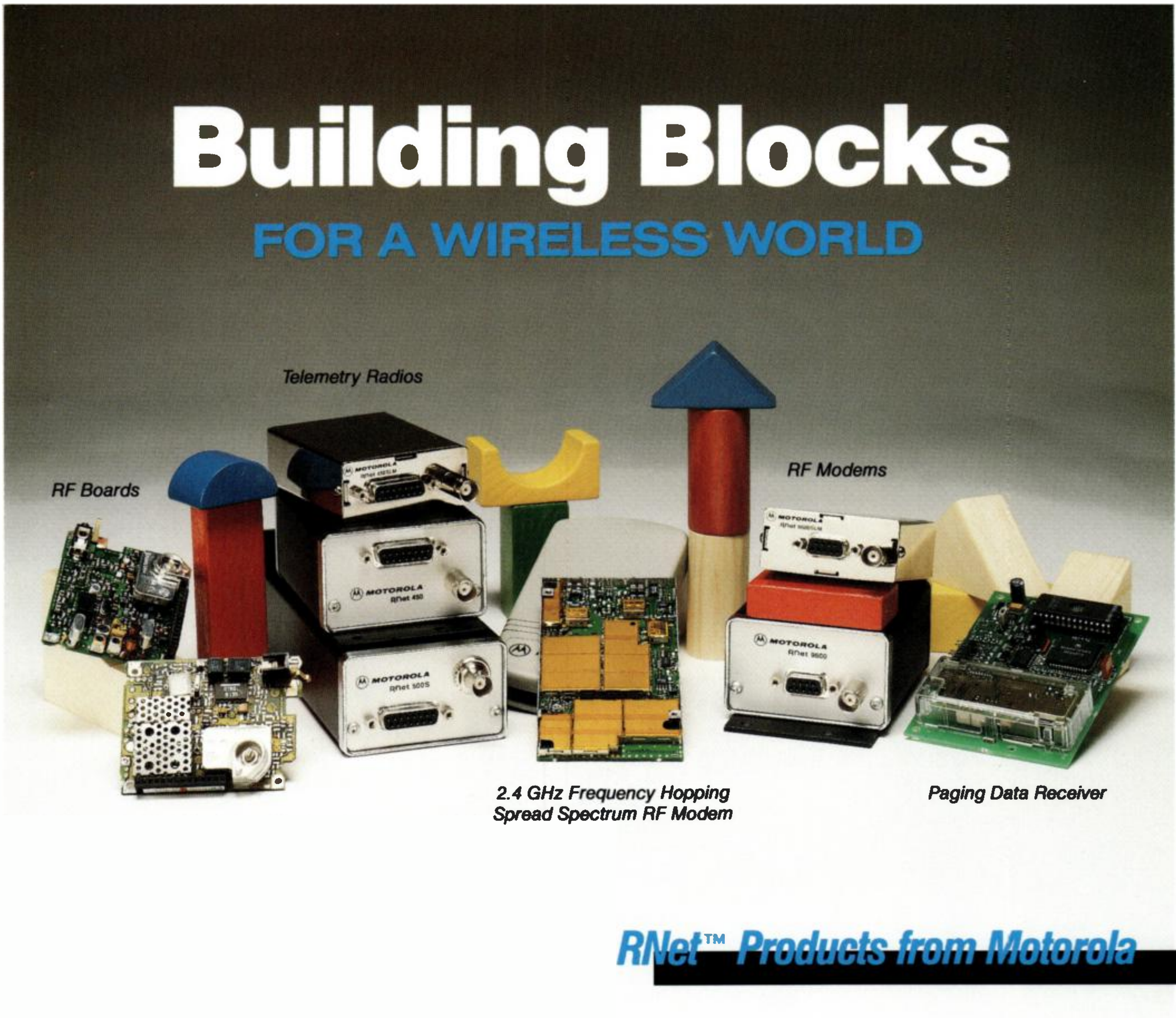
RF design™

engineering principles and practices

January 1994

Cover Story
New Test Set Simulates Multipath Propagation

Featured Technology
RF Applications in Science



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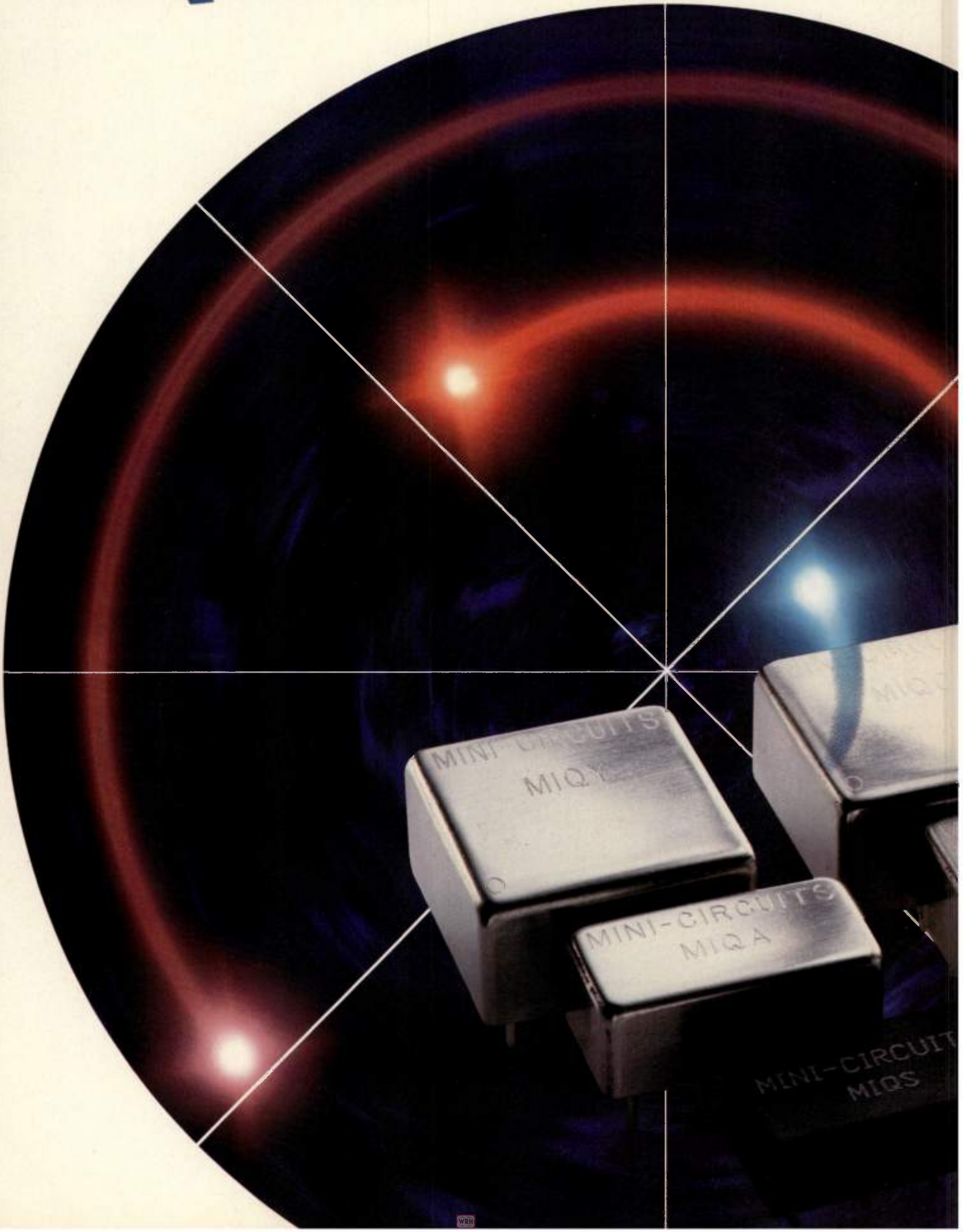
INFO/CARD 1

I&Q MODULATORS



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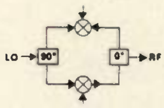
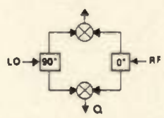


I/Q MODULATORS

MODEL NO	FREQ (MHz) f_L	f_U	CONV LOSS (dB) x	σ	CARRIER REJ (dBc) Typ	SIDEBAND REJ (dBc) Typ	HARM SUPPRESS (dBc) Typ 3xI/Q	5xI/Q	PRICE $ QTY (1-9)
MIQA-10M	9	11	5.8	0.20	41	40	58	68	49.95
MIQA-21M	20	23	6.2	0.14	50	40	48	65	39.95
MIQA-70M	66	73	6.2	0.10	38	38	48	58	39.95
MIQA-70ML	66	73	5.7	0.10	38	38	48	58	49.95
MIQA-91M	86	95	5.5	0.10	38	38	48	58	49.95
MIQA-100M	95	105	5.5	0.10	38	38	48	58	49.95
MIQA-108M	103	113	5.5	0.10	38	38	48	58	49.95
MIQA-195M	185	205	5.6	0.10	38	38	48	58	49.95
MIQC-88M	52	88	5.7	0.10	41	34	52	66	49.95
MIQC-176M	104	176	5.5	0.10	38	36	47	70	54.95
MIQC-895M	868	895	8.0	0.10	40	40	52	58	99.95
MIQC-1785M	1710	1785	9.0	0.30	35	35	40	65	99.95
MIQC-1880M	1805	1880	9.0	0.30	35	35	40	65	99.95
☐ MIQY-70M	67	73	5.8	0.20	40	36	47	60	19.95
☐ MIQY-140M	137	143	5.8	0.20	34	36	45	60	19.95



I/Q DEMODULATORS

MODEL NO	FREQ (MHz) f_L	f_U	CONV LOSS (dB) x	σ	AMP UNBAL (dB) Typ	PHASE UNBAL (Deg.) Typ.	HARM. SUPPRESS (dBc) Typ 3xI/Q	5xI/Q	PRICE $ QTY (1-9)
MIQA-10D	9	11	6.0	0.10	0.15	1.0	50	65	49.95
MIQA-21D	20	23	6.1	0.15	0.15	0.7	64	67	49.95
MIQC-895D	868	895	8.0	0.20	0.15	1.5	40	55	99.95
☐ MIQY-1.25D	1.15	1.35	5.0	0.10	0.15	1.0	59	67	29.95
☐ MIQY-70D	67	73	5.5	0.25	0.10	0.5	52	66	19.95
☐ MIQY-140D	137	143	5.5	0.25	0.10	0.5	47	70	19.95

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INFO/CARD 3

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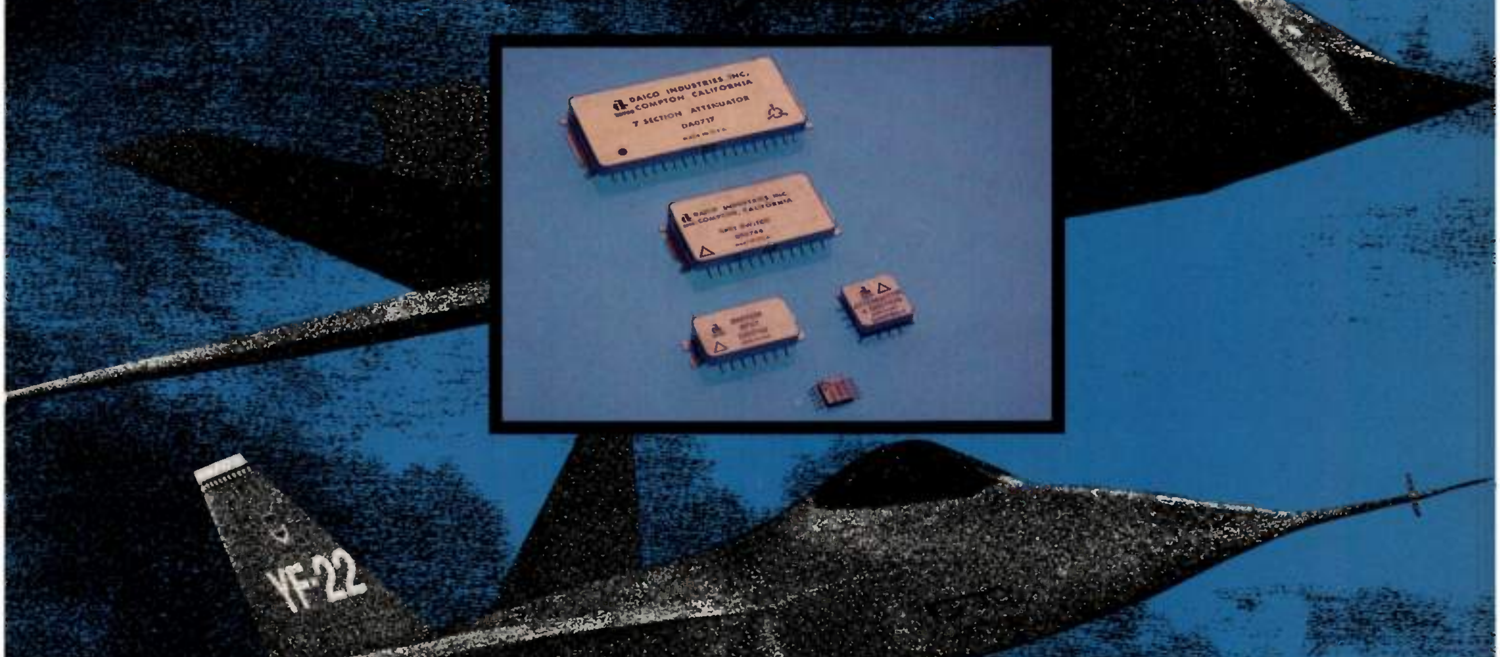
F 156 Rev. A

DAICO'S SURFACE MOUNT OFFENSIVE



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Surface Mount Switches

Config	Freq MHz	IL dB	Iso dB	Switch Speed µSEC Max	Control	Package	Part No.
SPST	5-1000	0.7	45	0.035	TTL	12 Pin SMP	DSO799
SPST	10-2000	1.8	53	0.035	TTL	14 Pin SMP	DSO790
SP2T	DC-2000	0.6	25	0.003	-	SOIC 8	DSO702
SP2T	10-1000	0.7	32	0.050	TTL	12 Pin SMP	DSO712
SP2T	5-1000	1.6	58	0.070	TTL	14 Pin SMP	DSO742
SP2T	50-1100	1.4	48	0.150	TTL	10 Pin SMP	DSW25030
SP4T	10-1000	1.0	57	3.0	TTL	24 Pin SMP	DSO744
SP4T	50-500	1.1	52	1.0	TTL	24 Pin SMP	DSO778
SP5T	10-400	1.0	43	0.100	TTL	24 Pin SMP	DSO705

Surface Mount Attenuators

# of Sections	Freq MHz	LSB Range dB	IL dB	Switch Speed µSEC Max	Control	Package	Part No.
1	20-700	10/10	1.0	0.035	TTL	12 Pin SMP	DAT15015
4	10-1000	1/15	1.9	0.030	TTL	12 Pin SMP	DAO784-1
5	300-1000	1/31	3.4	0.500	TTL	24 Pin SMP	DAO769
5	10-1000	2/62	5.6	0.050	TTL	24 Pin SMP	DAO757
6	10-1000	1/63	6.3	0.050	TTL	24 Pin SMP	DAO786
7	30-500	0.5/63	4.5	20.0	TTL	38 Pin SMP	DAO795
7	30-250	0.5/63.5	6.1	0.035	TTL	38 Pin SMP	DAO717
7	30-150	0.1/12.7	4.0	0.035	TTL	38 Pin SMP	DAO775
VCA	20-300	-/18	0.8	-	Analog	14 Pin SMP	DAO735

Surface Mount Bi-Phase Modulators

Freq MHz dB	LSB Range DEG	IL dB	Switch Speed µSEC Max	VSWR	Package	Part No.
10-500	0-180	1.0	0.070	1.4	12 Pin SMP	DBPO738

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INFO/CARD 4

SWITCHES ATTENUATORS PHASE SHIFTERS MMICS BIT DETECTORS COUPLERS MODULATORS AMPLIFIERS

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January 1994

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This issue's Featured Technology is Scientific Applications of RF. In the photo above, researcher Bob Chufo of the U.S. Bureau of Mines operates a test setup that analyzes reflected RF signals to measure coal seam depth. This method was reported at RF Expo East 1992. Photo courtesy U.S. Bureau of Mines.

departments

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RF is the King of Electronics in 1994

By Gary A. Breed
Editor

This is a monumental year for the RF industry. After all the talk and all the planning, the marketplace will finally get an opportunity to evaluate a whole host of new radio-linked products. Although there has been increasing action in the RF market over the past couple of years, 1994 will show a big jump in product introductions and sales.

A trickle of RF products has begun in the area of personal digital assistants (PDAs). The best analysts refer to these products as communicators rather than computers. The emerging perception of the PDA is as a combination of pager, fax machine and cellular phone.

Add-on RF links for notebook and laptop computers generally fall into the same category, except for wireless LAN interconnections. WLAN products have slowly been working their way into offices, mainly in those special places where the higher cost and slower speed of an RF-linked network connection is offset by the need for portability. As computer users seek greater flexibility, this market will grow.

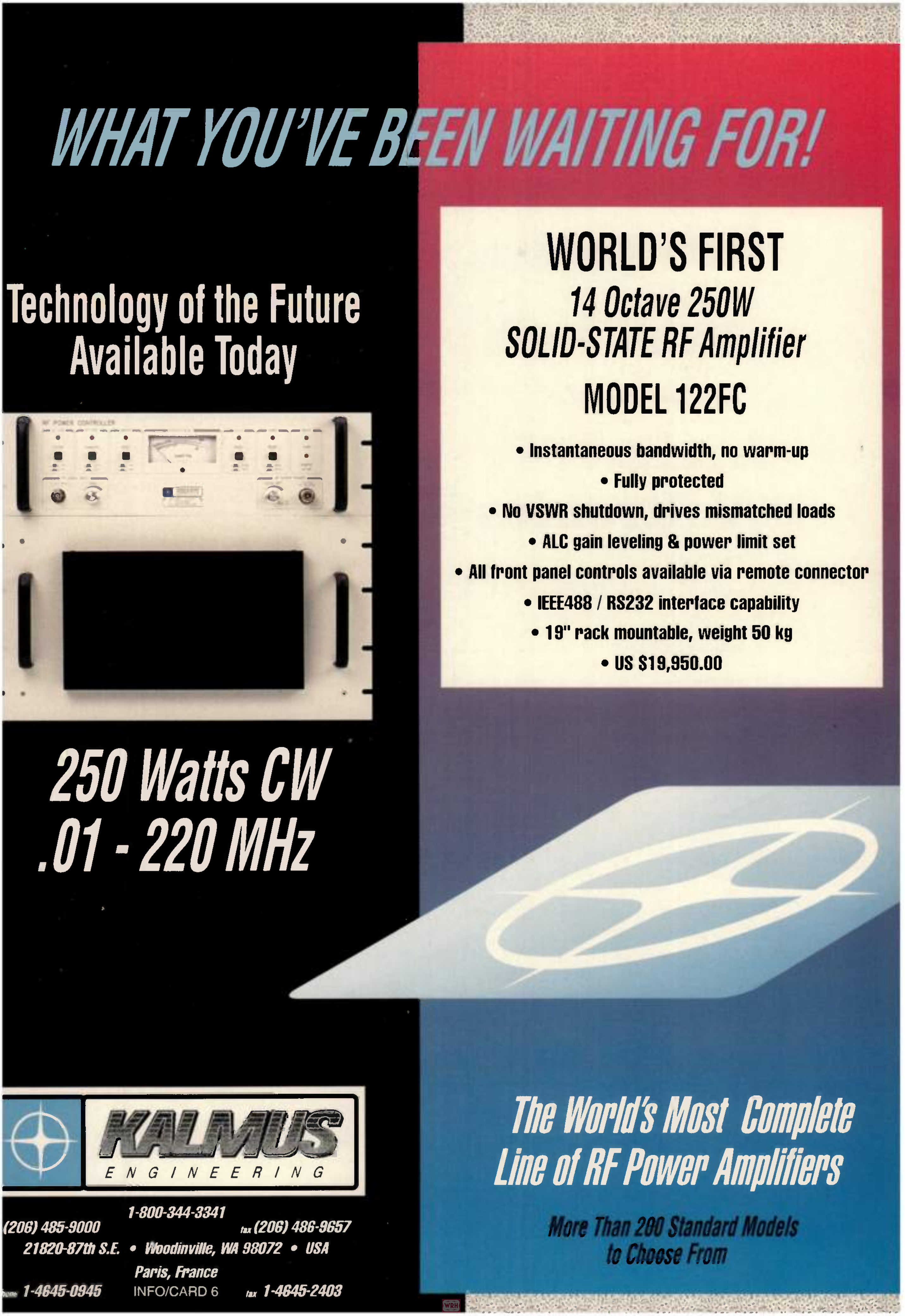
I predict that this year will see the debut of "one person, one number" telephone services. Test operations are in place now, and the early response is positive. Smart networks that know when the user needs to be in the company PBX, a corporate campus microcell, or the general cellular network should be promoted heavily later this year. Personal systems should follow shortly that will switch between a cordless phone at home and a cellular phone elsewhere, reached by calling just one number.

2450 MHz technology is developing rapidly, which should remove some of the concern over the overcrowded ISM/Part 15 band at 900 MHz. Those $750 spread spectrum cordless phones will come down in price, and plenty of new wireless gadgets will appear this year. I expect to see everything from cordless CD-quality headphones to household audio/video systems that function as VCR broadcasters, video intercoms and alarm systems.

Worldwide, GSM digital cellular should grow dramatically. Europe will lead the way, followed shortly by many other countries which have adopted that standard. The U.S. cellular market will grow, but expansion with new technology will likely be NAMPS (a narrower bandwidth analog system), with digital cellular a few years away from large-scale installation.

The personal communication system (PCS) market will not have an impact on product sales in 1994; the FCC rules and spectrum distribution will come too late for products to reach the market. However, test marketing of new systems will whet the consumer's appetite with prototype and pre-production models of products for 1995.

We who are observers of the RF industry have been predicting the coming boom. Now, financial analysts, business publications, plus the general print and television media have recognized the importance of these developing markets. Their reports will get the message to the general public, who will finally understand that these new capabilities for communications, convenience and information are not futuristic, but ready now for their evaluation.



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

For the very first time, you can make CW *and* peak power

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

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.

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.

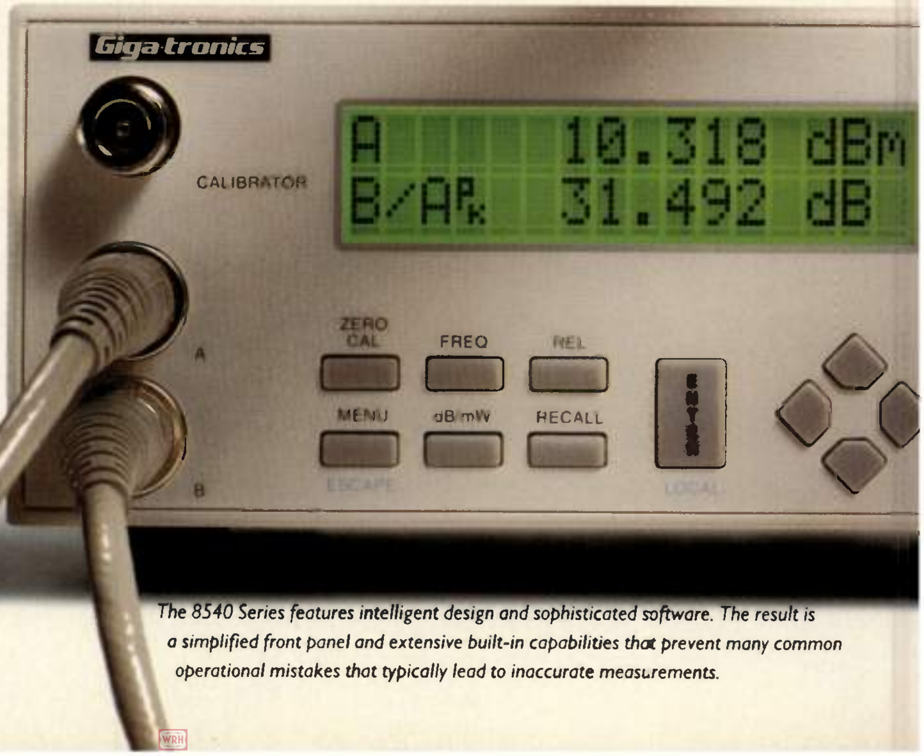
Think about what all this will do for your ATE productivity as well as for your company's bottom line.

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.

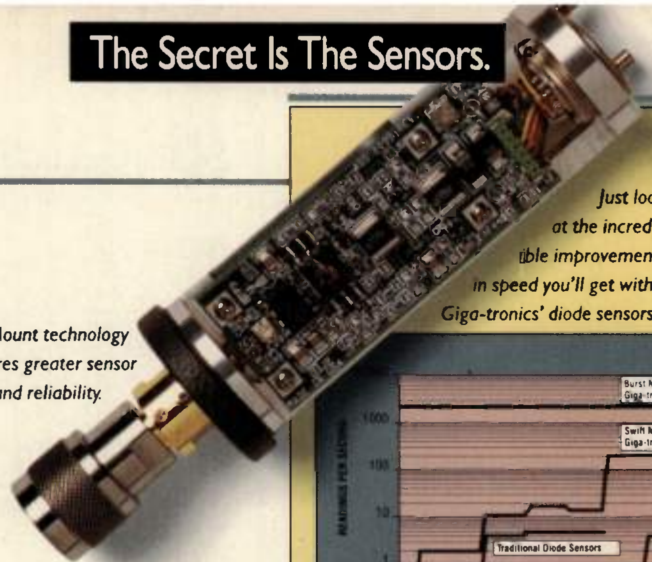
There are no time-consuming, unreliable duty cycle corrections, and you'll get the same accuracy and speed you'd get with a much-more-expensive dedicated peak power meter.

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.



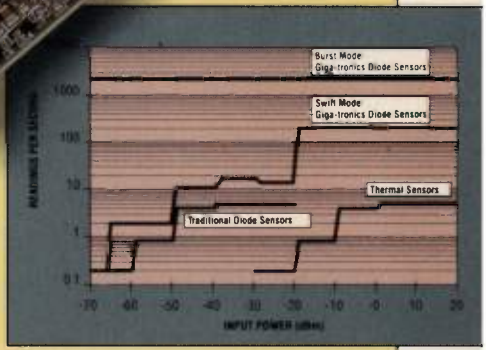
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

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.



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

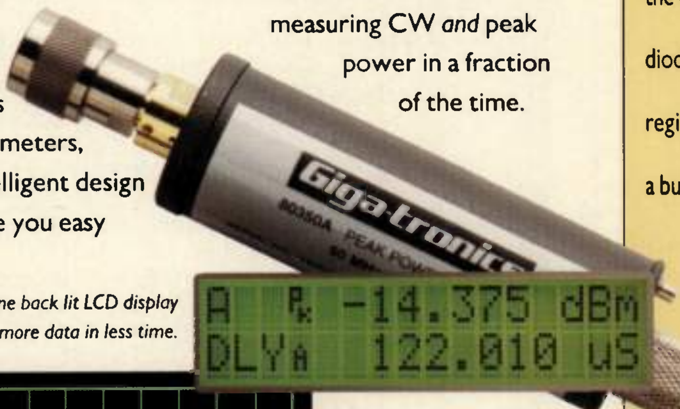
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

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.

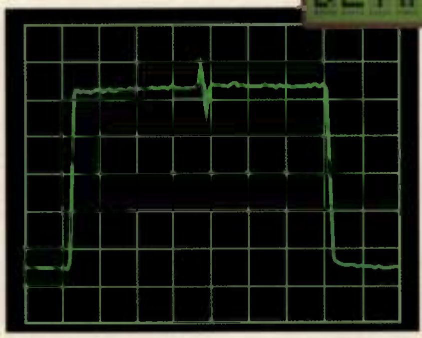
Imagine all this power and performance. But why just imagine? Get the truly incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.

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.



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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INFO/CARD 10

RF calendar

January

12-13 **San Diego Electronics Show**
San Diego, CA
Information: Epic Enterprises, Show Management, 8989 Rio San Diego Dr., Suite 160, San Diego, CA 92108-1647. Tel: (619) 294-2999. Fax: (619) 294-6699.

27-28 **Measurement Science Conference**
Pasadena, CA
Information: Measurement Science Conference, John Schulz, Registrar, 1280 Bison Ave, Suite B9-530, Newport Beach, CA 92660. Tel: (714) 863-9031. Fax: (714) 863-1723.

February

8-11 **EXPO Comm Mexico 1994**
Mexico City, Mexico
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

March

8-11 **Communications 1994 - Australian International Communications and Office Technology Exhibition**
Melbourne, Australia
Information: TWI, International Exhibition Logistics, 3190 Clearview Way, San Mateo, CA 94402. Tel: (415) 573-6900. Fax: (415) 573-1727.

14-17 **The Second International Symposium on Digital Audio Broadcasting**
Toronto, Canada
Information: DAB Symposium '94, 126 York Street, #401, Ottawa, Ontario, Canada, K1N 5T5. Tel: (613) 594–8226. Fax: (613) 565–2173.

20-24 **National Association of Broadcasters '94**
Las Vegas, NV
Information: NAB '94 Convention, 1771 N Street, NW, Washington, DC 20036–2891. Tel: (800) 342–2460, (202) 775–4972. Fax: (202) 775–2146.

21-26 **Applied Computational Electromagnetics Society 1994 Conference**
Monterey, CA
Information: Jodi Nix, Symposium Facilitator, Veda Incorporated 5200 Springfield Pike, Suite 200, Dayton, OH 45431. Tel: (513) 476-3550. Fax: (513) 476-3577.

