January 2000

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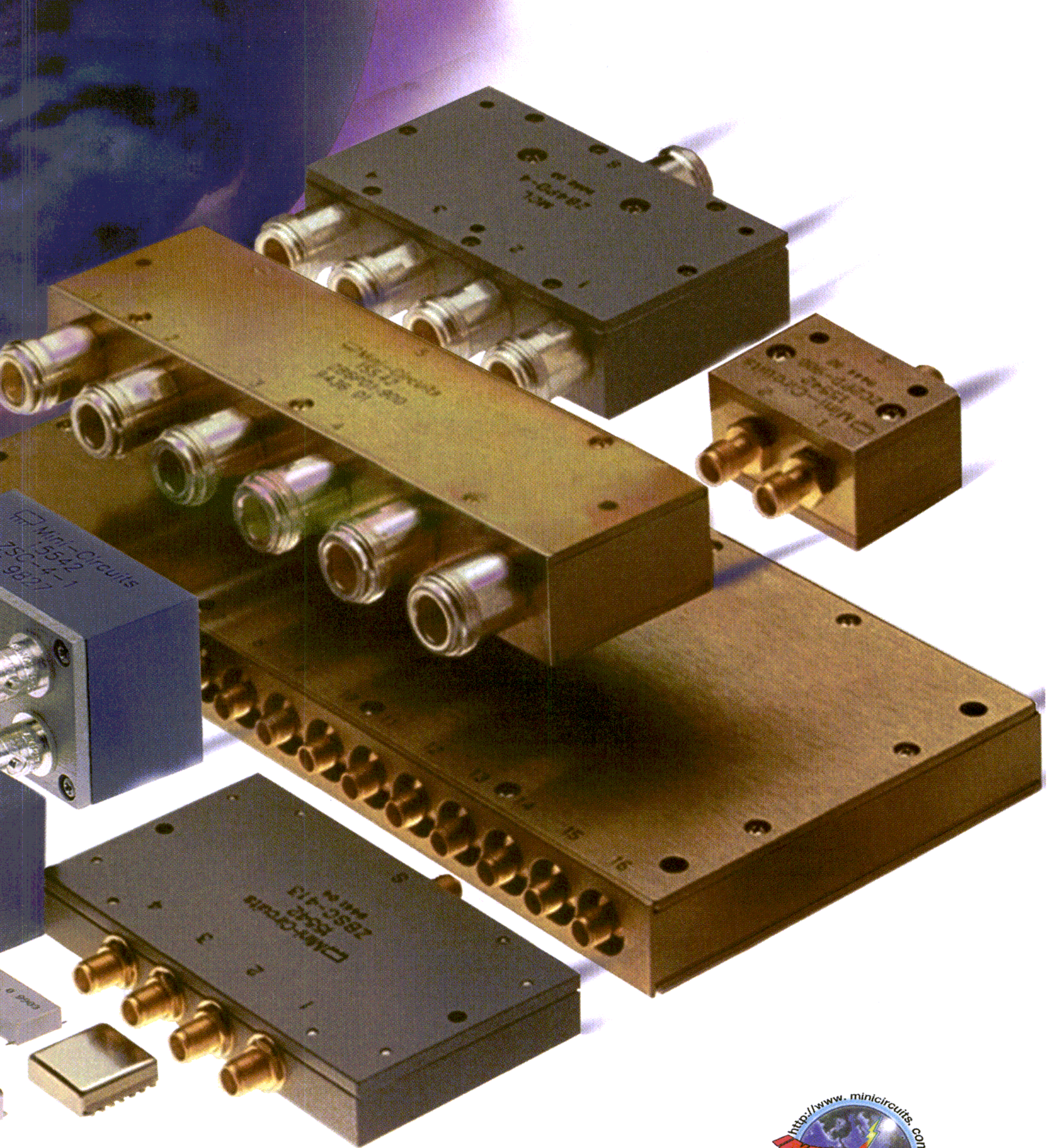
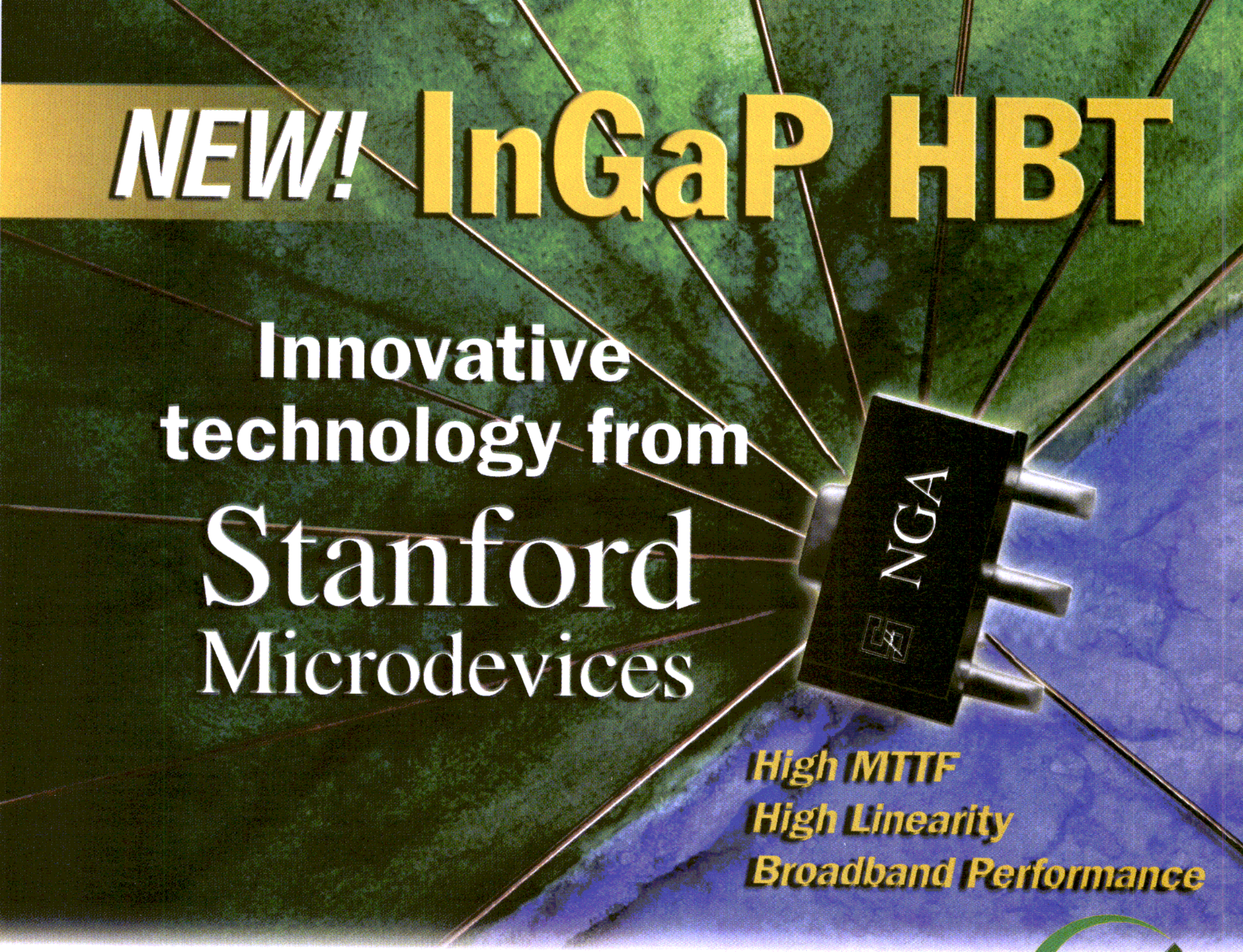
Circle 34


Mini-Circuits
15542
ZSC-4-1
ZBSC-413



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NGA-489 DC-8 GHz

Designed with InGaP process technology for greater reliability, this Darlington configured, high gain, heterojunction bipolar transistor MMIC amplifier offers value and performance for all wireless and broadband communication applications. Outstanding features are:

- Cascadable 50Ω: 1.5:1 VSWR
- Low positive voltage supply
- Low thermal resistance package
- High linearity

SPECIFICATION MATRIX

	NGA-489	NGA-589
Frequency (GHz)	DC-8.0	DC -6.0
Gain (dB)	14.5	19.0
TOIP (dBm)	38.5	38.0
N.F. (dB)	4.5	4.5
P1dB (dBm)	17.5	19.0
Supply Voltage	4.2	5.0
Supply Current	80	80

All data measured at 900MHz and is typical.
MTTF @ 150C T_j = 2 million hrs. (R_{TH} = 110 C/W typ.)

NGA-589 DC-6 GHz

High gain and high output make this heterojunction bipolar transistor MMIC amplifier ideal for use in all wireless applications. InGaP HBT technology improves the reliability and performance and minimizes leakage current between junctions. Other features include:

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- Low thermal resistance package
- High linearity
- High gain
- High P1dB

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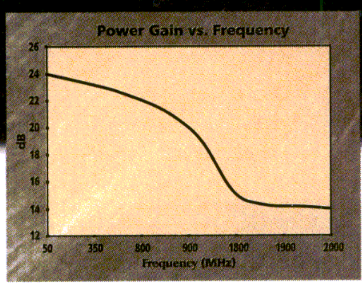
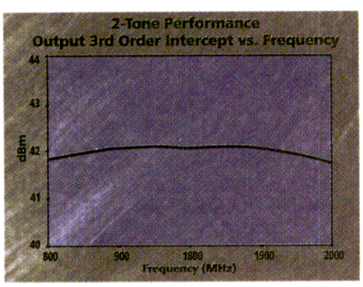
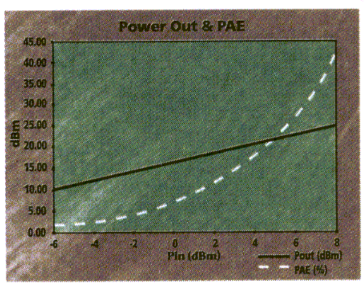
Circle 44

NEW! SXA-289 High Linearity Power Amplifier







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The **SXA-289** is a high efficiency GaAs Heterojunction Bipolar Transistor power amplifier, housed in a low cost surface-mountable plastic package and uses Darlington pair technology for broadband performance from 50-2000 MHz. It's fabricated using molecular beam epitaxial growth technology which produces reliable and consistent performance from wafer to wafer and lot to lot. The SXA-289 amplifiers were specifically designed for use as driver stages for infrastructure equipment in the 50-2000 MHz cellular, ISM and narrow band PCS bands. Its high linearity makes it an ideal choice for multi-carrier as well as digital applications.

Performance Matrix
SXA-289

Freq. (MHz)	Gain (dB) Typ.	S11 Typ.	S22 Typ.	P1dB (dBm)	TOIP (dBm)	Voltage	Current (mA) Typ.
800-900	20.0	1.5:1	2.0:1	24.0	40.5	5.0	115
1800-2000	14.5	1.5:1	2.0:1	24.0	42.0	5.0	115

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Circle 25

Applied MICROWAVE & WIRELESS

January 2000
Vol. 12 No. 1

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On Our Cover
Selectable-Gain Amplifiers Launch Integrated Subsystem Line

Cougar Components Corporation

Representing the company's move to the development and manufacture of of subsystems, three high dynamic range amplifiers with selectable gain offer high performance design options for demanding systems in communications and instrumentation.
Photo provided by Cougar Components Corporation.

TECHNICAL FEATURES

PRODUCTS & TECHNOLOGIES

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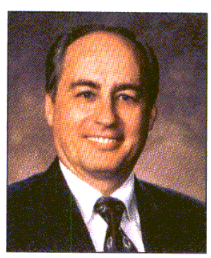
Applied MICROWAVE & WIRELESS

January 2000
Vol. 12 No. 1

PRODUCTS & TECHNOLOGIES

GUEST EDITORIAL



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AMPLIFIERS

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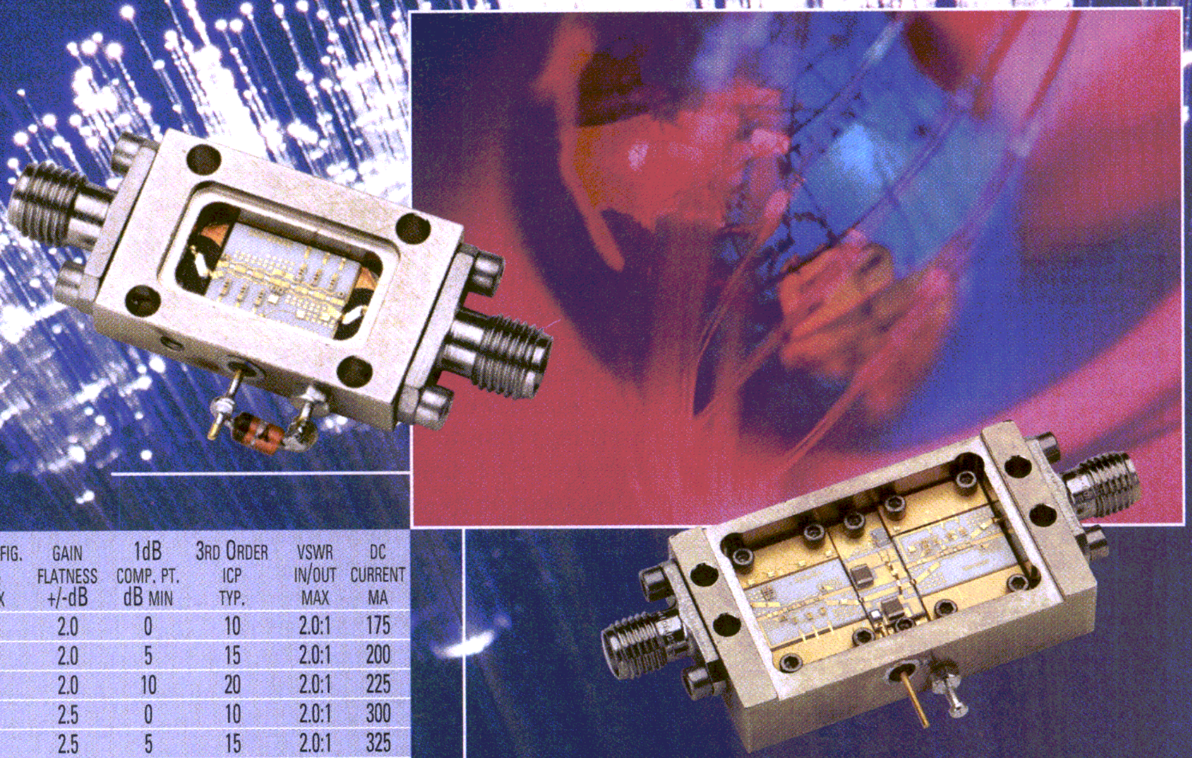

MODEL NUMBER	FREQ. RANGE GHZ	GAIN dB MIN	NOISE FIG. dB MAX	GAIN FLATNESS +/-dB	1dB COMP. PT. dB MIN	3RD ORDER ICP TYP.	VSWR IN/OUT MAX	DC CURRENT MA
JCA008-201	.01-8.0	25	*5	2.0	0	10	2.0:1	175
JCA008-202	.01-8.0	24	*5	2.0	5	15	2.0:1	200
JCA008-203	.01-8.0	22	*5	2.0	10	20	2.0:1	225
JCA008-301	.01-8.0	35	*5	2.5	0	10	2.0:1	300
JCA008-302	.01-8.0	34	*5	2.5	5	15	2.0:1	325
JCA008-303	.01-8.0	32	*5	2.5	10	20	2.0:1	350
JCA0010-201	.01-10.0	24	*5	2.0	0	10	2.0:1	175
JCA0010-202	.01-10.0	22	*5	2.0	5	15	2.0:1	200
JCA0010-203	.01-10.0	20	*5	2.0	10	20	2.0:1	225
JCA0010-301	.01-10.0	34	*5	2.5	0	10	2.0:1	300
JCA0010-302	.01-10.0	32	*5	2.5	5	15	2.0:1	325
JCA0010-303	.01-10.0	30	*5	2.5	10	20	2.0:1	350
JCA0012-201	.01-12.0	23	*5	2.0	0	10	2.0:1	175
JCA0012-202	.01-12.0	21	*5	2.0	5	15	2.0:1	200
JCA0012-203	.01-12.0	20	*5	2.0	10	20	2.0:1	225
JCA0012-301	.01-12.0	33	*5	2.5	0	10	2.0:1	300
JCA0012-302	.01-12.0	31	*5	2.5	5	15	2.0:1	325
JCA0012-303	.01-12.0	30	*5	2.5	10	20	2.0:1	350
JCA0118-201	.1-18.0	22	**5	2.5	3	13	2.0:1	200
JCA0118-202	.1-18.0	20	**5	2.5	5	15	2.0:1	250
JCA0118-203	.1-18.0	20	**5	2.5	7	17	2.0:1	300
JCA0118-301	.1-18.0	31	**5	2.5	3	13	2.0:1	250
JCA0118-302	.1-18.0	29	**5	2.5	5	15	2.0:1	300
JCA0118-303	.1-18.0	29	**5	2.5	7	17	2.0:1	350

* Noise Figure is specified above 300 Mhz
** Noise Figure is specified above 500 Mhz

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Circle 49

Editorial

Thinking Well Beyond the Year 2000

By Gary A. Breed
Publisher

It's January 1, 2010 — What wireless technologies are you using? What new technologies are you working on for your company's next major product release? What components do you use? What tools help you get the design finished, tested and manufactured?

I didn't think it was ambitious enough to just look ahead to the first year of the new millennium! We need to see as far into the future as possible. A few years ago, we finally reached the development of wireless technology where we could say, "We can build more things than we know what to do with." There are more potential applications for wireless technology than there are dollars in the economy to buy them.

The challenge is to identify the most valuable, the most useful ways to put radio waves to work for us. What will ten years' evolution do for LMDS' and cable's ambitious bandwidth-to-the-home efforts? Will fully interactive video and high speed data drive that market, or will entertainment continue to be the primary use of communications bandwidth? Will cellular technology expand further and grow in capability, or will it stabilize like the terrestrial phone system did for many years?

Will we continue to have diverse, individual solutions to specific communications needs, or will more and more services be integrated into a giant Internet as the medium through which most communication takes place?

How will society respond to faster, cheaper, and universal communications? We are already overloaded with information, so it will be interesting to see whether we adapt to accommodate the avalanche of data, or whether some kind of backlash arises as portrayed in science fiction stories. It's fascinating (and scary) to think about all the possible ramifications of major social change. As I've noted before, we are clearly in the midst of changes every bit as dramatic as the Industrial Revolution of the 1800s.

Maybe communications will move to the background of high tech in five years. Biology and medicine are on their way to becoming breakthrough technologies that are at least as far-reaching as recent communications developments. Materials science has also been moving ahead rapidly, starting with high temperature superconductors, then advanced ceramics and, more recently, molecular-scale tools, actuators and sensors.

But don't worry about your job; wireless will still be here in a big way. ■

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Circle 20

Comments on scientific integrity and the disputed calcium studies

Editor:

I read your editorial regarding "Fraudulent Science" in the November issue with great interest. I fully share your concern that fake results in one or two studies can sometimes tarnish the image of the whole area. Recently, I reported on this issue in my column in the October *IEEE Antennas & Propagation Magazine,* and I thought the following comments would be complementary to your editorial.

As far as the mainstream media were concerned, the hot news this summer was the allegation that Dr. Robert Liburdy, employed until recently as a cell biologist by the Lawrence Berkeley Laboratory in Berkeley, California, "faked what has been considered crucial evidence of a tie between electromagnetic radiation and cancer." (*The New York Times*, July 24, 1999). The federal Office of Research Integrity (ORI) declared that Dr. Liburdy had committed scientific misconduct by "intentionally falsifying and fabricating data" in two 1992 papers about EMF effects on calcium changes in rat blood cells.

In a voluntary settlement with ORI, Dr. Liburdy agreed to retract the three disputed graphs and not to receive any federal grant money for the next three years. However, in a letter published in *Science* (July 16, 1999), Dr. Liburdy claimed,

"The raw data for my two calcium studies are valid. Thus, these papers are not being retracted, and my scientific conclusions stand as published. I admit no wrongdoing. I could not afford a protracted legal battle with the federal Office of Research Integrity (ORI), and a settlement was reached in which I admit no liability. … My error was in not describing these procedures [baseline adjustment, normalization, etc.] in the methods section… There was no intent to deceive."

However, the ORI lawyers argue (and as the *Applied Microwave & Wireless* editorial emphasized) that "Liburdy simply chose data points to express his desired experimental outcome." On the other hand, it should be noted that the ORI does not challenge any of Dr. Liburdy's other work, including his studies purporting to show that environmental magnetic fields can block tamoxifen and melatonin action in human breast cancer. According to Dr. Liburdy, "Four independent replications of these findings have been reported at scientific meetings in 1998 and 1999."

Rajeev Bansal
University of Connecticut

Send your letters to "Editor" at the mailing or e-mail address on page 8.

Calendar

CONFERENCES

FEBRUARY

February 15-17, 2000
IZS 2000 — 2000 International Zurich Seminar on Broadband Communications
Zurich, Switzerland
Information: Professor R. Vahldieck
Tel: +411 632 27 42; Fax: +411 632 11 98
E-mail: vahldieck@ifh.ee.ethz.ch

MARCH

March 20-25, 2000
Hannover Fair 2000
Hannover, Germany
Information: Hannover Fairs USA Inc.
Tel: 609-987-1202; Fax: 609-987-0092
E-mail: info@hfusa.com
Internet: http://www.hfusa.com

March 28-30, 2000
RTS 2000
Paris, France
Information: BRIP
Tel: + 33 01 53 17 11 44
E-mail: crey@birp.fr
Internet: http://www.birp.com/rts

APRIL

April 8-13, 2000
NAB2000
Las Vegas, NV
Information: NAB
Internet: http://www.nab.org.conventions

April 10-11, 2000
2000 IEEE Emerging Technologies Symposium on Broadband Wireless Internet Access
Dallas, TX
Information: Dr. Jon Velhl
Tel: 972-952-4190
Internet: http://www.ieeedallas-els.org

MAY

May 1-4, 2000
2000 GaAs MANTECH Conference International Conference on Gallium-Arsenide Manufacturing Technology
Washington, DC
Information: Wes Mickanin
Tel: 503-615-9253; Fax: 503-615-8903
E-mail:wem@tqs.com
Internet: http://www.gaasmantech.org

May 20-26, 2000
IEEE International Conference on Phased Array Systems and Technology
Dana Point, CA
Information: Dr. Michael Thorburn
Tel: 310-336-2197; Fax: 310-336-6225
E-mail: m.a.thorburn@IEEE.org
Internet: http://www.ieee.org

JUNE

June 7-9, 2000
2000 IEEE/EIA International Frequency Control Symposium and Exhibition
Kansas City, MO
Information: IEEE Ultrasonics, Ferroelectrics and Frequency Control Society
Internet: http://www.ieee.org/uffc/fc

June 11-13, 2000
2000 IEEE Radio Frequency Integrated Circuits Symposium
Boston, MA
Information: Jyoti Mondal
Tel: 847-259-9600, ext. 4130
E-mail: mondajy@mail.northgrum.com
Internet: http://www.ims2000.org/rfic.htm

June 11-16, 2000
MTT-S International Microwave Symposium
Boston, MA
Information: LRW Associates
Tel: 704-841-1915; Fax: 704-845-3078
E-mail: lrwassoc@sprintmail.com
Internet: http://www.ims2000.org

June 15-16, 2000
Automatic RF Techniques Group 55th Conference
Boston, MA
Information: D. Michael Fennelly
Tel: 978-258-4101; Fax: 978-258-4102
E-mail: m.fennelly@ieee.org
Internet: http://www.arftg.org

June 18-20, 2000
IMAPS Europe Prague 2000 — European Microelectronics, Packaging and Interconnection Symposium
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Calendar

SHORT COURSES

University of California at Los Angeles Extension

Digital Signal Processing: Theory, Algorithms, and Implementation
Los Angeles, CA . . . . .January 10-14, 2000
Introduction to Software Radio and Wireless Networks
Los Angeles, CA . . . . .January 24-26, 2000
Digital Avionics Systems
Los Angeles, CA . . . . .Feb. 28-March 3, 2000
Kalman Filtering
Los Angeles, CA . . . . .March 13-16, 2000
Information: UCLA Extension, Short Course Program Office, Tel: 310-825-3344; Fax: 310-206-2815.

RTT Programmes Limited

RF Design
London, England . . . . . . .January 17-19, 2000
Cordless Design
London, England . . . . . . .January 24-26, 2000
Cellular Design
London, England . . . . . . .February 7-9, 2000
RF Amplifier Design
London, England . . . . . . .February 14-16, 2000
RF Oscillator Design
London, England . . . . . . .March 13-15, 2000
RF/IF Processing
London, England . . . . . . .March 20-22, 2000
Information: Lorraine Gannon, Tel: +44 181 844 1811; Fax: +44 181 751 2616; E-mail: seminars@rttsys.com; Internet: www.rttsys.com

Georgia Institute of Technology

Low Cost Flip Chip Processing and Analysis with Hands-on Application
Atlanta, GA . . . . .January 19-20, 2000
Information: Georgia Tech Distance Learning, Continuing Education and Outreach, Tel: 404-894-2547; Fax: 404-894-7398; E-mail: conted@gatech.edu.

Besser Associates

Digital RF Communication Systems
Mountain View, CA . . . . . .January 24-28, 2000
RF Component Modeling
Mountain View, CA . . . . . .Jan. 31-Feb. 4, 2000
Wireless Measurements: Theory and Practice
Mountain View, CA . . . . . .February 14-18, 2000
Signal Integrity of High-Speed Digital Design
Mountain View, CA . . . . . .February 28-29, 2000
Multitone Amplifier Design
Mountain View, CA . . . . . .March 2-3, 2000
Introduction to RF Transceivers and Systems Components
Mountain View, CA . . . . . .March 6-7, 2000
RF Test Equipment Operation (laboratory course)
Mountain View, CA . . . . . .March 8, 2000
RF Testing for the Wireless Age (laboratory course)
Mountain View, CA . . . . . .March 9-10, 2000
Behavioral Modeling
Mountain View, CA . . . . . .March 13-15, 2000
RFIC Techniques for Wireless Applications
Mountain View, CA . . . . . .March 20-22, 2000
EMC/EMI and Thermal Issues for Electronic Packages and Systems
Mountain View, CA . . . . . .March 23, 2000
Applied RF Techniques II: Nonlinear RF and Wireless Circuit Design
Phoenix, AZ . . . . .March 26-31, 2000
Applied RF Techniques I
Phoenix, AZ . . . . .March 27-31, 2000
Information: Annie Wong, Tel: 415-949-3300; Fax: 415-949-4400; E-mail: info@bessercourse.com; Internet: www.bessercourse.com/

University of Wisconsin at Milwaukee

EMC Printed Circuit Board Design
Tampa, FL . . . . .Feb. 28-March 1, 2000
Information: Mark Schmidt, Program Assistant, Tel: 1-888-545-4700; Fax: 1-888-545-4600; E-mail: dschmidt@uwm.edu; Internet: www.uwm.edu/dept/ccee

Applied Technology Institute

Fundamentals of Radar Technology
Washington, DC . . . . .February 8-10, 2000
Radar Systems Design and Engineering
Washington, DC . . . . .February 15-17, 2000
Mathematical and Physical Wavelets
Newport, RI . . . . .March 13-16, 2000
Satellite Communication Systems Engineering
Los Angeles, CA . . . . .March 21-23, 2000
Information: ATI Defense Systems Training, Tel: 1-888-501-2100 or 410-531-6034; Fax: 410-531-1013; E-mail: ATIcourses@aol.com; Internet: www.ATIcourses.com

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Niagra Falls, NY . . . . . .March 20-21, 2000
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Niagra Falls, NY . . . . . .March 22-24, 2000
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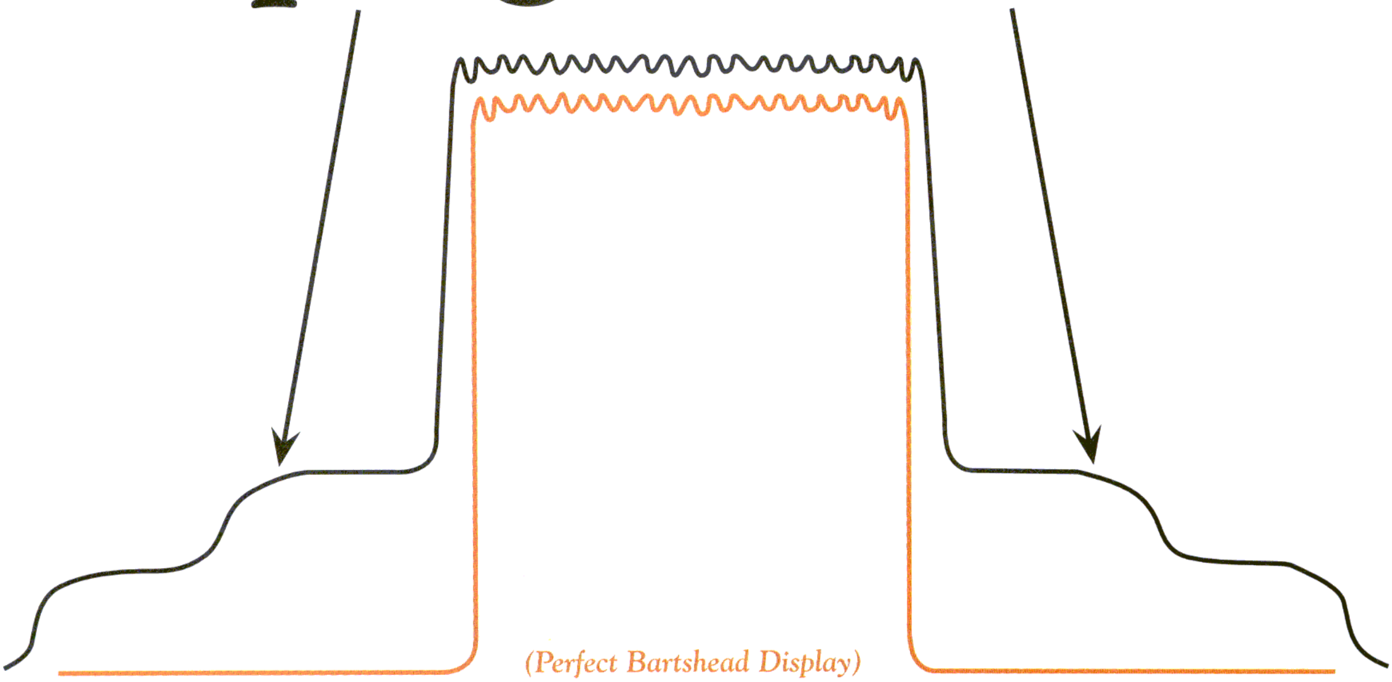
Danger. Sloping Shoulders.



Spectral growth can compromise the most precise CDMA technologies. Often, the difficulty lies in the transmission equipment, a potent source of distortion or the "shoulders" you see on your display. Now there are amplifiers that help you eliminate shoulders, keep the space between carriers narrow and optimize bandwidth.

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Circle 40

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E-mail: rumelhard@cnam.fr

Deadline: March 17, 2000

GAAS 2000 — European Gallium Arsenide and Other Semiconductors Application Symposium

October 2-3, 2000 — Paris, France

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Authors should submit an author information form (call or e-mail to receive a form) and either two hard copies or an electronic file (.doc or .pdf) of a summary for each paper. Send to:

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Deadline: March 17, 2000

Report: Global RF semiconductor market to nearly double by 2004

The worldwide market for radio frequency (RF) semiconductor devices in cellular telephones will reach $7.7 billion in 2004, up from $3.9 billion in 1999, according to a report from Strategies Unlimited, a Silicon Valley-based market research firm.

