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engineering principles and practices

September 1996

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Microstrip techniques

Tutorial—
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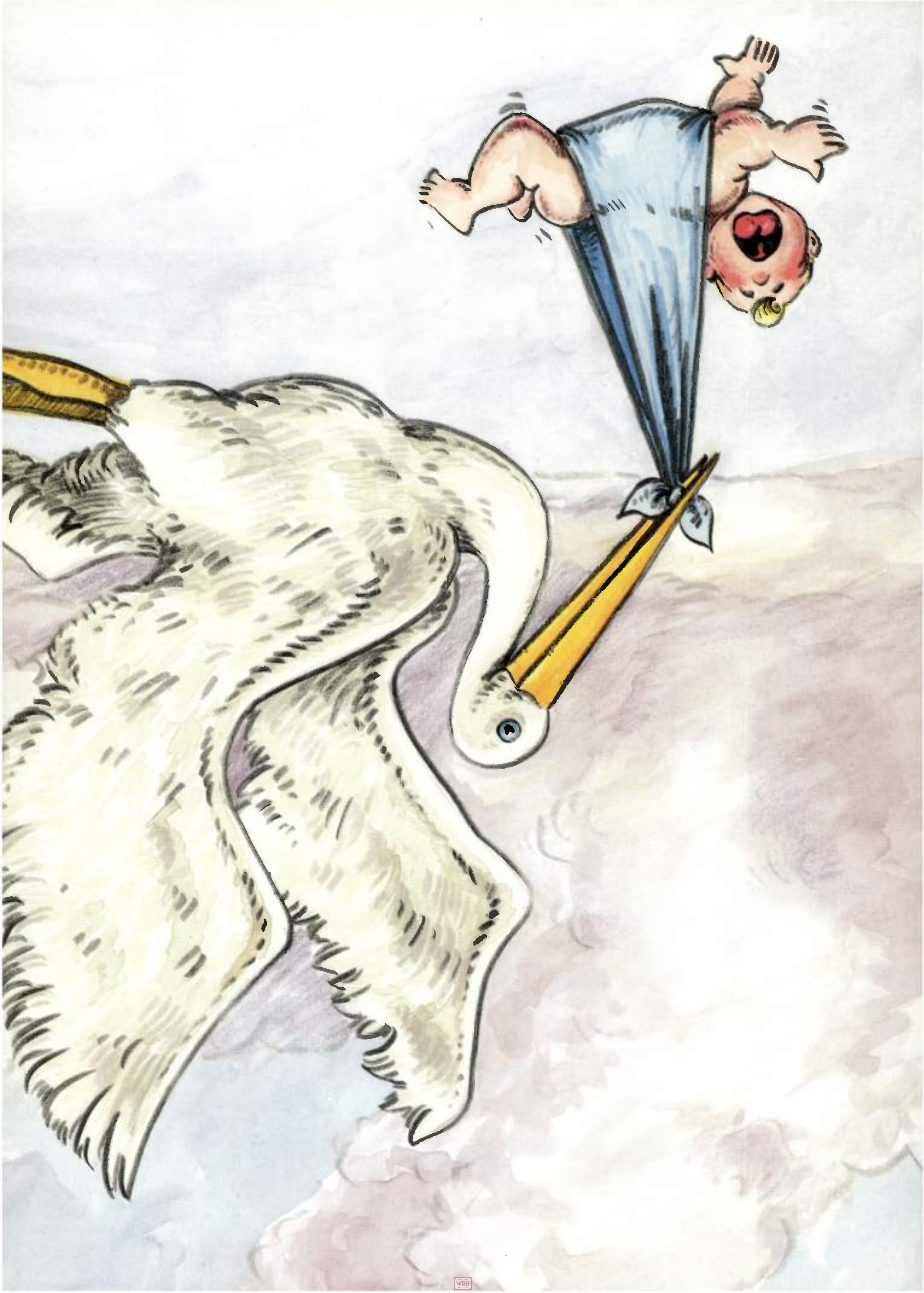

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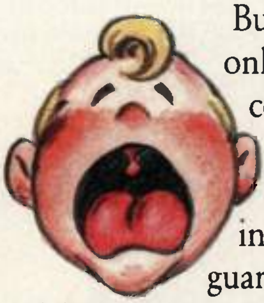

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ZOS-200	100-200	-106	0.015	-25
ZOS-300	150-280	-103	0.017	-27
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RF design

contents

September 1996

featured technology — microstrip techniques

cover story

tutorial

engineer's notebook

cover story — p.64

departments

Coming in October

- System-level design
- Phase noise tutorial

RF editorial

R & D: On the rise despite the risks

By Don Bishop
Editorial Director

Statistics indicate that, historically, research and development (R&D) spending by companies in other nations has outstripped efforts by American companies. "Foreign Participation in U.S. Research and Development: Asset or Liability," published by National Academy Press, explains why the *trend* for U.S. R&D spending is now on the *rise*. Many U.S.-based companies have been acquired by overseas businesses that have a predisposition to support R&D. During the past decade, foreign ownership of U.S. manufacturing companies moved up from 7.2% to 19.2%, nearly tripling. Along with the ownership increase, the percentage of R&D spending by U.S. private industry accounted for by foreign-owned U.S. businesses advanced from 9.3% in 1982 to 15.5% in 1993.

U.S. companies should exploit the trend of foreign involvement in R&D, rather than fight it, said Alan Schriesheim, the chairman of the report and director of the Argonne National Laboratory. Whatever the source of funds, R&D spending is good news for RF engineers. (A copy of the report can be purchased from the publisher; call 800-624-6242 or 202-334-3133.)

R&D risks

Regardless of how exciting R&D projects may seem to RF engineers, the projects serve an indisputable master: the customer. High-tech companies with a track record of leading-edge product development may devote substantially more resources to R&D, reflecting their managers' determination to be market leaders. There is a downside: aggressive competitors may reverse-engineer successful products and may market cheaper versions before their originators can recoup costs. If there is any advantage to the acceleration of the product life cycle and product obsolescence, it might be the way it limits opportunities for copycats.

Despite the risks connected with new product development, it is clear that innovations sell. RF engineers play an important role in serving one of the fastest-growing segments of the electronics industry.

* * *

Back to the future

If Tele-TV can hold its cable TV programming alliance together, next year's launch of 130 channels of digital TV programming beamed to home TV receivers from nearby transmission towers may become a reality. At stake are projected sales of a million downconverters, antennas and related accessories. Digital television, also known as multichannel, multipoint distribution service (MMDS), reminds me of the rollout of Home Box Office (HBO) service. HBO originally beamed its movies by microwave from nearby towers to home rooftop antennas, as Tele-TV will do. HBO later moved to cable TV, as Tele-TV is likely one day to move to telephone line delivery. If Tele-TV can match HBO's success, another market for RF equipment will burgeon.

RF

RF design

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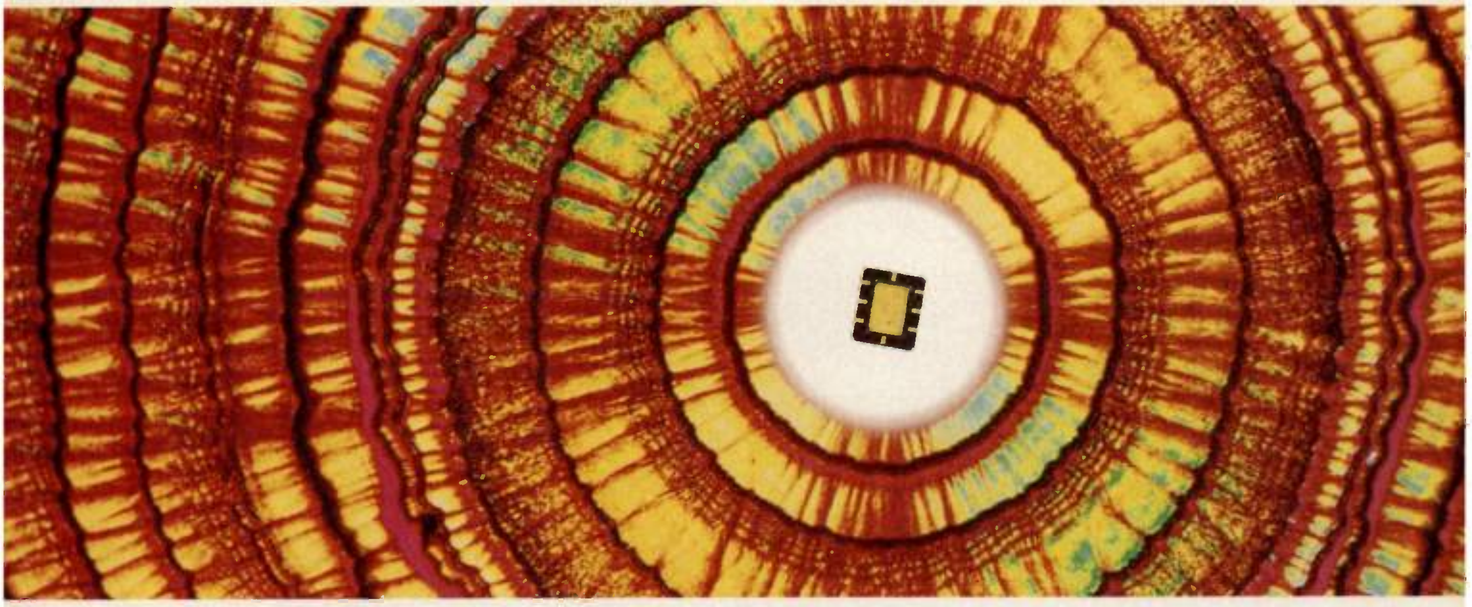

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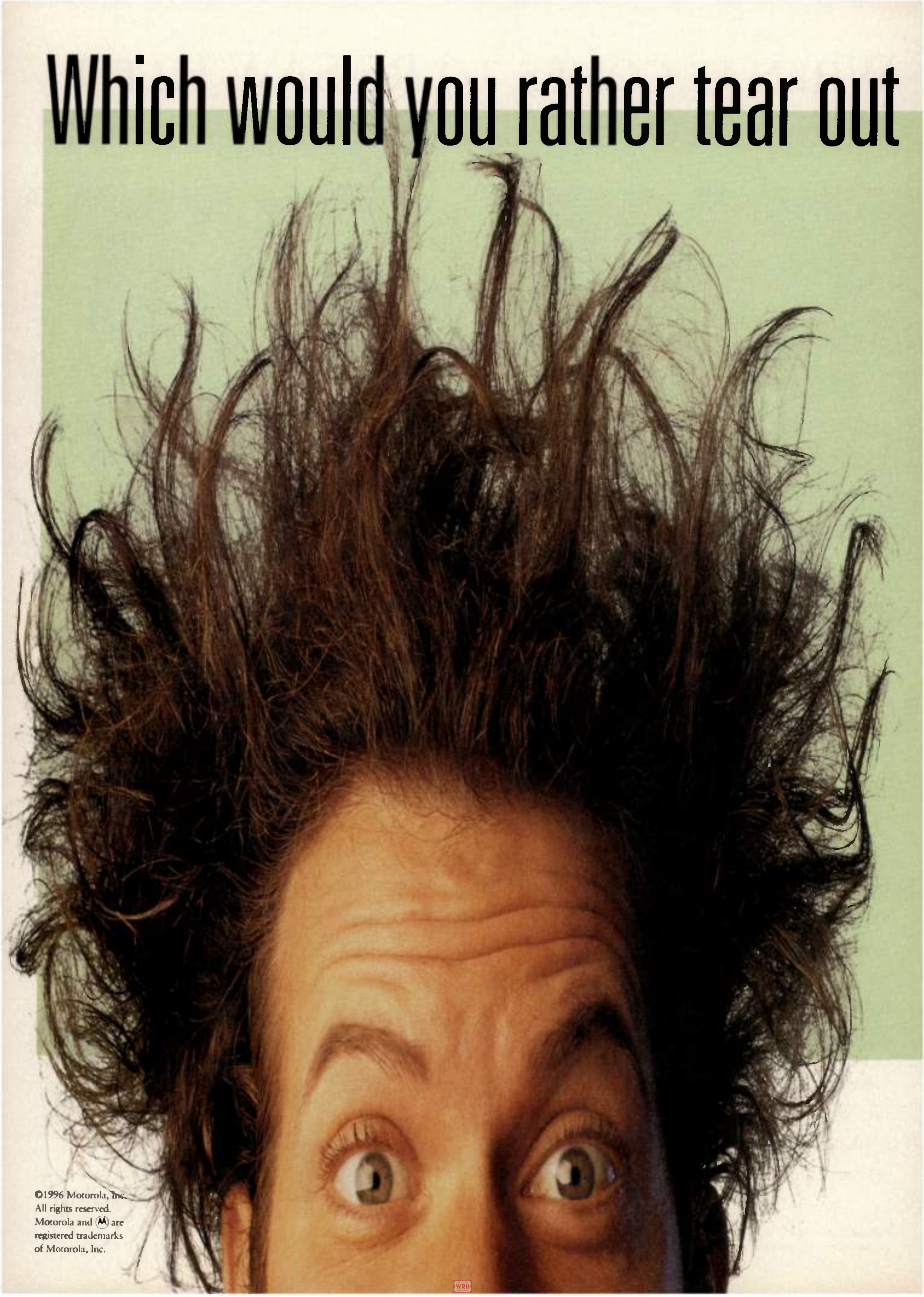
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RF letters

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Editor,

The letter criticizing the high-frequency active auroral research program (HAARP), in the June 1996 issue of *RF Design*, while giving anexcellent scientific criticism of it, neglected a more powerful political concern.

There is a widespread belief that HAARP is a cover for weather modification similar to that believed to have been carried on earlier in the Soviet Union. Whether this is true will of course never be acknowledged by the government. However, even laymen sense that there is serious misconduct afoot.

Whether justified or not, this belief is spreading like wildfire among the world's food production community. Many religious groups also are condemning the program. Both groups get most of their news via short-wave stations that allow people to tell their stories without censorship. Unprecedented changes in weather over the last year or two is being blamed on HAARP. For that reason as well as the bankrupt state of the U.S. government, HAARP ought to be shut down.

The program makes no scientific sense and follows on the heels of released news from the former Soviet Union that it did similar weather experiments with some success over the decade preceding its split up. Because no reputable scientist can defend HAARP to the public, the claim that it is a cover for weather control spreads without check.

An unforgivable failing of some scientists is that they will pursue a personal pipe dream without regard for the consequences. A classic example is Edward Teller and his ego-driven quest to create the H-bomb. Competent historians now more or less agree that this forced the Soviet Union to also pursue that weapon, when it otherwise would not have.

Other scientists beside the fellow who wrote the letter you published have a duty to speak out against bad science, funded or not, lest they end up tarred with the same foul brush.

Eugene Dusina
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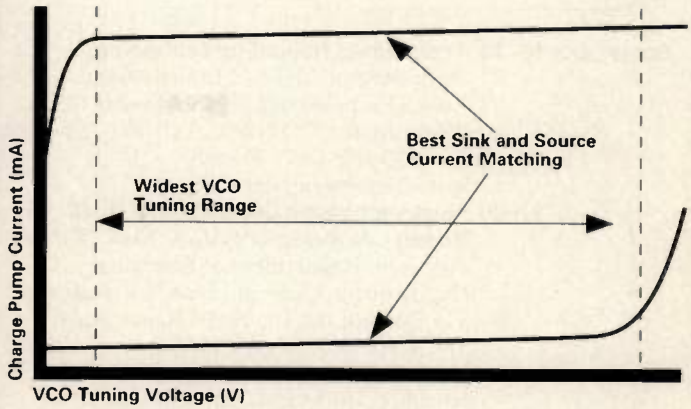

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September 16–20 Accelerated Reliability Technology Symposium—*Denver.* Information: Hobbs Engineering, 10218 Osceola Court, Westminster, CO 80030. Tel. 303-465-5988; Fax 303-469-4353; E-mail learn@hobbsengr.com.

17–20 Electromagnetic Compatibility (EMC '96 Roma)—*Rome.* Information: Prof. Mauro Feliziani, Department of Electrical Engineering, Univ. of Rome "La Sapienza," Via Eudossiana 18, 00184 Rome, Italy. Tel. +39 6 44585.809/44585. 810; Fax +39 6 4883235/4825380; E-mail emc96rom@elettrica.ing.uniroma1.it.

19–20 Electromagnetic Compatibility: Planning for Compliance in the U.S., Europe and Japan—*Phoenix.* Information: Seminars Department, Underwriters Laboratories, 333 Pfingsten Road, Northbrook, IL 60062-2096. Tel. 847-272-8800 ext. 43481; Fax 847-509-6235; E-mail seminar@ul.com.

24–26 Electrical Manufacturing & Coil Winding EMCW '96—*Chicago.* Information: Electrical Manufacturing & Coil Winding '96, Dept. 77-5053, Chicago, IL 60678-5053. Tel. 708-260-9700 or 800-323-5155; Fax 708-260-0395.

25–26 Chesapeake Electronics Show—*Chantilly, VA.* Information: Bonnie Lasky, MACC, P.O. Box 513, Colmar, PA 18915. Tel. 215-822-6319; Fax 215-822-3332.

29–Oct. 2 Wireless Workshop—*Sedona, AZ.* Information: Sharon Aspden, Rogers, 100 S. Roosevelt Ave., Chandler, AZ 85226. Tel. 602-961-1382; Fax 602-961-4533.

October 1–4 Eighth Annual Digital Audio and Video Workshop—*Philadelphia.* Information: Lisa Fasold, Consumer Electronics Manufacturers Association, 2500 Wilson Blvd., Arlington, VA 22201-3834. Tel. 703-907-7669; Fax 703-907-7690; E-mail lfasold@eia.org.

2–3 NEPCON Exhibition and Conference—*San Antonio.* Information: Customer service, Tel. 800-467-5656; Web site http://nepcon.reedexpo.com.

7–10 Signal Processing Applications and Technology—*Boston.* Information: Megan Forrester c/o Miller Freeman, 600 Harrison St., San Francisco, CA 94107. Tel. 415-356-3391; Fax 415-905-2220; E-mail dsp@mfi.com.

7–11 Wireless Technology '96—*Providence.* Information: Dawn Averyt. Tel. 407-878-8200; Fax 407-879-7388; E-mail Expo96@aol.com.

8–10 Microwaves & RF Conference and Exhibition—*London.* Information: Beverley Lucan, Nexus Information Technology, Nexus House, Swanley, Kent, BR8 8HY,

Continued on page 18

Continued from page 16

United Kingdom. Tel. +44-(0)1322-660070; Fax +44-(0)1322-661257.

13–16 Power Requirements for Mobile Computing and Wireless Communications—*Santa Clara, CA.* Information: Giga Conference Group, Tel. (in the U.S.) 800-874-9980; (in the U.K.) 44-1582-405678; Fax 617-982-1724; E-mail conferences@gigasd.com.

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.

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

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.

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.

November 4–6 Northcon Conference & Exhibition—*Seattle* Information: Electronics Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045. Tel. 800-877-2668 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.

18–22 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.

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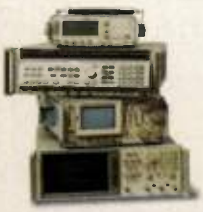

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6	TE5020	6	3.75	60	12.5	-	-	4	2.0	70	1500//+3
8	TE5030	6	3.75	60	10.0	90	12.5	5	2.0	80	1500//+3
2	TE5040	3	6.50	20	30.0	-	-	1	1.0	50	2700//0
4	TE5050	3	6.50	30	15.0	-	-	2	2.0	75	3100//0
6	TE5060	6	6.50	60	19.5	-	-	3	2.0	90	3100//0
8	TE5070	6	6.50	60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

21.4 MHz

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

45.0 MHz

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

70.0 MHz

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

90.0 MHz

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

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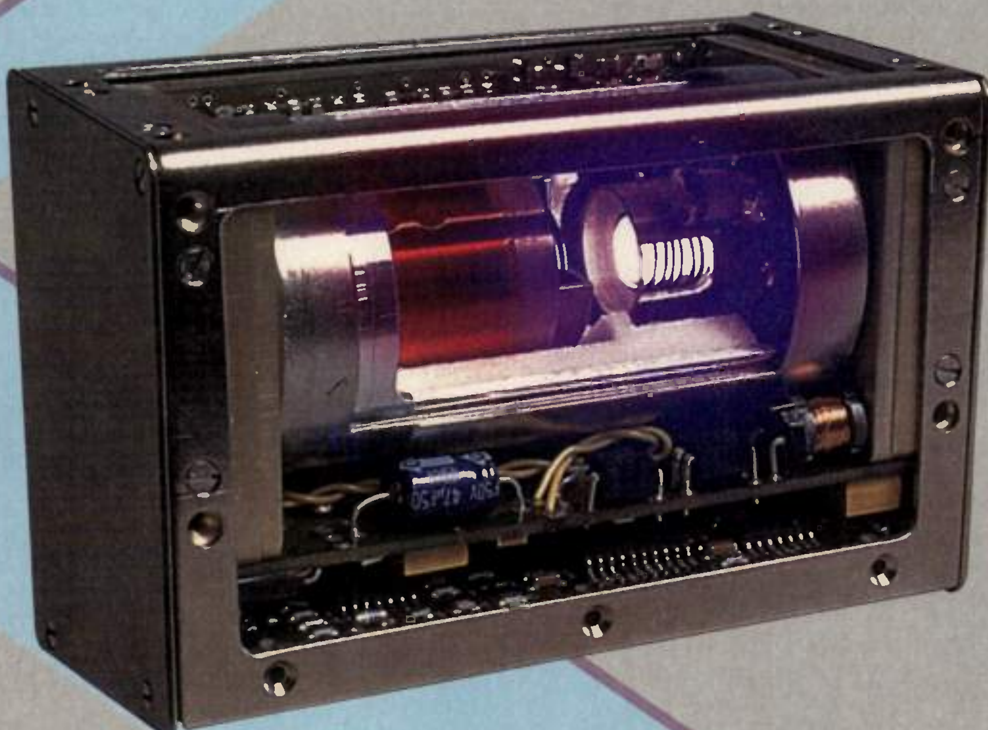

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Technology International—*Achieving and Maintaining Compliance with the Medical Devices Directive*—Oct. 15–16, Boston; Nov. 20–21, Anaheim, CA; January 14–15, Denver; February 11–12, Dallas. Information: Kristin Eckhardt, Marketing Manager, Technology 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.

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.

Communications Tech Training—1996 schedule for Orland Park, IL, Oct. 8–10, 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.

1996 CEI-Europe in Cambridge, United Kingdom—*Wireless Digital Communications: Mobile, Cellular, Personal, Voice and Data Networks*—Sept. 30–Oct. 4; *Applied RF Techniques: Linear Circuits*—Sept. 30–Oct. 4; *Adaptive Synchronous Receiver Structures for Mobile Communications*—Sept. 30–Oct. 4.
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.

DSP Without Tears—Sept. 30–Oct. 2, San Jose, CA. Information: Z Domain Technologies, 325 Pine Isle Court, Alpharetta, GA 30302. Tel. 800-967-5034 or 770-587-4812; Fax 770-518-8368; E-mail dsp@mindspring.com.

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.

Georgia Tech Continuing Education—*Near-field Antenna Measurements and Microwave Holography*—Sept. 10–13, Boulder, CO; *Advanced Electronic Warfare Principles*—Sept. 24–26, Atlanta; *Wireless Communication*—Sept. 24–27, Atlanta; *Radar Cross-section Reduction*—Oct. 29–Nov. 1, Atlanta. 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.

Johns Hopkins University Whiting School of Engineering—courses in Washington, DC. *Cellular and Personal Communications Services (PCS): Systems Engineering and Perspective with Emphasis on CDMA Technology*—Sept. 16–19; *Wireless and Personal Communications Systems*—Oct. 7–9; *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.

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.

Learning Tree International—*Wireless Networks and Mobile Communications*—Sept. 3–6, Washington; Sept. 10–13, Ottawa; Oct. 15–18, Washington; Dec. 3–6, Toronto; Dec. 17–20, 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.

