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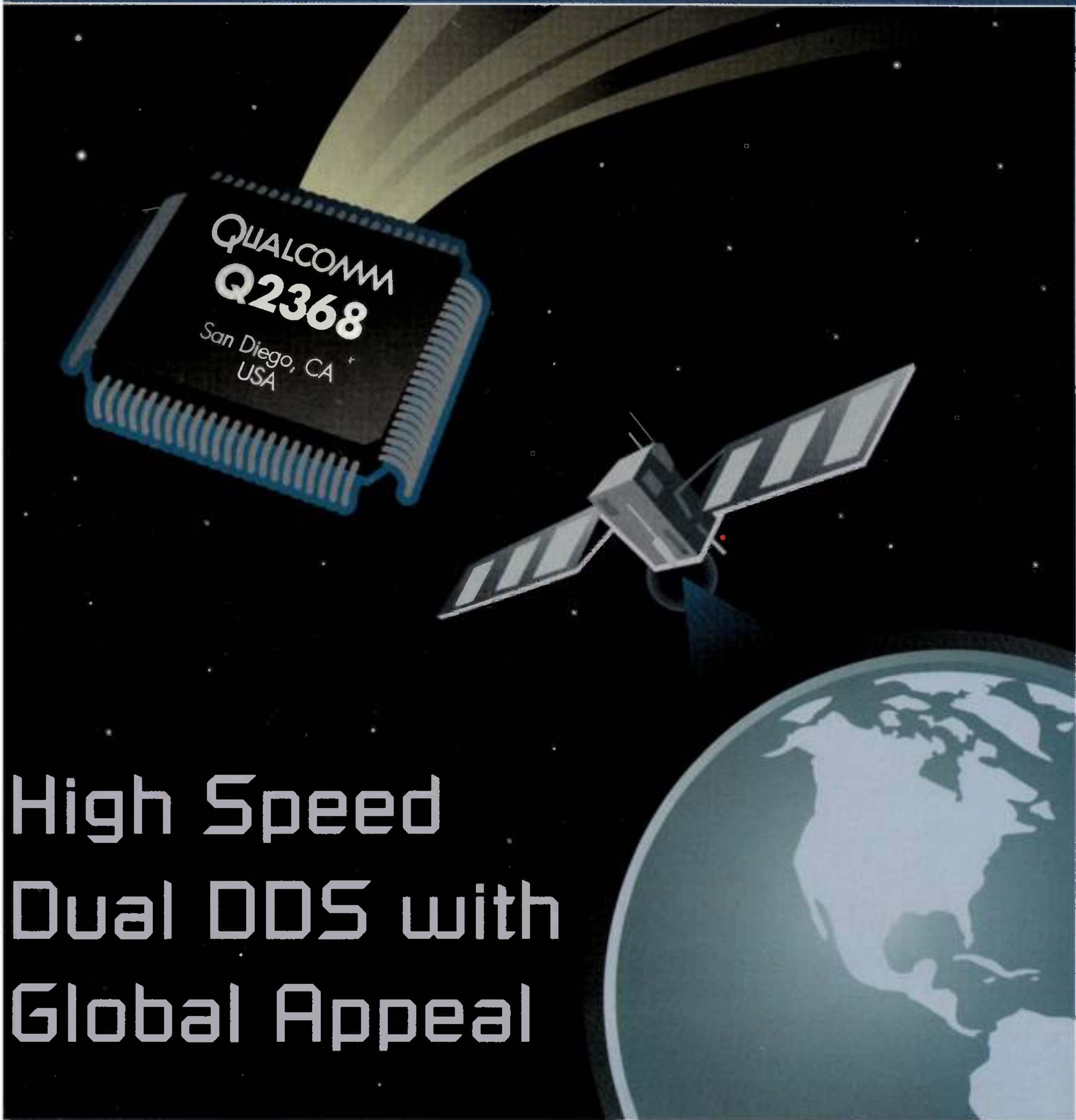
WIRELESS 17 YRS LEADERSHIP

RF design™

engineering principles and practices

May 199



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*This device has not been approved by the Federal Communications Commission.

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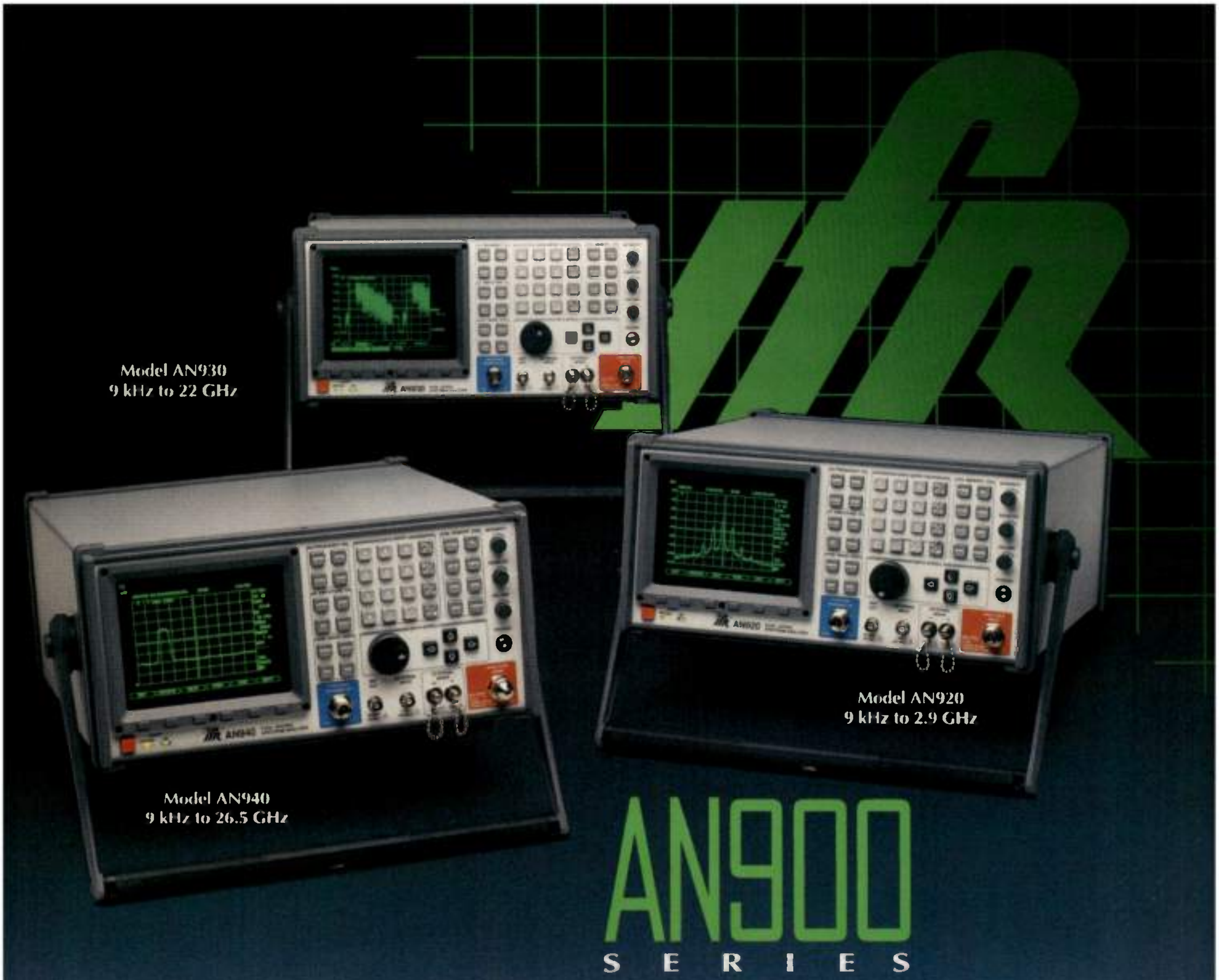

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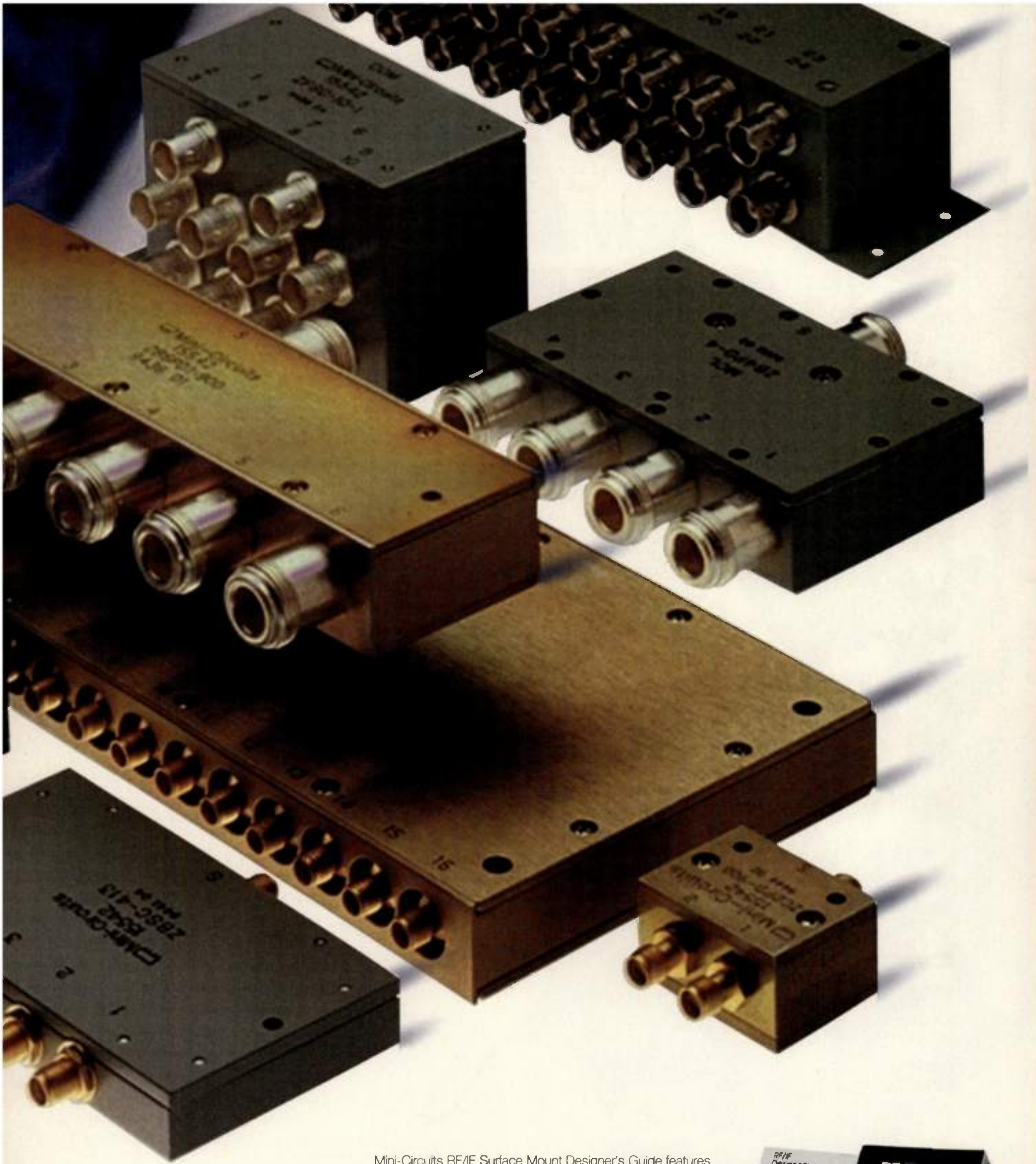

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F 194 Rev Orig

RF design

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departments

RF Design's Mission and New Editorial Team

By Don Bishop
Editorial Director

Working as an editor for *RF Design* is the best job in RF engineering. Know who told me that? Gary Breed, the previous editor for more than 10 years. He's correct... even though he left... but he hasn't left entirely. Gary continues as our consulting editor while he takes up new responsibilities for another company, publishing technical books.

This month, I'm introducing myself. Next month, I'll be introducing our new technical editor. Stacey O'Rourke, our assistant editor, joined us a few months ago. We're the new editorial team.

My enthusiasm for RF dates from my teenage years, when I became a radio amateur and began working for broadcast stations. I was a manager of a consulting engineering firm for a time, and for the past 13 years, I've been an editor with *Mobile Radio Technology*. In fact, many companies that purchase components advertised in this magazine, advertise their finished goods in *Mobile Radio Technology* and other Intertec Publishing magazines. I'm a joiner. I'm a member of IEEE and a life member of ARRL, QCWA and Radio Club of America.

Intertec, by the way, is the new owner of this magazine, as of Dec. 15, 1995. Intertec has been in the publishing business for 110 years, and *RF Design* fits well with many of the company's other titles, such as *Broadcast Engineering, Video Systems, Millimeter, Cellular Business, WirelessWorld, Cellular & Mobile International* and *Dealers' Product Source*. All of these publications serve industries that rely on RF to move information and images form one place to another.

So, that's something about me, and something about Intertec. I'm looking forward to meeting you and our advertisers through phone conversations, correspondence and meetings at conferences.

I'm especially interested in papers to publish. Maybe that goes without saying, but I'm saying it, anyway. From our mission statement:

RF Design is the only technical publication written exclusively for engineers working in radio frequencies where the electronic behavior of circuits and systems requires specialized design techniques. We want to present both classical and leading edge RF design methods using a practical, instructional approach—bridging the gap between general concepts taught in the engineering curriculum and the specific practices required for the successful design of products using RF technology. The broad range of RF applications includes communications, navigation, medical and industrial systems for commercial, consumer, aerospace and military uses.

Ready? Let's go to it!

To reach the editors at *RF Design*, use the addresses and phone numbers on page 10 or the following E-mail addresses:

Don Bishop
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don_bishop@intop.ccmail.compuserve.com
On Compuserve, enter this at the address prompt: CCMAIL: DON BISHOP AT INTOP.

Stacey O'Rourke
Internet:
stacey_orourke@intden.ccmail.compuserve.com
On Compuserve, enter this at the address prompt: CCMAIL: STACEYOROURKE AT INTDEN.

INFO/CARD 6

RF design

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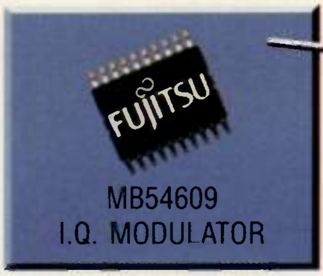
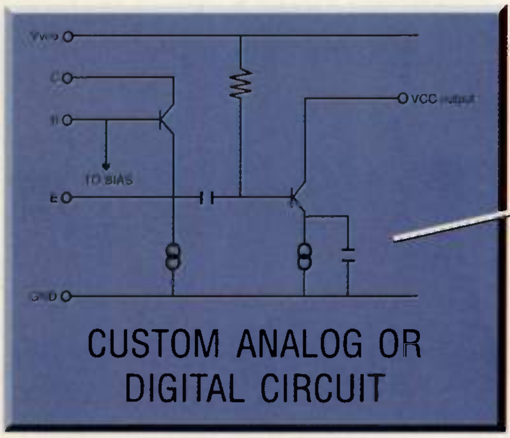
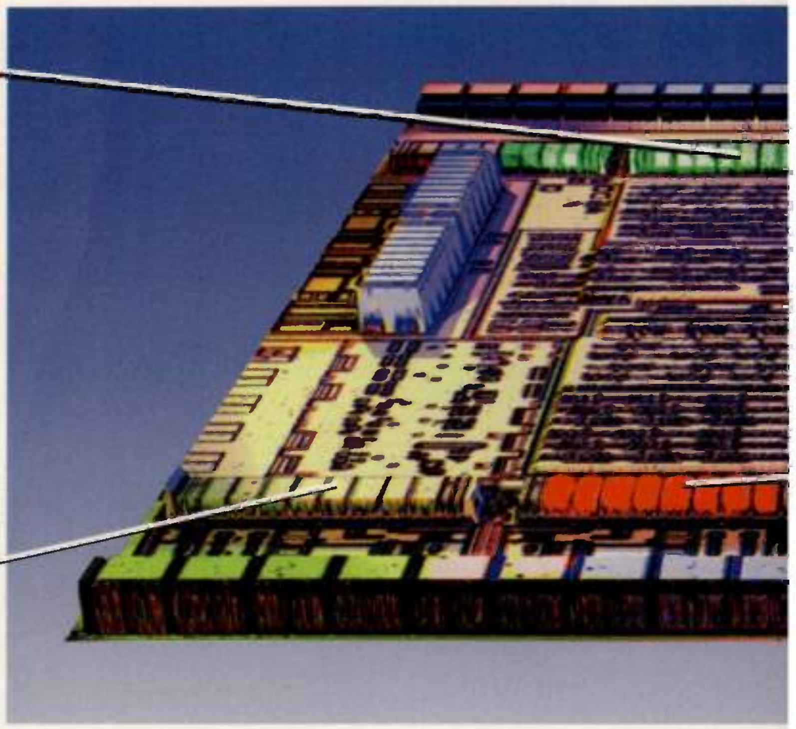
THE ROADMAP TO SEAMLESS RF INTEGRATION







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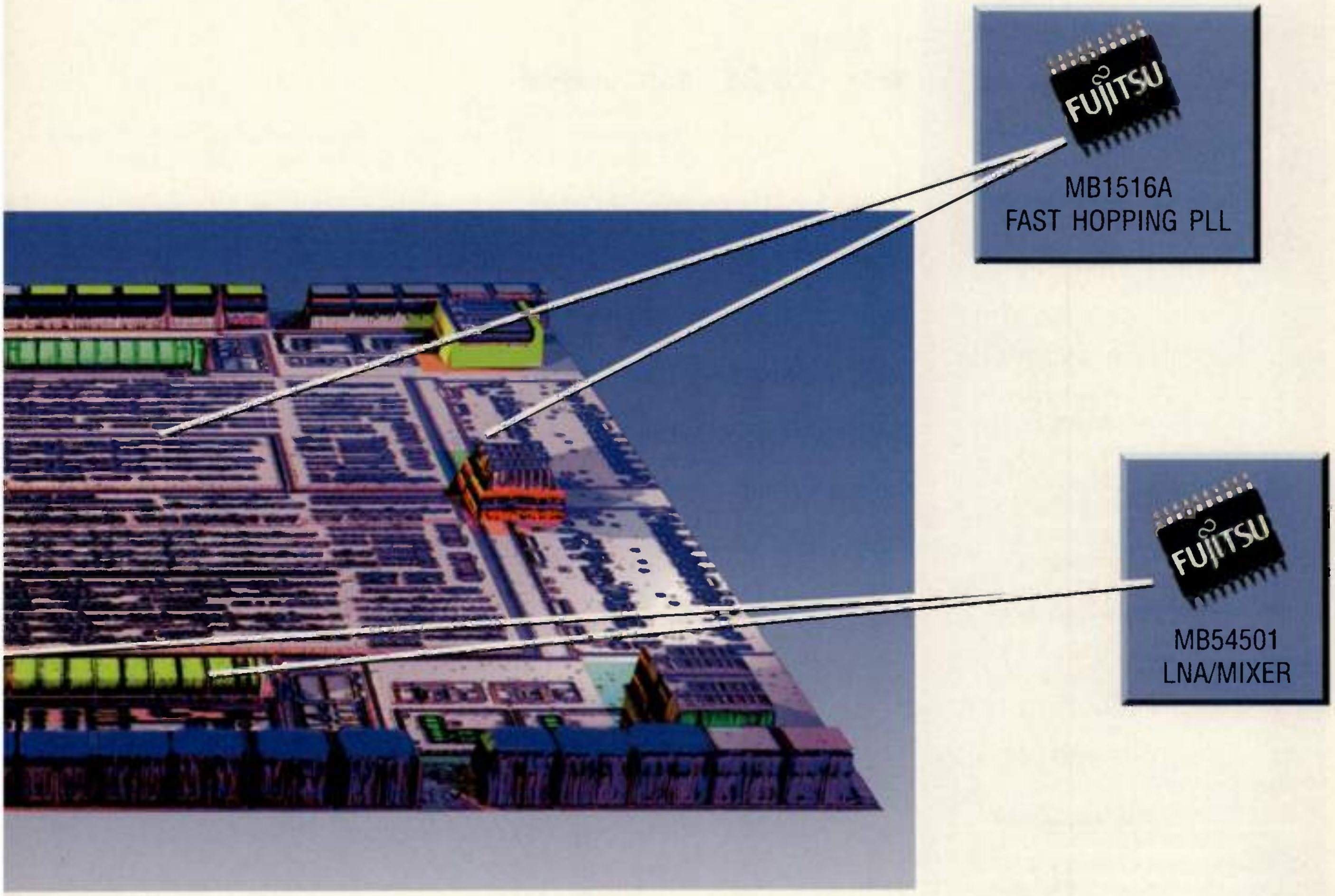

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INFO/CARD 10

RF calendar

May

23-24 **ESD Tutorials**
Bloomington, MN
Information: ESD Association, 7902 Turin Rd., Suite 4, Rome, NY 13440-2069. Tel: (315) 339-6937; Fax: (315) 339-6793.

June

5-7 **Wireless Personal Communications**
Blacksburg, VA
Information: Jack Lilly, Donaldson Brown Hotel and Conference Center. Tel: (540) 231-4849; Fax: (540) 231-9886; E-mail: jacklily@vt.edu.

11-12 **Radio Data Solutions Europe**
Amsterdam, Netherlands
Information: Radio Data Solutions Europe. the old Vicarage, Haley, Hill, Halifax, HX3 6DR, UK. Tel: 44 (0) 1422 380397; Fax: 44 (0) 1422 355604.

16-21 **MTT-S International Microwave Symposium**
San Francisco, CA
Information: Derry Hornbuckle, Hewlett-Packard; Tel: (707) 577-3658; Fax: (707) 577-2036, or Jerry Fiedziusko, Space Systems/Loral Corp. Tel: (415) 852-6868. Fax: (415) 852-5068.

24-25 **Time and Frequency Seminars: Introduction-Level 1**
Boulder, CO

26-28 **Time and Frequency Seminars: Fundamentals-Level 2**
Boulder, CO
Information: Wendy Ortega Henderson, National Institute of Standards and Technology, 325 Broadway, Boulder, CO 80303-3328. Tel: (303) 497-3593; Fax: (303)497-6461. E-mail: ortegaw @boulder.nist.gov

July

21-26 **1996 IEEE AP-S International Symposium and URSI Radio Science Meeting**
Baltimore, MD
Information: Mr. Jon Moellers, Steering Committee Chair, 445 Hoes Lane, P.O. Box 1331, Piscataway, NJ 08855-1331. Tel: (410) 993-6774; Fax: (410) 993-7432.

August

19-23 **IEEE International Symposium on Electromagnetic Compatibility**
Santa Clara, CA
Information: Gherry Pettit, Intel Corporation. Tel: (503) 696-2994; Fax: (503) 640-6411.

21-23 **Wireless Communications Workshop**
Boulder, CO
Information: Dr. Roger Marks, National Institute of Standards and Technology, 325 Broadway, MC 813.06, Boulder, CO 80303. Tel: (303) 497-3037; Fax: (303) 497-7828; E-mail: marks@nist.gov

RF courses

Communications Technology Training
June 4-6, 1996, Orland Park, IL
July 9-10, 1996, Orland Park, IL
August 6-8, 1996, Orland Park, IL
September 3-5, 1996, Orland Park, IL
October 8-10,1996, Orland Park, IL
November 5-7, 1996, Orland Park, IL
December 3-5, 1996, Orland Park, IL
Information: Andrew Corp., Dept 355, PO Box: 9000, San Fernando, CA 91341-9978. Tel: (800) 255-1479, ext. 117.

Microwave Antenna Measurements: Far-Field, Near-Field, Compact Ranges and Anechoic Chambers Topics
June 11-14, 1996, Northridge, CA
Information: Shirley Lang, School of Engineering and Computer Science, California State University, Northridge, 18111 Nordhoff St., Northridge, CA 91330-8295. Tel: (818) 885-2146; Fax: (818) 885-2140; E-mail: shirley.lang@csun.edu

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Intorduction to Global Positioning System (GPS)
June 12,1996, New Brunswick, NJ
Information: Jill Baun, Cook College – Office of Continuing Professional Educations, Rutgers, The State University of New Jersey, PO Box: 231, New Brunswick, NJ 08903-0231. Tel: (908) 932-9271; Fax: (908) 932-1187; E-mail: ocpe@aesop.rutgers.edu.

Wired and Wireless Telecommunications Networking
May 20-24, 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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Grounding & Shielding Electronic Systems, and Circuit Board Layout
June, 1996 (dates TBA), Chicago, IL
August 14-16, 1996, San Jose, CA
Information: Continuing Education, University of Missouri-Rolla, 103 ME Annex, Rolla, MO 65409-1560. Tel: (314) 341-4132; Fax: (314) 341-4992.

Training Program for Cellular, PCS Staff
Independent Learning Program
Information: Virginia Polytechnic Institute and State University, Mobile and Portable Radio Research Group, 840 University City Blvd., Pointe West Commons, Suite 1, Blacksburg, VA 24061-0350. Tel: (540) 231-2970; Fax: (540) 231-2968.

RF and Wireless Made Simple
July 8-9, 1996,. Los Altos, CA
Applied RF Techniques I
July 15-19, 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.

Antennas: Principles, Design and Measurements
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.

DSP Without Tears
June 19-21, 1996, San Jose, CA
Information: Z Domain Technologies, Inc., 325 Pine Isle Court, Alpharetta, GA 30202. Tel: (800) 967-5034; (770) 587-4812; Fax: (770) 518-8368; E-mail: dsp@mindspring.com.

Maximize Performance of Transmission Lines
June 4-6, 1996, Orland Park, IL
Information: Andrew Corporation, Dept. 355, P.O. Box: 9000, San Fernando, CA 91341-9978. Tel: (800) 255-1479 ext. 117.

Modern Receiver Design
Mobile Satellite Communications Systems
May 13-15, 1996, Washington, DC
Modern Digital Modulation Techniques
May 13-17, 1996, Washington, DC
Modern Digital Video Processing
May 15-17, 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
Mobile Communications Engineering
May 22-24, 1996, Washington, DC
Analyzing Communicatoins System Performance
June 10-13, 1996, Washington, DC
Radio Frequency Spectrum Management
June 10-14, 1996, Washington, DC
Spread-Spectrum Communicatons Systems: Commercial and Government Applications
June 10-14, 1996, Washington, DC
Hazardous RF Electromagnetic Radiation: Evaluation, Control, Effects, and Standards
June 12-14, 1996, Washington, DC
Digital and Analog Communication Systems for Non-Engineers: The Fundamentals
June 17-19, 1996 Washington, DC
Digital Cellular and PCS Communications: The Radio Interface
June 24-28, 1996, Washington, DC
Communications Satellite Systems: The Earth Station-A Practical Approach to Implementation
August 5-8, 1996, Washington, DC
Cellular and Wireless Telephony
August 12-16, 1996, Washington, DC
Wireless Infrastructure Network Engineering for Cellular, PCS, LEO, and WPBX
October 21-25, 1996, Washington, DC
Satellite Communications Engineering Principles
November 5-8, 1996, Washington, DC
Grounding, Bonding, Shielding, and Transient Protection
November 11-14, 1996, Washington, DC
Satellite Communications WIth Emphasis on Mobile Systems
December 2-4, 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; E-mail ceepinfo@ceep.vpaa.gwu.edu.

Japanese Electronics Industry Forecasts

In 1995, total production by the Japanese electronics industry is estimated to climb 4.7% from the previous year, to ¥22,413.9 billion, primarily owing to expanded internal demand. In 1996, continued growth in internal demand and increased demand for semiconductors are forecast to support a 5.6% production rise from 1995, to ¥23,680.2 billion. The 1996 forecast by product segment, the domestic market for consumer electronic equipment is likely to continue to be characterized by further moves toward off-shore production, marked expansion in imports and decreasing prices. With the expected introduction of new-concept products such as digital video cameras (DVDs) and digital video disks (DVDs), domestic production in this segment is seen rising 1.7% to ¥2,543.8 billion.