The report, RF Semiconductors for Cellular/PCS Handsets, Market Review and Forecast 2000, examines the semiconductor technology and trends behind the evolution to third-generation (3G) cellular chipsets. It projects worldwide demand for handsets through the year 2004 and the accompanying RF semiconductor market.

The number of cellular and PCS subscribers is forecast to reach 1.3 billion in 2004.

Led by the strong global demand for mobile communications, the number of cellular and PCS subscribers is forecast to reach 1.3 billion in 2004. Annual handset demand is projected to grow from 240 million units in 1999 to 600 million in 2004. New services such as instant messaging, wireless data and Internet access will provide continued momentum for strong growth over the next five years.

According to the report, GSM handsets are the largest market for RF semiconductors with device shipments of $2 billion in 1999, followed by CDMA, IS-136 TDMA, PDC and analog chipsets with $1.9 billion. Small signal amplifiers, frequency conversion devices and IF chipsets represented the largest part of the total market by chip type. Power amplifier and RF control chips also showed strong growth over 1998 levels.

Leading silicon bipolar technology companies such as Infineon, Motorola, NEC and Philips provide the majority of the silicon RF chips used in the handset market, the report says. Gallium arsenide (GaAs) chips made up 33 percent of the market in 1999, with strong growth in heterojunction bipolar transistor (HBT) sales by Conexant and RF Micro Devices. Other GaAs suppliers, including Alpha, Anadigics, Infineon, Motorola and TriQuint, demonstrated solid performance in 1999.

Over the next five years, increased integration of RF and IF circuits will reduce the number of chips per handset dramatically, with CMOS, BiCMOS and silicon germanium (SiGe) ICs playing the major roles. GaAs ICs will continue to lead in power amplifier and switch sockets. Bluetooth data links and GPS receivers for location-based services will appear beginning in 2000, leading to a $500 million chip market in 2004.

Motorola Labs announces world's thinnest transistor

Motorola Labs, based in Tempe, AZ, has built the world s thinnest functional transistor using a new class of semiconductor materials.

The company has built a working device that uses perovskite, a class of crystalline oxide materials with unique material properties. The new technology enables the development of a transistor with an effective thickness that is initially three to four times thinner than those built with conventional semiconductor materials.

While this technology is still at an early stage of development, it has already produced working devices that are electronically much thinner than those made with existing technology. The advance should allow computer consumption to continue to be significantly reduced in size and power consumption.

Motorola, based in Schaumburg, IL, provides semiconductors, integrated communications solutions, embedded electronic systems and components.

Omnipoint testing Ericsson GPRS equipment

GSM service provider Omnipoint has begun the first U.S. trials of high-speed wireless data technology using Ericsson s General Packet Radio Services (GPRS) solution. The new technology, being offered to customers in the New York, NY, metropolitan area, provides better wireless access to the Internet as well as a wide range of other IP-based services at speeds to reach 115 kbps, up to 12 times faster than current Omnipoint data rates and more than double the average speed of most dial-up modems today.

Omnipoint, a subsidiary of Omnipoint Corporation based in Cedar Knolls, NJ, provides advanced wireless communications services in more than a dozen states. Ericsson, based in Stockholm, Sweden, provides communications products and services worldwide.

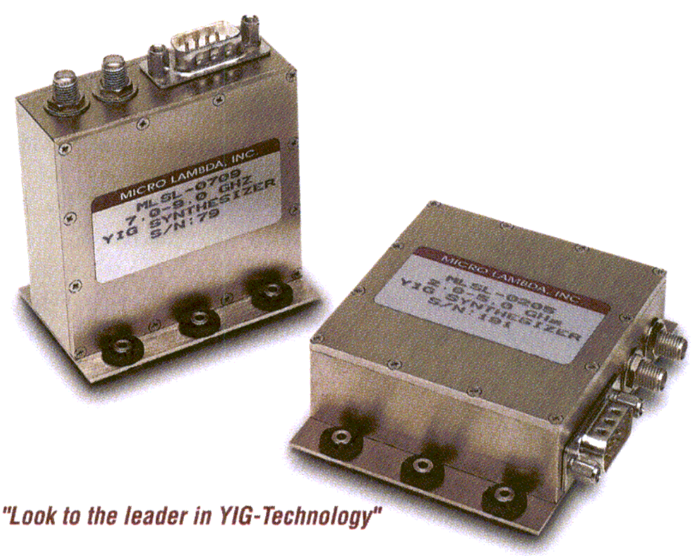
CTS establishes RF integrated modules unit

CTS Corporation of Elkhart, IN, has created a new business group, RF Integrated Modules, as part of its Wireless Group. The division will develop high frequency modules to be used in cellular handset phones.

CTS manufactures electronic components and custom electronic assemblies for OEM customers primarily in the communications equipment, automotive and computer equipment markets worldwide.

Companies, organizations and institutions may submit information for our News section to: Shannon O Connor, Managing Editor, Applied Microwave & Wireless, 4772 Stone Drive, Tucker, GA, 30084; Fax: (770) 939-0157; E-mail: amw@amwireless.com.

Low Power, Low Noise YIG-Based Synthesizers *for Digital Radios*



"Look to the leader in YIG-Technology"

MICRO LAMBDA, INC.

Micro Lambda, Inc. a leader in the development of next-generation YIG devices now offers YIG-Based Frequency Synthesizers covering the 2-12 GHz frequency range. Designed specifically for Digital Radio ODU's and harsh commercial environments, these synthesizers offer excellent integrated phase noise characteristics over carrier offset frequencies from 10 kHz to 10 MHz.

Tunable bandwidths of either 2 GHz or 3 GHz are available as standard products. This results in fewer numbers of synthesized sources required for a variety of Digital Radio frequency plans. Millimeter-Wave frequencies can easily be obtained using frequency multipliers to obtain output frequencies between 24 GHz through 44 GHz.

Applications include QAM and QPSK modulated Digital Radio's and a multitude of general purpose applications.

Features

- 2 - 12 GHz Frequency Coverage
- Excellent Integrated Phase Noise Characteristics
- Compact Size
- 3-Line Serial Interface
- Low Prime Power
- 500 kHz Step Size
- Internal Memory (last frequency programmed - recall)

MLSL-Series Synthesizers

This series of synthesizers utilize an external 1 to 50 MHz crystal reference oscillator to generate tunable frequencies covering the 2 - 12 GHz range. Output power levels of +12 dBm to +15 dBm are offered depending on frequency, with a standard tuning step size of 500 kHz. Input tuning commands are via 3-Line Serial interface. The size of these compact units is 2.5" x 2.5" x 1.0" without mounting plate and consume less than 6 watts of prime power. The units have an internal memory capability which "recalls" the last frequency programmed when the prime power is removed and reapplied. Standard models include 2-4 GHz, 4-6 GHz, 5-7 GHz, 7-9 GHz and 9-11 GHz. Specialized frequency ranges are easily implemented utilizing the versatile synthesizer architecture.

48041 Fremont Blvd. Fremont, CA 94538 (510)770-9221 mcrolambda@aol.com www.micro-lambda.com

Circle 79

BUSINESS AND FINANCE

Motorola NSS receives $228 million in new contracts

Motorola Inc. s Network Solutions Sector (NSS) has received six contracts from Unicom and China Mobile to upgrade the companies GSM networks in China. The contracts are worth a total of $228 million.

Five of the contracts, totaling $163 million, were awarded by China Unicom for the expansion of the GSM900 networks in the provinces of Jiangsu, Guangdong, Fujian, Shandong and Xinjiang. The work will increase the capacity of the networks by 1.65 million subscribers. The contract awarded by China Mobile, worth $65 million, is for GSM network expansion in the Sichuan province and will increase the capacity of that network by 700,000 subscribers.

Motorola, based in Schaumburg, IL, provides semiconductors, integrated communications solutions, embedded electronic systems and components.

Conexant to acquire Maker Communications

Conexant Systems Inc. has announced that it is acquiring Maker Communications of Framingham, MA, a provider of programmable, high-performance network processors, software solutions and development tools. The transaction is valued at $990 million.

Conexant, based in Newport Beach, CA, manufactures components, subsystems and semiconductor system solutions for digital wireless voice and data communications.

Andrew acquires Iowa-based Conifer

Andrew Corporation has annouced the acquisition of Conifer Corporation, which manufactures Multichannel Multipoint Distribution Service (MMDS) subscriber products, wireless LAN equipment, and Direct Broadcast Satellite (DBS) accessories. Terms were not disclosed.

Andrew, based in Orland Park, IL, supplies communications systems equipment and services. Conifer is based in Burlington, IA.

L-3 announces acquisition, contracts

L-3 Communications has signed an agreement to purchase the Teterboro, NJ-based Space and Navigation Systems business of the former AlliedSignal Inc., in a transaction valued at $55 million. Space and Navigation Systems provides navigation products for weapons systems and satellites and related systems.

L-3 also announced three new contracts for three of its divisions. L-3 s Power Systems Group, a unit of its SPD Technologies division, has been selected as the supplier of electric plant power-conversion equipment to Bechtel Plant Machinery Inc. for use in the first Virginia class attack submarine for the U.S. Navy. The value of the contract was not disclosed.

L-3 s Display Systems division has received a contract from Naval Air Systems Command (NAVAIR) to provide additional Programmable Tactical Information Displays (PTID) for the F-14D Tomcat aircraft. The contract modification is valued at $11.2 million.

L-3 s Ocean Systems division has signed a contract to supply ten Helicopter Long Range Active Sonars (HELRAS) to Agusta, a Finmeccanica Company. The contract, which includes logistics support items, is valued at more than $30 million.

L-3 Communications supplies secure communication systems and products, microwave components, avionics and ocean systems and telemetry, instrumentation, space and wireless products.

Cadence to acquire Diablo Research

San Jose, CA-based Cadence Design Systems Inc. has signed an agreement to acquire Diablo Research Company LLC, an electronics design services company based in Sunnyvale, CA. Terms were not disclosed.

Cadence supplies electronic design automation products and services for the design of semiconductors, computer systems, networking and telecommunications equipment and other electronics products.

SEZ to supply spin processing equipment

The SEZ Group has received an $18.2 million order from Nanya Technologies Corp., a microchip company based in Taiwan, to supply its semiconductor spin-processing equipment.

SEZ, based in Villach, Austria, manufactures semiconductor spin-processing equipment for the semiconductor chip manufacturing industry.

Micronetics receives microwave subassembly order

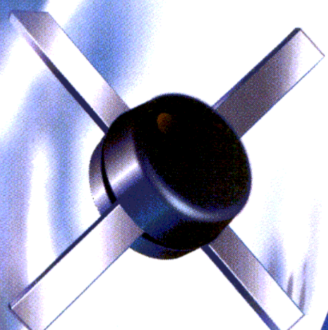
Micronetics Wireless Inc. has received the first phase of a U.S. Government contract for microwave integrated assemblies, switch/filter banks and PIN diode switches. The contract was obtained by Micronetics wholly-owned subsidiary, Microwave & Video Systems Inc. (MVS), of Danbury, CT. Terms were not disclosed.

Based in Hudson, NH, Micronetics Wireless manufactures broadband test equipment and components, as well as microwave components and subassemblies.

Atmel to acquire Thomson-CSF subsidiary

Atmel Corp, based in San Jose, CA, has agreed to acquire Thomson-CSF Semiconducteurs Specifique (TCS), a wholly owned subsidiary of Thomson-CSF. Terms were not disclosed.

TCS manufactures specific integrated circuits, including CCD and CMOS image sensors and RF ASICs. Atmel manufactures advanced logic, mixed-signal, nonvolatile memory and RF semiconductors.



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ERA-1	DC-8000	11.8	11.7	5.3	26.0	40	1.80
ERA-1SM	DC-8000	11.8	11.3	5.5	26.0	40	1.85
ERA-2	DC-6000	15.6	12.8	4.7	26.0	40	1.95
ERA-2SM	DC-6000	15.2	12.4	4.6	26.0	40	2.00
ERA-3	DC-3000	20.8	12.1	3.8	23.0	35	2.10
ERA-3SM	DC-3000	20.2	11.5	3.8	23.0	35	2.15
ERA-4	DC-4000	13.5	▲17.0	5.5	▲32.5	65	4.15
ERA-4SM	DC-4000	13.5	▲16.8	5.2	▲33.0	65	4.20
ERA-5	DC-4000	18.8	▲18.4	4.5	▲33.0	65	4.15
ERA-5SM	DC-4000	18.5	▲18.4	4.3	▲32.5	65	4.20
ERA-6	DC-4000	11.3	▲18.5	8.4	▲36.5	70	4.15
ERA-6SM	DC-4000	11.3	▲17.9	8.4	▲36.0	70	4.20

Note: Specs typical at 2GHz, 25°C. Exception: ▲ indicates typ. numbers tested at 1GHz.
* Low frequency cutoff determined by external coupling capacitors.
① Price (ea.) Qty.1000: ERA-1 $1.19, -2 $1.33, -3 $1.48, -4, -5 or -6 $2.95. SM option same price.

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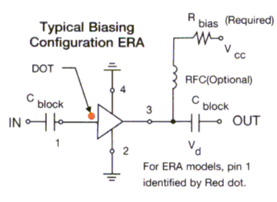
K3-ERASM: 10 each ERA-4SM,-5SM,-6SM (30 pieces) only $99.95

Chip Coupling Capacitors at 12¢ each (50 min.)

Size (mils)	Value
80x50	10, 22, 47, 68, 100, 220, 470, 680, 1000, 2200, 4700, 6800, 10,000 pf
120x60	.002, .047, .068, .1 µf

ERA-1 ERA-1SM
ACTUAL SIZE

Typical Biasing Configuration ERA



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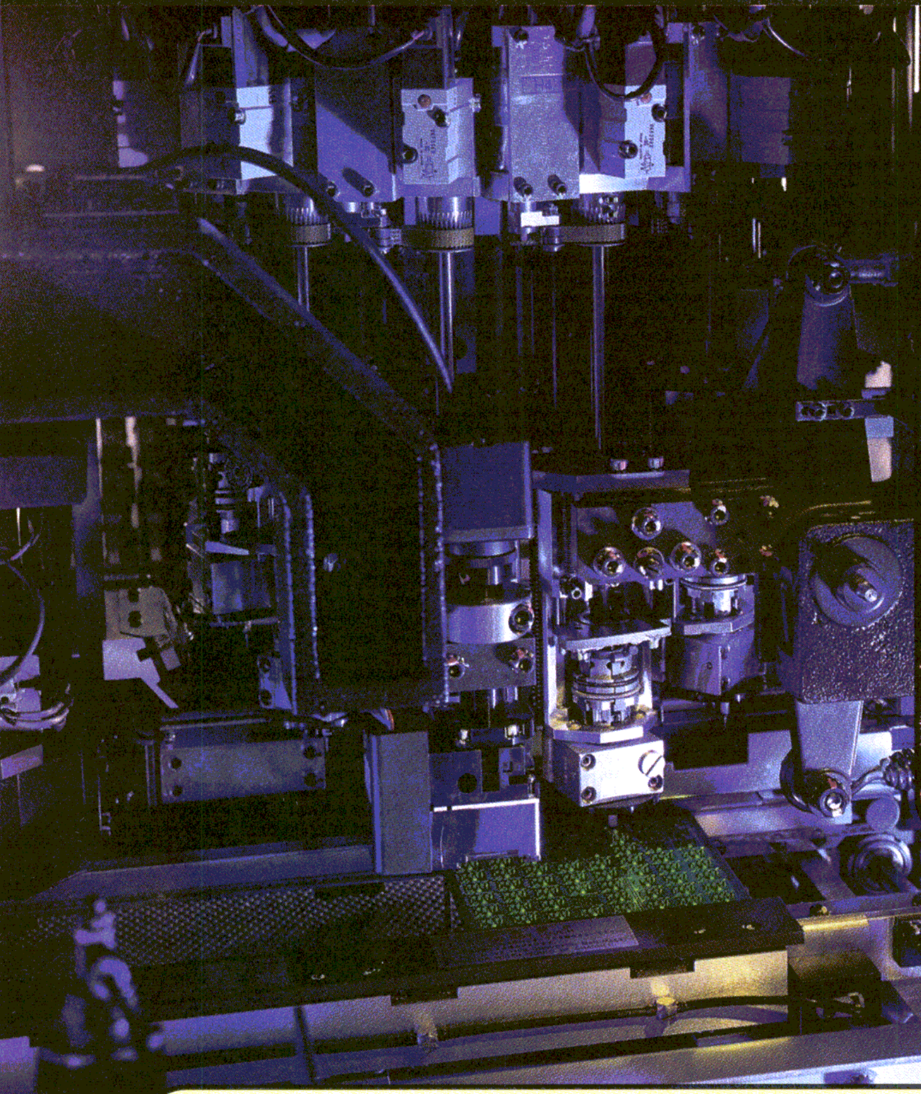

Everyday at Vari-L innovative engineering designs are coupled with state-of-the-art manufacturing technologies to create high performance, cost-effective RF and microwave solutions for the world's largest wireless equipment companies.

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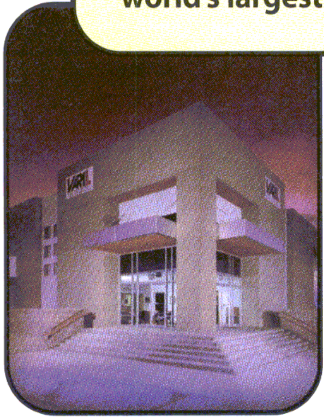
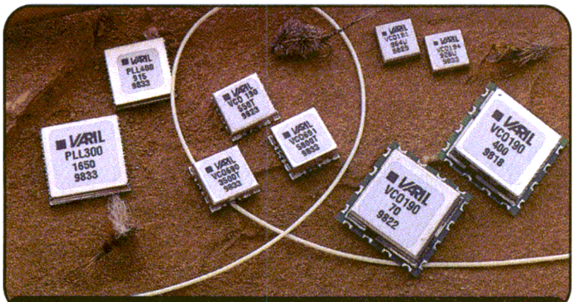


At Vari-L, we are dedicated to research and development. Our exceptional commitment to innovation has resulted in eight new patents in the last three years alone. Since 1953, our products have been utilized in many diversified and demanding applications ranging from satellite and aerospace systems, CATV and fiber optic transmission systems to wireless base stations and handsets.

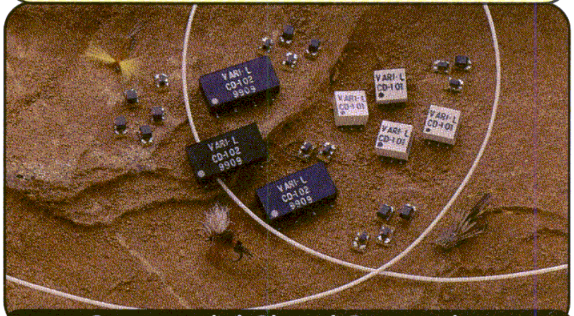
Our intensive investment in state-of-the-art manufacturing technology gives us the ability to offer low cost product realization through automated assembly, handling and testing. Recent expansions in our automated production systems have created a projected Year 2000 manufacturing capacity in excess of 12 million units housed within Vari-L's 80,000-square foot facility in Denver, Colorado.

Over 90% of our shipments are custom solutions tailored specifically to a customer's RF/microwave signal sources or signal processing project requirement. At Vari-L, our continued focus on research and development and commitment to enhancing manufacturing technology is intended to ensure that "We have a part in your future."



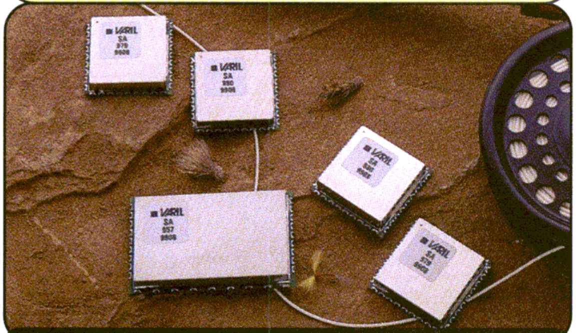
Commercial Signal Sources

- High Performance Voltage Controlled Oscillator Modules
- High Performance PLL Synthesizer Modules



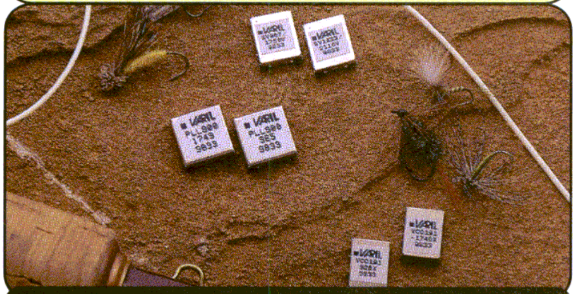
Commercial Signal Processing

- High Performance Wideband RF Transformers
- High Performance Power Dividers and Couplers
- High Performance Double Balanced Mixers
- High Performance RF Chokes
- High Performance Bias Tees



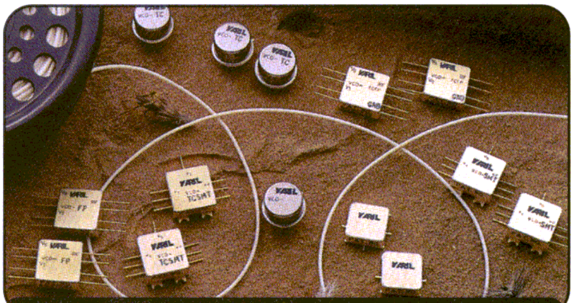
Commercial Special Assemblies

- Special Frequency Conversion Modules
- Special Frequency Generation Modules



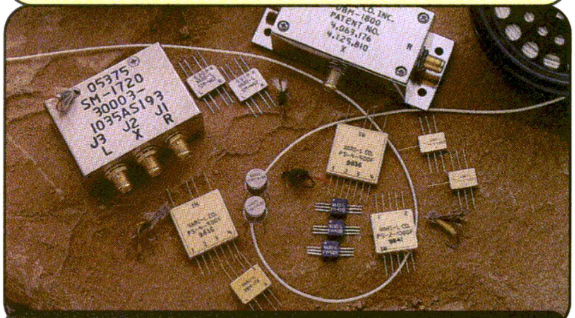
Subscriber Signal Sources

- Miniature Voltage Controlled Oscillator Modules
- Miniature PLL Synthesizer Modules



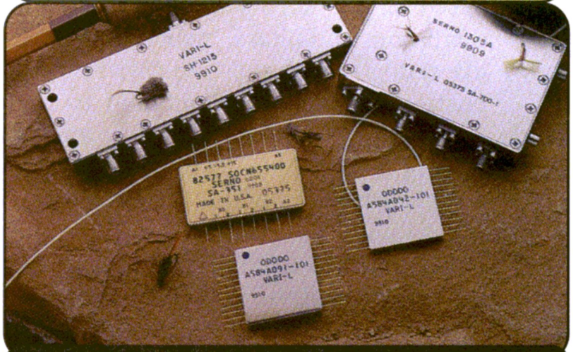
Military Signal Sources

- Ruggedized High Performance Hybrid
- Voltage Controlled Oscillators



Military Signal Processing

- Ruggedized Double Balanced Mixers
- Ruggedized Wideband RF Transformers
- Ruggedized Power Dividers and Couplers
- Ruggedized I/Q Modulators and Demodulators



Military Special Assemblies

- Ruggedized Special Frequency Generation Assemblies
- Ruggedized Special Frequency Conversion Assemblies
- Ruggedized Special RF Distribution Assemblies

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Circle 64

Wideband Low Phase Noise Push-Push VCO

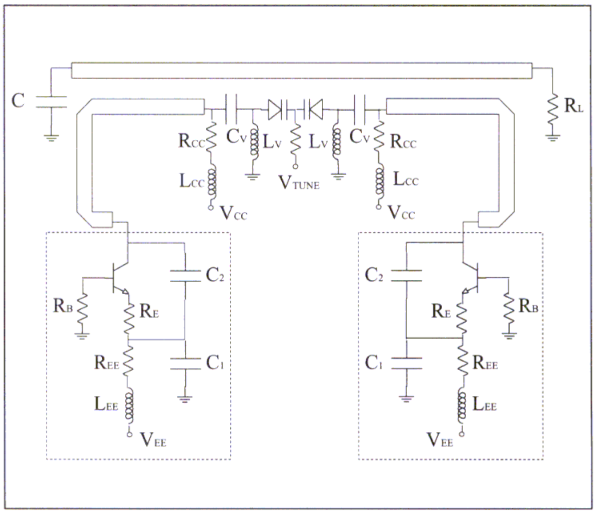
Push-push topology reduces noise compared to a conventional design

By Marco Gris
Alcatel Telecom

Many papers have shown the effectiveness of the push-push configuration in the reduction of oscillator phase noise and attempted to theoretically explain its behavior. This paper proposes an explanation based on the theory of phase noise in coupled oscillators and phase opposition and expects a 9 dB reduction of noise-to-carrier ratio for the push-push oscillator compared to that for the single oscillator. The simulation and experimental implementation of a wideband, low phase noise push-push VCO is also presented.

The push-push oscillator

A push-push oscillator consists of two identical oscillators that share one common resonator and oscillate 180 degrees out of phase with each other. According to existing literature, the push-push configuration can be used to realize a VCO characterized by a very wide tuning band, maintaining low phase noise levels. If we have a set of N identical oscillators and their output voltages are summed in phase, then the total carrier power is multiplied by N^2. However, the noise power increases by N if we assume that noise sources of different oscillators are independent [1]. Thus the phase noise relative to the carrier decreases by a factor of N. Theory of coupled oscillators shows that phase noise can be improved [2]. According to Chang et al., the output phase noise spectral density $S_\Phi(\omega)$ produced by two bilaterally coupled identical oscillators is 3 dB lower than the single oscillator. If the signals generated by two coupled oscillators in phase opposition are subtracted, the amplitude of the sinusoidal waveform generated by the push-push oscillator is doubled compared to the one generated by one single oscillator, and the output power is 6 dB higher. The noise-to-carrier ratio $L(f_m)$ for the push-push is then 9 dB lower than the single oscillator. This theoretical result agrees with the experimental results reported in [3].



▲ **Figure 1. A bipolar push-push VCO based on a microstrip resonator is used to evaluate the performance of the push-push configuration.**

Topology

The implementation of a push-push configuration by a microstrip resonator allows us to vary the oscillation frequency over a very wide band, using a couple of varactors for tuning. To evaluate the performance of the push-push configuration, a bipolar push-push VCO based on a microstrip resonator has been designed, as shown in Figure 1.

Two identical negative resistances share the same resonator coupled to the output microstrip

IC

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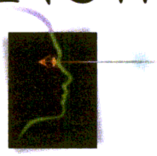
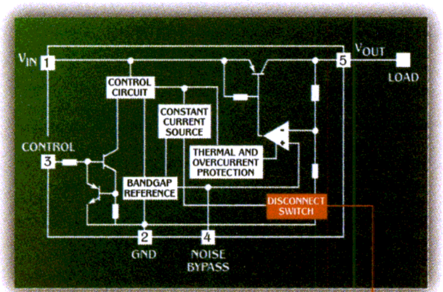


Block Diagram



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Toko's new TK716 family of LDO regulators prolongs battery life with an internal output disconnect switch that prevents low capacitance loads in the circuit from discharging. This reduces capacitance losses normally associated with standby or sleep mode in wireless systems.

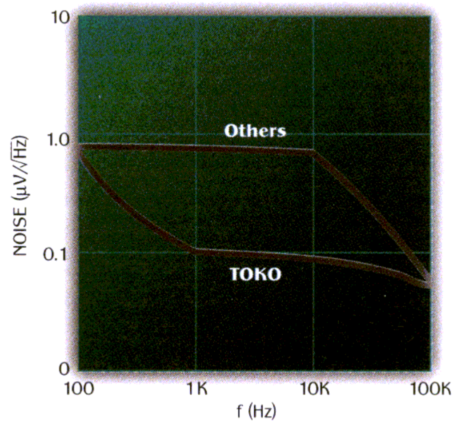
The TK716 family also offers excellent low-noise performance – less than $0.1\mu V\sqrt{Hz}$ at 10 KHz – due to its advanced design. Available in a SOT23-5 surface-mount package, the TK716 also features precise line and load regulation, and 30 standard output voltages in 0.1V increments...from 1.0V to 5V.

TK716's very low dropout voltage of 90mV (typical) at 50mA load current is essential to such products as **wireless communication devices, radio-controlled systems and electronic toys**...any portable product that runs on battery power.

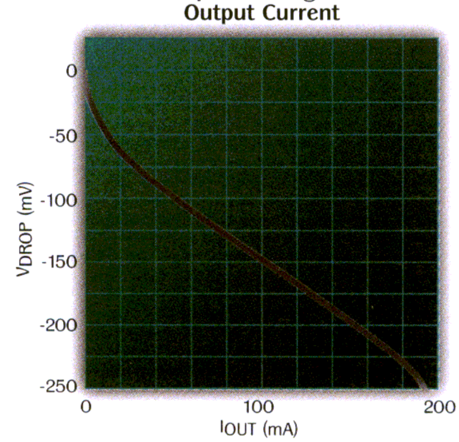
On the Roadmap: LDO regulators with a rated current output of up to 5 Amps.

For more information on the TK716 family or to order samples, visit www.ictoko.com.

Output Noise



Dropout Voltage vs. Output Current



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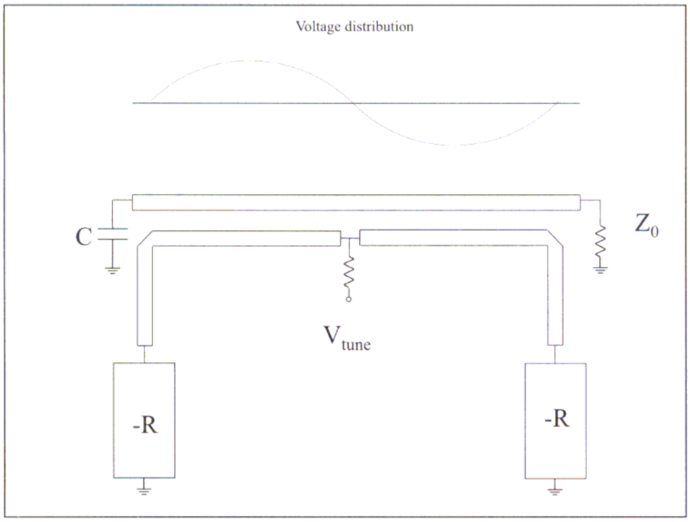
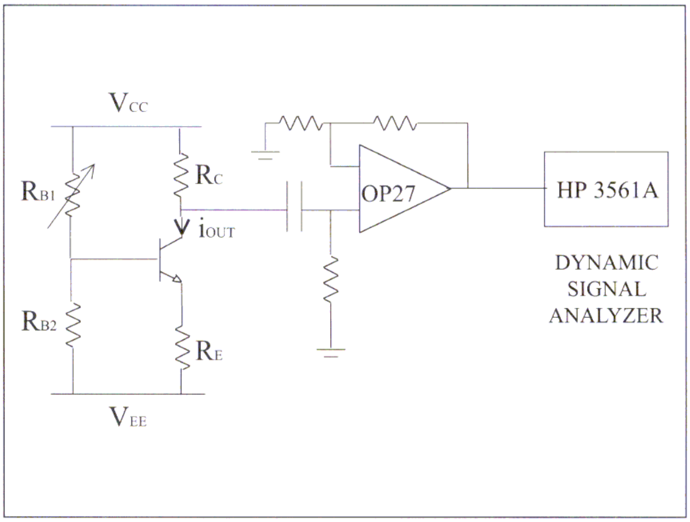
Circle 33



▲ **Figure 1b. A low value resistor placed on the common cathode node of the varactors, where there is the symmetry plane of the VCO.**



▲ **Figure 2. The test bench for the measurement of the corner frequency of the BJT, operated at rest.**

line. The voltages in phase opposition are summed at the output by the properly loaded and dimensioned coupler.

