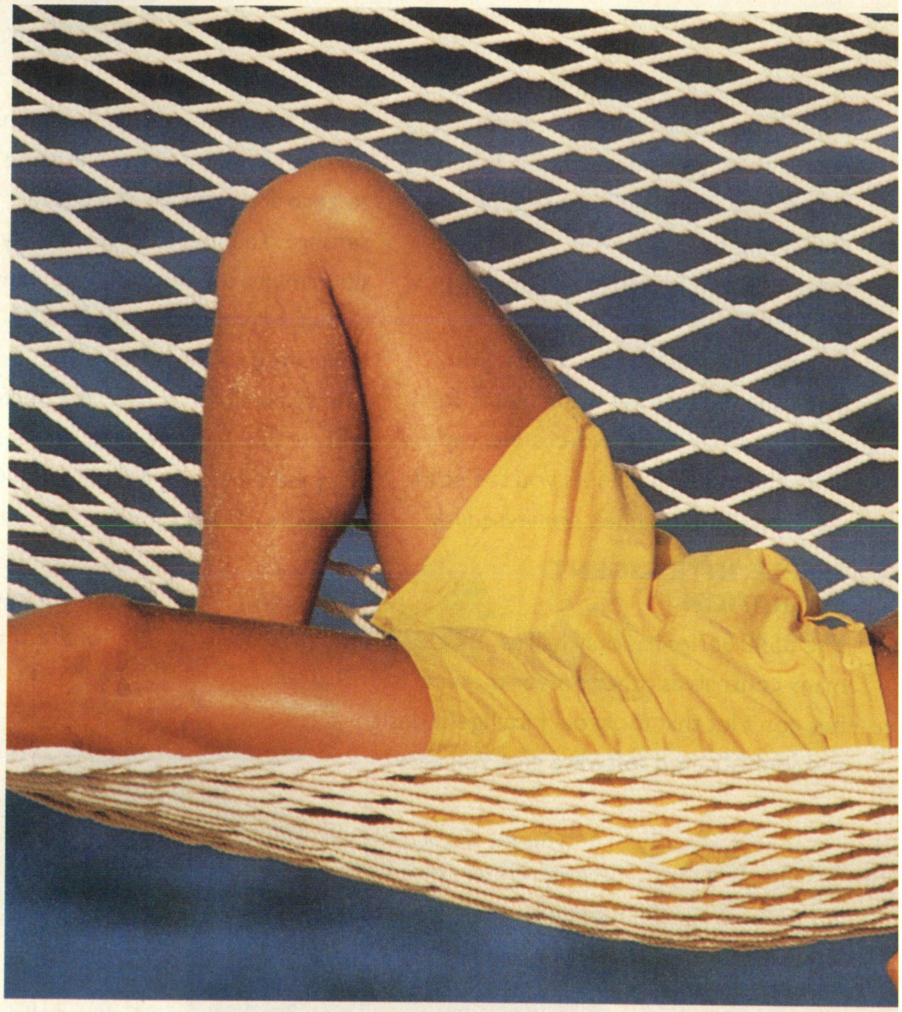

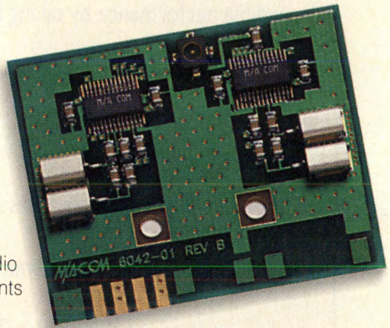

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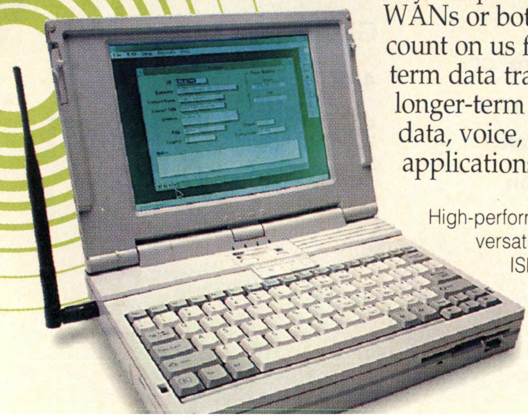

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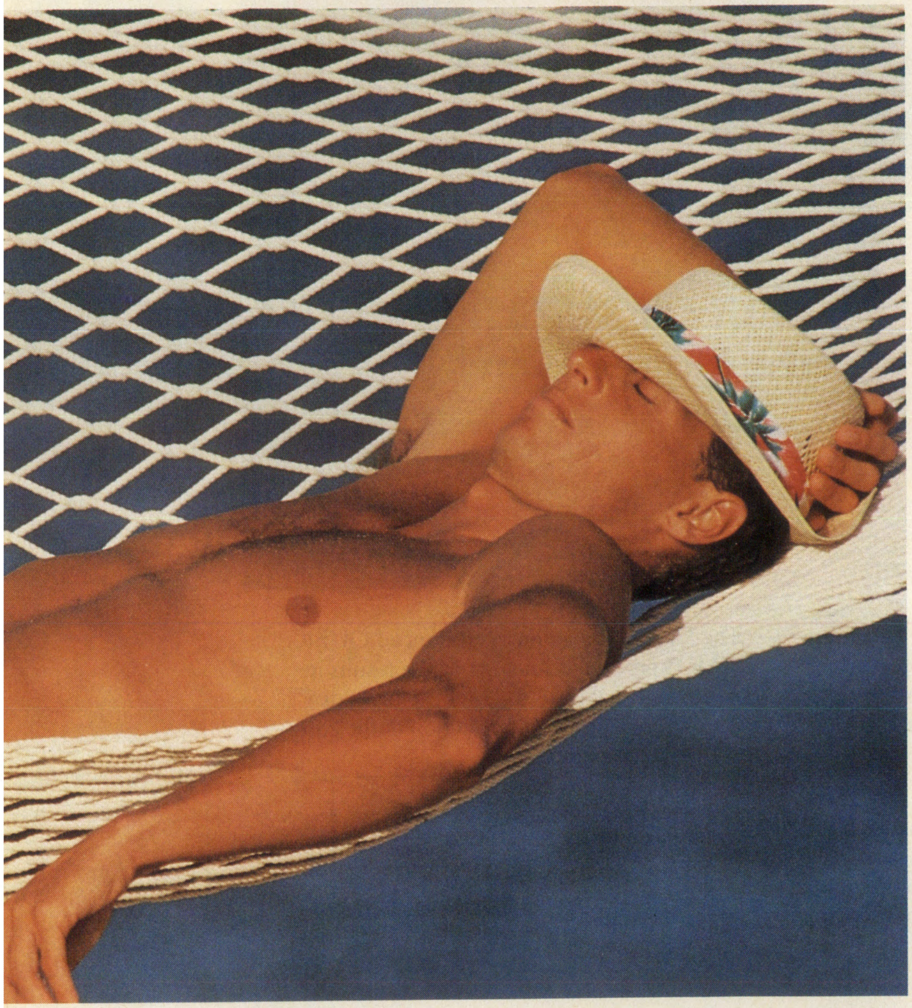

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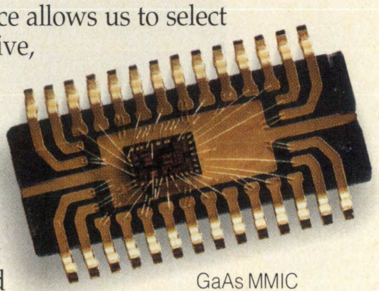

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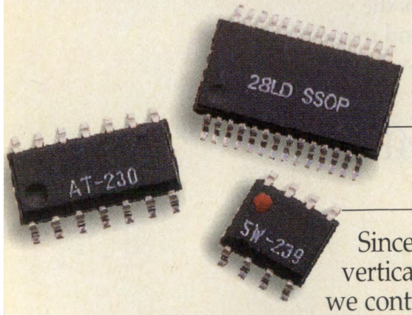

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INFO/CARD 11

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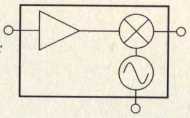
The SA620 is based on our advanced QUBiC BiCMOS process technology which results in performance that rivals today's best GaAs discrete and IC designs—at a fraction of the cost. Just look at the numbers:

The SA620 Integrated Front-End: 1 GHz LNA, Mixer and VCO. Available in 20-pin SSOP.



SA620 @900MHz	Vcc	Icc	LNA		Mixer	
			NF	Gain	NF	Gain
	3V	10.4mA	1.6dB	12dB	8.5dB	3dB

So instead of designing front-ends the way you used to, save time, space and money by selecting the Philips SA620 low-voltage Integrated Front-End.

1-800-447-1500 ext. 1054RF

Philips Semiconductors

PHILIPS

PHILIPS

RF letters

President's Letter

This month, readers of *RF Design* will see a new name, Argus Business, on the masthead.

Argus Business is a new subsidiary publishing company of Argus, Inc. This new subsidiary grew from a merger in July of Cardiff Publishing Company of Denver, publishers of *RF Design*, and Communication Channels, Inc. of Atlanta.

RF Design, one of Cardiff's 12 titles, joins with CCI to form a group consisting of 65 titles, plus 16 trade shows and conferences.

RF Design will retain its name and editorial independence, while bringing to our readers and advertisers the combined strengths of a group of publications serving more than two million readers in 10 industry segments. The merger will help identify Argus as one of the largest trade publishing groups in the United States.

As always, *RF Design* will maintain its support of the RF industry with a continued commitment to editorial excellence.

Jerrold France
President, Argus Business

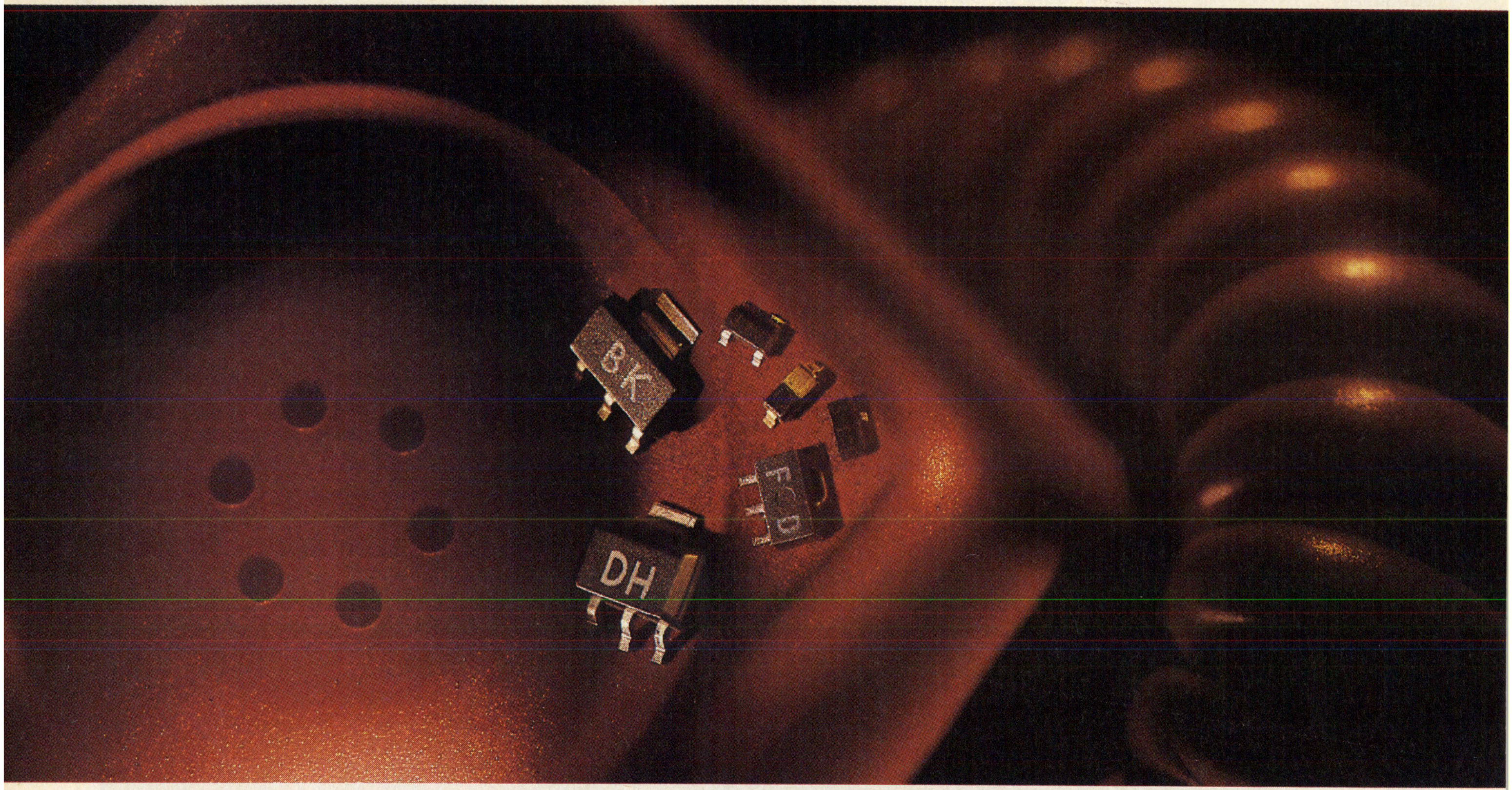
Correction²

The correction printed in July's "RF Letters" section was incorrect. Equation 3 in Andrzej Przedpelski's May 1993 tutorial "A Comparison of Simple Software Methods for RF Calculations" should be:

$$[(42+j2\pi f\cdot 0.01+(j2\pi f\cdot 0.342\times 10^{-15})^{-1})^{-1}+j2\pi f\cdot 2\times 10^{-12}]^{-1}$$

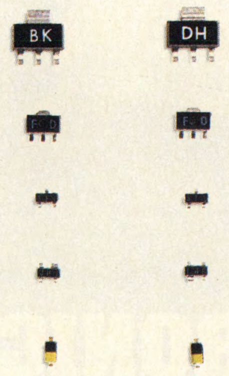
Thanks to Les Besser for pointing out the error.

SIEMENS



RF Performers

Announcing the most advanced RF transistors for mobile communications.



Siemens announces a full range of high-performance RF transistors offering increased gain, reduced noise, and lower power consumption. Best of all, these advanced devices *cost less* than devices you may currently be using. **Call 1-800-77-SIEMENS, ext. 310** for a complete data kit on mobile communications devices (including a data sheet about Siemens surface mount GaAs MMIC).

Type	Max Ratings		Characteristics		Application
NPN	U_{CEO} V	I_C mA	F 900 MHz 8 V dB	G_{pe} 900 MHz 8 V dB	U_{CE}/I_C f = 0 to above 3 GHz V/mA
BFP 180	8	3	1.9	15	1 to 3/0.2 to 2.5 paging system
BFP 280	8	10	1.4	18	1 to 5/0.2 to 8 low-noise pager
BFP 181	12	20	1.3	19	1 to 8/0.5 to 10 low-noise pre-amplifier
BFP 182	12	35	1.2	19	1 to 8/1 to 20 low-noise amplifier
BFP 183	12	65	1.2	19	1 to 8/2 to 28 low-noise amplifier
BFP 193	12	80	1.2	19	3 to 8/5 to 40 low-distortion output stage
BFP 196	12	120	1.4	18	3 to 8/30 to 100 low-distortion output stage

Our tightly graded range of devices ensures matching your application as closely as possible to the optimum operating point. (This data is for the SOT143 package. However, devices are available in all packages pictured.)

SCI-1009

INFO/CARD 15

RF calendar

September

22-23 **IEEE Broadcast Symposium**
Washington, DC
Information: Ed Williams, Symposium Chairman, (703) 739-5172.

23-26 **Microwave Update 93**
Atlanta, GA
Information: Jim Davey WA8NLC, 4664 Jefferson Township Place, Marietta, GA 30066.

27-30 **15th Annual Electrical Overstress/Electrostatic Discharge Symposium**
Orlando, FL
Information: 1993 EOS/ESD Symposium, P.O. Box 913, Rome, NY 13442-0913. Tel: (315) 339-6726. Fax: (315) 339-6793.

October

4-6 **15th International Electronics Manufacturing Technology (IEMT) Symposium**
Santa Clara, CA
Information: Electronic Industries Association, 2001 Pennsylvania Avenue, NW, Washington, DC 20006-1813, Mark V. Rosenker, Vice President, Public Affairs. Tel: (202) 457-4980. Fax: (202) 457-4985.

11-14 **Fourth European Conference on "Radio Relay Systems"**
Edinburgh, UK
Information: ECRR93 Secretariat, Conference Services, IEE, Savoy Place, London EC2R 0BL, United Kingdom. Tel: 071 240 1871. Fax: 071-497 3633.

12-14 **1993 Vehicle Navigation and Information Systems Conference**
Ottawa, Canada
Information: Hugh Reekie, Secretary, VNIS '93 Steering Committee, Box 3083, Station D, Ottawa, Ontario, Canada, K1P 6H7.

12-14 **Second International Conference on Universal Personal Communications: The Conference on Personal Communications Systems**
Ottawa, Canada
Information: Celia Desmond, Bell Canada, 800 Bay Street, 4th Floor,Toronto, Ontario K1G 3J4, Canada. Tel: (416) 353-4080. Fax: (416) 920-6689.

19-21 **RF Expo East**
Tampa, FL
Information: Renae Fierros, Cardiff Publishing, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 770-0253.

26-30 **Second International Conference on Signal Processing**
Beijing, P.R. China
Information: Prof. Yan Baozang, Institute of Information Science, Northern Jiaotong University, Beijing 100044, China.

31-3 **IEEE Ultrasonics Symposium**
Baltimore, MD
Information: Electronic Industries Association, EIA Components Group, 2001 Pennsylvania, Avenue N.W., Washington, DC 20006-1813. Tel: (202) 457-4930.

Radar Cross Section Reduction
October 26-29, 1993, Atlanta, GA
Computational Electromagnetics: The Finite Element Method, The Finite-Difference Time-Domain Method
October 26-29, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Wireless Personal Communications Services
September 15-17, 1993, Los Angeles, CA
Advanced Communication Systems Using Digital Signal Processing
November 29-December 3, 1993, Los Angeles, CA
RF and Microwave Circuit Design I: Passive and Linear Active Circuits
November 29-December 3, 1993, 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.

Electrostatic Discharge (ESD) in Integrated Circuits
October 18-19, 1993, Burlingame, CA
Information: Continuing Education in Engineering, University Extension, University of California, 2223 Fulton St., Berkeley, CA 94720. Tel: (510) 642-4151. Fax: (510) 643-8683.

Analyzing Communication System Performance
September 13-15, 1993, San Diego, CA
Digital Transmission for the Cellular Telephone
September 15-17, 1993, San Diego, CA
Global Positioning System: Principles and Practice
September 15-17, 1993, San Diego, CA
Future Telecommunications for Providers, Suppliers, Users, and Regulators
September 20-22, 1993, Washington, DC
Transmission Systems for Telecommunication
September 27-30, 1993, Washington, DC
Mobile Communication Engineering
October 6-8, 1993, Washington, DC
Modern Digital Modulation Techniques
December 13-17, 1993, Orlando, FL
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Navstar/GPS
September 22-24, 1993, Boston, MA
Modern RF & Microwave Techniques
October 26-29, 1993, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Boulevard, M/S C-752, Encino, CA 91436. Tel: (818) 995-6335. Fax: (818) 995-2932.

Basic Electrostatic Discharge Seminar
September 26, 1993, Lake Buena Vista, FL
Information: EOS/ESD Association, 200 Liberty Plaza, Rome, NY 13440. Tel: (315) 339-6937. Fax: (315) 339-6793.

High Speed Communication Networking
September 27-October 1, 1993, Cambridge, UK
Adaptive Signal Processing, with Applications to Communications, Radar, Control, Fiber Optics, and Neural Networks
September 27-October 1, 1993, Cambridge, UK
Personal Communication Networks: Cellular Systems, Wireless Data Networks, and Broadband Wireless Access
September 27-October 1, 1993, Cambridge, UK
Maintaining Signal Quality in High-Speed Digital Systems
September 27-October 1, 1993, Cambridge, UK
Fiber Optic Communication Technology, Systems and Networks
September 27-October 1, 1993, Cambridge, UK
RF and Microwave Circuit Design: Linear and Non-Linear (Theory and Applications)
November 15-19, 1993, Cambridge, UK
Modern Digital Modulation Techniques
November 17-19, 1993, Cambridge, UK
Far-Field, Compact & Near-Field Antenna Measurement Techniques
March 21-24, 1994, Switzerland
Aspects of Modern Military and Commercial Radar
March 21-25, 1994, Switzerland
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

Electromagnetic Compatibility Engineering
October 13-15, 1993, Longmont, CO
Information: Henry Ott Consultants, 48 Baker Road, Livingston, NJ 07039. Tel: (201) 992-1793. Fax: (201) 533-1442.

RF/MW Small Signal/Low Noise Amplifier Design
November 8-9, 1993, Indianapolis, IN
RF/MW Circuit Design I
November 29-December 3, 1993, Los Angeles, CA
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA 94022. Tel: (415) 949-3300. Fax: (415) 949-4400.

DSP Without Tears
September 20-22, 1993, Raleigh, NC
October 4-6, 1993, Scottsdale, AZ
October 18-20, 1993, Chicago, IL
November 8-10, 1993, Houston, TX
November 15-17, 1993, San Jose, CA
Advanced DSP With a Few Tears
November 18-19, 1993, San Jose, CA
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967-5034, (404) 664-6738. Fax: (404) 442-1210.

EMC Emission and Immunity Testing for Compliance in Europe
September 13, 1993, Newark, NJ
Information: Haefly, Inc., Dawne Fay, 2616 Morse Lane, Woodbridge, VA 22192. Tel: (703) 494-1900. Or, Rohde & Schwarz, Terry Palmer, 4425 Nicole Dr., Lanham, MD 20706. Tel: (301) 459-8800, ext. 252 or 254.

15th Piezoelectric Devices Conference Program Announced

The 15th Piezoelectric Devices Conference and Exhibition (PDC&E), sponsored by the Electronic Industries Association (EIA), will again be held in Kansas City, MO, September 21-23. The schedule of technical papers, tutorials and end-user sessions will include the following presentations:

Session A, Technical Papers — Topics include piezoelectric materials, various manufacturing process techniques, oscillator design, frequency references, SAW device packaging, basic and advanced crystal filter design, and measurement methods. These papers are directed to advanced users of piezoelectric products as well as engineers designing those products.

Session B, Tutorials — These sessions are directed to engineers and production personnel at companies which manufacture piezoelectric devices. Subjects scheduled include a tutorial on SAW resonator technology. In addition, crystal resonator, oscillator design and filter topics are planned.

Session C, End-Users — This track addresses both sides of the supplier/end-user relationship. A full-day tutorial on "Electronics and Piezoelectric Technology for Beginners and Non-Engineers" defines the role of the various products in the marketplace. Other papers cover basic oscillators, SAW device and ceramic filter applications, plus monolithic crystal filters. A panel of end-users is also scheduled, discussing their requirements for future products.

In addition to the above presentations, the conference also includes a product exhibition. For information on attending the 15th PDC&E, contact the EIA at (202) 457-4985.

NIST Seeks Calibration Experts — Assessors for a new program to accredit calibration laboratories are needed by the National Institute of Standards and Technology. Engineers, scientists, metrologists and technical experts in industry, universities and government with experience in calibrations and laboratory management are required to conduct on- site assessments of laboratories that perform calibration testing services. The National Conference of Standards Laboratories (NCSL), an association of more than 1100 organizations, asked NIST to develop the activity under the National Voluntary Laboratory Accreditation Program (NVLAP). The expert assessors will be under contract to NIST to help NVLAP establish the competence of laboratories that apply for accreditation. Experienced individuals should send a resume to: C. Douglas Faison, Operations Program Manager, NVLAP, A162 Bldg. 411, National Institute of Standards and Technology, Gaithersburg, MD 20899-0001; tel. (301) 975-4016, fax. (301) 926-2884.

Call for Papers: NPC94 — The 1994 Conference on Networks for Personal Communications is sponsored by the New Jersey Coast Section of the IEEE, and will be held in Long Branch, NJ, March 16-18, 1994. Authors are invited to submit papers on all topics related to telecommunications network support of Personal Communications Services. Extended abstracts should be submitted by September 15 to Technical Program Chairman, Joseph Rizzo, Exec. Dir., PCS, Room NVC-5A225, 100 Schultz Drive, Box 7050, Red Bank, NJ 07701-7050.

RFID Standard Sought for Small Animals — The AIM USA RFID Committee is developing a standard for small animal RFID. Small animals are defined as companion animals kept primarily for personal enjoyment. AIM seeks a standard so that the veterinarians, animal hospitals, humane societies and pet owners will be able to reliably use RFID tags, regardless of their manufacturer. Companies interested in participating should immediately contact the AIM USA Technical Department at 634 Alpha Drive, Pittsburgh, PA 15238-2802, tel. (800) 338-0206.

ARFTG Call for Papers — The 42nd conference of the Automatic RF Techniques Group (ARFTG) will be held December 2-3, 1993 in San Jose, Calif. The conference will focus on "RF and Microwave Testing for Commercial Applications," and a Call for Papers has been issued. Suggested topics are: optimization of test plans, design for testability, calibration and verification, fast test algorithms, fixturing, and others. Submit proposals by September 3 to John T. Grubb, Hewlett- Packard Co., MS: 1URM, 1212 Valley House Dr., Rohnert Park, CA 94928-4999.

VXIbus Consortium HQ Established — The VXIbus Consortium has established the position of Administrator and now has one central office. The Administrator position allows the members of the Consortium to spend more time on promotion of the VXIbus standard. Inquiries for technical assistance, membership requests, or Manufacturer's Identification Number should go to: VXIbus Consortium, 8380 Hercules, Suite P3, La Mesa, CA 91942; tel. (619) 697-6650, fax. (619) 697-5955.

Group Studies CVD Diamond — Producers and potential users of diamond films created using chemical vapor deposition (CVD) have formed a working group to evaluate thermal conductivity measurement methods. Coordinated by the National Institute of Standards and Technology (NIST), the group was formed to meet the need for an accepted method of measurement. Additional topics to be covered include optical absorption and scattering, plus electrical properties and contacts for electronic applications. A report on a recent NIST workshop is available: "Workshop on Characterizing Diamond Films II," available as PB 93-207157 from the National Technical Information Service, (703) 487-4650.

Report Predicts Antenna Market Growth — A market report from Allied Business Intelligence predicts over 11 percent growth in the U.S. antenna market. GPS antennas lead the way, followed by dishes for VSAT and

INMARSAT/INTELSAT. The full report has detailed information on many antenna types and includes market share data for 57 antenna manufacturers. Contact ABI for more information at (516) 624-3113.

NAB Seeks FCC Help With Local Building Laws — The National Association of Broadcasters (NAB) has asked the Federal Communications Commission (FCC) to take broad preemptive action against state and local laws that unreasonably keep broadcasters and other communications companies from building FCC-licensed transmission towers, or prevent consumers from using rooftop antennas to receive radio and television signals. According to the NAB, these practices amount to "non-federal authorities 'un-licensing' FCC-licensed facilities." NAB also suggested that such restrictive policies threaten to impede development of new terrestrial services such as High Definition Television (HDTV), CD-quality Digital Audio Broadcasting (DAB) and non-broadcast technologies like Personal Communications Services (PCS).

NASA Develops Broadband Patch Antenna — A researcher at NASA's Langley Research Center has developed a method for increasing the bandwidth of a conventional microstrip patch antenna. As reported in *NASA Tech Briefs*, the addition of strips of resistive sheet material at the edges of the patch reduces edge currents, making the edge effects a smoother function of frequency. Bandwidth can be optimized by controlling the shapes, size and resistance of the resistive strips. A working model of such a modified antenna had a bandwidth more than twice that of the same antenna, unmodified.