US Electronic Warfare Market to Rebound

Constrained by military cutbacks but propelled by awareness of the value of high-technology defense, US sales of electronic warfare equipment will grow from $1 billion in 1995 to $1.45 billion by the year 2001 rebounding after a decline to $946 million in 1997 as programs completing development in the mid-1990s enter production late in the decade projects a new study just release by Frost & Sullivan. Rotary-wing aircraft equipment will rise from 13 percent of total EW market revenues in 1995 to 20 percent in 2001 while dominant fighter/attack aircraft equipment dips from 72 to 61 percent in the same period forecast the report, US Electronic Warfare Market. Other segments include other fixed-wing aircraft equipment, with 8 percent of 1995 revenues, ground-based equipment, 5 percent and shipboard equipment, 3 percent, according to the study.

Decline in Semiconductor Content in Cellular Phones

International Data Corporation (IDC) released new forecasts which

Business Briefs

Prices Cut Across Entire Signal Processing Product Group—Harris Semiconductor announces sweeping price cuts across its signal processing product line of linear, data acquisition and digital signal processing building block chip for video/imagining and emerging communications applications. Hundreds of products effected, include op amps, data converters, industry-standard switches and mixers as well as hardwired DSP ICs for wireless communications and image processing.

National Semiconductor Research Center Joins UC San Diego Wireless Communications Consortium—National Semiconductor Corporation has joined the Center for Wireless Communications, a research center sponsored by the University of California, San Diego School of Engineering. The new National Semiconductor Research Laboratory (NSRL) will manage National's participation in the Center supporting the company's focus on moving and shaping information through researching advancements in networking and network access devices.

Expanded Facility—In response to increased sales and continuing growth, RF Power Components, Inc. has expanded their facility to accommodate additional personnel and expanded manufacturing capabilities. This will allow for the installation of new manufacturing equipment, an expansion in the production floor, shipping departments and the addition of much needed office space. The renovation is expected to be completed this Summer.

Silicon Valley Operations Expanded—RF Power Products, Inc. has relocated its corporate marketing staff to San Jose, CA, joining the sales, customer support and applications engineering functions already on site at this expanded location. Their new address is 780 Montague Expressway, Suite 307, San Jose, CA 95131.

Micro Extrusion Manufacturing Operations Established—Coors Ceramics Company's Electronic Products Group's facility in Grand Junction, CO has established a high volume and prototype manufacturing facility for its' micro extruded products. The ISO 9002 certified Grand Junction facility, which also manufactures thick film alumina substrates and ceramic ferrules for fiber optic connectors, produces micro extrusions for electronic applications.

Meta Wave Communications Has Moved—Their new address is: MetaWave Communications Corporation, 8700 148 th Ave. NE, Redmond, WA 98052. Tel: (206) 869-7499; Fax: (206) 869-2778.

European Presence Expanded—Ansoft Corporation has established a European sales office. Based in London, the office will support all of Europe with a goal of establishing additional sales and support sites in Europe, specifically in Germany and France.

Name Change—Electronic Technology Corp. (ETC), a leader in wireless communications and electronic manufacturing changed its company name to Digital Radio Communications Corporation. The new name reflect's the company's vision and central focus of operation. Digital Radio Communications Corporation will continue to use the ETC name through a newly-formed business unit: Electronic Technology Company.

European Compact Software, Inc has moved—Electronic Software Components (ESC), the European Compact Software, Inc. Representative announces their new address: ESC Electronic Software Components GmbH & Co. Trade KG, Ramersdorfer Strasse 1, D-81669 Muenchen, Germany. Tel: 49 89 680055 0; Fax: 49 89 680055 22; E-mail: 100407.1372@compuserve.com.

National Instruments to Acquisition—National Instruments of Austin, Texas, has agreed to acquire Georgetown Systems, Inc. Georgetown Systems is the privately held manufacturer of Lookout™, and industrial automation software package for Windows PCs. It is anticipated that Georgetown Systems employees will become National Instruments employees upon closing. Lookout development efforts will continue at the Georgetown Systems office.

show the semiconductor content of a typical cellular handset will decrease from an average cost of $75.64 in 1995 to $57.32 by the year 2000. Even as component costs decline, functionality, reliability and the ability to customize handsets will increase. The major contributor to lower costs is the growing integration of cellular ICs.

Cable Technology Markets to Pass $2 Billion

Paced by the upgrading of cable systems to offer telephony, video-on-demand and interactive services like home shopping, US sales of cable television-associated hardware and technology services will grow from $1.78 billion in 1994 to $2.36 billion by the year 1999 at a 6 percent compound annual rate, projects a new study just release by Frost & Sullivan. Services including engineering, construction and network management will rise from 10 percent of the combined markets' revenue in 1994 to 25 percent in 1999. Transmission and distribution lines rise from 125 to 24 percent while currently-dominant line accessories decline in share from 53 to 34 percent in the same period. Head-end and studio equipment accounted for 19 percent of cable technology market revenues in 1994 and security equipment another 3 percent, estimates the study.

Contracts

Direction Finder Selected for F-15 Flight Tests—Following an intensive, four-year down select competition, Litton's Amecom division, College Park, MD has been chosen by the US Air Force and McDonnell Douglas Aerospace, St. Louis, MO, to provide Litton's new LT-500 Precision Direction Finding (PDF) system for flight tests on an F-15 aircraft. The system will be delivered under a $2.4 million contract from McDonnell Douglas for flight test slated this summer.

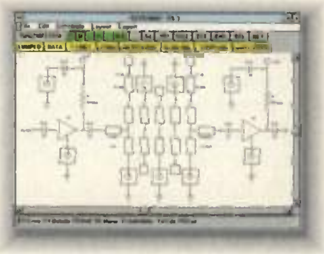
Exclusive Supplier of Front-End Technologies for TELE-TV Project Selected—ComStream has been selected by Thomson Consumer Electronics as their exclusive supplier for front-end technologies and products for the nation's largest wireless television system now under development. Thomson is prime contractor for TELE-TV Systems' MMDS (Multichannel Multipoint Distribution System) designed to deliver up to 100 television channels to some 3 million homes beginning in late 1996. The Thomson contract TELE-TV Systems is valued at %1 billion through the year 1999 and will include the delivery of some 3 million digital set-top receivers. The receivers will carry both the TELE-TV and the RCA brands.

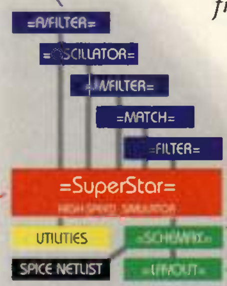
UTAM Selects Comsearch as their Prime Frequency Coordinator — UTAM has selected Comsearch as their prime frequency coordinator (PFC) to assess the interference potential for deployment of non-nomadic (coordinatable) unlicensed PCS products in the United States. UTAM, designated by the Federal Communications Commission as the frequency coordinator of the unlicensed PCS spectrum, will be managing the transition of this spectrum from fixed microwave use to unlicensed PCS. As PCS, Comsearch will perform all Zone 1 county coordination. In addition, Comsearch will perform oversight on all Zone 2 coordination.

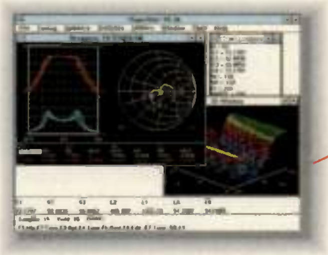
DESIGN FROM START-TO-ART

1 GENESYS synthesis modules make it easy to create initial designs for L-C filters, matching networks, active filters, oscillators, printed and machined filters, amplifiers and group delay equalizers. The many design alternatives and quick evaluation of each help you save time and achieve higher performance.

2 Schematics make it easy to enter your own designs, modify synthesis created designs and merge subcircuits. Schematics and simulation are integrated for back annotation. The GENESYS schematic module also writes Touchstone and Spice files.







3 =SuperStar= Professional redefines the circuit simulator standard. Step up to real-time tuning, statistical analysis, unrestricted global noise analysis, adaptive optimization, accurate models, unparalleled execution speed and the industry's easiest to use interface. Non-linear and time domain analysis is supported through the Spice interface.

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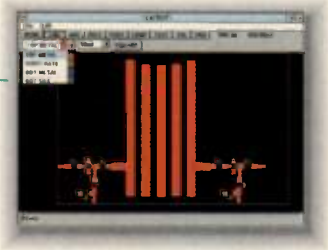

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INFO/CARD 16

Major Wireless Contract—ITS Corporation was awarded a major contract for two metropolitan area digital wireless (MMDS) systems. The contract with CAI Wireless Systems, Inc., involves construction of a large network of digital transmitters and boosters for Boston, Massachusetts, and Norfolk, Virginia. They are the first phase of a project to implement digital wireless cable service in 13 major metropolitan areas in the eastern United States. The orders for Boston and Norfolk encompass a minimum of thirty-three transmitter for each main and booster site, along with appropriate combiners, backup equipment and antennas, ITS will also perform system integration to provide equipment shelters and a full range of engineering services to CAI.

GeoSat Follow-On Satellite Selects GPS—Allen Osborne Associates, Inc. (AOA) will provide the Global Positioning System (GPS) receiver for the US Navy's GeoSat Follow-On (GFO) Satellite now being manufactured by Ball Aerospace and Technology Corporation (BATC). BATC Selected the TurboStar®, a satellite-borne GPS receiver based on AOA's TurboRogue® GPS receiver because of its exceptionally clean carrier phase data. The Navy's requirement for the GFO is to determine the satellite's orbit to 5cm using post processing techniques. A total of 4 TurboStar® receivers have been ordered by BATC. A TurboStar® receiver has been on orbit in the GPS Meteorological Satellite (GPS/MET) since April, and another was flown briefly on the Wake Field shuttle payload.

Spectrum Monitoring Solution Supplied to Colombia—Technology for Communications International (TCI) of Sunnyvale, California, and Hewlett-Packard Company will play a critical role in meeting Columbia, South America's spectrum monitoring needs for detecting and identifying unlicensed users of its airwaves. TCI has been awarded a $17 million contract from the Colombian Ministry of Communications for an International Telecommunications Union (ITU), complaint spectrum monitoring and management system. TCI will provide the operating software for the license database, administrative processing, engineering analysis tools and geographic map display, direction-finding equipment, overall system integration and turnkey installation, and commissioning of the system in Colombia. HP will supplying computers and networking hardware, test and measurement equipment and a portion of the operating software of the system. The system is scheduled to be operational in Colombia by 1997.

Radio Frequency Identification—As society grows increasingly dependent on computers and their ability to access data instantly, many companies have become concerned that some information may be a little too accessible. To battle this problem, CADIX, Inc. of Japan, has developed ID-007, an automatic, electronic information access control system that incorporates advanced TIRIS™ (Texas Instruments Registration & Identification System) technology. The ID-007/TIRIS system can be used to rapidly identify and authorize registered users, limiting access to sensitive information and applications. The system can also be used to log information and verify the identity of employees as they enter or exit the work environment.

Digital Cellular Technology Drives Component and Instrument Makers

By Gary A. Breed
Consulting Editor

Digital cellular, whether North American Digital Cellular (NADC), Code Division Multiple Access (CDMA), or the Global System for Mobile communications (GSM), is in a rapid growth mode at present. GSM is reaching maturity in European systems, and serves as the basis for new PCS systems yet to be constructed. NADC is in early implementation in the U.S. market, with CDMA systems just getting started.

Most industry news regarding digital cellular concerns is business implications — investments by major companies, numbers of subscribers using the new technology, or the rate of construction in various markets. For an RF engineer, this kind of data is useful, but more matters of direct impact are more important.

For engineers who are designing digital cellular telephone equipment, new components are probably the most interesting things. Semiconductors in particular are foremost in on an engineer's list of priorities. What functions are being integrated onto how few chips? How many external components are required to complete the design? Can all components be accommodated in a highly-automated manufacturing process?

Integrated circuits contain the active portions of a circuit, including speech processing, signal generation, up- and downconversion, system control, etc. Discrete transistors or modules perform the power amplification tasks and transmit/receive switching. The efforts made by companies in this part of the RF business have been well-documented as they strive to integrate more functions onto a single chip of silicon, gallium arsenide or silicon-germanium.

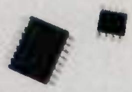
Passive components have gotten less attention, in part because they are used in all types of circuitry, not just cellular or other "hot" wireless applications. In digital cellular, even these more-or-less universal components may have unique requirements. Filters must have predictable group delay, attenuators and switches must be rated for the less-efficient class A power amplification used for digital cellular, and all passive components must contribute virtually no distortion to the system.

Passive components are also important in digital cellular design. Surface-mount filters like the Kel-Com device pictured above (top), and couplers such as the RF Power Components units (bottom) are needed to keep the cost of more complex digital cellular products in line with customer expectations.

Manufacturers of digital cellular subscriber equipment have a high interest in test equipment, like this CDMA Mobile Station Test System from Noise Com.

The manufacturing process is growing in importance. Cellular telephone and paging equipment have driven U.S. RF equip[ment manufacturers to higher quantity production, requiring greater automation. With the relative complexity of digital cellular, the production test equipment must include capabilities for testing the operation of a large number of signal and control functions. Nearly all RF test equipment makers have addressed this need, developing both new test instruments, and creating operating software to customize their general purpose units. In some cases, several units are integrated into a single test system, a job previously left to the customer.

In summary, digital cellular represents perhaps the largest design and manufacturing effort at this point in time. PCS, WLAN and other applications will come next, but GSM, NADC and CDMA are in the pipeline right now. Until those systems are fully implemented, digital cellular will continue to be a significant driver of RF component and test product development. *RF*



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	1510 - 1660	0.4	20	1.3:1	0.05	1	DS52-0004
	1700 - 1900	0.3	20	1.3:1	0.05	2	DS52-0005
	1850 - 1990	0.5	21	1.2:1	0.05	1	DS52-0002
	2200 - 2500	0.3	20	1.3:1	0.05	3	DS52-0003
4-Channel							
	824 - 960	1	23	1.2:1	0.30	2	DS54-0001
	1200 - 1660	1	23	1.2:1	0.30	2	DS54-0003
	1700 - 2000	1	23	1.4:1	0.30	3	DS54-0002
New	2200 - 2500	1	21	1.4:1	0.20	2	DS54-0004
6-Channel							
	824 - 960	1.3	25	1.4:1	0.30	6	DS56-0001

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INFO/CARD 19

Measuring Range and Reliability for Part 15 Systems

By Bernard Kasmir, P.E.
Consultant

FCC rules limit the power of Part 15 transmitters. This limits the range of a system. Part 15 receivers are also limited in performance by economic necessity especially in the choice of the accuracy of the local oscillator. This, in turn, affects the receiver bandwidth that ultimately defines the total systems range. Assuming that reasonable design compromises have been made (perhaps a dangerous assumption worthy of other articles), designers are therefore interested in defining the systems range and reliability of their systems. The basic question then becomes "how far will this system work?"

Perhaps the first question to consider, is how to define performance. Ultimately, performance is related to the ability to decode a received signal in terms of bit error rate, packet error rate or educated eyeballs. One axiom usually true is that performance is correlated to distance since in the limit, the signal will be too weak to be received and decoded.

The development of a method of systems performance evaluation would be desirable. Since distance is involved in signal reliability, first choose a distance. There may be a tendency to range the system out for maximum distance, but this leads to unreliable results. Figure #1 shows the relative signal level as a function of distance. This follows the 1/X curve for field intensity and the $1/X^2$curve for power. At large distances, the slope of this curve becomes very shallow and small level changes can be experienced over relatively large changes in distance.

More important is that there is no reason to measure performance at distances that the system does not use. It would be more useful to define the range the equipment will be operated at and then to define systems performance at that distance in terms of signal margin and systems reliability.

System Calibration

Some initial measurements will define the operating parameters for a useful, convenient graphical prediction of performance.

Initially, we measure the field intensity of the transmitter at some convenient distance, say three meters. Place the transmitter three meters from a receiver (or spectrum analyzer) using a calibrated antenna.

The field intensity is: in micro-

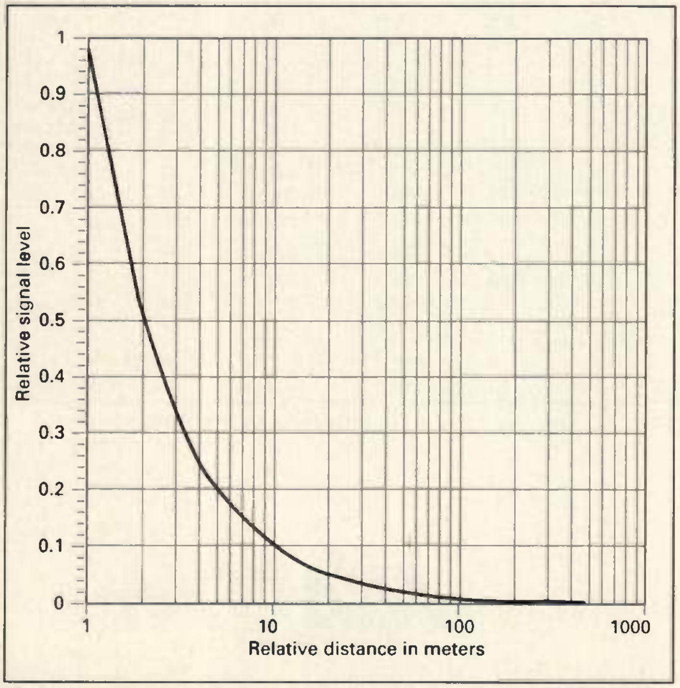

Figure 1. Theoretical relative signal level versus distance

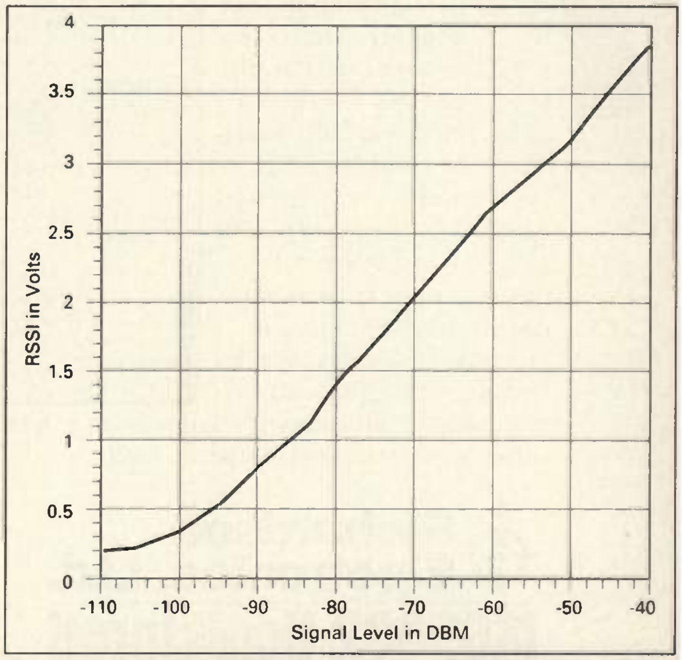

Figure 2. Receiver calibration curve – RSSI output voltage versus power at receiver input.

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	POS-50	25-50	-110	-19	20	11.95
	POS-75	37.5-75	-110	-27	20	11.95
	POS-100	50-100	-107	-23	20	11.95
	POS-150	75-150	-103	-23	20	11.95
	POS-200	100-200	-102	-24	20	11.95
	POS-300	150-280	-100	-30	20	13.95
	POS-400	200-380	-98	-28	20	13.95
	POS-535	300-525	-93	-26	20	13.95
	POS-765	485-765	-85	-21	22	14.95
	POS-1025	685-1025	-84	-23	22	16.95
NEW	POS-1060	750-1060	-90	-11	30*	14.95
NEW	POS-1400	975-1400	-95	-11	30*	14.95
NEW	POS-2000	1370-2000	-95	-11	30*	14.95

*Max. Current (mA) @ 8V DC.
Notes: Tuning voltage 1 to 16V required to cover freq. range. 1 to 20V for POS-1060 to -2000. Models POS-50 to -1025 have 3dB modulation bandwidth, 100kHz typ. Models POS-1060 to -2000 have 3dB modulation bandwidth, 1MHz typ. Operating temperature range: - 55°C to +85°C.

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volts/meter is calculated as:

$$E = 10^{((DBM + 107 + AF + CL)/20)} \quad (1)$$

Where DBM = value read on receiver, AF = antenna factor of the test antenna (value obtained from manufacturer or calculated from antenna gain) AF = 9.75/λX√gain, CL = Cable loss from test antenna to receiver. Given CL = 6dB E = 40,000 µv/m AF = 19.9 dB The value of DBM for this measurement would be –40.8 dBm.

Next, Calibrate the receiver. For Part 15 operation, most systems use pulse modulation. The receivers usually have a log video output known as the received signal strength indicator (RSSI)

The receiver is calibrated by plotting the RSSI output vs. DBM input. A typical plot is shown in Figure 2. It is also necessary to determine the receiver sensitivity and reducing the DBM into the receiver until (by whatever means listed above) decoding ceases.

This is defined as the threshold value and is also recorded.

From Figure 2, threshold sensitivity is –104 dBm

Plotting a Best-Case Line

Next, we go to the performance worksheet shown in Figure 3. The first significant anchor point on this curve is obtained by the DBM reading obtained when measuring transmitter field intensity and the plot of Figure 2 which is the receiver plot of RSSI versus DBM From Figure 2, determine the RSSI voltage for –40 dBM is 3.8 volts.

In the ideal case, the signal level decreases 20 dB/decade. Draw a line from this anchor point with this slope. From Figure 2 in this example, the slope is 50mv/dB, or 1 volt for 20 dB. This definees the best case condition on line AB.

Next, draw a horizontal line (CD) for the receiver decode threshold value previously measured. We then define the distance desired to characterize systems performance. In this example, we chose 100 meters or about 300 feet. One more line, this is a vertical line (GH) from the value of 100 meters.

The best case signal margin can now be obtained:

The point of interest is the intersection of the field intensity line (AB) and the vertical line at 100 meters (GH). This is the theoretical level that will produce a RSSI voltage of 2.3 volts Next, we take the vertical displacement between lines ICJ and lines CD This is the signal margin. This displacement is 2 volts. 20 dB/volt slope, this becomes a signal margin of 40 dB.

Reliability and signal margin are related in Table 1.

Plotting Actual Performance

Now the fun begins. The system is calibrated in terms of best case for parameters we have defined. We now perform some measurements with the total system of transmitter and receiver.

Margin	Reliability
0 dB	99%
30 dB	99. 9%
40 dB	99 99%

Table 1. Correlation between margin and reliability.

Back at the test site, place the systems transmitter and receiver three meters apart and record the RSSI produced by the receiver. In this example, the 40,000 µv/m produced a RSSI voltage of 3.3 volts

Referring to Figure 4, the anchor point of 3.3 volts is below the original anchor point of 3.8 volts indicating that the antenna system is not quite as good as the test antenna. This may be due to several factors such as inadequate antenna counterpoise, poorer antenna match, different antenna pattern, etc.

Next, take RSSI readings at various distances and draw a second curve (A´B´) This curve ideally should be identical to line AB, but in actuality shows a somewhat poorer anchor point at 3.3 volts, or 10 dB lower. Also, this curve A´B´ starts to depart from 20 dB/decade as the distance increases. Curve A´B´ intersects the 100 meter distance at 1.5 volts. The vertical dis-

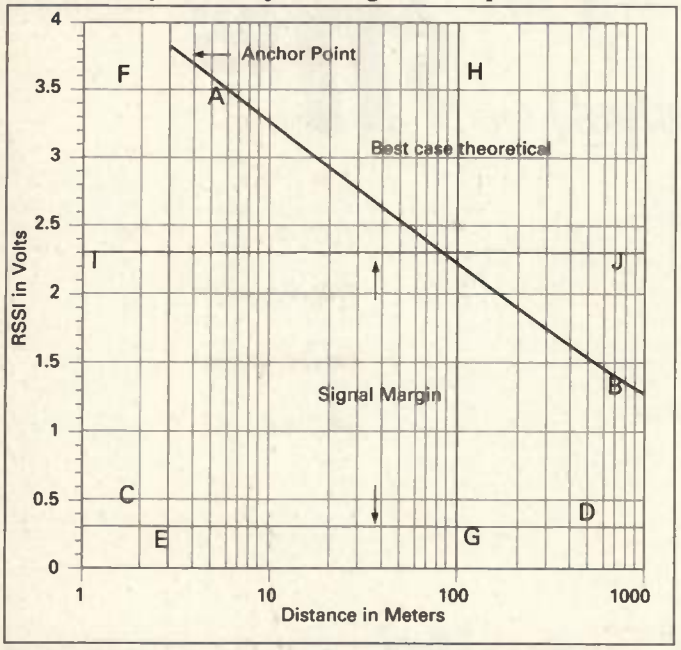


Figure 3. Worksheet for range measurements with just theoretical best case plotted for a particular receiver.

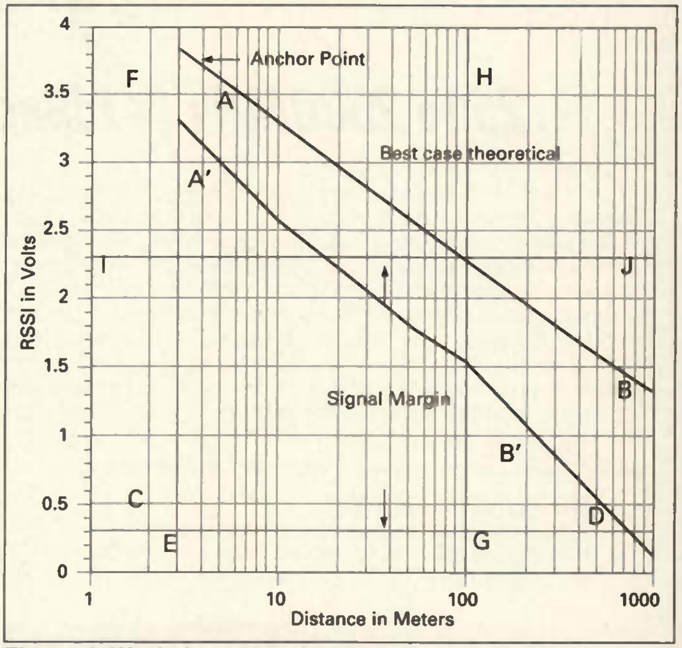
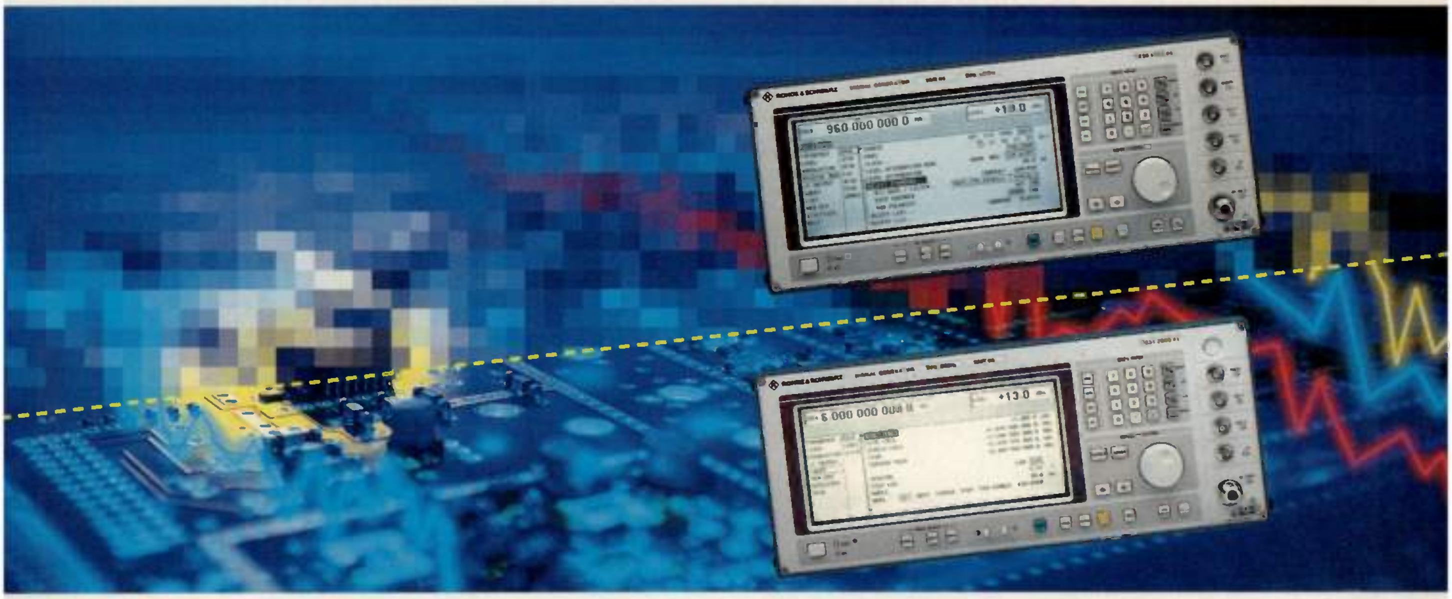


Figure 4. Worksheet showing both theortical best case and actual measured signal strength.