The coupler is very important for this topology of VCO. It has been designed by placing two sources in phase opposition where the negative resistances will operate. The termination load for the best summation of voltages is not a 50 ohm resistor but a capacitor with a value that varies with the gap and the length of the coupler. The impedance of the sources is an estimate of the average value presented by the active devices. Although this procedure is not rigorous, it is useful at least for a coarse design of the coupler. Tuning of capacitance, gap and length of the coupler will allow optimization during non-linear simulation of phase noise.

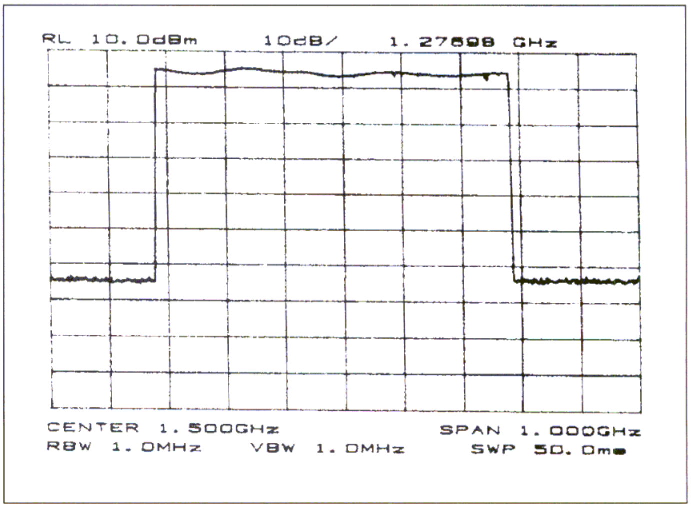
Two coupled oscillators can oscillate in the phase or antiphase mode and, in order to ensure the latter, the former can be suppressed by placing a low value resistor on the common cathode node of the varactors, where we have the symmetry plane of the VCO (Figure 1b).

Because of symmetry and phase opposition, the middle of the resonator behaves as a voltage node, and here the varactors experience the lowest RF voltage swing.

Models

The behavior of the circuit has been simulated with the software tool HP MDS. The BJT is the Avantek AT 42035. For the BJT, the nonlinear model based on the data reported in the datasheets has been used. This nonlinear model has been completed (for the noise analysis) with the A_f and k_f parameters for the BJT's flicker noise:

$$S_B(\omega) = \frac{d < i_B^{\,2} >}{df} = 2qI_B + k_f \frac{I_B^{\,A_f}}{f},$$



▲ **Figure 3. The tuning band results of the measurements.**

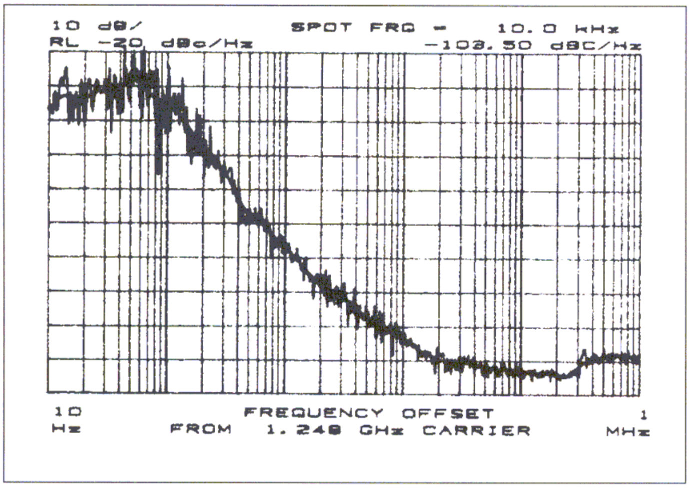
where I_B is the base current. The values for these parameters have been obtained by a measurement of the corner frequency of the BJT operated at rest. The test bench for this measurement is shown in Figure 2.

For the low noise measurement test bench, the OP27 opamp has been used with metal film resistors, ceramic capacitors and battery supply voltages [4]. Two measurements at two different collector currents allowed the solution of a nonlinear algebraic system for A_f and k_f.

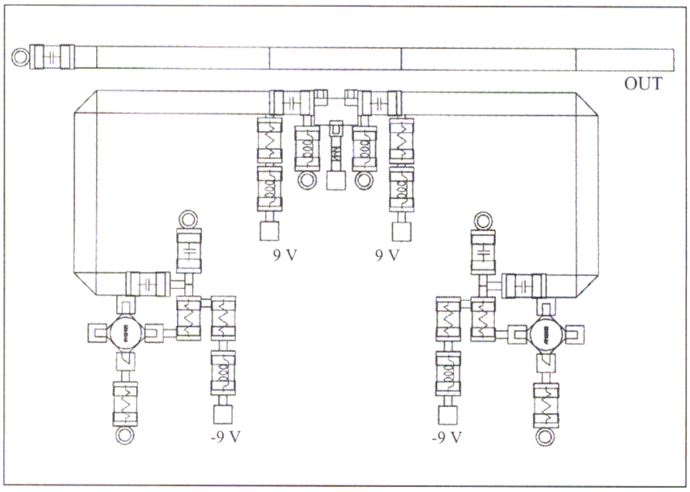
The values $A_f = 1.304$ and $k_f = 1.574.10^{-14}$ have been introduced in the model in MDS.

The varactor used is an Alpha SMV 1206-004 with junction capacitance, $C_j(V_A)$, modeled as indicated in [5], by fitting the curve reported in the datasheets; two devices are housed in SOT23 package.

Resistors (Philips), capacitors (ATC) and inductors



▲ **Figure 4. The noise-to-noise carrier ratio for VTUNE=3 V.**



▲ **Figure 5. The layout of the VCO.**

(AVX) have been described by the equivalent models suggested in the datasheets.

The simulations show that the oscillation frequency of the VCO ranges from 1.267 to 1.815 GHz, with tuning voltage from 0 to 22 V.

Careful dimensioning of negative resistance and resonator is necessary to intersect their reflection coefficients Γ_D^{-1} and Γ_R orthogonally, minimizing phase variations due to active devices for the whole voltage tuning range of varactors [6].

The results of the measurements agree with the results of the simulation. The tuning band ranges from 1.187 to 1.780 GHz (Figure 3): it is a 40 percent band around a 1.48 GHz center frequency. The output power of the VCO is between 4.4 and 5.6 dBm over the entire tuning band.

The measurement of the phase noise $L(f_m)$ has been realized by locking the oscillation frequency of the VCO with a narrow band PLL. Thus the phase noise levels measured represent the intrinsic phase noise limits of the VCO, and are not corrupted by contributions associated with the long term frequency fluctuations of the free-running VCO. The resulting phase noise level $L(f_m)$ @ f_m = 10 kHz is lower than –103 dBc/Hz over the entire tuning band. The noise-to-carrier ratio $L(f_m)$ for V_{TUNE} = 3 V is shown in Figure 4. The layout of this VCO is shown in Figure 5.

Conclusion

A theoretical explanation for the behavior of a push-push oscillator has been proposed. The good phase noise performance of this configuration has been experimentally proved by realizing a wideband push-push VCO (600 MHz tuning band around a 1.5 GHz center frequency) with less than –103 dBc/Hz over the entire tuning band.

More detailed simulations may be performed with low noise dynamic measurements of the devices becuase there are big differences in flicker noise distribution compared to static conditions. ■

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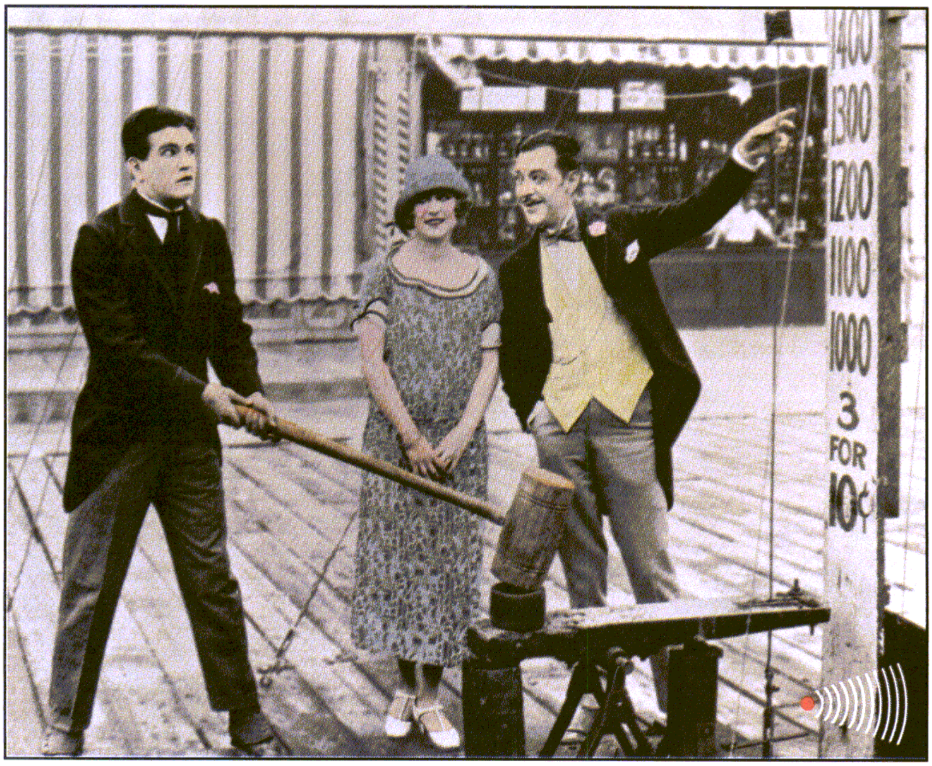
POWER



AMPLIFIERS

A Simple Analytic Method for Transistor Oscillator Design

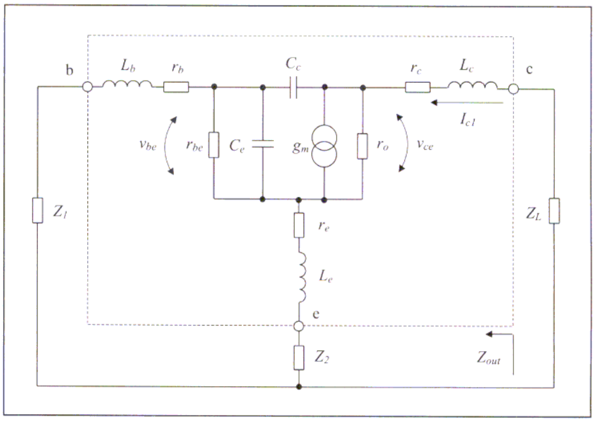
This straightforward mathematical technique helps optimize oscillator designs

By Andrei Grebennikov
Institute of Microelectronics, Singapore

A simple analytic method for transistor oscillator design has been developed. This technique defines explicit expressions for optimum values of feedback elements and load through bipolar transistor *z*-parameters. Such an approach is useful for practical optimization of a series feedback microwave bipolar oscillator.

Microwave oscillator design in general represents a complex problem. Depending on the technical requirements for designing an oscillator, it is necessary to define the configuration of the oscillation scheme and a transistor type, to measure the small-signal and large-signal parameters of a transistor-equivalent circuit and to calculate electrical and spectral characteristics of the oscillator. This approach is very suitable for implementing CAD tools if a transistor used in microwave oscillator circuits is represented by a two-port network. There are two ways to evaluate the basic parameters of the transistor equivalent circuit; one is by direct measurement and the other is by approximating based on experimental data with reasonable accuracy in a wide frequency range [1-3]. Furthermore, the equivalent circuit model can easily be integrated into a RF circuit simulator.

In large-signal operation, it is necessary to define the appropriate parameters of the active two-port network and the parameters of external feedback elements of the oscillator circuit. Therefore, it is desirable to have an analytic method to design a single-frequency optimal microwave oscillator. This helps to formulate the explicit expressions for feedback elements, load impedance and maximum output power in terms of transistor-equivalent circuit elements and their current-voltage characteristics [4]. Such an approach can be derived based on a two-step procedure. First, the optimal combination of feedback elements for realizing a maximum small-signal negative resistance to permit oscillations at the largest amplitude is defined. Second, for a given oscillator circuit configuration with maximal output power, by taking into account the large-signal nonlinearity of the transistor equivalent circuit elements, the realized small-signal negative resistance will be characterized to determine the optimum load.



▲ **Figure 1. The series feedback bipolar oscillator equivalent circuit.**

Recent progress in silicon bipolar transistors has significantly improved frequency and power characteristics. In contrast to the field-effect transistors (FETs), the advantages of reduced low-frequency noise and higher transconductance make bipolar transistors more appealing for oscillator design up to 20 GHz. A simple analytic approach used to design a microwave bipolar oscillator with optimized feedback and load will speed up the calculations of the values of feedback elements and simplify the design procedure.

General approach

Generally, in a steady-state large-signal operation, the design of the microwave bipolar oscillator is achieved by defining the optimum bias conditions and the values of feedback elements as well as the load that corresponds to the maximum power at a given frequency. Now, lets look into the generalized two-port circuit of the transistor oscillators as shown in Figure 1, where $Z_i = R_i + jX_i$, $i = 1, 2$, $Z_L = R_L + jX_L$. Such an equivalent oscillation circuit is used mainly by microwave and radio frequency oscillator design. The dotted-lined box (as shown in Figure 1) represents the small-signal SPICE2 Ebers-Moll model of the bipolar transistor in the normal region of operation. This hybrid-p model can accurately simulate both DC and high-frequency behavior up to the transition frequency $f_T = g_m/2\pi C_e$ [5]. For generic microwave bipolar oscillator design, the oscillation will arise under capacitive reactance in an emitter circuit ($X_2 < 0$), inductive reactance in a base circuit ($X_1 > 0$), and either inductive ($X_L > 0$) or capacitive ($X_L < 0$) reactances in a collector circuit.

For a single frequency of oscillation, the steady-state oscillation condition can be expressed as

$$Z_{out}(I,\omega) + Z_L(\omega) = 0 \qquad (1)$$

where $Z_L(\omega) = R_L(\omega) + jX_L(\omega)$ and $Z_{out}(I, \omega) = R_{out}(I, \omega) + jX_{out}(I, \omega)$.

The expression of output impedance, Z_{out}, can be written as

$$Z_{out} = Z_{22} + Z_2 - \frac{(Z_{12} + Z_2)(Z_{21} + Z_2)}{Z_{11} + Z_2 + Z_1} \qquad (2)$$

where Z_{ij} ($i, j = 1, 2$) are z-parameters of the hybrid transistor model.

To optimize the oscillator circuit, the negative real part of the output impedance Z_{out} has to be maximized. Based on expression (2), it is possible to find optimal values for X_1 and X_2 under which the negative value R_{out} is maximized by setting

$$\frac{\partial R_{out}}{\partial X_1} = 0, \quad \frac{\partial R_{out}}{\partial X_2} = 0 \qquad (3)$$

The optimal values X_1^0 and X_2^0 based on condition (3) can be expressed with the impedance parameters of the active two-port network in the following manner [4]:

$$X_1^0 = \frac{R_{21} - R_{12}}{X_{21} - X_{12}}\left(\frac{R_{12} + R_{21}}{2} - R_{11} - R_1\right) - X_{11} + \frac{X_{12} + X_{21}}{2}$$

$$X_2^0 = -\frac{(2R_2 + R_{12} + R_{21})}{2(X_{21} - X_{12})} - \frac{X_{12} + X_{21}}{2} \qquad (4)$$

By substituting X_1^0 and X_2^0 into equation (2), the optimal real and imaginary parts of the output impedance Z_{out} can be defined as follows:

$$R_{out}^0 = R_2 + R_{22} - \frac{(2R_2 + R_{12} + R_{21})^2 + (X_{21} - X_{12})^2}{4(R_{11} + R_2 + R_1)}$$

$$X_{out}^0 = X_2^0 + X_{22} - \frac{R_{21} - R_{12}}{X_{21} - X_{12}}\left(R_{out}^o - R_2 - R_{22}\right) \qquad (5)$$

Small-signal oscillator circuit design

At radio and microwave frequencies, the condition $r_{be} >> 1/\omega C_e$ is usually fulfilled. Besides, it is possible to ignore the effect of base-width modulation (the so-called Early effect) without a significant decrease of the final result accuracy, and to consider the resistance r_o as an infinite value. The parasitic lead inductances and substrate capacitance can be taken into account in the external feedback circuit. By doing so, the internal bipolar transistor in common-emitter small-signal operation can be characterized by the following real and imaginary parts of z-parameters:

$$R_{11} = R_{12} = a \times \left[\frac{1}{g_m} + r_b\left(\frac{\omega}{\omega_T}\right)^2\right]$$

$$R_{21} = R_{22} = R_{11} + \frac{a}{\omega_T C_c}$$

$$X_{11} = X_{12} = -a\frac{\omega}{\omega_T}\left(\frac{1}{g_m} - r_b\right) \qquad (6)$$

$$X_{21} = X_{11} + \frac{a}{\omega C_c}\left(\frac{\omega}{\omega_T}\right)^2$$

where $a = 1\Big/\left[1 + \left(\frac{\omega}{\omega_T}\right)^2\right]$ $\omega_T = 2\pi f_T$

By substituting the expressions for real and imaginary parts of the transistor z-parameters from the system of equations (6) to equations (4) and (5), the optimal www.ericsson.com/rfpower

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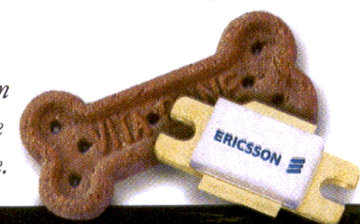

Part Number	Frequency	Power Out Watts	Gain dB	Supply Volts	IMD dBc	Efficiency %	Features
PTF 10049	470–860MHz	85	12.0	32	–32	50	Input Matched
PTF 10159	470–860MHz	120	12.0	32	–35	58	Input Matched
PTF 10160	860–960MHz	85	16.0	26	–30	54	I/O Matched
PTF 10036	860–960MHz	85	11.0	28	–30	55	Input Matched
PTF 10020	860–960MHz	125	11.0	28	–30	55	Push Pull
PTF 10100	860–960MHz	165	12.0	28	–30	47	Input Matched
PTF 10149	925–960MHz	70	16.0	26	–30	50	Input Matched
PTF 10021	1.4–1.6 GHz	30	11.0	28	–30	48	I/O Matched
PTF 10125	1.4–1.6 GHz	135	11.5	28	–30	45	I/O Matched
PTF 10035	1.9–2.0 GHz	30	11.0	28	–30	35	I/O Matched
PTF 10112	1.8–2.0 GHz	60	11.0	28	–28	41	I/O Matched
PTF 10120	1.8–2.0 GHz	120	10.0	28	–30	40	I/O Matched
PTF 10048	2.1–2.2 GHz	30	10.0	28	–30	39	I/O Matched
PTF 10122	2.1–2.2 GHz	50	9.5	28	–30	39	I/O Matched
PTF 10134	2.1–2.2 GHz	100	10.0	28	–30	36	I/O Matched

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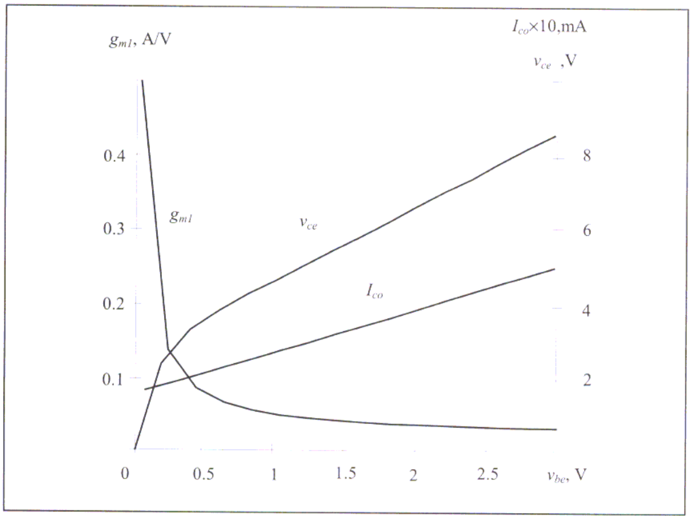


▲ **Figure 2. Amplitude dependencies of large-signal transconductance, constant bias collector current and fundamental amplitude.**

values of imaginary parts of the feedback elements $X_1{}^0$ and X_2^0 can be rewritten as follows:

$$X_1^0 = \frac{1}{2\omega C_c} - r_b \frac{\omega}{\omega_T}$$
$$X_2^0 = -\frac{1}{2\omega C_c} - r_e \frac{\omega}{\omega_T} \qquad (7)$$

In addition, the real and imaginary parts of optimum output impedance

$$Z_{out}^0 = R_{out}^0 + jX_{out}^0$$

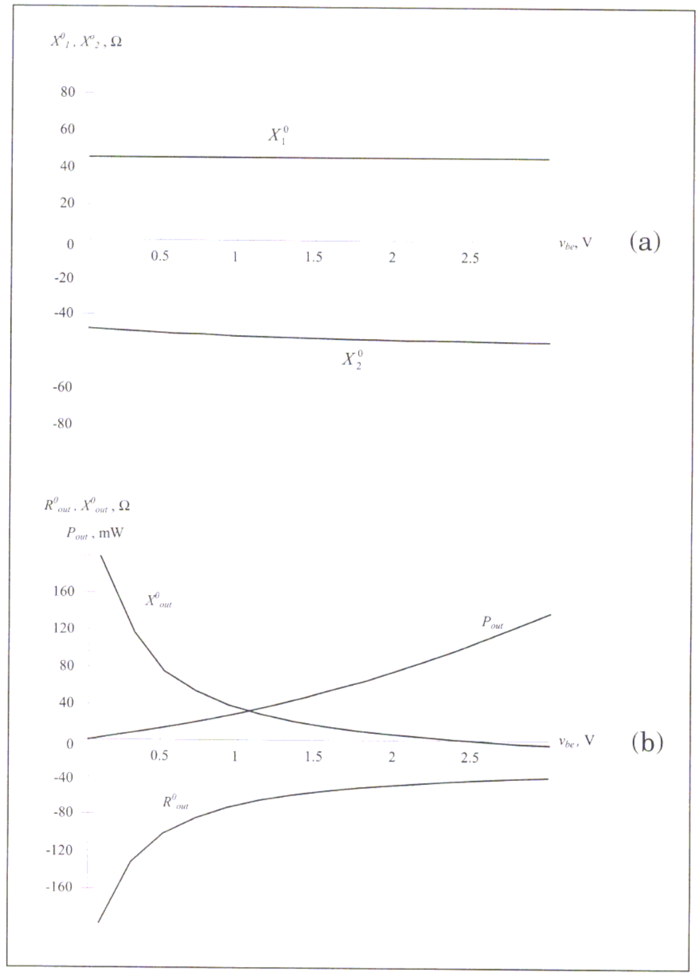
can be expressed as:

$$R_{out}^0 = r_c + \frac{r_b}{r_b + r_e + R_{11}}\left(r_e + R_{11} + \frac{a}{\omega_T C_e}\right) - \frac{a}{r_b + r_e + R_{11}}\left(\frac{1}{2\omega C_e}\right) \qquad (8)$$

$$X_{out}^0 = \frac{1}{2\omega C_c} - \left(R_{out}^0 - r_c\right)\frac{\omega}{\omega_T}$$

From equation (8), it follows that as frequency increases, the absolute value of the negative resistance R^0_{out} will be reduced and the maximum oscillation frequency f_{max} becomes zero. Without considering the parasitic series resistors r_e and r_c, the expression for f_{max} is

$$f_{\max} = \sqrt{\frac{f_T}{8\pi r_b C_c}} \qquad (9)$$

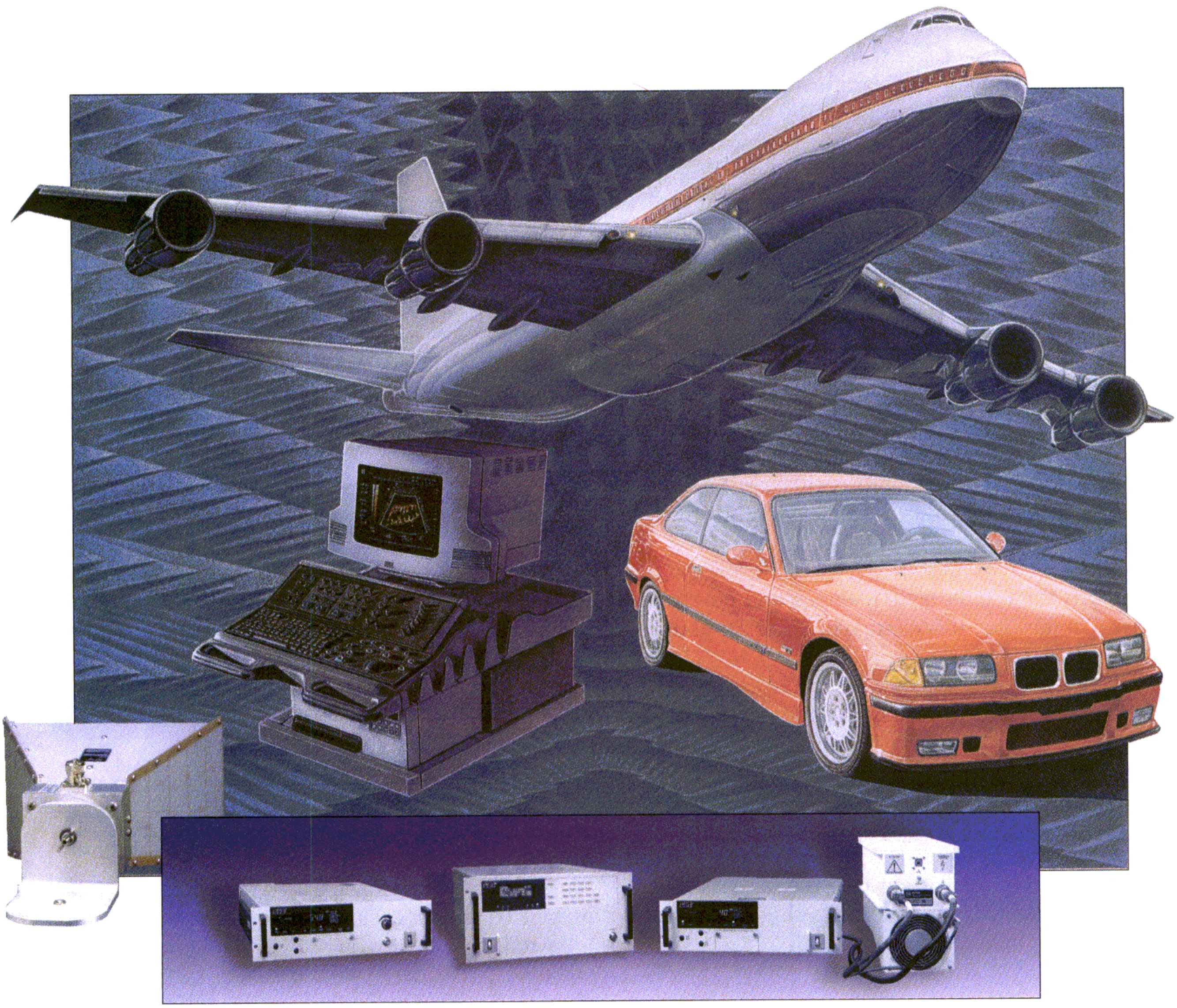


▲ **Figure 3. Amplitude dependencies of (a) optimum cicuit parameters and (b) the real and imaginary parts of output resistance and output power of the bipolar oscillator.**

Equation (9) matches with the well-known expression for f_{max} of the bipolar transistor, on which a maximum power amplification factor is unity, and the steady-state oscillation condition is carried out solely for the lossless oscillation system [6].

Large-signal oscillator circuit design

Generally, at least three steps are required for designing a large-signal oscillator circuit. First, it is necessary to choose an appropriate circuit topology, second, to determine the device large-signal characteristics and, third, to optimize the oscillator circuit to achieve the desired performances. To analyze and design the fundamental negative resistance bipolar oscillator, it is advisable to apply a quasi-linear method, based on the use of the ratios between the fundamental harmonics of currents and voltages as well as the representation of nonlinear elements by equivalent averaged fundamental linear ones. The derivation of equivalent linear elements in terms of signal voltages is based on static and voltage-capacitance bipolar transistor characteristics. In a com-



mon case, all elements of the transistor equivalent circuit are nonlinear and depend significantly on operation mode, especially for transconductance g_m and base-emitter capacitance C_e. However, in practice, it is enough to be limited to the nonlinear elements g_m, ω_T and C_c, as the base resistance r_b poorly depends on a bias condition. Besides, to calculate the constant bias and fundamental collector currents, it is sufficient to use a linear approximation of transition frequency ω_T and the small-signal value C_c at operating point [4]. Under the assumption of a sufficiently low collector-emitter bias value, the applied bipolar transistor model does not represent the effect of the forward-rectified current across the collector-base junction and the collector voltage-breakdown phenomenon. In this case, it is assumed that the transition from a soft-excited oscillation mode to a steady-state stationary mode is caused mainly by the nonlinear characteristic of the transconductance g_m in contrast to the FET oscillator. Here, under large-signal operation both the oscillator impedance and power prediction are primary due to a change in differential drain-source resistance [4].

When the base-emitter current is sufficiently small, the following exponential model can be applied to approximate the family of the experimental transfer I-V characteristics:

$$I_c = I_{cs}\left[\exp(V_{be} / V_T) - 1\right] \tag{10}$$

where V_{be} is the base-emitter junction voltage, I_{cs} is the reverse collector saturation current, and V_T is the temperature voltage. By restricting to the fundamental frequency, V_{be} can be expressed as

$$V_{be} = E_b \quad E_e \quad I_{co}R_e + v_{be}\cos\omega\, t,$$

where I_{co} is a constant bias collector current, E_b and E_e are the base and emitter bias voltages, and R_e is a self-bias resistor.

Taking into account that the voltage drops on collector and emitter transistor equivalent circuit resistors r_c and r_e are negligible, the values of the large-signal transconductance g_{m1} and the constant bias collector current I_{co} as the functions of the junction fundamental voltage amplitude v_{be} can then be defined as follows:

$$g_{m1} = \frac{2I_{cs}}{v_{be}} \exp\left(\frac{E_b - E_e - I_{co}R_e}{V_T}\right) \cdot I_1\left(\frac{v_{be}}{V_T}\right) \tag{11a}$$

$$I_{co} = I_{cs}\left[I_0\left(\frac{v_{be}}{V_T}\right)\exp\left(\frac{E_b - E_e - I_{co}R_e}{V_T}\right) - 1\right] \tag{11b}$$

where $I_0(v_{be}/V_T)$, $I_1(v_{be}/V_T)$ are the modified first order Bessel functions [7].

