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October 1996

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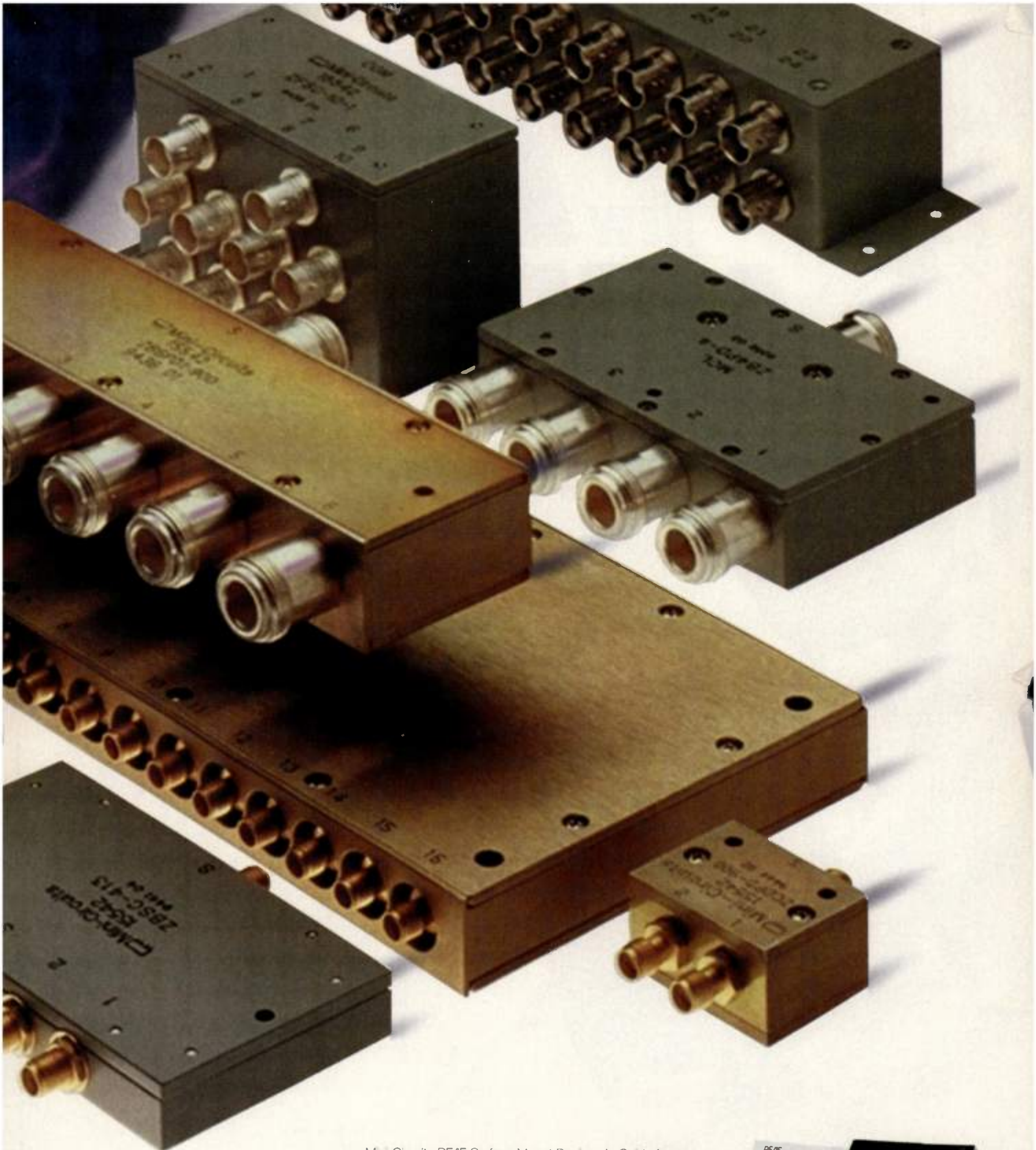

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ERA-3SM	DC-3000	19.4	11.0	4.5	23	2.15
ERA-4	DC-4000	13.9	▲19.1	5.2	▲36	4.15
ERA-4SM	DC-4000	13.9	▲19.1	5.2	▲36	4.20
ERA-5	DC-4000	19.0	▲19.6	4.0	▲36	4.15
ERA-5SM	DC-4000	19.0	▲19.4	4.0	▲36	4.20

Note: Specs typical at 2GHz, 25°C

▲ Typ. numbers tested at 1GHz. At 2GHz, Max. Pwr. Out may decrease by 0.4dB & IP3 by 3 to 4dB.

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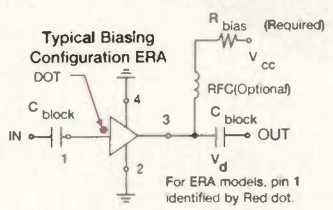
120x60	.002, .047, .068, .1 µf

ERA-1 ERA-1SM
ACTUAL SIZE



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Coming in November

- **Testing**
- **Shielding tutorial**

RF editorial

Education and information help engineers to remain active professionals

By Don Bishop
Editorial Director

Working engineers always are faced with the possibility of developing a knowledge gap compared to recent graduates. Some face the challenge by mapping a career path that includes an exit from front-line engineering work into sales or management. Many of those who achieve lifelong success as active engineers avail themselves of continuing education and immerse themselves, however selectively, in the enormous flow of information.

Associate Editor Patricia Werner gathers news about continuing education for posting in the "Courses" column. I'm always concerned that some courses that might be useful to you are not included, simply because we may not have heard about them. Won't you please let Pat know if you attend a specialized course that should be listed? Her E-mail address is pat_werner @intden.ccmail.compuserve.com. The postal address, telephone and fax numbers appear on page 10.

Information comes your way from sources too numerous to mention. I like to think that our magazine is one of your *better* sources. If you attend an event that isn't listed but should be, would you let Pat Werner know?

College courses, specialized courses, books, magazines, computer networks, conferences and expositions all vie for your attention and strive to be helpful to engineers. It can be a daunting challenge to spend time wisely in choosing and using the right mix of sources, but the difference is an extended career or an abbreviated one.

Spectrum wall chart

This column twice has mentioned the availability of a wall chart that displays spectrum allocations from 3 kHz to 300 GHz. Then the ordering information had to be updated. I'm mentioning it again because some readers still are having difficulty ordering the chart.

The National Telecommunications and Information Administration, an agency of the U.S. Department of Commerce, issued the chart. Here's how to obtain a copy: The product name is "1996 Spectrum Wall Chart"; the stock number is 003-000-00652-2; and the price is $3.25.

Orders can be placed by mail to "U.S. Government Printing Office, Superintendent of Documents, P.O. Box 371954, Pittsburgh, PA 15250-7954."

Inquiries and orders can be placed by telephone and facsimile; Tel. 202-512-1800; Fax 202-512-2250. Fax-on-demand orders can be placed with U.S. Faxwatch at 202-512-1716.

A word about ordering the chart: Some callers have been told that the stock number corresponds with a 1987 version of the chart. A representative of NTIA told us that the Government Printing Office (GPO) was *supposed* to discard remaining copies of the 1987 chart and replace them with the 1996 chart.

Computerized records that GPO representatives access when speaking with callers may not reflect the pairing of the stock number with the 1996 chart. If you are told that the stock number corresponds with the 1987 chart, now you know why. The folks at NTIA really have been helpful. Government procedures sometimes are difficult to overcome, though. But you know that going in, don't you? *RF*

INFO/CARD 6

RF design

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CDMA Mobile/Personal Station Testing

Minimum performance specifications such as IS-98 and PN3385 (ANSI-J-STD-018) ensure that the CDMA mobile (personal) stations perform satisfactorily under several channel impairments imposed on the system, namely noise, multipath fading, and interference.

Noise

To verify the performance of CDMA mobile station receivers, frame error rate (FER) is evaluated vs. carrier-to-noise density ($\hat{I}_{or}/I_{oc}$). See Fig. 1. This $\hat{I}_{or}/I_{oc}$ must be set accurately as it can dramatically change the FER measurements. It is important to note that it is not the accuracy of noise power density, but rather that of $\hat{I}_{or}/I_{oc}$ which is critical: ±0.2 dB $\hat{I}_{or}/I_{oc}$ accuracy is required for CDMA testing.

In order to set correct $\hat{I}_{or}/I_{oc}$ values, it is important to accurately set the ratios between $\hat{I}_{or}$ and traffic, sync, paging, or pilot power values of the base station.

The ability of the filters in the mobile station receiver to discriminate between random noise and the desired signal is tested by injecting Additive White Gaussian Noise (AWGN) into the forward channel as shown in Fig. 2 below. To properly test these filters, the bandwidth of the noise should cover the operating frequency range of the mobile receiver, typically about 15 MHz.

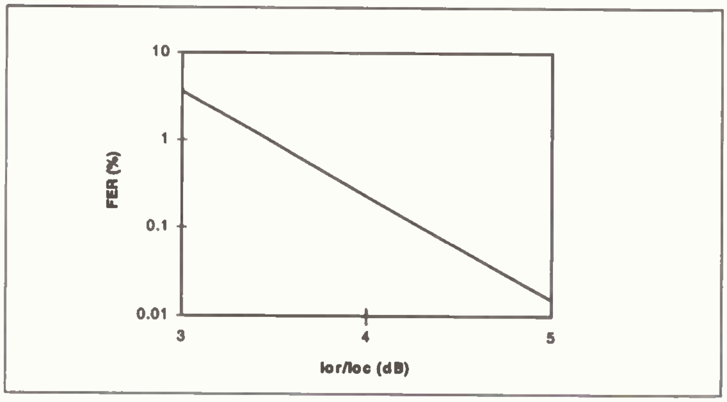

Figure 1. Typical CDMA FER vs. Ior/Ioc plot

Multipath Fading

A significant portion of the transmitted power can be lost due to fading (up to 30 dB). CDMA standards recommend using the Rayleigh model to test under various simulated multipath fading conditions.

During the FER measurements with multipath fading, confidence levels of 95% can require long test times, especially if the system nears the threshold rate. In order to eliminate ambiguity in the test results, the random fading statistics should not be repeated during the test.

Note from Fig. 2 that the AWGN Generator must be connected *after* the multipath fading emulator. If the AWGN Generator was placed before the fading emulator, then both the base station output signal and AWGN will undergo multipath fading. This will result in changes in the $\hat{I}_{or}/I_{oc}$ value, thus making it impossible to perform vaild tests .

CW Interference

Single-tone desensitization and intermodulation spurious tests measure the performance of a base station or mobile station receiver in the presence of CW interference. Attenuation on the reverse channel should be carefully chosen to establish calls, especially if you are working with base station simulators with a limited receiver dynamic range.

Open loop power control uses the power received at the mobile station as a reference and sets the sum of its receive and transmit powers for a constant level of -73 dBm. This means that if you are testing with high forward channel power, the reverse channel power will be very low, making it difficult for some base station simulators to receive the signal.

Under closed loop power control, because the power control bits are sent every 1.25 ms, the forward link should be maintained while the AWGN generator is calibrating to set the proper $\hat{I}_{or}/I_{oc}$ value. This is accomplished by a make-before-break bypass switch in the CDMA Mobile Station Interface Set.

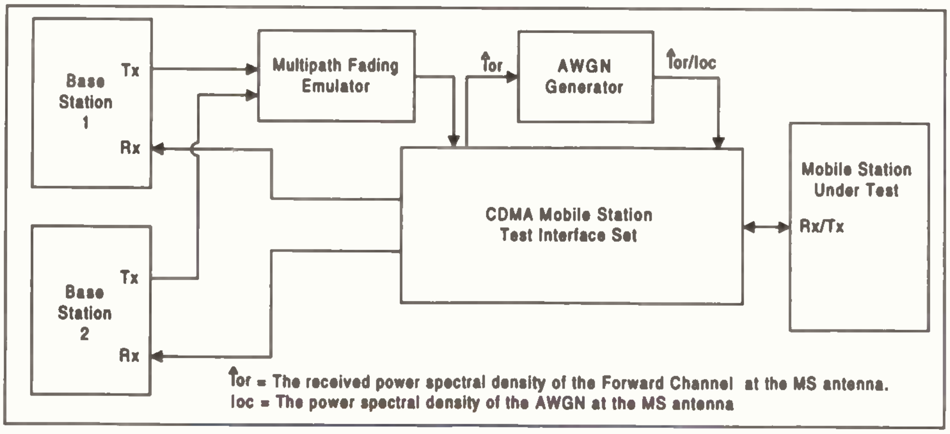

Figure 2. CDMA Mobile Station Test Setup

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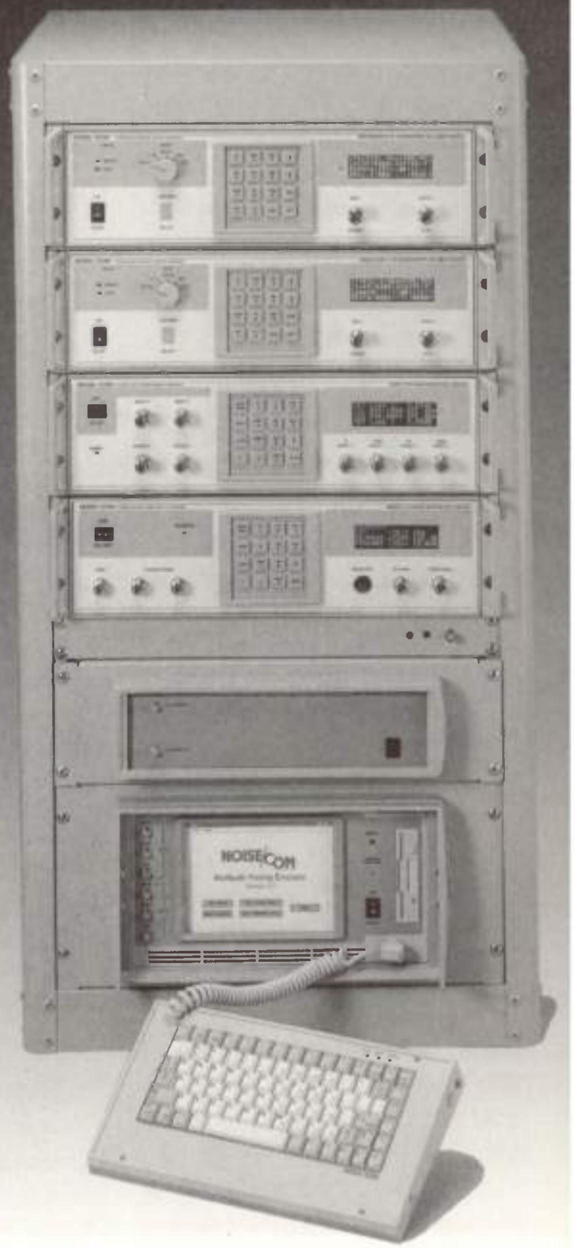
CDMA Mobile Station Test System (WIS-98/018)



CDMA Test Station for Cellular and PCS Mobile Subscriber Sets

The WIS-98 is an automated test station that evaluates the performance of CDMA mobile subscriber sets in design and production environments under realistic operating conditions.

It emulates channel impairments such as interference, multipath fading, and Additive White Gaussian Noise (AWGN) for base-to-mobile paths between one or two communication channels.

The WIS-98 includes a multipath fading emulator that models wireless channels and a precision AWGN generator that can precisely set E_b/N_o, C/N, C/N_o, and C/I ratios. The WIS-98 can be configured with a mobile station interference set to provide a complete impairment solution.

Ordering Information

WIS-98	CDMA Mobile Test System for Cellular (IS-98)
WIS-018	CDMA Mobile Test System for PCS (ANSI J - STD-018)
WIS-98/018	CDMA Mobile Test System for Cellular and PCS (Both standards

CDMA Mobile Station Test Systems Components:

- MP2500 Series Multipath Fading Channel Emulator with local oscillator
- UFX-BER Series Precision Carrier-to-noise Generator
- MS-800C/1900C Series Mobile Station Interface Set

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RF letters

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Transmission line standard

I have been involved in coaxial transmission line design since the mid-1930s. In your August tutorial on transmission lines, author Jim Weir presented some history of which I was unaware. I've long held a different belief regarding the *Zo* = 50 Ω standard being based upon British plumbing.

Minimum loss in a coaxial transmission line occurs at a diameter ratio of 3.591. This happens to be 76.7 Ω in the case of air, which was the most practical dielectric of an earlier era. Nevertheless, the 3.591 diameter ratio holds for all lossless dielectrics.

One of the earliest acceptable candidates for a low-loss flexible dielectric turned out to be polyethylene, which exhibited a dielectric constant of 2.26. When polyethylene is placed in a line with a 3.591 diameter ratio, the resulting *Zo* is 51 Ω.

Veteran RF designers will recall that in the pioneer days of flexible coaxial lines, there was a lot of 51 Ω as well as 52 Ω and 49 Ω line being manufactured. The reason is that impedances in this range exhibited the lowest loss for a given outer conductor diameter. The 50 Ω standard was almost certainly the eventual result of cooperation between the industry and the Armed Services Electromagnetics Standards Agency (ASESA).

Donald K. Goshay
Golden, MO

Mr. Weir's reply:

To answer the part about the plumbing reference, I quote from *Watts New From Bird,* Bird Electronic Company, Vol. 6, No. 2, August 1969.

"According to Consultant C. L. Rouault, who was present, an RMA committee in the 1940s recommended to the U.S. Navy that an impedance of approximately 50 Ω be selected the standard as a compromise between transmission parameters *as well as commercially available copper water tubing sizes.* One might muse on where the communication industry would be today *if the plumbers* had not come to the rescue." (Italics author.)

My further belief in the truth of this transcript was provided by Dr. R.C. Kent of Teledyne Ryan, San Diego, during my engineering "apprenticeship," who related his experiences in the first days of radar with British engineers who taught him the copper water pipe technique.

To answer the part about the various impedances would take another column of approximately the same length as the original. However, if you will permit a few jumps of math, we have:

Z(o) = (60 / sqrt (e)) * ln(D/d) as presented in the article. Remembering that

$P = E^2 / Z(o)$ we can get to $P = ((E^2)(D^2)(\ln(D/d))) / 480 (D/d)^2$

Differentiating for V with respect to D/d and setting equal to zero yields the inflection point 60 Ω (maximum voltage carrying capability).

Differentiating for P with respect to D/d and setting equal to zero yields the inflection point 30 Ω (maximum power handling capability).

Reader Goshay is precisely correct for the minimum attenuation point of 76.7 Ω, but the power and voltage handling capability of the coax at this point has deteriorated to the point where 72 Ω cable is used, as stated, in the cable TV industry, where significant power isn't a consideration. As to the selection of 72 Ω coax as the standard (instead of 77), my unconfirmed suspicion is that it was chosen to match to the 72 Ω impedance of a quarterwave dipole in free space.

Jim Weir VP Engineering
RST Engineering
Grass Valley, CA

More on coax

I work in TV broadcasting and found the August article on coaxial cables interesting. I have questions for which I've never been able to find answers.

We have a 1,000 ft tower, and we use cable with a line impedance of 75 Ω. I was told that the line is cheaper because it has a smaller inner conductor than 50 Ω cable. If the losses are less for 75 Ω cable, why do we use 50 Ω line to the satellite dishes with 250–300 ft runs?

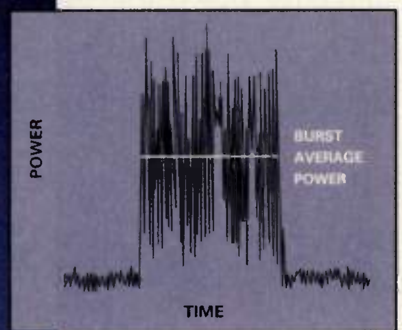
For practical reasons, 50 Ω or 75 Ω line can be used, but why not go for the best? In broadcasting, 50 Ω has been the standard for RF, whereas 75 Ω has been used for video. Why 50 Ω? Because this is the way we have always done it? Is it the impedance of an old broadcast vertical? Jim Weir's article may be more correct. A transmitter can be made to feed either, as well as the antenna. RG-8 and RG-11 look the same, but when they were mixed in sampling lines used for the tuning of a TV transmitter, the results were interesting.

Regarding high-definition television (HDTV), bit-error ratios and the tight tolerances that we must maintain in the transmitter plant, what sort of tolerances have to be maintained at the home receiver? FM and TV antennas could be either vertical or horizontal. My understanding is that this helps with the rejection of the unwanted signals when between towns.

For UHF, why not use FM as opposed to AM, the way they do in the satellites? In my mind, it would make things a lot easiser. I have never seen where this has been tested.

Dave Lawry
WHDH TV, Boston

Giga-tronics Introduces The 8540C Universal Power Meter



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RF calendar

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14–18 Test & Evaluation: Answering the Challenge—*Seattle.* Information: Chairman. Tel. 206-655-4832; Fax 206-655-7929; E-mail tethc@pony5.express.ds.boeing.com; Web site http://www.boeing.com/itea.

16–17 Practical Spectrum Measurements Seminar—*Portland.* Information: Brochure 9610, JMS, P.O. Box 25170, Portland, OR. Tel. 503-292-7035; Fax 503-292-0449.

21–23 RF Design Seminar Series—*Wakefield, MA.* Information: Intertec Presentations, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel. 303-220-0600; Fax 303-770-0253.

22–24 Wescon '96 Technical Conference—*Anaheim, CA.* Information: Wescon, 8110 Airport Blvd., Los Angeles, CA 90045. Tel. 800-877-2668 or 310-215-3976 ext. 243; Fax 310-641-5117; E-mail wescon@ieee.org; Web site http://www.wescon.com.

23 Battelle's Technology Intelligence Program—*Chicago.* Information: B-TIP. Tel. 614-424-4244; Fax 614-424-4260; E-mail BTIP@battelle.org.

28–29 Testing Product Power Quality—*Washington.* Information: Ergonomics, P.O. Box 964, Southampton, PA 18966. Tel. 800-862-0102 or 215-357-5124; Fax 215-364-7582.

29–30 Radio Solutions '96—*Birmingham, UK.* Information: LPRA Secretariat, Walker Mitchell Ltd., Brearley Hall, Luddenden Foot, Halifax HX2 6HS, United Kingdom. Tel. and fax +44 (0) 1422 88 69 50.

29–31 Signal Processing Applications and Technology—*Santa Clara, CA.* Information: DSP Associates, 49 River St., Waltham, MA 02154. Tel. 617-891-6000; Fax 617-899-4449; E-mail icspat@dspnet.com.

30 ELF/VLF Magnetic Fields and the New MPR3—*Washington.* Information: Ergonomics, P.O. Box 964, Southampton, PA 18966. Tel. 800-862-0102 or 215-357-5124; Fax 215-364-7582.

November 4–6 Northcon Conference & Exhibition—*Seattle* Information: Electronics Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045-3194. Tel. 800-877-2668 ext. 243 or 310-215-3976; Fax 310-641-5117; E-mail northcon@ieee.org; Web site http://www.northcon.org.

6–8 Plastics in Portable & Wireless Electronics—*Phoenix.* Information: Judy Wales, Donnelly. Tel. 520-321-7680; Fax 520-322-5635; E-mail judy@donnelly.ppp.theriver.com.

18–19 RF System Engineering—*Feltham, Middlesex, UK.* Information: Geoff Varrall, RTT Systems Limited, Enterprise House, Central Way, North Feltham Trading Estate, Middlesex TW14 0RX, United Kingdom. Tel +44 (0) 181 844 1811; Fax +44 (0) 181 751 2616.

18–22 IEEE Global Telecommunications Conference—*London.* Information: Vikki Pollard, Globecom, RT 10/5a, BT Laboratories, Martlesham Heath, Ipswich, Suffolk, IP5 7RE, United Kingdom. Tel. +44 1473 644799; Fax +44 1473 647488; E-mail pollarv@btlip10.bt.co.uk.

December 2–3 RF Design—*North Feltham, Middlesex, UK.* Information: Geoff Varrall, RTT Systems Limited, Enterprise House, Central Way, North Feltham, Middlesex TW14 0RX, UK. Tel +44 (0) 181 844 1811; Fax +44 (0) 181 751 2616.

8–11 International Electron Devices Meeting—*San Francisco.* Information: Melissa Widerkehr, IEDM, 101 Lakeforest Blvd., Suite 270, Gaithersburg, MD 20877.

9–11 DSP Without Tears—*Richardson, TX.* Information: Z Domain Technologies, 555 Sun Valley Drive, Suite A4, Roswell, GA 30076. Tel. 800-967-5034 or 770-587-4812; Fax 770-518-8368; E-mail dsp@zdt.com. Web site http://zdt.com/~dsp.

9–12 RF and Microwave Measurements and Applications—*Monterey, CA.* Information: Joleen Packman, University Consortium for Continuing Education, 16161 Ventura Blvd., M/S 752, Encino, CA 91436. Tel. 818-995-6335; Fax 818-995-2932; E-mail UCCE@AOL.COM.

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January 15–17 DSP Without Tears—*Long Beach, CA.* Information: Z Domain Technologies, 555 Sun Valley Drive, Suite A4, Roswell, GA 30076. Tel. 800-967-5034 or 770-587-4812; Fax 770-518-8368; E-mail dsp@zdt.com. Web site http://zdt.com/~dsp.

February 10–14 Wireless Symposium and Exhibition—*Santa Clara, CA.* Information: Penton Publishing. Tel. 201-393-6256.

March 3–5 CTIA Wireless—*San Francisco.* Information: Dobson & Associates. Tel. 202-463-7905.

13–19 CeBIT '97 World Business Center: Office, Information and Telecommunications—*Hannover, Germany.* Information: Mette Fisker Peterson, Hannover Fairs USA, 103 Carnegie Center, Princeton, NJ 08540. Tel. 609-987-1202; Fax 609-987-0092.

April 22–24 International Wireless Communications Expo—*Las Vegas.* Information: Intertec Presentations, 6300 S. Syracuse Way, Denver, CO 80111. Tel. 800-288-8606 or 303-220-0600.

RF courses

1996 CEI-Europe in Baveno, Italy—*Mobile and Wireless Personal Communications Networks* —Oct. 14–18; *Modern Digital Modulation Techniques*—Oct. 14–18; *Modeling and Simulation of Communication Systems*—Oct. 15–18; *Speech and Channel Coding for Mobile Communication*—Oct. 21–23; *Digital Cellular and PCS Communications: The Radio Interface*—Oct. 21–25; *Spread-Spectrum and CDMA*—Oct. 21–25. **in Barcelona, Spain**—*Mobile Cellular and PCS Telecommunications Systems*—Nov. 11–13; *RF Circuit Components*—Nov. 18–21; *Receivers and Transmitters*—Nov. 18–22; *Advanced Digital Receivers for Wireless Communications*—Nov. 18–21; *Cellular and Personal Communications Infrastructure*—Nov. 20–22. Information: CEI-Europe, P.O. Box 910, S-612 25 Finspong, Sweden. Tel. 46 122 175 70; Fax 46 122 143 47; E-mail cei.europe@one.se.

Celwave University—Marlboro, NJ. Modules offered include: *Antenna Basics, Advanced Theory, Towertop Amplifiers, Bidirectional Amplifiers, Filters* and *Combiners*. Information: Gail Magid, Celwave, 2 Ryan Road, Marlboro, NJ 07746-1899. Tel. Sales Engineering Dept. 800-235-9283.

CKC Laboratories—*Core EMC Design*—Oct. 22–23, Hillsboro, OR; Jan. 14–15, 1997, Orange County, CA; March 11–12, Fremont, CA; *Immunity to ESD*—Nov. 18, Fremont, CA; Feb. 3, 1997, Seattle; *CE Mark Design and Compliance Routes*—Nov. 19–20, Fremont, CA; Feb. 4–5, 1997, Seattle; *EMC for Medical Electronics*—Nov. 12–13, Orange County, CA. Information: Linda Grunow or Todd Robinson, CKC Laboratories, 5473-A Clouds Rest, Mariposa, CA 95338. Tel. 800-500-4362 or 209-966-5240; Fax 209-742-6133; E-mail lgrunow@ckc.com.

Communications Tech Training—1996 schedule for Orland Park, IL, Nov. 5–7, Dec 3–5, Information: Andrew Corp., Dept. 355, P.O. Box 9000, San Fernando, CA 91341-9978. Tel. 800-255-1479 ext. 117.

George Washington University—Washington. *Wireless Infrastructure Network Engineering for Cellular, PCS, LE, and WPBX*—Oct. 21–25. Information: George Washington University, Continuing Engineering Education, Academic Center, Room T-308, 801 22nd St. N.W., Washington, DC 20052. Tel. 202-994-6106 or 800-424-9773; Fax 202-872-0645; E-mail ceepinfo@ceep.vpaa.gwu.edu.

Georgia Tech Continuing Education—*RF and Wireless Engineering; Radar Cross-section Reduction*—Oct. 29–Nov. 1. *Principles of Modern Radar*—Nov. 4–8. *Infared Technology and Applications*—Nov. 5–8. *Far-field, Anechoic Chamber, Compact and Near-field Antenna Measurements*—Nov. 12–15. *Guidance, Navigation and Control*—Nov. 19–22. Information: Dept. of Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332-0385. Tel. 404-894-2547; E-mail conted@gatech.edu; Web site http://www.conted.gatech.edu.

Henry Ott Consultants—*Electromagnetic Compatibility Engineering*—Oct. 14–16, Palo Alto, CA. Information: Henry Ott Consultants, 48 Baker Road, Livingston, NJ 07039. Tel. 201-992-1793; Fax 201-533-1442.