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INFO/CARD 21

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60	13.0	80	17.5	4	2.0	100	3100//0
2	TE5080	3	7.50	20	35.0	-	-	1	1.0	50	3000//0
4	TE5090	3	7.50	30	17.5	-	-	2	2.0	75	3300//0
6	TE5100	6	7.50	60	22.5	-	-	3	2.0	90	3300//0
8	TE5110	6	7.50	60	15.0	80	20.0	3	2.0	100	3300//0
2	TE5120	3	15.0	20	70.0	-	-	1	1.0	35	5000//-1
4	TE5130	3	15.0	30	35.0	-	-	2	2.0	60	5000//-1
6	TE5140	6	15.0	60	45.0	-	-	2	2.0	90	5000//-1
8	TE5150	6	15.0	60	30.0	80	40.0	3	2.0	100	5000//-1

21.4 MHz

NO. POLES	TEMEX P/N	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM. Ω/PF
2	TE5180	3	3.75	15	12.5	-	-	2	1.0	50	850//+6
4	TE5190	3	3.75	30	12.5	-	-	3	2.0	70	850//+5
6	TE5200	6	3.75	60	12.5	-	-	4	2.0	90	850//+5
8	TE5210	6	3.75	60	10.0	80	12.5	5	2.0	100	850//+5
2	TE5220	3	6.50	15	20.0	-	-	2	1.0	50	1300//+2
4	TE5230	3	6.50	30	22.5	-	-	3	2.0	70	1400//0
6	TE5240	6	6.50	60	22.5	-	-	4	2.0	90	1400//0
8	TE5250	6	6.50	60	17.5	80	22.5	4	2.0	100	1400//0
2	TE5260	3	7.50	15	25.0	-	-	2	1.0	50	1500//0
4	TE5270	3	7.50	30	25.0	-	-	3	2.0	70	1600//0
6	TE5280	6	7.50	60	25.0	-	-	4	2.0	90	1600//0
8	TE5290	6	7.50	60	20.0	80	25.0	4	2.0	100	1600//0
2	TE5300	3	15.0	15	50.0	-	-	2	1.0	45	3000//0
4	TE5310	3	15.0	30	45.0	-	-	3	2.0	60	3000//-1
6	TE5320	6	15.0	60	45.0	-	-	3	2.0	90	3000//-1
8	TE5330	6	15.0	60	33.0	80	45.0	4	2.0	100	3000//-1

45.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	LOSS dB	RIPPLE dB-MAX	ULT. REJ. dB-MIN.	TERM. Ω/PF
2	TE9420	3-OT	3	3.75	18	16.0	3	1	40	2000//-1.0
4	TE9310	3-OT	3	3.75	30	12.5	3	1	70	2000//-1.0
2	TE7420	3-OT	3	7.50	18	28.0	2	1	40	3000//-1.0
4	TE7430	3-OT	3	7.50	40	30.0	3	1	70	3000//-1.0
2	TE7440	3-OT	3	15.0	15	47.0	2	1	40	8000//-1.5
4	TE7450	3-OT	3	15.0	30	50.0	3	1	70	8000//-1.5
2	TE7730	FUND	3	15.0	15	50.0	2	1	40	1100//+1.5
4	TE7740	FUND	3	15.0	40	60.0	3	1	70	800//+1.0

70.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10400	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10410	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10420	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10430	3-OT	3	10	35	40	80	-910	3	1	2500//-1

90.0 MHz

NO. POLES	TEMEX P/N	MODE	PASSBAND dB	PASSBAND ±KHz	STOPBAND dB	STOPBAND ±KHz	STOPBAND dB	STOPBAND KHz	LOSS dB	RIPPLE dB-MAX	TERM. Ω/PF
2	TE10440	3-OT	3	7.5	18	30	35	-910	2	1	2000//-1
4	TE10450	3-OT	3	7.5	35	25	80	-910	3	1	2000//-1
2	TE10460	3-OT	3	10	15	30	35	-910	2	1	2500//-1
4	TE10470	3-OT	3	10	35	40	80	-910	3	1	2500//-1
4	TE10480	3-OT	3	15	30	50	80	-910	3	1	4000//-1

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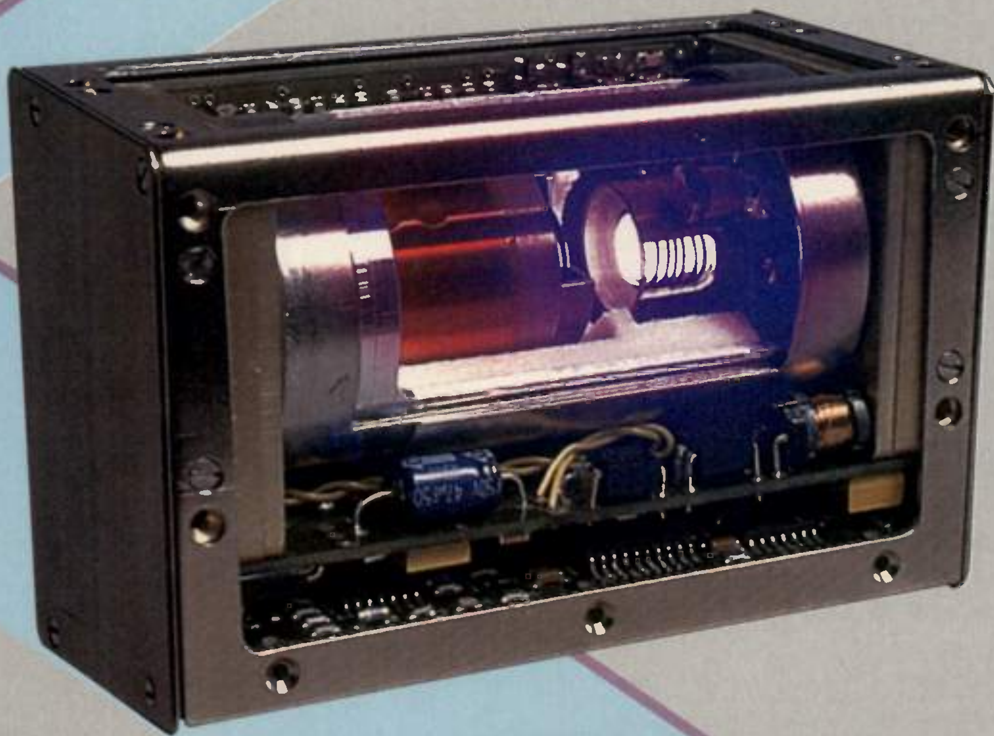

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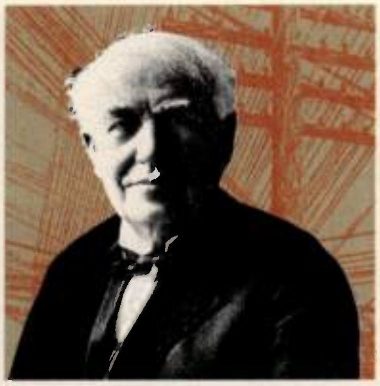

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placement (signal margin) from this intersection and the threshold value is 1.2 V (1.5-.3), or 24 dB as compared to 40 dB for the ideal curve.

As the environment becomes more hostile, range worksheets will show deteriorating values of signal margin. However, each site is different and have to be evaluated individually. Because of the variability of measurements caused by multiple reflections, measurements for curve A´B´ should be repeated several times. We than can either average out measurements, or reproduce worst case, best case by producing a thicker A´B´ line

The advantage of this graphical analysis is that an installer or field engineer will immediately know just how this particular system is functioning with respect to theory. Severe discrepancies between the ideal and measured curves can be evaluated and corrected once one knows what to expect.

Recent studies (1,2) have shown that the propagation path loss departs from the 1/X curve as the distance increases and as obstacles are encountered in the transmission path. The attenuation can deteriorate from 20 dB/decade to 30 and 40 dB/decade. This is another reason for evaluating systems performance at the distance or interest rather than to evaluate performance at some unused maximum distance. **RF**

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About the Author

Bernard Kasmir is a licesed professional engineer and works as an independent consultant. He has written several artic les previously for *RF Design*. He can be reached at Tel: (201) 568-4498 or E-mail: Bkasmir@aol.com.

A Low Voltage Monolithic FM/IF System for DECT and Other High Speed GFSK Applications

By Yanpeng Guo, Randall Yogi, and Alvin Wong
Philips Semiconductors

A Philips low voltage high performance monolithic FM/IF system, the SA639, is introduced to meet the increasing demand for high speed digital wireless PCS applications. Experimental performance evaluation including Bit error rate (BER), sensitivity to frequency off-set, and sensitivity to FM deviation variation of this system is presented. Results indicate that the low voltage SA639 FM/IF system provides superior performance for high speed digital wireless applications.

Standard	CT2/CT2+	DECT	PHS	PACS
Region	Europe/Canada	Europe	Japan	USA
Freq. Band (MHz)	CT2: 864-868 CT2+: 944-948	1880-1900	1895-1918	Tx: 1850-1910 Rx: 1930-1990
Duplex	TDD	TDD	TDD	FDD
Multiple Access	TDMA	TDMA	TDMA	TDMA
No. of channels	40	10	77	16 pairs
Ch. spacing (kHz)	100	1728	300	300
Users/channel	1	12	4	8/pair
Modulation	GFSK (FM dev. 14-25 kHz)	GFSK (FM dev. 288 kHz)	PI/4-DQPSK	PI/4-DQPSK
Bit Rate	72 kb/s	1.152 Mb/s	384 kb/s	384 kb/s
Speech coding	32 kb/s ADPCM	32 kb/s ADPCM	32 kb/s ADPCM	32 kb/s ADPCM
Frame duration	2 ms	10 ms	5 ms	2.5 ms
Peak power	10 mW	250 mW	80 mW	200 mW

Table 1. Summary of digital cordless standards.

To achieve the goal of wireless personal communications, allowing users access to the capabilities of the global communications network at any time without regard to location and mobility, cellular and cordless telephony have been taken as two major approaches. Cellular systems are evolving towards smaller cells (microcells) and lower power levels to provide higher overall capacity. Cordless telephones have evolved from home appliances towards wide spread "universal" low power personal communications systems. With the advent of digital cordless telephony, cordless systems with enhanced functionality have been developed that can support higher data rates and more sophisticated applications such as wireless private branch exchanges (WPBX) and public-access Telepoint systems. One of the first digital cordless standards is the Digital European Cordless Telecommunications (DECT) system, a pan-European standard designed to connect all of Europe with a common digital cordless system. DECT is also a flexible standard for providing a wide range of services in small cells.

In this paper, the SA639, a Philips low voltage FM/IF system, with several important features such as post filter amplifier and active data switch is proposed for DECT and other high speed digital wireless applications. A SA639-based DECT receiver evaluation board has been developed. Detailed description of the SA639 FM/IF system, structure of the evaluation board, design information, and experimental evaluation results are presented.

Review of the DECT Standard

DECT is designed as a flexible interface to provide cost-effective communications services to high user densities in small cells. This standard is intended for the applications such as domestic cordless telephony, Telepoint, cordless PBXs, and Radio Local Loop (RLL). It supports multiple bearer channels for speech and data transmission (which can be set up and release during a call), hand over, location registration, and paging. Functionally, DECT is closer to a cellular system than to a classical cordless telephone. However, the interface to PSTN or ISDN remains the same as for a PBX or corded telephone. Table 1 is a summary of the key specifications of DECT and other digital cordless telephone systems.

DECT is based on Time Division Duplex (TDD) and Time Division Multiple Access (TDMA) with 10 carriers in the 1880 - 1900 MHz band. Figure 1 illustrates the DECT TDD/TDMA frame structure. The completed frame is 10 ms in duration with 24 time slots. The first 12 slots are allocated for the transmission from base station to handsets, and the other 12 slots are for the transmission from hand sets to base station. Each slot is 417 μs long with 480 bits. The first 32 bits is a "1010..."

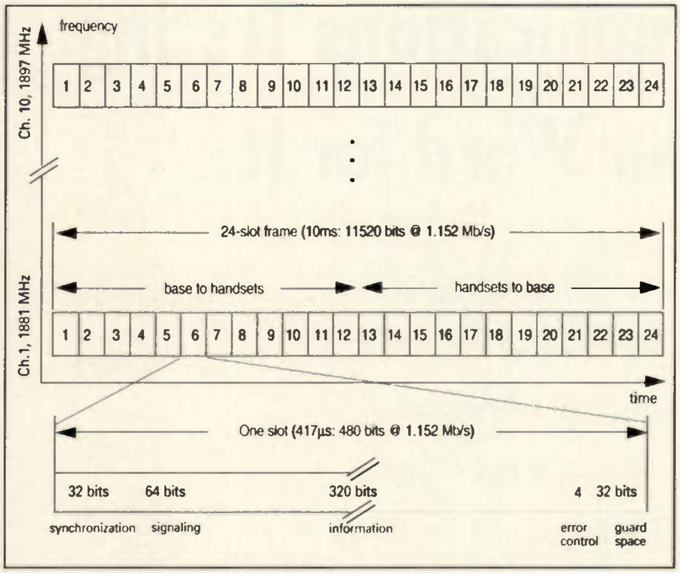

Figure 1. DECT TDD/TDMA frame structure.

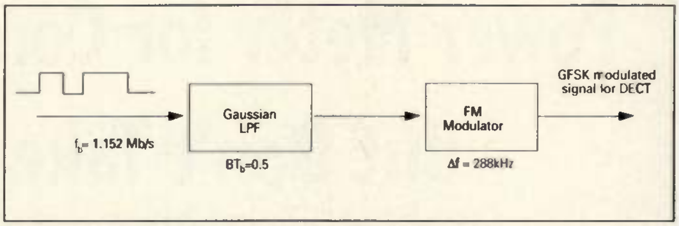

Figure 2. Block diagram of GFSK modulator.

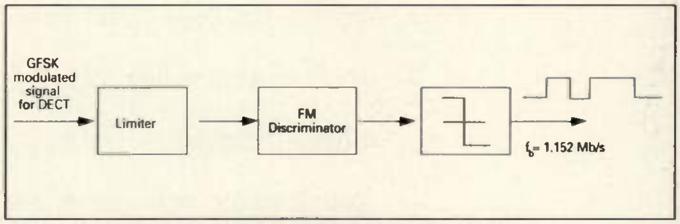

Figure 3. Block diagram of GFSK demodulator.

sequence for synchronization. The 32 kb/sADPCM CODEC is used for speech coding in DECT, which provides 320 bits during each 10 ms frame. When a call is made, two slots (oneis in the first 12 slots, the other is in the last 12 slots) are assigned to the user for transmit and receive.

Gaussian filtered FSK (GFSK) modulation scheme is employed in DECT. GFSK is a premodulation Gaussian filtered digital FM scheme. Figure 2 shows the block diagram of a GFSK modulator. The advantages of GFSK can be summarized as follows.

(i) Constant envelope nature: this allows GFSK modulated signal to be operated with class-C power amplifier without introducing spectrum regeneration. Therefore lower power consumption and higher power efficiency can be achieved.

(ii) Narrow power spectrum: narrow mainlobe and low spectral tails keep the adjacent channel interference to low levels and achieve higher spectral efficiency.

(iii) Non-coherent detection: GFSK modulated signal can be demodulated by the limiter/discriminator receiver as shown in Figure 3. This simple structure leads to low cost GFSK receivers.

The SA639 FM/IF System

The SA639 is a low-voltage high performance monolithic FM/IF system with high speed RSSI incorporating a mixer/oscillator, two limiting intermediate frequency amplifiers, quadrature detector, fast RSSI op amps, post detection filter amplifier, and a data switch. The block diagram of SA639 is presented in Figure 4.

The SA639 was designed specially for high data rate portable communications applications and will function down to 2.7 V. The data output provides a minimum bandwidth of 1 MHz to demodulate high speed data, such as in DECT applications. Figure 5 presents the quad tank S-curve of SA639, which indicates the linear range to be about 2 MHz. The measured RSSI characteristics of SA639 is presented in Figure 6. With more than 75 dB dynamic range, the SA639

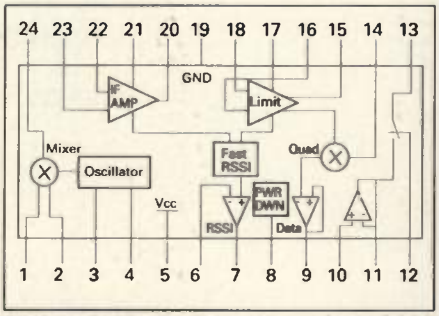

Figure 4. Block diagram of the SA639 FM/IF system.

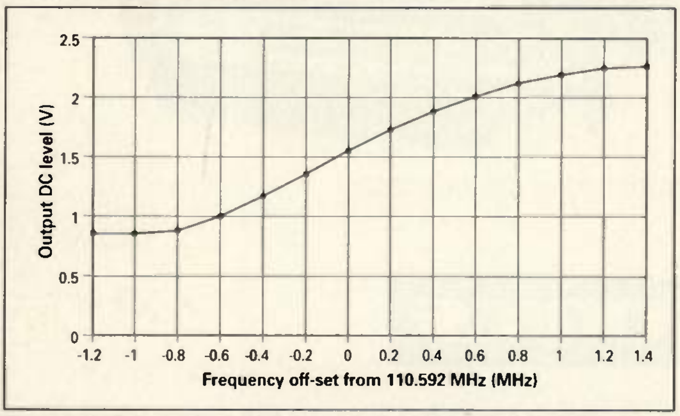

Figure 5. Quadrature tank S-curve for the SA639 board.

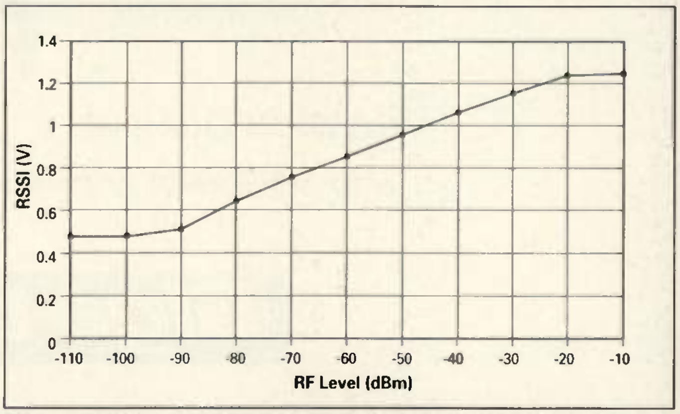

Figure 6. Measured RSSI characteristics of SA639.

RSSI rise/fall time is 0.8/2.0 μs at –45 dBm RF level.

The post-detection amplifier may be used to realize a group delay optimized low pass filter. The filter amplifier provides 0 dB gain and has a 3 dB bandwidth of at least 4 MHz in order to keep its frequency response influence on the filter group delay characteristics at a minimum. It can be configured for Sallen & Key low pass with Bessel characteristic and a 3 dB cutoff frequency of about 800 kHz.

The SA639 incorporates an active data switch to derive the data comparator reference voltage by means of routing a portion of data signal to an external integration circuit. The data switch is typically closed for 10 μs in the course of 32 bit synchronization sequence, and is open otherwise. The time constant of the external integration circuit is about 5 to 10 μs. This active switch provides excellent tracking behavior over a DC input range of 1.2 - 2.0 V. The slew rate is better than 1 V/ms. When the switch is opened, the output is in a tri-state mode with a leakage current of less than 100 nA. This reduces the discharge of the external integration circuit.

As compared to other similar FM/IF chips, another advantage of SA639 is that during power down mode (between data bursts) the data switch will output a reference of about 1.6 V to maintain a charge on the external RC circuit. This idea helps extract the reference voltage for the external capacitor in a shorter time and improves the accuracy of the voltage on the capacitor. The overall system is well suited for battery operated high quality products in digital wireless personal communications. Detailed specifications of SA639 can be found in [3].

Structure of the SA639 Evaluation Board

A SA639-based evaluation board has been developed based on DECT specifications. The structure of this board is illustrated in Figure 7 together with a VCO/FM discriminator based GFSK modem (modulator/demodulator). The demo board contains the entire demodulator as well as the Gaussian lowpass filter(LPF) for the modulator. The DECT modulated signal therefore can be generated either by a standard DECT signal generator, or by sending a 1.152 Mb/s data stream to the onboard Gaussian LPF (BTb = 0.5), then applying the filtered base band wave form to a FM signal generator with a modulation index of 0.5. The output is then the GFSK modulated signal (DECT). Baseband eye-diagram at the output of the Gaussian LPF is presented in Figure 8.

At the output of the limit/frequency discriminator, the post-detection amplifier is configured as a Sallen & Key LPF to eliminate noise. For the convenience of operation, the evaluation board is designed in such a way that the reference voltage for the data comparator can be obtained either from the switch controlled DC extraction circuit, or directly from the power supply. If the DECT Burst Mode Control circuit is available, the active data switch can be used to extract and

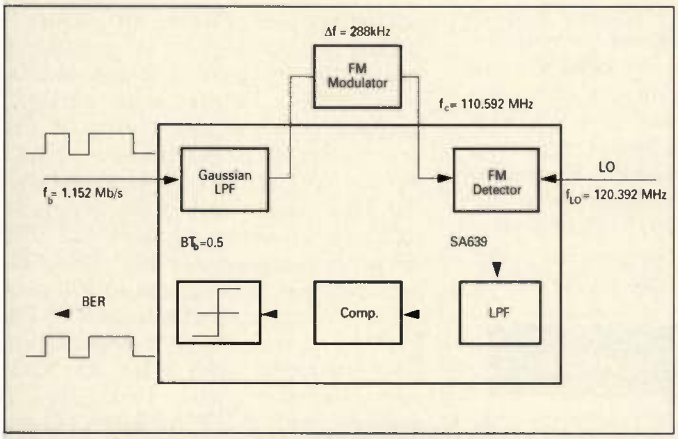

Figure 7. Structure of the SA639 GFSK evaluation board.

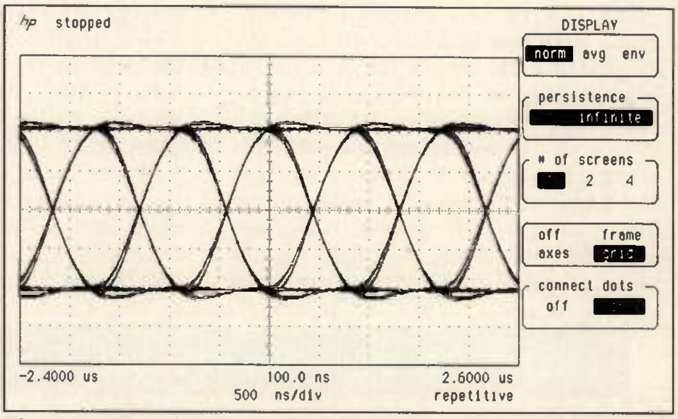

Figure 8. Measured eye-diagram at the output of the transmitter Gaussian LPF.

track DC level during the synchronization sequence. Otherwise the DC reference can be obtained from the power supply and manually adjusted for the comparator operation.

A 2-level threshold detector with sampling time adjustment circuit is implemented on the board for data regeneration. The phase of the data clock can be adjusted manually through a monostable multivibrator (74HC123) to achieve the optimal sampling time. The demo board is initially adjusted for a bit rate of 1.152 Mb/s. If a different data rate is used, the sampling time has to be re-adjusted. The output of the threshold detector is the regenerated binary data, which can be sent to a data error analyzer to evaluate the BER performance.

The symbol timing recovery (STR) circuit is not implemented on this evaluation board. Transmit data clock either hard-wire connected from the transmitter or from a separate STR circuit is required for the operation. The performance measurements presented in this paper were conducted with hard-wire connected data clock. However BER degradation caused by STR should not be more than 1 dB [6].

This SA639-based GFSK demo board is designed with DECT specifications at RF frequency of 110.592 MHz, LO frequency of 120.392 MHz, and intermediate frequency of 9.8 MHz. For different frequency plan applications, the step-by-step matching circuit design procedure can be found in [1].

Performance Evaluation

Performance of this SA639 based DECT GFSK system including BER and sensitivity to frequency off-set and FM deviation variation is experimentally evaluated. Measurement procedures and the measured results are presented in this section.

Figure 9 illustrates the measurement set-up with the SA639 DECT evaluation board. A data error analyzer is employed to generate a pseudo random binary sequence (PRBS) with length of 10^9-1 at a data rate of 1.152 Mb/s. This data sequence is sent to a DECT signal generator to generate a standard DECT modulated signal at 110.592 MHz. Another signal generator is employed to provide an LO signal at 120.392 MHz for the FM/IF system detection. The reference DC voltage for the data comparator is obtained from power supply for this evaluation. Data clock signal is directly from the data error analyzer. The sampling time is manually adjusted at the center of baseband eye diagram. Recovered data sequence is fed back to the Data Error Analyzer for BER measurement.

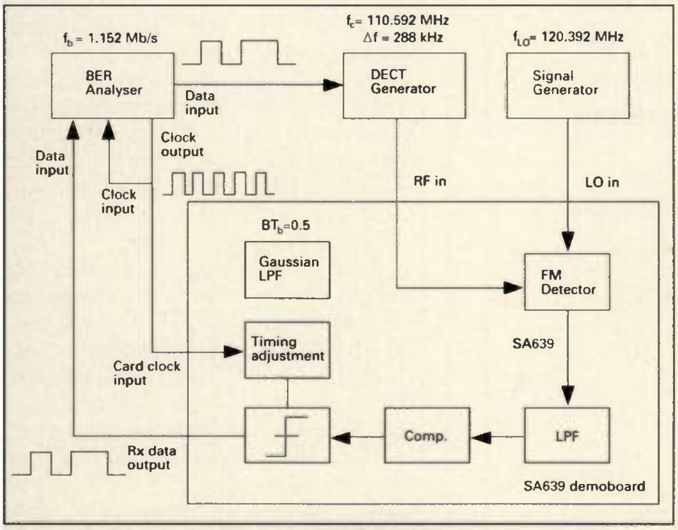

Figure 9. Block diagram of the BER evaluation set-up.