For the steady-state stationary oscillation mode, it is necessary to define the analytic relations between the load current amplitude and fundamental amplitude of collector voltage with input voltage amplitude v_{be}. These relations are expressed in terms of the transistor z-parameters and oscillator circuit parameters in the following manner:

$$I_{cl} = \frac{Z_{11} + Z_2 + Z_1}{Z_{11}Z_2 - Z_{12}(Z_2 + Z_1)} V_{be}$$

$$V_{ce} = \frac{Z_{22}(Z_{11} + Z_2 + Z_1) - Z_{21}(Z_2 + Z_{12})}{Z_{12}(Z_1 + Z_2) - Z_{11}Z_2} \tag{12}$$

In a steady-state operation, the output power of the oscillator is

$$P_{out} = I_{cl}^2 \operatorname{Re} Z_L / 2$$

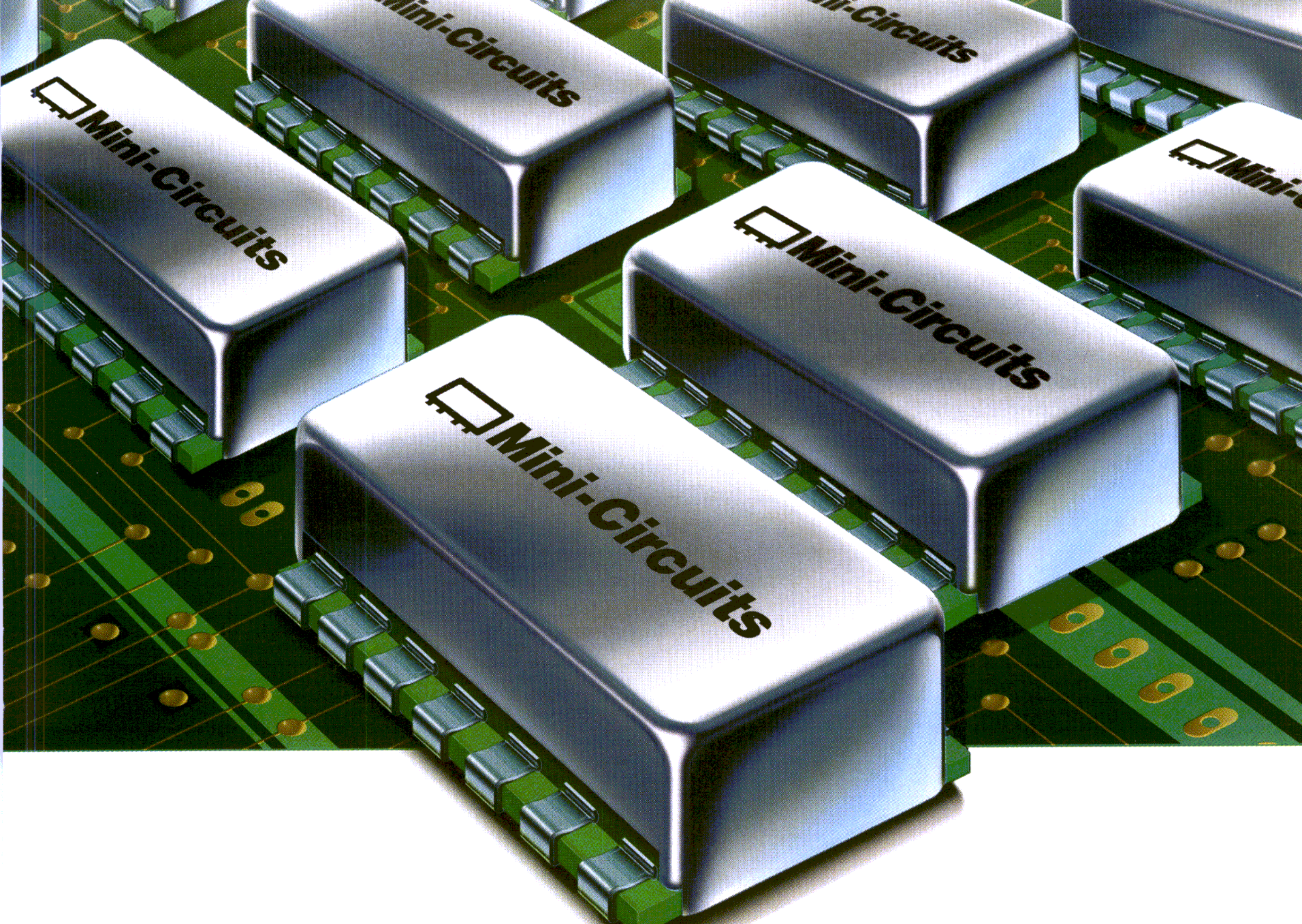
By replacing load resistance Z_L from equation (1) and using the parameters of transistor equivalent circuit from (6) to (8), P_{out} can be expressed as

$$P_{out} = -ag_{ml}^2(r_b + r_e + R_{11})\frac{R_{out}^0}{r_b + r_c - R_{out}^0} \cdot \frac{v_{be}^2}{2} \tag{13}$$

Results

An analytic approach was applied to the microwave bipolar oscillator design. The bipolar transistor has the following parameters of its equivalent circuit: f_T = 6 GHz, C_c = 0.5 pF, g_m = 1.6 A/V, r_b = 4 ohms, r_e = 0.3 ohm, r_c = 1.75 ohms, $L_e = L_c$ = 0.5 nH, L_b = 0.3 nH. The numerical calculation was performed using the values of the oscillation frequency f = 4 GHz, biases E_b = 0 V and E_e = 2 V, self-bias resistor R_e = 100 ohms and reverse saturation collector current I_{cs} = 10 µA. The numerical results obtained are shown in Figures 2 and 3. Figure 2 shows the amplitude dependence of large-signal transconductance g_{m1}, constant bias collector current I_{co}, and fundamental collector amplitude v_{ce}. Figure 3 shows the amplitude dependence of (a) optimum circuit parameters, X_1^0 and X_2^0 (b) the real and imaginary parts of the output resistance R_{out}^0, X_{out}^0 and output power P_{out}.

According to Kurokawa [8], as the negative resistance R_{out}^0 reduces with the increase of the base-emitter junction and collector voltage amplitudes, the stable oscillations are established in the oscillator. In that case, the value of the large-signal transconductance g_{m1} is significantly reduced as shown in Figure 2. In order to realize the maximal or required output power in a linear operation region without the saturation effect, it is necessary to choose the appropriate load value or to use the diode in parallel to load for restriction of collector voltage amplitude. The applied model of the microwave bipolar



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JTOS-1000W	500-1000	-94	-26	18V	25	21.95
JTOS-1025	685-1025	-94	-28	16V	22	18.95
JTOS-1300	900-1300	-95	-28	20V	30	18.95
JTOS-1550	1150-1550	-101	-20	***	30	19.95
JTOS-1650	1200-1650	-95	-20	13V	30	19.95
JTOS-1750	1350-1750	-101	-16	***	30	19.95
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JTOS-2000	1370-2000	-95	-11	22V	30 (@8V)	19.95
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transistor does not include a collector-base diode, as such an operation region is not recommended for the oscillator due to the significant deterioration of its noise properties. By means of the condition $v_{ce} \leq E_c$, the appropriate amplitude restriction at the numerical calculation can be considered.

From numerical calculations, it follows that v_{ce} and I_{co} are increased linearly with the increase of base-emitter junction amplitude v_{be}, starting from some significantly small values as shown in Figure 2. Therefore, analytically it is interesting to consider the influence of the emitter self-bias resistor R_e on the character of amplitude dependence of the fundamental output current I_{c1}. The fundamental output current can be expressed as

$$I_{cl} = 21_{cs}\left[\exp\left(\frac{E_b - E_e - E_{co}R_e}{V_T}\right) \times I_1\left(\frac{v_{be}}{V_T}\right)\right] \tag{14}$$

By differentiating, equation (11b) can be rewritten based on $d_{I0}(v_{be}/V_T)/d(v_{be}/V_T) = I_1(v_{be}/V_T)$ in the form of

$$\frac{dI_{co}}{dv_{be}} = \frac{I_{co} + I_{cs}}{V_T + (I_{co} + I_{cs})R_e} \cdot \frac{I_1(v_{be}/V_T)}{I_0(v_{be}/V_T)} \tag{15}$$

From the definition of modified first kind Bessel functions, it follows that for values $(v_{be}/V_T) \leq 5$ the following inequality is to be defined:

$$0.9 \leq I_1(v_{be}/V_T)/I_0(v_{be}/V_T) \leq 1$$

Hence, as it follows from (15), the value of constant bias collector current I_{co} varies practically linearly with the increase v_{be} under condition $v_{be} \geq 5V_T$, and the slope of the dependence $I_{co}(v_{be})$ is defined by value R_e. By substituting expression (11b) into equation (14), it allows us to rewrite the expression for fundamental output current I_{c1} as

$$I_{cl}(v_{be}) = 2\left[I_{co}(v_{be}) + I_{cs}\right] \times \frac{I_1(v_{be}/V_T)}{I_0(v_{be}/V_T)} \tag{16}$$

From equations (15) and (16) it follows that for a bipolar oscillator with an emitter self-bias resistor, the collector constant bias current I_{co} and the fundamental output current I_{c1} depend linearly on the base-emitter junction voltage amplitude v_{be} under the condition $v_{be}/V_T \geq 5$. Thus, the linearizing influence of the emitter self-bias resistor for the transistor oscillator is similar to the influence of the negative feedback resistor on a linearization of the amplifier gain-transfer characteristics. From Figure 3b, it follows that it is possible to realize a sufficiently high level of output power under the direct connection of standard load R_L = 50 ohms without use of a special load matching circuit.

Conclusion

A simple analytic method of microwave and RF bipolar oscillator design has been developed, allowing the definition of explicit expressions for optimum values of feedback elements and load through bipolar transistor z-parameters. A negative resistance concept is used to design series feedback microwave and RF bipolar oscillators with optimized feedback elements and maximum output power in terms of transistor impedance parameters. Based on its small-signal z-parameters and DC characteristics, a simplified large-signal model for a bipolar transistor is derived. The linearizing effect of the external emitter self-bias resistor is analytically shown. ■

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Author Information

Andrei Grebennikov received a Dipl-Ing. degree from the Moscow Institute of Physics and Technology and a Ph.D degree from Moscow Technical University of Communication and Informatics in 1980 and 1991, repectively. In 1983, he began working as a research assistant at the Scientific Research Department of Moscow Technical University of Communication and Informatics. Since 1998, he has worked with the Institute of Microelectronics, Singapore. He may be reached via telephone at +65-770-5494, via fax at +65-773-1915, and by e-mail at andrei@ime.org.sg.

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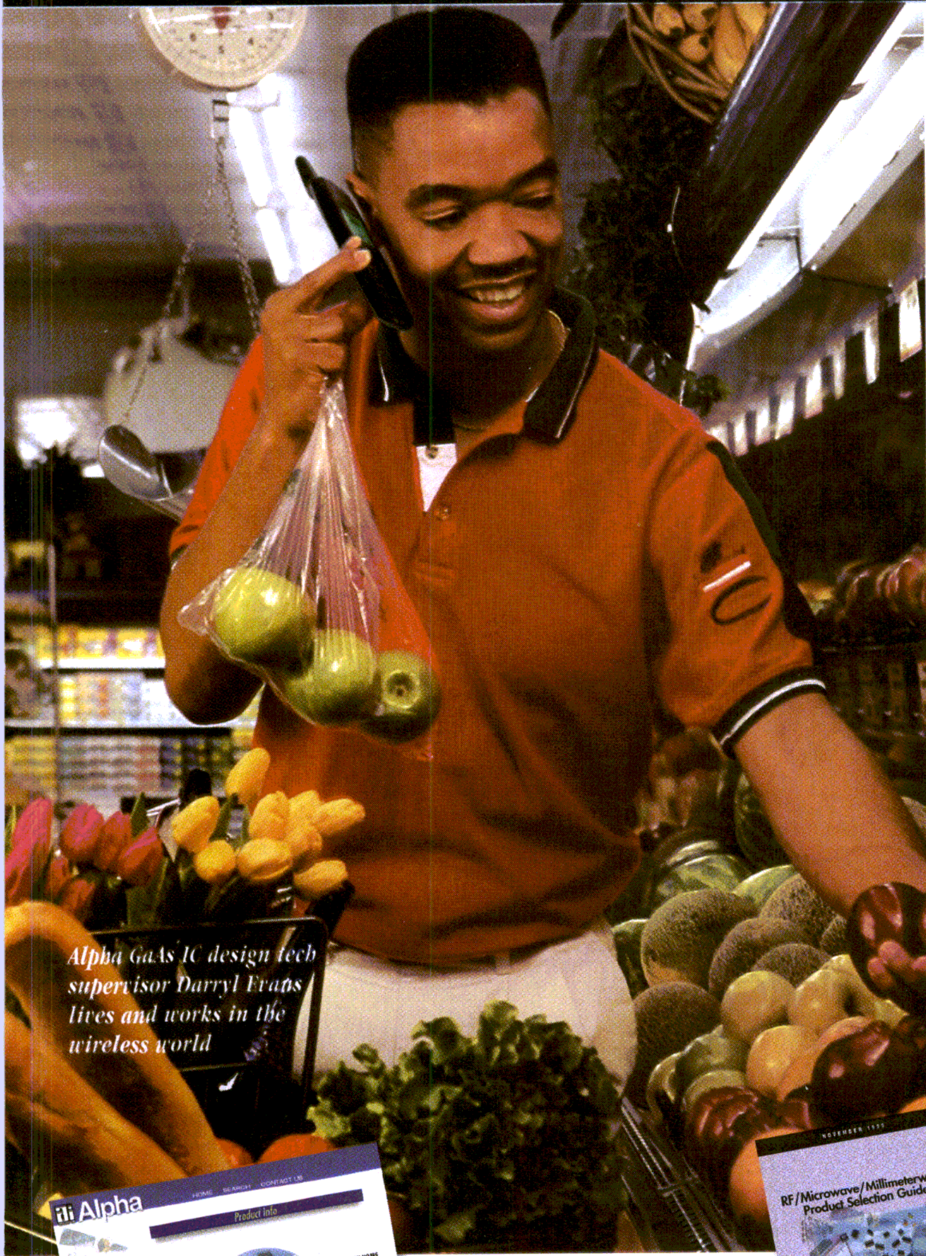

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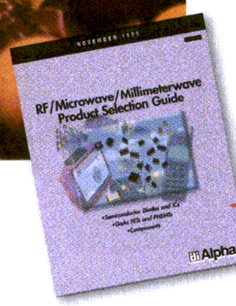

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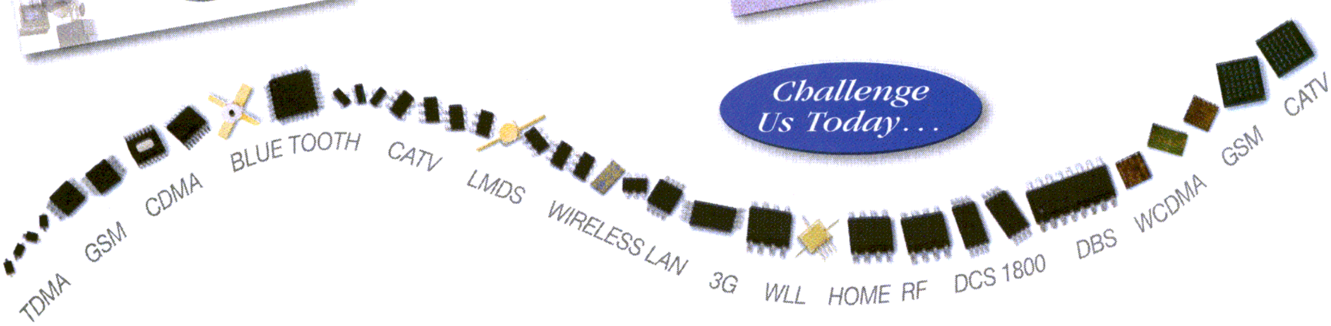

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Circle 71

Products

EMI/RFI CONTROL

RF shielding with removable cover

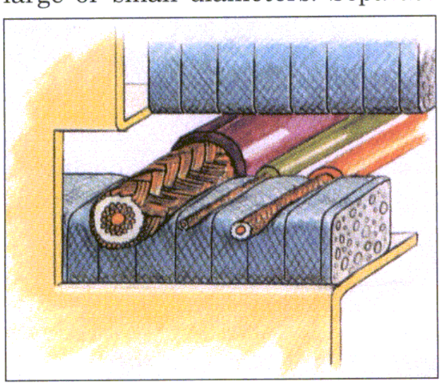
Photofabrication Engineering is now producing a line of photochemically etched RF shielding cans with a patented easy access removable cover. The unique design allows the

user to replace or repair internal components without damaging the surrounding shield. A tab connecting the lid to the shield can be snipped and resoldered quickly. The new shields are available in brass and solder-plated 10-10 steel. A free sample is offered by the company.

Photofabrication Engineering Inc.
Circle #160

EMC cable entry

Holland Shielding Systems has developed a cable entry system that simultaneously handles cables with large or small diameters. Separate

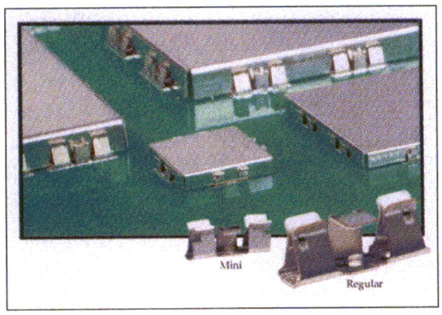


parts of the shielding strips can deflect separately to guarantee that every cable makes optimal contact. It permits new cables to be added easily without disassembly. The system also makes sure that there are no gaps in the shield. The entry system can be supplied ready-made or easily integrated into an enclosure. The shield strips are available for cable diameters from 1 to 25 mm, and are offered in high temperature or flame retardant versions.

Holland Shielding Systems BV
Circle #161

Clips hold shields in place

A new version of the Mini-Shield line from Autosplice, Inc. provides flexibility in the configuration and mounting of EMI/RFI shields. The



shield clips eliminate secondary hand-soldering operations. EIA packaging and ample surface area for vacuum pick-up allow for smooth handling and placement at up to 2,000 parts per hour on common equipment. The clips are self-aligning with as much as 20 percent correction during reflow soldering. Shield covers can then be placed into the spring loaded contacts at the end of the process after cleaning and inspection has been completed.

Autosplice, Inc.
Circle #162

TEST EQUIPMENT

Microwave analyzers get power boost

IFR Systems has extended the capabilities of its 6800 Series Microwave Systems Analyzers with the addition of a high output power source option. The option provides a minimum of +10 dBm of leveled power up to 24 GHz, making it suitable for local oscillator substitution. The higher power results in a 5 dB increase in dynamic range for scalar and tuned input modes. The 6820 line of analyzers provides full scalar analysis capabilities, while the 6840 line adds spectrum analysis features. The High Power Option 030 ranges in price from $1,620 to $4,050, depending on the analyzer model.

IFR Systems, Inc.
Circle #163

Communication analyzer gets software upgrade

Anritsu Company has developed a software option for its MT8802A Radio Communication Analyzer

that makes possible true soft handoffs with independent power control bits in a single instrument. The software eliminates the need to have a live base station in the lab, or to conduct handoff tests in the field. With the software, the MT8802A can simultaneously simulate two base stations, creating a real-world scenario in the laboratory. The software option is priced at $2,000. The MT8802A is $53,000.

Anritsu Company
Circle #164

New spectrum analyzers

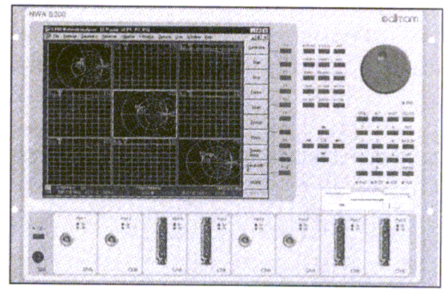
Tektronix, Inc. announces the FSP family of spectrum analyzers by Rohde & Schwarz, a pair of value instruments with high accuracy and measurement speed. Model FSP-3 covers 9 kHz to 3 GHz, while the FSP-7 spans 9 kHz to 7 GHz. The instruments are designed for high volume production testing of second and third-generation wireless handsets and base stations. Prices start at $18,950 for the FSP-3 and $25,950 for the FSP-7.

Tektronix, Inc.
Circle #165

Products

Multiple port vector network analyzer

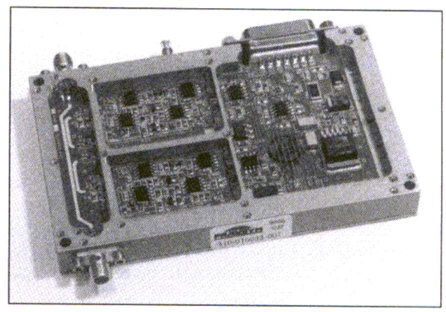
The S200 Vector Network Analyzer from Ballmann is a true multi-port VNA that uses RF heads on every port. Full 12-term error correction is used for all ports and all measurements have no residual error because no switching is performed. Up to 8 RF heads can be included in the base unit, with another 8 heads available in a second case. The main benefits of the S200 are high speed measurement of n-port devices, measurement accuracy and dynamic range of 140 dB with a 1 kHz IF. Other features include up to 500,000 points per sweep, log sweep, 50 or 75 ohm heads, internal or external head and two optional synthesized sources for intermodulation and frequency converter testing. A built-in Pentium PC operates with Windows NT and a 112-inch LCD TFT color display.



SJ Technologie
Circle #166

Cellular power monitor

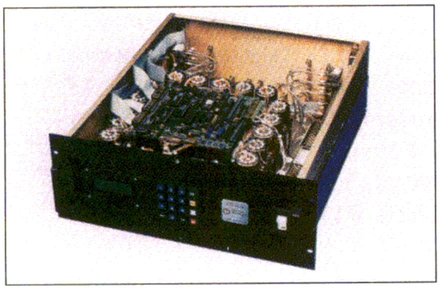
The Cellular Power Monitor from Praxsym has a wide dynamic range and immunity to multi-tone distortion, making it suitable for use with newer digital transmission formats. The unit monitors operation of a cellular transmission system, measuring the forward and reflected power to indicate transmitter and antenna system performance. Power levels are represented by a linear (mV/dB) signal that can be detected by an analog circuit, including portable voltmeters. The power monitor is priced at $795 each in quantities of 1,000.



Praxsym, Inc.
Circle #167

Bi-directional coaxial switch matrix

Dow-Key Microwave has released the Model 2101 bi-directional coaxial switch matrix. The switching system is designed for use in ATE systems, laboratory test systems and programmable interconnect systems. The matrix is configured as 10 inputs and 10 outputs, using Dow-Key electromechanical coaxial switches to cover DC to 18 GHz. Each unit has a solid state controller with front panel LCD display, a keypad for manual con-

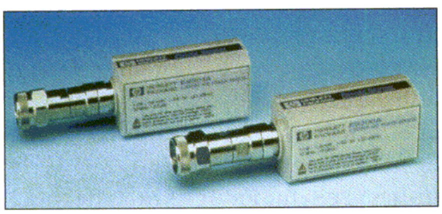


trol, and three remote interfaces (RS-232, RS-422 and IEEE-488/GPIB).

Dow-Key Microwave
Circle #168

Power sensor reaches for lower frequencies

To provide low frequency coverage for EMC and EMI test applications, Agilent Technologies announces the new E9304A power sensor for the EPM series E4418B and E4419B power meters. The new sensor's 9 kHz to 6 GHz frequency range and wide dynamic range (–60 to +20 dBm) allows EMC test engineers to capture the full range of field strengths during



testing, usually without the need to switch power sensors. The E9304A permits high speed measurements, reducing the time required for field uniformity measurements or EMC test receiver calibration. The power sensor is available now for $1,500.

Agilent Technologies
Circle #169

SEMICONDUCTORS

Set top tuner chipset

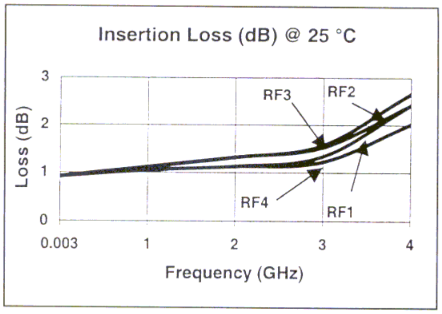
Mitel Semiconductor has introduced a double-conversion chipset solution for tuners in cable set top boxes. The chipset offers the benefits of alignment-free operation with improved performance. The use of bipolar technology reduces cost and increased immunity to electrostatic discharge, compared to GaAs devices in this application. The chipset includes the SL2030 and SL2035 up and down converters, along with the SP5848 dual phase locked loop. Together, they consume 0.4 watts of power. The three-chip set costs $3.50 in high volume quantities.

Mitel Semiconductor
Circle #170

SP4T switch MMIC

M/A-COM offers the new SW65-0440 MMIC, an absorptive SP4T GaAs switch with an integral TTL/CMOS driver. It is intended for diversity and channel switching in applications such as wireless telephony, WLAN and GPS receivers. Insertion loss is typically 1.2 dB with 35 dB isolation up to 2 GHz, 29 dB up to 3 GHz. The 1 dB

Products



compression point is +20 dBm from 0.5 to 3.0 GHz, with in input IP_3 of +46 dBm over the same frequency range. The SW65-0440 switch is priced at less than $6 each in 1,000 quantities.

M/A-COM
Circle #171

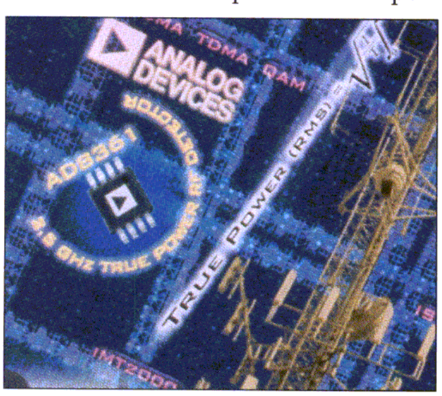
Chips support HDR

QUALCOMM announces the iMSM4500 and iMSM5500 Mobile Station Modem chips that add the company's High Data Rate (HDR) capability to their present CDMA handset products.

QUALCOMM
Circle #172

IC detects true RMS power

Analog Devices has developed a new IC offering true RMS-to-DC power detection over a frequency range of 0.1 to 2.5 GHz. It accurately measures complex waveforms such as CDMA and W-CDMA in wireless applications. The AD8361 has a frequency coverage that includes all worldwide cellular bands. The IC is specified for oper-

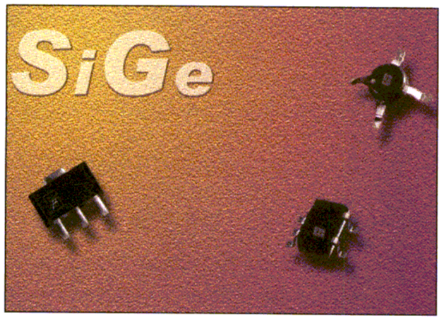


ation over a –30 to +85° C temperature range and operates from a 2.7 to 5.5 VDC supply. The AD8361 is provided in an 8-pin micro-SO package and is priced at $3.75 each in quantities of 1,000.

Analog Devices
Circle #173

High linearity, high efficiency SiGe MMICs

Stanford Microdevices' SGA-6000 Series high linearity SiGe RFIC MMIC amplifiers are offered in industry standard 85mil, SOT89 and SOT23-5 surface-mount plastic packages for design flexibility. These unconditionally stable amplifiers operate with single-supply voltages as low as 4.2 V, 75 mA of



current and are ideal for use applications such as HomeRF, Wireless Internet and IEEE 802.11. Model SGA-6386 has 1 dB compressed output power of +20 dBm, output third-order intercept point of +36 dBm and 15.5 dB of gain at 900 MHz. Pricing of the SGA-6386 is $1.21 in quantities of 10,000 pieces.

Stanford Microdevices, Inc.
Circle #174

WIRELESS SYSTEMS

Multilayer chip antenna

Murata Electronics North America announces availability of the LDA82 Chip Multilayer Antenna, using LTCC technology to achieve a small, surface mountable ceramic monolithic design. The LDA82 type is targeted to applications in the 900 to 2600 MHz range, including PDAs, pagers, cordless telephones, WLAN cards and ISM 2.4 GHz products. The primary advantage is mechanical robustness, resulting in an antenna with no protrusions that can be broken off. Surface mounting simplifies manufacturing of wireless products, while its 9.5 × 2 × 2 mm volume reduces their size. The LDA82 antenna is priced at $1 each in high volumes.

Murata Electronics North America
Circle #175

Spread spectrum transceiver operates at 2.4 GHz

RF Neulink introduces the SS9600 Spread Spectrum Data Transceiver, providing point-to-

point or point-to-multipoint communications at 9600 bps. The unit is a true frequency hopping design and large systems with as many as 238 units can be configured. The SS9600 operates in the 2.4 GHz unlicensed band. It is a true plug and play product for most small point-to-point applications.

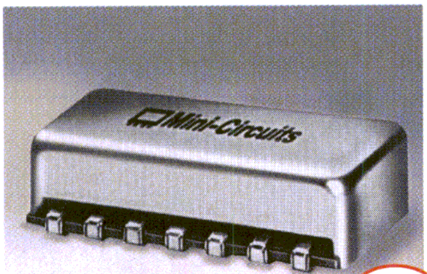
RF Neulink
Circle #176

Ceiling mount 7 dBi antenna array

Cushcraft recently engineered a directional PCS antenna for ceiling mounting in locations where directional coverage and low visibility are required. The S1857MD exhibits 7 dBi gain in the 1850 to 1990 MHz band. Its size is 3-3/4 × 7-1/4 × 2 inches and it weighs 3.2 oz. (90 g). The housing is vacuum formed UV-stabile thermoplastic in standard white color or with custom finishes. Applications include microcells, picocells and RF distrib-

NEW PRODUCTS NO.66

RF/IF MICROWAVE COMPONENTS

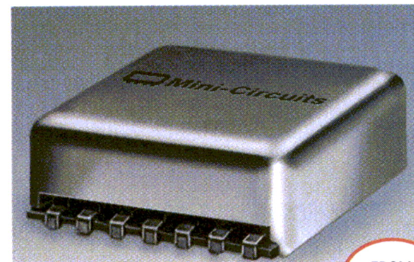


LOW COST 1150 TO 1550MHz VCO HAS LINEAR TUNING

Mini-Circuits has introduced a broad band 1150 to 1550MHz surface mount voltage controlled oscillator with 22 to 32MHz/V (typ) linear tuning characteristics. Typically, the JTOS-1550 operates with high 7dBm power output and features low 121dBc/Hz SSB phase noise at 100kHz offset, -20dBc harmonic suppression, and 0.5 to 20V (min. to max.) tuning voltage. Solder plated J leads provide superior mechanical integrity over temperature. Ideal for GPS applications.

BROADBAND MIXER REDUCES INTERMODS 10 TO 2400MHz

This level 17 (LO) SYM-25H surface mount frequency mixer from Mini-Circuits targets PCN, ISM, and cellular applications within 10 to 2400MHz. Typically at center band, the mixer exhibits high 25dBm IP3, low 6.1dB conversion loss, and high 40dB L-R and L-I isolation. Ruggedly constructed in a low cost plastic package and covered by a 5 year Ultra-Rel ® guarantee.



6WAY SPLITTER/COMBINER HAS HIGH ISOLATION 75 TO 425MHz

Mini-Circuits JCPS-6-3 is a 6way-0° surface mount power splitter/combiner for VHF receiver and instrumentation applications within the 75 to 425MHz band. Typically, important performance features are high 23dB isolation, excellent 0.1dB amplitude unbalance, good 2 degrees phase unbalance, plus excellent 0.9dB insertion loss (above 7.8dB). Maximum operating temperature is -20°C to +85°C, and power input is 0.25W (max.) as a splitter. Excellent price/performance value.

75 OHM DIRECTIONAL COUPLERS FOR CATV

Mini-Circuits announces off-the-shelf availability of their new 20 to 1000MHz ADC-12-4-75 directional couplers. These broad band 75 ohm units provide a nominal 12.6±0.5dB coupling value with excellent ±0.1dB (typ) flatness for CATV applications. Housed in a miniature 0.112" (max. height) water washable package, these 1W couplers typically exhibit low 0.9dB insertion loss, good 23dB midband directivity, and good 1.2:1 VSWR. Circuit and package patent pending.