IEEE—Huntsville, AL. *Modern Radar Fundamentals: Signal Processing, Components, Tracking and Detection*—Nov. 18; *Practical Phased-array Systems: Fundamentals,* *Errors, System Issues, Elements, Feeds, Limited Scan Techniques*—Nov. 19. Information: Boston IEEE AESS, 282 Marrett Road, Lexington, MA 02173. Tel. 508-440-4007 or (617) 862-7014; Fax 508-440-4040; E-mail Eli_Brookner@ccmail.ed.ray.com.

Johns Hopkins University Whiting School of Engineering—courses in Washington, DC. *Wireless Digital Communications Systems: Specification, Test and Evaluation*—Oct. 21–23; *The Telecommunications Revolution and its Impact on Organizational Planning*—Oct. 28–30. Information: OEI 14515 Barkwood Drive, Rockville, MD 20853. Fax 301-871-4942; E-mail info.oei@apl.jhu.edu.

Learning Tree International—*Wireless Networks and Mobile Communications*—Oct. 15–18, Washington; Dec. 3–6, Toronto; Dec. 17–20, Washington; Jan. 14–17, 1997, Washington. Information: Learning Tree International, 1805 Library St., Reston, VA. Tel. 800-850-9197 or 703-709-9119; E-mail uscourses@ learningtree.com; Web site http://learningtree.com.

National Institute of Standards and Technology—*Microwave and RF Measurements for Wireless Communications*—Dec. 3–4. Information: Robert Judish, NIST, 325 Broadway, Boulder, CO 80303. Tel. 303-497-3380; Fax 303-497-3970; E-mail judish@boulder.nist.gov.

Technology International—*Achieving and Maintaining Compliance with the Medical Devices Directive*—Oct. 15–16, Boston; Nov. 20–21, Anaheim, CA; January 14–15, Denver; February 11–12, Dallas. Information: Kristin Eckhardt, Marketing Manager, Technology International, 609 Twin Ridge Lane, Richmond, VA 23235, Tel. 804-560-5334; Fax 804-560-5342; E-mail Eckhardt@TechIntl.com.; Web site www.TechIntl.com.

Tektronix—*CDMA Modulation Technologies and Measurements; Deploying Digital Transmission in Cabled Networks; TDMA (IS-136 and PCS 1900) Technologies and Measurements*—Two-day seminars; Cities include Atlanta; Baltimore; Dallas; Los Angeles; Montreal; Orlando; Parsippany, NJ; Philadelphia; Phoenix; Raleigh, NC; Santa Clara, CA; Seattle; Toronto; Vancouver; Washington. Information: Tel. 800-763-3133; Fax 800-835-0025; E-mail TEKFORM2@TEK.COM. Web site http://www.tek.com/Measurement/Support/Seminars.

UCLA Extension—Los Angeles. *Spread Spectrum Wireless and IS-95 CDMA Digital Cellular Communications*—Oct. 21–23; *Radar Interferometry: Principles and Applications*—Nov. 18-20; *Communication Systems Using Digital Signal Processing*—Nov. 18–22; *Advanced Digital Communications: the Search for Efficient Signaling Methods*—Dec. 2–3. Information: Dept. of Engineering, Information Systems and Technical Management, UCLA Extenion, 10995 LeConte Ave., Suite 542, Los Angeles, CA 90024. Tel. 310-825-1047; Fax 310-206-2815; E-mail mhenness@unex.ucla.edu.

University of South Florida—Clearwater, FL. *Short Course on RF and Microwave Measurements for Wireless Applications*—Dec. 3–4. Information: Dept. of Engineering, 4202 E. Fowler Ave., ENB 118, Tampa, FL 33620. Tel. 813-974-2574; Fax 813-974-5250; E-mail dunleavy@eng.usf.edu.

Engineering experiment station receives funding

Very Large Scale Integration (VLSI) Telecommunications Research Center (VTRC) at Texas A&M University has received funding to improve the efficiency of cellular phones, pager systems and other wireless technology. The center will merge the disciplines of communications and microelectronics to enhance interaction between algorithm design and implementation teams.

One area of research will be the development of algorithms and VLSI chips for spread-spectrum digital cellular technology. Another goal is to reduce the current discrete components in wireless systems into more portable monolithic systems. Five research assistantships have been awarded to Ph.D. students.

ISA open certification testing period extended

The international society for measurement and control (ISA) certification board has extended the open testing period for the certified control systems technician (CCST) program from June 30, 1996, to June 30, 1997. Qualified measurement and control technicians will have an additional year to test at any of the three levels of certification without taking preceding examinations. Level I requires a five-year combination of education, training and experience; Level II requires seven years; and Level III requires thirteen years. After June 30, 1997, any technician who has not applied for the CCST program will have to begin with Level I testing, regardless of experience, education and training.

Contracts:

Alabama gets Astro trunked digital system—The Calhoun County Emergency Management Agency has awarded Motorola a $14.2 million contract for an Astro 800 MHz trunked digital communications system. Thirteen microwave sites and six 20-channel simulcast sites will serve Calhoun and Talladega counties. It also will aid response and recovery efforts should a chemical emergency occur at the Army Depot in Anniston, where 2,000 tons of nerve gas and other materials are stored. According to Motorola, this is the first Astro 800 MHz trunked digital system scheduled for installation in Alabama. It includes 2,260 base stations and mobile and portable Astro radios, giving local officials the capability to coordinate personnel quickly and efficiently. Those who routinely communicate with each other may do so privately using radio talkgroups. In an emergency, the talkgroups can be reconfigured. The mobile and portable radios' emergency dispatchers can identify radios and the people using them by the IDs that appear on the dispatchers' CRT screens. The portable units are designed to communicate effectively even when used within buildings. Dispatchers at participating agencies will work from new Centracom Series II Gold series consoles.

The communications system is being funded by the Chemical Stockpile Emergency Preparedness Program (CSEPP). Ram Communications Consultants, Woodbridge, NJ, will manage system installation and acceptance testing for the counties.

Micron to develop baggage matching system—Micron Communications signed a cooperative research and development agreement with the Federal Aviation Administration (FAA) to develop a positive passenger baggage matching (PPBM) system. The system will recognize baggage that has been placed on an aircraft without an associated passenger, an objective that received praise from the chairman of the Congressional and White House task force on terrorism.

Micron will use remote intelligent communications (RIC) technology to design a security system that will track passengers and baggage. RIC units are different from RFID tags because the RIC units have a central processing unit (CPU) memory and tiny microwave radio on board to receive, process, store and transmit data. This allows the RIC units to perform more applications than RFID tags that use older technology at lower frequencies.

Business Briefs

Plastic filter development alliance—Teledyne Electronic Technologies, which patented and commercialized the process of manufacturing silver-plated plastic filters, is collaborating with Allgon Enterprises to develop and manufacture light-weight plastic and dielectric filters and duplexers for the code-division multiple access (CDMA) market. The two companies will also produce plastic integrated assemblies of filters and low-noise amplifiers (LNAs) for advanced mobile phone systems (AMPS) and personal communications services (PCS).

CEMA calls for spectrum allocation to wireless area networks—In a filing before the Federal Communications Commission (FCC) this week, the Consumer Electronics Manufacturers Association (CEMA) declared that it supports allocating spectrum to permit the creation of wireless area networks. Backing a petition by Apple Computer, CEMA requested that spectrum in the 5.15–5.35 GHz and in the 5.725–5.875 GHz bands be set aside for a new category of unlicensed equipment, called NII-Supernet devices.

Unlicensed community networks would improve consumer access to the national information infrastructure (NII). The NII-Supernet devices would bring the benefits of wireless local area network (LAN) technology to consumers. The FCC proposes to subject NII-Supernet devices to only minimum technical standards, thus allowing them to operate at higher power ratings in the upper band, and to adopt Part 16 rules for these devices to more thoroughly protect them from interference.

CEMA will work with equipment designers and manufacturers to develop consensus protocols that will maximize efficient use of the allocated bands.

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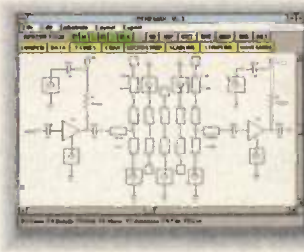

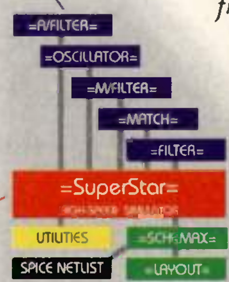

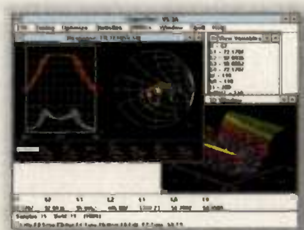

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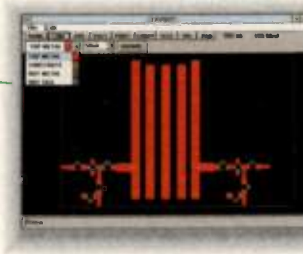

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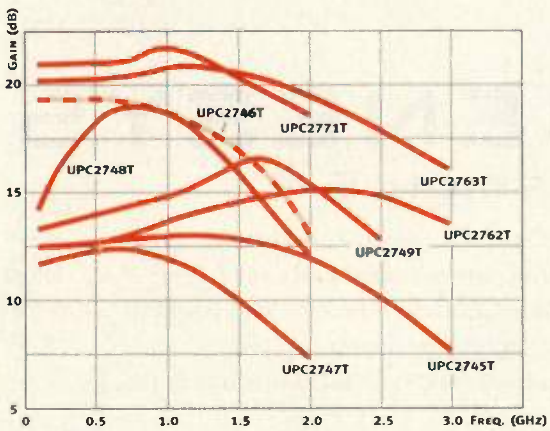

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UPC2709T	50MHz–2.3GHz	23	5	+7.5	25	1.0GHz
UPC2710T	50MHz–1.0GHz	33	3.5	+7.5	22	500MHz
UPC2711T	50MHz–2.9GHz	13	5	-3	12	1.0GHz
UPC2712T	50MHz–2.6GHz	20	4.5	-2.5	12	1.0GHz
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UPC2746T	50MHz–1.5GHz	19	4	-4.5	7.5	500MHz
UPC2747T	100MHz–1.8GHz	12	3.3	-11	5	900MHz
UPC2748T	200MHz–1.5GHz	19	2.8	-8	6	900MHz
UPC2749T	100MHz–2.9GHz	16	4	-12.5	6	1.9GHz
UPC2762T	100MHz–2.9GHz	14.5	7	7	27	1.9GHz
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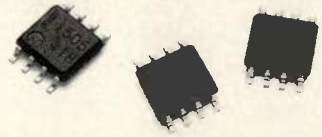

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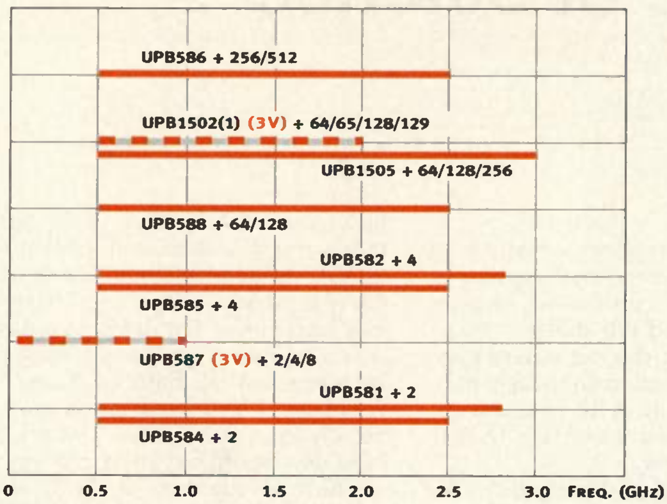

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UPC2758T[1]	100 - 2000	11	17	+6
UPC2753GR[1]	DC - 400	6.9	79	-17
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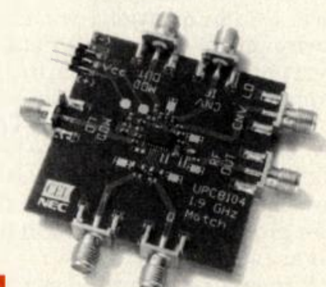

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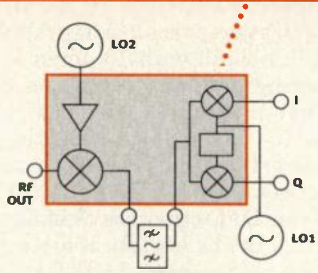

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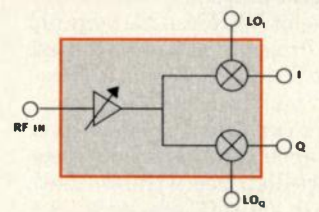

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RF technology on the cutting edge of science

By Ernest Worthman

Most of the development in computer technology is being applied in the performance area: faster central processing units (CPUs) and random access memory (RAM), more bandwidth in serial, parallel and disk input-output (I/O) and incremental improvements in standard technology. A number of technologies are being applied to wireless data communications within the PC environment. First come data communication over telephone lines. Then came cellular data transmission and satellite data transmission. There still are a number of bugs in the latter two, and they will take a bit more time to develop. The new area ready for explosive growth is wireless local area networks (wireless LANs).

Wireless LANs are a hot technology, and mobile offices are the current rage. The three largest players, Lucent Technologies, Aironet Wireless Communications and Proxim, have combined their efforts to create common specifications that will make each of their wireless networks compatible. The FCC has been talking about setting aside spectrum for short-distance wireless networks. Wireless network adapters that would connect mobile data users sell for about $1,000 but prices are expected to drop to around $300 in the next year or two.

Wireless LANs are expensive to implement, they have low performance and they are hampered by a lack of compatability standards. The current IEEE standard 802.11 isn't expected to be finished until mid-1997.

Cellular technologies such as circuit-switched cellular (CSC) and cellular digital packet data (CDPD) both use the cellular infrastructure. They both have coverage problems. Cellular modems are still expensive. Much of the personal computer card (PC card)—formerly Personal Computer Memory Card International Association (PCMCIA)—standards do not seem to work as well as they should.

Medicine

Because of tremendous advances in diagnostic and emergency medicine, real-time access to unique or specialized procedures and the ability to relay patient condition during emergency transport will become increasingly critical. It's a safe bet that RF applications in emergency medical services (EMS) will continue to grow.

Because of ongoing developments in digital technology, the amount of data that can be transferred over wireless links will increase. Current applications use mostly analog VHF, UHF and 800 MHz frequency bands and generally are limited to sending critical data such as electrocardiograms (ECGs) and vital statistics. With the continued development of low-power consumption integrated circuits (ICs), single-chip application-specific integrated circuits (ASICs) built with 10 µm or less circuit spacing and, one of the more exciting developments in wireless telephony, the 10-, 12- and 14-bit digital-to-analog converters (DACs) (such as Analog Devices' AD976X family of 10-, 12, and 14- bit DACs that are clocked up to 100 MHz), the bandwidth issue is becoming more manageable. Look for these technologies to give EMS teams the ability to transmit not only data, but video as well.

Video will open a new world of opportunity and direction for RF designers. The video data may be still photos from a fixed or hand-held external camera or real-time, streaming data from an endoscopic instrument used by the EMS team to probe the victim internally. This means that the attending physician can have real-time, actual, on-site video sent from the trauma site or from the transport vehicle, and video can place the trauma physician on-site with the emergency medical team.

Radio telemetry and wildlife

For the past six years, telemetry's greatest use has been for studying and managing wildlife. Most of the remote monitoring was done with analog radio at VHF frequencies, and much of the equipment was large and bulky and had high-power requirements. Advances in telemetry equipment design has been exponential. Both UHF and VHF frequencies are used much more diversely than ever before. Devices have been developed that cover a much wider range of applications. Technologies such as pulse-code modulation (PCM) and time-domain multiplexing (TDM) have been combined with DACs, and personal computers are being used to analyze the data.

The next generation of telemetry radios will use multiple stages, long-life modules with synthesized transmitters and receivers, analog or digital inputs and programmable modes, such as sleep and hot standby. The modules will switch power levels automatically to conserve power. They will use surface-mount technology (SMT) to make them reliable, miniature and full-featured. They will be field-programmable for changing functions, and these functions will include temperature-sensing, activity-sensing, sound transmission and mortality-sensing, to name a few.

Other wireless technologies that were too costly or too difficult to implement in the past will become more promising because costs will come down. Very large scale integration (VLSI), cost-effective manufacturing and off-the-shelf components will contribute to making their technologies appealing. Because of its relative immunity to noise and interference, spread-spectrum modulation is likely to become important in the science of wildlife management.

Once the low-earth orbit (LEO) technology is de-ployed and working, satellite use will offer yet another option for science to use RF to gain a better understanding of the animal world.

RF

Use adaptive digital predistortion to simulate a linearization system

By Roman Gloeckler

Non-public (municipal) mobile communication systems are confined to limited frequency resources. For this reason, adjacent-channel interference may be severe. Such interference must be suppressed to guarantee proper operation when all channels are busy. In the digital communication system considered here, the π/4-DQPSK (digital quadrature phase-shift keying) modulation scheme is used. This modulation has no constant envelope; therefore, unwanted intermodulation distortions are generated, especially in the transmitter's nonlinear power amplifier. These unwanted intermodulation distortions cause unwanted harmonic products known as *spectral regrowth* in the adjacent channels. Because of small channel bandwidths and close channel spacing, filtering is not an optimum choice for eliminating interferences. Transmitter linearization is a promising way to fight the problem of the aspect of demanded power efficiency. Each linearization method has specific advantages and drawbacks. Well-known methods include the *Cartesian loop feedback*, *feedforward linearization* and *predistortion* techniques. The simulation of a linearization system that uses adaptive digital predistortion is described in the following section.

Adaptive digital predistortion

Figure 1 shows the necessary components for a transmitter with linearization. The whole structure can be subdivided into two sections: a signal processing unit operating in the baseband and an RF section consisting of a transmitter and a receiving path. Signal processing usually is accomplished by a digital signal processor (DSP). Between both sections, the interfaces are analog-to-digital and digital-to-analog converters.

The object is to predistort the signal in such a way that at the transmitter output, a linear amplification for all possible amplitudes is achieved. In addition, the phase has to be constant over the entire amplitude range. Predistortion compensates for the nonlinear AM–AM and the AM–PM characteristic of the power amplifier and all other components in the transmitter chain.

An appropriate method of choosing predistortion limits is to use a look-up table that contains a row of coefficients. To modify the input signal, each specific signal level (the quantization depends on the table size) is multiplied by a certain coefficient. Similarly, the phase is altered by a second table. Both tables can be thought of as a single table with complex values.

To generate these tables, a portion of the transmitter output signal is fed back and transferred to the baseband. In the signal processing unit, the ideal input signal is compared with a sample of the output signal. By means of an iteration algorithm, every single table value is altered as long as an error limit is reached when comparing both signals. In the simulation, a table size of 512 values is sufficient.

An important feature of the adaptive predistortion is the necessity of two op-

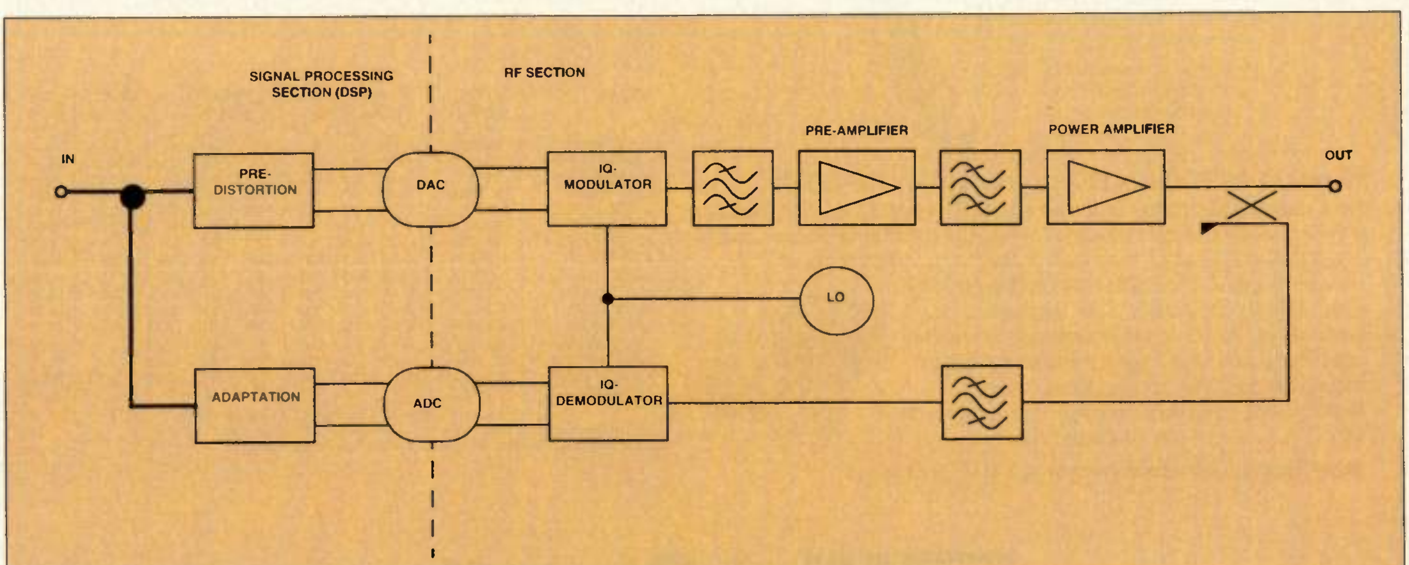


Figure 1. Block diagram of a transmitter using adaptive digital predistortion.

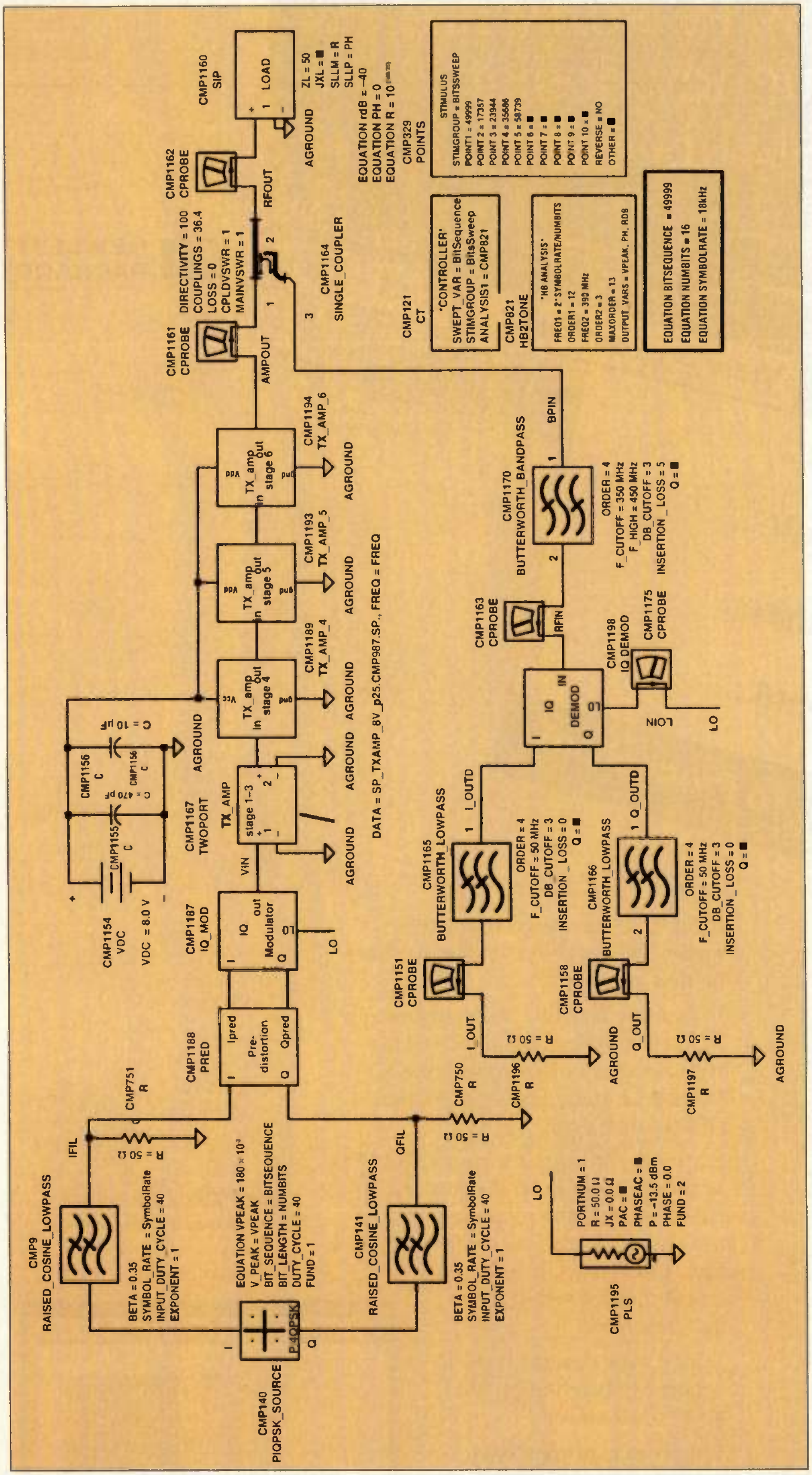


Figure 2. Block diagram of the RF-simulation set-up.

eration modes. The *table-generating mode* must be used to establish the proper table values before the regular *transmitter mode* can be turned on. During table generation, any transmitter operation must be avoided. Because of the two modes, predistortion exhibits an important advantage in comparison with other methods. The predistortion is unconditionally stable because no signal feedback occurs during the normal transmitting time interval. Unfortunately, this fact is coupled with the disadvantage of not being able to suppress interferences in real time. Therefore, the time interval between two table generating cycles has to be small enough to follow any disturbing influences.

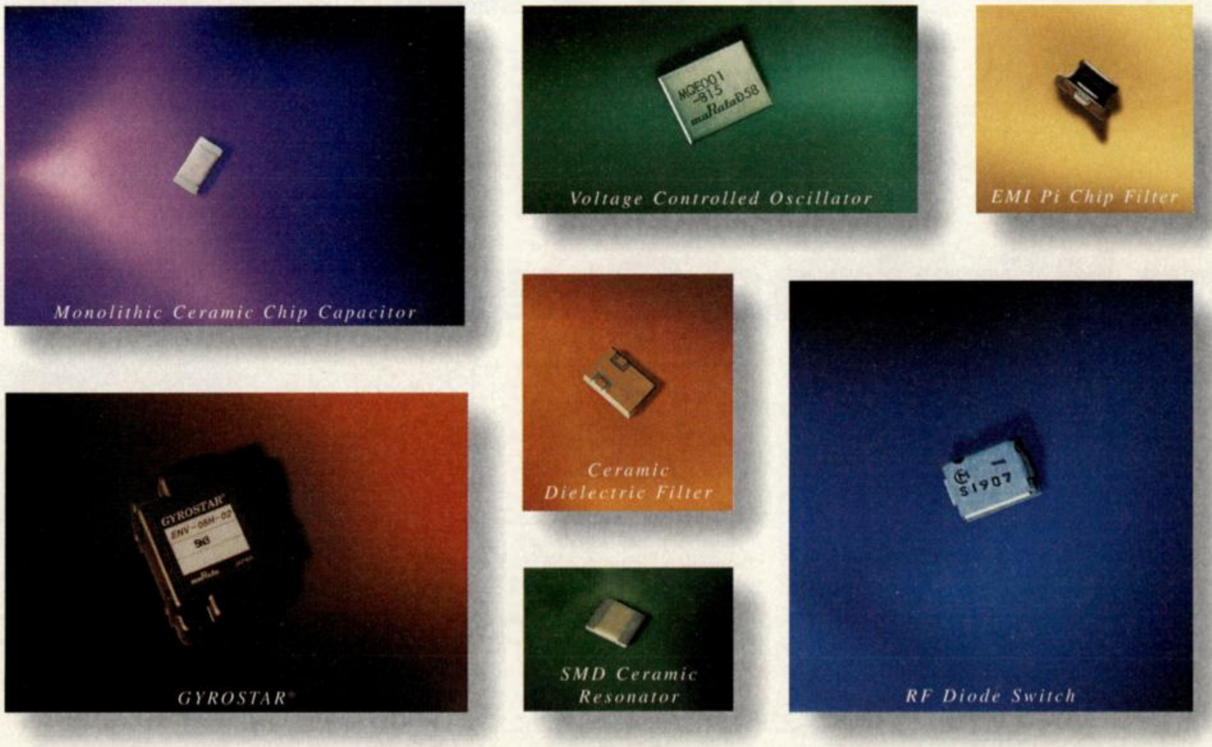
Simulation set-up

As already shown in Figure 1, the linearization system can be divided into two sections, a *signal processing unit* and an *RF section*. There have been no simulation tools available capable of modeling both sections in a single simulation environment with the required quality; therefore, two separate simulation tools were used. The algorithm for signal processing in the baseband was accomplished by means of a mathematics program in which the RF section was modeled by an RF-simulator tool. In principle, the following procedure should be applied to carry out a combined simulation.