Dow-Key Acquires Switch Line — Dow-Key Microwave expands its space-qualified switch capabilities with the cash purchase of the Transco Space Qualified Products from Datron Systems. All designs, patents and clean room equipment and test equipment for space programs were purchased. The acquisition makes Dow-Key one of the largest mechanical coaxial switch companies in the industry.

Proxim and PI Systems in Healthcare Mobile Computing Partnership — Proxim, Inc. has announced a business partnership with PI Systems that will result in a family of mobile information products incorporating Proxim's wireless data communications technology. Initially, PI will begin building ProxLink™ radio modules into its Infolio electronic clipboards. The radio-linked system will allow real-time connection to mobile-healthcare workers. Additional communications hardware is planned for other PI mobile hardware platforms.

Recent ISO-9000 Activity — The following companies have recently announced compliance with ISO 9001 international quality standards: Scientific-Atlanta's Network Systems Group (NSG), which includes both the Satellite Communications Division and Mobile Satellite Systems business unit; Linear Technology Corporation's Milpitas, Calif. facility, which manufacturers high performance

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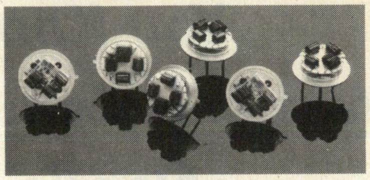
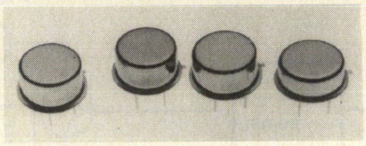
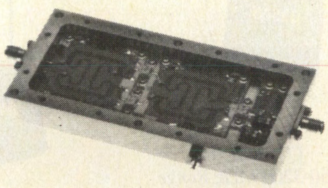
HYBRID *Guaranteed Specifications 25°C*

Model Number	Frequency Range	Gain 25°C	P1dB	VSWR In/Out	Noise Figure	Reverse Isolation	3rd/2nd OOIP	DC Power Volts/mA	Housing Style
QBH-101	5-500	13.0	7.0	1.5:1	2.4	25.0	20/28	15.0/18	TO-8
QBH-102	5-500	12.3	21.0	1.5:1	7.0	23.0	34/50	15.0/93	TO-8
QBH-103	5-300	11.3	22.0	1.5:1	6.8	22.0	37/51	15.0/91	TO-8
QBH-104	5-500	12.3	10.5	1.5:1	4.5	23.0	25/37	15.0/29	TO-8
QBH-105	5-300	12.2	8.0	1.5:1	3.0	27.0	22/30	15.0/18	TO-8
QBH-108	5-300	11.3	20.0	1.5:1	6.5	21.0	37/50	15.0/71	TO-8
QBH-109	10-500	10.6	12.0	1.5:1	4.5	24.0	28/40	15.0/35	TO-8
QBH-110	5-500	15.0	9.0	1.5:1	3.0	25.0	23/33	15.0/29	TO-8
QBH-115	10-500	12.3	26.0	1.5:1	7.8	25.0	35/42	15.0/165	TO-8
QBH-117	5-100	16.5	4.5	1.5:1	1.5	35.0	17/24	15.0/11	TO-8
QBH-118	3-100	16.3	13.0	1.5:1	1.9	35.0	27/38	15.0/21	TO-8
QBH-119	5-500	15.0	12.0	1.5:1	3.0	25.0	26/36	15.0/33	TO-8
QBH-121	10-500	13.5	12.0	1.5:1	3.5	24.0	27/39	15.0/37	TO-8
QBH-122	10-500	17.0	20.0	1.8:1	4.2	22.0	30/38	15.0/65	TO-8 LP
QBH-124	5-100	19.8	17.0	1.5:1	3.5	32.0	30/40	15.0/60	TO-8
QBH-125	10-100	19.6	23.0	1.5:1	4.5	33.0	38/50	15.0/132	TO-8
QBH-126	5-500	15.0	16.0	1.5:1	3.8	25.0	30/38	15.0/50	TO-8
QBH-131	5-1300	18.0	7.0	1.5:1	5.0	27.0	20/35	15.0/41	TO-8 LP
QBH-132	15-700	14.6	16.0	1.7:1	6.5	27.0	29/39	15.0/44	TO-8
QBH-133	10-500	10.3	16.0	1.5:1	4.5	25.0	29/45	15.0/57	TO-8
QBH-136	10-200	20.0	21.0	1.5:1	4.0	26.0	33/45	15.0/70	TO-8
QBH-137	10-200	12.7	21.0	1.5:1	3.5	26.0	38/48	15.0/94	TO-8
QBH-138	5-150	15.5	21.0	1.6:1	3.2	28.0	37/49	15.0/99	TO-8
QBH-147	20-1100	13.5	10.0	1.5:1	3.5	22.0	23/33	15.0/27	TO-8 LP
QBH-149	10-150	23.0	17.5	1.5:1	2.8	30.0	29/39	15.0/39	TO-8
QBH-150	10-300	20.0	18.0	1.5:1	3.5	25.0	30/41	15.0/46	TO-8
QBH-152	10-300	17.0	18.0	1.5:1	3.5	26.0	33/47	15.0/68	TO-8
QBH-154	200-1200	12.8	8.0	2.0:1	2.6	23.0	21/31	15.0/23	TO-8 LP
QBH-155	5-300	15.0	22.0	1.5:1	5.8	28.0	37/50	15.0/93	TO-8
QBH-171	10-150	13.5	27.0	1.5:1	6.5	27.0	40/50	15.0/105	TO-8
QBH-179	5-200	23.3	9.0	1.5:1	3.0	30.0	23/27	15.0/17	TO-8
QBH-180	5-150	29.0	18.0	1.6:1	3.8	50.0	32/42	15.0/59	TO-8
QBH-181	10-200	24.4	16.0	1.5:1	2.8	31.0	25/36	15.0/33	TO-8
QBH-804	10-100	19.8	24.0	1.5:1	4.0	27.0	38/48	15.0/82	TO-8
QBH-822	10-2000	20.0	11.0	2.0:1	5.0	24.0	24/35	15.0/60	TO-8 LP
QBH-824	10-2000	15.3	12.0	1.6:1	6.5	23.0	21/32	15.0/65	TO-8 LP







MODULAR *Guaranteed Specifications 25°C*

Model Number	Frequency Range	Gain 25°C	P1dB	VSWR In/Out	Noise Figure	Reverse Isolation	3rd/2nd OOIP	DC Power Volts/mA	Housing Style
QB-101	2-70	21.9	31.0	1.5:1	4.5	31.0	56/110	24.0/400	19044
QB-188	0.5-100.0	15.0	21.0	1.5:1	3.0	28.0	37/47	15.0/100	184
QB-300	1-300	24.5	22.0	1.5:1	3.8	36.0	37/52	20.0/154	181/182
QB-538	2-500	34.8	22.0	1.5:1	3.0	45.0	35/52	20.0/187	181/182
QB-761	806-870	23.0	18.0	1.5:1	3.5	42.0	32/0	15.0/140	187-2

SURFACE MOUNT *Guaranteed Specifications 25°C*

Model Number	Frequency Range	Gain 25°C	P1dB	VSWR In/Out	Noise Figure	Reverse Isolation	3rd/2nd OOIP	DC Power Volts/mA	Housing Style
QBH-5119	10-500	15.0	12.0	1.5:1	3.0	22.0	26/36	15.0/33	F-PACK .450 SQ SMD
QBH-5122	10-500	17.0	20.0	1.8:1	4.2	22.0	30/38	15.0/65	F-PACK .450 SQ SMD
QBH-5237	10-200	12.7	22.0	1.8:1	4.5	15.0	38/50	15.0/97	F-PACK .450 SQ SMD
QBH-5271	10-150	13.2	26.0	1.7:1	6.0	15.0	39/48	15.0/148	F-PACK .450 SQ SMD
QBH-5284	10-100	19.8	22.0	1.5:1	4.0	21.0	38/48	15.0/82	F-PACK .450 SQ SMD

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INFO/CARD 21

linear integrated circuits; Micro Networks, a supplier of data conversion components, custom hybrids and precision oscillators; John Fluke Mfg. Co., for its North American service and manufacturing facilities; and, AMP Incorporated's AMP Interconnection Components and Assemblies Products Group, which makes connectors and interconnection systems.

Wireless Modem Deal for Cincinnati Microwave and McCaw — Cincinnati Microwave, Inc. announces development of a series of wireless packet data modems utilizing Cellular Digital Packet Data (CDPD) networks. The company also announces that it is teamed with McCaw Cellular Communications, Inc. to market CDPD products and services to corporate data users.

Atlantic Microwave Moves — Atlantic Microwave Ltd. has moved to its new distribution center at 40A Springwood Drive, Braintree, Essex CM7 7YN England. The new telephone number is +44 0376 550220 and their fax is +44 0376 552145.

Larger Facilities for American Superconductor — New, larger facilities for American Superconductor Corp. are located at Two Technology Drive, Westborough, MA 01581. Telephone and fax numbers are (508) 836-4200 and (508) 836-4248, respectively.

Dataradio/IBM Partnership for Mobile Data — Dataradio announces that it has been selected as an IBM business partner to provide mobile data systems for the public safety market. The company's Vehicular Information Series (VIS) product line will be used in law enforcement, fire, emergency medical and other vehicles.

Steinbrecher Corp. Relocates — Steinbrecher Corporation announces completion of its move to 30 North Avenue, Burlington, MA 01803-3398. Their new telephone number is (617) 273-1400, and the fax number of (617) 273-4160.

Richardson Electronics Canada Ltd. Moves — As of September 6, 1993, Richardson's Candian sales and distribution operations are consolidated into a new larger location occupying a 7,350 square foot facility. The new address is: Richardson Electronics Canada Ltd., 6185 Tomken Road, Units 3-5, Mississauga, Ontario L5T 1X6. The component sales telephone is (416) 795-6300 or (800) 348-5580 and the fax number is (416) 795-6350. Readers: note that the area code will change to (905) after October 4.

Siemens Subsidiary in Digital Data TV Venture — U.S. based Wavephore, Inc. and Telecommunal AG, the cable and satellite subsidiary of Siemens, have announced an agreement to market Wavephore's Video 7500 System in Europe. The system transmits high speed digital data at 384 kbps within an active analog television signal, increasing the communications capabilities of existing satellite, microwave or broadcast transmission facilities. First application is distribution of subdivision and utility maps to offices of architects, engineers and contractors.

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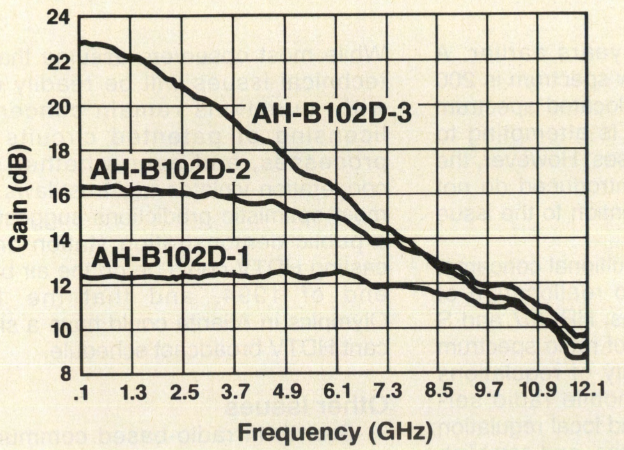
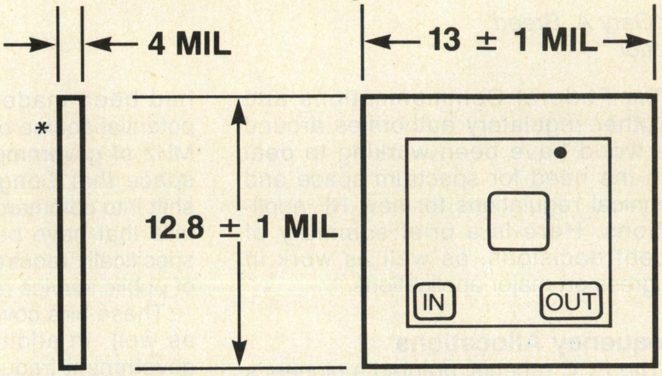
AH-B102D-1,-2.-3 Gain vs Frequency



AH-B102D-1 Chip



*Back of chip is ground

Note: Single dot indicates -1 part

Although the circuit diagram is the classical Darlington amplifier, it has been implemented as a MMIC (Microwave Monolithic Integrated Circuit) using HBT (Heterojunction Bipolar Transistor) technology on GaAs (Gallium Arsenide). This change from conventional Silicon Bipolar technology results in a greatly increased Gain-Bandwidth product. Since GaAs material is insulating, all of the circuit connections are on the surface. This amplifier uses a via to connect the circuit ground to the back of the chip (rather than the back contact being the common collector point).

All that is required for use is to die attach to ground and connect the input and output. The positive bias is connected to the amplifier through the output terminal as in a conventional Darlington amplifier.

The gain is determined by feedback resistors that are part of the chip; therefore, there are different dash numbers for the various gain versions. These devices are also available in standard 70 mil, 4-lead packages.

Typical Performance Characteristics (Ta = 25°C)

AH-B102D-1 (-70C)*	
Gain (dB)	12
Bandwidth (± 0.5 dB, GHz)	7
Bandwidth (3 dB, GHz)	10
P1 dB (dBm)	13
Reverse Insertion (dB)	17
VSWR, Input (≤ 7 GHz)	2.0:1
VSWR, Output (≤ 7 GHz)	2.0:1
Noise Figure (dB) (≤ 7 GHz)	6.5

* Commercial ceramic package data

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INFO/CARD 23

Update on Regulatory Activity

By Gary A. Breed
Editor

The Federal Communications and other regulatory authorities around the world have been working to deal with the need for spectrum space and technical regulations for new RF applications. Here is a brief summary of recent decisions, as well as work in progress on major applications.

Frequency Allocations

The FCC recently granted a pioneer's preference license to Mobile Telecommunications Technologies, Inc. (MTEL) for their Nationwide Wireless Network paging system. Three 50 kHz wide channels in the 930-931 MHz band will be used for a national paging system which utilizes a 24 kbps data rate to transfer large amounts of data (such as messages or even still picture video) quickly. The system includes a talk-back feature which allows individual pagers to confirm receipt of messages, or to send a brief reply.

Additional spectrum for the mobile satellite service has been proposed. 1530-1544 and 1626.5-1646.5 MHz would be co-primary service with the maritime mobile service, and 1525-1530 would be a primary allocation. These frequencies are for geostationary-based services, not for low earth orbit (LEO) uses. This proposal is based on a petition by American Mobile Satellite Corp. and Inmarsat.

Other major issues being discussed in the frequency allocation realm include final definition of the L-band segment set aside for Personal Communications Service (PCS) applications. The first rules to establish which particular frequency blocks will be assigned to which types of services are expected to be released in early 1994. Several major electronics firms are expected to introduce PCS products for the American market immediately following formal FCC authorization.

Also under discussion is crowding among public service users in the mobile radio bands. A recent report in *Communications* magazine noted that by 1992, public service radio users had exceeded by 70 percent projections that had been made ten years earlier. A potential source of new spectrum is 200 MHz of government-allocated spectrum space that Congress is attempting to shift into commercial uses. However, the bills that have been introduced do not specifically require attention to the issue of public service radio.

These bills cover additional concerns, as well. In addition to reallocation of government frequencies, HR 707 and S 335 address licensing of radio spectrum through auctions, parity of regulations among commercial mobile radio services, preempt state and local regulation of these wireless services, and establish deadlines for completion of PCS dockets at the FCC. The House bill has been unanimously approved by the Commerce Committee.

LEO Satellite Rules

So called "Big LEO" rules are due in October. The FCC's semiannual regulatory agenda indicates the Commission's intention to release final rules by that date, concerning low earth orbit (LEO) satellites operating above 1 GHz. The rules will be based on a negotiated rulemaking between various applicants that ended in March with concurrence on several major issues, including spectrum sharing.

Also on the agenda for Fall 1993 are the first rules concerning digital audio broadcasting (DAB) and a request by Norris Satellite Communications for reallocation of 20-30 GHz frequencies for a General Satellite Service, along with its request for a pioneer's preference license.

The Ongoing Story of HDTV

Pending technical rules for high definition television (HDTV) have been set aside once again, as this past spring's on-air testing of the competing systems provided inconclusive results. After that round of evaluations, the Japanese withdrew its analog-based transmission system from consideration. The FCC then asked the remaining U.S. applicants to get together and decide on a single system for terrestrial transmission of HDTV. While most observers assume that the technical issues will be readily dealt with, questions remain concerning licensing of patented circuits and processes, and even whether such cooperation violates anti-trust laws. The most optimistic predictions suggest that a public demonstration station broadcasting HDTV could be on the air by the end of 1994, and that the 1996 Olympics in Atlanta could have a significant HDTV broadcast schedule.

Other Issues

Growth in radio-based communications and control systems is creating a regulatory logjam. To work around the problem, many initial product offerings operate in the unlicensed industrial, scientific and medical (ISM) bands under the provisions of Part 15 of the FCC Rules and Regulations. However, it is assumed that these bands will rapidly become overcrowded if the trend continues.

The increased proliferation of radio frequency devices poses another problem that may ultimately require a regulatory solution, interference — both to other devices, and received interference. The European Community has included interference immunity standards in its electromagnetic compatibility (EMC) standards that will take effect in 1996. In the U.S., the FCC and the Standards Committee of the IEEE EMC Society has been studying susceptibility to interference, but no standards are in place. In order for an increased number of radio devices to operate reliably, such standards may be required. Many proposed products will be low-cost consumer items, which may not incorporate good EMC practices in their design unless required to do so. Other applications are for commercial and industrial users, which may involve harsh EMC environments. *RF*

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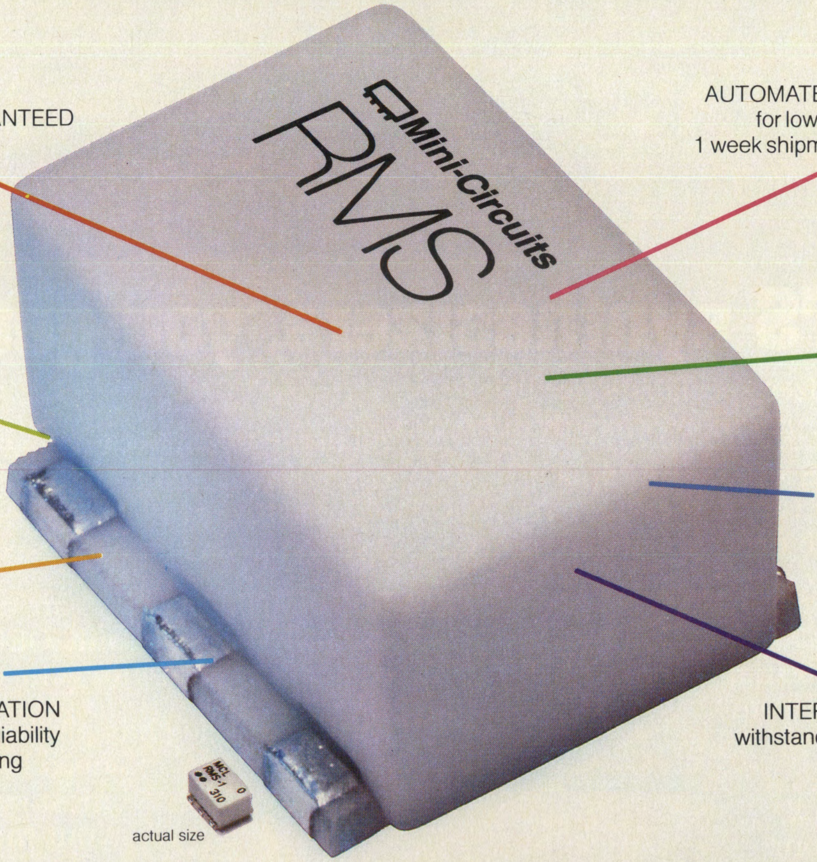

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INFO/CARD 24

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INFO/CARD 25

Using Aliased-Imaging Techniques in DDS to Generate RF Signals

By Allen Hill and Jim Surber
Analog Devices, Inc. Computer Labs Group

Direct digital synthesis (DDS) is an increasingly popular means of generating highly accurate, and harmonically pure digital representations of signals. Effectively, the architecture digitally divides the phase of a highly accurate clock by a digital tuning word (typically 32-bits) and uses a lookup table to convert the phase information to the desired frequency and signal output format. This conversion is accomplished with either a RAM lookup table or a direct phase-to-sine amplitude conversion stage. The latter type of DDS architecture will generate only a sine wave output, and may sometimes be referred to as a "numerically controlled oscillator", while the former may be programmed to generate any arbitrary waveform desired by the user (triangle, sawtooth, etc.). The digital output of a DDS circuit is usually reconstructed with an accurate high-speed D/A converter to provide an analog output signal. The inherent advantages in the DDS waveform generation technique is in its high degree of frequency resolution, frequency agility, and in the spectral purity of its digital output. DDS is also gaining popularity because of its computer-compatibility and ease of system design afforded by the large scale of functional integration contained in the DDS chip.

Figure 1 shows a block diagram of the Analog Devices AD9955 direct digital synthesizer chip which utilizes a phase-to-sine amplitude converter stage to generate a sinusoidal digital output. This highly-integrated DDS chip will accept a 32-bit tuning word and will generate a 12-bit sine output with a spurious-free dynamic range of >90 dB. This architecture uses a quarter wave sine/cosine computation circuitry, which takes advantage of the symmetry of a sine wave, to reduce the chip complexity and size. The output frequency resolution of this type of architecture is determined by the formula: clock/2^{32}.

The main function of DDS circuitry is to generate a high-speed output signal but there are several considerations that determine the maximum frequency that can realistically be synthesized by a DDS. Traditional Nyquist sampling theory dictates that at least two samples per cycle are required to accurately reconstruct a fundamental output frequency of interest. Under this convention, the maximum frequency that could be synthesized by a DDS system would be Clock/2. One factor that results from "violating" the Nyquist theorem is the appearance of aliased images which are generated as an interactive function of the clock frequency and the fundamental output frequency of the DDS. To get a clear understanding of this phenomenon, please refer to the model of the spectral output of a sampled system, as shown in Figure 2.

As depicted, the first aliased image appears at $F_{clock} - F_{fundamental}$ and additional images will appear at $(N)F_{clock} \pm F_{fundamental}$, where N=any integer. At fundamental output frequencies above clock/2, the first image response will fall within the DC-fundamental output bandwidth and it cannot be filtered out with a traditional anti-aliasing filter response. The rolloff in amplitude of the aliased images (gray area in Figure 2) is due to the SINX/X response of the quantized output of the DDS and the reconstruction DAC. The glitch energy and non-linearity errors contained in the DAC's transfer function, will appear as harmonics and spurious energy in the output spectrum. This energy does not have the SINX/X rolloff characteristic and will

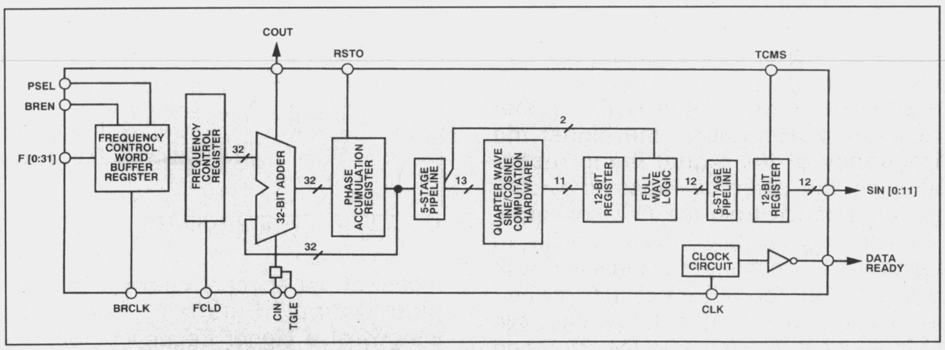

Figure 1. Block diagram of DDS device.

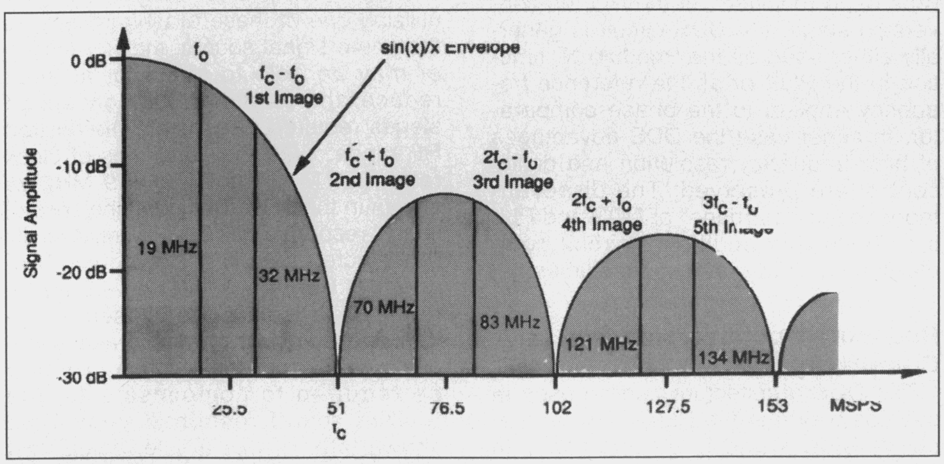

Figure 2. Spectral plot of sampled output.

generally be much lower in amplitude than the aliased images.

System clock noise is also a factor that limits the performance of high-speed DDS circuitry. As previously explained, standard sampling theory dictates that the clock must be two times the output frequency to avoid frequency aliasing. This means that high analog output frequencies require higher clock rates. However, high clock frequencies are difficult and expensive to generate and they are notorious for generating system noise that is virtually impossible to subdue. Clock noise is always present in the analog output and the fidelity of the DDS is directly related to the level of noise reduction techniques used in its circuit layout. Clock frequencies above 100 MHz are extremely difficult to implement and control which places a practical limit on the upper output frequency limit of a DDS system, under traditional Nyquist criteria.

Another limitation to using DDS architecture to synthesize high-speed RF signals is the relatively poor spectral purity of the signal reconstructed by the D/A converter, versus that of a traditional oscillator-generated analog RF signal. It is not uncommon to lose more than 20 dB of spurious-free dynamic range in the DDS-generated signal at the D/A converter stage, due to differential non-linearity and integral non-linearity errors, and glitch energy in the DAC transfer function. Furthermore, the higher the frequency of the signal being reconstructed, the lower the expected dynamic performance from the D/A converter. This has necessitated the technique of using a DDS circuit, in conjunction with phase-locked loop (PLL) circuits, or heterodyning mixer stages, to increase the output signal frequency to the RF range required for local oscillator stages in wide-band receivers. In the PLL up-conversion stage, the DDS circuit is generally either used as the "modulo N" function in the PLL, or as the reference frequency applied to the phase comparator. In either case the DDS advantages of high frequency resolution and digital control are preserved. The disadvantages to this "marriage" of DDS and PLL architecture is additional circuit complexity and slow loop response times.

