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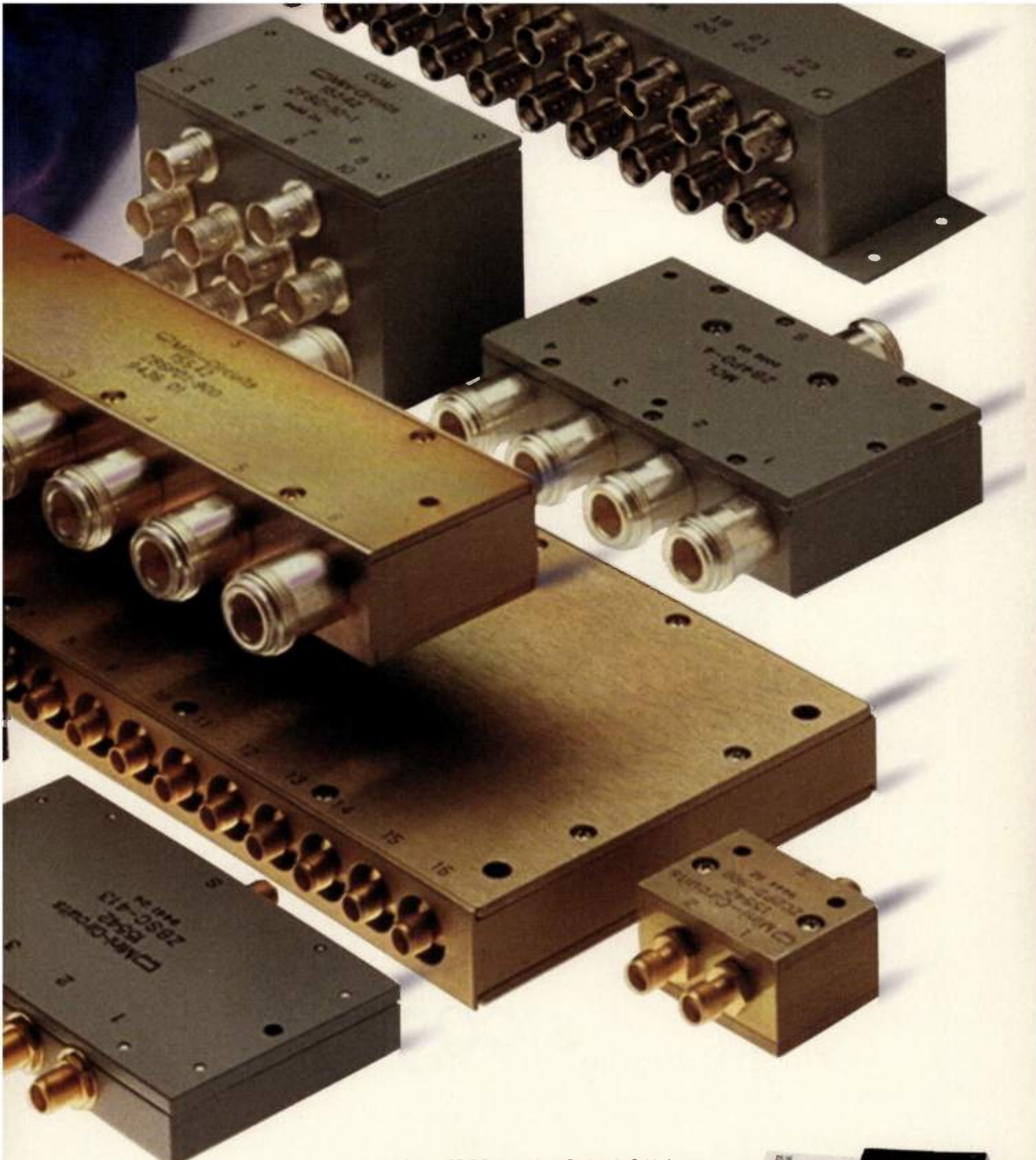

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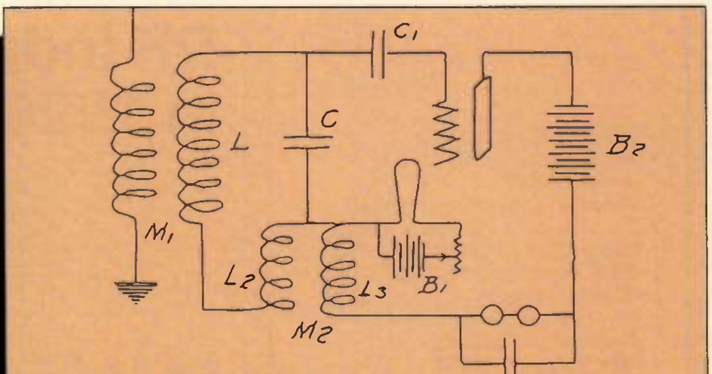

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RF editorial

Predictions for the RF Industry in 1996

By Gary A. Breed
Editor

I'm really not psychic, but there are a few things that will be big news in 1996 in our industry. A few other things aren't so big, but they promise to be interesting, as well.

The start of PCS — The first systems in the Personal Communications Services will begin operations. The interesting part will be watching customer reaction. Will PCS be perceived to be just like cellular service, or will its differences be the attraction? I think the early PCS systems will be the former, another service option for prospective cellular subscribers. Interest in new features will come along later. I also predict a slow start for PCS in any market that has unused cellular capacity. In places where analog cellular service is saturated, we will see strong head-to-head competition between PCS and digital cellular for new business.

Interactive cable — One of the cable giants (TCI or Time-Warner) will begin a large-scale project in a major market. Various trials have been completed, and more trials are underway to refine new technology. The companies' motivation is to get a two-way infrastructure in place to take advantage of the new Communications Act. Then, when will we see fiber-to-the home implemented?

The New Communications Act — After a lot of initial hoopla, the Act did not pass in 1995. Some supporters are having second thoughts — Will a more open market really create competition, or will the wealthiest players end up as de facto monopolies after a short period of increased competition? What are the appropriate levels of regulation and free-market factors? I don't think we'll get a new Communications Act until June or July — in time for its supporters to tout it in their fall campaigns, but late enough so no one can see the effects of its changes on the communications business.

The "one-chip" solution — My friends in the RF integrated circuit business tell me that only a few applications should be put on one chip. Cordless phones, wireless LANs and other devices with low transmitter power are the best candidates. I think the three-chip solution for PCS or cellular will become the standard in 1996, and will be the most common arrangement for two to three more years. I also expect multichip module (MCM) technology to be implemented that will create products much like single-chip radios. To make one-chip radios feasible, advances are needed in filter technology, allowing them to be integrated either on a chip, or in a hybrid or MCM. As long as filters are external to the chips, the advantage of a one-chip design is minimal.

The RF job market — Rapid business growth will continue for at least ten more years, so prospects look rosy. The thorn in the rose is industry's demand for very specific job skills and design objectives. The opportunities to design technically elegant, high-performance products are limited to instrumentation and perhaps some base station designs. The new chellenge for RF engineering is delivering designs quickly that can be easily manufactured for low consumer cost.

1996 promises to be another year of growth for RF products. I predict a great year for *RF Design* readers and their companies!

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Frank Kalmus
Technical Director

Frank Kalmus is the founder of Kalmus (formerly Kalmus Engineering Inc.), a division of Thermo Voltek Corporation. As Kalmus' principal design engineer, he has designed over 200 RF amplifiers used for EMC test, medical/MRI, general laboratory, and communications applications.

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Looking for Current Spectrum Usage Information

Editor,

Many of us would love to see an article describing the current use of the RF spectrum. Where are the garage door opener frequencies now? Where are the taxis and truckers? At one time the military took all of 225-339.95 MHz; now FCC Part 15 says unlicensed folk can use a large part of that. Spread spectrum is assigned to 902-928 MHz, but as an ISM band, it can have some very powerful CW users. Where are automotive keys transmitting? This information would be helpful to entrepreneurs.

Pete Lefferson, P.E.
St. Petersburg, FL

Who Knows About HARP?

Editor,

I would like to see something written about the monster RF activity taking place, HARP. This giant system, from what I've heard, is located some 240 miles from Anchrage, Alaska. It is reported to start out with a billion watts, with plans to go up to maybe 100 billion watts. Instead of the signal spreading out, it is beamed up to a point where it affects the ions. I have heard quotes from a book *Angels Don't Play This HARP* that say such a transmitter could change the weather, and make animals and people act strangely, and do other things.

I am sure many people would like to hear what is really happening with this system!

And, keep up the good work at *RF Design*.

G. Eldon Wright
Norton, KS

Over the past 25 or more years, there have been several "ionospheric heating" experiments to analyze behavior of the ionosphere, and to investigate man-made influences on ionospheric radio propagation. The HARP system in Alaska is the only active U.S. project I am aware of. I will contact engineers who have had involvement with this project, and perhaps other readers will offer information, as well. — Editor

Another Note on Experience vs. Credentials

Editor,

As a further comment to the letter by reader Klaus Spies — [*October 1995* Letters, *noting the difficulty companies have in evaluating applicants with excellent track records but few on-paper credentials — ed.*] — I may add that most paper qualifications have a "best used before...date," and, hence, need constant updates not only to replenish one's neuron cells with active ones, but also for creating new ones!

J. Ayer, P. Eng,
Victoria, BC

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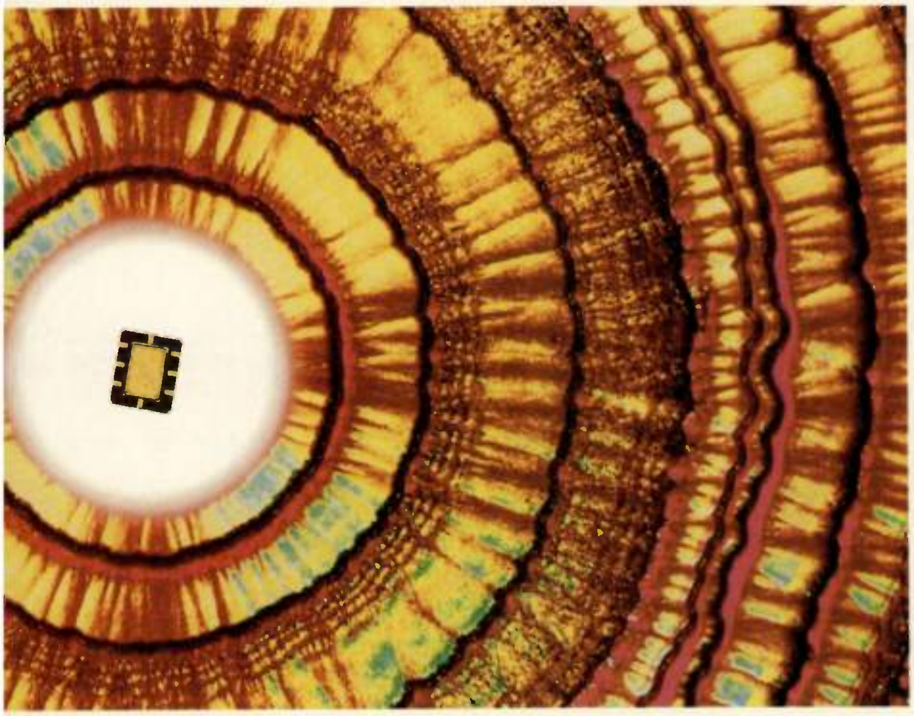

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16-18 **RF Design Seminar Series**
Dallas, TX
Information: Argus Trade Shows, 6151 Powers Ferry Road, N.W., Atlanta, GA 30339. Tel.: (800) 828-0420 or (770) 618-0499. Fax: (770) 618-0441.

February

12-16 **Wireless Symposium & Exhibition**
Santa Clara, CA
Information: Mary Begley, Penton Publishing, 611 Route 46 West, Hasbrouck Heights, NJ 07604. Tel: (201) 393-6289.

March

3-7 **IPC Printed Circuits Expo '96**
San Jose, CA
Information: Institute for Interconnecting and Packaging Electronic Circuits (IPC), 2215 Sanders Road, Northbrook, IL 60062-6135. Tel: (708) 509-9700. Fax: (708) 509-9798.

18-22 **12th Annual Review of Progress in Applied Computational Electromagnetics**
Monterey, CA
Information: Richard W. Adler, ECE Dept./Code ECAB, Naval Postgraduate School, 833 Dyer Road, Room 437, Monterey, CA 93943. Tel: (408) 649-1111. Fax: (408) 649-0300. Email: rwa@mcimail.com

April

15-18 **NAB '96**
Las Vegas, NV
Information: NAB Conventions, 1771 N St., NW, Washington, DC 20036. Tel: (800) 622-3976. Fax: (202) 429-4180.

24-26 **International Wireless Communications Expo (including the RF Design Seminar Series)**
Las Vegas, NV
Information: Argus Trade Shows, 6151 Powers Ferry Road, N.W., Atlanta, GA 30339. Tel.: (800) 828-0420 or (770) 618-0499. Fax: (770) 618-0441.

June

5-7 **Frequency Control Symposium**
Honolulu, HI
Information: Michael Mirarchi, Synergistic Management, 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 280-2024. Fax: (908) 681-9314.

11-12 **Radio Data Solutions Europe 1996**
Amsterdam, The Netherlands
Information: Radio Data Solutions Europe, The Old Vicarage, Haley Hill, Halifax, HX3 6DR, U.K. Tel: +44 1422 380397. Fax: +44 1422 355604.

5-7 **MTT-S International Microwave Symposium**
San Francisco, CA
Information: Derry Hornbuckle, Hewlet-Packard; Tel: (707) 577-3658; Fax: (707) 577-2036, or Jerry Fiedziusko, Space Systems/Loral Corp. Tel: (415) 852-6868. Fax: (415) 852-5068.

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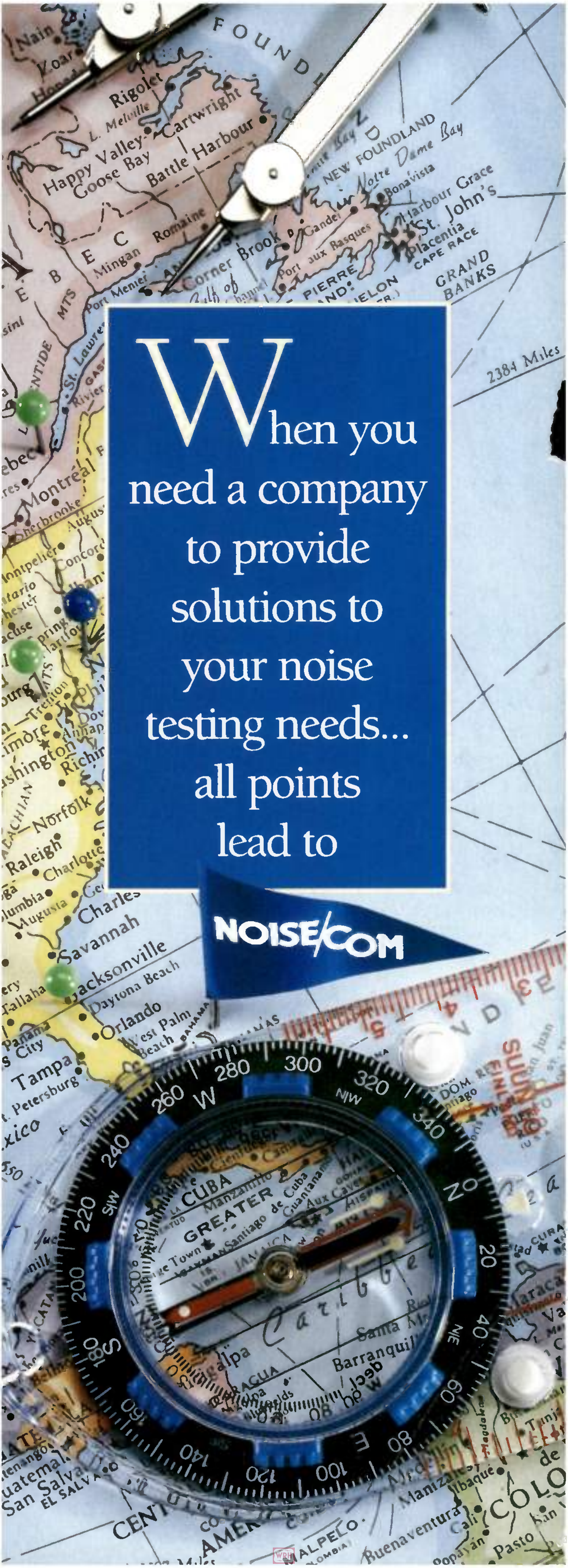
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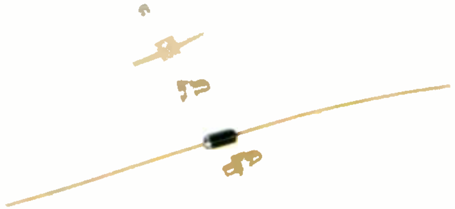

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NC 305	10 MHz - 11 GHz
NC 401	100 MHz – 18 GHz
NC 406	18 GHz – 110 GHz

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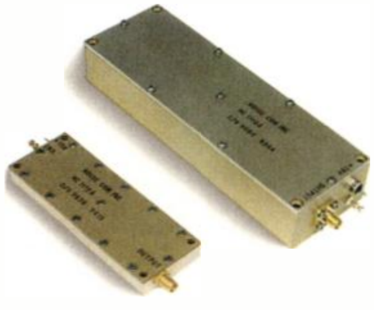
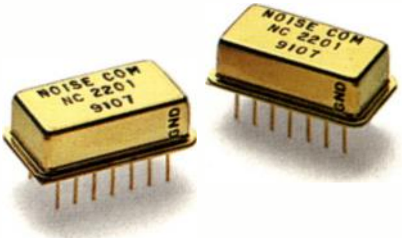
TYPICAL STANDARD MODELS		
MODEL	**FREQUENCY RANGE**	**OUTPUT ENR**
NC 501/15	0.2 MHz – 500 MHz	31 dB
NC 502/15	0.2 MHz – 1000 MHz	31 dB
NC 503/15	0.2 MHz – 2000 MHz	31 dB
NC 506/15	0.2 MHz – 5 GHz	31 dB
NC 511/15	0.2 MHz – 500 MHz	51 dB
NC 513/15	0.2 MHz – 2 GHz	51 dB

Broadband Amplified Modules



The NC 1000 series amplified noise modules produce white Gaussian noise from -14 dBm to +13 dBm at frequencies up to 6 GHz. They are designed for coaxial test systems, and are available with several bias voltages and connector options.

TYPICAL STANDARD MODELS		
MODEL	**FREQUENCY RANGE**	**OUTPUT**
NC 1101A	10 Hz – 20 kHz	+13 dBm
NC 1107A	100 Hz– 100 MHz	+13 dBm
NC 1112B	20 MHz – 2 GHz	0 dBm
NC 1126A	2 GHz – 6 GHz	-14 dBm



The NC 2000 series amplified noise modules are an excellent choice when a high level noise output is desired and the noise source is to be mounted on a circuit board. 24 pin packages are standard; 14 pins are also available.

TYPICAL STANDARD MODELS		
MODEL	**FREQUENCY RANGE**	**OUTPUT**
NC 2101	100 Hz – 20 kHz	0.15 Vrms
NC 2105	500 Hz – 10 MHz	0.15 Vrms
NC 2201	1 MHz – 100 MHz	+5 dBm
NC 2601	1 MHz – 2 GHz	-5 dBm

NOISE/COM

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Broadband Precision, Calibrated Coaxial

Noise Com's NC 346 series is designed for precision noise figure measurement applications. These products are available with coaxial or waveguide outputs. For OEM applications, the NC 3200 series provides high performance in a small ruggedized package.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT ENR
NC 346A	0.01 GHz – 18 GHz	6 dB
NC 346B	0.01 GHz – 18 GHz	15 dB
NC 346C	0.01 GHz – 26.5 GHz	15 dB
NC 346D	0.01 GHz – 18 GHz	25 dB
NC 346Ka	0.1 GHz – 40 GHz	15 dB

Broadband Calibrated Millimeter-wave

The NC 5000 series noise sources feature outstanding stability and convenience in waveguide bands up to 110 GHz.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	WAVEGUIDE
NC 5142	18 GHz – 26.5 GHz	WR-42
NC 5128	26 GHz – 40 GHz	WR-28
NC 5122	33 GHz – 50 GHz	WR-22
NC 5115	50 GHz – 75 GHz	WR-15
NC 5110	75 GHz – 110 GHz	WR-10

Broadband Noise Generators

The NC 6000 and NC 8000 series noise-generating instruments are designed for applications on the test bench or incorporated with other equipment to provide a wide variety of functions. Each instrument contains a precision noise source, amplification, and step attenuators to provide repeatable symmetrical white Gaussian noise with variable output power.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT POWER
NC 6107	100 Hz – 100 MHz	+13 dBm
NC 6108	100 Hz - 500 MHz	+10 dBm
NC 6110	100 Hz – 1500 MHz	+10 dBm
NC 6124	2 GHz – 4 GHz	-10 dBm
NC 8107	250 kHz – 100 MHz	+30 dBm
NC 8111	5 MHz - 1 GHz	+30 dBm

The new UFX-7000 series noise-generating instruments are extremely easy to use, combining dedicated keys for control of operations and programming, with a large 4 x 20-character LCD display. Control of output power, filter settings, and attenuator step size for both the noise and the signal (for units with internal combiners) is performed from the front panel or by remotely using the IEEE-488 interface.

TYPICAL STANDARD MODELS		
MODEL	FREQUENCY RANGE	OUTPUT POWER
UFX-7107	100 Hz–100 MHz	+13 dBm
UFX-7108	100 Hz – 500 MHz	+10 dBm
UFX-7110	100 Hz – 1500 MHz	+10 dBm
UFX-7218	2 GHz – 18 GHz	-20 dBm
UFX-7909	1 MHz – 300 MHz	+30 dBm

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Practical High Frequency Filter Design
January 16, 1996, Dallas, TX
April 24, 1996, Las Vegas, NV
Oscillator Design Principles
January 17, 1996, Dallas, TX
April 25, 1996, 1996, Las Vegas, NV
Digital Modulation and Spread Spectrum for Wireless Communications
January 16, 1996, Dallas, TX
April 25, 1996,
RF Power Transistors and Amplifiers
January 17, 1996, Dallas, TX
April 25, 1996, Las Vegas, NV
Wireless Communications for Non-Engineers
January 16-17, 1996, Dallas, TX
April 23, 1996, Las Vegas, NV
Information: RF Design Seminar Series, Argus Trade Shows, 6151 Powers Ferry Rd., N.W. Atlanta, GA 30339. Tel: (800) 828-0420.

51st Engineering + Management Program
March 24-29, 1996, Los Angeles, CA
Information: UCLA Extension, 10995 Le Conte Ave., Suite 542, Los Angeles, CA 90024. Tel: (310) 825-1047; Fax: (310) 206-2815. E-mail: mhenness@unex.ucla.edu

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Analog & RF Printed Circuit Design
February 14-16, 1996, Milwaukee, WI
Information: Center for Continuing Engineering Education, University of Wisconsin-Milwaukee, 161 W. Wisconsin Ave., Suite 6000, Milwaukee, WI 53203. Tel: (414) 227-3159.

RF & Microwave Measurements & Applications
January 29-February 1, 1996, Monterey, CA
Electronic System Design for Testability
January 22-23, 1996, Monterey, CA
Information: University Consortium for Continuing Education, 16161 Ventura Blvd., M/S 752, Encino, CA 91436. Tel: (818) 995-6335. Fax: (818) 995-2932.

Radar Signal Processing: Theory, Technology and Application
January 31-February 2, 1996, Atlanta, GA
Coherent Radar Performance Estimation
February 5-8, 1996, Atlanta, GA
Antenna Engineering
February 13-16, 1996, Atlanta, GA
Principles of Pulse Doppler Radar: High, Medium and Low PRF
February 13-15, 1996, Atlanta, GA
Radar Cross Section Reduction
March 26-29, 1996, Atlanta, GA
Information: Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel: (404) 894-2547.

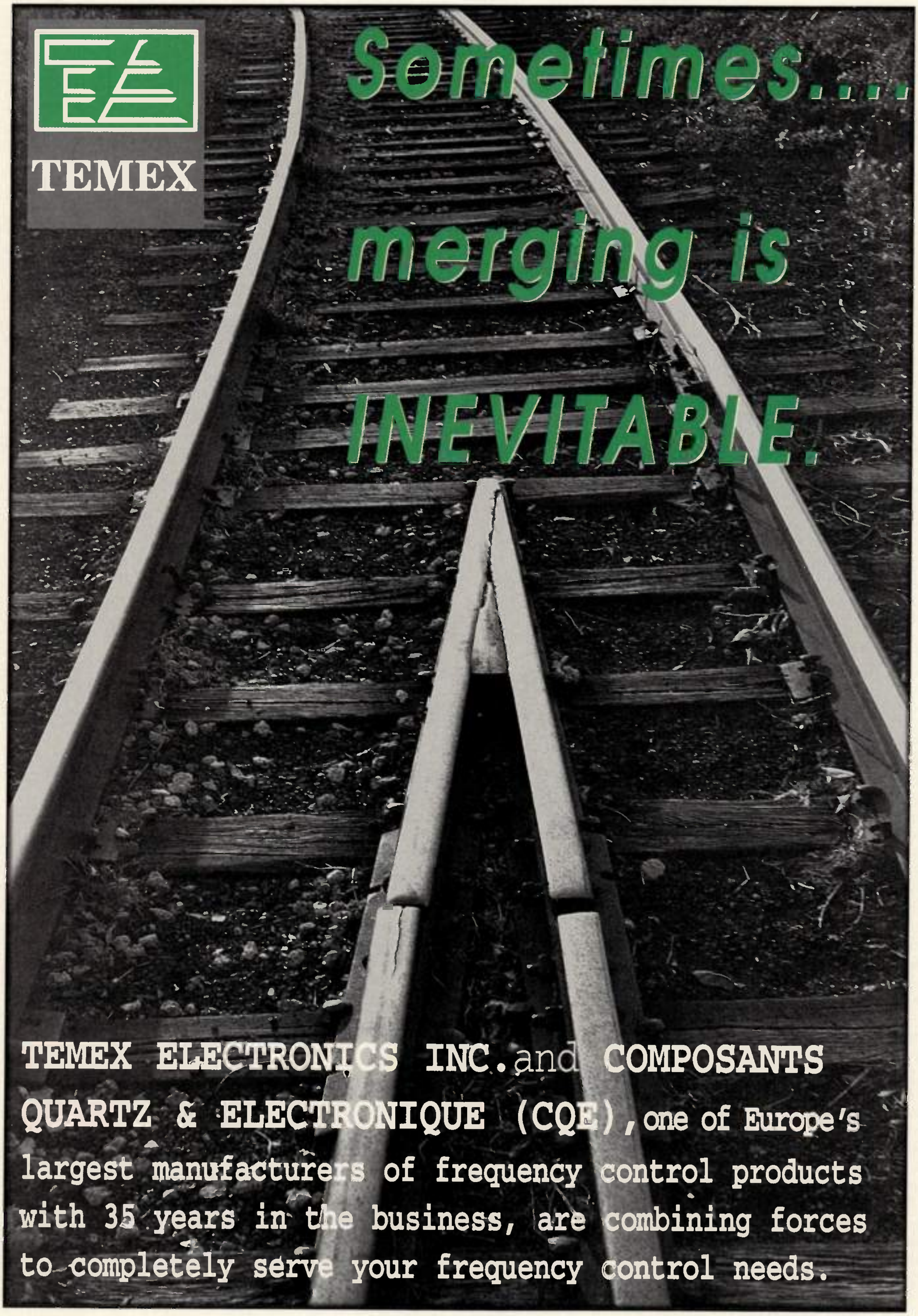
Wireless Digital Communications
February 26-March 1, 1996, Tempe, AZ
Information: College of Engineering and Applied Sciences, Center for Professional Development, Arizona State University, P.O. Box 877506, Tempe, AZ 85287. Tel: (602) 965-1740. Fax: (602) 965-8653.

Antennas: Principles, Design and Measurements
March 13-16, 1996, San Diego, CA
May 14-17, 1996, St. Cloud (Orlando), FL
Information: Kelly Brown, Northeast Consortium for Engineering Education, 1101 Massachusetts Ave., St. Cloud, FL 34769. Tel: (407) 892-6146. Fax: (407) 892-0406.

Digital Cellular and PCS Communications: The Radio Interface
January 22-26, 1996, Washington, DC
Mobile Cellular Telecommunications Systems
January 22-24, 1996, San Diego, CA
Cellular and Wireless Telephony
January 29-February 2, 1996, Washington, DC
Communications Satellite Systems: The Earth Station
February 5-8, 1996, Washington, DC
Electromagnetic Interference and Compatibility (EMI/EMC): A Practical Approach to Testing and Problem Solving
February 12-16, 1996, Washington, DC
Analog and Digital Cellular Networks: CDMA versus TDMA
March 6-8, 1996, Washington, DC
Lightning Protection
March 14-15, 1996, Washington, DC
Global Positioning System: Principles and Practice
May 20-23, 1996, Washington, DC
Electromagnetic Interference and Control in Modern Communications Systems
May 20-24, 1996, Washington, DC
Hazardous RF Electromagnetic Radiation: Evaluation, Control, Effects, and Standards
June 12-14, 1996, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd Street, N.W., Washington, DC 20052. Tel: (202) 994–6106 or (800) 424–9773. Fax: (202) 872–0645.

Modern Digital Modulation Techniques
February 5-9, 1996, France
Mobile Cellular and PCS Telecommunications Systems
February 7-9, 1996, France
Wireless Digital Communications
February 5-9, 1996, France
Applied RF Techniques: Linear Circuits
February 5-9, 1996, France
Mobile and Wireless Personal Communications Networks
February 12-16, 1996, France
Spread Spectrum/CDMA
February 12-16, 1996, France
Bandwidth-Efficient Coded Modulation: Theory and Application
February 12-14, 1996, France
Information: Tine Persson, CEI-Europe, P.O. Box 910, S-612 25 Finspong, Sweden. Tel: +46-122-175 70; Fax: +46-122-143 47.

RF and Wireless Made Simple
January 22-23, 1996, Los Altos, CA
Applied RF Techniques II
January 29-February 2, 1996, Los Altos, CA
Wireless RF System Design
January 22-26, 1996, Los Altos, CA
Information: Besser Associates, 4600 El Camino Real, Suite 210, Los Altos, CA, 94022. Tel.: (415) 949-3300. Fax: (415) 949-4400. E-mail: BesserCourse@delphi.com. World Wide Web: http://www.bessercourse.com.



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NIST Develops New Noise Design Rules

NIST scientists have developed a theoretical framework for describing the excess low-frequency phase modulation and amplitude modulation noise generation processes in RF and microwave amplifiers. They have produced a set of design rules that work exceptionally well for bipolar transistor circuits. A key element in this advance is the development of state-of-the-art systems for measuring PM and AM noise close to the carrier, allowing verification testing of the concepts. NIST has demonstrated high gain amplifiers where the excess phase noise is less than the thermal noise, even within a few Hz of the carrier. NIST is prepared to disseminate this new technology through intensive training sessions with limited attendance. Persons interested in the training sessions should contact Fred Walls, NIST, Boulder, CO 80303-3328; tel: (303) 497-3207; e-mail: walls@bldrdoc.gov

FCC Proposal: Drop Filings for UHF TV Noise Figure

The Federal Communications Commission, responding to a petition from the Electronic Industries Association, has proposed elimination of the requirement for filing UHF noise figure measurement of television receivers. The EIA's petition pointed out that compliance with the requirement was very high, and that the intent of the requirement to provide quality UHF television had been met. The FCC will continue to require TVs to have a 14 dB noise figure at UHF, and proposes that compliance be verified by routine follow-on testing by the manufacturers and through random sampling by the FCC.

Report Predicts 15% Annual Wireless IC Growth

Allied Business Intelligence predicts that the global market for integrated circuits in wireless communications will grow at a 15 percent annual rate through the year 2000. Growth in PCS, DBS, VSAT, GPS and WLAN will fuel IC market growth, and although some of these markets will grow by 20 to 30 percent, progressively lower costs for ICs will keep growth for those components at a lower rate. For more information on this report, contact Andy Fuertes at Allied Business Intelligence, tel: (516) 624-3113.