INFO/CARD 25

Cesium frequency standards supplied at WAAS sites

Hughes Information Systems has selected Frequency and Time systems of Beverly, MA, to supply precision cesium atomic frequency standards for the Federal Aviation Administration (FAA)'s wide-area augmentation system (WAAS) for civil aircraft navigation. The WAAS is designed to augment capabilities of the global positioning system (GPS) provided by the United States Department of Defense in order to provide an accurate navigation system that complies with the required navigation performance critical to flight safety. The first stage of the contract is to provide cesium frequency standards for the functional verification system to verify implementation, installation, functional and operational performance prior to implementation at all operational WAAS sites.

Frequency and Time Systems supplies cesium standards to the Department of Defense's Navstar satellites.

SpectrumMaster filter has successful field trial

Illinois Superconductor held a successful field trial of the SpectrumMaster filter for cellular base stations. The trial of the superconducting cellular filter in an operational time-division, multiple-access (TDMA) digital-cellular network was conducted at a Southwestern Bell Mobile Systems commercial cell site in Dallas. Initial results of this trial indicate that the filters improve cellular call quality and reduce dropped calls, extending cell range and expanding call carrying capacity on digital networks. Digital networks are expected to comprise an increasing proportion of new systems in both cellular and PCS systems.

Western Wireless launches nation's third PCS system

Western Wireless's VoiceStream celebrated Utah's statehood centennial by launching in Salt Lake City the nation's third personal communications service (PCS) system. Mayor Deedee Corradini demonstrated the wireless service with a call from Salt Lake City to U.S. Senator Robert Bennett in Washington, DC. Willy B. Hunsaker, a descendant of golden spike driver Abraham Hunsaker, also spoke to his nephew Hyram Hunsaker, who was in Washington with the senator. PCS technology, which allows both voice and data transmissions, is the digital equivalent of the linking of the transcontinental telegraph and railroad in the late 19th century when the golden spike was driven in the transcontinental railroad in 1869.

The first PCS system was launched in Washington, D.C. by Sprint Services. Western Wireless launched the second in Hawaii on Feb. 28 of this year.

All account information is programmed into a smart card, rather than the telephone. Users can access VoiceStream services from any VoiceStream phone and potentially from PCS phones in other countries using the same network technology.

Contracts:

Compact Software for Taiwanese Design Students—Compact Software has entered a 5-year agreement with the government-sponsored Chip Implementation Center (CIC) in Taiwan. The company will provide computer-aided engineering and computer-aided design (CAE/CAD) software to all CIC-affiliated universities and colleges as the microwave and RF design software of choice. Compact Software and its Taiwanese distributor, Evergo, will provide regular training on the software over the agreement period.

National Emergency Alert System—RF Industries' Neulink Telemetry Division has received approval of the second phase of its $1 million contract to supply wireless receivers for the national emergency alert system (EAS). The Federal Communications Commission (FCC) has mandated that the nation's radio, TV and cable stations must convert from the established emergency broadcast system to the EAS. RF Industries will be the sole supplier of AM, FM, UHF and VHF band receivers for the EAS products.

Business Briefs

TCSI Completes Software for Japanese Cellular Software Market—TCSI, based in northern California, has successfully completed development and testing of software for the Japanese personal digital cellular (PDC) standard's revision C. This milestone is a result of a 1996 award of more than $1 million from a major semiconductor company to sponsor TCSI's effort in the PDC cellular software market.

CAM Software Changes Name—CAM Software Research has changed its name to Unicam Software to reflect the name of the company's flagship product. Unicam Software develops and distributes Windows-based software tools that link the design engineering and electronic manufacturing process.

Analog Devices Acquires Mosaic Group—Analog Devices, Norwood, MA, has acquired Mosaic Microsystems, Kent, United Kingdom and its wholly owned U.S. subsidiary, Mosaic. The Mosaic group designs and develops radio frequency and intermediate frequency integrated circuit products used in digital-enhanced cordless telephone, direct-broadcast satellite and other cellular telephony standards.

Georgia Tech Tests RangeStar Designs—RangeStar Telecommunications has extended its antenna performance agreement with the Georgia Tech Research Institute (GTRI). GTRI will perform both analysis and testing of the antenna technology, using advanced computer modeling to evaluate and optimize RangeStar's antenna technology designs.

Motorola and Daniels Electronics Sign Licensing Agreement—Motorola's Land Mobile Products Sector has agreed to license digital technology to Daniels Electronics, Victoria, British Columbia for the development and production of Associated Public-Safety Communications Officials' Project-25 compliant public safety communications products.

The EAS features a digital signaling protocol, which functions as an electronic mailing label to route messages quickly to the targeted emergency area. Radio and TV broadcasters and cable-casters are able to distribute emergency messages in a universal signaling format. This structure enables state and local authorities and national agencies to issue emergency alert messages with a single transmission across the "emergency information superhighway."

RF Micro Devices and TRW Alliance—RF Micro Devices, Greensboro, NC, and TRW Space and Electronics Group, Redondo Beach, CA, have formed a strategic alliance to develop, produce and distribute low-cost, high-volume RF integrated circuits for wireless communications. The agreement grants RF Micro Devices an exclusive world-wide license to the GaAs heterojunction bipolar transistor (HBT) technology for commercial wireless communications below 10 GHz. TRW will double the current capacity at the Redondo Beach HBT chip production facility and will invest in the establishment of a new RF Micro Devices facility to produce these chips in high volume.

Litton's Guidance and Control Systems Division Wins Contract—Litton's Guidance and Control Systems division in Woodland Hills, CA, was awarded a $22.3 million contract by the U.S. Army to produce the AN/PPX-3B IFF interrogator sets for aircraft identification. The lightweight, ground-based, battery-operated units are linked with individual shoulder-fired or vehicle-mounted aircraft weapon systems.

The interrogator set's transmit-receive antenna is mounted on the weapon and is aligned with the weapon sight. A coded signal directed at an aircraft activates a transponder on the aircraft, which returns a signal audible to the weapon's operator on the ground. An aircraft not equipped with a properly coded transponder will cause a different audible signal, indicating a possible foe. Codes are changed frequently.

The interrogator is weatherproof, shockproof, and fungus resistant. A combination programmer and battery charger loads the code and charges the batteries that power the unit. The range of the interrogator is adjusted at the production facility to the range of the customer's weapon.

RF

Miniaturization — the shrinking circuit

By Ernest Worthman

While not new to most of us in the technology sector, the ability to miniaturize today's high-performance circuits has become an expected part of the development cycle. Just around the corner, the next century promises to take miniaturization even further. The process of miniaturization is largely responsible for many technologies being as prolific and as popular with consumers as they are today. The miniaturization process has made it possible for almost everyone to have a computer and cellular telephone. It has brought direct-broadcast satellite systems to the home. It has made electronic engine control practical. It has added special effects, unheard of 20 years ago, to audio and video technology. And it has made everyday appliances such as washers, dryers, stoves and refrigerators—"intelligent."

Of course, miniaturization is not the only factor in this evolution. In a more real sense, miniaturization is the natural progression of the market forces that squeeze product development to become cost-conscious and efficient. Efficiency and competitiveness have forced engineers to seek component integration and new methods of manufacturing that merge many different types of circuits into a single package.

One classic example is the cellular telephone industry. First-generation cellular telephones of the early 1980s had a volume of about 90-in^3. By 1994 that volume had been reduced to about 6-in^3. Much of the miniaturization was driven by the need for low-power consumption and long battery life. Si-bipolar technology was replaced by GaAs-FET technology. Class-C power amplifiers were replaced by Class-F. Operating voltages were reduced from five volts to three volts. And hybrid filters were replaced by monolithic IC filters. Additionally, other devices such as digital signal processing (DSP) chips underwent miniaturization. Very large scale integration (VLSI) technology reduced DSP power consumption from 900 watts in the early 1980s to under 500 mW by the early 1990s.

Another industry where technological development is highly visible is the computer industry. In 1970, when the 4004 series microprocessor was introduced as the world's first microprocessor, it contained a whopping 2,400 transistors. By 1993, the Intel Pentium chip had evolved into a 3.1 million transistor superscalar powerhouse with less than 15 times the original 4004 volume. Next came the P6, with 5.5 million transistors. In the near future (probably sometime in 1997), will come the P7, the next generation Intel chip, which likely will contain upwards of 7 million transistors. Again, much of this miniaturization is due to the evolution of other technologies, such as the use of bi-polar complimentary metal-oxide semiconductor (Bi-CMOS) material and thin-film wafer fabrication technology.

It is hard to find any industry or product that has gone untouched by the age of technology. One only needs to look at the watch that transfers its data to the computer via an infrared communications link. Or the pager that resides inside the form factor of a pen.

We have portable cassette tape recorders that are not much larger than the size of the actual cassette tape, and we can pack an entire record album or 600 plus megabytes of data onto a rugged plastic disk, less than 5 inches in diameter. As for the future? Even those of us on the cutting edge of technology have a hard time prognosticating accurately. One thing is for sure, however. Market forces, technology refinements, global competitiveness, and certainly, some time soon, the development of new technologies will present a myriad of opportunities for the world of miniaturization. Excuse me . . . I hear my wrist watch ringing. RF

Microstrip bandstop filter cuts spurious carrier frequencies

By Denis Jaisson

Where spurious carrier frequencies cause a problem, a bandstop filter can be "thrown" on a microstrip line to cut them by 20 dB or so. Although the filter's performance is not nearly as good as that of a structure with more poles, its narrow shape does not require changing the housing of the system into which it is integrated.

To reject unwanted carrier frequencies in an intermediate frequency (IF) processing unit of a digital communications system (DCS), a bandstop filter with multilayer microstrips was designed to operate around f_c = 1842.5 MHz. It involves a double-layer microstrip resonator coupled to a microstrip line. This coupler was modeled, and the filter's performance was predicted. An experiment with the filter documented its actual performance.

Figure 1 shows a microstrip line with a parallel open stub, which centers a stopband at frequency f_c. In microwave integrated circuits (MICs), where space is restricted, it might be preferable to have a somewhat narrower structure.

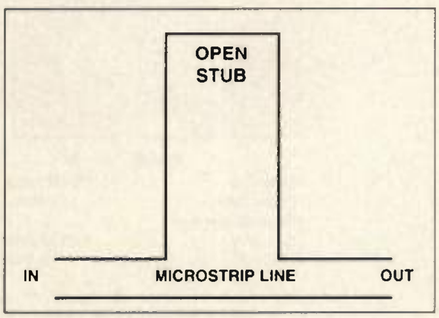

Figure 1. Open stub for bandstop effect.

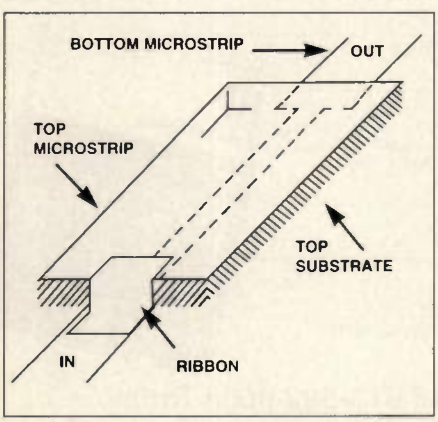

Figure 2. Bandstop filter with multi-layer microstrips.

One possibility is to rotate the open microstrip and to place it on top of the access line with an additional layer of substrate between them. (See Figure 2.) Broadside coupling has been used for tight coupling between microstrips, with a third cover layer for improved directivity [1]. Unlike broadside coupling, directivity is not critical where only two layers are needed. The structure in Figure 2 was used as a bandstop filter. Using Malherbe's model [2] and assuming a homogeneous cross-section in Figure 3 for this coupler, a simplified equivalent circuit was drawn. The behavior of the filter (its rejection bandwidth) was related to the coupler's parameters.

Simple model: homogeneous MCPW

Malherbe's static electrical equivalent circuit for a unit length of a homogeneous transverse electromagnetic (TEM) coupler is shown in Figure 4a. The model for the whole coupler shown in Figure 4b consists of two transmission lines with a quarter-wavelength at f_c and respective characteristic impedances normalized to 50 Ω Z_1 and Z_2 given by:

$$Z_1 = \frac{\sqrt{\varepsilon_r}}{c_o}\frac{C_{12}+C_{22}}{(C_{12})^2}$$
$$Z_2 = \frac{\sqrt{\varepsilon_r}}{c_o}\frac{C_{12}+C_{22}}{C_{11}C_{22}+C_{11}C_{12}+C_{22}C_{12}} \qquad (1)$$

where c_o is the speed of light in free space. The voltage ratio of the ideal transformer is:

$$n = 1+\frac{C_{22}}{C_{12}} \qquad (2)$$

By connecting the left end of the top strip to the bottom strip and by leaving its right end open (See Figure 2), the equivalent circuit in Figure 4b is further simplified to the equivalent circuit in Figure 4c. The parasitic effects brought about by the step in width and by the open end of the top strip have been neglected. At a frequency $f = f_c + \Delta f$ close to f_c, the input admittance of the 50 Ω loaded filter is:

$$Y_{in} \cong \frac{1}{(Z_1)^2} - j\frac{(n-1)^2}{Z_2}\frac{2}{\pi}\frac{f_c}{\Delta f} \qquad (3)$$

Assuming no losses, the transmission factor is given by [3]:

$$|S_{21}| = \sqrt{1-|S_{11}|^2} \qquad (4)$$

where:

$$S_{11} = \frac{1-Y_{in}}{1+Y_{in}} \qquad (5)$$

From equations 3, 4 and 5:

$$|S_{21}| = \pi\frac{Z_2}{(n-1)^2 Z_1}\frac{\Delta f}{f_c} \qquad (6)$$

The transmission factor is at most equal to $|S_{21}|_{max}$ over a bandwidth B

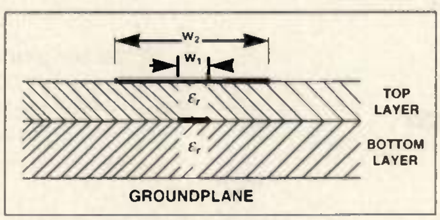

Figure 3. Cross-section of the filter.

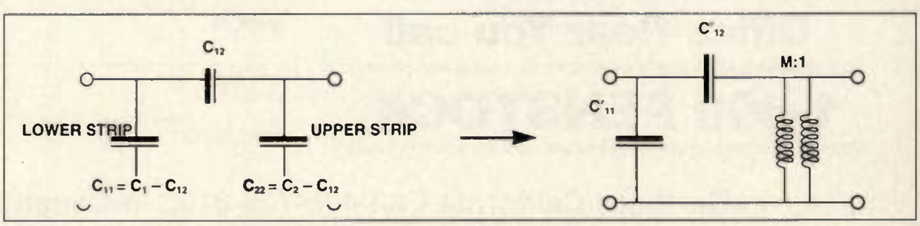

Figure 4a. Homogeneous TEM coupler, unit length equivalent circuit.

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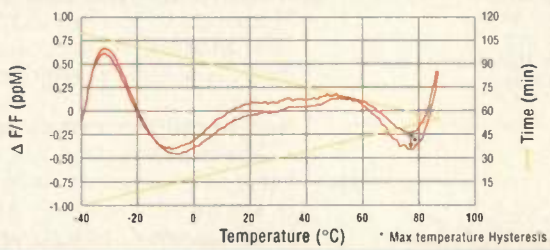
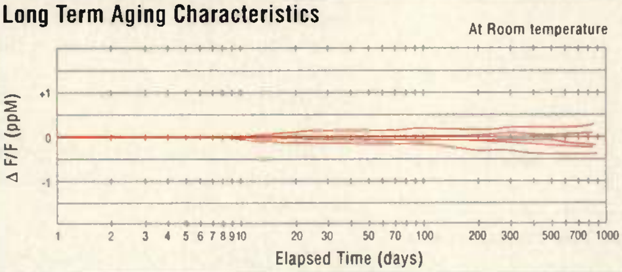
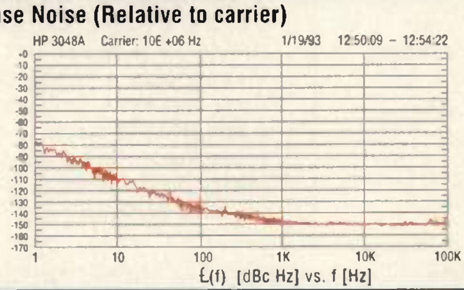


Long Term Aging Characteristics



SSB Phase Noise (Relative to carrier)



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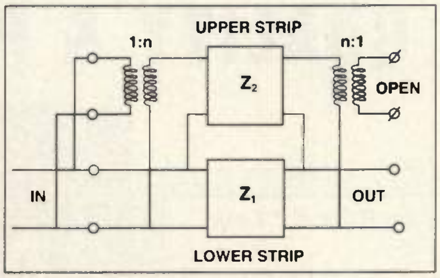

Figure 4b. Homogeneous TEM coupler, equivalent circuit of the coupled lines.

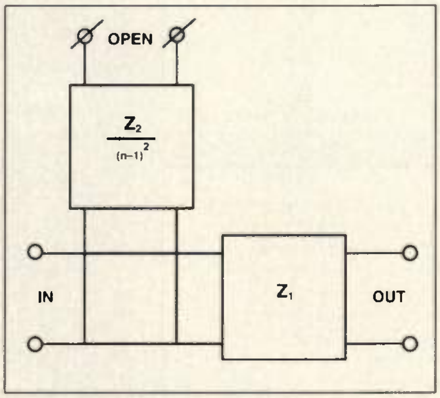

Figure 4c. Homogeneous TEM coupler, equivalent circuit of the filter

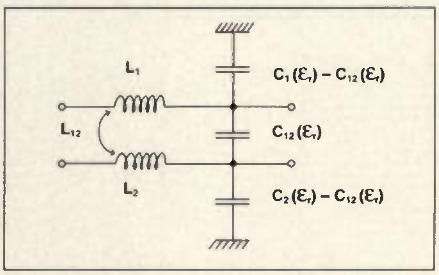

Figure 5. Non-homogeneous coupler.

defined as:

$$B = 2\frac{\Delta f}{f} \tag{7}$$

where $|S_{21}|_{max}$ replaces $|S_{21}|$ ($|S_{21}|$ in Equation 6 is frequency dependent, whereas $|S_{21}|_{max}$ is a constant arbitrarily chosen by the designer).

From equations 1, 2 and 6:

$$B = \frac{2}{\pi}\frac{(C_{22})^2}{C_{11}C_{22}+C_{11}C_{12}+C_{22}C_{12}}|S_{21}| \tag{8}$$

Referring to Equation 8, it appears that, for a given $|S_{21}|_{max}$, where $|S_{21}|_{max}$ replaces $|S_{21}|$ ($|S_{21}|$ in Equation 8 is frequency-dependent, whereas $|S_{21}|_{max}$ is a constant arbitrarily chosen by the designer), B is larger if C_{11} and C_{21} are small, and C_{22} is large. Consequently, the top strip was widened, and the bottom strip was narrowed where they are coupled to each other. (See Figure 2.)

The model in Figure 4c helps to relate qualitatively B to the parameters of the coupler, but it is valid only if the coupler's cross-section is homogeneous, which is not the case in practice. Moreover, it does not account for discontinuities. A more sophisticated model is, therefore, required.

Non-homogeneous coupler

The cross-section CS of the non-homogeneous coupler used in Figure 2 is shown in Figure 3, and its unit-length, quasi-static equivalent circuit is given in Figure 5 [4]. Capacitances are computed in terms of cross-section geometry, substrate permitivity ε_r and conductor potentials, using a finite element method [5]:

1) $C_i(\varepsilon_r)$ is computed for:

$$V_i = 1 \tag{9}$$

(All other conductors are grounded.)

2) Setting $V_1 = V_2 = 1$ yields a capacitance C_3, so that:

$$C_{12}(\varepsilon_r) = \frac{1}{2}\left(C_1(\varepsilon_r) + C_2(\varepsilon_r) - C_3\right) \tag{10}$$

The symmetry of CS about plane $x \equiv 0$ is taken advantage of by treating this plane as a magnetic wall and by making only a half of CS into discrete finite elements. Inductances in Figure 5 are given in terms of capacitances by removing the substrate ($\varepsilon_r = 1$). The following formulas have been derived in a previous paper [6]:

$$\frac{1}{L_{12}} = c_o^2 C_{12}(1)\left(\frac{C_1(1)C_2(1)}{C_{12}^2(1)} - 1\right) \tag{11}$$

and:

$$L_1 = L_{12}\frac{C_2(1)}{C_{12}(1)},\quad L_2 = L_{12}\frac{C_1(1)}{C_{12}(1)} \tag{12}$$

$C_1(1)$, $C_2(1)$ and $C_{12}(1)$ are obtained in the same way as $C_1(\varepsilon_r)$, $C_2(\varepsilon_r)$ and $C_{12}(\varepsilon_r)$ for $\varepsilon_r = 1$.

Once the unit length capacitances and inductances are known, the coupler is modeled into a network of 10 of the cells in Figure 5 connected in series, where all values are multiplied by one-tenth of the total coupled length l. A short FORTRAN program was written for the coupler which:

1) computes capacitances under the voltage conditions in Equations 9 and 10.

2) computes inductances according to Equations 11 and 12.

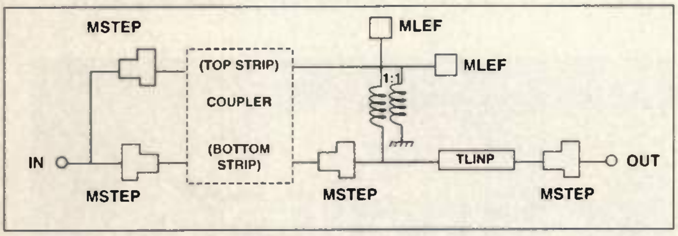

Figure 6. Touchstone network for the filter.

3) automatically updates a Touchstone circuit file for length l to be optimized at f_c.

The Touchstone network for the filter is shown in Figure 6. The resonator open end is modeled as two open-end microstrip line end effects (MLEFs). The one MLEF accounts for the capacitance to the groundplane, with width w_2–w′ (top view in Figure 7). The other, with width w′, accounts for the capacitance to the dielectric-covered, 50 Ω output line. The latter MLEF is connected through an ideal transformer because the MLEF model of Touchstone is referenced to ground.

Experiment

The bandstop filter in Figure 2 was designed for f_c = 1842.5 MHz on an FR4 substrate, with relative permitivity ε_r = 4.7 and substrate height h = 1.165 mm. Conductor thickness t = 18 μm was neglected. Width w' = 1.57 mm of the dielectric-covered 50Ω output line was obtained for an air-filled microstrip line with impedance [7]:

$$\sqrt{\varepsilon_r}\,Z_0 = \frac{1}{c_0 C(1)} = 108\Omega \qquad (13)$$

where C(1) is the unit length capacitance of the air-filled line. A B = 6.6% 20 dB rejection bandwidth was predicted around f_c for w = 5 mm, g = 0.5 mm and l = 19.44 mm. The following values were obtained for the equivalent circuit in Figure 5:

$C_1(\varepsilon_r)$ = 0.192 pF/mm	L_1 = 0.274 nH/mm
$C_2(\varepsilon_r)$ = 0.116 pF/mm	L_2 = 0.546 nH/mm
$C_m(\varepsilon_r)$ = 0.0681 pF/mm	L_m = 0.161 nH/mm

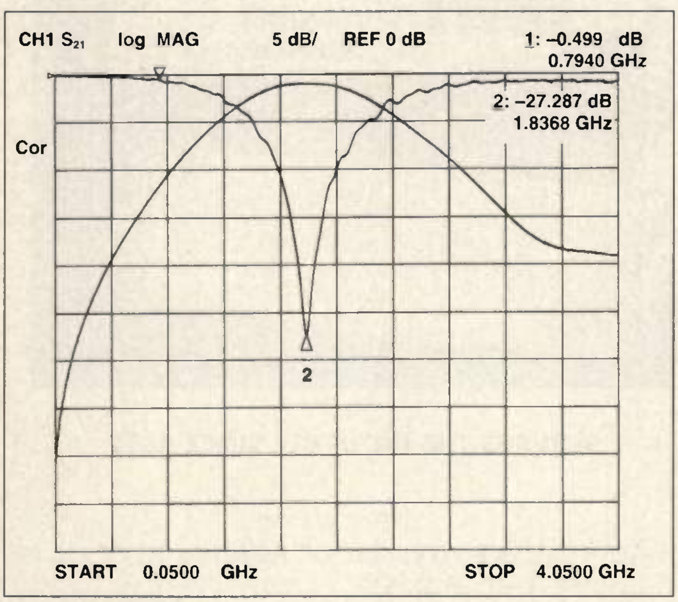

Figure 8. Filter measurements.

