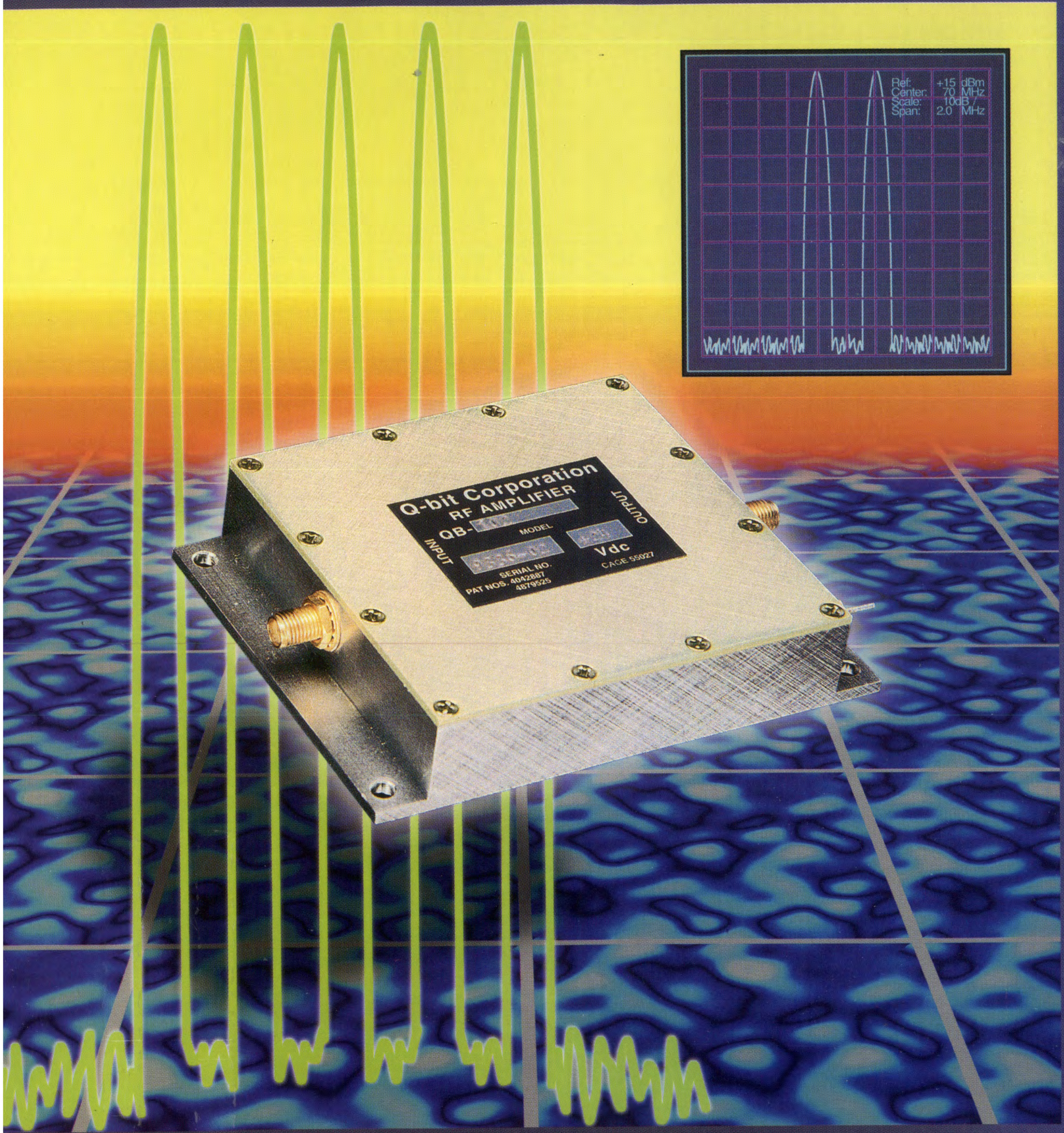
RFdesign™

engineering principles and practices

December 1993



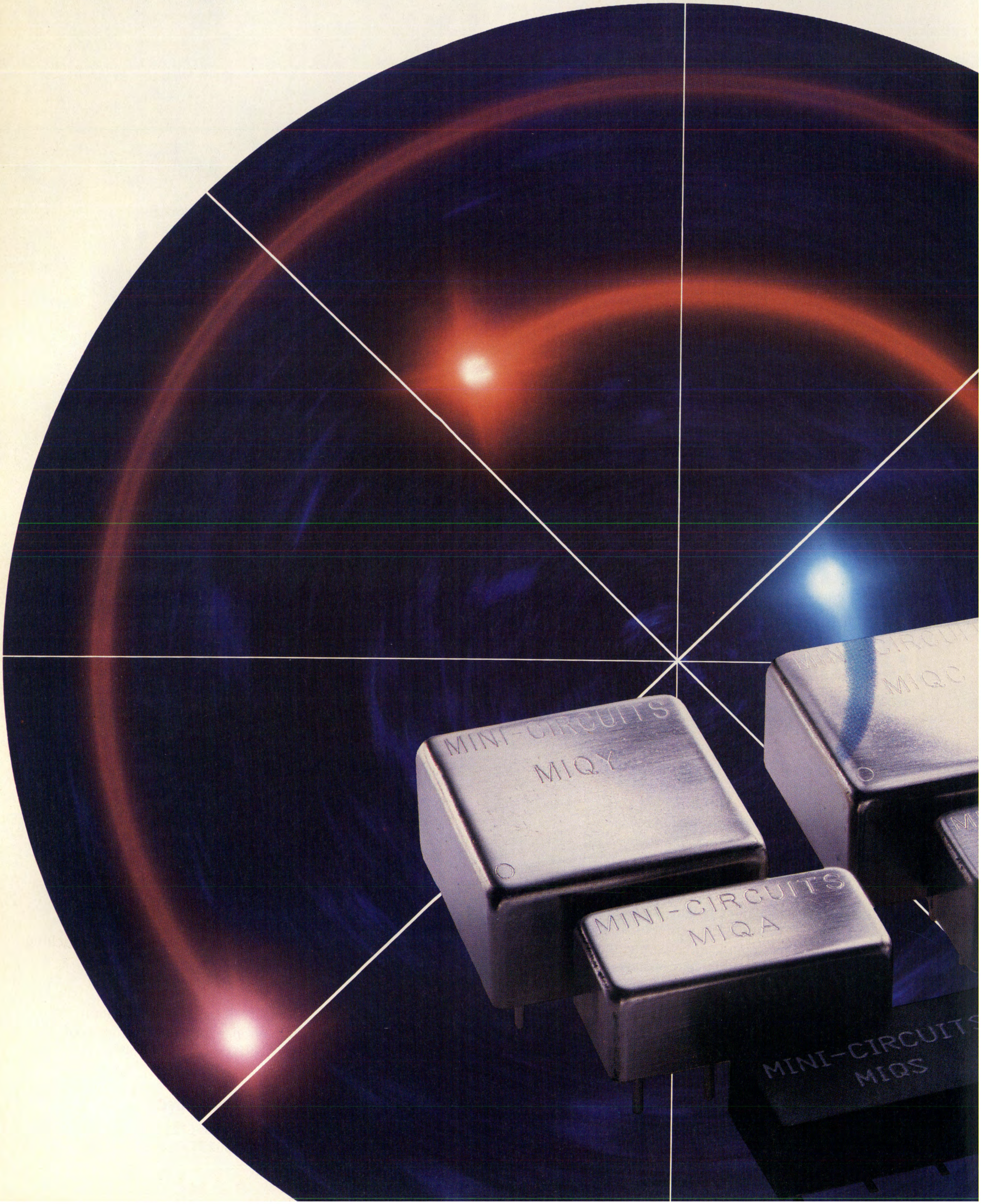
Cover Story

New High Dynamic Range Amplifiers

Featured Technology

Low Noise Design

I&Q MODULATORS



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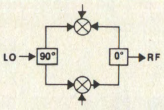
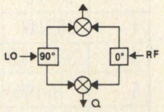


I/Q MODULATORS

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

□ NON-HERMETICALLY SEALED
MIQA case .4 x .8 x .4 in.
MIQY case .8 x .8 x .4 in.
MIQC case .8 x .8 x .4 in.
MIQS case .8 x .4 x .2 in.

INFO/CARD 3

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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 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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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

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SWITCHES ATTENUATORS PHASE SHIFTERS MMICS BIT DETECTORS COUPLERS MODULATORS AMPLIFIERS

RF design

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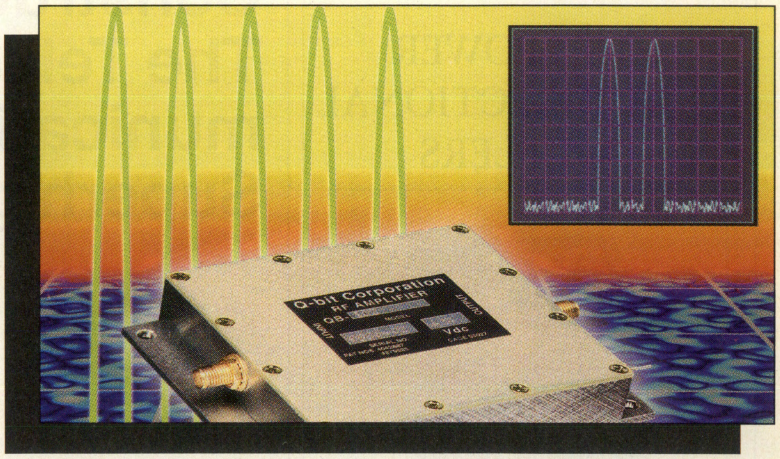

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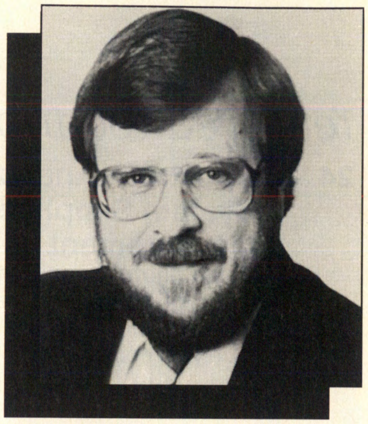
RF editorial

Under Construction: The Telecommunications Superhighway



By Gary A. Breed
Editor

Big corporate deals are being made, big money is being invested, and big plans are being drawn up for the so-called "information superhighway." Bell Atlantic has agreed to buy Telecommunications Inc. and form the country's sixth-largest corporation. US West has made a major investment in Time-Warner, and other telecommunications and programming companies are being bought, sold and refinanced. These new companies will have capabilities for voice, television and data communications. For the first time, single companies will be able to bring the entire range of telecommunications services into homes and businesses.

For proponents of the concept that years ago was called the "wired nation," their predictions are finally coming true. Fiber optic systems will eventually replace the twisted pairs of the phone company and the coaxial cables of the CATV companies. Radio links will provide jump-off points for new mobile and portable communications systems that are now in development.

To provide the hardware and operating systems for the proposed new infrastructure, major computer, radio and component companies are forming alliances. Motorola, IBM, Apple, AT&T, the "baby Bells" and many others have created a whole series of mix-and-match partnerships for hardware and software development.

What does this mean? The major result of this major investment is a boost in the overall electronics business. These companies believe that they are leading the way into the next boom in the electronics market. Their business activities have moved from evaluation to development. Their investments are intended to bring to market the pioneering work in new radio-linked applications that you have been reading about in *RF Design* and elsewhere. We will soon see if they are making a safe bet!

Recent Changes at *RF Design*

You may have read the letter from president Jerry France in September, and noticed some new names on our masthead over the past couple of issues (on page 10 this month). A reorganization by our parent company has given us a new company name, Argus Business, and some changes in corporate personnel. Our new Group Publisher, David Premo, took the reins of *RF Design* just before RF Expo East, and met some of you in Tampa.

You may also note that our corporate functions (Circulation, Accounting, Trade Show Management and Production) are now located in Atlanta, Georgia instead of Englewood, Colorado.

If you *haven't* noticed these corporate changes, it's because they have not resulted in changes to the magazine! Our editorial staff is the same, and remains in the Colorado offices. Our current advertising sales force will be expanding to better reach the growing RF industry.

We are planning a few small changes in *RF Design*, but only as part of our ongoing efforts to do the best job we can. The monthly reader survey has been discontinued, and you will see other means of research in the future to gauge your needs and interests. Also, our RF Design Awards Contest will be updated for 1994 — but it will still have the same kind of fabulous prizes that we have had all along. Watch the January issue for details!

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	210LC	10W CW	.008-225 MHz	40dB	$ 2,495
	710FC	10W CW	1-1000 MHz	40dB	$ 6,695
	*727LC	10W CW	.006-1000 MHz	44dB	$ 7,750
NEW	713FC	15W CW	10-1000 MHz	42dB	$ 4,250
	225LC	25W CW	.01-225 MHz	40dB	$ 3,295
	*737LC	25W CW	.01-1000 MHz	45dB	$ 9,995
	712FC	25W CW	200-1000 MHz	45dB	$ 6,950
NEW	714FC	30W CW	10-1000 MHz	45dB	$ 9,950
	250LC	50W CW	.01-225 MHz	47dB	$ 5,250
	715FC	50W CW	200-1000 MHz	47dB	$ 16,990
	707FC	50W CW	400-1000 MHz	50dB	$ 9,990
NEW	716FC	50W CW	10-1000 MHz	47dB	$ 17,950
	*747LC	50W CW	.01-1000 MHz	47dB	$ 19,500
	116FC	100W CW	.01-225 MHz	50dB	$ 8,800
	709FC	100W CW	500-1000 MHz	50dB	$ 16,990
	717FC	100W CW	200-1000 MHz	50dB	$ 19,500
NEW	718FC	100W CW	10-1000 MHz	50dB	$ 26,950
	*757LC	100W CW	.01-1000 MHz	50dB	$ 29,950
	122FC	250W CW	.01-225 MHz	55dB	$ 19,950
	723FC	300W CW	500-1000 MHz	55dB	$ 29,995
	LA500G	500W CW	500-1000 MHz	57dB	$ 53,500

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122C	200W CW	.01-220 MHz	53dB	$ 11,750
134C	500W CW	.01-220 MHz	57dB	$ 19,800
137C	1000W CW	.01-220 MHz	60dB	$ 27,550
140C	2000W CW	.01-220 MHz	64dB	$ 44,990

Warranty: Full 18 months all parts. Vacuum tubes 90 days.

* Indicates Dual-Band System (coaxial band switching)

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Group Publisher
David Premo (303) 220-0600

Editor
Gary A. Breed (303) 220-0600

Technical Editor
Andrew M. Kellett (303) 220-0600

Consulting Editor
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INFO/CARD 8

GIGA-TRONICS 8540 UNIVERSAL POWER METER

Incredibly Fast and Accurate CW *and* Peak Power Measurements At A Truly Incredible Price.

UNIVERSAL POWER MEASUREMENT

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

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

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

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

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.

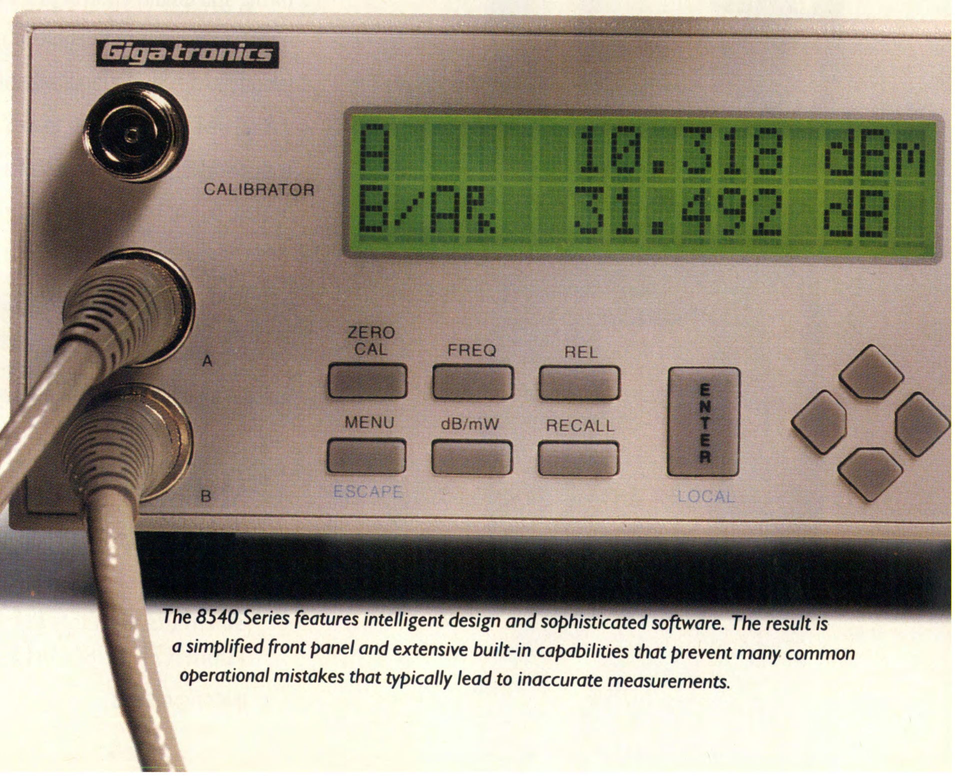
FAST, EASY PEAK POWER MEASUREMENT

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.

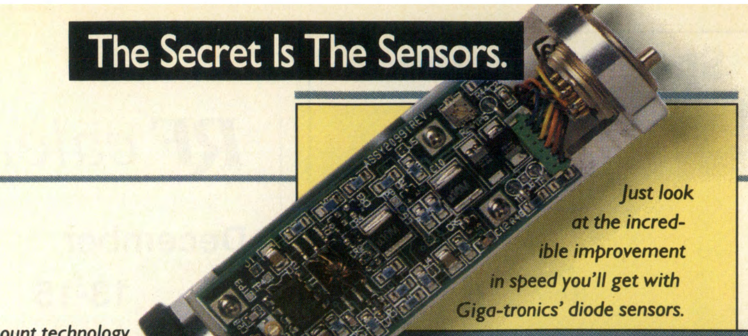
View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.

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



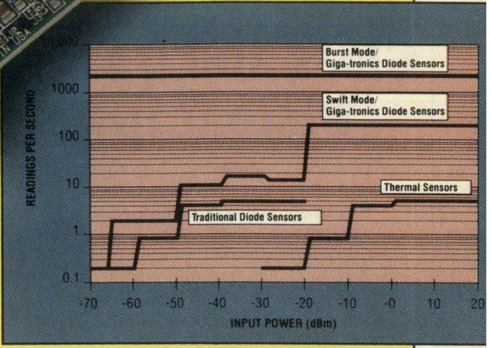
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

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



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

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

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

ONE OR TRUE TWO CHANNEL OPERATION

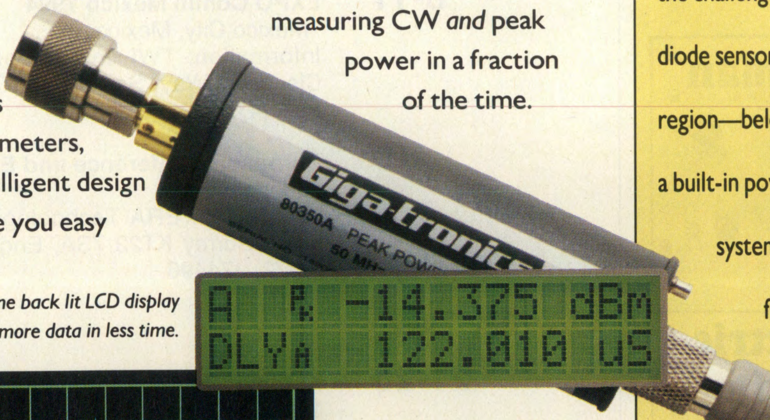
If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously*.

SIMPLE, INTELLIGENT OPERATION

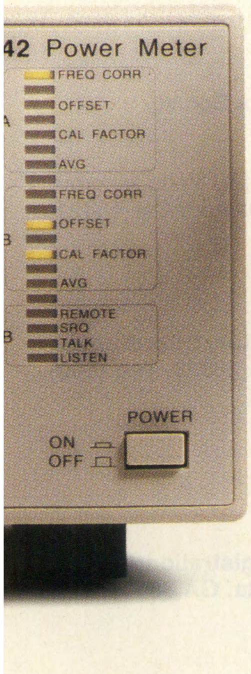
The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

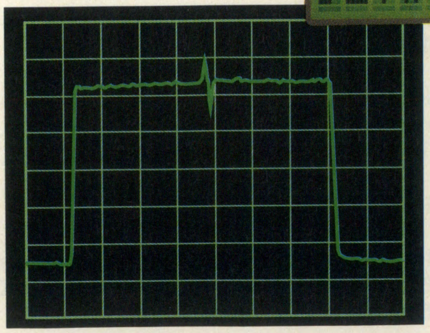
For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

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



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





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

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INFO/CARD 9

RF calendar

December

13-15 **Seventh International Conference on Mobile and Personal Communications**
Brighton, UK
Information: Conference Services, IEE, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071 344 5477. Fax: 071 497 3633.

15-18 **4th International Symposium on Recent Advances in Microwave Technology (ISRAMT '93)**
New Delhi/Agra, India
Information: Dr. Banmali Rawat, Technical Program Co-Chair, Department of Electrical Engineering, University of Nevada, Reno, NV 89557-0153. Tel: (702) 784-1457. Fax: (702) 784-6627.

1994

January

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.

10 **European Conference and Exhibition: EMC 94**
London, UK
Information: ERA Technology Ltd., Cleeve Road, Leatherhead, Surrey KT22 7SA, England. Tel: (0372) 374151. Fax: (0372) 374496.

March

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.

S H A R P

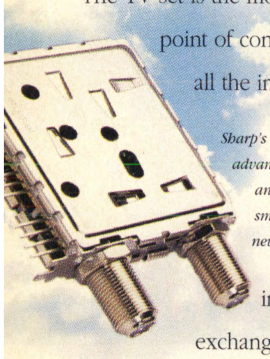
PROGRESS REPORT

Dedicated to the proposition that one good idea leads to another.

From the crystal radio to the wire

Fifty years ago, getting the "big picture" at home depended mostly on your powers of imagination. But the pictures to be delivered to homes within this decade may exceed what all but the most ardent futurist *ever* imagined.

The TV set is the most likely point of confluence for all the innovations in information exchange now converging on the home market, with every media from video disks to fiber optic highways jostling for position. At Sharp, we've always been partial to the path of least resistance — the airwaves — ever since we created and marketed the first mass-produced crystal radio sets back in 1925.



Sharp's ultra-compact DBS tuners — with their advanced designs, reduced component count and low power consumption — contribute to smaller system sizes and have established new paradigms for reliability and integration.

You might be surprised at the number of areas in which Sharp continues to be the pacesetter in wireless communications components. Whether your transmission is in RF or IR bandwidths, by terrestrial or satellite broadcast, across the continent or across the room, here's why it pays to have Sharp go to bat for you.

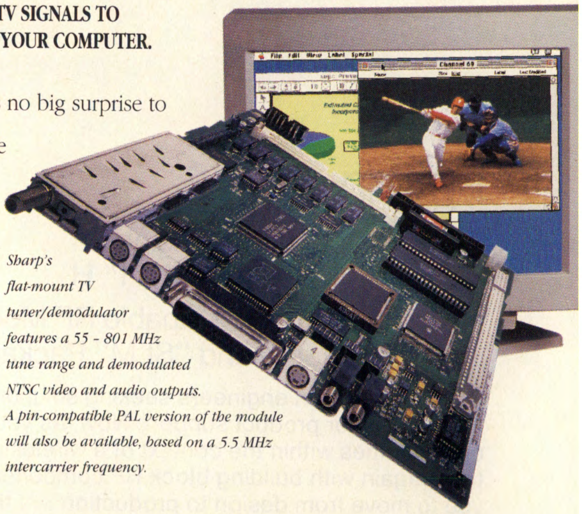
H/V linear and R/L circular polarization LNBs from Sharp provide superior cross-polar discrimination, plus a waterproof, airtight structure which ensures reliable performance in the most severe conditions.

A FINE LINE OF RF COMPONENTS, FINE-TUNED TO YOUR NEEDS.

For years, Sharp has led the field in the miniaturization, noise reduction and seamless integration of RF components, from Ku band LNBs — the industry's smallest — to 22cc DBS tuners. The breadth of Sharp's line and our ability to custom-configure help us meet your size, packaging or power requirements readily.

As North American DBS comes into its own, we'll be there to bring home whole *new* generations of products supporting, for example, the transmission of multiple signals and vast home programming options — the products of *your* imagination.

WE BRING TV SIGNALS TO THE HEART OF YOUR COMPUTER.



Sharp's flat-mount TV tuner/demodulator features a 55 - 801 MHz tune range and demodulated NTSC video and audio outputs. A pin-compatible PAL version of the module will also be available, based on a 5.5 MHz intercarrier frequency.

It should come as no big surprise to see that Sharp, the world's leading manufacturer of RF components, is one of the first to deliver a TV tuner/demodulator in a flat-mount package offering easy integration in system design.

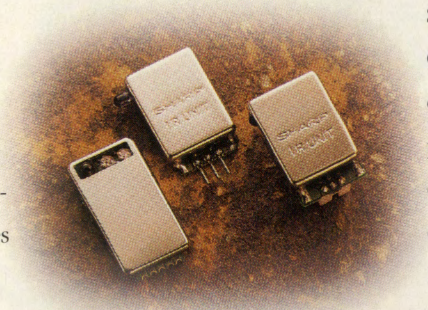
It's the start of a whole new ballgame. Manufacturers of desktop and portable PCs gain immediate entry into video and multimedia — two arenas where tuner size, mounting and power requirements precluded their entry until now. And because this is a complete component module, they can avoid design and development at the subassembly level — always an enviable competitive edge.

less revolution: we're pulling the future out of thin air.

SHARP IS ON YOUR WAVELENGTH IN THE IR SPECTRUM, TOO.

Sharp is the first company to deliver market-ready technology for short-range high-speed infra-red data transmission. Small size, low cost, low power consumption, immunity to EMI and no FCC licensing requirements make Sharp's IR product series a key player in a nearly limitless range of new applications, from personal communications systems to wireless joysticks and interactive TV. Transmission units providing ranges up to 3 meters, higher data rates and 3-volt operation are soon to follow as Sharp continues to evolve its expertise and experience in IR communications.

Sharp's RY5 Series includes receiver, transmitter and bi-directional IR units. They support 38.4kb transmission rates over a 1-meter range, operate from a 5V source, and can be readily interfaced with existing digital circuitry.



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And the fact is, your best connection is Sharp whether your venture calls for high-end digital signal processing, the power to store hundreds of pages of graphics within the smallest possible space, uncompromising flat panel display quality, or a manufacturer uncommonly committed to responding to your own design ideas.

For more information, simply contact your nearest Sharp Electronics Corporation sales office or call us at **1-800-642-0261, Ext. 902.**

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Coherent Radar Performance Estimation
January 23-25, 1994, Atlanta, GA
Radar Signal Processing: Theory, Technology and Applications
January 24-27, 1994, Atlanta, GA
Phased-Array Antenna Design
February 1-4, 1994, Atlanta, GA
Antenna Engineering
February 8-11, 1994, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894–2547.

System Engineering: Principles and Practice
January 10-21, 1994, Los Angeles, CA
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.

Analyzing Communication System Performance
December 13-15, 1993, Orlando, FL
Modern Digital Modulation Techniques
December 13-17, 1993, Orlando, FL
Modern Radar Technology: Monopulse Tracking Techniques and High-Performance Developments
December 13-17, 1993, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994–8522 or (800) 424–9773.

Frequency-Time Signal Processing
January 31-February 4, 1994,
Garmisch-Partenkirchen, Germany
Far-Field, Compact & Near-Field Antenna Measurement Techniques
March 21-24, 1994, Switzerland
Aspects of Modern Military and Commercial Radar
March 21-25, 1994, Switzerland
Cellular and PCS Communications - The Radio Interface
March 21-24, 1994, Davos, Switzerland
Combined Coding and Modulation Techniques
March 24-25, 1994, Davos, Switzerland
Speech and Channel Coding for Mobile Communication
March 28-30, 1994, Davos, Switzerland
Digital Receivers for Satellite and Mobile Communication
March 28-31, 1994, Davos, Switzerland
Personal Communication Networks
March 28-31, 1994, Davos, Switzerland
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122–175–70. Fax: (46) 122–143–47.

VOA and Radio Finland Perform DSB/SSB Broadcast Experiments

A joint experiment is underway to measure and document the HF broadcast audio performance of single sideband suppressed carrier (SSB) and double sideband (DSB) full carrier (A3) transmission. The experiment is designed to evaluate the signal delivery of SSB, compared to the existing method of DSB full carrier "AM" transmission.