Microwave Power Symposium Call for Papers

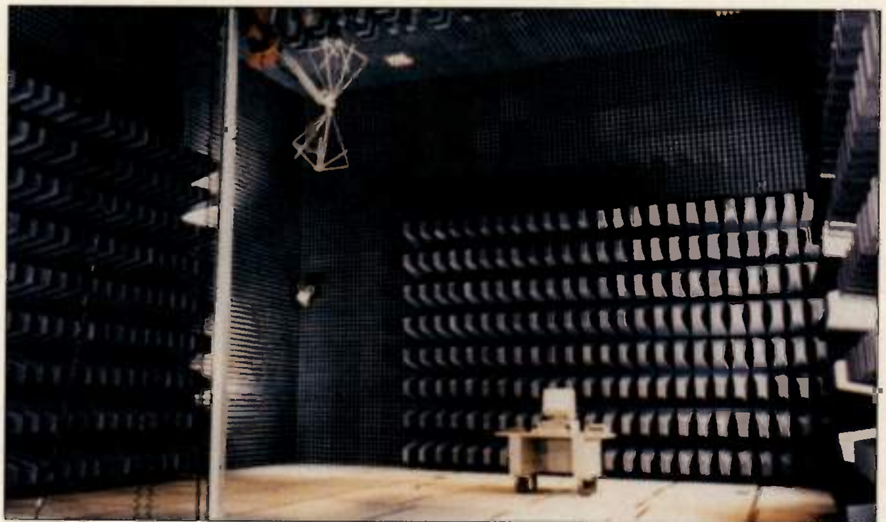
The 31st Annual Microwave Symposium sponsored by the International Microwave Power Institute (IMPI) is accepting proposals for papers. The conference will be held July 29-31, 1996 at the Park Plaza Hotel in Boston, Mass. Titles and abstracts are due by February 2. Send a 60-word abstract with complete author information to ISMI Section Symposium Papers, IMPI, 10210 Leatherleaf Court, Manassas, VA 22111. Tel: (703) 257-1415; Fax: (703) 257-0213.

Update on Bird Model 43 Quest Contest

The Quest 43 contest by Bird Electronic Corp. has yielded 327 entries by early December. Deadline for entries is January 31, 1996. Rewards will be offered for the 10 lowest serial number Model 43 wattmeters, with the owner of the lowest number unit receiving a gold-plated model 43 and $1000 gift certificate. The other nine winners will receive $250-$500 gift certificates and new model 43 meters. Winners will be announced at the International Wireless Communications Expo in Las Vegas, April 24-26, 1996. Contact Bird at (216) 248-1200 for an entry form.

NASA Develops A Wristband Person-Locator System

NASA Tech Briefs reports that a computerized system based on wristband RF passive transponders is being developed for real-time tracking of individuals in custodial institutions like prisons and mental hospitals. The system includes a monitoring system with a central transceiver that communicates with remote transceiver nodes that, in turn, interrogate the wristband transponders. Each wristband would be encoded with a unique digital code for identification of the wearer. The tracking system can warn of an individual's presence near the perimeter or other



Rantec installs new anechoic chamber at Compaq Computer — The first ten-meter chamber using Rantec FerroSorb^{TM} anechoic absorber has been successfully installed at Compaq Computer in Houston, Texas. The chamber will be used to perform product compliance testing for ANSI C63.4 emission and IEC 1000-4-3 immunity standards. FerroSorb is a multi-layered hybrid material offering broadband 30 MHz to 40 GHz performance in a 41-centimeter height. It replaces an eight-foot absorber that requires a much larger facility. Chamber size can be reduced by approximately 25 percent with no loss in interior test volume. Site attenuation can be correlated to theoretical Normalized Site Attenuation (NSA) values within ±4 dB for both horizontal and vertical polarizations.

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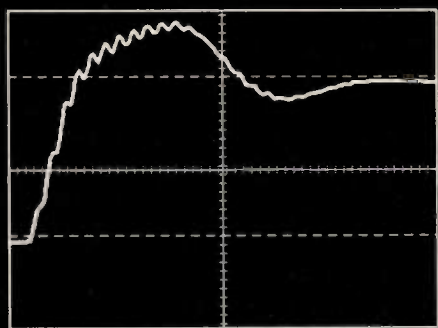

LMX2332A switching from 865MHz to 915MHz (the GSM band) in less than 350μs

	LMX1501A	LMX1511	LMX2314/15	LMX2320	LMX2325	LMX2330A	LMX2331A	LMX2332A	LMX2335	LMX2336	LMX2337
RF Input-Main PLL	1.1GHz	1.1GHz	1.2GHz	2.0GHz	2.5GHz	2.5GHz	2.0GHz	1.2GHz	1.1GHz	2.0GHz	550MHz
RF Input-Aux PLL						510MHz	510MHz	510MHz	1.1GHz	1.1GHz	550MHz
I_{cc} (typ) @3V	6mA	6mA	6mA	10mA	11mA	15mA	14mA	8mA	9mA	11mA	9mA
Powerdown (typ)	N/A	N/A	30μA	30μA	30μA	1μA	1μA	1μA	1μA	1μA	1μA

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INFO/CARD 18

off-limits locations, as well as provide real-time location information on a computer-generated map of the facility. For security, the wristbands contain an embedded wire loop that interrupts operation if the wristband is removed. In this case, the unit's absence would be detected and an alarm sounded. This work was done by Caltech for NASA's Jet Propulsion Laboratory. Inquiries for licensing this technology for commercial use should be addressed to William Callaghan, JPL-301-350, 4800 Oak Grove Drive, Pasadena, CA 91109.

Microwave Metrology Now Online at NIST

If you have an Internet e-mail address, you can subscribe to an electronic mailing list that provides an open forum on microwave metrology. To subscribe, send e-mail to: majordomo@central.bldrdoc.gov. The body of the message should read: subscribe mwave-meas [*your e-mail address*]. T cancel the service, substitute the com mand "unsubscribe" in place of sub scribe. To get help or to find out mor about the listserver, send e-mail to th same address with the word "help" i the message. Once subscribed, an message sent to: mwave-meas@cen tral.bldrdoc.gov will be forwarded t all members on the list. There is n charge for this service, which is pro vided by NIST's Electromagneti Fields Division. RI

Business Briefs

Motorola and Richardson disengage — Motorola's Semiconductor Products Sector and Richardson Electronics have agreed to disengage their supplier/distribution relationship due to changes in marketing strategies. A Motorola spokesman stressed that the disengagement was not a reflection of Richardson's performance, but an effort to reduce the number of distributors to a handful of global companies.

Illinois Superconductor start first filter field tests — Illinois Superconductor announces that a major cellular operator has installed six of the company's SpectrumMaster™ cellular B-band filters for field testing. Internal testing indicated that customer performance objectives had been met or exceeded, and the company is confident that field testing will confirm that the filters reduce interference and increase cell siting flexibility and call capacity.

Telecom Analysis Systems to merge with Bowthorpe plc — Telecom Analysis Systems (TAS), a manufacturer of test systems for telephone, digital network and wireless communications has agreed to a merger with Bowthorpe plc., which is involved in a number of focused niche markets in the electronics and electrical industries. TAS management will remain with the company following the merger.

Samsung licenses DSP Group technology — Samsung Semiconductor and DSP Group, Inc. announced that Samsung has licensed a key part of DSP Group's digital signal processor technology, the PineDSPCore™ engine. The technology is a 16-bit general-purpose DSP engine designed to be embedded with an ASIC for speech/audio processing, telecommunications, digital cellular and embedded control applications.

Texas Instruments acquires Savi Technology — Save Technology, a wireless communications company making RFID tags, tag readers and RF links, has been acquired by Texas Instruments. Prior to acquisition, the company has been teamed with TI on the Defense Department's Total Asset Visibility (TAV) logistics management initiative.

Amphenol, Richardson add Europe to distribution agreement — Electronic component distributor Richard Electronics and connector manufacturer Amphenol Corp. have added Europe to their distribution agreement. Richardson is now authorized to distribute Amphenol's line of RF connectors in Spain, Portugal, the U.K., Benelux and Scandinavia (excluding Norway). Previously, France was the only market besides the U.S. covered by a distribution agreement.

KeyTek moves to larger facilities — The new address for KeyTek is One Lowell Research Center, Lowell, MA 01852-4345. Their telephone number is (508) 275-0800; Fax: (508) 275-0850.

Nemal opens Brasil office — Nemal Electronics International announces the opening of its new sales and customer support facility in Sao Paulo, Brasil. The address of the facility is: 433 Alameda Campinas, 9 Andar, Sao Paulo, Brasil, CEP 01404-901, telefax: (55-11) 284-1769. Nemal is a manufacturer of cable assemblies and a distributor of cable and connectors.

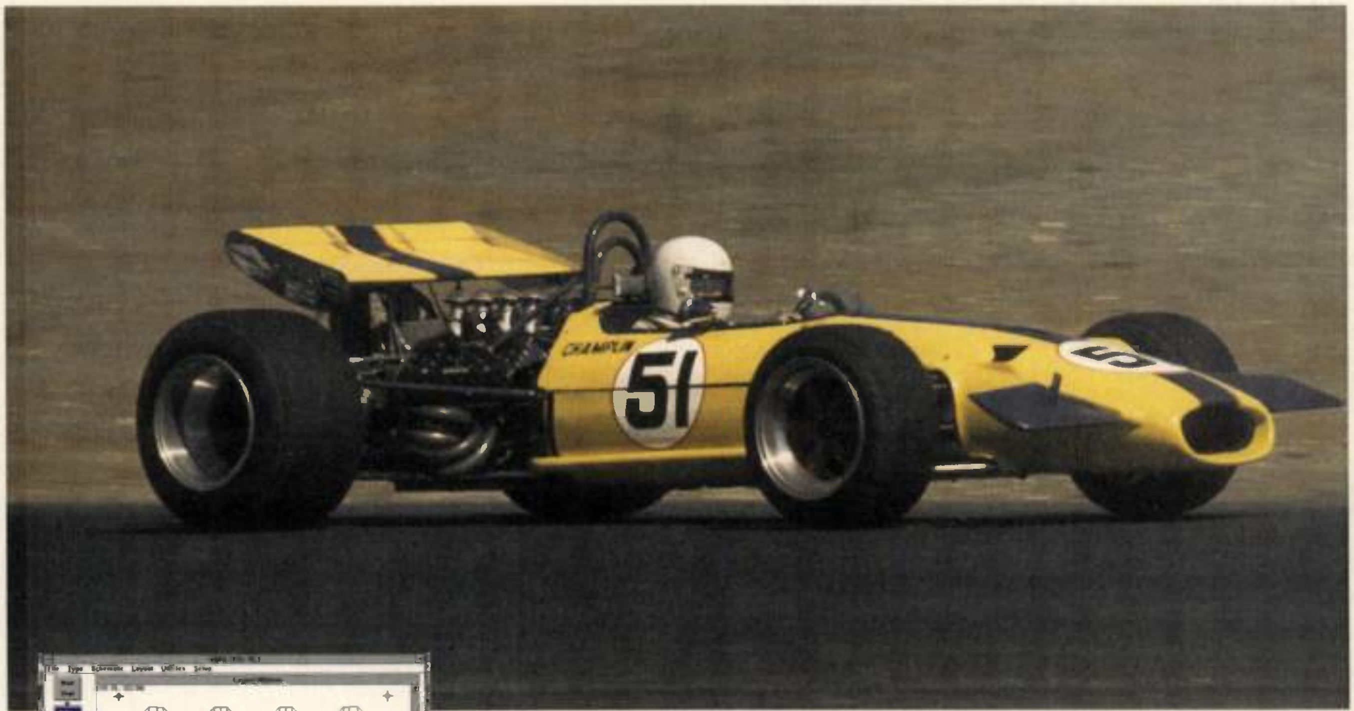
Coaxial Components moves sales office — The new National Sales Office for Coaxial Components Corp is at: 1707-32 Veterans Memorial Highway, Islandia, NY 11722-1531. Tel: (516) 234-4447 or (800) 583-9954.

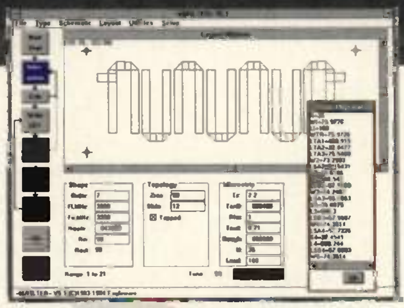
Amitron expands with a second U.S. plant — Amitron Corp., a printed circuit board manufacturer serving the telecommunications, computer, automotive, industrial and automotive markets, has increased manufacturing capabilities by 400 percent through the purchase of a 206,000 square foot plant in Elk Grove Village, Ill.

Anritsu Wiltron announces new service center — A full service repair and calibration facility has been opened by Anritsu Wiltron in Richardson, Texas. The new facility will service the company's complete line of Digital Mobile Radio and Wireless test and measurement instruments.

LNR gets contract for satellite terminals — LNR Communications, Inc. has received a multimillion dollar contract to supply flyaway INTELSAT Qualified Satellite Terminals (IQST). Under contract to GLS Associates, Inc., LNR will manufacture, integrate, test, deliver and support Ku-band terminals. The terminals comprise a 2.4 meter antenna subsystem, RF/IF transmit/receive electronics, digital modems and control systems.

TURBO-CHARGE TOUCHSTONE





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L-C Clapp
T-line and L-C VCO
VCO with xformer
Cavity bipolar and hybrid
Dielectric resonator
Terminal SAW bipolar
Port SAW hybrid
Port SAW MOSFET
Pierce and Colpitts crystal
Driscoll crystal
Butler overtone
Overtone with multiplier

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T-line pseudo lowpass
T-line stepped-Z
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INFO/CARD 17

No More Talk About Talk – Broadband PCS Hitting the Airwaves

By Andy Kellett
Technical Editor

Personal communications systems have been talked about for the better part of a decade, but finally the hand-waving and gee-whiz prognostication are giving way to actual hardware. Companies have invested millions of dollars in licenses and hardware; each company betting on their own combination of technology and services to woo customers. Whether companies make back their PCS investments in a few years or more than decade will depend not on dazzling savvy customers with new technology, but on luring customers to sign up who have not even heard of PCS.

PCS is ...

There is nothing like the investment of over seven billion dollars to bring a technology into sharp focus. The eighteen companies who together spent over seven billion dollars just to get broadband A and B block PCS licenses (and the three pioneers preference license holders who sank money into developing various PCS technologies) are quickly deciding what combinations of features and costs they will present to their potential subscribers as PCS.

The only PCS system up and providing service to paying subscribers at this writing is American Personal Communications' Sprint Spectrum service in the greater Washington DC-Baltimore area. This service uses GSM1900 technology and provides airtime for 12.5 cents to $1.00 per minute.

Standard features of the Sprint Spectrum service include caller ID, built-in answering-machine function and numeric paging. Other features such as voice mail, text messaging, call waiting, call forwarding, and call barring are also offered. "The Sprint Spectrum model is being looked at closely as a testbed for what works and what doesn't," says Rob Hoggarth, Director of Regulatory Relations at the Personal Communications Industry Association (PCIA).

One interesting thing to note is that Sprint Spectrum is not hyping PCS as "technology of the future, today" – service providers realize that consumers don't care how advanced a technology is as long as it works. Instead, PCS is being sold as the wireless phone service that does everything your wired phone can do.

Everything that has been seen as a flaw in cellular service is addressed: voice quality is touted as crisp, clear, static-free and without crosstalk. Privacy and security are selling points, and no long term service contracts are required. Even the concern of paying for unwanted incoming calls is addressed by Sprint Spectrum. According to promotional material you get the first minute of incoming calls free to determine if the person calling you is worth the additional cost of a wireless call.

Despite American Personal Communications' marketing, PCS does offer some new capabilities. Many PCS systems may be adapted to act as a wireless PBX when within a building (particularly DECT-based technology). When two PCS phones are within the coverage of each others handset, the two phones can operate as walkie talkies, completely bypassing the phone network (and tolls). The FCC doesn't specify what kind of service is to be provided on the PCS frequencies, so it is possible some non-telecommunications-network applications may appear, particularly in the unlicensed band from 1910 to 1930 MHz.

Status of PCS Build-Out

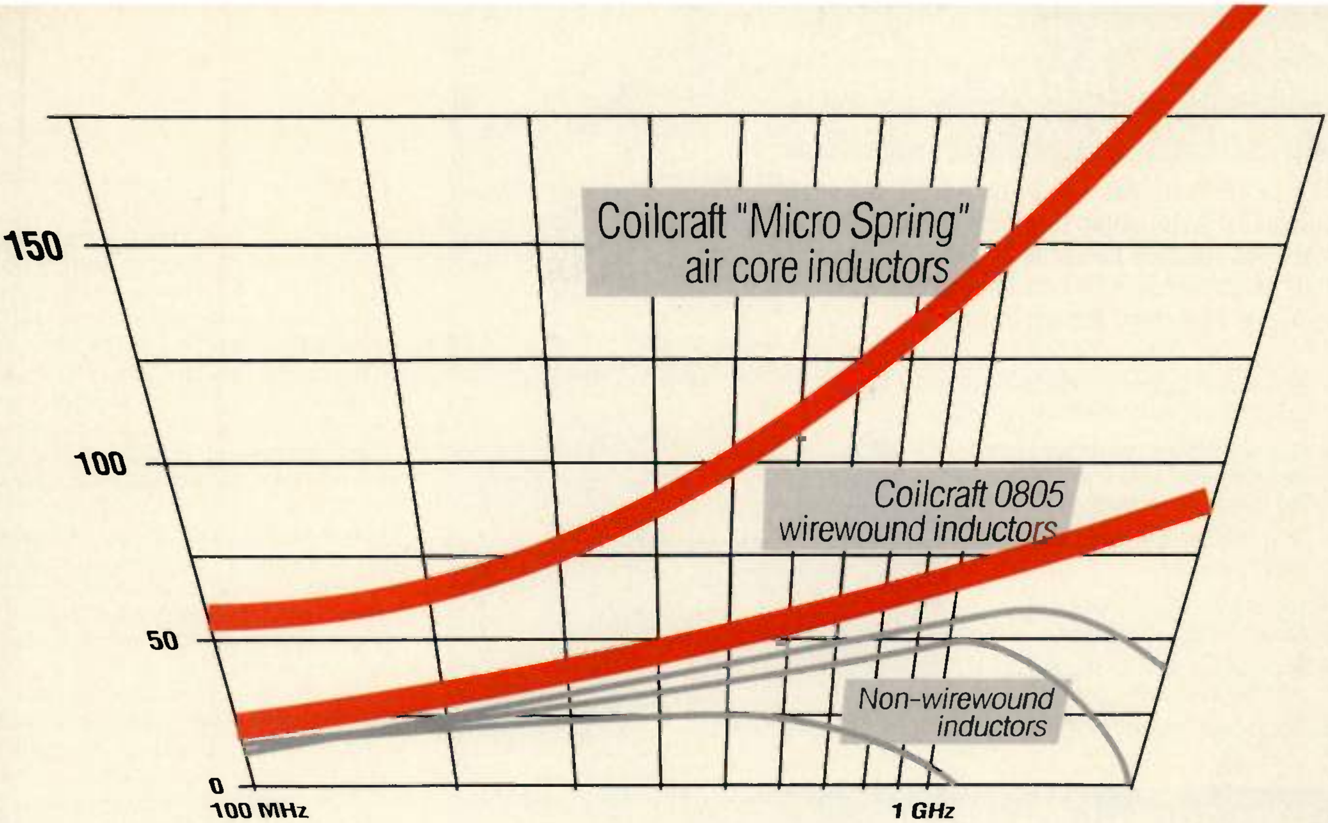
So how far along are PCS providers in their installation (or build-out) of PCS infrastructure? Most companies are hesitant to disclose the extent of their progress. However, most PCS providers have announced what technology they have selected, and pilot systems are in place in several large cities.

A particularly visible pilot system is being installed by Colorado-based Omnipoint Corp. in Manhattan's Wall

Name	IS-661	IS-95-based	PACS/UPACS	IS-136-based	GSM1900	DCTU
Technology	CDMA /TDMA	CDMA	TDMA	TDMA	TDMA	TDMA
Proposer	Omnipoint	Motorola, Qualcomm	Motorola, Hughes/PCSI, Panasonic, NEC, Hitachi	AT&T, Ericsson/GE Mobile	Nortel, Ericsson/GE Mobile, Siemens, Alcatel	Ericsson
Comments	Nortel is licensed to produce base-stations, JRC will produce handsets	Based on U.S. CDMA digital cellular standard (IS-95)	UPACS operates in both licensed and unlicensed PCS bands	Based on U.S. TDMA digital cellular standard (IS-136)	Based on European digital cellular standard (GSM)	Based on European cordless phone standard (DECT)

sources: OKI Semiconductor, Omnipoint Corp.

Table 1. PCS technologies. Most PCS service providers have selected one of these technologies for their systems, while generally each base station manufacturer has adopted several.



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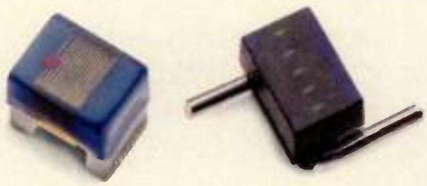

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Street district. Omnipoint begai installing their Nortel-built TDMA CDMA base stations in early Decem ber. Eventually the system will b expanded to cover all of the New Yorl MTA (major trading area), for whicl Omnipoint daughter company, Omni point Communications, has a license.

Industry wide, build-out has beei slowed down by the effort it takes t acquire locations for cell sites. Site that were made available to cellula operators for a little money and witl no second thought are now availabl to PCS providers for many times th money and only after much delibera tion, if not litigation.

Though not a threat to the buildin of PCS systems, providers still need t exactly resolve how they will split th costs of microwave relocation, say PCIA's Hoggarth. Those companie that currently hold licenses to use th PCS frequencies must be compensate for the costs of moving to a differen frequency.

Component Build-In

Many of the same companies tha produced hardware for cellular net works in the eighties are now ready t begin building equipment for PCS PCS basestations and handsets will b produced in very large numbers. "[PCS base station's] range is relatively mini mal, that's the purpose of it, so th quantity of required units is astronom ical." says Ken Wadors, Vice Presiden of Sales at electronics distributor Pen stock. Another hallmark of PCS hard ware will be its low power consump tion. Manufacturers are producing 3 V devices with PCS in mind. Oki Semi conductor has produced a whole PCS chipset – baseband and RF – tha operates from 3V.

Greg Peloquin, Business Unit Man ager for the RF Components Divisior at Richardson Electronics notes tha as a distributor, Richardson doesn' expect to supply a lot of components t high volume production lines, "but w will continue to play our role, which i to keep companies aware of what's lat est and greatest on the market and t help them design-in components.' Peloquin notes that PCS hardware will continue to be developed even after PCS systems come on line.

Though PCS is about to take defi nite shape, it will definitely be mal leable enough to allow new services and new technology – if it pleases sub scribers. *RF*

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INFO/CARD 22

Isolator for DECT Open Loop Modulation

By Rishi Mohindra
Philips Semiconductor

In defining a suitable architecture for a DECT double superheterodyne radio, there are various options for the choice of the synthesizer frequency plan and VCO modulation. To be cost effective, the same RF frequency synthesizer must be used both for the transmitter as well as the receiver. For the modulation, direct VCO modulation is the cheapest for GFSK, since accurate phase is not required. In GSM where accurate phase is essential, QAM modulation is used using I and Q baseband digitally-generated signals. Choice of the architecture is a trade-off between cost, design time and availability of compatible chip sets. This article describes in detail problems associated with open loop modulation in the DECT radio and gives comprehensive solutions for a low cost design.

Low cost DECT handset radios use single RF synthesizers that make use of blind-slot operation, which means that only every other TDMA time slot is used for communication. The unused time slots between active slots are used to program and settle the frequency synthesizer to the desired channel frequency. The first IF for most DECT radios is 110 MHz (220 MHz is less common). 110 MHz is also the difference between the TX (transmitter) and RX (receiver) frequency for the synthesizer. Most of the blind-slot synthesizer and modulation plans for DECT superhet radios fall under two categories, open loop and closed loop modulation.

Open Loop Modulation

In this scheme, the VCO output is directly at the channel frequency during transmission. Just before the beginning of a transmit RF burst, the synthesizer is switched off, and the VCO is made free running. The baseband gaussian-filtered data directly modulates the VCO. During receive, the same synthesizer is used with a receive VCO. It is not important if the VCO works in open loop (synthesizer off) during the receive data burst. A single band-switched or a wide-band VCO can work both for the TX as well as the RX.

A severe requirement for open-loop modulation is isolation of the VCO from output load impedance changes that lead to frequency jumps. A poorly designed synthesizer chip can also produce frequency jumps when the loop is opened. Wide-band VCOs are more susceptible to frequency jumps. Also, open loop modulation is sensitive to humidity that makes the loop filter capacitor leak, thereby producing a VCO frequency droop that must be contained within the required specification (formerly 13 kHz/millisecond).

Closed Loop Modulation

In this scheme, the RF VCO always works in closed loop. This modulation has two possibilities. The first is where the VCO is directly modulated at the transmit frequency in closed loop. The loop bandwidth must be very small during the transmit burst, in order not to be affected by the modulation. Up to 64 continuous ones or continuous zeros are allowed in the data, and they produce the worst-case frequency error due to the synthesizer correcting the modulation frequency deviation. The loop bandwidth can be larger in the settling time prior to the data burst, so that it can quickly settle to the programmed frequency. However, switching loop bandwidth can cause frequency jumps if not done properly.

Another possibility is in using a second VCO at IF for modulation, and mixing it with an RF VCO signal up to the TX frequency. Both the RF and IF VCOs and synthesizers are common to the transmitter as well as the receiver. The RF VCO is in closed loop and performs the blind-slot frequency hopping. The IF VCO could work with open loop modulation. Upconversion implies filtering of unwanted frequency components, and usage of image rejection mixers, and tends to be more expensive.

Philips DECT Superhet Radio

In a Philips DECT Radio (Figure 1), open loop modulation is used, with the TX VCO module running directly at

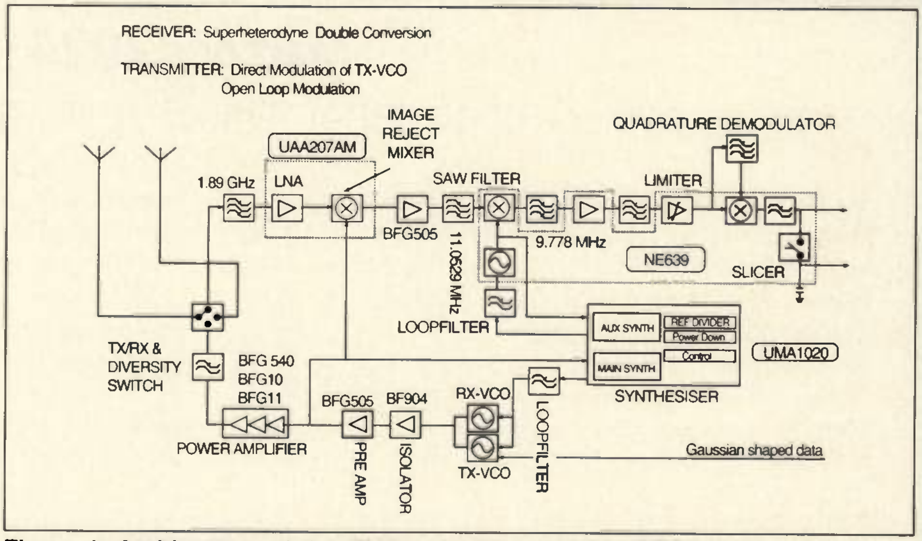

Figure 1. Architecture of the Philips double superhet DECT radio, which uses open-loop modulation during transmission.

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ZOS-50	25-50	-107	0.012	-22
ZOS-75	37.5-75	-110	0.016	-26
ZOS-100	50-100	-111	0.026	-29
ZOS-150	75-150	-107	0.017	-26
ZOS-200	100-200	-106	0.015	-25
ZOS-300	150-280	-103	0.017	-27
ZOS-400	200-380	-100	0.021	-24
ZOS-535	300-525	-96	0.018	-27
ZOS-765	485-765	-96	0.033	-27
ZOS-1025	685-1025	-92	0.051	-25

Notes: Tuning voltage 1 to 16V required to cover freq. range. Power output +9dBm Typ. (Main). Power 12V DC, +130mA (MAX.). Operating temperature range: -55°C to +85°C.

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the transmit frequency. The RX VCO module is physically connected in parallel to the TX VCO, at the RF output and tuning input (in the next release the two VCOs would be combined into a single VCO based on a 900 MHz discrete transistor band-switched oscillator, tuned to the 1.9 GHz second harmonic at the output). The PA (power amplifier) uses 3 bipolar surface mount RF transistors as shown in Figure 2a, and is designed for 26 dBm output power into 50 Ω at 3.6 volts supply. It needs an input drive of 3 dBm, has 40 percent effeciency, and works in a pulsed mode of up to 12.5 percent duty cycle.

Between the VCOs and the PA there is an active isolator and a buffer using two stages of discrete SMD RF transistors. Its main purpose is to reduce the load variation for the VCO, when the PA is power ramping at the beginning of a transmit slot, during which time the synthesizer is already powered down and the loop is open.

Secondly, the isolator serves as a buffer with 10 dB gain, to sufficiently drive the PA. The RF power at the isolator output is tapped to provide the LO signal (–6 dBm) to the UAA2077AM image-reject front-end, and the feedback signal (–10 dBm) to the UMA1020 synthesizer main prescaler input. The advantage of passive power tapping at the isolator output is to avoid the use of (expensive) RF PIN diode switches.

The modulation input signal to the TX VCO comes directly from the Philips ABC baseband chip that has a built-in digital Gaussian data shaper. The amplitude of the modulating signal is programmable in the ABC chip, in order to compensate for the spread in modulation sensitivity of the VCO.

The UMA1020 chip has a main synthesizer that is used both for the TX as well as the RX. The auxiliary synthesizer of the UMA1020 is used for the fixed 2nd LO of 120.36995 MHz. The first IF of the receiver is at 110.592 MHz, and the second IF is about 9.777951 MHz. A reference frequency of 13.824 MHz is used for the synthesizer, and it comes from the ABC chip's crystal oscillator.

The receiver uses a UMA2077AM silicon bipolar chip that has a low-noise amplifier and a 30 dB image-reject mixer, with a total RF to IF (110 MHz) power gain of 20 dB. A cheap but high insertion-loss SAW filter is used for channel selectivity. The Philips SA639 chip is used for the FM-IF strip, and also provides RSSI, demodulation, data filtering, and data slicing. A low-drop 3 volt regulator is used to power the VCOs, isolator and the SA639 chip.

TX VCO Frequency Jump

The DECT radio is a pulsed system, with various parts of the tranceiver being switched at different times. Many of these switchings cause disturbances that lead to changes in the TX VCO free-running frequency in the open loop modulation period. A typical timing diagram for the Philips DECT transmitter is shown in Figure 3.

Each slot has 480 bits, numbered from b0 to b479. There are 24 slots in a frame, of which the first 12 are used to transmit data from the base station to the portable, and the remaining 12

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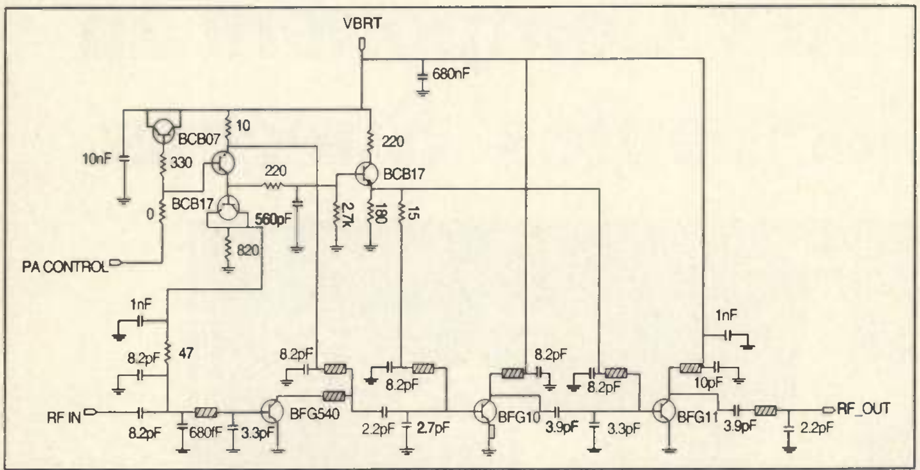

Figure 2a. Simplified schematic of the Philips discrete power amplifier (PA) for DECT. The PA produces 26 dBm into 50 Ω from a 3.6 V supply.

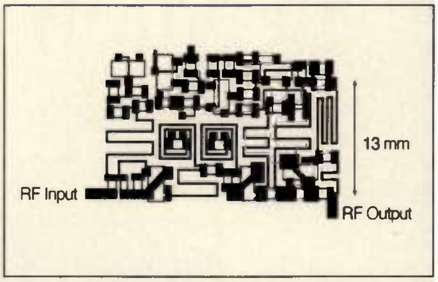

Figure 2b. Layout of the power amplifier of Figure 2a.