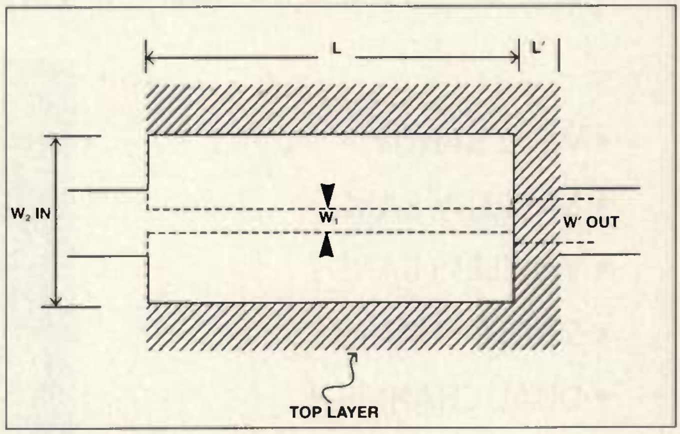

Figure 7. Top view of the filter.

Figure 8 shows measured $|S_{21}|$ and $|S_{11}|$ ($|S_{22}| \approx |S_{11}|$ by a half decibel up to f_c). A minimum $|S_{21}|$ = –27 dB was measured at f_m = 1836.8 MHz, that is 0.3% off design f_c. Rejection is more than 20 dB greater than B = 6.3%. Transmission loss is less than a half-decibel up to 794 MHz, where $|S_{11}|$, $|S_{22}|$ = –11 dB. $|S_{11}|$, $|S_{22}|$ < –20 dB for f < 410 MHz.

Permitivity ε_r of FR4, as this material is commonly manufactured, is known to suffer a larger spread than that of substrates that were first introduced for microwave applications such as filters, where low tolerance on ε_r is important. Fortunately, a good agreement was observed between simulated and measured values of f_c. Full-wave modeling of the filter with a 3D electromagnetic simulator, combined with an accurate measurement of ε_r, would give a better idea about the accuracy of the model in Figure 6.

Conclusion

For use when spurious carrier frequencies are a problem, a bandstop filter can easily be "thrown" on a microstrip line to knock them off by 20 dB or so. Although its performance does not approach that of structures with more poles, its narrow shape does not require any change in the housing of the system into which it is integrated.

RF

Acknowledgment

This project was funded by the Cost Reduction Program at Ericsson Radio Systems AB in Kista, Sweden.

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7. D. Jaisson, "Design Microstrip Components with a Dielectric Cover," *Microwaves & RF*, Vol. 32, No. 2, February 1993, pp. 71–80.

8. Touchstone is a trademark of EESof.

About the author

Denis Jaisson has worked as a microwave engineer since graduating from the Joint European Scheme in Electrical Engineering in 1985. While in industry, he carried out independent thesis work on coplanar structures and was awarded a Ph.D. from I. N. S. A. Rennes in January 1996. He has been involved in component and system design for DCS and DECT, antenna arrays, radar transponders, electronic countermeasures and test equipment including spectrum analyzers. He is now with T. R. T. (Lucent Technologies) in France. His main interest lies in the modeling and design of planar structures.

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INFO/CARD 32

Linear simulators offer successful microstrip modeling for Wilkinson power-splitters

By Sean R. Mercer, Ph.D.

Using a simple model compatible with most commercial linear simulators yields effective first-time designs. The technique avoids difficult implementations commonly shown in textbooks. Layout, resistor and fabrication technology choices make an important difference.

Successful Wilkinson power-splitter design results have been achieved for a two-way power-divider and combiner. The model for those microstrip circuits can be used with most commercial linear simulators, such as HP-EEsof's Touchstone and Compact Software's Supercompact. This model has resulted in first-time design success at frequencies up to and including X-band.

The design of a hybrid power-splitter was presented by Wilkinson [1] in 1960, and a functional description has appeared in many subsequent texts [2, 3, 4, 5, 6]. Many textbooks, unfortunately, show a physical implementation that is difficult to model with the library components available with most linear simulators [2, 4, 6].

A common textbook representation of the reciprocal, passive, Wilkinson power-splitter and combiner is shown in Figure 1. The 50 Ω terminating impedances at ports 2 and 3 each are transformed to 100 Ω impedances at node A. These two 100 Ω impedances appear in parallel at node A so that port 1 has a 50 Ω input impedance. Signals incident at port 1 are split equally between ports 2 and 3 so that half of the incident power appears at each of these ports.

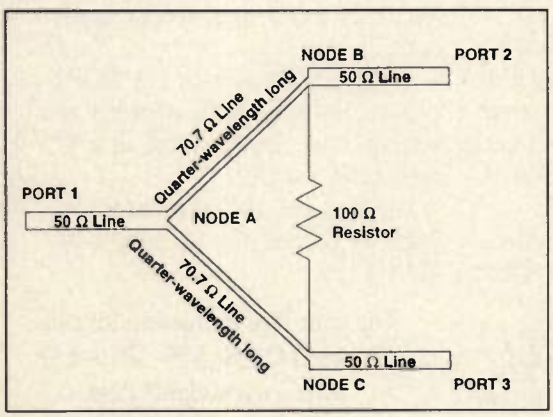


Figure 1. Wilkinson power splitter.

The 100 Ω resistor is included to isolate the two output ports. Any out-of-phase signals reflected back into the power splitter will be out-of-phase across the 100 Ω resistor and, therefore, will be dissipated in the resistor. Reflected signals that are in-phase across the resistor will, on the other hand, be passed back to the input port (port 1).

Layout Considerations

Electromagnetic simulators or finite difference techniques can be used to model complex and compact geometries accurately, but these tools are not available to all designers. The layout used by Gedney [7] is compact, but it employs junctions between the 50 and 70.7 Ω lines that cannot be modeled accurately using linear simulators such as Touchstone or Supercompact. Moreover, there will be significant coupling between the two closely spaced, parallel, quarter-wavelength transformer sections. To achieve good correlation between modeling with a linear simulator and actual circuit performance, the design should be restricted to use only elements that accurately can be modeled by the available software.

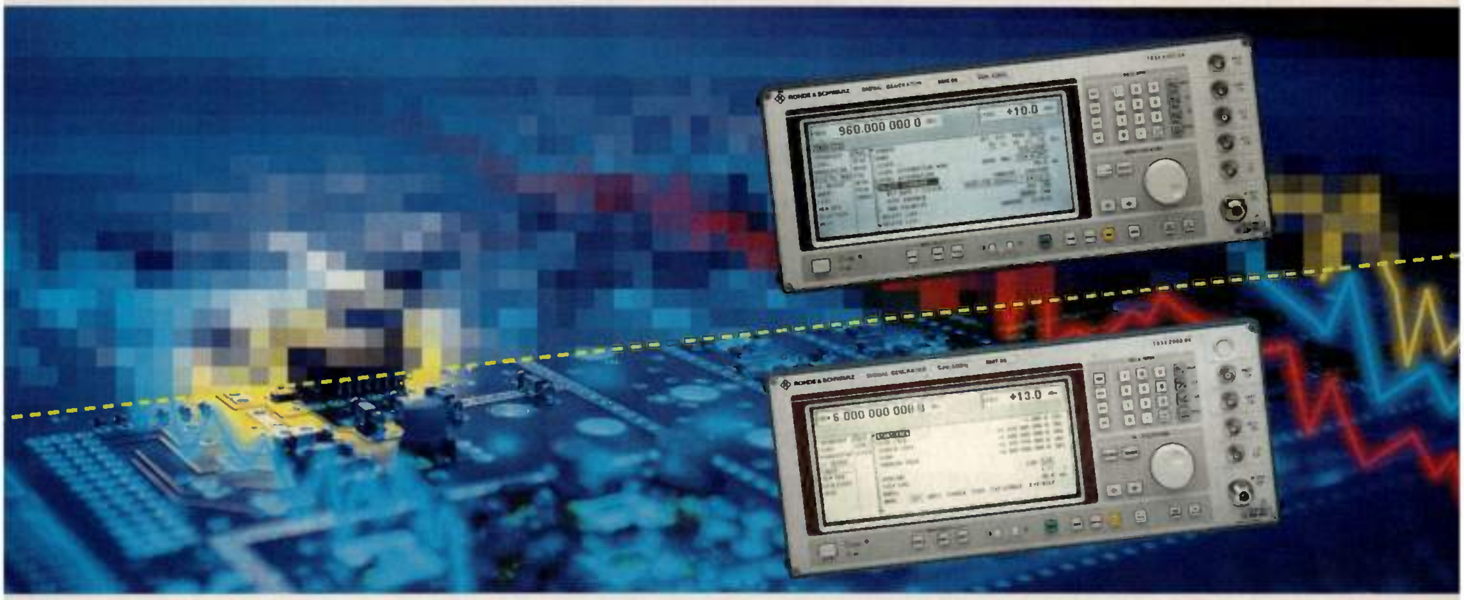
Furthermore, the layout should be arranged carefully to minimize unwanted coupling or interactions that do not appear in the simulation. To avoid significant undesirable coupling between two microstrip lines, a good rule of thumb is to maintain a distance of at least twice the substrate thickness between any two lines where coupling is unintended. No line should be closer than twice the substrate thickness to the edge of the substrate. This precautionary step will minimize changes in characteristic line impedance caused by fringing fields coupling to a ground plane. In addition, a reliable and repeatable fabrication process is essential to produce accurate microstrip circuits.

Resistor choice

The implementation of the 100 Ω resistor is crucial to the success of the power-splitter and combiner. Many past designs have provided no specific attachment points for the resistor. Although it is not impossible to achieve good performance with this technique, repeatable performance and correlation with the simulation can be harder to achieve if resistor placement can vary for lack of a precisely defined attachment point. This problem is easily overcome by providing resistor attachment pads and by modeling the effects of the discontinuities presented by these pads.

The reasoning assumes the use of an ideal resistor with no electrical length and ignores the resistor's physical size and power rating. At higher frequencies, the resistor's physical dimensions are important because they become a significant fraction of a wavelength. Higher-power resistors have a larger surface area and, therefore, they exhibit greater capacitance to ground than small, low-power resistors.

The optimum choice for the resistor is usually one that is as small as possible and that still meets the power-handling requirements of the design. For higher-power applications, where the resistor is of the type that is bolted to a heatsink, it can be beneficial to compute a set of s-parameters for the resistor and use those for the design. The use of this type of resistor in a Wilkinson power-splitter is rare at frequencies above 2.0 GHz because of their significant shunt capacitance and



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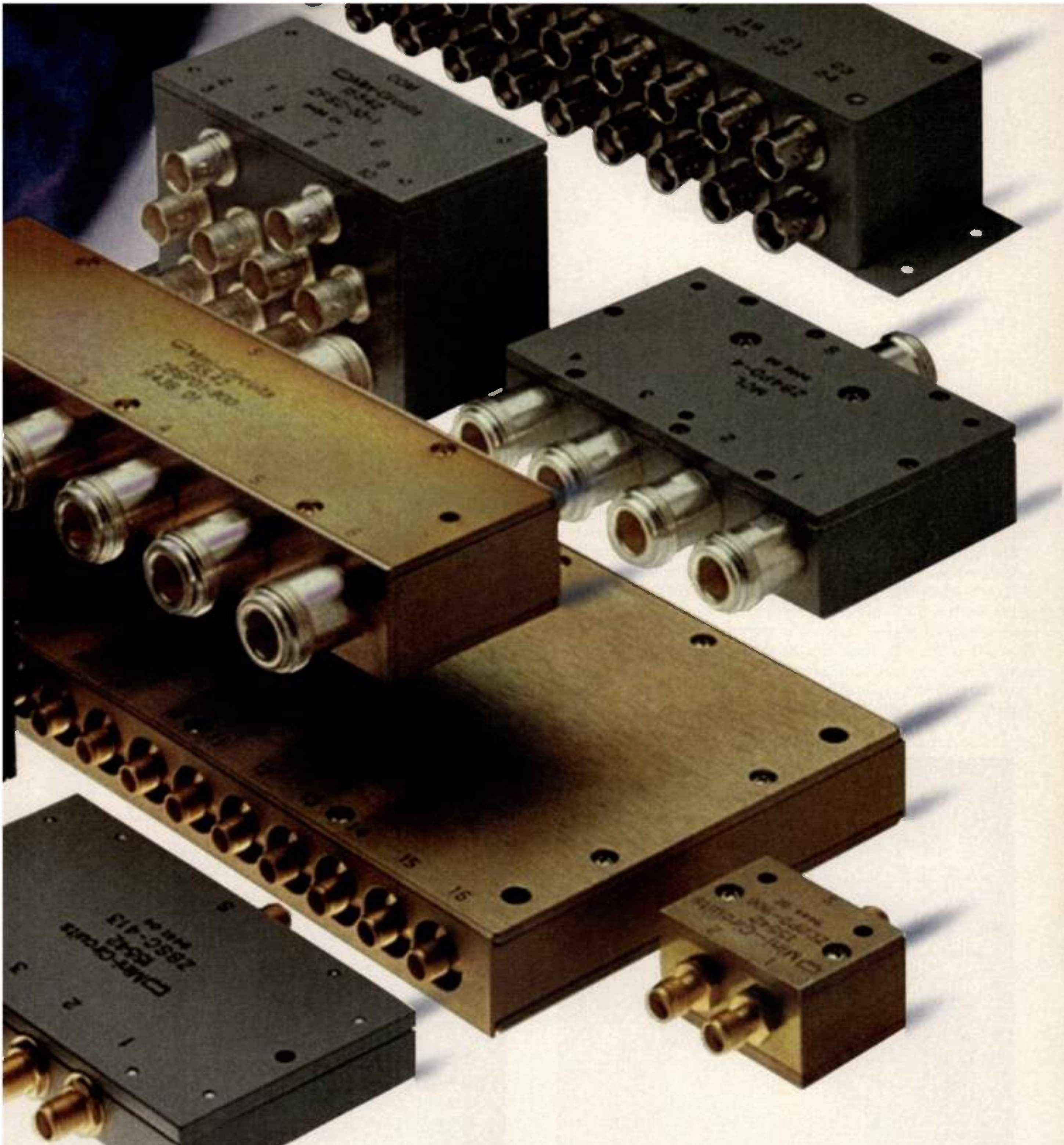

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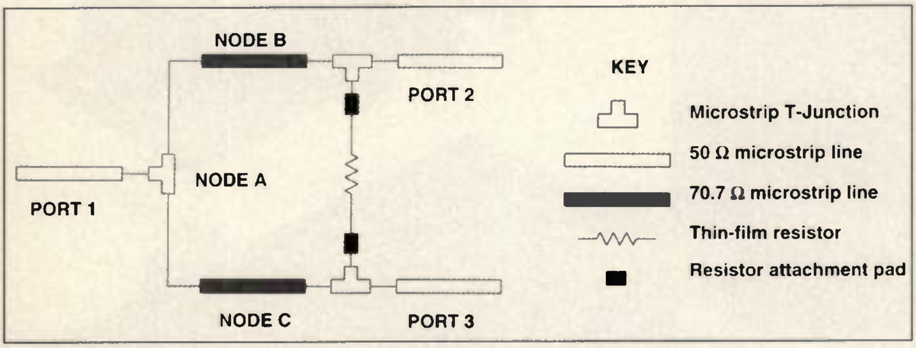


Figure 2. Simplified model for Wilkinson power-splitter.

relatively long electrical length at higher frequencies.

Fabrication technology

The choice of fabrication technology influences the design strategy. The use of 99.6% Al_2O_3 (Alumina) ceramic substrate with a 50 Ω-per-square resistive layer is probably the simplest to model. The 100 Ω resistor then can be implemented as a rectangular area of resistive material (typically Tantalum Nitride) with a length equal to twice the width. Resistor power-handling can be increased by increasing the resistor dimensions. This type of resistor can be modeled successfully using the thin-film resistor model in the linear simulator (the TFR model in Touchstone and Supercompact).

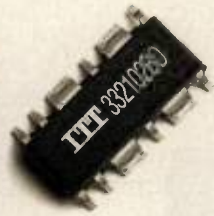
A model of a Wilkinson power-splitter for use in a linear simulator is shown in Figure 2. All of the elements in this model are available with most linear simulators. A 50 Ω microstrip line connects port 1 to a microstrip T-junction at node A. The two 70.7 Ω quarter-wavelength impedance transformers are connected to this T-junction. The opposite ends of these 70.7 Ω lines are connected to microstrip T-junctions at nodes B and C. The 50 Ω output lines are connected to these T-junctions, too, with the third leg of these T-junctions set to the width of the resistor termination pad. A short microstrip line of width equal to the width of the chosen resistor is connected to the third leg of these last T-junctions to form resistor attachment pads. A thin-film resistor with a width equal to the resistor attachment pads is connected between these two attachment pads.

Note that the models for a microstrip T-junction in Touchstone and Supercompact use different reference planes, so take this into account when trying to compare simulations of the same physical circuit with these software packages.

Curved vs. straight lines

The model shown in Figure 2 shows the 70.7 Ω quarter-wavelength imped-

Street Smarts.



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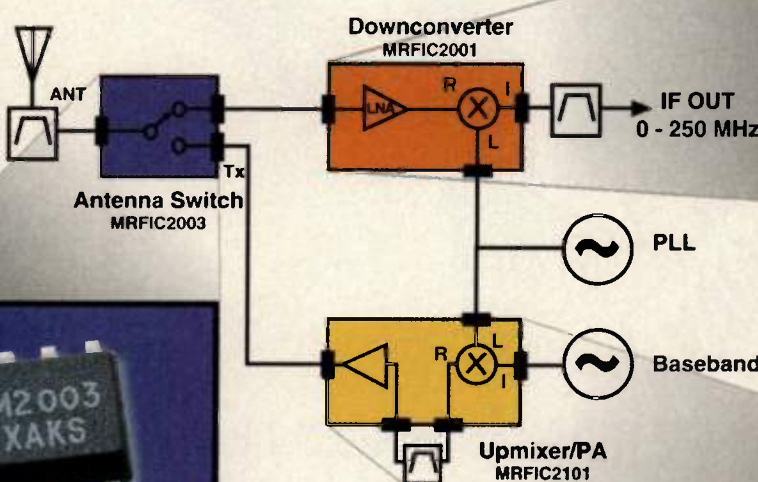
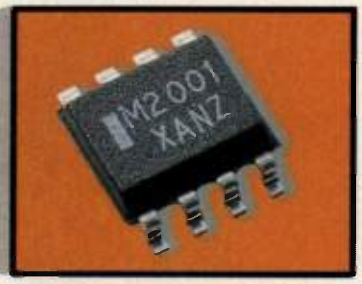


A Typical **900 MHz ISM** Block Diagram



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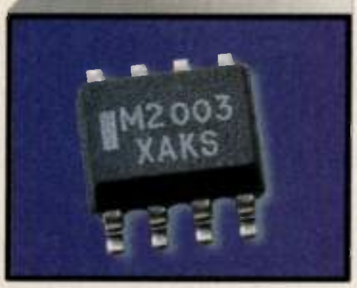

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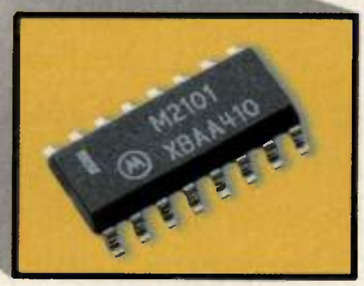

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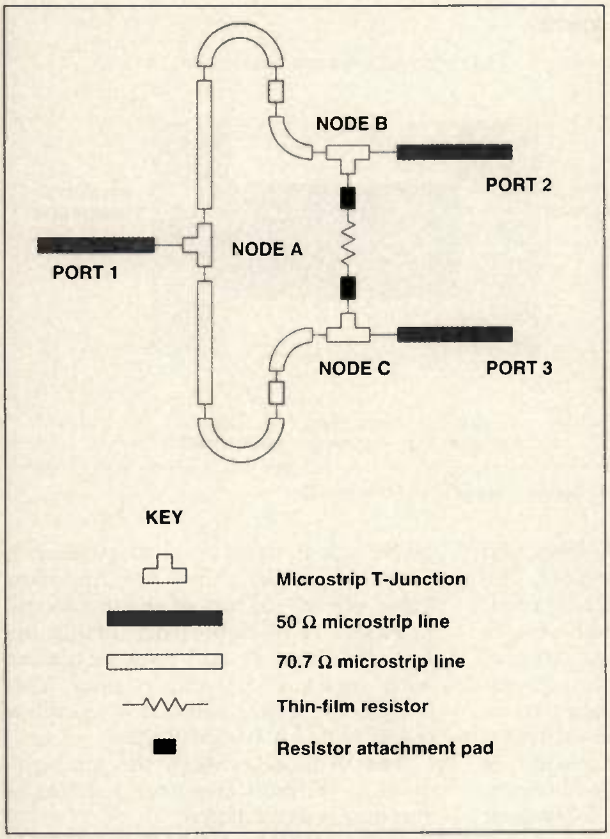

Figure 3. Complete model for Wilkinson power splitter.

ance transformers as single-line elements. In practice, though, each of these lines is made up of four sections as shown in Figure 3. The first 70.7 Ω line segments are orthogonal to the input T-junction and are 180° apart from each other. They need to be curved back together to allow the attachment of a physically short resistor between the resistor attachment pads. This can be achieved easily by using two straight-line sections and two curved sections as in Figure 3. The distance between the parallel straight sections is maximized (at least twice the substrate thickness apart) to minimize unwanted coupling between these traces. The use of curved microstrip lines makes this layout more compact than the layout shown in Figure 2. Moreover, the center frequency for maximum coupling between the shorter parallel 70.7 Ω line sections is much higher than for a layout where coupling can occur along the entire length of the 70.7 Ω lines; therefore, the use of the curved traces in the layout helps to reduce undesirable in-band coupling between the lines.

The curved traces do, nevertheless, add an extra complication because they appear to be electrically shorter than a straight conductor with the same physical length. This means that there will be poor correlation between simulation and practical results for a circuit modeled as shown in Figure 2, but implemented as shown in Figure 3. The modeling of curved traces has been documented in literature [8, 9], and some linear simulators do include models for curved traces. Use these models if they are available, and optimize the lengths of the straight sections of 70.7 Ω line to compensate for the contractions in electrical length in the curved traces. Tightly curved lines have a shorter effective length than gently curved lines with a greater bend radius.

If your linear simulator does not include a model for curved microstrip line, the 70.7 Ω lines should be modeled as in Figure 2. The layout still should be arranged as shown in Figure 3, but the total length of the 70.7 Ω lines has to be altered to compensate for the effects of the curved lines. Some formulas have been derived that account for the effects of curved lines [10], and using a ratio of arc radius-to-microstrip line width of five is advised to avoid the use of tightly curved lines. In this case, the effective arc radius is approximately equal to the actual arc radius reduced by 20% of the width of the curved trace. The difference between the actual arc lengths and the effective arc lengths should be added to the straight lengths of 70.0 Ω line to compensate for the effective length contraction caused by line curvature.

As an example, a curved, 0.23 mm-wide microstrip line with a ratio of arc radius to line width of five has a radius of 1.15 mm to the center of the line. The effective radius for this curved line will be 1.104 mm. This radius is used to calculate the effective length of the curved line. The effective length of a semicircular arc for this line would be 0.145 mm shorter than its actual physical length. Half of the difference of 0.145 mm must be added to each of the straight 70.7 Ω line sections to compensate for the effective reduction in line length of this curved line. Remember to apply this compensation to both curved lines in each 70.7 Ω arm of the layout shown in Figure 3.

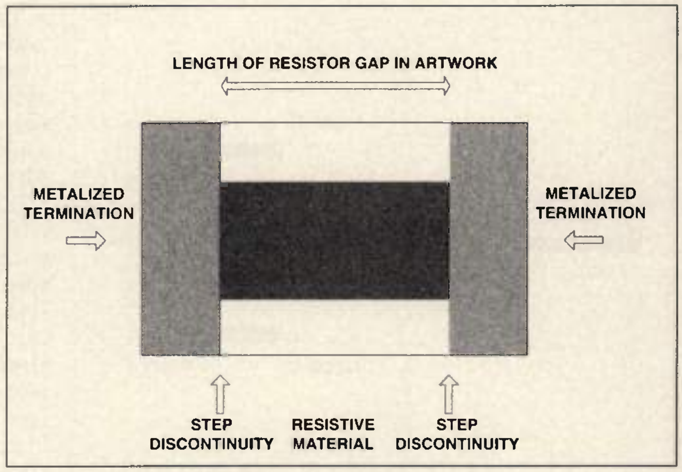

Figure 4. Bottom view of 100 Ω chip resistor.