The SSB modulation will use a –12 dB suppressed carrier. Transmission will take place from the Radio Finland facilities is Pori, initially on 15120 kHz (DSB) and 15240 kHz (SSB), using antennas directed at 225 degrees azimuth. The primary test receiving location will be in Munich, Germany. Tests began in September 1993 with once-per-day transmission.

To receive information on these tests and the data received, interested parties may contact either of the following:

Dr. Robert Everett
Voice of America
Systems Engineering, B/ESB
Bureau of Broadcasting
US Information Agency
330 Independence Ave., S.W.
Washington, DC 20547

Mr. Esko Huuhka
Finnish Broadcasting Company
HF Planning
OY YLEISRADIO AB
Engineering Division
P.O. Box 20, SF-00521
Helsinki, Finland

EIA Reports U.S. Export and Import Growth — The first quarter of 1993 saw an increase in exports of U.S. electronics products from $37.8 billion in 1992 to $41 billion this year. Imports saw an even greater increase, from $40.4 billion to $46.8 billion. The trade imbalance increased to $5.8 billion. Areas where exports exceeded imports include electron tubes, solid state products, telecommunications, defense communications, industrial equipment and electromedical equipment. EIA president Peter F. McCloskey attributes the increases in both exports and imports to strengthening of the U.S. economy, compared to continuing weakness in Asian and European economic conditions.

New Time and Frequency Services Paper from NIST — Persons interested in time, timekeeping, and precise time and frequency measurements can get a new paper from NIST. *Time and Frequency Services Offered by the National Institute of Standards and Technology* describes in detail shortwave radio stations WWV and WWVH, low-frequency radio station WWVB, satellite time services and the Automated Computer Time Service. It also discusses two services for persons needing the most accurate time and frequency signals possible. Future trends, including the use of GPS and optical communications are also discussed. The paper can be obtained by contacting Sarabeth Moynihan, Div. 104, NIST, Boulder, CO 80303-3328, tel. (303) 497-3237. Ask for paper no. 37-93.

Specifications Outlined for HDTV — The television industry's "Grand Alliance" has announced planned specifications for U.S. digital high definition television. These include and aspect ratio of 16:9, 1125 scanning lines (1080 with active picture information) with 1920 pixels per line, MPEG-2 video compression, and Dolby AC-3 audio compression (recommended). The exact transmission method choice between quadrature amplitude modulation (QAM) and vestigal sideband (VSB) has not been decided. A final proposal from the group may be ready late in the first quarter of 1994.

Bell Labs Makes Self-Focusing Lasers — New experimental devices called zone lasers (Z-lasers) have been developed at AT&T Bell labs Optoelectronic Device Research department. The new lasers require no lenses to focus light on a specific point. Using a vertical cavity geometry, they emit light vertically from their surface instead of horizontally from their edges, as conventional lasers now do. Potential applications include single-chip optical interfaces, chip-to-chip optical interconnections and future optical storage systems.

Earthmovers Guided by GPS — Using GPS receivers supplied by Allen Osborne Associates, the Center for Mapping at Ohio State University has successfully conducted tests where earthmoving equipment blades were guided to within a few centimeters accuracy using GPS as the reference. Such accurate control will lead to advances in automated earthmoving equipment, and greater precision in grading large areas such as airport runways or housing subdivisions. The system uses differential methods, comparing the moving unit to a fixed receiver located at a specified position on the site. This method compensates for the timing errors that are intentionally introduced into GPS.

Automatic Data Collection Lab Established — AIM USA, an organization of Automatic Identification Manufacturers, has helped establish the University of Pittsburgh's Automatic Data Collection (ADC) Laboratory within the Industrial Engineering department. The purpose of the lab is to educate students about ADC, using the latest products; to disseminate accurate and comprehensive data about ADC to practicing engineers and faculty; and to provide high quality laboratory facilities for researchers and industry users. Technologies used in ADC include bar code, magnetic stripe and radio frequency systems.

1994 Virginia Tech Symposium Call for Papers Issued — The 1994 Virginia Tech Symposium on Wireless Personal Communications will be held June 1-3, 1994 in Blacksburg, Virginia. Authors are invited to submit abstracts for technical papers, on topics such as: novel wireless products, applications of DSP, diversity and multiple access, performance simulation, network issues, propagation, support technologies and business issues. 200-400 word abstracts should be sent by December 15, 1993 to: Jeffrey H. Reed, Associate Director, MPRG – Bradley Dept. of Electrical Engineering, 340 Whittemore Hall, VPI&SU, Blacksburg, VA 24060-0111.

U.C.–Berkeley Studies Defense-Related Job Losses — According to a study by the University of California at Berkeley, the number of jobs lost in California due to defense cutbacks is higher than official estimates. The Berkeley study estimates that one-third of the state's total job losses are due to reductions in defense procurement and research, compared to official estimates that these reduction caused 22 percent of 1990-1992 job losses. The study indicates that a modest national economic recovery will not improve California's economic conditions without some kind of regional stimulus.

CATEL Acquired by MERET — CATEL Telecommunications, Inc. has been acquired by MERET Optical Communications, Inc. CATEL is a supplier of fiber optic and coaxial cable communications equipment, while MERET supplies fiber optic audio, video and data communications products.

Signal Technology Corp. Acquires Innowave — Innowave, Inc., a manufacturer of products for defense, space and commercial markets, has been acquired by Signal Technologies, Inc. in a cash transaction. Innowave manufactures RF and microwave detectors, limiters, gain equalizers, phase shifters, circulators, and isolators. Signal's current products involve low cost frequency generation, frequency management, power management and control products.

Radian Corporation in Profiler Radar Agreement With the University of Colorado — A five-year agreement for graduate-level educational activities associated with wind profiler radar systems has been made between Radian Corporation and the University of Colorado at Boulder. Radian will provide the university with a LAP™ -300 wind profiler, along with support, modifications and enhancements. In turn, the University of colorado will contribute basic research activities for commercial wind profiler development. The agreement is made possible by a grant from the National Science Foundation, and supported by Radian's existing Cooperative Research and Development Agreement with the Boulder laboratories of the National Oceanic and Atmospheric Administration (NOAA).

Active Noise Reduction Development Pact — Harris Semiconductor and Noise Cancellation Technologies (NCT), Inc. have signed an agreement to develop and manufacture proprietary silicon ICs for applications of NCT's active noise reduction technology. The technology is an electronic alternative to mechanical noise reduction and previous discrete component methods. NCT's methods involve digital signal processing to analyze noise and drive a high power amplifier and transducer which introduces "anti-noise" to cancel the unwanted noise or vibration.

H-P Opens Interface Standard — Hewlett-Packard Company announces that it has given I/O card suppliers open access to its standard instrument control library's driver-development specification. H-P believes that this license-free opportunity will make it easier for test engineers and technicians to interconnect their hardware.

Bell Atlantic Implements CDPD — Bell Atlantic Mobile (BAM) has selected equipment made by Hughes Network Systems for it cellular digital packet data (CDPD) system. The first equipment will be installed in Pittsburgh by the end of 1993, with commercial service beginning in early 1994. CDPD allows data transmission to mobile users for such applications as travel information, truck shipment information, fax transmission, or many other services.

Cellular Contracts Boost Motorola — Two recent contracts announced by Motorola total more than $100 million. $20 million in orders is for GSM base station equipment to be used in the United Kingdom (Cellnet) and Belgium (Belgacom). Additional contracts for $80 million were announced for digital cellular infrastructure equipment for Austria (PTV), United Arab Emirates (Etisalat) and Pakistan (Pakistan Mobile Communications Ltd.).

Saudis Make $10 Million Satellite Upgrade — Scientific-Atlanta has announced that it will begin shipment of equipment for a $10 million upgrade to Saudi Arabia's domestic satellite system. The improvements will increase channel capacity, and improve reliability and quality through use of digital IDR-type (Intermediate Data Rate) services.

H-P's EEsof Acquisition Approved — Final government approval for Hewlett-Packard Company's acquisition of EEsof, Inc. was received on October 5, clearing the way for H-P to complete the deal and begin merging EEsof's products with H-P's microwave and RF design, analysis and simulation products.

California Microwave Gets Earth Station Contract — A $3.8 million contract has been awarded to the Satellite Transmission Systems (STS) subsidiary of California Microwave. STS will design, build and commission an INTELSAT standard-A International Gateway earth station for the telecommunications authority in Burundi.

Recent ISO 9000 News — Among recent certifications to the ISO 9000 series of quality standards are the following companies: Scientific-Atlanta's Broadband Communications Group, Harris Semiconductor's Singapore and Malaysian test and assembly plants, Dale Electronics' Columbus, Nebraska plant, Analog Devices' Wilmington, Mass. facility, and Siliconix' headquarters facility.

Grace Licenses Copper Paste Technology — Grace Specialty Polymers has announced an agreement for exclusive licensing rights for the manufacture, sale and distribution of Mitsui Mining & Smelting Company's S-5000 solderable conductive paste.The product is used in additive printed circuit board manufacturing.

Flam & Russell Opens First European Office — A new branch office located near Eindhoven, The Netherlands, is Flam & Russell's first international expansion of operations. The office will provide RCS and antenna measurement applications support for the company's European customers. The new office can be reached by telephone at +31-40-63 11 79.

W-J Announces Contract — Watkins-Johnson Company announces that it has been awarded contracts valued at more than $12 million by Warner-Robins Air Logistics Center. The contracts cover ground support equipment, repair services, and provisioning spares for the APR-46 Threat Avoidance Receiving System used aboard the AC-130 gunship and MC-130 Combat Talon aircraft.

RF Fires Industry's New Foundries

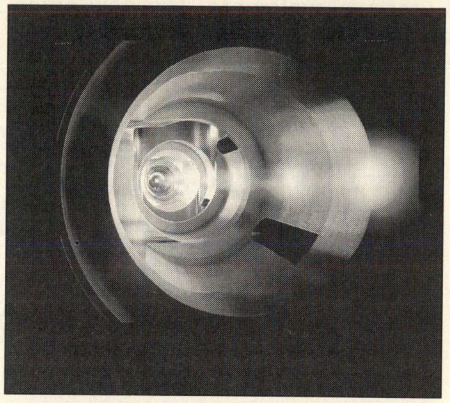
By Andy Kellett
Technical Editor

RF is a tool of industry in many respects; as a communications tool it helps coordinate the efforts and materials of industry; the topic of this report is its other use — as a source of power. RF finds use in induction heating, semiconductor processing, analytical chemistry, and even waste remediation. While these markets are not as large as the much talked-about wireless communications markets, their futures may be just as bright.

In an *Electronic News* article by Jeff Dorsch, Motorola's senior V.P. and assistant G.M., Murray Goldman, was quoted as saying semiconductor manufacturers may have to open more than sixty advanced wafer fabrication facilities in the next seven years to keep up with the projected growth in the semiconductor market. RF technology is used in a number of semiconductor processes.

"There are many different deposition techniques, and which one you use depends on your requirements," says Tam Pandhumsoporn, Applications Lab Manager at Alcatel. According to Pandhumsoporn, RF sputtering is the primary means of producing inductive disk heads, and is used to a smaller extent in semiconductor wafer processing. RF sputtering uses an RF-excited plasma to knock atoms off a target and onto a surface to be coated. This process produces both high quality conducting and dielectric films. The process is also used in the production of color flat-panel displays, where the plasmas are used both to deposit and etch away material. "Down the road you will hang your TV on the wall," says Christopher Ben, General Manger of RF Power Products (RFPP), "Of course right now, the flat panel display market is just starting to expand with their use in laptop computers. "

Another semiconductor process which uses RF is ion implantation. In this process, an RF field pushes ions down a linear accelerator and deep into a material to be implanted. This process is being used in the fabrication of the newest memory chips. "Ion implantation is of interest for a process called retrograde wells," says Stuart Denholm of Eaton's Semiconductor Equipment Division, "As you get to higher and higher density memories, cells interact with adjacent cells, and by putting ions deeply below them you can prevent this interaction." As semiconductor geometries shrink, another problem that becomes more acute is particle contamination. People within the industry claim that particles as small as half a micron can be removed by a new wet-cleaning process called megasonics. John Quinn, V.P. of Marketing for Intech says his company sells RF sources at 1.7 to 3 MHz to companies making equipment for this new process.



A modulated ion beam is accelerated by RF fields in an accelerator built by Eaton.

Many industries depend on analytical chemists to tell them, (to within parts per billion), what is in their samples. RF sources see plenty of use in the field of analytical chemistry. According to Dr. Tom Barnard, Director of Inorganic R&D at Perkin-Elmer Corporation, the elemental-analysis equipment market is around $800 million per year and growing, with most of that business accounted for in three types of analysis which use RF.

Atomic absorption (AA) , atomic emission and mass spectroscopy each use RF in one way or another to identify the concentrations of different species of atoms in a sample. Both inductively-coupled-plasma mass-spectroscopy (ICP-MS) and atomic emission spectroscopy use RF to generate plasmas which break samples down into free atoms. ICP-MS uses another RF source to drive the quardupole mass filter, which directs ions of a given mass to a detector. Atomic absorption (AA) spectroscopy depends on a source of photons of known wavelength to determine the types and concentrations of atomic species in a sample. Perkin-Elmer has developed a line of RF-powered, electrodeless discharge lamps for use in their AA equipment.

One of the oldest industrial uses for RF is induction heating. RF sources operating from below 100 kHz up to 1 MHz or so pump out kilowatts of power to anneal and even melt metal, to weld plastics and to cure glues. RF sputtering processes similar to those used in semiconductor manufacture are used to apply thin films of metals and other materials to a number of products. Car trim and CDs can both be metallized using RF sputtering processes.

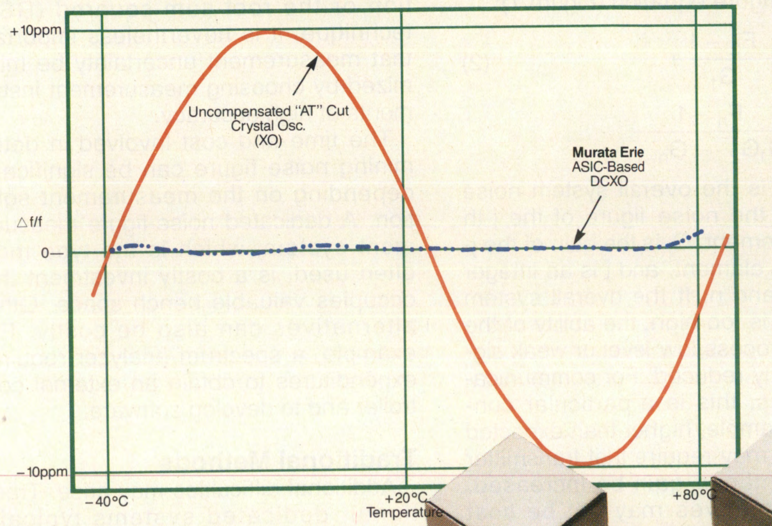
Waste remediation doesn't pop into one's head when talking about applications of RF, but it is an application into which RFPP sells RF sources. According to RFPP's Ben, an RFPP customer uses 25 kW RF sources to drive antennas placed within contaminated soil. As the RF energy heats the soil, the volatile chemicals diffuse to the surface where they are removed by large vacuums.

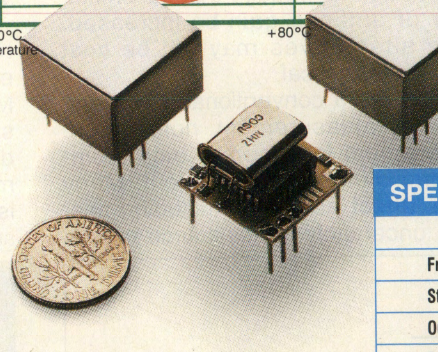
"We're paying a lot of money for a device that works screamingly well up at 150 MHz, and we're using it at 3 MHz because we have no alternative, " says Intech's Quinn. This sentiment was repeated by several other manufacturers of RF industrial equipment. Senior Vice-President at ENI, Ed Maier said that in addition to lower cost and better availability he would like to see, "more willingness on the part of vendors to work closely with customers on application specific requirements." In addition to the lack of components targeted at their applications, these manufacturers also face some of the more traditional design concerns. Compactness is important to semiconductor equipment manufacturers because cleanroom space is so valuable. Power conservation is also important because heat dissipation is expensive and bulky, and because the power bill for several multi-kilowatt devices can be rather large.

RF has proven itself to be more than just a medium for communication; it continues to be an important tool in other industries. *RF*

No compromise necessary.

Murata Erie's new ASIC-based DCXO





Excellent frequency stability has been a fact of life with crystal oscillators for years. So has small size (a relative term), low cost and flexibility. **But**, did you ever try to get everything that you needed in a single oscillator/package? No way. One, maybe two desirable characteristics but certainly not all. You had to compromise.

Now, compromise no more.

Murata Erie has the solution...our new ASIC-based, digitally compensated crystal oscillator is exceptionally stable, truly tiny, low in cost and extremely flexible. It is basically everything that you always wanted in a crystal oscillator but always had to compromise to get.

Check some of these features and specifications:

- To ±0.25ppm stability
- Less than 5.0 cubic centimeters volume
- 100 mW max power
- Built-in interface for reconfiguration capability and software
- Single +5VDC power supply

SPECIFICATIONS	DC2210 AH
Frequency Range	10MHz to 25MHz
Stability over −40°C to +85°C	±0.5ppm
Output	"HC" CMOS
Power Dissipation	+5V @ 15mA
Short Term Stability	1×10E−9 @ T=1 sec.
Package Size	0.79" × 0.79" × 0.45", PC PINS

Find out more about how you can take the compromise out of your crystal oscillator specifications. Write for complete details on Murata Erie's new ASIC-based DCXO or call 1-800-831-9172, FAX: 1-404-684-1541.

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INFO/CARD 17

Spectrum-Analyzer-Based System Simplifies Noise Figure Measurement

By Carla Slater
Hewlett-Packard Co.

Designers and manufacturers of receiver systems and components routinely face the task of noise figure measurement. This task is not an easy one, and it frequently raises concerns regarding manufacturing costs and measurement accuracy. Today, however, the integrated measurement capabilities contained within single pieces of equipment can lower total equipment costs and can make it easier to achieve measurement accuracy. In particular, a spectrum-analyzer-based system offers a convenient way to measure noise figure as well as other important receiver characteristics.

The noise figure is defined in linear terms as

$$F = \left.\frac{(S/N)_{in}}{(S/N)_{out}}\right|_{T=290\ K} \qquad (1)$$

where $(S/N)_{in}$ is the signal-to-noise ratio at the input of the device or system under consideration, $(S/N)_{out}$ is the signal-to-noise ratio at the output, and T is the temperature of the system. Noise figure may also be described in units of dB: F_{dB}=10logF. (To distinguish between noise figure in linear and logarithmic units, F in linear units is sometimes called "noise factor.") In the design and manufacture of active components and systems, it is crucial to determine the noise figure of individual devices. The noise figure of these devices in turn determines the noise figure of the overall system. We describe this by the cascade noise figure equation (Figure 1):

$$F_{total} = F_1 + \frac{F_2 - 1}{G_1} + \cdots + \frac{F_n - 1}{G_1 G_2 \cdots G_{n-1}} \qquad (2)$$

where F_{total} is the overall system noise figure, F_j is the noise figure of the j-th individual element, G_j is the gain of the j-th individual element, and j is an integer between 1 and n. If the overall system noise figure is too high, the ability of the system to process low level or weak signals is greatly reduced. For communication systems, this is a particular concern. For example, higher than expected noise levels may require that transmitter power or antenna gain be increased. These alternatives may not be cost effective or practical.

Compared to conventional power and frequency measurement, noise figure measurement is generally more complex and difficult to implement. Many factors affect accuracy, including mismatch uncertainty, noise-source excess noise ratio (ENR) uncertainty, and measurement-system uncertainty. [The excess noise ratio is the noise power ratio of the noise source (on/off) normalized to kTB.] Although these uncertainties cannot be combined simply by addition or the root-sum-squared (RSS) technique, it is nevertheless important that measurement uncertainty be minimized by choosing measurement instruments of high accuracy.

The time and cost involved in determining noise figure can be significant, depending on the measurement solution. A dedicated noise-figure measurement system, which is the type most often used, is a costly investment that occupies valuable bench space. Other alternatives can also be costly. For example, a spectrum analyzer requires expenditures to obtain an external controller and to develop software.

Traditional Methods

Additional difficulties may arise. Traditional, dedicated systems typically employ a fixed bandwidth of several MHz and are thus not suitable for measuring noise figure of narrow band devices. Errors are introduced when the measurement bandwidth of the system is wider than the bandwidth of the measured device. These errors can occur in

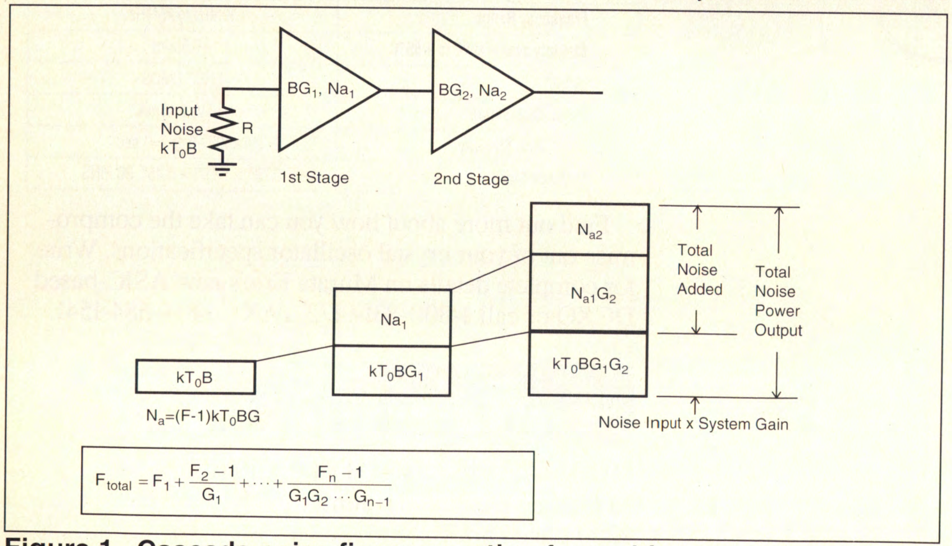


Figure 1. Cascade noise figure equation for multi-stage system.

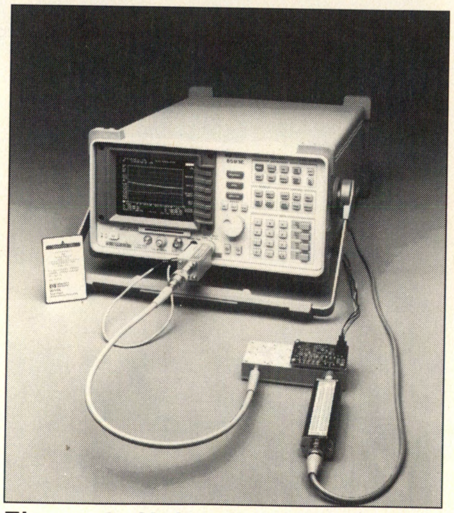


Figure 2. Noise figure measurement set-up

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

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

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

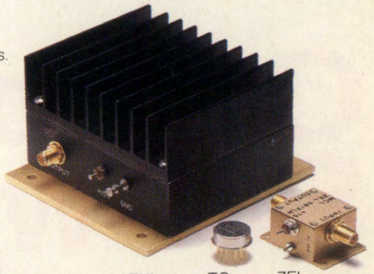
Available from stock, priced from $199 with better than 1.5dB NF performance, there's lots to shout about.

SPECIFICATIONS

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

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

*TO-8 includes test fixture loss.



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finding new ways ...
setting higher standards

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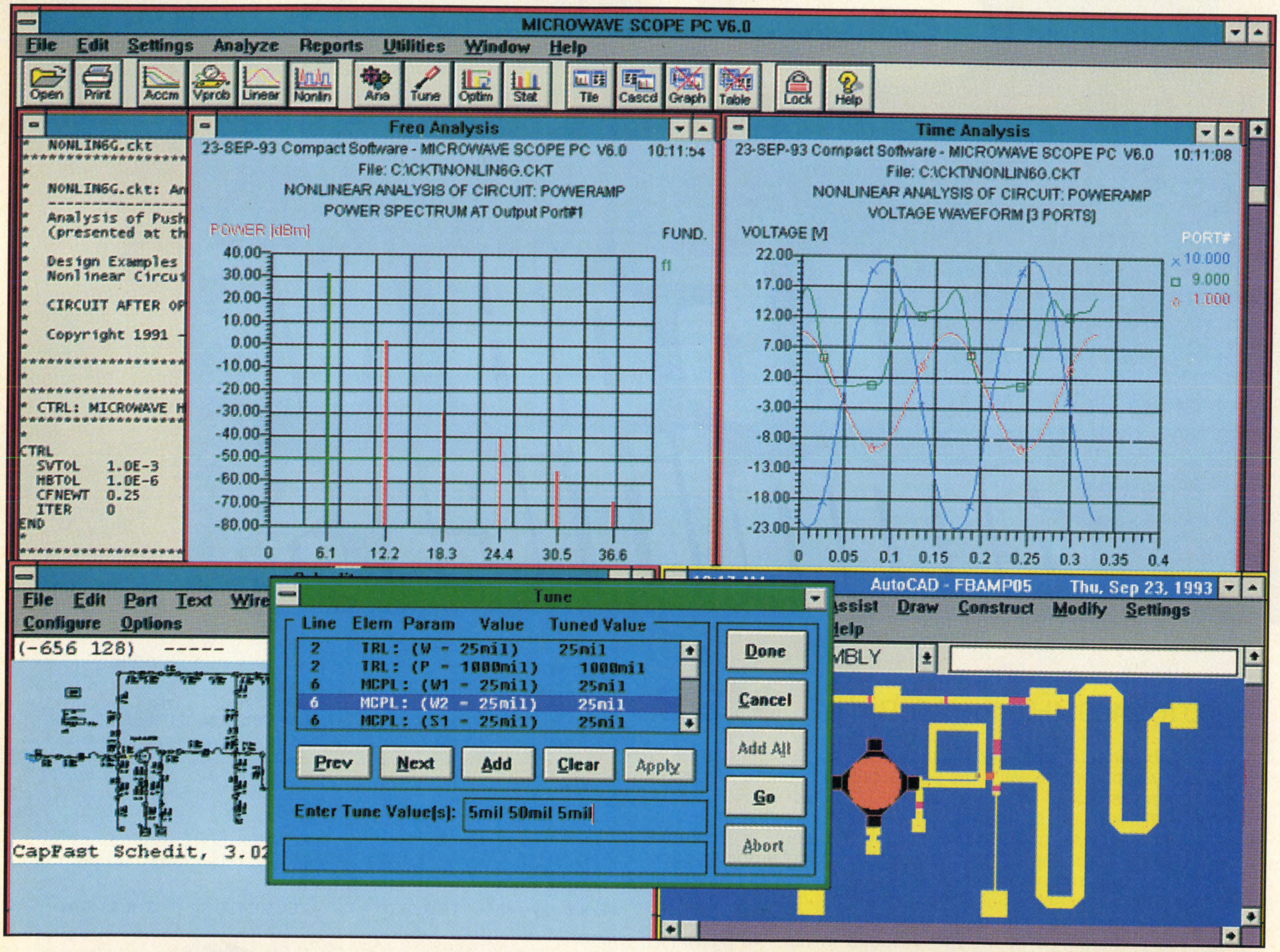

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both the noise figure and the gain measurement if the bandwidth used for calibration differs from the bandwidth used for measurement.

Errors in the gain measurement can be corrected by measuring the two bandwidths (device bandwidth and system bandwidth) using a signal generator and the measurement system as a detector. The following equation calculates the actual gain:

$$G_D = \frac{BW_S}{BW_D} G_M \tag{3}$$

where G_D is the device gain, BW_S is the noise bandwidth of the measurement system, BW_D is the noise bandwidth of the device under test, and G_M is the measured gain.

Errors in the noise figure measurement can be more difficult to overcome. If the device has high gain—several orders of magnitude greater than the bandwidth ratio—the error can be reduced. [The cascade noise figure equation (1) for a two-stage system can be arranged, as shown, to solve for the first stage (device) noise figure: $F_1 = F_{total} - (F_2 - 1)/G_1$. From this equation, it is apparent that if the gain of the device (G_1) is large, the second stage contribution will be reduced. Overall, this results in a reduction in error.]

If the device does not have high gain, error can be minimized by inserting, before the measurement system, a narrow band preamplifier followed by a bandpass filter. In this case, the filter center frequency becomes the measurement frequency and its bandwidth must be narrower than the device bandwidth. Also, the preamplifier must have sufficient gain to overcome the effective total noise power reduction caused by its narrow bandwidth, as we suggest in the following equation:

$$G_P = 10\log\left(\frac{BW_S}{BW_F}\right) + 10 \text{ dB} \tag{4}$$

where G_P is the gain of the preamplifier, BW_S is the bandwidth of the measurement system, and BW_F is the bandwidth of the filter.