First, the nonlinear transfer characteristic of the overall signal path from the in-phase quadrature (IQ) modulator to the IQ demodulator is determined by the RF simulator. This can be done by using a linear ramp sweep at the I-input of the modulator. Because of the AM–PM conversion of the power amplifier and to the phase imbalances of the modulator and demodulator, a complex output signal (I-and-Q channel) results. The number of the amplitude values measured must be equal to the desired table size.

These output data are transferred to the signal processing program that is generating the tables. Two separate tables are generated, one table for the predistortion of amplitude, the other for phase predistortion. Both tables are loaded into the RF-simulator and are made available there as datasets.

At this point, the simulation of the transmitter function can be started. Of main interest are the influence of the non-ideal components' characteristics



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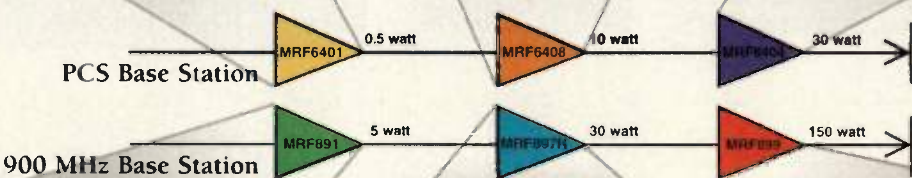

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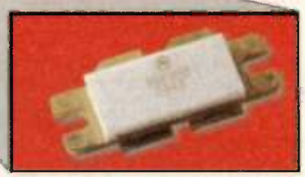

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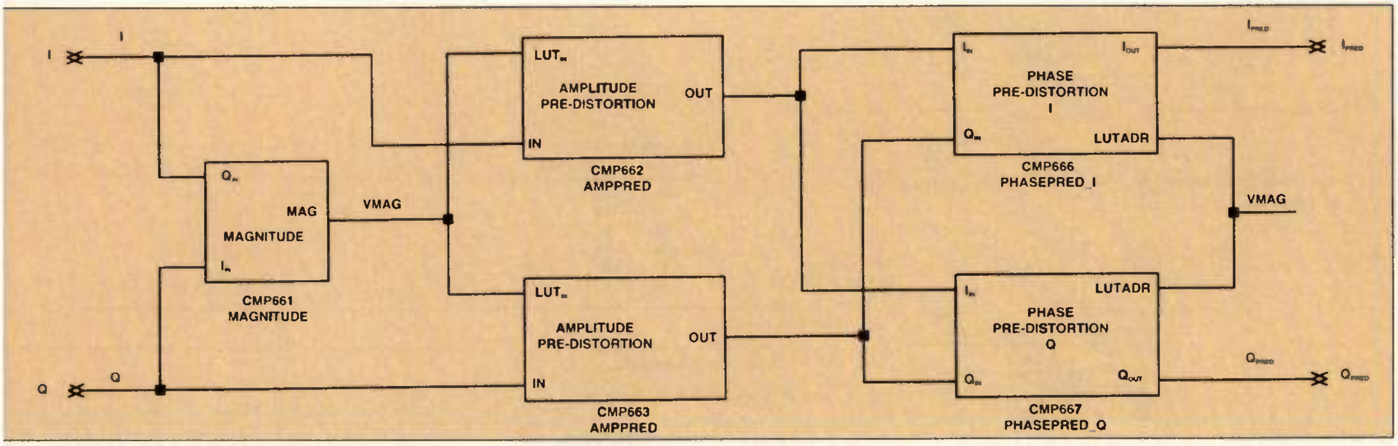

Figure 3. Structure of the predistortion block.

and the influence of variables such as surrounding temperature and supply voltage.

RF section

For modeling the system's RF components, a mixed procedure is applied. The precise modeling of all components at transistor level would be too demanding. It is not necessary for all components because the specifications of certain components are determined by means of simulations; therefore, some components are described in the form of behavioral models. Regardless, the power amplifier is realized as a transistor circuit because it influences the nonlinear characteristic of the system in a crucial way. In Figure 2, the simulation circuit is shown with all RF components.

The IQ modulator and IQ demodulator are built up as a behavioral model. This method allows the simple specification of some typical features of these components such as amplitude and phase imbalance, as well as the dependence on temperature and supply voltage. These dependencies are mathematically modeled by linear

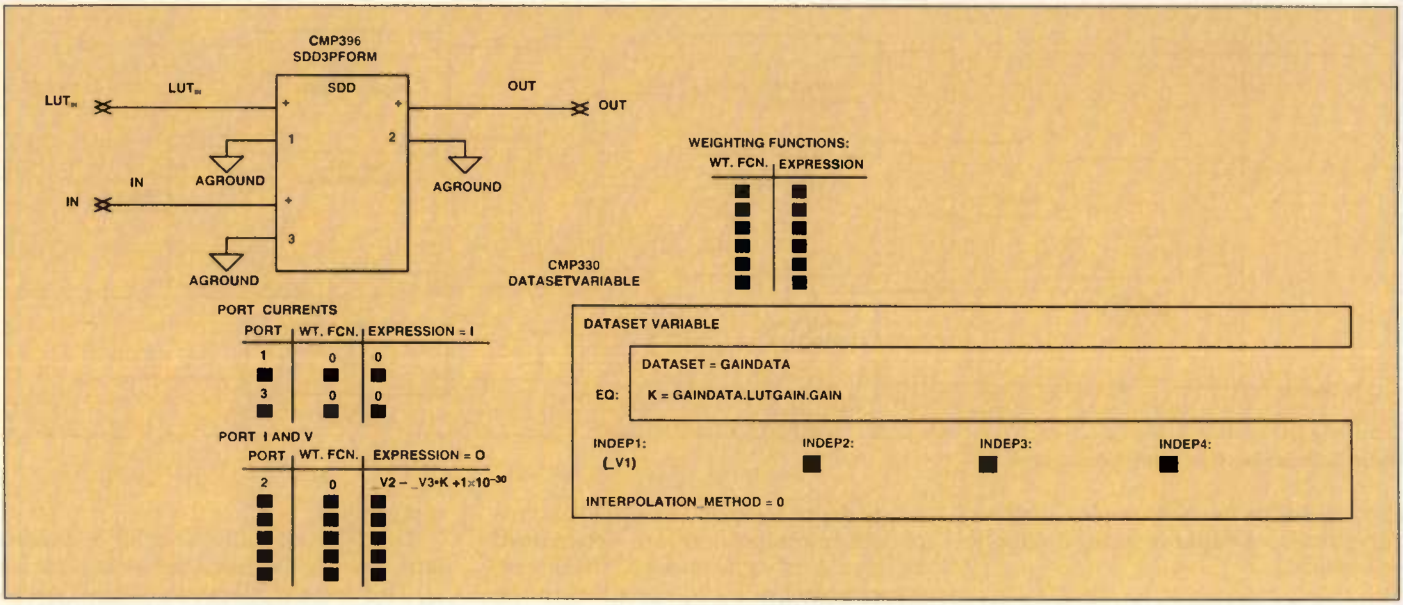

Figure 4. Amplitude predistortion.

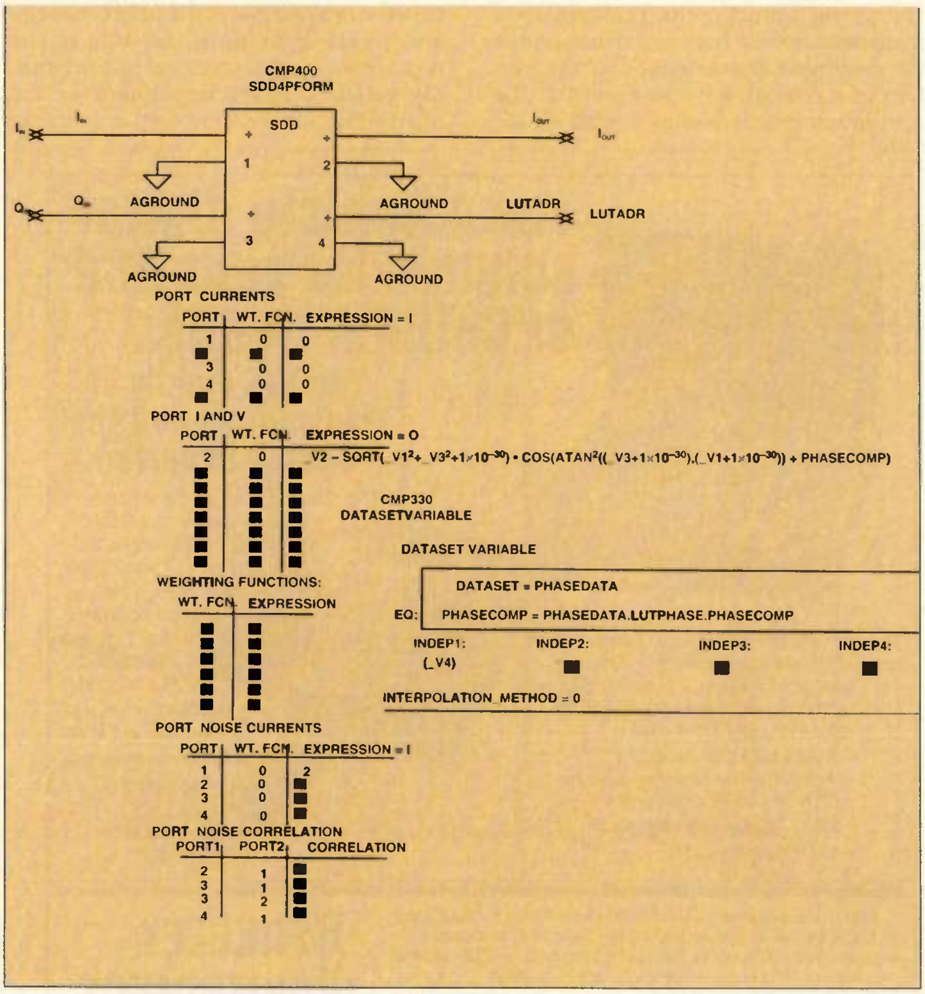

Figure 5. Modification of the I-channel for phase predistortion.

equations. These are a good approximation proved by simulations of a Gilbert cell structure, which is often used in such devices.

Altogether, six amplifier stages are used with a total gain of about 70 dB to achieve an output power of 33 dBm. The first four stages are built up with bipolar transistors, whereas the last two stages use gallium arsenide metal semiconductor field-effect transistors (GaAs MESFETs). The condition of linear operation is always fulfilled for the first three stages, even at high-input levels; therefore, these amplifiers are taken into account by a linear s-parameter block. The last three stages are implemented in detail, especially the power amplifier, which is the most important element for the nonlinear character of the complete transmitter. A special task is the implementation of the table-controlled predistortion inside the RF-simulator framework. Figure 3 shows the structure of the predistortion block.

The I- and Q-component of the signal is modified according to the following two equations:

$$I_{pred} = \sqrt{(kI(t))^2 + (kQ(t))^2} \bullet \cos\left(\arctan\left(\frac{Q(t)}{I(t)}\right) + \Delta\phi\right) \quad (1)$$

$$Q_{pred} = \sqrt{(kI(t))^2 + (kQ(t))^2} \bullet \sin\left(\arctan\left(\frac{Q(t)}{I(t)}\right) + \Delta\phi\right) \quad (2)$$

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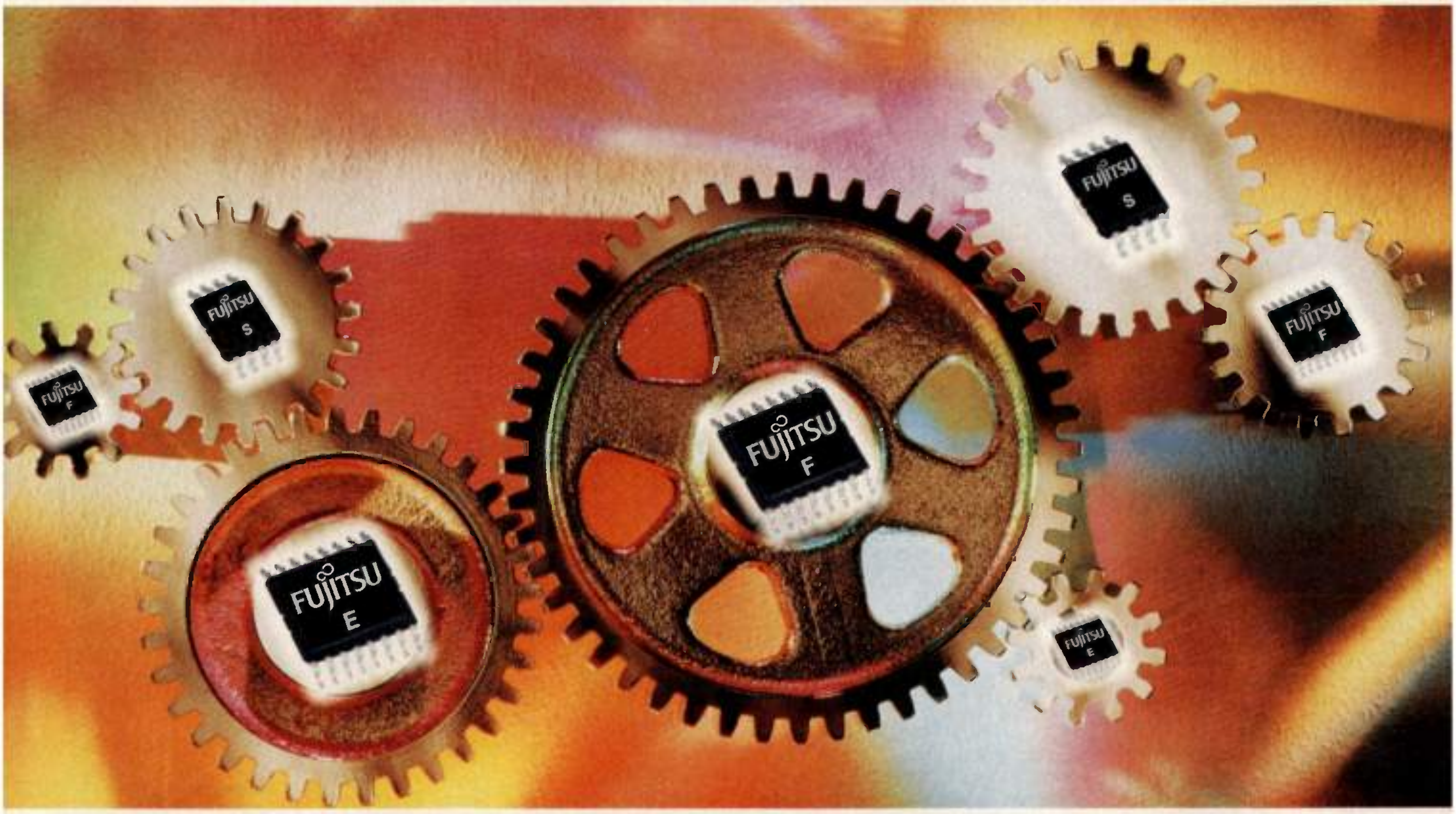

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2	TE5000	3	3.75	20	18.0	-	-	2	1.0	50	1800//+4
4	TE5010	3	3.75	30	14.0	-	-	3	2.0	60	1500//+3
6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	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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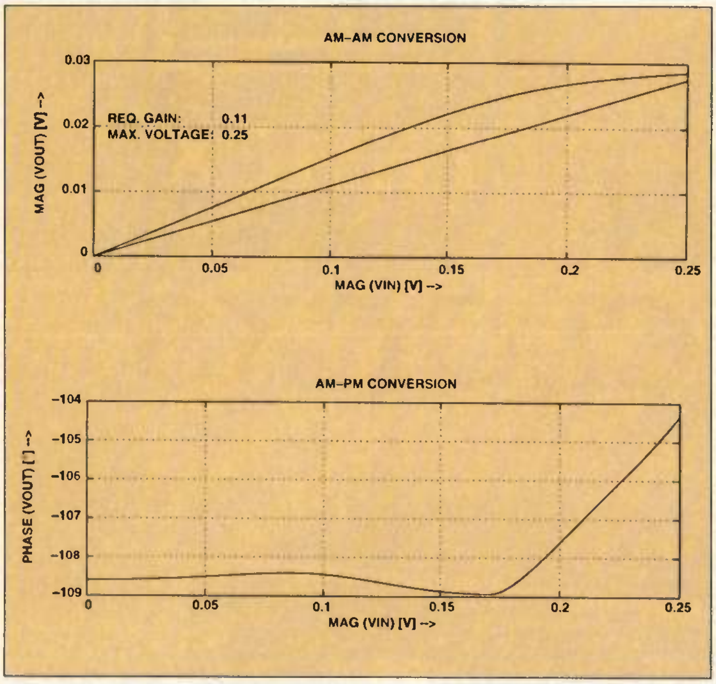


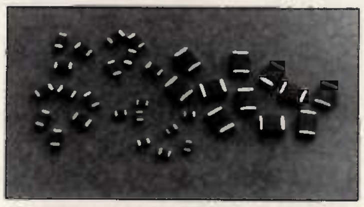
Figure 6. AM–AM and AM–PM conversion characteristic of the system.

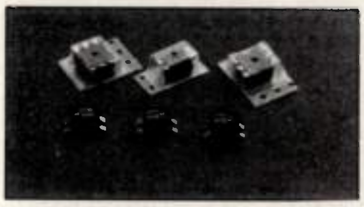
It is assumed that the nonlinear transfer characteristic of the amplifiers can be described by a complex gain function, thus enabling a decoupled amplitude- and phase-predistortion. In this case, the AM–AM and AM–PM conversion depends only on the amplitude of the complex modulation signal. The selection of the correct table value and the multiplication with the appropriate signal level is realized by the help of a symbolically defined device provided by the simulation tool. By means of this device, a mathematical description of port currents and voltages readily can be accomplished as shown in Figure 4.

The phase predistortion is done in a similar manner by using the fact that a complex RF-signal can be separated in an I-and-Q signal. In this case, the predistortion coefficient also is read out of a table. Figure 5 illustrates the necessary circuit for the modification of the I-channel.

Signal processing procedure

For signal processing, the nonlinear system reaction is required when the system is stimulated by a linear amplitude ramp. For each point of this ramp, the AM–AM and AM–PM conversion of the considered signal path is determined. With a root-finding algorithm, the appropriate predistortion coefficients for amplitude and phase are calculated to get a linear amplitude and a constant phase characteristic. This algorithm is programmed with a mathematics software tool. A table size of 512 values is sufficient.





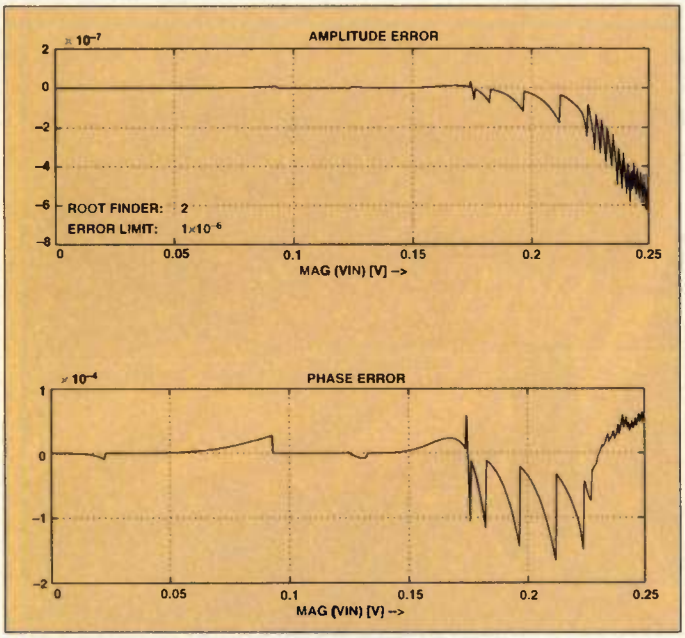


Figure 7. Remaining amplitude and phase errors.

The table size and the error limit of the root finding algorithm (e.g., 1×10^{-6}) depends on the requested performance of the system in regard to the adjacent channel interference. After generation, the tables are transferred in the form of a dataset to the RF simulator. The computation of one set of tables requires about two minutes (SUN-Sparc 10). In Figure 6, a typical example of a nonlinear AM–AM and AM–PM conversion characteristic is shown, together with the linear goal function for the amplitude.

This characteristic was taken from the complete signal path starting at the input of the IQ-modulator and ending at the output of the IQ-demodulator. The remaining error of amplitude and phase (the deviation from an ideal linear amplitude characteristic and a constant phase) can be seen in Figure 7.

Results of simulation

The main reason for simulating the linearization system is the particular interest in the behavior of the linearization when non-ideal parameters of components and global parameters (e.g. temperature) are changed. A wide variety of influences were examined in more detail:

- amplitude- and phase-imbalance of the demodulator
- variation of the load impedance at the output
- variation of the surrounding temperature

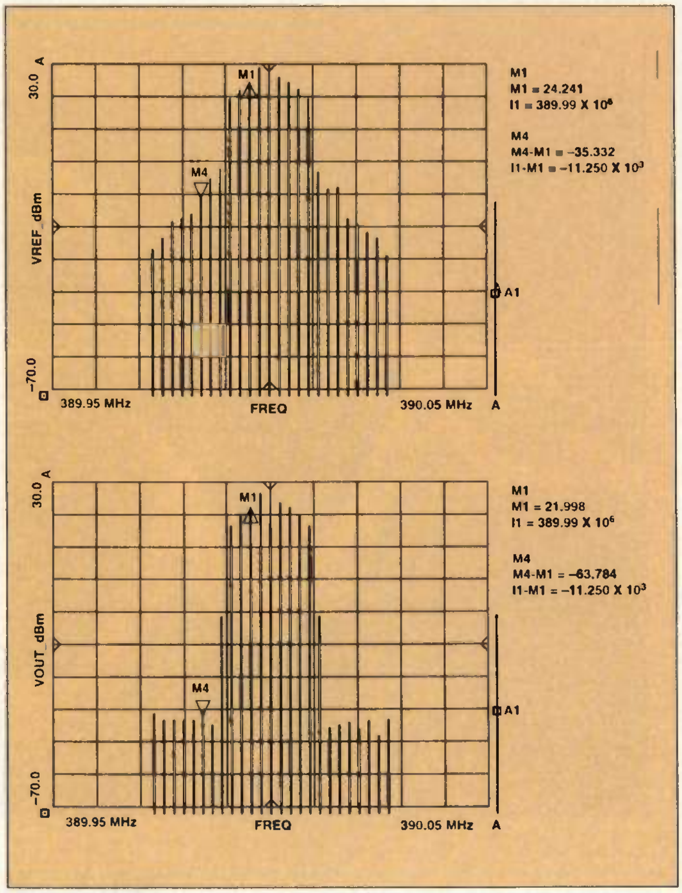


Figure 8. Output spectrum with and without linearization.

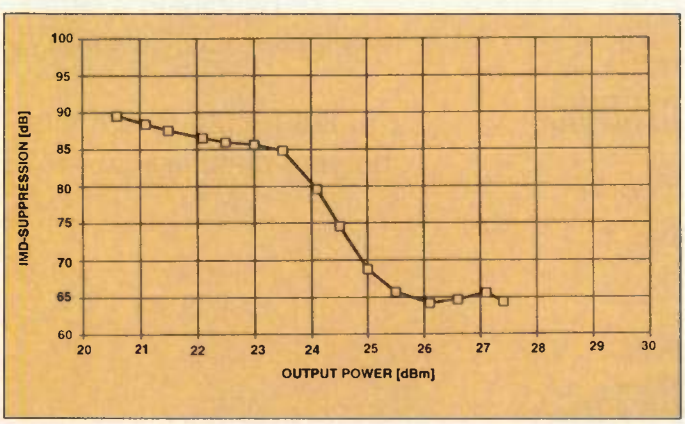


Figure 9. Decrease of intermodulation distortions due to power back-off.

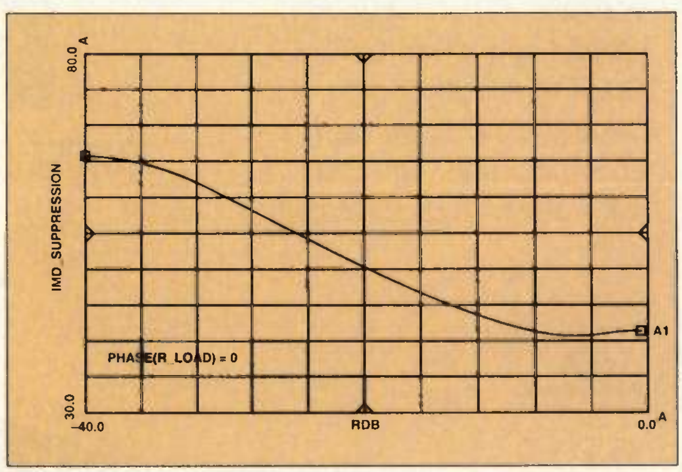


Figure 10. Change of interference suppression vs. output mismatch.

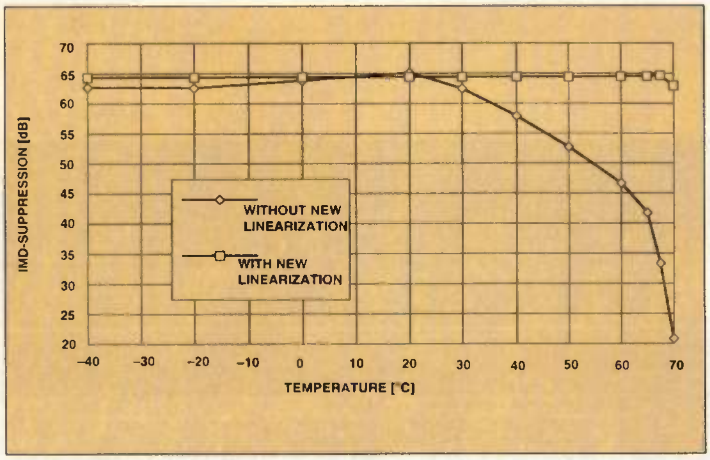


Figure 11. Effect of temperature variation.

- change of frequency
- influence of the 1 dB compression point of modulator and demodulator
- influence of the output power level

In all cases, the resulting performance of the system was compared to an ideal system without any perturbations. This comparison was accomplished by measuring the difference between the wanted signal and the adjacent-channel interferences (third-order intermodulation product) in the output spectrum. For modulation, a random bit sequence was used. Figure 8 shows a typical output spectrum for a linearized and for a non-linearized transmitter. The drive level for the power amplifier was in this case high enough that the voltage peaks of the output signal reached the 1 dB compression point of the amplifier. All system elements were assumed to be ideal.

The resulting suppression of interferences is 64 dB, meaning an improvement of 32 dB in comparison with the non-linearized case could be achieved. Another point of investigation was the change of interference suppression if the output power is reduced. The result is shown in Figure 9. The highest power level applied in this diagram corresponds to a peak voltage that reaches the 1 dB compression point of the power amplifier.

It clearly can be seen that a strong back-off of about 4 dB is needed to achieve an improvement of the interference suppression of about 20 dB. The demodulator plays an important role because its imperfections cannot be canceled out by the signal processing algorithm. The simulations showed that the amplitude imbalance has nearly no effect, and a severe phase imbalance of 4° causes a deterioration of only 1.5 dB. Another point of interest was the system behavior when the load impedance of the power amplifier was changed without a new linearization procedure. The result is illustrated in Figure 10.

This figure shows the interference suppression on the y-axis and the return loss of the load (in dB) on the x-axis. The

phase of the reflection coefficient is 0°. With increasing mismatch the interference suppression deteriorates continuously. The effect of a temperature variation was examined in Figure 11. For lower temperatures, only a slight deterioration can be seen, whereas higher temperatures show that a significant decrease of the interference suppression occurs. The second, nearly constant curve, shows the system behavior in the case of renewing the linearization after each temperature step.

Conclusion

A simulation arrangement is capable of dealing with an adaptive digital predistortion circuit for the linearization of an RF transmitter. This linearization method offers some promising features. Conventional simulation tools fail to handle this easily because of its complex structure—a signal processing unit and an RF section. As a result, a combination of a mathematics program for baseband calculations and an RF-simulator for the high-frequency portion of the circuitry was used. The most critical part of the structure in regard to the nonlinear characteristic is the power amplifier, which now can be modeled at transistor level. Other components (modulator, demodulator) are implemented by means of behavioral models. Nevertheless, a link is necessary between both simulation schemes. This is accomplished with mathematical function blocks in the RF simulator. The input data to this predistortion circuit is provided by look-up tables that were generated by the mathematics program and that were transferred to the RF simulator. After setting up the simulation structure, several examinations were carried out, such as the effect of components impairments, the influence of temperature, drive level and variation of load impedance. RF

Acknowledgments

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Manage a team of highly skilled engineers developing an RF module for the IS-95 based handset, including RF and mixed-signal ASICs and other key components. Requires an MS/PhD in EE with 12+ years of RF area experience. Must have exposure to all aspects of RF module development, including RF ASICs and mixed signal ASICs, as well as components such as PA, VCO, synthesizer and LNA. Job Code: 585

RF ANALOG DESIGN ENGINEERS

Participate in the design of an IS-95 cellular phone. Requires a BS/MS in EE and 4+ years of RF/analog circuits or systems design experience. Strong analytical skills, practical design experience with frequency synthesizers, VCO, PLL and ADC/DAC conversion necessary. Knowledge of communication theory and cellular phone structure is a plus. Job Code: 370

RF ANALOG IC DESIGNER

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 simulation and layout is essential. Understanding of design issues, low phase noise VCO, third intermodulation and power management circuit design a definite plus. Job Code: 371

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DIRECTOR OF SYSTEMS DEVELOPMENT

Head a systems development and integration team, whose initial focus will be the handset and will expand to include base station equipment. Requires an MS/PhD in EE/CS with 12+ years of experience in communication systems. Must be familiar with some cellular standards (AMPS, IS-95, GSM). Should understand system integration and test issues. Job Code: 583

MIXED-SIGNAL IC DESIGNER

Design mixed-signal ICs for cellular phone applications. Entails carrying the whole design from start to finish including design, simulation, layout supervision and testing. Requires a BS/MS in EE with 3+ years of Bipolar/BiCMOS mixed-signal ASIC design experience. Thorough knowledge of SPICE simulation required. Job Code: 309

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Wireless Communications Division

Standardized IS-95 CDMA system simulation aids advanced designs

By Paul Washkewicz

Code-division multiple-access (CDMA) is rapidly challenging other modulation techniques for dominance in personal communications services (PCS), as well as in other mobile communications systems. Accurate system simulation allows engineers to construct more advanced designs at the analog circuit, RF subsystem and digital signal processing (DSP) baseband level.