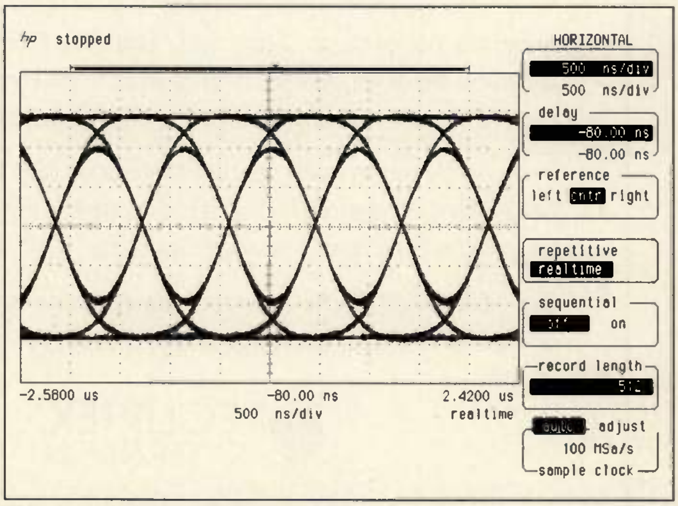

Figure 10. Recovered eye-diagram at the output of the SA639.

The recovered baseband eye-diagram is shown in Figure 10, and the measured BER versus RF input level is presented in Figure 11. It can be seen that less than –83 dBm RF power is needed to achieve the bit error rate of 10^{-3}. Since a typical front-end circuit has a better noise figure than FM/IF system, it is common to achieve more than 5 dB signal-to-noise ratio gain by the front-end circuit. Therefore, with the SA639 FM/IF the overall system sensitivity could be better than -88 dBm for the BER of 10^{-3}. Based on our measurements, by applying the Philips UAA2077AM 2 GHz image rejecting front-end to the SA639 FM/IF system the overall system sensitivity is –91 dBm for the BER of 10^{-3}. This performance compares very well to the DECT specifications for public access equipment (–86 dBm for 10^{-3} BER).

The performance degradatio caused by frequency off-set and th sensitivity to FM deviation variatio of this system are also evaluated. Figure 12 presents the measured BE versus frequency offset. Even wit (50 kHz offset, only minor degradation can be observed, and –82 dBn RF level is enough for 10^{-3} BER. Th sensitivity of this system to FM deviation variation is illustrated in Figure 13. Even with 10 percent deviation reduction (259 kHz), less tha –82 dBm RF signal is needed t achieve the BER of 10^{-3}. Thes results indicate that the Philips SA639 FM/IF system provides superior performance for DECT and othe high data rate GFSK applications.

About the Authors

Yanpeng Guo received his Ph.D. and MS degrees in Electrical Engineering from the University of California, Davis in 1994 and 1990, respectively. Since, 1994 he has been with the Wireless Communications Product Group at Philips Semiconductors. He has been doing research and development in digital cellular and wireless communications, especially in power and spectrally efficient modem/radio techniques, for more than six years and has published six IEEE Transaction and professional journal papers and more than 20 conference papers. Currently he is working on DS/FH spread spectrum techniques, RF/modem systems, and chip set development for PCS applications.

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Alvin K. Wong received the BSEE degree from San Jose State University in 1988. He has worked at Philips Semiconductors for seven years in the Wireless communications Product Group. He is the RF/Wireless Applications Manager.

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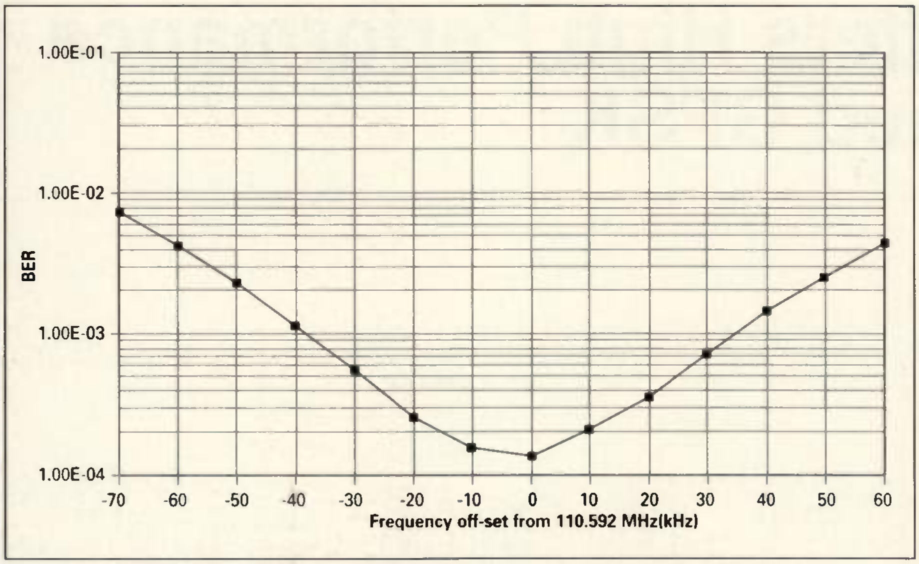

Figure 12. BER degradation caused by frequency offset (RF: –82 dBm).

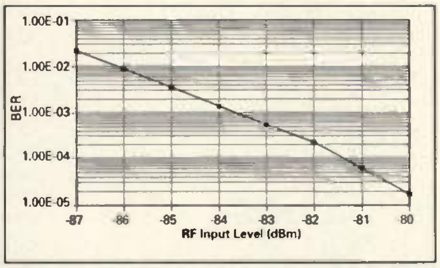

Figure 11. BER of the SA639 DECT demo board (RF: 110.592 MHz; LO: 120.392 MHz; fb: 1.152 Mb/s)

Conclusions

A Philips low voltage high performance FM/IF system (SA639) based GFSK modem evaluation board is presented. Experimental performance evaluation including bit error rate (BER), sensitivity to frequency off-set, and sensitivity to FM deviation variation of this system has been conducted based on DECT specifications. Results indicate that a superior performance can be achieved with the Philips FM/IF systems for high speed digital wireless applications. *RF*

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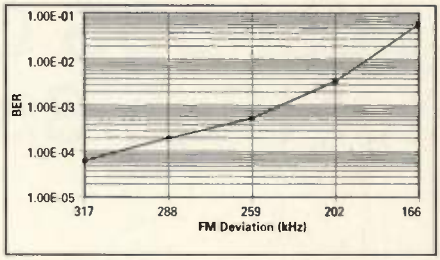

Figure 13. BER vs. FM deviation (RF: 110.592 MHz; –82 dBm; fb: 1.152 Mb/s).

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6. K. Feher, *Digital Communications, Satellite/Earth Station Engineering*, Prentice Hall, 1983.

Dual-DDS Offers High Performance for High-Speed GFSK

By Jonathan King
Qualcomm

This article begins with a brief tutorial of Direct Digital Synthesizers (DDS), followed by a functional description of QUALCOMM's new Q2368 dual DDS. Highlighted in this article is the use and application of the Q2368 in the Globalstar™(LP) low earth orbit satellite system.

March 1992 *RF Design* featured QUALCOMM's Q2220 K.i.S.S. (Keep it Simple Synthesizer) DDS. Although the technology was not new, it was considered exotic by many RF designers who felt comfortable using only analog synthesis techniques. Since then, direct digital synthesis has proliferated, including those "analog only" RF designers who appreciate the benefits of DDS in their synthesizer circuits. In fact, DDS are almost a required methodology in today's market-driven wireless systems.

Now, in 1996, the QUALCOMM Q2368 dual DDS offers improved capability and performance to better satisfy the demands of digital wireless communications and the exacting phase and frequency control required by high performance measurement systems.

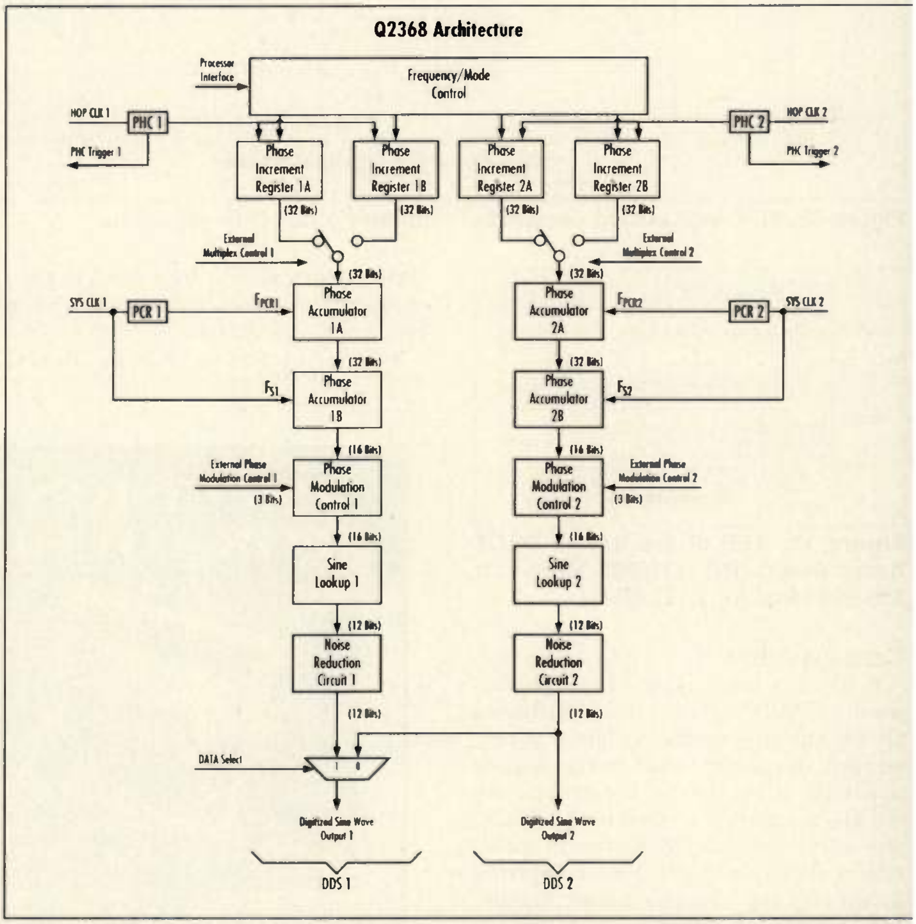

Figure 1. Functional block diagram of the Qualcomm Q2368 high-speed DDS. The two synthesizers on-chip can be operated independently with a clock speed of 60 MHz, or in tandem with an effective clock speed of 120 MHz.

Direct Digital Synthesizers in Practice

DDS can be practically defined as a means of generating highly accurate and harmonically pure digital representations of signals. This digital representation is then reconstructed with a high-speed Digital-to-Analog Converter to provide an analog output signal, typically a sinusoidal tone or sequence of tones. Many high performance DACs with greater clock speed and resolution capability are now available to obtain extremely low spurious from the DDS-DAC output. Since the DDS device output is digitally processed, the DDS functionality is easily software-configurable, making built-in utility more flexible.

DDS techniques offer unique capabilities in contrast to other synthesis methods. Although limited by the Nyquist criteria, (up to one-half the frequency of the applied clock reference), DDS allows frequency resolution control on the order of milli-hertz step size and can likewise allow milli-hertz or even nano-hertz of phase resolution control. Additionally, DDS imposes no settling time constraint for frequency changes other than what is required for digital control. This results in extremely fast frequency switching speeds, on the order of nanoseconds or a few microseconds. All frequency changes are automatically completed in a *phase continuous* fashion; that is, a change to a new frequency continues in-phase from the last point in the previous frequency. Since the signal being generated is in the digital domain, it can be manipulated with exceptional accuracy. This allows precise control of frequency or

phase and can readily accommodate frequency and phase modulation, i.e., FSK or PSK, for example. If desired, a microprocessor-controlled system can be utilized to store open loop compensation data vs. frequency for a particular device or system parameter. The microprocessor can then simply control the DDS to dynamically correct the frequency as needed. With a suitable frequency detector in a receive system, a closed loop system can easily be constructed.

Q2368 DDS Description

A functional block diagram of the Q2368 is shown in Figure 1. The device can be configured as a single high-speed DDS capable of operating at 120 MHz clock speed when set for doubling mode, or as two independent DDS devices each capable of operating at 60 MHz clock speed when set for dual DDS mode. Configuring the Q2368 for either mode is accomplished by a simple pin setting. The Q2368 provides 32-bit digital input resolution for both frequency and phase control.

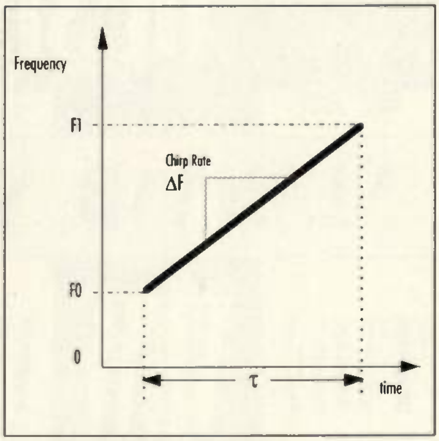

Figure 2. Linear chirp using programmable start and stop frequencies and chirp time.

This translates to ≤0.028 Hz minimum frequency step size depending on the frequency of the clock reference and 84 nano-degrees minimum resolution of phase control. The 12-bit output amplitude resolution also includes QUALCOMM's patented noise reduc-

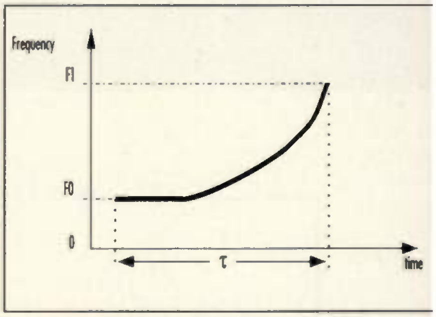

Figure 3. Piecewise linear chirp waveform generated with dynamic chirp-rate control.

tion circuit (NRC) for reducing discrete spurious levels while only slightly increasing the wideband noise floor.

All phase, frequency and operating modes are controlled via a single microprocessor interface with user options for 8-bit bus control or serial control. The control interface selection is accomplished by a simple pin setting; serial data output is provided to enable daisy-chaining of serial-controlled devices.

The Q2368's capabilities includes the following list of built-in functions:

- Programmable Hop Clock mode
- Programmable Hop Clock Trigger
- Chirp mode
- Programmable Chirp Rate
- Hold control
- Power-down mode
- Modulation control

User control of all functions apply identically to either single DDS (doubling mode) or dual DDS (dual mode) operation. A brief description of functions is offered here:

Programmable Hop Clock mode (PHC) – The Hop Clock command is used to activate new data information or enable the various operating modes. The Programmable Hop Clock mode takes the basic function of the Hop Clock one step further by introducing a programmable 32-bit duration counter derived from the DDS clock reference. This is used by the Q2368 as a built-in timer function which allows precise timed intervals of the Hop Clock command to be automatically and continuously re-asserted at the pre-programmed time intervals. This significantly eases the control burden and "housekeeping" function of the microprocessor.

Programmable Hop Clock Trigger – This feature provides an output pulse to simplify designing systems which are synchronously locked to the Q2368's clock. The Hop Clock trigger is an output pulse that is automatically generated each time the Programmable Hop Clock resets itself, according to the programmed time interval. A programmable 4-bit register is also included to adjust the Hop Clock trigger up to ±7 clock cycles of delay to compensate for any circuit path delay matching. This greatly facilitates precise triggering of other system events that require concurrent timing with DDS operation.

Chirp mode – The Chirp mode implements two phase accumulators serially to generate a linearly changing frequency output. In this mode, the output frequency changes at a constant rate, either increasing or decreasing the frequency direction. To generate a chirp waveform, a linear frequency change is added to the first phase accumulator to produce an instantaneous frequency, f. The first accumulator's output is then integrated by the second accumulator to produce an instantaneous phase, $\phi = \int f\,dt$.

Programmable Chirp Rate (PCR) – This feature provides for independent control of the chirp sweep rate and the output frequency. The limitations on the resolution of the sweep rate are due to the system clock frequency and the size of the phase accumulator. The chirp rate control allows the user to vary the slope of the chirp signal by changing the clock frequency used in the chirp mode. The chirp rate control introduces a programmable 20-bit counter derived from the DDS clock reference. This is used by the Q2368 to achieve <1 Hz/sec minimum chirp sweep rate over the entire clock speed range. If a linear sweep only is desired, the chirp output is described in terms of a start and stop frequency and the time (τ) interval for the signal to travel between the two frequencies as shown in Figure 2. Alternatively, if the desired output response cannot be implemented in one chirp waveform, the chirp rate control can be used to dynamically adjust the slope of the chirp response to achieve a near parabolic or piecewise-linear frequency output as shown in Figure 3.

Hold Control – The Q2368's Hold control function allows instant interruption of operating modes to produce a stationary (fixed) output. This is accomplished by an external pin enable. When operating in the chirp mode, enabling the Hold control will produce a constant output frequency corresponding to the exact position of where the chirp waveform was interrupted. Enabling the Hold control during any other DDS operation will result in a zero frequency (DC) output until the Hold signal is disabled. As expected, the Programmable Hop Clock mode is simultaneously interrupted when the Hold control is enabled. This allows the programmed time interval to be maintained irrespective of the Hold interruption time as illustrated in Figure 4. All DDS operations transition in a phase-continuous fashion when the Hold control is enabled or disabled making it attractive for many uses such as burst mode, auto-scan, or seek control.

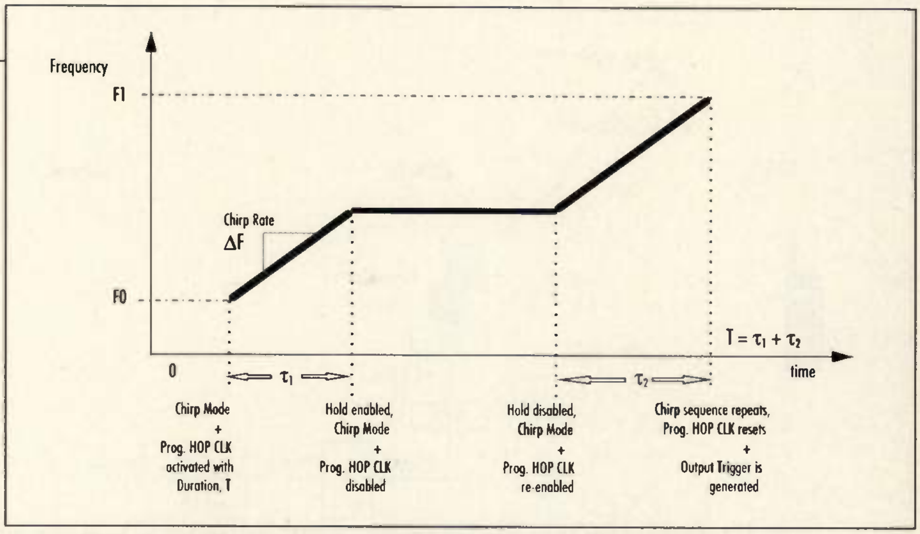

Figure 4. A chirp interrupted using the Q2368's hold control function.

Power-Down mode – This feature provides an independent power-down function for each DDS when operating in dual mode or a combined power-down function while operating the Q2368 as a single DDS (in doubling mode). To allow for power efficient standby operation, DDS current consumption is reduced to within the 0.1 to 10 mA range, depending on the clock frequency and operation mode. All data residing in the registers is retained during power down, although new information can still be addressed via the processor interface.

Modulation Control – Aside from the 32-bit control available for frequency and/or phase manipulations using the microprocessor interface, external FSK and PSK modulation inputs are also provided as convenient control ports. The Q2368 can generate BFSK modulation up to 15 Mbps using the external multiplex control to toggle between two pre-loaded frequencies. Using the 3-bit external phase modulation control, either BPSK, QPSK, or 8-PSK modulation can be generated up to 15 Mbps for continuous data transmission. Additionally, when operating in dual DDS mode, quadrature I and Q channels can be generated simply by using the external phase control to produce two signals with a 90° offset.

Device Parameters

The Q2368 DDS is specified over the industrial temperature range (–40° to +85°C) and operates from a single 5V ±10% supply voltage. Nominal power dissipation is 0.42 W per DDS at 60 MHz clock speed in dual mode and 0.8 W when operating as a single DDS at 120 MHz clock speed in doubling mode. The device is available in a 14mm × 14mm, 100-pin PQFP style package.

Globalstar LEO Satellite Description

The Globalstar LEO satellite system consists of a constellation of 48 satellites that employ CDMA technology to deliver low-cost global telephone and data services. The satellites orbit at a height of 876 mi., providing coverage areas as great as 3100 miles in diameter, compared to the 12 mile typical range accommodated by terrestrial analog cellular systems. User terminals include CDMA hand sets similar to dual mode terrestrial cellular telephones, with communication up to the Globalstar satellites at L-band (1610 to 1626 MHz) and retransmitted to the Gateway using the S-band (2483.5 to 2500 MHz). The ground segment consists of Gateways which use up to four bi-directional C-band antennas for

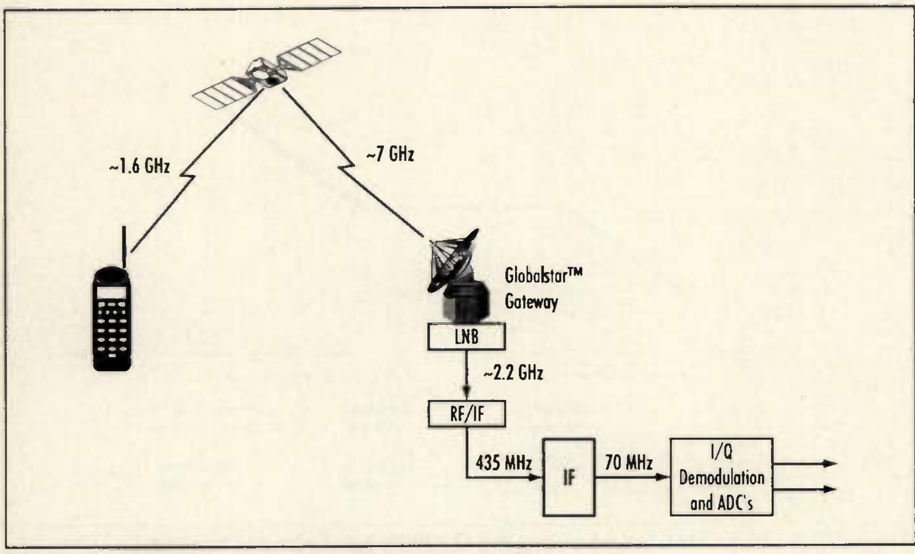

Figure 5. Globalstar reverse link frequency plan.

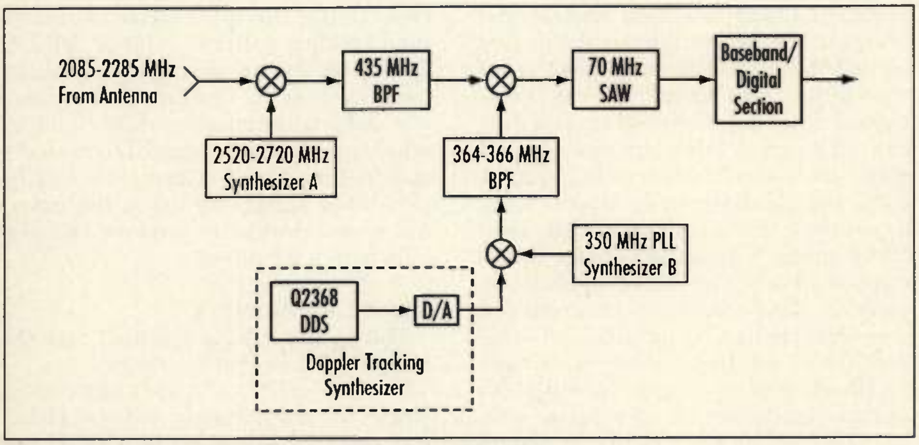

Figure 6. Gateway receiver card block diagram.

satellite communication. Within each coverage area, these gateways connect the satellite signals with the terrestrial telephone infrastructure.

Gateway Reverse Link

With regards to application with the Q2368 DDS, a parochial view of the Gateway reverse link will be used to highlight its utility in the Globalstar system. Figure 5 illustrates how the reverse link is configured. The received signals from the terminals in the 1.6 GHz range are upconverted within the satellite to the 7 GHz range for re-transmission to the Gateways.

In the Gateway, the Gateway Receiver Card (GREC) downconverts and digitizes the re-transmitted signal (IF) for the Transceiver Subsystem reverse link. The GREC performs the following functions:

- Selection of the C-band sub-beam to be demodulated
- Downconversion from the interfacility IF to baseband
- I and Q sampling and demodulation of the received baseband signal.
- Distribution of the digital samples for controller post processing
- Automatic gain control
- Full C-band Doppler correction and partial L-band Doppler correction
- Matched filtering by means of a SAW filter
- Generation of all local oscillators for the downconversion

The GREC contains two independent receivers, each with the ability to downconvert and sample a single subbeam. Figure 6 shows the conceptual block diagram for a single receiver path. The RF section translates the interfacility IF (2085 MHz-2285 MHz) to the final IF (70 MHz). Synthesizer A produces 2520 MHz to 2720 MHz in 2 MHz steps to tune the C-band downlink center frequency to the high side of the 435 MHz IF. Synthesizer B produces a 350 MHz signal which is used to generate the offset for the Doppler tracking synthesizer. The Q2368's digital output is reconstructed with a high speed, 12-bit D/A Converter (Harris HI5731) for optimal spurious-free-dynamic-range at the Doppler tracking synthesizer's analog output. The output of this mixer is filtered to reject the 350 MHz LO feedthrough and the lower sideband. The resulting 365

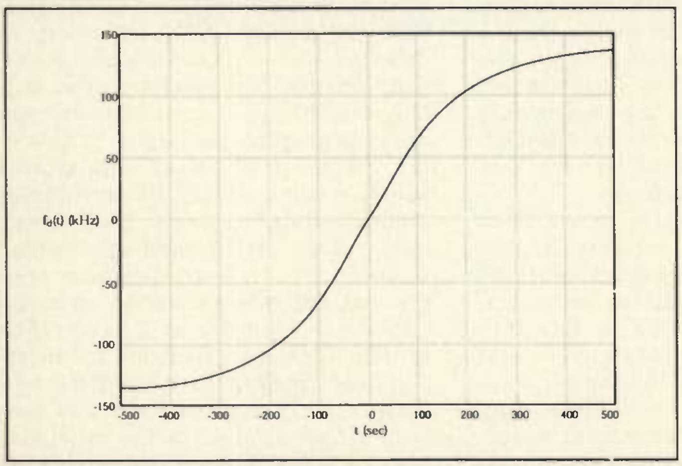

Figure 7. Magnitude of Doppler frequency versus time.

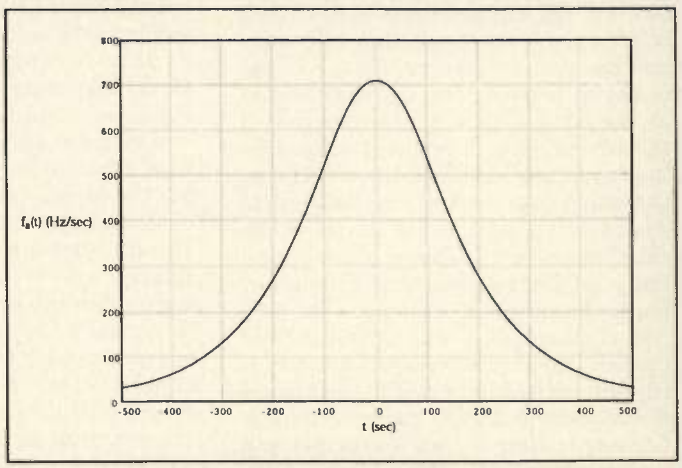

Figure 8. Doppler frequency rate of change versus time.

MHz LO tunes the 435 MHz IF signal to exactly 70.0 MHz. This architecture provides fine frequency resolution (0.01 Hz) with very fast tuning and low spurious content. The 70 MHz SAW filter rejects adjacent interference and functions as a pseudo-matched filter. The SAW nearly matches the spectral shape of the reverse link CDMA waveform. The baseband and digital section produces samples of the in-phase (I) and quadrature (Q) baseband waveforms and then distributes the digital samples on a single serial multiplexed data stream for post processing by a controller unit.