Correcting for the errors associated with this technique requires a lengthy calculation and knowledge of the actual filter pass-band shape and the out-of-band noise levels. An alternative measurement system provides a better solution to the difficulties described above.

An Alternative Method

For some time, industry has used spectrum analyzers to measure noise figure, largely because these instruments also can perform the multiple measurements so often required in a test system. Although this leverage of spectrum analyzer measurement capability offers an advantage, disadvantages still exist. Generating the software needed to automatically measure noise figure is a lengthy procedure that requires a great deal of knowledge of the factors involved in noise figure measurement error correction. Accuracy, as determined by measurement uncertainty, remains a major concern.

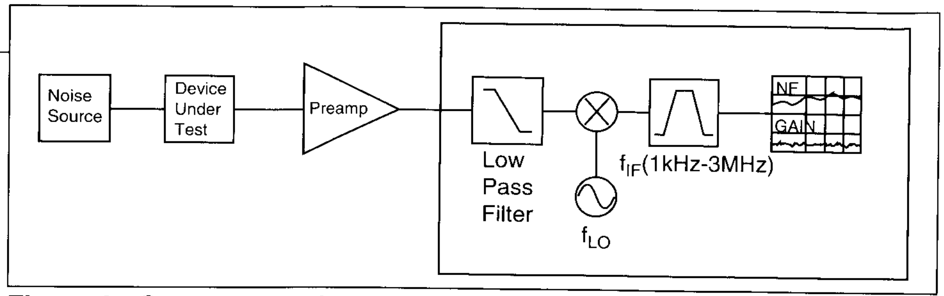

Figure 3. Spectrum-analyzer-based noise figure measurement system.

In a different approach, we integrate a portable spectrum analyzer with a software measurement "personality" for noise figure measurements (Figure 2). The software, provided as a downloadable program on a ROM card, controls the spectrum analyzer's measurement functions and automatically calculates the noise figure and gain of a device under test. These values are displayed as a function of frequency, offering swept noise figure and gain measurements from 10 MHz to 2.9 GHz. Noise figure can be measured from 0 to 20 dB and gain from 0 to 40 dB.

The complete system (Figure 3) requires a preamplifier, which is connected at the input to reduce system noise figure, and a diode noise source driven by a 28-volt modulated signal from the spectrum analyzer. A menu-driven interface makes this particularly easy to use. If measurement speed is an issue, as it often is, single-point measurements can be made at the frequencies of interest. This feature, called "one-point measurement mode," allows users to make measurements at a single point and utilizes the capability to display PASS/FAIL information to speed up the measurement process (Figure 4).

This system uses the Y-factor measurement technique, in which measurements are made with the noise source turned on and off. The Y-factor is the ratio of the device's output noise power with the noise source on to the device's output noise power with the noise source off. Noise figure and Y-factor are related by the following equation:

$$F = \frac{ENR}{Y - 1} \tag{5}$$

where F is the noise factor (noise figure in linear units), ENR is the excess noise ratio, and Y is the Y-factor.

The measurement system is calibrated by measuring the noise level with the device removed from the measurement set-up. This allows calculation of the system noise figure. The device is then reconnected and a second measurement is taken to calculate the total noise figure. From the total noise figure, the effect of the system noise figure is removed and the resulting device noise figure is displayed. The cascade noise figure equation shown here (in linear terms) provides this calculation:

$$NF_T = NF_D + \frac{NF_S - 1}{G_D} \tag{6}$$

where NF_T is the total noise figure, NF_D is the device noise figure, NF_S is the system noise figure, and G_D is the device gain.

The gain of the device is calculated by taking the ratio of (a) the difference between the noise power measured with the noise source turned on and turned off (with the device inserted in the measurement path), and (b) the difference between the noise power measured with the noise source turned on and turned off during calibration (with the device removed from the measurement path). In other words,

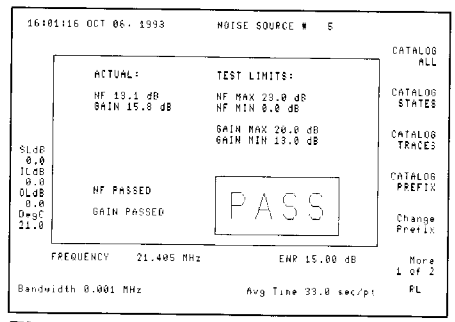

Figure 4. Noise figure and gain measured in "one-point measurement" mode.

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$$G_D = \frac{(N_D)_{on} - (N_D)_{off}}{(N_C)_{on} - (N_C)_{off}} \quad (7)$$

Calibration is not required to make a noise figure measurement; however, without calibration, a gain measurement cannot be made. As can be seen from Equation (6), in such cases, the noise figure measurement accuracy will not be sacrificed if the device has high gain.

Our spectrum-analyzer-based system offers several distinct advantages. It is the only system available with selective measurement bandwidths in the range of 1 kHz to 3 MHz. Because of this, narrow band devices can be measured directly. We demonstrate this ability with an example. Using the components of a cellular base station front end, a 1-MHz bandpass cavity filter and a broadband low noise amplifier, we complete an evaluation. Ordinarily, the filter would precede the broadband amplifier to remove any unwanted signals that might otherwise pass through the system (Figure 5), but here a problem arises in measuring noise figure.

Measurement Bandwidth

The measurement of interest is the pass band noise figure. However, if a system with a measurement bandwidth wider than the device bandwidth is used, the broadband noise out of the amplifier also will be measured. This out-of-band noise adds a significant amount of error, distorting the "true" pass band noise figure. To achieve the true pass band noise figure, we must use a measurement bandwidth narrower than the device bandwidth, as we can with our spectrum analyzer. Table 1 shows the results of the filter and amplifier measured using the spectrum analyzer system at varying bandwidths and a noise figure meter with a fixed 4-MHz measurement bandwidth. (The measurements were made at the device bandwidth center frequencies, which were measured first using the spectrum analyzer with a built-in tracking generator.)

As we expected, the results show distinct differences between measurement systems. The true passband noise figure is measured with the narrow bandwidths. In this case,

$$F_T = F_M, \quad (BW_S < BW_D) \quad (8)$$

where F_T is the true noise figure, F_M is the measured noise figure, BW_S is the system bandwidth, and BW_D is the device bandwidth.

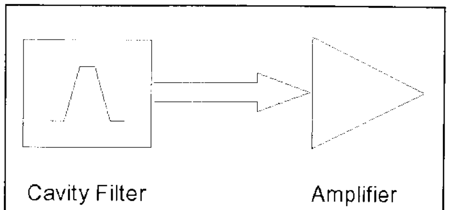

Figure 5. Narrow band cavity filter precedes broadband amplifier to remove unwanted signals.

Using a 4-MHz measurement bandwidth (four times the device bandwidth) results in measurement error. In this case, we can correct for the measurement error and approximate the true pass band noise figure in linear terms using the following equation [This equation assumes that the device has high gain, the pass band shape is rectangular, and the filter mismatch effects on out-of-band noise are ignored]:

$$F_T = \frac{BW_D}{BW_S} F_M, \quad (BW_D < BW_S) \quad (9)$$

To prove this, we use the 14.1 dB measured in the 4-MHz bandwidth to calculate the true noise figure according to Equation (9):

$$F_T = \frac{1\ \text{MHz}}{4\ \text{MHz}} (10^{14.1/10}) = 6.4 \quad (10)$$

or

$$F_T = \frac{1\ \text{MHz}}{4\ \text{MHz}} (10^{14.1/10}) = 6.4 \quad (11)$$

As we see in Table 1, with narrow measurement bandwidths, the true noise figure of 8.0 dB is measured. The error in the gain measurement that occurs with the 4 MHz bandwidth can be attributed to the different bandwidths used for calibration and for measurement. Using Equation (2), we can approximate the true gain as

$$G_D = \frac{4\ \text{MHz}}{1\ \text{MHz}} (10^{18.2/10}) = 264.3 \quad (12)$$

or

$$(G_D)_{dB} = 10 \log(264.3) = 24.2\ \text{dB} \quad (13)$$

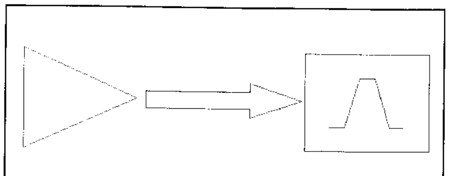

Figure 6. Narrow band filter follows broadband amplifier to measure true noise figure.

Figure 6 illustrates a case in which the narrow band filter follows the broadband amplifier in the system. Here, the out-of-band noise does not contribute to the measured noise figure because it is outside the filter's pass band. As a result, the "true" noise figure can be measured. The measurements for this case are shown in Table 2. For both measurement systems the results are similar, as expected, and the gain values remain fairly consistent.

Narrow measurement bandwidths are also required for making low frequency measurements. The presence of 0 Hz LO feed-through in the spectrum analyzer requires that a sufficiently narrow measurement bandwidth be used to reject this signal at the lowest measurement frequency. The measurement bandwidth must be about 1/30, or less, of the lowest measurement frequency. For a 3-kHz measurement bandwidth, the lower frequency measurements are limited to approximately 100 kHz.

Measurement Speed and Repeatability

Measurement speed can become an

Measurement System	Measurement Bandwidth (MHz)	Measured Noise Figure (dB)	Measured Gain (dB)
Spectrum Analyzer Based System	3	11.7	20.8
	1	9.2	23.4
	0.3	8.2	24.7
	0.1	8.0	25.0
	0.03	8.0	25.0
	0.01	8.0	25.0
Noise Figure Meter	4	14.1	18.2

Table 1. Noise figure and gain for amplifier with filter in front of it.

issue specifically when narrow measurement bandwidths are used. If measurement repeatability is not to be affected, then a reduction in the measurement bandwidth means an increase in the measurement time. [Measurement repeatability (variation from measurement to measurement due to jitter) is dependent on the measurement bandwidth and measurement time. These are related by the proportionality $r_{dB} \propto 1/\sqrt{(BW_S)t}$, where r_{dB} is the repeatability in dB, BW_S is the system measurement bandwidth, and t is the effective averaging time of the noise signal. If the measurement bandwidth decreases, the measurement time would need to increase if the measurement repeatability is not to be sacrificed.]

To understand the trade-off between measurement time and repeatability, we turn to another feature of the spectrum

Measurement System	Measurement Bandwidth (MHz)	Measured Noise Figure (dB)	Measured Gain (dB)
Spectrum Analyzer Based System	3	5.8	20.3
	1	5.6	23.2
	0.3	5.5	24.7
	0.1	5.6	24.9
	0.03	5.6	24.9
	0.01	5.7	25.0
Noise Figure Meter	4	5.8	18.6

Table 2.. Noise figure and gain for amplifier followed by a filter.

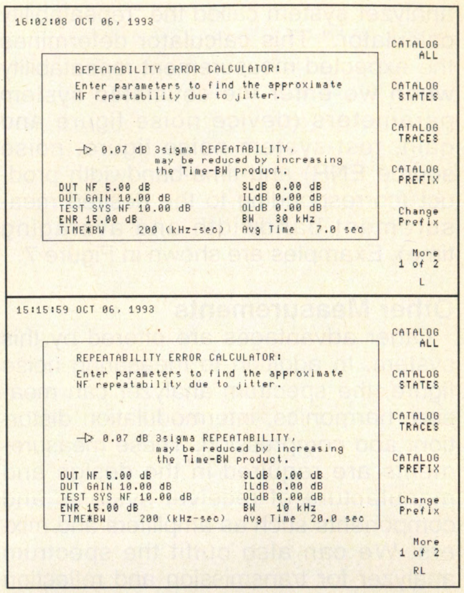

Figure 7. Calculation showing equal repeatability (0.07 dB) for measurement bandwidths of 30 kHz (top), and 10 kHz (bottom). Reduced measurement bandwidth yields longer test times.

analyzer system called the "repeatability calculator." This calculator determines the expected measurement repeatability when we enter the estimated system parameters (device noise figure and gain, test-system noise figure, noise source ENR) and time-bandwidth product (corresponding to the desired measurement bandwidth and averaging time). Examples are shown in Figure 7.

Other Measurements

Other advantages are offered by this system. In addition to measuring noise figure, the spectrum analyzer can measure harmonics, intermodulation distortion, and compression. These measurements are required in the design and manufacture of receiver systems and components such as amplifiers and mixers. We can also outfit the spectrum analyzer for transmission and reflection measurements. The gain/insertion loss, center frequency, and n-dB bandwidths can be measured easily. Some of these measurements are made by the press of a single button.

The spectrum analyzer's ability to measure spurious responses can become important in noise figure measurement. Often stray signals get coupled into a noise figure measurement and affect the measurement results. These can be easily detected by switching the system from noise figure measurement to spectrum analyzer mode. Once a signal has been identified and its frequency is known, noise figure can be measured at other appropriate frequencies. Detection of spurious signals ensures measurement accuracy.

Because the spectrum analyzer has a wide range of measurement capabilities, it is particularly attractive when cost is an issue. The capabilities can be added to the system relatively inexpensively. For a manufacturer of UHF remote satellite amplifier systems, cost was the main factor in choosing the spectrum-analyzer-based system. To fully test the amplifier system, which incorporates both the transmit and receive sections used with UHF satellite communication systems, multiple measurements were required. Tests required on the transmit section were transmit power and gain. On the receive section, noise figure and gain had to be tested. Out-of-band rejection and VSWR were tested for both sections.

While harmonics could be measured by the spectrum analyzer alone, transmit power, out-of-band rejection, and VSWR (calculated from the measured return loss) could not. Adding a built-in tracking generator as well as the noise figure and gain measurement capability allowed all measurements to be made. Figures 8 through 10 present a sampling of these measurements made with the spectrum analyzer system. Traditionally a number of instruments would have been used to make these measurements. However, in this case the spectrum-analyzer-based system limited the number of instruments required and provided the most economical solution.

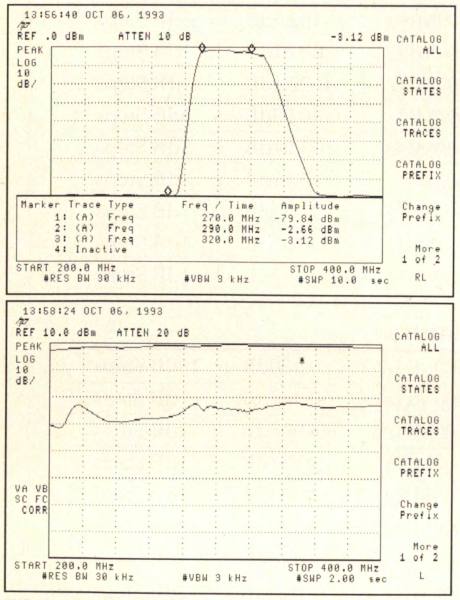

Figure 8. Transmit power (a) and return loss (b) measured on the amplifier system's transmit section.

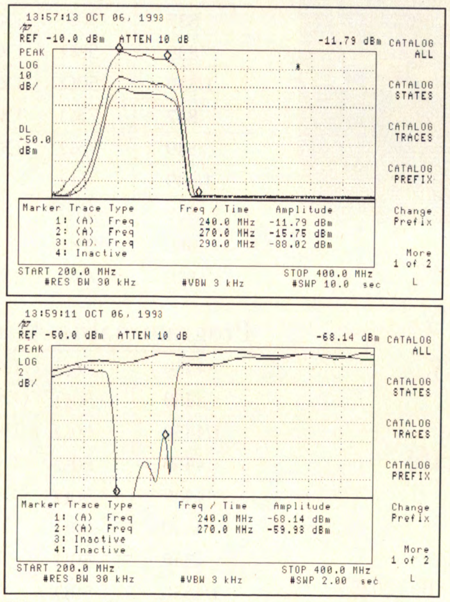

Figure 9. Out-of-band rejection and gain (a) as well as return loss (b) measured on the amplifier system's receive section.

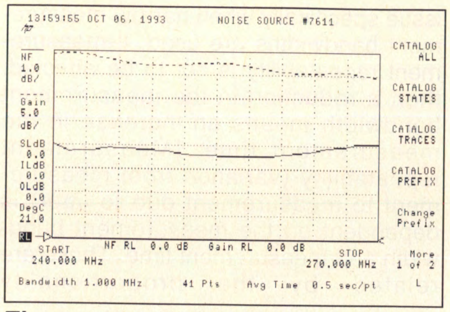

Figure 10. Noise figure and gain measured on the amplifer system's receive section.

Conclusion

Both cost and measurement accuracy are major concerns for engineers who make measurements. However, with the versatility of a low-cost, integrated measurement system and the capability to accurately and directly measure narrow band devices, the spectrum-analyzer-based system offers an attractive solution. *RF*

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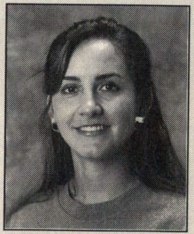
3. Product Note 85719A-1, Hewlett-Packard Company, 1992.

About the Author



Carla Slater is currently a Product Marketing Engineer for the Microwave Instrument Division of Hewlett-Packard in Santa Rosa, California. She is currently supporting the scalar network analyzers and power meter products. Prior to her Product Marketing position, Ms. Slater was a Production Engineer supporting MMS (Modular Measurement System) spectrum analyzers. Ms. Slater received the BSEE from the University of Arizona. For more infprmation on this sytem, write: Hewlett-Packard Co., MID, 1400 Fountaingrove Pkwy., M/S 4USK, Santa Rosa, CA 95403-1799.

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Single Tone Intermodulation Testing

By Steve Winder
BT Laboratories

If a signal is applied to a non-linear device, harmonics of the original signal are produced at the output. If two or more signals are applied, intermodulation products are produced due to the signals mixing. Intermodulation testing is carried out to determine the degree of non-linearity. The results of this testing are expressed in terms of the second and third order intermodulation intercept points. These are the imaginary amplitudes at which the fundamental and intermodulation product signal levels are equal. If devices are compared, the more linear one will have higher amplitude intermodulation intercept points.

Two tone testing is a widely used method for measuring intermodulation. Another method is to apply the rule of thumb -- the third order intercept point is 10 to 15 dB above the 1 dB compression point. Presented here is an alternative method, that of single tone testing. Mathematical analysis will be presented later to show that there is a direct relationship between the amplitude of harmonics and the intermodulation intercept points. Single tone testing cannot be applied to frequency converting systems such as radio receivers; however, for amplifiers the test can be applied, and it provides a cheap, quick and simple solution.

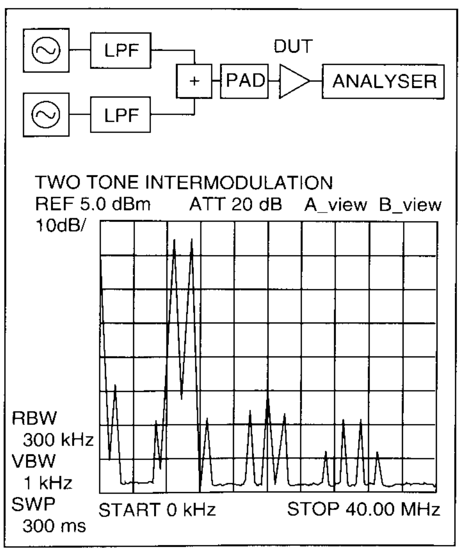


Figure 1. Two-tone intermodulation test set-up and results

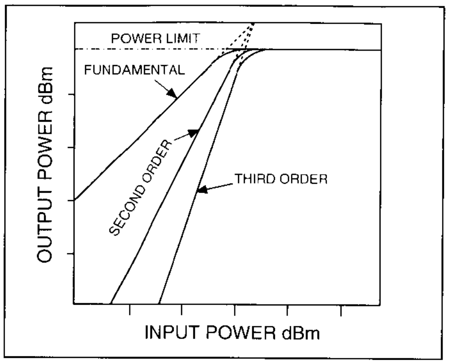


Figure 2. Schematic illustration of second- and third-order intercept points.

Two Tone Testing

The standard method of intermodulation testing requires two oscillators and a spectrum analyzer. Figure 1 shows a test set-up that can be used; variations on this may be found in texts. The oscillators must be isolated from each other so that intermodulation in their output stages does not occur. An attenuator placed after the directional coupler provides a constant 50 ohm termination, preventing reflection of power from one oscillator into the other. Any impedance mismatch into the active device causes reflections, but the attenuator reduces their amplitude. Intermodulation in the oscillator output stage is small compared with the DUT's.

Intermodulation ratio is the difference in dB between the wanted signal level, P_{out}, and the intermodulation products, measured at the output. This is written as IMR_2 or IMR_3 depending on whether the second or third order products are being considered. The second and third order intercept points, IP_2 and IP_3, are found utilizing the following equations:

$$IP_2 = P_{out} + IMR_2 \quad (1)$$
$$IP_3 = P_{out} + 0.5\, IMR_3 \quad (2)$$

To visualize an explanation for these formulae, a graph can be drawn (Figure 2). The fundamental intermodulation products are all straight lines when the graph axes are scaled in dB's. The fundamental has a slope of one and is dependent on V_{in}. Second order products have a slope of two because they are dependent on V_{in} . Third order products have a slope of three due to their dependence on V_{in} . Intercept points are found by extending these lines and reading their crossing point off the graph axis.

Single Tone Testing

There are times when two oscillators are not available. Fortunately, for amplifiers and other "linear" devices, it is possible to carry out tests to determine intermodulation intercept points using just one oscillator. A typical test arrangement is shown in Figure 3. The arrangement is simple and the measuring equipment does not necessarily have to be a spectrum analyzer; a frequency selective voltmeter would also serve the purpose since only harmonics of the applied signal are to be measured. Figure 3 shows the spectrum obtained by testing the same device as used in the two tone test.

Proof

To justify use of such a simple test arrangement some mathematics is necessary. Trigonometric formulae will be given before they are used.

Consider an input signal that is applied to an amplifying device,

$$V_{in} = A\cos xt + B\cos yt \quad (3)$$

The device will have an output, V_{out}, given by the transfer function:

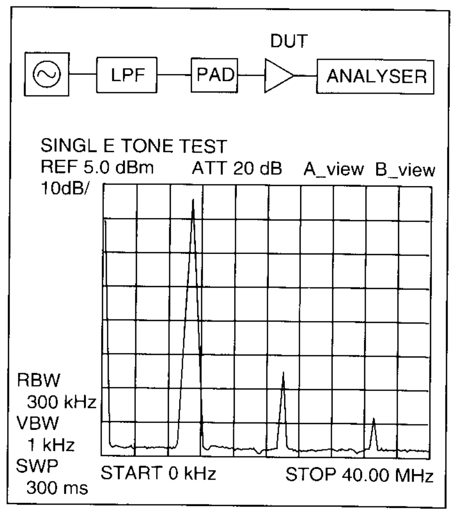
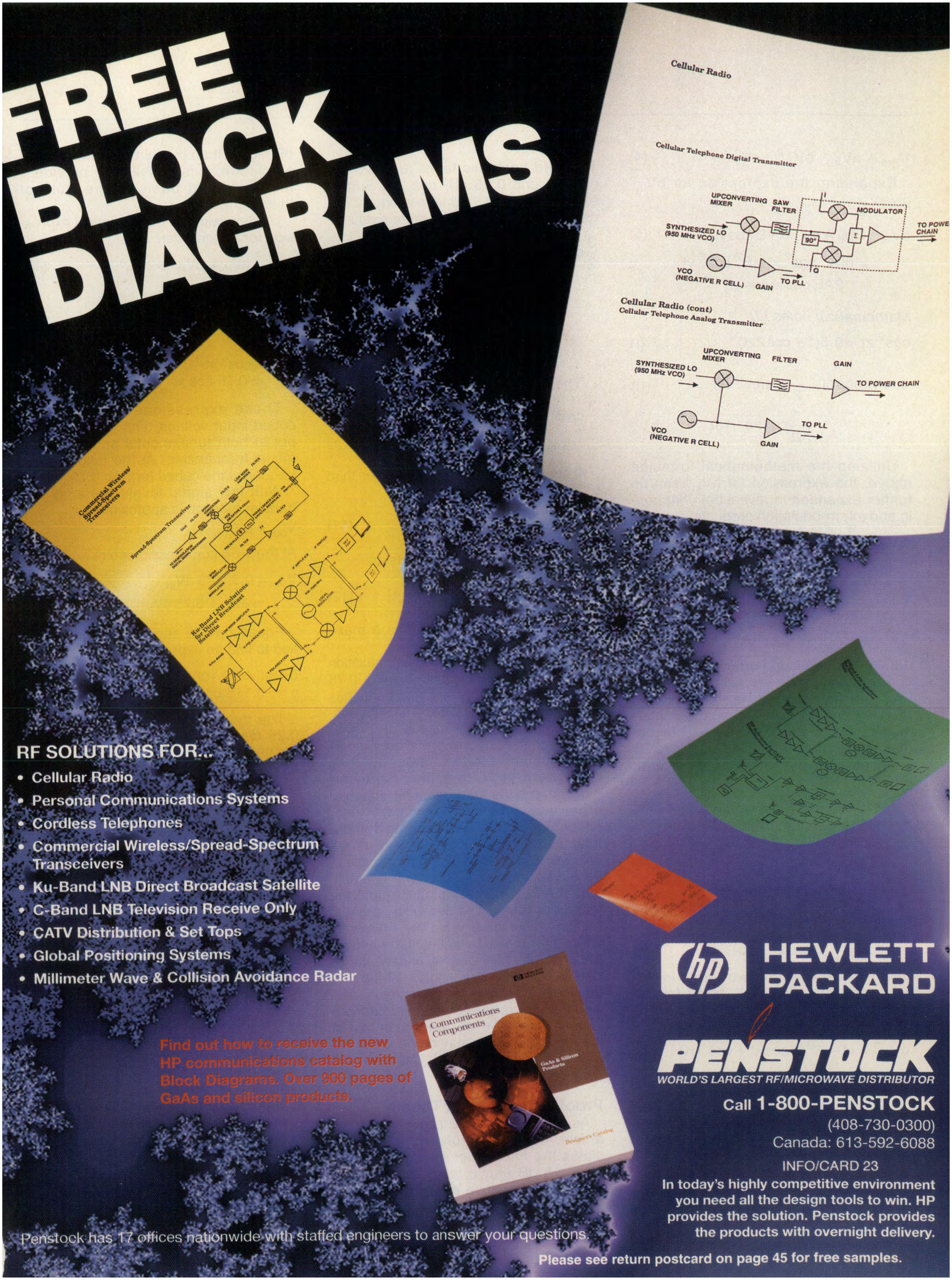


Figure 3. Single-tone test set-up and results



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$$V_{out} = aV_{in} + bV_{in}^2 + cV_{in}^3 + \cdots \quad (4)$$

Expanding the expression for bV_{in}^2 gives:

$$bV_{in}^2 = b[(A\cos xt + B\cos yt)^2] \quad (5)$$
$$= b[A^2\cos^2 xt + B^2\cos^2 yt + 2AB\cos xt\cos yt]$$

Mathematical notes I,II:

$$\cos^2 zt = 0.5(1+\cos 2zt) \quad (I)$$

$$\cos mt\cos nt = \frac{1}{2}\cos(m+n)t + \frac{1}{2}\cos(m-n)t \quad (II)$$

Utilizing the mathematical formulae above, the expression for bV_{in}^2 can be further expanded to reveal both harmonic and intermodulation products:

$$bV_{in}^2 = b[\frac{1}{2}A^2(1+\cos 2xt) + \frac{1}{2}B^2(1+\cos 2yt) + AB\cos(x-y)t + AB\cos(x+y)t] \quad (6)$$

If A = B the second order intermodulation products have a maximum level of bA^2 and the second harmonic signal level is 0.5 bA^2. The second order intermodulation products will be 6dB higher than the second harmonic signal level.

Let us also expand the cubed term, cV_{in}^3:

$$cV_{in}^3 = c[(A\cos xt + B\cos yt)^3] \quad (7)$$
$$= c[A^3\cos^3 xt + B^3\cos^3 yt + 3AB^2\cos xt\cos^2 yt + 3A^2B\cos^2 xt\cos yt]$$

Mathematical note III:

$$\cos^3 zt = \frac{1}{4}(3\cos zt + \cos 3zt) \quad (III)$$

Using *III* for the first two terms in the expansion of equation 7, and *I* for the last two yields,

$$cV_{in}^3 = c[\frac{1}{4}A^3(3\cos xt + \cos 3xt) + \frac{1}{4}B^3(3\cos yt + \cos 3yt) + \frac{3}{2}AB^2\cos xt(1+\cos 2yt) + \frac{3}{2}A^2B(1+\cos 2xt)\cos yt] \quad (8)$$

The third harmonic terms are 0.25 cA^3 cos 3xt and 0.25 cB^3 cos 3yt. Further expansion of the third term in equation 8, using *II*, reveals the third order products:

$$\frac{3}{2}cAB^2\cos xt(1+\cos 2yt) = \quad (9)$$
$$= \frac{3}{2}cAB^2\cos xt + \frac{3}{2}cAB^2[\frac{1}{2}\{\cos(x+2y)t + \cos(x-2y)t\}]$$

The third order products are:

$$\frac{3}{4}cAB^2[\cos(x+2y)t + \cos(x-2y)t] \quad (10)$$

A similar expansion of the other third order term,

$$\frac{3}{2}cA^2B(1+\cos 2xt)\cos yt \quad (11)$$

gives other third order products as:

$$\frac{3}{4}cA^2B[\cos(2x+y)t + (2x-y)t] \quad (12)$$

Each third order product has a maximum amplitude of 0.75 c.