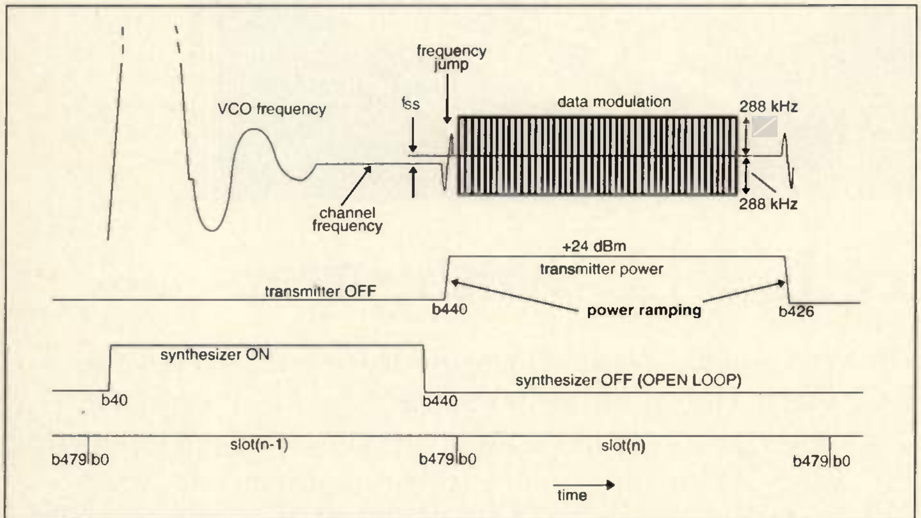

Figure 3. Diagrams showing typical frequency settling occuring in the slot before data transmission takes place, typical power levels before and during data transmission, and receiver synthesizer on/off switching.

are used for the reverse transmission.

The data is transmitted at a bit rate of 1152 kbits/sec. Of the 480 bits in a slot, only the first 424 are used, while the remain ning 56 are used as guard space. There are 10 frequency channels available, with a channel spacing of 1.728 MHz. The lowest frequency is 1881.792 MHz. In blind slot operation, only alternate slots are used for transmission. The slot preceding a transmission slot is used to program and settle the frequency synthesizer to the required channel.

The UMA1020's main synthesizer is turned ON at b40 in slot (n–1). It takes about 250 µs for the frequency to settle within 5 kHz of the steady-state programmed channel frequency. At b440, the synthesizer is turned OFF, and the loop is now open (VCO free-running). The charge-pump outputs do not load the loop filter when the synthesizer is OFF.

At b464, the transmitter power is ramped up to 24 dBm (maximum), starting from an OFF state. During power-ramping, the transmitted VCO free-running frequency changes due to various disturbances. The peak frequency change during the ramping, may be much larger than the steady- state frequency jump error, f_{ss}, at the end of the power ramping. The total frequency error, f_{error} is given by:

$$f_{error} = f_{ss} + f_{synth} + f_{noise} \leq 50 \text{ kHz} \qquad (1)$$

A maximum of 50 kHz (26 ppm at 1.89 GHz) is allowed for f_{error} relative to the ideal programmed channel frequency during transmission. f_{synth} is the error in the synthesizer frequency due to the reference frequency error, and is less than 20 kHz if the reference crystal frequency error is better than about 10 ppm. f_{noise} is the jump in VCO frequency when the loop is opened (which occurs before power ramping) and is due to synthesizer frequency jitter from noise and reference break-through. Noise and reference break-through produce voltage fluctuations in the loop filter, and when the loop is opened the instantaneous tuning voltage is frozen. This is seen as a small step in the VCO mean frequency.

Part of the step can also be due to digital break-through due to switching OFF the synthesizer. f_{noise} is less than a few kilohertz for blind-slot operation, using the UMA1020 synthesizer, with a VCO phase noise of –95 dBc/Hz at 25 kHz offset. The largest source of frequency error is f89, and is due to VCO disturbances as described next.

VCO Disturbance Sources

There are three sources of disturbance that change the VCO frequency during power ramping of the PA. The frequency change occurs as a transient plus a stady state error. While the transient error is not important as far as the 50 kHz frequency error is concerned, it is of significance in the test for spurious emission into adjacent channels due to transmitter transients.

Supply Voltage Change (Pushing Figure for VCO)

The PA draws about 280 mA when it is turned ON, and this drops the supply voltage by up to a few hundred milivolts in case of the portable, when the batteries are nearly discharged. The TX VCO must be protected from

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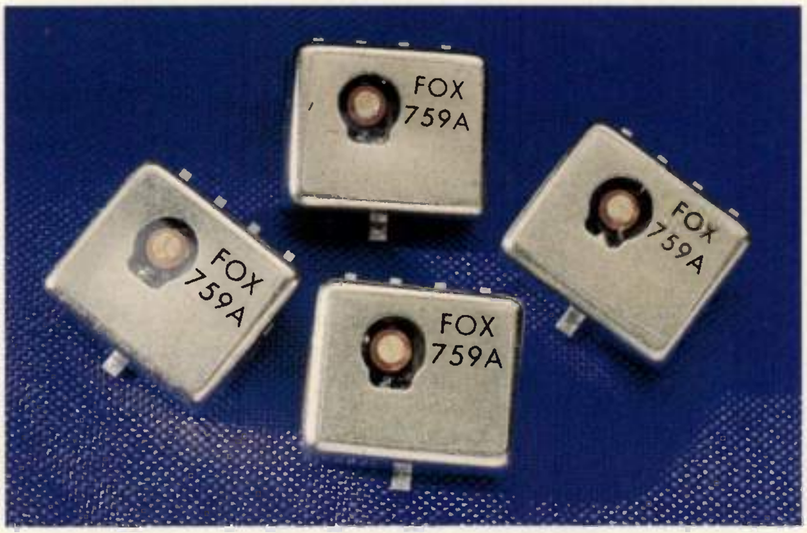

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this supply voltage change by the use of a fast-settling low-drop voltage regulator. For example, the pushing figure for the VCO module used in the Philips radio, is 800 kHz for a 150 mV supply change. With a line regulation of 0.12 %/V, the frequency jump due to a 200 mV supply change is 1.28 kHz, which is small. A larger transient frequency change occurs due to overshoot/undershoot in the regulator output during settling, but this decays quickly.

Radiation Leakage

Any leakage of transmitted RF power to the VCO (which is oscillating at the same frequency) has the effect of changing the internal feedback conditions of the oscillator, thereby changing the oscillation frequency. If the TX VCO is not shielded properly, it can lead to disastrous frequency changes during power ramping and will vary significantly with external reflection and absorbtion.

To protect the TX VCO from transmitted signal pickup, it must be placed in a shielded enclosure. Also, supply, tuning and modulation input lines should not inject RF signal into the VCO. Philips microwave NPO 8.2 or 6.8 pF capacitors may be placed in parallel with these lines to ground to attenuate the radiation pickup. These capacitors are series resonant near 1.89 GHz and act as a short circuit. Input line sections (microstripline or stripline) may be used as series inductance to form a π-section filter along with the shunt capacitors. Special antenna structures can also reduce this leakage effect. Using an oscillator at half the frequency helps greatly in reducing the frequency change.

In the Philips DECT radio, adequate shielding has kept the frequency jump contribution from radiation leakage, to less than 10 kHz. In an earlier design, tiny chip EMI filters (40 dB notch at 1.9 GHz) were used in series with the supply, tuning and modulation input lines that entered the shielded enclosure of the VCO and isolator. No observable frequency jump (due to radiation leakage) was present even when the antenna was placed next to the VCO shield.

Load Pulling

This is the most severe cause of frequency change during power ramping, and occurs due to change in input impedance of the PA when it goes from OFF to ON state and vice versa. The frequency change, both transient and steady state, can be reduced by placing an isolator between the TX VCO and the PA. The steady state frequency error can further be reduced by choosing an appropriate electrical length between the VCO and the PA input. A detailed description is given in the following section.

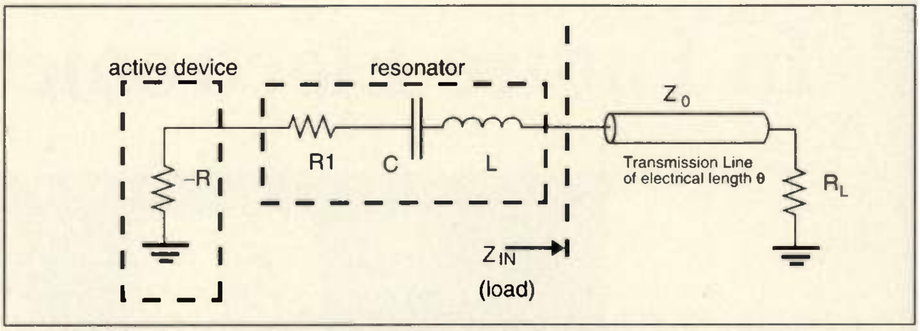

Figure 4. Negative resistance model for oscillator load-pull calculations.

VCO Load Pulling

An important VCO specification is it's load pulling. It indicates the peak-peak frequency ($\Delta f_{p\text{-}p}$) change when its load moves on a constant VSWR circle through all phases. Figure 4 shows an oscillator model using the negative resistance concept. The negative resistance (–R) increases with oscillation amplitude, and at steady state, R = R1 + real(Z_{IN}). When Z_{IN} is real (no reactance), the frequency of oscillation is:

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \tag{2}$$

At this frequency, the resonator reactances cancel, i.e.:

$$jX_L + jX_C = 0, \text{ or } 2\pi fL = (2\pi fC)^{-1} \tag{3}$$

When the load reactance (imaginary part of Z_{IN}) changes from 0 to ΔX, the resonant frequency changes by Δf from the original series resonance that is given by equation 3.

$$|\Delta X| = \left|\frac{2X}{f_0}\Delta f\right| \tag{4}$$

where X is the reactance X_L or X_C at resonance frequency f_0.

To see the effect of load pull, the impedance Z_{IN} is varied on a constant VSWR circle with reflection coefficient ρ, as shown in Figure 5. The peak-peak change in reactance of Z_{IN} can be shown to be:

$$\Delta X = X_A - X_B = Z_0\left(\frac{4\rho}{1-\rho^2}\right) \tag{5}$$
$$= Z_0\left(S - S^{-1}\right)$$

where S is the VSWR.

X_A and X_B are the constant reactance curves that are tangent to the constant VSWR circle at A and B respectively.

A change in load reactances shifts the oscillation frequency in such a way that the total loop reactance is always zero. Therefore, equating (4) and (5), the total change in frequency can be seen as:

$$\Delta f_{p-p} = \frac{f_0}{2X_L} Z_0\left(\frac{4\rho}{1-\rho^2}\right) \tag{6}$$
$$= \frac{f_0}{2X_L} Z_0\left(S - S^{-1}\right)$$

By putting $X_L/Z_0 = Q_{ext}$ (external Q), equation 6 becomes:

$$\Delta f_{p-p} = \frac{f_0}{2Q_{ext}}\left(S - S^{-1}\right) \tag{7}$$

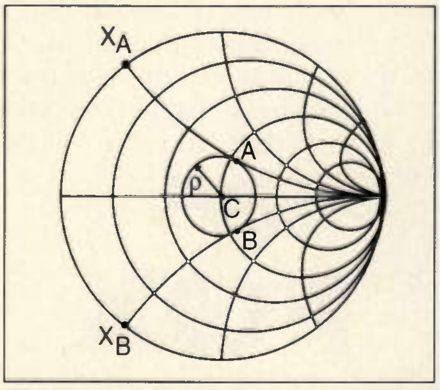

Figure 5. Peak-peak reactance change looking into transmission line terminated in R_L. Reactance change is on a circle of constant VSWR.

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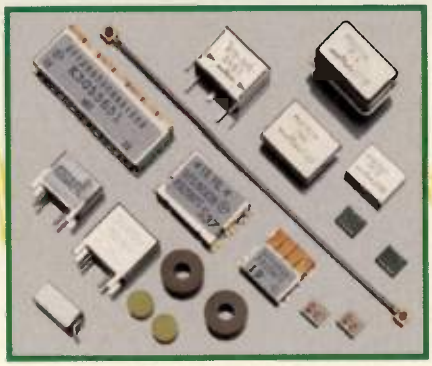

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$$\Delta f_{p-p} = \frac{f_0}{2Q_{ext}}\left(\frac{4\rho}{1-\rho^2}\right) \quad (8)$$

or

$$\Delta f_{p-p} = \frac{f_0}{2Q_{ext}} 2\rho \quad (9)$$

when $\rho << 1$.

Normally, the oscillator load is buffered from the resonant circuit as in Figure 6. S_{ij} represents the buffer's S parameters. If the buffer's load Z_L moves on a constant VSWR circle of radius (reflection coefficient) $\rho_L = 1$, it can be shown that the input impedance of the buffer also moves on a circle with a radius ρ_{IN} given by:

$$\rho_{in} = \frac{|S_{12}S_{21}|}{1-|S_{22}|^2}\rho_L \quad (10)$$

Using a matching circuit at the output does not change the radius of the input impedance circle, in terms of the original S-parameters of the buffer. Matching at the input increases the radius. In terms of the buffer's original S parameters, the new radius becomes:

$$\rho_{in} = \frac{|S_{12}S_{21}|}{\left(1-|S_{11}|^2\right)\left(1-|S_{22}|^2\right)}\rho_L \quad (11)$$

In the case when S_{11} and S_{22} are zero, and the load moves on a constant VSWR circle of arbitary radius (reflection coefficient magnitude) ρ_L, the input impedance of the buffer also moves on a constant VSWR circle with radius ρ_{IN} given by

$$\rho_{in} = |S_{12}S_{21}|\rho_L \quad (12)$$

Equation 12 is also valid when the buffer is conjugately matched at the input and output, and transformed to 50 Ω. In this case, S_{12} and S_{21} terms in equation 12 are resultant values after the matching and transforming to 50 Ω.

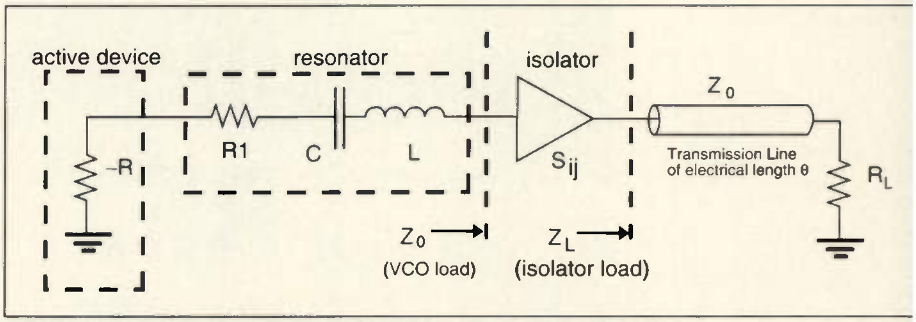

Figure 6. Model for buffered oscillator load-pull calculations.

Equations 10 to 12 show that the oscillator can be isolated from load variations by making $|S_{12}S_{21}|$ very small. Considering the case where the buffer's S_{11} and S_{22} are zero, and substituting ρ_{IN} from 12 for ρ in equation 9:

$$\Delta f_{p-p} = \frac{f_0}{Q_{ext}} 2|S_{12}S_{21}|\rho_L = \frac{f_0}{Q'_{ext}} 2\rho_L \quad (13)$$

assuming $|S_{12}S_{21}|\rho_L << 1$, where

$$Q'_{ext} = \frac{Q_{ext}}{|S_{12}S_{21}|} \quad (14)$$

Equations 13 and 14 show that buffering of the oscillator can be interpreted as an increase in effective Q as seen by the output load, in terms of load pulling.

A Definition for Isolation

From 12 it can be seen that an isolator reduces the load reflection coefficient variation by the factor $|S_{12}S_{21}|^{-1}$ and this factor may be termed isolation. In dB it is $-20\log|S_{12}S_{21}|$. In the special case where an external attenuator is put in series to make the total gain 0 dB, the isolation becomes S_{21}/S_{12}. This is because the forward and reverse gain of the attenuator is $|S_{21}|^{-1}$.

Example Isolation Calculation

For example, the 1.89 GHz TX VCO module used in the Philips radio, has a peak-peak frequency change of 2.5 MHz, when the load varies along a VSWR = 2 circle (corresponding to ρ = 0.33). Using equation 8, Q_{ext} = 560. This includes an increase in Q due to any internal buffering.

The PA has an OFF state input VSWR of about 40 corresponding to a reflection coefficient of $\rho_{OFF} \approx 1$. In the ON state, it is has an input impedance of 50 Ω ($\rho_{ON} \approx 0$). An active isolator is used between the PA and the VCO. The worst case reactance change as seen by the VCO would correspond to moving from A or B to C as in Figure 5. This is half the reactance change for the case when going from A to B. Therefore the worst case frequency from A to C is also half that of when going from A to B.

To keep the frequency change within 5 kHz when using this PA, Δf_{p-p} of 10 kHz must be used in equation 13.

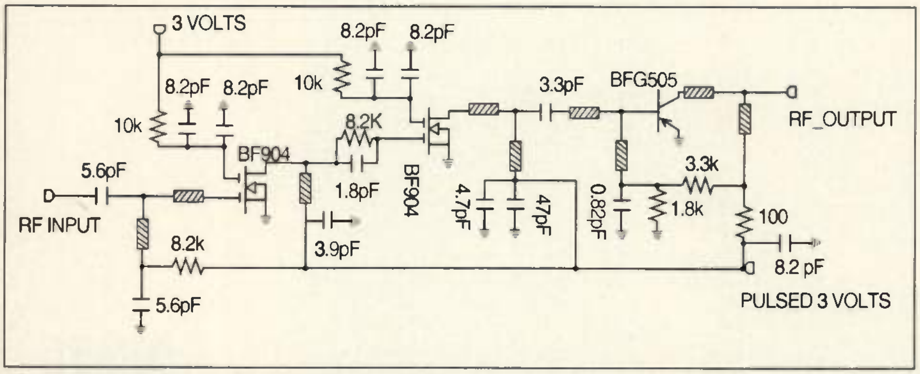

Figure 7a. Schematic for a three-stage, narrowband isolator for the transmit VCO.

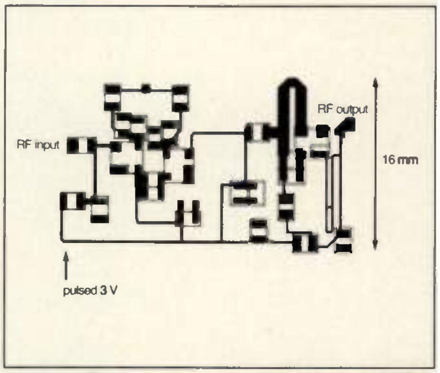

Figure 7b. Layout for isolator of Figure 7a.

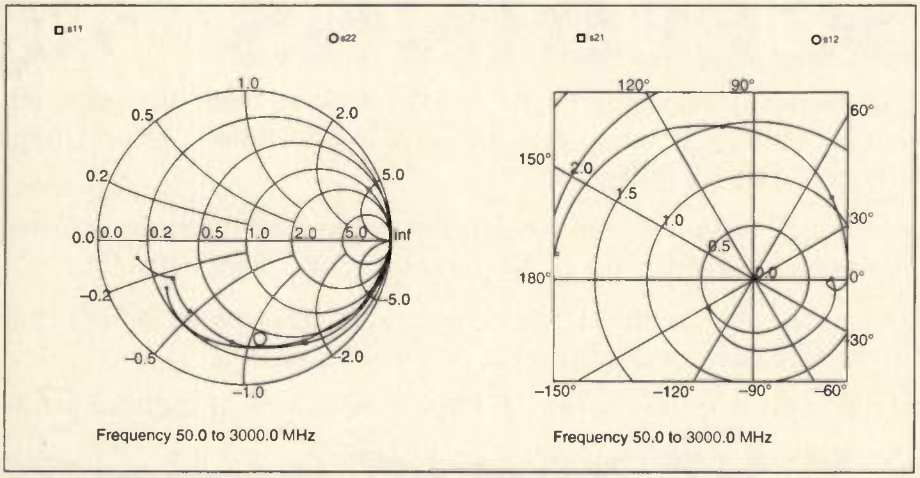

Figure 8. Measured S-parameters for the Philips BF904 at 3 volts, 5 mA.

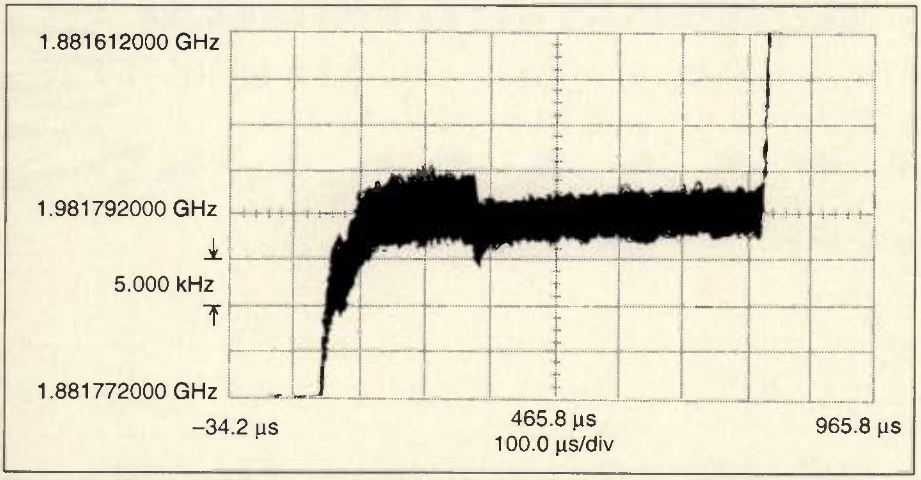

Figure 9. Measured frequency jump with the UMA1020 synthesizer and three-stage isolator.

Putting $\rho_L = 1$ and $Q_{ext} = 560$ in equation 13, we get $|S_{12}S_{21}| = 0.00148$, i.e. −56.6 dB.

3-stage Narrow Band Isolator Design

For an earlier version of the DECT radio, a 3-stage narrow band isolator was designed (Figure 7a) using Eesof Jomega version 3.0, in order to obtain an isolation that was better than $|S_{12}S_{21}| = -56.6$ dB. The isolator was used only for the TX VCO. Active devices with a small S_{12} were used.

Because the PA needs 3 dBm input drive and the TX VCO is specified to give −3 dBm nominal output power, the isolator must have 6 dB gain. Due to the power tapping to the RX LO and the synthesizer RF input, there is about 1 dB loss, and therefore the isolator gain should be about 7 dB.

The output stage of the isolator must source 3 dBm power. A Philips BFG505 bipolar transistor was selected for this stage. Collector DC feedback stabilized the bias at 5 mA collector current. A 3 V regulator supplies the entire isolator and VCO.

The BFG505 output stage was designed for 50 Ω input and output, with S_{21} of 12 dB, and S_{12} of −18 dB. To obtain the required overall isolation, two Philips BF904 low cost dual gate MOSFETS were used for the first and second stages. This transistor has an internal biasing circuit. Though its application is intended up to 1 GHz at 5 volts with superior cross-modulation during AGC, in the isolator it was found suitable as a buffer with high reverse isolation at 1.89 GHz. Its measured S-parameters with 5 mA drain current at 3 V supply, is shown in Figure 8.

The S-parameters of this device are strongly influenced by RF decoupling on gate 2. This gate 2 is normally used for the AGC control voltage, and does not draw any bias current. In the isolator it is tied to the static positive 3 V supply, and two 6.8 pF capacitors are used for RF bypass directly at the pin.

Gate 1 is the RF input, and it must be connected through a resistor to the pulsed positive 3 V supply. The drain bias current can be set by this resistor. For 5 mA collector current, this resistor must be about 100 kΩ. The drain supply need not be pulsed.

The BF904 can be turned ON or OFF by pulsing to 3 V and 0 V respectively, the supply voltage to the 100 kΩ resistor. Alternatively, the voltage to gate 2 (AGC setting) can be pulsed between 3 V and 0 V. Each of the first two stages were designed for a total matching gain of about 5 dB.

The S_{21} of this transistor is −2 dB, and therefore the overall gain of each stage is 3 dB. The reverse transmission gain, S_{12}, of each transistor is −32.4 dB. The S_{12} of each stage with matching gain is therefore −27.4 dB. The $S_{12}S_{21}$ product of the combined three stages is:

$$20\log|S_{12}S_{21}| = -18 - 27.4 + 12 + 3 + 3 = -54.8 \text{ dB} \qquad (15)$$

i.e. an isolation of 54.8 dB.

The forward gain is $20\log|S_{21}| = 12 + 3 + 3 = 18$ dB. Since approximately only 7 dB gain is actually required, a 10 dB pad was placed at the isolator input. The attenuator S_{12} and S_{21} is −10 dB. The overall isolation now is 54.8 + 20 = 74.8 dB. The circuit and layout of the isolator is shown in Figures 7a and 7b respectively. The actual frequency change due to the PA ramping was about 5 kHz with this three-stage isolator and 10 dB pad. It is shown in Figure 9.

Though an isolation of 74.8 dB has been calculated, in reality there is some extra leakage of signal from input to output of the isolator layout inside the shielded enclosure housing the VCO and isolator, even without the isolater transistors being mounted. By reducing the BF904 gate1 bias resistors from 100 kΩ to 82 kΩ, the overall forward gain achieved was 10 dB (3 dB more than the required value of 7 dB). This extra gain compensated for up to 3 dB possible decrease in VCO output power that was due to spread in production.

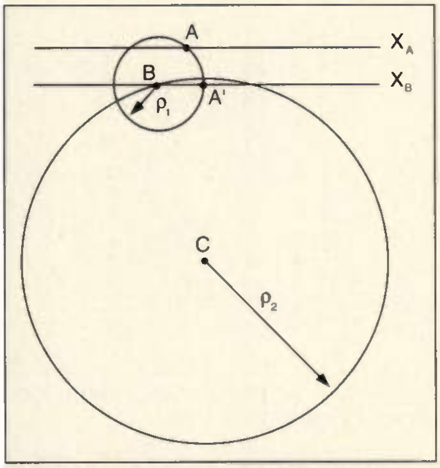

Figure 10. Constant VSWR circles for input impedance of PA and isolator.

Line Length Based f_{ss} Reduction

By carefully selecting the electrical line length between the isolator and the PA, the reactance change of the VCO load during power ramping can be made much smaller than the worst case as given by (13). By doing this, the extremely small $|S_{12}S_{21}|$ requirement for the isolator can be relaxed.

Figure 10 shows a mapping of the PA input impedance at the isolator input port. Point A represents the PA input impedance in the OFF state. By varying the 50 Ω line length between the PA input and the isolator output, point A moves on the small circle with radius ρ_1. Point B is for the ON state of the PA. Since the PA input impedance is close to 50 Ω in the ON state, point B is almost at the center of the small circle. Point C is the center of the isolator input S_{11} plane. ρ_2 is the radius of isolator input VSWR circle. By varying the 50 Ω line length between the VCO and the isolator input, the small circle moves along the large circle. X_A and X_B are constant reactance curves through point A and point B respectively. Since both ρ_1 and ρ_2 are << 1, the constant reactance curves X_A and X_B are nearly horizontal.

By selecting an appropriate line length between the isolator output and PA input, point A can be rotated clockwise and made to coincide with point A' so that the same constant reactance curve passes through both points A and B, resulting in no net reactance change before and after PA ramping.

2-stage Wider-Band Isolator Design

The Philips DECT radio as shown in Figure 1, uses a single stage BF904 for the main isolation, followed by a BFG505 stage for the gain and output power. This isolator serves as a buffer, both for the TX as well as the RX VCOs, and therefore has a wider bandwidth (more than 150 MHz). This avoided the use of a separate BFG505 buffer that was present in the earlier version of the radio, for the RX VCO. Also, only two stages were used in the isolator. Additional isolation was to be achieved later by proper selection of the line length between the PA and the isolator, as described previously.

Figures 11a and 11b shows the layout

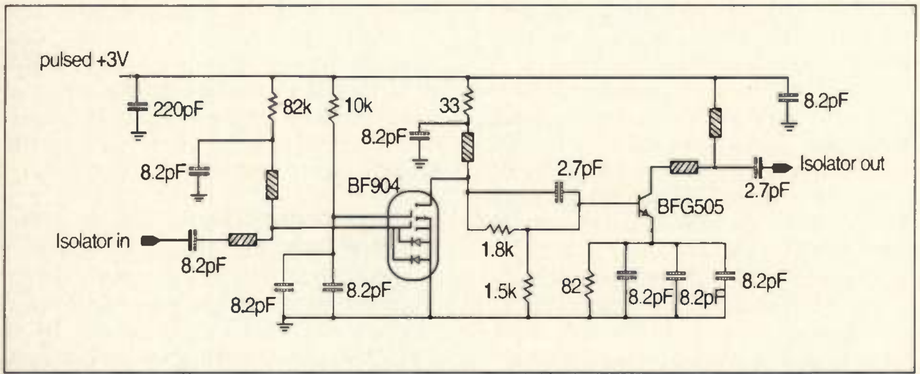

Figure 11a. A two-stage isolator with wider bandwidth (> 150 MHz) than that of Figure 7a.

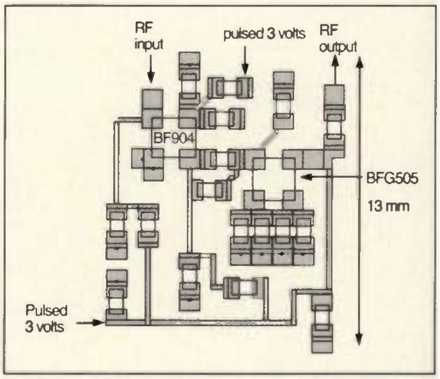

Figure 11b. Layout for two-stage isolator of Figure 11a.

and the simplified schematic (without full details of all microstrip elements that are present in the layout). Jomega was used for making the layout. The actual schematic was design-synchronised to the layout, and vice-versa.

The supply lines and decoupling capacitors were part of the design simulations. Measured S-parameters were used for the large supply decoupling capacitor. Equivalent circuit models have been used for the Philips surface mount 0603 resistors and NPO capacitors. The circuit was optimised for unconditional stability from 50 to 3000 MHz, and the gain difference between TX and RX frequencies was kept within 2 dB, in spite of small resonances in the BF904 transistors.

Another optimization criterion was good matching at the TX frequency and reasonable matching at the RX frequencies. The gain requirement was kept at more than 12 dB. High yield was obtained in the simulation, for spreads in the NPO capacitors, etching and substrate properties. FR4 material was used for the board, with 0.5 mm substrate height for the microstrip elements. The ε_r is 4.3 at 1.89 GHz, with a tanδ of 0.02. The simulated isolator results are shown in Figure 12. The measured result was close to the simulation.

Calculation of Frequency Jump

An attenuation of 4 dB exists between the paralleled VCO outputs and the isolator input. For the two-stage isolator itself, S_{12} = –49 dB, and S_{21} = 13.7 dB. The overall $|S_{12}S_{21}|$ is given by

$$|S_{12}S_{21}| = -4 - 4 - 49 + 13.7 \quad (16)$$
$$= -43.3\text{dB} = 0.006839$$

From (13), $\Delta f_{p\text{-}p}$ = 46.16 kHz when the PA input reflection coefficient moves on an infinite VSWR circle. The actual value of the frequency jump is half of this i.e. 23 kHz, because the PA input impedance moves from the high VSWR circle to it's center at 50 Ω. This frequency jump can be reduced further by selection of appropriate line length between the isolator output and the PA input.

Measured frequency jump was about 30 kHz i.e. more than the worst case computed value, because chip EMI filters were not used on the input lines to the shielded isolator and VCO enclosure. This resulted in increased VCO pulling from radiation leakage.