Doppler Effect in Gateway Reception

Globalstar LEO satellites orbit the earth approximately every 114 minutes and are therefore available to a user for up to 14 minutes. This transitory cycle requires seamless hand-offs between satellites and between antenna beams within a single satellite to maintain continuous phone calls. Doppler shift is calculated by the Gateway based upon the received signal from a Globalstar satellite moving along its orbital trajectory. The rate at which the correction must be performed for a given maximum error in the received frequency is considered in the Appendix.

The worse case scenario for the Doppler shift is when the Globalstar satellite is passing directly overhead (see Figure 7). In this figure the Doppler frequency, f (t), is expressed in kilohertz and is plotted versus time in seconds as the satellite passes from horizon to opposite horizon. Note that the frequency offset is an odd function of time, where t = 0 corresponds to the satellite passing directly above the Gateway. A negative frequency means that the satellite is moving towards the Gateway while a positive one means that the satellite is moving away. As an example, the center of the Gateway downlink band is 7075 MHz. The Doppler shift for this frequency as the satellite nears the horizon is about

$$(7075 \text{ MHz})(6 \text{ km/sec}) \div (300{,}000 \text{ km/sec}) = 141.5 \text{ kHz.}$$

The rate at which the Doppler shift changes is directly proportional to the acceleration of the satellite as shown in Figure 8. In this figure, the Doppler frequency change rate, $f_a(t)$, is maxi-

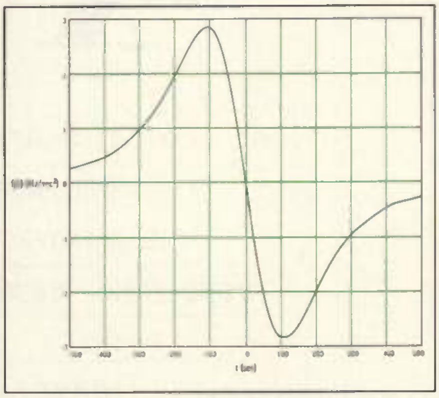

Figure 9. Doppler frequency jerk versus time.

mum as the satellite passes overhead and is an even function of time. As an example, at time t=0 when the Doppler shift vanishes, $f_a(t)$ is about

$$(7075\ \text{MHz})(30.036\ \text{m/sec}^2) \div (300000\ \text{km/sec}) = 708\ \text{Hz/sec}.$$

The third derivative of relative satellite position, sometimes referred to in classical mechanics as jerk, is denoted as $j_s(t)$. The rate at which $f_a(t)$ is changing is directly proportional to the jerk of the satellite and is shown in Figure 9. In this Figure, $f_j(t)$ indicates the Doppler frequency jerk in Hz/sec^2 and is an odd function of time.

These functions ($f_d(t)$, $f_a(t)$, and $f_j(t)$) have a direct bearing on the rate at which correction information must be communicated within the gateway transceiver subsystem reverse link.

Doppler Tracking Method

C-band instantaneous Doppler signature is a continuous function of time unique to a particular satellite pass and sub-beam. In addition, the mean L-band Doppler is added to the C-band correction to accommodate the uncertainty due to L-band Doppler. In order to reduce the correction update rate, the C-band Doppler is approximated by a piecewise-linear polynomial. The slope of the Doppler correction and the initial correction are sent to the GREC once every spline interval. A linear spline is simply a piecewise linear approximation to a given function as shown in Figure 10.

The Doppler shift is corrected by the GREC to both reduce the stress on the Gateway demodulator frequency tracking loop and to avoid increasing the 70MHz SAW noise bandwidth. The GREC Doppler correction is performed by a Q2368 DDS offset synthesizer operating in Chirp mode. This mode produces a linearly varying frequency output by pre-accumulating Accumulator 2 with Accumulator 1. Fine slope resolution is provided through control of the Q2368's PCR clock, FPCR . The synthesizer output frequency is a stair-step approximation to the C-band piecewise linear approximation of the Doppler as shown in Figure 11. At every clock period of Accumulator 1, the output frequency is increased by an amount equal to:

$$\Delta s = \text{PIRA}\ (F_S/2^N)$$

where F_S is the system clock, PIRA is the value in Phase Increment Register A which controls the amount of output frequency change with each Accumulator 1 clock cycle, and N is the number of bits in Accumulator 2.

The slope of the output is equal to:

$$\Delta F = \text{PIRA}\ (F_S/2^N \times F_{PCR})$$

The smallest non-zero slope for a constant clock period is obtained when PIRA=1, resulting in a minimum slope resolution of:

$$\Delta F_{min} = (F_S/2^N)(F_{PCR})$$
$$= (0.01\ \text{Hz})(F_{PCR})$$

The system clock is 43.75 MHz as dictated by the GREC frequency plan, hence the frequency slope can be programmed with a resolution of 0.01 Hz. The amount of error caused by the slope resolution is calculated below: (The unprimed symbols refer to the desired frequency versus time curve, and the primed symbols refer to the output of the synthesizer.)

$$F1 = F0 + (\Delta F)(\Delta t)$$
$$F1' = F0' + (\Delta F)(\Delta t)$$
$$F1 - F1' = F0 - F0' + (\Delta F - \Delta F')(\Delta t)$$

For the case when the starting point is correct (F0=F0′), the maximum error at the end of a Δt second interval occurs when the difference in the slopes is equal to half the slope resolution:

$$\text{Slope Error}_{max} = (1/2)(F_S/2^N)(F_{PCR})(\Delta t)$$

The error due to the slope resolution assumes that F_{PCR} is fixed for all spline intervals. If F_{PCR} is optimized for each spline interval to match the C-band Doppler slope produced by the piecewise-linear approximation, the

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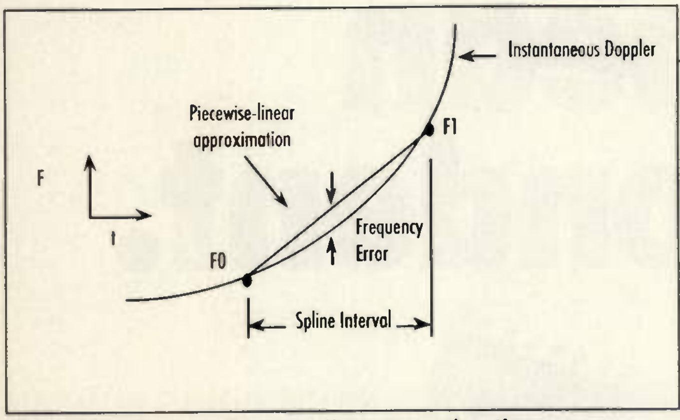


Figure 10. Doppler linear spline approximation.

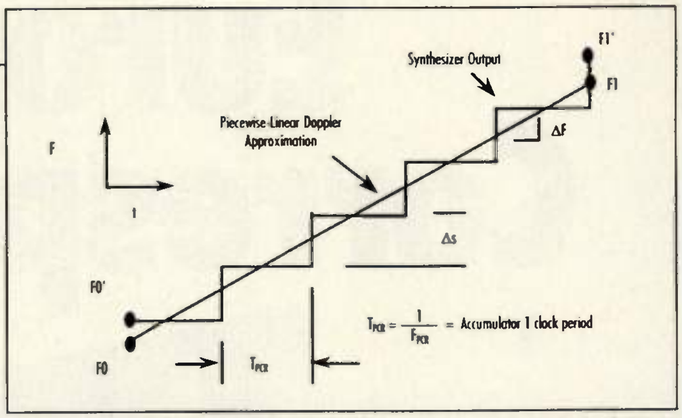


Figure 11. DDS stair-step approximation of Doppler shift.

error is reduced.

Sawtooth error is the maximum difference between the desired instantaneous frequency and the output of the synthesizer. The sawtooth error is equal to one half of the step size of the output. For a frequency change of F1–F0 in a time Δt, the error is:

$$\text{Sawtooth Error} = (1/2)(F1-F0)/[(F_{PCR})(\Delta t)]$$

The maximum frequency change however, is related to the maximum slope of the Doppler and the time interval:

$$F1-F0 \approx (\Delta t)(f_a(t)_{max})$$

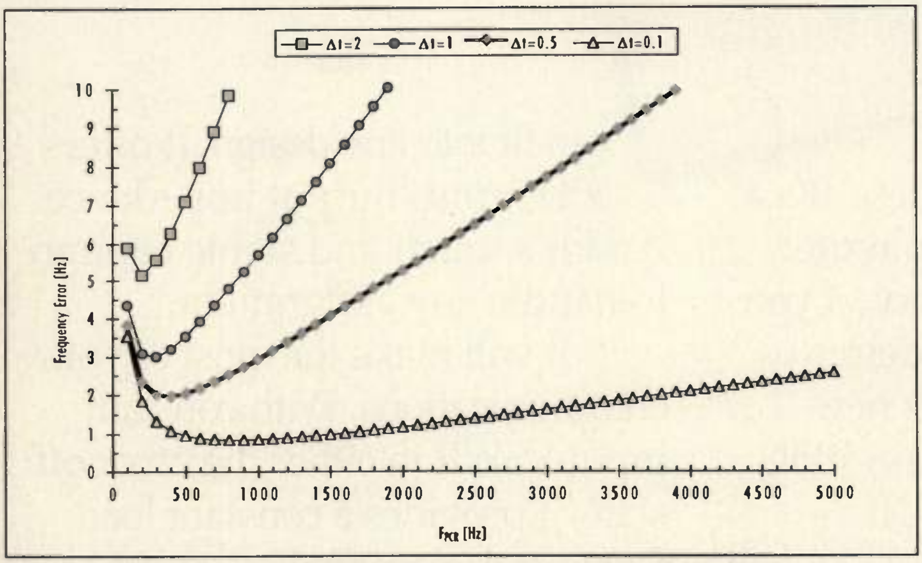


Figure 12. Maximum frequency error versus accumulator 1 clock rate.

Since the maximum satellite acceleration is less than 0.03 km/sec^2, the maximum rate of change of the Doppler shift is about 0.1 ppm/sec.

The resulting sawtooth error is:

$$\text{Sawtooth Error}_{max} = (1/2)[(f_c)(0.1)/F_{PCR}]$$

where f_c is the C-band reverse link frequency in MHz.

Q2368 Doppler Tracking Implementation

The Q2368 can be programmed to track the C-band Doppler in a number of ways. The one described here involves updating the value of PIRA every 500 msec. The period of F_{PCR} is fixed for the entire satellite pass. The total error in the frequency is the sum of the error due to both the slope resolution and the sawtooth approximation. Maximum frequency error versus F_{PCR} is shown in Figure 12.

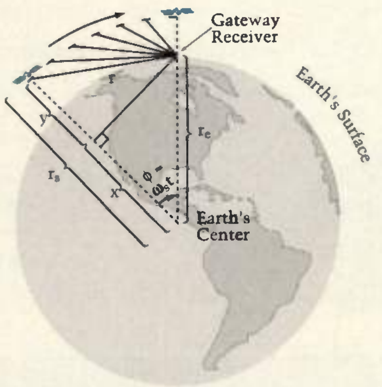
The F_{PCR} clock for Accumulator 1 is determined by the relation:

$$F_{PCR} = F_S/P = 43.75\ \text{MHz}/P$$

where P is the preset value of the 20-bit counter. The range of F_{PCR} is from 43.75 MHz to 41.72 Hz, therefore the ΔF slope resolution can range from 437.5 kHz/sec to 0.42 Hz/sec by changing the value of P. The update interval is chosen to be 500 msec to strike a balance between tolerable frequency error and software complexity. The Programmable Hop Clock mode is utilized to provide the continuous auto-sequencing of the 500 msec update. The 500 msec time interval is produced by programming the 32-bit PHC counter with a divide value of 21,875,000. F_{PCR} is selected to be 400 Hz to minimize frequency error with the chosen 500 msec update interval and is produced by programming P with a divide value of 109,375. Exactly

Doppler Calculations

Given Constants	
Time interval	t = -500 sec, -499 sec...500 sec
Speed of Light	$c = 2.998 \cdot 10^8 \frac{m}{sec}$
Radius of the Earth	$r_e = 6378 \cdot 10^3$ m
Radius to the Satellite	$r_s = 7778 \cdot 10^3$ m
Angular Velocity of the Satellite	$\omega_s = 0.00092068613 \frac{rad}{sec}$
Maximum Reverse Link Frequency	$f_c = 7075$ MHz



$$\text{Given } x = r_e \cos \omega_s t$$
$$y = r_s - r_e \cos \omega_s t$$
$$r(t) = [(r_s - r_e \cos \omega_s t)^2 + (r_e \sin \omega_s t)^2]^{1/2}$$
$$r^2(t) = r_s^2 + r_e^2 (\sin^2 \omega_s t + \cos^2 \omega_s t) - 2\, r_s\, r_e \cos \omega_s t$$

Doppler Effect Parameters	
Distance of the Satellite to the Gateway	$r(t) = \sqrt{r_s^2 + r_e^2 - 2r_e\, r_s \cos(\omega_s t)}$
Velocity of the Satellite	$v_s(t) = \frac{\omega_s\, r_e\, r_s \sin(\omega_s t)}{\sqrt{r_s^2 + r_e^2 - 2r_e\, r_s \cos(\omega_s t)}}$
Acceleration of the Satellite	$a_s(t) = \omega_s^2\, r_e\, r_s \cdot \frac{(\cos(\omega_s t)\, r_s^2 + \cos(\omega_s t)\, r_e^2 - r_e\, r_s \cos(\omega_s t)^2 - r_e\, r_s)}{(r_s^2 + r_e^2 - 2r_e\, r_s \cos(\omega_s t))^{3/2}}$
Doppler Frequency	$f_d(t) = v_s(t) \cdot \frac{f_c}{c}$
Doppler Frequency Change Rate	$f_a(t) = a_s(t) \cdot \frac{f_c}{c}$
Satellite Acceleration Change Rate (jerk)	$j_s(t) = -\omega_s^2\, r_e\, r_s \sin(\omega_s t) \cdot \frac{(r_s^4 - r_e^2 r_s^2 - r_e r_s^3 \cos(\omega_s t) + r_e^4 - r_e^3 r_s \cos(\omega_s t) + r_e^2 r_s \cos(\omega_s t)^2)}{(r_s^2 + r_e^2 - 2r_e\, r_s \cos(\omega_s t))^{5/2}}$
Doppler Frequency jerk	$f_j(t) = j_s(t) \cdot \frac{f_c}{c}$

Appendix. Summary of Doppler calculations.

200 cycles of the F_{PCR} occur during each spline interval. The frequency slope resolution is 4 Hz/sec and the tracking error is bounded at 2 Hz with the selected implementation.

The Gateway Transceiver Subsystem software updates the DDS slope value every spline interval. In order to maintain phase continuity, the initial frequency is not updated every interval. Because of the limited slope resolution of the Q2368, there may exist a residual frequency error as large as 1 Hz which carries over into the next spline interval. In order to keep this error from growing, the software selects the slope for interval j that minimizes the instantaneous frequency error at the end of interval j based on the following criteria:

(a) residual error of interval j–1
(b) desired slope for interval j
(c) under the constraint that the frequency slope resolution is 4 Hz/sec.

The following example will be used to clarify this. The desired slope for interval #1 is 18 Hz/sec and the initial frequency is 100 Hz. The Q2368 is programmed to generate a slope of 20 Hz/sec from an initial 100 Hz frequency. At the end of the first 500msec interval, the DDS output is 110 Hz, or 1 Hz too high. The desired slope for interval #2 is 19 Hz/sec. The software has the option of generating a slope of 16 Hz/sec or 20 Hz/sec. In order to reduce the error at the end of interval #2, the Q2368 is programmed to produce a slope of 16 Hz/sec. At the end of interval #2, the DDS output is less than the desired output by 0.5 Hz.

Conclusion

The Q2368 dual DDS offers a high-utility design solution for frequency synthesis in digital wireless communications. In the Globalstar LEOSAT system, a method has been described for using the Q2368 DDS to correct the Gateway reception for the Doppler effect. Since the rate of the Doppler shift varies numerically more slowly as a function of time than the Doppler shift itself, the time intervals between updates on the rate of change of Doppler can be much greater, for the same error in transmitted frequency than the time intervals required between updates of the absolute Doppler shift. This is equivalent to approximating the Doppler frequency as a function of time with a piecewise-linear polynomial. Therefore, the CPU controlling the Q2368 needs to provide only the rate of change of frequency as opposed to a new Doppler frequency correction for each time instant.

For more information on the Q2368 DDS, please contact QUALCOMM Incorporated at 6455 Lusk Boulevard, San Diego, CA 92121-2779. The telephone number is (619) 658-5005, and the fax is: (619) 658-1556; send email to: asic-products@qualcomm.com, or circle Info / Card #250. **RF**

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About the Author

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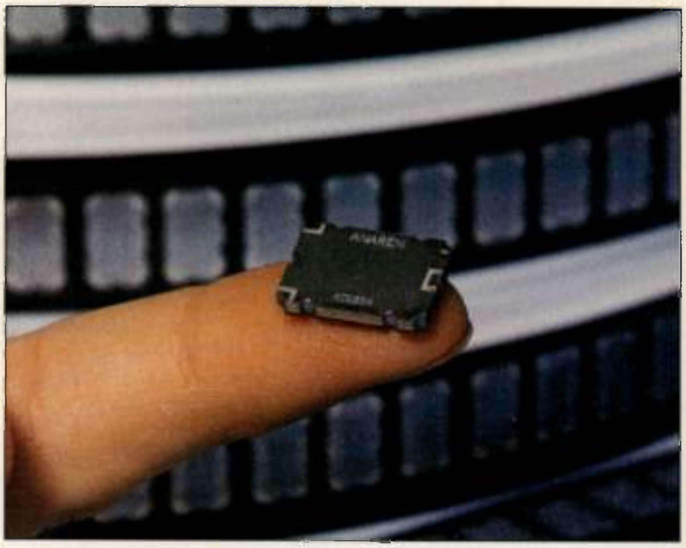
Surface-Mount Power Dividers Cover 660-2000 MHz

A line of 2-way and 3-way in-phase power dividers from Anaren are now available for many wireless applications. Four off-the-shelf models are offered that cover the following frequency bands: 660-1180 MHz, 800-1000 MHz, 1400-2300 MHz and 1750-2000 MHz. These surface-mount power dividers offer typical specifications that include 16 dB isolation, 0.35 dB insertion loss, and VSWR of 1.30:1 (input) and 1:40:1 (output). Additional performance specifications are 0.2 dB amplitude balance and 2 degrees phase balance. Maximum input power handling is 5 watts average or CW when terminated with a 1.2:1 or better VSWR.

The two surface mount package styles used by the devices have footprint areas of 0.196 and 0.312 square inches, with edge-plated channels for reliable soldering and inspection. Prices are as low as $3.50 each in large quantity orders. Tape and reel packaging for automated assembly is available.

Anaren Microwave, Inc.
INFO/CARD #212



IF Filters in kHz Range for Wireless Applications

Toko America has introduced the LF series, a kHz-range ceramic filter for use in a variety of wireless communication applications. The devices are available in either 4- or 6-element types, connected in a ladder form. Provided in 450 or 455 kHz center frequencies, the filters offer 6 dB bandwidths of 7 to 31 kHz with 3.0 dB maximum insertion loss. The 4-element filters carry the

LFC/LFCM model numbers, with 6-element filters designated LFY/LFYM. The filters are optionally available with 460 or 468 kHz center frequencies, and with group delay variations specified. Standard 6 dB bandwidths are ±3, ±4.5, ±6.0, ±7.5, ±10, ±12.5 and ±15 kHz. Center frequency tolerance is available as either ±1 kHz of ±2 kHz.

Toko America, Inc.
INFO/CARD #213

10 MHz OCXO Introduced as Standard Product

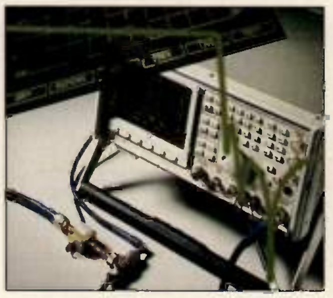
Micro Crystal introduces their new 10 MHz oven-controlled crystal oscillator (OCXO) as a standard, off-the-shelf design to reduce the user's development time. In the past, OCXOs have typically been custom orders, with attendant time delays, high cost and engineering fees. The

new OCXO has frequency stability specified at ±0.2 ppm over a temperature range of –20 to +70°C, and operates from a +12 VDC supply. Remote frequency setting is included, which does not require manual adjustment. The OCXO is provided in a 14-pin DIL package, and features very fast warm up time and needs only 20 mA typical for operation at 25°C. Applications include base stations, GPS, high-performance hand-held units, RF data collection and medical communications. It can be used to replace high-end TCXOs in many applications, as well.

Micro Crystal
INFO/CARD #214

Spectrum Analyzers Offer Value for Bench and Field

The new 2390 Series of general-purpose spectrum analyzers from Marconi Instruments are introduced to complement existing products, including the 6200B Series. The 2390 series comprises three models: The 2392 for RF applications in the 9 kHz to 2.9 GHz range, the 2390 for measurements from 9 kHz to 22 GHz, and the 2393 for extended microwave coverage to 26.5

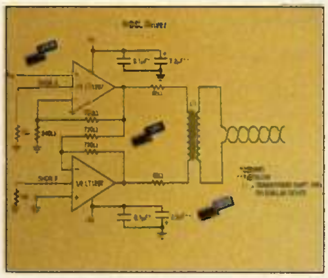


GHz. All three instruments include built-in AM and FM receivers 1 Hz resolution frequency counter for easy identification of interfering signals. Resolution bandwidths from 3 Hz to 30 MHz allow a wide variety of signals to be examined. An optional quasi-peak detector and filters to CISPR specifications allow the 2390 instruments to be used for EMC pre-compliance testing. RS-232 and IEE-488.2, SCPI interfaces are standard.

Marconi Instruments, Inc.
INFO/CARD #215

Current Feedback Amplifier for Video, Telecom

Linear Technology announces the LT1207, a dual 250 mA, 60 MHz current feedback amplifier targeted at video and telecom applications. The device has excellent video characteristics of 900 V/µs slew rate, 0.02% differential gain and 0.17° differential phase (typical). The LT1207 includes a pin for an optional



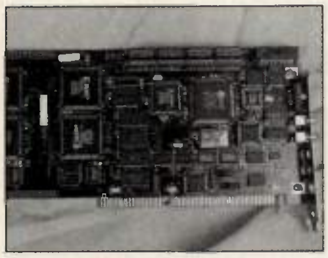
compensation network for heavy capacitive loads. Each amplifier has thermal and current limiting circuitry. Operation is specified from ±5 to ±15 volt supplies, with a current draw of 20 mA per amplifier, typical, adjustable with an external resistor. These features allow the device to drive cables in video or digital communications systems. The LT1207 is provided in a SO-16 package with flow-through pinout to simplify RF circuit board layout. Pricing in 1000+ quantity is $6.90.

Linear Technology Corp.
INFO/CARD #216

SUBSYSTEMS

Burst PSK Demod Bit Sync

Sigtek introduces the ST-105 Burst Demod Bit Sync for satellite, point-to-point microwave



and cable applications. The PC/AT compatible unit employs digital IF processing at 60 megasamples/second to demodulate TDM, TDMA, and SCPC signal with data rates from 2400 bps to 4.8 Mbps. BPSK, QPSK, π/4 DQPSK or OQPSK burst signals can be acquired with little or no preamble. The ST-105 handles input frequencies from 200 kHz to 22 MHz. The optional ST-512 IF translator allows the use of standard 70, 140 or 160 MHz IFs.
Sigtek Inc.
INFO/CARD #219

PCMCIA 2.4 GHz Antenna

TELECOM Industries announces the first shipment of a PCMCIA card antenna for wireless LAN applications. The low-profile antenna occupies just 1.0 × 2.0 × 0.2 inches, replacing the protruding external whip commonly used in these applications. Other versions can be designed for specific card configurations.
TELECOM Industries
INFO/CARD #220

Cellular Diversity Receive LNA

AML Communications announces additions to the Link Balancer™ product line, the R100-E and R100 diversity LNAs. Both operate in the 824-849 MHz cellular band and operate from 100 VAC or from optional DC power supplies. Gain is adjustable from –4 to +14 dB, with a noise figure of 1.0 dB for the R100-E and 3.0 dB for the R100. The units are available for rack mount or optional mast head installation.
AML Communications
INFO/CARD #221

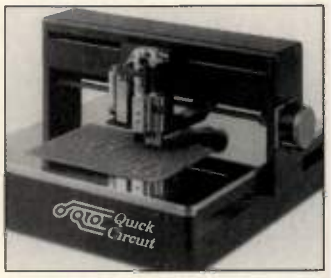
In-Line Switch Assemblies

Sierra Microwave Technology introduces an expanded line of In-Line Switches. The switches have been developed to cover a frequency range of 500 MHz to 18 GHz (wider range optional), with standard models having single- to five-throw configurations. Insertion loss ranges from 2.0 to 3.1 dB maximum, with 1.6 to 2.7 dB typical performance. Isolation of 60 dB minimum, 70 dB typical is standard. Both absorptive and reflective types are available.
Sierra MicroWave Technology
INFO/CARD #222

TOOLS, MATERIALS & MANUFACTURING

Smaller PC Board Prototype Maker

T-Tech introduces the Model 5000 circuit board prototyping product. The new model fea-



tures a smaller footprint than its predecessors, and is able to make double-sided printed circuit boards up to 9 × 12 inches. The unit accepts Gerber files from standard ECAD packages, or converts DXF files to Gerber using an optional translator. Model 500 offers 0.00025 inch resolution and 24,000 RPM spindle. Price is under $10,000.
T-Tech, Inc.
INFO/CARD #217

Thermoplastic PPO Laminates

NorCLAD Laminates have been introduce by Polyflon, using a thermoplastic PPO (polyphenylene oxide) dielectric. With a dielectric constant of 2.55, NorCLAD has uniform electrical properties and is stable dimensionally and over temperature. PPO has become a viable material as aqueous processing methods have replaced solvent-based chemicals. Pricing of the new laminate is 10 to 50% lower than existing material with comparable performance.
Polyflon Company
INFO/CARD #218

SIGNAL PROCESSING COMPONENTS

Custom Surface-Mount LC Filters

PTI offers surface-mount LC products designed for easy handling and manufacturing. Custom ESD carriers simplify handling, and the filters have been

developed to meet soldering requirements in large-scale production. Custom filter designs are available in wide range of types to meet customer needs.
Piezo Technology, Inc.
INFO/CARD #224

1087.5 MHz Surface-Mount Filter

KP Microwave Components introduces a 1087.5 MHz band-pass filter to its line. The maximum insertion loss is 2.0 dB in the passband, with a bandwidth of 237 MHz. Stopband rejection is 40 dBc at 769 and 1406 MHz. VSWR is 1.5:1 max.
KP Microwave Components
INFO/CARD #225

SAW Filters for IEEE 802.11 WLAN

RF Monolithics has introduced a new low-loss SAW filter for 2.4 GHz wireless LAN applications. The SF1067A has a center frequency of 350 MHz and is designed for IEEE 802.11 frequency-hopping spread spectrum systems. Maximum insertion loss is 10 dB, with a minimum 3 dB bandwidth of 1.05 MHz. Group delay variation across the 3 dB bandwidth is less than 250 ns_{p-p}. Ultimate rejection is greater than 50 dB.
RF Monolithics, Inc.
INFO/CARD #226

SIGNAL SOURCES

Glass SMT Crystals

ECLIPTEK Corporation announces the ECCM3 and ECCM4 crystals, offering high stability in a glass surface-mount package. The crystals are available in 11 MHz to 120 MHz frequencies, and are able to sustain ±10 ppm tolerance and ±5 ppm stability over –20 to +80°C. The ECCM3 and ECCM4 differ only in package type and pinout.
ECLIPTEK Corporation
INFO/CARD #227