If A = B the third order intermodulation products have a maximum level of 0.75 c, whereas the third harmonic signal has a maximum level of 0.25 c. The intermodulation products will be 9.5 dB higher than the third harmonic signal amplitude.

The primary products, from the third order term in the amplifier's transfer function, are the third harmonic and third order intermodulation signals. A secondary product is also generated at the fundamental frequency. To keep errors below 0.5 dB, this secondary product should be at least 12 dB lower in amplitude than the primary product (from the first order term). The secondary product has an amplitude of 1.5 cAB^2. This is six times the amplitude of the third harmonic signal, or +15.6 dB. Secondary effects due to the third order term will have a negligible effect on measurements, provided the amplitude of the third harmonic is at least 27.6 dB lower than that of the fundamental. It may be necessary to reduce the amplitude of V_{in} to meet this condition.

Proof of the Rule of Thumb

As discussed at the beginning of this article, there is a "rule of thumb" that states the third order intercept point is 10 to 15 dB above the 1 dB compression point. Because it has been proved that intermodulation products will be 9.5 dB higher than the third harmonic signal amplitude, it is possible to provide a mathematical basis to the "rule of thumb" method.

Starting with the mathematical expression for an amplifier:

$$V_{out} = aV_{in} + bV_{in}^2 + cV_{in}^3 \quad (13)$$

If V_{in} = cos xt,

$$V_{in}^2 = \cos^2 xt = 0.5\,(1 + \cos 2xt) \quad (14)$$

$$V_{in}^3 = \cos^3 xt = 0.75\cos xt + 0.25\cos 3xt. \quad (15)$$

These expressions show that the cubed input voltage term produces output at both fundamental and third harmonic frequencies. At the fundamental frequency the cubed term is: 0.75 cV_{in}.

With low signal levels the amplifier transfer function is approximately, $V_{out} = aV_{in}$.

At the 1 dB compression point:

$$V_{out} = 0.89125\,aV_{in} \quad (17a)$$
$$= aV_{in} + 0.75\,cV_{in} \quad (17b)$$

Equations 17a and 17b can be simplified to find an expression for c :

$$0.75c = -0.10875a \quad (18)$$
$$c = -0.145a$$

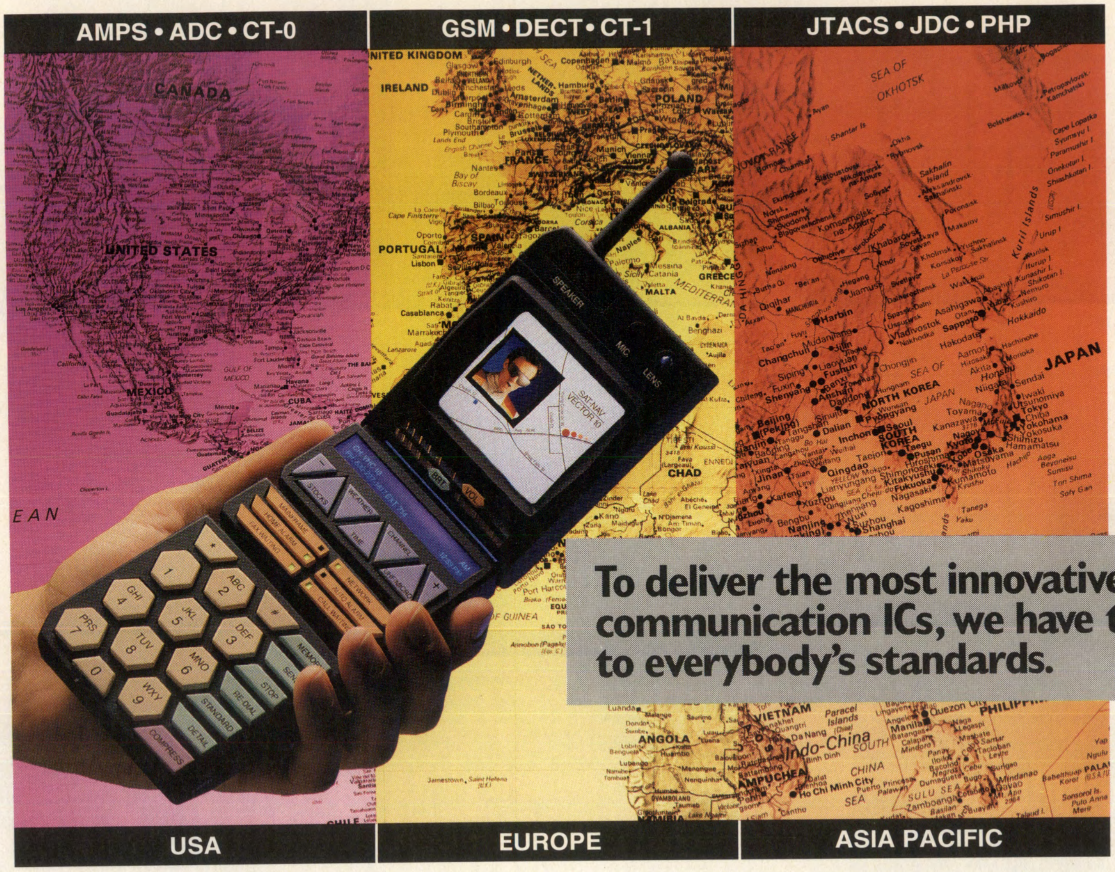
The third harmonic was found to be 0.25 c cos3xt. Because c = –0.145 a, the third harmonic magnitude is 0.03625 a cos 3xt, which is 28.8 dB below aV_{in}. However, we know from previous results that the third order intermodulation products are 9.5 dB above the third harmonic signal level, so the ratio $IMR_3 = aV_{in} -$ 19.3 dB, and third order intercept is:

$$IP_3 = aV_{in} + 0.5\,IMR_3 \quad (19)$$
$$= aV_{in} + 0.5 \times 19.3$$
$$= aV_{in} + 9.66\text{ dB}.$$

But aV_{in} is 1 dB above the 1 dB compression point, so IP_3 = 10.66 dB above 1 dB compression point. What do these results show us?

- If we know the 1 dB compression point we can calculate the third order intercept point, IP_3;
- If we measure signal levels at fundamental, second and third harmonic frequencies, generated by passing a single tone through a linear amplifier, we can calculate its 1 dB compression point, as well as second and third order intercept points;

$$IP_2 = F - (H_2 + 6)\text{ dBm} \quad (20)$$



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$$IP_3 = F - 0.5(H_3 + 9.5) \text{ dBm} \quad (21)$$

$$P_{1dB} = F - 0.5(H_3 + 9.5) - 10.66 \text{ dBm} \quad (22)$$
$$= F - 0.5\,H_3 - 15.41 \text{ dBm}$$

Where F = fundamental,
H_2 = fundamental/2nd harmonic,
H_3 = fundamental/3rd harmonic.

As an example:
$F = 5$ dBm, $H_2 = -50$ dB, $H_3 = -60$ dB
$IP_2 = 5 - (-50 + 6)$ dBm $= 49$ dBm
$IP_3 = 5 - 0.5(-60 + 9.5)$ dBm $= 30.25$ dBm
$P_{1dB} = 5 - 0.5(-60) - 15.41$ dBm
$= 35 - 15.41$ dBm $= 19.59$ dBm

Conclusions

The second order intermodulation ratio, IMR_2, can be calculated by adding 6 dB to the second harmonic signal level and comparing with a known P_{out}. The third order intermodulation ratio, IMR_3, can be calculated by adding 9.5 dB to the third harmonic signal level and comparing with a known P_{out}. Only one signal needs to be applied to the DUT in order to determine the second and third harmonic signal levels. The measuring equipment does not need to be a spectrum analyzer, although it is highly probable that one would be used for other tests on the device. This technique is a useful tool, provided the amplifier's frequency response is flat over the measured range. Errors will occur if the response at harmonic frequencies is not the same as at the fundamental frequency.

This technique has been applied in a few practical tests and found to produce second and third order intercept points within 0.5 dB of results using two tone testing. The input signal level should be adjusted so that the output power is at least 10 dB below the 1 dB compression point; an output level of 0 dBm simplifies the mathematics. Because the test is simple, mistakes are less likely to occur.

As an example, using the two tone test results in Figure 1, the amplitude of second order products is –44.4 dBm and that of third order products is –52.8 dBm, for a 0 dBm fundamental output signal level. So $IMR_2 = 44.4$ dBm and $IMR_3 = 52.8$ dBm, therefore $IP_2 = +44.4$ dBm and $IP_3 = +26.4$ dBm.

The results of a one tone test are given in Figure 3, using the same wide-band amplifier as in Figure 1. The second harmonic signal level is –50.6 dBm and the third harmonic signal level is –62.8 dBm, for a 0 dBm fundamental signal level. Thus $IMR_2 = 0-(-50.6 + 6) = 44.6$ dBm and $IMR_3 = 0-(-62.8 + 9.5) = 53.3$ dBm. Using these results we find that $IP_2 = 44.6$ dBm and $IP_3 = 26.65$ dBm. The error is within 0.25 dB of the two tone test.

Although the above example validates the single tone testing method, it is based on analysis using just the first three terms of the amplifier's transfer function. A quantified analysis is needed to find the conditions for which higher order products have little effect. To meet these conditions it may be necessary to reduce the amplitude V_{in}.

Fourth order terms have negligible effect on second harmonic or intermodulation levels, provided that the amplitude of the 4th harmonic is at least 31.6 dB lower than the 2nd harmonic.

Fifth order terms have negligible effect on fundamental, third harmonic or intermodulation levels, provided that the amplitude of the 5th harmonic is at least

52 dB below the fundamental and 41.25 dB below the 3rd harmonic.

Effect of High Order Distortion Products

Do higher order terms in the amplifier's transfer function have an effect on the results obtained so far?

The first order term only generated one primary product, at fundamental frequency. No secondary products were generated.

The second order term produced second harmonic and second order intermodulation signals. These are the primary products of a second order term. The secondary product was a DC output.

The third order term in the amplifier's transfer function was shown to generate primary products: the third harmonic and third order intermodulation signals. A secondary product was also generated, at the fundamental frequency.

Thus, at the fundamental frequency, the first order term produced primary signals and the third order term produced secondary signals. If secondary products generate an error of 1/16 of the total power, this becomes 1.0625 times the correct value. The error is 0.26 dB. Secondary products have little effect provided their amplitudes are at least 12 dB below the primary products.

In order to determine whether secondary products are low enough to be neglected, consider the series:

$$V_O = aV_{in} + bV_{in}^2 + cV_{in}^3 + dV_{in}^4 + eV_{in}^5$$

Where V_{in} = Acos xt + Bcos yt. The first three terms in this expression have already been expanded to find their products.

Fourth Order Effects

Expanding the fourth order term:

$$dV_{in}^4 = d\{A^4\cos^4 xt + 4A^3\cos^3 xt\, B\cos yt + 6A^2\cos^2 xt\, B^2\cos^2 yt + 4A\cos xt\, B^3\cos^3 yt + B^4\cos^4 yt\} \quad (23)$$

Each term in this expression must be expanded to reveal harmonic and intermodulation products.

Mathematical note IV:

$$\cos^4 zt = \frac{1}{4}\left(\frac{3}{2} + 2\cos 2zt + \frac{1}{2}\cos 4zt\right) \quad (IV)$$

Using mathematical notes I to IV, AV_{in}^4 becomes:

$$A^4\cos^4 xt = A^4\left\{\frac{3}{8} + \frac{1}{2}\cos 2xt + \frac{1}{8}\cos 4xt\right\} \quad (24)$$

$$4A^3\cos^3(xt)\,B\cos(yt) = A^3B\{\frac{3}{2}\cos(x-y)t + \frac{3}{2}\cos(x+y)t + \frac{1}{2}\cos(3x-y)t + \frac{1}{2}\cos(3x+y)t\} \quad (25)$$

$$6A^2\cos^2(xt)\,B^2\cos^2(yt) = A^2B^2\{\frac{3}{2} + \frac{3}{2}\cos 2yt + \frac{3}{2}\cos 2xt\ \frac{3}{4}\cos(2x-2y)t + \frac{3}{4}\cos(2x+2y)t\} \quad (26)$$

$$4A\cos(xt)\,B^3\cos^3(yt) = AB^3\{\frac{3}{2}\cos(x-y)t + \frac{3}{2}\cos(x+y)t + \frac{1}{2}\cos(3y-x)t + \frac{1}{2}\cos(3y+x)t\} \quad (27)$$

$$B^4\cos^4 yt = B^4\left\{\frac{3}{8} + \frac{1}{2}\cos 2yt + \frac{1}{8}\cos 4yt\right\} \quad (28)$$

Note, from the above equations, that there are no third order intermodulation products present. But note that secondary 2nd harmonic and 2nd order intermodulation products are present.

The second harmonic products are:

$$\frac{1}{2}A^4\cos 2xt + \frac{1}{2}B^4\cos 2yt \quad (29)$$

The fourth harmonic terms are:

$$\frac{1}{8}A^4\cos 4xt + \frac{1}{8}B^4\cos 4yt \quad (30)$$

The second order intermodulation products are:

$$\frac{3}{2}A^3B\cos(x-y)t + \frac{3}{2}A^3B\cos(x+y)t + \frac{3}{2}AB^3\cos(x-y)t + \frac{3}{2}AB^3\cos(x+y)t \quad (31)$$

Let A = B, so that $A^3B = A^4$:

Distortion due to 2nd order term:
- 2nd harmonic = 0.5 bA^2
- Intermodulation (2) = bA^2

Distortion due to 4th order term:
- 2nd harmonic = 0.5 dA^4
- Intermodulation (2) = 3 dA^4
- 4th harmonic = 0.125 dA^4
- Intermodulation (4) = 0.5 dA^4

The secondary 2nd harmonic product is four times, or 12 dB, greater in amplitude than the 4th harmonic. To minimize errors, the secondary product must be at least 12 dB lower than the second order term primary product. The 4th order term will have a negligible effect on 2nd harmonic levels, provided that the level of the 4th harmonic is at least 24 dB below the level of the 2nd harmonic.

Secondary 2nd order intermodulation products are 24 times (or 27.6 dB) greater in amplitude than the 4th harmonic. This secondary product should have an amplitude at least 12 dB lower than the second order term primary product, as explained above. Calculations, using the second order term, have shown that the second order intermodulation level is 6 dB greater than the 2nd harmonic. Therefore, secondary intermodulation produced by the 4th order term should be kept 6 dB below the 2nd harmonic. If the 4th order contribution to intermodulation is to be negligible, the level of the 4th harmonic must be at least 31.6 dB lower than the 2nd harmonic.

Fifth Order Effects

The fifth order term must be examined to see if this affects the third order intermodulation products. Expanding the fifth order term:

$$eV_{in}^5 = e\{A^5\cos^5 xt + 5A^4\cos^4 xt\, B\cos yt + 10A^3\cos^3 xt\, B^2\cos^2 yt + 10A^2\cos^2 xt\, B^3\cos yt + 5A\cos xt\, B^4\cos^4 yt + B^5\cos^5 yt\} \quad (32)$$

Harmonic and intermodulation products in the above expression are revealed by expanding each term.

Mathematical Note V:

$$\cos^5 zt = \frac{1}{8}\left(5\cos zt + \frac{9}{2}\cos 3zt + \frac{1}{2}\cos 5zt\right) \quad (V)$$

Using mathematical notes I to V, AV_{in}^5 becomes:

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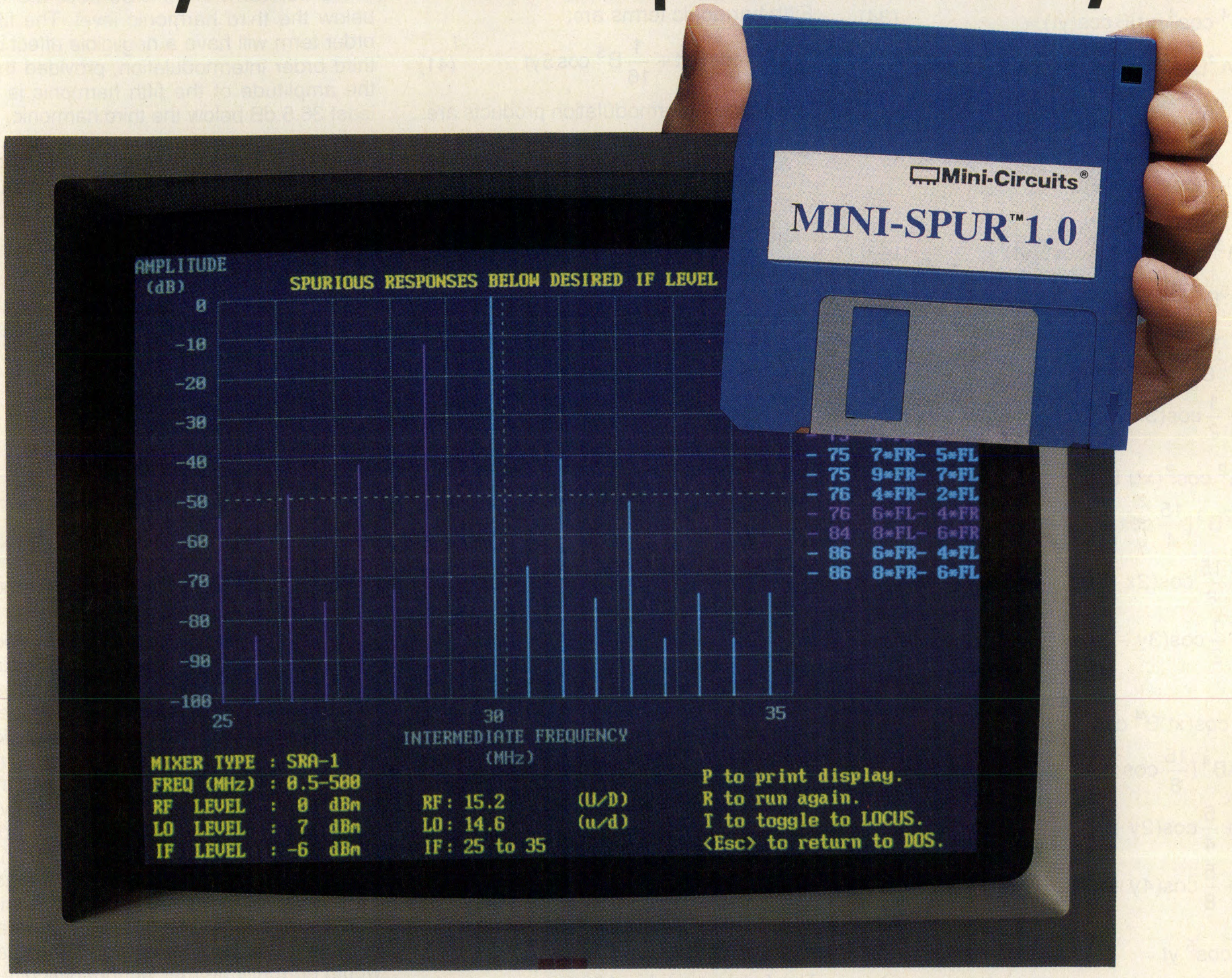

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$$A^5 \cos^5 xt = \quad (33)$$

$$A^5\{\frac{5}{8}\cos xt + \frac{9}{16}\cos 3xt + \frac{1}{16}\cos 5xt\}$$

$$5A^4 \cos^4(xt) B \cos(yt) = \quad (34)$$

$$A^4B\{\frac{15}{8}\cos yt + \frac{5}{4}\cos(2x-y)t + \frac{5}{4}\cos(2x+y)t + \frac{5}{8}\cos(4x-y)t + \frac{5}{8}\cos(4x+y)t\}$$

$$10A^3 \cos^3(xt)\ B^2 \cos^2(yt) = \quad (35)$$

$$A^3B^2\{\frac{15}{4}\cos xt + \frac{5}{4}\cos 3xt + \frac{15}{8}\cos(2y-x)t + \frac{15}{8}\cos(2y+x)t + \frac{1}{2}\cos(3x-2y)t + \frac{1}{2}\cos(3x+2y)t\}$$

$$10A^2 \cos^2(xt)\ B^3 \cos^3(yt) = \quad (36)$$

$$A^2B^3\{\frac{15}{4}\cos yt + \frac{5}{4}\cos 3yt + \frac{15}{8}\cos(2x-y)t + \frac{15}{8}\cos(2x+y)t + \frac{1}{2}\cos(3y-2x)t + \frac{1}{2}\cos(3y+2x)t\}$$

$$5A \cos(xt) B^4 \cos^4(yt) = \quad (37)$$

$$AB^4\{\frac{15}{8}\cos xt + \frac{5}{4}\cos(2y-x)t + \frac{5}{4}\cos(2y+x)t + \frac{5}{8}\cos(4y-x)t + \frac{5}{8}\cos(4y+x)t\}$$

$$B^5 \cos^5 yt = \quad (38)$$

$$B^5\{\frac{5}{8}\cos yt + \frac{9}{16}\cos 3yt + \frac{1}{16}\cos 5yt\}$$

The fifth order term generates products at the fundamental and at the third and fifth harmonics. Third order intermodulation is also produced.
The fundamental products are:

$$\frac{5}{8}A^5 \cos xt + \frac{5}{8}B^5 \cos yt + \frac{15}{8}A^4B \cos xt + \frac{15}{8}AB^4 \cos yt + \frac{15}{4}A^3B^2 \cos xt + \frac{15}{4}A^2B^3 \cos yt \quad (39)$$

Third harmonic terms are:

$$\frac{9}{16}A^5 \cos 3xt + \frac{9}{16}B^5 \cos 3yt + \frac{5}{4}A^3B^2 \cos 3xt + \frac{5}{4}A^2B^3 \cos 3yt \quad (40)$$

Fifth harmonic terms are:

$$\frac{1}{16}A^5 \cos 5xt + \frac{1}{16}B^5 \cos 5yt \quad (41)$$

Third order intermodulation products are:

$$\frac{5}{4}A^4B \cos(2x-y)t \quad (42)$$

$$+\frac{5}{4}A^4B \cos(2x+y)t$$

$$+\frac{15}{8}A^3B^2 \cos(2y-x)t$$

$$+\frac{15}{8}A^3B^2 \cos(2y+x)t$$

$$+\frac{15}{8}A^2B^3 \cos(2x-y)t$$

$$+\frac{15}{8}A^2B^3 \cos(2y+x)t$$

$$+\frac{5}{4}AB^4 \cos(2y-x)t$$

$$+\frac{5}{4}AB^4 \cos(2y+x)t$$

Let A = B, then:
Distortion due to 3rd order term:
3rd harmonic = $0.25\ cA^3$
Intermodulation (3) = $0.75\ cA^3$
Distortion due to 5th order term
Fundamental = $6.25\ eA^5$
3rd harmonic = $1.8125\ eA^5$
Intermodulation (3) = $3.125\ eA^5$
5th harmonic = $0.0625\ eA^5$

Secondary products due to the 5th order term should be more than 12 dB below the amplitude of primary products due to the other terms.

Fundamental and harmonic products due to the 5th order term are related. The fundamental is 100 times greater in amplitude than the fifth harmonic. The fifth order term will have negligible effect on the fundamental level, provided that the 5th harmonic is at least 52 dB below the level of the fundamental.

The third harmonic due to the 5th order term is 29.25 dB greater in amplitude than the fifth harmonic. The fifth order term will have negligible effect on the third harmonic, provided that the amplitude of the fifth harmonic is at least 41.25 dB below the third harmonic.

Secondary 3rd order intermodulation products due to the 5th order term are 50 times, or 34 dB, greater in amplitude than the fifth harmonic. The third order term was expanded earlier and intermodulation products were found to have an amplitude that is 9.5 dB above the third harmonic. To minimize errors, by keeping a 12 dB differential between primary and secondary products, those due to the 5th order term should be kept 2.5 dB below the third harmonic level. The fifth order term will have a negligible effect on third order intermodulation, provided that the amplitude of the fifth harmonic is at least 36.5 dB below the third harmonic.

Effect on 1 dB Compression Point

Expanding the expression for the output from an amplifier:

$$V_{out} = aV_{in} + bV_{in}^2 + cV_{in}^3 + dV_{in}^4 + eV_{in}^5 \quad (43)$$

The first three terms in this expression have been calculated previously. If $V_{in} = \cos xt$,

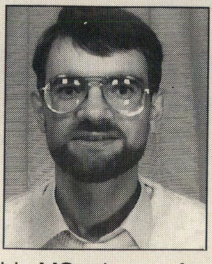
$$V_{in}^4 = \cos^4 xt = \frac{1}{4}\left(\frac{3}{2} + 2\cos 2xt + \frac{1}{2}\cos 4xt\right) \quad (44)$$

$$V_{in}^5 = \cos^5 xt = \frac{1}{8}\left(5\cos xt + \frac{9}{2}\cos 3xt + \frac{1}{2}\cos 5xt\right) \quad (45)$$

The fifth order term, 0.625 e cos xt, will be in phase with the fundamental and tend to increase the 1 dB compression point. Its amplitude will be ten times (+20 dB) greater than the fifth harmonic. At the 1 dB compression point the third harmonic will have an amplitude 28.8 dB less than the fundamental, using previous analysis. This will be unaffected by the fifth order term, provided that the amplitude of the fifth harmonic is at least 3.2 dB lower than the third harmonic (and therefore 32 dB below the fundamental). This is a trivial condition that would normally be met without difficulty.

RF

About the Author



Steve Winder is employed by BT Laboratories, in Ipswich, England. The mathematics behind this article was developed during his preparation for his MSc degree from Essex University. Mr. Winder can be reached at 4 Evabrook Close, IPSWICH, Suffolk, IP2 9JJ, England, UK

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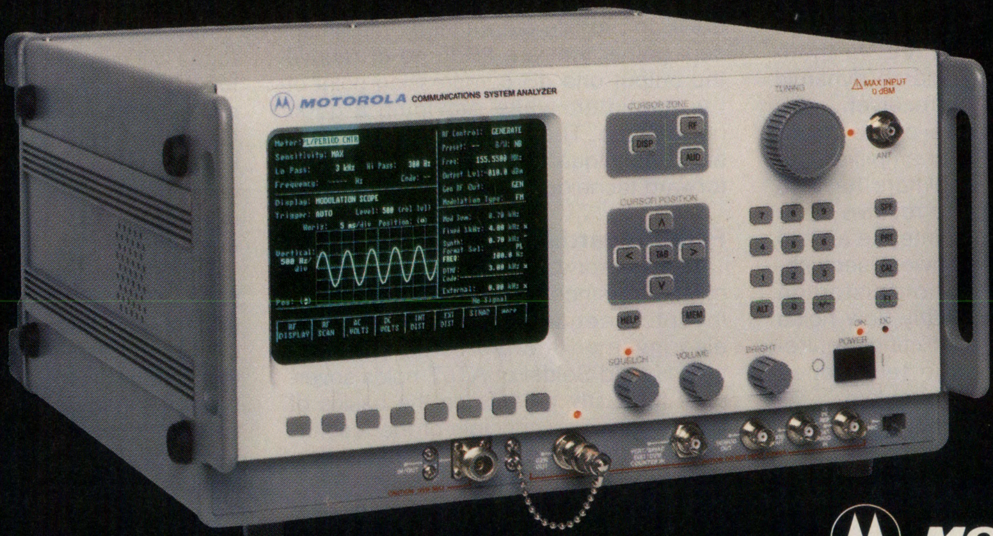

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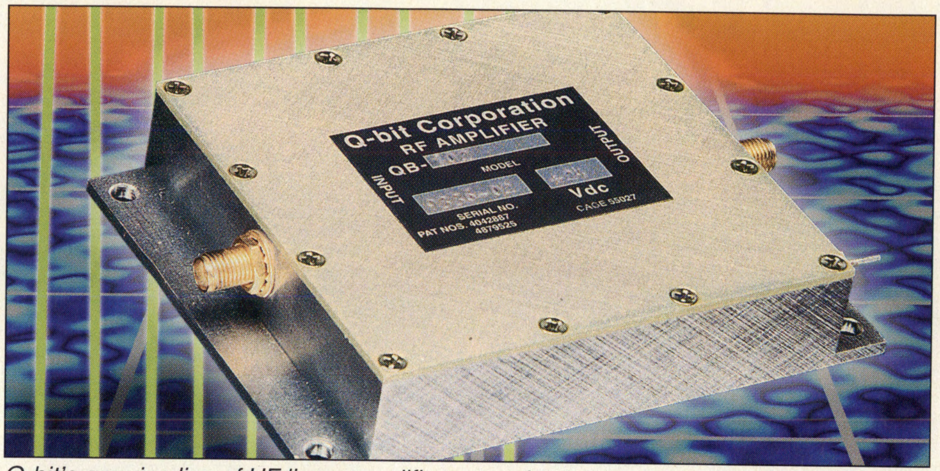
Techniques to Achieve Linear Amplification at HF

By Chris Rice
Q-bit Corporation

Receiver front end amplifiers for HF (2-32 MHz) systems must be capable of handling a wide range of signal levels with low distortion and high linearity. This article describes the design schemes used by Q-bit Corporation to achieve exceptional linearity in their amplifiers for the HF band and beyond, up to 70 MHz. Both feedback and feedforward techniques are described, and some typical performance curves are presented.