Reducing Upconversion Complexity

The DDS architecture can be used in such as way that the PLL, or mixer, upconversion stage is unnecessary in some applications. As was shown in the

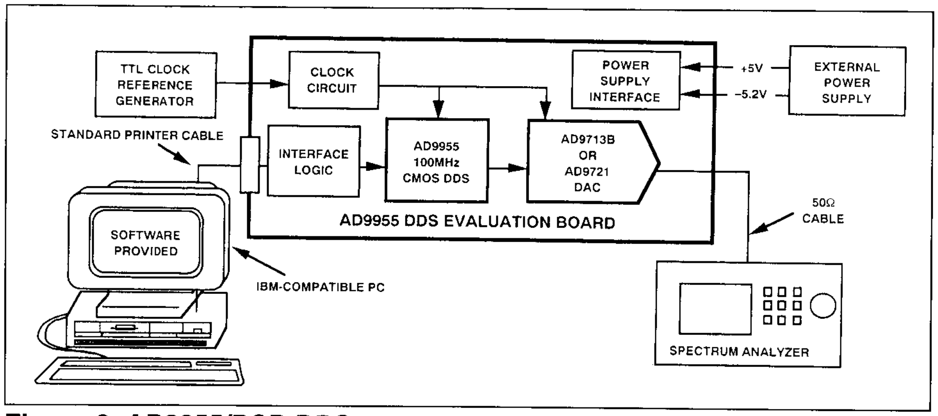

Figure 3. AD9955/PCB DDS test system block diagram.

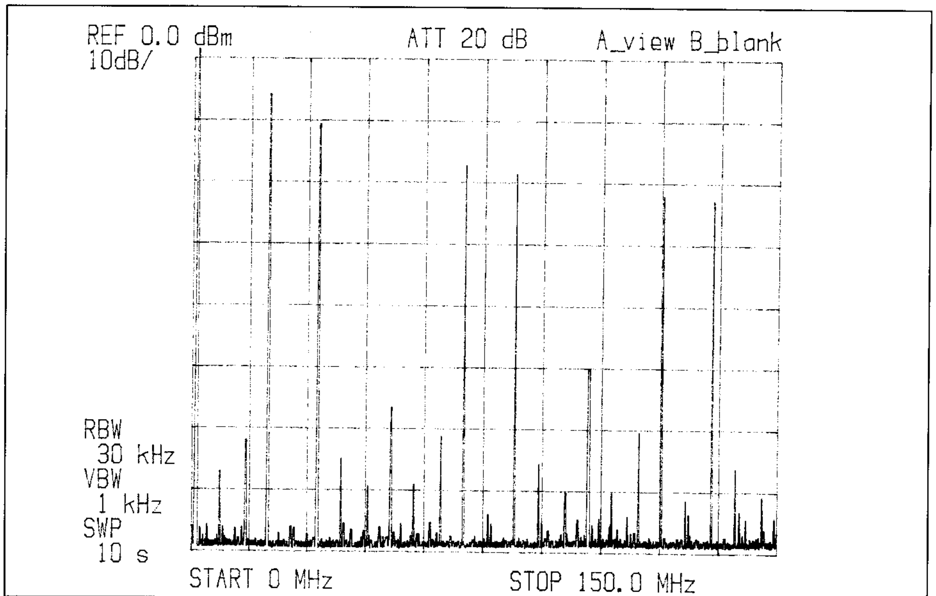

Figure 4. AD9955/PCB wideband spectral plot.

model of the spectral output of a sampled system (Figure 2), there are images of a higher frequency than the fundamental that appear at $(N)F_{clock} \pm F_{fundamental}$, where N = any integer. By utilizing one of these aliased images as the prime signal source, the RF designer may be able to avoid, or at least reduce, the additional up-conversion stages required to generate the desired RF range. With the conditions of clock rate = 51 MHz and F_{out} = 19 MHz as shown in the spectral model, the second image response of 70 MHz could easily be selected as the desired output frequency. Bandpass filtering would be required to isolate the aliased image from the adjacent harmonic and spurious signals and amplification may also be required to compensate for the SINX/X rolloff response. As will be shown later in this article, actual test results from a DDS/DAC combination operating under these conditions can yield a 70 MHz sine wave that has a spurious-free dynamic range of 66 dB, over a frequency span of at least 500 kHz. The amplitude of the 70 MHz aliased image will be about −17 dBm.

Real-World Test Results

For observing the output of a DDS system and characterizing the potential of the aliased-imaging technique, we used the Analog Devices AD9955/PCB, DDS evaluation board, as our test platform. This evaluation board is a complete, and fully PC-compatible, DDS system which greatly facilitates benchtop experimentation. The block diagram of the AD9955/PCB, and our test setup, is shown in Figure 3.

We selected a 10-bit D/A converter, the AD9721BBN, as the reconstruction DAC for the DDS system. This DAC gives the highest dynamic performance

available at frequencies above 20 MHz. The output was observed on a spectrum analyzer which had the necessary filtering and frequency resolution features to fully analyze the aliased images. The following plot of a spectrum analyzer display of the test setup output, shows the detailed spectral analysis of the second aliased image at 70 MHz. Figure 4 shows the wideband spectral plot from DC to 150 MHz which verifies the presence and amplitudes of the output fundamental, clock, aliased images, spurious signals, harmonics, and noise floor, of the test setup. The locations of the various frequency components in the output spectrum are exactly per the model.

The next series of plots show the spurious-free dynamic range of the second aliased image as the fundamental is swept over the range of 19 MHz ±250 kHz, in five increments. These plots verify that the aliased image has an SFDR of greater than 66 dB within this frequency span. This technique of sweeping the fundamental through a frequency band, provides a measure of the range of frequency agility that is available to the aliased image, while maintaining a minimum level of SFDR. This level of dynamic performance at 70 MHz is generally as good as that which is afforded by adding an up- conversion stage to the DDS, and operating under similar conditions.

Phase Noise Performance

System designers who are using DDS to generate carrier frequencies are generally interested in spurious signals that is very close in to the carrier frequency. Unfortunately the photographs of the spectrum analyzer output (refer to Figures 5a-e) indicate a very wide spread of spurious energy at the carrier. The unusually high level of spurious energy, manifested in the width of the carrier on the spectrum analyzer display, is mostly due to jitter in the local oscillator of the spectrum analyzer itself; it is not generated within the DDS system. The phase noise of a carrier synthesized in a DDS system, is inherently low. It is actually determined by the phase noise of the system clock and, as will be revealed, the DDS output actually contains less phase noise than the clock itself. The reason for the lower phase noise in the DDS output is straightforward. The function of the DDS architecture is to perform a divide-by-n function on the clock. The clock's phase jitter is also subjected to the divide-by function and will appear reduced on the output of the DDS.

In order to verify the level of spurious energy close-in to the carrier, we actually measured the phase noise of the output of the AD9955 DDS chip at various frequencies. This measurement was accomplished in the evaluation board test setup with a 50 MHz clock, generating frequencies of 7.220 MHz, 9.999 MHz, and 19.780 MHz. The phase noise measurement was performed with a Hewlett Packard 3048A phase noise measurement system. Figure 6 is the plot of the three output frequencies showing the phase noise over a span of up to 10 kHz from the carrier. This measurement technique provides a much more accurate characterization of the close-in spurious energy than the spectrum analyzer display. As can be seen, the phase noise, at a 1 kHz offset from the carrier, is greater than 130 dBc/Hz at all of the selected output frequencies, with this particular clock frequency. This performance characteristic is not directly measurable on a typical spectrum analyzer display. It is also apparent in the plot that the phase noise of the clock is of higher magnitude than that of the output frequencies because of the divide-by function of the DDS, as discussed earlier.

Making it Work

Establishing the proper DDS operating conditions for creating a desired aliased image response requires consideration of several variables. Obviously, aliased images are created by interaction of the clock and fundamental output frequencies so the optimum combination of the two must be determined through calculation and empirical observation. As a basic guideline for selecting the operating points, the following procedure can be followed:

1. Determine the aliased image output frequency of interest and the bandwidth requirement.

2. Select the fundamental frequency, for a given clock frequency, that will generate the desired aliased image. Sketching the spectral plot of $(N)F_{clock} \pm F_{fundamental}$ may be helpful in visualizing and selecting the best combination of clock frequency and fundamental output frequency. For best results, N should not be a multiple of the clock and the fundamental output frequency should not exceed 40 percent of the clock frequency. As a practical rule, lower clock frequencies will generate relatively lower levels of harmonic distortion related to

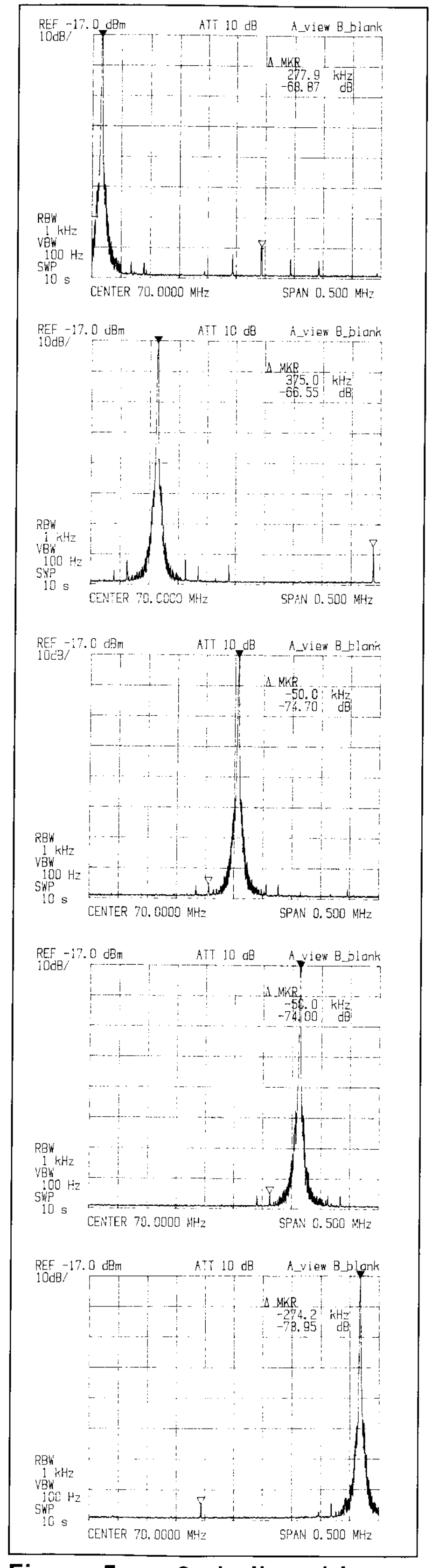

Figure 5a-e. 2nd aliased image swept over 70 MHz ±250 kHz.

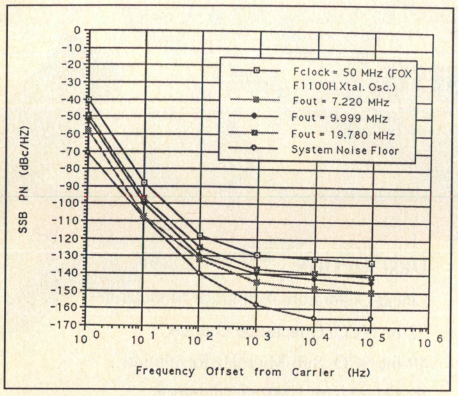

Figure 6. Phase noise performance of the AD9955 DDS test system.

non-linearity and glitch energy in the DAC's output transfer function. As previously discussed, the clock frequency should be as low as possible to minimize system noise problems.

3. Calculate the expected amplitude of the selected aliased image by inserting the clock and aliased output frequency into the following formula:

$$A = [\sin(\pi F_{fundamental}/F_{clock})]/(\pi F_{fundamental}/F_{clock})$$

4. All of the variables that establish an aliased image in a DDS system are interactive. It is best to use a DDS test platform and actually observe the actual spectral output of the selected conditions. Examine the aliased image for SFDR over the desired frequency span and experiment for best results.

The aliased imaging technique offers frequency hopping capabilities that are more restricted than that of the traditional A_{out}<Clock/2 application. However, we did demonstrate that swept-frequency is viable for even aliased frequency carriers. Changes in the output fundamental frequency will result in changes to both the aliased image frequency and its harmonics. Depending on the frequency conditions, this may result in unexpected changes in the width of the usable frequency span of the aliased image.

The amplitude of the aliased image will move along the SINX/X response function which means its amplitude is directly related to frequency. Even small changes in frequency about the fundamental will impose changes in the aliased image. This must be compensated for in frequency hopping applications.

While the aliased imaging DDS application does have limitations, it certainly holds the potential for eliminating one or more up-conversion stages in receiver applications. Careful selection of operating conditions is essential along with a set of valid expectations of the performance results, based on calculation and test results. *RF*

About the Authors

Allen Hill is an applications engineer for the Greensboro, NC division of Analog Devices. He has been with Analog Devices for 12 years. Jim Surber is currently a business development engineer, and has been with the company for 15 years. The authors can be reached at (919) 668-9511 or by fax at (919) 668-0101.

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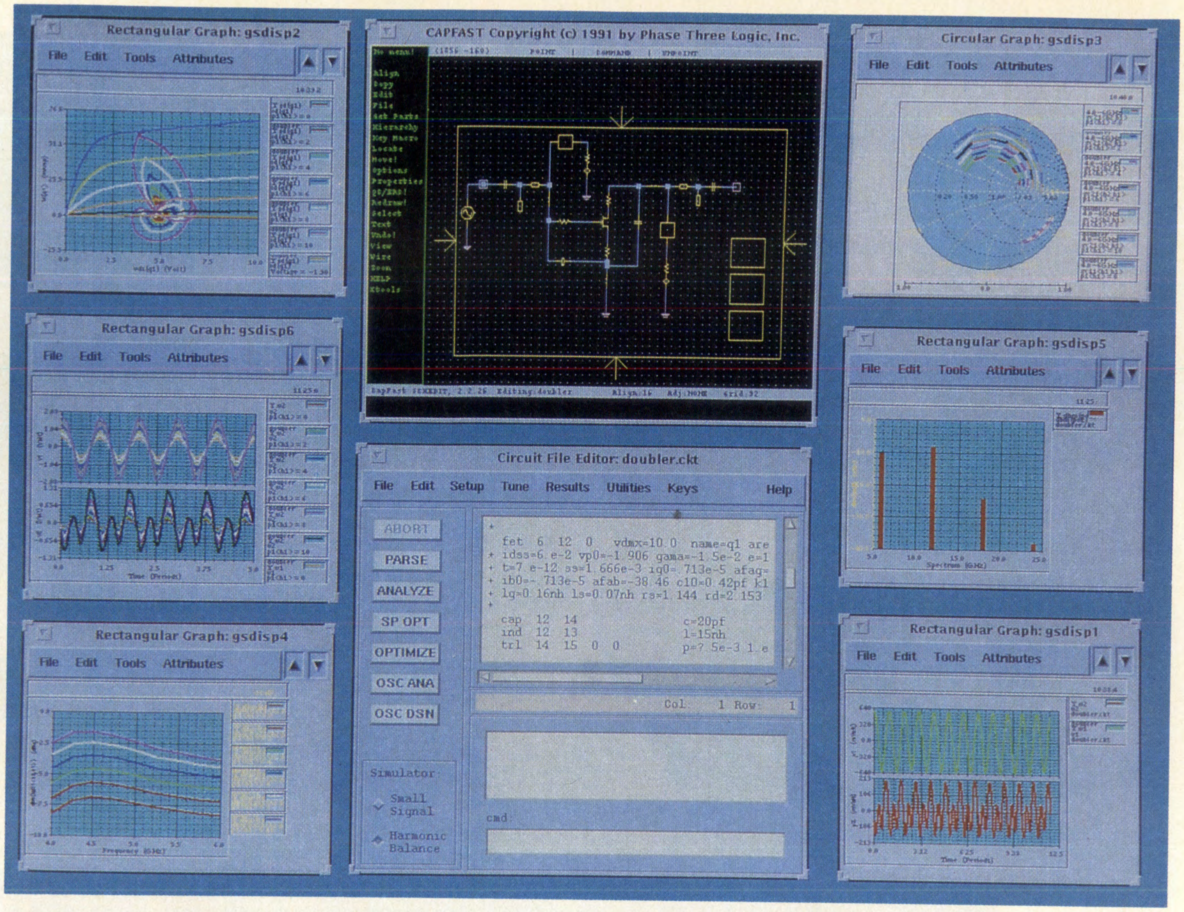

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INFO/CARD 30

Linear Frequency Modulation — Theory and Practice

By Bar-Giora Goldberg
Sciteq Electronics, Inc.

This article addresses high-performance applications for chirp waveforms, and describes the Sciteq DCP-1, an all-digital mechanism for generating chirp signals with a combination of linearity and operating bandwidth not previously achieved.

A linear frequency modulation (FM) signal, or "chirp" is a swept frequency change over some frequency range. Ideally, the rate of change is perfectly constant (time vs. frequency), without ambiguity or anomaly. Originally stimulated by emerging radar requirements to distribute energy across a range of frequencies, virtually all early LFM signals were generated with analog devices such as a voltage controlled oscillator (VCO) or surface acoustic wave (SAW) resonator. Analog ambiguities were introduced by typical RF component and environment uncertainties, and therefore system performance was far from theoretical, but designers had no choice but to attempt to work around these problems. The cost of linearizing and compensating such circuitry in some cases was as expensive as the rest of the RF subsystem, so the high cost of approaching theoretical numbers, and the unavoidable errors caused by analog solutions, made such applications useful primarily for experimentation and scientific exploration.

The advantage of a chirp waveform drove its adoption by certain high performance radar, sonar, and communication systems, particularly in military systems. A chirp signal's energy is spread across a wide band rather than focused at one point, which helps reduce emitter spectral density and makes it harder to jam. Such an approach permits communication or system operation to continue even if part of the spectrum is blocked. Most importantly, it allows high resolution and is therefore applicable to imaging, altimetry, and fuzing where resolution is an important parameter.

In the late 40s and early 50s, as pulse radar technology grew out of its infancy, a problem became visible. Radar range depends on E/N_0, where E is the pulse energy, $E = P \cdot T$, (P is the received power proportional to the transmit power, T is the pulse duration) and N_0 is the receiver noise density. Resolution depends on the signal bandwidth. These two requirements are related diagonally; improving E makes it necessary to increase T but an increase in T reduces the pulse bandwidth and therefore resolution. The most visible solution was to modulate the pulse, which makes the pulse bandwidth dependent upon the modulating waveform rather than the pulse length of (approximately) 1/T. One of the early waveforms suggested for this application was linear FM, and it remains optimum for many radar applications.

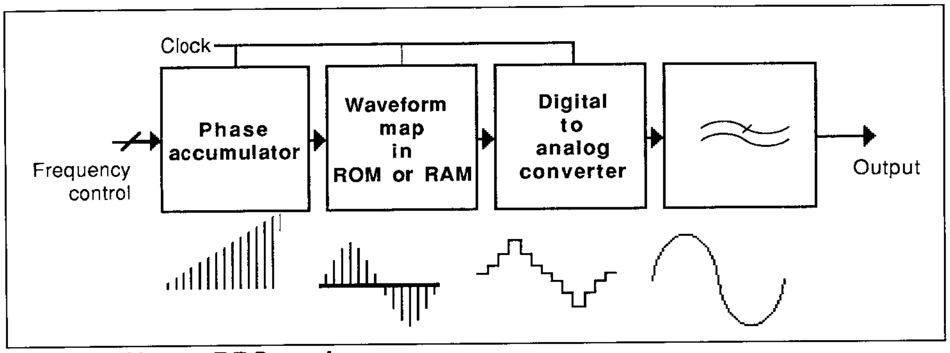

Figure 1. How a DDS works.

Recently, the utility of linear FM in non-radar applications has been recognized. Use of this technique was explored in various other systems including communication, semiconductor process control, sonar, seekers, simulation, and test equipment. In 1989, Sciteq began work on an all-digital technique for producing such systems that would exploit the firm's experience with direct digital synthesis (DDS) technology. DDS (Figure 1) had evolved into a useful signal generation tool, but existing implementations were not appropriate for chirp generation, except at near-audio frequencies for sonar, due to latencies in the designs. Nevertheless, it was clear that the best approach to an idealized chirp generator would exploit some derivative of DDS techniques.

By 1990, key digital and data conversion components had already been developed or were being developed, either by Sciteq or other agencies (Sandia National Laboratories, for one), and for the first time it appeared possible to produce the industry's first wideband digital chirp synthesizer. Accordingly, Army Research Laboratory (then Harry Diamond Labs), let a Small Business Innovative Research (SBIR) Phase I contract to Sciteq to determine the practicality of meeting theoretical goals for the Army's next-generation battlefield surveillance radars. Specifically, the group working on Synthetic Aperture Radar believed that a digital signal generation solution was important to program objectives. Sciteq's interface to this program has been Barry Scheiner of ARL.

The Phase I report was positive and defined a practical path to the desired goal, leading to a Phase II contract award. Though most SBIR projects are considered high risk, the development was successful and first prototypes were delivered during the past year to the Army and to other agencies interested in the exploitation of ideal linear FM signals. The nature of the SBIR program is to permit the contractor to retain control over the new technology, and to commercialize the results where possible, hence the availability of Sciteq's direct-digital chirp synthesizer (DDCS) to the industry. That product, designated the DCP-1, is a DDCS that produces linear FM signals over a band of more than 230 MHz, with linearity approaching an

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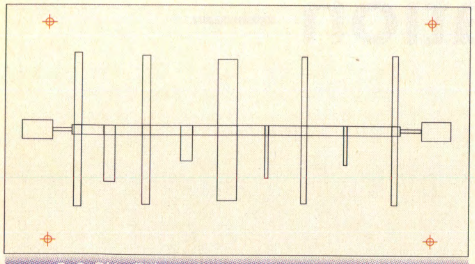
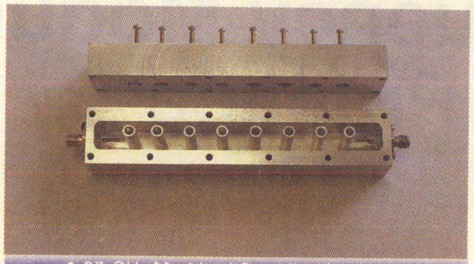
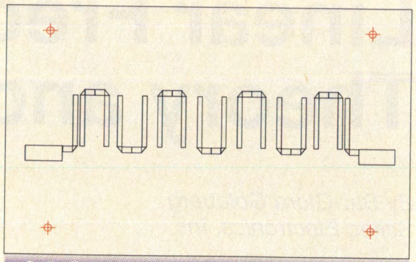
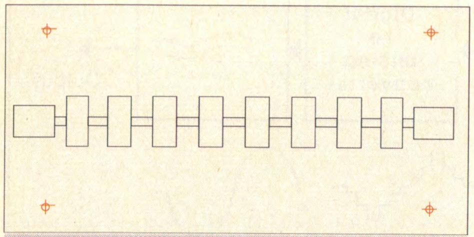
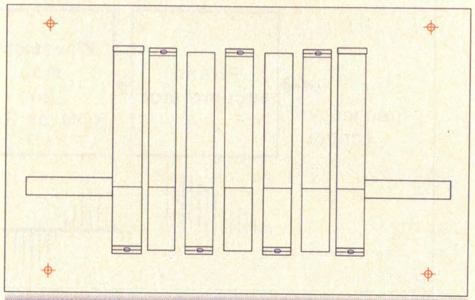
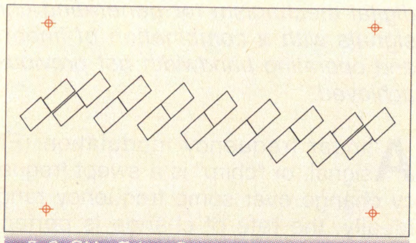

5.6 GHz Elliptic Bandpass on Teflon Board

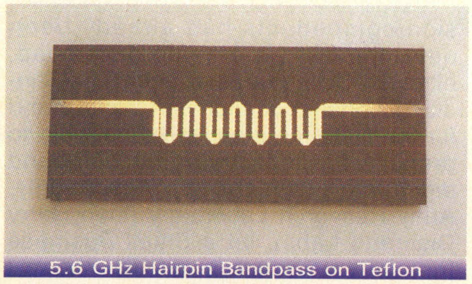

1.27 GHz Machined Combline Bandpass

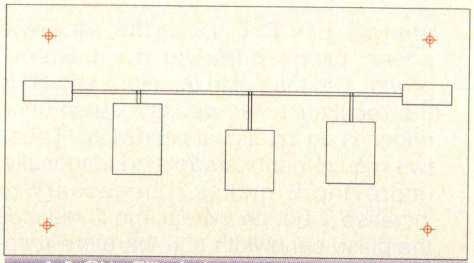

2.2 GHz Hairpin Bandpass on Teflon

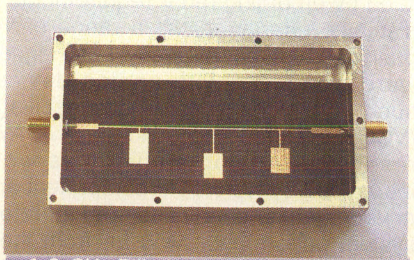

6.4 GHz High-Order Stepped-Z Lowpass


10 GHz Wideband Interdigital BP


5.6 GHz Edge-Coupled BP on Teflon


5.6 GHz Hairpin Bandpass on Teflon


1.2 GHz Elliptic Lowpass on G-10/FR-4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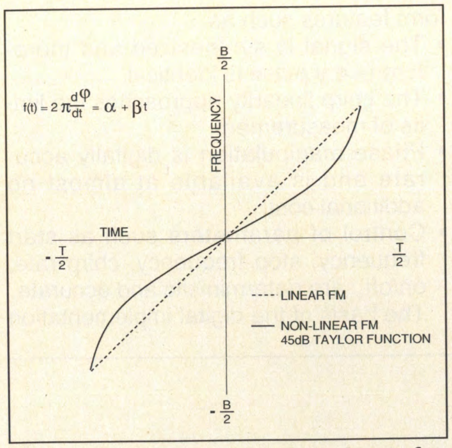

Figure 2. Frequency-time laws for typical linear and non-linear chirp signals.

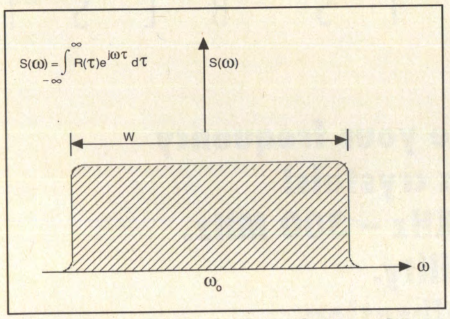

Figure 3. Chirp signal spectrum for W·T>>1.

ideal level previously defined only by mathematics and never actually observed.

Waveform Properties

A substantial amount of theoretical work was done to calculate the properties of such signals, i.e., power spectrum, auto-correlation, etc., and the characteristics of that theoretically perfect LFM signal were defined.

A general description of such a signal is given by:

$$s(t) = A \cdot \sin\left(\alpha t + \frac{\beta}{2}t^2 + \varphi_0\right)$$

and some of its properties are shown in Figure 2.

If the signal bandwidth is designated as W, then the product, W·T = TB, the time bandwidth product for radar applications or processing gain for spread spectrum communications.

For W·T = TB >> 1,, the power spectrum is flat (Figure 3) and the signal energy is distributed almost equally across the bandwidth, W. The signal auto-correlation is given by:

$$R(\tau) = \left(\frac{\sin X}{X}\right)^2$$

where,

$$X = \pi \cdot \tau \cdot W = \pi \cdot \tau \cdot \frac{TB}{T}$$

It can be shown that the signal auto-correlation has relatively high sidelobes with the worst case of –13.5 dB (approximately 20log [sin X/X] for X = 1.5π); see Figure 4. This parameter can be improved by adding the complexity of either amplitude modulation or phase modulation (pulse weighting), and there is an extensive body of literature on this issue. Unfortunately, while there is a great deal of mathematical extrapolation from ideal data, the experimental results were derived from analog mechanisms and therefore were limiting.

Applications

LFM is extensively used for both pulse Doppler radars and synthetic aperture imaging radars. As always, the more linear the LFM signal the better the performance of the system, but even the relatively crude chirps generated by analog circuitry are better than non-chirped operation. With the advent of DDCS technology, with its inherent deterministic linearity, resolution of SAR and performance of other chirp-dependent systems is improved.

In range measurement systems, for altimeters, control, flight control, etc., the principle and the advantages of digital LFM are shown in Figure 5.