Conclusions

It may be said that reducing the TX VCO frequency jump to well within the TBR6 requirement, is the most difficult part of a low cost DECT radio design using open loop modulation. This article describes in detail the reasons for the frequency jump, with quantitative analysis of oscillator load-pulling and isolation requirement. Two active isolator solutions from a Philips DECT radio are given, with details on layout and simulation results. Actual measurement of the frequency jump shows that the requirements can be met using these solutions. *RF*

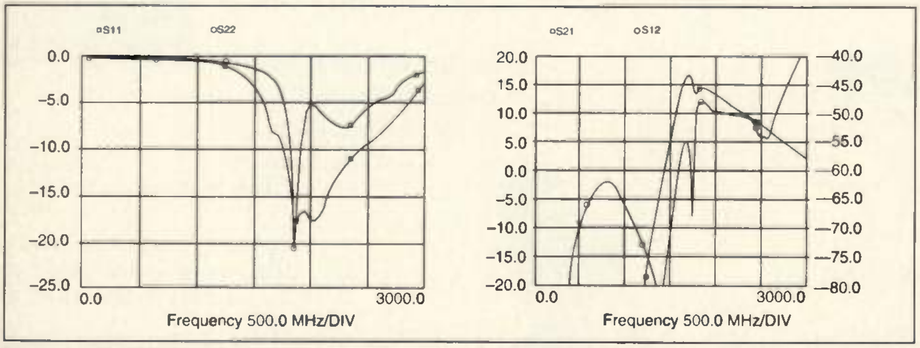

Figure 12. S-parameters for two-stage isolator. S_{11} and S_{22} are shown in the left graph, and S_{12} and S_{21} are shown in the right.

About the Author

Rishi Mohindra received his MEE (Master of Electronic Engineering) from PII, NUFFIC, at Eindhoven, The Netherlands, in 1990, and received his BE (ECE) (Bachelor of Engineering in Electrical Communication) from the Indian Institute of Science, at Bangalore, India, in 1988. For the last six years he has worked for Philips Semiconductors at the Product Concept and Application Laboratory in Eindhoven. He hold two patents and has part-time responsibility as superuser for Omnisys and jOmega microwave circuit and system simulators. He can be reached by E-mail at: mohindra@sce.philips.nl

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INFO/CARD 31

A Preview of the Dallas RF Design Seminars

The first of new RF Design Seminar Series will be held January 16-18, 1996 at the Sheraton Park Central Hotel & Towers in Dallas, Texas. The program emphasizes a series of short courses on fundamental, intermediate and advanced RF design techniques, along with a one-day program of technical papers. A group of sponsoring RF companies will also be on hand, presenting their latest products. For information on attending, call (800) 828-0420, ext. 323

Schedule-at-a-Glance

Tuesday, January 16

8:00 am – Noon	Wireless Communications for Non-Engineers (Part I)
8:00 am – 5:00 pm	RF and Wireless Engineering – Part I: Fundamental Concepts
	Practical High Frequency Filter Design
	Digital Modulation and Spread Spectrum for Wireless Comm.

Wednesday, January 17

8:00 am – Noon	Wireless Communications for Non-Engineers (Part II)
8:00 am – 4:00 pm	RF and Wireless Engineering – Part II: Amplifier Fundamentals
	Oscillator Design Principles
	RF Power Transistors and Amplifiers
Noon – 6:00 pm	Exhibit viewing opportunities
4:00 pm – 6:00 pm	Poster session and Ham Radio Reception in Exhibit Area

Thursday, January 18

8:00am – 5:00 pm	RF and Wireless Engineering – Part III: Amplifier Design
8:30 am – 4:30 pm	Technical paper presentations
9:00 am – Noon	M/A-COM Design Seminar
7:30 am – 5:00 pm	Exhibit viewing opportunities

SHORT COURSE DESCRIPTIONS:

RF and Wireless Engineering (3-day tutorial course)

Part I: Fundamental Concepts

The first day reviews fundamental ideas and terms that are important in all RF and wireless systems. Concepts such as gain, bandwidth, noise figure, dynamic range, resonance and Q are presented. Practical components are discussed and models for them are developed. Fundamentals of electromagnetic theory are reviewed, transmission lines are introduced, and practical realization of transmission lines is discussed. The Smith chart is developed from transmission line theory and its use in RF design is introduced.

Part II: Fundamentals of Amplifier Design

The second day begins by designing lumped element and distributed impedance transformation networks using the Smith chart. Next, the course presents a unique approach to impedance matching network design that facilitates the design of T-, PI- and L-networks for specified phase shift or Q. The procedure is extended to resistive attenuators and balanced networks. Active device models are introduced and concepts of power gain, noise figure and stability are reviewed. The theory, meaning and measurement of S-parameters are presented. Graphical and analytical techniques for amplifier design using S-parameters are developed. Fundamentals of computer-aided analysis and optimization are summarized.

Part III: Amplifier Design

The third day uses all the theory and techniques developed during the first two days to design several representative RF amplifiers. In addition to the basic RF design, practical topics such as bias network design, out-of-band stability, decoupling network design, the effect of microstrip discontinuities and the use of RF design software are presented.

Instructors for this course are Dr. Robert Feeney and Dr. David Hertling of the Georgia Institute of Technology. These professors bring both academic substance and practical design experience to the classroom. CEUs are available from the Continuing Education department at Georgia Tech.

RF Applications Courses (Intermediate level)

Practical High-Frequency Filter Design

HF Filter Design is a detailed review of L-C, printed and machined filter design and specification. Basic terminology and principles are discussed, followed by design procedures. Element models and unloaded Q are studied, then applied to case studies which match filter type to real-world applications. Some topics considered include: elliptic filters, determining the required order, component realizability, insertion loss, coupled resonator bandpass, ceramic resonators, filter match, controlled-phase filters, group-delay equalization, bandpass symmetry, printed filters and required tolerances.

The instructor is Randy Rhea, founder and President of Eagleware Corporation. Mr. Rhea is author of two books, *Oscillator Design & Computer Simulation* and *HF Filter Design and Computer Simulation*.

Oscillator Design Principles

A unified approach to oscillator design is presented which describes how to create high-performance oscillators using any type of resonator and any type of active device. Oscilators are de-mystified and fully understood for thyat design is no longer based on copying or modifying existing units. A complete understanding provides for known oscillation margins and design optimization. Both negative-resistance and open-loop Bode response design techniques are described. Gain margin, matching, starting, limiting and output level and harmonics are discussed.

The instructor for this short course is Randy Rhea.

RF Design Courses (Advanced Level)

Digital Modulation and Spread Spectrum for Wireless Communications

This advanced class covers the essential topic of digital communications for wireless applications. The course begins with principles and system architectures for complex I-Q, QPSK, PI/4 DQPSK, FQPSK and other modulation schemes. Standards for specific cellular and PCS systems are introduced. A theoretical analysis of spectral and power efficiency follows, including adjacent channel interference, linear/nonlinear amplification and BER performance. Advanced modulation techniques such as 16-QAM, trellis coding and multi-level FM are presented. A section on spread spectrum covers pseudo-noise sequences, correlation, direct-sequence, CDMA and slow frequency-hopping (TDMA) principles. The near-far problem and operation in mobile environments are discussed. The course assumes substantial experience in RF design.

The instructor is Dr. Kamilo Feher, professor at the University of California, Davis, and Director of DIGCOM, Inc. consulting group. He has authored five books on digital communications.

RF Power Transistors and Amplifiers: Principles and Practical Applications

RF power amplifier design is rarely taught in the depth provided by this course. Topics include the unique characteristics of RF power transistors compared to small-signal transistors, performance characteristics of bipolar and field-effect transistors (FETs), and the choice of devices and amplifier configurations. The principles and techniques of practical amplifier design include wideband impedance matching, frequency compensation, negative feedback, and guidelines for amplifier construction. The course assumes substantial RF design experience, and knowledge of small-signal amplifier design.

The instructor for this course is Norman Dye, an engineering consultant with long experience in Motorola's power transistor division. Along with Helge Granberg, he has recently authored the book *Radio Frequency Transistors*.

Wireless Communications for Non-Engineers

Part I: Communications Concepts and Common Uses

The first session of this course introduces the key concepts of signals, circuits, systems, and the radio spectrum. These technical concepts are presented in an intuitive, visual manner, relating them whenever possible to familiar activities of everyday life. The session continues with descriptions of common communications systems: AM and FM broadcasting, television, cable TV, and traditional two-way radio.

Part II: Modern Wireless Communications Systems

The second session covers modern wireless communications systems, including remote control systems, cordless telephones, the cellular phone system and satellite systems, Personal Communications Services (PCS), wireless local-area networks (WLAN), automatic toll collection and RF identification (RFID) tags are described.

The instructor for this class is Gary Breed, editor of *RF Design* magazine.

I&Q MODULATORS

Poster Session and Reception — Wednesday, January 17, 4:00–6:00 pm — Featured Presentation:

On-line Resources for RF Engineering

Charles Wenzel, Wenzel Associates and Russell Fish, Consultant

The following technical papers are scheduled Thursday, January 18, 1996:

8:30 – 11:30 am:

A Total Integrated EMC Test: An Essential Means to Meet the Challenge of the Automated Highway
Dr. Maqsood A. Mohd, Sverdrup Technology

Spurious and Harmonic Emissions Measurements from RF Transmitters in a WB TEM Cell
John D.M. Osburn, EMCO

Application of Network Synthesis to Capacitor Modeling
Joe Tillo, Ph.D., Ford Motor Company

1:30 – 4:30 pm:

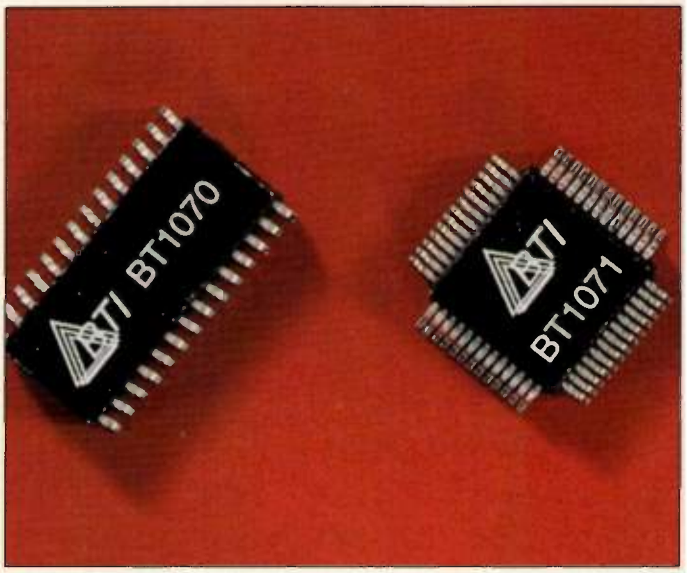
The Method of Moments: Applications to RF Engineering
Dr. W.D. Rawle, Axonn Corp.

Design and Fabrication of A Microwave Coaxial Resonator VCO
Ashok Nawarnage, Recoton

Low Noise Amplifiers for Wireless Applications
(tentative)

Spread Spectrum Chipset

A low cost, highly integrated spread spectrum/frequency hopping (SS/FH) two chip solution for cordless phone applications is being introduced by BethelTronix. The BT1070 is a BiCMOS RF transceiver front end for a 900 MHz SS/FH system. The RF transceiver includes a low noise (5.0 dB) amplifier, a down-conversion mixer, an IF amplifier driving a 50 Ω load, an up-conversion mixer, a class C power amplifier delivering from –3 dBm to +13 dBm, a local oscillator, and a divide-by-16 prescaler. The BT1071 ia a CMOS SS/FH IF transceiver. During recieve, the IC will despread an IF signal and downconvert to an FM signal. On transmit, it will spread a signal over a 6 MHz bandwidth. The hop rate is programmable for 5 to 20 khops/second. To minimize power consumption, time division duplexing is employed. The chip set operates at 2.7 to 3.6 V. Both ICs have a sleep mode option, and both chips are housed in a 48-pin TQFP package. [The BT1070 is in production in the TQFP package; the photo at right shows an early version of the part in an SOIC package.]
BethelTronix, Inc.
INFO/CARD #204



High-Frequency Design for PCs

Hewlett-Packard announces the release of HP Series IV/PC high-frequency design software for PC platforms. HP Series IV/PC EDA software provides a fully integrated schematic editor, multiple circuit simulator, component libraries, analysis presentations, and physical design tools for RF and microwave-circuit designers. Modules are available separately, and a Series IV/PC Touchstone Lite package and Series IV/PC Libra Design Suite are also available for users requiring only part of the capability of the full design suite. Series IV/PC software is priced from $8,000 (for a basic Touchstone Lite suite) and will be available in Spring 1996. Series IV/PC is supported on Intel 80486 and Pentium CPU-based personal computers using the Microsoft Windows 95 or Windows NT v3.51 operating systems.

HP EEsof
INFO/CARD #205

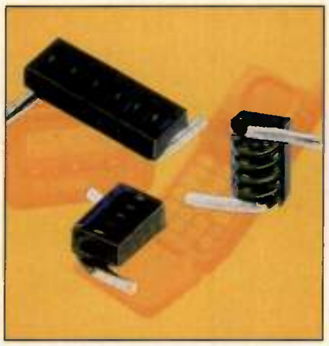
Surface Mount High-Power Couplers

RF Power Components' new line of surface mount 90° hybrid couplers and power terminations are now available from stock at Richardson Electronics. RF Power Components has developed SMD 90° hybrid couplers for use in high volume cellular and wireless applications. Frequency bands of operation are 815 - 960 MHz and 1750 - 1980 MHz with power ratings of 100 and 200 W CW. The couplers are constructed using bonded stripline circuitry with plated through technology. This allows for a small size of only 0.35 x 0.56 inches for 100 W and 1.0 × 0.5 inches for 200 W. The couplers are offered in bulk and in tape-and-reel. Terminations for the isolation port of the couplers are also available.

RF Power Components, Inc.
Richardson Electronics, Ltd.
INFO/CARD #206

Surface Mount Air Core Inductor

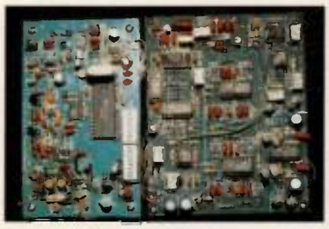
Coilcraft has introduced surface mount air core spring inductors measuring just 0.56 by 0.087 inches and 0.56 by 0.159 inches. Their height is 0.054 inches. The 0906/1606 Series have inductances from 1.65 to 12.54 nH. Tolerances are as good as 2 percent. Minimum Q of a six-turn part is more than 200 at 1.8 GHz. Minimum self resonance is rated at 4.9 GHz or higher. The two series are jacketed in a high temperature material that assures mechanical stability and forms a flat top for auto insertion. The leads are tinned for reliable soldering, and the parts are packaged in 12mm EIA tape-and-reel configuration. Coilcraft Designer's Kit C108 contains samples of all values. The parts are shipped from stock and cost $0.16 each in 10,000 quantities.



Coilcraft
INFO/CARD #207

Retrofit FM Exciter/Stereo Generator

JT Communications introduces replacement products for the FM RF chain. The model PLFM-100A replacement exciter module is a retrofit for obsolete FM exciters, eliminating the problem areas associated with old exciters. The PLFM-100A is a programmable, synthesized direct FM exciter on a 3 × 5 inch PC board and meets –80 dBc spurious response. The PLFM-100A contains composite and pre-emphasis audio inputs and operates at 12 VDC. The CSG-10 is a replacement stereo generator module containing a fully adjustable composite clipper. The module also includes active filtering, a digitally unbalanced modulator, and a crystal controlled reference. The CSG-10 measures 4 × 5 inches and operates from from bipolar 12 to 15 VDC supplies. Both modules meet all part 73 broadcast specs.



JT Communications
INFO/CARD #208

TEST EQUIPMENT

CDMA Propagation Analysis

The BVS Duet is a wireless transmitter and receiver system designed for characterization of RF-transmitted signals. The system consists of two pieces: a transmitter and a portable receiver. The transmitter uses direct sequence CDMA modulation. The receiver displays signal strength and bit error rate and can demodulate CW, AM, FSK, BPSK and QPSK. A GPS navigational system is also built in.

Berkeley Varitronics Systems, Inc.
INFO/CARD #209

Antenna Monitor

Narda's Model 72000 Tx/Rx antenna monitor offers precise Tx power measurements and Tx/Rx VSWR measurements for up to twelve individually-monitored antennas (the standard configuration holds six Tx antennas and six Rx antennas). Remote and local operation is available through the RS232 interface which provides access to all functions and controls.
Loral Microwave-Narda
INFO/CARD #210

GSM Equipment Test

Telecom Analysis Systems announces the TAS 4500 FLEX/GSM™ RF channel emulator. The instrument implements all six and twelve path propagation models required by GSM standards. Each path performs fading, propagation delay, log-normal shadowing, and attenuation functions. Price starts at $59,950.
Telecom Analysis Systems, Inc.
INFO/CARD #211

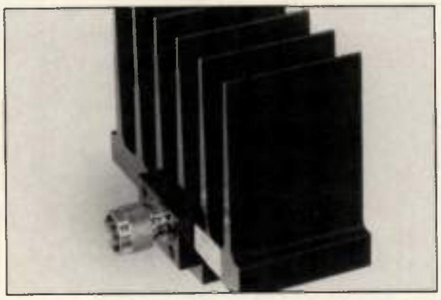
Frequency Distribution

Absolute Time™ has introduced the Model 512 frequency distribution unit. The Model 512 provides up to twelve precise reference signals to multiple users requiring low phase noise and extremely high port-to-port isolation. The device accepts a 10 MHz signal and sends out a user-definable combination of 10, 5, and 1 MHz signals. A secondary input port accepts and frequency between 1 and 15 MHz and distributes the same frequency on the output.
Absolute Time Corp.
INFO/CARD #212

SIGNAL PROCESSING COMPONENTS

High Power Attenuator

Model 65 from Weinschel is a high power fixed coaxial attenuator. The attenuator uses type N connectors and can handle an average 150 W (unidirectional) and 10 kW peak. Maximum SWR is 1.20, and the frequency range is DC to 2 GHz. Nominal attenuations of 3, 6, 10, 20, and 30 dB, with ±1.0 dB deviation.



Weinschel
INFO/CARD #213

Transmit/Receiver Duplexer

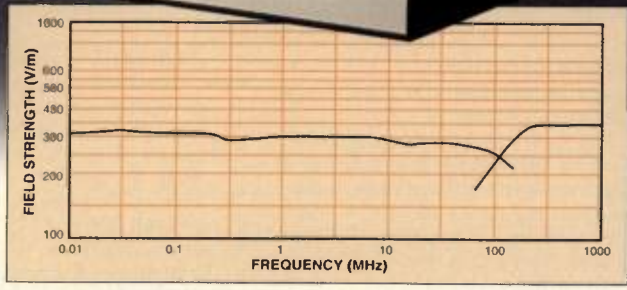
The DMT Division of Jay-El Products has developed a transmit/receive duplexer consisting of a transmit and a receive channel cavity filter. Tx passband is 5625-5850 MHz; Tx power is typically +42 dBm; max. Tx insertion loss is 0.7 dB; and Tx rejection is > 50 dB at 5475 MHz. Rx passband is 5250-5475 MHz, max. Rx insertion loss is 1.2 dB; and Rx rejection is typically 100 dB at 5625 MHz.
DMT
INFO/CARD #214

Phase Comparator

The PC 210 series of phase comparators from Technical Research and Manufacturing provides a 15 percent bandwidth at center frequencies from 10 to 2,000 MHz with quadrature IF outputs. The PC 210 comparators are available in flatpack, surface mount, and through-hole packages. Prices start at $599 in unit quantities.
Technical Research and Manufacturing, Inc.
INFO/CARD #215

Solderable Remote Switches

To facilitate connection directly to microstrip or stripline circuitry, RLC Electronics now offers solderable pin or tab versions of their SMA coaxial connector switches. These switches have greater than 60 dB isolation from DC to 18 GHz and insertion loss less than 0.4 dB. Prices start at $180.00 in small quantities.
RLC Electronics, Inc.
INFO/CARD #216



The AT5000 broadband transmission line matched to the 500A100 amplifier, and the AT1080 log-periodic antenna matched to the 1000W1000M7 amplifier, provide 200+ V/m testing from 10 kHz to 1000 MHz. Only one amplifier/antenna crossover is required. Other 200 V/m amplifier/antenna combinations are available up to 18 GHz.

One-stop 200 V/m susceptibility testing: 10 kHz to 1 GHz.

Sweep testing at 200 V/m in a highly reflective shielded room will tell you quickly what your test-equipment limitations are. For example, if your amplifier folds back or shuts down in the middle of a sweep, it probably can't tolerate the reflected rf power it's seeing. This happens often.

Or, if you must interrupt your testing repeatedly to exchange antennas or amplifiers, you're probably suffering from inadequate bandwidth and poor crossover matching.

The performance curve (above right) shows why we've been known for years as the broadband high-power people. It's a field-strength curve for amplifier/antenna combinations used for testing mid-size EUTs. The curve indicates the need for just two amplifier/antenna combinations, and only one mid-sweep crossover to achieve 200+ V/m fields from 10 kHz to 1 GHz.

Our all-solid-state Model 500A100 and 1000W1000M7 amplifiers (shown here) are extremely rugged and load tolerant. Neither will shut down in any load-VSWR condition; both provide 100% rated power up to 6.0:1 VSWR, and the 500A does so up to infinite VSWR.

If you're considering three or four amplifiers and as many antennas for 200 V/m testing over this frequency range, you're looking at a lot more hardware than you need, and more sweep-test interruptions for equipment changeovers.

Call one of our applications engineers today for similarly matched test-equipment combinations from dc to 18 GHz. **1-800-933-8181.**

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INFO/CARD 38

Harmonic Generators

KW Microwave has added 100 and 325 MHz harmonic generators to its product line. The 100 MHz harmonic generator requires P_{in} = +22 to +27 dBm input signal, and it outputs signals from 100 MHz to 2.2 GHz. The 325 MHz generator requires Pin = 0 dBm and f_{out} = 325 MHz to 4.5 GHz.
KW Microwave Corp.
INFO/CARD #217

AMPLIFIERS

High Power 2.1 GHz Amplifier

Model MSD-3438902 offers an output power of +41 dBm with an IP3 of +50.0 dBm in the 2.1 ±0.250 GHz band. The amplifier offers minimum gain of 35.0

dBm, with a maximum noise-figure of 8.0 dB. This solid-state amplifier has reverse bias and open/short circuit protection for any phase angles.
Microwave Solutions Inc.
INFO/CARD #218

Power Modules for GSM

SGS-THOMSON Microelectronics has introduced a hybrid RF module designed for use in GSM mobile transmitters. The STM915-14 is a quasi-linear RF power amplifier module that operates at 12.5 V and delivers minimum output power of 14 W over the 890-915 MHz range. The module is an AB class amplifier with maximum harmonic distortion of –45 dBc at full output power. The STM915-14 is available for $19.00 in quantities of 50k.
SGS-THOMSON Microelectronics, Inc.
INFO/CARD #219

400 W S-Band Amplifier

Chesapeake Microwave Technologies has introduced its APB 3135-400, a 400 W, silicon, Class C power amplifier. This unit is designed for short pulse operation – 2 us pulse width at 10 percent duty factor. The module has nominal gain of 26 dB and operates from a 36 VDC power source. Nominal size is 5.0 × 6.0 × 2.0 inches.
Chesapeake Microwave Technologies, Inc.
INFO/CARD #220

Multi-Channel Feed Forward Amplifier

The FFPA8689-12 from Microwave Power Devices is a solid-state, linear amplifier for cellular digital and analog applications. Using feed-forward techniques, this amplifier has IMD of 60 dBc. The amplifier operates over the full 869-894 MHz band at 100 W total average output power (4 × 25 W, fully redundant modules and an active combiner). Gain is 60 dB, and the amplifier is stable to infinite VSWR.
Microwave Power Devices
INFO/CARD #221

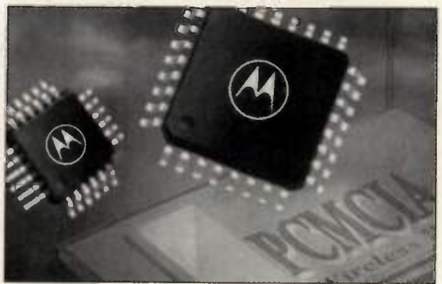
C-Band Amplifiers

Communications & Power Industries offers the SSCI series of 30 to 500 W high-efficiency C-band solid state amplifiers. These single-rack amplifiers offer single and multi-carrier satellite service in the 5.850 to 6.450 GHz band. Maximum intermodulation is –33 dBc with two equal carriers at total output power 7 dB below the rated single-carrier output. Remote monitoring and control are possible over the RS422/485 serial bus.
Communications & Power Industries
INFO/CARD #222

SEMICONDUCTORS

Narrowband FM IF

Motorola has introduced a narrowband FM IF subsystem with a coilless detector and excellent audio level output. The MC13150 has an onboard Colpits VCO for crystal controlled second LO in dual conversion receivers. The mixer is double balanced, with excellent third order intercept. It is useful to beyond 200 MHz. The quadrature detector does not require a tunable quadrature coil. The device operates from 2.5 to 6.0 VDC. Pricing in 10k quantities is $1.90.



Motorola Semiconductor
INFO/CARD #223

Frequency Counter Chip

Radio Adventures announces the availability of the C5 advanced CMOS frequency counter chip. The 28-pin DIP chip, along with a standard 74HC02 gate and three low-cost driver transistors, drives a six-digit, seven-segment LED display to 100 Hz resolution. Frequency range is DC to beyond 50 MHz. Unit price is $14.95.
Radio Adventures Corp.
INFO/CARD #224

CDPD Reference Design

VLSI Technology has unveiled its Geode™ CDPD reference design solution. The design incorporates VLSI's Ruby™ II chip and a separate radio interface IC, called Topaz. VLSI has compiled and tested a complete modem, giving designers a complete platform upon which to develop their application. Ruby II is a 32-bit ARM (Advanced RISC machine) processor with a complete set of communications peripherals. The Topaz radio interface IC provides flexible interfacing to a variety of radios. The Geode CDPD modem is available for $24.50 in 10,000 unit quantities.
VLSI Technology, Inc.
INFO/CARD #225

INFO/CARD 40

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3V Chip Set for PHS

PCSI's highly integrated chip set has)een designed for use in the Japanese Personal Handy Phone System (PHS) and :an be adapted to other TDMA systems such as DECT, PACS and WCPE. The :hip set consists of the PC 11101/PC 11102 baseband processors, PC 11501 nodulator, PC 11601 upconverter, PC 15301 frequency synthesizer, and PC 11301 power amplifier, low noise amplifi-er and Tx/Rx switches.

PCSI
INFO/CARD #226

ASIC Cells

American Microsystems has doubled its negacell offerings with the addition of 20 new cells. Among the cells are two 8-bit microcontrollers, an 8-bit microprocessor, a communication interface USART, and a programmable interval timer

American Microsystems, Inc.
INFO/CARD #227

Quadrature Mod/Demod

The MAX2450, a quadrature modulator/demodulator with local oscillator, and the MAX2451, a quadrature demodulator, operate from a single +3 V supply. The MAX2451 typically draws 5.8 mA in receive mode, while the MAX2450 typically draws 6.7 mA in transmit/receive mode. Both include a CMOS compatible shutdown mode. Prices start at $3.99 for the MAX2450 and $3.23 for the MAX2451 (in quantities of 1,000 an up).

Maxim Integrated Products
INFO/CARD #228

Wideband Limiting Amp

Anadigics has introduced a GaAs wide-band limiting amplifier, which converts data input voltage ranging in amplitude from 30 to 1000 mV to a constant output voltage of 500 mV. Features include operation from single +5 VDC supply, 25 dB small signal gain, 50 Ω output impedance, low pulse-width distortion, and limited output swing.

ANADIGICS
INFO/CARD #229

1 Watt Power Amplifier

TriQuint Semiconductor has introduced the TQ9142, a monolithic, high-power amplifier operating in the 824 to 849 MHz frequency range. The TQ9142 is optimized for use as a front-end transmit amplifier in AMPS telephones and CDPD WAN applications. Typical efficiency is 60 percent for a power output of 30.5 dBm while operating from 4.8 V. The three-stage amplifier is matched to 50 Ω at the input and is packaged in a power SO-16 plastic package. Unit price for orders of 50,000 is $4.25.

TriQuint Semiconductor, Inc.
INFO/CARD #230

SIGNAL SOURCES

High Frequency Digital Clocks

RF Monolithics has introduced a line of low-cost, high-frequency digital clocks. The new RFM clocks, available in frequencies from 250 to 700 MHz, are supplied in industry-standard metal dual in-line packages. Worst-case symmetry is 48 to 52 percent, and typical period jitter is 15 ps (peak-peak). In the 250 to 450 MHz range, the clocks cost less than $24 each in 1,000 piece quantities. In the 450 to 700 MHz range, the clocks are less than $26 each in 1,000 piece quantities.

RF Monolithics, Inc.
INFO/CARD #231

Digitally Compensated XOs

Fox introduces the FOX790 digitally temperature compensated crystal oscilla-

tor (DTCXO) series. An ultra compact quartz crystal unit, with an LSI chip providing digital processing temperature compensation circuits, are contained in a 28 pad LCC ceramic package measuring 11.4 × 11.4mm with a 2.2mm height. Frequency range is 12.000 to 19.200 MHz. Temperature stability is ±1 ppm max over –30 to +80 °C.
Fox Electronics
INFO/CARD #233

VCO for Portable Satellite

The V808ME01 VCO is designed for portable satellite applications in the 2,560 to 2,970 MHz range. This range is tunable with 2 to 20 V – the VCO exhibits an average gain of 50 MHz/V. The V808ME01 also

exhibits typical second harmonic suppression of –15 dBc. Typical phase noise performance is –87 dBc/Hz at 10 kHz offset. This VCO is housed in a package 0.50 × 0.50 × 0.20 inches.
Z-Communications, Inc.
INFO/CARD #234

Base Station OCXO

The 251-1521 model OCXO from MTI uses an industry standard package and produces an output frequency of 13 MHz. The oven-controlled crystal oscillator makes use of an AT-cut resonator to achieve thermal stability of 1.0 × 10^{-7} over –30 to +70 °C. Phase noise at 1 Hz offset is –75 dBc/Hz, and noise floor is –155 dBc/Hz. Output is either sine or HCMOS. The 251-series measures 2.00 × 1.61 × 0.98 inches. Versions using SC-cut crystals are also available.
MTI - Milliren Technologies, Inc.
INFO/CARD #235

TCXO Series

Champion Technologies introduces the K1600 Series of temperature compensated crystal oscillators (TCXOs), providing stability of ±2 ppm over –40 to +85 °C. The K1602 provides a sinewave output for a frequency range from 16.0 to 30.0 MHz. Other models are available with frequency outputs from 2.0 to 30 MHz. The TCXO is driven by a 5 V supply. Aging is less than 1 ppm per year. Control voltage is 0.5 to 4.5 V, centered at 2.5 V. Pricing for the K1600 Series starts around $30.00 (1,000 pieces).
Champion Technologies, Inc.
INFO/CARD #236

TCXO for GPS

A temperature compensated crystal oscillator (TCXO), model TQSMTV, from Tele Quartz has a frequency range of 10 to 26 MHz and a maximum can height of 4.5mm. Designed for mobile and satellite GPS applications, the oscillator has maximum current consumption of 2 mA with a supply voltage of 5 V. Frequency stability is better than 2.5 ppm within a temperature range of –25 to +70 °C. Frequency of the TQSMTV may be trimmed by about ±5 ppm via an an external voltage.
Tele Quartz GmbH
INFO/CARD #237

SUBSYSTEMS

HDTV/NTSC Antennas

UHF HDTV/NTSC antennas from Micro Communications allow broadcasters to transmit both NTSC and HDTV from one antenna. The antennas are broadband, covering 470-860 MHz, and have high power capability, making them useful for multichannel applications. VSWR is less than 1.1 from channel 14 to 69. The antenna is designed in doublett pairs to minimize downward radiation and to satisfy OSHA requirements. Using new offset techniques, low ripple omnidirectional patterns are obtained. Directional patterns are also available.
Micro Communications, Inc.
INFO/CARD #238

Narrowband PCS Antennas

Andrew announces the availability of its PG 900 Omni series of base station antennas. Designed for narrowband PCS two-way paging applications, these antennas have high power capability, high durability and low intermodulation generation. Antennas are available in the 901-902 MHz band, 928-944 MHz band, and in a full-band 901-944 MHz version.
Andrew Corp.
INFO/CARD #239

Low-Profile C-Band Antenna

A low profile antenna (PN 702410) from Electronics Development Corp. is a linearly polarized and operates over more than 12 percent bandwidth at C band. The antenna is 3/8-inch thick and is 1.5 inches in diameter. Operating frequency is 5350 to 6100 MHz, and boresight gain is 5 dBi. The antenna is available with SMA female connector or solder tabs. Price for 25 to 49 units is $170 each.
Electronics Development Corp.
INFO/CARD #240

CABLES & CONNECTORS

FCC Certified Connectors

RF Industries offers four connectors that comply with an FCC edict calling for a connector which prevents the replacement of certain antennas with antennas with higher gain. These four new connectors, which appear to be TNC but are not, include a male clamp for RF-174/316, an antenna mount, a right-angle male-to-female adapter, and a female bulkhead.
RF Industries, Ltd.
INFO/CARD #241

N to 3.5mm Adapters

Four new precision Type N to 3.5mm adapter have been introduced by Inmet. Connector configurations include Type N male to 3.5mm male, Type N male to 3.5mm female, Type N female to 3.5mm male, and Type N female to 3.5mm female. Frequency range is DC to 18 GHz, VSWR is 1.12:1, and impedance is 50 Ω. The 3.5mm connectors mate both electrically and mechanically with all SMA, 2.9mm, and Wiltron "K" series connectors.
Inmet Corp.
INFO/CARD #242

Field-Replaceable SMA

E.F. Johnson offers field replaceable SMA connectors, designed to mate with inexpensive 50 Ω hermetic seals. Featuring mechanically captivated contacts in economical brass bodies, this design eliminates excessive RF leakage that occurs in connectors with epoxy

fill holes. The series comes in plug and jack styles and in 0.500 × 0.375 inch and 0.625 × 0.223 inch flange configurations. The connectors will mate with 0.036, 0.020, 0.018, 0.015, and 0.012-inch pin diameters.
E.F. Johnson Co.
INFO/CARD #243

Low Profile Connector

Hirose has developed the E.FL connector series of surface-mount, low-profile, subminiature coaxial connectors. The mated height from the PCB is only 3mm, and the footprint is 9.4mm2 on the receptacle side. VSWR is 1.3 or less from DC to 2,000 mHz. Pricing is $1.94 per mated pair in quantities of 100.
Hirose Electric, Inc.
INFO/CARD #244

Capacitor Behavior at Radio Frequencies

By Gary A. Breed
Editor

The capacitor is an important component in RF circuits. It is used in resonant circuits, matching networks, coupling and decoupling. Like all components, real capacitors are not identical to the ideal components used in basic design equations. This tutorial describes the behavior of practical capacitors used in RF circuits.