Q-bit's growing line of HF linear amplifiers uses feedback and feedforward techniques to achieve high dynamic range and low distortion.

Linear amplification in the 2 to 32 MHz high frequency (HF) band is a critical system specification. There are numerous military and commercial radio services in this band, plus nine amateur radio bands. HF users include short wave AM broadcasting, low band land mobile and fixed communications, plus other aeronautical, maritime, science, and even space uses. For a given communications link, these signals can become unintentional "jammers," interfering with the desired communications. Even if these other users are out of the intended band, it is possible for their signals to mix or beat against the desired signals, creating spurious products that fall in-band, interfering with desired communications. This problem is compounded at HF since the desired link may be several hundred feet to thousands of miles, placing a large dynamic range requirement on the receiver.

System requirements place great demands on the front end amplifiers for these systems. The amplifiers have to be low enough in noise to receive weak signals, yet, at the same time, be able to handle a nearby higher power user without adding intermodulation distortion (IMD) to the system. Clearly, low noise and low distortion are conflicting requirements, and novel design techniques must be used to solve the problem.

Design Techniques

A common rule of thumb in amplifier design is that the output third order intercept point, OIP3, is 10 dB above the output 1dB compression point, P_{1dB} [1]. An overview of existing products from various manufacturers shows this to be fairly accurate for reactively matched amplifiers. Lower frequency, resistive feedback amplifiers have OIP3 about 10 to 15 dB above P_{1dB}; however, the resistive feedback adds undesirable additional noise to the system. This type of performance is not acceptable in the HF systems discussed earlier.

Q-bit has two techniques that it uses at HF to provide output third order intercept points that are 25 to 40 dB above the output one dB compression point of the amplifier, while providing a good noise figure. The first is a patented feedback technique and the other is a feedforward technique.

Feedforward Design

The feedforward amplifier has been in existence since 1924. H.S. Black, who invented negative feedback in 1927, also invented the feedforward correction. Harold Seidel of AT&T holds several of the early patents on the basics of the design. The principle of the feedforward amplifier operation is straightforward, and is described as follows:

Referring to Figure 1, there are two loops that make up the feedforward amplifier. The output from the first loop is fed into the error amplifier, G2; it has two paths: T1-D1-TS1-T3 (path 1a) and T1-ϕ1-G1-T2-TA-T3 (path 1b). Assuming that the RF input consists of two CW RF tones and no distortion, the output of path 1a would have these same tones but at a lower level. The output of path 1b would contain not only the CW RF tones, but also any distortion generated by the primary amplifier, G1. The values of the components in path 1a and path 1b are selected such that the CW RF tones will cancel at the output of the first loop (the input to G2). Ideally, the only signals that are present at the input of G2 are the distortion products generated by G1. In practice, the CW RF tones will only partially cancel, usually by 25 to 40 dB, but this fact will be ignored for this analysis.

The second loop also consists of two paths: T2-D2-T4 (path 2a) and G2-ϕ2-T4 (path 2b). The signals in path 2a consist of the two CW RF tones and the distortion products produced by G1. The signals in path 2b consist only of the distortion products generated in G1 (G2 can be selected such that its distortion is very small; and will be ignored for this analysis). The components in the second loop are selected such that the distortion

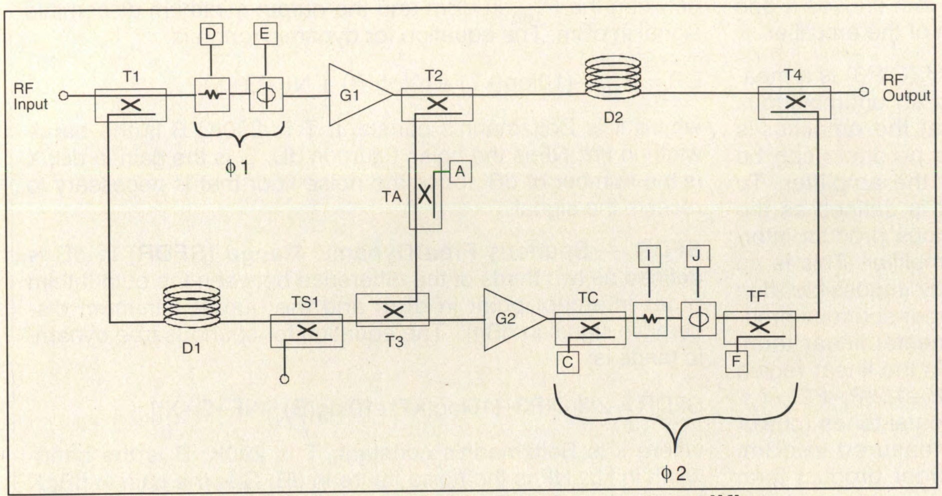


Figure 1. Block diagram of the feedforward power amplifier.

terms in both paths cancel, producing (ideally) only amplified versions of the original CW RF tones at the output.

Fundamentally, the above paragraphs describe the operation of the feedforward amplifier. However, the partial cancellation of the CW RF tones in loop 1 was ignored as was the partial cancellation of the distortion at the RF output; distortion in the error amplifier (G2) was also ignored as was the operation of the amplifier under varying environmental conditions (temperature, vibration, humidity, etc.). Overcoming these challenges is the key to providing superior feedforward performance.

Practical Implementations

Several HF band high dynamic range amplifiers using feedforward methods and the Q-bit patented feedback technique have been fabricated. These include the QBS-101, which is available now, and the QB-7223 and the QB-105, which will be available in the first quarter of 1994.

The QBS-101 utilizes feedforward technology to simultaneously provide low distortion and low noise figure. The unit has approximately 12 dB of gain, a 4.0 dB noise figure, and an output third/second order intercept point of greater than 60/100 dBm. The unit is only 6×3×1.5 inches (L×W×H) and operates from 24 Volts, drawing less than 350 mA. Performance is specified over a 2 to 70 MHz frequency band.

The QB-105 utilizes a patented feedback technique to provide low noise figure with extremely high output power over the 2 to 32 MHz HF band. The unit has approximately 22 dB of gain, a 4.5 dB noise figure, an output 1 dB compression point of 37 dBm, and an output third/second order intercept point of

Definitions

P_{1dB} — The output one dB compression point (P_{1dB}) is defined as the output power of an amplifier that causes a one decibel drop in the fundamental (linear) term of the amplifier.

OIP3 — The output third order intercept point (OIP3) is a measurement of linearity of an amplifier and is an unambiguous figure of merit. By knowing OIP3 and that the amplifier is operating in its linear region, the spurious products can be determined at any output power level for the amplifier. To determine OIP3, a two tone test is used; it is defined as the imaginary point where the third order spurious product intercepts the fundamental linear term of the amplifier. This is an imaginary point since it would be theoretically impossible, due to Parseval's power theorem, for the third order spurious product to be at the same level as the fundamental linear term. OIP3 is measured with two tones (x and y) in the linear region of the amplifier and is defined as $OIP3 = P_f + 1/2(P_f - P2f_x - f_y)$, where P_f is the power in one of the fundamental tones (power in tone x and tone y are equal) and is measured in dBm, $P2f_x - f_y$ is the power in the highest third order product term (ideally these terms are equal in power) and is measured in dBm.

DR — Dynamic Range (DR) in dB is defined as the difference between the P_{1db} in dBm and the output minimum discernible signal in dBm. The equation for dynamic range is:

$$DR = P_{1dB} - (10\log(kT) + 10\log(B) + NF + G + X)$$

where k is Boltzmann's constant; T is 290K; B is the bandwidth in Hz; NF is the noise figure in dB; G is the gain in dB; X is the number of dB above the noise floor that is necessary to receive the signal.

SFDR — Spurious Free Dynamic Range (SFDR) in dB is defined as two thirds of the difference between the output third order intercept point in dBm and the output minimum discernible signal in dBm. The equation for spurious free dynamic range is:

$$SFDR = 2/3[OIP3 - (10\log(kT) + 10\log(B) + NF + G + X)]$$

where k is Boltzmann's constant; T is 290K; B is the bandwidth in Hz; NF is the noise figure in dB; G is the gain in dB; X is the number of dB above the noise floor that is necessary to receive the signal.

greater than 50/90 dBm. The unit requires a 24 Volt supply and draws less than 1500 mA.

These new amplifiers augment Q-bit's connectorized HF band amplifiers listed in Table 1. All of the amplifiers demonstrate the low noise/high intercept point required for linear amplification in the front end of HF receivers. The dynamic range (DR) and the spurious-free dynamic range (SFDR) [2] of the amplifiers are presented in Figure 2. Note that all of the amplifiers have SFDR greater than 90 dB and DR greater than 115 dB.

Amplifiers in this article represent the state-of-the-art in HF band dynamic range performance. While the requests from customers for better noise and intermodulation distortion are not likely to end, the amplifiers described in this article provide performance advantages over existing amplifiers in this class. For the future, HF band amplifiers with 65 to 70 dBm third order intercept points and 135 to 140 dBm second order intercepts are presently being researched, and will be available.

Readers interested in these products can contact the author at the address below, or circle Info/Card #207 ***RF***

AMPLIFIER	FREQUENCY (MHz)	GAIN (dB)	NF (dB)	P1dB (dBm)	3RDS/2NDS (dBm)
QB-101	2 - 70	22	4.5	31	55/110
QB-102	2 - 32	12	6.2	28	50/100
QB-105	2 - 32	22	4.5	37	50/90
QB-7205	2 - 70	22	4.0	27	51/65
QB-7223	2 - 32	22	4.5	33	50/65
QBS-101	2 - 70	12	4.0	20	60/100

Table 1: Typical HF High Dynamic Range Amplifiers.

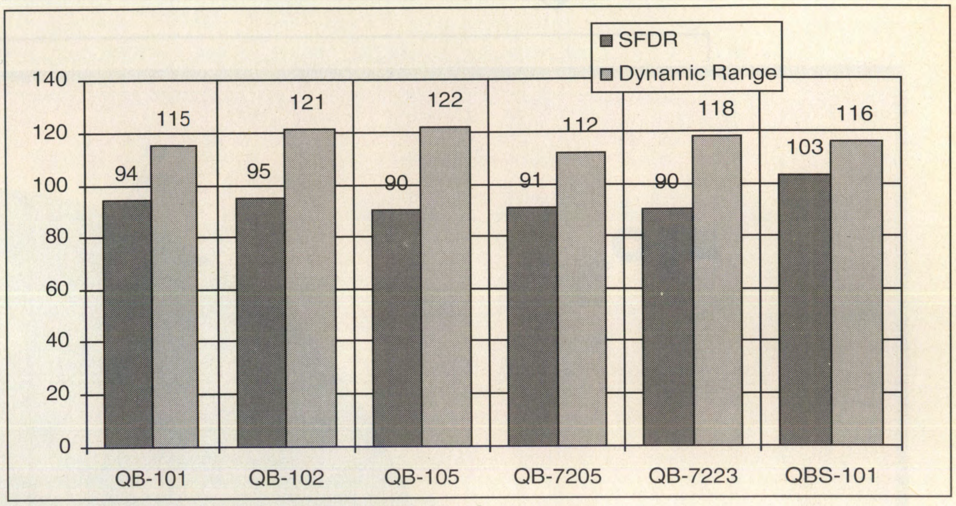


Figure 2. Typical dynamic range performance.

References

1. Guillermo Gonzalez, Microwave Transistor Amplifiers: Analysis and Design, Prentice-Hall, Inc., Englewood Cliffs, NJ, pp. 178-179.
2. Gonzalez, pp. 174-180. Defined in a 1 MHz bandwidth with a minimum discernable signal of 3 dB above the noise floor.

About the Author

Chris Rice is Engineering Manager at Q-bit Corporation, 2575 Pacific Avenue NE, Palm Bay, FL 32905; tel. (407) 727-1838, fax. (407) 727-3729.

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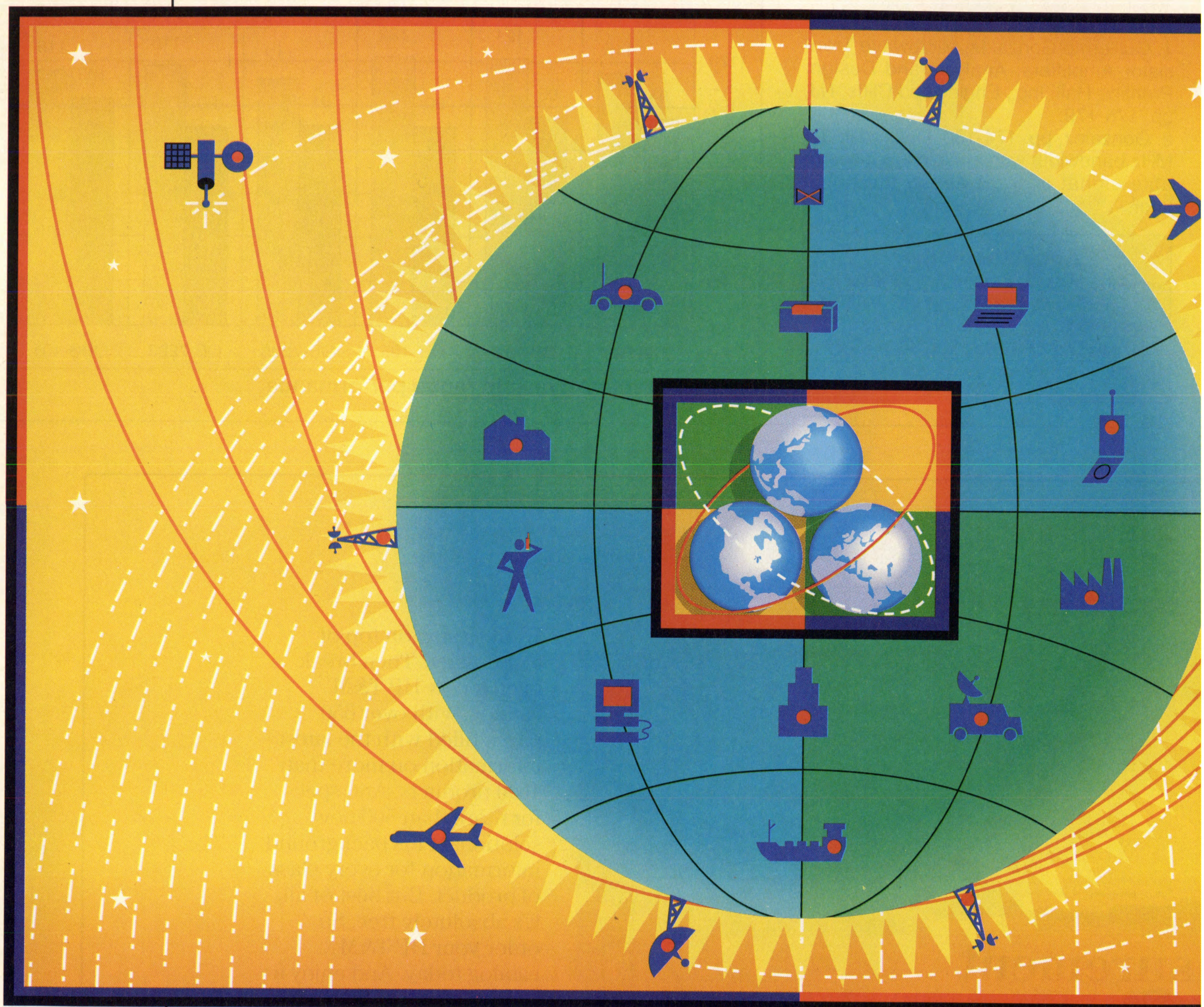

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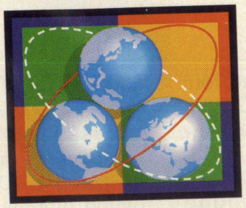

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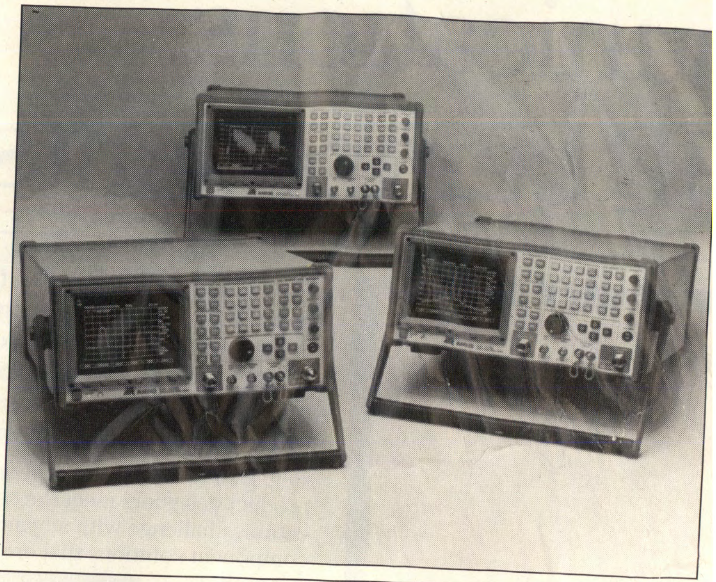
Spectrum Analyzer Line Expands With New Models

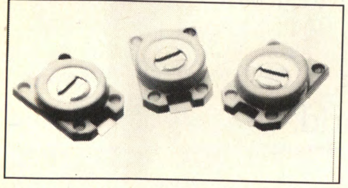


A high speed 200 ns/div sweep with pre/post trigger capability, a sensitive FM/AM receiver with modulation measurement scales, 3 Hz to 30 MHz resolution bandwidth, and a + 30 to −135 dBm measurement range are standard features on three new models from IFR Systems. Other standard features include a 5 MHz digital storage oscilloscope, DC to 20 kHz FFT analyzer, and automatic trace limits test function. For remote or field service use, all three models are designed to meet Mil-T-28800 Class 5 and can be powered from an optional rechargeable battery pack. Standard RS-232 and IEEE-488 interfaces allow remote control operation or direct hard copy output to a plotter. Optional, built-in features include a 2.9 GHz tracking generator, quasi-peak detector, and 0.02 ppm time base. Frequency coverage of 9 kHz to 2.9 GHz for the model AN920, 9 kHz to 22 GHz for the model AN930, and 9 kHz to 26.5 GHz for the model AN940 can be exteded with external mixers for higher frequency measurements. All three instruments are covered by the standard two-year IFR warranty.

IFR Systems, Inc.
INFO/CARD #250

Chip Trimmer Capacitors

Voltronic's JZ line of surface mount, chip-size trimmer capacitors been expanded with the addition of a 15 pF part. Members of the JZ line measure 0.177" (4.5mm) long, 0.126" (3.2mm) wide and 0.055" (1.5mm) high. It has six capacitance ranges from 3 to 20 pF maximum and DC working voltages of 100 and 50 V. Temperature coefficients are as low as 0 ±200 ppm/°C. The self inductance of all parts is constant at 1.9 nH. At 10 pF and below, the SRF is above 1 GHz. The high-temperature, liquid-crystal poly-

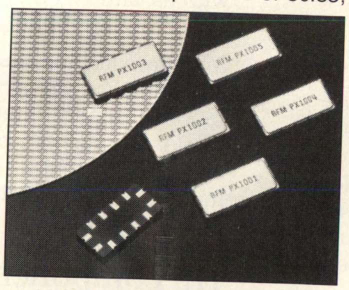


mer case is dust-proof and can be washed with water or solvents as advised. The parts are also easilly vacuum-placed because of their lack of air leakage. Under 1000 hour load life tests, 1000 hour humidity tests and temperature cycling, the capacitance changes were substantially below the ±5% allowed, measuring 1% max vs. competitive parts which change over 10%. Quantities under 1000 are in stock, with larger qunatities available in 30 to 60 days. Costs are below $0.40 in quantity. An engineering sample kit is $40.00 for 25 parts. Tape and reel packaging is at no extra cost.

Voltronics Corp.
INFO/CARD #249

SAW IF Filters

RF Monolithics introduced four new member of its line of SAW IF filters at RF Expo East. The new RFM filters, PX1002, PX1003, PX1004 and PX1005, are designed for radiotelephone applications. The PX1002, PX1003, PX1004 and PX1005 have center frequencies of 86.85,

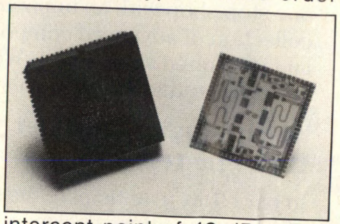


150.005, 82.20 and 86.01 MHz, respectively. The PX1002/1004/1005 are designed for North American cellular applications including AMPS, IS-54 (TDMA) and CDPD. They have maximum insertion loss of 4 dB, a minimum 3 dB BW of 30 kHz, and 70 dB of image rejection at -910 kHz. In-band group delay deviation is less than 6 µs p-p. The PX1003 is designed for CT-2 and PCN IF applications. The PX1003 filter features a maximum insertion loss of 4 dB, a minimum 3 dB BW of 70 kHz, and 65 dB of image rejection at −21.4 MHz. In band group delay deviation is less than 5 µs p-p. Each of the four new filters is provided in a 13.3 x 6.5 x 2 mm hermetic surface mount package.

RF Monolithics, Inc.
INFO/CARD #248

Cellular Components

AT&T Microelectronics has released two high performance components in JEDEC standard, 84-pin PLCC surface mount hybrid modules. The 2121B RF amplifier is designed to complement the 2121A complex vector attenuator (CVA), but can be used as a stade alone amplifier in applications requiring unconditional stability, linearity, excellent impedance match and low noise figure. The 2121B is a balanced amplifier featuring wide dynamic range with typical third order



intercept point of 42 dBm. The companion 2121A CVA is a compact second-generation combined endless phase shifter and attenuator. The part can control phase and amplitude without introducing intermodulation distortion, dispersion or variation in group delay. The 2121A is the industry's only single package, surface mounted, endless phase shifter and attenuator. Samples of the new products will be available in 1Q94, followed by production shortly thereafter. In quantity orders above 5k, the 2121A will be available at $46 each and the 2121B at $33 each.

AT&T Microelectronics
INFO/CARD #247

Sub-Miniature VCO

Z-Communications has introduced the model SMV2500 VCO. Developed for wireless LAN applications, the SMV2500 measures 0.3" x 0.3" x 0.117" (7.62 x 7.62 x 2.97mm), minimizing the space it occupies on the industry standard "credit card" format. The device tunes the frequency range of 2400 to 2485 MHz with an applied tuning voltage of 0 to 3 Vdc. The power output is 10 dBm (10 mW) across the frequency bandwidth. Tuning sensitivity is 65 MHz /V. Other performance characteristics include excellent phase noise performance of −85 dBc @ 10 kHz offset and second harmonic suppression of −20 dBc min. The SMV2500 has a low

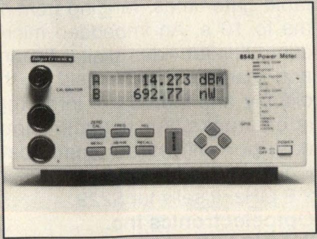
sensitivity to power supply variation (pushing) of 5 MHz/V. Operational temperature range is 0 to 70 °C. The SMV2500 is priced at $5.75 at 100,000 units, and samples are available at $150 per VCO from stock. Delivery of production VCOs will commence in January, 1994.

Z-Communications, Inc.
INFO/CARD #246

Product Spotlight: Power Meters

Meter Has Dual-Channel Outputs

The Gigatronic 8542 Universal Power Meter now can provide dual channel analog



recorder outputs. This capability allows simultaneous peak and average power measurement, power and power ratio outputs, or extended range of –10 to +10 Volts. Both channels are independently controllable via a GPIB interface. Also introduced area series of peak power sensors,with six models covering 45 MHz to either 18, 26.5 or 40 GHz. The second channel output option is priced at $350, and the new sensors range from $2800 to $3650.
Giga-tronics Inc.
INFO/CARD #245

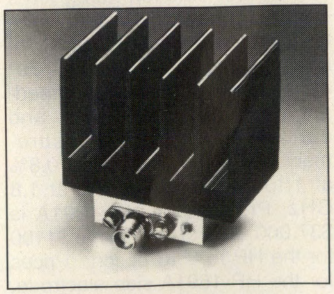
Peak Power Analyzer

A new peak power analyzer from Hewlett-Packard features an extended dynamic power range of >45 dB, with accuracy to within one percent of H-P's average power meters. The new HP 8991A is suitable for mobile communications transmitter testing, and digital cellular radio transmitters which require accurate peak power measurements. The HP 8991A is priced at $13,491, with a lower cost single channel version (Option 001) at $8991.
Hewlett-Packard Company
INFO/CARD #244

AMPLIFIERS

Rugged 2-8 GHz Amplifier

Mini-Circuits has introduced a tough, medium power microwave amplifier covering the wide 2-8 GHz frequency range. The



ZRON-8G amplifier typically provides 22.5 dBm outptput power, 2.0:1 max. VSWR and linear gain of at least 20 dB over the whole band. The amplifier is unconditionally stable. The ZRON-8G's SMA-F connectors are field removable to a pin connection for mounting on circuit board. The ZRON-8G is priced at $495 each, with a 1 week shipment
Mini-Circuits
INFO/CARD #243

Dual Channel Power Amplifier

Model PA03-800 is a single input, dual output RF power amplifier that operates in the 700 700 kHz to 2.7 MHz frequency range. Both outputs are rated at 400 W pulse or CW and are matched in both phase and amplitude. Both channels are adjustable from 0 to 400 W output and can withstand infinite load mismatch for an indefinite period with no damage.
Intech
INFO/CARD #242

Hybrid Amplifier

The model AC2577 thin-film cascadable amplifier operates over the frequency range of 10 to 2500 MHz with a minimum gain of 10.0 dB (0 to 50 °C). Noise figure is typically 3.3 dB, and output power at the 1 dB compression point is +21.0 dBm. Current drain is 100.0 mA at 15 V.
Cougar Components
INFO/CARD #241

Microcell Amps.