RF TRANSFORMERS HAVE 4:1 IMPEDANCE 500 TO 2500MHz

Mini-Circuits broad band TCM4-1W surface mount RF transformers operate in the 3 to 800MHz band with 4:1 impedance ratio. Referenced to midband loss (0.8dB typ), insertion loss is 1dB from 10MHz to 100MHz, 2dB in the 5 to 400MHz range, and 3dB band wide. Input return loss is 20dB (typ) at midband and maximum RF power is 250mW. Uses include CATV, VHF/UHF receivers, balanced amplifiers, and impedance matching. Shipped from stock.

2W SMA ATTENUATORS AVAILABLE IN DESIGNER'S KIT

Ten different DC to 18GHz fixed attenuators from Mini-Circuits "BW" series are now available at a special evaluation price in designer's kit form. Kit number K-BW3 contains 1 each of models BW-S1W2 to BW-S10W2 (nominal attenuation values 1 to 10dB), and each supplies precision accurate attenuation from a small stainless steel package measuring only .85" in length. Built tough to handle 2W average, 125W peak power, this 10 unit kit ships worldwide within 1 week.

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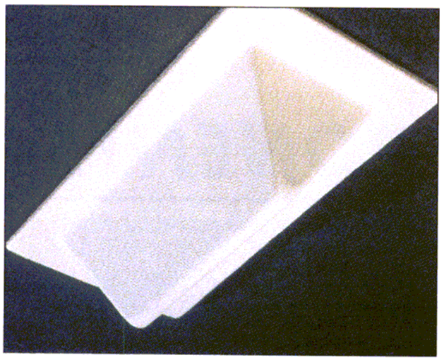


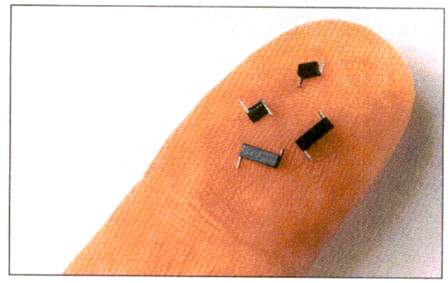
ution systems in hallways, tunnels and subways, as well as malls, parking garages, airports, office buildings and campuses.

Cushcraft Corporation
Circle #177

PASSIVE COMPONENTS

SMD air inductors

Coilcraft has introduced the Micro Springs™ line of extremely small surface mount air core inductors. Measuring as small as .06 × .08 × .05 inches (1.4 × 2.2 × 1.4 mm), these components provide typical Q factors of 200 at 1.8 GHz. The inductors are available in values from 1.65 nH to 12.55 nH with tolerances as low as 2 percent. An



acrylic jacket with a flat top makes them suitable for automatic placement and reflow or vapor phase soldering processes. Samples are available in designer's Kit C108. In 10,000-unit quantities, the Micro Spring inductors are priced at 16 cents each.

Coilcraft
Circle #178

Low cost chip attenuators

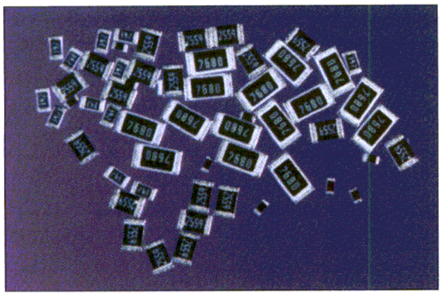
EMC Technology announces the TS0400 series of low cost chip attenuators for operation from DC to 2.5 GHz. Input power is rated at 1 watt to 85° C with a VSWR of 1.25:1. The chips are available in attenuation values of 1, 2, 3, 6, 10 and 20 dB. Attenuation accuracy is as good as ±0.3 dB. They are constructed on an alumina planar substrate that is 0.100 × 0.125 × 0.021 inch in size.

EMC Technology
Circle #179

Thin film resistors

Venkel Corp. manufactures thin film chip resistors in three different types. Rectangular Thin Film Chip Resistors are designed for high precision with exceptional stability and reliability. They are available in 0402, 0603, 0805 and 1206 sizes. Ultra Precision Thin Film Chip Resistors feature minimized aging

Products



with TCR and tolerances of ±5 ppm/°C and ±0.05 percent respectively. This line is available in 0805 and 1206 in 100 to 100k ohm values. MELF Thin Film Chip Resistors are made of a tantalum alloy, cylindrical in shape, with high mechanical body strength.

Venkel Corp.
Circle #180

SOFTWARE

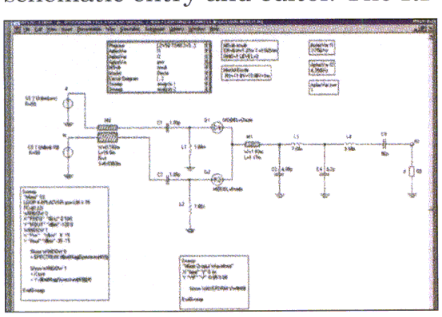
HFSS optimization module

Optimetrics is a new optimization and parametrics module for Ansoft's 3D simulation tools, beginning with HFSS Version 7.0. Optimetrics allows users to perform optimization, parametric analysis, sensitivity analysis and related design-of-experiments (DOE) studies for manufacturability. These features can be applied to magnetic, electromechanical and microwave products and circuits.

Ansoft
Circle #181

Tools analyze systems and circuits

The APLAC Circuit Simulator and Design Package helps design and analyze RF circuits. The basic package includes linear and harmonic balance simulation with schematic entry and editor. The RF



Design Tool supports microwave components and special methods for both PCB and IC-level design. The System Simulation Tool is optimized for path analysis of digital communications, and the Electromagnetic Tool permits the study of complex structures and multiple radiation sources.

APLAC Solutions Corporation
Circle #182

EM software update

CST now offers CST Microwave Studio version 1.1. This version includes the new CST MWS Eigenmode solver to allow fast calculation of filters. Also included is improved import of CAD data (IGES, SAT, STL and DXF files). Automatic "healing" of invalid EM models has been improved. A new SPICE extractor assists in the

design of high speed digital devices, and version 1.1 also has improved auto meshing, far field calculations, edge correction and de-embedding.
CST of America
Circle #183

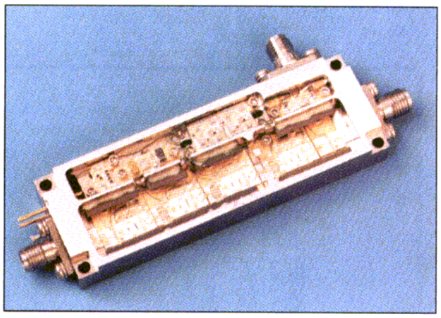
New version of HP Basic

Agilent Technologies (formerly Hewlett Packard) has introduced Version 7 of HP BASIC for Windows®. The new release combines a new graphical interface with new Rocky Mountain BASIC (RMB) support. RMB is a 20-year-old development language for test and measurement that runs on UNIX® workstations. Version 7 of HP BASIC allows users to preserve existing codes originally developed for RMB. The new release also incorporates new drivers for National Instruments' line of data acquisition cards. The new software is designated model E2060C and is priced at $1,115.
Agilent Technologies
Circle #184

SIGNAL PROCESSING

Miniature log detector

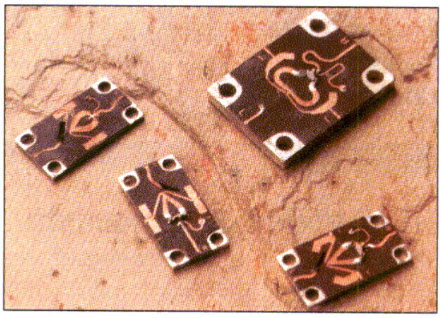
Model STDLVA-15015 from Signal Technology Corporation is a 2 to 6 GHz successive detection log



video amplifier with a linear RF output gain of 50 dB. The dynamic range at the video output is –65 to +5 dBm, tangential sensitivity is –75 dBm and total logging accuracy over temperature, frequency and power is ±3 dB. The logging slope can be set to between 15 mV/dB and 50 mV/dB. The STDLVA-15015 is provided in a 3 × 1 × 4 inch connectorized machined housing.
Signal Technology Corporation
Circle #185

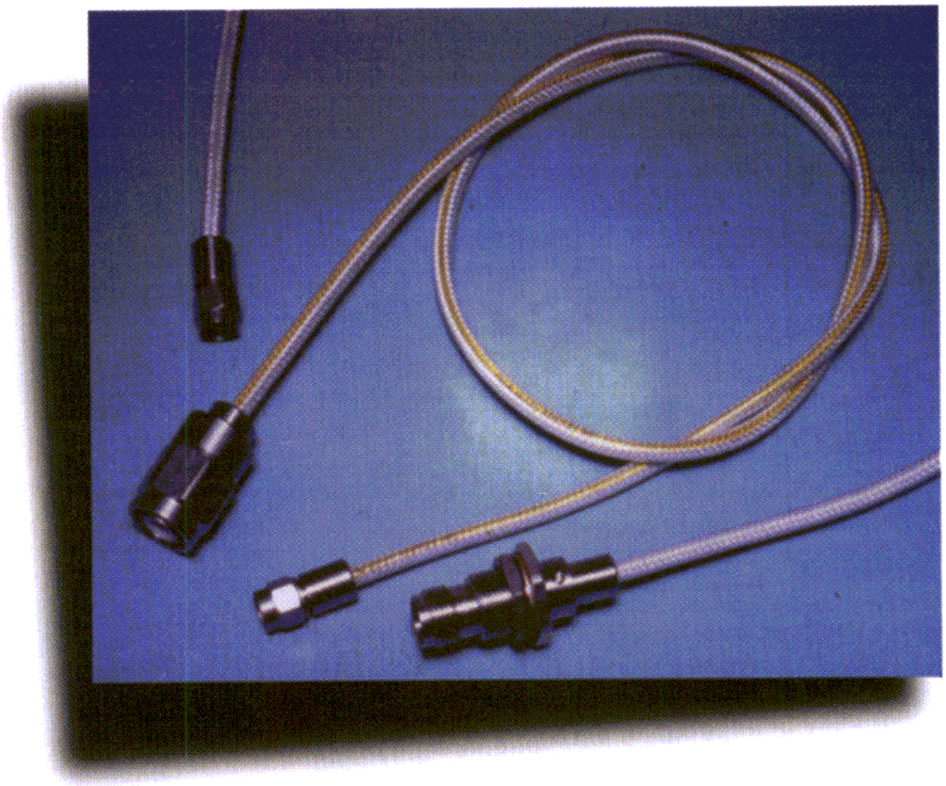
Open-carrier mixers

Stellex Microwave Systems has a new line of mixers featuring operation in frequency bands of 2.0 to 10.0 GHz, 4.0 to 22.0 GHz and 10.0 to 26.5 GHz. Typical IF bandwidths



OVERNIGHT DELIVERY

Semflex SIDEWINDER FLEXIBLE INTERCONNECT Cable Assemblies



SIDEWINDER FEATURES THE FOLLOWING:

- Frequency: D.C. to 18.0 Ghz
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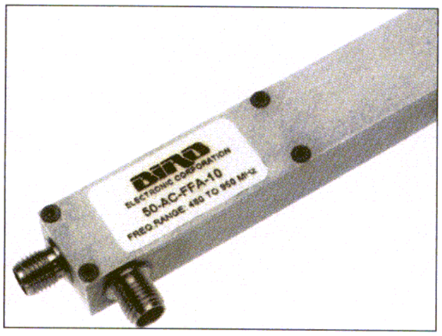
Bird Component Products (BCP) offers the 50-AC-FFA-10, a compact, cost-effective solution to UHF and cellular applications. This coupler operates from 0.5 to 1 GHz with available coupling factors of 10, 20 or 30 dB. Other specifications include VSWR of 1.15:1 maximum, reflected power of 5 watts



average and frequency sensitivity of ±0.75 dB maximum. The units are 3.61 × 0.86 × 0.42 inches in size and weigh 1.3 oz. (37 g).

Bird Component Products
Circle #187

LITERATURE

Products resource guide

Jensen Tools, a division of The Stanley Works, has just released an updated version of its *Communication Products Resource Guide* for Fall 1999. This 100-page, full color catalog offers a wide range of tool kits, specialty tools, diagnostics, and service aids for the electrical, telecom, cable television,

data, wireless and audio visual alarm communications industries. Many new products are featured, including the new line of JTS test sets. Jensen's customer benefits include same day shipping (free for orders over $100), a 30 day unconditional guarantee on all products, lifetime guarantee on Jensen brand hand tools, free technical support, a 24 hour/7 day FaxBack® service, and a complete online catalog with secure ordering at www.jensen-tools.com. Jensen also specializes in the design and development of custom tool kits and cases.

Jensen Tools
Circle #188

Products

Insulated tuning tools

Sprague-Goodman Electronics, Inc. offers a new Engineering Bulletin featuring its complete line of Insulated Tuning Tools. The bulletin includes data on several models which were recently added to the line, designed to fit the narrower slots in the newest miniaturized trimmer capacitors. Steel tips on all Sprague-Goodman insulated tuning tools are hardened and plated, and feature tough molded plastic bodies. The economical GTT Series includes both antistatic and nylon body styles with either metal or plastic tips.

Sprague-Goodman Electronics
Circle #189

New buyer's guide product catalog

Hub Material Company (HMC) introduces a new buyer's guide product catalog containing a variety of technical supplies for designing, assembling, testing and repairing electronic products and subassemblies. Products include precision hand tools, soldering and desoldering supplies, static-control products, test equipment, measurement and inspection instruments, workstation and PC board handling items, adhesives and cleaning chemicals, tool cases and custom tool kits. The guide offers color-coded comparison tables of product features, and large product photos.

Hub Material Company
Circle #190

Product summary brochure

Applied Signal Technology, Inc. has recently released its new 2000 Product Summary brochure. The products are arranged according to common processing functionalities. For example, Section A presents Voice Grade Channel Processing products, Section B presents Wideband Digital Processing products, and so on. The summary is designed to help customers better understand the capabilities of the products and the performance differences among similar products. The company features hardware and software products developed through aggressive research and development efforts as well as from hundreds of development contracts. Included in the brochure is a brief overview of Applied Signal Technology history.

Applied Signal Technology, Inc.
Circle #191

Website catalog update includes newest products

Microlab/FXR has completed a website update to include data sheets on all catalog products.

Information is available in Adobe Acrobat™ format for convenient viewing and/or downloading. The catalog includes the newest products for wireless designs, including unequal power dividers, DC blocks and lowpass filters. The address is www.microlab.fxr.com.

Microlab/FXR
Circle #192

EMI/RFI Radiation and Susceptibility from Cables and Enclosures

By Gary A. Breed
Publisher

Here are a number of notes on the control of unwanted radiation from electronic equipment as well as protecting that equipment from the effects of outside interference. In the U.S., FCC regulations place limits on the field strength of radiation from electronic devices of all types. In Europe, additional regulations specify that equipment should operate properly when exposed to fields of 1 V/m to 3 V/m.

Compliance with regulations can require significant investment in manpower, test equipment or services, EMI shielding and other components. Time-to-market can suffer greatly if the initial design is not robust from an EMI standpoint. A friend who designs high-speed DSP-based products claims that achieving regulatory compliance takes longer than the functional design. As he has learned methods for reducing radiation at the component, board and system level, the extensive testing and rework time is gradually diminishing.

Cables and connectors

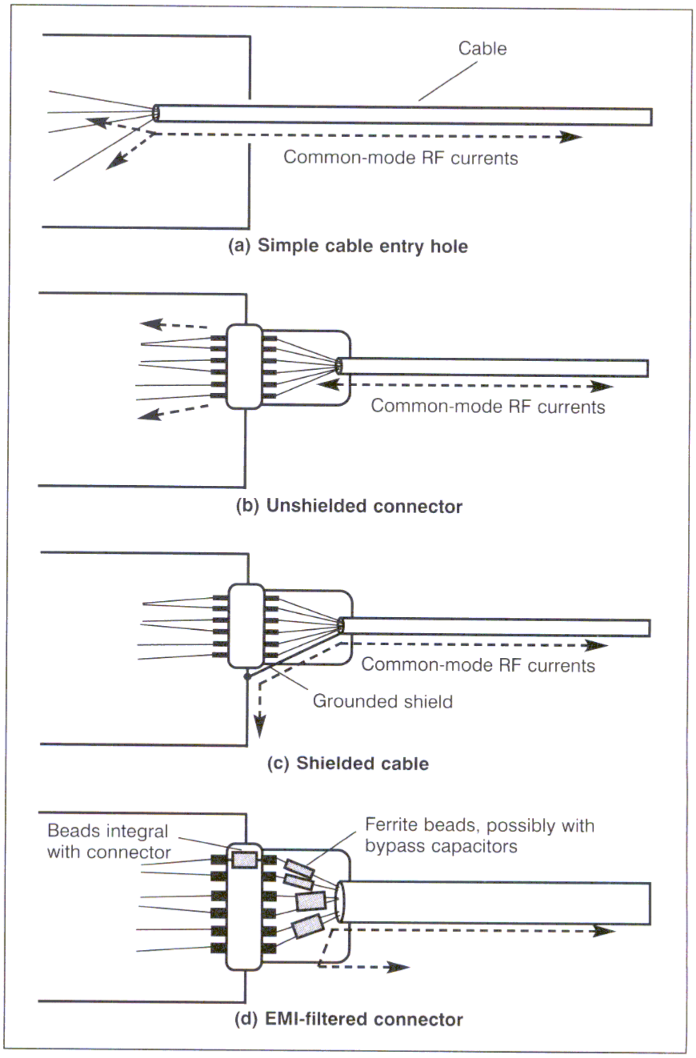
Cables bear a similarity to antennas, which is both a problem and the key to understanding their radiation and susceptibility. These interconnections between modules or enclosures, from power sources and to antenna systems are extremely important, as they represent one of the main ways that unwanted energy gets radiated from within an otherwise "shielded" enclosure.

Common-mode currents are the primary issue in cables. These are the currents that flow on the outside of a shield, or along with an unshielded "twisted pair" cable. The desired signals carried by the cable are in differential mode, and unless the conductors of the cables are somehow separated to interrupt their transmission line behavior, they will not be radiated.

Common-mode currents must be dealt with at the point where they enter or leave an enclosure. Figure 1 illustrates the most common problems and possible solutions. In (a), which is the worst case, a cable simply enters an enclosure through an access hole. The cable can easily couple to energy contained in the enclosure and carry it outside where it can radiate, or bring currents generated by external fields into the enclosure to cause interference.

Figure 1(b) is a more subtle case where a connector is used for unshielded cables, or with shielded cables that do not have their shields grounded at the connector. This installation gives the appearance of being robust, and the connector may even have a metal shell that provides shielding. However, common-mode currents will not be impeded and the performance is the same as that shown in Figure 1(a).

The first, and simplest solution is the use of a shielded cable, typically a braid that encompasses the entire bundle of individual cables. This overall shield is bonded



▲ **Figure 1. EMI/RFI performance of common cable entry methods.**

to ground at the entrance to the enclosure, keeping currents on the shield from entering the box. Figure 1(c) is a sketch of this technique.

A more elaborate technique is shown in Figure 1(d). Here, the individual cables within the bundle are filtered at the connector. Special connectors are available that include ferrite materials, or even L-C decoupling circuits, avoiding the need to hand-assemble filters adjacent to the connector.

Figure 1 is a simplified description of the general problem of cable entry, which applies to signal lines, AC and DC power connections and all other cables that are attached to the enclosure. There are also other methods for reducing the radiation and susceptibility of cables outside the enclosures that they are serving.

One simple, yet effective practice is routing the cables along a grounded metal surface, such as a metal equipment rack. Cables that simply cross open areas between rack systems will be much more prone to unwanted radiation and pickup. If they are routed along the metal support members or flat sides of a rack, common-mode behavior is reduced, since the combination of cable and surface acts like a transmission line, just like a microstrip's conductor-over-ground plane configuration. Thus, the interconnecting cable is no longer an "antenna" since currents are contained between the "transmission line" conductors.

EMI control inside enclosures

These same rules apply to the circuit elements within an enclosure, especially the interconnections between circuit boards, power supplies, etc. Similar techniques regarding shielded cables, grounding at the point where the cable leaves a p.c. board or applying EMI filtering will reduce the energy circulating within the enclosure. This may not change the external radiation that is measured for regulatory compliance, but it will reduce the possibility of self-interference.

In addition, enclosures require attention to the interruptions in their shielding, including all openings, gaps, panels, doors and mechanical bonds. Put simply, any gap in a metal surface is a slot antenna that can couple energy between the inside and outside of an enclosure. The size and shape of the gap determine its frequency response. A 6-inch long slot will pass 900 MHz cellular energy with very little attenuation!

Be careful with painted enclosures. The layer of paint may be an insulator that prevents conductive contact between panels, resulting in an unexpected "slot antenna" that is 12 or 18 inches long.

At a minimum, removable or assembled metal parts should have clean, bare metal surfaces where they contact one another. Access doors or panels that may be removed regularly should have EMI control gaskets that are designed to maintain electrical contact after repeated operations.

Why worry so much about RFI?

The most obvious reason for EMI/RFI control is to meet the regulatory requirements of various governmental bodies. For the past ten years, interest in standards and regulations has increased as the European Union implemented some of the world's most stringent standards.

Presently, all industrialized countries, plus many developing nations, are working toward "harmonization" to avoid dramatically different standards. The global economy is dependent on access to many different markets. If products must be built to meet widely varied EMI standards, then it may not be feasible to offer them in all markets.

These regulatory requirements are neither arbitrary nor designed to restrict market access. They have a solid practical basis: the reduction of interference. Interference among electronic devices is regulated for reasons ranging from simple consumer protection to public safety. Not only do we want to assure a minimum level of protection against products that are poorly designed and constructed, we also want to avoid interference with communications.

The European Union has added a protection factor that is not yet present in the U.S. — immunity to interference from external RF fields. There are many sources of significant RF energy in our environment, and the EU has determined that it is important to assure that electronic products have a reasonable level of protection from their effects.

For example, in the U.S., the single largest source of interference complaints to the Federal Communications Commission is interference to telephones by radio and television stations, CB operators, two-way radio, ham operators and other licensed radio services. Telephones are not intended to be radio receivers, and if they were required to comply with EU standards for immunity to interference from radio signals, the complaints would almost completely cease.

Summary

As more wireless communication services are offered to the public, the potential for interference will increase. The growth in other types of electronic equipment such as computers adds many more possible sources and victims of interference. It is both in the public interest and good business to design and build equipment that does not radiate unwanted signals and is robust enough to avoid being an unwanted "receiver" of interference. ■

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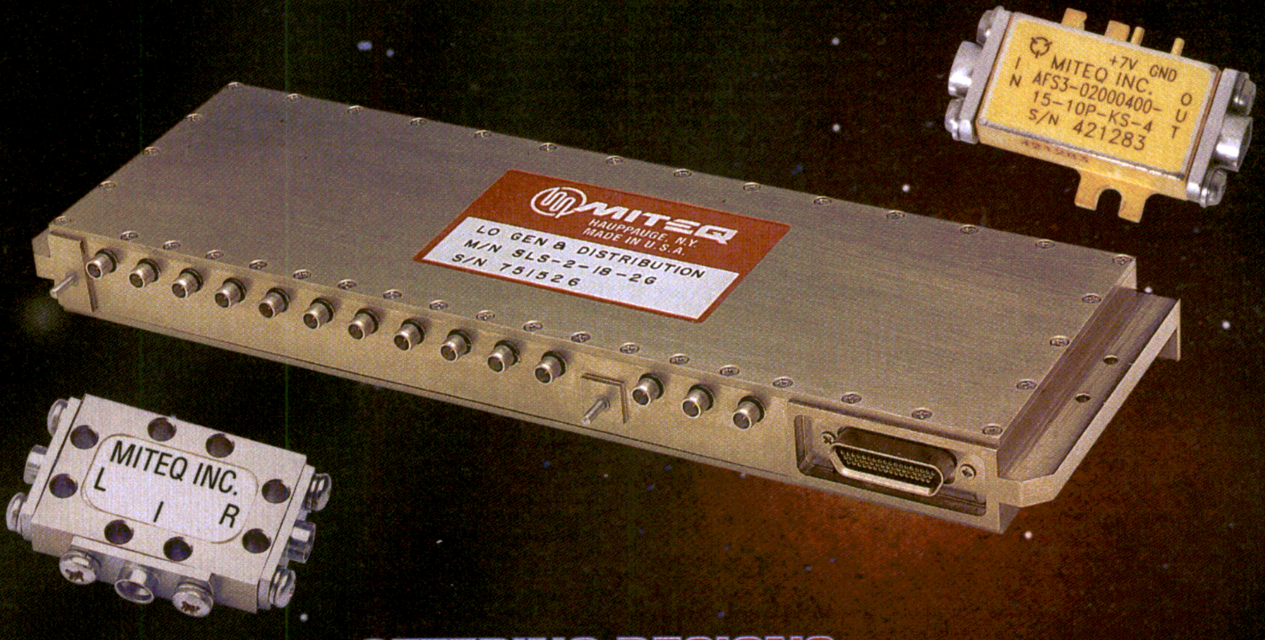

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Circle 14

Second Order Effects in Feedforward Amplifiers

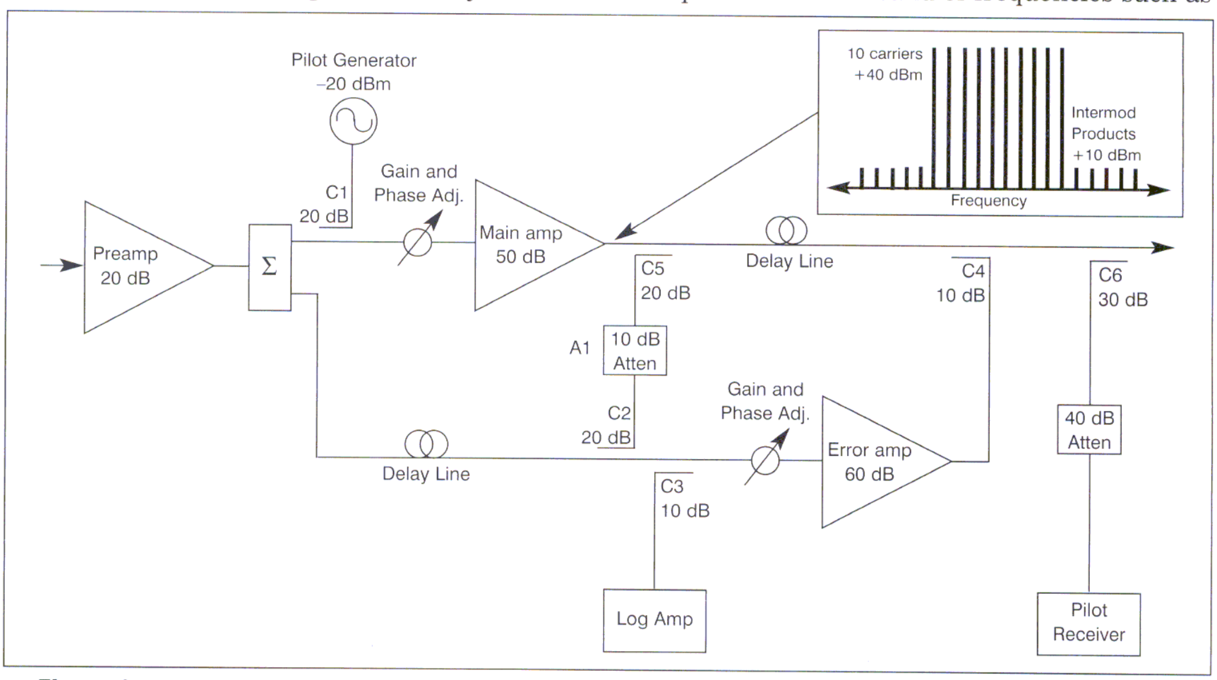
This article looks at the more subtle challenges in implementing feedforward linearization

By Barney Arntz
Arntz Design

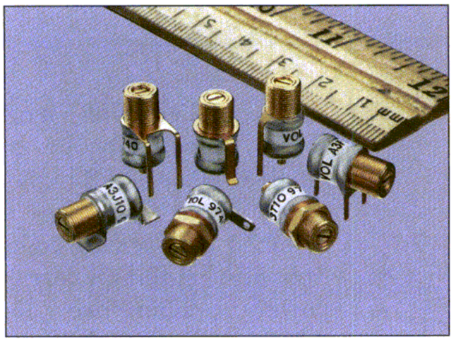
The use of feedforward linearization in power amplifiers has become widespread in the wireless arena, because it offers the ultimate performance in terms of intermods and spectral regrowth. It also provides the ultimate in distortion correction. It is often used with some form of predistortion. Feedforward can be used for amplifiers from under a watt to several hundred watts and beyond, and with almost any modulation technique. In the early 1990s it was used in wireless applications primarily with FM multicarrier modulation. Now the feedforward technique is being used for multiple CDMA channels through one amplifier, GSM, TDMA, and has potential for use with wideband CDMA (third generation) systems.

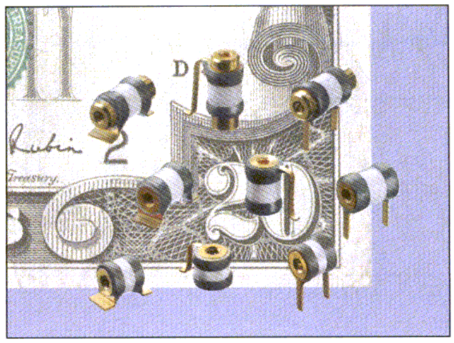
While the architecture of the standard feedforward amplifier is fairly straightforward on paper, significant challenges must be overcome in practical amplifiers. Controlling phase to less than a degree over a band of frequencies is just one problem if distortion terms are to be 60 or 70 dB down from the main carriers. Considering that a typical regrowth spec of –60 dBc means that the distortion energy is on the order of 0.1 percent of the signal energy, in voltage units (if the distortion bandwidth and the signal bandwidth were the same), this distortion performance is equivalent to the best audio amplifiers, but the frequency of operation is five or six orders of magnitude higher.