Compressive receivers are used to scan the spectrum, and unlike typical spectrum analyzers, which sweep the spectrum and find *all* signals that are within the analyzer bandwidth (during the time of the sweep only), these receivers see all signals that occur within the sweep time. In this respect, a compressive receiver operates like a real time DFT analyzer. An ideal compressive receiver depends upon a linear and fast chirp over the desired band of reception.

In semiconductor production, wafer perfection is a critical issue that influences yield, and therefore has serious economic impact. Surface anomalies and contamination are usually detected using a laser technique, where the laser is reflected off the wafer surface and scatter characteristics evaluated to determine surface characteristics, resulting in a map of the wafer showing imperfections, contamination, and therefore usable dice. In most such applications, the wafer is moved mechanically, and the laser is modulated by a Bragg cell driven by a linear FM signal (produced from an expensively-compensated VCO). The linearity of the FM signal is one of the factors that determines resolution of the system, and also affects the "guard" area around imperfections that are detected. The more accurate the signal, (hence, the more deterministic the laser position), the smaller the guard area and the higher the yield.

Digital LFM Synthesis

Evolution of digital technology has allowed certain Direct Digital Synthesizer (DDS) implementations to operate at sufficient speed to produce bandwidths sufficient for the above applications.

The use of DDS disciplines bring with it cardinal improvements in the waveform features such as:

- The signal is synthesized and therefore every pulse is identical.
- The chirp linearity approaches the limits of measurement.
- Phase manipulation is digitally accurate and is available at almost no additional cost.
- Control of parameters such as start frequency, stop frequency, chirp rate, on/off, are deterministic and accurate.

The basis of the digital implementation

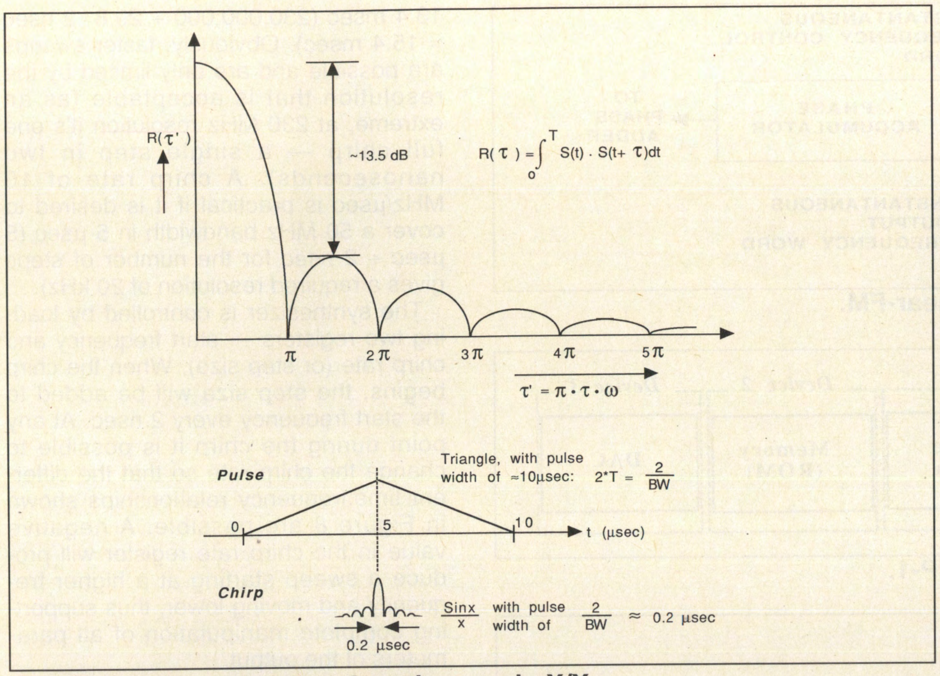

Figure 4. Auto-correlation function as sin X/X.

is shown in Figure 6. The output of this digital process can be represented as:

$$s(t) = \sin\left(\frac{\beta \cdot t^2}{2} + \alpha \cdot t + \varphi_0\right)$$

The general structure is that of a dual accumulator. Since an accumulator is a discrete integrator and we need to generate a quadratic function, two accumulators are necessary.

The output of the first accumulator is the instantaneous frequency and the frequency adder allows the setting of a start frequency. The F output allows the monitoring of the instantaneous frequency. The input of the first accumulator is therefore β in the equation and the frequency input is α. The output of the second accumulator is the signal's phase and therefore can be phase modulated by another adder. This input is equivalent to the term, φ_0, in the equation.

Such structure can be implemented in CMOS technology at clock frequencies up to 100 MHz, and in high speed ECL and GaAs technologies up to 800 MHz, though it must be remembered that the nature of a DDS allows an output frequency of less than half the clock.

In the block diagram, the chirp chip and phase adder (implemented in one device) are followed by a SINE ROM, a DAC (digital to analog converter), and a low pass filter. Because of the nature of the output spectrum (sin X/X) and the group delay of the filter, amplitude and group delay equalization improves the result.

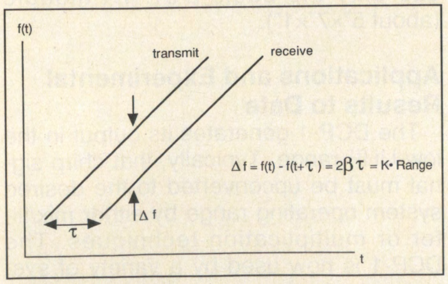

Figure 5. Range meaurement using linear-FM.

Practical Implementation: the DCP-1

The Sciteq model DCP-1 is an all-GaAs direct-digital chirp synthesizer (DDCS) clocked at 500 MHz and therefore generating output frequencies from DC to 230 MHz (limited by Nyquist and the low pass filter considerations). Though the design involves several support functions, the basic chirp generation function is achieved by three devices, a double-accumulator, a memory, and a digital-to-analog converter.

In the double accumulator, both accumulators are 24 bits in size, thus yielding a minimum step size of ≈29.8 Hz. The frequency and phase accumulator functions are integrated into one device, developed by a Sandia National Laboratories program under the leadership of Bruce Walker. The part includes not

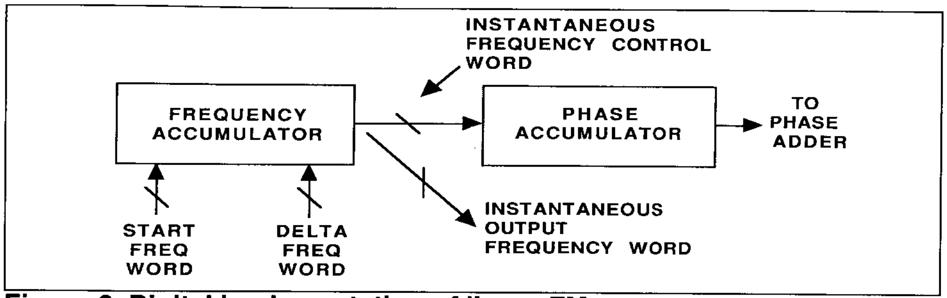

Figure 6. Digital implementation of linear-FM.

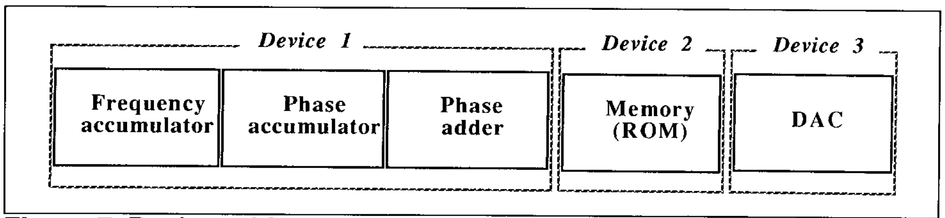

Figure 7. Basic architecture of the DCP-1.

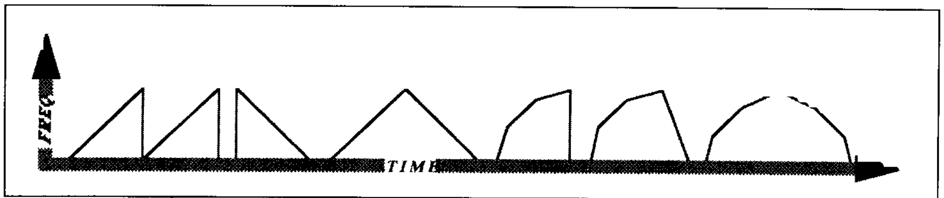

Figure 8. Different time/frequency relationships possible.

only the specified accumulation functions, but also provides 12 bits of phase control plus a time-equalized 8-bit frequency output that supports system timing. The memory device uses a patent-pending algorithm (Sciteq's) to map the phase output of the accumulator to digitally-defined amplitudes, also at a 500 MHz rate or better. The digital output of the memory is considered near-perfect, with digital error supporting a spurious response better than 70 dB below the carrier, so it is the digital-to-analog converter (DAC) that limits the spectral purity of the system. In the initial DCP-1s, the DAC is a 12-bit GaAs part developed by a consortium including GE, Sciteq, Motorola, and Hughes. The DCP-1 basic architecture is shown in Figure 7.

In the LFM mode, the DCP-1 updates frequency at a 500 MHz rate, which means that a new frequency is synthesized every two nanoseconds. The slowest chirp rate is ≈15 kHz/µsec (≈30 Hz times 500, since there are 500 steps in each microsecond). For a full band sweep, this would take approximately 13.4 msec (230,000,000 ÷ 29.8 2 nsec = 15.4 msec). Obviously, faster sweeps are possible and are only limited by the resolution that is acceptable (as an extreme, at 230 MHz resolution it's one full chirp — a single step in two nanoseconds). A chirp rate of 10 MHz/µsec is practical if it is desired to cover a 50 MHz bandwidth in 5 µsec (5 µsec ÷ 2 nsec for the number of steps gives a required resolution of 20 kHz).

The synthesizer is controlled by loading two registers — start frequency and chirp rate (or step size). When the chirp begins, the step size will be added to the start frequency every 2 nsec. At any point during the chirp it is possible to change the chirp rate so that the different time:frequency relationships shown in Figure 8 are possible. A negative value in the chirp rate register will produce a sweep starting at a higher frequency and moving lower, thus supporting complete manipulation of all parameters of the output.

In addition, 12 bits of phase control are available for compensation of the response during the sweep to reduce side lobes. This may be updated at a rate limited only by speed limitations of TTL logic. Phase control adds another dimension of flexibility by permitting the control of phase from pulse to pulse, which permits accurate matching of signals.

Basic specifications for the DCP-1 are given in Table 1.

Subsystem Configurations

The DCP-1 is shipped in a conventional 5.25 in. chassis that includes the basic chirp synthesizer (designated the DCP-1A) and the support functions shown in Figure 7. Users are encouraged to begin work with the complete system, as shown, to reduce risk and speed integration time. In production, however, some combination of the power supply, output filter, control interface, reference, and cooling blocks will ordinarily be integrated into the user's system. Production systems therefore use only the basic DCP-1A module (about 5"×7"×1").

Applications and Experimental Results to Date

The DCP-1 generates its output in the low-UHF range. Typically, that chirp signal must be upconverted to the desired system operating range by either mix/filter or multiplication techniques. The DCP-1 is now used by a variety of systems, including two millimeter-wave seeker programs, four synthetic aper-

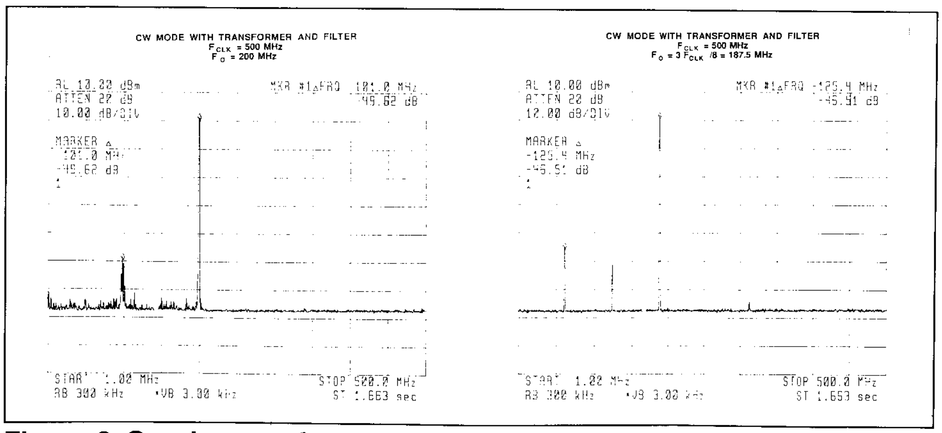

Figure 9. Spurious performance plots.

Parameter	Specification
Maximum Clock	500 MHz
Maximum Output Frequency	≈230 MHz
Frequency Resolution	24 bits (~29.8 Hz)
Frequency Update Rate (Sweep mode)	2 ns
Output Power	0 dBm ±1.5 dB
VSWR	<2.0:1
Harmonics (CW)	–45 dBc
Spurious (CW)	–50 dBc, typical
Phase Noise (CW)	
10 Hz offset	–80 dBc/Hz
100 Hz offset	–95 dBc/Hz
1 kHz offset	–110 dBc/Hz
10 kHz offset	–120 dBc/Hz
100 kHz offset	–130 dBc/Hz
1 MHz offset	–140 dBc/Hz
Group Delay Variation	
DC to 100 MHz	<0.5 nsec
DC to 150 MHz	<1 nsec
DC to 230 MHz	<4 nsec
Phase Modulation	12 binary bits
Power (DCP-1 chassis)	28VDC, 1.6A
Power (DCP-1A module)	–5.2VDC, 2.5A
	–2VDC, 500mA
	+5VDC, 100mA
	–12VDC, 125mA

Table 1. DCP-1 specifications.

ture radar systems, one electronic warfare program, and (in baseband) at least one wafer process control system.

So far, system developers have reported very favorable results. Spurious signal level was initially a concern, but one unpredicted result of experimentation is that discrete spurious signals seem to be integrated into the general output, and have little result on overall performance. Figure 9 shows typical spurious plots.

Linearity is within quantization levels, therefore for broadband chirps the errors are smaller than measurement techniques can detect. Initially, repeatability was evaluated using the configuration shown in Figure 10. The output of the DCP-1 was delayed and then compared with itself, and the result measured on an HP 3561A FFT analyzer.

Linearity testing was conducted using the Racal-Dana 2351 Time Interval Analyzer and the HP 5373A Modulation Domain Analyzer. The results are shown as Figure 11, which includes both time vs. frequency data and a histogram displaying frequency distribution.

First users report that controllability is a potential issue due to the speeds involved. Therefore, an interface card was developed to speed initial integration and provide a benchmark to support design of the user system's interface to the DCP-1.

Most users require that the DCP-1 be eventually reduced in size, price, and power consumption. At least one high-volume program requires the basic UHF chirp signal generation to be accomplished on a hybrid of about one square inch. Power consumption is a function of the digital process, and new devices are now being considered to reduce dissipation by about half. As usage/production increase, and the level of integration is improved, costs will improve as well.

The future of the DCP-X (and systems like it) depends upon two factors: yield/cost of digital and data conversion hardware, and performance of data conversion devices. As costs go down and performance improves, digital chirp generation will become a practical solution for more and more categories of systems, in both military and commercial markets. Like spread spectrum, which was once a classified and expensive approach to certain communication requirements, direct-digital chirp synthesis will be exploited in many predictable and also unforeseen niches of the RF industry.

Availability

The DCP-1 and DCP-1A are produced by Sciteq Electronics, Inc., San Diego, CA, tel. 619-292-0500. Availability is 6-8 weeks, at $25k (DCP-1 chassis) and $12.5k (DCP-1A module). *Readers may obtain further information by circling Info/Card #199.* *RF*

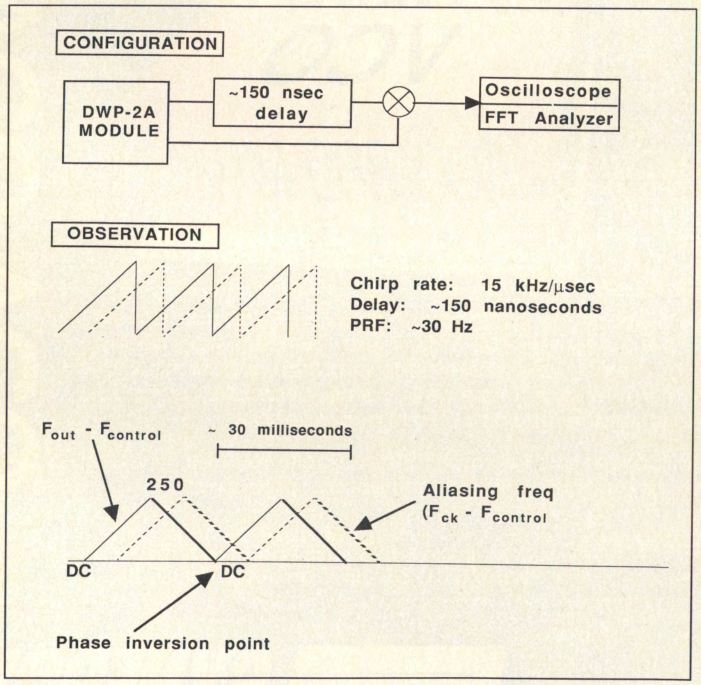

Figure 10. Repeatability testing method.

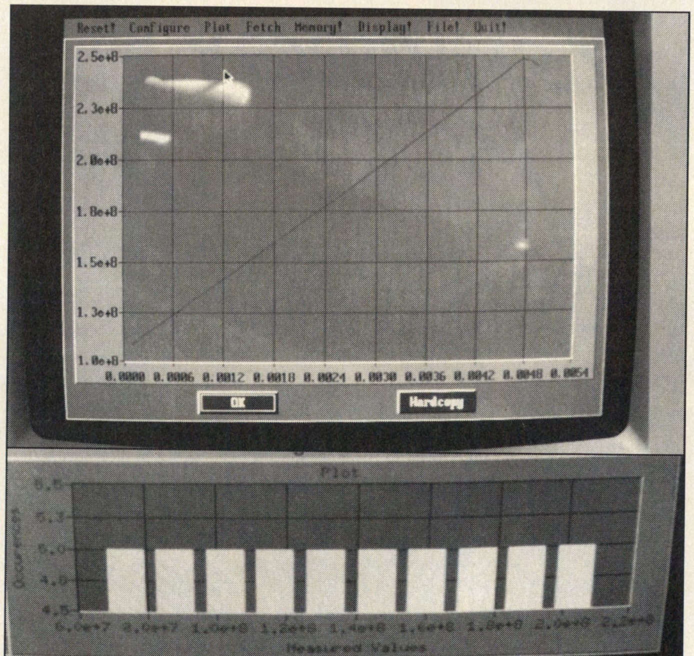

Figure 11. Linearity test results

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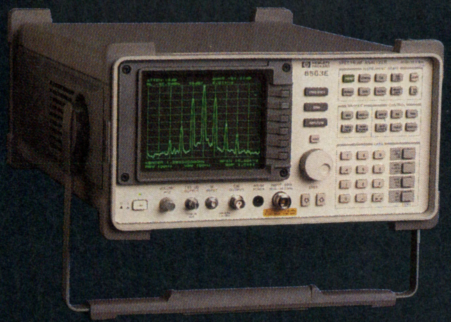

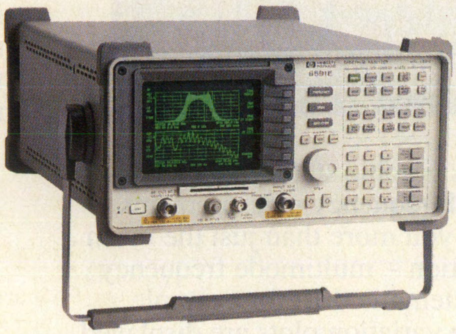

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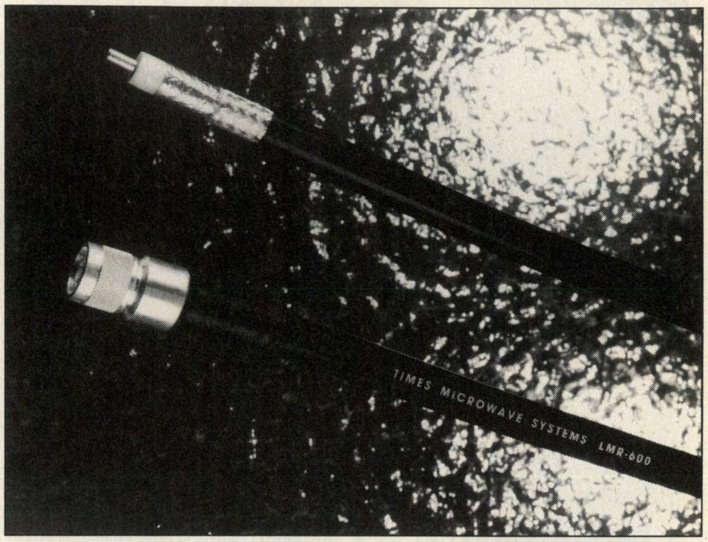

A new high performance flexible coaxial communications cable that is well suited for use as an antenna feeder in outdoor cellular and land mobile communications applications is being introduced by Times Microwave Systems. LMR™-600 flexible communications cable features a foam polyethylene dielectric to provide 2.5 dB attenuation per 100 ft. at 900 MHz. Incorporating a polyethylene jacket with a vapor sealed aluminum tape outer conductor for moisture protection, this weather resistant cable is easy to terminate in the field. Offering better than 90 dB shielding efficiency and ±10 ppm per degree C phase stability, LMR-600 flexible communications cable can be supplied in bulk or preterminated with Type N connectors. This coaxial cable is 0.590 inch diameter and is also available in 0.500 and 0.405 inch diameter sizes. LMR-600 flexible communications cable sells for $1.20 per ft. (list); Type N connectors cost $14.50 (list) each. Literature and price quotations are provided on request.

Times Microwave Systems
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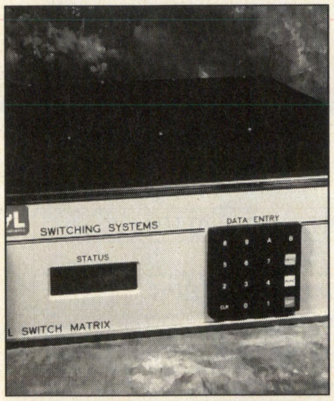
Dual Switch Matrix

K&L's dual switch matrix is designed specifically for the A.T.E. user who requires two switch matrices with operation from DC to 18 GHz but only has rack space for one. Two 1 × 24 switch matrices are integrated

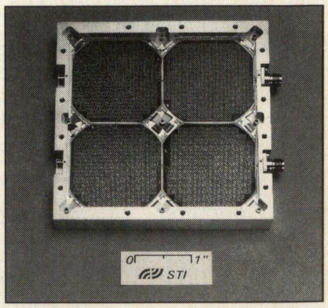


with one controller housed in a 5.25 inch high, 19 inch rack mountable enclosure with rack slides. The controller allows independent control of each matrix both via the front panel keypad and remote interface. The LCD display gives a complete status update. Both GPIB and RS-232 interfaces for computer control are included on the matrix. Because low loss coaxial switches are used, insertion loss is typically only 2 dB at 18 GHz. The system can be operated on AC power from 85 to 264 VAC.

K&L Microwave Inc.
INFO/CARD #249

Superconducting Delay Line

Superconductor Technologies Inc. (STI) has constructed a high temperature superconductor microwave delay line longer than 100 ns. The STI delay line measures a mere 3.8 × 3.8 × 0.5 inches, a dramatic reduction from the 70 feet of RG-141 stainless-steel coaxial cable typically needed for a 100 ns delay line. The new delay line takes advantage of the ultra-low electrical loss properties of HTS to improve insertion loss by a factor five, compared to RG-

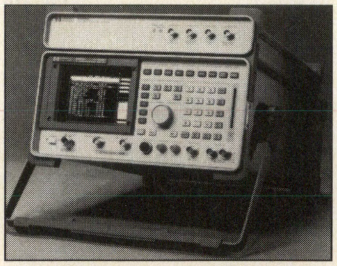


141 stainless-steel coaxial cable. This improvement enable a system to use fewer amplifiers to recover the signal, resulting in improved signal integrity. The monolithic construction of the STI delay line allows for excellent phase slope tracking between units, a critical feature in determining the angle-of-arrival of signals in direction finding equipment.

Superconductor Technologies Inc.
INFO/CARD #248

Cellular Test Set

Hewlett-Packard has introduced the HP 8920D dual-mode cellular mobile test system, which reduces test times in the manufacturing, installation and mainte-

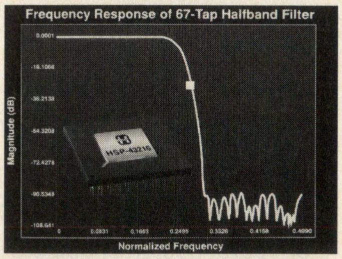


nance of AMPS and TDMA digital cellular radios. The HP 8920D performs analog and digital NADC measurements for parametric verification based on TIA IS-55 dual-mode standards. It performs call processing tests from call setup to digital-to-digital handoffs. In addition to analog testing, the HP 8920D offers $\pi/4$ DQPSK signal generation, $\pi/4$ modulation analysis, baseband data generation and analysis, BER meter and $\pi/4$ DQPSK power metering. The HP 8920D consists of the HP 8920A with options 001, 003, 004, 005, 013, 050 and the HP 83201A dual-mode cellular adapter. The HP 11807A radio test software can be used in conjunction with the HP 8920D to automate test procedures. Cost of the HP 8920D is $27,800, and cost of the HP 11807A software is $2000.

Hewlett-Packard Company
INFO/CARD #247

Digital Filter Chip

A 16-bit, 52 MSPS digital halfband filter that lets communications and video system designers implement interpolation- or decimation-by-2 with the option of quadrature up/down conversion is now available form Harris Semiconductor. The new chip, called the HSP43216, costs under $40.00. The HSP43216 incorporates a 67-tap halfband filter that processes 16- bit data with 20-bit coefficients to yield better than 90 dB stopband attenuation and better than 0.0005 dB



passband ripple. The chip includes four modes of operation: decimate- or interpolate-by-2 halfband filtering, digital down-conversion by $f_s/4$ followed by a decimate-by-2 halfband filtering, and interpolate-by-2 quadrature filtering of a complex input followed by $f_s/4$ up-conversion with real output. Engineering samples of the HSP43216 are currently avaialable, with full production slated for October. The 84-pin, PLCC, 1000-unit price is $32.55 each. The 85-lead PGA price is $62.99 each.