Microwave Power Devices has developed a multi-channel cellular microcell amplifier. Model LWA 880-30/14288 is a solid-

state, class A amplifier module designed for low intermodulation, multi-carrier operation. The amplifier operates over the 869-894 MHz frequency band with linear output power of 44 dBm at 1 dB compression. The module operates at 28 Vdc and 5.5 A.
Microwave Power Devices, Inc.
INFO/CARD #240

SIGNAL SOURCES

One Time Programmable Oscillator

The new Fox Programmable Oscillator is designed as a direct replacement for conventional metal can oscillators through identical packaging, pinout and characteristics. However, the F6233 output frequency can be programmed one time to values in the range of 0.4 to 135 MHz, eliminating the need to stock multiple frequencies.
Fox Electronics
INFO/CARD #239

OCXOs

NEL Frequency Controls announces its new line of oven controlled crystal oscillators. The NEL 108D and 108E have <5x10-10/day aging and short term stability of <5x10-12/sec. Phase noise is <–160 dBc at 10.0 kHz. The "D" version uses a printed circuit connector board, while the "E" uses filter feed-through terminals. The NEL 108D costs $650.00 each, while the NEL 108E costs $775.00 each.
NEL Frequency Controls, Inc.
INFO/CARD #238

SIGNAL PROCESSING COMPONENTS

SP4T Switch

Daico introduces the model DSO874 SP4T GaAs MIC non-reflective switch, featuring an operating frequency of DC-2000 MHz. The switch provides transistion time of 16 ns (90%/10% RF) with typical switching speed of 55 ns (50% control to 10%/90% RF). Switching transients are 25 mV and isolation is 70 dB at DC-100 MHz, 60 dB at 100-300 MHz, 46 dB at 300-1000 MHz and 35 dB at 1000-2000 MHz. The switch is TTL driven and comes in 14 pin DIP, flatpack or surface mount packages.
Daico Industries, Inc.
INFO/CARD #237

Octave Band Mixer

The MO49MN/PN from Magnum Microwave is a double balanced, octave band mixer. Available with removable SMA connectors (PN version) and without (MN version), this mixer has conversion loss near 5 dB for 100 MHz IF and near 6 dB for 1500 MHz IF. For LO at 8.5 GHz, both LO-RF and LO-IF isolation are near 30 dB; for LO at 3.5 GHz, LO-RF and LO-IF isolations are about 35 and 18 dB, respectively. LO drive power is +19 dBm.
Magnum Microwave Corp.
INFO/CARD #236

SEMICONDUCTORS

Dual Op-Amp

Comlinear announces the CLC412, a high performance dual video operational amplifier with closely matched AC and DC performance and high channel isolation. Employing a current-feedback topology, the CLC4123 provides a 250 MHz, -3 dB small-signal BW at a gain of +2 and a 1300 V/ms slew rate while consuming only 55 mW per amp. using a ±5 V supply. The device is available now in several different packages, or as die. Cost is $5.95 for PDIP and SOIC packaging in 1000 piece qty.
Comlinear Corp.
INFO/CARD #235

DQPSK Modulators

Analog Devices has introduced two differential quadrature phase shift key (DQPSK) modulators for use in North American and Japanese digital and cellular phone equipment. The AD7010 is compatible with Japan's JDC cellular standard with 25 kHz channel BW. The AD7011 is for 30 kHz North American IS-54 digital cellular networks. Both ICs generate a quadrature, 45°-shifted signal with root raised cosine filtering. Operation is from a +5 V supply.
Analog Devices
INFO/CARD #234

High Speed Buffers

Calogic announces additions to its high speed buffer product line. Models CLM4102, 4202 and 4302 are low power devices with 250 MHz bandwidth, capable of 150 mA output current. The CLM4122, 4222 and 4322 are ultra low power versions, with 180 MHz bandwidth and 800 µA operating current. These devices will provide 60 mA peak output current. Pricing in 100s ranges from $1.39 to $1.95, depending on speed and package.
Calogic Corp.
INFO/CARD #233

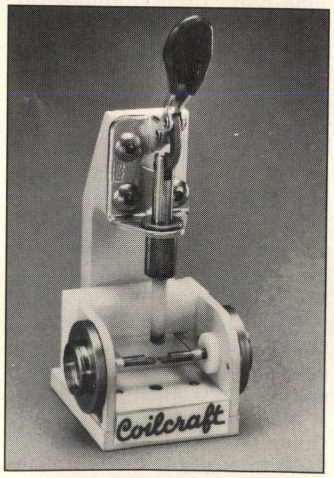
TEST EQUIPMENT

Field Strength Meter

Model R-501P from Z Technology is a handheld, battery operated, fully synthesized RF field strength meter. The unit covers 0.3 to 1000 MHz with frequency steps as fine as 2 kHz. The instrument can measure input power levels from as low as 0.32 µV up to 320 mV. It has input impedance of 50 ohms, and power measurement accuracy is guaranteed to within ±2 dB. Data logging and instrument control can be accomplished via an RS-232C interface. Price in the U.S. is $5,950.
Z Technology Inc.
INFO/CARD #232

Test Fixture

Coilcraft has introduced a precesion test fixture which measures surface mount inductor parameters at high frequencies.



The SMD-D fixture allows non-destructive L, Z and SRF testing of surface mount inductors from DC to 4 GHz. With an optional high Q tuning capacitor, you can test Q at frequencies as high as 1 GHz. The fixtuer accomodates body sizes from 0805 to 1812.
Coilcraft
INFO/CARD #231

Frequency Counter

Optoelectronics announces a high performance, pocket sized addition to their line of Handi Counters, the model M1. Unique features include ten, user-selectable gate-times, ranging from 13 ms to 10 s. An imbedded micro-controller provides digital filtering and readings can be held on the display and stored ion three register stacks. An asynchronous data port allows automatic data logging. The M1 covers 10 Hz to 2.8 GHz. it sells for $229.
Optoelectronics Inc.
INFO/CARD #230

Counter/Timer

Keithly Instruments' model 776 programmable timer/counter offers two independent input channels for measuring frequencies to 225 MHz. The 2.4G version offers a third channel for measuring 50 MHz to 2.4 GHz. Thirteen measurement functions are available, including peak voltage, time interval A to B, ratio C/B, event counting and phase measurements. Price for the standard model 776 is $1995, the 776/2.4G costs $2495.
Keithly Instruments, Inc.
INFO/CARD #229

SMD Testing

Hewlett-Packard has introduced a surface mount device (SMD) test solution using HP's 1.8 GHz impedance/material analyzer (HP 4291A) and a family of integrated test fixtures. The HP 4291A directly measures impedance as a ratio of voltage and current. Impedance measurement accuracy varies from 0.8% at 1 to 100 MHz and 4% at 1.8 GHz. Price of the HP 4291A is $37,000, with an additional $1100 for the HP IBASIC option. Prices for the HP 1691A side electrode, HP 1692A parallel electrode and HP 1693A small side electrode SMD test fixtures are $2200, $1760 and $1980, respectively.
Hewlett-Packard Co.
INFO/CARD #228

300 MHz DSO

Model 9361 from LeCroy Corp. is a 300 MHz digital oscilloscope with sufficient sample rate to accurately characterize input signals up to the full bandwidth.

Waveforms are digitized at 2.5 Gs/s, maintaining an 8:1 ratio between sample rate and maximum bandwidth. Full oscilloscope features, including "smart" triggering functions, are included. The 9361 is priced at $8990, with waveform math and FFT processing options available for $1250 each.
LeCroy Corporation
INFO/CARD #227

Low Cost Function Generators

Two low cost functions generators are introduced by Analogic Corp. for general-purpose testing, laboratory or classroom applications. The 2310, priced at $695, covers frequencies up to 10 MHz and FM modulation capability. The 2311 is $995, and adds AM and sweep capabilities.
Analogic Corporation
INFO/CARD #226

Rugged VXI Chassis

Racal Instruments announces the Model 550 rugged VXIbus mainframe. The 550 accepts AC or DC power, with supplies that are interchangeable on-site. 800W of usable power is supplied to a 12-layer backplane. Temperature control consists of fans and heating elements in a closed-loop system. Positive pressurization is maintained, with equal airflow to each slot. Shock mounts and a rugged equipment frame provide shock attenuation up to 30G. Vibration is controlled by the 550's resonant frequency of 9-11 Hz, well below the typical module resonance of 40 Hz. Prices begin at $17,000.
Racal Instruments
INFO/CARD #225

CABLES AND CONNECTORS

Mini-UHF Connector is PC Mounted

RF Industries has introduced the RFU-606 and RFU-606-4 mini-UHF connectors for pc board installation. the RFU-606 has ploypropylene insulation, while the RFU-606-4 uses PTFE. Each has a gold socket and are interchangeable with other mini-UHF units. VSWR is under 1.2:1 past 1.5 GHz. Pricing in production quantities is as low as $1.00.
RF Industries, Ltd.
INFO/CARD #224

PC Board Cable Termination

An inexpensive cable termination connection is announced by ITT Cannon/Sealectro. The Coaxial Terminators consist of two pieces, using a crimp process to attach the cable brain to the terminator. Easy assembly lends itself to a production environment. Styles are available for a variety of RG series cables and cables of similar dimensions.
ITT Cannon/Sealectro
INFO/CARD #223

Miniature Coaxial Cables

Micro-Coax has introduced three UTiFLEX miniature flexible coaxial cables designed to serve wireless applications. The three cables, UFF070A (0.070 in. dia.), UFF092A (0.092 in. dia.), and UFF165A(0.165 in. dia.), employ low density PTFE dielectric and have static bend radius of 0.25 in. (0.50 in. for UFF165A), 77% velocity of propagation and more than 100 dB sheilding effectiveness.
Micro-Coax
INFO/CARD #222

Ruggedized SMA Connectors

Model 1587, 1588 and 1589 from Lucas Weinschel are precision SMA connectors. The connectors have high repeatability and are designed for DC to 26.5 GHz. Auxiliary wrench flats aid in torquing connections without "chain reaction" loosing of multiple component hookups. Inner and outer conductors are gold plated, heat treated, beryllium copper. Models 1587, 1588 and 1589 are SMA female-female, female-male and male-male, respectively.
Lucas Weinschel
INFO/CARD #221

Impedance-Matched 75 Ω Connectors

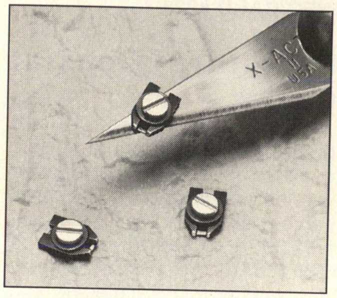
SMB style (snap-on) and SMC style (srew-on) connectors for 75-ohm systems are announced by Applied Engineering Products. Numerous configurations allow replacement of 50-ohm connectors often used in 75-ohm systems. Nickel and gold plated versions are available, with prices starting at $2.54 in 1000s.
Applied Engineering Products
INFO/CARD #220

INFO/CARD 34

DISCRETE COMPONENTS

Chip Trim Cap

Johanson Mfg. announces the introduction of Tiny-Trim™ ceramic chip trimmer capacitors. The 9343 series is available in capacitance ranges from 1.0-3.0 pF to 5.0-40 pF with rated voltage of 25 VDC. Mechanically, the 9343 series capacitors are 0.125



x 0.177 x 0.070 inches. Price is $0.63 each in 1000 piece quantities.
Johanson Mfg. Corp.
INFO/CARD #219

SMD Inductors

The Ecliptek® EC1210 series of high-quality, surface-mount inductors offers precision performance and high reliability in a miniature package. The devices come in molded epoxy cases and in inducatnces from 0.005 to 220.0 µH. The devices are fit, form and function equivalent to the de facto industry standard. Prices start from $0.35.
Ecliptek Corp.
INFO/CARD #218

Precision Molded Chip Resistor

Vishay's new molded SMR3 precision molded resistor uses the company's Bulk Metal® Foil resistive element for low temperature coefficient and long term stability. Construction of the SMR3 offers high moisture resistance, tolerances as tight as 0.01%, plus low noise, low thermal EMF and high frequency operation.
VISHAY Resistors
INFO/CARD #217

High Voltage Ceramic Chip Capacitors

Johanson Dielectrics has developed a new line of surface mount high voltage ceramic chip capacitors featuring working voltages from 500 to 6000 VDC. They are suitable for noise suppression and surge protection applications in high speed data communications circuits and other circuits operating in, or exposed to high voltages. NPO and X7R dielectrics are offered in six EIA standard chip sizes. Pricing ranges from $0.25 to $0.60 in production quantities of 100,000.
Johanson Dielectrics
INFO/CARD #216

TOOLS, MATERIALS & MANUFACTURING

Premixed Epoxy

A premixed two-part frozen epoxy systemis introduced by Syon Corporation. The compounds are packaged in 2 – 60 cc syringe kits, in standard of custom formulations — electrically and thermally conductive, as well as copper- or silver-filled.
Syon Corporation
INFO/CARD #215

Absorber Material

New pyramidal radar absorbers for anechoic chambers are built up from thin layers of absorbing material and punched with cooling holes for high power handling. Forced air cooling can be used.
Cuming Corporation
INFO/CARD #214

Crystal Blanks

Bliley Electric Company has made available quartz blanks of its pure Z growth cultured quartz. AT cut blanks are available rough lapped, finish lapped or polished; in square, round or plano-curved configuration. They can be ordered to a specific thickness, frequency or mode. All have x-ray angle orientation.
Bliley Electric Company
INFO/CARD #213

Metallic-Clad Ferrite

Metallic-clad ferrite substrates are now available from Ceramic Magnetics. These substrates can be used in integrated magnetic and electronic circuits, in tiles for microwave absorption, and other applications. Most commonly used metallization is copper and solder-coated copper.
Ceramic Magnetics
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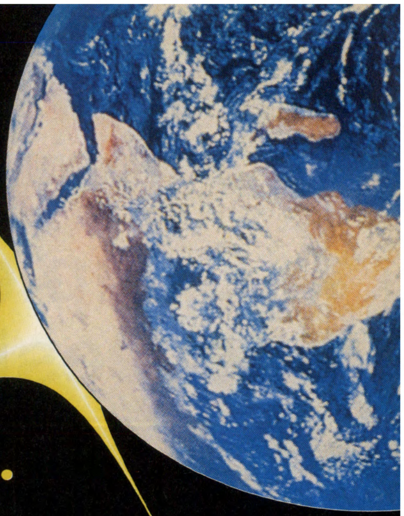

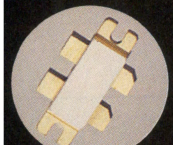

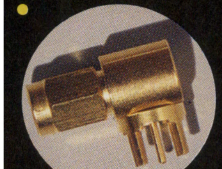

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New Products 1993

M/A-COM

Wireless Bipolar Power Transistors

Device	Frequency (GHz)	Cnfg	Class	Vsup (V)	Pout (W)	Gain (dB)	Eff. (%)	IMD3 (dBc)
PH0810-4	0.850-0.960	CE	AB	24	4	14.0	45	-30
PH0810-15	0.850-0.960	CE	AB	24	15	12.0	50	-30
PH0810-35	0.850-0.960	CE	A/AB	24	35	10.0	55	-30
PH0810-60	0.850-0.960	CE	AB	26	60	10.0	55	-30
PH0810-75	0.850-0.960	CE	AB	26	75-2t	10.0	35-2t	-28
PH0810-150	0.850-0.960	CE	AB	26	150-2t	10.0	35-2t	-28
PH1516-2	1.450-1.600	CE	A/AB	25	2	10.0	35	-32
PH1516-10	1.450-1.600	CE	A/AB	25	10	10.0	40	-30
PH1516-30	1.450-1.600	CE	AB	25	30	10.0	40	-28
PH1516-60	1.450-1.500	CE	AB	26	60-2t	8.0	30-2t	-30
PH1516-100	1.450-1.500	CE	AB	26	100-2t	8.0	30-2t	-28
PH1617-2	1.600-1.700	CE	A/AB	25	2	10.0	35	-32
PH1617-10	1.600-1.700	CE	A/AB	25	10	10.0	40	-30
PH1617-30	1.600-1.700	CE	AB	25	30	10.0	40	-28
PH1819-2	1.780-1.900	CE	A/AB	25	2	10.0	35	-32
PH1819-10	1.780-1.900	CE	A/AB	25	10	9.0	40	-28
PH1819-30	1.780-1.900	CE	AB	25	30	9.0	40	-28

Wireless Power Modules

Device	Frequency (GHz)	Cnfg	Class	Vsup (V)	Pout (W)	Gain (dB)	Eff. (%)	IMD3 (dBc)
PHA1516-2	1.500-1.600	CE	AB	25	2	10.0	35	-32
PHA1516-10	1.500-1.600	CE	AB	25	10	10.0	40	-30
PHA1516-30	1.500-1.600	CE	AB	25	30	9.5	40	-27
PHA1617-2	1.600-1.700	CE	AB	25	2	10.0	35	-30
PHA1617-10	1.600-1.700	CE	AB	25	10	10.0	40	-30
PHA1617-30	1.600-1.700	CE	AB	25	30	9.5	40	-27
PHA1819-2	1.800-1.900	CE	AB	25	2	10.0	35	-32
PHA1819-10	1.800-1.900	CE	AB	25	10	9.0	40	-28
PHA1819-30	1.800-1.900	CE	AB	25	30	9.0	40	-28
PHM960-16	0.920-0.960	CE	AB	26	16	27.0	35	-35

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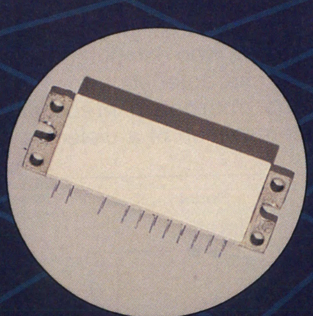

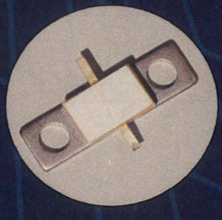

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A Coupled Microstrip Line Review and Design Program

By Brian Brewster
Satellite Microwave & Communications Ltd.

Understanding coupled microstrip lines is important to RF and microwave designers. In addition, high speed digital designers can benefit from knowing how the microstrip characteristics of circuit board traces affect crosstalk between adjacent lines. This article reviews the equations that define coupled microstrip lines, and presents a FORTRAN program that executes coupled line calculations.

A structure consisting of a pair of edge-coupled microstrip transmission lines, as shown in Figure 1, is widely applied to directional couplers, or provides the intrinsic element for a variety of circuit functions and filter networks. Characterization may be accomplished by representing the equivalent circuit (Figure 2) in the form of an ABCD matrix, and defining the voltage and currents in two distinct impedances of even and odd order coupling modes arising from the opposite field polarities [1]. Assuming a perfectly matched condition on all four ports, the coupling between the two lines may be simply written in terms of the two characteristic impedances, Z_{oe} (even mode) and Z_{oo} (odd mode), expressed in dB as:

$$C = 20\log\left[\frac{Z_{oe} - Z_{oo}}{Z_{oe} + Z_{oo}}\right] \qquad (1)$$

and the characteristic impedance Z_0 equals:

$$Z_o = \sqrt{Z_{oe} \cdot Z_{oo}} \qquad (2)$$

Equation (2) applies strictly to a pure TEM-mode transmission medium, however for the quasi TEM-mode operation of a microstrip line where the phase velocities are not equal in the two lines, Z_0 may be written in a frequency dependent form [2]:

$$Z_o = Z_{oe} \cdot Z_{oo} \frac{Z_{oe} \sin\theta_e + Z_{oo} \sin\theta_o}{Z_{oe} \sin\theta_o + Z_{oo} \sin\theta_e} \qquad (3)$$

However for practical applications, and with relatively loose couplings, equations (1) and (2) can be generally applied, especially when considering realizable geometries and the limitation of small values of line spacing, s, which occur for tight coupling values.

A coupler may be realized in the form of Figure 3. The principle design considerations are the determination of w, s and l for a desired coupling ratio and frequency of operation, as a function of substrate relative dielectric constant, E_r, and height, h. This requires the evaluation of the even and odd order mode impedances Z_{oe} and Z_{oo} and the effective dielectric constants E_{effe} and E_{effo}. For maximum coupling, line l, is a quarter wavelength at the mid band frequency, whose length is based on the mean of the guide wavelength, λ_g, for the even and odd order modes.

$$\lambda_{g(e,o)} = \frac{\lambda_o}{\sqrt{E_{eff(e,o)}}} \qquad (4)$$

$$\frac{\lambda_g}{4}\text{mean} = \frac{\lambda_{ge} + \lambda_{go}}{8} \qquad (5)$$

where, λ_0 is the free space wavelength (c/f) and suffixes e,o refer to the even and odd modes respectively.

The accurate analysis of Z_{oe}, Z_{oo}, E_{effe} and E_{effo} was perhaps first done by Bryant and Weiss [3] who computed the static frequency characteristics for an inhomogeneous microstrip medium through the determination of a dielectric

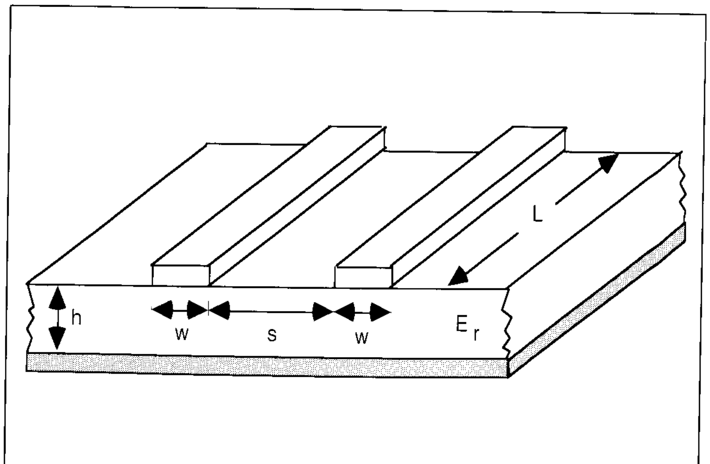

Figure 1. Symmetric microstrip transmission lines.

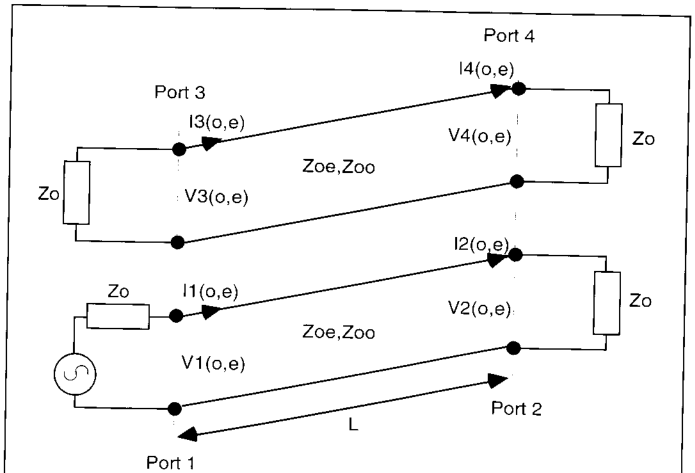

Figure 2. Parallel coupled line equivalent circuit.

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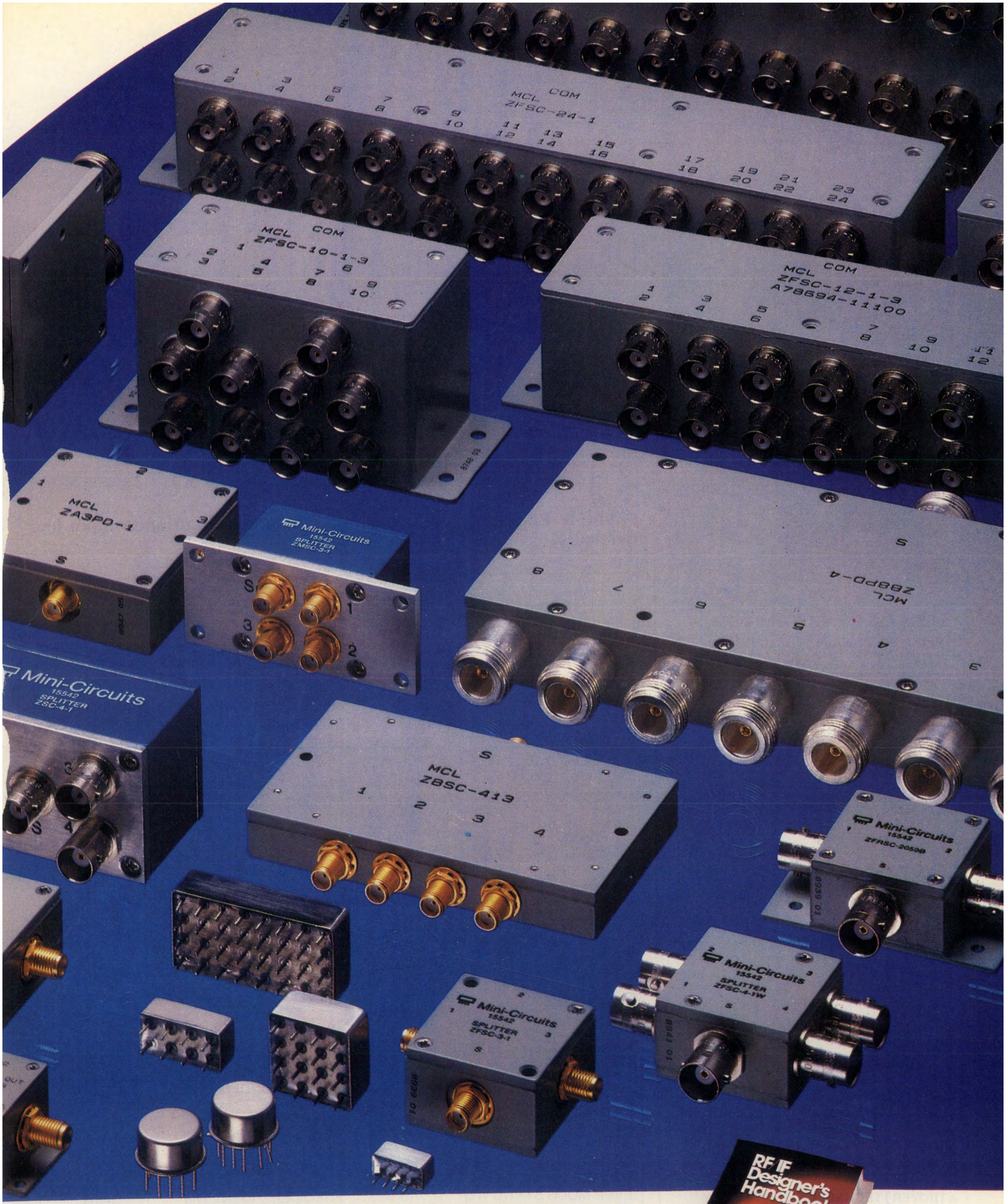
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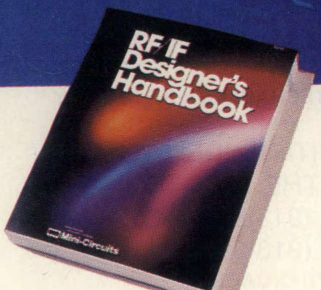
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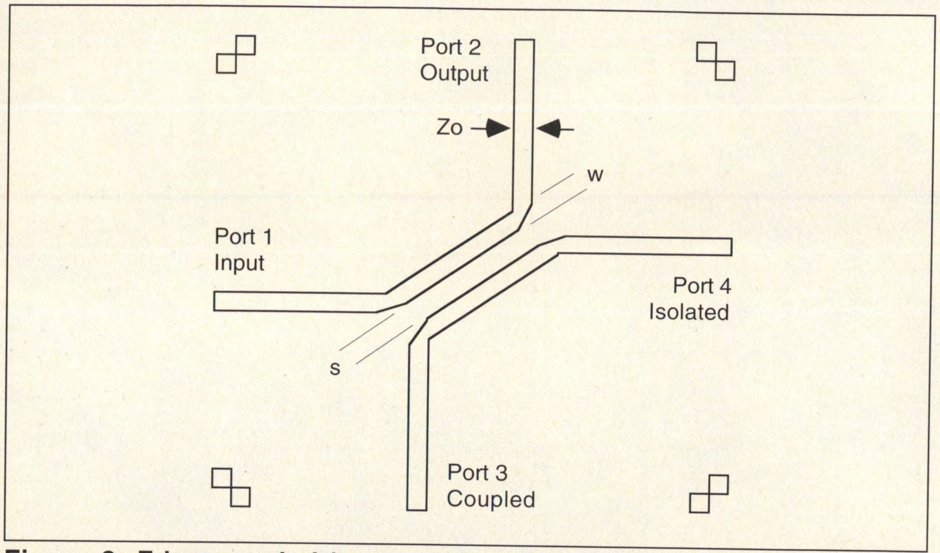

Figure 3. Edge-coupled layout with symmetric input and output line geometry.

Er 10.500 h .6350 w/h step .5000 s/h step 1.0000

w/h	s/h	Zoe	Zoo	Coupling	Zo
2.00	2.00	34.8673	30.6453	-23.82	32.6882
2.00	1.00	36.0765	29.2084	-19.56	32.4613
1.50	2.00	41.1180	36.6432	-24.80	38.8162
1.50	1.00	43.1249	34.2914	-18.85	38.4553
1.00	2.00	50.8221	45.3812	-24.95	48.0246
1.00	1.00	54.0835	41.7244	-17.79	47.5037
.50	2.00	68.5863	61.0615	-24.73	64.7147
.50	1.00	74.1116	55.2591	-16.73	63.9949

Pause.
Please press <return> to continue.

Figure 4. Coupled line anaysis results from the program COUPLER.

W/H .8418 W .5345 S/H .4961 S .3150 Er 10.50
FREQUENCY GHz 1.70

Zo (50 ohms) W/H .9372 W .5951

EVEN ORDER MODE Zoe	63.6747	ODD ORDER MODE Zoo	39.2631
COUPLING C, dB	-12.50	COUPLED LINE IMPEDANCE	50.00
EVEN MODE STATIC Eeff(0)	7.4974	FREQ DEPENDENT Eeff(f)	7.5605
ODD MODE STATIC Eeff(0)	6.0405	FREQ DEPENDANT Eeff(f)	6.0472
FREE SPACE WL (mm)	176.35	GUIDE WL (mm)	67.02
EVEN MODE STATIC Wlg	16.10	FREQ DEPENDENT Wlg	16.03
ODD MODE STATIC Wlg	17.94	FREQ DEPENDENT Wlg	17.93
COUPLED STATIC Wlg/4	17.0194	FREQ DEPENDENT Wlg/4	16.9808

Figure 5. Design optimization result for a –12.5 dB coupler, at 1.7 GHz.