22-24 **RF Expo West**
San Jose, CA
Information: RF Expo West, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828–0420. Fax: (404) 618–0441.

April

12-15 **EMC/ESD International**
Anaheim, CA
Information: EMC/ESD International, Registration Coordinator, 6151 Powers Ferry Rd. NW, Atlanta, GA 30339. Tel: (800) 828–0420. Fax: (404) 618–0441.

RF letters

Letters should be addressed to: Editor, RF Design, *6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Letters may be edited for length or clarity.*

ISO 9000 Responses

Editor:

I read the letter "ISO 9000: Barking up the Wrong Tree?" [*RF Design*, October 1993, p. 14] and felt compelled to clarify some of the misconceptions Mr. Lohrer expressed.

I am an ISO 9000 registered lead assessor recognized under both the United States (Registrar Accreditation Board) and United Kingdom (International Quality Assurance) systems. I assure your readers that "36-hour whiz kids" are not being used by registrars in systems audits as Mr. Lohrer suggests. Participating in a Lead Assessors course and passing a written examination could qualify one as a provisional assessor; however, to conduct a registration audit, one must have experience, education and have been supervised in at least 5 audits of greater than 2-day duration. Registrars are required to abide by these qualifications for their auditors as well as maintain a management system in compliance to EN 45012.

Dennis Omanoff
SynOptics Communications, Inc.

Editor:

I was disappointed to read the tirade against ISO 9000 by George Lohrer in your October issue. It does not have to be costly, ineffective or regimented.

Fortuately, in our company we are not burdened by a separate QC/QA heirarchy — quality is everybody's responsibility, including mine as Engineering Manager. Last year, I was given responsibility for developing a Quality Management System to meet the requirements of ISO 9001, to cover R&D through installation and service. It was achieved in July this year, ahead of schedule and with minimal hassle.

If Mr. Lohrer truly believes that he runs a good company which provides high quality products, then he must have good quality procedures already in place. This was the view we took with our analytical instrument business — we only had to write down what we did and have the documentation and processes examined by a "third party" in order to complete the assessment (first time!).

Do not be fooled by so-called consultants who will offer the write this stuff for you; they get paid by the word! It's your business, run it your way, write it down simply — and get a certificate (to go with the UL, CSA, etc. ones) to prove that what you say is true. This has to be the big advantage of ISO 9000; your business processes have been approved by an independent organization and not just claimed by your marketing operation. This process of self-examination is invaluable, and you will certainly come across improvements which need to be made — and for these, you can take Mr. Lohrer's "time-honored problem-targeted pragmatic approach."

Paul Knight
Finnigan MAT Ltd.

Editor:

I found George Lohrer's letter to the editor to contain arrogant, biased and just plain inaccurate information. As an ASQC Certified Quality Auditor with ten years experience in the "QC/QA establishment" he refers to, I decided it was only fair to answer Mr. Lohrer.

the ISO 9000 series of quality standards are nothing but a set of generic quality standards, not much different from the ones we are familiar with — military or otherwise. Unlike previous schemes, claiming compliance to an ISO standard is usually interpreted as being also "registered," which means that your company has satisfactorily demonstrated to a Registrar its compliance one of the three ISO standards. Registration is the only cost aspect which is different between complying to an ISO standard or any other generic quality standard.

The current feverish activity to become registered stems from customer requirements more than anything else. I imagine that educated customers prefer to do business with a company that has a stable quality system (usually one of the benefits of being registered), which in turn can produce a product with a certain expected quality level.

I would seriously suggest to the readers that they ignore Mr. Lohrer's letter and do one very important thing: Listen to your customers' requirements. If the requirements are for you to comply with any of the standards in order to do business, then you do not have much of a choice, do you? It is a case where a simple rule applies: The customer is always right.

Mario Vaz de Medeiros
Quad Tech, Inc.

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

RF and Microwave Circuit Design II: Linear/Nonlinear Techniques and Applications
January 31-February 4, 1994, Los Angeles, CA
Information: UCLA Extension, Engineering Short Courses, 10995 LeConte Ave., Ste. 542, Los Angeles, CA 90024. Tel: (310) 825–1047. Fax: (310) 206–2815.

Coherent Radar Performance Estimation
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Principles of Pulse Doppler Radar: High, Medium, and Low PRF
February 15-17, 1774, Atlanta, GA
Phased-Array Radar System Design
April 19-22, 1994, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894–2547.

Radio Frequency Engineering
January 17-21, 1994, Colchester, United Kingdom
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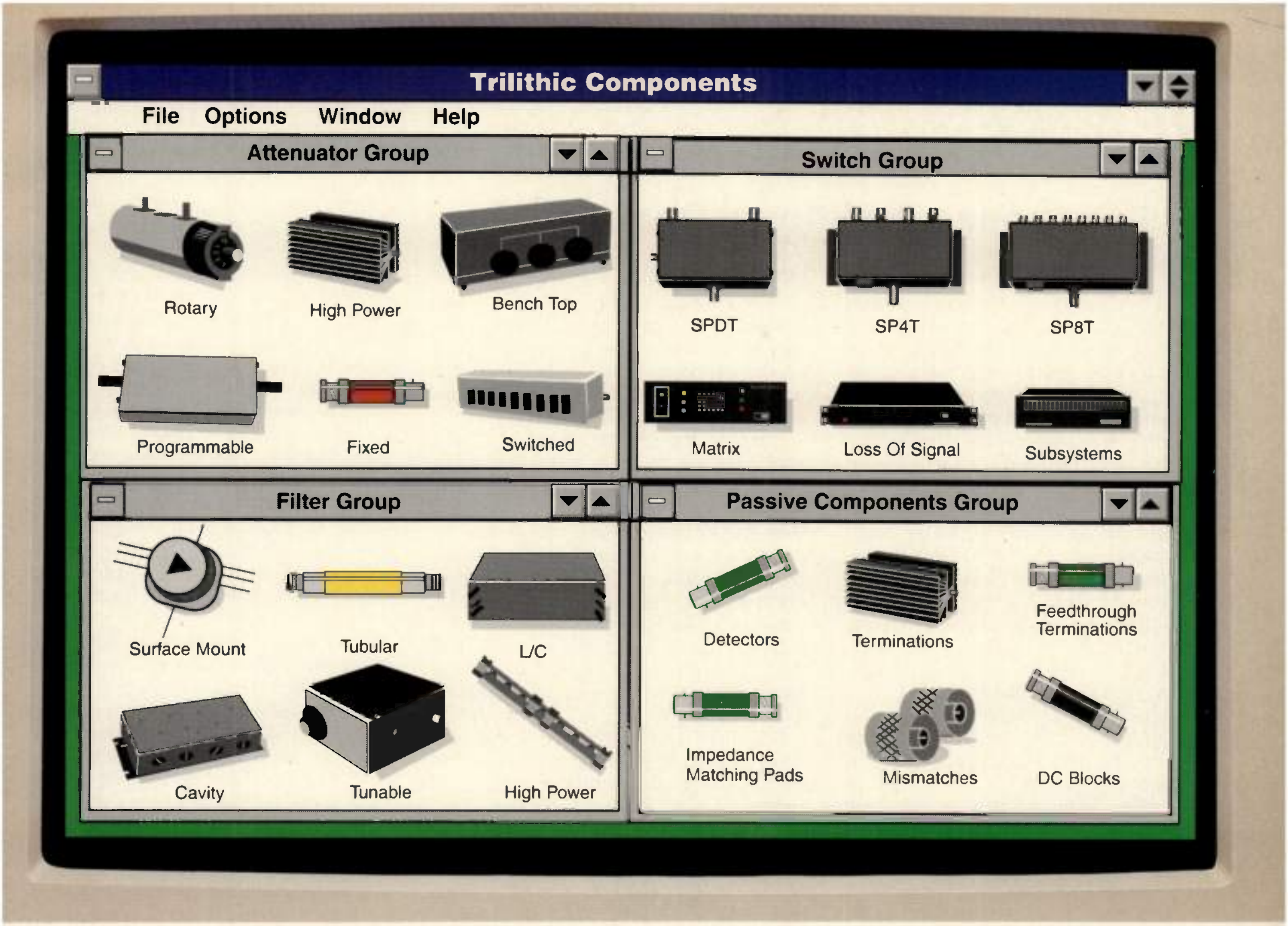

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INFO/CARD 15

FCC Proposes Additional Channels for Cordless Telephones

The Federal Communications Commission (FCC) has released RM-8094, a Notice of Proposed Rule Making which would add 15 new channel pairs to the existing ten channels allocated for cordless telephone applications. The new channels will use frequencies in the ranges of: 43.71-44.49 MHz (base transmitter) and 48.75-49.51 MHz (handset transmitter). Existing channels operate on 46.60-46.98 MHz (base transmitter) and 49.66-50.0 MHz (handset transmitter). The current channels use specific base/handset frequency pairs, but equipment using the new channels may pair any base frequency with any handset frequency, an option intended to reduce interference.

Although new services offered in the 915 and 2450 MHz include cordless telephones, the FCC feels that the "enormous popularity" of lower cost of units offered in the 46/49 MHz range will continue for the forseeable future. Also, increased use of baby monitors which operate on five of the existing channels has made cordless telephone operation on these channels impossible in some areas.

Because the are approximately 2800 stations operating in this frequency range as part of the Forest Products Radio Service, Motor Carrier Radio Service and Petroleum Radio Service, anti-interference measures are required. the proposed rules include a provision that cordless telephones operating on the newly proposed channels "must incorporate an automatic channel selection mechanism that will prevent establishment of a link on an occupied frequency."

Comment and reply comment deadlines passed in November, and action on this proposal is anticipated by mid-year.

AT&T Demonstrates 0.1 Micron Silicon Devices — Silicon ICs with 0.1 micron transistor geometries have been developed by AT&T's Silicon Electronics Research Laboratory. The new process solves the problem of temperature rise in small-geometry devices that occurs because the lower doping level required at these dimensions increases the current leakage. For this reason, previous devices using 0.1 micron channels required cooling. Using this process, dividers operating to 10 GHz have been demonstrated, along with individual transistors operating to 116 GHz for NMOS and 51 GHz for PMOS. Devices using the new AT&T process can run from 1.5 volt power supplies.

November ITU Conferences Plan for the Future — In November, the World Radiocommunication Conference (WRC 93) closed with a draft agenda for the next two conferences to be held in 1995 and 1997. WRC 93 is seen as a strategic planning exercise for future radio regulations, and a means of testing the new bi-annual system of World Radiocommunication Conferences. The draft agenda for WRC 95 includes: Review of committee work started in 1990 that will simplify the Radio Regulations; review of technical constraints associated with Mobile Satellite Services (MSS) under 3 GHz, intended to help facilitate their use; review additional matters regarding MSS and other satellite services, including feeder links, earth exploration satellites, broadcasting satellite services and other space services; and address the use of new HF bands recently allocated to broadcasting.

WRC 97 has the following items on the draft agenda: Coordination of space and terrestrial services between 1 and 40 GHz; allocation of unplanned space services; wind profiler radars; further work on HF bands allocated to broadcasting; and discussion of the Global Maritime Distress and Safety System.

The first ITU Radiocommunication Assembly, successor to the Plenary Assembly of the International Consultative Radio Committee (CCIR), approved a recommendation to assure compatibility between VHF broadcasting and aeronautical navigation, and continued progress on discussion of future mobile communications. Under the new structure, this body can be characterized by its unofficial mission statement, to aim at "an efficient contribution to technical standards and the most effective utilization of the radio spectrum in the context of the world's economies and needs."

Analog Devices and IBM Team Up on SiGe Technology — Analog Devices and IBM have announced an agreement regarding RF and mixed-signal integrated circuits based on IBM's ultra-high vacuum chemical-vapor deposition (UHV/CVD) silicon-germanium (SiGe) process. The agreement calls for the companies to design, produce and market ICs intended for high-volume radio and high-speed applications. The SiGe process has been shown to produce transistors with F_T to 60 GHz, roughly a factor of three higher than previous silicon IC processes. At the recent International Electron Devices Meeting (IEDM), ADI and IBM reported on a 1 GHz 12-bit digital-to-analog converter fabricated using this process. The 3000-transistor size of the DAC indicates that the process will be suitable for single-chip implementation of many RF functions.

NIST Describes EM Materials Measurement Services — A new publication, NIST Measurement Service of Electromagnetic Characterization of Materials (NISTIR 5006), presents an overview of special test and measurement services for characterizing dielectric and magnetic properties of materials at radio and microwave frequencies, from 100 kHz to 26.5 GHz. Measurements can be made of permittivity and permeability, as well as loss tangents of low-loss materials. Copies of the publication can be obtained from the National Technical Information Service, Springfield, VA 22161; (703) 487-4650. Order by: PB 94-110186.

Telephone-Over-Cable TV System Announced — Scientific-Atlanta has announced a system of telephone service delivery over broadband cable TV networks, enabling switched telephone, video, data and interactive TV services to be integrated into a single communications network. Interfacing equipment is located in a box placed at the customer end of the cable drop, where telephone and cable TV wiring then enters the build in the same manner as the existing service. Delivery of the CoAccess™ system hardware in commercial quantities is expected in the first half of 1994.

Conductive Polymers are Water-Soluble — IBM research scientists have invented a new class of polymeric materials which are both electrically conductive and water soluble. These derivatives of polyaniline, called PanAquas™, the polymers can be made insoluble by exposure to radiation such as electron beams, x rays and ultraviolet light. Advantages of these materials include

processing without organic solvents, curability and removability. They may be most valuable as ESD charge dissipators in both removable and permanent applications, either as a coating or blended into plastics and fabricated directly into structural components.

IPC Develops School for PWB Technology — To be held during the Printed Circuits Expo, April 1994 in Boston, the School for PWB Technology is presented by the Institute for Interconnecting and Packaging Electronic Circuits (IPC). Seventeen tutorials or workshops are currently planned, which are intended to cover topics ranging from laminating and flexible circuits to costing and ISO 9000 implementation. For more information call IPC at (708) 677-2850.

Hughes Promotes new CBiCMOS Process — Offering custom and semi custom integrated circuits, Hughes Microelectronics announces its CBiCMOS process. The process supports 2 micron CMOS gates, 2 GHz PNP transistors and 5 GHz NPN transistors. Operating voltages as low as 1.5 volts can be obtained. Full custom and standard analog building block semi-custom ICs are offered, based on this process.

D.L.S. Electronic Systems Receives TUV Appointment — The first Certificate of Appointment given by TUV Rheinland for EC EMC testing has been received by D.L.S. Electronic Systems, Inc., and EMC testing laboratory and consulting firm in Glenview, Illinois. The appointment means that D.L.S. may issue a Certificate of Conformance which their customers can then use to prove compliance and put the CE mark on equipment to be sold into the European Economic Community.

Signal Technology Acquires Systron Donner — Signal Technology Corp. will acquire the Microwave/Instrument Division of Systron Donner Corp., a Thorne EMI company. The division manufactures microwave and instrumentation products for defense and commercial markets, with sales of approximately $12.5 million annually. The instrument division will operate under the name ST Systron Donner.

New Thin Film Company Formed — Spectrum Thin Films is a newly formed company which manufactures standard and custom thin film devices for commercial and military applications. The company, based in Hudson, NH, will make thin film resistor, conductor and multilayer products for hybrid microelectronics, using an environmentally sound laser process.

Proxim Announces Public Stock Offering — Proxim, Inc.has announced that has filed a registration statement with the Securities and Exchange Commission relating to the proposed initial public offering of 2,000,000 shares of common stock. 1,500,000 shares will be sold by the company, and 500,000 shares sold by certain stockholders. Initial pricing is expected to be between $9 and $11 per share. Proxim designs, manufactures and markets spread-spectrum based wireless networking products.

SpecTran to Acquire Ensign-Bickford Optics — SpecTran Corp. has announced that it will acquire substantially all of the assets of Ensign-Bickford Optics Co. and Ensign-Bickford Optical Technologies, Inc. The acquisition will result in SpecTran's ability to offer a wide range of specialty fiber and value-added fiber optic products.

ESSCO Receives FAA Contract — Electronic Space Systems Corp. (ESSCO) has been awarded the Federal Aviation Administrations (FAA) Radome Replacement contract to manufacture and install sandwich radomes for primary and secondary surveillance systems at FAA facilities nationwide. The new sandwich radomes will accommodate the more sophisticated and larger beacon phased array en route surveillance radar systems.

Flight Test Contract Awarded to Aydin Vector — Lockheed Aeronautical Systems Co. has selected Aydin Vector Division of the Aydin Corporation to provide the Flight Test Data Acquisition System for the F-22 Advanced Tactical Fighter. Nine vehicles will be built and rigorously tested through the year 2000. The contract, valued at $20 million, includes micro-miniature MMSC-800 Narrow Band and Wide Band Data Acquisition units, high speed digital tape recorders, data handling equipment and ground support equipment.

Varian to Acquire Quality Hermetics — Varian Associates announces that its Canadian subsidiary, Varian Canada, Inc., has acquired the assets of Quality Hermetics Co. of Toronto. Quality Hermetics is a privately held company which designs and manufactures glass-to-metal hermetic seals for a variety of applications in solid state amplifiers, avionics equipment, high vacuum systems and semiconductor manufacturing. Varian Canada manufactures microwave tubes and electronic equipment for medical applications.

Analog Devices Announces RF Product Group — A new product line has been formed at Analog Devices, Inc. to continue development of products targeted at new wireless applications. Components for these applications have previously been part of the company's Analog Signal Processing product line. The new group will combine RF functions with DSP, voice-band and baseband functions to address all portions of RF-based communications products.

CMOS Technology License Agreement — TEMIC TELEFUNKEN microelectronic GmbH and Mitsubishi Electric Corp. have reached a licensing agreement on the transfer of submicron CMOS technology as part of long-term cooperation agreement between the companies. Initially, a 0.5 micron process will be used at TEMIC's French affiliate, MATRA-MHS, which makes ASICs, microcontrollers and microprocessors. The TEMIC microsystems group makes HF components, control and sensor systems, MCM modules and power hybrids.

Superconductor Technologies Adds Manufacturing Capacity — A new deposition system purchased by Superconductor Technologies will double the company's high temperature superconductors capabilities. The new chemical vapor deposition system will be used in conjuction with an existing laser ablation system to produce superconducting thin films used in computer, telecommunications and medical MRI applications.

Vendors Adopt HP Instrument Control Library — Six I/O product suppliers have adopted the standard instrument control interface library which Hewlett-Packard Company recently released as an open standard. The companies are IOTech Inc., Data Translation Inc., Capital Equipment Corp., Keithly Metrabyte and RadiSys Corp. HP also announced that Microsoft Corp. has made a commitment to bring the library to the control, engineering and manufacturing community.

Why DSP Could Be the Key to Your Next RF Design

By Andy Kellett
Technical Editor

Digital Signal Processing (DSP) has become faster and less expensive in the last decade, following a familiar pattern: new technology enables new applications, drives new technology, enables new applications... New DSP chips are faster, more self-contained, and increasingly, they are tailored for specific applications — applications such as digital cellular telephony, wireless LANs and digital video transmission.

DSP is crucial to many potentially large RF markets. "This whole infrastructure of satellite based personal communications, cellular-based communications and PCS require some DSP," says Doug Mitchell, Vice President of Marketing and Sales for array Microsystems, a Colorado Springs, CO manufacturer of DSP chips and boards for vector processing of signals. "I don't know that anybody disagrees when you say all those markets could be very large."

DSP processing power is up to the challenge offered by the emerging wireless markets. High-end DSP chips (for instance, the AT&T DSP1617) operate around 50 MIPS (Million Instructions Per Second), and chips are available with 16- to 32-bit, and fixed and floating point resolutions. What specification is the figure of merit depends on your applications says Thomas Hack, a Senior Design Engineer at Siliconix, and a frequent contributor to RF Design of articles on digital RF technology .

When a single processor can't do the job alone, parallel processing can be employed. Multiprocessor DSP is currently used in the most demanding applications for DSP. Radar and video signal processing often require the added power of parallel processing, as do digital cellular base stations, which have to handle several channels of information at once. Several chips are designed to make connecting several in parallel easier. "The TMS320C40 has a glueless interface," says TI's DSP Marketing Manager, Kun Lin , "We have customers who have a thousand TMS320C40s in one system." However, there are diminishing returns as more processors are connected. "At some point it's more work computing how to split the work, than it is to do the work itself," notes array Microsystems' Mitchell. Still, that point of diminishing returns is well above the performance level of a single DSP chip.

"DSP architectures are heading towards higher integration, with larger on-chip memory and on-chip peripherals," says Bill Windsor, Product Manager for DSP Processors at Analog Devices. Analog's newest 32-bit floating-point DSP, the ADSP-21060, reflects this trend, containing 4M of RAM on-chip and extensive on-chip memory management. Increased integration reduces size and cost; both are important for . The increased inclusion of DSP in handheld devices also drives increased demand for power-conserving DSPs. Many DSPs have wait and stop modes, and many of the next round of DSP device introductions will include 3 V devices.

DSPs are also becoming more application specific. Examples of this type of DSP are devices designed to code and/or decode video signals using JPEG, MPEG and H.261. Such encoding decoding schemes could be implemented with sets of general purpose DSPs, says array Microsystem's Mitchell, but, "no one could afford to buy it." Instead a number of companies, including array Microsystems, are readying monolithic DSP chips dedicated to performing these video encode/decode tasks. Another type of application specific DSP chips are those which perform traditional analog tasks, such as Harris' HSP50016. This chip performs digital down conversion and filtering, simply taking in a sampled signal and spitting out I and Q data streams.

Nobu Okuyama, Senior Product Marketing Engineer for DSP products at NEC, points to the various competing digital cellular and wireless datacom systems when he says, "... wireless communication is not yet firmed up. So, in a sense, it is not the time to talk about customized DSP solutions."

Designing with DSPs has also become easier. Algorithms no longer have to be hand-entered as machine code. "There are companies now that provide high level design tools which allow a designer to graphically pick and choose blocks to represent the functions in the application, and then the system will compile all the way down to DSP code," says TI's Lin. Companies such as Comdisco and Hyperception, among others, provide such software.

C code is another popular way to implement DSP algorithms, but even here, the designer doesn't have to worry about becoming an ace programmer. Robert DeRobertis, Product Manager for Wireless Signal Processors at AT&T, offers an example, "We offer VSELP code, so all the person has to do is write an executive and call those libraries. We've had customers integrate this into their phone in two or three months, as opposed to the many months it would take to do a complete software development."

Complete chip sets incorporating both RF analog and DSP functions are another way manufacturers are helping reduce design times. "Where RF and DSP never touched before, we are now seeing customers who require products that are gluelessly interfaced together," says AT&T's DeRobertis. The complementary nature of DSP and analog RF circuits is also seen by Charles Fadel, Corporate Marketing Manager for Analog Devices' new Communications Sector. "The advantage of providing a complete system is the ability to interplay the advantages of each section," says Fadel, "For instance, if you have a DSP, you can offload the non-linearity of an amplifier with it."

DSP has become less expensive in the last decade. For instance, TI's TMS320C10 was originally priced around $500 when it was introduced in 1982. Now, unit quantities sell for about $5. If this trend continues, RF designers should have some very useful and affordable DSP devices in the next decade. Says TI's Lin, "What looks expensive today will be very inexpensive tomorrow." *RF*

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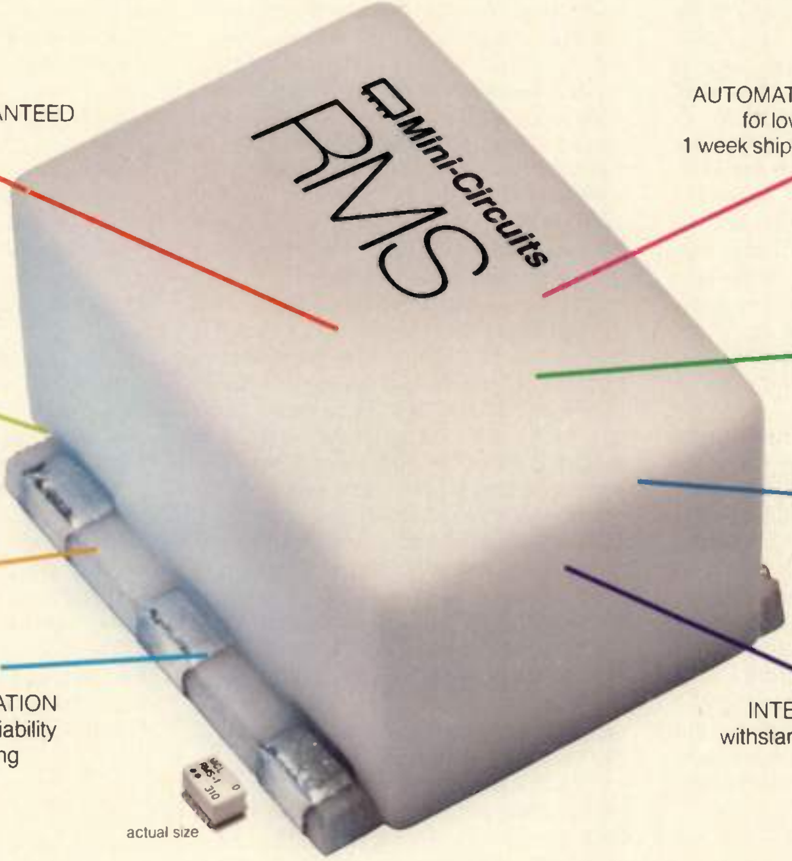

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INFO/CARD 18

The Flux-Gate Magnetometer: A Very Sensitive ELF Magnetic Detector

By Peter Vizmuller

Interference between a switching power supply and a receiver's VCO prompted the magnetometer described in this story; a magnetic flux modulated at audio frequencies was permeating the VCO's compartment through three layers of shielding and extensive bypassing. A way to analyze low frequency magnetic fields was required. This article describes the magnetometer's theory of operation in sufficient detail to allow anyone to build an inexpensive magnetic sensor of nearly 10 V/gauss sensitivity. (In case you are still wondering about the problem that triggered this investigation, the interference between the VCO and power supply was solved by changing the switching transformer construction, preventing excess flux from leaving its ferrite core.)

The flux-gate magnetometer has found use in an amazing range of applications. Interplanetary magnetic field mapping, submarine detection, non-contacting current measurements, ELF detection and mineral exploration are some of the uses for this versatile device [1]. Until the discovery of the superconducting SQUID and other quantum mechanical detectors, flux-gate magnetometers were among the most sensitive, and certainly remain the most cost-effective devices for measuring DC and low frequency magnetic-fields.

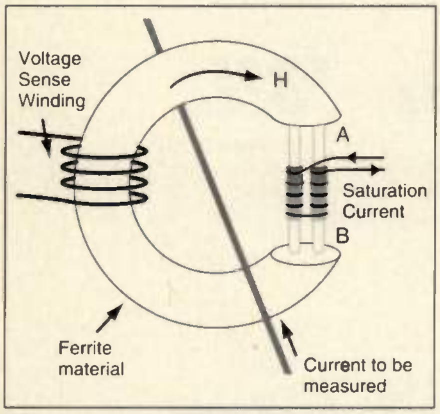

Figure 1. Simplified magnetic flux gate for measuring DC current.

The HP428A clip-on milliammeter was probably the first commercial instrument which utilized a flux-gate device to sense the magnetic field of a current-carrying wire passing through the probe head, thus allowing the measurement of DC currents below 1 mA without breaking the circuit. A brief note in the instrument's manual, warning about errors introduced by the Earth's magnetic field, hinted at the remarkable sensitivity of this little device [2]. The current-carrying wire was enclosed by a ferrite toroid of special shape, portions of which were periodically saturated by a sinusoidal gate signal, as shown in Figure 1.

The two saturation windings are wound in such a way as to confine the saturation flux in the two thinner ferrite sections between A and B. Both pieces of ferrite between points A and B saturate twice per each cycle of the saturation current, (because there are two current peaks in each sinusoidal cycle, one positive, the other negative). This action gates the magnetic flux which is produced in the toroid by the current to be measured.

Because the magnetic path is interrupted, or gated, at twice the rate of the saturation current, the flux density through the voltage sense winding will also change at the same rate, inducing a voltage in the sense winding at the second harmonic of the saturation gate signal.

An easy way to visualize this flux gating process is to realize that the ferrite between points A and B is effectively removed at twice the rate of the gate signal so that the overall effective permeability of the whole toroid changes between two states: high permeability when unsaturated, and low permeability when partially saturated. The low permeability is due to the presence of an 'air gap' between A and B when that portion is saturated.

Yet another way to explain flux-gate operation is to quantify the various relationships mathematically. Looking at the flux through the sense winding:

$$V = -N_s \frac{d\Phi}{dt} \qquad (1)$$

where,
V = voltage induced in sense winding
N_s = number of turns in sense winding
Φ = flux through cross-section of sense winding

But we know that $\Phi = Ba$ and $B = \mu H$, therefore:

$$V = -N_s a H \frac{d\mu}{dt} \qquad (2)$$

where,
a = ferrite cross-sectional area inside sense winding
H = magnetic field, proportional to the measured current
μ = effective permeability of whole toroid

Thus the induced voltage is directly proportional to current to be measured, and also depends on how fast the effective permeability is changing. Because the permeability changes at twice the rate of the saturation signal, the sense voltage has a strong component at the second harmonic of the saturation current as well. We can immediately see that the sensitivity can be increased by using high permeability material, lots of turns in the sense winding and a high frequency for the saturation gating signal. There are no inherent frequency limitations on H in the above equations,

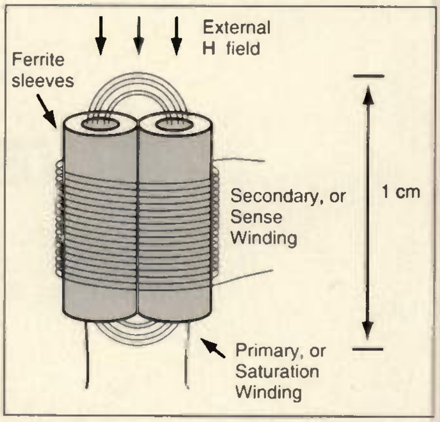

Figure 2. A practical flux-gate magnetometer using two 2673002201 ferrite sleeves and two orthogonal windings.