Substrate

It is possible to use an Alumina substrate without a resistive layer and attach a high-quality chip resistor. The thin-film resistor model can be used, but some additional discontinuities must be considered. An examination of a chip resistor under a microscope reveals a construction similar to that in Figure 4. Solder terminations at either end of the resistor are the full width of the body of the resistor. A thinner, lossy transmission line runs between these terminations. In most cases, this line is approximately twice as long as its width for a 100 Ω resistor, indicating the use of 50 Ω, square-shaped resistive material.

Resistor attachment pads are provided for the metalized ends of the resistor as shown in Figure 2. The attachment pads do not need to be any longer than the metalized ends of the chip resistor. The length of the lossy transmission line on the chip resistor is used to define the length of the gap in the artwork between the resistor attachment pads. The result is step discontinuities between the resistor attachment pads and the lossy transmission line. The thin-film resistor model still is used to model the resistive material, but step discontinuities (MSTEP in Touchstone or STEP in Supercompact) are included at either end of the resistor to model the discontinuity in width between the attachment pads and the resistive material.

The same technique can be applied to soft substrates. At lower frequencies (below 3 GHz), it is possible to use a simple 0805-package 100 Ω chip resistor with a 125 mW power rating for low-cost applications. At these frequencies, it is possible to model the resistor as though it were an ideal resistor in series with an ideal inductor. An inductance of 0.15 nH/mm of resistor length provides reasonable correlation between simulation and implementation when using a 0805-chip resistor. This simplistic model fails at higher frequencies and is best avoided.

The design strategy for a single-section Wilkinson power splitter is summarized as follows:

1. Choose the substrate material and the 100 Ω isolation resistor. The physical dimensions of the resistor must be known to determine the correct gap length to be left in the artwork for the resistor.

2. Determine the line widths for 50 and 70.7 Ω lines on the chosen substrate. Determine the length of a quarter-wavelength 70.7 Ω line at the center frequency of operation. Choose the dimensions for the resistor attachment pads equal to the width of the resistor.

3. Implement the model shown in Figure 3 in your linear simulator. If a chip resistor is to be used, remember to include the step discontinuities between the resistor attachment pads and the resistive material as shown in Figure 4.

4. Optimize the length of the 70.7 Ω lines to compensate for the effects of the resistor, curved lines and junction discontinuities. If models for curved microstrip lines are not available, the power-splitter should be simulated as shown in Figure 2. The curved line lengths for the artwork represented in Figure 3 then are calculated as previously described.

5. Lay out the design to minimize

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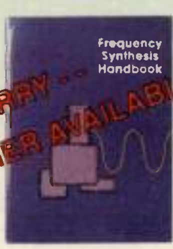

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unwanted couplings and substrate edge effects.

Conclusion

This model has been used successfully to construct single-section power-splitters at frequencies as high as X-band. It should be a simple matter to expand the method to design broadband, multisection power-splitters in circuits with more resistors and, consequently, more discontinuities that can lead to poor circuit performance when they are modeled incorrectly. *RF*

Acknowledgment

The author thanks F. Traut for useful discussions.

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About the author

Sean Mercer is a principal engineer at Racal Canada where he is involved in the design and development of HF transceivers. He received a Ph.D. from the University of Cape Town in 1990. He has co-authored several patents, and he is a member of the Institute of Electrical Engineers and the Institute of Electrical and Electronic Engineers. He can be contacted via e-mail at mercer@racalcanada.com or at Racal Canada, 1495 Franklin, British Columbia, V5L 5B6, Canada.

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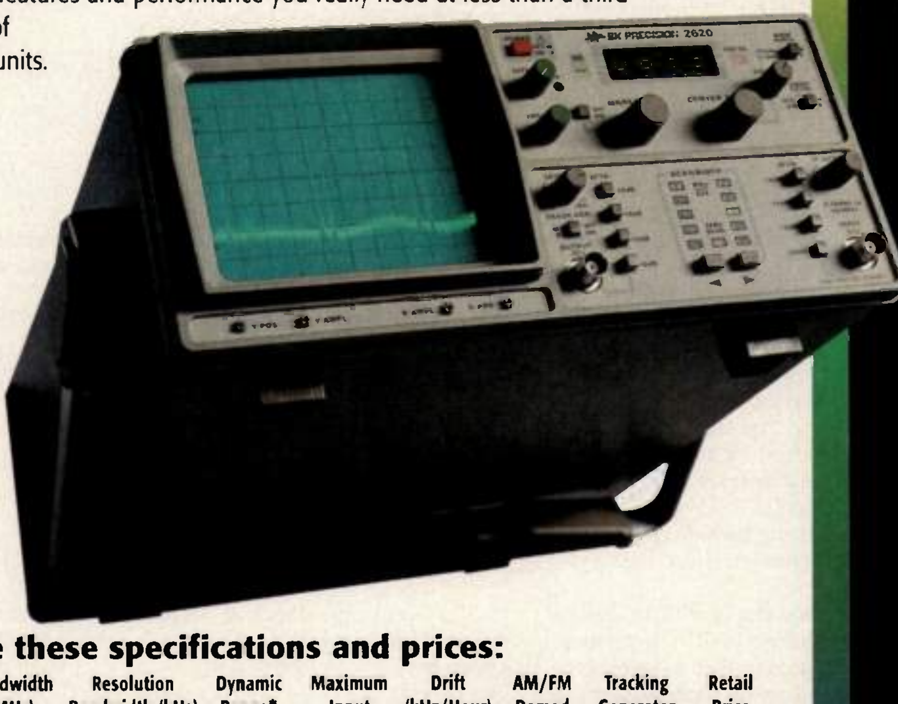

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Use matrix models to make analysis easy for microstrip matching circuits

By R. Partha

Despite the availability of many computer-aided design (CAD) tools, many RF product designers still prefer the old-fashioned Smith chart for starting a design. Matrix models offer a way to represent four common microstrip networks with more sophistication than Smith chart methods, yet with easy computability.

Microstrip matching circuits traditionally have been designed using the normalized impedance-and-admittance Smith chart. Although the simple Smith chart method is sufficient for checking a result at a single frequency, it is inadequate for an analysis over a range of frequencies. Such analysis is provided by powerful microwave computer-aided engineering (CAE) packages that offer accuracy and versatility. Unfortunately, these software packages are expensive, and they call for a dedicated investment in a laboratory's infrastructure. A straightforward evaluation technique offers greater sophistication than operations on a Smith chart and allows easy computability.

The following information describes common impedance-matching configurations in terms of cascade connections expressed in matrix form. These descriptions will enable RF or microwave designers to develop their own analysis programs. Such programs, based on matrix formalism specific to a matching circuit, need not be elaborate and may be confined to the task at hand. Matrix representation for microstrip networks involving the following elements are described:

1. Series microstrip transmission line.
2. Shunt-connected, open-circuited transmission line.
3. Shunt-connected, short-circuited transmission line.
4. Quarterwave transformer.

I am assuming that you already have the ability to design a microwave circuit to perform a particular task. The focus is on analyzing and optimizing any particular microstrip-matching circuit. The need arises when ready-pattern layouts with dimension details are obtained from microwave transistor catalogs for amplifier or oscillator applications. It becomes necessary to derive the matching load or source impedance point (R ± jX) from the layout dimensions and, possibly, to evaluate the response across a frequency band. It may be necessary to cross-check a matching network developed with graphical aids such as the Smith chart or with any cut-and-try approach.

While bench-testing a particular design, it is often normal practice to add tuning stubs or to shorten stub lengths for optimum performance. The circuit then requires reappraising with the modifications in place. Analytical computations using the matrix tool come in handy in all of these situations. Ultimately, an understanding of the design enhances an RF engineer's efficiency, and the result is reflected in the hardware performance.

Transmission line representation

The ABCD matrix for a transmission line of length l, characteristic impedance Z_o, shown schematically in Table 1, is well-known and is expressed as:

$$\begin{bmatrix} \cosh(\gamma_l) & Z_o \sinh(\gamma_l) \\ \dfrac{\sinh(\gamma_l)}{Z_o} & \cosh(\gamma_l) \end{bmatrix}$$

where $\gamma = \alpha + j\beta$, γ is the propagation constant, α the attenuation constant, and β the phase constant per unit length.

If the transmission line is assumed to be lossless, i.e., $\alpha = 0$, then the matrix is written as:

$$\begin{bmatrix} \cos(\beta_l) & jZ_o \sin(\beta_l) \\ \dfrac{j\sin(\beta_l)}{Z_o} & \cos(\beta_l) \end{bmatrix}$$

The input impedance Z_{in}, expressed in terms of ABCD parameters for any two-port network terminated in a load Z_L is given by:

$$Z_{in} = \frac{AZ_L + B}{CZ_L + D} \tag{1}$$

From Equation 1, the input impedance of the transmission line of length l terminated in a short circuit ($Z_L = 0$) is:

$$Z_{in} = \frac{B}{D} = jZ_o \tan(\beta_l) \tag{2}$$

When the transmission line is left open-circuited ($Z_L = \infty$),

$$Z_{in} = \frac{A + \dfrac{B}{Z_L}}{C + \dfrac{D}{Z_L}} = \frac{A}{C} = -jZ_o \cot(\beta_l)$$

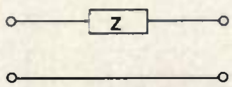
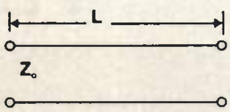
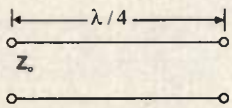
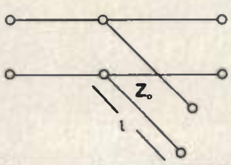
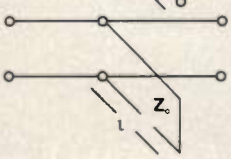
 SERIES IMPEDANCE	$\begin{bmatrix} 1 & Z \\ 0 & 1 \end{bmatrix}$	
 SHUNT ADMITTANCE	$\begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix}$	
 $\gamma = \alpha + j\beta$ TRANSMISSION LINE SEGMENT	$\begin{bmatrix} \cosh(\gamma l) & Z_o \sinh(\gamma l) \\ \frac{\sinh(\gamma l)}{Z_o} & \cosh(\gamma l) \end{bmatrix}$ $\gamma = \alpha + j\beta$	$\cong \begin{bmatrix} \cos(\beta L) & Z_o \sin(\beta L) \\ \frac{j\sin(\beta L)}{Z_o} & \cos(\beta L) \end{bmatrix}$ $\alpha = 0$
 λ / 4 line	$\begin{bmatrix} 0 & jZ_o \\ \frac{j}{Z_o} & 0 \end{bmatrix}$ $l = \frac{\lambda}{4}$	
 OPEN-ENDED SHUNT-CONNECTED STUB	$\begin{bmatrix} 1 & 0 \\ \frac{j\tan(\beta l)}{Z_o} & 1 \end{bmatrix}$	
 SHORTED SHUNT-CONNECTED STUB	$\begin{bmatrix} 1 & 0 \\ \frac{1}{jZ_o \tan(\beta l)} & 1 \end{bmatrix}$	

Table 1.

The chain matrix representation of the transmission line can be extended to the microstrip line because the mode of propagation on the microstrip is almost transverse electromagnetic (TEM). For the limited purpose of analyzing commonly used matching configurations (description of series and shunt elements), open and short stubs as listed in Table 1 are considered to be sufficient.

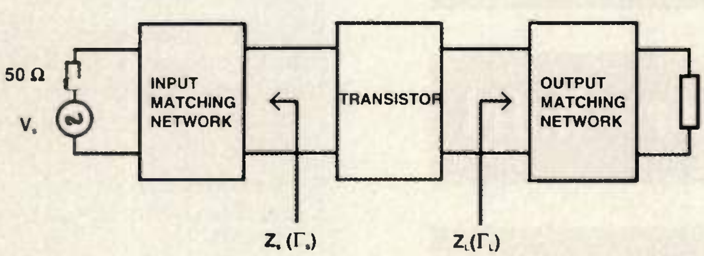


Figure 1. Block diagram of RF amplifier.

Applications

Supplying maximum power to the load in an amplifier is important. It calls for the design of optimum matching networks at the input and output of the active device (Figure 1). Load networks that cause instability are necessary to make devices oscillate, which in turn requires the load point to be located in the unstable region of the relevant stability circle. In all of these applications, combinations of microstrip line lengths, open or short stubs and single- or multiple-section impedance transformers are used to construct the networks. Some

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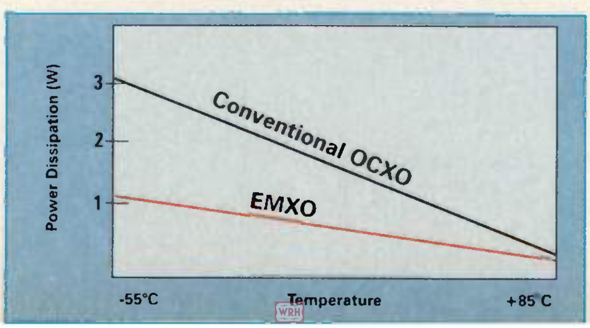

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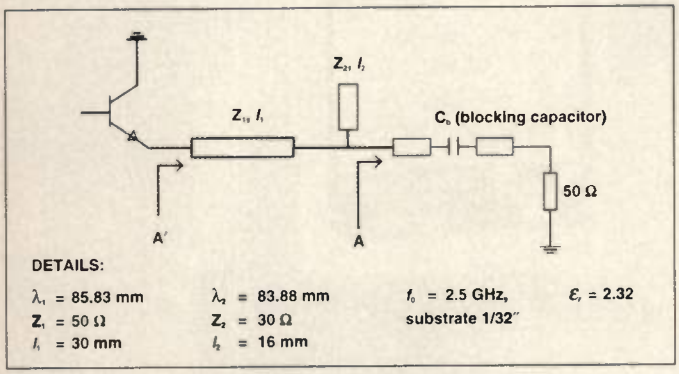

Figure 2. Matching with series line and shunt open stub.

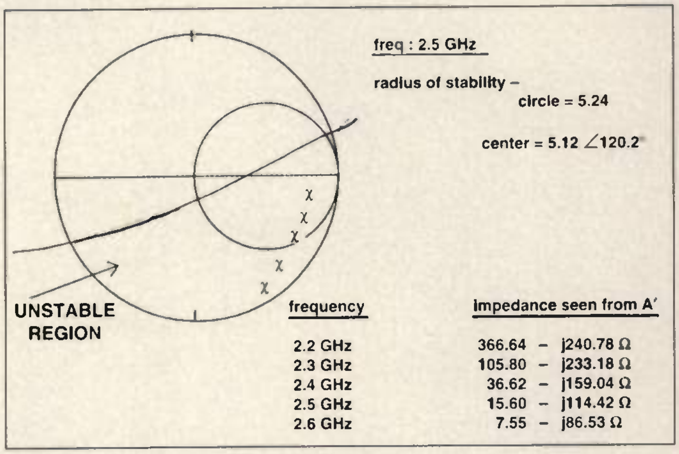

Figure 3. Impedance trace of matching circuit.

examples of matching circuits analyzed with the aid of relatively easy computer calculations are based on matrix models.

• *Example 1: series line with shunt open stub* — Consider a bipolar transistor that is potentially unstable at 2.5 GHz. Its output stability circle between the emitter and collector is shown in Figure 3. A load circuit (corresponding to Γ_L) located in the unstable region of the stability circle would ensure a negative resistance at the input port. The layout in Figure 2 has been designed graphically and is taken as a good starting point for verification and further optimization. Begin by calculating the impedance the circuit presents to the emitter.

At any plane from A toward the 50 Ω termination R, the impedance seen is 50 Ω itself. This length of line is not important to the analysis. The impedance seen from A′ is to be determined. The relevant microstrip configuration consists of a series transmission line cascaded with a shunt open stub. The overall matrix for this combination is derived from Table 1.

$$\underbrace{\begin{bmatrix} \cos(\beta_1 l_1) & jZ_1\sin(\beta_1 l_1) \\ \dfrac{j\sin(\beta_1 l_1)}{Z_1} & \cos(\beta_1 l_1) \end{bmatrix}}_{\text{series microstrip line}} \underbrace{\begin{bmatrix} 1 & 0 \\ \dfrac{j\tan(\beta_2 l_2)}{Z_2} & 1 \end{bmatrix}}_{\text{shunt open stub}} =$$

$$\begin{bmatrix} \cos(\beta_1 l_1) - \dfrac{Z_1\sin(\beta_1 l_1)\tan(\beta_2 l_2)}{Z_2} & jZ_1\sin(\beta_1 l_1) \\ \dfrac{j\sin(\beta_1 l_1)}{Z_1} + \dfrac{j\tan(\beta_2 l_2)\tan(\beta_1 l_1)}{Z_2} & \cos(\beta_1 l_1) \end{bmatrix}$$

$$Z_{A'} = \frac{AR + B}{CR + D}$$

$$A = \cos(\beta_1 l_1) - \frac{Z_1\sin(\beta_1 l_1)\tan(\beta_2 l_2)}{Z_2}$$

$$B = jZ_1\sin(\beta_1 l_1)$$

$$C = \frac{j\sin(\beta_1 l_1)}{Z_1} + \frac{j\tan(\beta_2 l_2)\tan(\beta_1 l_1)}{Z_2}$$

$$D = \cos(\beta_1 l_1)$$

Computing with a simple program, the impedance seen from the emitter terminal works out to 15.6 – j114.42 Ω, which is deep into the unstable region as desired. The impedance trace for a frequency sweep from 2.2–2.6 GHz is plotted on the Smith chart.

Computing for different lengths and impedances allows you to construct an optimum circuit, keeping overall oscillator (or amplifier) design considerations in mind. Clearly, such visualization is not possible with the use of graphical techniques that give no guarantee that the best circuit has been achieved.

• *Example 2: series line with balanced open stub* — Designers prefer to lay out shunt stubs in balanced configurations to minimize transition interactions between the shunt line and the series transmission line. This example offers a study of balanced stub representation.

The schematic of a load circuit for a 2 GHz oscillator is shown in Figure 4. The load circuit is required to locate the reflection coefficient seen by the output port. The cascade of the microstrip line length with the balanced stub needs to be considered, so the layout has been redrawn in Figure 5 to show a hypothetical length l_3, separating the two sections of balanced stubs. The overall matrix is written as:

$$\begin{bmatrix} \cos(\beta_1 l_1) & jZ_1\sin(\beta_1 l_1) \\ \dfrac{j\sin(\beta_1 l_1)}{Z_1} & \cos(\beta_1 l_1) \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{j\sin(\beta_2 l_2)}{Z_2} & 1 \end{bmatrix} \bullet$$

$$\begin{bmatrix} \cos(\beta_1 l_3) & jZ_1\sin(\beta_1 l_3) \\ \dfrac{j\sin(\beta_1 l_3)}{Z_1} & \cos(\beta_1 l_3) \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{j\sin(\beta_2 l_2)}{Z_2} & 1 \end{bmatrix} =$$

$$\begin{bmatrix} \cos(\beta_1 l_1) & jZ_1\sin(\beta_1 l_1) \\ \dfrac{j\sin(\beta_1 l_1)}{Z_1} & \cos(\beta_1 l_1) \end{bmatrix} \begin{bmatrix} 1 & 0 \\ \dfrac{j2\sin(\beta_2 l_2)}{Z_2} & 1 \end{bmatrix}$$

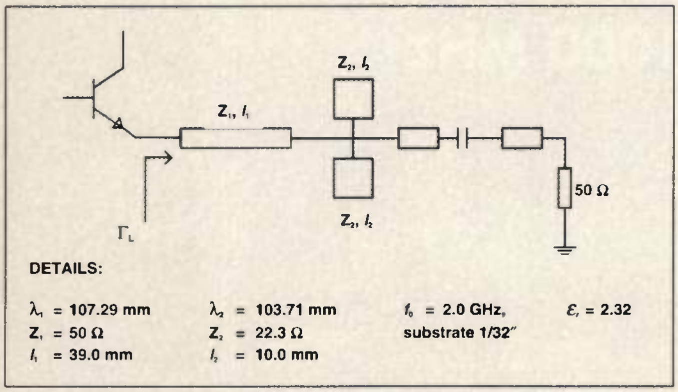

Figure 4. Matching with constant VSWR line and balanced stub.

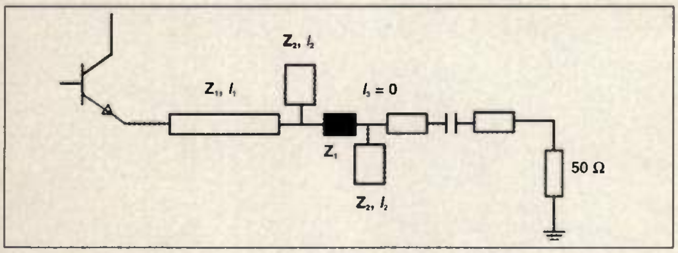

Figure 5. Balanced stubs separated by hypothetical line 13.

$$A = \cos(\beta_1 l_1) - \frac{2Z_1 \sin(\beta_1 l_1)\tan(\beta_2 l_2)}{Z_2}$$

$$B = jZ_1 \sin(\beta_1 l_1)$$

$$C = \frac{j\sin(\beta_1 l_1)}{Z_1} + \frac{j2\cos(\beta_1 l_1)\tan(\beta_2 l_2)}{Z_2}$$

$$D = \cos(\beta_1 l_1)$$

Applying Equation 1, the impedance seen from the emitter works out to be 24.32 – j105.24 Ω, which corresponds to a load reflection coefficient Γ_L of $0.84\angle -48.9°$.

• *Example 3: shunt stub with quarter-wave transformer* — An input match for an amplifier involving a shunt-shorted stub and a quarterwave transformer is shown in Figure 6. The first step is to determine the source reflection coefficient.

The overall matrix is simply:

$$\begin{bmatrix} 1 & 0 \\ \dfrac{-j}{Z_1 \tan(\beta_1 l_1)} & 1 \end{bmatrix} \begin{bmatrix} \cos(\beta_2 l_2) & jZ_2 \sin(\beta_2 l_2) \\ \dfrac{j\sin(\beta_2 l_2)}{Z_2} & \cos(\beta_2 l_2) \end{bmatrix}$$

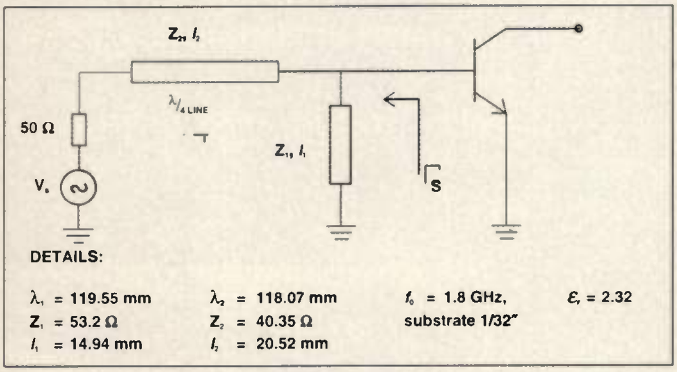

Figure 6. Matching with shunt-shorted stub and λ/4 transformer

$$A = \cos(\beta_2 l_2)$$

$$B = jZ_2 \sin(\beta_2 l_2)$$

$$C = \frac{-j\cos(\beta_2 l_2)}{Z_1 \tan(\beta_1 l_1)} + \frac{j\sin(\beta_2 l_2)}{Z_2}$$

$$D = \frac{Z_2 \sin(\beta_2 l_2)}{Z_1 \tan(\beta_1 l_1)} + \cos(\beta_2 l_2)$$

The impedance (Equation 1) is computed to be 23.69 + j14.5 Ω, which corresponds to a Γ_s of 0.4∠140°.