Designers working in today's strict time-to-market environment can use this standardized IS-95 system as a starting point for accurate system simulation before focusing design efforts on the proprietary circuit design tasks at hand. The following information describes a CDMA system that conforms to IS-95 and that was designed using HP EEsof's OmniSys communication system simulator. The CDMA system simulation includes a programmable Walsh code generator, a forward and reverse link modulator, an RF transceiver, a delay-locked loop and a matched filter. Once the standard system is functional, the modulated spectrum is propagated through a Rayleigh-faded channel and is sent through a Rake receiver, where bit-error-rate performance can be verified. Figure 1 shows the standardized system based on the IS-95 CDMA specification. The baseband section performs DSP convolutional encoding, interleaving, Walsh-code spreading and in-phase and quadrature short-code spreading. The mixer and oscillator perform an analog upconversion. The amplifier symbol is a high-level representation of the high-frequency components, such as the power amplifier, transmit and receive switch and power-control functions. After power amplification, the transmitted spectrum travels through the RF propagation environment, receives, downconverts, acquires the code, tracks the carrier phase and decodes the data.

Despite the difficulties inherent in performing so many tasks, the standardized system is able to accurately model the following: the finite numeric effects of the DSP components; transient and steady-state effects of the analog components; and the voltage

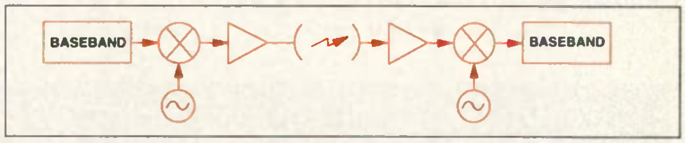


Figure 1. CDMA system block diagram.

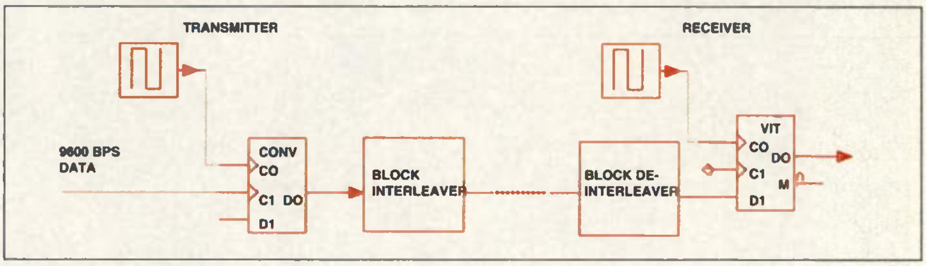


Figure 2. Coding and interleaving.

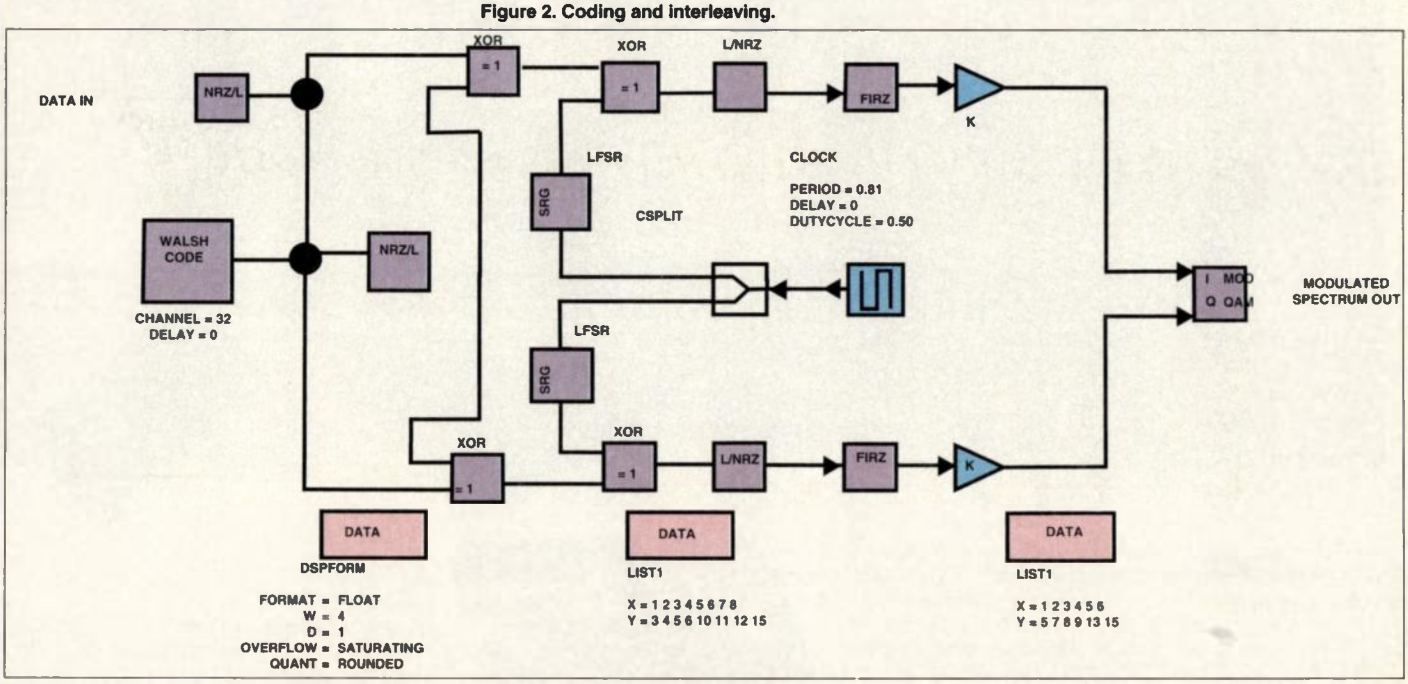


Figure 3. CDMA I-Q modulator block diagram.

dc to 3GHz from $\$11^{45}$

low pass, Plug-in, dc to 155MHz

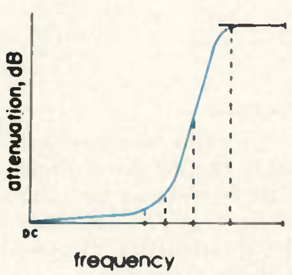
Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
•★LP-1.9	DC-1.9	3.4-4.7	4.7-200
•★LP-2.5	DC-2.5	3.8-5.0	5.0-200
★LP-5	DC-5	8-10	10-200
★LP-10.7	DC-11	19-24	24-200
★LP-21.4	DC-22	32-41	41-200
★LP-30	DC-32	47-61	61-200
★LP-50	DC-48	70-90	90-200
★LP-70	DC-60	90-117	117-300
★LP-90	DC-81	121-157	157-400
★LP-100	DC-98	146-189	189-400
★LP-150	DC-140	210-300	300-600

dc to 1200MHz

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
★LP-200	DC-190	290-390	390-800
★LP-250	DC-225	320-400	400-1200
★LP-300	DC-270	410-550	550-1200
★LP-450	DC-400	580-750	750-1800
★LP-550	DC-520	750-920	920-2000
★LP-600	DC-680	840-1120	1120-2000
★LP-750	DC-700	1000-1300	1300-2000
★LP-800	DC-720	1080-1400	1400-2000
★LP-850	DC-780	1100-1400	1400-2000
★LP-1000	DC-900	1340-1750	1750-2000
★LP-1200	DC-1000	1620-2100	2100-2500

All models priced qty. 1-9 ($ea.), Conn. Type P = 11.45, B = 32.95, S = 34.95, N = 35.95
• Exceptions: ★LP-1.9 P = 13.95, B = 34.95, ★LP-2.5 P = 14.95, B = 35.95
On both models, add following to B price: $3.00 for N, $2.00 for S
75 ohm versions available

LOW PASS



Surface-mount

dc to 108MHz

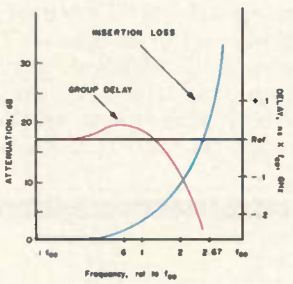
Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
SCLF-5	DC-5.0	8-10	10-200
SCLF-8	DC-8.0	12.5-16.5	16.5-200
SCLF-10.7	DC-11	19-24	24-200
SCLF-21.4	DC-22	32-41	41-200
SCLF-25	DC-25	36-47	47-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-95	DC-95	146-189	189-400

dc to 1200MHz

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-225	DC-225	340-440	440-1200
SCLF-380	DC-380	580-750	750-1800
SCLF-420	DC-420	750-920	920-2000
SCLF-550	DC-550	800-1050	1050-2000
SCLF-700	DC-700	1000-1300	1300-2000
SCLF-1000	DC-1000	1620-2100	2100-2500

Price: SCLF 21.4–SCLF 420 $11.45 ea. SCLF-8, 10.7, 550, 700, 1000 $12.95 ea. SCLF-5 $14.95 Qty. (1-9)

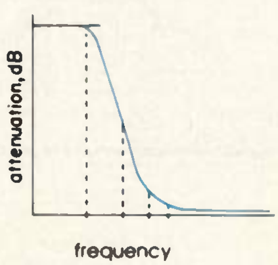
Flat Time Delay, dc to 1870MHz



Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss >10dB	Stopband MHz loss >20dB	VSWR Freq. Range, DC thru 0.2fco X	VSWR Freq. Range, DC thru 0.6fco X	Group Delay Variations, ns Freq. Range, DC thru fco X	Group Delay Variations, ns 2fco X	Group Delay Variations, ns 2.67fco X
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.70	4.0	5.00
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.90
★BLP-156	DC-94	312-416	416	1.3:1	1.1:1	0.30	1.1	1.50
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.40	1.3	1.60
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.20	0.6	0.80
★BLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
▲BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
▲BLP-1870	DC-850	3740-5000	5000	1.45:1	2.9:1	0.05	0.1	0.15

Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE: ▲ -933 and -1870 only with N and SMA connectors.

HIGH PASS



high pass, Plug-in, 13 to 1200MHz

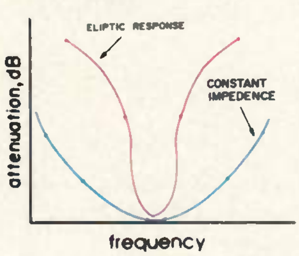
Model No.	Stopband MHz loss > 40dB	Stopband MHz loss > 20dB	Passband, MHz loss < 1dB	VSWR Pass-band Typ.
★HP-25	DC-13	13-19	27.5-200	1.7:1
★HP-50	DC-20	20-26	41-200	1.5:1
★HP-100	DC-40	40-55	90-400	1.5:1
★HP-150	DC-70	70-95	133-600	1.8:1
★HP-175	DC-70	70-105	160-800	1.5:1
★HP-200	DC-90	90-116	185-800	1.6:1
★HP-250	DC-100	100-150	225-1200	1.3:1
★HP-300	DC-145	145-190	†290-1200	1.7:1

210 to 2200MHz

Model No.	Stopband MHz loss >40dB	Stopband MHz loss > 20dB	Passband, MHz loss < 1dB	VSWR Pass-band Typ.
★HP-400	DC-210	210-290	395-1600	1.7:1
★HP-500	DC-280	280-365	500-1600	1.9:1
★HP-600	DC-350	350-440	600-1600	2.0:1
★HP-700	DC-400	400-520	700-1800	1.6:1
★HP-800	DC-445	445-570	780-2000	2.1:1
★HP-900	DC-520	520-660	910-2100	1.8:1
★HP-1000	DC-550	550-720	1000-2200	1.9:1

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95. For ★HP-25, Add $2 ea. †Loss 1.5 dB max.

BANDPASS



bandpass, Elliptic Response, 10.7 to 70MHz

Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max. (MHz)	3 dB Bandwidth Typ. (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
★BP-10.7	10.7	9.5-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
★BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
★BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
★BP-60	60.0	55.0-67.0	49.8-70.5	44 & 79	4.6 & 190-1000
★BP-70	70.0	63.0-77.0	58.0-82.0	51 & 94	6.0 & 193-1000

Price, (1-9 qty), all models: plug-in $18.95, BNC $40.95, SMA $42.95, Type N $43.95

Constant Impedance, 21.4 to 70MHz

Model No.	Center Freq. MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR 1.3:1 Total Band MHz
★IF-21.4	21.4	18-25	1.3 & 150	DC-220
★IF-30	30.0	25-35	1.9 & 210	DC-330
★IF-40	42.0	35-49	2.6 & 300	DC-400
★IF-50	50.0	41-58	3.1 & 350	DC-440
★IF-60	60.0	50-70	3.8 & 400	DC-500
★IF-70	70.0	58-82	4.4 & 490	DC-550

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

NOTE: **★Add Prefix P, B, N, or S for Pin, BNC, N, or SMA connector requirement.**

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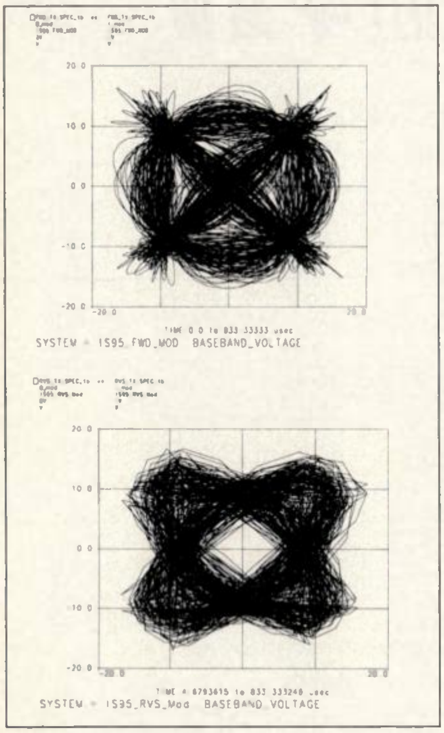

Figure 4. Forward and reverse channel constellations.

standing wave ratio (VSWR) and frequency-domain distortion of the RF components. This difficulty is further exacerbated by the requirement that the system must be simulated in an RF environment with delay, Doppler effects and attenuation at the transmitted frequency.

System modeling description

Various sub-blocks make up the standard system. The first contains baseband functions such as *convolutional encoding* and *block interleaving*. (See Figure 2.) The convolutional encoder is a rate 1/3, with a constraint length of 9. The generator functions are Co = 557, C1 = 663 and C2 = 711. The block interleaver is defined as loading a 24 × 16 matrix by columns and as reading out the data by rows. In this manner, burst errors caused by deep fading can be randomized and corrected by convolutional encoding and by Viterbi decoding.

The next block is the CDMA I/Q modulator, which is shown in Figure 3. Data enter from the left. They first are spread by the Walsh code, which is currently set to channel 32. This channel 32 signal is then spread by the in-phase (I) and quadrature (Q) short pseudorandom (PN) codes, which are set by the linear-feedback shift registers (LFSR). The polynomials for the LFSR, which can be seen in the figure, are:

$$P_1(x) = x^{15} + x^{13} + x^9 + x^7 + x^5 + 1 \qquad (1)$$

$$P_Q(x) = x^{15} + x^{12} + x^{11} + x^{10} + x^6 + x^5 + x^4 + x^3 + 1 \qquad (2)$$

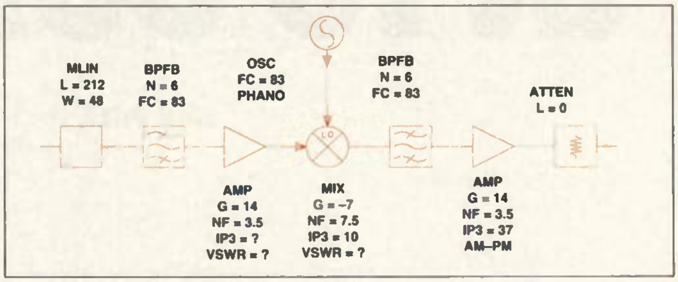

Figure 5. Analog-RF transmitter section.

After this baseband signal is filtered using the 48 finite-impulse response (FIR) coefficients from the IS-95 specification, it is modulated to 83 MHz using the quadrature phase-shift keying (QPSK) modulator.

If this were the reverse channel modulator, a delay equal to one-half the chipping rate would be inserted into the Q channel. Notice that the designer has complete control over the finite numeric effects of each of the digital blocks in the modulator. These effects include fixed point, saturation and quantization of the numeric signals. The digital effects on the signal-to-noise ratio (SNR) are estimated for uni-

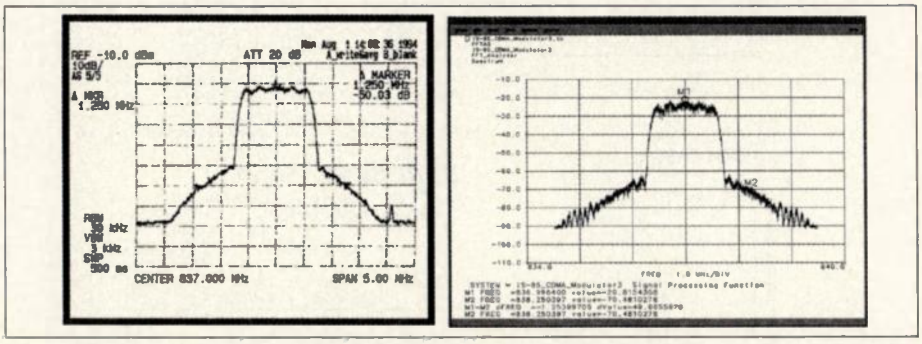

Figure 6. Measured vs. simulated transmitted spectrums.

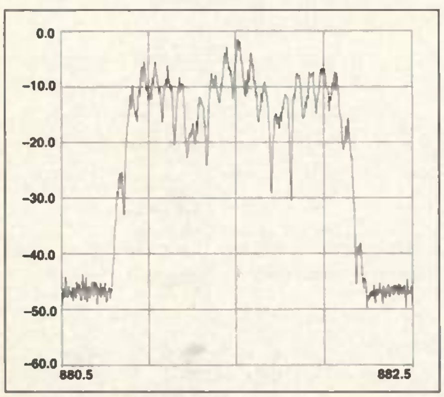

Figure 7. Received faded spectrum.

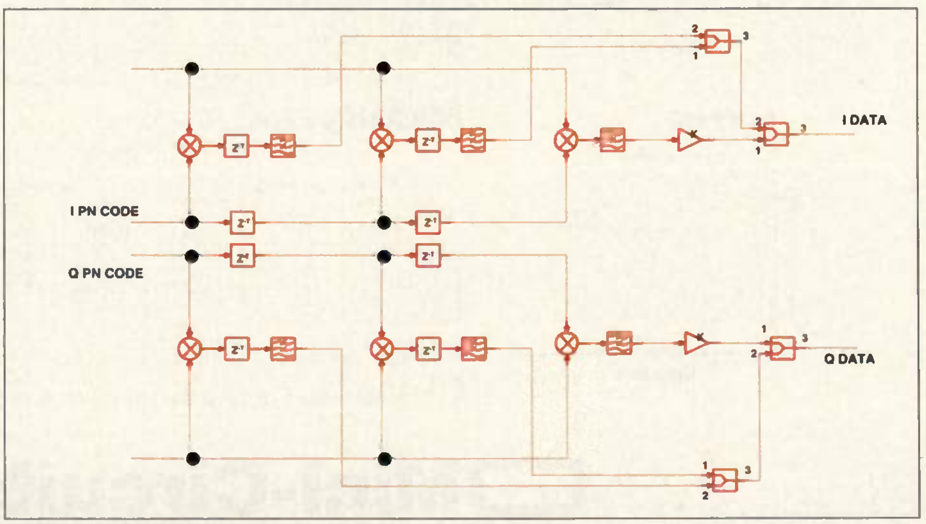

Figure 8. Three-arm Rake receiver.

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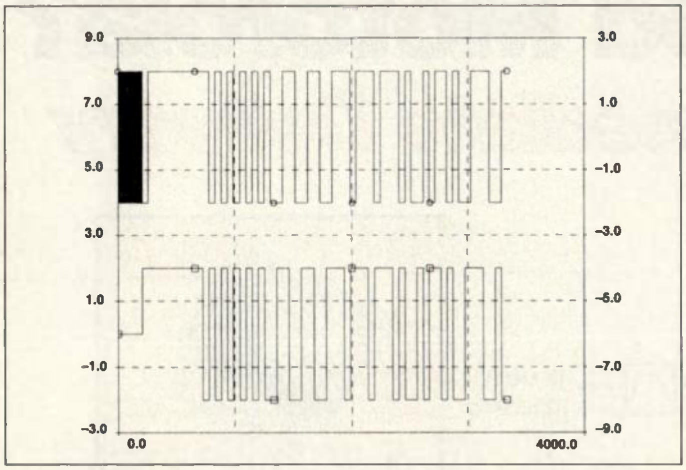

Figure 9. Rake receiver transmitted and recovered data.

form quantizers as follows:

$$SNR = 10\log_{10}\left(\frac{\sigma_x^2}{V^2}\right) + 4.77 + 6.02N \qquad (3)$$

This effect can be included by setting the numeric format from floating point to fixed point. Refer to Figure 4 for the constellation diagram of the QPSK and digital quadrature phase-shift keying (DQPSK) signals for the forward and reverse channels.

The intersymbol interference (ISI) present on these signals arises solely from the FIR filters, but will be further distorted after the analog RF transmitter section.

Using the analog high-frequency system models, an engineer finishes construction on the proposed design in software. This typical RF receiver, shown in Figure 5, includes amplifiers, mixer, oscillator and filter. What makes this block diagram accurate during simulation is the ability to include high-frequency RF effects, such as input and output VSWR, phase-noise and amplifier and mixer nonlinearities. In fact, the actual measured AM–AM and AM–PM effects of the power amplifier can be included. This RF modeling allows for accurate predictions of critical parameters, such as adjacent channel power ratio (ACPR). The parameters of each of the component may be tuned or optimized automatically while the system architecture is changed. This allows designers to find the best design space while monitoring system quality criteria such as bit-error rate (BER).

Figure 6 shows plots of transmitted spectrum predicted by a simulator and by that of actual hardware measured by a CDMA vendor. As can be seen from the plots of each spectrum, the adjacent-channel power is down –35 dB in the measured system, which correlates closely with the –35.5 dB down of the simulated system.

The next stage in the system is to send the spectrum in Figure 6 through propagation channel models. These models, available in the simulator, accept information on location, elevation, vehicle speed and direction. These elements are based on statistical models that simulate multipath effects by incorporating multitap delay networks and by incorporating Rayleigh and Rician probability distribution functions. For this case, ignoring Doppler effects, the power delay profile can be represented by:

$$P(\tau) = \sum_{n=1}^{3} \sigma_n^2 \delta(\tau - \tau_n) \qquad (4)$$

Transmission path loss for a typical urban environment based on Hata's model is:

$$\begin{aligned} L_{TU} = {} & 69.55 + 26.16\log_{10} f_c - \\ & 13.82\log_{10} H_{bs} + \\ & \left(44.9 - 6.55\log_{10} H_{bs}\right)\log_{10} R \end{aligned} \qquad (5)$$

For this simulation, the channel model is set for a three-path Rayleigh fade per the IS-95 specification. Figure 7 shows the faded signal at the input of the receiver.

On reception, the distorted spectrum is downconverted to a suitable intermediate frequency (IF) and then demodulated. The despreading occurs when the local PN sequence is in phase with the

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received PN sequence. The acquisition of the proper PN code phase is achieved by using either matched-filter correlators (parallel acquisition) or by using sliding-block correlators (serial acquisition). After acquisition, a code-tracking circuit is used to maintain the proper code phase. Rake receivers take advantage of the signal present in each multipath component by receiving each component individually. In this system, there are three fingers in the Rake corresponding to the three-path fading environment specified in IS-97. Each finger has a different phase and delay matched to one of the received multipath signals. The cross-correlation of this spread-spectrum signal with a delayed version of itself is low, so that the output will be a function of only the desired multipath component. Equal weighting has been used on each finger, whereas, in practice, a more complicated weighting system based on received signal power would be employed. At the output in Figure 8, the received signals of each finger are added coherently and are ready for data detection.

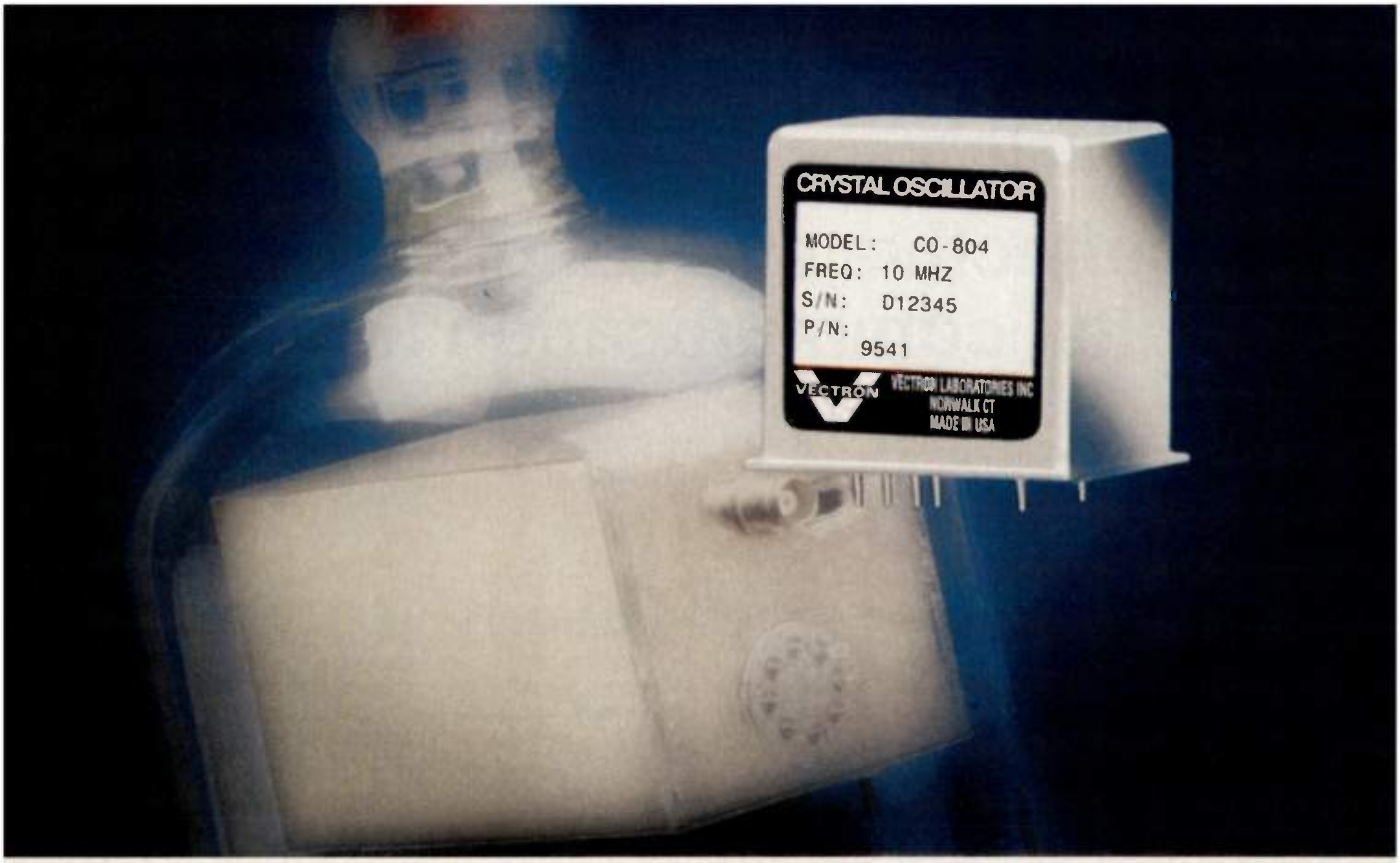
As can be seen in Figure 9, after an initial delay, while the delay-locked loop locks up to track the PN codes, the data is correctly recovered. At this point, the simulation of the standard system is operating correctly, as per specification, and is now ready for the various design groups to use for detailed analog, DSP and RF designs.