Harris Semiconductor
INFO/CARD #246

MORE

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RNet™ RF Modems

Frequency	403–430 MHz, 450–470 MHz
RF Output-TX	2 Watts or 4 Watts
Selectable Data Rate	1200/2400/4800/9600 BPS Asynchronous
Modulation	1200,2400-MSK; 4800,9600-DGMSK
Operation	Half-Duplex or Simplex
Protocol	Transparent to the User
Data Format	7 bits with even, odd, mark or space parity 8 bits with even, odd, mark, space parity or no parity
DTE Interface	RS-232 or TTL
Standby/RX Current Drain	32mA (RS-232) or 27mA (TTL)

MOTOROLA

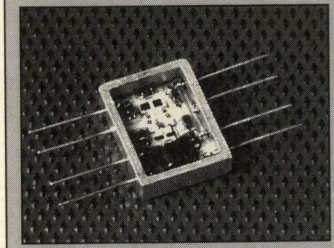
1301 East Algonquin Road
Schaumburg, IL 60196

INFO/CARD 38

Product Spotlight: Attenuators

Minimal Phase Shift

With a frequency from 10 to 1000 MHz, Merrimac's AEF-



35A series provides a minimum of 30 dB attenuation across the frequency band. The design uses PIN attenuator diodes within a precisely trimmed thick-film resistive pi network. Careful matching of the diodes for amplitude and phase balance characteristics ensures excellent flatness and minimal phase shift with attenuation.
Merrimac
INFO/CARD #245

Programmable Attenuator

A series of programmable attenuators from JFW Industries features P.C. through-hole mounting and a low package profile. Several configurations are available including a phase constant unit with frequency range of 10 to 600 MHz, attenuation steps of 0.5,1,2,4,8,16 and 32 dB, maximum VSWR of 1.4:1 and phase shift of ±3 degrees from 10 to 400 MHz and ±5 degrees from 400 to 600 MHz. Control voltage is +5 VDC at 350 mA maximum. Size is 4.25 × 1.75 × 0.500 inches.
JFW Industries, Inc.
INFO/CARD #244

Continuously Variable

Model 3100-127 from Lucas Weinschel is a precision, continuously variable attenuator covering DC to 2.0 GHz. Attenuation ranges between 6 dB and 133 dB, providing a 127 dB range. The attenuator is 1.794 inches in diameter, has 2 W average power rating at 25 degrees C, and uses SMA connectors.
Lucas Weinschel
INFO/CARD #243

UHF Programmable

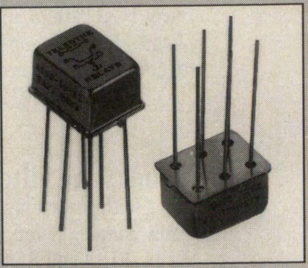
The RF2410 from RF Micro Devices is a single supply, programmable attenuator for the 700 to 1700 MHz range. The monolithic, GaAs device can be programmed over a 38 dB range in 2 dB steps. Input and output are matched to 50 ohms, and the device can drive up +10 dBm. The RF2410 is available in a 16 lead SOIC package and costs $9.50 in quatities of 1000, $4.50 in quatities of 100,000.
RF Micro Devices
INFO/CARD #242

Coaxial Attenuators

Two series of miniature 50 ohm coaxial attenuators are available in 3,6,10 and 20 dB steps. Units using SMA connectors have 0.5 dB accuracy from DC to 2.5 GHz and maximum VSWR of 1.35:1 to 3 GHz. Mini-UHF units have 0.5 dB accuracy and maximum VSWR of 1.35:1 to 1.0 GHz. They contain MIL HI REL networks with nickel or gold plated connectors. Model AT-55-dB/Conn. costs $21.00 to $22.00 each in ten piece quantities.
Elcom Systems, Inc.
INFO/CARD #241

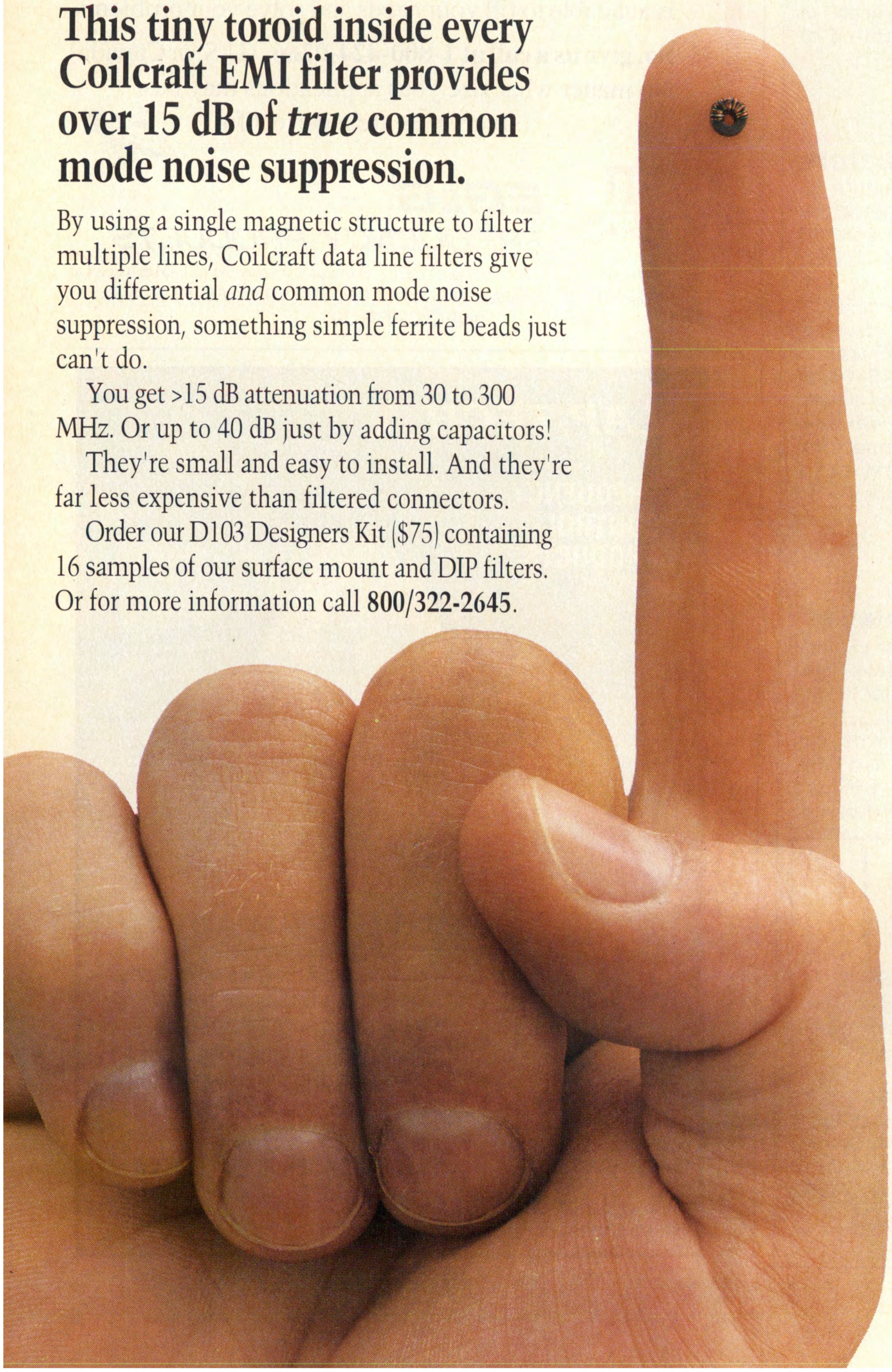
Attenuator Relay

The A150 series of electromechanical relays from Teledyne Relays are available with attenuations of 1,2,4,8,10,16 and 20 dB. The 50 ohm devices are useable from DC to 3 GHz. The devices handle 1 W and have maximum VSWR of 1.4.



The attenuator relay package is 0.375 × 0.475 × 0.280 inches. Nominal coil voltages are 5,12,15 and 26 V.
Teledyne Relays
INFO/CARD #240

Make your common mode EMI disappear into this black hole.



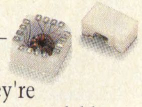
This tiny toroid inside every Coilcraft EMI filter provides over 15 dB of *true* common mode noise suppression.

By using a single magnetic structure to filter multiple lines, Coilcraft data line filters give you differential *and* common mode noise suppression, something simple ferrite beads just can't do.

You get >15 dB attenuation from 30 to 300 MHz. Or up to 40 dB just by adding capacitors!

They're small and easy to install. And they're far less expensive than filtered connectors.

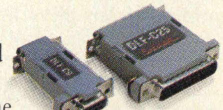
Order our D103 Designers Kit ($75) containing 16 samples of our surface mount and DIP filters. Or for more information call **800/322-2645**.



Surface Mount Filters Our ultra-miniature surface mount filters come in 10, 8, 4, 3 and 2 line configurations. They're packaged in a flat top ceramic case and are available in tape and reel format for auto insertion.

DIP Filters These filters reduce conducted noise by a factor of 32 from nearly 1 MHz to 300 MHz while passing frequencies below 100 kHz without attenuation. They come in 8, 4, 3 and 2 line versions

SIP LC Filters With capacitors on each line, these filters achieve up to 40 dB attenuation. We offer versions to filter from 1 to 8 lines.



Cable Filters Plug one of these modules into a cable and its common mode magnetics will filter all lines including the frame ground. EMI is attenuated as much as 19 dB while leaving data signals undisturbed. Choose DB-9, 15, 25 or IEEE 488 versions

Tip and Ring Filters These low cost filters help telecom designers meet FCC Part 15 and 68 requirements. Available in 2 and 4 wire versions, they provide 20 dB attenuation of common mode noise over a 30 to 250 MHz range; 15 dB out to 300 MHz.

Phone Line Filters Fix noisy phone lines with these plug-in filter modules. They come in 2 and 4 line versions, in RJ-11, RJ-14 and RJ-45 configurations.

Coilcraft

1102 Silver Lake Rd., Cary IL 60013
800/322-2645 Fax 708/639-1469
INFO/CARD 41

SIGNAL PROCESSING COMPONENTS

Quadraphase Power Divider

Model HDL 405 provides 4-way power division with outputs in phase quadrature, ie. 0°, 90°, 180° and 270° Operation is over 1 to 2 GHz, with maximum VSWR of 1.3:1 and minimum isolation of 20 dB. The divider will handle CW power levels of 60 W. Other frequency ranges are available. Size is 1.5 × 4.0 × 0.36 inches.
Technical Research and Marketing, Inc.
INFO/CARD #235

Bi-Phase Modulators

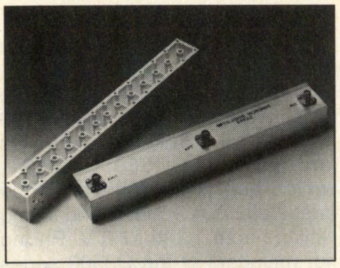
ST Olektron's BPM-1047 series of bi-phase modulators provide 180° phase shift through an RF path and feature broad bandwidth, low insertion loss, and excellent phase accuracy while consuming just 450 microwatts. Utilizing a GaAs FET switch structure with CMOS drivers, they achieve a maximum switching time of 4 ns and a modulation rate of 87 Mbits/sec.
ST Olektron Corp.
INFO/CARD #234

180° Hybrid Junction

The HT22 from Tele-Tech is a magic-tee covering 0.3 to 300 MHz. The amplitude imbalance is less than 1 dB, phase balance is less than 3° and VSWR is better than 1.5:1. The HT22 will handle input power of 0.5 W. A detailed data sheet is available.
Tele-Tech Corp.
INFO/CARD #229

Plated Plastic Filters

Teledyne Microwave has introduced a line of plated plastic filters which overcome the problems of plating adhesion and temperature stability. The filters replace cavity or combline filters



in the 500 MHz to 18 GHz with bandwidths from < 1% to > 50%. This product line has been tested to 18 W CW without degradation, with higher powers available.
Teledyne Microwave
INFO/CARD #232

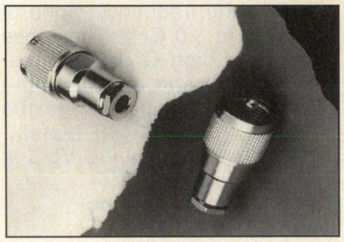
CABLES & CONNECTORS

Type N Tees

New T-connectors from Delta Electronics are currently produced in two styles of cable plugs and two styles of bulkhead mounted jacks. The connectors are available with crip attachments for flexible cable or soldered attachments for semi-rigid cable. The connectors are manufactured to MIL-C-39012 specifications, and all characteristics are equivalent to MIL-C-39012 type N right angle connectors for coaxial cable.
Delta Electronics
INFO/CARD #238

UHF Plug

A UHF series straight plug featuring a mechanical back end rather than soldered screw-down type construction is being introduced by Tru-Connector. The plug features all machined



nickel plated brass contruction with PTFE insulators and heavy silver plated contacts for use with RG-55, 58, 142 and 223/U cables. The UHF series plug will perform satisfactorily to 200 MHz. List price for the plug is $7.50.
Tru-Connector Corporation
INFO/CARD #236

SIGNAL SOURCES

Phase Locked Oscillators

T and M Microwave introduces its PL 3500 series of surface mount phase locked oscillators covering the frequency range of 600 MHz to 3.0 GHz. The series features a high Q ceramic resonator VCO and accepts input reference frequencies of 5 to 100 MHz at 0 dBm ±3 dB typical. Operating voltage is +15 VDC at 200 mA maximum.
T and M Microwave, Inc.
INFO/CARD #227

Expanded VCO Line

Qualcomm has announced ten new members of their Q3500 series of wide-band microwave VCOs. The series now covers 100 MHz to 3.5 GHz. The new VCOs are led by the Q3500C-2235T, which has an output frequency range of 2.2 to 3.5 GHz and tuning voltage from 1.2 to 17.0 V. Pricing for members of the Q3500 series is as low $29.90 each in 1000 piece quantities.
Qualcomm Inc., VLSI Products
INFO/CARD #226

VHF Oscillators

Wenzel Associates introduces the Sprinter series of VHF oscillators, featuring low phase noise and reduced size. Phase noise performance for the 100 MHz SC-cut crystal oscillator is –125 dBc/Hz at 100 Hz offset and –165 dBc/Hz at 10 kHz offset. Package size is as small as 2 × 2 × 0.6 inches. The series is available with SMA connectors, or in PC mount form.
Wenzel Associates, Inc.
INFO/CARD #223

UHF TV Amplifier

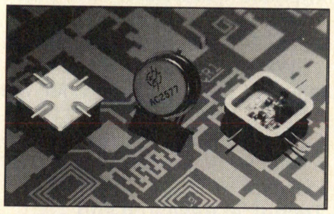
Motorola has introduced the ATV6060 broadband, linear RF amplifier assembly. The ATV6060 device is designed to operate with fully instantaneous bandwidth in the television broadcast bands IV and V (470 to 860 MHz). The assembly provides 9 dB of small signal gain is characterized by a 3-tone IMD specification of –50 dB at a reference power of 40 W, and is capable of 60 W, class A output power.
Motorola Semiconductor Products
INFO/CARD #222

SDLVA

Model SDLVA-0120-70 (0.1 to 2.0 GHz) and SLVA-06135 (600 MHz to 1.35 GHz) are hybrid MIC/MMIC DC-coupled successive detection log video amplifiers (SDLVA) with 65/70 dB dynamic range. TSS is ≤ –67 (20 MHz video BW), 25 mV/dB log slope, ±1.0 dB log accuracy with rise-time <= 20 ns, fall-time ≤ 25 ns and delay times of 8 ns at –10 dBm limited IF output. Size is 2 × 1.75 × 0.4 inches.
American Microwave Corp.
INFO/CARD #221

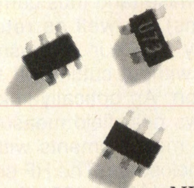
Hybrid Amplifier

The model AC2577 thin-film cascadable amplifier operates over the frequency range of 10 to 2500 MHz with minimum gain of



10.0 dB (0 to 50 degrees C). Noise figure is typically 3.3 dB. Typical output power at the 1 dB compression point is +21.0 dBm. The unit is available in SMA, flatpack and surface mount housings.
Cougar Components
INFO/CARD #220

FINALLY! 3 VOLT RF ICs FOR HANDHELD WIRELESS PRODUCTS



NEC'S NEW LOW CURRENT SILICON ICs not only reduce power consumption in your 3 Volt designs, they reduce the parts count, improve the reliability, and simplify the assembly.

NEC is committed to 3 Volt Silicon technology; these parts are just an introduction. There are more on the way. Soon.

If 5 Volt ICs are the solution, NEC has a wide selection of them as well, all available right now.

And don't forget NEC's huge family of discrete devices. Many are characterized at 2.5 Volts and they're available in a variety of packages, including a new ultra-small SMD that's a third the size of a SOT 23!

For a ***FREE Product Selection Guide***, call your nearest CEL Sales Office. Or circle the number below.

New 3 Volt NEC Silicon ICs

UPC2745
Ultra Broad Band Amplifier
DC to 2.7 GHz
12 dB Gain
-1 dBm PSAT
7.5 mA Icc

UPC2747
Broad Band Low Noise Amplifier
100 MHz to 1800 MHz
12 dB Gain
3.3 dB NF, 5 mA Icc

UPC2746
High Gain Amplifier
DC to 1500 MHz
19 dB Gain
0 dBm PSAT
7.5 mA Icc

UPC2748
High Gain Low Noise Amplifier
200 MHz to 1500 MHz
19 dB Gain
2.8 dB NF
6 mA Icc

UPB1502
Frequency Divider
DC to 2.2 GHz
Divide by 64/65, 128/129
6 mA Icc

UPB587
Frequency Divider
DC to 1000 MHz
Divide by 2/4/8
5.5 mA Icc

...plus dozens of Bipolar Transistors characterized at 2.5 Volts, including:

f = 1500 MHz	*f = 2.5 GHz*	*f = 1000 MHz*
NE68519	**NE68030**	**NE68039**
1.3 dB NF	1.90 dB NF	1.0 dB NF
10.0 dB GA	7.2 dB GA	11.6 dB GA
@ only 3 mA Ic	*@ only 3 mA Ic*	*@ only 1 mA Ic*

Surface Mount packages shown actual size

...and a wide variety of 5 Volt RF ICs including:

UPC2708
Wide Band Amplifier
DC to 2.9 GHz
15 dB Gain
10 dBm PSAT

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
Down Convertor
RF=0.4 to 3.0 GHz
IF=50 - 600 MHz
Conversion Gain=15 dB

UPC2726
Differential Amplifier
400 MHz to 1400 MHz
15 dB Gain

NEC

Typical Performance at 25° C

California Eastern Laboratories

CEL Headquarters, 4590 Patrick Henry Drive, Santa Clara, CA 95056-0964; (408) 988-3500 FAX (408) 988-0279 □ Santa Clara, CA (408) 988-7846 □ Los Angeles, CA (310) 645-0985
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INFO/CARD 44

200 W VHF

The model GRF0501 is a high powered solid state linear amplifier (class AB) capable of delivering 200 W in the frequency range of 100-500 MHz. The gain is adjustable from 3 dB to 8 dB via front panel control; other gain requirements are available. The amplifier is packaged in a standard 19 inch rack-mount enclosure.

GTC, RF Products Group
INFO/CARD #218

AMPLIFIERS

Optical to Coax Converter

The BGY887BO optical receiver operates over the frequency range of 40 to 860 MHz. The receiver module plugs directly into existing equipment and has an integral fiber pigtail to couple it to a fiber-optic distribution cable. The BGY887BO is packaged in a SOT115 power package. The device will sell for between $250 and $300, depending on volume.

Philips Semiconductor
INFO/CARD #216

Harmonic Generator/Filters

KW Microwave introduces a series of integrated harmonic generator switchable filters covering frequencies from 31.25 to 343.75 MHz. It consists of a 31.25 MHz harmonic generator and a six channel switch filter, selecting frequencies in the range of 31.25 to 343.75 MHz with 62.5 MHz frequency step.

KW Microwave Corp.
INFO/CARD #215

SEMI-CONDUCTORS

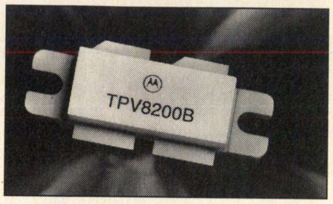
Power MESFETs

Twenty-two, internally matched power FETs from California Eastern Laboratories provide high efficiency and high gain over 4.4 to 8.8 GHz. Available in typical power outputs of 4.5, 9 and 18 W, the NEZ5964, 6472, 4450 and 7785 have efficiencies ranging from 31 to 40 percent and gains from 7.5 to 10.5 dB.

California Eastern Laboratories, Inc.
INFO/CARD #206

Class AB for UHF TV

Motorola has announced its highest power linear transistor for television transmitter applications. The push-pull TPV8200B



linear transistor is a class AB device with 190 W peak sync output power and 8 dB min. gain across 470 to 860 MHz. It has a minimum efficiency of 45 percent.

Motorola Semiconductor
INFO/CARD #205

TEST EQUIPMENT

Antenna Analyzer

The ESTAR2110 antenna measurement analyzer uses an RF marking process which allows characterization of both amplitude and phase (and thus current distributions), as well as return loss, of any device in a transmitting chain without cutting off the transmission. An optically-linked probe allows near field measurements and measurements within sheilded enclosures. The RF sensitivity of the 2110 is –120 dBm.

Antenna Research Associates
INFO/CARD #204

Switch Matrices

AML introduces a line of RF switch matrices. Model AC820500-8 is non-blocking 20-500 MHz, has eight inputs and three outputs. Any input may be switched to any one or more outputs via RS232 and IEEE busses. The matrices' third order intercept point, referenced to the input, is +24 dBm.

AML, Inc.
INFO/CARD #202

RF Sweep Generator

The LI10A covers 20 to 1200 MHz and has a sweep width variable from 2 MHz to full range. A single crystal time base provides 1, 10 and 50 MHz marker grids. Dual channel, $|S_{11}|$ and $|S_{21}|$ measurements can be made, and the two detector inputs are chopper-amplifier stabilized. Price for the LI10 is $1685. Accessories are available.

Lurie Instruments
INFO/CARD #200

M-tron Industries Announces ISO 9001 Registration
A superior manufacturer of quartz crystals and quartz crystal oscillators, M-tron Industries has been registered to the following quality systems: ISO 9001, EN 29001 and BS 5750: Part 1.
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INFO/CARD 47

Introduction to S-Parameters

By Theodore Grosch, Ph.D.
MIT/Lincoln Labs

Scattering-parameters, or S-parameters, have been used since the 50's to describe microwave circuits. At high frequencies it is difficult to measure the voltage and current at a terminal of a device or network. On the other hand, directional couplers can easily measure the power flow into or out of the circuit. This creates a situation where it is convenient to describe the electrical properties of a circuit by some means of power flow or ratios thereof.

S-parameters are steady-state linear coefficients used to express an n-port at a particular frequency. Since S-parameters are linear combinations of voltages and currents in a network, they can be manipulated much like any other linear matrix form of network characterization.

S-parameters represent ratios between incident and reflected wave amplitudes in a transmission line.These are direct expressions of reflection and transmission that are especially useful in high frequency analysis and design. They are particularly suitable for problems of power transfer in networks involving insertion gain or loss. Since they express power flow, they are also especially useful in problems dealing with the conservation of energy in passive networks. This makes them valuable for finding the realizability of networks in the frequency domain for network synthesis.

The Wave Vectors

S-parameters are ratios derived from transmission line concepts of incident and reflected voltage and current waves. The popular transmission line solution gives these quantities as forward and reverse time-harmonic traveling waves. Consider the circuit in Figure 1 with a generator of open circuit voltage E and a positive real internal impedance of Z_0 connected to a load of impedance Z_L. This represents a termination on a transmission line of characteristic impedance Z_0. The current and voltage in Figure 1 are:

$$I = \frac{E}{Z_0 + Z_L} \tag{1}$$

$$V = \frac{EZ_L}{Z_0 + Z_L} \tag{2}$$

where there is an implied factor of ejwt along with these quantities. These are vectors that define the amplitude and phase of the voltage and current at a specific frequency referenced to the open circuit voltage E. In a high frequency circuit, these quantities are difficult to measure. It's easier to put a directional coupler between the generator and the load and measure the power flow that is incident upon, and reflected from, the load.

Let's define the incident current and voltage as those quantities that would exist at the load if the load were perfectly matched to the generator ($Z_L = Z_0$).

$$I^+ = \frac{E}{2Z_0} \tag{3}$$

$$V^+ = \frac{E}{2} \tag{4}$$

This makes it easy to define the maximum power, or incident power, that can be delivered to any load.

$$V^+(I^+)^* = P^+ = \frac{|E|^2}{4Z_0} \tag{5}$$

Let's also define the reflected voltage and current as the difference between their actual and incident values.

$$\begin{aligned} I^- &= I - I^+ = \frac{E}{Z_0 + Z_L} - \frac{E}{2Z_0} \\ &= \frac{E(Z_0 - Z_L)}{2Z_0(Z_0 + Z_L)} \end{aligned} \tag{6}$$

$$\begin{aligned} V^- &= V - V^+ = \frac{EZ_L}{Z_o + Z_L} - \frac{E}{2} \\ &= \frac{E(Z_o - Z_L)}{2(Z_o + Z_L)} \end{aligned} \tag{7}$$

From the ratio of either of the two voltages or currents one gets the expression for reflection coefficient.

$$\frac{V^-}{V^+} = \frac{I^-}{I^+} = \frac{Z_o - Z_L}{Z_o + Z_L} \tag{8}$$

Since we can easily measure incident power with directional couplers, one of our variables will be related to Equations 5. Going back to the incident voltage and current, it would be more convenient to normalize these by the characteristic impedance Z_0. We can define a normalized incident vector a in terms of the incident voltage and current.

$$a = \frac{V^+}{\sqrt{Z_o}} = \sqrt{Z_o}I^+ = \frac{V + Z_oI}{2\sqrt{Z_o}} \tag{9}$$

We can also define a normalized reflected vector b that is a function of the reflected voltage and current.

$$b = \frac{V^-}{\sqrt{Z_o}} = -\sqrt{Z_o}I^- = \frac{V - Z_oI}{2\sqrt{Z_o}} \tag{10}$$

So, in terms of a and b, the incident power is

$$P^+ = |a|^2, \tag{11}$$

the reflected power is

$$P^- = |b|^2, \tag{12}$$

the reflection coefficient is

$$\Gamma = \frac{b}{a}, \tag{13}$$

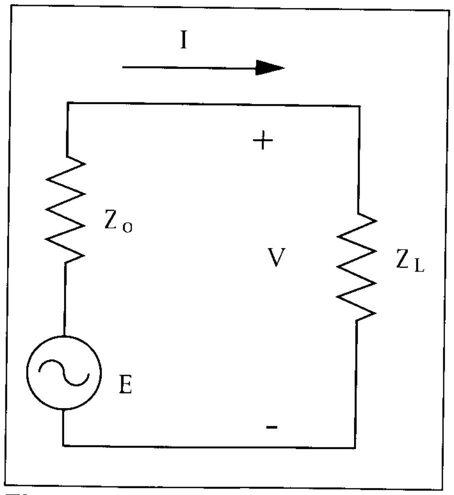
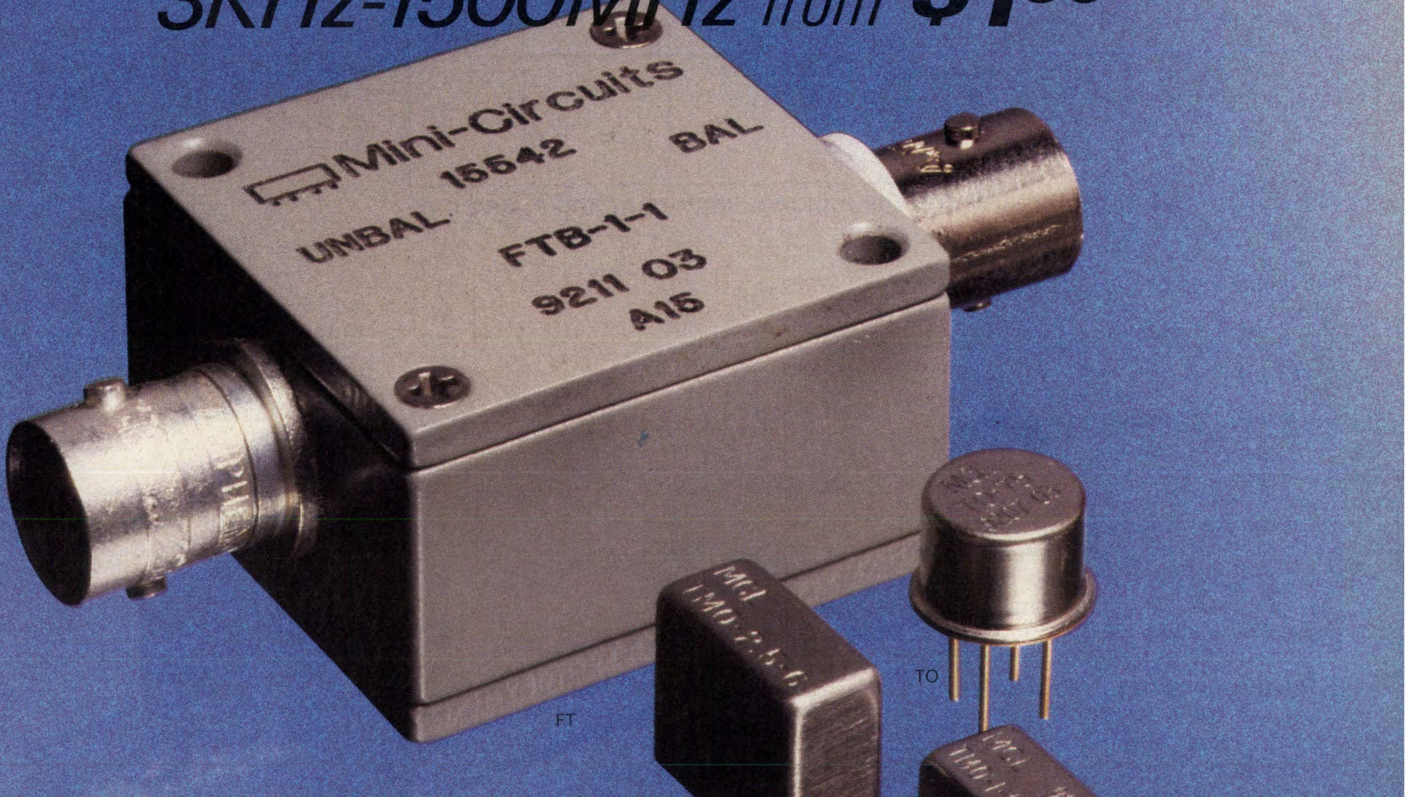


Figure 1. A one-port network Z_L connected to a generator with a positive real internal impedance Z_0.