Conclusion

Despite the availability of many CAD tools, many RF deigners still prefer the old-fashioned Smith chart for starting a design. Operations on the Smith chart easily can become obscure, especially when line impedance changes are affected. The expressions derived for the A, B, C and D matrix elements for various microstrip element combinations provide an analytical method for verification and optimization. Matrix models can be used to compute the impedance presented by common microstrip matching networks. Although only two matching elements were used in each example, the method can be applied to analyze more complex schemes comprising several cascade stages of open or shorted stubs and line lengths. *RF*

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About the author

In 1983, R. Partha, M.E., joined the research and development division of ITI, Bangalore, India, where he primarily was responsible for the design of microwave phase-locked oscillators (PLOs) and synthesizers. He has worked extensively on digital microwave radio systems, and presently is working on network design for ITI's basic services division.

Microwave integrated circuits meet HF project demands for high-dynamic range

By Adrian I. Cogan, Donald Apte, Frank Sulak, Thomas Wei, Jim Spear and Lee B. Max

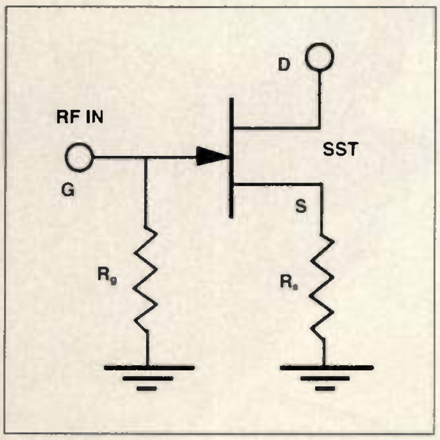

Figure 1. The 50 and 75 Ω power modules.

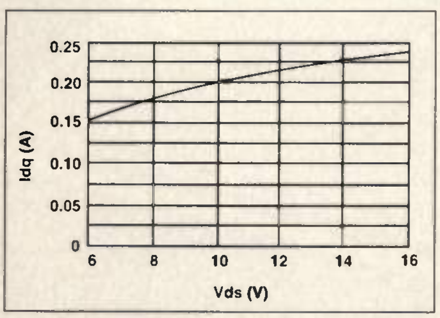

Figure 2. The I-V characteristics.

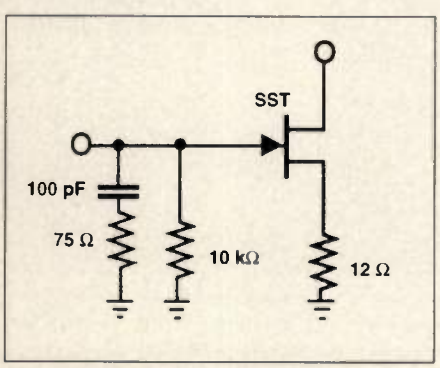

Figure 3. The AM-FM antenna amplifier.

RF engineers who need cost-effective, medium-power amplifiers with a wide dynamic range for applications at frequencies as high as 100 MHz may find a suitable device among a family of surface-mount, silicon field-effect transistor (FET) modules.

Requirements for higher capacity in next-generation wireless systems means that RF system designers must find active components with higher-than-ever linear dynamic ranges and the lowest noise figures. Oh yes . . . and the project must be *low-cost*.

A new family of low-cost, high-performance silicon FET gain blocks in SO8 packages meet such requirements. They deliver power levels from 0.2–2 W across the HF band. They are designed for 50 or 75 Ω operation and are available in single-ended and push-pull configurations.

Applications include reverse cable TV (CATV) amplifiers, RF cable modems, intermediate frequency (IF) amplifiers for personal communications service (PCS) base stations, windshield antenna preamplifiers and receiver front ends, including those used for magnetic resonance imaging (MRI) receivers.

The series was designed to meet the critical and challenging performance requirements defined for second- and third-order intercept points, third-order intermodulation distortion products, spectral purity, harmonic products and noise figure.

The internally matched, silicon FET, high-dynamic-range gain modules use thin-film, hybrid-circuit technology. The self-biased modules are internally temperature-compensated. They have suitable linearity with low-noise figures across the HF and VHF frequency range in applications at frequencies as high as 100 MHz.

Because the modules are self-biased and internally matched, they require fewer external components. Circuit simplicity conserves board space, improves reliability and reduces cost.

Circuit and operation

Transistor chips used by the modules are silicon FETs called *solid-state triodes* (SSTs) [1]. These FETs require the same biasing polarities as GaAs metal semiconductor field effect transistors (MESFETs). As shown in Figure 1, a source resistor inside provides the gate-

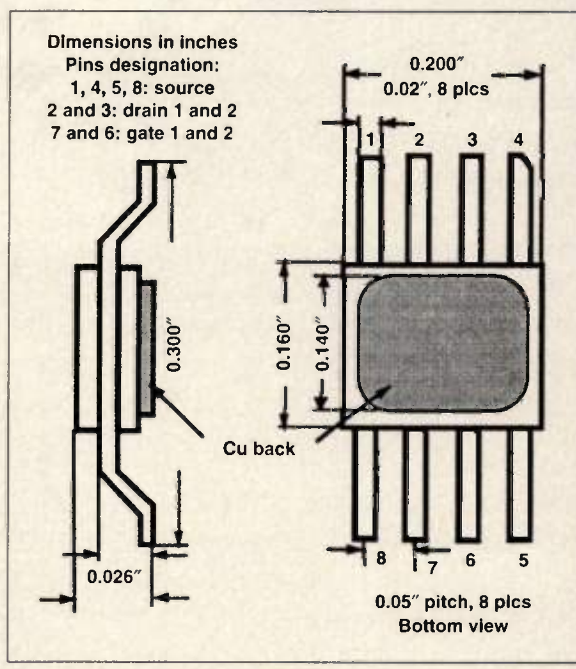

Figure 4. The push-pull power modules.

MODEL	f (MHz)	Gp (dB)	NF (dB)	P-1 (dBm)	IP3 (dBm)	IP2 (dBm)	Vdd (V)	Idd (mA)	Z (Ω)
single-ended, 50 Ω	≤50	12	4.5	33	44	46	12	250	50
single-ended, 75 Ω	≤50	11	4.5	31	42	44	12	150	75
single-ended, AM/FM antenna	0.1/100	11	4.5	26	39	41	12	100	var.
push-pull, 50 Ω	≤50	12	4.5	33	46	80	12	175	50
push-pull, 75 Ω	≤50	12	4.5	30	44	76	12	175	50
SHO3PP	<100	12	1.2	27	41	74	9	200	50/75

Table 1. Performance characteristics of for the family. The data is specified at tamb. 25°C.

biasing voltage required to select the operating point. This operating point is optimized for the best trade-off among RF power gain, linearity and dynamic range. The source resistor provides negative feedback, which results in improved linearity and improved temperature stability. The input resistor is selected to provide the required input-matching level.

The I-V characteristics are shown in Figure 2, which highlights the relatively small current vs. Vds variation. Remember that the single-sideband transmission has nonsaturated I-V characteristics [1]: The strong negative feedback developed across the source resistor is responsible for the drain-current saturation.

Another circuit version is shown in Figure 3. The passive elements were selected to provide an input impedance that varies with frequency in a predetermined manner. In the example, the resulting impedance values meet the requirements of a broadband AM-FM antenna match: high-magnitude capacitive impedance over the 500–1,700 MHz AM range and approximately 50 Ω over the 85–108 MHz FM range.

A third group of modules is configured as push-pull amplifiers, assembled in miniature surface-mount SO8 packages. (See Figure 4.) The internal circuit of the push-pull amplifier is shown in Figure 5.

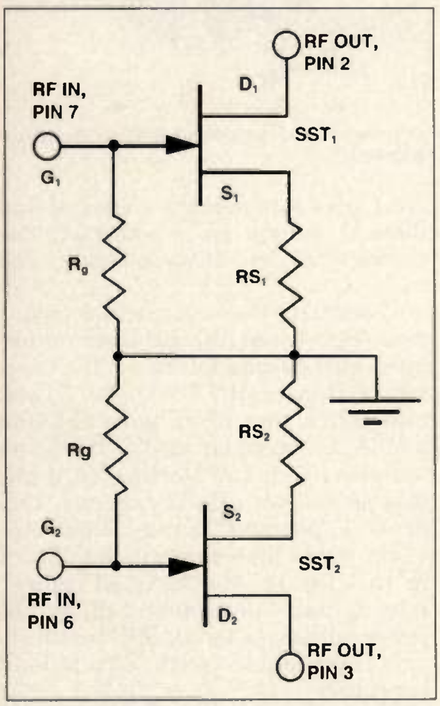

Figure 5. The internal circuits of the push-pull power amplifiers.

Performance summary

The family consists of five models, as follows: single-ended, 50 Ω power module (Figure 1); single-ended, 75 Ω power module (Figure 1); single-ended, AM-FM antenna amplifier (Figure 3); push-pull 50 Ω power module (Figure 5); push-pull 75 Ω power module (Figure 5). Table 1 lists the significant electrical characteristics.

In addition to the self-biased amplifiers, a "transistor-only" version is available in single-ended and push-pull configurations. The last line in Table 1 indicates typical performance achieved with a low-noise, push-pull, transistor-only amplifier. Its low noise figure makes these devices suitable for HF receiver application. Other applications include MRI front-ends and IF amplifiers.

Application examples

The single-ended circuit is shown in

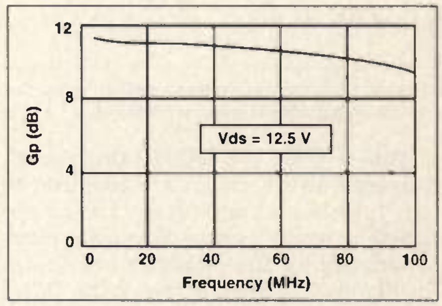

Figure 7. The power gain vs. frequency for a single-ended 50 Ω amplifier.

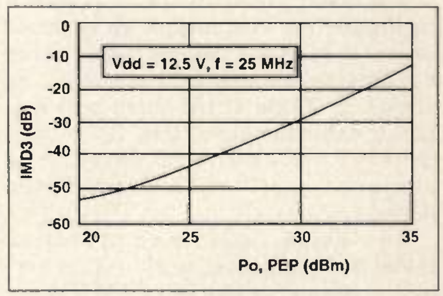

Figure 8. The IMD_3 vs. output power performance for the single-ended 50 Ω amplifier.

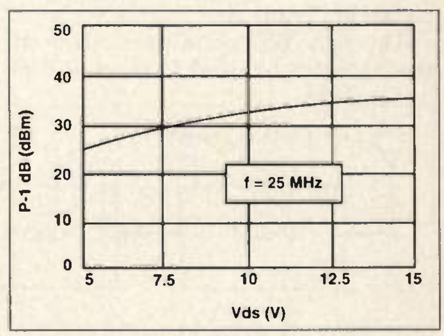

Figure 9. The P-1 dependence upon the drain biasing voltage for a single-ended 50 Ω amplifier.

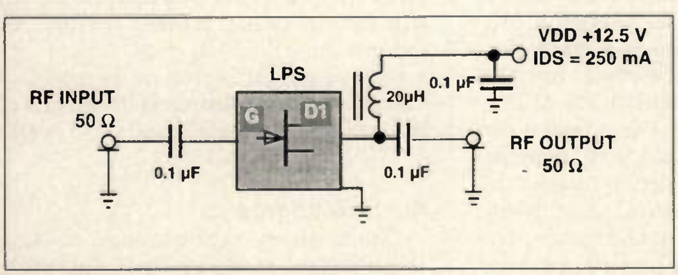

Figure 6. The single-ended circuit.

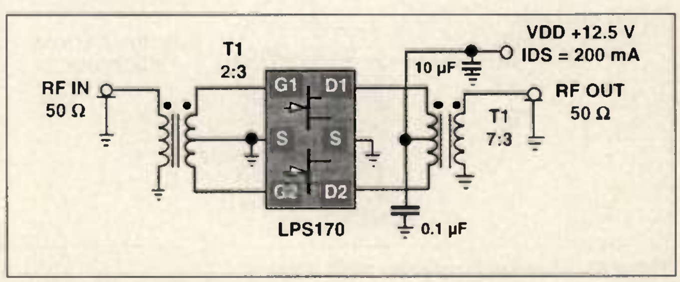

Figure 10. The push-pull circuit.

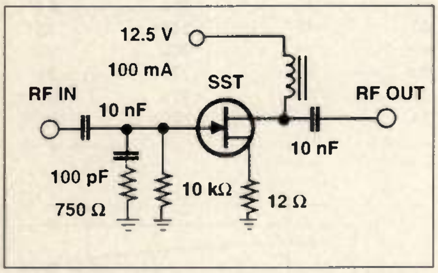

Figure 11. A typical antenna circuit using the single-ended AM-FM antenna amplifier.

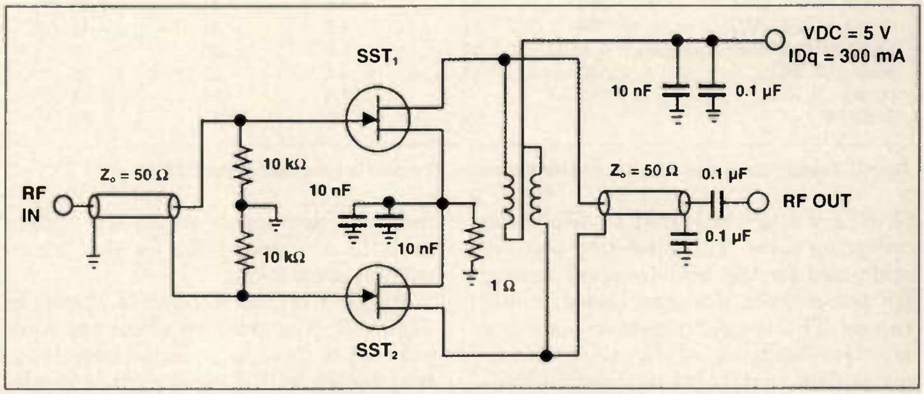

Figure 12. A front-end receiver using the SST chips for a medical device.

Figure 6. Only two DC-blocking capacitors and an RF choke are required to put together a single-stage HF amplifier. The power gain vs. frequency characteristics for the amplifier are shown in Figure 7. Figure 8 shows the IMD3 vs. output power performance. The P-1 dependence upon the drain-biasing voltage is shown in Figure 9.

The basic push-pull circuit is shown in Figure 10. The output transformer also is used in the biasing path, further simplifying the application circuit. As shown in Table 1, the push-pull amplifier exhibits satisfactory IP2 performance. Levels above 80 dBm can be achieved easily with prematched devices.

Low-noise, broadband operation make this component work well in low-cost preamplifier stages in AM-FM car radios, especially when used as part of the windshield antenna. A typical antenna circuit using the amplifier is shown in Figure 11.

The same SST chips used inside the modules can be used to fabricate HF amplifiers with 1 dB noise figures in single-ended, impedance-matched gain stages. These devices also are used as push-pull circuits. The example in Figure 12, which operates over the 2–50 MHz band, was designed as an input stage for front-end receivers used in medical equipment. The amplifier, biased at 8V, draws 300 mA, and it has a P-1 of 32 dBm, with an IP2 of over 70 dBm. The noise figure is less than 2 dB.

The module is used in wireless microphones and active LF and VHF whip antennas. Figure 13 is a wireless microphone block diagram [2]. Figure 14 is another application in which the device is used in a biomedical instrumentation amplifier [3].

Only a few years ago, the HF-VHF applications engineer had only one "linear" transistor choice: the bipolar junction transistor (BJT). Today, the difficult task involves selecting among a variety of bipolar junction transistors (BJTs), heterojunction BJTs (HBTs), metal oxide semiconductor field-effect transistors (MOSFETs), MESFETs and recently, SSTs [1]. Some of these transistors are fabricated on silicon, others on GaAs. The selection process is further complicated by the availability of low-cost integrated circuits [5, 6] supplied with or without internal matching. Furthermore, the selection process must take into account electrical and thermal performance, electrostatic discharge (ESD) sensitivity, reliability and cost.

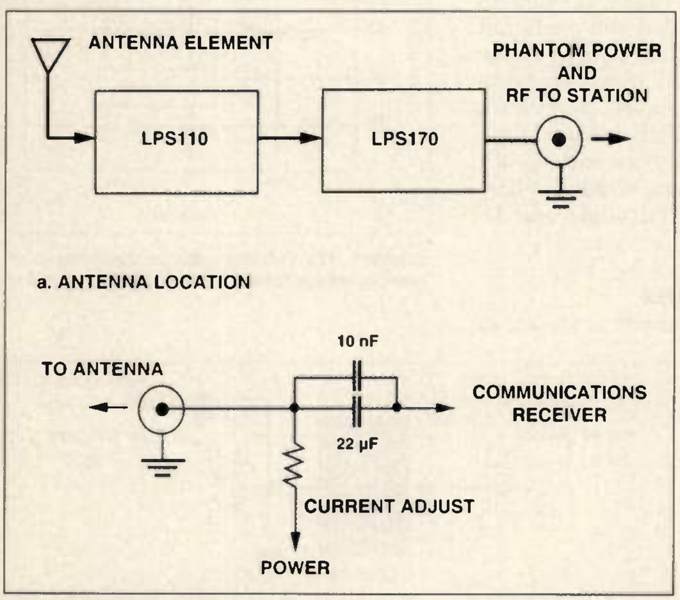

Figure 13. A wireless microphone block diagram.

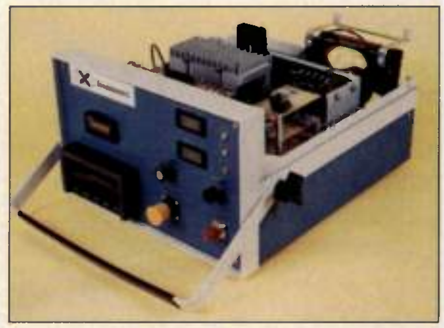

Figure 14. A biomedical instrumentation amplifier.

Presently, the competition comes from GaAs-based monolithic microwave integrated circuits (MMICs). The main versions are an HBT Darlington [5] and a Si Darlington [7] (Figure 15), and a MESFET-based circuit [6]. The figure indicates that the Darlington input biasing resistor network (Figure 15a) draws a "parasitic" current: some 20% of the total bias current. As shown in this figure, the "wasted" power roughly translates into a 2 dB output power advantage for an SST-based circuit (Figure 15b) with a 75 Ω load impedance.

Another advantage of the series is its low ESD sensitivity. In addition, when comparing thermal characteristics, silicon devices exhibit a lower thermal resistance than GaAs devices.

A final product selection is made, of course, on the performance trade-offs of the competing technologies. RF

Acknowledgments

The authors acknowledge critical contributions made by Neill Thornton,

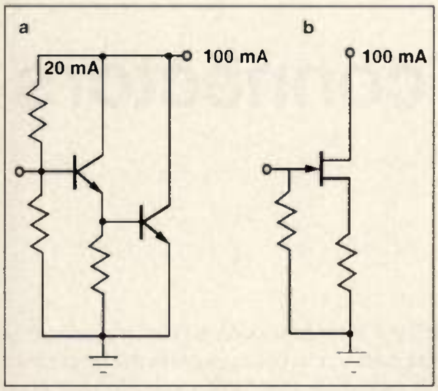

Figure 15. a) Darlington circuit. b) Single sideband transmission (SST) based circuit.

Jeff Phillips and Ken Sooknanan. The support and guidance provided by Masa Omori is appreciated.

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Microwave Technology manufactures high linearity devices and amplifiers for telecommunications, medical and military electronics, including the LPS family of products that is discussed in this article.

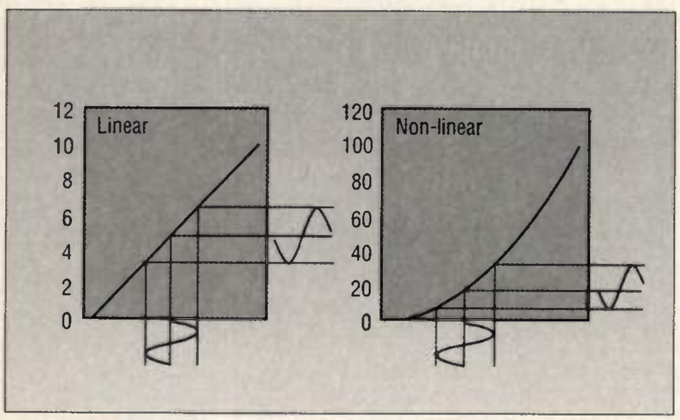
Intermodulation in coaxial connectors

By John King

Increased demand from the mobile communications industry for greater channel capacity coupled with the increased sensitivity of receivers has exposed a condition within RF coaxial connectors referred to as intermodulation distortion *(IMD). This condition occurs when non-linearities within the connectors act as imperfect diodes to generate other frequencies known as* intermodulation products *(IMP). Some of these frequencies appear within the receive band and effectively block the channel. This tutorial outlines basic causes of intermodulation and presents some techniques than can be undertaken to minimize the problem.*

Modern developments in base stations built for global system for mobile communications (GSM), digital communications systems (DCS) 1800 and personal communications service (PCS) 1900 have necessitated the use of 7-16, 4.1/9.5 and type N connectors because of the increased power requirements. The performance requirements typically are in the order of –160 dBc to –163 dBc (when working in dBc) or –120 dBm (when working in dBm), both with 2 × +43dBm tones. The requirement is so stringent because the connectors are used in post-filtering sections of the transmit path (between the diplexer and the antenna), and also because the system is full-duplex where the multiple-carrier transmit path is also the receive path. In a truly linear system, the output is directly proportional to the input, following the form of y = mx + c. (See Figure 1.) Coaxial connectors traditionally have been viewed as following this pattern. In reality, non-linearities always have been present in coaxial connectors. These were not readily apparent because the resultant IMPs were significantly below the system noise floor as a result of relatively weak carrier signals. This situation becomes apparent when the incident power is raised above 30 dBm.

The small non-linearities have a characteristic similar to a square law. (See Figure 2.) The distortion to the waveform is evident, the positive one-half-cycle being significantly greater in amplitude than the negative one-half-cycle. When converted to the frequency domain, this waveform consists of the desired fundamental plus a decaying series of related harmonics that, in themselves, interact with other carriers present on the transmission line.



Figures 1 and 2. Linear vs. non-linear response.

The effect of this interaction produces additional frequencies, some of which occur where they are least wanted. (See Figure 3.) The $2F_1 - F_2$ (third-order IMP, IMP_3), $3F_1 - F_2$ (fifth-order IMP, IMP_5) and $4F_1 - F_2$ (seventh-order IMP, IMP_7) products all can manifest themselves in the receive band and, if sufficiently large, they can effectively block a channel by making the base-station receiver perform as though a carrier were present when one is not.

Potential causes of IM in coaxial connectors

Identified below are the most likely factors that can affect intermodulation performance in RF coaxial connectors:

- Contaminated plating solution
- Insufficient plating thickness
- Corrosion
- Dissimilar metals in intimate contact
- Magnetic materials in the signal path
- Low contact pressure
- Less than 360° contact
- Poor surface finish
- Debris and dust within the connector
- Convoluted signal path

Remedies for IM in RF coaxial connectors

To combat these IM sources, the following precautions should be taken during the design and manufacture of the product:

- *High-quality plating to 6µm for IM-sensitive products* — The plating must be free from contaminants and must be properly *passivated* with a chromate passivate. Silver has been the preferred plating material because it possesses the lowest practical resistivity, thereby minimizing interface contact resistances. An alternative material is a white bronze plating finish, which provides excellent durability, tarnish resistance and non-magnetic properties that are ideal for low intermodulation. During testing with a system noise floor of –145 dBm, the difference in performance between silver-plating and white bronze is not discernible.
- *Restrict materials to copper and its alloys* — This ensures maximum plating adhesion and minimum electrochemical potential difference between the base materials and their overplatings.
- *Avoid the use of stainless steel, nickel and ferrites in the signal path* — Magnetic and para-magnetic materials only compound non-linearites and give poorer interface contact re-

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ERA-2SM	DC-6000	13.1	13.0	6.0	27	2.00
ERA-3	DC-3000	20.2	11.0	4.5	23	2.10
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

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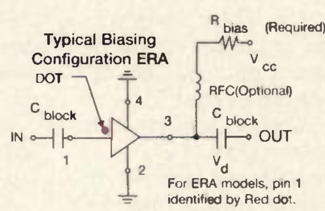

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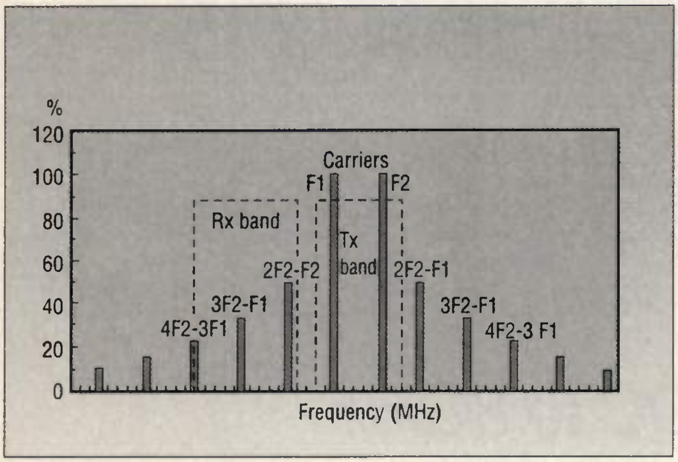

Figure 3. Spectrum of products in a two-tone system.