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In many cases, the magnetic field to be measured is not produced by a current carrying wire. Leakage from a cathode-ray tube, the Earth's ambient field, magnetization of a mineral rock sample are all examples where it is impossible to interrupt a closed magnetic path, as in the above example. In order to measure open field lines, the basic design can be modified as shown in Figure 2, but the principle of operation remains exactly the same:

The primary, or saturation winding has the same function as before, except that in this case, *all* of the ferrite material is saturated twice per cycle, rather than only a portion of the material. Voltage is induced in the secondary because ferrite effectively 'appears' and 'disappears' inside the secondary winding. The secondary, or sense winding will contain energy at the second harmonic of the primary frequency, only if there is an external magnetic field present. In the absence of an external magnetic field the secondary voltage will consist only of a small leakage signal from the primary. Because the two windings are wound at right angles, there is no transformer action and the primary signal does not get to the secondary, except for a small imbalance signal plus a capacitively coupled signal. In any case, the input (primary) and output (secondary) are at different frequencies, so they are easy to filter and separate.

Note that the magnetometer is now sensitive to direction and polarity; maximum output is obtained when an external magnetic field is along the long axis of the ferrite sleeves, and produces no output if the external field is perpendicular to it. Again, there is no lower frequency limit and the detector will respond to a DC field as well as to a changing field.

Resonating both the primary and secondary winding at their respective frequencies greatly improves the basic detector. Resonating the primary increases the current available for saturation and suppresses second harmonic energy. Resonating the secondary winding at the second harmonic filters out first harmonic energy and increases the available voltage amplitude, if the secondary is terminated in high enough impedance. A parallel capacitor will also absorb any of the inevitable stray capacitance at the input and output connections (Figure 3).

The effective permeability is different for the two windings. The permeability for the primary is very high, because primary flux is almost entirely confined to the ferrite material, whereas the secondary flux passes mostly through free air. If we assume that there is no hysteresis in the ferrite and that the B-H curve (Figure 4) can be modeled as a hyperbolic tangent, we can derive some equations to describe detector performance.

$$B = B_{sat} \tanh\left(\frac{\mu_i}{B_{sat}} H\right) \qquad (3)$$

B = magnetic flux density
H = magnetic field
μ_i = permeability at $H = 0$, initial permeability
B_{sat} = saturation flux density

The above formula can be verified by calculating dB/dH at H = 0:

$$\left.\frac{dB}{dH}\right|_{H=0} \qquad (4)$$

$$= B_{sat}\left(\frac{\mu_i}{B_{sat}}\right)\operatorname{sech}^2\left.\left(\frac{\mu_i}{B_{sat}} H\right)\right|_{H=0}$$

$$= \mu_i$$

The value of μ_i must be known for the ferrite sleeve, and is generally available in suppliers' catalogs [3].

Expanding equation 2 and noting that

$$\mu = \frac{dB}{dH},\ I = I_0 \cos(\omega t),\ H_{int} = \frac{N_p I}{d} \qquad (5)$$

then,

$$V = -N_s a H_{ext} \frac{d}{dt}\left(\frac{dB}{dH}\right) \qquad (6)$$

After much manipulation, (left to the reader as an interesting challenge), we get the following expression for instantaneous detector output voltage when measuring external magnetic field H_{ext} along the long axis of detector:

$$V = -2H_{ext} N_p N_s a \mu_i^2 I_0 \omega \cdot \qquad (7)$$

$$\frac{\sin(\omega t)}{B_{sat} d} \cdot \frac{\sinh\left[\dfrac{N_p \mu_i I_0 \cos(\omega t)}{B_{sat} d}\right]}{\cosh^3\left[\dfrac{N_p \mu_i I_0 \cos(\omega t)}{B_{sat} d}\right]}$$

V = secondary output voltage [V]
B_{sat} = saturation flux density of ferrite [T]
I_0 = peak value of primary current [A]
ω = frequency of input current [rad/s]
N_p = number of turns in primary winding
N_s = number of turns in secondary winding
a = cross-sectional area of both ferrite sleeves [m^2]
d = inside diameter of ferrite sleeve hole [m]
H_{ext} = external magnetic field to be measured [A/m]
H_{int} = internal, saturating magnetic field [A/m]
μ_i = initial permeability of ferrite sleeves (absolute, not relative) [H/m]

The above equation states that the amplitude of the secondary signal is directly proportional to the magnetic field parallel to the long axis of the detector. It is not immediately obvious that equation 6 is periodic in 2ω, but we can substitute some practical numbers as shown below, and then plot it, (Figure 5).

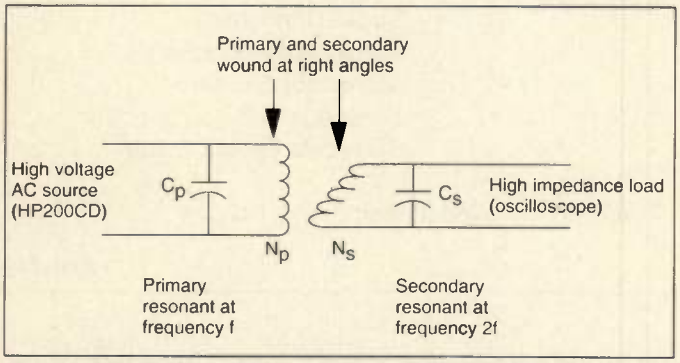

Figure 3. Schematic of flux-gate magnetometer including important resonating capacitors.

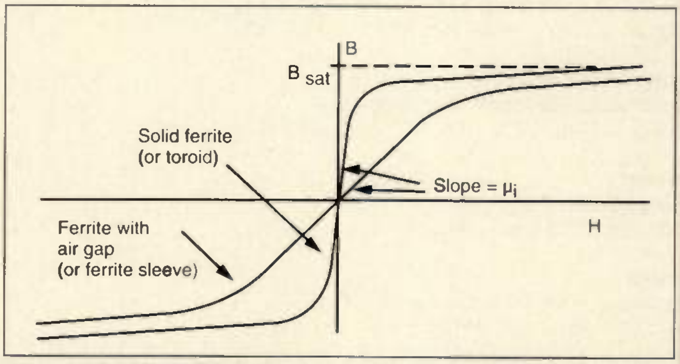

Figure 4. Simplified B-H curves for toroid and sleeve of the same ferrite material.

$B_{sat} = 0.2$ T
$I_0 = 0.2$ A
$\omega = (2\pi)100$ kHz
$N_p = 5$, $N_s = 200$
$a = 2\times10^{-6}$ m^2 (2 square millimeters)
$d = 0.8\times10^{-3}$ m (0.8 mm)
$H_{ext} = 39.8$ A/m (0.5 gauss)
$\mu_i = 2\times10^{-4}$ (= 160 x μ_0)

The output is almost 4 volts peak to peak in a 0.5 gauss field (approximately equal to Earth's magnetic field), and there definitely is a strong second harmonic component in the waveform.

Construction

A practical detector can be constructed using two ferrite sleeves, (Fair-Rite part number 2673002201 or 2643002201), and using the number of turns and frequency specified in the example above. The 43 material gives higher output voltage than 73 material, but its temperature stability is worse. The resonating values of the two capacitors can be found by experiment, using the detector's response to the Earth's magnetic field. First, decide on the input operating frequency (usually in the high 100's of kHz) and determine the output capacitor value C_s required for resonance of the secondary at twice this frequency. Then feed the required signal into the primary and adjust C_p for peak output voltage. There is plenty of output voltage, even using the Earth's magnetic field, so that this procedure can be carried out by orienting the detector with its long axis parallel to the ambient field (the orientation is almost vertical in U.S. and Canada).

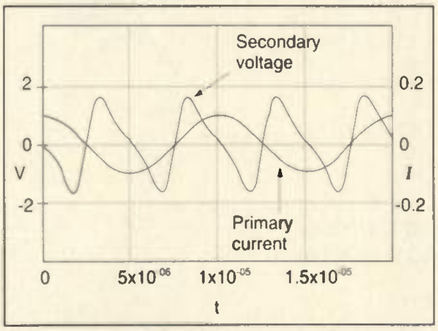

Figure 5. Theoretical detector voltage in 0.5 gauss field showing its second harmonic relationship to primary saturating current.

Despite the sleeves' small size (they are approximately 0.4 inches long and 0.072 inch in diameter), the amount of signal required for good saturation is surprisingly high. An HP200CD oscillator at full output, (producing 80 V p-p into an open circuit!), was required for satisfactory operation. The culprit is the high primary inductance; high input voltage is required to force sufficient current into the primary winding for saturation.

Equation 6 shows that the operating frequency should be as high as possible for a sensitive detector, but in practice the frequency is limited to several MHz or below, because of the self-resonance of the secondary winding.

Desirable qualities for the ferrite sleeves are high permeability, low saturation flux density, and for the sleeves to be long. Large cross-sectional area would also be beneficial, but it would also increase primary inductance, which is not desirable. In addition, the ferrite material should have high resistivity in order to limit eddy currents, and low hysteresis to minimize heating. Perminvar ferrites,

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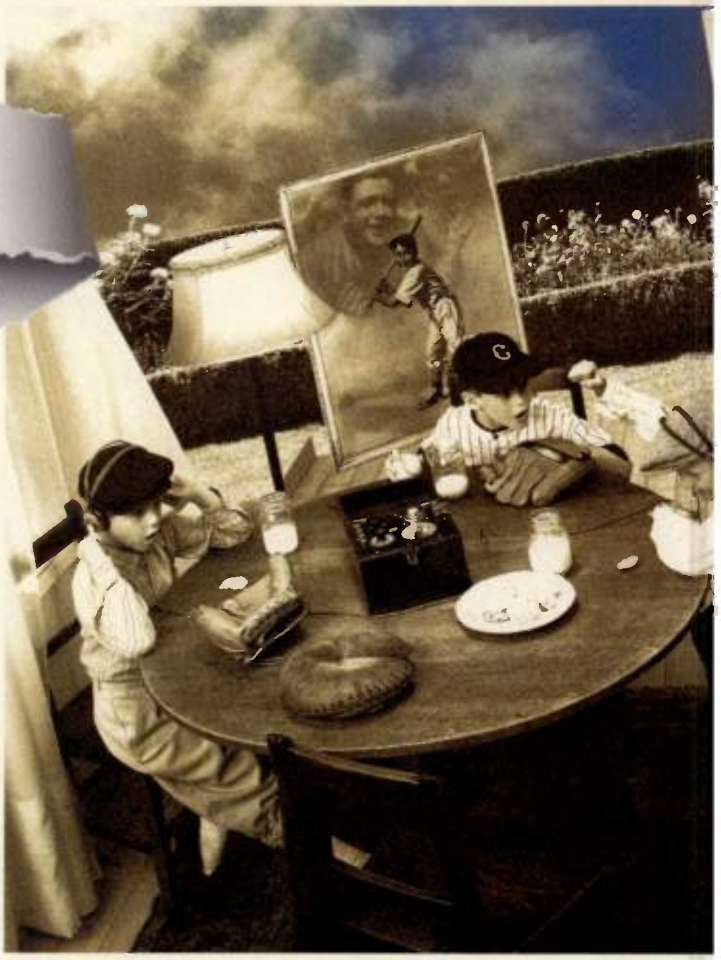

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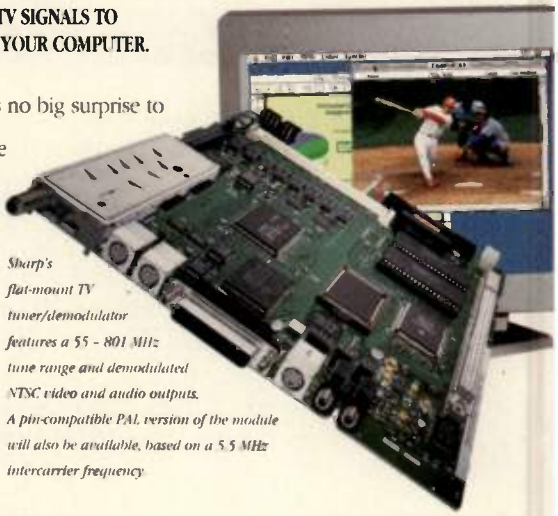

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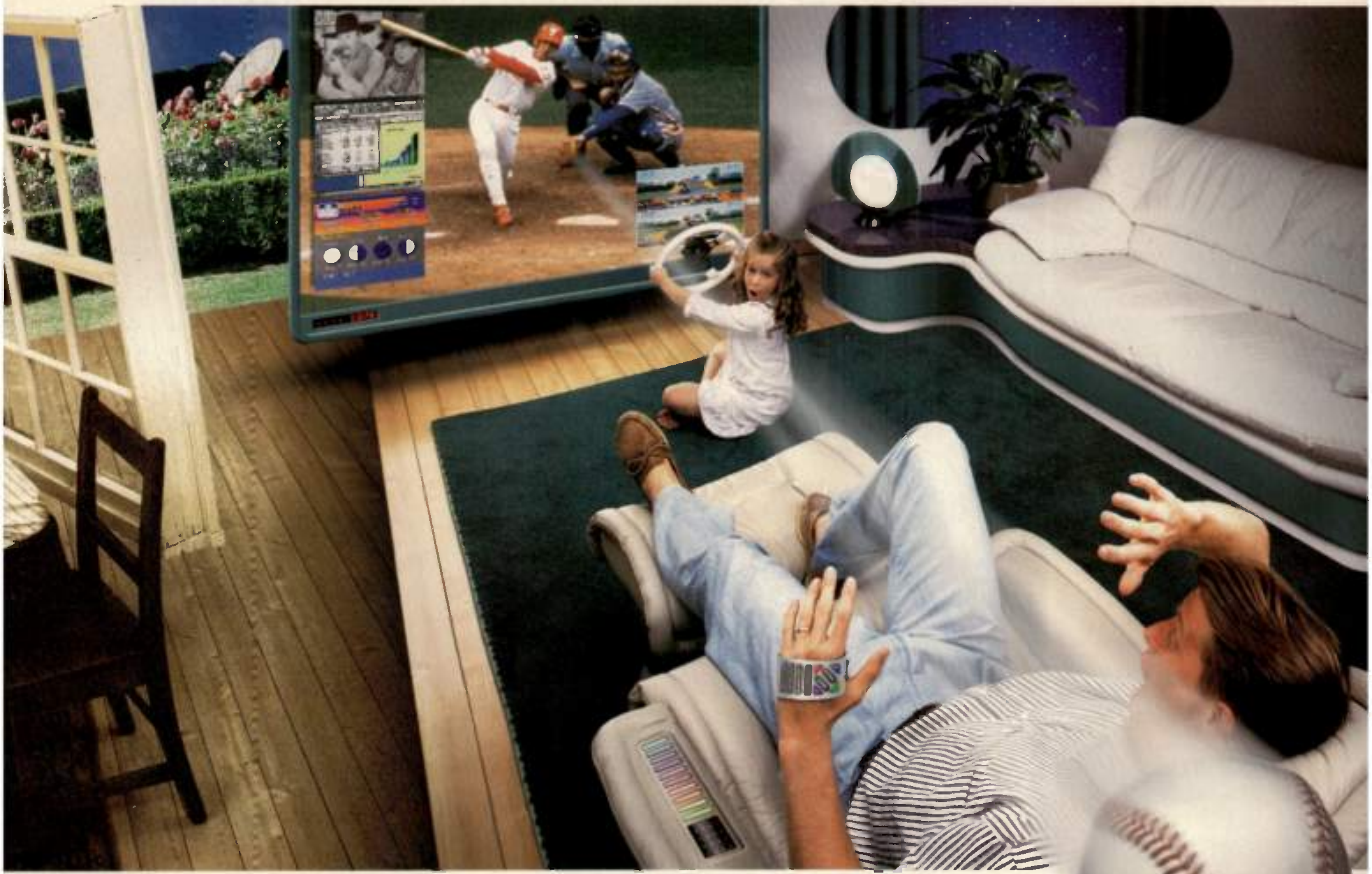

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which have nickel-zinc-cobalt composition, are not suitable for flux-gate magnetometers because their domain structure is semi-permanently changed by high magnetic fields [4]. The consequence of this is an effect similar to dielectric absorption in capacitors: the detector output voltage tends to drift toward its previous value following a disturbance.

Performance over temperature is probably the single most important challenge in designing a flux-gate magnetometer. Stable operation will invariably require some form of thermal compensation.

The most astonishing fact about flux-gate magnetometers is that under normal circumstances, their sensitivity is not limited by thermal noise. A permanent magnet, or a separate current-carrying coil can be suitably placed to cancel the ambient field, and the output signal amplified to increase the sensitivity even further. Even with field canceling and 30 dB gain, the detector output still does not contain much thermal noise. However, in addition to the ubiquitous 60 Hz signal, the output reveals events known as microfluctuations [5]: seemingly random variations of the Earth's magnetic field itself. It is difficult to imagine an object as large as the Earth changing its magnetic field in less than a second, but this is exactly what happens. These microfluctuations have been correlated to ionospheric disturbances, ground currents produced by distant thunderstorms, and even air tides [6].

Flux-gate magnetometers are easy to build and despite their simple structure, offer a rich set of design tradeoffs. Many variations of the basic design presented here are possible, and doubtless many more are yet to be discovered by interested experimenters. **RF**

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About the Author

Peter Vizmuller is a Staff Engineer with Motorola Canada, where he has worked with RF circuits and radio communication systems since obtaining his M. Eng. degree from the University of Toronto in 1981. He authored a book on helical resonator filters and published several articles. He has used the device described in this article to verify the daily variation in the Earth's magnetic field. He can be reached at (416) 756-5894.

Solar Source Antenna Gain Degradation Measurements

By Glenn W. Hurley
Naval Sea Support Center Atlantic

The topic of antenna gain measurements stirs up visions of near and far field measurement set-ups, well equipped indoor and outdoor ranges, and special normalizing software and computers to support these tests. This is to be expected for manufacturers and systems integrators, but how does the end user (field personnel) determine whether an antenna is degrading gracefully?

The Navy, who is one of the microwave industry's largest customers, has had to answer this question for many years. The sea, ever unforgiving in its relentless abuse of every product man has known, has no pity on large microwave antennas used in the Navy's many radars. Salt air corrosion, vibration, and radome abuse are only a few of the causes of gradual degration. With the aid of several microwave installations on each coast and a few overseas, the Navy has been able to check the critical parameters of many of these systems on a regular basis, during overhauls and pre-deployment periods. With the "down-sizing" of the Defense Department, and specifically the Navy, most of these microwave checking installations have been decommissioned and replaced with portable towers. Also, changes in system specifications have eliminated some of these mandatory checks, further reducing the availability of these systems for testing. This, along with reduced manpower, has forced the Naval Sea Support Center to look for some simple method of establishing a confidence level of performance for the antennas used on these systems. The following discussion will describe the results of one simple method found for checking antenna sensitivity.

In an effort to verify the gain parameter with minimal equipment, shore facilities, and manpower, the Naval Sea Support Center, Atlantic has been testing a procedure using the sun's energy as a signal source and a spectrum analyzer as the measurement device. The equipment chosen for this test was a ship based, G-band (5.0 - 8.0 GHz) tracking radar. In order to produce the required signal-to-noise ratio (S/N) at the spectrum analyzer, the introduction of a low noise amplifier (LNA), at the antenna terminal, was necessary on this radar. Many surveillance radars have LNAs already incorporated in their design. The LNA was specified by the Naval Sea Support Center and two were built by Microwave Solutions, of National City, CA.

Becuase only one technical person is needed to make the measurement, this procedure solves the manpower and minimum shore support requirements. The big advantage to this method is that the sun is a universal source of RF energy, and is in the antenna's far field, eliminating any near field corrections.

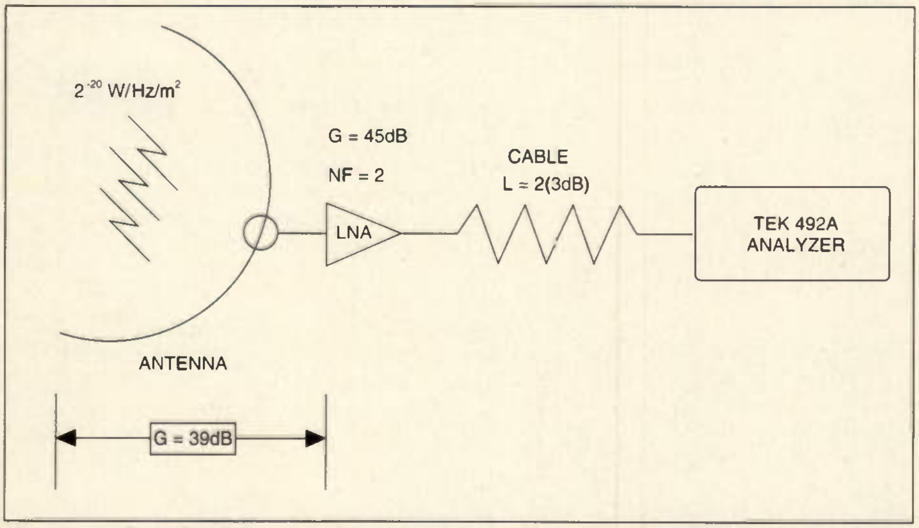

Figure 1. Block diagram of test set-up showing gains, losses and noise figures contributing to the antenna gain measurement.

Test Methodology

This procedure is a modified version of one previously published [1]. The older procedure requires the measurement of power per unit bandwidth and a comparison of cold-sky to solar flux density ratio. Measuring the power per unit bandwidth is relatively simple with the new generation of spectrum analyzers, which use digital processing and incorporate dBm/Hz (power per 1 hertz bandwidth) normalizing. This mode eliminates the need to keep track of resolution bandwidth changes and the need to renormalize the measurements when resolution bandwidth is changed. The specifications for the LNA were driven by the requirement to provide a S/N ratio 10 dB or more above the spectrum analyzer noise floor. This criteria resulted in the selection of an LNA with a gain of 45 dB and a noise figure of 3 dB, as seen in Figure 1.

The modification of the procedure [1] involved eliminating the sky noise measurement. The older procedure required a cold-sky measurement because the procedure made no assumptions about the antenna gain or effective aperture. However, the antenna referenced in this article is factory specified to be 39 dB (including waveguide losses). With the sun's energy specified in watts/bandwidth/area, one can compute the received power based on antenna effective area, so no knowledge of the background sky noise is required. The

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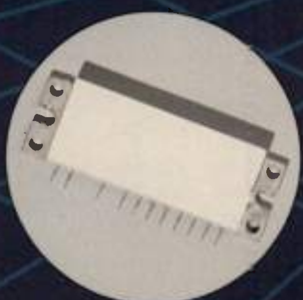

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RFP-88-108-90-800	88-108	±0.3	±1.5	800	23.0	1.15:1	0.2
RFP-101-501-90-200	100-500	±0.6	±1.5	200	20.0	1.15:1	0.5
RFP-101-501-90-400	100-500	±0.6	±1.5	400	20.0	1.15:1	0.5
RFP-251-102-90-200	250-1000	±0.5	±1.5	200	20.0	1.15:1	0.5
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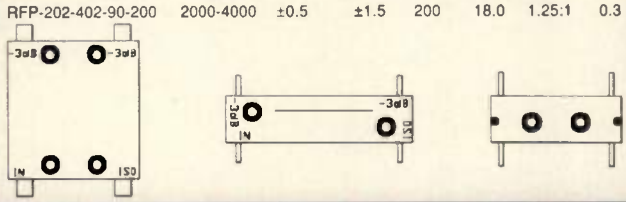
RFP-102-202-90-400	1000-2000	±0.5	±1.5	400	20.0	1.20:1	0.3
RFP-202-402-90-200	2000-4000	±0.5	±1.5	200	18.0	1.25:1	0.3



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Latin America and other international inquiries, contact Corporate Headquarters at (708) 208-2200 measured power, corrected for hardware/cable losses, (which are known system constants), and LNA gain is then compared to the measured solar data from the National Institute for Standards and Testing (NIST).

NIST serves as the data bank for several solar flux density measurement institutions around the world. Large calibrated sensors, located in Maine, Hawaii, Australia and Italy, collect data and pass it to NIST on a daily basis. This data is available from their Colorado location by phone (303-497-3171). This data is given at several bands from VHF to EHF, and is expressed in watts/hertz/square meter. All of the measurements in this article were taken on the East Coast, therefore solar flux density data from the Maine site was requested, since it represented approximately the same longitude. Figure 2 contains both a table and a graph showing the solar flux density as a function of frequency for two different time periods. Since NIST only specifies energy at certain frequencies, interpolation must be performed to find the solar flux for the frequency of interest.

The solar flux density, during normal

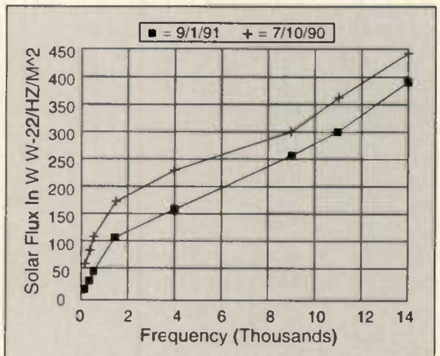


Table 1
Solar Flux Density Wx10^{-22}/Hz/M^2

Freq. MHz	09/01/89	07/10/90
200	15	60
400	30	85
600	50	110
1500	110	175
4000	160	230
9000	255	300
11000	300	360
14000	390	440

Figure 2. Solar flux measurements taken at Maine measurement location at two different times.

non-flare up periods, increases in power versus frequency in a linear fashion, approximately 3 dB per octave and follows the relationship,

$$Fx = 3.6 \times 10^{-30} \times f \quad (1)$$

where Fx is the flux density in watts per Hz per square meter and f is the frequency in hertz [3]. The sun subtends 0.5 degrees in the sky, therefore measurements of antenna gain where the beam width is less than 0.5 degrees should have a correction factor applied [1].

Using the relationship of antenna gain found in Skolnik [2],

$$\text{Gain} = A_e \times \frac{4\pi}{\lambda^2} \quad (2)$$

and solving for A_e, the effective antenna area, was found to be 1.6 square meters in the G-band.

The LNA noise contribution (3 dB), for the antennas tested using this procedure, was found to be +0.51 dB. With a noise figure of 3 dB the LNA contributes noise power:

$$P = kTB \quad (3)$$

$$P = (1.38 \times 10^{-23} J/K)(290K)(1\ Hz)$$

$$= 0.4002 \times 10^{-20} W/Hz$$

$$3.6/3.2 = 1.125 \text{ or } 0.51\ dB$$

which is the power referenced to the input of the LNA. The solar flux multiplied by the antenna's effective area of 1.6 m^2 A_e would yield an additional 3.2 × 10^{-20} W/Hz of noise power (signal). The total, 3.2 + 0.4 × 10^{-20} = 3.6 × 10^{-20} W/Hz, would yield an LNA input power that is 0.51 dB higher than predicted.

$$3.6/3.2 = 1.125 \text{ or } 0.51\ dB \quad (4)$$

With larger reflectors, for example one having an A_e of 6 meters, an error of only 0.142 dB would be realized.