Miniature OCXO

The 230-series from MTI–Milliren Technologies, Inc. uses both AT and SC-cut resonators to achieve performance usually found only in larger OCXOs. Stability is 2.5×10^{-8} over –30 to +70°C for models using SC-cut resonators. AT-cut performance

over the same range is 2.0×10^{-7}. The industry-standard package is 1.42 × 1.07 × 0.76 inches.
Milliren Technologies, Inc.
INFO/CARD #228

TCXO for Wireless Utility Applications

Networks International Corp. introduces a TCXO for the wireless utility market. The standard operating range is –30 to +60°C with stability of ±2.5 ppm. The frequency range is 9-25 MHz, with TTL or clipped sine wave outputs. Packages re compatible with 14-pin DIP sockets.
Networks International Corp.
INFO/CARD #229

Phase Locked Multipliers Offer Low Noise

The PLX700 Series of phase locked low-noise multipliers from Techtrol Cyclonetics provides output frequencies from 10 to 250 MHz, and will accept input references of 5 MHz or 10 MHz. Non-coherent spurious rejection is –90 dBc, with coherent spurious rejection of –60 dBc. Operating voltage is +15 VDC at 175 mA and –15 VDC at 5 mA; operating temperature is –20 to +60°C. The package is a 3.0 × 2.0 × 0.75 inch steel can.
Techtrol Cyclonetics, Inc.
INFO/CARD #230

SEMI-CONDUCTORS

Wireless Quadrature IF Transceiver

Analog Devices announces the AD6432 3-volt transceiver for quadrature-modulated wireless communication systems. The IC by itself can be used for low power systems up to 300 MHz, or for up- and downconversion for higher carrier frequencies. The receiver converts a signal up to 300 MHz to an IF of 10-30 MHz, with –20 to +60

Product Focus — Cables and Connectors

Reel Coax Packaging

Andrew Corporation an-nounces REEL PAX™, a new cable dispenser for easier handling of HELIAX® superflexible cable and braided cable. One carton holds a

500-foot reel of 1/4 or 3/8 inch HELIAX or braided cable, or a 250-foot reel of 1/2 inch HELIAX.
Andrew Corporation
INFO/CARD #233

TNC for Semi-Rigid Line

Coaxicom introduces a "pre-assembled" TNC bulkhead jack for RG402 (0.141 inch diameter) semi-rigid cable. Model 4539CC-430-1 is directly soldered to the cable jacket after stripping back the center contact and plugging it into the captive female contact. The connector is also accepted by ULTRAFLEX MDC8141 by MIDISCO and other types of reformable and rigid cable.
Coaxial Components Corp.
INFO/CARD #234

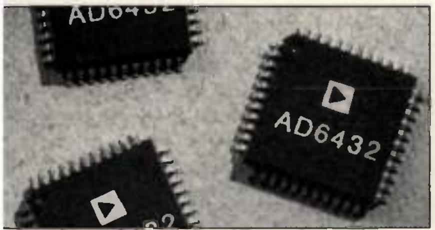
Two-Piece Signal and Power Connectors

A new two-piece DIN multipurpose PCB connector is introduced by Methode Electronics. Each piece has six power and 42 signal contacts. Signal contacts are positioned in three parallel rows of 14 contacts at 0.100 inch spacing. The header is offered in both DIP solder and press fit configurations to accommodate a 0.125 inch thick board. The receptacle has optional tail lengths for either 0.062 or 0.164 inch boards.
Methode Electronics
INFO/CARD #235

Solderless Connector for LMR-400

Times Microwave Systems introduces the first Type N crimp-style solderless connector for LMR-400 low-loss cable. Using the ST-400-EZ stripper and a standard 0.429 inch hex crimp tool, a connector can

be installed in minutes. The connector body is silver plated brass, the center contact is gold plated beryllium copper, with the center pin captivated in PTFE dielectric. The EZ-400-NMH connector is $8.50 and the ST-400-EZ tool is $50.00.
Times Microwave Systems
INFO/CARD #236



dB gain and linear-in-dB analog gain control. The transmitter include I-Q modulators accepting modulation bandwidths up to 1 MHz, –15 dBm output power, and a power control function. in 1000s, the AD6432 is $7.25.
Analog Devices, Inc.
INFO/CARD #231

10- and 12-Bit High- Speed A/D Converters

Burr-Brown offers the new ADS800 series of monolithic A/D converters. The devices are fabricated using a 0.6 μm process, and feature advanced pipeline techniques and +5 volt supply operation. The devices in the family are: ADS800, 12-bit, 40 MHz sampling; ADS801, 12-bit, 25 MHz sampling; ADS802, 12-bit, 10 MHz sampling; ADS820, 10-bit, 20 MHz sampling; and the ADS821, 10-bit, 40 MHz sampling. Pricing (1000s) ranges from $7.45 for the ADS820 to $35.00 for the ADS800.
Burr-Brown Corp.
INFO/CARD #237

Complete GSM Device Family

Lucent Technologies (formerly part of AT&T) has introduced the Sceptre™ semiconductor GSM hardware platform. The first part of the system is the DSP1618 with 24k of on-chip ROM for full capability of implementing the GSM physical layer. The DSP1628 is an improved voice compression device, scheduled for release in the 3rd quarter of 1996. The currently-available W2020 transceiver IC will be upgraded to support GSM offshoots, DCS1800 and PCS1900.
Lucent Technologies
INFO/CARD #238

IF Amplifier Designed for CDMA

A low-cost adjustable gain GaAs IF amplifier handles a wide range of signals, including CDMA, PCS and spread spectrum. The KGF2441 from Oki Semiconductor offers a gain/attenuation range of 90 dB, and operates over the frequency range of 70 to 250 MHz. The device operates from a single +5 VDC supply, with an input impedance of 1000 ohms and output impedance of 250 ohms. Pricing for 1000+ quantities is $4.45.
Oki Semiconductor
INFO/CARD #239

SPDT Switches for 100 MHz to 2.5 GHz

California Eastern Laboratories announces the availability of the NEC UPG133G, a GaAs T/R switch for WLAN, digital cordless phones, cellular and PCS applications. the device operates from 100 MHz to 2.5 GHz with 0.6 dB insertion loss (typical, at 2 GHz). Off isolation is 22 dB, and the UPG133G will handle signal levels up to 30 dBm. Operation is from a 3 volt supply, with the UPG132G version available for –3 volt operation. Pricing is from $1.00 in production quantities.
California Eastern Labs
INFO/CARD #240

TEST EQUIPMENT

Enhanced Digital Communications Test Sets

Tektronix announces major new capabil-

ities for the CMD57 and CMD80 Digital Communications Test Sets. The CMD57 is a self-contained test system for GSM, DCS1800 and PCS1900 base stations, now incorporating an in-service test capability that allows network operators to test without disconnecting from the network. The CMD, a complete CDMA test solution, now includes Rate Set 2 PCS (13K vocoder) data rates, enabling mobile phones to be tested for full parameters. The CMD57 is a product of Rohde & Schwarz, and the CMD80 is a joint development of Tektronix and Rohde & Schwarz. Both are marketed and supported in the U.S. and Canada by Tektronix.
Tektronix. Inc.
INFO/CARD #243

Generator Tests Multiple Formats

Anritsu-Wiltron announces the MG3660A, a digital modulation signal generator adaptable to changing standards. Modular architecture and 3 kHz to 2.75 GHz frequency range allow the MG3660 to accommodate new operating parameters. Modulation modules are currently available for: NADC, PHS, PDC, GSM, DECT, PACS, WCPE, TETRA, CT2, and PCS1800. Price of the instrument is $29,950.
Anritsu Wiltron
INFO/CARD #244

Field Strength Meter

IC Engineering introduces the DIGI-FIELD digital display field strength meter with a frequency response of DC to 12 GHz. The unit is useful for measuring antenna patterns, making RFI/EMI measurements, and other measuring and monitoring applications. The new Model "C" combines the sensitivities of previous units: 150 nanowatts at 100 MHz, and 2 nanowatts. Price is $229.00.
IC Engineering
INFO/CARD #245

EMC/ESD PRODUCTS

Extrusion Grade Composites for ESD

LNP Engineering Plastics now provides custom formulated conductive, static-dissipative and anti-static thermoplastic composites for the extrusion market. Base resins available include polycarbonate, ABS and poly propylene. The conductive filler can be carbon powder optionally reinforced with stainless steel.
LNP Engineering Plastics
INFO/CARD #246

ESD Footwear Promotes Comfort

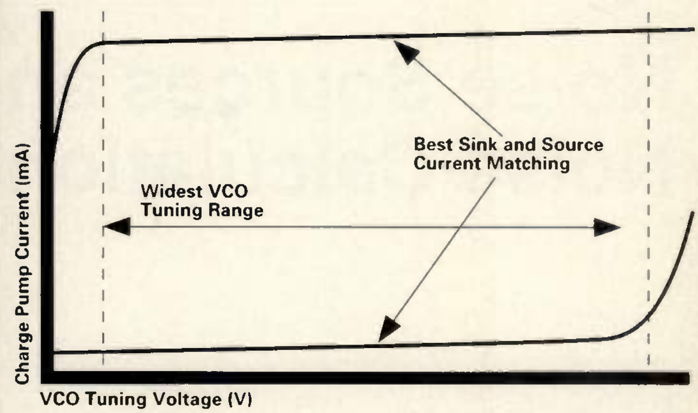
Birkenstock announces its new line of ESD footwear, bringing ergonomic and orthopedic design to the workplace. Available in adult men's and women's sizes, the shoes offer a resistance of 10^5 to 10^8 ohms and a fully dissipative outsole. They also feature long life and repairability.
Birkenstock Footwear
INFO/CARD #248

CHARGE PUMPS GO PLLatinum.

SINGLES	LMX1501A	LMX1511	LMX2315	LMX2320	LMX2325
RF Input-Main PLL	1.1GHz	1.1GHz	1.2GHz	2.0GHz	2.5GHz
(typ) @3V Powerdown (typ)	6mA N/A	6mA N/A	6mA 30μA	10mA 30μA	11mA 30μA*

DUALS	LMX2330A	LMX2331A	LMX2332A	LMX2335	LMX2336	[illegible]
RF Input-Main PLL RF Input-Aux PLL	2.5GHz 510MHz	2.0GHz 510MHz	1.2GHz 510MHz	1.1GHz 1.1GHz	2.0GHz 1.1GHz	550MHz 550MHz
(typ) @3V Powerdown (typ)	13mA 1μA	12mA 1μA	8mA 1μA	9mA 1μA	13mA 1μA	9mA 1μA

PLLatinum™: the industry's highest performing PLLs.



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INFO/CARD 52

Noise Sources and Noise Calculations for Op Amps

By Stephan Baier
Burr-Brown Corp.

Designing electronic systems means dealing with noise because noise is inevitable in electronics. If not accounted for right from the definition phase through component selection and design, the results can be disastrous. Especially in wide band systems designed to process very weak signals noise considerations are mandatory. Often, noise is viewed as a mysterious by-product of electronics and therefore designers tend to ignore it because of the involved supposedly "complex" calculations, or they make some critical generalizations.

The scope of this article is to provide a short, easy-to-use reference for those designers who use IC amplifiers and want to be knowledgeable about the noise performance of their circuits. The emphasis is on the noise that is directly related to the integrated circuit, which are the internal noise sources, and the noise that is generated in the components directly attached to the IC.

Kinds of Noise

First, we make the distinction between noise and interference, with interference being unwanted signals induced into the circuit from external unrelated sources, like power lines, motors, computers, etc. Noise is related to all passive and active components within the circuit and is the portion of an error signal that could not have been predicted by the DC error analysis. Noise can be random or repetitive, voltage or current, narrow- or wide-band, high or low frequency.

Colors of Noise

It is very common to characterize different forms of noise either as "white" or "pink" noise. Thermal noise (= Johnson noise) and shot noise is frequency independent, up to about 100 GHz, and therefore referred to as 'white' because of its similarity to white light, being a composite of all colors, and containing all frequencies. Pink noise, has an amplitude that changes over frequency; the flicker (1/f) noise or Popcorn noise are examples here. Pink noise is often formed by passing white noise through a filter with a 3 dB per octave roll-off.

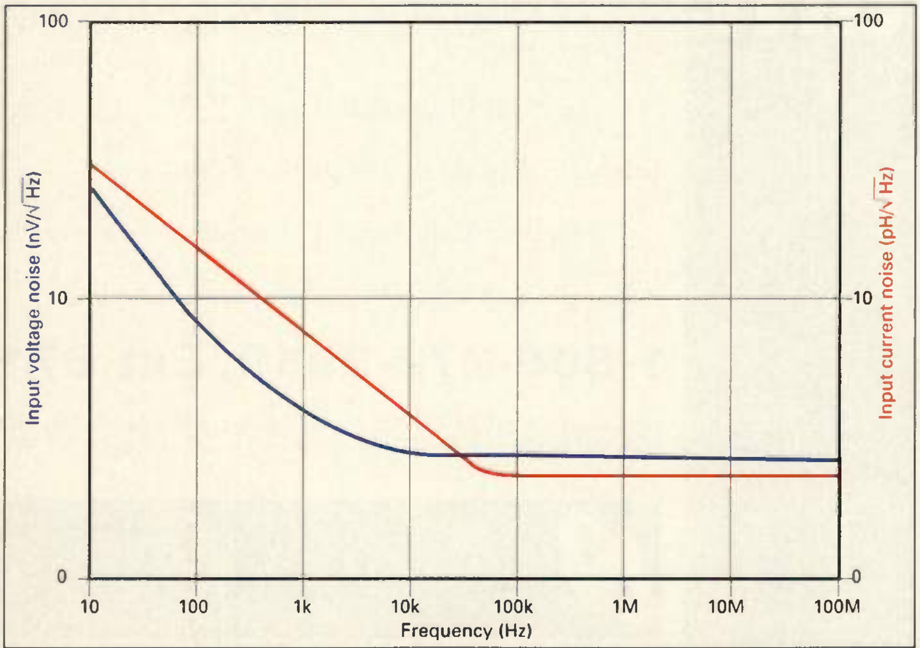

Figure 1. Voltage and current noise spectral density of OPA628, a voltage feedback op amp with a 160 MHz bandwidth.

Noise Density Spectrum

The characterization of noise in terms of its 'spectral density' makes the specification and calculation of noise in differing bandwidths relatively easy. For this reason it is widely used among IC manufacturer to specify the noise performance of their products. An example of a graphic representation of a noise density spectrum is shown in Figure 1.

The noise spectral density is simply the rms value of a noise voltage (e_n) or noise current (i_n) expressed as a voltage or current per root Hertz ($\sqrt{\text{Hz}}$). The short mathematical derivation of this is as follows: the power spectral density is defined as the derivative of noise power over frequency with the units watts per hertz:

$$\rho_u = \frac{dP_u}{df} \tag{1}$$

Because power is proportional to the square of rms voltage or current, the following equations for voltage and current noise spectral density can be derived:

$$e_n = \sqrt{\frac{d\overline{E}_n^2}{df}} \approx \frac{E_n(\text{rms})}{\sqrt{\Delta f}};\frac{V}{\sqrt{Hz}} \tag{2}$$

$$i_n = \sqrt{\frac{dI_n^2}{df}} \approx \frac{I_n(\text{rms})}{\sqrt{\Delta f}};\frac{A}{\sqrt{Hz}} \tag{3}$$

Another name for the noise spectral density that can be found in various

publications is spot noise, which is essentially the same, because it usually defines the frequency "spot" with 1 Hz bandwidth.

Thermal Noise

Thermal noise results from random motion of free electrons in a conductor caused by thermal agitation. This leads to the expression 'Thermal noise power', which is directly proportional to temperature and frequency:

$$P_u = kTB; \frac{J}{s} \tag{4}$$

where k is Boltzmann's constant ($1.38{\times}10^{-23}$ joules/kelvin), T is the absolute temperature (K) and B is the bandwidth (1/s) of the system. The units of kTB are usually joules/second. Because this is the same as watts, the term power becomes meaningful here.

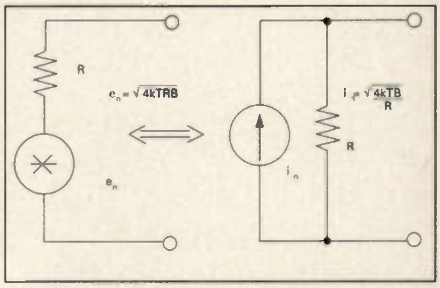

Figure 2. Interchangable resistor noise models.

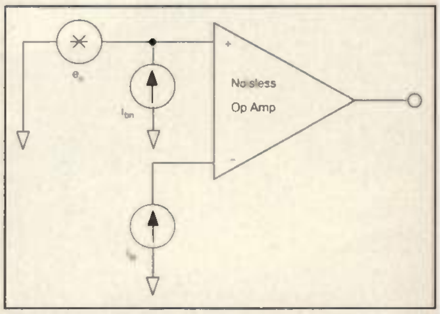

Figure 3. General operational amplifier noise model.

In conductors and semiconductors thermal noise is always present. For example, at temperatures above absolute zero an ohmic resistor is a voltage noise source and its root-mean-square (rms) thermal noise is defined by:

$$e_n = \sqrt{4kTRB} \tag{5}$$

Although noise is most often given as a voltage source, it can be easily transferred into a current source, as shown in Figure 2.

A more common way to express the thermal noise of a resistor is to use the spectral noise density. Here, the rms value of noise is normalized to a 1 Hz bandwidth:

$$\frac{e_n}{\sqrt{Hz}} = 4kTR \tag{6}$$

The units are then usually given as nV/√Hz for a voltage source, or pA/√Hz for a current source. This form is advantageous since for white noise the spectral density is constant; by simply multiplying it by the square-root of the noise bandwidth (NBW) of the system the total rms noise is obtained.

A useful number is that a 1kohm resistor has noise of 4nV/√Hz at room temperature (+25 °C = 298 K). In an attempt to reduce thermal noise, three options are available: reduction in

FREQUENCY SYNTHESIZERS

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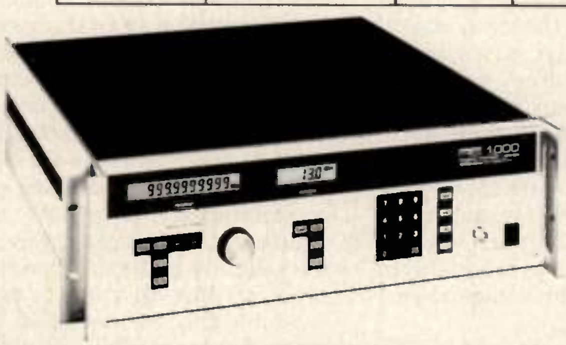
Accurate, stable frequencies on command, fast switching. For NMR, SATCOM, Surveillance, ATE, Laser, Fluorescence, Clock Sources. Low noise/jitter. Sources adapting to your needs with options. High demonstrated reliability. 20,000 + delivered in 20 years.

	Frequency Range	Resolution	Switching Time[1]	Phase-Continuous Switching[2]	Rack-Mount Cabinet Dim.[3]	Remote-Control Interface	Price Example[4]
PTS 040	.1-40 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$5,330.00 (1 Hz resol., OCXO freq. std.)
PTS 120	90-120 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$5,330.00 (1 Hz resol., OCXO freq. std.)
PTS 160	.1-160 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$6,495.00 (1 Hz resol., OCXO freq. std.)
PTS 250	1-250 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$7,440.00 (1 Hz resol., OCXO freq. std.)
Type 1 **PTS 310** Type 2	.1-310 MHz	1 Hz	1-20μs	standard	3½″H×19″W	BCD (std) or GPIB (opt)	Type 1: 1 Hz resol., OCXO: $6,425.00; Type 2: 1 Hz resol., OCXO: $5,850.00
PTS 500	1-500 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$8,720.00 (1 Hz resol., OCXO freq. std.)
PTS 620	1-620 MHz	optional .1 Hz to 100 KHz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$9,625.00 (1 Hz resol., OCXO freq. std.)
PTS 1000	0.1-1000 MHz	optional .1 Hz to 100 KHz	5-10μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$11,830.00 (1 Hz resol., OCXO freq. std.)
PTS 3200	1-3200 MHz	1 Hz	1-20μs	optional	5¼″H×19″W	BCD (std) or GPIB (opt)	$14,850.00 (1 Hz resol., OCXO freq. std.)
PTS x10	user specified 10 MHz decade	1 Hz	1-5μs	standard	3½″H×19″W	BCD (std) or GPIB (opt)	$3,000.00 (1 Hz resol., OCXO freq. std.)
PTS D310	two channels .1-310 MHz	.1 Hz	1-20μs	standard	5¼″H×19″W	BCD (std) or GPIB (opt)	$8,560.00 (.1 Hz resol., OCXO freq. std.)
PTS D620	two channels 1-620 MHz	.1 Hz/.2 Hz	1-20 μs	standard	5¼″H×19″W	BCD (std) or GPIB (opt)	$13,240.00 (.1 Hz/.2 Hz resol., OCXO freq. std.)



1 Switching Time is dependent on digit (decade) switched; see detailed instrument specifications.

2 For applicable digits, see detailed instrument specifications.

3 Bench cabinets are 17″ wide.

4 Prices are U.S. only and include Manual and Remote (BCD) Control; PTS 3200 Digital Front Panel.

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P.O. Box 517, 9 Beaver Brook Rd., Littleton, MA 01460 Tel: 508-486-3400 FAX: 508-486-4495 resistance, bandwidth or temperature. However, temperature reduction has generally a minor impact since noise power is proportional to the absolute temperature, T = °C + 273 K.

Shot Noise

Shot noise (or Shottky noise) is a white current noise caused by the quantized and random nature of current flow.

The shot noise current spectral density is defined by:

$$\overline{i_s}^2 = 2qI_{DC}B \qquad (7)$$

with: q = electron charge (1.6x10^{-19} coulombs), I_{DC} = DC current B = noise bandwidth.

Flicker (1/f) noise

Flicker noise is caused by contamination and defects in the silicon lattice structure resulting in the action of combination-recombination of carriers in the emitter-base area of a transistor. Even though it is mainly associated with bipolar processes similar noise has been found on CMOS processes as well. One formula that describes the flicker noise current spectral density is:

$$\overline{i_f}^2 = \frac{2qI_{DC}^{\lambda}f_cB}{f} \qquad (8)$$

with: q = electron charge, I_{DC} = DC current, f_C = corner frequency, f = frequency of interest, Δf = noise bandwidth, Y = an exponent between 1 and 2.

Popcorn noise

Another peculiar form of noise is the popcorn or burst noise. It has its name from the sound it makes in an audio system and the pulse shape appearance when viewed on an oscilloscope. The noise bursts appear randomly with different amplitudes and duration of up to milliseconds, which makes it a low frequency effect.

The source of this kind of noise is described as localized punch-through of emitter-base junction (also called emitter piping) and contamination in the emitter-base region by metallic ions. It is like flicker noise, process related and was a more serious problem at one time than it is today due to modern process control capabilities. The popcorn noise spectral density is given by the equation:

$$\overline{i_p}^2 = K\frac{I_c}{1+\left(\frac{f}{f_c}\right)^2}B \qquad (9)$$

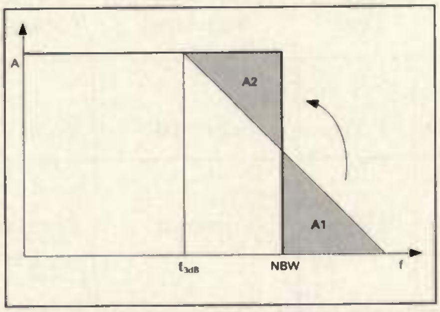

Figure 4. Brickwall and linear-rolloff system responses which enclose the same noise power.

with: K = a constant for a particular device, I_c = a direct current, C = a constant between 0.5 and 2, f_C = a corner frequency related to the process, f = frequency of interest, B = noise bandwidth.

Op Amp Noise Model

A simplistic way of describing the total noise of an op amp is to extract the internal noise generators and locate them at the input of the now "noiseless" op amp. As shown in Figure 3, the model uses one noise voltage source located in series with the non-inverting input and two noise current sources between each input and ground. It is important to note that all noise generators are independent and provide uncorrelated, random noise. Even though the current sources show an arrow in their symbol the generated noise does not have a defined flow direction. Keeping in mind that noise sources, because they are uncorrelated (i.e., one noise signal cannot be transformed into the other), have to be summed together by their root-sum-of-their-squares (RSS), a negative sign associated with the signal would become positive. Therefore, signs have no meaning in noise analysis.

Unfortunately, designers often tend to compare only the specified voltage noise (e_n) figures on the data sheets and make their selection based on the "if it's lower, it's better" rule. The problem with this is that it ignores the contribution from the current noise (i_n) which can become the dominant noise signal, depending on the type of op amp and the source resistance. To obtain the total noise the current noise has to be multiplied by the source resistance and added to voltage noise:

$$e_{nt} = \sqrt{e_n^2 + (i_nR_s)^2} \qquad (10)$$

The consideration of the noise cur-

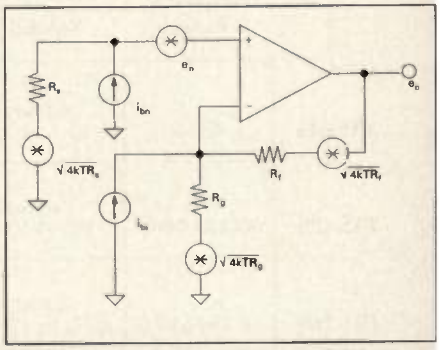

Figure 5. Op amp noise model including external resistor noises.

rent is also important for an accurate calculation of the noise figure, as we will discuss later.

Bipolar vs FET op amps

The current noise of an op amp is directly associated with the bias current. In fact, in absence of noise current specifications taking the bias current and applying it to the shot noise equation ($i_n = \sqrt{2q\ Ib\ B}$) will lead to a good approximation. For example, an op amp with a bipolar transistor input stage the current noise spectral density will be in the order of 0.56pA/√Hz, for a bias current of Ib = 1μA. This noise signal does not vary much over temperature. Different is the situation with FET op amps: because FET input type op amps provide input bias currents magnitudes lower than bipolar types, the noise current contribution is also much smaller, e.g. 10pA of bias current result in only 1.8fA/√Hz noise. However, because the bias current of FET op amps typically doubles for each 10 K increase of the junction temperature, the current noise increases by a factor of √2 with each 10 K rise. Another note of caution should be made; these current noise calculations hold true only for those op amps that do not have an internal bias current compensation scheme. For bias compensated op amps the specified bias current is the difference between the base current and the compensating current. Although, the DC-bias current would ideally cancel, from the noise point of view two sources of equal value have to be rms-summed together accounting for a √2 (or 40%) increase in noise current. As a general guideline for the selection of the appropriate op amp type: bipolar op amps traditionally exhibit higher current noise than FET types, but are superior in their voltage noise. Therefore they are more suited for applications with low source imped-

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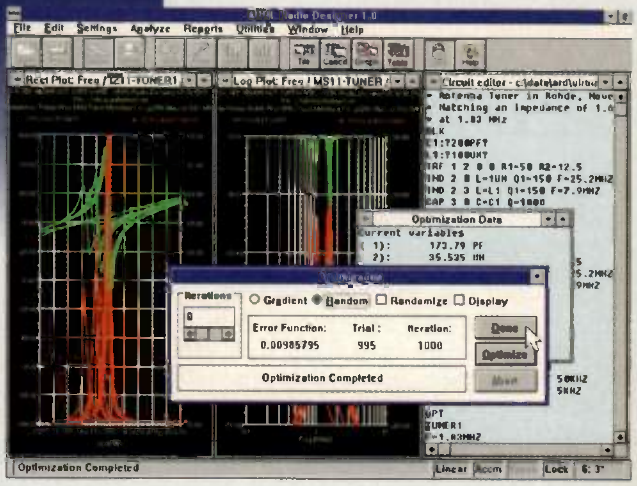

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- Gain, gain matching and noise parameters; and
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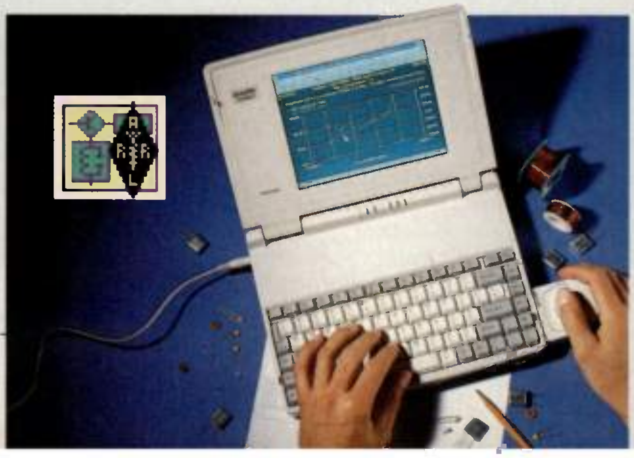
Maximum number of ports per circuit block	30
Maximum number of frequency steps	512
Maximum number of statistical histograms	20

* The number of circuit blocks allowed depends on their complexity; generally, 50 or more circuit blocks may be defined in a single *ARRL Radio Designer* netlist.