Green's function. Although generally accepted to be very accurate (within 1 percent), the rigorous solution is not suitable for practical CAD and many workers have since developed closed form expressions which approximate the above numerical solution, with varying degrees of accuracy. In this computer program, the results of Zehenter [4] and Kirschning and Jansen [5] are used for E_{effe}, E_{effo} and the corresponding frequency dependent forms to include the effects of dispersion, $E_{effe}(f)$, $E_{effo}(f)$, are employed.

The Coupler Program

COUPLER is a program for the analysis, synthesis and design of coupled microstrip lines. The routine will calculate the even and odd order mode impedances, effective dielectric constants, coupling and microstrip line geometries, from specified substrate parameters and frequency. Optimization of line parameters for specific coupler design values can be evaluated. A simple menu allows user selection of iteration limits and range. The routine is written in standard FORTRAN, implemented with double precision variables for high numerical accuracy. Results are computed within a few seconds on a '386 processor based desktop PC. Error flags are included for input variables which lead to calculation at threshold values.

Figure 4 shows the results obtained from the analysis section with a dielectric constant, E_r, of 10.5. These results agree very well with results calculated using a Green's function built using finite element methods from [6]. E_r values of 10.0 compare with percentage errors in general less than 0.5 percent, with certain combinations of w/h, s/h approaching 1 percent, with those from Bryant and Weiss [3].

The synthesis function implements adjustment of the w/h ratio to achieve a characteristic impedance Z_0 equal to 50 ohms employing eq. (2) over a range of s/h values, while presenting coupling values obtainable.

The optimization section of the program will evaluate w/h, s/h, $\lambda_{g(e+o)/8}$ and associated parameters for a desired input coupling ratio. Figure 5 shows the results obtained for a –12.5 dB coupler design example at 1.7 GHz on RT-Duroid 6010 (E_r = 10.5).

Summary

A simple routine, COUPLER, written in FORTRAN and employing closed form expressions is able to provide a good engineering approximation for the analysis, design and realization of coupled microstrip lines over a wide range of substrate parameters. Microstrip dimensions for directional couplers of the edge-coupled configuration are obtained directly. The results can be extended to form the basis for Lange coupler and filter designs. These approaches are equally valid for considering the degree of crosstalk between runs of "microstrip lines" represented by circuit board traces in high frequency digital logic circuits.

Acknowledgment

The support of Dr. I. Robertson of the Communication Research Department, Kings College London is acknowledged.

Various coupler designs operating at 4 GHz where constructed and tested there.

The COUPLER program is available from the RF Design Software Service. See page 78 for ordering information. **RF**

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About the Author

Brian Brewster has worked in design and project organization roles with satellite communication systems for both terminal equipment suppliers and network carriers. He is currently concluding a research program concerning VSAT modems for the Mphil degree in London, England. He can be reached at Satellite Microwave & Communications, Postbus 10416, 60000GK Weert, The Netherlands.

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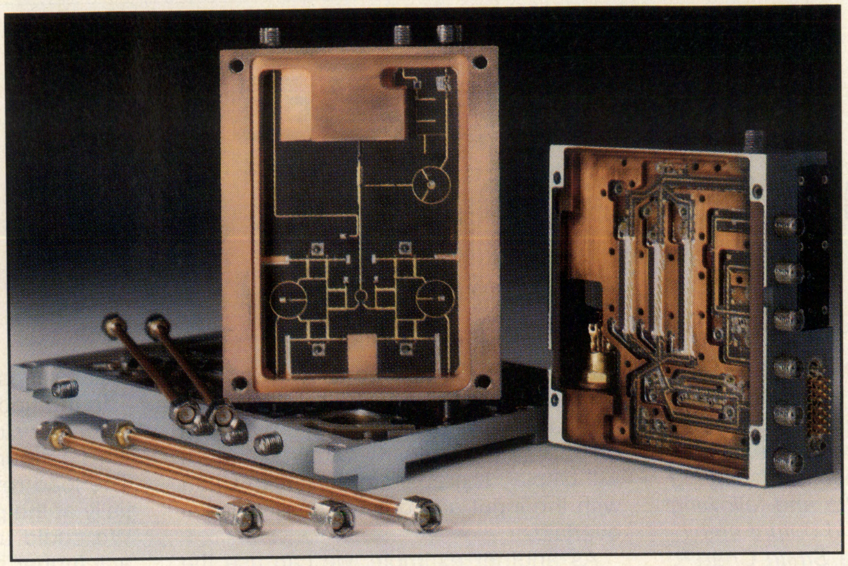

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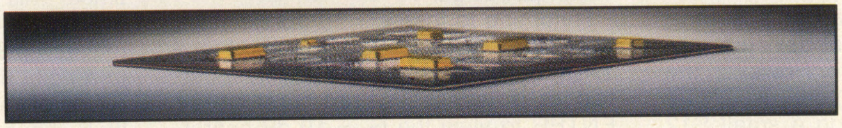

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An S-Parameter Based Amplifier Design Program

By Dale D. Henkes
Philips Consumer Electronics

This program was written to aid in the design of UHF/Microwave amplifier circuits where S parameters are generally used to specify the active device. Stability analysis is performed by displaying the values of K and Delta and following with plots of the input and output stability circles overlaid on the Smith Chart®. Constant gain circles and noise circles can be plotted together on the Smith Chart, allowing tradeoffs between gain and noise to be made visually and interactively.

MAMP was written using Visual Basic for DOS and has a Windows-like interface. The program accepts S parameters only for a bilateral device where S12 is not identically zero. However, the unilateral case (S12 = 0) can usually be approximated by entering a small S12 (S12 < 0.005 typically). The source and load reflection coefficients, corresponding to a user selectable point on the gain circles, are displayed for design of the input and output matching networks.

The operation of the program can best be explained by presenting three examples. First we investigate the stability of a GaAs FET transistor with the following S parameters at f = 1 GHz: S11 = 0.5/180°, S12 = 0.08/30°, S21 = 2.5/70°, S22 = 0.8/–100°. Make sure the "Enter S Parameters" button is highlighted, then press ENTER. One will now be in a position to enter the S parameters directly in polar form as above, or if the data requires, in rectangular form (S real + jS imaginary). After entering the S parameters, press TAB to move to the EXIT button and press ENTER. Now press TAB to highlight the "Stability" button and press ENTER. The values for K and Delta will be displayed as well as the locations and radius of the input and output stability circles. In addition, the message "the stable region is outside the stability circle", is displayed to indicate which area in the Smith Chart is the stable region. In this case K = 0.399 < 1, and so the device is potentially unstable. A message indicating the potentially unstable condition is displayed on the screen under the values for K and Delta. At this point, pressing ENTER or clicking on the "View Stability Circles" button, will display a graphic of the Smith Chart with the input and output stability circles overlaid.

As another example, we can use the same S parameters already entered and design an amplifier with operating power gain of G_p = 10 dB. To do this, simply return to the main menu screen and TAB to the "Power Gain" button and press ENTER. At this point the cursor is positioned to take the given value of 10 dB for power gain.

Returning to the G_p = 10 dB example, type 10 at the cursor prompt and press TAB. Since the "View Gain Circles" button is now highlighted, just press ENTER to view the G_p = 10 dB operating power gain circle on the Smith Chart. Notice the small red box on the gain circle and a small green box on the left hand half of the real axis. These are initial values of Γ_L and Γ_S respectively.

In the upper right corner of the screen are cursor control instructions for moving Γ_L around on the power gain circle. The load reflection coefficient, Γ_L, is initially at the point 0.0984 at angle 97.18°. We could use this point without going any further since we have Γ_L and Γ_S with their corresponding normalized load and source impedances which, when connected to the device, will result in the desired 10 dB power gain.

However, to illustrate moving Γ_L on the gain circle, press "F" for fast and press the down arrow key repeatedly nine times. This moves Γ_L to the point 0.4824 at angle 149.71°. In the process, Γ_S moved upward slightly off the real axis to a point near Γ_L in the upper left hand region of the Smith Chart. Notice how little Γ_S moved in comparison to Γ_L. This is indicative of the relatively small value of S12. The other values related to the movement of Γ_L are displayed together with Γ_L in the bottom right corner of the

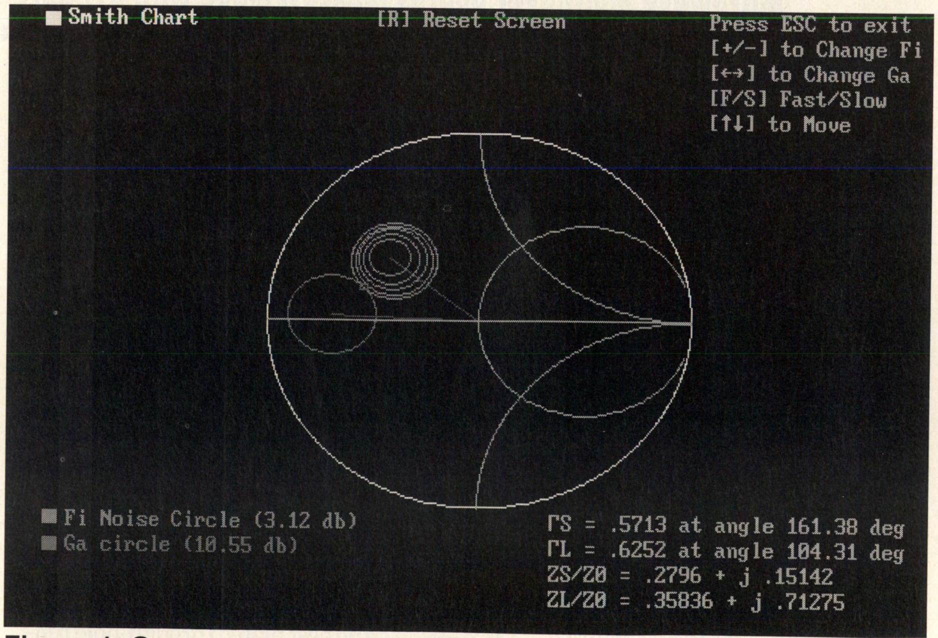

Figure 1. Screen showing intersection of 10.5 dB gain circle and minimum noise circle. The matching impedances shown here are used to design the amplifier of Figure 3.

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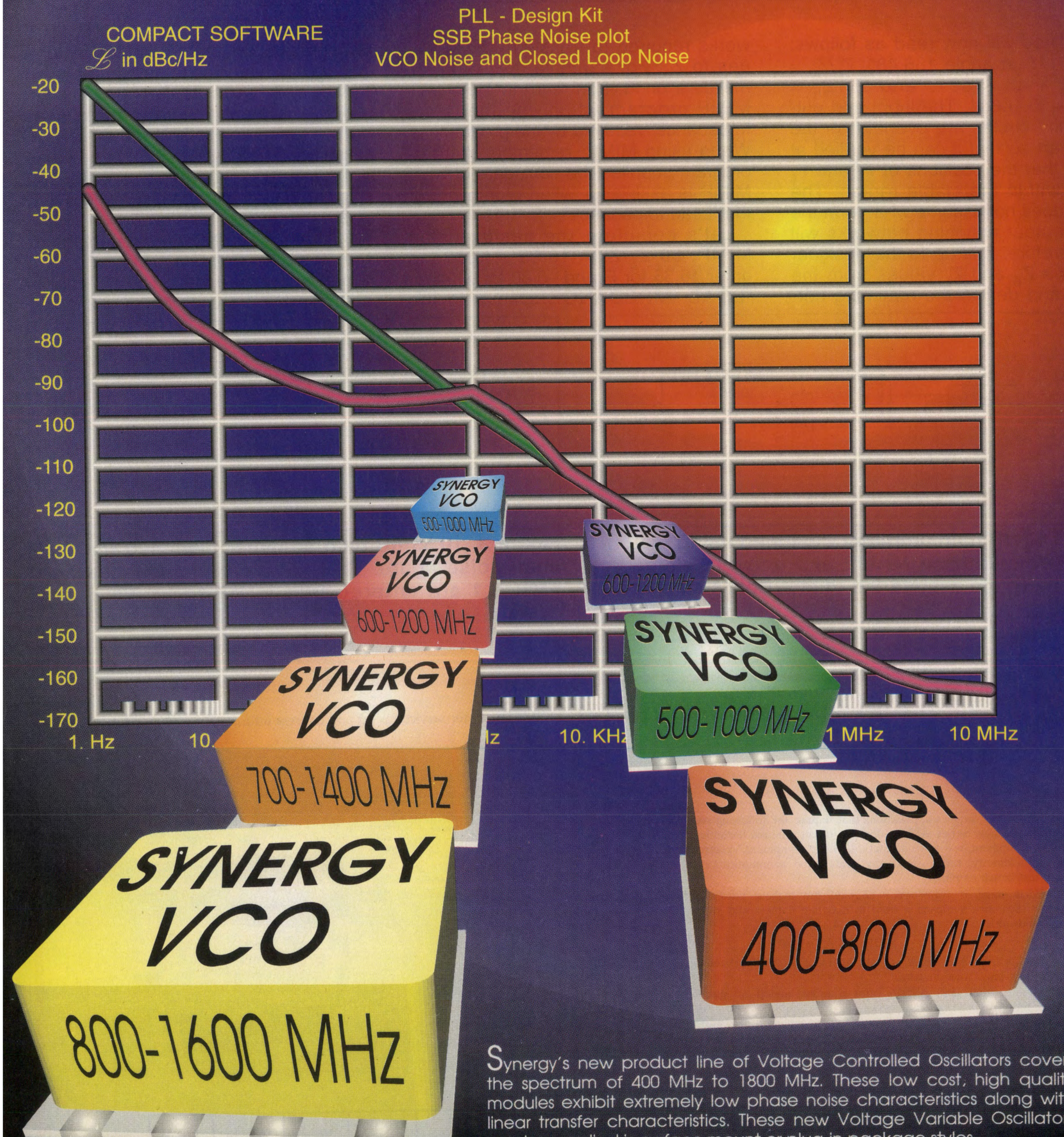

VCO VCO VCO VCO VCO VCO

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screen. They will now read as follows: Γ_S = 0. 5118 at angle 166.48°, Z_L/Z_0 = 0.37139 + j0.23561, and Z_S/Z_0 = 0.32695 + j0.10604. The input and output stability circles are overlaid on the Smith Chart so that it can be verified at a glance that the positions of Γ_L and Γ_S are well within the stable region. The program does not assume what form the input or output matching networks will take. Therefore, it is up to the designer to take the Z_S and Z_L values from the program and design the matching networks required to transform these values to Z_0. For example, this program makes a good "front end" to Motorola's MIMP program, supplying the "Load Impedance" required by the program [1]. When using MIMP, remember that this program supplies the conjugate of what MIMP is asking for when prompting for the "Load Impedance".

Let's complete the design using MIMP: 50 Z_S/Z_0 = 16.35 + j5.30, so enter 16.35 for R and –5.30 for X under "LOAD IMPEDANCE". Move to the schematic entry screen and first enter a microstrip line from the device to a open shunt microstrip line at the 50 ohm input. The exact length of these lines can be tuned on the Smith Chart screen, resulting in lengths of 0.139 and 0.1010 wavelength for the open stub and microstrip transmission line respectively. A similar procedure for Z_L = 18.57 + j11.78 results in 0.1271 wavelength line from the drain to an open stub of 0.1326 wavelength at the load. The AC amplifier schematic is shown in Figure 2.

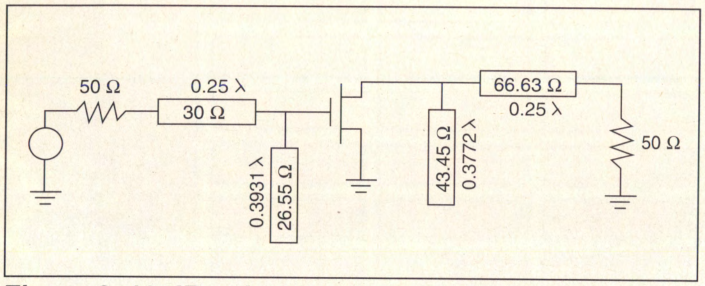

Figure 2. 10 dB gain amplifier designed using matching impedances from MAMP and network elements selected with the assistance of Motorola's MIMP program.

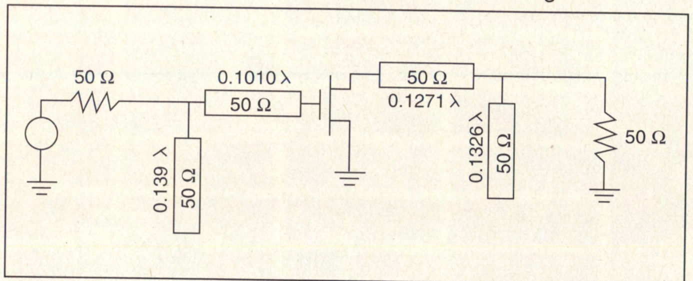

Figure 3. 10.55 dB gain amplifier designed for minimum noise figure. The matching impedances from Figure 1 are entered into MIMP to arrive at this design.

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As a final example, we will consider a 6 GHz GaAs FET amplifier designed for available power gain G_a = 10.55 dB and lowest noise figure [2]. The device parameters are: Sll = 0.641/–171.3°, S12 = 0.057/16.3°, S21 = 2.058/28.5°, S22 = 0.572/–95.7°, F_{min} = 2.9 dB, r_O = 0.542)/141°, R_n = 9.42Ω (R_n/Z_0 = 0.1884).

After entering the S parameters, select "Stability" on the main menu screen and verify that the device is unconditionally stable (K = 1.5 > 1 and Delta = .30 < 1). Then EXIT and select "Noise" from the menu screen. Enter 10.55 for G_a, 2.95 for F_i, 2.9 for F_{min}, .542/141 for Γ_0 and .1884 for R_n. Select "View Noise and Gain Circles". Press "+" 4 times until the noise circle just intersects the 10 dB Ga circle. Now press "F" and then press the up arrow key eight times until the Γ_S indicator moves to the intersection of the two circles. We now have the needed input and output impedances. They are: Z_S/Z_0 = 0.2796 + j0.15142 and Z_L/Z_0 = 0.3584 + j0.71275 corresponding to Z_S = 13.98 + j7.57 and Z_L = 17.92 + j35.64 ohms. Again, using MIMP or manual Smith Chart techniques we arrive at the AC amplifier schematic of Figure 3.

Alternately, we could have designed for F_{min} by entering 2.901 for F_i (use F_i slightly larger than F_{min} so that the noise circle will have a finite but small radius). Then while viewing the G_a circle on the Smith Chart press the left arrow key repeatedly until the G_a circle grows (less gain) and intersects with the center of the noise circle. Now move Γ_S to the point of intersection. Of course $\Gamma_S = \Gamma_0$ = 0.54/141° at this point, and we can read the Z_S and Z_L values for completion of the design. The final value of Available Power Gain can be read directly from the screen as G_a = 9.35 dB.

This program is available from the RF Design Software Service, see page 78 for ordering information. **RF**

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1. Dan Moline, "MIMP Analyzes Impedance Matching Networks", *RF Design*, January 1993.

2. Design based on Hewlett-Packard Application Note 970, February 1978, as presented by G. Gonzalez in *Microwave Transistor Amplifiers*, Prentice-Hall, 1984.

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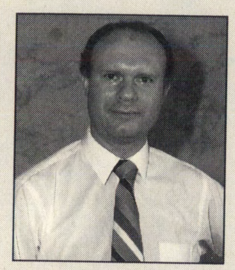

Dale Henkes is a Project Engineer with Philips Consumer Electronics Co. in Knoxville, TN. He currently works on tuning systems design for high volume television products. He recently has taken graduate course work in at the University of Tennessee, on the topic of microwave networks. He received a BS degree in Engineering from Walla Walla College in 1977 Prior to his employment with Philips, he worked as a design engineer for the Boeing Aerospace Co. in Seattle, WA. He may be reached at 1916 Plumb Ridge Rd., Knoxville, TN 37932.

The Search for Smaller, Cheaper Components

By Gary A. Breed
Editor

Passive components are not glamorous when compared to the newest RF synthesizer IC or super low noise transistor. However, the inductors, capacitors and resistors that make up the majority of passive components are undergoing the same kind of intensive product development as the attention-getting active devices. Despite their reputation as "commodity" components, passives can make or break a design, particularly when it goes into production.

Emphasis on cost is a result of growth in commercial and consumer products, where the cost of the finished product must be competitive. A few companies have always addressed this issue in the RF realm, such as Toko America, Murata Erie, Coilcraft, TDK, Philips, Siemens and others. In some cases, two lines have existed for similar products: one for general purpose devices, the other for RF-specific devices. In the past, components with performance requirements and additional testing for RF characteristics were reserved for a special place in the component catalog.

Customers now want RF-capable components at general purpose prices, and the companies are responding. Automated component manufacturing now includes testing, which allows all devices to come from the same line. For discrete components, many "commodity" grades are acceptable for RF use.

Integrated Passive Components

Components with specific RF performance or functionality are still required, in spite of demands for mass-market pricing. Also, some of the large-scale manufacturers have developed integrated passive devices such as EMI filters, RF bandpass filters, LC resonators, and RC networks. If a designer can find the right specifications among these devices, or if a design can be adapted to include a standard component, these assemblies can greatly reduce design time and cut down on the number of manufacturing operations.

Some examples include using an integrated RC network as the feedback element in a PLL loop filter or in an audio or control operational amplifier stage. Or, a manufactured resistive attenuator can replace three discrete resistors in a controlled-impedance coupling network. A series of untuned, wideband amplifier stages can include filters between stages instead of just coupling capacitors, reducing response to unwanted signals for a lower cost than a traditional multi-element filter.

Not all engineers are fully aware of the level of integration that is now available in passive components. At the recent RF Expo East, Mark Brooks of Thin Film Technology described a process which allows resistors, capacitors, spiral inductors and small-geometry microstrip lines to be combined using low-cost ceramic multilayer construction. Using such technology, SMT filters, matching networks, and other configurations can be built, limited only by the designers imagination and the physical constraints of the process (e.g. capacitor size, inductor value, microstrip line length).

RF-Specific Performance

There are few performance areas that are specific to RF applications, where proper circuit operation can not be achieved using general-purpose components. The most obvious application in in high power RF, where companies like American Technical Ceramics and Dielectric Laboratories have developed a specialty. Other companies with significant high power capacitor lines include Murata Erie, AVX and Philips Components

RF current handling is not an issue in small-signal circuits and at power levels under one watt or so. But at high power, several factors must be considered. First, high currents require large conductor areas, which may conflict with requirements for small capacitance values. The heat generated must not affect the plate metallization or the bonds within the capacitor. To cope with these requirements, RF capacitor manufacturers have developed optimum ceramic and porcelain dielectric materials (or still use mica dielectric), plus unique metallization and assembly techniques to minimize failures when the device is stressed by high RF currents.

High power inductors are a problem in two areas: high inductance values and microstrip inductors. Large-value inductors wound in solenoid form are too large for many applications, and usually have large inter-winding capacitance. They also generate sizeable magnetic fields. The typical solution today is to use an iron power or ferrite toroidal core to increase the amount of inductance per turn of the coil winding. A toroidal core contains the magnetic field, as well, reducing radiation to, or pickup from, adjacent circuitry. However, the thermal characteristics of these ceramic materials needs close attention. Temperature rise due to the RF power level varies with the composition of the core. A tradeoff is almost always required between increased inductance per turn and core losses (and, therefore, core heating).

Microstrip line places physical limitations on RF power handling, as well. The cost savings of implementing a circuit in microstrip can not be obtained if the conductors on the substrate can not handle the power. Alternatives usually include coaxial line sections, cavity structures, or stripline assemblies built piece-by-piece. The only way to avoid the necessity of these costly alternatives is to avoid the use of inductors whenever possible. The active device manufacturers are working on this problem by constantly improving their methods for internal matching in power devices.

Summary

Passive components manufacturers are addressing the issue of cost through improved manufacturing techniques and with greater functionality. Designers can cut component cost and assembly time with these advanced passive components. *RF*

Balanced Meissner Oscillator Circuits

By Nick Demma
Honeywell Technology Center

Though less familiar than the Colpitts and Hartley circuits, Meissner oscillators deserve wider recognition for their many desirable characteristics. This article describes their general advantages and also lists some features that make them particularly well suited to some special applications.

The improbable oscillator shown in Figure 1 has many virtues in addition to simplicity: there are no biasing resistors reducing the in-circuit Q of the resonator; there are no active devices generating noise; there is no supply voltage ripple creating phase noise; and there are no voltage-dependent semiconductor capacitances causing frequency drift. Its high quality can be observed by hitting it with a charged capacitor and watching it ring. Although its lack of a sustainable output renders it useless in most applications, this circuit does illustrate an important design principle; the way to build a good oscillator is to start with a good resonator and then to avoid ruining it with its associated electronics.

Obvious though this principle may be, it is not exemplified by the typical Colpitts oscillator shown in Figure 2. This circuit diminishes the capabilities of the resonator in many ways. If the values of the biasing resistors are small enough to provide good bias point stability, then they also reduce the Q. The amplitude of the waveform in this circuit is usually limited when the collector-base junction

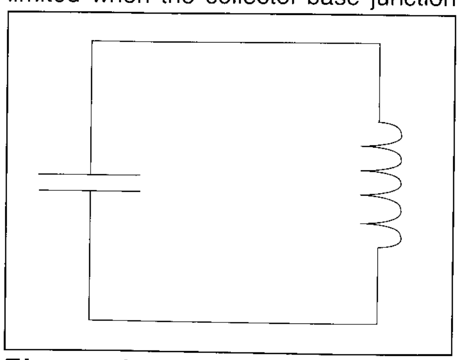

Figure 1. A high quality, but unsustainable oscillator.

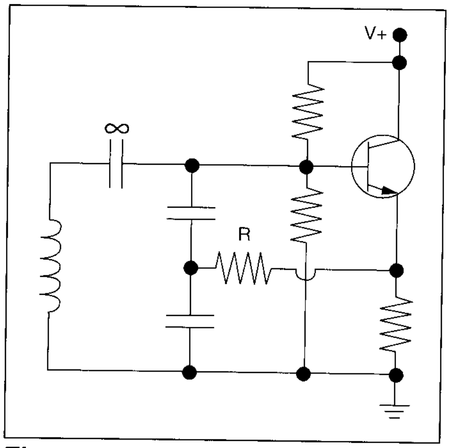

Figure 2. Colpitts oscillator

becomes forward biased. This low impedance momentarily removes energy from the resonator and further reduces the Q. The collector-base capacitance, which varies with supply voltage, appears directly across the resonator as a frequency-determining component. Although it is easy to get this circuit to oscillate, the distorted waveforms usually observed in this kind of oscillator suggest that the transistor is using a rather heavy handed approach to sustaining the oscillation. The regeneration-limiting resistor R can reduce the feedback to obtain a nice sine wave but only at some risk of quenching the oscillation.

Circuits

Unbalanced Meissner oscillators are not the subject of this article, but the circuit shown in Figure 3 is included anyway as a simple example of what a Meissner oscillator is. It shows a clear attempt to avoid harassing the resonator in accordance with the design principle mentioned earlier.

A fully balanced 7.1 MHz Meissner oscillator with 0 and 180 degree sine waves is shown in Figure 4. This is a positive feedback oscillator that works by injecting energy into the resonator through the three-turn links and extracting the feedback from the single turn links at the bases of Q1 and Q2. The dual current mirrors serve the aesthetic purpose of keeping the schematic symmetrical in addition to pulling about 4 mA from the differential pair formed by transistors Q1 and Q2. If the base of Q1 is about 0.1 Volt more positive than the base of Q2, most of the 4 mA will flow through the three-turn link on Q1's collector and will couple energy into the resonator while Q2 is cut off. Very little energy is taken from the resonator to provide the voltages that drive the bases and sustain the oscillation. Excellent resonator isolation is obtained both by using large turns ratios when coupling energy into and out of the resonator and by using Q1 and Q2 as current sinks that have high impedances even when they are on. The bi-phase outputs can be taken from the taps that provide the signals for the bases of Q1 and Q2, from across capacitor C, or from various tap points in between. Suitable FET buffers should be used when taking the outputs from further up the resonator or when using the outputs to drive low impedance loads. Although the amplitude of the outputs can be increased by tapping further up on the resonator, it is better to keep the tap

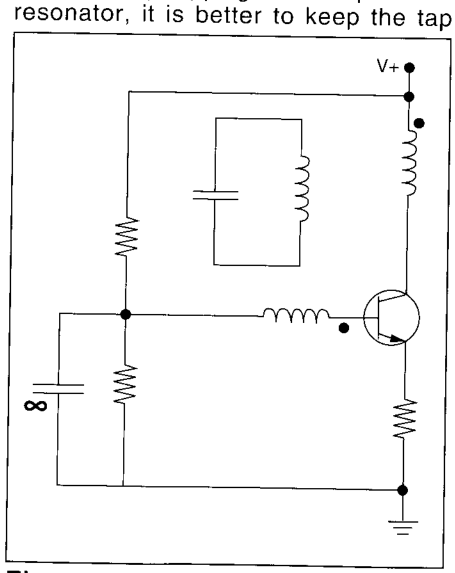

Figure 3. An unbalanced Meissner oscillator. All coils are coupled

point low and to increase the amplitude by increasing the DC bias currents. It is easy to create 50 Vpp sine waves across the resonator in this kind of circuit.