Mini-Circuits
15542
BAL
UMBAL
FTB-1-1
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FT
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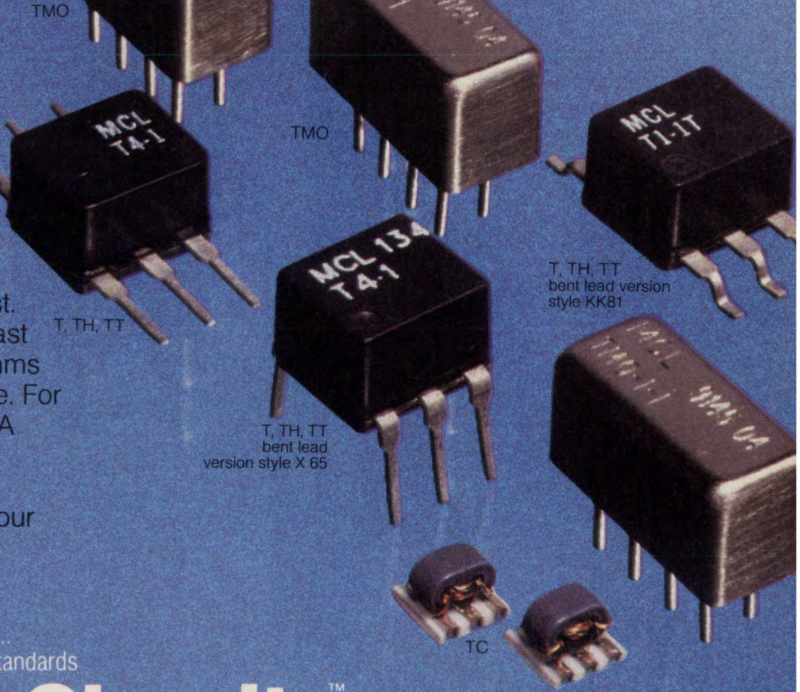

TMO
MCL
T4-1
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bent lead version
style KK81
MCL 134
T4-1
T, TH, TT
bent lead
version style X 65
TC

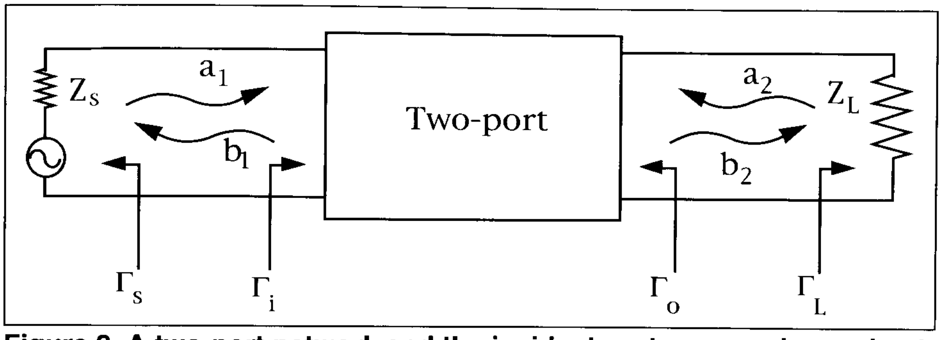

Figure 2. A two-port network and the incident vectors a_1 and a_2 and outgoing waves b_1 and b_2 at the two ports.

the actual power delivered to the load will be

$$P = |a|^2 - |b|^2, \tag{14}$$

and the VSWR of the termination is

$$\rho = \frac{|a| + |b|}{|a| - |b|}. \tag{15}$$

Armed with a new set of variables a and b, it is simpler to compare the calculated circuit response with the power flow measured with directional couplers and power meters.

S-Parameters of Linear Networks

Usually, there is very little interest in the actual values of a and b. The interest lies in their ratio defining reflection and transmission of power in a network. The reflection coefficient G in Equation 13 is the S-parameter representation of a one port. Just like Z and Y parameters, S-parameters can be extended to n-port networks by defining incoming and outgoing waves at each port,

$$a_i = \frac{1}{2}\left(\frac{V_i + Z_o I_i}{\sqrt{Z_o}}\right), \tag{16}$$

$$b_j = \frac{1}{2}\left(\frac{V_j - Z_o I_j}{\sqrt{Z_o}}\right), \tag{17}$$

where i and j are the port number. For the two-port network shown in Figure 2, the S-parameters are defined as,

$$S_{11} = \frac{b_1}{a_1}\bigg|_{a_2=0}, \tag{18}$$

$$S_{12} = \frac{b_1}{a_2}\bigg|_{a_1=0}, \tag{19}$$

$$S_{21} = \frac{b_2}{a_1}\bigg|_{a_2=0}, \tag{20}$$

$$S_{22} = \frac{b_2}{a_2}\bigg|_{a_1=0}, \tag{21}$$

where

$$b_1 = S_{11}a_1 + S_{12}a_2, \tag{22a}$$

$$b_2 = S_{21}a_1 + S_{22}a_2, \tag{22b}$$

or in matrix representation

$$\begin{pmatrix} b_1 \\ b_2 \end{pmatrix} = \begin{pmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{pmatrix}\begin{pmatrix} a_1 \\ a_2 \end{pmatrix}. \tag{23}$$

In Equation 22, one can see that b_1 is a linear combination of the reflected wave at the input due to S11 and the wave transmitted from output to input due to S12.

Since the waves a_1 and b_1 are functions of voltage and current vectors, the S-parameters their ratios form can be related to the Z, Y, and ABCD parameters of the network. To see how this works, we look at the matrix of S-parameters S and column vectors a' and b' in Equation 23,

$$\mathbf{b'} = \mathbf{Sa'} \tag{24}$$

where

$$\mathbf{a'} = \frac{1}{\sqrt{Z_o}}(\mathbf{V} + Z_o\mathbf{I}), \tag{25a}$$

$$\mathbf{b'} = \frac{1}{\sqrt{Z_o}}(\mathbf{V} - Z_o\mathbf{I}), \tag{25b}$$

and **V** and **I** are the column vector made of the voltages on the two-port terminals and the current flowing into the terminals. The impedance matrix of the two port is defined as,

$$\mathbf{V} = \mathbf{ZI}. \tag{26}$$

Substituting Equation 26 in to the Equations 25,

$$\mathbf{a'} = \frac{1}{\sqrt{Z_o}}(\mathbf{ZI} + Z_o\mathbf{I}), \tag{27a}$$

$$\mathbf{b'} = \frac{1}{\sqrt{Z_o}}(\mathbf{ZI} - Z_o\mathbf{I}), \tag{27b}$$

we can find that the S-parameter matrix can be expressed as a function of the Z-parameter matrix and the characteristic impedance.

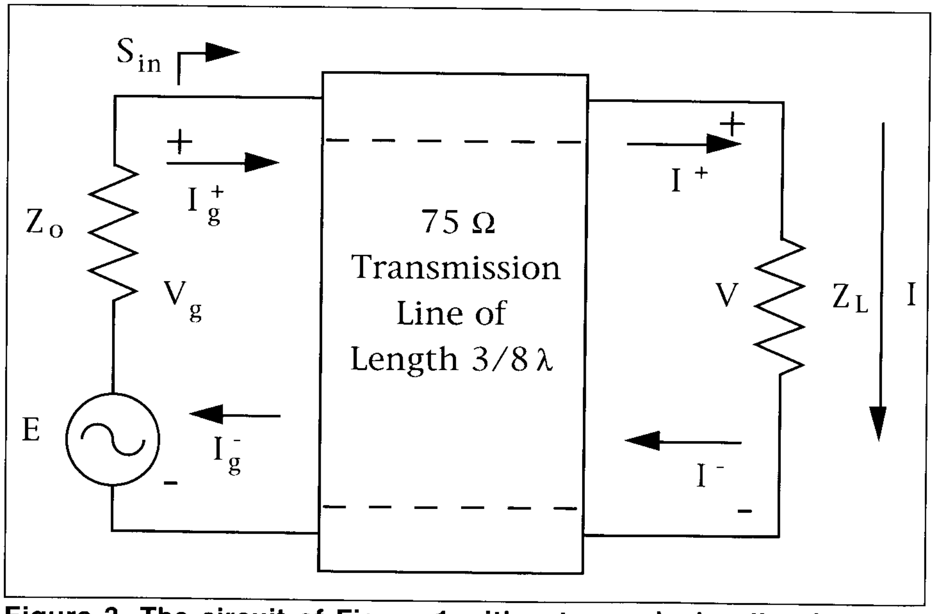

Figure 3. The circuit of Figure 1 with a transmission line inserted between the generator and load.

$$\mathbf{S} = \frac{\mathbf{b'}}{\mathbf{a'}} = \frac{(\mathbf{ZI} - Z_o\mathbf{I})}{(\mathbf{ZI} + Z_o\mathbf{I})} = (\mathbf{Z} - \mathbf{Z_o})(\mathbf{Z} + \mathbf{Z_o})^{-1} \tag{28}$$

Note that the constant Z_0 in Equation 28 was converted to a diagonal matrix Z_0 by multiplying Z_0 times the identity matrix. S-parameters are related to other network parameters by performing similar derivations to the one above.

Examples

The following examples will illustrate the use of the wave vectors and S-parameters. Consider the situation where a 75 Ohm transmission line terminated with a 35 + j120 Ohm load. We want to find the reflection coefficient and the VSWR using the wave vectors.

Figure 1 shows this circuit with Z_0 = 75 W and Z_L = 35 + j120. Normalizing the voltage source to 1, the incident voltage and current are (as if the transmission line were perfectly matched),

$$V^+ = \frac{1}{2}, I^+ = \frac{1}{150}. \tag{29}$$

The reflected voltage and current are found to be

$$\begin{aligned} V^- &= \frac{75-(35+j120)}{2(75+(35+j120))} \\ &= \frac{40-j120}{220+j240} = \frac{4}{-5+j9}, \end{aligned} \tag{30}$$

$$\begin{aligned} I^- &= \frac{75-(35+j120)}{150(75+(35+j120))} \\ &= \frac{4}{-375+j675}. \end{aligned} \tag{31}$$

The vectors a and b are

$$a = \frac{1}{2\sqrt{75}}, \; b = \frac{4}{\sqrt{75}(-5+j9)}, \tag{32}$$

and so the reflection coefficient and the VSWR are

$$\Gamma = S_L = \frac{8}{-5+j9}, \tag{33}$$

$$\rho = \frac{\frac{1}{2}+\frac{4}{\sqrt{106}}}{\frac{1}{2}-\frac{4}{\sqrt{106}}} = 7.97{:}1. \tag{34}$$

For the next example, consider the circuit shown in Figure 3. The termination in Example 1 is connected to the same 75 ohm source through a transmission line of length 3/8λ. At the load, the parameters calculated above are the same. At the generator, the phases have been shifted by the length of transmission line. In terms of incident voltage and current at the generator,

$$V_{g^+} = \frac{1}{2}\exp(j2\pi\frac{3}{8}) = -\frac{1}{2\sqrt{2}} + j\frac{1}{2\sqrt{2}}, \tag{35}$$

$$I_{g^+} = \frac{1}{150}\exp(j2\pi\frac{3}{8}) = \frac{1}{150}\left(-\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right). \tag{36}$$

The reflected voltage and current are

$$V_{g^-} = \frac{4}{-5+j9}\exp(-j2\pi\frac{3}{8}) = \frac{2\sqrt{2}+j2\sqrt{2}}{5-j9} \tag{37}$$

$$\begin{aligned} I_{g^-} &= \frac{4}{-375+j675}\exp(-j2\pi\frac{3}{8}) \\ &= \frac{2\sqrt{2}+j2\sqrt{2}}{375-j675}. \end{aligned} \tag{38}$$

The vectors a and b are

$$a' = \frac{1}{2\sqrt{75}}\left(-\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right), \tag{39}$$

$$b' = \frac{2\sqrt{2}+j2\sqrt{2}}{\sqrt{75}(+5-j9)} \tag{40}$$

and the reflection coefficient and the VSWR are

$$\begin{aligned} S_{in} = \Gamma = \frac{b'}{a'} &= \frac{\frac{4}{\sqrt{75}(-5+j9)}\exp(-j2\pi\frac{3}{8})}{\frac{1}{2\sqrt{75}}\exp(j2\pi\frac{3}{8})} \\ &= \frac{8}{-5+j9}\exp(-j4\pi\frac{3}{8}) \\ &= \frac{8\exp(-j\pi\frac{3}{2})}{-5+j9} = \frac{j8}{5-j9} \end{aligned} \tag{41}$$

$$\rho = 5.644{:}1\left(\text{since}\left|\left(-\frac{1}{\sqrt{2}} + j\frac{1}{\sqrt{2}}\right)\right| = 1\right). \tag{42}$$

Notice that the phase of the reflection coefficient has been shifted by 2 times the electrical length because the waves have to get to the load and then come back again through the length of transmission line.

For the final example, consider a two port network with port two terminated as shown in Figure 2. Suppose we want to know the input reflection coefficient Gi. The ratio of the reflected wave from the load, a_2, to the incident wave on the load, b_2, is

$$\Gamma_L = \frac{a_2}{b_2}. \tag{43}$$

The incident wave on the output of the two-port is

$$a_2 = \Gamma_L b_2. \tag{44}$$

Substituting this in Equation 22, the response of the two-port is given by

$$b_1 = S_{11}a_1 + S_{12}\Gamma_L b_2, \tag{45}$$

$$b_2 = S_{21}a_1 + S_{22}\Gamma_L b_2. \tag{46}$$

By solving the second equation for b2,

$$b_2 = \frac{S_{21}a_1}{1-S_{22}\Gamma_L} \tag{47}$$

we can solve the first equation and find the input reflection coefficient.

$$\Gamma_i = \frac{b_1}{a_1} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}, \tag{48}$$

The S-parameters can be manipulated algebraically like any of the other linear parameters.

Closing Remarks

This has been a basic nuts-and-bolts discussion of S-parameters and their relation to the other linear network parameters. The voltage across a load or network port is split into two components. One component being the voltage (incident voltage) that would exist there if the load or port were perfectly matched to the transmission line. The second voltage is the reflected or difference between the actual and incident voltages. The concept is the same for current, and with these quantities, we define the incident and reflected waves a and b.

A few things have been simplified in this discussion. Most all applications will be in a characteristic impedance of 50 Ohms. But in general, the characteristic impedance on the ith port of a network can be any complex number Zsi other than 0 or infinity. In this case, we have to define the column vectors a' and b' for an n-port as

$$\mathbf{a'} = \mathbf{R_s}(\mathbf{V} + \mathbf{Z_s}\mathbf{I}) \tag{49}$$

$$b' = R_s(V - Z_s * I) \qquad (50)$$

where Rs and Zs are diagonal matrices whose ith components are given by 1/2{Re(Zsi)}-1/2 and Zsi respectively. We must then derive new formulas for V?, I? and the conversion formulas for Y and Z parameters. Starting with the basics given here, one can continue onto discussions of power gain, insertion loss, load matching, and stability needed for circuit design. *RF*

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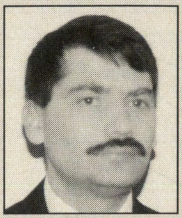
About the Author



Theodore Grosch received BS, MS and Ph.D. degrees in Electrical Engineering from Pennsylvania State University. He has worked at Hughes Aircraft, General Electric Astrospace, and was president of SSDD Research Corp. He currently is a member of the Technical Staff at MIT/Lincoln Laboratory. He can be reached at 242 Sugar Rd, Bolton, MA 01740, or by phone at (617) 981-0130.

A Tiny Electromagnetics Simulator

By Jonathon Y.C. Cheah
Hughes Network Systems

This article describes a tiny simulation program based on the Transmission Line Model [1] for solving the electromagnetic wave equation. The Transmission Line Model (TLM) method provides a visual numerical solution to problems in electromagnetics that is more intuitive than purely numerical results. This program was among the top entries in the 1993 RF Design Awards Contest.

Electromagnetics is one of the more difficult subjects in the RF engineering discipline. However, demands for modern day RF design engineers are in the circuit design arena, so the fundamentals of electromagnetics take a back seat in most instances. Engineers gladly use pre-packaged closed-form equations for most problems of this nature. The popular use of closed-form quasistatic approximations for the complicated inhomogeneous microstrip line impedance and discontinuities are good examples. However, problems in electromagnetics are normally not as easily visualized as in other circuit analysis.

TLM puts visibility into the solution by using a time domain impulse over a transmission line structure. If a time impulse is asserted in a transmission line, it is logical that the frequency response of the transmission line can then be fully characterized by its impulse response.

Following the equivalent circuit of a shunt node as shown in Figure 1, the well known resultant wave equation from the Maxwell's equations for E_y propagating in the z and x directions, and its equivalent circuit counterpart respectively are:

$$\frac{\partial^2 E_y}{\partial x^2} + \frac{\partial^2 E_y}{\partial z^2} = \mu\varepsilon \frac{\partial^2 E_y}{\partial t^2} \qquad (1)$$

and

$$\frac{\partial^2 V_y}{\partial x^2} + \frac{\partial^2 V_y}{\partial z^2} = 2LC \frac{\partial^2 V_y}{\partial t^2} \qquad (2)$$

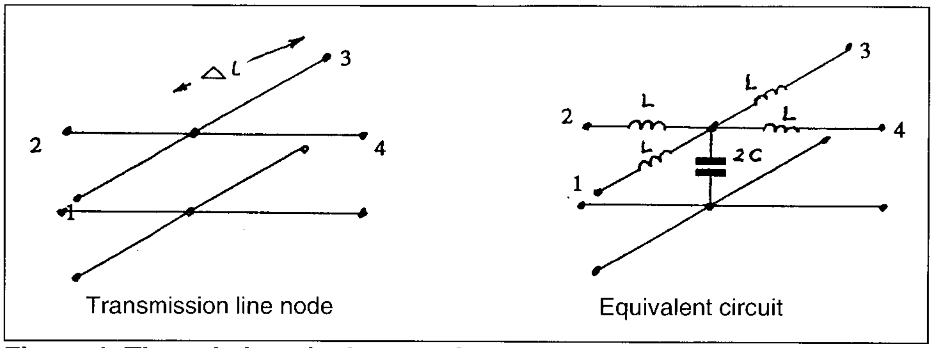


Figure 1. The unit length shunt node.

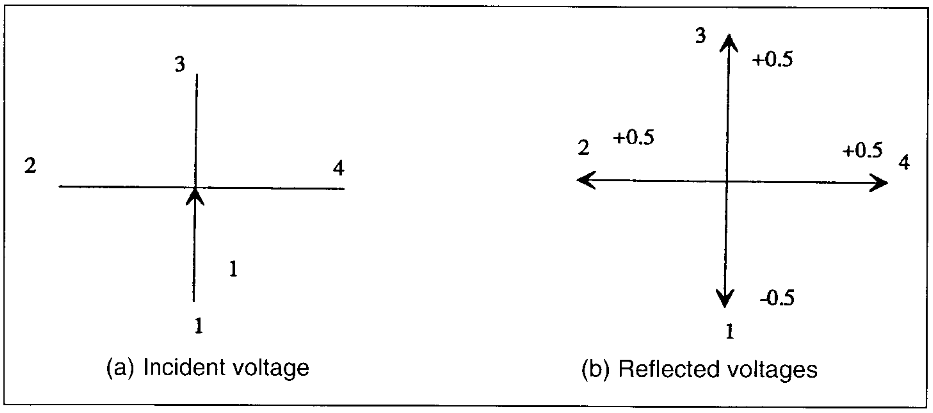


Figure 2. The unit node incident /reflected voltage relationship.

It follows that,

$E_y \equiv V_y \qquad \mu \equiv L \qquad \varepsilon \equiv 2C$

Thus, the propagation velocity at the shunt node is:

$$v = \frac{1}{\sqrt{\mu_0 \varepsilon_0}} = \frac{1}{\sqrt{2LC}} \qquad (3)$$

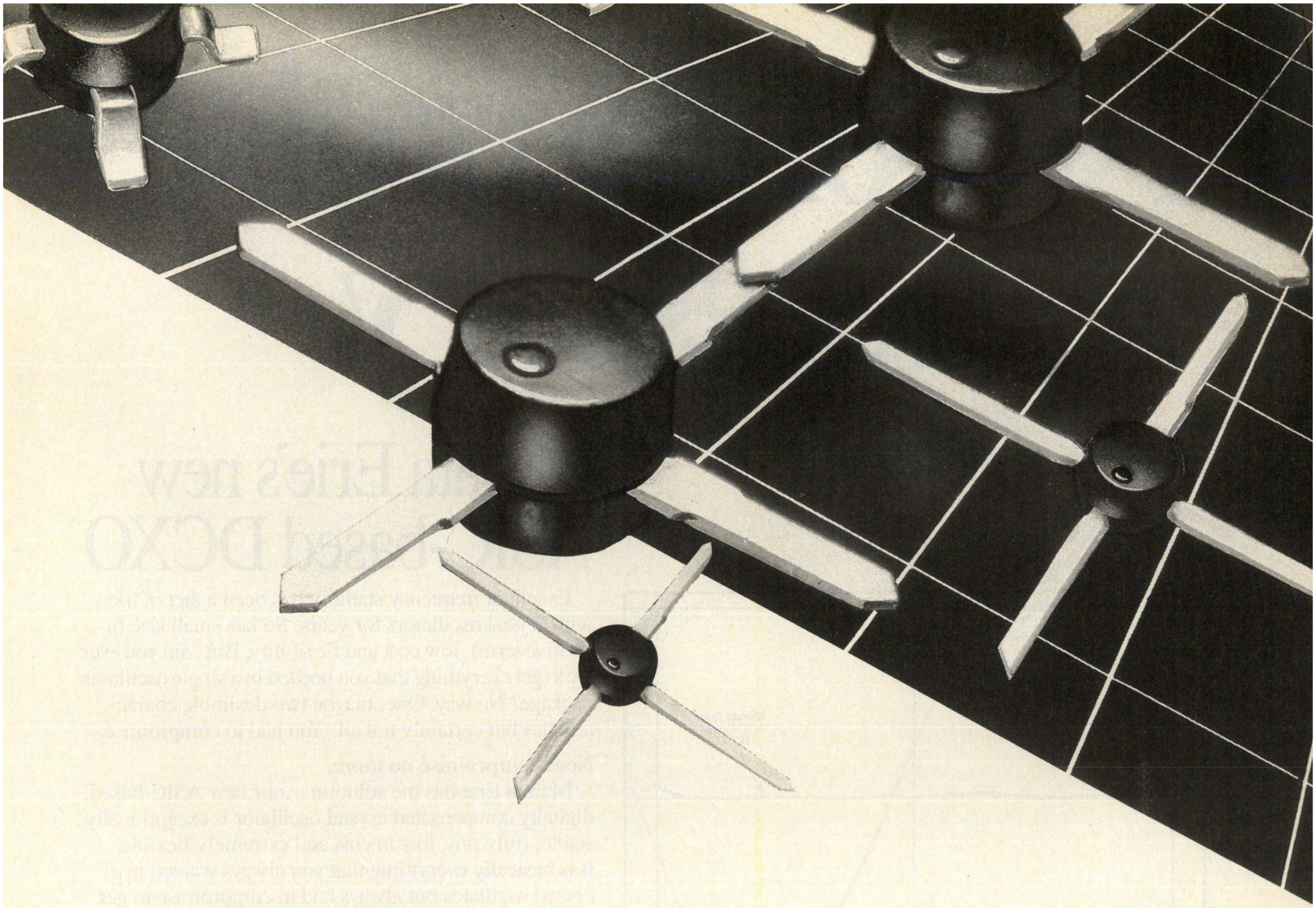
Similarly the intrinsic impedance is,

$$\eta = \sqrt{\frac{\mu_0}{\varepsilon_0}} = \sqrt{\frac{L}{2C}} \qquad (4)$$

It can be seen that the above equivalence introduces a propagation velocity factor of $1/\sqrt{2}$ in TLM simulation, and it will be more obvious in the simulation program. Indeed, the velocity is a function of the granularity of the physical dimension step size used in the computation with respect to the propagation wavelength. In this program, this variation is avoided by enforcing the total granularity to be no larger than 0.05 of a wavelength.

Considering the mesh point configuration shown in Figure 2a, a voltage impulse of magnitude 1 arriving from line 1 will generate a reflection of –1/2, by virtue of the transmission line mismatch of the termination of three parallel lines, with resultant impedance of 1/3,

$$\Gamma = \frac{1/3 - 1}{1/3 + 1} = -\frac{1}{2} \qquad (5)$$



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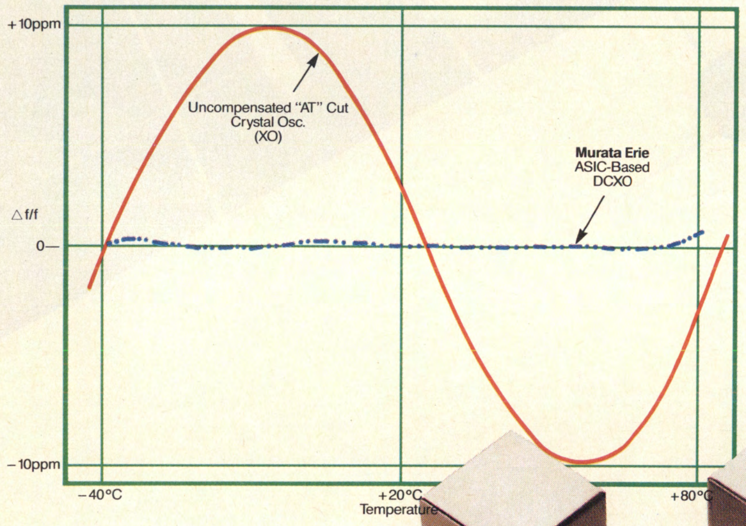

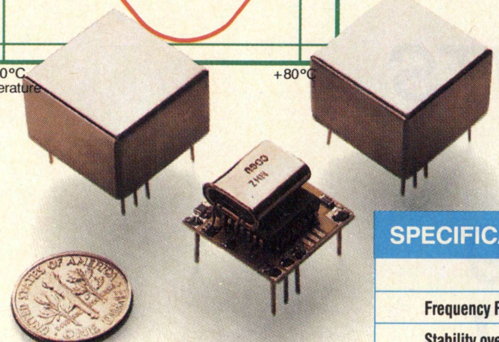

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Commands	Purpose
start	This command must be the first command of the file.
movie n	Movie enables the simulation in 3 dimensions to be shown. If n=0, 100 frames of the simulation in time beginning with the assertion of the impulse are displayed. This is the most educational part of TLM. n= any number 1 to 100 will cause that particular frame to be displayed only.
scatter	This command enables the scattering matrix computation. Input and output must be specified, for the S-plots to be displayed. Totally zero value functions will not be plotted.
input, z,x	This command indicates the location of the input coordinates.
output, z,x	This command indicates the location of the output coordinates.
limit f,s	This command indicates the lowest initial value f in $\Delta l/\lambda$. The step size is s. The program calculates 100 steps automatically.
end	This command ends the instruction section.
boundary	This command starts the boundary section.
z z,xl,xh	This command defines the boundary limits. The first is the location where the impulse will be asserted. xl, xh are the low x coordinate and the high x coordinate boundaries.
end	This command terminates the circuit file.