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ances. With new process technologies, FET op amps are approaching bipolar voltage noise performance while retaining their very low current noise. One example is the OPA655 from Burr-Brown, a 400 MHz FET op amp with 10 fA/√Hz noise current and only 5 nV/√Hz of noise voltage. For applications with high source impedances, e.g. transimpedance amplifiers (I/V converter) with high gains, FET op amps are the prime choice.

Noise Figure

Traditionally, the noise figure has been the standard noise parameter for RF components, such as transistors. Nowadays IC manufacturers have added noise figure specifications to the noise spectral density numbers, and this for a good reason. Integrated op amps achieve now bandwidth in excess of 1GHz (the OPA640 from Burr-Brown has a 1.3 GHz bandwidth) and RF designer can easily replace several discrete components with the IC. Therefore, op amps have to perform in the standard RF environment of 50ohm or 75ohm impedances, for which the noise figure originally was developed. This already indicates that noise figure not only reflects the noise contributions of the IC itself, but rather describes the IC with its attached components (feedback network, source and load resistance) as an entity. Using the noise figure a gain block can be completely characterized in terms of its noise. Therefore, total system noise calculations can easily be done by adding the noise figure numbers of each processing stage together.

Calculating Noise Figure

The noise figure itself is the logarithm of the signal-to-noise ratio on the input of the amplifier to the signal-to-noise ratio at the output.

$$NF = 10\log\frac{(S/N)_{in}}{(S/N)_{out}} \tag{11}$$

To calculate the noise figure for an op amp gain stage the following equation is more appropriate:

$$NF = 10\log\left(1+\frac{e_n^2+(i_nR_s)^2}{4kTR_s}\right) \tag{12}$$

It can be seen that the noise figure includes the voltage and current noise power from the amplifier. The noise current (i_n) flows through the source impedance (R_s), therefore the dependency of the noise figure to the source resistance. If given in a graphic from the noise figure is plotted over a range of source impedance.

The noise figure plot has a very characteristic shape: it has a minimum at a certain source resistance and increases for lower and higher resistances. Towards lower resistance the main noise contribution comes from the amplifier's noise voltage. For the rising side (increasing resistances) the noise current takes over and accounts for the increase in the noise figure. For a given source impedance the selection of the right op amp is a practical way to design a low-noise amplifier.

Comparing Noise in VFAs and CFAs

In some wideband applications not only the noise performance of an op amp is important but also its distortion. Then current feedback amplifiers (CFA) have an advantage because of their higher loop gain at higher frequencies compared to voltage feedback amplifier (VFA). Because current feedback amplifier have asymmetrical input impedances the bias current on the inverting input us typically higher than on the non-inverting input. Therefore, the inverting noise current on a CFA can be up to 10 times higher compared to voltage-feedback amplifier. Another restriction with CFAs is that the feedback resistor value has to stay within a specified range to insure stable operation, and cannot be arbitrarily reduced to improve the noise performance.

Noise Bandwidth

Before actually measuring or calculating the total noise of a system one question to ask is "What is the bandwidth of my system?". This is a very important determination because noise is proportional to the bandwidth. In order to calculate the total integrated noise in volts rms, the total output noise spectral density (e_{no}) given in nV/√Hz is multiplied by the square-root of the bandwidth. The implication of this is that 'bandwidth' in this case means that the upper frequency corner has an infinite roll-off, resembling a "Brickwall" filter. Generally, this is not the case and the transfer function of the measurement channel is more like a low-order Butterworth low-pass filter. By adding a factor into the equation of the integrated noise, the bandwidth of the theoretical brickwall filter now encloses the same noise power (or area) as the real filter transfer function, see Figure 4. This method is only applicable as long as the gain response is flat, within ±1 dB, over the bandwidth of interest. If the system response exhibits higher gain peaking a different approach, e.g. piecewise integration of the noise, should be considered to obtain accurate noise predictions. The contribution of the 1/f-noise is usually neglected since it represents an insignificant term in wideband systems.

Filter Order	Noise Bandwidth (NBW/f–3dB)	Roll-Off (dB per octave)
1	1.57	6
2	1.22	12
3	1.15	18
4	1.13	24
5	1.11	30

Table 1. Conversion Chart for Noise Bandwidth using Butterworth filter responses.

Noise calculation example

The following discusses the calculation of op amp noise in a simple non-inverting configuration. Shown in Figure 5 is the op amp noise model with the external noise sources. To derive the total output noise, each term is multiplied by its gain and taken to the output as a voltage. All terms are added together as the square root of the sum of their squares. The individual terms are:

$$R_S \rightarrow \sqrt{4kTR_S}\left(1+\frac{R_f}{R_g}\right) \tag{13}$$

$$i_{bn} \rightarrow i_{bn}R_S\left(1+\frac{R_f}{R_g}\right) \tag{14}$$

$$e_n \rightarrow e_n\left(1+\frac{R_f}{R_g}\right) \tag{15}$$

$$i_{bi} \rightarrow i_{bi}R_f \tag{16}$$

$$R_g \rightarrow \sqrt{4kTR_g}\,\frac{R_f}{R_g} \tag{17a}$$

$$R_f \rightarrow \sqrt{4kTR_f} \tag{17b}$$

$$e = \left(\left(4kTR_S + (i_{bn}R_S)^2 + e_n^{2}\right)\left(1+\frac{R_f}{R_g}\right)^2 + (i_{bi}R_f)^2 + 4kTR_f\left(1+\frac{R_f}{R_g}\right)\right)^{\frac{1}{2}} \tag{18}$$

Crest Factor p-p/rms	Percent of time peak is exceeded
2	32%
3	13%
4	4.6%
5	1.2%
6	0.3%
6.6	0.1%
7	0.05%

Table 2. Crest factors for converting rms noise into peak-to-peak noise.

Equation 18 provides the general form for the total output voltage noise. Note that this represents the noise spectral density per √Hz.

Measuring Noise

So far noise has been referred to in rms or spectral density form. Whenever measuring noise with an oscilloscope the crest factor has to be included. The crest factor provides a statistical measure of the relationship between the rms noise and its peak value. The reason for statistics in noise measurements is that noise is a random event and its amplitude has a Gaussian probability distribution. This means that voltage noise peaks are many times greater than the rms noise, but the highest peaks occur very infrequently and are therefore difficult to measure repeatedly. Table II lists the different crest factors that should be used to estimate peak-to-peak noise.

For example, using a crest factor of 6.6 means that 99.9% of the existing noise falls within the limits of the peak-to-peak values. This is also a good number for comparing calculated noise data with the noise observed on the oscilloscope. *RF*

About the Author

Stephan Baier is a senior application engineer with Burr-Brown Corp. in Tuscon, Arizona. He recieved his BSEE from the Acadamy of Telecommunication in Dieburg, Germany, in 1990. Before relocating to the US headquarter of Burr-Brown in 1993 he worked as an applications engineer in the german Filderstadt sales office.

He can be reached at Burr-Brown Corp., 6730S. Tucson Blvd., Tucson, Arizona 85706, Phone: (800) 548-6132.

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Measured Effect of Seasonal Foliage Growth on a Short UHF Telemetry Link

By Noel E Evans
University of Ulster, N. Ireland

Users of VHF/UHF mobile radios in wooded regions are familiar with increased incidence of signal fade-out in spring and summer as leaf growth takes place. This is often not a serious problem for vehicular systems since natural movement will restore satisfactory operation in a short time frame; special purpose, fixed links may present real difficulties, however, and low power telemetry systems are especially prone to failure with even small increases in path loss. Accordingly, a specific and naturally obstructed radio path was studied for a full year in a part of Northern Ireland subject to an annual rainfall of over 750 mm.

Short radio paths of 1-5 km are often used in rural areas for animal telemetry and for agricultural monitoring and control purposes. Remote transmitters/receivers are battery powered, are generally constrained to using a minimum amount of hardware and must operate to a strict power budget; output signal-to-noise ratios may be marginal and the reliability of uncoded digital links will deteriorate rapidly with relatively small increases in path loss [1]. For example, as shown in Figure 1, an additional 4 dB loss in a non coherent FSK system can drive the bit error rate from 2E10-4 (a just-acceptable design level) to 2EI0-2 (poor). Antenna heights are often necessarily low in telemetry and telecommand applications, so paths for this type of link are liable to offer significant excess loss due to obstructions such as hills and vegetation. The diffraction loss for multiple edges and the loss due to wooded regions may be calculated with reasonable accuracy, but changes due to leaf generation and subsequent depletion require experimental measurement.

The Propagation Path

The year's monitoring was carried out on a 462 MHz fixed link operating over the ground profile shown in Figure 2. This plot does not indicate the foliage present and ignores the relatively insignificant amount of Earth curvature which occurs at a transmitter-receiver spacing of 2.4 km. The signal path lies across sparsely-populated farmland, with traditional thorn hedging used for field separation. The hedges are on average 1.5 m high and 0.75 m wide; in total, they account for less than 1 percent of the exposed path length and have little effect on propagation. The dominant obstacle is a drumlin close to the receiver which supports a grove of naturally seeded, mature beech and oak trees. The tree heights range from 10 - 20 m, with corresponding trunk diameters of 25 - 60 cm. The grove is approximately 150 m in diameter with a floor of thick undergrowth, again mainly of hawthorn and up to 1.5 m in depth.

Link Equipment

Incoming carrier levels were recorded at 17:00 GMT on alternate days,

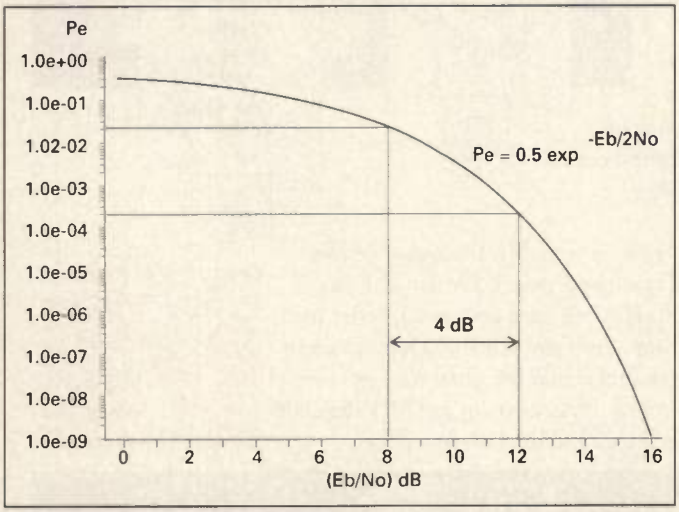


Figure 1. Error probability Pe in a non-coherent FSK link. Eb is the signal energy per bit and No is the average noise power density in W/Hz.

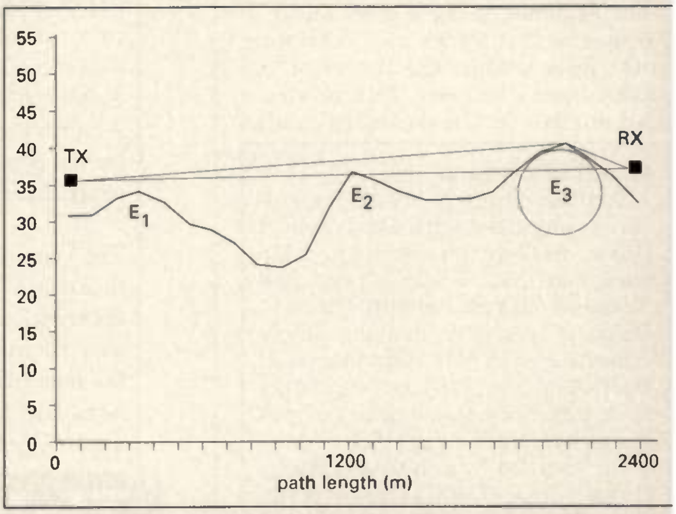


Figure 2. Elevations between the link's transmitter and receiver. Losses include free-space loss, knife edge losses for each hill, rounded obstacle loss for E3 and foliage loss.

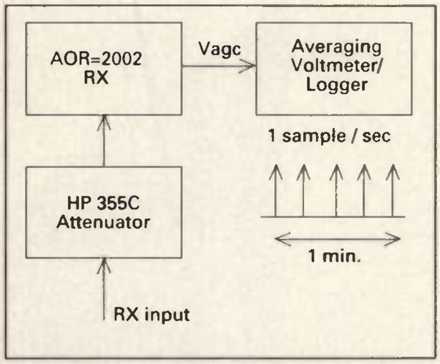


Figure 3a. Receiver hardware block diagram. Recorded signal levels were averaged from 60 samples taken at 1 second intervals.

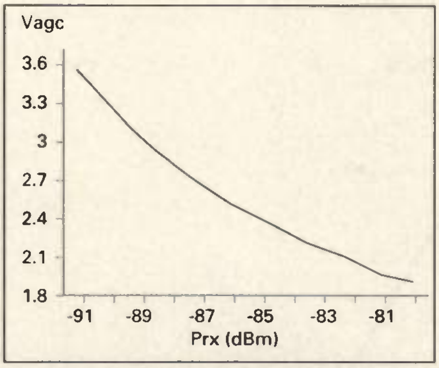


Figure 3b. AR2002 AGC characteristic, with 19 dB of antenna port attenuation.

	E_1	E_2	E_3
Fresnel Parameter	–0.22	–0.12	0.38
Edge Loss (dB)	3.9	5.0	10.0
Rounded obstacle loss (dB)	-	-	3.5

Table 1. Calculated values of signal path ground loss, totalling 22.4 dB.

using a measurement receiver (a modified AOR 2002 preceded by a variable attenuator) with its AGC voltage fed to an averaging voltmeter/logger built around a Psion Organiser II: Figure 3 summarizes the sampling details and shows the AGC calibration curve used. The custom transmitter had a regulated power output of +34.5 dBm and the receiver noise floor was -130 dBm. Both units were checked for parameter drift on a weekly basis; the combined power output/sensitivity stability was maintained at ±0.2 dB throughout the measurement period. Vertical antenna polarization was in use, with a three element (6 dBd) Yagi at the transmitter and a 5 dBd collinear at the receiver. Both antennas were mounted 5 m above ground level, well clear of nearby obstacles; the feeder loss was 2.5 dB in total.

Path Loss Calculations

For comparison with practical measurements, the link's attenuation (dB) was calculated by summing the free-space loss of 93.4 dB with excess ground loss, L_G, and foliage loss, L_F. L_G was made up of individual contributions from each of the three hills present, E_1, E_2 and E_3. For analysis purposes each was treated as a knife edge. Table 1 gives loss figures derived using Deygout's method; this returns good accuracy for three, well spaced obstacles, one of which is dominant [2]. A rounded obstacle loss [3] was calculated for the main edge E_3 and included in L_G The foliage loss was calculated using an exponential decay model [4] for dense, dry, in leaf trees, according to:

$L_F = 1.07\ d_F^{0.6}$ (1) where d_F is the foliage depth in meters. For the grove in question this gave a value of 21.6 dB which, when added to the other loss factors, gave a theoretical total of 137.4 dB. The corresponding minimum received signal power expected was therefore -90.2 dBm; this figure was validated in practice.

Experimental Findings

Figure 4(a) is the receiver's input power profile obtained during the measurement period; Figure 4(b) shows the maximum/minimum air temperatures prevailing, while Figure 4(c) indicates the rainfall noted on a daily basis. Climate data was derived from a weather station at the transmit end of the link The received power profile was biphasic in nature, with well-defined slopes between the winter (no leaf) and the late summer (full leaf) conditions. The maximum difference, summer to winter, was 8 dB. Maximum signal levels over the winter period correlated with the months during which the minimum air temperatures regularly fell to freezing point and below, and there was no active growth. The transition from maximum attenuation at the end of the growing season was quite sharp, at 6.7 dB per month. It is likely that leaf drop was accelerated by the dry period in October/November. Although tree leaves reached their maximum area and density by the end of July 1993, the underlying vegetation continued to thicken throughout the growing period and contributed to overall signal attenuation, attaining maximum impact in September. In 1994, new-leaf generation was observed in May; this was followed by a steady fall in received signal power, at about 1.8 dB per month. Periods of moderate rainfall (over 30 mm per week) followed by higher than normal day temperatures promoted fast foliage growth and transient reductions in signal level of up to 2 dB. This was very noticeable during the May-June and August-September periods; the effect is attributed to surges in leaf/stem moisture caused by increased transpiration [5]. Diurnal

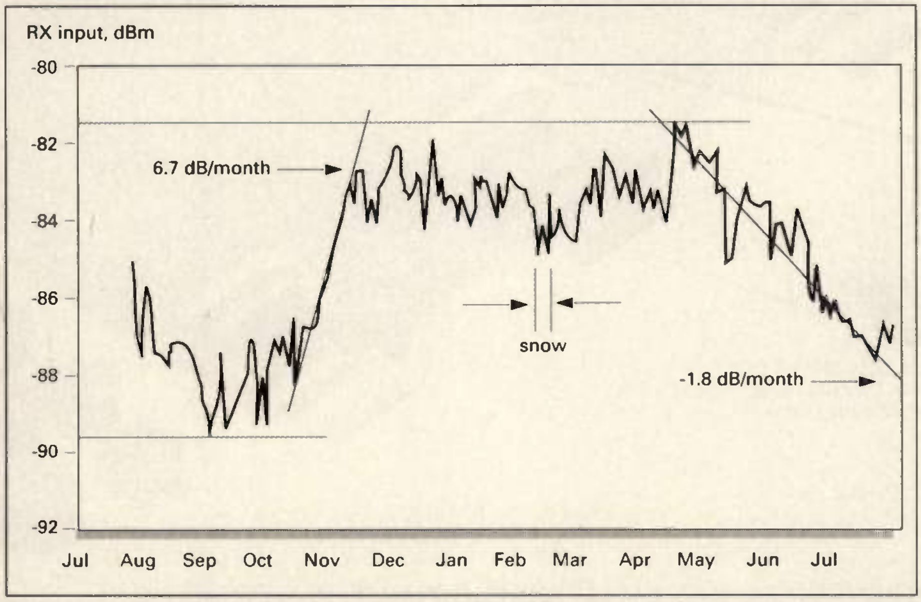
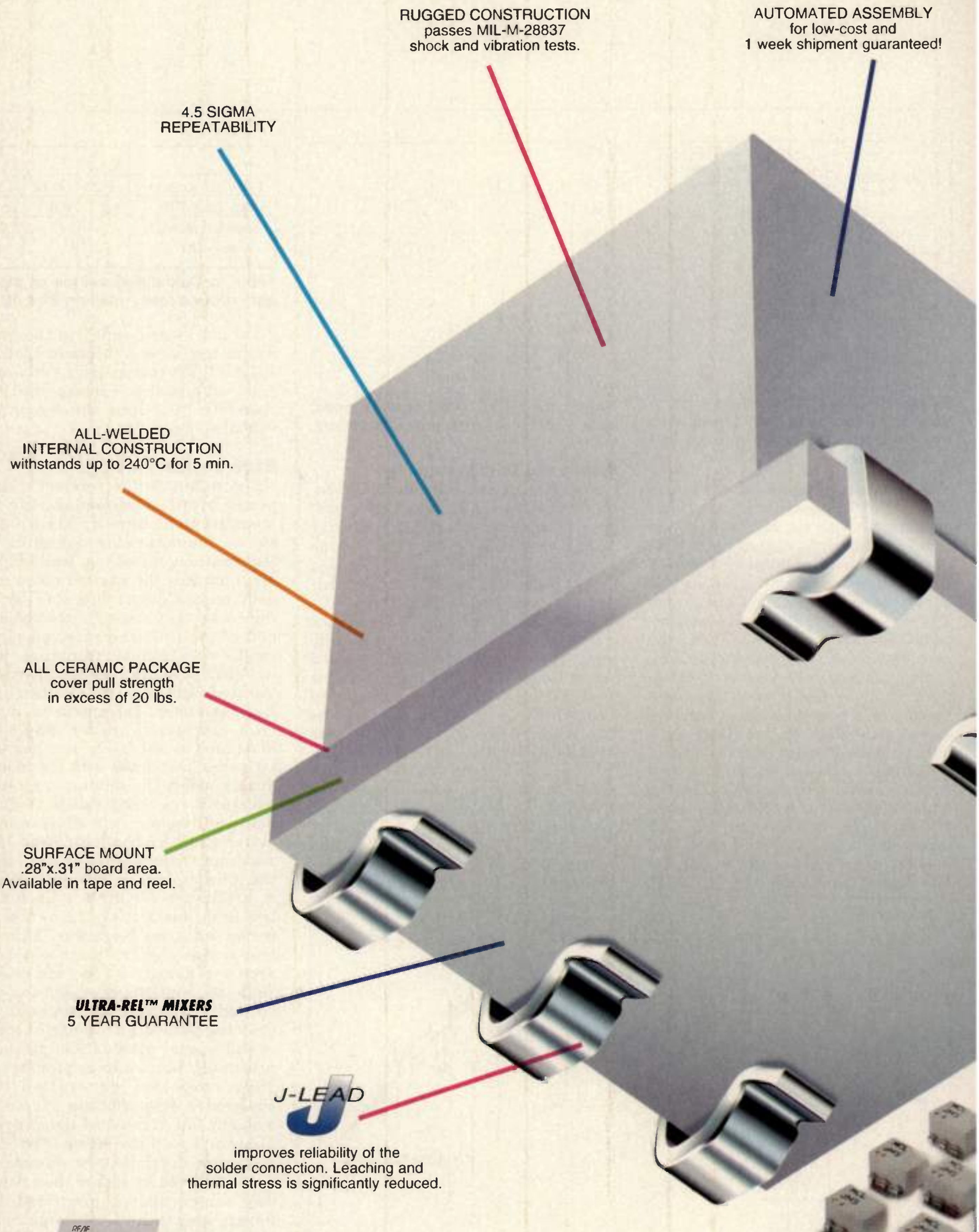


Figure 4a. Received signal levels at 17:00 GMT, recorded using apparatus of Figure 3a, from July 1993 to July 1994.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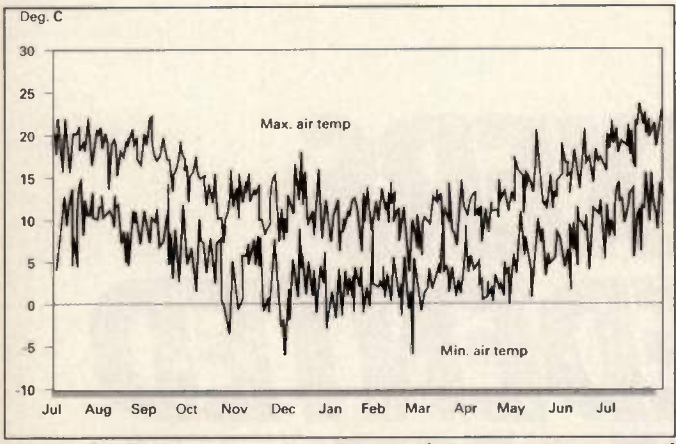

Figure 4b. Maximum and minimum air temperatures noted daily.

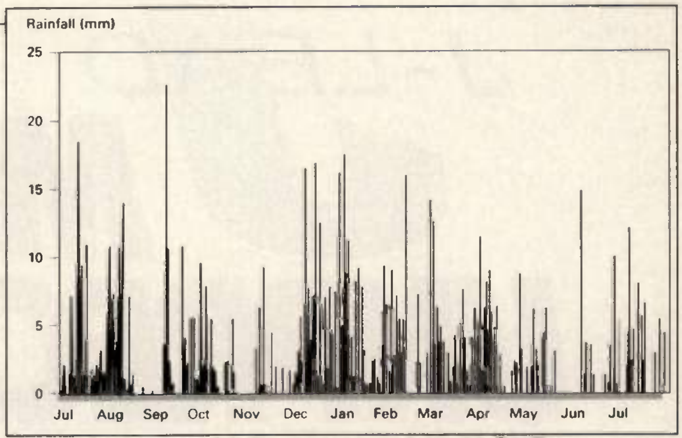

Figure 4c. Daily rainfall measured at 09:00 GMT.

variations from wet to dry conditions (including mist) resulted in signal 4 strength fluctuations of 1 - 1.5 dB. Such changes occurred during the year but increased in late summer when foliage volume was at a maximum. A 6 cm fall of wet snow in mid-February resulted in an additional 0.5 dB loss during the short period it persisted.

Conclusion

It is likely that isolated (single day) sampling of foliage attenuation is not adequate in situations where a relatively small change in path loss may adversely affect link reliability. This is particularly true when weather conditions promote fluctuating rates of leaf growth and the link is low-power-conscious. In the type of path discussed, up to 2 dB daily variation in attenuation can occur on a year round basis, depending largely on the rainfall; weekly totals of up to 15 mm restrict the variability to about 0.5 dB. Dry spells outside the growing period show best signal strength stability. The exact slopes of biphasic signal strength plots measured from year to year will ultimately depend on local climatic conditions; the information presented here should, however, be representative of behavior in a Cool Temperate zone since the weather data collected in 1993/94 showed no significant anomalies. *RF*

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An Introduction to Class-F Power Amplifiers

By Frederick H. Raab, Ph.D.
Green Mountain Radio Research Co.

High-efficiency power amplifiers (PAs) can be implemented by true switching-mode techniques (classes D and E) at frequencies from VLF through HF and sometimes even VHF. At UHF and higher frequencies, however, it is difficult to find transistors capable of switching fast enough for class-D and class-E operation. In addition, the drain/collector capacitance, leadinductance, lead length (including bond wires), and dispersion (frequency-dependent propagation velocity) make implementation of ideal tuned output circuits difficult. As a result, there is considerable recent interest inclass F for improving the efficiency of PAs in cellular- and personal-communication systems.

A class-F power amplifier (PA) uses a multiple-resonator output filter to control the harmonic content of its drain-voltage and/or drain-current waveforms, thereby shaping them to reduce dissipation and to increase efficiency. It is probably the oldest technique for improving the efficiency of an RF PA. This paper introduces class F through five basic circuits:

- Third-harmonic peaking,
- Second-harmonic peaking,
- Transmission line with parallel-tuned output,
- Transmission line with series-tuned output, and
- Arbitrary number of correct harmonic impedances.

In the first four configurations, at least one of the waveforms is a perfect half sine wave or square wave and therefore contains a complete set of even or odd harmonics. In the more general case, neither waveform is complete.