Applications

For each of the oscillator's following virtues, a practical application is described in which the virtue is important.

First, the frequency of oscillation is determined predominantly by the resonator's inductance and by capacitor C. The transistor capacitances and the inductance of the coils on the collectors of Q1 and Q2 have little influence on the frequency, and this circuit's sensitivity to supply voltage variations was only 1.2 ppm/volt. These qualities are useful whenever a stable, low phase noise oscillator is needed.

Second, there are no transistor capacitances or fixed capacitors across the resonator capacitor. This is important in battery powered electronically tuned AM radios where a wide ranging oscillator must be tuned with low voltage varactors. The various turns ratios would have to be changed to prevent overdriving the varactors.

Third, high purity sine waves are available with third harmonic levels that are typically more than 60 dB below the level of the fundamental even when hard switching the differential pair to guarantee oscillation. This is important when using linear multipliers like the AD 734 for synchronous demodulation in homodyne radios or lock-in amplifiers. In these applications, the harmonic sensitivity of hard-switching mixers is unacceptable. Low distortion sine waves are needed to make good use of the multiplier's linearity.

Fourth, the amplitude of the resonator's sine wave can be controlled either by the differential pair's non-linearity (using low loop gain) or by the level of current in the current sink (using high enough loop gain to cause hard switching). Hard switching with differential base voltages of about 2 Vpp is generally preferable for both short and long term frequency stability. If soft limiting is used, the signal levels at the bases of Q1 and Q2 are suitable for using switching demodulators like the LM1496 as inexpensive linear multipliers.

Fifth, the circuit can be converted into an amplifier. By using a ferrite rod instead of a toroid, one can construct a fully balanced regenerative antenna for use on AM broadcast or shortwave frequencies. Variable current sinks can be used for regeneration control.

Sixth, the oscillator can be crystal controlled. Figure 5 shows the modified circuit in which the dual current sinks are actually needed. The crystal's source and load impedances are low for good in-circuit Q while the 0.1 Vpp drive level across the crystal results in only 17 microwatts of power dissipation. By changing the bias current level and the various turns ratios, it is possible to independently adjust the source impedance (small signal emitter resistance divided by emitter current = $.025/I_e$), the drive level across the crystal, the signal level across the resonator, and the output signal level. The resonator must be tuned to the crystal's series-resonant

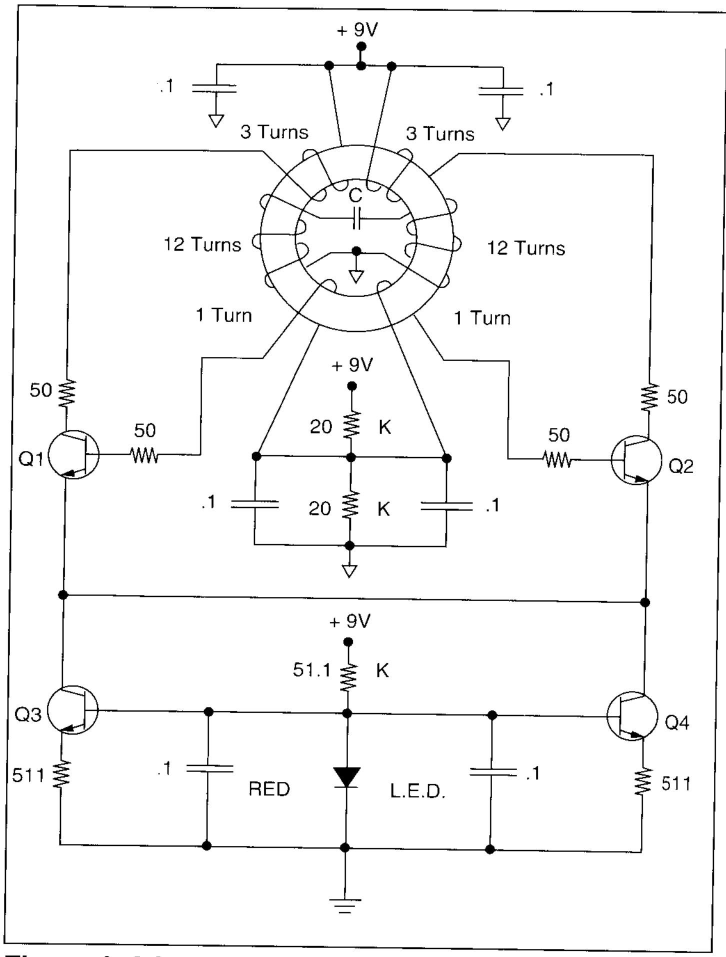

Figure 4. A balanced, 7.1 MHz Meissner oscillator.

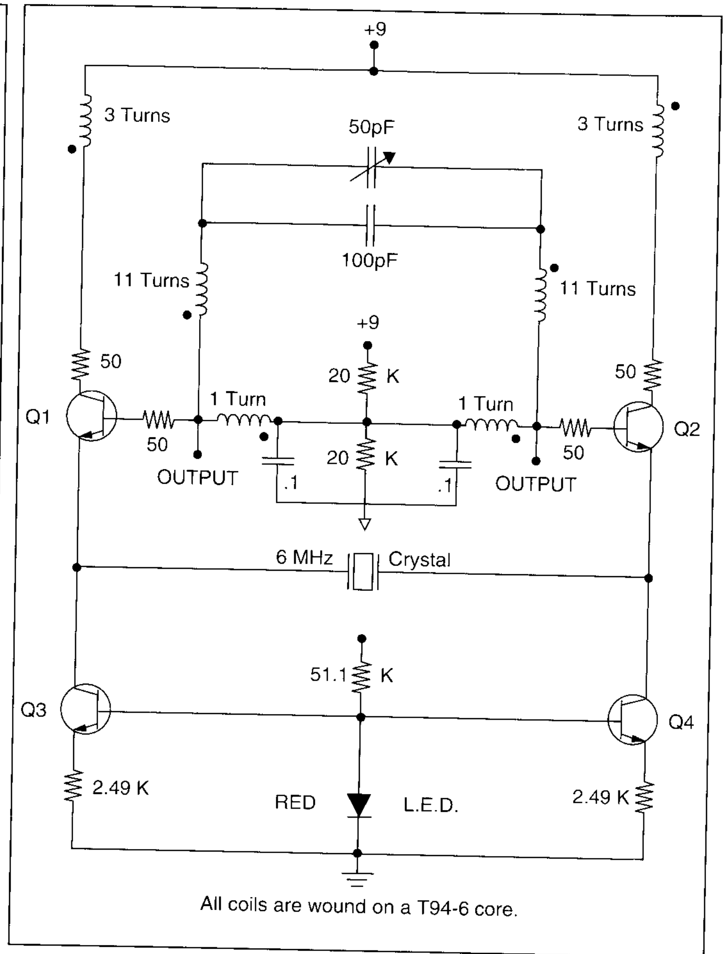

Figure 5. A crystal controlled, balanced Meissner oscillator.

frequency and it must remain stable because a dispute over the operating frequency will kill the oscillation. Due to the excellent isolation, the in-circuit Q of the resonator in this circuit is 225. It is worthwhile to reduce this Q to increase the crystal's authority over the frequency.

Construction Tips

When using low-permeability iron powder toroids, the winding technique can have a profound influence on the behavior of the oscillator. The recommended arrangement is shown in Figure 4. If hard switching is used, the short rise time of current through the coils can generate significant voltage spikes at the collectors of Q1 and Q2, so these coils are best kept away from everything else to avoid capacitively coupling the spikes. The 50 ohm resistors prevent parasitic VHF oscillations from occurring across the coupling inductances, which can be a problem in Meissner and Hartley oscillators. As the operating frequency is increased, balancing the physical layout of the circuit becomes increasingly important. A perfectly symmetrical layout is good practice anyway both for artistic reasons and for rejection of common mode disturbances. If only one sine wave output is needed, an unbalanced design can simplify the winding of the toroid while retaining many of the oscillator's inherent virtues (see reference 2 for details). The winding can also be simplified by using the non-isolated scheme shown in Figure 5, but this is not a true Meissner oscillator. *RF*

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1. I do not have Alexander Meissner's original paper but would like to get a copy of it for my recreational research files. It was probably published somewhere in the 1920's.

2. W. Hayward and D. Demaw. *Solid State Design for the Radio Amateur*, The American Radio Relay League, 1986, pp. 125-128

About The Author

Nick Demma is a research scientist at the Honeywell Technology Center. He designs and builds a wide variety of circuits and systems in support of the various research projects. He can be reached at Honeywell, MN65-2600, 3660 Technology Drive, Minneapolis, MN 55418, or by phone at (612) 951-7714.

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Transmitters/Receivers

An Improved FM Exciter Using Direct Digital Synthesis, Thomas Hack, March 1992, p. 95.
A Simple Road Information Transmitter, Thomas Hack, July 1992, p. 31.
Fundamentals of Receiver Design for Part 15 Applications, Bernard Kasmir, September 1992, p. 35.
Algorithms for Computing Phase and AGC in Digital PLL Receivers, Guglielmo A. Ferrari and Roberto Cali, October 1992, p. 33.
The Design of a Sampling Gate Receiver, Jonathan Y.C. Cheah, November 1992, p. 66.
The Mechanical Filter in HF Receiver Design, William E. Sabin, March 1993, p. 43.
Implementation of a C-QUAM AM-Stereo Receiver Using a General-Purpose DSP Device, Dion Messer Funderburk and Sangil Park, June 1993, p. 32.
Broadcast Radio Uses Home-Made Components, Jim Escoffier, June 1993, p. 84.
A Program for Design and Analysis of Receivers, John Donohue, October 1993, p. 101.

RF software

AT&T Libraries

AT&T Microelectronics and Cadence Design Systems have announced the availability of a complete analog design solution optimized for AT&T's bipolar integrated circuit technologies. AT&T will offer comprehensive design kits for use with Cadence's Analog Artist™ Design System and an interface to AT&T's Advice™ simulator.
AT&T Microelectronics
Cadence Design Systems, Inc.
INFO/CARD #211

Model Library

A new library of more than 65 function blocks for designing communications systems of any complexity is now available to support the DSP Framework™ design environment from Comdisco Systems. The library contains modulator and demodulator, adaptive equalizer, wireless channel and other models for digital cellular phones, pagers, mobile communications, satellite-based systems and other wireless design applications. The communications library priced at $5000 for the first node.
Comdisco Systems
INFO/CARD #210

Simulators

The new RFDesigner® v.1.6 provides post-processing of the simulated circuit's s-parameter set, allowing fast generation of any single (or any combination of) circuit performance function, as well as its time domain response in graphical and numerical form. RFLaplace™, also from ingSOFT, is a versatile electromagnetic simulator for uniform TEM mode structures. Unit prices for RFDesigner v.1.6 and RFLaplace are $1500 and $1200, respectively.
ingSOFT Ltd.
INFO/CARD #209

Network Design

Comsearch introduces a fully integrated family of engineering software for wireless communications networks. Netwok Planning, Spectrum Sharing, Microwave Interconnection, and Cell Design will help to design PCS, mobile, microwave satellite and other communications systems. Each module is portable across all UNIX workstations, can operate stand-alone and has an advanced graphical user interface (GUI).
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RF design

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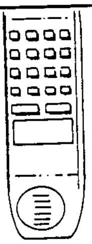
RF recruitment



...YOUR CAREER

Cellular Engineers: Design/develop RF and analog circuits for high capacity cellular systems. Requires minimum of 2 years experience in any of the following: DSP, ASIC Design, CAE Development, Digital Modulation, Digital Mobile Communications, Channel Equalizers, Transmitter-Receiver-Synthesizer or Audio Design, Digital Signal Processing.

ASIC Digital H/W Design Engineers: Responsible for creating a variety of designs for TDMA, CDMA and mixed signal applications. Successful candidates will have a BSEE with 4 or more years circuit-level development experience in ASIC, VISI, CMOS and GaAs top down design including interfacing to microcontrollers. Duties will include system integration/test from product conception to manufacturing. Experience with commercial DSPs in hardware and system integration into communications products is a plus.

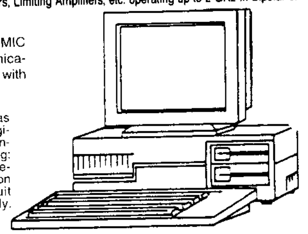
Analog Design EE • BSEE min., MSEE perf. • 5 to 7+ years design exp., professional maturity • Strong (85% of project history) in analog board level design, DC to RF (500 MHz to 4 GHz) with module level exp., use of RF and microwave test equip. at these frequencies. • Exp. in analog/digital interfacing and in design tradeoffs in processing signals in analog and digital sections • Self-starter; able to plan, design, and prototype proof-of-concepts, a hands-on person • Adaptable personality (a specialist in one or more application areas but welcoming any opportunity to dive into another) • Analog and digital circuit breadboarding and debugging exp.

RF Design Engineer: Responsible for design of analog and RF systems and circuits for consumer and commercial digital wireless products. Five to ten years experience in RF systems analysis and design. Experience with low-cost design techniques for frequency synthesizers, power amplifiers, up/down converters and baseband circuits for digital communications systems. Must be able to derive RF systems and module requirements to meet overall performance and cost goals. Familiarity with time division duplex or CDMA a plus.

Satcom Systems: Familiarity with equipment and systems utilized by broadcast organizations for the distribution of analog and digital audio programming via satellite. Familiarity with the following specific technical areas: communications systems design, data communications principles, modem design, satellite communications receiver design, and satellite link budget calculations.

RFIC Design: MS or PhD in Electrical Engineering with minimum 5 years related experience is preferred. The candidate should have a good knowledge and experience in Linear Bipolar High Frequency IC design and measurement techniques to design IC's like Amplifiers, Mixers, Oscillators, VCO's, Prescalers, Synthesizers, Limiting Amplifiers, etc. operating up to 2 GHz in Bipolar or BiCMOS technologies.

MMIC Design Engineer: Develop L/S band GaAs MMIC power amplifiers for commercial wireless communications. Requires:M.S. or BSEE, +2 years experience with GaAs MMIC design, simulation, packaging and test.



RF Manager: Two-way interactive multimedia service has a need for "hands-on" senior radio frequency (RF) engineer with 10 years experience for complete project responsibility. Individual must have experience in the following: project mgmt; RF test equipment and standard procedures; and RF system design: spurious, noise, modulation and demodulation. Desired skills include: RF circuit design; digital logic H/W; and manufacturing assembly. BSEE

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INFO/CARD 60

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Several of our growing commercial major electronic companies located in the southeast have asked us to identify candidates in the mid-level experience range.

1) BSEE – to design analog/RF circuitry for receiver/transmitters in 300-900 MHTZ – P-Spice
2) BSEE – RF Test Engineer – 50 MHTZ to 1 GHTZ. Spectrum analyzers, net work analyzers; HP Basic
3) BSEE – High power amplifier design engineer – Class AB 100-200 watts; 800-900 MHTZ
4) BSEE – Power supply design engineer – switch mode; power factor correction

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Stan Deckelbaum

INFO/CARD 61

RF career opportunities

SENIOR RF DESIGN ENGINEER

Five to ten years experience in RF systems analysis and design. Experience with low cost techniques for frequency synthesizers, receiver design, and transmitter design in the 900 and 2400 MHz bands. Must be proficient with RF modeling software such as Touchstone or Eagleware and familiar with surface mount technology.

AXONN

AXONN CORP. is a rapidly growing RF Spread Spectrum product R&D company. 101 W. Robert E. Lee Blvd., Suite 202, New Orleans, LA 70124 FAX: 504-282-0999 Phone 504-282-8119.

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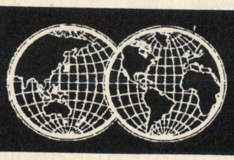

RF DESIGNERS

Northern Telecom and Bell-Northern Research offer comprehensive wireless voice and data communications systems solutions that serve both global cellular and personal communications services (PCS) markets. Northern Telecom Companion PCS systems are recognized on three continents as the leading wireless business solution for office and campus environments. Our digital cellular products are meeting demanding customer needs from North America to Europe and the Pacific Rim.

We have several opportunities available for engineering professionals to work in a dynamic team environment, in the following roles: Analog IC Designer, RF IC Designer, Surface Acoustic Wave Designer, Radio Systems Designer, and H/W RF Designer. As a member of our scientific staff, your responsibilities will cover a wide spectrum from prototype simulation, design and testing to circuit pack design and manufacturing introduction. Design activities are supported by various IF/RF/CAD tools, workstations and well-equipped laboratories.

Successful candidates will have a degree - at the master's level for some positions - in a related field, plus excellent analytical, problem-solving and team skills, and two to five years of relevant, combined experience. We are looking for candidates with experience to include Low Noise Amplifier Design, Power Amplifier Design, IF Circuit Design, Up/Down Converter Design, and Microwave Design.

Due to the international deployment of these products, we would also be interested in candidates with fluency in languages such as Mandarin, Spanish, French, Portuguese and/or German.

We offer competitive salaries and the opportunity to join a company committed to developing new initiatives in communications technology. Please forward your resume, quoting Stop Number 11012, to: Northern Telecom/Bell-Northern Research, National Resourcing Center, P.O. Box 3511, Station C, Ottawa, Ontario, Canada, K1Y 4H7. *If you have applied for employment with Northern Telecom and Bell-Northern Research through your campus placement office, there is no need to re-apply.*

Northern Telecom and Bell-Northern Research are equal opportunity employers and support a smoke-free environment.

Northern Telecom and Bell-Northern Research have the vision, the imagination and the resources to take on the world of telecommunications ... and become the world's leading suppliers of telecommunications equipment by the year 2000. Join our team and explore the infinite possibilities of invention and innovation.

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Advertiser Index

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B. Paid and/or requested circulation		
1. Sales through dealers and carriers, street vendors and counter sales.	N/A	N/A
2. Mail subscription (Paid and/or requested)	40,415	40,611
C. Total paid and/or requested circulation (Sum of 9B1 and 9B2)	40,415	40,611
D. Free distribution by mail, carrier or other means (samples, complimentary and other free copies)	1,639	2,955
E. Total Distribution (Sum of C and D)	42,054	43,566
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10. I certify that the statements made by me above are correct and complete.
(Signed) Arthur Sweum, Senior Vice President

RF literature

Cable Selection Guide

C.E. Precision Assemblies has released guidelines for microwave engineers concerning relative merits and problems associated with silicon dioxide, PTFE, and Isocore coaxial cables. Characteristics compared include: radiation hardening, phase stability, reliability, cost and availability.

C.E. Precision Assemblies
INFO/CARD #206

Amplifier Catalog

JCA Technology has released a 36-page catalog listing their 10 MHz to 26 GHz GaAs FET amplifiers. Standard amplifier models include ultra-low noise, standard octave bandwidth, multi-octave bandwidth EW and ECM, and narrow radar and communications radars. Custom amplifiers are also listed.

JCA Technology
INFO/CARD #205

Switch Catalog

Narda's complete family of RF and microwave electromechanical switches is covered in their 100-page catalog. Both stocked and custom electromechanical switches are indexed within the catalog.

Loral Microwave – Narda
INFO/CARD #204

RF companies

Index of companies and institutions mentioned in this issue's editorial pages.

Frequency Control Guide

SaRonix has announced a new a six-page Frequency Control Products Selection Guide, featuring crystal and oscillator products, including traditional quartz crystal and crystal oscillator components, SMT packages, and advanced technology oscillators and VCXOs with frequencies up to 622 MHz.

SaRonix, Inc.
INFO/CARD #203

Frequency Multipliers

Miteq's 24-page catalog contains information on our family of passive and active frequency multipliers. The catalog presents both product specifications and typical applications for standard products. Detailed electrical and mechanical specifications are provided.

Miteq
INFO/CARD #202

Crystal Catalog

A new design guide an catalog of frequency control crystals has been published by JAN Crystals. The twelve-page catalog is oriented to designers, engineers, and manufacturers, and includes extensive charts, graphs, formulas, and technical information.

JAN Crystals
INFO/CARD #201

Amplifier Handbook

AML announces immediate availability of a revised edition of their Communication High Intercept Amplifier Handbook. The new edition contains performance and mechanical information on over 160 RF wide band products that operate from 100 kHz to 2000 MHz. The book also contains an expanded application note section, with discussions of IMD cancellation and 2-tone IMD measurements.

AML Inc.
INFO/CARD #200

Cable Brochures

Time Microwave announces the availability of two brochures and a price list covering the LMR™ and T-COM™ series of flexible communications cables and connectors. The T-COM cables feature an all silver RF path; the LMR cables are low loss, cost effective alternatives to corrugated copper cables.

Times Microwave Systems
INFO/CARD #199

Tunable Filter Catalog

The Tunable (Solid State) RF Bandpass Filter Catalog is now available from Pole/Zero. This catalog presents their modular, digitally-tuned, RF filters and pre/post-selectors. The 18-page catalog contains information on their Mini-Pole™, Maxi-Pole™ and 10 W Power-Pole™ hopping filter product lines. Additionally, it covers their digitally-tuned HF filters, PC/AT RF pre/post selectors for personal computers, and PIN diode switches.

Pole/Zero Corp.
INFO/CARD #198

Crystal/Oscillator Catalog

A 79-page catalog from Fox Electronics describes their lines of crystals, oscillators, real-time clocks, and crystal filters. Included in the catalog are descriptions of Fox's 3.3 V tri-state SMT oscillator and their programmable oscillators. Engineering definitions and design notes are also included.

Fox Electronics
INFO/CARD #197

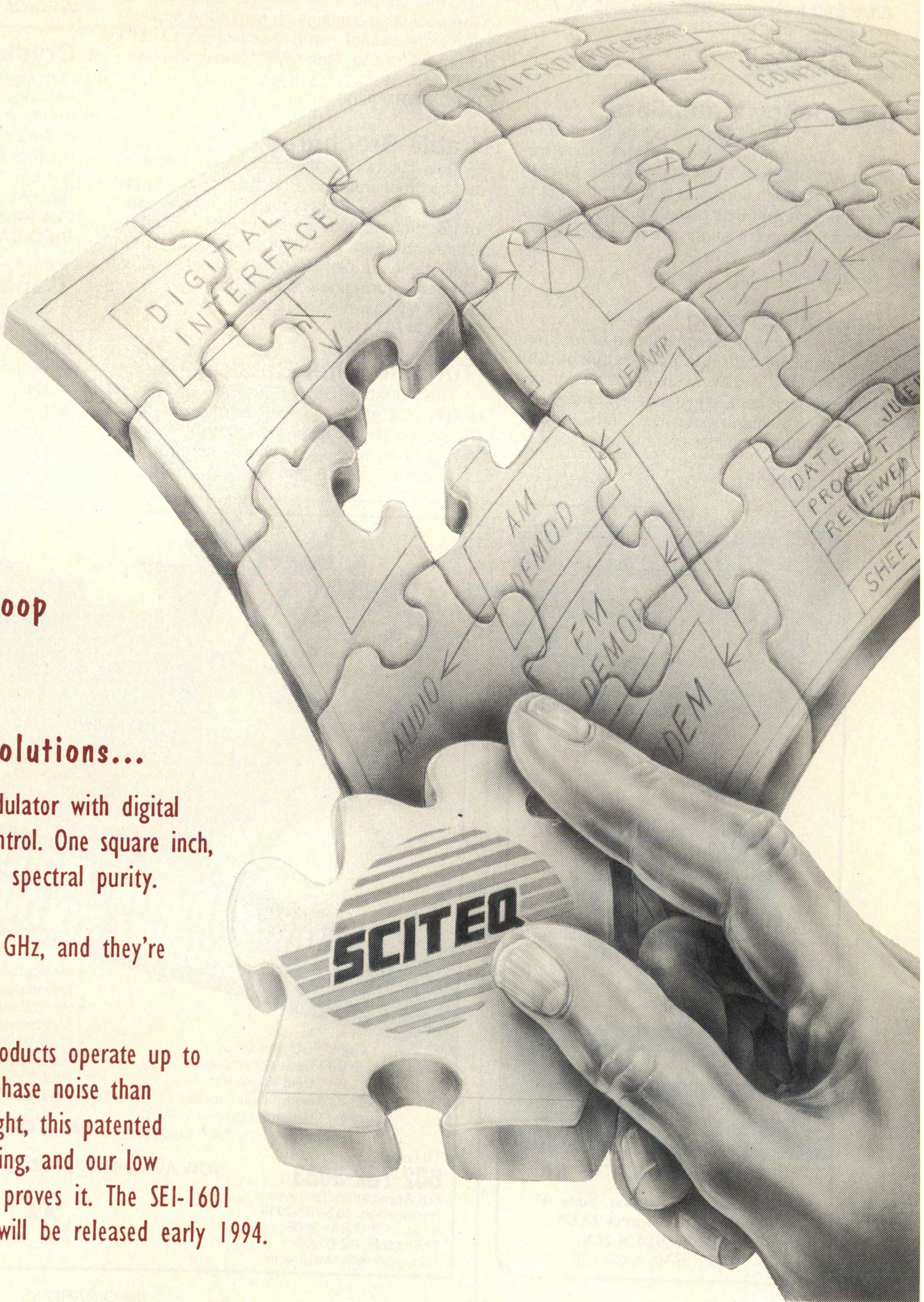
Capacitor Catalog

Cornell Dubilier Electronics (CDE) introduces its 44-page Mica Capacitors catalog. CDE's silvered mica capacitors are available in radial-leaded dipped, radial-leaded boxed, surface mounted, and transmitting cast or potted cases. Capacitances range from 0.5 pF to 1 µF; voltage range is 50 to 2500 VDC.

Cornell Dubilier Electronics
INFO/CARD #196

The Frequency Synthesizer is *Always* a Problem...

Sciteq is *Often* the Solution...



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Air-Cooled High Power Amplifiers, Class AB Linear:

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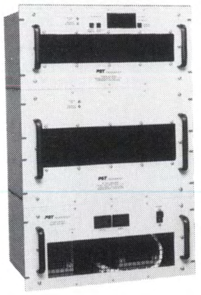


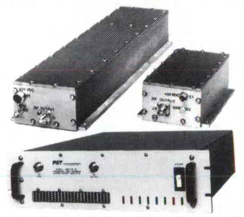
- High Reliability
- Wide Bandwidth
- Thermal Protection
- Load VSWR Protection
- Graceful Degradation
- Low MTTR
- Built-In Test Diagnostics
- IEEE 488 Bus (optional)
- RS232 (optional)

For complete specifications, request Product Data 2010

Series AM/AR

Class A Linear Amplifier Modules & Rack-Mount Instruments:

Frequency 1 to 4000 MHz; Power 2 to 100 Watts



- Multi-Octave Bandwidths
- Modular Maintainability
- Reduced Spares
- High Gain Ratings
- Low Bandpass Ripple
- Automatic Circuit Protection
- Built-In Thermal Protection
- IEEE 488 Bus (optional)

For complete specifications, request Product Data 1010

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Up-Down Frequency Converters:

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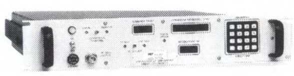
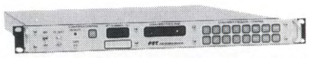


Series 3500:
1KHz Step Size



Series T-3500 "Thin-Line":
Single Rack Unit Package
125KHz Step Size

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- No Spectral Inversion
- 10 Pre-Programmed Frequencies

For complete specifications, request Product Data 4010 & 4020

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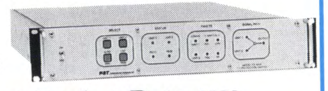
1:1 and M:N Converter Protection Switches:

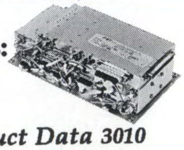
Direct, Continuous Communication Between On-Line & Standby Converters; Automatic, Manual Mode



For complete specifications, request Product Data 4030

Series 3 and 5

Sub-System Microwave Synthesizers:



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INFO/CARD 55

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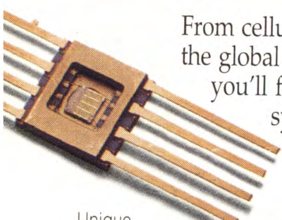

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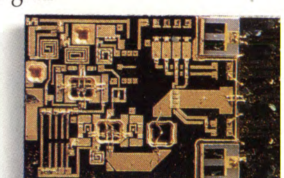

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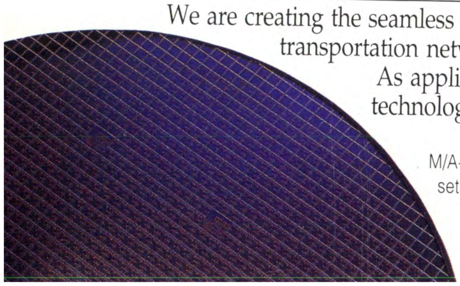

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