Procedure

a. Record the spectrum analyzer power as "X" (dBm)

b. Calculate received power =

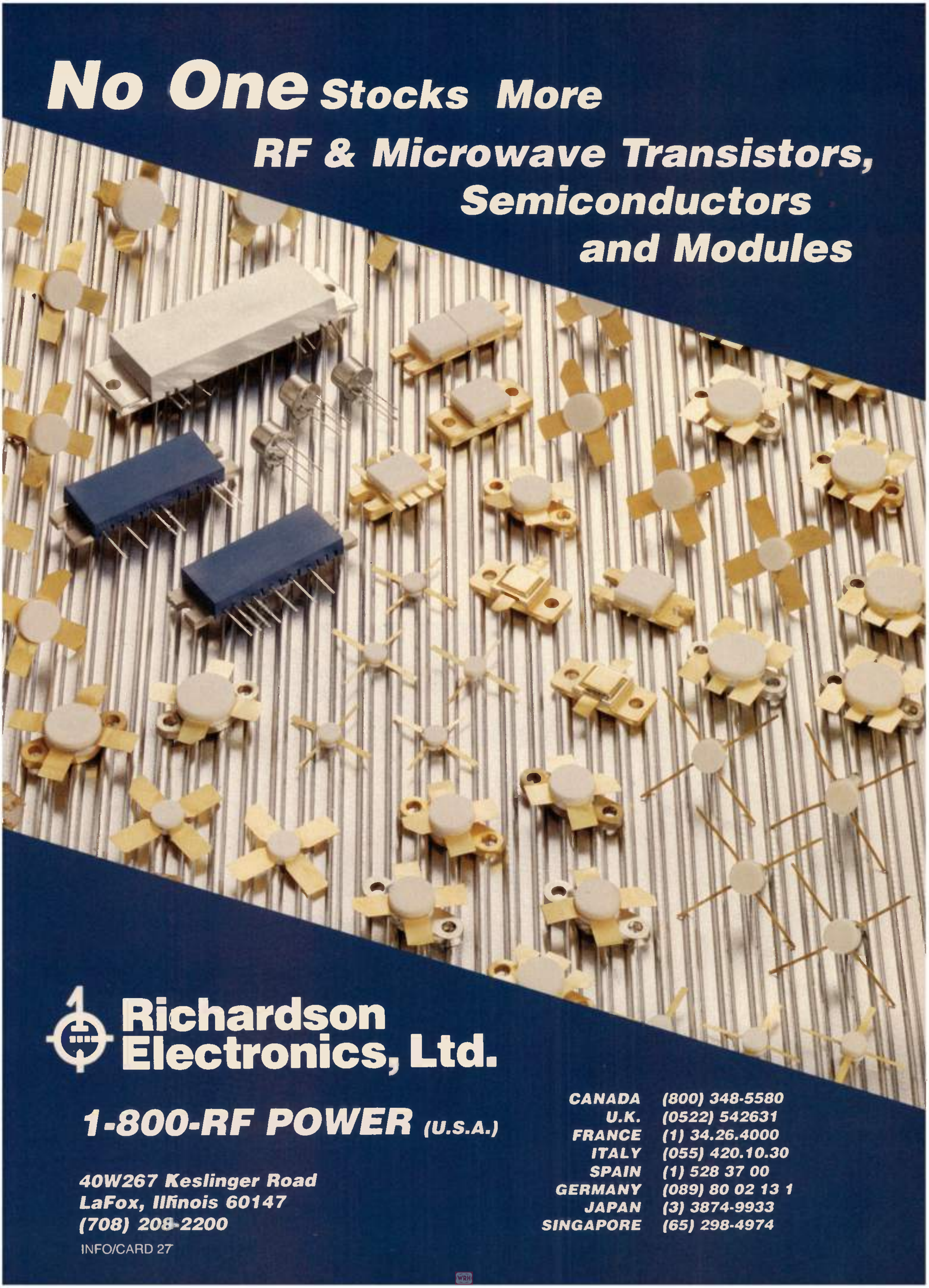
$$= \frac{Fx(A_e)(G)}{Loss} \quad (5)$$

G = LNA gain

Loss = coax loss

c. Convert the result of step B to dBm.

d. Calculate difference between power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measured in step A and power calculated in step C +/- any corrections

Example

Measured power (X) = −123.2 dBm/Hz

The solar flux from NIST, (F_x), at 4 GHz is 2 x 10^{-20} W/Hz/m^2. The effective area, (A_e), of the antenna at 4 GHz is 1.6 m^2, the low noise amplifier gain is 31622.8 (45 dB), and cable loss is 2.0 (3 dB). Calculating the theoretical power received we get:

$$\frac{2\times10^{-20}(1.6)(31622.8)}{2.0} = 5.1\times10^{-16} \quad (6)$$

Converting 5.1×10^{-16} W/Hz to dBm/Hz, we get −122.9 dBm/Hz.

Subtracting the power due to the noise contributed by the LNA from the measured power (X):

$$-123.2 \text{ dBm/Hz} - 0.51 \text{ dB} = -123.71 \quad (7)$$

The difference between measured and calulated powers is −0.81 dB. Therefore:

$$\text{gain} = 39 - 0.81 = 38.19 \text{ dB} \quad (8)$$

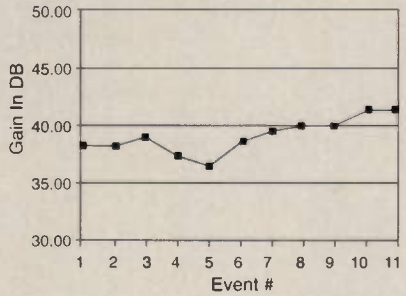

Table 2
Measured Antenna Gain

Event	Gain	
1	38.5	
2	38.3	**Mean**
3	39.1	39.06
4	37.4	
5	36.6	
6	38.5	
7	39.4	**Std. Dev.**
8	39.9	1.32
9	40	
10	41	
11	41	

Figure 3. Test results of 11 antennas using the solar source measurement technique.

Results

The test results from 11 systems, taken over 18 months, are displayed in Figure 3. Using the factory specified 39 dB gain as reference, the mean of the data was 0.06 dB high, with a standard deviation of 1.32. No effort was taken to check these systems using alternate procedures due to location and availability. This method of field testing, using solar energy, only applies when the factory gain is previously known. When testing gain on antennas where the effective aperture area is unknown, the cold sky to solar flux density ratio method is requirec.

Conclusions

Even though this procedure was accomplished at “G” band, it could also be applied at any frequency that NIST data is available and on any system where the antenna/LNA figure of merit provides a signal to the spectrum analyzer greater than 10 dB above the analyzer noise floor.

The author wishes to thank those who helped prepare th s paper, especially J. Laulainen of Naval Surface Weapons Center, Port Hueneme, CA. *RF*

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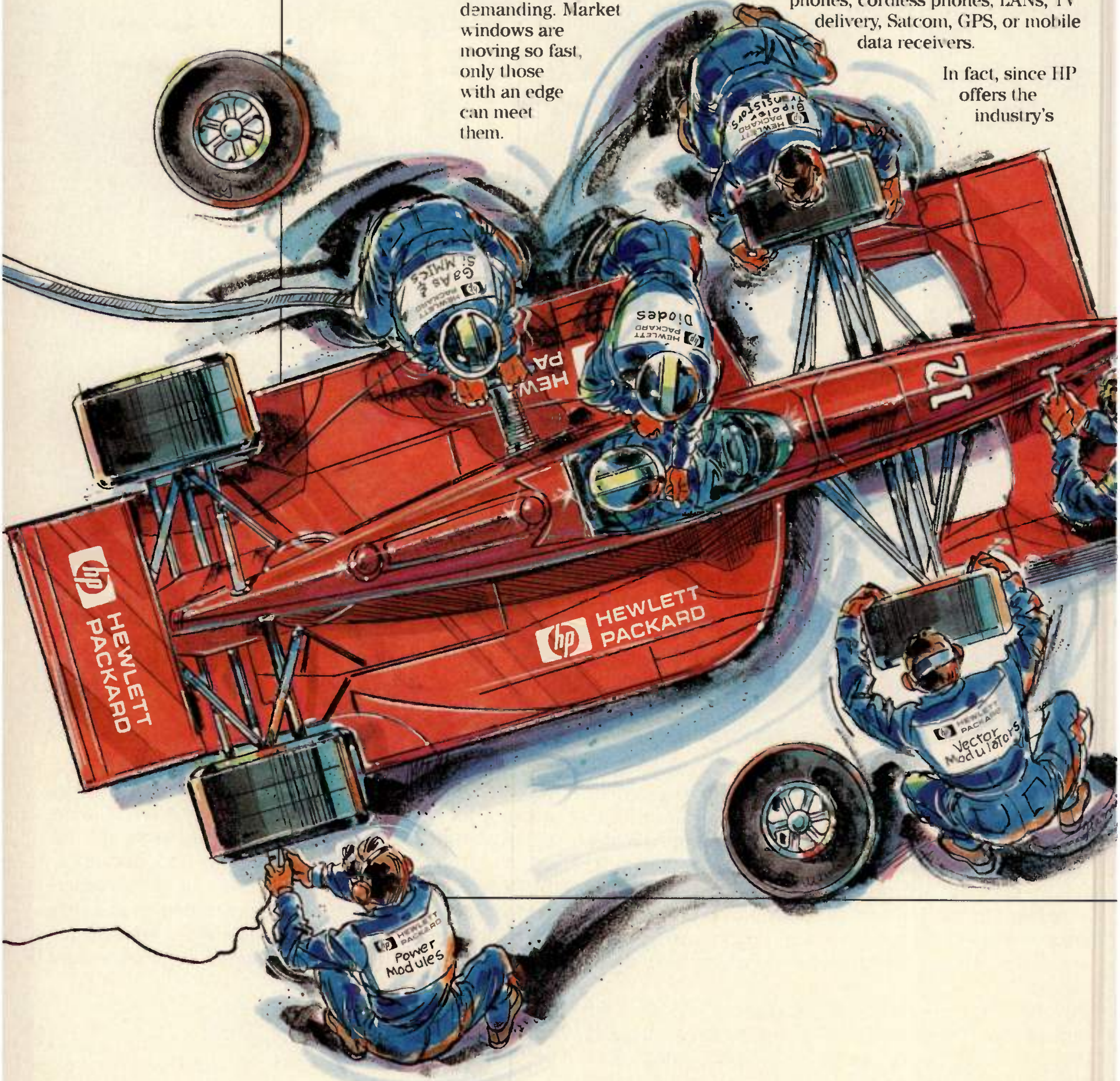
About the Author

Glenn Hurley received his degree from the DeVry Institute and began working for Sperry Gyroscope in 1963. Since then he has passed the EIT and now works for the Dept. of the Navy, Naval Sea Support Center in Norfolk Va. He can be reached at (804) 396-0435.



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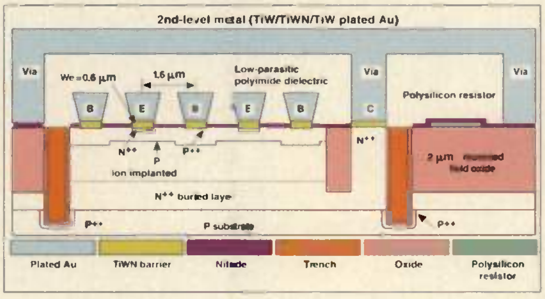

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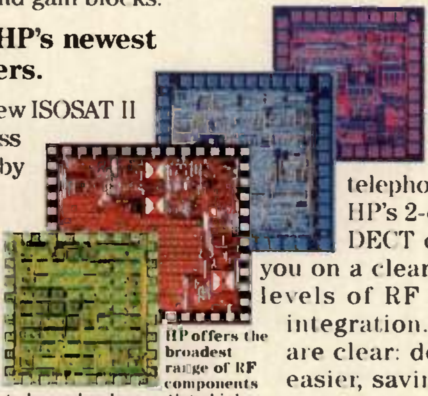

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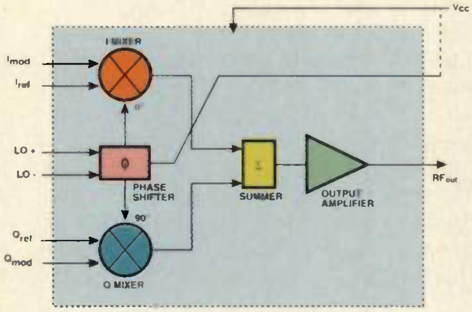

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Multipath Fading Emulation for Wireless Communications

By Bent Hessen-Schmidt
Noise Com, Inc.

The operating environment of a mobile digital radio system requires designers to allow for variations in propagation. The greatest propagation effect is multipath fading, which occurs when two or more replicas of a radio signal arrive to the receiver with different time delays. Testing for multipath performance is essential for reliable system operation — it is the mobile telephone user who experiences this multipath fading and, therefore, will associate it with the quality of the product he or she is using.

A typical multipath scenario is a mobile phone moving through an environment consisting of buildings, trees, ground, clouds, etc. The transmitter is omni-directional, and its energy is imposed on each of these objects. Objects with irregular surfaces scatter the reflected signal with varying amplitude and a random phase shift that is usually evenly distributed from 0 to 360°. As they arrive at the receiver, the direct and reflected signals ad as vectors, causing the amplitude of the received signal to vary, and in the worst case, approach zero and interrupt the communication.

Wireless communication systems such as GSM, DCS1800, CDMA, IS-54, DECT, and CT-2 mobile telephones are required to be tested while subjected to multipath fading [1]. Truly, most line-of-sight communication systems can benefit from multipath fading tests because the design margin for multipath fading is relatively large compared to the allowances for noise figure, carrier to noise ratio, and antenna gain (see Table 1). While all the other allowances usually are tested, the emulation of multipath has just recently become practical with the introduction of new instruments.

Operation of the MP-2400

Multipath fading can today be emulated in the lab and at the manufacturing facility by use of a multipath fading emulator, such as Noise Com's MP-2400. The RF link between the transmitter and receiver affects the signal in ways that can be emulated by multiple paths, normally up to 12 paths, each having an associated Doppler shift, time delay, attenuation, and Rician or Rayleigh (classical) scattering. The multipath fading emulator accepts inputs directly from a transmitter at up to +43 dBm (20 W). It downconverts the signal to an intermediate frequency where it is passed through a large dynamic range, 12 bit, A/D converter and thereafter split into 12 paths each emulating a reflected signal. The signal is split after A/D conversion to assure that each path is fed with exactly the same signal.

Each path will have a signal strength that most likely is less than the power of the direct path. The average power of

Transmitted power (2 W)	(Px)	+43 dBm
Antenna gain, transmitter	(Gt)	+3 dB
Antenna gain, receiver	(Gr)	+3 dB
Loss due to distance between Tx and Rx	(L)	−100 dB
Nom nal Received Carrier Power	(C)	101 dBm
Data Rate (64 kbit/s)		48 dB/Hz
Nominal Received Bit Energy	(Eb)	−149 dBm/Hz
Noise Floor	(kT)	174 dBm/Hz
Noise figure	(NF)	7 dB
Receiver Noise Density	(No)	−167 dBm/Hz
Bit Energy to Noise Density @ 10-9 BER	(Eb/No)	18 dB
Fading margin	(M)	50 dB
Nominal Received Bit Energy	(Eb)	−149 dBm/Hz

Table 1. Typical effect budget for a radio link.

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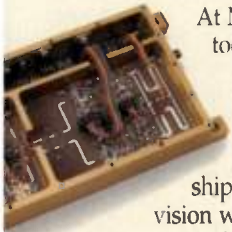

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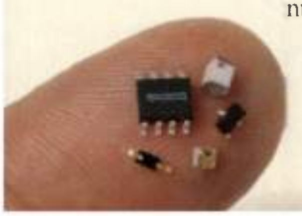

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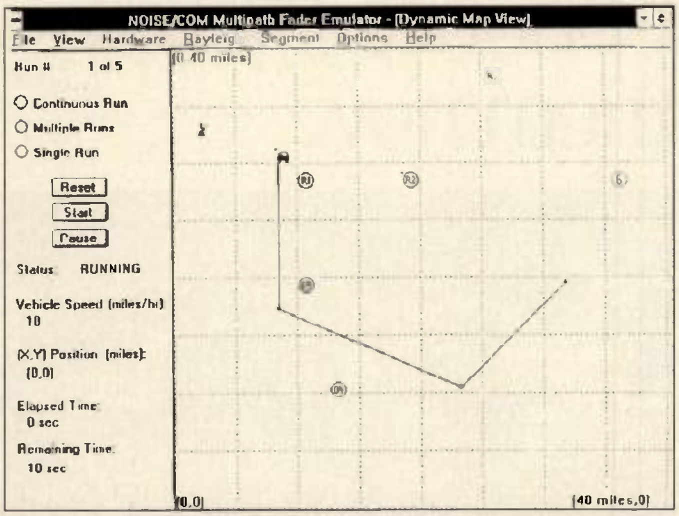

Figure 1. Multipath fading emulator, display for entry of fading path parameters.

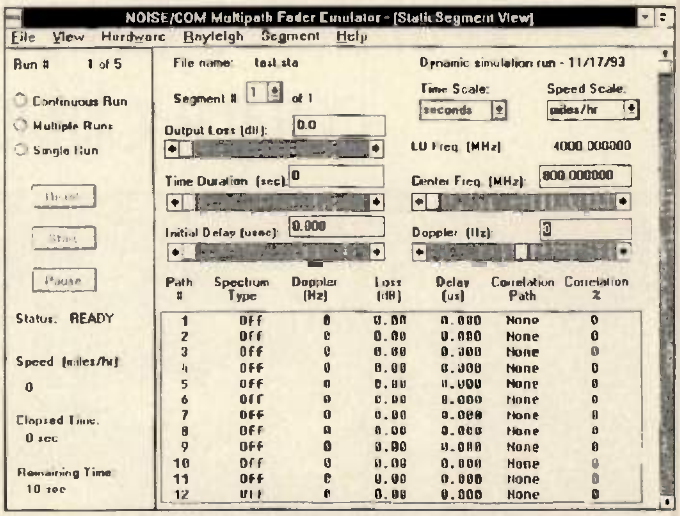

Figure 3. Display for dynamic mode emulation of fading from a mobile car telephone.

each path can therefore be set relative to that of the direct or strongest path. The reflected signals are then delayed relative to the direct because the paths of the reflected signals are longer. Again, this is all done digitally in order to obtain the best accuracy.

Paths with identical delay arise from reflectors located on an ellipse. All these paths will exhibit a Doppler shift due to the speed of the receiver, but the amount of Doppler shift depends on the angle of the reflecting object relative to the direct path, as:

$$F_d = \frac{2\,v\cos(a)}{c} F_c \qquad (1)$$

where v is the speed of the vehicle, F_c is the center frequency of the radio transmitter, c is the speed of light, and a is the angle of arrival.

If the mobile phone is moving directly away from the transmitter, then the Doppler shift of the direct path will reduce the frequency directly in proportion to the speed. If a reflector happens to be in front of the receiver, then the shift for that path will have equal magnitude but opposite sign of the direct path. Objects to the side of the receiver will cause a Doppler shift between these two extremes. For a scenario with reflected signals whose angle of arrival is evenly distributed over 360° the result is a spectrum which spreads the signal energy over a band which is the transmitted frequency (F_c) plus and minus the Doppler shift (F_d). The spectral power distribution is not flat, but has peaking at the Doppler offset since the summation of power from reflectors at or near 0° and 180° will contain more power. This Doppler spreading, which may be of

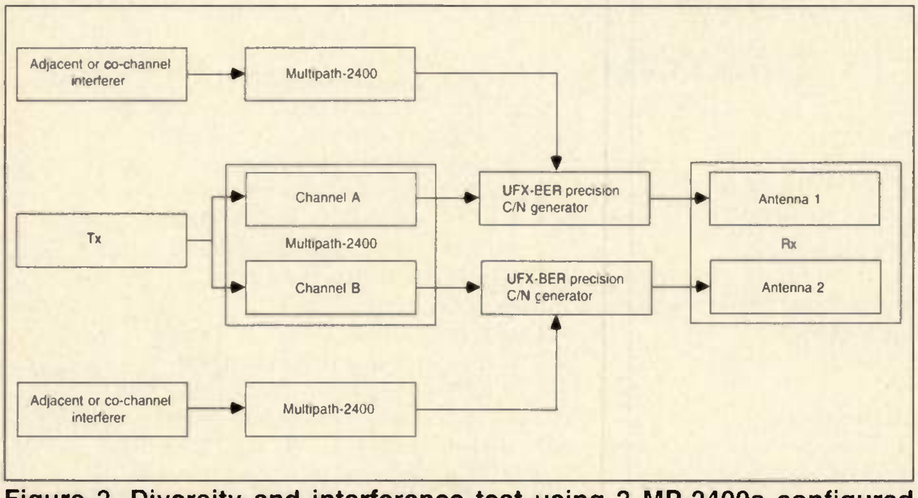

Figure 2. Diversity and interference test using 2 MP-2400s configured for 2 channels with 6 paths each.

Rayleigh or Rician nature, is obtained by digital modulation of the A/D converted signal of each paths. Even though each path has a random and independent phase distribution, the paths may be correlated due to the relationship to each other in the scenario. This correlation is achieved in the MP-2400 by summing a portion of the modulation signal from one path with a portion from another.

The required differential path delay, attenuation, and correlation are programmable for each path (Figure 1). The MP-2400 includes modules of identical hardware each containing 3 paths. Each path includes a MAC (multiplier/accumulator) which performs the function of the Doppler spreading by mixing the I/Q signal samples with two filtered Gaussian signal sources. Each source is shaped using a digital FIR filter with a bandwidth equal to the Doppler rate. The result is then delayed and attenuated digitally before being converted to an analog signal using a D/A converter. The paths (up to 12) are summed at the instrument output, emulating the summation which occurs at the receiver antenna. This summation of multiple signals with random phase causes the random amplitude fluctuations or fading. The maximum rate of fading is exactly the Doppler rate.

he composite signal is upconverted to a microwave frequency using dual conversion, and then downconverted to the received frequency in order to maintain an output free of spurious responses without local oscillator leakage and with images. The final output is purely the desired signal and the output power can

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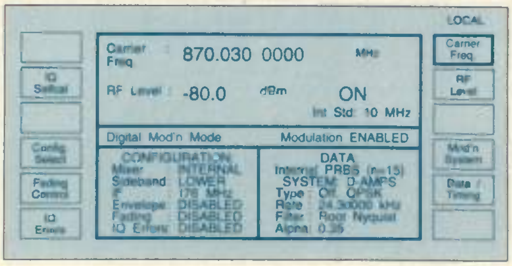

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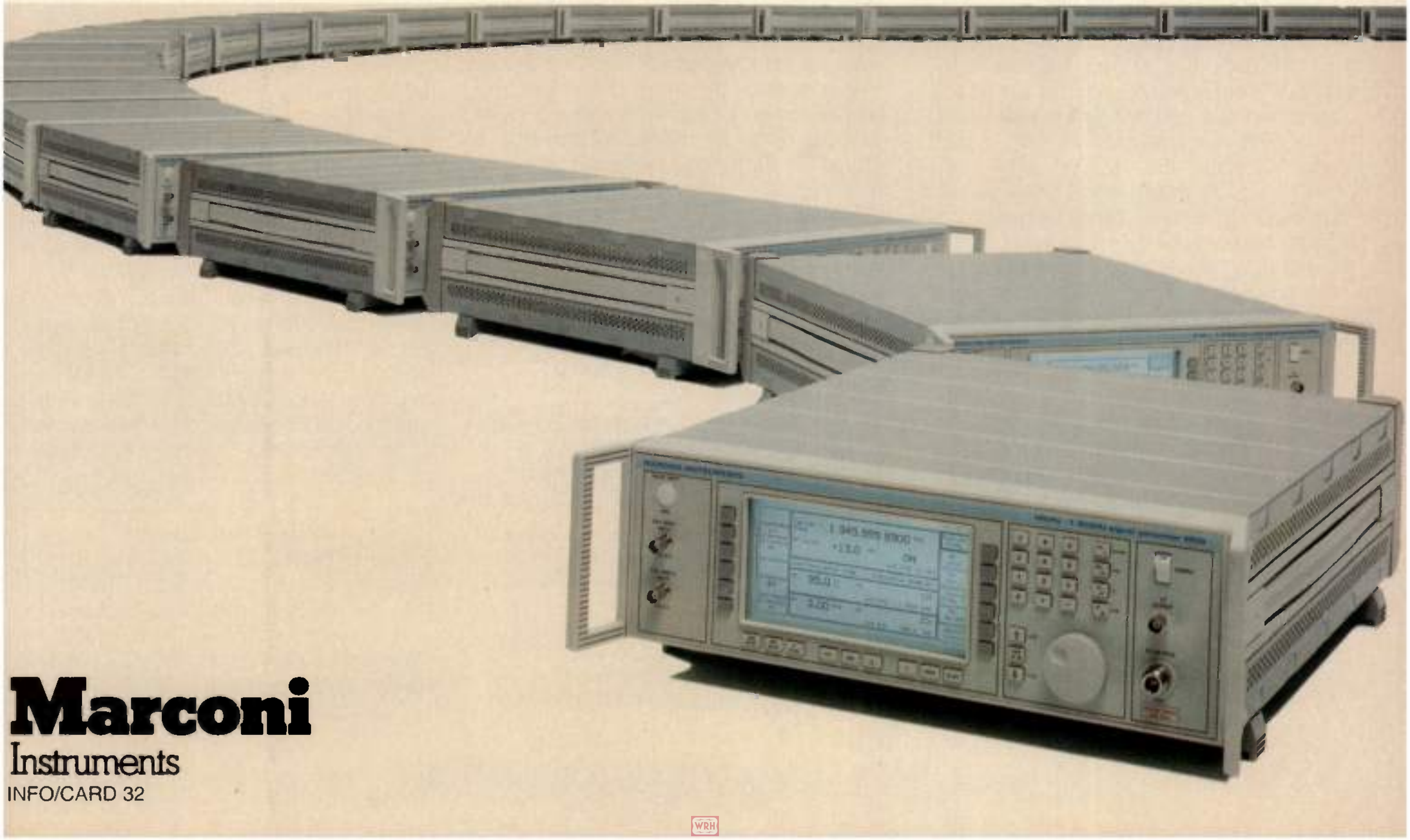

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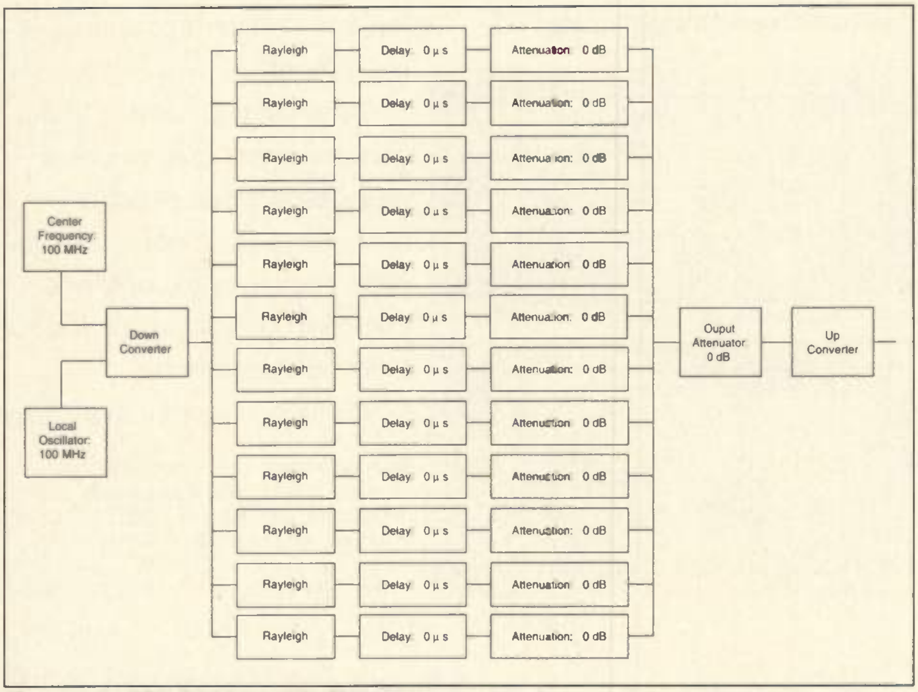

Figure 4. MP-2400 block diagram.

be set between –10 and –100 dBm to match the level expected at the receiver input.

Instrument Features

The MP-2400 emulates a wireless communication channel with up to 12 paths between a base station and a mobile station. It contains an embedded 486DX-33 microprocessor with PC compatible interfaces including an IEEE 488.2 interface to provide an easily transportable self-contained instrument. The built-in AT style keyboard, mouse and flat panel color LCD display make the MP-2400 an extremely powerful tool as a stand alone instrument or as a controller for an entire test station.

MP-2400 is easily upgraded to contain two independent channels of 6 paths each. These two channels are used to emulate multipath fading for the tests of diversity systems or to independently fade an interferer and a desired signal. The interferer may be an adjacent or co-channel signal as required for the test of GSM and DCS1800 mobile telephones. Figure 2 shows a block diagram of a diversity and interference test with two MP-2400s emulating 4 channels with 6 paths each.

Multipath fading can be emulated in both static and dynamic modes:

1) *STATIC MODE* — Here the user individually sets each path with the desired parameters and that scenario is run continuously. Standard test conditions for GSM, DCS, NADC (CDMA), etc. are factory stored and can be recalled from memory. Up to nine scenarios can be linked and run together.

2) *DYNAMIC (GRAPHIC) MODE* — The user enters the initial and final parameters for a dynamic scenario. For example, a mobile phone starts at rest, then accelerates to 96 km/h (60 Mph) over a 5 minute period. The user inputs the scenario including the reflecting objects on a graphic display, see Figure 3. Software calculates intermediate states for Doppler, delay and attenuation.

Figure 4 is the MP-2400 block diagram, showing a 12 path system. All signal processing is done digitally with 12 bit representation of the data. This provides very consistent and drift free results.

User Interface

After selection of mode (STATIC or DYNAMIC) the user will supply the following information to run a scenario:

- Center Frequency
- Rayleigh (Classical) or Rician distribution
- Vehicle speed or Doppler
- Start and Stop time
- Path characteristics: ON/OFF, Delay, Attenuation, Correlation

The path settings are displayed on the VGA monitor in a tabular form. The instrument status and graphic representation of certain scenarios, the Doppler spectrum, and the hardware configuration may also be displayed.

Other available features are a help key for setting up path characteristics for desired scenarios and Recall/Save of setups. The user has the option to select the best matching shape of the Doppler spreading from a library if the antenna patterns are different from the standard omni-directional model.

Thus, besides being a requirement, effective multipath testing of modern communication systems can also help make them more efficient. It is now possible to make modern communication systems which tolerate severe multipath fadings. These new wireless systems employing sophisticated modulation, diversity antennas, adaptive equalization, forward error correction coding, phase locked synchronization, and adaptive transmitter power control techniques. They may soon be tested using multipath fading emulators.