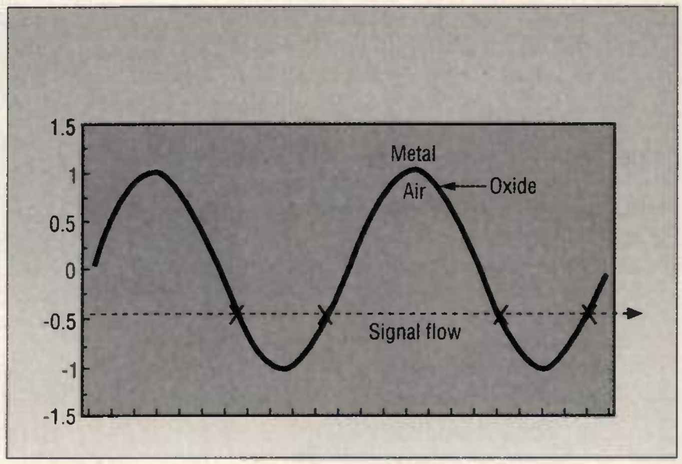

Figure 4. Effect of surface roughness.

sistances. During experimentation, a degradation in performance of 20 dB was discovered when nickel plate was used. Moreover, the presence of magnetic or para-magnetic materials cause the forward IMP figure to differ from the reverse IMP.

• *Quality machining* — Surface finish is paramount. The signal propagates within a "skin." If this skin is too rough, the signal repeatedly transitions through metal- and surface-oxide layers, thereby creating the same effect as a poor panel contact. (See Figure 4.) For IM-sensitive designs, 0.4 mm is the maximum.

• Contact design — This primarily affects the connector interface. Repeated matings can generate small amounts of plating from the individual parts. These oxidize and interfere with the mechanical (and therefore electrical) mating of connectors. The oxidized debris gives further rise to metal- and surface-oxide junctions and, consequently, to higher IM products.

• *Ensure, by design, a properly defined contact interface at connector, panel and contact interfaces* — Insufficient contact force gives rise to metal-oxide junctions. The classic rectifiers were metal oxide by composition. Axial-maximum material condition at the connector interface is critical to ensure minimum mismatch and maximum potential of a butt-contact. Panel interfaces generally concern the physical contact of the connector body to the panel. In this case, it has been determined that a protruding feature as close as possible to the body bore gives the best IM performance. The applied mounting force is concentrated in the surface area of the protrusion which, on engagement with the panel, punctures the existing oxide layer to give a metal-to-metal, gas-tight junction.

• *Avoidance of crimps* — Crimps, by nature, can only give multiple-point-contact, rather than 360° contact, and can cause a variability in the position of electrical contact during dynamic testing. IM products, therefore, will be greater. It has been found that soldered-center contacts and clamp-and-solder outer contacts give the best static and dynamic IM performance.

Commonly-asked questions

1. *Why is intermodulation such a concern for cellular infrastructure equipment?*

The primary concerns for cellular service providers are *channel efficiency* and *clarity of transmission*. Growth in demand for mobile communications has created a need to operate equipment at greater capacities and reliability to serve the competitive market. Intermodulation deteriorates or limits the ability of the service provider to operate at optimal performance levels and ultimately may cause subscribers to experience poor call quality. Intermodulation has become an important factor in system selection to ensure the best possible network service.

2. *Where is intermodulation most likely to occur in cellular infrastructure equipment?*

Intermodulation typically is of greatest concern between the filtering elements of the system and the antenna. The introduction of higher power levels for the transmit side of the equipment creates greater potential for intermodulation. This is why the primary focus for intermodulation concerns 7-16, type N, surface-mount and 4.1/9.5 connector interfaces.

3. *Is intermodulation a recent development?*

Intermodulation always has been present in RF coaxial connectors, but it may be relatively imperceptible in some devices for a variety of reasons. The amount of power applied to an RF connector determines the relative IM threshold that can be observed. Intermodulation, therefore, is more likely to cause concern in a higher-power system using as an example, a 7-16 connector interface, rather than an equivalent low-power product. The trend toward higher-power, digital cellular systems creates the need for greater intermodulation sensitivity.

4. *What is the best method of cable attachment for IM-sensitive cable assembly applications?*

Soldering and clamping are preferred methods because of the 360° point of contact created at the cable-to-connector interface. Such intimate contact improves the overall contact resistance leading to improved IM characteristics. It also is better to solder the center conductor of the connector to the cable rather than to use crimping because of the improved contact resistance path and because of the elimination of voids.

5. *Are there ways to test for intermodulation in an RF coaxial connector?*

Sophisticated methods are needed to test for intermodulation in RF connectors. The test system must use extremely sensitive filtering or clean amplification so that the equipment itself has a low intermodulation noise floor. There is no standard approach to testing, although an international committee has been formed in the connector industry to address the situation. A state-of-the-art test facility exists where designs are optimized for low-intermodulation performance and where further analysis on the effects of this phenomenon can be studied.

6. *Is intermodulation in coaxial connectors frequency-dependent?*

No. Because coaxial connectors are broadband, there is no frequency dependency. Some apparent variability can be detected during testing, but not because of the connector. The impedance matches of the output diplexer and triplexer and terminations are the causes of the variations. Causes should not be incorrectly attributed to the connector or assembly. By varying the impedance match of the test-station termination, a device under test (DUT) can show 15 dB better IMP_3 than what really exists.

7. *IMP_3 in mixers follows a 3dB/dB relationship. What is it for connectors?*

The relationship is identical. Taking the third-order ($2F_1 - F_2$): Varying the power of F_2 gives an IMP_3 relationship of 1dB/dB, whereas varying the power of F_1 gives a relationship of 2dB/dB because the IMP is derived from the second harmonic of F_1. This gives a total of 3dB/dB when symmetrically varying both carrier powers.

8. *I am buying a complete cable assembly. How do I interpret the IMP result now?*

With caution! When testing devices, some companies have a policy to move away from the normal static test to a dynamic test where the cable termination interfaces are mechanically exercised during live IM conditions. Furthermore, it is also a good indicator to customers of the "build quality" of the assemblies. A dynamic evaluation has shown 15dB degradation in IMP performance for poor assemblies and even as much as 50 dB for bad ones. It therefore is strongly advised to state IM-performance figures in the context of a dynamic measurement.

Improving IM connector design

The pursuit of design techniques that improve intermodulation performance to address the emerging tele-communications market needs continues. A state-of-the-art intermodulation test facility and participation on the international committee (IEC SC46D WG5) to develop standard test practices ensures a commitment to the understanding of intermodulation characteristics. This applied technology base is instrumental in developing innovative low-intermodulation products for 7-16, Type N, surface-mount and 4.1/9.5 connector interfaces. *RF*

About the author

John King is a technical manager in charge of intermodulation and lightning protection at M/A Com in Harlow, United Kingdom.

CDMA signals: A challenge for power amplifiers

By Klaus D. Tiepermann

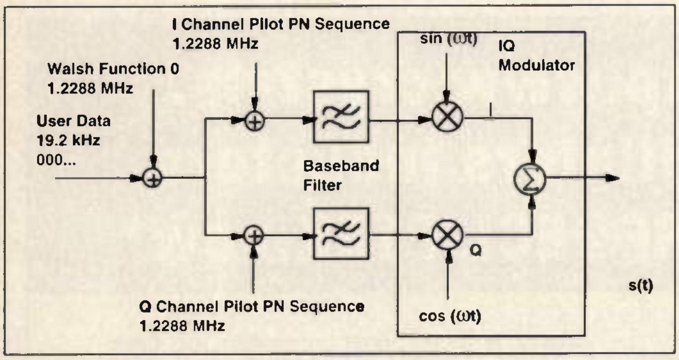

Figure 1. Generating the pilot signal of a base station.

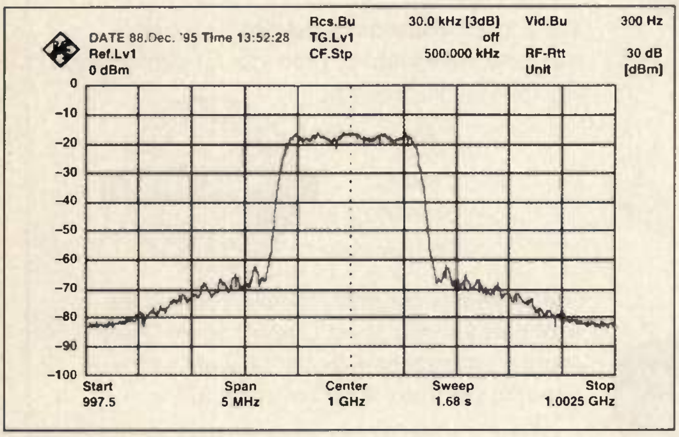

Figure 2. CDMA spectrum.

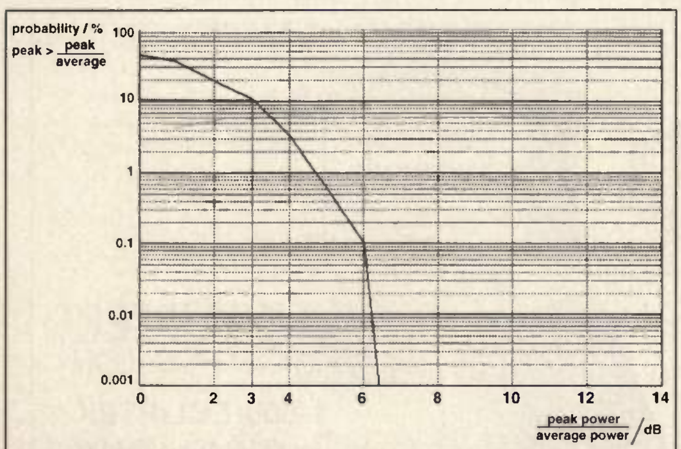

Figure 3. Statistical power distribution of a pilot signal where the probability of occurance for instantaneous envelope power values is given in decibels above the average power.

Code-division, multiple-access (CDMA) signals place stringent requirements on the linearity of power amplifiers. Theoretically, the peak power may attain 100 times the average power, depending on the number of CDMA channels transmitted at the same time. In addition, the crest factor and the statistical distribution of the power amplitudes is of particular interest. These aspects have to be taken into consideration when selecting a suitable signal source for performing measurements on power amplifiers.

The following signal characteristics are important for the development of CDMA power amplifiers: *frequency and bandwidth; average power, peak power* or the statistical distribution of power amplitudes in general; and *effects of nonlinear distortions.*

The most important CDMA mobile radio network from a commercial point of view was defined in Interim Standard IS-95 [1] by the Telecommunications Industry Association (TIA). This standard specifies the air interface parameters, including modulation, baseband filtering and spread of the spectrum. The following discussion is in reference to this standard.

Analysis of CDMA signals

Figure 1 shows the generation of the pilot signal of a base station (IS-95, forward channel). Quadrature phase-shift keying (QPSK) modulation is used for the RF carrier. Baseband filtering is defined in IS-95 in terms of a digital finite impulse response (FIR) filter. The user data in the pilot channel and the bits of the Walsh function are all zero, causing only the in-phase (I) and quadrature (Q) pilot sequence to be modulated.

Figure 2 shows the resulting RF spectrum. The baseband filtering additionally comprises a phase equalizer to simplify the design of the receiver filter in the mobile station. Because the filter is an allpass filter, the spectrum is not influenced.

Figure 3 shows the probability of the occurrence of power peaks to average power for an individual CDMA channel (pilot signal). The graph shows that, for example, the instantaneous power value exceeds the average power (0 dB) by 3 dB for 10% of the time. For the power signal, only power peaks of more than 6 dB above the average power occur with less than 0.1% probability. The highest peaks are about 6.5 dB above the average power. This value is called the *crest factor* of the signal. The baseband signal consists of additively superimposed 64 code channels at maximum, with each channel having a different Walsh function for the spectrum spread. (See Figure 1.) The spectrum is not changed because of this superimposition, but the crest factor is considerably increased.

Figure 4 shows the measured distribution functions for two and nine CDMA channels. The signals were generated with a Rohde & Schwarz model SMHU 58signal generator. A Helwett-Packard peak power analyzer was used for mea-

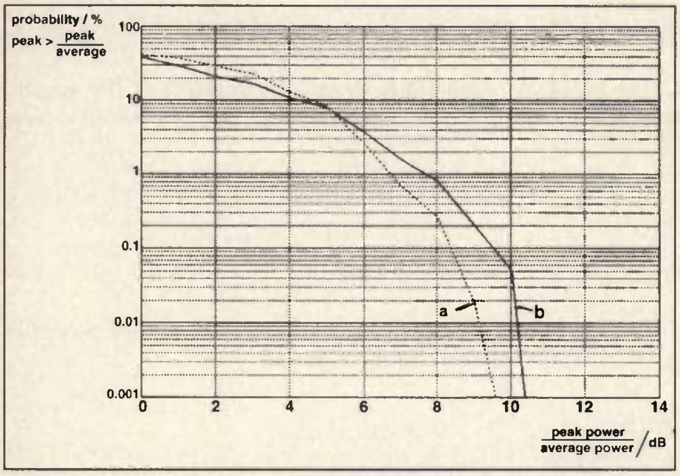

Figure 4. Statistical power distribution for a signal with two and nine CDMA channels. Curve (a) represents the pilot and traffic channels ($P_{traffic} = P_{pilot} - 3dB$). Curve (b) represents the nine-channel CDMA signal.

suring the power. The analyzer determines the magnitude of the instantaneous power by continually measuring and statistically evaluating the samples of the RF signal envelope. For the power amplifier design engineer, the peak power range of 6 to 10 dB above the average power is of particular interest. For nine channels, peak levels above 10 dB only occur with probabilities less than 0.05%.

In comparison, the conditions at the mobile station are not as demanding. Only one CDMA channel is sent. Because offset QPSK modulation is used, a crest factor of only 4.5 dB is attained, although the power amplifier of a mobile station has to cope with the so-called *power gating*. For data rates below 9,600 bps, the transmission is interrupted for the length of individual power control groups (1.25 ms). For data rates of 4,800 bps, for example, a duty cycle of only 50% is obtained for a mobile station's transmit signal.

Noise as a test signal

Spectrally, CDMA signals are similar to band-limited white noise. Therefore, a noise generator that can generate band-limited noise with Gaussian amplitude distribution (known as amplitude gaussian white noise or AWGN) is often recommended as a test signal source.

The probability density function f_G of an average-free Gaussian noise signal $s_G(t)$ is:

$$f_G(x) = \frac{1}{\sigma\sqrt{2\pi}} e^{-\frac{x^2}{2\sigma^2}}$$

where:

σ^2 = variance

The envelope of a band-limiting noise signal can be derived

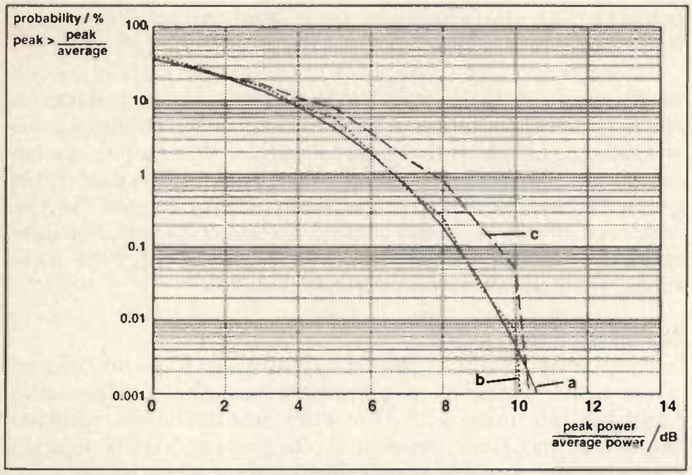

Figure 5. A comparison of statistical power distributions. Curve (a) represents the exponential distribution as theoretical distribution function of AWGN. Curve (b) represents a measured distribution of an AWGN signal. Curve (c) represents a nine-channel CDMA signal.

from the formulation of a complex Gaussian process. In this case, the real part s′(t) and the imaginary part s″(t) are to be taken as uncorrelated Gaussian-distributed random processes. The magnitude of the complex envelope is:

$$|s(t)| = |s'(t) + js''(t)| = \sqrt{s'(t)^2 + s''(t)^2}$$

The resulting probability density function f_R of the amplitude x = $|s(t)|$ then becomes a Rayleigh distribution (also see [2]):

$$f_R(x) = \frac{2x}{\sigma^2} e^{-\frac{x^2}{\sigma^2}} \quad \text{for } x \geq 0$$

$$f_R(x) = 0 \quad \text{for amplitudes } x < 0$$

The probability density function f_E of the envelope power P = $|s(t)|^2$ is an exponential distribution:

$$f_E(P) = \frac{1}{P_{avg}} e^{-\frac{P}{P_{avg}}} \quad \text{for } P \geq 0$$

where P_{avg} = average power.

The statistical distribution function F_E for the envelope power of band-limited Gaussian noise is then:

$$f_E(P) = 1 - e^{-\frac{P}{P_{avg}}} \quad \text{for } P \geq 0$$

Figure 5 shows 1- F_E with logarithmic scaling for the envelope power on the X-axis and the associated probability as a percentage on the Y-axis. The representation corresponds to that of Figure 3. In addition to the theoretical noise distribution, a measured distribution is also shown. The associated AWGN signal was generated using the SMHU 58 with a CDMA coder. The distribution function of the nine-channel CDMA signal is illustrated for comparison.

It is clear to see that peak values that are 6 to 10 dB above the average value (the critical range for power amplifiers) occur with a far higher probability for the CDMA signal than for the AWGN signal. This applies even though the crest factor of an AWGN is higher. Theoretically, the crest factor can even approach infinity. A comparison with Figure 4 shows that the AWGN distribution in the interesting range of 6 to 10 dB roughly corresponds to the distribution with only two CDMA channels.

Two conclusions can be drawn:

1. A noise generator with a Gaussian amplitude distribution is not a complete substitute for a multichannel CDMA signal, even if the AWGN has a high crest factor.

2. For measurements with AWGN or with a two-channel CDMA source, a test level of at least 1 dB higher should be used to approximate the statistical distribution of the nine-channel signal in the range of 3 to 10 dB.

Linear distortions

Linear distortions cause amplitude response and nonlinear phase response (group-delay distortion). A limit value for deviations from the ideal phase trajectory in a base station is defined in IS-95. Accordingly, the root mean square deviation from the ideal phase trajectory in the complete transmitter from the baseband filter to the RF output may be 0.01 rad^2. This parameter is determined by the integration of the baseband range from 1 to 630 kHz:

$$\varphi_{err} = \int_{1\ \text{kHz}}^{630\ \text{kHz}} \left[\varphi(f) - \varphi_{ideal(f)}\right]^2 df \le 0.$$

This parameter must be considered, in particular, when designing baseband lowpass and RF bandpass filters.

Nonlinear distortions in amplifiers

Nonlinear distortions are, however, of much greater importance for power amplifiers. Nonlinear distortions of the third order cause intermodulation products in the adjacent-frequency channel. Two measurements on a power amplifier with a maximum power of 10 W (1 dB compression point) clearly underline this.

Trace *a* in Figure 6 indicates the output spectrum of an individual CDMA channel (pilot) at an output power of 1 W (average). Trace *b* shows the spectrum with two CDMA channels again at an output power of 1 W (sum of the two channels). It can clearly be seen how intermodulation products are generated as a result of higher peak amplitudes despite a 10 dB margin to the maximum power. The adjacent-channel power (ACP) thus increases.

Limit values for the permissible

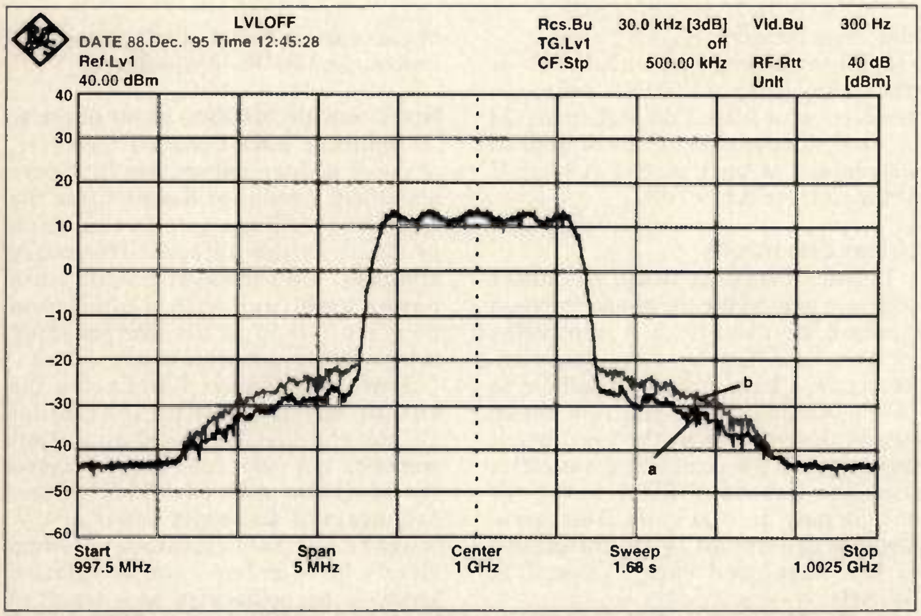

Figure 6. CDMA spectrum with intermodulation sidebands at the output of a power amplifier. Curve (a) represents a pilot signal with an output power of 1 W. Curve (b) represents pilot and traffic channels (–3 dB) with a total power of 1 W.

ACP are stipulated by FCC regulations as well as by IS-97 [3] and IS-98 [4]. IS-97 stipulates for the first adjacent channel, for example, that the ACP measured with a resolution bandwidth of 30 kHz shall be at least 45 dB below the total average output power.

All distortions affect the signal quality of the modulation signal. For the measurement of the signal quality, the waveform quality factor ρ (pronounced *rho*) for the pilot signal has been defined. ρ is a measure for the correspondence of the ideal and the actual transmit signal. It is evaluated from the cross-correlation between a calculated ideal signal and a measured transmit signal. In the ideal case, ρ = 1.

CDMA test source

Based on the above discussions, the following requirements are to be fulfilled by a signal generator for measurements on CDMA power amplifiers:

- CDMA modulation according to IS-95
- Forward-link transmit signal with several code channels and QPSK modulation
- Reverse-link transmit signal with offset QPSK modulation and power-gating
- Waveform quality ρ > 0.966 (IS-98 specification for standard test equipment)
- High-level accuracy with error < 1 dB, level of individual code channels can be set separately
- Large frequency range, including 900 MHz band and personal communications service (PCS) band (1,900MHz), as well as intermediate frequencies
- Spectral purity for ACP measurements
- One suitable signal generator includes a CDMA coder as a one-box solution. It has two freely configurable CDMA channels (forward channel) and an integrated noise generator. A reverse channel signal can be generated as well.

A measuring system consisting of a signal generator, an arbitrary waveform (ARB) generator and software for generating CDMA test signals with more than two code channels can be configured. The ARB generator generates I and Q signals that can then be used as input signals for the IQ modulator of the signal generator. The software is used for the calculation of IQ signals [5].