We tend to think of components as individual elements in a circuit. At DC and low AC frequencies, this assumption is very nearly true. Those kinds of circuits will behave very close to mathematically derived performance. As the frequency of operation increases, incidental inductance and resistance increase in importance, becoming significant at RF.

The equivalent circuit of a capacitor is shown in Figure 1. C_0 is the nominal capacitance value, ESL is the effective series inductance contributed by the capacitor's wire leads and the body of the device, and ESR is the equivalent series resistance (loss). ESL and ESR will normally be frequency-dependent, and along with C_0, are also temperature-dependent. The magnitudes of ESR and ESL depend on the configuration of the capacitor (disc, chip, etc.), the material it is made from (mica, ceramic, porcelain, etc.), and the method of construction (multilayer, single layer, etc.). When analyzing an entire circuit, it is also necessary to include mounting features surrounding the capacitor, such as circuit board pads and traces, and solder joints.

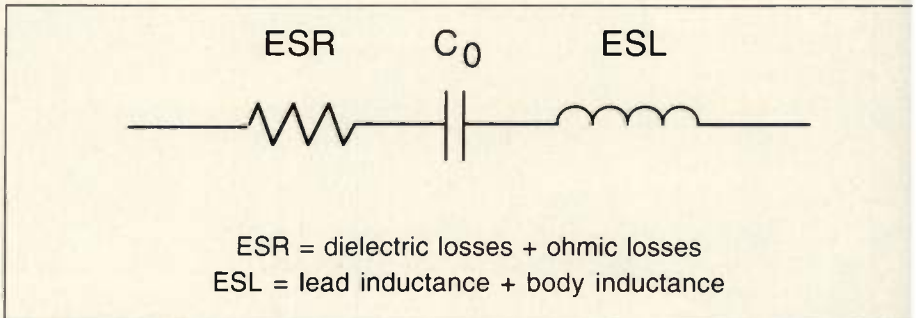


Figure 1. Capacitor equivalent circuit.

Effective Series Inductance

All wire leads are small inductors. Although components with wire leads are being rapidly replaced by surface-mount types, there are still plenty of them around. The inductance of a straight, round wire is given by [1]:

$$L = 0.0002b[\ln(2\,b/a) - 0.75]$$

where L is the inductance in μH, a is the wire radius in mm, and b is the wire length in mm

These values are for a wire in free space. Although more complex analysis might be more accurate in representing a particular installation, the above equation is sufficient to give us an idea of the magnitude of the lead inductance. For example, a 1/4-inch (6.3 mm) length of #20 AWG (0.8 mm dia.) wire has an inductance of 3.4 nH.

The body of a capacitor has a finite length and contributes to the inductance, as well. A chip capacitor may have a series inductance in the range of 1 nH or more, depending on its size. A capacitor with leads must include this value in addition to the inductance of the wires.

Effective Series Resistance

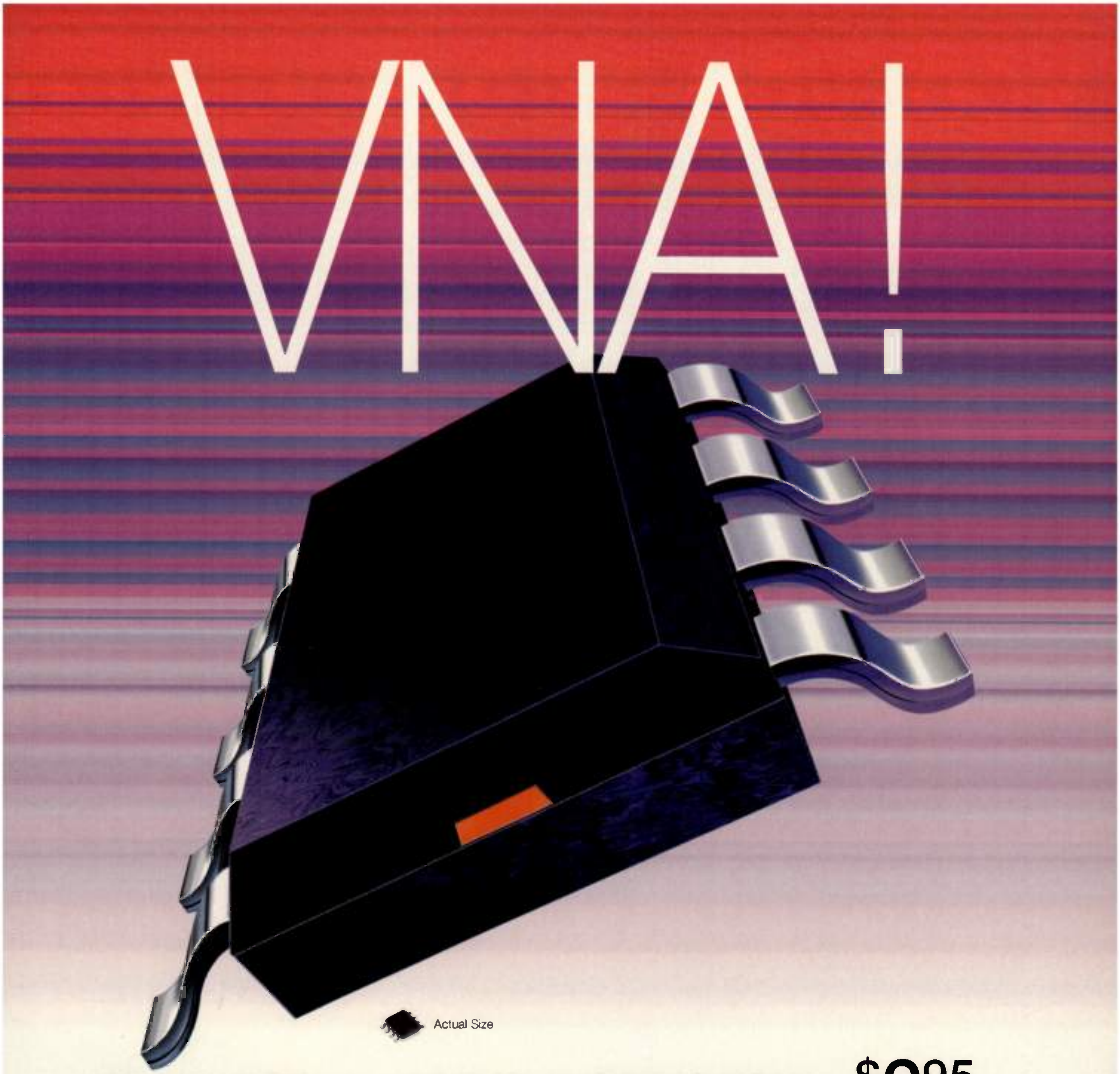
ESR consists of the combination of ohmic losses, such as contacts, fingers, leads, and bonded joints, plus dielectric losses. ESR is frequency-dependent, and will generally be in the range of 0.1 ohm to a few ohms. In some cases, ESR can be as high as a hundred ohms or more. Capacitor manufacturers provide specifications of ESR for the various styles of their devices. A very low ESR may be needed for a resonant circuit or matching network in order to keep the Q high and losses low. In a bypass or coupling application, ESR may be less important. To avoid surprises, be sure to include a realistic value of ESR when modeling and simulating RF circuits.

Series Resonance

Whenever inductors and capacitors are present in the same circuit, they create one or more resonant frequencies. At high frequencies, the effect of resonance on the impedance of a

Frequency (MHz)	X_C (Ω)	X_L (Ω)	X_{SUM} (Ω)	Equivalent Value
40	–39.8	+1.1	–38.7	103 pF
80	–19.9	+2.2	–17.7	112 pF
120	–13.3	+3.3	–10.0	133 pF
160	–9.9	+4.4	–5.5	181 pF
200	–8.0	+5.5	–2.5	318 pF
240	–6.6	+6.6	0	Short
280	–5.7	+7.7	+2.0	1.14 nH
320	–5.0	+8.8	+3.8	1.89 nH

Table 1. Performance of a "real" 100 pF capacitor versus frequency.



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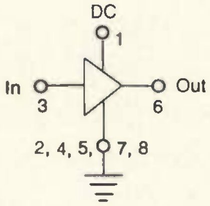

Freq.(GHz)	.5-.8	.8-1.0	1.0-2.0	2.0-2.5
Gain (dB) typ.	14.0	17.0	18.0	16.0
Max. Output (dBm) @1dB Comp. typ.	+18.0	+18.5	+17.5	+17.0
I P 3rd Order (dBm) typ.	+27	+27	+27	+27
VSWR Output typ.	1.5:1	1.7:1	1.7:1	1.5:1
VSWR Input typ.	6.4:1	2.8:1	2.0:1	1.4:1

DC Power.: +5.0V for specified performance.
Current,(mA): 85typ., 105 max.
Thermal Resistance, Junction-to-case: 125°C/W
Price ($) ea. : 2.95 (qty. 1000), 4.95 (qty. 10).

- All specs at 25°C (case temp. 35°).
- Available in Tape and Reel.
- MTTF at 150°C max. junction temp.: 3×10^7 hrs. typ.

"Case" is defined as mounting surface of leads.

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:apacitor can be dramatic! Using the previous example, a 100 pF capacitor vith 3.4 nH of lead inductance plus 1 ıH of body length inductance will ıave a series resonant frequency of !40 MHz.

The series inductive reactance is :ombined with the capacitive reac- ance, and as the frequency increases, he apparent capacitance value ncreases. Above the series resonant requency, the "capacitor" actually ehaves like an inductor! Table 1 lemonstrates this effect for the above .00 pF capacitor example.

For coupling and bypassing, we may e able to use series resonance to our dvantage. The impedance is mini- num at resonance, and our capacitor vill look like a short circuit. Caution is equired when attempting to do this — he mounting and circuit board inter- onnection inductance must be includ- d in series resonance calculations. The variation in capacitance value due o manufacturing tolerances and tem- perature swings must also be includ- d. In general, it is safest to select a apacitor for series resonance a bit ıigher than calculated, so it is still apacitive, not inductive.

Q or Dissipation Factor

This is a poorly understood quality neasurement. It relates stored energy pure capacitance) to dissipated ener- gy (loss). Dissipation factor is the reci- procal of Q.

Remembering that all values are fre- quency dependent:

$$Q = (X_{C_0} - X_{ESL}) / ESR$$

In other words, Q is the total reac- ance over the resistance *at the fre- quency of interest*. Q specifications for a capacitor are of little value unless he frequency at which Q was deter- nined coincides with the desired oper- ating frequency. Don't use manufac- urers' Q data without understanding ıow it was measured.

Q is a necessary specification for lesigning a practical oscillator or fil- er, since these circuits must include all non-ideal component characteris- ics. In a bypass or coupling applica- ion, Q is much less important.

Other Characteristics

Piezoelectric effects may be present in ceramic capacitors, causing changes in reactance at resonant frequencies in the 100s of kHz.

Although not described in this tutorial, RF current handling is a specific area that should be understood when designing amplifiers, high-power matching networks and filters.

In all cases, capacitors cannot be treated as ideal devices. At RF, the variations from idealized values can be significant and must be accounted for in all designs. *RF*

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1. *1990 ARRL Handbook*, American Radio Relay League, 1990, pp. 2-18, 2-19.
2. Mark W. Ingalls, Perspectives on Ceramic Chip Capacitors, *RF Design*, November 1989, pp. 49-53.
3. Les Besser, RF Component Modeling for CAD/CAE, *RF Design*, July 1993, pp. 64-70.

ENGINEER'S NOTEBOOK

Build A One-Tube Regenerative Receiver

By Mark Starin
New Hampshire Office of Emergency Management

In today's world of one-chip receivers and transmitters, surface-mount components, and digitally-controlled tuning, it is hard for us to imagine the revolutionary impact Edwin H. Armstrong's regenerative receiver design had on the communications industry of 1915. Armstrong's first regenerative circuit used a triode vacuum tube (see Figure 1) with a "tickler coil" to provide feedback and thus oscillation. This ability to oscillate meant that Armstrong's circuit could be used to either detect or generate continuous wave (CW) signals, a major technological breakthrough for the day.

To understand the nature of this breakthrough, we must first understand the state-of-the-art radio receiver prior to 1915. The typical receiver consisted of a crystal detector feeding a pair of high impedance headphones. Well-equipped stations used receivers with a triode vacuum tube audio amplifier stage (usually an Audion) after the detector for increased headphone volume. These receivers, however, could only detect signals from spark gap transmitters and offered little in the area of sensitivity or selectivity (interference was a major problem in the spark-gap era).

It is at this point that Armstrong's regenerative receiver circuit entered the scene, with these advantages:

- High selectivity
- Ample audio output with good quality
- Simple design
- Multimode detection capability

With the demise of the spark-gap transmitter and the advent of practical vacuum tube CW transmitters in the 1920's, these advantages were immediately utilized by the commercial radio industry. Like all designs, certain shortcomings became apparent after some use of the regenerative receiver (Armstrong himself corrected these later with his invention of the superheterodyne receiver). The disadvantages included the following:

- Detector leakage via the antenna (usually fixed with ar RF stage ahead of the detector)
- Detector "lock-on" to strong RF signals
- Frequent re-adjustment when receiving AM

Despite these disadvantages, regenerative detectors wer used in some HF receiver designs until the 1960s. Althougl the regenerative receiver is rarely seen today, the supe regenerative detector (another Armstrong invention), is stil used where simplicity of design and low-cost are important.

Circuit Description

This article describes an electron-coupled regenerativ receiver (using a 6SK7 pentode tube) which can be easil duplicated (Figure 2). The grid of the tube corresponds t the diode plate and the rectifying action is the same as i the diode. The DC voltage resulting from the rectified cur rent flow through the grid leak resistor (Rl) biases the gri negatively, and the audio frequency variations in voltag across R1 are amplified through the tube as in a normal AI amplifier. The filter choke in the plate circuit serves as th plate load resistance while C3 and the RF choke act as a fil ter to eliminate RF in the output circuit. Feedback in thi circuit results from connection of the tube cathode to the ta on the input coil; the potentiometer in the screen grid of th tube controls the level of feedback or regeneration.

I wish to thank Thomas Taylor, N1HCI, for his invaluabl assistance in the preparation of this article. RI

About the Author

Mark Starin is the Communication & Warning Officer for the New Hampshire Office of Emergency Management in Concord, NH. Prior to his state service he was a technical writer in the defense and computer industries for 15 years. He is also amateur radio operator KB1KJ.

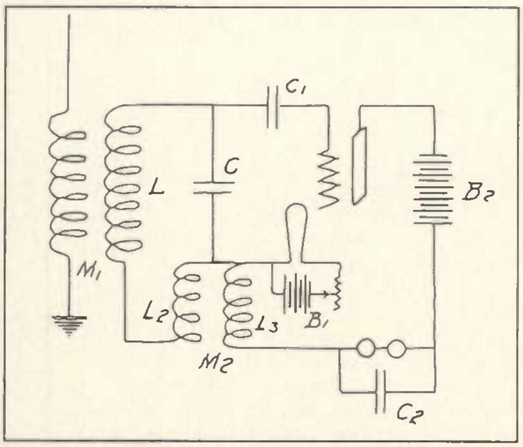

Figure 1. Armstrong's original circuit.

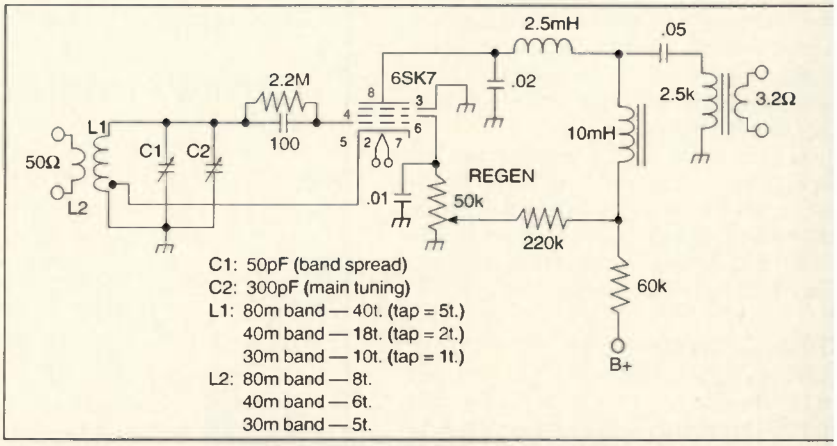

Figure 2. The author's updated regenerative receiver circuit.

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Operating Frequency (MHz)	Model Number	Gain (dB, Min.)	Gain Flatness (±dB, Max.)	Input/Output VSWR	Noise Figure (dB, Max.) Fl	Fm	Fh	Output Power (dBm, Min.)	1-9 Piece Price*
.01–500	AU-1310	30	0.5	2.0:1	1.3	1.4	1.5	8	$325
.01–500	AU-1332	45	0.5	2.0:1	1.3	1.4	1.5	10	$350
.01–1000	AM-1300	25	0.75	2.0:1	1.4	1.6	1.8	6	$375
1–100	AU-2A-0110	30	0.5	2.0:1	1.2	1.2	1.3	3	$275
1–100	AU-3A-0110	50	0.5	2.0:1	1.2	1.2	1.3	10	$300
1–500	AU-1A-0150	14	0.5	2.0:1	2.7	2.8	2.9	10	$200
1–500	AU-2A-0150	30	0.5	2.0:1	1.3	1.4	1.5	8	$275
1–500	AU-3A-0150	45	0.5	2.0:1	1.3	1.4	1.5	10	$300
1–500	AU-4A-0150	60	0.5	2.0:1	1.3	1.4	1.5	10	$325
1–1000	AM-2A-000110	25	0.75	2.0:1	1.4	1.6	1.8	8	$300
1–1000	AM-3A-000110	35	0.75	2.0:1	1.4	1.6	1.8	9	$350
20–200	AU-1158	30	0.5	2.0:1	2.7	2.7	2.7	17	$275
50–90	AU-1001	14	0.25	1.3:1	5.0	5.0	5.0	18	$200
50–90	AU-2A-1158	30	0.25	1.3:1	2.7	2.7	2.7	20	$275
50–90	AU-3A-1263	43	0.25	1.3:1	1.5	1.5	1.5	20	$325
50–350	AU-1210	18	0.5	2.0:1	2.6	2.7	2.8	10	$200
100–1000	AM-1331	35	0.75	2.0:1	1.4	1.6	1.8	15	$400
100–2000	AMMIC-1348	14	1	2.2:1	5.0	4.3	4.6	14	$395
100–2000	AMMIC-1318	6	1	2.2:1	4.5	4.0	4.0	12	$350
500–1000	AMMIC-1141	10	0.5	1.5:1	6.0	6.0	6.0	10	$200
500–1000	AM-2A-0510	24	0.5	2.0:1	1.4	1.5	1.6	0	$300
500–1000	AM-3A-0510	37	0.5	2.0:1	1.4	1.5	1.6	9	$350
500–1500	AM-3A-0515	30	0.5	2.0:1	1.5	1.8	2.2	4	$375
500–2000	AM-2A-0520	19	0.75	2.0:1	1.4	1.9	2.4	-4	$350
500–2000	AM-3A-0520	30	0.75	2.0:1	1.4	1.9	2.4	5	$400
500–2000	AM-4A-0520	40	1	2.0:1	1.4	1.9	2.4	5	$450
1000–2000	AM-2A-1020	19	0.5	2.0:1	1.8	2.1	2.4	4	$325
1000–2000	AM-3A-1020	30	0.5	2.0:1	1.8	2.1	2.4	10	$375
1000–2000	AM-4A-1020	40	0.75	2.0:1	1.8	2.1	2.4	10	$400

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RF Transformers Part 2: The Core

By Nic Hamilton
RAF Signals Engineering Establishment

In the first part of this article (June 1995, pg. 36) the different winding topologies for wideband, ferrite-loaded transformers were introduced. This article concentrates on selecting core shapes and material types. [Note: Figure numbering is continued from part 1 of this article, ed.]

Some ferrite manufacturers sell "designer's kits" of cores with a wide variety of shapes, sizes and materials. I suspect that these are not really used as designer's kits, but as bodger's kits, where the idea is to keep on winding transformers until one works. This part of the article will encourage a better engineering approach to the subject.

Transformer Loss

Imagine a transmitter connected to a load using a length of coaxial cable. It is clear that winding the cable onto a ferrite core will cause no additional signal loss: this is because the current flowing on the inner conductor exactly cancels the current flowing in the outer conductor so there is no magnetic flux in the core.

Now imagine that the case of the load is RF 'hot'. There will now be a small out of balance current flowing through the windings. This is referred to in the text books as the magnetizing current. Because it causes a magnetic flux in the core, it will have losses associated with it.

For simplicity, a simple Guanella phase inverter wound on a Philips type RCC14/5-4C65 core will be discussed first. The core is torroidal, with an external diameter of 14.5mm and is made of 4C65 grade material. The circuit diagram of this transformer is shown in Figure 17. Physically, this transformer consists of a few turns of miniature coaxial cable (type MCX) wound on a torroidal core. At the center of the winding, the inner conducto is connected to the outer of the seconc half of the winding, and vice-versa.

Figure 17 also shows a lumped ele ment equivalent circuit of the trans former. Its low frequency loss is dom inated by the low reactance of L_P, th winding inductance. This is well cov ered in the literature. In contrast authors have had considerable diffi culty in predicting the mid-banc insertion loss and high frequency performance of transformers. Th mid-band loss is dominated by R_P the parallel equivalent core loss. Th high frequency loss was discussed i Part 1, but it is sometimes dominat ed by C, the overall winding capaci tance. To predict the transforme loss, values for L_P, R_P and C must b estimated. In the following para graphs, the values of L_P and R_P change with frequency.

Predicting anything about inductors and transformers is notoriously diffi-

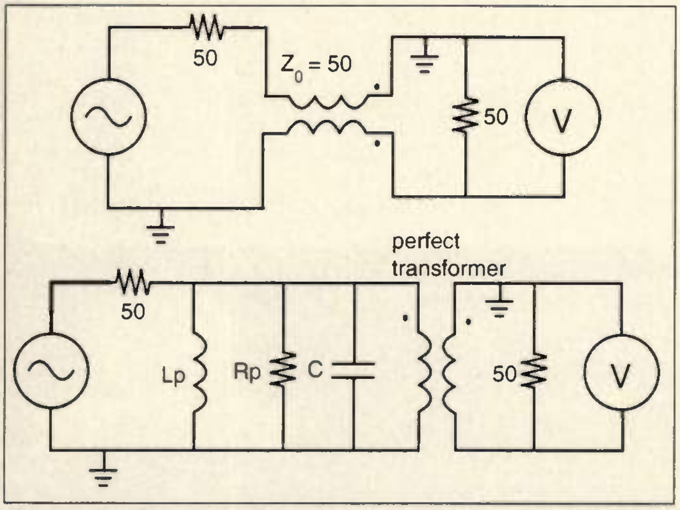

Figure 17. The Guanella inverter with a lumped element equivalent circuit.

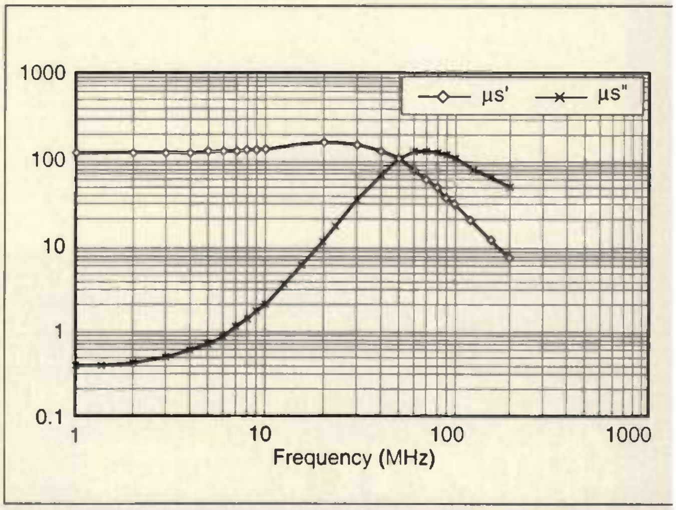

Figure 18. Complex permeability spectrum of Philips 4C65 material.

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cult. To simplify the problem, make some assumptions:

a) The Q of the coaxial choke is dictated solely by the core material
b) The windings have no losses
c) There is no dielectric loss in the core
d) Small signal conditions apply at all times

Be aware that this is skating on rather thin ice, but proceed with caution.

Normalized L and R

To estimate L_P and R_P of an inductor, start with the manufacturer's graph of the complex permeability spectrum for the material to be used. Amongst others, Neosid, Philips and Siemens publish their data in this format.

Put in simple terms, the impedance of an inductor, Z_L is proportional to $+j\mu$, where μ is the permeability of the material, and is one of the factors which determines the inductance of a coil with a magnetic core. The permeability can itself be represented as a complex number. This complex permeability is given by:

$$\mu = \mu' - j\mu''. \qquad (1)$$

Substituting one formula into the other gives,

$$Z_L \propto \mu'' + j\mu'. \qquad (2)$$

So μ_s' is related to the coil's inductance and $\omega\mu_s''$ is related to the coil's resistance, where the subscript s indicates that the two components are connected in series. These can also be recalculated and expressed as parallel components μ_p' and μ_p''. For this, and subsequent calculations, the use of a computer spreadsheet takes a lot of work out of the job.

Figure 18 shows the graph of the series complex permeability spectrum for Philips 4C65 ferrite. The data have been extracted from [10], but the LF data for μ_s'' have been extrapolated below 8 MHz. This will be discussed later.

The curves for μ_s' and μ_s'' cross at f_x, which for Figure 18 is 50 MHz. At this frequency, an "inductor" wound on this core would have a Q of about 1. At a higher frequency, the impedance of the "inductor" would be almost purely resistive. However, for a low power transmission line transformer, the only requirement is that the bifilar choke should have a high impedance, never mind whether the impedance is resistive or reactive.

The manufacturer has not specified the full complex permeability above 200 MHz, however the ferrite, used in a transformer core, will work up to at least GHz. There are two reasons for this:

a) The complex permeability data are measured on a core of stated size, however the permeability becomes more dependent on the core shape and size as the frequency is raised above f_x.

b) The data on which Figure 18 is based were intended for designers of inductors. Above f_x, an inductor will have a Q of less than 1, and therefore be useless.

Figure 19 shows the same data as Figure 18, but re-calculated in terms of normalized parallel components, R_{norm} and X_{norm} (in Ωmm/turn²). The values are given by:

$$R_{norm} = 8\pi^2 f\left(\mu_S'' + \frac{\mu_S'^2}{\mu_S''}\right) \times 10^{-4} \qquad (3$$

$$X_{norm} = 8\pi^2 f\left(\mu_S' + \frac{\mu_S''^2}{\mu_S'}\right) \times 10^{-4} \qquad (4$$

where f = frequency in MHz.

(All the formulae given in this arti cle are based on those given b Snelling [2].)

Some manufacturers (e.g. Fair-Rit and Ferronics) give data for trans former core materials in graphs simi lar to Figure 19.

Note that $X_{norm} \propto$ f. This is becaus the normalized parallel inductance i approximately constant. Also not that, above 20 MHz, the value of R_{norr} is approximately constant with fre quency. This is important.

The shape of the graph in Figure 1 is typical of a ferrite designed fo inductors with a moderate Q. Fo lower Q ferrites (e.g. for transformers the bulge in the spectrum of Rnorn above 100 Ωmm/turn² may be less pro nounced. For ferrites for RFI suppres sion beads, the Q of the device is of n importance, and the spectrum of R_{norr} may rise continuously, without havin any maximum value.

Constanst R_{norm}?

Many manufacturers publish com plex permeability data for their fer rites over a limited frequency range Some do not publish them at all. I either case, a crude, but useful rule o thumb is that near, and above f_x, th value of R_{norm} is approximately con stant at about 60 Ωmm/turn². Thi does not apply for frequencies much

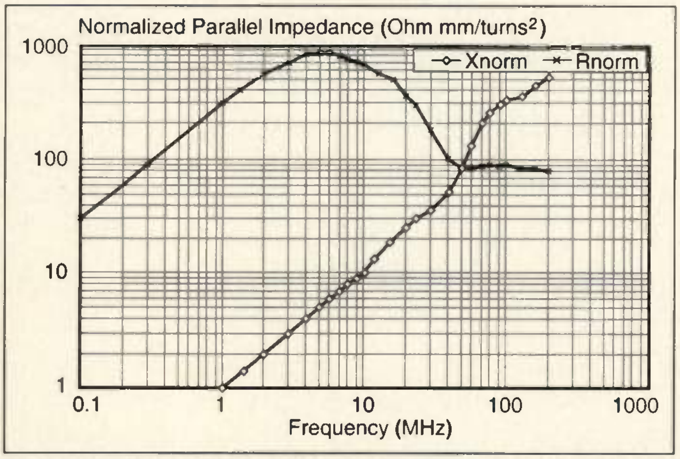

Figure 19. Normalized resistance and reactance for Philips 4C65 material.

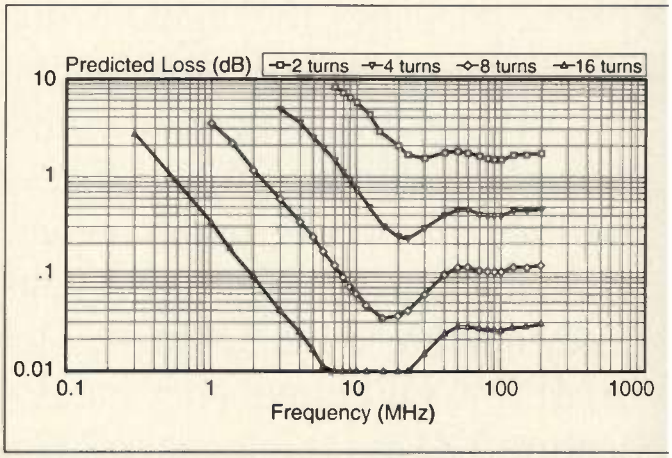

Figure 20. Predicted loss for different numbers of turns or RCC 14/5-4C65 toroid.