Conclusion

A CDMA system was assembled using a CDMA system simulation that includes a programmable Walsh code generator, forward and reverse link modulator, RF transceiver, delay-locked loop and matched filter. The system was made more realistic by the addition of a propagation model and models for mobile and base-station antennas. After RF reception, the system used a delay-locked loop and a three-tap Rake receiver to demodulate the transmitted data. The benefit of using simulation software with a standardized system is to allow engineers to focus on their proprietary analog circuit, RF or DSP design tasks and to verify these designs against the standardized IS-95 system before prototyping begins. RF

About the author

Paul Washkewicz received a B.S.E.E. from Cal Poly University in 1984. For the past five years, he has been working at Hewlett Packard's EEsof Division where he is the product manager for the OmniSys communication system simulator. Previously, he worked at General Dynamics for eight years in the RF design section where his primary duties were the design of analog and digital receiver systems and circuits.



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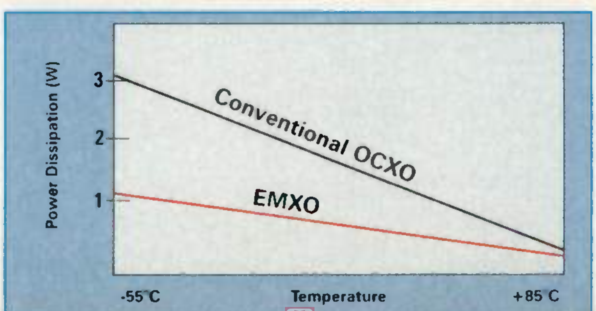

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INFO/CARD 14

Cost-effective evaluation of digital communication ICs

By Marc Brendan Judson

Use simple digital communication circuitry to facilitate evaluation of RF receiver architectures under various conditions. This can be a tool for isolating and avaluating factors that affect digital communications receiver designs. The result can speed your progress in bringing wireless data transmission products to market.

The Federal Communications Commission (FCC) allows unlicensed operation in three designated frequency bands for industrial, scientific and medical (ISM) applications with power levels of less than one watt, if the long-term average power is minimized through spread-spectrum technology. This regulatory provision sets the stage for the development of a vast array of low-power, short-distance wireless digital communications products for untethered computer and printer links (wireless local area networks (WLANs), cordless telephony, wireless inventory control systems and other applications. The constraints on physical size and power consumption require that the RF receiver of these products be highly integrated, while still exceeding the sensitivity and bit-error rate (BER) performance specifications for the application. Taking into account link losses, path losses and especially the limitations placed on transmission power and antenna gain by governing authorities, the transmission range determines the sensitivity requirements for the application. Along with the maximum acceptable number of occurrences causing the re-transmission of data, this will determine a maximum BER at a minimum reference power level.

To be successful in bringing a low-cost, highly integrated, high-performance digital receiver to market in the shortest time possible, an RF designer must be able to work quickly through a series of impedance-matching and filtering problems associated with the frequency down-converting front-end (FE) and intermediate frequency (IF)-processing circuit blocks of the receiver. The designer also must be able to evaluate the BER performance of these blocks in conjunction with variations in the modulated RF carrier signal and be able to evaluate the impact on BER by various first local oscillator (LO) and symbol timing recovery implementations. Often, because of the fast pace at which the digital communications market is growing, all of this design work must be done prior to the formal industry-wide acceptance of the communications protocol to be used. The following information describes a reference design for digital gaussian-filtered shift keying (GFSK) modulation receivers operating in the lower 902–928 MHz industrial, scientific and medical (ISM) band. It provides quick, cost-effective evaluation of front-end circuitry performance in this band and demonstrates the high data-rate capabilities of the IF-processing integrated circuit (IC) that can be applied to other frequency bands. The evaluation of these circuit blocks, when used in a digital communications receiver application, is demonstrated by presenting performance data in relation to current receiver BER specifications.

Typical receiver architecture

An example of a typical receiver architecture is a superheterodyne receiver intended to receive GFSK-modulated signals such as those specified in the digital European cordless telephone (DECT) and preliminary 802.11 committee standards.

Front-end circuitry includes (see Figure 1):

- a low-noise amplifier (LNA) that reduces the overall noise contribution of the receiver circuitry, while increasing the power of the received signal.
- an image reject surface acoustic wave (SAW) bandpass filter that rejects RF input frequencies that are twice the first IF away from the wanted RF signal and that would otherwise also mix down to the first IF.
- a mixer for performing the fre-

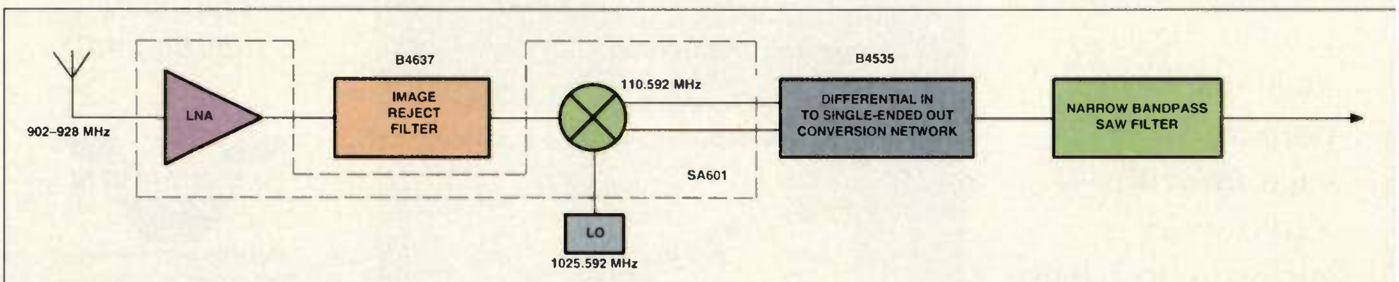

Figure 1. Receiver fromt-end circuitry.

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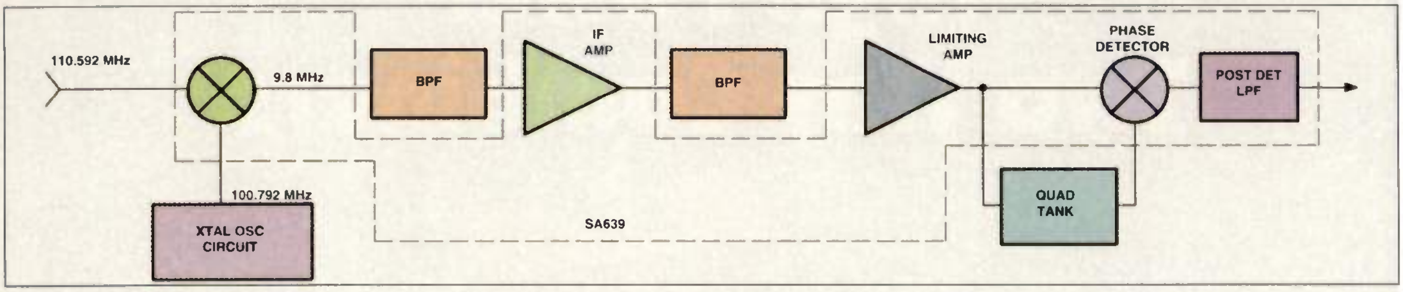


Figure 2. Receiver FM-IF circuitry.

quency down-conversion that translates the original carrier frequency to a level where parasitic impedances are less of a concern.

• a spurious-reject SAW bandpass filter that rejects the intermodulation frequency components falling outside the narrow bandwidth of this filter.

Impedance-matching issues for all single-ended, high-frequency path interfaces can be solved with simple inductor-capacitor (LC) matching networks. Often this can be done with only one shunt capacitor by implementing the series inductance with printed inductors on the printed circuit board. This is useful in board development stages because the windings of the printed inductor can be shorted to obtain a continuous range of impedance values, which simplifies the task of optimizing the noise figure and the gain of the LNA and the mixer.

The first IF mixer outputs are differential open-collector mixer outputs. A balun circuit must be implemented to bring the current of one collector in phase with the other at the chosen IF frequency and to present the complex conjugate impedance to match the spurious-reject filter input. The output of the spurious-reject filter also must be simultaneously matched to the second IF mixer input of the IF-processing circuitry. This is done most easily by implementing matching networks with variable series capacitance on both sides of the filter. These then can be iteratively tuned and later can be replaced by fixed values.

The FM-IF processing circuitry includes (see Figure 2):

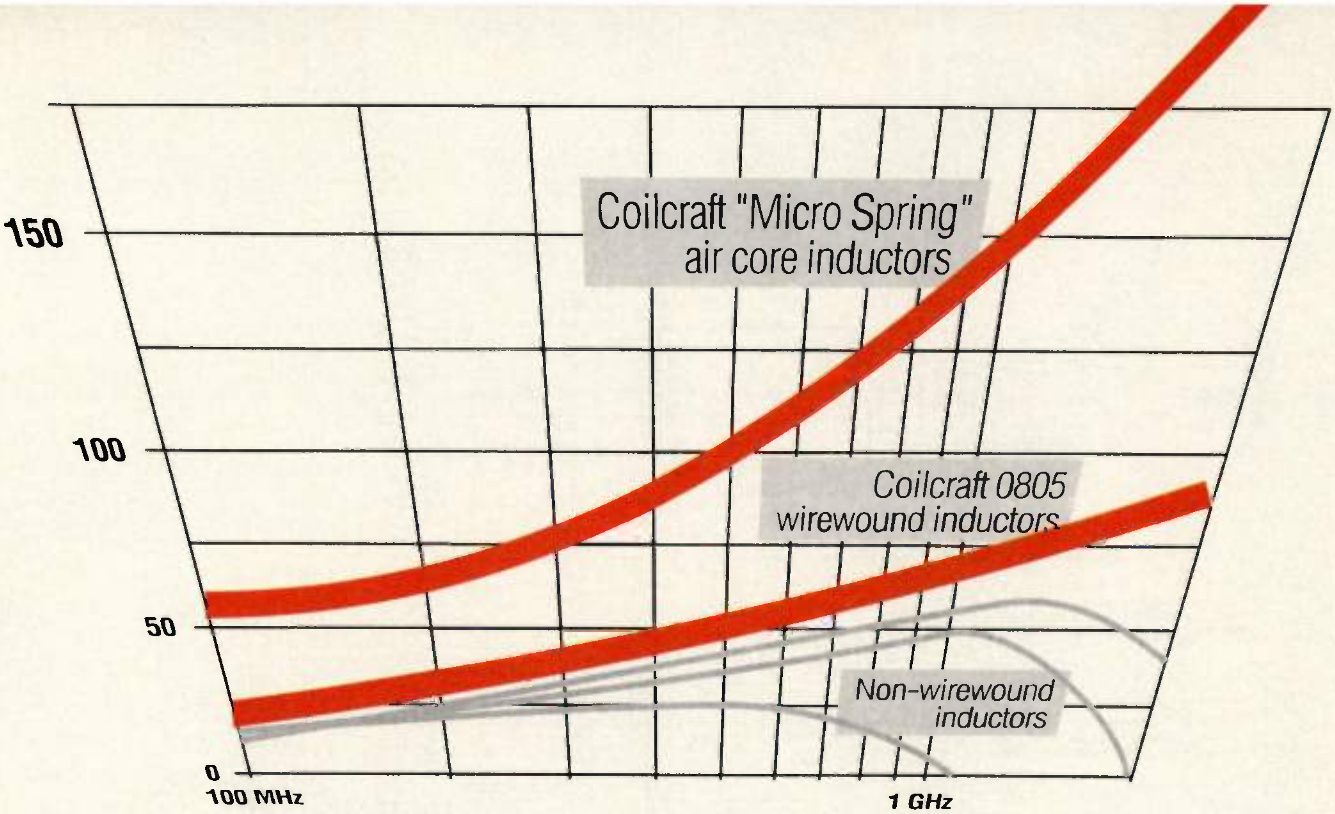
• a Gilbert cell mixer that down-converts the signal to the second IF frequency.

• a crystal oscillator circuit that generates the second LO signal.

• a chain of amplifiers that recovers weak signals and provides a hard-limited amplitude invariant signal to the phase detector.

• interstage filters that reject excess noise outside the required bandwidth.

• a phase detector that demodulates



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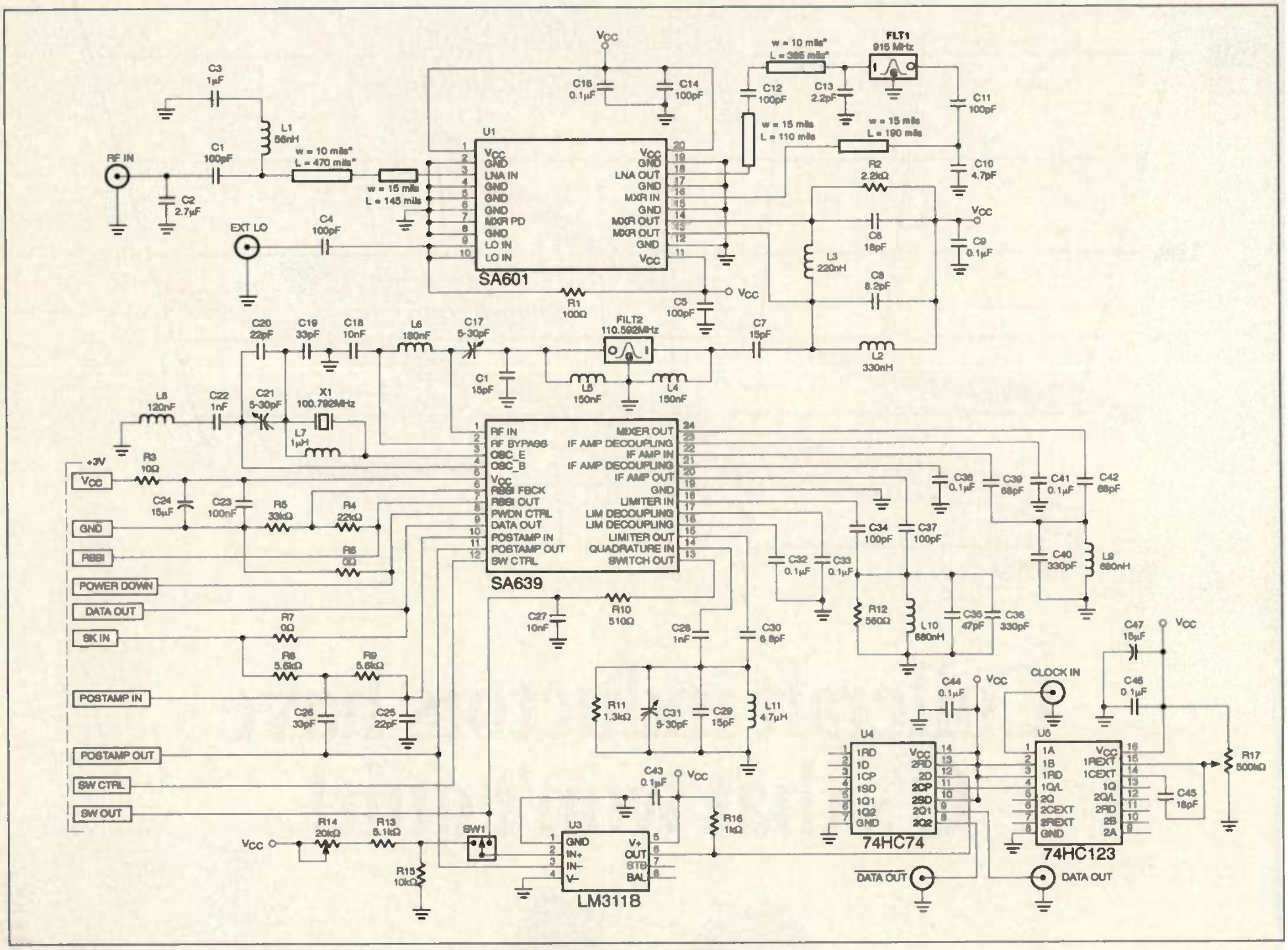

Figure 3. Digital receiver for GFSK-modulated signals in the 902–928 MHz ISM band.

the signal.

- a low-pass post-detection filter that recovers the transmitted information.

The interstage filters can be implemented with simple parallel resonant circuits between the signal path and ground. At resonance, this circuit presents a high impedance-to-ground and passes the wanted signal to the next stage, while all signals and noise outside the bandwidth of this filter are shunted to ground.

The quadrature (Q) tank circuitry can be implemented with simple discrete components, but care must be taken in setting the optimum selectivity of this circuit. The tuning and selectivity of the quadrature tank circuit has a significant effect on the demodulated analog output signal. If the Q of this circuit is set too high, the linear range of the phase detector is compromised, and it will distort the demodulated output signal if the FM deviation exceeds this linear range. If the Q is set too low, it will decrease the level of the demodulated output.

Evaluating BER performance

RF receiver sensitivity most commonly is specified by the input power level at which the signal-to-noise and distortion (SINAD) ratio is 12 dB. However, achieving 12 dB SINAD at a specified input power does not guarantee a good BER at this input power. The SINAD test is typically performed at a single, 1 kHz tone and at a relatively low FM deviation; therefore, it is not sufficient to guarantee proper reception of digital signals that have a much more complex harmonic content and that are modulated with much higher FM deviation. For this reason, BER testing of a receiver is necessary to ensure that the receiver performance is satisfactory in a digital communication application. This testing is done by transmitting data to the receiver and by comparing the data recovered at the receiver output with the data originally transmitted. It is the challenge of the designer to isolate and to evaluate the various sources contributing to the deterioration of the BER. A few simple solutions can help with this task.

Testing without a protocol

In the communication protocol preamble, a specified number of alternating ones and zeros is sent to the receiver. These ones and zeros present an average DC level at the demodulated output of the receiver. This DC level is sampled during the preamble and is used to set a comparator threshold voltage that regenerates the data. If this level is set incorrectly, the BER is severely impaired. The receiver's BER performance can be evaluated, even in the absence of the protocol, by disabling the sampling circuitry of the receiver and by imple-

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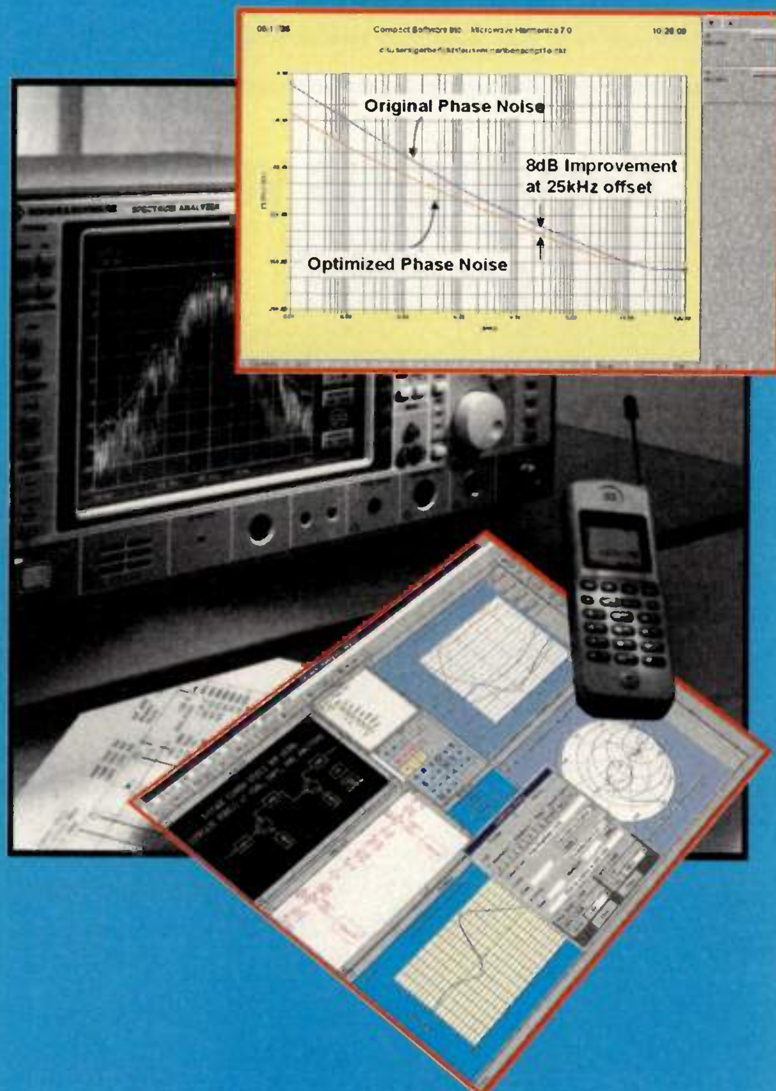

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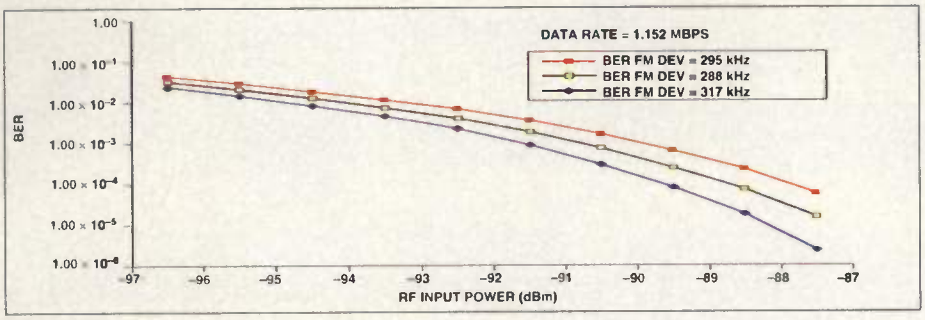

Figure 4. Bit-error rate vs. RF input power, DECT specifications.

menting a manual comparator threshold voltage adjustment, as shown in Figure 3. This allows a designer to evaluate the RF receiver circuitry without having to wait for published communications protocol specifications or having to purchase expensive testing equipment, steps that often slow product development.

Timing recovery and first LO baseline

Receiver synchronization is important too. A designer must choose among symbol timing recovery circuits ranging in performance and complexity and must evaluate the effect on the BER for each option. A manual adjustment of receiver synchronization allows the evaluation of the RF receiver circuitry under optimum conditions. It allows BER deterioration, as a function of synchronization error, to be characterized for a given set of modulation conditions. It provides a valuable benchmark by which the various symbol timing recovery solutions can be evaluated. This manual timing synchronization can be implemented with a simple D flip-flop and a monostable multivibrator as shown in Figure 3.

Another factor that potentially can affect the receiver's BER performance is the phase noise of the LO of the first down converting mixer (first LO). This LO signal can be brought in externally to the RF receiver circuitry. This allows a high-quality, low-phase noise signal generator to be used as a benchmark by the various voltage-controlled oscillator (VCO) and frequency synthesizer implementations of this first LO signal.

Benchmarking to BER specs

The characteristics of the modulation scheme implemented for a particular application such as the type of modulation, the modulation filter bandwidth and the FM deviation all have a significant effect on the BER performance of the receiver architecture being evaluated. There are no accepted high-data-rate protocol and performance standards for receivers in the lower 902–928 MHz band; therefore, as examples of the evaluation of a receiver under a given set of modulation conditions, the performance of the receiver architecture as previously described is compared to current high frequency industry standard specifications.

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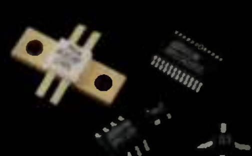

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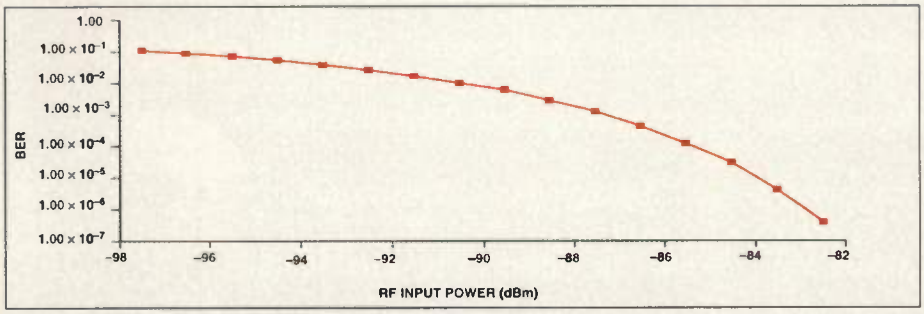


Figure 5. Bit-error rate vs. RF input power, 802.11 specifications.

ating in the 1,880–1,990 MHz band use GFSK modulation and an FM deviation of 288 kHz, and they provide a data rate of 1.152 Mbps. This standard requires that the BER be less than 1×10^{-3} at a reference power level of –86 dBm. Figure 4 shows BER performance as a function of RF input power and FM deviation. This graph indicates that under these modulation conditions, the receiver can hold a BER of 1×10^{-3} for RF input power levels down to –90 dBm, which exceeds the specification by a 4 dB margin. This graph also shows that the degradation in BER with a 10% change in FM deviation is approximately 1 dB.

The preliminary 802.11 standard specifications for WLAN operate in the 2.4–2.5 GHz band, use Btb = 0.5 GFSK modulation, require an FM deviation greater than 110 kHz (160 kHz) and provide a data rate of 1.0 Mbps. This standard requires a framing-error rate (FER) of 3% at a reference power level of –80 dBm, which is equivalent to a BER of approximately 1×10^{-5}. Figure 5 shows BER performance as a function of RF input power. This graph indicates that under these modulation conditions, the receiver can hold a BER of 1×10^{-5} for RF input power levels down to –84 dBm, which again exceeds the specification by a 4 dB margin.

Conclusion

The circuitry as described can be used to evaluate the performance of integrated circuits and FM receiver architectures in a variety of different conditions. It is a useful tool for isolating and evaluating the relative effect of the sources that contribute to a deterioration in the performance of digital communications receiver implementations. As a result, this circuitry can reduce the time needed to bring new wireless data transmission products to market. RF

About the author

Marc Brendan Judson is an applications engineer with the communications product group at Philips Semiconductors in Sunnyvale, CA.

Test set measures phase noise within 5 Hz of the RF carrier

By Charles Luke

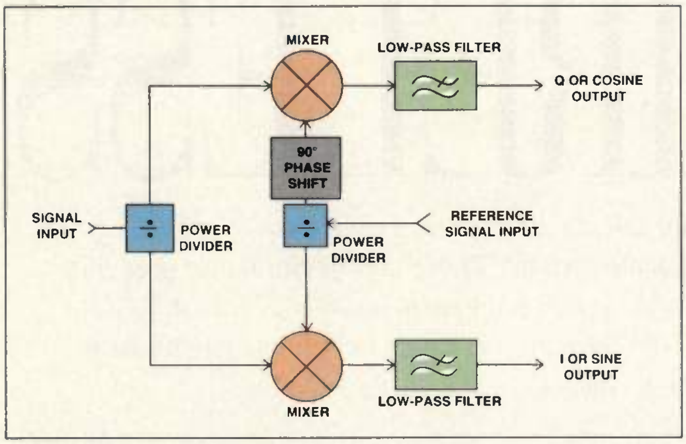


Figure 1. IQ detector schematic.