A multichannel signal with as many as nine channels can be defined by means of a convenient menu (Figure 7). The menu was created in line with the specification of a base-station test model in IS-97. After defining the channel levels and the desired Walsh functions, the IQ signals can be calculated and transmitted to the ARB generator. Lowpass filters should follow the ARB generator to suppress aliasing products. The analog IQ signals generated in this way are modulated in the signal generator. An RF frequency range from 1 to 2000 MHz is then available. *RF*

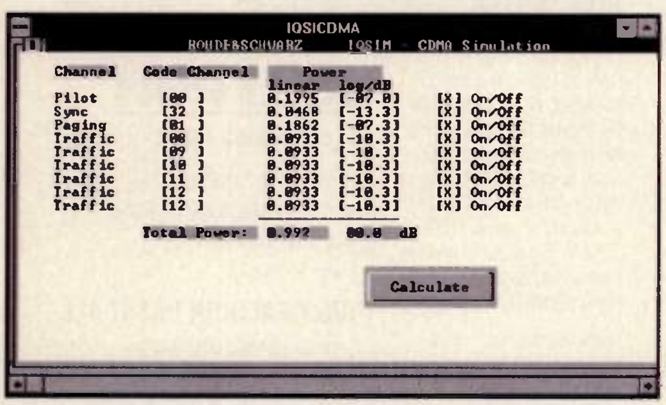

Figure 7. Configuration menu for multichannel CDMA signal.

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About the author

Klaus D. Tiepermann is with Rohde & Schwarz. The company makes the SMHU 58 signal generator, the ADS arbitrary waveform generator, and IQSIM-K software described in this article. Alternatively, a Tektronix AWG 2021 can be used as the ARB generator.

ENGINEER'S NOTEBOOK

Inductive tuned oscillator

By Dien M. Nguyen

In establishing inexpensive modem links for communications, low-cost resistive-capacitive (RC) oscillators are popular. RC oscillators have some disadvantages, such as temperature drift and frequency-pushing susceptibility. Frequency drifts can be as high as 1,500 PPM per degree Celsius. In designing products for outdoor environments such as cable TV distribution equipment, circuits are required to operate over an extended temperature range. A small drift in center frequency beyond the deviation frequency results in the failure of the communications link. Another problem that haunts CATV systems is the low regulation of ferro-resonant, coaxial power supplies. Thus, voltage-controlled oscillators (VCOs) must have good "pushing" or power supply rejection capability. To improve these deficiencies, we must resort to a higher-Q oscillator.

Choosing an oscillator topology

The oscillator used (Figure 1) is a capacitive-feedback Clapp oscillator. Depending on bias constraints, other topologies can be used. This oscillator requires a single positive supply. Resistors R_1, R_2, and R_3 set the DC bias point for Q_1. You can optimize these values to maximize transisitor gain. The dominant tuning mechanisms are set by L_1 and C_4. (C_2 and C_3 are set to be approximately 10 times the value of C_4.) The inductor choice has the greatest effect on the quality factor of the oscillator. The traditional method of tuning a VCO circuit uses a varactor diode. Varactors pose disadvantages, such as lowering the Q factor of the VCO, because of their inherent series resistance. Also, at lower frequencies, VCO tuning range is limited by varactors' small capacitance values and dynamic range. Overall, active devices and their associated components are mostly capacitive, thus the variation of *capacitance* with a varactor is mostly a *secondary* effect in varying frequency. The primary choice for any broadband VCO would be to vary *inductance*.

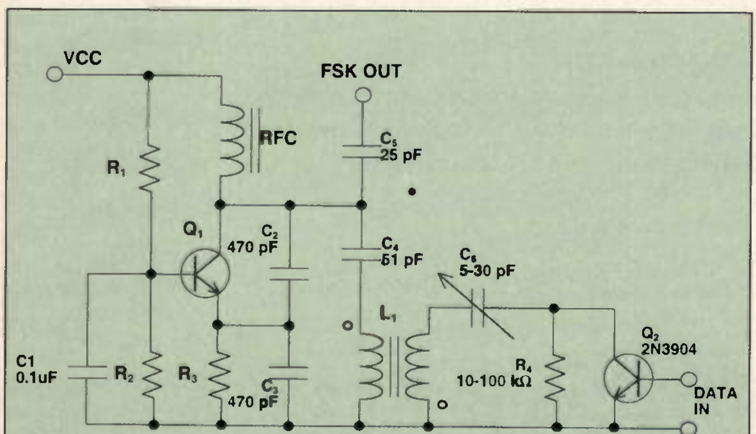

Figure 1. A capacitive-feedback Clapp oscillator.

Inductive coupling of LC tank

In a frequency-shift keying (FSK) application, when modulating high data rates, coupling data to the varactor can pose a problem. A diplex filter is necessary to prevent RF loading or leaking of the oscillator tank while transferring the fast data edges to the varactor. These two constraints contradict one another, especially when the edge's rise time approaches that of the carrier frequency. If an RF choke or a large-value resistor is used to inject data to the varactor, then the edges of the data will be severely distorted by RC or inductive-capacitance (LC) time constants. The loss of these rise times may result in duty-cycle distortion as seen by the threshold level detector in the receiver.

To improve upon the traditional VCO techniques, varying the inductor provides increased tuning sensitivity and better resonator Q. Inductive tuning uses the coupling or leakage inductance characteristics of a transformer. Thus, tuning capacitor C_4 can be changed to a fixed, high-quality capacitor. A use of a negative temperature coefficient capacitor here can balance out the active device's capacitance variation with temperature. The tuning inductor L_1 is now a transformer. At about 48 MHz, it is a 9-turn, center-tapped coil of AWG28 wound on a T25-10 Micrometals core. Other transformers also can be implemented, depending on operating frequencies and performance requirements. The Q for this inductor is about 100, which is more than sufficient to isolate the active device from pulling the oscillator tank over the extended temperature range of –40°C to 85°C. The inductance variation or stability with temperature is dependent on the core material of L_1.

Temperature stability

Inductance of toroidal transformer L_1 may be calculated as follows:

$$L = \frac{4\pi\mu N^2 A_e}{l_e} \text{ nH} \tag{1}$$

where:

A_e = effective core cross sectional area (cm^2)
l_e = effective magnetic path length (cm)
N = number of turns
π = pi
μ = effective permeability of core material

Inductance can be calculated using toroidal dimensions of OD, ID, and Ht, but it is more accurate to use the manufacturer's A_e and l_e (or A_L) values. Most manufacturers have already accounted for the toroid edge radius or roundness factors in their data. Because the physical parameters (l_e, A_e, and N) do not change with tempera-

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

*Max. Current (mA) @ 8V DC.
Notes: Tuning voltage 1 to 16V required to cover freq. range. 1 to 20V for POS-1060 to -2000. Models POS-50 to -1025 have 3dB modulation bandwidth, 100kHz typ. Models POS-1060 to -2000 have 3dB modulation bandwidth, 1MHz typ.
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ture, it is safe to say μ is the only temperature-dependent variable. The toroid core material, Carbonyl iron powder (μ = 6.0), was specified to have a temperature stability of 150 PPM/°C. Relative to RC oscillators, this is an order of magnitude better in temperature stability.

Operational theory

When the secondary winding of transformer L_1 is left unconnected or open (or if Q_2 is off), the primary inductance sets the oscillation frequency with C_4. When it is shorted or grounded (Q_2 is on), the primary inductance will change to a lower value that power supply engineers call *leakage inductance*. Ideally, this value is equal to zero, but because of parasitic inductance, the oscillation frequency can be varied. Thus, transistor Q_2 serves to short out or load the secondary winding of L_1 at the rate of the data to vary the output frequency for modulation. Coupling capacitor C_6 serves to prevent heavy Q-loading of the primary tank, and it also serves as a secondary resonator tank oscillating with L_1's secondary inductance. The second resonator is set at a higher frequency than the primary because the value of L_1's primary inductance can be lowered only when the secondary is loaded. Thus, when Q_2 turns on, the primary tank is pulled toward the higher frequency of the secondary tank. As an additional benefit, the high impedance of Q_2's collector is excellent for isolating the data line from the high-impedance RF oscillator node.

FSK amplitude balance

The amplitude balance between the two modulating frequency states can be adjusted by resistor R_4. When Q_2 is open, the primary resonator is undisturbed and has the highest resonating Q. When Q_2 is closed, the deviated frequency will have lower output amplitude because of the decrease in total Q. The lower Q is attributed by core losses, copper losses and Q_2's parasitic losses. The purpose of resistor R_4 is to set the Q for both oscillating states equal by slightly lowering the Q of the primary tank when Q_2 is open. This way, the output amplitudes of the ones' and zeroes' state (frequency) will be balanced.

Frequency deviation

The frequency deviation using this implementation was measured to be as high as ±2 MHz for a center frequency of 48 MHz. Adjustment of frequency deviation can be accomplished in two ways. One way is to limit how hard Q_2 turns on; the other way is to vary the coupling level between the primary and the secondary windings. To accomplish the latter, you must physically vary (compress or spread) the turn spacing of the windings on the core of L_1. You can see that most of the design was optimized in the laboratory. My experience with magnetics using iron powder cores tells me that time is better spent characterizing a transformer on laboratory equipment than on simulation or by number crunching on the computer. Because of numerous variations between physical winding techniques and material tolerances, a closed-form equation may not be practical.

To address the power supply sensitivity and temperature variation, you must pick a transistor (Q_1) with a high enough f_T so that oscillation is sustained throughout the operating temperature range. A general-purpose transistor can work sufficiently, but any RF transistor similar to the MRF581 will guarantee temperature tolerances at 48 MHz.

Linear FM applications

Although the application of this circuit is used for digital FSK modulation where linearity is not so important, it also can be applied to linear FM systems. If used in this way, Q_2 would need to be biased to its linear operating point, and the slope of KVCO (modulation sensitivity) can be controlled by the gain of the amplifiers that drive Q_2.

For better load-pulling performance or output-matching, a buffer is highly recommended. Capacitor C_5 is a –10 dB tap for the output port. I have measured as high as –5 dBm (Vcc of 5 V) of output power into a 50 Ω load at 48 MHz.

Conclusion

Based on designs by the grandfathers of all oscillators, a few inexpensive transistors still can compete in both cost and performance with today's semiconductor manufacturers. I hope this idea has helped to open another avenue of thought regarding the established or mature theory of VCOs. On the other hand, the subject of naming oscillator topologies has been disputed for many years between Colpitts, Clapp, Armstrong and Hartley. Maybe I'll throw another wrench in the pool of names and call this design the "Nguyen VCO." *RF*

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About the author

Dien M. Nguyen has a B.S.E.E. from the University of California at San Diego. His experience has mostly been in the field of optical fiber communications. He worked at PCO, and TACAN, designing RF-to-optical fiber interfaces for cable television and telecommunications systems. He also worked in the field of switch-mode power supplies, where power magnetic design experience was invaluable. Presently, he is an MTS at Nippondenso America, where he develops RF designs for cellular and PCS phones.

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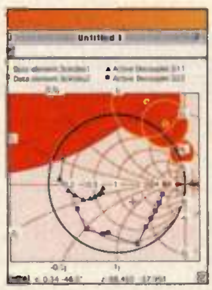
Amplifier Research's model AT1080 log-periodic antenna sweeps the full 80–1,000 MHz frequency band required by IEC-1000-4-3. The AT1080 features 6.5 dBi minimum gain, maximum power input of 750 W and suitable for up to 600 V/m susceptibility testing.
Amplifier Research
Souderton, PA
800-933-8181: www.ar-amps.com
INFO/CARD 175

A 20-page catalog of quartz crystals and clock oscillators features HC-49UP, SMX-500 and -301, HC-49 and 50/U, HC-49/US, HC-43U, HC-25 and -18CU, UM-1, TO-5, LPO, HCMOS, TTL, FTTL, ECL and VCXO.
Standard Crystal, El Monte, CA
818-443-2121: FAX 818-443-9049
INFO/CARD 176



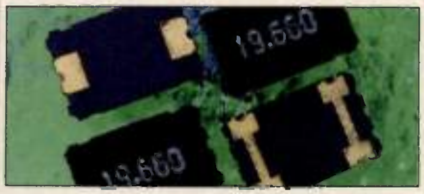
Dragon Wave circuit simulation software for PowerMac by GigaSim is a filter synthesis package that goes from specification to physical dimensions in one step. Features include schematic entry, full-noise analysis, custom models, optimization and data sheet layout.
GigaSim
Melbourne Beach, FL
407-725-5987: FAX 407-725-6071
www:http://www.gigasim.com
INFO/CARD 177

Surftrim sapphire dielectric surface-mount trimmer capacitors (GPY series) from Sprague-Goodman measure 5.2 × 4.3 × 3.2 mm and features O-ring seals, four capacitance range, a Q of greater than 450 at 100 MHz, a temperature range of –40°C to 85°C, a rated voltage of 65 VDC and are available in carrier-and-reel or bulk pack.

Sprague-Goodman Electronics
Westbury, NY
516-334-8700: Fax 516-334-8771
INFO/CARD 178

Coaxial connector types offered by Alpha Products include SMA/B/C, BNC, TNC, F, MCX and nano-coaxial. The connectors are manufactured to stringent Mil-C-39 or -12 specs. Specialized cable assemblies are available that are tested at frequencies as high as 26 GHz. Custom connectors can be made to order. Ask for catalog #238.

Alpha Products
INFO/CARD 179



VF/FE and VF/FH series miniature ceramic SMT crystals are only 1.8 mm high. The frequency range is 10,000–100,000 MHz, and the J-leaded molded plastic package is available in two land patterns. Prices are less than $1 each in 1,000-piece quantities.

Valpey-Fisher
508-435-6831 ext. 234
Fax 508-497-6377
INFO/CARD 180

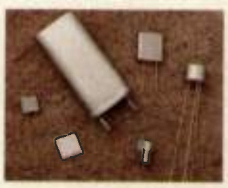
Model PA2500 spectrum analyzer allows any PC to function as a full-featured spectrum analyzer. Basic coverage is from 100 kHz to 1.2 GHz, extendable to 2.5 GHz. Battery operation allows field use with a notebook host for as long as one hour.

DKD Instruments
INFO/CARD 181

B+K PRECISION
RF PRODUCTS GROUP

Programmers for PROM, EPROM, EEPROM, Flash, PALs, GALs and microcontrollers are undated by software. Models 840 (one device) and 841 (four devices) handle more than 500 EPROM devices. Models 842 (plug-in board) and 843 (parallel port) handle more than 1,800 devices of all types. Prices begin at $229.

B+K Precision
INFO/CARD 182



ICM manufactures special and hard-to-find frequency control products including low frequency crystals from 8 kHz to 200 MHz. Unusual holder styles and sizes such as the TO-5 (HC-35) and gull wing surface mount, HC-49/U, HC-45/U are available in small quantities for quick delivery. Also available are pager crystals, crystal filters, channel elements and TCXO, VCXO, TCVCXO, 8- and 14-pin dip oscillators.

International Crystal Mfg.
1-800-725-1426
INFO/CARD 183

NEW PRODUCTS NO.28

RF/IF MICROWAVE COMPONENTS

DUAL OUTPUT VCO IS LOAD INSENSITIVE

200 to 380MHz linear tuning, low -100dBc/Hz (typ) phase noise at 10kHz offset, and excellent -24dBc (typ) harmonic suppression are important characteristics of Mini-Circuits new ZOS-400 connectorized voltage controlled oscillator. Main and auxiliary ports have excellent isolation from the main oscillator (50dB typ), and frequency settings are virtually independent of the load. Applications include PLL circuitry and production testing. The operating temperature range spans -55°C to +85°C and worldwide shipment is guaranteed within one week.

12dB FIXED ATTENUATOR IMPROVES IMPEDANCE MATCH

Mini-Circuits low cost PAT-12 fixed attenuator operates in a wide DC to 7000MHz bandwidth offering a 12dB± 0.6 nominal attenuation value with 0.6dB maximum midband flatness. VSWR is excellent throughout the entire band (example: VSWR 1.4 max. midband at +25°C case temp.). The surface mount device is housed in a rugged ceramic package and can be used in 50 ohm environments for power leveling and impedance match improvement. Available in tape and reel.

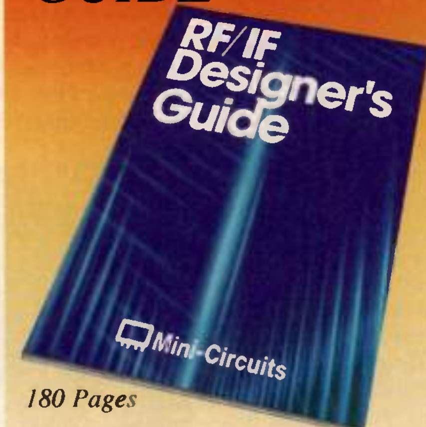
SMALL, LEVEL 13 (LO) MIXER IS USABLE TO 7.5GHz

Mini-Circuits SKY-60MH microwave mixer has the technology to travel the broad 2500MHz to 6000MHz frequency range with typically low 6.2dB conversion loss, high 28dB L-R, 14dB L-I isolation (IP3; 19dBm), and features IF response usable to 2.5GHz. Low in cost, this versatile surface mount component stands only 0.10" high, allowing for greater board density and design efficiency. Includes a 5 year Ultra Rel® guarantee. Available tape and reel.

I&Q DEMODULATOR HAS GOOD AMPLITUDE/PHASE UNBALANCE

This low cost JCIQ-895D I&Q demodulator from Mini-Circuits spans the 868MHz to 895MHz frequency range with typically good amplitude unbalance of 0.2dB, phase unbalance of 1 degree (both measured at -5dBm RF), and 8.6dB conversion loss. Typical 3rd and 5th order harmonic suppression is excellent at -45dBc and -65dBc respectively. The surface mount device is housed in a shielded metal case measuring only 0.80x0.87x0.25 inches and is equipped with solder plated J leads for strain relief. Available in tape and reel.

DESIGNER'S GUIDE



180 Pages

FREE!

Call Today
718 934-4500
Fax 718 332-4661

LOW COST MMIC AMPLIFIERS FOR COMMERCIAL USE

The wide DC to 4000MHz bandwidth of Mini-Circuits ERA-4 and surface mount ERA-4SM monolithic amplifiers eliminates the need for costly compensation networks and extra gain stages while providing up to 19.1dBm power output (at 1dB comp), and 38.5dBm IP3 (both typ. at 1GHz). The drop-in version exhibits 1.62:1 in, 1.38:1 out VSWR (typ) and 1.60:1 in, 1.33:1 out for the surface mount design. Both typically deliver high 13.9dB gain (±0.10dB flat to 2GHz). The SM unit is available in tape and reel packaging.

Mini-Circuits®

US 68 INT'L 78
CIRCLE READER SERVICE CARD

P.O Box 350166, Brooklyn, New York 11235-0003 (718) 934-4500 Fax (718)332-4661 *INTERNET* http://www. minicircuits. com

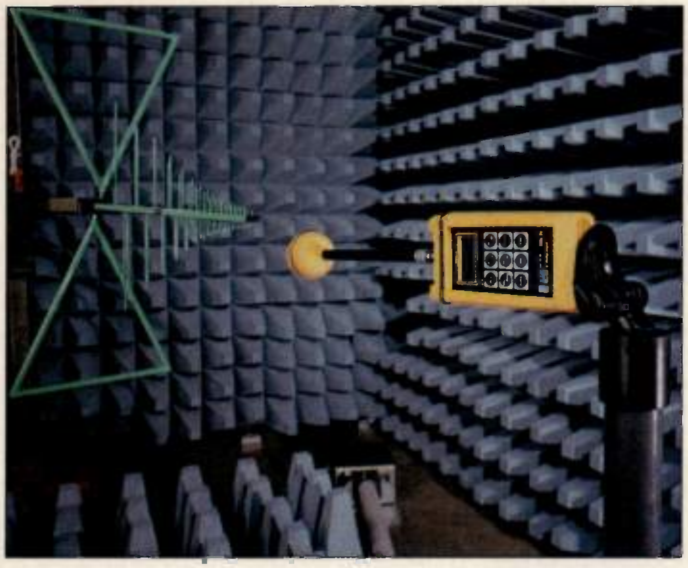
For detailed specs on all Mini-Circuits products refer to • 740- pg. HANDBOOK • INTERNET • THOMAS REGISTER • MICROWAVE PRODUCT DATA DIRECTORY • EEM

CUSTOM PRODUCT NEEDS...Let Our Experience Work For You.

F 231 Rev Orig

RF products

Electric field probe monitors RF fields from 1–800 V/m



The EMC 20 electric field probe from Chase EMC monitors RF electric fields from 1–800 V/m over the frequency range 100 kHz to 3 GHz without bandswitching. The EMC 20 has an isotropic sensor measuring three orthogonal axes. Data from each axis is internally combined to give a non-directional summation, making the EMC 20 satisfactory for field monitoring. The probe is compact and battery-operated and has a built-in optical interface allowing each of the three axes to be evaluated separately. It has complete remote control of all instrument functions via a standard RS232 personal computer (PC) serial data communications port. The EMC 20 uses a zeroing facility that is valid even in the presence of high field strengths. By using the Chase data transfer windows software, data can be tabulated or recorded in spreadsheet form for further processing and presentation.
Chase EMC
INFO/CARD 184

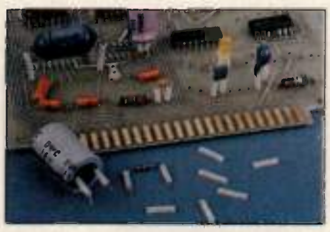
SAW oscillators for workstations

The CMD 5000 series DIL surface acoustic wave (SAW) clock oscillators from C-MAC are designed for use as high-speed processor clocks in high-performance workstations or as pixel dot clocks on graphics cards in high-performance workstations. The CMD 5000 series is available in a range of frequencies from 250–800 MHz. These

frequencies are generated directly, thereby minimizing possible degradation to performance, which typically results when starting with a lower frequency and using multipliers. These devices feature low jitter achieved through the use of SAW resonators operating at the output frequency, frequency stabilities of ±200 ppm and a high degree of reliability.
C-MAC Quartz Crystals
INFO/CARD 185

PC board standoffs protect components

M. M. Newman offers a line of Teflon tubing in a wide variety of sizes that can



be cut to length and supplied as PC board standoffs for raising and protecting components and for improving solder flow. These standoffs are used for raising components on plated through-hole PC boards and for drawing solder into the holes by capillary action. Helping to prevent electrical discontinuities, they can be supplied cut to length in 56 standard sizes ranging from AWG #30 to AWG #0 with varying wall thicknesses. Featuring a 2.1 dielectric constant and 1,400 v/mil strength, these standoffs are chemically inert and nonflammable, and they operate from –450°F to 500°F.
M. M. Newman
INFO/CARD 186

Digital transmitter chip for wireless

The GC4114, a four channel, all-digital transmitter chip from Graychip, digitally up-converts four external signal sources to user-programmable center frequencies. The GC4114 is suitable as a transmitter building block in digital cellular telephony, personal communications systems (PCS) and other wireless telecommunication systems. The GC4114 features four identical up-converter channels. Input sig-

nals can be supplied in a variety of formats. The GC4114 also features an 80 dB spur-free dynamic range (SFDR) and a 70 dB noise power ratio (NPR). The unit is priced at $60.63 in quantities of 10,000.
Graychip
INFO/CARD 187

Chip low-pass filters keep unwanted harmonics out

The LTF series of chip low-pass filters for personal communications systems is designed to keep unwanted harmonics out of the system. It offers a miniature

1206 footprint (3.2 × 1.6 mm) and a low profile of 1.4 mm. The LTF series from Toko America features a typical insertion loss of 0.35 dB, an impedance of 50 Ω and a carrier frequency range of 800 MHz to 2.6 GHz with standard part numbers for most common systems including cellular and personal communications systems (PCS).
Toko America
INFO/CARD 188

TEST EQUIPMENT

Frequency synthesizer supports wireless

The PTS W800 is the first in a series of high-performance frequency synthesizers from Programmed Test Sources. Spanning 800–960 MHz, the synthesizer provides coverage for all current and emerging wireless communications technologies. The instruments feature fast frequency switching (20 µs), low phase noise (–110 dBc at 1 kHz offset), fine-frequency resolution (0.1 Hz) and low spurious outputs (–65 dBc). The unit is priced at $4,800 in OEM configuration and at $5,100 with manual control.