Readers desiring more information on the Noise Com MP-2400 may contact the author at the address or telephone number given below, or by circling Info/Card #150. ***RF***

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About the Author

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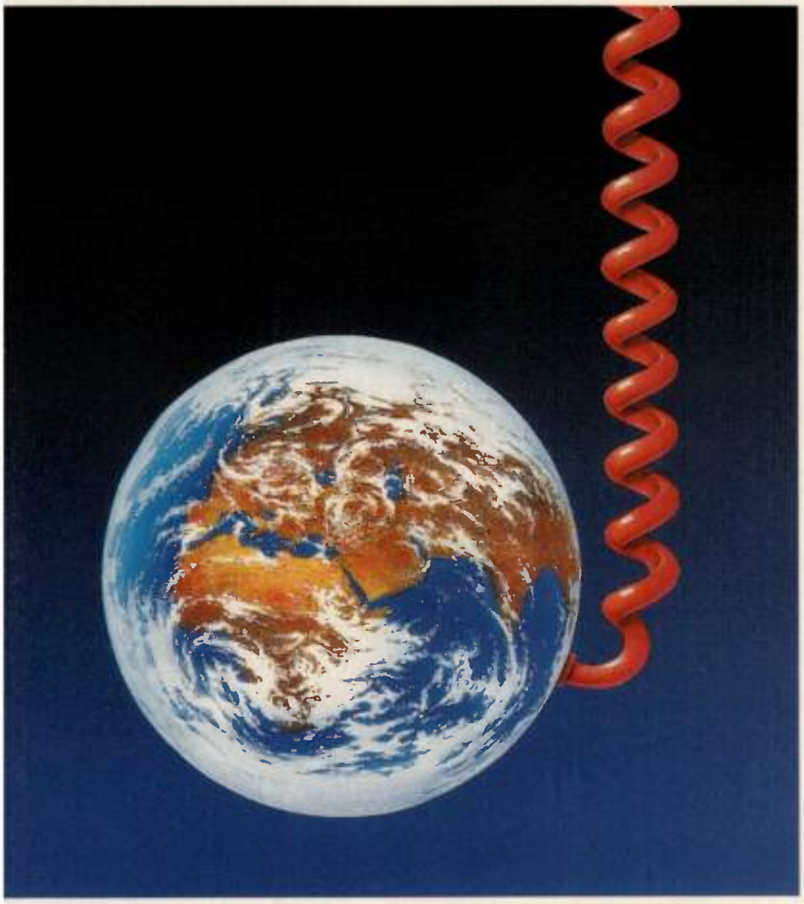

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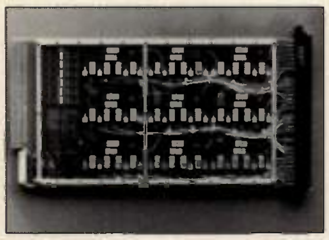
A new series of switching modules for RF applications is now available from Racal Instruments. The 1277-540 series permits switching of signals up to 1 GHz through the ARINC 608 interface. The switch is configured as nine 1x4 tree multiplexers and up to six RF splitters. With insertion loss of less than 0.5 dB, crosstalk of less than –95 dB, isolation of better than –80 dB and VSWR of 1.2:1, all at 100 MHz, the 1277-540 is ideal for switching of low-level RF signals. Maximum switchable RF power per channel is 60 W. The switch is available in three different models to facilitate different pin-out requirements. 1277-540C provides six RF splitters accessible through the interface connector. Model 1277-540B offers three RF splitters, while 1277-540A contains no RF splitters. The 1277-540 is designed for use in Racal Instruments' 1277 and 1262-77 mainframes. The former is an ARINC 608A mainframe and interface, while the latter is a VXIbus mainframe with an ARINC 608A sub-chassis. Prices begin at $5775, with delivery in 10 to 12 weeks.



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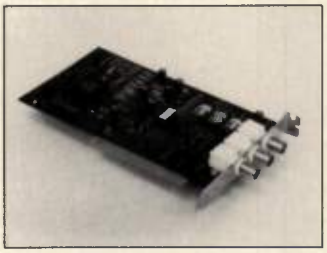

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INFO/CARD #248

40 MIPS DSP

The ADSP-21060 from Analog Devices is the industry's first DSP in the class of Super Harvard Architecture Computer (SHARC). It integrates the industry's fastest general purpose floating-point core (the ADSP-21020) with a dual-ported 4-megabit SRAM, communication ports, and a sophisticated DMA (direct memory access) controller on a single chip. The 32-bit single precision (or 40-bit extended precision) IEEE floating-point DSP core executes 40 MIPS per second, with 120 MFLOPS peak and 80 MFLOPS sustained. The I/O processor includes a DMA controller, memory mapper and communications which handles transfers between memories and ports without impacting DSP core performance. A software development system, including an instruction level simulator, an assembler, a linker, a C compiler and an extensive C runtime library is currently available. The ADSP-21060 will begin sampling 1Q94. Target pricing for 1000 piece quantities is $296.00.

Analog Devices, Inc.
INFO/CARD #247

250 W MOSFET Amplifier

A 250 watt, 14 octave MOSFET RF amplifier with an instantaneous bandwidth from 0.01 to 220 MHz has been introduced by Kalmus Engineering. Using special broadband ferrite circuitry in conjunction with new high power field effect tranaisitors, the 122FC Offers 55 dB minimum gain, with ±1.8 dB max. deviation from flatness (±0.6 dB when leveled). Three mW input is typically required for full output. Spurious outputs are -60 dBc minimum, and full power harmonics are typically -23 dBc. The class A amplifier is unconditionally stable and has open/short, temperature and overdrive protection circuitry. Forward and reflected RF power are metered, and temperature, VSWR, overdrive and blanking have their own indicators. All major front panel functions are available for remote control and monitoring. via IEEE-488 and RS-232 interfaces. The 122FC measures 48 x 51 x 39 cm and weighs 35 kg. Cooling is by forced air. Price for the 122FC is $19,950.

Kalmus Engineering Inc.
INFO/CARD #246

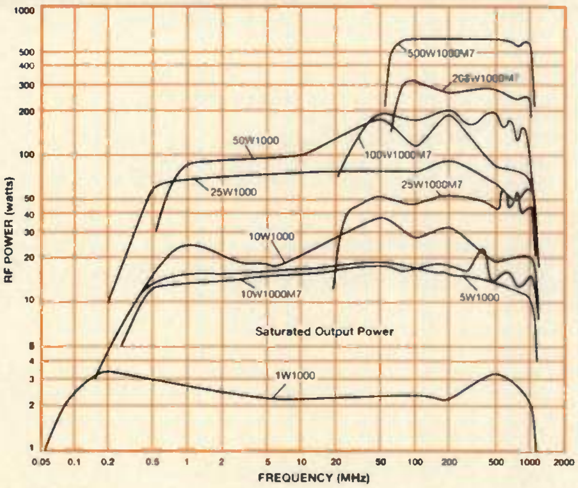
When it gets lonely out there at 1 GHz, you'll be glad you have all this power.

The only amplifier that can deliver 500 watts through that balky decade from 100 to 1,000 MHz is our new Model 500W1000M7. That's the majestic one on the left, with the front-panel bi-directional power meter, ALC, and a bagful of other controls that will let you actually *enjoy* automatic sweep testing in your lab.

The 500W1000M7 presently heads our well-known "W" Series of all-solid-state linear amplifiers that cover four crucial decades of bandwidth from 100 kHz to 1 GHz. Today, you may need only 1 watt (the little portable on the top), or 5, or 10, or 25, or 50, or even 100 or 200 watts *minimum*—all with that fantastic bandwidth instantly available without tuning or bandswitching—a combination of power and bandwidth that's comforting when you work the outer reaches of the rf spectrum. Wherever your power requirements are today, they're sure to go up. And one of these days you're going to want a 500W1000M7.

The "W" Series is part of a complete line of AR amplifiers offering rf power up to 10 kilowatts for rf susceptibility testing, nmr spectroscopy, plasma/fusion research, and a host of other test situations that demand rf power of unconditional stability—immunity to even the worst-case load mismatch, shorted or open cable, wild swings of VSWR—with no fear of oscillation, foldback, or system shutdown.

You might learn a lot from our free booklet, "Guide to Broadband Power Amplifiers." Send for a copy. Or call our toll-free number (800-933-8181), which will connect you directly to one of our applications engineers.



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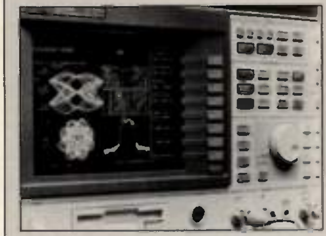
INFO/CARD 34

5351

Product Spotlight: Signal Analyzers

Vector Signal Analyzers

Hewlett-Packard has announced a vector-modulation-analysis option that gives designers of RF digital commu-



nications systems a single tool with which to analyze a wide variety of modulation formats. The new option works with the HP 89440A (DC to 1.8 GHz) and 89410A (DC to 10 MHz) vector signal analyzers and provides complete digital demodulation of several varieties of QPSK, DQPSK, MSK and QAM. Option AYA is priced at $4500.

Hewlett-Packard Co.
INFO/CARD #245

Vector Network Analyzers

Tektronix has introduced the Advantest R3762AH and R3763B vector network analyzers that combine exceptional performance from 300 kHz to 3.6 GHz. The Advantest analyzers are marketed and supported throughout North America by Tektronix under a strategic alliance formed earlier this year. The R3762AH and R3763B make measurements at up to 0.5 ms per point. The R3762AH and R3763B series vector network analyzers are priced at $24,000 and $28,000 respectively.

Tektronix, Inc.
INFO/CARD #244

TEST EQUIPMENT

Field Strength Meter w/ GPS and PCMCIA

"The Champ" RF field strength meter now is available with GPS navigation and PCMCIA memory options, allowing automatic logging of location and field strength information. The handheld meter is available for the 900 to 932 MHZ and 1.8 to 2.4 GHz bands, with other frequencies available on request. Sensitivity is –120 to –30 ±1 dBm (with 10 kHz IF BW). BER demodulation is also available.

Berkeley Varitronics
INFO/CARD #243

Extended BW Calibrators

A wideband option for the 4800 series of calibrators from Wavetek extends the instrument's frequency range to 30 MHz, enabling the calibration of a wide range of RF voltmeters. Option 70 uses direct digital synthesis to generate rms output for amplitudes up to 3.5 V, with response flatness better than 0.35 % over the entire bandwidth.

Wavetek Corp.
INFO/CARD #242

Counter

B+K Precision has improved their model 1856A multifunction counter. Its bandwidth is now 5 Hz to 2.4 GHz, with a temperature compensated crystal oscillator with 0.5 ppm stability from 18 to 28° C. Sensitivity at 2.4 GHz is 50 mV. Retail price remains at $499.

B+K Precision
INFO/CARD #241

Noise Sources w/ Isolators

Noise Com has introduced a series of coaxial noise sources with built-in isolators for ATE and RADAR applications. The NC 3400 series can be calibrated more accurately and has better flatness, (available to ±0.25 dB), due to its low VSWR. The built-in isolators also allow the noise source to withstand 100 W peak incident RF power. The series NC 3400 produces white Gaussian noise with 26 to 36 db ENR in octave bands up to 18 GHz.

Noise Com, Inc.
INFO/CARD #240

SEMI-CONDUCTORS

Amplifier/Mixer IC for 900 MHz

The HFA3600, from Harris Semiconductor, integrates a low noise amplifier (LNA) and single balanced mixer into one silicon chip to amplify and downconvert 900 MHz RF signals to a 100 MHz IF. For the combined ampli-

fier/mixer at 900 MHz, power gain is 19.1 dB, noise figure is 3.1 dB , and third order intercept point is –7.0 dBm, referred to the input. The HFA3600 draws 13 mA at 5 V. The device is packaged in a 14-pin SOIC. Cost is $3.56 in quantities of 1000.

Harris Semiconductor
INFO/CARD #239

Wideband, Low Distortion Op Amp

Burr-Brown's OPA628 is a low distortion, wideband operational amplifier that features 160 MHz bandwidth and 90 dB spur free dynamic range. The OPA628 is a unity gain stable, voltage feedback, monolithic operational amplifier with high input impedance, high common mode rejection and symmetrical differential input flexibility. Gain is flat to 30 MHz. Price is $6.30 in 100 piece qty.

Burr-Brown Corp.
INFO/CARD #238

DBC Tuner IC

Anadigics announces the availability of a new 950 to 1450 MHz direct broadcast satellite (DBS) tuner IC for digital applications. The ADC20014 incorporates a local oscillator with phase noise at -80 to -73 dBc at 10 kHz offset. The tuner has 10 dB noise figure, 5 dB conversion gain and operates from a single 5 V supply. The device is supplied in a 16 pin SOIC package.

Anadigics
INFO/CARD #237

1 to 2 GHz Power Transistor

Motorola has announced the availability of a microwave power transistor designed for class A, common emitter power amplifiers. Operating from a 20 VDC supply, the MRF2000-5L microwave power transistor delivers 5 W of RF output power for 1 W RF input power.

Motorola Semiconductor
INFO/CARD #236

Cellular DSP

AT&T Microelectronics has disclosed that it is sampling a 16-bit digital signal processor for cellular applications. The DSP1617 features 50 mega-instructions per second (MIPS), 4k RAM and 24k ROM on one chip, making it sufficient for both IS-54 speech processing and radio modem functions. Power consumption is typically 10 mW/MIPS at 5 V and less than 2.7 mW/MIPS at 2.7 V. The DSP1617 is packaged in 100-pin PQFP and TQFP packages, and is pin-for-pin compatible with the DSP1616 family. Pricing is $49 in PQFP and $53.90 in TQFP, in lots of 10k.

AT&T Microelectronics
INFO/CARD #235

AMPLIFIERS

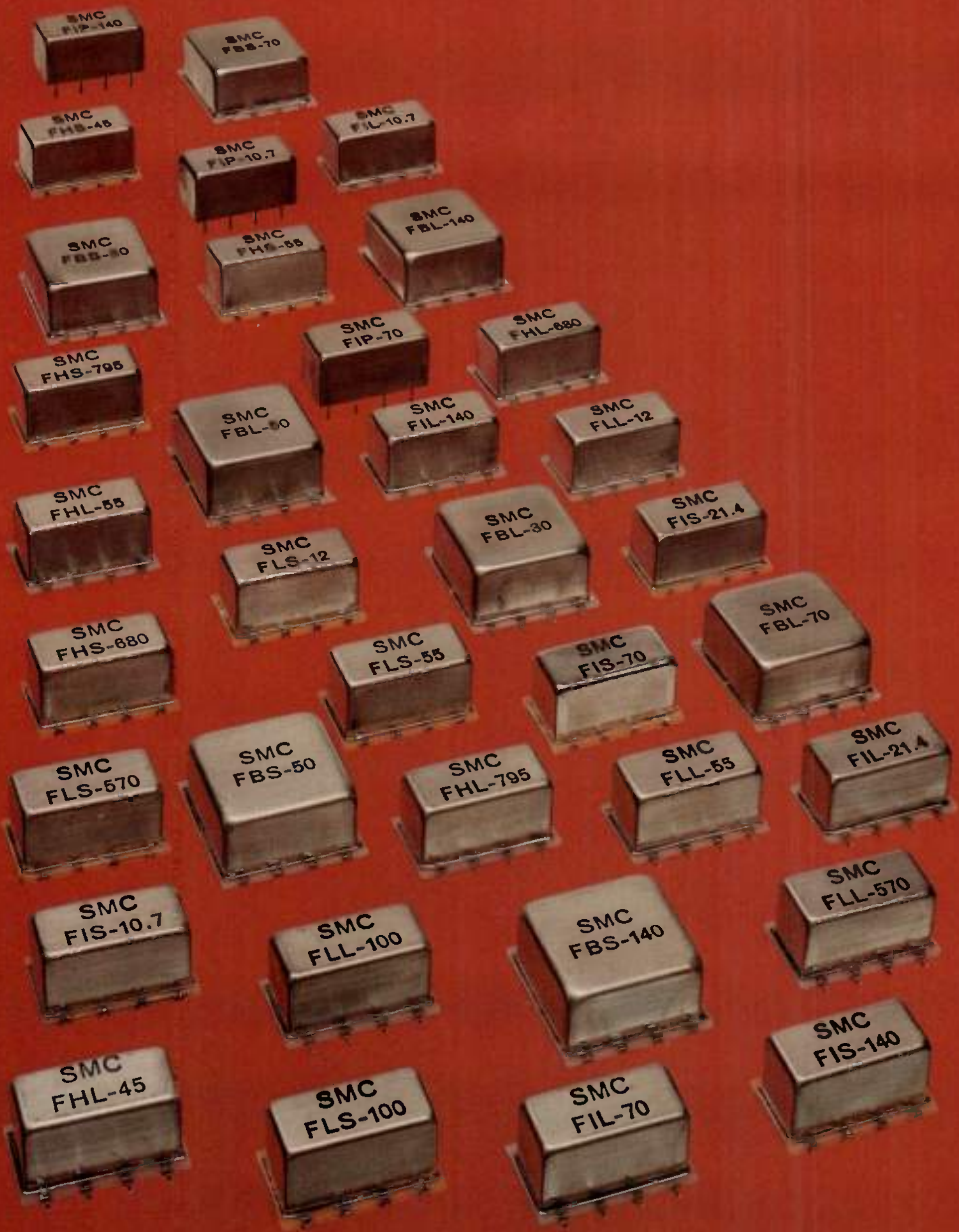
Cable Equalizing LNA

Model VMEA 118-106 amplifiers provide gain which rises from 6 dB to 11 dB at 18 GHz to compensate for 20' of cable loss, resulting in flat system IICP and gain. DC may optionally be supplied via the RF cable. NF is < 6 dB over most of the band, IICP3 is > +7 dBm. Size is 1.0 x 1.0 x 0.22 inches. DC power is +12 V at 80 mA.

Veritech Microwave, Inc.
INFO/CARD #234

100 W Class A Amp

ENI's model 3100LA RF power amplifier produces 100 W of linear class A output over the frequency range of 250 kHz to 150 MHz. The amplifier features a nominal gain of 55 dB (±1.5 dB) and will withstand a +13 dBm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FLS-570
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FBS-50
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FHL-795
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FLL-100
SMC
FBS-140
SMC
FLL-570
SMC
FHL-45
SMC
FLS-100
SMC
FIL-70
SMC
FIS-140

input signal for all output load conditions. The 3100LA's dimensions are 8.75 x 17 x 17", and it weighs 60 lbs. List price is $7885.
ENI
INFO/CARD #233

100 W Benchtop Amp

The model 100W000M1 delivers a minimum of 100 watts of cw power through a frequency range of 80 MHz to 1000 MHz. Minimum linear output measured at less than 1 dB gain compression is 60 watts through the same bandwidth. This amplifiers bandwidth precisely matches the 80-1000 MHz bandwidth of the AR model AT1080 log-periodic antenna. Price is $25,000
Amplifier Research
INFO/CARD #232

5W, VHF Power Module

Motorola has introduced a new RF power module designed for portable radio applications operating in the VHF frequency band of 68 to 88 MHz. The MHW105 RF power module operates from 7.5 V and requires only 1 mW of RF input power to deliver 5 W of output power. Epoxy glass PCB construction gives this device consistent performance and reliability.
Motorola Semiconductor
INFO/CARD #231

1 to 1000 MHz Amplifier

Power Systems Technology announces the development and availability of model AR1619-50/3040, a broadband solid state amplifier covering the frequency range of 1 to 1000 MHz in two bands. Each band has independent input and output connectors with power output of 50 W for the 1 to 500 MHz band, and 25 W for the the 400 to 1000 MHz band. The amplifier has a built-in power supply and measures 5.25 x 19 x 22 inches.
Power Systems Technology Inc.
INFO/CARD #230

SIGNAL SOURCES

L-band Synthesizer

The VDS-6000 L-band synthesizer is suited for INMARSAT-M, SCADA applications, and as a local oscillator in C and Ku-band VSAT systems. Its single loop design exploits Sciteq's patented Arithmetically Locked Loop (ALL). The ALL combines digital signal processing with fractional-n to preserve 20log(n), but reducing n to improve phase noise for a given step size. The VDS-6000 dissipates 4 W. Price of single units is $1395.
Sciteq
INFO/CARD #229

Low Profile Ovenized Oscillator

Piezo Crystal announces the availability of model 2920155. This oscillator utilizes Piezo's SC cut crystals and is available in the frequency range of 80 to 110 MHz.Typical phase noise is -150 dBc at 1 kHz offset, and aging rate is 1x10-8/day at the time of shipment. The size is 3.00 x 3.00

x 0.50 inches. Approximate price for the model 2930200 is $320 to $350 in quantities of 500.
Piezo Crystal Co.
INFO/CARD #228

SAW-based VCO

AT&T Microelectronics has introduced a surface acoustic wave (SAW) based, voltage controlled oscillator specially designed for applications ranging in frequency from 155 MHz to 1.1 GHz. Output is at 10k ECL levels. The VCO600 comes in a minia-

ture 28-lead ceramic surface-mount package measuring only 0.7 x 0.3 inches. In addition, it offers an output disable and clock-through feature, which improves on-board testing.
AT&T Microelectronics
INFO/CARD #227

VCOs for Wireless

Analop Engineering is pleased to announce two low cost, high performance VCOs designed particularly for the wireless applications in the 902 to 926 MHz and the 2400 to 2500 MHz bands. These 5 V devices consume only 8 mA and deliver 0 dBm. Phase noise at 100 kHz is 110 dBc/Hz typ.
Analop Engineering
INFO/CARD #226

SUBSYSTEMS

Limiter-Mixer Preamps

Alpha Industries announces its MX1011 and MX1010 miniature broadband limiter-mixer preamps covering 0.5 to 2 GHz and 2 to 18 GHz EW bands, respectively. A mere 0.375 cubic inches in volume, these mixers integrate a PIN diode limiter (50 W CW), a double balanced mixer and a low-noise/high-power (+15 dBm) IF preamp in a hermetic package with field replaceable connectors.
Alpha Industries, Components and Subsystems Div.
INFO/CARD #225

CDPD Modem

Cincinnati Microwave has announced the first in a family of wireless modems for the Cellular Digital Packet Data (CDPD) network. The MC-DART 100 has a 19.2 kbps data rate, up to three watts of RF power output, received signal strength indicator and optional switched antenna diversity. The MC-DART 100 is priced at $495, with volume quantities available in 1Q94.
Cincinnati Microwave, Inc.
INFO/CARD #224

Modem for CDPD

Motorola's Mobile CDPD™ has a 9-pin RS-232 serial interface which supports speeds from 300 to 19,200 bps. The airlink data rate is 19,200 bps. The new modem will operate in full duplex mode and use Motorola's 3 W cellular telephone platform.
Motorola Cellular Subscriber Group
INFO/CARD #223

CABLES & CONNECTORS

3/8 Inch, Flexible Cable

Andrew Corp. announces the introduction of three new 3/8 inch superflexible 50 ohm HELIAX® coaxial cables. Type FSJ2-50 is similar to Andrew's existing 1/4 inch (FSJ1-50A) and 1/2 inch (FSJ4-50B) superflexible cables. Types ETS2-50T and ETS2-50 are for higher power or plenum applications. Type ETS2-50T is jacketed; Type ETS2-50 is unjacketed.
Andrew Corporation
INFO/CARD #222

MCX Connector

ITT Cannon/Sealectro announces availability of its line of MCX RF connectors. MCX connectors provide a smaller, lighter alternative to an SMB connector when space and weight savings are a consideration.Snap-on mating allows for rapid connects and disconnects. The MCX performs well in 50 ohm applications up to 6 GHz. Durability is specified as a minimum of 500 matings.
ITT Cannon/Sealectro
INFO/CARD #221

2.92 mm Connectors

A family of 2.92 mm coaxial connectors from Southwest Microwave are flange mounted and use field-replaceable one-piece construction. The connectors launch to 0.012 diameter pins. Maximum VSWR is 1.15:1 over the frequency range of DC to 40.0 GHz.
Southwest Microwave, Inc.
INFO/CARD #220

SIGNAL PROCESSING COMPONENTS

Wideband Quadrature Hybrid

Synergy Microwave Corp. announces a new, miniature wideband quadrature hybrid. The DQP-3-33 covers an 11:1 bandwidth between 3 and 33 MHz. Across the unit's 30 MHz bandwidth insertion loss is typically 0.5 dB and amplitude imbalance is typically 0.7 dB. Phase unbalance is 90 ±4 degrees. Minimum isolation is 20 dB and VSWR is 1.5:1 on all ports. These hybrids are available from stock and are packaged in 8-pin relay headers. Prices start at $69.95 in quantities of 1 to 9 pieces.
Synergy Microwave Corp.
INFO/CARD #219

Mixer

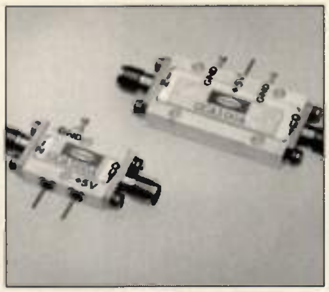
The MA27LLX from Western Microwave is a triple balanced, multi-octave mixer. The input (RF) is 2 to 20 GHz and output (IF) is 1 to 10 GHz. Isolation is typically 23 dB, conversion loss is typically 6.5 dB, and third order intercept is typically +17 dBm. Required LO drive is +10 dBm. Price is $161 in quantities between 10 and 24.
Western Microwave, Inc.
INFO/CARD #218

Delay Lines

A new generation of fixed and variable delay lines have been designed for use in high speed computers, but also have application in high speed analog circuits. Delay times are available from 0 to 30 ns. Minimum 3 dB bandwidth, depending on delay time, is as high as 1 GHz.
Toko America, Inc.
INFO/CARD #217

Comb Generators

A line of harmonic/comb generators from Herotek have integral preamplifiers. Wideband outputs go up to 26 GHz when tuned to input frequencies between 30 Mhz to 2.0 GHz. Features include optional input powers of 0 dBm or +10 dBm with outputs as high as −10 dBm at 18 GHz and −20 dBm at 26 GHz, 5 V DC supply, and removable connectors for drop-in assembly. Hermetic packaging as small as 0.75 x 0.56 inches is available.



Herotek, Inc.
INFO/CARD #216

SP8T Switch

The GL8T RF switch is part of a series of low cost switches offered in SPST through ST8T versions. The series operates over the 25 to 250 MHz range with maximum VSWR of 1.5:1. Prices range from $45 (SPST) to $225 (SP8T)
Geoffroy Labs
INFO/CARD #215

Frequency Multipliers

KW Microwave has introduced x4 multipliers to its multiplier product line. The x4 multipliers have input frequency fin = 714 MHz and 2.16 GHz, with output frequency of fout = 2.856 GHz and 8.64 GHz, respectively. Input power is 0 dBm while output power is +5 dBm with −60 dBc spectral purity. Typical size is 3 x 0.4 x 1 inches.
KW Microwave Corp.
INFO/CARD #214

TOOLS, MATERIALS & MANUFACTURING

RF Test Contacts

Johnstech Intl. has introduced high performance RF contactors for plunge-to-board style gravity feed test handlers. Johnstech's Short Contact™ test sockets are based on patented technology that combines outstanding electrical performance with superior contact reliability in automated and manual testing of RF and microwave devices. Each contact is individually replaceable.
Johnstech International Corp.
INFO/CARD #213

Benchtop Coax Stripper

The CTEL 1066 from Coastel Cable Tools is an A/C powered benchtop coax stripper. The CTEL 1066 has the ability to perform all three strips simultaneously in four seconds. The foot switch operated stripper has built-in automatic brake stop and easily replaceable cutter heads.
Coastel Cable Tools Inc.
INFO/CARD #212

Using RF Channel Simulators to Test New Wireless Designs

By Karl B. Fielhauer
Litton Amecom

Wireless digital communications systems are experiencing spectacular commercial growth. Digital communications using RF is among the most difficult environments to design for reliable performance. Yet, customers will demand 100% reliability from their new wireless digital communications products, despite RF propagation conditions. This article presents the propagation effects encountered by these systems that are replicated by RF channel simulators in the laboratory.

With the implementation of systems like: the North American Dual-Mode Cellular System (NADC), Japanese Digital Cellular (JDC) telephone system, Personal Communications Network (PCN), Global System for Mobile Communications (GSM), Digital European Cordless Telephone (DECT), Advanced Cordless Telephones and High Definition Television (HDTV), the need to realistically test these new products is tremendous. RF channel simulators aid in the design of new exotic products and can statistically test how robust a new design is compared to a baseline, theoretical or even a competitor's performance. Presently, the biggest growth in the wireless industry is mobile communications, with cellular leading the way and other technologies to follow soon.

Present cellular systems use the Advanced Mobile Phone System (AMPS) that has been around since the early 1980's. However, AMPS is an analog system requiring voice bandwidth. Because of an overcrowded spectrum due to limited frequency allocation, an AMPS user may not be able to use his cellular phone during peak hours, or may experience a dropped call. This is especially prominent in large metropolitan areas such as New York and Washington, DC. To get around this problem, the FCC approved the NADC system, which increased the number of phone calls by reducing the information bandwidth by digitally encoding the channel, namely $\pi/4$ DQPSK. This system can increase the numbers of users over the present AMPS by a factor of three, with more sophisticated voice CODECs, up to six — and with Extended TDMA, up to 15 times. However, along with the benefits of increased capacity due to digital encoding, comes the difficulties of demodulation in the presence of fading, multipath propagation, co-channel interference, adjacent channel interference and of course Additive White Gaussian Noise (AWGN).