Table 1. List of available commands.

File name	Variables	Explanation
EHI.dat	E(t),H(t)	Time values of E_y, H_x at the input port
EHO.dat	E(t),H(t)	Time values of E_y, H_x at the output port
EHR.dat	E(t),H(t)	Time values of E_y, H_x at the reference port
EMFEHI.dat	$\Delta l/\lambda, E_f, H_f$	Complex frequency values of E and H at the input port
EMFEHO.dat	$\Delta l/\lambda, E_f, H_f$	Complex frequency values of E and H at the output port
EMFEHR.dat	$\Delta l/\lambda, E_f, H_f$	Complex frequency values of E and H at the reference port
EMZEHI.dat	$\Delta l/\lambda, \|E/H\|$	Frequency values of the input port impedance
EMZEHO.dat	$\Delta l/\lambda, \|E/H\|$	Frequency values of the output port impedance
MAGS11.ps	|S11|	Postscript plot file of the magnitude of S11
ARGS11.ps	argS11	Postscript plot file of the phase of S11
MAGS21.ps	|S21|	Postscript plot file of the magnitude of S21
ARGS21.ps	argS21	Postscript plot file of the phase of S21

Table 2. ASCII files produced by the simulation.

and similarly, impulses of magnitude 1/2 are launched in the terminating lines as shown in Figure 2b.

Thus, if Vi^1_k and Vr^1_k denote the incident voltage and the reflected voltage at k node line 1 respectively, then at node k+1, the reflected voltages of line n can be written as:

$$Vr_n^{k+1} = \frac{1}{2}\left[\sum_{m=1}^{4} Vi_m^k\right] - Vi_n^k \qquad (6)$$

where,

$$Vi_1^{k+1}(z,x) = Vr_3^{k+1}(z,x-1)$$
$$Vi_2^{k+1}(z,x) = Vr_4^{k+1}(z-1,x)$$
$$Vi_3^{k+1}(z,x) = Vr_1^{k+1}(z,x+1)$$
$$Vi_4^{k+1}(z,x) = Vr_2^{k+1}(z+1,x)$$

For open circuit boundaries, the reflection component of a point inside the boundary is mirrored by the same value immediately outside a boundary. For example, a point (i,j) just outside the boundary of j–0.5, is:

example1.dat	example2.dat
start	start
movie	scatter
0	input
input	14,10
14,6	limit
limit	0.001,0.0001
0.002,0.0002	movie
end	75
boundary	output
z	45,10
4,2,10	end
z	boundary
24,2,10	z
end	4,2,19
	z
	30,8,12
	z
	46,8,12
	end

Table 3. Data files for example 1 and example 2.

$$Vr_1^k(i,j) = Vr_3^k(i,j-1) \qquad (7)$$

likewise, for boundary of i–0.5, the same point (i,j) just outside the boundary, is:

$$Vr_2^k(i,j) = Vr_4^k(i-1,j) \qquad (8)$$

The relationship indicated above, allows the propagation voltage values at each node to be calculated throughout the mesh points within the boundaries. Now the values of E_y and H_x can also be evaluated:

$$E_y^k(m,n) = 1/2[Vi_1^k(m,n) + Vi_2^k(m,n) + Vi_3^k(m,n) + Vi_4^k(m,n)] \qquad (9)$$

and,

$$H_x^k(m,n) = Vi_4^k(m,n) - Vi_2^k(m,n) \qquad (10)$$

The corresponding frequency response S(f) can now be obtained by FFT in the following manner:

$$S\left(\frac{\Delta l}{\lambda}\right) = \sum_{k=1}^{N} S^k \cos\left(\frac{2\pi k \Delta l}{\lambda}\right) + j\sum_{k=1}^{N} S^k \sin\left(\frac{2\pi k \Delta l}{\lambda}\right) \qquad (11)$$

where S^k is the time function of E_y or H_x generated before. $j=\sqrt{-1}$, N is the total number of time intervals and Δl is the physical length between two nodes.

$\Delta l/\lambda$	$|\eta|$ simulated	$|\eta|$ calculated
0.002	0.98712	0.9747
0.004	0.89798	0.9077
0.006	0.79714	0.8185
0.008	0.71550	0.7256
0.010	0.65234	0.6414
0.012	0.59169	0.5731
0.014	0.53148	0.5255
0.016	0.48980	0.5018
0.018	0.48909	0.5057

Table 4. Comparison of results.

In this way, the complete solution of E_y and H_x within the simulation boundaries are now known. Therefore, the traditional definitions of important circuit parameters such as impedance, S11 and S21 can be applied directly. S22 is S11 running in the –z direction. In this program, the above computations are incorporated for a computation area of 20 Δl in the x direction and 50 in the z direction of propagation. A unity magnitude impulse is automatically asserted along the x co-ordinate at the lowest value z boundary. Further extension of the simulation area can be easily made by the interested readers.

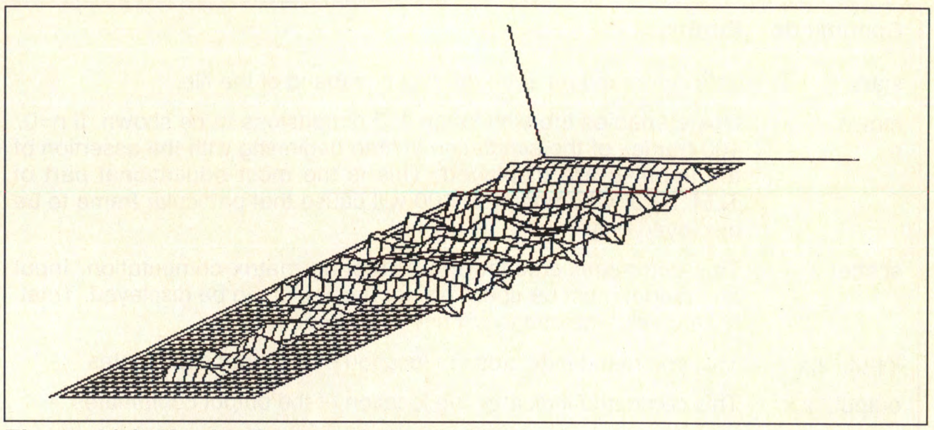

Figure 3(a). Movie display of simulaltion.

The simulation program begins by asking for a simulation circuit file name. The construction of the circuit file is divided into two sections. Section one contains the simulation instructions. The commands in this section can be placed in any sequential order. Section two contains the boundaries. In this case, the boundaries must be entered in the ascending order of the z co-ordinate. Table 1 shows the available commands. Each command starts on column one of the ASCII program file, there must not be any leading spaces.

This simulation program is computation intensive, one should use this program on a machine which is better than a 386 25 MHz class IBM PC/AT, otherwise the computation is painfully slow. VGA video is required. The simulation

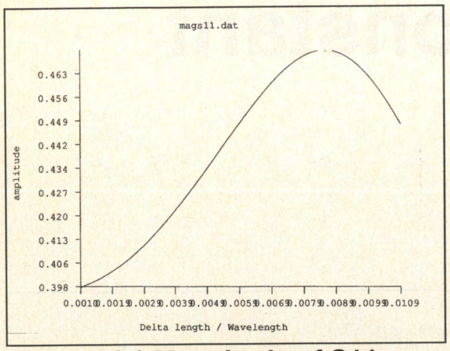

Figure 3(b) Magnitude of S11.

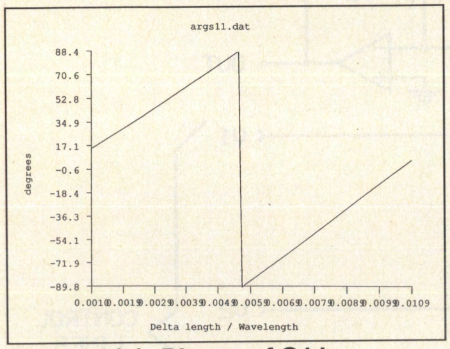

Figure 3(c). Phase of S11.

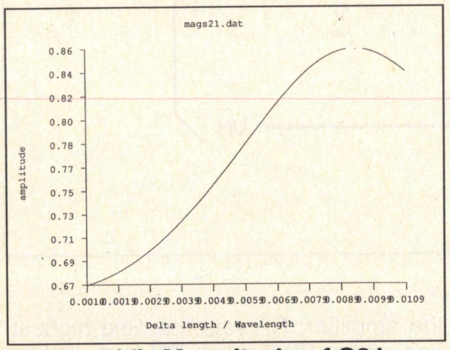

Figure 3(d). Magnitude of S21.

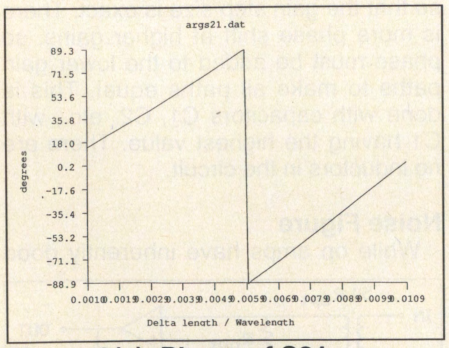

Figure 3(e). Phase of S21.

will produce the ASCII files listed in Table 2 for the convenience of further investigation.

An interesting simulation example used in [1] is shown here of a TEM

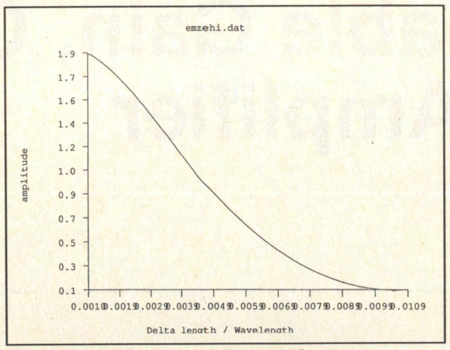

Figure 3(f). Input Impedance.

wave on a 25×11 rectangular matrix. The field boundaries are x=2, x=10, at z=4 and z=24 with input/output at z=14, x=6. Using the command set above, EXAMPLE1.DAT circuit file is generated (Table 3).

The theoretical solution for EXAMPLE1.DAT can be obtained very simply by using the relationship established in Equations (3) and (4) and the standard transmission line equation such that,

$$\eta = \eta_s \frac{1 - j\eta_s \tan\left(2\sqrt{2}\pi l_s \frac{\Delta l}{\lambda_s}\right)}{\eta_s - j\tan\left(2\sqrt{2}\pi i_s \frac{\Delta i}{\lambda_s}\right)} \qquad (12)$$

where,

$\eta_s = 1/\sqrt{2}$, $l_s = \sqrt{2}\lambda$, $l_s = 10.5$

the distance between z=14 of the input port and z=24.5 of the boundary. From EMZEHI.dat, the comparison of the results are shown in Table 4.

EXAMPLE2.DAT uses all the commands available to show a transmission line step discontinuity with more complicated boundaries. At z=4, the width is bounded by x=2, x=19. At z=30, x=8, x=12 and the boundary terminates at z=46, x=8 and x=12. The input port resides at z=14, x=10, and the output port, z=45, x=10. Only the 75th frame of

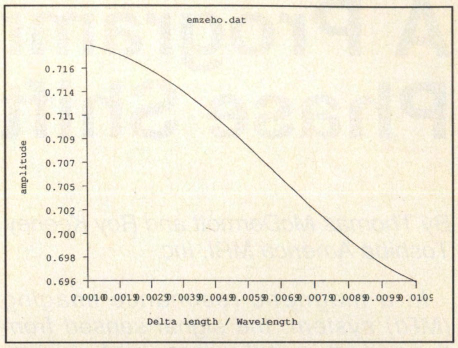

Figure 3(g). Output Impedance.

the simulation is shown to save computation time. Figures 3(a)-3(g show the displays for this file. However, it is interesting to see the full "movie" by replacing 75 with 0.

The purpose of this tiny simulation program is to provide a means of visualizing the solution of electromagnetics in terms of time domain impulse response, which is less abstract and more intuitive than other methods. ***RF***

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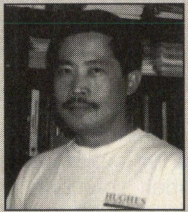
1. P. B. Johns and R. L. Beurle, *Numerical Solution of 2- Dimensional Scattering Problems Using a Transmission-line Matrix*, Proc. IEE, 1971, 118, pp. 1203-1208.

About the Author



Dr. Jonathon Y.C. Cheah is currently the project engineer of the Hughes Network Systems Cellular Digital Packet Data (CDPD) Mobile End System (MES) development. He can be reached at Hughes Network Systems, 10450 Pacific Center Ct., San Diego, CA 92121, tel. (619) 452-4847.

A Programmable Gain, Constant Phase Shift Amplifier

By Thomas McDermott and Roy Keeney
Toshiba America MRI, Inc.

In a magnetic resonance imaging (MRI) system, the signal sensed from the patient is at the resonant frequency of protons for the particular field strength of the system's magnet (1). Typical resonant frequencies for clinical MRI systems are from about 2 to 64 MHz, which correspond to 0.06 to 1.50 Tesla magnets. The RF receiver mixes the signal down from the resonant frequency to DC where it is converted to digital form and read by the system computer. An MRI receiver must have variable gain to accommodate a wide range of input levels. However, as the gain is varied, the phase shift through the receiver must be kept constant. This paper describes a programmable gain, constant phase shift amplifier for use in an MRI receiver.

The goal for the entire receiver is to switch gain over a 63 dB range in 1.0 ±0.2 dB steps, with a phase change of less than ±1.5 degrees. This can be achieved with several cascaded amplifiers. Further, the amplifiers must have noise and distortion characteristics that allow the receiver to have the highest possible dynamic range.

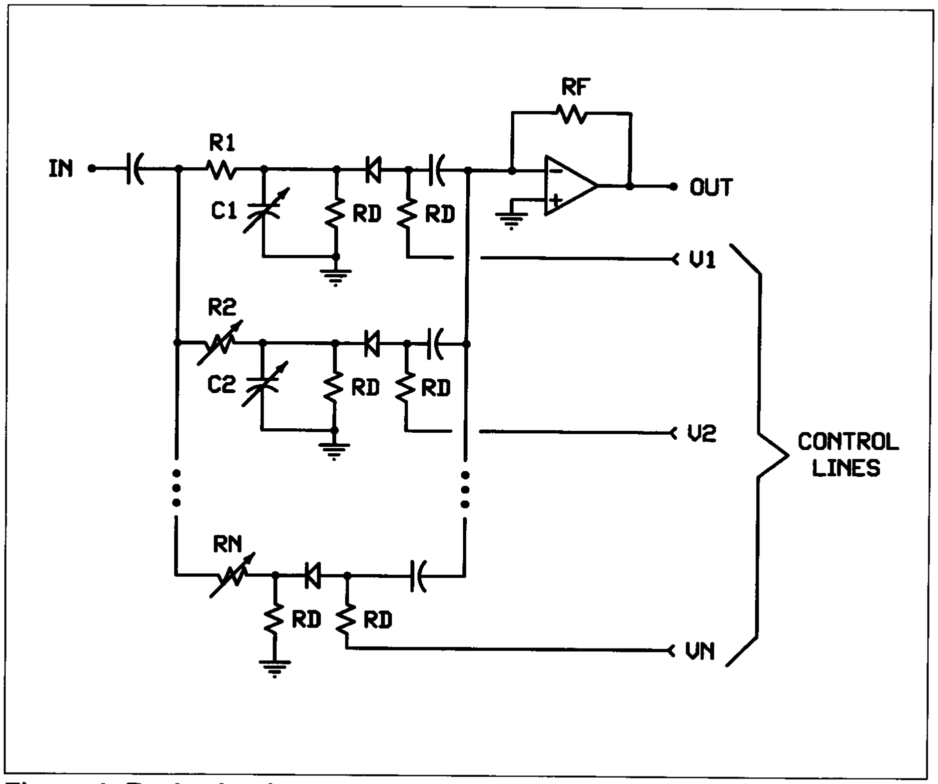


Figure 1. Basic circuit.

Basic Circuit

Several options exist for implementing this function. Off-the-shelf solutions are available, but typically their bandwidth greatly exceeds what is needed in MRI, and at several hundred dollars, can be quite expensive for this application. A straightforward way to change signal level is to switch an attenuator in and out of the signal path. This is commonly done with two SPDT switches (diode or relay). However, switching 63 dB of attenuation in and out can have a noticeable effect on the receiver noise figure. Switching an amplifier in and out of the signal path is one alternative. MMIC amplifiers would be simple and inexpensive to use, but their power capability would limit the overall receiver distortion performance. A transistor amplifier might overcome this, but can be unnecessarily complicated.

Finally we settled on an op amp circuit in which the gain itself is switched, as opposed to the signal path. This allows multiple bits of gain control per amplifier. An op amp is simple and inexpensive, can handle high signal levels, and can operate in the low MHz frequency range. As will be discussed in the next section, there is a way to compensate for its relatively poor noise performance.

Figure 1 shows the basic circuit. An op amp is configured as an inverting amplifier with multiple gain-determining resistors (R1, R2, ..., RN). Current from a control line flows through resistors RD to turn on a diode, thus selecting a gain. The resistance of the diode separates the upper end of the variable capacitor from AC ground, allowing phase adjustment with a small capacitor. The diode resistance simply adds to the gain-determining resistor for gain calculations. R1 sets the lowest gain, G1, for the amplifier; R2 sets the next highest, and so on. R2 through RN are adjusted to give the desired gain relative to G1, so that the gain step size is exact. There is more phase shift at higher gains, so phase must be added to the lower gain paths to make all paths equal. This is done with capacitors C1, C2, etc., with C1 having the highest value. There are no inductors in the circuit.

Noise Figure

While op amps have inherently good

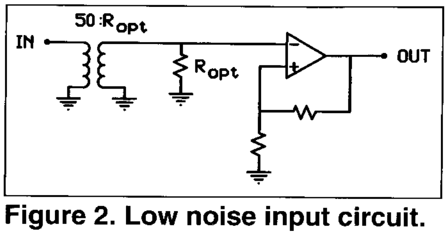


Figure 2. Low noise input circuit.

800 to 2400MHz, to +26dBm from **$199**

In a broadband amplifier, excellent low-noise performance is generally synonymous with low power output because of design compromises. Not so with Mini-Circuits' TO-, ZEL-, and ZHL- low-noise amplifier series, where a noise level less than 1.5dB is accompanied with up to +26dBm power output. In front-end applications, it is undesirable for the amplified output to appear back at the input; these amplifiers effectively isolate the output signal from the input by as much as 35dB and use the shortest possible lead lengths to minimize parasitics and optimize NF performance. Detailed performance specs are included in our 740-pg RF/IF Designer's Handbook.

Low-noise amplifiers are available in a rugged hermetically-sealed TO-8 package, or in a tiny (less than one cubic inch) EMI-shielded case, or a hefty EMI-shielded case for high power models.

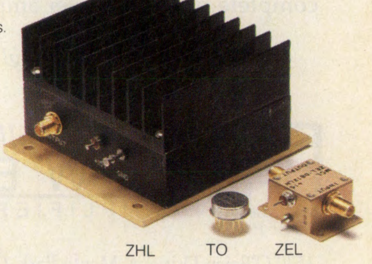
Available from stock, priced from $199 with better than 1.5dB NF performance, there's lots to shout about.

SPECIFICATIONS

Pin Model	TO 0812LN	TO 1217LN	TO 1724LN			
Connector Version	ZEL 0812LN	ZEL 1217LN	ZEL 1724LN	ZHL 0812HLN	ZHL 1217HLN	ZHL 1724HLN
Freq. (GHz)	0.8-1.2	1.2-1.7	1.7-2.4	0.8-1.2	1.2-1.7	1.7-2.4
NF, db, max*	1.6 1.5	1.6 1.5	1.6 1.5	1.5	1.5	1.5
Gain dB, min.	20	20	20	30	30	30
Output Pwr., dBm 1dB Comp.	+8	+10	+10	+26	+26	+26
Intercept Pt. 3rd order, dBm typ.	18	25	22	36	36	36
Price $ (Qty. 1-9)	TO $199.00		ZEL $274.00		ZHL $349.95	

NOTES:
1. NF max. at room temperature. Increases to 2 dB typ. at +85°C.
2. SMA connectors only, ZEL and ZHL units.
3. Operating temperature: -54°C to +85°C.
4. DC power, 15V, 70mA for TO and ZEL, 15V, 725mA for ZHL.
5. VSWR 2.5:1

*TO-8 includes test fixture loss.



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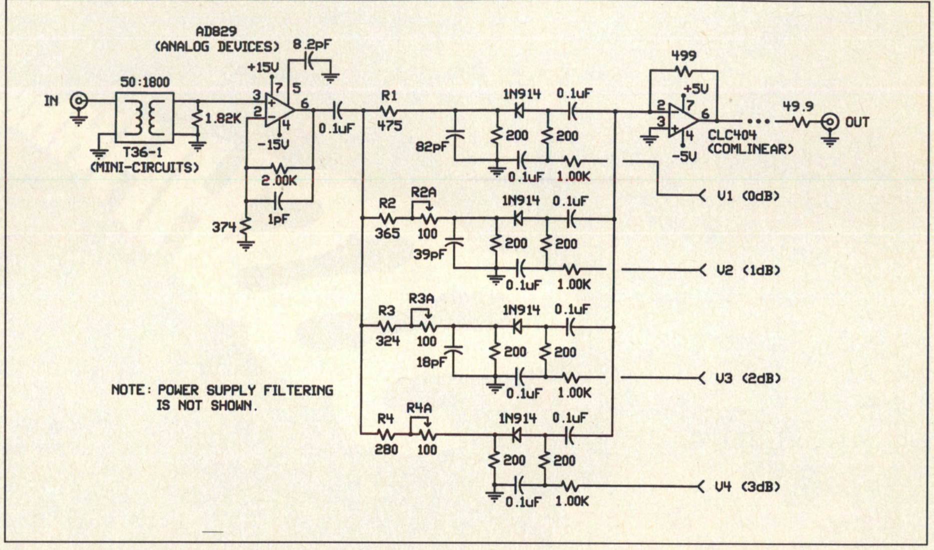

Figure 3. Programmable gain, constant phase shift amplifier with low noise input circuit.

capability to handle high signal levels with little distortion, their noise figure can be quite poor. A typical noise figure might be as high as 20 dB for an inverting amplifier. Since it is easier to maintain constant phase at lower frequencies, we decided to place the programmable gain amplifiers in the receiver's final IF strip, which is at 2 MHz. Towards the back end of a receiver, noise characteristics play a less critical role than distortion characteristics in determining overall dynamic range, so the op amp lends itself well to this application. To keep the overall noise figure of the receiver low, we placed a high gain, low noise amplifier immediately before the programmable amplifiers.

We used an op amp for this amplifier as well, as shown in Figure 2. The noise figure of an op amp circuit is determined not only by the amplifier equivalent input noise, but also by the source resistance from which the amplifier is driven, and by the feedback resistor values (2). To minimize noise figure and to match the amplifier to a 50 Ohm system, we transformed the driving impedance to an optimum value for the chosen op amp.

Final Circuit

Figure 3 shows the final circuit design. The transformer and first op amp comprise the low noise input circuit. The AD829 was selected for its very low equivalent input noise voltage and current (2.0 nV/Hz, 1.5 pA/Hz). A transformer with a turns ratio of 6 provides the desired source resistance of 900 ohms to the op amp. The gain of this section is +14 dB due to the transformer (including losses) plus 16 dB due to the op amp. The noise figure of this section alone is about 5 dB.

The second section can have a gain of 0, 1, 2, or 3 dB. The CLC404 was select-

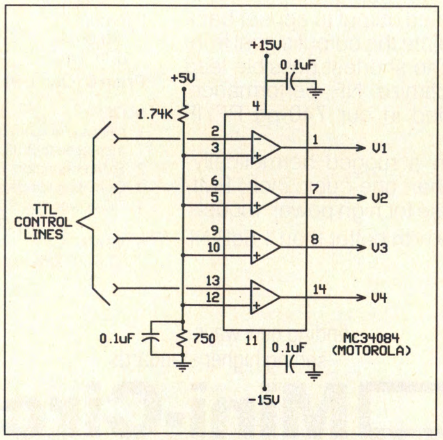

Figure 4. Diode driver circuit.

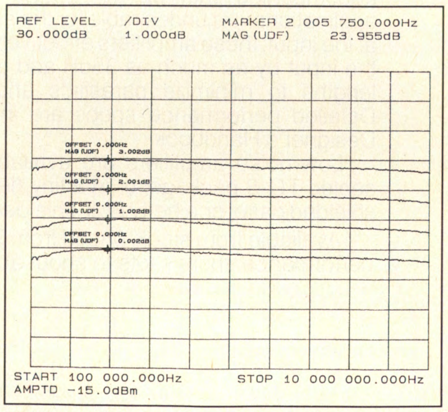

Figure 5. Gain vs. frequency for gain settings of 0, 1, 2, 3 dB.

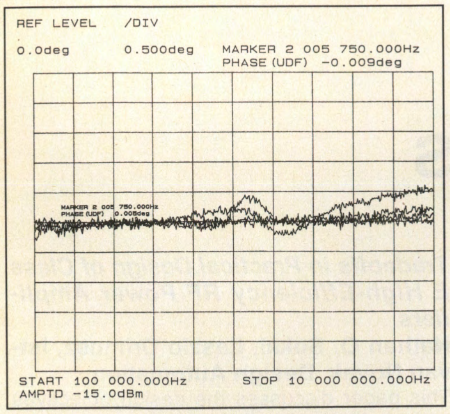

Figure 6. Phase differential vs. frequency for gain settings of 0, 1, 2, 3 dB.

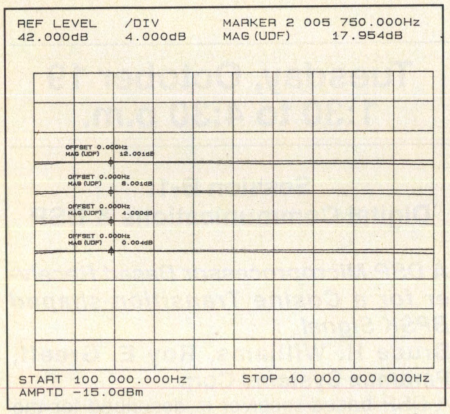

Figure 7. Gain vs. frequency for gain settings of –6, –2, 2, 6 dB.