Programmed Test Sources
INFO/CARD 189

Communication analyzer tests PHS and PDC formats

The MT8801A radio communication analyzer from Anritsu Wiltron quickly evaluates all transmitter and receiver parameters of personal digital communications (PDC) and personal handyphone systems (PHS) digital cellular formats. The single instrument platform performs dozens of tests that previously required several types of interconnected instruments. In 1.5 seconds or less, the MT8801A tests the following: adjacent channel power, occupied bandwidth, origin offset, frequency and frequency error, leakage power during carrier off, modulation accuracy, antenna power and go or no-go decision of rise-and-fall edge characteristics with template. The MT8801A features a frequency range of 300 kHz to 3 GHz and is priced at $46,100.

Anritsu Wiltron
INFO/CARD 190

EMI test receiver can cover 9 kHz to 2.5 GHz

The ESPC, a compact, lightweight electromagnetic imaging (EMI) test receiver for pre- and post-certification measurements from Tektronix, covers a frequency range of 150 kHz to 1 GHz with optional frequency extensions from 9 kHz to 2.5 GHz. The unit fits electronic certification applications including measurement and control engineering and mechanical engineering. It also can be used in medical information technology equipment and in the automotive and communications industries.

Tektronix
INFO/CARD 191

DISCRETE COMPONENTS

Current-sensing resistor suits high-volume use

The BVS low-ohmic, surface-mount, current-sensing resistor from Isotek is designed for high-volume applications that are price-sensitive and that can accomodate a resistance tolerance of ±5%. The BVS features a thermal resistance of less than 6°C/W and is available with resistance values of 0.3, 0.5, 1, 2 and 3 mΩ. BVS prices start at 38 cents each in quantities of 100,000. The minimum order is 1,000.

Isotek
INFO/CARD 192

Box capacitors offer 5.0 and 7.5 mm spacings

The MMKS (7.5 mm) and MMKT (5.0 mm) series of metalized polyester box capacitors feature operating voltages from 50 to 630 Vdc. Manufactured by Seacor, they offer tolerances from ±5% to ±20% and are supplied with tinned copper leads from 6–18 mm in length. Rated values are from 0.001–1.0 mfd with custom modifications on request.

Seacor
INFO/CARD 193

SMD crystal for wireless data-com applications

The CX1 series of quartz crystals from Micro Crystal is designed for use in wireless data-communications including modems, pagers, local area networks (LANs) and Personal Computer Memory Card International Association (PCMCIA) applications. The crystals come in a ceramic package and are available in the frequency range of 10 kHz to 2.1 MHz. The crystals feature low power dissipation, good stability

over their temperature range and high shock and vibration resistance.
Micro Crystal
INFO/CARD 194

High-frequency inductors for wireless communications

Models IMC-1210-100 from Dale have a wide range of high-frequency circuit applications including impedance matching, bandpass filtering, amplifier biasing, resonant circuits and noise filtering. They are available in an inductance range from 0.01–0.1 μH with self-resonant frequencies as high as 2.5 GHz. They are available with tolerances ranging from ±5% to ±20%, and they measure 0.098″ × 0.126″ × 0.087″.
Dale Electronics
INFO/CARD 195

SIGNAL PROCESSING COMPONENTS

OCXO covers 10–30 MHz range

Model 4598T from Oak Frequency Control has many commercial applications including personal communications service (PCS) base stations, cellular base stations, synthesizers and test equipment. The 4598T features a 10–30 MHz frequency range, aging of ±0.05 ppm per year and a temperature stability specification of $\pm 5 \times 10^{-9}$ over 0 to 70°C. It measures 38.1 × 27.2 × 20.0 mm. At 10 Hz, the phase noise is –110 dBc/Hz.
Oak Frequency Control
INFO/CARD 196

Sub-miniature oscillators cover 1.5–66.667 MHz range

Model EC2500 and EC2600 sub-miniature ceramic oscillators from Ecliptek are available at working voltages of 3.3 or 5.0 V, and they measure 7.5 × 5.0 × 1.8 mm. These oscillators are suited for Personal Computer Memory Card International Association (PCMCIA) cards, disc drives, personal digital assistants (PDAs), laptop computers, or any application where space and low-power consumption are critical. The EC2500 and EC2600 series have a frequency range of 1.5–66.667 MHz.
Ecliptek
INFO/CARD 197

VCXO frequency error never exceeds ±25 ppm

VCXO models M2031/2/3 from MF Electronics are tested and are individually guaranteed to have less than 25 ppm frequency error (deviation) at 2.5 V control voltage. Frequency capture ranges of 50 ppm, 100 ppm and 150 ppm are guaranteed from 0°C to 70°C. They feature a frequency range of 1–175 MHz and a first year aging of 3–5 ppm, dropping to 1 ppm per year thereafter.
MF Electronics
INFO/CARD 198

SIGNAL SOURCES

Surface mount LC filters feature 1.3 dB insertion loss

Model 8342 from PTI has a center frequency of 600 MHz and offers a 1.5 dB bandwidth of 300 MHz. The unit features an insertion loss of 1.3 dB and high-side rejection of 20 dB at 833 MHz. The unit is packaged in a 1.45″ × 0.50″ × 0.42″ enclosure.
Piezo Technology
INFO/CARD 200

Bias tee covers 45–6,000 MHz range

A bias tee from Sierra Microwave Technology features a frequency range of 45 MHz to 6 GHz. It has an insertion loss of 1 dB maximum, a voltage standing-wave ratio (VSWR) of 1.4:1 and a DC current capability of 2.5 A. In addition, a 2–18 GHz and a 0.1–26.5 GHz model is available.
Sierra Microwave Technology
INFO/CARD 201

400-18,000 MHz directional coupler for test systems

A directional coupler from Merrimac features a nominal 13 dB coupling that has been designed for an ultra-wide

bandwidth of 400 MHz to 18 GHz and for exceptionally high directivity, typically in excess of 20 dB. A directional coupler is a passive device that samples a small fraction of the power from a signal, with minimum disturbance to the transmission path. It differs from a simple resistive tap in that the coupled power is always proportional to the signal power being transmitted in one direction only. The package measures 5.8" × 0.8" × 0.53" and weighs 4.4 ounces.
Merrimac Industries
INFO/CARD 199

Software service provides shareware and freeware

Dynacomp's software service supports mathematical, scientific and engineering computing. A technically oriented collection of more than 550 volumes on more than 750 disks of public domain shareware and freeware for MS-DOS and Windows is available on CD-Rom, or it may be ordered by specific volume.
Dynacomp
INFO/CARD 153

Software for Macintosh simulates circuits

GigaSim has acquired the DragonWave circuit simulation software for Macintosh from Nedrud Data Systems and offers DragonWave 7.0 for Power Macintosh. GigaSim also has acquired Microwave Lab from Lightwave Technologies. Microwave Lab's extensive element model library will be incorporated into a DragonWave upgrade later this year. GigaSim is moving into the filter synthesis market with its acquisition of the Filtroid design software from Geesaman Software. Filtroid goes from specification to physical dimensions in one step.
GigaSim
INFO/CARD 154

Block diagam simulator comes in Windows version

Tesoft's PC telecom simulator Tesla is available to Windows users. Tesla lets users simulate modems, synthesizers, spread spectrum and other telecom systems at the block diagram level. It supports nonlinear, mixed-analog and digital-time simulation. Spectrum analysis also is included.

The Windows interface displays results graphically. After a simulation, users can plot spectra, waveforms, bit-error ratio (BER) test results, eye-patterns and logic analyzer style graphs. Multiple plot windows can remain open on the screen along with the block diagram.
Tesoft
INFO/CARD 155

Custom DSP technology becomes portable

Hyperception, a provider of digital signal processing (DSP) technology, has enhanced its Hypersignal real-time integrated development environment (RIDE) visual DSP design tool with a Personal Computer Memory Card International Association (PCMCIA) interface adapter for portable DSP applications. The combination of the visual programming software and Communication Automation and Control's Bulletdsp PC card allows engineers to develop real-time signal processing applications with a floating point DSP processor using a graphical interface. Custom DSP applications can be designed and executed on laptop computers at remote sites and in mobile environments.
Hyperception
INFO/CARD 156

Products support 136 cellular interface standard

TCSI's simulation, DSP algorithm and microprocessor software supports the North American time-division multiple access (TDMA) 136 cellular interface standard (IS-136). The DSP algorithms and simulations software enables developers to determine how best to process the cellular signal when it comes off the air. Developers tune the numbers on a computer rather than on the actual device to achieve the best performance possible.

The simulation software is written for the COSSAP signal processing environment. The baseband DSP software is the implementation of the algorithms. The software sits on the chip and functions as a low-level assembler language. It converts real-world (analog) signal to digital technology and back again at high speeds.

The protocol microprocessor software ensures that the cellular device is compatible with and can talk to any other cellular device in the world. It complies with applicable standards. The software configures and controls how the DSP converts signals at high speed. The microprocessor determines responses to input such as when the device rings, when the screen light goes on and whether the call will be analog or digital.
TCSI
INFO/CARD 157

Software supports wireless, networking, products

Software from Alta Group of Cadence Design Systems addresses the wireless communications arena including personal communications services (PCS), wireless local area network (LAN), advanced messaging, satellite communication and digital cellular standards such as IS-95, IS-136 and global system for mobile communications (GSM). The software, EnWave, focuses on specifications for networking, wireless and multimedia product design.

The software is built upon Alta's Convergence Simulation Architecture and provides continuous development from system-level specifications through links to hardware and software implementation. The software includes core modeling and simulation environment; application-tuned communications libraries containing hundreds of algorithmic elements for exploration of wireless communications designs in both floating and fixed-point simulation; a radio-frequency library allowing designers to model distortions and nonlinearities in the RF portion of a wireless design; analysis tools; and standard verification environments. The user can move from initial design capture, simulation and analysis through hardware and software implementation.

Previous designs may be reused, or the user may access off-the-shelf intellectual property, saving time with each wireless design. The software test benches allows the user to create prototypes.
Alta Group
INFO/CARD 158

Software automates measurements

EMI Consulting's commercial measurement program, EMICMP, Version 2.0 is a DOS-based, menu-driver program that operates with an Anritsu MS2601B spectrum analyzer to automate EMI-type measurements as prescribed by all current regulations. The EMICMP performs both precompliance and final compliance measurements for E-field, H-field and conducted-type measurements over the frequency range of 9 kHz to 2 GHz.

Measurement options permit the selection of three measurement levels: trace only; trace and peak of individual signals and trace; peak and quasi-peak or average of individual signals. Other options include automatic calibration at four levels. EMI offers custom installation, training classes and customized versions of the program.
EMI Consulting
INFO/CARD 159

RF literature

Chip inductor engineering bulletin

Sprague-Goodman is offering a 10-page engineering bulletin covering Surf-coil SMT chip inductors. Bulletin SG-800D contains specifications, outline drawings, Q, current ratings and DC resistance on its present line and includes its new GLW, GLX and GLY series.
Sprague-Goodman Electronics
INFO/CARD 160

Surface-mount clock oscillator data sheet

MF Electronics has a comprehensive data sheet detailing the company's ClockChip ASIC for building time-and-frequency reference clock oscillator circuits. It includes specifications, dimensions and packaging.
MF Electronics
INFO/CARD 161

Coaxial connector catalog includes installation tips

San-tron has a 28-page catalog of coaxial connectors with supplemental performance and material data, dimensional specifications and outline drawings for plugs, jacks, receptacles and dust caps. Also included are detailed cable assembly instructions and recommended panel cutout diagrams.
San-tron
INFO/CARD 162

Catalog has electromechanical switches

Loral Microwave-Narda's RF and microwave electromechanical switches are covered in a 108-page catalog. New to this catalog are low-cost models, hot-switching models, switches for wireless applications and a matrix section complete with worksheet.
Loral Microwave-Narda
INFO/CARD 163

Catalog lists connectors for telecom, broadcast use

Hirose's *Connectors for Communications and Broadcast Applications* features connectors for exchange and transfer as well as for broadcast and cable television (CATV). Main-edge connectors for exchange and transfer applications include inter-board connectors for high-speed transfer. Backplane cable connectors feature 50Ω and 75Ω coaxial ribbon connectors for high-speed transfer. Connectors for broadcasting and CATV include optical and electrical coupling connectors using a self-shutter mechanism, 75Ω set-top coaxial connectors and high-frequency devices, including micro-miniature signal distributors for surface-mount technology.
Hirose
INFO/CARD 164

Wireless data handbook covers integrated circuits

The *1996 Semiconductors for Wireless Communications Data Handbook* (IC-17) provides technical information on Philips Semiconductors' wide range of integrated circuits for wireless communications. The handbook includes data sheets, application notes and system solution references. Also covered is information on wireless system chipsets, RF front-ends and paging receivers, amplifiers, intermediate-frequency (IF) systems, frequency synthesizers and prescalers, transmitters, baseband processors, compandors, discrete transistors and power modules. Application notes describe products used in cellular radio, high-performance receivers, cordless telephones, two-way communications and wireless local area networks. The handbook is free and is available on CD-Rom.
Philips Semiconductors
INFO/CARD 165

Miniature surface-mount capacitors catalog

A new issue of Voltronic's catalog of miniature, one-half-turn, surface-mount trimmer capacitors includes improved documentation. The publication lists technical specifications and mechanical dimensions.
Voltronics International
INFO/CARD 166

Book details microstrip antennas

CAD of Microstrip Antennas for Wireless Applications by Robert A. Sainati details why microstrip antennas are suitable for low-profile, low-cost commercial applications using RF and microwave systems. The book includes a description of microstrip antennas; pertinent technology; analysis and modeling techniques to sharpen insight into the operation of various elements and arrays; step-by-step coverage of the design of various single-element radiators; discussion of the interaction between design parameters and antenna performance; an in-depth explanation of simple array design, including performance and calculations, types of architecture and specific design aspects; and a discussion of advanced feeding techniques that provide increased design flexibility and performance.

The 256-page hardback comes bundled with IBM-compatible diskette containing 18 stand-alone computer programs and a guide illustrating sample design cases and expected results. A 286 processor or higher and VGA monitor are required for the software.
Artech House
INFO/CARD 167

Fault-finding guide has tips for using basic gear

Practical Electronic Fault-finding and Troubleshooting, a 240-page paperback by Robin Pain, describes the fundamental principles of analog and digital fault-finding and gives practical tips, hints and rules of thumb for fault-finding using only the basic equipment—a digital multimeter and an oscilloscope.
Butterworth Heinemann
INFO/CARD 168

Handbook aids inductor specifications

The Inductor Handbook by Cletus J. Kaiser combines inductor and transformer theory and construction with practical circuit-application information. Chapters cover fundamentals and applications of all inductors; ferrites and transformers, including application information on pot cores, toroids, beads, chokes and slugs; electromagnetic interference (EMI) suppression; data-line filtering and power applications; power and signal transformer applications, with a section on switching-power magnetics. Also included are a glossary, symbols and equations, conversion tables, a ferrite-material constant chart, ferrite-selection guidelines and a comprehensive index. This 170-page, softcover reference book complements the author's other passive-component handbooks on capacitors and resistors. The price is $15.95 plus $4 (U.S.) shipping.
CJ Publishing
INFO/CARD 169

GaAs MMIC SWITCHES

3 VOLT - INTEGRATED DRIVER - EXTENDED FREQUENCY RANGE

Configuration	Operating Frequency (GHz)	Insertion Loss (dB)	Isolation (dB)	Features	Part Number
SPDT	DC- 2.0	.5	23	3 volt positive control; economy priced	SW-373
SPDT	DC- 2.0	.5	33	Extended frequency; industry standard	SW-239
SPDT	.8 -2.0	.9	38	Integral CMOS Driver	SW-335
SPDT	DC-2.5	.5	39	Terminated internally	SW-338

*Typical parameters at 1GHz.

HIGH POWER HANDLING

Configuration	Operating Frequency (GHz)	*Insertion Loss (dB)	*Isolation (dB)	Features	Part Number
SPDT	DC- 2.5	.5	32	+33dBm, P-1dB	SW-277
SPDT	DC- 2.0	.5	17	2 Watt power handling; single neg/pos control	SW358 /SW359A

*Typical parameters at 1GHz.

In quantities of 1000: $2.29

In quantities of 1000: $2.50

The GaAs IC Name You Know, From The Distributor You Trust

You know that M/A-COM has earned a reputation as a world leader in GaAs IC control devices. And at Penstock, we make it our business to be the leader in product availability, ease of ordering, on-time delivery and customer service. It's a combination that OEMs of any size demand today What's more, our technical support and product knowledge are unmatched in the industry.

AUTHORIZED M/A-COM DISTRIBUTOR

24 Hour Delivery

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THE ONLY FULL LINE M/A-COM DISTRIBUTOR

In addition to these products we also carry the full line of M/A-COM GaAs IC control devices, including all GaAs digital attenuators, amplifiers and mixers.

THE KNOW-HOW COMPANY

Our field support is the most highly regarded in the industry. We can give you fast expert advice, whether you're developing applications like CDPD, interactive CATV, wireless telemetry or MMDS.

ISO 9001 SUPPLIER CERTIFIED

We are certified in accordance with the highest international standards of ISO 9001.

GaAs MMIC ATTENUATORS

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.

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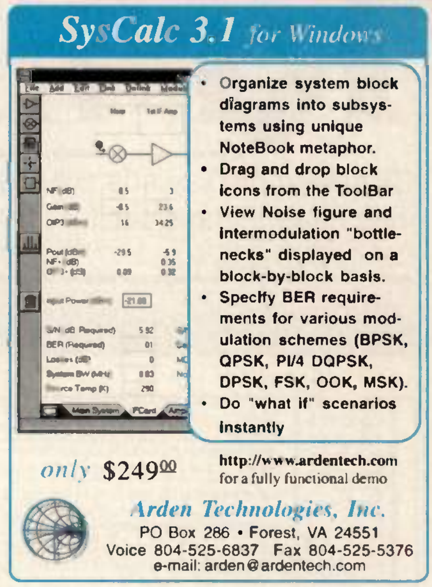

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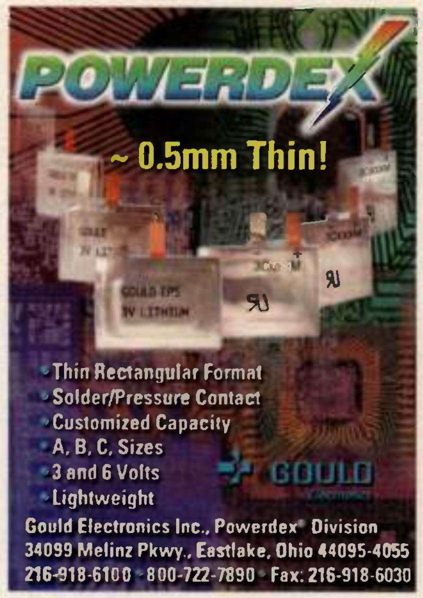

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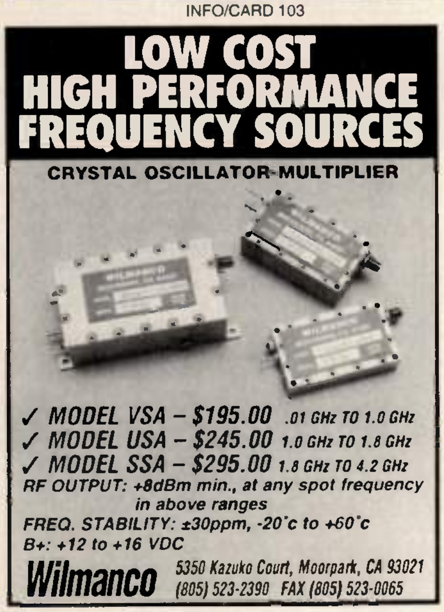

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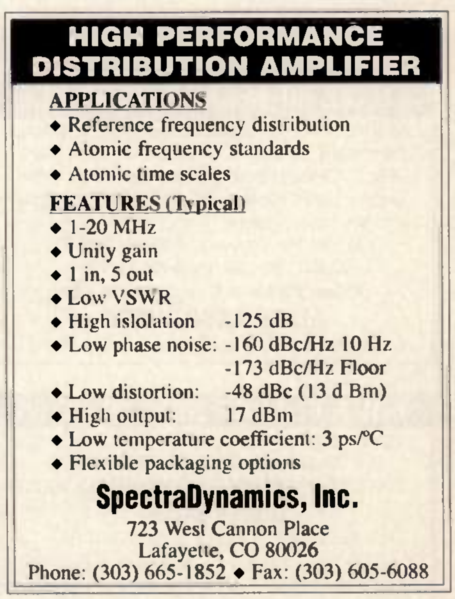

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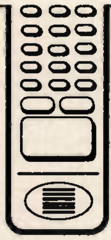

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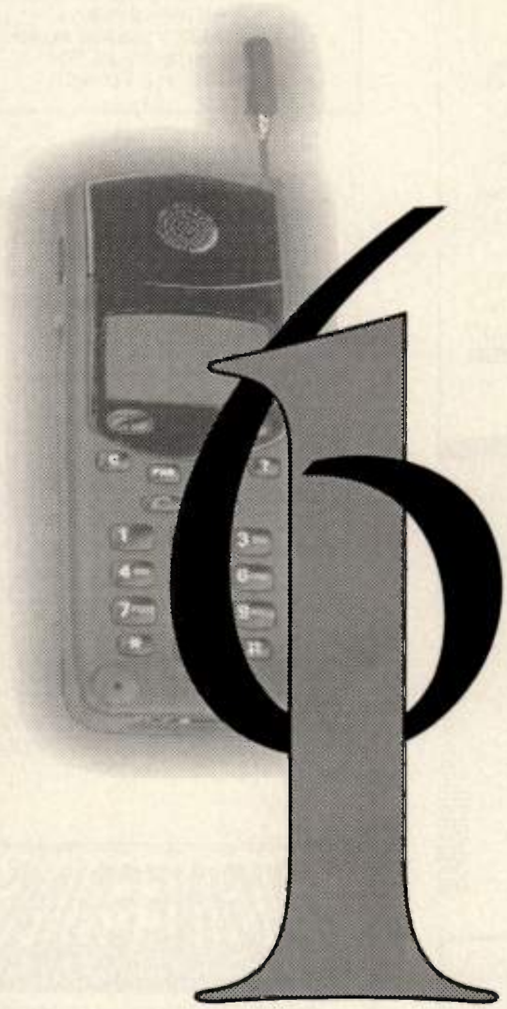

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RF guide to editorial coverage

RF advertising index

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NEW!

J-LEAD

J MIXERS UNPRECEDENTED IN VALUE.

All Ceramic 2 to 1900MHz

$4.25 from (qty. 10-49)

JMS mixers mean unparalleled performance and reliability. Solder plated J leads provide improved reliability of the solder connection while significantly reducing thermal stress and leaching. All-welded internal construction withstands reflow temperatures up to 240°C for 5 minutes, and rugged construction enables JMS mixers to pass tough MIL-M-28837 shock and vibration tests. Additionally, the all-ceramic surface mount package has a cover pull strength of 20 pounds! Other quality features include 4.5 sigma repeatability unit-to unit, automated assembly for low-cost and fast, guaranteed 1 week shipment, tape and reel availability plus a 5 year Ultra-Rel™ guarantee. If value is a must...specify JMS mixers from Mini-Circuits.

Mini-Circuits...we're redefining what VALUE is all about!

SPECIFICATIONS

MODEL	LO (dBm)	FREQUENCY (MHz) LO/RF	I F	MIDBAND (dB, typ.) Conv. Loss	Isolation L-R	L-I	$ea. (qty. 1-9)
JMS-1	+7	2-500	DC-500	5.75	45	45	4.95
JMS-1LH	+10	2-500	DC-500	5.75	55	45	8.45
JMS-1MH	+13	2-500	DC-500	5.75	60	45	9.45
JMS-1H	+17	2-500	DC-500	5.90	50	50	11.45
JMS-2L	+3	800-1000	DC-200	7.0	24	20	7.45
JMS-2	+7	20-1000	DC-1000	7.0	50	47	7.45
JMS-2LH	+10	20-1000	DC-1000	6.5	48	35	9.45
JMS-2MH	+13	20-1000	DC-1000	7.0	50	47	10.45
JMS-2H	+17	20-1000	DC-1000	7.0	50	47	12.45
JMS-2W	+7	5-1200	DC-500	6.8	60	48	7.95
JMS-11X	+7	5-1900	5-1000	6.7	35	37	4.25*

Note: *10-49 qty.

Mini-Circuits®

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For detailed specs on all Mini-Circuits products refer to • 740- pg. HANDBOOK • INTERNET • THOMAS REGISTER • MICROWAVE PRODUCT DATA DIRECTORY • EEM

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