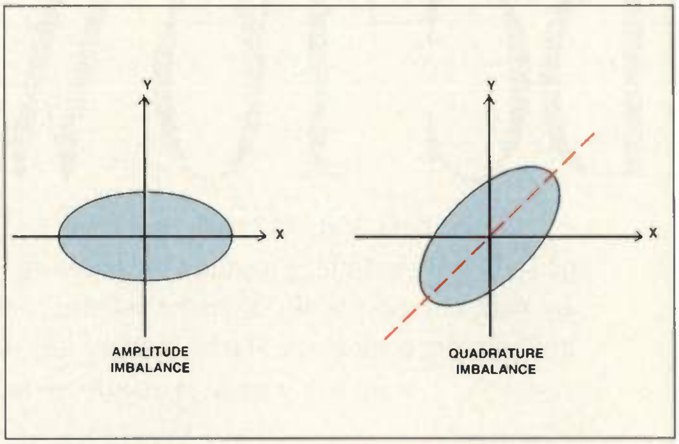


Figure 2. IQ elliptic response

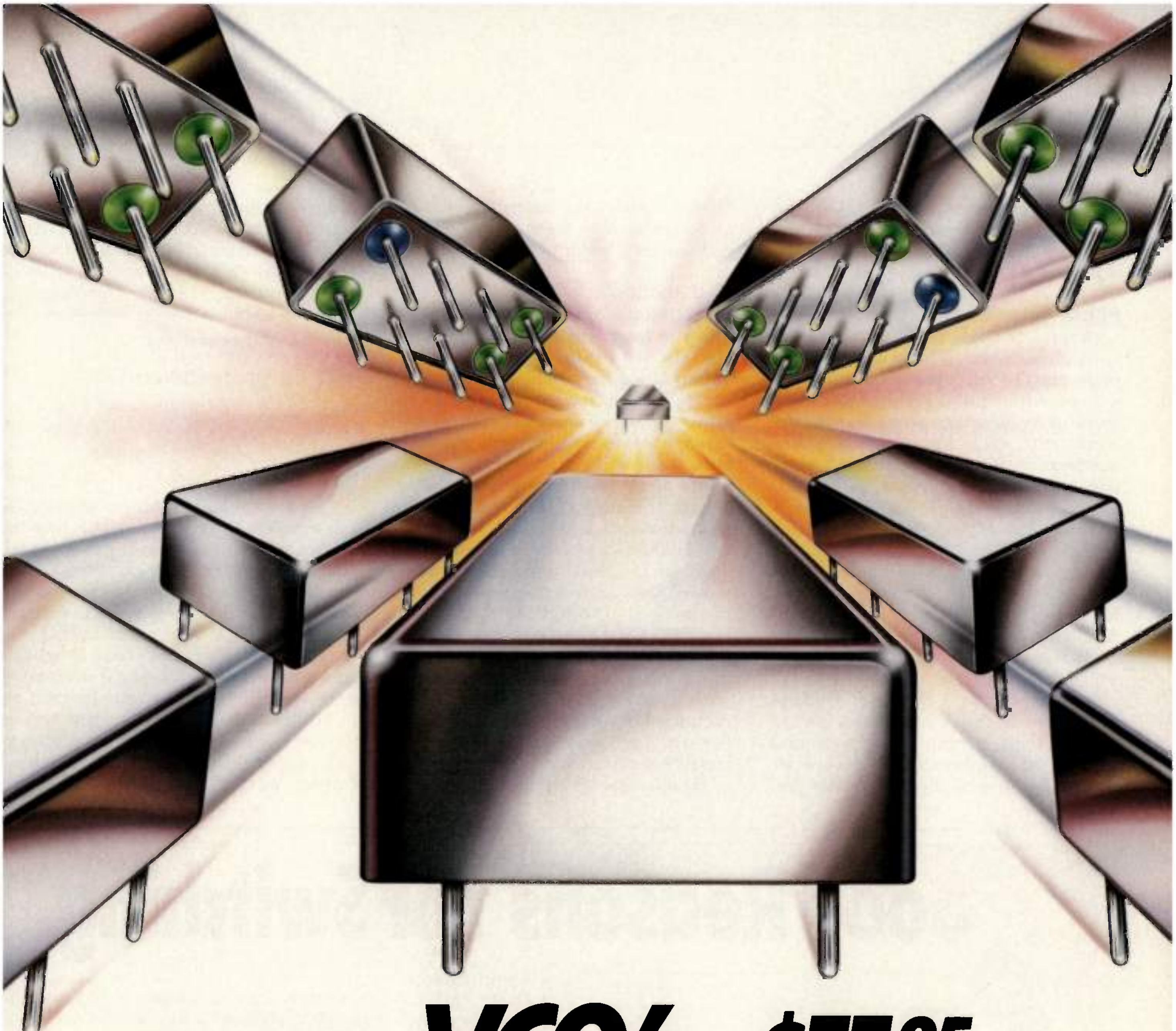
Measuring close-in phase noise with an accuracy better than 0.5° maximum quadrature phase error and 0.05 dB maximum amplitude imbalance over a 175 Hz lowpass bandwidth tests a ground station's contribution to phase noise. A suitable instrument also evaluates individual subsystems as the individual elements of the station are assembled. Measuring the phase noise helps to keep it to a minimum, which in turn makes it possible to transfer the necessary frequency-phase stability reference signal to orbiting satellites in a radio telescope system.

The phase-noise instrumentation system described in the following information was part of a satellite-tracking ground station designed by Scientific Atlanta for the Orbiting Very Long Baseline Interferometer (OVLBI) system, under contract to the National Aeronautics and Space Administration Jet Propulsion Laboratory (NASA-JPL). The OVLBI system is a radio interferometry or radio telescope system used for high-resolution imaging of celestial bodies. An interferometer uses multiple coherent antenna-receivers and their received phases to determine the angle of a signal's arrival. As the receivers are separated farther in distance, the angular resolution of the instrument increases. Limitations imposed by a finite earth diameter restrict the angular resolution of ground-based systems. One solution to this problem is to use orbiting satellites in conjunction with earth-based receivers for extending the baseline distance between receivers, thereby increasing the angular resolution. In a normal terrestrial system, phase and frequency stability between receivers is maintained by using hydrogen maser frequency standards as the basis for all frequency and timing references within the receiving system. In the OVLBI system, this frequency-phase stability reference signal must be transferred from the ground up to the satellite. For this scheme to work, the system that transfers this signal to the satellite must minimize its *phase contamination contribution* or *phase noise*. The measurement of the degree of contamination of phase stability is a necessary qualification for the OVLBI system. The purpose of this phase-noise instrument is to provide the ability to test the ground station's contribution to phase noise. It also is used to evaluate the individual subsystems as the individual elements of the station are assembled.

The phase-noise test instrument is designed to measure the close-in phase noise at an offset of less than 5 Hz from the carrier. Phase testing is performed at 375 MHz. In the ground station, a 375 MHz source first is phase-locked to the hydrogen maser 10 MHz reference. This 375 MHz source is used as the first intermediate frequency (IF) in an upconverter. This upconverted carrier from the ground station is transmitted to the satellite. Then the satellite transponder transmits a translated frequency carrier back to the ground-station receiver. This signal is then downconverted back to the original 375 MHz frequency. Mixing of the original 375 MHz IF signal with the received 375 MHz IF signal furnishes the baseband phase information.

The instrument also functions in a wide bandwidth mode. In this mode, it is used to measure residual doppler in this reference signal. The ground station has numerically controlled oscillators (NCOs) that are used to cancel doppler shift caused by motion of the satellite and by motion of the earth as it rotates. The final full implementation of the instrument also includes this ability to measure residual doppler. Although the doppler compensation is not a part of this article, it was of consequence in the initial design of the instrument.

Specifications for this system are somewhat different than a typical communication system. Most communications systems have data rates between 1 kHz <AL0> and 500 MHz. For this reason, close-in phase noise usually is not a concern, at least not below a 1 kHz offset from the carrier. Also, be-



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cause a hydrogen maser is used for the station-frequency reference, this frequency reference has low phase-noise sidebands.

Phase noise test systems

Several methods can be used to measure phase noise. These basically involve one of four types of instruments: a spectrum analyzer; a phase-lock quadrature detector; a quadrature detector with delay line; or an in-phase quadrature (IQ) detector. Each has its advantages and disadvantages. The differences between methods are listed in Table 1. The type detector chosen for this design is the IQ detector. Initial design called for an instrument that could be used for laboratory phase-noise testing as well as for the final doppler residual measurements.

The spectrum analyzer approach was not chosen because of the small offset (less than 5 Hz) of the phase noise and because of the low phase noise of the hydrogen maser reference. The quadrature detector types were not chosen because of the close offset frequencies where the measurements had to be made. Also these types were not suitable for doppler measurements.

IQ detection

• *Dynamic range limitations* — If a mixer is used as a detector, and if two signals are input to the mixer, the output of the mixer will be the sum and the difference frequencies of those two signals. If the outputs are passed through a lowpass filter, only the difference term will remain. If the signals are of equal frequency but of different phase, the output will be a DC level proportional to the cosine of the phase angle between them. If one introduces a 90° phase shift in one of the paths, the DC level then becomes proportional to the sine of the phase angle difference. Figure 1 shows a practical realization of this detector. By using one mixer with a sine output and one mixer with a cosine output, one can obtain rectangular coordinate (X, Y) outputs for a polar (amplitude and phase) input. A common way of describing this is *IQ outputs*. The phase relationship between the signals can be determined from arctan (Y/X), and the amplitude can be determined by taking the square root of the sum of the squares of the Y and X outputs.

$$I = Y = Amp \bullet \sin(phase) \quad (1)$$

$$Q = X = Amp \bullet \cos(phase) \quad (2)$$

$$Phase = \arctan\left(\frac{y}{x}\right) \quad (3)$$

$$Amp = \sqrt{y^2 + x^2} \quad (4)$$

One of the disadvantages of using IQ detectors is their more limited dynamic range compared to phase-lock or quadrature detectors. A quadrature detector is a mixer operated in the sine output mode with the signal locked or phase shifted to be at the zero DC point. For small variations in the phase about this point, the sine of the angle is equal to the angle. This will be true for angular variations much less than one radian. What is measured is the deviation of the phase about this point. With no phase perturbations, the phase de

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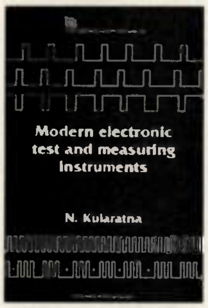

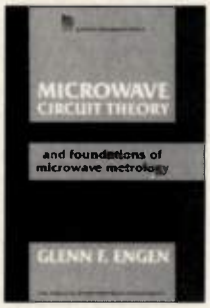

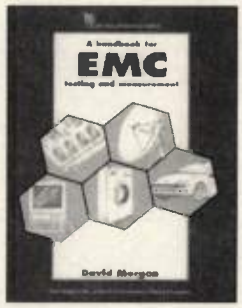

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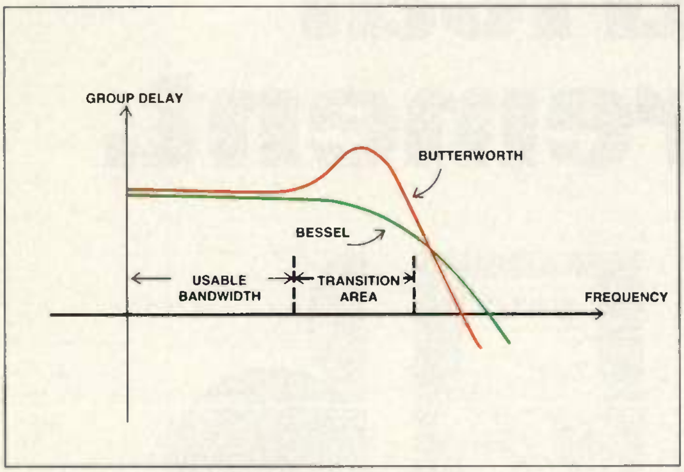

Figure 3. Group delay variation.

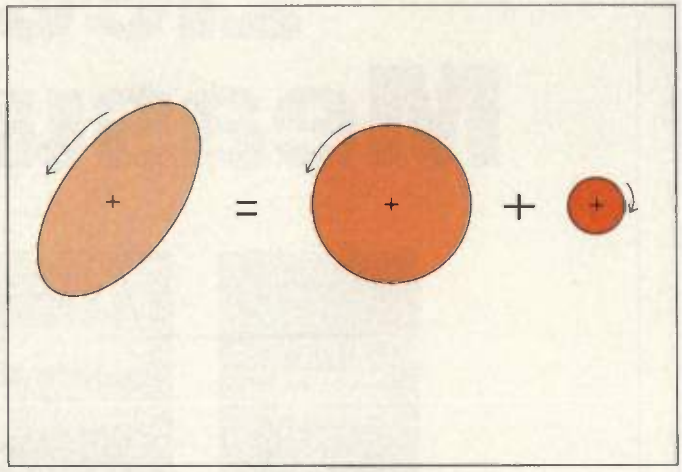

Figure 4. IQ ellipticity as sum of two circular terms.

tector's output voltage is zero volts, so the measurement is that of variations away from this point—the phase perturbations. The post-detection amplifiers have to handle only the dynamic range of the phase variations. Because of this, amplifiers with a lot of gain can be used after the detector, and the variations can be maximized into the analog-to-digital (A/D) converter. For IQ detectors, this is not the case. Since the detectors have no phase lock, the signal can be at any phase in the detector. The voltages out of the detectors are proportional to the sine and cosine of the phase of the signal. If the phase is at 0°, the cosine detector will be at maximum amplitude with its amplitude proportional to the signal's amplitude. This relationship is the limitation of this type of detector scheme. The detector has to handle the full signal level amplitude in the amplifiers after the mixer. The measurement becomes a case of being able to measure small fluctuations, while still accommodating a large signal amplitude.

• *Circularity and balance* — If one of the two inputs to the IQ detector is shifted slightly in frequency, and if one

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Provide RF/wireless system chipset applications support, develop evaluation boards, conduct system performance evaluations based on standard specifications and write system application notes and technical articles. Requires experience with system design and performance evaluation, and familiarity with RF/wireless system measurement equipment, including digital communication instruments. Knowledge of current analog/digital cellular standards is needed. Dept. CPG-SN/AW

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looks at the IQ outputs on an oscilloscope in XY mode, a circle will be displayed. One is really looking at the beat note between the two signal frequencies. These two IQ outputs are at the same beat-note frequency and have the same amplitude, but they are in quadrature to each other because of the sine and cosine outputs. For the detector to measure true phase, the sin and cosine detectors have to be 90° with respect to one another, and they have to be equal in amplitude. If they are not, there will be errors in the measured phase. In the case of errors, the formulas for I and Q become:

$$I = (Amp + Error) \bullet \sin(phase + Phi) \quad (5)$$

$$Q = Amp \bullet \cos(phase) \quad (6)$$

where "Phi" and "Error" represent the error terms. "Phi" is the phase error from 90° or quadrature, while "Error" is the departure from the true amplitude, Amp.

Referring to Figure 1, if the 90° term is not exactly 90°, the difference will cause phase imbalance or a "Phi" term. In the same schematic, if the input power-divider has imbalance in amplitude so that equal signals are not input to the two mixers, there will be an "Error" term or amplitude imbalance. The circular response, caused by a beat note and displayed on the scope in XY mode, becomes an elliptic response. This ellipse will have its major axis on the X or Y axis if the imbalance is in amplitude, and it will have its major axis at 45° or 135° if the imbalance is in phase. (See Figure 2.) Combinations of amplitude and phase imbalance can place the major axis of the ellipse at any angle.

- *Frequency dependent components (antialiasing filters)* — Not only are there problems with balancing the detector at the operating frequency, but because the detector has a frequency bandwidth, this balance has to be maintained over the full operating frequency range of the detector. Referring to Figure 1, all the components in the input signal path have to maintain amplitude and quadrature balance over the full bandwidth of the lowpass filters. This includes the input power divider, the mixers and the lowpass filters. Generally, the amplitude and phase balance between the filters are the most difficult properties to maintain. Because the reference signal does

not vary, the reference divider and the 90° phase shift only have to maintain their accuracy at one frequency.

- *Correcting the imbalance* — To correct the problem of imbalance, a threefold approach was taken. First, broader bandwidth than that required was chosen. Second, mechanical trims were added for coarse adjustment. Third, correction algorithms were used to measure and to cancel the remaining imbalance. This threefold approach proved successful in meeting the required specification of 0.5° maximum quadrature phase error and 0.05 dB maximum amplitude imbalance over a 175 Hz lowpass filter bandwidth.

In the design of the detectors, the frequency response of the components and filters was chosen to be much broader than what was required for the measurement. Typically, problems with matching phase and amplitude occur at the band edges where filtering occurs. For this reason, a wider bandwidth of 660 Hz was chosen for the filters to move these transition points outside the required measurement bandwidth of 175 Hz. (See Figure 3.) Group delay is the derivative, or the slope, of the phase response with respect to frequency. For a matched phase response, one wants the difference in group delay between two filters to be equal to zero. Achieving this is easy where the group delay response is flat (phase is linear with respect to frequency), but it is more difficult where the delay has variations. For this reason, the filter bandwidth was chosen to be broader than the usable bandwidth. For the analog filters, filter types were chosen that had minimal variations in phase or the flattest group delay. This meant that Bessel filters were chosen instead of Butterworth filters. To reduce the effects of added noise caused by broader bandwidths, further filtering was performed with digital filters. Because the same algorithm is used for the I and Q channel, the digital filters are perfectly balanced in I and Q.

Mechanical trims were added in the reference path for the adjustment of phase quadrature. Because the reference frequency does not vary, trimmer capacitors were added to tweak the phase quadrature. Amplitude balance adjustments were included as gain trims in the baseband amplifiers after the IQ detector.

Correction algorithms were used as a final measure to remove the effects of quadrature and amplitude imbalance. The end result of quadrature and amplitude imbalance is an elliptic response in the XY plane. What is wanted is a means of converting this elliptic response into a circular response. By way of example, if there is a frequency difference between the signal input and the reference input, and assuming there is imbalance in the detector, an elliptic response will be generated on an XY scope display. This elliptic response can be thought of as being the sum of two circular responses rotating in opposite directions. (See Figure 4.)

Because a complex representation (I + jQ) is being used, and not merely a

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real representation, the positive frequencies (counter-clockwise rotation) can be distinguished from the negative frequencies (clockwise rotation sense). The larger of the circulars is the fundamental term, while the smaller of the two is the image term. By varying the phase between the two circular terms, one can rotate the major axis of the ellipse anywhere in the XY plane. By varying the ratio of their magnitudes, one can arbitrarily change the degree of ellipticity. The advantage of this representation is that a Fourier transform can be used to determine the fundamental and image terms from the elliptic response. If the ellipticity were caused only by amplitude imbalance, the correction would consist of only multiplying one of the axis by a scaler to make the I and Q channel gains equal. Quadrature imbalance is not as easy to remove. This type of imbalance can cause the ellipticity to be at any angle. What is needed is a way of rotating the ellipse so that its major axis is on the X axis. Then a simple multiplication of one axis by a scaler would remove the imbalance. The ellipse can be rotated by multiplying the IQ pair by a 2 × 2 rotation matrix. (See Figure 5.) The math is straightforward and can be found in Equation 7: (7)

$$\begin{bmatrix} I_{rotated} \\ Q_{rotated} \end{bmatrix} = \begin{bmatrix} \cos(\theta_{tilt}) & \sin(\theta_{tilt}) \\ -\sin(\theta_{tilt}) & \cos(\theta_{tilt}) \end{bmatrix} \begin{bmatrix} I_{measured} \\ Q_{measured} \end{bmatrix}$$

The tilt angle is the angle the major axis of the ellipse forms with the X axis. The tilt axis angle can be found through the Fourier transform.

If one samples N uniformly spaced points (0 to $N - 1$) around the IQ circle, and if the offset frequency is equal to the sample rate divided by N, the N samples will correspond to one cycle of the offset frequency. If a fast Fourier transform (FFT) is taken of the data set, the fundamental term will be in bin 1, and the image term will be found in bin $N - 1$. Each term has an amplitude and a phase. The sum of the phases from these two terms, divided by 2, is equal to the tilt angle of the ellipse. The amplitude of the major axis is the sum of the amplitudes of the bin 1 and bin $N - 1$ terms in the FFT. The amplitude of the minor axis is the difference between the amplitude of the fundamental bin 1 term, and of the amplitude of the image, or bin $N - 1$ term. The correction then corresponds to applying the rotation matrix according to Equation 7 using the tilt angle of the ellipse and then by applying the necessary scaling to the X and Y axis. To correct the major axis, which has now been rotated

Spectrum analyzer	
advantages	• simple to implement
disadvantages	• can't distinguish between amplitude and phase noise • low sensitivity • mplitude detector doesn't directly measure phase
Phase lock or quadrature detectors	
advantages	• high sensitivity
disadvantages	• amplitude variations degrade the calibration • may require removal of lowpass response of PLL • extra phase lock circuitry • requires phase deviation <<<< 1 radian for sin(theta) = theta approximation
Quadrature detector with delay line	
advantage	• high sensitivity
disadvantages	• amplitude variations degrade the calibration • requires phase deviation <<<< 1 radian for sin(theta) = theta approximation • can only be used for sources with low frequency drift to maintain phase quadrature
IQ detection	
advantages	• simple to implement • directly measure phase with arctan function • requires no compensation of lowpass phase-lock loops (PLL) response because detector doesn't require locking
disadvantages	• moderate sensitivity • may require calibration of amplitude and quadrature balance

Table 1: Summary of phase noise test instruments.

o the X axis, one multiplies the X axis r I-channel data by the scaler:

$$\frac{\text{magnitude}(\text{Bin }1)}{\text{magnitude}(\text{Bin }1)+\text{magnitude}(\text{Bin N}-1)} \quad (8)$$

To correct the minor axis, which has ow been rotated to the Y axis, one nultiplies the Y axis or Q-channel data y the scaler:

$$\frac{\text{magnitude}(\text{Bin }1)}{\text{magnitude}(\text{Bin }1)-\text{magnitude}(\text{Bin N}-1)} \quad (9)$$

This completes the correction process. When measured data with ellipticity are assed through this algorithm, the imbalances are measured and removed vith the end result being circular data. The actual physical realization of the system includes noise-controlled oscillators (NCOs) for doppler compensation in he upconverter and the downconverter. These oscillators can be used with the ignal path in an internal closed loop to enerate the required difference frequency or beat note that is required for he calibration of the IQ detector.

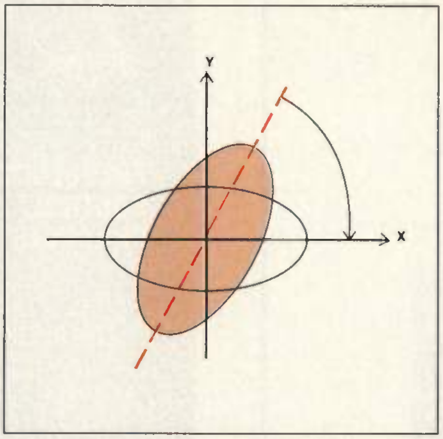

Figure 5. Rotation of major axis to X-axis

Digital averaging and processing

Digital filtering also is required for he processing of phase. IQ data are sampled at a 5 kHz rate, corrected and then block-averaged for 100 data points, such that the decimation results in a 50 Hz data rate. After this, the arctan is taken of this data to yield true phase data. Data are accumulated for 300 seconds.

The data now includes a random and a seemingly non-random term. When noise is sampled for a finite length of time, the low-frequency components of noise will appear to be deterministic. Also, because of temperature and other cyclic effects, systematic errors in the measurement will occur. These effects were removed from the phase noise measurement by fitting a third-order power series to the data and then by subtracting this curve from the data. Specifications include a bound on the level of the quadratic coefficient of this power series as well as on a limitation on the standard deviation of the residual random phase noise left after the subtraction.Further filtering of this random residual data was performed with a 5 Hz finite impulse response (FIR) digital filter. The final equivalent sample rate out of the FIR filter is 10 Hz. Appropriate noise scaling was performed for each filtering process to guarantee that the final noise power was equivalent to a 1 Hz bandwidth. The final spectrum was plotted as phase spectral density in the units of rad^2/Hz relative to the carrier, and the variance of the time domain data was calculated.

Conclusion

An instrument has been designed for the measurement of close-in phase noise. Measurement accuracies in the detector of better than 0.5° maximum quadrature phase error and 0.05 dB maximum amplitude imbalance over a 175 Hz lowpass filter bandwidth have been achieved. RF

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About the author

Charles Luke received a B.S.E.E. from Georgia Institute of Technology in 1975. Since then, he has worked for the Federal Aviation Administration, the Georgia Tech Experiment Station, Gulf Applied Research, Scientific Atlanta and the Amoco Oil Research Center. In 1988 he began Luke Engineering, a consulting company in the Atlanta area (7915 Robin Road, Cumming, GA 30131. Tel. 770-781-9443). Past design experience has included radar systems, nuclear magnetic resonance (NMR) systems, phase-noise test instruments, RF hyperthermia systems and the design of RF test instrumentation. He works part-time with his wife as the missions pastors of Christ Fellowship Church in Dawsonville, GA.

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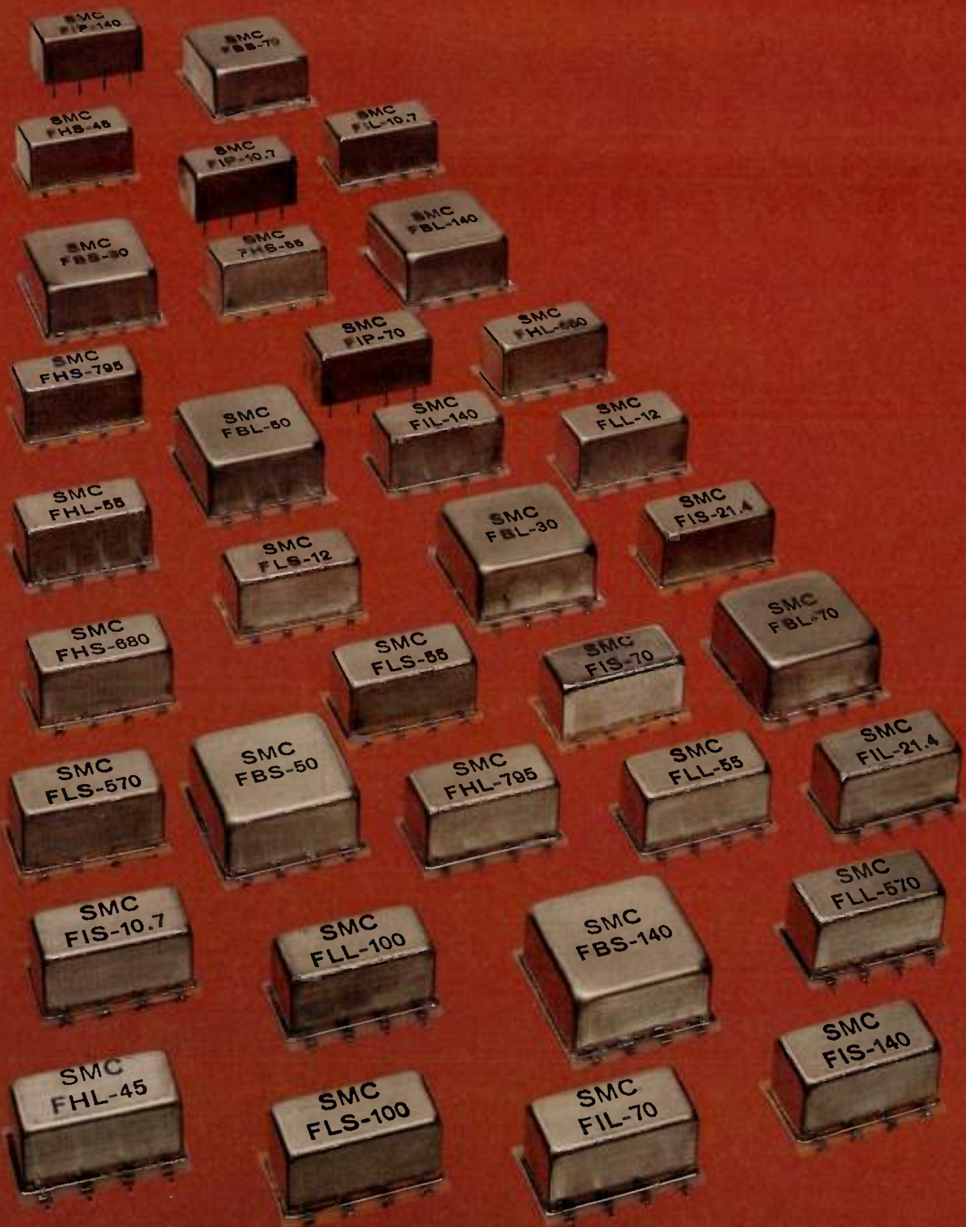

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A simple Chebychev bandpass impedance-matching program

By Yves Borlez

When it comes to designing amplifiers or antenna matching networks, the requirements for voltage standing wave ratio (VSWR) are, most of the time, given for a specified bandwidth. On the basis of the Chebychev synthesis, it is possible to create a filter that can absorb the reactive part of the source and load—if they can be represented by a series or parallel connection of a resistor with a reactive impedance. One interesting feature of the program described below is that it gives the theoretical limits relative to an infinite order network. This is helpful in selecting the filter order required or to eliminate poor load conditions. This program allows the user to match complex sources to complex loads (interstage matching).

All loads and sources are represented by a series or parallel combination of a resistor with an imaginary impedance changing with frequency. When the frequency is swept across the useful bandwidth, such an impedance can be represented by an arc in the Smith chart that is located on the circle corresponding with the resistance or the conductance. As a first step, it is necessary to make the load and source resonate at the center frequency:

$$f_0 = \sqrt{f_l f_h} \tag{1}$$

This resonance will be achieved by nulling the imaginary part of the impedance at f_0. In the case of lumped elements, one will try to obtain a resonance at f_0 by adding a capacitor or an inductance. For distributed elements, resonance could be acheived using lines.

Once load and source are resonant, two parameters must be calculated: the source Q and the load Q. They can be calculated easily from the equivalent circuit's components values or by looking at the Smith chart. In that case, the impedance Q is given by the ratio:

$$\frac{f_0}{|f_2 - f_1|} \tag{2}$$

where f_1 and f_2 are the two frequencies, the resistive part of the impedance or conductance is equal to the imaginary part and f_0 is the frequency, and the imaginary part of the impedance or conductance is 0.

Subsequently, Q_S and Q_L will be normalized to the bandwidth Q defined as:

$$Q_T = \frac{\sqrt{f_l f_h}}{f_h - f_l} \tag{3}$$

This is necessary to normalize frequency for the lowpass prototype:

$$q_s = \frac{Q_s}{Q_T}, \qquad q_L = \frac{Q_L}{Q_T} \tag{4}$$

Chebychev filter properties

Low pass prototype — The matching procedure is based on Chebychev polynomial synthesis. This requires the calculation of the g_i coeffficents of the lowpass prototype. Closed-form expressions (Equations 5–10) exist for that calculation [1].

$$a = \operatorname{arcsinh}\left[\sin\left(\frac{\pi}{2N}\right)\left(\frac{1}{q_{out}} + \frac{1}{q_{in}}\right)\right] \tag{5}$$

$$b = \operatorname{arcsinh}\left[\sin\left(\frac{\pi}{2N}\right)\left(\frac{1}{q_{out}} - \frac{1}{q_{in}}\right)\right] \tag{6}$$

$$g_0 = 1 \tag{7}$$

$$g_1 = \frac{2\sin\left(\frac{\pi}{2N}\right)}{\sinh(a) - \sinh(b)} \tag{8}$$

$$g_i g_{i+1} = \frac{4\sin\left(\frac{2i-1}{2N}\pi\right)\sin\left(\frac{2i+1}{2N}\pi\right)}{\sinh^2(a) + \sinh^2(b) + \sin^2\left(\frac{i\pi}{N}\right) - 2\sinh(a)\sinh(b)\cos\left(\frac{i\pi}{N}\right)} \tag{9}$$

$$g_N g_{N+1} = \frac{2\sin\left(\frac{\pi}{2N}\right)}{\sinh(a) + \sinh(b)} \tag{10}$$

where N is the network order and q_{in}, q_{out} are the input output quality factors.