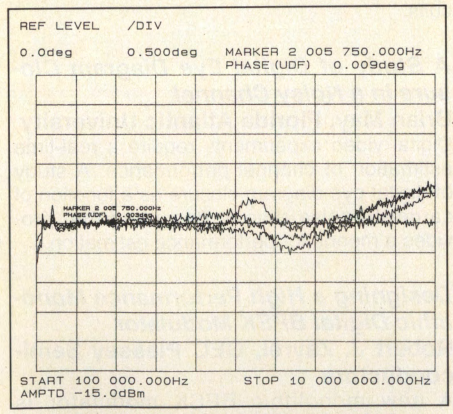

Figure 8. Phase differential vs. frequency for gain settings of –6, –2, 2, 6 dB.

ed for its high slew rate (2600 V/μs), which keeps phase shift to a minimum. The diode is an ordinary silicon diode which is turned on with about 10 mA of current. The "on" resistance of the diode is about 10 Ohms which is incorporated into the calculation of gain. The values of the phase compensation capacitors are highly dependent on the PC board layout, which should be as tight as possible. We hand-picked fixed value capacitors in the final implementation of the circuit. Several CLC404 sections can be cascaded, each with different gain step sizes, to cover a wide range. For example, a four bit binary word can be decoded into eight control lines: four for a 0, 1, 2, 3 dB section and four for a 0, 4, 8, 12 dB section to give a range of 16 dB in 1 dB steps. Other gain steps are also possible, as shown by the resistor values in Table 1. Because of control line decoding and circuit parasitics, there is a practical limit of two to four diode circuits per op amp. The 49.9 ohm series resistor at the output transitions back into a 50 Ohm circuit, with a gain of –6 dB. Figure 4 shows a driver circuit in which a comparator senses the TTL control lines and supplies current to one diode.

GAIN (dB)	R1	R2	R2A	R3	R3A	R4	R4A
0,1,2,3	475	365	100	324	100	280	100
–6,–2,2,6	953	536	100	324	100	205	50
–8,8	1.21K	158	50	X	X	X	X
0,16	475	52.3	50	X	X	X	X

Table 1. Resistor values for various gain configurations. "R" designators correspond to those in Figure 3.

Figures 5 and 6 show gain and phase verses frequency for gain settings of 0, 1, 2, and 3 dB. Figures 7 and 8 show the same for gain settings of –6, –2, 2, and 6 dB. In both cases, the circuit maintains accurate gain step size and constant phase shift at 2 MHz within the measurement ability of the network analyzer. Even at 10 MHz, the phase differential is about 0.5 degrees. Noise figure for an AD829 section followed by one or more CLC404 sections is under 7 dB.

Summary

A programmable gain, constant phase shift amplifier for use in an MRI receiver was presented. The circuit is based on an inverting op amp amplifier with diode-switched gain-determining resistors and phase compensating capacitors. By cascading several of these amplifiers, gain can be switched over a 63 dB range in accurate 1 dB steps. Phase variation from gain setting to gain setting stays within 1.5 degrees. A transformer input circuit keeps the overall noise figure low. The inherent high intermodulation intercept point of op amps allows the circuit to be used near the back end of the receiver where signal levels are highest. The circuit presented provides a simple and inexpensive, yet stable and accurate solution to a fundamental problem in MRI receivers.

Acknowledgments

The authors gratefully acknowledge the assistance of Ernest Greenwald and Bonifacio Capistrano in the development of this project. *RF*

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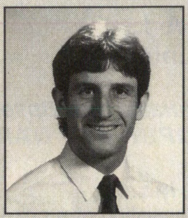
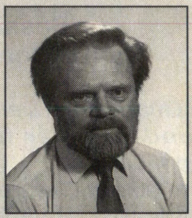
1. T. Farrar and E. Becker, *Pulse and Fourier Transform NMR*, Academic Press, 1971.
2. M. Steffes, "Improving Amplifier Noise Figure for High 3rd Order Intercept Amplifiers", Comlinear Corporation Application Note OA-14.

About the Authors

Tom McDermott is an RF design engineer and received BS and MS degrees in electrical engineering from Rensselaer Polytechnic Institute. He has worked in MRI electronics design for seven years.

Roy Keeney received his BSEE degree from Western States College of Engineering in Los Angeles in 1982. He is the manager of the RF engineering group at Toshiba, where he has been employed for the past one and a half years. He previously worked at Eaton Corporation's Electronic Instrumentation Division for 20 years in Los Angeles. His specialty is in low noise synthesizer designs and noise figure instrumentation.

They can be reached at Toshiba America MRI, 280 Utah Ave., South San Francisco, CA 94080, or by phone at (415) 872-2722.

RF Expo East Technical Paper Abstracts

The RF Expo East engineering conference and product exhibition will be held in Tampa, Florida, October 19-21. Full-day short courses will also be held each day from October 18-21. Once again, top RF engineers will be presenting papers covering all aspects of technology — from basic to advanced, from universal topics to specialized applications.

A special addition to this year's RF Expo East is RF Expo PLUS, a single track dedicated to one topic. This year, Commercial Space Applications is the special subject. Each morning will include an RF Expo Plus session in addition to the usual full schedule of technical papers.

Tuesday, October 19 8:30 to 11:30 a.m.

Session A-1: Digital Communications

A Quadrature Demodulator and Baseband Sampler for QPSK and QAM Applications
Jim Marsh, Tektronix, Inc.
A fully integrated ASSP consisting of a quadrature demodulator and dual baseband digitizers for QPSK and QAM applications is presented. RF input is in the 400-700 MHz range with greater than +5 dBm input IP3.

Integrated Modem/RF Design Architectures for Reduced Power, Increased Capacity F-QPSK Wireless Systems
Kamilo Feher, University of California, Davis
Integrated modem and RF design architectures for F-QPSK wireless systems applications are presented. Direct baseband to RF and RF to baseband (zero-IF) coherent and discrimination detected designs are compared.

Session A-2: System Performance

Methods for Estimating and Simulating the Third Order Intercept Point
Carl Stuebing and Mojy C. Chian, Harris Semiconductor
Circuit nonlinearity is commonly characterized by third order intercept. This paper reviews the Pip3 concept and reviews methods to estimate and simulate it.

Low Cost Phase Noise Measurement Technique
Jim H. Walworth, Tampa Microwave Lab, Inc.
This paper describes a proposed measurement technique which utilizes a phase locked oscillators's own VCO curve and a low frequency spectrum analyzer to measure FM noise and convert that measurement to phase noise.

Externally-Induced Transmitter Intermodulation: Measurement and Control
Ernie Franke, E-Systems ECI Division
Power amplifiers are typically designed for efficiency and output power, not linearity. This paper describes a wideband technique for evaluation of externally-induced intermodulation performance, necessary for amplifiers operating near other high power transmitters.

Session A-3: Amplifier Design

Tutorial on Current-Feedback Amplifiers
Anthony D. Wang, Burr-Brown Corporation
The history of current-feedback amplifiers, circuit operation and comparisons with voltage-feedback amplifiers begin this tutorial. Design considerations, data sheet specs and examples illustrate these devices' usage.

The SLAM: A New Ultralinear Power FET Module Concept for HF Applications
Adrian I. Cogan, Lee B. Max, MicroWave Technology, Inc.
A new FET technology for low cost HF ultralinear power amplifier blocks is described. The SLAMs are self-biased for minimum complexity in supporting circuitry.

Tradeoffs in Practical Design of Class E High-Efficiency RF Power Amplifiers
Nathan O. Sokal, Laszlo Drimusz, Istvan Novak, Design Automation
This paper discusses the design objectives and resulting tradeoffs for efficiency, harmonic content, power output and component losses in class D and E high efficiency power amplifiers.

Tuesday, October 19 1:30 to 4:30 p.m.

Session B-1: Digital Communications & DSP

A DSP Microprocessor Based Receiver for a Cosine Transition-shaped BPSK Signal
Bruce H. Williams, Roy E. Greeff, Paramax Systems Corp.
A DSP based receiver is described for the cosine transition-shaped waveform. The carrier recovery loop and matched filter functions are performed by TI TMS320C40 DSP circuits.

A Study of Digital Eye Diagram Closure in a Noisy Channel
Brian May, Florida Atlantic University
Digital video experiments require a real-time estimation of channel performance. A study of digital eye diagram closure as a function of carrier-to-noise ratio is described that provides a means for performance estimation.

Designing a High Performance Monolithic Digital BPSK Modulator
Robert J. Zavrel, GEC Plessey Semiconductors
A new monolithic BPSK modulator is described, emphasizing the functions performed, measured specifications, applications and additional devices for future development.

Session B-2: Test Methods and Equipment

A 3 GHz 50 ohm Probe for PCB Measurements
Joel Dunsmore, Robert Kornowski,

Chuck Tygard, Hewlett-Packard Co.
This paper describes a 3 GHz probe for making connections to printed circuit boards using spring loaded contacts for signal and ground contacts.

Low Cost RF Tuner System for JDC Load Pull and SSPA Design
C. Tsironis, Focus Microwaves Inc.
A new RF tuner system is presented with load/source pull capability for gain, power, efficiency and IMD. The load pull contours are processed by analysis and optimization software for accurate designs.

Session B-3: RF Power

High Power, Low Frequency Microstrip Switches
S. Irons, E. Higham, M/A-COM
This paper describes the design concept and results for a family of microstrip compatible switches operating at 150 watts CW in the 30-88 MHz range.

Class-E Power Amplifier Delivers 24 W at 27 MHz at 89-92% Efficiency, Using One $1.05 Transistor
Nathan O. Sokal, Ka-Lon Chu, Design Automation
A circuit is described which illustrates the performance of a class E amplifier at high frequencies, where the switching transition time can be as much as 15 percent of the period.

The CAM: A UHF/L-Band FET Module for Pulsed Power Avionics Applications
Frank Sulak, Ken Sooknanan, Adrian I. Cogan, MicroWave Technology, Inc.
Very small size, high pulsed power, broadband and narrowband amplifier modules for avionics applications are described. These modules use Solid State Triode silicon FETs for 50-150 watt peak power in the 800-1250 MHz range.

Wednesday, October 20 8:30 to 11:30 a.m.

Session C-1: Wireless Communications Systems

Frequency Synthesizer Strategies for PCS
Bar-Giora Goldberg, Sciteq Electronics, Inc.
This paper identifies some of the general requirements of Personal Communications Systems, then discusses specific frequency synthesis issues and mechanisms to address them.

TDMA Transmitters — Characterizing Power, Timing and Modulation Accuracy
Helen Chen, Hewlett-Packard Co.
This paper explores the test requirements of digital RF communication transmitters, presenting practical techniques that reduce system development time and installation costs. Examples from the NADC, JDC, GSM and CT-2 systems are presented.

Performance Simulation of a Low-Power In-Building Wireless Centrex System
Douglas Alston, BellSouth Telecommunications
A software simulation program is described which models the performance of an advanced prototype in-building wireless centrx system based on CT2 radio technology.

Session C-2: RF Applications

FMCW Radar Architecture
Ken Puglia, M/A-COM
The architecture of the basic FMCW ranging radar is investigated from the antenna input to the digital signal processor output. Microwave component requirements are discussed relative to system operation.

Filter Comparator Network for Beam Position Monitoring
Michael Ferrand and Mark McWhorter, Lorch Electronics
This paper describes a filter/comparator developed for Argonne National Laboratories for use in the Advanced Photon Source, providing boresight accuracy measurements in a particle accelerator ring.

Digital Temperature Compensation of Oscillators Using a Mixed Mode ASIC
Steve Fry, Murata Electronics North America
This paper discusses the design, construction and operation of crystal oscillators which are digitally temperature compensated using a custom mixed mode ASIC.

Session C-3: Modeling for CAD

RF Active Device Modeling for CAD, A Coming Necessity
Gary Roberts, Hewlett-Packard Co.
The need for accurate, timely active device models has grown as the use of CAD tools for RF design has spread. Various types of models are discussed, along with model libraries for commercial CAD tools.

Regression Based Algorithms for Inductor Modeling
Edmund (Joe) Tillo, Ford Motor Company
This paper describes inductor models developed using generalized linear regression, with data obtained using either an impedance meter or Q-meter. Resulting models accurately predict self-resonance, Q and impedance versus frequency.

RF Expo *PLUS*: Special Emphasis on Commercial Space Applications

RF Expo PLUS, a new addition to RF Expo East, is a special focus on a single technology. The subject for this year's conference is Commercial Space Applications. Look for detailed information on these papers in the October issue.

Expo *PLUS* Session 1 — Satellite and Space Systems
"An Overview of Current Commercial Satellite Systems"
"Global Positioning System — A Review"
"New Space Systems for Voice and Data Communications"

Expo *PLUS* Session 2 — Components for Space Applications
"Designing Microwave Circuits for Geosynchronous Space Applications"
"A Lightweight Plated Plastic Filter for Space Applications"
"High Power Amplifier Module for Satellite Transponder"

Expo *PLUS* Session 3 — Satellite and Space System Performance
"Unique Performance Requirements (and Limitations) in the Space Environment"
"Estimate of Channel Capacity of a Satellite Link in the Presence of Rain Attenuation"
"Hardware Verification of Communication System Simulations"

Computer Aided Design Tools for RF Circuits

Michael Rothery, Sam Ritchie, Madjid A.Belkerdid, University of Central Florida

A PC-based computer-aided design tool set is presented, for both small amplifiers and oscillators. Emphasis is on circuits for VHF and UHF, including matching, stability and other key parameters.

Wednesday, October 20 1:30 p.m. to 4:30 p.m.

Session D-1: Components for Wireless

Practical Applications of a Low Cost Low Noise GaAs PHEMT MMIC for Commercial Markets

Al Ward and Henrik Morkner, Hewlett-Packard Co.

A high performance broad band packaged GaAs MMIC for 1.5 to 8 GHz applications is described. The PHEMT device provides 20 dB gain, 2 dB noise figure and +7 dBm power output.

Highly Integrated GaAs MMIC RF Front End for PCMCIA PCS Applications

Thomas Kotsch, Andy Laundrie, Steve Geske, Howard Fudem, Jim Blubaugh, Sanjay Moghe, Northrop Corp.

This paper describes the technologies and development activities for a wideband 800 MHz to 1.8 GHz fully integrated RF front end using GaAs MMIC and multilayer MIC techniques to achieve PCMCIA card size.

Low Power Transmitter Design Using SAW Devices

Earl Clark, RF Monolithics, Inc.

This paper discusses the design of low power pulse modulated SAW stabilized transmitters, with discussion of the differences in requirements between U.S. and European markets.

Session D-2: Synthesizers

A Synthesizer Design Program With Detailed Noise Analysis

Terrence Hock, National Center for Atmospheric Research

The software winner in the 1993 RF Design Awards Software Contest describes a program which aids in the design and analysis of synthesizers using current monolithic PLL ICs.

PLL Settling Time: Phase vs. Frequency

Donald E. Phillips, Rockwell International

Analysis of PLL phase settling time versus frequency settling time is presented, along with a discussion of the significance of frequency to phase error ratio.

Digital Linear FM: Synthesizing a New Frequency Each Two Nanoseconds

Bar-Giora Goldberg, Sciteq Electronics, Inc.

This paper compares an ideal chirp with the capability of specialized VCOs and with digital synthesizers. It then explores the generation of such digital chirps and a chirp synthesizer project with the Army and Sandia National Labs.

Session D-3: CAD Methods

SAW Resonator Oscillator Design Using Linear RF Simulation

Alan R. Northam, RF Monolithics, Inc.

The designer faces two problems when designing a SAW oscillator: how to connect the SAW resonator and how to model the oscillator. This paper describes the two connection options and shows how linear analysis can be used to closely approximate actual performance.

Embedding RF Design Tools in an IC Design System

Mojy C. Chian, Steve S. Majors, Alan G. Whittaker, Harris Semiconductor

This paper describes efforts to enhance and modify an IC design framework to provide RF-specific design and data representation capabilities for development of silicon ASICs.

Electromagnetic Simulation of Arbitrary, Multilayer Planar Patterns

Charles Plott, Hewlett-Packard Co.

The goal of this paper is to convey an understanding of currently available electromagnetic technology and discuss realistic applications. The basics of the Method of Moments technique are presented, highlighting the most recent advances.

Thursday, October 21 8:30 to 11:30 a.m.

Session E-1: Wireless Applications

A Monolithic 915 MHz 20 dBm Direct Sequence Spread Spectrum Transmitter

Stephen Press, Tektronix, Inc.

This paper reports on the design and measured results for a 915 MHz Direct Sequence Spread Spectrum transmitter. This monolithic design was implemented on a bipolar process for all RF circuit elements.

A Low Power RFID Transponder

Raymond Page, Wenzel Associates, Inc.

This minimum-complexity transponder circuit is powered by an illuminating signal, returning data on a carrier at twice that frequency. This is the design winner in the 1993 RF Design Awards Contest.

Session E-2: RF System Topics

Multi-Component Module for High Speed Passive Design

Mark Brooks, Thin Film Technology

A novel means for making multilayer ceramic is described which facilitates design of multiple passive elements, bonded mechanically and electrically into a single package.

Design of a Search Based PLL
Mike Black, Texas Instruments
Design of a phase locked loop is approached from the standpoint of the search mechanism; how the VCO is moved close enough to the input signal for acquisition and subsequent phase-locking.

Analysis of Transversely Coupled SAW Resonator Filters Using COM Techniques
V.Narayanan and S.M. Ritchie, University of Central Florida
This paper describes the COM analysis technique and how it allows straightforward analysis of SAW device elements such the reflective grating and interdigital transducer.

Session E-3: Practical RF CAD

Basics of CAD at RF for Wireless Circuits and Subsystems
Compact Software Staff
Linear, nonlinear and system design tools are described for designing circuits and systems for new wireless applications.

Defining Circuit and Subsystem Specifications for Cordless Telephone Applications
Compact Software Staff
A description of system-level simulation methods, with the specific example of a cordless telephone design.

Design Examples of Small-Signal Circuits Operating from 3.3 to 4.5-Volt Supplies
Compact Software Staff
Amplifiers and switches are the emphasis in this small-signal CAD tutorial.

Design Examples of Large-Signal Circuits
Compact Software Staff
Oscillators, mixers and power amplifiers require specific large-signal models. Technology choices are reviewed for efficiency, low noise and low phase-noise.

Using CAD for Circuit Layout and Packaging Design
Compact Software Staff
Modeling circuits exactly as they are to be built is the topic of this paper, including prediction of performance changes with temperature.

RF Expo East Short Courses

Digital Modulation and Spread Spectrum (for Personal Wireless Communications)
October 18 – Dr. Kamilo Feher, Instructor

Filter and Matching Network Design: L-C and Distributed Circuits – HF to Microwaves
October 18 – Randy Rhea, Instructor

Introduction to RF Circuit Design, Part I: Fundamental Concepts
October 19 – Dr. Robert Feeney and Dr. David Hertling, Instructors

Introduction to RF Circuit Design, Part II: Active Circuit Design
October 20 – Dr. Robert Feeney and Dr. David Hertling, Instructors

Oscillator Design Principles
October 21 – Randy Rhea, Instructor

Manufacturability Shapes Boards and Substrates

By Andy Kellett
Technical Editor

Only rarely do printed circuit boards and hybrid substrates limit RF designs. However, PC boards and substrates do have a bearing on the manufacturability, reliability and price of RF devices. While most of the materials remain the same, PC boards and substrates have changed in other ways in order to meet the demands of new RF designs.

The RF designs getting the most attention right now are for "wireless" communications. These devices, meant to be sold in high volumes and to be operated in the low GHz range, will require printed circuit boards that can hold lots of components, handle high frequencies and hold down costs. "This trend will be part of everyone's product, especially if we are to take away some of the business that Japan currently posses," says Charles Dexter, Staff Scientist at Watkins Johnson (W-J).

Manufacturing Demands

The goal of design-for-manufacturability is a minimum number of steps requiring individual attention. "The ideal product would be one huge pc board completely covered with surface mount components," says W-J's Dexter, "You could slap it between two pieces of metal, test it and ship it." Manufacturers are working towards this goal by using multilayer boards, by using more surface mounted components and by keeping microstrip and lumped element circuits on the same board.

The same devices that are to be manufactured in high volumes are also going to operate at higher frequencies. For example, the current generation of analog cordless phones operates around 46 MHz. The next generation will operate above 800 MHz. At these frequencies, stripline and microstrip techniques are used for filters and other frequency selective circuits. This means increased attention to the effects of circuit board and substrate parameters on circuit performance.

Perhaps the most strident demand heard by circuit board and substrate manufacturers is for lower cost. "The people that have been suppliers to this market for a number of years, like Rogers, are really focusing efforts on reducing cost of manufacture while maintaining electrical performance," observes Jim Carroll, Product Line Manager for Rogers' Substrate Division.

What's Available

FR-4's low cost makes it popular for low frequency designs. At higher frequencies, PTFE boards are used. PTFE's relatively high coefficient of thermal expansion and sodium etch requirements present problems for plated-through holes and multilayer boards. Vias can "pop" when exposed to expansion and contraction. However, PTFE/ceramic composites stabilize PTFE board performance both mechanically and electrically. PTFE/ceramic composites perform well but have traditionally been relatively expensive.

Materials with properties somewhere between PTFE circuit board materials and alumina circuit substrates are being developed to accommodate increasing chip-on-board construction. For instance, Temperature-stable Microwave Material (TMM™) from Rogers Corporation can be processed just like other plastic circuit board materials. In addition, TMM is a thermoset plastic/ceramic composite which is rigid enough to allow reliable wire-bonding and whose surface readily accepts metalization. The material also exhibits stable thermal coefficients of expansion and dielectric constant.

Alumina (Al_2O_3) is still the most prevalently used material for substrates. Available in several grades, alumina can be tailored for firing properties, hardness and other properties. "For instance, for applications where low dielectric loss is important, 99.6% grade alumina is used; for devices using buried conductors, 90 or 92% grade alumina is preferred," noted Chong-il Park, R&D Manager for Kyocera America.

Alumina must give way to other substrate materials when high thermal conductivity is needed. Berylia (BeO) substrates are very thermally conductive — and very expensive. Aluminum Nitride falls between Alumina and Berylia in both thermal conductivity and cost. "More and more people are interested in it," noted Paul Garland, Production Manger for Metallizing at Kyocera America's Print Division.

Design and Manufacture

Design and manufacturing changes have accompanied the changes in printed circuit boards and substrates themselves. Designers working on complex, multilayered, single-board devices must make sure different layers and even different areas on the same layer do not interfere with each other says W-J's Dexter. "Now the engineer sits down with the printed circuit designer who's looking at all the different layers and is saying 'beef this up' or 'make this smaller.'"

Tools are available to make the transition from lumped-element design to distributed element design easier. A distributed element filter design program from Eagleware allows designers to enter circuit board parameters to accurately design microstrip and stripline filters. There are even desk-top "prototyping" mills by manufacturers such as T-Tech and Capex which can take the computer generated designs and create accurate prototypes in minutes.

Selling More For Less

If there is one target printed circuit board manufacturers are aiming for, it is cost reduction. "The materials we have provide the performance that the emerging wireless market needs," observes Rogers' Carroll. "Price levels have historically been high, reflecting the low volumes associated with prior applications. As the market moves to higher volumes, the price will come down."

The manufacturers of substrates for hybrid circuits have a less well defined path to success. "There are a lot of alternative technologies for hybrid circuit manufacture, and no clear winners," says Kyocera America's Gibson. "I think a lot of people are picking their own direction, setting a standard of their own. So for the supplier end of it, it is a mess."

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RF software

Electromagnetic Analysis

MacNeal-Schwindler has released version 2.5 of two programs which provide electromagnetic analysis. MSC/EMAS for RF and Microwaves calculates modal fields and resonant frequencies for lossless and lossy cavities, S-parameters and losses for microwave devices, and EM fields, currents and energy densities. MSC/EMAS for Antennas calculates antenna input impedance, directivity, polarization and radiation patterns. Both programs can handle anisotropic and complex tensors.
MacNeal-Schwindler Corp.
INFO/CARD #182

Antenna Pattern Plotting

EEsof and Sonnet Software have announced two new products, patgen™ and patvu™, which provide antenna pattern analysis and plotting capability, respectively. This capability includes the calculation of radiation loss for non-surface wave antennas.
EEsof Incorporated
INFO/CARD #181

Antenna Calculations

The Antenna Specialists Co. has developed a set of programs to aid communications system designers and operators. Disk #1, titled DXPLOT™, permits precise calculation of beamtilt coverage. Disk #2, titled PATPLOT™, displays and plots digitized base antenna patterns. Disk #3, ANTPLOT™, develops patterns for sidemounted base antennas. The programs are available free-of-charge on 5.25 inch disks.
Antenna Specialists Co.
INFO/CARD #180

Phase Locked Loop Design

PLL3 Version 1.6 and PLL2 Version 1.3 are interactive tools for developing 3rd or 2nd order phased lock loop circuits. Graphic screens can now be dumped to a printer. PLL Version 1.6 or PLL2 Version 1.3 are IBM™ compatible and sell for $49.95 each plus $3 shipping. Requires DOS 3.3 or greater and EGA 256K.
SW.I.F.T. Enterprises
INFO/CARD #179

Design Program Adds Breakpoints

The updated ICAP/4 design package from Intusoft offers a new version of IsSpice3 based on Berkeley SPICE 3F.2, which includes breakpoints and several other new features. The expanded SPICE library is twice the size of the previous version .
Intusoft
INFO/CARD #178

800-900 MHz Amplifiers

Q-bit's line of connectorized amplifiers for the 800-900 MHz commercial and cellular bands is detailed in their 20 page, "Commercial Radio and Cellular Base Station Amplifier Specification Guide."
Q-bit Corporation
INFO/CARD #198

Mixers

ST Olektron announces the introduction of a new catalog for double-balanced RF mixers covering frequency ranges up to 2 GHz. Fifteen new standard products are introduced, including flatpack and TO-5 devices.
ST Olektron Corp.
INFO/CARD #191

Switch Products

K&L Microwave is pleased to announce the new RF & Microwave Switch Products Short Form Catalog. This 48-page catalog features information on electromechanical switches, solid state switches and switching systems.
K&L Microwave Inc.
INFO/CARD #190

Electromagnetic Waves

The IEEE Press has announced a book series on electromagnetic waves which will include reprints and reissues of classic EM texts as well as new books. The first new titles will be *Dyadic Green's Function in Electromagnetic Theory*, by C.T. Tai, scheduled for late 1993; and *Mathematical Foundations of Electromagnetic Theory*, by D.G. Dudly, scheduled for 1994.
IEEE Press
INFO/CARD #197

84-Page Catalog

The new catalog from JFW Industries features attenuators, RF switches, power dividers, terminations and related components. New products in this edition include power dividers, rotary attenuators, low cost programmable attenuators and an introduction to JFW's matrix switch capability.
JFW Industries, Inc.
INFO/CARD #193

RF Signal Processing Components

M/A-COM announces a 120-page catalog describing their line of E-Series RF signal processing components for commercial applications. The catalog offers detailed specifications on a broad line of products including mixers, I/Q modulators, power splitters/combiners, couplers and transformers.
M/A-COM, Inc.
INFO/CARD #187

Design Parameter Library

California Eastern Labs' design parameter library set, version 6.0 contains ASCII data files (S and noise parameters and non-linear model data) for use with popular CAE programs. The single 5 1/4 inch, 1.2M floppy disk contains NEC GaAs and Silicon products.
California Eastern Labs
INFO/CARD #192

IF/RF Catalog

Daico Industries announces the publication of its 1993-1994 IF/RF Catalog. This 236-page, four-color catalog features Daico's complete family of IF and RF components. Daico specializes in the manufacturing of switches, attenuators, phase shifters, MMICs, bit detectors, couplers, modulators and amplifiers.
Daico Industries, Inc.
INFO/CARD #189

VLSI Products

QUALCOMM, VLSI Products Group has announced the Master Selection Guide. The four-color, 40-page catalog covers the complete line of VLSI products including Viterbi Decoders, Trellis Codecs, Variable-rate Vocoders, DDSs, DACs, PLL Frequency Synthesizers, VCOs, and Synthesizer Boards.
QUALCOMM Inc., VLSI Products Group
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RF design

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