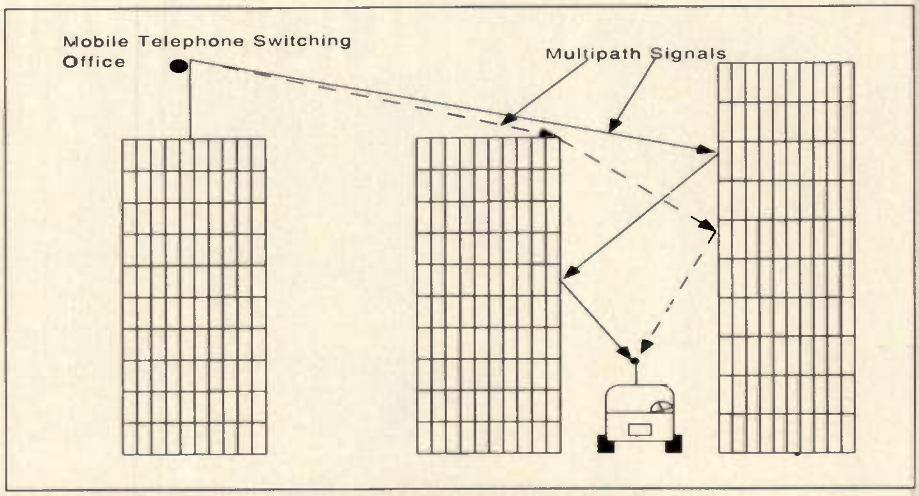


Figure 1. Multipath propagation of a cellular signal.

What is a RF Channel Simulator?

The characteristics of mobile radio channels, although complex in nature, can often be adequately represented by known statistical distributions. RF channel simulation, therefore, amounts to producing in the laboratory, signals that have appropriate statistical properties.

Many companies manufacture RF channel simulators for different mobile and stationary systems from HF into VHF/UHF. RF channel simulators now replicate the conditions of microwave channels conditions under rain fade and atmospheric multipath conditions. The simulators can produce ghost signals with which to test ghost cancellation designs for the emerging HDTV market. They are also available to test satellite aided mobile applications. Finally, RF channels simulators are now available to simulate mobile cellular conditions with multipath fading, Doppler, delay spread and interference. Each RF channel simulator for each RF application has its own statistical distribution which best represents the channels propagation characteristics.

The RF channel simulator is a versatile tool to improve the quality of digital (and also analog) wireless designs before a new product is sent to market. In addition the simulator can be used in the production phase to measure and identify performance variations from unit to unit. By running different propagation simulations on the production line, the radio can be tested for Automatic Gain Control (AGC) performance, error correction performance, etc., in a real environment. This can quickly be accomplished using Automatic Test Equipment (ATE) and will insure quality before the radio is sent to market.

This article focuses on RF channel simulators used for mobile communications and the effects that will be encountered in that environment, especially

ZRON

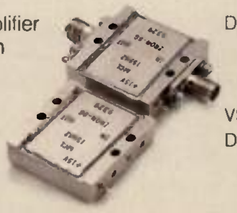
2-8GHz AMPLIFIERS

Up to 100mW, 20dB Gain, 2-8GHz, only 495^{00}

Mini-Circuits fused the science of economy and technology together to create the very broad band, 2 - 8GHz ZRON-8G amplifier. Each SMA connector on this amplifier can be easily removed for quick conversion to drop-in assembly capabilities...and with a price tag of just $495, the ZRON-8G will fit as well into your budget as it will your project. But, don't let the small price fool you. Each hermetically sealed unit is built extremely tough. Even if the power supply voltage is shorted under full RF power, the ZRON-8G amplifier will survive without damage.

Suitable for both lab and systems applications, this multi-functional GaAs amplifier delivers 20dBm minimum power output at 1dB compression and has a linear gain of at least 20dB over the full frequency range. To that, add a VSWR of 2.0:1 maximum over the whole band and you have a unit that's unconditionally stable for all load and source impedance magnitude and phase.

So, call your Mini-Circuits distribution center for guaranteed one week shipment and seize the competitive power packed in the new ZRON-8G amplifier!



Available with or without heat sink

INFO/CARD 39

SPECIFICATIONS Model: ZRON-8G

Freq. Range GHz	2-8
Gain (dB) Min.	20
Power (dBm)	
Min. Output (at 1dB Comp.)	+20
Dynamic Range	
NF(dB) Typ.	6
Intercept Point (dBm) 3rd Order Typ.	30
VSWR In/Out	2.0:1
DC Power	
Voltage	+15
Current mA Max.	310

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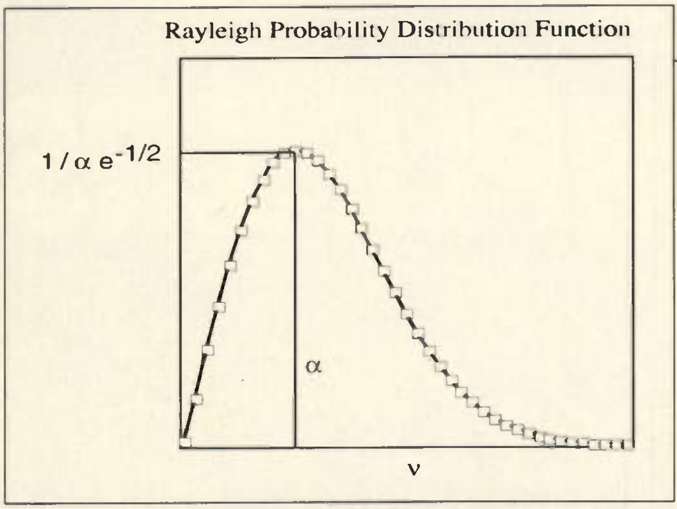

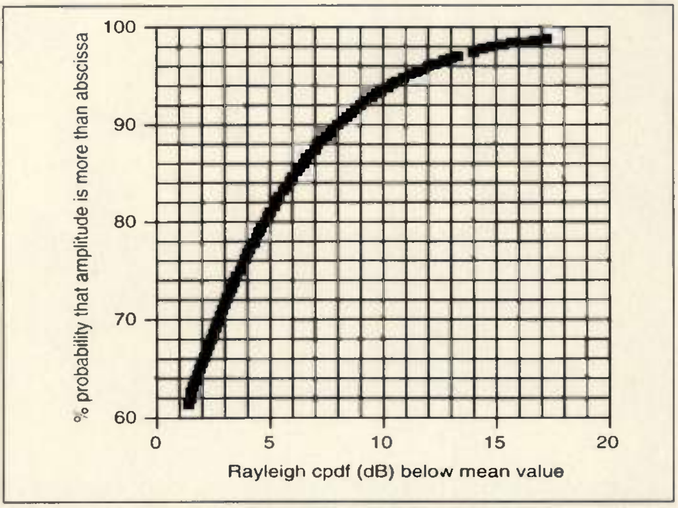

Figure 2. Rayleigh Probability Distribution Function.

Figure 3. Rayleigh Cumulative Distribution Function.

with the implementation of the NADC, JDC, DECT and GSM systems. Understanding how an RF channel simulator works for mobile applications will enable a basic understanding of the other RF environments in which these simulators have applications.

Why Use a Simulator?

Before the advent of RF channel simulators, field testing was the only way to test new wireless designs. However, using conventional field testing requires time to setup and take down, collect the data, and, worst of all, it is not repeatable. With the use of RF channel simulators this is no longer true. The use of simulators enables design engineers to test their product under favorable and poor conditions in the laboratory and compare their results to previous designs. The outcome of test results are rapid, allowing a testing team to communicate the results quickly to a design team for changes and revisions in DSP algorithms before ASIC's are sent to be fabricated and before diversity combining techniques are agreed upon. In addition coding and modulation schemes and the evaluations of equalizations can be tested using a RF channel simulator.

With a simulator it is possible to write a test plan incorporating such important parameters as fade margin, E_b/N_o, type of fading to be tested, Doppler, the addition of adjacent and co-channel interference and diversity combining techniques. With field testing, the environment where the test is conducted cannot be controlled, therefore propagation effects cannot be tested individually. The effects of adjacent channel, co-channel interface, intersymbol interference and other effects can add upon each other, thus the isolation of each effect upon the radio is not possible. With the use of a RF channel simulator, the bit error rate performance in fading and interference can be compared to theoretical performance when the propagation effects are isolated thus establishing a baseline of performance. The same test procedure can be rerun to measure design improvements after design changes are implemented. A final benefit is that the data acquisition and test equipment control (including the RF channel simulator) can be, and usually is, fully automated to produce reliable and repeatable results quickly.

What Does an RF Channel Simulator Do?

The mobile radio will have to work in the harshest of RF environments. The major problem is that operation often is in built-up areas, where the mobile antenna is well below or located between surrounding buildings, and there is no line-of-sight path to the transmitter (Figure 1). Propagation is, therefore, mainly by way of scattering from the surfaces of the buildings and by diffraction over and/or around them. The most pronounced effect in mobile cellular communications is that of multipath signal propagation caused by the scattering signals. The multipath reflections grow more likely as the frequency of transmission increases. The three major effects of multipath propagation are multipath fading (Rayleigh fading), delay spread and Doppler-frequency shifts.

The RF channel simulators for mobile applications model multipath propagation caused by the antenna being well below the transmitter (Mobile Telephone Switching Office, or MTSO). In addition, it is possible to incorporate Additive White Gaussian Noise (AWGN), adjacent channel and co-channel interference to the RF channel simulator and test in the presence of fading. In the new markets of digital cellular radio, understanding and designing systems which work reliably in multipath propagation conditions will be critical to winning the market.

Multipath fading occurs when an RF carrier is reflected over more than one path producing multiple signals with different arrival times (see Figure 1). The signals produce apparent phase differences between the multiple reflections and the original, or the shortest path. Thus for a constant amplitude CW signal, if a multipath reflection is 180° out of phase from the shortest path and equal amplitude,a total cancellation will occur. In some cases, there is constructive addition when two or more signals arrive in phase with one another. However, in the real world the multipath reflections have random arrival times and random phase shift between multipath signals, the fading is then called Rayleigh. This is because the fading is best describe by the Rayleigh probability density function (PDF). A Rayleigh distribution is the root-mean-square (RMS) addition of two independent Gaussian or normal distributions. The Rayleigh PDF is shown in Figure 2 and is defined as:

$$p(\nu) = \frac{\nu}{\alpha^2} \exp\left(-\frac{\nu^2}{2\alpha^2}\right), \quad (0 \le \nu \le \infty) \quad (1)$$

$$0, (\nu < 0)$$

where $v^2 = x^2_1 + x^2_2$, two independent variants, x_1 and x_2 and Gaussian-distributed both having zero mean and the same standard deviation, and:

$2\alpha^2$ = the mean square value of v over the distribution.

The corresponding Cumulative Probability Distribution Function (CPDF)

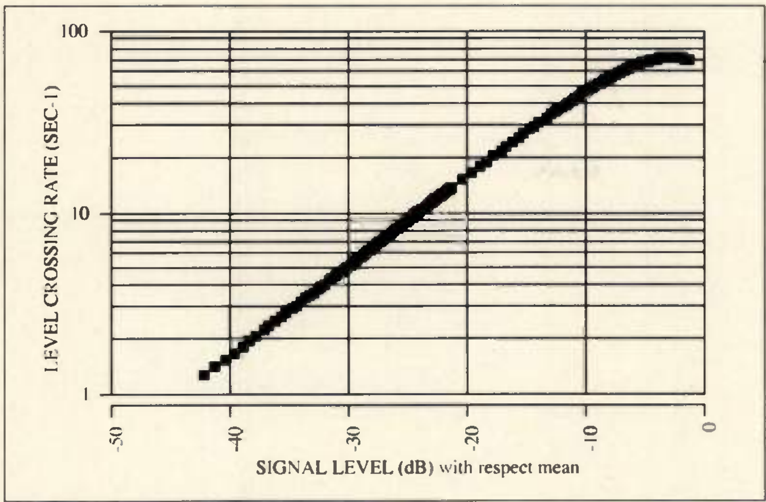

Figure 4. Level crossing rate.

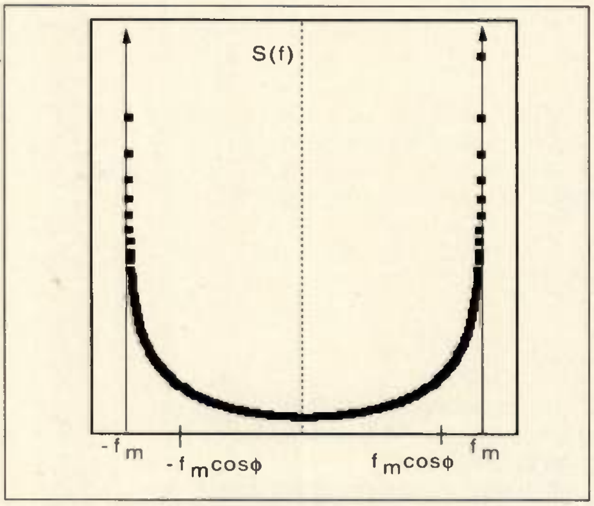

Figure 5. Rayleigh power spectra.

obtained by integrating p(v). The CPDF represents the probability that v is less than level A for an E-Field signal and defined by:

$$P(\nu \le A) = \int_0^A \frac{\nu}{\alpha^2} \exp\left(-\frac{\nu^2}{2\alpha^2}\right) \qquad (2)$$
$$= 1 - \exp\left(\frac{A^2}{2\alpha^2}\right)$$
$$0, (\nu < 0)$$

Plotting the Rayleigh CPDF on a logarithmic scale as shown in Figure 3, the received radio signal has smaller attenuation (less severe fade) than 10 dB from its mean value during 94% of the time and a less severe fade than 18 dB during 99% of the time. When evaluating the performance of an RF channel simulator capable of Rayleigh fading, the measured CPDF and the theoretical CPDF should match thus producing true Rayleigh characteristics for your simulation which is critical in mobile cellular applications. Rayleigh fading is relatively short term in duration yet can yield the deepest fades which are a significant impairment to all digital communications. In the NADC system, a π/4 DQPSK transmitted signal with constant amplitude in the presence of Rayleigh fading is converted into one with randomly varying amplitude and thus is a large problem for a demodulator to recover data error free if the fade depth exceeds the fade margin. Rayleigh fading only exists when there is no line-of-sight to the receiver only multiple reflections. Rayleigh fading can be dealt with by adding error correcting codes, interleaving and diversity combining techniques. Since most RF channel simulators are dual channel, therefore diversity improvement factors are measurable.

For a satellite aided mobile link and some cellular links the received envelope obeys Rician statistics rather than Rayleigh and some RF channel simulators include both. Rician fading occurs where there may be a line-of-sight path, or a least a dominant specular component to the mobile receiver. It is intuitively to be expected that there will be fewer deep fades and that the specular component will be a major feature of the spectrum.

A mobile receiver does not experience Rayleigh fading until the mobile is moving through a multipath environment. A measure of the number of times per second that a mobile will encounter a multipath fade is represented by the Level Crossing Rate (LCR), an important consideration to the designer of a mobile digital communications systems. The LCR is the number of times per second (Hertz) the carrier envelope crosses a level relative to the mean. A plot of this is shown in Figure 4 for a mobile receiver moving at a rate of 100 km/hour at a frequency of 900 MHz. The expected number of level crossing at a given signal amplitude v = A is:

$$n\left(\frac{\nu}{\sqrt{2\alpha^2}} = R\right) = \frac{\beta V}{\sqrt{2\pi}} R \exp\left(-R^2\right) \qquad (3)$$

where:

$$R = \frac{A}{\sqrt{2\alpha^2}}$$

$\beta = 2\pi/\lambda$

V = Speed of the vehicle (m/s)

Careful attention should be paid both to the theoretical LCR and theoretical Rayleigh PDF to the measured performance when evaluating RF channel simulators for true Rayleigh fading. These statistics should be part of the specifications for the mobile RF channel simulator.

Another propagation effect that is modeled by an RF channel simulator is Doppler frequency shift in a multipath environment. The Doppler frequency f is:

$$f = \frac{V}{\lambda} \cos\phi = f_m \cos\phi, \quad 0 < \phi < \pi \qquad (4)$$

where f_m is the maximum Doppler frequency. It is clear that in any particular case, the change in path length will depend on the spatial angle between the wave and the direction of motion. Generally, waves arriving from ahead of the mobile have a positive Doppler shift, i.e., an increase in frequency, while the reverse in the case for waves arriving from behind the mobile. Waves arriving from directly ahead of, or directly behind the vehicle are subjected to the maximum rate of change of phase. If the mobile unit is moving through an area of multiple reflections and it is assumed that the angles of arrival are evenly distributed over 360°, then the Doppler shift will be on each of the multipath scattering signals from the MTSO. The multipath scattering signals will be arriving to the moving mobile with positive shifts and with negative shifts causing a spread in the received power spectra. If there are a large number of scattering signals, as in the case of mobile cellular phone, then the receive spectrum will be widened by the multiple Doppler shifts and is referred to as Doppler spread. The limits of the Doppler spectrum can be quite high, for example in a vehicle moving at 30 m/s (approx. 70 mph) receiving a signal at 900 MHz the maximum Doppler shift is 90 Hz and is shown in Figure 5. Frequency shifts of this magnitude can cause interference with the message information. The theoretical power spectral density S(f) for a mobile in multipath environment is :

$$S(f) = \frac{1}{f_m \sqrt{1-\left(\frac{f}{f_m}\right)^2}} \qquad (5)$$

This spreading of the signal energy is an important characteristic of Rayleigh channel simulator.

Another propagation event affecting largely the mobile receiver is delay spread. Delay spread occurs when the base station transmits an impulse signal to the mobile unit and because multipath scattering the received impulse signal is significantly lengthened. In a high data rate mobile communications environment, a bit time can be significantly lengthened thus overlapping onto another bit period which is the cause for intersymbol interference.

Long term fading such as that caused by an obstruction or terrain absorption are characterized using the log-normal probability distribution function and is found on most RF channel simulators. Long term fading is also referred to as shadow fading because the mobile moves through the shadow of an obstruction which can take a considerable amount of time. Long term fading is usually compensated by Automatic Gain Control (AGC) circuits. Of the three statistics (Rayleigh, log normal and Rician) that are represented in RF channel simulators, Rayleigh fading with Doppler is the most difficult environment for a mobile receiver to operate.

Implementation of an RF Channel Design in Hardware

Figure 6 represents a basic design of an RF channels simulator for mobile communications receivers which simulates Rayleigh fading. This is a simple hardware configuration for a Rayleigh multipath fading simulator that consists of two Gaussian noise generators, two variable low pass filters and two balanced mixers. The cut-off frequency of the two low pass filters is selected on the basis of the frequency and the assumed average speed of the mobile vehicle. The output of the simulator represents the envelope and phase of a Rayleigh-fading signal. Tests on such simulators show the output signal envelope to be a close approximation to a Rayleigh distribution and the level-

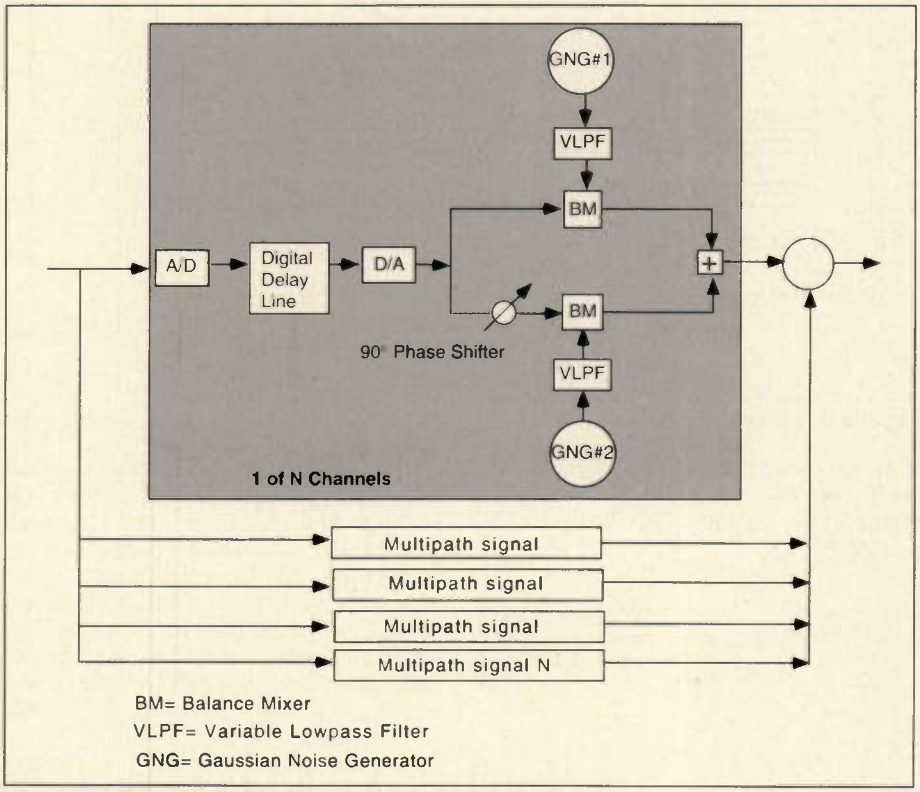

Figure 6. Rayleigh RF channel simulator hardware.

crossing rates agree well with theoretical predictions.

This hardware version is duplicated several times depending on the number of multipath signals that are to be simulated for the environment under consideration. The simulator uses digital delay lines to replicate the multipath delay on each of the multipath signals. The several delayed output signal from the Rayleigh-fading simulators are summed together to form a multipath signals. The NADC systems requires receivers to operate with at least two Rayleigh-faded signals spaced by a delay of up to 42 µs. These standards have been implemented to guarantee functionality of the various mobile radio manufactures equipment. Usually the RF channel simulator replicates this design to form two separate RF channels in which the fading ccrrelation can be controlled between separate RF channels. This aids investigations of diversity reception techniques and the effects of co-channel and adjacent-channel interference.

Conclusion

The RF channel simulator is a valuable new addition of test equipment for the evaluation of digital modulation and demodulation designs. The RF channel simulator can be used through the initial design stages through quality control on the production line and can simulate mobile as well as stationary systems. The RF channel simulator can be used to verify design improvements without field testing and performance variations, if any from unit to unit in the production process. *RF*

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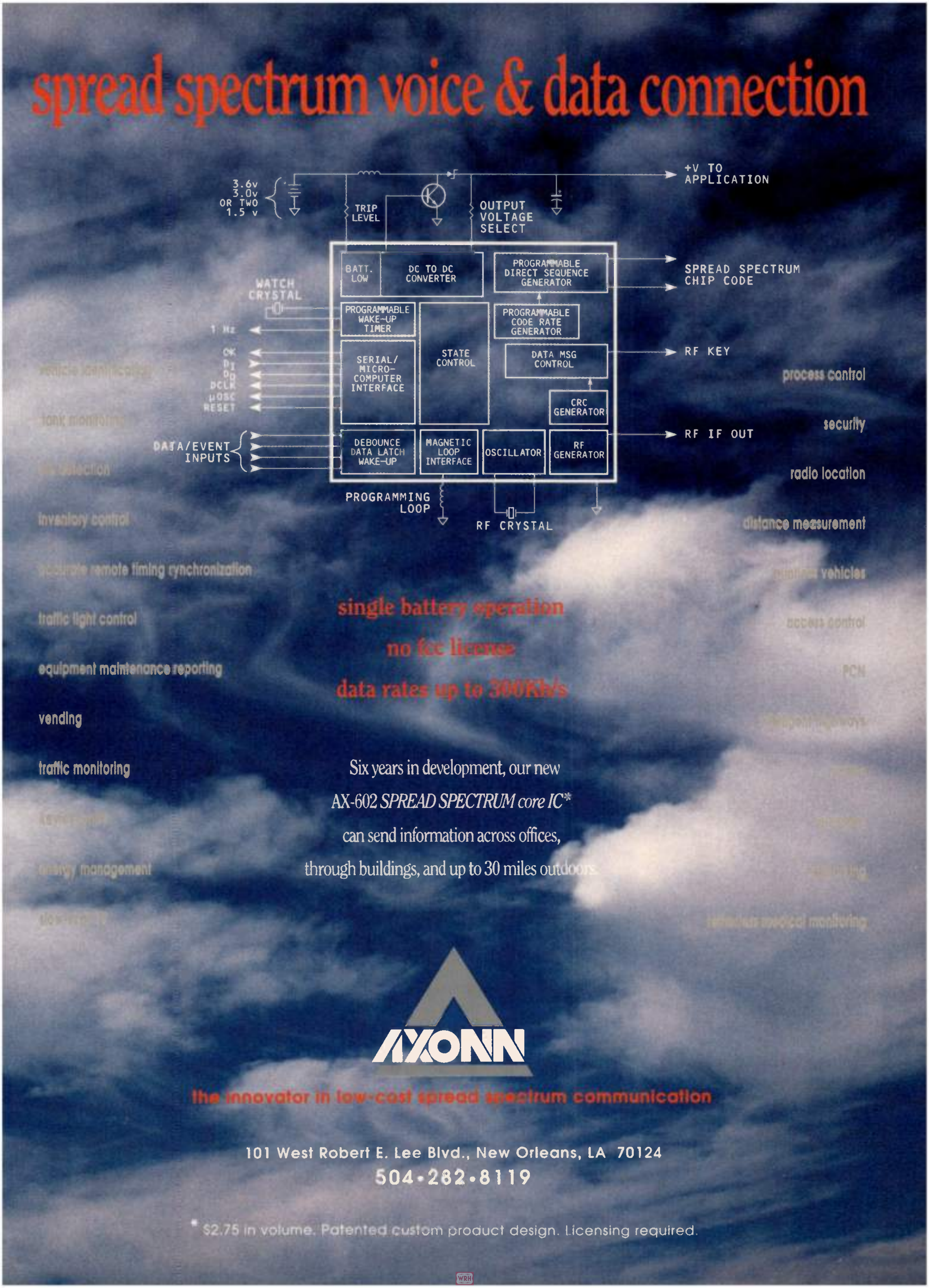
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Programs Help Design Vacuum Tube Amplifier Output Circuits

By Thomas F. Crawford
Electronic Communications and Service

High power amplifiers still commonly use vacuum tubes, despite the growth of solid state technology. Power triodes and tetrodes remain economical choices, especially for power levels above 1000 watts in the HF, VHF and UHF ranges. The author's entry in the 1993 RF Design Awards Contest was a program which computes the component values for tuned output circuits in Pi, L and Pi-L configurations. This article describes the Pi-L network program.

The Pi-L network program was originally written to assist amateur radio operators in the design and construction of high power amplifiers operating in Class AB1, AB2, B or C. It is intended to guide the user through the design process, allowing either the user or the program (actually, the program's author) to determine various design criteria, depending on the user's level of experience and ability. The program begins with several information screens that may be bypassed after they have been read. The equivalent schematic of the network is shown in Figure 1.

Before entering data for calculation, the program addresses plate resistance. The user may enter a calculated resistance value, or the program can calculate an approximate value based on plate voltage and current, using the following equations:

$$e_{RMS} = \frac{e_{PEAK}}{\sqrt{2}} \qquad (1)$$

$$i_{RMS} = \frac{I_{DC} \cdot 1.57}{\sqrt{2}} \qquad (2)$$

$$R_L = \frac{e_{PEAK}}{I_{DC} \cdot 1.57} \qquad (3)$$

In practice, the voltage swing, ePEAK, will be less than the DC voltage applied, depending on the tube characteristics, loading and drive. With 3000 volts applied, the voltage swing might be

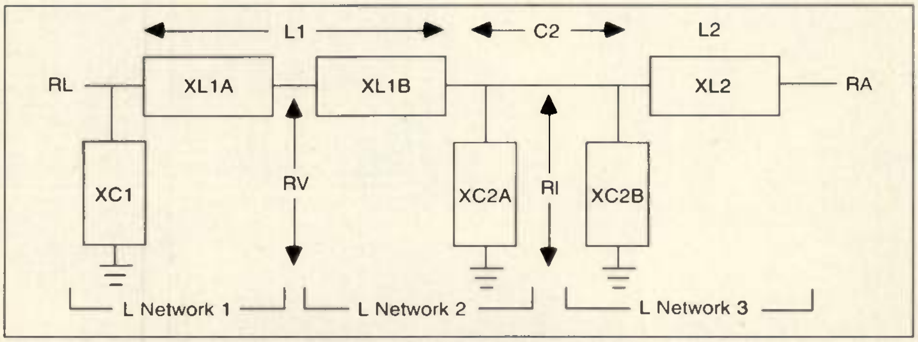

Figure 1. Equivalent schematic of the PI-L output network.

2500 volts. There is no way for an amateur to measure e_{PEAK}. The peak fundamental component of current is one-half the peak current, or $I_{DC} \times \pi/2$:

In the program, the $\pi/2$ current multiplication factor ($\pi/2 = 1.57$) is replaced with 1.8. This is the same as declaring e_{PEAK} to be 0.87 times the full DC voltage. This estimation was determined empirically over many years of evaluation of electron tubes, and should be a good approximation for most applications.

$$R_P = \frac{E_P}{I_P \cdot 1.8} \qquad (4)$$

Next comes selection of the operating Q, which requires a tradeoff between harmonic suppression and critical tuning. Typically, Q values from 10-30 offer acceptable harmonic rejection, but higher Q results in a sharper peak at resonance and very "touchy" tuning characteristics. Select the desired Q for the highest frequency that the network will cover, and since Q increases at lower frequencies, harmonic suppression will not be compromised.