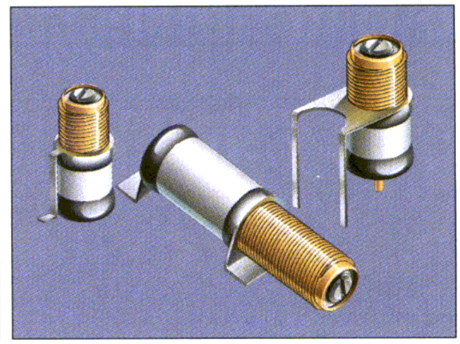
Operation over a band of frequencies such as



▲ **Figure 1. A typical feedforward amplifier and basic signal levels.**







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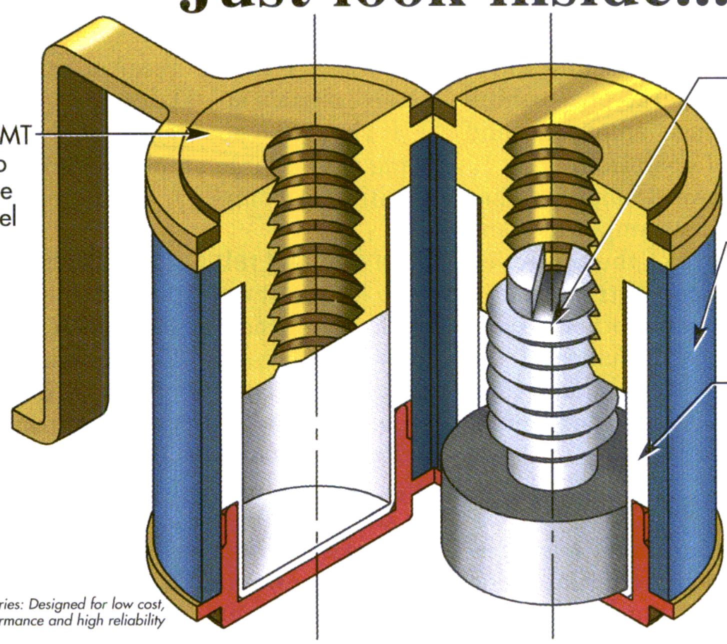

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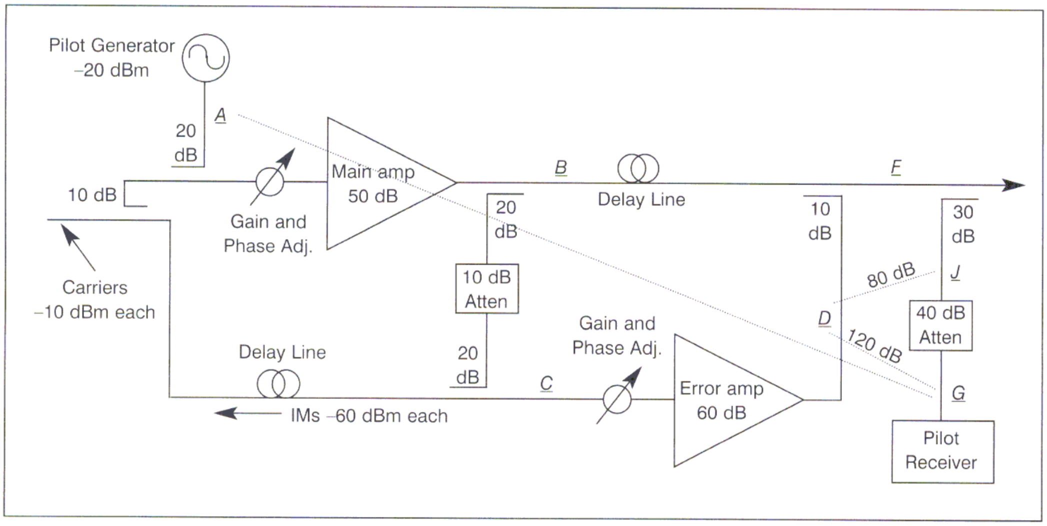

▲ **Figure 2. Pilot isolation issues and preamp s_{22} issues.**

20 to 60 MHz becomes difficult in light of the fact that the core power sections of the amplifier have to be run with tuned (rather than broadband) matching networks. This requirement is driven by efficiency, and this tuning results in band edge rolloff of both gain and phase. Also, delay lines are used in both the low power and high power paths, whose electrical lengths are many wavelengths long. These delay lines have large phase shifts over the operating band and produce amplitude and phase ripple throughout the band of operation.

Much of the literature on feedforward amplifiers has concentrated on issues of efficiency, amplitude and phase ripple, delay mismatch, the size of the correction amplifier, and more. This paper addresses the issues of shielding, coupler directivities and some other second order effects which can limit the performance of the amplifier. It will be shown that in some cases more than 100 dB of shielding may be required between some of the circuit elements.

Analysis

Figure 1 shows a general block diagram for a feedforward ampliflier, as well as typical signal and intermod levels. For analysis, this example uses ten equal power CW (or FM) tones, each at 10 watts, for a total power of 100 watts (+50 dBm). The intermods are shown coming out of the main power amplifier at –30 dBc. The pilot signal, injected before the main power amplifier, is set to be equal to the intermods. This is a convenient level, high enough to be measurable but not so high that if it becomes partially re-established, it will not show up significantly at the output. The gain of the error amplifier is 60 dB, equal to the sum of couplers C2, C5, C4 and A1.

A log amplifier/detector is connected to a tap off point at the first cancellation point. This would be used to adjust the first loop gain and phase, for a minimum signal at that point. When unadjusted with full carrier power applied, this signal could conceivably be as much as 80 or 90 dB higher than when the loop is adjusted for good cancellation with minimum carrier power. Therefore, log detection is normally used to cover a wide range.

It is common to use a pilot generator and receiver for loop adjustment. Like the log amp, the pilot receiver has to cover a wide range, and it has to detect a cancelled pilot that is perhaps 70 dB down from a group of carriers. A full fledged superhet receiver is generally required (sometimes with two down conversions), with a mixer input level set to trade off IP_3 and noise performance. A 30 dB coupler and 40 dB pad are shown before the receiver, to prevent it from overloading and creating its own intermods. In this example, the maximum mixer level is set to –20 dBm for all carriers combined. The pilot receiver would then be required to receive signals at –90 dBm.

It should be apparent that the amplifier will be operating with a large amplitude difference between signals at different points.

Shielding and return loss issues

Figure 2 shows some of the potential pitfalls from inadequate shielding, and also from an inferior return loss on the pre-amp. The configuration of the input splitting is different from that of Figure 1, to highlight a problem when that type of splitting is used.

First, we look at the pilot isolation. Suppose the pilot generator were operating at –20 dBm as shown. This is by definition operating at the same frequency as the pilot receiver. If the receiver is to detect a signal at –90 dBm, the unwanted crosstalk through the air or through the ground system should be below –100 dBm, at least. This means that the shielding between the pilot generator and receiver (points *A* to *G*) must be at least 80 dB.

Now lets look at points *D* to *G*. The pilot level at point *D* is +20 dBm, to cancel the pilot signal coming out of the main amplifier at +10 dBm (there is a 10 dB loss in the output coupler, from the coupled port to the output). The isolation between points *D* and *G*, then, must be 120 dB. This is not achievable without considerable care, and manufacturing methods that are optimized for cost or ease of assembly rather than isolation might not be adequate. The isolation requirement from point *D* to *J* is only 80 dB, because of the 40 dB attenuator.

An interesting problem occurs with inadequate isola-

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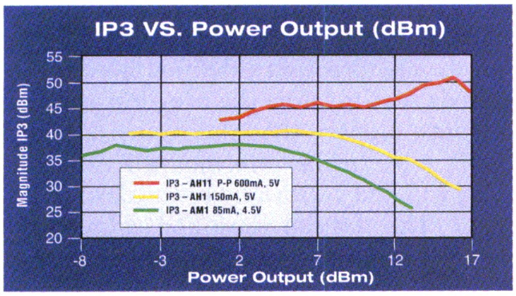

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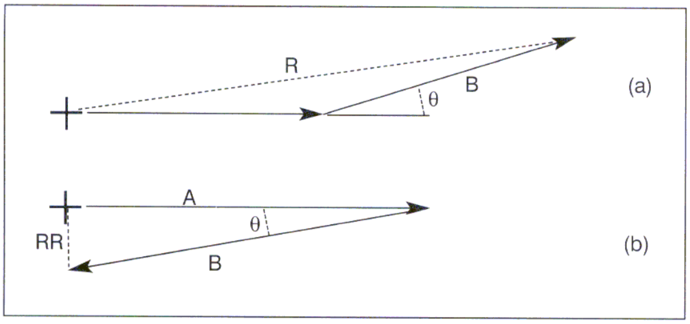
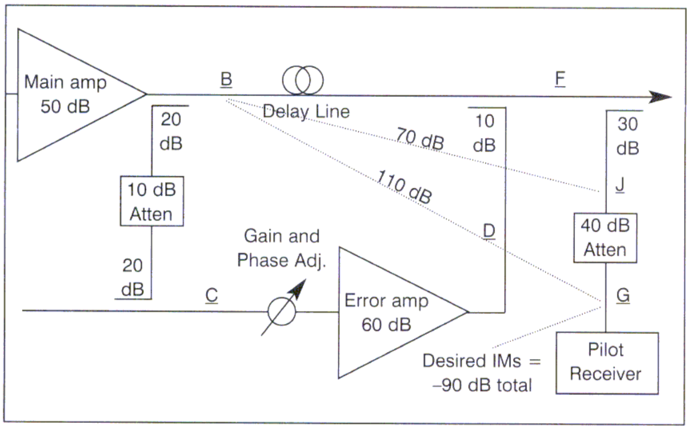


▲ **Figure 3. Vector relationships for signal addition and subtraction.**



▲ **Figure 4. Pilot receiver isolation issues.**

tion from points F to G. At the pilot frequency, if there were leakage between those two points that were 70 dB down, this would combine with the desired "coupler plus attenuator" coupling to create a signal anywhere from zero to twice the correct signal, in volts (6 dB above the correct signal). This would become a problem if the undesired leakage path changed in dB value, as a result of a pilot frequency change, connector tightness, housing screw pressure, etc. Even if the undesired leakage signal were 10 dB below that of the desired path, the resultant could vary from –3.29 to 2.38 dB from its nominal, as a function of the relative phases of the two signals combined. This is shown in Figure 3. The lower valued signal, B, is 10 dB below that of A (a voltage factor of, or 0.316).

The two dotted lines show the two extreme cases of the vector addition, where the resultant signal R is 0.864 times the main signal amplitude in the minimum case, and 1.316 times the main signal in the maximum case. In general, the magnitude of the resultant R, in voltage terms is

$$R := 2 \cdot \frac{\sqrt{(A + B \cdot \cos(\theta))^2 + (B \cdot \sin(\theta))^2}}{A + B} \quad (1)$$

where A and B are real and θ is the deviation from the ideal. The maximum and minimum values of R over all phases can be found with θ equal to 0 and 180 degrees, or simply

$$R_{MIN} = A - B \text{ and } R_{MAX} = A + B, \text{ if } A > B. \quad (2)$$

Limits of cancellation

Figure 3b shows two vectors which are close to cancellation. This typifies what would be at point C with a single carrier applied, with imperfect gain and phase match. The limit on cancellation is represented by the length of the resultant, which is equal to

$$RR := 2 \cdot \frac{\sqrt{(A + B \cdot \cos(\theta + \pi))^2 + (B \cdot \sin(\theta + \pi))^2}}{A + B} \quad (3)$$

in voltage terms. In dB it is 20 log(RR). For example, if the two unity amplitude vectors differ by 1 degree, the cancellation is 35.2 dB (in general, the cancellation is referred to the vector amplitudes). For small angles and unity amplitude vectors, the cancellation is approximately

$$RR = 20 \log (\sin \theta) \text{ in dB.}$$

Although unrelated to the problem, Figure 2 also shows how inadequate return loss from the pre-amp and imperfect directivity in coupler C2 can combine to cause a subtle problem. The intermods come out of the main amplifier at –30 dBc or +10 dBm (each intermod). They couple through C5 and C2 and the 10 dB attenuator will cause a signal at C of –40 dBm. The reverse coupled signal, which happens because of the imperfect directivity in C2, is down by the directivity, defined here to be 20 dB. The intermods sent towards the pre-amp are thus at –60 dBm. Now lets suppose that the return loss of the pre-amp is 15 dB. A particular intermod will reflect back at –75 dBm, and will add vectorially to the desired intermod at point C. The desired intermod coupled down from the power amp is –40 dBm at that point. Therefore, the "leakage" intermod is 35 dB down from the desired intermod (this figure is also equal to the sum of the directivity and the return loss of the pre-amp). Although this seems low enough in value, it must be recognized that there is hopefully 30 dB cancellation of the desired intermods at point F, leaving the leakage intermods only 5 dB down from the residual intermod left by imperfect cancellation. In effect, this limits cancellation to 35 dB. In practice that is about as far as one can expect under the best of conditions, but this error mechanism is pointed out as an example, which in other configurations could be a problem.

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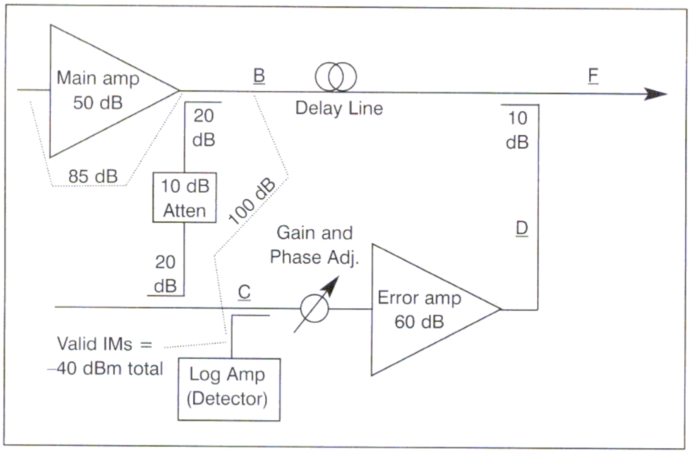


▲ **Figure 5. Log amplifier/null detector isolation and main power amplifier self-isolation.**

Power amplifier to receiver isolation

Figure 4 shows a configuration with no pilot generator. Here, the receiver acts as a spectrum analyzer, constantly sweeping the spectrum and using programmed intelligence and knowledge of the carrier characteristics to ascertain the overall distortion, which is used in the controller to adjust the loops. The isolation from B to G must be 110 dB to have a 10 dB margin between the desired intermod at –90 and the leakage intermod, which starts out at +10 dBm and couples down through leakage to –100 dBm.

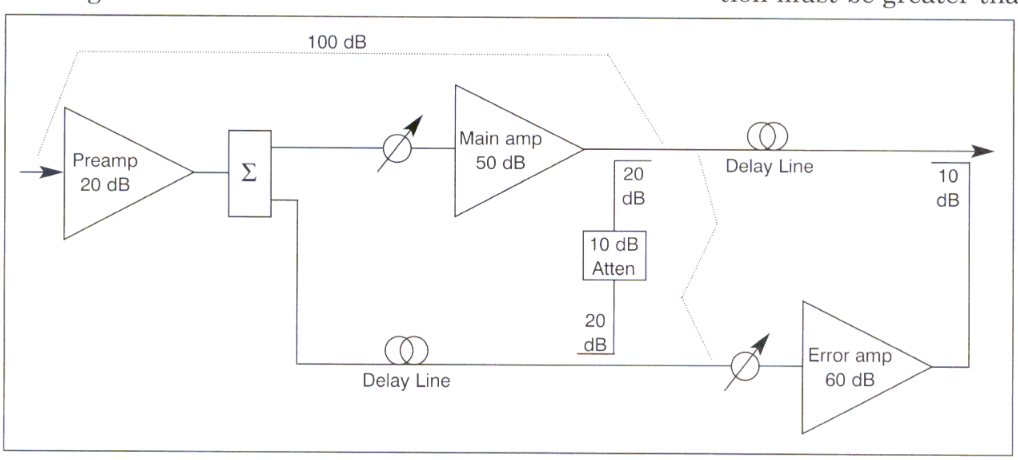
Figure 5 relates to the main amplifier output shielding. If some of the output leaks back into the input, it will not oscillate unless the leakage is severe. However, it can cause amplitude ripple across the operating frequency, because the phase of the two signals adding at the input will typically go through many cycles of phase as the frequency is changed. There is constructive/destructive addition as shown in Figure 3. For the case shown, with 60 dB of gain and 90 dB of output to input leakage, the B term in Figure 3 would be down 30 dB or a voltage factor of

$$10^{-\left(\frac{30}{20}\right)} = .0316$$

relative to A, which is unity. According to Equation 2, the resultant R has maximum and minimum values of 1.0316 and 0.968 respectively, or approximately ±0.27 dB. This is quite a lot. For cancellation in the first loop, the gain of the amplifier from point E to point C has to match that of the passive path from the same two points. With amplitude ripple, the best adjustment might be where cancellation is optimized for the average gain at all frequencies. Then, at a particular frequency, the gain could be high by 0.27 dB. This would limit the isolation to 30 dB, a figure which equals the "loop" gain from the input of the main amplifier back to the same point via the leakage path.

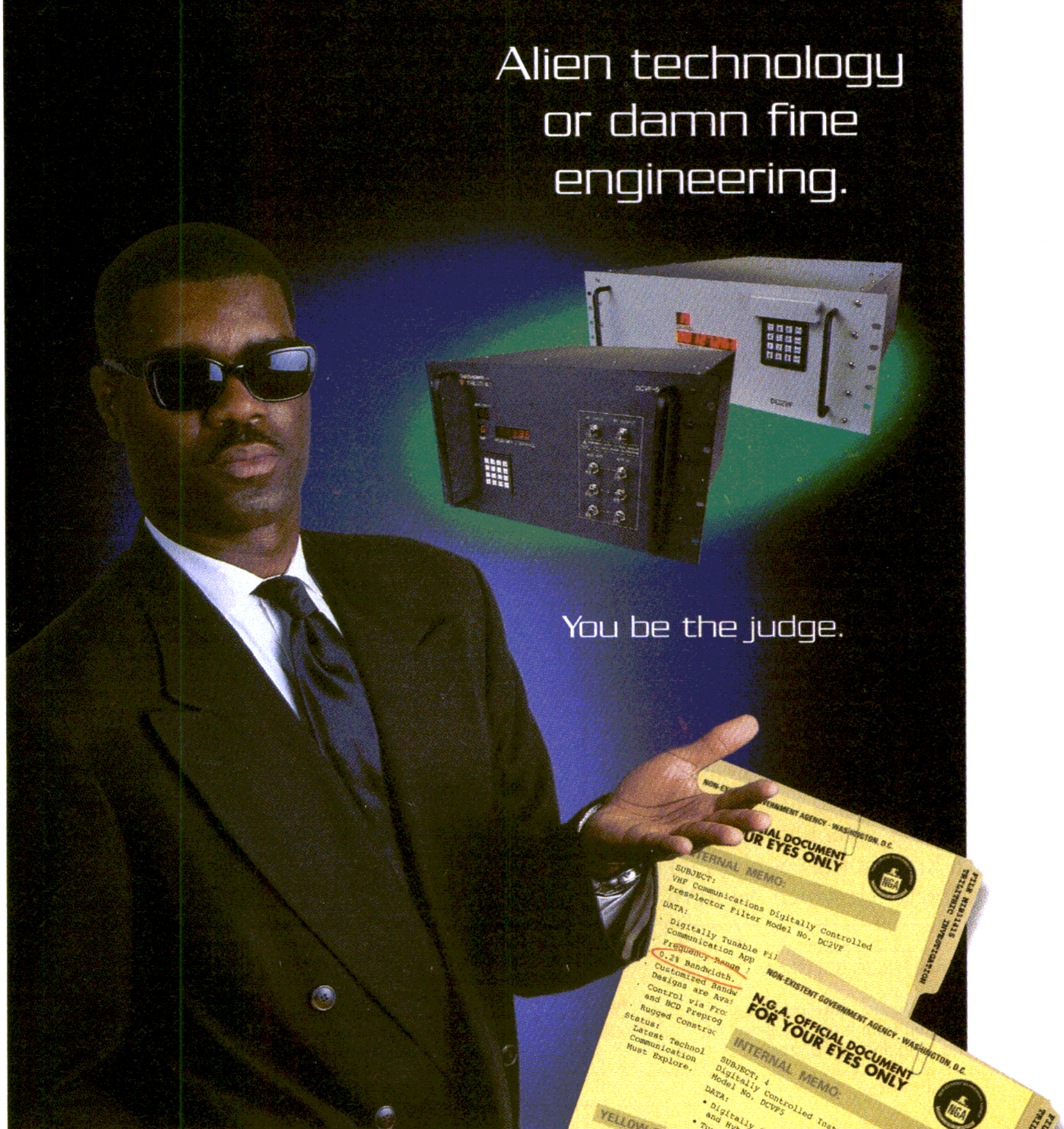
Figure 5 also shows the requirement for the main amplifier output shielding to the log amplifier. With full carrier power applied, and 30 dB cancellation in the first loop, the "valid" intermods are –40 dBm for all carriers. The total power at point D is +50 dBm. Therefore, for an interference-to-desired signal ratio of –10 dB, the required isolation is 100 dB.

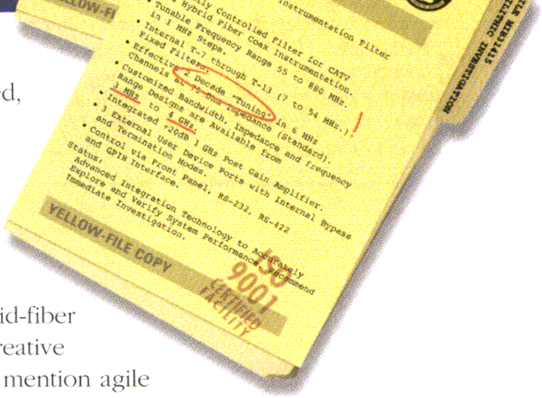
Figure 6 shows the requirement for isolation from the main amplifier (point D) to the input of a 20 dB preamp. Intermods that reach that point will come out of the feedforward amplifier, since that point represents the "clean" reference for the error amplifier. Also in Figure 6, there is an issue of leakage from point D to the input of the error amplifier. The carrier leakage causes the error amplifier to amplify and inject carriers into the coupled side of C4. This is a minor problem for the output of the feedforward amplifier (the carriers just add), but a larger issue is that the power in the error amplifier goes up. This amplifier is sized as small as possible to get the job done, and extra carrier power through it may cause intermods to be generated in it. (This problem can also be caused by insufficient cancellation at point C.) For the carrier power due to leakage to be 5 dB below that of the intentional intermods at point C, the isolation must be greater than 85 dB.



▲ **Figure 6. Main amplifier output leakage.**

An additional problem with leakage from point D to C is that the error amplifier sees intermod terms that the log amplifier does not see, and vice versa. The log amp will thus not be adjusted (via the control processor) to a point of cancellation/null at C. Finally, the frequency flatness from D to C becomes an issue because of the two independent paths, whose phase will not likely track with frequency, as mentioned previously.





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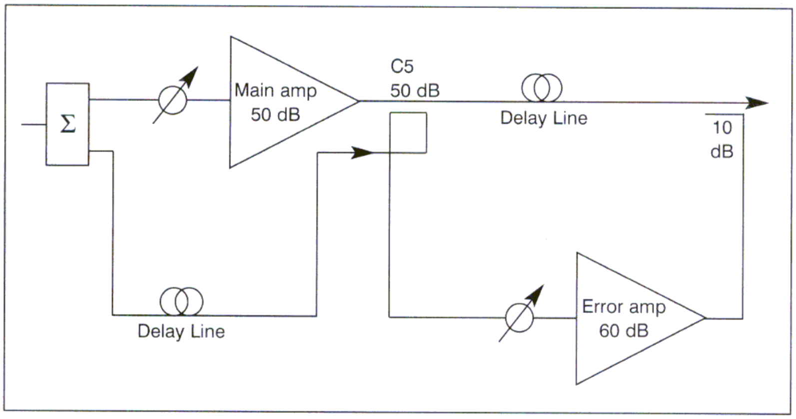


▲ **Figure 7. Alternate feedforward architecture.**

Figure 7 shows an alternative arrangement of couplers which is sometimes seen. This configuration saves one coupler. The addition of the tapped off distorted signal from the main amplifier to the clean reference happens in that coupler. This puts tremendous constraints on the coupler. First, it has to have a coupling of 50 dB without using resistive attenuators. The wide spacing required of the RF lines in the coupler makes it difficult to achieve good directivity, which we have seen to be a problem. Also, adjusting the coupling value is not easily done, due to the lack of an attenuator.

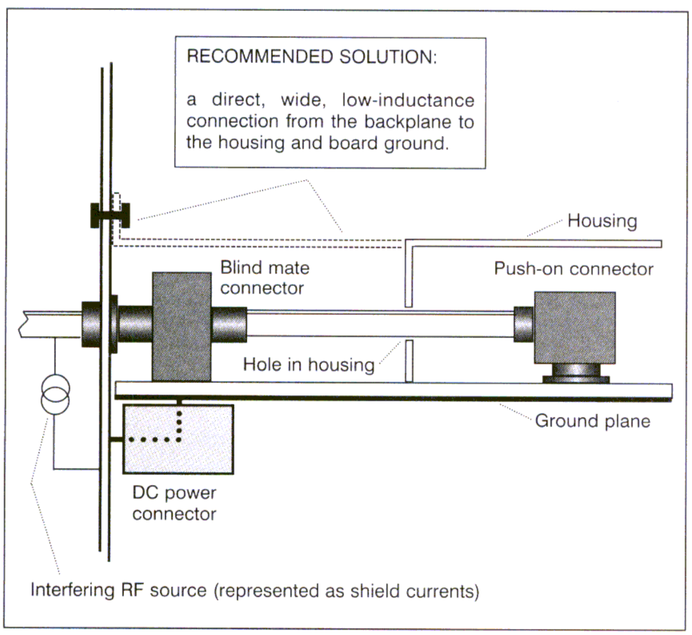
Typical grounding issues

Achieving isolations on the order of 100 to 120 dB does not happen automatically, especially in cost-sensitive applications, which minimize extra screws, housings, bulkhead connectors and so forth. Simplifications such as board-mounted (rather than housing-mounted) connectors and backplane connectors are a particular problem if not implemented correctly.

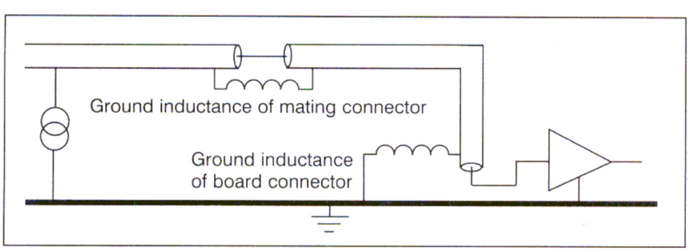
Figure 8 shows a situation with both of these issues. There is connector ground inductance with push on connectors because the number of points of contact is limited. This may amount to tens of picohenries, and this inductance is different from the rest of the cable and presents a discontinuity. A shield current caused by an interfering source (Figure 9) causes a drop across that inductance. This shield current may be due a nearby radiator, but more likely (if everything is shielded) from a difference in potential between two housings or ground planes, caused by heavy RF currents elsewhere.

The shield current and connector inductance create a voltage drop which trickles down to the destination on the center conductor. The destination amplifier or circuit block is sensitive to voltages relative to its local ground connection.

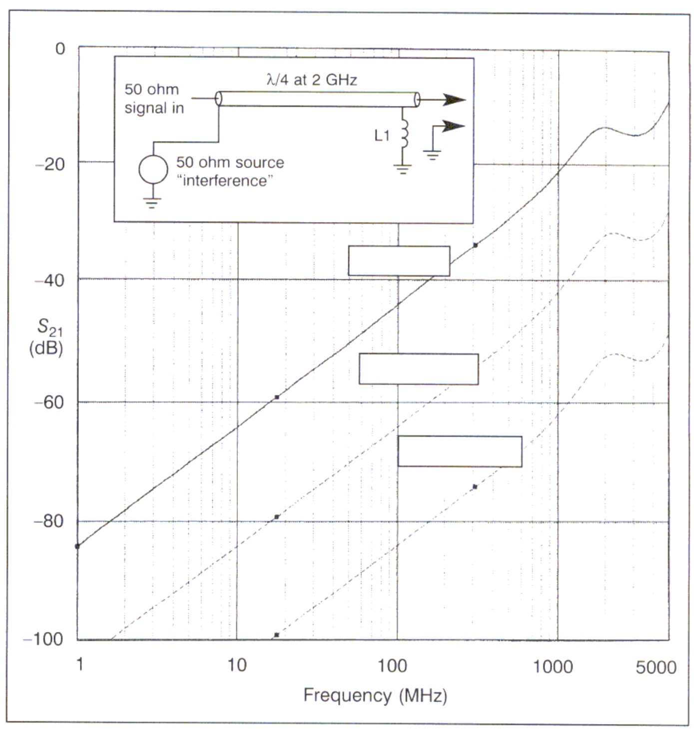
An inductance of 20 pH is 0.25 ohms. If an interfering source of 0 dBm (0.226 V) from 50 ohms were applied to the shield as shown in the figure, the entire current



▲ **Figure 8. Ground problems in typical construction.**



▲ **Figure 9. Electrical model of ground inductance.**



▲ **Figure 10. Analysis of a single ground inductance.**

would make its way to ground. This current is 0.226/50 =.0045 A, and the IZ drop is .0045 * 0.25 = 1.12 mV. This is equal to –46 dBm.

The board connector shown in Figure 8 poses a similar problem. A surface mount connector has to be grounded through vias. There are sometimes mechanical requirements such as thermal breaks which limit the number of vias at the connector, especially right under the connector where they are needed most. The via inductance then becomes an issue.

A solution to much of this is to provide a connection from the backplane ground to the board ground that is more solid and larger than connectors provide. This is shown in the figure. Although this creates a local ground loop, in the sense that there is more than one ground path, the performance is much better. Measurements show many tens of dB of improved shielding with this method.

Simulation shows similar results as shown in Figure 10. This simulates just one of the ground inductances at the board connector. The plot shows s_{21} in dB.

Conclusion

The shielding, isolation, and directivity requirements in feedforward amplifiers present challenges in terms of manufacturing methods, component packaging, and cabling. Although the input signal may be 0 dBm and the output may be +50 dBm, there are signals inside the typical pilot-corrected amplifier which are in the region of –90 dBm. This low level is the result of how log amplifiers and superhet receivers operate, and also the result of the dynamic range of the signal often entering the feedforward amplifier.

Isolation requirements between two specific points in the system can be as much as 120 dB. This is not normally achieved with ordinary inexpensive construction methods, especially at the higher frequencies now being used in cellular systems. One example of a backplane to board interface and its potential pitfalls has been discussed, with remedies to minimize the amount of crosstalk picked up and inadvertently injected into the board.

Author information

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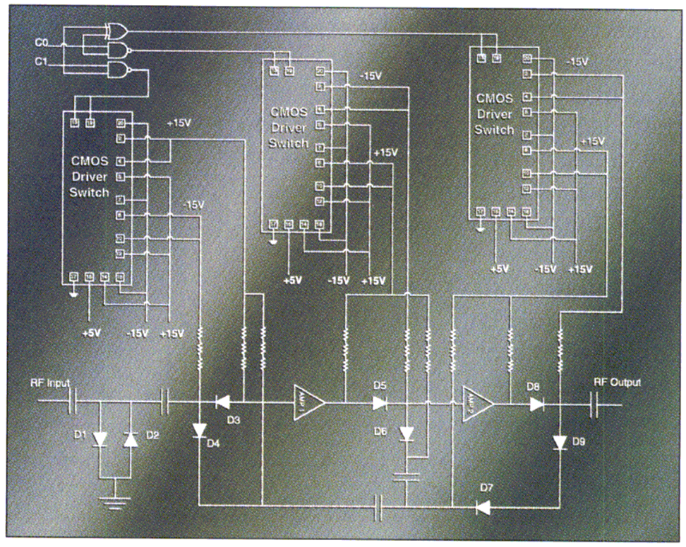
Selectable-Gain Amplifiers Launch Integrated Subsystem Line

This month's cover features multi-function integrated modules that help designers maintain performance and control quality

Cougar Components Corporation

As part of its subsystems development plan, Cougar Components Corporation has designed, manufactured and delivered three different high dynamic range selectable gain amplifiers. The three designs cover the frequency bands of 0.5 to 30, 30 to 100 and 100 to 1000 MHz. Each amplifier has three different gain states. The two lower frequency bands use balanced class A push-pull bipolar amplifiers, obtaining high 3rd order and 2nd order intercept points. Cougar uses bipolar transistors to obtain noise figures lower than GaAs FETs provide at frequencies below 100 MHz. The amplifier for the 100 to 1000 MHz band unit uses a single-ended GaAs FET design.



▲ **Figure 1. PIN diodes control the signal path through two amplifier stages, selecting one, two or none.**

A major advantage engineers achieve by using selectable gain designs is minimizing dynamic range degradation for each gain state. Mode 1 provides 20 dB gain, and Mode 2 provides 10 dB gain while Mode 3 serves as a bypass mode. Each mode provides the same output linearity. The higher an amplifier's gain the lower the dynamic range on a dB-for-dB basis. The amplifier remains effectively linear over an additional 20 dB range by providing the same output intercept point for each gain state. If the amplifier's gain remained constant at 20 dB, the amplifier would require +95 dBm 2nd order intercept and +64 dBm 3rd order two-tone intercept points, which are not practical levels.