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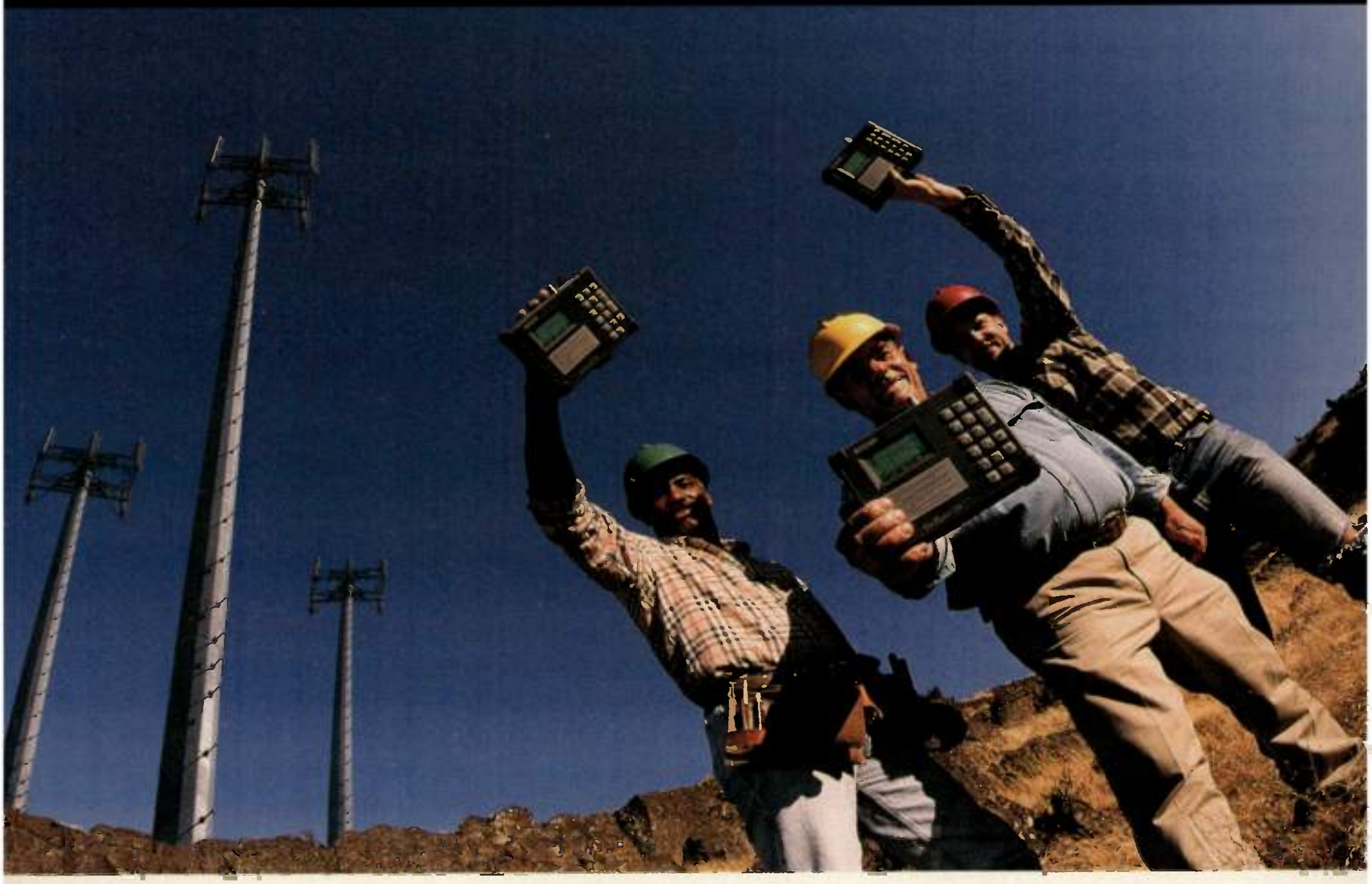

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less than f_x.

For narrow-band use, (one decade or less), more careful selection of the material type based on the complex permeability data can result in designs with much higher values of R_{norm}. This gives a higher resistance per turn, but will only be necessary for very high power or high impedance transformers.

The complex permeability data indicate that ferrites with a high Cobalt content [11] have $R_{norm} > 60$, but these cannot be used for power applications. I have not tested them.

It can now be seen why radio amateurs have experimented with such a wide range of different materials for their transformers, and with so little definite conclusion: the main factor affecting transformer loss is Rp, and this is approximately independent of the core material.

Conventional wisdom says that a ferrite material with a small value of μ_i should be selected if the transformer loss is to be small. However, this means that the number of turns must be increased in order to keep the same value of primary inductance. This leads to a satisfactory result, because $Rp \propto N^2$. However, the same number of turns with a higher μ_i ferrite would have given a similar reduction in loss, but would give superior low frequency performance.

Finding L_P and R_P

To de-normalize R_{norm} and X_{norm}, use

$$R_P = \frac{N^2 R_{norm}}{C_1}\,\Omega \qquad (5)$$

$$X_P = \frac{N^2 X_{norm}}{C_1}\,\Omega \qquad (6)$$

where N = number of turns

C1= core factor in mm^{-1} (2.84 for RCC14/5 core in example). (Note that there is another core factor: C2. It is not used in this article)

The inductance is given by

$$L_P = X_P/\omega. \qquad (7)$$

It should be roughly equal to the inductance calculated from the manufacturer's published value of A_L.

Some manufacturers use alternative symbols for C1, for example C1, $\Sigma(1/A)$ or I_e/A_e. Some manufacturers do not give values for C1 at all, however it can be calculated:

$$C_1 = \frac{0.4\pi\mu_i}{A_L}\,mm^{-1} \qquad (8)$$

where μ_i = initial permeability of the ferrite, and A_L = inductance factor in $nH/turn^2$.

Predicting the Guanella Inverter's LF and Mid-Band Loss

To calculate A_i (in dB), the insertion loss of the Guanella inverter, use:

$$A_i = 10\log\left[\left(1+\frac{Z_0}{2R_P}\right)^2+\left(\frac{Z_0}{2X_P}\right)^2\right] \qquad (9)$$

where Z_0 = line characteristic impedance.

Figure 20 shows the final prediction for the loss of Guanella inverters with various numbers of turns on the RCC14/5-4C65 core. Figure 21 shows measurements of the real thing. The measurements were made on four different RCC14/5-4C65 cores, and were compensated for the loss of the coaxial cable used in the windings. The graphs become rather noisy below 0.1dB due to measuring equipment limitations.

The measured values fit the predictions quite well. Residual inaccuracies are due to the assumptions made at the start of the calculations, and because the published A_L has a ±25% tolerance. This tolerance is due to variability of both the material performance and of the core's physical dimensions. The tolerance is even greater at HF because μ_p' and μ_p'' vary with differing core size and shape.

Earlier, I stated that I had extrapolated the manufacturer's complex permeability data below 8 MHz. At low frequency, the transformer loss is dominated by X_P, the reactance of L_P. This is calculated from μ_p' which is constant at low frequencies. It can be extrapolated without fear of error. The value of R_P, which is calculated from μ_p'' has only a very small effect on the transformer loss. I have extrapolated the spectrum of μ_p'' below 8 MHz using a curve with a shape that is typical of ferrites.

LF and Mid-Band Loss of Other Guanella Transformers

The insertion loss of the complex Guanella transformer (i.e. composed of sub-transformers, and not just an inverter) can be calculated by finding the values of R_P and L_P across each sub-transformer. The differential voltage is then calculated across each of

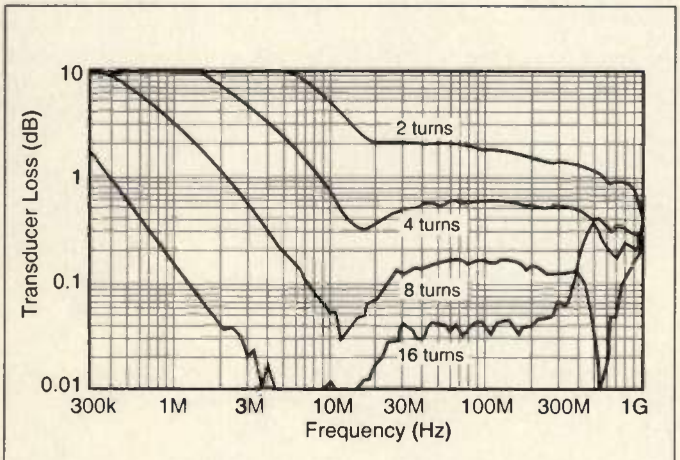

Figure 21. Measured loss for 2, 4, 8, and 16 turn Guanella inverters on RCC 14/5-4C65 toroid.

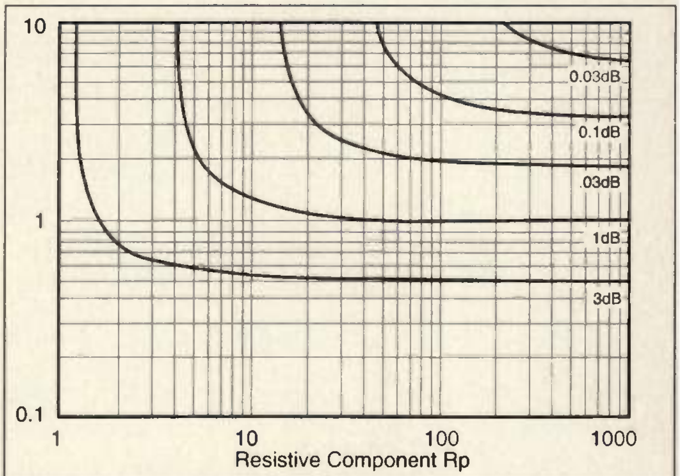

Figure 22. Transformer insertion loss as a function of Rp and Xp where Z_0 = 1 Ω


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these components, and so the insertion loss can be found. This is another of the tasks that is best done with a computer simulation. Where L_P and R_P are not constant, S parameters must be used for each transformer component, and simulated using one of the newer sons of SPICE 2G.6.

It is possible to calculate the Guanella transformer's loss using matrices [12], but it is still best to use a computer for this.

The Insertion Loss Formula

It is hard to visualize the implications of equation 6, so Figure 22 shows a graph of the loss for various values of R_P and X_P, where Z_0 = 1Ω. This shows that the transformer's insertion loss is much more dependent on the value of R_P than on the reactance due to the winding inductance or the capacitance. As an example, for an insertion loss of 0.3 dB, the reactance should be about four-times the circuit impedance, whereas the resistance should be about 20-times the circuit impedance. An intuitive explanation for this is that reactance is non-dissipative, and serves only to increase the loss by increasing the VSWR. Resistance both increases the VSWR and dissipates power.

Auto-Transformer Loss

Figure 23 shows the graph of the insertion loss of a 50 to 75 Ω transformer wound on a Siemens B62152-A8-X30 double aperture core with a turns ratio of 9:11. The core is made of N30 grade material. Unfortunately, the manufacturer does not give complex permeability data for N30 grade for frequencies above 4 MHz. So, using R_{norm} = 60 Ωmm/turn^2, N=11, and the manufacturer's figure of C_1 = 1.78 mm^{-1}, equation 3 gives the result R_P = 4 kΩ. Because the Q of the winding will be very low, the term X_P due to the parallel inductance can be ignored, and equation 9 can be simplified to:

$$A_i = 20\log\left(1+\frac{Z_N}{2R_P}\right)\text{dB} \qquad (10)$$

where Z_N = Impedance connected across the N turns (75 Ω in this case).

This gives an estimated loss value of 0.08 dB, which is close to the value seen in Figure 23. The insertion loss rises at high frequencies towards a maximum at 1.4 GHz. This is difficult to model using SPICE, as a full model of this transformer uses 55 transmission lines. The core is small: it has two holes of 0.9mm diameter separated by 1.5mm. It is a good test of hand-eye coordination threading 11 turns through it.

HF Performance: Core Material

From my experiments, transformers with Manganese-Zinc ferrite cores give higher values of overall winding capacitance C (as defined in Figure 17) than transformers with cores of other materials. This is probably because Manganese-Zinc ferrite has such a high dielectric constant, and such low resistivity, that the core acts as a conductor. As a result, HF performance can be dominated by C. For this reason, these cores should have a plastic coating. This interposes a layer with a relatively small ε_r between the windings and the core, and thus reduces the winding capacitance. Some early transformer designs used un-coated cores, and called for an extra insulating layer to be wrapped round the core.

If C has been determined, the calculated capacitive reactance may be substituted for values of X_P in equation 9.....

For Nickel-Zinc or dust-iron cored transformers, it is less likely that the HF performance of a transformer will be determined by C. There are other sources of loss which include radiation loss and the loss due to the impedance of the spurious transmission line between the line and ground.

Finding C: The Impedance Spectra of Inductors

Figure 24 shows a graph of the impedance of a coil consisting of a ferrite toroid with various numbers of bifilar turns. The values of impedance were calculated from the values of series insertion loss measured in a 50 Ω system.

At all frequencies, doubling the number of turns gives an impedance that is increased by a factor of four. This is as expected of an inductor: if N turns give an inductance L, then 2N turns give an inductance of 4L. This is composed of the self inductance of 2 lots of N turns, which is 2L, plus their mutual inductance which is also 2L. This assumes that the coupling coefficient is 1. From this it may be seen that, at low frequencies where capacitances can be ignored, a core with 16 turns will have an impedance equal to 16^2 = 256 single turn cores in series (e.g. 256 beads threaded on a wire). For this reason it is better to use multiple turns round a single core, providing the end-to-end capacitance does not degrade the high frequency performance too much.

Figure 25 shows the impedance achieved for various values of bifilar choke inductance and capacitance. It can be concluded from this that the 24 turn choke in Figure 24 had a capacitance of about 1 pF. This caused the sharp roll-off of impedance from 20 to 200 MHz. Above 200 MHz, resonance of the wire lengths caused the erratic behavior up to 1 GHz.

Core Shape: Guanella Transformers

To take full advantage of the Guanella transformer's good high frequency performance, the inter-turn capacitance between the transmission-line windings must be minimized, so the use of beads/sleeves /tubes for a '1 turn winding' is to be preferred over toroids with multi-turn windings.

The beaded line is useful where the minimum frequency is above 30 MHz. Choose a thin transmission line, and beads that are a tight fit over it. This will maximize each bead's impedance. Many short beads are better than a few long ones with the same overall length, because the gaps between the beads reduce the end to end capacitance of the transformer.

The toroid is useful where good LF performance is required. It should have the smallest possible hole through the middle, consistent with the turns being reasonably separated. This will give the highest value of C_1, and thus the best low frequency performance.

The twin-hole core usually gives an inter-turn transmission-line with too low an impedance for use on Guanella transformers.

Core Shape: Ruthroff Transformers

Toroids are usually the best shape for N-filar Ruthroff transformers. As shown in Part 1, the best high frequency performance is to be had by using the shortest possible total wire length; so use the core that gives the shortest winding length for 1 turn (I_w, in mm), and use as few turns as possible. The best low frequency performance is achieved using the greatest

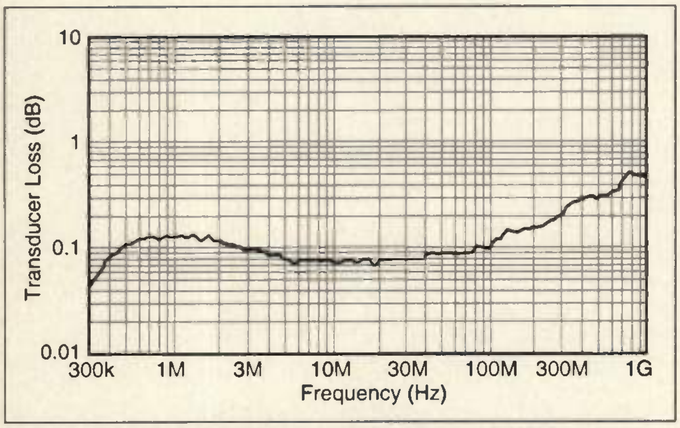

Figure 23. Loss of 50:75 Ω transformer on A8X30 core.

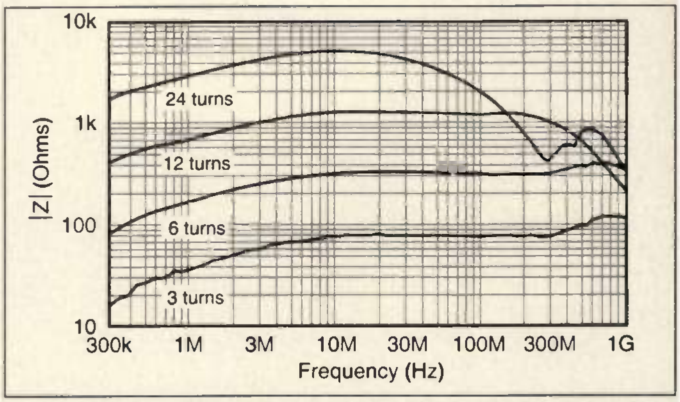

Figure 24. Impedance vs. frequency for choke wound on K37X38 toroid.

number of turns and a core with the lowest core factor. There is thus a trade-off between the winding length, the core factor, and the potential bandwidth of the transformer. This is related to the form factor (F), which is calculated from:

$$F = I_w C_1 \qquad (11)$$

The core with the smallest value of F should be selected.

Core Shape: Auto-Transformers

For auto-transformers, the same considerations as for the Ruthroff transformers apply, with the exception that the high frequency performance is sacrificed for winding convenience. For low power transformers, use twin-hole cores. Large twin-hole cores are hard to come by, so for high power designs, use two tube cores side by side. The tube cores can be made from a stack of suitable torroids.

Windings

The wire used should be as thin as possible. For some microwave directional couplers, coaxial cable the thickness of a human hair is used. On Nickel-Zinc or dust-iron cores, try to leave a gap of at least half a wire width between adjacent turns where the winding passes through the center of the toroid. The windings should extend over 330°, leaving a gap of 30° between the winding ends. These measures all reduce the capacitance. There is a limit to this: if the wires are made too small, the required transmission line characteristic may be difficult to achieve, and the line's loss may be increased due to the decreasing skin depth with frequency.

Core Heating

For low power transformers, the only effect of R_P is on the insertion loss, which is an inconvenience which can usually be removed with a little extra amplification. (The exception to this being low-noise designs, which must have low-loss transformers.) However, R_P also sets a limit to the transformer's power handling capacity due to core heating. To handle high powers R_P must be quite large. For example, take a 400 W 200:50 Ω HF balun. The core dissipation is to be 1 W maximum. The required bifilar choke impedance can be calculated to be 50 × 400/1 = 20 kΩ.

In this article R_P has been calculated as a small-signal value derived from the complex permeability data. At high power there will be additional hysteresis loss, which must be calculated from the manufacturer's data. This reduces the actual value of R_P, however the small-signal value provides a useful practical guide to the expected core heating.

Intermodulation Products

It has been stated that transmission line transformers do not suffer from intermodulation products because the magnetic fluxes of the two windings cancel. This is not true. In general, for each bifilar winding there is a difference between the average voltages at the input and output. This voltage is connected across the bifilar choke, and causes a small out of balance current to flow. This is true even where the transformer is being used solely as a delay element: the voltage difference is provided by the phase shift of the RF.

This magnetization current will exercise the ferrite's hysteresis loop. Unfortunately, the only simple way to

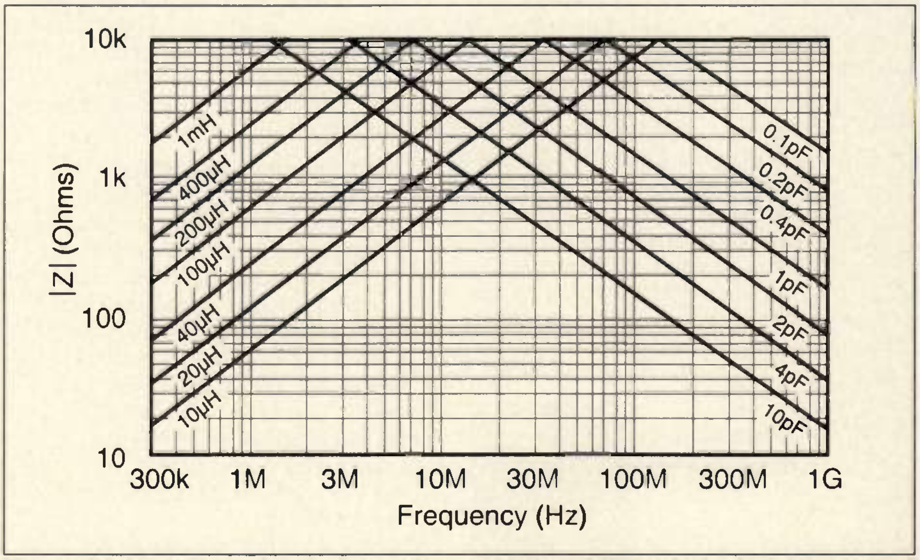

Figure 25. Impedance levels for various values of bifilar choke inductance and capacitance.

predict the resultant intermodulation products is by using CAD calculations. I have not tried this.

The Non-Working Transformer

In Part 1 of this article, I made an unsupported statement regarding Figure 3. I said that the transformer consisting of a few turns on a ferrite ring would not work at RF. The time has come to both qualify and justify this statement.

The transformer in Figure 3 is obviously a pure Faraday type, and is clearly the wrong shape, as defined above. To work efficiently, it must operate on the LF part of the permeability characteristic where the resistive losses are low. However, more turns would be needed to achieve sufficient primary inductance to make the transformer work. The only way that the transformer in Figure 3 could work was if the input and output impedances were small, say less than 1 Ω. I have measured a transformer like the one in Figure 3 in a 50:12.5 Ω circuit. The center of its pass-band was at 5 MHz, where its insertion loss was 2 dB. Its bandwidth was very small.

Conclusion

It has been argued that data for core materials that are given for conventional transformers and inductors are of no use when trying to predict the performance of transmission-line transformers. This is not true. To select a grade of ferrite for a transformer, the manufacturer's complex permeability or impedance spectra should be consulted. These, after mathematical manipulation, can give a good estimate of transformer loss.

At the start of this article, I pointed to the apparent disagreement between Snelling's and Sevick's view of transmission line transformers. I hope that it is now obvious that there is no disagreement because:

a) Following Snelling's advice on 'conventional' transformer design leads to the construction of transmission-line transformers

b) Sevick did not claim the auto-transformer as a transmission-line device.

I hope that this article has encouraged RF designers to build their own transformers. There is much still to be discovered about them. They are also one of the last components that the circuit designer can design for himself [13]. *RF*

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About the Author

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RF Circuits For Communication Applications

Steve Andrezyk and Raphael Matarazzo
Harris Semiconductor

Commercial applications in the 800-MHz to 1-GHz band have been growing at a rapid pace. Products such as cellular/cordless phone, PCS, WLAN and wireless remote equipment highlight the need for flexible, integrated, low-cost, front-end RF devices. Integrated circuits reduce board size, improve reliability and optimize the manufacturing process. The higher level RF IC will be characterized for all critical system specifications (1 dB compression, noise figure, third order intercept, etc.). This article reviews several application circuits using ICs for RF transceiver front-end design, including a low noise amplifier (LNA)/mixer, a Gilbert cell down-converter, an up-converter, and a cascode amplifier.

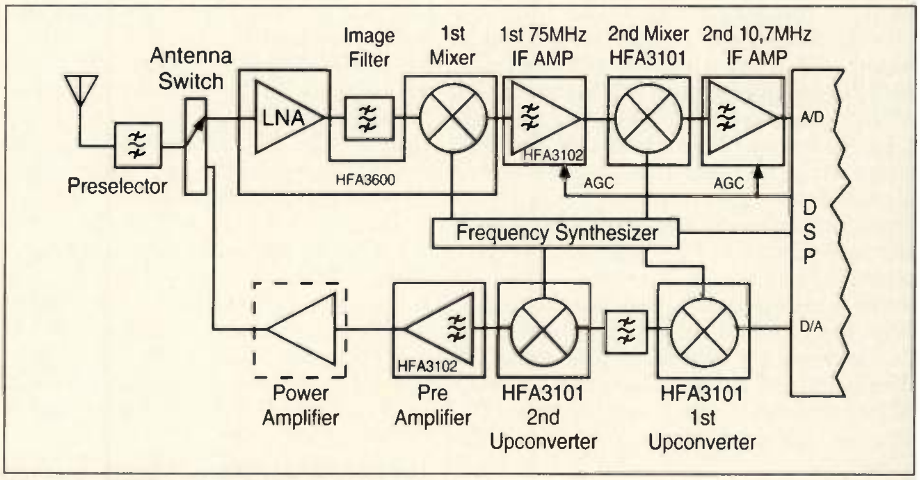

Figure 1. RF front end block diagram.

The ICs used for these designs constitute four of Harris' integrated RF building block components for analog front end transceiver design needs. They are fabricated in a 8-GHz f_T UHF-1 silicon based complementary bipolar SOI process. These devices can be used with input signals up to 1.2 GHz. Figure 1 highlights three of the four devices in a generic dual conversion RF front-end block diagram. The HFA3600 low noise amplifier/mixer offers a low minimum noise figure for sensitive, wide dynamic range receivers, as well as a power down feature. The HFA3102 dual long tail pair is a flexible RF building block used as an amplifier, multiplier, or AGC. The HFA3101 Gilbert cell can be used for up- and down-conversion. The HFA3XXX series of transistor arrays (not shown) are uncommitted high frequency NPNs and PNPs utilized for custom RF designs.

The dielectric isolated (DI) process used to fabricate these devices improves the RF performance over standard junction isolated (JI) processes. A JI NPN transistor is isolated from the substrate by a reverse-biased pn junction which, like a diode, has a leakage current and parasitic capacitance. With the DI process, each transistor is isolated from other transistors by an oxide layer. This allows fast PNPs as well as fast NPN transistors. UHF-1 is a bonded wafer DI process. Two wafers with a thin layer of thermal oxide are chemically bonded together in a controlled process environment. One side of the wafer is used for the active devices while the other serves as the "handle" wafer or substrate. Low doping of the DI handle wafer raises the resistance of this insulating material to reduce mutual capacitance and dielectric losses. Devices with an on-board spiral inductor (fabricated on the HFA3600 using metal 2) have an improved Q or quality factor as well as self-resonant frequency due to the high resistivity of the handle wafer.

LNA/Mixer

This critical first stage element in the receiver chain establishes system noise figure and dynamic range. The mixer down converts the RF input to a useable intermediate frequency (IF). Figure 2 shows the LNA/mixer and the evaluation board components operating with a 900 MHz input. The device has LNA and mixer gain of 19 dB, not including signal loss from the image filter. Current consumption at 5 V is 11.2 mA during normal operation, and 250 μA in the power-down mode. The HFA3600 is also characterized down to 4 volt supply operation for the portable market.

The next two sections review design considerations for the LNA and mixer elements of the HFA3600, which can be used cascaded or independently.

I) LNA — Matching the impedances of the various RF front end functional blocks to minimize reflections and signal loss throughout the system is critical. Characterization of Γ_{opt}, or the source impedance for minimum noise figure, lets engineers design a matching network between the LNA input and the antenna T/R switch stage that provides the best noise performance. The LNA input Γ_{opt} at 900 MHz is close to 50 ohms. This makes it easy to match the input port to a 50-ohm antenna switch or preselector with a simple coupling capacitor, eliminating the need for impedance matching networks at the input for many applica-

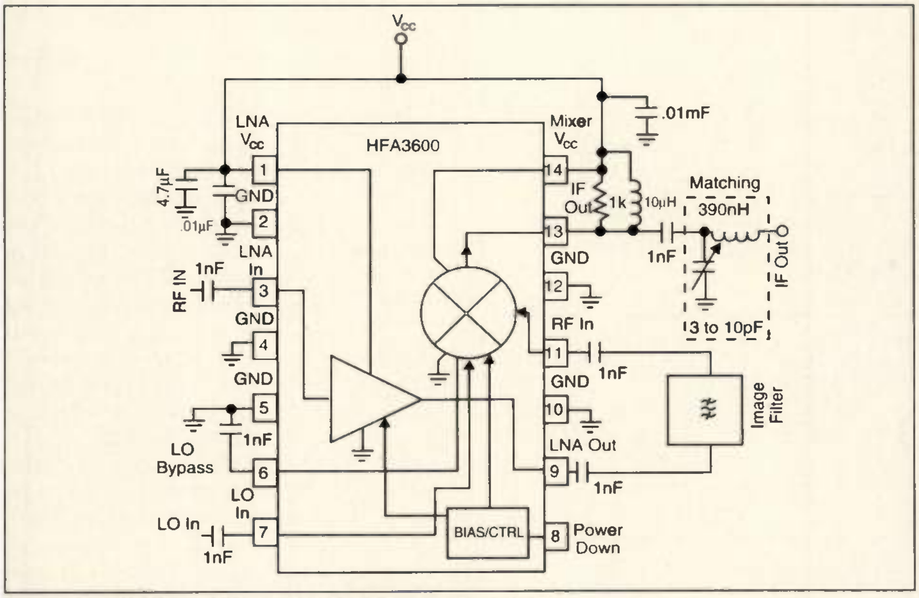

Figure 2. 900 MHz front end LNA/mixer circuit.

tions. Under these conditions, the LNA achieves a 2.4 dB noise figure and 12 dB gain.

Designers whose applications require more demanding noise figure and gain performance can use the noise and gain circle information provided with the device (Figure 3). Noise and gain circles are used to establish a compromise between gain and noise figure at a fixed frequency. The source impedance in front of the LNA determines gain and noise figure per the plotted circles on the Smith chart. For example, a minimum 2.2 dB noise figure and an associated 11.5 dB gain requires the source impedance matching network specified at the intersection of those gain and noise circles. Higher gain can be achieved at the cost of noise figure with the appropriate source impedance network.

Controlling package and bondwire parasitics, along with transistor geometry helps establish good 50-ohm ports in the HFA3600. The input of the LNA and mixer, and the output of the LNA can directly interface with a 50-ohm source and load. This allows them to be directly coupled to common 50-ohm SAW or helical image filters using a single coupling capacitor to each port, as shown in Figure 2.

An image filter rejects incoming RF signals that could generate the same IF frequency (images) after mixing with the local oscillator (LO). Use of a high quality image filter along with a high IF frequency (>75 MHz) reduces the demands on selective filters at the antenna for the limited frequency bands in the cellular spectrum. The HFA3600 has 40 dB isolation between the LNA-out port and the mixer RF-in port. The device is primarily characterized for a 75 MHz IF.

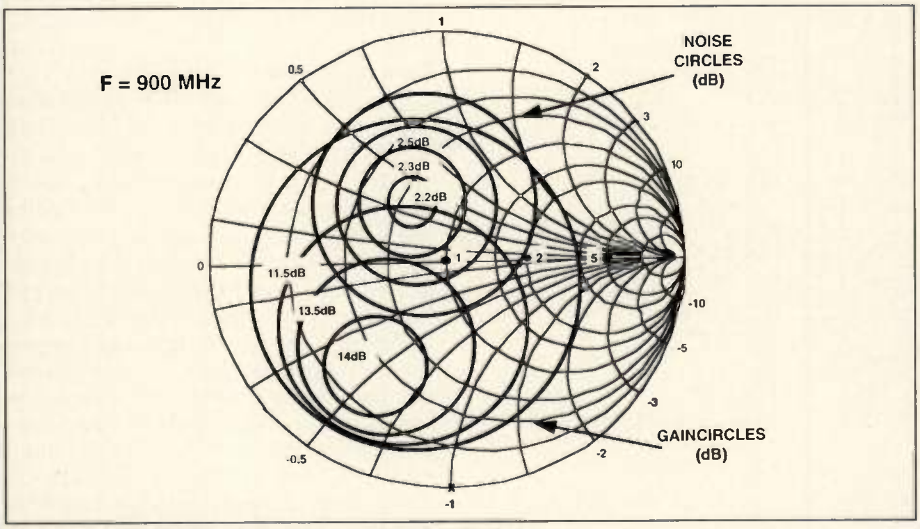

Figure 3. HFA3600 LNA noise and gain circles.

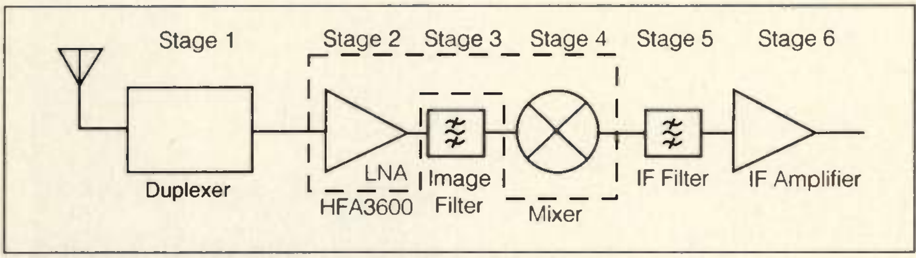

Figure 4a. Typical front end cascade block diagram.