The most difficult choice in the design is the selection of image resistance (RI). Since the program designs the Pi-L network as a cascade of three L-networks, an impedance for the junctions of the networks is required. This is somewhat arbitrary, but limited by the peak voltage

```
——————————— ENTER PROGRAM DATA ———————————

Enter the selected PLATE VOLTAGE. (100-.5KV)        [EP] = [3000] Volts
Enter the PLATE CURRENT in AMPERS. (.1 to 2.5)      [IP] = [.75 ] Amps
The PLATE LOAD RESISTANCE is ...                    [RP] = 2222.222 Ohms
The PLATE POWER INPUT is ...                        [PIN] = 2250 Watts
Desired Plate Tank Oerating Q (1 to 30)             [Q1] = [15 ]

Select a trial value of IMAGE RESISTANCE
(Values between 150 & 1500 Ohms)          [RI] = [750 ] Ohms

Remember —        RA < RI
Enter the Ant. or LOAD RESISTANCE (10 to 800)  [RA] = [50 ] Ohms

Would you like to change any one VALUE before continuing (Y or N) [ ]
```

Figure 2. Main program screen

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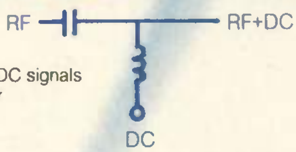

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across the loading capacitor, where:

$$E_P = \sqrt{Z_P \cdot P_o \cdot 2.414} \qquad (5)$$

Design constraints require RI > RA, so values in the range of 150-1500 ohms are practical.

The program allows the user to examine the results for various values of Q and RI. Selecting analysis frequencies close together shows how critical the tuning is likely to be. Observing the component values shows what parameters will result in a realizable design.

Program Operation

After the initial information screens, the main input screen (Figure 2) is reached. Limits have been established for each entry, based on values deemed within a practical range. Since it was written for amateur applications, the program halts and asks for re-entry of data when the voltage and current selected are likely to result in power output exceeding the maximum permissible power of 1500 watts. [Modification of the source code to allow higher power calculations for commercial applications should be straightforward — editor].

Plate tuning capacitor reactance:
XC1 = RP/Q1 (Q1 is selected tank Q)
Virtual Resistance:
RV = EP/(Q2 + 1)
First half of tank coil:
XL1A = R1/Q2
First half of the loading capacitor:
XC2A = RI/Q2
Second half of tank coil:
XL1B = RV x Q2
Third network figure of merit:
Q3 = SQR((RI/RA)–1)
Second half of loading capacitor:
XC2B = RI/Q3
Loading coil:
XL2 = RA x Q3
Total reactance of L1 is the sum of:
XL1A + XL1B.
Total reactance of the loading capacitor is the algebraic sum:
XC2 = (XC1A)(XC1B)/(XC1A + XC1B)
Component values are calculated according to normal reactance formulas.

Figure 3. Program calculations.

Frequency entry can be anywhere from 1-500 MHz, but the results for very high or very low frequencies can be very interesting! See the note above on practical component values. The program output includes both reactances of the network components and the inductance and capacitance values. Operation of the entire program is easily accomplished following the on-screen instructions. When all data have been entered, the computation proceeds as shown in Figure 3.

At the end of the calculations, the program gives the choices for printout of the data, calculation of more frequencies, check Q at the low frequency end of the range, do a new calculation, or exit.

The program was written in GWBASIC and compiled in Quick Basic 4.5. Both an executable file and source code are included. This program, along with additional programs for Pi-networks, L-networks and a coil-winding utility, is available from the RF Design Software Service. See page 75 for ordering information. *RF*

About the Author

Thomas Crawford is a retired engineer whose career included the Coast Guard, Air Force, and naval shipyards. He is part of a small company of retirees, Electronics Communications and Service, which designs and builds custom receiver front ends. He can be reached at W490 Bulb Farm Road, Shelton, WA 98584.

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	240L	A	20kHz -10MHz	40W	50dB (±1.5dB)
	2100L	A	10kHz -12MHz	100W	50dB (±1.5dB)
	HF				
	A-150	A	300kHz -35MHz	150W	55dB (±1dB)
	A-300	A	300kHz -35MHz	300W	55dB (±1dB)
	A-500	A	300kHz -35MHz	500W	60dB (±1dB)
	A-1000	A	300kHz -35MHz	1000W	60dB (±1.5dB)
	VHF				
	325LA	A	250kHz -150MHz	25W	50dB (±1.5dB)
	3100LA	A	250kHz -150MHz	100W	55dB (±1.5dB)
	3200L	A	250kHz -120MHz	200W	55dB (±1.5dB)
	UHF				
	400AP	A	150kHz -300MHz	3W	37dB (±1dB)
	403LA	A	150kHz -300MHz	3W	37dB (±1dB)
	411LA	A	150kHz -300MHz	10W	40dB (±1.5dB)
	525LA	A	1MHz -500MHz	25W	50dB (±1.5dB)
	550L	A	1.5MHz-400MHz	50W	50dB (±1.5dB)
	5100L	A	1.5MHz-400MHz	100W	50dB (±1.5dB)
	603L	A	800kHz -1GHz	3W	40dB (±1.5dB)
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Double Feedback Circuit Analysis

By Frank Egenstafer
American Meter Co.

Double feedback circuits are those circuits incorporating both series and shunt feedback paths. The double feedback topology is a byproduct of many amplifier designs, but the interactions among the two paths and the active device have not been adequately characterized in the past. It is the purpose of this paper to afford the engineer an understanding of how the double feedback stage operates and to supply him with some simple equations that will aid him in making circuit decisions.

To design an amplifier of the type used in the CATV field it was previously necessary to initiate that design without the use of engineering design equations, mainly because there were none. In most cases, the design was pursued by varying the emitter resistance (R_e), shunt feed back resistance (R_f) and the transformer turns ratio until satisfactory results were achieved (Figure 3). This procedure was followed first on a single stage, then on two stages together, then on three, etc. Mostly, by the time three stages were completed, the interaction had become complex and frustrating.

The designer was plagued by all sorts of amplifier behavior, or rather misbehavior, that escaped explanation and reason. It was no simple task to design an amplifier with sufficient gain, wide frequency response, exceptional flatness, good input match, good output match, low noise figure, high output capability, linear distortion characteristics; especially since when all was done, one of the elements had to be made variable (tilt control), which tended to degrade many of the above characteristics.

Throughout this article, the terms β and h_{fe} are used interchangeably for the high-frequency current gain of a transistor, and the constant $R_f \cdot R_e = R_L^2$ is often used because this equality puts order into the analysis. It permits the elimination of either R_f or R_e from a complex equation and hence the solving of a single quantity in terms of another, and it eliminates an infinite number of value relationships that would exist between R_f and R_e.

Basic Feedback Circuits

Take a transistor, put a resistor in the emitter circuit and a resistor from collector to base and you have a feedback ampifier. It sounds very simple but is, in essence, a very complex circuit, as will soon be shown. Analysis through the use of simple equations, logic, and confirming experimentation indicates that the circuit obeys the family of laws described in the equations presented here. For this analysis, the transistor is considered to be a black box, and β is the only parameter which varies with frequency. The close agreement between theory and practice validates this assumption.

First let us examine, separately, shunt and series feedback and see how each affects the action of a simple transistor stage. Shunt voltage feedback tends to lower both the input and output impedance of a circuit and to stabilize the current gain under control of R_f and R_L (Figure 1). Series current feedback tends to increase both the input and output impedance and to stabilize the voltage gain under control of R_L and R_e (Figure 2).

If we combine these two circuits we have a circuit that is both current and voltage gain stabilized under control of external resistances. If the gains are equal,

$$\frac{R_f}{R_L} = \frac{R_L}{R_e} \therefore R_L^2 = R_f \cdot R_e \qquad (1)$$

which is the expression for checking the bridged-T equation under matched conditions. The gain equations in Figures 1 and 2 are also simplified forms of the complex gain equation.

If we now produce a model embracing both R_f and R_e, equations may be derived which describe the input impedance, the output impedance and the gain under control of R_e, R_f, and β.

$$A = 2R_L(n\beta R_f - (\beta+1)R_e) \cdot [R_S((\beta+1)R_e + R_L(n\beta+1) + R_f) + (\beta+1)R_e(R_f + R_L)]^{-1} \qquad (2)$$

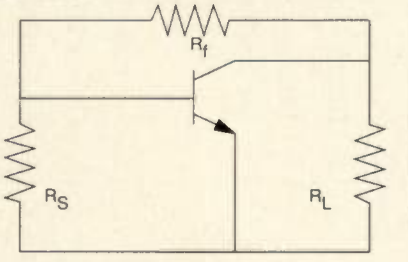
Letting $R_f \cdot R_e = R_L^2$, and $R_s = R_L$:

$$A = 2[n\beta R_f - (\beta+1)R_e) \cdot [2(\beta+1)R_e + (n\beta+1)R_L + (\beta+1)R_L + R_f]^{-1} \qquad (3)$$

Letting $\beta = \infty$:

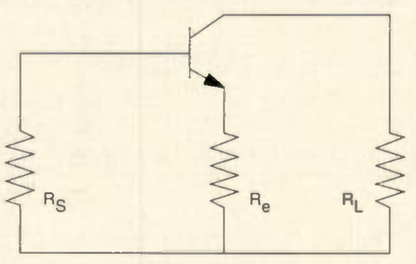
$$A = \frac{2[nR_f - R_e)}{R_L(n+1) + 2R_e} \qquad (4)$$

Letting n = 1:



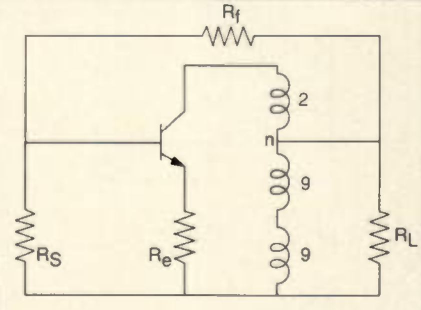
$A_i = R_f/R_L$

Figure 1. Shunt feedback circuit showing load, source and feedback resistors.



$A_v = R_L/R_e$

Figure 2. Series feedback circuit showing load, source and emitter resistors.



TR = n = 20/18 = 1.11

Figure 3. Double feedback circuit.

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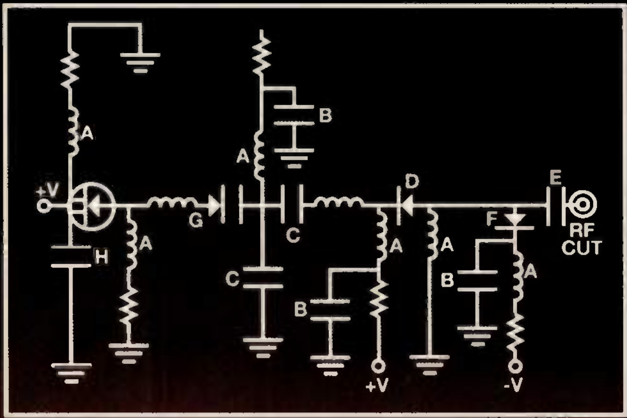

Typical switch with FSI semiconductors. (A) GC9002 Spiral Inductor; (B) GC84020 MNS Chip Capacitor; (C) GC86001 MNS Chip Capacitor; (D) GC4901 Beam Lead; (E) GC84030 MNS Chip Capacitor; (F) GC4221 PIN Diode; (G) GC15004 Linear Tuning Varactor; (H) GC84001 MNS Chip Capacitor.

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$$A = \frac{R_f - R_e}{R_L + R_e} \qquad (5)$$

Let $R_f = R_L^2/R_e$:

$$A = \frac{R_L}{R_e} - 1 \qquad \text{(Voltage Gain)} \qquad (6)$$

Let $R_e = R_L^2/R_f$:

$$A = \frac{R_f}{R_L} - 1 \qquad \text{(Current Gain)} \qquad (7)$$

Under matched conditions the voltage gain equals the current gain.

Since this feedback circuit has elements connected directly between the output and input, there exists a finite and measurable signal path opposite to the designed direction and it is called the reverse signal path. By analyzing the circuit for gain from output to input, an equation may be realized that sets the order of magnitude for the reverse path. Amplifier circuits involved with oscillators, mixers, frequency doublers and isolation stages require gain stages with controlled or predictable reverse gain values.

$$A_r = 2R_sR_e(\beta+1)\cdot \qquad (8)$$
$$[R_LR_e(\beta+1)+R_sR_L(n\beta+1)+$$
$$R_e(\beta+1)(R_s+R_f)+(R_sR_f)]^{-1}$$

Beta vs. Gain

These graphs were plotted from a computer run of the complex gain equation 3 with n = 1. Since there were many different values of R_e, two graphs were required in order to avoid confusion (only one is shown here). The computer program varied β from 3 to 21 for values of R_e from 6 to 28 ohms, with the constraining value of R_f, determined by $75^2/R_e = R_f$, also computed.

The equations and the graphs indicate that:
(a) Gain increases with increasing β.
(b) Gain increases with increasing R_f.
(c) Gain increases with decreasing R_e.
(d) Gain variation with β decreases with increasing R .
(e) Gain variation decreases with increasing β.

Beta was discontinued after 21 because the variation of gain from 21 to infinity is very small and the variation may easily be projected.

The graphs may be used by assuming a transistor with an f_t of say 1400 and a frequency of interest (f_m) of 200 MHz. This says that β or current gain will be 7 ($\beta = f_t/f_m$). Go into the graph at β = 7, and intersect, for instance, the line of the equation of R_e = 16. The gain may be read from either side of the graph. The notation at the left of each line indicates what the gain would be at $\beta = \infty$

We now have a group of equations that describe the operation of a feedback amplifier in terms of gain but, as yet, they are not useful. This is to say, in their present form they cannot help in designing an amplifier. All the equations do is to permit one to plug in values after completion of the design.

Up to now, we have no way to predetermine the values of R_f and R_e. One way to start is to use the simplified gain equation where gain is described in terms of R_e or R_f (with $R_f \cdot R_e = R_L^2$).

Solving equations 6 and 7 for R_e and R_f, respectively:

$$R_e = R_L/A+1 \qquad (9)$$
$$R_f = R_L(A+1) \qquad (10)$$

With R_L = 75 and a required gain of say 3 (about 9.5 dB),

$$R_e = 75/4 = 18.7 \text{ ohms} \qquad (11)$$
$$R_f = 75(4) = 300 \text{ ohms} \qquad (12)$$

Low-Frequency Gain

Instead of solving for many values of R_e, a graph was constructed which enables one to select the correct value of R_f and R_e for a given gain or vice-versa.

Figure 5 was plotted from the simplified gain equation 5 where $R_f \cdot R_e = R_L^2$, $\beta = \infty$, and n = 1. Since β is not in the simplified equation, no frequency dependability is implied; however, in the design of most amplifiers it is found that the low-frequency end of the response is higher than the high-frequency end. This effect is caused by variations of β and distributed circuit capacitance with frequency. The result is a tilted or sloping response, so that figure 5 can be used to set the low-end response only.

This graph proved very accurate and useful in many instances, even though the equations were calculated with n = 1 and the circuits all had transformers with n greater than 1. The reason for the accuracy will become evident later when a new graph will be plotted for different values of n.

The accuracy was noted when the graph was used to calculate gain with β at infinity. The calculated gain was very nearly the gain of a transistor amplifier with the same R_e and R_f but with a transformer. The increase in gain because of the transformer was in many cases similar to the gain when β was at infinity.

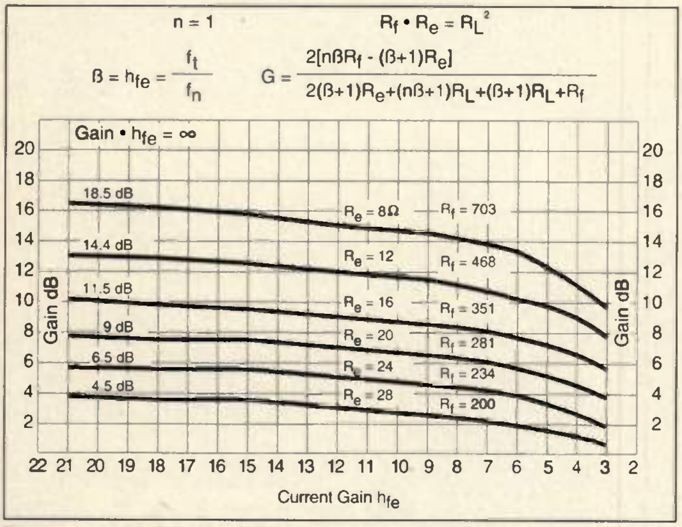

Figure 4. Gain versus different betas.

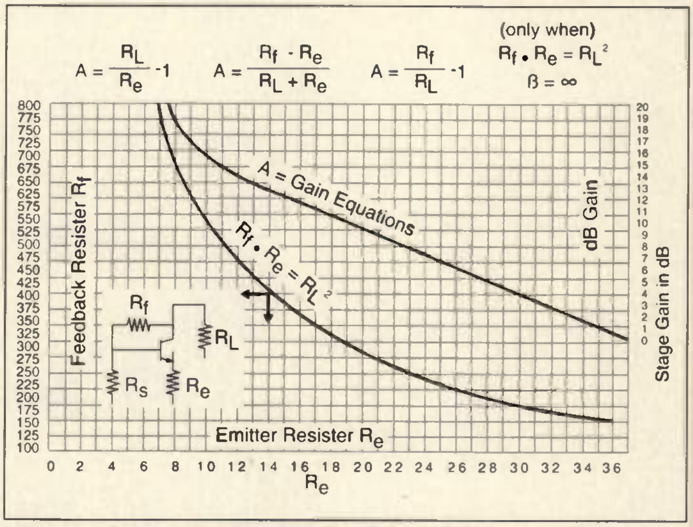

Figure 5. Graph relating R_f, R_e and gain.

To use the graph, select a gain at the right side of the graph, move to the left and intersect the gain equation line, move down to emitter resistor R_p value, move up to the curve of $R_f \cdot R_e = R_L^2$, at the intersection go left to the feedback resistor R_f value.

Since this graph represents the low frequency end of the response some degree of tilt will be exhibited. In order to insure that the slope will not be severe, refer the R_e value chosen to figure 4. In practice, with or without a transformer there will always be some inductance in the circuit that will hold up the gain at the high frequency end of the response curve. This inductance effect will be explained subsequently.

Input and Output Impedances

The equations for input impedance are:

$$Z_{in} = \frac{(\beta+1)R_e(R_f + R_L)}{(\beta+1)(R_e) + R_L(n3+1) + R_f} \qquad (13)$$

Let $\beta = \infty$

$$Z_{in} = \frac{R_e(R_f + R_L)}{R_e + nR_L} \qquad (14)$$

Let $n = 1$ and $R_f \cdot R_e = R_L^2$

$$Z_{in} = R_L \qquad (15)$$

The equations for output impedance are:

$$Z_{out} = \frac{(\beta+1)R_e(R_s + R_f) + (R_f R_s)}{(\beta+1)(R_e) + (r\beta+1)R_s} \qquad (16)$$

Let $\beta = \infty$

$$Z_{out} = \frac{R_e(R_f + R_s)}{R_e + nR_s} \qquad (17)$$

Let $n = 1$ and $R_f \cdot R_e = R_L^2$

$$Z_{out} = R_s \qquad (18)$$

Equations 13 and 16 are very similar and are equal to each other when β is at infinity, and equal to the input and output terminations R_s and R_L when $R_f \cdot R_e = R_L^2$ and $n = 1$. It was found that the easiest way to evaluate the action of R_f, R_e and β on Z_{in} and Z_{out} was to select appropriate values and look at the results. In order to obtain many reliable numerical values, equations 13 and 16 were programmed into a computer with R_f varying from 150 to 300 ohms, R_e varying from 10 to 40 ohms and β varying from 3 to 40. The action of the parameters of the equations for Z_{in} and Z_{out} may be summed up as follows:

1. Z_{in} (in ohms) increases from a value below 75 ohms with:
(a) Increasing β
(b) Increasing R_f
(c) Increasing R_e

2. Z_{out} (in ohms) decreases from a value above 75 ohms with:
(a) Increasing β
(b) Decreasing R_f
(c) Decreasing R_e

Experience and analysis of the input and output impedance values obtained from the computer run indicate that match and gain are direct trade-offs. This means that when an amplifier design is initiated it is not gain or match alone that is to determine the constants,

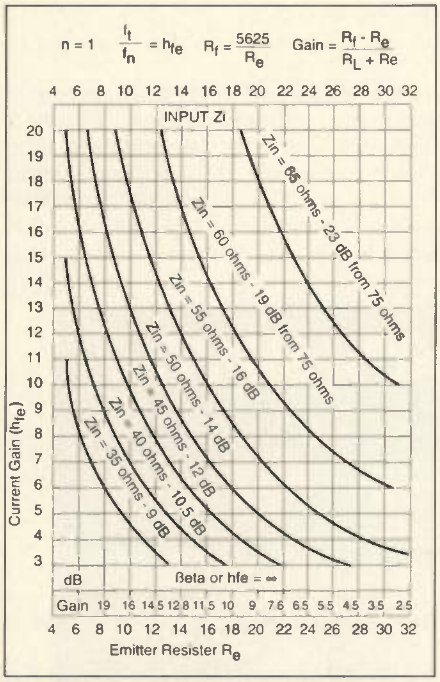

Figure 6. Graph relating beta, gain and input impedance.

but what match at what gain. This constraint is inescapable. With this in mind, it was decided to rearrange the input equation, put on the constraint.

$R_f = R_L^2/R_e$,

and solve the equation for R_e.

$$R_e^2(Z_{in}(\beta+1) - R_L(\beta+1)) + R_e((n\beta+1)R_L Z_{in} - 75^2 Z_{in} = 0 \qquad (19)$$

The purpose of this quadratic equation was to combine the relationship between gain and match. Although gain is not part of the equation it is implied since gain varies inversely with R_e. A convenient graph may be constructed from the equation by plotting different values of Z_{in} with β and R_e as the coordinates. In essence this graph permits the designer to go from theory to practice by relating R_e to β as frequency. In practice all the input impedance vs. R_e graphs were found to be very accurate, within 10% of Z_{in} in ohms. The degree of accuracy is reasonable when you consider that the input of the double feedback circuit is resistive with reactance secondary in nature.

Input Impedance (with n = 1),

The purpose of figure 6 is to permit a designer to obtain a first approximation of a particular design in the least possible time. The graph shows that with a particular f_t and a particular desired input match one set of R_e and R_f will be approximated that will match the curves. When the transistor (and therefore the f_t) is changed, the graph shows that other values of R_e and gain are optimal. However, the graph will show the varying input match with varying f_t and a constant R_e quite accurately .

The graph also indicates that gain and match are divergent and that a trade-off is required in the design of an amplifier.

Further, the graph indicates that with a constant R_e the input match will vary with f_t. The extent of the variation is also indicated.

Design Example

Given a transistor with f_t = 1300, and the highest frequency of operation is 200 MHz, then: h_{fe} = 1300/200 = 6.5.

Go into graph on the left hand side at h_{fe} = 6.5. As you traverse to the right, a number of lines labeled Z_{in} will be intersected; these curves represent input impedances. Choose Z_{in} and traverse down to the abscissa labeled "Re". For example, let us choose a Z_{in} of 50 ohms or –14dB. An h_{fe} of 6.5 intersects Z_{in} at an R_e of 16 ohms.

This says that a transistor with an f_t of 1300 and an R_e of 16 ohms, along with the constraining R_f = 5626/16 = 354, will give an input match at 200 MHz of 16 dB. Since R_f, R_e, R_L are related in the gain equation by $(R_f-R_e)/(R_L+R_e)$, equation (4), the gain obtained (in this case 11.5 dB) is included along with R_e.

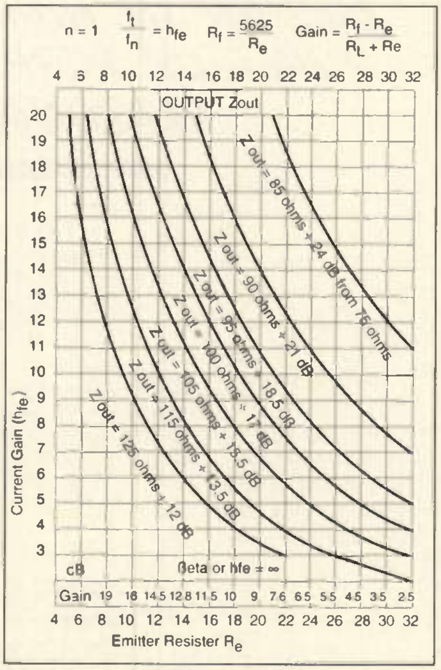

Figure 7. Graph relating beta, gain and output impedance.

Working with the general equations indicates that Z_{in}, Z_{out} and gain change very little after an h_{fe} of 20. With h_{fe} in the general equation, the response of an amplifier is usually shown to be falling off at the high-frequency end because of the reduced h_{fe}. This effect is noted in practice, and makes ingenuity a must for designing a flat amplifier.

A number of transistors were checked for f_t where the values ranged between 1000 and 1800. These transistors were then inserted in a typical feedback circuit.

By converting the f_t to β or frequency it was possible to predict with reasonable accuracy the input match at various frequencies, and the gain of the circuit associated with a particular emitter resistor. The match values obtained were accurate to 1 dB in the test circuits used.

A plot of the input or output impedance curve may be con structed by selecting a value of R_e from graph #5 or 6 and moving up that line noting the intersection between the plots of Z and current gain on the left Use graph #7 as a guide for assigning coordinates and values.

Output Impedance (with n = 1)

Figure 7 was constructed for Z_{out} using the same principle. However, after the graph was made it became obvious that Z_{in} and Z_{out} are symmetrical about 75 ohms and, therefore, are equal in dB at all points if $R_f \cdot R_e = R_L^2$. This means that graphs #6 and #7 may be used for both Z_{in} and Z_{out}, in dB only.

All input impedances are below 75 ohms in dB and all output impedances are above 75 ohms in dB.

For practical use, convert any dB from the output graph for input use by converting the dB value to ohms below 75; for using the input graph for output match, converting the dB value to ohms above 75.

The quadratic equation used for figure 7 is:

$$R_e^2((\beta+1)(R_s - Z_o)) + R_e(75^2(\beta+1) - R_s Z_o(n\beta+1)) + 75^2 R_s = 0 \qquad (20)$$

The accuracy of the output impedance graphs 7, 9b and 9d is limited by the shunting effect of the output capacitance of the transistor and associated circuitry which is in the order of 2 to 3 pF. In practice the effect of this is to lower the actual output impedance (in ohms) by approximately 20%. As β increases the error reduces in magnitude.

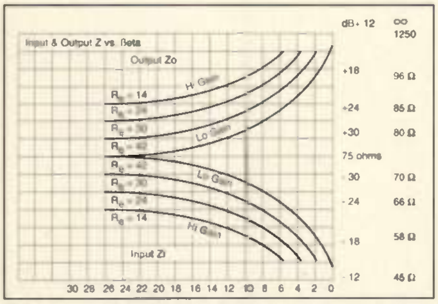

Figure 8. Graph showing how input and output impedances vary with beta.

Input and Output Impedance vs. Beta

This graph shows how the input and output impedances vary with β and with respect to the 75-ohm reference. Note that the impedances are equal and opposite, in dB, with respect to β and that the input and output curves for a given R_e are equidistant from the reference. This occurs only for n = 1. This graph clearly indicates that a trade-off of match vs. gain is mandatory in the design of an amplifier stage.

The graph was derived from the complex input and output equations with $R_f \cdot R_e = R_T^2$ and n = 1

This graph also indicates what happens when two similar stages are cascaded. At a β of 10 and an R_e of 24 the output impedance is about +19dB and the input impedance is about −19dB. Measured individually the stages appear to function very well but when they are coupled together the resulting performance is degraded because you are coupling 94 ohms to 60 ohms matched in dB but not in ohms.

The solution is to use a transformer that will make the transformation from high output Z to low input Z more compatible. Since both impedances vary with β the transformer turns ratio will have to be a compromise which lowers the high frequency output impedance closer to 75 ohms and lowers the low frequency impedance below 75 ohms to a value close to the existing input impedance of the following stage.

"n" and the Design Equations

With n appearing in both the numerator and denominator of the complex gain equation, its function is not immediately obvious. Let $R_f \cdot R_e = R_L^2$

$$A = 2(n\beta R_f - (\beta+1)R_e)\cdot [2(\beta+1)R_e + (n\beta+1)R_L + (\beta+1)R_L + R_f]^{-} \quad (21)$$

Let $\beta = \infty$

$$A = \frac{2(nR_f - R_e)}{R_L(n+1) + 2R_e} \quad (22)$$

In the simplified equation 17, n appears in both the numerator and denominator but, since R_f is always larger than R_L, the net result of increasing n will be an increase in gain. A few calculations will confirm this fact.

In both the input and output equations n appears in the denominator, so that increasing n reduces the value of Z_{in} or Z_{out}. Since the input impedance for practical values of β and R_e is always less than 75 ohms, making n greater than 1 always takes the input impedance further away from 75 ohms.

The output equation is another story.