From those expressions, it is important to note that th specification of both q_{in} and q_{out} allows the calculation of al the g_i.

Prototype performances — The two parameters *a* and *b* ca also give the value of the return loss inside the passband. Th insertion loss and return loss of the lowpass prototype ar given by equations 11 and 12. For the Chebychev lowpas prototype, the insertion loss vs. frequency follows alternat maxima and minima. There can be a non-zero minimum los inside the passband. This is equal to K^2. The passband rippl is equal to ε^2. As a consequence, the minimum and maximur return loss in the passband will be given by Equations 13 an 14 from [4]. Expressions are also available to determine I and ε from a and b (Equations 15–18).

$$|IL|^2 = 1 + K^2 + \varepsilon^2 T_N^2\ (\Omega) \tag{11}$$

$$|\rho|^2 = \frac{K^2 + \varepsilon T_N^2\ (\Omega)}{1 + K^2 + \varepsilon T_N^2\ (\Omega)} \tag{12}$$

$$|\rho_{min}|^2 = \frac{K^2}{1 + K^2} \tag{13}$$

$$|\rho_{max}|^2 = \frac{K^2 + \varepsilon^2}{1 + K^2 + \varepsilon^2} \quad (14)$$

$$\varepsilon = \frac{1}{\sqrt{\sinh^2(N_a) - \sinh^2(N_b)}} \quad (15)$$

$$K = \frac{\sinh(N_b)}{\sqrt{\sinh^2(N_a) - \sinh^2(N_b)}} \quad (16)$$

$$|\rho_{min}| = \left|\frac{\sinh(N_b)}{\sinh(N_a)}\right| \quad (17)$$

$$|\rho_{max}| = \left|\frac{\cosh(N_b)}{\cosh(N_a)}\right| \quad (18)$$

Once more, if *a* and *b* are known (q_{in} and q_{out} are known), all the matching performances of the filter will be computable. The principle of matching will consist of finding the best (q_{in}, q_{out}) choice to achieve a minimum insertion loss over the useful bandwidth.

Limit values for N infinite — When increasing the filte order, it is possible to lower the insertion loss, but it will b limited to an asymptotic value depending only on the (q_{in}, q_{ou} choice. By replacing *a* and *b* as functions of q_{in} and q_{out}, an taking the limit for N going to infinite, equations 19 and 2 give the limit values. In the special case where one of th $q \to 0$, it is simple to obtain equations 21 and 22:

$$\lim_{N\to\infty} K = \frac{1}{\sqrt{\dfrac{\sinh^2\left[\dfrac{\pi}{2}\left(\dfrac{1}{q_{out}} + \dfrac{1}{q_{in}}\right)\right]}{\sinh^2\left[\dfrac{\pi}{2}\left(\dfrac{1}{q_{out}} - \dfrac{1}{q_{in}}\right)\right]} - 1}} \quad (19)$$

$$\lim_{N\to\infty} \varepsilon = \frac{1}{\sqrt{\sinh^2\left[\dfrac{\pi}{2}\left(\dfrac{1}{q_{out}} + \dfrac{1}{q_{in}}\right)\right] - \sinh^2\left[\dfrac{\pi}{2}\left(\dfrac{1}{q_{out}} - \dfrac{1}{q_{in}}\right)\right]}} \quad (20)$$

$$\lim_{\substack{N\to\infty \\ q_{out}\to 0}} K = \frac{1}{\sqrt{e^{\frac{2\pi}{q_{in}}} - 1}} \quad \text{and} \quad \lim_{\substack{N\to\infty \\ q_{out}\to 0}} \varepsilon = 0 \quad (21)$$

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$$|\rho_{max}| = e^{-\frac{\pi}{q_{in}}} \tag{22}$$

which is precisely the theoretical limit given by Fano [2,3]. Those asymptotic values will be compared to the ones obtained for a finite filter order N. It is possible to choose the matching effectiveness compared with the theoretical limits.

Matching procedure

It has been shown that all the Chebychev filter properties may be calculated from (q_{in}, q_{out}). The matching between source and load will consist in finding the optimum q_{in} and q_{out} with respect to the following conditions :

1. q_{in} and q_{out} will always be greater than or equal to q_S and q_L respectively. This is necessary to ensure that the reactive absorption will be made on both sides of the filter.

2. If q_S is greater than q_L, then q_{in} will be equal to q_S and q_{out} will be optimized. If q_L is greater than q_S, then q_{out} will be equal to q_L, and q_{in} will be optimized.

3. The matching program will adjust the remaining q_{in} or q_{out} to minimize the total insertion loss over the passband.

This last step requires an optimization procedure that consists of finding the value of the remaining q_{in} or q_{out} to minimize the maximum insertion loss. It can be shown [1] that the optimum values of *a* and *b* satisfy equation 23:

$$\frac{\tanh(Na)}{\cosh(a)} = \frac{\tanh(Nb)}{\cosh(b)} \tag{23}$$

Because the function is well conditioned, a simple optimization routine is able to find the optimum value. That value will be compared with the q_L ($q_S > q_L$ case) or q_S ($q_L > q_S$ case), and the maximum will be taken. In some cases where there is a big difference between q_S and q_L, it will be necessary to increase the smaller q to achieve the optimum matching (see Example 2). There will be exact absorption on one side and excess absorption on the other side.

Up to now, nothing has been mentioned about the filter components nor about the load and source resistors' values. To realize a practical matching network, one will use the Chebychev prototype filter coefficients to obtain an actual filter. It appears that, most of the time, the filter requires an ideal transformer to adjust the impedance level to the load and source values. This ideal component will be further removed by the use of network transformations (Tapping, Norton transformations and Kuroda transformations)[5]. The location of the transformer will determine the impedance level of each component of the filter during denormalization. So, the program provides the components' values for two transformer locations; close t the source and close to the load.

The peculiar case of both q_S and q being equal to zero is used for tradition al filter synthesis. In this case, ε and K are specified, and the program calcu lates q_{in} and q_{out} to match the require values of ε and K.

Lumped components realization of the matching network

From the filter prototype coefficients all the filter elements can be calculate using traditional denormalizing tech niques. It is then easy to obtain th inductance and capacitance actual val ues. The program gives these values an includes an ideal transformer necessar for resistive impedance scaling. The pro gram output gives the solutions for th transformer on each side of the network so it is possible to have the values o each component independent of its posi tion relative to the transformer.

This ideal transformer will be mos often removed by several design tech niques. The most simple way would b to use an inductance tapping to realiz the transformation and the inductanc at the same time. This complicate slightly the inductance realization Other type of transformations, calle Norton transformations, can be used They consist of transforming a couple o inductances or capacitances into thre components of the same type, that hav an impedance scaling property. The provide truly broadband equivalencies A complete set of those transformation is given with Figure 1.

On the first results window, the pro gram gives the maximum Norton trans formation ratio available with the filte components. In [4], a matching progran has been given to optimize the use of th Norton transformation. By using th maximum transformation ratio, th number of components can be lowere because the Norton network only pro duces two components instead of thre The drawback is non-optimum perfor mance in terms of return loss.

Distributed realization of the matching network

When the g_i values of the prototype fil ter are known, other matching to pologies may be used to build the filte As an example, it is possible to construc a matching filter based on the use o quarterwave shorted stubs separated b quarterwave lines (see Figure 2).

This topology presents some limita

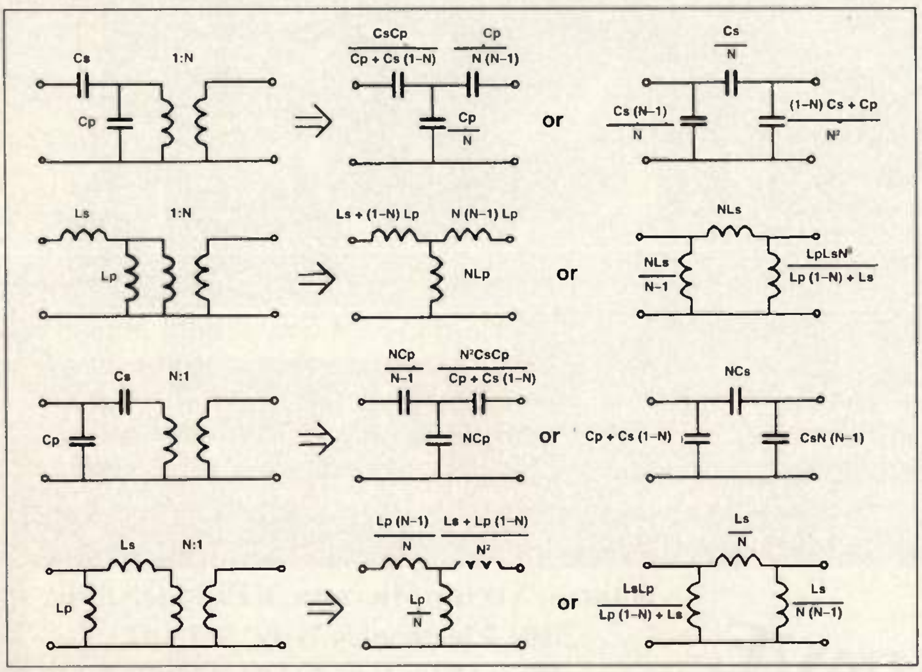

Figure 1. Norton transformations.

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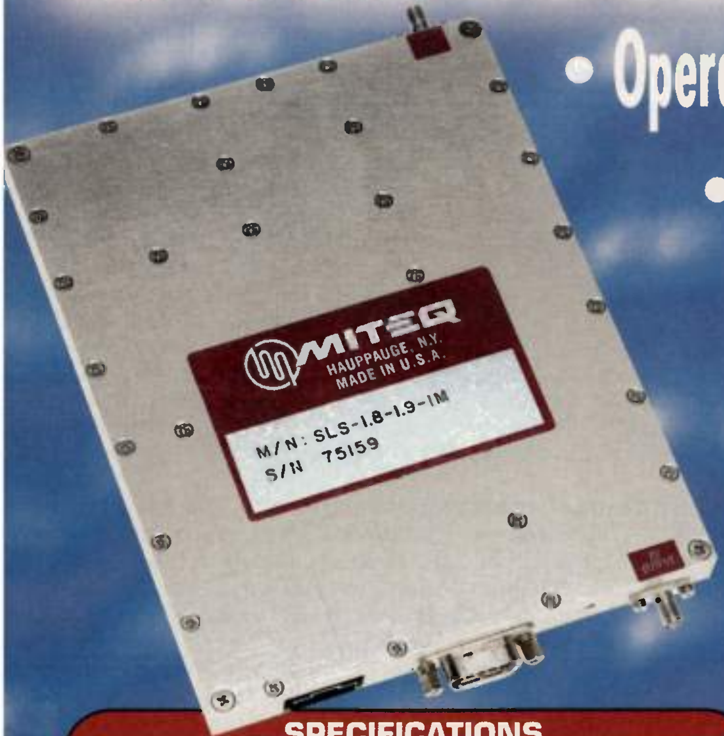

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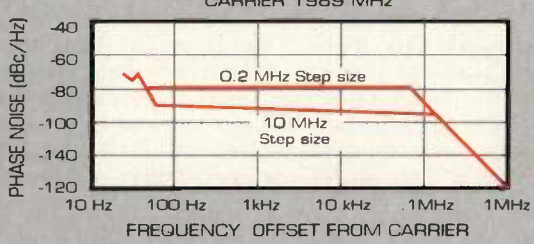

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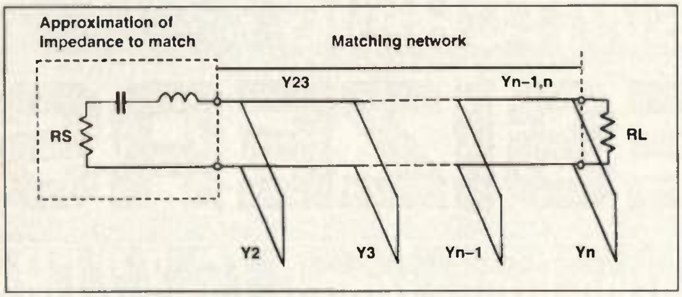

Figure 2. Distributed matching network.

BEST PERFORMANCES FOR N = 3			
Source R:	50.000	Load R:	25.000
Source Q:	0.000	Load Q:	5.000
Input Q:	2.890	Output Q:	5.000
IL min:	0.058 dB	IL max:	0.117 dB
Ripple: 0.060 dB			
RL max:	18.791 dB	RL min:	15.736 dB
ULTIMATE THEORETICAL LIMITATIONS (N INFINITE)			
IL min:	0.029 dB	IL max:	0.029 dB
Ripple: 0.000 dB			
RL max:	21.842 dB	RL min:	21.842 dB
RL/RS ratio:		0.500	
Impedence transformation requested:		62.986×10^{-2}	
Maximum norton down transformation:		25.679×10^{-4}	

Table 1. Results with a network order of three.

tions on the impedances that must be matched [6]. The first is that matching is possible only between a complex impedance and a resistive one. The reason for this limitation is that it will be difficult to increase, artificially, impedance Q to adjust the filter. For programming convenience, only the source is allowed to be complex when using the program. The second constraint is that the impedance must be series resonating at the center frequency. This problem can be solved for shunt resonating impedances by putting a quarterwave line between the source and the matching circuit to get a series resonance. This is made at the expense of a slight increase on impedance Q that is, most of the time, acceptable.

Such a filter allows one more degree of freedom, which is

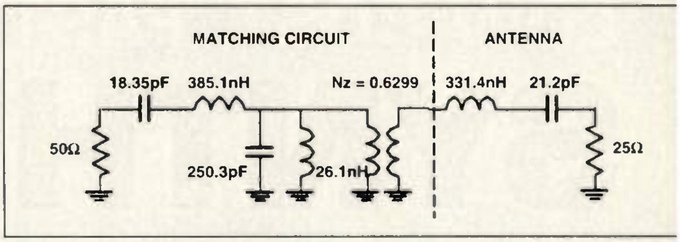

Figure 4. Antenna matching network.

the possibility of matching the resistance value at each side Moreover, it is possible to adjust the impedance level of each filter stage to get realizable lines. This adjustment will be made using the parameter D. Following [6 pg 703], one should choose a value of D between 0.5 and 1. The program proposes a starting value at 0.75.

The designer must pay attention to the fact that this realization is only an approximation convenient for transmission lines. It is possible to see some differences between the predicted performances and the one obtained through simulation of the complete circuit. For bandwidth up to 25%, the difference will be relatively small. Nevertheless, some adjustments can be done on the circuit elements' values to obtain good results quickly. An example of matching procedure following this method is explained as Example 3.

Example 1: antenna matching with lumped elements

A VHF antenna of a handheld transmitter consists of a 70 cm whip antenna. The frequency bandwidth is between 53 and 68 MHz. The antenna has a series impedance that has been made resonant at 60 MHz by an adjustable inductor. By reading the markers on the network analyzer, the measurement indicates that this antenna exhibits a resonance resistance of 25Ω and that the antenna quality factor is equal to five. The high value of the antenna Q is due to the fact that the antenna length is far from quaterwave length resonance. Its impedance is then capacitive, and a series inductor is needed to achieve resonance at mid frequency. When only the series inductance is present, the return loss curve follows Figure 3.

A better matching could be achieved at the center frequency by using an ideal transformer to transform 25Ω into 50Ω. Nevertheless, the return loss values at the band edges are poor. We are going to choose a network order of three. This gives the results shown in Table 1.

With an infinite network order, it is possible to achieve a return loss of –21.8 dB. For the present application, it has been decided that the performances of the third order network were sufficient. The prototype Chebyche filter coefficients are given:

$$\begin{aligned} g_0 &= 1 \\ g_1 &= 0.7221 \\ g_2 &= 1.179 \\ g_3 &= 0.9917 \\ g_4 &= 1.260 \end{aligned} \qquad (23$$

The program then calculates all the lumped elements of the filter, giving the results shown in Table 2, which corresponds to the drawing of Figure 4.

In this case, the ideal transformer can be replaced by a tapping on the 28.1 nH

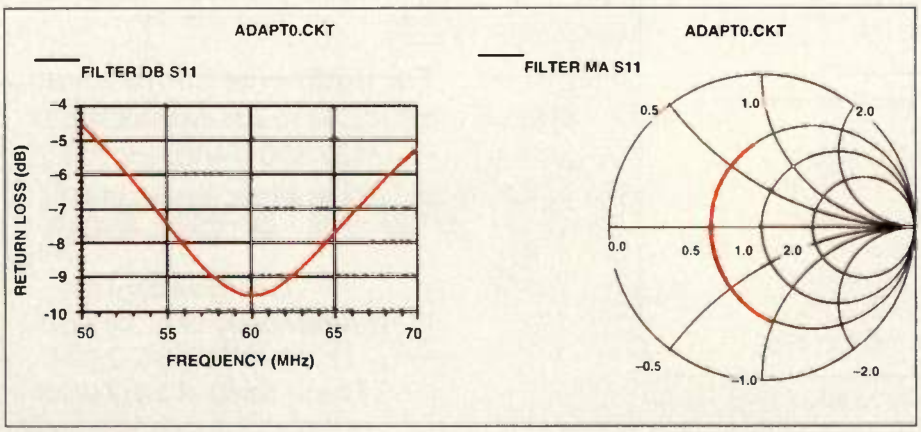

Figure 3. Antenna impedance locus.

Source R: $0.5000\times10^2\ \Omega$		
C_1:	0.1835×10^2 pF	S
L_1:	0.3831×10^3 nH	S
C_2:	0.2503×10^3 pF	P
L_2:	0.2808×10^2 nH	P
Transformer: Nz = 0.6299, Nv = 0.7936		
Load C:	0.2121×10^2 pF	S
Load L:	0.3314×10^3 nH	S
Load R:	$0.2500\times10^2\ \Omega$	

'able 2. Transformation with discrete compo- ients and ideal transformer.

nductance. The tapping ratio is simply ;iven by :

$$\sqrt{0.6299} = 0.7936 \qquad (24)$$

vhich can be easily realized on such a ımall inductance because it is com- ›osed of one turn. That circuit gives the esults of Figure 5.

One can see clearly the exact absorp- ion of the imaginary parts of the anten- ıa impedance. The VSWR values ob- ained are identical to the one predicted.

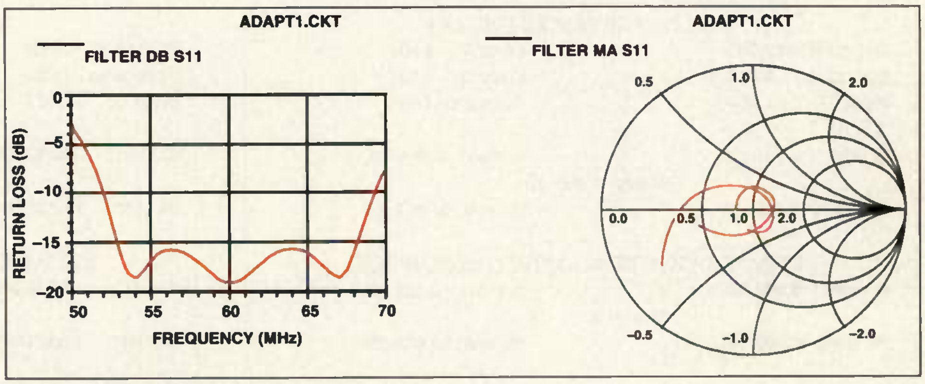

Figure 5. Antenna matching network results.

Example 2: interstage matching network with lumped elements

In this example, the program is used o match the complex output imped- ınce of a FET with the complex input mpedance of a subsequent stage using he same transistor. Between 130 and 200 MHz, it has been calculated from S parameter data that, for a simultane- ous conjugate match, a given field effect ransistor (FET) has input and output mpedance equivalent circuit at 161 MHz given by Figure 6.

The source and load Q's are calculat- ed for the center frequency of 161 MHz. They are equal to 0.5142 and 1.925, espectively. As the source and load ıave different configurations, we must hoose an even order filter. We take N = 4. From those values, the program gives the predictions shown in Table 3.

In this case, it is important to see that he optimum value of the input Q is higher than the source Q. This high- ights the fact that perfect absorption on both sides is not necessarily the best solution. The reason why we have chosen N = 4 is to ensure adequate Norton ransformation ratio to match the two esistors, which are quite different. After network calculation, the program gives he two solutions shown in Table 4.

The matching procedure can be achieved with Figures 7 and 8. On the eft-hand side of Figure 8, the two shunt nductances have been unified as one. A Norton transformation has been made with the 64 nH inductance, the ideal transformer, and the series inductance of 17.1 nH. First the two inductances are put on the same side of the trans- former and the Norton transformation is calculated using three inductances. The simulation results relative to that net- work are given in Figure 9.

Example 3: designing matching network with distributed elements

An X-band FET transistor has an output impedance that can be repre- sented on the Smith chart by the fol- lowing locus (Figure 10). The first step consists of a simple line transformation to obtain a series resonant equivalent

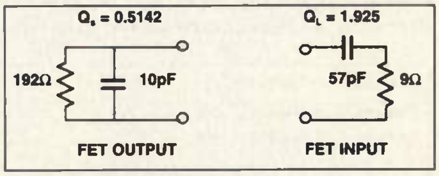

Figure 6. FET equivalent circuit.

BEST PERFORMANCES FOR N = 4		
Source R: 192.000		Load R: 9.000
Source Q: 0.514		Load Q: 1.925
Input Q: 1.324		Output Q: 1.925
IL min: 0.007 dB		IL max: 0.014 dB
	Ripple: 0.008 dB	
RL max: 28.240 dB		RL min: 24.878 dB
ULTIMATE THEORETICAL LIMITATIONS (N INFNITE)		
IL min: 0.002 dB		IL max: 0.002 dB
	Ripple: 0.000 dB	
RL max: 32.653 dB		RL min: 32.653 dB
RL/RS ratio:		0.047
Impedence transformation requested:		52.545×10^{-3}
Minimum norton down transformation:		13.272×10^{-4}

Table 3. Source and load Q's for the center frequency of 161 MHz.

BEST PERFORMANCES FOR N = 3		
Source R: 83.000		Load R: 50.000
Source Q: 7.000		Load Q: 0.000
Input Q: 7.000		Output Q: 3.986
IL min: 0.063 dB		IL max: 0.126 dB
	Ripple: 0.063 dB	
RL max: 18.407 dB		RL min: 15.421 dB
ULTIMATE THEORETICAL LIMITATIONS (N INFINITE)		
IL min: 0.032 dB		IL max: 0.032 dB
	Ripple: 0.000 dB	
RL max: 21.351 dB		RL min: –21.351 dB
RL/RS ratio:		0.602
Impedence transformation requested:		47.320×10^{-2}
Minimum norton down transformation:		13.849×10^{-4}

Table 5. Computer results for a network order three.

circuit. This is achieved by placing a 20Ω line with 50° electrical length followed by an 80Ω quarterwave line. The purpose of this line is to transform a parallel resonance into a series resonance and to adjust the impedance level to allow practical values for the matching network. Here, a high real impedance is desirable to have practical shunt stubs.

Source R: 0.1920×10^3 Ω			Source R: 0.1920×10^3 Ω	
Source C: 0.2643×10^1 pF			Source C: 0.2643×10^1 pF	
Source L: 0.3686×10^3 nH			Source L: 0.3686×10^3 nH	
Transformer: Nz = 0.5254×10^{-1}			Transformer: Nz = 0.5254×10^{-1}	
	C_1: 0.7919×10^2 pF	P	C_1: 0.4161×10^1 pF	
	L_1: 0.1230×10^2 nH	P	L_1: 0.2341×10^3 nH	
	C_2: 0.3688×10^2 pF	S	C_2: 0.1938×10^1 pF	
	L_2: 0.2642×10^2 nH	S	L_2: 0.5027×10^3 nH	
	C_3: 0.2897×10^3 pF	P	C_3: 0.1522×10^2 pF	
	L_3: 0.3363×10^1 nH	P	L_3: 0.6400×10^2 nH	
Load C: 0.5697×10^2 pF			Load C: 0.5697×10^2 pF	S
Load L: 0.1710×10^2 nH			Load L: 0.1710×10^2 nH	S
Load R: 0.9000×10^1 Ω			Load R: 0.9000×10^1 Ω	

Table 4. Solutions after network calculation of the figures in Table 3.

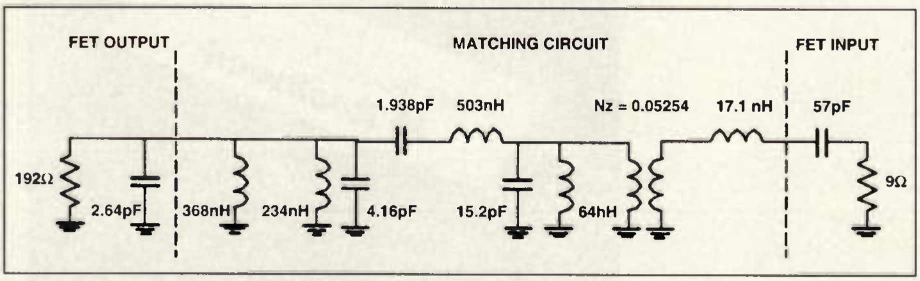

Figure 7. First amplifier matching circuit.

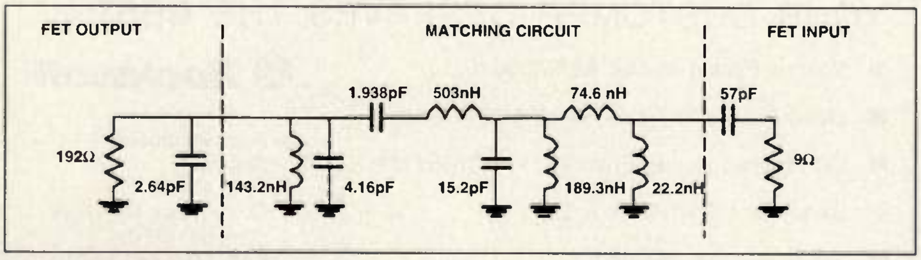

Figure 8. Final amplifier matching circuit.

From the Smith chart data, it has been seen that this new load can be approximated by a series resonant impedance, which has a real part equal to 83Ω. On the Smith chart, this load has been compared to an RLC circuit having a Q = 7. This value is the closest to the real Q value. Also observe that the curve is more closed with the impedance to match. This would be the case with the addition to the RLC circuit of a parallel shorted stub of impedance equal to 50Ω. By subtracting this influence from the first shorted stub, it will be possible to compensate. By introducing that data as being the source impedance data in the program the computer gives the results shown in Table 5 for a network order three. In the worst case, the theoretical return loss will always be better than 15 dB. Choosing the distributed realization of the filter with D = 0.75, the program gives the results shown in Table 6.

Circuit topology is shown in Figure 11 and simulation results from this circuit are given in Figure 12. All the values are as computed. The only exception is the modification of the 13Ω line to subtract the effect of the 50Ω line in parallel. This gives a final value of 18Ω.

Conclusion

The software program presented here, though relatively simple, can be used in various situations where broadband matching is needed. The program output gives first the theoretical limitations due to the impedance's Qs to allow the designer to choose the appro

Source R: 0.8300×10² Ω
Source Q: 0.7000×10¹

Shunt shorted stub 2: 0.1303×10² Ω, 90°
Quarter wave line 23: 0.7269×10² Ω, 90°
Shunt shorted stub 3: 0.1001×10² Ω, 90°

Load Q: 0.0000
Load R: 0.5000×10² Ω

'able 6. Transformation with shunt-shorted stubs and quarterwave lines.

›riate filter order. Then the lumped element values for the filters are avail- ıble. If the g_i coefficients of the Chebychev filter are known, other :opologies can be used. Distributed realizations have also been demonstrat- ed. Simulated results show good agree- nents between theoretical predictions and actual circuit performances. The simple series or parallel RLC equiva- ent circuits seem to be, in many appli- :ations, a good approximation.

The software program is available from the author. Please contact the ıuthor via E-mail. *RF*

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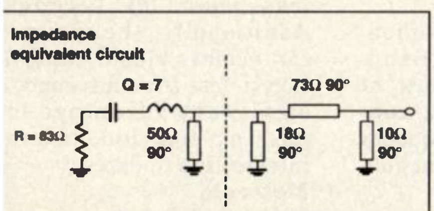


Figure 11. Matching circuit.