	Duplexer	LNA	Image Filter	Mixer	IF Filter	IF Amplifier
Stage:	1	2	3	4	5	6
Noise Figure (dB)	3.00	2.30	3.00	12.10	8.00	3.00
Gain (dB)	–3.00	12.80	–3.00	7.00	–8.00	20.00
IP3 (dBm)	100.00	12.80	100.00	3.20	100.00	100.00
dNF/dT (dB/ °C)	0.00	0.00	0.00	0.00	0.00	0.00
dG/dT (dB/ °C)	0.00	0.00	0.00	0.00	0.00	0.00
System Temp. (°C)	25.00		Reference Temp. (°C)		25.00	
Input Power (dBm)	–30.00		Noise Bandwidth (MHz)		1.00	
Pout (dBm)	–33.00	–20.20	–23.20	–16.20	–24.20	–4.20
dNF/dNF (stage)	0.30	0.50	0.003	0.50	0.04	0.08
dNF/dG (stage)	–0.70	–0.50	–0.48	–0.06	–0.03	0.00
Cascade Noise Figure (dB)	8.55		Cascade Gain (dB)		25.80	
Noise Temperature (°K)	1785.0		Input IP3 (dBm)		–10.80	
Signal-to-Noise Ratio (dB)	75.5		Output IP3 (dBm)		15.01	
Spur-free Dynamic Range (dB)	63.1		IM3 O/P Level (dBm)		–42.63	
Nominal Detectable Signal (dBm)	–105.5					

Figure 4b. Nominal cascade performance using Hewlett-Packard APPCAD NoiseCalc (V1.02).

II) Mixer — The IF output of the mixer is an open collector, which offers maximum flexibility in IF stage impedance-matching. An RF choke in parallel with the collector resistance improves the DC biasing of the open-collector transistor. The inductor short-circuits the collector resistor at DC, preventing it from changing the quiescent bias point on the load line. Use of the inductor is strongly recommended for high values of collector resistance, to avoid saturation in the mixer current-switching stages.

At high frequencies, when the choke is an open circuit, the output impedance of the mixer is set by the collector resistance. The collector resistor should typically be between 330 ohms and 1 kohms. Greater conversion gain is achieved with a higher resistance. The HFA3600 mixer has 7 dB conversion gain with a 1 kohm resistor.

The upper end of the receiver dynamic range is determined by the mixer's output third-order intercept ($IP3_{OUT}$), a function of the available LO drive level. This mixer has a 4.4 dB $IP3_{OUT}$ for a –1 dBm LO input power. Portable, power-sensitive designs can use a lower power LO drive level with a proportional drop in the $IP3_{OUT}$ (2.2 dB at –6 dBm).

The Harris LNA/mixer can be evaluated in a receiver cascaded arrangement (Figure 4a) using readily available software programs such as Hewlett Packard's APPCAD NoiseCalc. The example in Figure 4b shows a typical screen from this software package with six stages of the receiver front end. The user enters typical values for noise figure, gain and $IP3_{OUT}$ for each stage from the duplexer to the first IF stage. (Between-stage matching is assumed to have been kept within acceptable limits.)

System performance is calculated and presented on the bottom half of the screen as in Figure 4b. The nominal detectable signal for this cascaded receiver arrangement is –105.5 dB, with an 8.55 dB cascade noise figure. The system $IP3_{OUT}$ is 15.01 dB. Designers may modify the electrical characteristics of an individual stage to obtain the desired receiver performances. For example, if an active filter with 0 dB gain, 9 dB noise figure and 7 dB $IP3_{OUT}$ is used to replace the passive filter given in Figure 4b, the cascaded noise figure is improved from 8.55 to 7.33 dB with a degradation of the input third order intercept point from –10.8 to –14.8 dBm.

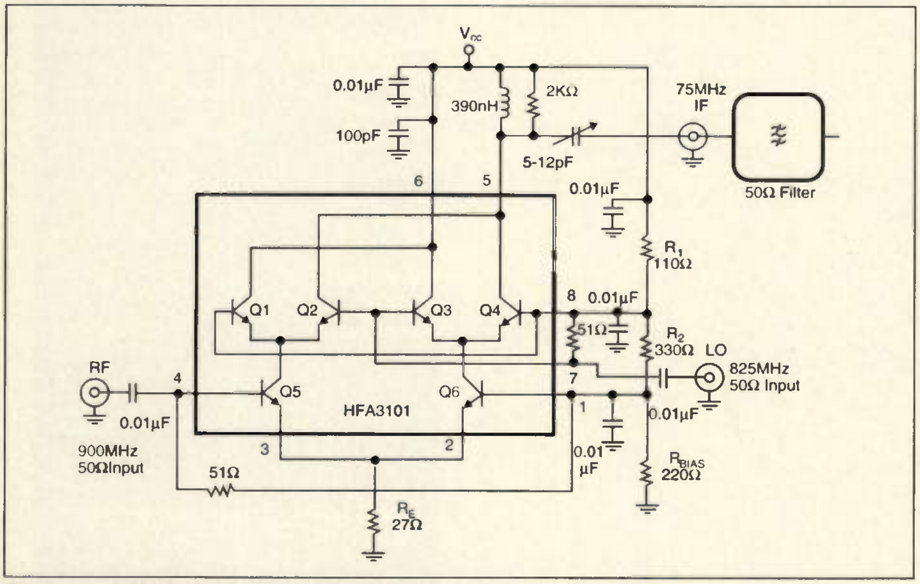

Figure 5. HFA3101 down converter circuit implementation.

Down-Converter

The HFA3101 all-NPN transistor array is configured as a Gilbert cell, which enables four-quadrant multiplication. It can be used for direct down-conversion to a much lower IF. Unlike the LNA/mixer, this topology is a double-balanced architecture, which generally improves feed-through performance over the single-balanced type.

Figure 5 shows the internal transistor configuration, as well as the external circuitry required for a down-converter or mixer application. Test conditions included a 900 MHz RF input and 825 MHz LO, for a 75 MHz IF. The circuit was tested on a carefully laid out evaluation board of 0.031 inch

G-10 material. Quality chip resistors and capacitors and very short board trace lines minimized reflections.

Bias levels for the transistors in the Gilbert cell are established using the resistive network R_1, R_2 and R_{bias}. The voltage drop across R_2 must be large enough to ensure that the base-to-collector junction of the lower pair is reverse biased. 51 ohm resistors bias the bases of the upper and lower differential pairs, and provide terminations for the LO and RF signals.

The emitter resistor R_e sets the bias current through the cell, which impacts available voltage overhead for the differential pairs and affects the AC input impedance. The emitter current is equal to :

$$I_e = \left[\frac{V_{cc} \cdot R_{bias}}{R_1 + R_2} - V_{be}\right] \cdot \frac{1}{R_e}$$

The 27 ohm emitter resistor is optimum for biasing the cell current for an expected input signal level less than 26 mV and a 3- to 4-V bias supply. An output coil aids both voltage overhead and conversion gain.

The 50 ohm RF and LO inputs are capacitively coupled to the Gilbert cell IC input ports. Good quality decoupling capacitors on the resistive bias network become AC terminations at RF frequencies.

Test data was recorded for the circuit biased at 3 V and 8 mA and –6 dBm LO input power. Results were 8.5 dBm power gain, 11.5 dBm $IP3_{OUT}$, and 14.5 dB SSB noise figure.

Down-converter spectrum results are shown in Figure 6. The IF-band spectrum shown on the left achieves an 8.5 dBm conversion gain at 75 MHz for the –26 dBm RF input. This conversion gain allows direct connection to a 50 ohm antenna, along with a high gain IF strip for certain low-cost receiver applications. The out-of-band spectrum shows feed-through at the IF port. This can be improved by adding a low-pass filter.

Up-Converter

Similarly, the Gilbert cell IC can be used as an up-converter in the transmitter portion of a transceiver. The same evaluation board layout for the down-converter can be used with some modifications for matching components. The up-converter circuit in Figure 7 was tested with a 76 MHz modulating RF input and an 825 MHz LO

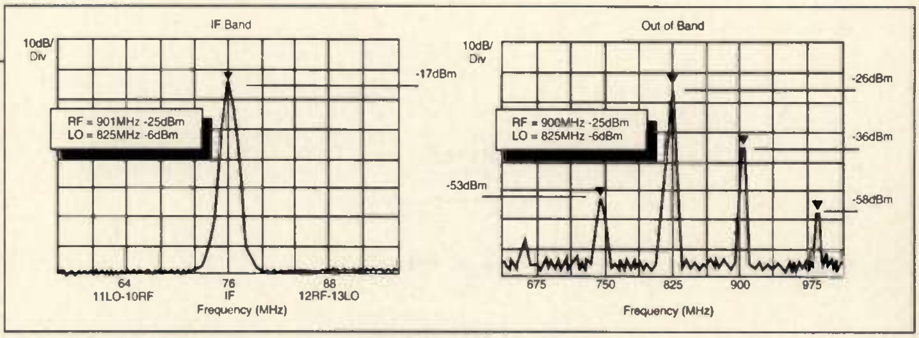

Figure 6. HFA3101 down converter frequency spectrum.

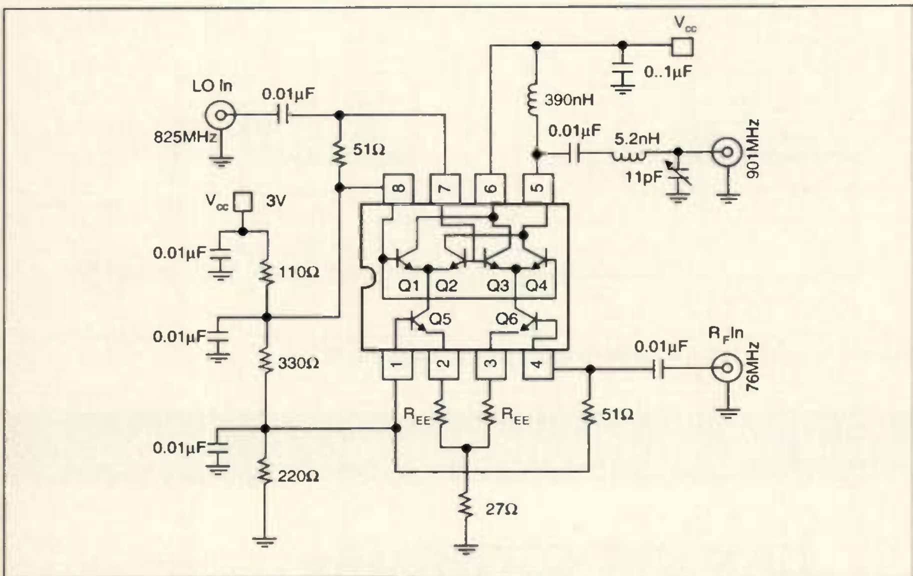

Figure 7. HFA3101 up converter circuit.

or carrier frequency. The desired up-converted RF output for the transmitter is 901 MHz.

Obviously, signal levels in the transmitter path are higher than incoming signals in the receiver path. If the modulating RF input signal in the transmitter exceeds 26 mV (peak), the linear range of operation for the lower differential pair amplifier must be extended. An increase in linearity for higher level signals can be obtained by using emitter degeneration resistors (R_{ee}) off the emitters of the lower transistor pair, Q5 and Q6. This increases the linear range of operation for the lower differential pair by a factor of approximately I_eR_{ee}. The linear range improvement is obtained at the cost of a proportional reduction in the voltage gain by the same factor. Figure 8 illustrates the overall improvement in dynamic range and linearity when using 4.7 ohm emitter degeneration resistors. Adjacent intermodulation distortion terms were reduced by 22 dB. The tradeoff for this improvement in dynamic range is a 2.5 dB gain reduction.

Up-converter circuit performance data was taken with a 4 V supply and 18 mA bias current. Power gain was 7.2 dB for a 0 dBm LO power level. With a matched output network, measured LO isolation and RF isolation were 28 dBc and 22 dBc, respectively.

Cascode Amplifier

The HFA3102 dual long tail pair consists of two NPN emitter-coupled pairs with an NPN-transistor current source. A two-transistor cascode amplifier offers a higher output bandwidth and larger collector voltage capability than the traditional common emitter amplifier. In Figure 9, two of the transistors in the dual long tail pair are configured as a cascode amplifier. The cascode amplifier consists of a common-emitter (the tail transistor) input stage feeding a high output impedance common-base stage. The collector of the common-emitter (CE) stage is heavily loaded by the low input impedance emitter of the common base (CB) stage, reducing any feedback voltage. This configuration prevents high frequency feedback through the CE collector capacitance back to the input.

The test circuit was set for a 10 mA output current with a transistor h_{FE} of 40. As in the Gilbert cell, a resistive divider provides proper DC bias to the cascode transistors. It also sets the base voltage of Q2 which sets V_{CE2}. A few mA is sufficient to set the divider and provide the necessary base current. The presence of two additional parts, resistor R_b and a shunt capacitor across R_3 of the divider, offers a low-noise alternative to the traditional cascode resistive divider biasing approach.

The AC signal is capacitively coupled to the base of the common emitter. Shunt capacitors at the divider nodes provide an AC short circuit which removes noise generated from the divider resistors (R_1, R_2, R_3). The

only contributor of thermal noise in the bias circuit is R_B. In this configuration, however, the thermal noise is attenuated by R_s (source impedance of the AC input) divided by R_B. In this case, making R_B large actually lowers the thermal noise transferred to the amplifier. Some test results will illustrate this effect.

At the output stage, an RF choke in parallel with the collector resistance provides part of the output matching network. It also affects the DC biasing of the amplifier in that it changes the classic DC load line slope, making it vertical. The new quiescent point or Q-point is moved to the right of the I_C V_{CE} load line, where the base current response is spaced further apart. This Q-point is also set at V_{CC}, allowing the AC peak voltage to exceed the rail voltage. This is possible because the inductor in parallel with the load resistor stores energy (V = L di/dt).

This cascode amplifier circuit achieves 15.1 dB gain and a 3.68 dB noise figure at 800 MHz. Changing the bias resistor R_b to 50 ohms improves the input return loss, but raises the noise figure to 5.35 dB. This illustrates the need to keep R_b high to reduce thermal noise. Figure 10 shows the two-tone (799 MHz and 801 MHz) intermodulation distortion of the circuit. The measured output third order intercept was 15.8 dBm, with a 5.5 dBm compression point. S-parameters for the tuned cascode, recorded between 700 MHz and 1 GHz are shown in Figure 11.

Conclusion

The three devices discussed, along with transistor arrays, facilitate an

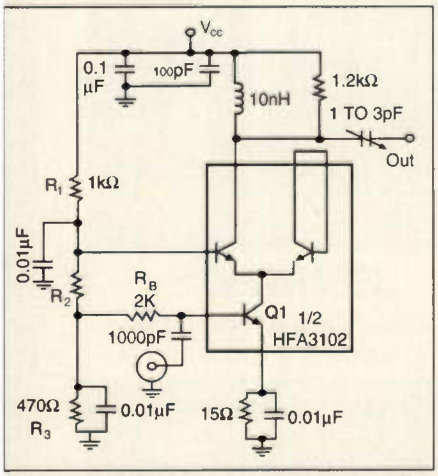

Figure 9. HFA3102 cascode amplifier circuit implementation.

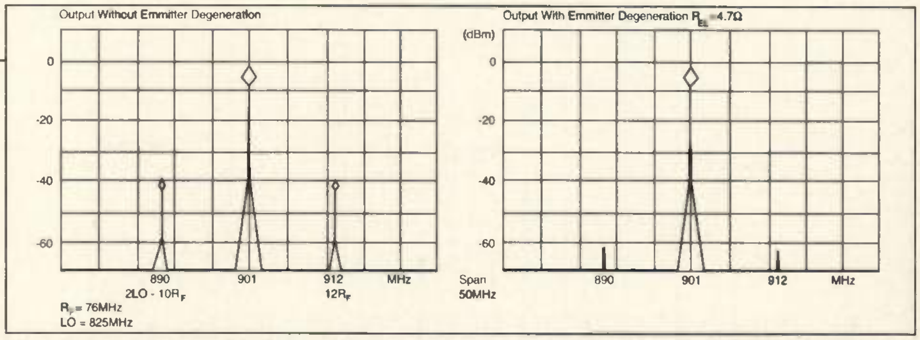

Figure 8. Up converter frequency spectrum with and without emitter resistors.

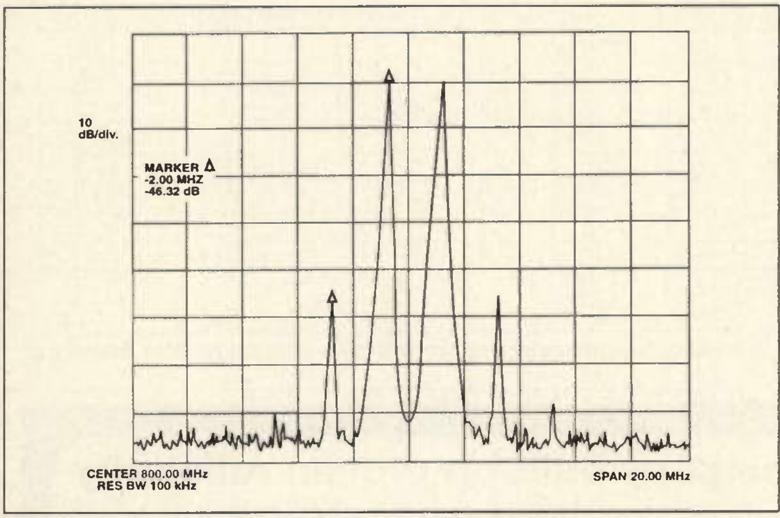

Figure 10. Two-tone (790 and 801 MHz) IMD distortion measurement. IMD products are at 797 and 803 MHz.

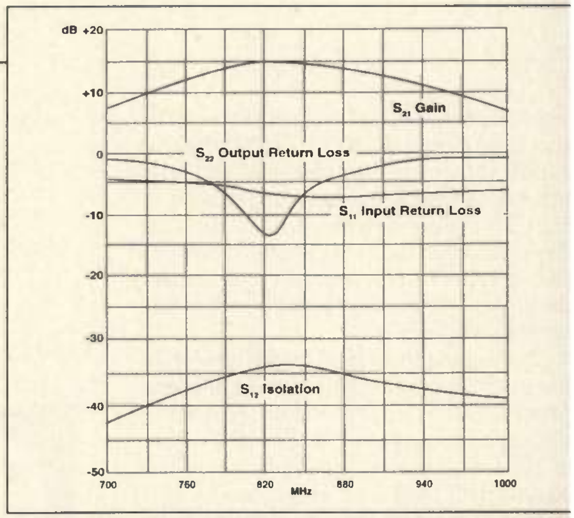

Figure 11. S parameters for the tuned 820 MHz cascode amplifier circuit.

integrated design approach to front-end transceiver functions in the 800 MHz to 1 GHz spectrum. In addition to the characterization data described above for the LNA/mixer, Gilbert cell and Long-Tail Pair ICs, Spice models for the transistor arrays enable designers to analyze circuit performance before breadboarding. *RF*

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Using RF Channel Sounding Measurements to Determine Delay Spread and Path Loss

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Time dispersion in RF channels, induced by multipath caused by reflectors and scatterers in the propagation environment, must be considered when designing high-rate digital radio systems and wireless modems. This article describes the parameters used to characterize time dispersion and path loss in an RF channel from measured data. Following a discussion of the time dispersion parameters, the analysis and results of measurement data taken in a large train yard in the 902 - 928 MHz ISM band are presented.

A previous article on propagation measurements described a system used by Virginia Tech researchers to sound RF channels to determine the propagation delays and attenuation of the channel. The measurement results obtained with the system are used by research laboratories to validate propagation prediction models, and by engineers who are designing new wireless applications and systems. Because of the importance of characterizing a channel to determine the reliability of a wireless system design, this article presents a brief tutorial on how to characterize time dispersion of a radio channel. A description of a recent measurement program, conducted in an active train yard in the 902 - 928 MHz ISM band, is then presented along with an explanation of the data analysis and some results derived from the measurement data.

Measured Propagation Delays in an RF Channel

The system used to measure propagation delays for this research was a spread spectrum sliding correlator system, a channel sounding technique used by many researchers to measure time delay spread [1,2,3]. This measurement system records snapshots, called power delay profiles, of received power versus time. (These snapshots actually show the correlation of two rectangular pulses convolved with the impulse response of the channel.) Figure 1 illustrates a power delay profile recorded during the train yard measurement campaign.

Figure 1 shows the power at the receiver relative to the peak power of the first arriving signal. The time axis shows absolute propagation delay, where t=0 on the axis corresponds to zero propagation delay between the transmitter and receiver. The peaks

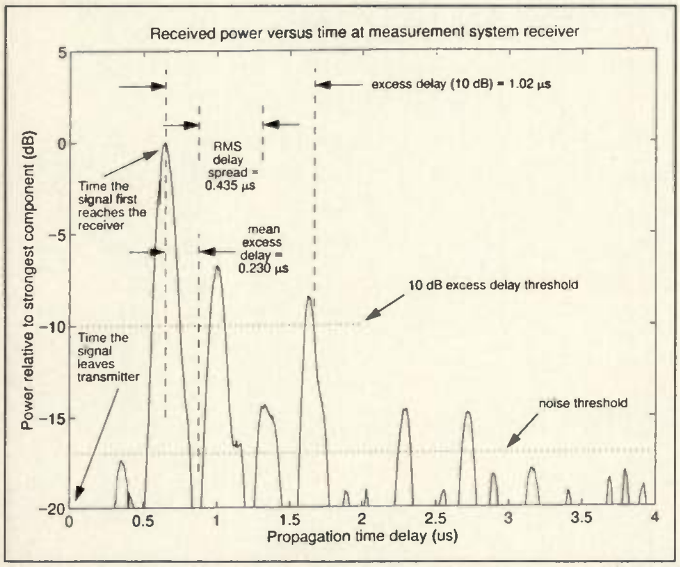

Figure 1. A power delay profile showing the delays induced by an RF channel.

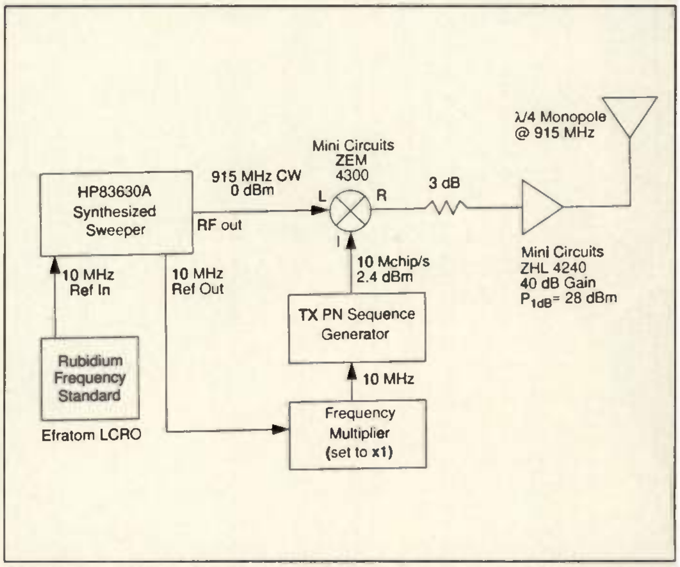

Figure 2. The 915 MHz wideband propagation measurement system transmitter used for the train yard research.

within approximately 17 dB of the peak of the strongest signal represent multipath delay caused by the RF channel. Peaks below the 17 dB noise threshold are caused by thermal noise and are not multipath components. This noise threshold is a function of the specific measurement system and is related to the noise figure, detector gain, and dynamic range.

The power delay profile in Figure 1 shows six peaks above the noise threshold, the first and strongest of which is the line-of-sight (LOS) signal. The LOS signal has a propagation delay of 645 ns, implying that the transmitter and receiver (T-R) separation was approximately 645 feet, since RF energy travels at roughly 1 foot per nanosecond in air. The five peaks following the LOS signal are delayed copies of the received LOS signal caused by excess distance travelled due to at least five multiple propagation paths. These delayed peaks, although several dB below the line-of-sight, can cause intersymbol interference (ISI) and affect the reliability of wireless systems.

Researchers use parameters which describe the delay exhibited by an RF channel. These parameters are excess delay (X dB), mean excess delay (τ) and RMS delay spread (σ_τ) [4]. Because the actual shape of power delay profiles has little effect on wireless system performance [5], these parameters may be used to characterize channels for system design and determine reliable transmission rates.

Characterizing the RMS delay spread of an RF channel helps determine if the radio channel is a flat-fading or frequency-selective fading channel. Flat fading means that all frequency components transmitted through the channel within the channel bandwidth will experience the same degree of fading. Frequency-selective fading, however, means some portions of the band within the channel bandwidth will experience more severe fading than other parts of the channel band. If the channel is not flat, then an equalizer may be required at the receiver to achieve the desired system performance for a given data rate. A rule of thumb is that a channel can be considered flat when σ_τ/T is less than 0.1 (where σ_τ is the RMS delay spread and T is the symbol period).

For example, United States Digital Cellular (USDC) forward and reverse channel data rate is 24,300 symbols per second (48.6 kbps). If a USDC system can tolerate a maximum σ_τ/T value of 0.1, then the maximum tolerable RMS delay spread σ_τ is 4.12 µs. If the RMS delay exceeds 4.12 µs in an environment, then equalization is required to achieve satisfactory performance. This is why some cellular phone manufacturers use equalizers in their design, particularly on the West-coast where many mountains surround the cities.

Excess delay (X dB) is the time delay between the first arriving component (peak) and the time past which no other components exist within a given level of X dB relative to the strongest component. For example, the power delay profile in Figure 1 illustrates excess delay for a 10 dB threshold. The excess delay is the time between the first component and the fourth component, beyond which no peaks exist within 10 dB of the strongest component. In this case, the first component is the strongest component, but this is not necessarily the case, especially when there is an obstructed LOS path between the transmitter and receiver.

Mean excess delay $\langle\tau\rangle$ is defined to be the first moment of the power delay profile, given by

$$\langle\tau\rangle = \frac{\sum_k P_k \tau_k}{\sum_k P_k} \qquad (1)$$

where P_k is the absolute instantaneous power at data point k on the power delay profile, and τ_k is the time between the first detectable signal at the receiver and data point k. Figure 1 illustrates the mean excess delay on the power delay profile. For this profile, the mean excess delay is 235 ns.

RMS delay spread is the square root of the second central moment of the power delay profile, defined to be

$$\sigma_\tau = \sqrt{\langle\tau^2\rangle - \langle\tau\rangle^2} \qquad (2)$$

where

$$\langle\tau^2\rangle = \frac{\sum_k P_k \tau_k^2}{\sum_k P_k} \qquad (3)$$

where P_k is the absolute instantaneous power at data point k on the power delay profile, and t_k is the time between the first detectable signal at the receiver and data point k. RMS delay spread is actually independent of the time reference point; therefore, when calculating $\langle\tau\rangle$ and $\langle\tau^2\rangle$ for equation 2, the delay of each data point can be referenced to time zero if so desired.

The Train Yard Measurement Campaign

Propagation research conducted in a train yard provided data for an environment which had not been characterized by delay spread parameters. Because of the number of good reflectors and obstructions (large, metal rail cars), and the geometry of the yard (long, parallel, closely spaced trains), train yard propagation characteristics could not be confidently extracted from other propagation studies. The train yard research results may be used to develop applications intended to track individual rail cars and for track-to-terminal communications.

Parameter Name	Symbol	Value	Units
Transmitter chip rate	α	10.00	Mchip/s
Receiver chip rate	β	9.990	Mchip/s
PN sequence length	M	2047	chips
Slip rate	R_{slip}	10.00	kHz
Slide factor	γ	1000	unitless
Spectrum analyzer ResBW	ResBW	10 or 30	kHz

Table 1. Measurement system parameters for the train yard measurements

Location	Max σ_τ (ns)	Avg σ_τ (ns)	StDev σ_τ (ns)	n	σ (dB)
Site 1	280.7	94.6	58.3	-	-
Site 2	731.5	331.0	145.0	2.69	3.43
Site 3	1161.2	565.8	164.3	3.41	2.49
Sites 2 & 3	1161.2	454.3	194.3	3.08	8.78
All Sites	1161.2	412.9	216.8	-	-

Table 2. Delay spread and path loss results of train yard measurements.

The train yard chosen for the propagation study was the Norfolk Southern rail yard in Roanoke, VA, because of its size and activity. The yard extends several miles through the city of Roanoke, having a span of 54 tracks wide across the classification yard. With over 100 trains passing through daily, the environment was typical for large train yards, and RF propagation measurements may be extended to other large rail yards across the country.

The propagation measurements focused on three sites within the train yard. The first site was one used to observe the effect of receiver antenna location on delay spread for three types of rail cars. The average T-R separation at site one was approximately 17 meters. The second site was used to investigate propagation paths parallel to the tracks and the effect of changing yard conditions on delay spread, and the average T-R separation was approximately 189 meters. The third site was used to characterize the effect of changing yard conditions on delay spread for propagation paths across the entire yard, perpendicular to the tracks. The average T-R separation at site three was approximately 228 meters.

The Measurement System and Measurement Parameters

The sliding correlator measurement system used for the measurements is shown in Figures 2 and 3. This wideband system, explained in detail in a previous article [6], was configured with system parameters shown in Table 1. The transmitter chip rate α of 10 MHz was chosen to sound the 902 - 928 MHz ISM band using a carrier frequency of 915 MHz. Using this chip rate, the time resolution obtained was approximately 100 ns, meaning that multipath components having a separation in time of more than 100 ns could be identified as separate components. The receiver chip rate β was 9.99 MHz,

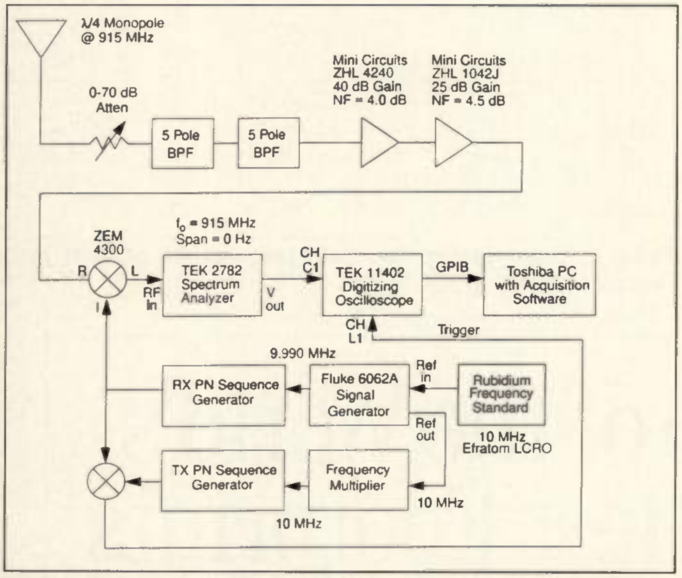

Figure 3. The 915 MHz wideband propagation measurement system receiver used for the train yard research.

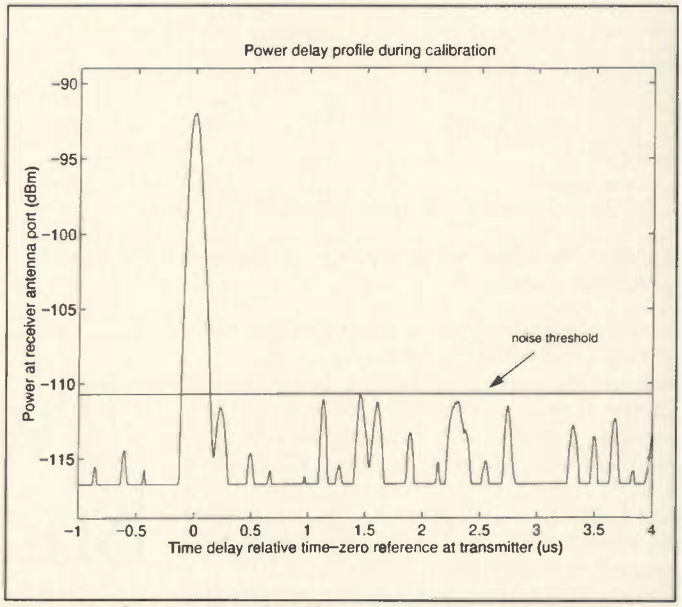

Figure 4. A sample calibration power delay profile.

making the slip rate R_{slip}, or the difference between the transmitter and receiver chip rates, equal to 10 kHz. The resolution bandwidth (ResBW) of the spectrum analyzer was varied between 10 kHz and 30 kHz for all measurement sites. A chip rate of 10 MHz and a slip rate of 10 kHz yielded a slide factor of 1000, meaning time delays observed on the oscilloscope were 1000 times greater than the actual propagation time delays. The profile in Figure 1 has already been corrected for the slide factor, and shows actual propagation time in air.

With the measurement system in this configuration, 396 power delay profiles were recorded at the three sites. Each power delay profile was recorded in a separate data file for post-processing.