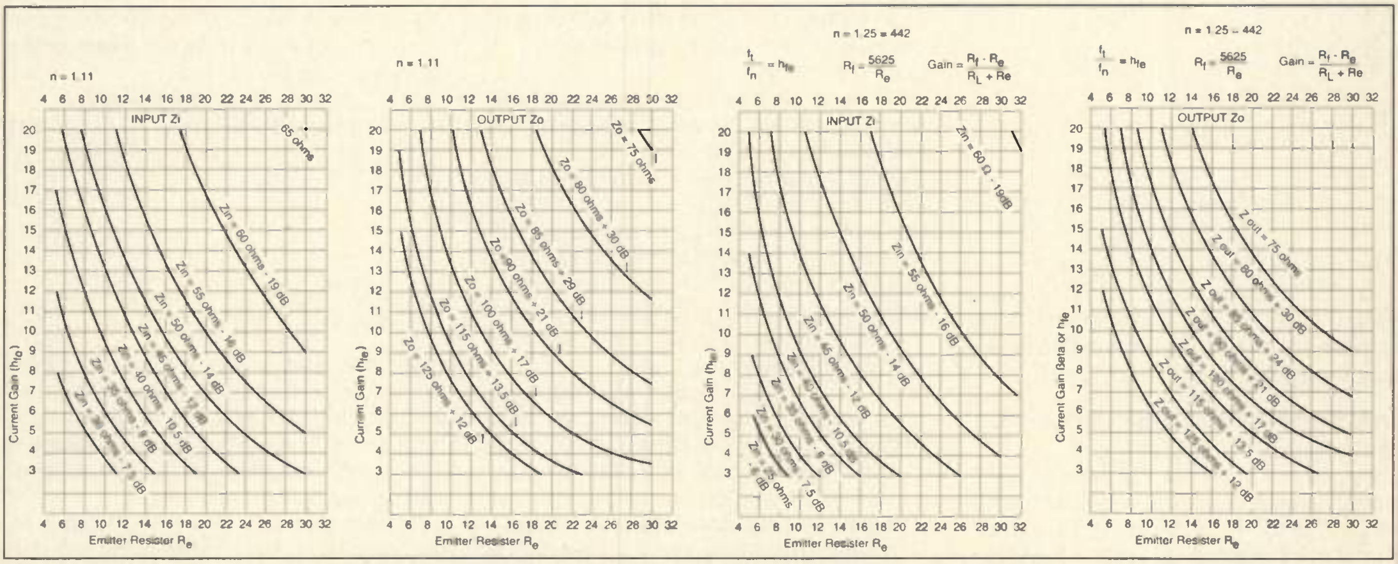

Figure 9a, 9b, 9c, 9d. Input impedances (9a, 9c) and output impedances (9b, 9d)for n = 1.11 (9a, 9b) and for n = 1.25 (9c, 9d).

Here the impedance for practical values of β and R_e is always above 75 ohms, so that making n greater than 1 moves the output impedance toward 75 ohms.

The action of, n greater than one, on figure 8, "Input and Output Impedance vs. Beta", is to move all curves down from their present position. What this does is to change the asymptote, now 75 ohms, to a value lower than 75 in accordance with the value of n. Any transformer then tends to make the amplifier input and output impedances lower in value. With 75 ohms as a reference this improves the output impedances but worsens the input impedances.

The simplified input and output equations are the same except for R_L and R_s being interchanged.

Let $\beta = \infty$

$$Z = \frac{R_e(R_f + R_L)}{R_e + nR_L} \qquad (23)$$

With n = 1 and $R_f \cdot R_e = R_L^2$

$$Z = R_L \qquad (24)$$

Input and Output Impedance (with n > 1)

These graphs were plotted from the same quadratic equations, 19 and 20, as were figures 6 and 7; however, for figures 9a-d, values of n greater than 1 were used. The values of n used were 1.25 (trifilar 4-4-2 winding) and 1.11 (trifilar 9-9-2 winding). By comparing some equivalent R_e vs. h_{fe} points on each of the graphs n = 1, n = 1.11 and n = 1.25 it becomes evident how n affects the impedances.

These graphs may be used in the same manner as were figures 6 and 7.

High Frequency Gain vs. Beta (n Variable)

Figure 10 was derived from the complex gain equation with $R_f \cdot R_e = R_L^2$, n = 1, n = 1.11, and n = 1.25.

As mentioned previously in a practical amplifier with a transformer in the collector, the leakage inductance of the trans former starts to increase the gain above 75 MHz to a greater value than the turns ratio a one would indicate.

Within the frequencies of interest and with the transistors available, a β of 21 was used in the gain equation calculation for constructing the graph. This value will be, for most transistors, below the point frequency where leakage inductance needs to be considered. This means that the graph will be useful as a first approximation for design.

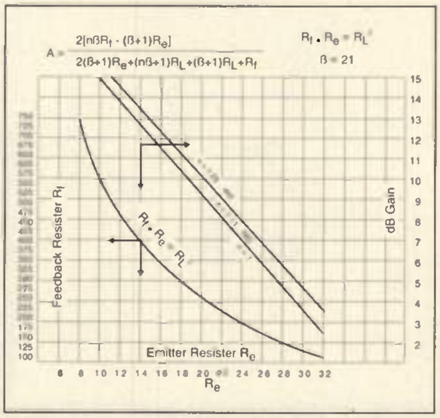

Figure 10. Gain versus beta in high frequency operation and with variable n.

Notice that the maximum difference in gain from n = 1 to n = 1.25 is about 1.5 dB, which is approximately equal to the difference between the gain at β = 21 and the at β= ∞ on the β vs. Gain graph, (figure 4). Compare figure 10 with the low frequency gain graph (figure 5) and note the numerical similarity of gain for a given value of Re as was mentioned in the discussion of graph #3. ***RF***

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About the Author

Frank Egenstafer's background in electronics has been circuit design from DC to 2 GHz, and from relays and stepper switches to vacuum tubes and transistors. He has mostly worked on the "bench" even in the several managerial positions he has enjoyed. He attended Drexel Institute in Philadelphia. He can be reached at (304) 562-6582.

Frequency Independent Phase Tracking System Creates A True Phase Locked Loop

By Glen A. Myers, Ph.D.
Kintel Technologies, Inc.

In a phase-locked loop, the phase of the VCO output voltage relative to that of the input voltage varies with the frequency of the input voltage. This article considers the realization of a circuit for which this phase is selectable with an applied DC voltage and held constant independent of the frequency of the input voltage. The operation of this patented circuitry and some possible applications are presented here.

The block diagram of a phase-locked loop (PLL) is shown as Figure1. The phase detector consists of a voltage multiplier followed by a lowpass filter (LPF). Most PLLs are contained in systems as dedicated integrated circuits (ICs). And, most of these are configured using digital techniques; the voltage-controlled oscillator (VCO) may be a multivibrator (flip-flop), and the phase detector multiplier an XOR or XNOR gate (1). A limiter, which is commonly part of the IC, converts the analog input into a two-level voltage suitable for one input of the XOR gate. Apart from all the advantages of digital ICs (cost, reproducibility, stability, operability), this realization of PLLs provides linear characteristics of the VCO and also the phase detector (Figures 2 and 3).

A PLL is usable when the frequency of the VCO output voltage equals that of the PLL input voltage. This is called the lock condition. Suppose the frequency of the input moves from a value of f_A to a new value f_B as shown in Figure 2. To maintain lock, the frequency of the VCO output must move in a like manner. This will occur only if the VCO input voltage moves from v_A to v_B (Figure 2). This voltage must come from the phase detector, and this means that the difference in phase of its two input signals must move from ϕ_A to ϕ_B as shown in Figure 3. Clearly, the loop is not phase locked (fixed in phase). Rather, a PLL is frequency locked.

This operation can be explained mathematically for the characteristics of Figures 2 and 3 by writing:

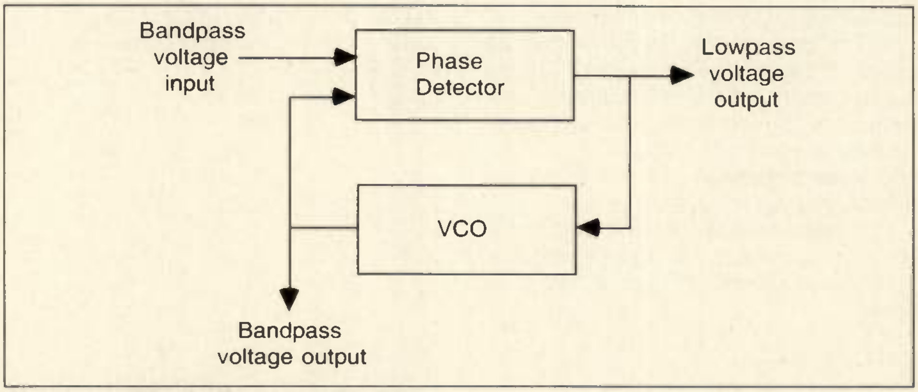

Figure 1. Block diagram of a PLL.

$f_0 - f_r = k_f v$ for the VCO

$v = k_p(\phi - \phi_0)$ for the phase detector

where:

f_0 = frequency in Hz of the VCO output voltage
f_r = constant = "at rest" frequency of the VCO (f_0 when $v = 0$)
k_f = slope of the VCO characteristic in Hz per volt (constant)
v = VCO input voltage = phase detector output voltage
k_p = slope of the phase detector characteristic in volts per radian (constant)
ϕ = relative phase of the phase detector inputs
ϕ_0 = that value of phase for which $v = 0$

Combining these two equations by eliminating v and setting $(k_f)(k_p) = k_L$ gives:

$$(\phi - \phi_0) = (f_0 - f_r)/k_L = (f_{in} - f_r)/k_L$$

which shows dependence of ϕ on $f_0 = f_{in}$ (lock condition) where f_{in} is the frequency of the input

A few comments are in order here before proceeding:

1) The constant k_L includes any voltage gain in the loop.

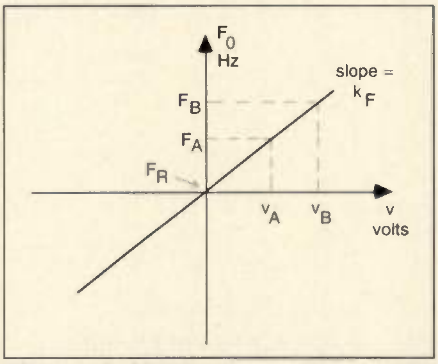

Figure 2. VCO voltage/frequency characteristics.

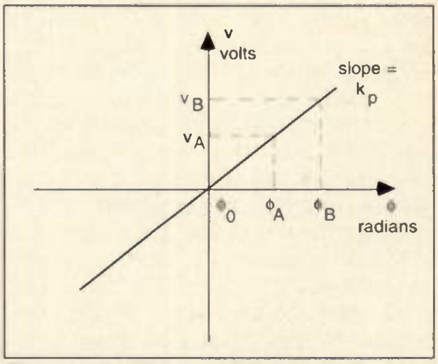

Figure 3. Phase detector characteristics.

2) In the use of a PLL, f_r is usually set by suitable choice of a resistor and capacitor external to the IC.

3) The value of ϕ_0 depends on the design of the IC. In analysis, ϕ_0 is commonly taken as $\pi/2$ radians. In practice, unipolar logic is common and ϕ_0 has other values. See Figures 2 and 5 of Reference 1 for example.

4) The equations as presented are for the static case. They are also useful in the case of a changing frequency (modulation) of the loop input voltage provided the loop LPF can accommodate (output follow) the dynamics of the input.

5) The loop has two usable outputs as shown in Figure 1. The lowpass voltage is used when the loop is functioning as a frequency demodulator. The bandpass voltage is used when the loop is functioning as a bandpass filter or as part of a frequency synthesizer.

Now, let us open the loop and apply an external voltage v_x by means of a summer as shown in Figure 4. We then have:

$$f_0 - f_r = k_f(v + v_x)$$

$v = k_p(\phi - \phi_0)$ as before.

Therefore,

$f_0 - f_r = k_f v_x + k_f k_p(\phi - \phi_0)$ or,

$$\phi - \phi_0 = (f_0 - f_r - k_f v_x)/k_f k_p$$

Previously ($v_x = 0$), $\phi = \phi_0$ when $f_0 = f_r$. Now ($v_x \neq 0$), ϕ has to adjust to accommodate this external voltage. The amount of the adjustment (from ϕ_0) is $-v_x/k_p$ radians. This is a well known method of using a PLL as a phase modulator (2).

We have now set the stage for the presentation of a true phase-locked loop — a patented circuit whose output maintains a constant phase relation with the input voltage independent of the frequencyof the input voltage (3), see Figure 5. We have added a second PLL-whose behavior is affected by both outputs of the first PLL. Let the subscripts 1 and 2 be applied to parameters of the first and second PLL respectively. With both loops in lock, then $f_{02} = f_{01} = f_{in}$. As with any one PLL, this condition must be satisfied for the phase tracking circuit (PTC) to be usable. By coupling the correct voltage ("DC") from loop 1 into loop 2, we can force the second loop to compensate for this voltage by adjusting its phase ϕ_2 such that $\phi_1 + \phi_2$ does not vary with frequency. Therefore, we have realized a frequency independent phase tracking system.

This operating principle of the PTC is readily revealed by example. Suppose fin moves from a value of f_A Hz to a new value of f_B Hz. Consider that this requires a 0.75 volt change at the input to the VCO of the first PLL to maintain lock. Let this correspond to a 20 degree increase in ϕ_1. Assume $k_{f1} = k_{f2}$ and $k_{p1} = k_{p2}$. This is reasonable if identical ICs and identical external circuitry (LPFs) are used to realize loops 1 and 2. For this case, we need a 0.75 volt change at the input to the VCO of the second PLL for it to maintain lock. But we have injected into loop 2 the value of (2)(0.75) = 1.5 volts. This means the phase detector of loop 2 must create –0.75 volts which it does by causing ϕ_2 to decrease by 20 degrees or the same amount that phi1 increased. The net result is a value of $\phi_1 + \phi_2$ which remains constant with fin. The circuit is, then, a frequency independent phase tracking circuit — a circuit which is locked in frequency and set in phase.

A general equation for $\phi_p = \phi_1 + \phi_2$ involving all the parameters of loops 1 and 2 can be derived for the static case in a straightforward manner. This does, perhaps, provide no more insight than

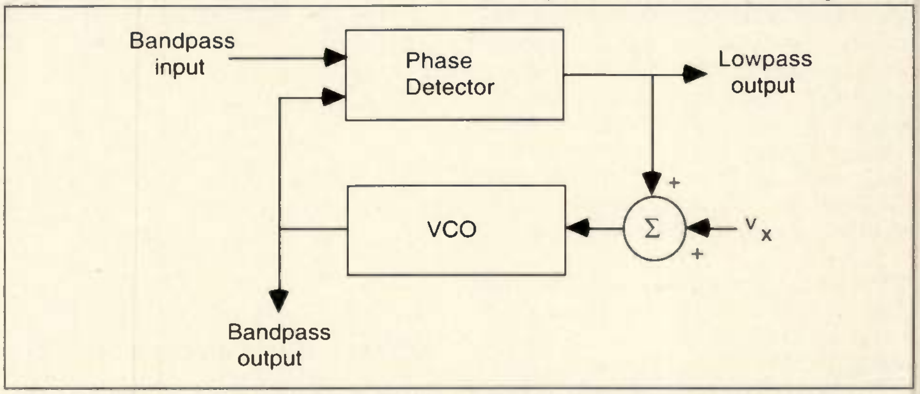

Figure 4. Block diagram of a PLL with an external applied voltage, v_x.

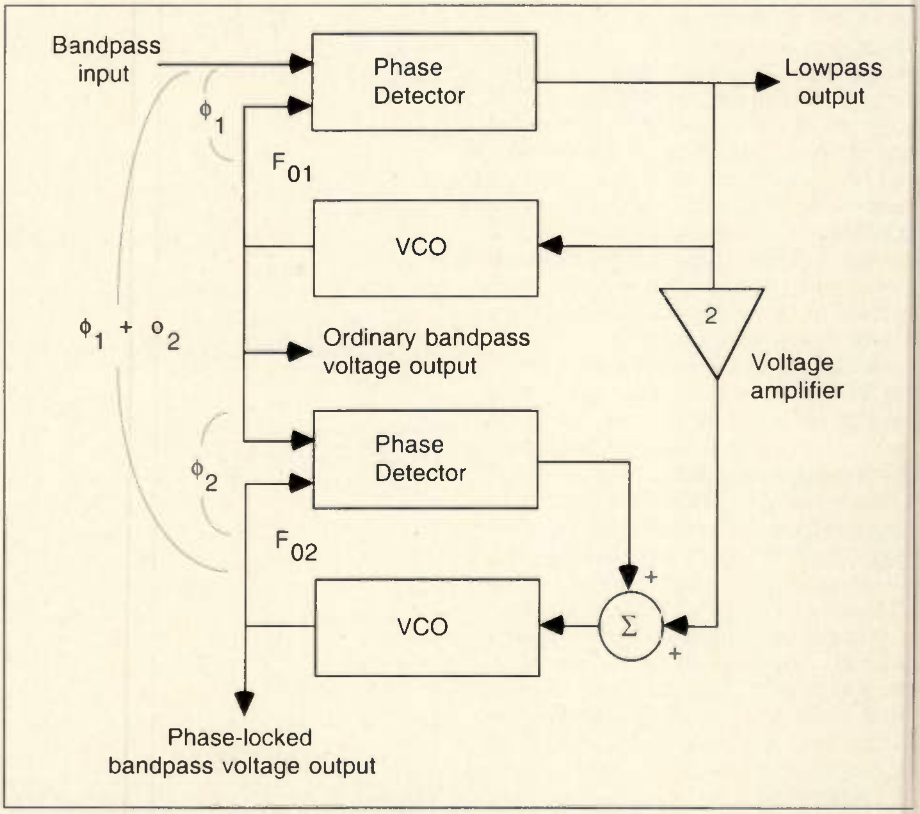

Figure 5. Block diagram of the phase tracking circuit.

the numerical example.

We have built, tested and operated various such phase tracking circuits. Excellent phase tracking is maintained under dynamic conditions (frequency modulated [FM] input) over the entire lock range of the PLL.

In our practice, an external DC voltage is applied to the summer of Figure 5 to set the desired value of ϕ_p and to compensate for differing values of the rest frequencies of the two PLLs. This is a static adjustment. With a FM input, the gain of the voltage amplifier is then adjusted to make k_L the same for both loops. This ensures that phip does not vary with frequency. These adjustments are readily made by observing a suitable oscilloscope presentation.

A constant value of $\phi_p = 0$ is desirable for some applications (frequency synthesizers, coherent demodulators). We have used a value of $\phi_p = \pi$ radians to suppress a strong interfering carrier. This same value provides us with a patented Power Multiplexing™ capability (multiple frequency reuse) which we have successfully demonstrated (4, 5).

PLLs are also used as bandpass filters. In this case, the PTC provides a constant-phase bandpass filter. When the input is FM, for example, no incidental phase modulation is created by the filtering action of the PTC.

The ability to vary ϕ_p in a known, controlled manner with an external voltage (third input to the summer of Figure 5) has useful applications. For example, in the interferometry method of direction finding, ϕ_p can compensate for the phase shift between two antenna elements and thereby create constructive or destructive interference when their outputs are summed. The condition of interference when detected can be related to wave angle of arrival (patent pending)(6). Conversely, in a transmitter feeding multiple antenna elements, use can be made of the voltage-controlled phase to sweep the beam spatially in a prescribed time-dependent manner. Clearly, using reciprocity, the same beam-forming system can be used to create a phased array for reception (search mode of the receiver). So, use of a PTC in series with each radiating element creates a simple, rapid, highly flexible and easily programmable (computer control of the tuning voltage) phased-array antenna.

Use of multiple phase tracking circuits connected in series or other combinations can serve other useful purposes such as suppression of strong interference and measurement of angle of arrival of each of several FM carriers in the same frequency band. We continue to explore applications of the frequency independent phase tracking capability. We invite the involvement of others in the development of this technology. *RF*

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companies and institutions mentioned in this issue's editorial pages

RF product report

Power Transistor Manufacturers Aim for New RF Markets

By Gary A. Breed
Editor

RF power transistors continue to be an active part of RF research and product development. At this time, the primary force behind these products is the potential for large new markets in the 900-2450 MHz range. However, new satellite systems also represent a major area of development activity.

The continued move toward massive new markets in personal and business communications has been included in the strategy of nearly all RF power transistor makers. The traditional large manufacturers are extremely active, and at the same time, smaller companies are growing.

Young Companies Look for A Piece of Growing RF Business

Among the newer entrants into the RF power marketplace is RF Micro Devices, which is developing Heterojunction Bipolar Transistor technology (HBT) for digital cellular and other digital transmission applications, initially in the 800-1000 MHz range.

RF Products is developing MOSFET technology for linear amplification in the 900 MHz range, as well. A recent announcement was the WRLS0941 power amplifier module, a 1 watt device operating from a 12.5 volt supply.

MicroWave Technology continues its development of Solid State Triode (SST) technology with devices for HF, VHF and UHF/L-band. The company's SLAM linear internally-biased devices target medium and high power applications from 1-100 MHz, and their L-band pulsed power products have been developed for transponder and radar applications.

Modular products, devices with supporting circuitry to make complete amplifier modules, is another growing area of RF power transistor development. The demands for shortening the design cycle and for reducing manufacturing complexity are both served by value-added products. Power amplifiers using discrete components have traditionally been engineering-intensive, involving matching network design, broadband optimization, allowances for variations in production specifications, plus mechanical considerations like mounting and heat dissipation. These issues are solved almost completely by designing a RF product with a drop-in power amplifier module.

Motorola, M/A-COM, MicroWave Technology, and Philips all offer amplifier modules that incorporate their transistors. Efforts are underway to meet additional needs for low voltage operation, low standby power consumption, good linearity for digital modulation and of course, lowest possible cost.

Applications and Requirements

The RF power applications seen as most important range from digital cellular to PCS to satellite communications. Other applications that are getting transistor manufacturers' close attention are MRI systems, RF heating and sputtering, television broadcasting, VHF/UHF mobile radio, collision avoidance radar. Military applications in voice and data communications at HF through microwaves, plus various countermeasures and radar systems, represent significant business at some companies, and continue to get special attention.

All digital transmission systems, including digital cellular and PCS require several performance features, including linear amplification for the complex digital modulation envelope, power control capability, and good efficiency, although this conflicts directly with the linearity requirement. Complicating product development and designer's choice of an appropriate device is the fact that the 900-2000 MHz range represents a transition region. Bipolar technology is not as simple as it is at lower frequencies, especially at higher powers needed for base stations, but GaAs FET devices are expensive and may have special DC power requirements. HBT devices and SST technology are other alternatives, as mentioned above.

Satellite communications remains a strong market as more companies establish worldwide networks, maritime communications changes from HF to satellite, and as more aircraft are fitted with satellite communications links. California Eastern Laboratories is actively marketing NEC power MESFETs for satellite systems, including a new line for C-band (4.4-8.5 GHz) transponders, with per-device power output to 18 watts.

MRI has faded as a headline item, but these medical systems represent a significant market niche. Vacuum tubes and transistors share this market, which covers the 10-100 MHz range, with amplifier outputs in the hundred watt to kilowatt range.

Industrial applications at 13.56, 27.12 MHz and other ISM-bands are quietly growing. These heating, sputtering and plasma applications require high efficiency techniques, often using switching-mode amplifiers, although class C is the dominant technique. High power handling and efficiency are key transistor specifications.

VHF/UHF mobile radio systems are a well-established market, and many regions of the country are nearly saturated with users. Among companies active in this area is SGS-Thomson, which has made particular efforts to develop high power transistors for base stations. Inland and coastal marine radio is in this same part of the spectrum, and is a steadily growing market. These systems are virtually all FM, requiring straightforward class C amplifiers.

Military systems is a shrinking market, but fewer companies are targeting military applications for their products. The remaining business is a significant number of dollars, and companies which are successful at mixing a military and commercial customer base can realize continued profits.

This last example may be the best indicator of the RF market. There are many different applications to be served, and flexibility is as much a marketing asset as a good product line. *RF*

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Active Filter Design

AFDplus is an active filter design and analysis package. The package creates lowpass, highpass, bandpass, bandstop and allpass filters, using any of several response functions. The menu-driven program includes sensitivity and worst case analyses, PLL synthesis and analysis, netlist generation and cascadable filter designs. AFDplus ships from stock for $995.

Webb Laboratories
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RF literature

Standard Parts Catalog

The Standard Parts Catalog from Synergy Microwave contains descriptions of Synergy's lines of surface mount mixers, surface mount power dividers, surface mount 90° hybrids, surface mount 180° hybrids, wideband 90° hybrids, and passband filters. Also included in the 38-page catalog are a surface mount tutorial and technical articles on SMD-MIL packaging and a metal hermetic SMT package.

Synergy Microwave Corporation
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Switch/Attenuator Catalog

M/A-COM has published a new GaAs MMIC Switch and Attenuator catalog. The catalog provides product specifications, application notes and outline drawings for components designed for cellular, PCN/PCS, IVHS, GPS, medical and military applications. Among the products in this expanding product line are eight new plastic components.

M/A-COM, Inc.
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UHF Radio Products

An updated guide to RF Monolithics' (RFM) low-power UHF radio products is now available from the company. The guide lists RFM's wireless products for original equipment manufacturers. SAW components within the 224 to 928 MHz range are matched to frequencies used around the world. Surface mount hybrid AM transmitters, resonators, and delay lines are among the devices listed.

RF Monolithics, Inc.
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Full Line Catalog

Active and passive devices operating in the 5 MHz to 26.5 GHz range are described in Microwave Communications Laboratories' 85-page catalog. Dividers/combiners, couplers, hybrids, D.R.O.s, filters, switches, terminations and attenuators are among the products described.

Microwave Communications Laboratories, Inc.
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Trimmer Catalog

Sprague-Goodman offers a four-page engineering bulletin featuring their complete line of trimmer capacitors and inductors for microwave tuning. Bulletin SG-670 incorporates features and specifications, plus photos, outline drawings and charts for the six elements available: metallic tuning elements, dielectric tuning elements, metallic and dielectric tuning rotors, dielectric resonator tuners, LC tuning elements and resistive tuning elements.

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Antenna Catalog

Antenna Research's 130-page catalog of transmit and receive antennas and accessories for communication, direction finding, GPS, telemetry, search and surveillance is now available. The catalog covers a full range of omnidirectional, directional and hemispherical coverage antennas from 100 Hz to 40 GHz. Accessories include field intensity meters, tripods, preamps, rotators, masts and an antenna analyzer.

Antenna Research Associates, Inc.
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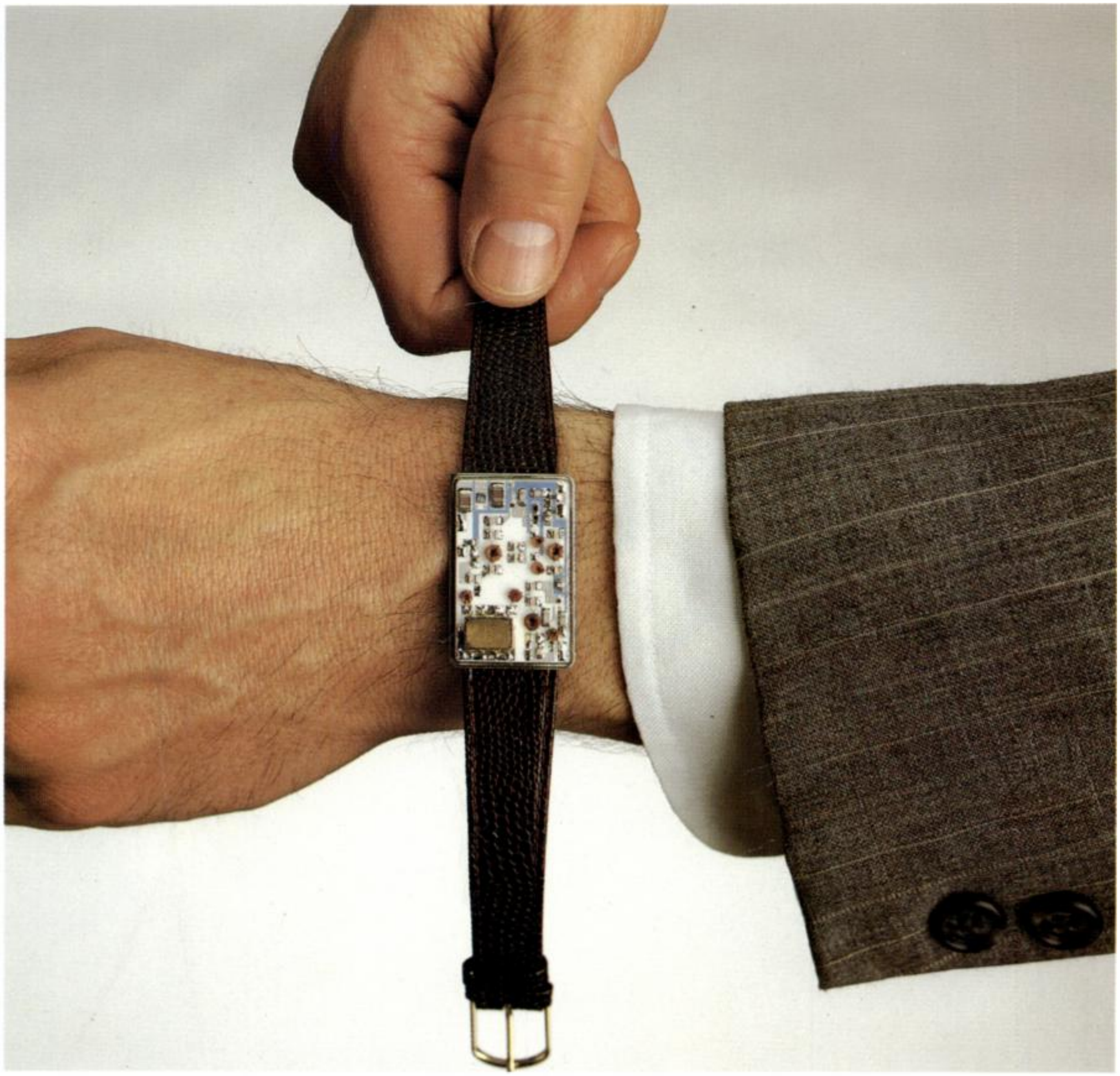

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