Cougar optimizes the bias points to achieve the best 2nd order and 3rd order two-tone suppression. The two lower band amplifiers' output power is nearly +27 dBm, and the high band amplifier is +25 dBm. The typical 2nd order two-tone linear intercept point is +75 dBm for the two low band amplifiers. The typical 3rd order two-tone intercept point for the low band amplifiers is +44 dBm. The high band amplifiers' typical 2nd order two-tone linear intercept point is +60 dBm, with +39 dBm 3rd order two-tone products.

Figure 1 shows PIN diodes D3 through D9 used to control the RF signal path. The diode's transit time is very long for the low band amplifiers in order to maintain linearity and PIN chactericties to 0.5 MHz. The switching waveform does not remain clean if the diode lifetime is too short. The CMOS high-level switch driver must supply sufficient bias current when the diodes are in the "on" state to keep the distor-

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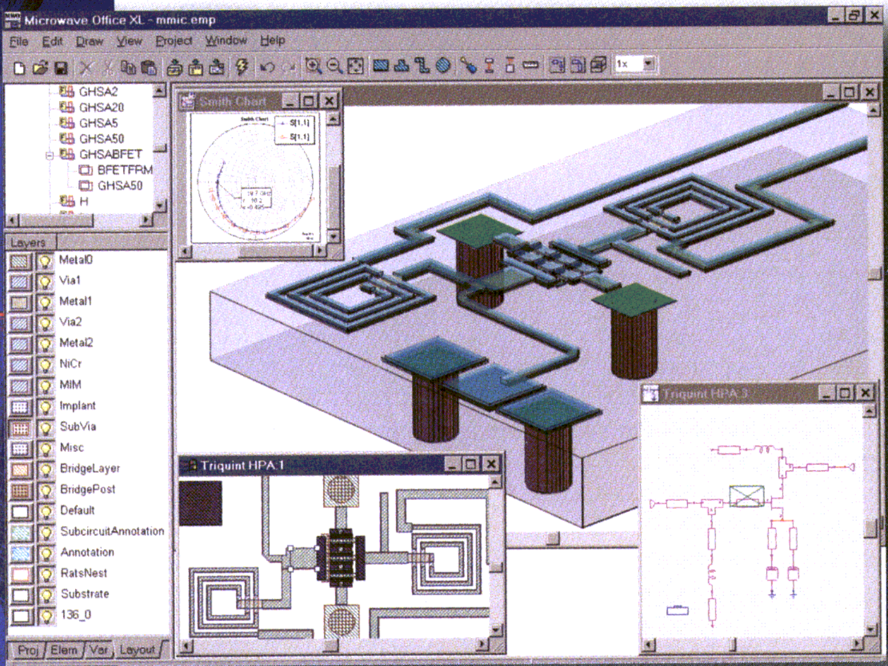

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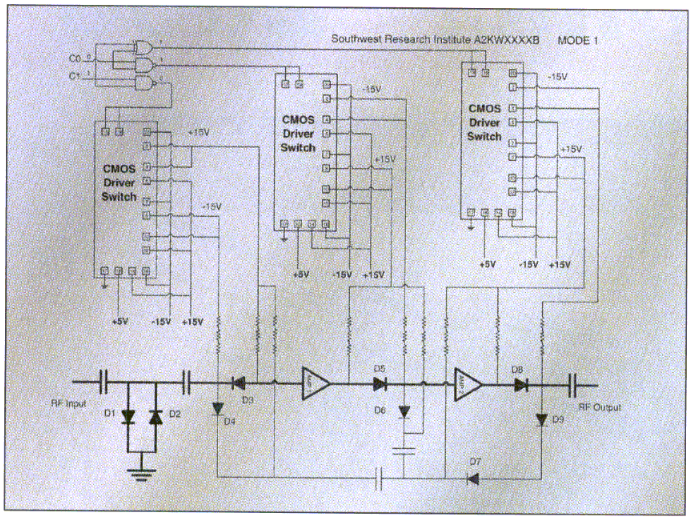
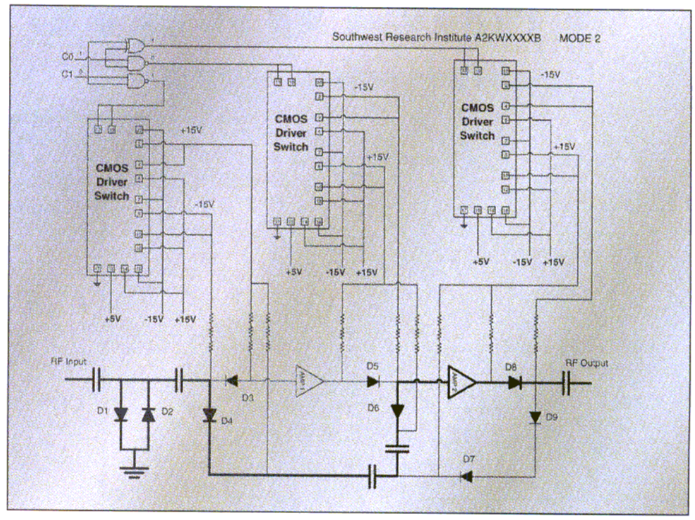


▲ **Figure 2. Mode 1 routes the signal through both amplifiers, providing 20 dB gain.**



▲ **Figure 3. Signals are amplified by one stage in Mode 2, providing 10 dB gain.**

tion products well below the level generated in the amplifiers. At the same time, the "off" diodes must be held off by a sufficiently high back bias voltage so that the large forward RF signal swing does not cause diode conduction. Even very small amounts of conduction on a back-biased diode will have deleterious effects on the linearity.

Figure 2 illustrates the Mode 1 RF path. Diodes D3, D5 and D8 are all biased "on" to route the signal through both amplifier stages. D4, D6 and D9 are all biased "off." The total gain in this configuration is a nominal 20 dB, less the losses of the internal switching circuitry. Diodes D1 and D2 are used for signal limiting to protect the amplifiers from high input power levels.

Figure 3 shows the Mode 2 RF path. Diodes D4, D6 and D8 are biased "on." Diodes D3, D6, D7 and D9 are biased "off." These settings allow only the 10 dB gain of Amplifier 2 to be in the circuit.

Figure 4 shows the bypassed condition of Mode 3. Diodes D3, D5, D6 and D8 are biased "off" to disconnect the amplifiers from the circuit, while D4, D7 and D9 are all biased "on" to create the desired signal path.

The figures also show the logic gates, which operate from straight TTL input. Table 1 is a summary of the amplifier family's specifications.

Cougar Components has built, RF aligned, and tested several hundred of these mini-subsystem modules. This design demonstrates ther company's capability in com-

Model	Freq. Range MHz	Small Signal Gain dB			Gain Flatness ±dB		Noise Figure dB			Power Output at 1dB Compression dBm			Rev. Iso. dB	Intercept Point dBm	SWR In/Out		DC	
		Typ.	Min. 0/50°C	Min. –55/85°C	Typ.	Max. 0/85°C	Typ.	Max 0/85°C	Max. –55/85°C	Typ.	Min. 0/50°C	Min. –55/85°C	Typ.	3rd/2nd Typ.	Typ	Max 0/85°C	Volts Nom.	mA Typ.
A2KW5233	0.5-30																	
Mode 1		18.0	17.0	19.0	0.5	1.0	5.5		6.2	26.0		25.0	32	42/70	1.5	1.6	+15	500
Mode 2		10.0	9.0	11.0	0.5	1.0	5.5		6.2	24.0		21.0	16	42/70	1.5	1.6	–15	40
Mode 3		–1.5	–2.0	–2.0	0.3	1.0	1.5		2.0	15.0		11.0	2	42/70	1.3	1.6	+5	20
A2KW5232	30-100																	
Mode 1		19.5	18.5	19.5	0.5	1.0	5.0		6.0	27.5		26.0	32	42/66	1.5	1.6	+15	530
Mode 2		10.0	9.0	11.0	0.5	1.0	5.5		6.8	25.5		22.0	16	42/66	1.5	1.6	–15	40
Mode 3		–1.8	–2.0	–2.0	0.3	1.0	1.8		2.0	15.0		11.0	2	42/66	1.3	1.6	+5	20
A2KW1331	10-100																	
Mode 1		21.0	20.4	21.7	0.5	1.4	4.7		5.0	26.0		25.0	32	35/57	1.8	1.9	+15	520
Mode 2		9.5	9.0	11.0	0.5	1.0	5.1		5.4	26.0		25.0	16	35/57	1.8	1.9	–15	40
Mode 3		–2.3	–2.5	–2.5	0.3	1.0	2.3		2.5	15.0		11.0	2	35/57	1.3	1.9	+5	20

▲ **Table 1. Specifications of the three selectable gain amplifiers.**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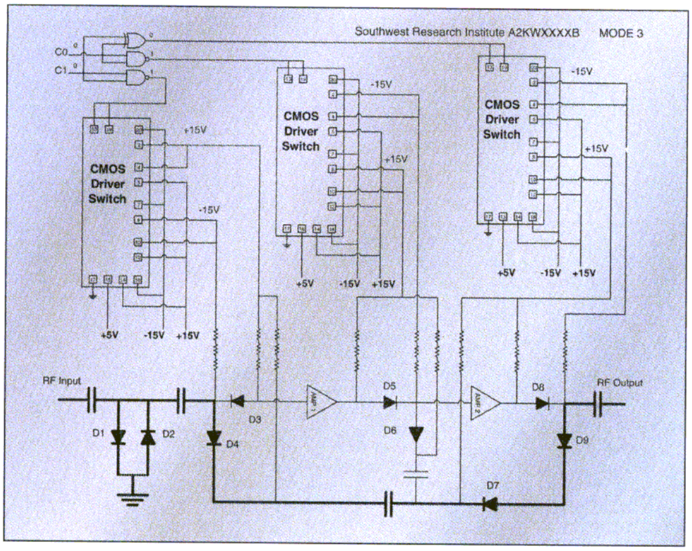

▲ **Figure 4. Mode 3 is the bypass mode, routing the signal around both amplifiers.**

bining state-of-the-art high performance hybrid amplifiers with high dynamic range RF switches and switch drivers, including the digital logic. Each of these designs is incorporated into very small packages devoid of screw-in carriers, requiring high levels of assembly precision and manufacturing capability.

As these selectable gain amplifiers demonstrate, Cougar has expanded its capabilities to include subsystem and multi-function solutions. Their component and subsystem engineering staff can also help with analog and mixed analog-digital solutions of hybrid, MMIC and mixed hybrid-MMIC configurations. Cougar develops designs built upon the strengths of its component expertise to provide practical and efficient subsystem solutions. Its subsystem development offers enhanced functionality in reduced space, improved efficiency, increased reliability, reduced cost, and improved schedules. ■

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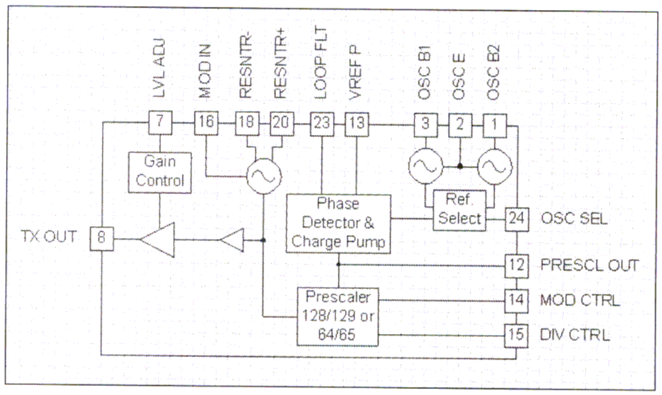
Low Cost UHF/VHF Transmitter ICs

Integrated circuits simplify design of low power FM/FSK or AM/ASK transmitters

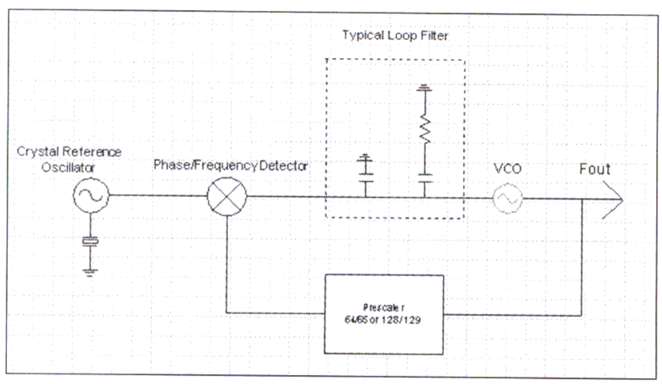
By Alan Nicol
RF Micro Devices

The RF2512 and RF2513 are low-cost monolithic frequency synthesizer and transmitter ICs that provide all the functions necessary to implement low power FM/FSK or AM/ASK transmitter or local oscillator operation in commercial wireless products. The devices can be used in the US 915 MHz ISM band and European 433 MHz or 868 MHz ISM band. Typical applications include wireless security systems, wireless meter reading and wireless data link.

The parts are provided in a 24-pin plastic SSOP package and operate from a 2.2 to 5 volt DC supply. The Optimum Technology Matching™ approach taken by RF Micro Devices in all of its designs leads to the choice of a 15 GHz silicon bipolar process technology for these parts. The process features a f_T of 15 GHz at less than a 3 volts and a 0.5 mA operating point. The only difference between the RF2512 and RF2513 is that the RF2513 has on-chip tuning varactors. The RF2513 is a more cost effective integrated solution but has lower output power, smaller tuning range and higher phase noise than the RF2512.



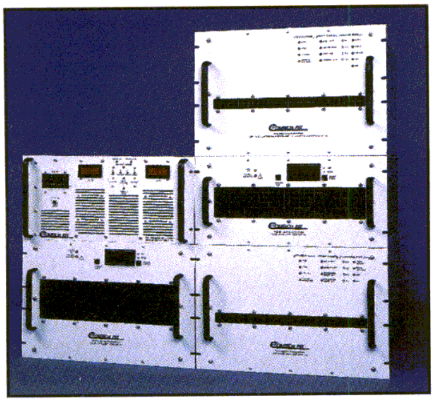
▲ **Figure 1. RF2512 functional block diagram.**



▲ **Figure 2. PLL block diagram.**

Operating features

The RF2512 and RF2513 operate as FM/FSK or AM/ASK transmitters or local oscillators. The integrated VCO, dual modulus/dual divide prescaler and reference oscillator only require the addition of an external crystal oscillator to complete a fully integrated phase-locked loop (PLL) system. A functional block diagram of the RF2512 is shown in Figure 1. A second reference oscillator is available to support two channel applications. In sleep mode, when only V_{cc} is applied, the part draws less than 1 μA. The PLL enable pin powers the VCO and PLL while the TX enable powers the transmitter stages. The level adjust control allows the transmitter output power to be varied over a 15 dB range. The dc transmitter current is also reduced with output power. The level adjust pin



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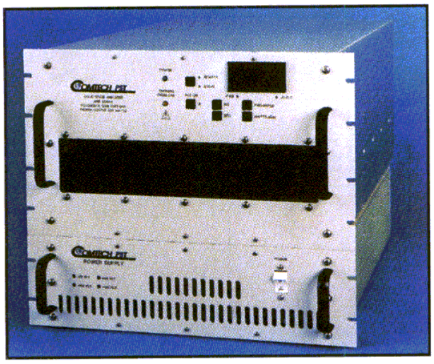

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must be set low if the transmitter output is disabled. The dual modulus/dual divide prescaler provides divider ratios of 64/65 or 128/129. This allows for a certain amount of flexibility in the choice of reference frequency.

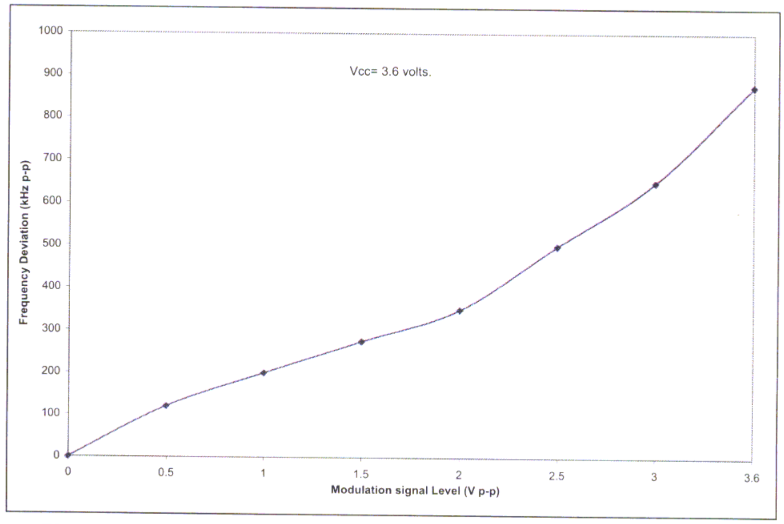
PLL functionality

A block diagram of the PLL section is shown in Figure 2. The VCO design is based on a balanced configuration that utilizes cross coupling between the bases and collectors of a transistor pair. A pair of external inductors and a hyperabrupt varactor diode form the tank resonator for the RF2512. The RF2513 uses on-chip varactors, which are the base-collector junctions of bipolar devices. The base collector junction was chosen because although the base emitter junction has more capacitance, the breakdown voltage of the base-emitter is much lower than the base-collector. The VCO tuning ranges at the ISM band frequencies are shown in Table 1, where the reduced range of the on-chip varactors is evident. A balanced configuration was used instead of a single ended design since it provides differential drive to the prescaler section and has better rejection of unwanted modulation signals. There must be a DC path to V_{cc} on the VCO resonators.

The dual modulus prescaler is implemented using a master/slave flip-flop divider architecture. A current mode phase/frequency detector using D-type flip-flops and charge pumps provide the control signal to the varactors. A 2nd or 3rd order passive loop filter is typically used to set the loop parameters.

Crystal oscillator

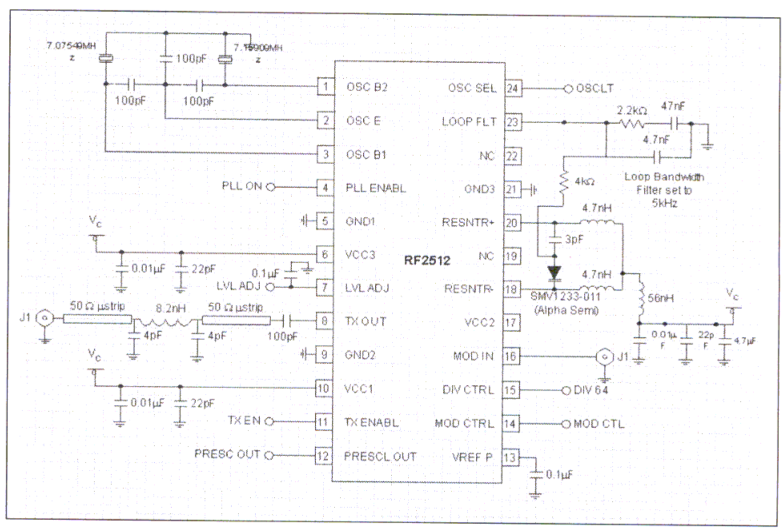
The crystal reference oscillator is a fundamental mode common emitter Colpitts design that operates in a parallel resonant circuit. The crystal is calibrated with a 32 pF load. The on-chip amplifier is an emitter follower, which gives a voltage gain of one. The values of the external feedback capacitors can be adjusted for optimal performance at different frequencies. The capacitors provide most of the phase shift and set the loop gain. The crystal can be replaced by an external source. The signal must be AC coupled and the drive level should be at least 200 mV peak to peak.



▲ **Figure 3. Frequency deviation vs. modulation signal level.**

Frequency (MHz)	VCO Sensitivity (MHz/volt)	
	RF2512	RF2513
915	45	23
868	44	23
433	27	10

▲ **Table 1. VCO tuning ranges at Cvv = 3.6 volts.**



▲ **Figure 4. 915 MHz application schematic.**

Modulation and PLL characteristics

The RF2512 and RF2513 can support wireless applications with data rates up to 1 Mb/s using FSK or ASK digital modulation or linear modulation using FM or AM. The FM/FSK modulation is imparted directly to the VCO using on-chip modulation varactors. The diodes are formed in a similar way to the tuning varactors on the RF2513. The main difference is that the base-collector junction of a smaller bipolar device is used since less capacitance is required. The TX enable and level adjust control can be used to implement AM modulation. The level adjust control and TX enable need to be tied together to provide on-off keying functionality when ASK operation is required. The level adjust control does not provide enough range. The frequency deviation is

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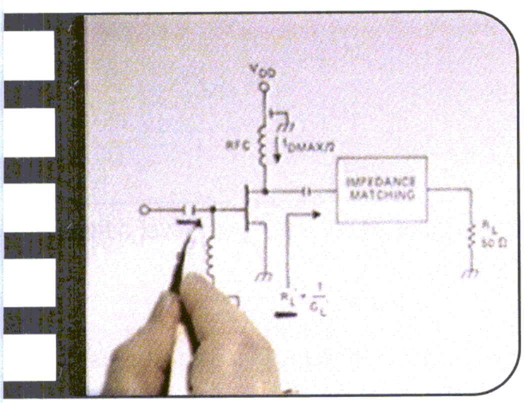

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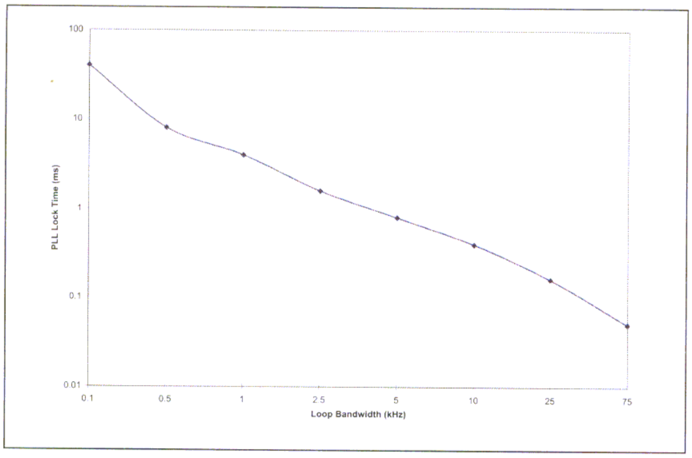
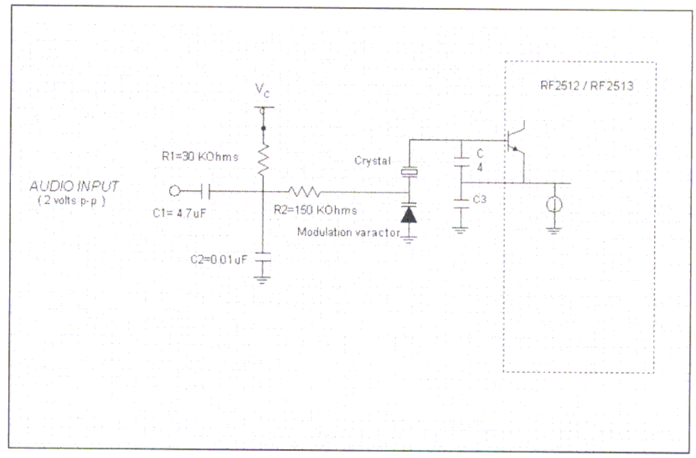
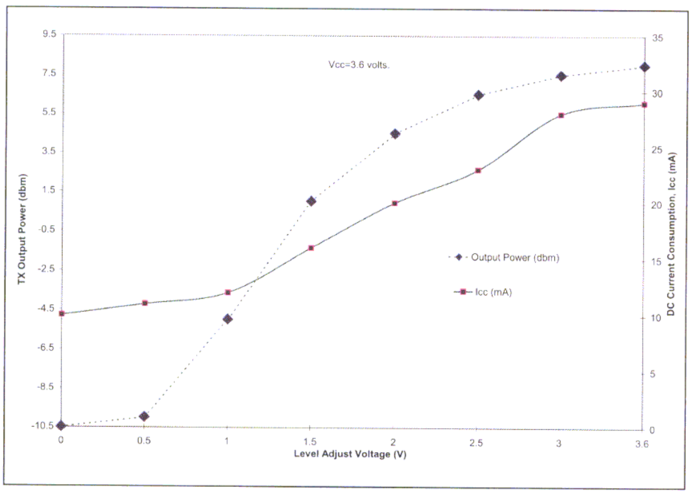


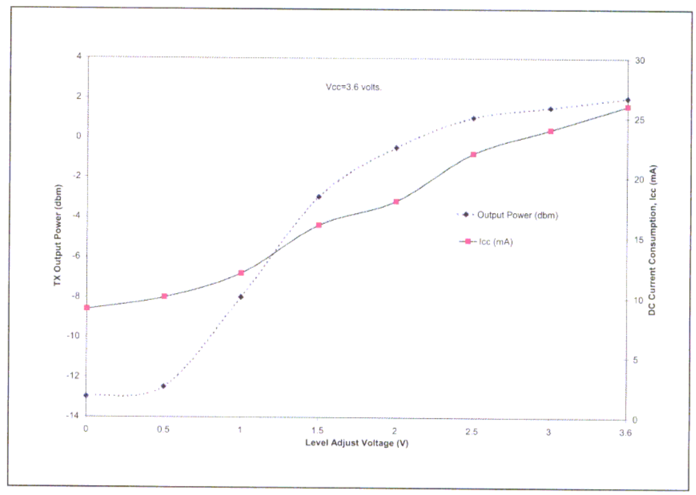
▲ **Figure 5. PLL lock time vs. loop bandwidth.**



▲ **Figure 6. Audio modulation using the reference frequency crystal.**



▲ **Figure 7. Transmitter output power and I_{cc} vs level adjust voltage for the RF2512.**



▲ **Figure 8. Transmitter output power and I_{cc} vs. level adjust voltage for the RF2513.**

proportional to the amplitude of the modulation signal. A typical graph of frequency deviation versus input modulation level for a 40 kHz signal with no DC offset is shown in Figure 3.

Direct modulation is a simple method of applying FM/FSK modulation. However, care must be taken since the modulation rate needs to be higher than the PLL bandwidth otherwise the PLL will track out the modulation. Another situation that can cause these errors is when the modulation data has long strings of ones or zeros. The minimum modulation frequency should always be greater than the PLL bandwidth and a Manchester encoding scheme is recommended to avoid the modulation tracking problem.

The external second order passive filter using capacitors and a resistor as shown in the 915 MHz application schematic typically sets the PLL bandwidth, Figure 4. The series resistor and capacitor set the loop parameters while the single capacitor helps to suppress the reference sidebands. Classical PLL loop analysis can be used to determine the loop component values for a specific loop bandwidth. The graph of PLL lock time versus loop bandwidth is shown in Figure 5.

This graph shows that if direct modulation is used for low frequency modulation, such as audio, then the PLL lock time could become excessively large for certain applications. A solution to this problem is to modulate outside the loop by modulating the reference crystal. A simple implementation with typical component values is shown in Figure 6. Low frequency modulation can then be applied without having a very small loop bandwidth.

The turn on time of the RF2512 or RF2513 is affected by the PLL lock times but mainly by the start-up time of the reference oscillator. The appropriate choice of crystal and prescaler parameters can help reduce the turn-on time. The start-up time is inversely proportional to the crystal oscillator reference frequency. This means that if start-up time is a critical parameter then

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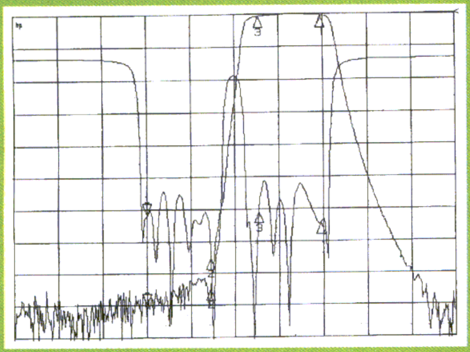

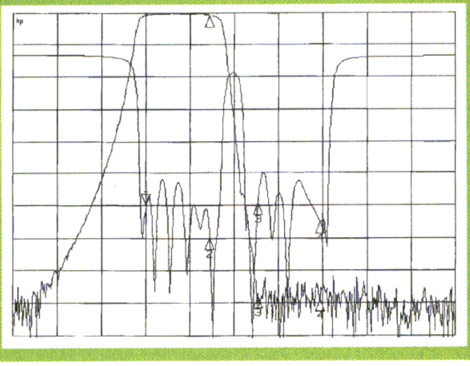

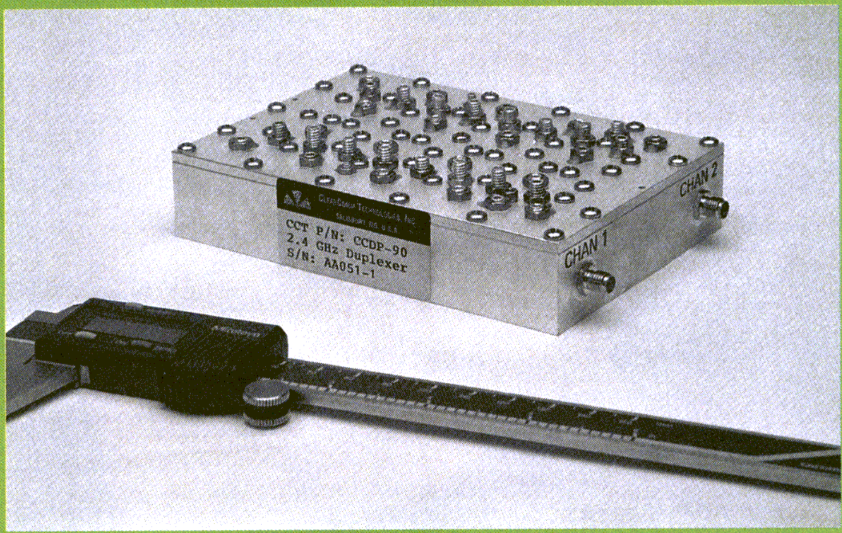

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the lowest possible divider ratio in the prescaler should be used since this would use the highest possible crystal reference frequency. The amplitude of the reference oscillator signal is another factor, which can affect the start-up time. The on-chip phase/frequency detector requires a certain amplitude of reference signal before locking can occur. It is possible to decrease the start-up time by changing the feedback capacitors on the crystal oscillator to increase the reference signal swing. The maximum drive level of the crystal must be taken into consideration, however, when doing this. A start-up time of around 1 ms is achievable with these parts.

Transmitter output characteristics

The RF2512 and RF2513 both consume less than 1 µA in the sleep mode at 3.6 volts and around 10 mA at 3.6 volts and level adjust low. This rises to around 28 µA when the level adjust is brought up to 3.6 volts. The RF2513 draws about 2 µA less than the RF2512 since the drive level from the VCO to the output PA is lower than in the RF2512.

Transmitter output levels and dc current consumption versus level control voltage for the RF2512 and RF2513 at 915 MHz are shown in Figures 7 and 8 respectively. The RF2513 is typically 3 to 5dB lower than the RF2512 with maximum output levels being around +8 dBm for the RF2512 and +2 dBm for the RF2513. The second and third order harmonic levels are typically 23 dBc, therefore the evaluation boards have a low pass filter on the transmitter output. Higher output levels are obtainable when the part is used at lower frequencies, 433 MHz for instance. The phase noise at a 10 kHz offset and a 10 kHz loop bandwidth is –80 dBc/Hz for the RF2512 and –75 dBc/Hz for the RF2513. These two parts are now available in volume at a low cost. ■

Author information

Alan Nicol received his B.S. degree in Electrical Engineering in 1979 from Edinburgh University. From 1983 to 1986, he served as a Microwave Research Assistant at Paisley University, Scotland, where he received his M.Phil. in Microwave Engineering. He currently is a staff engineer for RF Micro Devices in Greensboro, NC. He may be reached by telephone at 336-931-6654.