All analyses in this paper are based upon ideal transistors, which are shown here as MOSFETs. They are assumed to have neither saturation voltage nor on-state resistance, and to be ideal current sources or switches. The effects of nonzero saturation voltage or resistance can be included through the use of an effective supply voltage, as described in Chapter 14 of [1].

Third-Harmonic Peaking

The circuit and waveforms for a third-harmonic-peaking class-F PA areshown in Figure 1. The circuit is identical to that of a single-ended-class-B or C PA except for the addition of third-harmonic resonator L3-C3. This is the classic Tyler circuit [2]; other examples are given in [3] and[4].

The third-harmonic resonator is a parallel-tuned circuit and therefore represents the drain with an infinite impedance at 3f. The impedance at the fundamental frequency is load resistance R. The impedance is zero at other frequencies. There is therefore no impediment to the half-sine-wave drain current of class-B operation, which consists of dc, fundamental frequency,and even harmonics. The drain-voltage waveform has the form

$$v_D(\theta) = V_{DD} + V_{0m}\sin\theta + V_{3m}\sin 3\theta \tag{1}$$

where $\theta = \omega t$, $\omega = 2\pi f$, and f is the fundamental (angular) frequency.

At low drive levels, the level of the driving signal determines the magnitude i_{Dmax} of the drain current, which in turn determines the fundamental-frequency output current I_{om}. As in a class-B PA,

$$V_{0m} = I_{0m}R \tag{2}$$

and

$$P_0 = \frac{V_{0m}^2}{2R} \tag{3}$$

At lower drive levels, the third harmonic is generally small and the drain-voltage waveform remains mostly sinusoidal.

As drive is increased past the point at which $V_{om} = V_{DD}$, the MOSFET enters saturation near the minimum voltage. The high impedance at 3f allows the presence of a third harmonic, which flattens the drain-voltage wave form. Odd-harmonic symmetry causes the flatness near the minimum drain voltage to be mirrored near the maximum voltage. The requirement for flatness near the minimum voltage causes the phase of the third harmonic to be as shown in (1).

The amplitudes of the fundamental-frequency and third-harmonic components are determined through a process much like that in a saturating class-C PA (Section 13-2 of [1]). For a maximally-flat waveform [1], [5], $V_{3m} = V_{om}/8$ and the corresponding fundamental-frequency at peak output voltage is $V_{om} = (9/8)V_{DD}$. The peak power output

$$P_0 = \frac{81V_{DD}^2}{128R} = 0.633\frac{V_{DD}^2}{2R} \tag{4}$$

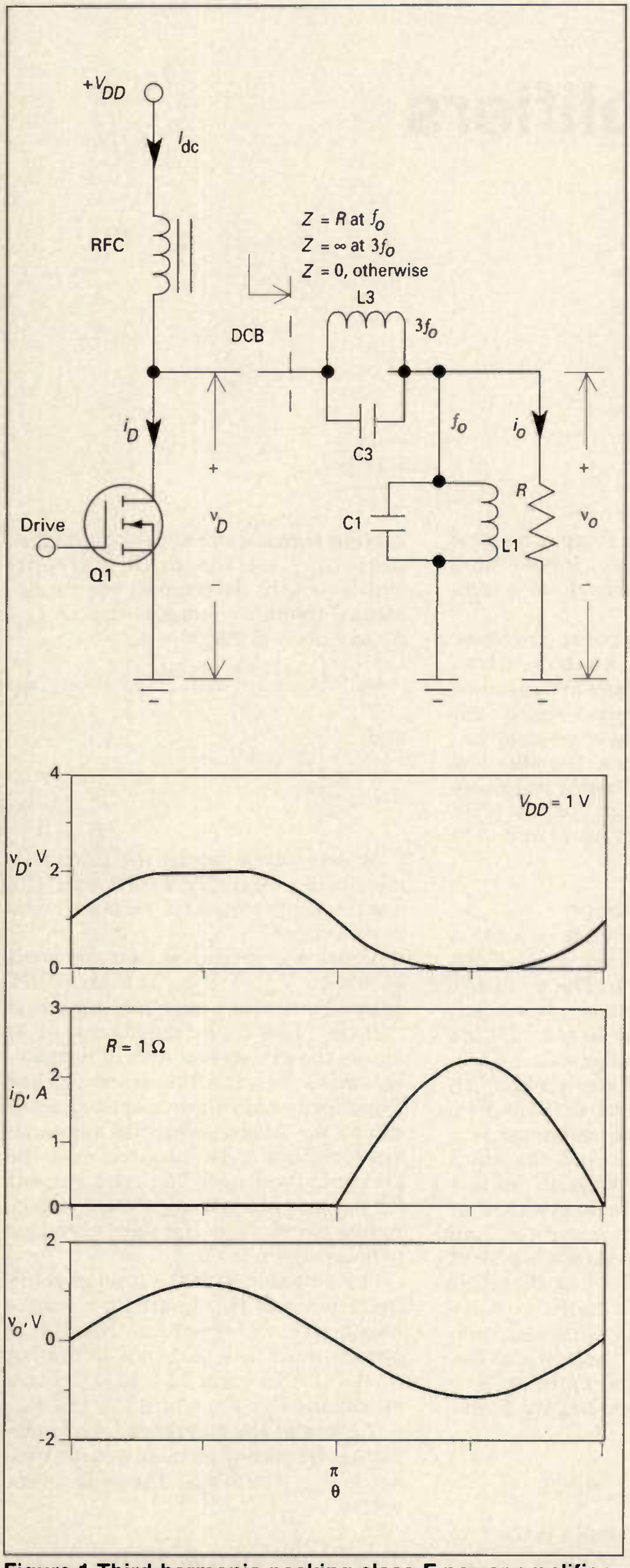

Figure 1 Third-harmonic-peaking class-F power amplifier

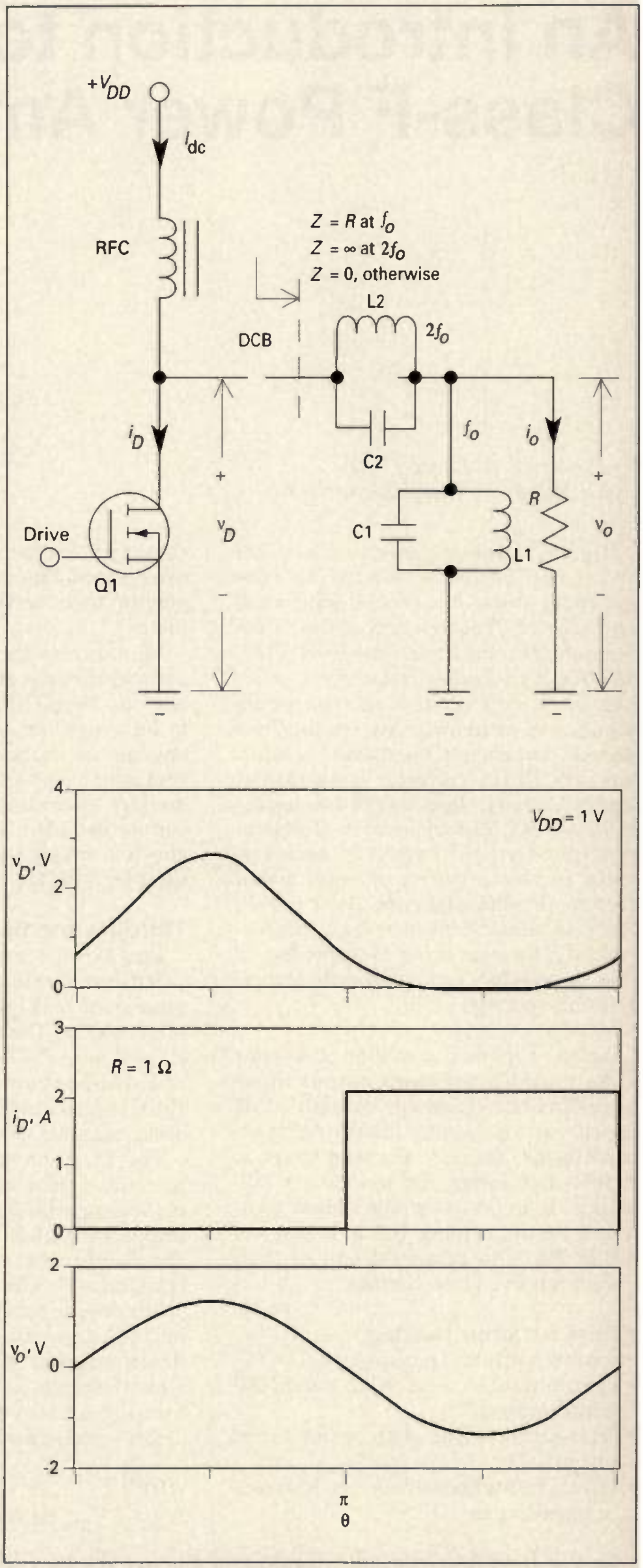

Figure 2. Second-harmonic-peaking class-F power amplifier.

is 27-percent greater than that for class-B operation.

The dc-input current is related to peak drain current and output current as in a class-B PA, hence $I_{dc} = (1/\pi)$ $i_{Dmax} = (2/\pi)I_{om}$. The dc-input power is therefore (9V2DD/4πR) and the efficiency at PEP is

$$\eta = \frac{P_i}{P_o} = \frac{9\pi}{32} = 0.884 \tag{5}$$

Second-Harmonic Peaking

The circuit for the second-harmonic peaking PA (Figure 2) is similar to that of the third-harmonic peaking PA, the additional resonator creates a high impedance at the second harmonic. The drain current is a square wave, containing only dc, fundamental-frequency, and odd-harmonic components. PAs of this type are described by McAlpin [6] and Zivkovic [7].

The principles of operation are generally the same as those of third-harmonic peaking. The drive causes the MOSFET to act as a current source. Harmonic components of the drain current are bypassed to ground by parallel-tuned output tank L1-C1. The fundamental-frequency voltage created on the load also appears on the drain. As drive is increased past the maximum output for class-B operation, the MOSFET begins to saturate. This creates a second-harmonic voltage, which flattens the drain-voltage waveform, causing it to approximate a half sine wave.

Drain-Voltage Waveform – The drain-voltage waveform is described by

$$v_D(\theta) = V_{DD} + V_{0m}\sin\theta + V_{2m}\cos 2\theta \tag{6}$$

Flattening of the waveform requires the second-harmonic cosine rather than sine.

Maximum flatness is achieved with V2m = $-V_{om}/4$. For maximum output, the minimum drain voltage is zero, hence Vom = $(4/3)V_{DD}$. The power output is therefore

$$P_0 = \frac{V_{0m}^2}{2R} = \frac{8}{9}\frac{V_{DD}^2}{R} \tag{7}$$

$$= 0.888\frac{V_{DD}^2}{R}$$

The peak drain voltage is $v_{Dmax} = (8/3)V_{DD}$.

Drain-Current Waveform – The magnitude of the output current at PEP is $I_{om} = (4/3)V_{DD}/R$. Since the drain current waveform is a square wave, $I_{om} = (2/\pi)i_{Dmax}$. The peak drain current is therefore

$$i_{D\,max} = \frac{2\pi}{3}\frac{V_{DD}}{R} \tag{8}$$

and the dc-input current is

$$I_{dc} = \frac{1}{2}i_{D\,max} = \frac{\pi}{3}\frac{V_{DD}}{R} \tag{9}$$

Efficiency and Power-Output Capability – The efficiency of the second-harmonic peaking PA is

$$\eta = \frac{8/9}{\pi/3} = \frac{8}{3\pi} = 0.849 \tag{10}$$

The power output capability is

$$P_{max} = \frac{P_0}{v_{D\,max}\,i_{D\,max}} = \frac{1}{2\pi} = 0.159 \tag{11}$$

which is the same as the transmission-line class-F PAs described in the nexttwo sections.

Transmission Line with Parallel-Tuned Output

In theory, it is possible to include an infinite number of additional resonators in the harmonic-peaking circuit of Figure 1, thus allowing a true square-wave voltage and therefore 100-percent efficiency. Such a PA is implemented in practice (Figure 3) by using a quarter-wavelength transmission line and a parallel-tuned output circuit. This type of PA [8], [4] is used at VHF where implementation of lumped-element networks (e.g., Figures 1 and 2) is difficult, but the inductance between the drain and the case remains relatively small.

The quarter-wavelength transmission line (Figure 13-10, [1]) transformsthe load impedance according to\sq

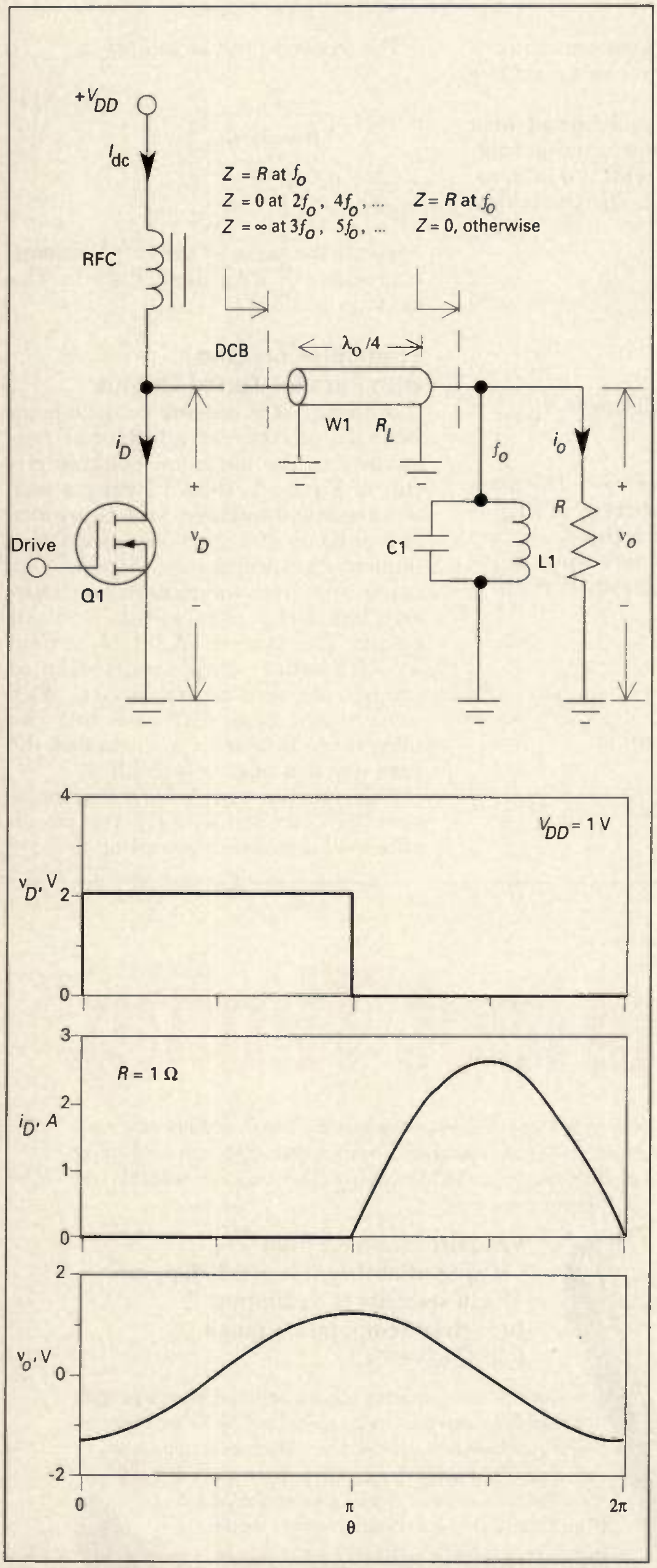

Figure 3. Class-F power amplifier with quarter-wavelength line and parallel-tuned output.

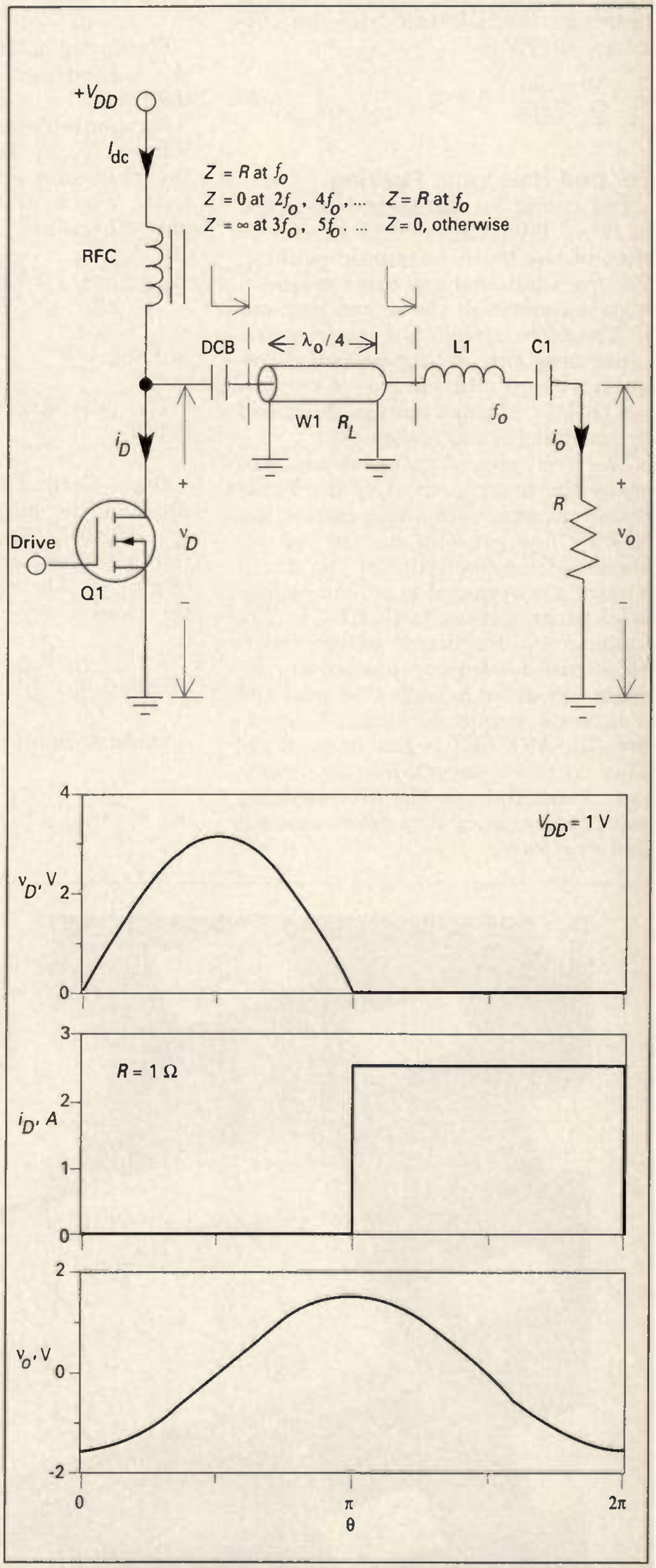

Figure 4. Class-F PA with quarter-wavelength line and series-tuned output.

$$R = \frac{R_o}{R_L^2} \quad (12)$$

at the drain. For an even harmonic, the short circuit on the load side of the transmission line is repeated, producing a short circuit at the drain. However, for an odd harmonic, the short circuit at the load is inverted, producing an open circuit at the drain. A resistive load is produced at the fundamental frequency.

At low drive levels, the FET acts as a current source. As drive increases, the FET enters saturation and harmonics are generated. Since the transmission line presents a high impedance to all odd harmonics, all odd harmonics can be present. The output can therefore increase until the drain-waveform is a complete square wave and the FET is saturated for a full half cycle. In this case, the FET acts as a switch rather than a saturating current source.

It is most convenient to analyze operation in terms of the fundamental-frequency drain-load impedance R. The fundamental-frequency component of the square-wave drain voltage is

$$V_{1m} = \frac{4}{\pi} V_{DD} \quad (13)$$

hence the power output is

$$P_o = \frac{8}{\pi^2} \frac{V_{DD}^2}{R} \quad (14)$$

The fundamental-frequency component of the drain current is

$$I_{1m} = \frac{4}{\pi} \frac{V_{DD}}{R} \quad (15)$$

Current flows only during half of the RF cycle and is composed of dc and even harmonics. It must therefore be a half sinusoid with peak

$$I_{dc} = \frac{1}{\pi} i_{D\,max} = \frac{8}{\pi^2} \frac{V_{DD}}{R} \quad (16)$$

The dc-input current is therefore

$$i_{D\,max} = 2I_{1m} \quad (17)$$

Comparison of (17) and (14) shows that the efficiency of an ideal PA of this type is unity at PEP.

The output voltage and current are delayed 90° from the fundamental-frequency components at the drain. The output voltage and current are TV_{1m} and I_{1m}/T, where

$$T^2 = \frac{R_o}{R} = \frac{R_o^2}{R_L^2} \quad (18)$$

Transmission Line with Series-Tuned Output

Kazimierczuk [9] describes a class-F PA with a quarter-wavelengthtransmission line and a series-tuned output circuit. This circuit (Figure 4) is the dual of that of Section 4 and represents the limit to improvement of the second-harmonic-peaking PA (Section 3) by additional resonators. At VHF, a series-tuned output network is generally considerably more practical than a parallel-tuned network.

The FET or other active device is driven to act as a switch. The series-tuned output circuit (L1-C1) presents to the transmission line a resistance at the frequency of operation and an open circuit at all harmonics.

The quarter-wavelength transmission line transforms the load impedance according to (12). For an even harmonic, the open circuit on the right side of the transmission line is repeated, producing an open circuit at the drain. However, for an odd harmonic, the quarter-wavelength line inverts the open circuit, producing a short circuit at the drain.

When the switch is open, $i_D = 0$. Since the drain-current waveform can be composed only of dc, fundamental-frequency, and odd-harmonic components, the only possible drain-current waveform is a square wave (Figure 4). When the switch is closed, $v_D = 0$. Since the drain-voltage waveform can be composed only of dc, fundamental-frequency, and even-harmonic components, the only possible drain-voltage waveform is a half sinusoid. The drain current and voltage waveforms are identical to those in the current-switching class-D PA (Figure 14-3 of [1]).

The average drain current must be the dc input current entering the PA through the RF choke, hence

$$i_{D\,max} = 2I_{dc} \quad (19)$$

Similarly, the average voltage on the drain must be the dc voltage applied to the input of the RF choke, hence

$$v_{D\,max} = \pi V_{DD} \tag{20}$$

From standard Fourier analysis, the amplitude of the fundamental-frequency component of the drain voltage is

$$V_{1m} = \frac{\pi}{2} V_{DD} \tag{21}$$

The power output is therefore

$$P_o = \frac{V_{1m}^2}{2R} = \frac{\pi^2}{8}\frac{V_{DD}^2}{R} \tag{22}$$

assuming that the transmission line is lossless. The quarter-wavelength line causes the output voltage to be phase-shifted by 90° relative to the fundamental-frequency components of the drain voltage and current. The amplitude V_{om} of the output in the actual load R_o can be determined by applying the impedance transformation (18) to (21).

The amplitude of the fundamental-frequency component of the drain currentis that of the drain voltage (21) divided by R. The application of standard Fourier-series analysis to the drain-voltage waveform also yields

$$I_{1m} = \frac{4}{\pi} I_{dc} = \frac{2}{\pi} i_{D\,max} \tag{23}$$

Putting (21) and (23) together yields

$$I_{dc} = \frac{\pi^2}{8}\frac{V_{DD}}{R} \tag{24}$$

Comparison of (22) and (24) again yields an efficiency of 100 percent for an ideal PA of this type.

General Load Network

At UHF and above, it is seldom possible to provide the right harmonic impedance (open or short) at more than a few frequencies. A number of recent class-F designs, for example, use transmission lines to control the second and third harmonics [10], [11]. In one case, dielectric resonators are used to implement a Tyler-type fifth-harmonic-peaking PA [12] at 900 MHz.

The characteristics are derived through a Fourier-series analysis [5] that relates the peak and fundamental-frequency component of the voltage orcurrent to its dc component. The resultant efficiencies and power-output capabilities for various combinations of contolled harmonics are given in Table 1. *RF*

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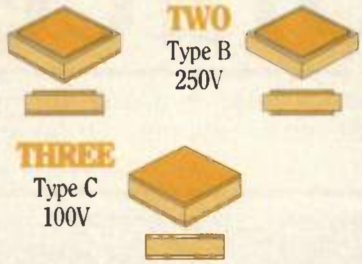

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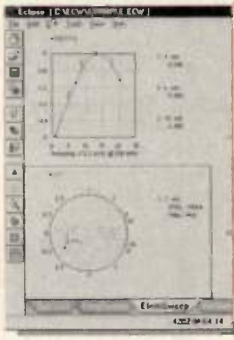

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Participate in the design of the RF and analog portion of a CDMA cellular phone. Requires a BS/MS in EE and 5+ years of related experience. RF circuit design experience, knowledge of RF components desirable. Board level design experience preferred. An exceptional new graduate will be considered. Job Code: JC-467

RF Associate Engineer or Technician

Assist RF design engineers to develop, build and test the RF/analog portion of the CDMA cellular phone. An AA in EE and 3+ years of related experience are necessary. Must be familiar with RF measurement and test equipment like network and spectrum analyzers. Must be able to learn new cellular test equipment. Job Code: JC-465

Sr. RF/Analog ASIC Engineer

Develop Bipolar/BiCMOS RF and analog ASICs for low power wireless applications like the IS-95 CDMA cellular phone. A BS/MS/PhD in EE and 5+ years of experience required. Must be familiar with all phases of RF and analog IC design. Experience with design tools of simulation and layout is essential. Understanding of design issues low phase noise VCO, third intermodulation, power management circuit design a definite plus. Job Code: JC-468

RF/Analog ASIC Engineers

Develop RF and analog ASIC. Requires a BS/MS in EE and 3+ years of experience and expertise in simulation tools. ADC and DAC design experience is preferred. IC modeling experience is advantageous. Job Code: JC-469

Product Managers

Define product features for the IS-95 cellular phone system, understand the market trend and monitor activities in the cellular industry. A BS/BA/MBA in EE, CS, Business or equivalent and 7+ years of cellular industry or product experience are essential. Must be familiar with the cellular industry in general. Knowledge of any specific area of the industry like CDMA is a definite plus. Job Code: JC-472

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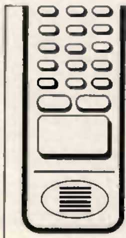

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Baseband Analog Circuit Design: Baseband analog circuit designers for cordless telephony systems at 900 MHz and 46/49 MHz. BSEE 5 yrs.

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RF Design Manager: Lead a team of RF engineers from initial design and implementation through product intergration and testing into high volume production. 8+ years of RF design with emphasis on low cost radio design. BS/MS

Sr. MMIC Design: Design highly integrated GaAs MMICs for advanced cellular products.Circuits to be designed include: power amplifiers, LNAs, mixers, IF amplifiers, buffer amplifiers.RF frequencies are 900 and 1800 Mhz

Product Line Manager Wireless: Specific responsibilities include product line strategic planning, establishing revenue and price objectives, setting internal cost targets and oversight of internal product realization schedules.

Sr. MMIC Design: Design highly integrated GaAs MMICs for advanced cellular products.Circuits to be designed include: power amplifiers, LNAs, mixers, IF amplifiers, buffer amplifiers.RF frequencies are 900 and 1800 Mhz.

Sr. Analog IC Designers: Responsible for conceptual circuit design and developing new analog/mixed signal ic's. BS/MS experience in A/D D/A ASIC's bipolar and BiCMOS.

Rf Systems Engineer PCS: Responsible for developing radio system performance requirements including modulations/demodulations, coding, channel models, deployment models, hardware performance requirements, interference rejection, blocking power control, handover ect BS/MS

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Senior RF Test Engineer (San Jose)
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Sr. RF Field Apps. Engineer (San Diego)
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ELECTRONIC COMPONENTS AND EQUIPMENT

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LABORATORIES

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CONSULTANTS

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