Calculating Channel Parameters from Measurement Data

Since the recorded oscilloscope waveforms contain voltage versus time data, a relationship must be made between scope voltage and received power to reconstruct the power delay profile. This is done with calibration data taken before and after a measurement run. Calibration data consists of power delay profiles recorded when the system transmitter and receiver antenna ports are connected back-to-back with a cable and attenuators. Since the cable used for calibration is short compared to the spatial resolution of the system, the recorded profiles show a single pulse at time zero (see Figure 4). During calibration, the power into the receiver antenna port is known, so a relationship can be established between the power at the receiver antenna port and the voltage peak observed on the oscilloscope. When operating in the linear range of the amplifiers and mixers used in the system, the received power (in dBm or dBW) at each data point of a power delay profile can be mapped to an absolute power level. For the measurement system used at the train yard, the ratio of a change in received power to the change in scope voltage was 59.1 dB/volt.

Because recorded power delay profiles have a noise floor caused by either thermal noise or PN (pseudo noise) leaked through the mixer, a threshold must be applied to keep noise peaks from being interpreted as multipath components during data analysis. This may be done by setting the threshold to the peak of the noise where a multipath component is known not to exist, such as in a range on the time axis far ahead of the LOS component or just behind the LOS component on the profile (which is actually far ahead in delay time on the profile because of the cyclical nature of the sliding correlator). The threshold shown in Figure 1 is 17 dB below the LOS peak and derived from noise peaks later in the profile. All data points below this threshold are assumed to be noise and are not included in the calculation of RMS delay spread or received power.

RMS Delay Spread from Measured Data

The RMS delay spread for each power delay profile is calculated from the displayed data points by using equations (1), (2), and (3). RMS delay spread results for the train yard measurements are summarized in Table 2. The last two rows in the table summarize results for a combination of the sites. Cumulative distribution functions (CDFs) shown in Figures 5, 6 and 7 also summarize the RMS delay spread results. These figures show the probability that the RMS delay spread will be less than a particular σ_τ.

Received Power from Measured Data

The power delay profiles recorded by the measurement system were also used to determine received power and path loss between the transmitter and receiver. Calculating the received power uses the same noise threshold technique used for calculating RMS delay spread, and points below the established threshold are not included in the calculations.

The average received power for a given profile is proportional to the area under the curve. Because a single signal component in the power delay profile has a finite width, and is not an ideal infinitely narrow pulse, the pulse width must be normalized out of the

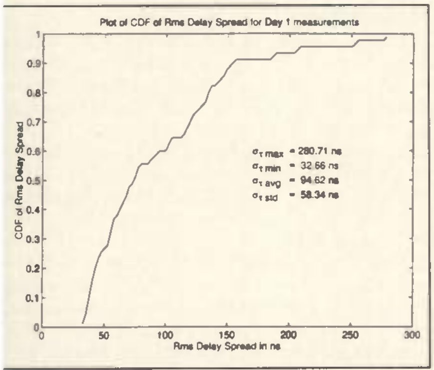

Figure 5. Cumulative distribution function for RMS delay spread measured at site one.

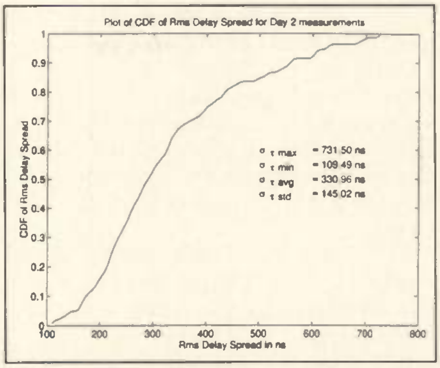

Figure 6. Cumulative distribution function for RMS delay spread measured at site two.

power delay profile to determine the true received power. As shown in [7], local averaged received power is not a function of bandwidth.

To determine power from the power delay profile, define the parameter A_N as the area under a power delay profile. With a periodic sounding signal, the area represents the received power. If A_{Nrec} is the area under the power delay profile received during a channel sounding, and A_{Ncal} is the area under the curve of a calibration power delay profile (Figure 4), then the power received during the channel sounding is

$$P_{rec}[\mathrm{dBm}] = P_{cal}[\mathrm{dBm}] + 10\log\left(\frac{A_{Nrec}}{A_{Ncal}}\right) \tag{4}$$

where P_{cal} is the known power at the receiver antenna port during back-to-back or close range calibration. If $P_{abs}(k)$ is the power at data point k expressed in watts and Δt is the time between two data points, then the area under a power delay profile is given by

$$A_N = \sum_k P_{abs}(k)\cdot\Delta t \tag{5}$$

where k includes all data points above the noise threshold.

Now that the received power has been calculated, the path loss can be calculated by making some simplifying assumptions. Using the Friis free space transmission formula in logarithmic form, received power in dBm at the antenna port is given as

$$P_r[\mathrm{dBm}] = P_t + U_t(\theta) + U_r(\theta) - L_P \tag{6}$$

where P_t is the transmit power in dBm at the antenna port, $U_t(\theta)$ and $U_r(\theta)$ are the gain functions of the transmitter and receiver antennas in dB, respectively, and L_P is the path loss between isotropic antennas in dB, given by

$$L_P[\mathrm{dB}] = 10\log\left(\left(\frac{\lambda}{4\pi d}\right)^2\right) \tag{7}$$

in free space. In real propagation environments where L_P is measured, equation 6 is still valid.

The antenna gain functions of the monopole antennas used for this research can be expressed as

$$U(\theta) = G - D(\theta) \tag{8}$$

where G is the gain of the antenna and maximum value of $U(\theta)$, and $D(\theta)$ is the discrimination function of the antenna in the θ direction (with the monopole oriented on the z-axis of the conventional spherical coordinate system). Combining (6) and (8) yields

$$P_r[\mathrm{dBm}] = P_t + G_t + G_r - (L_P + D_t(\theta) + D_r(\theta)) \tag{9}$$

Propagation measurements in [8] simplify this equation by defining the isolation (L_I) variable to include the effect of both path loss and the discrimination function of the transmitter and receiver antennas, where

$$L_I[\mathrm{dB}] = L_P + D_t(\theta) + D_r(\theta) \tag{9}$$

Therefore, the equation for received power can be written as

$$P_r[\mathrm{dBm}] = P_t + G_t + G_r - L_I \tag{10}$$

For realistic antennas, isolation L_I equals path loss L_P only when anten-

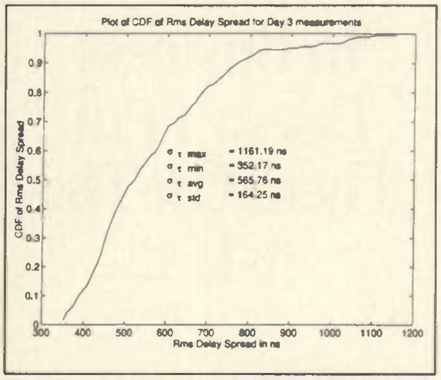

Figure 7. Cumulative distribution function for RMS delay spread measured at site three.

nas are aligned in the direction of maximum gain. When the discrimination functions of the antennas are not known or θ cannot be accurately determined, only isolation L_I can be determined from measurements, but path loss can be computed with assumptions.

To obtain an estimate of path loss statistics from the train yard data, unity antenna gains were used along with unity discrimination functions during data post-processing. These assumptions yielded reasonable results when compared to statistics derived from past research in different environments, as will be shown.

A common path loss model is the d^n model (log-distance path loss model) which gives mean (ensemble average) path loss as a function of T-R separation d as:

$$\langle L_P(d)\rangle = \langle L_P(d_0)\rangle + 10n\log\left(\frac{d}{d_0}\right) \quad (11)$$

where d_0 is a close-in reference distance and $L_P(d_0)$ is the path loss in dB at the reference distance. The path loss exponent n is usually used to characterize path loss in different environments; the free-space path loss exponent is 2. Using a reference distance of 1 meter and a free-space path loss assumption at the reference distance, the reference path loss $\langle L_P(d_0)\rangle$ is 31.7 dB at 900 MHz.

Best-fit path loss exponent estimates were calculated from site two and site three data using a 1 meter reference distance. Table 2 summarizes the results. Path loss exponents were found to be 2.7 for site two and 3.4 for site three. These exponents fall into a reasonable range when compared to propagation research in other environments [4], and they suggest that mean path loss characteristics in the train yard are similar to those found in urban areas, which have path loss exponents in the range of 2.7 to 3.5 [4, 9]. Table 2 also shows the path loss exponent determined for site two and site three data combined.

The log-normal shadowing model includes the effect of shadowing in the d^n model by including a random variable, or

$$\langle L_P(d)\rangle = \quad (12)$$
$$\langle L_P(d_0)\rangle + 10n\log\left(\frac{d}{d_0}\right) + X_\sigma$$

where X_σ is a zero-mean log-normally distributed random variable with standard deviation σ expressed in dB. The value of σ was computed for site two and site three data, and results are shown in Table 2.

Conclusion

Advances in RF technology coupled with the demand for wireless developments have made it necessary for RF engineers to broaden their knowledge in radio propagation. Hardware designers are finding it necessary to have a base of knowledge in propagation to keep current with technology and research. This article was intended to discuss RF propagation measurements and how measured data is used to compute fundamental channel characteristics.

RF propagation measurements provide data from which time dispersion effects and path loss can be calculated. Time dispersion can be characterized by excess delay (X dB), mean excess delay, and RMS delay spread. The path loss exponent n and standard deviation σ can characterize path loss with a simple statistical model in a log-normal shadowing environment.

RF measurements were made in a large train yard to study its unique propagation characteristics. The largest RMS delay spread was 1161 ns and occurred during cross track measurements. Parallel-track measurements showed a maximum RMS delay spread of 732 ns, and close-in measurements at site one resulted in a maximum RMS delay spread of 281 ns. Path loss exponents were 2.7 for the parallel-track measurements and 3.4 for the cross-track measurements. *RF*

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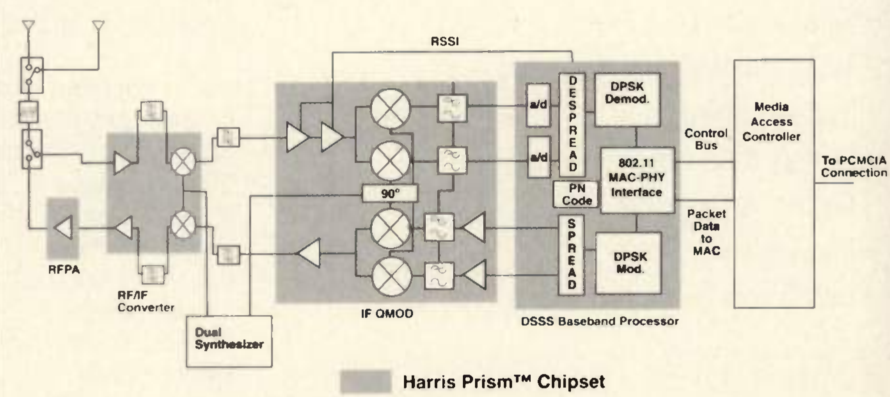
Spread Spectrum ICs Aimed at ISM Band Applications

This month's Product Forum features integrated circuits for spread spectrum applications. As markets grow, and as demand increases for high performance combined with economy, highly-integrated devices are being developed to simplify the design of wireless communications products using spread spectrum transmission.

To help you gain insight into current and future trends in spread spectrum, the following companies offer their comments on the markets and technologies for these products, and the outlook for future requirements.



Prism (™) consists of 1.) 2.4 Ghz RF/IF Converter, 2.) 2.46GHz RFPA, 3.) IF/QMOD, 4.) DSSS Baseband Processor

RF Design **readers were introduced to the Prism^{TM} chip set from Harris Corporation in the Cover Story feature of the October 1995 issue. This set of integrated circuits is designed to support development of a baseband-though-RF transceiver that meets the standards of the soon-to-be-finalized IEEE 802.11 standard for 2.4 GHz frequency-hopping spread spectrum data transmission.**

Stanford Telecom Sees Niches Growing into Markets

As an early innovator in spread spectrum technology, Stanford Telecom has found that this technology can be successfully applied to a wide range of wireless communication challenges. In many instances, as the market matures, what started out as technology niches are starting to look like real markets in their own right. As an example, mobile data transmission applications — such as remote meter reading and vehicle positioning — are turning into very significant businesses which rely on spread spectrum technology for high performance at affordable prices.

Stanford Telecom recently patented an orthogonal code-division-multiple-access (OCDMA) architecture that greatly reduces the effect of signal multipath interference, and enables greater utilization of bandwidth. We see OCDMA technology as perfect for cordless telephones and wireless public-branch-exchange (WPBX) key systems. We support our customers with both ASIC and board-level products, to enable commercial OEMs to build products utilizing this exciting technology.

Loral Communication Systems Goes for Higher Performance

Loral Communication Systems produces high performance spread spectrum integrated circuits, board-level spread spectrum modems and proprietary products utilizing spread spectrum technology.

We expect the market for spread spectrum technology to undergo tremendous growth over the next two years, spurred by growing awareness of the capabilities of wireless systems. Up to this point, the applications for spread spectrum technology consisted primarily of fixed rate systems that were minimally compliant with FCC Part 15 and ISM Band rules.

The emerging applications take advantage of the power and flexibility available in devices such as the Loral PA-100 spread spectrum demodulator ASIC. The pace of the market growth will accelerate as less expensive, highly integrated RF devices emerge for use in the upper ISM bands of 2.4 and 5.7 GHz.

Evolving markets for high performance spread spectrum products include wireless telephony, CDMA satellite channels, T1/E1 and higher rate point-to-point data links, as well and ranging and location systems.

Betheltronix, Inc. Responds to Market Expansion

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the cordless telephone, wireless LAN, and data capture market segments will expand more rapidly as cost reductions in spread spectrum ICs are achieved. This expansion will be realized through increasing consumer public demands for the technology benefits of security, noise reduction, minimal interference, extended range, and immunity to multipath fading.

Betheltronix (BTI) responded to the need for cost-effective spread spectrum ICs by achieving higher levels of integration and low cost silicon designs. Our innovative BT1071 CMOS chip operates from 900 MHz to 5.8 GHz in a direct sequence (DS) or frequency-hopping (FH) mode. BTI continues to advance designs for lower cost and increased performance.

Zilog Offers SS Wireless Development Kit

The Zilog Z2000/Z182 Development Kit offers engineers a means of examining the performance and applications of the Z2000 spread spectrum transceiver and the Z182 controller. These ICs combine to form a complete baseband system ready for upconversion to RF and demodulating down-converted RF signals.

The kit contains the Z0200000ZCO Evaluation Board, development software for command and debug functions, the Zilog EPM^{TM} Electronic Programmer's Manual, and complete product documentation. Also available is the Z0200001ZCO Loopback Evaluation Board.

Zilog also will supply appropriate RF modules for development of a complete spread spectrum communications system. Available devices and boards are made by various vendors. An example is a 902-928 MHz RF section supplied by Utilicom for direct sequence spread spectrum in that ISM band. This unit supports both BPSK and QPSK modulation, 16-256 kbps data rates, and offers programmable output power from +10 to +28 dBm.

American Microsystems Has DSSS Transceiver IC

The S20043 from American Microsystems is a direct sequence spread spectrum (DSSS) transceiver IC. An evaluation kit is available (the S20043ERU) that transmits and receives data messages in the 915 MHz ISM band at data rates of 2.4, 100 and 400 kbits/sec. The S20043 radio is priced at $3000 per unit including license to use the radio design and complete Gerber files and schematics.

Axonn Provides the AX602 Spread Spectrum core IC

A spread spectrum baseband and RF IC is available from Axonn. The AX602 can be part of a system requiring few additional components to transmit data at a 300 kbits/sec in a variety of applications. Potential uses include control systems, keyless entry, industrial process control, image transmission/slow-scan TV, inventory control, access control, and other voice or data applications. The AX602 has a volume price of $2.75. RF

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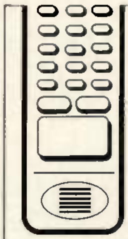

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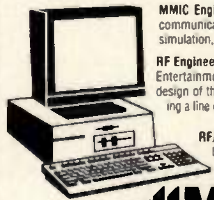

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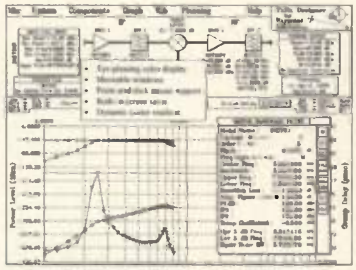

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MBF Microwave, Inc., Rt. 2, Box 252A, Hardy, AR 72542 ....(501) 856-2685

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United Glass to Metal Sealing, Inc, 11A Executive Park Dr., North Billerica, MA 01862... (508)670-6494
Components/Hybrid Packages
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PC BOARDS

Laminates, Polyester Copper Clad
Glasteel Industrial Laminates, P.O. Box 910 Collierville, TN 38027 ....(901) 853-5070

TEST EQUIPMENT

Racal Instruments, 4 Goodyear St., Irvine, CA 92718 ....(800) 722-2528
Instrument Repair Labs, Inc., 2100 W. 6th Ave., Broomfield, CO 80020 ....(800) 345-6140

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RHR Laboratories, 207 Harding Blvd. W., Richmond Hill, Ontario,CN L4C 8x6 ......(905) 884-2392
Tesoft, Inc., 205 Crossing Creek Ct., Roswell, GA 30076 ....(800) 631-1113
Makers of TESLA Com Simulator ....Fax (770) 664-5817 Intl (770) 751-9785

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ESD AND SURGE CONTROL COMPONENTS

Lightning Arrestors

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CABLE AND CONNECTORS

Connectors And Adapters

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Optical Fibers And Connectors

ADC Telecommunications, 4900 W. 78th St., Minneapolis, MN 55435 .... (800) 366-3891 X3000

EMC TEST EQUIPMENT - EMISSIONS

Absorbers

Advanced Electromagnetics, Inc., P.O. Box 711719, Santee, CA. 92072- 1719 .... (619) 449-9492

Microsorb Technologies, Inc., 14A Airport Dr. Hopedale, MA 01747 .... (508)634-1900

Rantec, P.O. Box 1546, Austin, TX 78767 .... (512) 835-4684

Anechoic Chambers

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Rantec, P.O. Box 1546, Austin, TX 78767 .... (512) 835-4684

Antennas

Antenna Research Associates, Inc., 11317 Fredrick Ave., Beltsville, MD 20705 .... (301) 937-8888

EMCO, P.O. Box 1546, Austin, TX. 78767 .... (512) 835-4684

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Current Probes

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Ion Physics Corp., 11 Industrial Way, Atkinson NH 03811 .... (603) 893-6687

Ferrite Absorber Tiles

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Surface & Volume Resistivity Meters

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Conductive Fiber/Fabric

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Ferrite Absorber Tiles

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Gasketing Materials

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Laminates

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Positioning Controls

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Sheilding Foils and Tapes

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3-D Electromagnetic Analysis

Sonnet Software announces the North American Release of Micro-Stripes, from KCC Ltd of Nottingham, U.K. Micro-Stripes may be used to simulate the electromagnetic properties of structures with arbitrary three dimensional shapes. The program uses the transmission-line matrix (TLM) technique to obtain a large number of frequency points from one time-domain simulation. Micro-Stripes version 2.3 starts near $30k and can be obtained from Sonnet Software.
Sonnet Software, Inc.
INFO/CARD #245

Spectrum Management

Softworks has introduced RF-SpecTrac™, a Windows™-based software package for radio frequency spectrum management and direction finding (DF) applications. When combined with appropriate RF measurement hardware, RF-SpecTrac allows you to record general and specific frequency activity using several factors such as time used, length of use, and frequency. The software provides a range of tools to interpret the data. RF-SpecTrac also offers both omnidirectional and direction finding/line of bearing analysis functions.
Softworks Inc.
INFO/CARD #246

Planar/3D EM Simulation

Ansoft announces the release of Maxwell® Strata, a design tool for simulation of RF and microwave simulation of 2D, 2.5D and true 3D traces. Using the space domain method of moments, the program solves for s-, y-, and z- parameters. All radiation effects are considered. Antenna parameters such as gain, directivity, efficiency, input impedance and beamwidth can be calculated. Maxwell Strata is available on all popular UNIX workstations and on high-performance personal computers. Stand alone pricing, including the advanced post-processor, for a perpetual license is $29,900. If bundled with another Ansoft product, the price is $19,900.
Ansoft Corp.
INFO/CARD #247

Circuit Simulation/Layout

Insight$_{RF}$ is a PC-based design system for RFICs and MMICs, as well as etched circuit boards. Insight$_{RF}$ offers easy schematic capture and extensive simulation capability. The program also performs physical circuit layout design and calculates parasitics. The effects os these parasitics can be inserted in the circuit simulation so that circuit layout can be optimized. Finally, Insight$_{RF}$ can write the layout into a GDS2 stream file which can read directly by mask fabrication equipment. Insight$_{RF}$ is available immediately for $2,495 plus shipping and handling.
Intercept Software Inc.
INFO/CARD #248

Spectrum Sharing Software

Comsearch has announced new enhancements to its Spectrum Sharing software. Key features include implementation of carrier-to-interface (C/I) objectives, manual frequency assignment, threshold degradation calculations and threshold-to-interference (T/I). Spectrum Sharing's open architecture allows importation of cell layouts and assigned frequencies from RF design tools, allowing the PCS operator to select their software solutions freely.
Comsearch
INFO/CARD #249

RF literature

Wireless Semiconductor Selection

Motorola has released a comprehensive selector guide for its wireless semiconductor device portfolio. The Wireless Communications Resource Guide (BR3006/D is now available free-of-charge from Motorola's Literature Distribution Center. The guide contains information on devices from multiple Semiconductor Sector product groups.

Motorola Semiconductor
INFO/CARD #200

Updated Trimmer Bulletin

Sprague-Goodman Electronics announces the latest edition of their ceramic dielectric trimmer capacitor bulletin, Engineering Bulletin SG-305E, is now available. The new bulletin provides updated product data for their complete line of ceramic dielectric trimmer capacitors. Included is a new SURFTRIM® capacitor featuring small size and multilayer construction.

Sprague-Goodman Electronics, Inc.
INFO/CARD #199

Filter Catalog

Microlab/FXR announces the publication of a new catalog of high-power coaxial lowpass filters. The models featured are available for use in the 100 to 2,000 MHz region and with average power levels of up to 750 W.

Microlab/FXR
INFO/CARD #198

Amplifier Product Guide

A 25-page product guide from Cougar Components provides detailed specifications for their lines of high-power, TO-8B, high-gain, low-noise, high dynamic range, high efficiency, and low voltage amplifiers. Also covered are limiters, attenuators and AGC amplifiers. The catalog also contains information about Cougar's manufacturing capabilities.

Cougar Components Corp.
INFO/CARD #197

High-Frequency Design Guide

The Institute for Interconnecting and Packaging Electronics Circuits (IPC) offers a book, High-Frequency Design Guide, that addresses packaging and interconnect design for circuits operating in the frequency range of 100 MHz to 30 GHz. Document IPC-D-316 is available to IPC members for $20; $40 for non-members.

The Institute for Interconnecting and Packaging Electronic Circuits
INFO/CARD #196

BER Tester Info

Noise Com has released a four-page black-and-white data sheet on its UFX-BER Series of precision carrier-to-noise generators designed to test the bit error rate of digital transmission systems. Explained in the data sheet are features, specifications, applications, and ordering information.

Noise Com, Inc.
INFO/CARD #195

Note on Gluing Ferrite Cores

Philips Components has introduced a new application note titled, "Gluing of Ferrite Cores." Philips Magnetic Products recently investigated the relative strengths of 17 glues commonly used to bond ferrite core halves together. The note details the test conditions and presents the results of aging tests.

Philips Components
INFO/CARD #194

New and Reconditioned Test Equipment

The new 41-page Lectronic Research Labs catalog offers savings of up to 70 percent on high-quality reconditioned equipment as well as substantial discounts on hundreds of hard-to-find items including amplifiers, analyzers, generators, component kits, oscilloscopes, signal generators, power supplies, meters, tools and connectors.

Lectronic Research Labs
INFO/CARD #193

RF Telemetry/Range Safety

Hi-rel transmitters, receivers, power amplifiers, transponders, and data links in P, U, L, TV, S, and C-bands are described in a new product selection guide available from Aydin Vector Division. High efficiency and synthesized models are among those described.

Aydin Vector Div.
INFO/CARD #192

Antenna Catalog

A 24-page, four-color catalog from Antenex offers information on the company's whip, portable and Yagi antennas along with antenna mounts, cable assemblies and mounting hardware. Technical specifications and pricing information are listed for each product.

Antenex
INFO/CARD #191

Modulation Analyzer Literature

A two-page data sheet on the Model 8201 modulation analyzer has been released by Boonton Electronics. The black and white literature provides a complete description of the modulation analyzer, including specifications, operation, and applications.

Boonton Electronics Corp.
INFO/CARD #190

Crystals and Crystal Oscillators

The "International Sourcebook: Crystals & Oscillators" from Ecliptek provides a company background and describes Ecliptek's through-hole and surface-mount crystals. Technical specifications are also given for Ecliptek surface-mount and through-hole oscillators.

Ecliptek Corp.
INFO/CARD #189

Cable Assembly Catalog

Applied Specialties' 78-page cable assembly catalog contains part number and current prices for 14 different coaxial connector types. Thirty-four cable types are offered in 56 configurations. The catalog includes specifications and drawings and is available free-of-charge from Applied Specialties.

Applied Specialties Inc.
INFO/CARD #188

Personal Radio Systems

Modern Personal Radio Systems, edited by RCV Macario of the University of Wales, is published by the Institution of Electrical Engineers (IEE) as part of their IEE Telecommunications Series. The 336-page book is derived from a recent IEE Vacation School and is aimed at the engineer entering this field or seeking a good grounding in it. It includes new material on methods of coverage prediction and cellular network planning management. Price is $89.00.

IEE/INSPEC Dept.
INFO/CARD #187

Site Management Guide

TESSCO, a supplier to the cellular, paging, PCS, and two-way communications industries, has published the 1995-96 edition of its Site Management Buyer's Guide. This specialized guide contains only site construction and maintenance products for SMR through PCS frequencies. The guide consolidates the offerings of more than 95 manufacturers of site construction and maintenance products.

TESSCO
INFO/CARD #186

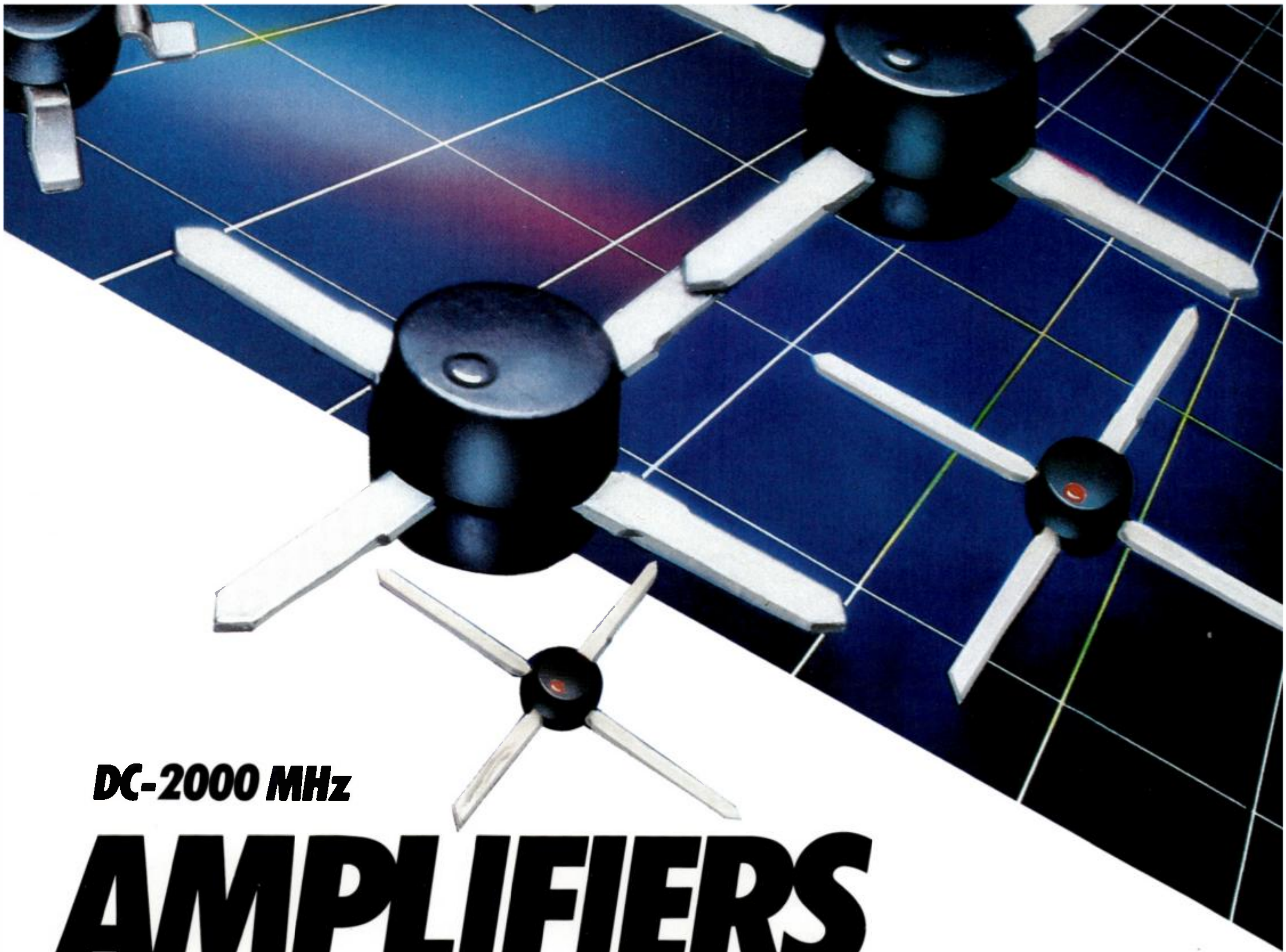
Filter Selection

RLC Electronics introduces its E-Z Filter Finder disk. The E-Z Filter Finder is a DOS/Windows compatible program that will allow the user to input their filter requirements and select an appropriate RLC filter. The program can provide frequency response curves and package dimensions. If a standard RLC filter cannot be specified, the program will generate a quote sheet to be completed and and sent to RLC's Engineering Department.

RLC Electronics, Inc.
INFO/CARD #185

RF guide to editorial coverage

RF advertising index



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	MAR-2	2000	12.5	4.5	6.5	1.12
	MAR-3	2000	12.5	10.0	6.0	1.19
	MAR-4	1000	8.3	12.5	6.5	1.29
MAR SM	MAR-6	2000	20.0	2.0	3.0	1.16
	MAR-7	2000	13.5	5.5	5.0	1.31
	MAR-8	1000	32.5	12.5	3.3	1.27
RAM	RAM-1	1000	19.0	1.5	5.5	*4.80
	RAM-2	2000	12.5	4.5	6.5	*4.80
	RAM-3	2000	12.5	10.0	6.0	*4.80
	RAM-4	1000	8.5	12.5	6.5	*4.80
	RAM-6	2000	20.0	2.0	2.8	*4.80
	RAM-7	2000	13.5	5.5	4.5	*4.80
	RAM-8	1000	32.5	12.5	3.0	*4.80
MAV	MAV-1	1000	18.5	1.5	5.5	.99
	MAV-2	1500	12.5	4.5	6.5	1.12
	MAV-3	1500	12.5	10.0	6.0	1.19
	MAV-4	1000	8.3	11.5	7.0	1.29
MAV SM	MAV-5SM	50-1500	8.0	18.0	6.5	1.90
	MAV-11	10-1000	12.7	17.5	3.6	1.57
VAM	VAM-3	2000	11.5	9.0	6.0	1.19
	VAM-6	2000	19.5	2.0	3.0	1.16
	VAM-7	2000	13.0	5.5	5.0	1.31

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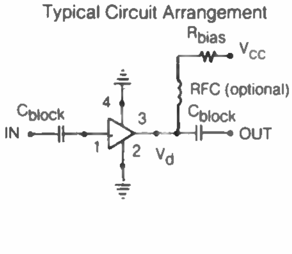
Typical Circuit Arrangement



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r detailed specs on all Mini-Circuits products refer to • THOMAS REGISTER • MICROWAVE PRODUCT DATA DIRECTORY • EEM • MINI-CIRCUITS' 740- pg. HANDBOOK.

INFO/CARD 69

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