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Biasing LDMOS FETs for Linear Operation

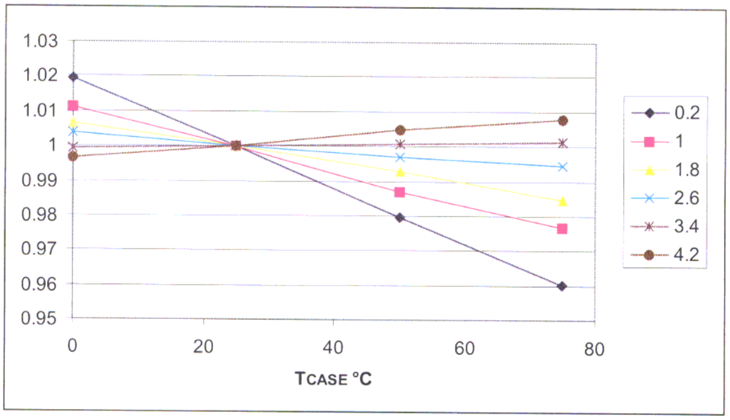
Bias choices determine linearity, gain and efficiency, but also require attention to thermal effects

By Cindy Blair
Ericsson RF Power Products

The new CDMA and WCDMA wireless standards have dramatically impacted the optimization criteria that a wireless base station designer might choose in designing or selecting RF ower amplifiers. Now, more than ever before, linearity is of paramount importance. Lateral DMOS devices have provided superior performance at a reasonable price per watt. RF power amplifier designers must choose the bias point that will give them the best trade-offs between linearity, gain and efficiency. In order to provide these high degrees of linearity under all reasonable conditions, this bias point must be reasonably maintained over time and temperature.

A FET device has three parameters that change with increasing temperature: the gate threshold shifts; g_m drops; and R_{DS} (on) goes down. The combined result of this is shown in Figure 1, displaying a standard typical temperature plot given by most manufacturers of commercial LDMOS devices. The gate bias values shown are normalized to 1 volt at 25° C and depict the bias values needed to maintain the graphed current point over temperature.

At low drain currents, the LDMOS device has a positive temperature coefficient. As the drain current increases, the positive coefficient becomes progressively less positive, until at very high currents it goes negative. This is wonderful protection against thermal runaway, but it makes the designer's job a little more difficult. These plots serve as a good basic guideline, but the designer should be aware that normal process variations might affect these curves slightly.



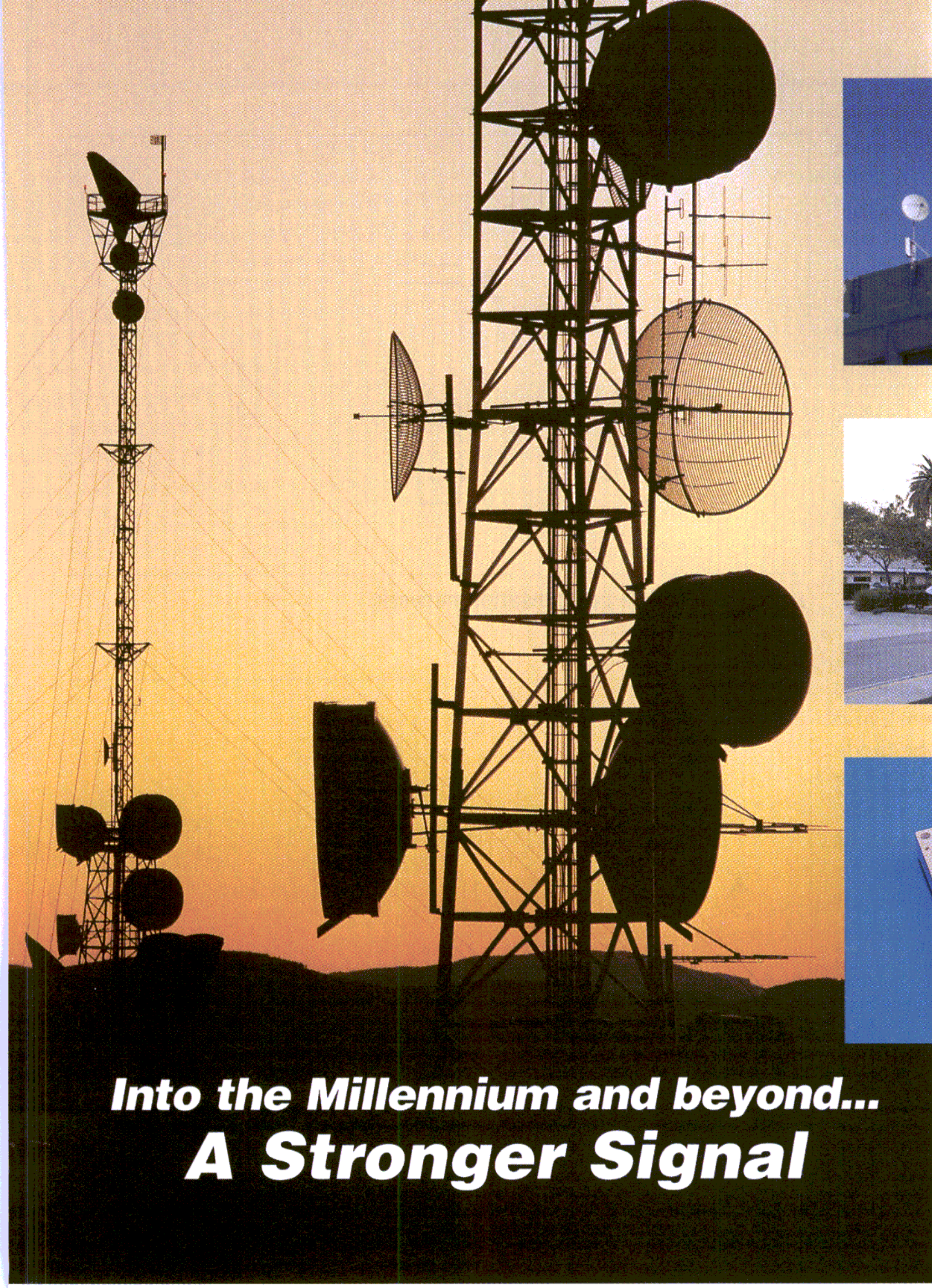
▲ **Figure 1. A standard typical temperature plot given by most manufacturers of commercial LDMOS devices.**

Basic considerations

In all FET bias circuits, it is of critical importance that the gate bias voltage be derived from a very well regulated source. It is also assumed that whatever sense element is used should be thermally linked to the device for which it is providing the correction reference. With the low currents present in the gate circuitry, this can be accomplished simply and very inexpensively. Furthermore, in linear operation, any "corrective" bias circuit will reduce the bias voltage with temperature to retain the same operating point. While at the same time g_m is dropping, this will appear as less amplifier gain at high temperatures. For most applications the gain change is slight, and most system designers provide a system ALC loop, which should mask this effect in the end product.

The diode compensation network

Diode compensation networks (as shown in



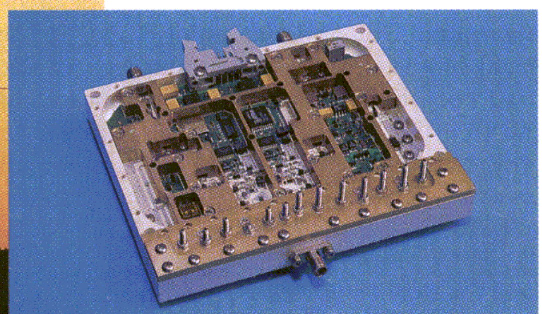


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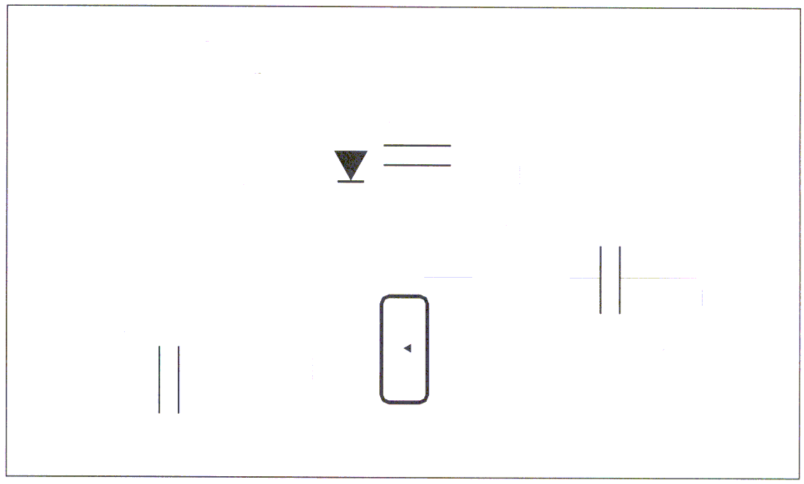
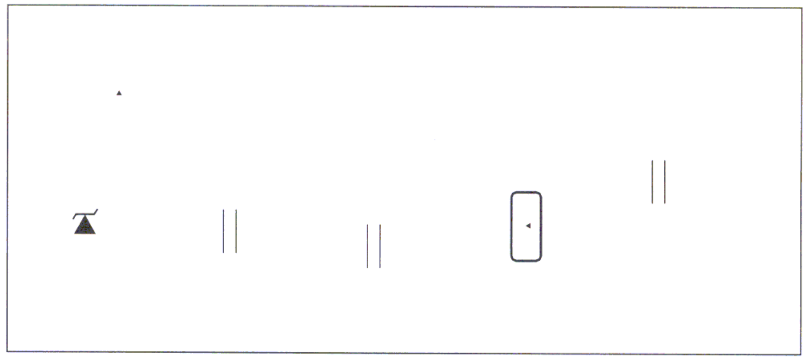
Figure 2) are perhaps the simplest and most widely used bias circuit. They have been used for years in bipolar circuits to prevent thermal runaway. There are excellent application articles available for their use in FET products as well. Operation is very straightforward. As D1 becomes hotter, its enhanced conduction will increase the current through R2, thus altering the divide ratio of R1/R2/D1. Single diode networks are, however, somewhat limited, in that it is more difficult to tailor the compensation curve to a specific device curve without additional circuitry. One may also series two or more diodes to increase the amount of compensation available.



▲ **Figure 2. Diode compensation network.**

The thermal sensor

If the designer desires a bit more sophistication, a thermal sensor (as shown in Figure 3) may be used. This eliminates the guesswork and extra characterization associated with the diode compensator. The designer simply looks up the thermal coefficient of the device at the specific desired operating current, then compares that to the 10 mV/K slope of the temperature sensor (in this case, LM335). A simple op-amp circuit takes care of providing the desired slope conversion. In the sensor shown in Figure 3, R1 can be replaced by automated methods of main bias control for the device (e.g., I_{DQ} set from the system controller). R4 will adjust the amount of temperature slope at the output of IC1. R2, R3, and R5 may be tailored to the bias controls of individual systems. This is a well behaved configuration that is easily modeled in Spice.



▲ **Figure 3. Thermal sensor.**

The "Like material" reference

When the designer wants the mirroring effect in a compensation circuit as close to perfect as possible, there is no substitute for a "like material" reference (as shown in Figure 4). Q2 contains an RF transistor die that is similar in semiconductor processing characteristics to the die contained in Q1, but is much smaller (for this discussion, 1/56 size). Q2 is kept small to maximize system efficiency. The reference device may be any scalable size. R5 and R6 set the gate bias on Q2, typically set so that the drain current is scaled down to the same ratio as the difference in die size. For example, if the normal operating current in Q1 is 2.6 amperes, Q2 would be set to run at 1/56 of that value (assuming a 56:1 Q1/Q2 size ratio). As Q2 changes temperature, the change in current causes the drain voltage of Q2 to change. This voltage feeds the correction input of the op-amp Summing circuit above.

▲ **Figure 4. "Like material" reference.**

A second, less obvious advantage to this topology is compensation for hot electron effects. High frequency (short gate length) lateral DMOS devices exhibit over time a gate oxide charging phenomenon. This oxide charging leads to an upward shift in threshold voltage in the enhancement mode device. For example, assume that the gate bias of an FET is set so that the quiescent drain current is 1 A. If the FET is operated for 20 years and the quiescent current is re-measured as 0.95 A, the FET has drifted 5 percent, due to this oxide charging effect. One ampere was the value of quiescent current

that provided optimal linearity performance; consequently the linearity performance (and small signal gain) of our device has degraded over time. These effects are minimized if the gate bias is increased over time to compensate for the charging effect. In the ideal case, V_{DD}* will be above V_{DD} so that with the current drawn the actual V_{DD} of Q2 will approximate V_{DD} on Q1.

One very important assumption for this is that both transistors are "new" when installed. (On a mass production base station line, this will not be an issue.) When properly implemented, the "like material" reference compensates for many drift and thermal effects. This is certainly not a perfect solution, but it allows offsetting negative characteristics to subtract; even if material types are not perfectly matched, you are still subtracting errors.

Although device manufacturers are continually reducing this drift effect, it will be a factor for some time to come. The "like material" reference eliminates the need for burn in, but requires the amplifying and compensating devices be of identical "age." This may add substantial cost to any field replacement of devices.

The correction circuits presented here are generic in nature, and may be modified to individual system requirements. This paper is intended as a general reference only. Equations for the sum/difference circuits are available in the references. Most of these correction circuits are easily modeled. If a higher degree of manufacturing ease is needed a custom integrated circuit could be produced for method 1 or 2.

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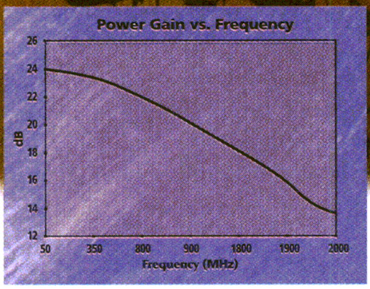

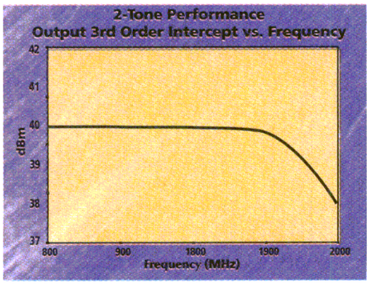

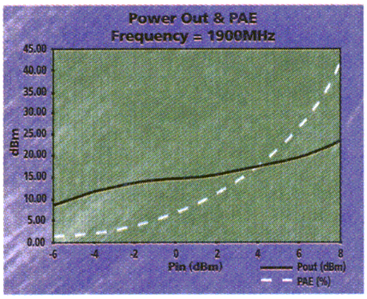

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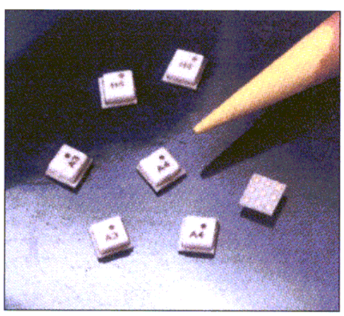

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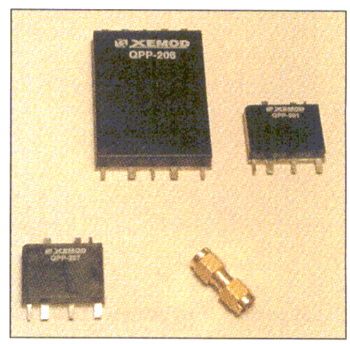

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Plexus Corp. has developed an inexpensive radio module which can operate in any of the three U-NII bands. The radio is capable of operating at 20 Mbits/s in both directions over ten miles line-of-sight or several hundred feet within a building. The first in a series of GaAs

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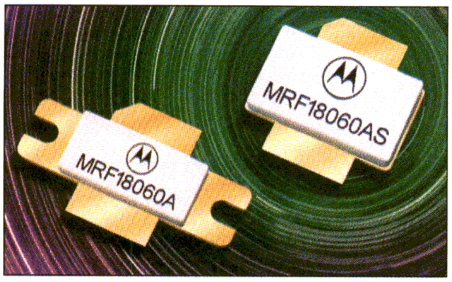

cations from 1.8 to 2.0 GHz. Using fourth generation (HV4) device technology, the family includes models offering power output from 60 watts to 90 watts CW. The MRF18060A/MRF18060AS (60 watts, 1805 to 1880 MHz) is priced at $75.00 each in quantities of 10,000.

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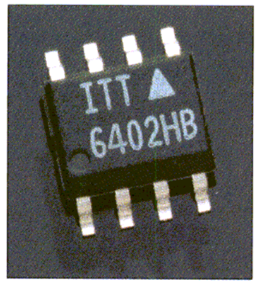
GaAsTEK announces a GaAs power amplifier for the 5 to 6 GHz range. The ITT6402HB is a two-stage PA featuring 30 dBm power output at 7 volts, and 28 dBm at 5 volts DC power, with 20 dB typical small signal gain. Power added efficiency is 39 percent at P_{1dB}. The amplifier is packaged in a standard SOIC-8 with a simple stub-tuned microstrip output match.



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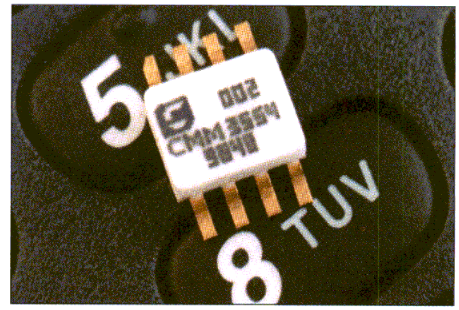

put at 3.5 GHz. An SO-8 plastic power package keeps cost low and maintains electrical and thermal performance. The device is biased to meet the linearity requirements of W-CDMA. In volume quantities, the price is $10.00 each, as individual units or in tape-and-reel packaging. Evaluation boards are available.

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Telemetry Transceiver Operates in New European Unlicensed Band

A versatile, credit-card-size transmitter-receiver to the new 868-870 MHz license exempt telemetry band has been introduced by Wood & Douglas Ltd. The SX850 transceiver provides either analog or digital modulation modes for a wide variety of short range applications in on-site monitoring, process control, data exchange, control systems and security.

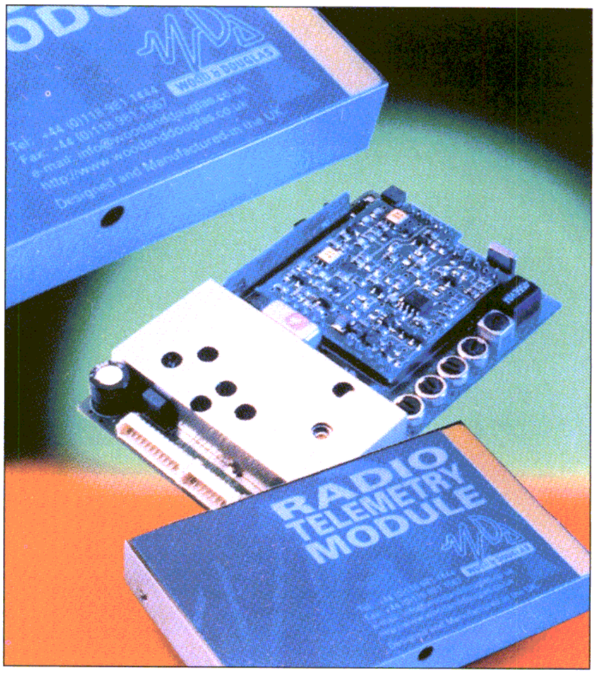
The SX850 is type approved to the European EN 300 220 (telemetry) and the more demanding ETS 300 086 (mobile radio) specifications. Manufacturers of systems requiring short range data transmission can add that capability easily with this transceiver module.

The pan-European 868-870 MHz band provides an alternative to the UK's crowded 418, 433 and 458 MHz unlicensed bands, which have become prone to interference after the introduction of the TETRA trunked radio allocation around 420 MHz.

The SX850 uses a TCXO referenced synthesizer that can be tuned to any channel in the band, programmable by serial interface. It has a low power (5 mW) output for short range use, plus a high power mode (500 mW) for longer range telemetry needs. With a 7.2 VDC supply, power consumption is 75 mA while receiving, and typically 450 mA when transmitting in the high power mode.

Analog and digital modulation modes are supported, along with an optional matching Gaussian minimum shift keying (GMSK) modem for high data rates up to 4800 or 9600 baud. Other features include a received signal strength indicator (RSSI) and squelch outputs.

The SX850 has also received ETS 300 683 electromagnetic compatibility approval.



▲ **The SX850 provides telemetry connection in the 868-870 MHz European unlicensed band.**

For more information, contact:

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System Planning Software Helps Determine Electrical and Mechanical Requirements

Andrew Corporation has introduced the *Powertools* telecommunications system planning CD-ROM, containing an integrated, easy-to-use suite of seven RF system planning tools. *Powertools* includes programs to speed the RF design process for cellular, PCS, DTV and NTSC broadcast, terrestrial microwave and other wireless communication systems. The CD-ROM contains the following programs.

Andrew Antenna System Planner — for the design of antenna systems, checking components for frequency compatibility, calculating insertion loss and VSWR for selected cables and generating a bill of materials.

Andrew Broadcast System Planner — to help consultants and engineers configure digital and analog television broadcast systems, providing printed antenna patterns, power analysis data, electronic files of antenna patterns for propagation studies. It has data on 35 popular broadcast transmission lines.

Andrew Microwave System Planner — for complete terrestrial microwave system design, helping to select the components for a microwave site and generating a bill of materials.

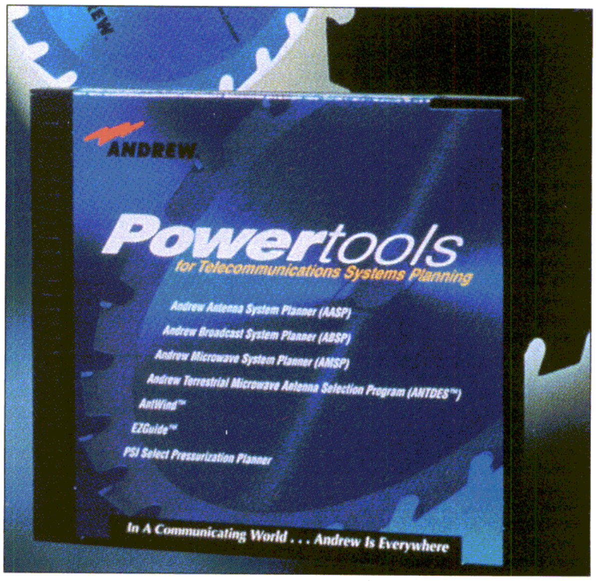
ANTDES™ — an antenna selection program detailing the entire Andrew line, including printed radiation patterns.

AntWind™ — a program to simplify the calculation of wind forces applied to towers and supporting steelwork by the installed Andrew parabolic antenna.

ezGuide™ — software to guide the ordering and design process for microwave transmission lines, with products pictured on-screen and diagrammed for reference.

PSI Select Pressurization Planner — for complete information on Andrew pressurization products and accessories and performing system calculations and dehydrator selections.

The individual software tools can also be downloaded from the company's Web site. ■



▲ **The Powertools CD-ROM from Andrew Corporation aids in the design and specification of transmission systems.**

For more information, contact:

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Tel: 708-349-3300 or 800-255-1479
Fax: 708-349-5222
Internet: www.andrew.com

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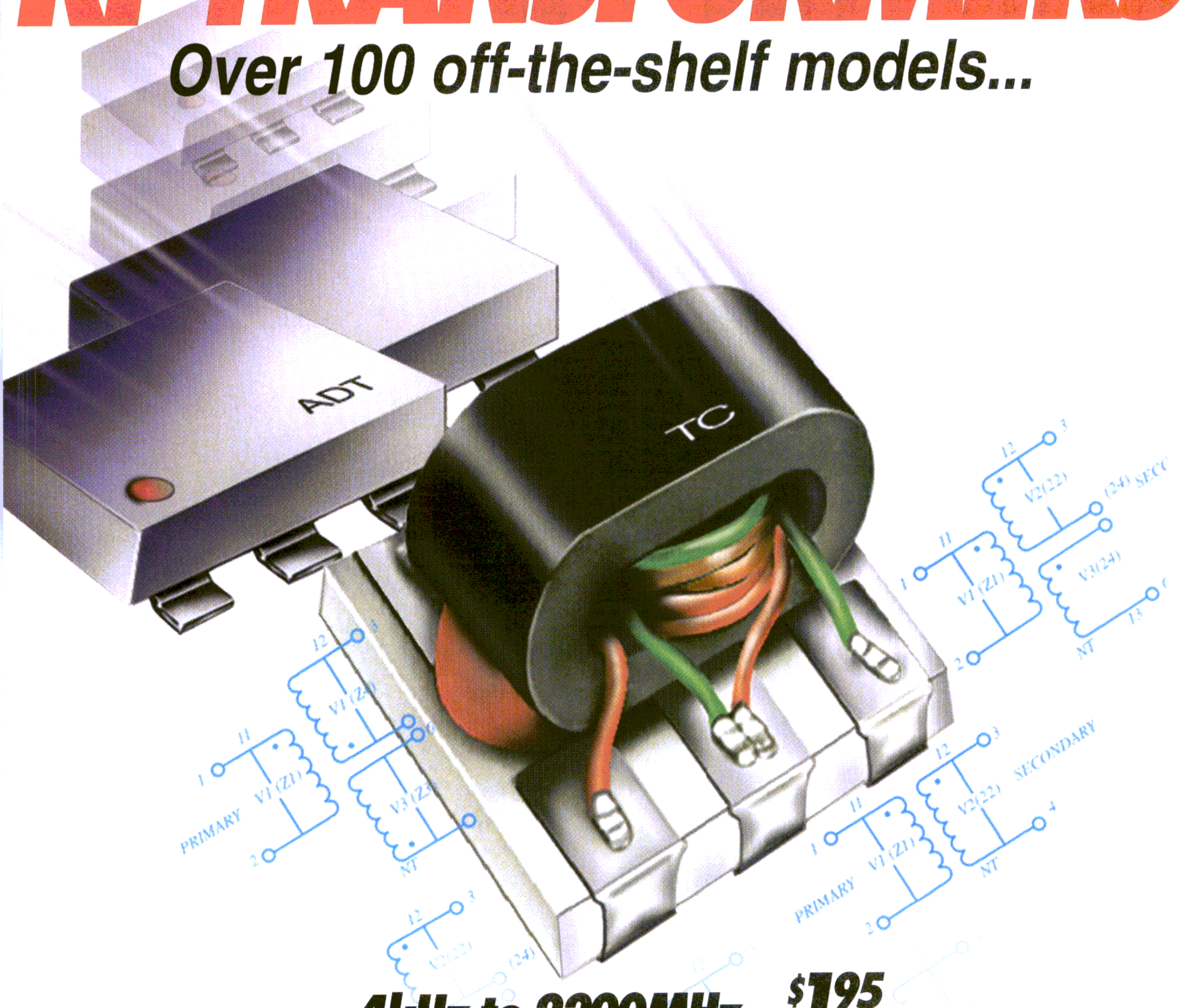
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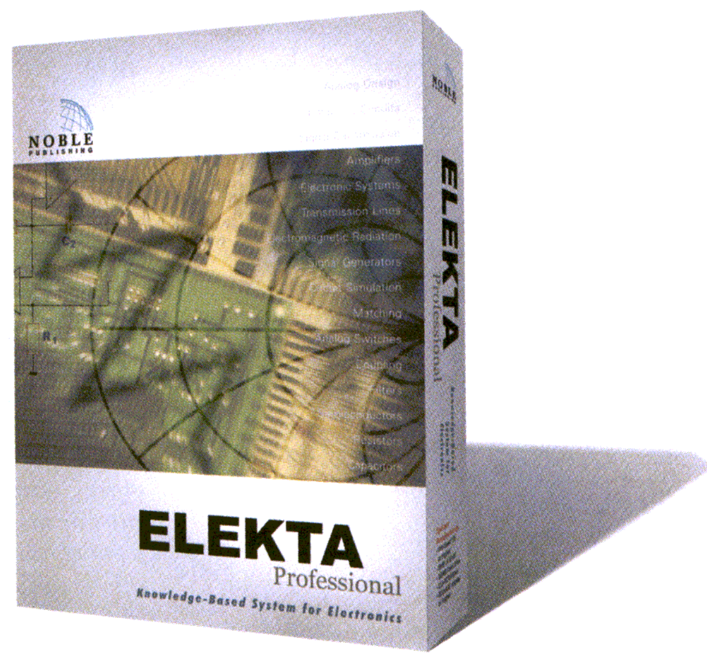

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Circle 39

Repeaters Can Solve Wireless System Growth Problems

By David McKay
CI Wireless Inc.

Many wireless system engineers are still unaware of the capability of repeaters for the extension of wireless communications networks. Traditionally, repeaters have simply been "bi-directional RF amplifiers" used to extend the reach of rural base stations, or to allow in-town base stations to cover areas shadowed by hills, valleys or buildings.

Here at CI Wireless, we have added two major new dimensions to repeater operation — fiber optic interconnection using our EkoCel™ repeaters, and adaptive interference cancellation (AIC) in our recently-released Eko-BTS™ products.

The low loss characteristics of fiber allow a carrier to centralize base station radios and distribute radio signals to desired coverage areas where the size or cost of deploying additional base stations is not feasible. These repeaters are designed to enhance coverage and capacity in airports, malls, office campuses, industrial facilities, dense metropolitan areas, tunnels and highways. The system performance is compatible with CDMA, TDMA/IS-136 and PCS1900/GSM, and is forward-compatible with 3G/cdmaOne and W-CDMA. The fiber link not only carries radio traffic but also includes alarm and control signals for the base station and system diagnostics that can be accessed anywhere in the system.

Clearly, a repeater is no longer just antennas, amplifiers and connecting cables. The technology between the base station communication end of the system and the remote transmitters/receivers has evolved to a new, more capable level. Nowhere is this more evident than in our new Eko-BTS products using AIC technology. CI Wireless' implementation of DSP-based technology in the interconnecting link has been accomplished with the lowest possible time delay in the signal path, with the added ability to select only the desired base station transmission for rebroadcast. With these advantages, our repeaters can have higher power, since AIC achieves much greater input/output isolation by cancelling out the repeaters' own transmissions. Self-interference, which formerly limited the power and range of a repeater, is reduced by 30 dB or more.

It is also important to note that AIC technology has been implemented with full compatibility with the major wireless protocols. The Eko-BTS repeaters can be used with AMPS, CDMA, TDMA, GSM and IDEN. CI Wireless is dedicated to offering the most advanced wireless distribution products available. The system can be selective down to the PN code for CDMA transmissions, or channelized to minimize the problem of time delay in TDMA systems.

David McKay is president of CI Wireless Inc. He has more than 30 years experience in the wireless industry, from equipment design, operations and product marketing to business development, product strategy and hardware/software specification and development. Prior to the launch of CI Wireless in 1997, McKay created, staffed and led a new wireless division at Ortel Corp. He may be reached at CI Wireless Inc., 1211 Ira E. Woods Ave., Grapevine, TX 76051; tel: 817-416-0583; fax: 817-488-1949. The company's Web site is www.ciwireless.com.

The business value of repeaters

Repeaters are far more economical than a new base station when coverage must be extended beyond the range of existing base stations. With a growing user base, wireless communications providers want service extensions that increase coverage, not just distance. Coverage expansion can be inside buildings where the structure shields signals from the nearest base stations or reduces them to unreliable levels. It may also be underground, in congested areas or anywhere where better signals are required.

Repeaters also improve system capacity. This is not obvious, since no more base stations are added to the system! Repeaters allow each base station to operate closer to its capacity by filling "blind spots" in its coverage. More locations are served by the same number of base stations by eliminating unusable capacity.

To conclude, CI Wireless is totally focused on the distribution of RF signals, with products for cellular and PCS, SMR/ESMR, paging, LMDS, MMDS and conventional two-way radio. Our products are the tools our customers need to get the coverage they want. This is our expertise and that's what we will continue to do for the wireless industry. ■