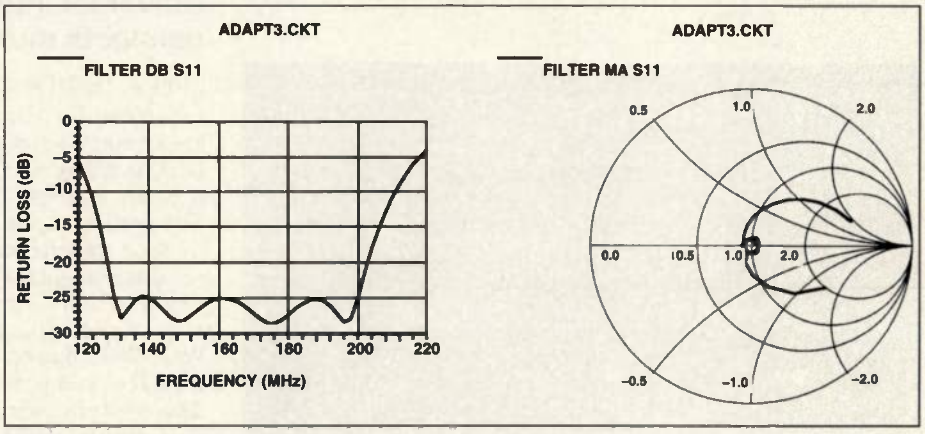


Figure 9. Amplifier matching results.

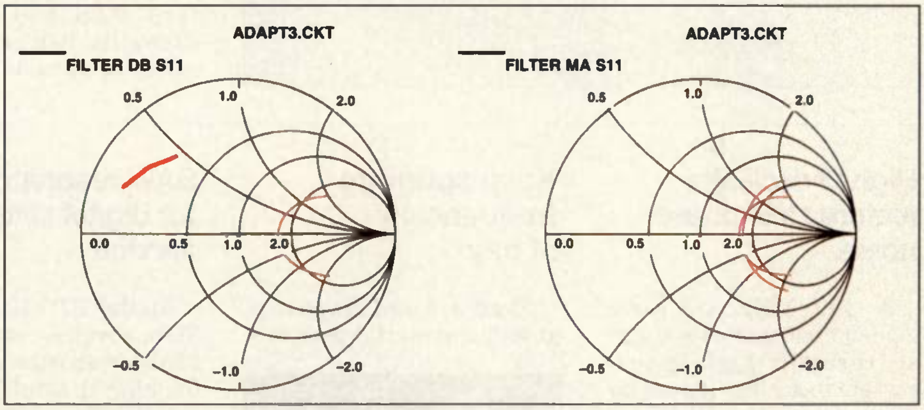


Figure 10. Initial and modified transistor impedance.

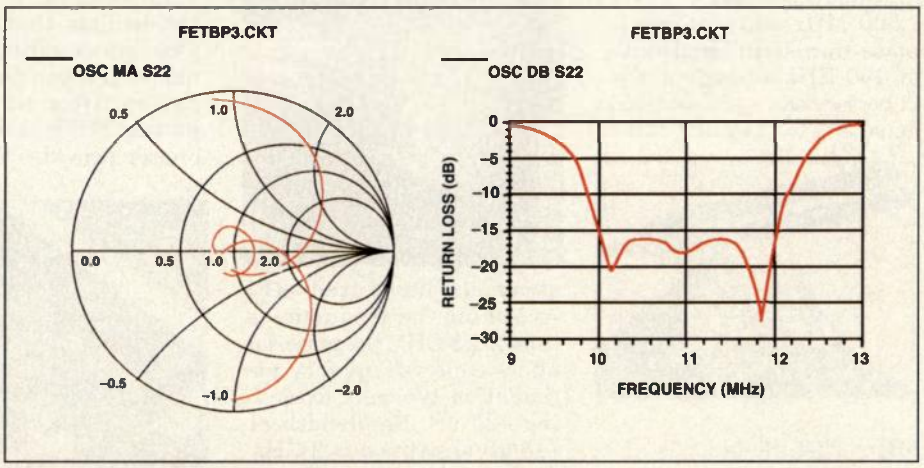
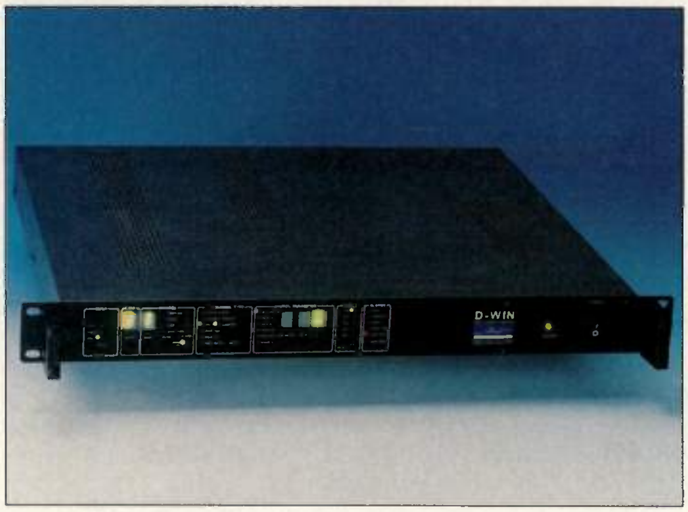


Figure 12. Example 3 matching results.

About the author

Yves Borlez received a B.S.E.E. from the University of Liege, Belgium in 1988. After earming a master's degree in microwaves and telecommunication at the University of Louvain-la-Neuve Belgium, he joined S.A.I.T. Electronics, where he has been involved in many aspects of RF circuit design in the VHF and UHF bands. Amplifiers, antennas, filters, oscillators and synthezisers are his main fields. In 1992, he joined B.E.A., where he is responsible for the research and development of microwave radar antennas and detectors for door-opening appli- cations. He may be reached at (32) 41/33.30.21., by mail at B.E.A., 47 Boulevard Frere Orban, B-4000 Liege, Belgium, or by E-mail at berb@innet.be.



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*Typical parameters at 1GHz.

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DIGITAL ATTENUATORS

Configuration	Operating Frequency (GHz)	*Attenuation (dB)	*Insertion Loss (dB)	Features	Part Number
5 Bit Digital	DC- 2.0	1,2,4,8,16	1.6	Highly accurate attenuation; low power consumption	AT-210
4 Bit Digital	DC- 2.0	2,4,8,16	1.6		AT-220
3 Bit Digital	DC- 2.0	4,8,16	1.6		AT-230

*Typical parameters at 1GHz.

VOLTAGE VARIABLE ATTENUATORS

Configuration	Operating Frequency (GHz)	*Attenuation (dB)	*Insertion Loss (dB)	Features	Part Number
VVA	.5 - 2.0	0 - 35	3.2	Best linearity, single positive control	AT-108
VVA	DC- 2.0	0 - 35	7.2	18 dBm IP3	AT-635
VVA	DC- 2.0	0 - 15	3.2	Economical; small size (50T-14B)	AT-259

*Typical parameters at 1GHz.

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CABLES AND CONNECTORS

Multiple cable simplifies antenna runs

Designed for multiple antenna installations, the Multi-LMR cable bundles eight LMR coaxial cables into a single cable. The cable incorporates an outer braid for grounding and an outer jacket for weather protection.

Times Microwave Systems
INFO/CARD 201

Integrated cable for multi-antenna sites

Micro-Coax has expanded its line of cable-integrated products to include flexible cable assemblies. Equalizers, attenuators and filters are combined with flexible cable assemblies into a small, lightweight package called UtiFlex.
Micro-Coax
INFO/CARD 202

Solderless N connector makes installation simple

A new, solderless, type N connector uses an easy 3-step assembly process, and it can be serviced in the field. The connector has a 50 Ω nominal impedance and a VSWR of 1.25:1, as high a[s] 1 GHz.
Trompeter Electronics
INFO/CARD 203

Push-on RF connectors speed cable assembly

The POD1 is a small-footprint, push on, locking RF coaxial connector fo[r] BNC and TNC applications. The push on technology allows the connector t[o] be installed with minimal effort an[d] alignment. The product provides a[n] option for space-constrained applications and features a 50 Ω nomina[l] impedance and a VSWR of 1.3:1 fro[m] DC to 3 GHz.
Hirose Electric
INFO/CARD 204

Bias tree for cellular and PCS applications

The EP-1075 bias tree, which cover[s] 800 KHz to 2 GHz, can be used to injec[t]

Product Focus: Power Meters

New and fast power meter

Model 4230 RF power meter features include a sampling rate of as many as 200 readings per second and

either an RS-232-C serial or IEEE-488 GPIB interface. The meter measures power from –70 dBm to 44 dBm and covers the frequency range from 10 KHz to 100 GHz. Sensor calibration data resides in the sensor adapter, eliminating the need for manual sensor setup. All of the principal instrument functions are menu-driven and are orchestrated via front-panel, touch-sensitive controls. A four-line, 20-column liquid crystal display (LCD) with five-digit resolution makes the measurements easy to read.
Boonton Electronics
INFO/CARD 220

Low-cost, high-bandwidth power meter

A diverse, wide-spectrum power meter that needs only three sensors to cover the frequency range from 50 KHz to 325 GHz can measure CW power from 100 μW to 1 W. The single millimeter wave sensor's typical insertion loss extension is 0.1 dB. The sensor incorporates a heat sink for temperature stability across the measurement range. The coax sensor covers the range from 50 MHz to 15 GHz with a measurement range of 1 μW to 100 mW. It uses a type N connector and has a maximum voltage standing wave ratio (VSWR) of 1.2:1. The sensor has a sensitivity of 120 mV/mW with a measurement time of 20–50 ms and an overall accuracy of 5%.
Dorado International
INFO/CARD 221

Digital and analog RF power meter

Model APM 16 measures RF power in both analog and digital systems. The compact, portable wattmeter can be used in cellular, personal communications services (PCS), industrial, scientific and medical (ISM), aviation, broadcast, paging and conventional trunked radio system measurements. The unit can measure signal from 100 mW to 10 kW for the frequency range of 450 KHz to 2.3 GHz. The unit features a shock-mounted linear scale meter to withstand abuses typically encountered in the field.
Bird Electronics
INFO/CARD 222

a DC bias current into an RF transmission line to bias an RF amplifier without loading the line. The device features a typical insertion loss of less than 0.1 dB over the entire frequency band, a voltage standing wave ratio (VSWR) of 1.2:1, a power rating of 80 W and a DC current carrying capacity of 1 A.
KDI/Triangle Electronics
INFO/CARD 205

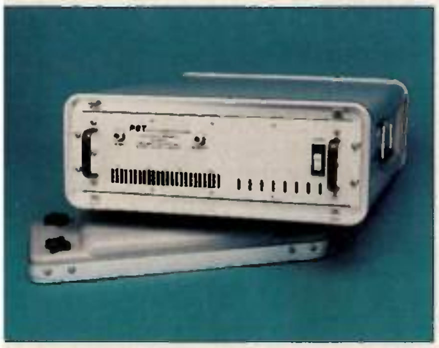
Multiple connectors made obsolete

One panel that accommodates a variety of RF connectors includes a panel mounting adapter that may be permanently mounted at any convenient location. It allows the user to change RF connectors for any particular need. The panel can accommodate 7/16DIN, BNC, C, HN, LC/HV, SC, SQ TNC and other varieties of common RF connectors. The panel is constructed of nickel-plated brass and has a Teflon insulator and brass conductor. Custom engineering is available for special requirements.
Tru-Connector
INFO/CARD 206

AMPLIFIERS

Power amplifier operates from 800–2,000 MHz

Model AR8829-50 class A high-power amplifier operates in the 800–2,000 MHz frequency band with linear output power of 50 W. Gain is 50 dB nominal, with ±1.5 dB flatness across frequency. The amplifier is protected from either open- or shorted-load conditions. Operating voltage is 115/230 VAC, 50/60 Hz. The AR8829-



50 has an operating temperature range of 0°C to 50°C and measures 5.25″ × 19″ × 22″.
Comtech Microwave
INFO/CARD 207

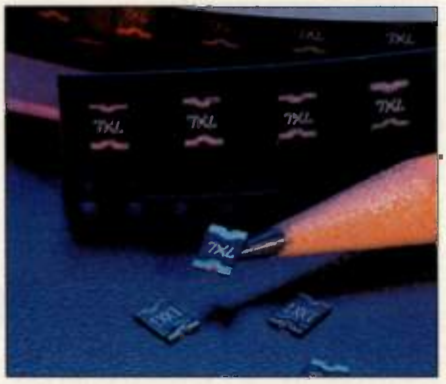
DISCRETE COMPONENTS

Fine tune components in critical installations

Microgoniometrics arcs allow precise alignment of position-sensitive components such as lasers, photo diodes and crystals in board mounting. The arcs provide a drive-key type adjustment that allows for as much as ±15° of travel. The arcs can be mounted individually or stacked concentrically with a common center to allow uniform adjustment of the stack. The arcs use nonmagnetic dry or permanently lubricated stainless-steel drive screws.
Charles Supper
INFO/CARD 208

Surface-mount resettable fuses offer reliability

Raychem has expanded its line of surface-mount technology (SMT) resettable fuses to include four new models.



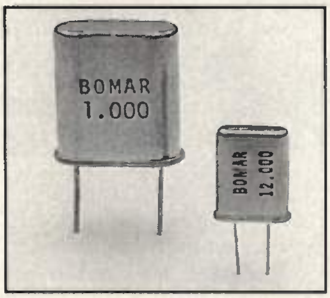
The SMDC series extends available ranges from 0.2 to 1.1 A. The products are polymeric PTC devices that break when overcurrent is applied, then reset when the condition is removed.
Raychem
INFO/CARD 209

Hybrid CATV amplifiers with tight performance

Models MHW8272 and MHW8292 are hybrid CATV amplifiers operating at 860 MHz. The amplifiers feature 27 dB and 29 dB gain, respectively. They are specified at 24 V for 128 channels.
Motorola Semiconductors
INFO/CARD 210

LC crystal filters embed band reject properties

A series of LC crystal filters for band rejection use LC filter designs in such a manner that a matched-crystal, band



reject filter is embedded in the circuit. Properly designed circuits have a bandpass insertion loss of less than 3 dB and a voltage standing wave ratio (VSWR) of less than 1.35:1.
Piezo Technology
INFO/CARD 211

SMT toroidal inductors are cost effective

Electromagnetic interference (EMI) and radio-frequency interference (RFI) power supply inductors that use surface-mount technology (SMT) are overmolded in high-temperature plastic designed to withstand surface-mount soldering procedures. The devices are encapsulated to resist moisture and to provide a more cost-effective and rugged design than glued-on or potted shell. The package features both a rounded, flat-top body for pick-and-place and flat sides for mechanical placement.
Standex Electronics
INFO/CARD 212

Timing recovery unit offers multiple features

The TRU-050 is a quartz-based phase-locked loop (PLL) with a custom monolithic microwave integrated circuit (MMIC) that combines a sophisticated phase detector, operational amplifier, crystal-drive circuitry and a voltage-controlled crystal oscillator (VCXO) drive-by feature matched with a high-Q crystal. The device is suited for data recovery and alignment, frequency translation and clock-smoothing applications.
Vectron Technologies
INFO/CARD 213

Cap voltage inverters have low voltage loss

The ADP3603 and ADP3604 high-frequency, switched-capacitor voltage inverters deliver a regulated output with low voltage loss. They eliminate the need for external inductors. The units deliver 50 mA and 120 mA, respectively, of output current with ±3% output error over the 120 kHz output-switching frequency. Output ripple is 15 mV and 25 mV, respectively.
Analog Devices
INFO/CARD 214

Chip inductors save space

Toko America's FSLU series miniature surface-mount chip inductors has an EIA standard 1008 footprint and uses a proprietary wire-wound structure with welded terminations. It uses a ferrite core for temperature stability. The inductance ranges from 10 nH to 220 µH in 5%, 10% and 20% tolerances. Operating temperatures are from –40°C to 85°C, and the components are supplied on tape and reel in 2,000-piece quantities.
Toko America
INFO/CARD 215

SUBSYSTEMS

Chip makes GPS more cost effective

A one-chip global positioning system (GPS) receiver using monolithic microwave integrated circuit (MMIC) technology allows manufacturers of GPS receivers to reduce costs. Model UPB1004GS is a 3 VDC chip that combines a double-conversion RF-IF downconverter block and a phase-locked loop (PLL) frequency synthesizer in a single package. The device has a superheterodyne architecture that features a single oscillator serving the on-chip downconverters. This architecture enables lower power consumption, high reliability, compact packaging and good out-of-band spurious signal rejection. The device is designed to process a 1,575.42 MHz GPS l1-band spread-spectrum input signal and only draws 37.5 mA for extended battery life.

California Eastern Laboratories
INFO/CARD 216

Cellular antenna routes best signal automatically

A multiple-beam array antenna with improved carrier-to-interface ratio, talkback link margin and increased coverage area operates in the 824–894 MHz cellular band and provides four separate 30° beams within each 120° sector. A controller automatically routes the signal from the beam with the best signal quality back to the base-site receiver.

Andrew
INFO/CARD 217

Broadband power amp with high gain, efficiency

A 100 W VHF and UHF broadband power amplifier is a high-power module designed to work in the 100–500 MHz frequency range. It offers 40 dB of gain (typical), 40% efficiency, 50 Ω input-output impedance, 24 VDC supply voltage at 9 A and –50° to 70°C operating temperature. The unit is suitable for radar and airborne systems and is available as a rack-mount system.

LCF Enterprises
INFO/CARD 218

Combination antenna for paging applications

A unique three-antenna combination positions three antennas under one radome. Each antenna has 5.5 dB gain, and an omnidirectional emission pattern for the 860–960 MHz frequency range. The AC90555-3 Combi has more than 40 dB of isolation between sections and can handle 400 W of power.

Celwave
INFO/CARD 219

Software accelerates design process

Spurious Analysis Professional is a computer-aided engineering program to assist the receiver and transmitter designer. The software locates where the internally generated spurious frequencies are. It is capable of calculating the spurious frequencies inherent to single-, dual- or triple-conversion design. A search option assists in finding non-tunable local oscillator and intermediate frequency (IF). The program also can perform dual modulus phase-locked loop (PLL) A & N counter calculations. Other features include: graphical plot of spurious frequencies (low-order harmonics are highlighted); spurs in the table selectable for data about an individual spurious frequency; search table enhanced to provide more information about calculations; capability to switch between search table and individual spurs while in the search mode; enhanced graphics; error trapping performed on all inputs; parameters and options menu values that can be read and saved from file; as many as 600 spurious frequencies displayed per calculation; table output that can be filtered to display only certain harmonics; ability to print output tables. An illustrated user's manual is included.
Orion Software
INFO/CARD 180

Integrated package suits circuit design projects

HP Basic RFIC Simulation Suite provides the simulation tools needed for radio-frequency integrated circuit (RFIC) design. The package was developed to shorten the design cycle in wireless applications and to enhance performance and yield by multiple simulation technologies combined with models, optimization and statistical design. When analytical models exceed their range of validity or are nonexistent, as in the case of spiral inductors on silicon, HP's electromagnetic simulators fill the void to maintain the overall RFIC simulation accuracy. Users may also develop their own linear and nonlinear models to enhance accuracy for in-house RFIC foundry processes and to fit them to measured data through a parameter-extraction system.

Features include: linear simulator that analyzes and optimizes low-noise amplifiers, matching networks and oscillators; nonlinear simulator that analyzes and optimizes mixer intermodulation products, frequency-conversion noise figure, amplifier compression, power-added efficiency, and oscillator phase noise; transient and convolution simulators that analyze transient responses of circuits containing lumped and distributed components; statistiical design package that ensures robust RFIC design to tolerate process variations with the goal of achieving acceptable production yields; custom model development kit that conveniently enables the addition of user-defined and foundry models to the simulation environment; SPICE netlist translator that imports Berkeley 2G6, PSpice and HSPICE netlists.

HP Basic RFIC Simulation Suite is available in both series IV and microwave design system (MDS) on the UNIX platform, and in series IV on the pc platform. The HP Professional Simulation Suite, which extends the basic RFIC capabilities, is available in only MDS for UNIX. Prices for both suites start at $46,000.
Hewlett Packard
INFO/CARD 181

Software features digital simulation capability

MicroSim PLSyn version 6.3 for Windows allows designers to simulate a system with programmable logic, discrete digital and analog parts on the same schematic. Designers can describe programmable logic using systhesis language, logic symbols or both. Improved error-handling in this version allows users to detect design systhesis language (DSL) syntax errors and leads directly to the source of the problem.

Language portions of the design can be specified from a list of DSL templates so that designs can be created without the user having to memorize programming language syntax.

The program is available for $2,750 and requires MicroSim PSpice or MicroSim PSpice A/D with schematics.
MicroSim
INFO/CARD 182

Program customizes probe station control

Cascade Microtech's Prober Control version 2.1 allows customized control of a probe station for on-wafer measurements and characterization. The status window displays the positions for any active MS1-8 programmable micropositioners, including remote control via standard commands for programmable instruments (SCPI). The control pad is adjustable; movement is proportional to the velocity changes in the joystick movement. Edge sense control ensures that the type of switch detected matches the hardware configuration. New dialog boxes for joystick and control pad mean more control over device movement.

Owners of earlier versions of the software may upgrade for no charge.
Cascade Microtech
INFO/CARD 183

Supercomputer software analyzes complex models

MCS/Nastran for both the NEC SX-3 and for the SX-4 series supercomputers is a finite element analysis program used for normal modes, transient response, frequency response, nonlinear systems, design sensitivity and optimization. The software provides DMAP, a macro language for creating custom applications. It also provides the capability to describe design objectives and constraints through user-written equations, making it possible to continuously update the computer model to match empirical test data.

The mathematical techniques of the large-scale program coupled with the SX series' supercomputer speed allows users to quickly analyze large, complex models.
MacNeal-Schwendler
INFO/CARD 184

Free software aids circuit design

DOS-based software from Rogers saves time in the design of prototype circuits for wireless communications applications that use Rogers laminates. If the user supplies other physical property data including dielectric constant, loss tangent and thermal conductivity, the program can assist with another substrate materials as well.

The free software may be downloaded from the company's web site http://www.rogers-corp.com/mwu/, or a diskette may be obained directly from the company.
Rogers
INFO/CARD 185

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RF literature

Study on wireless telecom highlights competitors

Kenneth W. Taylor and Associates offers a comprehensive study of the telecommunications industry. The 505-page study, entitled "Worldwide Digital Wireless Communications — Competitors, Strategies and Partners," focuses on the competitive nature of the industry. The report analyzes 94 current and future competitors that hope to become leaders in large, emerging segments of the digital wireless communications industry. The study looks at the years between 1996 and 2000.
Kenneth W. Taylor and Associates
INFO/CARD 186

Brochure covers connectors

A brochure from Applied Engineering Products details the company's latest line of connectors, including subminiature connectors for 75 Ω applications.
Applied Engineering Products
INFO/CARD 187

Handbook aids capacitor specifications

The second edition of *The Capacitor Handbook* by Cletus J. Kaiser combines capacitor theory with practical circuit application information and expands the reliability, radiation, AC motor start and nonpolar capacitor-style information. This edition also contains more charts with application information and selection factors. The first chapter covers the fundamentals of all capacitors, followed by individual chapters on ceramic, plastic film, aluminum electrolytic, tantalum, glass and mica types of capacitors.

Included are a glossary, symbols and equations, capacitor selection guidelines and a comprehensive index. This 134-page, soft-cover reference book complements the author's other passive-component handbooks on inductors and resistors. The book is available for $15.95 plus $4 shipping.
CJ Publishing
INFO/CARD 188

Full-color antenna catalog offers technical data

A catalog from Ace Antenna covers base-station antennas, mobile and portable antennas, cavities, filters, combiners, isolators, amplifiers and satellite TV antennas. Products are illustrated in color. A technical section includes worldwide Ku-band frequency allocations, noise figure to noise temperature conversion chart, rainfall γ_R attenuation coefficient, voltage standing wave ratio (VSWR) conversion chart, characteristics of standard rectangular wave guide and other formulas and data. Booklets on specific products are also available upon request.
Ace Antenna
INFO/CARD 189

Brochure addresses testing requirements

Wavetek's color 28-page *Cable Television Selection Guide* provides information on products for cable television testing. Included are the CMS 1000 central monitoring system, a CATV headend remote monitoring system and the CLI-1450 combination signal level and leakage detection meter. A separate color matrix brochure clarifies product use and details technical capabilities of the testing instruments.
Wavetek
INFO/CARD 190

Measurement booklet covers practical applications

Racal's application notes and technical papers booklet covers applications ranging from an introduction to VXI to enhanced pulse measurement and generation using VXI-based instruments. Articles also cover VXIbus system integration. The booklet is a resource for all VXIbus products and is free.
Racal Instruments
INFO/CARD 191

CAD tutorial emphasizes schematic drafting

Inside OrCAD by Chris Schroeder contains an overview and introduction to drafting, with hands-on exercises to help the reader master the software program. Chapters include installation and configuration; hierarchical design; postprocessing; library editor; advanced features; user buttons and macros; creating SPICE netlists; netlists for printed circuit board design; and editing bill of materials. OrCAD commands and functions are listed and explained. Appendices provide tips, techniques and information about linking OrCAD to other computer-aided design (CAD) and computer-aided engineering (CAE) tools. The book has an index and comes with a disk containing a parts library for the tutorial exercises as well as several utilities. The 408-page paperback is $39.95.
Butterworth Heinemann
INFO/CARD 192

On line:

Electronic components web site launched—Internet users can now access information on Vishay Intertechnology's family of passive component organizations. The web site (http://www.vishay.com) provides company and product information covering operations in North America, Asia, Europe and Israel. Companies covered include Dale, Draloric, Measurements Group, Roederstein, Sfernice, Sprague, Vishay Resistive Systems and Vitramon. Hypertext links enable users to locate product news and data for a range of passive components including capacitors, connectors, crystals, displays, inductors, multi-component networks, resistor networks, oscillators, resistors, rheostats, thermistors, transformers, trimmers and potentiometers. Data sheets, news releases, cross-reference guides, technical bulletins, distributor locations and field sales locations can be downloaded.
Vishay
INFO/CARD 193

Coax selection guide—Times Microwave Systems' web site includes its *Communications Coax Selection Guide*. An interactive attenuation calculator enables visitors to compute the loss of any of the company's LMR coaxial cable at the frequency of interest. The *Cable Handbook* also can be requested for free (http://www.timesmicrowave.com).
Times Microwave Systems
INFO/CARD 194

Fiber-optic video, audio and data links—Force's web site catalog (www.forceinc.com) details such product lines as CATVLinx; CATVLite VSB and AM multichannel video links; and AM and FM video links. A corporate profile provides information on the company, and a response form allows the user to request additional information.
Force
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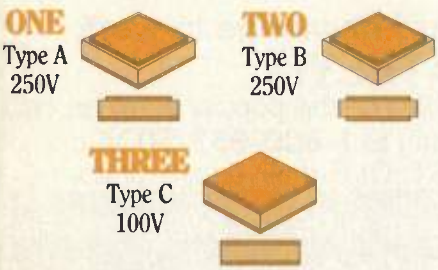

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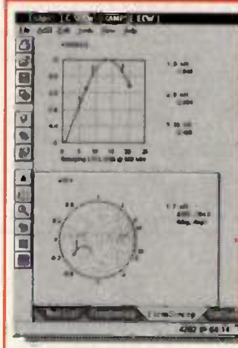

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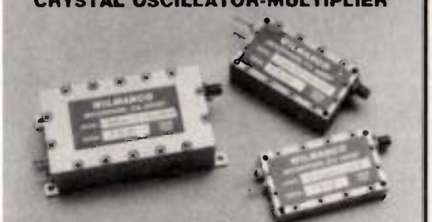

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Parameter	Test Conditions	Unit	Typ
Insertion Loss	1MHz-0.5GHz 1MHz-1GHz 1MHz-2GHz	dB dB dB	0.5 0.6 0.8
Isolation	[illegible]	[illegible]	[illegible]
VSWR	1MHz-1GHz 1MHz-2GHz	dB dB	1.2:1 1.3:1
Input Power	[illegible]	[illegible]	[illegible]

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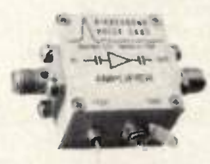

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RF Literature/ Product Showcase

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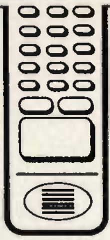

...YOUR CAREER

Baseband Circuit Design Engineer: Baseband circuit design for development of wireless phones. Requires significant experience in analog and digital design, particularly as they relate to telephone products. Working knowledge of AMPS, TDMA or CDMA is desirable.

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JMS-2L	+3	800-1000	DC-200	7.0	24	20	7.45
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JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
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DUALS	LMX2330A	LMX2331A	LMX2332A	LMX2335	LMX2336	LMX2337
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
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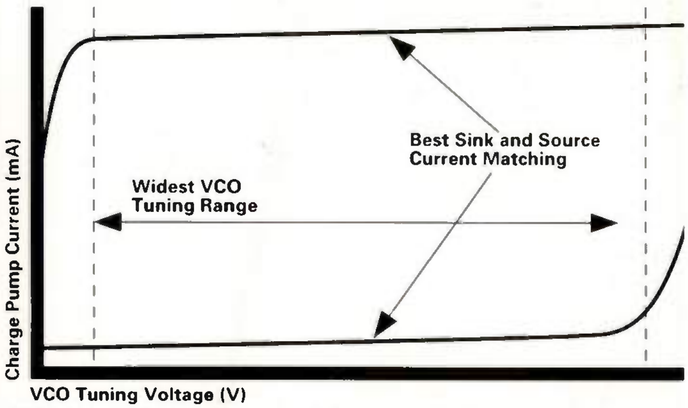

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